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Video Transmission over Wireless CDMA Networks

By

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

Ottawa-Carleton Institute for Electrical Engineering
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The undersigned recommend to
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submitted by **Xiao Wen Wu, B.Eng.**

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Abstract

Real time Variable Bit Rate (VBR) video transmission over wireless channel has been a great challenge to telecommunication researchers. It has high, variable bandwidth demand and real time constraints. Code Division Multiple Access (CDMA) is a wireless cellular system which has many advantages. Utilizing these advantages can help handling VBR real time video. In this thesis, we first verified statistical characteristics of coded video sequences. Then we proposed a simple, yet efficient multiplexing scheme for transmitting VBR traffic over wireless CDMA channel. Finally we analyzed the system capacity for proposed CDMA videophone system and proved that it is very competitive compared to other cellular systems.

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Acronyms

AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BPSK	Binary Phase Shift Key
CBR	Constant Bit Rate
CDF	Cumulative Distribution Function
CDMA	Code Division Multiple Access
CIF	Common Intermediate Format
CODIT	COde DIvision Testbed
COFDM	Coded Orthogonal Frequency Division Multiplexing
DCT	Discrete Cosine Transform

DMPR	Discrete Mean to Peak Ratio
DPCM	Discrete Pulse Code Modulation
DS-CDMA	Direct Sequence Code Division Multiple Access
FH-CDMA	Frequency Hopping Code Division Multiple Access
HDTV	High Definition Tele-Vision
ISI	Inter-Symbol Interference
MAI	Multiple Access Interference
MIC	Multipath Interference Coefficient
MMSE	Minium Mean Square Error
MPEG	Motion Picture Expert Group
NTSC	National Television System Committee
OIC	Other-cell Interference Coefficient
PAL	Phase Alternating Line
PCS	Personal Communication System
PDF	Probability Density Function
PMR	Peak to Mean Ratio
PSNR	Peak Signal to Noise Ratio
QCIF	Quanter Common Intermediate Format

RGB	Red-Green-Blue
RLMS	Recursive Least Mean Square
SDMR	Standard Deviation to Mean Ratio
SIR	Signal to Interference Ratio
SMI	Self Multi-path Interference
TDMA	Time Division Multiple Access
VBR	Variable Bit Rate
YIQ	luminance/chrominance signal, American standard
YUV	luminance/chrominance signal, European standard

CHAPTER 1

Introduction

Real time Variable Bit Rate (VBR) video, whose major characteristic is constant quality, has been treated as the most difficult traffic in communication systems, regardless of wireline or wireless transmission because it has both high, variable bandwidth demand and real time requirement. MPEG [1] [2] is a widely used video compression standard while H.261/263 [3] [4] standard is mainly designed and good for videophone applications. There has been some statistics analysis [5~8] and modelling [9~13] made about coded video. For a long time, researchers has been studying how to transmit video over computer communication networks. Recently communication researchers are working on wireless/cellular multimedia communication of which the most difficult part is to handle VBR video [14-16].

CDMA is a multiple access version of Spread Spectrum (SS) technique. SS was originally used in military wireless communication. CDMA has been found to have good performance in commercial wireless/mobile communication both in theory [17] [18] and in practice [19]. It has been proposed for both indoor and outdoor applications, also for both cellular phone and wireless computer communication. In the case of cellular phone application, it's said to have far more capacity than TDMA given the same bandwidth. This is mainly due to the spread spectrum technique and the coding gain which provide the following: elegant degradation of service with

increased users, small frequency reuse factor, soft handoff, and easy to make use of voice activity factor, etc. Besides, it also provides some inherent features such as user privacy and security.

While wireline is mainly limited by Additive White Gaussian Noise, the wireless channel is characterized by shadowing and multipath fading which deteriorates the signal significantly if no combat measures are taken. Fading can be handled by various kinds of diversity. Depending upon the signal bandwidth and the extent of multipath spread, Inter Symbol Interference (ISI) may happen, which will further degrade the service. This ISI can be either avoided by reducing the signal symbol rate while maintaining the same bit rate, or using equalization techniques at the expense of receiver complexity.

1.1 Thesis Objectives

The objectives of this thesis are: (1) to investigate the statistical characteristics of VBR videophone signals such as Probability Distribution Function (PDF) of bits or packets per video frame, Standard Deviation to Mean Ratio (SDMR), Peak to Mean Ratio (PMR) and Autocovariance; (2) to provide a statistical base for videophone traffic analysis and performance evaluation; (3) to present the theoretical basis of CDMA signaling; (4) to take into account the adverse effects of wireless mobile channels and present combating techniques; (5) to exploit the advantages of CDMA to find suitable transmission schemes to accommodate VBR traffic; (6) to analyze the capacity of a basic CDMA videophone system.

1.2 Thesis Organization

In this chapter we give a brief introduction of the thesis contents, our objectives and the thesis organization.

In chapter 2, the following are included: basic concepts of image, video and video coding are introduced, the difference between VBR and CBR (Constant Bit Rate) coded video sequence is discussed, various video compression standard and schemes are reviewed and compared, VBR video statistics and modelling are analyzed, H.263 type compression video used as an example to analyze statistical characteristics even though our conclusions is not confined to it, data obtained is compared to extensive literatures to seek common beliefs and explain difference, SDMR, PMR and autocovariance are the main statistics we are concerned with, conclusions on these statistics are presented and used to build VBR video models.

Chapter 3 presents the background knowledge of wireless mobile communications. We concentrate on microcellular Personal Communication Systems (PCS) but our analysis can be easily modified to include other kinds of channels. We analyze the shadowing and fading characteristics and the combating techniques. Then, we study the kind of channel we are interested in, i.e. CDMA. By applying the above and special schemes for CDMA, like fast power control etc., we build a system model for CDMA.

Chapter 4 focus on a CDMA system with specific parameters chosen based on the observations made in previous chapters. We discuss how to multiplex VBR over CDMA. Possible schemes are discussed to find a suitable one for our system without sacrificing the advantages of CDMA. Lastly, maximum user capacities of CDMA videophone systems are analyzed for different system conditions.

In chapter 5, we summarize all the conclusions derived from this work and suggest topics for future work.

Our contributions are described in detail in section 2.3 and in chapter 4.

CHAPTER 2

Video Coding, Statistics and Modelling

In this Chapter, we first present the basic concepts of image, video and their entropy. We emphasize the comparison of Variable Bit Rate (VBR) and Constant Bit Rate (CBR) coding from the view of bit rate vs. quality. Then various video coding standards and schemes are discussed and some comparisons are made. They are followed by an extensive study of statistical characteristics and modelling of coded video sequences.

2.1 Image and video

Basically, both image and video are presented or displayed in the format of pixels. Images are presented by a block of pixels with 8 bits for each pixels indicating luminance of a black and white picture, or 24 bits indicating RGB (Red-Green-Blue) component of each color pixel. Videos are presented by sequence of images which are called frames sampled usually at 50 frames per seconds (fps) for HDTV (1440x1152 pixels), 30 (fps) for TV broadcasting (720x576 pixels), or 10 fps for applications like videophone and some videoconferences. Since human eyes are more sensitive to luminance distortions than chrominance distortions and the color component contributes just about one tenth [20] to the information entropy, the video data inputs before coding are

often represented in a 4:1:1 luminance/chrominance signal. e.g. YUV of European standard PAL, (Phase Alternating Line) or YIQ of American standard NTSC (National Television Systems Committee). If we choose a video that has a raw input format of QCIF (Quarter-Common Intermediate Format, 176x144 pixels for luminance and 2x88x72 for chrominance) at 12.5 fps (H.263 codec output rate) and using eight bits for each luminance and chrominance pixel, we end up with a bit rate of $176 \times 144 \times 12.5 \times 1.5 = 3.8$ Mbps at the video encoder input.

Due to both the insignificance and the small contribution to entropy rate by chrominance component, we will only focus on the analysis of luminance component while the video sequences under analysis are color sequence.

From a pixel by pixel point of view, the entropy of any source pixel can be represented as:

$$H = \sum_{i=0}^{255} p(i) \log \frac{1}{p(i)} \quad \text{per pixel} \quad (\text{EQ 2.1})$$

where

$$P(i) = \text{Prob}(APixelValue = i) \quad (\text{EQ 2.2})$$

When we just consider each pixel separately and assume an equal probability case, we get an entropy of 8 bits for each pixel and the overall entropy rate of the video source is simply the same as the bit rate. However, there are strong correlations among both the pixels in the same frame (Spatial correlation) and the ones that belong to contiguous frames (Temporal correlation). Thus the conditional probability a pixel take a value doesn't equal the unconditional one:

$$(P_{t_1, x_1, y_1}(i) | Pixel_{t_2, x_2, y_2} = j) \neq P_{t_1, x_1, y_1}(i) \quad (\text{EQ 2.3})$$

where t, x and y stand for time, horizontal and vertical coordinates of the 3-D space of the video sequence. And the closer the two pixels are in the 3-D space, the more likely the above inequality will hold. Without this kind of correlation, video compression will not exist. When $t_1 = t_2$,

the correlation is called spatial correlation, and when $t_1 \neq t_2$, it's referred to as temporal correlation.

For different types of video the entropy rates are certainly different. A full-action movie like "Star War" has much higher entropy rate than a "head and shoulder" type videophone sequence even if they are made at the same resolution, i.e. same frame size and rate. Also, even within the same sequence, the instantaneous entropy rate varies. In a "head and shoulder" sequence the entropy rate of the person talking or moving around is higher than when the person is just listening to the other side quietly. Let alone the difference between the action clips and the static clips of the "Star War". Generally speaking, video source is characterized by variable entropy bit rate.

2.2 Video Coding/Compression

The aim of video coding/compression is to reduce the information redundancy of raw video input so that transmission and storage of video won't take too much resources such as bandwidth, time and storage space. Before discussing various coding standards and schemes, we want to emphasize the difference between VBR coding and CBR coding.

2.2.1 VBR Coding vs. CBR Coding

Both VBR and CBR appear very often in various video coding/compression literatures. The basic difference between these two kind of coding is the trade off between quality variations and bit rate variations. Figure 2.1 illustrates this point. Figure 2.2 and Figure 2.3 provide data from real H.263 coded "Carphone" Sequences to support the same point.

To evaluate the quality of coded video, PSNR (Peak Signal to Noise Ratio) performance is often used. PSNR is the signal power to distortion (noise) power ratio where the distortion/noise is introduced by the entropy coding. Equation 2.4 gives mathematical definition of PSNR where x and $\hat{x}$ stand for the original and reconstructed value of a pixel, and x_{max} stands for the maximum

value of a pixel, e.g. 255 for 8 bit image, *Rows* and *Columns* are the pixel counts of the image's row and column respectively.

$$PSNR = 10\log \left(\frac{x_{max}^2 Rows \times Columns}{\sum_{AllPixels} (x - \hat{x})^2} \right) dB \quad (EQ 2.4)$$

(a) *VBR*

(b) *CBR*

Figure 2.1 CBR vs. VBR

Figure 2.2 Verifications of Trade-off between Bit Rate Variation and PSNR

From Figure 2.2, we can see that (b), which is the CBR case, is just an effort of H.263 codec to maintain constant output bit rate. Although its effort is not quite successful, it is a strong validation of Figure 2.1. Because in this figure, even when the bit rate is not quite constant, and for almost the same average bit rate, the PSNR of the CBR case shows a significant change of 3 dB degrade for the high activity frames around frame 105 and the average PSNR drops by 1.7 dB. Having these numbers in mind, let's have a look at Figure 2.3. which is the PSNR of the same sequence of Figure 2.2 (a), but the sequence is corrupted by a single bit error in frame 105. We can tell that the worst PSNR here is only 1 dB lower while the average PSNR is 1 dB higher than that of CBR. Note that they have the same average bit rate. Images from the VBR sequence and CBR sequence in Figure 2.2 and the corrupted sequence in Figure 2.3 are plotted in Figure 2.4 to give visual verifications.

