DESIGN AND ANALYSIS OF A LAND MOBILE CELLULAR RADIO SYSTEM UNDER THE EFFECTS OF INTERFERENCE

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1984

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

Department of Systems and Computer Engineering

Carleton University
Ottawa, Ontario
June, 1984

ACKNOWLEDGEMENTS

I offer sincere thanks to my thesis supervisor, Dr. S.A. Mahmound, for his technical guidance, enthusiastic support and encouragement. He has created an atmosphere which facilitates productive research and solution of problems through innovative approaches.

I also thank the Department of Communications for granting me educational leave and financial support.

Finally, I thank my wife, Ada, for her patience, understanding and support, over the past few years, without which it would not have been possible for this study to be successfully completed on a part-time basis.

ABSTRACT

This dissertation is concerned with the design and analysis of a Land Mobile cellular radio system. Two problems are examined. The first is the investigation of the impact of interference in land mobile channels on a cellular radio system. The second one is the design of a spectrum-efficient frequency allocation scheme.

The effects of fading and shadowing of the mobile channel on the required transmitter output power are studied and a model for obtaining the output power for a given cell radius is derived. This tool is later used to analyse the adjacent channel and intermodulation interference encountered in a cellular system. In studying the capacity of the cellular system, the number of voice and digital control channels required are obtained as a function of the cell radius. Special attention has been given to the influences of the Pure Aloha random access scheme and channel fading on the probability of successful reception of data packets over the control channel.

The interference implications in a land mobile radio channel are examined. The co-channel interference, being the most important source of interference encountered in a cellular system, is studied under different conditions of fading and shadowing. Results have indicated that the corner-illuminated

configuration suffers less co-channel interference and requires a smaller reuse distance than the centre-illuminated configuration. The adjacent channel, as well as transmitter and receiver intermodulation interference in a cellular system are also analysed. In addition, the two cases of intra-cell and inter-cell interference on both uplink and downlink channels are considered. Numerical results show that inter-cell interference may be ignored while intra-cell interference should be avoided by proper selection of frequency-cell assignment.

The problem of designing a novel frequency allocation scheme which would provide an interference-free operation and a maximum spectrum efficiency is then addressed. The algorithm developed receives as input parameters, the number of cells in a cluster, the number of channels in a cell and the minimum separation between channels, and generates the required frequency allocation plan. Two sub-optimal schemes are also proposed to complement the first scheme in cases where the first scheme fails to provide solutions. Numerical results for various input values are presented.

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GLOSSARY OF SYMBOLS

Chapter 1

m	=	number of radio channels
n	=	number of users
t	=	traffic load per user
p	=	blocking probability of one channel
P_b	=	blocking probability of all trunked channels
T	=	offered traffic in trunked system

Chapter 2

U	=	co-channel reuse distance
D_c	=	distance between two co-channel base stations
r	=	cell radius
N	=	number of cells in a cluster
g,h	=	shift parameters
X	=	power level of received signal
X_o	=	local mean power of signal
$p(X)$	=	probability density function of X
$p(X_o)$	=	probability density function of X_o
X_m	=	median received power level
X_{md}	=	X_m in dB
σ	=	standard deviation of X_o

X_t	=	threshold power level
X_{td}	=	X_t in dB
P_f	=	failure rate
P_{fd}	=	P_f in dB
P_t	=	base station transmitter output power
G_t	=	base station transmitter antenna gain in excess of circuit losses
G_r	=	mobile station receiver antenna gain in excess of circuit losses
P_l	=	path loss
d	=	distance
f	=	frequency
H_r	=	receiver antenna height
H_t	=	transmitter antenna height
t	=	traffic load per mobile subscriber
V	=	mobile density
P_b	=	probability of blocking
T	=	offered traffic
m	=	number of available channels
B	=	average level crossing rate
F_o	=	average fade duration
Z_o	=	average signal to noise ratio
Z	=	signal to noise ratio
f_d	=	maximum Doppler frequency shift
V	=	vehicle speed
w	=	wavelength of carrier frequency
I_o	=	average interfade duration

J_o = average non-fade duration

J = non-fade duration

P_1 = probability that the first bit of the packet is in a non-fade duration

P_2 = probability that the remaining bits are in non-fade duration

P_3 = probability that all the bits in the non-fade duration are received correctly

P_c = probability of correct reception of a packet

M = number of bits in the packet

b = bit duration

$P_{eo}(Z_o)$ = average bit error rate in a non-fade duration

$P_e(Z)$ = bit error rate

S = system throughput

G = offered channel traffic

λ = number of calls made per second

τ = packet duration

E = spectrum efficiency

L = number of cells in the service area

A = area of service area

C = number of different clusters in the system

f_s = channel bandwidth

$n(T,P_b)$ = number of channels per cell

R = radius of service area

K = $10/\ln 10$

Chapter 3

X	=	power level of received signal
X_t	=	threshold power level
Z'	=	sum of the received power levels of six interfering co-channel signals
p	=	inverse power law index
Y_i	=	power level of an interfering signal
q	=	protection ratio
Q	=	q in dB
K	=	propagation constant
D	=	distance between desired base station and mobile
n	=	number of co-channel interferors
U	=	co-channel reuse distance
U_{min}	=	minimum value of U required
Y_o	=	mean power level of interfering signal
P_{F1}	=	probability of co-channel interference under fading with one interferor
X_o	=	mean power level of desired signal
Z	=	sum of the power levels of n interfering co-channel signals
$f(Z)$	=	probability density function of Z
$P(Y_i)$	=	probability density function of Y_i
$f(s)$	=	Laplace Transform of f(Z)
$P(s)$	=	Laplace Transform of $P(Y_i)$
P_{Fn}	=	Probability of co-channel interference under fading with n interferors

$P_{Fn,k}$ = k^{th} term of P_{Fn}

X_d, Y_d = X_o, Y_o in dB

X_m, Y_m, Z_m = median power levels of X, Y, Z

X_{md}, Y_{md}, Z_{md} = X_m, Y_m, Z_m in dB

σ_X, σ_Y = σ = standard deviation of X_d, Y_d

c = correlation coefficient

$f(X_d, Y_d)$ = joint probability density functions of X_d and Y_d

$f(X_d)$, $f(Y_d)$ = probability density function of X_d, Y_d

P_{Bn} = probability of co-channel interference under fading and shadowing

$P_{Bn,k}$ = k^{th} term of P_{Bn}

R = ratio of X_m to Y_m

P_{S1},P_{Sn} = probability of co-channel interference under shadowing with one interferor, with n interferors

L = lower limit of integration for P_{Sn}

U_c = co-channel reuse distance for corner-illuminated configuration

H = constant for the spectrum efficiency expression

V = no. of vehicles per Sq. Km

t = traffic load per mobile user

M = no. of channels per cluster required in the corner-illuminated configuration

Chapter 4

f_s	=	channel spacing
m	=	no. of channels per cell
N	=	number of cells per cluster
X, Y	=	power level of desired signal, interfering signal
Q	=	protection ratio
J	=	amount that the interfering adjacent channel signal is below its carrier
P_d, P_{adj}	=	power level of received desired signal, received adjacent channel signal
P_m	=	mobile station transmitter output power
P_{1d}, P_{1i}	=	path loss of desired signal, interfering signal
n	=	number of channels allocated to the base station
r	=	cell radius
d	=	distance
P_a	=	probability that one of (n-1) channels is active within interference area
V	=	mobile density
t	=	traffic load per mobile subscriber
P_{AD}	=	probability of adjacent channel interference
f_i	=	carrier frequency
f_{IM}	=	frequency of IM product
a	=	positive integer coefficient
S	=	frequency separation
S_{max}	=	maximum value of S

$P(S)$ = distribution of IM products with frequency separation S

T, T_{av} = total, average number of IM products that fall in the band

L = lowest channel number such that frequencies on the lower side of the channel would form an IM product to fall on it

U = highest channel number such that frequencies on the upper side of the channel would form an IM product to fall on it

T_o, T_e = T corresponding to odd, even values of n

$T_{o,av}$, $T_{e,av}$ = T_{av} corresponding to odd, even values of n

P_{ti} = output power of TX IM product

B = coupling loss

C = TIM conversion loss

G_t, G_r = transmitter, receiver antenna gain

θ = vertical angle relative to antennae position

P_{ti} = probability that two transmitters are active in the formation of a TIM product to cause interference

P_{tim} = probability of TIM interference

$P_{o,tim}$, $P_{e,tim}$ = P_{tim} corresponding to the odd and even values of n

P_{ri} = received power level RX IM product

K = RIM conversion loss

P_N, P_F = power level of received signal from near, far interfering transmitters

df = frequency separation between near and far transmitter frequencies

P_{mt} = mobile transmitter power

G_{mt} = mobile transmitter antenna gain in excess of circuit losses

G_{br} = base receiver antenna gain in excess of circuit losses

P_{1N}, P_{1F}, P_{1d} = path loss from near transmitter signal, far transmitter signal, desired signal

X_{td} = threshold power level

P_{fd} = failure rate

= standard deviation of P_N, P_F, P_d

K_{im} = RIM criterion

d_N, d_F = distance of near transmitter, far transmitter from base station receiver

$P_{im}(S)$ = probability of RIM interference for a given frequency separation of S channels

P_{rim} = probability of RIM interference

$P_{ri}(S)$ = probability that two transmitters are active in the formation of a RIM product to cause interference

Chapter 5

N	=	number of cells in a cluster
D	=	frequency separation
m	=	number of channels per cell
f_i, f_x, f_y, f_z	=	carrier frequencies
i	=	channel number
f_s	=	channel bandwidth
J	=	cell number
C_{iJ}	=	channel number in the row i and cell J
S	=	shift
h	=	difference in channel spacing between 2 rows
$W_j(h)$	=	relative possible shift of row j with respect to row i
S_i	=	shift of row i
A_j	=	actual possible shift of row j
D_{ji}	=	minimum separation between rows i and j
L_k	=	list of relative possible shifts of row k
$W_j = W_j(1)$	=	relative possible shift of row j with respect to row j-1
W	=	relative possible shift for any row
M	=	maximum number of channels at which the optimal frequency allocation algorithm can find a solution
H_k	=	dimension of the list of rejected channels for row k

T_k = total number of rejected channels from row M to and including row k

"A" = addition

"M" = multiplication

"C" = comparison

"O" = number of operations

Chapter 6

σ = standard deviation of received power level

CHAPTER 1 - INTRODUCTION

1.1 Background

Eighty-nine years have passed since the invention of radio telegraph by Marconi in 1895. Today, our daily life benefits to a great extent by the utilization of radio waves. In its early stages of development, radio was essentially used by ships in what later became known as the Maritime Mobile Service. Since then, it has proliferated in many directions leading the way to the definition of a multitude of so-called radio services. The explosion of the radio use in all spheres of human activity is witnessed by the tremendous increase in the world total of radio stations from about 14,000 sixty years ago to at least 20 million today.

The steady expansion of radio communications, in particular of the mobile radio service, requires guidelines and regulations in order to make efficient use of the radio spectrum which is a scarce resource. Hence, spectrum management has become the art of adequately blending administrative and technical procedures to ensure an interference-free and equitable use of radio communication services. It is also gradually becoming a complex task requiring advanced and methodical approaches.

Over the past few decades, the problems associated with the management of the radio spectrum have been increasing rapidly. The ever increasing number of radio users, especially in the land mobile bands, coupled with the interference between operating radio frequencies has caused a shortage of assignable frequencies in the allocated bands.

In response to the land mobile demand for increased levels of usage of the spectrum, new frequency bands have been allocated (or reallocated) for mobile use in the 1979 World Administrative Radio Conference (WARC) of the International Telecommunications Union (ITU). In Canada and the U.S., the frequency band 806-890 MHz has been allocated in recent years for mobile communications. While the new band seems to be able to satisfy current needs, further shortage of the spectrum would soon occur again. In order to prevent this, modern engineering techniques must be applied to develop new radio systems, to increase efficient utilization of the spectrum and minimize radio interference.

The work described in this dissertation is directly relevant to a specific portion of the radio spectrum, namely, the land mobile bands for which new spectrum-efficient concepts are presented.

1.2 Solutions to the Spectrum Congestion Problem

The radio spectrum may be viewed as an entity with three dimensions: frequency, space and time. As such, frequencies may be assigned in one of the three methods:

(1) Frequency sharing in which the same frequency may be used at the same location but at different times;

(2) Multiple frequency assignment in which different frequencies may be used simultaneously at the same location;

(3) Frequency reuse in which the same frequency may be used simultaneously but at different locations.

Over the past decade or so, research and development efforts on new technologies and techniques have been in active progress with the intention of coming up with new spectrum-efficient systems. The following are some of the underlying system concepts behind current approaches for solving the spectrum congestion problem.

1.2.1 The Frequency Sharing Method

This method calls for the time-sharing of a given portion of the spectrum between a number of users at the same location. In order to increase the usage of the radio spectrum over a given period of time, the sharing mechanism must be efficient. The method is also based on the assumption that no single user would occupy the whole spectrum 100% of the time. Efficient spectrum sharing techniques include:

- Trunking
- Packet Radio
- Data and Voice Integration

1.2.1.1 Trunking

Trunking may be defined as an automatic method of temporarily assigning radio communication channels to users, dynamically, from a central pool of channels.

The trunking technique gains its merit of high spectrum efficiency by allowing the total available resources to be shared between all the users. This can be illustrated in a simple example. Suppose m radio channels are evenly assigned to n users with $n > m$ and each single user generates a traffic load of t Erlangs. If the user has access to only one channel, the probability that the channel is blocked is p where $p = nt/m$ which is also the amount of traffic carried on each channel.

On the contrary, if all the m channels are shared between all the n users, the probability of blocking may be obtained from the Erlang B expression [KLEI75A]:

$$P_b = \frac{T^m}{m!} \Big/ \sum_{i=1}^{m} \frac{T^i}{i!} \qquad \text{..... (1.1)}$$

where P_b is the blocking probability and T is the total traffic offered by all the users in Erlangs.

Keeping the blocking probability constant, i.e. $P_b = p$, T may be found from the Traffic Tables [FRAN76]. In the example where $n = 30$, $t = 0.01$ Erlangs and $p = 0.05$, m has to be equal to 6 resulting in $T = 2.96$ Erlangs. It can be seen that the amount of traffic carried in the m channels has increased with a ten-fold gain from 0.3 with no trunking to about 3 with trunking. In general, the advantages of trunking are greatest when the channels in question are lightly loaded. As loading increases, the gain per channel levels off.

While the trunking technique is spectrum-efficient, it requires more complexed circuitry and extra control channels to allocate and supervise channel assignments. This technique is also based on the assumption that:

(a) the average message over the system is brief and

(b) the probability that many stations will need to communicate simultaneously is small.

1.2.1.2 Packet Radio

The single most important factor that determines how well a mobile radio channel is shared among its data users is the scheme that is used to gain access to it. Conventional fixed multiple access schemes such as frequency division multiple access and time division multiple access methods or some kind of polling scheme could be used to share the channel among the users. However, these are wasteful of bandwidth and under certain conditions polling may require additional system complexity and delay.

A number of random access schemes have been proposed for packet radio communication systems. They are attractive because all these schemes suggest an efficient sharing of the spectrum resource. Using these schemes, more users could be accommodated on the same channel resulting in a better utilization of the channel.

The four widely-known random access schemes are Pure Aloha, Slotted Aloha, Carrier Sense Multiple Access (CSMA) and Reservation.

In the Pure Aloha scheme as described in [ABRA77], each user may randomly access the channel and transmit the data packets. Packets may collide and destroy each other. Loss of packets may be avoided by retransmitting the packet if an acknowledgement message is not received after a certain time-out period. While having the virtue of simplicity, the maximum theoretical channel utilization or throughput of this scheme is only 0.184.

Throughput of the system can be improved if a time base is established for the transmission of the packets. This is accomplished by dividing the channel into time slots of length equal to the packet transmission time. All stations have to transmit packets in the beginning of the time slot and the probability of packet collision would be reduced to half of that of the Pure Aloha scheme. This technique is termed Slotted Aloha and has a maximum throughput of 0.368 [ABRA77].

In the above two schemes, the system throughput is limited by the fact that data packets collide and destroy each other. If the channel can be sensed before a data packet is transmitted, the probability of collision would be minimized and the throughput increased. Kleinrock and Tobagi [KLEI75B] found that the capacity of such a Carrier Sense Multiple Access (CSMA) channel depends on the propagation delay and for large delays, the performance of a CSMA system could be even worse than that of a Pure Aloha system.

The fourth well-known access scheme is the Reservation Scheme in which a number of variations exist. Essentially, these schemes were designed in some optimal manner to satisfy a particular environment or requirement.

1.2.1.3 Data and Voice Integration

In a normal two-way telephone conversation, the transmission of voice signals is very inefficient because of the intra- and inter- voice gaps. It was reported by Bullington and Fraser [BULL59] that the telephone circuit is busy no more than 35 to 40 % of the time.

The utilizaton of the telephone circuits may be improved by TASI (Time Assignment Speech Interpolation) which is a high-speed transmission and switching system based on the principle of using the free channel time to interpolate additional talkers. The concept has been applied successfully in international submarine cable and satellite systems.

TASI [BULL59] is a technique in which the idle time between telephone calls and the conversation pauses during calls is used to accommodate additional calls. With a sufficiently large number of channels, most of the idle time on the transmission link can be filled, giving an enhancement in transmission capacity greater than two.

TASI exploits the low speech spurt activity by assigning transmission channels only when a speech spurt is present. It is evident that the process becomes more efficient as the number of channels increases. When a large number of independent conversations compete for some smaller number of channels, there is a finite probability that the number of conversations demanding service will exceed the number of available channels. This competition manifests itself in the form of clipping of the initial portion of a speech spurt. The percentage of time that speech is lost due to such competition is called the percent freeze-out. The effect of the freeze-out on transmission quality is negligible as long as the percent freeze-out is less than 0.5 percent [BULL58].

The concept of TASI may be adopted in land mobile radio systems. Since a speech pattern is composed of a series of talkspurts interspersed by a number of pauses, digital data can be transmitted in the gaps allowing operation of more mobile radios per channel and hence can provide a more efficient use of the spectrum.

Based on the assumption that the probability density function of both a talkspurt length and a pause length may be represented by exponential functions, daSilva [DASI81] studied the possibility of integrating voice and data packets in the same channel. They analysed the performance of such a system by assuming that the arrival processes of voice calls and data packets arrivals are Poisson processes with different mean arrival rates. Their simulation results show that the number of data sources that can be supported on a single integrated voice-data radio channel is in the order of a few hundred users.

Some other workers, Marsan and Pent, described in [MARS80] a mixed voice and data mobile radio system using two types of multiple random access protocols: Pure ALOHA and CSMA. They have also derived data throughput expressions for both protocols.

An integrated voice/data system for land mobile radio was recently proposed by Mahmoud et al [MAHM83]. This system makes use of the natural pauses in conversational speech by interleaving data packets and talkspurts from different voice sources. A speech detector, designed specifically for the land mobile environment, is used to suppress the mobile transmitter carrier during the silence gaps, so that the channel is freed for other uses. Data packets are transmitted using either the Pure Aloha or slotted Aloha random access schemes. Analysis shows that such a system can result in bandwidth reduction of 30-35% relative to conventional radio telephony. Simulation results indicate that the volume of data traffic that can be transmitted during the silent gaps is limited by the requirement to reduce clicks in the speech resulting from collisions with data packets.

It can be seen from the results of the above studies that voice and data integration would lead to improved utilization of the mobile channel and higher spectrum efficiency of land mobile radio systems.

1.2.2 The Multiple Frequency Assignment Method

In this method, a certain band of spectrum is made available for simultaneous usage to users located in the same area. Because of co-channel interference, no two users are normally allowed to use the same radio frequency. The objective of this method is therefore to maximize the number of users served by the given band. This can be accomplished by two entirely opposite methods. The first one is the conventional method of reducing the channel bandwidth hence increasing the number of available channels. The widely used techniques belonging to this method are the Voice Digitization Techniques and the Single Side-Band Modulation Techniques. The second method is the spread spectrum techniques which employ specific encoding techniques to multiplex users in the same area on the undivided band of frequencies. The following is a summary description of:

- Voice Digitization Techniques
- Single Side-Band Modulation Techniques
- Spread Spectrum Techniques

1.2.2.1 Voice Digitization Techniques

The primary purpose of digital voice systems is to provide secure, unintercepted voice messages and compatibility with digital networks. Many systems implemented in the field require a bandwidth at least equal to the bandwidth necessary for analog voice systems. These digital voice systems do not provide spectrum savings unless their bandwidth requirement is reduced. The improvement in digital voice encoding and decoding systems over the past five years has however led to their incorporation in land mobile transmitters and receivers. New promising voice encoding techniques with a smaller bandwidth requirement may indeed be economically feasible for land mobile communications in the near future.

There are basically two general categories of voice digitization techniques [OCCH78]. They are the less expensive, high digitization rate, waveform reconstruction technique and the more expensive, low digitization rate speech analysis synthesis technique.

Waveform reconstruction is a straight-forward reproduction of the acoustic time waveform by means of discrete-time, discrete-amplitude representations. There are two classes of modulation in this technique: the pulse code modulation (PCM), and the delta modulation (DM). In PCM, the speech signal is sampled at a rate equal to at least twice the highest frequency of the signal and each sample value is coded into an 8-bit word. This system is not very efficient, typically requiring a bandwidth of 64 KHz, but provides high quality speech at relatively low cost.

In DM, the voice signal is sampled at a rate much higher than the highest frequency of the signal so as to increase the adjacent sample correlation. Each sample is compared with the previous one. In the simplest arrangement, a logic ´1´ is transmitted if there is an increase in level and a logic ´0´ is transmitted if there is a decrease in level. The bit rate of this class of modulation is about 10-12 KHz and up.

The speech analysis-synthesis technique is different from the waveform reconstruction technique in that the exact shape of the orignal waveform is not preserved. The output speech would however sound like the original input speech. In this technique, the voice encoder, or vocoder, analyses and digitizes the basic speech parameters: voice or unvoiced sound, pitch, sound intensity, spectral envelope, etc. The destination vocoder can then synthesize a replica of the input speech based on the received speech parameter values.

A number of different vocoding strategies exist: channel vocoder, cepstrum and formant vocoders and the linear predictive vocoder with the voice digitization rate ranging from 0.6 Kbits per second to about 9.6 Kbits per second. The major difference between these vocoders is in the manner in which the basic speech parameters are extracted.

The selection of the voice digitization strategy between the waveform reconstruction technique and the speech synthesis analysis technique is a trade-off between cost, quality and transmission bandwidth. As the technology improves and the cost of analog-to-digital and digital-to-analog conversion hardware declines, low voice digitization rate will become realizable at an acceptable cost and quality in the future.

With lower voice digitization rates, the bandwidth of the channel for digital voice communications may be reduced, hence increasing the number of channels in a given frequency band. If interference between the digital signals in the narrower channels can be minimized, more users can be accommodated in the spectrum hence increasing its efficiency.

1.2.2.2 Single Side-Band (SSB) Modulation Techniques

Single Side-Band (SSB) has been used for almost all speech channels in the HF band (2-30 MHz) for many years. However, this technique has been largely ignored for mobile radio in the VHF and UHF bands because of problems such as frequency stability and the rapid fading of the signal at these frequencies. Only until recently has this technique been studied by various parties for application in the land mobile bands. It has been claimed that SSB systems with a 5 KHz bandwidth has a five-fold advantage in terms of spectrum efficiency over conventional 25 KHz and 30 KHz land mobile FM channels [HERR83]. The number of available channels can therefore be significantly increased.

SSB takes several forms. The type used at HF is suppressed carrier SSB where only a single modulation side-band is transmitted. In applications in the VHF and UHF bands, a pilot signal is transmitted to overcome the automatic gain control (AGC) and automatic frequency control (AFC) problems. The various forms of SSB are: pilot carrier; in-band pilot tone; above-band pilot tone, analog Lincompex and digital Syncompex. The first three systems have roughly similar performance and may be generally categorized as pilot SSB. The last two provide amplitude compression and expansion and are referred to as amplitude compandored single side-band (ACSB) systems.

The Lincompex (Linked Compressor and Expander) is a speech processing system which is designed to preserve maximum clarity of speech under poor signal-to-noise conditions. The system is used over HF radio circuits and uses a rapid-acting 'compressor' to maintain constant output, even at syllabic rate, to drive the transmitter. To link the action of the transmitter compressor and the compensating receiver expander, a control tone is added to the speech channel. This tone is at a few dBs below the constant-level speech and at a frequency which varies over a small range (120 Hz) above the band used by the speech signal. The frequency is a function of the compressing ratio and provides the receiver with the necessary information to expand the signal.

The Syncompex (Synchronous Compressor and Expander) system is the digital version of Linkcompex. Instead of transmitting a frequency tone, a digitally modulated signal is transmitted to carry the compressing information [HAFE84].

One important concern with the development of SSB systems in the land mobile bands is the determination of the interference criteria (i.e. SSB-to-SSB and SSB-to-FM). This leads to the questions such as: what separation distances are required between

FM and SSB systems and between SSB systems? What is the improvement in spectrum efficiency of SSB systems with considerations given to the required distance and frequency separations? Active research has been going on in different areas in SSB systems and so far it has indicated that the introduction of SSB systems into the land mobile bands would provide relief to spectrum congestion.

1.2.2.3 Spread Spectrum Techniques

The Spread Spectrum Technique is considered to be different from the conventional frequency assignment technique. Instead of improving spectrum efficiency by reducing the necessary bandwidth and keeping the transmission path clean by eliminating unnecessary and spurious radiations, spread spectrum spreads the transmitted bandwidth to hundreds of times the data bandwidth. According to the Shannon-Hartley theorem, the signal-to-noise ratio can be greatly reduced.

There are basically two classes of spread spectrum techniques: pseudo-noise (PN) modulation and frequency hopped (FH) modulation.

In PN modulation, the data signal is multiplied by a binary pseudo-random sequence which has a symbol rate or chip rate many times, say n, the binary data bit rate. The effect of PN sequence modulating the data stream is to increase the digital rate going into the PSK modulator. Consequently, the occupied RF bandwidth of the resulting waveform is increased by a factor of n.

In FH modulation, the carrier frequency pseudo-randomly hops in discrete increments among n frequencies as determined by a suitable algorithm. The spreading signal remains at a given carrier frequency for only one bit or for several bits.

The principal feature of a spread spectrum system is that more users could be accommodated. Each user would have a discrete access code which would be used during either transmitting or receiving. This code determines the pseudo-random sequence in PN modulation or the hopping sequence in FH modulation. In this technique, separation in the three dimensions of frequency, time or space is not required. Since more users could be accommodated in the given band, spread spectrum is claimed to be more spectrum-efficient than conventional narrow band systems [COOP79], [NETT80].

In addition, the spread spectrum techniques also provide a number of other advantages. Some of these are:

- capability of reducing interference
- capability of reducing degradation due to fading because of its use of a broad frequency band which effectively provides frequency diversity
- random access feature to enable users to initiate calls without waiting for a free channel

- privacy in communications

1.2.3 The Frequency Reuse Method

The third method to assign frequencies for land mobile use is to reuse or reassign the same frequency at a distance far enough away from the first frequency assignment so that the co-channel interference at one site caused by the other is at an acceptable level. Because of its frequency reuse capability, cellular radio has generally been recognized as a spectrum-efficient technique which allows many more simultaneous calls per frequency slot than present systems within a given service area.

1.2.3.1 Cellular Radio Systems

The cellular radio system is different from the conventional land mobile system in a number of ways. In conventional systems, single high power transmitters are usually used to cover the entire service area with a coverage distance of 30-50 km. This inhibits the reuse of the same frequency for a great distance and a large number of frequencies would be needed to cover the entire service area. The conventional system is therefore spectrally inefficient to operate.

The cellular system improves the situation by segregating the service area into a number of non-overlapping areas or cells with a base station located either at the centre of each cell or at alternate cell corners. Low power transmitters are used since the cell radius is only a fraction of the original coverage distance. The same frequencies may therefore be reused within the same service area resulting in a smaller number of frequencies required, and hence a higher spectrum efficiency.

However, in return for the opportunity of large capacity and high spectrum efficiency, the design of cellular system is more technically complex and expensive, requiring large amounts of spectrum to make it economically viable.

The cellular system has been studied by a large number of workers over the past few years [BSTJ79], [ECLJ77]. As frequency assignment is one of the most important topics in the design of cellular systems, numerous workers have studied this topic in depth using simulation methods. A brief survey of this work shows that the frequency assignment schemes have only been analysed with respect to traffic loading and co-channel interference considerations but no reference has been made to other forms of interference such as adjacent channel or intermodulation interference. This survey is provided in Appendix A.

A more comprehensive description of the land mobile cellular radio system is given in Section 2.2 in the next chapter.

1.3 Commercial Cellular Radio Systems and Future Developments:

The mobile telephone, used by subscribers in cars or trucks, has been generally considered as a limited service. Spectral crowding has prohibited expansion of the service. But cellular mobile radio telecommunications are changing this situation and with this new technology, several hundred thousand users can be offered better service than is now available for hundreds.

In the U.S., the AMPS (Advanced Mobile Phone Service) system began its field trial in Chicago in 1978 with a 10-cell system and 2000 customers covering an area of 5,400 sq. km. Another cellular system which is a Motorola-designed DYNA TAC system (dynamic adaptive total area coverage) was tested in the Washington/Baltimore area at about the same time to provide a full-scale demonstration of the viability of portable and mobile radiotelephones in a single system. Both systems have been reported to have successfully demonstrated the usage of cellular telephones in commercial service.

In Japan, a high capacity land mobile cellular radiotelephone system was put into service in 1979 in the Tokyo area. Since then, the service area has been extended to other major cities. In 1982, there were more than 13,000 subscribers and the total service area covered was about 8,500 sq. km.

In Europe, the Nordic countries have jointly developed a cellular mobile telephone system for commercial operation in Denmark, Norway, Finland and Sweden. The system is operating in the 400 MHz frequency range with up to 440 radio channels. On the other hand, Britain has selected an advanced version of the American AMPS standard known as TACS (Total Access Communications System). This system was thought to be capable of handling densely populated areas better than the Nordic system.

In Canada, July 1, 1983 saw the official inauguration of a 450 MHz Aurora (Automatic Roaming Radio) system in Edmonton. Aurora may be called a 'quasi-cellular' system as it does not incorporate sophisticated 'hand-off' algorithms (or techniques used to change the channels when the mobiles move into a neighbouring cell). It however provides a running test-bed for the next phase of the true 800-MHz cellular system. Furthermore, the Department of Communications has recently announced the award of the national cellular radio licences to the Telephone Common Carriers and CANTEL to develop the world's largest national cellular mobile radio telephone service across Canada. Twenty-three communities will be served by cellular radio beginning in 1985 with full implementation spreading over about 30 months.

From the above, it can be seen that cellular systems are slowly coming out of a prototype testing phase and entering into a full commercialised phase. This phase is expected to be succeeded by two major developments. The first development is portable radio and the second is Mobile Satellite Service (MSAT).

(1) Portable Radio: Due to the advanced development of LSI technology in recent years, the manufacturing of miniaturized transceivers and 1000-frequency synthesizers for consumer user is a reality. Office equipment manufacturers in the near future will be pressuring to extend the 800-MHz cellular radio techniques to cordless telephones. Such a concept is also a recognition of and response to the changing mobility pattern which places less emphasis on the private automobile. Portable radio telephone complements the service to the subscriber who is on the move by either private or public transportation.

To be portable, the unit must be easy to carry, rugged and weather resistant. Key items are size and weight. Low current CMOS circuits and custom integrated circuits and hybrid technology will be extensively used. Due to power limitation, the portable unit has to be voice activated. Furthermore, because of shielding effect when the portable cellular phone is brought into an office building, hand-off and location algorithms of a portable system are much more complex than those of a mobile-only system.

(2) Mobile Satellite Service: Currently, high capacity cellular radio systems have been the prime target for development and implementation for large metropolitan centres. Areas in between

these centres have been largely ignored because of the low population density (the Canadian Northern Territories is a good example). While it may not be economically advisable to install terrestrial systems to serve thinly populated areas, satellites appear to offer a cost-effective means to serve these areas. It is highly desirable that mobile stations are equipped with equipment which is compatible to both systems for economic reasons [NICH83], [GLEN83]. Indeed, a combined terrestrial and satellite system would provide an ubiquitous mobile telephone service to users in all areas.

This satellite-aided system would provide frequency reuse by having many separate antenna beams, each with its own transmitter-receiver. The footprint of each antenna beam would serve a cell on a nationwide basis just as each fixed transmitter-receiver serves a cell in an urban system. This direct link to the satellite from mobile transceivers will open the door for future use of pocket-sized mobile telephones in far-flung areas where communication services are lacking.

1.4 Objectives of the Research

In the previous sections, we have summarized the effort of a number of feasible technologies and system concepts that can be used to solve the spectrum congestion problem. We have also discussed the trends of development of the commercial cellular radiotelephone systems which have received increasing attention over the past few years.

Among these system concepts and technologies, trunking is currently being used and is recognized as a spectrum-efficient technique with some limitations. Packet radio, data and voice integration and digitized voice techniques are well established concepts. Single Side-Band Modulation and Spread Spectrum techniques are fairly well understood and are presently subjects of active research and development.

Despite the proliferation of cellular radio systems, two technical areas in the performance of such systems remain largely unresolved and could benefit from further investigation. First, the effects of multi-path fading, shadowing, multi-source co-channel, adjacent channel and intermodulation interference on a cellular system remain poorly understood. Second, even though a large volume of work has been devoted to searching for a better frequency assignment scheme to handle varying traffic loads, as described in Appendix A, very little has been accomplished in designing a spectrum-efficient interference-free frequency allocation scheme for the cellular system.

Based on the above discussions, we have formulated the following two main objectives of this dissertation:

(1) To provide a better understanding of the impact of

interference in land mobile channels on a cellular radio system;

(2) To investigate and design a spectrum-efficient frequency allocation scheme which will improve the performance by providing interference-free operations in the cellular system.

The first objective is dealt with in Chapters 2, 3 and 4 in which a number of models for predicting interference are described. This provides the necessary knowledge and background that will be used to address the second objective, namely, the design of a frequency allocation scheme. The details of such a scheme are presented in Chapter 5.

It is worth noting that the complete design of a cellular mobile radio system involves considerations in many other different areas. Examples of these major areas include: the design of the hardware interface between the network of base stations and the telephone switching network, the communications protocol between the base and mobile stations, the design of the call sequences, the switching of information and control messages between the telephone network and the mobile stations through the base stations, the optimal equipment selection, etc. This dissertation will be limited to the design and analysis of a land mobile cellular radio system under the effects of interference, fading and shadowing and the design of a frequency allocation scheme which will solve the interference problems.

1.5 Thesis Contributions

The major contributions of this dissertation are outlined in the following:

- Determination of Co-channel Reuse Distance:

A detailed analysis of the multiple-source co-channel interference encountered in a cellular radio system is described. We conduct the analysis under different conditions: (1) no fading and no shadowing, (2) shadowing only, (3) fading only and (4) a combination of fading and shadowing. We also derive a model for determining the co-channel reuse distance for the centre-illuminated configuration as well as for the corner-illuminated configuration and compare the performance of the systems with different cellular structures.

Some previous work was performed in this area by French [FREN79] and Muammar [MUAM82]. French analysed the co-channel interference situations under fading and shadowing with only a single co-channel interferor while Muammar considered multiple-source co-channel interferors by using only approximate methods. Both of these analyses were performed for the centre-illuminated configuration only. Our research work differs from theirs in that our analysis is based on an accurate analytical derivation of the model. As well, our work extends the analysis to the corner-illuminated systems.

- Interference Analysis of a Cellular System:

A detailed analysis of the adjacent channel, transmitter and receiver intermodulation interference in a cellular system has been conducted. The analysis considers inter-cell and intra-cell interference situations for voice transmissions on both the uplink and downlink channels. Worst case assumptions are used such as the allocation of consecutive channels to a cell site, location of desired mobile station being on the edge of the cell coverage area, etc.

- The Design of a novel Spectrum-efficient Frequency Allocation Scheme:

A novel frequency allocation scheme for the cellular system has been designed to provide radio operation which is free from co-channel, adjacent channel and intermodulation interference and at the same time to maximize spectrum efficiency. Two other schemes which provide sub-optimal solutions in terms of spectrum-savings are also developed to complement the first scheme. All these schemes are fundamentally different from the existing cellular frequency assignment schemes in which traffic loading is considered without taking interference into account.

There are also a number of minor contributions which could be considered as extensions or improvements of research results obtained previously:

- Development of the Transmitter Output Power Model:

The transmitter output power is derived with consideration given to the effects of fading, shadowing and the system reliability factor. A similar model was described by Hata et al in [HATA81]. Our model is different from theirs in terms of certain aspects of the derivation as well as the completeness of the solution.

- Capacity of the Digital Control Channel:

The capacity of the digital control channel using the Pure Aloha random access scheme is derived. Consideration is given to the fading characteristics of the channel and the type of modulation technique used. This may be considered as an extension of the work by daSilva [DASI81].

- Definition of the Spectrum Efficiency for Cellular System

The definition of spectrum efficiency proposed by Hatfield in [HATF77] is extended and is expressed in terms of the essential parameters of the cellular system. The types of cellular structure, system size and channel bandwidth are reflected in the new definition. We have also established limitations of the cell radius based on cellular reuse requirements.

1.6 Dissertation Organization

The dissertation is organized as follows:

Chapter 1 - Introduction

The background, some suggested solutions to the spectrum congestion problem and trends of the commercial cellular radio system developments are described. The objectives of the dissertation and contributions of the research are also outlined.

Chapter 2 - The System Model

The basic cellular concept is described. The derivations of the transmitter output power model, system capacity in terms of the number of channels required and the spectrum efficiency definition are also provided.

Chapter 3 - Determination of Co-channel Reuse Distance

The co-channel interference under the conditions of fading and shadowing encountered in a cellular system is studied. Co-channel reuse distance requirements for centre- and corner-illuminated systems are also obtained.

Chapter 4 - Interference Analysis of a Cellular Radio System

Inter-cell and intra-cell adjacent channel and intermodulation interference encountered in a cellular system is analysed in detail. Uplink and downlink transmissions are considered in every interference situation.

Chapter 5 - Design of a Spectrum-efficient Frequency Allocation Scheme

Existing frequency allocation schemes are surveyed and briefly outlined. A spectrum-efficient frequency allocation scheme is derived and two other sub-optimal allocation schemes which are used to complement the first scheme are described.

Chapter 6 - Conclusions and Suggestions for Further Work

The conclusions of the research work are provided and further work for future research is suggested.

CHAPTER 2 - THE SYSTEM MODEL

2.1 Introduction

In this chapter, we provide a description of a land mobile cellular radio system model which will be used as a basis for analyses and discussions in later chapters. The first part of the chapter describes the general cellular concept, while the second part deals with a number of essential system parameters.

The cellular concept is briefly described in summary form in Section 2.2 so as to provide the framework and overall view of the operation of the system. In Section 2.3 we develop the transmitter power model with considerations given to fading and shadowing. The next important parameter that requires careful consideration is the cell radius. Since this element directly determines the number of subscribers in a cell and hence the traffic, we devote Section 2.4 to investigate the number of voice and control channels required for a given cell radius. The Pure Aloha random access method is used in the control channels. Subsequently, we derive an expression for the spectrum efficiency of a cellular radio system in Section 2.5 and compare the efficiencies of different cellular architectures.

The results of this chapter will be used in Chapters 3, 4 and 5 for the interference analysis as well as for the design of the frequency allocation schemes.

2.2 The Cellular Concept

A cellular system may be defined as a high capacity mobile radio system in which radio channels are assigned to one or more geographic cells within a defined service area. The major advantage of cellular systems over a conventional mobile radio system is the possibility of reusing the same channels within the same service area. It is this capability that makes such a system more spectrum-efficient than the other systems. The price paid for this advantage is added complexity in the control and operation of the cellular system.

In a cellular system, the entire service area is divided into non-overlapping zones or cells as shown in Fig. 2.1(a). The hexagon is preferred over other geometric shapes such as the equilateral triangle and the square which also provide non-overlapping coverage because hexagon is the closest approximation to a circular shape dictated by propagation considerations. A base station transmits its signals within its own cell which is only a small fraction of the entire service area. The transmitter power can therefore be significantly reduced from what is required in a conventional system, making it possible for the same channel to be used somewhere else in the same area served by the cellular system.

