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MULTIUSER SCHEDULING USING EMBEDDED MODULATION IN
WIRELESS NETWORKS

by

Shermeen Nizami

B.Sc. University of Engineering and Technology, Lahore, Pakistan, 1996

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MULTIUSER SCHEDULING USING EMBEDDED MODULATION IN
WIRELESS NETWORKS

submitted by

Shermeen Nizami

in partial fulfillment of the requirements for the degree of

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Abstract

This thesis proposes a new adaptive use of embedded modulation (EM) for multiuser cellular wireless networks with feedback channels. In the proposed system multiple users with different carrier-to-interference ratios are allowed to share a single downlink channel simultaneously. Information bits meant for two different users are combined in a single adaptive coded modulation signal constellation using multi-level modulation and coding. To exploit the benefits of EM, two novel multiuser scheduling algorithms are also proposed that provide for high system throughput while maintaining an acceptable level of fairness amongst all the system users. The first algorithm is based on round-robin scheduling such that all users are allocated equal time slots. Computer simulation performance evaluation results show that throughput gains of up to 33% over traditional round-robin techniques are possible. The second algorithm is based on the principle of minimal fairness, and has been shown to deliver up to a 60% improvement in system performance.

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List of Symbols

Symbol	Description
$\alpha(t)$	time-varying channel attenuation
$\hat{\alpha}(t)$	estimated channel attenuation
A_{max}	maximum attenuation at BTS antenna
A^2	line-of-sight signal power
β	fairness control factor
β_{min}	minimum allowable value of fairness control factor
χ	maximum C/I limiting parameter
$\mathcal{C}_i$	signal constellation
C	intermediary constant assignment
$(C/I)_{effective}$	effective C/I at the receiver
$(C/I)_{inst}$	instantaneous C/I at the receiver antenna
$(C/I)_{max}$	maximum achievable C/I at the receiver
$\delta_{n,k}$	the mobile bearing, it is the angle between a line directed due East from BTS # n and the line connecting BTS # n and mobile k in the horizontal plane
Δ	min. distance for M -QAM signal constellation
Δ_1	min. distance for M_1 -QAM signal constellation
Δ_2	min. distance for M_2 -QAM signal constellation

d	transmitter-receiver separation
d_0	close-in reference distance
d_1	min. distance for embedded M_1 -QAM signal constellation
d_2	min. distance for embedded M_2 -QAM signal constellation
ϵ_i	antenna bearing for the i th sector in each cell
η	effective ACM throughput
$\bar{\eta}$	average channel throughput
η_n	average channel throughput for user n
$\bar{\eta}_n$	normalized throughput for user n
$\eta_{max-C/I}$	average channel throughput of the Max C/I scheduling scheme
η_{RR}	average channel throughput of the RR scheduling scheme
E_s	average symbol energy
$E_{s(c_i)}$	average symbol energy of constellation C_i
$E_{s(EM)}$	average symbol energy of an EM constellation
f_D	Doppler frequency of the fading channel
F	Ricean fading component in the channel
F_n	number of slots assigned to user n
γ	user C/I
γ_i	C/I of user i
γ'_i	C/I of user i after embedding
G_r	mobile unit receiving antenna gain
G_t	BTS antenna gain

$I_{n,i}$	interference signal power from interfering sector #i in cell #n
κ	Boltzmann's constant
λ_i	percentage of symbol energy to embedded user i
m	ACM mode
M_i	number of signal points in the constellation for user i
$n(t)$	AWGN process
N	number of concurrent system users
N_i	total number of interfering sectors
N_0	single-sided thermal noise power spectral density
NF	noise figure
NT	total channel noise power
$\mathcal{N}_T$	thermal noise
$\overline{PL(d)}$	average large-scale path loss at a distance d metres
$Pilot\ C/I$	received pilot signal at the mobile unit
$PL(d)$	path loss and shadowing model
P_r	received signal power at the mobile unit antenna
P_t	transmitted signal power at the BTS antenna
SH_σ	lognormal random variable with zero mean and standard deviation σ
Q	Ricean fading line-of-sight component
ρ	propagation exponent
$r(t)$	received signal in time domain
R_F	Ricean random variable

R_n	transmission rate for user n
R_y	Rayleigh random variable
σ	standard deviation of log-normal shadowing
σ_F	standard deviation of multipath fading components
$s(t)$	transmitted signal in time domain
S_i	number of signal points in an EM constellation
θ	horizontal transmit angle at BTS antenna
θ_{3dB}	3 dB beamwidth at BTS antenna
$\theta_{n(i,k)}$	the angle between the i th sector antenna of the n th cell and the line connecting BTS $\#n$ and mobile k in the horizontal plane
τ	propagation delay
τ_f	feedback delay
τ_t	adaptation delay
T_o	thermal noise temperature
ω_M	energy coefficient for M-QAM signal constellation
Ω_n	minimal fairness criteria for user n
W	transmission bandwidth
X_1, X_2	uniform random variables
Y_n	throughput embedded for user N with user n
ζ	nominal ACM throughput
Z_1, Z_2	gaussian random variables

List of Acronyms

1xEV-DV	1xEVolution-Data/Voice
3G	3rd Generation
3GPP2	3rd Generation Partnership Project 2
ACM	Adaptive Coding and Modulation
ARQ	Automatic Repeat Request
AWGN	Additive White Gaussian Noise
BER	Bit Error Rate
BICM	Bit-Interleaved Coded Modulation
BPCHU	Bits Per Channel Use
BPSK	Binary Phase Shift Keying
BTS	Base Transceiver Station
CDF	Cumulative Density Function
CDMA	Code Division Multiple Access
C/I	Carrier-to-Interference Ratio
CSI	Channel State Information
DAB	Digital Audio Broadcast
DVB	Digital Video Broadcast
EM	Embedded Modulation

GHz	Giga Hertz
HIPERLAN	High Performance Radio Local Area Network
LOS	Line-Of-Sight
Max-C/I	Maximum C/I
MF	Minimally Fair algorithm
MF-ACM	Minimally Fair ACM algorithm
MF-EM	Minimally Fair EM algorithm
MHz	Mega Hertz
MS	Mobile Station
PDF	Probability Density Function
PER	Packet Error Rate
RR	Round Robin
QAM	Quadrature Amplitude Modulation
T-R	Transmitter-receiver
TDMA	Time Division Multiple Access

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Chapter 1

Introduction

Wireless networks, such as those based on the evolving 1xEVolution-Data/Voice (1xEV-DV) standard for third generation cellular systems [1], require communication over a hostile and time-varying radio channel. Wireless communication channels are characterized by several random time-varying attributes, including path loss depending on the transmitter-receiver separation, shadowing caused by partial blockage of the transmission path due to the surrounding clutter and fading due to multipath propagation. Moreover, signal interference is introduced by other users and base transceiver stations (BTS) on the same network. As a result, the link carrier-to-interference ratio (C/I) would be different for each user, it would also be time-variant.

Furthermore, since the available wireless radio spectrum is limited, wireless systems are constrained to operate in a fixed bandwidth to support an ever-increasing number of users. Spectrally efficient modulation techniques are therefore a vital component in the development of wireless networks. The anticipation for a communication system that provides high-speed access to information has led to widespread research on information transmission over fading channels. One possible method to achieve robust and spectrally efficient communication is to adapt the transmission

scheme to the instantaneous channel characteristics using current channel state estimates. These channel state estimates are made available through feedback links between the receiver and transmitter units in adaptive communication systems. Link adaptation techniques compensate for the time-varying nature of the wireless channel and provide the best possible operating modes under constantly changing channel conditions.

Adaptive coded modulation (ACM) has been widely applied in third generation wireless systems as a powerful tool for improving spectral efficiency. Such a scheme adjusts the constellation size and code rate of the transmitted symbols according to channel variations. ACM achieves improved spectral efficiency by transmitting at high speeds under favourable channel conditions as well as appropriately responding to channel degradation through a smooth reduction of the data throughput.

For the purpose of achieving high spectral efficiency, it is the function of a scheduler to allocate BTS resources in such a fashion that each of the users receive a fair share of system resources. The two main objectives of a scheduler are to maximize the average system throughput while maintaining system fairness amongst all concurrent system users. These two objectives may not necessarily compliment each other, and in fact there is always a tradeoff involved between the two. To understand this better, we can consider, for example, two existing scheduling schemes namely, the Maximum C/I (max-C/I) and the Round Robin (RR) scheduling schemes. In case of the max-C/I scheduler, all system resources are allocated to the user experiencing the best link C/I. Although this makes the system throughput exceptionally high

with max-C/I scheduling producing an upper bound on system throughput, but this occurs with the total loss of system fairness. The RR scheduler, on the other hand, would allocate equal system resources to all users. This results in a highly fair system but at the cost of lowered system throughput.

1.1 Thesis Objectives

The aim of this thesis is to design a novel scheduling scheme which improves the average throughput of the system over that of the Round Robin scheme, while still maintaining acceptable fairness. An attempt is made to achieve this goal by applying an evolving link improvement technique called embedded modulation (EM).

Traditionally, an embedded modulation symbol is made up of sublevels of different coding and modulation schemes. EM is promoted for digital audio and video broadcasting (DAB and DVB), where the signals are coded in such a way that they can be received with varying levels of resolution. Layered video/image compression techniques typically encode “high level” and “low level” features into different segments of the bit-stream, producing information with varying levels of importance. Low signal strength receivers can decode only a low audio/video bandwidth containing the most important part of the data while high signal strength receivers can get a proportionally higher decodable bandwidth containing both the “high level” and “low level” features of the data, all of these derived from the same received signal.

This traditional use of EM also leads into one of its derived uses for cellular wireless networks employing feedback links. We have the familiar problem of huge

C/I variations amongst users located differently in the cellular network. If a mobile user cannot indicate to the BTS what the correct link C/I would be for the next packet download, EM could be employed. In such a situation, the graceful degradation of the coding and modulation levels within a multi-level embedded symbol can be used to allow robust “best effort” communications. In this application, the embedded symbol transmission would be destined for a single user on each link.

This thesis proposes a new adaptive use of EM for multiuser communication in time division multiple access (TDMA) cellular wireless networks with feedback channels. In the proposed application multiple users with widely different C/I’s are allowed to share a single downlink channel simultaneously. In this kind of modulation, the information bits meant for two different users are combined in a single ACM signal constellation using multi-level modulation and coding. This application is new and differs from the prior DAB/DVB broadcast applications in that each mobile user would receive information different from the other users whose information has been embedded in the same transmission symbol. Hence, one of the users would retain some of the bits received during a single channel use and discard the others, while the other user would only retain the bits discarded by the first user. The use of EM is adaptive in this context, because depending on the embedded users’ C/I’s not only the the interpoint symbol separation but also the code rate within the constructed EM constellation would be different for different pairs of users.

Having proposed the use of adaptive EM in this thesis, the next objective is to design a scheduling algorithm which enables fair system throughput maximization

by means of EM. For this purpose, two scheduling algorithms have been proposed in this thesis.

1.2 Summary of Thesis Contributions

The contributions made by this thesis can be summarized as follows:

- A new application of EM on wireless network multiuser links has been proposed for TDMA systems.
- This application simultaneously serves users having different C/I's on the same channel.
- Different information is delivered to each user by means of the same transmitted signal.
- In this thesis it is shown how to construct an EM constellation and how it can be adapted according to the users' C/I's under fixed power constraints.
- Two EM scheduling algorithms have been designed, namely the RR-EM and MF-EM.
- The performance of both these algorithms has been evaluated by means of computer simulations, with respect to system throughput and fairness.

1.3 Thesis Organization

The thesis consists of a total of five chapters. This introductory chapter is followed by Chapter 2, in which the two link adaptation techniques, namely adaptive coded modulation and embedded modulation, are discussed in full detail. This chapter also describes how to construct embedded constellations. Chapter 3 explains the objectives in scheduling users for data communication. Examples of existing scheduling schemes such as the Maximum C/I and the Round Robin algorithms are cited for setting system performance bounds. A fairness criteria is defined for assessing the different scheduling schemes. The second half of Chapter 3 presents the thesis proposal in the form of two EM scheduling schemes, namely Round Robin EM and Minimal Fairness EM. The system simulation results follow in Chapter 4. Chapter 5 summarizes our research and concludes the thesis with recommendations for future work.

