Mobile Cellular Telecommunications Systems

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Preface

As the number of cellular subscribers increases, the interference that will be experienced by the systems will also increase. This means that many large cellular systems will, sooner or later, have to handle interference problems. This is a lucrative field that is ripe for research and that will soon be begging for more advanced applications.

This all-inclusive and self-contained work, consisting of fifteen chapters, is a basic textbook that supports further exploration in a new communications field, cellular communications. Since it is the first in its field, this book may be considered a handbook or building block for future research.

For years it has been my desire to write a book on the technical aspects of cellular systems. Since it is a new field, the theory has to be developed and then verified by experiment. I am seeking to adhere to the progression of learning that I described in *Who's Who in America:*

1. Use mathematics to solve problems.
2. Use physics to interpret results.
3. Use experiments and counterexamples to check outcomes.
4. Use pictures to emphasize important points.

Since I have accumulated many pictures in my mind, I would like to share them with my readers. In this field many new applications and theories have been discovered. Thus, my findings will help the reader to assimilate this new knowledge and accelerate learning time. The many mistakes that have been made in the past in designing cellular systems can now be avoided. Engineers who work in other communication systems will appreciate the many diverse concepts used in cellular systems. The reader should be aware that it is possible to apply the various theories improperly and thereby create many serious problems. I would like to hear from readers about their cellular systems experiences, both successful and unsuccessful.

Overall, I have written this book for technical engineers who would like to explore options in the cellular industry. However, Chapters 1 and 2 are for *executives* and for anyone who would like to familiarize himself or herself with key concepts of the field. Chapter 3 describes the specification of cellular systems. The North American specification works in Canada, the United States, and Mexico, so a cellular phone will work anywhere in this territory because of the standardized specification.

Chapter 4 introduces the point-to-point model I developed over the last 15 years. It can be used as a core to develop many design tools. Chapters 5 and 6 deal with cochannel interference problems, and Chapter 7 deals with noncochannel interference problems. Chapters 8 through 13 offer detailed material for engineers to solve problems concerning improved system performance. Chapter 14 describes the digital systems which may become the next-generation cellular systems, and presents many key issues in order to alert readers to possible future developments. Chapter 15 highlights some miscellaneous topics related to cellular systems.

Much of my unpublished work is included in these 15 chapters. I welcome feedback from readers about how I can better meet their needs in the second edition of the book.

I have always felt that cellular technology should be openly shared by cellular operators. Competition only occurs in a saturated market, and the cellular market is almost unlimited. Therefore, competition is not to be feared in this early stage of the cellular industry. We need to promote this industry as much as possible by involving more interested engineers and investors.

In the last six years, I have taught 3-day seminars sponsored by George Washington University. I am trying to convince the cellular industry that if we have narrow-minded attitudes and do not share our experiments or knowledge, the whole industry will not advance fast enough and could be replaced by other new industries, such as wireless communications or in-building communications.

Let us join together to allow the cellular industry its optimum potential and set our goal that one day a pocket cellular phone will carry our calls to any place in the world.

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Irvine, California

Chapter

1

Introduction to Cellular Mobile Systems

1.1 Why Cellular Mobile Telephone Systems?

1.1.1 Limitations of conventional mobile telephone systems

One of many reasons for developing a *cellular* mobile telephone system and deploying it in many cities is the operational limitations of conventional mobile telephone systems: limited service capability, poor service performance, and inefficient frequency spectrum utilization.

Limited service capability. A conventional mobile telephone system is usually designed by selecting one or more channels from a specific frequency allocation for use in autonomous geographic zones, as shown in Fig. 1.1. The communications coverage area of each zone is normally planned to be as large as possible, which means that the transmitted power should be as high as the federal specification allows. The user who starts a call in one zone has to reinitiate the call when moving into a new zone (see Fig. 1.1) because the call will be dropped. This is an undesirable radio telephone system since there is no guarantee that a call can be completed without a handoff capability.

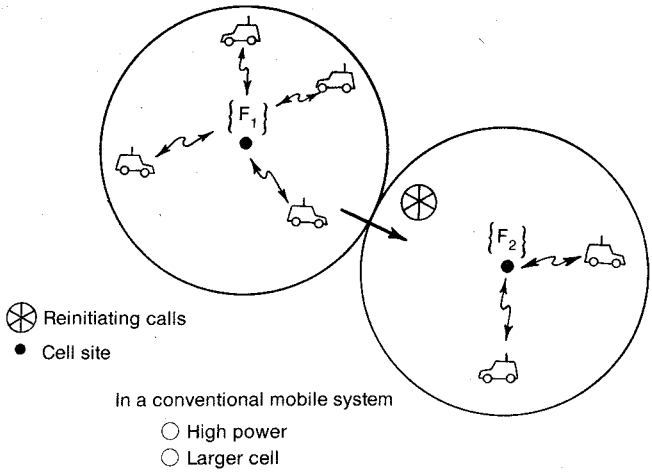


Figure 1.1 Conventional mobile system.

The handoff is a process of automatically changing frequencies as the mobile unit moves into a different frequency zone so that the conversation can be continued in a new frequency zone without redialing. Another disadvantage of the conventional system is that the number of active users is limited to the number of channels assigned to a particular frequency zone.

Poor service performance. In the past, a total of 33 channels were allocated to three mobile telephone systems: Mobile Telephone Service (MTS), Improved Mobile Telephone Service (IMTS) MJ systems, and Improved Mobile Telephone Service (IMTS) MK systems. MTS operates around 40 MHz and MJ operates at 150 MHz; both provide 11 channels; IMTS MK operates at 450 MHz and provides 12 channels. These 33 channels must cover an area 50 mi in diameter. In 1976, New York City had 6 channels of MJ serving 320 customers, with another 2400 customers on a waiting list. New York City also had 6 channels of MK serving 225 customers, with another 1300 customers on a waiting list. The large number of subscribers created a high blocking probability during busy hours. The actual number of blockings will be shown later. Although service performance was undesirable, the demand was still great. A high-capacity system for mobile telephones was needed.

Inefficient frequency spectrum utilization. In a conventional mobile telephone system, the frequency utilization measurement M_0 is defined as the maximum number of customers that could be served by one channel

at the busy hour. Equation (1.1-1) gives the 1976 New York City data cited earlier.

$$M_0 = \frac{\text{no. of customers}}{\text{channel}} \qquad \text{(conventional systems)} \qquad (1.1\text{-}1)$$

or

$$M_0 = \begin{cases} 53 \text{ customers/channel} & \text{(MJ system)} \\ 37 \text{ customers/channel} & \text{(MK system)} \end{cases}$$

Assume an average calling time of 1.76 min and apply the Erlang B model (lost-calls-cleared conditions). Calculate the blocking probability as follows: Use 6 channels, with each channel serving the two different numbers of customers shown in Eq. (1.1-1). The offered load can then be obtained by Eq. (1.1-2).

$$A = \frac{\text{av calling time (minutes)} \times \text{total customers}}{60 \text{ min}} \qquad \text{erlangs} \qquad (1.1\text{-}2)$$

$$A_1 = \frac{1.76 \times 53 \times 6}{60} = 9.33 \text{ erlangs} \qquad \text{(MJ system)}$$

$$A_2 = \frac{1.76 \times 37 \times 6}{60} = 6.51 \text{ erlangs} \qquad \text{(MK system)}$$

Given that the number of channels is 6 and the offered loads are $A_1 = 9.33$ and $A_2 = 6.51$, read from the table in Appendix 1.1 to obtain the blocking probabilities $B_1 = 50$ percent (MJ system) and $B_2 = 30$ percent (MK system), respectively. It is likely that half the initiating calls will be blocked in the MJ system, a very high blocking probability.

If the actual average calling time is greater than 1.76 min, the blocking probability can be even higher. To reduce the blocking probability, we must decrease the value of the frequency spectrum utilization measurement M_0 as shown in Eq. (1.1-1).

As far as frequency spectrum utilization is concerned, the conventional system does not utilize the spectrum efficiently since each channel can only serve one customer at a time in a whole area. A new cellular system that measures the frequency spectrum utilization differently from Eq. (1.1-1) and proves to be efficient is discussed in Sec. 1.1.2.

1.1.2 Spectrum efficiency considerations

A major problem facing the radio communication industry is the limitation of the available radio frequency spectrum. In setting allocation

policy, the Federal Communications Commission (FCC) seeks systems which need minimal bandwidth but provide high usage and consumer satisfaction.

The ideal mobile telephone system would operate within a limited assigned frequency band and would serve an almost unlimited number of users in unlimited areas. Three major approaches to achieve the ideal are

1. Single-sideband (SSB), which divides the allocated frequency band into maximum numbers of channels
2. Cellular, which reuses the allocated frequency band in different geographic locations
3. Spread spectrum, frequency-hopped, which generates many codes over a wide frequency band

In 1971, the cellular approach was shown to be a spectrally efficient system.[1] The comparison of a cellular approach with other approaches is given in Chap. 12.

1.1.3 Technology, feasibility, and service affordability

In 1971, the computer industry entered a new era. Microprocessors and minicomputers are now used for controlling many complicated features and functions with less power and size than was previously possible. Large-scale integrated (LSI) circuit technology reduced the size of mobile transceivers so that they easily fit into the standard automobile. These achievements were a few of the requirements for developing advanced mobile phone systems and encouraging engineers to pursue this direction.

Another factor was the price reduction of the mobile telephone unit. LSI technology and mass production contribute to reduced cost so that in the near future an average-income family should be able to afford a mobile telephone unit.

On Jan. 4, 1979, the FCC authorized Illinois Bell Telephone Co. (IBT) to conduct a developmental cellular system in the Chicago area and make a limited commercial offering of its cellular service to the public. In addition, American Radio Telephone Service, Inc., (ARTS) was authorized to operate a cellular system in the Washington, D.C.–Baltimore, Md., area. These first systems showed the technological feasibility and affordability of cellular service.

1.1.4 Why 800 MHz?

The FCC's decision to choose 800 MHz was made because of severe spectrum limitations at lower frequency bands. FM broadcasting ser-

vices operate in the vicinity of 100 MHz. The television broadcasting service starts at 41 MHz and extends up to 960 MHz.

Air-to-ground systems use 118 to 136 MHz; military aircraft use 225 to 400 MHz. The maritime mobile service is located in the vicinity of 160 MHz. Also fixed-station services are allocated portions of the 30- to 100-MHz band. Therefore, it was hard for the FCC to allocate a spectrum in the lower portions of the 30- to 400-MHz band since the services of this band had become so crowded. On the other hand, mobile radio transmission cannot be applied at 10 GHz or above because severe propagation path loss, multipath fading, and rain activity make the medium improper for mobile communications.

Fortunately, 800 MHz was originally assigned to educational TV channels. Cable TV service became a big factor in the mid-70s and shared the load of providing TV channels. This situation opened up the 800-MHz band to some extent, and the FCC allocated a 40-MHz system at 800 MHz to mobile radio cellular systems.

Although 800 MHz is not the ideal transmission medium for mobile radio, it has been demonstrated that a cellular mobile radio system[2,3] that does not go beyond this frequency band can be deployed. Needless to say, the medium of transmitting an 800-MHz signal, although it is workable, is already very difficult. Section 1.6.1 briefly describes the transmission medium.

1.2 History of 800-MHz Spectrum Allocation

In 1958, the Bell System (FCC Docket 11997) proposed a 75-MHz system at 800 MHz, quite a broadband proposal. In 1970, the FCC (Docket 18262)[13] tentatively decided to allocate 75 MHz for a wire-line common carrier. In December 1971 the Bell System assured technical feasibility by showing how a cellular mobile system could be designed.[1] In 1974, the FCC allocated 40 MHz of the spectrum, with one cellular system to be licensed per market. There was considerable uncertainty in predicting the cellular market. However, the FCC strategically placed spectrum reserves totaling 20 MHz in proximity to the cellular allocation.

In 1980, the FCC reconsidered its one-system-per-market strategy and studied the possibility of introducing competition into the previous one-carrier markets. Although cost savings make one cellular system per market attractive, balancing the benefits of economies of scale against the benefits of competition, two licensed carriers per service area was more in line with emerging FCC policies.

Trunking efficiency degradation using two carriers per service area will be discussed in Sec. 1.3. It was the FCC's view that such an approach, while not gaining the full competitive market structure, would

TABLE 1.1 Mobile and Base Transmission Frequencies*

Band†	Mobile	Base	Two systems/market
A	824–835, 845–846.5	869–880, 890–891.5	Non-wire-line‡
B	835–845, 846.5–849	880–890, 891.5–894	Wire-line

* On July 24, 1986, an additional 5 MHz was allocated to each band. Therefore, an additional 83 channels were added to each band. The system accommodating an additional 83 channels may be ready in mid-1988.
† 416 channels per band, 30 kHz per channel.
‡ Majority are non-wire-line companies.

provide some competitive advantages. The frequencies will be assigned in 20-MHz groups identified as block A and block B, or called band A and band B.

Two bands serve two different groups in the standard situation: one for wire-line (telephone) companies* and one for non-wire-line (non-telephone) companies. Each company designs its own system and divides the area into geographic areas, or cells. Each cell operates within its own bands (see Table 1.1).

Since 30 kHz is the specified bandwidth, each band operating nowadays consists of 333 channels.† How to utilize these limited resources to provide adequate voice quality and service performance to an unrestricted population size presents a challenge.

1.3 Trunking Efficiency

To explore the trunking efficiency degradation inherent in licensing two or more carriers rather than one, compare the trunking efficiency between one cellular system per market operating 666 channels and two cellular systems per market each operating 333 channels. Assume that all frequency channels are evenly divided into seven subareas called *cells*. In each cell, the blocking probability of 0.02 is assumed. Also the average calling time is assumed to be 1.76 min.

Look up the table of Appendix 1.1 with $N_1 = {}^{666}/_{7} = 95$ and $B = 0.02$ to obtain the offered load $A_1 = 83.1$ and with $N_2 = {}^{333}/_{7} = 47.5$ and $B = 0.02$ to obtain $A_2 = 38$. Since two carriers each operating 333 channels are considered, the total offered load is $2A_2$. We then realize that

$$A_1 \geq 2A_2 \tag{1.3-1}$$

* Let telephone companies operate mobile radio telephone systems.

† Most analyses in this book are based on the present channel numbers of 333 channels per system.

By converting Eq. (1.3-1) to the number of users who can be served in a busy hour, the average calling time of 1.76 min is introduced. The number of calls per hour served in a cell can be expressed as

$$Q_i = \frac{A \times 60}{1.76} \text{ calls/h} \tag{1.3-2}$$

Then

$$Q_i = \begin{cases} 2832.95 \text{ calls/h} & \text{(1 carrier/market)} \\ 1295.45 \times 2 = 2590.9 \text{ calls/h} & \text{(2 carriers/market)} \end{cases}$$

The trunking efficiency degradation factor can be calculated as

$$\eta_e = \frac{2832.95 - 2590.9}{2832.95} = 8.5\% \tag{1.3-3}$$

for a blocking probability of 2 percent. Figure 1.2 shows η_e by comparing one carrier per market with more than one carrier per market situations with different blocking probability conditions. The degradation of trunking efficiency decreases as the blocking probability increases. As the number of carriers per market increases the degradation increases. However, when a high percentage of blocking probability, say more than 20 percent, occurs, the performance of one carrier per market is already so poor that further degradation becomes insignificant, as Fig. 1.2 shows.

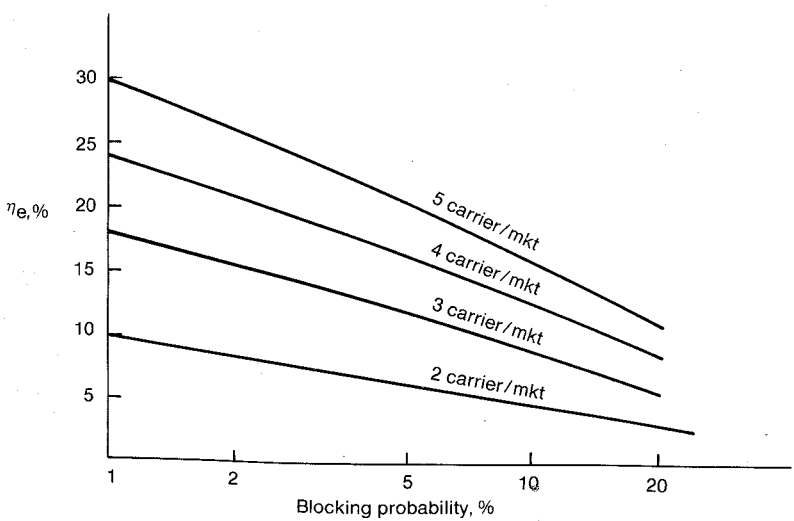


Figure 1.2 Degradation of trunking efficiency—comparing one carrier/market and other-than-one-carrier/market.

For a 2 percent blocking probability, the trunking efficiency of one carrier per market does show a greater advantage when compared to other scenarios.

1.4 A Basic Cellular System

A basic cellular system consists of three parts: a mobile unit, a cell site, and a mobile telephone switching office (MTSO), as Fig. 1.3 shows, with connections to link the three subsystems.

1. *Mobile units.* A mobile telephone unit contains a control unit, a transceiver, and an antenna system.
2. *Cell site.* The cell site provides interface between the MTSO and the mobile units. It has a control unit, radio cabinets, antennas, a power plant, and data terminals.
3. *MTSO.* The switching office, the central coordinating element for all cell sites, contains the cellular processor and cellular switch. It interfaces with telephone company zone offices, controls call processing, and handles billing activities.

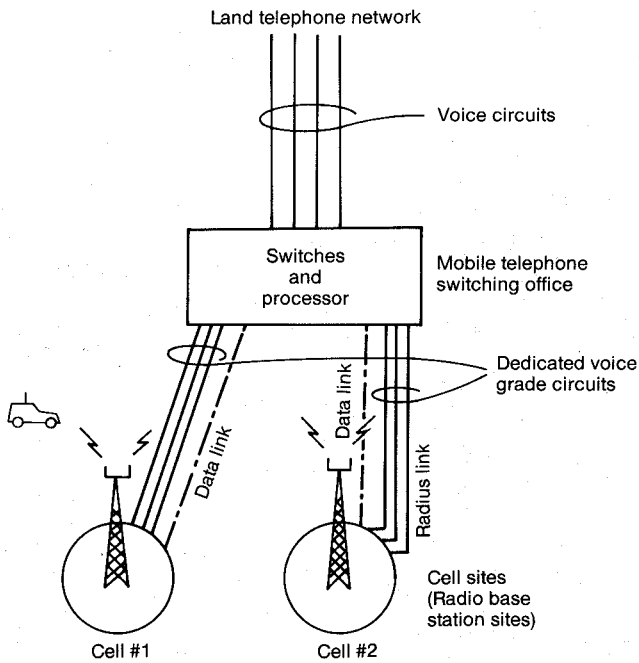


Figure 1.3 Cellular system.

4. *Connections.* The radio and high-speed data links connect the three subsystems. Each mobile unit can only use one channel at a time for its communication link. But the channel is not fixed; it can be any one in the entire band assigned by the serving area, with each site having multichannel capabilities that can connect simultaneously to many mobile units.

The MTSO is the heart of the cellular mobile system. Its processor provides central coordination and cellular administration.

The cellular switch, which can be either analog or digital, switches calls to connect mobile subscribers to other mobile subscribers and to the nationwide telephone network. It uses voice trunks similar to telephone company interoffice voice trunks. It also contains data links providing supervision links between the processor and the switch and between the cell sites and the processor. The radio link carries the voice and signaling between the mobile unit and the cell site. The high-speed data links cannot be transmitted over the standard telephone trunks and therefore must use either microwave links or T-carriers (wire lines). Microwave radio links or T-carriers carry both voice and data between the cell site and the MTSO.

1.5 Performance Criteria

There are three categories for specifying performance criteria.

1.5.1 Voice quality

Voice quality is very hard to judge without subjective tests from users' opinions. In this technical area engineers cannot decide how to build a system without knowing the voice quality that will satisfy the users. In military communications, the situation differs: armed forces personnel *must* use the assigned equipment.

For any given commercial communications system, the voice quality will be based upon the following criterion: a set value x at which y percent of customers rate the system voice quality (from transmitter to receiver) as good or excellent, the top two circuit merits of the five listed below.

CM5	excellent (speech perfectly understandable)
CM4	good (speech easily understandable, some noise)
CM3	fair (speech understandable with a slight effort, occasional repetitions needed)
CM2	poor (speech understandable only with considerable effort, frequent repetitions needed)
CM1	unusable (speech not understandable)

As the percentage of customers choosing CM4 and CM5 increases, the cost of building the system rises.

1.5.2 Service quality

Three items are required for service quality.

1. *Coverage.* The system should serve an area as large as possible. With radio coverage, however, because of irregular terrain configurations, it is usually not practical to cover 100 percent of the area for two reasons:
   a. The transmitted power would have to be very high to illuminate weak spots with sufficient reception, a significant added cost factor.
   b. The higher the transmitted power, the harder it becomes to control interference.

   Therefore, systems usually try to cover 90 percent of an area in flat terrain and 75 percent of an area in hilly terrain. The combined voice quality and coverage criteria in AMPS cellular systems[3] state that 75 percent of users rate the voice quality between good and excellent in 90 percent of the served area, which is generally flat terrain. The voice quality and coverage criteria would be adjusted as per decided various terrain conditions. In hilly terrain, 90 percent of users must rate voice quality good or excellent in 75 percent of the served area. A system operator can lower the percentage values stated above for a low-performance and low-cost system.
2. *Required grade of service.* For a normal start-up system the grade of service is specified for a blocking probability of .02 for initiating calls at the busy hour. This is an average value. However, the blocking probability at each cell site will be different. At the busy hour, near freeways, automobile traffic is usually heavy, so the blocking probability at certain cell sites may be higher than 2 percent, especially when car accidents occur. To decrease the blocking probability requires a good system plan and a sufficient number of radio channels.
3. *Number of dropped calls.* During Q calls in an hour, if a call is dropped and $Q - 1$ calls are completed, then the call drop rate is $1/Q$. This drop rate must be kept low. A high drop rate could be caused by either coverage problems or handoff problems related to inadequate channel availability.

1.5.3 Special features

A system would like to provide as many special features as possible, such as call forwarding, call waiting, voice stored (VSR) box, automatic

roaming, or navigation services. However, sometimes the customers may not be willing to pay extra charges for these special services.

1.6 Uniqueness of Mobile Radio Environment

1.6.1 Description of mobile radio transmission medium

The propagation attenuation. In general, the propagation path loss increases not only with frequency but also with distance. If the antenna height at the cell site is 30 to 100 m and at the mobile unit about 3 m, and the distance between the cell site and the mobile unit is usually 2 km or more, then the incident angles of both the direct wave and the reflected wave are very small, as Fig. 1.4 shows. The incident angle of the direct wave is θ_1, and the incident angle of the reflected wave is θ_2. θ_1 is also called the *elevation angle*. The propagation path loss would be 40 dB/dec,[4] where "dec" is an abbreviation of *decade,* i.e., a period of 10. This means that a 40-dB loss at a signal receiver will be observed by the mobile unit as it moves from 1 to 10 km. Therefore C is inversely proportional to R^4.

$$C \propto R^{-4} = \alpha R^4 \tag{1.6-1}$$

where C = received carrier power
R = distance measured from the transmitter to the receiver
α = constant

The difference in power reception at two different distances R_1 and R_2 will result in

$$\frac{C_2}{C_1} = \left(\frac{R_2}{R_1}\right)^{-4} \tag{1.6-2a}$$

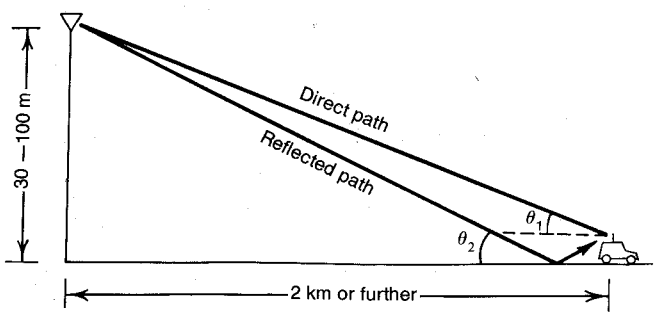


Figure 1.4 Mobile radio transmission model.

and the decibel expression of Eq. (1.6-2) is

$$\Delta C \text{ (in dB)} = C_2 - C_1 \text{ (in dB)}$$

$$= 10 \log \frac{C_2}{C_1} = 40 \log \frac{R_1}{R_2} \qquad (1.6\text{-}2b)$$

When $R_2 = 2R_1$, $\Delta C = -12$ dB; when $R_2 = 10R_1$, $\Delta C = -40$ dB.

This 40 dB/dec is the general rule for the mobile radio environment and is easy to remember. It is also easy to compare to the free-space propagation rule of 20 dB/dec. The linear and decibel scale expressions are

$$C \propto R^{-2} \qquad \text{(free space)} \qquad (1.6\text{-}3a)$$

and

$$\Delta C = C_2 \text{ (in dB)} - C_1 \text{ (in dB)}$$

$$= 20 \log \frac{R_1}{R_2} \qquad \text{(free space)} \qquad (1.6\text{-}3b)$$

In a real mobile radio environment, the propagation path-loss slope varies as

$$C \propto R^{-\gamma} = \alpha R^{-\gamma} \qquad (1.6\text{-}4)$$

γ usually lies between 2 and 5 depending on the actual conditions.[5] Of course γ cannot be lower than 2, which is the free-space condition. The decibel scale expression of Eq. (1.6-4) is

$$C = 10 \log \alpha - 10\gamma \log R \qquad \text{dB} \qquad (1.6\text{-}5)$$

Severe fading. Since the antenna height of the mobile unit is lower than its typical surroundings, and the carrier frequency wavelength is much less than the sizes of the surrounding structures, multipath waves are generated. At the mobile unit, the sum of the multipath waves causes a signal-fading phenomenon. The signal fluctuates in a range of about 40 dB (10 dB above and 30 dB below the average signal). We can visualize the nulls of the fluctuation at the baseband at about every half wavelength in space, but all nulls do not occur at the same level, as Fig. 1.5 shows. If the mobile unit moves fast, the rate of fluctuation is fast. For instance, at 850 MHz, the wavelength is roughly 0.35 m (1 ft). If the speed of the mobile unit is 24 km/h (15 mi/h), or 6.7 m/s, the rate of fluctuation of the signal reception at a 10-dB level below the average power of a fading signal is 15 nulls per second (see Sec. 1.6.3).

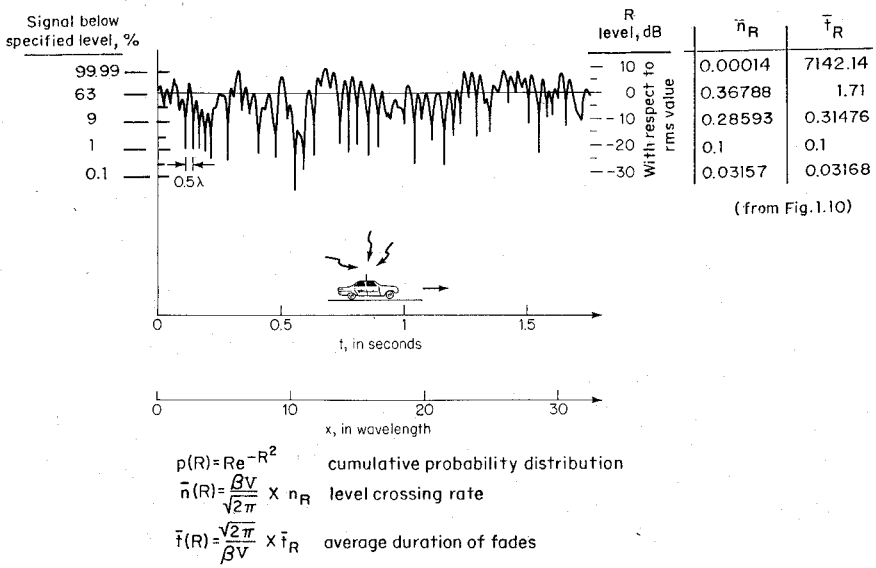


Figure 1.5 A typical fading signal received while the mobile unit is moving. (*Reprint after Lee, Ref. 4, p. 46.*)

1.6.2 Model of transmission medium

A mobile radio signal $r(t)$, illustrated in Fig. 1.6, can be artificially characterized by two components $m(t)$ and $r_0(t)$ based on natural physical phenomena.

$$r(t) = m(t)r_0(t) \tag{1.6-6}$$

The component $m(t)$ is called *local mean, long-term fading,* or *log-normal fading* and its variation is due to the terrain contour between the base station and the mobile unit. The factor r_0 is called *multipath fading, short-term fading,* or *Rayleigh fading* and its variation is due to the waves reflected from the surrounding buildings and other structures. The long-term fading $m(t)$ can be obtained from Eq. (1.6-7a).

$$m(t_1) = \frac{1}{2T}\int_{t_1-T}^{t_1+T} r(t)\,dt \tag{1.6-7a}$$

where $2T$ is the time interval for averaging $r(t)$. T can be determined based on the fading rate of $r(t)$, usually 40 to 80 fades.[5] Therefore, $m(t)$ is the envelope of $r(t)$, as shown in Fig. 1.6a. Equation (1.6-7a) also can be expressed in spatial scale as

$$m(x_1) = \frac{1}{2L}\int_{x_1-L}^{x_1+L} r(x)\,dx \tag{1.6-7b}$$

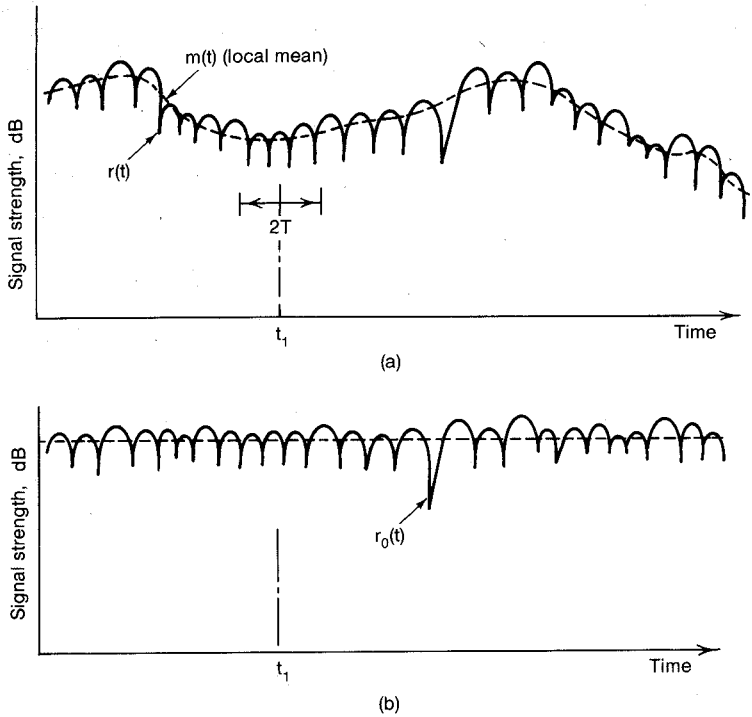


Figure 1.6 A mobile radio signal fading representation. (*a*) A mobile signal fading. (*b*) A short-term signal fading.

The length of $2L$ has been determined to be 20 to 40 wavelengths.[5] Using 36 or up to 50 samples in an interval of 40 wavelengths is an adequate averaging process for obtaining the local means.[4]

The factor $m(t)$ or $m(x)$ is also found to be a log-normal distribution based on its characteristics caused by the terrain contour. The short-term fading r_0 is obtained by

$$r_0 \text{ (in dB)} = r(t) - m(t) \quad \text{dB} \tag{1.6-8}$$

as shown in Fig. 1.6*b*. The factor $r_0(t)$ follows a Rayleigh distribution, assuming that only reflected waves from local surroundings are the ones received (a normal situation for the mobile radio environment). Therefore, the term Rayleigh fading is often used.

1.6.3 Mobile fading characteristics

Rayleigh fading is also called multipath fading in the mobile radio environment. When these multipath waves bounce back and forth due to the buildings and houses, they form many standing-wave pairs in

space, as shown in Fig. 1.7. Those standing-wave pairs are summed together and become an irregular wave-fading structure. When a mobile unit is standing still, its receiver only receives a signal strength at that spot, so a constant signal is observed. When the mobile unit is moving, the fading structure of the wave in the space is received. It is a multipath fading. The recorded fading becomes fast as the vehicle moves faster.

The radius of the active scatterer region. The mobile radio multipath fading shown in Fig. 1.7 explains the fading mechanism. The radius of the active scatterer region at 850 MHz can be obtained indirectly as shown in Ref. 12. The radius is roughly 100 wavelengths. The active scatterer region always moves with the mobile unit as its center. It means that some houses were inactive scatterers and became active as the mobile unit approached them; some houses were active scatterers and became inactive as the mobile unit drove away from them.

Standing waves expressed in a linear scale and a log scale. We first introduce a sine wave in a log scale.

$$y = 10 \cos \beta x \qquad \text{dB} \tag{1.6-9}$$

A log plot of the sine wave of Eq. (1.6-9) is shown in Fig. 1.8*a*. The linear expression of Eq. (1.6-9) then is shown in Fig. 1.8*b*. The sym-

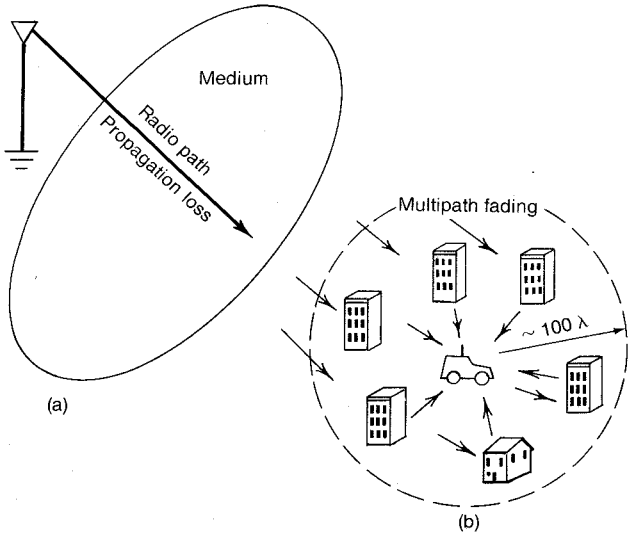


Figure 1.7 A mobile radio environment—two parts. (1) Propagation loss; (2) multipath fading.

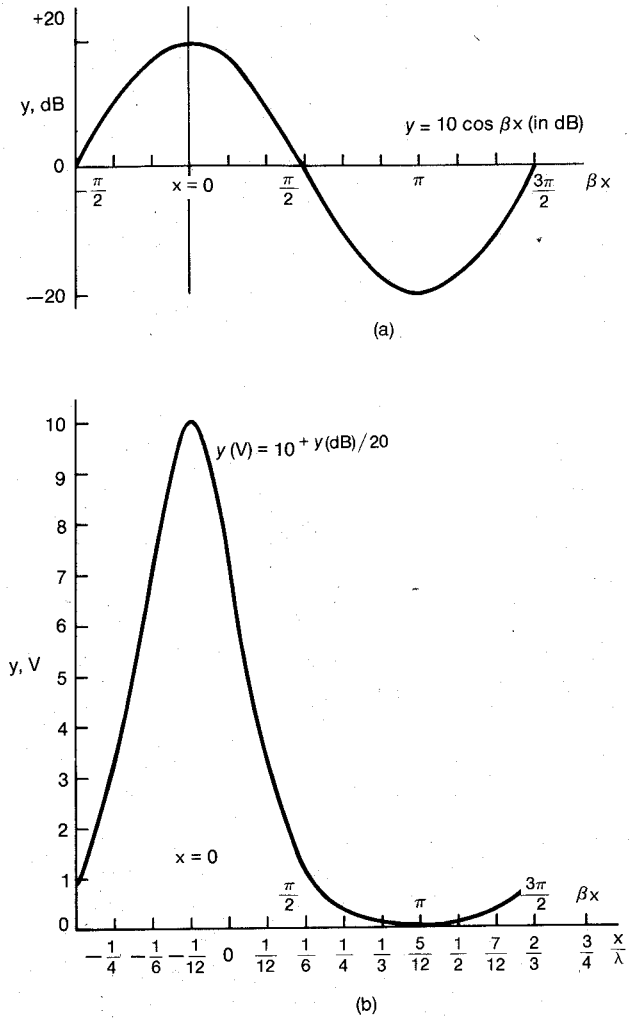


Figure 1.8 The linear plot and the log plot of a sine wave. (*a*) In linear scale; (*b*) in log scale.

metrical waveform in a log plot becomes an unsymmetrical waveform when plotted on a linear scale. It shows that the sine wave waveform in a log scale becomes a completely different waveform when expressed on a linear scale and vice versa. Two sine waves, the incident wave traveling along the x-axis (traveling to the left) and the reflected wave traveling in the opposite direction, can be expressed as

$$e_0 = E_0 e^{j(\omega t + \beta x)} \tag{1.6-10}$$

and
$$e_1 = E_1 e^{j(\omega t - \beta x + \delta)} \tag{1.6-11}$$

where ω = angular frequency
β = wave number ($= 2\pi/\lambda$)
δ = time-phase lead of e_1 with respect to e_0 at $x = 0$

The two waves form a standing-wave pattern.

$$e = e_0 + e_1 = R \cos(\omega t - \delta) \tag{1.6-12}$$

where the amplitude R becomes

$$R = \sqrt{(E_0 + E_1)^2 \cos^2 \beta x + (E_0 - E_1)^2 \sin^2 \beta x} \tag{1.6-13}$$

We are plotting two cases.

Case 1. $E_0 = 1, E_1 = 1$; that is, the reflection coefficient $= 1$,

$$\text{Standing wave ratio (SWR)} = \frac{E_0 + E_1}{E_0 - E_1} = \infty$$

and
$$R = 2 \cos \beta x \tag{1.6-14}$$

Case 2. $E_0 = 1, E_1 = 0.5$; that is, the reflection coefficient $= 0.5$, SWR $= 3$, and

$$R = \sqrt{(1.5)^2 \cos^2 \beta x + (0.5)^2 \sin^2 \beta x} \tag{1.6-15}$$

The linear expression of Eqs. (1.6-14) and (1.6-15) are shown in Fig. 1.9*a*. The log-scale expression of Eqs. (1.6-14) and (1.6-15) are shown in Fig. 1.9*b*. The waveform of Fig. 1.9*b* is the first sign of the fading signal which resembles the real fading signal shown in Fig. 1.5.

First-order and second-order statistics of fading. Fading occurs on the signal reception when the mobile unit is moving. The first-order statistics, such as average power probability cumulative distribution function (CDF) and bit error rate, are independent of time. The second-order statistics, such as level crossing rate, average duration of fades, and word error rate, are time functions or velocity-related functions. The data signaling format is based on these characteristics. The description of the fading characteristic can be found in detail in two books, Refs. 4 and 5.

Some data can be found from Fig. 1.10*a*, the cumulative distribution function (CDF), and Fig. 1.10*b*, the level crossing rate. In Fig. 1.10*a*, the equation of CDF for a Rayleigh fading is used as follows:

$$P(x \le A) = 1 - e^{-A^2/\overline{A^2}} \tag{1.6-16}$$

and
$$P(y \le L) = 1 - e^{-L/\overline{L}} \tag{1.6-17}$$

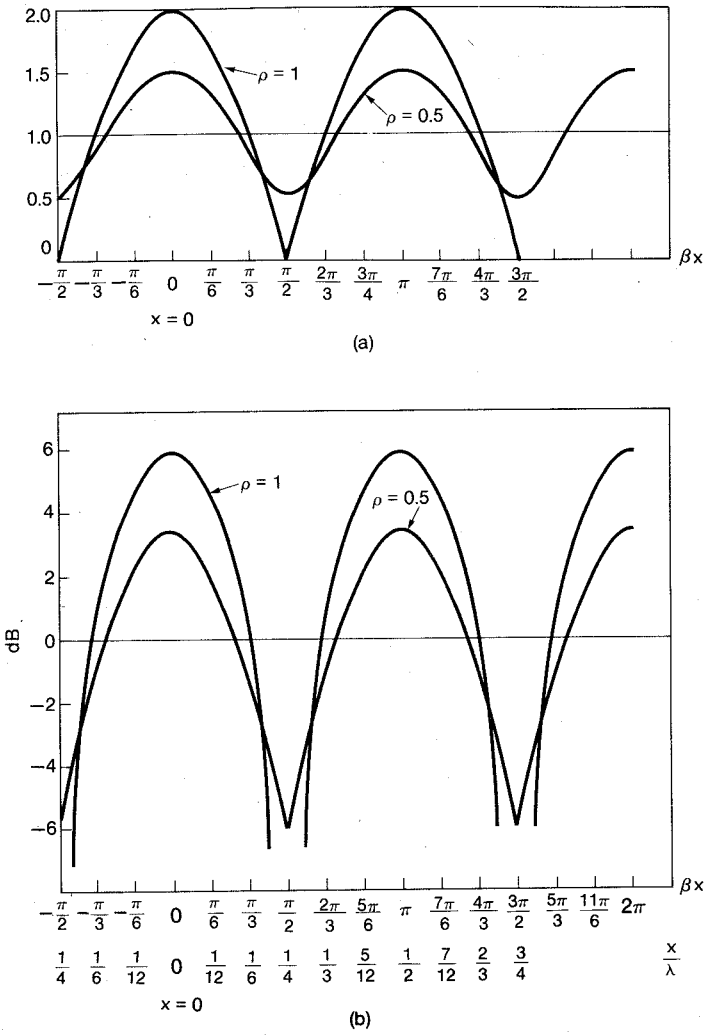


Figure 1.9 The linear plot and the log plot of a standing wave. (*a*) In linear scale; (*b*) in log scale.

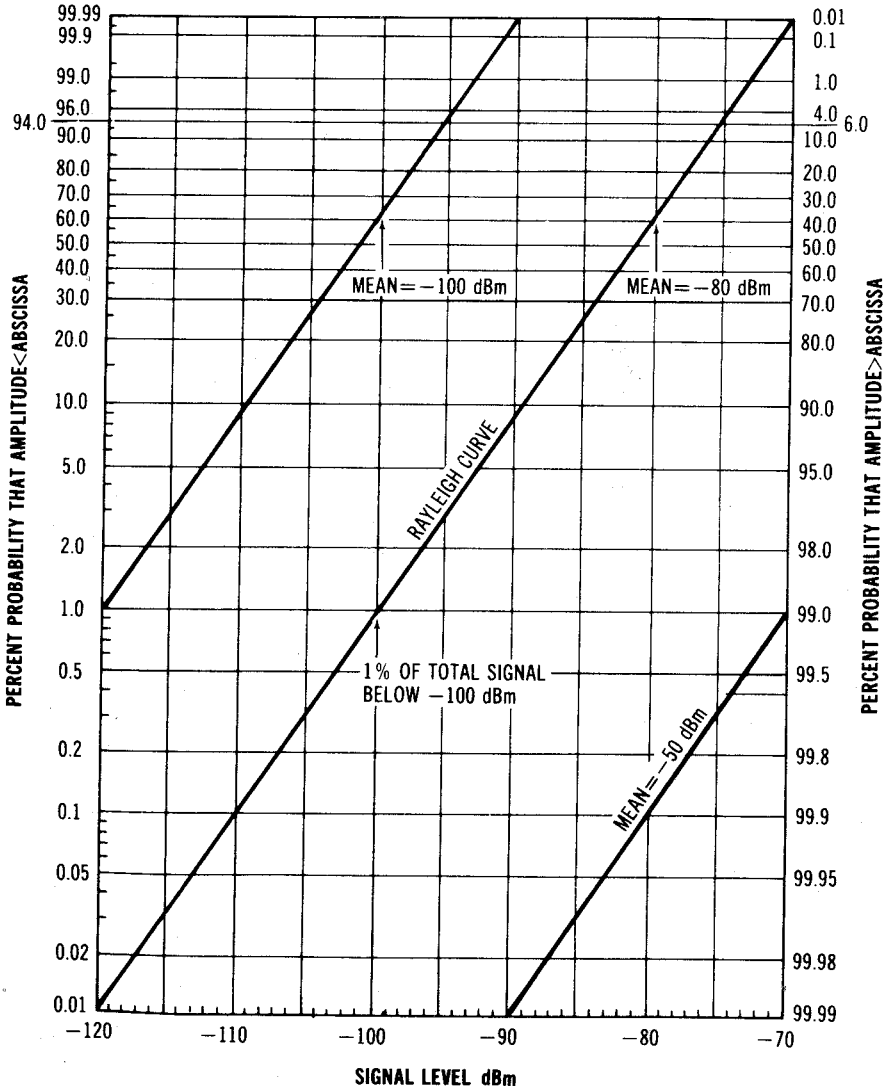
where $\overline{A^2}$ and $\overline{L}$ are the mean square value and the average power, respectively. In Fig. 1.10*a*, about 9 percent of the total signal is below a level of -10 dB with respect to average power. In Fig. 1.10*b*, the level crossing rate (lcr) at a level A is

$$\overline{n}(A) = \frac{\beta v}{\sqrt{2\pi}} n_R$$

where n_R is the normalized lcr which is independent of wavelength

and the car speed. At a level of -10 dB, $n_0 = 0.3$ can be found from Fig. 1.10*b*. Assume that a signal of 850 MHz is received at a mobile unit with a velocity of 24 km/h (15 mi/h). Then

$$n_0 = \frac{\beta V}{\sqrt{2\pi}} \approx 50$$

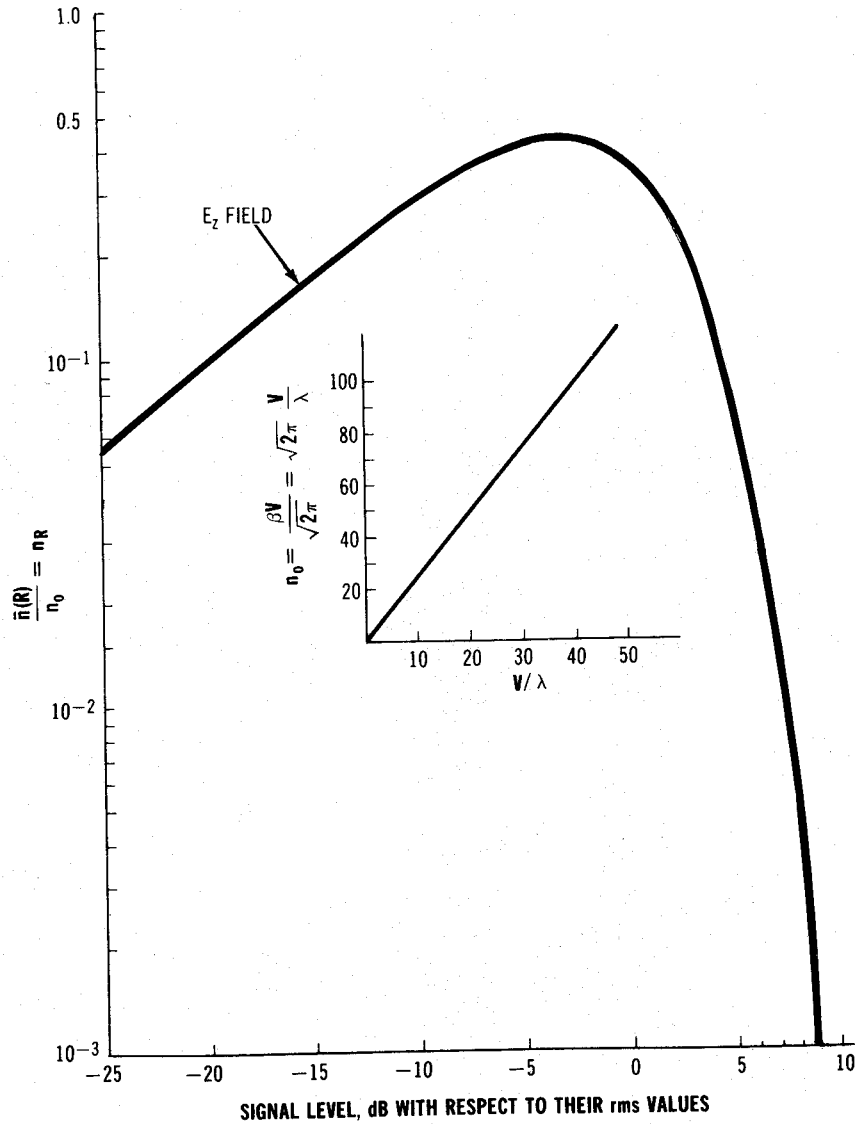


(a)

Figure 1.10 Fading characteristics. (*a*) CDF. (*After Lee, Ref. 5, p. 33.*) (*b*) Level crossing rate. (*c*) Average duration of fades.

and $$\overline{n} = 50 \times 0.3 = 15$$

Therefore, at a cellular frequency of 800 MHz and a vehicle velocity of 15 mi/h, the level crossing rate is 15 per second. It is easy to remember.