Figure 2.3 PSNR Curve of Error Corrupted Video Sequence

Frame 105 (a1)

Frame 115 (a2)

Frame 105 (b1)

Frame 115 (b2)

Frame 105 (c1)

Frame 115 (c2)

Frame 125 (a3)

Frame 155 (a4)

Frame 125 (b3)

Frame 155 (b4)

Frame 155 (c3)

Frame 155 (c4)

Figure 2.4 Visual Effect of VBR, CBR and Corrupted Video

The notations of x1, x2, x3 and x4 stand for frame No. 105, 115, 125 and 155 respectively where the value of x can take a, b and c which represent VBR, CBR and corrupted video.

Although the printing is not very good, we can still tell that compared to the VBR series, the head and the right hand of the person and the scenes outside the window, which are the moving parts of the picture, are always blurred in the CBR sequence. The corrupted sequence has one bit error in frame 105 which results in the distortion around the person's right hand while other region in the scenes remains unaffected. The distortion recovers slowly according to the c1~c4 sequence. We can also tell this from the PSNR curve in Figure 2.3.

It's obvious that if the bit rate is more constant compared to entropy-base-coded VBR, both the worst case PSNR and the average PSNR will degrade further. It's equivalent to that even before being transmitted over any channel, the video is corrupted to some extent, maybe in a subtle way. The average PSNR will be several dB lower than its VBR counter part. Once being sent over the channel, it will be further corrupted. The equivalent effect has already been demonstrated in Figure 2.4.

By now we can have the following conclusions:

- CBR has overall significant worse performance in average PSNR and worst PSNR given the same average bit rate. In CBR coding, bits are wasted on static scenes and not enough for the active scenes. The performance of CBR can be worse when subjected to further channel errors during transmission.
- VBR, on the other hand, distributes the bits reasonably and results in a better overall PSNR performance with less visual distortion.
- Transmission errors propagates in the video scenes, but will recover slowly to some extent when the update information coming from subsequent frames.

As mentioned in section 2.1, the entropy rate is variable for the video sequence. So if constant quality (to some extent can be reflected by constant PSNR) is required, the video encoder output will be inevitably variable as in (a) of Figure 2.1. On the other hand, if constant output bit rate is required, the quality/PSNR of the coded stream will be reverse proportional to the entropy rate of the input which is variable. In reality, most users prefer constant quality that can satisfy their visual enjoyment while communication service providers like constant bit rate because it's much easier to handle.

Many published papers addressed the topic of VBR over wireline ATM [5~8] [9] [10] [21] [22] etc. But only a few papers have proposed constant bit rate or Quasi-Variable bit rate video transmission over wireless channels [14] [15] [23]. Quasi-variable means that the output rate can be one of two or few selected bit rates. While it's not necessary for the coded video to have 100% constant quality or continuous variable bit rate, it's certainly necessary to investigate the nature of VBR traffic transmission over wireless channel before we quantize it into Quasi-VBR. This is one of the important aims of this work.

2.2.2 Video Compression Standards and Schemes.

To achieve the goal of video coding or compression, all algorithms are implemented in such way that after some transformation, the information contents are concentrated in a small range of a certain domain or domains. e.g. frequency domain, then the unimportant range can be dropped without much sacrifice of the video quality and the important range is coded with entropy coding. i.e. lossless coding to ensure that the compressed information can be reproduced completely.

Video compressions can be categorized based on source coding techniques as Prediction coding, Transform coding, Layered coding, Object Oriented coding. They can also be categorized based on entropy coding techniques as Run-length coding, Huffman coding, Arithmetic coding and etc... Practical video compression schemes are usually a combination of source coding which is often lossy, and entropy coding which is often lossless. The source coding or entropy coding of

a compression scheme also are often hybrid of two or more techniques. For example, the scheme chosen as an example in this work, H.263, is a hybrid of Prediction coding, Transform coding and Arithmetic coding and Run-length coding [24].

The following is a brief introduction of various video compression standards and schemes and a comparison of the chosen H.263 and other schemes that is possible for videophone applications.

2.2.2.1 MPEG Video Compression.

Moving Picture Experts Group (MPEG) is a standard group established in 1988. It aims at coding moving pictures for digital storage and transmission. There are two widely used standards: MPEG-1 [1] and MPEG-2 [2], and there is another one coming up: MPEG-4.

Generally speaking, MPEG targets at wide range of applications for storage and transmission of digital video. This led to advantages such as random access, fast forward/reverse search, audio-visual synchronization, good error behavior, etc. MPEG has three kinds of coded frame, Intra frame (I), Predicted frame (P) and Bidirectional predicted frame (B) with the sequence of I, P and B frames depends on application. The source coding of MPEG is a combination of prediction coding and transform coding (DCT with Motion Estimation, generally) while variable length coding including run-length are used for entropy coding. The compression ratio is in the range of 50:1 to 200:1.

MPEG-1 aims at a coding rate of 1.5-3 Mbps while MPEG-2 handles Broadcast TV type video and HDTV in the range of 4-60Mbps. MPEG-4 was proposed for very low bit rate video compression in the range of 4.8-64 kbps for CIF or QCIF type input and it's under development now. Note that the above bit rates include audio.

2.2.2.2 H.261/263 Video Compression.

CCITT Recommendation H.261 [3] is a guideline for video codec for audiovisual services at px64 kbps with input of CIF or QCIF at 30 fps/10fps. The coding/decoding system is a hybrid of

inter-picture prediction, DCT transform coding and motion compensations. Its entropy coding is variable length coding such as Huffman coding. The compression ratio is in the range of 100:1 to 1000:1.

With the input format and compression ratio mentioned above, H.261 caters for videophone and desk-top video-conference application with the multiple of p being 1 or 2 and for group video-conference application with higher value of p .

ITU-T recommendation H.263 (Draft) [4] has been proposed in July 1995 for very low bit rate compression of moving picture component of audio-visual service like videophone and video-conference. It basically uses the same compression algorithm of H.261 while adding half pixel predicting and negotiable options. Together with less overhead and improved variable length coding (including Arithmetic coding and Run-Length coding), H.263 can achieve the same quality of H.261 with 30-50% of the bit usage.

H.263 can code video ranging from SQCIF(128x96) to 16CIF (1408x1152) with frame rate ranging from 10fps to 30 fps.

2.2.2.3 Other Compression Standard and Schemes.

There are other widely used video compression standards like Intel's Digital Video Interactive (DVI), Phillips, Sony and Microwave 's cooperation product Compact Disc Interactive (CD-I) and two well known software-based compression standards: Apple's QuickTime Movie (QT or MOV) and Microsoft's Audio Visual Interleaved (AVI). These standards were defined by commercial companies, each with their advantages and limitations. A detailed discussion of them is beyond the scope of this work.

Besides these standards, there are various proposed schemes for very low bit rate video coding. Here we mention a few schemes: Layered coding schemes which includes sub-band coding [23] [25] and wavelet coding [26~28] has the advantage of multi-resolution coding which can be

used for joint source-channel video coding. Objected Oriented coding, which includes Segmentation coding [29] [30] and Model coding [31] [32], is a novel approach that achieves very low bit rate. But it has strict application limitations.

There are also various undergoing researches concerned about new and break-through compression schemes with many of them Heading toward MPEG-4.

MPEG-4 is expected be either a standard relying on some break-through of coding theory or a standard being hybrid of absorbing advantages of many proposed schemes.

2.2.3 H.263 Compared to Other Standards and Schemes.

2.2.3.1 H.263 vs. MPEG.

H.263 is mostly designed for low bit rate and low resolution applications such as Videophone and Video-Conference. The fact that H.263 has many more negotiable options and yet much less overhead than MPEG makes it a more suitable candidate for videophone communication in our work. Another characteristic of H.263 (also H.261) type compression is that the Peak to Mean ratio of bit rate is quite low compared to other schemes. Further discussions are in section 2.3, VBR Video Modelling.

2.2.3.2 H.263 vs. Other Possible Candidates for Videophone.

While H.263 is less flexible than MPEG, it's more flexible than the Object Oriented compression schemes. It's also a more state-of-art compression standard compared to layered coding and can easily be modified to form layered bitstreams. Even just in the sense of PSNR vs. bit rate performance, H.263 is better than these schemes in many cases. But it also has weaknesses. One of it's weaknesses is that it often codes the first frame as Intra frame and the following as Predictive or Bidirectional frames which will result in poor error performance and a big first frame. The

solution can be adding Partial Forced Update mentioned in [15]. This will increase the bit rate by 30 to 50 percent. But it gets higher PSNR. Another inconvenience of H.263 is that its PB frame option which code two little motion frames as one PB frame is not practical in real-time application. Discarding this option will result in some increase of bit rate.

Note that all the above data about H.263 are obtained by coding the real sequences. H.263 generally achieve 3-4 dB higher PSNR for the color components compared to luminance component, but for the reason stated in section 2.1, the above and later PSNRs used, unless specified, are just those of luminance component.

Overall speaking, H.261/263 is a good candidate for videophone applications. So we utilize the state-of-art H.263 software codec to produce video bitstreams for statistical analyses and modelling. We also use them for some of the CDMA system input.

2.3 VBR Video Modelling.

Following the emergence of video compression standards and schemes, studies has been carried out in the field of modelling the coded video traffic in the communication networks to simplify the simulations and evaluations of network performance. Although most of the works was done for applications in wireline computer networks, they are valuable guidelines for wireless video communications. We summarize the video data statistics in section 2.3.1 and video modelling in section 2.3.2. Several coded videophone sequences will be used to verify both the statistics and modelling. Conclusions are also stated in this section.

2.3.1 Video/Videophone Data Statistics.

During the past, there were investigations about the statistics of the variable bit rate / packet-rate output of video coder. Our concerns are the variance to mean ratio, peak to mean ratio, bits/

packets per frame distribution and autocovariances. They were measured and analyzed in many literatures [5~8] [9-13] [21] [22].

[9] gave a study of DPCM/DCT coded 4CIF (TV), 25fps video with no Motion Compensation (MC). The video features three people talking for 30 minutes long. The standard deviation to mean ratio (SDMR) and peak to mean ratio (PMR) are 0.335 and 5~6 respectively. The probability density function (pdf) of packets per frame was verified to be Gamma distributed. And the unitized autocovariance of the packets per frame sequence is quite smooth and is about 0.2 at 100 frames.

In [11], a 10 second interval of 500x500 frame video was DPCM coded with interframe difference coding and conditional replenish. The sequence is a 30 fps "Head & Shoulder" (H&S) type video. SDMR equals 0.48 and PMR equals 2.71. The pdf is described as "Bell Shaped" and the autocovariance has some periodic peak which was explained as either an effect of computing truncation or people in the scene moved in a periodic way. The autocovariance coefficient at 10 frame is about 0.18.