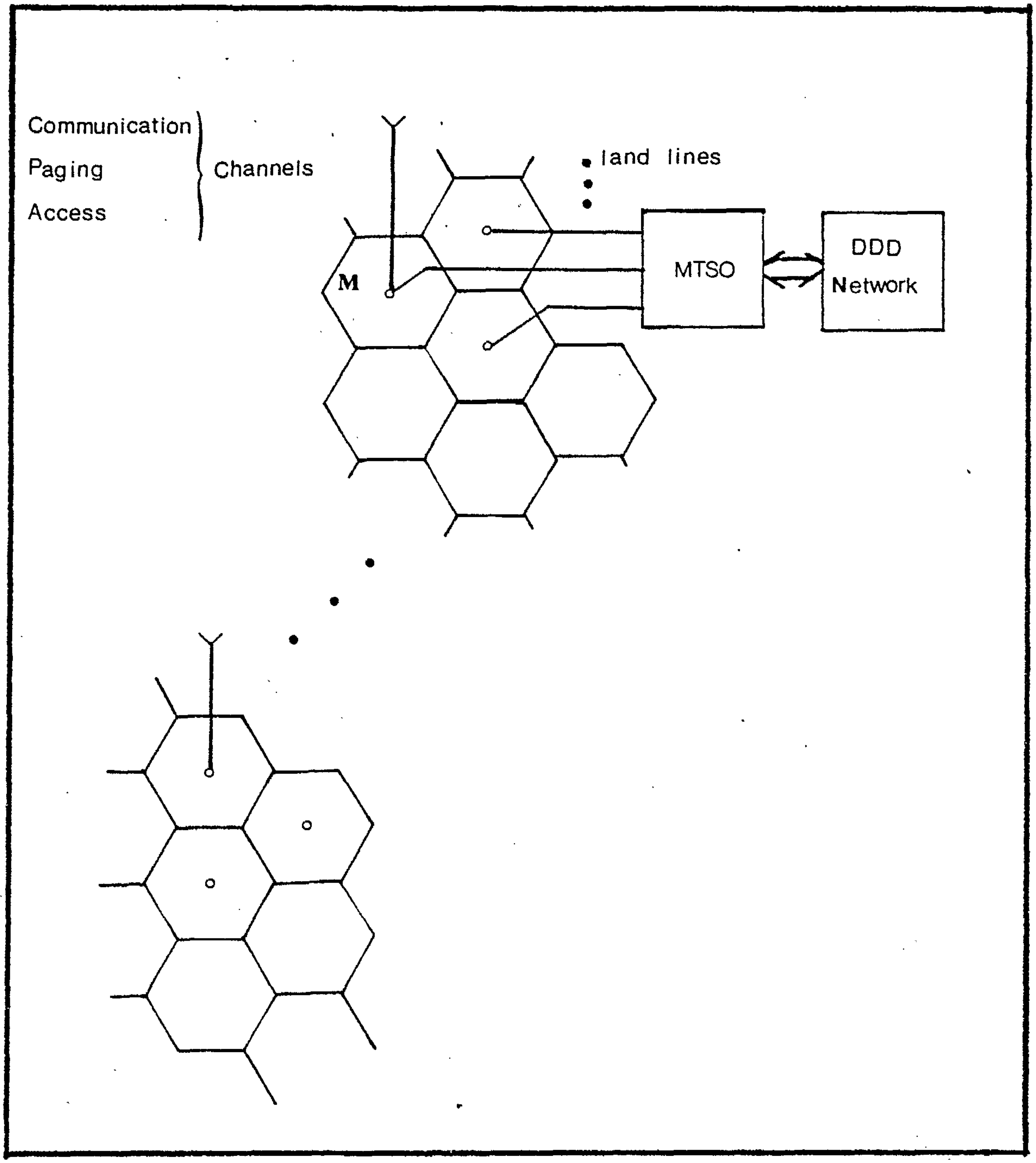


FIGURE 2.1a) THE CELLULAR CONCEPT

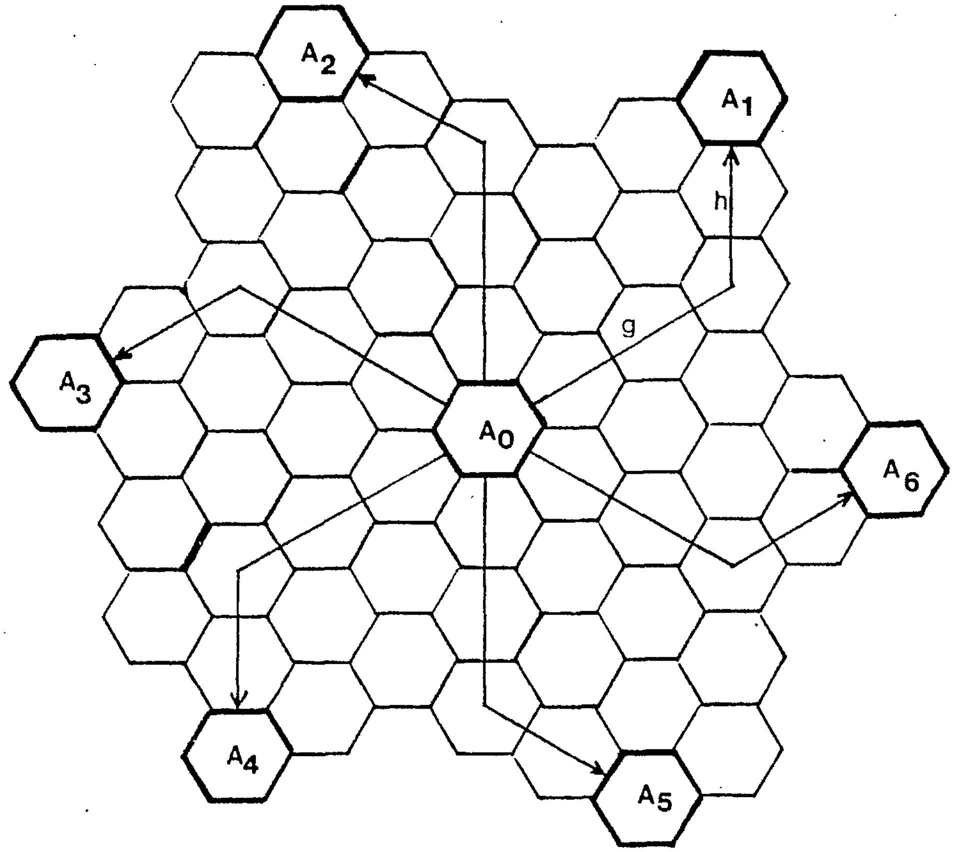
Due to interference, co-channel base stations have to be separated from each other by a certain distance which we designate as D_c depending on the minimum signal to interference ratio that is acceptable at the radio receiver. If r is the cell radius, that is the distance between the centre of the cell to any one of the six vertices, the ratio of D_c to r is called the reuse distance. Hence,

$$U = D_c/r \qquad \ldots\ldots (2.1)$$

The reuse distance is also related to N which is the number of channel sets in a cluster. N may take only selected values such as 3, 4, 7, 9, 12, 13, etc. as defined by the following expression:

$$N = g^2 + gh + h^2 \qquad \ldots\ldots (2.2)$$

where g and h are integers with $g \geq h$ and are called "shift parameters" by MacDonald in [MACD79]. Fig. 2.1(b) illustrates how the shift parameters are used as "coordinates" to locate co-channel cells. As shown, a co-channel cell such as A_1 may be reached from cell A_0 by moving g cells along the appropriate chain of hexagons; turning counter-clockwise 60 degrees and moving h cells along the chain that lies on this new heading. Furthermore, MacDonald in [MACD79] derived the relationship between the reuse distance and the number of cells, N, using geometry and arrived at the following simple relationship:



SHIFT PARAMETERS: $g = 3$, $h = 2$

$$N = g^2 + gh + h^2$$

$$U = \sqrt{3N}$$

FIGURE 2.1 b) ILLUSTRATION OF THE DETERMINATION OF CO-CHANNEL CELLS

$$U = \sqrt{3N} \qquad \ldots\ldots (2.3)$$

Therefore, g and h determine the value of N which in turn determines the structure of the cellular system.

In the mobile telephone network, each of the base stations in the coverage area contains a group of low-power transmitters/receivers that communicate with mobiles in its cell over the channels assigned to it. The number of channels required in each cell depends on the amount of traffic and the channels are shared by all the mobile units in the cell. This concept of resource sharing or more specifically trunking in this case provides lower blocking probability and greater traffic carrying capability for the system.

In addition to these communication channels, each base station is also assigned one or more control channels. These channels carry control commands in the form of data messages. These messages are used to inform the base station of mobile-originated calls and to inform the mobiles of land-originated calls through paging. The control messages are also used for mobile channel assignments. Each base station is connected to a mobile telephone switching office (MTSO) over land transmission facilities which, in turn, is connected to the nationwide telephone network.

In this mobile telephone radio network, customers in mobile vehicles may communicate with other customers in the same cell or those served by the DDD (Direct Distance Dialing) network. The mobile call sequences were described in detail by Fluhr and Nussbaum in [FLUH73]. In brief, they operate as follows.

Land-originated call: The call is first received at the MTSO from the telephone network. The MTSO then pages the desired mobile unit via the base stations over the downlink paging control channels. The mobile unit, if not busy, can then respond to the base station using the uplink paging control channel. Following this, the mobile unit tunes to the downlink access control channel and waits for the base station to send over the channel assignment. The mobile unit then switches to the communications channel and conversation may begin.

Mobile-originated call: If the call is originated at the mobile station, an origination message is transmitted via the base station to the MTSO on the uplink access control channel. When a communications channel is selected, the channel assignment will be sent to the mobile unit on the downlink access control channel..

While a telephone call is in progress, the mobile unit may cross the cell boundary and enter into another cell. This requires the MTSO to know at all times during a call the vehicle's approximate location and to reassign the mobile a channel from the new cell's channel set. This process is called hand-off. One of the problems associated with this hand-off process is the interruption in the conversation and it has to be kept to a level acceptable to the customer. It can easily be seen that systems with small cell radius may lead to an undesirable number of hand-offs during a single telephone conversation.

Cell radius is one of the most important parameters that have to be considered in the design of a cellular system. Cellular systems with small cell sizes have high spectrum efficiency since the same channel may be reused a larger number of times in the same service area. However, this leads to higher system cost since a larger number of cell base stations would be required. Furthermore, this also gives rise to more frequent hand-off procedures per single cell, leading to noticeable deteriorations in the quality of the conversations on the communications channels and consuming a significant portion of the MTSO's central processor's capacity.

On the other hand, systems with large cell radius have low spectrum efficiency. They also require high-powered transmitters or highly-elevated antennas to maintain an acceptable signal to noise ratio at the receivers. However, as there is a smaller number of base stations in the system, the installation cost will be smaller.

Another consideration in the determination of cell sizes is the growth of the system. As the system grows, the demand can only be matched by either increasing the number of channels in each cell or by splitting the cell. As the frequency band allocated to the service is limited, the number of channels per cell cannot be increased indefinitely. In the case of cell splitting, MacDonald in [MACD79] indicated that each stage of cell splitting multiplies the number of cell sites in the desired coverage area by a factor of about 4 and the system's total traffic-carrying capacity is also increased by essentially the same factor. In the same article, it was reported that the practical minimum cell radius is about 1 mile for the AMPS system.

2.3 The Transmitter Output Power Model

The transmitter output power P_t at the base station is another important parameter in the design of cellular systems. Its prediction is complicated by the fading and shadowing characteristics of the mobile channel. In the following, we first obtain a relationship between the failure rate, P_f, of the received signal, the median received signal power level, X_m, and the received signal threshold power level, X_t. Then we refer to the simple Egli formula for calculating path loss to obtain the formula of the required P_t. It should be noted that Hata et al [HATA81] also described a model for calculating transmitter output power. Our model is different from theirs in terms of certain aspects of the derivation as well as the completeness of the model.

In an urban environment where mobile vehicles travel on streets between buildings, the direct line-of-sight link between the base station transmitter and the mobile receiver is seldom attainable. Instead, the received signal is composed of a combination of reflected signals bounced off from buildings, lamp posts and other obstructions. This multipath transmission causes the phases of some signals to add constructively in some locations and others to interfere destructively in other locations. As the vehicle moves, the received signal strength varies erratically and unpredictably over a range of 20 to 30 dB over distances of about a metre. In most urban and suburban locations, the fluctuations result in a received signal envelope with a Rayleigh probability density function, i.e. the signal suffers Rayleigh fading. This has been confirmed by a number of authors such as Okumura et al [OKUM68], Meno [MENO77], Jakes [JAKE74] and French [FREN76] who conducted numerous experiments over the last decade on the mobile channel. Theoretically, it was shown by Hansen [HANS77] that the instantaneous power of the signal X under Rayleigh fading is exponentially distributed with the local mean power, X_0, as its mean:

$$P(X) = \begin{cases} 1/X_0 \cdot \exp(-X/X_0), & X > 0 \\ 0, & X \le 0 \end{cases} \qquad \ldots\ldots (2.4)$$

As the mobile vehicle moves along, the local mean power changes gradually. This effect is referred to as shadowing which is caused by buildings and hills. The changes observed in the local mean power are slow in comparison to the Rayleigh fades, since 5-dB changes in mean signal level in less than 30 metres of vehicle travel are typical. The local mean power has been found in [OKUM68], [OTT78], [MENO77], [FREN76], etc. to be lognormally distributed, i.e. the dB value of X_o is normally distributed with mean 10 log X_m or X_{md} and standard deviation σ (dB) [LEE82]. This mean is also referred to as the long term median of the signal. Hence,

$$P(X_o) = \begin{cases} \dfrac{K}{\sqrt{2\pi}\,\sigma\, X_o} \exp \dfrac{(10 \log X_o - 10 \log X_m)^2}{2\sigma^2}, & X_o > 0 \\ 0, & X_o \leq 0 \end{cases} \qquad \ldots\ldots (2.5)$$

where $K = 10 / \ln 10$

We now consider the effect of fading superimposed on shadowing on the received signal. The derivation is based on the assumption that X_o is much larger than X_t (i.e. high signal to noise ratio) which is generally true for reliable reception of signal.

From (2.4) and (2.5), the failure rate or the probability that the received signal power is less than X_t is:

$$P_f = P(X \leq X_t) = \int_0^{\infty} \int_0^{X_t} P(X \text{ given } X_o) \cdot P(X_o \text{ given } X_m) \cdot dX \cdot dX_o$$

$$= \int_0^{\infty} \int_0^{X_t} \frac{1}{X_o} \exp\left(-\,{}^{X}\!/_{X_o}\right) \cdot \frac{K}{\sqrt{2\pi}\sigma X_o} \cdot \exp\left[\frac{-(10\log X_o - 10\log X_m)^2}{2\sigma^2}\right] dX \cdot dX_o$$

$$= \int_0^{\infty} (1 - e^{-X_t/X_o}) \times \frac{K}{\sqrt{2\pi}\sigma X_o} \exp\left[-\frac{(10\log X_o - 10\log X_m)^2}{2\sigma^2}\right] \cdot dX_o$$

for $X_o \gg X_t$, $(1 - e^{-X_t/X_o}) \simeq \frac{X_t}{X_o}$

Hence,

$$P(X \leq X_t) \cong \int_0^{\infty} \frac{X_t}{X_o^2} \cdot \frac{K}{\sqrt{2\pi}\,\sigma} \cdot \exp\left[-\frac{(10\log X_o - 10\log X_m)^2}{2\sigma^2}\right] \cdot dX_o$$

$$= \int_0^{\infty} \frac{X_t \cdot K}{\sqrt{2\pi}\,\sigma X_o^2} \times \exp - \left(\frac{10\log \frac{X_o}{X_m}}{\sqrt{2}\sigma}\right)^2 \cdot dX_o$$

Let $a = X_o/X_m$, then

$$P(X \leq X_t) \cong \frac{X_t \cdot K}{\sqrt{2\pi}\,\sigma X_m} \int_0^{\infty} \frac{1}{a^2} \exp - \left[\frac{(10\log a)^2}{2\sigma^2}\right] \cdot da$$

and let $\bar{y} = \log a$, then

$$P(X \leq X_t) \cong \frac{10 . X_t}{\sqrt{2\pi}\ \sigma X_m} \int_{-\infty}^{\infty} e^{-y\ln 10}\ e^{\frac{-y^2 100}{2\sigma^2}} \ . \ dy$$

After some algebraic manipulations,

$$P(X \leq X_t) \cong \frac{10 . X_t}{\sqrt{2\pi} . \sigma . X_m} \ . \ \exp\left[\frac{\sigma^2}{2}\ \frac{(\ln 10)^2}{100}\right] . \int_{-\infty}^{\infty} \exp - \frac{1}{2\sigma^2}\ (10y + \frac{\sigma^2 \ln 10}{10})^2 \ . \ dy$$

$$\cong \frac{X_t}{X_m} \ . \ \exp\left[\frac{\sigma^2 (\ln 10)^2}{200}\right]$$

$$P_f \cong \frac{X_t}{X_m} \ . \ \exp\left[\frac{\sigma^2}{200(0.43)^2}\right] \qquad \text{...............(2.6)}$$

Expressing in dB units,

$$P_{fd} \cong X_{td} - X_{md} + 0.117\,\sigma^2 \qquad \text{..... (2.7)}$$

But X_{md} is related to P_t, P_l, G_t and G_r as follows:

$$P_t + G_t + G_r - P_l = X_{md} \qquad \text{..... (2.8)}$$

where P_l = path loss in dB,

G_t = base station transmitter antenna gain in excess of circuit losses in dB,

G_r = mobile station receiver antenna gain in excess of circuit losses in dB.

By combining (2.7) and (2.8), we have

$$P_t + G_t + G_r - P_1 = X_{td} - P_{fd} + 0.117\sigma^2 \qquad \ldots\ldots (2.9)$$

Using the Egli model [EGLI57],

$$P_1 = 40\log d + 20\log f + 83.64 - 20\log H_r.H_t \qquad \ldots\ldots (2.10)$$

where d = distance in Km.,

f = frequency in MHz,

H_r = receiver antenna height in metres and

H_t = transmitter antenna height in metres.

and from (2.9) and (2.10),

$$P_t \cong X_{td} - P_{fd} + 0.117\sigma^2 - G_t - G_r + 40\log d$$

$$+ 20\log f + 83.64 - 20\log H_r.H_t \qquad \ldots\ldots (2.11)$$

If we assume H_r = 2 m., H_t = 45 m., $G_t + G_r$ = 10 dB, (2.11) becomes:

$$P_t \cong X_{td} - P_{fd} + 0.117\sigma^2 + 34.6 + 40\log d + 20\log f \qquad \ldots\ldots (2.12)$$

It should be noted that when there is no fading and no shadowing, P_t can simply be obtained from (2.8) and (2.10).

The values of P_t for different values of P_{fd}, σ and d with f = 850 MHz and X_{td} = -132 dBW are given in Table 2.1 and plotted in Fig. 2.2. It can be seen that for a cell radius of 6 Km. at 90 % reception reliability or P_f = 0.1, P_t has to be increased by 10 dB to compensate for fading loss and an extra 4 dB to compensate for shadowing loss when σ = 6 dB and 16.5 dB when σ = 12 dB.

While the above only provides the calculation of the transmitter output power for the base station, the calculation of the transmitter output power for the mobile station is essentially the same. The mobile transmitter power is usually lower than the base transmitter power and is compensated by the higher sensitivity of the base station receiver. Other methods of compensation such as the use of repeaters or directional base station receiver antennae, etc. may also be used.

2.4 System Capacity

Cellular systems are high capacity radio systems capable of supporting up to 500,000 subscribers. The traffic handling capability is basically determined at the cell level. In this section, we are concerned with finding the number of voice and control channels that are required in a cell to provide a grade

σ (dB)	d (Km)	P_t (dBW) $P_f = 0.1$	P_t (dBW) $P_f = 0.02$
0	3	-9.72	-2.72
(fading, but no	6	2.32	9.33
shadowing)	10	11.19	18.20
	14	17.04	24.05
6	3	-5.51	1.49
	6	6.54	13.54
	10	15.41	22.41
	14	21.26	28.26
12	3	7.13	14.13
	6	19.18	26.18
	10	28.05	35.05
	14	33.89	40.90
(no fading and	3	-19.7	
no shadowing)	6	- 7.7	
	10	1.2	
	14	7.0	

TABLE 2.1

TRANSMITTER OUTPUT POWER REQUIREMENTS

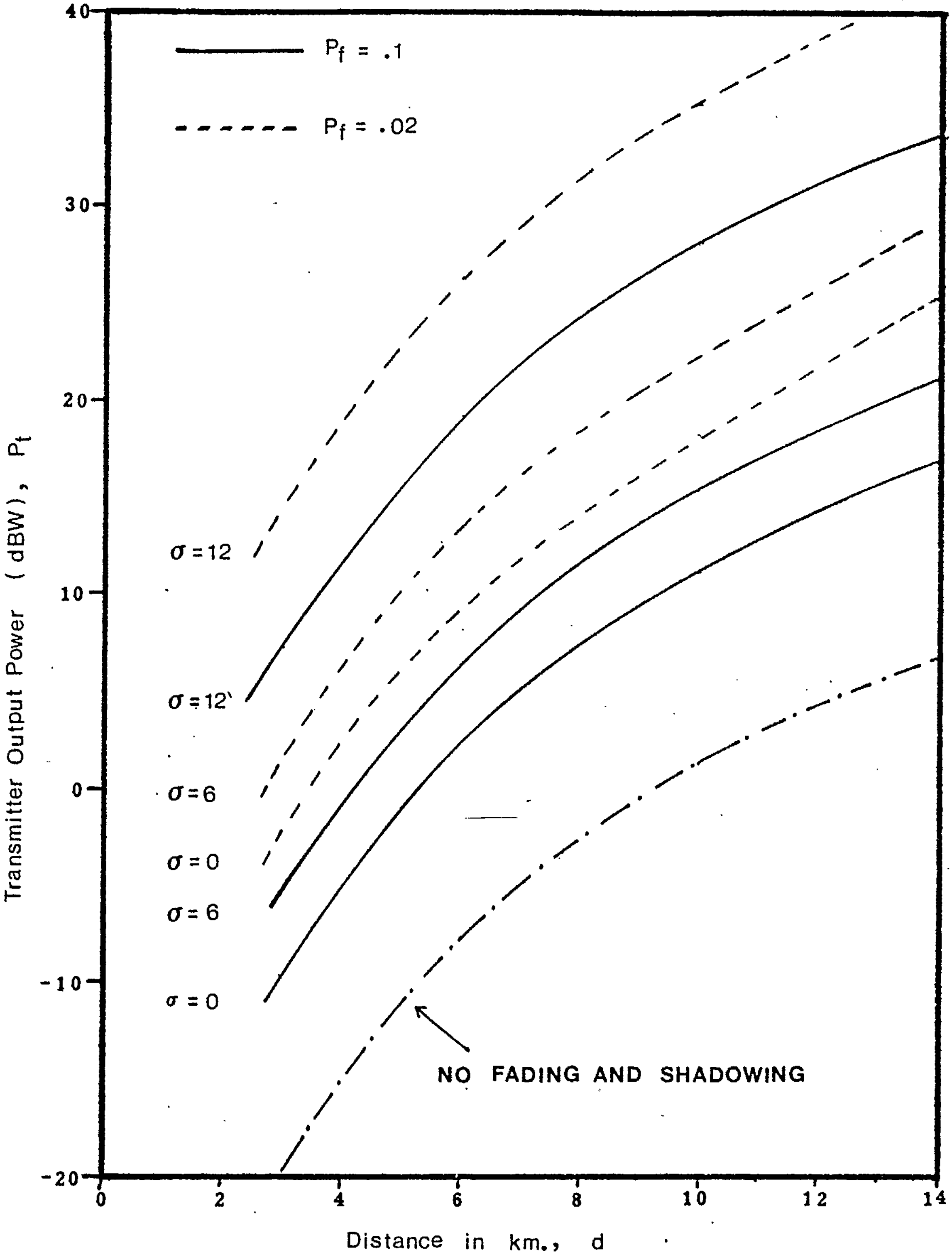


FIGURE 2.2 TRANSMITTER OUTPUT POWER REQUIREMENT vs DISTANCE

of service of 0.02. If the traffic load per mobile subscriber in the busy hour is t (Erlangs) and the mobile density is V (mobiles per sq. km.), then the traffic density is Vt (Erlangs per sq. km.). Based on these assumptions, we now determine the number of channels required by the cellular system.

2.4.1 Voice Channels

Using the Erlang B expression in (1.1), the probability of having blocked calls is given by:

$$P_b = \frac{T^m}{m!} \Big/ \sum_{i=1}^{m} \frac{T^i}{i!} \qquad \text{..... (2.13)}$$

where P_b is the probability of blocking or loss,

T is the traffic offered in Erlangs and

m is the number of available channels,

In land mobile radio telephone systems, all blocked calls are cleared. The relationship between T and m for P_b = 0.02 is extracted from the Traffic Tables in [FRAN76] and is shown in Fig. 2.3. The relation between cell radius and traffic load may also be obtained easily:

If r is the cell radius in km., then the offered traffic T in the cell is the product of the area of the cell and the traffic density. Hence,

T = (area of cell).V.t

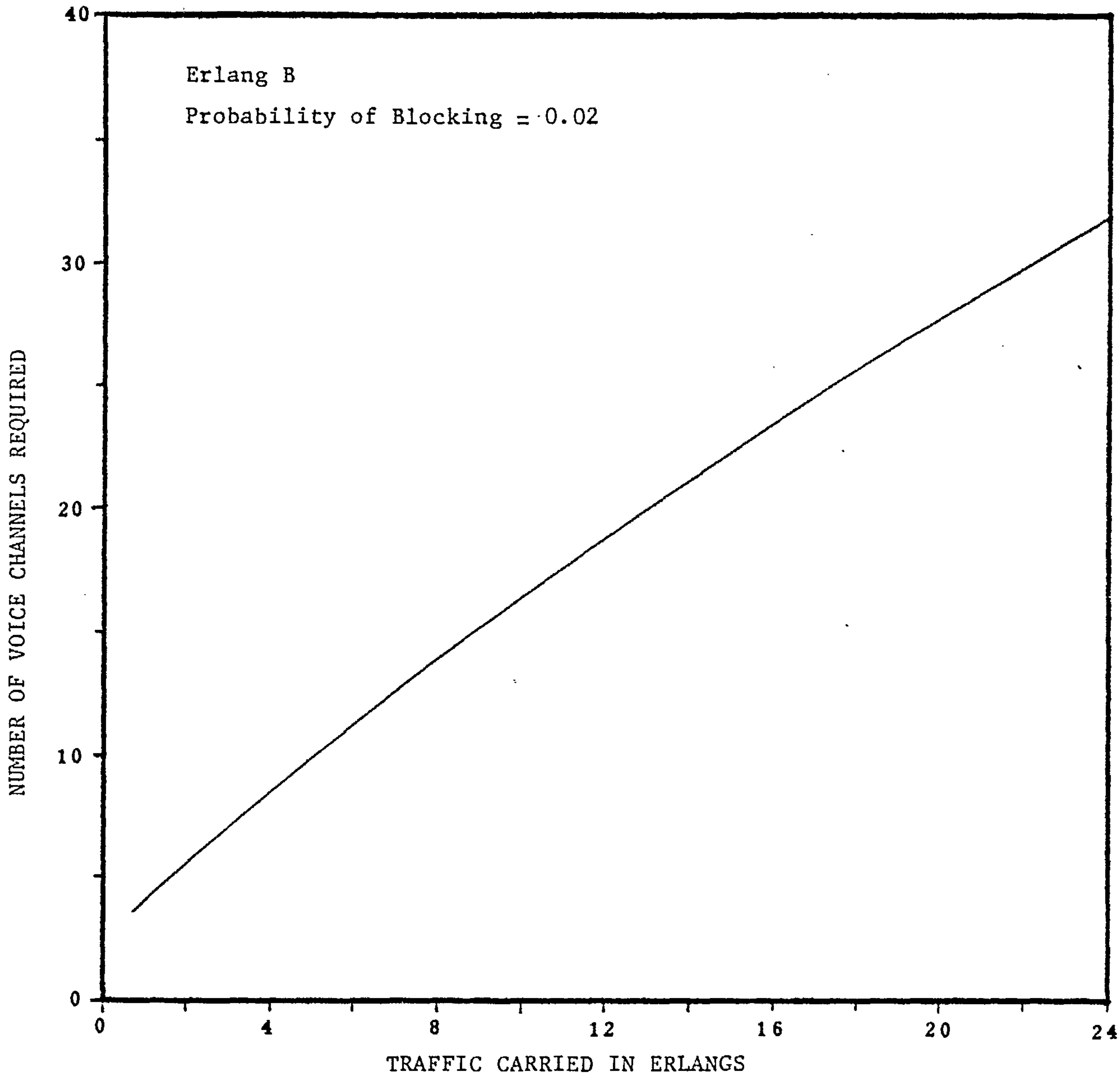


FIGURE 2.3 THE NUMBER OF VOICE CHANNELS REQUIRED VS TRAFFIC CARRIED

$$T = 3\sqrt{3} \; . \; r^2.V.t/2 \qquad \text{..... (2.14)}$$

For a start-up system, Vt may take on values such as 0.04 Erlangs per sq. km. with $t = 0.02$ Erlangs and $V = 2$ mobiles per sq. km. We have also chosen values of $Vt = 0.2$ and 0.4 Erlangs per sq. km. for the medium size and mature cellular systems, respectively, for later analysis. We assume that the amount of traffic per mobile subscriber remains unchanged while the number of mobiles per sq. km. in a cell would increase as the system size grows.

2.4.2 Control Channels

As mentioned in Section 2.1, the control channels carry signalling information between the base and mobile stations such as paging, acknowledgement, mobile call request, channel assignment data, etc. Since mobile stations share the use of one or more uplink (mobile to base) channels, some kind of contention of channel usage would exist. The downlink (base to mobile) channels do not have the same problem since its usage is under the direct control of the base station.

In the AMPS system as described by Fluhr and Porter in [FLUH78], messages on the downlink and uplink control channels are repeated 5 times with the receiver performing a bit-by-bit, 3-out-of-5 vote to determine the received message. The repetitions are used to combat burst errors caused by fading in the mobile channel.

In this section, we propose to organize the messages in the form of fixed-length packets and use the random access Pure Aloha scheme on the uplink channel with automatic repeat request (ARQ). Considering the call sequence and the type of messages carried on the control channel as described in Section 2.2, we assume that the maximum number of packets is 4 per call on the uplink as well as on the downlink channel. Before we proceed to determine the number of control channels required, we have to look at the packet error rate and the capacity of the data channel under the conditions of fading.

The two important parameters of the Rayleigh fading process are the average level crossing rate, B, and the average fade duration, F_o. Jakes in [JAKE74] defined B as the expected rate at which the envelope of a received signal crosses a threshold signal level, X_t, in the positive direction and F_o is the average duration the signal stays below the threshold level. These two parameters are given by:

$$B = f_d . \sqrt{2\pi / Z_o} . \exp(-1/Z_o) \qquad ..... (2.15)$$

$$F_o = \frac{\exp(1/Z_o) - 1}{f_d \sqrt{2\pi / Z_o}} \qquad ..... (2.16)$$

where Z_o = average value of the signal to noise ratio, Z

$$= \frac{X_o}{X_t} = \frac{\text{local mean power level}}{\text{threshold power level}}, \quad \ldots(2.17)$$

and f_d = maximum Doppler frequency shift = v/w , ... (2.18)

v = vehicle speed,

w = wavelength of carrier frequency.

Note that both B and F_o are functions of X_t.

The average interfade duration, I_o, (which is the sum of the fade duration and the non-fade duration) and the non-fade duration, J_o, may be obtained and they are given by:

$$I_o = \frac{1}{B} = \frac{\exp(1/Z_o)}{f_d\sqrt{2\pi/Z_o}} \quad \ldots\ldots (2.19)$$

$$J_o = I_o - F_o = \frac{1}{f_d\sqrt{2\pi/Z_o}} \quad \ldots\ldots (2.20)$$

Furthermore, the non-fade duration J was shown by daSilva [DASI81] to have an exponential distribution if Z_o is in the range of 10 to 30 dB:

$$P(J) = 1/J_o \,.\, \exp(-J/J_o) \quad \ldots\ldots (2.21)$$

We can now derive the probability P_c that a packet may be received correctly over a fading mobile radio channel. We assume here that if the packet falls into a fade duration, P_c is zero. Hence,

$$P_c = P_1 \cdot P_2 \cdot P_3 \qquad \ldots\ldots (2.22)$$

where P_1 = the probability that the first bit of the packet is in a non-fade duration, J

$$= J_o / I_o \qquad \ldots\ldots (2.23)$$

P_2 = the probability that all the remaining bits are in J

P_3 = the probability that all the bits in J are received correctly

Because of the memoryless property of the exponential distributed function,

$$P_2 = \left(\text{probability that } J \geq b \right)^{M-1}$$

where M = number of bits in the packet and

b = length of 1 bit

Using (2.21), we have,

$$P(J \geq b) = \exp(-b/J_o)$$

$$P_2 = \exp(-b/J_o).(M-1) \qquad \text{..... (2.24)}$$

For P_3, we first let $P_{eo}(Z_o)$ be the average bit error rate when the signal is in J, then

$$P_3 = \left[1 - P_{eo}(Z_o)\right]^M \qquad \text{..... (2.25)}$$

where $P_{eo}(Z_o) = \int_1^{\infty} P_e(Z).\ P(Z).dZ$ ..... (2.26)

From (2.4), we have,

$$P(Z) = 1/Z_o.\ \exp(-Z/Z_o) \qquad \text{..... (2.27)}$$

$P_e(Z)$ is the bit error rate expression for the selected modulation scheme. The choice of the modulation scheme affects the error performance of the packet transmissions. In this section, we deal with the differential phase shift keying (DPSK) scheme mainly because of its simplicity. Therefore,

$$P_e(Z) = 1/2\ .\ \exp(-Z) \qquad \text{..... (2.28)}$$

and $P_{eo}(Z_o) = \dfrac{1}{2(1 + Z_o)}\ .\ \exp -(1 + 1/Z_o)$ ..... (2.29)

Hence P_c can be obtained from (2.23), (2.24), (2.25) and (2.29) and its values are plotted in Fig. 2.4 for different values of M, signal to noise ratio and transmission speed. The vehicle speed and the carrier frequency are assumed to be 60 km./hr. and 850 MHz respectively.

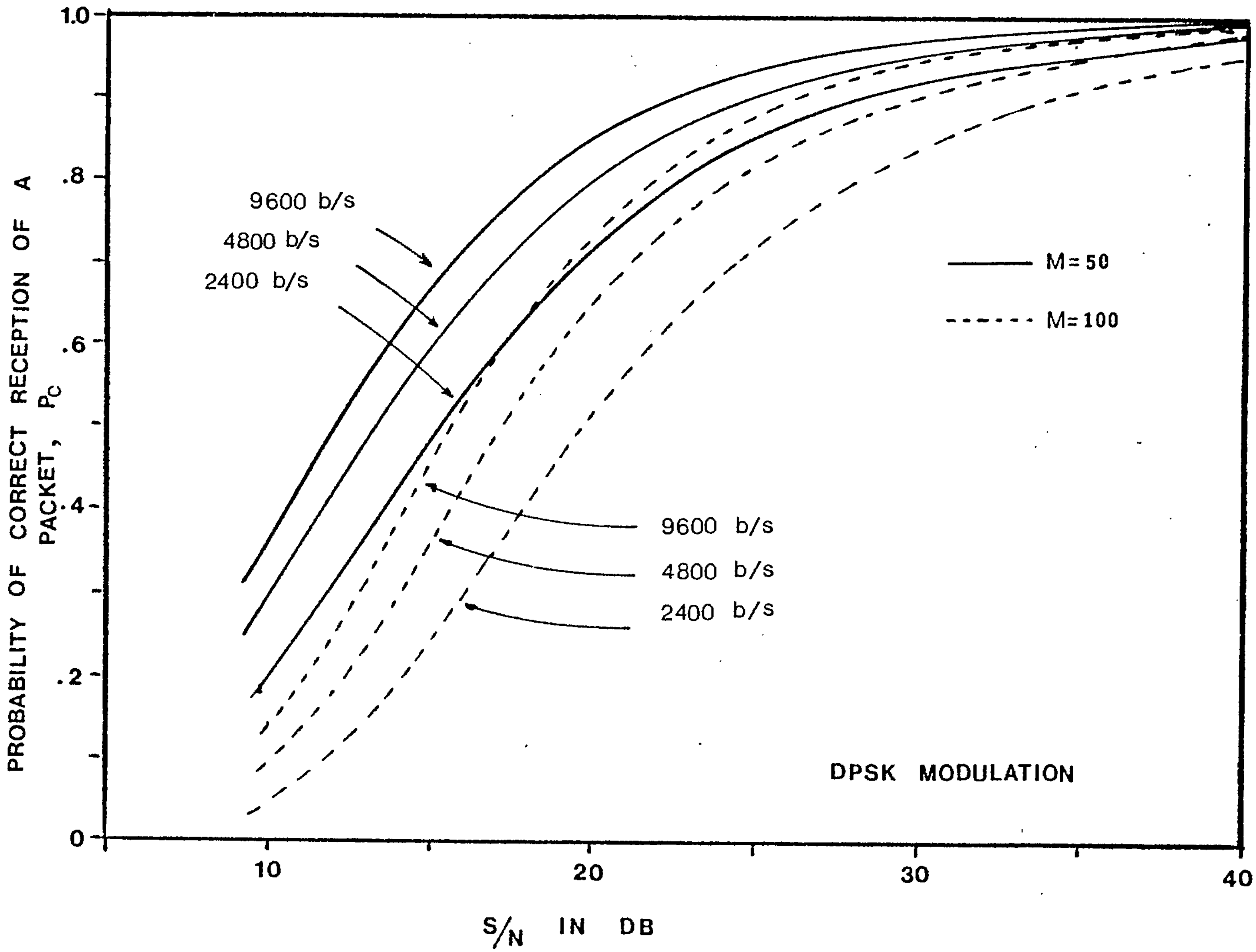


FIGURE 2.4 PROBABILITY OF CORRECT RECEPTION OF A PACKET VS S/N RATIO UNDER FADING

Using the Pure Aloha random access scheme as described by Abramson in [ABRA77], the throughput, S, of the system or the amount of traffic (information) that can be carried by the system is related to the offered channel traffic, G, by:

$$S = G \exp(-2G) \qquad \text{..... (2.30)}$$

The exp(-2G) factor is the probability that no two packets would collide with each other. As the packet is transmitted in the fading channel, the actual system throughput would be smaller than the one described in (2.30) and must be equal to:

$$S = G.\ P_c.\ \exp(-2G) \qquad \text{..... (2.31)}$$

Equation (2.31) is plotted in Fig. 2.5 and 2.6 for M = 50 bits and 100 bits respectively at a transmission speed of 4800 bits per second.

In order to estimate the number of control channels required, it is necessary to predict the amount of carried traffic S on the channel. Assuming a mean call holding time of 120 seconds, the number of calls made per second, λ , is equal to:

$$\lambda = \frac{\text{total voice traffic in cell}}{120} \qquad \text{..... (2.32)}$$

From (2.14),

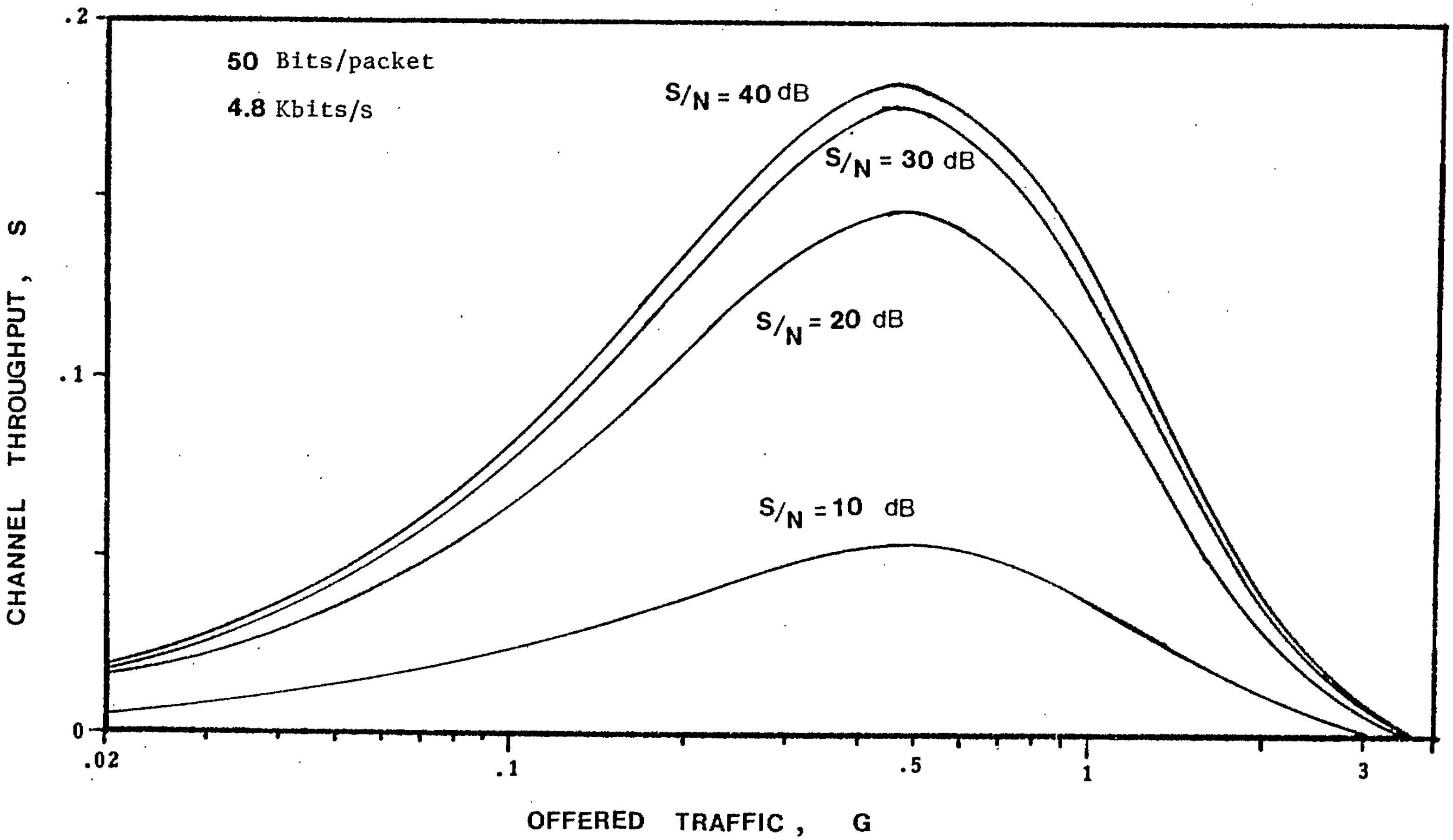


FIGURE 2.5 CHANNEL THROUGHPUT VS OFFERED TRAFFIC USING PURE ALOHA RANDOM ACCESS SCHEME, M=50 BITS

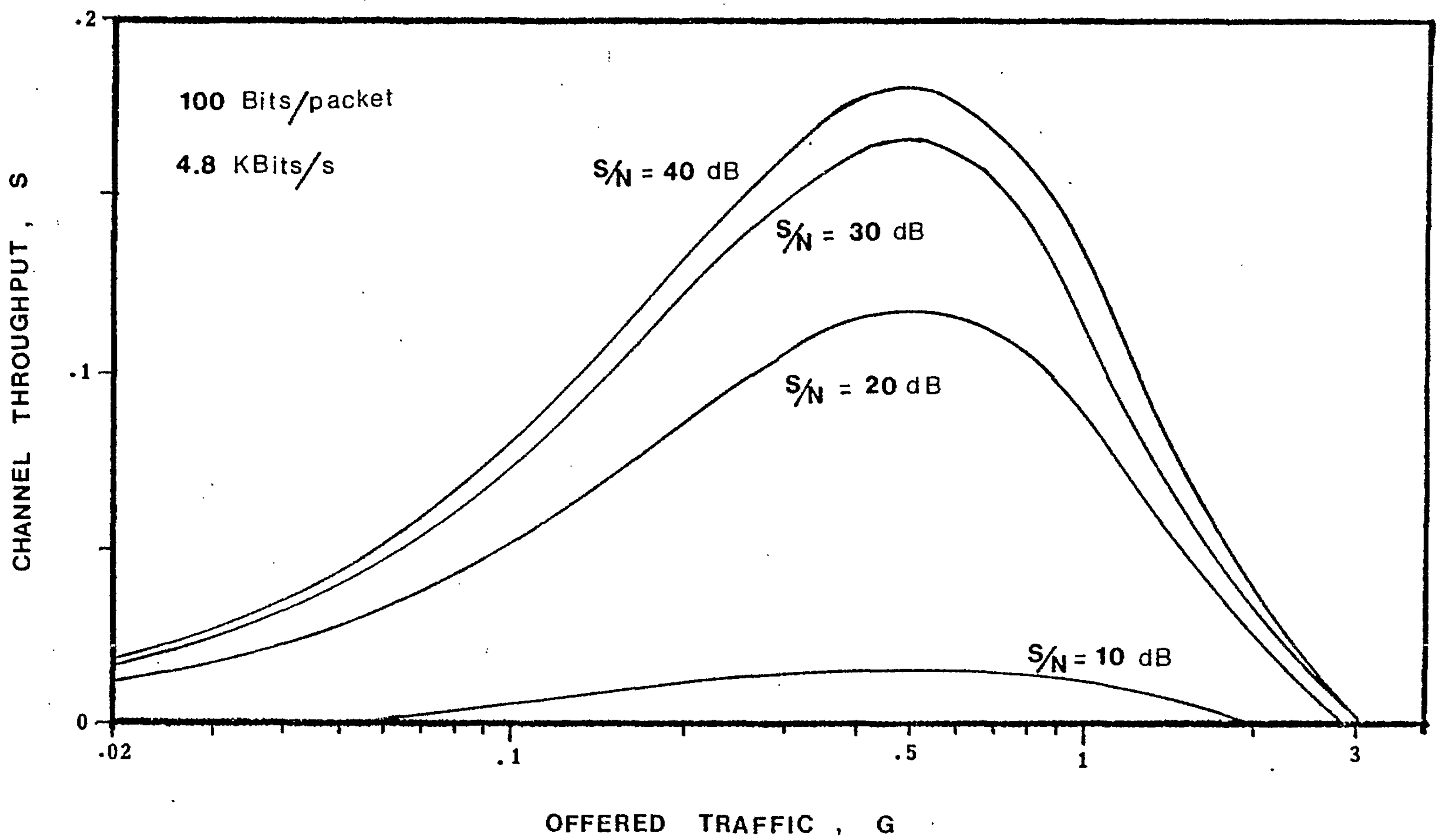


FIGURE 2.6 CHANNEL THROUGHPUT VS OFFERED TRAFFIC USING PURE ALOHA RANDOM ACCESS SCHEME, M=100 BITS

$$\lambda = \frac{T}{120} = \frac{3\sqrt{3}\ r^2.V.t}{2 \times 120} \qquad \text{..... (2.33)}$$

For a data rate of 4800 bits per second and packet length of 100 bits, the packet duration, τ , is 0.021 sec. Assuming a maximum of 4 packets per call,

$$S = 4\lambda\tau$$

$$= 0.0018\ r^2\ Vt \qquad \text{..... (2.34)}$$

The number of control channels may then be determined from Figs. 2.5 and 2.6 if the values of r, Vt and S/N are known. The results are shown in Table 2.2 and it can be seen from the Table that except for a mature system with a cell radius of 14 km. which is much too large in this case, one control channel using the Pure Aloha random access scheme would be sufficient.

