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Inverse Multiplexing of IS95 Traffic for Video Transmission

by

Mazda Salmanian

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

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Faculty of Engineering
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The undersigned hereby recommend to the Faculty of Graduate Studies
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submitted by Mazda Salmanian, B.Sc.
in partial fulfillment of the requirements for
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Abstract

This thesis studies the transmission of video signals over an IS95 CDMA communication system through Rayleigh fading channels. The video encoder's (H.263-compliant) bit rate ranges from 16Kbps to 64Kbps with which the air interface (IS95) system copes by adapting and assigning more channels to the video users dynamically. We consider up to 8 channels at 9600bps per video user. A channel in the reverse direction (mobile to the base-station) is identified by the ESN (Electronic Serial Number) mask of the mobile; whereas a channel in the forward direction (base-station to the mobile) is identified by its Walsh code spreader. Effects of power control, slow and fast fading, and channel utilization are studied through simulation of several scenarios involving audio and video users. Power control is required to maintain a reliable communication channel, especially at low mobile speeds; however it is observed that video-users do not require power control and synchronization per channel like audio users do. The base-station manages the channels assigned to a video user as one, since their origins are co-located. The system operates reliably in fast fading environments where the fade duration is several times shorter than the interleaver's depth. It is also found that a video-user's channel utilization is more efficient than that of an audio user. It is concluded that such system can reliably support 2 video users in the forward and reverse traffic channels with the presence of 6 to 8 audio interferers.

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List of Acronyms

		Page
AIM	ATM Inverse Multiplexing	4
Aimux	ATM Inverse Multiplexing	3
AMPS	The advanced Mobile Phone System	5
AR	Activity Ratio	40
ARMA	Auto Regressive Moving Average	12
ATM	Asynchronous Transfer Mode	3
AWGN	Additive White Gaussian Noise	73
BER	Bit Error Rate	5
BPSK	Bipolar Phase Shift Keying	5
BW _c	Coherence Bandwidth of the channel	80
CBR	Constant Bit-Rate	12
CCITT	International Telephone and Telegraph Consultation Committee	10
CDMA	Code Division Multiple Access	1
CODIT	Code Division Testbed	1
CPU	Computing Processor Unit	11
DCT	Discrete Cosine Transform	12
DPSK	Differential Phase Shift Keying	56
ESN	Electronic Serial Number	iii
IID	Independent Identically Distributed	16
ISDN	Integrated Services Digital Networks	4
ISO	International Standards Organization	11
ITU-T	International Telecommunication Union - Telecommunication Standardization Sector	11
MIN	Mobile Identity Number	5
MMPP	Modified Markov Poisson Process	13
MPEG	Motion Picture Experts Group	9

NNI	Network-to-Network Interface	3
PCS	Personal Communication Systems	1
PN	Pseudo Random Noise	5
QQCIF	Quarter, Quarter Common Interface Format	11
RACE	Research and development in Advanced Communication Technologies for Europe	34
SNR	Signal to Noise Ratio	10
TDD	Time Division Duplex	35
TDMA	Time Division Multiple Access	35
telco	Telephone Company	1
UNI	User to Network Interface	3
VBR	Variable Bit Rate	10
VR	CELP Variable Rate Codebook Excited Linear Predictive coding	9
W	System Bandwidth	80
WAN	Wide Area Network	3

Chapter 1: Introduction

1.1 Perspective

Code Division Multiple Access has received much attention in recent years. This technology is also a candidate as an access technique for the third generation Personal Communication Systems [1]. The move toward this higher and more complex technology is demand driven. The customers' need for more services from the telephone companies (telcos) and other providers has lead researchers to search for more capacity efficient techniques, such as CDMA.

In 1994 authors of [1] gave an overview of the CODIT system with an introduction which stated that “present second generation CDMA systems are primarily designed as low-rate voice and data microcellular systems and fall short of meeting important requirements of third generation systems.” They added that the third generation systems go further, especially with respect to the range of services, service bit-rates, quality of services, mixed-cell operation, and flexibility in frequency and radio resource management, and system provisioning. They emphasized the fact that a “multiple RF channel bandwidth” system is the appropriate way “to implement an open and flexible radio interface as required for third generation systems.”

One of the services that has been a subject of extensive research is video telephony [2]-[8]. With the growing need for multimedia networks, researchers attempt to find better ways for providing good quality video images for personal communication. At the same time, demand for mobility has increased. Wireless networks and mobile personal communicators, such as pagers and cellular phones, are in such demand that in some countries the demand outstrips the supply. Amalgamation of mobility and video

telephony is the latest trend of the demanded services, and this thesis addresses just that.

The following sections provide overviews of mobile video telephony, inverse multiplexing, and of CDMA for introducing the issues addressed in this thesis.

1.2 A Brief Review of Mobile Video Telephony

Video telephony is one of the demanded services in the next generation mobile networks. Mobile networks have different challenges than the traditional wired networks. Telephone communication has been for fixed transmitters and receivers through wired networks. Although microwave towers provide convenient wireless line-of-sight communication links between two nodes, the transmitter or receiver nodes of cellular networks are neither connected with wires, nor do they always have a line-of-sight communication link. The challenge in cellular networks is to provide a reliable method of communication through channels and links which continuously change with time because of mobility through different obstacles and environments.

Video telephony brings other challenges to mobile networks. Telephone communication has been for voice and low bit-rate data services (less than 10 Kbps). Most of the hardware, software, and services developed to this time have been for such low bit-rate applications. The next generation networks, however, demand applications, such as video, which require high bit-rate transmissions. Faster transmission of data is especially needed for video traffic to keep the real-time nature of the communicated images. The challenge for mobile networks is to evolve and provide high bit-rate services with the least cost incurred. Inverse multiplexing is one method which allows for video transmission through existing voice channels.

1.3 An Overview of Inverse Multiplexing

Inverse multiplexing is a data transfer method which allocates traffic from a high-capacity link to few low-capacity links. This is in contrast with the working definition of multiplexing which implies that few low-capacity users get access to communicate through a high-capacity link. Inverse multiplexing is a technique capable of allocating bandwidth on demand which makes access more flexible from a user point of view. This technique has much support in the literature and it is becoming standardized for the industry use.

1.3.1 Examples of Existing Inverse Multiplexing Schemes

ATM Inverse Multiplexing (Aimux) [9] is a method under discussion for standardization which gives bandwidth flexibility to the ATM users in the area of User to Network Interface (UNI) through T1 (1.5 Mbps) and T3 (45 Mbps) circuits. Aimux provides additional affordable bandwidth in T1 increments up to eight T1s, after which T3 becomes more cost-effective. Aimux is a physical layer protocol for cell¹ delivery in-sequence and within tolerable delays with no more than 1% overhead for initialization and connection, sequencing, re-synchronizing (for dropped links), and differential circuit delay management. Aimux may also be utilized to interconnect ATM networks over a Network-to-Network Interface (NNI) for WAN access. Aimux is flexible to support many types of services: variable bit rate, constant bit rate, and available bit rate for voice, data, and video traffic. The industry's demand for such an interface with strong financial justification [10] continues to drive the standardization of ATM Inverse Multiplexing.

1. ATM data frames are referred to as CELL in ATM terminology.

Authors in [11] propose the use of inverse multiplexing with multi B-channel ISDN lines for Internet dial-up access - a scenario that the Internet Service Providers and the users benefit in speed of access and in terms of cost. "With inverse multiplexing and dial-on-demand functionality, the individual user or network router appears to have a dedicated circuit."

The European ATM Inverse Multiplexing interface (AIM) [12] provides "a high-speed connection between two locations by using multiple point-to-point E1 links (2.048 Mbps) and then manages them as a single bundle making it easier to achieve bandwidth efficiency and to carry high bandwidth applications." According to Bandwidth On Demand Interoperability Group (BONDING) specifications bit-level inverse multiplexing is attractive for traffic applications such as video conferencing; however, the data rates cannot be readily extended to higher speeds. AIM is fully compatible with existing ATM interface standards because of cell-level inverse multiplexing. AIM is "implemented as a sublayer within the transmission convergence layer" of ATM. The benefits of this method include easy upgrade in E1 increments, multimedia support, reliability by link load sharing, efficiency, and versatility because of cell switching.

The common point in these examples is the fact that a high capacity variable bit-rate application communicates its information through a number of low bit-rate channels which are either fixed or dynamically allocated. The method of inverse multiplexing is used in this project to communicate high bit-rate (variable rate) coded video signals (16 to 64 Kbps) through a few (up to 8) low bit-rate IS95 CDMA voice channels (9.6 Kbps).

1.4 An Overview of CDMA

Code Division Multiple Access scheme (CDMA) is one technology with which users

gain access to the communication channel with their own key-like codes. In CDMA the communication channel is usually open to all users so long as the encryption of the user specific data is decodable and retrievable. An equivalent analogy to this access scheme is the example of a conversation in a crowded room, where the listener is “tuned” to the speaker and filters out the background and surrounding conversations that are simultaneously taking place.

CDMA is chosen as the multiple access scheme of choice for this project because of its high capacity applications [13][14] in cellular networks and because it is standardized by the industry [15]. Spread spectrum is the technology utilized by the IS95 CDMA standard to resolve mobility problems of cellular telephony. Because of high capacity [16], privacy, multiple services, multiple rates [15], and low average transmit power, CDMA is also a candidate for the third generation personal communication systems' access technology.

The theoretical Erlang capacity of CDMA is 20 times that of AMPS [17] - based on a blocking definition established as the total interference-to-background noise ratio of above 10 dB. Privacy is inherent by spread spectrum and coding - user-specific coding (MIN and ESN masks) and use of Pseudo random Noise (PN) codes. The correlation properties of the PN coded transmitted signals make CDMA resistant to interference and user limited, for a minimum quality of service and error in the system. The bit error rate (BER) of a bipolar phase-shift-keyed (BPSK) system in [18] stayed in the order of 10^{-5} even when the received signal power was lower than the Direct Sequence waveform by as much as 2 dB. These properties have made CDMA the access scheme of choice for this project.

1.5 Thesis Description and Organization

The following chapters provide background and support for one proposed implementation of mobile video telephony through existing cellular networks. A simulation of “variable bit-rate video transmission with inverse multiplexing and dynamic channel allocation in an IS95 CDMA system” is implemented to demonstrate the system under various conditions. The research concludes that video transmission as proposed and demonstrated makes efficient use of the physical resources, and that Inverse Multiplexed video data is more receiver friendly than audio data and does not require as much supervision and maintenance. These key points along with their supporting information assist in a better understanding of the capacity behavior of such systems.

The focus of this research is on the system capacity issues of an IS95 system with video users, as shown in Figure 1. In Chapter 2 a brief background on real-time video signals, source coding and modeling, and other related concepts to this thesis are presented. Overviews of the IS95 standard, call setup, and system specifications are given in Chapter 3. The key points and focus of this project are reiterated in Chapter 4. In Chapter 5 the system simulation and assumptions are discussed in detail. This chapter presents the modifications to the current IS95 standard which allow for functional video users in the system. Chapter 6 demonstrates the system under study with various test-case scenarios and simulations and Chapter 7 concludes the project and suggests future research areas.

Figure 1: System Diagram

Chapter 2: Video Source Modelling

2.1 Perspective

Wireless video transmission with CDMA access technology is the subject of this thesis. In order to demonstrate a video telephony system employing IS95 CDMA technology, a brief description of video signal characteristics and modeling is presented in this chapter; explanations on the basic characteristics and concepts of the multipath mobile fading channel are provided in APPENDIX A.

2.2 Real Time Video Signals

Once a still image is digitized, the information regarding every pixel, luminance and color, is recorded in one or two bytes, depending on the coding algorithm. The Common Interface Format (CIF) [19] has 325 pixels per line, and 288 lines per picture. The corresponding information may be stored in 100 KBytes of memory. Video images are made of continuously changing still images; they change fast enough that the stillness is not visible to the human eye - 5 to 24 times a second. One may note that transfer, reconstruction, and error control of 1 MBytes of information per second is quite costly. It is for this reason that researchers take advantage of the image characteristics of video signals to compress the information.

The pictorial changes of subsequent video images are correlated and the common sections from one image to another need only to be coded once and reused in the other. This results in compression of the video data and decreasing its bit-rate. Therefore the scene changes are the key areas which require detailed coding; these changes increase the output bit-rate of the video coder. One may note that the resulting coded

(compressed) video data is quite bursty in time.

Video signals are not tolerant of large delays, of more than a few milliseconds. If the image, or part of the image, is not reconstructed at the receiver, the viewer would see dark and erroneous pixels, or blocks of pixels, for a short period of time till the next scene change. However, if the image is delayed, the received information may be difficult to comprehend for the viewer. Therefore image compression assists in reducing the pictorial redundancies, and in turn reduces the transferable data which can result in faster transmission and lower delay.

