Advanced Power Control Algorithm for CDMA Based Systems

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Summary

Code Division Multiple Access (CDMA) has been selected as the main technology for third generation (3G) mobile communication system. Traffic expectations of 3G are high whilst only limited spectrum is available. Any system enhancement that would increase capacity and improve spectral efficiency is highly desired.

Power control is the single most important requirement of any CDMA based mobile communication system. Power control directly impacts the quality of service as well as system capacity. It is for this particular reason that adaptive power control schemes have been the subject of many researchers in recent years.

In this thesis, a novel Environment Detection Algorithm (EDA), an essential component of adaptive communication systems is presented and extensively evaluated. EDA is an important enabling component of Speed Adapted Closed Loop Power Control (SA-CLPC). An advanced SA-CLPC is presented and its superiority in both terrestrial and satellite communication systems is clearly demonstrated. The work follows to introduce a novel interference management technique in the form of a distributed power control algorithm. The proposed algorithm outperforms all its rivals by maximising $C/I$ in response to system load without the need to any knowledge of users in the neighbouring cells.

Key words: CDMA, Power Control, Speed Estimation, Open Loop Power Control, Closed Loop Power Control, Adaptive Power Control, Environment Detection Algorithm, Delay Spread Estimation, Distributed Power Control, centralised Power Control
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I dedicate this work to my mother who

"Sacrificed her life to better ours".
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<td>PSD</td>
<td>Power Spectral Density</td>
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<tr>
<td>QoS</td>
<td>Quality of Service</td>
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<tr>
<td>QPSK</td>
<td>Quadrature Phase Shift Keying</td>
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<tr>
<td>RMS</td>
<td>Root Mean Square</td>
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<td>RNC</td>
<td>Radio Network Controller (Equivalent to 2G BSC)</td>
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<td>RTD</td>
<td>Round Trip Delay</td>
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<td>SA-CLPC</td>
<td>Speed Adapted Closed Loop Power Control</td>
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<tr>
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<td>Spreading Factor</td>
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<td>SIR</td>
<td>Signal-to-Interference Ratio</td>
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<td>SNR</td>
<td>Signal-to-noise Ratio</td>
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<td>Satellite Personal Communication Network</td>
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<td>Time Averaged Distributed Balancing Algorithm</td>
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</tr>
<tr>
<td>UMTS</td>
<td>Universal Mobile Telecommunication System</td>
</tr>
<tr>
<td>UTRAN</td>
<td>UMTS Terrestrial Radio Access Network</td>
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<tr>
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<td>World Radio Conference</td>
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Chapter One

1. Introduction

Prior to the use of Spread Spectrum (SS) techniques in civilian communication systems, its use in military communications had become common thanks to its high level of privacy (security) and inherent robustness against interference and jamming.

Code Division Multiple Access was originally proposed to support simultaneous digital communications among a large community of relatively uncoordinated users. Yet, as recent as 1985 a straightforward comparison [1] of the capacity of CDMA to that of conventional TDMA and FDMA for satellite applications suggested a reasonable edge in the capacity for the latter to more conventional techniques.

History was made in 1985, when Federal Communications Commission (FCC) allocated three frequency bands for commercial SS applications. The FCC rule changes in 1985, combined with the continuing evolution of digital technology catalysed the development of spread spectrum communication systems. Since then, SS has been successfully utilised in several commercial radio systems all the way from Terrestrial Mobile Networks to Satellite Systems.

In short, SS is a technique that takes a narrow band signal and spreads it over a broader portion of the radio frequency band.

Of the many potential uses for spread-spectrum communications in civilian applications, Code Division Multiple Access (CDMA) is the main one. This is especially true in what is arguably the hottest communication topic of the decade, the Third Generation Mobile Systems.

Code Division Multiple Access was originally proposed to support simultaneous digital communications among a large community of relatively uncoordinated users. In mobile environment, multipath is often the fundamental limitation to system performance, and CDMA is
a well-known technique for combating multipath. In addition to multiple accessing capability and multipath diversity, one of the major opportunities that arise when using spread spectrum techniques is the possibility of overlaying the low level direct sequence (DS) waveforms on top of the existing narrow-band users, and hence increasing the overall spectrum capacity even more so than just through the use of CDMA network.

The CDMA technique come into life only when various technologies such as optimal sequence generation, code acquisition, tracking, modulation, forward error-control coding, power control, channel estimation, RAKE combining, interference cancellation, soft handover, and variable rate vocoders fit together to push the technological performance boundaries beyond what can be achieved with conventional FDMA/TDMA systems. CDMA is conceptually more complex than Frequency Division Multiple Access (FDMA) and Time Division Multiple Access (TDMA), but not necessarily more difficult to implement considering the advances in microelectronics.

The presented work in proceeding chapters addresses the tight power control requirements of CDMA systems. But before going into detail, the following section provides an overview of the main motivation for this work.

1.1 A Brief Overview of UMTS

1.1.1 UMTS Terrestrial Radio Access Network (UTRAN)

The research findings and topics presented in the proceeding chapters have all been carried out with applicability to UMTS in mind. That is a matured and advance UMTS system beyond today’s phase-I (release ‘99) [1].

On that note, the following section provides a brief and step-by-step history on development of UMTS to date:

- In 1991 - 1995 two European Commission funded research projects called Code Division Testbed (CODIT) and Advanced Time Division Multiple Access (ATDMA) were carried out by the major European telecom manufacturers and network operators. The CODIT and ATDMA projects investigated the suitability of Wideband Code Division Multiple Access (W-CDMA) and Time Division Multiple Access (TDMA) based radio access technology for 3G systems. This work was later continued in the FRAMES (Future Radio Wideband Multiple Access System) [1] project and became the basis of the further ETSI UMTS work until decisions were taken in 1998.

- February 1992 World Radio Conference in Malaga (WRC-92) allocated frequencies for future UMTS use. Frequencies 1885 - 2025 and 2110 - 2200 MHz were identified for IMT2000 use.
• The 1994 “Green Paper on Mobile” identified the evolution towards personal communications.

• February 1995 The UMTS Task Force was established, "The Road to UMTS" report.

• December 1996 The UMTS Forum was established at the inaugural meeting, held in Zurich, Switzerland.

• June 1997 the UMTS Forum produced its first report entitled "A regulatory Framework for UMTS"

• October 1997 ERC decided on UMTS core band.

• January 1998 ETSI SMG meeting in Paris both W-CDMA and TD-CDMA proposals (generated by the FRAMES project) were combined to form the Terrestrial-UMTS air interface specification, or what is commonly referred to as UTRAN.

• June 1998 Terrestrial air interface proposals (UTRAN, WCDMA(s), CDMA2000(s), EDGE, EP-DECT, TD-SCDMA) were handed into ITU-R.

• September 1998 the first call using a Nokia WCDMA terminal in DoCoMo's trial network was completed at Nokia's R&D unit near Tokyo in Japan.

• December 4, 1998 ETSI SMG, T1P1, ARIB TTC, TTA created 3GPP in Copenhagen, Denmark.

• December 7 and 8, 1998 the first meetings of the 3GPP Technical Specification Groups in Sophia Antipolis, France.

• December 14 1998 The decision of the European Parliament and Council of Ministers requested that Member States take all necessary measures to allow the coordinated and progressive introduction of UMTS services by 1st January 2002 at the latest.

• February 1999 Nokia Oyj said that it has completed what it claims to be the first WCDMA call through the public switched telephone network in the world. The calls were made from Nokia's test network in Finland using a WCDMA terminal, WCDMA base station subsystem and Nokia GSM Mobile switching centres connected to the PSTN.

• March 16, 1999 Finland gave out the world's first 3G mobile technology licenses. Four licenses awarded to Sonera, Radiolinja, Telia and Suomen Kolmegee. Technically speaking though, some operators in USA and elsewhere already had the licenses and frequencies to operate third generation networks (technology independent).
• March 1999 ITU approved radio interfaces for third generation mobile systems in the meeting in Fortaleza, Brazil. Also Ericsson and Qualcomm agreed to share access to each other's technology ending a two-year patent dispute.

• April 27 and 28, 1999 Lucent Technologies, Ericsson and NEC announced that they have been chosen by NTT DoCoMo to supply W-CDMA equipment for NTT DoCoMo's next-generation wireless commercial network in Japan. This was the first announced WCDMA 3G infrastructure deal.

• 1999 World Radio Conference (WRC-99) handled spectrum and regulatory issues for advanced mobile communications applications in the context of IMT-2000. The aim was to identify additional frequency bands to satisfy market demand by 2005-2010. World wide roaming issues were also discussed.

• December 1999 in Nice ETSI Standardisation finished for UMTS Release 1999 specifications both for FDD and TDD


• July 2000 responsibility for maintenance and development of the GSM specifications was transferred from ETSI TC SMG to 3GPP to achieve harmony with UMTS.

• January 1, 2001 was originally planned for first commercial networks operational. No 3G networks were operating in January 1, 2001.

• March 2001 in Palm Springs 3GPP approves UMTS Release 4 specification (spec version 4.y.z). This is the version based on which all the first generation of UMTS infrastructure was developed.

• April 17, 2001 Ericsson and Vodafone UK claim to have made the world's first WCDMA voice call over commercial network.

• June 28, 2001 NTT DoCoMo launched a trial 3G service; an area-specific information service for i-mode. NTT DoCoMo has announced that it definitely plans to hit its October 1st target for a full commercial launch.

• September 25, 2001 NTT DoCoMo announced that three 3G-phone models are commercially available.

• October 1, 2001 NTT DoCoMo launched the first commercial WCDMA 3G mobile network.
• December 1, 2001 Telenor launched in Norway the first commercial UMTS network. UMTS terminals were expected to be available 3Q 2002.

• December 19, 2001 Nortel Networks and Vodafone in Spain (formerly Airtel Movil) completed first live international UMTS 3GPP standard roaming calls between Madrid (Vodafone network) and Tokyo (J-Phone network). Calls were made with a QUALCOMM MSM5200 chipset-based handset and J-Phone SIM technology.

• February 8, 2002 Nokia claims to have made the first end-to-end 3G WCDMA standard level 3GPP Release 99 June 2001 packet data calls between its commercial network infrastructure and terminals in its laboratories in Finland. The Nokia 3G WCDMA network and terminal used were based on the commercial version.

• February 18, 2002 Motorola unveils the company’s first GSM/GPRS and 3G/UMTS product, the A820. Motorola is "one of the first vendors to introduce a dual-mode enabled UMTS mobile phone".

• February 20, 2002 Nokia and Omnitel Vodafone claims to have made the first rich call in an end-to-end All-IP mobile network at the 3GSM World Congress in Cannes, France.

• March 2002 (Freeze date) UMTS Release 5 (the initial target date was December 2001).

• September 24, 2002 Ericsson announces the first live, dual mode WCDMA/GSM calls with seamless handover between the two modes and high data rate in live networks.

• September 25, 2002 Mobilkom Austria launches "Europe's First UMTS-Network" when Boris Nemsic CEO of Mobilkom "video-phoned" the Austrian politician Waltraud Klasnic. (see December 1, 2001 note above).

• September 26, 2002 Nokia introduces the "world's first handset (6650) for WCDMA (UMTS) and GSM networks".

• October 1, 2002 Qualcomm announces world's first Bluetooth WCDMA (UMTS) and GSM Voice Calls.

• October 3, 2002 Nokia and Vodafone Omnitel claims to "have carried out the world's first VoIP call completed in a 3GPP release 4 compliant network that transports circuit-switched voice and data calls through an IP backbone".

• October 10, 2002 Nortel Networks and Qualcomm claim to have "completed the industry's first UMTS voice and data calls demonstrating mobility across commercial cell
sites using live 1900 MHz radio spectrum, Qualcomm chipsets in commercial-form-factor handsets, and a live, end-to-end 3GPP UMTS network from Nortel Networks”.

- October 17, 2002 Nortel Networks claim to have "demonstrated the world's first UMTS calls using an IP-based UTRAN" using form factor handsets and an IP backbone based on Nortel Networks Optical Ethernet equipment. (Announced on October 22, 2002).

- November 29, 2002 Nokia and Vodafone Omnitel carry out 3G WCDMA call handover to commercial GSM network.

- June 2003 is a target date for UMTS Release 6.

- The decision of the European Parliament and Council of Ministers dated 14 December 1998 requires that Member States take all necessary measures to allow the coordinated and progressive introduction of UMTS services by 1st January 2002 at the latest. The EU's suggestion is that operators must cover 80% of the national population by the year 2005. This target is under serious question!

The future of UMTS to a large extend depends on availability of affordable handsets, a reliable network and above all, availability of multimedia services. For UMTS to be commercially successful, bandwidth efficiency and ability to handle large amounts of traffic through the network efficiently and cost effectively becomes paramount. Power control and inference management in CDMA systems are some of the most effective mechanisms for capacity improvement. Both of these topics are thoroughly treated in the proceeding sections.

1.1.2 Satellite UMTS (S-UMTS)

Whilst the second and third generation digital cellular networks can already cope with a large variety of requirements, the inherent bandwidth limitations make these networks less suitable for high-speed broadcast and push services.

The S-UMTS component will make the outdoor coverage globally seamless whilst maintaining terminal compatibility and service portability. More specifically, the S-UMTS component major objectives are:

- “Gap Filling” To enable global roaming for UMTS users, particularly in the early years of terrestrial UMTS roll out.

- To act as the broadcast component of UMTS allowing for efficient delivery of a range of multicast services to large geographical areas.
• To provide a quality of service commensurate to that of the terrestrial at an affordable cost.

• To provide rapid and cost effective deployment of UMTS services over large geographical regions and to augment the development of telecommunication services in developing countries.

On that note a multi-regional drive [2] to define and demonstrate S-UMTS has and continues to strive towards provision of a cost effective system that truly complements the terrestrial UMTS.

The key to success of any such system is the ability to deliver cost effective handsets. S-UMTS commonality in terms of frequency band and air interface technology can greatly reduce the terminal complexity, size and weight. The dual-mode (terrestrial and satellite) handset should eventually reach a cost comparable to the high-end T-UMTS mobile terminals if it is to be successful.

1.2 Aims and Objectives

The main aims and objectives of the presented research work were to develop new power control techniques that address a wide range link and system level issues. In doing so, several objectives were at the heart of the research:

• Introduction of speed intelligence as an adaptation mechanism.

• Operational environment awareness and adaptability.

• Adaptability to different operational environments and regional propagation environments.

• Optimal performance in both terrestrial and satellite operational environments.

• High performance, yet practical.

• Maximisation of system capacity without sacrificing performance.

• Practical power control algorithm that can adapt to a varying system load as well an inter-cell interference.
1.3 Original Achievements

Amongst the original achievements of the presented work, the following are the most significant:

**Shadow Correlation Distance in Satellite Communication Environment:** Based on statistical evaluation of actual channel recordings for L-band in two different operational environments, effective correlation distance of shadowing was established for the first time [17].

**Environment Detection Algorithm (EDA):** An intelligent algorithm that provides an essential component of any adaptive communication systems has been developed. The algorithm supports adaptive systems by accurately estimating the local operational environment and the speed of a user from a range of available information at the receiver. The algorithm has never been published since in preparation for a patent application.

**Speed Adapted Closed Loop Power Control:** The proposed algorithm effectively utilises the EDA to adjust the power control step size to the propagation environment and speed of the mobile. The algorithm’s superiority has been demonstrated and its performance thoroughly investigated. For the first time, performance of conventional closed power control algorithm in recorded propagation channels was also reported [29]. The work was further extended by adaptation of SA-CLPC to the satellite environment and subsequently enhanced by introduction a speed-adapted averaging scheme at the Fixed Earth Station.

**Near-Optimum Distributed Power Control Algorithm:** An advanced interference management algorithm in the form of distributed power control was developed and presented. The algorithm takes advantage of a set of information available in a single cell to dynamically adjust the target C/I for each individual user based on the system load. As shown in chapter 6 the proposed algorithm outperforms a range of similar algorithms as is significantly less complex.

1.4 Structure of the Thesis

Chapter 2 establishes some of the fundamental CDMA theorem and highlights the impact of power control on link level performance. The work proceeds to provide a comprehensive treatment of different power control schemes together with their corresponding application.

Chapter 3 describes the mobile propagation environment. In particular, this chapter focuses on second-order (time variant) characteristics of the propagation channel, an essential understanding for development of EDA and SA-CLPC.
Chapter 4 presents the all-important Environment Detection Algorithm. This algorithm forms the basis on which any adaptive communication system can “adapt”. The work successfully addresses one of the fundamental requirements of SA-CLPC algorithm of Chapter 5.

Chapter 5 establishes the performance of conventional fixed step CLPC. The concept of SA-CLPC is presented and extensively evaluated and optimised. The work covers SA-CLPC for both terrestrial and satellite communication systems.

Chapter 6 proceeds to address the system-wide interference management by means of distributed power control. An advance distributed power control algorithm is proposed and comprehensively evaluated against a number of competing schemes.

Finally, Chapter 7 summarises the major findings of each section, provides concluding remarks and suggestions for further work in each area.
Chapter Two

2. Power Control Principles

In TDMA-based systems, users within a cell transmit in different slots/frequencies and hence they do not interfere with each other. In such systems, power control is effectively used to minimise the co-channel interference and conserve the all-important mobile battery. Variations of the signal power due to the changing propagation conditions are taken care of by relatively large fade margins in the link budgets. Therefore, in TDMA-based systems the power control overheads can be kept to a minimum through the use of low speed power control loops that would adjust the power of each individual user for the reasons stated above.

In CDMA-based systems, however, all the users transmit on the same frequency using different spreading sequences thereby generating what is commonly referred to as the Multiple Access Interference (MAI). Since the uplink capacity of a CDMA system is interference limited, minimising the effects of MAI is of great importance. Power control is used as an effective mechanism for keeping the MAI under control and is therefore considered as the single most important requirement in CDMA-based mobile systems.

There are essentially two main types of interference generated in a mobile CDMA system:

1. Interference within a given cell, generated by multiple users transmitting on the same frequency.
2. Other cell interference, generated by users of the neighbouring cells also transmitting on the same frequency.

Several power control schemes are used to control the MAI. These can be broadly categorised as Uplink and Downlink power control schemes.

Chapter 2 establishes the CDMA principles and identifies the power control requirements in a mobile CDMA system. In doing so, the cause of power variations in a system followed by different mechanisms designed to overcome the problem are discussed.

### 2.1 CDMA Capacity

In CDMA systems, the Uplink and Downlink system capacity has to be treated separately as they essentially have different limitations.

The capacity of a cellular CDMA system is MAI limited in the uplink. In the downlink, however, capacity limitation is given by maximum BS/Node-B output power, i.e. the total power of all active channels (both traffic and control channels) has to be less or equal to the maximum transmitter (BS/Node-B) output power at any given instance. Depending on traffic asymmetry in a given cell both extremes could be experienced.

#### 2.1.1 Single-cell, single-service CDMA capacity

In this section, a single cell in which each CDMA user occupies the entire allocated spectrum, is considered.

In order to maximise the number of users in a cell it is important that all the Mobile Stations (MS') transmit at power levels such that the received power at the Base Station (BS) from each of the MS' is, to a good approximation, the same. This is clearly demonstrated in the proceeding sections of this chapter. However, for establishing the fundamental of the capacity evaluations, perfect uplink power control is assumed.

Each BS receiver at the centre of the cell receives and processes a composite waveform containing signals from $N$ users within the cell of interest. Assuming perfect power control, each user's signal will arrive at a power of $S$, and there could be up to $N-1$ same-cell interfering signals with a given user of interest. Thus the signal-to-noise (interference) ratio of the desired signal at reception is,

$$SNR = \frac{S}{(N-1)S} = \frac{1}{N-1} \quad \text{Equation 1}$$
The energy-to-noise density ratio \( \frac{E_b}{N_0} \) of the received signal can be obtained by simply dividing the numerator of Equation 1, by the information bit rate, \( R \), and dividing the noise (or interference) by the total bandwidth \( W \), resulting in,

\[
\frac{E_b}{N_0} = \frac{S/R}{(N-1)S/W} = \frac{W/R}{N-1}
\]

Equation 2

Equation 2, ignores the background noise, \( \eta \), due to thermal noise as well as other interference combined in the total spread bandwidth. This can be modified to bring this noise into consideration as follows,

\[
\frac{E_b}{N_0} = \frac{W/R}{(N-1)+\left(\frac{\eta}{S}\right)}
\]

Equation 3

Hence the capacity of the system in terms of the number of supported users, \( N \), is

\[
N = 1 + \frac{W/R}{\frac{E_b}{N_0} - \frac{\eta}{S}}
\]

Equation 4

where \( W/R \) is generally referred to as processing gain (denoted as \( G_p \)) and \( E_b/N_0 \) is the value required for adequate receiver performance, which for digital voice transmissions implies a BER of \( 10^{-3} \) or better. For data this could be as low as \( 10^{-7} \) to \( 10^{-11} \).

It is important to point out that in practice, the MS RAKE receiver can resolve not all multipath echoes of the received signal. This gives rise to a parameter commonly known as the orthogonality factor, \( \rho \). For the fully orthogonal case \( \rho=1 \) (best case) and the worst case is when \( \rho=0 \). By simple substitution of \( \rho \) in Equation 3, it is possible to derive a new relationship taking the orthogonality factor into consideration as shown below:

\[
\frac{E_b}{N_0} = \frac{W/R}{(N-\rho)+\left(\frac{\eta}{S}\right)}
\]

Equation 5

Therefore:

\[
N = \rho + \frac{W/R}{\frac{E_b}{N_0} - \frac{\eta}{S}}
\]

Equation 6

As it can be seen from the above any improvements in the required \( E_b/N_0 \) or any reduction in the interference from other users results in an increased capacity.

2.1.1.1 Sectorisation

Sectorisation is a commonly used technique, which employs usage of directional antennas at the cell-site (BS) for both receiving and transmitting. For example, with three antennas per cell-site,
each having an effective beam width of 120°, the interference sources seen by any antenna are approximately one-third of those seen by an omni directional antenna. This reduces the multiple access interference \((N-I)\) in the denominator of Equation 3, by a factor of three, which directly translates to approximately a three-fold increase in the capacity of Equation 4,

\[
N \approx \left[ \frac{G_p}{E_b/N_o} \right] \cdot \frac{\eta}{S} \cdot G_s
\]

Equation 7

where \(G_s\) denotes the sectorisation gain, i.e. the number of sectors.

However, this can only be assumed in theory. In practice it is not possible to have sectorisation without overlapping of different coverage sectors on the borders. It is also important to take into account the side lobes of the antennas when calculating the exact value of the sectorisation gain \(G_s\). In CDMA systems, this overlap could create additional interference. To bring this interference under control, in CDMA-based systems narrower beam width antennas (about 65°) are commonly used. Practical sectorisation gains of 2.4-2.6 are commonly achieved.

**2.1.1.2 Traffic Activity Factor**

Monitoring the traffic activity allows the signal transmission to be suppressed for a particular user when no voice/data bits are present, thereby reducing the average level of MAI in a given cell. Extensive studies show that in the case of a real time voice conversation either speaker is active only 35% to 45% of the time [3] (voice activity factor of 0.35 to 0.45). For other data applications such as email, web browsing, file transfer and video conferencing the activity factor would vary significantly, reaching highs of up to 100% in some cases.

The traffic activity factor is denoted by \(d\), and is inversely proportional to the capacity,

\[
N \approx \left[ \frac{G_p}{E_b/N_o} \right] \cdot \frac{\eta}{S} \cdot G_s \cdot \frac{1}{d}
\]

Equation 8

**2.1.2 Multiple-cell, Single-service CDMA Capacity**

In the previous section, derivation of the capacity formula was carried out ignoring the interference from the neighbouring cells. Since the same frequency band is reused in all the neighbouring cells, the total interference at the cell-site is comprised of interference from other MS's in the same cell plus interference from MS's in the neighbouring cells.

Let us assume a large number of equal sized cells with a uniform density of mobile stations. The path loss are considered to be function of the well known forth power law of distance (discussed in section Error! Reference source not found. of Chapter 3), if the system is assumed to exist in an area of uniform flat topography with relatively low antennas. This is however, fortunate since without the path loss, an unacceptable level of interference is received from far distant stations in
a very large system area. In the following analysis, perfect power control within the boundaries of a cell is assumed. This means that each cell controls the power of all the MS’ relative to its own centre and hence there is no overall control over the interfering power from the neighbouring cells.

The capacity formula for a multiple-cell CDMA system is essentially Equation 8, with additional interference from other cells. The other-cell interference is introduced by means of an additional parameter referred to as the frequency reuse efficiency factor $F$. The frequency reuse efficiency of omnidirectional cells is defined as the ratio of the interference from mobile stations within a cell to the total interference from all other system cells. The contribution of neighbouring cells is denoted by $k_1$ to $k_m$ as shown in Figure 1.

![Figure 1: Interference contribution from neighbouring cells](image)

Hence, the frequency reuse efficiency factor is,

$$F = \frac{\text{Interference from the mobile stations within a cell}}{\text{Total Interference from all the system cells}} = \frac{1}{1 + 6k_1 + 12k_2 + 18k_3 + \cdots}$$

Equation 9

where $k_1$ to $k_m$ represent the interference contribution from the corresponding neighbouring ring cells and the multiples, 6, 12, 18 ..., indicates the number of contributing cells.