In [13], four types of video sequences, each of 11 minutes and 15 fps and resolution of 384x288, were compressed using DCT plus MC type of MPEG-1 coding. The video types are commercial clips, TV show, sceneries and video conference. SDMRs are 0.59, 0.59, 0.58 and 0.71. PMRs are 4.55, 2.99, 3.16 and 2.52 respectively. This paper claimed that Gamma pdf or any "Bell Shaped" pdf does not apply to conference type compressed video when both intra and inter frame coding are used. This is not true in [21] and our sequence analysis.

In [21], inter and intra DPCM were employed to compress four video sequence at 25 fps. They are one 30 minutes H&S (videophone), one 30 minutes two-people-talking conference scene and two television scenes of 12 and 3 hours long respectively. For videophone, conference and TV sequences, SDMRs are 0.69, 0.47 and 0.26 and PMRs are 4.44, 3.08 and 2.66. As for the

autocovariance coefficients, they are 0.1, 0.2 and 0.65 respectively at 25 frames. Similar to [9], the curves are relative smooth.

In [5], 9 minutes of H&S type video was coded with H.261 at 30 fps. SDMR is 0.28 and PMR is 1.75.

In [6], three types of videos, each lasting 10 seconds, were H.261 type code at 50 fps with 360x228 resolution. The types are: H&S type news, low/High motion and zooming combined scenes and a sports sequence with three scene cuts. The PMRs, in the same order, are 1.95, 1.97 and 2.56. They were expected to be higher for longer sample sequence.

In [7], three different coding methods were applied to the same two 20 seconds, 15fps and 256x240 resolution conference sequences, one inactive and one active. The three coding schemes are: DPCM, DPCM plus MC and DCT plus MC. For the inactive scene, SDMRs are 0.2, 0.3 and 0.3 and PMRs are 3.1, 2.6 and 1.9, in the same order of above coding schemes. Then for the active one, SDMRs are 0.4, 0.4 and 0.3; PMRs are 3.4, 2.5 and 1.8. The authors also presented the pdf and autocovariance data, both of which are claimed to be similar for different coding schemes. The pdfs are bell shaped. The autocovariance curve of the active scene is smoother than the inactive one. At 0.5 second separation, the coefficient of inactive scene is around 0.3~0.4 while it's almost zero for active scene. At 1 second separation, coefficients for both scenes are around 0.1~0.3.

Lastly, we get some useful statistics from coding various video scenes using H.263 codec. The sequence format is QCIF and input frame rate is 25 fps and output 12.5 fps with one exception. Quite a few video sequences were used for this purpose. They are listed in Table 2.1 (name, description, length, SDMR and PMR) together with their SDMR and PMR characteristics. Their pdf and autocovariance curves are presented in figure 2.5 and figure 2.6.

By comparing all the mentioned references as well as our statistics, we reached the following conclusions:

Table 2.1: Video Sequences Info and Statistics

Name	Descriptions	Frames	SDMR	PMR
Mother and Daughter	A mother caressing her daughter.	479	0.51	2.96
Grandmother	An old lady talking.	434	0.4	2.17
Salesman	TV indoor commercial, salesman behind table, taking and waving arms while sitting on a chair.	224	0.43	2.32
Suzie	A lady listening to phone, turning head, close up.	74	0.51	2.36
Claire	Single conference speaker, relative static.	246	0.27	1.84
Carphone	A gentleman driving a car, talking, laughing, waving hands, scenes outside window backing.	190	0.49	2.34

- Both SDMR and PMR depend on both scene contents and compression algorithms. Active scenes result in higher peak to mean ratio (Table 2.1, [6] and [13]), at least for transform coding. The PMR of DPCM coding is larger than that of DPCM plus Motion Compensation (MC), and the PMR of DPCM plus MC coding is larger than that of DCT plus MC (compare Table 2.1, [5], [7], [13], [21]).
- The pdf of bits or packets per frame can be approximated as a gamma distribution (Figure 2.5, [9]), rather than Gaussian as used in [7] and [11], at least for videophone/conference type video. This conclusion is not confined to H.263, not even DCT + MC. It holds for both Prediction coding and Transform coding. Results for other types of coding are unknown but expected to be similar because they reflect the nature of entropy variations of video sequences.

(a)

(b)

(c)

(d)

(e)

(f)

Figure 2.5 PDFs of Coded Video

(a)

(b)

(c)

(d)

(e)

(f)

(g)

(h)

Figure 2.6 Autocovariances of Coded Video

-
- The autocovariance depends mostly on scene contents and is independent of coding frame rate or output bit rate (f, g and h of Figure 2.6). While some of the scenes have repeated peak of autocovariance, the reason is more likely due to short sequence and computation truncations (compare a, b and d, f of Figure 2.6, [9], [11] and [21]). Exponential autocovariance is a proper model ([7], [9] and [11])

Gamma distribution is a continuous version of the Negative Binomial distribution. The pdf function of Gamma distribution is as follows:

$$f_X(x) = \frac{\lambda (\lambda x)^{S-1} e^{-\lambda x}}{\Gamma(S)} \quad (\text{EQ 2.5})$$

where $\Gamma(S)$ is the so-called gamma function which is:

$$\Gamma(s) = \int_0^{\infty} x^{s-1} e^{-x} dx \quad (\text{EQ 2.6})$$

The relations of λ and S to the mean m and variance σ^2 are:

$$S = \frac{m^2}{\sigma^2} \quad \lambda = \frac{m}{\sigma^2} \quad (\text{EQ 2.7})$$

or:

$$m = \frac{S}{\lambda} \quad \sigma^2 = \frac{S}{\lambda^2} \quad (\text{EQ 2.8})$$

The characteristic function of Gamma pdf is:

$$\Phi_x(\omega) = \left(\frac{\lambda}{\lambda - j\omega} \right)^S \quad (\text{EQ 2.9})$$

We can see that when s is an integer, say l , the Gamma distribution becomes l -Erlang distribution which is defined as:

$$f_X(x) = \frac{\lambda (\lambda x)^{l-1} e^{-\lambda x}}{(l-1)!} \quad (\text{EQ 2.10})$$

It's well known that a l-Erlang random variable is a sum of l Independent Identical Distributed (i.i.d.) exponential random variables with the same rate λ . Given that the characteristic function is the Foulrier transform of the pdf, and when the sum of the several independent random variables has a characteristic function which is simply a multiplication of their characteristic functions, we can reach the following conclusion:

- sum of N independent Gamma random variables with the same parameter λ and different parameter S_n ($n=1,2,\dots,N$) is still a Gamma random variable with parameter λ and $S = \sum_{n=1}^N S_n$.

We can see that when S is an integer or when S is very large, a Gamma random variable can be replaced with s $[S]$ -Erlang random variable without much loss of accuracy, where $[S]$ stands for the biggest integer that is not greater than S . The advantage of doing so is that a l-Erlang random variable can be easily generated from l i.i.d uniform random variables. Explanation follows.

Given l i.i.d. uniform random variable Y_i ($i=1,2,\dots,l$) with pdf as:

$$f_{Y_i}(y_i) = 1 \quad 0 \leq y_i \leq 1 \quad (\text{EQ 2.11})$$

Then the cdf of random variable $X_i = -\frac{1}{\lambda} \ln(Y_i)$ can be deduced as:

$$F_{X_i}(x_i) = P(X_i < x_i) = P(Y_i > e^{-\lambda x_i}) = 1 - e^{-\lambda x_i} \quad (\text{EQ 2.12})$$

This is one the familiar exponential cdf. There is no doubt that the X_i s are i.i.d. Then $X = \sum_{i=1}^l X_i$ inevitably leads to a l-Erlang random variable. So we get another conclusion here:

- A Gamma random variable with integer or large parameter S can be modelled as a summation of $[S]$ i.i.d exponential random variables. And the exponential random variables can be easily generated by taking the logarithm of a uniform random variable.

2.3.2 Video/Videophone Modelling.

In [33], it's stated that in order to achieve any statistic multiplexing gain, buffer of at least one frame coding output is needed and according to [6], smoothing between frames is claimed not necessary. So we will model coded videos frame by frame.

In [11], the basic models of Autoregressive (AR) and Multi-states Markov Chain were proposed. Later various combinations of them with different order were proposed to model different kind of videos [5~10] [12] [22] [34]. Among these, [10] give a good summary, [9] tested several kinds of models with a real long sequence and proposed a DAR model as the most suitable for video conference and videophone type applications. AR(1) model is claimed not good in residue test and big frame test [9] although it modeled the autocovariance really good.

The idea of AR(1) model is that we assume the random process is an autoregressive process Markov process with the following definition:

$$X(n) = aX(n-1) + b + e(n) \quad (\text{EQ 2.13})$$

where $e(n)$ is a sequence of independent zero mean random variable with variance σ_e^2 , not necessarily to be Gaussian. Then the mean and autocovariance of the process can be expressed as:

$$m_X = E[X] = \frac{b}{1-a} \quad C_X(n) = \sigma_X^2 a^n = \frac{\sigma_e^2}{1-a^2} a^n \quad (\text{EQ 2.14})$$

Given a sequence of N samples of a random process, say X_i ($i=1,2,\dots,N$), we can estimate the three parameters, i.e. a , b and σ_e^2 using EQ 2.14. The residues of the AR(1) modelling is defined as:

$$R(n) = \bar{X}(n) - X(n) = aX(n-1) + b - X(n) \quad (\text{EQ 2.15})$$

for $n=2,3,\dots,N$. An ideal stochastic model which catches all the characteristic of a random process will lead to zero mean and independent residues [9]. Practically, residues should have a negligible mean and very small autocovariance, i.e. $C_R(n) \approx 0$ for $n \neq 0$. This will depend on how large t N is.

We performed fit tests of the AR(1) modelling for some coded video. Due to one of the conclusions in 2.4.1, i.e. small peak to mean ratio, AR(1) model doesn't have the big frame problem mentioned in [9], and it's also verified good in residue test in Figure 2.7 given the short video sequences we have in Table 2.1. The AR(1) parameters are presented in Table 2.2. The AR(1) model is a very simple, but yet good model for video statistical modelling and simulations.

(b)

(c)

Figure 2.7 Residue Test of AR(1) Model

Table 2.2: AR(1) Model Parameters for Video Sequences

Video Sequences	m_x in bits	σ_x^2 in bits ²	$C(1)/\sigma_x^2$	a	b	σ_e^2
“Carphone”	2507	1422500	0.38	0.9255	186.77	204057
“Mother and Daughter”	1119	324840	0.44	0.9364	71.19	40006
“Grandmother”	582	54184	0.16	0.8636	79.39	13773

Now we have one more conclusion:

- For video coded with small peak to mean ratio, the AR(1) model has good independent residue performance. and it is a very good and simple candidate for statistical modelling and simulation.

CHAPTER 3

Mobile Channel, Microcellular and CDMA Systems

In this chapter, we will first discuss general characteristics and receiving techniques of wireless mobile channel. It will be followed by microcellular system discussion. Then we will concentrate on CDMA specific signal, channel receiver and other related issues.