Following the last chapter are three appendices. Appendix A contains the complete mathematical derivation of the symbol energy for EM constellations. Appendix B begins by describing the time-varying characteristics of the wireless channel as a medium of transmission, followed by the path loss, shadowing, fading, AWGN and interference model used in this research. The cell layout, user C/I selection, adaptive communication system model and sample system parameters are all included in this appendix. Appendix C carries information on how to generate random variables for the system model, such as shadowing, fading and thermal noise.

Chapter 2

Link Adaptation Techniques

High-speed wireless data transmission networks require robust and spectrally efficient communication techniques to counter the effects of time-varying channels. Severe multipath fading, shadowing due to surrounding objects, path loss and co-channel interference from neighbouring cells can lead to serious degradation in the link carrier-to-interference ratio (C/I), consequently increasing the bit error rate (BER). Transmission techniques which do not adapt to the fading channel require increased link budget margins to maintain acceptable performance during deep fades. However such systems are effectively designed for worst-case channel conditions, resulting in inefficient utilization of the channel. Adapting certain parameters of the transmitted signal to the fading channel leads to more efficient channel use [2, 3, 4, 5].

The basic concept of adaptive transmission is real-time balancing of the link budget through adaptive variation of the symbol transmission rate (variable-rate transmission), constellation size, error control coding rate/scheme, or any combination of these parameters. These schemes provide a higher average link throughput by compensating for the time-varying nature of the wireless channel i.e., transmitting at high speeds under favourable channel conditions and responding to channel degradation through a smooth reduction of their data throughput.

Section 2.1 details the methodology employed for ACM. The rest of the chapter focuses on *embedded modulation* (EM), introducing the theory of EM in detail, describing how to construct an embedded constellation, and giving an algorithm to create the optimal EM constellation.

2.1 Adaptive Coded Modulation

Variable rate adaptive channel coding and modulation has proven to perform with significant gains in channel throughput relative to fixed-rate coding and modulation. Many researchers have contributed to the development of adaptive coded modulation (ACM) [6, 7, and the references therein]. In ACM, under fixed power constraints, the code rate and modulation level are varied according to the instantaneous channel condition. The receiver estimates the current channel C/I for the link and conveys this information to the transmitter through a low capacity feedback link. When the channel state is good, the transmitter employs a higher level modulation and code rate, increasing the instantaneous throughput. And as the receiver enters a fade the constellation size and/or the code rate is decreased to a value which provides an acceptable system error rate at a fixed average transmit power. This results in a variable throughput, expressed as the number of information bits transmitted per channel use (bpchu). It is the average value of this throughput which has been chosen as the metric for evaluating system performance in this thesis.

For ease of implementation, the transmitter is limited to selecting the code rate and modulation scheme from a finite set of ACM operating modes, for example those

listed in Table 2.1. These ACM modes are taken from the HIPERLAN/2 standard, which is a high speed wireless local area network standard that has been developed by the European Telecommunications Standard Institute [8, 9, 10]. A key feature of the physical layer of HIPERLAN/2 is that it provides several ACM modes, which can be dynamically selected during link adaptation. The variable code rates cited here are achieved by puncturing a 64-state convolutional code with generator $(133, 171)_8$. Bit interleaved coded modulation (BICM) is used for transmission [11]. Although each ACM mode has its own nominal throughput ζ , as shown in Table 2.1, due to packet errors and consequent retransmissions the link actually operates at an effective throughput η that is less than the nominal one. For a given ACM mode, m , operating at a C/I γ , η is given by¹ the expression:

$$\eta(\gamma, m) = \zeta(m)[1 - PER(\gamma, m)] \quad (2.1)$$

where $PER(\gamma, m)$ is the packet error rate when mode m is used at a user C/I of γ . Plotting the packet error rate for different ACM schemes versus the C/I (dB) required to achieve it generates a family of curves called *error probability performance curves*. Fig. 2.1 shows the error probability performance curves corresponding to the HIPERLAN/2 ACM modes of Table 2.1, assuming a packet size of 192 bits. Based on these curves the effective throughput can be calculated according to (2.1), and plotted as shown in Fig. 2.2. For a given C/I, the transmitter selects the ACM mode that gives the highest effective throughput. Based on Fig. 2.2, Table 2.2 shows the ACM mode that would be selected for a given C/I.

¹ assuming selective-repeat ARQ with unlimited transmit and receive buffers, and reliable acknowledgement transmissions.

Mode m	Modulation Scheme	Modulation Level (M)	Code Rate	Nominal Throughput ζ (bpchu)
1	BPSK	2	1/2	$\frac{1}{2}$
2	QPSK	4	1/2	1
3	QPSK	4	3/4	$1\frac{1}{2}$
4	16-QAM	16	9/16	$2\frac{1}{4}$
5	16-QAM	16	3/4	3
6	64-QAM	64	3/4	$4\frac{1}{2}$

Table 2.1: ACM modes and their corresponding nominal throughputs.

Figure 2.1: Packet error rate of ACM modes as a function of the C/I [12].

Figure 2.2: Effective throughput of ACM modes at different C/I's.

C/I Range (dB)	Mode m	Modulation Scheme	Code Rate
< 1.0	1	BPSK	1/2
1.0 - 4.0	2	QPSK	1/2
4.0 - 7.7	3	QPSK	3/4
7.7 - 10.3	4	16-QAM	9/16
10.3 - 15.3	5	16-QAM	3/4
> 15.3	6	64-QAM	3/4

Table 2.2: User C/I's and their corresponding maximum throughput modes.

2.2 Adaptive Embedded Modulation

Having talked about the advantages of using ACM in single-user channels, we now move on to present one of the proposals behind this thesis. The thesis proposes a link improvement technique which can be viewed as a new application of ACM. This technique employs *embedded modulation* (EM) [13, 14]. EM is typically used to simultaneously transmit bits with differing levels of reliability to multiple users having different C/I's. Traditionally, EM is promoted for digital audio and video broadcasting (DAB and DVB), without any feedback, where the signals are coded in such a way that they can be received with varying levels of resolution. An EM symbol is made up of sublevels of different coding and modulation schemes. Layered video/image compression techniques typically encode “high level” and “low level” features into different segments of the bit-stream, producing information with varying levels of importance. Low signal strength receivers can decode only a low audio/video bandwidth containing the most important part of the data while high signal strength receivers can get a proportionally higher decodable bandwidth containing both the “high level” and “low level” features of the data, all of these derived from the same received signal.

This thesis proposes a new adaptive use of EM for multiuser communication in cellular wireless networks with feedback channels. In the proposed application, multiple users with widely different C/I's are allowed to share a single downlink channel simultaneously, wherein each user experiences different channel conditions such as path loss, shadowing and interference. The power transmitted to a single

user is “rolled back” so that an additional user can be served with this conserved energy in the same channel transmission. Although it is theoretically possible to embed more than two users at the same time, the results of this thesis are limited to the case where only two users are embedded in one ACM signal constellation.

This application is new and differs from the existing DAB/DVB broadcast applications in that each mobile user would receive information different from the other user whose information has been embedded in the same transmission symbol. Hence, the information bits embedded for one user would only be useful for that particular user, and the other embedded user would discard those bits and retain only the next couple of bits that were transmitted for its use only.

2.2.1 Theory

By means of embedded modulation, information meant for two different users is superimposed in the same channel symbol. The signal points of an EM constellation are grouped into “clouds”, with each containing an equal number of points. The clouds represent the constellation meant for the user which requires a higher minimum distance (d_1) between its signal points (i.e., the user with the lower C/I) and the points embedded within a cloud represent the constellation meant for the user with the lower required minimum distance (d_2) (i.e., the user experiencing better channel conditions). The EM constellation is created under the condition that $d_1 > d_2$, [13, 14]. To demonstrate this, a 4×4 EM constellation is shown in Fig. 2.3. The 16 points of the embedded constellation are grouped into four “clouds”, each containing four signal points. Fig. 2.3 also shows the bit-to-signal mapping, which is done in

Figure 2.3: 4×4 Embedded modulation constellation, with signal point labels.

such a fashion that the first two bits specify the cloud, and the last two bits specify the point within the cloud. Observe that a user with a low C/I may only be able to determine the cloud containing the transmitted point, but not the actual point within the cloud. Hence it could reliably detect the first two bits, but not the last two. A user with a high C/I, on the other hand, could determine all four bits, although it would only be interested in the last two. The use of EM in this case therefore involves pairing two users, one with a high C/I and one with a low C/I. A control channel on the forward link would inform each user beforehand of its prospective bit positions within the EM symbol.

This general idea is easily extended to different EM constellations, such as 2×4 EM, where only one bit is sent to the low C/I user and two bits are sent to the high

Figure 2.4: 4×16 Embedded modulation constellation.

C/I user. Similarly a 4×16 EM constellation results when a 4-QAM constellation for one user is embedded with a 16-QAM constellation for the second user. This 4×16 EM constellation is shown for example in Fig. 2.4. Different coding rates can also be applied to the data for the different users. It is also important to note that the inter- and intra-cloud separation (d_1 and d_2 respectively) are constrained to ensure fixed average transmit power. In Appendix A it is shown that the average signal energy of an $M_1 \times M_2$ EM constellation with separations d_1 and d_2 is given by

$$E_{s(EM)} = \omega_{M_1} d_1^2 + 2\omega_{M_1} \left(\sqrt{M_2} - 1 \right) d_1 d_2 + \left[\omega_{M_1} \left(\sqrt{M_2} - 1 \right)^2 + \omega_{M_2} \right] d_2^2 \quad (2.2)$$

where

$$\omega_M = \begin{cases} \frac{1}{4} & \text{for } M = 2 \text{ (BPSK)} \\ \frac{1}{6} (M - 1) & \text{for } M = \text{even power of 2 (square } M\text{-QAM)} \end{cases}. \quad (2.3)$$

For a fixed average transmit energy of $E_{s(EM)} = E_s$ and a given M_1 and M_2 , there is a one-to-one relationship between d_1 and d_2 . That is, specifying d_1 gives d_2 and vice versa.

2.2.2 EM Constellation Construction

In systems employing ACM, the code rate and number of modulation levels are adapted to the channel C/I to maximize the effective throughput. In the proposed EM system, not only are the code rates and number of modulation levels for both users adapted, but so are the inter- and intra-cloud separations. This can lead to higher throughput and better control of fairness.

To optimize the EM constellation we consider the case where two users, with C/I's of γ_1 and γ_2 , are to be embedded, where $\gamma_1 < \gamma_2$. For the time being, let us suppose that the embedded ACM operating modes for the two users have been selected, and with each EM channel use, that $\log_2 M_1$ and $\log_2 M_2$ code bits are to be transmitted to user 1 and user 2 respectively. We want to determine the separations d_1 and d_2 that maximize the effective throughput. Repeating this procedure for all possible embedded ACM modes for both users will then determine the optimal EM constellation.

The effective throughput of an EM constellation depends not only on the ACM modes for each user, but also on the link packet error rate. Determining the ex-

act PER's from an EM constellation is not possible, so instead we make use of an assumption for simplification. It is assumed that, in terms of PER, transmitting a $M_1 \times M_2$ EM constellation with separations d_1 and d_2 is equivalent to separately sending an M_1 -QAM constellation with separation d_1 to user 1, and M_2 -QAM constellation with separation d_2 to user 2. This is actually a pessimistic assumption because when EM is used the PER's will be slightly lower than for the equivalent non-embedded transmissions with the same d_1 and d_2 .