There are two points that should be noted. The first one is that since the downlink channel is under direct control of the base station, no collision of packets would occur on it. Hence the probability of receiving a packet correctly on this channel is higher than that on the uplink channel. If one channel is required for the uplink transmissions, the same holds true for the downlink transmissions. The second one is that shadowing is not considered in this analysis. Since shadowing is a slowly-varying process, the packet duration of 21 msec. is too short for shadowing to have much effect on the signal level. For the range

Vt (Erlangs per sq. km.)	r (km)	S	No. of control channels required at S/N = 20
0.04	3	0.001	1
	6	0.003	1
	10	0.007	1
	14	0.014	1
0.2	3	0.003	1
	6	0.013	1
	10	0.036	1
	14	0.071	1
0.4	3	0.006	1
	6	0.026	1
	10	0.072	1
	14	0.142	2

TABLE 2.2

CONTROL CHANNEL REQUIREMENTS FOR A CELLULAR SYSTEM of S/N ratio (10-30 dB) we are considering, it is reasonable to assume that the mean power level of the data packet is constant leaving fading as the predominant factor that would affect the signal level of the packet.

2.5 Spectrum Efficiency of Cellular Systems

In order to be able to develop a spectrum-efficient system, it is necessary to define what spectrum efficiency is. Berry [BERR77] and later Tou and Roy [TOU80] generally described the spectrum efficiency of a system as the ratio of the information delivered to the amount of spectrum space used. Spectrum space is equal to the product of bandwidth, physical space and time. Berry [BERR77] also indicated that the definition of Erlangs/MHz/square mile proposed by Hatfield [HATF77] is the most useful. Since an Erlang is a measure of traffic per unit time, the definition becomes

$$\text{spectrum efficiency, } E = \frac{\text{traffic}}{(\text{bandwidth})(\text{area})(\text{time})} \qquad \ldots. \ (2.35)$$

This definition is very suitable for a cellular system where the same frequency is reused a number of times within a service area.

In comparing spectrum efficiencies of systems using the Hatfield definition, the traffic distribution is assumed to be uniform as is the case in this dissertation. However if the traffic is non-uniform, the area will have to be partitioned into sections with uniform traffic and the average spectrum efficiency for the whole system must be obtained. Furthermore, systems must be compared under similar conditions such as equal signal to noise ratio, bit error rate, blocking probability, etc.

In the following, we further expand the definition described in (2.35). We assume that the traffic carried is approximately equal to the traffic offered since the blocking probability P_b is low and

$$\text{traffic carried} = \text{traffic offered} \times (1 - P_b)$$

But total traffic carried in the system = traffic carried per cell T x number of cells L in the system. Therefore, if A is the area of the service area, then

$$L \cong \frac{A}{\text{area of a cell}}$$

This is only an approximation because of irregularities of the boundaries of the service area. Therefore,

$$L \cong \frac{2A}{3\sqrt{3}\;\; r^2} \qquad \text{..... (2.36)}$$

The bandwidth required by the system is equal to the product of the channel spacing, f_s, the total number of channels per cell, m, the number of cells per cluster, N, and the number of different clusters, C in the system:

$$\text{Bandwidth} = f_s \times m \times N \times C \qquad \ldots\ldots (2.37)$$

where C = 1 with frequency reuse. Also note that m is equal to the sum of the number of control channels and the number of voice channels, $n(T, P_b)$, which is a function of T and P_b.

From (2.33), (2.35), (2.36) and (2.37), the spectrum efficiency

$$E = \frac{T \times L}{\text{Bandwidth} \times A}$$

$$= \frac{Vt}{f_s . m . N}$$

To summarize, with the results obtained from Section 2.4 and this section, we have tabulated the essential parameters of traffic T, the number of channels required and the spectrum efficiency as a function of r in Table 2.3. It should be noted that the unit of E is Erlangs per KHz per Sq. Km.

The tabulations are also plotted in Fig. 2.7. The three groups of curves with Vt = 0.04, 0.2 and 0.4 Erlangs per sq. km. correspond to the start-up system, the medium capacity system and the high capacity mature system respectively. From these results, we can make a number of observations.

Firstly, the high capacity system has higher spectrum efficiency than the medium capacity and start-up systems because

Vt Erlangs per sq. km.	r km.	T Erlangs	$n(T, P_b)$	No. of channels req'd per cell, m	E** N=3	N=7	N=12	N=19
0.04	2	0.4	3	4	111.1	47.6	27.8	17.5
	3	0.9	4	5	88.9	38.1	22.2	14.0
	6	3.6	8	9	49.4	21.2	12.3	7.8
	10	10.0	17	18	24.7	10.6	6.2	3.8
	14	19.6	28	29	15.3	6.6	3.8	2.4
0.2	2	2.1	6	7	317.5	136.1	79.4	50.1
	3	4.5	10	11	202.0	86.6	50.4	31.9
	6	18.0	26	27	82.3	35.3	20.6	13.0
	*10	50.0	61	62	35.8	15.3	8.9	5.7
	*14	98.0	111	112	19.8	8.5	4.9	3.1
0.4	2	4.2	9	10	444.0	190.5	111.1	70.2
	3	9.0	15	16	278.0	119.0	69.5	43.9
	* 6	36.0	46	47	94.6	40.5	23.7	14.9
	*10	100.0	113	114	39.0	16.7	9.7	6.2

Notes:

* Situations in which cell radius becomes impractically large

** E is in Erlangs per KHz per sq. km.

TABLE 2.3

SPECTRUM EFFICIENCY OF A START-UP, MEDIUM-SIZE AND MATURE

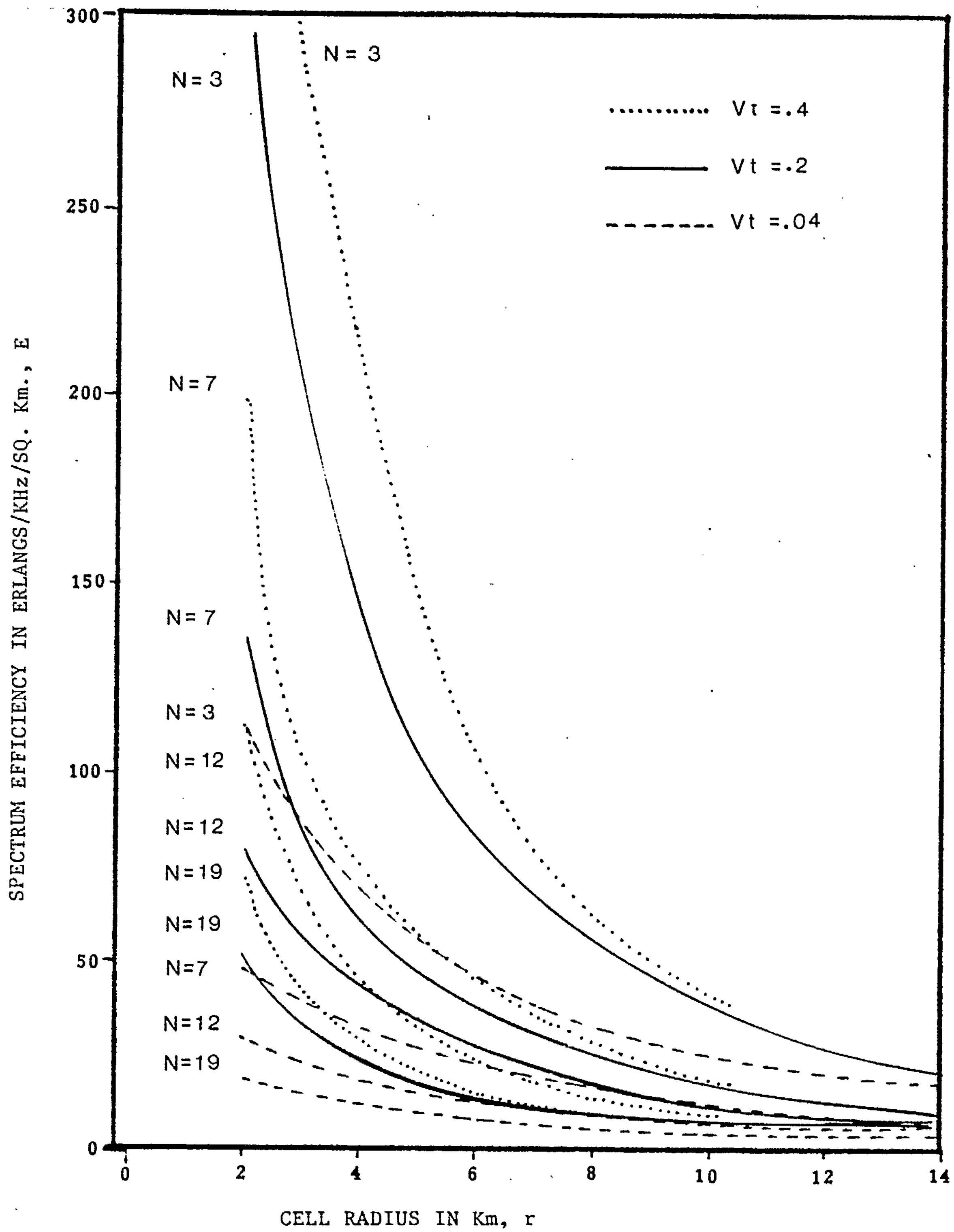


FIGURE 2.7 SPECTRUM EFFICIENCY OF A START-UP, MEDIUM-SIZE AND MATURE CELLULAR SYSTEM VS CELL RADIUS

of the gain due to trunking. It can be seen that the ratio of traffic density to the number of channels required for a given r is higher in the high capacity system than in the medium and low capacity systems.

Secondly, as N increases, the efficiency decreases. Since more channels are required for a larger N to cover the same area, the system is less efficient.

The last observation is that when cell radius increases, the efficiency decreases. This may be explained by the fact that even though r does not appear in the expression for E, as it increases, the number of channels required increases, hence more spectrum is required. However, as shown in Table 2.3, the traffic carried per channel increases when r increases. Due to the trunking effect, resources can be used more efficiently when they are shared between users than when they are used separately. Therefore for large r, channel utilization is higher, but spectrum efficiency for the system is lower. Conversely, if r is reduced, the spectrum efficiency increases. However, as mentioned in Section 2.2, there is a lower limit for the cell radius. In this case, we have assumed the minimum value of r to be 2 km.

The above is based on the assumption that A is large enough to sustain frequency reuse at all values of r. However, in practice, if A is not large enough, the cell radius cannot increase beyond a certain limit and yet maintain the reuse capability. If we have a service area which is approximately circular, we may observe that in order to make frequency reuse possible,

$$D_c < 2R \quad \text{..... (2.40)}$$

where R is the radius of the service area.

In terms of reuse distance and N,

$$r < 2R/U \quad \text{..... (2.41)}$$

or

$$r < 2R/\sqrt{3N} \quad \text{..... (2.42)}$$

This inequality is plotted in Fig. 2.8 in which r must be below the straight line for a given N in order to have frequency reuse. Otherwise reuse is not possible and $\underline{C}$ would be larger than 1, giving a lower spectrum efficiency.

2.6 Summary of Results

In this chapter, we have laid down the ground work for further analysis and investigation of cellular radio systems which will be conducted in later chapters. In Section 2.2, we describe the cellular concept and the model which we shall adopt in this research work.

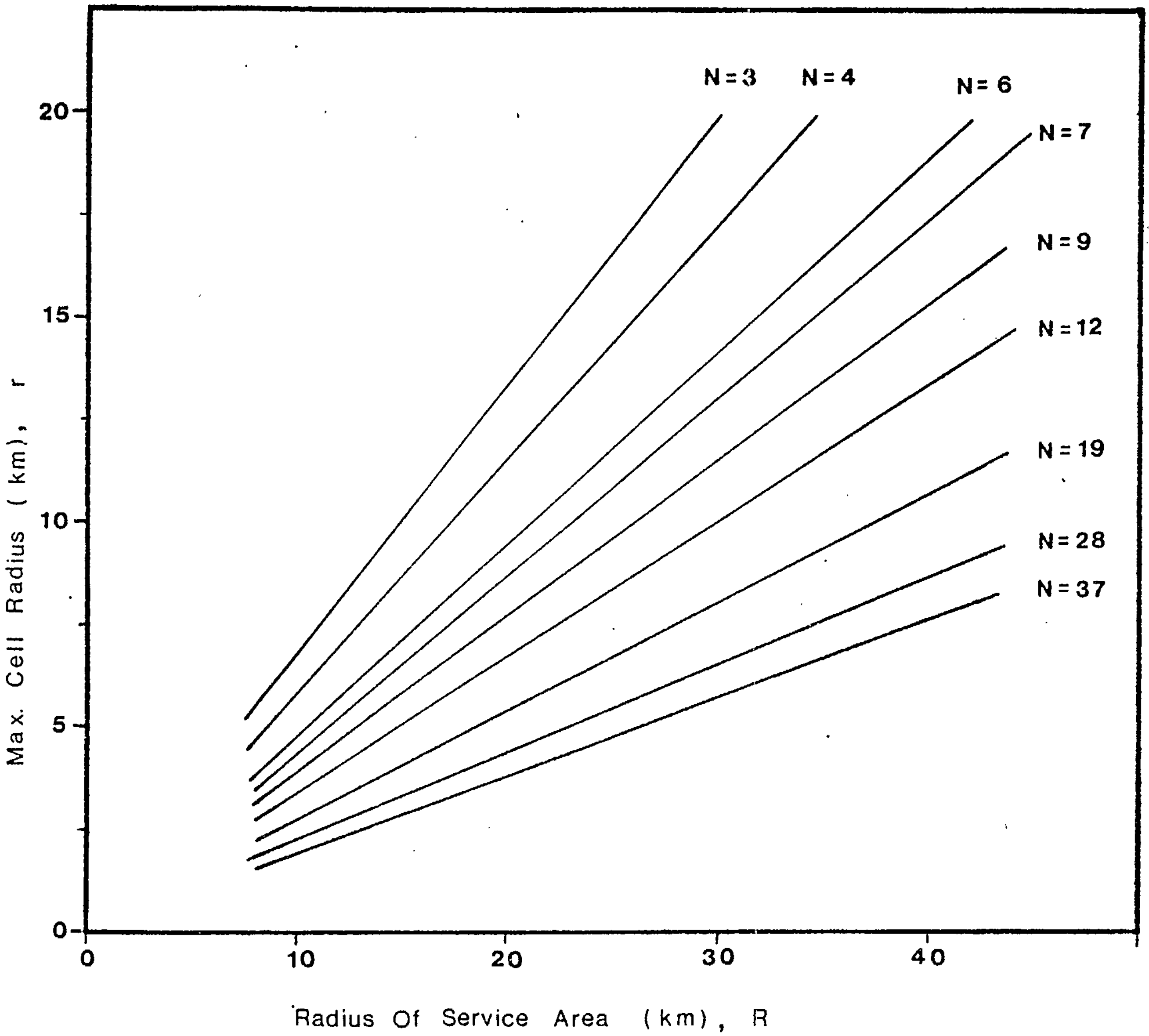


FIGURE 2.8 MAXIMUM CELL RADIUS vs SERVICE AREA RADIUS

The transmitter output power model is developed in Section 2.3 with considerations of fading and shadowing. The output power required is a function of the distance or cell radius, the system reliability factor, the carrier frequency, the threshold power level and σ . This model will be used in Chapter 4 where adjacent channel and intermodulation interference is analysed.

In Section 2.4, we derive the methodology to obtain the number of voice and control channels required for a given cell radius and system size. The results will be needed in Chapter 5 in which a frequency allocation scheme is developed. We also investigate in the same section the capacity of the digital control channel using Pure Aloha random access scheme and the probability of receiving a data packet correctly over a fading mobile channel. The number of such channels required does not exceed one in all practical situations.

The last section of the chapter recommends a suitable definition of spectrum efficiency for cellular system. This parameter may be used to compare performance of different system structures in later chapters of this work.

Finally, it must be pointed out that the cell radius is an important parameter that would affect the design of the system. We have already indicated that it has an impact on the transmitter output power, the number of voice channels required and finally the efficiency of the system. It will be shown in Chapter 4 that it also has a role to play in formulating the interference criteria of the system.

CHAPTER 3 - DETERMINATION OF CO-CHANNEL REUSE DISTANCE

3.1 Introduction

In addition to fading and shadowing, the performance of the mobile radio channel is further degraded by interference caused by other radio users. The most common form of interference is that due to co-channel operation. As one of the methods to increase the overall spectrum efficiency of a system is to increase its frequency reuse capability, it is desirable to reuse the same channel in two cells which are as close to each other as possible. However, this increases the probability of co-channel interference. Therefore a trade-off exists between the increased efficiency of the system and the higher probability of co-channel interference.

Study of co-channel interference in mobile radio systems was conducted by some earlier workers. Engel [ENGE69] studied the effects of fading only using analytical methods while Lundguist and Peritsky [LUND71] studied the effects of fading and shadowing using simulation methods. The first theoretical analysis on the combined effect of fading and shadowing on the mobile channel was conducted by Hansen and Meno [HANS77]. They derived a simple expression for calculating the probability, $P\ (X \leq X_t)$, that the power levels of a received signal, X, is less than or equal to the threshold power level, X_t under fading and shadowing. This is based on the assumption that the mean received signal power level, X_o, is much larger than X_t. French in [FREN79] obtained a more accurate value for the probability of co-channel interference under fading and shadowing using numerical integration techniques and derived the relationship between reuse distance and the probability of co-channel interference.

In a cellular system, a cell is surrounded by rings of cells operating on the same channel. The total interference coming from all these stations must be considered. Muammar and Gupta in [MUAM82] took the six closest surrounding stations into consideration and approximated the distribution of the sum of the six interfering signals, Z', by a normal distribution. All other co-channel stations located further away were ignored.

As received power level is inversely proportional to the p^{th} power of distance separation, with p ranging from 3 to 4, the interfering signals from stations beyond the first ring may be ignored. Based on the same assumption, Chan et al in [CHAN82] obtained a more accurate expression for the probability density function for Z'. It was shown that Z' has a Gamma distribution. The result however was only derived under conditions of fading. Effect of shadowing was not taken into consideration.

This chapter is concerned with the study of the multiple-source co-channel interference under fading and shadowing and the determination of the reuse distance. In Section 3.2, we study the situation with co-channel interference in the absence of fading and shadowing. In Section 3.3, we investigate the characteristics of co-channel interference in a cellular system under fading and in Section 3.4, we include the effects of shadowing. Then in Section 3.5, we evaluate the situation when there is only shadowing. The phenomena of having fading only, shadowing only, fading and shadowing together and no fading and no shadowing at all are discussed in Section 3.6 using numerical examples. The performance of the system with the corner-illuminated configuration is also analysed in Section 3.7.

The analysis provided in the following sections is conducted only for the downlink channel. The same results would be obtained if the analysis is conducted for the uplink channel with the base station employing a single receive antenna. Since on the uplink channel, the mobile transmitter power is lower and is normally compensated by the higher sensitivity of the base station receiver, the analysis is basically the same for the uplink channel as for the downlink channel.

3.2 Co-channel Interference With No Fading And No Shadowing

Consider a land mobile cellular radio system model similar to the one described in Section 2.2. The mobile radio receives a desired signal of power, X, from its own base station as well as a number of unwanted signals of power, Y_i's, from the surrounding distant co-channel base stations. The mobile suffers co-channel interference if X does not exceed the sum of the Y_i's by the protection ratio q. The probability of co-channel interference is defined as the probability that:

$$X \leq q \sum_i Y_i \qquad \text{..... (3.1)}$$

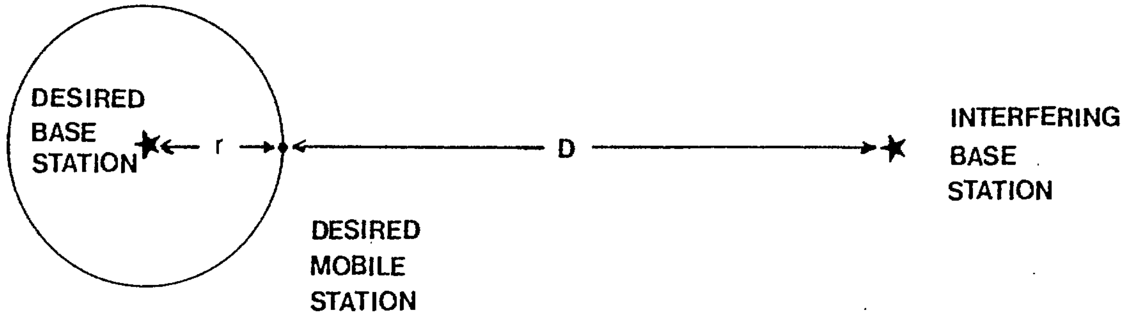
With no fading and no shadowing, X and Y_i may be found from a simple propagation law which states that the received signal power is inversely proportional to the p^{th} power of distance. In Fig. 3.1, the mobile station is on the edge of the coverage area of the cell. At this location, the desired signal is weakest while the interfering signal is strongest. If the cell radius is r and the distance between the interfering base station and the mobile station is D, then

$$X = K / r^P \qquad \text{..... (3.2)}$$

and $$Y = K / D^P \qquad \text{..... (3.3)}$$

where K is a propagation constant depending on antenna heights and the operating frequency and in this case is assumed to be the same for both transmitters.

Since D >> r, the distance between the desired mobile station and the other six co-channel interferors on the first ring around the desired base station is approximately equal to D



$$U = \frac{D + r}{r}$$

FIGURE 3.1 RELATIONSHIP BETWEEN CELL RADIUS r AND REUSE DISTANCE U as shown in Fig. 3.2. Hence all the Y_i's have the same value and (3.1) becomes,

$$D^p \leq q \cdot n \cdot r^p \qquad ..... (3.4)$$

where n is the number of co-channel interferors.

In terms of reuse distance U, we have from (2.1),

$$U = D_c / r = (D + r)/r$$

therefore (3.4) becomes

$$(U - 1)^p \leq q \cdot n$$

or $$U \leq 10^{(Q + 10 \log n)/10p} + 1 \qquad ..... (3.5)$$

where Q = 10log q.

It is interesting to study the variations of the minimum value of U or U_{min} with p for fixed values of Q and n. In Table 3.1, such variations are shown with p ranging from 3 to 4 for Q = 17 dB and n = 1 and 6.

Table 3.1 shows that the required reuse distance decreases with increasing values of p but increases with increasing values of n. It would be interesting to point out that p = 2 corresponds

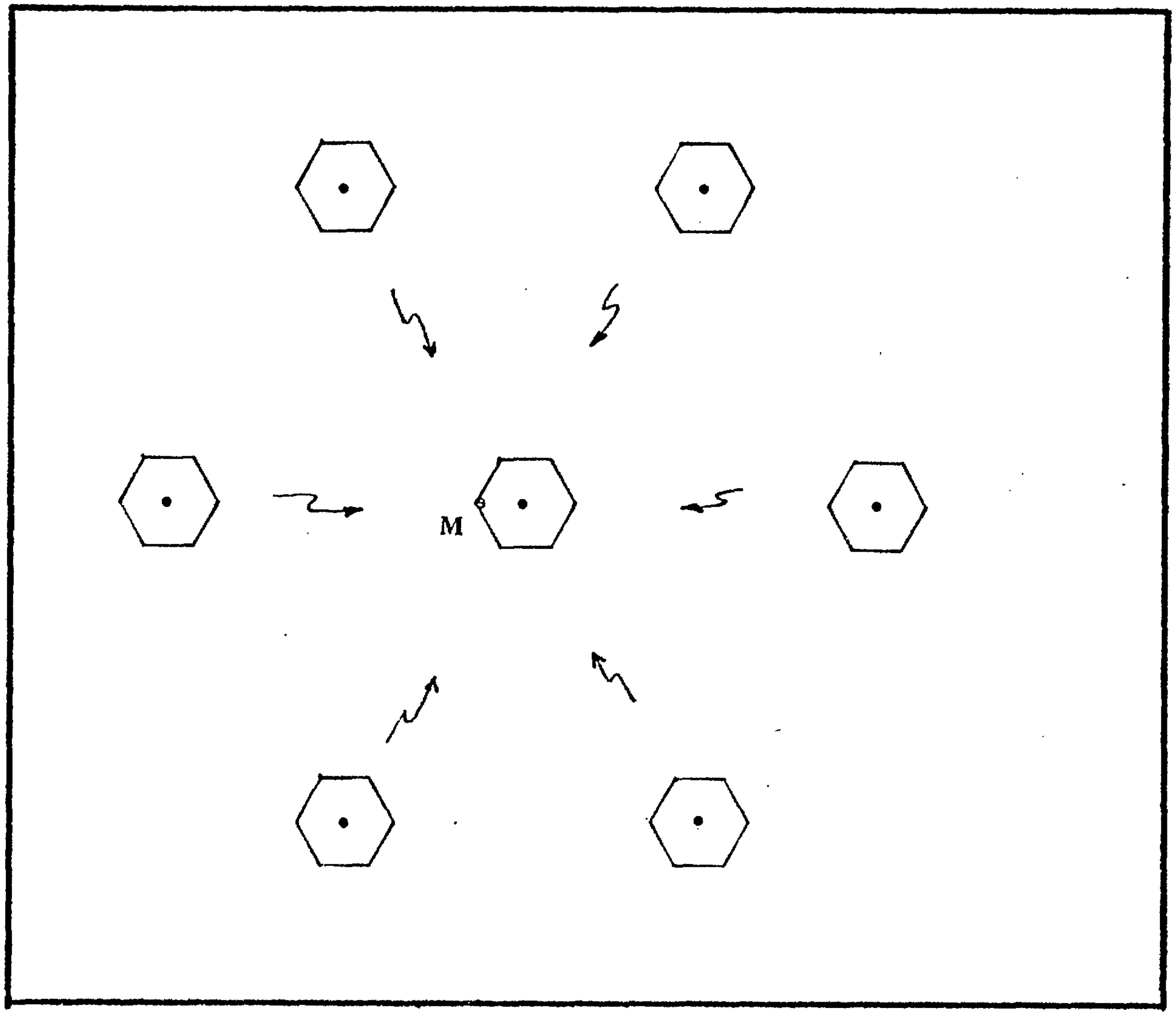


FIGURE 3.2 MULTIPLE-SOURCE CO-CHANNEL INTERFERENCE IN

A CELLULAR SYSTEM

to free space path loss which certainly cannot be applied to the land mobile environment. According to Hata in [HATA80], the value of p is quite independent of frequency but largely dependent on the height of the base station antenna. Low base station antenna heights result in large values of p.

p	Umin n = 1	Umin n = 6
3.0	4.69	7.70
3.2	4.40	6.95
3.4	4.16	6.36
3.6	4.00	5.88
3.8	3.80	5.49
4.0	3.66	5.16

TABLE 3.1

REUSE DISTANCE REQUIREMENTS AS A FUNCTION OF THE INDEX OF THE INVERSE POWER LAW

3.3 Probability of Co-channel Interference With Fading Only

In the presence of fading alone, the amplitude of the signal at the mobile receiver is Rayleigh distributed, and the power of the received signal is given by the probability density function [HANS77]:

$$P(X) = \begin{cases} 1/X_o \, . \, \exp(-X/X_o), \; X > 0 \\ 0, \; X \leq 0 \end{cases} \qquad \text{..... (3.6)}$$

where X = received signal power level

X_o = mean received signal power level

We consider first the case when the desired signal is interfered by a single co-channel base station. Since the interfering signal also suffers from fading, the probability that the interfering signal power level would exceed X/q for a given X is:

$$P(Y \geq X/q \text{ given } X) = \exp(-X/qY_o) \qquad \text{..... (3.7)}$$

where Y_o = mean power level of the undesired signal.

The probability of interference is obtained by integrating the above conditional probability over all values of X:

$$P_{F1} = \int_o^{\infty} P(Y \geq X/q \text{ given } X) . \, P(X) . \, dX$$

$$= \int_o^{\infty} \exp(-X/qY_o) \, . \, 1/X_o \, . \, \exp(-X/X_o).dX$$

$$= \quad qY_o/(X_o + qY_o) \qquad ..... (3.8)$$

In the case when there is more than one co-channel interferor, it is necessary to find the cummulative effect of all the n co-channel interferors. It is reasonable to assume that Y_o would be the same for all the interfering signals since they are at approximately the same distance from the mobile station and that each of the n interfering power levels is independent and identically distributed. Hence, the probability density function f (Z) is given by the convolution of the n probability density functions [PAPO65]:

$$f(Z) = \quad P(Y_1) * P(Y_2) * .... * P(Y_n) \qquad ..... (3.9)$$

where $Z = Y_1 + Y_2 + ... + Y_n$

and $P(Y_i) = 1/Y_o . \exp(-Y/Y_o)$

Taking the Laplace Transform,

$$f(s) = P(s) \, . \, P(s) \, ... \, P(s)$$

$$= \frac{1}{Y_o{}^n} \, . \, \frac{1}{(s + 1/Y_o)^n}$$

Hence, $$f(Z) = \frac{1}{Y_o{}^n} \, . \, \frac{Z^{n-1} \, . \, \exp(-Z/Y_o)}{(n-1)!} \qquad ..... (3.10)$$

which is a Gamma density function.

The probability that Z would exceed X/q given X is:

$$\text{Prob }(Z \geq X/q \text{ given } X) = \int_{\frac{X}{q}}^{\infty} f(Z) \cdot dZ$$

$$= \frac{\exp(-X/q\ Y_o)}{Y_o^n} \cdot \sum_{k=0}^{n-1} \left(\frac{X}{q}\right)^{n-1-k} \cdot Y_o^{k+1} \cdot \frac{1}{(n-1-k)!} \qquad \text{..........(3.11)}$$

Hence the probability P_{Fn} of co-channel interference under fading from n stations is:

$$P_{Fn} = \int_o^{\infty} P(X) \cdot \text{Prob}\left[Z \geq \frac{X}{q} \cdot \text{ given } X\right] \cdot dX$$

$$= \int_o^{\infty} \frac{1}{X_o} \exp(-X/X_o) \exp\left(-\frac{X}{qY_o}\right) \sum_{k=0}^{n-1} \left(\frac{X}{qY_o}\right)^{n-1-k} \cdot \frac{1}{(n-1-k)!} \cdot dX$$

$$= \sum_{k=0}^{n-1} \frac{q(Y_o/X_o)}{\left(1+ q\frac{Y_o}{X_o}\right)^{n-k}}$$

$$= \frac{q(Y_o/X_o)}{\left[1+ q(Y_o/X_o)\right]} + \text{---} + \frac{q(Y_o/X_o)}{\left[1+ q(Y_o/X_o)\right]^k} + \text{---} + \frac{q(Y_o/X_o)}{\left[1+ q(Y_o/X_o)\right]^n}$$

..............(3.12)

It may be interesting to note that the first term is the probability of interference due to the first interferor, the second term is the probability of interference due to the second interferor and so on.

3.4 Probability of Co-channel Interference with Fading and Shadowing

We consider now the effect of shadowing in addition to fading. For simplicity, we first consider a single term, say, the k^{th} term from (3.12) and express it in terms of Y_d and X_d, the dB values of Y_o and X_o respectively;

$$P_{Fn,k} = \frac{q(Y_o/X_o)}{(1+ q \frac{Y_o}{X_o})^k} = \frac{q\ 10^{\frac{Y_d-X_d}{10}}}{\left(1+ q\ 10^{\frac{Y_d-X_d}{10}}\right)^k} \qquad \text{..............(3.13)}$$

Under shadowing conditions, the mean power levels of the desired and interfering signals are both lognormally distributed. If we let X_{md} and Y_{md} be the medians in dB for X and Y respectively, the joint and individual probability density functions of X_d and Y_d will be given by:

$$f(X_d, Y_d) =$$

$$\frac{1}{2\pi\sigma_X\sigma_Y\sqrt{1-c}} \quad \exp - \frac{1}{2(1-c^2)} \left[\frac{(X_d-X_{md})^2}{\sigma_X^{\ 2}} - \frac{2c(X_d-X_{md})(Y_d-Y_{md})}{\sigma_X\sigma_Y} + \frac{(Y_d-Y_{md})^2}{\sigma_Y^{\ 2}} \right]$$

..............(3.14)

$$f(X_d) = \frac{1}{2\pi\ \sigma_X^{\ 2}} \quad \exp - \left[\frac{(X_d-X_{md})^2}{2\sigma_X^{\ 2}} \right] \qquad \text{..............(3.15)}$$

$$f(Y_d) = \frac{1}{2\pi\ \sigma_Y^{\ 2}} \quad \exp - \left[\frac{(Y_d-Y_{md})^2}{2\sigma_Y^{\ 2}} \right] \qquad \text{..............(3.16)}$$

where c is the correlation coefficient of X_d and Y_d, and σ_X, σ_Y are the standard deviations of X_d and Y_d respectively.

If $c = 0$, the desired and interfering signals are uncorrelated and $f(X_d, Y_d)$ becomes the product of $f(X_d)$ and $f(Y_d)$. In practice, the signals are slightly correlated since a small portion of the transmission path is common to both signals. The assumption that $c = 0$ gives the worst interference prediction. We further assume that $\sigma_X = \sigma_Y$, since shadowing is primarily a function of the topography near the mobile.

The probability of co-channel interference contributed by the k^{th} term under both fading and shadowing, $P_{Bn,k}$ is then:

$$P_{Bn,k} = \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} P_{Fn,k} \cdot f(X_d) \cdot f(Y_d) \cdot dX_d \cdot dY_d$$

$$= \frac{1}{2\pi\sigma^2}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} q.dX_d.dY_d \cdot \frac{10^{\frac{Y_d-X_d}{10}}}{\left(1+ q.10^{\frac{Y_d-X_d}{10}}\right)^k} \cdot \exp\left[-\frac{(X_d-X_{md})^2}{2\sigma^2} - \frac{(Y_d-Y_{md})^2}{2\sigma^2}\right]$$

By letting $Y_1 = Y_d - Y_{md}$

$Y_2 = X_d - X_{md}$

$$P_{Bn,k} = \frac{q}{2\pi\sigma^2} \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} dY_1 \cdot dY_2 \cdot \frac{10^{\frac{Y_1+Y_{md} - Y_2 - X_{md}}{10}}}{\left(1+q\ 10^{\frac{Y_1+Y_{md} - Y_2 - X_{md}}{10}}\right)^k} \cdot \exp\left(-\frac{Y_1^{\ 2}+Y_2^{\ 2}}{2\sigma^2}\right)$$

$$= \frac{q}{2\pi\sigma^2} \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} dY_1 \cdot dY_2 \cdot \frac{Y_m}{X_m} \cdot \frac{10^{\frac{Y_1-Y_2}{10}}}{\left(1+ q\frac{Y_m}{X_m}\ 10^{\frac{Y_1-Y_2}{10}}\right)^k} \cdot \exp\left(-\frac{Y_1^{\ 2}+Y_2^{\ 2}}{2\sigma^2}\right)$$

By rotational transformation and letting

$$Y_1 = \frac{X_1+X_2}{\sqrt{2}}$$

$$Y_2 = \frac{X_1-X_2}{\sqrt{2}}$$

$$R = X_m/Y_m$$

the Jacobian of the transformation $J(Y_1\ ,\ Y_2)$ according to [PAP065] is given by:

$$J(Y_1, Y_2) = \begin{vmatrix} \frac{\partial X_1}{\partial Y_1} & \frac{\partial X_1}{\partial Y_2} \\ \frac{\partial X_2}{\partial Y_1} & \frac{\partial X_2}{\partial Y_2} \end{vmatrix} = \begin{vmatrix} \frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \\ -\frac{1}{\sqrt{2}} & \frac{1}{\sqrt{2}} \end{vmatrix} = 1$$

$$\therefore \quad P_{Bn,k} = \frac{q}{2\pi\sigma^2} \cdot \frac{1}{R} \cdot \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} dX_1 \cdot dX_2 \cdot \frac{10^{\frac{\sqrt{2}X_2}{10}}}{\left(1+\frac{q}{R}\cdot 10^{\frac{\sqrt{2}X_2}{10}}\right)^k} \cdot \exp\left(-\frac{X_1^{\,2}+X_2^{\,2}}{2\sigma^2}\right)$$

$$= \frac{q}{2\pi\sigma^2 R} \int_{-\infty}^{\infty} \frac{10^{\frac{\sqrt{2}X_2}{10}}}{\left(1+\frac{q}{R}\cdot 10^{\frac{\sqrt{2}X_2}{10}}\right)^k} \cdot \exp\left(-\frac{X_2^{\,2}}{2\sigma^2}\right) dX_2 \int_{-\infty}^{\infty} \exp\left(-\frac{X_1^{\,2}}{2\sigma^2}\right) \cdot dX_1$$

$$= \frac{q}{\sqrt{2\pi}\ \sigma R} \int_{-\infty}^{\infty} \frac{10^{\frac{\sqrt{2}X_2}{10}}}{\left(1+\frac{q}{R}\, 10^{\frac{\sqrt{2}X_2}{10}}\right)^k} \cdot \exp\left(-\frac{X_2^{\,2}}{2\sigma^2}\right) \cdot dX_2$$

To simplify, we let $u = \dfrac{X_2}{\sqrt{2}\ \sigma}$

$$P_{Bn,k} = \frac{q}{\sqrt{\pi}\ R} \int_{-\infty}^{\infty} \frac{10^{\frac{\sigma u}{5}}}{\left(1+\frac{q}{R}\cdot 10^{\frac{\sigma u}{5}}\right)^k} \cdot \exp\ (-u^2) \cdot du \qquad \ldots\ldots\ldots\ldots\ldots(3.17)$$

$$\therefore \quad P_{Bn} = \sum_{k=1}^{n} P_{Bn,k}$$

$$= \sum_{k=1}^{n} \frac{q}{\sqrt{\pi}\ R} \int_{-\infty}^{\infty} \frac{10^{\frac{\sigma u}{5}}}{\left(1+\frac{q}{R}\, 10^{\frac{\sigma u}{5}}\right)^k} \cdot \exp\ (-u^2) \cdot du \qquad \ldots\ldots\ldots\ldots\ldots(3.18)$$

(3.18) can also be expressed in terms of the reuse distance U. According to the inverse p^{th} power law which states that the received power level is inversely proportional to the p^{th} power of distance,

$$R = X_m/Y_m = (D/r)^p \qquad \text{..............(3.19)}$$

and $$U = 1 + R^{1/p}$$

or $$R = (U-1)^p \qquad \text{..............(3.20)}$$

(3.18) therefore becomes:

$$P_{Bn} = \sum_{k=1}^{n} \frac{1}{\sqrt{\pi}} \frac{q}{(U-1)^p} \cdot \int_{-\infty}^{-\infty} \frac{10^{\frac{\sigma u}{5}} \exp(-u^2)}{\left[1 + \frac{q}{(U-1)^p} \cdot 10^{\frac{\sigma u}{5}}\right]^k} \cdot du$$

..............(3.21)

The integral in (3.1.21) can only be solved by numerical integration. The values of P_{Bn} for different values of σ, n and q (or Q in dB) are plotted against the reuse distance U in Fig. 3.3, 3.4 and 3.5 with p = 4.

Figs. 3.3 and 3.4 show the relationship between the probability of co-channel interference and reuse distance for n = 1 and n = 6 respectively. In both cases, the co-channel interference is largely affected by σ. It should be noted that σ = 0 corresponds to the situation where there is only fading and no shadowing since the local mean power deviation from the median power is zero.

The other parameter that has an impact on the co-channel interference is the protection ratio. From Fig. 3.4, with n = 6 and σ = 6 dB, the probability of co-channel interference increases about 5-fold when Q increases from 8 to 17 dB.

Fig. 3.5 shows the effect of the number of co-channel interferors on the probability of interference for Q = 17 dB and σ = 6 dB. In a mature cellular radio system with the exception of the outer cells, there are normally six interferors on the first ring of co-channel cells around the desired base station.