Numerical integration techniques show [4] these contributions, for a system with the specified specifications, as indicated in Figure 1, result in a frequency reuse factor efficiency of,

$$F = \frac{1}{1 + (6 \times 0.6) + (12 \times 0.002) + \cdots} \approx 0.65 \text{ or } 65\%$$

Equation 10

Hence, the capacity of the multiple-cell CDMA system is determined by,

$$N \approx \left[ \frac{G_p}{E_s/N_o} - \eta \right] \cdot G_s \cdot \frac{1}{d} \cdot F$$

Equation 11
were, $G_p$ is the processing gain (typically between 4 to 512), $E_b$ is the bit energy, $N_0$ is the noise power spectral density, $\eta$ is thermal noise, $S$ is the signal power, $G_s$ is the sectorisation gain (typically with 3 sectors of 120°, $G_s=2.55$), $d$ is the activity duty cycle (typically 0.35 to 0.40 for voice), and $F$ represents the frequency reuse efficiency factor (typically 65%).

Equation 11, defines the upper capacity bound or "the pole capacity" since it simply assumes perfect power control is operational in the reverse-link (uplink) and does not take the power control imperfections into account. On the other hand, the above also ignores the possible improvements when forward-link (downlink) power control is employed in order to reduce the interference from the neighbouring cells.

Power Control Error, PCE (the difference between the target and the received $E_b/N_0$) has been shown to have a lognormal distribution. Knowing the mean and the standard deviation of this error, for a given probability, the required $E_b/N_0$ margin can be read from the PCE Cumulative Distribution Functions. To take the effect of imperfect power control into consideration in the capacity calculations above, this margin can simply be added to the required value of $E_b/N_0$ in Equation 11. As it can be seen, any increase in the required $E_b/N_0$ will result in a capacity reduction.

### 2.2 CDMA Performance Under Perfect Power Control

#### 2.2.1 CDMA Performance in AWGN Channel

The BER probability of an uncoded single user system using BPSK modulation is determined by:

$$P_b = \frac{1}{2} \text{erfc}\left(\sqrt{\frac{E_b}{N_0}}\right)$$

Equation 12

It is demonstrated [5] that in the case of a conventional CDMA detector and in the absence of coding, the variance of the white Gaussian noise source can be adjusted by the factor $F$, shown in Equation 13 in order to account for the $k-1$ interfering CDMA users for a given spreading factor of $G$.

$$F = 1 + \frac{2}{3} \cdot \frac{k-1}{G} \frac{E_b}{N_0}$$

Equation 13

The above assumes that all the users within the considered cell are all perfectly power controlled. Hence the probability of error in AWGN channel can be calculated as follows,

$$P_b = \frac{1}{2} \text{erfc}\left[\sqrt{\frac{E_b}{N_0}}\right] \left[\frac{E_b}{N_0}\right]$$

Equation 14
Chapter 2

Power Control Principles

Figure 2, and Figure 3, illustrate the analytical performance of a conventional single user CDMA detectors in an AWGN channel with an adjusted variance to take all the interfering users into account as shown in Equation 14.

![Figure 2](image1.png)

**Figure 2**: The analytical performance of a conventional CDMA detector in AWGN, G=128, [solid]: k=1, [*]: k=10, [+] : k=30, [o]: k=60

![Figure 3](image2.png)

**Figure 3**: The analytical performance of a conventional CDMA detector in AWGN, G=256, [solid]: k=1, [*]: k=10, [+] : k=30, [o]: k=60

### 2.2.2 CDMA Performance in Rayleigh Channel

It is also known that the probability of error of a single user QPSK-CDMA system in any channel is simply the probability of error of a QPSK in that channel since the spreading/dispersing process does not improve the modulation performance. As discussed in [6], this probability in a Rayleigh channel is,

\[
P_e = \frac{1}{2} \left[ 1 - \sqrt{\frac{\bar{\gamma}_b}{1 + \bar{\gamma}_b}} \right]
\]

**Equation 15**

where,

\[
\gamma_b = \alpha^2 \cdot \frac{E_b}{N_0}
\]

**Equation 16**

and \( \bar{\gamma}_b = E(\gamma_b) \), the Expected or the average value of \( E_b/N_0 \), defined mathematically as,

\[
\bar{\gamma}_b = \frac{E_b}{N_0} E(\alpha^2)
\]

**Equation 17**

and \( E(\alpha^2) \) the expected (average) fading amplitude \( \alpha \) (Rayleigh-distributed), and has a normalised average of unity in our case.
By substituting $\gamma_b$ into Equation 18, the single user performance can be written as,

$$P_e = \frac{1}{2} \left[ 1 - \sqrt{\frac{E_b}{N_o}} \right]$$

Equation 18

The performance of a conventional CDMA detector in the presence of MAI from multiple users in a Rayleigh channel can also be analytically driven from the performance of the single user system, Equation 15. It is shown in [6] that our treatment of modelling the multiple access interference as AWGN is also valid in a Rayleigh channel. Hence, the $\gamma_b$ can be modified to take the MAI into account as follows,

$$\gamma^k_b = \frac{\gamma_b}{1 + \frac{2}{3} \frac{k-1}{G} \gamma_b}$$

Equation 19

By replacing Equation 19, into Equation 15 the following is derived,

$$P_e = \frac{1}{2} \left[ 1 - \sqrt{\frac{\gamma^k_b}{1 + \gamma^k_b}} \right] = \frac{1}{2} \left[ 1 - \sqrt{\frac{E(\gamma^k_b)}{1 + E(\gamma^k_b)}} \right]$$

Equation 20

Figure 4 illustrates the analytical performance of QPSK-CDMA in Rayleigh channel as stated in Equation 20. The results were confirmed through simulation as shown in Figure 5.

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**Figure 4:** The analytical performance of a conventional CDMA detector in Rayleigh Channel, [solid]: single user bound, [---]: 5 users, [- - -]: 10 users, [- - -]: 20 users

**Figure 5:** The simulated performance of a conventional CDMA detector in Rayleigh Channel, [solid]: single user bound, [---]: 5 users, [- - -]: 10 users, [- - -]: 20 users
2.3 Why Power Control?

The capacity of a CDMA system is interference limited as demonstrated in section 2.1 of this chapter. There are several contributory elements that determine the overall system interference. Power control mechanisms can minimise system interference, however, as the total system interference is broken down to several different elements, each has to be treated separately and remedied with the appropriate power control mechanism.

2.3.1 Uplink Power Control

The use of asynchronous CDMA in the return/up-link gives rise to tight power control requirements as the spreading sequences of different user terminals lose their orthogonality with respect to each other to a large extent. This can be clearly seen in the example of Figure 6.

In the above example, two orthogonal spreading codes $C_1$ and $C_2$ assigned to two different users are considered. Figure 6-(a) shows that as long as the two codes remain synchronised, their cross correlation is kept to a minimum resulting in no interference between the two codes. Figure 6-(b), however, shows that if one of the codes is shifted in time by a single chip with respect to the other, the two codes would look identical (cross correlation of 4) and thereby producing maximum interference to each other. This can commonly be experienced in the uplink of a mobile CDMA system. For example assuming a typical chip rate of 3.84MC/s, to experience one chip difference between the received signals of two different uplink users, their path needs to be separated by a distance of 78m only. In practice, a shift equal to a third of a chip would remove the orthogonality of the two codes with respect to each other.

The level of this interference can, however, be minimised if all the signals arriving from different mobiles (within a single-cell), at the base station's receiver have the same fixed power level, regardless of the position of the mobiles and the characteristics of the channels in the cell. This effect in cellular communications is known as the Near-Far effect, where MS' closer to the BS (in
the absence of power control) will have a stronger signal arriving at the base station compared to those farther away, as shown in Figure 7.

Figure 7: The Near-Far effect

In order to assess the limitations of the synchronous and asynchronous CDMA reception under Near-Far conditions, the simulation model of Figure 8, was developed [7].

Figure 8: Evaluation model for assessing the performance of the conventional detector under Near-Far Interference

The model is based on a conventional single user detector (correlation receiver) and assumes a static AWGN channel, which although not of very practical interest, is extensively used as a baseline for comparative studies in this area. In this model, $b_1$, $b_2$ and $c_1$, $c_2$ are the BPSK modulated bits and the code sequences of the two users. $P_I$ and $P_J$ are the power levels of the user of interest and the interfering user respectively, and $n$ is the White Gaussian Noise. By increasing the power level of the interfering user near-far interference can be created. The simulations were carried out for two cases of Asynchrony-CDMA (A-CDMA) and Synchronous-CDMA (S-CDMA).

Figure 9, compares the performance results of a conventional A-CDMA and S-CDMA [8]. The use of Gold codes of length 31 has been assumed. The number of active users has been taken equal to 10. For the S-CDMA transmission, a maximum time synchronisation error of one third of a chip has been assumed.
As it can be observed under perfect power control conditions, performance of the S-CDMA system is very close to that of the single user bound. When the Average Interfering Energy Ratio (AIER), which corresponds to the ratio of the average energy per bit of the interfering users over the energy per bit of the user of interest, is equal to 6 dB (or 4 times), the $E_b/N_0$ degradation to the single user bound is kept below 1.6 dB for a BER of $10^{-3}$. On the other hand, assuming perfect power control, A-CDMA requires an 18 dB $E_b/N_0$ (some 11.2 dB degradation) to achieve a BER of $10^{-3}$.

The impact of Near-Far effect on link level performance can be minimised through the use of reverse-link (mobile-to-base station) power control. This proves to be the most difficult type of power control to achieve, since the mobile stations move independently within the boundaries of a cell, resulting on a series of received asynchronous signals, each experiencing independent fast fading.

### 2.3.2 Downlink Power Control

The BS generates and transmits signals to all users of the cell using the same clock and hence all the spreading sequences are synchronous at generation. Every MS in the cell receives a composite waveform that contains spreading codes of all users within that cell. Since all the codes from different users have travelled the same distance to get to the MS of interest, they remain

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1 For the purpose of this discussions in the chapter AIER and Power-Ratio are interchangeably used.
synchronous at the input to the mobile station receiver, regardless of the MS within a cell, as shown in Figure 10.

**Figure 10:** In the downlink, spreading codes from different users remain synchronous with each other at the input to MS receiver regardless of the travelled distance.

Furthermore, since the composite waveform containing signals from all other users within the system, has travelled through one propagation channel to get to the MS of interest, the relative power between different users remains the same. Therefore, at the input to the MS receiver no power imbalance between different users will be experienced and hence no power control mechanism between different users in the downlink is required.

As discussed in section 2.1.2, in a CDMA based cellular environment the users within a given cell also experience external interference from other cells as shown in Figure 11.

**Figure 11:** In the downlink, spreading codes from different users remain synchronous with each other at the input to MS receiver regardless of the travelled distance.

Downlink power control becomes necessary in a multiple-cell system in order to limit the interference to and from the neighbouring cells. The base station only transmits at the minimum power level required for satisfactory communication quality of each mobile. This effectively reduces the transmission coverage of each mobile to the minimum required, resulting on a reduction in the interference to the neighbouring cells. Downlink power control can be achieved through the use of both open and closed loop power control.
2.4 Power Control Parameters

2.4.1 Power Control Criteria

It is important to point out that in digital communication systems power control is actually based on the estimate of the received signal quality. That is commands and decisions to increase and decrease the transmitted power should be based on the signal quality and not just the power of the received signal. Quality can be measured through estimation of SIR or $E_{b}/N_{0}$\(^2\). However, different SIRs can correspond to the same Frame Error Rate (FER) in different radio environments, a function that maps the desired FER into the required SIR target is needed. This is performed by continuous measurement of the FER and SIR, and then adjusting the SIR targets to produce the desired FER. Accurate estimation of the $E_{b}/N_{0}$ or SIR at the receiver is of great importance and is therefore comprehensively treated in Chapter 5.

2.4.2 Power Control Dynamic Range

The dynamic range of power control describes the range within which the power of a single user can vary in a given cellular system.

The uplink dynamic range requirement is determined by the maximum possible residual distance between two mobile stations and the base station (i.e. one at centre and the other at cell edge):

$$\frac{P_{\text{rx}1}}{P_{\text{rx}2}} = \left(\frac{d_2}{d_1}\right)^a$$

Equation 21

where $P_{\text{rx}i}$ is the received signal power from user $i$, $d_i$ is the distance between the $i^{th}$ MS and the BS and the exponent $a$ is the path loss attenuation factor. In an extreme situation, if MS\(_1\) is 100m from the BS and MS\(_2\) is 2km away, and assuming a path loss attenuation factor of 3.5 (as in section 3.1.3), using Equation 21, the required dynamic range is about 50dB. Note that in realistic propagation conditions, one experiences fast fading as well as shadowing which have to be considered when determining the dynamic range of the power control. Considering these parameters could increase the required dynamic power control range to figures as high as 80dB. However, for the purpose of establishing the reasoning used to determine this parameter, the above argument is valid.

In the downlink, dynamics cannot be as high, since transmissions come from a single source (base station). Considering the nonlinearity limitations of the BS amplifiers as well as the requirements of the MS for increased power levels at the edge of the cell, dynamic power control ranges of up to 30dB are commonly used in the downlink.

\(^2\) In the definition of $E_{b}/N_{0}$ it is assumed that the $N_{0}$ represents both noise and the interference.
2.4.3 Power control step size

An absolute power setting at the transmitter would require extremely accurate, and thus expensive, power control circuitry beyond practicality. Power control adjustments are hence relative to the previous power settings with an increase or decrease equal to the power control step size. The step size is a parameter that may differ between different users of a cell. If the step size employed is too small then power control commands fail to compensate for the signal fades. On the other hand, a too large step size can cause distortions. Adaptive step size power control can be effectively used to optimise power control performance in different radio propagation channels as shown in the proceeding sections of the thesis.

2.4.4 Power Control Command Rate

Power control command rate is defined as the rate at which commands are transmitted. Power control command rate effectively determines how fast the control loop can operate. It is very difficult to track fast fading at high mobile speeds due to measurement delay, signalling of the power control commands and processing delay caused by extracting the power control at the receiver. In general, the optimal power control rate depends on the mobile speed and carrier frequency. Therefore, it would be beneficial to have a variable power control rate according to the mobile speed in order to minimise system overhead.

Power control command rates of 800-1500k command/s are commonly used in CDMA based cellular systems. It is important to point out that each command can be more than 1 bit, thereby increasing the power control symbol rate by a corresponding factor.

2.4.5 Power Control Headroom

It is well known that fast power control is effective at low mobile speeds [5], [7] [11]. At higher speeds interleaving becomes an effective tool for combating the effects of fast fading. As a consequence the required $E_b/N_0$ for slow moving mobiles experiences a significant improvement in presence of fast power control. In such cases where power control is really effective, improvement figures of up to 8.2dB reduction in the required $E_b/N_0$ have been reported by [9], [40]. These gains are gradually reduced for higher speed mobiles. Whilst such figures do not truly represent the actual gains in a realistic system, they do indicate the relative performance improvement for various speeds.

The improvements above are at the cost of additional transmitted power for combating fast fading conditions. When a user reaches the edge of the cell it may therefore hit the maximum allowed transmit power at certain points in time? Power control headroom is a margin taken into consideration during the link budget analysis that allows a few extra dBs to ensure slow moving mobiles at the edge of the cell do not reach their maximum transmit power, when they can still combat fast fading effectively. Values of 1.5-5dB are in common use [9].
In many instances, power control headroom is taken into consideration in conjunction with the required \( E_i/N_0 \) or the fading margin and may not appear as a separate figure in the link budgets.

### 2.4.6 Power Control Error

Power Control Error (PCE) is referred to as the difference between the actual and the desired power control criterion (PCC) as shown in Equation 22:

\[
\text{Power Control Error (dB)} = \text{Desired PCC (dB)} - \text{Actual PCC (dB)} \quad \text{Equation 22}
\]

where, PCC can be SIR, \( E_b/N_0 \) or the signal strength.

Research has shown [10] that the PCE (dB) has a normal distribution that can be characterised by its mean and standard deviation.

### 2.5 Power Control Mechanisms

In this section a brief review of various power control mechanisms applicable to uplink and downlink is given.

It is important to note that all power control algorithms between the base station (or Node-B) and the mobile station are referred to as the Inner Loop power control whilst the algorithm that sets the desired quality target levels for each user is referred to as the Outer Loop power control. Outer Loop power control algorithm resides in the Base Station Controller (BSC) or the Radio Network Controller (RNC).

#### 2.5.1 Inner Loop Power Control

##### 2.5.1.1 Open Loop Power Control

The Open Loop Power Control (OLPC) estimates and compensates the power variations due to shadowing and path loss (slow variations). OLPC can operate at both the BS and the MS (up and downlink), during the call set-up procedure as well as the entire duration of the call. In the call set-up procedure, the OLPC sets the initial transmitting power of the traffic channel as well as other channels such as the random access channel.

In OLPC, the receiver measures the average \( E_i/N_0 \) of the downlink pilot channel (continuously transmitted over the entire cell) and compares this measurement with the desired \( E_i/N_0 \). In some systems, the actual transmitted power of the pilot channel can be known to the mobile station through the broadcast channel, which carries cell related information including, the cell interference levels or the base station's transmitting power.
Figure 12, depicts a typical open loop power control [7].

![Diagram of Open Loop Power Control](image)

**Figure 12:** Open Loop Power Control mechanism employed by a MS

The major disadvantage of open loop power control is the assumption that the uplink and the downlink channel characteristics are the same. In FDD based systems this clearly is not the case in fast fading (multipath) environments. However, the channel characteristic in slow varying environments (shadowing and path loss) exhibit reciprocity in the up and downlink, making the OLPC the solution suitable for tracking the slow variations of power.

It is also important to note that in OLPC, the adjustment of the transmitted MS's power is entirely controlled by the mobile unit. This could be very costly in the case of a faulty mobile station whose transmitted power is at a level much higher than required, hence posing as interference to other CDMA users.

It is important to point out that as the OLPC operates on slow fading only, its performance would to a great extent depend on accurate separation of the slow and fast fading at reception. In practice this is subject to inaccuracies, which could cause major difficulties in the system performance.

### 2.5.1.2 Closed Loop Power Control

The received signal level is subject to rapid variations in a channel caused by multipath propagation. Assuming an FDD system whereby the uplink and the downlink carrier frequencies are not the same, these variations are uncorrelated and hence cannot be controlled by OLPC as described in the previous section.

Figure 13, depicts a typical Closed Loop Power Control (CLPC) scheme base on $E_b/N_0$ estimation technique designed to track such fast fading. Note that this particular closed loop is intended for uplink power control.
As shown above, the value of the $E_b/N_0$ is estimated at the BS receiver. Power control commands are then issued and transmitted back to the mobile station, after the estimated $E_b/N_0$ is compared with the required threshold. The mobile station receives the commands through a dedicated channel and acts upon them by varying its transmitting power.

As illustrated in Figure 14, in fast fading environments, an ideal estimation of the channel envelope variations at the BS receiver, and zero delay transmission of decisions to the MS, will result in complete cancellation of power variations.

However, Figure 14, is an academic example and perfect power equalisation can never be achieved due to the finite accuracy of the commands issued, and the loop delay constraint. The fact that a feedback loop exists in the system, imposes an inherent hysteresis phenomenon. Therefore, as common practice [11] power commands are transmitted unprotected, since the usual long delay due to coding/interleaving is inconsistent with the need for fast power control schemes.
That is, by the time the received power is measured (at BS), and the actual power control commands are received by the mobile station, the fading has changed and the command to increase or decrease the power may no longer be valid. In theory one can increase the power command rate to combat this problem. However, there are a few major problems in doing so,

a. The resources will be inefficiently used, i.e. high rate power control command channels are required.

b. The command rate is limited by the round trip delay, i.e. the command intervals must be large enough so that the effect of one command is measured before another one is sent. For example, in an environment where the round trip delay is about 100μs (cell size ≈ 15 km), maximum power control command rate of 10 kb/s can be achieved.

c. In conventional power control algorithm, the step size is fixed regardless of the depth of fade.

d. Power control dynamic range is limited.

e. The SIR or $E_b/N_0$ estimation is inherently inaccurate unless advanced and complex techniques are used.

f. Power control commands may be detected in error in the absence of coding protection.

Thus, it is very important for success of the power control technique to design closed loop power control systems with emphasis on the followings:

1. Hysteresis minimisation of the loop.

2. Employment of accurate techniques based on which power control commands are issued on the transmitter side and reliable algorithm for interpretation of the power control commands on the receiving side.

3. Adaptation of the step size to the experienced propagation conditions.

### 2.5.2 Outer Loop Power Control

As described earlier, the users within a given cell are power controlled to a desired SIR or CIR\(^3\). However as mobiles moves with different speeds in various propagation environments, they will require different values of SIR to achieve the same quality of service (BER/FER). A simple mechanism for ensuring that the mobiles are power controlled to the minimum required SIR is for an outer loop to dynamically change the target value of SIR for each user. The BSC (RNC) monitoring the ongoing FER can simply achieve this. If the achieved FER is above the required,

\[^3\text{The terms Signal to Interference Ratio (SIR) and Carrier to Interference Ratio (CIR) are used interchangeably throughout the various chapters.}\]
the SIR threshold of a particular user can be reduced and if the achieved FER is lower than the required, the SIR for that particular user will be increased.

This is a simple outer loop power control algorithm that only takes into consideration the requirements of the users within a given cell. Under this scheme decision on the actual SIR level for each user is taken only based on the QoS of the user of interest. This type of power control is generally referred to as “distributed system” and is the most practical of all. Distributed power control is fully discussed in Chapter 6.

Under realistic conditions, the level of interference experienced at the input to the BS receiver could vary due to a varying load within a given cell in addition to interference from the neighbouring cells as shown in Figure 15.

![Figure 15: Non-uniform user distribution](image)

Conditions in which mobiles within a cell are requested to increase their power to combat fading will certainly occur. This in turn could increase interference to other users within a cell as well as the neighbouring cells thereby reducing the experienced SIR for each user. The interfered mobile stations could be commanded to increase their transmitted power. An increased transmission power from other users will increase the interference to the user of interest. This effect is referred to as the “party effect” (where one would raise voice to overcome background noise) and could result in an unstable system in which the users in all cells are constantly increasing power to combat the neighbouring cell’s interference.

The outer loop algorithm resides at BSC/RNC level and has visibility to QoS and interference levels of different users of all the cells under its control (normally several 50-300 cells). An advance algorithm maximise system capacity if it sets the desired SIR in each cell by taking the users within a cluster of cells into consideration. Whilst such a “centralised” approach brings improved system performance, there are practicality issues.
Under a centralised scheme, outer loop power control can be used as an effective mechanism for managing interference at a system level as described above. Outer loop power control can be used in both up and downlink in conjunction with closed loop power control.
Chapter Three

3. Propagation Channel

Propagation channel plays an important role in the overall design of any mobile communication system. It directly impacts the design of many system aspects from the choice of access scheme, channel coding and modulation, all the way to higher layer signalling flow and protocols. It is therefore important to fully understand and base the design and optimisation of various system parameters on accurate propagation channel models, representing the intended communication environment realistically.

A more comprehensive treatment of the channel is required within the context of power control studies presented in the proceeding chapters. The validity of any power control algorithm and its performance can only be verified if correct propagation assumptions are made.

In this chapter both the narrowband terrestrial and satellite propagation channels are examined. Initially, some of the well-known mobile propagation theories are established. These theories are primarily established in the context of terrestrial cellular communication systems. However, in the proceeding sections of this chapter these concepts are applied to the mobile satellite communication environments. Through this chapter, particular attention is given to much less well-known second-order (time-variant) statistics of the channel. In cases where such channel characteristics were not available in the literature, fresh analysis of measurement data have been made as a result of which new models were proposed.
3.1 Mobile Cellular Propagation Channel

Power control techniques are designed to specifically combat amplitude variations of the received signal. It is therefore necessary to identify the causes of these variations in the mobile radio environment, namely, multipath propagation (Rayleigh), shadowing (Log-Normal), and the path loss. Each of the above propagation phenomenon can be treated separately in the propagation analysis. However, under realistic operational environments they are superimposed over each other. The overall propagation effect on the received signal power as a function of distance between the MS and the BS is shown in Figure 16.

![Figure 16: The overall impact of various propagation elements on the received signal strength](image)

For the purpose of power control analysis, the narrowband characteristics of the channel are mainly considered, as power control is performed over the received signal SIR or power only. However, in the proceeding sections of this chapter the wideband characteristics of the channel are also reviewed for the environment detection algorithm. Figure 17 shows the end-to-end narrowband propagation chain.

![Figure 17: The overall cellular narrowband propagation chain](image)

The three identified fading elements are multiplicative in nature, however, noise and interference are additive features due to the fact that they contain energy.
3.1.1 Multipath Propagation

In environments where the antenna height of the MS is lower than its typical surroundings, and the carrier frequency wavelength is much less than the dimensions of the surrounding structures, multipath waves are generated. These conditions best describe the terrestrial cellular environments. Nevertheless, links between a MS and a low elevation angle satellite are very similar to that of the terrestrial and would come under the same category. At higher elevation angles, however, the presence of the dominant Line-of-Sight (LOS) signal reduces these fluctuations, resulting in a less hostile environment.

3.1.1.1 The Rayleigh Channel

In the absence of the LOS, the signal power received from a moving MS fluctuates rapidly (according to the speed of the mobile and the carrier frequency) within a range of 40dB (10dB above and 30dB below the average level) [12], as the MS travels through a formation of standing waves in its path. Figure 18, shows a typical profile of the received fading envelope.