Since the average transmitted energy of a square M -QAM² constellation with separation Δ is

$$E_s = \omega_M \Delta^2 \quad (2.4)$$

with ω_M as defined in (2.3), the use of EM is equivalent to transmitting an M_1 -QAM constellation to user 1 with effective energy

$$E'_{s,1} = \omega_{M_1} d_1^2 \quad (2.5)$$

and transmitting an M_2 -QAM constellation to user 2 with effective energy of

$$E'_{s,2} = \omega_{M_2} d_2^2 . \quad (2.6)$$

Let λ_1 be the fraction of the total transmitted energy (E_s) that is assigned to the M_1 -QAM constellation, and λ_2 be the fraction of energy assigned to the M_2 -QAM constellation, so

$$\lambda_1 = \frac{E'_{s,1}}{E_s} = \frac{\omega_{M_1} d_1^2}{E_s} \quad (2.7)$$

$$\lambda_2 = \frac{E'_{s,2}}{E_s} = \frac{\omega_{M_2} d_2^2}{E_s} . \quad (2.8)$$

² For BPSK, $M = 2$ and $\omega_M = \frac{1}{4}$

The effective C/I of user 1 after embedding is now given by γ'_1 and that of user 2 is given by γ'_2 , such that

$$\gamma'_1 = \lambda_1 \gamma_1 \quad (2.9)$$

$$\gamma'_2 = \lambda_2 \gamma_2 . \quad (2.10)$$

The effective throughput to user 1 can be found by evaluating the appropriate curve from Fig. 2.2 for the ACM mode m_1 at a C/I of γ'_1 . Similarly the effective embedded throughput to user 2 is evaluated, and adding the two gives the combined throughput delivered by the EM constellation.

2.2.3 Throughput Maximization

Maximizing the combined throughput of the EM constellation involves finding the best choices for d_1 and d_2 , or alternatively, λ_1 and λ_2 , since it follows from (2.7) and (2.8) that

$$d_1 = \sqrt{\frac{E_s \lambda_1}{\omega_{M_1}}} \quad (2.11)$$

$$d_2 = \sqrt{\frac{E_s \lambda_2}{\omega_{M_2}}} . \quad (2.12)$$

Substituting for d_1 and d_2 in (2.2) gives

$$\begin{aligned} E_{s(EM)} &= \omega_{M_1} \frac{E_s \lambda_1}{\omega_{M_1}} + 2\omega_{M_1} \left(\sqrt{M_2} - 1 \right) \sqrt{\frac{E_s \lambda_1}{\omega_{M_1}}} \sqrt{\frac{E_s \lambda_2}{\omega_{M_2}}} \\ &\quad + \left[\omega_{M_1} \left(\sqrt{M_2} - 1 \right)^2 + \omega_{M_2} \right] \frac{E_s \lambda_2}{\omega_{M_2}} \\ &= E_s \lambda_1 + 2E_s \sqrt{\frac{\omega_{M_1}}{\omega_{M_2}}} \left(\sqrt{M_2} - 1 \right) \sqrt{\lambda_1 \lambda_2} \\ &\quad + E_s \left[\frac{\omega_{M_1}}{\omega_{M_2}} \left(\sqrt{M_2} - 1 \right)^2 + 1 \right] \lambda_2 \end{aligned} \quad (2.13)$$

which implies

$$0 = \left[\frac{\omega_{M_1}}{\omega_{M_2}} \left(\sqrt{M_2} - 1 \right)^2 + 1 \right] \lambda_2 + 2 \sqrt{\frac{\omega_{M_1}}{\omega_{M_2}}} \left(\sqrt{M_2} - 1 \right) \sqrt{\lambda_1 \lambda_2} + (\lambda_1 - 1). \quad (2.14)$$

Eq. (2.14) is a quadratic equation in $\sqrt{\lambda_2}$. If we define

$$K = \frac{\omega_{M_1}}{\omega_{M_2}} \left(\sqrt{M_2} - 1 \right)^2 \quad (2.15)$$

then (2.14) can be written as

$$(K + 1) \lambda_2 + 2 \sqrt{K \lambda_1 \lambda_2} + (\lambda_1 - 1) = 0. \quad (2.16)$$

Solving for $\sqrt{\lambda_2}$ gives

$$\sqrt{\lambda_2} = \frac{-2\sqrt{K\lambda_1} + \sqrt{4K\lambda_1 - 4(K+1)(\lambda_1 - 1)}}{2(K+1)}. \quad (2.17)$$

Therefore

$$\lambda_2 = \left[\frac{-\sqrt{K\lambda_1} + \sqrt{K+1-\lambda_1}}{K+1} \right]^2. \quad (2.18)$$

With this expression for λ_2 in terms of λ_1 , it is now possible to fully describe the optimization algorithm.

2.2.4 Optimization Algorithm

Step 1: For each possible pair of ACM modes (m_1, m_2) selected from the modes shown in Table 2.1 do the following:

Step 2: For each λ_1 in the range of $[0,1]$, increasing λ_1 by very small increments, do the following:

- a) calculate ω_{M_1} and ω_{M_2} from (2.3) where (M_1, M_2) are based on the selected ACM modes (m_1, m_2) ,

- b) calculate K using (2.15),
- c) calculate λ_2 using (2.18),
- d) evaluate γ'_1 and γ'_2 using (2.9) and (2.10), and
- e) determine $\eta(\gamma'_1, M_1)$ and $\eta(\gamma'_2, M_2)$ from Fig. 2.2.

Step 3: The combined throughput of the optimal EM constellation for a given pair of user C/I's γ_1 and γ_2 , is denoted by $\eta_{EM}(\gamma_1, \gamma_2)$ and is defined as

$$\eta_{EM}(\gamma_1, \gamma_2) = \max_{m_1, m_2, \lambda_1} (\eta(\gamma'_1, m_1) + \eta(\gamma'_2, m_2)) . \quad (2.19)$$

The values of (m_1, m_2) and λ_1 that give the optimal throughput are then used to construct the EM constellation for the given user C/I pair of (γ_1, γ_2) , by determining the signal separations as in (2.11) and (2.12).

As an example, Fig. 2.5 shows the embedded throughputs delivered by the optimal EM constellation as a function of λ_1 , for a pair of users having $\gamma_1 = 15$ dB and $\gamma_2 = 20$ dB. We observe that for λ_1 values approaching 0, there is ample margin to embed the high C/I user and provide a fairly good throughput to it. However, due the constraint of system fairness, the high C/I user cannot be allotted single-user ACM throughput because a minimal number of information bits must be embedded for the low C/I user. When λ_1 starts to increase, the throughput gradually starts to shift from the higher C/I user to the lower C/I one. Such that with λ_1 approaching 1, all the throughput goes to the lower C/I user with no possible embedding of the higher C/I user.

Figure 2.5: Embedded user throughputs as a function of λ_1 , for $\gamma_1 = 15$ dB, $\gamma_2 = 20$ dB.

Chapter 3

Scheduling Algorithms

In the previous chapter it was shown how an embedded constellation is constructed to simultaneously transmit data to two users. In a multiuser systems with more than two users the question arises of which two users to pair together at a given time. This is addressed by the scheduling algorithm used by the BTS.

In a system employing single-user ACM transmission the purpose of the scheduler is to decide which user to serve next, whereas an EM scheduler must decide which pair of users to serve. In either case the scheduler must make its decision based on balancing two competing objectives: namely, to maximize the average channel throughput and to guarantee a degree of fairness to all the users. For example, the system capacity can be maximized by serving only those users which can support the highest coding and modulation rates. However, doing so will make the system highly unfair as the users experiencing poor channel conditions will not be served. Alternatively, the scheduler can cycle through the list of all active users in a round-robin fashion and serve them in equal time slots. Although this method of scheduling is highly fair, the overall system throughput decreases when an equal proportion of time is allocated to low-rate users as to high-rate users.

More generally, balancing of throughput and fairness can be achieved by having the scheduler allocate different numbers of channel uses to each user. It is easy to see that the two examples given above are special cases of this more general approach. The first involves allocating 100% of the channel uses to just one user, and the second involves allocating an equal number of channel uses to all users.

This chapter contains a detailed description of the two well-known scheduling algorithms mentioned above. Then, after providing a formal definition of fairness, we also describe a scheduling algorithm designed to maximize throughput while meeting a minimal acceptable level of fairness, based on the aforementioned general balancing approach. This is followed with the proposal of two new scheduling algorithms, which are designed specifically to work in conjunction with EM.

3.1 Scheduling Algorithms for ACM Systems

3.1.1 Maximum C/I Scheduling Scheme (max-C/I)

When the proportion of time allocated to low rate users is increased, the overall throughput decreases. Therefore, if the throughput is to be maximized without considering the fairness allocation, the obvious solution is to allocate all the time to the maximum-rate user. Since the user with the maximum rate is also the user with the maximum C/I, this scheme is called Maximum C/I (max-C/I) scheduling. In this scheme, the BTS only serves the user with the highest C/I [15]. This scheme tends to only ever use the highest possible ACM mode for transmission, so the average throughput is very high. On the other hand, users experiencing poor C/I are seldom

if ever served, thus this scheme is very unfair and would starve some users in real situations.

For a system with N users, having C/I's $\{\gamma_i | i \in \{1, 2, \dots, N\}\}$, the max-C/I algorithm will choose to serve only the user with the highest C/I. The system throughput is therefore given by

$$\eta_{max-C/I} = \max_i \eta(\gamma_i) = \eta(\max_i \gamma_i) \quad (3.1)$$

3.1.2 Round Robin Scheduling Scheme (RR-ACM)

It is clear that a system operator would not want to provide an unfair scheme such as the max-C/I by only serving the user which is experiencing favourable channel conditions. Instead it would be preferable that all users are provided with some throughput, no matter how good or bad their instantaneous C/I is. Therefore all users would be treated equally irrespective of their location and channel conditions. If the subscribers pay an equal amount for the service, it is fair that they receive equal system resource allocation (in this case time slots).

Such a scheduling scheme would be based on the round robin (RR) algorithm in which the BTS serves all users in turn. The users are allotted equal time slots. Depending on a user's C/I, it will be assigned one of the ACM transmission modes, hence we term this scheme as the RR-ACM scheme. Although the BTS resources are allocated fairly amongst users, less efficient ACM modes must be employed to serve the users with poor C/I's. This results in a reduced system throughput. For a system with N users, having C/I's $\{\gamma_i | i \in \{1, 2, \dots, N\}\}$, the average system throughput

delivered by a RR algorithm is

$$\eta_{RR} = \frac{1}{N} \sum_{i=1}^N \eta(\gamma_i) \quad (3.2)$$

3.1.3 Fairness Criteria

In a system using ACM, different users may receive different throughput due to the difference in their C/I's, raising the question of fairness. The two scheduling schemes described above provide us with enough room in between to adjust system throughput versus its fairness. Fairness is defined as the disparity in per user throughput. Although there are many ways to describe this disparity, this thesis focuses on the fairness criteria proposed by Qualcomm and Motorola in the 3GPP2 standards forum for 1xEV-DV system evaluation [1, 16, 17]. This fairness criteria is defined in Table 3.1. The fairness is evaluated by determining the cumulative distribution function (CDF) of the user throughput normalized to the average system throughput for all the users. A system is classified as fair if the user CDF lies on or to the right of the curve given by the three points in Table 3.1. This criteria states, for example, that no more than 10% of the users can have a throughput that is less than 10% of the average system throughput.

Normalized Throughput	User CDF
0.1	0.1
0.2	0.2
0.5	0.5

Table 3.1: Fairness Criteria.

3.1.4 Minimally Fair Scheduling Scheme (MF-ACM)

In general, the user CDF for systems employing max-C/I scheduling lies to left of and above the fairness criteria curve (i.e., it is unacceptably unfair), whereas the user CDF when RR scheduling is used lies to the right of and below the curve (i.e., it is acceptably fair). By carefully controlling the number of channel uses allocated to each user it is possible to create a scheduling algorithm that just meets the minimal acceptable level of fairness as defined by the fairness criteria. The calculations required to determine the number of channel uses allocated to each user for this minimally fair scheduling algorithm for use with single-user ACM transmission (called MF-ACM), are derived below.

Suppose that at a given scheduling decision time, the algorithm needs to schedule N users with corresponding transmission rates $R_1, R_2, \dots, R_N$. Let $F_1, F_2, \dots, F_N$ be the respective number of channel uses allocated for transmission to each user. The MF-ACM scheduling algorithm needs to determine $\{F_1, F_2, \dots, F_N\}$ such that the minimum fairness criteria is just met.