The compression and coding algorithms take advantage of the redundancies when scenes are not changing fast - this is especially useful for “talking heads” on videophones’ screens. Examples of popular compression techniques are Moving Picture Experts Group (MPEG), and H.261 [19][20]. But even after sophisticated compression techniques, the compressed video data rate varies from 16 Kbps to 2 Mbps.

Video transmissions through wireless channels have been achieved [21] with peak rates below 64 Kbps. Video signals of QCIF (quarter CIF: 176 x 144 pixels) resolution were studied in [21]. The source encoder chose a rate dynamically, from a set of specified rates, based on its Automatic Repeat Request (ARQ) buffer occupancy which is an indication of the channel condition. The authors used two modes of coding: one at 65.6 Kbps and the other at 54 Kbps. They achieved the lower rate by using a larger quantizer step size.

There are also techniques to cope with the bursty nature of voice data which may eventually be redesigned and applied to video. The Variable Rate Codebook Excited Linear Predictive coder (VR CELP) of the CODIT (RACE) project [22][23] chooses 7

different rates of output dynamically based on the tiered Voice Activity Detector, Voicing Decision, and SNR. Although such source coder is not available for video to be used in this project, a source may be assumed which generates coded random data based on the statistical characteristics of video signals.

The following sections elaborate on the video coders of interest in the industry and research literature which may be used for this project, and provide an overview on video data's statistical characteristics and modeling.

2.2.1 Source Coding

The area of source coding for video signals is aggressively under research to meet the demands of different types of networks, such as ATM and cellular wireless. MPEG and H.261 are the most utilized standards in the teleconferencing field, and researchers continuously attempt to enhance and modify them to match their system needs.

The International Telecommunication Union's (ITU former CCITT) H.261 standard is an algorithm for video coding at multiple integers rates of 64 Kbps up to 30x64 Kbps [19]. This algorithm is especially used for video conferencing because of its design for low bit rates, and its fine-scale quantization levels [24]. Authors of [24] describe a method of object-selective quantization that forces the rate control algorithm to "transfer a fraction of the available bit rate to the facial areas of the head and shoulders in video conferencing for better eye-contact and lip-sync.

A modified H.261 codec is introduced in [25] that produces variable bit rate (VBR) output for Frame Relay transmissions, as opposed to the integer multiples of 64 Kbps. This is accomplished by "disabling the rate feedback loop from the codec output buffer to the quantizer," keeping the quantization number constant, and by applying a rate

control based on the network congestion. The authors downloaded the H.261 codec software from the public domain internet, developed by the Portable Video Research Group at Stanford University, and acquired another codec proprietary to the Ottawa based Dedicated Technologies Corporation (DTC). The proposed rate controller either reduces the video frame rate or increases the video codec quantization number. With 350 ms cell delay tolerance of video conferencing [25], and packet loss rates of around 2% - due to buffer overflow or excessive delay - the authors modified the standard and demonstrated a VBR video conferencing over an integrated services Frame Relay network.

In [4] the advantage of the “talking head” shot characteristic of video conferencing, as opposed to a conference room, is presented along with an H.261-compatible video decoder based on a reduced spatial resolution for the display. This is “achieved by decoding the low-frequency components of the received bit-stream and by producing a decimated image.” The decoder is implemented in software that adds flexibility and saves on hardware. Because of lower but acceptable resolution, the software may be run on a desktop CPU. The encoder's complexity is easily reduced by omitting the CPU intensive functions such as the Motion Compensator which is permitted by the H.261 recommendation. The software's images are QQCIF (88 X 72 pixels) instead of QCIF (172 X 144 - Quarter Common Interface Format).

The International Standards Organization (ISO) and the ITU are concentrating on standards (under completion) for very low bit-rate video conferencing [3]. ISO's MPEG-4 is a generic video coding standard which may be operating in very low bit-rates. However, ITU-T has established a draft of H.263 specifically for very low bit-rate video transmission of less than 64 Kbps. This standard is based on H.261 but its algorithm

suffers from “blocking” at low bit rates because of truncation of the high frequency discrete cosine transform (DCT) coefficients for achieving low bit rates [3]. The authors in [3] reduce this “blocking” effect by weighting the DCT coefficients with a visual quantization matrix before quantization.

The source coder of this project is assumed to follow or be a modified version of the H.263 standard for very low bit-rate video transmission. The peak rate is assumed to be 64 Kbps. Although actual real-time video data is not used in this thesis, the coded video data is generated randomly (via simulation) in accordance to its statistical characteristics.

2.2.2 Video Model

Real-time video traffic may be constant bit-rate (CBR) or variable bit-rate (VBR), depending on the coding algorithm. The CBR coders keep the rate constant by generating periodic streams of data during the transmission which keeps the inter-cell delay constant as well. VBR coders' rates are bursty due to the nature of video compression [20]. Variable bit-rate video coding, unlike constant bit rate coding, does not need large buffering and memory to optimize and account for bit-rate changes in frames with different contents [5]. CBR traffic however is predictable and is easier to manage, and because of more redundancy it is less susceptible to loss [20]. In this project variable bit-rate codecs are considered because of their efficiency, lower overhead, and shorter delays.

VBR video models proposed in the literature generally match in the first two moments (mean and autocovariance) of their bit-rate or inter-arrival time with that of the actual experimental data. Models based on Auto Regressive Moving Average (ARMA) process are generally used for simulations and not for analytical study [20]. However models that

estimate the aggregate video traffic by multi-stage birth and death Markov processes for the bit-rate [7] are good models for traffic studies and analysis.

A Modified Markov Poisson Process (MMPP) is proposed in [5], which also considers batch arrivals, for packet video modeling. Packet generation oscillates between two states, in one of which packets are generated by a Poisson process at a known rate, and in the other packet generation is idle.

The video traffic model used in [6] and [8] is based on a number of discrete-time two-state Markov chains, each of which alternate between an ON and an OFF state. The holding time on each state is geometrically distributed and all the two-state Markov chains are defined on a common time duration, equal to the frame length of the transport protocol. It is noted that the basic difference in voice and video source models is in the “time epoch” under which the two-state Markov chains are defined. Also for bursty video the independent two-state Markov chains do not necessarily have to be identically distributed.

Authors of [7] modeled uniform activity scenes, such as a talking person, via correlated Markov process models. For faster changing pictures, they allowed sources with different bit-rate statistics (mean and variance) in their model and approximated an exponential correlation decay for the data-rate process. Their overall aggregate state transition model is like a matrix of states, where every row and (every column) is a Markov chain - “a superposition of independent ON-OFF mini-processes”. They note two important correlations from their model: A short-term (fast-decaying) correlation corresponding to the uniform activities on a videophone, in the order of few hundred milliseconds, and a long-term (slow-decaying) correlation corresponding to scene

changes, in the order of a few seconds.

The two-state Markov chains model was adopted for this project because bursty data is easily generated and analyzed. The arrival process of the video model is briefly explained below.

2.2.2.1 The Arrival Process

The arrival process is modeled as a superposition of M independent identically distributed (IID) two-state Markov sources [20], as shown in Figure 1. A source changes from its ON state to OFF with probability $1-p$, and stays in OFF state with probability q . While in ON state, with probability p , the source generates γ frames during a time slot equal to the IS95 frame size¹.

1. It is assumed that the 172 input bits to IS95 frame include 12 bits of header for inverse multiplexing protocol and the rest are coded video.

Figure 1: The Variable-rate Video frame generator

The transition matrix of the source, and the state probabilities (π_{state}) are as follows [26]:

$$T = \begin{bmatrix} p & 1-p \\ 1-q & q \end{bmatrix} \quad (\text{EQ 1})$$

$$\pi_{ON} = \frac{1-q}{2-q-p} \text{ and } \pi_{OFF} = \frac{1-p}{2-q-p}. \quad (\text{EQ 2})$$

The mean of the process is the probability of being in ON state, times the mean of the source:

$$\bar{\epsilon} = \frac{1-q}{2-q-p} \cdot \gamma. \quad (\text{EQ 3})$$

And finally, because the sources are independent, the mean cell arrival rate is

$$\bar{\mu} = M \cdot \frac{1-q}{2-q-p} \cdot \gamma. \quad (\text{EQ 4})$$

The output of a variable bit rate video modeled by aggregate traffic produced with 16 IID Markov sources is displayed in Figure 2. The burstiness is noted as the packet arrival rate increases suddenly and returns back to the average. The Figure was generated with 283 20-milli second frames which correspond to 5.5 seconds of compressed coded-video transmission.

16 sources @ 4 Kb/s, 20 ms frame, => 80 bits/source per frame.
There are 50 IS95 frames in 1s.

Figure 2: The output of a VBR Video modeled as a superposition of 16 IID two-state Markov sources

Chapter 3: An Overview on IS95

3.1 Perspective

IS95 standard is utilized and adopted in this project as the physical layer of the communication system; therefore an overview description of this standard, which includes call setup protocol and system specifications, is presented in this chapter. This overview is followed by explanations on the transmission environment and not the reception - decoding techniques that correspond to the system specification and channel characteristics are not documented in the standard.

3.2 An Overview on IS95 Standard

IS95 is the North American CDMA standard revised by the Telecommunication Industry Association (TIA) and the Electronic Industry Association (EIA). It is a compatibility standard of the air interface between the Mobile Station (MS) and the Base-Station (BS) for dual-mode wideband spread spectrum cellular systems. The standard's coverage is parallel to the physical and the datalink layers of the Open System Interconnection (OSI) model. The standard is an integral part of this thesis; the physical layer of the IS95 system is implemented in detail in this project. Thus, the following overview is provided as a description of the system simulated in this thesis. In the sections below an explanation is given as to how a call is setup in an IS95 system, followed by the block diagram descriptions of the traffic channels.[15]

3.2.1 Channel Descriptions

In the IS95 system the Mobile Station (MS) and Base Station (BS) communicate through 6 group of channels: paging channel, access channel, forward traffic channel, reverse

traffic channel, sync channel, and the pilot channel. Reverse traffic and access channels (MS to BS) are identified by user long code and access long code sequences, whereas the pilot, sync, paging, and the forward traffic channels (BS to MS) are coded by unique Walsh functions.

Once powered on, the MS obtains time and frame synchronization information from the pilot and the sync channels, then registers with the BS. The pilot channel also provides a phase reference to the MS for coherent demodulation. After registration, the mobile monitors the paging channel for incoming calls from the BS, and transmits through the access channel for call originations or reply to pages. Once the call is established, the voice signals are communicated through the forward (BS to MS) and the reverse (MS to BS) traffic channels.

3.2.2 Call Setup Procedure

Upon powering on, the MS first goes through the registration procedure and provides its location, status, identification, and other characteristics to the BS to ensure that the BS is aware of the MS's capabilities, and can page the MS efficiently. After the MS is registered, it is ready to originate calls or receive (be terminated on) calls.

3.2.2.1 Mobile Originated Calls

Mobile originated calls, as shown in Figure 3, start by the *Origination* message sent from the MS to the BS through the access channel. The BS sets up a traffic channel and begins sending *null traffic data* frames to the MS. The *null traffic data* transmission is meant for the mobile to start and maintain connectivity with the BS. The BS also informs the MS of the assigned forward traffic channel with the *Channel Assignment* message via the paging channel. The MS then begins to transmit the *traffic channel preamble* frames to

the BS in order to aid the BS in acquiring the set reverse traffic channel. Once a connection is established, the BS sends a *(Base Station Acknowledgment) Order* message to the MS through the connected forward traffic channel. At this point the MS begins to transmit *null traffic data* frames to keep the connection “alive”. The BS sends a *(Service Option Response) Order* message to the MS through the forward traffic channel and the MS begins to process the traffic according to the Service Option parameter of the message. This parameter is governed by the TSB58 (Technical Services Bulletin) [38]. The MS also sends an *Origination (Continuation)* message to the BS through the reverse traffic channel and receives an *Alert With Information* message, through the forward traffic channel, to apply ring-back to itself on the audio path. Once the called party answers, the MS receives another *Alert With Information* message to stop ring-back and begins to transmit/receive traffic frames.

Figure 3: Sample Call Flow for Mobile Originated Calls

3.2.2.2 Mobile Terminated Calls

In mobile terminated calls, as shown in Figure 4, the MS receives a *Page* message from the BS while monitoring the paging channel. The MS responds with a *Page Response* message through the access channel. The BS then sets up the forward traffic channel and begins sending *null traffic data* frames. The MS receives a *Channel Assignment* message through the paging channel, sets up the reverse traffic channel, and begins sending the *traffic channel preamble* frames. The BS acquires the traffic channel and sends a (*Base Station Acknowledgment*) *Order* message to the MS through the forward traffic channel. The MS begins to transmit null traffic data and receives a (*Service Option Response*) *Order* message through the forward traffic channel. The MS starts to process traffic according to the service option parameter. The MS also receives an *Alert With Information* message and starts ringing. Once the call is answered, ringing stops and the MS sends a (*Connect*) *Order* message to the BS through the reverse traffic channel and begins to send/receive traffic frames.