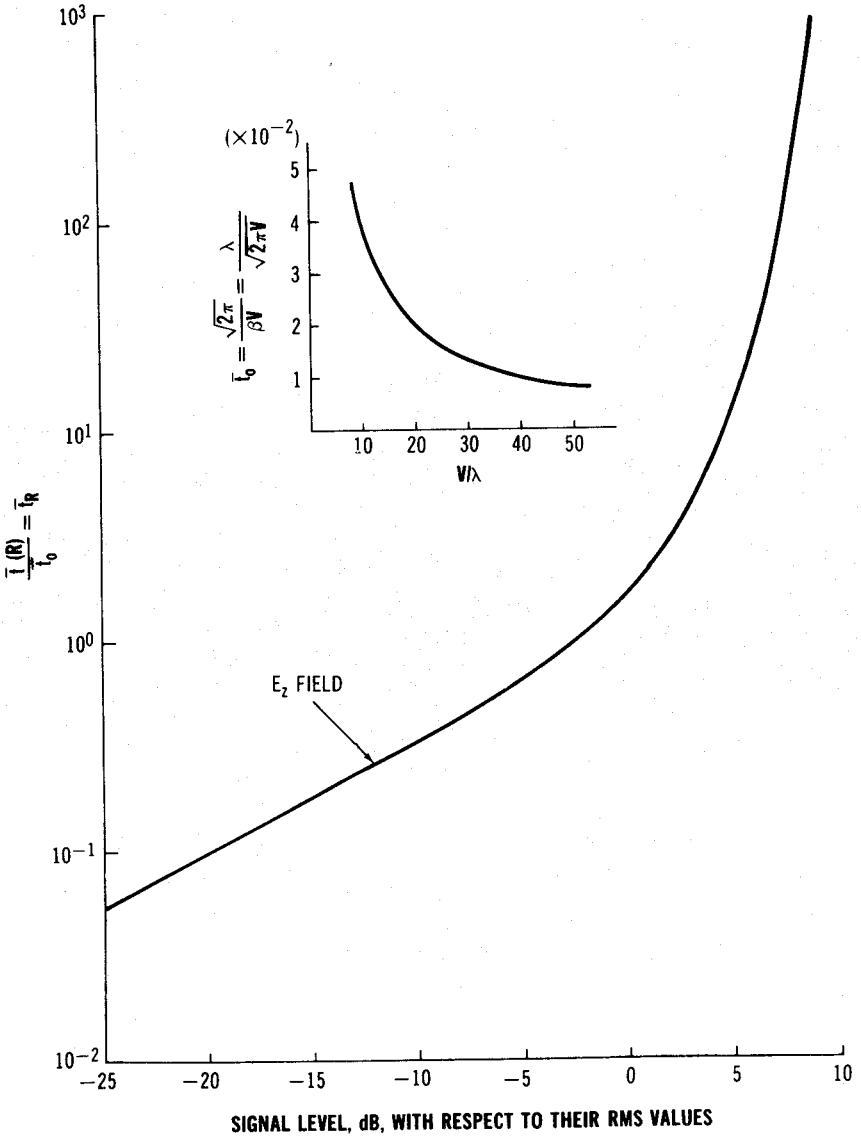
(b)

Figure 1.10 *(Continued)*

The average duration of fade is

$$\bar{t} = \frac{\text{CDF}}{\bar{t}_0 \bar{t}_R} = \frac{\text{CDF}}{\bar{n}} \tag{1.6-19}$$

Equation (1.6-19) is plotted in Fig. 1.10c, where t_0 and t_R are also shown.



(c)

Figure 1.10 (*Continued*)

At -10 dB, the average duration of fades is

$$\bar{t} = \frac{\text{CDF}}{\bar{n}} = 0.0066 \text{ s} = 6.6 \text{ ms}$$

Now the average power level plays an important role in determining the statistics. Therefore, it should be specified by the system design. The second-order statistic of fading phenomenon is most useful for designing a signaling format for the cellular system. As soon as the signaling format is specified, we can calculate the bit error rate and the word error rate and find ways to reduce the error rates, which will be described in Sec. 13.2.

Delay spread and coherence bandwidth

Delay spread. In the mobile radio environment, as a result of the multipath reflection phenomenon, the signal transmitted from a cell site and arriving at a mobile unit will be from different paths, and since each path has a different path length, the time of arrival for each path is different. For an impulse transmitted at the cell site, by the time this impulse is received at the mobile unit it is no longer an impulse but rather a pulse with a spread width that we call the *delay spread*. The measured data indicate that the mean delay spreads are different in different kinds of environment.

Type of environment	Delay spread Δ, μs
Open area	<0.2
Suburban area	0.5
Urban area	3

Coherence bandwidth. The coherence bandwidth is the defined bandwidth in which either the amplitudes or the phases of two received signals have a high degree of similarity. The delay spread is a natural phenomenon, and the coherence bandwidth is a defined creation related to the delay spread.

A coherence bandwidth for two fading amplitudes of two received signals is

$$B_c = \frac{1}{2\pi\Delta}$$

A coherence bandwidth for two random phases of two received signals is

$$B_c' = \frac{1}{4\pi\Delta}$$

1.6.4 Direct wave path, line-of-sight path, and obstructive path

A *direct wave path* is a path clear from the terrain contour. The *line-of-sight path* is a path clear from buildings. In the mobile radio environment, we do not always have a line-of-sight condition.

When a line-of-sight condition occurs, the average received signal at the mobile unit at a 1-mi intercept is higher, although the 40 dB/dec path-loss slope remains the same. It will be described in Sec. 4.2. In this case the short-term fading is observed to be a rician fading.[14] It results from a strong line-of-sight path and a ground-reflected wave combined, plus many weak building-reflected waves.

When an out-of-sight condition is reached, the 40-dB/dec path-loss slope still remains. However, all reflected waves, including ground-reflected waves and building-reflected waves, become dominant. The short-term received signal at the mobile unit observes a Rayleigh fading. The Rayleigh fading is the most severe fading.

When the terrain contour blocks the direct wave path, we call it the *obstructive path*. In this situation, the shadow loss from the signal reception can be found by using the knife-edge diffraction curves shown in Sec. 4.7.2.

1.6.5 Noise level in cellular frequency band

The thermal noise kTB at a temperature T of 290 K (17°C) and a bandwidth B of 30 kHz is -129 dBm.* Assume that the received front-end noise is 9 dB, then the noise level is -120 dBm. Now there are two kinds of man-made noise, the ignition noise generated by the vehicles and the noise generated by 800-MHz emissions.

The ignition noise. In the past, 800 MHz was not widely used. Therefore, the man-made noise at 800 MHz is merely generated by the vehicle ignition noise.[10] The automotive noise introduced at 800 MHz with a bandwidth of 30 kHz can be deduced from Ref. 15, as shown in Fig. 1.11.

The 800-MHz-emission noise. As a result of the cellular mobile systems operating in all the major cities in the United States and the spurious energy generated outside each channel bandwidth, the early noise data measurements[9, 10] are no longer valid. The 800-MHz-emission noise can be measured at an idle channel (a forward voice channel) in the 870- to 890-MHz region while the mobile receiver is operating on a car battery in a no-traffic spot in a city. In this case, no automotive ignition

* k is Boltzmann's constant, and $kT = -174$ dBm/Hz at $T = 290$ K.

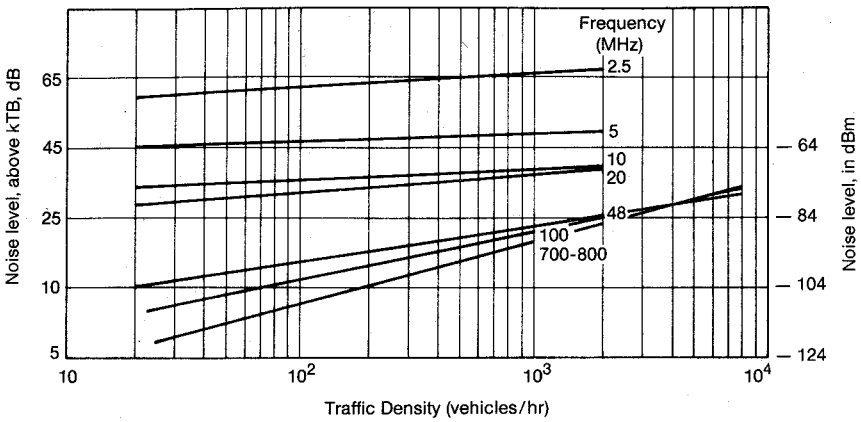


Figure 1.11 Average automotive-traffic-noise power for various traffic densities and frequencies. Detector noise bandwidth 30 kHz at room temperature (17°C). (*After Lee, Ref. 16.*)

noise is involved, and no cochannel operation is in the proximity of the idle-channel receiver. We found that in some areas the noise level is 2 to 3 dB higher than −120 dBm at the cell sites and 3 to 4 dB higher than −120 dBm at the mobile stations.

1.6.6 Amplifier noise

A mobile radio signal received by a receiving antenna, either at the cell site or at the mobile unit, will be amplified by an amplifier. We would like to understand how the signal is affected by the amplifier noise. Assume that the amplifier has an available power gain g and the available noise power at the output is N_0. The input signal-to-noise (S/N) ratio is P_s/N_i, the output signal-to-noise ratio is P_o/N_o, and the internal amplifier noise is N_a. Then the output P_o/N_o becomes

$$\frac{P_o}{N_o} = \frac{gP_s}{g(N_i) + N_a} = \frac{P_s}{N_i + (N_a/g)} \tag{1.6-20}$$

The noise figure F is defined as

$$F = \frac{\text{maximum possible S/N ratio}}{\text{actual S/N ratio at output}} \tag{1.6-21}$$

where the maximum possible S/N ratio is measured when the load is an open circuit. Equation (1.6-21) can be used for obtaining the noise figure of the amplifier.

$$F = \frac{P_s/kTB}{P_o/N_o} = \frac{N_o}{(P_o/P_s)kTB} = \frac{N_o}{g(kTB)} \tag{1.6-22}$$

Also substituting Eq. (1.6-20) into Eq. (1.6-22) yields

$$F = \frac{P_s/kTB}{P_s/[N_i + (N_a/g)]} = \frac{N_i + (N_a/g)}{kTB} \tag{1.6-23}$$

The term kTB is the thermal noise as described in Sec. 1.6.5. The noise figure is a reference measurement between a minimum noise level due to thermal noise and the noise level generated by both the external and internal noise of an amplifier.

1.7 Operation of Cellular Systems

This section briefly describes the operation of the cellular mobile system from a customer's perception without touching on the design parameters.[17, 18] The operation can be divided into four parts and a handoff procedure.

Mobile unit initialization. When a user sitting in a car activates the receiver of the mobile unit, the receiver scans 21 set-up channels which are designated among the 333 channels. It then selects the strongest and locks on for a certain time. Since each site is assigned a different set-up channel, locking onto the strongest set-up channel usually means selecting the nearest cell site. This self-location scheme is used in the idle stage and is user-independent. It has a great advantage because it eliminates the load on the transmission at the cell site for locating the mobile unit. The disadvantage of the self-location scheme is that no location information of idle mobile units appears at each cell site. Therefore, when the call initiates from the land line to a mobile unit, the paging process is longer. Since a large percentage of calls originates at the mobile unit, the use of self-location schemes is justified. After 60 s, the self-location procedure is repeated. In the future, when land-line originated calls increase, a feature called "registration" can be used.

Mobile originated call. The user places the called number into an originating register in the mobile unit, checks to see that the number is correct, and pushes the "send" button. A request for service is sent on a selected set-up channel obtained from a self-location scheme. The cell site receives it, and in directional cell sites, selects the best directive antenna for the voice channel to use. At the same time the cell site sends a request to the mobile telephone switching office (MTSO) via a high-speed data link. The MTSO selects an appropriate voice channel for the call, and the cell site acts on it through the best directive antenna to link the mobile unit. The MTSO also connects the wire-line party through the telephone company zone office.

Network originated call. A land-line party dials a mobile unit number. The telephone company zone office recognizes that the number is mobile and forwards the call to the MTSO. The MTSO sends a paging message to certain cell sites based on the mobile unit number and the search algorithm. Each cell site transmits the page on its own set-up channel. The mobile unit recognizes its own identification on a strong set-up channel, locks onto it, and responds to the cell site. The mobile unit also follows the instruction to tune to an assigned voice channel and initiate user alert.

Call termination. When the mobile user turns off the transmitter, a particular signal (signaling tone) transmits to the cell site, and both sides free the voice channel. The mobile unit resumes monitoring pages through the strongest set-up channel.

Handoff procedure. During the call, two parties are on a voice channel. When the mobile unit moves out of the coverage area of a particular cell site, the reception becomes weak. The present cell site requests a handoff. The system switches the call to a new frequency channel in a new cell site without either interrupting the call or alerting the user. The call continues as long as the user is talking. The user does not notice the handoff occurrences.

1.8 Marketing Image of Hexagonal-Shaped Cells

We have to realize that hexagonal-shaped communication cells are artificial and that such a shape cannot be generated in the real world. Engineers draw hexagonal-shaped cells on a layout to simplify the planning and design of a cellular system because it approaches a circular shape that is the ideal power coverage area. The circular shapes have overlapped areas which make the drawing unclear. The hexagonal-shaped cells fit the planned area nicely, as shown in Fig. 1.12, with no gap and no overlap between the hexagonal cells. The ideal cell shapes as well as the real cell shapes are also shown in Fig. 1.12.

A simple mechanism which makes the cellular system implementable based on hexagonal cells will be illustrated in later chapters. Otherwise, a statistical approach will be used in dealing with a real-world situation. Fortunately, the outcomes resulting from these two approaches are very close, yet the latter does not provide a clear physical picture, as shown later. Besides, today these hexagonal-shaped cells have already become a widely promoted symbol for cellular mobile systems. An analysis using hexagonal cells, if it is desired, can easily be adapted by the reader.

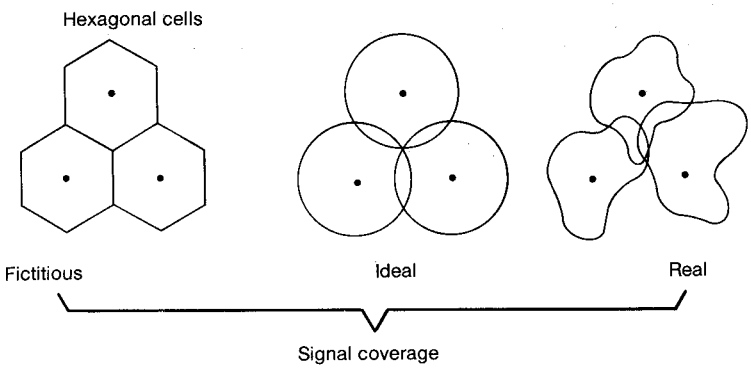


Figure 1.12 Hexagonal cells and the real shapes of their coverages.

1.9 Planning a Cellular System

1.9.1 How to start planning

Assume that the construction permit for a cellular system in a particular market area is granted. The planning stage becomes critical. A great deal of money can be spent and yet poor service may be provided if we do not know how to create a good plan. First, we have to determine two elements: regulations and the market situation.

Regulations. The federal regulations administered by the FCC are the same throughout the United States. The state regulations may be different from state to state, and each city and town may have its own building codes and zoning laws. Become familiar with the rules and regulations. Sometimes waivers need to be applied for ahead of time. Be sure that the plan is workable.

Market situation. There are three tasks to be handled by the marketing department.

1. *Prediction of gross income.* We have to determine the population, average income, business types, and business zones so that the gross income can be predicted.
2. *Understanding competitors.* We also need to know the competitor's situation, coverage, system performance, and number of customers. Any system should provide a unique and outstanding service to overcome the competition.
3. *Decision of geographic coverage.* What general area should ultimately be covered? What near-term service can be provided in a limited area? These questions should be answered and the decisions passed on to the engineering department.

1.9.2 The engineer's role

The engineers follow the market decisions by

1. Initiating a cellular mobile service in a given area by creating a plan that uses a minimum number of cell sites to cover the whole area. It is easy for marketing to request but hard for the engineers to fulfill. We will address this topic later.
2. Checking the areas that marketing indicated were important revenue areas. The number of radios (number of voice channels) required to handle the traffic load at the busy hours should be determined.
3. Studying the interference problems, such as cochannel and adjacent channel interference, and the intermodulation products generated at the cell sites, and finding ways to reduce them.
4. Studying the blocking probability of each call at each cell site, and trying to minimize it.
5. Planning to absorb more new customers. The rate at which new customers subscribe to a system can vary depending on the service charges, system performance, and seasons of the year. Engineering has to try to develop new technologies to utilize fully the limited spectrum assigned to the cellular system. The analysis of spectrum efficiency due to the natural limitations may lead to a request for a larger spectrum.

1.9.3 Finding solutions

Many practical design tools, methods of reducing interference, and ways of solving the blocking probability of call initiation will be introduced in this book.

1.10 Cellular Systems

1.10.1 Cellular systems in the United States

There are 150 major market areas* in the United States where licenses for cellular systems can be granted by the FCC. They have been classified by their populations into five groups. Each group has 30 cities.

1. Top 30 markets—very large cities
2. Top 31 to 60 markets—large-sized cities

* There are 305 defined MSAs (metropolitan statistical areas) and more than 400 proposed RSAs (rural statistical areas) in the United States.

3. Top 61 to 90 markets—medium-sized cities
4. Top 91 to 120 markets—below medium-sized cities
5. Top 121 to 150 markets—small-sized cities

Each market area is planned to have two systems. The status of each system in each area of groups 1 to 3 as of December 1985 appears in Appendix 1.2. The specifications of the system are described in detail in Chap. 3.

1.10.2 Cellular systems outside the United States

Japan.[6] Nippon Telegraph and Telephone Corporation (NTT) developed an 800-MHz land mobile telephone system and put it into service in the Tokyo area in 1979. The general system operation is similar to the AMPS system. It accesses approximately 40,000 subscribers in 500 cities. It covers 75 percent of all Japanese cities, 25 percent of inhabitable areas, and 60 percent of the population. In Japan, 9 automobile switching centers (ASCs), 51 mobile control stations (MCSs), 465 mobile base stations (MBSs), and 39,000 mobile subscriber stations (MSSs) were in operation as of February 1985.

The Japanese mobile telephone service network configuration is shown in Fig. 1.13. In the metropolitan Tokyo area, about 30,000 subscribers are being served.

The 1985 system operated over a spectrum of 30 MHz. The total number of channels was 600, and the channel bandwidth was 25 kHz. This system comprised an automobile switching center (ASC), a mobile control station (MCS), a mobile base station (MBS), and a mobile subscriber station (MSS). At present there is no competitive situation set up by the government. However, the Japanese Ministry of Post and Telecommunication (MPT) is considering providing a dual competitive situation similar to that in the United States.

United Kingdom.[7] In June 1982 the government of the United Kingdom announced two competing national cellular radio networks. The UK system is called TACS (Total Access Communications System). The total number of channels was 1000, with a channel bandwidth of 25 kHz per channel. Among them, 600 channels are assigned and 400 are reserved. Two competing cellular network operators, Cellnet and Vodafone, are operating in the United Kingdom. Each network system has only 300 spectral channels. The Cellnet system started operating in January 1985. Cellnet has over 200 cell sites, covering 82 percent of the United Kingdom. Vodaphone, though, which started operations late, has served the same areas as Cellnet.

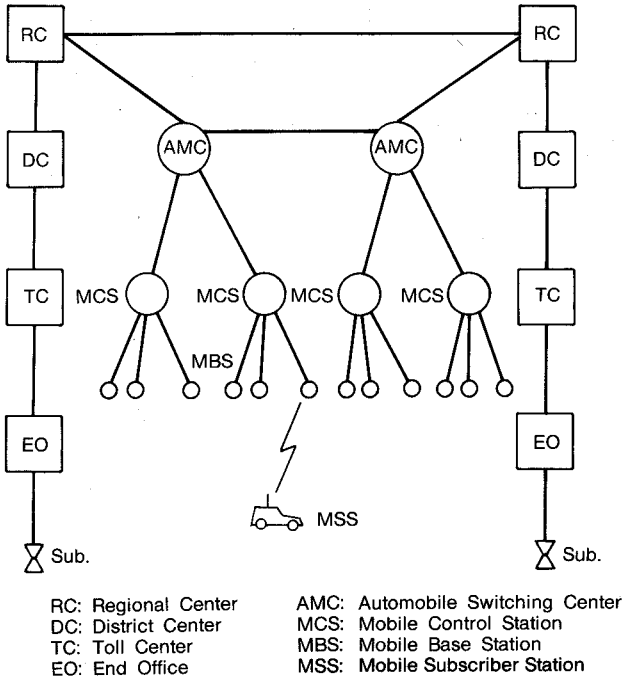


Figure 1.13 Japanese mobile telephone service network configuration.

Canadian system.[8] In 1978, a system called AURORA was designed for the Alberta government telephone (AGT). The system provides provincewide mobile telephone service at 400 MHz. Ongoing developmental work on the AURORA is underway at 800 MHz.

AURORA 400 system. It is aimed at 40,000 subscribers living in an area approximately 1920 km × 960 km. The AURORA 400 system initially has 40 channels and is expected to add an additional 20 channels with frequency reuse and a seven-cell cluster plan. A fully implemented system has 120 cells. The 400-MHz system does not have a handoff capability.

AURORA 800 system. The AURORA 800 system is truly frequency-transparent. By repackaging the radio frequency (RF) sections on the cell site, the mobile unit can be operated on any mobile RF band up to 800 MHz. The handoff capability will be implemented in this system.

Nordic system. This system was built mostly by Scandinavian countries (Denmark, Norway, Sweden, and Finland) in cooperation with Saudi Arabia and Spain and is called the *NMT network*. It is currently

a 450-MHz system, but an 800-MHz system will be implemented soon since the frequency-transparent concept as the AURORA 800 system is used to convert the 450-MHz system to the 800-MHz system.

The total bandwidth is 10 MHz, which has 200 channels with a bandwidth of 25 kHz per channel. This system does have handoff and roaming capabilities. It also uses repeaters to increase the coverage in a low traffic area. The total number of subscribers is around 100,000.

European cellular systems.[11] All the present generation of European cellular networks is totally lacking in cross-border compatibility. Besides the United Kingdom and NMT networks, the others include the following.

Benelux-country network. The Netherlands served on their ATF2 network (the same as the NMT 450 network) at the beginning of 1985. It has a nationwide coverage using 50 cell sites with two different cell sizes, 20- and 5-km radii. The capacity of the present system is 15,000 to 20,000 subscribers. Dutch PT&T is using a single Ericsson AXE10 switch. Luxembourg came on air in August 1985. In 1986, Belgium joined the network. It operates at 450 MHz. The network is compatible among the three countries.

France. A direct-dial car telephone operating at 160 MHz can access the system in 10 regional areas. The network serves 10,000 subscribers. By the end of 1984, 450 MHz was in operation. In the meantime Radicom 2000 (digital signaling) was introduced, operating at 200 MHz but with no handoff feature.

Spain. It uses an NMT 450-MHz cellular network introduced in 1982. It was the first cellular system in Europe. The number of cells in service is 13. There are three separate networks operating 104 channels. Each channel bandwidth is 25 kHz.

Austria. A new NMT cellular network called *Autotelefonnetz C* has two mobile switching exchanges and has enough capacity for 30,000 subscribers.

The Austrian PT&T has allocated 222 duplex channels in ranges 451.3 to 455.7 MHz and 461.3 to 465.7 MHz, with a channel bandwidth of 20 kHz.

Although both Austria and Spain are using NMT 450 systems, their systems are not compatible because of different frequency allocations, channel spacings (bandwidth), and protocols by different PT&Ts.

Germany. A full national coverage, including West Berlin, using a C-450 cellular system was installed in September 1985 with 100 cell sites. Another 75 cell sites were completed in mid-1986. Also, Germany

and France are working on cross-border compatibility in cellular radio systems and have proposed a CD-900 digital system. The details of the CD-900 system will be described in Chap. 14.

Switzerland. Swiss PT&T decided to install an NMT 900-MHz cellular network that had a capacity of 12,000 subscribers. A pilot scheme with 20 transmitters (cell sites) was installed in the Zurich area in late 1986.

Cellular systems in the rest of the world. Australia is installing a system using Ericsson's AXE-10 switching network and will operate at 800 MHz with 12 sites concentrated in three big cities.

Kuwait's cellular system uses NEC's switches and provides 12 sites. It operates at 800 MHz.

Hong Kong has three systems. The United Kingdom's TACS system is installed with Motorola switches. The United States' AMPS system and Japanese NEC systems were also installed in Hong Kong. It is a very competitive market. All systems are penetrating the markets of both portable sets and car sets.

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APPENDIX 1.1
Blocked-Calls-Cleared (Erlang *B*)

	A, erlangs												
	B												
N	1.0%	1.2%	1.5%	2%	3%	5%	7%	10%	15%	20%	30%	40%	50%
1	.0101	.0121	.0152	.0204	.0309	.0526	.0753	.111	.176	.250	.429	.667	1.00
2	.153	.168	.190	.223	.282	.381	.470	.595	.796	1.00	1.45	2.00	2.73
3	.455	.489	.535	.602	.715	.899	1.06	1.27	1.60	1.93	2.63	3.48	4.59
4	.869	.922	.992	1.09	1.26	1.52	1.75	2.05	2.50	2.95	3.39	5.02	6.50
5	1.36	1.43	1.52	1.66	1.88	2.22	2.50	2.88	3.45	4.01	5.19	6.60	8.44
6	1.91	2.00	2.11	2.28	2.54	2.96	3.30	3.76	4.44	5.11	6.51	8.19	10.4
7	2.50	2.60	2.74	2.94	3.25	3.74	4.14	4.67	5.46	6.23	7.86	9.80	12.4
8	3.13	3.25	3.40	3.63	3.99	4.54	5.00	5.60	6.50	7.37	9.21	11.4	14.3
9	3.78	3.92	4.09	4.34	4.75	5.37	5.88	6.55	7.55	8.52	10.6	13.0	16.3
10	4.46	4.61	4.81	5.08	5.53	6.22	6.78	7.51	8.62	9.68	12.0	14.7	18.3
11	5.16	5.32	5.54	5.84	6.33	7.08	7.69	8.49	9.69	10.9	13.3	16.3	20.3
12	5.88	6.05	6.29	6.61	7.14	7.95	8.61	9.47	10.8	12.0	14.7	18.0	22.2
13	6.61	6.80	7.05	7.40	7.97	8.83	9.54	10.5	11.9	13.2	16.1	19.6	24.2
14	7.35	7.56	7.82	8.20	8.80	9.73	10.5	11.5	13.0	14.4	17.5	21.2	26.2
15	8.11	8.33	8.61	9.01	9.65	10.6	11.4	12.5	14.1	15.6	18.9	22.9	28.2
16	8.88	9.11	9.41	9.83	10.5	11.5	12.4	13.5	15.2	16.8	20.3	24.5	30.2
17	9.65	9.89	10.2	10.7	11.4	12.5	13.4	14.5	16.3	18.0	21.7	26.2	32.2
18	10.4	10.7	11.0	11.5	12.2	13.4	14.3	15.5	17.4	19.2	23.1	27.8	34.2
19	11.2	11.5	11.8	12.3	13.1	14.3	15.3	16.6	18.5	20.4	24.5	29.5	36.2
20	12.0	12.3	12.7	13.2	14.0	15.2	16.3	17.6	19.6	21.6	25.9	31.2	38.2

21	12.8	13.1	13.5	14.0	14.9	16.2	17.3	18.7	20.8	22.8	27.3	32.8	40.2
22	13.7	14.0	14.3	14.9	15.8	17.1	18.2	19.7	21.9	24.1	28.7	34.5	42.1
23	14.5	14.8	15.2	15.8	16.7	18.1	19.2	20.7	23.0	25.3	30.1	36.1	44.1
24	15.3	15.6	16.0	16.6	17.6	19.0	20.2	21.8	24.2	26.5	31.6	37.8	46.1
25	16.1	16.5	16.9	17.5	18.5	20.0	21.2	22.8	25.3	27.7	33.0	39.4	48.1
26	17.0	17.3	17.8	18.4	19.4	20.9	22.2	23.9	26.4	28.9	34.4	41.1	50.1
27	17.8	18.2	18.6	19.3	20.3	21.9	23.2	24.9	27.6	30.2	35.8	42.8	52.1
28	18.6	19.0	19.5	20.2	21.2	22.9	24.2	26.0	28.7	31.4	37.2	44.4	54.1
29	19.5	19.9	20.4	21.0	22.1	23.8	25.2	27.1	29.9	32.6	38.6	46.1	56.1
30	20.3	20.7	21.2	21.9	23.1	24.8	26.2	28.1	31.0	33.8	40.0	47.7	58.1
31	21.2	21.6	22.1	22.8	24.0	25.8	27.2	29.2	32.1	35.1	41.5	49.4	60.1
32	22.0	22.5	23.0	23.7	24.9	26.7	28.2	30.2	33.3	36.3	42.9	51.1	62.1
33	22.9	23.3	23.9	24.6	25.8	27.7	29.3	31.3	34.4	37.5	44.3	52.7	64.1
34	23.8	24.2	24.8	25.5	26.8	28.7	30.3	32.4	35.6	38.8	45.7	54.4	66.1
35	24.6	25.1	25.6	26.4	27.7	29.7	31.3	33.4	36.7	40.0	47.1	56.0	68.1
36	25.5	26.0	26.5	27.3	28.6	30.7	32.3	34.5	37.9	41.2	48.6	57.7	70.1
37	26.4	26.8	27.4	28.3	29.6	31.6	33.3	35.6	39.0	42.4	50.0	59.4	72.1
38	27.3	27.7	28.3	29.2	30.5	32.6	34.4	36.6	40.2	43.7	51.4	61.0	74.1
39	28.1	28.6	29.2	30.1	31.5	33.6	35.4	37.7	41.3	44.9	52.8	62.7	76.1
40	29.0	29.5	30.1	31.0	32.4	34.6	36.4	38.8	42.5	46.1	54.2	64.4	78.1
41	29.9	30.4	31.0	31.9	33.4	35.6	37.4	39.9	43.6	47.4	55.7	66.0	80.1
42	30.8	31.3	31.9	32.8	34.3	36.6	38.4	40.9	44.8	48.6	57.1	67.7	82.1
43	31.7	32.2	32.8	33.8	35.3	37.6	39.5	42.0	45.9	49.9	58.5	69.3	84.1
44	32.5	33.1	33.7	34.7	36.2	38.6	40.5	43.1	47.1	51.1	59.9	71.0	86.1
45	33.4	34.0	34.6	35.6	37.2	39.6	41.5	44.2	48.2	52.3	61.3	72.7	88.1
46	34.3	34.9	35.6	36.5	38.1	40.5	42.6	45.2	49.4	53.6	62.8	74.3	90.1
47	35.2	35.8	36.5	37.5	39.1	41.5	43.6	46.3	50.6	54.8	64.2	76.0	92.1
48	36.1	36.7	37.4	38.4	40.0	42.5	44.6	47.4	51.7	56.0	65.6	77.7	94.1
49	37.0	37.6	38.3	39.3	41.0	43.5	45.7	48.5	52.9	57.3	67.0	79.3	96.1
50	37.9	38.5	39.2	40.3	41.9	44.5	46.7	49.6	54.0	58.5	68.5	81.0	98.1

APPENDIX 1.1
Blocked-Calls-Cleared (Erlang *B*) (*Continued*)

	A, erlangs												
	B												
N	1.0%	1.2%	1.5%	2%	3%	5%	7%	10%	15%	20%	30%	40%	50%
51	38.8	39.4	40.1	41.2	42.9	45.5	47.7	50.6	55.2	59.7	69.9	82.7	100.1
52	39.7	40.3	41.0	42.1	43.9	46.5	48.8	51.7	56.3	61.0	71.3	84.3	102.1
53	40.6	41.2	42.0	43.1	44.8	47.5	49.8	52.8	57.5	62.2	72.7	86.0	104.1
54	41.5	42.1	42.9	44.0	45.8	48.5	50.8	53.9	58.7	63.5	74.2	87.6	106.1
55	42.4	43.0	43.8	44.9	46.7	49.5	51.9	55.0	59.8	64.7	75.6	89.3	108.1
56	43.3	43.9	44.7	45.9	47.7	50.5	52.9	56.1	61.0	65.9	77.0	91.0	110.1
57	44.2	44.8	45.7	46.8	48.7	51.5	53.9	57.1	62.1	67.2	78.4	92.6	112.1
58	45.1	45.8	46.6	47.8	49.6	52.6	55.0	58.2	63.3	68.4	79.8	94.3	114.1
59	46.0	46.7	47.5	48.7	50.6	53.6	56.0	59.3	64.5	69.7	81.3	96.0	116.1
60	46.9	47.6	48.4	49.6	51.6	54.6	57.1	60.4	65.6	70.9	82.7	97.6	118.1
61	47.9	48.5	49.4	50.6	52.5	55.6	58.1	61.5	66.8	72.1	84.1	99.3	120.1
62	48.8	49.4	50.3	51.5	53.5	56.6	59.1	62.6	68.0	73.4	85.5	101.0	122.1
63	49.7	50.4	51.2	52.5	54.5	57.6	60.2	63.7	69.1	74.6	87.0	102.6	124.1
64	50.6	51.3	52.2	53.4	55.4	58.6	61.2	64.8	70.3	75.9	88.4	104.3	126.1
65	51.5	52.2	53.1	54.4	56.4	59.6	62.3	65.8	71.4	77.1	89.8	106.0	128.1
66	52.4	53.1	54.0	55.3	57.4	60.6	63.3	66.9	72.6	78.3	91.2	107.6	130.1
67	53.4	54.1	55.0	56.3	58.4	61.6	64.4	68.0	73.8	79.6	92.7	109.3	132.1
68	54.3	55.0	55.9	57.2	59.3	62.6	65.4	69.1	74.9	80.8	94.1	111.0	134.1
69	55.2	55.9	56.9	58.2	60.3	63.7	66.4	70.2	76.1	82.1	95.5	112.6	136.1
70	56.1	56.8	57.8	59.1	61.3	64.7	67.5	71.3	77.3	83.3	96.9	114.3	138.1

71	57.0	57.8	58.7	60.1	62.3	65.7	68.5	72.4	78.4	84.6	98.4	115.9	140.1
72	58.0	58.7	59.7	61.0	63.2	66.7	69.6	73.5	79.6	85.8	99.8	117.6	142.1
73	58.9	59.6	60.6	62.0	64.2	67.7	70.6	74.6	80.8	87.0	101.2	119.3	144.1
74	59.8	60.6	61.6	62.9	65.2	68.7	71.7	75.6	81.9	88.3	102.7	120.9	146.1
75	60.7	61.5	62.5	63.9	66.2	69.7	72.7	76.7	83.1	89.5	104.1	122.6	148.0
76	61.7	62.4	63.4	64.9	67.2	70.8	73.8	77.8	84.2	90.8	105.5	124.3	150.0
77	62.6	63.4	64.4	65.8	68.1	71.8	74.8	78.9	85.4	92.0	106.9	125.9	152.0
78	63.5	64.3	65.3	66.8	69.1	72.8	75.9	80.0	86.6	93.3	108.4	127.6	154.0
79	64.4	65.2	66.3	67.7	70.1	73.8	76.9	81.1	87.7	94.5	109.8	129.3	156.0
80	65.4	66.2	67.2	68.7	71.1	74.8	78.0	82.2	88.9	95.7	111.2	130.9	158.0
81	66.3	67.1	68.2	69.6	72.1	75.8	79.0	83.3	90.1	97.0	112.6	132.6	160.0
82	67.2	68.0	69.1	70.6	73.0	76.9	80.1	84.4	91.2	98.2	114.1	134.3	162.0
83	68.2	69.0	70.1	71.6	74.0	77.9	81.1	85.5	92.4	99.5	115.5	135.9	164.0
84	69.1	69.9	71.0	72.5	75.0	78.9	82.2	86.6	93.6	100.7	116.9	137.6	166.0
85	70.0	70.9	71.9	73.5	76.0	79.9	83.2	87.7	94.7	102.0	118.3	139.3	168.0
86	70.9	71.8	72.9	74.5	77.0	80.9	84.3	88.8	95.9	103.2	119.8	140.9	170.0
87	71.9	72.7	73.8	75.4	78.0	82.0	85.3	89.9	97.1	104.5	121.2	142.6	172.0
88	72.8	73.7	74.8	76.4	78.9	83.0	86.4	91.0	98.2	105.7	122.6	144.3	174.0
89	73.7	74.6	75.7	77.3	79.9	84.0	87.4	92.1	99.4	106.9	124.0	145.9	176.0
90	74.7	75.6	76.7	78.3	80.9	85.0	88.5	93.1	100.6	108.2	125.5	147.6	178.0
91	75.6	76.5	77.6	79.3	81.9	86.0	89.5	94.2	101.7	109.4	126.9	149.3	180.0
92	76.6	77.4	78.6	80.2	82.9	87.1	90.6	95.3	102.9	110.7	128.3	150.9	182.0
93	77.5	78.4	79.6	81.2	83.9	88.1	91.6	96.4	104.1	111.9	129.7	152.6	184.0
94	78.4	79.3	80.5	82.2	84.9	89.1	92.7	97.5	105.3	113.2	131.2	154.3	186.0
95	79.4	80.3	81.5	83.1	85.8	90.1	93.7	98.6	106.4	114.4	132.6	155.9	188.0
96	80.3	81.2	82.4	84.1	86.8	91.1	94.8	99.7	107.6	115.7	134.0	157.6	190.0
97	81.2	82.2	83.4	85.1	87.8	92.2	95.8	100.8	108.8	116.9	135.5	159.3	192.0
98	82.2	83.1	84.3	86.0	88.8	93.2	96.9	101.9	109.9	118.2	136.9	160.9	194.0
99	83.1	84.1	85.3	87.0	89.8	94.2	97.9	103.0	111.1	119.4	138.3	162.6	196.0
100	84.1	85.0	86.2	88.0	90.8	95.2	99.0	104.1	112.3	120.6	139.7	164.3	198.0

APPENDIX 1.1
Blocked-Calls-Cleared (Erlang *B*) (*Continued*)

	A, erlangs												
	B												
N	1.0%	1.2%	1.5%	2%	3%	5%	7%	10%	15%	20%	30%	40%	50%
102	85.9	86.9	88.1	89.9	92.8	97.3	101.1	106.3	114.6	123.1	142.6	167.6	202.0
104	87.8	88.8	90.1	91.9	94.8	99.3	103.2	108.5	116.9	125.6	145.4	170.9	206.0
106	89.7	90.7	92.0	93.8	96.7	101.4	105.3	110.7	119.3	128.1	148.3	174.2	210.0
108	91.6	92.6	93.9	95.7	98.7	103.4	107.4	112.9	121.6	130.6	151.1	177.6	214.0
110	93.5	94.5	95.8	97.7	100.7	105.5	109.5	115.1	124.0	133.1	154.0	180.9	218.0
112	95.4	96.4	97.7	99.6	102.7	107.5	111.7	117.3	126.3	135.6	156.9	184.2	222.0
114	97.3	98.3	99.7	101.6	104.7	109.6	113.8	119.5	128.6	138.1	159.7	187.6	226.0
116	99.2	100.2	101.6	103.5	106.7	111.7	115.9	121.7	131.0	140.6	162.6	190.9	230.0
118	101.1	102.1	103.5	105.5	108.7	113.7	118.0	123.9	133.3	143.1	165.4	194.2	234.0
120	103.0	104.0	105.4	107.4	110.7	115.8	120.1	126.1	135.7	145.6	168.3	197.6	238.0
122	104.9	105.9	107.4	109.4	112.6	117.8	122.2	128.3	138.0	148.1	171.1	200.9	242.0
124	106.8	107.9	109.3	111.3	114.6	119.9	124.4	130.5	140.3	150.6	174.0	204.2	246.0
126	108.7	109.8	111.2	113.3	116.6	121.9	126.5	132.7	142.7	153.0	176.8	207.6	250.0
128	110.6	111.7	113.2	115.2	118.6	124.0	128.6	134.9	145.0	155.5	179.7	210.9	254.0
130	112.5	113.6	115.1	117.2	120.6	126.1	130.7	137.1	147.4	158.0	182.5	214.2	258.0
132	114.4	115.5	117.0	119.1	122.6	128.1	132.8	139.3	149.7	160.5	185.4	217.6	262.0
134	116.3	117.4	119.0	121.1	124.6	130.2	134.9	141.5	152.0	163.0	188.3	220.9	266.0
136	118.2	119.4	120.9	123.1	126.6	132.3	137.1	143.7	154.4	165.5	191.1	224.2	270.0
138	120.1	121.3	122.8	125.0	128.6	134.3	139.2	145.9	156.7	168.0	194.0	227.6	274.0
140	122.0	123.2	124.8	127.0	130.6	136.4	141.3	148.1	159.1	170.5	196.8	230.9	278.0

142	123.9	125.1	126.7	128.9	132.6	138.4	143.4	150.3	161.4	173.0	199.7	234.2	282.0
144	125.8	127.0	128.6	130.9	134.6	140.5	145.6	152.5	163.8	175.5	202.5	237.6	286.0
146	127.7	129.0	130.6	132.9	136.6	142.6	147.7	154.7	166.1	178.0	205.4	240.9	290.0
148	129.7	130.9	132.5	134.8	138.6	144.6	149.8	156.9	168.5	180.5	208.2	244.2	294.0
150	131.6	132.8	134.5	136.8	140.6	146.7	151.9	159.1	170.8	183.0	211.1	247.6	298.0
152	133.5	134.8	136.4	138.8	142.6	148.8	154.0	161.3	173.1	185.5	214.0	250.9	302.0
154	135.4	136.7	138.4	140.7	144.6	150.8	156.2	163.5	175.5	188.0	216.8	254.2	306.0
156	137.3	138.6	140.3	142.7	146.6	152.9	158.3	165.7	177.8	190.5	219.7	257.6	310.0
158	139.2	140.5	142.3	144.7	148.6	155.0	160.4	167.9	180.2	193.0	222.5	260.9	314.0
160	141.2	142.5	144.2	146.6	150.6	157.0	162.5	170.2	182.5	195.5	225.4	264.2	318.0
162	143.1	144.4	146.1	148.6	152.7	159.1	164.7	172.4	184.9	198.0	228.2	267.6	322.0
164	145.0	146.3	148.1	150.6	154.7	161.2	166.8	174.6	187.2	200.4	231.1	270.9	326.0
166	146.9	148.3	150.0	152.6	156.7	163.3	168.9	176.8	189.6	202.9	233.9	274.2	330.0
168	148.9	150.2	152.0	154.5	158.7	165.3	171.0	179.0	191.9	205.4	236.8	277.6	334.0
170	150.8	152.1	153.9	156.5	160.7	167.4	173.2	181.2	194.2	207.9	239.7	280.9	338.0
172	152.7	154.1	155.9	158.5	162.7	169.5	175.3	183.4	196.6	210.4	242.5	284.2	342.0
174	154.6	156.0	157.8	160.4	164.7	171.5	177.4	185.6	198.9	212.9	245.4	287.6	346.0
176	156.6	158.0	159.8	162.4	166.7	173.6	179.6	187.8	201.3	215.4	248.2	290.9	350.0
178	158.5	159.9	161.8	164.4	168.7	175.7	181.7	190.0	203.6	217.9	251.1	294.2	354.0
180	160.4	161.8	163.7	166.4	170.7	177.8	183.8	192.2	206.0	220.4	253.9	297.5	358.0
182	162.3	163.8	165.7	168.3	172.8	179.8	185.9	194.4	208.3	222.9	256.8	300.9	362.0
184	164.3	165.7	167.6	170.3	174.8	181.9	188.1	196.6	210.7	225.4	259.6	304.2	366.0
186	166.2	167.7	169.6	172.3	176.8	184.0	190.2	198.9	213.0	227.9	262.5	307.5	370.0
188	168.1	169.6	171.5	174.3	178.8	186.1	192.3	201.1	215.4	230.4	265.4	310.9	374.0
190	170.1	171.5	173.5	176.3	180.8	188.1	194.5	203.3	217.7	232.9	268.2	314.2	378.0
192	172.0	173.5	175.4	178.2	182.8	190.2	196.6	205.5	220.1	235.4	271.1	317.5	382.0
194	173.9	175.4	177.4	180.2	184.8	192.3	198.7	207.7	222.4	237.9	273.9	320.9	386.0
196	175.9	177.4	179.4	182.2	186.9	194.4	200.8	209.9	224.8	240.4	276.8	324.2	390.0
198	177.8	179.3	181.3	184.2	188.9	196.4	203.0	212.1	227.1	242.9	279.6	327.5	394.0
200	179.7	181.3	183.3	186.2	190.9	198.5	205.1	214.3	229.4	245.4	282.5	330.9	398.0

APPENDIX 1.1
Blocked-Calls-Cleared (Erlang *B*) (*Continued*)

	A, erlangs												
	B												
N	1.0%	1.2%	1.5%	2%	3%	5%	7%	10%	15%	20%	30%	40%	50%
202	181.7	183.2	185.2	188.1	192.9	200.6	207.2	216.5	231.8	247.9	285.4	334.2	402.0
204	183.6	185.2	187.2	190.1	194.9	202.7	209.4	218.7	234.1	250.4	288.2	337.5	406.0
206	185.5	187.1	189.2	192.1	196.9	204.7	211.5	221.0	236.5	252.9	291.1	340.9	410.0
208	187.5	189.1	191.1	194.1	199.0	206.8	213.6	223.2	238.8	255.4	293.9	344.2	414.0
210	189.4	191.0	193.1	196.1	201.0	208.9	215.8	225.4	241.2	257.9	296.8	347.5	418.0
212	191.4	193.0	195.1	198.1	203.0	211.0	217.9	227.6	243.5	260.4	299.6	350.9	422.0
214	193.3	194.9	197.0	200.0	205.0	213.0	220.0	229.8	245.9	262.9	302.5	354.2	426.0
216	195.2	196.9	199.0	202.0	207.0	215.1	222.2	232.0	248.2	265.4	305.3	357.5	430.0
218	197.2	198.8	201.0	204.0	209.1	217.2	224.3	234.2	250.6	267.9	308.2	360.9	434.0
220	199.1	200.8	202.9	206.0	211.1	219.3	226.4	236.4	252.9	270.4	311.1	364.2	438.0
222	201.1	202.7	204.9	208.0	213.1	221.4	228.6	238.6	255.3	272.9	313.9	367.5	442.0
224	203.0	204.7	206.8	210.0	215.1	223.4	230.7	240.9	257.6	275.4	316.8	370.9	446.0
226	204.9	206.6	208.8	212.0	217.1	225.5	232.8	243.1	260.0	277.8	319.6	374.2	450.0
228	206.9	208.6	210.8	213.9	219.2	227.6	235.0	245.3	262.3	280.3	322.5	377.5	454.0
230	208.8	210.5	212.8	215.9	221.2	229.7	237.1	247.5	264.7	282.8	325.3	380.9	458.0
232	210.8	212.5	214.7	217.9	223.2	231.8	239.2	249.7	267.0	285.3	328.2	384.2	462.0
234	212.7	214.4	216.7	219.9	225.2	233.8	241.4	251.9	269.4	287.8	331.1	387.5	466.0
236	214.7	216.4	218.7	221.9	227.2	235.9	243.5	254.1	271.7	290.3	333.9	390.9	470.0
238	216.6	218.3	220.6	223.9	229.3	238.0	245.6	256.3	274.1	292.8	336.8	394.2	474.0
240	218.6	220.3	222.6	225.9	231.3	240.1	247.8	258.6	276.4	295.3	339.6	397.5	478.0

242	220.5	222.3	224.6	227.9	233.3	242.2	249.9	260.8	278.8	297.8	342.5	400.9	482.0
244	222.5	224.2	226.5	229.9	235.3	244.3	252.0	263.0	281.1	300.3	345.3	404.2	486.0
246	224.4	226.2	228.5	231.8	237.4	246.3	254.2	265.2	283.4	302.8	348.2	407.5	490.0
248	226.3	228.1	230.5	233.8	239.4	248.4	256.3	267.4	285.8	305.3	351.0	410.9	494.0
250	228.3	230.1	232.5	235.8	241.4	250.5	258.4	269.6	288.1	307.8	353.9	414.2	498.0
	.976	*.982*	*.988*	*.998*	*1.014*	*1.042*	*1.070*	*1.108*	*1.176*	*1.250*	*1.428*	*1.666*	*2.000*
300	277.1	279.2	281.9	285.7	292.1	302.6	311.9	325.0	346.9	370.3	425.3	497.5	598.0
	.982	*.984*	*.990*	*1.000*	*1.016*	*1.044*	*1.070*	*1.108*	*1.174*	*1.248*	*1.428*	*1.668*	*2.000*
350	326.2	328.4	331.4	335.7	342.9	354.8	365.4	380.4	405.6	432.7	496.7	580.9	698.0
	.982	*.988*	*.994*	*1.004*	*1.020*	*1.046*	*1.070*	*1.108*	*1.176*	*1.250*	*1.430*	*1.666*	*2.000*
400	375.3	377.8	381.1	385.9	393.9	407.1	418.9	435.8	464.4	495.2	568.2	664.2	798.0
	.986	*.990*	*.996*	*1.004*	*1.018*	*1.046*	*1.072*	*1.110*	*1.176*	*1.250*	*1.428*	*1.666*	*2.000*
450	424.6	427.3	430.9	436.1	444.8	459.4	472.5	491.3	523.2	557.7	639.6	747.5	898.0
	.988	*.994*	*.998*	*1.006*	*1.022*	*1.048*	*1.070*	*1.108*	*1.176*	*1.250*	*1.428*	*1.668*	*2.000*
500	474.0	477.0	480.8	486.4	495.9	511.8	526.0	546.7	582.0	620.2	711.0	830.9	998.0
	.991	*.994*	*1.000*	*1.008*	*1.022*	*1.047*	*1.073*	*1.110*	*1.176*	*1.249*	*1.429*	*1.666*	*2.000*
600	573.1	576.4	580.8	587.2	598.1	616.5	633.3	657.7	699.6	745.1	853.9	997.5	1198.
	.993	*.997*	*1.002*	*1.010*	*1.024*	*1.049*	*1.073*	*1.110*	*1.176*	*1.250*	*1.428*	*1.665*	*2.00*
700	672.4	676.1	681.0	688.2	700.5	721.4	740.6	768.7	817.2	870.1	996.7	1164.	1398.
	.994	*.998*	*1.004*	*1.011*	*1.025*	*1.050*	*1.073*	*1.110*	*1.176*	*1.250*	*1.433*	*1.67*	*2.00*
800	771.8	775.9	781.4	789.3	803.0	826.4	847.9	879.7	934.8	995.1	1140.	1331.	1598.
	.997	*1.000*	*1.004*	*1.013*	*1.025*	*1.050*	*1.074*	*1.111*	*1.172*	*1.249*	*1.42*	*1.67*	*2.00*
900	871.5	875.9	881.8	890.6	905.5	931.4	955.3	990.8	1052.	1120.	1282.	1498.	1798.
	.997	*1.001*	*1.006*	*1.013*	*1.025*	*1.046*	*1.077*	*1.112*	*1.18*	*1.25*	*1.43*	*1.66*	*2.00*
1000	971.2	976.0	982.4	991.9	1008.	1036.	1063.	1102.	1170.	1245.	1425.	1664.	1998.
	.998	*1.000*	*1.006*	*1.011*	*1.03*	*1.05*	*1.07*	*1.11*	*1.18*	*1.25*	*1.43*	*1.67*	*2.00*
1100	1071.	1076.	1083.	1093.	1111.	1141.	1170.	1213.	1288.	1370.	1568.	1831.	2198.