The average throughput (in bits per channel use) can be defined as

$$\begin{aligned} \bar{\eta} &= \frac{\text{total \# of bits send to all users}}{\text{total \# of channel uses required to send those bits}} \\ &= \frac{\sum_{k=1}^N F_k R_k}{\sum_{k=1}^N F_k} \end{aligned} \tag{3.3}$$

and the average throughput for user n is given by

$$\eta_n = \frac{\text{total \# of bits send to user } n}{\text{total \# of channel uses to serve all users}}$$

$$= \frac{F_n R_n}{\sum_{k=1}^N F_k} . \quad (3.4)$$

The normalized throughput for user n will then be

$$\begin{aligned} \bar{\eta}_n &= \frac{\text{throughput to user } n}{\text{average throughput per user}} \\ &= \frac{\eta_n}{\frac{1}{N} \bar{\eta}} = \frac{N F_n R_n}{\sum_{k=1}^N F_k R_k} \\ &= \frac{N F_n R_n}{C} \end{aligned} \quad (3.5)$$

where C is a constant defined as

$$C = \sum_{k=1}^N F_k R_k . \quad (3.6)$$

The fairness criteria described in Section 3.1.3 provides a lower bound on the normalized throughput that can be delivered to each of the users. In particular, we need

$$\bar{\eta}_n \geq \Omega_n \quad (3.7)$$

where

$$\Omega_n = \begin{cases} 0 & \text{for } \frac{n}{N} < 0.1 \\ \frac{n}{N} & \text{for } 0.1 \leq \frac{n}{N} < 0.5 \\ 0.5 & \text{for } \frac{n}{N} = 0.5 \\ 0.5001 & \text{for } \frac{n}{N} > 0.5 \end{cases} . \quad (3.8)$$

To maximize the throughput while still meeting the minimum fairness requirements, we want to send as few bits as possible to the $N - 1$ users with low C/I, and

as many bits as possible to the max-C/I user (user N). That is, we want

$$\bar{\eta}_n = \Omega_n \quad \text{for all } n \in \{1, 2, \dots, N-1\}. \quad (3.9)$$

To determine $\bar{\eta}_N$ (i.e., the normalized throughput to the max-C/I user) note that

$$\sum_{n=1}^N \bar{\eta}_n = \frac{N \sum_{k=1}^N F_k R_k}{C} = N \quad (3.10)$$

so

$$\sum_{n=1}^{N-1} \bar{\eta}_n + \bar{\eta}_N = N \quad (3.11)$$

or

$$\bar{\eta}_N = N - \sum_{n=1}^{N-1} \bar{\eta}_n. \quad (3.12)$$

To determine the number of channel uses to allocate to each user note that from (3.5) we have

$$F_n = \frac{C \bar{\eta}_n}{N R_n}. \quad (3.13)$$

To maximize the throughput we therefore need

$$F_n = \begin{cases} \frac{C \Omega_n}{N R_n} & \text{for } 1 \leq n \leq N-1 \\ \frac{C \left(N - \sum_{k=1}^{N-1} \Omega_k \right)}{N R_N} & \text{for } n = N \end{cases}. \quad (3.14)$$

By selecting a value for C , which denotes the total number of bits delivered to all the users, (3.14) determines the number of channel uses to allocate to each user to maximize throughput while maintaining minimally acceptable fairness.

3.2 Proposed EM Scheduling Schemes

Taking perspective of the scheduling tradeoff explained at the beginning of this chapter, two embedded modulation scheduling algorithms are proposed in this thesis. The performance of these scheduling algorithms lies in between that of the max-C/I and the RR-ACM algorithms such that they are more fair than the former and produce a higher average throughput than the latter.

Both schemes work with a set of N simultaneous users. The users are sorted in an ascending order of C/I's, such that user 1 has the lowest C/I, and user N has the highest. Section 2.2 explains how to embed two users in an EM constellation. Each EM constellation is formed by pairing one of n users, $n \in \{1, 2, \dots, N-1\}$, with the max C/I user.

Always embedding the max C/I user with one of the lower C/I users offers the following advantage. Eq. (2.10) implies that for any given value of λ_1 and λ_2 , the higher the embedded user's original C/I (i.e., γ_2), the higher would be the effective C/I γ'_2 after embedding. Hence we can get the maximum gain if we choose to embed the user which has the highest C/I.

This section gives the definition of the two proposed EM schedulers. The RR-EM scheduler assigns an equal number of time slots to each of the N users in a round robin fashion, whereas the MF-EM algorithm assigns time slots which allow the system to be just fair.

3.2.1 Round Robin Scheduling Scheme (RR-EM)

In each time slot $n \in \{1, 2, \dots, N-1\}$, information for the user having the highest C/I in the set (i.e., user N) is embedded with the information for user n . In the last (N^{th}) time slot, only the max C/I (N^{th}) user is served, using single-user ACM. This algorithm has the effect of increasing the throughput in each embedded channel use over that of single-user RR-ACM.

To control fairness in this scheme, there is a restriction placed on the minimum throughput going to the low C/I user. This restriction is defined in terms of a variable β , such that the minimum throughput delivered to a low C/I user cannot be lower than β percent of the effective throughput it would have received had single-user ACM been used instead. Without this constraint the algorithm given in Section 2.2.4 would often produce an EM constellation that involved sending no bits at all to user 1, thereby violating the fairness criteria. By introducing the fairness control parameter β in this way, the scheduler can control the fairness of the scheme. If β is set to 100% then EM will not be used, and the scheduler will work as a RR scheduler. If $\beta = 0\%$, then all of the throughput will be delivered to the max C/I user alone.

3.2.2 Minimally Fair Scheduling Scheme (MF-EM)

The second proposed algorithm is called the Minimally Fair EM Scheduling algorithm (MF-EM). It is an extension of the MF-ACM algorithm, but for use with EM constellations. It is also similar to RR-EM in that the max C/I user is paired with each of the other users. The difference is that the EM constellation associated with

each pair is transmitted over a variable number of channel uses, with the goal of maximizing the total throughput while maintaining minimally acceptable fairness. The calculations to determine the number of channel uses allocated to each pair are derived below.

Suppose that the scheduler needs to schedule N users, and that user $n \in \{1, 2, \dots, N-1\}$ is paired with the max-C/I user (user N), the resulting EM constellation would deliver R_n bits to user n and Y_n bits to user N . When the max-C/I user is served by itself it receives R_N bits. Let $F_1, F_2, \dots, F_{N-1}$ be the respective number of channel uses allocated for transmission to each pair, F_N is the number of channel uses dedicated solely to the max-C/I user.

The average throughput is given by

$$\begin{aligned}\bar{\eta} &= \frac{\sum_{k=1}^{N-1} F_k (R_k + Y_k) + F_N R_N}{\sum_{k=1}^N F_k} \\ &= \frac{C}{\sum_{k=1}^N F_k}\end{aligned}\tag{3.15}$$

where

$$C = \sum_{k=1}^{N-1} F_k (R_k + Y_k) + F_N R_N\tag{3.16}$$

is the total number of bits transmitted to all the users. The average throughput for

user n is given by

$$\eta_n = \begin{cases} \frac{F_n R_n}{\sum_{k=1}^N F_k} & \text{for } 1 \leq n \leq N-1 \\ \frac{F_N R_N + \sum_{k=1}^{N-1} F_k Y_k}{\sum_{k=1}^N F_k} & \text{for } n = N \end{cases}. \quad (3.17)$$

The normalized throughput per user will then be

$$\begin{aligned} \bar{\eta}_n &= \frac{\eta_n}{\frac{1}{N} \bar{\eta}} \\ &= \frac{N \eta_n \sum_{k=1}^N F_k}{C} \\ &= \begin{cases} \frac{N F_n R_n}{C} & \text{for } 1 \leq n \leq N-1 \\ \frac{N F_N R_N + N \sum_{k=1}^{N-1} F_k Y_k}{C} & \text{for } n = N \end{cases}. \end{aligned} \quad (3.18)$$

It follows that

$$F_n = \frac{C \bar{\eta}_n}{N R_n} \quad \text{for } 1 \leq n \leq N-1 \quad (3.19)$$

and

$$F_N = \frac{\frac{C \bar{\eta}_N}{N} - \sum_{k=1}^{N-1} F_k Y_k}{R_N} \quad (3.20)$$

$$= \frac{C \bar{\eta}_N}{N R_N} - \frac{1}{R_N} \sum_{k=1}^{N-1} F_k Y_k \quad (3.21)$$

$$= \frac{C\bar{\eta}_N}{NR_N} - \frac{C}{NR_N} \sum_{k=1}^{N-1} \frac{\bar{\eta}_k Y_k}{R_k}. \quad (3.22)$$

The total number of channel uses is therefore

$$\begin{aligned} \sum_{n=1}^N F_n &= \sum_{k=1}^{N-1} F_k + F_N \\ &= \frac{C}{N} \left[\sum_{k=1}^{N-1} \frac{\bar{\eta}_k}{R_k} + \frac{\bar{\eta}_N}{R_N} - \frac{1}{R_N} \sum_{k=1}^{N-1} \frac{\bar{\eta}_k Y_k}{R_k} \right]. \end{aligned} \quad (3.23)$$

Since

$$\sum_{n=1}^N \bar{\eta}_n = \frac{N}{C} \left[\sum_{k=1}^{N-1} F_k R_k + F_N R_N + \sum_{k=1}^{N-1} F_k Y_k \right] = \frac{N}{C} [C] = N \quad (3.24)$$

we have

$$\bar{\eta}_N = N - \sum_{n=1}^{N-1} \bar{\eta}_n. \quad (3.25)$$

Substituting (3.25) for $\bar{\eta}_N$ in (3.23) gives

$$\begin{aligned} \sum_{n=1}^N F_n &= \frac{C}{N} \left[\sum_{k=1}^{N-1} \frac{\bar{\eta}_k}{R_k} + \frac{N}{R_N} - \sum_{k=1}^{N-1} \frac{\bar{\eta}_k}{R_N} - \frac{1}{R_N} \sum_{k=1}^{N-1} \frac{\bar{\eta}_k Y_k}{R_k} \right] \\ &= \frac{C}{N} \left[\frac{N}{R_N} + \sum_{k=1}^{N-1} \bar{\eta}_k \left(\frac{1}{R_k} - \frac{1}{R_N} - \frac{Y_k}{R_N R_k} \right) \right] \\ &= \frac{C}{N} \left[\frac{N}{R_N} + \sum_{k=1}^{N-1} \bar{\eta}_k \left(\frac{R_N - R_k - Y_k}{R_N R_k} \right) \right]. \end{aligned} \quad (3.26)$$

By substituting (3.26) into (3.15), the average throughput can be expressed as

$$\bar{\eta} = \frac{C}{\frac{C}{N} \left[\frac{N}{R_N} + \sum_{k=1}^{N-1} \bar{\eta}_k \left(\frac{R_N - R_k - Y_k}{R_N R_k} \right) \right]} \quad (3.27)$$

$$= \frac{N}{\frac{N}{R_N} + \sum_{k=1}^{N-1} \bar{\eta}_k \left(\frac{R_N - R_k - Y_k}{R_N R_k} \right)}. \quad (3.28)$$

To maximize $\bar{\eta}$, we need to minimize the denominator term

$$\sum_{k=1}^{N-1} \bar{\eta}_k \left(\frac{R_N - R_k - Y_k}{R_N R_k} \right). \quad (3.29)$$

This is accomplished in part by minimizing the normalized throughput to the low C/I users. That is, we need $\bar{\eta}_k = \Omega_k$ for all $k \in \{1, 2, \dots, N-1\}$, where Ω_k is given by (3.8). Hence

$$\bar{\eta} = \frac{N}{\frac{N}{R_N} + \sum_{k=1}^{N-1} \Omega_k \left(\frac{R_N - R_k - Y_k}{R_N R_k} \right)}. \quad (3.30)$$

Examination of (3.30) reveals that maximization of the average throughput also requires careful construction of the EM constellation. Maximization of $\bar{\eta}$ requires that

$$\left(\frac{R_N - R_k - Y_k}{R_N R_k} \right) \quad (3.31)$$

be minimized for all $k \in \{1, 2, \dots, N-1\}$. It is not sufficient to merely choose the EM constellation with the highest total throughput (i.e., the largest value of $R_k + Y_k$). The optimal constellation can still be found using the algorithm given in Section 2.2.4, except that

$$\eta_{EM}(\gamma_1, \gamma_2) = \min_{m_1, m_2} \left(\frac{R_N - \eta(\gamma'_1, m_1) - \eta(\gamma'_2, m_2)}{R_N \eta(\gamma'_1, m_1)} \right) \quad (3.32)$$

is used in place of (2.19).

3.3 Discussion on System Complexity

This section describes the increase in system complexity for both the transmitter and receiver when EM transmissions are used in place of ACM.