Figure 4: Sample Call Flow for Mobile Terminated Calls

3.2.3 The System Overview

The IS95 standard is based on a packet switching system operating in the 800 MHz range. The packets¹ are 20 ms long and are communicated via base-stations with 1.23 MHz bandwidth for transmission and 1.23 MHz bandwidth for reception. The wide bandwidth is required for the Direct Sequence Spread Spectrum technology used in Code Division Multiple Access because the transmission rate is much faster in time (spread in frequency domain over a larger spectrum) than the information rate, due to coding and added redundancy. IS95 is a variable-rate system with information bit-rates of 9600, 4800, 2400, and 1200 bps. The systems spreading gain is 128 to 1024, depending on the bit-rate. For brevity, this overview is limited to the forward and reverse traffic channels of the system.

Block diagrams of the forward traffic channel and the reverse traffic channel are shown in Figure 5 and Figure 6 respectively [15]. Both forward and the reverse traffic channels utilize convolutional encoders and interleavers for data spreading and redundancy; a short description of these blocks are noted below.

1. referred to as frames in this thesis

Figure 5: The Forward Traffic Channel

Figure 6: The Reverse Traffic Channel

3.2.3.1 Convolutional Coders

Both the forward and reverse traffic channels use convolutional coders of constraint length of 9 as shown in Figure 5 and Figure 6. These encoders perform modulo-2 addition on selected taps of a serially time-delayed shift register as shown in Figure 7. Every input bit produces a 2-bit and a 3-bit output code word through the 1/2 rate and the 1/3 rate encoders of the forward and reverse traffic channels respectively.

Figure 7: Block Diagram of A Convolutional Encoder

The base-stations' generator functions (Forward Channel Encoder) are as follows in octal

format:

$g_0=753$ Octal, which translates to 1 1110 1011 in Binary, and
 $g_1=561$ Octal, which translates to 1 0111 0001 in Binary notation.

The mobile's generator functions (Reverse Channel Encoder) are as follows in octal

format:

$g_0=557$ Octal, which translates to 1 0110 1111 in Binary,
 $g_1=663$ Octal, which translates to 1 1011 0011 in Binary, and
 $g_2=711$ Octal, which translates to 1 1100 1001 in Binary.

3.2.3.2 Interleavers

Both forward and reverse traffic channels utilize interleavers to scatter possible bit errors in the frame and avoid error bursts as result of fading. The scattered errors may be corrected by forward error correction capabilities of the convolutional decoder. The read and write operations of the interleavers are shown in Figure 8 for the forward and reverse traffic channels. Both interleavers hold 20 *ms* length of data. The mobile's interleaver is 32x18 bits while the base-station's interleaver is 24x16. The mobile's interleaver writes the data in columns and reads the data out in rows. Although the base-station's interleaver writes and reads the data in columns, it performs the reading task in an encrypted scheme as shown in Figure 8.

For better description of the IS-95 system, the rest of the components of the block diagrams are explained separately under Forward and Reverse Traffic Channel headings.

Figure 8: The read and write operations of the forward and reverse traffic channels' interleavers

3.2.3.3 The Forward Traffic Channel

The coded data symbols of the forward traffic channel from the interleaver are scrambled by a 1/64 decimated long-code sequence, which is generated at 1.228 Mcps, masked with the mobile station's Electronic Serial Number (ESN). The power control sub-channel operates at one bit every 1.25 ms (800bps); the bit is to signal the mobile station to increase or decrease its output power.

The data symbols are then spread by a 64-bit Walsh code, designated to the forward traffic channel. Following the (Walsh) orthogonal spreading, the symbols are transmitted through a QPSK modulator. In the quadrature and the in-phase branches of the modulator, the data chips are masked by the corresponding PN sequence of the base-station, and filtered by a symbol shaping baseband filter. The correlation characteristics of the PN (short-code), inherent by m-sequences, ensure the transmitted data (and its delayed versions) are non-correlated to each another and to those of other base-stations. The PN code is generated via the linear feedback shift register shown in Figure 9 with the following generator polynomials:

$$P_I(x) = x^{15} + x^{13} + x^9 + x^8 + x^7 + x^5 + 1 \quad \text{for the in-phase sequence, and} \quad (\text{EQ } 5)$$

$$P_Q(x) = x^{15} + x^{12} + x^{11} + x^{10} + x^6 + x^5 + x^4 + x^3 + 1 \quad (\text{EQ } 6)$$

for the quadrature sequence.

3.2.3.4 The Reverse Traffic Channel

In the reverse traffic channel the coded data symbols from the interleaver are read in groups of six-symbol words, each of which is mapped to a 64-bit Walsh code [15] indexed by

$$index = b_0 + 2b_1 + 4b_2 + 8b_3 + 16b_4 + 32b_5 \quad (\text{EQ } 7)$$

where b_0 is the oldest binary valued code symbol, and b_s is the most recent (the last in).

The 64-ary mutually orthogonal waveforms are randomized by the masking patterns of the data burst randomizer on the redundant data of the lower rate symbols, when repeated for variable rate transmission. Then the symbols are (modulo-2) multiplied to the long code masked with the mobile station's Electronic Serial Number (ESN), which is unique to every mobile in the cell. For this multiplication every symbol is repeated 4 times to match the spreading factor of the long code. The long-code is generated via the shift register and the masking algorithm presented in Figure 9. The modulo-2 feedback polynomial of the long-code shift-register is as follows:

$$P(x) = x^{42} + x^{35} + x^{33} + x^{31} + x^{27} + x^{26} + x^{25} + x^{22} + x^{21} + x^{19} + x^{18} + x^{17} + x^{16} + x^{10} + x^7 + x^6 + x^5 + x^3 + x^2 + x + 1. \quad (\text{EQ } 8)$$

The existing element of the shift register is fed back to the elements represented by the polynomial of the above equation with modulo-2 addition.

The symbols are then transmitted through an O-QPSK modulator. In the quadrature and the in-phase branches of the modulator, the data chips are masked by the corresponding PN sequence of the base-station, and filtered by a symbol shaping baseband filter.

Figure 9: Long-Code and Short-Code (PN) Generators

Chapter 4: Dynamic Channel Allocation in IS95

4.1 Synopsis and Description

In Chapter 1, the review of the Wireless Telephony highlighted the demand of video as a new service in the next generation cellular networks. Inverse multiplexing was reviewed, examples were presented, and its role in wireless video telephony was discussed against current implementation of voice networks. An overview of CDMA, and its advantages, were presented as the multiple access scheme of choice for this project.

In Chapter 2 real-time video signals and models were described along with the IS95 standard system specification in Chapter 3 for the purpose of simulation and modeling in this project. The channel environment, and its corresponding receiver properties were also reviewed

In this chapter a concise statement of this thesis is presented and established in light of the previous chapters' overview and background. This thesis demonstrates, via simulation, that video data may be transmitted through inverse multiplexing of existing voice channels of a CDMA system governed by IS95. It demonstrates that wireless video transmission with this method is efficient with respect to the use of the voice traffic channels. It observes that the inverse multiplexed video data requires less operations and maintenance to the receiver than current audio signals do. It concludes, through several test-case scenarios, how the capacity of the current system may be affected with video user(s) roaming in the cell.

4.2 Motivation

The driving force in this project has been the upcoming migration towards Personal Communication Systems and the next generation wireless cellular networks [1]. The services demanded by the industry and offered in PCS are still under research and very few are developed. The ones under development are mainly for voice applications [39]-[41].

Authors of [39] and [40] simulate and study the reverse traffic and forward traffic channels of a CDMA system, respectively, for capacity purposes with diversity and power control techniques. They observed that “the path loss exponent has a major impact on the system capacity”; it gets reduced significantly when path loss exponent is decreased from 3 to 2. Capacity was defined as the maximum number of variable rate voice users for a target BER.

In [41] a technique is observed for coherent detection on the uplink channel for the European RACE project R2020 Code Division Testbed. The authors separate control data (frame ID) and user data in two physical channels on the same frequency with different spreading sequences. Control channel's data rate is fixed at 4Kbps whereas the user data (speech) rate is variable. The user-data's spreading factor is determined from the multiplexed stream of data and is fed to the control channel. The information bits on the control channel are detected with high reliability and powerful coding. These bits are also used for uplink channel estimation which allows for coherent detection.

CDMA, a candidate for the PCS access, is a versatile technology for wireless mobility problems of establishing a reliable communication channel. Moreover, it provides privacy and higher capacity. However in the literature CDMA applications for video data

are very limited [2],[21].

In [2] a wireless multimedia network is observed that employs CDMA to transmit low speed human interface signals on the uplink, and TDMA to transmit high speed video data on the downlink on the same frequency band in an indoor multipath Rayleigh fading environment. The authors take advantage of the unbalanced data rates; the impulse response of the channel is determined for each symbol interval of the uplink and when the chip rate of the downlink is equal to the symbol rate of downlink, the determined impulse response is applied to the downlink. The FEC encoded CDMA signal is transmitted in a time slot of the TDD frame.

The closest work to this project was presented in [21]. However, the methods applied for the video transmission were not through inverse multiplexing. Instead, the authors used an Automatic Repeat Request (ARQ) protocol for frame retransmissions in case of error detection. They further concluded that ARQ method is very efficient that for the same performance, forward error correction techniques, such as Reed-Solomon, require extensive amount of overhead.

In summary, a wireless variable rate video telephony with an IS95 CDMA system has not been demonstrated nor documented in the literature before with inverse multiplexing technique. In this thesis, this task is accomplished with existing and developing standards (H.263 and IS95) which is of significant value for industry use and implementation.

4.3 Applications

The issues under study here are the system behaviors of adding video users to a cell in a CDMA cellular network with other voice users. The research aim in this project is a

comparative study of interference and capacity issues in a cell when a mobile uses a video terminal as the medium of communication. In the following chapter the system implementation and the results of the research are presented to shed some light on this focus.

It is observed that video coders' rate increase, during background scene changes, is met with a number of dynamically allocated traffic channels to transport the data to its destination. At the receiver, the data is put back in sequence and is passed to the video decoder. It is demonstrated that IS95 protocol is capable of accepting and transporting frames with video telephony images; the protocol is modified to take advantage of Inverse Multiplexing for video transmission.

Chapter 5: System Simulation

5.1 Perspective

In this chapter, inverse multiplexing of traffic channels of an IS95 system are implemented for video telephony. Simulation design and assumptions are presented below with detailed explanations on iterations, setup, and calculations for the results which follow in the next chapter. The simulation was written and executed in MATLAB under different scenarios. All signals and system components were represented in their low-pass equivalent format.

5.2 System Design and Assumptions

We consider a single radio cell in which mobile handsets communicate with a single base-station according to the IS95 CDMA standard. The handsets can generate audio or video signals. A very low bit-rate video compression scheme (as proposed in H.263) is assumed whose codecs' maximum bit rate is 64Kbps. The mobile-station and base-station transmitters are designed to adapt to the variable rate video data by inverse multiplexing of the IS95 traffic channels (9.6 Kbps each). The corresponding receivers are also designed to collect the data and output it sequentially as it was transmitted.

Many assumptions were made in the preparation and implementation of the simulation programs. These have been explained in the designated sections of the system component about which the assumptions are made. However in this section, some fundamentals and specifications are presented and observed based on which the simulated system functions.

5.2.1 Call Setup

In Chapter 3 a review of the IS95 protocol was presented along with message flows for call setup through the reverse, access, forward, and paging channels. The Service Option parameter plays an important role in establishing a traffic service for the connection between the MS and the BS. For example, Service Option with value 1 is designated for 2-way variable rate speech service. This section presents areas and parameters of the IS95 call setup procedure which need to be modified in order to support mobile and base-station video transceivers.

5.2.1.1 Dynamic Reverse Channel Allocation

When a mobile station is registered as a video user, it does not occupy extra resources in the cell until it has established a call and has bursty traffic which require the extra traffic channels. The Service Option Parameter of the *Origination* message can be modified to indicate the type of traffic that is being requested for transport, i.e. video telephony calls. The call can be set up with the allocation of 8 traffic channels to the video user as identified by the Service Option parameter. These traffic channels are only used dynamically based on the video coder's instantaneous rate. In reply to the *Origination* message on the Access Channel, the BS sends a *Channel Assignment* message on the paging channel with the assigned channels' unique user long code sequences.

The eight channels in the project were chosen to match the peak rate of the video coders, 64Kbps. The assumption is that IS95 traffic channels would operate at its maximum rate of 9.6 Kbps. Therefore a video encoder with a peak rate of 64 Kbps would require up to 7 traffic channels. The number 8 was chosen for possible overflows, and as a worst case scenario from a system resource perspective.

5.2.1.2 Dynamic Forward Channel Allocation

The base-station can request a specific Service Option while paging the mobile station. The mobile station can accept or reject the Service Option. It can also request a new Service Option through the *(Service Option Request) Order* message; the service negotiation may continue till an agreement is reached via the *(Service Option Response) Order* message. Therefore the call may be setup by the BS for up to 8 forward channels. The channels' unique Walsh codes can be sent with the *Channel Assignment* message. The mobile station, in response, communicates its long code masks (Private_LCMs) to the BS.