SOURCE: After Lee, Ref. 5, pp. 258–265.

APPENDIX 1.2
Status of First 90 Cities (September 1986)

Key: W—wire-line carrier. NW—non-wire-line carrier. CPG—construction permit granted. Information available as of July 23, 1986. Boldface type indicates markets on line.

MSA no./name	System operators	Status	No. of cells	Switching equipment
1. New York	W—Nynex Mobile	On line 6/15/84	48	AT&T
	NW—Metro One	On line 4/5/86	24	Motorola
2. Los Angeles	W—PacTel Mobile Access	On line 6/13/84	29	AT&T
	NW—LA Cellular Telephone	**CPG 12/4/84**	**24**	**Ericsson**
3. Chicago	W—Ameritech Mobile	On line 10/13/83	44	AT&T
	NW—Cellular One	On line 1/3/85	18	Ericsson
4. Philadelphia	W—Bell Atlantic Mobile	On line 7/12/84	32[a]	AT&T
	NW—Metrophone	On line 2/12/86	18	Motorola
5. Detroit	W—Ameritech Mobile	On line 9/21/84	19	AT&T
	NW—Cellular One	On line 7/30/85	15	Ericsson
6. Boston	W—Nynex Mobile	On line 1/1/85	20	AT&T
	NW—Cellular One	On line 1/1/85	10	Motorola
7. San Francisco	W—GTE Mobilnet	On line 4/2/85	38[b]	Motorola
	NW—Cellular One	**CPG 8/9/84**	**27**	**Ericsson**
8. Washington	W—Bell Atlantic Mobile	On line 4/2/84	38[c]	AT&T
	NW—Cellular One	On line 12/16/83	25[c]	Motorola
9. Dallas	W—Southwestern Bell Mobile	On line 7/31/84	41	AT&T
	NW—MetroCel	On line 3/1/86	28	Motorola
10. Houston	W—GTE Mobilnet	On line 9/28/84	43	Motorola
	NW—Houston Cellular Telephone	On line 5/16/86	29	Ericsson
11. St. Louis	W—Southwestern Bell Mobile	On line 7/16/84	17	AT&T
	NW—CyberTel	On line 7/16/84	13	Motorola
12. Miami	W—BellSouth Mobility	On line 5/25/84	23[d]	AT&T
	NW—Cellular One	**CPG 4/26/85**	**16**	**NTI/GE**
13. Pittsburgh	W—Bell Atlantic Mobile	On line 12/10/84	20	AT&T
	NW—Cellular One	**CPG 3/6/84**	**17**	**Ericsson**
14. Baltimore	W—Bell Atlantic Mobile	On line 4/2/84	38[c]	AT&T
	NW—Cellular One	On line 12/16/83	25[c]	Motorola
15. Minneapolis	W—NewVector Communications	On line 6/6/84	13	NTI/GE
	NW—Cellular One	On line 7/23/84	11	NTI/GE
16. Cleveland	W—GTE Mobilnet	On line 12/18/84	20	Motorola
	NW—Cellular One	On line 5/31/85	7	NTI/GE
17. Atlanta	W—BellSouth Mobility	On line 9/5/84	12	AT&T
	NW—Gencell	**CPG 1/18/85**	**10**	**Motorola[f]**
18. San Diego	W—PacTel Mobile Access	On line 8/15/85	8	AT&T
	NW—Vector One	On line 4/1/86		Motorola

APPENDIX 1.2
Status of First 90 Cities (September 1986) (*Continued*)

MSA no./name	System operators	Status	No. of cells	Switching equipment
19. Denver	W—NewVector Communications	On line 7/10/84	10	NTI/GE
	NW—Cellular One	**CPG 1/31/85**	**11**	**NEC**[f]
20. Seattle	W—NewVector Communications	On line 7/12/84	13	NTI/GE
	NW—Cellular One	On line 12/12/85	15[e]	AT&T
21. Milwaukee	W—Ameritech Mobile	On line 8/1/84	9	AT&T
	NW—Milwaukee Telephone Co.	On line 6/1/84	7	Motorola
22. Tampa	W—GTE Mobilnet	On line 11/30/84	26	Motorola
	NW—Bayfone	**CPG 4/26/85**		
23. Cincinnati	W—Ameritech Mobile	On line 11/5/84	13	AT&T
	NW—Cellular One	**CPG 1/9/85**		**Ericsson**
24. Kansas City	W—Southwestern Bell Mobile	On line 8/14/84	13	Motorola
	NW—Cellular One	On line 2/14/86	11	AT&T
25. Buffalo	W—Nynex Mobile	On line 4/16/84	7	Motorola
	NW—Buffalo Telephone	On line 6/1/84	9	Ericsson
26. Phoenix	W—NewVector Communications	On line 8/15/84	9	NTI/GE
	NW—Metro Mobile CTS	On line 3/1/86	10	Motorola
27. San Jose	W—GTE Mobilnet	On line 4/2/85	38[b]	Motorola
	NW—Cellular One	**CPG 8/9/84**	**27**	**Ericsson**
28. Indianapolis	W—GTE Mobilnet	On line 5/3/84	8	Motorola
	NW—Cellular One	On line 2/3/84	9	Motorola
29. New Orleans	W—BellSouth Mobility	On line 9/1/84	5	Motorola
	NW—Radiofone	On line 9/6/85	5	Motorola
30. Portland OR	W—GTE Mobilnet	On line 3/5/85	8	Motorola
	NW—Cellular One	On line 7/12/85	7	AT&T
31. Columbus OH	W—Ameritech Mobile	On line 5/30/85	5	NTI/GE
	NW—Cellular One	**CPG 1/28/85**		**Ericsson**
32. Hartford CT	W—Southern New England Tel.	On line 1/31/85	6	AT&T
	NW—Metro Mobile CTS	**CPG 2/14/85**		**Motorola**
33. San Antonio TX	W—Southwestern Bell Mobile	On line 1/28/85	12	AT&T
	NW—Cellular One	**CPG 1/30/85**		
34. Rochester NY	W—Rochester Telephone	On line 6/4/85	5	AT&T
	NW—Genesee Telephone Co.	**CPG 1/30/85**	**7**	**Ericsson**
35. Sacramento CA	W—PacTel Mobile Access	On line 8/29/85	5	NEC
	NW—Cellular One	**CPG 2/13/85**	**5**	
36. Memphis TN	W—BellSouth Mobility	On line 5/1/85	5	Motorola
	NW—Cellular One	**CPG 2/13/85**		**AT&T**
37. Louisville KY	W—BellSouth Mobility	On line 1/3/85	6	Motorola
	NW—Louisville Telephone	On line 2/15/85	5	AT&T

APPENDIX 1.2
Status of First 90 Cities (September 1986) (*Continued*)

MSA no./name	System operators	Status	No. of cells	Switching equipment
38. Providence	W—Nynex Mobile	On line 8/22/85	4	AT&T
RI	**NW—Metro Mobile CTS**	**CPG 9/21/84**	**8**	**Motorola**
39. Salt Lake City	W—NewVector Communications	On line 1/29/85	6	AT&T
UT	**NW—Cellular One**	**CPG 3/6/85**		
40. Dayton	W—Ameritech Mobile	On line 5/31/85	5	NTI/GE
OH	**NW—Cellular One**	**CPG 2/27/85**		**Ericsson**
41. Birmingham	W—BellSouth Mobility	On line 9/26/85	3	Motorola
AL	**NW—Birmingham Cellular Tel.**	**CPG 2/14/85**		
42. Bridgeport	W—Southern New England Tel.	On line 5/20/85	5	AT&T
CT	**NW—Connecticut Cellular Tele.**	**CPG 1/28/85**		**Motorola**
43. Norfolk	W—Contel Cellular, Inc.	On line 5/3/85	4	AT&T
VA	NW—Cellular One	On line 11/1/85	5	Motorola
44. Albany	W—Nynex Mobile	On line 6/25/85	4	NTI/GE
NY	**NW—Cellular System One**	**CPG 9/4/84**		
45. Oklahoma City	W—Southwestern Bell Mobile	On line 1/14/85	9	AT&T
OK	NW—Cellular One	On line 1/17/86	8	AT&T
46. Nashville	W—BellSouth Mobility	On line 6/10/85	8	Motorola
TN	**NW—Cellular One**	**CPG 1/30/85**		
47. Greensboro	W—Centel	On line 5/15/85	8	Motorola
NC	NW—Cellular One	On line 12/27/85	9	Motorola
48. Toledo	W—United TeleSpectrum	On line 7/25/85	9	Motorola
OH	NW—Cellular One	On line 4/15/86	7	Ericsson
49. New Haven	W—Southern New England Tel.	On line 3/4/85	6	AT&T
CT	**NW—Metro Mobile CTS**	**CPG 2/14/85**		**Motorola**
50. Honolulu	**W—GTE Mobilnet**	**CPG 3/26/84**	**4**	**Motorola**
HI	NW—Honolulu Cellular Tel.	On line 6/1/86	13	Ericsson
51. Jacksonville	W—BellSouth Mobility	On line 6/12/85	6	Motorola
FL	**NW—Cellular One**	**CPG 2/21/85**		
52. Akron	W—GTE Mobilnet	On line 10/31/85	4	Motorola
OH	**NW—Cellular One**	**CPG 2/13/85**		**NTI/GE**
53. Syracuse	W—Nynex Mobile	On line 1/24/86	3	NTI/GE
NY	NW—Cellular One	On line 12/31/85	4	Motorola
54. Gary	W—Ameritech Mobile	On line 3/11/85	3	AT&T
IN	NW—Cellular One	On line 4/21/86	2	Ericsson
55. Worcester	W—Nynex Mobile	On line 11/18/85	5	AT&T
MA	NW—Worcester Cellular Tel.	On line 11/18/85		
56. Northeast	W—Commonwealth Telephone	On line 7/2/85	8	NTI/GE
PA	NW—Cellular One	On line 12/31/85	3	NTI/GE

APPENDIX 1.2
Status of First 90 Cities (September 1986) (*Continued*)

MSA no./name	System operators	Status	No. of cells	Switching equipment
57. Tulsa	W—United States Cellular	On line 8/30/85	8	NEC
OK	NW—Cellular One	On line 5/22/86	10	Astronet
58. Allentown	W—Bell Atlantic Mobile	On line 3/18/85	32[a]	AT&T
PA	NW—Cellular One	On line 10/18/85	5	NTI/GE
59. Richmond	W—Contel Cellular, Inc.	On line 5/10/85	5	AT&T
VA	**NW—Cellular One**	**CPG 2/4/85**	**7**	**NEC**
60. Orlando	W—BellSouth Mobility	On line 2/27/85	4	Motorola
FL	**NW—Orlando Cellular Tel.**	**CPG 2/27/85**		
61. Charlotte	W—Alltel	On line 4/15/85	6	Motorola
NC	NW—Metro Mobile	On line 3/1/86	8	Motorola
62. **New Brunswick**	**W—Nynex Mobile**	**CPG 9/26/84**	**3**	**AT&T**
NJ	**NW—Cellular One**	**CPG 2/7/85**		**Motorola**
63. **Springfield**	**W—Nynex Mobile**	**CPG 4/19/84**		**AT&T**
MA	**NW—Springfield Cellular Tel.**	**CPG 1/30/85**		**Motorola**
64. **Grand Rapids**	**W—GTE Mobilnet**	**CPG 10/17/84**	**4**	**AT&T**
MI	**NW—Grand Rapids Cellular Tel.**	**CPG 1/30/85**	**5**	**Ericsson**
65. Omaha	W—Centel	On line 4/15/85	4	Motorola
NE	NW—Omaha Cellular Telephone	On line 12/23/85	3	CTI/EFJ
66. Youngstown	W—United TeleSpectrum	On line 9/19/85	2	Motorola
OH	NW—Cellular One	On line 12/23/85	3	Astronet
67. **Greenville**	**W—GTE Mobilnet**	**CPG 11/1/84**		**Motorola**
SC	**NW—Metro Mobile**	**CPG 2/21/85**		**Motorola**
68. Flint	W—Ameritech Mobile	On line 7/12/85	2	AT&T
MI	NW—Cellular One	On line 7/30/85	5	Ericsson
69. Wilmington	W—Bell Atlantic Mobile	On line 3/27/85	32[a]	AT&T
DE	**NW—Cellular One**	**CPG 1/30/85**	**7**	**Motorola**
70. **Long Branch**	**W—Nynex Mobile**	**CPG 9/26/84**	**3**	**AT&T**
NJ	**NW—Cellular One**	**CPG 1/30/85**	**4**	**Motorola**
71. Raleigh-Durham	W—United TeleSpectrum	On line 11/11/85	10	Motorola
NC	NW—Cellular One	On line 9/16/85	10	NTI/GE
72. W. Palm Beach	W—BellSouth Mobility	On line 5/23/85	23[b]	AT&T
FL	**NW—Cellular One**	**CPG 2/19/85**		
73. Oxnard	W—PacTel Mobile Access	On line 10/30/85	3	AT&T
CA	**NW—Oxnard Cellular Telephone**	**CPG 2/14/85**		
74. **Fresno**	**W—Contel Cellular, Inc.**	**CPG 10/22/84**	**3**	**AT&T**
CA	**NW—Fresno Cellular Telephone**	**CPG 2/26/85**		
75. Austin	W—GTE Mobilnet	On line 9/27/85	7	Motorola
TX	NW—Cellular One	On line 12/27/85	8	AT&T

APPENDIX 1.2
Status of First 90 Cities (September 1986) (*Continued*)

MSA no./name	System operators	Status	No. of cells	Switching equipment
76. New Bedford	W—Nynex Mobile	On line 12/9/85	2	AT&T
MA	**NW—New Bedford Cellular Tel.**	**CPG 2/13/85**		**Motorola**
77. Tucson	W—NewVector Communications	On line 8/6/85	4	NTI/GE
AZ	NW—Metro Mobile CTS	On line 4/1/86	4	Motorola
78. **Lansing**	**W—GTE Mobilnet**	**CPG 10/9/84**	**2**	**AT&T**
MI	**NW—Cellular One**	**CPG 2/21/85**	**6**	**Ericsson**
79. Knoxville	W—United States Cellular	On line 7/23/85	7	NEC
TN	**NW—Cellular One**	**CPG 2/7/85**		
80. Baton Rouge	W—BellSouth Mobility	On line 7/2/85	3	Motorola
LA	**NW—Cellular One**	**CPG 1/30/85**		
81. El Paso	W—Contel Cellular, Inc.	On line 2/25/85	2	AT&T
TX	NW—Metro Mobile	On line 5/2/86	4	Motorola
82. Tacoma	W—NewVector Communications	On line 4/18/85	3	NTI/GE
WA	NW—Cellular One	On line 12/12/85	17[e]	AT&T
83. Mobile	W—Contel Cellular, Inc.	On line 9/3/85	6	AT&T
AL	**NW—Bay Area Telephone Co.**	**CPG 1/30/85**		
84. Harrisburg	W—United TeleSpectrum	On line 10/18/85	4	Motorola
PA	NW—Cellular One	On line 9/18/85	4	NTI/GE
85. Johnson City	W—United TeleSpectrum	On line 10/3/85	6	Motorola
TN	**NW—Cellular One**	**CPG 2/1/85**		
86. Albuquerque	W—NewVector Communications	On line 8/13/85	2	NTI/GE
NM	NW—Metro Mobile	On line 11/1/85	3	Motorola
87. **Canton**	**W—GTE Mobilnet**	**CPG 10/17/84**	**2**	**Motorola**
OH	**NW—Canton Cellular Telephone**	**CPG 1/30/85**		**NTI/GE**
88. Chattanooga	W—BellSouth Mobility	On line 8/1/85	4	Motorola
TN	**NW—Cellular One**	**CPG 2/1/85**		
89. Wichita	W—Southwestern Bell Mobile	On line 2/11/85	4	Motorola
KS	NW—Cellular One	On line 1/24/86	2	AT&T
90. Charleston	W—United TeleSpectrum	On line 9/11/85	5	Motorola
SC	**NW—Charleston Cellular Tel.**	**CPG 1/28/85**		**Motorola**

[a]Includes Philadelphia, Allentown, PA, and Wilmington, DE.
[b]Includes San Francisco and San Jose, CA.
[c]Includes Washington, D.C., and Baltimore.
[d]Includes Miami and W. Palm Beach, FL.
[e]Includes Seattle and Tacoma, WA.
[f]Indicated in filing but no contract.
SOURCE: From *Cellular Business*, September 1986.

Chapter

2

Elements of Cellular Mobile Radio System Design

2.1 General Description of the Problem

Based on the concept of efficient spectrum utilization, the cellular mobile radio system design can be broken down into many elements, and each element can be analyzed and related to the others. The major elements are (1) the concept of frequency reuse channels, (2) the co-channel interference reduction factor, (3) the desired carrier-to-interference ratio, (4) the handoff mechanism, and (5) cell splitting. The purpose of this chapter is to introduce a simple methodology which will enable us to better understand how each element affects a cellular mobile radio system.

Since the limitation in the system is the frequency resource, the challenge is to serve the greatest number of customers with a specified system quality. We may ask ourselves three questions.

1. How many customers can we serve in a busy hour?
2. How many subscribers can we take into our system?
3. How many frequency channels do we need?

2.1.1 Maximum number of calls per hour per cell

To calculate the predicted number of calls per hour per cell Q in each cell, we have to know the size of the cell and the traffic conditions in the cell. The calls per hour per cell is based on how small the theoretical cell size can be. The control of the coverage of small cells is based on technological development.

We assume that the cell can be reduced to a 2-km cell, which means a cell of 2-km radius. A 2-km cell in some areas may cover many highways, and in other areas a 2-km cell may only cover a few highways.

Let a busy traffic area of 12 km radius fit seven 2-km cells. The

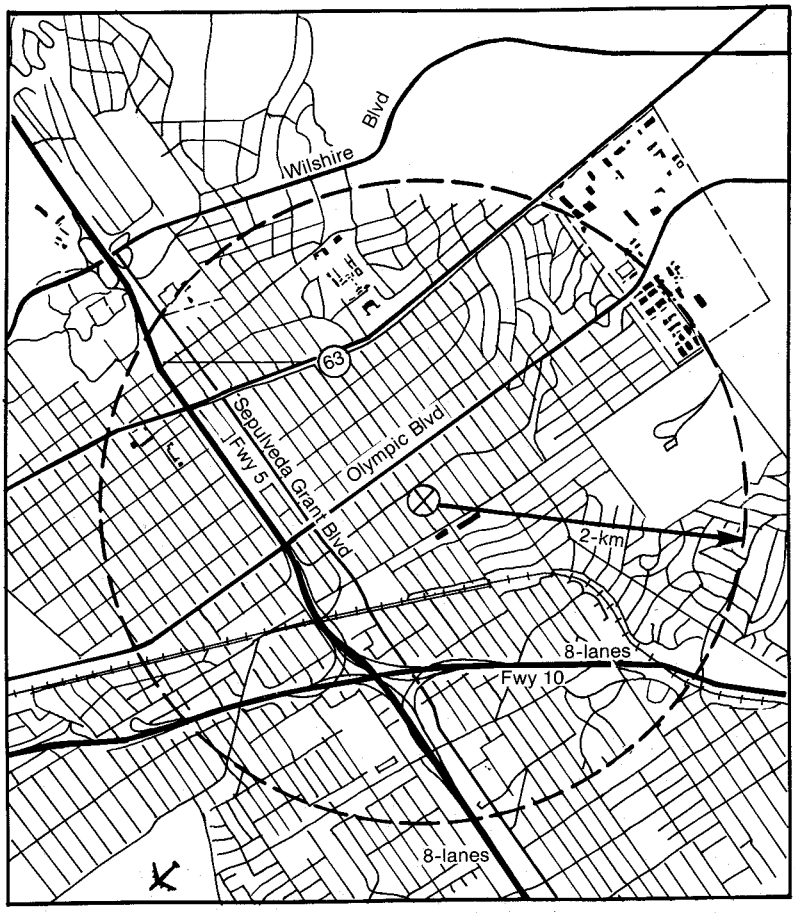

Figure 2.1 To establish the traffic capacity from a geographic map (west Los Angeles).

heaviest traffic cell may cover 4 freeways and 10 heavy traffic streets, as shown in Fig. 2.1. A total length of 64 km of 2 eight-lane freeways, 48 km of 2 six-lane freeways, and 588 km of 43 four-lane roads, including the 10 major roads, are obtained from Fig. 2.1. Assume that the average spacing between cars is 10 m during busy periods. We can determine that the total number of cars is about 70,000. If one-half the cars have car phones, and among them eight-tenths will make a call ($\eta_c = 0.8$) during the busy hour, there are 28,000 calls per hour, based on an average of one call per car if that car phone is used.

The maximum predicted number of calls per hour per a 2-km cell Q is derived from the above scenario. It may be an unrealistic case. However, it demonstrates how we can calculate Q for different scenarios and apply this method to finding the different Q in different geographic areas.

2.1.2 Maximum number of frequency channels per cell

The maximum number of frequency channels per cell N is closely related to an average calling time in the system. The standard user's calling habits may change as a result of the charging rate of the system and the general income profile of the users. If an average calling time T is 1.76 min and the maximum calls per hour per cell Q_i is obtained from Sec. 2.1.1, then the offered load can be derived as

$$A = \frac{Q_i T}{60} \quad \text{erlangs} \tag{2.1-1}$$

Assume that the blocking probability is given, then we can easily find the required number of radios in each cell.[1]

Example 2.1 Let the maximum calls per hour Q_i in one cell be 3000 and an average calling time T be 1.76 min. The blocking probability B is 2 percent. Then we may use Q from Eq. (2.1-1) to find the offered load A.

$$A = \frac{3000 \times 1.76}{60} = 88$$

With the blocking probability $B = 2$ percent, the maximum number of channels can be found from Appendix 1.1 as $N = 100$.

Example 2.2 If we let $Q_i = 28{,}000$ calls per cell per hour, based on one scenario shown in Sec. 2.1.1, $B = 2$ percent, and $T = 1.76$ min, how many radio channels are needed? The offered load A is obtained as

$$A = \frac{28{,}000 \times 1.76}{60} = 821$$

Inserting the above known figures into the table of Appendix 1.1, we find that $N = 820$ channels per cell.

Example 2.3 If there are 50 channels in a cell to handle all the calls and the average is 100 s per call, how many calls can be handled in this cell with a blocking probability of 2 percent? Since $N = 50$ and $B = 2$ percent, the offered load can be found from Appendix 1.1 as

$$A = 40.3$$

The number of calls per hour in a cell is

$$Q_i = \frac{40.3 \times 3600}{100} = 1451 \text{ calls per hour}$$

Example 2.4 If the maximum number of calls per hour per cell is 1451 and there is a seven-cell reuse pattern* in the system ($K = 7$), and assuming that $B = 2$ percent and $T = 100$ s as in Example 2.3, then $N = 50$ as indicated. The total number of required channels for a $K = 7$ reuse system is

$$N_t = 50 \times 7 = 350 \text{ radios}$$

If a large area is covered by 28 cells, $K_t = 28$; the total number of customers $M_t = \Sigma_{i=1}^{K_t} M_i$ in the system increases. Therefore, we may assume that the number of subscribers per cell M_i is somehow related to the percentage of car phones used in the busy hours (η_c) and the number of calls per hour per cell Q_i as

$$M_i = f(Q_i, \eta_c) \tag{2.1-2}$$

where the value Q_i is a function of the blocking probability B, the average calling time T, and the number of channels N.

$$Q_i = f(B, T, N) \tag{2.1-3}$$

If the $K = 7$ frequency reuse pattern is used, the total number of required channels in the system is $N_t = 7 \times N$. We must realize that it is the maximum number of calls per cell Q_i that determines the total required channels N_t, not the total number of subscribers M_t. In this case ($K_t = 30$ and $K = 7$), the total number of channels N_t has been used four times in the system.

2.2 Concept of Frequency Reuse Channels

A radio channel consists of a pair of frequencies, one for each direction of transmission that is used for full-duplex operation. A particular radio channel, say F_1, used in one geographic zone to call a cell, say C_1, with

* Its pattern is shown in Fig. 2.3 and described in Sec. 2.2.1.

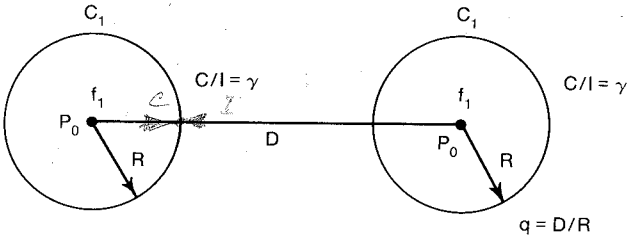

Figure 2.2 The ratio of D/R.

a coverage radius R can be used in another cell with the same coverage radius at a distance D away.

Frequency reuse is the core concept of the cellular mobile radio system. In this frequency reuse system, users in different geographic locations (different cells) may simultaneously use the same frequency channel (see Fig. 2.2). The frequency reuse system can drastically increase the spectrum efficiency, but if the system is not properly designed, serious interference may occur. Interference due to the common use of the same channel is called *cochannel interference* and is our major concern in the concept of frequency reuse.

2.2.1 Frequency reuse schemes

The frequency reuse concept can be used in the time domain and the space domain. Frequency reuse in the time domain results in the occupation of the same frequency in different time slots. It is called *time-division multiplexing* (TDM). Frequency reuse in the space domain can be divided into two categories.

1. Same frequency assigned in two different geographic areas, such as AM or FM radio stations using the same frequency in different cities.
2. Same frequency repeatedly used in a same general area in one system[2]—the scheme is used in cellular systems. There are many cochannel cells in the system. The total frequency spectrum allocation is divided into K frequency reuse patterns, as illustrated in Fig. 2.3 for $K = 4, 7, 12$, and 19.

2.2.2 Frequency reuse distance[2, 3, 7, 8]

The minimum distance which allows the same frequency to be reused will depend on many factors, such as the number of cochannel cells in the vicinity of the center cell, the type of geographic terrain contour, the antenna height, and the transmitted power at each cell site.

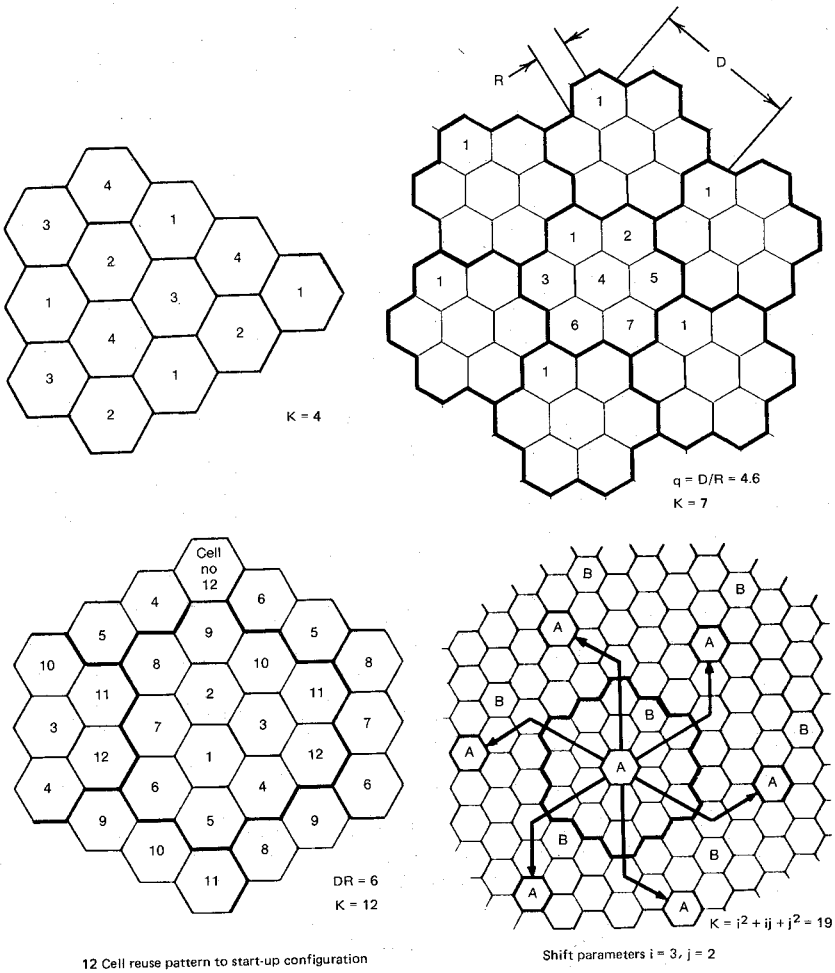

Figure 2.3 N-cell reuse pattern.

The frequency reuse distance D can be determined from

$$D = \sqrt{3K}\,R \tag{2.2-1}$$

Where K is the frequency reuse pattern shown in Fig. 2.2, then

$$D = \begin{cases} 3.46R & K = 4 \\ 4.6R & K = 7 \\ 6R & K = 12 \\ 7.55R & K = 19 \end{cases}$$

If all the cell sites transmit the same power, then K increases and the frequency reuse distance D increases. This increased D reduces the chance that cochannel interference may occur.

Theoretically, a large K is desired. However, the total number of allocated channels is fixed. When K is too large, the number of channels assigned to each of K cells becomes small. It is always true that if the total number of channels in K cells is divided as K increases, trunking inefficiency results. The same principle applies to spectrum inefficiency: if the total number of channels are divided into two network systems serving in the same area, spectrum inefficiency increases, as mentioned in Sec. 1.3.

Now the challenge is to obtain the smallest number K which can still meet our system performance requirements. This involves estimating cochannel interference and selecting the minimum frequency reuse distance D to reduce cochannel interference. The smallest value of K is $K = 3$, obtained by setting $i = 1, j = 1$ in the equation $K = i^2 + ij + j^2$ (see Fig. 2.3).

2.2.3 Number of customers in the system

When we design a system, the traffic conditions in the area during a busy hour are some of the parameters that will help determine both the sizes of different cells and the number of channels in them.

The maximum number of calls per hour per cell is driven by the traffic conditions at each particular cell. After the maximum number of frequency channels per cell has been implemented in each cell, then the maximum number of calls per hour can be taken care of in each cell. Now, take the maximum number of calls per hour in each cell Q_i and sum them over all cells. Assume that 60 percent of the car phones will be used during the busy hour, on average, one call per phone ($\eta_c = 0.6$) if that phone is used. The total allowed subscriber traffic M_t can then be obtained.

Example 2.5 During a busy hour, the number of calls per hour Q_i for each of 10 cells is 2000, 1500, 3000, 500, 1000, 1200, 1800, 2500, 2800, 900. Assume that 60 percent of the car phones will be used during this period ($\eta_c = 0.6$) and that one call is made per car phone. Summing over all Q_i gives the total Q_t

$$Q_t = \sum_{i=1}^{10} Q_i = 17{,}200 \text{ calls per hour}$$

Since $\eta_c = 0.6$, the number of customers in the system is

$$M_t = \frac{17{,}200}{0.6} = 28{,}667$$

2.3 Cochannel Interference Reduction Factor

Reusing an identical frequency channel in different cells is limited by cochannel interference between cells, and the cochannel interference

can become a major problem. Here we would like to find the minimum frequency reuse distance in order to reduce this cochannel interference.

Assume that the size of all cells is roughly the same. The cell size is determined by the coverage area of the signal strength in each cell. As long as the cell size is fixed, cochannel interference is independent of the transmitted power of each cell. It means that the received threshold level at the mobile unit is adjusted to the size of the cell. Actually, cochannel interference is a function of a parameter q defined as

$$q = \frac{D}{R} \tag{2.3-1}$$

The parameter q is the cochannel interference reduction factor. When the ratio q increases, cochannel interference decreases. Furthermore, the separation D in Eq. (2.3-1) is a function of K_I and C/I,

$$D = f(K_I, C/I) \tag{2.3-2}$$

where K_I is the number of cochannel interfering cells in the first tier and C/I is the received carrier-to-interference ratio at the desired mobile receiver.[3]

$$\frac{C}{I} = \frac{C}{\sum_{k=1}^{K_I} I_k} \tag{2.3-3}$$

In a fully equipped hexagonal-shaped cellular system, there are always six cochannel interfering cells in the first tier, as shown in Fig. 2.4; that is, $K_I = 6$. The maximum number of K_I in the first tier can be shown as six (i.e., $2\pi D/D \approx 6$). Cochannel interference can be experienced both at the cell site and at mobile units in the center cell. If the interference is much greater, then the carrier-to-interference ratio C/I at the mobile units caused by the six interfering sites is (on the average) the same as the C/I received at the center cell site caused by interfering mobile units in the six cells. According to both the reciprocity theorem and the statistical summation of radio propagation, the two C/I values can be very close. Assume that the local noise is much less than the interference level and can be neglected. C/I then can be expressed, from Eq. (1.6-4), as

$$\frac{C}{I} = \frac{R^{-\gamma}}{\sum_{k=1}^{K_I} D_k^{-\gamma}} \tag{2.3-4}$$

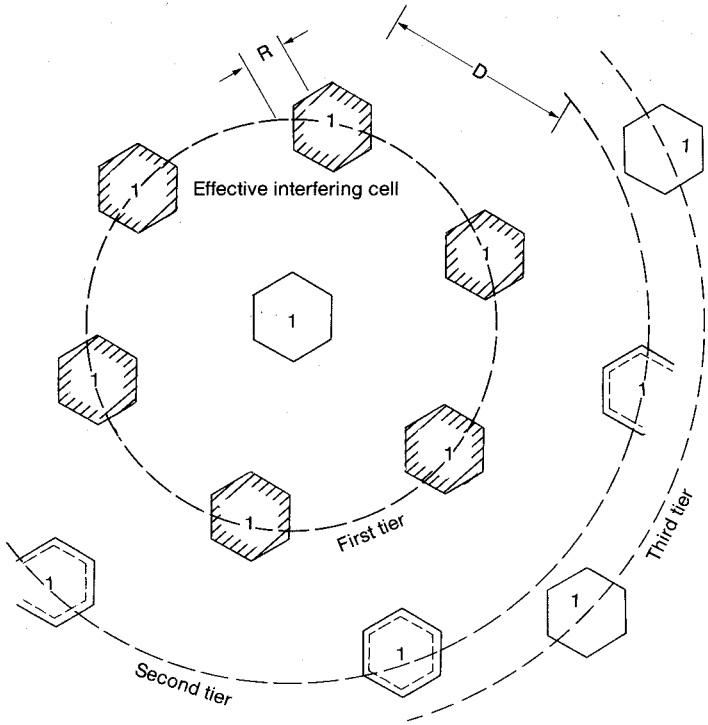

Figure 2.4 Six effective interfering cells of cell 1.

where γ is a propagation path-loss slope[5] determined by the actual terrain environment. In a mobile radio medium, γ usually is assumed to be 4 (see Sec. 1.6). K_I is the number of cochannel interfering cells and is equal to 6 in a fully developed system, as shown in Fig. 2.4. The six cochannel interfering cells in the second tier cause weaker interference than those in the first tier (see Example 2.6 at the end of Sec. 2.4.1).

Therefore, the cochannel interference from the second tier of interfering cells is negligible. Substituting Eq. (2.3-1) into Eq. (2.3-4) yields

$$\frac{C}{I} = \frac{1}{\sum_{k=1}^{K_I} \left(\frac{D_k}{R}\right)^{-\gamma}} = \frac{1}{\sum_{k=1}^{K_I} (q_k)^{-\gamma}} \tag{2.3-5}$$

where q_k is the cochannel interference reduction factor with kth cochannel interfering cell

$$q_k = \frac{D_k}{R} \tag{2.3-6}$$

2.4 Desired C/I from a Normal Case in an Omnidirectional Antenna System

2.4.1 Analytic solution

There are two cases to be considered: (1) the signal and cochannel interference received by the mobile unit and (2) the signal and cochannel interference received by the cell site. Both cases are shown in Fig. 2.5. N_m and N_b are the local noises at the mobile unit and the cell site, respectively. Usually N_m and N_b are small and can be neglected as compared with the interference level. The effect of the cochannel interference on spectrum efficiency systems will appear in Sec. 13.4. As long as the received carrier-to-interference ratios at both the mobile unit and the cell site are the same, the system is called a *balanced system*. In a balanced system, we can choose either one of the two cases to analyze the system requirement; the results from one case are the same for the others.

Assume that all D_k are the same for simplicity, as shown in Fig. 2.4; then $D = D_k$, and $q = q_k$, and

$$\frac{C}{I} = \frac{R^{-\gamma}}{6D^{-\gamma}} = \frac{q^{\gamma}}{6} \tag{2.4-1}$$

Thus

$$q^{\gamma} = 6\frac{C}{I} \tag{2.4-2}$$

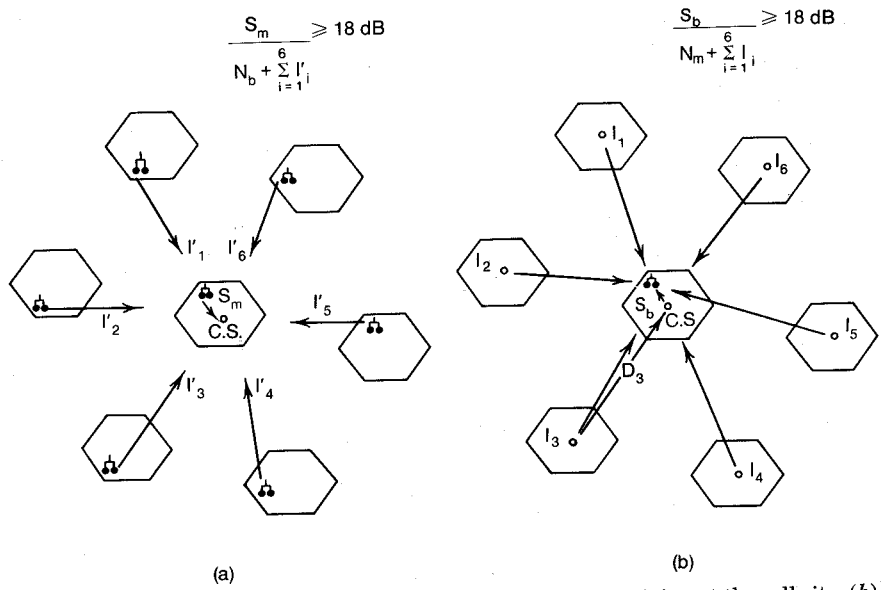

Figure 2.5 Cochannel interference from six interferers. (*a*) Receiving at the cell site; (*b*) receiving at the mobile unit.

and
$$q = \left(6\frac{C}{I}\right)^{1/\gamma} \tag{2.4-3}$$

In Eq. (2.4-3), the value of C/I is based on the required system performance and the specified value of γ is based on the terrain environment. With given values of C/I and γ, the cochannel interference reduction factor q can be determined. Normal cellular practice is to specify C/I to be 18 dB or higher based on subjective tests and the criterion described in Sec. 1.5. Since a C/I of 18 dB is measured by the acceptance of voice quality from present cellular mobile receivers, this acceptance implies that both mobile radio multipath fading and cochannel interference become ineffective at that level. The path-loss slope γ is equal to about 4 in a mobile radio environment.[5]

$$q = D/R = (6 \times 63.1)^{1/4} = 4.41 \tag{2.4-4}$$

The 90th percentile of the total covered area would be achieved by increasing the transmitted power at each cell; increasing the same amount of transmitted power in each cell does not affect the result of Eq. (2.4-4). This is because q is not a function of transmitted power. The computer simulation described in the next section finds the value of $q = 4.6$, which is very close to Eq. (2.4-4). The factor q can be related to the finite set of cells K in a hexagonal-shaped cellular system by

$$q \triangleq \sqrt{3K} \tag{2.4-5}$$

Substituting q from Eq. (2.4-4) into Eq. (2.4-5) yields

$$K = 7 \tag{2.4-6}$$

Equation (2.4-6) indicates that a seven-cell reuse pattern* is needed for a C/I of 18 dB. The seven-cell reuse pattern is shown in Fig. 2.3.

Based on $q = D/R$, the determination of D can be reached by choosing a radius R in Eq. (2.4-4). Usually, a value of q greater than that shown in Eq. (2.4-4) would be desirable. The greater the value of q, the lower the cochannel interference. In a real environment, Eq. (2.3-5) is always true, but Eq. (2.4-1) is not. Since Eq. (2.4-4) is derived from Eq. (2.4-1), the value q may not be large enough to maintain a carrier-to-interference ratio of 18 dB. This is particularly true in the worst case, as shown in Chap. 6.

* In this seven-cell reuse pattern, the total allocated frequency band is divided into seven subsets. Each particular subset of frequency channels is assigned to one of seven cells.

Example 2.6 Compare interference from the first tier of six interferers with that from twelve interferers (first and second tiers) (see Fig. 2.4).

From the first tier,

$$\frac{C}{I} = \frac{C}{\sum_{i=1}^{6} I_i} = \frac{R_1^4}{6D_1^{-4}} = \frac{a_1^4}{6} \tag{E2.6-1}$$

From the first and second tiers,

$$\frac{C}{I} = \frac{C}{\sum_{i=1}^{6} (I_{1i} + I_{2i})} = \frac{1}{6(a_1^{-4} + a_2^{-4})} \tag{E2.6-2}$$

Since we have found $a_1 = 4.6$, then from the second tier, $a_2 = D_2/R_1 \approx 2D_1/R_1 = 2a_1 = 9.2$. Substituting a_1 and a_2 into Eqs. (E2.6-1) and (E2.6-2), respectively, yields

$$\left(\frac{C}{I}\right)_{\text{1st tier}} = 18.72 \text{ dB}$$

$$\left(\frac{C}{I}\right)_{\text{1st and 2nd tiers}} = 18.46 \text{ dB}$$

We realize that a negligible amount of interference is contributed by the six interferers from the second tier.

2.4.2 Solution obtained from simulation

The required cochannel reduction factor q can be obtained from the simulation also. Let one main cell site and all six possible cochannel interferers be deployed in a pattern, as shown in Fig. 2.4. The distance D from the center cell to the cochannel interferers in the simulation is a variable. $D = 2R$ can be used initially and incremented every $0.5R$ as $D = 2R, 2.5R, 3R$. For every particular value of D, a set of simulation data is generated.

First, the location of each mobile unit in its own cell is randomly generated by a random generator. Then the distance D_k from each of the six interfering mobile units to the center cell site (assuming $K_I = 6$) is obtained. The desired mobile signal as well as six interference levels received at the center cell site would be randomly generated following the mobile radio propagation path-loss rule, which is 40 dB/dec, along with a log-normal standard deviation of 8 dB at its mean value. Summing up all the data from six simulated interferences,

$$I = \sum_{k=1}^{K_I=6} I_k$$

and dividing it by the simulated main carrier, value C becomes C/I.

This C/I is for a particular D, the distance between the center cell site and the cochannel cell sites (cochannel interferers). Repeat this process, say 1000 times, for each particular value of D, based on the criterion stated in Sec. 1.5 (that 75 percent of the users say voice quality is "good" or "excellent" in 90 percent of the total covered area). Then from 75 percent of the users' opinion, $C/I = 18$ dB needs to be achieved[4] with a proper value of D. Assuming that mobile unit locations are chosen randomly and uniformly, then 90 percent of the area corresponds to 900 out of 1000 mobile unit locations.

To find a proper value for D, each mobile unit location associates with its received C/I. Some C/I values are high and some are low. This means that the lowest 100 values of C/I should be discarded. The main C/I value should be derived from the remaining 900 C/I values. This associates a particular C/I for a particular separation D. Repeating this process for different values of D, the corresponding mean C/I values are found. The C/I versus D curve can be plotted, depicting $C/I = 18$ dB as corresponding to $D = 4.6R$, as illustrated in the Bell Lab publication.[4] Then

$$q = \frac{D}{R} = 4.6 \tag{2.4-7}$$

Comparing the values of q obtained from an analytic solution shown in Eq. (2.4-4) and q obtained from a simulation solution shown in Eq. (2.4-7), the results are surprisingly close.

Although a simulation (statistical) approach deals with a real-world situation, it does not provide a clear physical picture. The two agreeable solutions illustrated in this section prove that the simple analytic method is implementable in a cellular system based on hexagonal cells.

2.5 Handoff Mechanism

The handoff is the process mentioned in Sec. 1.7. It is a unique feature that allows cellular systems to operate as effectively as demonstrated in actual use. To clearly describe the handoff concept, it is easy to use a one-dimensional illustration as shown in Fig. 2.6, although a real two-dimensional cellular configuration would cover an area with cells. The handoff concept as applied to a one-dimensional case will also apply to two-dimensional cases.

Two cochannel cells using the frequency F_1 separated by a distance D are shown in Fig. 2.6a. The radius R and the distance D are governed by the value of q. Now we have to fill in with other frequency channels such as F_2, F_3, and F_4 between two cochannel cells in order to provide a communication system in the whole area.

