SPREAD SPECTRUM RECEIVER
ARCHITECTURES FOR MOBILE CHANNELS
SUBJECT TO MULTIPATH FADING

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A thesis submitted for the degree of
Doctor of Philosophy

University of Edinburgh
1992
To my wife Lorraine,
and also my daughter Joanne who has been a most enjoyable distraction.
ABSTRACT

Recently there has been considerable interest in spread spectrum techniques for use in commercial communication systems. One particular area where there may be considerable benefit is in mobile radio systems. Many of the difficulties presented by mobile radio communications result from the multipath fading which is present on the channel. This can lead to large reductions in the available channel capacity if measures are not taken to combat its effects.

The use of diversity is highly effective in ameliorating the effects of fading. This research is concerned with an anti-multipath receiver architecture, the RAKE receiver, which uses the inherent multipath diversity of the received spread spectrum signal. As a precursor to this work we describe a spread spectrum correlator architecture, the serial-parallel receiver, which is capable of resolving the multipath diversity signals which are subsequently used in the RAKE receiver.

A digital serial-parallel correlator is developed which has negligible implementation loss, fast PN code synchronisation, and can provide very large processing gains. The performance obtained from this spread spectrum receiver architecture is demonstrated using a DSP based implementation, and is shown to improve upon previous analogue implementations in most respects. It is particularly attractive for mobile communications since it can resolve the multipath signal components without the need for a large time-bandwidth product matched filter, or multiple active correlators.

The performance of an adaptive RAKE receiver, using simple alpha tracker profile estimation filters, is analysed for a simulated UHF mobile channel (vehicular and hand-held) in a typical urban environment. Bit error rate graphs indicate that the performance approaches the theoretical upper bound when the signal fading is slow relative to the bit rate of the system (e.g. hand-held operation). For an error rate of $10^{-3}$, up to 20dB of signal power may be saved by the use of 6 orders of multipath diversity. For a fast moving mobile the results are poorer due to the much larger phase lag on the channel estimate. This leads to an irreducible bit error rate, but can still provide a worthwhile improvement in performance for just a few orders of diversity.

Improvements in the performance for the fast fading signal will be possible if the phase lag on the channel estimate can be reduced or eliminated. It is suggested that a forward linear prediction of the channel's response can be made based on the past responses and a spectral estimate of the fading characteristics. We show that it is possible to form a good spectral estimate of the fading signal, even for poor SNRs (e.g. 0dB), using an efficient multirate FFT technique.
ACKNOWLEDGEMENTS

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I would like to thank Steve Mowbray for many lively and useful discussions, Ed Warner for reading the draft of this manuscript, and members of the Signal Processing Group for their useful comments.
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<td>ACF</td>
<td>Auto-Correlation Function</td>
</tr>
<tr>
<td>ADC</td>
<td>Analogue-to-Digital Converter</td>
</tr>
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<td>AMF</td>
<td>Analogue Matched Filter</td>
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<tr>
<td>AWACS</td>
<td>Airborne Warning And Control System</td>
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<td>AWGN</td>
<td>Additive White Gaussian Noise</td>
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<tr>
<td>BER</td>
<td>Bit Error Rate</td>
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<tr>
<td>BS</td>
<td>Base Station</td>
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<tr>
<td>CCD</td>
<td>Charge Coupled Device</td>
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<tr>
<td>CCF</td>
<td>Cross-Correlation Function</td>
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<tr>
<td>CDF</td>
<td>Cumulative Distribution Function</td>
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<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>CELP</td>
<td>Codebook Excitation Linear Prediction</td>
</tr>
<tr>
<td>CODORAC</td>
<td>Coded Doppler Radar Command</td>
</tr>
<tr>
<td>COST</td>
<td>European Cooperation in the Field of Scientific and Technical Research</td>
</tr>
<tr>
<td>CW</td>
<td>Continuous Wave</td>
</tr>
<tr>
<td>DAC</td>
<td>Digital-to-Analogue Converter</td>
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<td>DFT</td>
<td>Discrete Fourier Transform</td>
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<tr>
<td>DMF</td>
<td>Digital Matched Filter</td>
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<tr>
<td>DPC</td>
<td>Differential Phase Combining</td>
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<tr>
<td>Abbreviation</td>
<td>Description</td>
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<td>DPSK</td>
<td>Differential Phase Shift Keying</td>
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<td>DS</td>
<td>Direct Sequence</td>
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<td>DSP</td>
<td>Digital Signal Processor</td>
</tr>
<tr>
<td>DSSS</td>
<td>Direct Sequence Spread Spectrum</td>
</tr>
<tr>
<td>EGC</td>
<td>Equal-Gain Combining</td>
</tr>
<tr>
<td>EPROM</td>
<td>Erasable Programmable Read Only Memory</td>
</tr>
<tr>
<td>FDM</td>
<td>Frequency Division Multiplexed</td>
</tr>
<tr>
<td>FDMA</td>
<td>Frequency Division Multiple Access</td>
</tr>
<tr>
<td>FFH</td>
<td>Fast Frequency Hop</td>
</tr>
<tr>
<td>FFT</td>
<td>Fast Fourier Transform</td>
</tr>
<tr>
<td>FH</td>
<td>Frequency Hop</td>
</tr>
<tr>
<td>FIR</td>
<td>Finite Impulse Response</td>
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<td>FM</td>
<td>Frequency Modulation</td>
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<td>FSK</td>
<td>Frequency Shift Keying</td>
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<tr>
<td>GPS</td>
<td>Global Positioning System</td>
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<tr>
<td>GSM</td>
<td>Groupe Speciale Mobile (or Global System for Mobile Communications)</td>
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<tr>
<td>HF</td>
<td>High Frequency</td>
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<tr>
<td>IBER</td>
<td>Irreducible Bit Error Rate</td>
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<td>IIR</td>
<td>Infinite Impulse Response</td>
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<td>ISI</td>
<td>Intersymbol Interference</td>
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<tr>
<td>JPL</td>
<td>Jet Propulsion Laboratory</td>
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<tr>
<td>Acronym</td>
<td>Definition</td>
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<tr>
<td>JTIDS</td>
<td>Joint Tactical Information Distribution System</td>
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<tr>
<td>LAN</td>
<td>Local Area Network</td>
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<tr>
<td>LPD</td>
<td>Low Probability of Detection</td>
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<tr>
<td>LPI</td>
<td>Low Probability of Intercept</td>
</tr>
<tr>
<td>MAC</td>
<td>Multiply and Accumulate</td>
</tr>
<tr>
<td>MACD</td>
<td>Multiply and Accumulate with Data Move</td>
</tr>
<tr>
<td>MIT</td>
<td>Massachusetts Institute of Technology</td>
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<td>MRC</td>
<td>Maximal-Ratio Combining</td>
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<td>MS</td>
<td>Mobile Station</td>
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<td>NOMACS</td>
<td>Noise Modulation and Correlation System</td>
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<tr>
<td>PAMF</td>
<td>Programmable Analogue Matched Filter</td>
</tr>
<tr>
<td>PC</td>
<td>Personal Computer</td>
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<tr>
<td>PCM</td>
<td>Pulse Coded Modulation</td>
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<td>PCN</td>
<td>Personal Communications Network</td>
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<tr>
<td>PG</td>
<td>Processing Gain</td>
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<td>PN</td>
<td>Pseudo-Noise</td>
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<tr>
<td>PSD</td>
<td>Power Spectral Density</td>
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<td>PSK</td>
<td>Phase Shift Keying</td>
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<td>RAKE</td>
<td>This is not an acronym</td>
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<td>RF</td>
<td>Radio Frequency</td>
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<tr>
<td>RMS</td>
<td>Root Mean Square</td>
</tr>
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<td>Description</td>
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<tr>
<td>SAW</td>
<td>Surface Acoustic Wave</td>
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<tr>
<td>SCW</td>
<td>Scrambled Continuous Wave</td>
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<tr>
<td>SFH</td>
<td>Slow Frequency Hop</td>
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<td>SIK</td>
<td>Sequence Inversion Keying</td>
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<tr>
<td>SNR</td>
<td>Signal-to-Noise Ratio</td>
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<td>SPR</td>
<td>Serial-Parallel Receiver</td>
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<tr>
<td>STDCC</td>
<td>Swept Time-Delay Cross-Correlation</td>
</tr>
<tr>
<td>TACS</td>
<td>Total Access Communications System</td>
</tr>
<tr>
<td>TDM</td>
<td>Time Division Multiplexed</td>
</tr>
<tr>
<td>TDMA</td>
<td>Time Division Multiple Access</td>
</tr>
<tr>
<td>TH</td>
<td>Time Hop</td>
</tr>
<tr>
<td>TOA</td>
<td>Time of Arrival</td>
</tr>
<tr>
<td>UHF</td>
<td>Ultra High Frequency</td>
</tr>
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<td>US</td>
<td>Uncorrelated Scattering</td>
</tr>
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<td>VHF</td>
<td>Very High Frequency</td>
</tr>
<tr>
<td>VLSI</td>
<td>Very Large Scale Integration</td>
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<td>WHYN</td>
<td>Wobbulated Hyperbolic Navigation</td>
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<tr>
<td>WSS</td>
<td>Wide-Sense Stationary</td>
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<tr>
<td>WSSUS</td>
<td>Wide-Sense Stationary Uncorrelated Scattering</td>
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Chapter 1

INTRODUCTION

1.1 Background

Communication systems have come a long way since the days of telegraphy, Bell's first telephone, and Marconi's early radio broadcasts. The ease with which we can use modern communication systems tends to obscure the underlying technology and complexity.

Since the early 1980s there has been a large amount of investment in upgrading public telephony networks. In the UK all of the fixed network backbone is now digital and the use of optical fibre is widespread. Many of the considerable advances in the public switched telephone network (PSTN) have, however, been introduced virtually unnoticed. One fairly visible change in modern telephony, over the past decade, has been the emergence of mobile cellular telephones [1]. The penetration of this technology still remains fairly low in relation to its potential. This situation is set to change rapidly with the introduction of lower cost personal communication networks (PCNs) [2].

Mobile communications must use radio frequency (RF) transmission for the links with the mobile stations (MS) so that they may roam freely. RF bandwidth is a precious resource and must be used efficiently. The current bandwidth allocation cannot continue to sustain the growing demands now being placed on it for mobile communications. The UK's current analogue frequency division multiplexed (FDM) system, the total access communication system (TACS), operates around 900MHz in ultra high frequency (UHF) radio band. Additional bandwidth is not readily available and hence it must be utilised in a most efficient manner in order to maximise its communication potential. The available frequency spectrum must be divided into a suitable number of channels for use by active subscribers. In cellular systems these frequency bands are reused as often as possible in a close geographical proximity in order to maximise the capacity.
The new pan-European Groupe Speciale Mobile (GSM) [3] cellular system addresses some of these issues and uses an efficient time division multiple access (TDMA) scheme. The demand for additional services and the ever increasing number of subscribers will require further expansions of the network. One of the problems in upgrading such systems is that we must live with the legacies of the previous ones for some time. These networks, once established, cannot be changed hastily. New systems or networks must either coexist, or be backwards compatible, with the previous technology for some time.

It has been suggested that spread spectrum code division multiple access (CDMA) cellular systems [4] can be made more spectrally efficient than any of the systems currently used [5]. It is also possible for them to share the bandwidth with existing narrowband subscribers [6]. There are many other advantages which can be obtained by using spread spectrum. However many problems still exist such as capacity limitations due the co-channel interference of non-ideal spreading codes [7].

Much of the recent interest in spread spectrum systems has centred on the multiple access aspects of these systems. In the mobile radio environment we must also consider the effect which the multipath fading channel will have on the coding and modulation scheme used, and hence the receiver performance. The current TACS narrowband system uses analogue frequency modulation (FM), and so can tolerate these effects. TDMA systems, such as GSM, use a high symbol rate and so the multipath dispersion causes severe intersymbol interference (ISI). Equalisation techniques are normally employed in such systems. This requires training sequences to be included within the signalling timeslots. Spread spectrum systems are inherently capable of resolving the multipath profile of the channel. This is a useful feature which can subsequently be used to equalise the channel, and so improve the receiver signal-to-noise ratio (SNR) and shorten power fade durations.

1.2 Research Areas

In this discourse we wish to examine methods by which we can best limit the transmitter power. This is especially important in cellular environments since it helps to minimise
the physical size of the transceiver, it limits co-channel interference, and also improves the frequency reuse efficiency. In operation, CDMA cellular systems must have effective power control management [8] and hence the mean received SNR can be maintained at a fairly constant rate. This can also be beneficial for secure spread spectrum communication systems which may require the signal to be hidden in the ambient noise.

The specific areas of the system which we examine are associated with anti-multipath spread spectrum receiver designs, and the design of the despreading pre-processors which are complementary to their function. The research programme has been divided into two distinct parts. The first investigates architectures for despreading correlation receivers which have minimal complexity, fast synchronisation, and are compatible with anti-multipath post-processors. The second part looks at the implementation of an adaptive channel matched filter to combat the time-variant multipath fading.

![Figure 1.1 - Receiver Block Diagram](image)

The two stage receiver can be viewed as two matched filters, figure 1.1. The first correlation stage is a filter matched to the transmitted spread spectrum coded waveform. The second stage is the adaptive channel matched filter which is constructed using estimates of the communication channel impulse response. The net result is ideally a matched filter receiver which is optimally matched to the received waveform.

It is considered to be beyond the scope of this research programme to investigate a complete spread spectrum system. We concern ourselves predominantly with a single user communication channel. We further simplify matters by assuming that we have an ideal receiver front end which provides us with a complex baseband digital signal. In other words, we will concentrate our efforts on the digital signal processing requirements within the receiver.
CDMA co-channel interference components can be represented by independent random variables. The effect can be considered as a loss value which represents the increase in SNR required to overcome this interference. If we assume this interference to be Gaussian (by applying the central limit theorem for multiple users) we may instead simply add additional noise power to the effective received signals. Since we are only concerned with the spread spectrum channel, and not the overall system, we do not aim to provide numerical values for this SNR loss or noise power. The results obtained are exclusively for a single simplex link. However it is possible that CDMA performance could be estimated using a suitable adjustment in the received SNR.

Direct sequence spread spectrum will be considered since it minimises the system complexity, works well with anti-multipath techniques, and limits the transceiver power requirements. The majority of the complexity of the proposed system is contained within the receiver, and so we do not consider the transmitter which is uncomplicated in comparison. A summary, table 1.1, shows the primary system requirements and some of the target parameters which have been selected for this study.
<table>
<thead>
<tr>
<th>System Summary</th>
</tr>
</thead>
<tbody>
<tr>
<td>Low transmitted power</td>
</tr>
<tr>
<td>Physically small</td>
</tr>
<tr>
<td>Rapid synchronisation</td>
</tr>
<tr>
<td>Mobile operation</td>
</tr>
<tr>
<td>(max. velocity) 120km/hr</td>
</tr>
<tr>
<td>Carrier frequency 900MHz</td>
</tr>
<tr>
<td>Minimum data rate 4.8kbit/s</td>
</tr>
<tr>
<td>Mean received SNR &gt; -20dB</td>
</tr>
<tr>
<td>Bit error rate &lt; 10^{-3}</td>
</tr>
</tbody>
</table>

Table 1.1 - System Requirements Summary

1.1 Thesis Structure

The following chapter provides the reader with an introduction to spread spectrum systems and briefly documents the historical development of spread spectrum techniques in communication systems. This chapter should provide the reader with a basic understanding of spread spectrum principles and the reasons for their deployment in communications. It also provides a context for this current research work in relation to previous developments.

Chapter 3 concisely introduces the concepts of correlation in spread spectrum receivers. Serial, parallel, and the novel serial-parallel, receiver architectures are described. The development of a new digital serial-parallel correlator is documented and the principles and performance of this design is proven using a DSP demonstration system. The use of dedicated and general purpose DSP hardware for implementing these designs are discussed, and the possibility of using virtual DSP architectures for reconfigurable
receivers is introduced. The serial-parallel receivers described have been designed so that they are compatible with the anti-multipath techniques described in the following chapters.

In Chapter 4 we look at the RAKE receiver operation. This is the channel matched filter used to combat multipath fading losses. The optimal maximal-ratio combining of the resolved multipath components used by the RAKE receiver is detailed. We also take a brief look at alternative sub-optimal RAKE receiver designs. The UHF mobile radio channel is discussed and a general model is presented based upon previous research work on channel sounding and analysis. A software based channel model is developed based on these studies and is used in the following system performance simulations. Two mobile scenarios, one for a hand-held mobile, and the other for a fast moving vehicle are built into the software model. The operation and shortcomings of this channel simulator are examined, and the output results are illustrated and compared with measured data.

Chapter 5 is concerned with the adaptive RAKE receiver. First of all we describe how to obtain a reliable estimate of the channel impulse response when it is subjected to degradation by noise. This estimate is applied to the adaptive RAKE receiver to assess the improvement in performance which can be gained through the use of the channel matched filter. A simple mathematical model can be used to predict, first of all, the upper bound on the RAKE performance (assuming a perfect channel estimate). This model is modified to take account of the imperfections in the practical channel estimate. The predicted and ideal performances are compared with Monte Carlo bit error rate (BER) simulation results, and the robustness of the receiver is demonstrated by using simulated burst error characteristics. We also present BER results for sub-optimal combining techniques as a comparison, and illustrate an efficient multirate FFT spectral estimation technique aimed at future improvements in the channel estimation.
Finally we present the conclusions which can be drawn from this research programme and indicate its potential and limitations. Areas which may be worthy of future investigation are highlighted.

Also included at the front of this thesis is a glossary of acronyms and abbreviations used. At the end are the references and the appendices. Appendix A contains reprints of publications which are directly related to this work. Appendix B details the DSP programmes which are contained in the 5.25-inch floppy disc attached to the inside rear cover. The disc is in standard MS-DOS 360kbyte format and also includes the explanatory file "readme.txt" which is printed in Appendix B.
Chapter 2

SPREAD SPECTRUM COMMUNICATION SYSTEMS

In this chapter we introduce the concept of spread spectrum modulation and describe its use in communication systems. We begin by defining what a spread spectrum system is, and then explain how it works and discuss why we may wish to use such a system. All of the basic types of spread spectrum system are concisely described. A summary of the significant historical developments is included as a background to this current research programme.

2.1 Definition of Spread Spectrum

A spread spectrum system must, by definition, satisfy two basic requirements [9].

- The transmitted bandwidth must be greater than the bandwidth of the information being sent.
- The spreading process must be independent of the information being sent.

In order to successfully operate such a system the spreading function used in the transmitter must also be known to the receiver. Note that other wideband modulation techniques such as frequency modulation (FM) and pulse code modulation (PCM) are not examples of spread spectrum, since they do not satisfy the second requirement above.

2.2 Spread Spectrum Communication

The basis of a spread spectrum communication system can be explained by Shannon’s theory on channel capacity [10] which can be expressed as

\[ C = W \cdot \log_2 \left( 1 + \frac{S}{N} \right) \]  

(2.1)
where \( C \) - channel capacity (bits/s)  
\( W \) - channel bandwidth (Hz)  
\( N \) - channel noise power (W)  
\( S \) - signal power (W)

The channel capacity is therefore a function of the SNR present on the channel and the channel bandwidth. If we reduce the SNR of the channel we can maintain the channel capacity by increasing the bandwidth. For low SNR (i.e. \( \leq 0.1 \)) the required channel bandwidth is well approximated by the linear relationship given in equation 2.2 [11].

\[
W \propto \frac{NC}{S}
\]  

(2.2)

This indicates that, if we wish to spread our bandwidth by some factor, we can reduce the SNR by the same factor without altering the channel capacity. Thus it is possible to communicate effectively in wideband channels with very low SNR values. It also reveals that, at least in terms of signal power, spread spectrum is no less efficient than equivalent narrowband systems.

The spreading process used in the transmitter must be reversed at the receiver before the data information can be obtained. This "despreading" process effectively reduces the signal back to the original information bandwidth. This is illustrated in figure 2.1. Modern spread spectrum systems use a pseudo-random binary sequence as a spreading signal. These sequences have noise-like properties and are termed pseudo-noise (PN) codes.

Figure 2.1 - Basic Spread Spectrum System
2.3 Why Spread Spectrum?

Spread spectrum modulation results in a transmission bandwidth far in excess of the minimum required to carry the information content. This, at the outset, seems to be an extremely inefficient use of the available spectrum, and appears to contradict the philosophy of modern communications. There must obviously be some significant advantages to such a scheme to warrant its use.

The advantages of spread spectrum have long been utilised in military systems [12], and are now of growing interest for commercial systems [13]. Recently there has been considerable academic as well as commercial activity involving spread spectrum communication systems. This is due to the increasing awareness and understanding of these systems, and the availability of low cost VLSI technology. It is argued that spread spectrum may help solve some of the constraints on bandwidth allocation [14]. Mobile communications [15] and indoor wireless systems [16] are areas of particular interest at the present.

Spectral spreading is achieved by modulating the slow data sequence with a much faster spreading code. Typically the spreading ratio (i.e. the ratio of spread bandwidth to original data bandwidth used in figure 2.1) would be of the order of 100 times or greater, figure 2.2. This wideband signal, amongst other things, gives a signal to noise advantage at the output through the despreading process whereby the wideband signal is reduced to the
original data bandwidth. However, unlike wideband FM and PCM systems which have a similar effect on the signal to noise ratio, the spread signal bandwidth \( W_{ss} \) is independent of the message bandwidth \( W_m \). The difference between the output and input SNR of a system is called the "processing gain" (PG). In a spread spectrum system we can estimate the available processing gain using the ratio between \( W_{ss} \) and \( W_m \), i.e.

\[
PG = \frac{S}{N}_{out} - \frac{S}{N}_{in} \quad (\text{dB})
\]

\[
PG \approx 10 \log_{10} \left( \frac{W_{ss}}{W_m} \right) \quad (\text{dB})
\]

The spreading of the message signal, and the associated processing gain, give rise to a number of advantages for such systems [17]. Of course there are some disadvantages to counter these.

2.3.1 Advantages

- **Interference Rejection**

The despreading process uses apriori knowledge of the transmitted signal to enhance the signal content relative to the noise interference. This gives the receiver a signal-to-noise advantage which is directly proportional to the degree of spectral spreading or processing gain. The amount of spreading can be controlled allowing weak signals buried deeply in noise to be recovered and message information bits to be extracted.

- **Low Signal Detectability**

The noise rejection advantage can be used to deliberately bury the transmitted signal below the ambient noise level of the transmission medium. This makes the presence of the signal difficult to detect by the casual listener and is termed "covert" communications. It is necessary that the processing gain is high enough to recover this hidden signal and the processing gain is only available to receivers with a complete knowledge of the
spreading process. Because there is little chance of "eavesdropping", this type of signal
is said to have a low probability of intercept (LPI). There are, however, techniques which
can be used to detect hidden spread spectrum signals [18].

• Privacy

The spreading process is independent of the message signal, and so it is necessary for
the receiver to have knowledge of this process in order to despread the signal. This
indicates that, even if a casual listener could detect the presence of such a signal, it would
be difficult for the signal to be decoded to yield any information. This privacy is desirable
in broadcast methods for communicating sensitive data and, in fact, is sometimes
sufficient reason for utilising spread spectrum. However spread spectrum does not offer
the high security of cryptographic techniques.

• Jam Resistance

The correlation process used in spread spectrum receivers means that uncorrelated
jamming signals are cancelled out to a certain degree and the system exhibits resistance
to intentional and non-intentional jamming. This property can be quantified, and is known
as the jamming margin, \( M_j \). It is given by

\[
M_j = PG - \left[ L_{\text{sys}} + \left( \frac{S}{N} \right)_{\text{out}} \right] \quad (\text{dB})
\]  

(2.5)

where \( PG \) is the processing gain of the system, and \( L_{\text{sys}} \) is the system implementation
loss. This takes into account the output SNR which is required at the receiver. The
jamming margin indicates the in-band interference power which a hostile jammer would
need to exceed, relative to the desired signal power, in order to jam the communication
link.
- Multipath Rejection

Interference caused by multipath echo effects, figure 2.3, in high frequency (HF), very high frequency (VHF), and UHF communications, is a significant source of signal degradation. This is demonstrated graphically in the ghosting effect which can be seen in television pictures when there are reflections from the surrounding objects such as large buildings or hills. Since the despreading process can discriminate against time shifted versions of its own template, only one of the multipath signals is normally selected (usually the stronger signal) and the weaker signals can be rejected.

- Range and Velocity Measurement

Spread spectrum systems have been utilised to provide accurate range and velocity information. This property has found application in many systems such as radar and electronic navigation. The satellite based global positioning system (GPS) is a widely used example of such a navigation system. GPS has both civilian and military channels and will have world-wide coverage by mid 1993. The civilian band can provide accuracy to within a 100m radius. As well as the US based GPS Navstar system there is also a similar Soviet based system called Glonass.
2.3.2 Disadvantages

- Complexity

Due to the inclusion of the spreading process at the transmitter and the despreading process at the receiver the complexity, and therefore the cost, of the spread spectrum system is increased. This can be minimised by careful design, and may be partially offset by the fact that filtering requirements are less stringent than in traditional band-minimised systems.

- Bandwidth Requirements

The wide bandwidth required for spread spectrum signals can, of course, be a problem. At the present time legislation may limit the amount and precise allocation of bandwidth which is available for RF spread spectrum signals. In theory it should be possible to bury the spread signal so deeply in the ambient noise that a single user’s effect on the other narrowband users, which it overlays, is negligible. However the regulations are an important consideration in the design process and it may be some time before the spread spectrum concept is fully taken into account.

2.4 Spread Spectrum Techniques

Spread spectrum systems can be categorised by the techniques used in the signal spreading and despreading processes. Four distinct categories can be identified:

- Direct Sequence (DS)
- Frequency Hop (FH)
- Time Hop (TH)
- Chirp

DS and FH are the most widely used techniques. TH is less common and tends be most prevalent in hybrid systems. Chirp is by definition a spread spectrum technique and is used mainly for radar applications, but it does not afford all of the advantages listed in
the previous section since the spreading process is a linear frequency sweep pattern and is not governed by a PN code. Chirp is rarely used for communication systems. Hybrids of these systems are possible by cascading two or more of the techniques. The most popular hybrid systems have been those employing DS/FH techniques, although FH/TH and DS/TH are not unknown.