3.1 Mobile Channel Classifications and Receiving Techniques

Mobile communication channels, are characterized by fading and time varying impulse response [35-38]. They are caused by multipath propagation and movements in the environments. Depending on the actual system, the channel can be classified as slow or fast fading channel and selective or non-selective fading channel. To understand these, we re-produce a commonly used, nomalized frequency-time correlation function [38] as EQ 3.1:

$$\rho(\Delta f, \tau) = \frac{J_0^2(2\pi f_m \tau)}{1 + (2\pi \Delta f \sigma)^2} \quad (\text{EQ 3.1})$$

Here Δf is the frequency difference, τ is the time difference, f_m is the maximum doppler shift which depends on moving speed and carrier frequency, σ is the power delay spread of the channel. In [39], coherent bandwidth B_c and coherent time T_c are defined as following:

- a. Two sinusoids with frequency separation greater than B_c are affected differently by channel.
- b. Two identical signals separated by a time delay larger than T_c will be affected independently.

If we choose some value of correlation, usually one half, as the separation of coherence and non-coherence as in EQ 3.2, we get B_c and T_c are proportional to the reciprocal of σ and f_m respectively. We can classify channels according to Figure 3.1. When the symbol period

Figure 3.1 Channel Classifications

If the symbol period T of the system is relatively smaller than T_c , successively transmitted symbols undergo similar transformation which is called slow fading (time flat). Otherwise they get almost independent transformation, this is fast fading.

$$\rho(B_c, 0) = \rho(0, T_c) = \frac{1}{2}\rho(0, 0) = \frac{1}{2} \quad (\text{EQ 3.2})$$

For fast fading slow channels, fast adaptive channel estimation and signal process are required to receive signal successfully. Usually both indoor and outdoor channels in urban environment only suffer slow fading. This means that we can estimate the channel response accurately for each symbol period and the estimation only changes slowly with time.

3.1.1 Slow-Flat Fading Channels and receiving technique

In addition to Additive White Gaussian Noise, slow-flat fading channel has the problem of fading. It can be Rayleigh distributed if just scattered signals are received or Rician distributed if a LOS path exists. Usually the channel can be modelled as a multiplicative process having the diagram as illustrated in Figure 3.2. A simple receiver is also shown in this diagram and we assume BPSK modulation. $U(t)$ is the channel input and $R(t)$ is the output. $Z(t)$ is the AWGN. $\alpha e^{-j\varphi}$ represents the low pass equivalent of the multiplicative process. α is the envelop which can be Rayleigh or Rician distributed and φ is the phase, usually uniform distributed. Y_n is the decision variable.

Figure 3.2 Multiplicative Model and Receiver for Fading

The performance of the system above is not as good as that of AWGN-only system, but we can use diversity to improve the performance. The diversity can be done in space, frequency or in

hybrid, etc. The prerequisite for achieving improvements is that signals coming from different diversity branch must experience independent or at least only slightly correlated fading. Many kinds of combining methods exist, such as maximum ratio combining, selective combining, and equal gain combining, etc. A diagram of maximum ratio combining diversity receiver, which is the optimal receiver for slow-flat channel, is shown in Figure 3.3.

Figure 3.3 Diversity Receiver (Maximum Ratio Combining)

3.1.2 Slow frequency-selective fading channels and receiving technique

When the transmission bandwidth W is much larger than B_c , the channel is classified as non-flat fading channel and there are two kinds of them. Together with this sub-classification and similar to [38] [39], we present some other definitions:

Figure 3.4 Pulse Duration and Symbol Period

- a. **Transmission bandwidth** W is $1/\tau$ where τ is pulse duration as illustrated in figure 3.4.
- b. **Information bandwidth** B_i is $1/T$ where T is symbol period as illustrated in figure 3.4.
- c. When $W \gg B_c \gg B_i$, the channel is called **spread spectrum channel**.
- d. When $W = B_i \gg B_c$, the channel is called **broadband channel**.

3.1.2.1 ISI and equalization

Before we go into details of broadband and spread spectrum channels, It is necessary to discuss inter-symbol interference (ISI) and the according combatting technique, i.e. equalization.

First, what is ISI? The general definition is that previous symbols are still “ringing” during current symbol’s reception. ISI add irreducible errors if we just use the diversity receiver in section 3.1.1. However, any ISI can be handled no matter how difficult it is theoretically. It’s just a matter of complexity and cost. Here we first present a discrete model of ISI and a simple linear equalizer as shown in figure 8. We can see that the effects of ISI and equalizer can be treated as two filters.

Figure 3.5 Discrete Model for ISI and Linear Equalizer

The output of channel, r_n , can be expressed as:

$$r_n = \sum_{i=0}^N h_i S_{n-i} \quad (\text{EQ 3.3})$$

When $N=0$, it's according to the special case of no ISI. And the final output is expressed as:

$$y_n = \sum_{i=0}^{M+N} x_i S_{n-i} \quad (\text{EQ 3.4})$$

where

$$x_n = \sum_i h_i c_{n-i} \quad \text{for } n=0,1,2,\dots,M+N. \text{ (} c(n) \text{ has } M+1 \text{ taps)} \quad (\text{EQ 3.5})$$

Since our goal is to make $y_n=a_n$, we would like to let x_n equals 1 for $n=0$ and 0 otherwise. So we get $M+1$ variables to solve $M+N+1$ equations. Generally speaking, we can only achieve zero ISI when we M tends to infinity. Practically, it's acceptable to just decrease ISI to some extent. Thus we just adjust the $M+1$ taps according to some algorithms such as zero forcing, minimum mean square error (MMSE) and maximum likelihood algorithms. In the sense of minimize BER, maximum likelihood algorithm is the optimal one. The famous Viterbi algorithm is an efficient way to implement it for short channel impulse response [39]. When the channel coefficients are unknown, we can use adaptive equalizer plus training sequence and if we wish to let the algorithm to converge faster and achieve smaller MMSE, decision feedback can be used. Moreover, when the $h(n)$ s are time varying which is the case for mobile channel, recursive least mean square (RLMS) equalizer can handle this. A detailed discussion of all these varieties of equalizers is beyond the scope of this paper.

3.1.2.2 Spread spectrum channel and RAKE receiver

As we have defined, for spread spectrum channel $W \gg B_c \gg B$, which means $T \gg \sigma \gg \tau$. So there is no ISI but within one symbol period, there are as many as $N+1$ delayed versions of the signal pulse where $N = \lceil \sigma/\tau \rceil = \lceil W/B_c \rceil$. Here for simplicity, we just assume $W = 1/\tau$ and $B_c = 1/\sigma$.

Moreover, since these received signals go through different propagation paths, we can treat their fading as independent (uncorrelated scattering assumed here). This reminds us the application condition for diversity technique in section II. The receiver utilizing this inherent diversity is so called RAKE receiver. Its diagram for BPSK is illustrated in the following figure.

Figure 3.6 Diagram of RAKE Receiver (BPSK)

Here $R(t) = \sum_{k=0}^N \alpha_k e^{-j\phi_k} U(t - \frac{k}{W}) + Z(t)$. We assume the estimator can estimate the $\alpha_k e^{-j\phi_k}$ accurately because the fading is slow. In this approach we can get similar performance as the case of $N+1$ branches diversity. (maybe a little worse because the mean power of $\alpha_k e^{-j\phi_k}$ for different k are weighted differently. A common case is that it decays exponentially.)

Here maximum ratio combining is employed and the delay tap is chosen to be $1/W$ because the resolution of a W bandwidth can be chosen as $1/W$. Practically equal gain combining is often used in order to avoid the complexity of the channel estimator. The resultant performance is degraded due to possible noise and interference amplifying when signal suffers deep fading

3.1.2.3 Broadband channel and receivers

In broadband channels, we are facing both fading and ISI. This is the most troublesome case. If we just make hard decision upon what's received, we will get large irreducible BER even when E_b/N_o is quite high. However, some combatting technique exist. Most of them are modifying the channel to get the situation of spread spectrum so that we can avoid ISI and make use of the inherent diversity. There are

- a. Diversity-Equalizer combatting ISI plus fading. [40]
- b. Increasing modulation alphabet.
- c. Commutating Signalling. [41]
- d. Coded Orthogonal Frequency Division Multiplexing (COFDM). [42]
- e. Hybrid receiving technique which is a combination of two or more of above.

A detailed explanation of all the above techniques are beyond the scope of this literature. A summary table is given in Table 3.1. Here we give some introduction to Commutating Signalling because it's close related to CDMA. Solution c can be regarded as a smart improvement of 3.3.2. And there are two version of it as below. Similar to the last section, we define the resolvable delayed rays as $N + 1$ where $N = W/B_c$.

- 1) Commutation among $M \times N$ signals. i.e. N sets distinct M -ary signals.
- 2) Commutation among $M + N$ signals.

A simple choice of signal sets involves the use of sine waves; an $M \times N$ code would then consist of N distinct sets of M frequencies each, while an $M + N$ code would consist of $M + N$ frequencies, of which only M are used at any time. A more complicated choice involves sets of co-spectral signals with large time-bandwidth products (spread spectrum signals). Because spread

spectrum signals allow easy exploitation of multipath diversity gains, we usually prefer these signals. Here M is the alphabet size of each signal set. Let R be the data rate in bits/s. Then the signal duration is $T = \log_2 M / R$ and let $K = \lceil P \sigma / \log_2 M \rceil + 1$, i.e. $\lceil \sigma / T \rceil + 1$. So $M \times N$ codes require $N = K$ signal sets resulting in a total of MN signals and $M + N$ requires $M + (K - 1)$ signals. For example, if $M = 2$ and $K = 4$, these two commutation methods will require eight and five signals respectively while solving this just by enlarging alphabet size demands sixteen signals. This is why commutation signalling is smarter.

Table 3.1: Comparison of Broadband Receiving Techniques

Receiving techniques	Ways of cancelling ISI	Ways of suppressing fading	Advantage	Disadvantage
(1) Diversity and Equalizer combining	by diversity equalizer combining	diversity	little or no expanding of bandwidth.	complexity. channel estimation. time-correlated fading can destroy converging.
(2) Enlarging modulation alphabet	enlarge symbol duration to achieve spread spectrum channel for bits	inherent diversity in broadband	direct to implement	waste bandwidth or energy
(3) Commutating signalling	signal orthogonality (frequency or code)	inherent diversity in broadband	relative simple compared to (1) and (4), relative efficient compared to (2)	complex code and/or bandwidth expansion
(4) COFDM	reduce effective W and B as defined in 3.1.2 for each bit to achieve flat fading channel per bit	coding & interleaving	attractive modulation scheme. little or no bandwidth expansion	complex code must needed and delay by interleaving

To illustrate how ISI is avoided in this scheme, we take the example of $M \times N$ for $M = 2$ (BPSK) and $K = 4$. The receiver diagram is shown below. Since the matched filters are matched to a sets of orthogonal frequencies or codes, other three filters will have no outputs when one signal set is chosen to transmit. And since the 4 sets are chosen alternately and $N=K$, when a set is chosen for the second time the ringing of previous bit in this set has died out. And each MF output can be connected to a RAKE receiver to make use of the inherent diversity to combat fading.

Figure 3.7 Diagram for $M \times N$ Commutation Receiver (BPSK)

The $M - N$ commutation seems to have advantage over $M \times N$ commutation in the sense of lit requires less signal sets. But the price we have to pay for this advantage is the complexity. We have to decide which set of M-ary signals to choose for the next transmission and this involves the problem of error propagation [41].