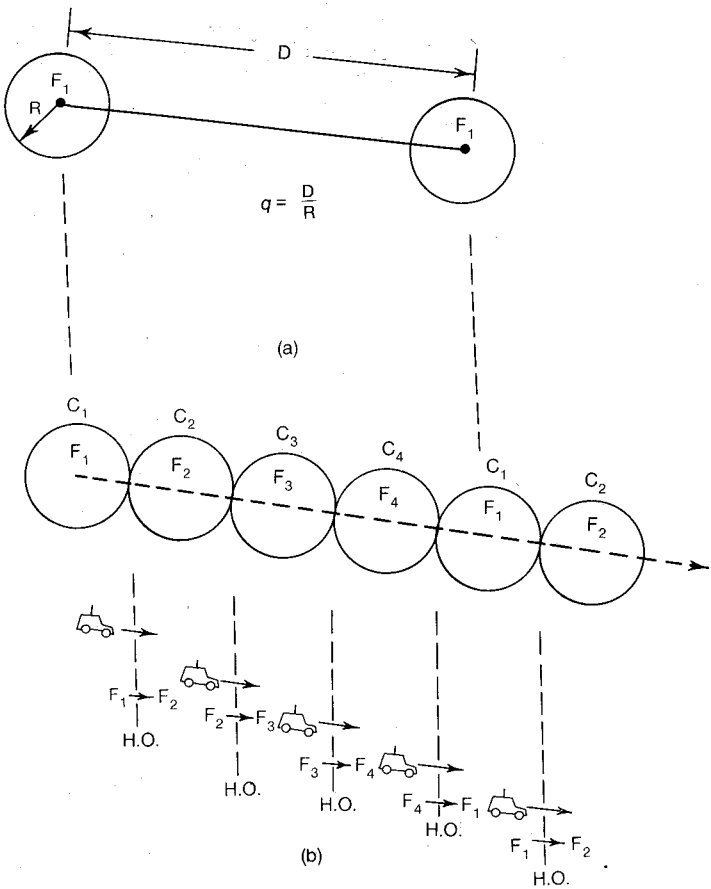

Figure 2.6 Handoff mechanism. (*a*) Cochannel interference reduction ratio q. (*b*) Fill-in frequencies.

The fill-in frequencies F_2, F_3, and F_4 are also assigned to their corresponding cells C_2, C_3, and C_4 (see Fig. 2.6*b*) according to the same value of q.

Suppose a mobile unit is starting a call in cell C_1 and then moves to C_2. The call can be dropped and reinitiated in the frequency channel from F_1 to F_2 while the mobile unit moves from cell C_1 to cell C_2. This process of changing frequencies can be done automatically by the system without the user's intervention. This process of handoff is carried on in the cellular system.

The handoff processing scheme is an important task for any successful mobile system. How does one make any one of the necessary handoffs successful? How does one reduce all unnecessary handoffs in the system? How is the individual cell traffic capacity controlled by

altering the handoff algorithm? All these questions will be answered in Chap. 9.

2.6 Cell Splitting

2.6.1 Why splitting?

The motivation behind implementing a cellular mobile system is to improve the utilization of spectrum efficiency.[5, 9] The frequency reuse scheme is one concept, and cell splitting is another concept. When traffic density starts to build up and the frequency channels F_i in each cell C_i cannot provide enough mobile calls, the original cell can be split into smaller cells. Usually the new radius is one-half the original radius (see Fig. 2.7). There are two ways of splitting: In Fig. 2.7*a*, the original cell site is not used, while in Fig. 2.7*b*, it is.

$$\text{New cell radius} = \frac{\text{old cell radius}}{2} \tag{2.6-1}$$

Then based on Eq. (2.6-1), the following equation is true.

$$\text{New cell area} = \frac{\text{old cell area}}{4} \tag{2.6-2}$$

Let each new cell carry the same maximum traffic load of the old cell; then, in theory,

$$\frac{\text{New traffic load}}{\text{Unit area}} = 4 \times \frac{\text{traffic load}}{\text{unit area}}$$

2.6.2 How splitting?

There are two kinds of cell-splitting techniques:

1. *Permanent splitting.* The installation of every new split cell has to be planned ahead of time; the number of channels, the transmitted power, the assigned frequencies, the choosing of the cell-site selection, and the traffic load consideration should all be considered. When ready, the actual service cut-over should be set at the lowest traffic point, usually at midnight on a weekend. Hopefully, only a few calls will be dropped because of this cut-over, assuming that the downtime of the system is within 2 h.
2. *Dynamic splitting.* This scheme is based on utilizing the allocated spectrum efficiency in real time. The algorithm for dynamically splitting cell sites is a tedious job since we cannot afford to have

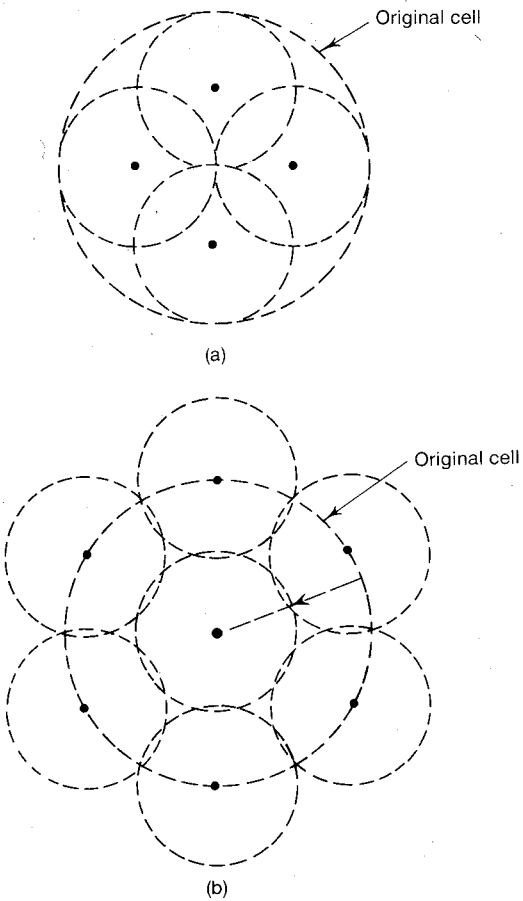

Figure 2.7 Cell splitting.

one single cell unused during cell splitting at heavy traffic hours. Section 10.4.2 will discuss this topic in depth.

2.7 Consideration of the Components of Cellular Systems

The elements of cellular mobile radio system design have been mentioned in the previous sections. Here we must also consider the components of cellular systems, such as mobile radios, antennas, cell-site controller, and MTSO. They would affect our system design if we do not choose the right one. The general view of the cellular system is shown in Fig. 2.8. Even though the EIA (Electronic Industries Association) and the FCC have specified standards for radio equipment at the cell sites and the mobile sites, we still need to be concerned about

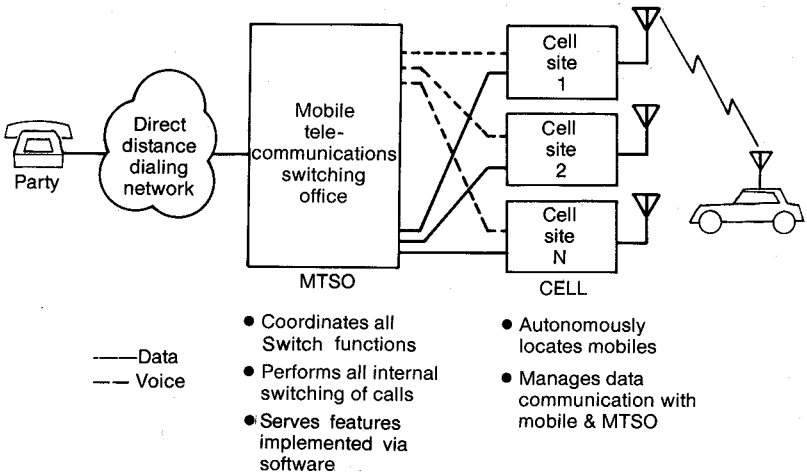

Figure 2.8 A general view of cellular telecommunications systems.

that equipment. The issues affecting choice of antennas, switching equipment, and data links are briefly described here.

2.7.1 Antennas

Antenna pattern, antenna gain, antenna tilting, and antenna height[6] all affect the cellular system design. The antenna pattern can be omnidirectional, directional, or any shape in both the vertical and the horizon planes. Antenna gain compensates for the transmitted power. Different antenna patterns and antenna gains at the cell site and at the mobile units would affect the system performance and so must be considered in the system design.

The antenna patterns seen in cellular systems are different from the patterns seen in free space. If a mobile unit travels around a cell site in areas with many buildings, the omnidirectional antenna will not duplicate the omnipattern. In addition, if the front-to-back ratio of a directional antenna is found to be 20 dB in free space, it will be only 10 dB at the cell site. An explanation for these phenomena is given in Chapter 5.

Antenna tilting can reduce the interference to the neighboring cells and enhance the weak spots in the cell. Also the height of the cell-site antenna can affect the area and shape of the coverage in the system. The effect of antenna height will be described in Chap. 4.

2.7.2 Switching equipment

The capacity of switching equipment in cellular systems is not based on the number of switch ports but on the capacity of the processor

associated with the switches. In a big cellular system, this processor should be large. Also, because cellular systems are unlike other systems, it is important to consider when the switching equipment would reach the maximum capacity.

The service life of the switching equipment is not determined by the life cycle of the equipment but by how long it takes to reach its full capacity. If the switching equipment is designed in modules, or as distributed switches, more modules can be added to increase the capacity of the equipment. For decentralized systems digital switches may be more suitable. The future trend seems to be the utilization of system handoff. This means that switching equipment can link to other switching equipment so that a call can be carried from one system to another system without the call being dropped. We will discuss these issues in Chap. 11.

2.7.3 Data links

The data links are shown in Fig. 2.8. Although they are not directly affected by the cellular system, they are important in the system. Each data link can carry multiple channel data (10 kbps data transmitted per channel) from the cell site to the MTSO. This fast-speed data transmission cannot be passed through a regular telephone line. Therefore, data bank devices are needed. They can be multiplexed, many-data channels passing through a wideband T-carrier wire line or going through a microwave radio link where the frequency is much higher than 850 MHz.

Leasing T1-carrier wire lines through telephone companies can be costly. Although the use of microwaves may be a long-term money saver, the availability of the microwave link has to be considered and is described in Chap. 12. The arrangement of data links will be described in Chap. 11.

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Chapter

3

Specifications

In this chapter, we concentrate on U.S. cellular mobile specifications.[1] Also, we touch on the differences in other foreign cellular mobile systems.

3.1 Definitions of Terms and Functions

1. *Home mobile station (unit).* A mobile station that is subscribed in its cellular system.
2. *Land station.* A station other than a mobile station, which links to the mobile station.
3. *Control channel.* A channel used for the transmission of digital control information from a land station to a mobile station, or vice versa.
4. *Forward control channel* (FDCC). A control channel used from a land station to a mobile station.
5. *Reverse control channel* (RECC). A control channel used from a mobile station to a land station.
6. *Forward voice channel* (FVC). A voice channel used from a land station to a mobile unit.
7. *Reverse voice channel* (RVC). A voice channel used from a mobile station to a land station.

8. *Set-up channels.* A number of designated control channels.
9. *Access channel.* A control channel used by a mobile station to access a system and obtain service. The access channel always accesses from the mobile station to the cell site.
10. *Paging channel.* The act of seeking a mobile station when an incoming call from the land line has been placed to it.
11. *Digital color code* (DCC). A digital signal transmitted by a forward control channel to detect capture of an interfering mobile station. There are four codes. (See Sec. 3.2.8.)
12. *Flash request.* A message sent on a voice channel from a mobile station to a land station indicating a user's desire to invoke special processing, such as an emergency.
13. *Signaling tone.* A 10-kHz tone transmitted by the mobile station on a voice channel. It serves several functions.
14. *Handoff.* The act of transferring a mobile station from one voice channel to another voice channel. There are two kinds of handoffs:
    a. Interhandoff, from one cell to another cell
    b. Intrahandoff, within a cell
15. *Numeric information.* Used to describe the operation of the mobile station.

Numeric indicators

MIN	mobile identification number
MIN1	24 bits that correspond to the seven-digit directory number assigned to the mobile station
MIN2	10 bits that correspond to the three-digit area code
BIS	identifies whether a mobile station must check an idle-to-busy transition on a reverse control channel when accessing a system. In a forward control channel busy-idle bit inserts in every 10-bit interval of a transmitted bit stream.
CCLIST	scanned by a mobile station on a list of control channels
CMAX	maximum number of control channels to be scanned by the mobile station (up to 21 channels)
MAXBUSY	maximum number of busy occurrences allowed on a reversed control channel
MAXSZTR	maximum number of seizure attempts allowed on a reversed control channel
NBUSY	number of times a mobile station attempts to seize a reverse control channel and finds it busy
NSZTR	number of times a mobile station attempts to seize a reverse control channel and fails
PL	mobile station RF power level

SCC — a digital number that is stored and used to identify which SAT (see item 20 below) frequency a mobile station should be received on

16. *Paging.* The act of seeking a mobile station when an incoming call from the land station has been placed to it.
17. *Paging channel.* A forward control channel which is used to page mobile stations and send orders.
18. *Registration.* The procedure by which a mobile station identifies itself to a land station as being active.
19. *Roamer.* A mobile station which operates in a cellular system other than the one from which service is subscribed.
20. *Supervisory audio tone* (SAT). One of three tones in the 6-kHz region; there is one SAT frequency for each land station. In certain circumstances, there is one SAT frequency for each sector of each land station.
21. *System identification* (SID). A digital identification uniquely associated with a cellular system.
22. *Electronic serial number* (ESN). Each mobile station has an ESN assigned by the manufacturer.
23. *Group identification.* A subset of the most significant bits of SID that is used to identify a group of cellular systems, such as NYNEX systems, PacTel systems, Southwestern Bell systems.
24. *Channel spacing.* 30 kHz per one-way channel. As an example, channel 1 is 825.030 MHz (mobile transmit) and 870.030 MHz (land transmit). Additional spectrum allocation of 10 MHz for the cellular industry changes the channel numbering order.

3.2 Specification of Mobile Station (Unit) in the United States[1–3]

3.2.1 Power

Let P_0 be the specified power and f_0 be the specified frequency channel. P and f are the operating power and frequency, respectively.

Power level (carrier-off condition) requires $P < -60$ dBm in 2 ms

Power level (carrier-on condition) within 3 dB of specified power (P_0) within 2 ms

Power level (off-frequency condition), if $|f - f_c| > 1$ kHz, do not transmit; then $P < -60$ dBm

Power transmitted levels are maximum effective radiated power (ERP) with respect to a half-wave dipole

Mobile stations have three station class marks.

Power class	P, power level = 0	Tolerance
I	6 dBW (4.0 W)	(8 dBW $\leq P \leq$ 2 dBW)
II	2 dBW (1.6 W)	(4 dBW $\leq P \leq$ −2 dBW)
III	−2 dBW (0.6 W)	(0 dBW $\leq P \leq$ −6 dBW)

Each mobile station power class I has eight full power levels (0 to 7), with power level 0 being the highest. Each level has a 4-dB drop. The total power control range for power class I (mobile) is 28 dB. The names CMAC and VMAC indicate the maximum control and maximum voice attenuation codes, respectively. For all three mobile station power classes, power level 7 is −22 dBW (or 8 dBm or 6.3 mW).

3.2.2 Modulation

1. *Compressor/expandor (compander).* A 2:1 syllabic compander is used. Every 2-dB change in input level converts (compresses) to 1 dB at output (at the transmitted side). Then reverse the two numbers (expand) at the received side. It serves two purposes:
   a. To confine the energy in the channel bandwidth
   b. To generate a quieting effect during a speech pulse
2. *Preemphasis/deemphasis* (see Fig. 3.1). The preemphasis network and its response are shown in Fig. 3.1. The improvement factor ρ_{FM} is[4]

$$\rho_{FM} = \frac{(f_2/f_1)^3}{3[f_2/f_1 - \tan^{-1}(f_2/f_1)]}$$

For $f_2/f_1 < 2$, ρ_{FM} approaches 1.

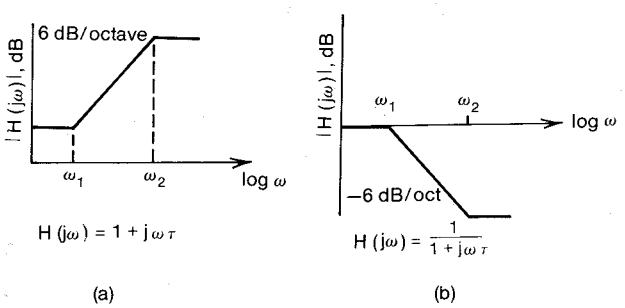


Figure 3.1 Preemphasis/deemphasis response. (*a*) Preemphasis response $\omega_1 = 1/\tau$, $\omega_2 = 2\pi f_2$. (*b*) Deemphasis response $\omega_1 = 1/\tau$, $\omega_2 = 2\pi f_2$.

For wideband, $f_2 \gg f_1$. Thus

$$\rho_{FM} = \frac{(f_2/f_1)^2}{3}$$

For $f_2 = 3$ kHz and $f_1 = 300$ Hz, the improvement factor ρ_{FM} is

$$\rho_{FM} = 10 \log \frac{100}{3} = 15.23 \text{ dB}$$

3. *Deviation limiter.* A mobile station must limit the instantaneous frequency deviation to ±12 kHz. The deviation limiter of the frequency-to-voltage characteristic is shown in Fig. 3.2.
4. *Wideband data signal.* A NRZ (non-return-to-zero) bindary data stream is encoded to a Manchester (biphase) code, as shown in Fig. 3.3. It modulates to a ±8-kHz binary FSK with a transmission rate of 10 kbps. The advantage of using a Manchester code in a voice channel is that the energy of this code is concentrated at the transmission rate of 10 kHz. Therefore, a burst of signals transmitted over the voice channel can be detected. The Manchester code is applied to both control channels and voice channels.

3.2.3 Limitation on emission

It is very important that each mobile station have a limit on its emission. An RF signal emitted by any receiver has to be less than some value, which in turn depends on different frequency bands. The values are shown in the box on the following page.

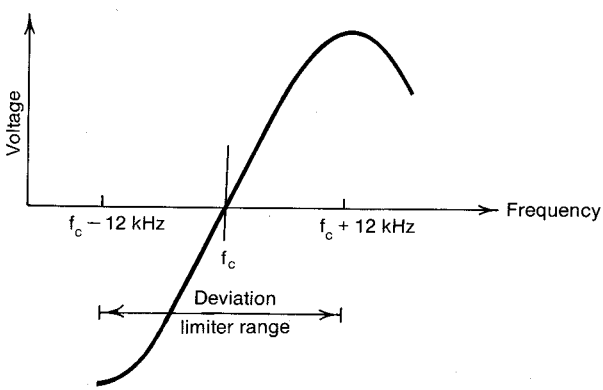


Figure 3.2 Deviation limiter characteristics.

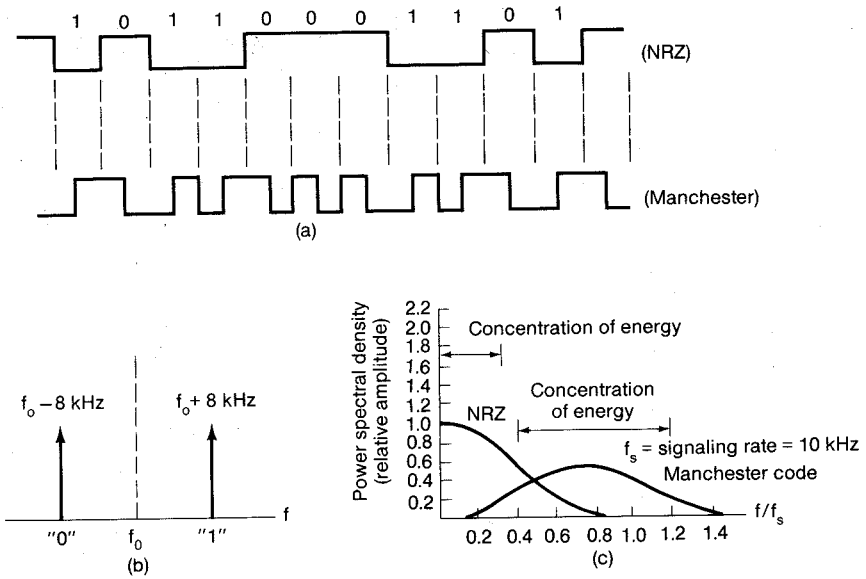
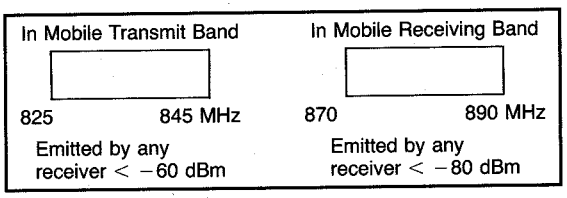


Figure 3.3 Waveforms and power spectral densities of NRZ and Manchester code. (*a*) Waveforms of NRZ and Manchester code. (*b*) Use of two frequency deviations to represent two levels. (*c*) Power spectral densities of NRZ and Manchester code.



3.2.4 Security and identification

1. *Mobile identification number* (MIN). A binary number of 34 bits ($2^{34} \approx 1.7 \times 10^{10}$) derived from a 10-digit directory telephone number.

2. *Electronic serial number* (ESN). A 32-bit binary number that uniquely identifies a mobile unit and must be set by the factory. Attempts to change the serial number circuit should render the mobile unit *inoperative*.

   The manufacturer should not overlook this requirement, and each operating system has to check that this requirement is enforced from different manufactured units. This number will be used in the system for security purposes.

   An amount of revenue can be lost to the system operation if the manufacturer fails to implement this requirement in the mobile unit and a false ESN can be used in place easily.

3. *First paging channel* (FIRSTCHP). An 11-bit system which iden-

tifies the channel number of the first paging channel when the mobile station is "home." It is stored in the mobile unit.

4. *Home system identification* (SID). Fifteen-bit system used to identify the home station. The least significant bit is 1 for a Block A system, otherwise 0 for a Block B system.
5. *Preferred system selection.* Provided as a means for selecting the preferred system as either system A or system B within a mobile station.

3.2.5 Supervision

SAT (supervisory audio tone)

1. *SAT function.* There are three SAT tones: 5980, 6000, and 6030 Hz. The tolerance of each tone is ±20 Hz. The features of the SAT tones are
   a. Each land station is assigned to one SAT among the three listed above.
   b. One SAT tone is added to each forward voice channel (FVC) by a land station. The mobile station detects, filters, and modulates on the reversed voice channel (RVC) with this same tone.
   c. SAT is suspended during transmission of wideband data (a burst of signaling of 10 kbps), information, or other control features on the reverse voice channel.
   d. It is not suspended when a signaling tone (10 kHz) is sent.
   e. The received audio transmission must be muted if the measured SAT and SCC do not agree with each other.
2. *SAT transmission.* The tone modulation index is ⅓. It is a narrowband FM. The deviation $\Delta F = \pm 2$ kHz is centered around each SAT tone.
3. *Fade timing status of SAT.* The transmitter is turned off if no valid SAT tone can be detected or the measured SAT does not agree with the SAT color code (SCC) of each cell site transmitted by the control signal, or if no SAT is received when the timer counts to 5 s. SCC is indicated in the set-up channel.

Signaling tone. Signaling tones must be kept within 10 kHz ± 1 Hz and produce a nominal frequency deviation of ±8 kHz of the carrier frequency. It is used over the voice channel. It serves three functions:

1. Flush for special orders
2. Terminate the calls
3. Order confirmation

Malfunction timer set at 60 s. The transmission will cease when the timer exceeds 60 s. The timer never expires as long as the proper sequence of operations is taking place.

3.2.6 Call processing

Initialization. When the user turns on the mobile station, the initialization work starts.

1. Mobile station must tune to the strongest dedicated control channel (usually one of 21 channels) within 3 s. Then a system parameter message should be received.
2. If it cannot complete this task on the strongest dedicated control channel, it turns to the second strongest dedicated control channel and attempts to complete this task within the next 3-s interval.
3. Check whether it is in Enable or Disable status. Serving system status changed to Enable if the preferred system is System A. Serving system status changed to Disable if the preferred system is System B.

Paging channel selection

1. The mobile station must then tune to the strongest paging channel within 3 s. Usually paging channels are control channels.
2. Receive an overhead message train and update the following: Roam, system identification, local control status.

Idle stage. A mobile station executes each of four tasks at least every 46.3 ms.

1. *Responds to overhead information.* Compare SID_s (stored) with SID_r (received). If $SID_s \neq SID_r$, return to initialization stage.
2. *Page match.* If the roam status is Disable, the mobile station must attempt to match for one-word message or two-word messages. If the roam status is Enable, the mobile station must attempt to match two words.
3. *Order.* After matching the MIN, respond to the order.
4. *Rescan access channels.* Because the vehicle is moving, information must be updated. Rescan at 2- to 5-min intervals, depending on different manufacturers.

Call initiation. Origination indication (system access task).

1. *System access task.* When the system access task is started, an

access timer is set for

Origination	max 12 s
Page response	max 6 s
An order response	max 6 s
Registration	max 6 s

2. *Scan access channels.* Choose one or, at most, two channels with the strongest signals. If the service request cannot be completed on the strongest signal, the mobile station can select the second strongest access channel (called the *alternate access channel*).
3. *Seize reverse control channel* (RECC).
   a. BIS = 1 status: The mobile station is ready for sending. The land station may ask the mobile station to check and wait for overhead message (WFOM) bit.

WFOM = 1	The mobile station waits to update overhead information (see item 5).
WFOM = 0	Delay a random time (0 to 92 ms) and send service request. It is an access attempt.

   b. BIS = 0 status: This is the status of a "busy" condition. The mobile station increments NBUSY by 1, and then has to wait a random time interval of 0 to 200 ms to check the BIS status (0 or 1) again. When NBUSY exceeds MAXBUSY the call is terminated.
4. *Access attempt parameters.* Maximum of 10 attempts. There is a random delay interval of 0 to 92 ± 1 ms for each attempt at checking the status of BIS. If BIS = 1, the mobile station just waits for the transmitting power to come up and sends out the service request message. The random time delay is used to avoid two or more mobile stations requesting services at the same time.
5. *Update overhead information.* Update overhead information should be completely received by the mobile station within 1.5 s after a call is initiated. Update overhead information is as follows.
   Overhead information (*OHD*). The OHD will be sent by the land station and updated by the mobile station.
   a. Overhead control message (whether the system is overloaded or not)
   b. Access type parameter message sets the busy-idle status bits in the BIS field.
   c. Access attempt parameters message provides the following parameters.
      (1) Maximum number of seizure tries allowed
      (2) Maximum number of busy occurrences

   After the update overhead information has been completely received, the mobile station waits a random time interval of 0- to

750-ms and enters the seize reverse control channel task stated in item 3.

6. *Delay after failure.* The mobile station must examine the access timer every 1.5 s; if it does not expire it reenters the access task after failure. The three failure conditions are as follows.
   a. Collision with other mobile station messages. If the collision occurs before the first 56 bits, the BIS changes from 1 to 0.
   b. The land station does not receive the signaling bits. The BIS remains 1 after the mobile station has sent 104 bits.
   c. The land station receives all the signaling bits but cannot interpret them and respond.

   When these conditions occur, the mobile station must wait a random time before making the next attempt. A random delay should be in the interval of 0 to 200 ms.
7. *Service request message.* A whole package of service request messages must be continuously sent to the land station. The format of each signaling word is shown in Fig. 3.4. There is a maximum length to the message consisting of five words: A, B, C, D, E. After a complete message is sent by the mobile station, an unmodulated carrier follows for 25 ms to indicate the end of the message.
8. *Await message response.* If there is no response after the request is sent for 5 s, the call is terminated, and a 120-impulse-per-minute fast tone is generated to the user. If decoded MIN bits match within 5 s, the mobile station must respond with the following messages
   a. If access is an *origination* or *page response*
      (1) Initiate voice channel designation message. Update the parameters as set in the message.
      (2) For a directed-retry message the mobile station must examine the signal strength on each of the retry channels and choose up to two channels with the strongest signals. The mobile station must then tune to the strongest retry access channel.
   b. If the mobile station encounters the start of a new message before it receives the directed-retry message, the call has to be terminated.

3.2.7 Mobile station controls on the voice channel

Loss of radio-link continuity. While the mobile station is tuned to a voice channel, a fade timer must be started when no SAT tone is re-

ceived. If the fade timer counts to 5 s, the mobile station must turn off.

Confirm initial voice channel. Within 100 ms of the receipt of the initial voice channel designation, the mobile station must determine that the channel number is within the set allocated to the home land station or from the other source.

Alerting

I. Waiting for order
   A. If an order cannot be received in 5 s, terminate the call.
   B. Order received. If order is received within 100 ms, the action to be taken for each order is
      1. *Handoff*
         a. Turn off the home land station.
            (1) a 10-kHz signal tone is on for 50 ms after the SAT tone.
            (2) Turn off signaling tone.
            (3) Turn off transmitter.
         b. Turn on the new site, adjust power level.
            (1) Turn to new channel.
            (2) Adjust to new SAT.
            (3) Set SCC (signaling color code).
            (4) Turn on new transmitter.
      2. *Alert.* Turn on signaling tone, run 500 ms, and enter the Waiting for Answer task.
      3. *Release*
         a. Send signaling tone for 1.8 s.
         b. Stop sending signaling tone.
         c. Turn off the transmitter.
      4. *Audit.* Send order confirmation message to land station, remain in Waiting for Answer task, and reset the order timer for 5 s.
      5. *Maintenance.* Turn on signaling tone, run for 500 ms, and enter the Waiting for Answer task.
      6. *Change power*
         a. Adjust the transmitter to new ordered level.
         b. Send order confirmation to the land station.
         c. Local control. If the local control option is enabled in the mobile station, the local control order can be enabled if the group identification matches the SID_p in the mobile station's permanent security memory. A system operator can have a "local control" order for several markets and order under a group identification.

II. Waiting for answer. After requesting orders from the land station, the mobile station is in the Waiting for Answer status. An alert time must be set to 65 s. If no answer comes back in 65 s, the call is terminated. Events occur in the same order as listed in the Waiting for Order section above.

Conversation. A release-delay timer must be set to 500 ms during the conversation. The task can be used for the following conditions.

1. If the user terminates the call.
2. If the user requests a flush.
3. Within 100 ms of receipt of any orders, action will be taken by the mobile station for each order.

3.2.8 Signaling format

Signaling rate. The signaling rate is 10 kbps ± 1 bps. It is slow enough to not cause the intersymbol interference. The Manchester code waveform is applied so that the energy of this signaling waveform is concentrated at 10 kHz, which can be distinguished from the energy concentrated around the carrier frequency for the baseband voice. (See Fig. 3.3*c*.)

Signaling format. The reverse control channel (RECC) data stream is shown in Fig. 3.4*a*. The first word of 48 bits is called the *seizure precursor,* which consists of 30 synchronization bits, 11 frame bits, and 7 coded DCC (digital color code) bits.

Function	Coding
30 synchronization bits	10101010 – –
11 frame bits (word synchronization)	11100010010
7-bit coded DCC (00)	0000000
(01)	0011111
(10)	1100011
(11)	1111100

Each information word contains 48 bits. Each word block contains 240 bits, where each word is repeated five times.

The maximum data stream is one seizure precursor plus five word blocks: A, B, C, D, and E. The total number of bits is 1248 bits as shown in Fig. 3.4*a*.

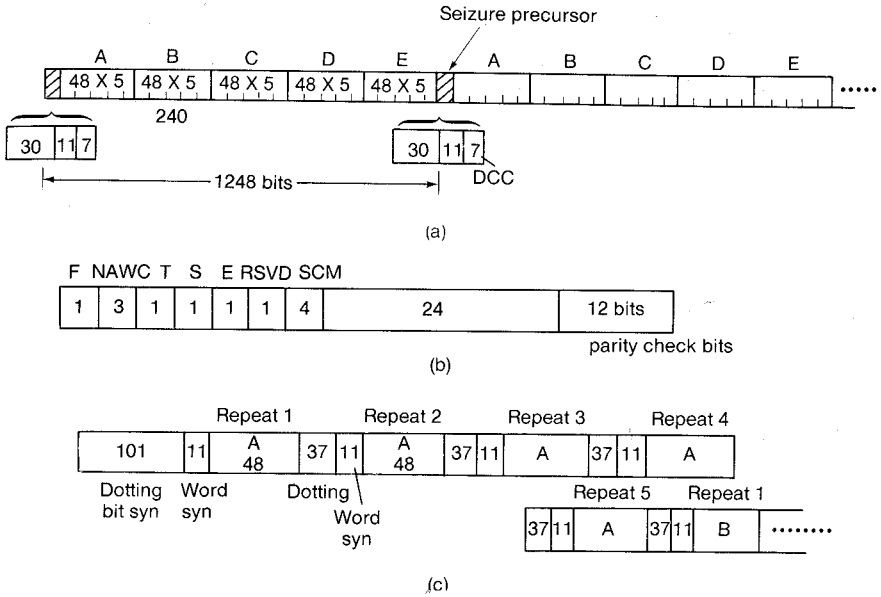


Figure 3.4 Signaling format of both RECC and RVC. (*a*) Reverse control channel (RECC) data stream. (*b*) A message word of RECC. (*c*) Signaling format of RVC.

In each information word, 36 bits are information bits, and the other 12 bits are parity check bits, formed by encoding 36 bits into a (48, 36) BCH code that has a Hamming distance of 5 (described in Chap. 13). The format is shown in Fig. 3.4*b* for the first word. The interpretation of the data field is as follows.

F	First word indication field, 1—first word, 0—subsequent words
NAWC	Number of additional words coming
T	T field, 1—indicates an origination, 0—indicates page response
S	S field, 1—send serial number word, 0—otherwise
E	Extended address field, 1—extended address word sent, 0—not sent
SCM	Station class mark field (see Sec. 2.2.1)

Types of messages. The types of messages to be transmitted over the reverse control channel (RECC) are:

- Page response message. When the mobile station receives a page from the land station, the mobile station responds back.
- Origination message. The mobile station originates the call.
- Order confirmation message. The mobile station responds to the order from the land station.

- Order message. The mobile station orders the tasks which should be performed by the land station and the mobile transmission switching office (MTSO).

Function of each word

Word A	An abbreviated address word. It is always sent to identify the mobile station.
Word B	An extended address word. It will be sent on request from the land station or in a roam situation. In addition, the local control field and the other field are shown in this word.
Word C	A serial number word. Every mobile unit has a unique serial number provided by the manufacturer. It is used to validate the eligible users.
Word D	The first word of the called address.
Word E	The second word of the called address.

Reverse voice channel (RVC). The reverse voice channel (RVC) is also used by a wideband data stream sent from the mobile station to the land station. A 10 kbps ± 1 bps data stream is generated. A word is formed by encoding the 36 content bits into a (48, 36) BCH code, the same as the RECC.

1. *Signaling format.* The first 101 syn bits are used for increasing the possibility of successful syn. The signaling format of RVC is shown in Fig. 3.4*c*. There are two words: The first word repeats five times, and then the second word repeats five times.
2. *Types of messages.* There are two types of messages:
   a. Order confirmation message (one word) responds to the land station to confirm the order, e.g., handoff confirmation.
   b. Called-address message (two words) establishes a three-party call.

3.3 Specification of Land Station (United States)[1–3]

Most parts in the specification of the land station are the same as the specification of the mobile station, such as the modulation of voice signals (Sec. 3.2.2), security and identification (Sec. 3.2.4), and supervision (Sec. 3.2.5). These sections will not be repeated in the specification of the land station.

3.3.1 Power

Maximum effective radiated power (ERP) and antenna height above the average terrain (HAAT) must be coordinated locally on an ongoing

basis. Maximum power is 100 W at a HAAT of 500 ft. Normally, the transmitting 20 W at an antenna height of 100 ft above the local terrain is implemented.

3.3.2 Limit on emission

The field strength limit at a distance of 100 ft or more from the receiver is 500 μV/m.

3.3.3 Call processing

Call processing is the land station operation that controls the mobile station.

Overhead functions for mobile station initiation. The overhead message train contains the first part of the system identification (SID1) and the number of paging channels (N).

On control channel

1. Overhead information is sent on the forward control channel and requires all mobile stations to either update or respond with new information during a system access.

   Update the following information

   - First part of the system identification (SID1)
   - Serial number (S). If S = 1, all mobile stations send their serial numbers during a system access; if S = 0, no need to send serial number.
   - Registration (REGH, REGR). The land station is capable of registering the mobile stations.

     REGH = 1 Enables registration for home mobile stations
     REGH = 0 Otherwise
     REGR = 1 Enables registration for roaming mobile stations
     REGR = 0 Otherwise

   - Extended address (E)

     E = 1 Both MIN1 and MIN2 required
     E = 0 Otherwise

   - Discontinuous transmission (DTX)

     DTX = 1 Let mobile stations use discontinuous transmission mode on the voice channel (reducing power consumption for portable units)
     DTX = 0 Otherwise

- Number of paging channels (N)
- Read control-filler message (RCF) (see Sec. 3.3.4)

  RCF = 1 ask the mobile unit to read control-filler message before accessing a system on a reverse control channel

  RCF = 0 Otherwise

- Combined paging/access (CPA)

  CPA = 1 Paging channel and access channel are the same

  CPA = 0 Paging channel and access channel are not the same

- Number of access channel (CMAX)

  Respond with the following information

- Local control. A system operation for home mobile stations and for the roaming mobile stations that are members.
- New access channels (NEWACC). Send NEWACC information along the first access channel.
- Registration increment (REGINCR). Each time the mobile station increments a fixed value received on FOCC for its updated registration ID if it is equipped for autonomous registration.
- Registration ID (REGID). The last registration number received on FOCC and stored at the mobile station. Every time an increment occurs REGID (new) = REGID (old) + REGINCR, the mobile station identifies itself to a land station.
- Rescan. The rescan global action message must be sent to require all mobile stations to enter the initialization task.

2. The land station will use a control message (one word or two words) to page a mobile station through its home land station. The roaming mobile station must be paged with a two-word message.
3. Orders must be sent to mobile stations with a two-word control message. The orders can be audit and local control. By sending local orders with the order field set to local control and using system identifications (SID) that have identical group identifications, a home mobile station or a roaming mobile station which is a member of a group can be distinguished.

Land station support of system access

1. *Overhead information.* The following information must be sent on a forward control channel to support system access which is used by mobile stations.
   - *Digital color code* (DCC). The mobile station uses DCC to identify the land station.

- *Control mobile attenuation code* (CMAC). When a control-filler message is transmitted, the mobile station receiving the code has to adjust its transmitter power level before accessing a system on a reverse control channel.
- *Wait for overhead message* (WFOM). Set WFOM to 1 in the control-filler message; then the mobile station must wait for WFOM before accessing a system on a reverse control channel.
- *Overload control* (OLC). The mobile stations that are assigned to one or more of the 16 overload classes (N = 1 to 16) must not access the system for originations on the RECC.
- *Access-type parameters.* When the access-type parameters' global action message with the BIS field set to 0 is appended to a system parameter overhead message, the mobile stations do not check for an idle-to-busy status.
- *Access-attempt parameters* are the limit on the number of "busy" occurrences for mobile stations or the default values for the number of seizure attempts.

2. *Reverse control channel seizure by a mobile station.* When this equals 1 all mobile stations must check for an idle-to-busy status when accessing a system. A seizure precursor (48 bits including coded DCC) sent by a mobile station and received by the land station should match its encoded form of DCC.

   It must set the status of the busy-idle bits on the forward control channel between 0.8 and 2.9 ms of receipt of the last bit of 48 bits of the seizure precursor. The busy-idle bits must remain busy until
   - 30 ms after the last word of message has been received
   - (24 N + 55) ms otherwise, where N is the maximum number of words. It will not exceed 175 ms.
3. *Response to mobile station messages.* It is not required that the land station respond to the mobile station message. During periods of system overload or high usage, it may be desirable to permit mobile stations to "time-out" rather than sending release or other orders which use system capacity. The usual time-out period is 5 s. It means that after 5 s, if the mobile station does not receive any response from the land station, the mobile station terminates the transmitted power. The following responses to mobile stations may be sent:
   a. Origination message. Send one of the following orders.
   (1) Initial voice channel designation
   (2) Directed retry—direct to other cell site
   (3) Intercept—priority feature
   (4) Reorder—initiate again
   b. Page response message. Send one of the following orders.
   (1) Initial voice channel designation

(2) Directed retry
(3) Release—turn off signaling tone and release the channel

c. *Order message.* Send one of the following orders.
(1) Order confirmation
(2) Release

d. *Order confirmation message.* "No message is sent."

Mobile station control on voice channel. The change of status of the supervisory audio tone (SAT) and signaling tone (ST) are used to signal the occurrence of certain events during the progress of a call, such as confirming orders, sending a release request, sending a flash request, and loss of radio-link continuity. In addition to the analog signaling (SAT and ST) to and from the mobile station, digital messages (in a burst mode with 10 kbps transmission rate) can be sent to and received from the mobile station. Response to the digital message is either a digital message or a status change of SAT and ST.

We use the notation "(SAT, ST) status" to describe the signaling condition.

SAT		ST		(SAT, ST)	
On	Off	On	Off	status	Conditions
1			0	(1,0)	Mobile off-hook
1		1		(1,1)	Mobile on-hook
	0	1		(0,1)	Mobile in fade
	0		0	(0,0)	Mobile transmitter off

1. *Loss of radio-link continuity.* A designated SAT tone is continuously sent to the mobile station; the same SAT should be sent back on a reverse voice channel. If within 5 s the SAT has not been received, the land station would assume that the mobile station is lost and terminates the call.
2. *Initial voice channel confirmation*
   a. Confirmation will be received by the land station as a change in the SAT, ST status from (0,0) to (1,0).
   b. If the confirmation is not received, the land station must either resend the message or turn off the voice channel transmitter.
   c. If the mobile station was paged, the land station must enter the Wait for Order task or Conversation task.
3. *Alerting*
   a. *Waiting for Order task.* After being paged, the mobile station confirms the initial voice channel designation.
      (1) *Handoff.* The mobile station confirms the order by a change

in the (SAT, ST) status from (1,0) to (1,1) for 50 ms. The land station must remain in the Waiting for Order task.

(2) *Alert.* The mobile station confirms the order by changing (SAT, ST) status from (1,0) to (1,1). The land station must then enter the Waiting for Answer task.

(3) *Release.* The mobile station confirms the order by a change of (SAT, ST) status from (1,0) to (1,1) and holds the (1,1) status for 1.8 s. The land station must then turn off the transmitter.

(4) *Audit.* The mobile station confirms the order by a digital message. The land station remains in the Waiting for Order task.

(5) *Maintenance.* The mobile station confirms the order by a change in (SAT, ST) status from (1,0) to (1,1). The land station remains in the Waiting for Order mode.

(6) *Change power.* The mobile station confirms the order by a digital message.

(7) *Local control.* The confirmation and action depends on the message.

b. Waiting for Answer task. When this task is entered, an alert timer must be set for 30 s. The following orders can be sent:

(1) *Handoff.* The mobile station confirms the order by a change of (SAT, ST) status from (1,1) to (1,0) for 500 ms followed by (1,0) to (1,1) held for 50 ms on the old channel. Then (1,1) status is sent on the new channel.

(2) *Alert.* If no confirmation is received, the land station must reset the alert timer to 30 s.

(3) *Stop alert.* The mobile station confirms the order by a change of (SAT, ST) status from (1,1) to (1,0).

(4) *Release.* The mobile station confirms the order by changing (SAT, ST) status from (1,1) to (1,0) for 500 ms followed by a change of (SAT, ST) from (1,0) to (1,1), which is then held for 1.8 s. The land station must turn off the transmitter.

(5) *Audit.* The mobile station confirms the order by a digital message.

(6) *Maintenance.* If no confirmation is received, the land station resets the alert timer to 30 s.

(7) *Change power.* The mobile station confirms the order by a digital message (see Sec. 3.2.1).

(8) *Local control.* The confirmation and action depends on the message.

4. *Conversation.* The mobile station signals an answer by a change in the (SAT, ST) status from (1,1) to (1,0). The land station enters the conversation task.

a. *Handoff.* The mobile station confirms the order by a change in the (SAT, ST) from (1,0) to (1,1), which is then held for 50 ms. Then the land station must remain in the Conversation task.
b. *Send called address.* The called mobile station confirms the order by a digital message with the called address information. This feature would save the established link if the called address were in error because of the transmission medium.
c. The functions *alert, release, audit, maintenance,* and *local control.* Same as in the Waiting for Order task.
d. *Change power.* The mobile station confirms the order by a digital message.
e. *Flash request.* The mobile station signals a flash by changing (SAT, ST) from (1,0) to (1,1) then holding (1,1) for 400 ms, then following with a transition to (1,0).
f. *Release request.* The mobile station signals a release by changing the (SAT, ST) status from (1,0) to (1,1), which is then held for 1.8 s. The land station must turn off the transmitter. This would be used for the mobile user who dials a called number and decides to terminate for any reason.

3.3.4 Signaling formats

Forward control channel (FOCC). The FOCC is a continuous wideband 10 kbps ± 0.1 bps data stream sent from the land station to the mobile station. Each forward control channel consists of three discrete information streams (see Fig. 3.5):

Stream A (least significant bit of MIN = 0)

Stream B (least significant bit of MIN = 1)

Busy-idle stream (busy = 0, idle = 1); it is at a 1 kbps rate, i.e., one busy-idle bit every 10 data bits.

The 10-bit dotting sequence (1010101010) is for bit syn. The 10-bit length is assumed to be sufficient for bit syn because the mobile station is always monitoring the FOCC after initialization (see Sec. 3.2.6). The frame syn bits are the Barker sequence (11100010010). A word is formed by coding 28 control bits into a (40, 28, 5) BCH code. The total number of bits is 40, the number of information bits is 28, and the hamming distance is 5. The hamming distance d can be translated to the capability of error correction bits, t, as follows:

$$t = \frac{d-1}{2} = 2$$

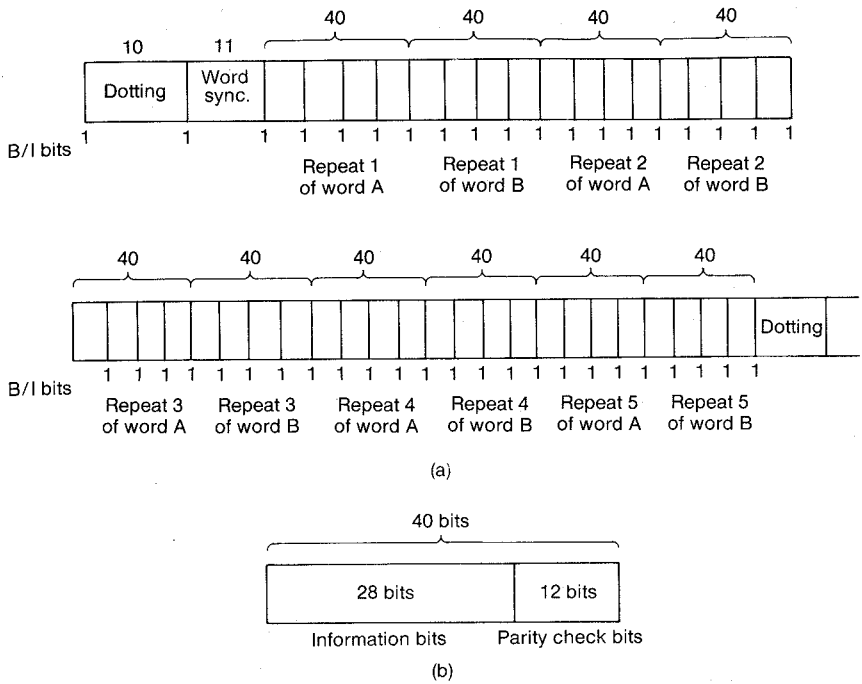


Figure 3.5 Forward control channel message stream (land-to-mobile). (*a*) Signaling format of FOCC. (*b*) A message word of FOCC.

Since this code will detect errors as well as correct them, it reduces to correct one bit in error and assure detection of two bits in error. The code is a shortened version of the primitive (63,51,5) BCH code. It has 12 parity check bits. As long as the 12 parity check bits are retained, the shortened version can be any length. We use (40,28,5), for which a message of 28 bits is suitable (see Fig. 3.5*b*). The transmission rate is 10 kbps. The throughput is 1200 bps (see Fig. 3.5*a*).