An RR-EM transmitter would need to search for the max C/I user amongst concurrent system users. A similar search is carried out by the max-C/I transmitter, but not the RR-ACM transmitter. A BTS employing an MF-EM scheduler would need to sort all the users with respect to their C/I's, same would be the case in MF-ACM scheduling. Constructing an EM constellation for a given pair of user C/I's could either be done at each scheduling decision time by finding m_1 , m_2 , d_1 and d_2 using the algorithm in Section 2.2.4, or a look-up table could be created for this purpose. The size of this table would depend on the range of serveable C/I's for a system and the C/I quantization. For example, for the system studied in this thesis, the C/I range is [4,20] dB and the quantization size is 0.1 dB. So the resultant look-up table would contain $(160)^2$ entries, one for each possible pair of C/I's.

For communication involving EM, the BTS also needs to convey additional information on the forward control with regards to both the embedded modes m_1 , m_2 and the adaptive interpoint separations d_1 , d_2 . In case of ACM, the receiver only needs information regarding the transmitted mode, since the interpoint separations are fixed for a given mode.

The lower C/I receiver operates identically to an ACM system receiver. On the other hand, the max C/I user needs to perform some initial operations on the received samples. In particular it has to take the absolute values of the in-phase and quadrature-phase channel outputs and subtract a value equal to $\lambda_1/2$ from both. This result is then fed to a standard ACM decoder. An EM receiver must also be able to distinguish between both these operating assignments as sometimes it would

be the max C/I user whereas at other times it could be a low C/I user.

Chapter 4

Simulation Results and Discussion

This chapter presents the performance evaluation results obtained by means of computer simulation for the two EM scheduling algorithms described in Chapter 3. The average channel throughput for these EM schemes is compared to the average channel throughput provided by single-user ACM transmission. To further explore the tradeoff between throughput maximization and system fairness, this chapter carries the fairness curves obtained by each of the schemes. They are based on the cellular system model given in Appendix B. The system parameter values used in the simulations, which also correspond to the 1xEV-DV system [1], are summarized in Table B.1. Based on this system model, the cumulative density function (CDF) of the user C/I was generated by means of computer simulation. This CDF is plotted in Fig. 4.1. We note that the C/I's are limited to the range [4,20] dB, and that the majority of the user C/I's lie in the lower portion of the range. In fact there may be a large number of potential users whose C/I's fall below 4 dB. Such users cannot be served regardless of which link adaptation technique we employ, hence they remain unserveable with both ACM and EM.

To evaluate the performance of these scheduling algorithms, the C/I's are drawn randomly from this CDF for N users. Transmissions for these N users are then sched-

Figure 4.1: Cumulative Density Function of User C/I's

uled at the BTS and the average throughput delivered per channel use is determined. This process is repeated 100,000 times. In other words, the process undergoes 100,000 scheduling decision times, called runs, and the average over all these runs is then presented. It is assumed that all the users have some data to send at all times, i.e., they have full buffers during the run. We do not consider traffic modelling, so the number of concurrent users in the system (N) does not represent the total number of users present. In fact it represents only the actual number of users which have some data to send at the scheduling time.

4.1 Existing ACM Schedulers

Fig. 4.2 shows the average system throughput for the max-C/I and RR-ACM scheduling schemes. It is apparent that the throughput for the max-C/I scheme

depends on the number of concurrent system users. When there are only a few users it is quite likely that the C/I of the max C/I user will not be very high, so one of the lower-rate ACM modes will be used on that link. However, when there are many users it is almost certain that the C/I of the max C/I user will be sufficiently large, ensuring that a higher-rate ACM mode is used.

The RR-ACM scheduler, on the other hand, provides a constant average throughput regardless of the number of users. Since all the users are served, including the low-rate ones, the average throughput is much lower than that for the max-C/I scheme.

Fig. 4.3 contains the fairness curves for these schedulers for a system with $N = 10$ users. The fairness criteria from Table 3.1 is also plotted as a reference curve. In general, a fairness curve shows the CDF of the normalized throughput, for a system with N users. This gives the fraction of users receiving a throughput equal to or less than the corresponding value along the abscissa. The abscissa represents user throughput normalized to the average value so that any evaluation is independent of the absolute throughput numbers. The CDF for a fair system must fall on, below or to the right of the indicated fairness criteria. From Fig. 4.3 we see that the RR-ACM curve demonstrates high throughput fairness amongst the system users, whereas the max-C/I curve tells us that although 10% of the users are served with a high throughput, the remaining 90% are not served at all. Hence we can see that the max-C/I scheduler attains a high average throughput by transmitting to users employing high ACM modes, but on the other hand, the system becomes highly

Figure 4.2: Upper and lower bounds on the system throughput given by the max-C/I and RR-ACM schemes respectively, as a function of the number of users N .

Figure 4.3: Fairness Curves showing the CDF of the normalized average throughput for the max C/I and RR-ACM schemes, for a system with 10 users.

unfair by not serving the lower C/I users. The RR-ACM scheduler transmits on low C/I links using low ACM modes an equal number of times as it transmits on higher C/I links using high ACM modes. Although this reduces the average throughput as compared to the max-C/I scheduling, it helps in maintaining system fairness.

4.2 *RR-EM Scheduler*

Fig. 4.4 shows the average throughput obtained by the RR-EM scheme as a function of the number of system users, for various values of the fairness control factor β , as defined in Section 3.2.1. As can be seen, decreasing β leads to improved throughput. This is because decreasing β results in less data being sent to low C/I users, allowing more embedded throughput to be sent to the max C/I user. With an appropriate choice of β it would be possible to attain any desired throughput between the system bounds established by the max-C/I and the RR-ACM schedulers.

Although it is desirable in terms of average throughput, to make β as small as possible, doing so will lead to a violation of the fairness criteria. For any given number of users there is a specific minimum allowable value of β (β_{min}) that will provide the minimum required fairness. This implies that the embedded throughput to the lower C/I user should not be less than $\beta_{min}\%$ of its ACM throughput. In RR-EM scheduling, the value of β_{min} varies with the number of concurrent system users N . This value of β_{min} has been determined by means of simulations for different numbers of concurrent users and plotted in Fig. 4.5.

It can be seen in Fig. 4.5 that as N is increased, β_{min} increases too. With

Figure 4.4: Effect of varying β on the average throughput of the RR-EM scheme as a function of the number of users N .

an increased number of users the total embedded throughput to the max-C/I user increases. Furthermore, there is a higher probability of having users with good C/I's, which will embed the max-C/I user and provide higher total embedded throughput. Hence in both cases, the average system throughput would increase such that the normalized throughput to the lower C/I users decreases due to increase in embedded throughput destined for the max C/I user, and the system violates the fairness criteria at some point. The solution to this problem would be to increase the percentage of throughput assigned to the lower C/I users such that system fairness is maintained, which effectively means increasing the value of β_{min} .

Figure 4.5: β_{min} values as a function of the number of users, N .

Fig. 4.4 also contains the curve for average throughput obtained at those values for β_{min} which were shown in Fig. 4.5 for the different number of users. Any RR-EM throughput curve above the curve for β_{min} is unfair, whereas any throughput curve below the β_{min} curve is fair but provides lower throughput than the β_{min} curve. Thus the curve for β_{min} represents optimal RR-EM system performance in terms of both the throughput and fairness. This graph shows that, for example, the RR-EM scheme is able to deliver a fair average throughput of 2.92 bpchu for a system serving 50 simultaneous users. This represents a gain of 33% over the RR scheme average channel throughput of 2.22 bpchu as shown in Fig. 4.2.

The fairness curves of the RR-EM schemes as well as max-C/I and RR-ACM

for $N = 6, 10, 20, 50$ are given in Fig. 4.6 to Fig. 4.9. Each of these curves show the fairness provided by the RR-EM scheme at different values of β . For the sake of analysis, let us consider Fig. 4.7, where $N = 10$. With β set to 10%, the average throughput is 2.81 bpchu but the system violates the fairness criteria, whereas for a β value equal to 50% the fairness curve lies just to the right of the fairness criteria, giving an average system throughput equal to 2.76 bpchu. A higher value of β , such as 75%, would provide a more fair system but with a reduced throughput of 2.55 bpchu. This so happens because with increased fairness, the poor C/I users are now getting more bits and the embedded bits to the max C/I users are decreased. So the fairness curve starts to shift towards the RR-ACM curve because eventually if β is set to a 100%, then there is no embedding and RR-EM would degenerate into RR-ACM. Through these fairness curves we can observe the tradeoff between system throughput and system fairness and deduce the result that $\beta_{min} = 50\%$. With the RR-EM scheduler now operating with a β value of 50%, the gain in average throughput is equal to 25.5% over the RR-ACM throughput of 2.2 bpchu.

4.3 MF Schedulers

Fig. 4.10 shows the performance of the Minimal Fairness schemes in terms of their average channel throughput as a function of the number of users (N). The throughput curves for max C/I, RR-EM and RR-ACM are also plotted for reference. There is improvement in system throughput when the Minimal Fairness schemes are used as opposed to the RR schemes. As anticipated, the MF-EM scheme outperforms

Figure 4.6: Fairness Curves showing the CDF of the normalized average throughput for the max-C/I, RR-ACM and RR-EM schemes with varying β values, for a system with 6 users.

Figure 4.7: Fairness Curves showing the CDF of the normalized average throughput for the max-C/I, RR-ACM and RR-EM schemes with varying β values, for a system with 10 users.

Figure 4.8: Fairness Curves showing the CDF of the normalized average throughput for the max-C/I, RR-ACM and RR-EM schemes with varying β values, for a system with 20 users.

Figure 4.9: Fairness Curves showing the CDF of the normalized average throughput for the max-C/I, RR-ACM and RR-EM schemes with varying β values, for a system with 50 users.

the MF-ACM scheme. When compared with RR-ACM, MF-EM demonstrates a gain of 60.5% for $N = 100$, delivering 3.53 bpchu. For the same N , the throughput delivered by MF-ACM is 3.26 bpchu which is equal to a gain of 48% over the RR-ACM throughput of 2.2 bpchu.

Fig. 4.11 show the fairness curve for the MF schemes $N = 50$ system users. The mark for the last N th user occurs at the coordinates (32,1), but is not visible on the shown scale. This means that the normalized throughput for the max C/I user is 32 times the average system throughput. Hence this algorithm tends to produce a high average throughput by maximizing the number of time slots assigned to the max C/I user, while serving the lower $N - 1$ users in a minimally fair number of time slots. In this way the highest possible mode is scheduled for transmission a fairly larger number of times than the lower modes for the poor C/I users thus the average throughput gets maximized.

The fairness curve for both the Minimal Fairness schemes is identical. This curve demonstrates the principle of Minimal Fairness, by following the fairness criteria as closely as possible. But it is interesting to note that although there is equal fairness amongst both MF schemes, MF-EM delivers a higher throughput of 3.27 bpchu versus an MF-ACM throughput of 3.09 bpchu, as shown in Table 4.1. This proves the obvious advantage of embedded modulation. When we make a comparison between RR-EM and MF-EM, we notice that just under 10% of the users (for which $\Omega_n < 0.1$) remain unserved with MF-EM, whereas RR-EM would serve all the users. Secondly the number of time slots to achieve instantaneous fairness in case of MF-EM may

Figure 4.10: Average throughput of the proposed EM scheduling schemes as compared to max-C/I, MF-ACM and RR-ACM schemes, as a function of the number of users N .

Figure 4.11: Fairness Curves showing the CDF of the normalized average throughput for the MF schemes, for a system with 50 users.

be large as compared to round robin scheduling in RR-EM, because where RR-EM assigns equal slots to each of the users, MF-EM assigns only a fraction of the number of slots assigned to the max C/I user to each of the lower users. Hence the scheduler would have to run a longer time in order to assign enough slots to the lower C/I users to actually achieve fairness. On the other hand, it is significantly noticeable in Fig. 4.10 that the throughput performance of the MF-EM scheme lies closer to the max-C/I curve whereas that of the RR-EM is much closer to the RR-ACM curve. Hence both these EM scheduling schemes demonstrate different levels of link improvement.

Table 4.1 summarizes the average system throughput delivered by all the different schemes discussed in this thesis, along with the percentage gains achieved by each scheduling scheme over the RR-ACM scheme.