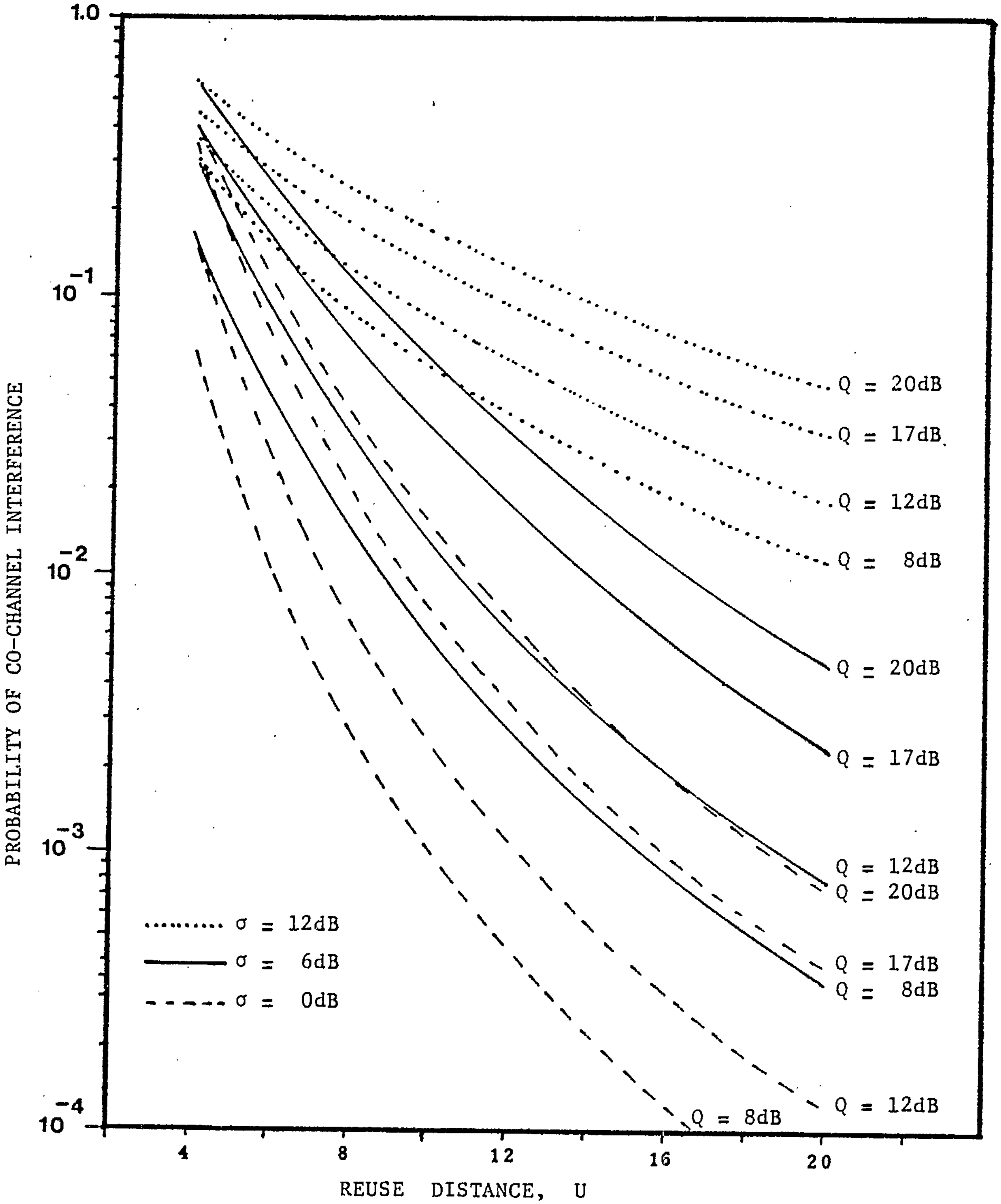


FIGURE 3.3 PROBABILITY OF CO-CHANNEL INTERFERENCE VS REUSE DISTANCE UNDER FADING AND SHADOWING WITH n=1

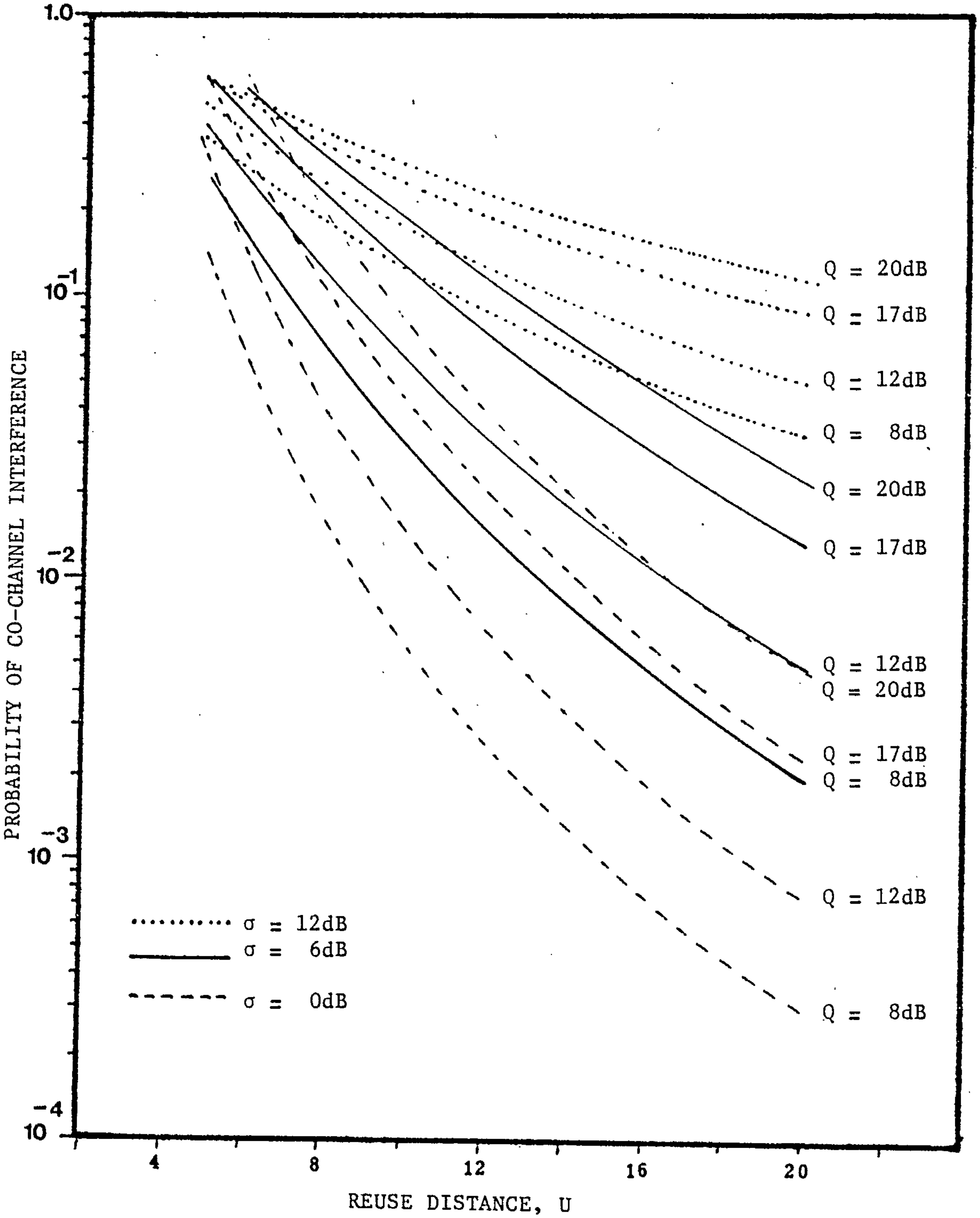


FIGURE 3.4 PROBABILITY OF CO-CHANNEL INTERFERENCE IN A CELLULAR SYSTEM UNDER FADING AND SHADOWING WITH n=6

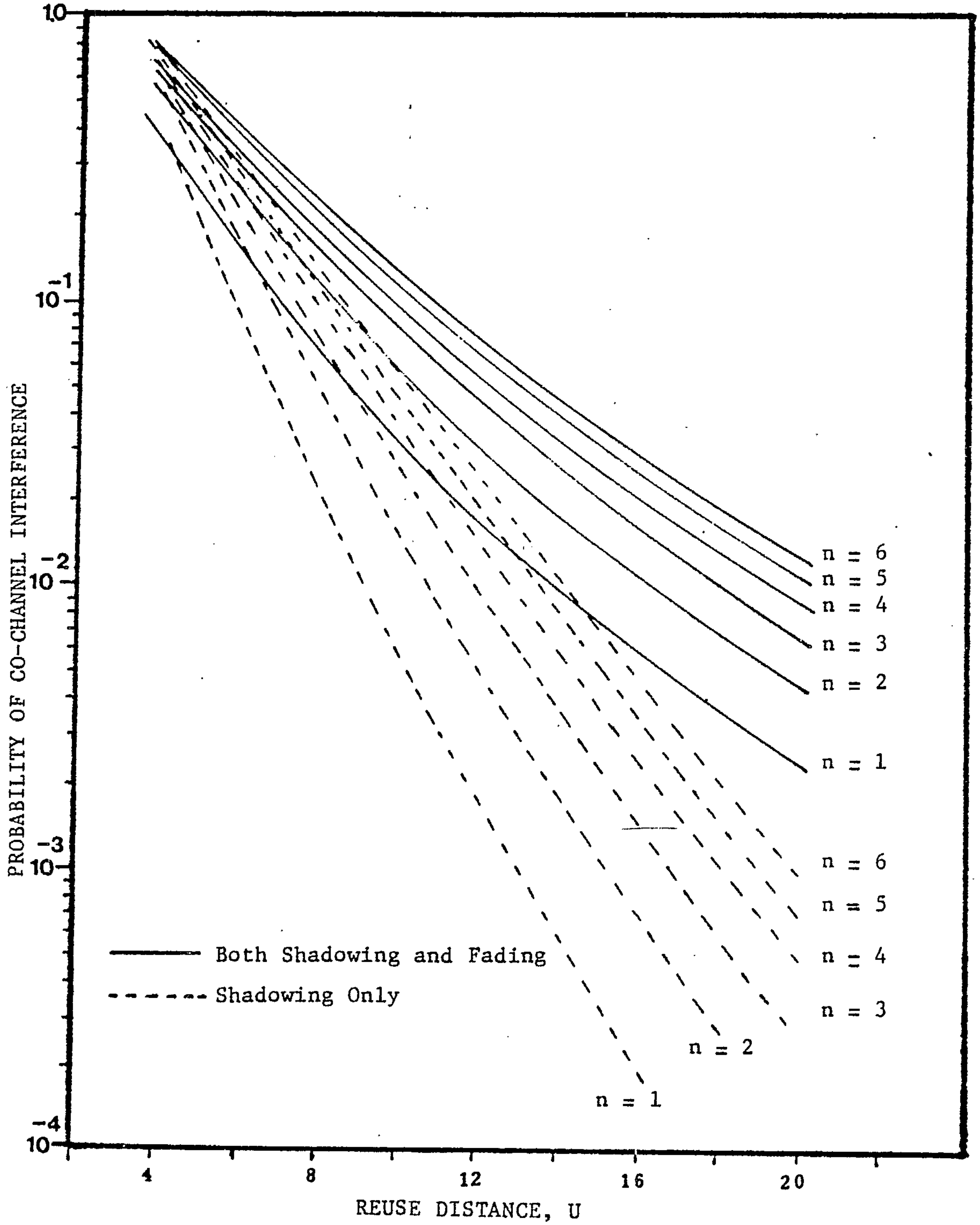


FIGURE 3.5 PROBABILITY OF CO-CHANNEL INTERFERENCE IN A CELLULAR SYSTEM UNDER FADING AND SHADOWING AND SHADOWING ONLY WITH Q=17dB AND σ = 6dB

3.5 Probability of Co-channel Interference with Shadowing Only

With shadowing only, the instantaneous received power level X is the same as the local mean power level X_o, and has a lognormal distribution. We first derive the expression for co-channel interference for one interferor and then extend it to include multiple interferors.

For a single interferor, interference occurs if $X \leqq Y + Q$ or $Y \geqq X - Q$. Therefore the probability of interference P_{S1} under shadowing only is,

$$P_{S1} = \int_{-\infty}^{\infty} P(X_d) \cdot \int_{X_d-Q}^{\infty} P(Y_d \text{ given } X_d) \cdot dY_d \cdot dX_d \qquad \text{..............(3.22)}$$

$$= \int_{-\infty}^{\infty} \frac{1}{\sqrt{2\pi\sigma^2}} \cdot \exp \frac{-(X_d-X_{md})^2}{2\sigma^2} \int_{X_d-Q}^{\infty} \frac{1}{\sqrt{2\pi\sigma^2}} \exp\left[\frac{-(Y_d-Y_{md})^2}{2\sigma^2}\right] . dY_d . dX_d$$

By changing variable and letting $w = Y_d - (X_d - Q)$,

$$J(X_d , Y_d) = \begin{vmatrix} \frac{\partial X_d}{\partial X_d} & \frac{\partial X_d}{\partial Y_d} \\ \\ \frac{\partial W}{\partial X_d} & \frac{\partial W}{\partial Y_d} \end{vmatrix} = \begin{vmatrix} 1 & 0 \\ -1 & 1 \end{vmatrix} = 1$$

$$\therefore \quad P_{S1} = \frac{1}{2\pi\sigma^2} \int_{-\infty}^{\infty} \exp\left[-\frac{(X_d-X_{md})^2}{2\sigma^2}\right] \int_{o}^{\infty} \exp\left[\frac{-(X_d+w-Y_{md}-Q)^2}{2\sigma^2}\right] . dw . dX_d$$

$$= \frac{1}{2\pi\sigma^2} \int_{-\infty}^{\infty} dX_d . \int_{o}^{\infty} \exp -\left[\frac{(X_d-X_{md})^2}{2\sigma^2} + \frac{(X_d+w-Y_{md}-Q)^2}{2\sigma^2}\right] . dw$$

$$= \frac{1}{2\pi\sigma^2} \int_{-\infty}^{\infty} dX_d . \int_{o}^{\infty} dw . \exp - \frac{1}{\sigma^2}\left[(X_d-X_{md}) + \frac{(w+X_{md}-Y_{md}-Q)}{2}\right]^2 - \frac{(w+X_{md}-Y_{md}-Q)^2}{4\sigma^2}$$

We further let $t = (X_d-X_{md}) + \dfrac{(w+X_{md}-Y_{md}-Q)}{2}$

$$J(X_d,w) = \begin{vmatrix} \frac{\partial t}{\partial X_d} & \frac{\partial t}{\partial w} \\ \frac{\partial w}{\partial X_d} & \frac{\partial w}{\partial w} \end{vmatrix} = \begin{vmatrix} 1 & \frac{1}{2} \\ 0 & 1 \end{vmatrix} = 1$$

$$\therefore \quad P_{S1} = \frac{1}{2\pi\sigma^2} \int_{-\infty}^{\infty} dt \int_{o}^{\infty} dw . \exp\left[-\frac{t^2}{\sigma^2} - \frac{(w+X_{md}-Y_{md}-Q)^2}{4\sigma^2}\right]$$

$$= \frac{1}{2\pi\sigma^2} \int_{-\infty}^{\infty} \exp\left(-\frac{t^2}{\sigma^2}\right) . dt \int_{o}^{\infty} \exp \frac{-(w+X_{md}-Y_{md}-Q)^2}{4\sigma^2} . dw$$

$$= \frac{1}{2\sqrt{\pi}\sigma} \int_{o}^{\infty} \exp\left[\frac{-(w+X_{md}-Y_{md}-Q)^2}{4\sigma^2}\right] . dw$$

By letting $v = \dfrac{w+X_{md}-Y_{md}-Q}{2\sigma}$

$$P_{S1} = \frac{1}{\sqrt{\pi}} \int_{\frac{X_{md}-Y_{md}-Q}{2\sigma}}^{\infty} \exp(-v^2) \cdot dv \qquad \text{..............(3.23)}$$

For multiple interferors, co-channel interference would occur if $Z_{md} \geq X - Q$ where Z_{md} is the sum of the median power levels of the n interfering signals, then

$$Z_{md} = 10\log(nY_m)$$

$$Z_{md} = 10\log n + Y_{md} \qquad \text{..............(3.24)}$$

From (3.23) and (3.24), we get

$$P_{Sn} = \frac{1}{\sqrt{\pi}} \int_{\frac{X_{md}-Y_{md}-10\log n - Q}{2\sigma}}^{\infty} \exp(-v^2) \cdot dv \qquad \text{..............(3.25)}$$

$$= \frac{1}{\sqrt{\pi}} \int_{L}^{\infty} \exp(-v^2) \cdot dv \qquad \text{..............(3.26)}$$

The lower limit of integration L may be expressed in terms of the reuse distance U,

$$L = \frac{X_{md}-Y_{md} - 10\log n - Q}{2\sigma}$$

$$= \frac{10\ p\ \log\ (U-1) - 10\log n - Q}{2\sigma} \qquad \text{..............(3.27)}$$

Values of P_{Sn} are plotted in Figs. 3.5, 3.6, 3.7 for different values of Q, n, and σ against the values of U with p = 4. These Figures show that the behaviour of co-channel interference with shadowing only is similar to that with both fading and shadowing.

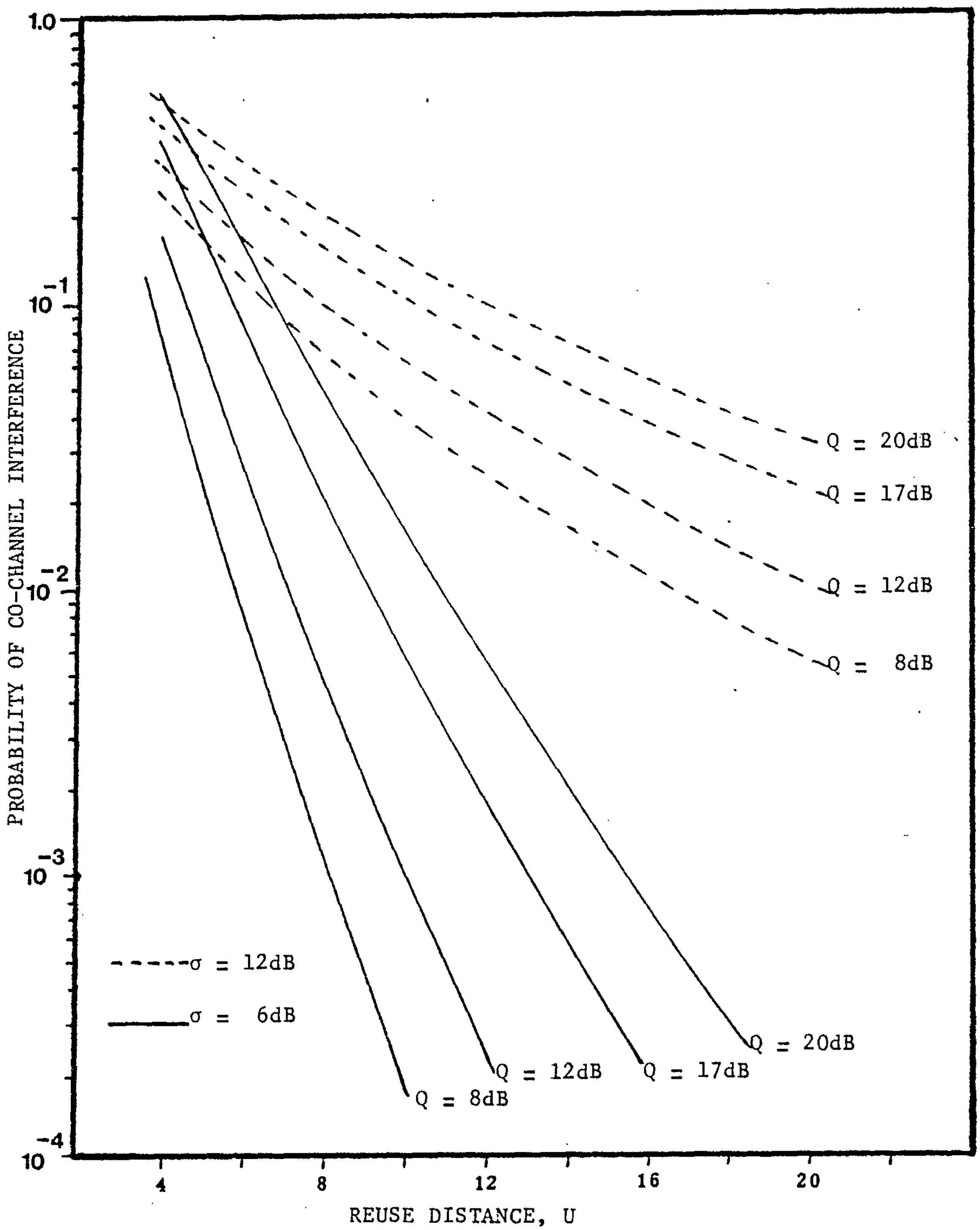


FIGURE 3.6 PROBABILITY OF CO-CHANNEL INTERFERENCE IN A CELLULAR SYSTEM UNDER SHADOWING ONLY WITH n=1

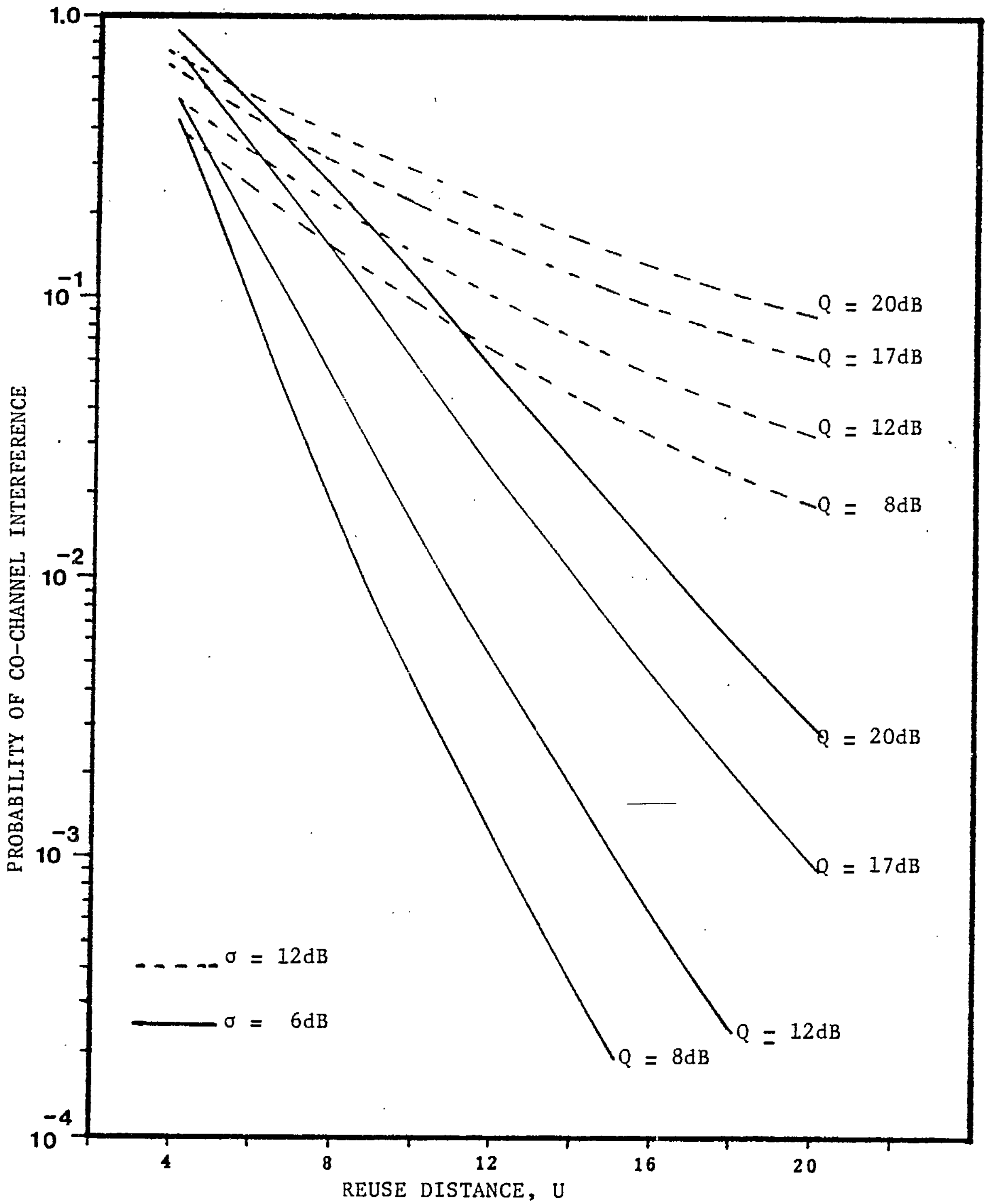


FIGURE 3.7 PROBABILITY OF CO-CHANNEL INTERFERENCE IN A CELLULAR SYSTEM UNDER SHADOWING ONLY WITH n=6

3.6 Discussion of Numerical Results

In the previous sections, we have derived expressions to calculate the probability of co-channel interference for the cases in which there is:

(1) no fading and no shadowing;
(2) fading only;
(3) both fading and shadowing;
(4) shadowing only.

For comparison purposes, we have extracted the data from Figs. 3.4 and 3.7 and plotted the behaviour of cases (2), (3) and (4) for n = 6 and Q = 17 dB in Fig. 3.8. It can be seen that the shadowing only curve with σ = 6 dB is in close proximity to the fading only curve (σ = 0 dB). But the shadowing only curve with σ = 12 dB shows much severe co-channel interference than the fading only curve. Furthermore, the fading and shadowing curve is above the shadowing only curve by about the same amount for σ = 6 dB as well as for σ = 12 dB.

We can now compare the four cases with fixed values of Q, n, σ and probability of co-channel interference using reuse distance, number of cells per cluster and spectrum efficiency as parameters. We know from Chapter 2 that N may take on only certain values as specified by (2.2). We therefore need to establish the appropriate values of N with respect to U. Table

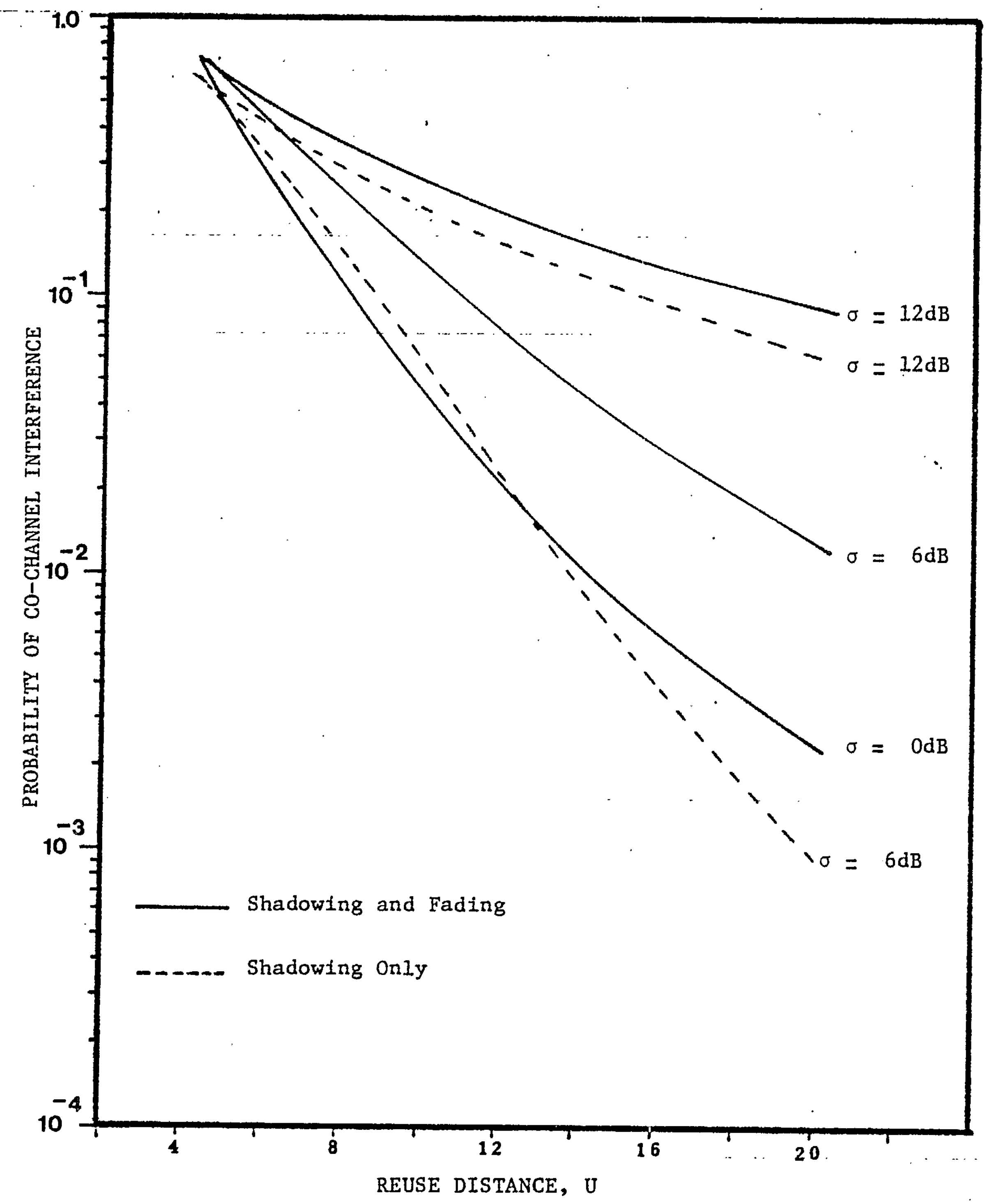


FIGURE 3.8 PROBABILITY OF CO-CHANNEL INTERFERENCE IN A CELLULAR SYSTEM UNDER SHADOWING AND FADING AND SHADOWING ONLY WITH n=6 AND Q = 17dB

g	h	$N = g^2 + gh + h^2$	U
0	0	0	0
0	1	1	1.73
1	1	3	3
0	2	4	3.46
1	2	7	4.58
0	3	9	5.20
2	2	12	6
1	3	13	6.24
0	4	16	6.92
2	3	19	7.55
1	4	21	7.94
0	5	25	8.66
3	3	27	9
2	4	28	9.17
1	5	31	9.64
0	6	36	10.39
3	4	37	10.54
2	5	39	10.82
1	6	43	11.35

TABLE 3.2

VALUES OF SHIFT PARAMETERS, CLUSTER SIZE AND REUSE DISTANCE

3.2 tabulates the values of the shift parameters, g and h, and the corresponding values of N and U in ascending order of N varying from N = 0 to N = 43.

For the comparison, we have constructed Table 3.3 for Q = 17 dB, σ = 6 dB, probability of co-channel interference $\leq$ 0.1 and n = 1, 2 and 6. The spectrum efficiency E is expressed in terms of the constant H which is a function of the traffic density Vt, the channel spacing f_s and the number of channels per cell, m as in (2.38).

3.7 Performance Of The Corner-Illuminated Cellular System

The above analysis has been conducted for the centre-illuminated cellular system. The other alternative mentioned in the literature [MACD79] is the corner-illuminated system which has a smaller reuse distance requirement. In this configuration, the base station transmitters are located at alternate corners of a cell 120 degrees apart as shown in Fig. 3.9. This provides the system with the capability to eliminate some shadowing loss since a mobile may be switched to another channel if it is in a shadowed area. In the following, we attempt to investigate the performance of the corner-illuminated system in terms of reuse distance and compare its performance with that of the centre-illuminated system in terms of reuse distance.

Cases	n = 1			n = 2			n = 6		
	U	N	E *	U	N	E	U	N	E
(1)	3.66	7	14.29	4.16	7	14.29	5.16	9	11.11
(2)	5.60	12	8.33	6.30	16	6.25	8.20	25	4.00
(3)	7.20	19	5.26	8.60	25	4.00	11.00	43	2.33
(4)	6.00	12	8.33	6.80	16	6.25	8.70	27	3.70

Notes: For Q = 17 dB, σ = 6 dB and probability of co-channel interference $\leq$ 0.1,

Cases (1) = with no fading and no shadowing

(2) = with fading only

(3) = with both fading and shadowing

(4) = with shadowing only

* From (2.38), $E = Vt/f_s.m.N = 100.H/N$ (Erl./MHz/sq. km.)

where $H = Vt/100.f_s.m$ (Erl. /MHz/sq. km.)

TABLE 3.3

VALUES OF REUSE DISTANCE AND CLUSTER SIZE

AND SPECTRUM EFFICIENCY

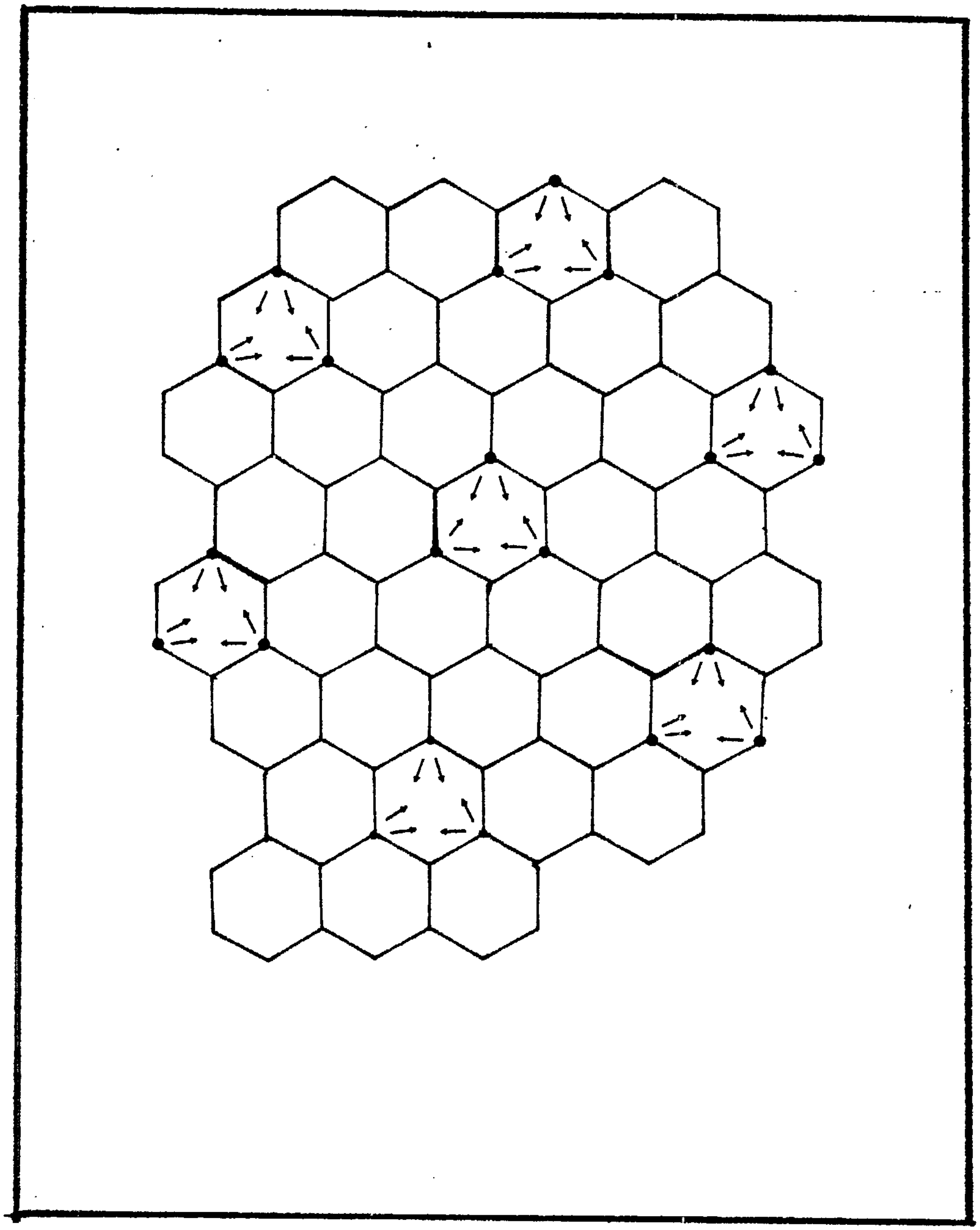


FIGURE 3.9 THE CORNER-ILLUMINATED CELLULAR SYSTEM

In a corner-illuminated configuration, the relationship between the reuse distance and the ratio R ($R = X_m/Y_m$) is different from that in the centre-illuminated configuration. As shown in Fig. 3.10, the position at which the mobile station receives the weakest desired signal is at the centre of the cell, assuming that the transmitter covers only one-third of the cell. Hence,

$$R = \frac{X_m}{Y_m} = \left(\frac{\text{Distance between interfering base station and desired mobile station}}{\text{cell radius}} \right)^p$$

$$\leq \left(\frac{D_c' + r}{r} \right)^p = (U_c + 1)^p \qquad \text{..... (3.28)}$$

as opposed to $R = (U - 1)^p$ in (3.20). U_c is the reuse distance required in the corner-illuminated configuration.

Since the locations of the interfering base station, desired base station and mobile station are not collinear, (3.28) represents the upper bound of the value of R. By comparing (3.28) and (3.20), it can be seen that (3.28) has the same effect as sliding the scale of U in Figs. 3.3 to 3.8 to the right by up to two units for a given probability of co-channel interference. This is equivalent to obtaining the required reuse distance U_c by deducting up to two units from the value of U obtained from the graphs for the centre-illuminated system.

Furthermore, the number of interferors is reduced to only

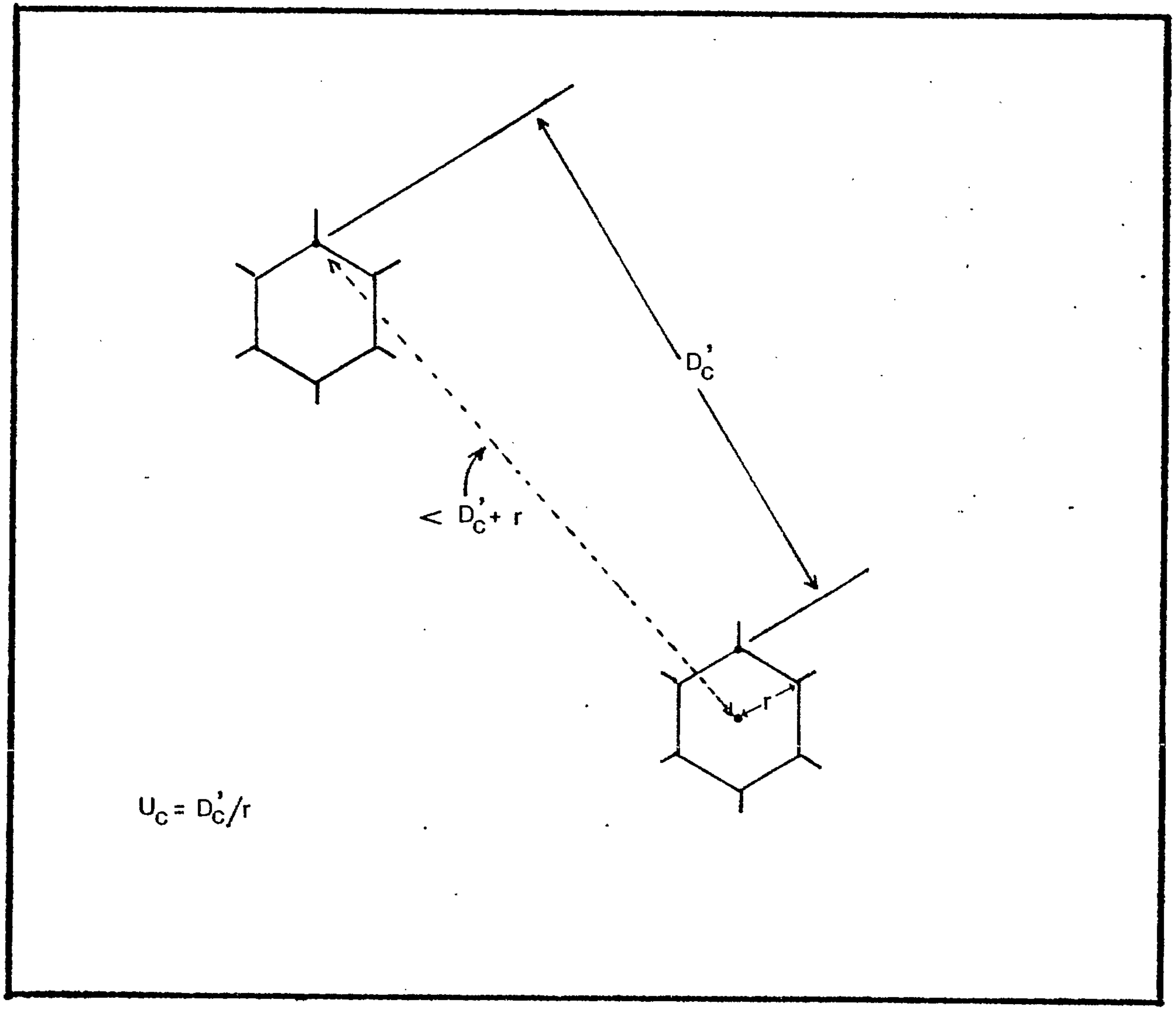


FIGURE 3.10 RELATIONSHIP BETWEEN CELL RADIUS AND REUSE DISTANCE IN A CORNER-ILLUMINATED SYSTEM

two as can be seen from Fig. 3.9. The data for the required reuse distance are obtained from Table 3.3 and are shown in Table 3.4.

Unfortunately, despite the fact that the conversion from the centre-illuminated configuration to the corner-illuminated configuration produces a lower reuse distance requirement, it does not reduce N , the number of cells in a cluster. Instead, the number of channels required per cluster would increase, bringing the spectrum efficiency down. This is illustrated in an example provided by MacDonald in [MACD79] in which the 12-cell structure requires a 21-channel set with the corner-illuminated configuration while it requires only a 12-channel set with the centre-illuminated configuration. Other examples which demonstrate this behaviour are shown in Table 3.5.

3.8 Summary of Results

In the above sections, we have examined the system performance in terms of the reuse distance requirements for the four different cases in the centre-illuminated and corner-illuminated configurations. These cases are: (1) no fading and no shadowing; (2) fading only; (3) fading and shadowing and (4) shadowing only. The results are summarized in Tables 3.3 and 3.4. From the results obtained, we can see that both fading and shadowing increase the probability of co-channel interference. In order to limit the co-channel interference to a certain level

Cases	approximated value of U_c	N
(1)	2.16	7
(2)	4.30	16
(3)	6.6	25
(4)	4.8	16

Notes: $n = 2$, $\sigma = 6$ dB, $Q = 17$ dB, and probability of co-channel interference ≤ 0.1

Cases (1) = with no fading and no shadowing

(2) = with fading only

(3) = with both fading and shadowing

(4) = with shadowing only

TABLE 3.4

VALUES OF REUSE DISTANCE AND CLUSTER SIZE FOR A CORNER-ILLUMINATED CELLULAR SYSTEM

N	No. of channels per cluster, M required in the corner-illuminated configuration	Ratio M/N
3	3	1
7	9	1.29
9	12	1.33
12	21	1.75
19	36	1.89

TABLE 3.5

CHANNEL REQUIREMENTS IN A CORNER-ILLUMINATED CELLULAR SYSTEM under fading and shadowing conditions, the reuse distance, and hence N, has to be increased. This in turn results in a lower spectrum efficiency of the system.