2.4.1 Direct Sequence

Direct sequence spread spectrum (DSSS) modulation is the simplest of all systems. Each message bit is modulated by the many bits (usually termed "chips") of the PN sequence. This is achieved using simple modulo-two addition (i.e exclusive-or arithmetic). It is common, in civilian systems, for the period of the PN sequence to be the same as the message bit period. In this case we have either an inverted or non-inverted PN sequence depending on whether the data bit was a one or a zero respectively. This is termed sequence inversion keying (SIK). A simple DSSS transmitter using SIK is shown in figure 2.4.

![Figure 2.4 - Direct Sequence Signal Spreading](image)

It is also possible to use a sequence which is much longer than the data bit period, and use the same modulo-two modulation technique. Not only must the very long PN code be known at the receiver, but a good estimate of the part of the code currently being used must also be known before it can be successfully decoded. This type of system requires more complex synchronisation, but it does have greater security, hence it is used more widely in military systems.
To decode a DSSS signal at the receiver we must compare the likeness of the incoming signal with the same PN code which is used in the transmitter. This process, which causes the despreading, is achieved by correlating the received signal with the known PN code. If a non-inverted PN signal was transmitted, we obtain a large positive correlation peak at the output of a correlator when the received signal aligns correctly with the PN code in the receiver. Similarly if an inverted PN signal is transmitted we obtain a large negative peak at the output from a correlator, which is due to the received PN signal and the receiver PN code being opposite in polarity when aligned.

The power spectral density (PSD) of a DSSS SIK signal consists of two elements. First of all, the periodic nature of the PN code leads to discrete spectral lines spaced by $1/T_m$, where $T_m$ is the message bit period. A $\text{Sinc}^2$ envelope with spectral nulls spaced $1/T_c$ apart defines the amplitude of these spectral lines. $T_c$ is the PN code chip period, i.e. $T_c \times M = T_m$, where $M$ is the PN code length. Secondly, if we assume the message data to be random, the message signal will appear as a continuous $\text{Sinc}^2$ envelope with nulls spaced $1/T_m$ apart. The DSSS signal is therefore obtained by convolving these two signals. In the time domain a PN signal multiplied by a slower random data signal has the appearance of a completely random signal of the same chip rate. It is not, therefore, surprising that in the frequency domain the DSSS signal looks like a continuous $\text{Sinc}^2$ envelope equivalent to that of a truly random sequence of chip period $T_c$. The baseband PSD, $S(\omega)$, is approximated by

$$S(\omega) = T_c \text{Sinc}^2 \left( \frac{\omega T_c}{2} \right)$$

(2.6)

It is difficult to detect the message bit period or deduce the PN code length from the spectrum when the system is modulated by random data, especially for large spreading factors. As $M \to \infty$ the power envelope approaches that of the continuous function given in equation 2.6.

\[ ^{\dagger} \text{Sinc}(x) \text{ is given by } \frac{\sin(x)}{x} \]
2.4.2 Frequency Hop

Frequency hop spreading, shown in figure 2.5, results in a wideband signal with a well defined set of frequency peaks in its spectra, each separated by \( f_h \). However it differs from frequency shift keying (FSK) in the fact that it is the deterministic random spreading process and not the data that selects the frequency of transmission. It is then necessary to apply some modulation to the stepped carriers in order to include the message data content. This data modulation of the spreading carriers is usually provided by conventional techniques such as PSK and FSK but can also include DS spread data such as in FH/DS hybrid systems.

Frequency hop systems were developed for military use in an attempt to avoid enemy jamming or interception. The idea was simply to jump the carrier frequency around through a seemingly random selection of predetermined frequencies at a fast enough rate so that enemy jamming transmitters or listening receivers could not reliably lock onto the signal. This is an acceptable technique provided that the hop period is less than the round trip propagation delay to the enemy jammer which will prevent the jammer following the hopping signal.

The spreading process is usually defined by the states of a PN sequence. The frequency hops are spaced evenly within the spread bandwidth and each frequency is occupied uniquely for only a short period of time before hopping to another frequency in the
sequence. Each frequency occurs only once within a complete hopping sequence. This fact can be utilised by the receiver to produce a local synchronisation on the carrier sequence by searching for the start of the hopping sequence.

The process used to hop the carrier frequencies is independent of the data signal, and so it is possible to convey data bits with as many hops as is felt necessary. It is also possible to convey more than one data bit on an individual hop. When the centre frequency is hopping at a faster rate than the data bit rate then the system is termed fast frequency hop (FFH), otherwise it is known as slow frequency hop (SFH). SFH systems are useful in combating in-band signal jamming as long as the hopping rate is fast enough to avoid any jammer locking on to the signal frequency (repeat jamming). SFH is compatible with conventional frequency synthesised transceiver equipment and has been used in military combat network radio equipment such as SYNCgars or UK Scimitar/Jaguar systems [19].

Frequency hop spread spectrum systems can also support multiple access. If a suitably long frequency hop sequence is used, the probability of frequency "collisions" between adjacent channels is low.

**2.4.3 Time Hop**

![Figure 2.6 - A Time Hopped Signal](image)

Time hop is the least utilised of the spread spectrum techniques available. The data symbol period is divided into a set number of time hop periods, which in turn is subdivided into a number of timeslots, figure 2.6. Spectral spreading is accomplished by pulsing
the transmission of the data symbol in an individual timeslot within the particular time hop period. The allocation of which timeslot the data symbol is pulsed in, is determined by the state of a PN sequence generator. The processing gain of the system comes from the averaging of the total number of pulses used to convey a single bit of information. In many TH systems the time hop period and the data symbol period are the same (unlike in figure 2.6), resulting in the reception of only one pulse per data bit. This gives no processing gain but is useful as a method of enciphering TDM communications and the resultant system is more secure (if more than one channel is being utilised).

2.4.4 Chirp Systems

Chirp modulation involves sweeping the frequency of the carrier during a pulse period and this modulation scheme is used almost exclusively by radar systems. It is not based on the use of PN code sequences but the bandwidth produced is greater than the minimum required by the signal and a processing gain enhancement is achieved. Its use within communications systems is limited [20] as there are only two available codes (the up chirp and down chirp) but it is nevertheless a valid spread spectrum system.

In radar chirp is utilised mainly to produce a significant power reduction and is normally implemented by pulsed RF signals whose frequency content varies in a known, usually linear, manner. Receivers for this type of signal are of the form of a matched filter for the linear FM coding and this is a major application area for surface acoustic wave (SAW) devices which are further weighted for time sidelobe reduction [21].

Normally the frequency sweep of the chirp signal is linear but any pattern which can support the use of a matched filter is possible. The despreading process is performed by matched filtering to the FM swept signal. The output signal is effectively enhanced and the processing gain is given by the compression ratio or time-bandwidth product of the carrier sweep.
2.5 Multiple Access

Multiple access in communication systems is achieved by dividing the bandwidth of the system in a manner such that separate channels can exist. The interference between these channels must be low so that the system can operate satisfactorily. These separate channels can be created by various means.

The system bandwidth has traditionally been split into different channels by dividing it into smaller frequency channels - frequency division multiple access (FDMA), or by dividing it into channels separated in time - time division multiple access (TDMA). With spread spectrum techniques it is possible to divide the system bandwidth into separate channels, with each using a unique PN code. If these PN codes have low correlation with each other, the receiver can discriminate between the different channels using the different PN codes. This is known as code division multiple access (CDMA). The degree of discrimination between these codes (their orthogonality) will determine the degree of co-channel interference, and hence the capacity of this system.

CDMA systems have the advantage of degrading gracefully with the amount of channel loading. Each additional channel appears as an extra additive noise term on the other channels. The capacity of FDMA and TDMA systems is fixed and once all channels are utilised, any additional communications will be blocked.

The PN codes used are of fundamental importance. Research into PN code properties has resulted in the identification of superior families of codes for multiple access purposes [22]. These PN code sequences provide "channels" for multiple access transmission within the same bandwidth. The number of users that the channel can support is dependent on the length, and also the orthogonality of these codes with respect to each other.

Time shifted versions of the same code can also be used for multiple access, due to the fact that the correlation process rejects its own sequence if the template is not aligned to within less than one chip period. With a novel protocol the time shift "channels" can be used to convey multiple access communications. Simcock et al. [23] describe a system
which achieves the multiple access capability of sixty nodes within a single code sequence of length 1024 chips used for a local area network (LAN). This type of protocol is not suitable for highly dispersive channels.

All spread spectrum techniques can be used for multiple access except for chirp systems which do not use PN sequence coding techniques.

### 2.6 Correlation Functions and PN Codes

The properties of PN codes used in a spread spectrum system can be of considerable importance. Perfectly random sequences of infinite length have an autocorrelation function (ACF)\(^\dagger\) which is a single infinite energy impulse at time zero, with zero energy at all other time delays, figure 2.7.

![Autocorrelation Function of a Random Sequence](image)

**Figure 2.7 - Autocorrelation Function of a Random Sequence**

Pure random signals or sequences have a cross-correlation function (CCF)\(^\ddagger\ddagger\) which is zero at all time delays. The correlation properties of random sequences are desirable (especially for CDMA), but clearly cannot be achieved for practical sequences which must be of finite length. Figure 2.8 shows a typical ACF for a repeating finite length PN code, where the width of the peak is controlled by the PN code chip period.

\(^\dagger\) An ACF is defined as a signal correlated with a time shifted version of itself.

\(^\ddagger\ddagger\) A CCF is defined as one signal correlated with a time shifted version of another signal.
Figure 2.8 - Autocorrelation Function for a Repeated PN Code

Much of the research effort in this area is concerned with obtaining "good" sets of PN codes with universally low CCFs with respect to all other codes in the set. This ensures that co-channel interference in CDMA systems is minimised. Near zero sidelobe levels on the ACF are particularly desirable when the receiver is in the synchronisation acquisition phase to avoid false synchronising to the sidelobes. The ACF sidelobes can also provide useful information about the channel such as noise levels and multipath information. Multipath effects can cause minor intersymbol interference (ISI) for non-zero ACF sidelobes.

There are several PN code families which are used, or have been proposed, for spread spectrum communications.

2.6.1 M-Sequences

Maximal length sequences (m-sequences) are generated using shift registers with feedback, figure 2.9. The output, if the feedback taps are correctly set, is a periodic sequence which exhibits pseudo-random properties. The symbol "D" represents a unit delay. The PN code produced has almost an equal number of ones and zeros (there is always one less zero than there are ones). The runs of the same symbol follows an exponential distribution such that 1/2 are runs of length one, 1/4 are runs of length two, 1/8 are of length three, and so on.

These sequences are termed "maximal" since the shift register will cycle through all of the state permutations except the all zeros state. All zeros in the shift register would result in there being no feedback term and so the shift register would become stuck in this state. Starting the shift register from any other initial condition will cause it to
generate an m-sequence which is determined by the feedback polynomial. Since the shift register cycles through all permutations except the all zeros state, we can deduce that the length of the output sequence (before it repeats) is:

\[ M = 2^m - 1 \]  

where \( M \) is the sequence length, and \( m \) is the effective shift register length or code order.

For each length of shift register, \( m \), there can be a number of different m-sequence codes. However not all feedback polynomials will produce m-sequences. The polynomial must be irreducible, i.e. it will always produce a remainder when the non-zero shift register state polynomial is divided by the feedback polynomial. Also the PN code length must be \( M \), otherwise the sequence is not maximal. Not all irreducible polynomials produce m-sequences, some produce shorter sequences instead.

Equation 2.7 indicates that very long m-sequences can be produced with relatively few shift register stages. There are a number of different codes which can be obtained from the same number of shift register stages. The position of the feedback taps in figure 2.9 (i.e. the feedback polynomial) determines which code will be generated. Table 2.1 shows the first few irreducible polynomials which can be used to generate m-sequences, these are sometimes referred to as the primitive polynomials. Note that, in general, the number of possible codes increases as the number of shift register stages increases, table 2.2. The feedback polynomial for figure 2.9 is given by \( x^3 + x + 1 \), and is equivalent to \( 1 + x^{-2} + x^{-3} \).
M-sequences have been widely used, mainly because they are very easy to generate. The ACF of these codes are two-valued with a peak of $M$, and a sidelobe value of -1. However this seemingly ideal property is lost when data modulation is applied in the form of SIK and we effectively have an aperiodic ACF. In general m-sequences display relatively high CCF values when correlated with other m-sequences of the same length. They are not, therefore, considered to be desirable for CDMA applications. However, a small subset of m-sequences which have low CCF values have been identified. These codes are too few to be useful for CDMA in themselves. From these near orthogonal m-sequences we can select "preferred pairs" which can be used to generate a large family of new codes which also display near orthogonal properties. These codes are called "Gold codes".

<table>
<thead>
<tr>
<th>order(m)</th>
<th>irreducible polynomial</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>$x^2 + x + 1$</td>
</tr>
<tr>
<td>3</td>
<td>$x^3 + x + 1$</td>
</tr>
<tr>
<td></td>
<td>$x^3 + x^2 + 1$</td>
</tr>
<tr>
<td>4</td>
<td>$x^4 + x + 1$</td>
</tr>
<tr>
<td></td>
<td>$x^4 + x^3 + 1$</td>
</tr>
<tr>
<td>5</td>
<td>$x^5 + x^2 + 1$</td>
</tr>
<tr>
<td></td>
<td>$x^5 + x^3 + 1$</td>
</tr>
<tr>
<td></td>
<td>$x^5 + x^4 + x^3 + x^2 + 1$</td>
</tr>
<tr>
<td></td>
<td>$x^5 + x^3 + x^2 + x + 1$</td>
</tr>
<tr>
<td></td>
<td>$x^5 + x^4 + x^2 + x + 1$</td>
</tr>
<tr>
<td></td>
<td>$x^5 + x^4 + x^3 + x + 1$</td>
</tr>
</tbody>
</table>

Table 2.1 - M-Sequence Irreducible Polynomials

M-sequences have been widely used, mainly because they are very easy to generate. The ACF of these codes are two-valued with a peak of $M$, and a sidelobe value of -1. However this seemingly ideal property is lost when data modulation is applied in the form of SIK and we effectively have an aperiodic ACF. In general m-sequences display relatively high CCF values when correlated with other m-sequences of the same length. They are not, therefore, considered to be desirable for CDMA applications. However, a small subset of m-sequences which have low CCF values have been identified. These codes are too few to be useful for CDMA in themselves. From these near orthogonal m-sequences we can select "preferred pairs" which can be used to generate a large family of new codes which also display near orthogonal properties. These codes are called "Gold codes".
2.6.2 Gold Codes

A method of isolating preferred pairs of m-sequences is described by Gold [24]. When a preferred pair is obtained, they can be combined using modulo-2 addition, figure 2.10. The resulting sequence is a Gold code which has approximately the same degree of orthogonality as the preferred pair m-sequences. Further, if one of the codes is shifted by some number of chips, the resulting combined code is another Gold code of the same family and with similar properties. Therefore there are $M$ possible shifted combinations of the preferred pair. We can include the original two m-sequences in the Gold code family and so there are a total of $M+2$ Gold codes available for any code length of $M$.

The near orthogonal properties of these Gold codes, the ease with which they can be generated, and the large code sets which exist, have made them very useful for CDMA systems.

<table>
<thead>
<tr>
<th>order</th>
<th>length</th>
<th>number of m-sequences</th>
</tr>
</thead>
<tbody>
<tr>
<td>m</td>
<td>M</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>3</td>
<td>1</td>
</tr>
<tr>
<td>3</td>
<td>7</td>
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<tr>
<td>5</td>
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</tr>
<tr>
<td>6</td>
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<tr>
<td>8</td>
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<td>1023</td>
<td>60</td>
</tr>
<tr>
<td>12</td>
<td>4095</td>
<td>144</td>
</tr>
<tr>
<td>16</td>
<td>65535</td>
<td>2048</td>
</tr>
</tbody>
</table>

Table 2.2 - Numbers of Valid M-Sequences
2.6.3 Other PN Codes

The number of bit patterns available, even for short code lengths, is extremely large. For example, the number of permutations for length 15 is given by:

\[2^{15} = 32,768\]

Many of the possible permutations will not exhibit good pseudo-random properties, however a great number will. The number of m-sequences and Gold codes for a code of length 15 (2 and 16 respectively) constitutes only a small fraction of those available. There are many other code families which can be successfully used in spread spectrum communication systems. These are too numerous to try and document in this concise text.

Amongst the other PN code families are Kasami sequences which were developed for use in error correction circuits. They produce similar ACF and CCF performances to those of the m-sequences and Gold codes.

Composite codes, such as Kronecker sequences [25], use several (usually two) shorter codes to produce a much longer code. Composite codes generally use an outer PN code of length \(K\), which is used to "modulate" a further inner code of length \(L\). This results in a PN code of length \(K \times L\). The correlation properties of these codes tends to be poorer than those obtained using other families. The benefits which can be gained by using
them lies in the generation and detection circuits where the complexity is minimised. More rapid synchronisation can usually be achieved since acquisition may be obtained first for the inner code and then for the outer code. This limits the search to $K+L$ chips rather than $K \times L$, but reduces the privacy of the codes.

The choice of PN codes for spread spectrum systems is dependent on the requirements of the particular system. PN codes selected by a system designer are invariably a compromise since no finite length PN code exhibits the ideal correlation properties.

### 2.7 Historical Review of Early Spread Spectrum Systems

Spread spectrum techniques grew out of the need to protect military communications from unwanted reception and intentional jamming. The first methods employed in the 1940s were crude frequency hop style systems, where both transmitter and receiver stepped through a known sequence of frequency bands for signal transmission. From these early steps, research into the use of spread spectrum, most notably by the American armed forces, led to the development of various working systems from the 1950s onwards [26].

Towards the end of World War II missile guidance systems were being developed. There was a fear that the Germans would quickly learn how to jam the remote link of the guidance systems, and so anti-jamming techniques were investigated. One early anti-jam system, the AN/ARW-4, employed a dual-band frequency hop system. This system could still operate successfully when one band was completely jammed. Another system, also developed for missile guidance around this time, was ITT’s Rex system which used a changing pulse repetition frequency to combat jamming signals. These systems were forerunners to the early spread spectrum systems which were to follow. Neither of these systems used the concept of pseudo-random noise sequences.

WHYN was a navigation system originally developed for missile guidance, the acronym standing for wobbled hyperbolic navigation. A carrier was simply frequency modulated by a sinewave - the term FM wobulation was used. This produced a
bandwidth expansion which was independent of the signal content. Although noise modulation was still not used, correlation was employed at the receiver to compress the bandwidth back to the message bandwidth. This system clearly can be categorised as a true spread spectrum system, since it satisfies the definition given in section 2.1.

Shannon’s paper [27] on the subject of communicating in the presence of noise was published just after these very early spread spectrum systems appeared. This work indirectly implies the concept of using coded wideband signals as a method of reducing the bit error likelihood due to noise interference. At the same time as Shannon’s pioneering work, Rogoff, working at ITT, demonstrated the fundamentals of spectral spreading systems using a novel approach. He built a "noise wheel" for storing a random noise-like signal generated from a selection of telephone numbers. These numbers were transferred onto a circular film, and the radius of transparent slits etched in this film corresponded to the randomly selected telephone numbers. This stored noise-like signal could be transferred into an electrical voltage by rotating the wheel past a light source, and the resulting light intensity was sensed by a photocell. Rogoff then mounted two identical wheels on a single axis. The signal generated by one of the wheels was used to modulate data, resulting in a spread spectrum type signal. This spread signal plus interference was then correlated with the unmodulated noise-like signal derived from the second duplicate wheel, which recovered the data content. Working at 1 bit per second, this crude method highlighted the feasibility of transmitting information hidden in noise-like signals.

One of the very first operational systems utilising noise coded spread spectrum techniques was called NOMACS (Noise Modulation and Correlation System). This was a joint project by MIT’s Lincoln Laboratory and the Army Electronics Command, and was built and tested in the late 1940s. Based on direct sequence, it employed a spread bandwidth of 10 kHz transmitted in the HF frequency band and gave an overall processing gain of 25 dB. Unfortunately due to the complexity of the early technology necessary
in its construction at the time, it was never used operationally. However the system was
tested and worked very well in prototype form, communicating between the two
laboratories involved in its development.

Another DS system developed was that of the Hush-Up programme which used
m-sequence codes and led to the production of the ARC-50 by Magnavox. This system
again used the HF band for ground to air communication links of up to 100 miles. It did,
however, suffer from multipath effects when operated over long ranges. To overcome
some of the problems associated with the ARC-50 system FH was used for the subsequent
system, BLADES, which again used m-sequences. FSK was used for the data modulation
and error correcting codes were employed. BLADES has seen service in long distance
ship to shore communications.

Further developments of the NOMAC system became what was called the F9C. Again
this was a DS system and initially the message data was applied using FSK. The system
provided a very useful jamming margin of about 17dB and went into production as the
F9CA. The FSK data modulation was eventually changed to SIK. The main limitations
of this system were due to multipath dispersion losses. However it was found that the
time of arrival (TOA) resolution of this system was sufficient to separate out different
multipath components using a bank of active correlators. It was realised that these
components could be coherently combined in order to significantly improve the receivers
SNR and help combat the severe multipath effects. The resulting receiver was called a
RAKE. Several RAKE receivers were constructed and used to enhance the F9CA system.

In 1952 the Jet Propulsion Laboratory (JPL) in California were involved in the
development of communication and control systems for the Corporal rocket. They
decided to use DS for the rocket guidance system and appear to have coined the term
"pseudo-noise". The PN system had a jamming margin of over 15dB. Synchronisation
acquisition was achieved by connecting the transmitter and receiver together prior to
launch. Another system used for the same rocket was the scrambled continuous wave
(SCW) system. It used a narrowband CW modulated by a PN code. Further developments led to the coded Doppler radar command (CODORAC) system. This was a guidance system used for the Sergeant and Jupiter missiles.

Table 2.3 shows the development of these early spread spectrum systems in an approximate chronological time frame. It is significant that true SS systems started to appear just after Shannon publish his works on communication in noise. Both frequency hop and time hop type systems were tried prior to this in an attempt to avoid jamming. It was perhaps as a result of the development of information theory that PN based systems appeared in the early 1950s.

<table>
<thead>
<tr>
<th>1940</th>
<th>1945</th>
<th>1950</th>
<th>1955</th>
<th>1960</th>
<th>1965</th>
</tr>
</thead>
<tbody>
<tr>
<td>SYLVANIA</td>
<td>AN/ARW-4</td>
<td>WHYN</td>
<td>HUSH-UP ARC-50</td>
<td>BLADES</td>
<td></td>
</tr>
<tr>
<td>ITT</td>
<td>REX FACSIMILE SYS.</td>
<td>NOISE WHEELS</td>
<td></td>
<td></td>
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</tr>
<tr>
<td>MIT</td>
<td>NOMAC</td>
<td>F9A-A</td>
<td>RAKE</td>
<td></td>
<td></td>
</tr>
<tr>
<td>JPL</td>
<td>PN</td>
<td>SCW CODORAC</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>MAGNAVOX</td>
<td>ARC-50</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>SS CONCEPT</td>
<td>FH &amp; TH</td>
<td>SHANNON'S THEORY</td>
<td>DS-SS</td>
<td>SS SYSTEM DEVELOPMENT</td>
<td></td>
</tr>
</tbody>
</table>

Table 2.3 - Chronology of Early Spread Spectrum Systems†

In 1959 Costas expanded the wideband approach to include the concept of multiple access systems employing spread spectrum techniques overlaid on existing narrowband channels [28]. Costas suggested that the capacity of the spectrum is a fixed quantity and cannot be exceeded, but more elegant methods of sharing its allocation are available through spread spectrum.

More recent technological advances such as SAW device developments in the 1970s [21] have facilitated the use of spread spectrum in more sophisticated communication

† Source of data is Scholtz [26].
systems. These include the Joint Tactical Information Distribution System (JTIDS), and the Airborne Warning and Control System (AWACS) which has been operational from the mid 1970s and employs a hybrid combination of FH and DS to provide a number of secure and jam resistant communication channels.

Nowadays spread spectrum concepts are widely understood, and the signal processing technology required for its implementation is less specialised and becoming less expensive, for example hand-held Navstar GPS receivers now retail for a few hundred pounds. However it was not until very recently that interest has been directed at the spread spectrum alternative for use in commercial communication systems such as mobile cellular radio networks.
Chapter 3

THE SERIAL-PARALLEL RECEIVER

Systems can often be categorised by their mode of operation or the process used, each type of system has advantages and disadvantages relating to the type chosen. In order to achieve a better compromise it is sometimes possible to produce a hybrid system which is more suitable. The serial-parallel architecture discussed in this chapter is such a system. It will be seen that the system outlined can benefit from some of the merits of both serial and parallel receivers, while not suffering greatly from their disadvantages.

3.1 Spread Spectrum Receivers

A spread spectrum signal must be "despread" using the same PN code as that which is employed in the transmitter. No decisions as to the data content can be made prior to this process as the signal will appear noise-like. The signal may in fact have a power level which is considerably less than that of the noise contained within the signal bandwidth. The optimal method for despreading the signal when the channel is corrupted by white Gaussian noise is by correlating the received signal with the known PN code. This is a form of matched filtering and is based on simple majority decision decoding [29]. However this method is not optimal if the channel contains any correlated interferers. Under these conditions maximum likelihood decoding will provide better performance [30]. Our study will be restricted to the basic matched filtering approach for two reasons. Firstly the urban channel is largely subject to Gaussian or impulsive types of noise mixtures [31] which have low cross-correlations with respect to a PN type code. Secondly it will be assumed that any co-channel interferers will be seen as AWGN if the channel is used for multiple access (i.e. CDMA).
Ideally the output from the receiver’s matched filter will be an ACF of the PN coded waveform, $x(t)$, given by,

$$\Phi_{xx}(\tau) = \int_{0}^{\tau} x(t) \cdot x(t + \tau) \cdot dt \quad (3.1)$$

where $\tau$ is the time shift between the two coded waveforms.