In fact when commutation is realized by ways of orthogonal codes, it's a sort of Direct Sequence Spectrum technique (DS-SS). And when the commutation is realized by selecting among many carrier frequencies (randomly or not), it's a sort of Frequency Hopping Spread Spec-

trum technique (FH-SS). The difference between such commutating signalling and the so called Spread Spectrum technique is that the latter may have longer and more complex code and/or more bandwidth and much lower information rate so that it may not need to worry about ISI and change every bit as the commutating signalling does. Another fact is that the spread spectrum channel introduced in section 3.1.2 can be regarded as a simple case of the Spread Spectrum techniques mentioned here. We will discuss more about Spread Spectrum and so called CDMA in section 3.3

3.2 Microcellular Channel and Macro/Micro Cellular Overlay

Microcellular channel (200m-1km) has been proposed for PCS applications. A lot of measurements have been done to characterize the channel [43–56].

The shadowing distribution has been commonly accepted as Lognormal pdf. But there are some disagreement on fading pdf. In [43], the indoor fading PDF is claimed to be Lognormal too. In [45] it's Rician with factor 7 dB. We study the measurement in details and conclude that the disagreement come from the measurement resolution. In [43], the time multipath resolution is so small that not enough multipath was presented in each bin so that the sum mainly exhibits shadowing. We will see in chapter 4, our system bandwidth is similar to that of [45]. In [48] urban microcellular channel is also claimed to be Rician with factor 6-8 dB.

It's been measured that by a bin time of 100 ns, the second bin and third bin are significantly weaker than the first LOS one, the difference range from 6-8 dB in [52], 10dB in [45] to nearly 20 dB in [46] [49]. While this makes the multipath RAKE receiver very inefficient, it contribute little interferences (ISI) in CDMA channel. Plus considering the break-point effect or Fresnel Zone effect verified in [50] [51] [53], the adjacent channel interference can be reduced by a factor up to one half.

The above summary is based on both indoor and outdoor measurements. The differences between these two kinds of channels are as follows: first, the cell size of indoor channel is even smaller than outdoor channel. (with radius below 100m, so called pico cell); second, fading in indoor environment is not as obvious as outdoor and for wide bandwidth, as we discussed before, shadowing will dominate the channel; furthermore there are usually more bandwidth available for indoor wireless communications. Since our system bandwidth is not big enough to exhibit shadowing domination, we don't discriminate between indoor and outdoor cellular channels.

Microcellular is not suitable for high speed mobile communications due to frequent handoffs. So the Doppler frequency is usually just 2 Hz (1km/h). This leads to highly correlated fading with correlation time equals to video frame duration. Good thing is that Rician fading is not as severe as Rayleigh fading, and antenna diversity can make it even less. As we will see in chapter 4, this will not decrease the service very much.

Although LOS Rician fading in microcellular system provide better fading performance and other advantages as mentioned above. When there is a LOS obstructed user, his reception will be very bad and it may be solved by just using power control to increase that user's transmission power. But some time increasing one user's transmission power too much may results in system instability because other users suffer more interference. In this case, the system administration will hand it over to another nearby base station which can provide a LOS channel. If no LOS exist in nearby microcells, the system can order the user to use the Macrocell overlaid channel. The overlaid Macrocell channel can also be used to handle those fast moving users who will result in too frequent handoffs.

So with above cellular channel schemes, the system will accommodate various users while most of them only suffer Rician fading. Given that microcellular users move relative slowly (walking speed), both Rayleigh and Rician fading can be reduced greatly in CDMA with power control schemes. Details in following sections.

3.3 Code Division Multiple Access (CDMA) System

Spread Spectrum technique was originally used in military communications to eliminate both broadband and narrowband jamming and to hide the signal in background noise to achieve low detectability. Recently, it has proposed and implemented [17~19] in cellular multiple access communication systems, i.e. CDMA, to provide commercial voice phone service.

CDMA (Code Division Multiple Access) is a multi-users access scheme to use different long Pseudo-Noise (PN) binary sequences to either multiply each user's information sequence which is called Direct Sequence CDMA (DS-CDMA), or control each user's carrier frequency hopping pattern which is called Frequency Hopped CDMA (FH-CDMA). In either case, the overall frequency bandwidth used by each user is much larger than the user's information bandwidth so CDMA channel is a kind of Spread Spectrum (SS) channel. Another aspect of CDMA is that multiple users are using the same frequency band meaning that users are co-channel interference to each other. But this co-channel is still separated by a firewall of PN code which makes the interference noise like. By using long PN sequence, it achieves the advantages of anti-jamming, inherent multipath or frequency diversities and user privacy and etc. And by transmitting all users' signal in the same frequency-time domain, DS-CDMA achieves flexibilities such as soft service degradation with the increase of user number, small frequency reuse factor, very easy to make use of voice activities and antenna sectorization to improve capacity, soft handoff and etc.

3.3.1 CDMA Signal, Channel and Receiver

Similar to the approaches of [56~61], we first give a mathematical formula description of CDMA BPSK signal, channel and coherent receiver. The diagram for these parts is shown in Figure 3.8.

The k th user's signal come out from his transmitter with initial transmit power of P , power control coefficient of β_k , carrier frequency of ω_c and initial phase of θ_k can be described as:

$$s_k(t) = \sqrt{2P}\beta_k c_k(t) d_k(t) \cos(\omega_c t + \theta_k) \quad (\text{EQ 3.6})$$

where

$$c_k(t) = \sum_{n=-\infty}^{\infty} c_k^{(n)} P_{T_c}(t - nT_c) \quad (\text{EQ 3.7})$$

$$d_k(t) = \sum_{n=-\infty}^{\infty} d_k^{(n)} P_{T_d}(t - nT_d) \quad (\text{EQ 3.8})$$

and $P_T(t)$ is a pulse waveform of period T , T_c and T_d are chip duration and data duration respectively, $c_k^{(n)}$ and $d_k^{(n)}$ are the chip bit and data bit respectively, with value of either 1 or -1.

Figure 3.8 CDMA BPSK Signal, Channel and Receiver (User k)

Assuming there are K users and each user's signal experiences up to L multipath fading, then the received signal at the a th diversity branch will be:

$$r^a(t) = \sum_{k=1}^K \sum_{l=1}^L \sqrt{2P} h_{kl}^a \beta_k c_k(t - \tau_{kl}^a) d_k(t - \tau_{kl}^a) \cos(\omega_c t + \phi_{kl}^a) + z^a(t) \quad (\text{EQ 3.9})$$

here h_{kl}^a is the fading envelop, τ_{kl}^a is the path time delay and ϕ_{kl}^a is the composite phase which is:

$$\phi_{kl}^a = \theta_k - \omega_c \tau_{kl}^a - \phi_{kl}^a \quad (\text{EQ 3.10})$$

ϕ_{kl}^a is the path phase delay. Then the output from the correlator of the j th multipath of the i th user, according to the 0 th data bit, is:

$$Y_{ij}^a = \int_{\tau_{ij}^{(a)}}^{T+\tau_{ij}^{(a)}} r^a(t) \sqrt{2} \cos(\omega_c t + \phi_{ij}^a) c_i(t - \tau_{ij}^a) dt \quad (\text{EQ 3.11})$$

which can be deduced to:

$$Y_{ij}^a = \sqrt{P} h_{ij}^a \beta_i d_i^{(0)} T + SMI + MAI + AWGN \quad (\text{EQ 3.12})$$

where SMI denotes Self Multi-path Interference which is:

$$SMI = \begin{cases} \sum_{l=1, l \neq j}^L \sqrt{P} h_{il}^a \beta_i [d_i^{(-1)} R_{ii}(\Delta\tau) + d_i^{(0)} \hat{R}_{ii}(\Delta\tau)] \cos \Delta\phi, & (\Delta\tau > 0) \\ \sum_{l=1, l \neq j}^L \sqrt{P} h_{il}^a \beta_i [d_i^{(0)} R_{ii}(T + \Delta\tau) + d_i^{(1)} \hat{R}_{ii}(T + \Delta\tau)] \cos \Delta\phi, & (\Delta\tau < 0) \end{cases} \quad (\text{EQ 3.13})$$

here $\Delta\tau = \tau_{il}^a - \tau_{ij}^a$, $\Delta\phi = \phi_{il}^a - \phi_{ij}^a$ and

$$R_{pq}(\tau) = \int_0^\tau c_p(t - \tau) c_q(t) dt \quad \hat{R}_{pq}(\tau) = \int_\tau^T c_p(t - \tau) c_q(t) dt \quad (\text{EQ 3.14})$$

and MAI denotes Multiple Access Interference which is:

$$MAI = \begin{cases} \sum_{k=1, k \neq i}^K \sum_{l=1}^L \sqrt{P} h_{kl}^a \beta_k [d_k^{(-1)} R_{ki}(\Delta\tau) + d_k^{(0)} \hat{R}_{ki}(\Delta\tau)] \cos \Delta\phi, & (\Delta\tau > 0) \\ \sum_{k=1, k \neq i}^K \sum_{l=1}^L \sqrt{P} h_{kl}^a \beta_k [d_k^{(0)} R_{ki}(T + \Delta\tau) + d_k^{(1)} \hat{R}_{ki}(T + \Delta\tau)] \cos \Delta\phi, & (\Delta\tau < 0) \end{cases} \quad (\text{EQ 3.15})$$

here $\Delta\tau = \tau_{kl}^a - \tau_{ij}^a$, $\Delta\phi = \phi_{kl}^a - \phi_{ij}^a$ and the definition of $R_{pq}(\tau)$ and $\hat{R}_{pq}(\tau)$ are the same as above.

the last item is the modulated Additive White Gaussian Noise coming with the signal:

$$AWGN = \int_{\tau_{ij}^a}^{T + \tau_{ij}^a} z^a(t) \sqrt{2} \cos(\omega_c t + \phi_{ij}^a) c_i(t - \tau_{ij}^a) dt \quad (\text{EQ 3.16})$$

3.3.2 CDMA Power Control.

From the previous section, we can see that CDMA is mainly interference limited instead of noise limited in TDMA and FDMA. Interferences to one user are valuable information signals to other users. So from a system point of view, we can't just improve the performance of a certain user by reducing all other users' signal power. Besides, there is already a near/far problem existing in CDMA. The solution to overall capacity improvement is power control [17] [18], i.e. the transmitting power is controlled such that signal received from and by each user are equalized, or in other words, balanced.

Although power control scheme was originally intended to overcome the near/far problem in CDMA, it's also used to mitigate multipath fading as well as shadowing and it has been proven to work just fine for low Doppler frequencies [62~65]. For our microcellular CDMA system, slow Rician fading plus a little Rayleigh fading characterizes the channel. There is no doubt that fast power control can be used here to cancel fading.

Theoretically, power control adjustment should be made upon the received signal power after eliminate the interference components which is difficult to implement. Practically, power control based on received signal-to-interference power ratio (SIR) [62] [64] [66] will serve the purpose.

The adjustment of transmitted power is done by multiply the original signal with a adjustment factor β as in section 3.3.1.

Ideally, after implementing power control, we can face a much better AWGN like channel. However, due to the fact that power control signal is also transmitted through fading and shadowing channel and other practical reasons like time lag, etc., the adjustment can be received wrong. So the actual received SIR after power control is still a random variable and has been measured to be Log Normal distributed [62] [63] [67]. In the following paragraph, we give some useful parameters about these Power Controlled CDMA systems for future references.