Based on the discussion above, the system can allocate 8 channels at the call-setup stage but it uses the channels based on the video encoders' rate, so that the average dynamic interference of the cell is minimized. If more than one channel is used, the frames are sequence-stamped and the information contained in the generated frames are multiplexed through additional traffic channels in parallel. For instance at a 4 x 9.6 Kbps mode, the first frame is sent through the first channel, and the 4th frame through the 4th channel to avoid buffering and real-time delay in long queues.

5.3 Video Codec

The video source is modeled as 16 independent and identically distributed two-state continuous-time Markov chains [7],[20] as shown in Figure 1 in Chapter 2. This model provides the two important correlation characteristics of video signals as reported in [7]: the short-term exponential-decaying correlation characteristics of video data from uniform activities in the image (few hundred milliseconds decay), and the long-term correlation from sudden scene changes which increase the rate of arrival (in the order of

few seconds decay). The interval time between scene changes is assumed to be exponentially distributed:

$$p_x(x) = \lambda e^{-\lambda x} \quad x \geq 0, \lambda > 0 \quad (\text{EQ 9})$$

where $1/\lambda$ is the mean, and $1/\lambda^2$ is the variance. Every source is assumed to be in the ON state, independently, 32% of time [20]; this parameter is referred to as the Activity Ratio (AR) and was chosen from typical multimedia modeling parameters presented in [20] and its references. With the above discussion, the video packet generator's parameters are summarized as the following:

- The number of sources in the model $M=16$ @ 4Kbps each, which result in maximum rate of 64Kbps,
- The Activity Ratio, $AR=32\%$, and
- The ON duration, $T_{ON}=20$ ms, the IS95 frame length duration.

The activity ratio was determined from [20] based on actual traffic parameters. In the simulation for video traffic generation, the mean was set to 0.32. In other words, if the random number (with exponential distribution) happened to be less than the AR, the corresponding mini Markov process, in the ON state, would generate random binary data at constant rate of 4Kbps (bits for approximately half of an IS95 frame); and if the number were more than the mean, the process would shut off. The ON and OFF durations of a mini Markov source are geometrically distributed:

$$P_k = P(1-P)^k \quad k = 0, 1, 2, \dots \quad (\text{EQ 10})$$

with mean $(1-P)/P$ and variance $(1-P)/P^2$, where P is the probability of being in ON state, and k is the number of trials before which the ON event occurs. The output of this module of the simulation was to dictate the instantaneous number of IS95 frames

generated by the source, as shown in Figure 10. As explained below, every frame was set to be filled with binary random data in the program.

Figure 10: The Output of the VBR Video Source

5.4 The Program

Several test-cases were created to run the Matlab programs under different scenarios that would push the system to its capacity limits. The simulations were run for 5 seconds of real-time transmission of video data, approximately 250 IS95 frames (43 Kbits of compressed video data). All simulations consider audio-users and some consider audio- and video-users as interferers. The users are assumed to start transmission at the same

time. Transmitter and receivers are perfectly synchronized and the receivers have perfect estimates of the channel for coherent detection.

A block diagram of the software modules in the simulation is presented in Figure 11.

The **Setup & Iterator** module staged the environment for the simulation to run. Its functions included assigning the number of users in the simulation, designating unique user masks, and preparing a generic matrix of 2^8 input combinations to the convolutional encoder states. This matrix is used in the Viterbi decoder as its steady-state list of possible input states. The user-masks of the reverse link (42 bits long) were chosen from the rows of a (48 by 48) Hadamard matrix to identify the reverse traffic channels. The user-masks of the forward link (64 bits long) were chosen from the rows of a (64 by 64) Hadamard matrix to identify the forward traffic channels; since only one cell was assumed for the simulation, the Walsh codes were used to identify different users in the cell, instead of applying the mobiles' Electronic Serial Numbers (ESN) as masks on the long code.

This module of the program was also responsible for managing the Case Operators (transmitting modules) based on the video coder's rate. For example in an instance during a change of a scene, when the data-rate is high, seven 172-bit frames are given to the inverse multiplexer, Case Operator number 7. Immediately, 7 masks are used to deliver the data through 7 traffic channels within one 20 *ms* time interval. After collecting the bit-error rate and interference power from the receiver module, the program ends the simulation.

Figure 11: The Software Flow Diagram of the Simulation

A **Case Operator** creates random binary input data for every user in the simulated scenario supported by that case, as dictated by the Video Source Generator through the Setup & Iterator module. This module then resets the state of the shift-registers responsible for short-code (PN) and long-code, and calls the transmitter module for the IS-95 operations explained in Section 5.5. The chip stream sequences of every user in the scenario is convolved through independent complex samples of a Rayleigh

distributed multipath channel explained in Section 5.6. The transmission power of every user is obtained by multiplying every users' complex data with its conjugate for interference and power control calculations. Then the average power of all the users (or in the case of video-users, every channel) are normalized based on perfect power control assumption. The power control in the forward link is performed very slowly, one update per frame every 20 ms, for the near-far positioning of the mobile stations. This assumption is clearly separate from the fast closed-loop power control in the reverse link, which also contributes to signal recovery against fast fading.

The program then combines the transmitted signals in the pseudo air interface. The transmitted signals with CDMA act as interferers in the air medium against others and the user itself. The interference from audio and/or video users, additive white Gaussian noise (AWGN), and the signal itself (from the user under study) are accounted for in the signal to (total) interference ratio (SIR) calculations. In the case of multipath channel, the total signal-to-interference ratio is the sum of the SIRs of all branches of the RAKE receiver [30].

$$SIR = 10.\log \left[\sum_{k=1}^3 SIR_k \right] \quad (\text{EQ 11})$$

$$SIR_k = \frac{Signal_k}{Signal_k|_{AutoCorrelation} + Noise_k + Interference_k|_{CrossCorrelation}} \quad (\text{EQ 12})$$

The auto-correlation terms include the correlation of each resolvable ray of the same signal with the receiver's synchronized code, whereas the cross-correlation terms include the correlation of each resolvable ray of each interferer with the receiver's synchronized code. For example, the signal-to-interference ratio of the second path, SIR_2 , from a three finger RAKE receiver is calculated as:

$$SIR_2 = \frac{Signal_2}{Signal_1 + Signal_3 + Noise_2 + Interference} \quad (EQ 13)$$

For video users, $Signal_2$ is assumed to be the sum of the second rays of up to 8 channels which may be utilized for the transmission of the desired signal. All multipath rays of the interferers and the first and third multipath rays of the desired signal act as interference for the second ray which (is assumed to) get recovered by the second branch of the RAKE receiver. The SIR of the individual branches are added when the receiver performs maximal ratio combining¹.

The AWG Noise was generated from a normal distributed random process with zero mean and unity variance. The *Noise* relative power was calculated from the random stream multiplied by an arbitrary value that was set to 0.16. This value, obtained experimentally, was large enough to add additive noise to one audio user data (zero interference) for an average SIR=-0.25dB, and was small enough to be insignificant amongst interference. The *SIR* was calculated and stored for every 20 ms iteration and then time averaged over the number of iterations.

The combined coding and spreading gain of both the reverse link and forward link are 128 which correspond to $10 \cdot \log(128) = 21.07$ dB gain after despreading and decoding. In the reverse link, the effective spreading gain is 32 which contributes to interference immunity by 15.05 dB, whereas the coding gain lowers the minimum required SIR for reception. The the effective spreading gain of the forward link is $10 \cdot \log(2) = 3$ dB.

Every traffic channel's data, corresponding to the user under study, is then received by

1. To compensate for the phase shift in the channel and to weight the signal by a factor that is proportional to the signal strength, the maximal ratio combiner multiplies the received signal to the corresponding complex-conjugate channel gain [28].

the Receiver, explained in Section 5.7, and checked for errors against the corresponding original random binary data before transmission. The sum of errors is divided to the total number of video-user's bits (for example for seven IS95 frames) which were transmitted and received in 20 ms. A summary of the bit error rate is also sent back to the Setup & Iterator module.

5.5 Transmitter

The **Transmitter** module is called by a Case Operator to prepare the user data according to the block diagrams of the forward traffic channel and the reverse traffic channel shown in Figure 12 and Figure 13 respectively. The diagrams, as shown, are slightly modified from the standard for the purpose of this thesis. The standard's modulation format is QPSK through an inphase and a quadrature signal summation, whereas for simplicity in the project, the modulated signal is BPSK. Since QPSK and BPSK result in the same probability of error (as explained in Chapter 2), choosing BPSK allows for keeping the receiver complexity low. Also in the diagrams the feedback and feedforward links related to power-control are omitted because perfect power control is assumed.

Both forward and the reverse traffic channels utilize convolutional encoders and interleavers for data spreading and redundancy; a short description is noted below.

Figure 12: The Forward Traffic Channel

Figure 13: The Reverse Traffic Channel

5.5.1 Convolutional Encoder

The encoders implemented in the simulation were identical to those described in the IS95 standard (also Chapter 3). The implementations were performed through matrix manipulations of the respective generator functions [42].

5.5.1.1 The Forward link Encoder

The base-stations' generator functions were presented in Chapter 3. The generator matrix for this encoder may be written as

(EQ 14)

$$G = \begin{bmatrix} 11 & 10 & 11 & 11 & 01 & 10 & 00 & 10 & 11 \\ & 11 & 10 & 11 & 11 & 01 & 10 & 00 & 10 & 11 \\ & & 11 & 10 & 11 & 11 & 01 & 10 & 00 & 10 & 11 \\ & & & \ddots & & & & & & & \\ & & & & 11 & 10 & 11 & 11 & 01 & 10 & 00 & 10 & 11 \end{bmatrix}$$

Every row in the general format is written as

$$w = \{a_{1g1}a_{1g2} \quad a_{2g1}a_{2g2} \quad \dots \quad a_{ig1}a_{ig2}\} \quad (\text{EQ 15})$$

where a_{igm} is the i th symbol of the m th generator function. Every row is padded on the header or the trailer by a multiple of n (in this case $n=2$ for 1/2 rate encoder) up to the maximum length of the input data that is being encoded. In other words the size of the matrix G is

*the length of the input data (number of rows of G)-by-the
length of $w+(n*(\text{length of input data}-1))$ (number of columns of G).*

The statement in the parentheses, $(n*(\text{length of input data}-1))$ is what is used for padding w in the matrix G . This register has a constraint length of 9; it is a $k=1$ bit

register, meaning that it shifts the data 1 bit at a time; and $n=2$ for the number of generators, or the number of output bits for every k input bits.

The coded data is the result of multiplying (in modulo-2) the input binary sequence and the generator matrix G , plus ignoring the data past the $n*(length\ of\ input\ data)$ element of the result. So in this case, if we consider the contents of the interleaver matrix as the output of the encoder, a 192-bit input to the encoder, would result in a 384-bit array:

$$1-by-192 * 192-by-(9*2)+2(191) = 1-by-400 \rightarrow 1-by-384$$

([9-1]*2=16 bits are ignored).

5.5.1.2 The Reverse Link Encoder

The same equivalent matrix operations were performed for the reverse link encoder. The mobile's generator functions were presented in Chapter 3. The generator matrix for this encoder may be written as

(EQ 16)

$$G = \begin{bmatrix} 111\ 011\ 101\ 110\ 010\ 101\ 100\ 110\ 111 \\ 111\ 011\ 101\ 110\ 010\ 101\ 100\ 110\ 111 \\ 111\ 011\ 101\ 110\ 010\ 101\ 100\ 110\ 111 \\ \ddots \\ 111\ 011\ 101\ 110\ 010\ 101\ 100\ 110\ 111 \end{bmatrix}$$

and the 192-bit input to the encoder, would result in a 576-bit array:

$$1-by-192 * 192-by-(9*3)+3(191) = 1-by-600 \rightarrow 1-by-576$$

([9-1]*3=24 bits are ignored).

5.5.2 The Interleavers

The interleavers in the project were identical to those of IS95 at maximum data rate of 9600 bps. As with the rest of the simulation, the interleavers were also implemented

through matrix manipulation of the data in MATLAB. The read and write operations were implemented according to the specifications noted in Chapter 2. The interleaved data of the forward and reverse traffic channels were handled by different Transmitter modules, therefore they are explained separately.

The Forward Traffic Channel Transmitter

The coded data symbols of the forward traffic channel from the interleaver were spread, with a factor of 64, and masked with a 64-bit Walsh identity of the base-station. In this simulation, since only one cell was assumed, the Walsh codes were also used to identify different users in the cell, instead of applying the mobiles' Electronic Serial Numbers (ESN) as masks on the decimated long code. The Walsh codes were utilized for their orthogonal characteristics which ensure the transmitted data is (almost) non-correlated to those of other users. The PN (short code) is used for its auto-correlation characteristics [28],[37]. These characteristics, inherent by the m-sequences, ensure the transmitted data (and its delayed versions) are non-correlated to each another; the PN code was generated via the shift register shown in Figure 9 of Chapter 3.