Types of messages

1. *Mobile station control message.* Consists of one, two, or four words.
   a. *Word 1.* Abbreviated address word (see Fig. 3.6*a*), T_1T_2 = type field.
      $T_1T_2 = 00$ Only word 1 is sent
      $T_1T_2 = 01$ Multiple words are sent
      DCC—digital color code field, 2 bits sent on FOCC. Then it is received by the mobile station and translated to a 7-bit coded DCC on RECC (see Fig. 3.6*d*).
   b. *Words 2 to 4.* Extended address word $T_1T_2 = 10$, set in each additional word (see Fig. 3.6 *b* and *c*). Let SCC = 11 or SCC $\neq$ 11 (see two bits indicated in Fig. 3.6*d*).

2	2	24	12
$T_1 T_2$	DCC	MINI $_{23\text{-}0}$	P

(a)

2	2	10	1	5	3	5	12
$T_1 T_2 = 10$	SCC = 11	$MIN2_{33\text{-}34}$	RSVD = 0	Local	OrdQ	Order	P

(b)

2	2	10	3	11	12
$T_1 T_2 = 10$	SCC $\neq$ 11	$MIN2_{33\text{-}24}$	VMAC	Chan	P

(c)

DCC (from FOCC)	7 bit coded DCC on RECC	SCC		
0 0	0 0 0 0 0 0 0	SCC $\neq$ 11	0 0	5970 Hz
0 1	0 0 1 1 1 1 1		0 1	6000 Hz
1 0	1 1 0 0 0 1 1		1 0	6030 Hz
1 1	1 1 1 1 1 0 0		1 1	Not a channel designation

(d)

Figure 3.6 Mobile station control message. (*a*) Word 1—abbreviated address word (the busy-idle stream is not shown); (*b*) word 2—extended address word (SCC = 11); (*c*) word 2—extended address word (SCC $\neq$ 11); (*d*) DCC and SCC codes.

(1) *Word 2*

(*a*) By combining "order code" (ORDER) and "order qualification code" (ORDQ) in this word (see Fig. 3.6*b*), we can describe 11 functions.

Page (or origination)	Registration
Alert	Intercept
Release	Maintenance
Reorder	Send called address
Stop alert	Direct-retry status
Audit	

(*b*) Also changes the mobile station power levels in eight levels.

(2) *Word 3.* First directed-retry word.

(3) *Word 4.* Second directed-retry word.

2. *Overhead message (OHD).* A 3-bit OHD field (see Fig. 3.7*a*) is used to identify the overhead message types. It locates just before the 12

Overhead message types

Code	Order
000	registration ID
001	control-filler
010	reserved
011	reserved
100	global action
101	reserved
110	word 1 of system parameter message
111	word 2 of system parameter message

(a)

2	2	20	1	3	12
T_1 T_2	DCC	REGID	END	OHD	P

(b)

Figure 3.7 Overhead message types and format.

parity check bits shown in Fig. 3.7*b*. The overhead types are grouped into the following functional classes.

a. System parameter overhead message. The system parameter overhead message must be sent every 0.8 ± 0.3 s. It consists of two words: word 1 contains the first part of the system identification field, and word 2 contains the number of paging channels and number of access channels.

b. Global action overhead message. There are many global action overhead messages. Each of them consists of one word. The actions are registration increment, new access channel starting point, maximum busy occurrences, maximum seizure tries, etc.

c. Registration ID message. It consists of one word containing the registration ID field.

d. Control-filler message. It consists of one word and is sent whenever there is no other message to be sent on FOCC. It is used to specify a control mobile attenuation code (CMAC) which is used by mobile stations accessing the system, and a Wait for Overhead Message bit (WFOM) indicating whether mobile stations must read an overhead message train before accessing the system.

3. Data restriction

a. The overhead message transmission rate is about once per second.

b. Design the control-filler message to exclude the frame-sync (word sync) sequence.

c. Restrict the use of certain control office codes.

Forward voice channel (FVC). During the call period, FVC is used for signaling. At the beginning, the 101-bit dotting sequence is used for bit sync then for all the repeat dotting sequences; each of them only contains 37 bits, as shown in Fig. 3.8. The word length is 40 bits and repeats 11 times. The reason for repeating 11 times is to be sure that the handoff message would reach the mobile station before the signal dropped below the unacceptable level at the mobile station. The FVC signaling is mainly used for handoff, and the signal level when the handoff occurs is usually very weak. Therefore, the purpose of the FVC is to be sure that the mobile station will get the message and not have a chance to send back a response on receipt because of a weak signal condition.

3.3.5 Additional spectrum radio (ASR) issues

The FCC has allocated an additional 83 voice channels to each system (Band A and Band B). ASR will have 832 channels, half of them, 416 channels (333 channels plus 83 channels), are operated for each system. The 21 control channels still remain the same, but the total number of voice channels becomes 395. The new numbering scheme is shown below.

		Numbering scheme				
(Base Tx) →	869	870	880	890	891.5	894 MHz (Base Tx)
(Mobile Tx) →	824	825	835	845	846.5	849 MHz (Mobile Tx)
(# of channels) →	(New) 33(A)	333(A)	333(B)	(New) 50(A)	(New) 83(B)	
(Ch. numbering) →	991	1 1023	333 334	666 667	716 717	799

ASR is identified by using the station class mark field as shown in Fig. 3.4*b*. The station class mark (SCM) consists of 4 bits and is specified as shown below.

Power class	SCM	Transmission	SCM	Bandwidth	SCM
Class I (4 W)	XX00	Continuous	X0XX	20 MHz	0XXX
Class II (1.2 W)	XX01	Discontinuous	X1XX	25 MHz (ASR)	1XXX
Class III (0.6 W)	XX10				
Reserved	XX11				

The new frequency management charts for Block A and Block B are shown in Table 3.1.

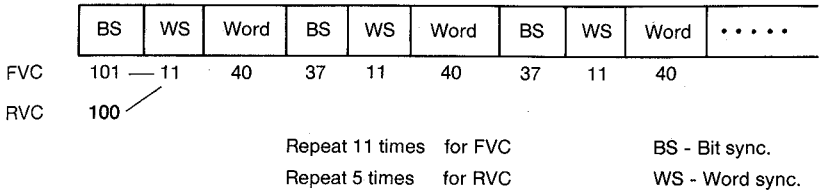


Figure 3.8 Signaling format of FVC and RVC.

3.4 Different Specifications of the World's Cellular Systems

In general, cellular systems can be classified by their operating frequencies: 450-MHz or 800-MHz. Also, they can be distinguished by the spacing between their channels (also called the *channel bandwidth*): 30, 25, or 20 kHz. In the future, Japanese NTT may propose a 12.5-kHz channel spacing in their cellular system.

The large-capacity cellular telephones used in the world are listed in Table 3.2. There are five major systems, and their message protection schemes are different. The major differences are

1. The principle of majority decision (PMD)
2. The automatic repeat request (ARQ)

These schemes each have their merits. In a very severe fading environment, PMD is a good candidate. In a fairly light fading environment, ARQ is a good candidate. Table 3.3 lists the system of each country's use. There are 25 countries which have cellular systems.

Today there is no compatibility among the analog cellular systems serving European countries. In 1987, they agreed to set up a pan-European digital cellular standard, called GSM (special mobile group) specification, which is to be implemented in 1991 and is described as follows.

Multiple access—Time division multiple access (TDMA)

Channel bandwith—200 kHz

Time slots—8

Modulation scheme—Gaussian minimum shift keying (GMSK)

Speech code—RPE-LPC (regular pulse excited-linear predictive code)

Speech coder rate—13,000 bps

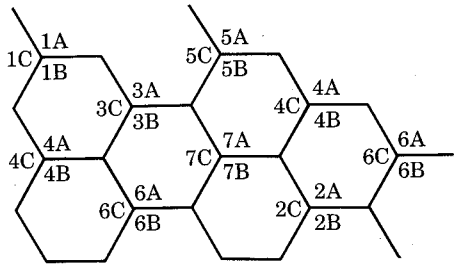
Channel code—Convolutional code

Channel coder rate—22.8 kbps

Equalizer required—Eliminates the delay spread in the received signal due to the mobile radio environment.

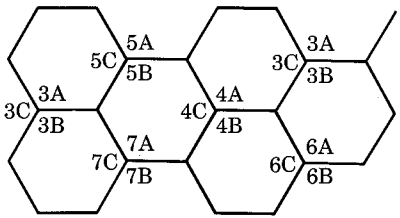
TABLE 3.1 New Frequency Management (Full Spectrum) Block A

1A	2A	3A	4A	5A	6A	7A	1B	2B	3B	4B	5B	6B	7B	1C	2C	3C	4C	5C	6C	7C
1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21
22	23	24	25	26	27	28	29	30	31	32	33	34	35	36	37	38	39	40	41	42
43	44	45	46	47	48	49	50	51	52	53	54	55	56	57	58	59	60	61	62	63
64	65	66	67	68	69	70	71	72	73	74	75	76	77	78	79	80	81	82	83	84
85	86	87	88	89	90												102	103	104	105
106	107	108	109	110	111												123	124	125	126
127	128	129	130	131	132												144	145	146	147
148	149	150	151	152	153												165	166	167	168
169	170	171	172	173	174												186	187	199	190
190	191	192	193	194	195												207	208	209	210
211	212	213	214	215	216												228	229	230	231
232	233	234	235	236	237												249	250	251	252
253	254	255	256	257	258												270	271	272	273
274	275	276	277	278	279												291	292	293	294
295	296	297	298	299	300												312	X	X	X
313*	**314**	**315**	**316**	**317**	**318**	**319**	**320**	**321**	**322**	**323**	**324**	**325**	**326**	**327**	**328**	**329**	**330**	**331**	**332**	**333**
667	668	669	670	671	672	673	674	675	676	677	678	679	680	681	682	683	684	685	686	687
688	689	690	691	692	693	694	695	696	697	698	699	700	701	702	703	704	705	706	707	708
709	710	711	712	713	714	715	716	X	991	992	993	994	995	996	997	998	999	1000	1001	1002
1003	1004	1005	1006	1007	1008	1009	1010	1011	1012	1013	1014	1015	1016	1017	1018	1019	1020	1021	1022	1023



Block B

1A	2A	3A	4A	5A	6A	7A	1B	2B	3B	4B	5B	6B	7B	1C	2C	3C	4C	5C	6C	7C
334	**335**	**336**	**337**	**338**	**339**	**340**	**341**	**342**	**343**	**344**	**345**	**346**	**347**	**348**	**349**	**350**	**351**	**352**	**353**	**354**
355	356	357	358	359	360	361														375
376	377	378	379	380	381	382														396
397	398	399	400	401	402	403														417
418	419	420	421	422	423	424														438
439	440	441	442	443	444	445														459
460	461	462	463	464	465	466														480
481	482	483	484	485	486	487														501
502	503	504	505	506	507	508														522
523	524	525	526	527	528	529														543
544	545	546	547	548	549	550														564
565	566	567	568	569	570	571														585
586	587	588	589	590	591	592	593	594	595	596	597	598	599	600	621	622	623	624	625	606
607	608	609	610	611	612	613	614	615	616	617	618	619	620	621	622	623	624	625	626	627
628	628	630	631	632	633	634	635	636	637	638	639	640	641	642	643	644	645	646	647	648
649	650	651	652	653	654	655	656	657	658	659	660	661	662	663	664	665	666	717	718	719
720	721	722	723	724	725	726	727	728	729	730	731	732	733	734	735	736	737	738	739	740
741	742	743	744	745	746	747	748	749	750	751	752	753	754	755	756	757	758	759	760	761
762	763	764	765	766	767	768	769	770	771	772	773	774	775	776	777	778	779	780	781	782
783	784	785	786	787	788	789	790	791	792	793	794	795	796	797	798	799				



*Boldface numbers indicate 21 control channels for Block A and Block B respectively.

TABLE 3.2 Large-Capacity Cellular Telephones Used in the World

	Japan	North America	England	Scandinavia	West Germany
System	NTT	AMPS	TACS	NMT	C450
Transmission frequency (kHz):					
Base station	870–885	870–890	935–960	463–467.5	461.3–465.74
Mobile station	925–940	825–845	890–915	453–457.5	451.3–455.74
Spacing between transmission and receiving frequencies (kHz)	55	45	45	10	10
Spacing between channels (kHz)	25	30	25	25	20
Number of channels	600	666 (control channel 21 × 2); interleave used	1000 (control channel 21 × 2); interleave used	180	222

Coverage radius (km)	5 (urban area) 10 (suburbs)	2–20	2–20	1.8–40	5–30
Audio signal:					
Type of modulation	FM	FM	FM	FM	FM
Frequency deviation (kHz)	±5	±12	±9.5	±5	±4
Control signal:					
Type of modulation	FSK	FSK	FSK	FSK	FSK
Frequency deviation (kHz)	±4.5	±8	±6.4	±3.5	±2.5
Data transmission rate (kb/s)	0.3	10	8	1.2	5.28
Message protection	Transmitted signal is checked when it is sent back to the sender by the receiver.	Principle of majority decision is employed.	Principle of majority decision is employed.	Receiving steps are predetermined according to the content of the message.	Message is sent again when an error is detected.

SOURCE: Report from International Radio Consultative Committee (CCIR).

TABLE 3.3 World's Cellular Systems

NTT	AMPS	TACS	NMT	C450	NEC
Japan	U.S. Canada South Korea Hong Kong*	England Hong Kong*	4 Nordic countries Spain The Netherlands Belgium Oman Saudi Arabia Tunisia Malaysia Australia Ireland	West Germany	Australia Singapore Hong Kong Jordan Colombia Mexico Kuwait

*Under construction.
SOURCE: Report from International Radio Consultative Committee (CCIR).

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Chapter

4

Cell Coverage for Signal and Traffic

4.1 General Introduction

Cell coverage can be based on signal coverage or on traffic coverage. Signal coverage can be predicted by coverage prediction models and is usually applied to a start-up system. The task is to cover the whole area with a minimum number of cell sites. Since 100 percent cell coverage of an area is not possible, the cell sites must be engineered so that the holes are located in the no-traffic locations. The prediction model is a point-to-point model which is discussed in this chapter. We have to examine the service area as occurring in one of the following environments:

Human-made structures
In an open area
In a suburban area
In an urban area

Natural terrains
Over flat terrain
Over hilly terrain
Over water
Through foliage areas

The results generated from the prediction model will differ depending on which service area is used.

There are many field-strength prediction models in the literature.[1–28] They all provide more or less an area-to-area prediction. As long as 68 percent of the predicted values from a model are within 6 to 8 dB (one standard deviation) of their corresponding measured value, the model is considered a good one. However, we cannot use area-to-area prediction models for cellular system design because of the large uncertainty of the prediction.

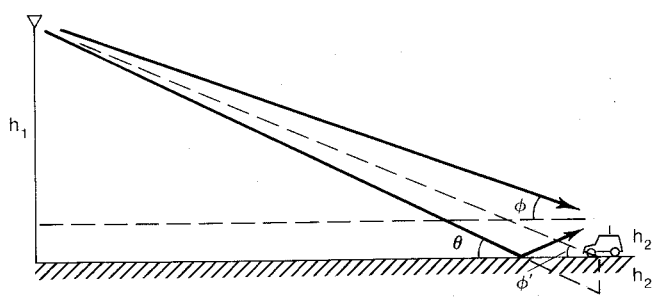
The model being introduced here is the point-to-point prediction model which would provide a standard deviation from the predicted value of less than 3 dB. An explanation of this model appears in Refs. 23, 24, and 41. Many tools can be developed based upon this model, such as cell-site choosing, interference reduction, and traffic handling.

4.1.1 Ground incident angle and ground elevation angle

The ground incident angle and the ground elevation angle over a communication link are described as follows. The ground incident angle θ is the angle of wave arrival incidently pointing to the ground as shown in Fig. 4.1. The ground elevation angle ϕ is the angle of wave arrival at the mobile unit as shown in Fig. 4.1.

Example 4.1 In a mobile radio environment, the average cell-site antenna height is about 50 m, the mobile antenna height is about 3 m, and the communication path length is 5 km. The incident angle is (see Fig. 4.1)

$$\theta = \tan^{-1}\frac{50\text{ m} + 3\text{ m}}{5\text{ km}} = 0.61°$$



θ is the incident angle
ϕ is the elevation angle

Figure 4.1 A coordinate sketch in a flat terrain.

The elevation angle at the antenna of the mobile unit is

$$\phi = \tan^{-1}\frac{50\text{ m} - 3\text{ m}}{5\text{ km}} = 0.54°$$

The elevation angle at the location of the mobile unit is

$$\phi' = \tan^{-1}\frac{50\text{ m}}{5\text{ km}} = 0.57°$$

4.1.2 Ground reflection angle and reflection point

Based on Snell's law, the reflection angle and incident angle are the same. Since in graphical display we usually exaggerate the hilly slope and the incident angle by enlarging the vertical scale, as shown in Fig. 4.2, then as long as the actual hilly slope is less than 10°, the reflection point on a hilly slope can be obtained by following the same method as if the reflection point were on flat ground. Be sure that the two antennas (base and mobile) have been placed vertically, not perpendicular to the sloped ground. The reason is that the actual slope of the hill is usually very small and the vertical stands for two antennas are correct. The scale drawing in Fig. 4.2 is somewhat misleading; however, it provides a clear view of the situation.

Example 4.2 Let $h_1 = 50$ m, $h_2 = 3$ m, $d = 5$ km, and $H = 100$ m as shown in Fig. 4.2.

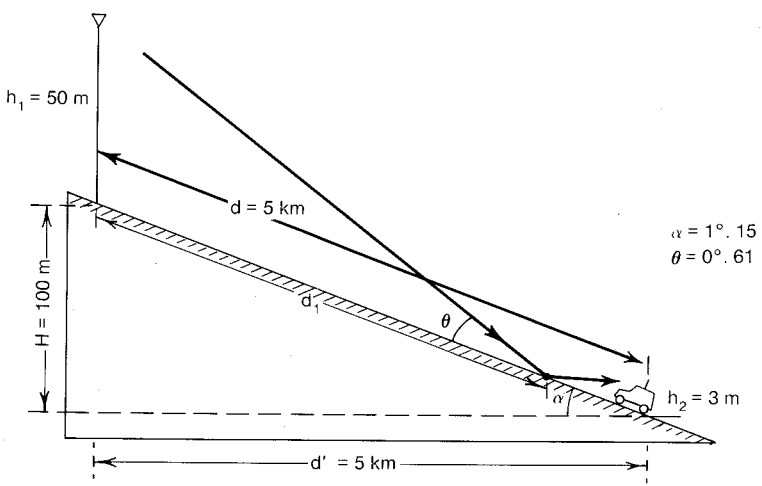


Figure 4.2 A coordinate sketch in a hilly terrain.

(*a*) Using the approximate method ($d \approx d' \approx 5$ km), the slope angle α of the hill is

$$\alpha = \tan^{-1} \frac{100 \text{ m}}{5 \text{ km}} = 1.14576°$$

the incident angle is

$$\theta = \tan^{-1} \frac{50 \text{ m} + 3 \text{ m}}{5 \text{ km}} = 0.61$$

and the reflection point location from the cell-site antenna

$$d_1 = 50/\tan \theta = 4.717 \text{ km}.$$

(*b*) Using the accurate method, the slope angle α of the hill is

$$\alpha = \tan^{-1} \frac{100 \text{ m}}{\sqrt{(5 \text{ km})^2 - (100 \text{ m})^2}} = \tan^{-1} \frac{100}{4999} = 1.14599°$$

The incident angle θ and the reflection point location d_1 are the same as above.

4.2 Obtaining the Point-to-Point Model

This point-to-point model is obtained in three steps: (1) generate a standard condition, (2) obtain an area-to-area prediction model, (3) obtain a point-to-point model using the area-to-area model as a base.

4.2.1 A standard condition

To generate a standard condition and provide correction factors, we have used the standard conditions shown on the left side and the correction factors on the right side[10] of Table 4.1. The advantage of using these standard values is to obtain directly a predicted value in decibels above 1 mW expressed in dBM.

4.2.2 Obtain area-to-area prediction curves for man-made structures

The area-to-area prediction curves are different in different areas. In area-to-area prediction, all the areas are considered flat even though the data may be obtained from nonflat areas. The reason is that area-to-area prediction is an average process. The standard deviation of the average value indicates the degree of terrain roughness.

Therefore, the differences in area-to-area prediction curves are due to the different man-made structures. We should realize that measurements made in urban areas are different from those made in suburban and open areas. The area-to-area prediction curve is obtained

TABLE 4.1 Generating a Standard Condition

Standard condition	Correction factors*
At the Base Station	
Transmitted power P_t = 10 W (40 dBm)	$\alpha_1 = 10 \log \frac{P_t'}{10}$
Antenna height h_1 = 100 ft (30 m)	$\alpha_2 = 20 \log \frac{h_1'}{h_1}$
Antenna gain g_t = 6 dB/dipole	$\alpha_3 = g_{t2}' - 6$
At the Mobile Unit	
Antenna height, h_2 = 10 ft (3 m)	$\alpha_4 = 10 \log \frac{h_2'}{h_2}$
Antenna gain, g_m = 0 dB/dipole	$\alpha_5 = g_m'$

*All the parameters with primes are the new conditions.

from the mean value of the measured data and used for future predictions in that area. Any area-to-area prediction model[1–28] can be used as a first step toward achieving the point-to-point prediction model.

One area-to-area prediction model which is introduced here[10] can be represented by two parameters. (1) the 1-mi (or 1-km) intercept point and (2) the path-loss slope. The 1-mi intercept point is the power received at a distance of 1 mi from the transmitter. There are two general approaches to finding the values of the two parameters experimentally.

1. Compare an area of interest with an area of similar man-made structures which presents a curve such as that shown in Fig. 4.3. The suburban area curve is a commonly used curve. Since all suburban areas in the United States look alike, we can use this curve for all suburban areas. If the area is not suburban but is similar to the city of Newark, then the curve for Newark should be used.
2. If the man-made structures of a city are different from the cities listed in Fig. 4.3, a simple measurement should be carried out. Set up a transmitting antenna at the center of a general area. As long as the building height is comparable to the others in the area, the antenna location is not critical. Take six or seven measured data points at the 1-mi intercept and at the 10-mi boundary. Then compute the average of the 1 mi data points and of the 10 mi data points. By connecting the two values, the path-loss slope can be obtained. If the area is very hilly, then the data points measured at a given distance from the base station in different locations can be far apart. In this case, we may take more measured data points to obtain the average path-loss slope.

   If the terrain of the hilly area is generally sloped, then we have

to convert the data points that were measured on the sloped terrain to a fictitiously flat terrain in that area. The conversion is based on the effective antenna-height gain as[29]

$$\Delta G = \text{effective antenna-height gain} = 20 \log \frac{h_e}{h_1} \quad (4.2\text{-}1)$$

where h_1 is the actual height and h_e is the effective antenna height at either the 1- or 10-mi locations. The method for obtaining h_e is shown in the following section.

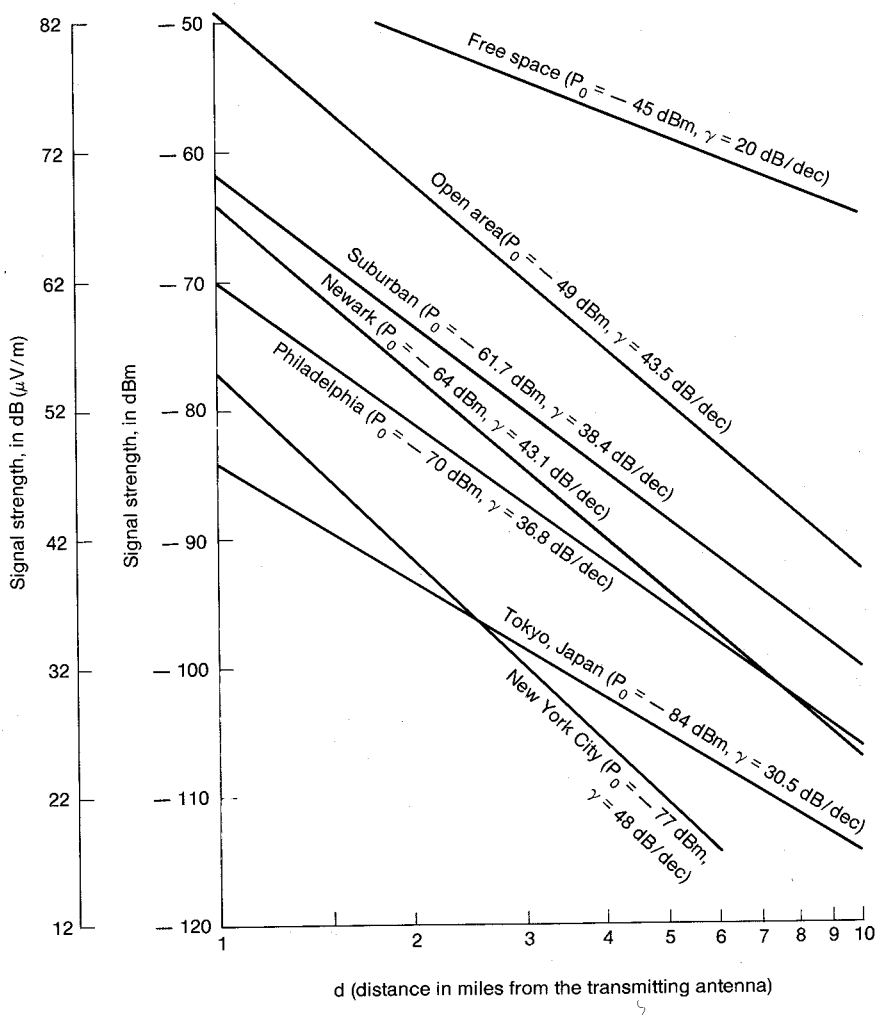


Figure 4.3 Propagation path loss in different areas.

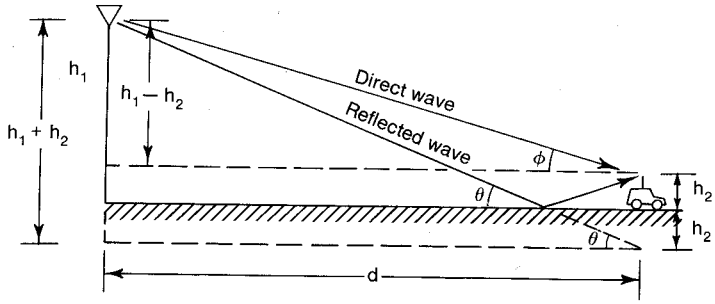


Figure 4.4 A simple model.

4.2.3 The phase difference between a direct path and a ground-reflected path

Based on a direct path and a ground-reflected path (see Fig. 4.4),

$$P_r = P_0 \left(\frac{1}{4\pi d/\lambda}\right)^2 \left|1 + a_v e^{j\Delta\phi}\right|^2 \tag{4.2-2}$$

where a_v = the reflection coefficient
$\Delta\phi$ = the phase difference between a direct path and a reflected path
P_0 = the transmitted power
d = the distance
λ = the wavelength

In a mobile environment $a_v = -1$ because of the small incident angle of the ground wave caused by a relatively low cell-site antenna height.

Thus

$$P_r = P_0 \left(\frac{1}{4\pi d/\lambda}\right)^2 \left| 1 - \cos\Delta\phi - j\sin\Delta\phi \right|^2$$

$$= P_0 \frac{2}{(4\pi d/\lambda)^2}(1 - \cos\Delta\phi) = P_0 \frac{4}{(4\pi d/\lambda)^2} \sin^2\frac{\Delta\phi}{2} \tag{4.2-3}$$

where

$$\Delta\phi = \beta\,\Delta d \tag{4.2-4}$$

and Δd is the difference, $\Delta d = d_1 - d_0$, from Fig. 4.4.

$$d_1 = \sqrt{(h_1 + h_2)^2 + d^2} \tag{4.2-5}$$

and

$$d_2 = \sqrt{(h_1 - h_2)^2 + d^2} \tag{4.2-6}$$

Since Δd is much greater than either d_1 and d_2,

$$\Delta\phi \approx \beta\,\Delta d = \frac{2\pi}{\lambda}\frac{2h_1h_2}{d} \tag{4.2-7}$$

Then the received power of Eq. (4.2-3) becomes

$$P_r = P_0\frac{\lambda^2}{(4\pi)^2d^2}\sin^2\frac{4\pi h_1h_2}{\lambda d} \tag{4.2-8}$$

If $\Delta\phi$ is less than 0.6 rad, then $\sin(\Delta\phi/2) \approx \Delta\phi/2$, $\cos(\Delta\phi/2) \approx l$ and Eq. (4.2-8) simplifies to

$$P_r = P_0\frac{4}{16\pi^2(d/\lambda)^2}\left(\frac{2\pi h_1h_2}{\lambda d}\right)^2 = P_0\left(\frac{h_1h_2}{d^2}\right)^2 \tag{4.2-9}$$

From Eq. (4.2-9), we can deduce two relationships as follows:

$$\Delta P = 40\log\frac{d_1}{d_2} \quad \text{(a 40 dB/dec path loss)} \tag{4.2-10a}$$

$$\Delta G = 20\log\frac{h'_1}{h_1} \quad \text{(an antenna height gain of 6 dB/oct)} \tag{4.2-10b}$$

where ΔP is the power difference in decibels between two different path lengths and ΔG is the gain (or loss) in decibels obtained from two different antenna heights at the cell site. From these measurements, the gain from a mobile antenna height is only 3 dB/oct, which is different from the 6 dB/oct for h'_1 shown in Eq. (4.2-10*b*). Then

$$\Delta G' = 10\log\frac{h'_2}{h_2} \quad \text{(an antenna-height gain of 3 dB/oct)} \tag{4.2-10c}$$

Example 4.3 The distance d = 8 km. The antenna height at the cell site is 30 m, and at the mobile unit it is 3 m. Then the phase difference at 850 MHz is $\beta \cdot \Delta d$, or 0.4 rad, which is less than 0.6 rad. Therefore, Eq. (4.2-9) can be applied.

4.2.4 Why there is a constant standard deviation along a path-loss curve

When plotting signal strengths at an given radio-path distance, the deviation from predicted values is approximately 8 dB.[10,12] This standard deviation of 8 dB is roughly true in many different areas. The explanation is as follows. When a line-of-sight path exists, both the direct wave path and reflected wave path are created and are strong (see Fig. 4.2). When an out-of-sight path exists, both the direct wave path and the reflected wave path are weak. In either case, according

to the theoretical model, the 40-dB/dec path-loss slope applies. The difference between these two conditions is the 1-mi intercept (or 1-km intercept) point. It can be seen that in the open area, the 1-mi intercept is high. In the urban area, the 1-mi intercept is low. The standard deviation obtained from the measured data remains the same along the different path-loss curves regardless of environment.

Support for the above argument can also be found from the observation that the standard deviation obtained from the measured data along the predicted path-loss curve is approximately 8 dB. The explanation is that at a distance from the cell site, some mobile unit radio paths are line-of-sight, some are partial line-of-sight, and some are out-of-sight. Thus the received signals are strong, normal, and weak, respectively. At any distance, the above situations prevail. If the standard deviation is 8 dB at one radio-path distance, the same 8 dB will be found at any distance. Therefore a standard deviation of 8 dB is always found along the radio path as shown in Fig. 4.5. The standard deviation of 8 dB from the measured data near the cell site is due mainly to the close-in buildings around the cell site. The same standard deviation from the measured data at a distant location is due to the great variation along different radio paths.

4.2.5 The straight-line path-loss slope with confidence

As we described earlier, the path-loss curves are obtained from many different runs at many different areas. As long as the distances of the radio path from the cell site to the mobile unit are the same in different runs, the signal strength data measured at that distance would be used to calculate the mean value for the path loss at that distance. In the experimental data, the path-loss deviation is 8 dB across the distance from 1.6 to 15 km (1 to 10 mi) where the general terrain contours are not generally flat. Figure 4.5 depicts this. The path-loss curve is γ. The received power can be expressed as

$$P_r = P_0 - \gamma \log \frac{r}{r_0} \tag{4.2-11}$$

The slope γ is different in different areas, but it is always a straight-line in a log scale. If $\gamma = 20$ is a free-space path loss, $\gamma = 40$ is a mobile path loss.

Confidence level.[31] A confidence level can only be applied to the path-loss curve when the standard deviation σ is known. In American suburban areas, the standard deviation $\sigma = 8$ dB. The values at any given distance over the radio path are concentrated close to the mean and have a bell-shaped (normal) distribution. The probability that 50 per-

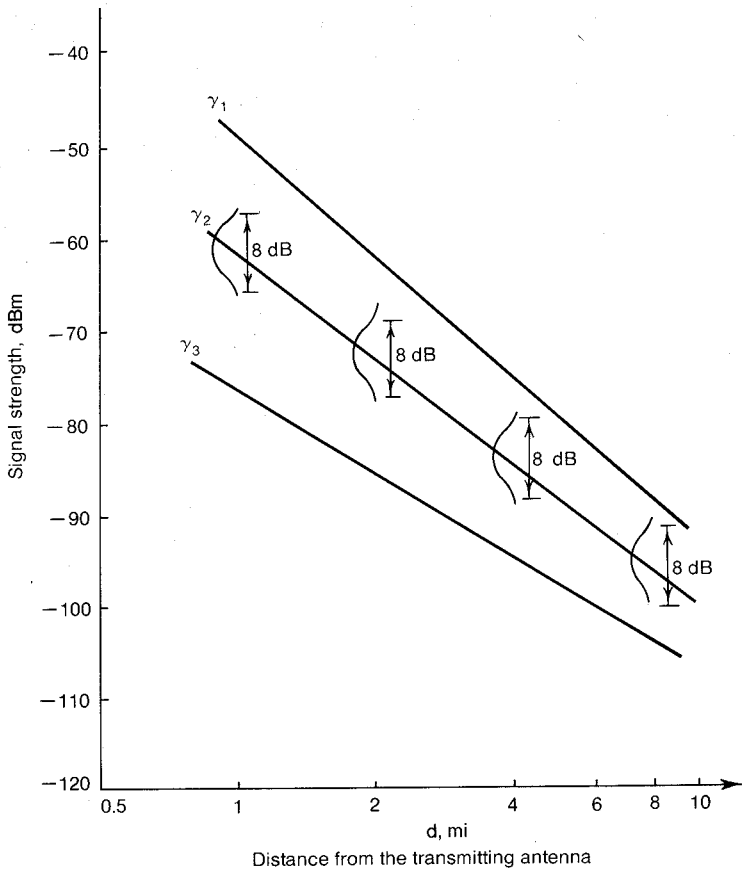


Figure 4.5 An 8-dB local mean spread.

cent of the measured data are equal to or below a given level is

$$P(x \geq C) = \int_C^{\infty} \frac{1}{\sqrt{2\pi}\sigma} e^{-(x-A)^2/2\sigma^2}\, dx = 50\% \tag{4.2-12}$$

where A is the mean level obtained along the path-loss slope, which is shown in Eq. (4.2-11) as

$$A = P_0 - \gamma \log \frac{r_1}{r_0}$$

Thus level A corresponds to the distance r_1. If level A increases, the confidence level decreases, as shown in Eq. (4.2-12).

$$P(x \geq C) = P\left(\frac{x - A}{\sigma} \geqslant B\right) \tag{4.2-13}$$

TABLE 4.2

$P(x \geq C)$, %	$C = B\sigma + A$
80	$-0.85\sigma + A$
70	$-0.55\sigma + A$
60	$-0.25\sigma + A$
50	A
40	$0.25\sigma + A$
30	$0.55\sigma + A$
20	$0.85\sigma + A$
16	$1\sigma + A$
10	$1.3\ \sigma + A$
2.28	$2\ \sigma + A$

Let $C = B\sigma + A$. The different confidence levels are shown in Table 4.2. We can see how to use confidence levels from the following example.

Example 4.4 From the path-loss curve, we read the expected signal level as -100 dBm at 16 km (10 mi). If the standard deviation $\sigma = 8$ dB, what level would the signal equal or exceed for a 20 percent confidence level?

$$P\left(\frac{x - A}{\sigma} \geq B\right) = 20\% \qquad x \geq B\sigma + A \tag{E4.4-1}$$

or from Table 4.2 we obtain

$$x \geq 0.85 \times 8 + (-100) = -93.2 \text{ dBm}$$

The log normal curve with a standard deviation of 8 dB is shown in Fig. 4.6.

4.2.6 Determination of confidence interval

The confidence interval is often confused with confidence level. This usually happens when dealing with a particular run in a particular terrain contour. The signal strength of a run is shown in Fig. 1.6. The local mean is the envelope of the received signal, which also follows a log-normal distribution as shown in Fig. 4.6. The standard deviation of the local mean curve is a reflection of how much variation there is in terrain contour. If we know the standard deviation, then we can estimate how often the local mean (average power of the signal) falls within given limits (confidence interval).

The confidence intervals are defined as

$$\begin{aligned} P(m - \sigma \leq x \leq m + \sigma) &= \int_{m-\sigma}^{m+\sigma} P(z)\, dz \\ &= \int_{m-\sigma}^{m+\sigma} \frac{1}{\sqrt{2\pi}\sigma} e^{-(x-m)^2/2\sigma^2}\, dx \\ &= \int_{-1}^{+1} \frac{1}{\sqrt{2\pi}} e^{-y^2/2}\, dy = 68\% \end{aligned} \tag{4.2-14}$$

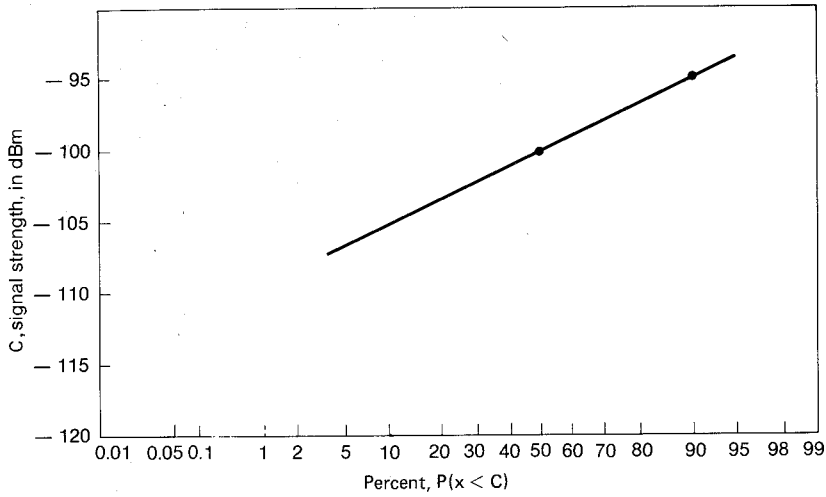


Figure 4.6 A log-normal curve.

or $P(m - 2\sigma \leq x \leq m + 2\sigma)$

$$= \int_{m-2\sigma}^{m+2\sigma} \frac{1}{\sqrt{2\pi}\sigma} e^{-(x-m)^2/2\sigma^2}\, dx = 95.45\% \quad (4.2\text{-}15)$$

where m is the mean of all the data and σ is the standard deviation of all the data.

Equation (4.2-14) indicates that 68 percent of predicted data will fall in the range between $-\sigma$ and $+\sigma$ around this mean value. In other words, we are 68 percent confident that a predicted data point will fall between $m - \sigma$ and $m + \sigma$.

The standard deviation from Fig. 4.6 can be found from Table 4.2 as

$$\sigma = \frac{C - A}{B} \qquad \text{[for a given percentage, } P(x \geq C)] \quad (4.2\text{-}16)$$

Example 4.5 Find the standard deviation of a local mean curve as shown in Fig. 4.6. The confidence level for -95 dBm is found to be 10 percent, and the mean is -110 dBm. Then $C = -95$ dBm, $A = -110$ dBm, $B = 1.3$ (from Table 4.2), and

$$\sigma = \frac{-95 - (-110)}{1.3} = 11.54 \text{ dB}$$

Example 4.6 If we do not have Fig. 4.6 in hand but we know both the average power and its standard deviation, we can determine the percentage of signal above any level.

Assuming that the average power is -90 dBm and the standard deviation is 9 dB, what would the signal level be if the confidence level is 30 percent?

The level would be (from Table 4.2)

$$0.55 \times 9 + (-90) = -85.05 \text{ dBm}$$

The confidence level and the confidence interval of a signal strength can be calculated from the predicted data applied to a point-to-point model in an area of interest. However, the confidence level and the confidence interval of a signal strength cannot be found from a simple path-loss slope. In other words, it cannot be obtained from an area-to-area model unless the standard deviation of the model from which the curves were generated is known.

F(50,70) is a common notation to indicate that a signal strength is predicted under a confidence level of 50 percent for time to 70 percent for coverage. A detailed description can be found in Ref. 30.

4.2.7 A general formula for mobile radio propagation

Here we are only interested in a general propagation path-loss formula in a general mobile radio environment, which could be a suburban area. The 1-mi intercept level in a suburban area is −61.7 dBm under the standard conditions listed in Table 4.1. Combining these data with the equation shown in Eq. (4.2-10*b*) from the theoretical prediction model, and Eqs. (4.2-10*c*) and (4.2-11) from the measured data, the received power P_r at the suburban area can be expressed as

$$P_r = (P_t - 40) - 61.7 - 38.4 \log \frac{r_1}{1 \text{ mi}} + 20 \log \frac{h_1}{100 \text{ ft}} + 10 \log \frac{h_2}{10 \text{ ft}} + (G_t - 6) + G_m \quad (4.2\text{-}17)$$

Equation (4.2-17) can be simplified as

$$P_r = P_t - 157.7 - 38.4 \log r_1 + 20 \log h_1 + 10 \log h_2 + G_t + G_m \quad (4.2\text{-}18)$$

where P_t is in decibels above 1 mW, r_1 is in miles, h_1 and h_2 are in feet, and G_t and G_m are in decibels. Equation (4.2-18) is used for suburban areas. We may like to change Eq. (4.2-18) to a general formula by using P_r at 10 mi as a reference which is −100 dBm, as shown in Fig. 4.3. Also the 40 dB/oct slope used is generous. Then Eq. (4.2-18) changes to

$$P_r = P_t - 156 - 40 \log r_1 + 20 \log h_1 + 10 \log h_2 + G_t + G_m \quad (4.2\text{-}19)$$

where the units of P_t, r_1, h_1, h_2, G_t, and G_m are stated below Eq. (4.2-18). Equation (4.2-19) can be used as a general formula in a mobile radio environment.

4.3 Propagation over Water

Propagation over water is becoming a big concern because it is very easy to interfere with other cells if we do not make the correct arrangements. Interference resulting from propagation over the water can be controlled if we know the cause.

In general, the permittivities ϵ_r of seawater and fresh water are the same, but the conductivities of seawater and fresh water are different. We may calculate the dielectric constants ϵ_c, where $\epsilon_c = \epsilon_r - j60\sigma\lambda$. The wavelength at 850 MHz is 0.35 m. Then ϵ_c (seawater) $= 80 - j84$ and ϵ_c (fresh water) $= 80 - j0.021$.

However, based upon the reflection coefficients formula[32,33] with a small incident angle, both the reflection coefficients for horizontal polarized waves and vertically polarized waves approach 1. Since the 180° phase change occurs at the ground reflection point, the reflection coefficient is -1. Now we can establish a scenario, as shown in Fig. 4.7. Since the two antennas, one at the cell site and the other at the mobile unit, are well above sea level, two reflection points are generated. The one reflected from the ground is close to the mobile unit; the other reflected from the water is away from the mobile unit. We recall that the only reflected wave we considered in the land mobile propagation is the one reflection point which is always very close to the mobile unit. We are now using the formula to find the field strength under the circumstances of a fixed point-to-point transmission and a land-mobile transmission over a water condition.

4.3.1 Between fixed stations

The point-to-point transmission between the fixed stations over the water can be estimated as follows. The received power P_r can be expressed as (see Fig. 4.8)

$$P_r = P_t \left(\frac{1}{4\pi d/\lambda}\right)^2 \left| 1 + a_v\, e^{-j\phi_v} \exp\,(j\,\Delta\phi) \right|^2 \tag{4.3-1}$$

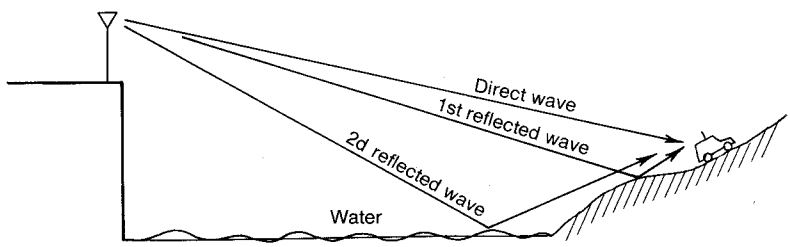


Figure 4.7 A model for propagation over water.

where P_t = transmitted power
d = distance between two stations
λ = wavelength
a_v, ϕ_v = amplitude and phase of a complex reflection coefficient, respectively

$\Delta\phi$ is the phase difference caused by the path difference Δd between the direct wave and the reflected wave, or

$$\Delta\phi = \beta\,\Delta d = \frac{2\pi}{\lambda}\,\Delta d \tag{4.3-2}$$

The first part of Eq. (4.3-1) is the free-space loss formula which shows the 20 dB/dec slope; that is, a 20-dB loss will be seen when propagating from 1 to 10 km.

$$P_0 = \frac{P_t}{(4\pi d/\lambda)^2} \tag{4.3-3}$$

The $a_v e^{-j\phi_v}$ are the complex reflection coefficients and can be found from the formula[32]

$$a_v e^{-j\phi_v} = \frac{\epsilon_c \sin\theta_1 - (\epsilon_c - \cos^2\theta_1)^{1/2}}{\epsilon_c \sin\theta_1 + (\epsilon_c - \cos^2\theta_1)^{1/2}} \tag{4.3-4}$$

When the vertical incidence is small, θ is very small and

$$a_v \approx -1 \qquad \text{and} \qquad \phi_v = 0$$

can be found from Eq. (4.3-4). ϵ_c is a dielectric constant that is different for different media. However, when $a_v e^{-j\phi_v}$ is independent of ϵ_c, the reflection coefficient remains -1 regardless of whether the wave is propagated over water, dry land, wet land, ice, and so forth. The wave propagating between fixed stations is illustrated in Fig. 4.8.

Equation (4.3-1) then becomes

$$P_r = \frac{P_t}{(4\pi d/\lambda)^{12}}\,|1 - \cos\Delta\phi - j\sin\Delta\phi|^2$$

$$= P_0(2 - 2\cos\Delta\phi) \tag{4.3-6}$$

since $\Delta\phi$ is a function of Δd and Δd can be obtained from the following calculation. The effective antenna height at antenna 1 is the height above the sea level.

$$h_1' = h_1 + H_1$$

The effective antenna height at antenna 2 is the height above the sea level.

$$h_2' = h_2 + H_2$$

as shown in Fig. 4.8, where h_1 and h_2 are actual heights and H_1 and H_2 are the heights of hills. In general, both antennas at fixed stations are high, so the reflection point of the wave will be found toward the middle of the radio path. The path difference Δd can be obtained from Fig. 4.8 as

$$\Delta d = \sqrt{(h_1' + h_2')^2 + d^2} - \sqrt{(h_1' - h_2')^2 + d^2} \tag{4.3-7}$$

Since $d \gg h_1'$ and h_2', then

$$\Delta d \approx d\left[1 + \frac{(h_1' + h_2')^2}{2d^2} - 1 - \frac{(h_1' - h_2')^2}{2d^2}\right] = \frac{2h_1'h_2'}{d} \tag{4.3-8}$$

Then Eq. (4.3-2) becomes

$$\Delta\phi = \frac{2\pi}{\lambda}\frac{2h_1'h_2'}{d} = \frac{4\pi h_1'h_2'}{\lambda d} \tag{4.3-9}$$

Examining Eq. (4.3-6), we can set up five conditions:

1. $P_r < P_0$. The received power is less than the power received in free space; that is,

$$2 - 2\cos\Delta\phi < 1 \qquad \text{or} \qquad \Delta\phi < \frac{\pi}{3} \tag{4.3-10}$$

2. $P_r = 0$; that is,

$$2 - 2\cos\Delta\phi = 0 \qquad \text{or} \qquad \Delta\phi = \frac{\pi}{2}$$

3. $P_r = P_0$; that is,

$$2 - 2\cos\Delta\phi = 1 \qquad \text{or} \qquad \Delta\phi = \pm 60^\circ = \pm\frac{\pi}{3} \tag{4.3-11}$$

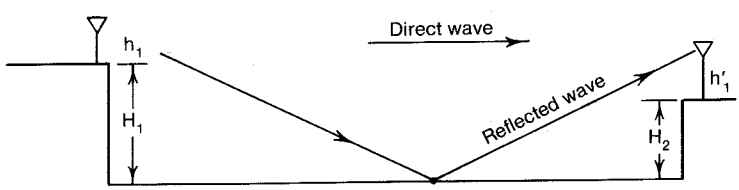


Figure 4.8 Propagation between two hills.