However since we transmit a binary "zero" as a non-inverted PN sequence and a "one" as a polarity inverted PN sequence (i.e. SIK), we have two types of ACF. An even ACF occurs when no data transition occurs, and an odd ACF corresponds with a data transition. In addition to this, the ACFs can exist as either positive or negative functions due to the polarity of the PN coding. This is illustrated in figure 3.1 for a 31 chip m-sequence. Trace (a) shows the message data signal, (b) is the PN signal which has been spread using the 31 chip code, and (c) is the ACF of this PN signal.

![Figure 3.1 - PN Code Autocorrelation Function (m-sequence)](image)

Since the received signal, $y(t)$, will now be corrupted by the channel and will also contain additive noise, it will not be identical to the transmitted PN coded signal, $x(t)$. We therefore have a CCF rather than an ACF. This is obtained using equation 3.2.
Cross-correlation also occurs between separate coded sequences in multiple access DS systems and the correlation coefficient $\Phi_{AB}(\tau)$ will indicate the level of interference between two coded waveform signals $A$ and $B$.

$$\Phi_{AB}(\tau) = \int_{0}^{T} x(t) \cdot y(t + \tau) \cdot dt$$  \hspace{1cm} (3.2)

The correlators used in the receiver are often classified as either serial or parallel architectures. These will be explained in the following sections.

### 3.2 Serial and Parallel Architectures

Early spread spectrum systems tended to use serial type receivers (sometimes called active correlators) mainly due to technology constraints and because they were relatively simple to construct using conventional analogue circuits. Parallel receivers are now starting to dominate modern spread spectrum systems. They first appeared in analogue form as analogue matched filters (AMFs) [32] using SAW devices [33] and charge coupled devices (CCDs) [34]. More recently digital matched filters (DMFs) have been employed, these are constructed using VLSI devices. The parallel architectures are undoubtedly more complex but they do offer significant advantages in speed of synchronisation and their ability to resolve multipath propagation components. Each of these architectures will be discussed in turn.

#### 3.2.1 Serial Architectures

Serial receivers operate by multiplying and accumulating (over one data bit interval $T$) the incoming signal with a locally generated version of the PN code on a continuous chip by chip basis. The accumulated total is dumped before the process is repeated over the next data bit interval, figure 3.2. Unfortunately serial receivers require correct synchronisation before the signal will be correctly despread and there is, therefore, an associated acquisition time for the receiver. For a repeated PN code where the code
length is $M$ and the data bit period is $T$ we will require up to $M$ serial correlations each of time $T$ before signal acquisition is achieved. When a correlation did not acquire the ACF peak, the code is slipped slightly (by a maximum of one chip period) such that a different time phase of the correlation function is obtained. This procedure is repeated until the correlation peak is acquired and the code is locked using a tracking filter. This is known as a serial search.

The maximum acquisition time for a serial receiver is thus $MT$ seconds (assuming that the search is on a chip by chip basis and that the peak is found on the first attempt). Analogue receivers are still widely used due to their speed and the large processing gains which they can achieve, however digital systems are starting to approach these performance levels. The output signal from a serial receiver is only really valid when an integration period is completed and the result is dumped when $t = T$. Thus the effective bandwidth of the serial post-correlation signal has been reduced back to that of the unspread narrowband message signal.

![Serial Receiver Diagram](image)

3.2.2 Parallel Architectures

The parallel receiver performs a similar function to that of the serial receiver except that it uses a tapped delay line or finite impulse response (FIR) filter architecture with an array of separate multipliers. It produces the full ACF waveform output shown in figure 3.1 trace (c), and requires just a single chip period to calculate one sample of the
correlation function. The receiver multiplies the received signal with a static version of the PN code and accumulates the result in a single chip period, figure 3.3. It is therefore an asynchronous process, but may require tens of individual devices to be cascaded before the required processing gain can be achieved. In most practical digital devices the number of multiply and accumulate stages is in the range 8-64.

![Diagram of Tapped Delay Line and PN Code Coefficients]

Figure 3.3 - The Parallel Receiver

The asynchronous nature of the parallel architecture allows acquisition to be achieved within one message bit period, $T$ seconds, however the hardware complexity is considerably greater than that of the serial architecture. A complete correlation is performed every chip period and so the effective post-correlation bandwidth for the parallel receiver is the same as that of its input. This implies that there is considerably more useful time domain resolution obtained from parallel architectures than from the equivalent serial form. The output signal envelope over a message bit period is known as the correlation profile and can be very useful for providing information about the communication channel statistics. The implications of this property will be discussed later.

Analogue implementations are useful in that the signal contains both amplitude and phase information and so they can process complex signals directly. In particular SAW
devices are capable of decoding bandpass signals directly at their carrier or intermediate frequency, and hence can be considered to be receivers in their own right. DMF based systems, on the other hand, can only process real baseband signals, the DMF element is therefore simply a correlator. To decode non-coherent complex signals two real channels must be used with a quadrature phase separation. The receiver structure for non-coherent detection of demodulated "band-pass" signals using real (in-phase, I) and imaginary (quadrature, Q) signal components, is widely used. This arrangement is illustrated in figure 3.4.

### 3.3 Hybrid Serial-Parallel Architectures

A hybrid serial-parallel architecture was first suggested by Hunsinger [35] in an attempt to overcome the time-bandwidth product limitations encountered when using SAW devices. The maximum time-bandwidth product for a single SAW AMF is limited to about 30dB, primarily due to physical limitations.

Hunsinger's basic design, figure 3.5, consisted of two sections. The first was the matched filter element, and the second was a recirculating integrator which performs the integration process. The operation can be described as follows. The received signal is a 1016 chip composite PN code composed of 8 identical 127 chip m-sequence codes whose relative phase (0° or 180°) is determined by an 8 bit PN code. The SAW device is matched to the 127 chip PN code, and so its output is an eighth partial correlation of the total received signal. The eight bit PN code then phase corrects these partial correlations. In
order to reconstruct the full correlation we sum the eight partial correlations coherently. This is performed by the recirculating integrator consisting of a 127 chip SAW delay line followed by an integrating element and the loop has a switchable feedback.

The resultant output from the serial-parallel receiver is obviously only valid once the integration is complete, and will thus provide only a section (1/8th) of the full correlation profile which is equivalent to the length of the parallel matched filter element (127 chips for Hunsinger's receiver). The resolution of this section is however identical to that which would be obtained from a full parallel correlator. Note also that the correlation profile can be seen accumulating at the output of the integrator.

One of the major disadvantages of using conventional SAW devices is that their structure (i.e. the tap coefficients) is fixed, and so the code diversity and security of the signal is greatly reduced. The ACF of the composite codes also have higher sidelobes than those of most other types code. These constraints have been partially overcome by the use of programmable SAW AMFs (PAMFs) [36] or SAW convolvers [37]. However these enhanced systems increase the system complexity and were not always reliable. All of the SAW systems were also limited by self-noise and bandlimiting in the recirculating integrator which constrains the maximum processing gain to about 40dB [38].
feedback path of the integration loop must have a gain of greater than one to compensate for the losses in the SAW delay line. This can lead to implementation losses as great as 2dB [39] due to the system noise.

Although the serial-parallel receivers are not asynchronous like the parallel architectures, they do offer considerably faster acquisition times than serial receivers. If we let $N$ be the parallel matched filter device length, and $M$ is the PN code length, with $T$ being the message bit period, - the maximum acquisition time will be $\frac{M}{N} \cdot T$. This is an $N$ fold reduction compared to that of the serial correlator since an $N$ chip section of the profile is searched each message period rather just a single chip. It is usual for the ratio $\frac{M}{N}$ to be an integer since this simplifies the hardware requirements.

3.4 The DMF Serial-Parallel Correlator

The DMF based implementation of the serial-parallel architecture does not differ greatly from that of the analogue architecture, however, only real signals can be processed directly. The system consists of a parallel DMF followed by a digital delay line equal in delay to that of the DMF stage. An accumulator sums the result at the output and the feedback gain can be switched to 0 or 1, figure 3.6. The main advance in the digital implementation of figure 3.6 is that the matched filter is repeatedly reprogrammed with the next N-chip segment of code making the correlator fully programmable, and it is able to receive any coded sequence reliably.

The analogue serial-parallel receiver implementations were largely limited by the performance of the analogue SAW devices, particularly the recirculating integrator. Digital implementations are limited by the signal quantisation and the finite arithmetic effects of the hardware employed. Once the signal has been quantised there will be no further noise introduced into the system. In the analogue system self-noise was a problem which limited the performance, there are no such problems with the digital system. It should therefore be apparent that given moderate arithmetic precision, the digital system will out perform the analogue system in terms of the available processing gain, e.g. 24
bit integer accumulator precision can produce processing gains of about 70dB - a 1000 fold increase over the equivalent analogue systems. With floating point arithmetic the processing gain is virtually limitless. It has been shown that the losses due to quantisation are minimal [40] [41] for linear quantisation in additive white Gaussian noise (AWGN).

At present the speed of the digital systems cannot match those of their analogue counterparts. The fastest commercially available correlator/DMF hardware (end-1991) will run at up to 30Mchip/s [42]. SAW devices are capable of speeds of around 50Mchip/s. If current trends are to be continued it seems highly likely that the speeds of the AMFs will soon be matched by DMFs with useful resolutions. There is also great potential for the miniaturisation of the hardware and the increase in reliability and reduced power consumption which usually accompanies this. The digital output of such systems may also be post-processed without introducing further noise into the system.

The limited length (or time-bandwidth product) of commercial DMF/correlator devices has led to some manufacturers designing their devices such that they are cascadable. A good example of this approach is the Inmos A100 chip [43] which is termed a "cascadable signal processor". This device enables a signal to be delayed (the delay is the same as that through the FIR filter) and added to the output of the filter element. Filter tap coefficients can also be updated using a serial-in and parallel-load technique, figure 3.7a.
This device was clearly intended to be cascaded with other identical devices to produce even longer parallel DMF structures, but is also ideal for the construction of serial-parallel correlators since it is easily cascaded with itself, figure 3.7b. The A100 therefore contains all of the main hardware elements of the DMF serial-parallel correlator shown in figure 3.6. Additional hardware is required for the interface and control functions of the system. The Inmos A100 is capable of processing at up to 10Mchip/s. It is also possible to perform complex correlations with a single device, however this is incompatible with the serial-parallel design and it also halves the time-bandwidth product of the device. Another particularly useful device is the Stanford Telecom STEL-3310 [44] which has 64 taps, runs at up to 11Mchip/s, is cascadable, and has both I & Q channels for processing complex signals in a single device.
Figure 3.7 a) The Inmos A100, b) The A100 Serial-Parallel Correlator
3.5 PN Code Selection for Serial-Parallel Receivers

There is considerable literature available detailing PN code selection for code division multiple access (CDMA) systems, however we require to select a code with essentially different requirements for the serial-parallel receiver. In common with the requirements of other systems, we are interested in codes which exhibit low sidelobe peaks so that they are unlikely to be mistaken for the main peak during the receiver code acquisition phase. As we may wish to use the sidelobe multipath information from the correlation profile, we also require the autocorrelation sidelobe energy to be significantly less than that of the multipath energy over its excess delay period to ensure that this information is not greatly distorted. The excess delay time of the signal is the time over which significant amounts of energy have been dispersed by the multipath channel.

Most common classes of PN code are of length $2^m-1$ where $m$ is an integer. The DMF serial-parallel correlator is greatly simplified if the code is of length $M$ and the DMF is of length $N$ such that $\frac{M}{N}$ is an integer. We cannot always satisfy these two criteria for a given processing gain requirement and this is compounded by the fact that commercially available DMF devices are of length $N=2^n$ where $n$ is an integer (typically $N=8,16,32,64$). The control of the serial-parallel correlator is further simplified if $\frac{M}{N} = 2^x$ where $x$ is an integer and so we introduce this constraint.

As we wish to operate over a channel which could be subject to -20dB SNR, a processing gain of about 30dB would seem a sensible figure. Therefore a correlator length of $M=1024$ chips was chosen, and the DMF device length was chosen to be $N=32$. The 32 chip window gives a 6.5μs' excess delay time sufficient for a typical urban multipath scenario (see Chapter 4). The 1024 chip code is chosen such that the even and odd ACF sidelobes 31 chips either side of the main peak are very low to minimise interference with the multipath information. We also wish to minimise the maximum overall ACF peak sidelobe such that the code acquisition receiver phase is reliable. A smaller $M=64$ chip

† This is based on a chip rate of 5Mchip/s (i.e. ~4.8kbit/s x 1024 chip/bit).
code is also used for test and demonstration purposes. The choice of this code follows a similar selection process. However only the choice of the 1024 chip code is documented here.

As m-sequences are easy to generate and have reasonably good ACF properties it was decided to use the 60 possible 1023 chip m-sequences as a starting point for generating a new class of 1024 chip codes. The 60 m-sequences were generated using the 30 shift register polynomials of order 10 found in tables [45], plus their reverse polynomials which produce identical m-sequence properties. These m-sequences have a slight code imbalance, there being 512 ones and 511 zeros. It was decided to "stuff" an extra zero into the code to balance it and produce a 1024 chip code for our receiver demonstration.

Computer simulations were run to find the best position in each code to augment or "stuff" this zero, figure 3.8a. The simulation yielded 60 "good" codes based on the lowest sidelobe peaks around the main peak, in an even ACF sense. The odd ACF is not only determined by the code but also by its relative framing position, figure 3.8b. The framing position does not effect the even ACF. The 60 "good" codes previously found were tested for all 1024 framing positions using the same criterion as for the even ACF simulations. The best code of this group of 60 has thus been selected using both the even and odd ACFs (there were in fact two codes with identical properties due to the reverse polynomial). Only positive ACFs were used since negative ACFs have identical properties only the function is reflected about the time axis.

From the simulations it was found that the maximum overall even ACF sidelobe peaks (significant during acquisition) were all fairly similar and so the code selection was based entirely upon the sidelobes from which the multipath information is obtained. The best codes found from the initial simulations had maximum sidelobe values of 8/1024 (even and odd ACF) within the 31 chip range around the main peak and an overall maximum sidelobe value of 52/1024 (even ACF). These results have produced significantly better results than would have been obtained from random codes. However, this method restricts the selection of code to the subset of 60 codes produced from the even ACFs.
Figure 3.8 - a) Even ACF Bit Stuffing, b) Odd ACF Frame Positioning

and these will not necessarily produce the best properties from the odd ACF simulations. An exhaustive search of all the zero "stuffed" positions and all framing positions would require over 60 million ACF simulations in the form of a two dimensional search. Instead it was decided to approach the problem by reframing the code first. Then an extra bit was simply added to the end of this code to see if better codes could be found. Note that the m-sequences were generated using an initial shift register polynomial which is set to one and provides a frame reference position. Before barrel shifting the codes all have a framing position of zero.

Each order 10 m-sequence was reframed by barrel-shifting one bit at a time and a zero was added as the 1024th chip. The best framing position was then found for each code, the even and odd ACFs being found from the same simulation. It was found that better results were obtained if the last chip was the same as the first rather than simply a zero. The results from these latter simulations did, as was hoped, produce better codes, and the simulation time was halved. The maximum sidelobe peak around the main peak is now only 6/1024 for an odd ACF and 4/1024 for an even ACF. This is for the code
polynomial $7F9$ (hex) and a framing position of $1B5$ (hex). Even and odd portions of the ACF are illustrated in figure 3.9, the 31 chip portions around the main peak are shown magnified.

Finally, it may be possible to obtain slightly better codes than those given here but there seems to be no analytical method to find them. The only certain method of finding the optimal codes would be by a totally exhaustive search of all the permutations and there are over $10^{18}$ of them!

### 3.6 DSP Implementations

In order to confirm and demonstrate the performance of the DMF serial-parallel correlator, the architecture was implemented on a general purpose DSP card fitted to an IBM compatible personal computer (PC). The DSP card uses the Texas Instruments TMS320C25 DSP chip which is a 16-bit fixed-point processor with a 32-bit accumulator and a 100ns instruction cycle [46] [47]. The DSP board also contains a 16-bit DAC and a 16-bit ADC (the ADC runs faster at 12 bit resolution). This system board was supplied
by Loughborough Sound Images Ltd. In addition to the software which will implement the serial-parallel correlator on the DSP, some form of signal source must provide the system with a suitable analogue input. A custom hardware simulator was constructed for this purpose.

3.6.1 DSP Software

The serial-parallel architecture is constructed using a single parallel DMF. This was the first element of the code to be implemented on the DSP. The DMF is a FIR filter type structure and is easily programmed on the DSP which has an efficient multiply and accumulate (MAC) cycle due to the Harvard architecture used [48] [49]. The basic DMF pseudo-code algorithm is given below for a filter of length $M$.

\[
\text{loop:  } \quad \text{IN sample, ADC} \quad \text{;Get data sample from ADC}
\]

\[
\text{MPYK 0} \quad \text{;Zero multiplier and}
\]

\[
\text{ZAC} \quad \text{;accumulator.}
\]

\[
\text{RPTK } M-1 \quad \text{;Repeat M times, multiply coeffs}
\]

\[
\text{MACD coeffs, samples} \quad \text{;by samples (autodecrement) and}
\]

\[
\text{;accumulate with data move.}
\]

\[
\text{APAC} \quad \text{;Accumulate final result and}
\]

\[
\text{SACH output, I} \quad \text{;store it, then output it}
\]

\[
\text{OUT output, DAC} \quad \text{;to the DAC.}
\]

\[
B \text{  loop} \quad \text{;Repeat for next sample.}
\]

Although the system implemented on the DSP is a parallel correlator, it should be noted that the operation has been performed sequentially due to the serial nature of the processor. The MACD (multiply and accumulate with data move) instruction must be repeated $M$ times for each chip of output. In all it takes about $10+M$ 100ns machine cycles to produce one chip of output from this implementation of a parallel correlator.
The parallel correlator length is limited to 128 chips using this algorithm - due to the DSP internal memory page size. Serial-parallel architectures can of course be longer than 128 chips as long as the parallel DMF section does not exceed this limit. They should also run faster due to there being less processing to perform. When referring to serial-parallel correlators we quote the total correlation length first, followed by the parallel correlator length (or virtual device length if we are using a serial DSP device). Thus a serial-parallel correlator which performs a 256 chip correlation with a parallel correlator of length 8 will be referred to here as a 256/8 chip correlator, and the parallel correlator section performs 32 partial correlations before the final result can be accumulated.

One of the initial problems encountered when developing the code on the DSP was that of providing a PN signal source which could be synchronised to the DSP code. Normally synchronisation would be achieved using an externally generated interrupt, however this would present considerable problems during the code development stages. Developing and debugging the code requires the use of breakpoints and single stepping which must avoid the use of external interrupts and cannot be used with real-time input samples. A TMS320C25 simulation system was not available so it was decided to provide the "samples" internally without recourse to interrupts. Although the samples were artificially provided by files which were linked into the DSP code, this did allow the code to be developed fairly easily as it temporarily avoided the synchronisation problem and permitted free use of single stepping. Once the code was running correctly the samples are sourced from the ADC rather than the internal files and the external interrupt is re-enabled.

The development method outlined was used first of all to test the parallel correlator. A 31-chip m-sequence was used and the ACF output is shown in figure 3.10. The code for the serial-parallel correlator was then developed based on the parallel correlator code. The algorithm used is summarised in the flow diagram of figure 3.11. Note that \( N \) chips are correlated consecutively before the coefficients are changed. There are \( \frac{M}{N} \) sets of coefficients which must be used before the full result will be obtained at the output. The
counter is then reset, the recirculating delay line is cleared and a new correlation can begin. Note that the maximum sample rate is determined by the worst case algorithm cycle time. This will be the cycle which includes resetting the counter, the delay line, and the code coefficients. In a dedicated hardware system these operations would occur in parallel with the correlation function and there would be no delay penalty.

Figure 3.12 shows the output of a 256/16 serial-parallel correlator where the "samples" are provided internally. Note that the main peak can be seen to grow linearly. The output is only valid during the final 16 chips of the period. Before the code could be converted for synchronous real-time operation, a hardware signal source had to be constructed.
Figure 3.11 - Serial-Parallel Correlator Flow Diagram
3.6.2 The Signal Source

The signal source provides the DSP with 8 possible code outputs. The gain of the signal can be adjusted and an inbuilt noise generator circuit permits AWGN to be applied to this signal.

In addition to the analogue signal which is generated, there are also signals for chip synchronisation and bit period synchronisation. Both of these are provided as TTL signals. The chip synchronisation signal is used as an external interrupt for timing the operation of the DSP algorithm. The message period signal is used as a trigger for an oscilloscope so that the output from the DSP can be displayed, figure 3.13.

The 64-chip and 1024-chip codes which were detailed in Section 3.5 are the main codes used in the source. In addition to these we have their reverse polynomial partner which is an alternative code with the same "good" properties, and there is also an additional
randomly chosen code which could be used to measure typical cross-correlations. Also provided is a 4096-chip code which can be used as random data, and a square wave used for source test purposes. These are detailed in table 3.1.

<table>
<thead>
<tr>
<th>Name</th>
<th>Length</th>
<th>Polynomial</th>
<th>Frame</th>
</tr>
</thead>
<tbody>
<tr>
<td>RAND12</td>
<td>4096</td>
<td>1053h</td>
<td>0h</td>
</tr>
<tr>
<td>POLY10</td>
<td>1024</td>
<td>7F9h</td>
<td>165h</td>
</tr>
<tr>
<td>REVP10</td>
<td>1024</td>
<td>4FFh</td>
<td>240h</td>
</tr>
<tr>
<td>RAND10</td>
<td>1024</td>
<td>408h</td>
<td>0h</td>
</tr>
<tr>
<td>POLY6</td>
<td>64</td>
<td>43h</td>
<td>1h</td>
</tr>
<tr>
<td>REVP6</td>
<td>64</td>
<td>61h</td>
<td>38h</td>
</tr>
<tr>
<td>RAND6</td>
<td>64</td>
<td>56h</td>
<td>0h</td>
</tr>
<tr>
<td>SQUARE</td>
<td>2</td>
<td>-</td>
<td>-</td>
</tr>
<tr>
<td>SPARE</td>
<td>-</td>
<td>-</td>
<td>-</td>
</tr>
</tbody>
</table>

Table 3.1 - Codes Available From Signal Source

The signal source consists of three elements, the noise source, the code generator, and the gain section. The design of each of these elements will be briefly described in turn.
The noise source employs a 12V Zener diode reverse biased by 15V followed by two wideband operational amplifier stages. Each stage is isolated using d.c. blocking capacitors. The last amplification stage contains a preset potentiometer for adjusting the noise level (in terms of the amount of clipping due to the rail voltages).

The code generator uses an 8k by 8-bit EPROM containing 8 different codes, i.e. one on each data output line. The codes (see table 3.1) are stored in the first 4k bytes the remaining 4k is blank. Each code is an integer multiple of 4096 and is simply repeated the required number of times to fill the full 4k bytes. The codes are clocked out of the EPROM using a 12-bit counter driven by an external clock. Switch settings determine the code selected via a 8 to 1 line selector. With 1024-chip codes selected, counter bit 9 is provided as the trigger signal (bit clock), while the 64-chip signals use bit 5. Message data can be added to the signal via an EX-OR gate and this is enabled via a latch using the trigger signal. The output is retimed by another latch such that it lags the external clock input by half a clock cycle.

The gain section employs another wideband operational amplifier. Note that the noise source can be switched in or out. The noise source was preset to give 1% clipping on the ±10v ADC employed on the DSP system board. The code signal gain can be adjusted from zero to fully clipped using a potentiometer. With the noise on, this adjustment is effectively controlling the input SNR.

The signal source was designed for a chip rate of 100kchip/s. This is the maximum sampling rate of the ADC (at 12 bit resolution) and it was estimated that the 1024/32 chip serial-parallel correlator would approach this rate, i.e. it will achieve about 100 message bit/s. Figure 3.14 shows an oscilloscope trace with a portion of the 1024-chip code (upper) at 50kchip/s and the same signal but with the noise switched on (lower) at a SNR level of ~-15dB. Figures 3.15a and 3.15b show the same two signals respectively, only this time they are displayed on a spectrum analyser. Note that the central peak of figure 3.15a is the d.c. reference marker of the spectrum analyser and is not part of the
signal. The chip rate used in figure 3.15a was 50kchip/s and it can be seen that the noise bandwidth of figure 3.15b is considerably greater than this. The 3dB noise bandwidth was measured to be around 500kHz.

The DSP code was then modified to accept the sampled data externally via interrupt control. Two versions of the code were produced, one for the 1024 chip code, and the other for the 64 chip code also described. The only difference between the two versions is the different values of constants $M$ and $N$ and the different code coefficient files which must be linked into the main code.
Figure 3.15 - Frequency Spectrum of Signal Source Output

(vertical 10dB/div.; horizontal 20kHz/div.)
3.6.3 Correlator Performance

Due to the large dynamic range of the DSP arithmetic and the resolution of the ADC and DAC, it was expected that the measured performance of the correlators would be indistinguishable from the theoretical values. Any significant deviations from the predicted performance would be largely due to any errors in measurement or non-stationary nature of the signal source statistics.

The 64/8 chip correlator was tested first of all. The processing speed was limited by the ADC conversion time, nevertheless data rates in excess of 1.5kbit/s could be processed and without this restriction the speed is over 6kbit/s. The correlator’s bit rate can be generalised to an approximation obtained by counting the number of instruction cycles per bit, \( I \), (equation 3.3) and dividing by the instruction cycle time.

\[
I = \frac{(3N^2 + N)^M}{N} + 28 \text{ instruction cycles} \tag{3.3}
\]

The receiver synchronisation acquisition was achieved synthetically. First the DSP and the signal source were reset. The DSP would idle until the source was set to run and the first interrupt was sent to start the processing at the correct position. Figure 3.16a shows the 64/8 chip serial-parallel correlator input and output signals with random message data. Figure 3.16b shows a similar picture only the input SNR \( \sim 9 \text{dB} \). The output from the correlator clearly illustrates how the correlation output grows linearly while the noise adds in a root mean square (RMS) sense. Note that the signal can be seen rising out of the noise at the output.