The spread data symbols are then modulated by BPSK and sent through a time-varying frequency-selective Rayleigh-distributed channel with three multipaths. The IS95 modulation schemes (QPSK and O-QPSK) were modified for the purpose of this thesis to be BPSK and equivalent receivers were designed for this purpose which are explained in detail in Section 5.7.

The Reverse Traffic Channel Transmitter

Following the interleaver, the data symbols are mapped (in groups of 6) to one row of a

64x64 Hadamard matrix for orthogonal modulation. The 64-ary mutually¹ orthogonal waveforms are then spread, by a factor of 4, and (modulo-2) multiplied to the long code, masked with the mobiles' Electronic Serial Number (ESN), which is unique to every mobile in the cell. The long-code is generated via the shift register and the masking algorithm shown in Figure 9 of Chapter 3. The masked and spread data symbols are then modulated in BPSK format and sent through a time-varying frequency-selective Rayleigh-distributed channel with three multipaths, explained in Section 5.6.

5.6 Rayleigh Distributed Multipath Channel

The Rayleigh distributed multipath fading channel samples were simulated based on two zero-mean, unity variance, independent identically distributed Gaussian random variables X_1 and X_2 [28]:

$$Y = X_1 + jX_2 \quad . \quad (\text{EQ 17})$$

The probability distribution function (PDF) of Y is given by:

$$p_Y(y) = \frac{1}{2\sigma^2} e^{-y/2\sigma^2} \quad (\text{EQ 18})$$

where σ^2 is the variance of X_1 or X_2 , and $\sigma^2 \sqrt{X_1^2 + X_2^2}$ is the power of one multipath ray channel sample.

The σ^2 portion of the rays' power were considered to be a vector whose number of non-zero elements were equal to the number of the multipath rays considered in the simulation, i.e. three. This was done to ensure an exponential decay of the received multipath powers through the simulated Rayleigh channel, as shown below:

-
1. Authors of [43] report a 2.38% difference in their simulation results using an 8x8 Hadamard matrix instead of the 64x64. However, in this project this was avoided for the purpose of keeping all structural assumptions as close to IS95 as possible.

$$\sigma^2 = \begin{bmatrix} 1 & e^{-\frac{1}{DS}} & e^{-\frac{2}{DS}} & 0 & 0 & 0 & 0 \end{bmatrix} \quad (\text{EQ 19})$$

where DS is the delay spread of the channel, which was set to 7 chip-durations long, or 5.7 microseconds.

5.7 Receiver

As explained in Chapter 2, the IS95 standard describes the transmission of CDMA frames and does not specify the receiver structure. In this simulation, the reception and decoding tasks were performed in the reverse order of the transmission, as shown in the block diagrams of Figure 12 and Figure 13.

A RAKE receiver was used, with 3 fingers, in the form of a tapped delay line, to collect the energy of the multipath signals from the frequency selective channel of the traffic links, as shown in Figure 12, and Figure 13. It was assumed that the base-station was perfectly synchronized with the designated mobile via Costas loops and Early-late Gate synchronizers, or via user specific pilots as in [41]. The number of fingers chosen for the RAKE receiver and the number of multipaths in the channel were not necessarily related; however in practice, increasing the number of fingers in the RAKE would have a diminishing return for the extra added complexity. In this simulation, it was assumed that each delay contains a signal component, and that all three correlators contribute to the decision. In practice, adaptive algorithms are used to exclude low-level noise-only contributions from the tap correlators [28].

After despreading with the (synchronized) short-code and/or the long-code, the chip matrix of user-data was de-interleaved and sampled for detecting the convolutional code-word symbols. In the simulation of the reverse channel, the RAKE receiver performed a

soft-decision (inverse orthogonal) Walsh demodulation, at the last de-spreading stage, based on the shortest distance of the received chip values and the Hadamard sequences. This technique allowed for higher signal-to-interference ratio for the decision device to operate, because the chip sequence of the fingers were first combined in soft-format and then mapped to the hard-formatted code word symbols. The coded symbols were then input to the Viterbi decoder to obtain the data bit stream of every frame.

5.7.1 Viterbi Decoder

The Viterbi decoder was implemented based on the algorithm explained in Chapter 2. At steady state, every input combination (node) was assigned a weight according to its corresponding output comparison with the received code-word. Then through the window of decision, which was assumed to be $9 \times 5 = 45$ stages of the trellis, and in the reverse order of the trellis, the corresponding node weights were added and the survived path of every state was determined - whether it was from an inputted one or a zero. Then at the beginning of the trellis, the path corresponding to the state with the largest cumulative weight was chosen as the state whose input is the decoded bit. This algorithm was repeated for every bit of the IS95 frame as the window of decision was shifted forward by one code-word (stage) at a time. The coded data was originally padded by deterministic code words before the decoder, in order to obtain the last bit of the frame. The padding consisted of a series of code-words resulting from a $45 + 9 - 1 = 53$ bit stream of ones inputted to the Convolutional Encoder.

5.8 Test-Cases Generation

The software architecture of the simulation was designed with the intention of minimizing the required changes to the program for different test-case scenarios. The

module which controlled the scenario of the simulation in a test-case was the Case Operator (Figure 11) which consisted of 8 (to 64) sub-modules that the Setup & Iterator would execute. Each sub-module was designed to perform transmission, reception, error detection, and BER calculation for one 20 *ms* period of time with its own portion of the designated characteristics of the scenario - because the number of the traffic channels assigned to the video-user changed dynamically frame-by-frame.

Chapter 6: Simulation Results

6.1 Introduction

In this chapter, inverse multiplexing of an IS95 system traffic channels is observed. It is demonstrated that such technique is reliable for video telephony in the simulated system. Simulation results are presented below from several test-cases with corresponding discussions that lead to a few conclusions and observations about the system with mobile video users.

6.2 Simulation Results

In the previous chapter the system-level design and assumptions of this project were presented. In this section the results of Matlab simulations based on the established system and assumptions are presented with discussions.

6.2.1 Non-Coherent Detection of the Reverse Link

The first hurdle in this research was to establish a reliable communication channel in a sense that the BER would stay below (or in the order of) 10^{-3} with a reasonable number of users as interferers¹. The non-coherent RAKE receiver of Figure A6 in APPENDIX A was not capable to recover the DPSK-coded user-data chips reliably without power control in the reverse link. The simulation was run with 10 iterations for 2 users (non-coherent detection) and the resulting average BER was 2.8×10^{-2} . The same simulation for 5 users gave random errors of 50%. It was evident that the receiver lost the data under medium-level interference even in fast fading cases where power control is not

1. The overall mean error tolerance of H.261 is documented as 0.0015 for accuracy.

required. In this reception scheme, the interference would bury the spectrum of the user-data so deeply that non-coherent detection could not de-spread it and retrieve it out from the interferers' spectrums. Therefore coherent detection was used in the rest of the test-cases instead. After the switch to the coherent RAKE of Figure 13, Chapter 5, the average decreased by an order of magnitude even when the simulation was run for 10 users.

6.2.2 Test-Case Scenarios

The results presented in this section were acquired and summarized from the Matlab simulated test-cases of Tables i and ii. Results of every one of the test-case scenarios is explained and shown in detail in APPENDIX B. The results are summarized in 2 sets of tables for the forward traffic and reverse traffic channel environments, respectively. The test-cases were created with the purpose of finding capacity limits of the system with variable-rate video roamer(s). Several scenarios were compared against the results from [18][21][31][33] for verification of the simulation. The results are not in any particular order, and they are presented here for the discussions in the sections that follow.

Table i: Result Summary for the Forward Link

Test-Case	Simulated Scenario	Ave. BER	Ave. ^a SIR _{dB}	Max. ^b SIR _{dB}	Min. SIR _{dB}
1F	1 video user with 0 audio interferers	1.5×10^{-3}	1.6964	1.7249	1.6119
2F	1 video user with 4 audio interferers	5.3×10^{-3}	-1.5531	-0.9180	-3.1015
3F	1 video user with 8 audio interferers	1.11×10^{-2}	-3.3800	-2.5103	-5.4498
4F	1 video user with 1 video interferer	6.8×10^{-3}	-2.2937	-0.6149	-4.3760
5F	1 video user with 0 audio interferers	0 ^c	1.6932	1.7493	1.6109
6F	1 video user with 8 audio interferers	7.55×10^{-4}	-3.5445	-1.1216	-6.6102
7F	1 video user with 16 audio interferers	7.5×10^{-3}	-5.8847	-3.2840	-9.7492
8F	1 audio user with 23 audio interferers	1.06×10^{-2}	-12.9716	-5.5275	-18.4107

a. Averaged over the SIR values of all iterations.

b. Maximum SIR value in the iterations.

c. Errors due to AWGN were corrected by the Viterbi decoder

High Doppler, Fast Fading

Slow Fading with Power control

Table ii: Result Summary for the Reverse Link

Test-Case	Simulated Scenario	Ave. BER	Ave. SIR _{dB}	Max. SIR _{dB}	Min. SIR _{dB}
1R	1 video user with 0 audio interferers	0	1.7043	1.7316	1.6262
2R	1 video user with 4 audio interferers	6.47×10^{-5}	-1.5514	-0.8821	-3.0397
3R	1 video user with 8 audio interferers	1.7×10^{-3}	-3.3758	-2.5210	-5.3368
4R	1 video user with 8 audio interferers (slow fading, no power control)	4.63×10^{-2}	-3.5409	-1.3682	-7.3896
5R	1 video user with 10 audio interferers	3.1×10^{-3}	-4.0606	-3.1694	-6.0042
6R	1 video user with 1 video interferer	7.76×10^{-4}	-2.2821	-0.6499	-4.2884
7R	1 video user with 1 video & 4 audio interferers	2.3×10^{-3}	-3.8750	-2.6356	-6.0902
8R	8 bursty audio users with 8 audio interferers	1.4×10^{-3}	-3.3766	-2.5333	-5.3466
9R	8 bursty audio users with 8 audio interferers (slow fading, no power control)	4.82×10^{-2}	-3.5408	-1.3724	-7.3971
10R	1 audio user with 15 audio interferers	1.2×10^{-3}	-11.9446	-11.6974	-12.1904

High Doppler, Fast Fading

Slow Fading with Power control

(Continued on next page...)

Table ii: Result Summary for the Reverse Link (... Continued from previous page)

Test-Case	Simulated Scenario	Ave. BER	Ave. SIR _{dB}	Max. SIR _{dB}	Min. SIR _{dB}
11R	1 video user with 0 audio interferers	0	1.7004	1.7568	1.6022
12R	1 video user with 8 audio interferers	2.80×10^{-4}	-3.5402	-1.3637	-7.3965
13R	1 video user with 16 audio interferers	1.03×10^{-2}	-5.8862	-3.4582	-9.0158
14R	1 audio user with 15 audio interferers	0	-12.4002	-7.0442	-22.5279
15R	1 audio user with 23 audio interferers	2.4×10^{-2}	-14.6021	-8.3941	-25.1675

High Doppler, Fast Fading
Slow Fading with Power control

6.2.3 Effects of Power Control

The simulation of testcases 11R, 12R and 14R of Table ii are the modified versions of testcases 1R, 3R, and 10R of Table ii, respectively. These testcases were simulated with perfect power control to explore its effects on the base-station's receiver. Because the received data originates from different mobiles (audio-users), maintaining control over every IS95 frame of every user makes the video-user more *receiver friendly* to the base-station. It is observed that the base-station is not required to perform operations and maintenance activities, such as power control, on a "per traffic channel basis". With video roamers in the cell, this activity is reduced to "per user basis".

Four test-cases were considered for comparison: 4R, 8R, 9R, and 12R. While simulating with one video-user and 8 audio-users, testcase 12R, the voice handsets sent their data at all times and acted as interferers for the video data from different parts of the cell.

However, all the channels allocated to the video-user shared the same originating location in the cell. While simulating with the bursty-attendant¹ audio-users, testcases 8R and 9R, at least 8 audio-users acted as interferers constantly, and the signals from the bursty-attendant users, who randomly appeared, were transmitted in bursts. The purpose of this comparison was to see the effect of multi-location data bursts on the base-station's power control, and reception.

In the simulation, perfect synchronization was assumed. In fact, the state of the shift-registers for the long-code and short-code (PN) were reset after every 20 ms of transmission during which 1 to 8 user frames would be sent through the channel. Test-

1. The same video burstiness was used on the audio users' attendance in the cell for comparison reasons. So at a time when the video user transmitted and utilized 6 channels, the equivalent audio simulation ran with 6 users.

cases 3R, 8R, and 9R were run with no power control, whereas test-case 12R was simulated with perfect power control.

As mentioned above, the system under consideration consisted of one CDMA radio cell and one base-station. In the reverse link with slow fading (testcase 12R, with 50 Hz Doppler or 28 Km/hr speed for 1.9 GHz¹ carrier frequency) and power control, the channel was well suited and reliable for communication at a bit error-rate of 2.80×10^{-4} . However, the same video user without power control (test-case 3R) must speed up to 10,913 Km/hr in order to achieve a bit error-rate of 1.7×10^{-3} .