![Typical Rayleigh Amplitude Envelope](image)

Figure 18: The Rayleigh envelope of the received signal for a mobile with the specified speed and carrier frequency

If each multipath component in the received signal is independent then the PDF (probability density function) of its envelope is Rayleigh [13]. The Rayleigh amplitude envelope can be mathematically expressed as,

\[ r(t) = \sqrt{x^2(t) + y^2(t)} \]  

Equation 23

where \( x(t) \) and \( y(t) \) are independent Gaussian random variables.
The movement of the MS, also introduces Doppler frequency, which translates into phase changes of the received signal at the base station. However, it is important to point out that the phase changes introduced have a uniform PDF, which can be expressed as follows,

$$\psi(t) = \tan^{-1}\left(\frac{y(t)}{x(t)}\right)$$  \hspace{1cm} \text{Equation 24}

The well-known PDF [13] and [15] of the received signal amplitude $r(t)$ is given by Equation 25, below,

$$p(r) = \frac{r}{\sigma^2}e^{-r^2/(2\sigma^2)}$$  \hspace{1cm} \text{Equation 25}

where $r^2/2$ is the short-term signal power, and $\sigma$ is the mean power. Figure 19 depicts this PDF, highlighting some important statistical relationships of the distribution.

Further first-order statistical (time-invariant) characteristics of a Rayleigh channel are extensively discussed in [14]. However, for the purpose of power control analysis, in particular the closed loop power control, the second-order (time-variant) statistics of the fast fading characterised by its autocorrelation or the PSD function, are required.

It is important to point out that the Rayleigh channel is the most hostile propagation channel. In the terrestrial environment, the main communication medium is the multipath as LOS is generally not available in the built up areas. The system is hence designed to operate under such conditions through the use of a combination of power control, interleaving and coding techniques. However, in the satellite environment, due to the long round-trip delays associated with such channels, and very limited satellite power budgets, it is extremely difficult to maintain acceptable quality of service in such channels. Dynamic satellite constellations are hence used in order to ensure
satellite visibility at generally high elevation angles, leading to presence of a dominant LOS signal.

### 3.1.1.2 The Ricean Channel

When the LOS signal is present in addition to the multipath signals reflected from the surrounding scatterers, the distribution of the received signal amplitude is shown to be Ricean. The ratio between the average power in the LOS to the average power in the scattered component is referred to as the Rice factor,

\[
k = \frac{P_{\text{LOS}}}{P_{\text{Multpath}}}
\]

**Equation 26**

### 3.1.1.3 The Fading Rate

Fading rate is referred to as the rate by which the received signal experiences deep fades [15]. In a mobile environment deep fades are experienced at about every half wavelength \((0.5\lambda)\), as the MS moves through standing wave formations caused by the surrounding structures. Hence this rate can be easily calculated as follows,

\[
\text{Fading Rate} = \frac{v_m}{\lambda/2}
\]

**Equation 27**

where \(v_m\) (m/s) is the speed of the mobile (MS) and \(\lambda\) (m) is the wavelength of the carrier signal. As the speed of the mobile increases or the frequency of the carrier increases, the fading occurs more frequently. In some worst-case scenarios, this rate could be as high as 400 fades/sec \((f_c=2\text{GHz}, v_m=100\text{km/h})\), or in other words, fades occurring every 2.5ms.

### 3.1.1.4 Level Crossing Rate

The average level crossing rate (LCR), \(\bar{n}(R)\), is referred to [12] and [15] the average rate at which the positive slope signals cross a given threshold \(R\), and is expressed as,

\[
\bar{n}(R) = n_o \cdot n_R
\]

**Equation 28**

where \(n_o\) is the normalisation factor, calculated as follows,

\[
n_o = \sqrt{2\pi} \frac{v_m}{\lambda}
\]

**Equation 29**
where \( v_m \) (m/s) is the speed of the MS, \( \lambda \) (m) is the wavelength, and \( n_R \) in Equation 28 is the normalised level crossing rate, independent of wavelength and the speed of the MS.

The value of \( n_r \) for a Rayleigh channel can be deduced from,

\[
R \cdot e^{-R^2} = R' Z
\]

Equation 30

where \( R \) is the envelope of the E-Filed, with respect to its \( rms \) value. As \( n_R \) is the normalised level crossing rate, it is easier to deduce it from the plotted graph of Equation 30, as shown in Figure 20.

It can be observed that the maximum level crossing rate occurs at 3dB below the average signal level. This maximum can also be obtained by,

\[
\frac{d(n_R)}{dR} = e^{-R^2} \left( 1 - 2R^2 \right) = 0.0
\]

Equation 31

\[ \Rightarrow R = 0.5 \text{ or } -3dB \]

For example, the expected value of the level crossing rate at 10 dB below its average power level from a mobile travelling at a speed of 50 km/h (13.89 m/s), and transmitting on a frequency of 1.5 GHz (\( \lambda = 0.2 \) m) can be calculated as follows.

In order to calculate \( \bar{n}(R) \) from Equation 28 \( n_o \) needs to be calculated from Equation 29:
and the value of the \( n_R \) for a level of -10 dB is deduced from the graph in Figure 20, to be 0.284. Hence,

\[
\bar{n}(R) = n_o \cdot n_R = 147 \times 0.284 \approx 42 \text{ crossings per second}
\]

Equation 33

This value gives an indication of the direction changes of the power control commands per second, when designing closed loop power control for fast fading environments.

### 3.1.1.5 Average Fade Duration

The average duration of the fade [15], \( \bar{t}(R) \), indicates the average duration of each fade below the given level \( R \), and can be calculated as follows,

\[
\bar{t}(R) = \frac{CDF}{LCR} = \frac{P(r \leq R)}{n(R)} = \left( \frac{1}{n_o} \right) \cdot \left( \frac{P(r \leq R)}{n_R} \right) = t_o \cdot t_R
\]

Equation 34

where \( t_o \) (the normalisation factor) and \( t_R \) (the normalised AFD) are defined as,

\[
t_o = \frac{1}{n_o} = \frac{\lambda}{v_m \sqrt{2\pi}}
\]

Equation 35

where \( v_m \) (m/s) is the speed of the mobile, \( \lambda \) (m) is the wavelength, and \( t_R \) in Equation 10 is the normalised fade duration, independent of wavelength and the speed of the mobile and is,

\[
t_R = \frac{P(r \leq R)}{n_R} = 1 - e^{-R^2} \approx \frac{1}{R \cdot e^{-R^2}} = \frac{1}{R \left(e^{R^2} - 1\right)}
\]

Equation 36

Figure 21, shows the plot of the above normalised fade duration, \( t_R \), for different power levels.
The average fade duration is by far the most important statistic of channel when designing the closed loop power control.

### 3.1.2 Shadowing (Log-Normal)

The Log-Normal fading is referred to the long term fading which is caused due to the terrain counter, surrounding structures, etc., between the base station and the mobile station.

It is called Lognormal since, experimentally one typically finds that data collected around a circular route (i.e., a route for which path length is kept constant) yield a received power distribution in decibels (hence the term \( \log \)) that plots linearly (hence the term normal) on normal
probability paper [16]. In other words the logarithm of the received signal amplitude will have a normal distribution.

Lognormal shadowing is characterised by its first and second order statistics. The mean of the lognormal shadowing effectively corresponds to the path loss at any given location of the cell and its standard deviation is an environmental dependent parameter. Second order statistic of the shadow fading determines the time-varying characteristics of the channel. As shown in Figure 22, the time varying nature of the shadowing is very much dependent on the geometry of the cell.

For the purpose of power control, correlation distance statistics are focused upon. As far as the simulation of power control is concerned, correlation distance of shadowing is one of the most important statistics required for both the mathematical and the simulation based analysis of these techniques. Correlation distance of the shadowing simply indicates likelihood of two samples separated by the measurement interval. Knowing the correlation distance between two samples, the sampling rate, the standard deviation of the shadow component around its mean, and the mean of the shadowed signal, a simple shadow-fading generator shown in Figure 23 can be developed [17].

![Shadow fading generator](image)

*Figure 23: Shadow fading generator*

Both the first and second order characteristics of the Lognormal shadowing for terrestrial systems have been widely reported [12] and [15].

### 3.1.3 Path loss

In free space, the causes of propagation path loss are merely frequency $f$, and distance $d$, as shown in equation (28).

$$\frac{P_r}{P_t} = \left(\frac{1}{4\pi df/c}\right)^2 = \frac{1}{[4\pi(d/\lambda)]^2}$$  \hspace{1cm} \text{Equation 37}

In practice however, the above formula will not apply directly as the path loss characteristics vary according to the environment. In fact it not only depends on the different terrain environment but also in different build structures and antenna heights.
The general empirical model for path loss is modelled in [15] as follows,

$$ P_r = P_{ro} \left( \frac{r}{r_0} \right)^{-\gamma} \cdot \left( \frac{f}{f_o} \right) \cdot \alpha_0 $$  \hspace{1cm} \text{Linear expression} \hspace{8cm} \text{Equation 38}

$$ P_r = P_{ro} - \gamma \log \left( \frac{r}{r_0} \right) - n \log \left( \frac{f}{f_o} \right) + \alpha_0 $$  \hspace{1cm} \text{dB expression} \hspace{8cm} \text{Equation 39}

where \( r \) is the distance in miles or kilometres and \( r_0 \) equals to 1 mile or 1.6 km. \( \gamma \) can be expressed as \( \gamma \)th power in linear expression and a \( \gamma \) dB/dec in dB expression. \( P_{ro} \) is the power at the 1-mile point of interception. The reason for taking 1-mile point of interception is that within a 1-mile radius very few streets are available. Therefore, data with limited runs that do not provide the statistical mean are avoided. \( f_0 \) is 900 MHz and \( f \) represents the operating frequency as long as it is above 30 MHz.

The value of \( n \) in Equation 39 is found from empirical data. Okumura [18] indicates that \( n=30 \) dB/dec, and Young [19] indicates that \( n=20 \) dB/dec. Hence, it can be assumed that,

$$ 20 \text{ dB/dec} < n < 30 \text{ dB/dec} $$

Note that the above is only true for frequency ranges from 30 to 2000 MHz and the distance range from 2 to 30 km.

\( \alpha_0 \) in Equation 39 is the adjustment factor and is calculated as follows,

$$ \alpha_0 = 20 \log \left( \frac{h_1}{30.48 m} \right) + 10 \log \left( \frac{P_t}{10 w} \right) + (g_1 - 4) + g_2 + 10 \log \left( \frac{h_2}{3.048 m} \right) $$  \hspace{1cm} \text{Equation 40}

$$ = 20 \log h_1 + 10 \log P_t + g_1 + g_2 + 10 \log h_2 - 43.36 $$

where \( P_t \) is the transmitted power in watts, \( h_1 \) and \( h_2 \) are the base antenna height and the mobile antenna height in meters respectively, and \( g_1 \) & \( g_2 \) are the base station and mobile station's antenna gain in dB.

In order to be able to calculate the path loss\(^4\) in different terrain environments from Equation 39, the parameters \( P_{ro} \) and \( \gamma \) are required. Field trials of [15] show that these values change according to different environments as shown in Table 1.

\(^4\) Equation 39, determines the received power. The transmitted power is used in calculation of the adjustment factor \( \alpha_0 \), therefore path loss can be directly calculated from Equation 39.
<table>
<thead>
<tr>
<th></th>
<th>$\gamma$ (linear)</th>
<th>$\gamma$ (dB/dec)</th>
<th>$P_m$ (m watts)</th>
<th>$P_m$ (dBm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Free space</td>
<td>2</td>
<td>20</td>
<td>$10^{-4.5}$</td>
<td>-45</td>
</tr>
<tr>
<td>Open area</td>
<td>4.35</td>
<td>43.5</td>
<td>$10^{-4.9}$</td>
<td>-49</td>
</tr>
<tr>
<td>Suburban area</td>
<td>3.84</td>
<td>38.4</td>
<td>$10^{-6.17}$</td>
<td>-61.7</td>
</tr>
<tr>
<td>Urban area (Philadelphia)</td>
<td>3.68</td>
<td>36.8</td>
<td>$10^{-7}$</td>
<td>-70</td>
</tr>
<tr>
<td>Urban area (Newark)</td>
<td>4.31</td>
<td>43.1</td>
<td>$10^{-6.4}$</td>
<td>-64</td>
</tr>
<tr>
<td>Urban area (Tokyo)</td>
<td>3.05</td>
<td>30.5</td>
<td>$10^{-8.4}$</td>
<td>-84</td>
</tr>
</tbody>
</table>

*Table 1: Measured $\gamma$ and $P_m$ for different terrestrial environments.*

From the above it can be seen that the value of $\gamma$ in a terrestrial environment varies between 2 and 4. Hence, the received power for each of the above environments (only considering path loss) can be calculated from Equation 39, e.g. for the suburban area,

$$P_r = -61.7 - 38.4 \log r - n \log \left( \frac{f}{900} \right) + \alpha_0 \text{ dBm (suburban)}$$

**Equation 41**

### 3.2 Land Mobile Satellite Propagation Channel

Due to an ever-increasing interest in the provision of Satellite Personal Communication Networks (S-PCN), channel models become an essential requirement for system design and dimensioning. Over the past few years there has been a great effort in characterisation of the Land Mobile Satellite Propagation Channel (LMSPC), [20], [21], [22]. Researchers have resorted to airborne measurement campaigns using aircraft and helicopter platforms to emulate a moving satellite. Such experiments are very costly and time consuming. Consequently, unlike the terrestrial cellular case, a very limited number of such campaigns have been carried out.

There are a few fundamental differences between the terrestrial and the LMSPC, the most important of which is an ever-changing propagation characteristic due to the dynamic nature of the considered Non-Geostationary constellations (N-GEO) as well as the long round-trip delays associated with such systems.

Furthermore, unlike the terrestrial propagation environment, wideband channel measurements show delay spreads of 100ns or less in satellite environments. This means that in CDMA system to take advantage of multipath diversity, a spreading bandwidth of 10MHz or above is required. Therefore, narrowband representation of the channel will be sufficient for all bandwidths of interest, including S-UMTS.
In this section, a brief overview of LMSPC, essential for analysis and simulations of proceeding chapters is presented. In the subsequent sections of this chapter the multi-state channel (Markov model) representation of the LMSPC is then reviewed followed by a detailed examination of the narrowband characteristics of the channel at L and S-band.

### 3.2.1 The Two State Markov Model

When a mobile user within a satellite communication system moves through an operational environment, the signal received by the user is blocked time to time due to LOS obstructions around the user. Furthermore, in non-geostationary satellite constellations the relative position of the satellite also changes causing similar shadowing impairments. This leads to a large drop in the signal strength commonly referred to as shadowing. Signal variation during shadowing is modelled using Loo’s model [23]. In the proceeding sections of this paper, the shadowed and the non-shadowed cases are referred to as bad and good states, respectively. The bad and good duration are measured in terms of distance for a required environment and the measured values are incorporated within a two state Markov model for a single user-satellite link. Within any given states, there are signal strength variation due to the multipath effect. *Figure 24*, shows the narrowband representation of a typical received signal, highlighting various fading elements and parameters [24].

The narrowband propagation parameters such as the good and bad state statistics for various operational environments have been widely reported on from experimental databases of University of Surrey [22] together with the DLR [20].
Chapter 3

Propagation Channel

It is important to point out that when in a complete block, the satellite channel could have a Rayleigh distribution, but due to the limited link margin in satellite environments, the average drop in the signal strength often means complete loss of communication.

3.2.2 Ricean Channel

In presence of line-of-site (LOS), the received signal is composed of two components; the multipath and the LOS. Narrowband satellite propagation channels are mainly represented by a Rice distribution together with a shadow component. The severity of a Ricean channel is determined by the ration of signal energy in the LOS to the energy in the multipath component. This factor could range from 0 to \( \infty \) representing a Rayleigh to an AWGN channel respectively.

The first order statistics of a Rice distribution can be found in [14], however, the second order statistics of a Ricean channel are harder to come by.

For the Rician channel the level crossing rate, \( N(R) \) at a specific level \( R \), is given by the expression.

\[
N(R) = \sqrt{\frac{\beta}{2\pi}} \rho_\xi(r)
\]

Equation 42

where \( \rho_\xi(r) \) is the Rician pdf and \( \beta = 2\sigma^2 (\frac{\nu f_d}{\Delta f_{\text{max}}})^2 \).

This can be used in similar fashion to the Rayleigh LCR in Chapter 4 for speed estimation.

3.2.3 Shadowing (log-Normal)

As stated before for the purpose of power control analysis, particular attention is given to the much less well-known second-order (time-variant) statistics of the channel. One such parameter is the correlation distance of shadow fading, an essential parameter for realistic evaluation of any power control scheme. Through the analysis of channel recording data, a very simple yet accurate shadow fading correlation model is proposed here [17]. These statistics have not yet been available for satellite channels, restricting research to only terrestrial values [25] reported by [26].

The proposed shadow fading correlation model is based on the narrowband measurement campaign recordings at L-band (1.3 GHz) and S-band (2.3 GHz), carried out by CCSR\(^5\) in spring'92 [27]. The measurements were carried out using a helicopter emulating a satellite moving in parallel with the mobile unit along a straight road. Different helicopter heights and

\(^5\) Centre for Communication Systems Research (CCSR), University of Surrey
distances from the road would therefore emulate different elevation angles. Specifically suburban and heavily wooded categories are paid particular attention as the urban environment is not the main S-PCN operational environment, and the open category is not a hostile environment.

### 3.2.3.1 Data Processing

The channel recordings were low-pass filtered in order to average out the multipath variations and hence separate the shadow fading. The low-pass filter was hence designed to be sufficiently narrow in order to average the fast variations, yet wide enough to pass the slower fading [26]. Such filters can be implemented using an integrate and dump, averaging the signal over a range of 20-80λ (depending on the environment) as suggested in [15] and [26]. For the purpose of our analysis, in both the heavily wooded and the suburban environments, a length of 20λ (= 4.5m at L-band and 2.5m at S-band) was proved sufficient as the multipath component in such environments with relatively high elevation angles are not the dominant element of the received signal. The distribution of the filtered received signal strength (slow variations) was found to approximate to lognormal. That is, the signal strength in dB is normally (Gaussian) distributed.

Figure 25 depicts the CDF of the received signal strength in dB, together with the Gaussian fit.

![Figure 25: Cumulative distribution function of the shadow fading in wooded environment, L-band, 60°](image)

### 3.2.3.2 The Proposed Correlation Model

In order to determine the effective correlation distance and in coherence with the approaches of [17] and [26], a simple exponential auto covariance fit is proposed,

\[
C_r(r) = E\{\xi(t)\xi(t+r)} - E^2\{\xi(t)} = \sigma_e^2 e^{-\frac{\pi r}{\lambda e}}
\]

Equation 43
where $X_c$ is the effective correlation distance of the shadow fading, $v$ is the velocity and $\tau$ is the measurement interval. A more appropriate notation of the above expression independent of the velocity can be expressed as,

$$C_c(\tau) = \sigma^2 e^{-|d_0-d_1|/X_c}$$  \hspace{1cm} \text{Equation 44}$$

where $|d_0- d_1|$ represents the measurement distance. The normalised correlation distance in the following diagrams is defined as the distance at which the correlation falls to $e^{-1}$. Also note that the results presented here are the normalised auto covariance, so that the co variances at zero lag are identically 1.

Figure 26 and Figure 27 show the normalised auto covariance of the shadow fading in the wooded environment. The route consisted of a 7 km long stretch along a small typical country road of 4 meters width. The route was considered to have high tree density of tall nature and predominantly non-deciduous coniferous trees much taller than those in the suburban environment.

![Figure 26: Normalised auto covariance in the wooded environment, L-band, $60^\circ$, [-] measurements and [-] exponential fit](image)
From the above it can be seen that the simple correlation model operates satisfactorily up to a certain distance. The proposed exponential fit is only applicable for maximum sample distances of up to 20m in Figure 26 and 14m in Figure 27. Due to the nature of the tree-shadowed environment, generally short correlation distances of 20m and 9m were measured respectively. It can also be observed that at 80°, the correlation distance of the shadowing is shorter than the 60° case as the obstructing trees are less dense at the top. It is important to point out that the type of structures and obstructions heavily influences the correlation behaviour of the shadowing.

Figure 28 and Figure 29 show the normalised auto covariance of the shadow fading in the suburban environment. The suburban route considered can be described as a typical English town suburb. The roadside trees were primarily of tall and deciduous type. The residential buildings along the road were generally two stories high. The distance of trees and buildings along the same section of the road varied from the roadside edge from only a few meters to some tens of meters. However, on average it was estimated to be between 5-10 meters.
From the above, longer correlation distances compared to that of the wooded environment can be seen.

In general, short correlation distances of the order of tens of meters were obtained for elevation angles of $60^\circ-80^\circ$, for both the environments. The results were quite expected as at such high elevation angles, any obstruction of the line-of-site will be momentarily, be it treetops or building roofs. However, preliminary analysis of the lower elevation angles indicate longer correlation
distances. In low elevation angles the situation is expected to get closer to the terrestrial environment as the signal travels through more obstructions of random dimension.
Chapter Four

4. Environment Detection

A mobile user by definition would be associated with a changing operational and propagation environment. The propagation environment plays an important role in the overall design of any mobile communication system. It directly impacts many low layer system aspects from the choice of access scheme, channel coding and modulation, all the way to higher layer signalling and protocol requirements. Traditional receiver designs are based on a single or average case of propagation environment, resulting in sub-optimum performance when user is in a different environment to the system design assumptions. It is therefore important to optimise various system parameters for a realistic range of expected propagation environments and adapt the receiver parameters to use most appropriate set.

The novel environment detection algorithm proposed in this chapter identifies the operational propagation environment and allows adaptation of various algorithms such as synchronisation, power control, voice codec rate, channel coding and modulation. The proposed algorithm is a fundamental requirement for any practical adaptive communication system.

In the proceeding sections of this chapter, the structure of the proposed algorithm together with all its associated sub-algorithms are presented and evaluated. It is important to point out that the proposed algorithm is primarily evaluated within the UTRAN framework, however, it can be applied to many mobile communication systems.
4.1 Environment Detection Algorithm Overview

In practice, the received signal in a mobile communication system is subject to degradation caused by the propagation channel (multipath fading, delay spread, shadowing and path loss). Severity and characteristics of the above propagation effects significantly vary under different operational environments. Various transceiver algorithms therefore can only be optimised for a particular operational environment at the design stage. Since in mobile communication systems the user terminal is required to operate in a range of environments, the conventional approach results in a sub-optimal performance in environments for which the algorithms have not been optimised. Knowledge of the propagation environment provides a valuable source of information for use in adaptation of various system parameters in particular the air-interface parameters.

The proposed Environment Detection Algorithm (EDA) blindly estimates the operational environment from a range of information available to the receiver. The operational environments can be categorised into several groups based on the differences in the propagation characteristics. These characteristics can be introduced as a series of look up tables specific to the region/country in which the system is deployed.

The environment detection algorithm consists of three main elements shown below:

![Diagram of the three main elements of the environment detection algorithm](image)

**Figure 30: The three main elements of the environment detection algorithm**

**Speed Estimation:** An essential part of the algorithm providing valuable information on user mobility at any point in time. The algorithm extracts this information from analysis of the received signal strength that has been subject to fast fading.
Shadow Correlation Estimation: Knowing the effective shadowing correlation distance in a given operation environment provides indications as to the type of operational environment. Each environment category (e.g. Urban, Suburban, etc.) is associated with a unique shadow correlation figure.

Delay Spread Estimation: Similar to the above case, delay spread provides valuable information on the type of operational environment. This complements the information provided by the shadow correlation estimation algorithm to provide an effective estimation of the operational environment.

4.2 The EDA Architecture

Having discussed various algorithms that constitute the EDA, this section focuses on the overall EDA architecture. The proposed algorithm (Figure 31) consists of a series of subsections closely interacting with each other.

![Diagram of the environment detection algorithm](image)

*Figure 31: The environment detection algorithm*

Input to this algorithm is the detected received signal (per RAKE Arm) and outputs of the algorithm are two $n$ bit binary numbers representing the operation environment and mobile speed. The length of $n$ (number of bits, e.g. 000 or 0000) is dependent on system requirements and the number of distinctly different environment categories (e.g. 8 or 16 environmental categories). The algorithm is subdivided into two main sections,

- Multiple RAKE Arm Units
- The Central Control Unit
4.2.1 The RAKE Arm Unit

EDA can work on single or multiple RAKE Arms as shown above. Performance of the algorithm would certainly be improved by taking additional diversity paths into consideration.

Various subsections of each algorithm that are discussed in the proceeding sections are all based on a single arm. Availability of sufficient signal processing and power resources would effectively determine how many such units could be utilised in a handheld unit. These restrictions are less of a problem at BS.

The operation of the algorithm is as follows,

1- Through an adaptive filtering stage the slow (shadowing) and fast (multipath) variations are separated. In order to do this optimally, experimental results indicate requirements for an averaging period of about 20-40\(\lambda\). Therefore the exact period of time over which this averaging would have to take place would depend on the speed of the terminal. This is why a feedback loop from the speed estimation algorithm is envisaged in the above diagram. At initiation of the algorithm an assumed speed will be used. The combined performance of the adaptive filtering and speed estimation would increase with time, as the algorithm converges to the actual speed. Subsequent speed changes will be closely tracked by the loop.