Number of Users	Scheduling Scheme	Average Throughput Value (bpchu)	%age gain over RR-ACM
$N = 6$	RR-ACM	2.2	-
	RR-EM	2.66	21%
	MF-ACM	2.67	21.4%
	MF-EM	2.77	26%
	Max-C/I	3.56	62%
$N = 10$	RR-ACM	2.2	-
	RR-EM	2.76	25.5%
	MF-ACM	2.87	30.5%
	MF-EM	3.01	37%
	Max-C/I	3.89	77%
$N = 20$	RR-ACM	2.2	-
	RR-EM	2.85	29.5%
	MF-ACM	3.09	40.5%
	MF-EM	3.27	48.6%
	Max-C/I	4.25	93%
$N = 50$	RR-ACM	2.2	-
	RR-EM	2.92	32.7%
	MF-ACM	3.23	47%
	MF-EM	3.47	57.7%
	Max-C/I	4.47	103%

Table 4.1: System Throughput Performance

Chapter 5

Conclusions and Future Research

5.1 Conclusions

In this thesis we presented a detailed analysis of embedded modulation when used as a link adaptation technique. A new adaptive use of EM for multiuser links in cellular wireless networks employing feedback channels has been proposed. In this proposed application multiple users with widely different C/I's are allowed to share a single downlink channel simultaneously, wherein each user derives separate information from the same received signal. In this kind of modulation, the information bits meant for distinctly located users are combined in a single ACM signal constellation using multi-level modulation and coding. It is shown how to derive the energy equation of an EM symbol constellation, and consequently how to construct the optimal constellation which maximizes throughput. Multiuser embedded modulation realizes higher average throughput per channel use than that achieved by single-user ACM. We also proposed two separate algorithms, RR-EM and MF-EM, to schedule data transmissions for mobile users by means of EM. Through these scheduling algorithms, we investigated the potential benefits in terms of increased system throughput of cellular wireless networks. Results show significant improvement in

system throughput when compared to the throughput provided by the previously existing RR-ACM scheduling scheme. Along with this benefit, there is the added advantage of the throughput being fairly distributed amongst system users. Both our schedulers comply with the fairness criteria which had been specified as a benchmark in this thesis.

Using embedded modulation with an appropriate scheduling algorithm provides a flexible means for increasing the aggregate channel throughput while guaranteeing a degree of throughput fairness to all the users. Always embedding the max C/I user with the lower C/I users is a simple method for acquiring higher throughput as well as delivering an acceptably fair number of bits to the lower C/I users. The lower C/I users are not as disadvantaged as they would be in a max-C/I scheme, and the average channel throughput increases over that of RR-ACM and MF-ACM. The fair average throughput delivered by the RR-EM scheme is up to 33% higher than the average throughput delivered by the RR-ACM scheme. And the fair average throughput delivered by the MF-EM scheme is up to 60% higher than the average throughput delivered by the RR-ACM scheme. Hence we conclude that EM scheduling provides a balanced and fair alternative to single-user ACM scheduling.

5.2 Proposed Future Research

This thesis raises a number of interesting issues which may be dealt with in future research work. Some propositions are given as follows:

- Multiuser embedding for more than two simultaneously served users in one EM

constellation under fixed and variable power constraints.

- To find an analytical expression for β_{min} of an RR-EM scheduler.
- To reduce the number of unserved users in MF.
- To derive an algorithm which matches the fairness criteria as closely as possible with just one channel use per user. This implies bridging the gap between RR-EM and MF-EM.

Appendix A

Energy Calculations for EM Constellations

An EM signal constellation can be constructed by combining two lower-level constellations, $\mathcal{C}_1$ and $\mathcal{C}_2$, containing M_1 and M_2 points with minimum interpoint separations of Δ_1 and Δ_2 , respectively. The points in $\mathcal{C}_1$ are the positions of the centres of the clouds in the EM constellation, and the points in $\mathcal{C}_2$ determine the relative positions of the points inside each cloud. For all $k \in \{0, \dots, M_1 - 1\}$ and $l \in \{0, \dots, M_2 - 1\}$, the points in the EM constellation are given by

$$S_{kM_2+l} = p_k + q_l \quad (\text{A.1})$$

where $p_k \in \mathcal{C}_1$ and $q_l \in \mathcal{C}_2$. Examples of $\mathcal{C}_1$ and $\mathcal{C}_2$, with minimum separations Δ_1 and Δ_2 , are shown in Fig. A.1. Combining $\mathcal{C}_1$ and $\mathcal{C}_2$ results in the EM constellation shown in Fig. A.2. The average energy per symbol in an EM constellation with signal points given by (A.1) is

$$\begin{aligned} E_{s(EM)} &= \frac{1}{M_1 M_2} \sum_{k=0}^{M_1-1} \sum_{l=0}^{M_2-1} |S_{kM_2+l}|^2 \\ &= \frac{1}{M_1 M_2} \sum_{k=0}^{M_1-1} \sum_{l=0}^{M_2-1} |p_k + q_l|^2 \\ &= \frac{1}{M_1 M_2} \sum_{k=0}^{M_1-1} \sum_{l=0}^{M_2-1} \left[|p_k|^2 + p_k^* q_l + p_k q_l^* + |q_l|^2 \right] \end{aligned}$$

Figure A.1: Arbitrary lower level M_1 -QAM and M_2 -QAM constellations.

Figure A.2: Construction of an $M_1 \times M_2$ EM constellation.

$$\begin{aligned}
&= \frac{1}{M_1} \sum_{k=0}^{M_1-1} |p_k|^2 + \left(\frac{1}{M_1} \sum_{k=0}^{M_1-1} p_k \right)^* \left(\frac{1}{M_2} \sum_{l=0}^{M_2-1} q_l \right) \\
&\quad + \left(\frac{1}{M_1} \sum_{k=0}^{M_1-1} p_k \right) \left(\frac{1}{M_2} \sum_{l=0}^{M_2-1} q_l \right)^* + \frac{1}{M_2} \sum_{l=0}^{M_2-1} |q_l|^2 . \tag{A.2}
\end{aligned}$$

In this thesis we only consider $\mathcal{C}_1$ and $\mathcal{C}_2$ being either square QAM or BPSK constellations. In either case, the constellations have zero mean, (i.e., the points are centered around the origin), so

$$\frac{1}{M_1} \sum_{k=0}^{M_1-1} p_k = \frac{1}{M_2} \sum_{l=0}^{M_2-1} q_l = 0 . \tag{A.3}$$

Therefore it follows that

$$\begin{aligned}
E_{s(EM)} &= \frac{1}{M_1} \sum_{k=0}^{M_1-1} |p_k|^2 + \frac{1}{M_2} \sum_{l=0}^{M_2-1} |q_l|^2 \\
&= E_{s(\mathcal{C}_1)} + E_{s(\mathcal{C}_2)} \tag{A.4}
\end{aligned}$$

where $E_{s(\mathcal{C}_1)}$ and $E_{s(\mathcal{C}_2)}$ are the average symbol energies for $\mathcal{C}_1$ and $\mathcal{C}_2$ respectively.

The average transmitted energy of a square M -QAM (or BPSK) constellation with separation Δ is

$$E_s = \omega_M \Delta^2 \tag{A.5}$$

where

$$\omega_M = \begin{cases} \frac{1}{4} & \text{for } M = 2 \text{ (BPSK)} \\ \frac{1}{6} (M - 1) & \text{for } M = \text{even power of 2 (square } M\text{-QAM)} \end{cases} . \tag{A.6}$$

Therefore

$$E_{s(\mathcal{C}_1)} = \omega_{M_1} \Delta_1^2 \tag{A.7}$$

and

$$E_{s(\mathcal{C}_2)} = \omega_{M_2} \Delta_1^2 . \tag{A.8}$$

It follows from (A.4) that

$$E_{s(EM)} = \omega_{M_1} \Delta_1^2 + \omega_{M_2} \Delta_1^2. \quad (\text{A.9})$$

Eq. (A.9) gives average energy of the EM constellation in terms of Δ_1 and Δ_2 , but we need an expression for $E_{s(EM)}$ in terms of the inter- and intra-cloud separations d_1 and d_2 . Note that

$$d_1 = \Delta_1 - \left(\sqrt{M_2} - 1 \right) \Delta_2 \quad (\text{A.10})$$

and

$$d_2 = \Delta_2. \quad (\text{A.11})$$

Solving for Δ_1 and Δ_2 gives

$$\Delta_1 = d_1 + \left(\sqrt{M_2} - 1 \right) d_2 \quad (\text{A.12})$$

and

$$\Delta_2 = d_2. \quad (\text{A.13})$$

Substituting (A.12) and (A.13) into (A.9) gives

$$\begin{aligned} E_{s(EM)} &= \omega_{M_1} \left[d_1 + \left(\sqrt{M_2} - 1 \right) d_2 \right]^2 + \omega_{M_2} d_2^2 \\ &= \omega_{M_1} d_1^2 + 2\omega_{M_1} \left(\sqrt{M_2} - 1 \right) d_1 d_2 \\ &\quad + \left[\omega_{M_1} \left(\sqrt{M_2} - 1 \right)^2 + \omega_{M_2} \right] d_2^2 \end{aligned} \quad (\text{A.14})$$

which is the average symbol energy of an $M_1 \times M_2$ EM constellation with minimum separations d_1 and d_2 .

Appendix B

Forward Link Design

In the design of communication systems used for transmitting information through physical channels, it is convenient to employ mathematical models that reflect the most important characteristics of the transmission medium. This appendix begins by describing the characteristics of the wireless channel when used as the medium of transmission. Generic channel models for the downlink in urban wireless networks are presented. This appendix describes the complete system model in detail and all system parameter values are tabulated at the end of the appendix.

B.1 The Wireless Channel

Wireless communication channels are characterized by several time-varying attributes, including path loss depending on the transmitter-receiver (T-R) separation, shadowing caused by partial blockage of the transmission path by, for example, foliage or rainfall, and fading due to multipath propagation.

Both theoretical and measurement-based radio propagation models indicate that the average received signal power decreases logarithmically with distance, whether it is an indoor or outdoor radio channel. The average large-scale path loss for an

arbitrary T-R separation is expressed as a function of distance as follows

$$\overline{PL}(d)_{dB} = \overline{PL}(d_0) + 10\rho \log \left[\frac{d}{d_0} \right] \quad (\text{B.1})$$

where ρ is the path loss exponent which indicates the rate at which the path loss increases with distance, d_0 is the close-in reference distance which is determined in a path loss model from measurements taken close to the transmitter, and d is the T-R separation. The overbars in (B.1) denote the ensemble average of all possible path loss values for a given value of d .

This equation, however, does not take into consideration the fact that the surrounding environmental clutter may also be vastly different at two different locations having the same T-R separation, so the measured signals can be vastly different than the average value predicted by (B.1). At any value of d , the path loss $PL(d)$ at a particular location is random and log-normally (normal in dB) distributed about its distance-dependent mean value $\overline{PL}(d)$. This implies that

$$PL(d) = \overline{PL}(d) - SH_\sigma \quad (\text{B.2})$$

$$= \overline{PL}(d_0) + 10\rho \log \left[\frac{d}{d_0} \right] - SH_\sigma \quad (\text{B.3})$$

where SH_σ is a zero-mean Gaussian random variable in dB with standard deviation σ dB. The Gaussian or log-normal distribution describes the random *shadowing* effects which occur over a large number of measurement locations which have the same T-R separation, but have different levels of clutter on the propagation path. This phenomenon is referred to as *log-normal shadowing*. Simply put, log-normal shadowing implies that the measured signal levels at a fixed T-R separation have a Gaussian distribution about the distance-dependent mean of (B.1) [18, 19].

Multipath propagation in the radio channel creates small-scale *fading* effects, such as time dispersion which causes echoes and loss of signal due to deep fades. Small-scale fading is used to describe the rapid fluctuations of the amplitude of a radio signal over a short period of time or travel distance. Fading is caused by interference between two or more versions of the transmitted signal which arrive at the receiver at slightly different times. These waves, called *multipath waves*, combine at the receiver antenna to give a resultant signal which can vary in amplitude and phase, depending on the distribution of the intensity and relative propagation time of the waves and the bandwidth of the transmitted signal. Even when a mobile terminal is stationary, the received signal may fade due to movement of surrounding objects in the propagation path. Even when a line-of-sight path exists, multipath occurs due to reflections from the ground and surrounding structures. The incoming radio waves arrive from different directions with different propagation delays. Hence a distorted version of the signal is received.