4. $P_r > P_0$; that is,

$$2 - 2\cos\Delta\phi > 1 \qquad \text{or} \qquad \frac{\pi}{3} < \Delta\phi > \frac{5\pi}{3} \tag{4.3-12}$$

5. $P_r = 4P_0$; that is,

$$2 - 2\cos\Delta\phi = \max \qquad \text{or} \qquad \Delta\phi = \pi \tag{4.3-13}$$

The value of $\Delta\phi$ can be found from Eq. (4.3-9). Now we can examine the situations resulting from Eq. (4.3-9) in the following examples.

Example 4.7 Let a distance between two fixed stations be 30 km. The effective antenna height at one end h_1 is 150 m above sea level. Find the h_2 at the other end so that the received power always meets the condition $P_r < P_0$ at 850-MHz transmission ($\lambda = 0.35$ m).

solution

$$\frac{4\pi h_1' h_2'}{\lambda d} \le \frac{\pi}{3} \tag{E4.7-1}$$

or

$$h_2' \le \frac{d\lambda}{12h_1'} = \frac{30{,}000 \times 0.35}{12 \times 150} = 6 \text{ m} \tag{E4.7-2}$$

Example 4.8 Using the same parameters given in Example 4.7, find the range of h_2 which would keep $P_r > P_0$, and find the maximum received power P_r for $P_r = 4P_0$.

solution

$$a.\ \frac{\pi}{3} \le \frac{4\pi h_1' h_2'}{\lambda d} \le \frac{5\pi}{3} \qquad \text{the range of } h_2 \text{ for } P_r > P_0 \tag{E4.8-1}$$

Substituting the values given in Example 4.7, we obtain

$$6 \text{ m} < h_2 < 30 \text{ m} \qquad 42 \text{ m} < h_2 < 66 \text{ m} \tag{E4.8-2}$$

b. $\Delta\phi = \pi$ for the maximum received power.

$$h_2 = 18 \text{ m} \qquad h_2 = 54 \text{ m} \qquad h_2 = 6 \text{ m}[3(2n - 1)] \tag{E4.8-3}$$

where n is any integer.

4.3.2 Land-to-mobile transmission over water

The propagation model would be different for land-to-mobile transmission over water. As depicted in Fig. 4.7, there are always two equal-strength reflected waves, one from the water and one from the proximity of the mobile unit, in addition to the direct wave. The reflected

wave, whose reflected point is on the water is counted because there are no surrounding objects near this point. Therefore the reflected energy is strong. The other reflected wave that has a reflection point proximal to the mobile unit also carries strong reflected energy to it.

Therefore, the reflected power of the two reflected waves can reach the mobile unit without noticeable attenuation. The total received power at the mobile unit would be obtained by summing three components.

$$P_r = \frac{P_t}{(4\pi d/\lambda)^2} |1 - e^{j\Delta\phi_1} - e^{j\Delta\phi_2}|^2 \tag{4.3-14}$$

where $\Delta\phi_1$ and $\Delta\phi_2$ are the path-length difference between the direct wave and two reflected waves, respectively. Since $\Delta\phi_1$ and $\Delta\phi_2$ are very small usually for the land-to-mobile path, then

$$P_r = \frac{P_t}{(4\pi d/\lambda)^2} |1 - \cos\Delta\phi_1 - \cos\Delta\phi_2 - j(\sin\Delta\phi_1 + \sin\Delta\phi_2)|^2 \tag{4.3-15}$$

Follow the same approximation for the land-to-mobile propagation over water.

$$\cos\Delta\phi_1 \approx \cos\Delta\phi_2 \approx 1 \qquad \sin\Delta\phi_1 \approx \Delta\phi_1 \qquad \sin\Delta\phi_2 \approx \Delta\phi_2$$

Then

$$P_r = \frac{P_t}{(4\pi d/\lambda)^2} |-1 - j(\Delta\phi_2 + \Delta\phi_2)|^2$$

$$= \frac{P_t}{(4\pi d/\lambda)^2} [1 + (\Delta\phi_1 + \Delta\phi_2)^2] \tag{4.3-16}$$

In most practical cases, $\Delta\phi_1 + \Delta\phi_2 < 1$; then $(\Delta\phi_1 + \Delta\phi_2)^2 \ll 1$ and Eq. (4.3-16) reduces to

$$P_r \approx \frac{P_t}{(4\pi d/\lambda)^2} \tag{4.3-17}$$

Equation (4.3-17) is the same as that expressing the power received from the free-space condition. Therefore, we may conclude that the path loss for land-to-mobile propagation over land, 40 dB/dec, is different for land-to-mobile propagation over water. In the case of propagation over water, the free-space path loss, 20 dB/dec, is applied.

4.4 Foliage Loss

Foliage loss is a very complicated topic that has many parameters and variations. The sizes of leaves, branches, and trunks, the density and

distribution of leaves, branches, and trunks, and the height of the trees relative to the antenna heights will all be considered. An illustration of this problem is shown in Fig. 4.9. There are three levels: trunks, branches, and leaves. In each level, there is a distribution of sizes of trunks, branches, and leaves and also of the density and spacing between adjacent trunks, branches, and leaves. The texture and thickness of the leaves also count. This unique problem can become very complicated and is beyond the scope of this book. For a system design, the estimate of the signal reception due to foliage loss does not need any degree of accuracy.

Furthermore, some trees, such as maple or oak, lose their leaves in winter, while others, such as pine, never do. For example, in Atlanta, Georgia, there are oak, maple, and pine trees. In summer the foliage is very heavy, but in winter the leaves of the oak and maple trees fall and the pine leaves stay. In addition, when the length of pine needles reaches approximately 6 in., which is the half wavelength at 800 MHz, a great deal of energy can be absorbed by the pine trees. In these situations, it is very hard to predict the actual foliage loss.

However, a rough estimate should be sufficient for the purpose of system design. In tropic zones, the sizes of tree leaves are so large and

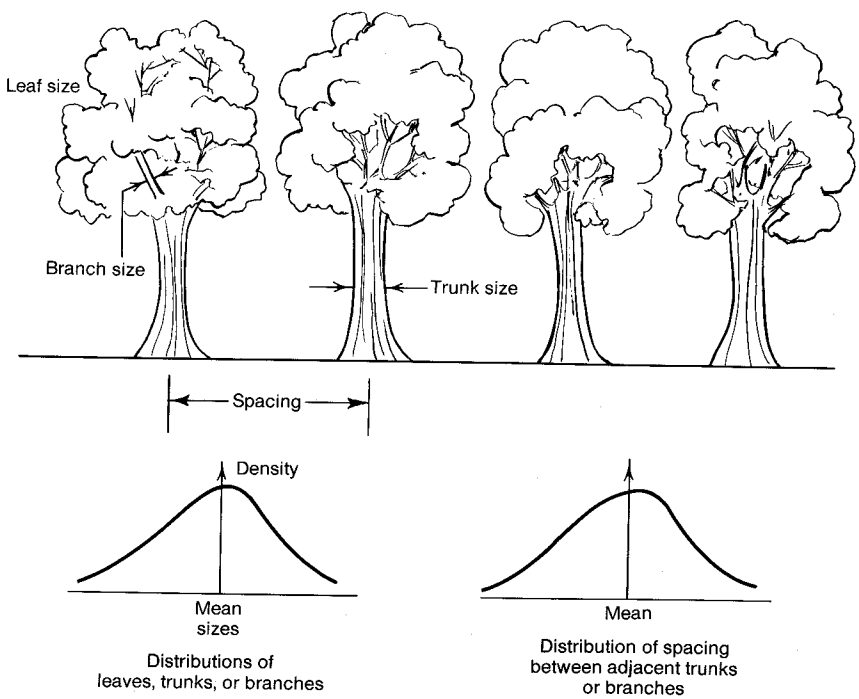


Figure 4.9 A characteristic of foliage environment.

thick that the signal can hardly penetrate. In this case, the signal will propagate from the top of the tree and deflect to the mobile receiver. We will include this calculation also.

Sometime the foliage loss can be treated as a wire-line loss, in decibels per foot or decibels per meter, when the foliage is uniformly heavy and the path lengths are short. When the path length is long and the foliage is nonuniform, then decibels per octaves or decibels per decade is used. Detailed discussion of foliage loss can be found in Refs. 34 to 38. In general, foliage loss occurs with respect to the frequency to the fourth power ($\sim f^{-4}$). Also, at 800 MHz the foliage loss along the radio path is 40 dB/dec, which is 20 dB more than the free-space loss, with the same amount of additional loss for mobile communications. Therefore, if the situation involves both foliage loss and mobile communications, the total loss would be 60 dB/dec (=20 dB/dec of free-space loss + additional 20 dB due to foliage loss + additional 20 dB due to mobile communication). This situation would be the case if the foliage would line up along the radio path. A foliage loss in a suburban area of 58.4 dB/dec is shown in Fig. 4.10.

Example 4.9 In a suburban area two places are covered with trees: 2 to 2.5 mi away from the cell site and 3 to 3.5 mi away from the cell site. The additional loss due to foliage is 3 dB, according to Fig. 4.10.

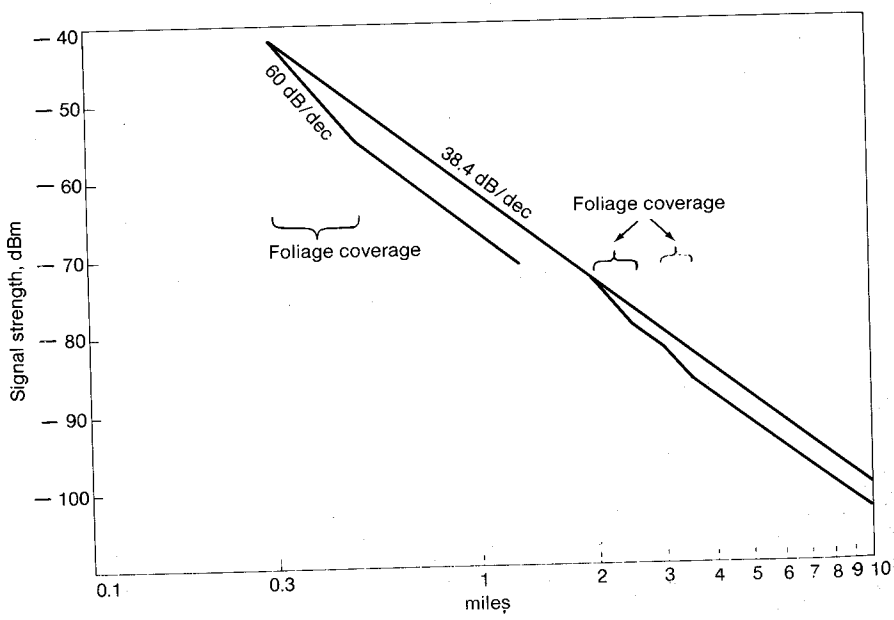


Figure 4.10 Foliage loss calculation in suburban areas.

Example 4.10 In a suburban area, one place is 0.3 to 0.5 mi (a distance of 1056 ft) from the cell site with additional trees. The additional path loss is 5 dB due to the foliage, according to Fig. 4.10.

As demonstrated from the above two examples, close-in foliage at the transmitter site always heavily attenuates signal reception. Therefore, the cell site should be placed away from trees. If the heavy foliage is close in at the mobile unit, the additional foliage loss must be calculated using the diffraction loss formula given in Sec. 4.7.3.

4.5 Propagation in Near-in Distance

4.5.1 Why use a 1-mi intercept?

1. Within a 1-mi radius, the antenna beamwidth, especially of a high-gain omnidirectional antenna, is narrow in the vertical plane. Thus the signal reception at a mobile unit less than 1 mi away will be reduced because of the large elevation angle which causes the mobile unit to be in the shadow region (outside the main beam). The larger the elevation angle, the weaker the reception level due to the antenna's vertical pattern, as shown in Fig. 4.11.
2. There are fewer roads within the 1-mi radius around the cell site. The data are insufficient to create a statistical curve. Also the road orientation, in-line and perpendicular, close to the cell site can cause a big difference in signal reception levels (10–20 dB) on those roads.
3. The near-by surroundings of the cell site can bias the reception level either up or down when the mobile unit is within the 1-mi radius. When the mobile unit is 1 mi away from the cell site, the effect due to the near-by surroundings of the cell site becomes negligible.

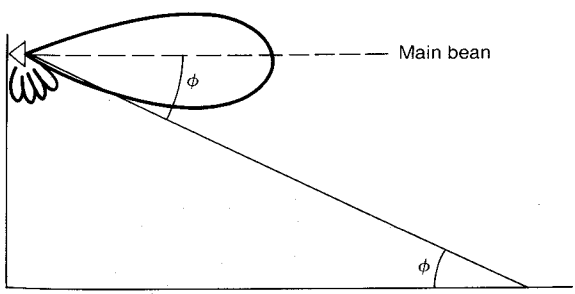


Figure 4.11 Elevation angle of the shadow of the antenna pattern.

4. For land-to-mobile propagation, the antenna height at the cell site strongly affects the mobile reception in the field; therefore, mobile reception 1 mi away has to refer to a given base-station antenna height.

4.5.2 Curves for near-in propagation

We usually worry about propagation at the far distance for coverage purposes. Now we also should investigate the near-in distance propagation. We may use the suburban area as an example. At the 1-mi intercept the received level is −61.7 dBm based on the reference set of parameters; i.e., the antenna height is 30 m (100 ft). If we increase the antenna height to 60 m (200 ft), a 6-dB gain is obtained. From 60 to 120 m (200 to 400 ft), another 6 dB is obtained. At the 120-m (400-ft) antenna height, the mobile received signal is the same as that received at the free space.

The antenna pattern is not isotropic in the vertical plane. A typical 6-dB omnidirectional antenna vertical beamwidth is shown in Fig 4.12. The reduction in signal reception can be found in the figure and is listed in the table below.

At d = 100 m (328 ft) [mobile antenna height = 3 m (10 ft)], the incident angles and elevation angles are 11.77° and 10.72°, respectively.

Antenna height h_1, m (ft)	Incident angle θ, degrees	Elevation angle ϕ, degrees	Attentuation α, dB
90 (300)	30.4	29.6	21
60 (200)	21.61	20.75	16
30 (100)	11.77	10.72	6

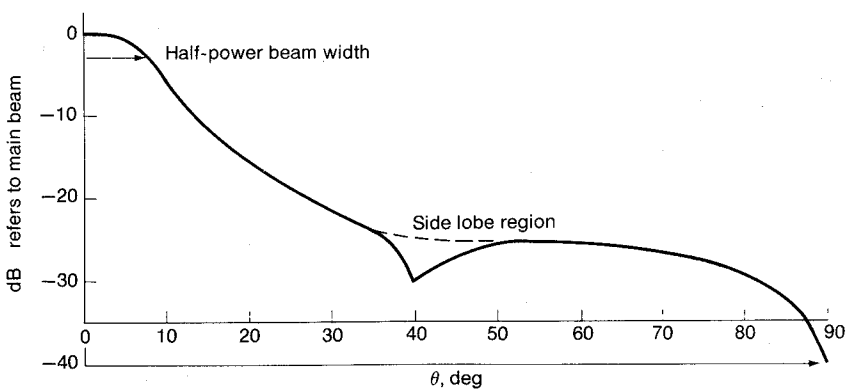


Figure 4.12 A typical 6-dB omnidirectional antenna beamwidth.

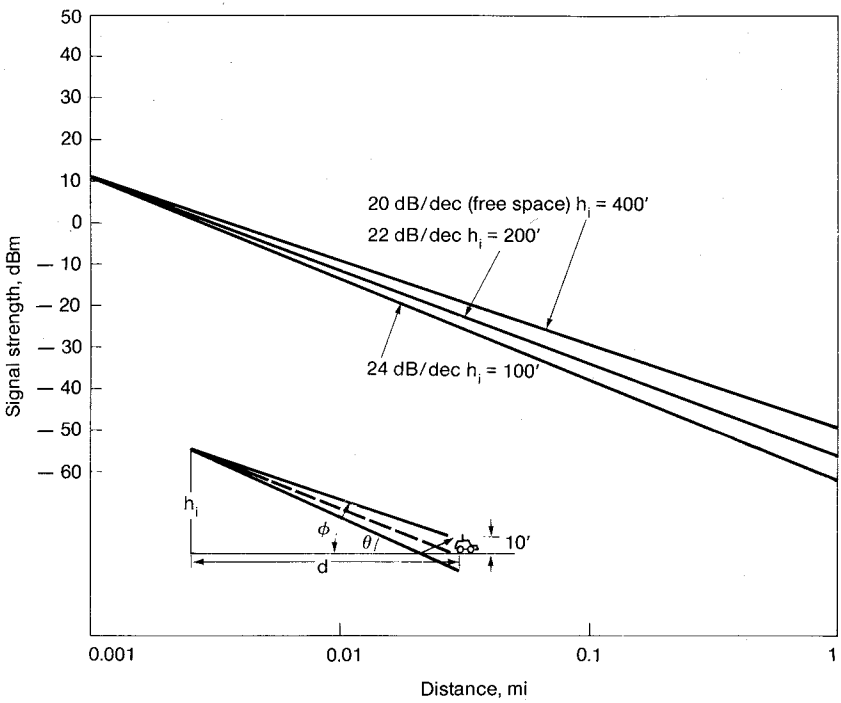


Figure 4.13 Curves for near-in propagation.

Since the incident angle becomes larger, the 40-dB/dec slope is no longer valid. If the antenna beam is aimed at the mobile unit, we will observe 24 dB/dec for an antenna height of 100 ft, 22 dB/dec for an antenna height of 200 ft, and 20 dB/dec for an antenna height of 400 ft or higher. The slope of 20 dB/dec is the free-space loss as shown in Fig. 4.13. The power of 11 dBm received at 0.001 mi is obtained from the free-space formula with an ERP of 46 dBm at the cell site as the standard condition (Table 4.1).

4.6 Long-Distance Propagation

The advantage of a high cell site is that it covers the signal in a large area, especially in a noise-limited system where usually different frequencies are repeatedly used in different areas. However, we have to be aware of the long-distance propagation phenomenon. A noise-limited system gradually becomes an interference-limited system as the traffic increases. The interference is due to not only the existence of many cochannels and adjacent channels in the system, but the long-distance propagation also affects the interference.

4.6.1 Within an area of 50-mile radius

For a high site, the low-atmospheric phenomenon would cause the ground wave path to propagate in a non-straight-line fashion. The phenomenon is usually more pronounced over seawater because the atmospheric situation over the ocean can be varied based on the different altitudes. The wave path can bend either upward or downward. Then we may have the experience that at one spot the signal may be strong at one time but weak at another.

4.6.2 At a distance of 320 km (200 mi)

Tropospheric wave propagation prevails at 800 MHz for long-distance propagation; sometimes the signal can reach 320 km (200 mi) away.

The wave is received 320 km away because of an abrupt change in the effective dielectric constant of the troposphere (10 km above the surface of the earth). The dielectric constant changes with temperature, which decreases with height at a rate of about 6.5°C/km and reaches −50°C at the upper boundary of the troposphere. In tropospheric propagation, the wave may be divided by refraction and reflection.

Tropospheric refraction. This refraction is a gradual bending of the rays due to the changing effective dielectric constant of the atmosphere through which the wave is passing.

Tropospheric reflection. This reflection will occur where there are abrupt changes in the dielectric constant of the atmosphere. The distance of propagation is much greater than the line-of-sight propagation.

Moistness. Actually water content has much more effect than temperature on the dielectric constant of the atmosphere and on the manner in which the radio waves are affected. The water vapor pressure decreases as the height increases.

If the refraction index decreases with height over a portion of the range of height, the rays will be curved downward, and a condition known as *trapping,* or *duct propagation,* can occur. There are surface ducts and elevated ducts. Elevated ducts are due to large air masses and are common in southern California. They can be found at elevations of 300 to 1500 m (1000 to 5000 ft) and may vary in thickness from a few feet to a thousand feet. Surface ducts appear over the sea and are about 1.5 m (5 ft) thick. Over land areas, surface ducts are produced by the cooling air of the earth.

Tropospheric wave propagation does cause interference and can only

be reduced by umbrella antenna beam patterns, a directional antenna pattern, or a low-power-low-antenna-mast approach.

4.7 Obtain Path Loss from a Point-to-Point Prediction Model—A General Approach[24,41]

4.7.1 In nonobstructive condition[42]

In this condition, the direct path from the cell site to the mobile unit is not obstructed by the terrain contour. Here, two general terms should be distinguished. The *nonobstructive direct path* is a path unobstructed by the terrain contour. The *line-of-sight path* is a path which is unobstructed by the terrain contour and by man-made structures. In the former case, the cell-site antenna cannot be seen by the mobile user whereas in the latter case, it can be. Therefore, the signal reception is very strong in the line-of-sight case, which is not the case we are worrying about.

In the mobile environment, we do not often have line-of-sight conditions. Therefore we use direct-path conditions which are unobstructed by the terrain contour. Under these conditions, the antenna-height gain will be calculated for every location in which the mobile unit travels, as illustrated in Fig. 4.14. The method for finding the antenna-height gain is as follows.

Finding the antenna-height gain

1. Find the specular reflection point. Take two values from two conditions stated as follows.
   a. Connect the image antenna of the cell-site antenna to the mobile antenna; the intercept point at the ground level is considered as a potential reflection point.
   b. Connect the image antenna of the mobile antenna to the cell-site antenna; the intercept point at the ground level is also considered as a potential reflection point.

   Between two potential reflection points we choose the point which is close to the mobile unit to be the real one because more energy would be reflected to the mobile unit at that point.
2. Extend the reflected ground plane. The reflected ground plane which the reflection point is on can be generated by drawing a tangent line to the point where the ground curvature is, then extending the reflected ground plane to the location of the cell-site antenna.
3. Measure the effective antenna height. The effective antenna height is measured from the point where the reflected ground plan and the cell-site antenna location meet. Between these two cases shown

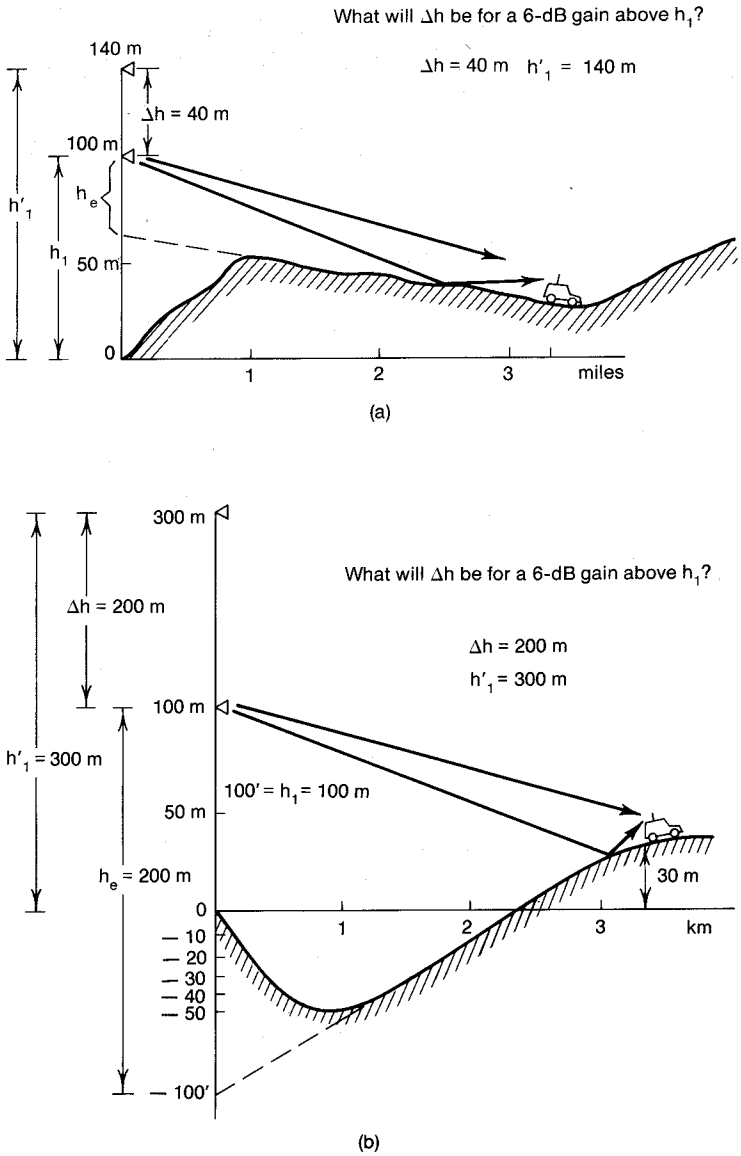


Figure 4.14 Calculation of effective antenna height: (*a*) case 1; (*b*) case 2.

in Fig. 4.14, h_e equals 40 m in Fig. 4.14*a* and 200 m in Fig. 4.14*b*. The actual antenna height h_1 is 100 m.

4. Calculate the antenna-height gain ΔG. The formula of ΔG is expressed as [see Eq. (4.2-10*b*)]

$$G = 20 \log \frac{h_e}{h_1} \tag{4.7-1}$$

Then the ΔG from Fig. 4.14*a* is

$$\Delta G = 20 \log \frac{40}{100} = -8 \text{ dB} \qquad \text{(a negative gain in Fig. 4-14}a\text{)}$$

The ΔG from Fig. 4.14*b* is

$$\Delta G = 20 \log \frac{200}{100} = 6 \text{ dB} \qquad \text{(a positive gain in Fig. 4.14}b\text{)}$$

We have to realize that the antenna-height gain ΔG changes as the mobile unit moves along the road. In other words, the effective antenna height at the cell site changes as the mobile unit moves to a new location, although the actual antenna remains unchanged.

Another physical explanation of effecitve antenna height. Another physical explanation of effective antenna height is shown in Fig. 4.15. In Fig. 4.15*a*, we have to ask which height is the actual antenna height

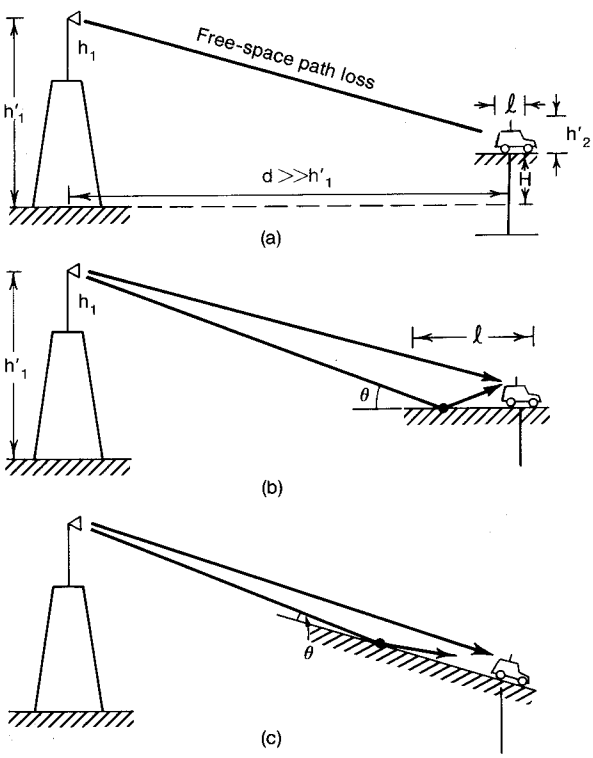


Figure 4.15 Physical explanation of effective antenna heights.

h_1, or is the actual antenna height very important in this situation? As long as the value of H is much larger than h_2 and the length l of the floor is roughly equal to the length of the vehicle, there is only one direct wave, and the free-space path loss is applied to the situation which provides a strong reception.

In Fig. 4.15*b*, the situation remains the same, except the length l is longer to allow a reflection point to be generated on the floor. Now two waves are created, one direct wave and one reflected wave. The stronger the reflected wave is, the larger the path loss is. The stronger reflected wave occurs at a very small incident angle θ. This means that a small incident angle corresponds to a large reflection coefficient because of the nature of the reflection mechanism, and the wave reflected from the ground is a 180° phase shift. Therefore, no matter what, the amount of reflected energy always becomes negative. The addition of a strong reflected wave to a direct wave tends to weaken the direct wave.

In Fig. 4.15*c*, as the incident angle θ approaches zero, the signal reception becomes very weak. The shadow-loss condition starts when both the direct wave and the reflected wave have been blocked. When the direction of the vehicle-site floor is reversed (i.e., going counterclockwise), the incident angle θ increases and the reflection coefficient decreases. The energy reflected from the floor becomes less, and so the direct wave would reduce the small amount of energy resulting from the negative contribution from the reflected wave. The larger the incident angle of the reflected wave, the weaker the reflected wave, and the signal reception becomes the free-space condition.

When the incident angle of a wave is very small, two conditions shown in Fig. 4.16 can be considered.

1. Sparse man-made structures or trees along the propagation path. When there are few man-made structures along the propagation path, the received power is always higher than when there are many. This is why the power level received in an open area is higher than that received in a suburban area and higher still than that received in an urban area.
2. Dense man-made structures along the propagation path. There are two conditions.
   a. A line-of-sight wave exists between the base station and the mobile unit. When the waves reflected by the surrounding buildings are relatively weak, less fading (rician fading)[40] is observed.
   b. The mobile unit is surrounded by the scatterers. If the direct reception is blocked by the surrounding buildings, Rayleigh fading is observed.

   In the above two conditions the average received powers are not the same. However, if the reflected waves from surrounding build-

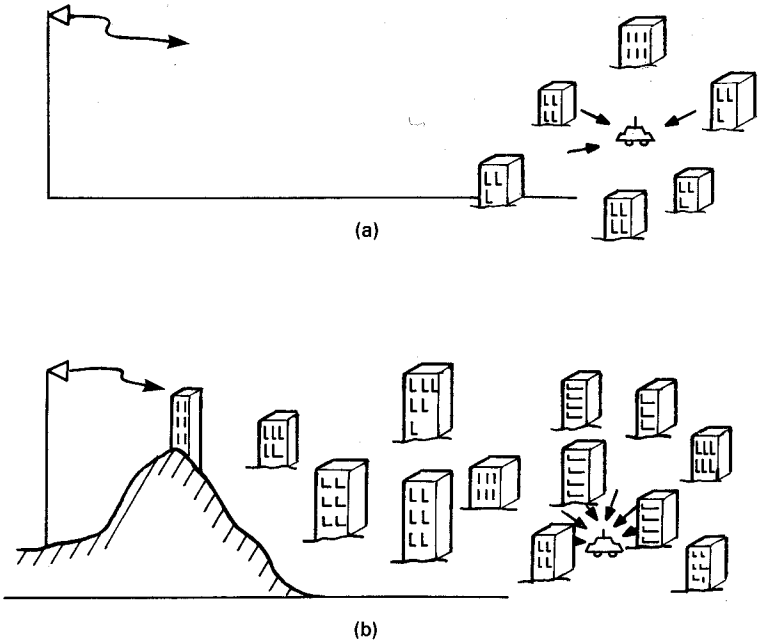


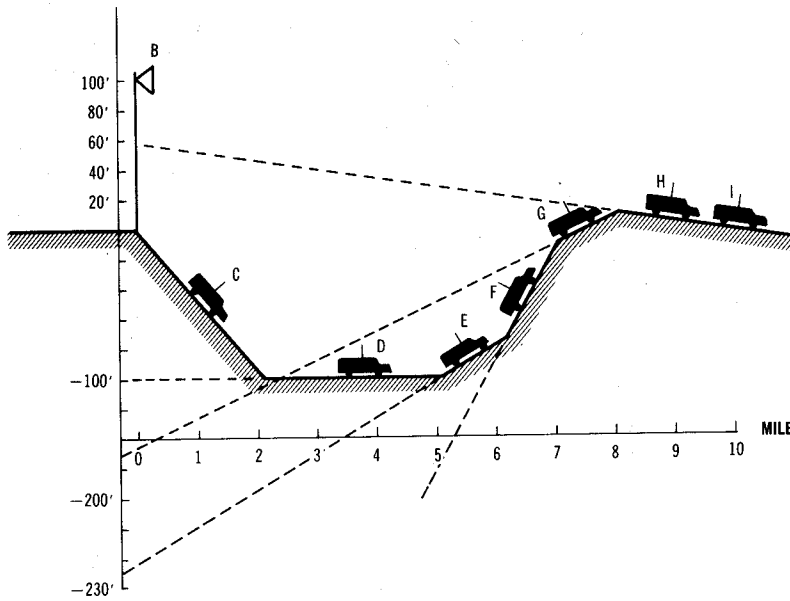
Figure 4.16 Man-made environment. (*a*) Sparse man-made structures. (*b*) Dense man-made structures.

ings are very strong, the average received power from the two different conditions can be very close. It can be seen as an analog to conservation of energy. The total signal received at the mobile unit (or at the cell site) either from a single wave or from many reflected waves tentatively remains a constant. In both conditions, the propagation path loss is 40 dB/dec because both conditions are in a mobile radio environment.

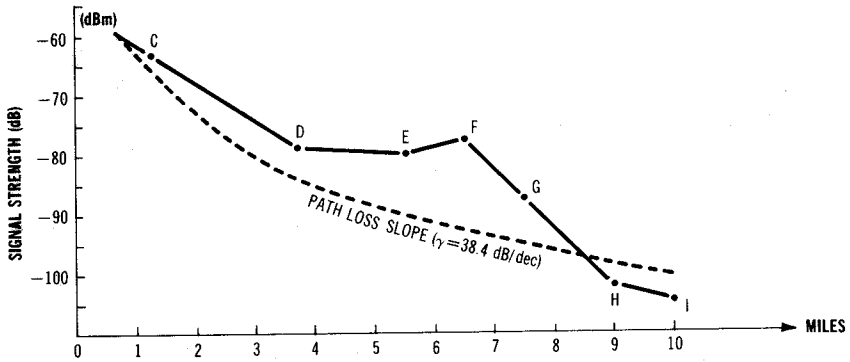
Comments on the contribution of antenna-height gain. If we do not take into account the changes in antenna-height gain due to the terrain contour between the cell site and the mobile unit the path-loss slope will have a standard deviation of 8 dB. If we do take the antenna-height gain into account, values generally have a standard deviation within 2 to 3 dB.

The effects of terrain roughness are illustrated in Fig. 4.17*a* as changing different effective antenna heights, h_e and h_e' at different positions of the mobile unit. Then the effective antenna gain ΔG can be obtained from Eq. (4.7-1) as

$$\Delta G = 20 \log \frac{h_e}{h_e'}$$



(a)



(b)

Figure 4.17 Illustration of the terrain effect on the effective antenna gain at each position. (*a*) Hilly terrain contour. (*b*) Point-to-point prediction. (*After Lee, Ref. 40, p. 86.*)

Assume that the mobile unit is traveling in a suburban area, say northern New Jersey. The path-loss slope of this suburban area is shown in Fig. 4.3 and then plotted in Fig. 4.17*b*. Then the antenna-height gains or losses are added or subtracted from the slope at their corresponding points. Now we can visualize the difference between an area-to-area prediction (use a path-loss slope) and a point-to-point pre-

diction (after the antenna-height gain correction). The point-to-point prediction is based on the actual terrain contour along a particular radio path (in this case, the radio path and the mobile path are the same for simplicity), but the area-to-area prediction is not. This is why the area-to-area prediction has a standard deviation of 8 dB but the point-to-point prediction only has a standard deviation of less than 2 to 3 dB (see Sec. 4.8.2).

4.7.2 In obstructive condition

In this condition, the direct path from the cell site to the mobile unit is obstructed by the terrain contour. We would like to treat this condition as follows.

1. *Apply area-to-area prediction.* First, just apply the same steps in the area-to-area prediction as if the obstructive condition did not exist. If the area is in Philadelphia, the Philadelphia path-loss slope applies. All the correction factors would apply to finding the area-to-area prediction for a particular situation.
2. *Obtain the diffraction loss.* The diffraction loss can be found from a single knife-edge or double knife-edge case, as shown in Fig. 4.18.
   a. Find the four parameters for a single knife-edge case. The four parameters, the distances r_1 and r_2 from the knife-edge to the

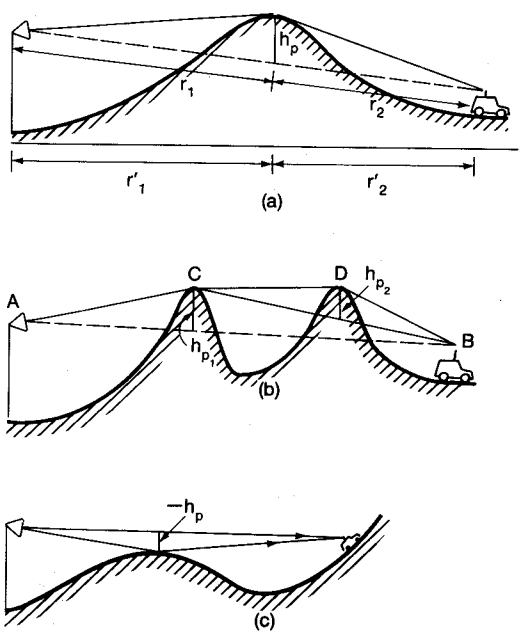


Figure 4.18 Diffraction loss due to obstructive conditions. (*a*) Single knife-edge; (*b*) double knife-edges; (*c*) nonclear path.

cell site and to the mobile unit, the height of the knife-edge h_p, and the operating wavelength λ, are used to find a new parameter v.

$$v = -h_p \sqrt{\frac{2}{\lambda}\left(\frac{1}{r_1} + \frac{1}{r_2}\right)} \tag{4.7-2}$$

h_p is a positive number as shown in Fig. 4.18*a*, and h_p is a negative number as shown in Fig. 4.18*c*. As soon as the value of v is obtained, the diffraction loss L can be found from the curves shown in Fig. 4.19. The approximate formula below can be used with different values of v to represent the curve and be programmed into a computer.

$$\begin{aligned}
1 \le v \quad & L = 0 \text{ dB} \\
0 \le v < 1 \quad & L = 20 \log (0.5 + 6.2v) \\
-1 \le v < 0 \quad & L = 20 \log (0.5e^{0.95v}) \\
-2.4 \le v < -1 \quad & L = 20 \log (0.4 - \sqrt{0.1184 - (0.1\,v + 0.38)^2} \\
v < -2.4 \quad & L = 20 \log \left(-\frac{0.225}{v}\right)
\end{aligned} \tag{4.7-3}$$

When $h_p = 0$, the direct path is tangential to the knife-edge, and $v = 0$, as derived from Eq. (4.7-2). With $v = 0$, the diffraction loss $L = 6$ dB can be obtained from Fig. 4.19.

b. A double knife-edge case. Two knife edges can be formed by the two triangles *ACB* and *CDB* shown in Fig. 4.18*b*. Each one can be used to calculate v as v_1 and v_2. The corresponding L_1 and L_2 can be found from Fig. 4.19. The total diffraction loss of this double knife-edge model is the sum of the two diffraction losses.

$$L_t = L_1 + L_2$$

4.7.3 Cautions in obtaining defraction loss

We always draw the scales of the y and x axes differently. The same intervals represent 10 m or 10 ft in the y-axis but 1 km or 1 mi in the x-axis. In this way we can depict the elevation change more clearly. Then we have to be aware of the measurement of r_1 and r_2 shown in Fig. 4.18*a*. The simple way of measuring r_1 and r_2 is based on the horizontal scale as shown in the figure, where $r_1 = r_1'$ and $r_2 = r_2'$. It

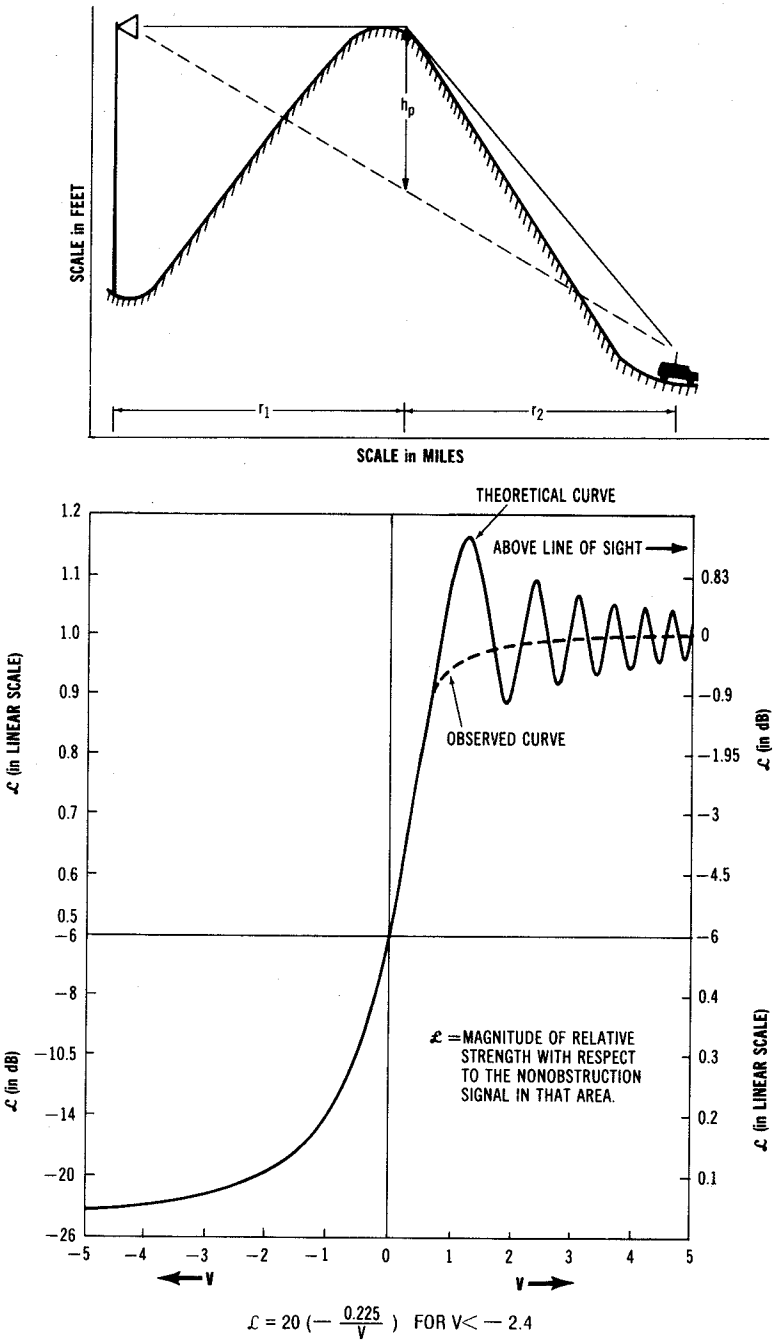


Figure 4.19 Shadow-loss prediction.

can be shown that the errors for $r_1 = r_1'$ and $r_2 = r_2'$ are insignificant if the scales used on both the x and y axes are the same.

When the heavy foliage is close in at the mobile unit, the loss due to foliage can be obtained from the diffraction loss. The average foliage configuration resembles the terrain configuration. Therefore the height of knife edge over the foliage configuration can be found. Then the diffraction loss due to the foliage is obtained.

4.8 Form of a Point-to-Point Model

4.8.1 General formula

We form the model as follows:

$$P_r = \begin{cases} \text{nonobstructive path} = P_{r_0} - \gamma \log \dfrac{r}{r_0} + 20 \log \dfrac{h_e'}{h_1} + \alpha \\ \text{obstructive path} = P_{r_0} - \gamma \log \dfrac{r}{r_0} + L + \alpha \\ \text{land-to-mobile over water} = \text{a free-space formula} \end{cases} \tag{4.8-1}$$

(see Sec. 4.3)

Remarks

1. The P_r cannot be higher than that from the free-space path loss.
2. The road's orientation, when it is within 2 mi from the cell site, will affect the received power at the mobile unit. The received power at the mobile unit traveling along an in-line road can be 10 dB higher than that along a perpendicular road.
3. α is the corrected factor (gain or loss) obtained from the condition (see Sec. 4.2.1).
4. The foliage loss (Sec. 4.4) would be added depending on each individual situation. Avoid choosing a cell site in the forest. Be sure that the antenna height at the cell site is higher than the top of the trees.

4.8.2 The merit of the point-to-point model

The area-to-area model usually only provides an accuracy of prediction with a standard deviation of 8 dB, which means that 68 percent of the actual path-loss data are within the ±8 dB of the predicted value. The uncertainty range is too large. The point-to-point model reduces the uncertainty range by including the detailed terrain contour information in the path-loss predictions.

The differences between the predicted values and the measured ones for the point-to-point model were determined in many areas. In the following discussion, we compare the differences shown in the Whippany, N.J., area and the Camden-Philadelphia area. First, we plot the points with predicted values at the x-axis and the measured values at the y-axis, shown in Fig. 4.20. The 45° line is the line of prediction without error. The dots are data from the Whippany area, and the crosses are data from the Camden-Philadelphia area. Most of them, except the one at 9 dB, are close to the line of prediction without error. The mean value of all the data is right on the line of prediction without error. The standard deviation of the predicted value is 0.8 dB from the measured one.

In other areas, the differences were slightly larger. However, the standard deviation of the predicted value never exceeds the measured one by more than 3 dB. The standard deviation range is much reduced as compared with the maximum of 8 dB from area-to-area models.

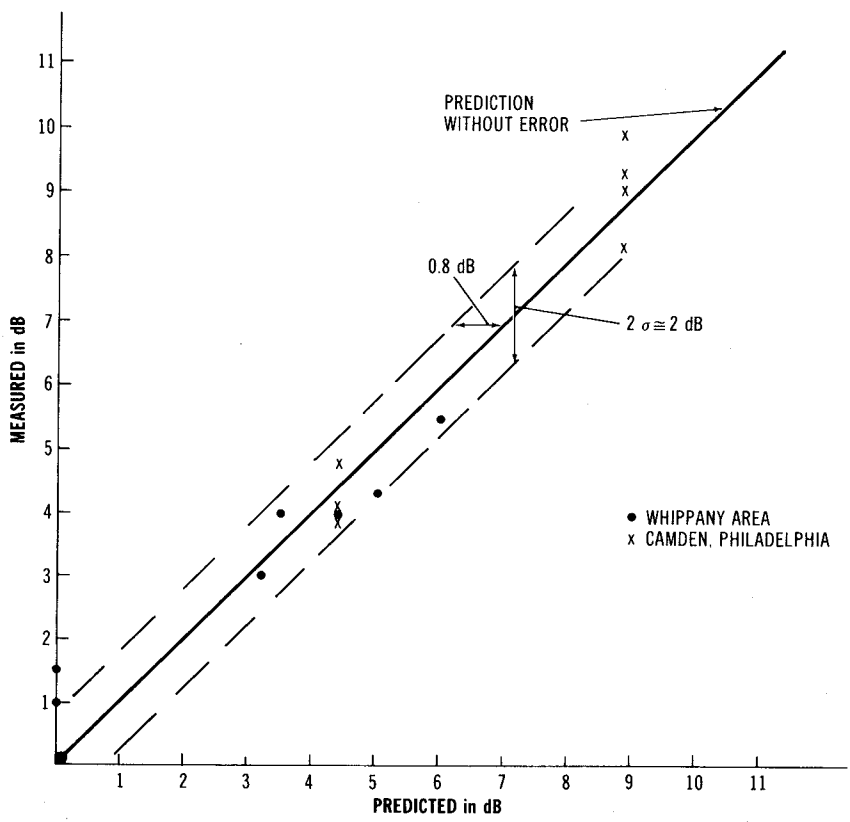


Figure 4.20 Indication of errors in point-to-point predictions under nonobstructive conditions. (*After Lee, Ref. 42.*)

The point-to-point model is very useful for designing a mobile cellular system with a radius for each cell of 10 mi or less. Because the data follow the log-normal distribution, 68 percent of predicted values obtained from a point-to-point prediction model are within 2 to 3 dB.

This point-to-point prediction can be used to provide overall coverage of all cell sites and to avoid cochannel interference. Moreover, the occurrence of handoff in the cellular system can be predicted more accurately.

The point-to-point prediction model is a basic tool that is used to generate a signal coverage map, an interference area map, a handoff occurrence map, or an optimum system design configuration, to name a few applications.