The SNRs of the input and output were measured and used to calculate the processing gain. The input PN signal was measured on the oscilloscope as was the output signal peak. The signal was then switched off and the noise turned on. The DSP input and output noise levels were measured on an RMS voltmeter while the system was running a 64 chip parallel correlation, since this provides a continuous output power. The results are given in table 3.2.
Figure 3.16 - 64/8 Chip Serial-Parallel Correlator Output

(a) no noise

(b) with noise (SNR=-9dB)

(upper traces 5v/div., lower traces 0.5v/div.; timebase 0.5ms/div.)
<table>
<thead>
<tr>
<th></th>
<th>Input</th>
<th>Output</th>
</tr>
</thead>
<tbody>
<tr>
<td>Signal (V)</td>
<td>3.3</td>
<td>1.68</td>
</tr>
<tr>
<td>Noise (V)</td>
<td>3.1</td>
<td>0.20</td>
</tr>
<tr>
<td>SNR (dB)</td>
<td>0.543</td>
<td>18.486</td>
</tr>
<tr>
<td>Processing Gain (dB)</td>
<td>17.94</td>
<td></td>
</tr>
</tbody>
</table>

Table 3.2 - Processing Gain for 64 Chip Correlator

Expected processing gain $10 \log_{10}(2^6) = 18.06$dB

A similar procedure was followed for the 1024/32 chip correlator, this was capable of processing data at about 100bit/s. Figure 3.17a and 3.17b show the input and output signals without and with noise, again the signal had a random message content. The peaks are very narrow and are a little difficult to see after photographic reproduction. The results for this correlator are given in table 3.3.

<table>
<thead>
<tr>
<th></th>
<th>Input</th>
<th>Output</th>
</tr>
</thead>
<tbody>
<tr>
<td>Signal (V)</td>
<td>6.00</td>
<td>6.05</td>
</tr>
<tr>
<td>Noise (V)</td>
<td>2.67</td>
<td>0.0833</td>
</tr>
<tr>
<td>SNR (dB)</td>
<td>7.033</td>
<td>37.225</td>
</tr>
<tr>
<td>Processing Gain (dB)</td>
<td>30.19</td>
<td></td>
</tr>
</tbody>
</table>

Table 3.3 - Processing Gain for 1024 Chip Correlator

Expected processing gain is $10 \log_{10}(2^{10}) = 30.10$dB.

There will be a loss of processing gain due to quantisation, however, this will be negligible for the 12 bit resolution. The errors due to the measurements are much more significant and can have a maximum of about 4%. The errors found were about -2.5% and +2%
Figure 3.17 - 1024/32 Chip Serial-Parallel Correlator Output

(a) no Noise

(b) with Noise (SNR=-20dB)

(upper traces 5v/div., lower traces 0.2v/div.; timebase 10ms/div.)
respectively. More accurate results could have been made by logging the DSP output, however, the results do prove that the correlators were operating very close to the theoretical values.

This test has illustrated the performance levels which can be achieved using general purpose DSP devices. The 1024/32 chip correlator with a data rate of 100bit/s can only be considered to be suitable for demonstration purposes, however if only about 18dB of processing gain were required, the data rate could be increased to over 6kbit/s with the 64/8 chip correlator. We must however remember that when complex I&Q processing is required this effectively halves these rates. If the system employed were upgraded to the new TMS320C50 with an instruction cycle of 50ns, the processing speed could be doubled. This rate would be close to those required for acceptable toll-quality speech based on codebook excitation linear prediction (CELP) techniques [50]. Even faster DSPs are planned for the future, so it may soon be possible to consider using DSPs for complete spread spectrum receiver designs.

Table 3.4 shows the performance obtained for the DSP based correlator in terms of its throughput, and the acquisition times for code synchronisation. We have ignored the limitations of the analogue interface. As the mode of operation tends towards a full parallel configuration, the bit rate decreases. This is due to the increase in computational complexity required for parallel operation. The code synchronisation timings are for a serial search and assumes that the ACF peak is always detected. These timings do not indicate a large reduction in synchronisation times for serial-parallel operation relative to serial operation. The reason for this is that the bit rate is reduced when we increase the parallel correlation length, i.e. we have reduced the maximum number bit periods over which the search must be made, but each bit period is now longer. This would not be the case for hardware implementations since the bit rate does not decrease with complexity, unlike the DSP system. Note that the performance of the 1024/32 correlator is improved by two orders of magnitude if we use the A100 device instead of the TMS320C25.
3.7 Reconfigurable Receivers

The serial-parallel correlator which was developed on the DSP is designed such that the correlation length $M$, and the virtual device length $N$ are set using two constants. A 64 chip correlator could be configured as 64/1, 64/2, 64/4, 64/8, 64/16, 64/32 or 64/64. Since the virtual device length of a 64/1 chip correlator is one, it is therefore just a serial correlator, figure 3.18a and figure 3.18b. Similarly a 64/64 chip correlator is in fact a parallel correlator, figure 3.19a and figure 3.19b. The modes of operation (i.e. serial/parallel/serial-parallel) are all achieved with the same DSP code. It was found that these modes can be altered during operation without a break in transmission. Thus it is possible to reconfigure the receiver’s architecture without loss of data (at most one bit will be invalid), since the receiver remains synchronised. Figure 3.20 shows the 64 chip PN code on the upper trace, the lower three traces correspond to 64/1 chip (serial) correlator, 64/64 chip (parallel) correlator, and 64/8 chip (serial-parallel) correlator shown respectively without any data content or noise.

---

Table 3.4 - DSP Correlator Performance

<table>
<thead>
<tr>
<th>$M/N$</th>
<th>Max. bit rate (approx.)</th>
<th>Max. sync time (ms)</th>
</tr>
</thead>
<tbody>
<tr>
<td>64/1</td>
<td>35kbit/s</td>
<td>1.8</td>
</tr>
<tr>
<td>64/8</td>
<td>6kbit/s</td>
<td>1.3</td>
</tr>
<tr>
<td>64/64</td>
<td>800bit/s</td>
<td>1.25</td>
</tr>
<tr>
<td>1024/1</td>
<td>2.4kbit/s</td>
<td>430</td>
</tr>
<tr>
<td>1024/32</td>
<td>100bit/s</td>
<td>320</td>
</tr>
<tr>
<td>1024/32$^\dagger$</td>
<td>9.6kbit/s</td>
<td>3.3</td>
</tr>
</tbody>
</table>

$^\dagger$ Predicted Inmos A100 receiver performance (10MHz sampling rate).
Figure 3.18 - 64 Chip Serial Correlator Output

(a) no noise

(b) with noise (SNR=-9dB)

(upper traces 5v/div., lower traces 0.5v/div.; timebase 0.5ms/div.)
Figure 3.19 - 64 Chip Parallel Correlator Output

(a) no noise

(b) with noise (SNR=-9dB)

(upper traces 5v/div., lower traces 0.5v/div.; timebase 0.5ms/div.)
The implications of having an on-line reconfigurable receiver could be far reaching. Such a receiver would be capable of adjusting itself to meet operating requirements such as data rate, code length for security, multipath interference, co-channel interference and noise, etc. The receiver is fastest when processing in a serial mode, however it is slow to synchronise and gives no information regarding multipath responses, co-channel interference, and noise. The parallel receiver mode is slower in operation but can synchronise more quickly and provides detailed information about the channel over the entire bit period. The serial-parallel receiver can be configured to operate almost anywhere between these modes to provide the necessary information while maintaining a suitable speed of operation. As the operational requirements change the receiver can adapt its architecture accordingly.
It is often desirable to synchronise a receiver quickly. During this synchronisation process no data can be received and any error correction circuits are redundant. Once synchronisation is acquired we may then desire error correction on the data and the synchronisation acquisition circuits would be redundant (only fine synchronisation tracking is required). If a reconfigurable receiver is employed, a parallel correlation can be used for a fast acquisition phase, the receiver can then be configured as serial or serial-parallel design which would free adequate processing time for error correction (or other tasks such as data decoding or diversity combining).

It is equally possible to adapt the correlation length of the receiver by changing the constant, $M$, which will adjust the processing gain. Of course a suitable code of that length must be in place before this can be done. However it is possible that a single code of suitable length could be used for all correlation lengths. If the code employed is larger than that desired, only a portion of it need be used for each correlation, the next portion is used for the next correlation. This eliminates the need to change to a new PN code which would inevitably lead to a short break in transmission and a need to resynchronise.

No experimental work was conducted on reconfigurable correlation lengths $M$, only work on the reconfigurable virtual device length $N$ was conducted. The main reason for this was that the signal source was designed to provide synchronisation for the oscilloscope at the PN code period and so would require to be modified to cater for this. Also the PN codes and the DSP code were not developed with a reconfigurable receiver in mind. The code requirements for a receiver with a variable processing gain are slightly different to those which were used. Very slight modifications would also be required to the DSP code such that the code length can be different from that of the correlation length $M$.

The virtual architectures which can be achieved using the reconfigurable receiver could be useful for systems with automatic fall back, or whose mode of operation also depends the type of transmitter being used. The use of post-processing such as equalisation (or a RAKE receiver) may also benefit from an adaptable receiver architecture. It also
appears feasible to multiplex different architectures such that the DSP could decode signals from several different types of source or from the same source employing a diversity scheme. Of course the benefits of DSP reconfigurable receivers need not be restricted to spread spectrum systems.

3.8 Summary

This chapter has explained the operation of the serial, parallel and in particular the serial-parallel architectures used for spread spectrum systems. It was shown that the serial-parallel receiver, first developed by Hunsinger, was capable of increasing the processing gain available from SAW AMFs by about 10dB but had limited code diversity. Synchronisation of such systems was however considerably faster than that of serial receivers. The problem of code diversity could be overcome by the use of programmable AMFs or SAW convolvers, however the processing gains were still constrained to about 40dB due the devices’ self-noise.

The DMF serial-parallel correlator architectures developed here have overcome the processing gain constraints imposed upon the analogue receivers. The speed of these digital systems is now approaching that of the SAW receivers. Special codes have been selected for use with these new DMF correlators such that the sidelobes are minimised and will not greatly interfere with any channel information contained within them. Table 3.5 summarises the properties of the various spread spectrum receiver types.

A DSP based serial-parallel correlator was constructed and fed with a "real" signal including noise. The measured performance was shown to be near ideal. The speed of the DSP system was considerably slower than that which can be achieved using dedicated fixed architecture hardware. However it is suggested that very soon spread spectrum receivers may be implemented on DSPs and produce respectable performances. One of the benefits of a DSP based receiver was found to be its ability to be reconfigured to a different virtual architecture without necessitating a break in data transmission.
In conclusion, it was seen that the DMF based serial-parallel correlator represents a very efficient architecture with which to construct a spread spectrum receiver. It benefits from the limited complexity of the serial correlator and the fast synchronisation of the parallel correlator. It is also capable of providing a portion of the channel’s delay profile which can be engineered to be large enough to capture all of the significant channel multipath energy. It will be seen, in the following chapters, that this is a very useful property.
Chapter 4

ANTI-MULTIPATH RECEIVER DESIGNS AND THE UHF MOBILE CHANNEL

This chapter explores receiver techniques which can be used to compensate for the severe losses which are attributed to the multipath dispersion in a communication channel [51]. First we describe the design of a RAKE receiver which can be used to increase the SNR and reduce the fade duration of a received multipath signal. Then the UHF communication channel itself is investigated in an urban type environment. This study includes the channel characterisation for a mobile receiver, and the development of a realistic software channel simulator.

4.1 Overview of the RAKE Receiver

The RAKE receiver concept was developed to enhance the performance of an early DS FSK spread spectrum system. The first RAKE system was constructed in the mid 1950s by MIT's Lincoln Laboratory. The idea of the RAKE receiver is simply to recombine the various multipath signal returns such that the overall SNR is improved. This is achieved by combining, or "raking", the strongest of the received signals such that they are brought back into alignment with the main component and are weighted according to their signal strength [52]. A tapped delay line or transversal filter may be used for this purpose [53]. If maximal-ratio combining (MRC) [54] is used the resultant receiver is, in fact, a form of channel matched filter. This assumes that there is no ISI, and that all of the resolved multipath signal is used. There will be no significant ISI if the delay spread caused by dispersion is less than the data symbol period, and this is normally the case.
The RAKE receiver uses the post-correlation signal to estimate the channel's impulse response. The preceding correlator will already have resolved the received signal into the multipath components each separated by the chip period $T_c$ (assuming a parallel or serial-parallel architecture). The RAKE filter uses a tapped delay line, figure 4.1, with delay elements of $T_c$ seconds, to recombine the multipath components which have been separated by the dispersive channel. Each tap must be weighted according to the estimated multipath component observed at that particular delay. Note that the multipath amplitude (given as $a_k$ for the $k^{th}$ component) must be estimated, as well as the phase ($\theta_k$). Both the signal phases and their delays must be known in order to combine the signals coherently.

We effectively correlate the received signal with the complex conjugate of the estimated impulse response. Estimated values in figure 4.1 are represented by the hat "^\wedge" symbol. The RAKE is a channel matched filter, and so the output waveform shape is of no significance. Its SNR is, however, maximised at the peak of the waveform if the filter is correctly matched.

The impetus for using an anti-multipath receiver architecture is based on our desire to transmit a very low signal power. If we are losing significant amounts of energy due to multipath effects it is desirable to retrieve some of this power rather than increase the transmitter output to compensate for it. Such a scheme is possible when operating over a wideband urban UHF channel. The RAKE receiver proposed has the benefit of independent fading such that the multipath returns provide an inherent signal diversity. This enables the error burst effects of signal fading to be significantly reduced due to a reduction in the fade durations. Thus we have two mechanisms by which the bit error
rate (BER) of the receiver can be improved, first of all we can improve the receiver SNR by using maximal-ratio combining, and secondly we have multipath diversity due to the independently fading multipath signal components [55]. The net effect is that the multipath fading channel can often provide better performance than could be obtained from a non-dispersive fading channel. Note that increasing the transmitter power will improve the receiver’s mean SNR but it will not combat the fading since no independent diversity technique has been applied. Any increase in transmitter power also adversely affects the channel reuse efficiency [56] in cellular systems.

4.1.1 Maximal-Ratio Combining

In order to understand fully the operation of a RAKE type receiver it is useful to consider the operation of maximal-ratio combining. First we consider the combining of two received signal components which are part of the same data bit and have been received on the same locally stationary channel. Thus they will both be subject to rms noise signals of strength $N_1$ volts. The first signal has a voltage of $S_1$, the second has a voltage of $S_2$. We wish to determine the ratio with which we should combine these two signals in order to maximise the overall SNR of the combined signal. We will assume that the signals combine linearly, while the noise will add as rms values. Assigning $R$ as the variable by which $S_2$ will be multiplied before the signal is combined, we obtain an expression for the combined SNR.

$$SNR = \frac{S_1 + S_2 \cdot R}{\sqrt{N_1^2 + (N_1 \cdot R)^2}}$$  \hspace{1cm} (4.1)

This is differentiated with respect to $R$ and set to zero to find the optimal ratio

$$\frac{d(SNR)}{dR} = \frac{N_1(\sqrt{R^2 + 1})S_2 - (S_1 + S_2R)N_1 \cdot \frac{2R}{\sqrt{R^2 + 1}}}{N_1^2(R^2 + 1)} = 0$$ \hspace{1cm} (4.2)

$$S_2(R^2 + 1) = R(S_1 + S_2R)$$ \hspace{1cm} (4.3)
This result indicates that the signals should be weighted by their relative signal strength prior to combining and hence it has been termed a "ratio squarer" [57]. This is also equivalent to combining according to the signal’s SNR since the noise power is constant. The result indicates that SNR improvements can be made by using diversity combining. Simply selecting the strongest signal (selective diversity) is sub-optimal.

The ratio squarer result can be generalised to show the effect of any linear combination of diversity signals where the noise powers need not have the same value. Consider the combining of \( N \) signals \( f_i(t) \) such that \( f_i(t) = s_i(t) + n_i(t) \), where \( s_i(t) \) and \( n_i(t) \) are the individual instantaneous signal and noise voltages respectively, then

\[
f(t) = a_1f_1(t) + a_2f_2(t) + \ldots + a_Nf_N(t) = \sum_{i=1}^{N} a_if_i(t)
\]

where \( a_i \) is the weighting coefficient for the \( i \)th term of the combiner.

Now let \( P \) be the SNR of \( f(t) \) and \( p_i \) be the SNR of \( f_i \). Using the Schwartz inequality it can be shown [58] that,

\[
P \leq \sum_{i=1}^{N} p_i
\]

The equality is satisfied if the weighting coefficients \( a_i \) are chosen such that,

\[
a_i = K \frac{\sqrt{s_i^2}}{n_i^2}
\]

where \( K \) is any arbitrary real scale factor common to all \( N \) samples \((K \neq 0)\).

The maximum SNR obtainable is therefore equal to the sum of the individual SNRs and we have obtained the optimal condition given by maximal-ratio combining. This result
is general enough to also hold true for complex signals and coefficients. The coefficient derivation in this case uses the complex conjugates of the signals components, and $K$ can be a non-zero complex number.

Note that if the noise powers on all diversity signals are not equal, then the optimal coefficient weights are not always equated such that the combining is in accordance with the individual SNRs. Equation 4.7 indicates that both the signal and noise powers must be known, or at least be known relative to the other signal and noise powers involved in the linear combining.

This result is related to the mechanism which describes the conditions for matched filtering, indeed the RAKE receiver, when satisfying the MRC conditions is a form of channel matched filter. It should be clear that the RAKE receiver relies upon optimal combining of the multipath signal components to improve the effective SNR of the receiver. We depend on these components being independent so as to obtain signal diversity. Note that MRC is only optimal if the interference and noise can be considered to be Gaussian in nature.

4.1.2 Other Combining Techniques

MRC when applied to the RAKE receiver assumes that the impulse response of the channel is known. This is rarely the case and so it must either be estimated, or if that is not possible other sub-optimal combining techniques must be used. If we estimate the channel the MRC condition is only approximated and so it too will be sub-optimal to some degree.

Selective diversity combining can be used to reduce the effects of fading. It operates on the principle that not all diversity branches (or multipath components) will fade at the same time. If the largest signal component is always selected, the others being discarded, then we have increased the mean signal SNR and reduced the fading. Selective diversity combining is fairly complex to apply to a mobile radio channel which has several diversity branches and fast fading signals. A more practical form is switched diversity combining.
whereby the strongest signal is only selected when the current one drops below a predefined level. These techniques are most commonly used in two branch diversity systems [59] such as a dual antenna receiver. It is possible to use selective diversity to choose more than one diversity branch. We can then apply some other form of diversity combining to the remaining branches. We therefore produce a hybrid diversity combining technique. Neither selective or switched diversity are particularly well suited to multipath combining.

Equal-gain combining (EGC) allows signals to be combined without them first being weighted. This technique is useful in that there is no need to form a channel estimate, however the phase of the each signal component must be obtained for coherent combining. Equal-gain combining works best if the signal components are of similar mean SNR or for very high SNR. Again this technique is particularly attractive for dual antenna systems but can also be used in a sub-optimal RAKE receiver. In the following chapter we will use a modified and simpler form of EGC which combines the signals taking only account of the polarity rather than the phase of the signal.

Another combining technique which can be used with the RAKE receiver is differential phase combining (DPC). This technique relies on DPSK signalling being used. The post-correlation signal which contains the multipath profile is used to set the RAKE tap weights directly without averaging or decision feedback. The multipath profile for following bit will then be fed into the RAKE filter which uses the weights set by the previous bit. The polarity of the output provides an indication of whether the two bits were the same or different thus providing the decoded DPSK data decision. DPC is particularly effective for high SNR signals where it approaches the ideal MRC performance. It does, however, suffer in high noise channels since there can be no averaging to form a channel impulse response estimate.
4.2 Alternative Architectures

![Diagram](image)

Figure 4.2 - Matched Receiver Block Diagrams

a) two stage matched filter receiver, b) single stage matched filter receiver

The digital RAKE receiver is preceded by a correlator, and so the complete anti-multipath receiver can be viewed as a correlator (or PN code matched filter) followed by a channel matched filter. The net result of this architecture is to produce a matched filter which is matched, not to the transmitted signal, but to the received signal. If the impulse responses of these two filters are convolved we obtain a single matched filter, figure 4.2. The filter stages are of length $p$ and $q$ respectively, so the single stage filter will be of length $p+q$. If, as we would expect, $p \gg q$ it is then possible to truncate the single stage filter to be of length $p$. This truncation would have a limited effect on the output, as we expect a decaying multipath delay profile, but it would certainly result in a sub-optimal structure.

It was hoped that having a single matched filter would reduce the receiver complexity. The problem with this approach is that the output from this filter gives little indication of the impulse response of the channel. When the receiver is properly matched the output signal should be symmetrical, however, there are many solutions which can provide this symmetry without it being a true matched filter design. Of course the solution is always constrained by the fact that it must be composed of pure PN codes only.
The shape of the output does give some indication as to the channel impulse response but adaptive algorithms would have to be given good initial conditions before having a reasonable chance of converging towards the correct solution. The only sure method of adapting such a receiver towards a good solution would be to use it first to sound the channel by using the "pure" PN code coefficients and then reprogramme it with the convolved PN code and estimated impulse response coefficients. This requires a signalling overhead for the sounding signal which may not always be acceptable.

Study has indicated that even if the system of figure 4.2b were capable of performing as well as the conventional two stage approach, the complexity in terms of the signal processing required would undoubtedly exceed that of the two stage system and no useful advantage could be obtained from its use.

A conventional adaptive filter approach to the problem is not really feasible if operated on the received signal due to the very poor SNR on the channel. The convergence of the filter would be limited by the noise floor on the signal and the amount of averaging required to produce comparable results would be at least equivalent to the processing gain already provided by the correlator. Operating an adaptive equaliser on the post-correlation signal would be more sensible but is largely unnecessary. There is already an estimate of the channel impulse response available from the instantaneous excess delay profile provided by the post-correlation signal. An adaptive equaliser is given a priori knowledge of the desired output and uses iterative techniques to converge upon the channel's impulse response. The RAKE on the other hand uses a deterministic technique based on a posteriori knowledge of the channel provided directly from the post-correlation signal.

4.3 The UHF Wideband Channel

Radio communication channels are subject to propagation path losses, noise and interference. Free space signal strength decays according to an inverse square law, however, urban channel power losses are generally more accurately modelled by an
inverse fourth power relationship. The path losses observed are characterised by the topography of the terrain which causes scattering and reflections of the signal. This multipath phenomenon causes the received signal to be characterised in delay, phase, and magnitude. Thus a signal transmitted over a multipath channel will arrive as a group of signals with delays determined by the lengths of the propagation path routes taken. The magnitude of the received signal is determined by the vector addition of all signals arriving at that time instant, and the resultant phase shift is governed by the path lengths and wavelength of the component signals. Each element of the multipath signal will also have a distinct angle of arrival at the receiver dependent on the locations of the scatterers.

For a wideband channel, such as that used in spread spectrum systems, the channel cannot be modelled as a single fading statistic for the received signal at the given frequency, as it often can be for narrowband signals. The model must take account of the individual fading multipaths rather than an unresolvable conglomerate of paths. The uplink and downlink on mobile channels tend not to be reciprocal, but both can be modelled by the same parameters if they are on the same carrier.

4.3.1 Channel Characterisation

Although the statistics of the channel are time-variant when viewed over a short period, they can be considered to be wide-sense stationary (WSS). This indicates that although we do not have local stationarity, when viewed over long periods the statistical characteristics will average to a stationary value. This is useful as it allows us to simulate the channel using our knowledge of these WSS parameters and the nature of the local variations.

The wideband multipath channel model can be represented by a simple mathematical model. If we transmit the real signal \( \text{Re}\{s(t)e^{j\omega_0 t}\} \) where \( s(t) \) is a bandpass signal transmitted on the carrier \( \omega_0 \). The isotropically received complex low-pass signal envelope centred on the carrier will be given by
\[ r(t) = \sum_{k=0}^{\infty} a_k \cdot s(t - \tau_k) \cdot e^{j0_k} + n(t) \]  

where \( a_k, \tau_k, \) and \( \theta_k \) are the amplitude, delay, and phase of the \( k \)th propagation path. The additive noise term is \( n(t) \) and is often assumed to be AWGN.

This simple model does not in itself describe the channel’s characteristics, only the mechanism by which it can be represented. The model is completed once the statistics of all of the variables have been defined. There have been various attempts to complete this model. The following subsection attempts to summarise concisely the notable work in modelling a wideband UHF channel. The main parameters of the channel model are highlighted.

### 4.3.2 UHF Wideband Channel Modelling

The first published data on the UHF multipath channel appears to be that of Young and Lacy [60]. This paper is useful in that it clearly illustrates the type of multipath profiles which are obtained in an urban environment. The profiles were obtained by transmitting pulses on a 450MHz carrier from the top of a building to a slow moving vehicle in New York City. Some of the multipath profiles obtained from this work are shown in figure 4.3. Subsequent work has attempted, not only to measure the channel, but to use this data to create a suitable mathematical model to describe it [61] [62] [63] [64] [65] [66]. The channels which were initially studied are predominantly those of urban areas of cities in the United States.

It is evident from these studies that the fading of the multipath signals can be accurately modelled by a Rayleigh distribution, i.e. \( a_k \) is governed by Rayleigh statistics. Alternatively a Rician distribution is widely used for channels with fixed path lengths (e.g. for fixed channels, and indoor office or factory environments). Long term fading effects including shadowing are better described by a log-normal distribution.
It was proposed that the average excess delay profiles could be modelled by a Poisson process [67], however this did not entirely agree with all of the analysed data so log-normal and modified Poisson statistics have also been proposed [68] [69]. The excess delay profiles tend to vary considerably from location to location which may explain some of the difficulty in defining them. However, all the proposed profile models imply some sort of exponential decay in average path strength, \( \{a_t\}_0^\infty \) against the excess delay times, \( \{\tau_t\}_0^\infty \). There also seems to be general agreement that the majority of the significant urban multipath power is contained within a 5μs delay period.