The error-rate for the case of 8 multi-location bursty audio users at high speed and with no power control, test-case 8R, is relatively the same as test-case 3R; it was measured at 1.4×10^{-3} . However, the reliability of the communication link drops to a bit error-rate of 4.82×10^{-2} if the audio-users slow down to 28 Km/hr without power control. Although it is an exaggeration to have no power control for 8 multi-location bursty audio users while communicating 252 IS95 frames in an environment with average SIR of -3.5 dB at the receiver, test-cases 12R and 9R portray that the base-station must improve the BER up to 2 orders of magnitude by controlling the transmission powers of 8 audio users at different locations. The power control information of (up to 8 channels of) a video user can be amalgamated to be sent to one location.

The increase in the performance of the forward link (test-cases 3F, and 6F) with the presence of power control is well noticeable. There is one base-station transmitter, and multiple mobile receivers. The BER is lower by 2 orders of magnitude when the base-

1. Although the operating frequency in IS95 is in the 800MHz range, the Doppler calculations are considered for the PCS applications at 1900 MHz range. The test-case simulation results are not affected by the carrier frequency.

station adjusts the powers of the individual frames prior to transmission, and does not favor one over others, but includes their near-far effects into the transmission. The average error rates in this study were 1.11×10^{-2} for test-case 3F and 7.55×10^{-4} for test-case 6F. Comparison of these test-cases with their reverse link counter parts (test-cases 3R and 12R) suggest that perhaps less coding in the forward link makes the resolution of the interferers more difficult. The error rate is consistently higher in the forward link when the same number of interferers are used in the simulation of both links.

6.2.4 Effects of Slow and Fast Fading

In all the fast fading test-cases, during the 20 ms duration of the frame the random multipath rays of the channel were changed 384 times in the simulation. In both the reverse and forward links, the channel remained the same for 0.52 μ s; this corresponds to a mobile speed of 10,913 Km/hr and 1,9200 Hz Doppler at 1.9 GHz carrier frequency. This high Doppler was chosen to apply sufficient redundancy to the system (as a black box) to account for diversity and power control. In contrast, for the slow fading test-cases 4R, 9R and 11R to 15R the channel was not changed during the period of the frame. The speed of the mobile was assumed to be 28 Km/hr (50 Hz Doppler). Slower changes of the channel result in long fade durations which make longer stream of data erroneous.

The test-cases 3R, 4R and 12R were considered for comparison where a video-user has 8 audio interferers. In high speeds (test-case 3R), the video user recovers the information reliably with a BER of 1.7×10^{-3} ; however, at slower speeds (test-case 4R) with no power control the error-rate increases to 4.63×10^{-2} . This is due to fact that the whole information frame falls in the fade, because the fade duration is also 20 ms. In other

words, the interleaver and the convolutional coder which are meant to randomize the errors [44] and make the symbols uncorrelated, become obsolete. It is also observed in [45] that the depth of the interleaver must be larger than the average fade duration to minimize correlation of the successive symbols, especially at low vehicle speeds.

Table iii provides the results of test-case 4R run at different mobile speeds; it shows the decrease of BER with higher Doppler. It is concluded that at low speeds power control and diversity are required to receive and to recover the data reliably, as it is evident by test-case 12R at $\text{BER}=2.8 \times 10^{-4}$.

Table iii: Test-case 4R run at different mobile speeds

BER	Speed (Km/hr)	Doppler (Hz)	Channel Transitions per Frame
2.8×10^{-4}	28	50	1 (with power control)
4.63×10^{-2}	28	50	1
3.29×10^{-2}	114	200	4
2.75×10^{-2}	227	400	8
1.7×10^{-3}	10,913	19,200	384

6.2.5 Channel Utilization and Efficiency

The results of the simulation suggest that inverse multiplexing is quite efficient for video transmission in cellular networks. Dynamic use of the system resources not only improves efficiency, but also decreases interference in the cell.

Average bit rate of the video-users of all the test-cases was found to be 52.6 Kbps.

Considering 45% voice activity [39][40], average bit-rate of audio users is estimated at

4.3 Kbps. A video-user's average is 12.2 times that of an audio user but it only occupies 8 times the number of resources. Moreover, a video-user is assigned up to 8 channels in a cell but the average number of utilized channels is found to be 6. This shows that the dynamic interference in the cell is also less when compared to 8 audio roamers.

In test-case 6R, two video-users were simulated with one acting as an interferer. In test-case 3R, 8 audio users provide interference to one video user. In test-case 6R the average SIR is higher by 1dB. An equivalent result may also be noted in test-cases 4R and 6R. This is directly related to the previous explanation that a video-user's channel utilization is more efficient than an audio user by 50%.

6.2.6 System Capacity

Most of the test-case scenarios were designed to explore the system's capacity limits. The results in Table i and Table ii allow for few concluding remarks on the capacity limits of the forward traffic and reverse traffic channels.

The forward traffic channel has lower capacity than the reverse traffic channel without power control. This is mainly due to the fact that the forward link uses less coding than the reverse link. Table i shows that the base-station can transmit the video data destined to one video user reliably with a BER of 1.5×10^{-3} and no power control. The other rates seem to be higher than the target error rate assumed in this project; test-case 2F (1 video-user with 4 audio interferers) has 5.3×10^{-3} BER when simulated without power control. Power control techniques improve these limitations to support scenarios such as 2F, and narrow down the performance margin of the reverse and forward links. The performance of the two links merges and becomes more similar at lower Doppler, where the fade duration is in the order of the frame length, and coding has diminishing return. Test-case

6F supports 8 audio interferers with power control at 7.55×10^{-4} BER. With power control, the mobile is able to recover the data with 16 interferers involved in the forward link at a BER of 7.5×10^{-3} . Considering the dynamic interference of another video user and the results from test-cases 4F and 6F, one could conclude that the forward link can reliably support two video users in the presence of up to 8 audio interferers.

The reverse traffic channel is able to serve up to 2 video-users and 4 audio-users simultaneously without power control at high Doppler with a BER of 2.3×10^{-3} , as shown in Table ii . One video user could communicate in a cell with up to 10 audio interferers reliably with a BER of 3.1×10^{-3} . Results of Table ii with power control show that it takes 16 audio interferers to cross over the target BER for reliable communication. In test-case 13R the BER was measured at 1.03×10^{-2} . Considering the dynamic interference of another video user, and the result of test-cases 6R and 12R, one could conclude that the reverse link can reliably support 2 video users in the presence of 6 to 8 audio interferers. Figure 14 and Figure 15 illustrate a summary of the test-cases with respect to their bit error-rates.

Figure 14: BER Graph of Table i (Forward Link)

Figure 15: BER Graph of Table ii (Reverse Link)

Chapter 7: Conclusion

The research aim in this project was a comparative study of interference and capacity issues of a single radio cell when a mobile uses a video terminal as the medium of communication. Chapter 6 presented the necessary results to show and to conclude that a video-user in a cell, with an IS95 CDMA system, employing inverse multiplexing to cope with the data-rate bursts, performs as well as an equivalent system with audio-users in both reverse and forward links. It is observed that in practice the video-users' data is more *receiver friendly* in the reverse link than the audio-users' data - friendly in a sense that the channels assigned to a video user do not require power control and synchronization individually. Because all transmissions originate from the same terminal, the corresponding channels may be power controlled and managed as one.

The total average interference in a cell with video-users is less than that of a cell with eight-times as many audio-users. This is due to the fact that at registration or call setup the video-user reserves (up to) 8 traffic links for inverse multiplexing, not all of which are utilized throughout its transmission. This dynamic utilization results in lower interference because audio users, even at 45% activity factor, must maintain the traffic channel with sub-rate frame transmissions. The channels (for video) are seized and the cell resources are occupied for an equivalent of (up to) 8 audio-users; however, they are maintained as one.

Power control is found to be crucial in the performance of slow-fading CDMA systems, especially in the reverse link where one mobile could potentially overload the base-station's receiver. Such systems require power control and diversity for data recovery and maintaining a reliable communication link. This requirement was evident as result of

fade durations that are similar in length as the 20 millisecond-long interleaver.

Average bit-rate of a video-user modeled in this thesis was found to be 12.2 times that of an audio user with 45% voice-activity factor. However, a video-user utilizes up to 8 channels for transmission.

In conclusion, two video-users may access and communicate through the forward and the reverse traffic channels with the presence of 6 to 8 audio interferers, at the cost of 3 to 6 audio channels. Researchers constantly attempt to increase the system capacity with more microcells in a PCS environment. These attempts may resolve possible blocking problems that an audio-user might face through a rejected registration resulted by a video user roaming in the cell and occupying related audio channels. With more cell coverage, CDMA audio users are rejected less frequently because of overlapped micro cells, and this opens the door for acceptance of a limited number of video users.

7.1 Future Research

Implementation of this application protocol for standards is left as a proposal for future research work. In fact, one suggested research area is optimization and standardization of a video codec with bit-rate increments of 9600 Kbps. This would enhance the dynamic channel allocation of the inverse MUX. This may be easier in the future since the IS95 standard, itself, is changing to higher rates with a maximum of 14.4 Kbps [TSB 74, 1995].

APPENDIX A: Background

A.1 Perspective

In this appendix; explanations on the basic characteristics and concepts of the multipath mobile fading channel are provided.

A.2 Mobile Communication Channel

The wireless channel, as opposed to the wireline, is affected and characterized by the external environment of the air medium. Movements of objects, signal reflections, and the mobility of the transmitter or the receiver can cause the signal to fade, become blocked, or even change because of doppler or destructive out of phase construction of the multipath reflections. The received signal in an urban environment experiences attenuation by distance, irregular envelope fluctuations, and deep fades. The envelope fluctuations are because of the movement of the mobile in the terrain of its travel, and the deep fades are due to the destructive out-of-phase additions of the signal reflections bounced from different barriers in the terrain. Thus the characteristics of the channel change in time, strength, and delay (of multipaths).

The power delay profile of the bounced reflections, or multipaths, is assumed to be exponentially distributed due to the propagation delay of the rays, as shown in Figure 1. The multipaths arrive at different times within the delay spread of the channel - a property of the environment related to the number of reflecting obstacles and their positions. For urban environments the delay spread is in the order of 1 to 5 μ seconds [27].

Figure A1: Multipath Reflections and the Power Delay Profile Model

Every one of the multipath received signals has a random and time-varying amplitude and phase which cause the received sum to add destructively at times. In such cases the received signal level becomes undetectable; this is called multipath fading. In this project the mobile communication channel is modeled as a Rayleigh¹ distributed random process that varies with time, has 3 multipaths, and a delay spread of 7 chips, along with Additive White Gaussian Noise(AWGN) [27]-[29].

A.2.1 Channel Characteristics

In general, the multipath Rayleigh fading channel is expressed by its equivalent lowpass impulse response as defined in [28]:

$$c(\tau, t) = \sum_n a_n(t) e^{-j2\pi f_c \tau_n(t)} \delta[\tau - \tau_n(t)] \quad (\text{EQ 20})$$

where f_c is the carrier frequency, $2\pi f_c \tau_n(t)$ is the time-varying phase of signal received on the n th path, $a_n(t)$ is the attenuation factor of the received signal on the n th path with delay $\tau_n(t)$, and $c(\tau, t)$ is the response of the channel at time t to an impulse applied at time $t - \tau$. When the impulse response is characterized by a wide-sense-stationary zero mean complex-valued Gaussian process, the signal envelope becomes Rayleigh distributed and the channel is referred to as a Rayleigh fading channel.

The time-dispersive nature of the channel due to multipath fading is characterized by the multipath power delay profile, as was shown in Figure 1. In practice, it is measured by transmission of a narrow pulse (wide band signal) and cross-correlating the received signal with a delayed version of itself:

1. In case of the presence of line-of-sight, the channel may be represented by the Ricean statistical model.

$$\phi_c(\tau_1, \tau_2; \Delta t) = \frac{1}{2} E[c^*(\tau_1; t) c(\tau_2; (t + \Delta t))] \quad (\text{EQ 21})$$

It is from this function where the delay spread of the channel is obtained, as shown in Figure 2.