We have also compared our results with the results obtained by both French [FREN79] and Muammar [MUAM82]. In comparison with French's results, with a single co-channel interferor, our results for fading only, fading and shadowing, and shadowing only agree extremely well with French's data. In comparison with Muammar's results, with a single co-channel interferor, our results for fading only and fading and shadowing agree extremely well with Muammar's data. However, with six co-channel interferors, Muammar's data are more pessimistic than our data. For example, for Q = 17 dB and both σ = 6 dB and 12 dB, Muammar predicts a probability of co-channel interference which is about three times higher than our predictions.

We indicated that by using the corner-illuminated configuration, the shadowing effect may be reduced. A number of workers such as Jakes [JAKE74], Kelly and Ward [KELL73], Leung [LEUN81] and Hata [HATA81] suggested diversity techniques to combat Rayleigh fading. They showed that the probability of error due to multipath fading is much lower if diversity is introduced.

Finally, it must be pointed out that cellular radio system in reality may perform better than indicated by the results of Tables 3.3 and 3.4. In deriving the probability of interference, a correlation coefficient of zero between the desired and interfering signals is assumed, representing the worst case situation. Another possibility is that the number of interferors may be less than six since the topography may not allow such an arrangement to exist.

CHAPTER 4 - INTERFERENCE ANALYSIS OF A CELLULAR RADIO SYSTEM

4.1 Introduction

In conventional land mobile systems, it has become difficult to assign an interference-free frequency to a radio user especially in populated areas with a large number of radio users. In order to assign a frequency within a reasonable time frame, computer assistance has been introduced to alleviate the tedious tasks of interference analyses and calculations that are needed in the assignment process. These efforts were described by Decouvreur et al in [DECO81], DeMercado et al in [DEME82] and Chan in [CHAN83].

In view of the amount of human and computing effort that has to be spent in processing a frequency application, studies have been conducted to investigate if such lengthy analyses could be avoided. This idea, which seems quite impossible for the random installations of existing radio stations, would indeed be possible with the introduction of new spectrum-efficient cellular systems in which advanced planning in the geographical distribution of low-powered transmitters must be made. A frequency allocation plan can also be developed for such a system so that subsequent interference analysis would not be necessary. However, before such a frequency allocation plan can be developed, the interference situations between cells and within each cell must be analysed.

The three most common types of interference occur in the form of co-channel, adjacent channel and intermodulation interference. Co-channel interference discussed in Chapter 3 may be controlled by selecting the appropriate reuse distance. In this chapter, we consider the adjacent channel and intermodulation interference which have significant impacts on the frequency allocation plan for the system.

Section 4.2 of this chapter deals with the analysis of adjacent channel interference while Sections 4.3, 4.4 and 4.5 deal with the analysis of transmitter and receiver intermodulation interference. In both types of interference, intra-cell and inter-cell interference situations are considered. The results of the analysis are used as inputs to Chapter 5 in which an interference-free frequency allocation scheme is presented.

4.2 Adjacent Channel Interference Analysis

The analysis of the effect of adjacent channel interference on a mobile radio channel was discussed by a number of authors over the past few years. McMahon in [MCMA74] provided an adjacent channel model in which the side-band noise of transmitters may be predicted. Gosling in [GOSL80] described a situation in which two base stations TX_1 and TX_2 separated by a certain distance from each other are transmitting to their respective mobile stations on adjacent channels. In this situation, the mobile station communicating with TX_1 would experience adjacent channel interference if it is operating in an area called a 'high risk zone' close to TX_2. In this section, we attempt to analyse the intra-cell and inter-cell adjacent channel interference situation in a cellular system. In the first situation, we assume that the base station at the centre of the cell is assigned n consecutive channels for communications with the mobile stations. In the second situation, we assume that the base stations in adjacent cells are assigned adjacent channels.

4.2.1 Intra-cell Adjacent Channel Interference Analysis

Adjacent channel interference in the intra-cell situation arises from the possibility that the desired signal path may be much longer than the interfering signal path. This phenomenon would happen when mobile stations are transmitting on the uplink channels from different locations in the cell to the base station. Because of the difference in distance, the interfering signal power level, Y, may be higher than the desired signal power level, X at the desired received carrier frequency. Adjacent channel interference would occur if:

$$Y \geq X - Q + J \qquad \text{...... (4.1)}$$

where Q is the protection ratio and J is the amount of attenuation that the interfering adjacent channel signal would suffer at the carrier frequency of the desired signal. All parameters are in dB's.

However, when the base station is transmitting to the mobile stations on the downlink channels, the desired signal path is equal to the interfering signal path and the propagation characteristics for both paths are the same. Therefore Y is same as X and adjacent channel interference would occur if:

$$Q \geq J \qquad \ldots\ldots (4.2)$$

The value of J varies from one transmitter to another. In this analysis, we have chosen J to be equal to 26 dB * [DOC83]. The signal to interference ratio (S/I) for acceptable signal quality as suggested by MacDonald in [MACD79] is 17 dB. Hence if we assume Q to be 20 dB including an allowance of 3 dB on top of the suggested 17 dB to account for other types of interference which would add to the total amount of interference, (4.2) will not be satisfied even with both adjacent channel signals on at the same time. We can therefore deduce that adjacent channel interference would not occur on the downlink channel.

* We also assume that the out-of-band emission power level starts to drop off at the edge of the 30-KHz-bandwidth channel.

The situation is different with the transmissions from the mobiles to the base station on the uplink channels. Because of the mobility of the vehicles, an interfering transmitter may be much closer to the base station receiver than the desired transmitter. Based on the system model described in Chapter 2 and assuming that all mobile stations have the same transmitter output power, P_m, the desired signal received power level, P_d, is (note that all the variables of power levels, path losses and antenna gains in this chapter are in dB):

$$P_d = P_m - P_{ld} \qquad \ldots\ldots (4.3)$$

and the interfering adjacent channel signal received power level, P_{adj}, is:

$$P_{adj} = P_m - P_{li} \qquad \ldots\ldots (4.4)$$

where P_{ld} and P_{li} are the path losses from the desired mobile and the interfering mobile to the base station respectively. For the worst case consideration, we assume that the desired mobile is on the edge of the cell at a distance r from the base station and the interfering mobile is at a distance d smaller than r from the base station. Using the path loss equation (2.10) in Section 2.3, (4.3) and (4.4) become:

$$P_d - P_{adj} = P_{li} - P_{ld}$$

or $\quad P_d - P_{adj} = 40\log d - 40\log r$

For adjacent channel interference to occur,

$$P_{adj} \geq P_d - Q + J$$

i.e. $\quad 40\log d - 40\log r \leq Q - J \qquad \text{..... (4.5)}$

or $\quad d \leq r.10^{(Q-J)/40} \qquad \text{..... (4.6)}$

This means that if the interfering mobile station is at a distance d from the base station, interference would occur.

Given that the desired mobile is transmitting, the probability, P_a, that one of the (n-1) remaining channels is active within the interference area with radius d may be found and is given by:

$$P_a = \frac{\text{total traffic in interference area}}{\text{total number of available channels}}$$

$$= \pi d^2 Vt/(n-1) \qquad \text{..... (4.7)}$$

It must be noted that the value of n in (4.7) is such that a grade of service of 0.02 is maintained for the cell with radius r. Under this condition, P_a would not exceed unity. Since the desired mobile may be operating in any one of the n consecutive channels as depicted in Fig. 4.1 where f_0, f_1, ... f_{n-1} are carrier frequencies of channels 0, 1, ..., n-1 respectively and each channel is flanked by two adjacent channels except at the two ends of the frequency band, the probability of adjacent channel interference, P_{AD}, is :

$$P_{AD} = P_a.2/n + (2.P_a - P_a^2).(n-2)/n \qquad ..... (4.8a)$$

$$= P_a . [(n-2)(2-P_a) + 2]/n \qquad ..... (4.8b)$$

The first term in (4.8a) is the probability that an adjacent channel is on when the desired channel is an end channel. The second term is the probability that either one or both adjacent channels are on when the desired channel is not an end channel.

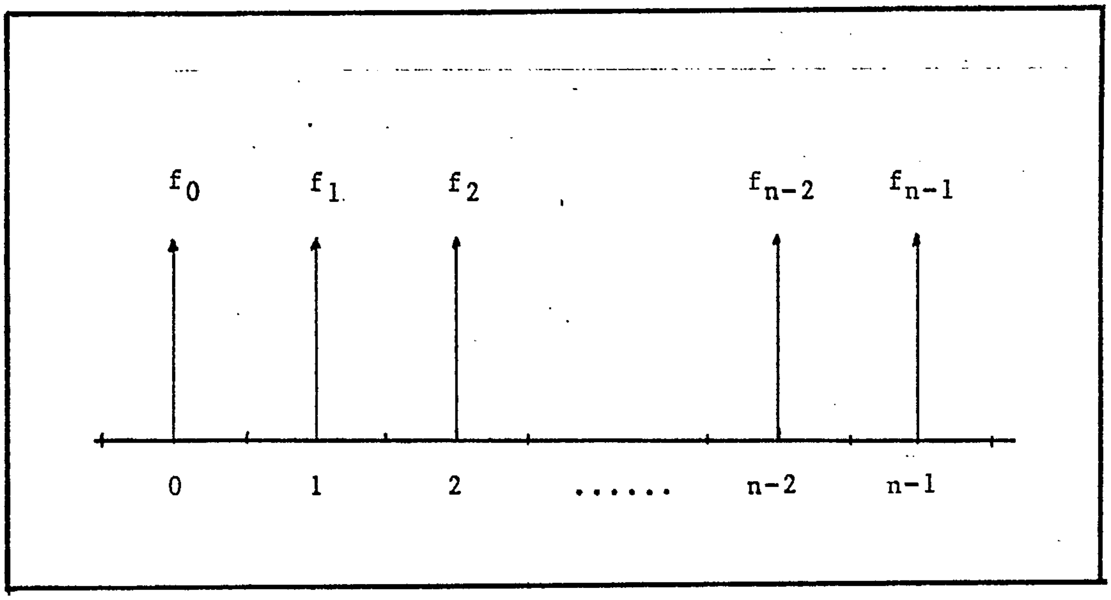


FIGURE 4.1

ALLOCATION OF n CONSECUTIVE CHANNELS TO A CELL SITE

It should be pointed out that the value of d is obtained based on the assumption that only one adjacent channel is on. With both adjacent channels on, the value of d is slightly larger. Hence P_{AD} should be slightly higher than the value predicted in (4.9).

With reference to Table 2.3 in Chapter 2, we can construct Table 4.1 with Q = 20 dB and J = 26 dB for P_{AD} with different values of Vt corresponding to the three types of systems, i.e. start-up, medium-size and mature.

The values in Table 4.1 are plotted in Fig. 4.2. It can be seen that P_{AD} increases as the cell radius increases and also as the system grows into maturity.

4.2.2 Inter-cell Adjacent Channel Interference Analysis

In this analysis, we consider the worst case situation in which the mobile station is located at the intersection of three cell sites, namely, point M in Fig. 4.3. The desired base station D and the interfering base stations I_1 and I_2 are assigned channels which are adjacent to the desired base station channel. Since the signal paths from D to M, I_1 to M and I_2 to M and vice versa are about the same, we may apply the same reasoning as in the downlink channel analysis in Section 4.2.1 and conclude that adjacent channel interference would not occur on both the downlink as well as the uplink channel.

Vt (Erlangs per sq. km.)	r (km)	n(T, 0.02)	P_{AD}
0.04	2	3	0.16
	4	5	0.36
	6	8	0.49
	8	12	0.56
	10	17	0.60
0.2	2	6	0.38
	3	10	0.49
	4	14	0.59
	6	26	0.68
0.4	2	8	0.53
	3	15	0.61
	4	24	0.66

TABLE 4.1

PROBABILITY OF INTRA-CELL ADJACENT CHANNEL

INTERFERENCE ON THE UPLINK CHANNEL

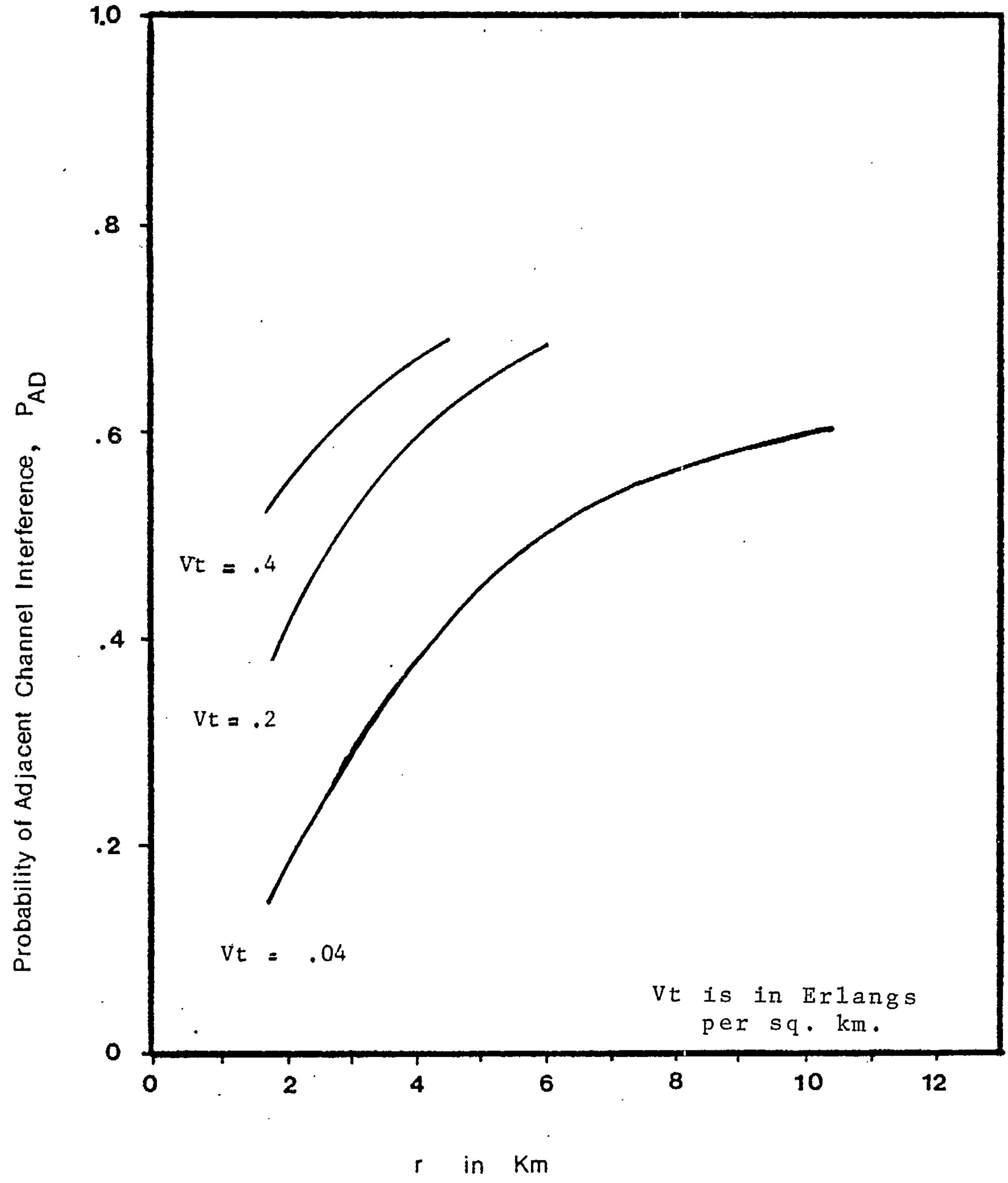


FIGURE 4.2 PROBABILITY OF INTRA-CELL ADJACENT CHANNEL INTERFERENCE ON THE UPLINK CHANNEL vs CELL RADIUS

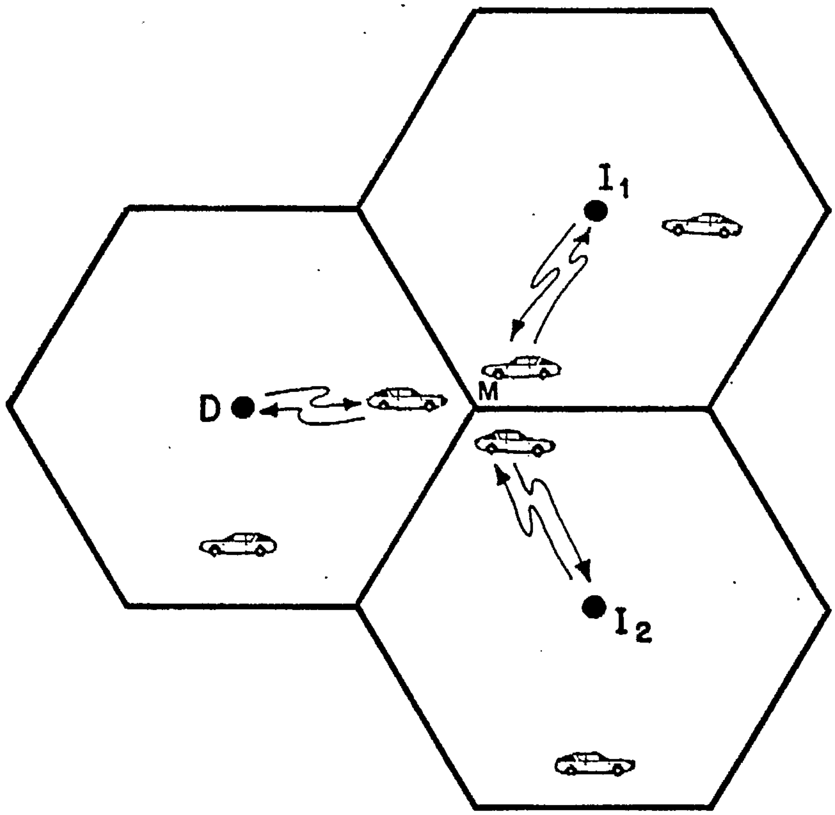


FIGURE 4.3 Inter-cell Adjacent Channel Interference

4.3 Intermodulation Interference Analysis

The second type of interference that is known to occur commonly is intermodulation. Due to the non-linear characteristics of radio equipment, both transmitter and receiver intermodulation occurs. The phenomena have been reported by a number of authors such as McKee [MCKE69], Lustgarten [LUST68], Maiuzzo [MAIU81], McMahon [MCMA74], Chan [CHAN83], etc. Unfortunately, no analysis on the intermodulation interference in a cellular system has been reported.

The basic intermodulation phenomena involve the mixing of two or more signals at different frequencies, in a non-linear circuit, which generate a signal at a different frequency. In general:

$$f_{IM} = a_1 f_1 + a_2 f_2 + a_3 f_3 \;\ldots\; \qquad \ldots\ldots (4.10)$$

where a_1, a_2, a_3, etc. are non-zero integers. f_{IM} is the frequency of the intermodulation product; the other f values represent frequencies of the interfering signals. The intermodulation order of this product is defined as the sum of a_1 , a_2 , a_3 , etc. Although many such products exist, only those for which

$$a_1 + a_2 + a_3 + \ldots\ldots = 1$$

lie in the frequency band of interest. These are called odd-ordered IM products.

In this section and the two following sections, we consider only two-signal third-order transmitter intermodulation (TIM) and receiver intermodulation (RIM) interference. Three-signal third-order intermodulation is less likely to occur since all three signals have to be on simultaneously before it would cause a problem. Higher-order intermodulations are of less significance due to their lower power levels of interference.

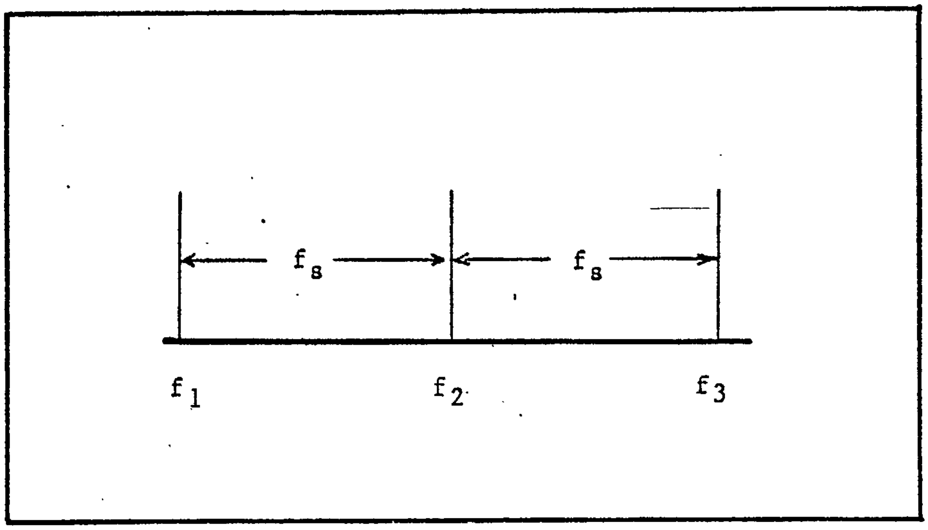


FIGURE 4.4 a)

CONDITION FOR FORMATION OF 2-SIGNAL THIRD-ORDER INTERMODULATION PRODUCTS

A two-signal third-order IM product is formed when:

(i) f_2 and f_3 combine to interfere f_1 so that

$$2f_2 - f_3 = f_1 \quad \text{..... (4.11)}$$

(ii) f_1 and f_2 combine to interfere f_3 so that

$$2f_2 - f_1 = f_3 \quad \text{..... (4.12)}$$

These two conditions are mathematically equivalent and can be expressed as follows:

$$f_s = f_3 - f_2 = f_2 - f_1 \quad \text{..... (4.13)}$$

where f_s is the frequency separation and $f_1 < f_2 < f_3$ as shown in Fig. 4.4a).

We now attempt to answer the following questions before we proceed to the interference analysis:

When a base station is assigned n consecutive channels in a given frequency band and S is the frequency separation in channel spacings between the closest two frequencies in a 2-signal third order IM product,

(a) what is the maximum frequency separation, S_{max}, in an IM product that falls within this given frequency band?

(b) what is the distribution, P(S), of the IM products which have frequency separation S and fall within the same frequency band?

(c) What are the total number, T, and average number, T_{av}, of the IM products that fall in the band?

(a) Calculation of S_{max}:

Referring to Fig. 4.1, since there are n-1 channel spacings between f_0 and f_{n-1}, if n is even,

$$S_{max} = (n-2)/2 \qquad \ldots\ldots (4.14)$$

and if n is odd,

$$S_{max} = (n-1)/2 \qquad \ldots\ldots (4.15)$$

(b) Calculation of P(S):

For all the IM products to fall inside the frequency band from f_0 to f_{n-1} , S must be between S = 1 and S_{max}. For channel i, such that $f_i = f_{IM}$, f_{IM} is either formed by frequencies that lie on both of its sides or by frequencies that lie only on one side depending on the location of i in the frequency band and the value of S.

For a certain S, we can see that the smallest carrier frequency f_i such that $f_i = f_{IM}$ where f_{IM} is formed by frequencies that lie on the lower side of f_i is f_L where

$$L = 2S \qquad ..... (4.16a)$$

since it requires a minimum of 2S channel spacings on its lower side. Similarly, the largest carrier frequency f_i such that $f_i = f_{IM}$ where f_{IM} is formed by frequencies that lie on the upper side of f_i is f_U where

$$U = n - 1 - 2S \qquad ..... (4.16b)$$

since it requires a minimum of 2S channel spacings on its upper side. The situation is depicted in Fig. 4.4b).

It can be seen from Fig. 4.4b) that for carrier frequency f_i where $L \leq i \leq U$, frequencies on both sides of f_i can form IM products to fall on it. However, for a certain $S' \leq S_{max}$, L could be bigger than U. When that happens, as depicted in Fig. 4.4c), only frequencies on one side of f_i can form an IM product to fall on it. We are going to prove in the following that in general the distribution P(S) of the number of IM products as a function of S is (2n - 4S).

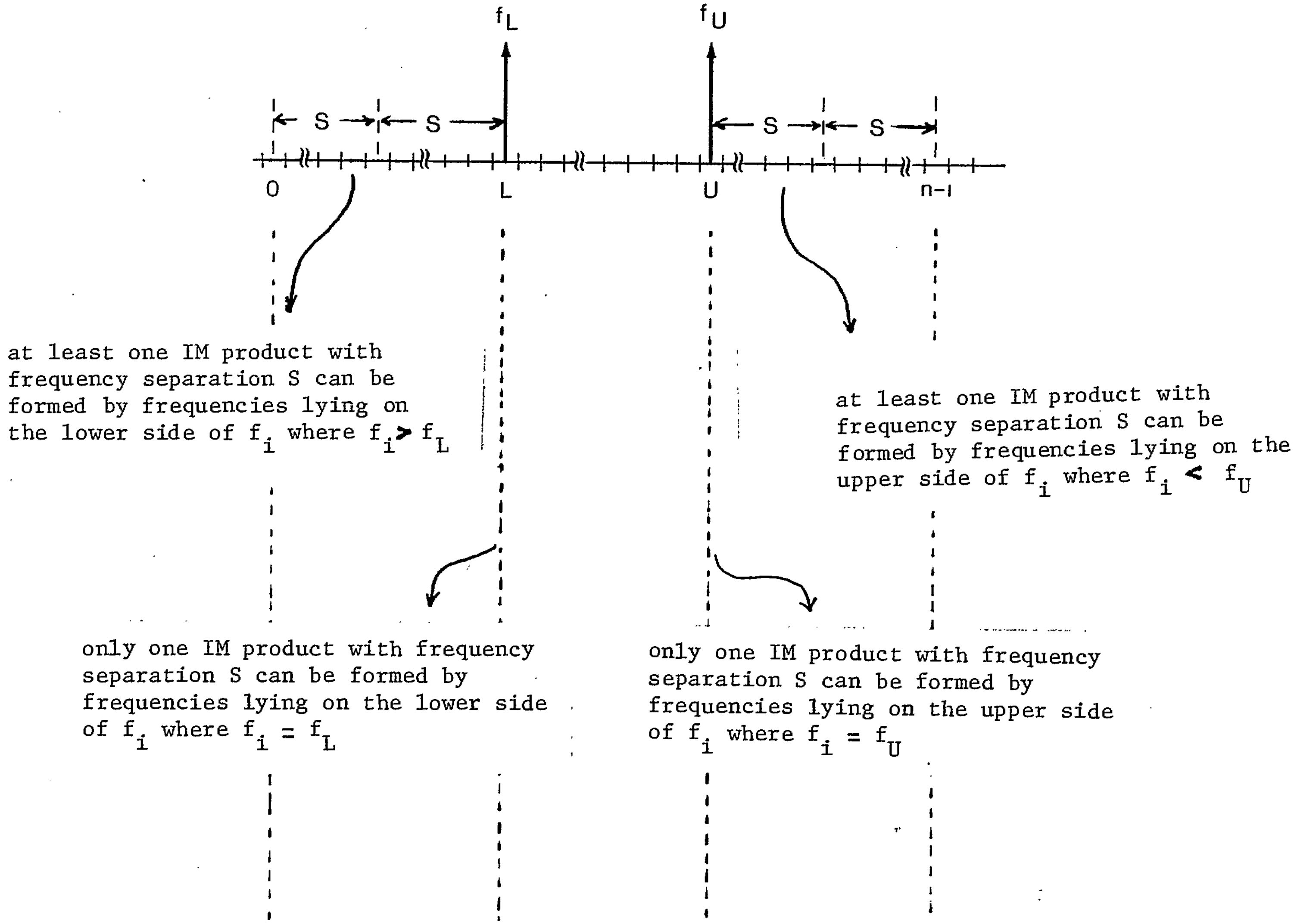


FIGURE 4.4b FORMATION OF IM PRODUCTS WHEN $L \leq U$

Case (i): $L \leq U$,

From Fig. 4.4b), the number of carrier frequencies which have IM products formed by frequencies on both of their sides for a given S is $(U - L + 1)$, or from (4.16a) and (4.16b), $n - 4S$. The remaining $n - (U - L + 1)$ or $4S$ carrier frequencies must have IM products formed by frequencies that lie only on either its upper or lower side. Therefore,

$$P(S) = 2(n - 4S) + 4S$$

or $$P(S) = 2n - 4S \qquad \ldots\ldots (4.16c)$$

The case where $L = U$ occurs when n is odd and $S = S_{max}/2$. An example with $n = 9$ is shown in Fig. 4.4d).

Case (ii): $L > U$

In this case, for a certain S, say S', an IM product can be formed by frequencies that lie only on one side of the carrier frequency. Referring to Fig. 4.4c), these carrier frequencies range from f_L to f_{n-1} and from f_0 to f_U. Therefore,

$$P(S') = (U + 1) + (n-1) - L + 1$$

or $$P(S') = 2n - 4S'$$

(c) Calculation of T and T_{av}:

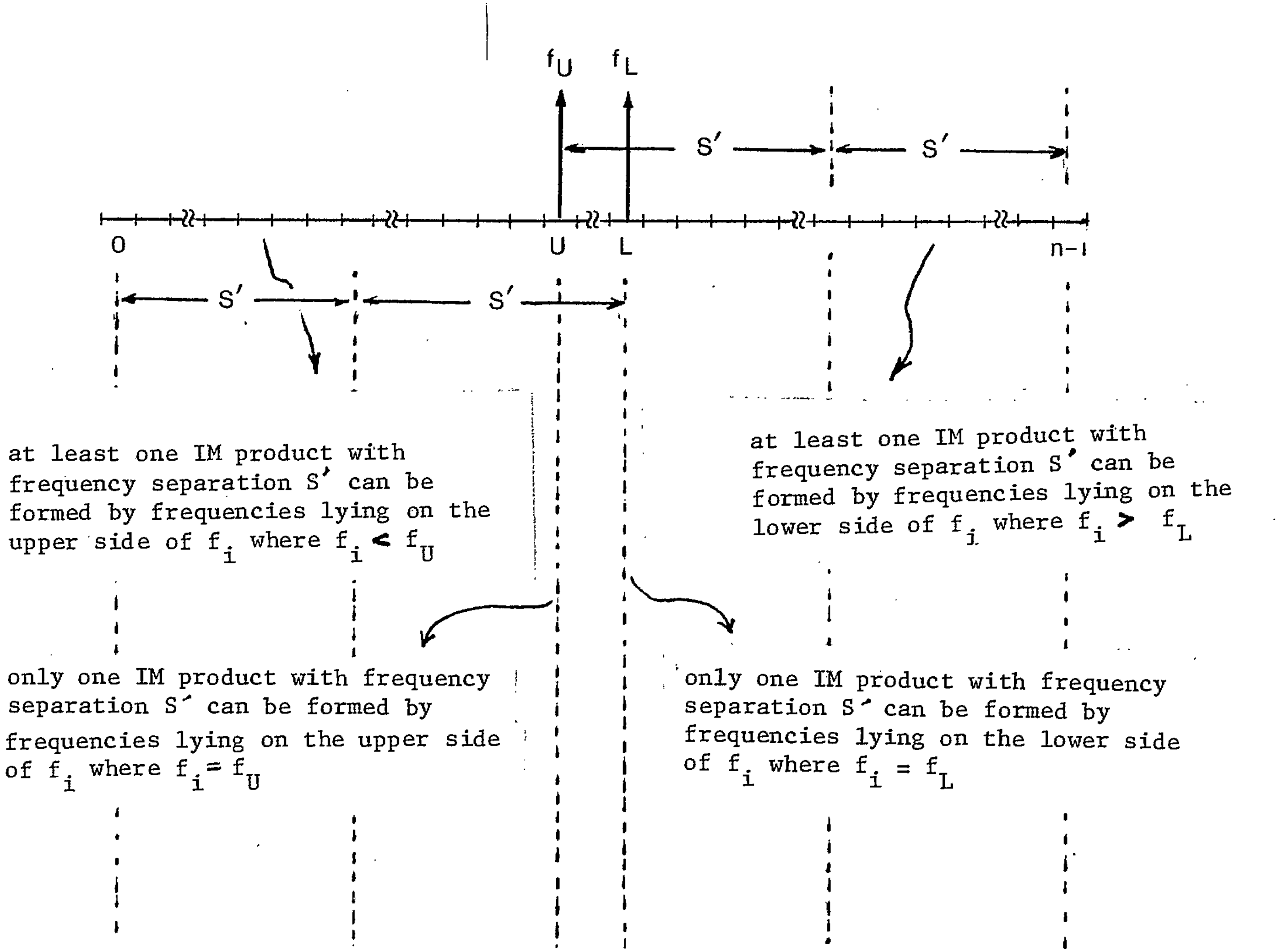
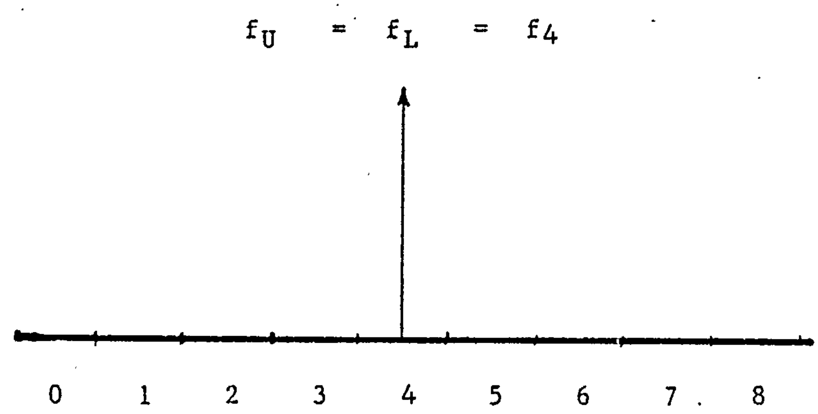


FIGURE 4.4c FORMATION OF IM PRODUCTS WHEN L > U



Note: S = 2, n = 9

FIGURE 4.4d)

FORMATION OF IM PRODUCTS: AN EXAMPLE WITH n = 9,

S_{max} = 4 AND L = U

The total number of IM products that fall within this given frequency band is then:

$$T = \sum_{S=1}^{S_{max}} 2n - 4S \qquad \text{..... (4.17)}$$

For even n, $$T_e = n(n-2)/2 \qquad \text{..... (4.18)}$$

For odd n, $$T_o = (n-1)^2/2 \qquad \text{..... (4.19)}$$

and $$T_{e,av} = (n-2)/2 \qquad \text{..... (4.20)}$$

$$T_{o,av} = (n-1)^2/2n \qquad \text{..... (4.21)}$$

4.4 Transmitter Intermodulation (TIM) Interference Analysis

TIM interference occurs when a signal from one transmitter (called the interfering transmitter) is combined with the output signal of a second transmitter (called the victim transmitter) in its output stage to produce an intermodulation product at a third frequency which is radiated by the second transmitter. We are going to deal with the intra-cell TIM interference in Sections 4.4.1 and 4.4.2 and the inter-cell TIM in Section 4.4.3.

4.4.1 Intra-Cell TIM Interference -- The Downlink Channel

In this analysis, n consecutive channels are assigned to the base station for transmissions to its mobile stations.

Referring to Fig. 4.5, transmitter TX_A emits a signal at frequency f_A which mixes in transmitter TX_B with its signal at frequency f_B. The intermodulation product is a signal at frequency $2f_B - f_A$ equal to frequency, f_C, emitted by transmitter TX_C. Note that one other IM product at $2f_A - f_B$ is also formed, but its magnitude is relatively lower and is usually ignored. The output power P_{ti} of the TIM product at $2f_B - f_A$ is given by:

$$P_{ti} = P_t - B - C + G_t \qquad \text{..... (4.22)}$$

where G_t is the transmitter antenna gain in excess of the circuit losses, C is the conversion loss in dB, defined as the difference in dB between the power levels of the interfering signal from an external source and the intermodulation product, both measured at the output of the transmitter. The value of C is typically in the range of 5 to 20 dB [CCIR739], and is estimated [MCMA74] to be about 11 dB for a frequency separation of 30 to 500 KHz.

The value of B which is the coupling loss in dB between the antennae is usually very low due to the close proximity of the antennae. In its evaluation, we base on the empirical formula given in CCIR Report 524-1 [CCIR524] for calculation of B between

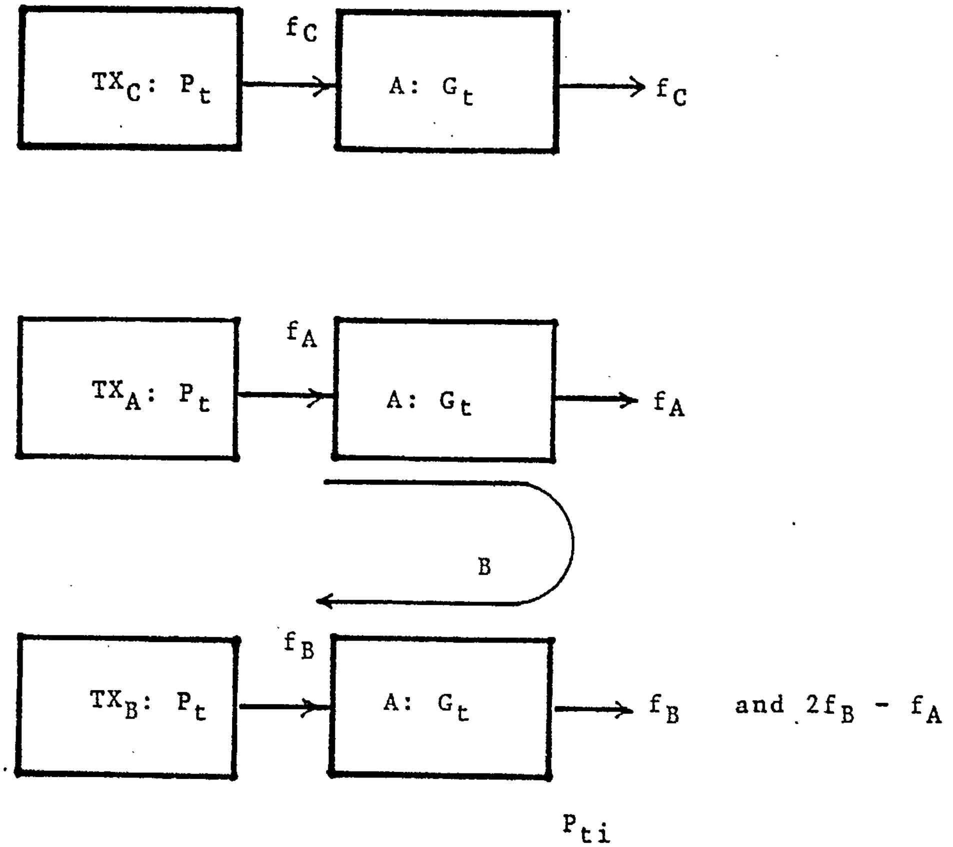


FIGURE 4.5

FORMATION OF TX IM PRODUCT

broadband monopole vertically polarized antennae which are commonly used in the lower UHF band. Hence,

$$B = - G_t - G_r - 27 + 20\log f + 20 \log d +$$

$$\sin^2 \theta (- 40 + 20\log f + 20\log d) \qquad ..... (4.23)$$

where G_t, G_r are normally average transmitter and receiver antenna gains respectively in excess of circuit losses in dB, but in this case, both of them are transmitter antenna gains, f is the frequency in MHz, d is the distance between the antennae in metres and θ is the vertical angle relative to the antenna positions in degrees.

If $G_t = G_r = 12$ dB, f = 460 MHz and d = 0.5 metre, the values of B may be estimated:

B = 7.5 dB, θ = 90 degrees (antennae vertically separated)

B = 3.9 dB, θ = 45 degrees (antennae diagonally separated)

B = 0.2 dB, θ = 0 degree (antennae horizontally separated)

Then from (4.22),

$$(P_t + G_t) - P_{ti} = B + C \qquad ..... (4.24)$$

$$= \begin{cases} 18.5 \text{ dB} \ ; & \theta = 90 \text{ degrees} \\ 14.9\text{dB} \ ; & \theta = 45 \text{ degrees} \\ 11.2 \text{ dB} \ ; & \theta = 0 \text{ degree} \end{cases} \qquad \ldots\ldots (4.25)$$

But intermodulation interference occurs when the interfering signal received power level, Y, is larger than or equal to the desired signal received power level, X, less the protection ratio Q, i.e. when $Y \geq X - Q$. Since the transmission characteristics from the co-located transmitters to the mobile station are the same, interference occurs when

$$B + C \leq Q \qquad \ldots\ldots (4.26)$$

If we consider a Q of 20 dB to be required for good quality reception as in Section 4.2.1, TIM interference must not be ignored.

The probability of TIM interference may be calculated as follows: We first assume that the amount of traffic generated by the base station on the downlink channel is same as that generated by the mobile station on the uplink channel. Then given that the desired transmission is on, the probability that the interfering transmitter is on is:

$$Vt.\ 3\sqrt{3}\ .\ r^2/2(n-1)$$

and the probability that the victim transmitter is on is:

$$Vt.\ 3\sqrt{3}\ .\ r^2/2(n-2)$$

Hence the probability, P_{ti}, that both transmitters are on is:

$$P_{ti} = \left(\frac{Vt.\ 3\sqrt{3}\ .\ r^2}{2}\right)^2 . \frac{1}{(n-1)(n-2)} \qquad \ldots\ldots (4.27)$$

This expression is governed by the assumption that the cell radius and the number of channels in the cell are related to each other such that the grade of service in the cell is 0.02.

Since there is an average of T_{av} IM products in the given frequency band of n channels, when we consider one or more IM products to be on at the same time,

Probability of TIM interference = P_{tim}

$$= 1 - (1 - P_{ti})^{T_{av}} \qquad \ldots\ldots (4.28)$$

It should be noted that this expression is only an approximation since some frequencies are shared between a number of IM products, hence the actual probability of interference is higher than that predicted in (4.28). The second term in this expression corresponds to the probability that no IM product is on in the given frequency band.

From (4.20) and (4.21), the values of P_{tim} for various values of Vt and r may be obtained for both odd and even values of n. The results are shown in Table 4.2 and are also graphically depicted in Fig. 4.6.

4.4.2 Intra-Cell TIM Interference -- The Uplink Channel

In the intra-cell uplink channel situation, mobile stations have to be extremely close to each other and to the base station before TIM interference would cause a problem.