The Rayleigh distribution is commonly used to describe the statistical time-varying nature of the received envelope of an individual multipath fading component. When there is a dominant stationary(nonfading) signal component present, such as a line-of-sight propagation path, the small-scale fading envelope distribution is Ricean. As the dominant signal becomes weaker, the composite signal resembles a noise signal which has an envelope that is Rayleigh. Thus, the Ricean distribution degenerates to a Rayleigh distribution when the dominant component fades away.

Eq. (B.3) alongwith the fading distribution statistically describe the path loss

model for any given T-R separation, and this model may be used in computer simulation to provide received power levels for randomly located mobile units in communication system design and analysis.

B.1.1 Received Power

In the downlink direction, the received signal strength at the receiver antenna is computed by means of the following link budget equation (B.4) [18]. It takes into account the effects of antenna gains, path loss, shadowing and multipath fading.

$$P_{r||dBm} = P_{t||dBm} + G_{t||dB} + G_{r||dB} - \overline{PL}_{||dB} + SH_{||dB} + F_{||dB} \quad (B.4)$$

where

P_r = received signal power at the subscriber station antenna

P_t = transmitted signal power at the BTS antenna

G_t = transmitting antenna gain

G_r = receiving antenna gain

$\overline{PL}$ = path loss as given by (B.1)

SH = lognormal shadowing in dB, see Appendix C

F = a random variable that accounts for the Ricean

fading in the channel, in dB, see Appendix C

B.1.2 System Parameters

The system parameters used during the simulation in order to calculate the power received on the forward link are described in this section.

Antenna Pattern

The BTS antenna gain used for each sector on the forward link is determined by

$$G_t [dB] = \text{Transmit Antenna gain} - \text{Cable Loss} - \min \left[12 \left(\frac{\theta}{\theta_{3dB}} \right)^2, A_{max} \right] \quad (B.5)$$

where

$$\theta = \text{horizontal transmit angle at BTS, } -180^\circ \leq \theta \leq 180^\circ$$

$$\theta_{3dB} = 3 \text{ dB beamwidth}$$

$$A_{max} = \text{maximum attenuation} = 20 \text{ dB}$$

The BTS antenna pattern is depicted by the third term in B.5. This antenna pattern has been employed in this research and is plotted in Fig. B.1. Determination of the horizontal transmit angle, θ , at the BTS has been described in section B.1.3.

Ricean Fading

It is assumed that the multipath fading environment is characterized by the Ricean probability density function (PDF). The Ricean model assumes that the channel induces an amplitude $\alpha(t)$ which varies in time according to the Ricean distribution. The fading envelope distribution is said to be Ricean when there is a dominant non-fading signal component present, such as a line-of-sight propagation path. In such a situation, random multipath components arriving at different angles are superimposed on a stationary (non-fading) dominant signal. At the output of an envelope detector, this has an effect of adding a DC component to the random multipath, [18].

Figure B.1: BTS antenna pattern for each sector

B.1.3 Cell Layout

The research in this thesis is based on a hexagonal cell layout containing 19 three-sectored hex cells. So a total of 57 sectors are modeled, out of which the center sector #1 is always the serving sector and the remaining fifty six act as interferers. The separation between adjacent base transceiver stations is selected as a constant value throughout the cell layout. This is also the hex cell diameter, measured by drawing a straight line from the center of one side to the center of the opposite side, passing through the BTS. This hex cell layout is drawn in Fig. B.2. The center cell orientation and corresponding sector antenna orientations at 0, 120 and 240 degrees are shown in Fig. B.3. The antenna directions are represented by arrows drawn in the center of each sector. The main antenna lobe center directions for each sector

point to the side of the hexagon. The main antenna lobe center directions of the other 18 surrounding cells are assumed parallel to those of the center cell.

Figure B.2: Hex cell layout

(a)

(b)

Figure B.3: (a) Orientation of the center cell #1; (b) Center cell antenna 0, 120, 240 degrees sector orientation diagram

Antenna Orientation

Antenna orientation for each sector of the center cell is shown in Fig. B.3(a). All antennas are drawn in the center of each sector as arrows pointing to the sides of the hexagon. All the other 18 cells lie parallel to cell #1. Antenna bearing is defined as the angle between the sector antenna direction and a line directed due East given in degrees. The bearing angle increases in a clockwise direction. Fig. B.3(b) shows the 3 sector antennas of the center cell, such that the sector #1 antenna has a bearing of zero degrees due East, the sector #2 antenna has a bearing of 120 degrees and the sector #3 antenna has a bearing of 240 degrees. All antenna main lobe directions shall point to the sides of the hexagon.

BTS Transmit Angle

Let ϵ_i represent the antenna bearing for the i th sector in each cell as shown in Fig. B.4. Hence

$$\epsilon_i = \begin{cases} 0^\circ & i = 1 \\ 120^\circ & i = 2 \\ 240^\circ & i = 3 \end{cases} \quad (\text{B.6})$$

We now need to know the BTS transmit angle θ in order to calculate the BTS antenna gain from B.5. Fig. B.4 shows the hex cell geometry used in calculating the transmit angles at each sector antenna of a given cell. Assume a mobile k is dropped at a position (mobile(k)_xpos, mobile(k)_ypos). At the beginning we shall study a generalized scenario for any given cell # n located at the coordinates (cell(n)_xpos, cell(n)_ypos) and present the equations for calculating the transmit angle from each

sector BTS in cell $\#n$ to the mobile k . Following that we shall study the example of bearing orientations of mobile k with respect to two cells, BTS $\#1$ and BTS $\#6$. For all notations, refer to Fig. B.4. The BTS transmit angle of sector i in cell $\#n$ to the mobile k is denoted by $\theta_{n(i,k)}$ and it is defined as follows

$$\theta_{n(i,k)} = \delta_{n,k} - \epsilon_i \quad (\text{B.7})$$

where

$\theta_{n(i,k)}$ = the angle between the i th sector antenna of the n th cell

and the line connecting BTS $\#n$ and mobile k in the horizontal plane

$\delta_{n,k}$ = the mobile bearing, the angle between a line directed due East from BTS

$\#n$ and the line connecting BTS $\#n$ and mobile k in the horizontal plane

$$= \arctan \left(\frac{\text{mobile}(k)\text{-ypos} - \text{cell}(n)\text{-ypos}}{\text{mobile}(k)\text{-xpos} - \text{cell}(n)\text{-xpos}} \right)$$

ϵ_i = the antenna bearing for the i th sector in cell $\#n$

Sector $\#1$ in the center cell $\#1$ is always the serving sector and the remaining 56 sectors act as the interferers. In cell $\#1$, the BTS transmit angle of sector $\#1$ is $\theta_{1(1,k)}$, the BTS transmit angle of sector $\#2$ is $\theta_{1(2,k)}$ and the BTS transmit angle of sector $\#3$ is $\theta_{1(3,k)}$. The mobile's bearing from the center cell is denoted by $\delta_{1,k}$. It is to be noted that in this cell layout $\theta_{n(1,k)} = \delta_{n,k}$, since $\epsilon_1 = 0^\circ$. The mobile bearing in cell $\#6$ is $\delta_{6,k}$, extending between the line directed due East from BTS $\#6$ and the line connecting BTS $\#6$ and mobile k . Also $\theta_{6(1,k)} = \delta_{6,k}$ is the BTS transmit angle of sector $\#1$ in this cell, the BTS transmit angle of sector $\#2$ is $\theta_{6(2,k)}$, extending between the sector $\#2$ antenna and the line connecting BTS $\#6$ and mobile k . Similarly the

Figure B.4: Mobile Bearing Orientation

BTS transmit angle of sector #3 is $\theta_{6(3,k)}$ and it extends between the sector #3 antenna and the line connecting BTS #6 and mobile *k*.

B.1.4 C/I Calculation

In evaluating system performance, the quantity of greatest interest is the carrier-to-interference ratio (C/I). This is so because the basic system design centers on the ability to detect the signal, with an acceptable error probability, in the presence of multiple-access interference and system noise.

This means that the signal is not only distorted by the channel, but it is also contaminated along the path by undesirable signals lumped under the broad term *additive noise* which are random and unpredictable signals from causes external and internal. Noise is one of the basic factors that sets a limit on the rate of communication. Physically, the internal additive noise process may arise from electronic components such as amplifiers at the receiver. Such noise is categorized as thermal noise, and it is statistically characterized as a *Gaussian noise process* with a flat (constant) power spectral density over the entire frequency range. Hence, the resulting mathematical model for the channel is called the *additive white Gaussian noise* (AWGN) model. The transmitted signal $s(t)$ gets disturbed by a simple AWGN process $n(t)$ and the resultant signal at the receiver is given by

$$r(t) = \alpha(t)s(t) + n(t) \tag{B.8}$$

where $\alpha(t)$ is the time varying channel attenuation factor. Assuming a *memoryless channel*, the samples of the AWGN process are uncorrelated implying that noise effects each transmitted symbol independently. The term “*additive*” means that this noise is simply superimposed or added to the signal and that there are no multiplicative mechanisms at work.

External noise can be both natural and man-made, such as atmospheric disturbances, lightning, spark-plug ignition noise, switching transients and other electromagnetic radiations. Additional interference is introduced by surrounding base transceiver stations (BTS). Multiple access interference is a general constraint in any communication system that shares channel resources. This is generated by the fact that every user receives some power from transmissions intended for other users, in addition to the signal that is specifically directed to it. For a particular subscriber unit, it's serving BTS will provide the desired forward channel while the surrounding co-channel BTS's will provide the *forward channel interference*.

The *carrier-to-interference ratio* (C/I), is defined as the ratio of the received signal power to the total interference and noise power,

$$C/I = \frac{P_r}{\sum_{n=1}^{N_i} \sum_{i=1}^3 I_{n,i} + NT} \quad (\text{B.9})$$

where

P_r = *received signal power from the serving sector*

$I_{n,i}$ = *interference signal power from interfering sector #i in cell #n*

N_i = *total number of cells in the layout*

NT = *total channel noise power*

Since the signal strength decreases with distance from the transmitter, the C/I is continuously decreasing along the channel. A severe penalty in C/I must be paid as a consequence of the fading characteristics of the received signal. This penalty in C/I can be dramatically reduced by means of efficient modulation and coding techniques,

as discussed in Chapter 2.

To determine the C/I, *AWGN* must be accounted for. Two factors are considered to be contributing to the noise in the communication link, namely thermal noise $\mathcal{N}_T$ and the noise figure NF . For determining $\mathcal{N}_T$ see Appendix C. Noise figure relates the C/I at the input of a network to the C/I at the output of the network. All electronic components generate thermal noise, and hence contribute to the noise level seen at the detector output of a receiver. The NF adds a noise floor in order to refer the noise at the output of a receiver to an equivalent level at the receiver input terminals where C/I measurements are taken in effect [1, 18, 20]. Therefore the total noise contributed by the link, NT , is given as follows:

$$NT_{||dBm} = \mathcal{N}_{T||dBm} + NF_{||dB} \quad (\text{B.10})$$

Eq. B.4 is valid for computing both desired and interfering signal powers. Eq. B.9 is then used to compute the received C/I measurement at the receiver front end.

Maximum achievable C/I

The maximum C/I achievable in the subscribing receiver unit is limited by several sources, including inter-chip interference induced by the baseband pulse shaping waveform, the radio noise floor, ADC quantization error, and adjacent carrier interference [1]. In the system level simulation, the maximum C/I limitation can be characterized by the parameter χ , such that

$$\chi = \frac{1}{(C/I)_{max}} \quad (\text{B.11})$$

where, $(C/I)_{max}$ is the maximum achievable C/I (in Watts) assumed for the mobile receivers in the system. The effective C/I observed by the mobile receiving station shall be given by

$$C/I_{effective} = \frac{1}{\chi + \frac{1}{(C/I)_{inst}}} \quad (B.12)$$

where $(C/I)_{inst}$ is the instantaneous C/I at the receiver antenna, which is further limited by a $(C/I)_{max}$ ceiling. $(C/I)_{effective}$ is a scaled-down version of $(C/I)_{inst}$, and it accounts for the interference sources associated with the maximum C/I limitation. $(C/I)_{effective}$ is used as the C/I value observed by the mobile receiver station.