The received carrier phases, \( \theta_k \), in all of the studies are assumed to be independent random variables uniformly distributed over \([0, 2\pi]\). There appears to be no evidence to suggest that this is not the case [70]. This is intuitively sensible considering the wavelengths of the UHF signals and the wide geographic areas in which they propagate.

In the 1980s there were several studies conducted in the United Kingdom. The work of Bajwa and Parsons attempted to describe the mechanics of multipath propagation [71]. A companion paper [72] presents a series of scattering plots obtained from a moving receiver, these show not only the excess delay, but also the Doppler shift. The scattering
plots are quite revealing and indicate that the Doppler shift is largely attributed to the vehicles speed at longer excess delay times. Whereas local scatterers cause a variety of Doppler shifts at most frequencies up to that given by the vehicle speed, for short delay times. Bajwa went on to produce a simulator based on these results [73]. This did allow for variations in the vehicle speed and used measured data directly using a sampling rate which was dependent on the vehicle speed.

High resolution channel sounding has been conducted using a swept time-delay cross-correlation (STDCC) technique [74] which automatically averages several delay profiles obtained at a given location. This method was used in a study which shows that the degree of time correlation between the different multipath delay separations are almost totally uncorrelated for small-scale amplitude fluctuations (independent Rayleigh fading) but more highly correlated for large-scale fluctuations (shadowing) [75].

A very useful and fairly simple model has been proposed under the European COST 207 programme [76] and was developed in order to provide standards against which the GSM [77] digital mobile radio system can be tested [78]. The model provides a number of different averaged excess delay profiles for characterising various types of terrain (e.g. urban, rural-hilly, etc.) and allows for the speed of the mobile in determining the fade (Doppler) rate.

4.3.3 Coherence Bandwidth, Frequency Selectivity, and Diversity

We have seen how the received signal components can be combined and used to improve the system’s SNR. It is assumed that each of the multipath components is largely independent of the others, i.e. there is negligible correlation between the relative powers of adjacent multipath components. Thus we have uncorrelated scattering (US) which is due to a fairly random topology of reflectors and scatterers around the transmitter and receiver terrain. We have already noted that the channel can be considered to be WSS, thus we have a wide-sense stationary uncorrelated scattering (WSSUS) channel. The inherent diversity of the multipath profiles may be considered to be time diversity since
the independent signal components are separated by different time delays. However the mechanism present which causes this inherent diversity on the channel can also be explained in terms of frequency diversity due to time-frequency duality.

The coherence bandwidth of a channel is given by the maximum bandwidth over which two CW signals will be highly correlated. If the coherence bandwidth of the channel, $B_c$, is much smaller than the signal bandwidth, $W$, the received signal will be severely distorted by the channel and we have a "frequency selective" channel. Frequency selectivity is of fundamental importance to the operation of the RAKE receiver which requires the wideband spread spectrum channel to satisfy the following condition

$$W \gg B_c$$

The coherence bandwidth can be determined from the excess delay profile [79] since it is linked by the Fourier transform

$$C(f) = \int_{-\infty}^{\infty} A(\tau)e^{-j2\pi f \tau} \cdot d\tau$$

$C(f)$ is the correlation coefficient between the two frequencies separated by $\Delta f$ hertz. $A(\tau)$ is the excess delay function and $\tau$ is the delay in seconds. This relationship is illustrated by figure 4.4. Note that the useful excess delay, $T_m$, is related to $B_c$ such that

$$B_c \propto \frac{1}{T_m}$$

$B_c$ is quoted for a given correlation coefficient, this is usually 0.5, but sometimes 0.9 is also used.

Consider the transmission of a message signal at a rate of $I/T$, where $T$ is the message period. If the bandwidth of the channel $W$ is greater than $B_c$ the message signal will be subject to different attenuations and phase distortion across this frequency band. The received message signal will therefore be subject to variations in power and if the channel
is time-variant, the signal will exhibit fading. If the channel is not frequency selective it will have the same attenuation and phase shift across the frequency band. Any multipath present will therefore fade together as a single signal and provides no inherent diversity.

We now consider the frequency selective channel. Individual multipath components will be resolvable since the frequency selective channel satisfies the diversity constraint [80]

\[ W > B_c \]  

(4.12)

For any given delay spread there will be \( W \cdot T_m \) independent resolvable signal components separated by \( I/W \). Thus the maximum order of diversity obtainable is given by [81]

\[ L = W \cdot T_m \]  

(4.13)

and thus

\[ L \propto \frac{W}{B_c} \]  

(4.14)

The inherent, or multipath, diversity as it is sometimes termed [82] can therefore be seen to be due to the frequency selective channel. Note that conventionally frequency diversity is achieved when the separate signal carriers are at least \( B_c \) apart. In wideband frequency selective channels the signal obtains an inherent diversity by the same mechanism, only
we do not physically separate the independent carriers. The diversity components of the signal can be resolved using a spread spectrum receiver. Higher chipping rates can resolve the signal into more multipath components. Each multipath signal component is a conglomerate of "rays" with path differences separated by no more than $\Delta d$ meters. Figure 4.5 shows two ellipses representing the positions of scatterers for path lengths $d_1$ and $d_2$. This diagram assumes that the rays are scattered only once, and that the transmitter and receiver form the foci of both ellipses. We can see that

$$\Delta d = d_2 - d_1$$

(4.15)

and this path difference is dependent on the chipping rate, such that

$$\Delta d = c \cdot T_c$$

(4.16)

where $c$ is the speed of electromagnetic wave propagation in free space. This gives a path difference of about 60m with a 5MHz chipping rate. This is equivalent to about 180 wavelengths of a 900MHz carrier.

Figure 4.5 - Multiple Ray Propagation Paths
4.4 The Channel Simulator

The delay profile which we will use in our model has been interpolated from the curves given in the GSM COST 207 literature [76] for a typical urban scenario. The GSM data uses 6 or 12 unevenly spaced taps. The model we have adopted uses up to 32 evenly spaced taps. In addition to the exponential decaying profile, the first two taps have been chosen such that there is no direct line-of-sight path, i.e. these two taps are of lower gain than the following one. Independent Rayleigh fading is multiplied by the excess delay gains for each tap. White Gaussian noise is then added to each tap where the noise gain is given by the mean SNR which is defined for the tap with the largest mean signal power.

The channel simulator which was constructed is based on software developed by Newson [83] for work on channel equalisation for the GSM mobile radio system, but was extensively modified for this study. A block diagram of our modified simulator is given in figure 4.6. Its operation can be described as follows. The simulator is based on a tapped delay line model, taps being spaced one chip period apart. The output from each tap is multiplied by a time varying shaped noise factor which characterises the fading. It is then weighted by a further gain, $a_k$ (for an $N$ tap model $1 \leq k \leq N$), which is the average

![Figure 4.6 - The Channel Simulator Model](image-url)
multipath signal strength expected at that delay. The signals from all of the weighted
taps are combined in a complex summer and complex channel noise is added according
to the noise power specified. Note that, even with a real input, the model produces a
complex output. The phase at the output will be uniformly distributed due to the uniform
phase distribution of the noise generators on the I and Q channels.

Rayleigh fading is produced by virtue of the fact that we have independent Gaussian
fading on both the real and imaginary weighting components. The filter used to
characterise the fading is a second order biquadratic IIR filter [84], figure 4.7, which
determines the rate and type of fading. We refer to this filter as the "noise shaping filter".
The transfer function for this filter is given by

\[ H(s) = \frac{\omega_n^2}{s^2 + 2\zeta\omega_n + \omega_n^2} \]  

In the z-domain the filter poles \( c_1 \) and \( c_2 \) are calculated for a given damping factor \( \zeta \)
using the bilinear transform. The transfer function of the filter can thus be given as

\[ H(z) = \frac{1 + 2z^{-1} + z^{-2}}{1 + c_1z^{-1} + c_2z^{-2}} \]  

The maximum delay spread which the model has been designed to simulate is 6.5\( \mu s \). There are 32 taps each separated by approximately 0.2\( \mu s \) which was derived from the
reciprocal of the 5Mchip/s system chip rate.

The excess delay profile used in our channel model is shown in figure 4.8. The frequency
correlation function in figure 4.9 was obtained by zero padding the excess delay profile
and then taking a discrete Fourier transform (DFT) of this extended profile. Note that
the coherence bandwidth for a correlation coefficient of 0.5 is about 600kHz, compared
with the signal bandwidth which is in excess of 5MHz, i.e it is frequency selective.
The delay profile used, figure 4.8, can only be considered to be typical. Different terrain topography will produce a different delay profile, essentially this profile is an average of numerous measured profiles. Thus individual profiles produced by the model may differ from the instantaneous measured ones on a channel, but should produce similar results when averaged.
Figure 4.9 - The Derived Frequency Correlation Function

Figure 4.10 - Average Excess Delay Power

Figure 4.10 shows the average excess delay power received on the 32 possible taps of the RAKE receiver. When the post-correlation signal of the simulated receiver is
averaged we can observe this WSS profile. The decay is exponential on a linear power scale and is approximately 5dB/μs. The first two taps have less amplitude than the third, since we assume that there is no direct line of sight between the transmitter and receiver.

Slow fading, often implemented as log-normal fading of the delay profile [85], has not been directly included in this model. This type of fading caters for the receiver shadowing and other long term effects, which vary widely in standard deviation and depend on the levels of urbanisation in the area [86]. For fast (100Hz Doppler) fading our model allows the noise shaping filters to pass a low level of slower fading. This will ensure that the received power levels from all delays will change slowly. Fading of the entire signal power will be much less significant than the fast independent multipath fading, which is the desired situation.

Figure 4.11 - Low-Pass Urban Scenario

In order to simplify the simulations it was decided to identify two urban channel model types, one for a hand-held receiver and the other for a vehicle based receiver, and to use these exclusively. The two channel scenarios are illustrated by figures 4.11 and 4.12 respectively. The hand-held channel of figure 4.11 will be subject to slow and unpredictable fading due to the MS being able to freely wander through the pattern of deep fades and signal peaks set up by the base station (BS) and the surrounding areas.
The spatial distribution statistics of the fades is governed by the transmitter frequency. Thus the fading is modelled as a low-pass characteristic which allows the signal to fade slowly but no single fading frequency will be predominant. The noise shaping filter is a second order IIR Butterworth characteristic with a 10Hz 3dB cut-off, and is based on a UHF carrier frequency in the region of 900MHz. The vehicle based channel scenario of figure 4.12 is characterised by a peaked band-pass characteristic which is "tuned" to the Doppler frequency, $f_d$, associated with the vehicle speed given by,

$$f_d = f_c \cdot \frac{v}{c}$$  \hspace{1cm}(4.19)\]

where $f_c$ is the carrier frequency, $v$ is the vehicle speed, and $c$ is the speed of electromagnetic propagation in free space. For the 900Mz carrier and a vehicle speed of 120km/hr, we obtain a 100Hz Doppler frequency.

It is a vehicle’s speed rather than its velocity relative to the BS which tends to determine the Doppler component of the signal fading for an urban environment. This is due to the line-of-sight path usually being blocked and therefore most of the signal is reflected from buildings and the road surface both in front of and behind the vehicle. The Doppler shift observed tends to be similar, in both a positive and negative sense, in the absence of any singularly large reflectors.
The noise shaping filter characteristics used for the low-pass and the Doppler scenarios are shown in figure 4.13. The low-pass filter uses a damping factor of $\zeta = \frac{1}{\sqrt{2}}$ to produce the Butterworth characteristic. The Doppler filter uses a very low damping factor of $\zeta = 0.02$ which gives the sharply peaked response due to the poles being located close to the unit circle, this characteristic also allows some slow fading to occur. The output power of the filtered AWGN has been calculated using the residues of the filter poles and is normalised to one prior to weighting.

![Figure 4.13 - Noise Shaping Filter Characteristics](image)

The urban channel noise indicated in figures 4.11 & 4.12 has been modelled as AWGN for simplicity. It is appreciated that this is not an ideal assumption and that urban noise can be impulsive and varies with the time of day and from city to city [87], thus it is difficult to quantify and model. It can, however, be shown that variable noise mixture signals can be sampled such that their effect is bounded to less than 2dB of the performance for Gaussian noise [88].

4.4.1 The Simulator Operation

The fading power envelope for the simulated 10Hz low-pass fading on a single tap is shown in figure 4.14. Note that the low-pass envelope is not characterised by any particular spectral component, but has a seemingly random wander. There are clearly
points in time where the signal is subject to severe fades. The power loss can be in excess of 40dB, with respect to the mean power, which is clearly going to result in error bursts if no counter measures are used. These severe losses tend to be short lived, but fades of say -10dB to -20dB can last for much longer periods. If the mobile receiver becomes temporarily stationary, it may of course become "stuck" in a fade. The simulator model, like most others, does not cater for such a situation, since we assume a constant but fairly random motion. This shortcoming is not a major problem, and should not greatly affect the system performance, once moderate degrees of diversity are applied. Figure 4.14a shows the signal power when no noise has been added to the channel. Figure 4.14b shows the same signal but AWGN is now present on the channel such that the mean SNR observed on that tap is 0dB. The overall shape of the fading profile appears to have been preserved but the severe fades are not so well defined.

In contrast to the 10Hz low-pass fading, the 100Hz Doppler fading, figure 4.15a, is highly characterised by the 100Hz fading frequency. The mean spacing between fades is $\lambda/2$, where $\lambda$ is the carrier frequency (we still assume a 900MHz carrier and a vehicle speed of 120km/hr). Slower and less dominant shadow fading can also be observed. The addition of noise, figure 4.15b, tends to make the Doppler frequency difficult to distinguish only the slower fading is still evident. Note that the mean and variance of both envelopes, figures 4.14a and 4.15a, will be the same, and the same is also true of figures 4.14b and 4.15b. Only the spectral components and the fade rate of these two sets of graphs differ.
a) no additive noise

b) with AWGN (mean SNR=0dB)

Figure 4.14 - Simulated 10Hz Low-Pass Fading Power Envelope
Figure 4.15 - Simulated 100Hz Doppler Fading Power Envelope

a) no additive noise

b) with AWGN (mean SNR=0dB)
The power profile envelopes with noise are typical of the type of signal power histories which will be presented to the receiver for each multipath component. The receiver, however, will normally obtain the fading multipath signal components as their I & Q voltage envelopes. These complex voltage signals provide not only signal magnitude information but also its phase. The phase is also severely effected by the channel noise.

Typical series' of fading multipath profiles, derived from the model, are shown in figure 4.16 and 4.17. Only every second profile is shown for clarity. Figure 4.16 illustrates that there is very little change in the profile over ten message bit periods (@4.8kbit/s) for the slow 10Hz low-pass fading, and so the profiles can be tracked using a fairly narrow bandwidth filter. Figure 4.17, on the other hand, illustrates how rapidly the profile changes for the 100Hz Doppler fading. There can be little correlation between the first and last profile in this series. These will be more difficult to track accurately when obscured by the channel noise. Typical profiles obtained from both low-pass and Doppler models are essentially no different in character. It is the rate and manner in which these profiles change that distinguishes the two models. If we compare the profiles obtained via the software model with those obtained using channel sounding (figure 4.3), we find a high degree of similarity.

Each time the software model is called a new profile is generated. Each tap of the profile can be generated in about 100μs on a SUN 4 IPX workstation. The rate at which simulations can be run is therefore dependent on the number of taps to be used. Standard 32 bit floating point arithmetic is used for all calculations. The channel simulator must be run for a number of iterations before it reaches a steady state. In general the simulator is run for 1000 dummy data bits prior to logging any simulation results.

The channel simulator, with its two preset urban scenarios, is used extensively in the following work on the adaptive RAKE receiver. It allows us to apply fairly realistic channel conditions to the receiver model, in order to analyse its performance and limitations.
Figure 4.16 - 10Hz Low-Pass Multipath Profiles

Figure 4.17 - 100Hz Doppler Multipath Profiles
4.5 Summary

We have shown that the RAKE receiver is optimal in Gaussian noise when correctly matched to the channel impulse response, due to the maximal-ratio linear combining techniques which are used. The theoretical improvement in SNR achievable, using these techniques, has been highlighted and the diversity mechanism which achieves the inherent multipath diversity of the signal, has been described. A review of previous work in UHF channel characterisation shows an agreement between these studies such that a realistic channel model can be constructed with a degree of confidence. The parameters used in constructing a software implementation of an urban channel model have been illustrated and the multipath and fading characteristics of the simulations have been shown to be similar to those measured on urban mobile channels. It was evident that the low-pass profiles can be considered locally stationary over about 10 bits whereas the Doppler channel cannot, at the same 4.8kbit/s rate.
Chapter 5

THE ADAPTIVE RAKE RECEIVER

In this chapter we consider a digital implementation of an adaptive RAKE receiver which uses the inherent diversity of the frequency selective wideband channel. The performance of such a system is first predicted using an semi-analytical model and then analysed via software simulation for a non-stationary mobile channel. Some of the performance deficiencies of this implementation are explored and compared with alternative sub-optimal RAKE architectures. We also consider methods for improving the performance when the channel is subject to fast fading.

5.1 RAKE Receiver Simulations

The first element of the RAKE simulations was the implementation of the channel simulator model as detailed in section 4.4. The channel model was used first to analyse the tracking filter performance in Rayleigh fading with noise. The simulations were next extended to the estimation of the multipath profile. The full system with pseudo-random data transmission and a DPSK modulation scheme was then constructed to determine the performance of the adaptive RAKE receiver. Perfect receiver synchronisation has been assumed throughout these simulations. The performance of the RAKE receiver is assessed by deriving BER curves and cumulative distribution functions (CDFs) for the burst errors.

5.1.1 The Receiver Operation

The channel model described was implemented such that the noise shaping filters are preset before the simulations are started. The number of taps, the data rate and the 3dB
cut-off frequency of the fading are all pre-determined by the operator. Each time the model is called a new multipath profile is generated. DPSK data is applied and noise is added in accordance with the mean SNR required.

The RAKE receiver circuits form an estimate of the channel's impulse response and this is then correlated with the signal being received. The complex conjugate of the channel estimate is used so that the signal components add coherently when combined. The output from the RAKE will have a peak SNR when the received waveform and the channel estimate are in alignment. This is easy to achieve since we are synchronised to the system chip rate, and the RAKE output need only be calculated every $M$ chips. This is useful since it limits the simulation complexity considerably.

![Figure 5.1 - Doppler Power Profile for a Moving Vehicle](image)

If a perfect channel estimate can be made then the RAKE receiver will be optimal. The quality of the estimate is dependent on how quickly the impulse response changes. This is heavily reliant on the movements of the MS and the wavelength of the carrier. Figure 5.1 shows the changes in received signal power against the distance travelled in wavelengths. The faster the mobile travels through this pattern of fades, the more difficult
it will be to obtain an accurate channel estimate since the profile will have changed before we are able to use it to "match" the RAKE receiver. The estimate will also be affected by the level of the channel noise.

5.1.2 Estimation of the Channel Impulse Response

The received set of fading post-correlation profiles when viewed as a voltage signal envelope will have sections which are reflected about the time axis when a data bit "one" is transmitted and non-reflected sections for a "zero", as shown in figure 5.2. Averaging the profile over several data bits would result in a zero mean and would be of no benefit to us. In order to track the fading observed at each delay we must correct the "phase" of each fading signal component so as to track the fading envelopes of the impulse response on each tap. The required correction term is given by the data content of the signal.

![Fading Multipath Signal Component and Envelope Component](image)

**Figure 5.2 - Multipath Fading Envelope and Signal With Data**

To remove the data from the fading signal we must employ decision feedback. One alternative to this is to sound the channel in the absence of data. Direct channel sounding, whereby an unmodulated PN signal is transmitted prior to a block of data bits, can be used for slow fading signals [89] but the extra bandwidth overhead becomes too severe for it to be sensibly used for tracking the fast Doppler fading. The direct approach will normally use each sounding signal as a complete channel estimation without recourse
to averaging. This makes the quality of estimation highly dependent on the SNR and type of noise observed on the channel. It is claimed that such a system is more robust than the indirect approach we are proposing [90]. This claim, it is assumed, is based on there being no requirement for decision feedback with the possibility of error propagation.

In this chapter we will look at the robustness of our chosen method by looking at the BER performance curves as well as burst error characteristics. The direct channel sounding system should perform less well in high noise conditions due to the poor channel estimate which will result, whereas the indirect decision directed method allows a degree of averaging since the "sounding" is more frequent. The improvement afforded by this averaging is of course dependent on the number of decision errors. Bit errors cause the channel estimation to diverge over the following bit period. The robustness of the two methods is a function of different mechanisms but in both cases it is limited by the noise on the channel over the estimation interval. The indirect method proposed is designed for a fast fading mobile channel which is not practical using the direct system since it dramatically increases the system bandwidth or reduces the available data rate.

The proposed German digital cellular radio system CD900 [91] employed a form of direct sounding, but as part of a time division multiplexed (TDM) scheme. In this system the preamble to each subscriber's timeslot is used as a channel sounding sequence as well as for frame synchronisation [92]. The preamble is 127 chips long and repeated every 31.5ms frame with a chip/bit rate of 4Mbit/s. The information bits in each timeslot are severely affected by ISI and so the impulse responses obtained via correlation are used to equalise the received signal. The quality of channel estimation however would be restricted by the relatively small processing gain (~21dB), and would not work well for fast moving mobiles due to the frame rate restrictions. GSM also incorporates a training sequence (26 bits) for use in adaptive equalisers which reduce the ISI [93], but more complex maximum-likelihood decoding is mainly used due to the short training sequence and an adequate received SNR.
5.1.3 Tracking of the Rayleigh Fading

To achieve maximal-ratio combining of the signals we require perfect knowledge of the channel's impulse response. Of course such knowledge is unavailable and we must therefore form an estimate which must be as accurate as possible so that the RAKE's performance is close to optimal. This requires us to estimate or predict the instantaneous magnitude for both real and imaginary components of the signal at each delay. The design of a filter to track each multipath component while rejecting most of the channel noise is now considered.

There are two basic requirements for the multipath signal tracker. It must allow the removal of the signal data content so that only the fading signal (plus noise) is left, and it must track the fading signal while rejecting the channel noise.

The multipath signals, which are to be tracked, are available as quantised levels and will be received serially as continuous multipath profiles. The tracking system must isolate the individual multipath components and track them independently after the data content has been removed. The model which has been adopted provides up to 32 independent multipath taps. If a single tracker is to be employed for all taps then it must be multiplexed between all of the taps. This is a sensible approach for the digital system since it minimises the hardware complexity. The multiplexed tracking filter operates on previous multipath profile measurements and the current one to try and construct a good noise-free estimate of the instantaneous channel impulse response.

A fast fading channel provides only a short period over which the tracker can "learn" the channel response before it changes significantly. The filter's memory need not necessarily be of high order but it must allow a reasonable amount of averaging to provide a usable SNR. This can be formalised by applying the constraint that the channel must have a frequency spread (Doppler fading rate) of much less than the data rate $R$ before the tracker can be expected to make reasonably accurate estimates. This is termed the "learning constraint" [80] given as
\[ R(\text{bit/s}) > f_D(\text{Hz}) \] (5.1)

where \( f_D \) is the frequency of the maximum Doppler spread on the channel.

At the data rate of \( R = 4.8 \text{kbit/s} \) the 10Hz low-pass model described in Chapter 4 satisfies this learning constraint. However the 100Hz Doppler model does not so easily satisfy this, which indicates that some problems may arise here. There are only 24 data bits transmitted in the time it takes the 100Hz Doppler signal to fade from peak to trough. Considerably fewer data bits than this can therefore be used to construct the estimate. The phase lag due to the decision feedback requirement will also cause a larger tracking error for the Doppler channel as opposed to the low-pass channel.

The mean SNR of the channel is obviously very significant since it will dictate the degree of averaging which is desirable. Essentially the desired tracking filter design will be governed by the rate of the fading observed on the channel and the mean SNR of this fading signal. The phase response of the tracking filter is critical to its performance and would ideally be zero over all fading frequencies. Such a filter is, of course, not realisable.

5.1.4 The Alpha Tracker

The simplest tracking filter to implement is the alpha tracker, figure 5.3, which is a single pole recursive filter given by equation 5.2.
\[ \hat{d}_k(n) = (1 - \alpha) \cdot \hat{d}_k(n) + \alpha \cdot d_k(n - 1) \quad 0 \leq k \leq (N - 1) \quad (5.2) \]

where \( a_k(n) \) is the received magnitude for the \( k^{th} \) multipath tap of the \( n^{th} \) profile, \( \hat{a}_k(n) \) is the tracked estimate, \( \alpha \) is the tracker pole coefficient, and \( N \) is the number of taps to be tracked. Note that the unit delay in figure 5.3, \( z^{-1} \), is for a bit (or symbol) delay rather than a chip delay.

This is a very efficient arrangement since the tracker only needs knowledge of the current profile estimate and the new received profile. It is therefore easily implemented as a multiplexed filter. The pole, \( \alpha \), can be used directly to control the filter's bandwidth which determines the amount of averaging to be applied to the profiles. The normalised frequency response is, however, fixed. With \( \alpha = 0 \) the bandwidth is infinite as the signal can pass through the filter unmodified. With \( \alpha = 1 \) we have zero bandwidth since the signal is blocked at the input to the filter. With intermediate values of \( \alpha \) we are thus able to set the bandwidth for the desired amount of exponential averaging.