2- The speed estimation algorithm periodically provides an estimation of the user speed based on a number of criterions the most important of which is the level crossing rate (LCR).

3- Through the use of Channel Estimation algorithm (CHEST) already available within CDMA based receivers, the experienced delay spread can be measured over time. Delay spreads for a wide range of typical operational environments are known through experimental measurement campaigns. Look up tables can be constructed based on which estimations of the operational environments would be made.

4- The correlation of the shadow fading contains a large amount of information about the physical dimensions and density of the obstructing structure. Using experimental data, look up tables containing effective correlation distance of different operational environments can be constructed. From the measured correlation of the shadowed signal, and the look up tables, it will be possible to have another estimation of the operational environment in addition to the delay spread information.

4.2.2 The Central Control Unit

The rate at which each one of the above algorithms converges could be different. Therefore in real time operations, the output of each algorithm will be associated with a different accuracy depending on measurement window size. The reliability factors of each parameter is passed to a
central controller block which effectively adjust the weighting factor for each estimate to
minimise the measurement errors before combining and decision making process.

As shown in Figure 31, in case of multiple RAKE Arm Units, reliability information on each path
can be obtained from the channel estimation unit to adjust the relative weighting factors of each
diversity path accordingly.

In the following sections, each algorithm is presented and evaluated individually.

4.3 Speed Estimation

This algorithm estimates the mobile speed by using a level crossing rate (LCR) detector. The
input to the algorithm is the received signal strength after disspreading in a single RAKE arm and
the output is simply the estimated mobile speed. It is important to point out that the effectiveness
of this algorithm is significantly reduced if the input to the speed estimation unit is the received
signal after RAKE combining. This is due to the fact that diversity combining reduces the
variations of the signal and hence the number of experienced level crossings as illustrated in
Figure 32.

![Figure 32: Diversity combining effect](image)

Full treatment of Level Crossing Rate in a Rayleigh channel was given in section 3.1.1.4 of
chapter 3. By substitution of Equation 29, and Equation 30, into Equation 28 the average LCR
can be re-written as:

\[ N(R) = \sqrt{2\pi} f_{d_{\text{max}}} \rho \exp(-\rho^2) \]

Equation 45

where \( \rho = R / \sigma \sqrt{2} \) is the normalised RMS value of the envelope and \( f_{d_{\text{max}}} \) is the maximum
Doppler shift.

The above expression indicates that the average LCR is a function of the minimum Doppler shift
\( f_{d_{\text{max}}} \) and therefore, a function of the speed of the mobile unit as it is:

\[ f_{d_{\text{max}}} = f_c \frac{\nu}{c} \]

Equation 46

where \( \nu \) is the speed of the mobile unit, \( f_c \) is the carrier frequency and \( c \) the speed of light.
The speed of the mobile unit can therefore be estimated, provided availability of the average Level Crossing Rate of the received signal. By substituting Equation 46, into Equation 45 and rearranging Equation 46 to make \( \hat{f}_{d_{\text{max}}} \) the subject, the following is obtained:

\[
\hat{f}_{d_{\text{max}}} = \frac{\hat{N}(R)}{\sqrt{2\pi} \rho \exp(-\rho^2)}
\]

and

\[
\hat{v} = \frac{c}{f_c} \hat{f}_{d_{\text{max}}}
\]

where \( \hat{f}_{d_{\text{max}}} \) and \( \hat{v} \) are estimates of the actual \( f_{d_{\text{max}}} \) and \( v \) respectively.

In order to be able to demonstrate accuracy of any such algorithm simulations would be necessary. In doing so, correlated fast fading generators have to be used. It is important to point out that before this algorithm was developed and optimised, accuracy of the theoretical LCR versus the simulated LCR was established.

### 4.3.1 LCR Estimation

In this section a simple “single-point” level crossing estimation algorithm is initially proposed. The algorithm is then further enhanced by introduction of a “multi-point” element and an error minimisation feature.

#### 4.3.1.1 LCR Estimation Algorithm-1

The natural way to identify the most effective LCR threshold is to determine where the signal level crosses the most in a Rayleigh channel. The maximum number of level crossings in a Rayleigh fading channel is given by the expression:

\[
(LCR)_{\text{MAX}} = f_{d_{\text{max}}} \sqrt{\frac{\pi}{e}}
\]

The normalised reference level \( \rho \) at which the maximum number of crossings occur is \( \rho = 1/\sqrt{2} \) or -3dB as shown in Figure 33.

---

\(^6\) derivation of the maximum level crossing rate for a Rayleigh channel is discussed in section 3.1.1.4.
Equation 49 can be rearranged and to make the mobile speed \( v \), subject of the equation:

\[
\nu = \frac{c}{f_c} (L_{CR})_{max} \sqrt{\frac{\pi}{e}}
\]

where \( c \) is the speed of light and \( f_c \) is the carrier frequency.

Using the above expression, the mobile speed can be estimated by measuring the maximum number of level crossings of the simulated signal.

**4.3.1.2 Performance of LCR Estimation Algorithm-1**

Performance of this model was primarily evaluated in terms of the speed estimation error (%), as described below:

\[
\text{percentage error} = 100 \times \left| \frac{\text{actual value} - \text{estimated value}}{\text{actual value}} \right|
\]

Equation 51

A series of 300 simulations were performed in order to compute the estimation error (in percentage) for a speed range of 5 to 120 Km/h. The sampling rate was chosen to be 64K samples/sec, a submultiple of the actual UMTS chip rate 3.84 Mchips/sec. The performance of the model was evaluated for an estimation duration range of 50msec to 300msec, which corresponds to 5 and 30 UMTS frames [1] respectively.

Figure 34 illustrates the error performance of the speed estimation model.
Figure 34: Error performance of the simple speed estimation model over 300 series of simulations. The sampling rate is 64K samples/sec and the frame duration is assumed to be 10msec.

Figure 34 indicates that the mean speed estimation error is considerably high at low speeds and decreases gradually as higher speeds are reached.

The estimation error was averaged over the considered speed range for each individual measurement period (No of frames) and plotted in Figure 35, to provide an indication as to maximum required measurement period.

Figure 35: Percentage error averaged over the entire simulated speed range for the simple speed estimation model.

From Figure 35, it can clearly be seen that the estimation error improves as the number of considered frames (measurement period) increases. As a statistical measure, this is an expected
result, however, the important observation from the above is that no major performance improvement is experienced for measurement durations beyond the 30-frame mark.

The distribution of the speed estimation error was investigated and was found to be Gaussian (Figure 36).

![Figure 36](image)

**Figure 36:** Distribution of the speed estimation error and Gaussian pdf fit, (a) Speed = 30 Km/h, number of frames = 10, (b) Speed = 60 Km/h, number of frames = 20

The mean estimated speed, together with the corresponding standard deviation for two measurement periods (5 and 30 frames) are plotted in Figure 37.

![Figure 37](image)

**Figure 37:** Mean and standard deviation of the estimated speed, (a) 5 frames, (b) 30 frames

Having observed the performance of this simple speed estimation model, the following conclusions can be drawn:
• For small measurement windows the performance of the model is unacceptable.

• As the measurement window increases, the mean estimated speed gets closer to the actual speed whilst the standard deviation of each estimate reduces significantly.

• The standard deviation of the estimate is larger at higher speeds.

This leads to investigation of alternatives to improve the accuracy of the estimation.

4.3.1.3 Extended LCR Estimation Algorithm-1

In this section, the simple speed estimation model is expanded so that 4 different average level crossing rates at -3 dB, -6 dB, -9 dB and -12 dB (Figure 38) are simultaneously measured in a bid to provide more accurate estimations. The final estimate of the mobile terminal speed will be the average of these four estimations.

Figure 38: The Rayleigh envelope and the four selected LCR thresholds

The above 4 level crossing thresholds were selected after careful examination of the measured LCR of the simulated Rayleigh channel versus the theoretical approximations given in Equation 49. Figure 39, shows that the simulated LCR at these levels is accurately represented by the theoretical approximations.
4.3.1.4 Performance of the Extended LCR Estimation Algorithm-1

The error performance of the extended speed estimation model for different measurement periods is illustrated in Figure 40.

Figure 41 lists the mean percentage error in the speed estimation for a different number of frames.

Figure 41: Error performance of the simple speed estimation model over 300 series of simulations, sampling rate of 64ks/s and a frame duration of 10ms
Chapter 4

Environment Detection

Figure 41: Percentage error averaged over the entire simulated speed range for the extended speed estimation algorithm-1

The following graphs illustrate the standard deviation of the speed estimations for both the simple and expanded model and two different measurement durations (5 and 30 frames).

Figure 42: Standard deviation of the speed estimation error. Comparison between the simple and the extended model for varying signal duration: (a) 5 frames and (b) 30 frames

Comparing like with like, both the mean estimation error and the standard deviation of the estimation error were found to be larger with the expanded model. The performance of the expanded (multi-level) algorithm is therefore slightly worse than that of the single level algorithm.

Careful analysis of the simulations results indicated that this is mainly due to the fact that the actual number level crossings at lower thresholds is significantly lower than the -3dB level. This means less reliable measurements around the additional thresholds. As the above algorithm
combined the measurements at each threshold with an equal weighting factor it resulted in an overall worse performance.

4.3.1.5 Extended LCR Estimation Algorithm-2

The performance of the speed estimation algorithm can be improved if, somehow, instead of using ‘fixed’ threshold levels, it is possible to identify ‘appropriate’ thresholds that provide the most accurate estimates.

An advanced speed estimation algorithm base on this concept is introduced and further described in this section.

The fundamental underlying speed estimation concept remains unchanged whilst more intelligence in the form of a Signal Processing Unit is introduced to the algorithm.

The functional block diagram of the Advanced Speed Estimation algorithm is shown in Figure 43.

![Figure 43: Functional block diagram of the advanced speed estimation model](image)

Under this scheme, the speed estimation is performed in the following five distinct stages:
**Stage 1**: The Level Crossing Rate Measurement Unit produces a vector of 30 estimations of the level crossing rate at 30 different levels all the way from +4 dB to −10.5 dB with a step size of −0.5 dB.

**Stage 2**: The Signal Processing Unit translates this vector into a vector consisting of 30 individual candidate speed estimations before identification of the true value.

**Stage 3**: A new vector is produced, each element of which is the least sum of the squared differences between a given speed estimate and all other 29 candidate estimates.

**Stage 4**: The three estimations of the true speed that produce the least sum of squared differences are chosen.

**Stage 5**: The estimated speed is the average of these three estimations.

Please note that in Figure 43, $\hat{v}_1$ to $\hat{v}_{30}$ represents the speed estimates at corresponding level crossing thresholds of 1 to 30 and $\hat{v}_i$, $\hat{v}_j$ and $\hat{v}_k$ represent the best 3 estimates.

**4.3.1.6 Performance of the LCR Estimation Algorithm-2**

Performance of the above advanced speed estimation algorithm was evaluated for sampling rates ranging from 64K bits/sec to 512K bits/sec (with steps that are submultiples of UMTS chip rate). Figure 44 illustrates the error performance of the advanced speed estimation algorithm for different number of measurement periods, at a sampling rate of 64K samples/sec.

![Figure 44: Error performance of the Advanced Speed Estimation model over 300 series of simulations, sampling rate of 64K samples/sec and the frame duration of 10msec](image_url)
Figure 45 presents the mean percentage error in the speed estimation for all data rates.

As expected, it can be observed that increasing the data rate only results in a proportional increase of the computational time whilst the performance of the algorithm remained the same for different data rates.

Comparing the results listed in Figure 35, Figure 41, and Figure 45, for a sampling rate of 64K samples/sec, it can be conclude that the advanced speed estimation algorithm outperforms all the other considered algorithms. It reduces the average speed estimation error between 2 – 3.5% across the wide range of simulated speeds.

Alternatively, the advanced model requires fewer samples from the received signal to produce a reliable estimation compared to the other proposed algorithms. For example, in order to estimate the speed with a mean estimation error of about 9.2% algorithm-1 requires \(25 \times 640 = 16000\) samples, whilst the advanced model requires only \(15 \times 640 = 9600\) samples, i.e. 40% faster.

Not only the mean estimation error has been found to decrease with increasing measurement duration, but also the estimation standard deviation is significantly decreased.

This was the case for all sampling rates. The above is clearly demonstrated in Figure 46 and Figure 47.
Figure 46: Mean estimation and standard deviation of the estimation error for sampling rate 64K samples/sec and signal duration equivalent to (a) 5, (b) 10, (c) 20 and (d) 30 frames, respectively.

Figure 47: Standard deviation of the speed estimation error, (a) Sampling rate = 64K samples/sec (5, 10, 20, 30 frames), (b) Sampling rate = 512K samples/sec (5, 15, 20, 30 frames)
To better demonstrate the introduced improvements by the advance algorithm, Figure 48, shows the comparative reduction in the standard deviation of the error compared to algorithm-1 and its extended version.

\begin{figure}
\centering
\includegraphics[width=0.5\textwidth]{figure48.png}
\caption{Standard deviation of the speed estimation error. Comparison between the advanced, the simple and the expanded model. The sampling rate is 64K samples/sec and the measurement duration is 30 frames.}
\end{figure}

\section{4.4 Delay Spread Estimation}

Wideband propagation channels are commonly experienced in CDMA systems. By definition, if the bandwidth of a system is greater than the coherence bandwidth of the environment, the propagation channel is classed as a "wideband channel". The wideband channel theory based on which the following analysis are carried out is fully discussed in [14].

Wideband channels can be modelled as linear filters with a time-varying impulse response. The impulse response of a given channel contains important information that can be used to determine the proximity of the surrounding reflectors and hence the operational environment.

By taking the spatial average of an instantaneous impulse responses of a channel over a local area, the "\textit{power-delay profile}" of the channel can be found [14]. Several multipath channel parameters can be determined by the power-delay profile of a channel amongst which "\textit{excess delay}", "\textit{mean delay}" and the "\textit{RMS delay spread}" are the most important.

In this section, an algorithm for estimating the RMS delay spread of a channel is proposed. The RMS delay spread is widely used to quantify the time depressive properties of a channel. The types of environment that will be considered are the ones that are classified as "\textit{Test Environments}" in the UMTS specifications (APPENDIX-A)
4.4.1 RMS Delay Spread Estimation Model

The simulation model considered for evaluation of the proposed algorithm consists of an Impulse Response Generator and coupled by an RMS Delay Estimator, as shown in Figure 49.

![Figure 49: RMS delay spread estimation model](image)

The presented model estimates the RMS delay spread by spatial averaging of a large number of impulse responses.

4.4.1.1 Impulse Response Generator

The Impulse Response Generator effectively creates the multipath environments where multiple versions of the transmitted signal (echoes) arrive at the receiver through different paths at different times and with different power levels.

The considered model employs a number of Rayleigh fading signal generators each representing a wideband channel tap. The Rayleigh fading of each tap is assumed to be uncorrelated with the others. The number of taps, their associated delays and the average power of each tap used in this study are in fact those specified by 3GPP as standard UMTS test environments (APPENDIX-A).

The number of Rayleigh samples that are being considered for the generation of an impulse response are equal to the number of chips that constitute the (in-band) pilot sequence in each UMTS slot. This number may vary from 32 to 1024 chips. The power level for each tap of the ‘instantaneous’ impulse response is simply the average power of pilot chips.

![Figure 50: Block diagram of the Impulse response-generating unit](image)

4.4.1.2 The RMS Delay Spread Estimator

The input to this unit is the generated ‘instantaneous’ impulse response. The function of this unit is to compute the RMS delay spread of the instantaneous impulse response. Since the received signal in a multipath channel consists of a series of attenuated, time-delayed, phase shifted...
replicas of the transmitted signal, the base band impulse response of a multipath channel can be expressed as:

\[ h(t, \tau) = \sum_{i=0}^{N-1} a_i(t, \tau) \exp\left[ j2\pi f_c \tau_i(t) + \phi_i(t, \tau) \right] \delta(\tau - \tau_i(t)) \]  

Equation 52

where \( a_i(t, \tau) \) and \( \tau_i(t) \) are the real amplitudes and excess delays, of the \( i \)th multipath component at time \( t \).

The phase term represents the phase shift due to the free space propagation of the \( i \)th multipath component, plus any additional phase shifts encountered in the channel. The tap delays are fixed and known. If the channel impulse response is assumed to be time invariant, or is at least wide sense stationary over a small-scale time or distance interval, then the channel impulse response may be simplified to:

\[ h(t, \tau) = \sum_{i=0}^{N-1} a_i \exp(-j\theta_i) \delta(\tau - \tau_i) \]  

Equation 53

For small-scale channel modelling, the power delay profile of the channel can be found by taking the spatial average of \( |h(t, \tau)|^2 \) over a local area. By making several local area measurements of \( |h(t, \tau)|^2 \) in different locations, it is possible to build an ensemble of power delay profiles, each one representing a possible small-scale multipath channel state. At the end of the \( i \)th slot, the estimated RMS delay spread \( \hat{S}_i \) is the average of all RMS delay spreads, \( S_1, S_2, S_3, \ldots, S_i \), computed from the impulse responses of the previous slots including that of the present \( i \)th slot:

\[ \hat{S}_i = \text{average}\{S_1, S_2, \ldots, S_i\} \]  

Equation 54

\[ \text{Figure 51: Block diagram of the Signal Processing Unit} \]

As the number of the computed RMS delay spreads \( S_i \) increases, the estimate of the RMS delay spread \( \hat{S}_i \) should be expected to converge to the actual RMS delay spread value \( S \) for the specific channel. That is, of course, due to the fact that the actual RMS delay spread is calculated from the power-delay profile which essentially, is the average of a large number of impulse responses of the channel over a local area.
4.4.2 Performance of the RMS Delay Spread Estimation Model

The average performance of the model described in the previous paragraph is illustrated in the following plots. Figure 52-(a) and (b) show, the convergence of the estimate of the RMS delay spread and the mean percentage error of the estimation, as a function of the measurement window (number of slots), respectively.

![Figure 52: Performance of the RMS delay spread estimation averaged over 50 simulations, (a) convergence of the RMS delay spread estimation as a function of the number of slots, (b) Mean percentage error of the estimated RMS Delay spread as a function of the number of slots](image)

As expected, the RMS Delay Spread estimate \( \hat{S} \) converges to the actual value \( S \) and the mean percentage error drops as the number of considered impulse responses increases.

Figure 52-(b) shows that there is a residual mean percentage error, which is of the order of only 1-2%. This is due to the fact that during calculation of the RMS Delay spread, instead of considering the power of only one sample in every slot period of 0.000625 sec, the average power of a small number of samples (i.e. the samples of the pilot signal) are considered.

4.4.3 Impact of the pilot signal length on the model performance

In UMTS the number of chips that constitute the pilot component of a slot vary from 32 to 1024 chips depending on the channel bit rate. Table 2 lists the number of pilot bits or chips used for different bearer rates, i.e. different services on a dedicated downlink physical channel [28]. The channel bit and symbol rates are immediately before spreading (data bits). After spreading, by applying the appropriate spreading factor, a chip rate of 3.84 Mcps is achieved.

The number pilot symbols in a frame provides an indication of the maximum tap delay that can be 'captured'. The longer the pilot sequence, the greater excess delays can be captured. Since various Test Environments are characterised by different excess delays, the choice of pilot sequence length depends upon the operational environment.
From Table 5.4, it can be observed that the shorter pilot signals are used for higher channel bit rates (e.g. 32 chips/pilot signal for 2.048Mbps channel bit rate), which mainly represent very slow or almost stationary terminals in pedestrian and indoor environments with small excess delays. On the contrary, for lower bit rate services, large excess delays and longer pilot signals of up to 1024 chips are used.

<table>
<thead>
<tr>
<th>Channel Bit Rate (Kbps)</th>
<th>Channel Symbol Rate (Kbps)</th>
<th>SF</th>
<th>Bits/Frame</th>
<th>Bits/Slot</th>
<th>Npilot</th>
<th>Npilot Chips/slot</th>
</tr>
</thead>
<tbody>
<tr>
<td>16</td>
<td>8</td>
<td>512</td>
<td>160</td>
<td>10</td>
<td>4</td>
<td>1024</td>
</tr>
<tr>
<td>16</td>
<td>8</td>
<td>512</td>
<td>160</td>
<td>10</td>
<td>4</td>
<td>1024</td>
</tr>
<tr>
<td>32</td>
<td>16</td>
<td>256</td>
<td>320</td>
<td>20</td>
<td>8</td>
<td>1024</td>
</tr>
<tr>
<td>32</td>
<td>16</td>
<td>256</td>
<td>320</td>
<td>20</td>
<td>8</td>
<td>1024</td>
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<tr>
<td>64</td>
<td>32</td>
<td>128</td>
<td>640</td>
<td>40</td>
<td>8</td>
<td>512</td>
</tr>
<tr>
<td>64</td>
<td>32</td>
<td>128</td>
<td>640</td>
<td>40</td>
<td>8</td>
<td>512</td>
</tr>
<tr>
<td>128</td>
<td>64</td>
<td>64</td>
<td>1280</td>
<td>80</td>
<td>8</td>
<td>256</td>
</tr>
<tr>
<td>128</td>
<td>64</td>
<td>64</td>
<td>1280</td>
<td>80</td>
<td>8</td>
<td>256</td>
</tr>
<tr>
<td>256</td>
<td>128</td>
<td>32</td>
<td>2560</td>
<td>160</td>
<td>8</td>
<td>128</td>
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<td>128</td>
<td>16</td>
<td>5120</td>
<td>320</td>
<td>16</td>
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<td>640</td>
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<td>64</td>
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<td>1024</td>
<td>4</td>
<td>20480</td>
<td>1280</td>
<td>16</td>
<td>32</td>
</tr>
</tbody>
</table>

Table 2: Various lengths of Pilot signal for different Channel Bit Rates [15]

The impact of the pilot signal length on the rate of convergence of the RMS Delay Spread estimation was investigated. As shown in Figure 53, it was found that for a given mobile speed, the rate of convergence of the RMS Delay Spread remains independent of the pilot sequence length.

![Figure 53: Impact of the pilot signal length on the RMS Delay Spread estimation, (vehicular test environment, channel A, speed = 40 Km/h)](image)

(a) convergence plot (b) zoomed version

### 4.4.4 Impact of the mobile unit speed on the model performance

The convergence rate of the RMS Delay Spread depends on the mobile terminal speed. Figure 54 illustrates the convergence of the estimation, using a specific pilot signal length, for three different speeds.
Chapter 4

Environment Detection

Convergence of the estimated RMS Delay Spread

The convergence of the RMS Delay spread estimation appears to become slower as the speed of the mobile unit decreases. This can be explained as follows:

The lower speeds lead to lower fade rates of the received signal and therefore more samples (and equally more slots) need to be processed before a stable statistical average power for each Rayleigh tap can be adequately represented.

4.5 Effective Correlation Distance Estimation

A signal transmitted in a mobile propagation environment experiences fast fading due to multipath propagation, as well as slow fading, also called ‘Shadowing’. Shadowing is mainly due to the varying nature of the obstruction configuration in the propagation path between the mobile unit and the base station and it is responsible for the slow variations of the received signal.

The time varying nature of the shadowing process [17] is characterised by correlation properties of the received signal, “effective autocorrelation distance”, as defined in Chapter 3. Different types of environments are characterised by different values of effective autocorrelation distance and, therefore, it is possible to enhance detection of the operational environment using this information. Effective Correlation Distance Estimation, provides another indicator of the operational environment, in addition to the RMS delay spread.
4.5.1 Effective Autocorrelation Distance Estimation Model

The effective autocorrelation distance estimation model consists of a signal generator, an adaptive filtering unit, a speed estimation unit and an effective autocorrelation distance estimation unit as shown in Figure 55, below.

![Figure 55: Block diagram of the effective autocorrelation distance estimation system](image)

4.5.1.1 Signal Generator

This unit consists of a Rayleigh (fast fading) and a Log-Normal (shadowing) generator, as described in chapter 3. For a given effective correlation distance, specific to an operational environment and a mobile speed of \( v \), the shadowing generator produces shadow fading samples at a predefined sampling rate \( S_s \). A correlated Rayleigh generator is also used to produce samples of zero mean for a mobile speed of \( v \), at a sampling rate \( S_r \).

The outputs of the two generators are combined so that the resulting signal 'experiences' both slow and fast variations (long and short term fading), similar to the real mobile propagation environments.

4.5.1.2 Filtering Unit

The function of this unit is to filter the received signal in order to average out the fast fading variations and separate the slow fading. The target 'filtering window' length should effectively be a period sufficient for accurate separation of the fast and slow fading from each other. Typical lengths of \( L \), were found [26] to be between \( 10-80\lambda_s \), depending on the propagation environment. The initial indication for the operational environment can be provided by the delay spread estimator of section 4.4, based on which an appropriate \( L \) can be selected. An estimate of the speed \( \hat{v} \), is also provided by the speed estimation unit. The filtering 'window duration' in time is therefore constantly adjusted to \( \frac{L}{\hat{v}} \) [sec].
**4.5.1.3 Speed estimation unit**

The output of the filtering unit is the fast fading and the slow fading signal, separated into two individual streams. The fast fading signal is processed by the advanced speed estimation system, as described in section 4.3. The estimated speed \( \hat{v} \) is used to ‘adjust’ the duration of the filtering window.

As the speed of the mobile is unknown at the beginning of the process, an arbitrary speed \( v_{\text{assumed}} \) will be selected based on which the starting averaging window length will be decided. As a result, the process would require an initial period to converge the assumed speed to the actual speed. After this period, the speed will be tracked accurately. The time taken for convergence depends on the assumed speed versus the actual speed.