4.9 Computer Generation of a Point-to-Point Prediction

The point-to-point prediction described in Sec. 4.8 can easily be used in a computer program. Here we describe the automated prediction in steps.[43]

4.9.1 Terrain elevation data

We may use either a 250,000:1 scale map, called a *quarter-million scale map,* or a 7.5-minute scale map issued by the U.S. Geological Survey. Both maps have the terrain elevation contours. Also, terrain elevation data tapes can be purchased from the DMA (Defense Map Agency). The quarter-million scale map tapes cover the whole United States, but the 7.5-minute scale maps are only available for certain areas. Let us discuss the use of these two different scale maps.

1. Use a quarter-million scale map. Each elevation contour increment is 100 ft, which does not provide the fine detail needed for a mobile radio propagation in a hilly area. The quarter-million scale elevation data tapes made by DMA come from the quarter-million scale maps. The elevation data for two adjacent terrain contours is determined by extrapolation. Although the tape gives elevations for every 208 ft (0.01 in on the map), these elevations are not accurate because they are from the quarter-million scale map. However, in most areas, as long as the terrain contour does not change rapidly, the DMA tape can be used as a raw data base. DMA provides two kinds of tapes: a 3-second arc tape and a planar 0.01-in tape. The former has an elevation for every 3-second interval (about 61 to 91.5 m, or 200 to 300 ft, depending on the geographic locations) on the map. The latter has an elevation at intervals of 0.01 in (63.5 m, or 208 ft). Since the arc-second tape

provides sample intervals of 3 seconds, a length of 1° has 1200 points. The advantage of using the arc-second tape is the continuity of sample points from one tape to another. However, for the arc-second tape the sample points are not equally far apart on the ground but are in the planar tape. Therefore, if the desired coverage is within one tape's geographic area, the planar tape is used. If the desired area is spread over more than one tape (see Fig. 4.21), the arc-second is used. The structure to a 250,000:1 scale digital elevation format is shown in Fig. 4.22.

2. Use a 7.5-minute map. A 7.5-minute map roughly covers 10 × 13 km^2, or 6 × 8 mi^2. The increment of elevation between two contours is 3 or 6 m (10 or 20 ft). The fine resolution of elevation data proves to be useful for propagation prediction. There are three ways to deal with the 7.5-minute map.
   a. Divide a 7.5-minute scale map into either a 30 × 45 grid map, where each grid is 300 × 300 m (1000 × 1000 ft), or into a 15 × 22.5 grid map, where each grid is 600 × 600 m (2000 × 2000 ft). An eyeball estimate of the elevation value in each grid is quite adequate.
   b. Use DMA tapes in the area if the 7.5-minute tape is available. The 7.5-minute tape can have 150 3-second points. Therefore, a quarter-million scale tape can be replaced by 8 × 8 = 64 7.5 minute maps. The 7.5-minute tapes are used in those areas of the quarter-million map tapes when the terrain contour changes rapidly.

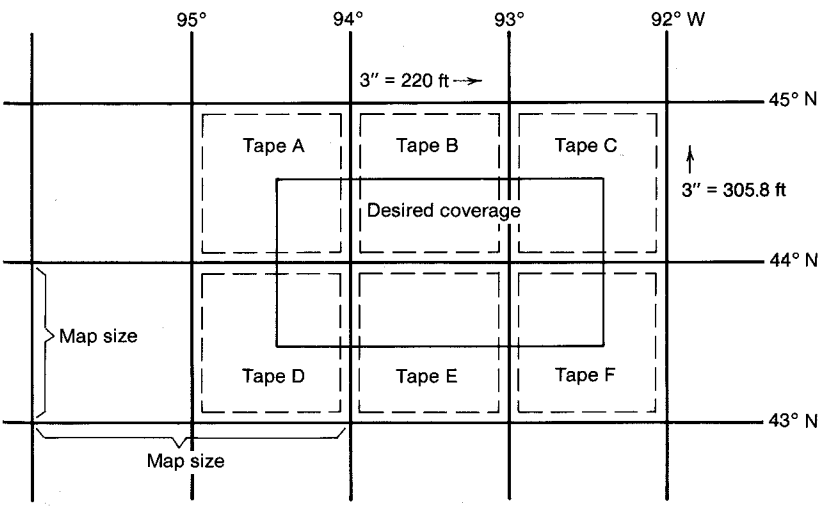


Figure 4.21 A coverage by six tapes.

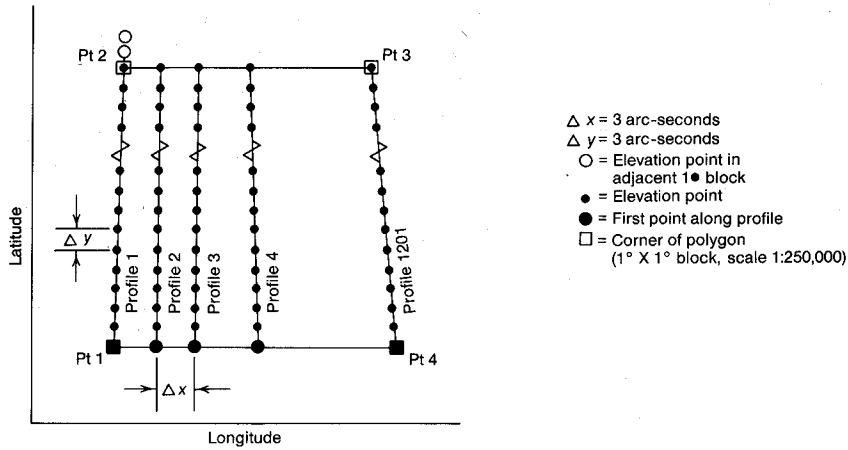


Figure 4.22 Structure of a 250,000:1-scale digital elevation format.

c. The elevation contour line of a 7.5-minute map with the same elevation contour can be digitized in different sample points on the map and stored in a database or on a tape. Then any two points along the terrain elevation contour can be plotted based on the actual contour lines. This is the most accurate method; however, sometimes the accuracy obtained from item *a* is sufficient for predicting the path loss.

4.9.2 Elevation map

We prefer to use the arc-second tape since the continuity of sample points from tape to tape simplifies the calculation. In the area of N40° latitude, the average elevation of a 2000 × 2000 ft (roughly) grid can be found by taking 7 samples in latitude (a length of 2141 ft) and 10 samples in longitude (a length of 2200 ft).

$$\text{Average elevation} = \frac{\sum_{1}^{70} \text{sample elevations}}{70} \qquad \text{(in one grid)}$$

Figure 4.23 shows an elevation map whose grids are approximately 2200 × 2200 ft in area. The average elevation of each grid is given along with a (Y, X) tag.

4.9.3 Elevation contour

Assuming that a transmitter is in a grid (20, 30) and the receiver is in a grid (23, 37), then an elevation contour can be plotted with an

$$H = \frac{\sum_{i=1}^{70} \text{sample elevations}}{70} = \text{average elevation}$$

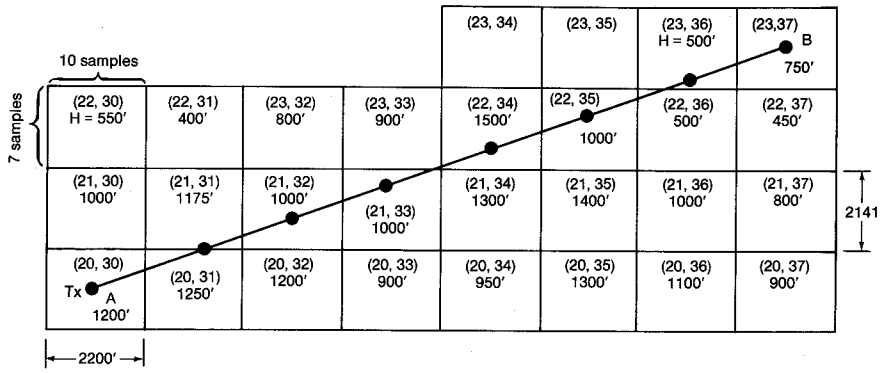


Figure 4.23 Elevation map.

increment of 2200 ft for every elevation point in its corresponding grid, as shown in Fig. 4.24.

Assume that the antenna height is 100 m (300 ft) at the transmitter and 3 m (10 ft) at the receiver. We can plot a line between two ends and see that shadow-loss condition exists. An example the path loss calculation for this particular path is as follows.

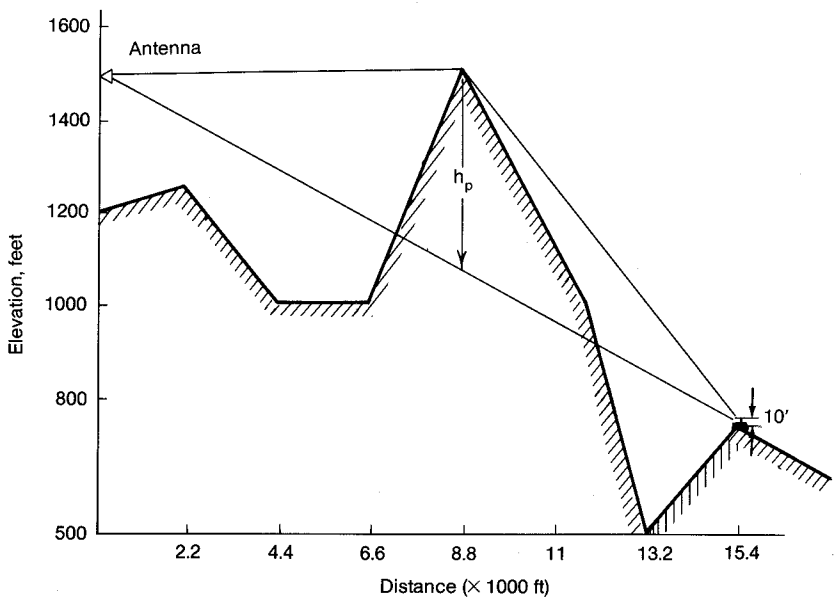


Figure 4.24 Elevation contour.

TABLE 4.3 Work Sheet for Example 4.11

Classification of area: suburban
Distance: 15,400 ft = 2.92 mi
From the curve: (see Fig. 4.3)
P_r = 79.5 dBm

New data:	*Corrections:*
Transmitter power = 5 W	− 3 dB
Antenna gain = 2 dB per dipole	− 4 dB
Antenna height = 300 ft, $20 \log \frac{300}{100}$ =	+ 9.5 dB
	+ 2.5 dB

For flat-terrain case:
New path loss $P_r' = -79.5 \text{ dBm} + 2.5 \text{ dB} = -77 \text{ dBm}$
For shadow-region case:
$r_1 = 8800$ ft $\quad r_2 = 6600$ ft $\quad h_p = 450$ ft $\quad f = 850$ MHz
$v = -10.35$
$L = -33 \text{ dB} \quad \left[L = 20 \log \left(-\frac{0.225}{v}\right)\right]$
New path loss $P_r' = -77 \text{ dB} - 33 \text{ dB} = -110 \text{ dBm}$

Example 4.11 If the transmitter power is 5 W, the base station antenna gain is 2 dB above dipole, and its height is 300 ft, calculate the path loss from the path shown in Fig. 4.24. The results are given in Table 4.3.

4.10 Cell-Site Antenna Heights and Signal Coverage Cells

4.10.1 Effects of cell-site antenna heights

There are several points which need to be clarified concerning cell-site antenna-height effects.

Antenna height unchanged. If the power of the cell-site transmitter changes, the whole signal-strength map (obtained from Sec. 4.9) can be linearly updated according to the change in power.

If the transmitted power increases by 3-dB, just add 3 dB to each grid in the signal-strength map. The relative differences in power among the grids remain the same.

Antenna height changed. If the antenna height changes ($\pm\Delta h$), then the whole signal-strength map obtained from the old antenna height cannot be updated with a simple antenna gain formula as

$$\Delta g = 20 \log \frac{h_1'}{h_1} \tag{4.10-1}$$

where h_1 is the old actual antenna height and h_1' is the new actual antenna height. However, we can still use the same terrain contour

data along the radio paths (from the cell-site antenna to each grid) to figure out the difference in gain resulting from the different effective antenna heights in each grid.

$$\Delta g' = 20 \log \frac{h_e'}{h_e} = 20 \log \frac{h_e \pm \Delta h}{h_e} \tag{4.10-2}$$

where h_e is the old effective antenna height and h_e' is the new effective antenna height. The additional gain (increase or decrease) will be added to the signal-strength grid based on the old antenna height.

Example 4.12 If the old cell-site antenna height is 30 m (100 ft) and the new one h_1', is 45 m, the mobile unit 8 km (5 mi) away sees the old cell-site effective antenna height (h_e) being 60 m. The new cell-site effective antenna height h_e' seen from the same mobile spot can be derived.

$$h_e' = h_e + (h_1' - h_1) = h_e + (h_e' - h_e) = h_e + \Delta h = 60 + (45 - 30) = 75 \text{ m}$$

Since the difference between two actual antenna heights is the same as the difference between the two corresponding effective antenna heights seen from each grid, the additional gain (or loss) based on the new change of actual antenna height is

$$\Delta g' = 20 \log \frac{h_e'}{h_e} = 20 \log \left(1 + \frac{h_1' - h_1}{h_e}\right) \tag{E4.12-1}$$

Location of the antenna changed. If the location of the antenna changes, the point-to-point program has to start all over again. The old point-to-point terrain contour data are no longer useful. The old effective antenna height seen from a distance will be different when the location of the antenna changes, and there is no relation between the old effective antenna height and the new effective antenna height. Therefore, every time the antenna location changes, the new point-to-point prediction calculation starts again.

Visualization of the effective antenna height. The effective antenna height changes when the location of the mobile unit changes. Therefore, we can visualize the effective antenna height as always changing up or down while the mobile unit is moving. This kind of picture should be kept in mind. In addition, the following facts would be helpful.

Case 1: The mobile unit is driven up a positive slope (up to a high spot). The effective antenna height increases if the mobile unit is driving away from the cell-site antenna, and it decreases if the mobile unit is approaching the cell-site antenna. (See Fig. 4.25*a*.)

Case 2: The mobile unit is driven down a hill. The effective antenna height decreases if the mobile unit is driving away from the cell-site

antenna, and it increases if the mobile unit is approaching the cell-site antenna. (See Fig. 4.25*a*.)

4.10.2 Visualization of Signal Coverage Cells

A physical cell is usually visualized as a signal-reception region around the cell site. Within the region, there are weak spots called *holes*. This is always true when a cell covers a relative flat terrain. However, a cell can contain a hilly area. Then the coverage patterns of the cell will look like those shown in Fig. 4.25*b*. Here the two cell sites are separated by a river. Because of the shadow loss due to the river bank, cell site A cannot cover area A', but cell site B can. The same situation applies to cell site B in area B'. Now every time the vehicle enters area A', a handoff is requested as if it were in cell B.

Therefore, in most cases, the holes in one cell are covered by the other sites. As long as the processing capacity at the MTSO can handle excessive handoff, this overlapped arrangement for filling the holes is a good approach in a noninterference condition.

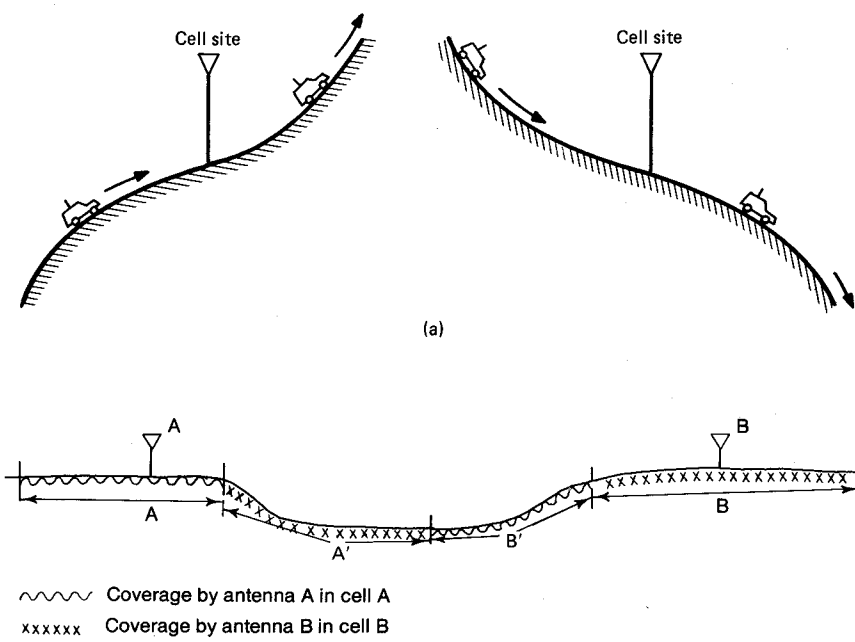


Figure 4.25 Different coverage concept. (*a*) Signal coverage due to effective antenna heights. (*b*) Signal coverage served by two cell sites.

4.11 Mobile-to-Mobile Propagation[44]

4.11.1 The transfer function of the propagation channel

In mobile-to-mobile land communication, both the transmitter and the receiver are in motion. The propagation path in this case is usually obstructed by buildings and obstacles between the transmitter and receiver. The propagation channel acts like a filter with a time-varying transfer function $H(f, t)$, which can be found in this section.

The two mobile units M_1 and M_2 with velocities V_1 and V_2, respectively, are shown in Fig. 4.26. Assume that the transmitted signal from M_1 is

$$s(t) = u(t)e^{j\omega t} \tag{4.11-1}$$

The receiver signal at the mobile unit M_2 from an ith path is

$$s_i = r_i u(t - \tau_i)e^{j[(\omega_0 + \omega_{1i} + \omega_{2i})(t - \tau_i) + \phi_i]} \tag{4.11-2}$$

where $u(t)$ = signal
ω_0 = RF carrier
r_i = Rayleigh-distributed random variable
ϕ_i = uniformly distributed random phase
τ_i = time delay on ith path

and

$$\begin{aligned}\omega_{1i} &= \text{Doppler shift of transmitting mobile unit on } i\text{th path} \\ &= \frac{2\pi}{\lambda} V_1 \cos \alpha_{1i}\end{aligned} \tag{4.11-3}$$

$$\begin{aligned}\omega_{2i} &= \text{Doppler shift of receiving mobile unit on } i\text{th path} \\ &= \frac{2\pi}{\lambda} V_2 \cos \alpha_{2i}\end{aligned} \tag{4.11-4}$$

where α_{1i} and α_{2i} are random angles shown in Fig. 4.26. Now assume that the received signal is the summation of n paths uniformly distributed around the azimuth.

$$\begin{aligned}s_r &= \sum_{i=1}^{n} s_i(t) = \sum_{i=1}^{n} r_i u(t - \tau_i) \\ &\qquad \times \exp\{j[(\omega_0 + \omega_{1i} + \omega_{2i})(t - \tau_i) + \phi_i]\} \\ &= \sum_{i=1}^{n} Q(\alpha_{i,t})u(t - \tau_i)e^{j\omega_0(t-\tau_i)}\end{aligned} \tag{4.11-5}$$

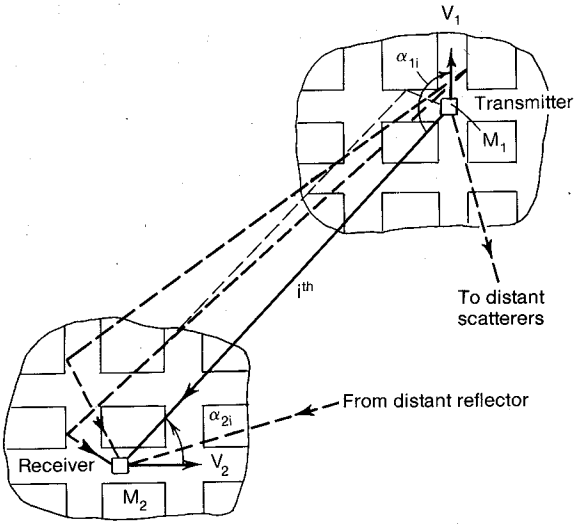


Figure 4.26 Vehicle-to-vehicle transmission.

where
$$Q(\alpha_i, t) = r_i \exp\{j[(\omega_{1i} + \omega_{2i})t + \phi_i']\} \tag{4.11-6}$$
$$\phi_i' = \phi - (\omega_{1i} + \omega_{2i})\tau_i \tag{4.11-7}$$

Equation (4.11-5) can be represented as a statistical model of the channel, as shown in Fig. 4.27.

Let $u(t) = e^{j\omega t}$, then Eq. (4.11-5) becomes

$$e_r(t) = \left[\sum_{i=1}^{n} Q(\alpha_i, t)e^{-j(\omega_0+\omega)\tau_i}\right]e^{j(\omega_0+\omega)t} = H(f, t)e^{j(\omega_0+\omega)t} \tag{4.11-8}$$

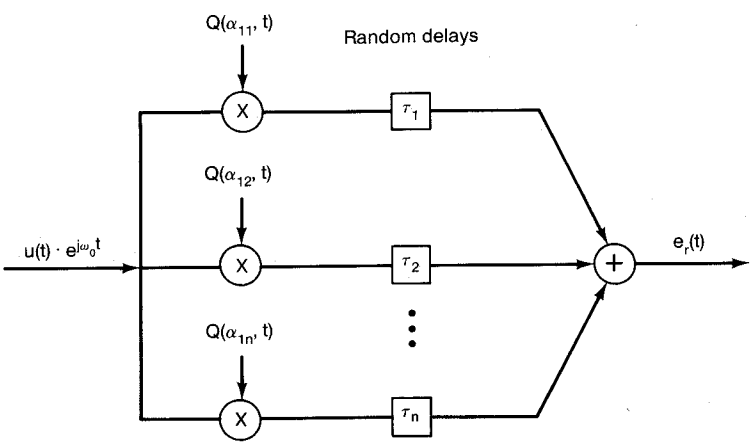


Figure 4.27 Statistical model for mobile-to-mobile channel.

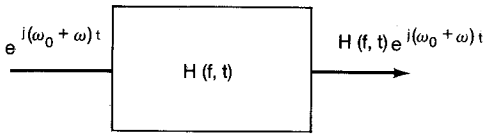


Figure 4.28 A propagation channel model.

Therefore
$$H(f, t) = \sum_{i=1}^{n} Q(\alpha_i, t)e^{-j(\omega_0+\omega)\tau_i} \tag{4.11-9}$$

where the signal frequency is $\omega = 2\pi f$. Equation (4.11-9) is expressed in Fig. 4.28.

Let $f = 0$; that is, only a sinusoidal carrier frequency is transmitted. The amplitude of the received signal envelope from Eq. (4.11-8) is

$$r = |H(0, t)| \tag{4.11-10}$$

where r is also a Rayleigh-distributed random variable with its average power of $2\sigma^2$ shown in the probability density function as

$$P(r) = \frac{r}{\sigma^2} e^{-r^2/2\sigma^2} \tag{4.11-11}$$

4.11.2 Spatial time correlation

Let $r_{x_1}(t_1)$ be the received signal envelope at position x_1 at time t_1. Then

$$r_{x_1}(t_1) = \sum_{i=1}^{n} r_i \exp j\left[(\omega_{1i} + \omega_{2i}) + \phi_i + \frac{2\pi}{\lambda} x_1 \cos \alpha_{1i}\right] \tag{4.11-12}$$

The same equation will apply to R_x at position x_2 at time t_2.

The spatial time-correlation function of the envelope is given by

$$R(x_1, x_2, t_1, t_2) = \tfrac{1}{2} \langle r_{x_1}(t_1) r^*_{x_2}(t_2)\rangle \tag{4.11-13}$$

assuming that the random process r is stationary. Then Eq. (4.11-13) can be rewritten as

$$R(\Delta x, \tau) = \sigma^2 J_0(\beta V_1 \tau) J_0(\beta V_2 \tau + \beta\, \Delta x) \tag{4.11-14}$$

where $\beta = 2\pi/\lambda$
$J_0(\cdot)$ = zero-order Bessel function
$\tau = t_1 - t_2$
$\Delta x = x_1 - x_2$

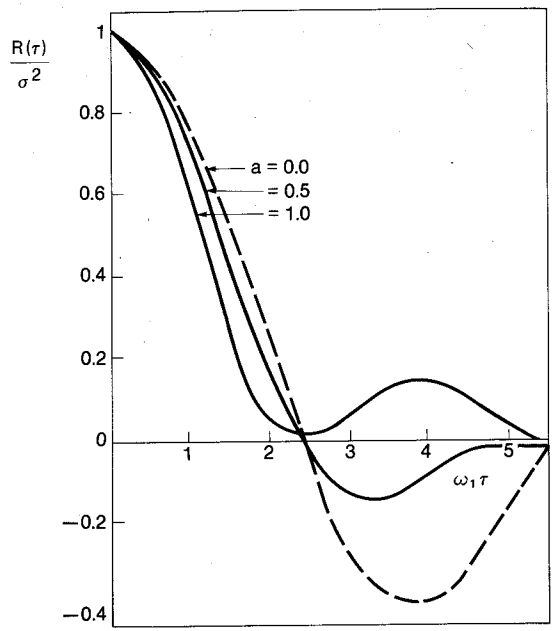


Figure 4.29 Normalized time-correlation function of the complex envelope for different values of $a = V_2/V_1$ versus $\omega_1 \Delta t$ where $\omega_1 = \beta V_1$. (*After Akki, Ref. 44.*)

The normalized time-correlation function is

$$\frac{R(\tau)}{\sigma^2} = J_0(\beta V_1 \tau) J_0(\beta V_2 \tau) \tag{4.11-15}$$

Equation (4.11-15) is plotted in Fig. 4.29. The spatial correlation function $R(\Delta x)$ is

$$\frac{R(\Delta x)}{\sigma^2} = J_0(\beta\ \Delta x) \tag{4.11-16}$$

Equation (4.11-16) is the same as for the base-to-mobile channel.

4.11.3 Power spectrum of the complex envelope

The power spectrum $S(f)$ is a Fourier transform of $R(\Delta t)$ from Eq. (4.11-15).

$$S(f) = \int_{-\infty}^{\infty} R(\Delta t) e^{-j2\pi f \Delta t}\, d(\Delta t) \tag{4.11-17}$$

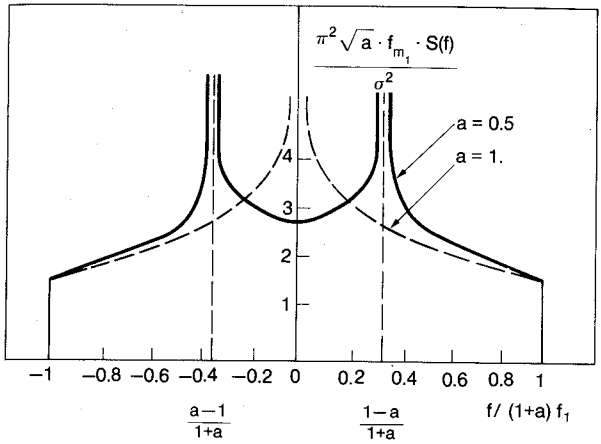


Figure 4.30 Power spectrum of the complex envelope for the case of $a = 0.5$ and $a = 1$ (where $a = V_2/V_1 = f_2/f_1$). (*After Akki, Ref. 44.*)

Substituting Eq. (4.11-15) into (4.11-17) yields

$$S(f) = \frac{\sigma^2}{\pi^2 f_1\sqrt{a}} K\left\{\frac{1+a}{2\sqrt{a}}\sqrt{1-\left[\frac{f}{(1+a)f_1}\right]^2}\right\} \tag{4.11-18}$$

where $a = f_2/f_1$
$\omega_1 = 2\pi V_1/\lambda$
$K(\cdot)$ = complete elliptic integral of the first kind.

Equation (4.11-18) is plotted in Fig. 4.30.

If $V_2 = 0$ and $a = 0$, Eq. (4.11-18) can be reduced to

$$S(f) = \frac{\sigma^2}{\pi\sqrt{f_1^2 - f^2}} \tag{4.11-19}$$

which is the equation for a base-to-mobile channel.

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Chapter

5

Cell-Site Antennas and Mobile Antennas

5.1 Equivalent Circuits of Antennas

The operating conditions of an actual antenna (Fig. 5.1*a*) can be expressed in an equivalent circuit for both receiving (Fig. 5.1*b*) and transmitting (Fig. 5.1*c*). In Fig. 5.1, Z_a is the antenna impedance, Z_L is the load impedance, and Z_T is the impedance at the transmitter terminal.

5.1.1 From the transmitting end (obtaining free-space path-loss formula)

Power P_t originates at a transmitting antenna and radiates out into space. (Equivalent circuit of a transmitting antenna is shown in Fig. 5.1*b*.) Assume that an isotropic source P_t is used and that the power in the spherical space will be measured as the power per unit area. This power density, called the *Poynting vector* ρ or the outward flow of electromagnetic energy through a given surface area, is expressed as

$$\rho = \frac{P_t}{4\pi r^2} \qquad \text{W/m}^2 \tag{5.1-1}$$

A receiving antenna at a distance r from the transmitting antenna

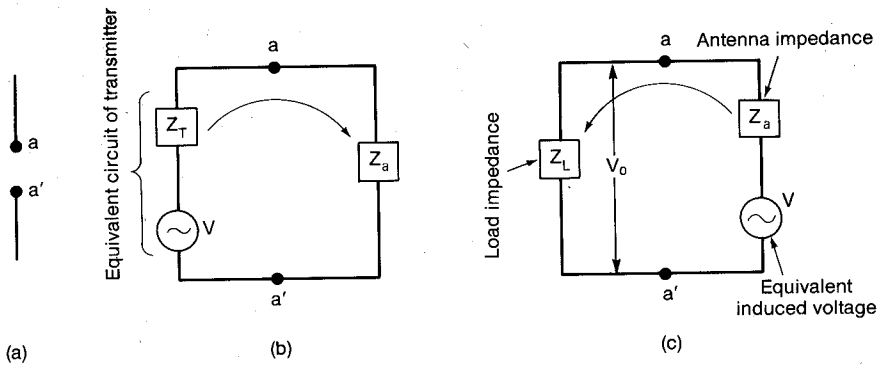

Figure 5.1 An actual antenna and its equivalent circuit. (*a*) An actual antenna; (*b*) equivalent circuit of a transmitting antenna; (*c*) equivalent circuit of a receiving antenna.

with an aperture A will receive power

$$P_r = \rho A = \frac{P_t A}{4\pi r^2} \quad \text{W} \tag{5.1-2}$$

Figure 5.2 is a schematic representation of received power in space.

From Eq. (5.1-2) we can derive the free-space path-loss formula because we know the relationship between the aperture A and the gain G.

$$G = \frac{4\pi A}{\lambda^2} \tag{5.1-3}$$

For a short dipole, $G = 1$. Then

$$A = \frac{\lambda^2}{4\pi} \tag{5.1-4}$$

Substitution of Eq. (5.1-4) into Eq. (5.1-2) yields the free-space formula

$$P_r = P_t \frac{1}{(4\pi r/\lambda)^2} \tag{5.1-5}$$

5.1.2 At the receiving end dBμV ↔ dBm (decibels above 1 μV ↔ decibels above 1 mW)

We can obtain the received power from Fig. 5.1*c*

$$P_r = \frac{V^2 Z_L}{(Z_L + Z_a)^2} \tag{5.1-6}$$

where V is the induced voltage in volts. For a maximum power delivery $Z_L = Z_a^*$, where the notation * indicates complex conjugate. Then we obtain $Z_L + Z_a = 2R_L$, where R_L is the real-load resistance. Equation (5.1-6) becomes

$$P_r = \frac{V^2}{4R_L} \tag{5.1-7}$$

Assume that a dipole or a monopole is used as a receiving antenna. The induced voltage V can be related to field strength E as[1]

$$V = \frac{E\lambda}{\pi} \tag{5.1-8}$$

where E is expressed in volts or microvolts per meter. Substitution of Eq. (5.1-7) into Eq. (5.1-8) yields

$$P_r = \frac{E^2\lambda^2}{4\pi^2 R_L} \tag{5.1-9}$$

If we set $R_L = 50\ \Omega$, P_r in decibels above 1 mW, and E in decibels (microvolts per meter), Eq. (5.1-9) becomes

$$P_r\,(\text{dBm}) = E\,(\text{dB}\mu\text{V}) - 113\ \text{dBm} + 10 \log \left(\frac{\lambda}{\pi}\right)^2 \tag{5.1-10}$$

The notation "dBμV" in Eq. (5.1-10) is a simplification of decibels above 1 μV/m, and has been accepted by the Institute of Radio Engineers. We can find the equivalent aperture A of Eq. (5.1-2) because the Poynt-

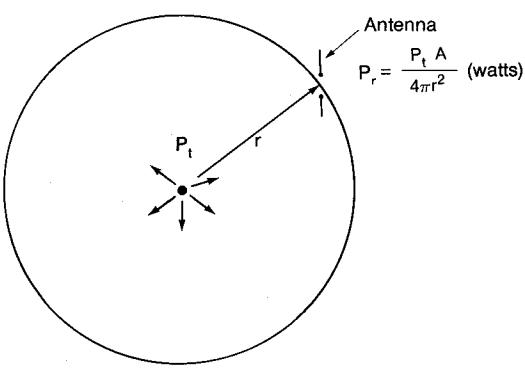

Figure 5.2 Received power in space.

ing vector ρ can be expressed as

$$\rho = \frac{E^2}{Z_0} \tag{5.1-11}$$

where Z_0 is the instrinsic impedance of the space ($=120\ \pi$). By substituting Eq. (5.1-11) into Eq. (5.1-2) and comparing it with Eq. (5.1-9) we obtain the equivalent aperture A.

$$A = \frac{\lambda^2 Z_0}{4\pi^2 R_L} \tag{5.1-12}$$

Equation (5.1-12) is the same as that given in Eq. 3.11 in Ref. 2.

5.1.3 Practical use of field-strength–received-power conversions

There is always confusion between the following two conversions.

Measuring field strength and converting it to received power. From Eq. (5.1-10), converting field strength in decibels above 1 μV/m to power received in decibels above 1 mW at 850 MHz by a dipole with a 50-Ω load is −132 dB.

$$P_r \quad \text{dBm} = E \quad \text{dB}\ (\mu\text{V/m}) - 132\ \text{dB} \qquad \text{at 850 MHz} \tag{5.1-13}$$

$$39\ \text{dB}\ (\mu\text{V/m}) = -93\ \text{dBm} \tag{5.1-14}$$

The notation "39-dBμV contour" is commonly used to mean 39 dB (μV/m) in cellular system design. Equation (5.1-13) is valid only at a given frequency (850 MHz), for a given antenna (monopole or dipole antenna), and for a given antenna load (50 Ω). Otherwise the field strength and the power have to be adjusted accordingly.

Measuring the voltage V_0 at the load terminal (Fig. 5.1c) and converting to received power. Given $P_r = (V_0^2/R_L)$, where $R_L = 50\ \Omega$, we can obtain a relationship

$$0\ \text{dB}\mu\text{V} \langle = \rangle -107\ \text{dBm} \tag{5.1-15}$$

For example, if a voltage meter at V_0 is 7 dBμV, then the received power is −100 dBm. Equation (5.1-15) expresses a voltage-to-power ratio which varies with the load impedance but is independent of the frequency and the type of antenna.

5.1.4 Effective radiated power

There is a standard way of specifying the power radiated within a given geographic area. The radiated power that can be transmitted should be converted to the amount of power radiated from an omnidirectional antenna (particularly a dipole antenna). If a high-gain antenna is used, the transmitted power should be reduced. Therefore, if the specification states that the effective radiated power (ERP) or the transmitted power multiplied by the antenna gain is 100 W (50 dBm) and if a 10-dB high-gain antenna is used, the actual transmitted power should be 10 W (40 dBm).

There is also a term called *effective isotropic radiated power* (EIRP). EIRP is referenced to an isotropic point source. The difference between ERP and EIRP is 2 dB.

$$\text{ERP} = \text{EIRP} + 2 \text{ dB}$$

5.2 The Gain-and-Pattern Relationship

5.2.1 The gain of an antenna or an antenna array

$$G = \frac{4\pi(\text{maximum radiation intensity})}{\text{total power radiated}}$$

$$= \frac{E^2_{\max}(\theta_m, \phi_m)}{\overline{E^2}(\theta, \phi)} \tag{5.2-1}$$

where E = electric field
$E_{\max}$ = maximum of E
$\overline{E^2}$ = average value of E^2 which is related to the radiation intensity
θ, ϕ = angles shown in Fig. 5.3

We can obtain the antenna pattern E from either a measurement or an analytic form. Then the gain G from the pattern E can be calculated.

5.2.2 The pattern and gain of an antenna array

Frequently, it is necessary to calculate the gain of an antenna array. The general field pattern can be expressed as[3]

$$E(\theta, \phi) = \frac{\sin[N\pi(d\cos\phi \cdot \sin\theta + \psi)]}{N\sin[\pi(d\cos\phi \cdot \sin\theta + \psi)]} \times \begin{pmatrix}\text{individual antenna}\\ \text{element pattern}\end{pmatrix} \tag{5.2-2}$$

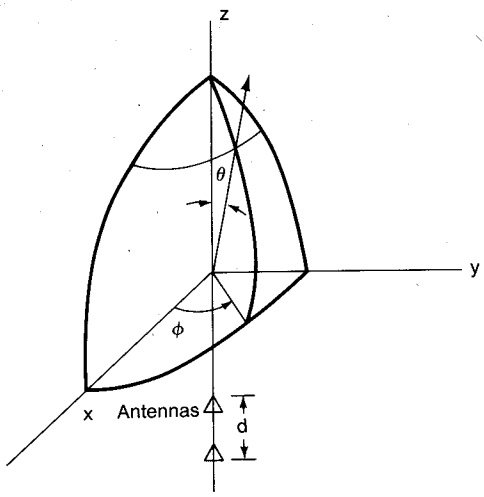

Figure 5.3 A coordinate of antenna arrays.

where N = number of elements in an array
d = spacing between adjacent elements in wavelength
ψ = phase difference between two adjacent elements
ϕ, θ = radiation angles are shown Fig. 5.3
$\theta = 90°$ = direction perpendicular to the array axis

The gain of the array can be obtained by substituting Eq. (5.2-2) into Eq. (5.2-1). Two cases are listed below.

1. *A broadside-array case* ($\psi = 0$). The individual elements are in parallel and lie in the y-z plane. In this case the gain related to a single element of any type of antenna is shown below.

	Gain G, dB	
N	$d = 0.5\lambda$	$d = 0.9\lambda$
2	4	4.6
3	5.5	6
4	7	8
6	9	10.4
8	10	12
12	12	14
16	13.2	15.4

2. *A collinear-array case* ($\psi = 0$). The individual elements are collinear in the z axis. In this case, the gain related to a single dipole element is as shown in the following table.

N	Gain G, dB	
	$d = 0.5\lambda$	$d = 0.9\lambda$
2	2.2	3.7
3	3.8	5.7
4	4.9	8.1
6	6.5	9.0
8	8.8	9.4
12	9.4	12.2
16	10.7	13.4

5.2.3 The relationship between gain and beamwidth

A general formula. Assume that the gain G and the directivity D are nearly the same. Then[4]

$$G \approx D = \frac{32{,}400}{\phi^\circ\theta^\circ} \qquad \text{for small } \phi \text{ and } \theta \tag{5.2-3}$$

where ϕ° and θ° are the 3-dB beamwidths in two planes. Elliott[4] points out that when ϕ and θ are small (<40°) the figure in Eq. (5.2-3) reported by Kraus[5] (41,253) is based on the incorrect assumption that the cross section formed by θ and ϕ is rectangular. It is, in fact, elliptical, so its area is $\pi/4$ that of Kraus' figure, or 32,400. However, for large (>40°) ϕ° and θ°, the Kraus equation

$$G \approx D = \frac{41{,}253}{\phi^\circ\theta^\circ} \qquad \text{for large } \phi^\circ \text{ and } \theta^\circ \tag{5.2-4}$$

has been shown to be valid. For a high-gain or high-directivity antenna, the gain G resulting from decreased efficiency through the system has only 50 to 70 percent of the directivity D.

For a linear element or collinear array. If a linear element is used, the approximate gain also can be obtained from a vertical 3-dB beamwidth. Assume that the array or the linear element has a doughnut pattern. The relationship between the 3-dB beamwidth θ_0 and the gain G for a linear element or a collinear array can be derived in decibels, above an isotropic source which is a unity gain from Eq. (5.2-3) provided the angle θ_0 is small (<40°).[6]

$$G \approx D = \frac{101.5^\circ}{\theta_0} \qquad \text{or} \qquad D = 10 \log\left(\frac{101.5}{\theta_0}\right) \qquad \text{for small } \theta_0 \tag{5.2-5}$$

Assume that the gain G and the directivity D are the same. The gain G always refers to an isotropic source which is 2 dB lower than the gain of a dipole. Equation (5.2-5), which is valid for small θ_0, may have an error of $\pm 1°$ as compared with the measured pattern. For large θ_0, the relationship between G and θ_0 can be found from Eq. (5.2-4) as

$$G \approx D = \frac{114.6°}{\theta_0} \tag{5.2-6}$$

Example 5.1 For gain of 9 dB with respect to a dipole antenna or 11 dB with respect to a short dipole, 3-dB beamwidth is

$$\theta_0 = \frac{101.5°}{10^{1.1}} = \frac{101.5°}{12.59} = 8°$$

For a gain of 6 dB with respect to a dipole antenna, the 3-dB beamwidth is

$$\theta_0 = \frac{101.5°}{6.32} = 16°$$

In reality, because of the inefficiency of power delivery, the gain would decrease by 1 dB as a result of a mismatch measured from a (voltage standing-wave ratio) (VSWR)

$$G = 10 \log \frac{101.5}{\theta_0} - 1 \quad \text{dB} \tag{5.2-7}$$

Equation (5.2-7) would be used for an antenna operating within its marginal frequency range.

5.3 Sum-and-Difference Patterns—Engineering Antenna Pattern

After obtaining a predicted field-strength contour (see Sec. 4.8), we can engineer an antenna pattern to conform to uniform coverage. For different antennas pointing in different directions and with different spacings, we can use any of a number of methods. If we know the antenna pattern and the geographic configuration of the antennas, a computer program can help us to find the coverage. Several synthesis methods can be used to generate a desired antenna configuration.

5.3.1 General formula

Many applications of linear arrays are based on sum-and-difference patterns. The mainbeam of the pattern is always known as the sum pattern pointing at an angle θ_0. The difference pattern produces twin mainbeams straddling θ_0. When $2N$ elements are in an array, equispaced by a separation d, the general pattern for both sum and differ-

ence is

$$A(\theta) = \sum_{n=1}^{N} I_n \exp j \frac{2n-1}{2} \beta d(\cos\theta - \cos\theta_0) + I_{-n} \exp\left[-j\frac{2n-1}{2} d(\cos\theta - \cos\theta_0)\right] \quad (5.3\text{-}1)$$

where β = wavenumber = $2\pi/\lambda$
I_n = normalized current distributions
N = total number of elements

For a sum pattern, all the current amplitudes are the same.

$$I_n = I_{-n} \quad (5.3\text{-}2)$$

For a difference pattern, the current amplitudes of one side (half of the total elements) are positive and the current amplitudes of the other side (half of the total elements) are negative.

$$I_n = -I_{-n} \quad (5.3\text{-}3)$$

Most pattern synthesis problems can be solved by determining the current distribution I_n. A few solutions follow.

5.3.2 Synthesis of sum patterns

Dolph-Chebyshev synthesis of sum patterns. This method can be used to reduce the level of sidelobes; however, one disadvantage of further reduction of sidelobe level is broadening of the mainbeam. The techniques are discussed in Ref. 7.

Taylor synthesis. A continuous line-source distribution or a distribution for discrete arrays can give a desired pattern which contains a single mainbeam of a prescribed beamwidth and pointing direction with a family of sidelobes at a common specified level. The Taylor synthesis is derived from the following equation, where an antenna pattern $F(\theta)$ is determined from an aperture current distribution $g(l)$

$$F(\theta) = \int_{-a}^{a} g(l) e^{j\beta l \cos\theta}\, dl \quad (5.3\text{-}4)$$

Symmetrical pattern.[8] For production of a symmetrical pattern at the mainbeam, the current-amplitude distribution $|g(l)|$ is the only factor to consider. The phase of the current distribution can remain constant.

A typical pattern (Fig. 5.4*a*) would be generated from a current-amplitude distribution (Fig. 5.4*b*).

Asymmetrical pattern. For production of an asymmetrical pattern, both current amplitude $|g(l)|$ and phase arg $g(l)$ should be considered.

5.3.3 Synthesis of difference patterns (Bayliss synthesis)[9]

To find a continuous line source that will produce a symmetrical difference pattern, with twin mainbeam patterns and specified sidelobes,

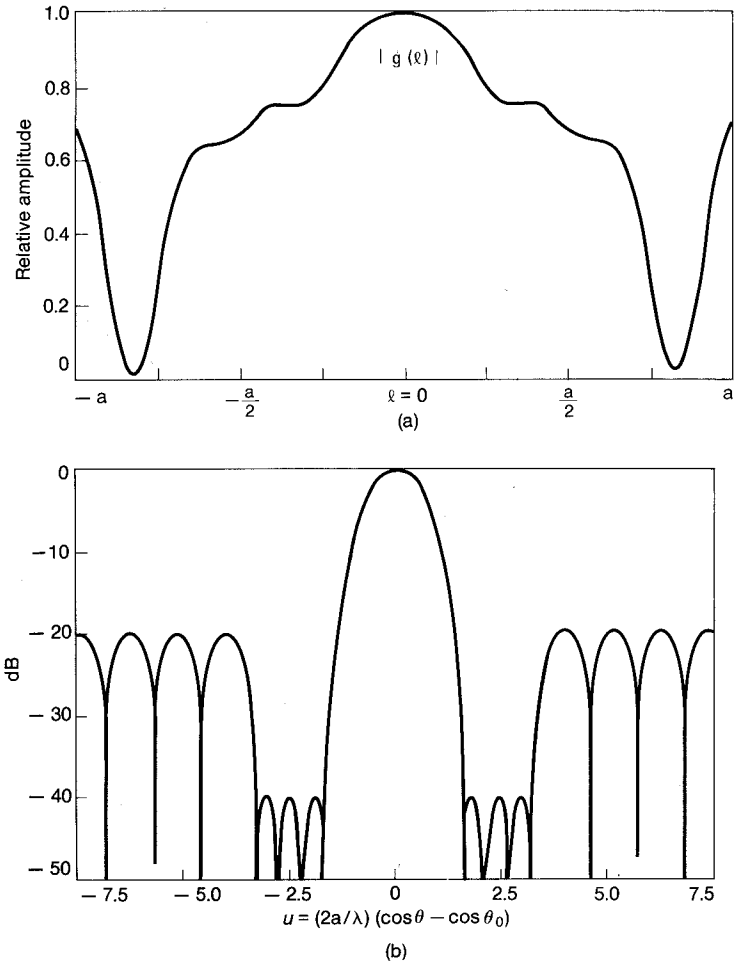

Figure 5.4 A symmetrical sum pattern (*reprinted from Elliot, Ref. 8*). (*a*) The aperature distribution for the two-antenna arrangement shown in Fig. 5.11*b* (© *1976 IEEE; reprinted from IEEE AP Transactions, 1976, pp. 76–83*). (*b*) The evolution of a symmetrical sum pattern with reduced inner sidelobes (© *1976 IEEE; reprinted from IEEE AP Transactions, 1976, pp. 76–83*).

we can set

$$D(\theta) = \int_{-a}^{a} g(l)e^{j\beta l \cos\theta}\, dl \tag{5.3-5}$$

For a desired difference pattern such as that shown in Fig. 5.5*a*, the

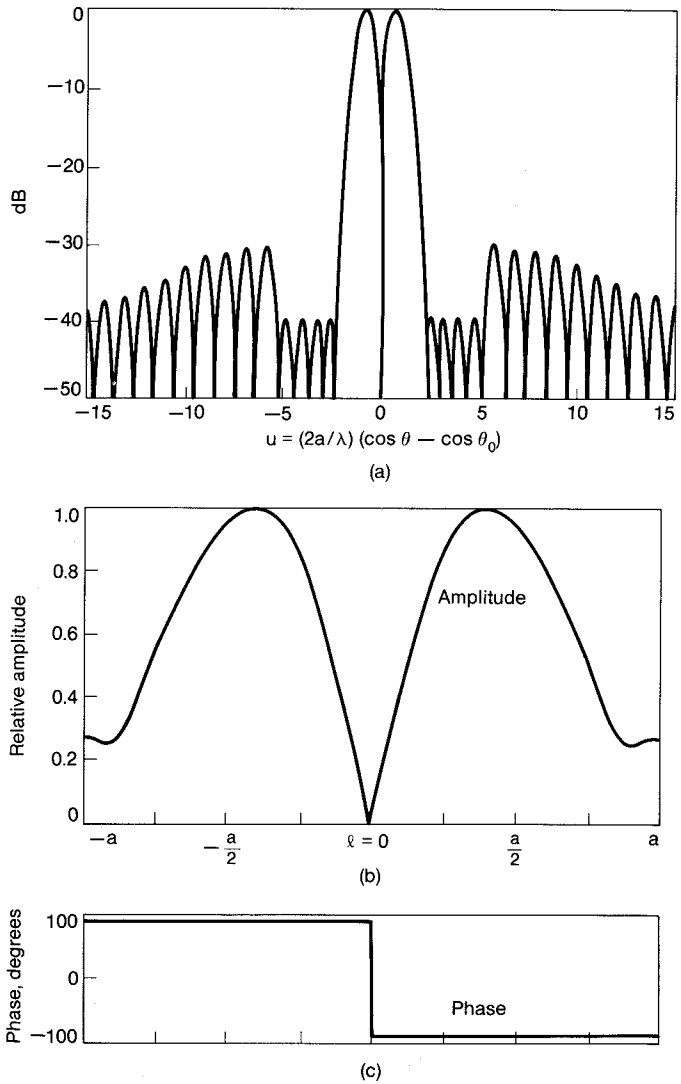

Figure 5.5 A symmetrical difference pattern (*reprinted from Elliot, Ref. 9*). (*a*) A modified Bayliss difference pattern; inner sidelobes symmetrically depressed. (© *IEEE; reprinted from IEEE AP Transactions, 1976, pp. 310–316*). (*b, c*) Aperature distribution for the pattern. (© *IEEE; reprinted from IEEE AP Transactions, 1976, pp. 310–316*).

current-amplitude distributions $|g(l)|$ should be designed as shown in Fig. 5.5*b* and the phase arg $g(l)$ as shown in Fig. 5.5*c*.