This design must be modified slightly before it can be fully implemented since the estimation process is decision-directed. The modified alpha tracker is shown in figure 5.4. Its input is now provided by the post-correlation signal \( x_k(n) \) which subsequently has the data removed using the data decision from the previous message symbol, \( m(n-1) \). A symbol delay has been inserted so that the received correlation profile corresponds with the correct message bit, equation 5.3.
\[ \alpha_k(n) = m(n - 1) \cdot x_k(n - 1) \quad \text{(5.3)} \]

and so equation 5.2 becomes

\[ \alpha_k(n) = (1 - \alpha) \cdot m(n - 1) \cdot x_k(n - 1) + \alpha \cdot \alpha_k(n - 1) \quad \text{(5.4)} \]

The magnitude and phase responses for the simple alpha tracker are shown in figure 5.5 curves (a) and (b) respectively. Curve (c) shows the additional phase lag which is caused by a one symbol delay. Only the phase response is altered by the use of the modified tracker of figure 5.4. The effect of \( \alpha \) on the magnitude response is shown in figure 5.6. The phase response for the modified alpha tracker is shown in figure 5.7 and is obtained from the addition of the unmodified filter phase added to the decision feedback lag. Adding curves (b) and (c) of figure 5.5 will produce curve (e) of figure 5.7. Note that the wider bandwidth of the filter used for the 100Hz Doppler signal enables it to track the faster signal. The penalty is that we permit more noise to enter the system with the subsequent errors in the estimate. The value of \( \alpha \) is therefore something of a compromise since narrowing the bandwidth would cause poorer signal tracking but would reduce the noise further.

\[ \text{Figure 5.5 - Magnitude and Phase Response for Alpha Tracker} \]
Figure 5.6 - Alpha Tracker Magnitude Response

Figure 5.7 - Modified Alpha Tracker Phase Response
The filter tracking error at any given frequency can be calculated. The absolute tracking error between any two sinusoids of identical period and amplitude one, is given through the trigonometrical identity as

\[ 2 \sin\left(\omega t - \frac{\Phi}{2}\right) \cdot \sin \frac{\Phi}{2} \]  

(5.5)

where \( \omega \) is the angular frequency of the sinusoids and \( \Phi \) is the phase angle between them. The root mean square (rms) tracking error due to the phase difference is therefore

\[ \sqrt{2} \sin \frac{\Phi}{2} \]  

(5.6)

The absolute rms tracking error, \( \Delta \), between a sinusoid (with a peak value of one) and its linearly tracked output is

\[ \Delta = \frac{1}{\sqrt{2}} (1 + |H(\omega)|) \sin \frac{\Phi}{2} \]

\[ = \frac{1}{\sqrt{2}} (1 + |H(\omega)|) \sin \left( \frac{\angle H(\omega) - \Phi_r}{2} \right) \]  

(5.7)

since \( \Phi \) consists of the filter's phase \( \angle H(\omega) \) and the symbol delay due to decision feedback \( \Phi_r \). For the modified alpha tracker the magnitude and phase responses are given by

\[ |H(\omega)| = \frac{1 - \alpha}{\sqrt{1 + \alpha^2 - 2\alpha \cos \Phi_r}} \]  

(5.8)

\[ \angle H(\omega) = \tan^{-1}\left( \frac{-\alpha \sin \omega}{1 - \alpha \cos \omega} \right) \]  

(5.9)

and

\[ \Phi_r = \omega T \]  

(5.10)
where $T$ is one message bit period and hence $\Phi_r$ is the angle through which the signal will travel in a single bit period.

Using the above equations we can deduce that when tracking a 100Hz sinusoid at a bit rate of 4.8kbit/s with $\alpha = 0.5$, we introduce an rms error of 0.182 (0.257 peak). On the other hand a 10Hz sinusoid at the same bit rate and $\alpha = 0.8$ introduces an error of only 0.046 (0.065 peak). These results are for pure tones only and do not take account of the effects of noise on the tracking performance. Figures 5.8a & 5.8b illustrates the instantaneous tracking errors for the above two cases. Note that the tracking error on the 100Hz signal is largely due to the phase response of the tracker which is primarily a result of the decision feedback delay.

It is clear that the alpha tracker is perhaps not ideal, especially for fast Doppler fading where it is debatable whether the learning constraint is entirely satisfied. The tracker however is simple and performs adequately for slow-fading. At this point it is conjectured that individual path estimations need not be very accurate in order for the RAKE to show an improvement in performance, assuming sufficient diversity is available. Of course the performance would be better if a more accurate estimate were available. What is unclear is what the loss in performance will be in terms of the BER.

The alpha tracker’s performance has been observed using its magnitude and phase responses for pure tone inputs. We now wish to determine the trackers ability to follow realistic Rayleigh fading signals with additive noise.

The modified alpha tracker was fed with the two types of signal scenario - 10Hz low-pass and 100Hz Doppler fading at varying noise levels. In the following figures the tracking error is calculated by subtracting the tracker output from the ideal noise-free fading signal. Figure 5.9a shows the low-pass fading signal from a single tap of the simulator with a mean SNR of 7dB. This fading signal is fed to the alpha tracker. Figure 5.9b shows the tracked signal along with the ideal noise-free fading signal which it is trying to track ($\alpha=0.8$), while figure 5.9c shows the magnitude of the absolute tracking error. The message period is given by the reciprocal of the bit rate and is approximately 0.2ms.
Figure 5.8 - Alpha Tracker Performance With Pure Tones

It is clear from figure 5.9b that the tracking filter is performing well although some noise is still present giving the slightly spiky envelope. Figure 5.9c effectively shows the noise on the estimate since the phase lag is negligible for the 10Hz low-pass signal. Note that
these graphs are the fading signal voltage envelopes which exist for both real and imaginary components of the received signal. Only the tracking of a single real component is illustrated.

Figures 5.10 shows the same signals but for a Doppler fading signal again for 7dB mean SNR but with $\alpha=0.5$. Figure 5.10a shows the signal with the dominant Doppler frequency. The number of zero crossings of this signal is about ten times that of the 10Hz low-pass signal and they are more evenly spaced. Figure 5.10b appears to indicate good tracking. However, comparing the tracking errors of figure 5.10c with that of the low-pass tracking, figure 5.9c, reveals that the error is significantly increased. This is due to two factors, firstly the bandwidth of the alpha tracker has been increased to cater for the higher frequency 100Hz Doppler signal and thus more noise is included in the estimate, secondly the phase lag at this frequency leads to a greater error. A small 100Hz component is evident on the tracking error signal of figure 5.10c which is due to this phase lag.

The plots shown in figure 5.11 are equivalent to those of figure 5.9, only the mean SNR is 0dB instead of 7dB. The noise is clearly visible in figure 5.11a. Figure 5.11b shows that the tracking follows the general envelope of the signal but the detail is lost in the noise. The tracking error signal, figure 5.11c, although noticeably worse than that for the low-pass 7dB mean SNR, figure 5.9c, is still better than that for the Doppler signal at 7dB mean SNR, figure 5.10c.

The Doppler signal with 0dB mean SNR is shown in figure 5.12a, note that the number of zero crossings has increased as a result of the high noise level. The tracked signal, figure 5.12b, has, however, reduced the number of these extra zero crossings. The tracking error, figure 5.12c, indicates a large mismatch with the original signal, again the 100Hz signal can be seen on the tracking error signal which highlights the phase error component.
Figure 5.9 - Tracking of 10Hz Low-Pass Fading (mean SNR=7dB)
Figure 5.10 - Tracking of 100Hz Doppler Fading (mean SNR=7dB)
Figure 5.11 - Tracking of 10Hz Low-Pass Fading (mean SNR=0dB)
Figure 5.12 - Tracking of 100Hz Doppler Fading (mean SNR=0dB)
These graphs clearly illustrate the superior tracking which is obtained for low-pass fading. Very large tracking errors occur for the Doppler case when the mean SNR is low. This unfortunately will often be the case for most of the less significant taps. It is not possible to improve the tracking performance significantly by adjusting $\alpha$ since more averaging will reject more noise but will also increase the phase lag on the filter as indicated previously by figure 5.7. The tracking filters must track the complex fading signal. When implemented in the adaptive RAKE receiver independent tracking filters are used for the in-phase (I-channel) and quadrature (Q-channel) signal components.

If we consider the effect of this tracking error for an entire multipath profile we will observe both over and under estimates on the independent taps. These errors when viewed over the entire profile estimate appear to have simply reduced the effective SNR of the estimate but have not destroyed the estimate’s inherent diversity. It is therefore possible to consider this tracking error as a noise component. This viewpoint has not considered the fact that the noise signal assumed for the tracking error has a large correlation to the signal itself for any given independent tap. The justification for this is that we are looking at the correlation of the estimates error to the ideal impulse response of the channel as a whole and not as individual taps. These are therefore largely uncorrelated. As a result of the central limit theorem and the independent nature of the taps which have phases that are randomly distributed about 0 to $2\pi$ radians, we can use a noise variance to represent the tracking error.

The extra noise component is dependent on both the actual mean SNR on the channel and the rate of fading (which determines the estimate phase lag). The implication of this is that large tracking errors may be tolerable if there is sufficient diversity and moderate mean SNR on the channel. Section 5.2.1 attempts to predict the performance of the adaptive RAKE using a model which introduces noise components to represent the tracking error due to the phase lag and the noise in the estimate.
5.1.5 Channel Profile Tracking

The individual alpha trackers estimate the signal at each discrete delay of the profile. So we are not only tracking the fading of the individual multipath components, we are also tracking the entire multipath profile. This produces an estimate of the impulse response of the channel. The complex conjugate of this estimate is subsequently used as the coefficients for the adaptive RAKE filter to form the channel matched filter. A typical sequential set of multipath profiles produced by the channel model are illustrated by figure 5.13.

![Figure 5.13 - Typical Simulated Multipath Profiles](image)

The estimated multipath profiles will be subject to error due to the magnitude response of the tracker - which is small, phase lag - which is large for fast fading, and noise tracking - which decreases with increasing SNR. Note the problem of tracking noise becomes more pronounced for the less significant taps. The average excess delay profile of the channel simulator decays at a rate of around $\frac{500}{11}$, thus if the main tap has a mean SNR of 10dB, the tap at a further delay of 2µs will have a mean SNR of only 0dB.

Maximal-ratio combining is achieved when weighting is performed with respect to the signal power. We get an unrealistically high estimate of the signal power for taps with very low SNR since they will be dominated by the noise. Thus, the signal combining
tends to become less ideal as more taps are added since the mean SNR of these taps rapidly decreases. The effect of the sub-optimal combining can be seen by observing the post-correlation profile and the tracked profile estimate for decreasing SNR values. Figures 5.14 a, b, & c illustrate the profile tracking for 10Hz low-pass fading ($\alpha=0.8$) with decreasing SNR, taken as a snapshot in time. The received multipath signal and the tracked signal are shown as well as the ideal (noiseless) multipath profile which we will ideally match. Note that for high mean SNR (100dB) the tracking is almost perfect whereas for higher channel noise the less significant taps have very poor estimates. The 0dB mean SNR graph shows that the noise floor of the post-correlation signal dominates the estimate. Figures 5.15 a, b, & c show the same information for a 100Hz Doppler fading profile. Note that similar trends can be observed but the tracking is poorer due to the phase lag on the estimates and higher equivalent noise bandwidth of the alpha tracker used ($\alpha=0.5$).

In general the estimate has more power than the multipath power due to noise tracking, but has a lower power than the received multipath signal as a result of the filter averaging. At low SNRs some of the mismatching is caused by decision feedback errors.
Figure 5.14 - 10Hz Low-Pass Multipath Profile Tracking
Figure 5.15 - 100Hz Doppler Multipath Profile Tracking
It is evident that the number of taps that the RAKE should ideally employ is dependent on the mean SNR observed on each tap and the rate of fading of the channel. Clearly there comes a point, at poor SNR values, where the tracking errors are so large that adding further taps into the RAKE receiver cannot be justified.

5.1.6 Intersymbol and Interpath Interference

We have already noted that intersymbol interference is not a problem if the delay spread of the channel is less than the period of the message signal. This is due to the post-correlation signal spread being restricted by the excess delay of the channel, which is normally shorter than the message symbol period used. This has assumed that the autocorrelation profile of the channel has a single peak and that the sidelobes are of zero magnitude. For a practical system this is not the case.

Any non-zero portion of the autocorrelation sidelobes, which is spread into the next period, will cause ISI. However, if the sidelobes are very small relative to the peak the problem will be minimal, but nonetheless present. The sidelobes which are spread into the next period will usually have a mean power considerably lower than the ones present in the correct symbol period.

Another problem which exists is the interference which the sidelobes cause to the multipath profile estimate. The sidelobes, at any given delay, do not belong with the multipath correlation peak observed at that same delay, they belong to preceding correlation peaks and will thus fade independently of the peak present at the same delay. This again will cause an error in the channel estimation. This phenomenon is called interpath interference.

To minimise both of these problems we chose codes which have low sidelobes both preceding and following the autocorrelation peak. This low sidelobe region should be equivalent to the delay spread expected on the channel. Such codes are reported in section
3.5, their near optimal characteristics for both odd and even correlation functions essentially eliminate this problem. The simulations have therefore assumed an ideal ACF.

The effects highlighted will be more significant if shorter or randomly chosen codes are used. The bias error which results from this [94] is deterministic. Since the ACF is known, an estimate of the multipath profile is available, and the data content of the signal can be determined, it is therefore theoretically possible to partially eliminate these effects by cancellation. However, the complexity required for this, and the improvement which would be gained, is unlikely to make this a viable proposition unless very short codes were employed.

5.2 The Performance of the Adaptive RAKE Receiver

The performance of the adaptive RAKE receiver design is predicted using a mathematical model which we define. This system is then simulated and the BER results are compared. The burst error characteristics for the simulated system are also discussed.

5.2.1 Predicted Performance

When considering the performance of the RAKE we will assume that the first multipath component used will be the one with the largest mean SNR, from the model defined in Chapter 4 this is given by tap 2 (figure 4.8). All subsequent taps, when required, will be used in order, i.e tap 2, 0, 1, 3, 4, 5,... Note that apart from the first two taps, each preceding tap will in general be the next largest and hence it is sensible to add them in this sequence. The instantaneous SNR, $\gamma_k$, on the $k^{th}$ diversity branch is given by

$$\gamma_k = \frac{E}{N_0} a_k^2$$

(5.11)

where $E$ is the energy in the largest diversity branch which is normalised to one, and $N_o$ is the single-sided power spectral density of the noise. Therefore the average or mean SNR for the $k^{th}$ branch is
\[ \bar{\gamma}_k = \frac{E}{N_0} E[a_k^2] \quad (5.12) \]

\[ E_k = \frac{E}{N_0} \]

where \( E_k = E \cdot \bar{a}_k^2 \)

\[ E[x] \] is the expected value of \( x \). The average probability of error for DPSK [81] obtained from \( L \) orders of diversity is given by

\[ P_e = \frac{1}{2} \sum_{k=1}^{L} \frac{\Pi_k}{1 + \bar{\gamma}_k} \quad \text{where,} \quad \Pi_k = \prod_{i=1}^{k} \frac{\bar{\gamma}_k}{\bar{\gamma}_k - \bar{\gamma}_i} \quad (5.13) \]

This equation describes the upper bound on performance achievable by using MRC. It represents the RAKE receiver performance when a perfect estimate of the channel is obtained. The theoretical bit error rate curves have been plotted for \( L=1 \) to \( L=5 \), figure 5.16. The mean SNR values, \( \bar{\gamma}_k \), are calculated using the tap gains, given by the channel model defined in Chapter 4. The mean SNR scale is with reference to the most significant tap and so the graph indicates the theoretical improvement in performance obtained when adding additional taps to a RAKE receiver which is perfectly matched to the channel's impulse response.

We now consider the non-ideal tracking whereby we have added an extra noise term to represent the average tracking error on the \( k^{th} \) diversity branch.

Recall from Chapter 4 that the multipath channel impulse response can be represented by a set of amplitude and phase components \( a_k \) and \( \theta_k \) each separated by a given delay. In the RAKE implementation we assume that the separation of these components is the reciprocal of the chip rate of the system \( 1/T_c \). The impulse response is assumed finite in duration and is thus restricted to a number of delay components or taps. For convenience we will use the complex multipath components given by
and we represent this set of components by the vector $\mathbf{C}$. The received channel vector, $\mathbf{R}$, will also have an additional AWGN vector $\mathbf{N}$, so

$$\mathbf{R} = \mathbf{C} + \mathbf{N}$$  \hspace{1cm} (5.15)

The complex conjugate estimate of this noisy channel impulse response is given by

$$\hat{\mathbf{C}}^* = \mathbf{C}^* + \mathbf{W}^*$$  \hspace{1cm} (5.16)

where $\mathbf{W}^*$ is the vector of the estimation noise and "*" denotes a complex conjugate.

Ideally the output from the RAKE will be $\mathbf{RC}^*$, but it is in fact given by

$$\mathbf{R}\hat{\mathbf{C}}^* = \mathbf{RC}^* + \mathbf{RW}^*$$  \hspace{1cm} (5.17)

The mean SNR of the $k$th tap of an $N$-tap adaptive RAKE, $\frac{E_t}{N_0 + E[|r_k w_k^*|^2]}$, will therefore be reduced by the noise term $\mathbf{RW}^*$, such that

$$\tilde{\gamma}_k = \frac{E_t}{N_0 + E[|r_k w_k^*|^2]}$$  \hspace{1cm} (5.18)
where \( r \) and \( w \) are the \( x^{th} \) components of the vectors \( R \) and \( W \). These components result in the addition of two extra noise variances, assuming that they have Gaussian statistics. Firstly we have a noise variance associated with the proportion of the post-correlation noise power which can pass through the tracking filter, \( N_f \). Secondly we have the self-noise generated by the tracking filters, \( N_e \), which is almost totally attributed to their phase response and is dependent on the fading rate of the signal. Thus we have

\[
\tilde{\gamma}_k = \frac{E_k}{N_o + N_f + N_e}
\]  

(5.19)

The value of \( N_f \) is proportional to the equivalent normalised noise bandwidth for the tracking filter, \( B_N \). So equation 5.19 becomes

\[
\tilde{\gamma}_k = \frac{E_k}{N_o + N_o K B_N + N_e}
\]  

(5.20)

where \( K \) is a constant and \( B_N \) is normalised to the equivalent noise bandwidth of the untracked input. The value of \( B_N \) was determined using Simpson's Rule for the calculation of the area under the tracking filter's magnitude response. Equation 5.20 can be rearranged to give

\[
\tilde{\gamma}_k = \frac{\tilde{\gamma}_k}{1 + K B_N + \tilde{\gamma}_k \bar{E}_k}
\]  

(5.21)

where \( \bar{E}_k \) is the normalised tracking error for the \( k^{th} \) branch

\[
\bar{E}_k = \frac{\varepsilon_k}{\alpha_k}
\]  

and

\[
\varepsilon = \frac{N_k}{E}
\]  

(5.22)
and $e$ is the normalised tracking error for the main diversity branch (whose power is assumed to be unity). This can be calculated if the fading statistics of the channel and the phase characteristics of the tracking filter are known. This parameter is often difficult to obtain analytically especially for time varying channel statistics.

Alternatively this parameter can be obtained from the irreducible bit error rate (IBER) of a non-diversity (or single tap RAKE) system employing the same tracking filter. This enables us to analyse the effect of adding diversity to the system under the condition of non-ideal signal tracking. For a DPSK system the probability of error for ideal signal tracking of Rayleigh fading is given by [96]

$$P_e = \frac{1}{2(1 + \Gamma)} \quad (5.23)$$

where $\Gamma$ is the mean SNR on a non-diversity channel. Thus,

$$P_{IBER} = \frac{1}{2(1 + \frac{1}{\epsilon})} \quad (5.24)$$

since the effective channel SNR is restricted by the reciprocal of the additional normalised noise power, $\epsilon$, of a non-diversity receiver. So

$$\epsilon = \frac{2P_{IBER}}{1 - 2P_{IBER}} \quad (5.25)$$

We can observe from equation 5.20 that there are two noise terms which are dependent on the noise on the channel, these terms will affect the slope of the BER curve. The third noise term is independent of the channel noise and will result in an upper bound on the effective mean SNR for each diversity branch, $\bar{\gamma}_k$. This leads to the an IBER when used in equation 5.13 in place of $\bar{\gamma}_k$. The IBER [97] [98] is noticeably reduced in value when we apply diversity to the system. The model has not considered the possibility of error propagation and so as it stands, it is not able to predict the robustness of the adaptive RAKE receiver.
An example of the predicted BER curves is illustrated by figure 5.17 for a value of $\varepsilon = 0.001$ which is equivalent to a mean irreducible channel SNR of about 30dB and gives an IBER of 0.0005 for DPSK signalling and no diversity. $K$ was set to 2.1 (this was found by simulation) and $B_N$ was calculated from equation 5.8 to be 0.24 for $\alpha=0.6$. Note that using just one tap would not be sensible since a better performance could be obtained with a non-tracking receiver.

![Predicted Performance for Adaptive RAKE](image)

**Figure 5.17 - Predicted Performance for Adaptive RAKE**

The equations developed are used in the next subsection to directly compare the simulated results with those predicted.

It is also possible to add further system loss terms to equation 5.19 by equating them as additional noise variance terms. CDMA co-channel interference, for example, can be catered for by calculating the variance of the multiple access noise [99] and adding it as if it were white Gaussian noise.

### 5.2.2 Simulated Results

The results for the adaptive RAKE receiver simulations are now considered. The channel model detailed in Chapter 4 was used to obtain these results. Random data was transmitted
over this channel using a DPSK modulation scheme. The data decision at the output of
the adaptive RAKE is compared with the known transmitted bits in order to determine
the probability of error.

The complex signal produced by passing the random data bits as single chip impulses
through the time varying multipath channel produces the equivalent of the
post-correlation signal at the output of the channel simulator. This has assumed that there
is negligible intersymbol and interpath interference caused by the sidelobes, i.e. a PN
code of length 1024 with an autocorrelation peak of ±1024 and sidelobes of zero over
31 chips either side of this peak is used. While such a code does not exist in practice,
we have already illustrated in Chapter 3 that codes with properties approaching this ideal
ACF can be generated.

The RAKE receiver implementation, figure 5.18, used in the simulations is limited to 32
taps with the same tap spacing as the channel simulator. The received signal (which
consists of the resolved multipath components \( c_k(n) \) and the post-correlation sampled
noise \( \eta_k(n) \)) is coherently multiplied with the complex conjugate of the estimated channel
impulse response, \( c_k^*(n) \), which has been obtained from the tracking filters.
If we consider a single tap, the complex received signal samples, $A + jB$, are multiplied by the estimated complex conjugate weight samples, $\hat{A} - j\hat{B}$, to produce the real number $A\hat{A} + B\hat{B}$. This represents the amplitude of the coherently detected sample value at that tap. When these values are summed for all active taps, the sign of the result will indicate whether the received data bit is a 0 or a 1. Thus with complex conjugate RAKE combining this receiver effectively operates as a coherent PSK detector. Subsequently two sequential data bits are compared in order to determine the message content of the DPSK signal.

Identical filters are used on each tap to form the channel estimate. In fact, it is the same filter which is multiplexed between all of the taps. The mean SNR values quoted in the results are assumed to be the average SNR values at the output of the correlator. Any losses due to the receiver front end and synchronisation processes are assumed to be included in this figure. As with the analytically predicted results, the mean SNR on the graphs is quoted for the tap with the largest mean SNR. It is not the combined SNR for all the taps which is used.

The number of errors logged for each point on the BER curves was dependent on the probability of an error at that point. For high probability of error the errors tended to occur in bursts and thus the BER calculations are based on a large number of logged errors. As an example, there were 30,000 errors counted for each point on the single tap curve of the 100Hz Doppler curve. The minimum error count for any point on the BER curves is 200. Low error counts obviously had to be used for very small probability of error points on the curves, especially when using a large number of taps on the RAKE, due to the simulation times involved. All points on the graphs have been obtained by randomly seeding the noise and data generators with a time variable. This ensures that the curves will not become prematurely smooth, and appear correct, prior to that point on the graph converging to a "stationary" value.

It was not feasible to extend some of the BER curves beyond the points given on the graphs due to the simulation times involved. In some cases in excess of one billion
complex multipath components were simulated (which takes about one week of computational time on a SUN 4 workstation) to produce one point on the curve. The next point in many cases would require an increase in processing complexity of about two orders of magnitude. It is highly unlikely that we would wish to operate in these low error rate regions and so this information would not be of great significance anyway. It was also suspected that the limited precision of the calculations would start to show a background error count in these low error rate regions. Single precision rather than double precision floating point arithmetic was used for the simulations in order to try and further limit the processing time.

The first simulated results are given for the 10Hz low-pass (hand-held) channel, figure 5.19. The adaptive RAKE was implemented for 1, 3, 6, 12 and 32 taps. The single tap curve would appear to be heading towards $P_{BER} \approx 2.0 \times 10^{-4}$ as a consequence of the tracking errors. It is also clear that adding an extra two taps gives a considerable increase in performance but as further taps are added the improvement is less marked. This can be explained by two factors. First of all there is less signal power contained in all subsequent taps and so the signal power increase is not so great. Secondly the lower power taps will contribute proportionately more noise into the system. This can be observed on the profile tracking results, figure 5.14. Indeed there is no net gain for up to 5dB mean SNR by increasing the number of taps from 12 to 32, this indicates that the losses due to noise tracking on the less significant taps nullifies the diversity gain obtained.

We noted earlier that the phase lag for the modified alpha tracker with 10Hz low-pass fading was small in comparison with that of the 100Hz Doppler. The noise which this introduces to equation 5.21 will be minimal and indicates that the IBER will be fairly low. We are also able to narrow the bandwidth of the tracker and can therefore reduce $B_N$. We have used a value of $\alpha = 0.8$ which gives an equivalent noise bandwidth of $B_N=0.12$. This indicates that the noise in the channel estimate should be small. This is confirmed by comparing the BER curves with the upper bound performance (or the lower
bound BER performance). This comparison is made in figure 5.20. It is clear that the simple alpha tracker provides a performance for the low-pass fading scenario which is close to the ideal.

Figure 5.19 - Simulated RAKE BER Performance for 10Hz Low-Pass Fading

Figure 5.20 - Upper Bound and Simulated RAKE Performance
Similar simulations have been carried out for the 100Hz Doppler (vehicle based) scenario. The results are shown in figure 5.21. The most notable difference between these curves and those for the low-pass fading is the emergence of the IBER limitation. This is clearly visible for both one and three taps, but has been considerably reduced on the latter curve. Again the performance of the 32-tap RAKE can be inferior to that with only 12-taps. In this instance the 32-tap RAKE is actually inferior to that of the 6-tap RAKE at 0dB mean SNR. It may be possible to improve this performance by discarding taps which fall below a noise threshold [100], however, this adds more complexity to the implementation.