In both [62] and [64], the power control signalling rate are 800 times per second while the former aimed at pure 1.25MHz IS-95 type system and the latter discussed some variations of it, including a 10 MHz broadband version. In fact, given higher information signal rate, the power control signal can be transmitted at higher rate accordingly like 2000 times per second in the European CODIT system [16]. The resulting performance will be bounded by that of the necessary rate needed to combat the variation speed of fading.

As for power control errors, the standard deviation of the Log Normal distributed received SIR was analyzed to be 1.1~1.5 dB in [68] and measured to be 1.5 dB for low mobility (slow fading) users and 2.5 dB for high mobility (fast fading) users in [62].

3.3.3 CDMA Coherent Reception

It's well understood that coherent reception of desired signal in noise/interference affected environment can provide better performance. When CDMA was proposed for commercial cellular systems in the early days [17] [18], coherent reception was chosen for downlink and noncoherent reception for uplink. It's commonly accepted that system capacity is limited by uplink mainly due to severe near/far problems in uplink. Noncoherent reception is not the major limiting factor

here and it will make theoretical analysis easier as well as offer performance improvement if uplink signal can be coherently received.

In fact CDMA uplink coherent reception has already been proposed [69] [70] and implemented in CODIT test bed [16]. All these schemes utilize control symbols that transmitted along with information signals to assist coherent detection of information signals. While the system described in [70] appeared to be IS-95 type system, i.e. 1.25 MHz bandwidth, the other two systems are all broadband, e.g. a 10 MHz CDMA system was discussed in [69] and both 5 MHz and 20 MHz systems are discussed in [16]. Given the same slow fading environment, higher signal data rate which is used in broadband CDMA system can surely make coherent detection more possible and more stable.

Inevitably, control symbols add overhead to the data stream. The ratio in the above three references can give us some idea of the trade-off. In [70] the control symbol is about one sixth of the total signal, one eighth in [69] and one fifth in [16]. In [16], the one fifth is not just for assisting coherent receiving, it's used for all control signalling including guarding period, training sequence, power control signalling and etc.

3.3.4 Multiple Access Interference (MAI) Modelling

MAI has been modelled as White Gaussian Noise in [39] [56~61] [71] with different conditions. General rule is that when processing gain is large (greater than 64) and number of multi-users is large where central limitation theory can be invoked, Gaussian Approximation has good accuracy [57~61]. There are also modified or improved Gaussian Approximations which perform good even with small PG and user numbers [71~73].

Given the characteristics of pseudo-random sequences [39] [67], they have close to zero autocorrelation and crosscorrelation except the zero point of autocorrelation. Usually the autocorrela-

tion is better or smaller than crosscorrelation. So we can model SMI in the same way of MAI to get conservative results.

Since we consider system capacity in the sense of maxim number of users allowable and our PG is relative large. Gaussian Approximation can be good enough for us. But considering accuracy of worst case performance, i.e. big standard deviation of the received SIR, we finally adopt improved Gaussian Approximations. A comparison of these two models is given in Figure 3.9.

Figure 3.9 Comparison of Standard Gaussian Approximation and Improved Gaussian Approximation

The formulas are reproduced here for single diversity branch receiving and no multipath and no fading case [71~73] which we can use for later analysis in chapter 4. In this case the decision variable of the receiver in Figure 3.8 is:

$$Y_i = \sqrt{P} T d_i^0 + \sum_{k=1, k \neq i}^K \sqrt{P} [d_k^{-1} R_{ki}(\Delta\tau_{ki}) + d_k^0 \hat{R}_{ki}(\Delta\tau_{ki})] \cos \Delta\phi_{ki} \quad (\text{EQ 3.17})$$

note AWGN is ignored here and without loss of generality, the delay of other users is assumed to be bigger than that of the i th user. Applying Improved Gaussian Approximation, the received SIR is a random variable with mean

$$\mu = \frac{3N}{K-1} \quad (\text{EQ 3.18})$$

and variance

$$\sigma^2 = (K-1) \left[N^2 \frac{23}{360} + N \left(\frac{1}{20} + \frac{K-2}{36} \right) - \frac{1}{20} - \frac{K-2}{36} \right] \quad (\text{EQ 3.19})$$

here N is the processing gain or PN code length. And the BER which is deduced [71] [73] to be

$$P_e = \int_0^\infty Q(\sqrt{SIR}) f(SIR) dSIR \quad (\text{EQ 3.20})$$

can be closely approximated as:

$$P_e \approx \frac{2}{3} Q(\sqrt{\mu}) + \frac{1}{6} Q(\sqrt{\mu + \sqrt{3}\sigma}) + \frac{1}{6} Q(\sqrt{\mu - \sqrt{3}\sigma}) \quad (\text{EQ 3.21})$$

which leads to the following result:

$$P_e \approx \frac{2}{3} Q\left(\sqrt{\frac{3N}{K-1}}\right) + \frac{1}{6} Q\left(\frac{N}{((K-1)N/3 + \sqrt{3}\sigma)^{0.5}}\right) + \frac{1}{6} Q\left(\frac{N}{((K-1)N/3 - \sqrt{3}\sigma)^{0.5}}\right) \quad (\text{EQ 3.22})$$

The result in [73] showed good accuracy for probability error greater than 10^{-5}

CHAPTER 4

CDMA Videophone System Capacity Analysis

So far we have discussed VBR coded video and CDMA channel & signal separately. We want to analyze the ability/capacity of DS-CDMA system to carry VBR video traffic, which is the most resource demanding traffic with real time requirement.

In this literature, we will concentrate on theoretical analysis of CDMA uplink capacity because it's claimed to be the limiting link of CDMA capacity [17] [18]. But even for theoretical analysis purpose, we need to consider some practical issues before we can go on. The system that we are considering is illustrated in Figure 4.1.

4.1 Practical Considerations

Given the video data analysis in section 2.3 and without loss of generalization, we assume each user sends coded (both video encode and source encode) video streams with 32 kbps mean rate, 80 kbps peak rate and 16 kbps standard deviation which results in a PMR of 2.5 and a SDMR of 0.5. The video data rate has about one eighth source coding overhead in it and has already interleaved on a frame by frame base. The reason for source coding and interleaving is given in a

later paragraph. As for the parameters of CDMA system, we consider uplink with bandwidth around 13 MHz and process gain of 127, which offers a data rate of 100 kbps. Minus 80 kbps video peak data rate, the remaining 20 kbps is used as control symbols to provide channel signaling including assisting coherent uplink reception. Since we are analyzing a basic model, cell sectorization is not considered.

**May including interleaving and de-interleaving*
***May be combination of PN coding and Error Control Coding*

Figure 4.1 CDMA Videophone System Diagram

Utilizing those microcellular measurement results we mentioned in section 3.2, we decided to employ diversity receiver (antenna, space, multipath, etc.) together with power control to combat fading. Now we are facing SMI and those multipath MAI. [39] mentioned that the PN sequence has better, i.e. smaller autocorrelation than cross-correlation. So if we model them the same way as the first path MAI, i.e. AWGN modelling which will just increase the variance of the equivalent noise, it will make the capacity smaller than it should be. Fortunately, the power of these interferences are 7-20 dB below those first path MAI as we summarized in section 3.2, they shouldn't affect the result too much. In special case that multipath becomes bigger than the slight multipath case, Rake receiver can be deployed to improve receiving. This is beyond the scope of this literature. As for AWGN, it's quite safe for us to ignore it since CDMA is interference limited. Since we evaluate our system capacity in a cellular environment, it's necessary to consider other cell interferences. similar to dealing with SMI and multipath MAI, we multiply the variance by a factor of the equivalent noise. The ratio of all the other cell interference power to intra-cell interference power was theoretically analyzed to be 0.56 in [17] and 0.55~0.65 in [62] [74]. Counting microcell breakpoint effect, this factor should be smaller for our system, but we use 0.55~0.65 anyway because we do not always have breakpoint.

The process gain of 127 could contain both PN spreading encoding and channel encoding. In [66], it mentioned that power control reduces the effect of coding/interleaving. The combination of power control and coding/interleaving is still the most effective measure to combat correlated fading, but to ease the analysis, we assume that we have a PN spreading gain of 127 (sequence length). So in actual CDMA system, a process gain achieved by both PN spreading and coding/interleaving can give better performance than ours. Even in source encoding, powerful coding/interleaving combination like RS(64,56) block code mentioned in [75] can reduce an input Bit Error Rate (BER) of 10^{-3} to an output BER of 10^{-6} besides that interleaving scatters the correlated errors around to make them independent of each other. This is the reason we leave one eighth overhead in the video data bit rate. Looking at the video quality (PSNR) vs. bit rate plot in Figure 4.2, we can see that in the range of 20 to 50 kbps, we have an average trade-off of 0.117 dB less

PSNR for 1 kbps less bit rate. So if we want to use a more powerful code of RS(64,48), we only need to sacrifice 0.5 dB PSNR. This plot is made based on the actual H.263 codec output for the “Carphone” sequence. Note that due to the reason specified in 2.4.2, i.e. there is no point to buffer more than one frame of coded video and source interleaving is done inside each video frame.

Figure 4.2 PSNR-Bitrate Relationship for “Carphone” Sequence

When considering a slow fading environment with about 2 Hz Doppler shift, we realized that the coherent time has the same magnitude as the video frame duration. While this questions the efficiency of intra-frame interleaving, it also reminds us that slow fading will make fast CDMA power control more efficient. In fact, fast power controlled CDMA can have much less correlated errors than that of slow fading. For perfect power control, we can cancel most of the fading, and the interference can be modelled as AWGN [39] [71], so there is no time correlation. For non-perfect power control, uplink BER correlation comes from correlated power control errors. But

power control signals are transmitted in the natural power balanced downlink which is a better AWGN like channel. So power control errors should be much less correlated than slow fading, so is the uplink BER. And intra-frame source interleaving for video should be good enough.

There is not much research done to characterize the tolerance of coded video signal to bit corruptions and correlated corruptions. We assume source coded video can tolerate 10^{-3} BER and leave the job of achieving this to the source encoder which is not covered in this literature. So similar to that of [14] and [15] we are interested in the BER of 10^{-3} after CDMA channel decoding.

4.2 Variable Bit rate Traffic over CDMA

There are different schemes for CDMA to accommodate variable bit rate:

- a. Multi-level modulation [76] [77]. It has the ability of accommodate different importance bit-stream. Complex power control, also complex handset transceiver structure.
- b. Multi-PG [77]. There is doubt in orthogonality of the multi-PG code and also complex power control.
- c. Multi-code [77~78]. One user can use several PN code upon request. Receivers (mostly base station) are complex.
- d. Slotted CDMA [79~80]. A hybrid of ALOHA and CDMA but with no collisions, or a hybrid of TDMA and CDMA with less coordinate requirement of TDMA. It can be treated as running CDMA inside each TDMA time slot. It doesn't make power control complex and requires not much complexity in transceiver either.

We choose slotted CDMA for the sake of maturity and simplicity. The rule we chose for a user to transmit is simple: randomly choose a number of slots that is needed for current video frame.