The frequency coherence of the channel is characterized by the auto-correlation function of the Fourier transform of $c(\tau; t)$ (in frequency domain $C(f; t)$) equivalent to the time-domain characterization above:

$$C(f; t) = \int_{-\infty}^{\infty} c(\tau; t) e^{-j2\pi f\tau} d\tau \text{ and} \quad (\text{EQ 22})$$

$$\phi_c(f_1, f_2; \Delta t) = \frac{1}{2} E[C^*(f_1; t) C(f_2; (t + \Delta t))] \quad (\text{EQ 23})$$

In practice, it is measured by transmission of two sinusoids separated by Δf and cross-correlating the received signals with a delay of Δt . It is from this cross-correlation function where the coherence bandwidth of the channel is obtained, as shown in Figure 2. The coherence bandwidth is also the reciprocal of the delay spread. Due to the original assumption that $c(\tau; t)$ is a wide-sense-stationary zero mean complex-valued Gaussian process, the two correlation functions are also related by the Fourier transform:

$$\phi_C(\Delta f) = \int_{-\infty}^{\infty} \phi_c(\tau) e^{-j2\pi \Delta f \tau} d\tau \quad (\text{EQ 24})$$

where $\Delta f = f_2 - f_1$ and $\Delta t = 0$. The coherence bandwidth is the minimum spectral separation between two sinusoids' frequencies that makes the two uncorrelated and affected differently by the channel.

Analogous to the relationship of the coherence bandwidth and the delay spread, the coherence time and Doppler spread are also channel parameters that are obtained by the correlation function $\phi_c(\Delta f; \Delta t)$ with time variations measured by Δt . These two parameters

are also reciprocal of each other. A slowly changing channel is represented by a small Doppler spread which corresponds to a large coherence time. A visual relationship of these parameters with their respective functions are shown in Figure 2.

A.2.2 Channel Conditions

When the bandwidth of the transmitted signal is larger than the channel's coherence bandwidth, the frequency components of the signal go through different attenuation and phase shifts in the channel. In this case the channel is said to be frequency selective. In contrast, if all the frequency components of the signal go through the same attenuation and phase shifts, the channel is called frequency non-selective; in this case if the multipaths add destructively and the signal fades, the condition is referred to as flat fading.

Figure A2: The Statistical Relationship of the Multipath Fading Channel Parameters

A frequency non-selective channel is said to be slowly fading if the signal interval (or the symbol duration) is much smaller than the channel's coherence time which implies that the Doppler spread is small. If the signal interval is larger than the channel's delay spread, intersymbol interference distorts the signal.

When the signal bandwidth is smaller than the channel's coherence bandwidth the received signal "appears to arrive at the receiver through a single fading path". In contrast, through the wideband frequency selective channel of use in this project, certain rays may be received for diversity advantages to reduce the effects of fading. In CDMA access technology, utilization of spread spectrum and the RAKE receiver allow for taking advantage of such diversity potential.

A.2.3 Spread Spectrum and Power Control

Spread spectrum is a method of transmission in a frequency selective channel that allows for resolution of certain multipath components of the received signal at the receiver. In order to make the signal bandwidth larger than the channel's coherence bandwidth, the data symbols in are coded time-domain and transmitted faster than the original symbol rate. This treatment in frequency domain makes the information spectra to appear as if it has been spread in frequency. This, however, does not affect the total average power of the information spectra.

Figure 3 illustrates the theory of data spreading in time-domain and frequency domain for a square pulse, whose frequency spectrum is $\sin(x)/x$ [30]. The average powers under both spectrums are the same; however, the bandwidth of the main lobe is larger for the coded and spread data with an equivalent lower magnitude. The despreading process at the base-station consists of decoding the user data (narrowing its spectrum, and

increasing its height) and filtering its energy, as shown. The filter also takes a small portion of noise and interference. If the power of the signal from every individual mobile is not adjusted by the base-station, the despread user-spectrum will be buried in interference beyond filtering. In fact one mobile with a high signal strength could overload the base-station receiver.

Spread spectrum is performed for resolving certain multipath components of the received signal at the receiver. This is accomplished by use of the RAKE receiver.

Figure A3: Spreading Process and the Received Power Spectrum at the Base-Station

A.2.4 The RAKE Receiver in Frequency Selective Channel

The RAKE is the optimum receiver in frequency selective environments. A comprehensive overview of RAKE receivers is given in [31] for coherent and non-coherent detection. The receiver is optimum in a sense that it collects the information of L selected multipaths and performs as an L th-order diversity communication system. Therefore given the channel conditions, the receiver's decision is based on the combined energy of the replicas of the signal paths which are received within the span of the RAKE's fingers and carry the same information [28].

Because the chip duration of the spread symbols is much shorter than the delay-spread of the channel, replicas of every symbol caused by multipath reflections, fall within the delay spread of the channel, with little or no inter-symbol interference. This is where the coherence bandwidth of the channel is much smaller than the signal bandwidth, $BW_c \ll W$. Symbolically, the received signal $r(t)$ is as follows

$$r(t) = \sum_{k=1}^L c_k(t) \cdot u\left(t - \frac{k}{W}\right) + z(t) \quad (\text{EQ 25})$$

where $c_k(t)$ represent the time-varying channel coefficients, $u(t)$ is the band-limited signal represented by the *sinc* function ($|f| \leq W/2$), and $z(t)$ is the AWGN. The model is represented as a tapped delay line with $1/W$ tap delays and $\{c_k(t)\}$ weight coefficients. In theory only the rays that are delayed by multiples of $1/W$ are the ones that are resolvable and receivable; this treatment is frequency diversity. Others rays are destructively added and are beyond the "fingers" of the RAKE.

The output of a frequency selective slowly fading channel during a 7.68 second time interval is shown in Figure 4 which demonstrates the fading characteristics of a

frequency selective channel. The radio receivers have a threshold (in dB) below which the signal is not recoverable or goes into a deep fade, however in practice they are diversified further with 2 antennas half wavelength apart for no correlation.

Figure A4: A typical output of the Rayleigh fading channel with 1 multipath during 384 frames (20 ms each)

A.2.5 Mobility and Doppler

The signal is said to be in a fade status when its signal strength goes below the receiver's threshold. This occurs when the reflected rays of the signal (multipaths) combine destructively and add errors to the information. A frequency non-selective channel is said to be slowly fading if the signal interval (or the symbol duration) is much smaller than the channel's coherence time which implies that the Doppler spread is small. In other words, the signal fades slowly when the speed of the mobile is low and the channel environments do not change quickly with time. In wireless communication slow and fast fades are modeled as a characteristic of the channel with Doppler frequency measurements:

$$f_{Doppler} = \frac{vf_c}{C} \quad (\text{EQ 26})$$

where v is the speed of the mobile, and $\frac{f_c}{C}$ is the carrier frequency divided by the speed of light.

A.2.6 Coherent and Non-Coherent Detection

Information is manipulated at the transmitter by different coding and modulation schemes in order to traverse the channel and be detected and decoded successfully. The channel itself, also manipulates the data while it passes through. Theoretically at the receiver, the inverse process of the transmission manipulations must be performed in order to obtain the original data symbols. If the channel's effect on the data is known at the receiver, or estimated through statistical processes and techniques, then the receiver is said to have coherent detection. Specifically, it is the carrier phase shift through the channel that the receiver requires to be able to compensate for it via coherent detection.

If such channel estimates are not available, nor estimated at the receiver, the data must be transmitted with certain techniques so that the receiver can perform non-coherent detection while ignoring the channel phase shift.

Coherent detection techniques require the carrier phase, for reception, and the symbol timing of the information-bearing signal for synchronous sampling of the output [28]. Carrier synchronization is performed on either a pilot signal, or the information-bearing signal itself. “The pilot signal allows the receiver to synchronize its local oscillator to the carrier frequency and phase of the received signal”; “the receiver also employs a phase-locked loop (PLL) to acquire and track the carrier component of the information-bearing signal.” On the other hand when there is no pilot signal, the receiver employs techniques such as squaring loop, Costas loop, and decision-feedback loop to acquire and track the carrier phase of the signal. Symbol synchronization also may be performed via a clock signal, but the approach is not economic for the transmitter’s power. In practice, the receiver employs a technique known as the early-late gate synchronizer [28].

Acquisition and tracking of the received signal make up the two stages of synchronization [30]. In the acquisition stage, the phase of the received signal is obtained and in the tracking stage, the timing uncertainty and variations are tracked. If the phase is lost, while tracking the timing drifts, the acquisition system takes control. Figure 5 shows part of a receiver that employs Costas Loop and the Early-late Gate synchronizer [30]. The input signal is correlated with an advanced, and a delayed version of a carrier. The delta of the two is inputted to a voltage controlled clock (VCC) that excites the symbol waveform generator which feeds the correlators (band pass filters). The aim is to control the delta to be zero and to keep the clock phase locked on the carrier.

Figure A5: Costas Loop and Early-late Gate of a Coherent Receiver

As mentioned above, non-coherent detection is made possible by transmission techniques to recover the received signal irrespective of the channel phase. One such technique is Differential Phase Shift Keying, DPSK, [27]-[29],[31]-[35] in which the data symbols are modulated as differential PSK in order to be received non-coherently. In this form of modulation, every transmitted bit is the mod-2 difference of two bits: one of the data sequence and one of the adjacent (previously) coded sequence, as shown in Figure 6. At the receiver this transmitted difference is matched to its delayed version and is correlated, just as coherent receivers. However, the information is in the form of signal

phase difference which could either be in-phase or out-of-phase with its delayed version in the correlator. This allows for the retrieval of the original data since the channel phase cancels out (and gets eliminated in the subtraction) when the difference of the received data and its delayed version are calculated (correlation); therefore it is unnecessary for the receiver to estimate the channel's phase.

Figure 6 shows the encoding scheme for the DPSK modulation. The reference bit is arbitrarily chosen [34][35]. The encoding operation is $c_n = \overline{d_n \oplus c_{n-1}}$ which includes modulo-2 addition and inversion of the coded sequence $\{C_n\}$ and the data $\{d_n\}$ [34]. The receiver is shown in the form of a RAKE whose individual fingers are DPSK demodulators, and the LPF filters eliminate the cross complex terms.

Figure A6: DPSK Transmission & A Non-Coherent RAKE Receiver

A.2.7 Quadrature Phase Shift Keying (QPSK)

A QPSK system consists of two orthogonal Bipolar PSK (BPSK) systems running in parallel [30], as shown in Figure 7. This combination results in doubling the bit rate through the channel on the same carrier frequency without requiring additional bandwidth or average energy per bit. The QPSK signal is represented as

$$S_{QPSK}(t) = \pm A \cos(\omega_c + \theta) \mp A' \sin(\omega_c + \theta) \quad 0 \leq t \leq T_b \quad (\text{EQ 27})$$

where A and A' are binary data from two sources and $T_b = T/2$. The signal constellation is shown in Figure 7; analogous to the data, the phase transitions are $\pm m \frac{\pi}{4}$ $m \in \{1, 3\}$.

The average probability of error (for one bit) is given by

$$P_e = \frac{1}{2} \text{erfc} \left[\sqrt{\frac{A^2 T_b}{N_0}} \right] \quad (\text{EQ 28})$$

where T_b is the bit duration of the data, and N_0 is the Additive White Gaussian Noise power (AWGN). This is also the probability of error for a PSK system with a peak

amplitude of A and bit duration of T where $\frac{(\sqrt{2}A)^2}{2} T_b = \frac{A^2 T}{2}$. Therefore while

considering equal powers, both QPSK and BPSK result in the same bit error probability.

The largest phase change in QPSK is 180 degrees because 2 bits map to one of 4 signal phases. This phase reversal in a band-limited system causes amplitude fluctuations since the signal is forced to cross a zero amplitude state. These distortions do not interact well with non-linear amplifiers [30]. The QPSK amplitude distortions may be avoided by delaying one (BPSK) branch of the system by $T/2$. This way, the signal changes in phase by 90 degrees, and the phase changes become dependent on one bit and not both. A system of this kind is referred to as Offset-QPSK (O-QPSK).

Figure A7: A QPSK Modulator and its Signal Constellation

A.3 Interleavers

Interleaving and de-interleaving techniques are used to scatter bursts of errors throughout the received signal (in time), especially those through the multipath fading channels.

Interleavers increase the reliability of transmission as errors become statistically independent. A block interleaver of m rows can break an error burst of length $l=mb$ into m bursts of length $b \leq [(n-k)/2]$ which can theoretically be corrected by error-correcting (n,k) codes with $n-k$ parity [28]. Error randomization provided by interleaving improves the performance of forward error correcting decoders.

A.4 Convolutional Codes

Convolutional codes are generated by passing the information sequence through a linear finite-state shift register of L (k -bit) stages and n polynomial function generators as shown in Figure 8 [28]. A k/n rate convolutional coder of constraint length L outputs n -bit code words for every k binary input bits and shifts the data through the register k bit at a time. Convolutional codes are represented with their generator polynomials' matrix format, where every polynomial is represented as a vector of 1s and zeros. "A one in the i th position of the vector indicates that the corresponding stage in the shift register is connected to the modulo-2 adder," and a zero means no such connection exists.

Convolutional codes are used for the error correcting capability of their corresponding decoders.