The performance of the adaptive RAKE for the 100Hz Doppler scenario is clearly worse than that for the 10Hz low-pass case. The curves do not fit closely to those of the upper bound performance. It is evident that both the phase lag of the modified alpha tracker and its equivalent noise bandwidth have contributed to this performance degradation. The bandwidth of the filter had to be widened in order to track the faster fading signal.
Alpha was set to 0.5 giving $B_N = 0.33$. The theoretical upper bound is not a good indicator of the performance in this case. Instead we may use the equations developed in the previous section.

Using the IBER result for the single tap we obtain a value of $\varepsilon = 0.020$. Inserting the values $K=2.1$, as before, $B_N$ and $\varepsilon$ into equation 5.21 & 5.22, and using equation 5.13 enables us to compare the simulated results with those predicted. Figure 5.22 compares the simulated and predicted results with the upper bound for one and three taps. We can observe that, while our predicted curves correspond well with the simulated results, the performance is well short of the upper bound on performance which closely modelled the 10Hz low-pass results.

![Figure 5.22 - BER Performance Comparison for 100Hz Doppler Fading](image)

In order to illustrate more clearly the restriction on performance imposed by the IBER we have simulated the same adaptive RAKE for 1, 2, 3, 4, & 5 taps, figure 5.23.

The predicted IBER performance, using the parameters defined and equations 5.21 and 5.13, is also given in figure 5.24 for comparison. Clearly the prediction is very consistent if given the appropriate set of constants.
The results and observations presented in the preceding sections have used alpha tracker values of $\alpha=0.8$ for the 10Hz low-pass fading model and $\alpha=0.5$ for the 100Hz Doppler fading model. These were chosen to try and optimise the adaptive RAKE receiver.
performance for the two different operating scenarios. They were arrived at initially in an attempt to minimise the noise bandwidth while still allowing the majority of the fading signal components to be tracked. To prove that these are sensible choices the relative tracking error loss has been plotted against alpha for both scenarios. The optimum value is dependent on the channel SNR and to a lesser extent on the order of diversity. Figure 5.25 illustrates that the performance is relatively insensitive to the values chosen for high mean SNR. At higher noise levels the optimum value is more distinct, figure 5.26.
5.2.3 Burst Error Performance

One aspect of the system which has not yet been discussed is the effect of decision errors which will cause the tracker to diverge rather than converge. The burst error characteristics of the adaptive RAKE receiver are important in that they indicate the extent to which error propagation occurs and show the nature of the errors which are obtained under different operating conditions. We have limited this analysis to the 100Hz Doppler case only, since this is where the effect of these errors is most pronounced. Two burst error CDFs are plotted. The first examines the effects of additive noise on a single tap system, figure 5.27. The second has constant mean channel SNR and observes the effect of increasing the diversity of the system, figure 5.28. We have defined an error burst as the number of bits between two consecutive error free regions which are at least 8 bits in length.

![Figure 5.27 - Burst Error CDF for AWGN (1-tap)](image)

Note, in figure 5.27, that with 20dB mean SNR over 90% of all errors are single bit errors. This is a very significant result considering that we are using a DPSK data modulation format. A single bit error indicates that it has been caused by a fading signal zero crossing where the signal estimate has not yet changed polarity (i.e. crossed through
zero) due to phase lag. If we compare the equivalent 20dB point on the single tap curve of the BER results in figure 5.21 we see that the IBER has almost been reached. We can thus conclude that the IBER is caused exclusively by single errors induced by the lag in the fading signal estimate.

As the mean SNR is reduced the burst errors increase in length. At 0dB the errors are predominantly bursts of two or more errors, however 60% of these errors exist as bursts of less than 10 and there are almost no bursts of over 100. This indicates that even at moderately low SNR the system is fairly robust given the low BER which is expected at these SNR values. At -10dB the results are poor but they still give confidence that the system will not suddenly become divergent due to error propagation effects.

The performance would appear to indicate that error propagation is not a major problem since the degradation in performance occurs smoothly and in accordance with that expected as the SNR decreases. Severe signal shadowing should not therefore lead to the receiver becoming stuck in an irrecoverable state. It is thought that the DPSK data modulation employed helps with this situation since it tends to break up consecutive errors such that continual divergent estimates are highly improbable even for very high
bit error rates. These conclusions have been drawn from the single tap case (without diversity) but they clearly illustrate the non-divergent nature of the decision-directed tracking filters.

The CDF curves for the increasing number of taps indicates that the diversity itself would appear to help reduce the error bursts. The improvement in burst error performance is due both to the increase in effective SNR and by the multipath diversity reducing the fade durations. Most errors at the mean SNR of 5dB, figure 5.28, are caused by noise and manifest themselves as bursts of 2 or more bits. The RAKE diversity also helps to reduce the proportion of single bit errors since the effective number of zero crossings of the RAKE output are reduced. If there is a zero crossing on one diversity branch there is a high probability that its effect will be compensated for by the other diversity branches, and so it is less likely to cause an error.

The ability to reduce the error burst lengths is useful if block or convolutional coding is to be used for error correction. These techniques cannot operate effectively during error bursts unless bit interleaving is used over lengths much greater than the burst length. This leads to considerable receiver complexity and significant delays on the data. Figure 5.28 indicates that adding diversity can contribute to the effectiveness of the error correction by shortening the error bursts.

5.2.4 Sub-Optimal RAKE Combining

We have observed that the performance of the RAKE receiver when applied to the 100Hz Doppler fading scenario is limited by the relatively poor tracking of the modified alpha tracker. We also noted that it is the time lag on the estimation and not the magnitude error which are significant. MRC is difficult to obtain under these severe conditions and so we obtain sub-optimal combining which only approximates that of MRC. We now compare these results with those which can be obtained using other sub-optimal RAKE combining techniques.

• Equal-Gain Combining
Equal-gain combining enables us to eliminate all of the phase lag directly associated with the tracking filter. All signal components are combined with equal weighting, regardless of their effective mean SNR. We still cannot eliminate the decision feedback delay since the components must be added with the correct polarity otherwise they will tend to cancel rather than add coherently. This indicates that we require estimates of the previous signal in terms of the multipath component polarities. This is achieved by zero thresholding both the I and Q signals to obtain the weights ‘-1’ or ‘1’ and applying these to the RAKE receiver for the next received symbol.

We would expect a reduced IBER value from the EGC RAKE due to the smaller phase lag. We would also, however, anticipate a poorer BER slope for large numbers of taps because the less significant taps will not be suppressed due to the equal gain. Figure 5.29 shows the BER performance for EGC. Figure 5.29a shows a broadly similar trend to that of the MRC RAKE (figure 5.23), although the IBER values are noticeably lower. Note that the first five taps will contain similar magnitudes of signal power (when viewed in a wide sense), and so the performance degradation should not be that great for these low orders of diversity. Figure 5.29b clearly illustrates that by adding more taps with equal gain is not always beneficial. The 32 tap curve has a considerably poorer BER performance than that of only 12 taps, and the performance with 6 taps can match that of 32 at 7dB mean SNR.

- Differential Phase Combining

With DPC each signal component is weighted using only the previous message signal’s multipath profile. Once combined, the output provides the decoded DPSK signal waveform, assuming the absence of channel noise. We are in effect comparing the previous received symbol with the current one using correlation. Again this reduces the lag on the estimate, but there is no possibility of improving the estimate magnitude. We also reduced the complexity of the receiver by eliminating the tracking filter and the additional DPSK decoding circuit. This method produces relatively poor results when the channel is noisy [101].
Figure 5.29 - BER Curves for Equal-Gain Combining RAKE
Figure 5.30a illustrates that the DPC RAKE, again, produces a similar type of result to that of MRC, and a lower IBER has been obtained due to the reduced estimation lag. Figure 5.30b shows the similar increase in performance due to additional diversity up to about 6 taps. The 12 tap curve has only a marginal improvement beyond this, and the 32 tap results are worse than those for just 12 taps. This worsening at 32 taps is not as prominent as it was for the EGC curves.

• Comparison of Results

To compare the results of EGC and DPC with MRC, the BER curves have been plotted on the same graphs for 1 & 3 taps - figure 5.31a, and 6 & 32 taps - figure 5.31b. For a single tap we obtain 3 similar curves characterised by the IBER value. Significantly DPC has the best IBER value followed closely by EGC. MRC is the worst of the three as a result of the additional lag of the tracking filter.
Figure 5.30 - BER Curves for Differential Phase Combining RAKE

a) 1-5 taps

b) 1-32 taps
Figure 5.31 - BER Comparison Between MRC, EGC, & DPC
For the 3 taps, MRC performs best between 0-10dB mean SNR. Beyond this it begins to be constrained by the IBER and DPC produces the lowest error rates. EGC performs poorly in comparison but manages to cross the MRC curve at 18dB as a result of its lower IBER value. The 6 tap curves show how the EGC performance becomes steadily worse in relation to the other curves as the order of diversity increases. The MRC is marginally better than that of DPSK between 0-12dB. With 32 taps MRC is clearly superior. The EGC result is particularly poor even at fairly high SNRs. At 10dB it is still performing notably worse than that of 6 tap DPC and MRC.

![Graph showing BER comparisons for 3 Tap RAKE](image)

Figure 5.32 - BER Comparisons for 3 Tap RAKE

The final BER graph, figure 5.32, compares the curves obtained for the 100Hz Doppler results with those for the 10Hz low-pass and the upper bound. A 3-tap RAKE is used for this comparison. Obviously no curve is lower than that of the upper bound which represents the performance which would be obtained with perfect tracking using the MRC RAKE and DPSK signalling. The closest result to this curve is that for the 10Hz low-pass scenario using the MRC RAKE. The result is within 1dB of the upper bound curve. At low SNR the 100Hz Doppler MRC performance is about a further 1-2dB worse, but gradually deviates away from this towards the IBER asymptote. The DPC and EGC
curves show a poorer performance at lower SNR but are more favourable as the SNR increases. The EGC result is consistently worse than that obtained for DPC. These trends tend to become further accentuated as more taps are added.

It is clear that although the MRC RAKE is not operating as close to its upper bound as we may have wished for fast Doppler fading, it is still providing a considerable improvement in performance for a moderate amount of taps. The EGC RAKE requires a good SNR to perform as well as MRC. EGC is not considered very useful since it represents a loss in performance over the region of particular interest and provides little compensation in view of the minor reduction in complexity. The DPC performance is somewhat better but still has a minor loss in performance at low SNRs. DPC is particularly attractive in view of the reduction in complexity. It is also evident from the results that improving the performance of the tracking filters will enhance the overall system performance which is constrained by the accuracy of their channel estimates.

• Selective Diversity Combining

We previously noted that by nulling taps whose signal fall below some predetermined threshold, we may alleviate some of the performance degradation at low SNRs with a large number of taps. A problem does, however, exist in the setting of these threshold values. Nulling all taps below a certain power value often leads to a loss rather than gain in performance. This is due to the noise peaks being given a lot of weight while low level signals are nulled.

Nulled taps must be held at zero until the averaged signal power on that tap again exceeds the null threshold. Averaging must be performed over several fade periods in order to reduce the noise floor asymptote and enhance the excess delay signal, figure 5.33. We can then determine which taps do not contain any useful multipath signals at that time.

This technique is a form of selective diversity. After the desired taps are selected we apply MRC in the usual manner. Unfortunately, if the channel SNR is particularly low it may not be possible to apply adequate averaging before the power profile has changed.
considerably. This problem is compounded if the mobile is travelling particularly fast which will cause the averaged multipath power profile to change rapidly. This results from the changing surroundings around the MS, and can limit the potential of this technique when it would be most desirable. The difficulties in applying selective diversity, and the complexity in calculating and applying the thresholds constrains its usefulness in comparison with other techniques.

5.2.5 Spectral Estimation of Channel Fading

Better tracking filters will inevitably be of a higher order. This, unlike the alpha tracker, allows us some control over the filter's frequency response shape, but adds complexity. Ideally the design of the tracking filters should take account of the statistics of the fading observed on the channel. These fading statistics will be time-variant and so it would be desirable to adapt the tracking filters to the channel. Real channels may have fading statistics which are not so clearly defined as the low-pass and Doppler ones used in the channel simulator.

Figure 5.34 illustrates the typical scattering function which has been measured from an urban channel. These results were obtained from measurements made using wideband
Figure 5.34 - Scattering Function for a Measured Channel

(10MHz) soundings at 436MHz where the speed of the mobile receiver was 2.8 m/s (~10km/hr). Note that the negative and positive Doppler spectra are not dissimilar. However the Doppler spectra observed on each delay, or group of delays, are quite variable in shape.

This same scattering function information can be obtained from the post-correlation signals. This is similar to calculating the range and Doppler bins used in pulse Doppler radar. We must use some spectral estimation technique over several message bits on each tap in order to obtain the Doppler information. This can be obtained using the fast Fourier transform (FFT).

If we assume that there will be no significant Doppler components beyond about 150Hz, we need only sample at the Nyquist rate of 300Hz in order to successfully measure this fading. The effective sampling rate of the signal is determined by the bit rate of the system which at 4.8kbit/s represents 16 times oversampling. We can reduce the sampling rate by decimation in time [102]. Rather than simply throwing away the oversampled

† After Bajwa & Parsons [72]
information, we may digitally filter the 4.8kbit/s signal to the narrower bandwidth required prior to decimation. This will help to reduce the effects of channel noise and incorrect data decisions.

A 15 tap FIR filter with a 160Hz cut-off is used to "block filter" 15 samples at a time. It reduces the effective sampling rate to 320Hz, i.e. we have produced a multirate decimation filter. The filter was designed by defining the ideal "brickwall" filter desired in the frequency domain. The filter coefficients were than obtained by taking the inverse FFT and applying a Hamming window function to this time series [103]. The decimation technique greatly reduces the complexity of the spectral estimation problem when using FFTs.

A 64-point complex FFT can provide a useful 5Hz frequency resolution. We have already noted that both positive and negative Doppler generally have a similar shape, and so we may combine them to produce a single-sided Doppler spectrum over the range 0 to 160Hz. This multirate FFT spectral estimation technique has been used to estimate the spectrum of the fading signals generated by our channel model. Ideally the estimated spectrum will match those of the noise shaping filters which were used to generate the fading.

Figure 5.35 shows the estimate obtained over a simulated 10 second (real-time) interval for both the 10Hz low-pass, figure 5.35a, and the 100Hz Doppler, figure 5.35b, types of Rayleigh fading. The magnitude response of the noise shaping filters are shown for reference purposes in each case. As each FFT requires 15x64 message periods (0.2 seconds), this is the average of 50 FFTs within the 10 second interval.

The low-pass result indicates a good overall estimation of the Butterworth shape, although the stopband appears at a higher level than the noise shaping filter used. Much of this can be attributed to the high sidelobes obtained when using no time windowing (effectively rectangular windowing) on the data samples. The 100Hz Doppler result clearly gives a good indication of the spectral shape. Again the effect of the FFT sidelobes are noticeable in the stopbands but are of little practical significance.
a) 10Hz low-pass

b) 100Hz Doppler

Figure 5.35 - FFT Estimation With 100dB SNR and Decimation Filtering Over a 10s Interval
Figure 5.36 shows the effect of the estimation interval on the quality of the spectrum obtained for the 100Hz Doppler case. Note that the absolute power levels are of no consequence, it is the shape and relative powers (e.g. peak to sidelobe) which are important. The curves were not normalised in power, but appear at different absolute power levels and consequently are more easily studied. The graphs are plotted for 0.2s, 1s, and 10s, which corresponds to averaging 1, 5, and 50 64-point FFTs respectively. The ideal noise shaping filter shape is again shown for reference. Figure 5.36a illustrates how the estimate improves with the degree of averaging with a 0dB mean SNR. Figure 5.36b shows a similar situation with 10dB mean SNR, however we can see that the peak to sidelobe amplitude is now much closer to that of the ideal spectrum. This is particularly clear if we compare the two 10s results.
Figure 5.36 - FFT Estimation With Decimation Filtering and no Windowing

a) 0dB SNR

b) 10dB SNR
Figure 5.37 - FFT Estimation Over a 1s Interval
Figure 5.37 is used to illustrate the effect of using a window function [104] (the Hamming window) on the quality of the estimation. We also analyse the effect of the multirate decimation block filtering. Figure 5.37a shows that the use of the window function has little effect other than to widen the main peak of the desired function. This actually causes a slight distortion to the spectral shape and so the use of this window is inadvisable when using this technique as the higher sidelobes are of minor consequence here. The noise floor will often obscure these sidelobes anyway. Figure 5.37 also demonstrates that the decimation filter improves the estimation by lowering the noise floor by about 5dB. Without this filter we are simply throwing away 14 out of every 15 samples. Figure 5.37b shows the same curves but for a mean SNR of 10dB. The main difference is that the noise floor is lower, but the rest of the estimation is almost unchanged. The improvements gained from the use of the decimation filter is not so noticeable at the higher SNR value.

It is interesting to note that fairly good spectral estimations can be formed using just 5 64-point FFTs and a 15 tap decimation filter, even for SNRs of 0dB. The spectral estimation complexity involved is therefore fairly minimal. Commercially available DSPs are currently capable of performing 64-point complex FFTs in less than 100μs (28μs for a specialist FFT device [105]). It looks likely that a single DSP could be used for the spectral estimation on all of the significant taps required in the RAKE receiver, and in addition to this it could implement the adaptive tracking filters derived from the spectral estimations. There may also be the possibility of implementing the entire adaptive RAKE receiver, including the adaptive tracking filters, on a single DSP device.

5.3 Integrated Receiver Design

In Chapter 3 we saw that the serial-parallel receiver is capable of providing a section of the full correlation profile. If the 1024/32 chip receiver is operated at 5Mchip/s this will provide 32 chips of ACF equivalent to a delay of 6.5μs. The main peak can be synchronised to appear at the start of the correlation profile such that we can obtain the urban multipath information within the post-correlation section provided. It is therefore
possible to successfully integrate the serial-parallel receiver with the adaptive RAKE receiver. Previous systems tended to use separate serial (active) correlators for each component of the multipath signal when long PN code lengths are used [106].

It is theoretically possible to obtain slightly improved performance by adding the RAKE's FIR in place of the 32 chip delay on the feedback of the serial-parallel correlator. The RAKE coefficients could then be updated every 32 chips rather than 1024. However it would still suffer from the decision feedback delay which is more significant. The hardware complexity would be greater since the serial-parallel section of the architecture could no longer be contained within a single device and the calculation of the RAKE coefficients becomes more involved.

5.4 Summary

We have seen that the simple alpha tracker can be used to estimate the time varying channel's impulse response. The accuracy of the estimate is dependent on the fade rate and the channel noise. Problems of intersymbol and interpath interference are negligible for a carefully chosen 1024 chip code.

The performance of the adaptive RAKE can be predicted by assuming that the tracker phase errors and the noise in the estimate are uncorrelated additive noise. The performance of the receiver for 10Hz low-pass fading is near to the theoretical upper bound. The performance obtained for the 100Hz Doppler channel was poorer and characterised by an IBER. Much of the error was due to the phase lag of the decision delay. Improvements to the tracking filter would improve the performance but may still be bounded by the decision delay. Even with the poorer performance, adding up to 6 orders of diversity greatly enhanced the capabilities of the spread spectrum receiver.

It is clear from the burst error performance of the adaptive RAKE that fears about its limited robustness to noise are largely unfounded since the performance degrades
gracefully and predictably as the channel SNR reduces. The mechanisms which reduce
the burst length are by the increase in effective SNR and the reduction in fade durations,
both of these result from the systems inherent multipath diversity.

We have also seen how the performances of sub-optimal combining techniques can
approach, and under certain circumstances exceed, those obtained for MRC applied to
the fast fading 100Hz Doppler scenario. This is a result of the noise tracking and the
filters phase lag which can cause large estimation errors.

Higher order filters are indicated as an obvious way to improve the RAKE's tracking
performance. We have highlighted an efficient multirate FFT spectral estimation
technique which can be used to help define the requirements of the improved tracking
filters which will ideally be adaptive.

Integrating the 1024/32 serial-parallel receiver described in Chapter 3 with the adaptive
RAKE described here is possible. This produces a very efficient design which is robust
and performs well over a multipath fading channel.
Chapter 6

CONCLUSIONS

6.1 Summary of Results

We have seen that the digital implementation of a serial-parallel receiver has a superior performance to that of an equivalent analogue system. Measured implementation losses of less than 0.1dB were achieved, compared with 1-2dB for the SAW receivers. The advantages of considerably greater processing gain, and the ability to operate with any type or length of code are also significant advances.

The speed of general purpose DSP devices, at the present time, severely limits their potential for use in spread spectrum receivers. A large processing gain with a high bit rate is not readily achievable using the parallel matched filter architecture. The serial-parallel receiver has, however, provided more flexibility to the design of DSP receivers. For example, we can achieve a 6kbit/s data rate with a processing gain of 18dB and synchronisation time of 1.3ms while obtaining access to 8 chips of the correlation profile for a 64/8 correlator. Larger processing gains such as the 30dB of the 1024/32 correlator result in much lower bit rates (100bit/s) and consequently have a poor synchronisation time (320ms). Implementing such a design on custom DSP hardware can dramatically improve this data rate and synchronisation time - by about two orders of magnitude in the case of the Inmos A100 device. The demonstration system did, however, enable us to prove the operation, performance, and limitations of the digital serial-parallel receiver using general purpose DSP devices. Also it showed how the large bandwidth of this spread spectrum receiver can provide access to the multipath components which are present in a dispersive channel.
The development of a realistic UHF wideband simulator allows us to create a variety of realistic channel conditions. The bit rate of the simulator was dependent on the number of multipath taps modelled. On a SUN 4 IPX workstation the simulator was capable of a peak update performance of 10,000 tap/s. The number of multipath components (taps) required determines the peak simulation bit rate. It was found to be fast enough to conduct Monte Carlo type BER simulations so long as very low error rate results were not required. These timings do not include the processing time required to model the receiver.

The adaptive digital RAKE receiver was found to operate very well for the simulated hand-held channel scenario. These 10Hz low-pass fading results show only a slight deviation (<1dB) from the upper bound performance. Irreducible BER characteristics did not degrade the performance significantly. The good performance is attributable to the faithful tracking which is possible using the modified alpha tracker filter at these fading frequencies. The phase error in the tracking filter’s estimation are low due to the slow fading. This is also true of the estimation lag caused by the decision feedback.

The BER curves for the 10Hz low-pass fading indicate the benefits which can be obtained by using the inherent diversity caused by the fading multipaths. To obtain a BER of $10^{-3}$ with no diversity, we must have a mean SNR of about 27dB on a Rayleigh fading channel using DPSK modulation. Using the adaptive RAKE with 3 taps we can reduce the SNR to 12dB, or to only 6dB with 6 taps. These relate to dramatic reductions in the transmit power requirement. In practical systems the reductions in power will not be quite so large since we have assumed perfect synchronisation. Combining the components from the less significant taps is not really required as only a minor increase in performance is achieved but there is a significant increase in complexity. Six taps gave very good results, doubling this to 12 can give a further 2dB of improvement, however, increases in complexity beyond this point do not appear to be worthwhile. The excess delay profile used in the simulator is an averaged profile and will not always correspond with that of a real channel for any specific location. This may lead to more variable results when using a non-simulated channel.
The equivalent BER results for the high speed mobile scenario did not produce such a good performance. Even so there were still significant improvements over conventional receivers with no diversity. The 100Hz Doppler BER curves did not agree well with the upper bound performance curves. There are two mechanisms which have caused this degradation. The tracking estimate includes a large proportion of the noise which leads to a degradation in BER performance seen as a change in the slope of the curves, and the tracking estimate has a large phase lag which leads to a further worsening of the BER slope and also it introduces the IBER asymptote. At a BER of $10^{-3}$ we have a loss of about 2-3dB compared with the 10Hz low-pass case. The difficulty in tracking the fading is largely attributable to the fast fading of the signal relative to the bit rate. If we were to increase the bitrate by a factor of ten, then we could expect a BER performance similar to that of the 10Hz low-pass results.

Modification of the BER equations showed that we can predict the IBER values fairly readily. For the 100Hz Doppler case we can reduce the IBER by at least one order of magnitude with the addition of one extra significant tap. From about 3 taps onwards the IBER becomes insignificant if we are interested in a BER of $10^{-3}$.

Burst error CDFs have served to show that the adaptive RAKE is robust in the sense that it degrades smoothly with noise and does not exhibit sudden divergent behaviour. The curves also illustrate that we are correct in indicating that the IBER results are caused exclusively by the zero crossings of the fading voltage envelope, since at close to the IBER the CDFs indicate mostly the presence of single bit errors which cannot be attributed to noise as we are using DPSK modulation. By adding more taps to the RAKE we observed a reduction in burst error length due to the reduced fade durations and increased SNR.

It is clear from the 100Hz Doppler results that although we were attempting to use maximal ratio combining of the signals - which should yield optimum performance, we did not achieve this ideal. The performance comparisons with the sub-optimal combining techniques have revealed that they can perform almost as well as the maximal-ratio
combining under certain conditions. EGC is, in most respects, poorer except at fairly high SNR values where there is an improved IBER value. DPC performs very well in comparison, only at fairly low SNR values or large orders of diversity does the MRC scheme show any obvious improvement. These comparisons have indicated that if a good channel estimate cannot be formed, then a poor attempt at MRC combining may sometimes be less desirable than some of the inherently sub-optimal combining techniques. This has further emphasised the significance of the phase errors within the channel estimates. The use of selective diversity is difficult to implement and is of limited benefit if sufficient performance can be obtained from the first few taps.

6.2 Future Directions

We have concluded that even if tracking filters with better phase responses were used the performance would not be greatly improved due the decision feedback error. What is now required is a forward prediction of the channel impulse response whereby we can eliminate the phase lag. This can probably be achieved using a linear prediction filter [107] [108]. We have shown how it is possible to determine the power spectrum of the fading using an efficient multirate decimation and FFT technique. The fading power spectrum changes with vehicle speed and the inhomogeneous topology of its surroundings. A fast spectral estimation technique allows us to obtain the autocorrelation coefficients via the inverse FFT. From this we can form the Yule-Walker equations [109] which can then be solved to determine the next forward predicted result. Separate predictors will be required for each tap. However it may be possible to calculate the prediction coefficients for a group of taps. Such an adaptive prediction filter will enable us to adapt the prediction coefficients in accordance with the fading statistics observed on the channel. They must, however, be updated at regular intervals since these statistics are dependent on the mobile’s speed and location.