With such a rule and large number of users, it's expected that overall bit rate will be statistically evenly-distributed among all time slots. More controlled slot allocation schemes can certainly improve capacity. But they are complex and will erase the most important beauty of CDMA, i.e. each user don't need to know information about others and not much system coordination except PN code allocation is needed.

4.2.1 Probability Model for Slotted Channel

It's not realistic to allocate slots on a bit by bit base. We give a discrete model below to ease the analysis we are going to perform. Given a video frame duration e.g. T_f , which can transmit B bits at maximum, we divide it into M slots. Then the possibility of a user needing m slots is

$$P_U(m) = \int \frac{\frac{mB}{(m-1)!} f(x) dx}{M} \quad (\text{EQ 4.1})$$

here $f(x)$ is the pdf of bits per frame for coded video which is Gamma distributed as we discovered in chapter 2. In practical engineering design, video data will be padded to fit into integer number of slots.

When a user needs m slots to transmit its data, he choose them randomly form the total M slots. In other words, for any slot to be chosen by a user to transmit one of its n data slots, the probability is the same. And it is:

$$P_{Sm} = \frac{C_{M-1}^{m-1}}{C_M^m} = \frac{\frac{(M-1)!}{(m-1)!(M-m)!}}{\frac{M!}{m!(M-m)!}} = \frac{m}{M} \quad (\text{EQ 4.2})$$

The $\frac{m}{M}$ result makes intuitive sense. Further more, the probability for a user to transmit in a slot is:

$$P_S = \sum_{m=1}^M P_{Sm} P_U(m) = \frac{1}{M} \sum_{m=1}^M m P_U(m) \quad (\text{EQ 4.3})$$

which is the discrete mean to peak ratio of the coded video data. So when N approach B , P_s will be the reverse of the continuous PMR which is 0.4 for our video input. A plot is given in the Figure 4.2 below to verify this.

$$\lim_{N \rightarrow B} P_s = \frac{1}{PMR} \quad (EQ 4.4)$$

Figure 4.3 Probability of a User Transmits in a Slot

So, given K multiple access users in a CDMA cell, the probability of one slot has k users transmitting is Binomial distributed as described in EQ 4.5:

$$P_s(K, k) = C_K^k P_s^k (1 - P_s)^{K-k} \quad (EQ 4.5)$$

Here is the conclusion we have for the discrete probability of slotted channel:

- For random multiplexing slotted channel, single slot single user transmitting probability is the Discrete Mean to Peak Ratio (DMPR) of the source. And single slot multiple user transmitting probability is Binomial distributed. This can be applied to any VBR traffic, not just video.

4.2.2 Error Probability of CDMA

Now we need to get the BER expression for CDMA. Given those practical considerations in section 4.1, first, we are going to deduce the expression for flat fading, single cell channel with diversity receiver and power control, then we count multipath and multicell interferences as a multiplicative factor to the variance of the equivalent noise. The AWGN from channel is ignored.

Combine EQ 3.9~3.12 and the above conditions, we can express the decision variable for the 0th information bit is at the diversity receiver as:

$$Y_i = \sum_{a=1}^A Y_i^a = \sum_{a=1}^A \sqrt{P} h_i^a \beta_i d_i^a T + \sum_{a=1}^A \sum_{k=1, k \neq i}^K \sqrt{P} h_k^a \beta_k [d_k^{-1} R_{ki}(\Delta\tau_{ki}^a) + d_k^0 \hat{R}_{ki}(\Delta\tau_{ki}^a)] \cos \Delta\phi_{ki}^a \quad (\text{EQ 4.6})$$

here $\Delta\tau_{ki}^a = \tau_k^a - \tau_i^a$, $\Delta\phi_{ki}^a = \phi_k^a - \phi_i^a$ and the definition of $R_{pq}(\tau)$ and $\hat{R}_{pq}(\tau)$ are the same as in section 3.3.1. Now we consider a same interference variance case in which all phases and delays from different diversity branches are the same for each user, i.e. $\Delta\tau_{ki}^a = \Delta\tau_{ki}$ and $\Delta\phi_{ki}^a = \Delta\phi_{ki}$ for any a. To demonstrate the effect of the power control, we rewrite EQ 4.6 as:

$$Y_i = d_i^0 \sqrt{P} T \left(\beta_i \sum_{a=1}^A h_i^a \right) + \sum_{k=1, k \neq i}^K \sqrt{P} [d_k^{-1} R_{ki}(\Delta\tau_{ki}) + d_k^0 \hat{R}_{ki}(\Delta\tau_{ki})] \left(\beta_k \sum_{a=1}^A h_k^a \right) \cos \Delta\phi_{ki} \quad (\text{EQ 4.7})$$

The reason that this is a equal variance case is as follows: if random variable, say Z, can be expressed as: $Z = \sum_{a=1}^A X_a Y_a$, X_a are i.i.d. distributed with zero mean and variance σ_x^2 and Y_a are i.i.d distributed with a second moment of δ_y^2 , furthermore, X_a and Y_a are independent of each others. Then it's easy to deduce that Z has a zero mean and variance of $A\sigma_x^2\delta_y^2$. If X_a becomes fully correlated, i.e. $X_a = X$ for any a, then Z still remains zero mean, and the variance stays as $A\sigma_x^2\delta_y^2$ too. In our application $X_a = [d_k^{-1} R_{ki}(\Delta\tau_{ki}^a) + d_k^0 \hat{R}_{ki}(\Delta\tau_{ki}^a)] \cos \Delta\phi_{ki}^a$ and $Y_a = h_k^a$.

The aim of power control is to make $\beta_k \sum_{a=1}^A h_k^a$ equal for all k users including the ith user. So assume perfect power control can normalize it to 1 for all users, the equation become much simpler:

$$Y_i = \sqrt{P} T d_i^0 + \sum_{k=1, k \neq i}^K \sqrt{P} [d_k^{-1} R_{ki}(\Delta\tau_{ki}) + d_k^0 \hat{R}_{ki}(\Delta\tau_{ki})] \cos \Delta\phi_{ki} \quad (\text{EQ 4.8})$$

Note that now we are at the ideal no fading situation. To make the model complete, we need to take MultiPath Interference (MI) from both desired user himself and the other users, and Other-cell Interference (OI) into account. From section 4.1, we know that the power of the second received signal ray is about 7~20 dB lower than that of the first ray. To avoid bringing in too much complexity, we will only consider the second ray. As for OI, we knew from section 4.1 that its power is about 55%~65% of Intra-cell Interference (II) which includes MAI and MI. So the received signal can be represented as:

$$RxSignal = DesiredSignal + MAI + MI + CI = DesiredSignal + II + CI \quad (\text{EQ 4.9})$$

Since all interferences can be modelled as AWGN, their variance should be proportional to their power. We already know the variance of MAI, σ_{MAI}^2 . So σ_m^2 , variance of MI, equals to $c_m \sigma_{MAI}^2$, where c_m is in the range of 0.01~0.2 (7~20 dB lower in power). σ_{II}^2 , variance of II, equals to $(1 + c_m) \sigma_{MAI}^2$. And σ_o^2 , variance of OI, equals to $c_o \sigma_{II}^2$ where c_o is in the range of 0.55~0.65 (55%~65%). Given all this, the variance of the total interference, σ^2 , equals to $(1 + c_m) (1 + c_o) \sigma_{MAI}^2$. Substitute σ_{MAI}^2 with EQ 3.19 we get:

$$\sigma^2 = C_m C_o (K - 1) \left[N^2 \frac{23}{360} + N \left(\frac{1}{20} + \frac{K-2}{36} \right) - \frac{1}{20} - \frac{K-2}{36} \right] \quad (\text{EQ 4.10})$$

where $C_m = 1 + c_m$ is the Multipath Interference Coefficient (MIC) which is in the range of 1.01~1.2, and $C_o = 1 + c_o$ is the Other-cell Interference Coefficient (OIC) which is in the range of 1.55~1.65. To get the BER, we just substitute the σ into EQ 3.22.

For imperfect power control the BER becomes a random variable because the SIR becomes Log Normal distributed as in mentioned in section 3.3.2. To incorporate this, EQ 3.22 changes into EQ 4.11 below:

$$P_e = \frac{2}{3} Q\left(\sqrt{\frac{3\rho N}{k-1}}\right) + \frac{1}{6} Q\left(\frac{\rho N}{((K-1)N/3 + \sqrt{3}\sigma)^{0.5}}\right) + \frac{1}{6} Q\left(\frac{\rho N}{((K-1)N/3 - \sqrt{3}\sigma)^{0.5}}\right) \quad (\text{EQ 4.11})$$

where ρ is a Log Normal random variable with mean 1 and variance σ_{ρ}^2 i.e. the variance of the SIR due to power control errors. Now P_e becomes a random variable which makes the system performance evaluation can be done in the same way as simple fading TDMA or FDMA systems, i.e. evaluating outage probabilities.

4.3 System Capacity Analysis.

After all the preparations and derivations, we can easily evaluate the system capacity. We choose the outage probability, as in EQ 4.12 and defined as the probability of a user suffer BER above a pre-determined threshold rate P_{th} , usually 10^{-3} , to be some value P_{out} , which is usually 0.01 in cellular voice communication. Then the system capacity is the maximum number of users that the system can accommodate while providing each user a outage probability smaller than P_{out} .

$$P_{out} = \int_{P_{th}}^1 f(P_e) dP_e = \int_0^{\infty} \left(\sum_{k=k_{th}(\rho)}^K P_S(K, k) \right) f(\rho) d\rho = \int_0^{\infty} \left(\sum_{k=k_{th}(\rho)}^K C_K^k P_S^k (1 - P_S)^{K-k} \right) f(\rho) d\rho \quad (\text{EQ 4.12})$$

The $f(P_e)$ here is the pdf of BER and we define $k_{th}(\rho)$ as the threshold number of users, which is the minimum k in EQ 4.11 satisfying $P_e \geq 10^{-3}$ given certain K, P_S, ρ, C_m, C_o . $f(\rho)$ is the Log Normal pdf of power control errors, i.e. standard deviation. We plot the outage probability vs. number of users to find out how many users we can support at given outage probability, i.e. the system capacity. Note that $P_S(K, k)$ is defined in EQ 4.5.

Since we have chosen slotted CDMA, we can compute outage probability on a per slot base. We need a list of variables that will affect the outage probability. As we can see from previous analysis, it comes down to the following conclusion:

- There are five factors that affect the outage probability: $K, P_s, \sigma_{PC}, C_m, C_o$ which are, by order, number of multiple access users, discrete mean to peak ratio (DMPR), standard deviation of the Log Normal power control error, Multipath Interference Coefficient (MIC) and Other-cell Interference Coefficient (OIC).

The first variable is used to determinate the system capacity, i.e. given some combination of the rest four variables, the maximum number of user that the CDMA system can accommodate without exceeding certain outage probability P_{out} is the capacity. To show how the system capacity is affected by these factors, we analyze them one by one.