The coupling loss B from one mobile transmitter to another is usually high. If the mobile transmitter antenna gain is 2 dB, d = 3 m. (i.e. the two mobile vehicles are side by side to each other), θ = 0 degrees, f = 460 MHz, B may be found to be equal to 55 dB from (4.23). With C = 11 dB and Q = 23 dB, the TIM signal at the base station receiver has to be 43 dB stronger than the desired signal before interference would occur. This corresponds to a distance ratio of 12 assuming the inverse fourth power law. Therefore, for all practical purposes, we do not consider the probability of TIM interference on the uplink

Vt (Erlangs per sq. km.)	r (km)	n(T, 0.02)	P_{tim}
0.04	2	3	0.06
	3	4	0.15
	4	5	0.34
	5	7	0.48
	6	8	0.70
	8	12	0.92
0.2	2	6	0.39
	3	11	0.71
	4	14	0.97
	6	26	1.00
0.4	2	9	0.73
	2.5	12	0.91
	3	15	0.99

TABLE 4.2

PROBABILITY OF INTRA-CELL TIM INTERFERENCE

ON THE DOWNLINK CHANNEL

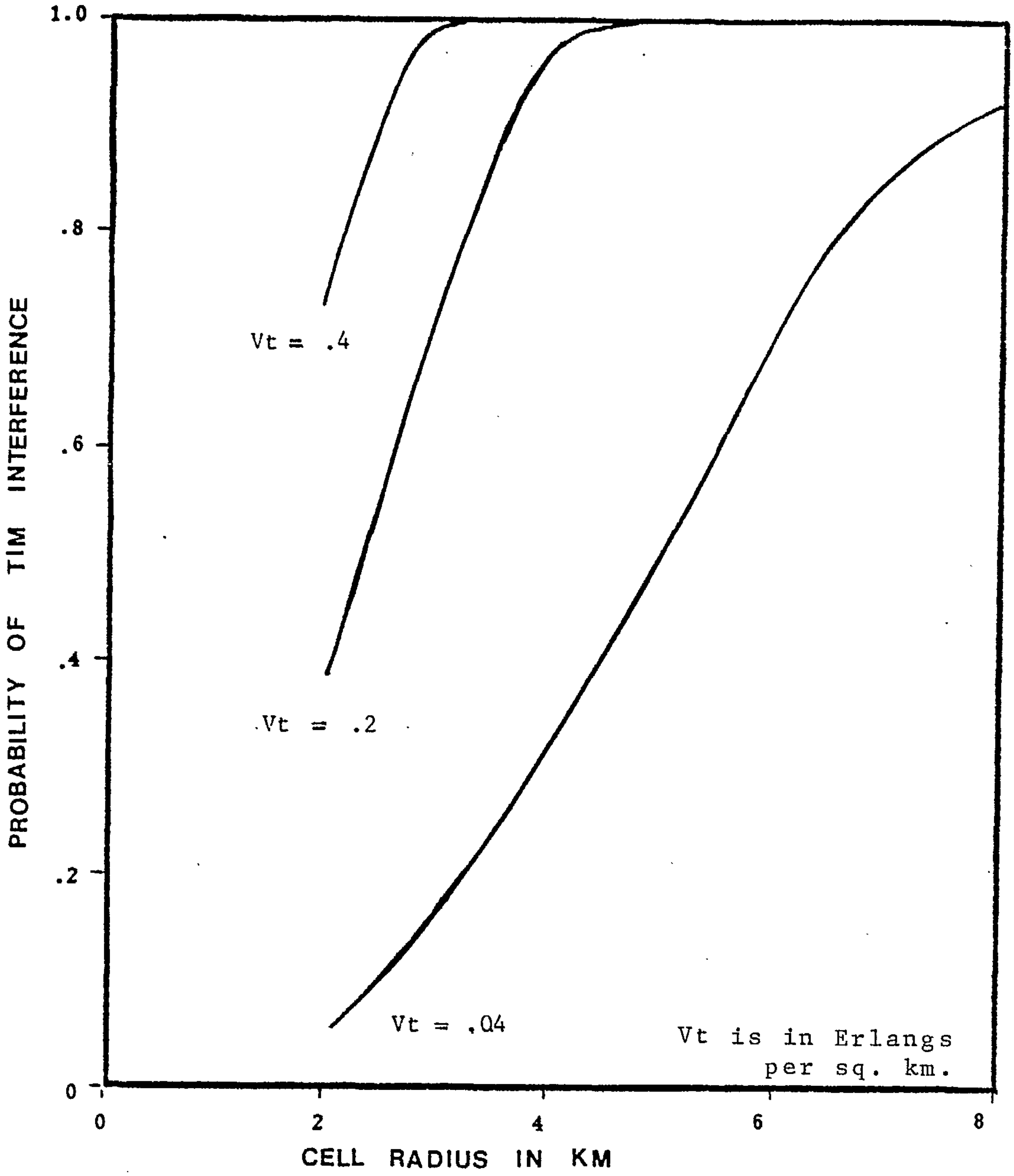


FIGURE 4.6 PROBABILITY OF INTRA-CELL TIM INTERFERENCE ON THE DOWNLINK CHANNEL VS CELL RADIUS

channel to be significant.

4.4.3 Inter-Cell TIM Interference

In this section, we consider the inter-cell uplink and downlink situations together. Since the assumptions are similar to those in the adjacent channel interference analysis described in Section 4.2.2, we refer to Fig. 4.3 in our discussions here.

On the downlink channel, the interfering and victim transmitters are separated by a distance almost equal to twice the cell radius, the coupling loss is high enough to dismiss any possibility of TIM interference.

On the uplink channel, similar to the intra-cell uplink channel situation, the interfering mobiles from different cells have to stay close together before any TIM interference would occur. An example of the possible location is at M as shown in Fig. 4.3. Since they would then be at the maximum distance from the desired base station, no TIM interference would be encountered here as well.

4.5 Receiver Intermodulation (RIM) Interference Analysis

RIM interference occurs when two signals are combined in the non-linear section of the RF stages of a receiver to form a third signal which falls within the frequency band accepted by the receiver. Similar to the TIM interference analysis, we deal with the intra-cell and inter-cell situations separately in the following sections.

4.5.1 Intra-Cell RIM Interference -- The Uplink Channel

In the intra-cell uplink channel situation, the desired mobile station is assumed to be located on the edge of the cell. The two interfering mobile stations are referred to as the near transmitter (with frequency f_N) and the far transmitter (with frequency f_F) since f_N is closer to the victim receiver frequency f_V than f_F in the IM product: $2f_N - f_F = f_V$.

To obtain the interfering received signal power level P_{ri}, we refer to the CCIR Report 522 [CCIR522]:

$$P_{ri} = 2P_N + P_F - K \qquad \text{..... (4.29)}$$

where K is the RIM conversion loss factor in dB, P_N and P_F are the received power levels of signals from transmitters N and F respectively. The value of K varies with different receivers. McMahon in [MCMA74] reported that the empirical value of K for equipment in the VHF and lower UHF bands is given by:

$$K = 60 \log df \qquad \text{..... (4.30)}$$

where df is the frequency separation between the near and far transmitter frequencies.

The received signal power levels may be obtained by subtracting the path loss from the total of the mobile transmitter power, P_{mt}, the mobile transmitter antenna gain, G_{mt}, in excess of circuit losses, and the base receiver antenna gain, G_{br}, in excess of circuit losses, hence,

$$P_N = P_{mt} + G_{mt} - P_{1N} + G_{br} \qquad \ldots\ldots (4.31)$$

$$P_F = P_{mt} + G_{mt} - P_{1N} + G_{br} \qquad \ldots\ldots (4.32)$$

$$P_d = P_{mt} + G_{mt} - P_{1d} + G_{br} \qquad \ldots\ldots (4.33)$$

where P_{1N}, P_{1F} and P_{1d} are path losses from the near transmitter, the far transmitter and the desired transmitter respectively to the base station receiver and P_d is the desired signal received power level. Interference occurs if:

$$P_d - P_{ri} \leq Q \quad \text{, or,}$$

$$-2P_{mt} - 2G_{mt} - 2\,G_{br} - P_{1d} + 2P_{1N} + P_{1F} + K \leq Q \qquad \ldots\ldots (4.34)$$

From (2.10),

$$2\,P_{1N} + P_{1F} - P_{1d} = 40\log f + 2(83.64) - 40\log H_t.H_r$$

$$+ 40\log d_F\,.d_N^{\,2} - 40\log r \quad \ldots\ldots (4.35)$$

From (4.34), (4.35), (4.30) and (2.11), and assuming σ to be the same for the median power levels of all the three signals, interference occurs when,

$$40\log d_F\,.d_N^{\,2} \leq Q - 60\log df + 2\,X_{td} - 2\,P_{fd} + 2(0.117)\sigma^2$$

$$+120\log r$$

where X_{td} is the threshold power level at the base station and P_{fd} is the failure rate in dB as defined in Section 2.3. Hence,

$$d_F\,.d_N^{\,2} \leq 10^{\left(\frac{Q - 60\log df + 2X_{td} - 2P_{fd} + 120\log r + 0.234\sigma^2}{40}\right)} \quad \ldots\ldots (4.36)$$

$$\text{or} \quad d_F\,.d_N^{\,2} \leq K_{im} \quad \ldots\ldots (4.37)$$

where K_{im} is the RIM interference criterion such that:

$$K_{im} = 10^{\left(\frac{Q - 60\log df + 2X_{td} - 2P_{fd} + 120\log r + 0.234\sigma^2}{40}\right)} \quad \ldots\ldots (4.38)$$

This means that RIM interference would occur if the mobile with frequency f_N transmits at distance d_N from the base station and a second mobile with frequency f_F transmits within distance d_F from the base station.

Assuming that the desired mobile is transmitting on the edge of the cell, the probability that a second mobile transmits with frequency f_F at distance d_F from the base station within a thin ring of width $d(d_F)$ as shown in Fig. 4.7a) is:

$$2 \quad d_F \,.\, Vt \,.\, d(d_F) \,/\, (n-1)$$

The probability that a third mobile transmits with frequency f_N within distance d_N from the base station is:

$$d_N^{\,2} \,.\, Vt \,/\, (n-2) \quad \text{or} \quad Vt \,.\, K_{im} \,/\, d_F.(n-2)$$

for RIM interference to occur.

As in the adjacent channel interference and TIM interference analyses, these two probabilities are based on the assumption that the grade of service in the cell is maintained at 0.02.

Fig. 4.7b) shows the relationship between d_N and d_F. Note that both d_F and d_N would not exceed the cell radius r and that $d_F = K_{im} / r^2$ when $d_N = r$.

The probability of RIM interference may be calculated as

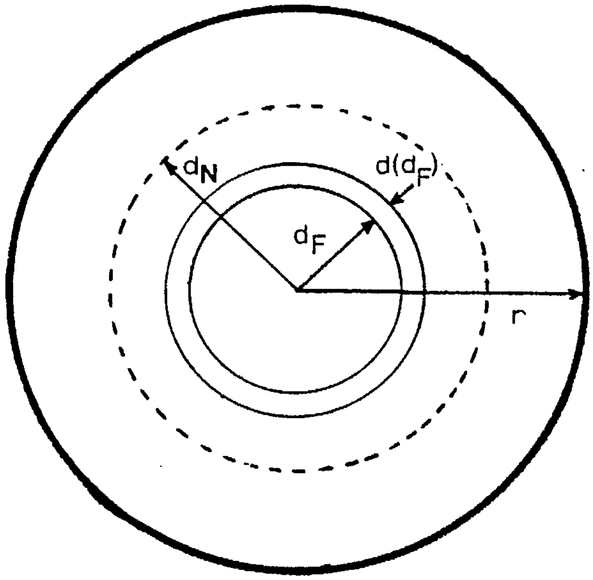


FIGURE 4.7 a) DIAGRAM OF d_N, d_F, $d(d_F)$ AND r

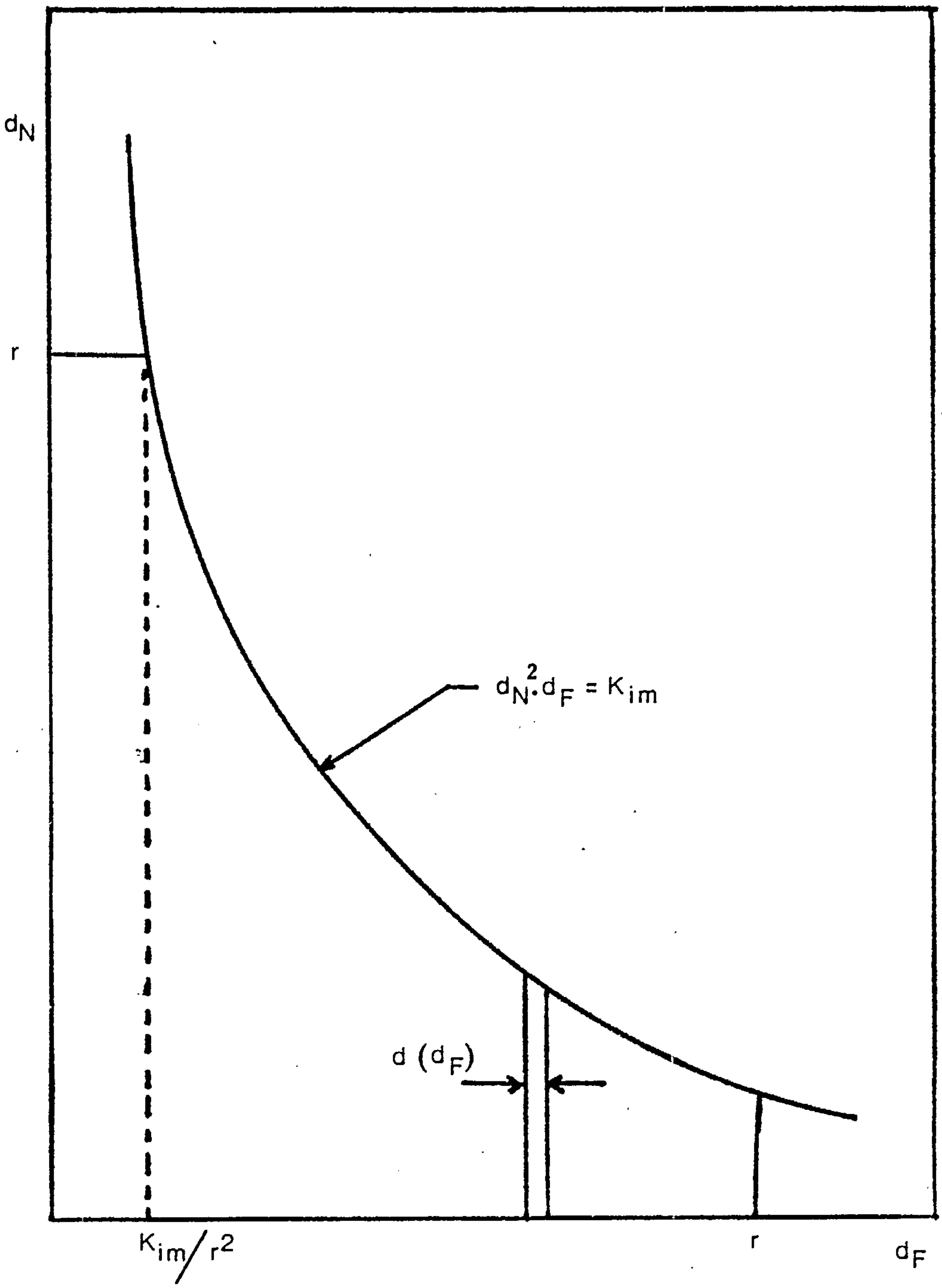


FIGURE 4.7b) RELATIONSHIP BETWEEN d_N AND d_F

follows:

For $K_{im}/r^2 \leq d_F \leq r$, the probability of interference at a given d_F is:

$$2\ (\pi Vt)^2 \cdot K_{im} \cdot d(d_F)\ /\ (n-1)(n-2)$$

For all values of d_F in this range , the probability of interference, P_{r1}, is:

$$P_{r1} = \int_{K_{im}/r^2}^{r} \frac{2\ (\pi Vt)^2 \cdot K_{im}}{(n-1)(n-2)} \cdot d(d_F)$$

$$= \frac{2\ (\pi Vt)^2 \cdot K_{im}}{(n-1)(n-2)} \left[r - \frac{K_{im}}{r^2} \right]$$

For $0 \leq d_F \leq K_{im}/r^2$, the probability of interference at a given d_F is:

$$2\ \pi\ d_F \cdot Vt \cdot d(d_F)/(n-1) \quad x \quad \pi Vt \cdot r^2\ /(n-2)$$

For all values of d_N in this range, the probability of interference, P_{r2}, is:

$$P_{r2} = \int_{0}^{K_{im}/r^2} \frac{2\ (\pi Vt)^2 \cdot r^2}{(n-1)(n-2)} \cdot d_F \cdot d(d_F)$$

$$= \frac{(\pi Vt)^2 \cdot K_{im}{}^2}{(n-1)(n-2) \cdot r^2}$$

Hence the total probability of interference,

$$P_{ri}(S) = P_{r1} + P_{r2}$$

$$= \frac{(\pi Vt)^2}{(n-1)(n-2)} \cdot \left[2 K_{im} \cdot r - \frac{K_{im}{}^2}{r^2} \right] \qquad \ldots\ldots (4.39)$$

Note that $P_{im}(S)$ is the probability of RIM interference due to one IM product with frequency separation S in channel spacings or df ($= S \times f_s$) in MHz and K_{im} is a function of df or S, among other parameters.

For n consecutive channels assigned to the cell, there is more than one way that an IM product is formed. For a given S, there are on the average P(S)/n IM products falling on the desired channel. The probability of RIM interference, $P_{im}(S)$, for a given S is therefore,

$$P_{im}(S) = 1 - [\, 1 - P_{ri}(S) \,]^{P(S)/n} \qquad \ldots\ldots (4.40a)$$

This probability takes into account the possibility that more than one RIM product may be active at one time. However, as in the TIM interference analysis, this expression is only an approximation since some frequencies are shared between a number of IM products and hence the actual probability of interference is higher than that predicted in (4.40a).

If $P_{im}(S)$ is small,

$$P_{im}(S) = P(S) \cdot P_{ri}(S)/n \qquad \text{..... (4.40b)}$$

Therefore, the total probability of RIM interference, P_{rim}, is,

$$P_{rim} = \sum_{S=1}^{S_{max}} P_{im}(S) \qquad \text{..... (4.40c)}$$

From (4.16c) and (4.40b),

$$P_{rim} = \sum_{S=1}^{S_{max}} P_{ri}(S) \cdot \frac{(2n - 4S)}{n} \qquad \text{..... (4.40d)}$$

From (4.14), (4.15),

$$P_{rim} = \sum_{S=1}^{(n-1)/2} P_{ri}(S) \cdot \frac{(2n - 4S)}{n} \qquad \text{..... (4.41a)}$$

for odd n,

$$P_{rim} = \sum_{S=1}^{(n-2)/2} P_{ri}(S) \cdot \frac{(2n - 4S)}{n} \qquad \text{..... (4.41b)}$$

for even n.

From (4.41), the values of P_{rim} can be obtained and tabulated in Table 4.3. We assume in the calculation that f_s = 0.03 MHz, X_{td} = - 132 dBW, Q = 20 dB and P_f = 0.1 and 0.02 separately. It can be seen that the probability of RIM interference increases when (i) the cell radius increases, (ii) the value of σ increases or (iii) the system size increases. But in general, RIM interference is insignificant in almost all of the situations.

4.5.2 Intra-Cell RIM Interference -- The Downlink Channel

In this situation, different signals from the base station are received at a mobile station in which an IM product may be formed. However, since the interfering signals travel the same distance as the desired signal from the base station to the mobile receiver,

$$P_N = P_F = P_d \qquad \ldots\ldots (4.42)$$

Therefore, from (4.34) and (2.9),

$$\begin{aligned} P_d - P_{ri} &= -2P_{mt} - 2G_{mt} - 2G_{br} + 2P_{1d} + K \\ &= K - 2(X_{td} - P_{fd} + 0.117\ \sigma^2) \\ &\gg Q \end{aligned}$$

for all practical values of K, X_{td}, P_{fd} and σ . So no RIM interference would occur.

P_f	Vt (Erl. per sq. km.)	r (km)	n(T, 0.02)	P_{rim} $\sigma = 6$	P_{rim} $\sigma = 12$
0.1	0.04	3	4	0.0004	0.0017
		6	8	0.0018	0.0076
		10	17	0.0035	0.0148
		14	28	0.0053	0.0227
	0.2	3	10	0.0018	0.0079
		6	26	0.0051	0.0220
		10	61	0.0080	0.0344
	0.4	3	15	0.0035	0.0151
		6	46	0.0071	0.0303
		10	113	0.0100	0.0427
0.02	0.04	3	4	0.0009	0.0038
		6	8	0.0040	0.0169
		10	17	0.0077	0.0331
		14	28	0.0119	0.0507
	0.2	3	10	0.0041	0.0176
		6	26	0.0115	0.0490
		10	61	0.0180	0.0769
	0.4	3	15	0.0079	0.0337
		6	46	0.0159	0.0678
		10	113	0.0222	0.0953

TABLE 4.3

PROBABILITY OF INTRA-CELL RIM INTERFERENCE ON THE UPLINK CHANNEL

4.5.3 Inter-Cell RIM Interference

For the inter-cell RIM interference, we again refer to Fig. 4.3. On the uplink channel, the worst situation would be for the interfering mobiles from I_1 and I_2 to be located at M, the intersection of the three cells. However, since the interfering mobiles at best are at approximately the same distance from base station D as the desired mobile station, no RIM interference would occur, based on the same reasoning as in Section 4.5.2.

On the downlink channel, the interfering signals from base stations I_1 and I_2 would not cause any problems to the mobile from D as it can be easily seen that the distance travelled by the interfering and desired signals are approximately the same.

4.6 Summary of Results

In this chapter, we have analysed the adjacent channel, transmitter and receiver intermodulation interference situations in a cellular system. We have also considered the intra-cell and inter-cell cases in which the downlink and uplink channel operations are discussed separately. The results of the analysis are summarized in Table 4.4 for easy comparison and reference.

As it can be seen from the Table, inter-cell interference situations may be ignored but not the intra-cell interference

INTERFERENCE TYPE	INTRA-CELL		INTER-CELL	
	UPLINK	DOWNLINK	UPLINK	DOWNLINK
Adjacent Channel	Y	N	N	N
TIM	N	Y	N	N
RIM *	Y	N	N	N

Notes:

Y indicates presence of interference

N indicates insignificant amount of interference

* RIM interference relatively mild

TABLE 4.4

SUMMARY CHART OF DIFFERENT TYPES OF INTERFERENCE IN A CELLULAR SYSTEM situation. Other observations are listed in the following:

- Interference increases when cell radius increases. This is due to the fact that a bigger cell radius increases the probability that a desired mobile station be located farther away from the base station than the interfering mobile station, thus increasing the chance for the interfering signal to exceed the desired signal.

- Interference increases when the system size increases. This is due to the fact that the amount of traffic Vt increases faster than the number of channels required, n, causing the ratio Vt/n to increase as the system expands.

- In the case of RIM interference, a larger value of σ gives a larger probability of interference. This may be explained by the fact that a larger transmitter output power is required for a larger σ to compensate for the shadowing loss. This in turn increases the probability of RIM interference.

There are a number of points that should be noted. The first one is that the above interference analysis is performed for voice channels only. Interference between digital control channel and voice channel is assumed to have similar behaviour as that between voice channels. This is based on the assumption that the out-of-band emission of the digital signal is confined in the same fashion as the voice signal.

The second one is that the RIM and TIM interference analysis is performed for the low UHF band due to the lack of empirical data for equipment in the upper UHF band. The interference characteristics is expected to show similar behaviour at the 800 MHz band.

The third point that is worth-mentioning is that in some parts of the analysis, we have considered the worst case situation while in other parts, the actual situation may be worse than that assumed in the analysis. The worst case situation includes: the location of the desired mobile station to be on the edge of the cell when normally the mobile station operates more closely to the base station; the allocation of n consecutive channels to the base station which normally does not take place; etc.

On the other hand, our analysis has considered only long term median power levels and the effects of fading and shadowing on the desired and interfering signals are not differentiated. In other words, the desired and interfering signals are assumed to be fully correlated and suffer the fading and shadowing losses in exactly the same fashion. This is not usually true for the transmissions on the uplink channels since the mobile stations transmit signals from different locations in the cell. But for downlink transmissions from the base to a mobile station, this assumption is correct. The effect of this on our analysis is that the uplink interference situation may be slightly worse than what the results have indicated as the desired and interfering signals may not be fully correlated.

In this chapter, we have identified the interference problems for the intra-cell situation. It is necessary to devise methods to avoid such interference. As pointed out earlier, consecutive channels are normally not allocated to the same cell. This is partly because of the difficulties and limitations in multicoupling the transmitters and also because of the fact that interference can normally be reduced using larger frequency separations. In practice, filtering, isolation techniques and variable mobile power features have been used to minimize interference problems. These solutions however add to the cost and complexity of the system. In the next chapter, we propose a frequency allocation scheme which can solve this specific problem by providing interference-free operation to the system.

CHAPTER 5 - DESIGN OF A SPECTRUM-EFFICIENT FREQUENCY ALLOCATION SCHEME

5.1 Introduction

The results of the interference analysis obtained in Chapter 4 indicated that intra-cell adjacent channel and two-signal third-order intermodulation interference in a cellular system need to be minimized while the inter-cell interference may be ignored. In this chapter, we describe a novel spectrum-efficient frequency allocation scheme that provides an operation free from the above-mentioned types of interference. We assume that the traffic is uniform for all the cells of the cellular system and there are m channels in each cell.

In solving the problem of designing this spectrum-efficient frequency allocation scheme, the following requirements must be satisfied:

(i) maximum utilization of the given spectrum must be achieved;
(ii) all channels must be free from co-channel interference;
(iii) all channels must be free from two-signal third-order intermodulation interference;
(iv) all channels must be free from adjacent channel interference;
(v) any two frequencies in a given cell must be separated from

each other by a frequency separation of at least D channels.

Requirement (i) is necessary to ensure that the scheme is spectrum-efficient. Requirements (ii), (iii) and (iv) guarantee the cellular system of an interference-free operation. Requirement (v) arises because of limitations of multicoupling equipment in which a certain minimum separation between operating frequencies sharing the same equipment needs to be maintained. In simpler terms, given m, D and N (the number of cells per cluster), the problem is to find a scheme to allocate frequencies to the cells of a cellular system so that all the stated requirements are satisfied and the maximum utilization of the spectrum is achieved.

In Section 5.2, we provide a brief description of the work of other researchers in this area. We then proceed to describe in Section 5.3 one possible optimal scheme that satisfies all the above requirements. Section 5.4 provides the results of this scheme and discussions of its limitations. In Section 5.5, we describe a sub-optimal scheme which provides solutions to cases not solved by the optimal scheme. We then compare this scheme with a simple channel rejection method in Section 5.6. Finally, in Section 5.7, we determine the computational complexity of the optimal algorithm.

5.2 Previous Channel Allocation Methods

A brief survey of the existing literature on the channel allocation methods reveals that there are four major approaches to solving this problem. Unfortunately, most of these methods do not provide a complete solution to our problem even though some succeed to tackle part of the whole problem.

Since some researchers have used the word "allocation" while others used "assignment" in the literature, these two words are used interchangeably in this chapter.

5.2.1 The Dynamic Frequency Assignment Schemes

Over the past decade, significant volume of research work has been devoted to the development of dynamic and hybrid frequency assignment schemes for cellular systems with the main objective of reducing the blocking probability of mobile telephone calls when the average traffic load increases in a cell. These include the work by Schiff [SCHI70], Cox and Reudink [COX73], Kahwa and Georganas [KAHW78], Sin and Georganas [SIN81], Nehme and Georganas [NEHM81], Singh [SING81], Elnoubi et al [ELNO82], etc.

However, these workers have not considered the effect of adjacent channel and intermodulation interference on the assignment schemes even though co-channel interference was not ignored in some cases. Therefore, these schemes do not address the problem formulated in the beginning of this chapter.

5.2.2 The IM-Free Frequency Lists

In this category of frequency allocation schemes, researchers are interested in obtaining intermodulation-free frequency lists. One of them is Babcock [BABC52] who obtained a frequency list containing up to ten third-order IM-free frequencies. The principle behind the method is based on the fact that in order to avoid this type of intermodulation, the frequency difference between a given pair of channels must be different from that between any other pair of channels. Edwards et al [EDWA69], Fang and Sandrin [FANG77] and Lustgarten [LUST68] also developed IM-free frequency lists based on a similar principle. One drawback that is common to these schemes is that a very large number of frequencies have to be rejected resulting in a very poor utilization of the spectrum. In a recent paper [GARD82] by Gardiner and Magaza, five techniques were proposed to mix channels with different bandwidths to avoid intermodulation. This has resulted in improvements in spectrum utilization.

5.2.3 The Graphical Method

The Graphical Method has been reported by Zoellner and Beall [ZOEL77] to be capable of providing solutions to the frequency assignment problem. They drew analogy between the frequency assignment problem and the classical node-colouring problem encountered in graph theory. In colouring the graph, inter-connected nodes may not receive the same colour. The objective is to find the minimum number of colours for all the nodes. In frequency assignment, any two stations which may not be assigned the same channel are represented by two nodes connected by a link.

In these schemes, the interference constraints and the stations are represented by a graph with the appropriate nodes and links. A graph decomposition procedure is then used to remove nodes with the largest number of links connected to them. This is used to rank the stations in order of decreasing assignment difficulty so that frequencies may be assigned first to the more difficult stations.

The Graphical Method is mainly used for frequency assignment with the co-channel interference constraint and in some cases, the adjacent channel interference constraint. It was pointed out in [ZOEL77] that the inclusion of complex cosite constraints such as intermodulation interference would result in an enormous increase in data requirements, that in turn, require more computer processing and storage. Up to the present, we have not encountered any description of work that uses the Graphical Method to eliminate IM interference in frequency assignment problems.

5.2.4 A Heuristic Technique

A heuristic technique capable of solving frequency assignment problems involving co-channel, adjacent channel and intermodulation interference, wide frequency separation requirements and previous frequency assignments is described by Box in [BOX78]. This technique comes closest to solving our frequency allocation problem stated in Section 5.1.

The basis of this procedure is in setting up an assignment sequence in descending order of "assignment difficulty", quite similar to the approach taken in the Graphical Method. A channel separation matrix is first set up in which the interference (excluding intermodulation) constraints are incorporated. Frequencies are assigned to the stations in a number of attempts. Stations that cannot be assigned a frequency in an attempt are called denials. After each attempt, the assignment order is rearranged so that the denials receive their assignments first in the next attempt. The process stops when no denial is encountered. Intermodulation check is performed only when the interference constraints are satisfied.

In the next section, we describe a frequency allocation scheme specially designed for a cellular radio system or a system which has multiple sites. It is different from the above schemes in terms of the general approach taken. While we do not use an ordering procedure like the Graphical Method or the Heuristic technique, we adopt the same principle in performing intermodulation interference checks as some of the schemes described since this is basically the fundamental check for such phenomenon.

5.3 A New Frequency Allocation Scheme

The objective of the frequency allocation scheme is to allocate a total of mN frequencies to the N cells of the cluster in the cellular system so that each cell has m frequencies with frequency separation of at least D channel spacings between them. The same pattern can then be repeated for other clusters in the system. Referring to the requirements stated in Section 5.1, we describe the derivation of the scheme in the following:

5.3.1 The Initial Allocation Plan

The first step taken in deriving the frequency allocation scheme is to establish an initial allocation plan. This plan has to satisfy requirements (i) and (ii) which state that maximum utilization of the given spectrum is to be attained and all the frequencies are free from co-channel interference.

Since the initial plan is used as a basis on which the ultimate frequency allocation scheme is developed, the choice of such a plan is important. While a number of possibilities exist, there are very few that are simple and systematic. The one we have chosen is the horizontal frequency arrangement plan as it is one of the simplest and the most suitable for our frequency allocation scheme. The plan is depicted in Fig. 5.1 in which a matrix is shown with the columns representing the cell sites and the rows representing the frequencies of the site. This initial allocation plan satisfies the first two requirements since exactly mN consecutive channels are used so that the given spectrum is fully utilized and no frequencies are used more than once so that co-channel interference would not occur.

It should be noted that in the following discussion, channel numbers and frequencies may be used interchangeably. The relationship between the two is:

$$f_i = f_o + i.f_s \qquad \text{..... (5.1)}$$

where f_s is the channel spacing, f_o is the lowest frequency in the given band of spectrum and f_i is the frequency of channel i.

In the initial allocation plan, channel i in cell J is represented by C_{iJ} where

$$C_{iJ} = N.i + J \qquad \text{..... (5.2)}$$

with $0 \leq i \leq m - 1$ and $0 \leq J \leq N - 1$.

cell no / channel number	0	1	2	3	....	N-1
0	0	1	2	3	....	N-1
1	N	N+1	N+2	N+3	....	2N-1
2	2N	2N+1	2N+2	2N+3	....	3N-1
3	3N	3N+1	3N+2	3N+3	....	4N-1
.	.	.	.	.	.	.
.	.	.	.	.	.	.
.	.	.	.	.	.	.
m-1	(m-1)N	(m-1)N+1	.	.	....	mN-1

FIGURE 5.1

INITIAL FREQUENCY ALLOCATION PLAN

Despite the fact that this plan satisfies the first two requirements, it does not fulfill the intermodulation interference-free requirement. Based on the condition stated in (4.13) in Chapter 4, two-signal third-order IM products formed from some frequencies in the cell will fall right on other allocated frequencies in the same cell. In order to avoid the formation of such products, any frequency in any group of three frequencies allocated to the cell must not be equidistant from the other two frequencies and this may be accomplished by cyclically shifting the channels on a particular row of channels in the initial plan from one cell site to another as described in the next section.

5.3.2 The Cyclic Shifting Technique

The cyclic shifting of channels on a particular row of channels in the initial allocation plan may be visualized as moving the channels from one cell to another cyclically so that for a shift of one,

C_{j1} assumes the value of C_{j0}

C_{j2} assumes the value of C_{j1}

.

.

.

.

$C_{j(N-1)}$ assumes the value of $C_{j(N-2)}$

C_{j0} assumes the value of $C_{j(N-1)}$

An example is shown in Fig. 5.2 to illustrate the situation where a shift of 1 is applied to row k of the initial plan. It is quite clear from the cyclic shifting property that the shift S of any row must lie between 0 and N - 1 inclusively to produce a unique channel arrangement. Therefore,

$$0 \leq S \leq N - 1 \qquad \ldots\ldots (5.3)$$

This shifting of channels solves the intermodulation problem. However, it causes the frequency separation between channels to vary. With no shifting, the separation between any two consecutive rows is constant at N. But shifting reduces this separation in some cells and increases it in others. Indeed, the maximum of the minimum separation between any two consecutive rows is N and is only attainable when the shift is zero. The fourth and fifth requirements in Section 5.1 imply that the separation in channel spacings must not be less than 2 in order to avoid adjacent channel interference and has to be greater than or equal to a given number D in order to provide the necessary frequency separation between channels. It should be quite obvious that D cannot exceed N if all the given channels are to be utilized otherwise the allocation plan would require more than mN consecutive channels.

If we let $W_j(h)$ be the relative possible shift of row j with respect to row i to provide the required separation D, then

	cell no.	0	1	2	....	N-1
before shift :	row (k-1) row k	(k-1)N kN	(k-1)N+1 kN+1	(k-1)N+2 kN+2	....	kN-1 (k+1)N-1
after shift:	row (k-1) row k	(k-1)N (k+1)N-1	(k-1)N+1 kN	(k-1)N+2 kN+1	....	kN-1 (k+1)N-2

Note: shift = 1 for row k

FIGURE 5.2

ILLUSTRATION OF THE CYCLIC SHIFTING TECHNIQUE

$$0 \leq W_j(h) \leq N - D \quad \text{for } h = 1 \qquad \text{..... (5.4)}$$

and $$0 \leq W_j(h) \leq N - 1 \quad \text{for } h \geq 2 \qquad \text{..... (5.5)}$$

where $h = j - i$ and $j > i$.

The proofs of (5.4) and (5.5) are provided in the following:

Proof of (5.4): Without loss of generality, we assume that row i has a shift of zero, i.e. $S_i = 0$. In considering the frequency separation between channels i and j, we only need to consider the smallest channel N.j in row j since if the smallest channel satisfies the minimum frequency separation requirement, any larger channel must satisfy the same requirement. If row j has a shift of $W_j(h)$, channel N.j would be moved to cell $W_j(h)$ from cell 0. From (5.2), the minimum channel separation, D_{ji}, between rows i and j is therefore,

$$D_{ji} = N.j - [\ N.i + W_j(h)\ ]$$

or $$W_j(h) = N.h - D_{ji} \qquad \text{..... (5.6)}$$

But $D_{ji} \geq D$, therefore,

$$W_j(h) \leq N.h - D \qquad \text{..... (5.7)}$$

For $h = j - i = 1$,

$$W_j(1) \leq N - D \qquad \ldots\ldots (5.8)$$

Hence, from (5.3) and (5.8),

for $h = 1$, $\quad 0 \leq W_j(1) \leq N - D \quad$ (Q.E.D.)

$W_j(1)$ may assume any integer value between 0 and N-D inclusively and the number of relative possible shifts is (N-D+1).

Proof of (5.5): Without loss of generality, we again assume $S_i = 0$. From (5.7),

$$W_j(h) \leq N.h - D$$

But $h = j - i \geq 2$ and $D \leq N$,

$$W_j(h) \leq N \qquad \ldots\ldots (5.9)$$

From (5.3) and (5.9),

for $h \geq 2$, $\quad 0 \leq W_j(h) \leq N - 1 \quad$ (Q.E.D.)

$W_j(h)$ may assume any integer value between 0 and N-1 inclusively and the number of relative possible shifts is N.

Since $W_j(h)$ is not restricted by the frequency separation D_{ji} when $h \geq 2$, we are only concerned with the case when $h = 1$. For simplicity reasons, we shall use W_j instead of $W_j(1)$ in later discussions.

Let us now introduce the actual possible shift A_j of row j. It is equal to the sum of S_i and W_j with modolo N which is needed because of (5.3). Therefore,

$$A_j \quad = \quad [\ S_i + W_j\ ] \bmod N \qquad \ldots\ldots (5.10)$$

and the number of actual possible shifts is N-D+1.

We now proceed to obtain an expression for C_{jJ} by taking the shifting of channels into consideration. If row j assumes a shift of S_j, i.e. S_j is one of the (N-D+1) A_j's and if $S_j = 0$, then from (5.2), we have for cell 0, $C_{j0} \quad = \quad N.j$

If $S_j = 1$, $\quad C_{j0} \quad = \quad N.(j + 1) - 1 \quad = \quad N.j + (N - 1)$

If $S_j = 2$, $\quad C_{j0} \quad = \quad N.(j + 1) - 2 \quad = \quad N.j + (N - 2)$

.

.

.

etc.

Then in general, $C_{j0} = N.(j + 1) - S_j \quad = \quad N.j + (N - S_j)$

Hence, with reference to (5.2),

$$C_{jJ} = N.j + [\ (N - S_j) + J\ ] \bmod N \qquad ..... (5.11)$$

Modulo N is necessary here since the term in square brackets would not exceed (N - 1) due to the cyclic shifting. The application of (5.11) is illustrated in Fig. 5.3 by an example with $N = 7$, $m = 3$ and $S_2 = 2$.

5.3.3 The Intermodulation Check

As mentioned in Section 5.3.1, the purpose of shifting the channels is to avoid the formation of IM products which fall on the allocated channels of the cell. This section deals with the intermodulation check which ensures that such products are not formed. The check is based on the property that for IM interference not to occur between the three frequencies f_x, f_y, f_z,

$$f_z - f_y \neq f_y - f_x \qquad ..... (5.12)$$

where $f_x < f_y < f_z$.

We now assume that a new row k with shift S_k, equal to one of the A_k's, is added to the frequency allocation plan which consists of IM interference-free channels. The check in (5.12), in this case, becomes:

m \ N	0	1	2	3	4	5	6
0	0	1	2	3	4	5	6
1	7	8	9	10	11	12	13
2	14	15	16	17	18	19	20

$$\Big\downarrow\; S_2 = 2$$

m \ N	0	1	2	3	4	5	6
0	0	1	2	3	4	5	6
1	7	8	9	10	11	12	13
2	19	20	14	15	16	17	18

$$C_{jJ} = C_{23} = 7(2) + (7 - 2 + 3) \bmod 7$$
$$= 14 + 1 = 15$$
$$C_{jJ} = C_{21} = 7(2) + (7 - 2 + 1) \bmod 7$$
$$= 14 + 6 = 20$$

FIGURE 5.3

FREQUENCY ALLOCATIONS : AN EXAMPLE WITH

$N = 7$, $m = 3$ AND $S_2 = 2$

$$C_{kJ} - C_{jJ} \neq C_{jJ} - C_{iJ} \quad \text{..... (5.13)}$$

where i and j are any two rows in the plan and k > j > i.

Using (5.11), (5.13) becomes,

$$[\, N.k + (N - S_k + J) \bmod N \,] - [\, N.j + (N - S_j + J) \bmod N \,]$$
$$\neq [\, N.j + (N - S_j + J) \bmod N \,] - [\, N.i + (N - S_i + J) \bmod N \,]$$

or $$N.(k - j) \neq N.(j - i) + 2(N - S_j + J) \bmod N$$
$$- (N - S_i + J) \bmod N - (N - S_k + J) \bmod N \quad \text{..... (5.14)}$$

This check is performed for three values of J, namely, $J = S_i$, $J = S_j$ and $J = S_k$ as these values represent the three unique relationships in this check.

5.3.4 The Algorithm

Based on the development described in the previous sections, the frequency allocation algorithm may now be described. It is basically divided into two parts. In the first part, it derives the relative possible shifts for all the rows in the plan. This is used to ensure that all the frequencies are free from adjacent channel interference and are separated from each other by at least D channels. In the second part, one of the relative possible shifts is selected for a particular row. Its actual possible shift is calculated and the IM check is activated. This ensures that all the frequencies are free from IM interference. The following briefly describes the basic steps taken in this algorithm. Fig. 5.4 provides the flow of the algorithm.

Step (1): Obtain values of N, m and D.

Step (2): Derive the relative possible shifts W which initially are identical for all rows:

$$0 \leq W \leq N - D$$

Also let L_j be the list of (N-D+1) relative possible shifts W_j of row j.

Step (3): Assign a shift of zero to row 0. Initialize row number k = 1.

Step (4): If k = m or if k=1 and L_k = 0, go to step (9).

Step (5): For row k, obtain the next W_k from L_k and calculate A_k. Let $S_k = A_k$.

Step (6): Perform IM check for row k:

$$N.(k - j) \neq N.(j - i) + 2(N - S_j + J) \bmod N - (N - S_i + J) \bmod N - (N - S_k + J) \bmod N$$

for all i, j less than k. If the check is successful, store S_k, increment k, go to step (4); If the check is unsuccessful, delete W_k from L_k, go to step (7).