As no experimental results have yet been obtained for this proposed technique, we cannot determine how successful it will be. The predictors must operate well in variable SNR signals, and their coefficients should be updated fast enough to cater for the time-variant
nature of the channel. We must also consider the complexity involved in such a technique. It is undoubtedly more involved than that of the modified alpha tracker. Determining the performance and complexity of an adaptive RAKE using linear prediction should be the aim of a future research programme which will hopefully take the adaptive RAKE receiver to a field trial stage.

It is suggested that RAKE techniques will be as commonplace in CDMA cellular systems of the future, as equalisation techniques are becoming in current TDMA cellular systems. Their effectiveness in reducing the power requirements, and their improvement of the frequency reuse efficiency, cannot be overlooked. It is still not clear what code lengths will be required for a commercial CDMA system. If these codes are to be greater than, say, 128 chips then it is definitely worthwhile considering the use of a serial-parallel receiver architecture which can interface easily with the RAKE receiver.

6.3 Summary

We have demonstrated the considerable performance advantages of digital serial-parallel receivers as compared to earlier analogue implementations, and how they can reduce the receiver complexity. We have shown how the adaptive RAKE can be used to combat multipath effects and so reduce the required transmitter power. The deficiencies of the RAKE's tracking filters have been highlighted and linear prediction techniques have been proposed to help overcome some of the performance degradation.

Thus, a considerable insight has been provided to the performance which can be expected from the application of direct sequence spread spectrum techniques to urban mobile communication systems.
REFERENCES


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Appendix A

ORIGINAL PUBLICATIONS

The following original publications have resulted from the research programme documented in this thesis.


* Reprinted in this appendix.
The operation and characteristics of serial, parallel and in particular serial-parallel correlators used in a spread spectrum receiver are described. The suitability of a DSP device for this application is highlighted and its operation is illustrated using the Texas Instruments TMS320C25 device. The potential for such systems to be reconfigured or adapted during operation is discussed and shown to be a major advantage over fixed architecture receivers.

INTRODUCTION

One of the primary functions of a spread spectrum receiver is that of correlation. The received signal is correlated with the same pseudo-noise (PN) code which was used in the transmitter, this effectively "despreads" the signal. The correlation may be performed serially or in parallel. A serial correlator, figure 1, integrates the product of the received signal and a locally generated PN code over a data bit period, the result is then dumped and a new integration started. The local PN code must be correctly synchronised such that it is aligned with the received signal before any data decisions can be made. Only the correlation peak is found from a serial correlation (assuming correct alignment). The parallel correlator, figure 2, performs a similar task only the local PN code is static and the integration is performed in parallel and so is completed within a single chip period. This operation is more numerically intensive but does not require any code alignment, also the entire correlation profile (i.e. the peak and sidelobes) is obtained and not just the peak.

The correlation function sidelobes can contain very useful information about the channel such as the multipath echoes and channel noise level, this information can often be used to enhance the performance of the receiver [1]. Thus a parallel receiver architecture is desirable but is not always the most practical since tens of correlator devices may be required in cascade to realise the desired correlation length. Commercial correlator devices tend to have 8-64 stages, spread spectrum systems can require lengths in excess of 1000 bits or "chips" as they are termed.

THE SERIAL-PARALLEL RECEIVER

It is possible to combine the serial and parallel architectures to produce a hybrid serial-parallel receiver [2] which has the advantage of fast synchronisation, partial sidelobe functions and relatively low complexity (or numerical intensity). In effect the characteristics of the two conventional structures may be combined in order to obtain a more suitable system configuration.

The operation of the serial-parallel correlator, figure 3, can be described as follows. The first \( N \) chips of the received signal are correlated with \( N \) chips of the PN code in parallel, the output of this stage which is a partial correlation is fed through the adder and into the digital delay line. The next \( N \) chips of the received signal will then be correlated with the next \( N \) chips of the PN code, this partial correlation will be added coherently to the previous one which has been delayed by \( N \) chips. This procedure is repeated until the a section of the full \( M \) chip correlation profile is obtained. The correlation peak can be observed growing linearly.
throughout the bit period. The feedback to the adder is disconnected for the first \( N \) chips so that the delay line is reset each bit period.

Figure 3 - The Digital Serial-Parallel Correlator

If there are \( M \) chips in a bit period of \( T \) seconds, the maximum time to acquire synchronisation (i.e. find the correlation peak) for a parallel correlation will be just \( T \) seconds. For a serial correlator this time will be \( M \times T \) seconds since \( M \) possible code alignments may have to be tested. This time is reduced to \((M/N) \times T\) for the serial-parallel correlator which can test for synchronisation over an \( N \) chip window during a single bit period. The system complexity has also been reduced since we require only one correlation device compared with \( M/N \) for a parallel correlator. It is often possible to engineer the system such that the \( N \) chip section obtained contains the required channel information so there is no major advantage in performing a full parallel correlation.

**DSP BASED SERIAL-PARALLEL RECEIVER**

To prove the operation and performance of the digital serial-parallel correlator hardware the system was implemented on a Texas Instruments TMS320C25 DSP system [3] running at 10 MIPS.

A serial correlator requires a total of \( M \) multiply and accumulates (MACs) in a data bit period, while the parallel correlator must perform \( M^2 \) MACs in the same period. The DSP processes serially so the number of MACs is proportional to the processing time for one bit. It therefore follows that the DSP is fastest at performing serial correlations, however a serial-parallel correlation will improve the acquisition time and provide channel information. The serial-parallel correlation requires \( M^2/N \) MACs which can be a significant saving over the parallel correlator.

The DSP implementation did suffer a processing overhead in changing the PN code coefficients after every \( N \) chips and managing the counter. For a 1024 chip correlation with \( N=32 \) (written as a 1024/32 i.e. \( M/M \), a bit rate of 100 bits was achieved. A 64/8 correlator was however capable of almost 10 kbits/s. These performance figures could be bettered by a factor of almost 3 simply by transferring the system to the faster TMS320C50 device which can operate at up to 25 MIPS and has more on-chip RAM where all the PN code coefficients can be held. The system was designed to be very flexible in that a single variable controls the virtual parallel device length \( N \). If \( N=1 \) we effectively have a serial correlator, if \( N=M \) it is a parallel correlator, and if \( M/N \) is an integer we obtain a serial-parallel correlator.

Figure 4a shows the operation of the 64/1 serial correlator, the upper trace is the PN signal which has been modulated with random binary data, the lower trace is the post-correlation output. The polarity of the output indicates its data content, both polarities appear on the output due to the random data and the persistence time of the photograph. Figure 4b shows the same system only Gaussian noise has been added to the input signal (input SNR=9dB). Figures 5a & 5b show the equivalent traces for the 64/64 parallel correlator. Figures 6a & 6b show the operation of the 64/8 serial-parallel correlator. Figures 7a & 7b (input SNR=-20dB) are for the 1024/32 correlator. Note that the partial correlation peaks grow linearly while the post-correlation noise grows in a root mean squared sense. This illustrates how the despreading process of the spread spectrum receiver is capable of extracting the data signal from what appears to be just noise but in fact contains the spread signal which has a low SNR. The theoretical improvement in SNR from the correlators is 18dB and 30dB for the 64 and 1024 chip correlators respectively. The only real factor which can cause a reduction in processing gain is the precision and accuracy of the ADC at the input and this becomes insignificant if more than about 4 bits are used [4]. A 12-bit conversion was used so this loss was negligible. There was no detected degradation of processing gain relative to the theoretical limit when this was measured statistically from the DSP output. The arithmetic resolution of the DSP constrains the maximum processing gain, but with over 90dB dynamic range for the 32 bit accumulator it is highly unlikely that such an application would approach this limit.
THE RECONFIGURABLE RECEIVER

The DSP system is extremely flexible since the receiver architecture is virtual and can be reconfigured even while active. The receiver may be adapted during operation to cater for channel conditions, security, data types, etc. Figure 8 illustrates the 3 fundamental modes of operation - serial, parallel and serial-parallel, the upper trace is the input PN signal. These modes of operation can be swapped during transmission without loss of data, it is also possible to adjust the correlator length although this can result in the loss of one bit of data. The ability to reconfigure the receiver architecture could be very powerful since different signal processing requirements can be performed using the same piece of hardware, thus the receiver need not be restricted to use with a single transmitter type or mode of operation. Possible applications include versatile modems, low-power telemetry systems, and secure or covert communications systems.

Figure 8 - The Reconfigurable Receiver

CONCLUSIONS

The speed and flexibility of the DSP system described indicates that it may be possible to implement some spread spectrum receivers directly onto general purpose DSP devices. The serial-parallel architecture has been shown to operate close to the optimal performance achievable using a general purpose DSP device. Speeds of up to 10kbit/s have been obtained for an 18dB processing gain using a TMS320C25, this figure can be bettered by using faster DSPs. The precision of fixed point DSP devices is adequate for very long correlations. The operation of a reconfigurable correlation receiver has been illustrated and a few applications suggested.

REFERENCES


REAL-TIME ADAPTIVE MULTIPATH COMPENSATION

J. S. Thompson, G. J. R. Povey and P. M. Grant

Indexing term: Mobile communications

Multipath propagation interference in urban mobile communications can be minimised by using spread spectrum modulation and a RAKE filter. The Letter describes the simulation and practical operation of an adaptive or programmable RAKE filter, employing maximal ratio combining, which can handle time varying multipath channels in either the base station or mobile receiver.

Introduction: Signal propagation in urban areas is subject to multipath effects which involve reflections from buildings and other objects. On average the secondary paths are of longer delay than the direct link, with more attenuation and the average urban channel impulse response generally follows a decaying exponential function [1] of up to 5 μs duration.

A direct sequence spread spectrum system [2] encodes digital narrowband data or bandwidth $W_1$ with a fast polar pseudo-noise (PN) code to give a much larger transmitted bandwidth, $W_2$. The receiver uses correlation to detect the incoming PN codes and, with a suitable choice of $W_2$, it is possible to resolve these multipath returns separately.

However, the received signal power for each symbol is distributed among the different multipath returns $x_0(n) ... x_3(n)$, which generally follow the main correlation peak, and these must be subsequently combined to enhance the signal-to-noise ratio (SNR) for decision making. This is achieved by the RAKE filter, Fig. 1 [3], which implements a filter matched [4] to the decaying multipath channel. The RAKE filter exploits fully the time diversity in the multipath environment in order to achieve improved performance compared with using only the largest of the received signal samples. It is designed with a sample rate appropriate to $W_2$ and filter length (32 in Fig. 1) to accommodate the (5 μs) multipath dispersion. It should be
noted that symbol duration or bit period (1/W) can be much greater than the filter delay in Fig. 1.

**Software design**: This RAKE finite impulse response filter was programmed onto a TMS320C25 processor which also allows the RAKE filter to become adaptive in order to handle time varying multipath conditions. Each of the RAKE filter impulse response coefficients, h0(n) ... h31(n), are controlled by an alpha tracker (Fig. 1 inset) so that we may approximate to maximal-ratio combining of the signal. This modifies each filter coefficient, h0(n), once per received symbol, using the corresponding multipath signal sample x(n) such that h0(n) approximates to x(n) to obtain the channel matched filter coefficients. The z value controls the bandwidth of the multipath signals passed to the RAKE filter coefficients: z varies from zero (infinite bandwidth) to one (zero bandwidth). Thus in comparison to previous RAKE filter designs [5], which integrated the returns from a stationary channel or employed a separate channel sounding to obtain the RAKE response, our implementation is fully adaptive and tracks the fast fading mobile response.

**System performance**: The operation of a three tap adaptive RAKE filter, incorporating three separate z trackers, is initially shown without data transitions by examining the response to a step change in the multipath profile, Fig. 2. Here three out of the 32 samples are processed in the filter to obtain each output response. The RAKE filter is poorly matched to the first symbol it receives and the main lobe of the filter output is not much bigger than the side lobes. As each training symbol is processed the filter coefficients converge to the correct values, enhancing the main lobe and reducing the side lobes.

Fig. 2a demonstrates that the low value of z has a better step response than a larger alpha factor (Fig. 2b). Therefore, z needs to be small to allow the RAKE filter to converge quickly to the multipath distribution. However, under high noise conditions z must be large to implement the required averaging operation, as in previous RAKE designs [5]. The performance of a RAKE filter over a fixed multipath channel was verified by the measured results shown in Fig. 3. Curve (i) shows the performance of a polar spread spectrum system without multipath and curve (iii) indicates the 3 dB degradation for a multipath channel where half the received signal power is present in the two delayed paths. Curve (ii) demonstrates the improvement due to the DSP hardware RAKE filter after it has combined the separate multipath components. The demonstration, as configured with the general purpose TMS320C25 processor, has a 13 kbit/s data rate capability. When simplified for data detection the rate would increase beyond 10 kbit/s. Faster DSPs or custom FIR chips, such as Inmos A100, would increase the data rate further.

**Fading channel conditions**: Wideband mobile communications systems are subject to frequency selective fading of the multi-path signal [1, 6], where the phase and amplitude of each multipath signal component are independently changing and the receiver is subject to alternating constructive and destructive interference. This complex signal requires the alpha tracker to be extended to include decision feedback. This necessitates a delay of one symbol for the decision, degrading the phase response thus making it difficult to coherently follow the absolute phase of the received signal. This deficiency was minimised by using differential phase shift keying (DPSK) data transmissions.
The performance of this complex spread spectrum demodulator was simulated and illustrated in Fig. 4 for a fast fading multipath channel with a 100 Hz Doppler component. The curves for a single tap indicate the poor performance without a RAKE filter. Adding extra taps to the RAKE filter considerably improves the performance of the system, but it is clearly limited at high SNR by the irreducible bit error rate (BER) caused by the phase lag in the alpha tracker. The BER improves considerably as the RAKE filter complexity is increased. Having three active taps, within the 32 processed samples, produces an acceptable BER of $5 \times 10^{-4}$. The mean SNR axis in Fig. 4 is the average SNR as observed on the most significant tap.

![Fig. 4 Simulated error rate performance for complex processing RAKE receivers of increasing complexity on fast fading multipath channel](image)

(i) one tap
(ii) two tap
(iii) three tap
(iv) four tap
(v) five tap

Conclusion: The adaptive RAKE filter, implemented on a DSP general purpose or custom processor, is a powerful device for overcoming multipath interference. Selective fading effects, common in mobile systems, can then be ameliorated by the use of a decision-feedback alpha tracker with DPSK data coding and only three or four taps should provide an acceptable BER performance for many applications.

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ABSTRACT

This paper details a digital RAKE receiver which uses the inherent post-correlation signal diversity and decision feedback to form a channel estimate. The fading multipath signals are tracked using simple IIR filters. Performance is shown to approach the theoretical limit for low-pass Rayleigh fading. For faster Doppler type fading the system is less robust particularly for low orders of diversity. New techniques are proposed to improve the tracking of fast fading signals which are based on adaptive filters with coefficients derived from frequency domain channel characteristics rather than from the time domain.

INTRODUCTION

There are essentially two elements of the spread spectrum receiver design which enable it to perform well in a multipath environment. The first element is the correlator which despreads the incoming signal and resolves the received multipath components by providing a delay profile for the channel. The second element is the RAKE receiver which attempts to recombine the multipaths such that we can use the majority of the received power rather than just the single main path component.

The correlators ability to resolve the individual multipath returns implies that intersymbol interference (ISI) is not a major problem at typical bit rates. There can, however, still be a large received power loss due to the multipaths diverting energy from the direct path signal. Narrowband systems have attempted to address this problem by using adaptive equalisers with training sequences or decision feedback. Such techniques are difficult to apply in spread spectrum systems since the high noise on the channel severely restricts the adaptive algorithms convergence. A RAKE type receiver, on the other hand, uses the aposteriori knowledge of the channel provided by the correlator. The post-correlation signal can be resolved into components each with a different associated delay, phase, and magnitude. The post correlation multipath delay profile thus provides a measure of the channel impulse response (or rather it is an estimate since it will contain an element of noise). This can be used to construct a second digital matched filter (DMF) receiver which is matched to the channel response. This method of post-processing the multipaths [1] uses maximal-ratio combining [2].

THE RAKE RECEIVER

To operate robustly the channel matched filter must be programmed with a good estimate of the channel's multipath delay profile. This estimate can either be obtained by using a channel sounding signal or it can be found directly from the post-correlation signal once the data content has been removed, figure 2. A problem with channel sounding techniques is that the bandwidth overhead becomes very large if we wish to make accurate estimates for a fast fading signal. The direct method on the other hand requires decision feedback with its associated delay and incorrect decisions may lead to bursts of errors. The divergent nature of the estimator is largely eliminated by providing a sufficient number of taps such that the errors are minimised. The use of a DPSK modulation
scheme also tends to break-up error bursts and reduces the divergent behaviour, especially for good SNRs.

Doppler (equivalent to a vehicle speed of 120km/hr on a 900MHz carrier).

![Figure 2 - The RAKE Receiver](image)

The performance obtained from this arrangement can be better than that achieved using indirect channel sounding when dealing with fast fading signals such as the Doppler fading observed on mobile radio systems. The fast fading signal estimate is updated every data bit period.

Channel estimates will be subject to noise and cannot directly provide a good representation of the channel impulse response. An improved estimate is obtained when each element of the profile is low-pass filtered in order to reject as much of the noise as possible while retaining the signal, this also reduces the effect of decision feedback errors. If the channel is subject to slow fading (i.e. the channel impulse response changes very slowly) then a narrow bandwidth filter should provide a good estimate. Often this is not the case, e.g. where there is Doppler fading due to a mobile transmitter or receiver. Here the ideal filter is the one which will minimise the RAKE's mismatch loss.

FADING CHANNEL MODELS

In order to test the performance of the RAKE receiver a Rayleigh fading channel simulator was used to model a typical urban UHF channel [3]. The most severe channel scenario, which the RAKE must contend with, is associated with a fast moving transceiver in an urban area with low signal power. A typical time history for a single path fading signal (including Doppler) is shown in figure 3. The model presented here has a message rate of 4.8kbit/s, Rayleigh fading and additive white Gaussian noise (AWGN). The fading is modelled by either 10Hz low-pass characteristics (worst-case hand held operation), or 100Hz Doppler (equivalent to a vehicle speed of 120km/hr on a 900MHz carrier).

![Figure 3 - Doppler Rayleigh Fading Signal with Noise](image)

**THE RAKE PERFORMANCE**

The low-pass tracking filter used was a single pole IIR alpha tracker, defined below.

\[
y(k) = ay(k-1) + (1-a)x(k-1)m(k) \quad (1)
\]

where \(x(k)\) is the channel measurement at time \(k\), \(y(k)\) is the channel estimate, and \(m(k)\) is the message data decision \((-1,+1)\) which is used to "correct" the phase of \(x(k)\). This decision feedback introduces an additional phase error on the estimate.

This filter was found to be effective for slow fading signals, provided by the 10Hz low-pass model, but it is less capable of tracking faster Doppler type fading since it introduces a severe phase lag. Figure 4 indicates that the simple alpha tracker introduces a irreducible bit error rate (IBER) of \(10^{-5}\) for a 100Hz Doppler Rayleigh fading signal with no diversity, whereas the 10Hz low-pass fading follows the theoretical curve (i.e. for a perfect estimate) more closely, the IBER is reduced to about \(5 \times 10^{-5}\). The IBER is caused by tracking errors which can cause single bit errors when the fading envelope crosses the zero threshold.

The theoretical error rate for Rayleigh fading is given below [4]. This is for ideal tracking and therefore it does not have an irreducible error rate.

\[
P_e = \frac{1}{2(1+\Gamma_0)} \quad (2)
\]

where \(\Gamma_0\) is the mean SNR for the Rayleigh fading signal.
Figures 5 and 6 illustrate the improvement in BER performance obtained by using maximal-ratio combining with independent alpha trackers on each of the taps, for low-pass and Doppler Rayleigh fading respectively. Note that each successive tap is assumed to contain the next largest mean SNR fading multipath signal for a typical urban environment. The mean SNR is defined for the largest fading signal component (i.e., the main correlation peak) this ensures that the improvement in performance afforded by the multipath diversity is clearly visible on the BER curves. DPSK signalling was used.

The nature of the bit errors can be analyzed by looking at the burst error cumulative distribution function (CDF), figure 7. These curves indicate that the diversity receiver does substantially reduce the length of error bursts which can cause the channel estimation to diverge. We define the error burst length as the interval between two error free regions of at least 8 bits. Single errors tend to be caused by the fading signal envelope zero crossings where the tracking filters phase lag is critical. Double errors are more likely to be caused by noise. The IBER is characterised by single errors only. The burst error curves of figure 8 (one tap only) can be seen to be tending towards the IBER with increasing SNR, this can be related to the single tap curve of figure 6.

AN IMPROVED RECEIVER DESIGN

To improve upon the performance of these alpha trackers it is necessary to look at higher order filter types. We basically wish to optimise their filter re-
Observations of an urban mobile channel [5] reveal fading in the time domain which is considerably faster than the changes in the channel frequency statistics. High frequency components of these changes will tend to be temporary and are caused by the mobile passing close to large objects at high speed. It is therefore desirable to track the multipath Doppler signal in both frequency and time, however the tracking in frequency need not be as fast as the tracking of the fading signal multipaths. This is useful since the two tracking loops will not interact to produce divergent behaviour resulting in error bursts. Both tracking loops will require decision feedback.

In order to track the channel frequency statistics we intend to employ FFTs which provide a frequency characteristic template to control the FIR tracking filter, figure 9. FFTs are performed on the time history of samples received at each tap after the inputs are phase corrected using decision feedback (following a bit period delay) and block filtered to reduce the data set and noise. Decimating the data by 16 gives a sampling rate of 300Hz. A 64 point FFT produces an estimate every fifth of a second with a frequency resolution of 5Hz. The FFT output is folded around the 150Hz point and added such that there is no distinction between positive and negative Doppler components. A simple persistence filter on the output helps to suppress any minor short term fluctuations. The FIR tracking filters are difficult to design due to the minimum phase requirement and still suffer from the decision feedback delay.

This technique can produce improved tracking for fast Doppler fading. A realistic channel model which takes account of the frequency statistics for a mobile moving through a inhomogeneous environment is difficult to characterise and hence the system performance will be determined using real channel data, compared with the non-adaptive RAKE performance using the same data source.

It has been estimated that the frequency domain processing could be performed on a fixed-point DSP device and would require about 0.1 MIPS per tap. This is based on 64 point radix-2 complex FFTs and the associated filter processing.

CONCLUSIONS

The RAKE's ability to track the independently fading multipaths is dependent on the RAKE filter design, and typically 3-12 tap complexity is required for a BER of 10^-4. The alpha-tracker is suitable for slow fading but performance is noticeably degraded for a severe Doppler environment. However, if the RAKE tap filters are designed in real-time using channel frequency statistics, the superior channel profile estimate will improve the receiver performance.

REFERENCES

Appendix B

DSP SOFTWARE

• The "README.TXT" File

The following information is simply a copy of the "readme.txt" file contained on the floppy disk attached to the inside of the back cover. The 5.25-inch disk has been formatted using the MS-DOS 360kbyte standard. Most IBM compatible PCs equipped with a 5.25-inch floppy disk drive should be capable of reading it. The file documents the contents of the disk and how they can be used to build DSP code for the DSP based serial-parallel correlator. It can be read by typing "type readme.txt / more" once the correct drive is selected.

DSP Code for the Serial-Parallel Correlator

The files in the directory "DEMOS" are self-contained executable files which require only the LSI TMS320C25 DSP system board for their execution. The output from the board can be observed on an oscilloscope. The input signals are simulated internally and thus no external hardware is required. No noise signal or data content is used in this limited demonstration. The files are:

- DEMO64S.OUT Serial 64 chip correlator (internal signals)
- DEMO64P.OUT Parallel 64 chip correlator
- DEMO64SP.OUT Serial-parallel 64/8 chip correlator
- DEMO1024.OUT Serial-parallel 1024/32 chip correlator

This disc also contains the files necessary to produce correlators for use with external signals and interrupts. The executable files for the serial-parallel correlators are held in the directory "EXES". These files are:

- SPC64.OUT Serial-parallel 64/8 chip correlator (external signals)
- SPC1024.OUT Serial-parallel 1024/32 chip correlator
The TMS320C25 source code required to produce the correlators are held in the directory "SOURCES", and the code coefficients (for 64 and 1024 chip codes) are held in the directory "CODE_LIB". Batch files to aid the building of the required correlator are held in the directory "BATCHES".

```
$SOURCES$
SPC64.ASM
SPC1024.ASM
SPC.CMD

$CODE_LIB$
COEF64.ASM
COEF1024.ASM
COEF64.OBJ
COEF1024.OBJ

$BATCHES$
SPC64.BAT
SPC1024.BAT
```

To build the correlator the assembly and coefficient sources must be assembled using the Texas Instruments assembler "dspa". The resulting ".obj" files are then linked along with the linker command file "SPC.CMD" using the TI linker "dsplnk". The easiest way of doing all of this is via a batch file. These are provided for the 64 and 1024 chip versions.

Example

To create a 64 chip parallel correlator (i.e. 64/64 serial-parallel correlator). Copy the files "SPC64.ASM", "SPC.CMD", "COEF64.OBJ", and "SPC64.BAT" into a new directory created on the hard-disc. Edit the "SPC64.ASM" file so that M=64 and N=64 and run the batch file by typing "SPC64". It is assumed that there is a path to the assembler and linker set in the "AUTOEXEC.BAT" file and that the DSP system board is setup for external interrupts. The system will run fastest if the ADC is set for a 9.8us conversion time and 12 bit resolution.

When running the code the correct analogue polar PN signal must be presented to the ADC. This can, if required, include noise, data and multipath dispersion. An external interrupt at the chip rate must also be provided as INT1 via the Analogue Control Connector (J2).