**4.5.1.4 Effective autocorrelation distance estimation unit**

The shadowing samples, that are discriminated by means of filtering, are processed by this unit so that the number of samples after which the correlation coefficient drops to the value of \( e^{-1} \), is determined. This number of samples is converted to time and subsequently to distance using the speed estimate. This distance is the estimated autocorrelation distance \( \hat{X}_C \).

**4.5.2 Performance of the Effective Correlation Distance Estimator**

Performance of the above system was evaluated under a typical simulation scenario. The Rayleigh samples were generated for a 120 Km/h mobile units at 2GHz. The sampling distance between successive generated Shadowing samples was chosen to be 15\( \lambda \) and the effective autocorrelation distance was set to 20m. The speed \( v_{\text{assumed}} \) needed for carrying out the first filtering of the received signal was assumed to be 96 Km/h, i.e. with a 20% error with respect to the actual value.

The number of shadowing samples per iteration, produced by means of filtering for estimation of \( X_C \) was taken to be equal to 2000 samples. Figure 56, provides a visual indication of how well the algorithm performs.
The system performance was evaluated for 10 consecutive iterations (estimations of $X_c$). The simulations were repeated 20 times with different starting seeds and similar statistical results were reproduced every time.

Figure 57, shows the estimated versus the actual effective correlation distance.
Chapter 4 Environment Detection

It is important to note that whilst the ‘average’ effective correlation distance in the shadowing signal generator was 20m, the actual generated correlation distance varies in time. For this reason, iteration-based comparison between the generated and the estimated $X_c$ was made. Furthermore, it can be seen that after the speed estimation converges from the assumed level to the actual level, the accuracy of the $X_c$ estimations increase. Figure 58, shows how rapidly the speed estimation converges.

![Graph showing speed estimation convergence](image)

**Figure 57:** Mean estimated and mean generated effective correlation distance

Figure 59, shows the accuracy of the $X_c$ estimates with respect to the actual.

![Graph showing accuracy of $X_c$ estimates](image)

**Figure 58:** Mean and standard deviation of speed estimation

Figure 59, shows the accuracy of the $X_c$ estimates with respect to the actual.
The above clearly demonstrates that after 6 iterations, the estimate $X_C$ converges to produce an error of just 1%.
Chapter Five

5. Adaptive CLPC

An effective method to overcome the latency and power control command rate limitations in CLPC is the use of Adaptive CLPC algorithm. As discussed in chapter 2, in conventional CLPC the step size by which the power of a mobile user is adjusted is fixed. This reduces the accuracy of the power control and limits the ability of a mobile user to track fast envelope changes rapidly.

In this chapter, a simple yet effective signal quality measurement technique based on which power control decisions can be made, is presented. The work then proceeds to establish the performance baseline for conventional fixed step CLPC in both terrestrial and satellite propagation environment. In the satellite case, performance of conventional CLPC algorithm in actual channel recordings is evaluated.

The above is then used as a base to which performance of the proposed Speed Adapted CLPC (SA-CLPC) algorithm is compared to. The performance of SA-CLPC is thoroughly examined under both terrestrial and satellite propagation environment and significant gains are identified.

The adaptation mechanism plays an important role in performance of any such algorithm. In Chapter 4, a speed and environment detection algorithm was proposed and evaluated for this purpose. In the case of terrestrial environment, the impact of imperfect Speed Estimation on the SA-CLPC is demonstrated.


5.1 Signal Quality Estimation

In digital communication systems, the quality of a link is determined by the average ratio of a bit Energy, $E_b$ to the Noise + Interference density, $N_0+I_0$. As MAI in CDMA systems has been shown to have a Gaussian distribution, the term interference is commonly referred to as noise rise. For simplicity, throughout this section it is considered that $N_0$ does include the noise rise due to MAI.

In controlling the quality of a single link, one only has control over the transmitted power as opposed to the noise and interference in the system. The term ‘power control’ refers to a mechanism for controlling the transmitted power of a BS/MS. The most appropriate criterion based on which control commands are issued is quality of the link, $E_b/N_0$.

As it can easily be concluded, an important component of the power control technique discussed here, is the estimation of $E_b/N_0$. The $E_b/N_0$ estimator proposed here, operates on the disspread samples $|x_j|$ in both the in-phase (I) and quadrature (Q) channels of a QPSK model [7].

The simplified algorithm involves the following calculations,

$$\frac{\hat{E}_b}{N_0} = \frac{\mu_I^2 + \mu_Q^2}{2(\sigma_I^2 + \sigma_Q^2)}$$

Equation 55

Where $\mu$ and $\sigma$ for any of the two channels can be estimated using the following equations,

$$\mu = \frac{1}{N} \sum_{j=1}^{N} |x_j|$$

Equation 56

$$\sigma^2 = \frac{1}{N} \sum_{j=1}^{N} |x_j|^2 - \mu^2$$

Equation 57

Note that Equation 55 can also be used for BPSK systems, where the quadrature channel (Q) is not present, by eliminating $\mu_Q$ and $\sigma_Q$.

Since we are only interested to examine the viability of the estimation technique, an AWGN channel with an average $E_b/N_0$ of 10 dB has been used, whilst the error statistics have been accumulated over 200 simulations.

In Figure 60, accuracy of the above algorithm is examined. The figure naturally shows, that the larger the number of samples used for an estimate the more accurate the estimation will be. The estimation error is defined as the difference between the actual and the estimated value.
Figure 60: $E_b/N_0$ estimation accuracy versus the number of QPSK symbols used in the estimation

Figure 61 shows the output of the estimated $E_b/N_0$ (for an actual $E_b/N_0$ of 10dB) versus the variations of the Rayleigh fading channel. The number of QPSK symbols used, for each estimation is 100 bits in a Rayleigh fading channel.

Figure 61: $E_b/N_0$ Estimation in a Rayleigh channel

As it can be seen from the above, the $E_b/N_0$ estimations closely follow the fading profile. The estimates were deliberately shifted several dBs to better illustrate the similarity in the profile. From a single user point of view, a single CLPC model would therefore behave quite similarly whether $E_b/N_0$ or signal strength thresholds are used as a decision making criterion. In the proceeding sections, the power threshold criterion is used for demonstration establishing superiority of the proposed algorithm.
5.2 Performance of Conventional CLPC

In Chapter 2, the concept of ‘perfect power control’ was introduced. However, perfect power control cannot be achieved under realistic propagation environments. In this section the performance of conventional power control in both terrestrial and satellite channel environments are evaluated. This will highlight limitations of the conventional CLPC and establishes comparative bases for performance evaluation of the advanced CLPC schemes presented in the proceeding sections of this chapter.

5.2.1 Performance of Conventional CLPC in Terrestrial Channel

Whilst fixed step size CLPC is designed to track fast fading, its true performance is highly dependent on the time varying behaviour of the propagation channel, the power control command rate, power control step size and the Round Trip Delay (RTD).

For evaluation of the CLPC performance in a terrestrial environment, Rayleigh channel has been selected to represent the worst-case flat fading propagation environment.

Figure 62 shows that in such fast fading environments, the upper half (around the mean) of a Rayleigh channel has a smaller rate of change compared to the deep fades of the lower half.

Consequently, some parts of the fading can be better tracked than others resulting in an overall performance that is somewhat imperfect. The performance would be different if the step sizes or the command rate were different as discussed in section 5.3. But before entering this topic, the performance limitations of conventional fixed step CLPC need to be discussed.
The fixed-step CLPC model used for performance evaluations is shown in Figure 63. The model was set-up to evaluate performance of the CLPC in the most hostile flat fading propagation environment, the Rayleigh channel. Due to the correlated nature of the channel, it is extremely difficult to analytically derive the power control error [11] and [25]. Simulation is therefore used to evaluate the performance of the loop.

The operation of the above model is similar to the actual CLPC mechanism described in section 2.5.1.2. The received power \( P_r \) during a power control command control is expressed in Equation 58.

\[
T_p = -\int_{-\infty}^{T_p} P(t) G(t) dt
\]

where \( P(t) \) is the transmitted power, \( T_p \) is the power control command period (1/power control command rate) and \( G(t) \) is the fading in a noise free environment. As shown above, the received signal strength is averaged over a period of time equal to \( T_p \) based on which power control commands are issued.

\[ P(t) \] can be expressed by Equation 59.

\[
P(t) = P(t-T_p) \pm \Delta_p
\]

where \( \Delta_p \) is the power control step size.

Each received command changes the transmitted power by a fixed step size relative to the previous transmitted level.

In this model, it is assumed that the MS receives all the power control commands correctly in the forward link. It is further assumed that the power control bits are not encoded and hence do not experience any additional decoding delay. Using the above model, performance of conventional CLPC was evaluated under a range of simulation parameters. Figure 64, Figure 65 and Figure 66
show the performance of a conventional CLPC control (signal strength method) in a correlated Raleigh channel operating at 900MHz. Note that in all the presented cases a traffic channel rate of 9.6 kb/s and a delay of 1bit (round trip delay=100μs which refers to a MS distance of about 15km).

Figure 64: (a) Transmitted amplitude and the channel variations, speed 20km/h, command rate of 800 b/s, fixed step of 1dB (b) The received amplitude envelope after power control

As it can be seen from the above even at low MS speed of 20km/h, the received amplitude still experiences deep fades. However, as shown in Figure 65, by increasing the power control command rate the power control can become much more effective.

Figure 65: (a) Transmitted amplitude and the channel variations, speed 20km/h, command rate of 2.4 kb/s, fixed step of 1dB (b) The received amplitude envelope after power control

The conventional closed loop power control reaches its limitations (deep fades of about 8dB are still experienced after power control) at higher MS speeds of about 40km/h even with high power control command rates as shown in Figure 66.
Chapter 5

Adaptive CLPC

Figure 66: (a) Transmitted amplitude and the channel variations, speed 40km/h, command rate of 2.4 kb/s, fixed step of 1dB (b) The received amplitude envelope after power control

The above provides an indication of the performance under a number of power control command rates extremes. The 800bps represent the lowest power control command rate used in today’s narrowband CDMA system, IS-95 and the 2.4kbps represents the upper bound limit that was being considered for 3rd Generation mobile systems some years ago.

Today’s terrestrial UMTS (UTRAN) utilises a power control command rate of 1.5kbps, operating in the 2GHz band. Given the importance of this technology, all the presented work in proceeding sections are based on the UMTS power control command rate and a centre frequency of 2GHz.

Figure 67, shows performance of the conventional fixed step size CLPC with a control command rate of 1.5kbps in a Rayleigh channel at 2GHz and a RTD, representing a 30km distance between MS and BS.

Figure 67: Performance of a conventional CLPC with fixed step sizes of 1 & 2dB
As it can be seen from the above, standard deviation of the PCE increases with higher speeds clearly indicating ineffectiveness of the CLPC at higher speeds. The above simulations were carried out for a fixed step size of 1 and 2dBs, as specified in UMTS.

In conventional CLPC, different MS power control step sizes may be negotiated during the call set-up. Nevertheless, the power control step size is fixed for the entire duration of the call. This is in fact quite limiting as the fixed step CLPC performance varies under different vehicular speeds, operating centre frequencies, propagation channels and RTDs.

Larger step sizes perform better at higher speeds and smaller step sizes can control the power much more accurately for lower mobile speeds.

It is important to point out that power control is essential for low speed mobile users. Interleaving is ineffective in low speeds hence a burst of bits received in error cannot be corrected by channel coding. At speeds above 40km/h, interleaving begins to randomise errors effectively thereby relaxing the requirement for tight power control. At very high speeds, the standard deviation of the PCE converges towards the standard deviation of the Rayleigh channel itself.

5.2.2 Performance of Conventional CLPC in Satellite Channel

As far as the accuracy of the CLPC is concerned, the long Round-Trip Delays (RTDs) associated with the satellite environment are by far the most limiting factor. This significantly limits the power control command rate (1/RTD), which in turn reduces the ability of the CLPC in tracking the fast fading. Nevertheless, CLPC could still prove very useful under circumstances such as a slow moving or a stationary MS which might be in fade for example in the return but not in the forward link.

Figure 68, shows the minimum and maximum RTDs for three typical non-geostationary orbits of,

- a 66 satellite polar LEO system with a minimum elevation angle of about 8°
- a 48 satellite inclined LEO system with minimum elevation angle of 10°
- a 10 satellite inclined MEO system with minimum elevation angle of 10°
When considering a particular satellite constellation the minimum and maximum RTDs could be slightly misleading as they represent the two extremes where both the user and the feeder links are either at the minimum elevation angle (maximum RTD) or at 90° (minimum RTD). It is therefore more appropriate to consider the average elevation angles for each constellation for estimation of the average RTD. The average elevation angle statistics of a constellation change with latitude and therefore knowledge of the average RTD for different latitudes is of vital importance (Figure 69).

Knowing the above, various CLPC algorithm parameters can be fine-tuned for a given constellation.
To demonstrate the impact of RTD on performance of CLPC simulations were carried. The conventional CLPC model of Figure 70 is similar to that of the terrestrial case. The only difference here is that instead of a simulated channel, the performance of the conventional CLPC was evaluated with actual channel recordings [29].

**Figure 70: Fixed step satellite CLPC model**

Simulations were carried out using the L-Band propagation data recorded at 8m/s, 1024 samples/s as discussed in Chapter 3. The measurements were carried out in a range of operational environments, but for the purpose of this study, wooded and suburban environments were selected. These environments are the primary operational environments for satellite communications systems, where the terrestrial coverage may be patchy.

As discussed in section 3.2, the recordings were processed and all large-scale variations (shadowing component) were removed. The channel was recorded at a fixed velocity and a constant sampling rate. It is possible to simulate different MS speeds using the same recordings. Playing the data faster can easily simulate higher speeds, however, as the speed is lowered, less samples/s will have to be processed which would then lead to the loss of resolution. To overcome this problem interpolation techniques (preferably sinc-interpolation) can be used on the data.

In order to demonstrate the performance degradation of the CLPC in the satellite environment with long RTDs, a range of simulations with different RTDs representing different satellite altitudes were carried out.

The standard deviation of the PCE was evaluated for a range of step sizes (0.25-1.25 dB). Best results were achieved using a fixed step size of 0.5 dB. Figure 71 shows the standard deviation of the PCE for a range of delays covering up to the highest LEO. Longer delays were not simulated as the standard deviation of PCE was approaching the standard deviation of the channel without CLPC. In all cases simulated, the PCE was found to be log-normally distributed.
It can be seen that for RTDs of more than 20ms, standard deviation of the PCE reaches its maximum. It is also interesting to see that the heavily wooded environment is a more hostile environment as far as CLPC is concerned. It is so, as the small dimension of the tree leaves and stems causes rapid fluctuations in the received signal level. Although these fluctuations could be considered as shadowing, it is neither possible to filter them out (unless averaged over a very short distance which almost certainly result in large estimation errors in other environments), nor to track them with CLPC.

As the power control command rate (1/RTD) is very limited in the satellite environment, the use of variable step size power control could improve the performance as shown in the proceeding sections.

But before describing and evaluating the proposed power control scheme, it is important to establish the reference conventional CLPC to which comparisons will be made. On that note, two Ricean channels (k=8dB and k=10dB) were used for simulations to establish the reference point. Furthermore, as discussed above RTD in satellite environment has a significant impact on the performance of CLPC. This together with a varying MS speed create a set of multidimensional criterion against which CLPC performance needs to be evaluated.

The term $F_d T$, where, $F_d$ represents the maximum Doppler frequency (combined speed and centre frequency representation), and $T$ denotes both the power control command period and the RTD assuming the received signal is averaged by the FES over a period of a single round trip delay. This implies a maximum power control command rate of $1/T$. 
Figure 72 and Figure 73, show the experienced power control error (PCE) versus $F_dT$.

**Figure 72:** Performance of fixed step size CLPC in a Rice channel, $k=10\text{dB}$

**Figure 73:** Performance of fixed step size CLPC in a Rice channel, $k=8\text{dB}$
As expected, the CLPC performance in satellite systems is generally much worse than that of the terrestrial cellular case. This is simply due to long round trip delays associated with such systems. Nevertheless, CLPC is still effective for lower values of $F_d T$. It is important to point out that the performance figures above show that for higher values of $F_d T$, an increased PCE compared to that of the Ricean channel with no CLPC (the dashed line) is experienced.

In order to provide a simple method for translation of $F_d T$ to RTD and MS speed, the following table has been generated assuming a centre frequency of 1.5GHz.

<table>
<thead>
<tr>
<th>MS Speed (km/h)</th>
<th>RTD or Power Control Command Period (ms)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>2.03 0.06 0.08 0.11 0.14 0.17 0.19 0.22 0.25 0.28 0.31 0.33 0.36 0.39 0.42 0.44</td>
</tr>
<tr>
<td>20</td>
<td>0.07 0.14 0.21 0.28 0.35 0.42 0.49 0.56 0.63 0.69 0.76 0.83 0.90 0.97 1.04 1.11</td>
</tr>
<tr>
<td>30</td>
<td>0.14 0.28 0.42 0.56 0.69 0.83 0.97 1.11 1.25 1.39 1.53 1.67 1.81 1.95 2.08 2.22</td>
</tr>
<tr>
<td>40</td>
<td>0.26 0.56 0.83 1.11 1.39 1.67 1.95 2.22 2.50 2.78 3.06 3.34 3.61 3.89 4.17 4.45</td>
</tr>
<tr>
<td>50</td>
<td>0.42 0.83 1.25 1.67 2.08 2.50 2.92 3.34 3.75 4.17 4.59 5.00 5.42 5.84 6.25 6.67</td>
</tr>
<tr>
<td>60</td>
<td>0.69 1.39 2.08 2.78 3.47 4.17 4.86 5.56 6.25 6.95 7.64 8.34 9.03 9.73 10.42 11.12</td>
</tr>
<tr>
<td>70</td>
<td>0.97 1.95 2.92 3.89 4.86 5.84 6.61 7.78 8.79 9.73 10.70 11.67 12.65 13.62 14.59 15.57</td>
</tr>
</tbody>
</table>

Table 3: $F_d T$ table, for different MS speed and RTDs at 1.5GHz

### 5.3 Speed Adapted CLPC (SA-CLPC)

#### 5.3.1 SA-CLPC

In previous sections of this chapter, the baseline performance of conventional CLPC with fixed step size was established. Furthermore, it was shown that for any given speed, different power control step sizes result in different standard deviation of PCE.

The idea behind Speed Adapted CLPC is to simply select the best power control step size for the given speed. Figure 74, shows the operation of any such models.
The performance of such algorithm not only depend on the performance of the speed estimation algorithm, but also on prior knowledge of which step size produces the best results at a given speed.

In proceeding sections of this chapter, SA-CLPC in the context of terrestrial and satellite propagation environment is examined and its superiority over conventional methods clearly demonstrated.

5.3.2 SA-CLPC Performance in Terrestrial Channel

The first step in development of SA-CLPC is to establish the optimum step size for a given speed. Due to the correlated nature of the channel, analytical approaches to evaluation of the optimum step size are not feasible. Therefore, in order to establish these figures a large number of simulations were carried out.

Figure 75, shows performance of the conventional fixed step size CLPC for range of step size, with a control command rate of 1.5kbps in a Rayleigh channel at 2GHz and a RTD, representing a 30km distance between MS and BS.

![Figure 75: CLPC performance in a Rayleigh channel for different step sizes](image)

From the above it is evident that CLPC performance is improved at lower speed when small steps are used and at higher speeds when larger steps are selected. By focusing on two different speed, 3km/h and 50km/h, Figure 76 and Figure 77 better illustrate the potential gains in selecting the optimum speed.
Figure 76: CLPC performance in a Rayleigh channel for different step sizes, zoomed at 3km/h

Figure 76 shows that by selection of the 0.5dB step size at 3km/h, an average reduction of PCE std equal to 0.3dB and 0.5dB could be achieved compared to 1 and 2dB step sizes, respectively.

Figure 77: CLPC performance in a Rayleigh channel for different step sizes, zoomed at 3km/h
Figure 77, similarly shows that at 50km/h a reduction of 0.8dB in the std of PCE could be achieved if the selected step size is increased from 1dB to 3dB.

Assuming perfect knowledge of MS speed, the most optimum step sizes can be selected from Figure 75 to form the following look up table.

<table>
<thead>
<tr>
<th>Speed (km/h)</th>
<th>3</th>
<th>5</th>
<th>10</th>
<th>20</th>
<th>30</th>
<th>40</th>
<th>50</th>
<th>70</th>
<th>100</th>
</tr>
</thead>
<tbody>
<tr>
<td>Step Size, $\Delta_p$ (dB)</td>
<td>0.5</td>
<td>1.0</td>
<td>1.5</td>
<td>2.0</td>
<td>2.5</td>
<td>2.5</td>
<td>3.0</td>
<td>3.0</td>
<td>3.0</td>
</tr>
</tbody>
</table>

*Table 4: Step size look up table*

The selection of appropriate step size for any given speed results in a significant reduction of PCE as shown in Figure 78.

![Figure 78: Performance of SA-CLPC in comparison to 1dB fixed step conventional CLPC](image)

However, knowledge of the exact speed would require perfect speed estimation, an impossible feature based on the findings of Chapter 4. From Figure 48 of section 4.3.1.6 it is evident that for an estimation period equal to 30 UMTS frames (300ms), the proposed advanced speed estimation algorithm would result in normally distributed speed estimations with standard deviations summarised in the following table:

<table>
<thead>
<tr>
<th>Speed (km/h)</th>
<th>3</th>
<th>5</th>
<th>10</th>
<th>20</th>
<th>30</th>
<th>40</th>
<th>50</th>
<th>70</th>
<th>100</th>
</tr>
</thead>
<tbody>
<tr>
<td>Std of Estimations (dB)</td>
<td>0.5</td>
<td>1.0</td>
<td>1.5</td>
<td>2.0</td>
<td>3.0</td>
<td>3.5</td>
<td>3.8</td>
<td>5.0</td>
<td>7.1</td>
</tr>
</tbody>
</table>

*Table 5: Standard deviation of speed estimation*
The above results were used in conjunction to establish the impact of imperfect speed estimation on the performance of SA-CLPC. Under practical circumstances where the MS/BS is estimating the speed of the user, knowing the exact speed is not critical, but knowing the speed category becomes important. A speed category is defined as a range of speed all of which will be represented by a single speed of Table 5. For example, under the proposed algorithm all estimated speeds from 0-3.5km/h are categorised under the 3.0km/h category. The following table shows all the speed ranges for different speed categories.

<table>
<thead>
<tr>
<th>Speed Category (km/h)</th>
<th>3</th>
<th>5</th>
<th>10</th>
<th>20</th>
<th>30</th>
<th>40</th>
<th>50</th>
<th>70</th>
<th>100</th>
</tr>
</thead>
<tbody>
<tr>
<td>Speed Range (km/h)</td>
<td>&lt;3.5</td>
<td>3.6-7</td>
<td>7.1-13</td>
<td>13.1-24</td>
<td>24.1-35</td>
<td>35.1-45</td>
<td>45.1-60</td>
<td>60.1-85</td>
<td>85.1-200</td>
</tr>
</tbody>
</table>

*Table 6: Speed categories and their corresponding range*

The above boundaries were optimised together in conjunction with the step size to ensure the highest probability of correct speed size selection for all speed categories.

Based on the above speed categories and their corresponding step sizes, simulations were carried out to determine the average probability of selecting the correct speed category and hence the probability of selecting the correct step size. Figure 79, presents the results.

*Figure 79: Probability of correct speed estimation vs. probability of correct step size selection*
To better understand the above, it is essential to illustrate the behaviour of the considered speed estimation algorithm. Figure 80 shows the PDF of each speed category assuming 30 frames per estimate as stated in Figure 47(a).

![Speed estimation probability density function](image)

**Figure 80: Speed estimation probability density function for different speeds**

The reasons for a reduction in probability of correct speed category estimation at 40km/h and 50km/h can be clearly seen from the above. Excessive overlap between the PDF of the above two categories brings about uncertainty as to which speed category the user is in.

Nevertheless, as per details of Table 4, a number of neighbouring categories (for example 30km/h and 40km/h) use the same step size (2.5dB). For this reason higher probability of correct step size selection compared to speed category selection is shown in Figure 79. In such circumstances, some incorrect speed category estimations will still result in correct step size selection.

Having established the probability of correct step size selection, the simulation model of Figure 74 was used to evaluate the impact of imperfect speed estimation on SA-CLPC performance.
As it is shown above, no significant performance degradation was observed. This is mainly due to the fact that in the event of incorrect step size selection, the neighbouring step size, which incidentally is also the second best step size, is selected.

It is important to point out that the above speed estimations were carried out using 30 UMTS frames (or 300ms period). The performance evaluations of Chapter 4 demonstrate no significant reduction in standard deviation of the estimated error for estimation periods beyond 30 frames. Also looking at realistic acceleration behaviour of a MS, it is safe to assume that the speed of the MS would remain constant over this period.

To conclude, superiority of the SA-CLPC algorithm was clearly demonstrated and its performance in conjunction with the advanced speed estimation algorithm of Chapter 4 was demonstrated.