4.3.1 System Capacity vs. Discrete Mean to Peak Ratio

First, we assume perfect power control, no multipath interference and no other cell interference. P_s is determined by video pdf and M as defined in section 4.2.1. Bigger M , number of slots per video frame, will result in smaller P_s , according to Figure 4.3 which will affect the outage probability. However, we leave the effect of M to be investigated in the next section and focus just on P_s . We give a plot of single slot random multiplexing gain and efficiency in Figure 4.4 as a function of DMPRs. The gain and efficiency are defined as:

$$g = \frac{U_{actual}}{U_{slot}} \quad (\text{EQ 4.13})$$

$$\eta = \frac{U_{actual}}{U_{ideal}} \quad (\text{EQ 4.14})$$

where U_{actual} is the actual number of users system can accommodate, given outage probability of P_{out} , and U_{slot} is the users that a single slot can accommodate with BER lower than 10^{-3} . U_{ideal} is

defined as U_{slot} divided by DMPR, i.e. the ideal capacity where slots are assigned deterministically in order to strictly average total user traffic over all slots. We chose the outage probability to be 0.01 in Figure 4.3. It's obvious that the efficiency increase with bigger mean to peak ratio meaning that constant bit rate is easy to multiplex. Here is one more conclusion:

- The single slot random multiplexing gain increases as DMPR decreases while efficiency decreases as DMPR decreases. The decrease of efficiency is the price we need to pay for random multiplexing.

Figure 4.4 Slot Multiplexing Efficiency and Outage Probability vs. DMPR

4.3.2 System Capacity vs. Power Control Error

Now we are counting power control errors. Since we mentioned in last section that M , number of slots is also affecting the outage, outage probabilities determined by power control errors ranging from 0 dB to 3 dB for different number of slots per video frame are plotted with $C_m = 1.55$ and $C_o = 1.1$ in Figure 4.5.

As a example of how to read system capacity from these figures, in Figure 4.5 (d), we first find the cross of 0.01 outage probability line and the outage curve of 1.5 dB σ_{PC} . The Number of users coordinate of the cross is 20. Thus for 1.5 dB power control error, using a 32 slot allocation scheme, we can support up to 20 users without exceeding 0.01 outage probability.

(a)

(b)

Figure 4.5 Outage Probability vs. Power Control Error for Different Slot Numbers

From all these plots in Figure 4.5, we can tell that capacity increases with bigger slot number. But bigger slots number means smaller slot length which is difficult to manage in practical systems. Also from Table 3.1, capacity given $0.01 P_{out}$ for different slots number, it's obvious that the capacity stop increasing for slot number bigger than 32. Since 32 also give us a reasonable slot length, we choose 32 as the slot number for later analysis.

Table 4.1: Number of Users Supported for $P_{out} < 0.01$

σ_{PC} in dB	0	0.5	1.0	1.5	2.0	2.5	3.0
$M = 1$	22	17	12	9	7	5	4
$M = 4$	30	27	21	16	12	9	6
$M = 16$	36	32	25	19	14	11	8
$M = 32$	37	33	26	20	15	11	8
$M = 64$	37	33	27	20	15	11	8
$M = 128$	37	33	27	20	15	11	8

The results for different P_{out} are listed in Table 3.2 for 32 slots case. From observation of Figure 4.5 and these two tables, we reach the following conclusions:

Table 4.2: Number of Users Supported for $M = 32$

σ_{PC} in dB	0	0.5	1.0	1.5	2.0	2.5	3.0
$P_{out} = 0.01$	37	33	27	20	15	11	8
$P_{out} = 0.005$	35	31	24	17	12	8	6
$P_{out} = 0.001$	32	28	21	14	9	6	4

- Multi-slot multiplexing efficiency increases with slot number but will stop increasing above some slot number.

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- Power control error affect system capacity greatly. e.g. for 1.5 dB error, capacity is 50-55% of perfect power control.
 - Given smaller outage probability, the capacity decreases relatively slowly. e.g. for a 10 dB difference, capacity of $0.001 P_{out}$ is 70% of $0.01 P_{out}$.
 - CDMA can provide better capacity when handling VBR compared to FDMA, or even to TDMA without any coordination of video activity factor or so called DMPR here. e.g. for $0.01 P_{out}$ and a normal 1.5 dB power control error, 32 slot allocation offer a single cell capacity of 20 users. Multiply by 7, which is commonly used frequency reuse factor in FDMA, we get a 7 cell capacity of 140 while FDMA can only offer 128 user capacity given the same bandwidth. TDMA also has a capacity of 128 if it can't make use of the video activity factor. As a matter of fact, it's much more difficult for TDMA to make use of it. And if we count the fact that antenna sectorization can increases CDMA capacity directly and it only increases TDMA and FDMA capacity through changing frequency reuse factor, CDMA system capacity can be further increased.

4.3.3 System Capacity vs. MIC and OIC

In last section, we analyzed system capacity for fixed MIC and OIC, i.e. $C_m = 1.55$ and $C_o = 1.1$. The following figure tell us how capacity will change with respect to different MICs and OICs.

We list the capacity for different P_{out} in Table 4.3. From EQ 4.9, we can see that MIC and OIC affect system capacity in exactly the same way. So we list the results in a increasing order of $C_m \times C_o$. Note the biggest value 2.2 can be treated as a $C_m \times C_o$ of 1.42×1.55 , or 1.1×2.0 , or 1.2×1.85 . We can see that a increase from 1.566 to 2.2, i.e. $MIC \times OIC$ is doubled, capacity is reduced by 32-38%. Certainly we can treat other value of $C_m \times C_o$ as different multiplication combinations of C_m and C_o .

(a) $MIC = 1.1, OIC = 1.6$

(b) $MIC = 1.1, OIC = 1.65$

(c) $MIC = 1.01, OIC = 1.6$

(d) $MIC = 1.2, OIC = 1.55$

Figure 4.6 Outage Probability vs. MIC and OIC

Table 4.3: Number of Users Supported for $M = 32$ and $\sigma_{PC} = 1.5$ dB

$C_m \times C_o$	1.01×1.55 (1.566)	1.1×1.55 (1.65)	1.1×1.6 (1.76)	1.1×1.65 (1.815)	1.2×1.55 (1.86)	2.2
$P_{out} = 0.01$	22	20	19	19	17	15
$P_{out} = 0.005$	20	17	17	17	16	13
$P_{out} = 0.001$	16	14	13	12	12	10

Here are the final conclusions we reach in this literature:

- MIC and OIC decrease system capacity, but not as big as power control errors. e.g. double powered $MIC \times OIC$ only decreases capacity by 32-38%.
- To increase CDMA system capacity, power control accuracy is the key factor.

4.4 CDMA System Capacity Analysis Summary

In this chapter we first discussed some practical issues like choosing system parameter, determining channel characteristics, then we present a random VBR video multiplexing scheme and improved Gaussian approximation model for BER evaluation, using the scheme and the model, we evaluate CDMA system maximum user capacity for all kinds of factors that affecting outage probability.

The capacity analysis was made for a CDMA videophone system with no cell sectorization, no RAKE receiver, equal gain diversity combining only, random multiplexing only and PN spreading code only. i.e. we have analyzed a basic model which is proper for theory analysis. In practical systems, more complex receiving, multiplexing and code spreading schemes can be employed to increase system capacity.

CHAPTER 5

Conclusions and Expectations

In this chapter, we first give a summary of the conclusions we have drawn from this work. Then suggestions about future work will be proposed. Please notice that the conclusions are described in details in section 2.3 and in chapter 4.

5.1 Conclusions

- CBR has overall significant worse performance in average PSNR and worst PSNR given the same average bit rate. It's almost equivalent to suffering some corruption from the CBR coding because bits are wasted on static scenes and there are not enough for the active scenes. The performance of CBR can be worse when subjected to further channel errors during transmissions.
- VBR, on the other hand, distributes the bits reasonably and results in a better overall PSNR performance with less visual distortion.
- Transmission errors propagates in the video scenes, but will recover slowly to some extent when the update information coming from subsequent frames.

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- Both SDMR and PMR depend on both scene contents and compression algorithms. Active scenes result in higher peak to mean ratio, at least for transform coding. DPCM, DPCM plus Motion Compensation (MC), and DCT plus MC results in a descending peak to mean ratio according to the order of the algorithm.
 - The pdf of bits or packets per frame is gamma distributed, not Gaussian distributed, at least for videophone/conference type video. This conclusion is not confined to H.263 coding. not even to DCT + MC coding. It holds for both Prediction coding, Transform coding. Results for other type of coding are unknown, but expected to be similar because they reflect the nature of entropy variation of video sequences.
 - The autocovariance depends mostly on scene contents, independent of coding frame rate and output bit rate. While some of the scenes have repeated peak of autocovariance, the reason is more likely due to short sequence and computation truncations. Exponential autocovariance is a proper model.
 - Sum of N independent Gamma random variables with the same parameter λ and different parameter S_n ($n=1,2,\dots,N$) is still a Gamma random variable with parameter λ and $S = \sum_{n=1}^N S_n$.
 - A Gamma random variable with integer parameter S or large S can be modelled as a summation of $[S]$ i.i.d exponential random variables. And the exponential random variables can be easily generated by taking the logarithm of a uniform random variable.
 - For video coded with small peak to mean ratio, AR(1) model has good independent residue performance. It's a very good and simple candidate among all the models.
 - For random multiplexing slotted channel, single slot single user transmitting probability is the Discrete Mean to Peak Ratio (DMPR) of the source. And single slot multiple user transmitting probability is Binomial distributed. This can applied to any VBR traffic, not just video.
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- There are five factors that affect outage probability: $K, P_s, \sigma_{PC}, C_m, C_o$ which are, by order, number of multiple access users, discrete mean to peak ratio (DMPR), standard deviation of the Log Normal power control error, Multipath Interference Coefficient (MIC) and Other-cell Interference Coefficient (OIC).
 - The single slot random multiplexing gain increases as DMPR decreases while efficiency decreases as DMPR decreases. The decrease of efficiency is the price we need to pay for random multiplexing.
 - Multi-slot multiplexing efficiency increases with slot number but will stop increasing above some slot number.
 - Power control error affect system capacity greatly. e.g. for 1.5 dB error, capacity is 50-55% of perfect power control.
 - Given smaller outage probability, capacity decrease relative slowly. e.g. for a 10 dB difference, capacity of 0.001 is 70% of 0.01.
 - CDMA can provide better capacity when handling VBR compared to FDMA, or even to TDMA without any coordination of video activity factor or so called DMPR here. e.g. under normal conditions, we get a 7 cell capacity of 140 while FDMA can only offer 128 user capacity given the same bandwidth. TDMA also has a capacity of 128 if it can't make use of the video activity factor. As a matter of fact, it's much more difficult for TDMA to make use of it. If we count the fact that antenna sectorization can increase CDMA capacity directly and it only increases TDMA and FDMA capacity through changing frequency reuse factor, CDMA system capacity can be further increased.
 - MIC and OIC decrease system capacity, but not as big as power control errors. e.g. double powered only decreases capacity by 32-38%.
 - To increase CDMA system capacity, power control accuracy is the key factor.
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5.2 Future Expectations

The world of telecommunication is evolving in a fascinating pace in all areas. One human expectation of wireless communication is to be able to use portable multimedia phone in the coming future. We have presented a theoretical capacity analysis of CDMA videophone system. Further work can be done in the near future to make one step further toward our dream. These are as follows:

- Design efficient joint video source and channel coding schemes to enable protect coded video from suffering the high-BER condition of wireless mobile channel.
- Design efficient CDMA fast power control schemes to increase capacity dramatically.
- Combine voice, video and data together, design a real wireless multimedia communication system.

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IMAGE EVALUATION TEST TARGET (QA-3)

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