Figure A8: Block Diagram of A Convolutional Encoder

A.4.1 Viterbi Decoder

The Viterbi algorithm is an optimum convolutional decoder in the maximum likelihood sense. The algorithm is based on finding the (trellis) path with the smallest distance between its code sequence and the received code word sequence. The Hamming distance is the number of digits that two code sequences of the same length differ with one another. "For example the sequence 011010111 differs from the sequence 111001101 in digits 1, 5, 6, and 8, so the Hamming distance is 4." [36]

The algorithm for decoding one coded word is based on

- Comparing the convolutionally encoded sequences of all possible input combinations to an L -bit shift register with the instantaneous received code word,
- Assigning a weight (Hamming distance) to every combination,
- Selecting a survivor path amongst the branches that terminate to the same state based on the branch's weight, and,
- Choosing the largest cumulative weighted path (smallest Hamming distance) through a window of decision which consists of sufficient number of stages of the trellis.

A convolutional encoder with rate k/n and constraint length L could have 2^k branches of input to each state on the trellis and 2^k branches of output at steady state, after the first $L-1$ stages of operation. Every one of the 2^{L-1} shift register input combinations represents a state in the trellis diagram of the Viterbi decoder [28], as shown in Figure 9. Every state has 2 ($k=1$) weights associated with it - one weight is due to a zero input to a state, and another is due to a 1 input. The algorithm chooses the (input) branch that has the larger weight, shortest Hamming distance, at steady state as a survivor. The weights of all survivor states are added cumulatively through the stages of the tree for obtaining the states' transition weights. This is performed till the end of a window of decision, at

which point the largest cumulative weighted path (the likeliest survived path) indicates what was inputted to the register at the beginning of the window - thus the n -bit word gets decoded back to a k -bit symbol.

The window is shifted forward by one n -bit word, and the procedure explained above is repeated. The decision window must be large enough to give an unbiased surviving sequence among the likeliest outputs of the combination states, in order to decide in favor of the bit at the beginning of a path with the largest cumulative weight. It must also be short enough to conserve expensive memory for calculation and storage of the states, their surviving weights, and their cumulative weights. Experimentally [28], a decision window of greater than or equal to $5 \times \text{constraint length}, L$ of the register “results in a negligible degradation in the performance relative to the optimum Viterbi algorithm” of infinite stage window. In the simulation of this project the decision window was set to 45 (5x9) transition stages.

Figure A9: A Viterbi Trellis

A.5 M-Sequences and PN generation

Pseudo random noise sequences are used because of their important auto-correlation and cross-correlation properties in telecommunication. These sequences' origin is from the maximum-length shift-register sequences, or m-sequences described in [37]. A PN of length $n=2^m-1$ is generated by an m-stage shift register with a linear feedback as shown in Figure 10 [28].

A PN's auto-correlation properties depend on its feedback polynomial; m-sequence PNs have sharp peaks in their auto-correlation function. Such property is especially desired in spread spectrum applications where the PN codes are employed for spreading the data. In CDMA applications because of multipath interference sharp auto-correlation peaks and low side-peaks are desired; and because of multiple access interference low cross-correlation is required. PN codes do not have low cross-correlation [28][30], however, a small subset of m-codes such as Gold and Kassami do possess such properties.

In CDMA applications, Hadamard matrices are used for their low (zero) cross correlation property. Every row of a Hadamard matrix, also known as a Walsh code, is orthogonal to other individual rows. This means that the integral sum of two bit-multiplied rows are always zero.

Figure A10: A PN Generator and its Auto-Correlation Function

APPENDIX B: Simulated Test-Cases

In this appendix results of the simulated test-case scenarios of this thesis are presented in more detail.

1F Result of 1 video-user with 0 interferers - forward link

In this case as shown in Figure 1F the SIR still follows the bursty packet generations. average BER is 1.5×10^{-3} at 1.6964 dB average SIR with 252 IS95 transported frames. With no interferers, the video user in the forward link performs better in BER by one order of magnitude than in the case with 8 interferers (test-case 3F). The effect of the time-varying fast-fading channel is especially noticed at the 2 points where the number of generated frames is as low as 3, but still the error-rate persists above zero.

Figure 1F: Result of 1 video-user with 0 interferers - forward link

2F Result of 1 video-user with 4 interferers - forward link

As shown in Figure 2F, the SIR graph follows the graph of the bursty frames. Average BER is measured to be 5.3×10^{-3} at an average SIR of -1.5531 dB. The same scenario in the reverse link (Figure 2R) shows a BER of 2 orders of magnitude less. At high Doppler, where power control does not contribute to the performance of the link, the forward traffic channel seems to be more susceptible to interference than the reverse link because of less coding and redundancy. The performance of the two links merges and becomes more similar at lower Doppler, where the fade duration is in the order of the frame length, and coding has diminishing return.

Figure 2F: Result of 1 video-user with 4 interferers - forward link

3F Result of 1 video-user with 8 interferers - forward-link

As presented in Figure 3F, the SIR graph adapts to the same bursty characteristics as the video frame generator has. The average error rate in this case was 1.11×10^{-2} but not as bursty as that of the reverse link - 252 IS95 frames were transported in an environment with average SIR of -3.3800 dB at the receiver. The same number of interferers were used in the forward link simulation as were simulated for the reverse link.

The transmissions in the forward channel are synchronous in a sense that the data is sent from one base-station and received by the mobiles. There is only one transmitter, and multiple receivers. This is especially advantageous at low Doppler for power control at the base-station.

Figure 3F: Result of one video-user with 8 interferers - forward-link

4F Result of 2 video-users - forward link

This scenario approaches the capacity limit of the forward link at high Doppler, considering a reliable error-free channel as the limiting factor. The BER in this case was 6.8×10^{-3} at -2.2937 dB SIR. As shown in Figure 4F, the SIR graph is not in harmony with the generated frames' graph, in contrast with the test-cases with less interference. This is due to the bursty nature of the second video-user.

Figure 4F: Result of 2 video-users - forward link

5F Result of 1F, slow fading with power control . forward link

At low Doppler and with power control, the performance of the forward link has much improved from test-case 1F. Figure 5F shows that all frames were decoded free of error against the bursty data and the channel conditions.

Figure 5F: Result of 1F, slow fading with power control - forward link

6F Result of 3F, slow fading with power control - forward link

As shown in Figure 6F the error rate has much improved as compared to that of test-case 3F. This improvement directly contributes to a higher capacity in the forward link. The BER is measured at 7.55×10^{-4} at -3.5445 dB SIR.

Figure 6F: Result of 3F, slow fading with power control - forward link

7F Result of 1 video-user with 16 interferers . forward link

As shown in Figure 7F, The error rate has just passed the acceptable limit of a reliable communication link. At -5.8847 dB SIR, the bit error rate is measured at 7.5×10^{-3} .

Considering the dynamic interference of another video user and the results from test-cases 4F and 6F, one could conclude that the forward link can reliably support two video users in the presence of up to 8 audio interferers.

Figure 7F: Result of 1 video-user with 16 interferers - forward link

8F Result of 1 audio user with 23 interferers - forward link

In this test-case, we have clearly reached the limit of a reliable forward link. Figure 8F shows that most errors are above the 5×10^{-3} mark. The average SIR was measured at -13 dB at the receiver. Both this test-case and test-case 7F made use of 24 user channels in the forward link; however, the dynamic interference of the video user (with up to 8 channels) contributes to the reliability of the link - unlike 8 audio interferers.

Figure 8F: Result of 1 audio user with 23 interferers - forward link

1R Result of 1 video-user with 0 interferers

As shown in Figure 1R the average SIR in this scenario is comparable to that of test-case 1F at 1.7043 dB (one video user and no interferers, forward link); however, there are no errors in the data recovery. The error correcting convolutional codes along with Viterbi decoder retrieve the data symbols at high Doppler. The SIR graph follows the frame generation in this scenario as well.

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Figure 1R: Result of 1 video-user with 0 interferers

2R Result of 1 video-user with 4 interferers

As shown in Figure 2R, the average BER is 6.47×10^{-5} at -1.5514 dB average SIR. This BER is 2 orders of magnitude less than that of the forward link, test-case 2F, because of powerful error correcting techniques used in the reverse link and their effectiveness at high Doppler.

Figure 2R: Result of 1 video-user with 4 interferers

3R Result of 1 video-user with 8 interferers

As presented in Figure 3R, the SIR graph resembles and follows the same general shape as the frame-generation graph. In other words, in such environment the total signal to interference (noise and interference) ratio adapts to the same bursty characteristics as the video frame generator has. The average bit error rate in this case was 1.7×10^{-3} - 252 IS95 frames were transported in an environment with average SIR of -3.3758 dB at the receiver.

Figure 3R: Result of 1 video-user with 8 interferers

4R Result of 3R - slow fading

Figure 4R shows that the average error-rate has increased substantially as compared to that in Figure 3R. This is mainly due to lack of power control for proper user-data retrieval, and the fact that the fade durations are as long as the interleaver's depth and the whole frame falls into fade and becomes erroneous.

Figure 4R: Result of 3R - slow fading

5R Result of 1 video-user with 10 interferers

This scenario was not tried in the forward link (at high Doppler) because experimental results pointed out that the total SIR would give a BER of higher than what was obtained with 1 video and 8 audio interferers.

As shown in Figure 5R the reverse link is still reliable for communication, but is quickly approaching its capacity limit with a BER of 3.1×10^{-3} at -4.0606 dB SIR.

Figure 5R: Result of 1 video-user with 10 interferers

6R Result of 2 video-users

Two bursty data generators were handled in this simulation. A BER of 7.76×10^{-4} was achieved at -2.2821 dB SIR. Figure 6R and the results show that the system with 2 video-users still has room for additional traffic.

Figure 6R: Result of 2 video-users

7R Result of 2 video-users and 4 interferers

With a BER of 2.3×10^{-3} at -3.8750 dB SIR, this test-case explicitly demonstrates the dynamic interference of video-users as compared to that of the audio users, Figure 7R. The forward link at high Doppler supports two video users in the presence of 4 to 6 audio interferers.

Figure 7R: Result of 2 video-users and 4 interferers

8R Result of 8 bursty audio-users with 8 interferers

The follow-up resemblance of the signal-to-noise ratio graph to the frame-generation graph is still noted in the voice-users' case, as shown in Figure 8R. In this experiment the average bit error rate was found to be 1.4×10^{-3} , after 252 IS95 communicated frames, in an environment with average SIR of -3.3766 dB at the receiver.

Figure 8R: Result of 8 audio-users with 8 interferers

9R Result 8R - slow fading

Figure 9R shows that the average error-rate has increased substantially as compared to that in Figure 8R. This is mainly due to lack of power control for proper user-data retrieval, and the fact that the fade durations are as long as the interleaver's depth and the whole frame falls into fade and becomes erroneous.

Figure 9R: Result of 8R - slow fading

10R Result of 1 audio-user with 15 interferers

This scenario was not tried in the forward link at high Doppler because experimental results pointed out that the total SIR would give a BER higher than what was obtained in test-case 3F (one video-user and 8 audio interferers).

As shown in Figure 10R, the SIR is more stable than bursty as compared to other scenarios. It varies within ± 0.3 dB of the average -11.9446 dB. The generated frames increase linearly with the number of iterations as every mobile transmits one frame. The reverse link is reliable for communication in this scenario at 1.2×10^{-3} BER, but the average interference is high at -12dB SIR. For the same number of reserved channels in test-case 3R (one video and 8 interferers) the interference is -3.3758 dB; for the same BER performance this scenario has 3 times interference than test-case 3R.

Figure 10R: Result of 1 audio-user with 15 interferers

11R Result of 1R - slow fading with power control

As shown in Figure 11R, the performance of the reverse link is far from its limits with power control at low Doppler. The results seem to be similar to test-case 1R.

Figure 11R: Result of 1R - slow fading with power control

12R Result of 3R - slow fading with power control

Figure 12R illustrates the bursts of errors. Power control has substantially decreased the error rate compared to that of test-case 3R. The bit error rate is measured at 2.8×10^{-4} with -3.5402 dB SIR.

Figure 12R: Result of 3R - slow fading with power control

13R Result of 1 video-user with 16 interferers

With this test-case we have just crossed the performance limit of the reverse link. Figure

13R illustrates the error rate which averages at 1.03×10^{-2} with -5.8862 dB SIR.

Considering the dynamic interference of another video user, and the result of test-cases 6R and 12R, one could conclude that the reverse link can reliably support 2 video users in the presence of 6 to 8 audio interferers.

Figure 13R: Result of 1 video-user with 16 interferers

14R Result of 10R - slow fading with power control

As illustrated in Figure 14R, power control has allowed for elimination of burst errors that occurred in test-case 10R.

Figure 14R: Result of 10R - slow fading with power control

15R Result of 1 audio-user with 23 interferers

In this test-case, we have clearly reached the limit of a reliable reverse link. Figure 15R shows that most errors are above the 1×10^{-3} mark. The average SIR was measured at -14.6021 dB at the receiver. Both this test-case and test-case13R made use of 24 user channels in the reverse link; however, the dynamic interference of the video user (with up to 8 channels) contributes to the reliability of the link - unlike 8 audio interferers.

Figure 15R: Result of 1 audio-user with 23 interferers

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IMAGE EVALUATION TEST TARGET (QA-3)

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