5.3.3 SA-CLPC Performance in Satellite Channel

In section 5.2.2, performance of conventional CLPC in recorded and simulated propagation channels were demonstrated. Similar to the SA-CLPC for terrestrial systems, the first step in development of SA-CLPC for satellite environment is the establishment of the optimum step size for a given $F_d T$ [30].
Chapter 5

Adaptive CLPC

Figure 82: Performance of fixed step power control in Ricean channel, Rice factor of 10dB

Figure 83: Performance of fixed step power control in Ricean channel, Rice factor of 8dB

From the above it is evident that the performance of CLPC after an $F_dT$ of 1.2 is worsened by the loop. In fact by switching off the CLPC in such circumstances the overall performance is improved. The area in which CLPC is effective can be better observed in the magnified version of same plot in Figure 84 and Figure 85.
Figure 84: Magnified performance of fixed step power control of Figure 82

Figure 85: Performance of fixed step power control in Ricean channel, Rice factor of 8dB

As it can be seen below, a single step size does not necessarily have the best performance at all the $F_dT$s. A slight degradation in the CLPC performance compared to the case of Figure 82, can also be observed as the considered channel in Figure 85, fluctuates relatively faster.
Assuming knowledge of the vehicular speed through a speed estimation algorithm, the MS would then be able to select the best step size by using the appropriate lookup tables as discussed in the terrestrial case.

<table>
<thead>
<tr>
<th>$F_d T$</th>
<th>0.02</th>
<th>0.04</th>
<th>0.08</th>
<th>0.17</th>
<th>0.25</th>
<th>0.42</th>
<th>0.58</th>
<th>0.83</th>
</tr>
</thead>
<tbody>
<tr>
<td>For K=10dB, Step Size, $\Delta_p$ (dB)</td>
<td>0.5</td>
<td>1.0</td>
<td>1.5</td>
<td>0.0</td>
<td>0.0</td>
<td>0.0</td>
<td>0.0</td>
<td>0.0</td>
</tr>
<tr>
<td>For K=8dB, Step Size, $\Delta_p$ (dB)</td>
<td>0.5</td>
<td>1.0</td>
<td>1.5</td>
<td>0.0</td>
<td>0.0</td>
<td>0.0</td>
<td>0.0</td>
<td>0.0</td>
</tr>
</tbody>
</table>

Table 7: Step size look up table for a Rice factor of 10dB and 8dB

It is important to point out that under the proposed scheme, CLPC algorithm can be switched-off after certain vehicular speeds ($F_d T$) where significant degradation in the performance is expected.

Figure 86, shows the performance comparison between the conventional fixed step and the speed-adapted scheme.

Figure 86: Comparison between the conventional 1.0 dB fixed-step and SA-CLPC, $k=10$dB
From the above significant performance gains of about 0.5dB for the lower values of $F_d T$ can be observed. Furthermore, reductions of PCE std, in excess of 1.0dB can be made by just switching the power CLPC loop off. This is mainly due to the fact that for high values of $F_d T$, the loop delay compared to channel variations is too long thereby resulting in incorrect commands executed.

Considering the long RTDs of the channel, it is beneficial to reduce the averaging periods (bin sizes) based on which the commands are issued at the FES, as shown in Figure 88.
That is to issue commands only based on a much smaller portion of the received signal between two consecutive issuing. Under the proposed scheme [30], two different approaches were investigated.

In the first approach, control commands are issued based on the last $0.25T$ seconds of the received signal, where $T$ is the power control command period or the round trip delay. This requires no additional estimation algorithm at the FES as the power control command rate is set for a given system.

In an alternative approach the decision is made over a varying averaging period which always corresponds to the time taken for the MS to travel $1/500$ of a wavelength. This approach can only be implemented with the aid of a speed estimation algorithm at the FES, as shown Figure 88.

Simulations results for both options in Ricean channels with factors of 8 & 10dB are shown in Figure 89 and Figure 90, respectively.
It can be observed from the above that by reducing the size or the averaging period in the FES, based on which power control commands are issued, further improvements can be made. From Figure 89 and Figure 90, it can be seen that this technique introduces a more significant gain for channels with lower values of Rice factor. This is mainly due to the fact that in such channels, the average fade duration is shorter, thereby the received commands at the MS have a higher probability of being invalid. Reduction of the averaging period at the FES therefore increases the accuracy of the commands resulting in a visible gain.

It can further be seen that in the case of the Ricean channel with a factor of 8dB (Figure 90), the first technique in which averaging periods of 0.25T, are taken improves the PCE by a further 0.3dB, compared to that of the SA-CLPC algorithm.

Although the second technique whereby an averaging period of 1/500 of the wavelength is taken does not show great improvement over the above technique, it highlights an interesting trend. On careful examination of results of Figure 90 and Figure 89, it can be seen that this technique, however small, has further improved the performance. In fact early results show a much greater improvement in Rice channels with Rice factors below 7dB.
Chapter 6

6. Distributed Power Control

As discussed in Chapter 3, power control is an essential feature of CDMA systems. Effective power control in mobile communications maximises system capacity, improves the quality of service and ensure efficient use of available radio resources (bandwidth and transmitter power). In Chapters 4, adaptive power control algorithms that control power of each user were introduced.

This Chapter focuses on system-wide impact of individual power control algorithm of each user in mobile terrestrial systems. In managing system interference, outer loop power control becomes an essential overriding mechanism that looks after the system QoS as oppose to the individual users only.

Proceeding sections of this chapter establish the theory behind centralised and distributed power control. An advanced distributed power control algorithm is then presented and its superiority is demonstrated. It is further shown how the proposed algorithm can make use of outer loop power control to achieve interference management (or C/I balancing) in UMTS.
6.1 System Model Fundamentals

In this section all system assumptions and fundamentals based on different system level power control approaches are discussed and evaluated are established.

6.1.1 Link Gain Matrices

A finite mobile system consisting of $N$ square grid cells is assumed. Assuming a CDMA system, frequency reuse pattern of "1" is applied to all users in all cells. Consequently, intra-cell interference as well as inter-cell interference are taken into account. The base stations are located at the centre of the cells and omni directional antennas are employed. $Q$ mobiles are randomly located in the system with a uniform density of mobiles per cell.

Cell membership is determined by the maximum pilot power amongst the cell sites the subscriber receives. Mobiles transmit at a constant information rate without the use of voice activity detection, which will increase the capacity of an actual system. Only fully loaded systems are investigated in order to derive the upper performance bound and thus all channels (codes) are assumed to be in use.

The normalised uplink gain matrix of the system $W = \{W_{ij}\} (Q \times Q)$ can be defined as:

$$W_{ij} = \begin{cases} \frac{G_{ij}}{G_{ii}} & i \neq j \\ 0 & i = j \end{cases}$$  

Equation 60

where $G_{ij} (Q \times Q)$ is the link gain on the path between the base station where mobile $i$ is assigned to and the mobile station $j$. The interference that user $j$ causes to the receiver of user $i$ is given by $G_{ij}P_j$ where $P$ is the $Q \times 1$ transmitter power vector of the active users. $G$ can be considered as a snapshot of the system. Figure 91 shows the uplink gain matrix.

![Figure 91: Link gain geometry](image)
The total interference experienced by any pair of base station and mobile is modelled as the sum of the powers of all the other active interferers. The background noise is negligible in a high capacity, interference-limited system. It is therefore assumed that the transmission quality to be dependent only on the carrier-to-interference ratio $C/I$. Using these notations the uplink $C/I$ for mobile $i$ in cell $k$ can be derived:

$$\Gamma_i = \frac{G_{ij}P_i}{\sum_{j \neq i} G_{ij}P_j}$$  

Equation 61

Mobile $j$ can be assigned to cell $k$ or to another cell. From Equation 61, $\Gamma_i$ can be expressed as:

$$\Gamma_i = \frac{P_i}{\sum_j P_j G_{ij}^{-1}} = \frac{P_i}{\sum_j P_j W_{ij}}$$  

Equation 62

where $W = \{W_{ij}\}$ is the normalised uplink gain matrix.

In the same manner an expression for the downlink SIR of mobile $i$ in cell $k$ can be formulated:

$$\tau_i = \frac{g_{ij}P_i}{\sum_{j \neq i} g_{ij}P_j}$$  

Equation 63

where $g_{ij}$ is the link gain on the path between mobile $i$ and the base station to which mobile $j$ is assigned at any time. $P_i$ in Equation 63, denotes the downlink transmitter power of the link between mobile $i$ and the base station of cell $k$. It can be further derived that:

$$\tau_i = \frac{P_i}{\sum_j P_j \frac{Z_{ij}}{Z_{ii}}} = \frac{P_i}{\sum_j P_j Z_{ij}}$$  

Equation 64

where $Z = \{Z_{ij}\}$ is the normalised downlink gain matrix:

$$Z_{ij} = \begin{cases} 1 & i \neq j \\ g_{ii} & i = j \end{cases}$$  

Equation 65
6.1.2 Channel Assumptions

Our study is concerned with power control schemes, which operate above the physical network layer. The aim of the algorithms investigated is the optimisation of system capacity and call quality. It is therefore assumed that the effects of motion-induced multipath fading (fast fading) are suppressed by means of synchronisation, rake receiver, modulation or the combined use of a fast closed-loop power control scheme. In order to form the uplink and downlink gain matrices only the slow variations in received signal strength due to path loss and shadowing are considered. The link gain $G_y$ is modelled as:

$$G_y = \frac{A_y}{d_y^\alpha}$$  \hspace{1cm} \text{Equation 66}

where $d_y$ is the distance between the base station of cell $k$, where mobile $i$ is assigned to, and mobile $j$. The $1/d^\alpha$ factor models the large-scale propagation loss whereas the factor $A_y$ models the power variation due to shadowing. At this stage it is assumed for all $A_y$ to be independent log-normally distributed random variables with 0dB expectation. It should be noted that assuming that $A_y$ are independent, implies that we do not take account of the decorrelation distance of the long-term fading. Parameter values of $\alpha$ in the range of 3-5dB and $\sigma$ in the range of 4-10dB usually provide good models for urban propagation. The downlink gain matrix $Z = \{Z_y\}$ is formed in a similar way.

The $C/I$ values given by Equation 61 and Equation 63 are local mean levels estimated over a certain averaging period. These measurements will more or less differ from the actual $C/I$ levels due to the limited suppression of fast fading. In order to analyse these effects the "corrupted" $C/I$ values can be used:

$$\Gamma'_i = n_i \cdot \Gamma_i$$  \hspace{1cm} \text{Equation 67}

where $n_i$ are all independent lognormal random variables with log-variance $\sigma_m$ in the range of 1-3dB. The limited suppression of Rayleigh fading will therefore deteriorate the system performance by increasing the local mean $C/I$ threshold.
6.2 Centralised Power Control

Centralised power control (or centralised $C/I$ balancing) implies the optimisation of system capacity is based on measurements of all radio links in the system. One such algorithm applicable to CDMA mobile systems has been investigated [31]. In the approaches discussed in [31], [32], [33], it is shown that the outage probability i.e. the probability that the uplink or downlink $C/I$ of a randomly chosen mobile station that is insufficient is minimised when $C/I$ balancing is applied. The focus of this chapter is however on the uplink.

**Theorem 1**: There always exists a unique maximum achieved mean uplink $C/I$ level $\gamma_U^*$ as expressed below:

$$\gamma_U^* = \frac{1}{\lambda^*}$$  \hspace{1cm} \text{Equation 68}

where $\lambda^*$ is the largest real eigenvalue of matrix $W$. The uplink transmitter power vector $P^*$ achieving this maximum is the eigenvector corresponding to $\lambda^*$ and $U$ denotes the uplink.

Theorem 1 provides an optimum solution to the problem of $C/I$ balancing of all users in a cellular system. Replacing the normalised uplink gain matrix $W$ by the normalised downlink gain matrix $Z$ in the above derivation, in the same manner the maximum achievable mean $C/I$ level $\gamma_D^*$ for the downlink and the corresponding transmitter power vector can be obtained.

The link gain measurements required for determining the BS transmitter powers are actually measured on the downlink through measurements of the individual BS pilot channels. Reciprocity between the uplink and the downlink forms a reasonable assumption only for the long-term channel variations (path loss and lognormal shadowing) and therefore these measurements cannot be used for the suppression of fast fading. However, an estimate of $\gamma_U^*$ and $\gamma_D^*$ could be of great importance for assessing the network load and determining the upper limits of QoS that a system is able to offer to all users. These estimates can be exploited in the context of an outer loop function that adjusts the target FER of the fast closed loop power control or used by a load control process which may adjust the bit-rates of the users according to the level of network congestion.

Attempting to estimate $\gamma_U^*$ and $\gamma_D^*$ would be difficult, since the path gain elements of the (time varying) link gain matrices are usually not known. Estimating these gains would require significant measurement effort. Even if this would be possible, the amount of data that would
have to be communicated and managed by the central controller would be enormous for any commercial network. The main contribution of Centralised Power Control theorem is the set the upper bound performance for any $C/I$ balancing.

### 6.3 Distributed Power Control

The main feature of distributed power control [34] is the adjustment of the transmitter power in each link on the basis, which does not require quality and path gain measurements on other links. A version of the distributed power control algorithm introduced in [33], which applies for CDMA cellular systems has been investigated. The only measurement required for the evaluation of the optimum transmitter power of each link is that of the received $C/I$.

#### 6.3.1 Zander’s Distributed Balancing Algorithm

This algorithm consists of three main steps:

**Step 1:** Set $P^{(0)} = P_0$, $P_0 \geq 0$. Set $\nu = 0$

**Step 2:** Set $P_i^{(\nu+1)} = \beta \cdot P_i^{(\nu)} \left(1 + \frac{1}{\Gamma_i^{(\nu+1)}}\right)$, $\beta > 0$ and $1 \leq i \leq Q$. \hfill (Equation 69)

**Step 3:** If $\nu < L$ repeat Step 2 otherwise stop.

The transmitter power at each iteration is derived from the previous iteration and the corresponding measured $C/I$ vector. It should be noted that this algorithm theoretically ($L \rightarrow \infty$) converges to the optimum $P$ vector and the achieved mean $C/I$ obtained when centralised $C/I$ balancing is employed.

The performance of Zander’s DBA was evaluated by means of simulation and numerical results are provided in the following subsection. It was observed that for limited suppression of the multipath fading (as specified in APPENDIX-B), this method fails to converge to an optimum $C/I$ vector as shown in Section 6.3.3. To eliminate non-convergence in the conventional algorithm, a time average balancing algorithm is used.

#### 6.3.2 Time Average Distributed Balancing Algorithm (TA-DBA)

The TA-DBA consists of three main steps

**Step 1:** Set $P^{(0)} = P_0$, $P_0 \geq 0$. Measure and store $C/I$ vector $\Gamma_i^{(0)}$. 

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Step 2: Operate Zander's DBA and set $p^{(v)} = P^{(v)}$ ($1 \leq v \leq L$) at each step. If $v \geq v_o$ then use Equation 70 as the transmitter power vector at the end of the iteration.

$$
P^{(v)} = \left\{ P_i^{(v)} = \left( \prod_{j \neq v} p_i^{(j)} \right)^{\frac{1}{\nu+1-v_o}}, \quad 1 \leq i \leq Q \right\} \quad \text{Equation 70}
$$

This technique demonstrated a relative robustness to mean $C/I$ estimation errors due to limited suppression of the fast fading and measurement errors.

### 6.3.3 Simulation Results

The performance of the distributed method was evaluated by means of simulation. The system was assumed to be initially completely unbalanced:

$$
P_i^{(0)} = 1, \quad 1 \leq i \leq Q \quad \text{Equation 71}
$$

The convergence of the uplink $C/I$ of a given user with the largest initial $C/I$ value and the user with the smallest initial $C/I$ values are shown in Figure 92-(a). Figure 92-(b) illustrates the convergence of transmitter powers to their optimum values.

![Figure 92: (a) Convergence of the maximum and minimum $C/I$ in Zander's DBA scheme and (b) convergence of the corresponding transmitter powers.](image_url)
Chapter 6 Distributed Power Control

It can be observed that convergence is achieved at about 20 iterations. If $C/I$ measurements are performed over an averaging period of e.g. 0.3 sec, then the convergence time is 6 sec.

In order to model the effects of limited suppression of multipath fading as well as measurement noise, the model described in Section 6.1.2 is used. This implies that the estimated SIR value used at each iteration will be given by:

$$\Gamma^{(v)}_i = n_i^{(v)} \cdot \Gamma^{(v)}$$

where $n_i^{(v)}$ represents all independent lognormal random variables with log-variance $\sigma_m$.

Figure 93-(a) and (b) show an example of the fluctuation of the received $C/I$ levels and transmitter powers respectively for $\sigma_m = 1$ dB. It can be observed that the conventional Zander's distributed balancing algorithm fails to converge to stable $C/I$ and transmitter power levels as pointed out in Section 6.3.1.

![Figure 93](image)

**Figure 93:** (a) Fluctuation of the maximum and minimum $C/I$ in Zander's DBA scheme and (b) fluctuation of the corresponding transmitter powers.

To eliminate non-convergence in the conventional algorithm, the TA-DBA discussed in the previous subsection was simulated. Figure 94-(a) and (b) show an example of the variations in the received $C/I$ level and transmitter power respectively with this scheme.
From the above, it can be observed that convergence is achieved in the presence of mean $C/I$ measurement errors.

As seen from the discussions in this Chapter, distributed balancing algorithms involve an estimation procedure performed in sequential iterations. Convergence is achieved if the link-gain matrix (containing the path loss and log-normal shadowing variations), remain approximately constant during the estimation procedure. However, this may not be the case especially for fast moving mobiles. The time between two subsequent iterations is determined by the averaging period over which the mean $C/I$ values are measured. For a fast moving mobile the averaging period required for fast-fading suppression is relatively short and the total convergence time needs to be short. Therefore, a more effective technique could attempt to relate the length of the averaging period with the speed of each mobile i.e. the second order characteristics of fast fading as discussed in Chapter 4.

Another practical problem is that the transmitter powers in the distributed balancing algorithm will all be increasing, unless parameter $\beta$ (Equation 73) is appropriately selected.

$$\beta = \beta(\nu) = \frac{1}{[P^{(\nu)}]}$$  \hspace{1cm} \text{Equation 73}

Selecting an appropriate value for Equation 73 would ensure a "constant" average power level. However, calculating this quantity may not be possible in a completely distributed concept since it would require knowledge of the power levels in all links.
6.4 Removal Power Control Algorithm

6.4.1 Stepwise Removal Algorithm (SRA)

If the system-wide balanced C/I stabilises below the required threshold for mainlining the communication link $y_0^*$, C/I balancing of all radio links can be detrimental. Therefore, C/I balancing can be used in combination with an algorithm, which removes as “few” mobiles as possible in order for $y_0^*$ to be achieved.

A “brute force” search would first check if $y_0^*$ is achievable for the original matrix $W$. If not the algorithm would try to remove one mobile station computing the Eigen values of each reduced system until the C/I requirements are fulfilled. If not the algorithm would try removing all combinations of 2 mobile stations and so on. Whilst such an algorithm could produce optimum results, its complexity would be beyond practical.

An alternative lower complexity approach is to remove one mobile at a time until the required C/I level is achieved for the remaining mobiles. Knowing that the row and column sums of a matrix provide bounds on the dominant eigenvalue of matrix $W$, a reasonable approach [31] would the following algorithm:

Step 1: Determine the achievable C/I level $y_0^*$ corresponding to $W$. If $y_0^* \geq y_o^*$ use the eigenvector $P^*$ corresponding to $W$, else set $Q' = Q - 1$ and perform Step 2.

Step 2: Remove mobile $k$ for which the maximum of the row and column sums

$$\sum_{j=1}^{Q} W_{kj}, \sum_{l=1}^{Q} W_{lk}$$

is maximised, then form the $(Q' - 1) \times (Q' - 1)$ matrix $W'$. If $y_0^* \geq y_o^*$ use the eigenvector $P^*$, else set $Q' = Q' - 1$ and repeat Step 2.

6.4.2 Limited Information Stepwise Removal Algorithm (LI-SRA)

The limited information removal algorithm introduced in [33] was investigated. This is clearly a centralised procedure based on the knowledge of the C/I values in each link in the cellular system. The amount of information that needs to be processed for the removal of users is, however, much smaller in comparison with the SRA.

Step 1: Set $P=1$. Measure and store the C/I vector $\Gamma_i^{(0)}$. If $\Gamma_i^{(0)} > y_o^*$ for all $i$ stop; otherwise,
Step 2: Operate the distributed balancing algorithm for at most \( L \) iterations. If at some step 
\[ \gamma_i^{(\nu)} > \gamma_o \text{ for all } i, \] stop, otherwise

Step 3: Remove the cell \( i \) that has the smallest initial \( C/I, \Gamma_i^{(0)} \). Go to step 1.

6.4.3 Simulation Results

The performance of centralised \( C/I \) balancing in terms of achieved capacity in a simple cellular system (APPENDIX-B) using SRA was evaluated by means of simulation. The configuration used for this test environment included \( N=9 \) cells, lognormal fading with \( \sigma=8 \text{dB} \), path loss exponent \( \alpha=4 \) and uplink required mean \( C/I \) threshold \( \gamma^*_o = -14.76 \text{dB} \). \( \gamma^*_o \) was obtained assuming \( W/R=150 \) and \( (E_b/I_o)_{\text{req}} = 7 \text{dB} \).

The achievable mean \( C/I \) level was determined for a number of mobiles ranging from \( Q=189 \) (21 mobiles /cell) to \( Q=117 \) (13 mobiles/cell), for 500 independent \( \mathbf{W} \) configurations. Assuming that different link gain matrix configurations are independent, time-dependent of second-order statistics of the slow fading, were not taken into account.

![Figure 95: Outage probability comparison for the SRA and LI-SRA schemes.](image)

Figure 95 shows the outage probability versus the number of mobiles/cell. It can be observed that only a small number of mobiles could cause a significant reduction in the outage probability. The achieved SRA capacity for \( P_{\text{out}} = 2\% \) is 18 mobiles/cell.

The performance of LI-SRA was evaluated using a similar simulation procedure. As it can be seen from Figure 95 the capacity achieved for \( P_{\text{out}} = 2\% \) is 16 mobiles/cell. By comparison of the two algorithms, complexity-performance trade-offs can clearly be seen.
Implementation of SRA could prove very difficult as similar to the case of optimum power control the amount of the required link-path information could prove overwhelming. LI-SRA offers a remedy to that, since the central controller simply needs to sort the initial $C/I$ values of users. On the other hand a basic assumption for LI-SRA is that transmitter powers are initially completely unbalanced, which may not hold in a real system situation. Further versions of this algorithm with slightly better performance are presented in [35]

## 6.5 Soft Dropping Power Control Algorithm

### 6.5.1 Distributed Power Control and the “Party Effect”

The algorithms described in Section 6.4 are based on quality information of the link of interest and do not require knowledge of the link-gain matrix. However, in order to achieve stability, each mobile station should have some knowledge of the transmitter powers of other users as discussed in Section 6.2. A fully distributed quality-based adjustment [36] of transmitter power is described below:

$$P^{(n+1)}_i = \frac{\Gamma_{\text{arg}}^{(n)}}{\gamma_i^{(n)}} P^{(n)}_i$$  \hspace{1cm} \text{Equation 74}

where $P^{(n+1)}_i$ and $P^{(n)}_i$ are the average transmitter powers of the link at step $n+1$ and $n$ respectively, $\gamma_i^{(n)}$ is the measured $C/I$ of the link at step $n$ and $\Gamma_{\text{arg}}$ is the target $C/I$. The main shortcoming of this simple algorithm is the need to assign an appropriate $\Gamma_{\text{arg}}$ that all users will strive towards. If this value is set too high, then it is not possible to support all the mobiles, and the “party-effect” may be experienced in the network. This occurs when one user increases its power and thereby the interference at other receivers. These users may react by increasing their powers which will increase the interference at the first user, who will find it necessary to increase its power and so on. If the transmitter power is bounded to the above by a maximum level determined by the physical limits of the system, some mobiles may increase their powers to this maximum, without achieving the specified target $C/I$.

### 6.5.2 The AAW Algorithm

An attempt to counteract the party effect and to employ graceful degradation in the system was proposed by Almgren, Andersson and Wallstedt in [37]. The main idea is that a user requiring a high transmission power has to accept a lower quality. The target $C/I$ of each user $\Gamma_{\text{arg}(i)}^{(q)}$ is not fixed and varies in time and from user to user. $\Gamma_{\text{arg}(i)}^{(q)}$ is a linear function of the average transmitter power of the user.
Chapter 6 Distributed Power Control

\[
\left( P_{\text{transmit}(t+1)} \right)_{dB} = k \left( P_{\text{transmit}(t)} \right)_{dB} + m
\]  
\text{Equation 75}

where \( k < 0 \) is the slope of the degradation and \( m \) is an offset. Equation 74 can now be written as:

\[
\left( P_{\text{transmit}(t+1)} \right)_{dB} = k \left( P_{\text{transmit}(t)} \right)_{dB} + m - \left( P_{\text{transmit}(t)} \right)_{dB} + \left( P_{\text{transmit}(t)} \right)_{dB}
\]  
\text{Equation 76}

and

\[
\left( P_{\text{transmit}(t+1)} \right)_{dB} = a \cdot \left( P_{\text{transmit}(t)} \right)_{dB} - \beta \cdot \left( P_{\text{transmit}(t)} \right)_{dB}
\]  
\text{Equation 77}

where

\[
a = \frac{m}{1 - k} \quad \text{and} \quad \beta = \frac{1}{1 - k}
\]  
\text{Equation 78}

The transmit power in the next iteration is only based on the current transmitted power and the current \( C/I \), therefore the AAW method is a fully distributed algorithm.