B.1.5 User Selection

Mobile users are dropped uniformly all over the 19 cell boundary. Then for each user the path loss from all 57 sectors is calculated. The link with the lowest path loss is chosen as the serving sector link. If this sector happens to be sector #1, the user is retained, otherwise the user is discarded. The next step to validate a retained user is by looking at its received pilot strength. The BTS assigns 12.5% of the total transmit power to the pilot signal, see [1]. The BTS average transmit power is fixed to a maximum value of 43 dBm. The pilot C/I received at the mobile station must not be lower than -5 dB. If the received pilot C/I from a sector is any lower, then the mobile is not to be served by that sector. The calculation of the received pilot from sector #1 is done by means of the following equation

$$Pilot\ C/I\ [mW] = \frac{(12.5\%)(Received\ power\ from\ sector\ \#1)}{Interference\ and\ Noise\ Power} \quad (B.13)$$

If the Pilot C/I from sector #1 satisfies the minimum pilot criterion of having a value equal to or greater than -5 dB, then the user is finally selected to be part of the system model, and its received C/I is accounted for in the probability distribution function (PDF) of the C/I's generated by this system.

The system parameter values used in the simulations, which also correspond to the 1xEVolution-Data/Voice (1xEV-DV) system [1], are listed in Table B.1. In the computer simulation approximately 100,000 users were successfully chosen as part of the C/I PDF. Fig. B.5 shows the actual positions of these selected users with respect to the center BTS sector #1 which is located at (0,0). It is interesting enough to note that this coverage area extends to the boundary of the layout in the direction of the serving sector antenna. Hence it is easily seen that the path loss is not only distance dependent, it is also affected by the path clutter and propagation conditions encountered on each individual transmission link. The generated C/I PDF is dependent on these parameters. The cumulative density function (CDF) of the user C/I's was obtained from the generated PDF, and is shown in Fig. B.6. For this CDF, a maximum achievable C/I of 20 dB has been used, making the scale parameter χ equal to 0.01. Appendix C contains a detailed explanation of how statistical random variables are generated to be used in the computation of lognormal shadowing $SH_{|dB}$ and Ricean fading $F_{|dB}$. These random variables effect the power received by the mobile unit on the forward link, see (B.4).

Figure B.5: BTS Sector #1 Coverage Area within 19 Hex Cell Boundary

B.1.6 Adaptive Communication System

Adaptive modulation and error control coding are well-known link adaptation techniques that have been applied to mobile wireless communication systems [2, 3, 4, 21]. These techniques, when designed to track the channel variations, yield a higher channel throughput as compared to fixed modulation and coding [5]. For complete details on Adaptive Coded Modulation (ACM), see Chapter 2. This section describes the system model which has been used for adaptive communication purposes in our

Figure B.6: Cumulative Density Function of User C/I's

research.

The translation of the user's information to a digital signal sent out by the BTS and then its reception and detection at the mobile subscriber unit can be represented by the block diagram in Fig. B.7. An essential component of the adaptive communication system is the feedback channel, shown in Fig. B.7. The feedback channel is used for updating the channel state information (CSI) made available to the BTS. The CSI determines the current fade level on the channel and it may be known to the receiver as well as the BTS. Knowledge of the CSI allows the transmitter to adapt $s(t)$, and the receiver to adapt its detection strategy, to the reported fade level.

Figure B.7: Adaptive Communication System Model

On the forward link a pilot tone made up of a channel sounding or training sequence, is continually transmitted so that the channel-induced envelope fluctuation $\alpha(t)$ can be extracted by the channel estimator at the receiver. Based on this channel gain estimated by $\hat{\alpha}(t)$, a decision device selects the constellation size to be used, configures the demodulator according to this constellation size, and finally informs the transmitter about the decision via the feedback path. The constellation size assignment has been discussed in more detail in Section 2.1. The BTS evaluates the feedback information that was sent by the mobile user and sends the payload with a coded modulation scheme that matches the current channel condition. The feedback information is assumed to be received correctly by the BTS. On the slow fading channel, the transmission system keeps its configuration unchanged for a duration

of τ_t sec. For satisfactory operation, the BTS modulator and the receiving unit demodulator must be configured for the same constellation size at any given instant. Efficient error control schemes are therefore required to insure error-free feedback communication. However such coding schemes inevitably introduce a certain time delay τ_f sec, which may include delays due to decoding and automatic repeat request (ARQ) as well propagation time delays on the feedback channel. Hence, even if perfect channel estimates α are available at the receiver, the system will not be able to adapt to the actual fading channel but rather at best to a τ_f sec delayed version of it. In practice, the adaptive scheme will operate with a delay τ such that $\tau_f \leq \tau \leq \tau_t$ [3]. The goal is to operate with the smallest possible τ_f in order to minimize feedback delay and the largest possible τ_t to minimize the rate of system reconfiguration or adaptation.

In this thesis, perfect channel estimates ($\hat{\alpha}(t) = \alpha(t)$) have been assumed at the receiver with zero delay τ_f in the feedback path. Researchers have analysed the effects of estimation error and feedback path delay on adaptive modulation, and it has been found that an estimation error less than 1 dB and a feedback delay of less than $0.001/f_D$ results in minimal performance degradation, where f_D is the Doppler frequency of the fading channel. The effect of estimation error and feedback delay for adaptive coded modulation is similar, yielding the same set of requirements for minimal performance degradation. These requirements are easily met on slowly varying channels [2, and the references therein].

Parameter	Value	Comments
Number of cells (3 sectored)	19	2 tier cell layout, 3-sector cell, 57 sectors
Antenna Horizontal Pattern	70 deg (-3 dB beamwidth) with 20 dB max attenuation	Refer Section B.1
Antenna Orientation	0 degree horizontal azimuth is East (main lobe)	No loss is assumed on the vertical azimuth
Propagation Model BTS Ant Ht=32m, MS Ant Ht=1.5m	$31.5 + 35\log_{10}(d)$ dB, d in metres	Modified Hata Urban (COST 231) Model @2 GHz Min. 35 metres separation between MS and BTS
BTS Site-to-site Distance	2.5 km	Separation between adjacent base stations
Log-Normal Shadowing	Standard Deviation = 8.9 dB	Independently generate shadowing per mobile See Appendix C
Mobile Noise Figure	10.0 dB	
Additive Thermal Noise Density	-174 dBm/Hz	See Appendix C
Carrier Frequency	2 GHz	
BTS Antenna Gain with Cable Loss	15 dB	17 dB BTS antenna gain; 2 dB cable loss
MS Antenna Gain	-1 dBi	
Fading Model	$Q = 3$	Ricean, See Appendix C
BS Maximum Transmit Power	43 dBm	20 W
Minimum Acceptable C/I	4 dB	
Maximum Achievable C/I	20 dB	See (B.12)
Transmission Bandwidth	1.25 MHz	

Table B.1: Forward Link System Parameters [1]

Appendix C

Random Variable Generators

The Box-Muller transformation [22] transforms random variables from a two-dimensional continuous uniform distribution to a two-dimensional Gaussian distribution. If X_1 and X_2 are uniformly and independently distributed between 0 and 1, then Z_1 and Z_2 as defined by

$$Z_1 = \sqrt{-2 \ln(X_1)} \cos(2\pi X_2) \quad (\text{C.1})$$

$$Z_2 = \sqrt{-2 \ln(X_1)} \sin(2\pi X_2) \quad (\text{C.2})$$

each have a Gaussian distribution with mean $\mu = 0$ and variance $\sigma^2 = 1$, and are statistically independent.

C.1 Log-normal Shadowing

The Log-normal distribution models the random *shadowing* effects which occur over a large number of measurement locations which have the same T-R separation, but have different levels of clutter along the propagation path. Simply put, log-normal shadowing implies that the measured signal levels (in dB) at a specific T-R separation have a Gaussian (normal) distribution about the distance-dependent mean of the path loss, where the measured signal levels have values in dBs. The standard

deviation σ of the Gaussian distribution that describes the shadowing also has units in dB. Eq. C.3 is used in our computer simulations to model shadowing for random locations in the communication system.

$$SH_{||dB} = \sigma \sqrt{-2 \ln(X_1)} \cos(2\pi X_2) \quad (C.3)$$

where X_1 and X_2 are independent random variables uniformly distributed over $[0,1]$.

C.2 Ricean Fading

The Ricean fading channel has been used in our system model. The Ricean distribution is often described in terms of a parameter Q , which is defined as the ratio between the deterministic or line-of-sight (LOS) signal power A^2 , and the variance of the component multipath complex Gaussian random variables σ_F^2 [18]. The parameter Q is known as the Ricean factor, and the higher the value of Q , the higher the dominant signal power. The following calculations were employed to include Ricean fading in the simulated received signal power given by (B.4).

$$Q = 10 \log_{10} \frac{A^2}{2\sigma_F^2} \text{ dB} \quad (C.4)$$

As $A \rightarrow 0$, $Q \rightarrow -\infty \text{ dB}$, the dominant path decreases in amplitude and the Ricean distribution degenerates to a Rayleigh distribution. It is well known that the envelope of the sum of two quadrature Gaussian noise signals obeys a Rayleigh distribution [18]. Let a random variable R_y be defined in terms of its component Gaussian random variables Z_1 and Z_2 , given by (C.1) and (C.2), each with variance σ_F^2 and zero mean.

$$R_y = Z_1 + iZ_2 \quad (C.5)$$

The square root of the sum of the square of the two Gaussian components then gives the envelope of the attenuated signal denoted by $|R_y|$.

$$|R_y| = (Z_1^2 + Z_2^2)^{1/2} \quad (\text{C.6})$$

This envelope $|R_y|$ has a Rayleigh distribution with $E[|R_y|^2] = 1$ (normalized power) [23]. A Ricean envelope would have the additional direct path (line-of-sight) component Q . Since Q is the power ratio of the direct path to the reflected paths and the reflected paths have normalized power, hence Q is equal to the power along the direct path.

$$R_F = \sqrt{Q} + Z_1 + iZ_2 \quad (\text{C.7})$$

Here, $|R_F|$ has a Ricean distribution with $E[|R_F|^2] = Q + 1$. Now normalize R_F so that $E[|R_F|^2] = 1$

$$R_{F_n} = \frac{\sqrt{Q} + Z_1 + iZ_2}{\sqrt{Q + 1}} \quad (\text{C.8})$$

The power attenuation in the received signal is given by $|R_{F_n}|^2$ such that

$$\begin{aligned} |R_{F_n}|^2 &= [\text{re}(R_{F_n})]^2 + [\text{imag}(R_{F_n})]^2 \\ &= \left(\frac{\sqrt{Q} + Z_1}{\sqrt{Q + 1}} \right)^2 + \left(\frac{Z_2}{\sqrt{Q + 1}} \right)^2 \\ F_{||dB} &= 10 \log_{10}(|R_{F_n}|^2) \end{aligned} \quad (\text{C.9})$$

where $F_{||dB}$ is the Ricean fading component of the received power in dB, see (B.4).

C.3 Thermal Noise Power

The internally created AWGN is also termed as *thermal noise*. It is caused by the thermal motion of electrons in all conductors, and is generated in the lossy coupling

between an antenna and receiver and the first stages of the receiver. The noise spectral density is constant at all frequencies up to about 10^{12} Hz, such that thermal noise is given the name *white noise*. The calculations for the thermal noise power are as follows, [20].

$$\mathcal{N}_T = \text{the thermal noise}$$

$$= N_0 W \text{ [watts]}$$

$$N_0 = \text{the single - sided thermal noise power spectral density}$$

$$= \kappa T \text{ [watts/Hz]}$$

$$\kappa = \text{Boltzmann's constant}$$

$$= 1.38 \times 10^{-23} \text{ W/Kelvin - Hz} = -228.6 \text{ dBW/Kelvin - Hz}$$

$$T = \text{the thermal noise temperature}$$

$$= 290 \text{ [Kelvin]}$$

$$W = \text{the transmission bandwidth, in Hz}$$

Thermal noise in dBm is given by

$$\mathcal{N}_{T|dBm} = 10 \log_{10} \left(\frac{\mathcal{N}_T}{0.001} \right) \quad (\text{C.10})$$

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