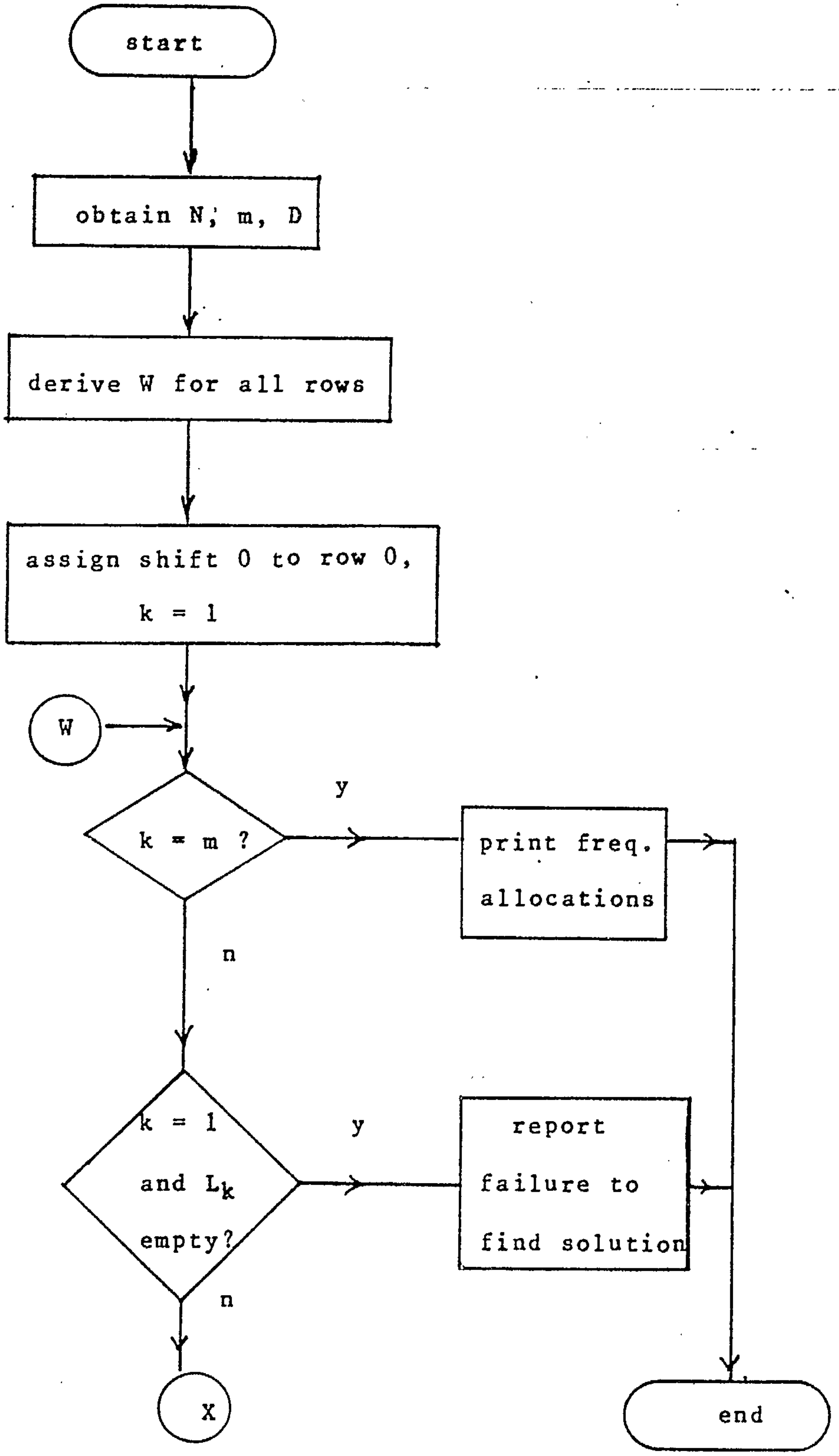


FIGURE 5.4 (to be continued)

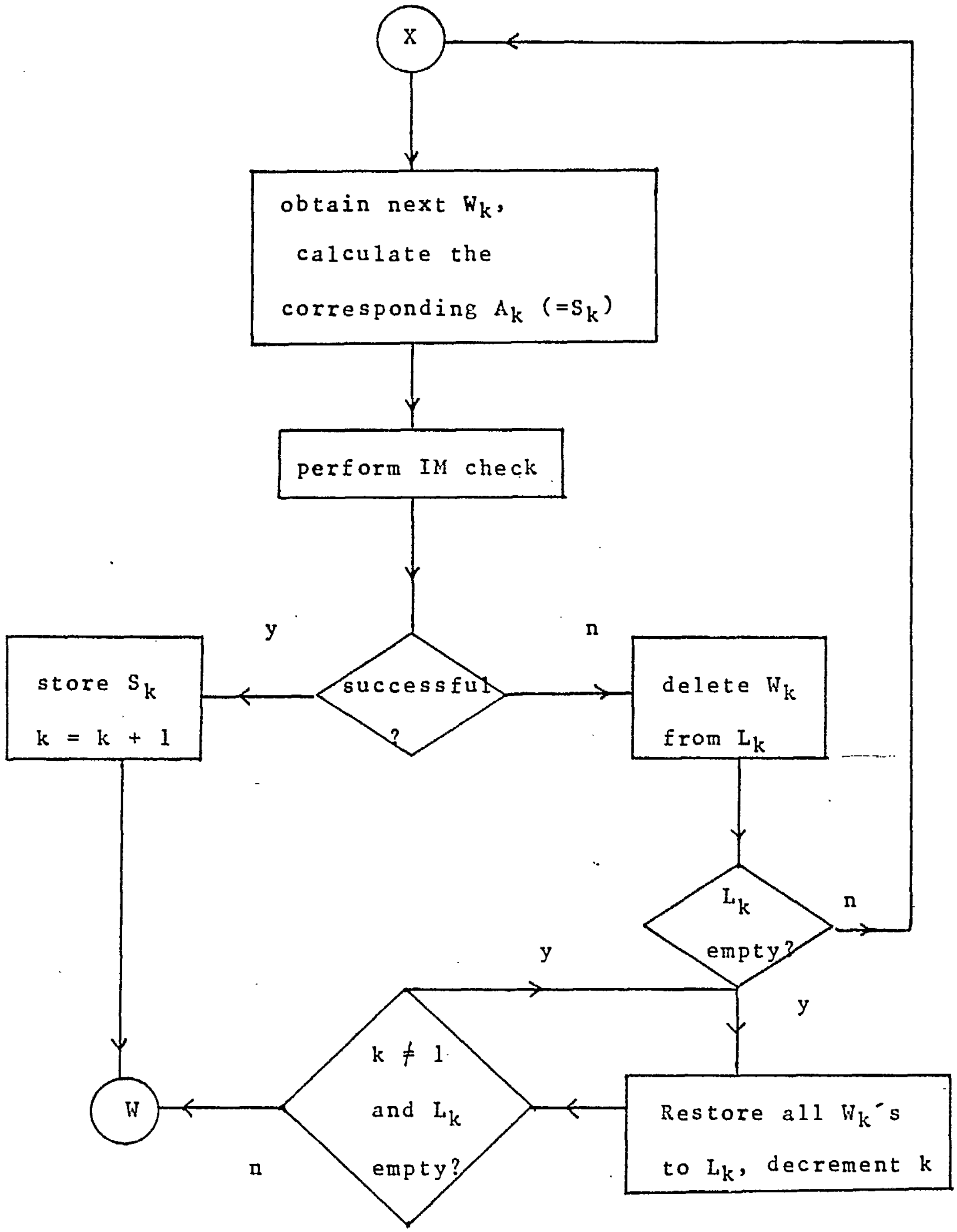


FIGURE 5.4 (cont.)

OPTIMAL FREQUENCY ALLOCATION ALGORITHM

Step (7): If L_k is not empty, go to step (5); otherwise go to step (8).

Step (8): Decrement k. If $k \neq 1$ and L_k is empty, repeat step (8); otherwise go to step (4).

Step (9): Print frequency allocations if solution is found, otherwise report no solution. End of algorithm.

5.4 Numerical Results of the Frequency Allocation Algorithm

The frequency allocation algorithm is run for a variety of cases with different values of N, m and D. These cases are shown in Table 5.1 with m varying from 3 to 35. No runs are performed for cases where D is greater than or equal to N since obviously no solution can be obtained.

Due to the large amount of data, we only provide the frequency allocations for one of the cases. This corresponds to the example where $N = 9$, $D = 6$ and $m = 8$ and is shown in Fig. 5.5. The results of the rest of the cases are given in Table 5A.1 in the Annex to this chapter. In this table, we only show the shifts of the channels; the corresponding channel numbers can be derived easily using (5.11).

In addition to Table 5A.1, we also summarize the data in a

N / D	3	4	6	7	9	12	19	21	28	31	37
2	V	V	V	V	V	V	V	V	V	V	V
4			V	V	V	V	V	V	V	V	V
6				V	V	V	V	V	V	V	V
8					V	V	V	V	V	V	V
10						V	V	V	V	V	V
12							V	V	V	V	V
16							V	V	V	V	V
20								V	V	V	V

Note: V = case where frequency allocation algorithm run was performed

TABLE 5.1

FREQUENCY ALLOCATION ALGORITHM RUNS

actual shifts	Channel Allocations for cells:								
	0	1	2	3	4	5	6	7	8
0	0	1	2	3	4	5	6	7	8
0	9	10	11	12	13	14	15	16	17
1	26	18	19	20	21	22	23	24	25
1	35	27	28	29	30	31	32	33	34
3	42	43	44	36	37	38	39	40	41
3	51	52	53	45	46	47	48	49	50
4	59	60	61	62	54	55	56	57	58
4	68	69	70	71	63	64	65	66	67

FIGURE 5.5

OPTIMAL FREQUENCY ALLOCATIONS: AN EXAMPLE WITH

$N = 9$, $D = 6$, $m = 8$

graphical form for easy presentation. Figures 5.6a) to 5.6h) indicate whether a solution can be found by using the algorithm for the different cases. We have selected values of D equal to 2, 4, 6, 8, 10, 12, 16 and 20 for different values of N as indicated in Table 5.1. The algorithm is run for each case by incrementing the number of channels, m, in steps of one from 3 to 35. In some cases, a solution may be obtained by only shifting channels in the newly-added row. But in other cases, it is necessary to change the shifts in previous rows. Whenever this occurs, the value of m is noted and is represented by a mark (-) on the vertical line in Fig. 5.6. Furthermore, an arrow in the figure represents the maximum value of m, or M, beyond which no solution can be found. It must be pointed out that the solutions are not unique and in fact this same algorithm can produce more than one solution for the same input parameters.

Two observations can be made from Figs. 5.6a) to 5.6h):

(1) The value of M increases for increasing values of N but decreases for increasing values of D.

Since from (5.4), $0 \leq W \leq N - D$, the number of relative possible shifts is $N - D + 1$. As N increases or D decreases, this number increases. This effectively increases the number of combinations that can be tried to obtain a solution and hence leads to a higher chance of getting a solution resulting in a higher value of M. For example, with $N = 3$, $D = 2$ and $m = 5$, $0 \leq W \leq 1$. There are only two possible shifts for each row giving a total of 2^4 or 16 combinations. But if $N = 9$, $D = 2$ and $m = 5$,

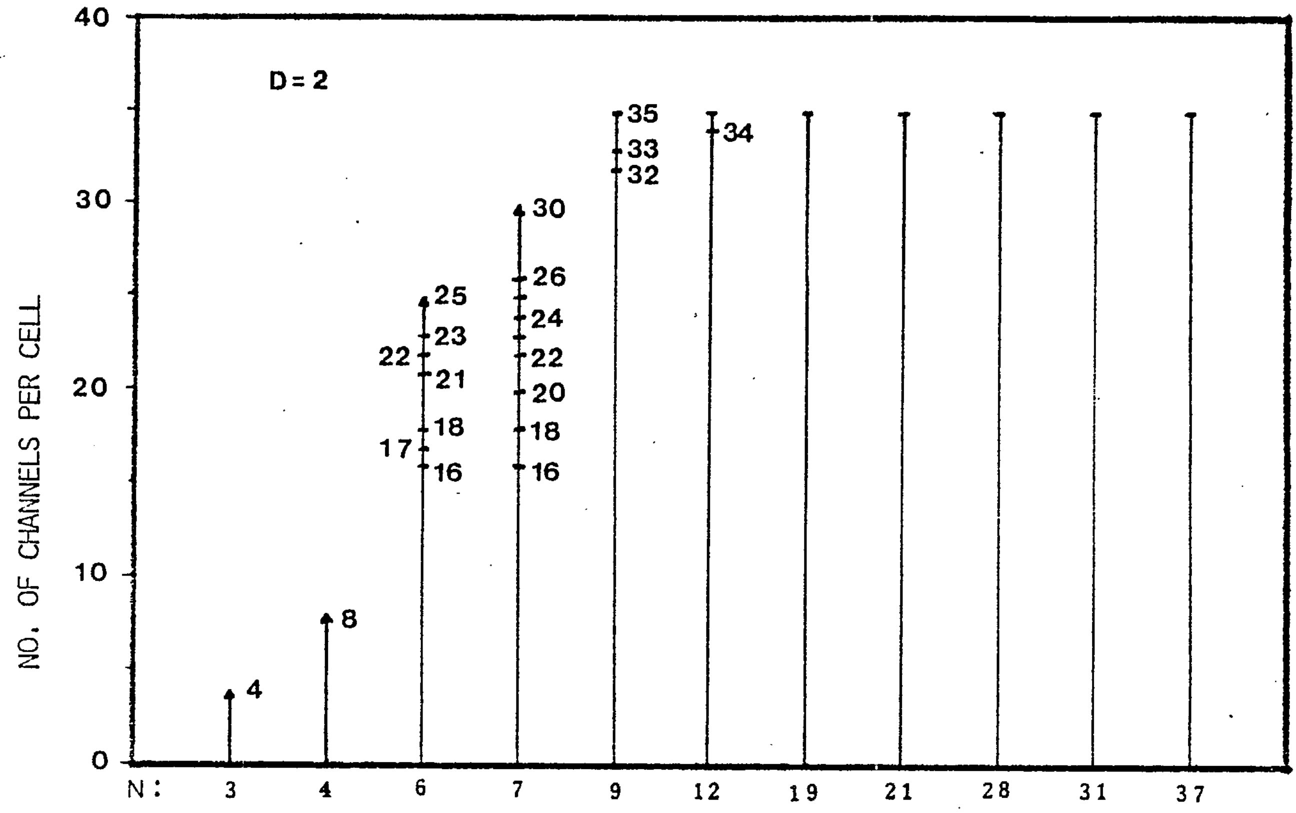


FIGURE 5.6 a) SUMMARY RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

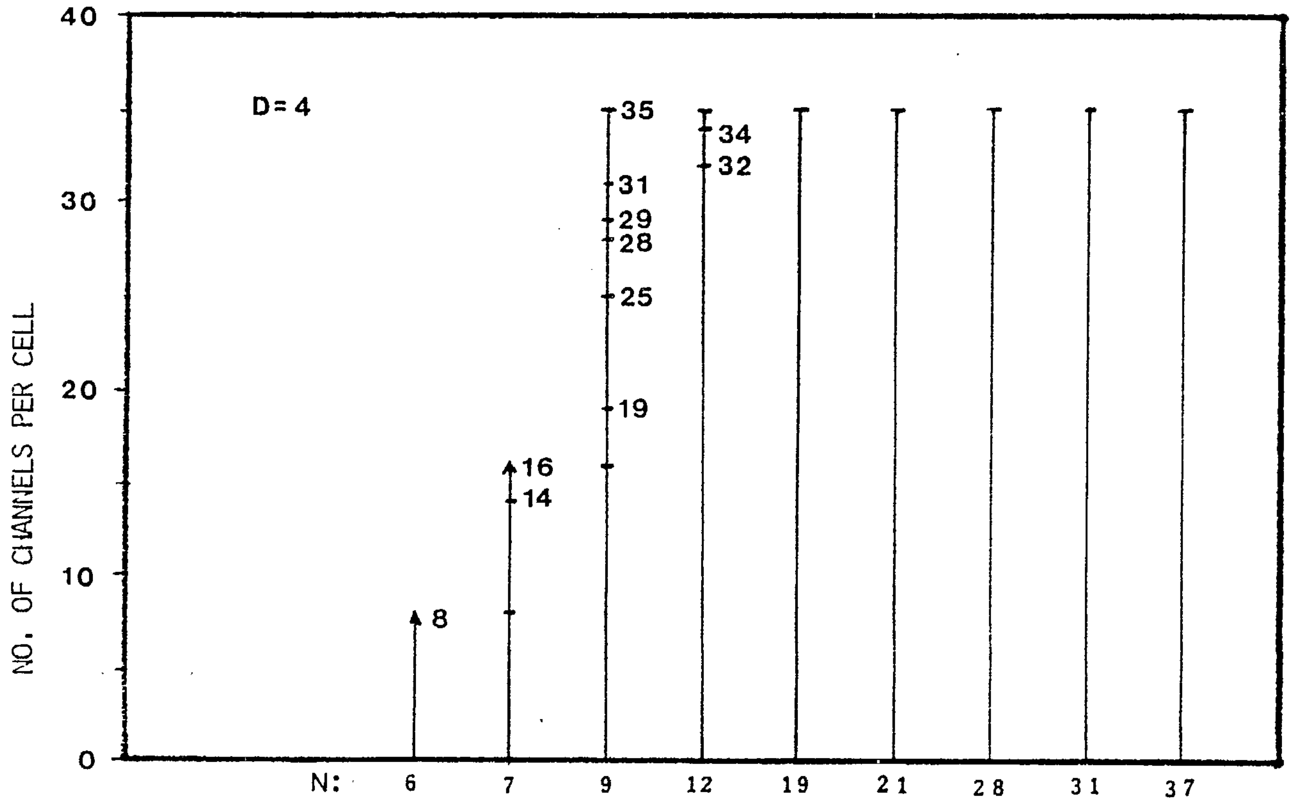


FIGURE 5.6 b) SUMMARY RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

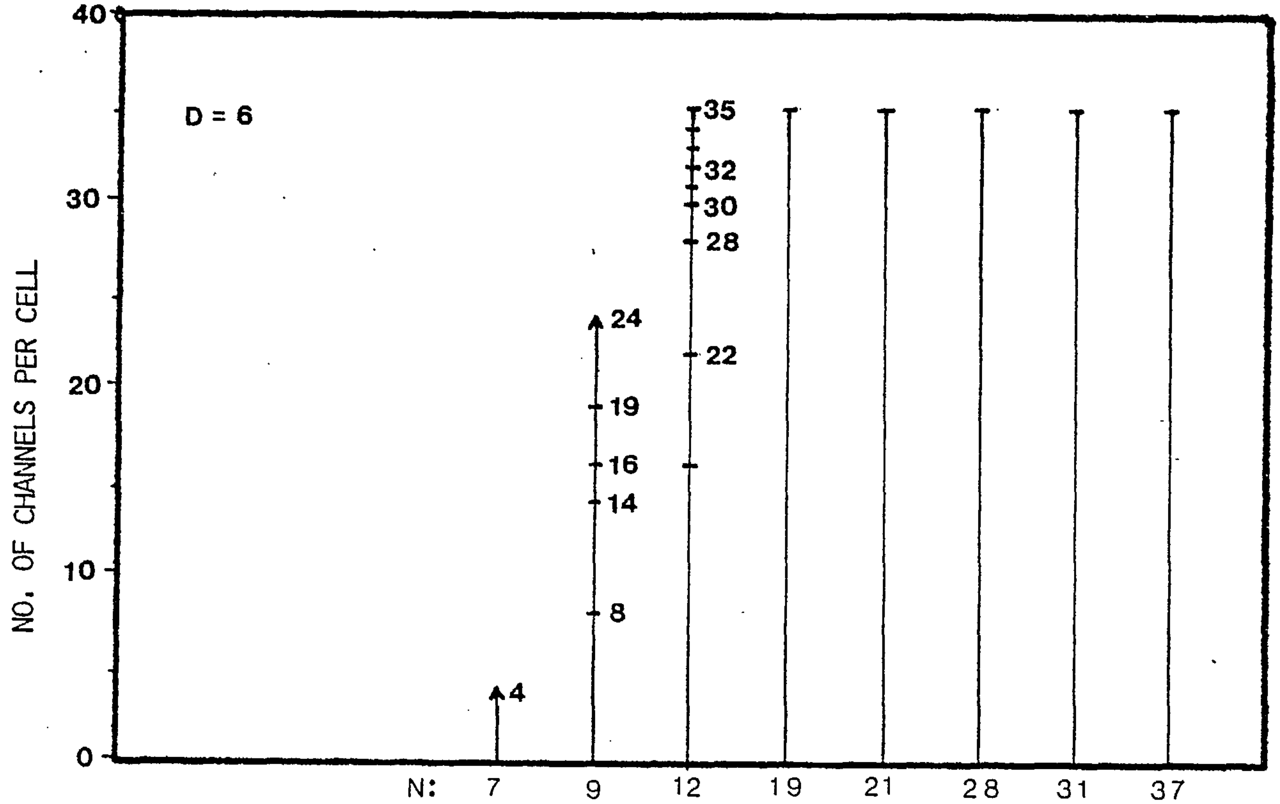


FIGURE 5.6 c) SUMMARY RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

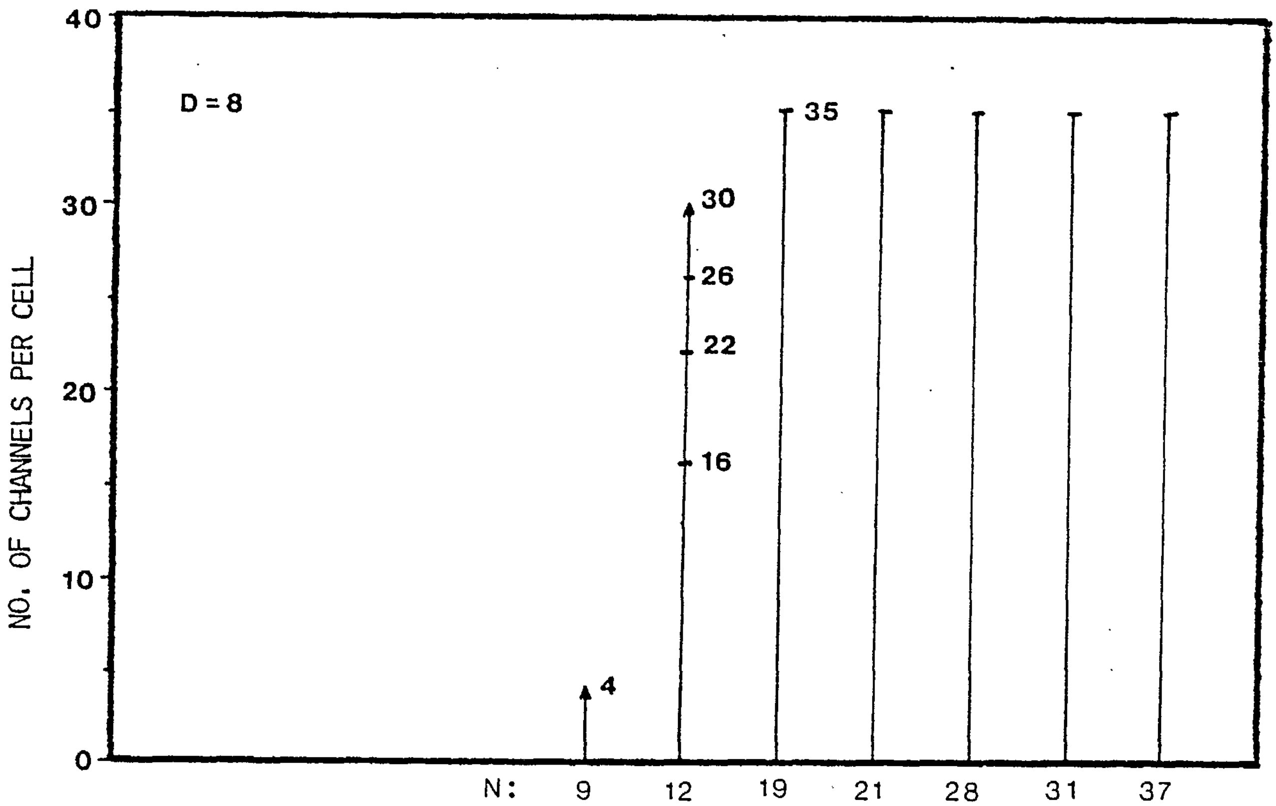


FIGURE 5.6 d) SUMMARY RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

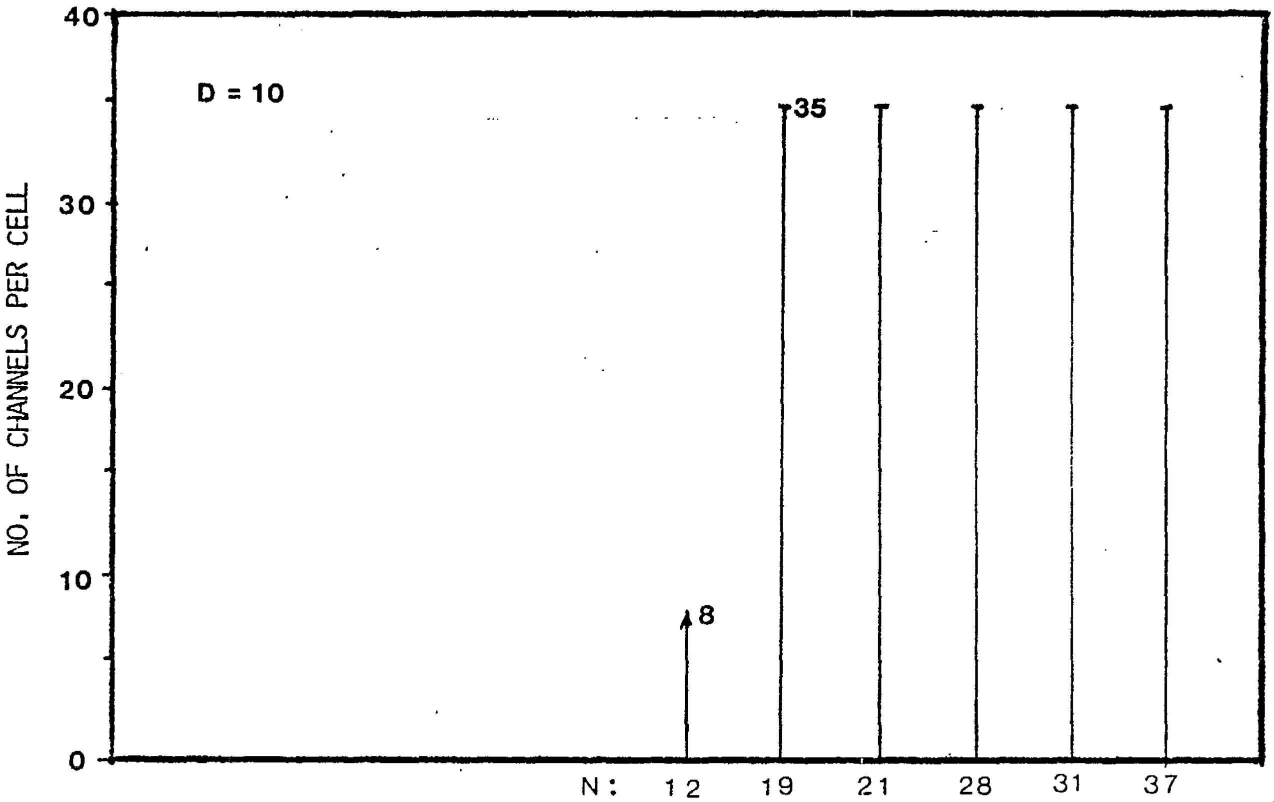


FIGURE 5.6 e) SUMMARY RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

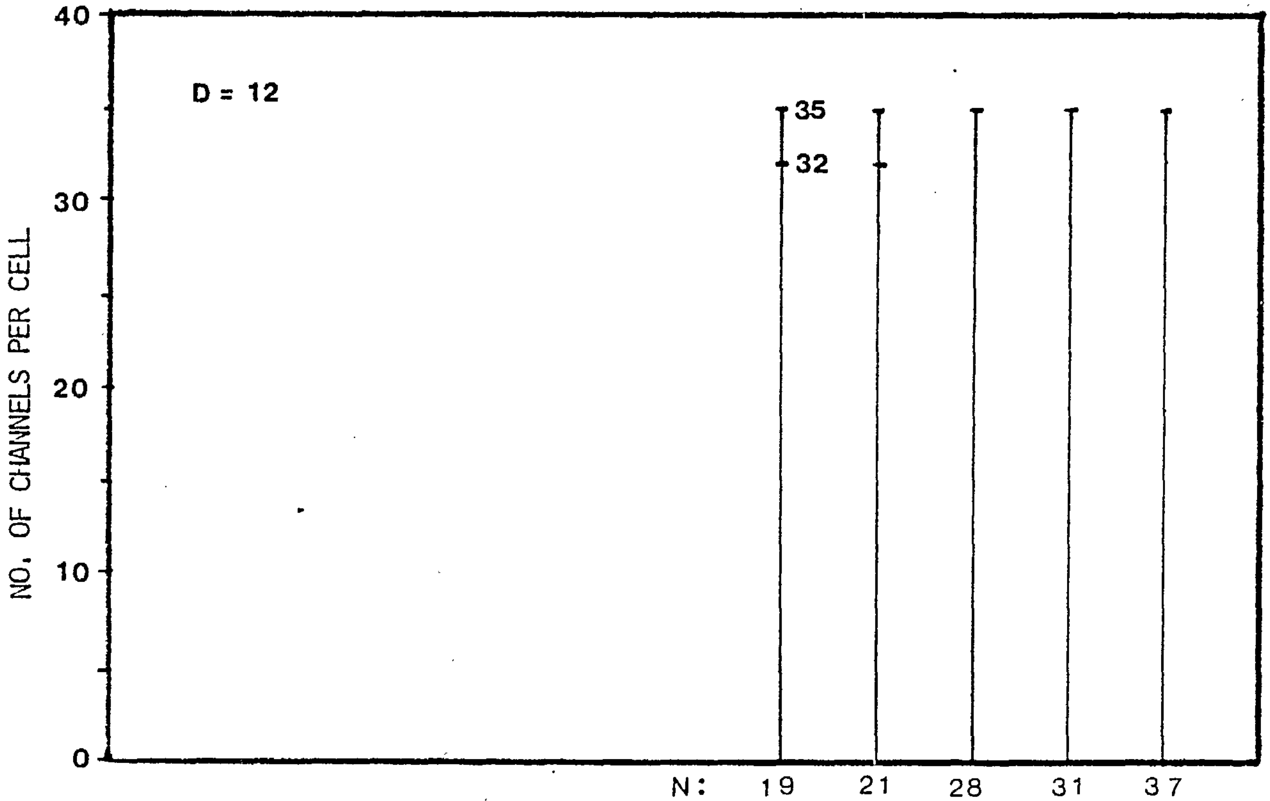


FIGURE 5.6 f) SUMMARY RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

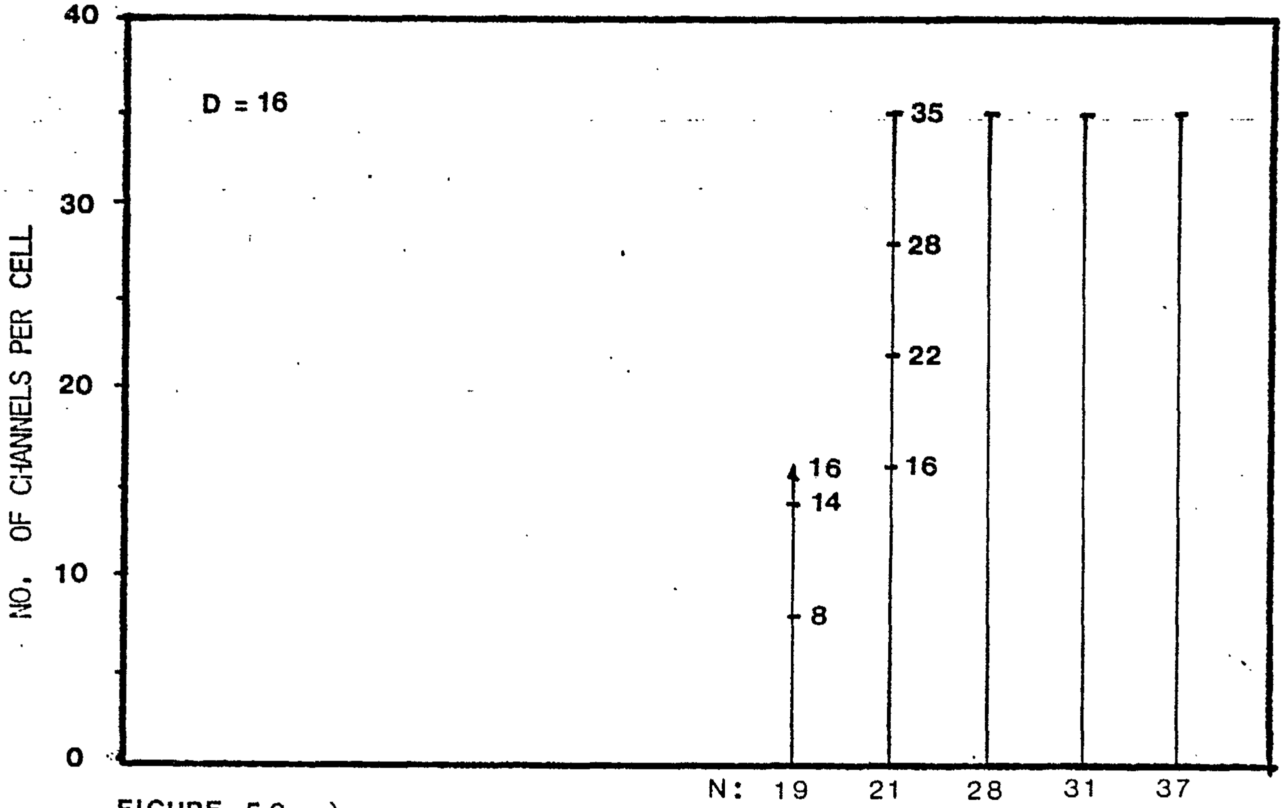


FIGURE 5.6 g)
SUMMARY RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

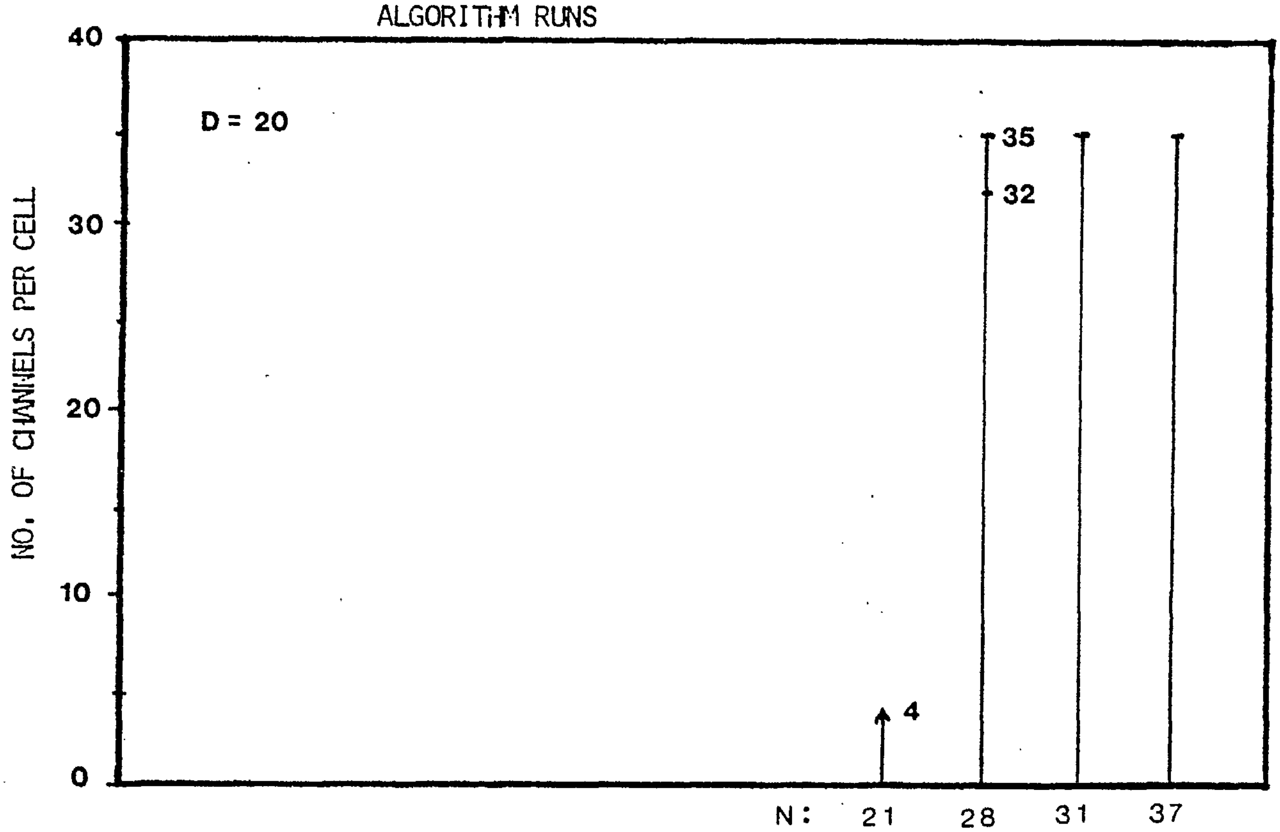


FIGURE 5.6 h)
SUMMARY RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

the number of possible shifts is 8 giving a total of 8^4 or 4096 combinations.

(2) As m is getting close to M for low values of N, the marks on the vertical line occur more frequently. This means that the shifts established for lower values of m are changed more often.

This behaviour may be explained by the fact that as m gets close to M, especially for small N, it becomes more difficult to find a solution. More tries are required and shifting channels alone in the last row may not produce a solution. Therefore it becomes more necessary to change the shifts that were previously established for lower values of m earlier in the run.

From the results obtained, it can be seen that an optimal solution may not always be attainable. Here, an optimal solution refers to one in which all the given mN channels are utilized to produce a frequency allocation which is interference-free. Out of the 55 runs we performed, 13 have M less than 35. In the next section, we attempt to find an algorithm which may find sub-optimal solutions for these cases for M = 35. A sub-optimal solution refers to one in which more than mN channels would be required for the cells so that interference-free operation is guaranteed. Since references would be made to the row numbers frequently, Fig. 5.7 is provided here to clarify the relationships between k, m and M.

1^{st} row	....	row 0,	$k = 0$		
2^{nd} row	....	row 1,	$k = 1$		
3^{rd} row	....	row 2,	$k = 2$		
	.			M rows	
	.				
	.				
$(M-1)^{th}$ row	....	row $(M-2)$,	$k = M-2$		m rows
M^{th} row	....	row $(M-1)$,	$k = M-1$		
$(M+1)^{th}$ row	....	row M,	$k = M$		
	.				
	.				
	.				
$(m-1)^{th}$ row	....	row $(m-2)$,	$k = m-2$		
m^{th} row	....	row $(m-1)$,	$k = m-1$		

FIGURE 5.7

RELATIONSHIP BETWEEN k, m AND M

5.5 The Sub-Optimal Algorithm

The optimal frequency allocation algorithm terminates either when a solution is found or when all combinations of shifts from different rows are tried and no solution can be found. The only possible way to obtain a solution in the latter case is to give up some of the channels in the new row. The sub-optimal algorithm does exactly that by eliminating the minimum number of channels from row M or the $(M + 1)^{th}$ row. As a result of this elimination, the sub-optimal algorithm finds a frequency allocation for the cells but has to use more than mN channels. As pointed out earlier, the optimal algorithm may produce more than one optimal solution at row (M-1), the process is repeated with other optimal solutions until the best sub-optimal solution is obtained. The steps of the sub-optimal algorithm is described in the following. The flow of the algorithm is shown in Fig. 5.8.

Step (1): Obtain values of N, m and D and the sequence of shifts from the optimal solution at row (M-1). Derive the relative possible shifts W_k for all rows.

Step (2): Initialize K=M, k = M and $H_k = 0$ for all k's. H_k is the dimension of the list of rejected channels for row k and row K is the row where all possible shifts tried to find a solution are exhausted.