6.5.3 Soft Dropping Power control

In [36] a more realistic framework allowing for graceful degradation according to transmit power levels, is discussed. In this work the uplink target \( C/I \) value ranges from a maximum value for ideal call quality to a minimum for just acceptable QoS. The graceful degradation of the \( C/I \) is termed \textit{Soft Dropping}. The parameter which quantifies QoS in [37] is the received \( C/I \). In this scenario \( \Gamma_{\text{transmit}(t)} \) is varied according to the users transmitter power as depicted in Figure 96.

![Figure 96: Target C/I of each MS according to its transmitter power](image)
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At all times, user $i$ aims for a target $C/I$ that is above a dropping threshold $\Gamma_T$. The target $C/I$ at time instant $t$ varies according to:

$$
\begin{align*}
\left(\Gamma^{(i)}_{\text{target}}\right)_{\text{dB}} &= \begin{cases} 
\left(\Gamma^{(i)}_{\text{max}}\right)_{\text{dB}} & P^{(i-1)}_i < P_u \\
\left(\Gamma^{(i)}_{\text{max}}\right)_{\text{dB}} + \delta \left( P^{(i-1)}_i - P_u \right) & P_u \leq P^{(i-1)}_i \leq P_d \\
\left(\Gamma^{(i)}_{\text{min}}\right)_{\text{dB}} & P^{(i-1)}_i > P_d
\end{cases}
\end{align*}
$$

Equation 79

where

$$
\delta = \frac{\log(\Gamma^{(i)}_{\text{max}}/\Gamma^{(i)}_{\text{min}})}{\log(P_u/P_d)}
$$

Equation 80

This approach can offer a significant gain in capacity as compared to a fixed target algorithm where $\Gamma^{(i)}_{\text{target}} = \Gamma_{\text{max}}$ and if $\Gamma_{\text{max}} > \gamma^*_U$, where $\gamma^*_U$ is the maximum achievable uplink $C/I$ for all users as defined in Section 6.2.

6.5.4 Simulation Results and Discussion

The performance of AAW algorithm was evaluated by means of simulation. The test environment of APPENDIX-B with the following configuration was used:

$N=9$ cells, lognormal fading with $\sigma=8$dB and path loss exponent $\alpha=4$.

The algorithm was applied for 500 independent $W$ configurations. Measurements of the link quality and transmitter power of each user, were taken after a sufficient number of iterations. The CLPC dynamic range was assumed to be infinite and the power control to be “perfect” i.e. no control delays or errors. No user was blocked due to unacceptable quality.

A number of $C/I$ and transmitter power cumulative density functions corresponding to different values of $\beta$ are shown in Figure 97. The parameter $\alpha$ in Equation 79 controls the mean value of the transmitter power and $\beta$ controls the system’s spread in transmitter power and $C/I$. Higher values of $\beta$ will lead to lower spread in the $C/I$ distribution and higher spread in the power distribution. When $\beta = 1$ which corresponds to $k=0$ in Equation 79 all connections will strive towards the same $C/I$. 

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The above results offer an effective mechanism for observing the system behaviour when graceful degradation of call quality is applied. However, certain parameters which would affect a real system's capacity and QoS such as cell radius, power control dynamic range and the limits within which $C/I$ can vary, have not been taken into account.

It is important to point out that by nature, soft dropping power control does not take the system load and traffic distribution into account. Therefore, probability of system instability is decreased but not eliminated. In the case of light traffic, this approach can lead to an unjustifiable reduction of call quality. This corresponds to the situation that the path gain between a user and the serving cell site is relatively low due to shadowing or a large separation distance. The user will require a high transmit power and consequently will be forced to decrease its target $C/I$. However, this would not be due to high levels of interference and therefore call quality would be degraded without any improvement in the overall system performance.

The above effects are better observed in next section whereby transmitter power is proven not to form an optimum criterion for call quality degradation.
6.6 A Novel Distributed Power Control Algorithm

Setting the target QoS of users too high can lead to instability of the transmitter powers resulting in decreased capacity and an unjustifiable increase in power consumption. A remedy to that could be to set the quality requirements of users at the lowest possible level for just acceptable quality of service. This on the other hand endangers the quality of a link.

The aim of this section is to develop an algorithm which adapts the quality requirements of users according to the network load, the distribution of users in the system coverage area and the channel conditions of each user. This reflects realistic CDMA conditions under which, QoS can be adapted to varying capacity requirements.

An important feature of any such algorithm would be its practicality. On that note the desired algorithm could operate with path gain information within a single cell only, a half way house between distributed and centralised schemes.

6.6.1 Evaluation of the maximum achievable uplink C/I

Zander in [31] introduces an evaluation model of the maximum achievable $C/I$, applicable to systems where only intercell co-channel interference is present (FDMA/TDMA systems). In this section an analytical approach for evaluating the maximum achievable uplink $C/I$ for each user in a CDMA system with a frequency reuse factor of one is used.

The path link, power and quality measurements that will be used in this section are assumed to be mean values affected only by the long-term variations of the fading channels (path loss and shadowing). Rayleigh fading is assumed to be averaged out. For this argument to be valid, averaging distances of 20 to 80 wavelengths are required as discussed in Chapter 3.

Let $G$ denote the $N \times N$ uplink gain matrix, where $N$ is the number of active users in the system and $G_y$ is the link gain on the path between the base station where mobile $i$ is assigned to and the mobile station $j$ as shown in Figure 91 of Section 6.1.1. The interference that user $j$ causes to the receiver of user $i$ is given by $G_y P_j$, where $P$ is the $N \times 1$ transmitter power vector of the active users. $G$ can be considered as a snapshot at a given moment in the system. Therefore, the uplink received $C/I$ of user $i$ can be expressed as:

\[
CIR_i = \frac{G_y P_i}{\sum_{j \neq i} G_y P_j}
\]

Equation 81

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where the background noise was considered to be negligible. The above equation can be written as:

$$CIR_i = \frac{P_i}{\sum_{j=1}^{N} P_j W_{ij}}$$  \hspace{1cm} \text{Equation 82}$$

where $W$ is the $N \times N$ normalised uplink gain matrix defined by:

$$W_{ij} = \begin{cases} 
  G_{ij} & i \neq j \\
  G_{ii} & i = j 
\end{cases}$$  \hspace{1cm} \text{Equation 83}$$

The maximum achievable uplink $C/I$ that can be simultaneously achieved by all users was defined in Section 6.2. The uplink transmitter power vector achieving this maximum is the eigenvector of $W$ corresponding to $\lambda^*$, denoted by $v^*$. From (Equation 81) it can be observed that in the absence of background noise there exists an infinite number of optimum power vectors achieving $\gamma_U^*$. Any power vector $P^*$, which satisfies the following equation, is an optimum vector:

$$P^* = av^*, \quad a > 0$$  \hspace{1cm} \text{Equation 84}$$

The maximum optimum power vector $P_{\text{MAX}}^*$ is constrained by the maximum transmitter power constraint $P_{\text{max}}^*$:

$$P_{\text{MAX}}^* = v^* \frac{1}{\max(v^*)} P_{\text{max}}$$  \hspace{1cm} \text{Equation 85}$$

In the presence of background noise $n_s$, the maximum achievable $C/I$ for user $i$ is:

$$\gamma_{U(i)}^* = \frac{G_{ii}P_{\text{MAX}(i)}^*}{\sum_{j=1}^{N} G_{ij}P_{\text{MAX}(j)}^* + n_s}$$  \hspace{1cm} \text{Equation 86}$$

However, empirical evidence has shown that in real systems considering $\gamma_U^*$ given by Equation 64 and $\gamma_{U(i)}^*$ given by Equation 86 as equal, is a realistic assumption.
Since, $W$ is a snapshot of the system and contains path loss and shadowing information of all the links in the system, the maximum achievable $C/I \gamma_U^*$ also varies with time. This was estimated every 1 sec for the Outdoor to Indoor and Pedestrian Test Environment described in APPENDIX-B. The speed of mobile users was 3km/h and their distribution was assumed uniform in the streets of the Manhattan-like urban model. The simulation was performed for a system load of 20 users/carrier/cell. The $\gamma_U^*$ samples corresponding to 1 hour of virtual simulation time are plotted in Figure 98. It can be observed that despite the low speeds, $\gamma_U^*$ varies quite rapidly within a range of approximately 2.5 dB. This demonstrates that the maximum achievable $C/I$ for all users should not be considered as a function of only the system load. Therefore, it would be beneficial to employ an algorithm, which "tracks" $\gamma_U^*$ or even adjusts the target $C/I$ in consistent way. Let $CIR_i^{(t)}$ denote the measured average $C/I$ of user $i$ at time instant $t$. On that note, it is desired to determine whether the quality requirements of user $i$ should be increased or decreased in the context of a network which provides the best possible quality of service to all users.

![Figure 98: Variations of the maximum achievable C/I versus time in the Pedestrian Test Environment with a system load of 20 users/carrier/cell.](image)

This decision should be ideally based on the knowledge of $\gamma_U^*$. The information required for the evaluation of $\gamma_U^*$ consists of the uplink path gains between all mobile stations and all base stations ($N_{MS} \times N_{BS}$ path gains, where $N_{MS}$ is the number of mobile users and $N_{BS}$ is the number of base stations). Even if the local path link measurements from a set of cells around user $i$ are sufficient, the amount of information that has to be exchanged between BSCs and MSCs
could be enormous. The processing power required, could also prove to be intolerable within the
time limitations imposed by the mobility of users. Thus, problem needs to solved in a more
decentralised manner.

6.6.2 Near-Optimum Power Control Algorithm

The algorithm discussed in this section attempts to provide a near-optimum criterion for assessing
whether the target signal quality of user $i$ should be increased or decreased to allow for an
improved overall network performance. This decision will be based on information concerning
only the links of users assigned to cell $c_i$ where user $i$ is located. Therefore, no link-path
information has to be exchanged between the base stations. Once again, the path link, power and
quality measurements used are assumed to be mean values affected only by the long-term
variations of the fading channels. Rayleigh fading is assumed to be averaged out.

Let $G^{c_i}$ and $W^{c_i}$ denote the uplink gain matrix and the normalised uplink gain matrix of the
one-cell system $c_i$, and $P_{c_i}$ denote the uplink transmitter power vector of users of cell $c_i$. $W^{c_i}$
can be defined if the link gains between the base station in $c_i$ and all the mobile stations in this
cell are known. If the uplink measured intercell interference at cell $c_i$ is denoted by $I_{intercell}^{c_i}$ and
$N^{c_i}$ is the number of intracell users, the average $C/I$ of user $i$ can be expressed as:

$$CIR_i = \frac{G^{c_i}_{ii} P_{c_i}^{i}}{n_{intercell} + I_{intercell}^{c_i} + n_s}$$  \hspace{1cm} \text{Equation 87}

In the preceding analysis, the index of the user of interest was assumed to be $i$ in both the sets of
$N$ users in the system and $N^{c_i}$ users in cell $c_i$.

Under the condition that only uplink path gain measurements of the users in cell $c_i$ are available,
intercell interference will be treated as background noise. However this contribution cannot be
treated as negligible. In the following section an attempt is made to provide a $C/I$ estimate
based on the optimisation of transmitter powers of intracell users and subject to the effect of
intercell interference. Let $\lambda^{c_i}$ denote the largest eigenvalue of $W^{c_i}$ and $v^{c_i}$ the corresponding
eigenvector. The reference $C/I$ used to provide a criterion for the increase or decrease of $CIR_i$
is given by:
Chapter 6 Distributed Power Control

\[ \gamma_{\text{REF}} = \frac{G_i^0 P_i^{5^*}}{\sum_{j=1}^{N_i} G_j^0 P_j^{5^*} + I_{\text{intercell}}^5 + n_z} \]  

Equation 88

where \( P_i^{5^*} \) is given by

\[ P_i^{5^*} = a^c v^* \quad , \quad a^c > 0 \]  

Equation 89

The selection of parameter \( a^c \) is of great importance. Since \( \gamma_{\text{REF}} \) needs to reflect the current network interaction between users, an efficient selection for \( a^c \) is:

\[ a^c = \frac{1}{\max(v^{5^*})} \max(P^5) \]  

Equation 90

where \( P_i^{5^*} \) is the measured vector of average transmitter powers of intracell users. The above value for \( a^c \) provides a method for weighting the intracell optimum power vector \( P_i^{5^*} \) against the currently measured \( I_{\text{intercell}}^5 \). It should be noted that \( \gamma_{\text{REF}} \) is equal for all the intracell users since the optimum vector itself provides received power balancing in the context of a one-cell system.

The algorithm could therefore operate as follows:

**Step 1:** Cell sites measure the average uplink received powers and average \( C/I \) of intracell users. Each mobile station measures the average power of a pilot signal transmitted on the downlink of the active base station, compares that with the fixed pilot transmitter power provided by the downlink broadcast channel and determines the link gain (reciprocity is assumed between the uplink and the downlink). The link gains are reported to the base station and the transmitter powers of users are estimated.

**Step 2:** \( \gamma_{\text{REF}} \) is estimated separately for each base station.

**Step 3:** The average \( C/I \) of each user is compared with the \( \gamma_{\text{REF}} \) of the active base station. If \( CIR_i > \gamma_{\text{REF}} \) the target FER of the user is decreased otherwise it is increased.
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Figure 99: Operation principle of the algorithm

The estimate $\gamma_{REF}$ provides an efficient criterion for assessing whether the transmitter powers of intracell users are optimum with respect to each other or not. However, intercell interference is treated as background noise. This can cause instability since the cells that suffer higher intercell interference decrease their quality requirements, while the other cells increase their target quality of service. Consequently, users in suffering cells are forced to reduce their transmitter powers and users in other cells are forced to transmit with higher powers leading to an even more unbalanced situation. A sensible approach for mitigating the effect described above is the selection of the minimum $\gamma_{REF}$ from the neighboring set of cells of each user, as the reference value used in Step 3. Since the $\gamma_{REF}$ of problematic cells will generally be lower, the algorithm provides a near-optimum way of adjusting the quality requirements of each user accordingly not only to the intracell interference but also to the interference that adjacent cells suffer.

In this approach, each base station has to be informed of the $\gamma_{REF}$ of its adjacent cells in order to determine whether the target QoS of each user should be increased or decreased. However, no link-path gain information needs to be exchanged between base stations. Another feature of the method is that the target $C/I$ is measured rather than imposed. Therefore, the algorithm could be realized as a part of the process that maps the average $C/I$ or SIR on the required FER or
BER (as described in Section 2.5.2). This can be achieved by adjusting the outer loop power control parameters such as the range within which the target SIR of fast closed loop power control can vary, or the target frame error rate. These parameters are exchanged via Layer 3 messages.

An example of the operation principle is shown in Figure 5.1 in which it is assumed that omni directional base station antennas are employed in a hexagonal cell layout. Let $CIR_i$ denote the measured average $C/I$ level of user $i$ in cell 1 (centre cell). It is now important to determine whether the target signal quality of this user should be increased or decreased. First the measurements listed in Step 1 of the algorithm are performed at each cell-site. Secondly the $\gamma^{(k)}$ values, $1 \leq k \leq 7$, are estimated for each cell-site (Step 2). Those values are then reported to a Central Controller (CC). The CC decision (Step 3) is based on the comparison of $CIR_i$ with the minimum $\gamma^{(min)}_{\text{ref}}$, $\gamma^{(min)}_{\text{ref}} = \min(\gamma^{(1)}_{\text{ref}}, \ldots, \gamma^{(7)}_{\text{ref}})$.

### 6.6.2.1 Convergence of the Algorithm

The $C/I$ balancing properties of the algorithm are investigated in this section. At each time instant the algorithm is demonstrated to lead to a more balanced system state. Indeed, if the intercell interference caused by a mobile station to its adjacent cells is decreased according to the criterion imposed on Step 3 of the algorithm, then $\gamma^{(min)}_{\text{ref}}$ is likely to increase, since the intercell interference term in the denominator of Equation 88 will decrease. In the case a mobile station is forced to increase its signal quality, then higher transmit power is likely to result in a smaller $\gamma^{(min)}_{\text{ref}}$. These observations can lead us to the conclusion that during the operation of the algorithm and in the absence of fading and path loss variations, the absolute difference between each user’s $C/I$ and the $\gamma^{(min)}_{\text{ref}}$ corresponding to the set of neighbouring cells of the user, gradually decreases with time.

In order to demonstrate the convergence of the algorithm, 50 users were distributed in the hexagonal cell layout of the Vehicular Test Environment Model described in APPENDIX-B. Omni directional base station antennas are deployed that all users are located in the area corresponding to seven hexagonal adjacent cells. $\gamma^{(k)}_{\text{ref}}, \ 1 \leq k \leq 7$, is estimated separately at each cell site, and the algorithm is invoked on the basis of the minimum $\gamma^{(min)}_{\text{ref}}$ value of the cells. It is considered that all users are “Stationary” and therefore the link gain matrix of the system is
assumed not to vary with time. Power control is performed in subsequent iterations and the transmitter power $P_i^{(n+1)}$ of each user $i$ is adjusted at iteration $n+1$ according to:

$$P_i^{(n+1)} = \frac{\Gamma_{\text{targ}(i)}^{(n+1)}}{\gamma_i^{(n)}} P_i^{(n)} \tag{Equation 91}$$

where $\gamma_i^{(n)}$ is the measured $C/I$ of the link at iteration $n$ and $\Gamma_{\text{targ}(i)}^{(n+1)}$ is the target $C/I$ of user $i$ at iteration $n+1$. The outer loop function which adjusts $\Gamma_{\text{targ}(i)}^{(n+1)}$ is invoked every 10 iterations. $\Gamma_{\text{targ}(i)}$ of each user is increased or decreased using a step of 0.2 dB. The initial values of the target $C/I$ levels of the users are randomly distributed in the range from -6 to -12 dB. Every 10 power control iterations, the $C/I$ of each user is recorded. Those values were plotted versus the number of preceding outer loop commands. Results are shown in Figure 5.3.

![Figure 100: Convergence of the received C/I of all users due to the constraints imposed by the algorithm](image)

From the above, it can be observed that after a sufficient number of outer loop commands, the imposed constraint on $\Gamma_{\text{targ}(i)}$ of each user forces the received $C/I$ to fluctuate around a fixed value.

Full evaluation of the this algorithm’s efficiency in comparison with other distributed power control schemes, is presented in the next section.
6.6.3 Simulation Approach

The system coverage area is considered rectangular with a size of $30\text{km} \times 30\text{km}$ and $2.5\text{km} \times 2.5\text{km}$ for the Vehicular and Pedestrian Environments respectively. In the Pedestrian environment it was observed that the interference experienced by a user was mainly due to the transmissions of mobile stations located along the same street with the user. The variable-target algorithms employed are adaptive to interference and therefore can perform better in comparison with fixed-target algorithms, when some cells suffer less interference. In order to decrease this influence, which could result in optimistic results, statistics are collected from the innermost cells.

The received signal strength in dB is calculated as the emitted power minus the path loss between the base station and the mobile station plus the base station antenna gain. The latter is $13\text{dBi}$ for the Vehicular Environment and $10\text{dBi}$ for the Pedestrian Environment (cable and connector losses of 2dB are also considered). The path loss is modelled as the sum of two terms, one due to the distance and one due to the lognormal shadow fading as stated in APPENDIX-B. Rayleigh fading is assumed to be averaged out and hence not considered.

The system performance is evaluated in terms of service quality and outage probability. The outage probability is defined as the probability that a user has a signal quality below a certain threshold. It is assumed that the criterion upon which call quality depends is the received mean $C/I$. This assumption does not always hold for digital radio systems since it is well known that a certain SIR or $C/I$ can correspond to different values of frame error rates according to the channel conditions. However, it can be claimed that a higher target frame error rate in general requires a higher mean received $C/I$. The algorithms under investigation are concerned with determining the target $C/I$ of each user at each time instant. In doing so, it is assumed that at all times each user aims for a target $C/I$ above a threshold $CIR_T$ for just acceptable signal quality. In order to trade-off between QoS and system capacity, the target of each user is considered to vary within a range defined by a maximum $CIR_{MAX}$ for ideal quality and a minimum $CIR_{MIN}$. The latter is set higher than $CIR_T$ to allow for a margin against degradation due to channel variations. The values chosen for the above parameters for the two Test Environments are shown in the following table.
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<table>
<thead>
<tr>
<th>Test Environment</th>
<th>CIR&lt;sub&gt;T&lt;/sub&gt;</th>
<th>CIR&lt;sub&gt;MIN&lt;/sub&gt;</th>
<th>CIR&lt;sub&gt;MAX&lt;/sub&gt;</th>
</tr>
</thead>
<tbody>
<tr>
<td>Vehicular</td>
<td>-20.7 dB</td>
<td>-19.0 dB</td>
<td>-16.0 dB</td>
</tr>
<tr>
<td>Pedestrian</td>
<td>-22.3 dB</td>
<td>-20.7 dB</td>
<td>-17.7 dB</td>
</tr>
</tbody>
</table>

*Table 8: C/I thresholds for the employed Test Environments*

The threshold value $CIR_T$ was chosen on the basis of the link-level simulations results provided in [38]. In this work, the required $E_b/N_o$ corresponding to a BER of $10^{-3}$ for speech (information rate $R_b = 8kbps$) is evaluated at $(E_b/N_o)_{req} = 6.4dB$ for the Vehicular Test Environment and at $(E_b/N_o)_{req} = 4.8dB$ for the Pedestrian Test Environment. The $E_b/N_o$ values include all overhead and $E_b$ contains all energy needed to transmit one information bit. The corresponding $C/I$ values are derived by the well-known formula:

$$\frac{C}{I} = \frac{E_b}{N_o} \frac{R}{W}$$

Equation 92

where $W/R$ is the processing gain. Considering an information bit rate of $8kbps$ and a chip rate of $4.096Mcps$, the processing gain becomes $W/R = 512$.

In [38] power control is assumed to be perfect. In order to simulate signal degradation due to imperfect power control the conclusions of [39] are considered. This approach suggests that $E_b/N_o$ is log normally distributed and that the standard deviation for the narrowband CDMA employing both open loop and fast closed-loop power control is between 1.5 and 2.5 dB. Wideband CDMA reduces the impact of imperfect power control due to improved diversity. Therefore, in the proposed approach a standard deviation of $1dB$ is used.

According to [1] a satisfied user is one that has sufficiently good quality (BER=$10^{-3}$) for more than 95% of the session time. Considering a lognormal variation of $E_b/N_o$ about the mean with a standard deviation of $1dB$, a confidence level of $1.64dB$ above $(E_b/N_o)_{req}$ should be used. $CIR_{MIN}$ corresponds to that value of $E_b/N_o$, therefore $CIR_{MIN}$ is set at least $1.64dB$ above $CIR_T$. The outage probability is defined as:

$$\text{Outage Probability} = \text{Probability}(E_b/N_o < (E_b/N_o)_{req})$$

Equation 93

The maximum average transmit power constraints considered are $30dBm$ and $14dBm$ for the Vehicular and Pedestrian Test Environments respectively [1]. The uplink power control dynamic range specified in [38] is $80dB$. This value is used by the fast closed loop power control
attempting to eliminate the near-far effect (long-term channel variations) as well as the fast fading. Since the proposed algorithm is concentrated on the long-term channel variations a dynamic range of 50 dB is assumed.

Transmitter powers of users are adjusted at a rate of 100 iterations/sec. The transmit power $P_i^{(n+1)}$ of each user $i$ at iteration $n+1$ is determined by:

$$
P_i^{(n+1)} = \frac{\Gamma_{\text{tag}(i)}^{(n+1)}}{\gamma_i^{(n)}} P_i^{(n)}
$$

Equation 94

where $\gamma_i^{(n)}$ is the measured $C/I$ of the link at iteration $n$ and $\Gamma_{\text{tag}(i)}^{(n+1)}$ is the target $C/I$ of user $i$ at iteration $n+1$. The algorithm described in Chapter 4 is invoked every 10 iterations. The target $C/I$ of each user is increased or decreased using a step of 0.5 dB. The performance of the algorithm is tested in comparison with two fixed-target methods whereby $\Gamma_{\text{tag}(i)}$ is set to $CIR_{\text{MIN}}$ and $CIR_{\text{MAX}}$. For comparison purposes, soft dropping power control is also simulated. The target $C/I$ of each mobile station is varied according to its transmitter power as depicted in Figure 101.

![Figure 101: Target C/I of each MS according to its transmitter power](image)

The parameters of Soft Dropping power control used in the two Test Environments are shown in the following table:

<table>
<thead>
<tr>
<th>Test Environment</th>
<th>$P_u$</th>
<th>$P_d$</th>
<th>$\delta$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Vehicular</td>
<td>5 dBm</td>
<td>30 dBm</td>
<td>3/25</td>
</tr>
<tr>
<td>Pedestrian</td>
<td>2 dBm</td>
<td>24 dBm</td>
<td>3/22</td>
</tr>
</tbody>
</table>

Table 9: Soft Dropping parameters for the employed Test Environments
At each simulation run, the number of mobile stations is provided as an input. No mobile is dropped during the call due to bad link even if this implies a very large outage probability.