Step (3): If k = m, go to step (8); if k = M-1, increment channel

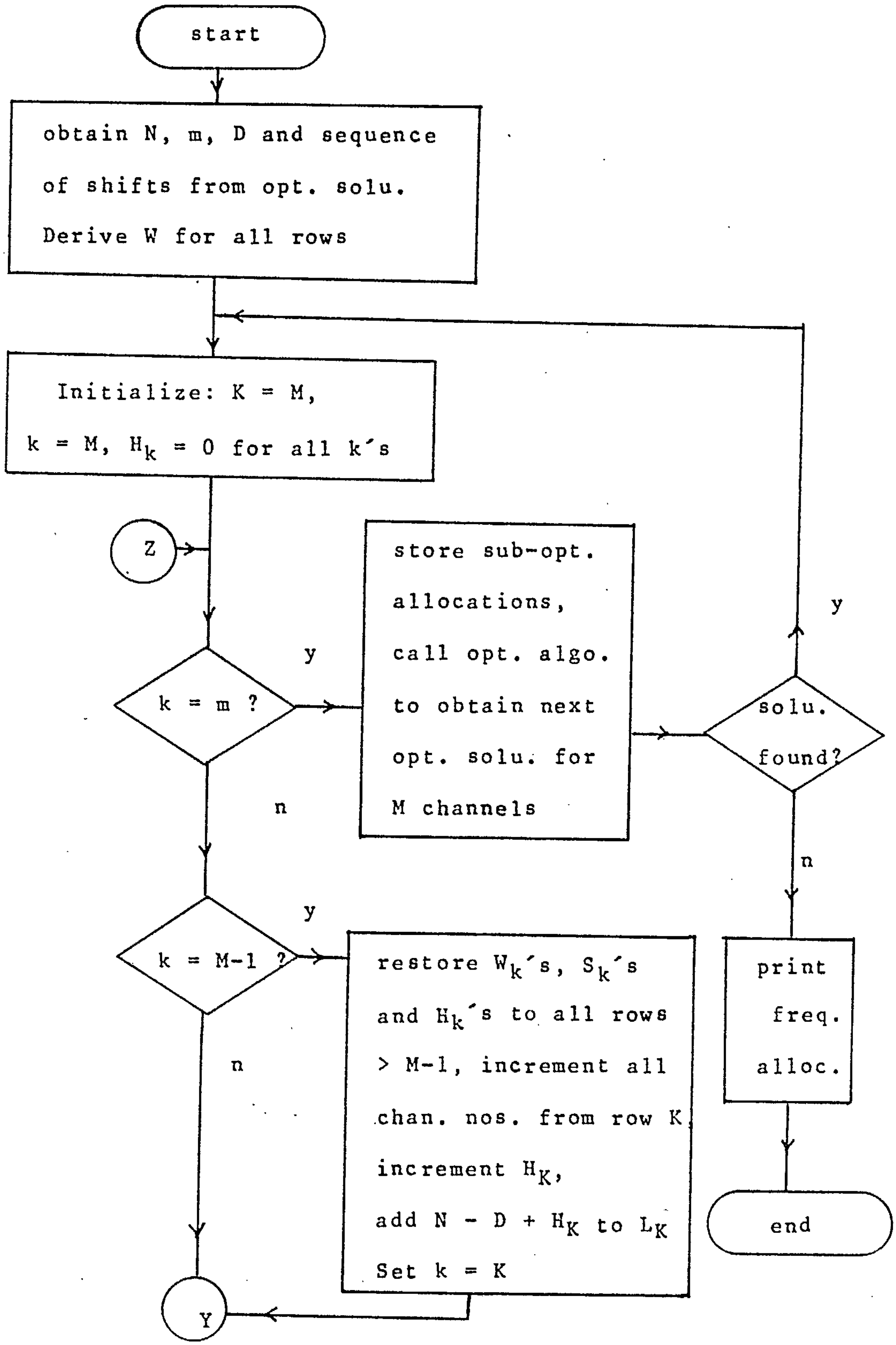


FIGURE 5.8 (to be continued)

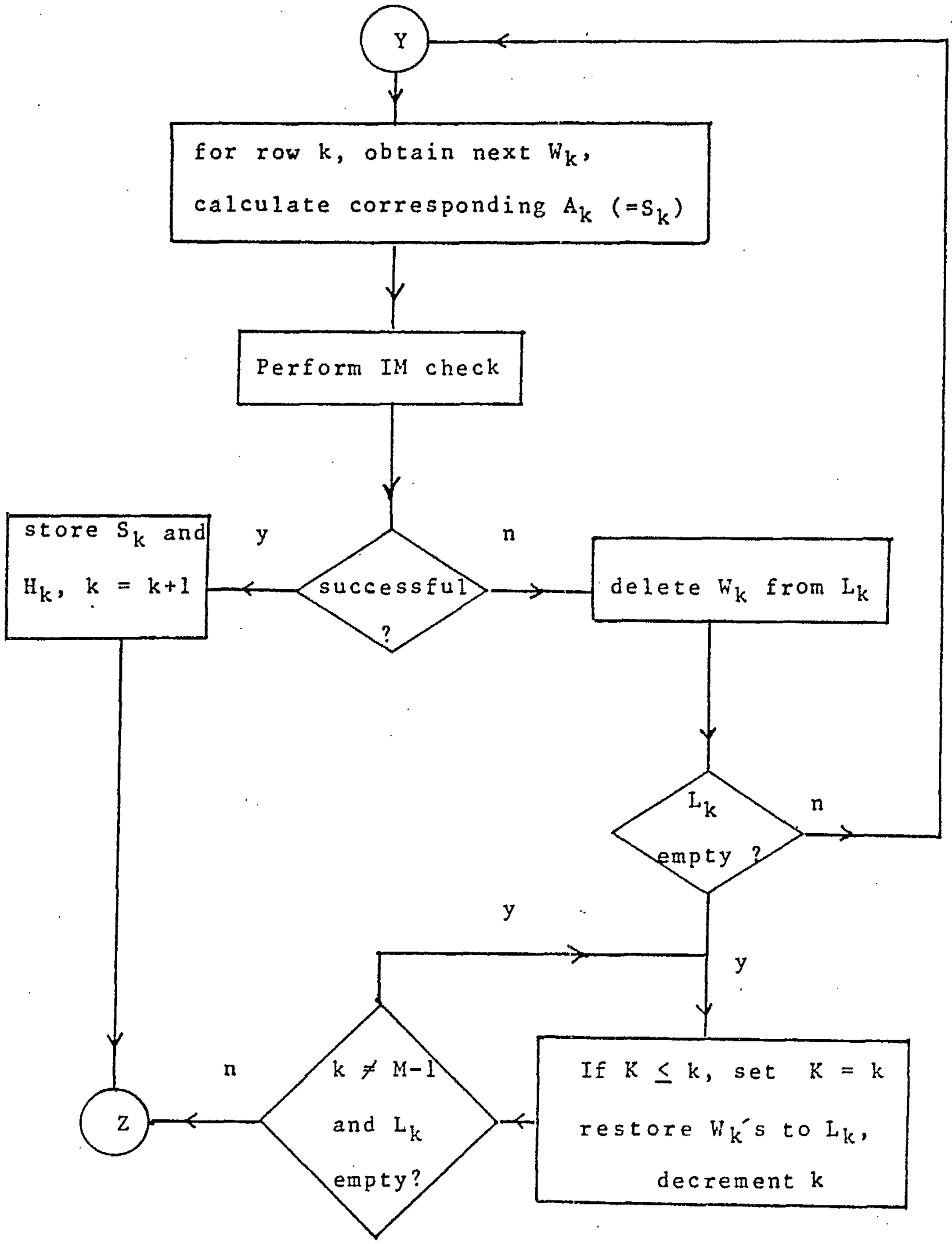


FIGURE 5.8 (cont.)

SUB-OPTIMAL FREQUENCY ALLOCATION ALGORITHM numbers for all rows starting from row K, increment H_K , restore the values of W_k's, S_k's and H_k's just before K is set to all rows starting from row M. Add $(N - D + H_K)$ to L_K, set $k = K$.

Step (4): For row k, obtain the next W_k from L_k and calculate the corresponding A_k. Let $S_k = A_k$.

Step (5): Perform IM check for row k for all i, j < k. If the check is successful, store S_k and H_k, increment k, go to step (3). If the check is unsuccessful, delete W_k from L_k, go to step (6).

Step (6): If L_k is not empty, go to step (4); otherwise go to step (7).

Step (7): Set $K = k$ if $K \leq k$, restore W_k's to L_k, decrement k. If $k \neq M - 1$ and L_k is empty, repeat step (7); otherwise go to step (3).

Step (8): Store the sub-optimal frequency allocations. Call the optimal algorithm to obtain the next optimal solution. If found, go to step (2); otherwise, print all frequency allocations and end.

It should be pointed out that in step (3), after all possible shifts are exhausted for row K, the first channel has to be deleted and replaced by the next channel. When this is done, the relative possible shift W_k becomes:

$$0 \leq W_K \leq N - D + 1$$

since the channel separation between rows K and K - 1 is automatically increased by one. In general, since H_K is the number of channels deleted from row K, for $H_K \leq D - 1$,

$$0 \leq W_K \leq N - D + H_K \qquad \text{..... (5.15)}$$

and for $H_K > D - 1$,

$$0 \leq W_K \leq N - 1 \qquad \text{..... (5.16)}$$

because of the requirement in (5.3).

The number of relative possible shifts hence becomes $(N - D + H_K + 1)$ with a maximum value of N when $H_K > D-1$.

The IM check also has to be modified because of the introduction of H_K . The channel C_{KJ} in row K now becomes,

$$C_{KJ} = NK + (N - S_K + J) \bmod N + T_K \qquad \text{..... (5.17)}$$

where $T_K = \sum_{k=M}^{K} H_k$

Therefore, the IM check stated in (5.14) becomes,

$$N(K - j) \neq N(j - i) + 2(N - S_j + J) \bmod N - (N - S_i + J) \bmod N$$
$$- (N - S_K + J) \bmod N + 2 T_j - T_i - T_K$$

where $K > j > i$, $T_j = \sum_{k=M}^{j} H_k$ and $T_i = \sum_{k=M}^{i} H_k$

It should be obvious that $T_i = T_j = 0$ for i and j less than or equal to M-1.

The sub-optimal algorithm is run for the 13 cases where no optimal solution can be found for m = 35. The results are shown in Table 5.2 and plotted in Fig. 5.9. The number of channels lost is the number of channels that have to be rejected by the algorithm in order to find a frequency allocation for all the cells. This is also the number of additional channels that have to be used, resulting in more spectrum required.

While only the minimum number of channels lost is presented in Table 5.2, our results show that the number of channels lost varies widely depending on the particular optimal solution on which the sub-optimal algorithm is based. The first optimal solution does not necessarily be the best root for the sub-optimal algorithm.

N	D	M	Total no. of sub-optimal solutions	Minimum no. of channels lost
3	2	4	1	235
4	2	8	6	132
6	2	25	4	69
	4	8	6	114
7	2	30	1	78
	4	16	4	59
	6	4	1	110
9	6	24	49	9
	8	4	1	42
12	8	30	26	2
	10	8	2	13
19	16	16	20	3
21	20	4	1	25

Note: The algorithm was run for m = 35

TABLE 5.2

RESULTS OF THE SUB-OPTIMAL FREQUENCY ALLOCATION ALGORITHM

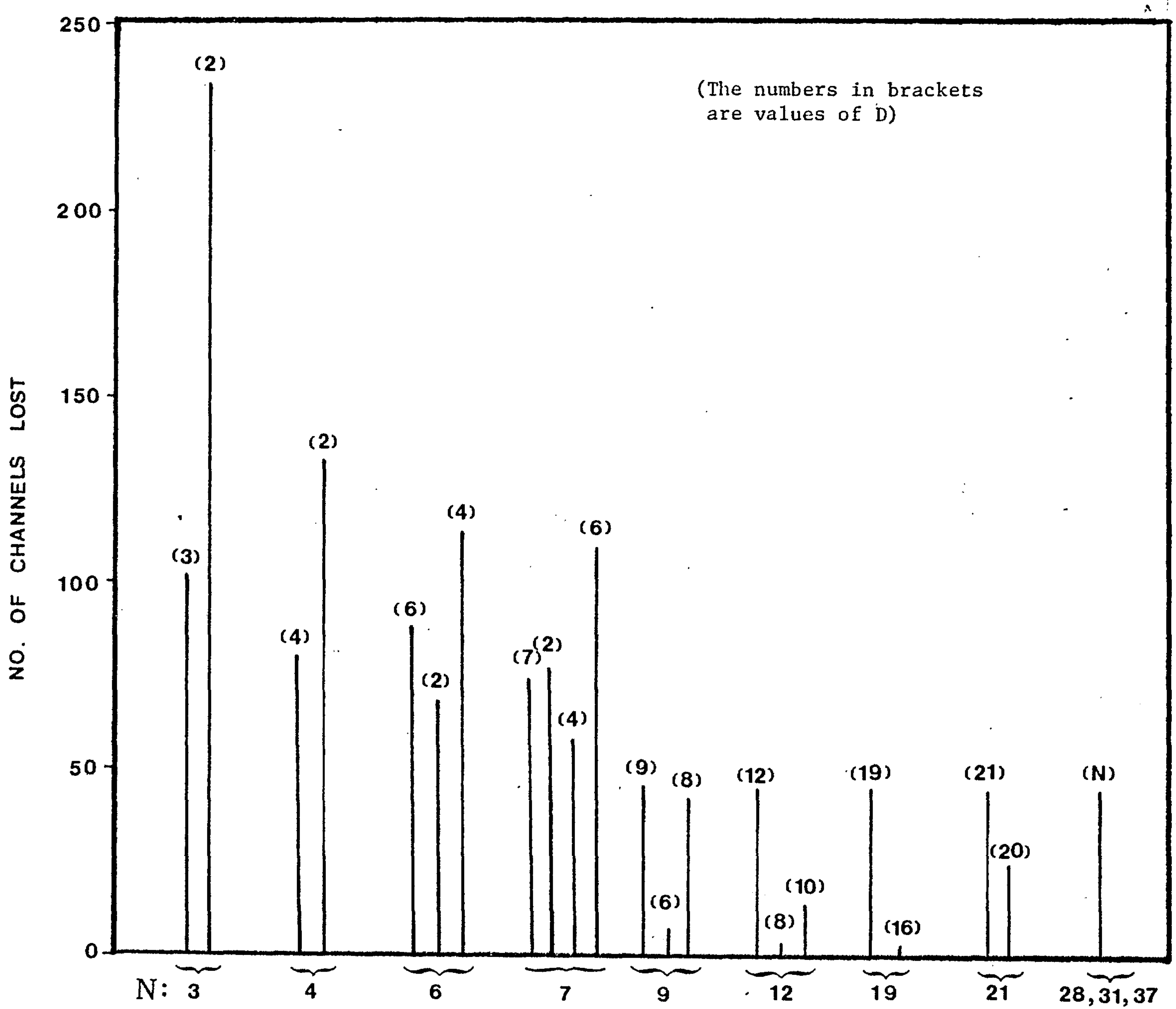


FIGURE 5.9 SUMMARY RESULTS OF THE SUB-OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

From Fig. 5.9, we can see that the number of channels lost for small N is much larger than that for large N. This leads to a question that cannot be answered from the obtained results and that is how well does the sub-optimal algorithm perform? To answer this question, we attempt to compare the performance of the sub-optimal algorithm with that of another method which we devise to obtain an IM interference-free allocation plan by simply rejecting the channels whenever an IM product falls on one of the existing frequencies. No shifting of channels is performed. This method is described in the next section.

5.6 Comparison with the Simple Channel Rejection (SCR) Method

In this method, the initial allocation plan is used as the basis. As the channels are checked for IM products, the first channel in the row is rejected whenever the IM check fails. All subsequent channels in the plan are then incremented by one. The IM check is needed only for the first cell since the frequencies in any other cell differ from those in the first cell by a constant. This method is used for all values of m from 3 to 35. The number of channels lost is plotted in Fig. 5.10 for N = 3, 4, 6, 7, 9, 12, 19, 21, 28, 31 and 37. The figure shows a certain degree of overlapping between the lines drawn for various values of N but the general trend indicates that the number of channels lost is smaller for larger values of N.

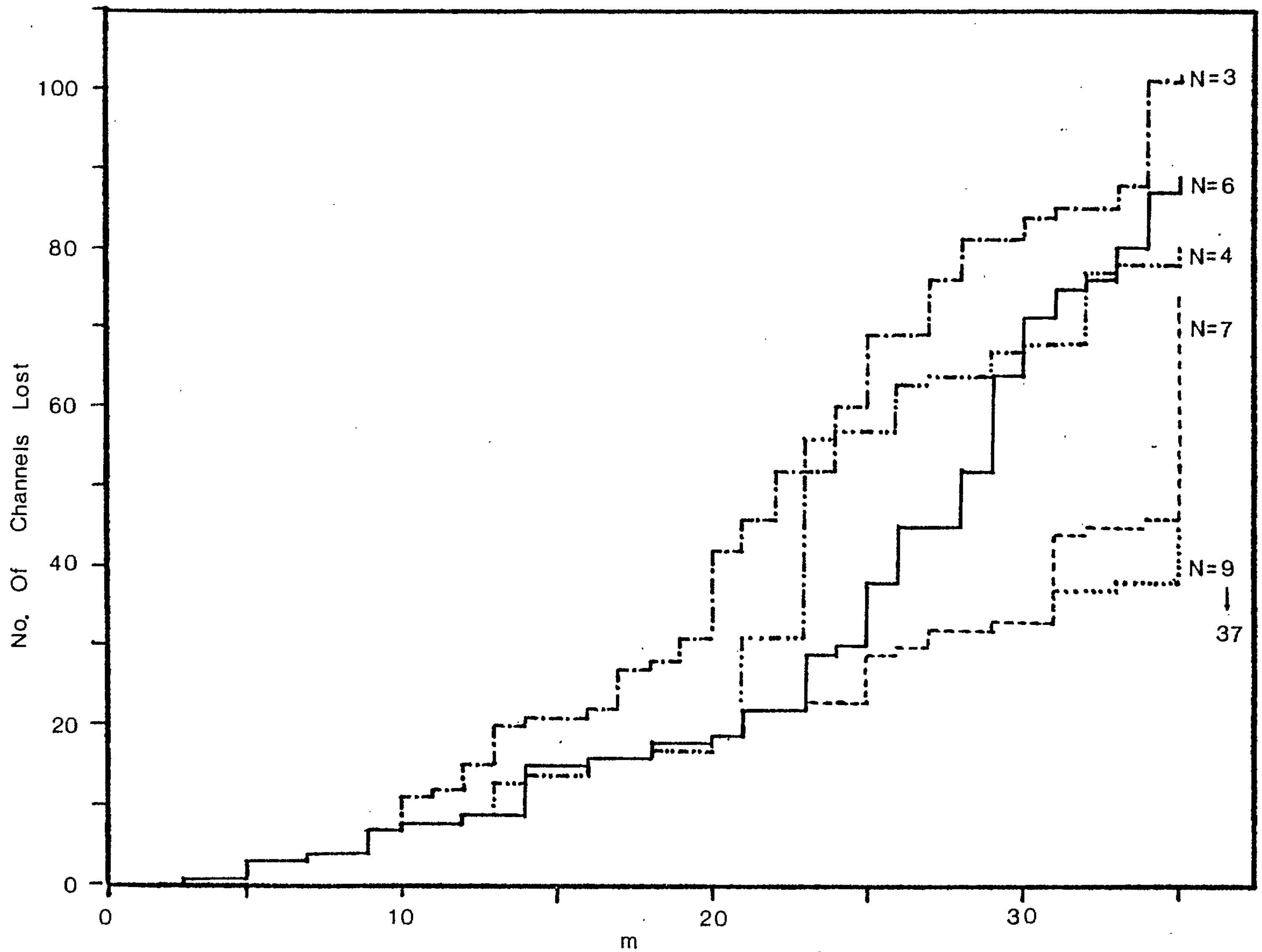


FIGURE 5.10 SUMMARY RESULTS OF THE SCR METHOD RUNS

This trend can be explained in the following. Suppose that the first k rows have been obtained using the SCR method for IM-free operation. The k channels form a total of k(k-1) IM products. A portion of these fall within the frequency band bounded by the first and the k^{th} channel and the remaining products fall either below or above the band. From (4.18) and (4.19), we can obtain the number of IM products that fall within the band, hence we can estimate the number of IM products that fall above the k^{th} channel. This number is of the order of $k^2/4$ [$k^2/4$ if k is even and $(k^2 - 1)/4$ if k is odd]. These products spread over a frequency band which is equal to the band occupied by the k channels and is in the order of mN channels. It can be seen that on the average there are about $k^2/4mN$ IM products falling on a single channel in this frequency band. As this number is inversely proportional to N, there are relatively fewer IM products falling on a single channel when N is large than when N is small. Hence the channel would experience less chance of rejection for larger values of N. This explains why there are less channels lost for large N than for small N.

In order to compare this method with the sub-optimal algorithm, we have included in Fig. 5.9 the number of channels lost at m = 35 obtained from the SCR Method. These numbers are represented by the lines where N = D in the figure. It can be seen that the sub-optimal algorithm performs better than the SCR Method when N is approximately larger than 7. But for small N, the sub-optimal algorithm rejects more channels. One of the possible reasons for this phenomenon is that while only one test is needed in the IM check for the SCR Method, three tests are required in the sub-optimal algorithm. This gives a higher probability of rejection by the IM check especially for small N in the sub-optimal algorithm.

5.7 Complexity of the Optimal Frequency Allocation Algorithm

The computational complexity of the optimal frequency allocation algorithm is a function of the three input parameters N, D and m. We may estimate the upper bound for the degree of complexity of the algorithm by calculating the total number of combinations and the number of additions, "A", multiplications, "M", and comparisons, "C", for each combination.

First, we calculate the total number of combinations formed from the possible shifts. Since there are $(N - D + 1)$ possible shifts for each row, there must be $(N - D + 1)^{m-1}$ combinations for the $(m-1)$ rows of channels. The first row always assumes a shift of zero.

Then, we assume that the frequency allocations for rows 0 to $k-1$ are successfully obtained. As row k is introduced, IM check has to be performed for the new row k with all rows j and i where $i < j < k$. We can see that for every k, j varies from 1 to $k-1$ and for every j, i varies from 0 to $j-1$. The check proceeds with $j = k-1$, $i = k-2, k-3, \ldots, 0$; $j = k-2$, $i = k-3, k-4, \ldots,$ 0 and so on until $j = 1$, $i = 0$. The check:

$$N(k - j) = N(j - i) + 2(N - S_j + J) \bmod N - (N - S_i + J) \bmod N - (N - S_k + J) \bmod N$$

as stated earlier is performed for $J = S_i$, S_j and S_k.

For each newly-introduced i, it requires 28 "A", 4 "M" and 3 "C" to obtain the terms and perform the checks. The total number of i´s for all j´s and k´s is:

$$\sum_{k=2}^{m-1} (k-1) + (k-2) + (k-3) + \ldots + 2 + 1$$

$$= \sum_{k=2}^{m-1} \sum_{w=1}^{k-1} w \quad = \sum_{k=2}^{m-1} 1/2 \, . \, k.(k-1)$$

$$= \quad 1/6 \, . \, m(m-1)(m-2)$$

For each newly-introduced j, it requires 1 "A" and 1 "M" to obtain the required term. The total number of j´s for all k´s is:

$$\sum_{k=2}^{m-1} (k-1) \quad = \quad 1/2 \, . \, (m-1)(m-2)$$

For each k, it requires 1 "A" to obtain the actual possible shift. The total number of k's is:

$$\sum_{k=2}^{m-1} k = (m-2)$$

The total number of operations "O" for a combination is therefore:

$$\text{"O"} = 1/6 \,.\, m(m-1)(m-2) \,[\, 28 \text{ "A"} + 4 \text{ "M"} + 3 \text{ "C"}]$$
$$+ 1/2 \,.\, (m-1)(m-2) \,[\, 1 \text{ "A"} + 1 \text{ "M"} \,] + (m-2) \text{ "A"}$$

$$= 1/6 \,.\, (m-2)(4m-3)(7m-1) \text{ "A"} + 1/6 \,.\, (m-1)(m-2)(4m+3) \text{ "M"}$$
$$+ 1/2 \,.\, m(m-1)(m-2) \text{ "C"}$$

Therefore the upper bound for the degree of computational complexity is $(N - D + 1)^{m-1}$ "O" . For $N = 7$, $D = 2$, $m = 10$, this value is (3404 "A" + 516 "M" + 360 "C") 10×10^6 .

5.8 Summary of Results

In this chapter, we have surveyed a number of frequency allocation methods. We have also presented a different approach to obtain a spectrum-efficient frequency allocation scheme which satisfies all the stated requirements. This scheme is suitable not only to cellular radio systems, but also to any multiple-site multiple-channel land mobile systems in any frequency band.

The mathematical derivation is given in Section 5.3 which also outlines the basic steps of the frequency allocation algorithm. We have tested the capability of this algorithm by running large number of cases with different values of N, D and m. While in most cases optimal solutions have been satisfactorily obtained, the algorithm failed to provide the required solution in a few cases.

To overcome the situations where no solution is found, we introduced a sub-optimal algorithm which presents sub-optimal solutions by rejecting a minimum number of channels. Hence a compromise is struck between maximum spectrum efficiency and interference-free operation. This sub-optimal algorithm is compared with a Simple Channel Rejection method which simply rejects a channel if the IM check fails and does not invoke channel shifting. We find that the sub-optimal algorithm is more spectrum-efficient than the SCR Method when N is larger than 7.

In Section 5.7, an estimate of the upper bound for the degree of computational complexity of the optimal algorithm is derived. While the number of computations, including additions, multiplications and comparisons per combination is reasonable, the number of combinations is extremely high. However, this upper bound is not reached in most runs since only a small percentage of the combinations are involved and the remaining combinations are rejected by the IM check early in the run.

ANNEX -- TABLE 5A.1

RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS:

N	D	M	shifts for rows 0-9	10-19	20-29	30-34
3	2	4	0 0 1 1			
4	2	8	0 0 1 1 3			
			3 0 0			
6	2	16	0 0 1 1 3	3 3 5 5 0		
			3 4 4 2 2	0		
		17	0 0 1 1 3	3 4 4 0 4		
			3 1 4 1 3	0 1		
		18	ditto	3 4 0 4 4		
				0 1 3		
		21	0 0 1 1 4	1 3 4 4 1	3	
			4 0 0 3 3	3 1 3 3 1		
		22	0 0 1 4 0	1 1 3 3 1	0 3	
			4 0 3 3 4	3 1 4 0 4		
		23	0 0 2 0 2	2 5 3 3 5	2 0 2	
			3 3 0 2 5	5 3 3 5 0		
		25	0 2 0 2 3	3 5 5 3 5	5 2 0 2 0	
			5 0 2 3 5	3 0 2 5 3		
6	4	8	0 0 1 1 3			
			3 4 4			

N	D	M	shifts for rows 0-9	10-19	20-29	30-34
7	2	16	0 0 1 1 3	2 2 4 4 5		
			3 4 4 1 1	5		
		18	ditto	2 2 4 2 4		
				4 5 5		
		20	ditto	2 4 4 2 3		
				3 6 6 1 6		
7	2	22	0 0 1 1 3	2 2 6 6 4	5 5	
			3 4 1 4 4	4 5 5 0 0		
		23	0 0 1 1 3	3 3 1 3 3	3 4 4	
			3 0 0 1 1	4 4 1 1 3		
		24	ditto	3 3 1 3 3	1 2 4 4	
				4 4 1 1 3		
		25	0 0 1 1 3	0 2 2 0 2	2 2 5 5 0	
			3 0 0 1 1	2 3 3 5 5		
		26	0 0 1 1 4	3 5 5 3 4	1 1 6 6 1	
			1 4 4 5 5	4 0 4 0 0	1	
		30	0 0 2 2 0	1 3 1 1 5	3 3 4 4 6	
			0 2 2 3 3	1 5 5 3 5	6 4 4 6 6	
7	4	8	0 0 1 1 3			
			3 4 4			
		14	0 0 1 1 3	3 3 4 4		
			3 4 0 1 1			

N	D	M	0-9	10-19	20-29	30-34
			shifts for rows			
		16	0 0 1 1 4	3 3 6 6 1		
			4 6 6 2 2	1		
7	6	4	0 0 1 1			
9	2	32	0 0 1 1 3	0 0 2 2 3	4 4 5 5 0	4 4
			3 4 4 8 8	3 1 1 2 2	0 1 1 3 3	
		33	ditto	0 0 2 2 0	4 4 8 8 1	3 4 4
				0 1 1 3 3	1 5 5 1 5	
		35	0 0 1 1 3	1 1 3 3 4	4 0 0 1 1	4 0 0 1 1
			3 4 4 0 0	4 1 3 3 4	3 3 4 4 0	
9	4	16	0 0 1 1 3	0 0 2 2 3		
			3 4 4 8 8	3		
9	4	19	ditto	1 1 2 5 1		
				1 3 3 4		
		25	ditto	1 1 2 5 1	8 8 1 1 5	
				1 4 4 5 5		
		28	ditto	1 1 5 5 1	4 4 6 6 2	
				1 4 4 0 0	2 6 6	
		29	0 0 1 1 3	1 1 3 4 0	6 0 0 1 1	
			3 4 4 0 0	1 1 4 4 6	4 4 6 6	
		31	ditto	1 1 4 0 0	4 0 0 4 0	4
				1 1 3 3 4	1 1 3 3 4	
		35	ditto	ditto	ditto	4 0 0 1 3
9	6	8	0 0 1 1 3			
			3 4 4			

N	D	M	shifts for rows 0-9	10-19	20-29	30-34
		14	0 0 1 1 3	1 1 2 2		
			3 4 7 8 8			
		16	0 0 1 1 4	1 1 4 4 6		
			4 6 6 0 0	6		
		19	ditto	1 1 4 6 0		
				1 1 4 4		
		24	0 0 1 1 4	4 6 6 0 0	1 1 4 4	
			6 0 1 1 4	1 1 4 6 0		
9	8	4	0 0 1 1			
12	2	34	0 0 1 1 3	9 9 11 11 0	10 10 11 11 3	7 7 5 5
			3 4 4 8 8	0 7 7 8 8	3 4 4 6 6	
		35	ditto	ditto	ditto	7 4 5 5 8
12	4	32	ditto	ditto	ditto	7 7
12	4	34	0 0 1 1 3	9 9 11 11 0	1 1 2 2 4	7 7 9 9
			3 4 4 8 8	0 7 7 8 11	0 0 4 6 6	
12	4	35	ditto	9 9 11 11 0	10 10 2 2 4	7 8 10 10
				0 7 7 8 9 9	4 0 2 3 3	11
12	6	16	ditto	9 9 11 11 0		
				0		
		22	ditto	9 9 11 11 0	10 10	
				5 7 7 8 8		
		28	ditto	ditto	0 0 1 1 3	
					3 4 4	

N	D	M	0-9	10-19	20-29	30-34
		30	ditto	9 9 11 11 0	0 0 1 1 5	
				5 9 9 10 10	5 6 6 8 8	
		31	ditto	9 9 11 11 0	1 1 3 3 6	10
				6 9 9 10 10	6 7 1 6 10	
		32	ditto	ditto	1 1 3 3 6	10 10
					7 1 1 6 7	
		33	ditto	ditto	1 1 3 3 6	10 1 1
					0 4 4 6 10	
		34	ditto	ditto	1 1 3 3 7	11 11 2 3
					1 4 4 10 10	
		35	ditto	ditto	ditto	11 11 3 3
						4
12	8	16	ditto	9 9 11 11 0		
				0		
		22	ditto	10 10 2 5 7	2 2	
				7 10 10 11 11		
12	8	26	0 0 1 1 3	9 9 11 11 9	4 4 6 6 7	
			3 4 4 8 8	9 11 11 3 3	7	
		30	0 0 1 1 3	10 10 0 0 4	3 3 4 4 6	
			3 7 7 9 9	7 9 9 1 1	10 10 0 0	
19	2	35	0 0 1 1 3	9 9 11 11 12	3 3 4 4 8	12 12 1 3 3
			3 4 4 8 8	12 0 0 1 1	8 9 9 11 11	
19	4	35	ditto	ditto	ditto	ditto

N	D	M	0-9	10-19	20-29	30-34
			shifts for rows			
19	6	35	ditto	ditto	ditto	ditto
19	8	35	ditto	ditto	ditto	ditto
19	10	35	ditto	ditto	ditto	ditto
19	12	32	ditto	ditto	3 3 4 4 8	12 12
					8 9 9 11 11	
19	16	35	ditto	ditto	3 3 4 4 8	3 3 7 7 10
					14 17 17 0 0	
21	2	35	ditto	9 9 11 11 12	1 1 2 2 6	10 10 20 20 2
				12 19 19 20 20	6 7 7 9 9	
21	4	35	ditto	ditto	ditto	ditto
21	6	35	ditto	ditto	ditto	ditto
	8	35	ditto	ditto	ditto	ditto
	10	35	ditto	ditto	ditto	ditto
21	12	32	ditto	ditto	ditto	10 10
		35	ditto	ditto	ditto	10 18 20 20 2
21	16	16	ditto	9 9 11 11 12		
				12		
		22	ditto	9 9 11 11 12	1 1	
				17 19 19 20 20		
21	16	28	0 0 1 1 3	9 9 11 11 12	3 3 4 4 6	
			3 4 4 8 8	17 19 19 20 20	6 7 7	
		35	ditto	9 9 11 11 12	2 2 7 10 15	5 5 7 7 8
				12 17 19 1 1	15 17 17 0 0	

N	D	M	shifts for rows 0-9	10-19	20-29	30-34
	20	4	0 0 1 1			
28	2	35	0 0 1 1 3	9 9 11 11 12	22 22 23 23 27	3 3 13 13 16
			3 4 4 8 8	12 19 19 20 20	27 0 0 2 2	
28	4	35	ditto	ditto	ditto	ditto
	6	35	ditto	ditto	ditto	ditto
	8	35	ditto	ditto	ditto	ditto
	10	35	ditto	ditto	ditto	ditto
	12	35	ditto	ditto	ditto	ditto
28	16	35	ditto	ditto	ditto	ditto
	20	32	ditto	ditto	ditto	3 3
		35	ditto	ditto	ditto	3 11 13 13
						16
31	2	35	ditto	ditto	22 22 23 23 27	0 0 10 10 13
					27 28 28 30 30	
31	4	35	ditto	ditto	ditto	ditto
	6	35	ditto	ditto	ditto	ditto
	8	35	ditto	ditto	ditto	ditto
	10	35	ditto	ditto	ditto	ditto
	12	35	ditto	ditto	ditto	ditto
	16	35	ditto	ditto	ditto	ditto
	20	35	ditto	ditto	ditto	ditto
37	2	35	ditto	ditto	ditto	31 31 4 4 7
37	4	35	0 0 1 1 3	9 9 11 11 12	22 22 23 23 27	31 31 4 4 7
			3 4 4 8 8	12 19 19 20 20	27 28 28 30 30	

N	D	M	shifts for rows 0-9	10-19	20-29	30-34
37	6	35	ditto	ditto	ditto	ditto
	8	35	ditto	ditto	ditto	ditto
	10	35	ditto	ditto	ditto	ditto
	12	35	ditto	ditto	ditto	ditto
	16	35	ditto	ditto	ditto	ditto
	20	35	ditto	ditto	ditto	ditto

TABLE 5A.1

RESULTS OF THE OPTIMAL FREQUENCY ALLOCATION ALGORITHM RUNS

CHAPTER 6 - CONCLUSIONS AND SUGGESTIONS FOR FURTHER WORK

6.1 Conclusions

The work described in this dissertation attempted to achieve two objectives:

(1) To provide a better understanding of the impact of interference in land mobile channels on a cellular radio system.

(2) To investigate and design a spectrum-efficient frequency allocation scheme which will improve the performance of the system by providing interference-free operations to the cellular system.

The first objective was achieved by conducting an analysis on the characteristics of the mobile channel together with the three most common types of interference and their effect on the system performance. The framework of the system was first laid down through the description of a system model. It was then shown that the cell radius has a large impact on the design of the system. Together with other parameters, the cell radius determines the transmitter output power, the number of voice and control channels required in each cell to meet certain traffic loading requirements and the spectrum efficiency. While the lower bound of the cell radius is determined by the cost factors associated with the large number of base stations and the capability of the hand-off algorithm, the upper bound is limited by the maximum transmitter power, the number of channels required, the higher probability of interference and the boundary of the service area for maintaining frequency reuse.

In the transmitter output power model, we showed that it is necessary to use higher transmitter power to compensate for fading and shadowing losses so as to maintain an acceptable S/N ratio. In determining the number of voice channels required in a cell, we used the Erlang B expression. As for the digital control channels, we studied the effect of fading on the probability of receiving a packet successfully. Using the DPSK modulation technique and the Pure Aloha random access scheme, we showed that in most situations, only one control channel would be required in each cell to support the mobile users.

The co-channel interference which has an important role to play in determining the spectrum efficiency in a cellular system was thoroughly analysed. The co-channel interference situations with fading only, fading and shadowing, shadowing only, and no fading and no shadowing were explored with considerations given to the presence of multiple co-channel interferors. It was shown that the degree of shadowing is largely determined by the standard deviation σ and that the co-channel interference is approximately the same for the case where there is only fading and the case where there is only shadowing with $\sigma = 6$. Previous work [JAKE74], [KELL73], [LEUN81], [HATA81] showed that diversity techniques are the best way to combat fading. Our work indicated that shadowing could be alleviated if the corner-illuminated configuration is used. It was also shown that this configuration requires a smaller reuse distance for the same probability of co-channel interference.

Two other types of interference, namely, the adjacent channel and intermodulation interference, were thoroughly analysed for a cellular system. The inter-cell and intra-cell situations with both uplink and downlink transmissions were considered. The analysis showed that only intra-cell interference poses a problem and that adjacent channel interference on the uplink channel and intermodulation interference on both the uplink and downlink channels need to be reduced.

The second objective of this dissertation was achieved by proposing an optimal algorithm which allocates interference-free frequencies to the cells of the cellular system. It is optimal in the sense that all channels are utilized and that there is no waste of spectrum. This algorithm accepts as inputs the three parameters, namely, the number of cells per cluster N, the number of channels required per cell m and the minimum separation in channel spacings D between frequencies. The algorithm generates as outputs solutions to the frequency-cell allocation problem. Satisfactory solutions for most cases tested were produced except for those in which the difference between the values of N and D is small. This is due to the fact that in these cases the number of relative possible shifts is small, resulting in a lower number of combinations which have to pass the stringent IM check.

In order to find solutions for these cases, we developed a sub-optimal algorithm which rejects some of the channels that fail the IM check. In comparison with the Simple Channel Rejection (SCR) method, this algorithm was shown to be more efficient in terms of the number of channels rejected when N is greater than 7. For cases where N is less than 7, the SCR method should be used.

The optimal frequency allocation scheme was shown to have a very large upper bound for its computational complexity which is mainly due to the large number of combinations. It was pointed out that most of the combinations were rejected early in the run by the IM check and in most cases, the algorithm found a solution in less than a few seconds.

6.2 Suggestions For Further Work

The work described in this report represents important contribution towards achieving a better understanding of the impact of interference in the mobile channels on a cellular system and towards the development of spectrum-efficient interference-free frequency allocation schemes. The following is a summary of the areas that require further research:

* Investigation of the capacity and spectrum efficiency of the cellular system when data and voice are transmitted.

* Investigation of the behaviour of co-channel, adjacent channel and intermodulation interference in a cellular system with data only transmissions as well as with both voice and data transmissions.

* Investigation of the improvement in spectrum efficiency when multiple source co-channel interference is reduced by the introduction of diversity techniques.

* In a cellular system with non-uniform traffic load in the cells, it is necessary to allocate different number of channels to different cells. An interference-free frequency allocation scheme is needed that will provide maximum utilization of the spectrum.

* In dynamic and hybrid frequency assignment schemes, channels are borrowed from another cell or drawn from a central pool of reserved channels if the traffic in a particular cell increases. It is highly desirable to incorporate this capability in the optimal frequency allocation algorithm to provide better traffic handling performance.

APPENDIX A - FREQUENCY ASSIGNMENT SCHEMES

A.1 Introduction

One of the most important aspects in the design of a cellular radio system is the choice of the frequency assignment scheme. The choice of the scheme involves specifying a set of channels and assigning them at each given point in time to the various cells of the system. The purpose of the assignment scheme is two-fold. The first one is to minimize the blocking probability of a call for a given acceptable grade of service. The second one is to avoid mutual interference between assigned frequencies.

Over the past few years, numerous researchers have been working on new frequency assignment schemes so as to maximize the traffic carrying capability of the cellular system. Very little work has been performed to come up with a scheme to deal specifically with the interference problem. There are basically three types of assignment schemes: the fixed, dynamic and hybrid schemes. They are briefly discussed in the following sections.

A.2 Fixed Frequency Assignment Schemes

In this scheme, frequencies are assigned to specific cells of the cellular system. While the number of cells in a cluster is a function of the reuse distance and the acceptable level of co-channel interference, the number of frequencies per cell depends on the expected level of traffic carried in each cell. The major drawback of this scheme occurs when the traffic loading condition varies from one cell to another resulting in the possibility that one cell may be experiencing high blocking probability while the neighbouring cell has a lot of idle channels.

A number of fixed frequency assignment schemes exist such as the ones described by MacDonald in [MACD79] and Gamst in [GAMS82]. The later used algorithms to generate frequency allocations with best adjacent channel distance from one cell to another.

A.3 Dynamic Frequency Assignment Schemes

A fixed frequency assignment scheme would be all that is needed if the amount of traffic in each cell is constant. However, due to the dynamic nature of the mobiles, the amount of traffic offered in each cell is constantly changing. Therefore, in order to maintain an acceptable grade of service, the number of channels required would have to be varied according to the traffic demand and the fixed frequency assignment scheme becomes inadequate. In the dynamic scheme, no channels are assigned to a specific cell. But rather, they are grouped together in a central pool and are assigned on demand.

The concept of using the dynamic frequency assignment scheme was first described by Schiff in [SCHI70]. He first derived an expression for calculating the blocking probabilities and then compared the performance of the dynamic frequency assignment system with the fixed frequency assignment scheme. The comparison showed that for a cellular system with N=7 and a given blocking probability, the dynamic frequency assignment scheme out-performs the fixed frequency assignment scheme in terms of the amount of load carried per frequency for low levels of traffic while the reverse is true for high levels of traffic.

A.4 Hybrid Frequency Assignment Schemes

These assignment schemes attempt to take advantage of both the fixed and dynamic schemes. They operate by dedicating a number of frequencies for fixed and the remaining for dynamic assignment. Cox and Reudink were among the earliest to introduce the concept of hybrid assignment. They described an improved dynamic channel reassignment scheme in [COX73]. The scheme was studied using computer simulations and the results showed that the hybrid arrangement performs better than either the fixed or the dynamic schemes for the range of offered traffic considered. Kahwa and Georganas [KAHW78] further studied the hybrid scheme by using simulation techniques and compared the performance of the systems with different ratios of the number of fixed to dynamic channels over a wide range of traffic load increases. In these studies, mobile telephone calls (blocked calls lost) and uniform average traffic among cells were assumed. In a later study [SIN81] by Sin and Georganas, similar system comparisons with dispatch calls (blocked calls delayed) and uniform average of traffic among cells were performed. The results obtained from the two studies are very similar. Both reports indicated that for small amount of traffic increases of up to 15%, systems with most dynamic channels gave the lowest probability of blocking or smallest average number of queued calls. For load increases between 15% and 35%, a medium number of dynamic channels performed better and for load increases beyond 50%, the performance is best for the fixed assignment scheme.

The above analysis was based on the assumption of uniform traffic in the cells. In the case of non-uniform traffic, Nehme and Georganas in [NEHM81] reported that the Hybrid Channel Assignment Scheme is also superior in terms of both traffic carried and spectral efficiency as compared to the Fixed Scheme for a given blocking probability and with the centre cell having a greater traffic demand than the surrounding cells.

A.5 Other Improved Frequency Assignment Schemes

The Hybrid Assignment Scheme was further improved by a number of workers. Singh in [SING81] very briefly described an adaptive channel assignment scheme which outperforms the fixed, dynamic and hybrid schemes. This new scheme constantly tries to adapt to the current traffic and channels can be borrowed from neighbouring cells if the demand exceeds the expected mean traffic in one cell. The scheme relies on the capability of the system to determine quickly and efficiently the traffic intensity of the arrival process which is assumed to be varying slowly with time.

In [ELNO82], Elnoubi et al described a technique which assigns priorities to the channels. The last ordered channels would be given priority to be borrowed by neighbouring cells while the first ordered channels would be given priority to be assigned to nominal calls. Hence switching of a nominal call from a low ordered channel to a high ordered channel is necessary if a high ordered channel somehow becomes available. It was shown that this scheme out-performs the fixed and the hybrid schemes.

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