The employed handover scheme is based on measurements performed by the mobile station. If a terminal finds that the path gain of a neighbouring base station is $h$ dB higher than that the user is presently connected to, it will attempt an intercell handover. The hysteresis parameter $h$ is set to $h=5dB$ for both the Test Environments. The simulated scheme (Mobile Assisted Handover) provides hard handover options. Consequently, the capacity figures provided in the following section are pessimistic in the sense that the levels of interference would be lower if a soft handover scheme was employed.

### 6.6.4 Simulation Results

The proposed near-optimum algorithm was evaluated in conjunction with two fixed target methods with target $C/I$ equal to $CIR_{MIN}$ and $CIR_{MAX}$, and the soft dropping algorithm. The performance of the algorithms were evaluated in terms of mean $C/I$ (for all users) and outage probability. Both of those measurements are necessary system efficiency assessment, since:

- a high average $C/I$ does not automatically imply acceptable quality for a sufficient fraction of users
- and a low outage probability does not always take account of a large percentage of users operating with just acceptable quality whilst radio resources (transmitter power and bandwidth) are not fully exploited.

Simulations were carried out in both the Vehicular and Pedestrian Test Environment for a range of system loads (users/MHz/cell). The spread bandwidth is assumed to be 5 MHz [1]. Omni directional BS antennas were used for both the Test Environments. This assumption applies to the micro cellular Environment only which provides sufficient cell separation. In the case of the macro cell (Vehicular) environment, system load figures provided in this section do not correspond to the actual achieved capacities since tri-sectored cells are nominally used.

A number of $C/I$ cumulative density functions corresponding to different values of system loads for the Vehicular Test Environment are shown in Figure 102-Figure 104.
Figure 102: CDF plots of mean C/I for the Vehicular Environment with a system load of 2 users/MHz/cell

Figure 103: CDF plots of mean C/I for the Vehicular Environment with a system load of 3 users/MHz/cell
Chapter 6 Distributed Power Control

Figure 104: CDF plots of mean C/I for the Vehicular Environment with a system load of 3.5 users/MHz/cell

As seen from Figure 101, in the case of a light traffic (2 users/cell/MHz) it can be observed that the outage probabilities achieved by all the employed methods are generally acceptable (lower than 5% required for sufficiently good quality in UMTS). The proposed method achieves C/I levels slightly lower than the fixed target method using the ideal target quality, whilst soft dropping exhibits a small degradation. The fixed target method using the minimum C/I has a poor performance as compared to the other schemes.

As the traffic load increases (3 users/MHz/cell) it is observed (Figure 102) that the fixed target algorithm aiming at a too high quality significantly deteriorates the performance of the users and hence the outage probability is increased. It can also be observed that the soft dropping method is relatively immune to the increase of the load and that the outage probability is not significantly increased as compared to the light-traffic case. The proposed near-optimum algorithm is also proved to provide an efficient scheme for reducing the quality requirements of users resulting in an acceptable outage probability.

If the system load is increased furthermore (3.5 users/MHz/cell), the fixed target method aiming at an ideal call quality and the soft dropping algorithm become very inefficient. This is due to the fact that the "party effect" has lead to instability of the transmitter powers and an increased variation in the levels of received C/I in the system. This unbalanced situation is reflected on the corresponding CDF plots shown in Figure 103, where it can be observed that the achieved C/I is "spread" as compared with the previous two Figures. In the case of Soft Dropping power control it is demonstrated that transmitter power does not comprise an optimum criterion.
for assessing the quality target of users. The proposed algorithm succeeds in reducing the target $C/I$ of users and the outage probability is not considerably degraded. Furthermore, as demonstrated in Figure 106, signal quality is superior to the one achieved by the fixed target algorithm aiming at the lowest acceptable QoS.

![Figure 105: System performance in terms of outage probability- Vehicular Test Environment](image)

![Figure 106: System performance in terms of average $C/I$ for the Vehicular Test Environment](image)
An equivalent simulations for the Pedestrian Test Environment were carried out. Outage probability plots are shown in Figure 107.

![Figure 107: System performance in terms of outage probability - Pedestrian Test Environment](image)

The corresponding achieved average $C/I$ levels are plotted in Figure 108.

![Figure 108: System performance in terms of average C/I for the Pedestrian Test Environment](image)

As the system load increases, the performance of the power control algorithms under investigation exhibits similar characteristics in both the Test Environment. However, it can be
observed that the proposed algorithm performs better in the case of a micro cell environment (Pedestrian Test Environment). This can be attributed to the fact that the method imposes an optimum power control criterion applicable to the intracell users of each cell-site. Intercell interference is measured and treated as a fixed value. Improved cell separation proves to counteract the effects of non-optimum intercell-interference management, providing an improved gain in quality and outage probability performance.

In general it can be observed that the proposed near-optimum power control scheme manages to "track" the limits up to which the quality demands of users without experiencing severe system performance degradations can be increased. The price paid is that the algorithm is not completely decentralised and requires intracell path-gain information along with some co-operation between base stations.
Chapter 7

7. Conclusions and Future Work

The presented work here addressed a large number of issues related to power control. Amongst these the following major conclusions can be drawn.

7.1 Effective Shadow Correlation Distance

One of the major results of Chapter 3 is the correlation distance of shadowing in satellite environments. Until now, the second order statistics of the shadow correlation in satellite environments were not very well known. Analysis of channel recordings at L-band can be summarised in the following Table:

<table>
<thead>
<tr>
<th></th>
<th>Correlation Distance (m) at 60° Elevation Angle</th>
<th>Correlation Distance (m) 80° Elevation Angle</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wooded Environment</td>
<td>20</td>
<td>9</td>
</tr>
<tr>
<td>Suburban Environment</td>
<td>16</td>
<td>20</td>
</tr>
</tbody>
</table>

*Table 10: Effective correlation distance summary table*

The results show that in wooded environment a significant reduction in the correlation distance is experienced for higher elevation angles. However, in the suburban environment correlation distance remained in a range between 16-20m despite the elevation angle change.

The presented results are highly dependent on the regional characteristics of the area in which measurements were carried out. Whilst the above provides a good indication of the order of magnitude, it is highly desired to expand the span of measurements to different countries and operational environments, as well range of the elevation angles.
7.2 Environment Detection Algorithm

The proposed environment detection algorithm (EDA) provides an essential component of any adaptive communication system. EDA is in fact one of the foundations of the proposed adaptive power control schemes of Chapter 5. Up to now, most of the adaptive systems research assumed perfect knowledge of speed and characteristics of the operational environment. The proposed algorithm simply brings true adaptive communications one step closer.

EDA performance evaluations demonstrated that:

- The advanced speed estimation algorithm is highly accurate and fast.
- The accuracy of the speed estimation algorithm reaches its maximum for an averaging window of 300ms.
- The proposed adaptive filtering used for separation of fast and slow fading is an accurate and essential feature of EDA without which estimation errors significantly increases.
- The combination of RMS delay spread estimation and shadow correlation distance provides extremely valuable information about the local operation environment.
- The algorithm can operate on single or multiple RAKE taps for increased performance.

EDA has shown that in a changing speed and channel conditions it accurately tracks the operational environment as shown below.

![Figure 109: EDA tracking changing RMS delay spread and speed simultaneously](image_url)
Going forward, all the lookup tables that define characteristics of the operational environment would need to be established. This would require extensive measurement campaigns and data processing. After this stage, various weighting factors highlighted in the algorithm would require optimisation for best performance.

7.3 SA-CLPC

The proposed speed adapted closed loop power control algorithm demonstrates significant gains over the conventional method. This gain can directly be converted into an increased system capacity and QoS. It has been demonstrated that by careful selection of different step size values combined with speed categories, the impact of speed estimation error on the performance of SA-CLPC can be managed. Using the advance speed estimation module of the EDA, it was clearly demonstrated that the SA-CLPC in terrestrial environment experiences no noticeable degradation in performance compared to the ideal case. The following table provides a summary of the SA-CLPC gains.

<table>
<thead>
<tr>
<th>Speed of 20 (km/h)</th>
<th>Speed of 30 (km/h)</th>
<th>Speed of 40 (km/h)</th>
<th>Speed of 50 (km/h)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain (dB)</td>
<td>0.6</td>
<td>0.7</td>
<td>0.8</td>
</tr>
</tbody>
</table>

Table 11: Summary of SA-CLPC gain over fixed step CLPC with 1dB step size in terrestrial propagation environment

As a part of this study, performance of conventional CLPC in actual satellite channel recordings was evaluated. The results of the conventional fixed step CLPC showed no significant improvement for RTDs of longer than 10ms.

As expected in satellite environments, long RTDs are the main limiting factor when it comes to CLPC. Nevertheless, the superiority of SA-CLPC can easily be demonstrated for low speed mobiles. For example, in an L-band LEO system with a Rice factor of 8dB and a RTD of 20ms, a 10km/h mobile using SA-CLPC will have over 0.5dB gain compared to a 1dB fixed step size CLPC.

In satellite environments, fast moving mobiles would in fact benefit from ignoring the power control commands. This is mainly due to the fact that by the time a power control command is received, the propagation channel has significantly changed. Failure to switch the power control loop off results in significant performance degradation at higher speeds. SA-CLPC takes advantage of this technique to reduce the PCE at higher speeds.

SA-CLPC’s gain is further extended in satellite systems by making the averaging period based on which power control commands are issued a function of mobile speed. This adds a further 0.2-0.5 dB to the gain of SA-CLPC.

By adding a predictive element to SA-CLPC, one can overcome some of the latency problem of the satellite channels and achieve further gains. Any such predictive scheme would heavily rely on speed estimations as a tool for determining how far ahead each prediction is referring to thereby issuing optimum commands well ahead of time.
7.4 Near-Optimum Distributed Power Control

Centralised power control schemes simply set the upper bound for performance, however, they are impractical and cannot be implemented due to complexity and the amount of information exchange required between the BS and BSCs. On the other hand, distributed power control algorithms do not require the link gain knowledge of all active users in the system providing a practical solution for managing system interference.

The proposed Near-Optimum Distributed Power Control algorithm of Chapter 7, operates on intracell link-gain knowledge of a given cell only. Therefore the algorithm does not require link gain information exchange between different base stations. Only a limited co-operation between base stations is required to avoid instability (the party effect) in the system.

The proposed method was tested in comparison with fixed-target and Soft Dropping power control algorithms. The target $C/I$ of the two fixed target algorithms were set to $CIR_{\text{MIN}}$ for "just acceptable" signal quality and $CIR_{\text{MAX}}$ for "ideal" call quality. The efficiency of the algorithms was assessed in terms of maximum achieved capacity and mean received $C/I$ levels of users. The maximum achieved capacities and average $C/I$ levels of users for the Pedestrian Test Environment are shown in Table 6.1. The $C/I$ levels correspond to a relatively light traffic of 2 users/MHz/cell.

<table>
<thead>
<tr>
<th>Employed Scheme</th>
<th>Maximum capacity (users/MHz/cell)</th>
<th>Achieved average $C/I$ level for Light Traffic (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$CIR_{\text{tar}}=CIR_{\text{min}}$</td>
<td>10.6</td>
<td>-19.05</td>
</tr>
<tr>
<td>$CIR_{\text{tar}}=CIR_{\text{max}}$</td>
<td>6.1</td>
<td>-16.10</td>
</tr>
<tr>
<td>Soft Dropping</td>
<td>8.0</td>
<td>-16.70</td>
</tr>
<tr>
<td>Near-Optimum</td>
<td>10.6</td>
<td>-16.68</td>
</tr>
</tbody>
</table>

Table 12: Performance of power control algorithms

From Table 12 it can be seen that maximum capacity is achieved by setting $CIR_{\text{tar}}=CIR_{\text{min}}$. Maximum $C/I$ levels under light traffic conditions are achieved when $CIR_{\text{tar}}=CIR_{\text{max}}$. The proposed algorithm is proved to adapt the target $C/I$ to the actual system load, thereby achieving the maximum capacity and the highest $C/I$ at the same time.

The basic assumption throughout this work was that $C/I$ is the criterion based on which call quality is measured. Although extensively used in the literature, this assumption may not hold under realistic system conditions. This is due to the fact that a certain level of $C/I$ or SIR can correspond to different values of BER or frame error rate (FER) for different users due to their local operational environment, speed and channel conditions. The "mapping" of different SIR levels to the required FER is performed by an outer loop function. A more realistic approach
should be able to cope with such variations and provide a mechanism for adjusting the target parameters of the outer loop power control according to channel fading conditions (Layer 1) and the experienced interference levels in the network (Layer 3). This may bring about the need for modelling the Rayleigh-channel variations, which in the current approach was assumed to be averaged out.

During the operation of the proposed algorithm, a Central Controller unit is simply required to determine a minimum value for of the optimum $C/I$ estimates with respect to the neighbouring cells of each user. Furthermore, the performed link-gain, power, and $C/I$ measurements are assumed to be perfect in the absence of noise. In order to provide robustness to measurement errors a more sophisticated method for processing the available link-path information would be required.
8. References


9. Appendix-A: UMTS Wideband Propagation Channel

UMTS is expected to operate in a large number of environments, large and small cities, with variations in building construction, as well as rural, desert, and mountainous areas. Since it impossible to consider all possible radio environments in the design of a mobile system and construct propagation models for them, a smaller set of more general models, which adequately span the overall range of possible environments, are required. As a result, the large number of possible radio environments has been condensed into the three following test radio environments which are intended to cover the range of the UMTS operating environments:

- Vehicular test environment
- Outdoor to indoor and pedestrian test environment
- Indoor office test environment

These test environments correspond to macro cell, micro cell and picocell cell types, respectively. The case of the Indoor Office test environment has not been considered in this work.

9.1 Vehicular Test Environment

The vehicular test environment is characterised by large cells (macro cells) and high transmit powers. The rate of the Rayleigh fading is determined by the speed of the mobile units which is high due to the fast moving vehicles. However, lower fading rates should also exist in application where the terminals are stationary. A typical value for the standard deviation of the shadowing is 10 dB.

9.2 Outdoor to Indoor and Pedestrian Test Environment

This environment is characterised by small micro cells and low transmit powers. The base station antennas are located below rooftops, while pedestrian users are located on streets and inside buildings. Both line-of-sight and non line-of-sight connections exist. Indoor coverage can also be provided from an outdoor base station. The rate of Rayleigh and/or Rician fading is set by walking speeds, however more rapid fading may be experienced due to reflections from moving vehicles. Typical values of the standard deviation of the shadowing is 10 dB for outdoors and 12 dB for indoors.

For each test environment, a channel impulse response model based on the tapped-delay line model is given. The model is characterised by the number of taps, the time delay relative to the first tap, the average power relative to the strongest tap, and the Doppler spectrum of each tap.
Most of the time, RMS delay spreads are relatively small, but occasionally, there are ‘worst case’
multipath characteristics that lead to much larger RMS delay spreads. Measurements in outdoor
environments show that RMS delay spread can vary over an order of magnitude, within the same
environment. Although large delay spreads occur relatively infrequently, they can have a major
impact on system performance.

For each environment, two types of channel are defined: **channel A** where the RMS delay spread
is low and which occurs frequently, and **channel B** where the value of RMS delay spread is
median and which also occurs frequently. Each of these two channels are expected to be
experienced for a certain percentage of time in a given environment. Table 13 lists the RMS
average delay spread for channel A and B for each environment, as well as the percentage of
time.

<table>
<thead>
<tr>
<th>Test Environment</th>
<th>Channel A</th>
<th>Channel B</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>RMS Delay (nsec)</td>
<td>P(A) [%]</td>
</tr>
<tr>
<td>Outdoor to Indoor and Pedestrian</td>
<td>46</td>
<td>40</td>
</tr>
<tr>
<td>Vehicular – High Antenna</td>
<td>370</td>
<td>40</td>
</tr>
</tbody>
</table>

**Table 13:** Parameters of the Channel Impulse Responses of the Test Environments

The tapped-delay-line parameters for each of the test environments that are being considered are
shown in the following tables. For each tap of the channel three parameters are given: the time
delay relative to the first tap, the average power relative to the strongest tap, and the Doppler
spectrum of each tap as shown in Table 14 and Table 15.

<table>
<thead>
<tr>
<th>Tap</th>
<th>Channel A</th>
<th>Channel B</th>
<th>Doppler Spectrum</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0</td>
<td>0.0</td>
<td>0</td>
</tr>
<tr>
<td>2</td>
<td>310</td>
<td>-1.0</td>
<td>300</td>
</tr>
<tr>
<td>3</td>
<td>710</td>
<td>-9.0</td>
<td>8900</td>
</tr>
<tr>
<td>4</td>
<td>1090</td>
<td>-10.0</td>
<td>12900</td>
</tr>
<tr>
<td>5</td>
<td>1730</td>
<td>-15.0</td>
<td>17100</td>
</tr>
<tr>
<td>6</td>
<td>2510</td>
<td>-20.0</td>
<td>20000</td>
</tr>
</tbody>
</table>

**Table 14:** Vehicular Test Environment, High Antenna, and Tapped-Delay-Line Parameters
<table>
<thead>
<tr>
<th>Tap</th>
<th>Channel A</th>
<th>Channel B</th>
<th>Doppler Spectrum</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>0</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>2</td>
<td>110</td>
<td>-9.7</td>
<td>200</td>
</tr>
<tr>
<td>3</td>
<td>190</td>
<td>-19.2</td>
<td>800</td>
</tr>
<tr>
<td>4</td>
<td>410</td>
<td>-22.8</td>
<td>1200</td>
</tr>
<tr>
<td>5</td>
<td>-</td>
<td>-</td>
<td>2300</td>
</tr>
<tr>
<td>6</td>
<td>-</td>
<td>-</td>
<td>3700</td>
</tr>
</tbody>
</table>

*Table 15: Outdoor to Indoor and Pedestrian Test Environment Tapped-Delay-Line Parameters*
10. ANNEX-B: System Level Test Environment

Simulations evaluating the efficiency of various algorithms of Chapter 6 were carried out using the Vehicular Test Environment and the Outdoor To Indoor and Pedestrian Test Environment specified in UMTS. The former is characterised by relatively large cells (radius of 2 km) and high transmitter powers. Assuming limited spectrum, higher cell capacity is important. The Outdoor To Indoor and Pedestrian Test Environment is characterised by small cells and low transmit powers. Base stations with low antenna heights (below rooftops) are used and pedestrian users are located on the streets of a Manhattan-like urban model.

10.1 Vehicular Simulation Environment

The cell radius of the deployment model is 2000 m for services up to 144kb/s and 500 m for services above 144kb/s. Only speech users are considered and therefore a maximum cell-radius of 2000 m will be used. The base station antenna height is 15 m above the average rooftop level. The deployment scheme shown in Figure 110 is assumed to be a hexagonal cell lay-out with distances between base stations equal to 6000 m. Tri-sectored cells are nominally used. The model defining the mobility of users in the Vehicular Test Environment is a pseudo-random mobility model with semi-directed trajectories.

![Figure 110: Vehicular Test Environment Deployment Model](image-url)
Mobile’s position is updated according to the decorrelation length of the lognormal shadowing and direction can be changed at each position update according to a given probability. The speed of mobile users is constant. The model is characterised by the parameters shown in Table 16. Mobiles are uniformly distributed on the map and their directions are randomly chosen at initialisation.

<table>
<thead>
<tr>
<th>Speed value</th>
<th>120 km/h</th>
</tr>
</thead>
<tbody>
<tr>
<td>Probability to change direction at position update</td>
<td>0.2</td>
</tr>
<tr>
<td>Maximum angle of direction update</td>
<td>45 degrees</td>
</tr>
</tbody>
</table>

*Table 16: Mobility parameters for the Vehicular Test Environment*

### 10.2 Outdoor to Indoor and Pedestrian Simulation Environment

The physical environment description includes only outdoor users located on the streets of a Manhattan-like urban model. The base station antennas are omni directional with heights below rooftops (micro-cellular environment). The system layout and the positions of the base stations are shown in Figure 111. The block size is $200m \times 200m$, the street width is $30m$ and the height difference between the base station and mobile station antenna is $10m$.

*Figure 111: Indoor To Outdoor and Pedestrian Deployment Model*
Mobiles move in the streets of the Manhattan-like structure according to a pseudo-random mobility model. Users are uniformly distributed in the streets and their direction is randomly chosen at initialisation. The turning probability is illustrated in Figure 112.

\[
\text{Figure 112: Turning probability for the Pedestrian Test Environment mobility model}
\]

10.3 Propagation Models

The propagation effects are divided into three distinct types of model. These are path loss, slow variation about the mean due to shadowing and the rapid variations due to multipath. In the sequel it is assumed that the Rayleigh fading is averaged out since our algorithms operate based on measurements performed over relatively long periods (the averaging lengths used are at least 20 wavelengths as discussed in Chapter 4).

10.3.1 Path Loss Models

10.3.1.1 Vehicular Test Environment

This model is applicable for the test scenarios in urban and suburban areas outside the high rise core where the buildings are of nearly uniform height.

\[
L = 40 \left(1 - 4 \times 10^3 \Delta h_b \right) \log_{10}(R) - 18 \log_{10}(\Delta h_b) + 21 \log_{10}(f) + 80 \quad (dB)
\]

Equation 95

where \( R \) is the separation distance between the base station and the mobile station in kilometres, \( f \) is the carrier frequency of 2000 MHz and \( \Delta h_b \) is the base station antenna height in metres, measured from the average rooftop level. The base station antenna height is fixed at 15 m above the average rooftop level (\( \Delta h_b = 15 \text{ m} \)). Considering the values given above becomes:
\[ L = 128.1 + 37.6 \log_{10}(R) \text{ (dB)} \] \hspace{1cm} \text{Equation 96}

\( L \) shall in no circumstances be less than free space loss. This model is valid for NLOS case only and describes worse case propagation.

### 10.3.1.2 Outdoor To Indoor and Pedestrian Test Environment

The model used is a recursive model [UMTS98] that calculates the path loss as a sum of LOS and NLOS segments. The shortest path along streets between the BS and the MS has to be found within the Manhattan environment. The path loss due to propagation through the streets is given by the well-known formula:

\[ L_{st} = 20 \log \left( \frac{4\pi d_n}{\lambda} \right) \] \hspace{1cm} \text{Equation 97}

where \( d_n \) is the "illusory" distance, \( \lambda \) is the wavelength and \( n \) is the number of straight segments between the BS and the MS (along the shortest path). The illusory distance is the sum of these street segments and can be obtained by recursively using the expressions

\[ k_n = k_{n-1} + d_{n-1}c \quad \text{and} \quad d_n = k_ns_{n-1} + d_{n-1} \]

where \( c \) is a function of the angle of the street crossing. For a 90° street crossing the value \( c \) should be set to 0.5. Further, \( s_{n-1} \) is the length in metres of the last segment. A segment is a straight path. The initial values are set according to \( k_0 = 1 \) and \( d_0 = 0 \). The illusory distance is obtained as the final \( d_n \) when the last segment has been added.

The model is extended to cover the micro cell dual slope behaviour, by modifying the expression to:

\[ L_{st} = 20 \log_{10} \left( \frac{4\pi d_n}{\lambda} \cdot D \left( \sum_{j=1}^{n-1} s_j \right) \right) \] \hspace{1cm} \text{Equation 98}

Where

\[ D(x) = \begin{cases} x/x_{br}, & x > x_{br} \\ 1, & x \leq x_{br} \end{cases} \] \hspace{1cm} \text{Equation 99}
Before the breakpoint \( x_{br} \) the slope is 2, after the breakpoint it increases to 4. The breakpoint is set to 300m.

To take account of the effects of propagation going above rooftop, it is also needed to calculate the path loss according to the shortest geographical distance. This is done by using the COST Walfish-Ikegami model and with antennas below rooftops:

\[
L_{\text{macro}} = 24 + 45 \log(d + 20)
\]

Equation 100

where \( d \) is the shortest physical geographical distance from the transmitter to the receiver. The final path loss value is the minimum between path loss from the propagation through the streets and the path loss based on the shortest geographical distance.

\[
L = \min(L_{st}, L_{\text{macro}})
\]

Equation 101

10.3.2 Shadowing Model

The long-term (log-normal) fading is characterised by a Gaussian distribution with zero mean on the logarithmic scale. Due to the slow fading process versus distance, adjacent values are correlated. The normalised autocorrelation function \( R(\Delta x) \) can be described with sufficient accuracy by an exponential function:

\[
R(\Delta x) = e^{-\frac{\Delta x}{d_{cor}}}
\]

Equation 102

where \( d_{cor} \) is the decorrelation length.

Suppose that the lognormal component of the path loss at position \( P_1 \) has been determined to be \( L_1 \) and it is intended to compute the lognormal component \( L_2 \) at the "next" position \( P_2 \), where \( P_2 \) is \( \Delta x \) metres away from \( P_1 \). Then \( L_2 \) is normally distributed with mean \( R(\Delta x)L_1 \) and standard deviation \( \sqrt{1-R(\Delta x)^2}\sigma \), where \( \sigma \) is the standard deviation of the lognormal fading for that environment. The values for the decorrelation length and standard deviation of the long-term fading for the two test environments are shown in the following table.
## System Level Test Environment

<table>
<thead>
<tr>
<th>Test Environment</th>
<th>Decorrelation Length $d_{cor}$</th>
<th>Standard Deviation $\sigma$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Vehicular</td>
<td>20 m</td>
<td>10 dB</td>
</tr>
<tr>
<td>Outdoor to Indoor and Pedestrian</td>
<td>5 m</td>
<td>12 dB</td>
</tr>
</tbody>
</table>

*Table 17: Parameters of lognormal fading for the employed Test Environments*