5.3.4 Null-free patterns

In mobile communications applications, field-strength patterns without nulls are preferred for the antennas in a vertical plane. The typical vertical pattern of most antennas is shown in Fig. 5.6*a*. The field pattern can be represented as

$$F(u) = \sum_{n=0}^{N} K_n \frac{\sin \pi u}{\pi u} \tag{5.3-6}$$

where $u = (2a/\lambda)(\cos\theta - \cos\theta_n)$. The concept is to add all $(\sin \pi u)/(\pi u)$ patterns at different pointing angles as shown in Fig. 5.6*a*. K_n is the maximum signal level. The resulting pattern does not contain nulls. The null-free pattern can be applied in the field as shown in Fig. 5.6*b*.

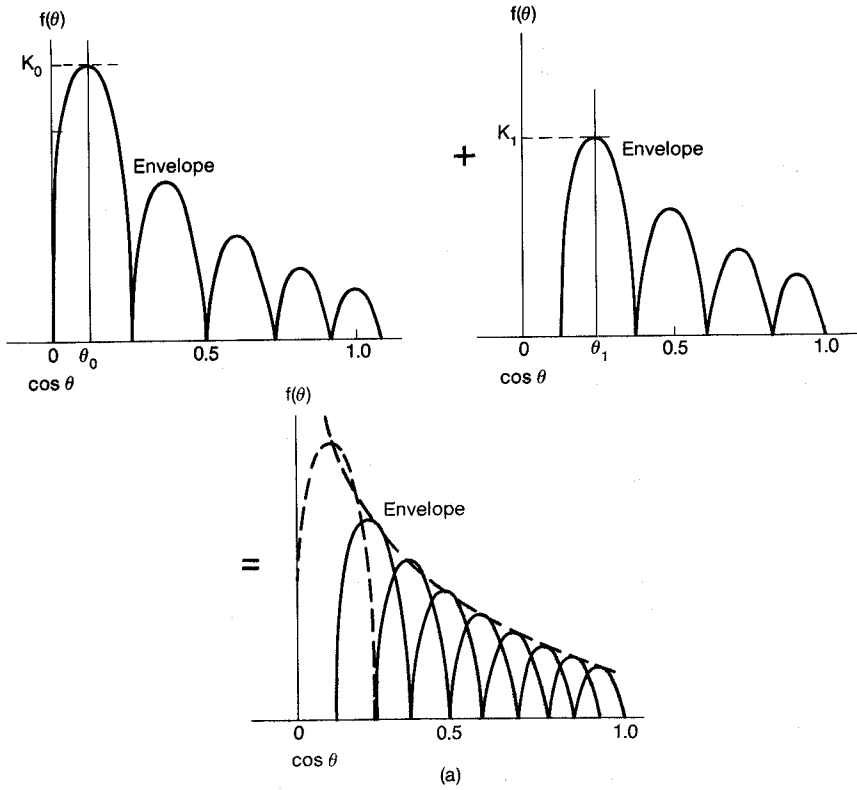

Figure 5.6 Null-free patterns. (*a*) Formation of a null-free pattern; (*b*) application of a null-free pattern (*reprinted from Elliott, Ref. 4, p. 192*).

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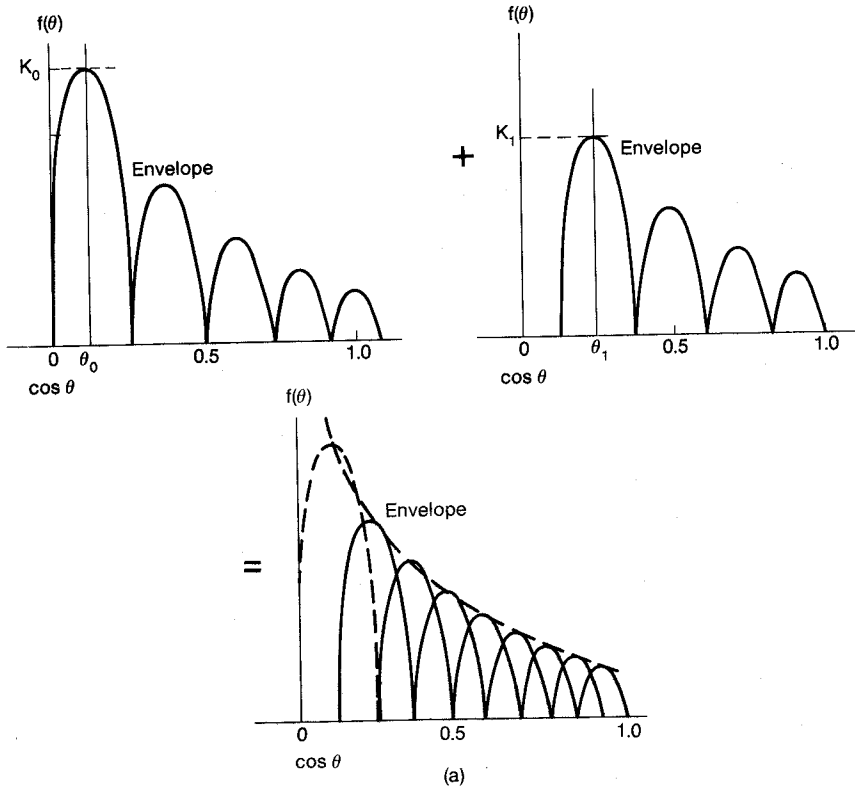


Figure 5.6 Null-free patterns. (*a*) Formation of a null-free pattern; (*b*) application of a null-free pattern (*reprinted from Elliott, Ref. 4, p. 192*).

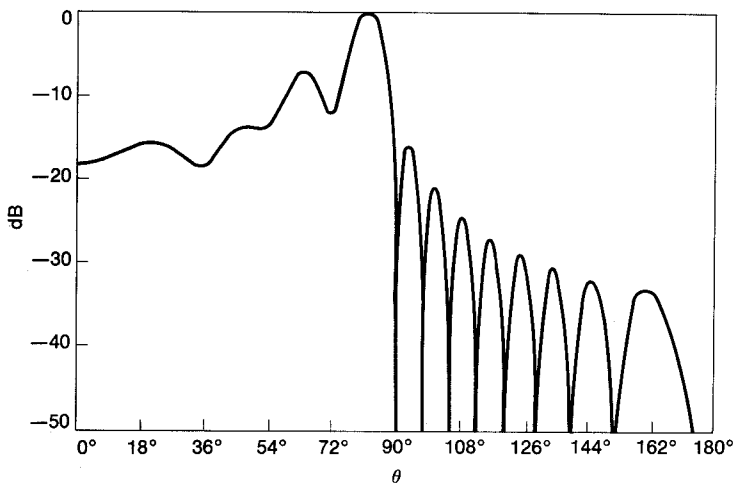
5.3.5 Practical applications

In designing a collinear array for the high-gain omnidirectional antenna, it is possible to control the current distribution by a microprocessor. This synthesis technique awaits further investigation by antenna research-and-development (R&D) specialists.

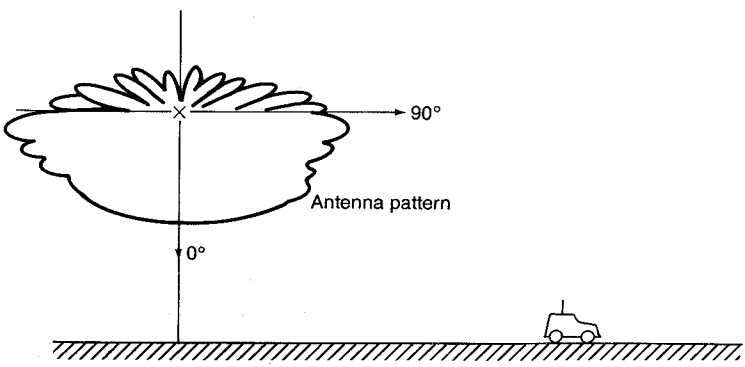
5.4 Antennas at Cell Site

5.4.1 For coverage use—omnidirectional antennas

High-gain antennas. There are standard 6-dB and 9-dB gain omnidirectional antennas. The antenna patterns for 6-dB gain and 9-dB gain are shown in Fig. 5.7.

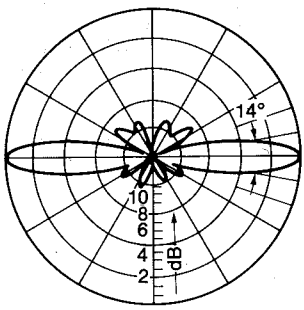


Null-Filled Cosecant Squared Pattern

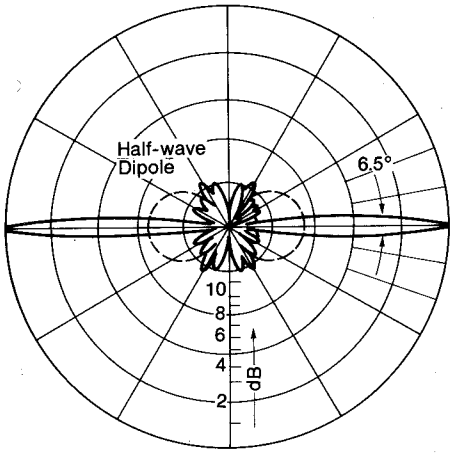


(b)

Figure 5.6 (*Continued*)



Vertical
(a)



Vertical
(b)

Figure 5.7 High-gain omnidirectional antennas (*reprinted from Kathrein Mobile Communications Catalog*). Gain with reference to dipole: (*a*) 6 dB; (*b*) 9 dB.

Start-up system configuration. In a start-up system, an omnicell, in which all the transmitting antennas are omnidirectional, is used. Each transmitting antenna can transmit signals from 16 radio transmitters simultaneously using a 16-channel combiner. Each cell normally can have three transmitting antennas which serve 45 voice radio transmitters* simultaneously. Each sending signal is amplified by its own channel amplifier in each radio transmitter, then 16 channels (radio signals) pass through a 16-channel combiner and transmit signals by means of a transmitting antenna (see Fig. 5.8*a*).

* The combiner is designed for combining 16 voice channels. However, the cellular system divides its 312 voice channels into 21 sets; each set consists of only about 15 voice channels. Therefore the dummy loads have to be put on some empty ports of a 16-channel combiner.

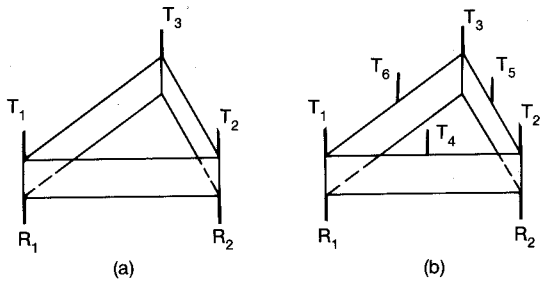


Figure 5.8 Cell-site antennas for omnicells: (*a*) for 45 channels; (*b*) for 90 channels

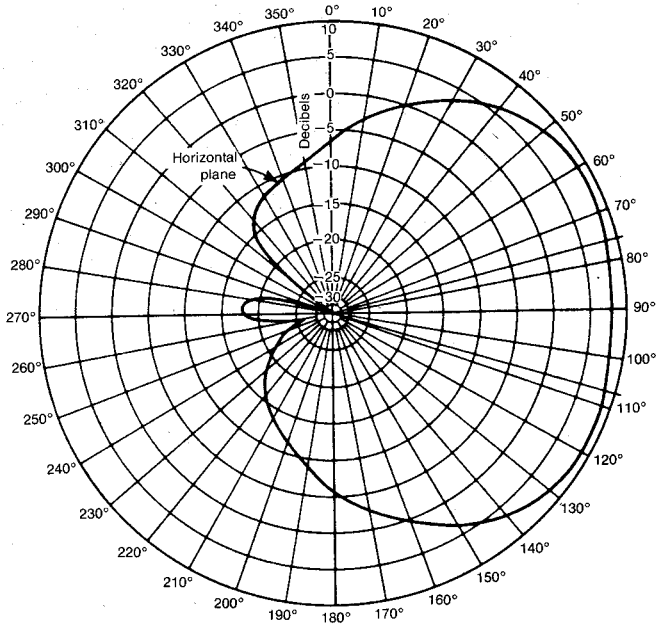
Two receiving antennas commonly can receive all 45 voice radio signals simultaneously. Then in each channel, two identical signals received by two receiving antennas pass through a diversity receiver of that channel. The receiving antenna configuration on the antenna mast is shown in Fig. 5.8. The separation of antennas for a diversity receiver is discussed in Sec. 5.5.

Abnormal antenna configuration. Usually, the call traffic in each cell increases as the number of customers increases. Some cells require a greater number of radios to handle the increasing traffic. An omnicell site can be equipped with up to 90 voice radios. In such cases six transmitting antennas should be used as shown in Fig. 5.8*b*. In the meantime, the number of receiving antennas is still two. In order to reduce the number of transmitting antennas, a hybrid ring combiner which can combine two 16-cannel signals is found.[10] This means that only three transmitting antennas are needed to transmit 90 radio signals. However, the ring combiner has a limitation of handling power up to 600 W with a loss of 3 dB.

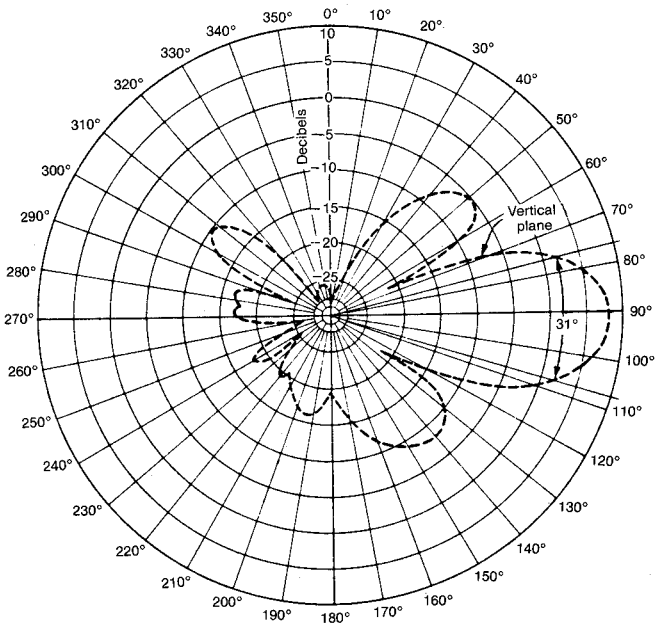
5.4.2 For interference reduction use—directional antennas

When the frequency reuse scheme must be used, cochannel interference will occur. The cochannel interference reduction factor $q = D/R = 4.6$ is based on the assumption that the terrain is flat. Because actual terrain is seldom flat, we must either increase q or use directional antennas.

Directional antennas. A 120°-corner reflector or 120°-plane reflector can be used in a 120°-sector cell. A 60°-corner reflector can be used in a 60°-sector cell. A typical pattern for a directional antenna of 120° beamwidth is shown in Fig. 5.9.



(a)



(b)

Figure 5.9 A typical 8-dB directional antenna pattern. (*Reprinted from Bell Systems Technical Journal, Vol. 58, January 1979, pp. 224–225.*) (*a*) Azimuthal pattern of 8-dB directional antenna. (*b*) Vertical pattern of 8-dB directional antenna.

Normal antenna (mature system) configuration

1. $K = 7$ cell pattern (120° sectors). In a $K = 7$ cell pattern for frequency reuse, if 333 channels are used, each cell would have about 45 radios. Each 120° sector would have one transmitting antenna and two receiving antennas and would serve 16 radios. The two receiving antennas are used for diversity (see Fig. 5.10*a*).
2. $K = 4$ cell pattern (60° sectors). We do not use $K = 4$ in an omnicell system because the cochannel reuse distance is not adequate. Therefore, in a $K = 4$ cell pattern, 60° sectors are used.[11] There are 24 sectors. In this $K = 4$ cell-pattern system, two approaches are used.
   a. Transmitting-receiving 60° sectors. Each sector has a transmitting antenna carrying its own set of frequency radios and hands off frequencies to other neighboring sectors or other cells. This is a full $K = 4$ cell-pattern system. If 333 channels are used, with 13 radios per sector, there will be one transmitting antenna and one receiving antenna in each sector. At the receiving end, two of six receiving antennas are selected for an angle diversity for each radio channel (see Fig. 5.10*b*).
   b. Receiving 60° sectors. Only 60°-sector receiving antennas are used to locate mobile units and hand off to a proper neighboring cell with a high degree of accuracy. All the transmitting antennas are omnidirectional within each cell. At the receiving end, the angle diversity for each radio channel is also used in this case.

Abnormal antenna configuration. If the call traffic is gradually increasing, there is an economic advantage in using the existing cell systems rather than the new splitting cell system (splitting into smaller cells). In the former, each site is capable of adding more radios. In a $K = 7$ cell pattern with 120° sectors, two transmitting antennas at each sector are used (Fig. 5.10*c*). Each antenna serves 16 radios if a 16-channel combiner is used. One observation from Fig. 5.10*c* should be mentioned

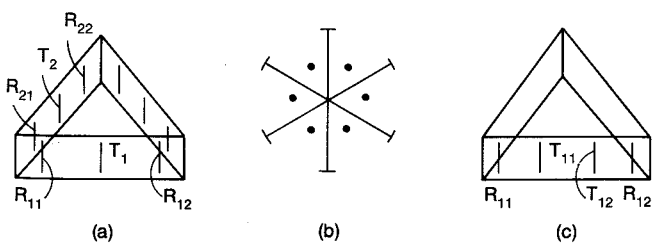


Figure 5.10 Directional antenna arrangement: (*a*) 120° sector (45 radios); (*b*) 60° sector; (*c*) 120° sector (90 radios).

here. The two transmitting antennas in each sector are placed relatively closer to the receiving antennas than in the single transmitting antenna case. This may cause some degree of desensitization in the receivers. The current technology can combine 32 channels[10] in a combiner; therefore, only one transmitting antenna is needed in each sector. However, this one transmitting antenna must be capable of withstanding a high degree of transmitted power. If each channel transmits 100 W, the total power that the antenna terminal could withstand is 3.2 kW.

The 32-channel combiner has a power limitation which would be specified by different manufacturers. Two receiving antennas in each 120° sector remain the same for space diversity use.

5.4.3 Location antennas

In each cell site a location receiver connects to the respective location antenna. This antenna can be either omnidirectional or shared-directional. The location receiver can tune a channel to one of 333 channels either upon demand or periodically. This operation is discussed in Chaps. 8 and 9.

5.4.4 Setup-channel antennas

The setup-channel antenna is used to page a called mobile unit or to access a call from a mobile unit. It transmits only data. The setup–channel antenna can be an omnidirectional antenna or consist of several directional antennas at one cell site. In general, in both omnicell and sector-cell systems, one omnidirectional antenna is used for transmitting signals and another for receiving signals in each cell site. Setup–channel operational procedures are discussed in Chap. 8.

5.4.5 Space-diversity antennas used at cell site

Two-branch space-diversity antennas are used at the cell site to receive the same signal with different fading envelopes, one at each antenna. The degree of correlation between two fading envelopes is determined by the degree of separation between two receiving antennas. When the two fading envelopes are combined, the degree of fading is reduced; this improvement is discussed in Ref. 12. Here the antenna setup is shown in Fig. 5.11*a*. Equation (5.4-1) is presented as an example for the designer to use.

$$\eta = \frac{h}{D} = 11 \tag{5.4-1}$$

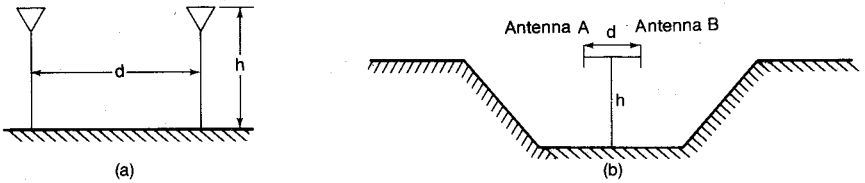


Figure 5.11 Diversity antenna spacing at the cell site: (*a*) $\eta = h/d$; (*b*) proper arrangement with two antennas.

where h is the antenna height and D is the antenna separation. From Eq. (5.4-1), the separation $d \geq 8\lambda$ is needed for an antenna height of 100 ft (30 m) and the separation $d \geq 14\lambda$ is needed for an antenna height of 150 ft (50 m). In any omnicell system, the two space-diversity antennas should be aligned with the terrain, which should have a U shape[13] as shown in Fig. 5.11*b*.

Space-diversity antennas can separate only horizontally, not vertically; thus, there is no advantage in using a vertical separation in the design.[13] The use of space-diversity antennas at the base station is discussed in detail in Ref. 13.

5.4.6 Umbrella-pattern antennas

In certain situations, umbrella-pattern antennas should be used for the cell-site antennas.

Normal umbrella-pattern antenna.[14] For controlling the energy in a confined area, the umbrella-pattern antenna can be developed by using a monopole with a top disk (top-loading) as shown in Fig. 5.12. The size of the disk determines the tilting angle of the pattern. The smaller the disk, the larger the tilting angle of the umbrella pattern.

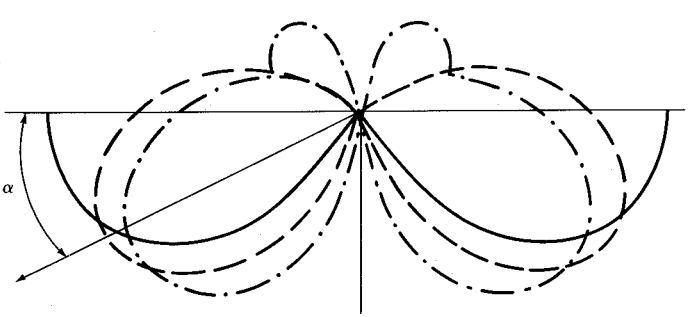


Figure 5.12 Vertical-plane patterns of quarter-wavelength stub antenna on infinite ground plane (solid) and on finite ground planes several wavelengths in diameter (dashed line) and about one wavelength in diameter (dotted line) (*after Kraus, Ref. 14*).

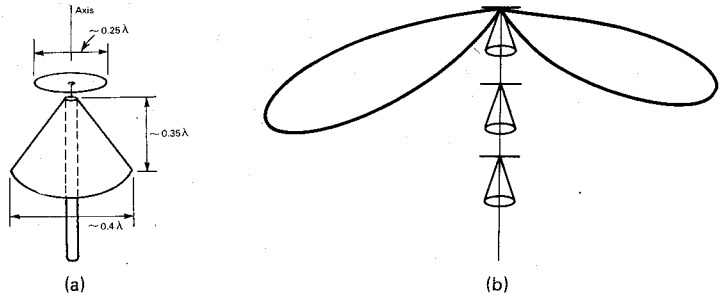


Figure 5.13 Discone antennas. (*a*) Single antenna. (*b*) An array of antennas.

Broadband umbrella-pattern antenna. The parameters of a *discone antenna* (a bioconical antenna in which one of the cones is extended to 180° to form a disk) are shown in Fig. 5.13*a*. The diameter of the disk, the length of the cone, and the opening of the cone can be adjusted to create an umbrella-pattern antenna as described in Ref. 15.

High-gain broadband umbrella-pattern antenna. A high-gain antenna can be constructed by vertically stacking a number of umbrella-pattern antennas as shown in Fig. 5.13*b*

$$E_0 = \frac{\sin\,[(Nd/2\lambda)\cos\,\phi]}{\sin\,[(d/2\lambda)\,\cos\,\phi]} \cdot \text{(individual umbrella pattern)}$$

where ϕ = direction of wave travel
N = number of elements
d = spacing between two adjacent elements

5.4.7 Interference reduction antenna[16]

A design for an antenna configuration that reduces interference in two critical directions (areas) is shown in Fig. 5.14. The parasitic (insu-

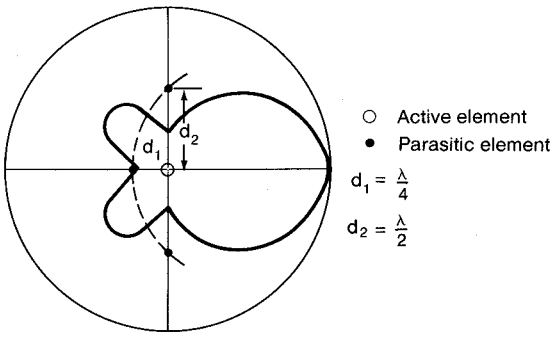


Figure 5.14 Application of parasitic elements (*after Jasik, Ref. 16*).

lation) element is about 1.05 times longer than the active element. The separation d and the various values of parasitic element reactance X_{22} were shown by Brown for this application.[17]

5.5 Unique Situations of Cell-Site Antennas

5.5.1 Antenna pattern in free space and in mobile environments

The antenna pattern we normally use is the one measured from an antenna range (open, nonurban area) or an antenna darkroom. However, when the antenna is placed in a suburban or urban environment and the mobile antenna is lower than the heights of the surroundings, the cell-site antenna pattern as a mobile unit received in a circle equidistant around the cell site is quite different from the free-space antenna pattern. Consider the following facts in the mobile radio environment.

1. The strongest reception still coincides with the strongest signal strength of the directional antenna.
2. The pattern is distorted in an urban or suburban environment.
3. For a 120° directional antenna, the backlobe (or front-to-back ratio) is about 10 dB less than the frontlobe, regardless of whether a weak sidelobe pattern or no sidelobe pattern is designed in a free-space condition. This condition exists because the strong signal radiates in front, bouncing back from the surroundings so that the energy can be received from the back of the antenna. The energy-reflection mechanism is illustrated in Fig. 5.15.
4. A design specification of the front-to-back ratio of a directional antenna (from the manufacturer's catalog) is different from the actual front-to-back ratio in the mobile radio environment. Therefore the environment and the antenna beamwidth determine how the antenna will be used in a mobile radio environment. For example, if a 60° directional antenna is used in a mobile radio environment, the actual front-to-back ratio can vary depending on the given environment. If the close-in man-made structures in front of the antenna are highly reflectable to the signal, then the front-to-back ratio of a low-master directional antenna can be as low as 6 dB in some circumstances. In this case, the directional antenna beamwidth pattern has no correlation between it measured in the free space and it measured in the mobile radio environment. If all the buildings are far away from the directional antenna, then the front-to-back ration measured in the field will be close to the specified antenna pattern, usually 20 dB.

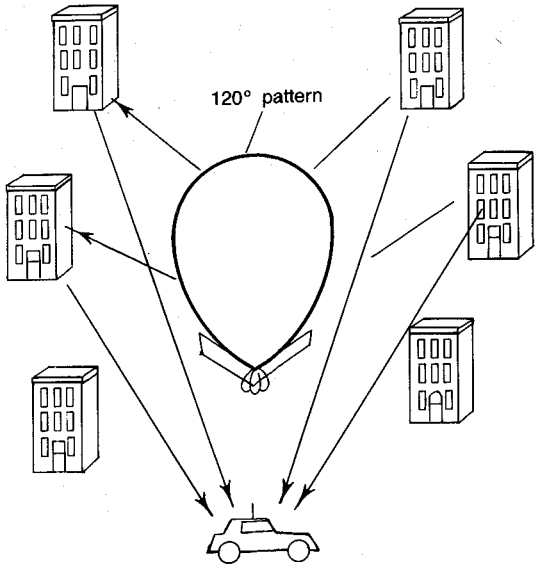


Figure 5.15 Front-to-back ratio of a directional antenna in a mobile radio environment.

5.5.2 Minimum separation of cell-site receiving antennas

Separation between two transmitting antennas should be minimized to avoid the intermodulation discussed in Chap. 7. The minimum separation between a transmitting antenna and a receiving antenna nec-

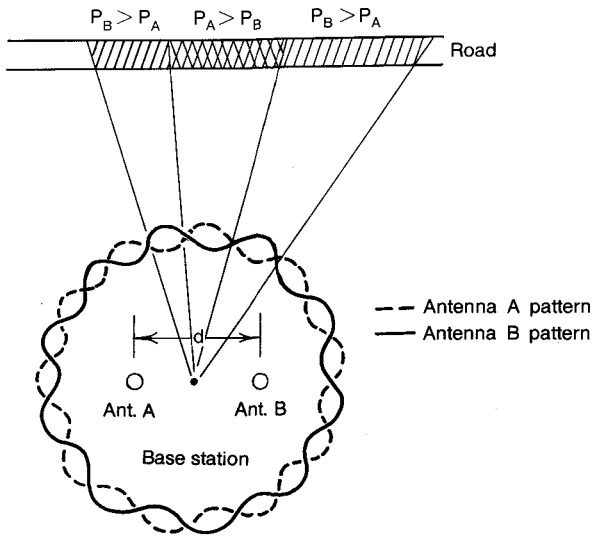


Figure 5.16 Antenna pattern ripple effect.

essary to avoid receiver desensitization is also described in Chap. 7. Here we are describing a minimum separation between two receiving antennas to reduce the antenna pattern ripple effects.

The two receiving antennas are used for a space-diversity receiver. Because of the near-field disturbance due to the close spacing, ripples will form in the antenna patterns (Fig. 5.16). The difference in power reception between two antennas at different angles of arrival is shown in Fig. 5.16. If the antennas are located closer, the difference in power between two antennas at a given pointing angle increases. Although the power difference is confined to a small sector, it affects a large section of the street as shown in Fig. 5.16. If the power difference is excessive, use of a space diversity will have no effect reducing fading. At 850 MHz, the separation of eight wavelengths between two receiving antennas creates a power difference of ±2 dB, which is tolerable for the advantageous use of a diversity scheme.[18]

5.5.3 Regular check of the cell-site antennas

Air-pressurized cable is often used in cell-site antennas to prevent moisture from entering the cable and causing excessive attenuation. One method of checking the cell-site antennas is to measure the power delivered to the antenna terminal; however, few systems have this capability. The other method is to measure the VSWR at the bottom of the tower. In this case the loss of reflected power due to the cable under normal conditions should be considered. For a high tower, the VSWR reading may not be accurate.

If each cable connector has 1-dB loss due to energy leakage and two midsection 1-dB loss connectors are used in the transmitted system as shown in Fig. 5.17, the reflected power P_b indicated in the VSWR would be 4 dB less than the real reflected power.

5.5.4 Choosing an antenna site

In antenna site selection we have relied on the point-to-point prediction method (discussed in Chap. 4), which is applicable primarily for coverage patterns under conditions of light call traffic in the system. Reduction of interference is an important factor in antenna site selection.

When a site is chosen on the map, there is a 50 percent chance that the site location cannot be acquired. A written rule states that[24] an antenna location can be found within a quarter of the size of cell $R/4$. If the site is an 8-mi cell, the antenna can be located within a 2-mi radius. This hypothesis is based on the simulation result that the change in site within a 2-mi radius would not affect the coverage pattern at a distance 8 mi away. If the site is a 2-mi cell, the antenna can be located within a 0.5-mi radius.

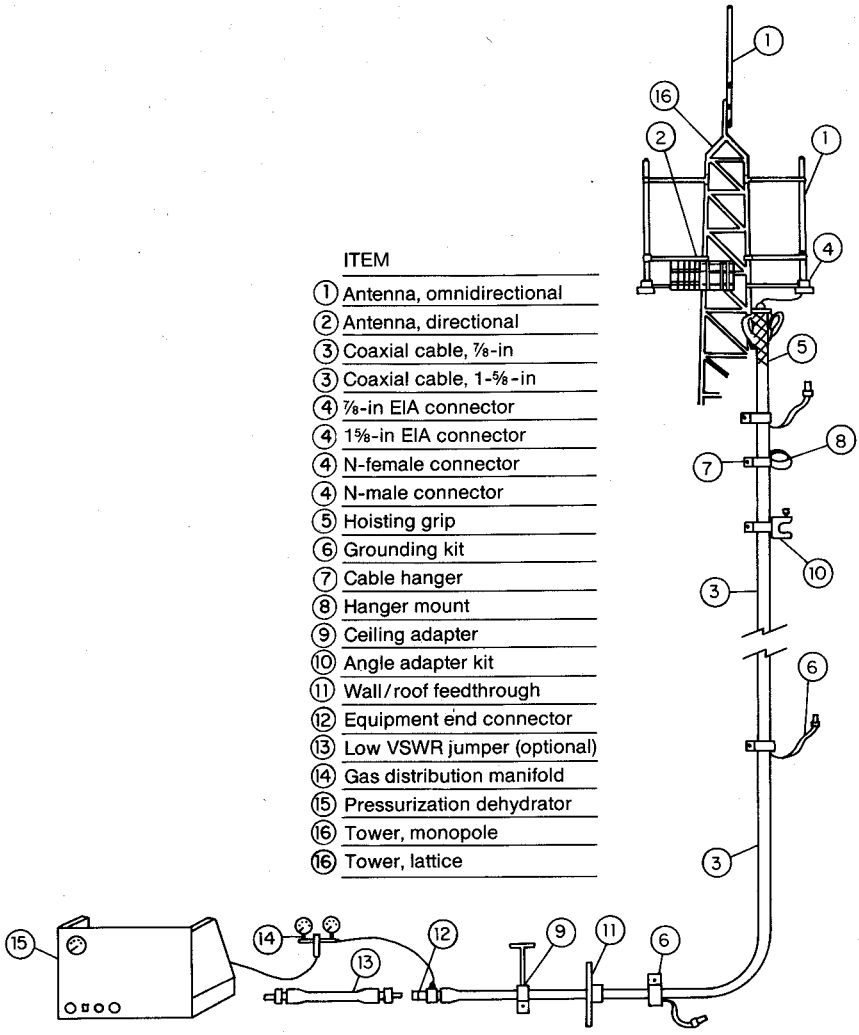


Figure 5.17 Antenna system at cell site.

The quarter-radius rule can be applied only on relatively flat terrain, not in a hilly area. To determine whether this rule can be applied in a general area, one can use the point-to-point prediction method to plot the coverage at different site locations and compare the differences. Usually when the point-to-point prediction method (tool) can be used to design a system, the quarter-radius rule becomes useless.

5.6 Mobile Antennas

The requirement of a mobile (motor-vehicle–mounted) antenna is an omnidirectional antenna which can be located as high as possible from

the point of reception. However, the physical limitation of antenna height on the vehicle restricts this requirement. Generally the antenna should at least clear the top of the vehicle. Patterns for two types of mobile antenna are shown in Fig. 5.18.

5.6.1 Roof-mounted antenna

The antenna pattern of a roof-mounted antenna is more or less uniformly distributed around the mobile unit when measured at an antenna range in free space as shown in Fig. 5.19. The 3-dB high-gain antenna shows a 3-dB gain over the quarter-wave antenna. However, the gain of the antenna used at the mobile unit must be limited to 3 dB because the cell-site antenna is rarely as high as the broadcasting antenna and out-of-sight conditions often prevail. The mobile antenna with a gain of more than 3 dB can receive only a limited portion of the total multipath signal in the elevation as measured under the out-of-sight condition.[19] This point is discussed in detail in Sec. 5.6.3.

5.6.2 Glass-mounted antennas[10,20]

There are many kinds of glass-mounted antennas. Energy is coupled through the glass; therefore, there is no need to drill a hole. However, some energy is dissipated on passage through the glass. The antenna gain range is 1 to 3 dB depending on the operating frequency.

The position of the glass-mounted antenna is always lower than that of the roof-mounted antenna; generally there is a 3-dB difference between these two types of antenna. Also, glass-mounted antennas cannot be installed on the shaded glass found in some motor vehicles because this type of glass has a high metal content.

5.6.3 Mobile high-gain antennas

A high-gain antenna used on a mobile unit has been studied.[19] This type of high-gain antenna should be distinguished from the directional antenna. In the directional antenna, the antenna beam pattern is suppressed horizontally; in the high-gain antenna, the pattern is suppressed vertically. To apply either a directional antenna or a high-gain antenna for reception in a radio environment, we must know the origin of the signal. If we point the directional antenna opposite to the transmitter site, we would in theory receive nothing.

In a mobile radio environment, the scattered signals arrive at the mobile unit from every direction with equal probability. That is why an omnidirectional antenna must be used. The scattered signals also arrive from different elevation angles. Lee and Brandt[19] used two types of antenna, one $\lambda/4$ whip antenna with an elevation coverage of 39° and one 4-dB-gain antenna (4-dB gain with respect to the gain of a

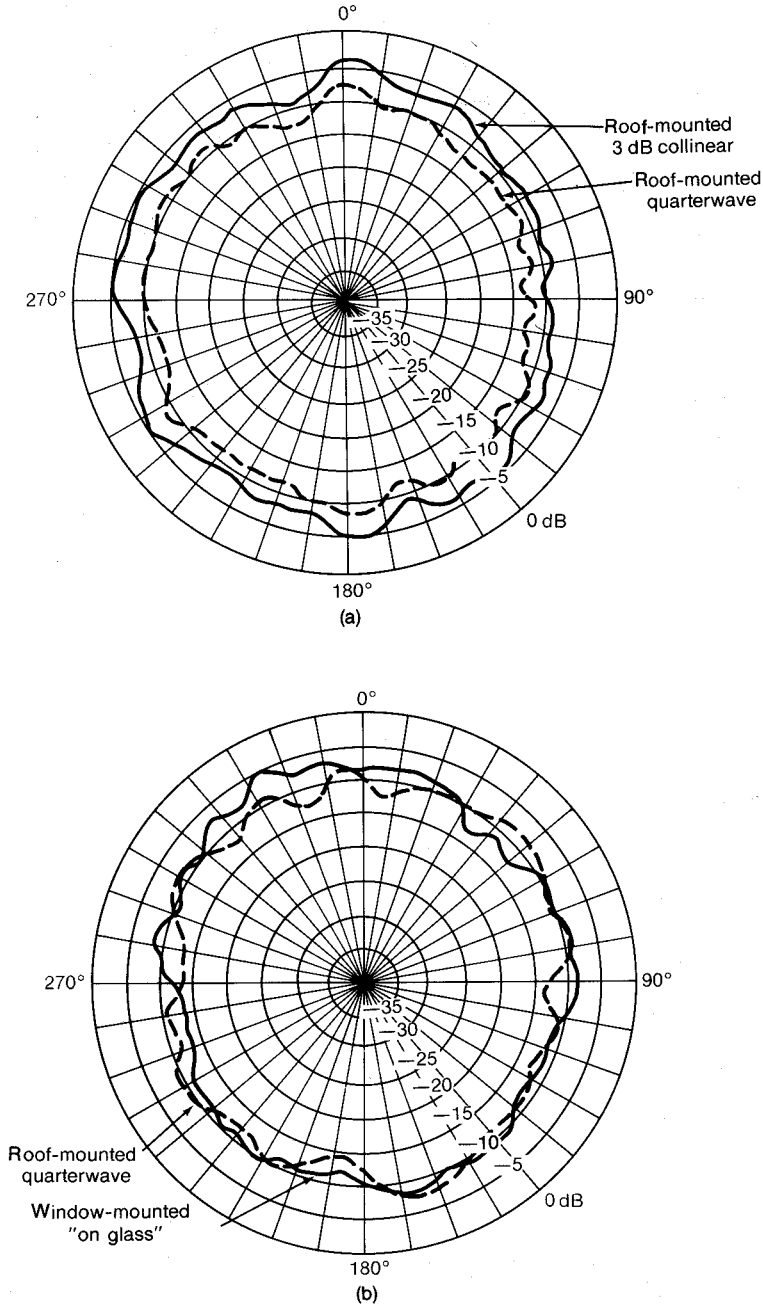


Figure 5.18 Mobile antenna patterns (*from Antenna Specialist Co., Ref. 10*). (*a*) Roof-mounted 3-dB-gain collinear antenna versus roof-mounted quarter-wave antenna. (*b*) Window-mounted "on-glass" gain antenna versus roof-mounted quarter-wave antenna.

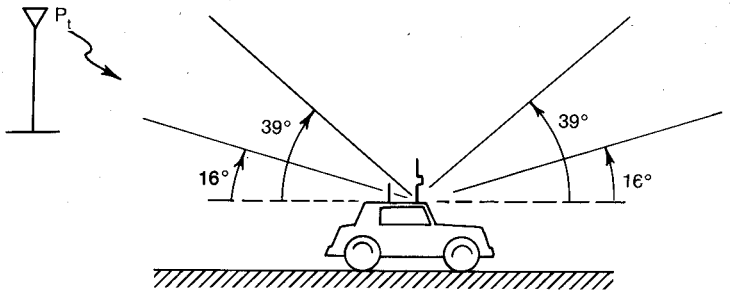


Figure 5.19 Vertical angle of signal arrival.

dipole) with an elevation coverage of 16°, and measured the angle of signal arrival in the suburban Keyport-Matawan area of New Jersey. There are two types of test: a line-of-sight condition and an out-of-sight condition. In Lee and Brandt's study the transmitter was located at an elevation of approximately 100 m (300 ft) above sea level. The measured areas were about 12 m (40 ft) above sea level and the path length about 3 mi. The received signal from the 4-dB-gain antenna was 4 dB stronger than that from the whip antenna under line-of-sight conditions. This is what we would expect. However, the received signal from the 4-dB-gain antenna was only about 2 dB stronger than that from the whip antenna under out-of-sight conditions. This is surprising.

The reason for the latter observation is that the scattered signals arriving under out-of-sight conditions are spread over a wide elevation angle. A large portion of the signals outside the elevation angle of 16° cannot be received by the high-gain antenna. We may calculate the portion being received by the high-gain antenna from the measured beamwidth [the beamwidth can be roughly obtained from Eq. (5.2-6)]. For instance, suppose that a 4:1 gain (6 dBi) is expected from the high gain antenna, but only 2.5:1 is received. Therefore, 63 percent of the signal* is received by the 4-dB-gain antenna (i.e., 6 dBi) and 37 percent is felt in the region between 16 and 39°. Consider the data in the following table.

	Gain, dBi	Linear ratio	$\theta_0/2$, degrees
Whip antenna (2 dB above isotropic)	2	1.58:1	39
High-gain antenna	6	4:1	16
Low-gain antenna	4	2.5:1	24

* For a Rayleigh fading signal, 63 percent will be below its power level.

Therefore, a 2- to 3-dB-gain antenna (4 to 5 dBi) should be adequate for general use. An antenna gain higher than 2 to 3 dB does not serve the purpose of enhancing reception level. Moreover, measurements reveal that the elevation angle for scattered signals received in urban areas is greater than that in suburban areas.

5.6.4 Horizontally oriented space-diversity antennas

A two-branch space-diversity receiver mounted on a motor vehicle has the advantage of reducing fading and thus can operate at a lower

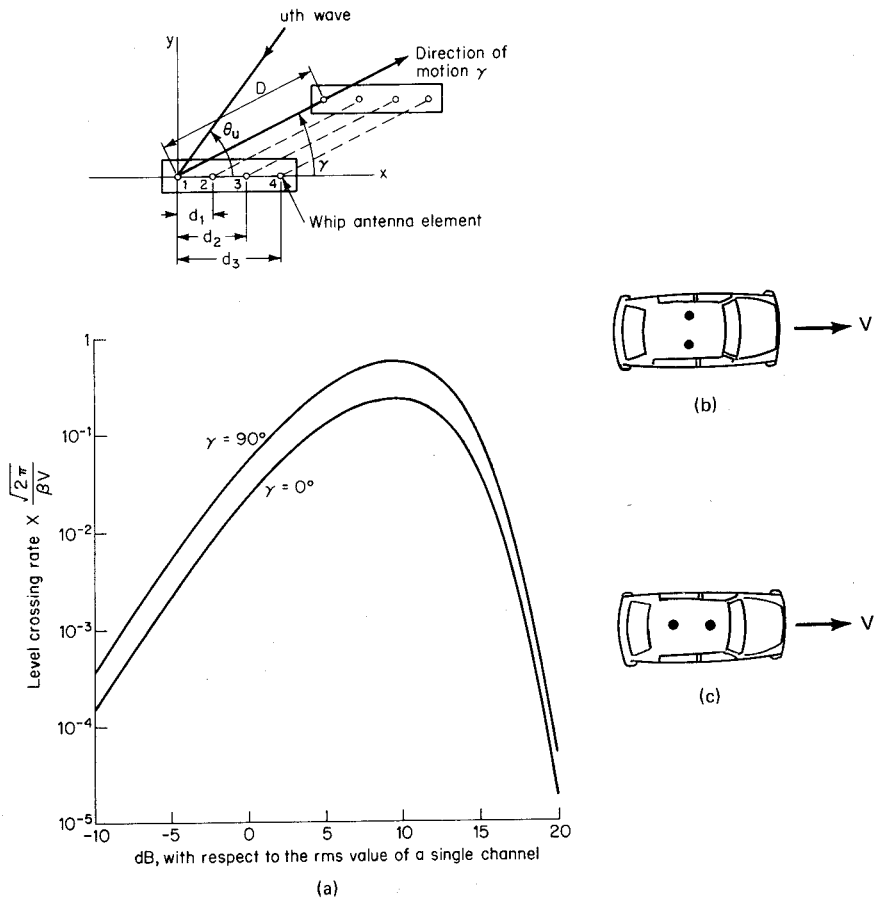


Figure 5.20 Horizontally spaced antennas. (*a*) Maximum difference in lcr of a four-branch equal-gain signal between $\alpha = 0$ and $\alpha = 90°$ with antenna spacing of 0.15λ. (*b*) Not recommended. (*c*) Recommended.

reception level. The advantage of using a space-diversity receiver to reduce interference is discussed in Chap. 7. The discussion here concerns a space-diversity scheme in which two vehicle-mounted antennas separated horizontally by 0.5λ wavelength[21] (15 cm or 6 in) can achieve the advantage of diversity.

We must consider the following factor. The two antennas can be mounted either in line with or perpendicular to the motion of the vehicle. Theoretical analyses and measured data indicate that the in-line arrangement of the two antennas produces fewer level crossings, that is, less fading, than the perpendicular arrangement does. The level crossing rates of two signals received from different horizontally oriented space-diversity antennas are shown in Fig. 5.20.

5.6.5 Vertically oriented space-diversity antennas[22]

The vertical separation between two space-diversity antennas can be determined from the correlation between their received signals. The positions of two antennas X_1 and X_2 are shown in Fig. 5.21. The theoretical derivation of correlation is[23]

$$\rho\left(\frac{d}{\lambda}, \theta\right) = \frac{\sin[(\pi d/\lambda)\sin\theta]}{(\pi d/\lambda)\sin\theta} \tag{5.6-1}$$

Equation (5.6-1) is plotted in Fig. 5.22. A set of measured data was obtained by using two antennas vertically separated by 1.5λ wavelengths. The mean values of three groups of measured data are also shown in Fig. 5.22. In one group, in New York City, low correlation coefficients were observed. In two other groups, both in New Jersey, the average correlation coefficient for perpendicular streets was 0.35

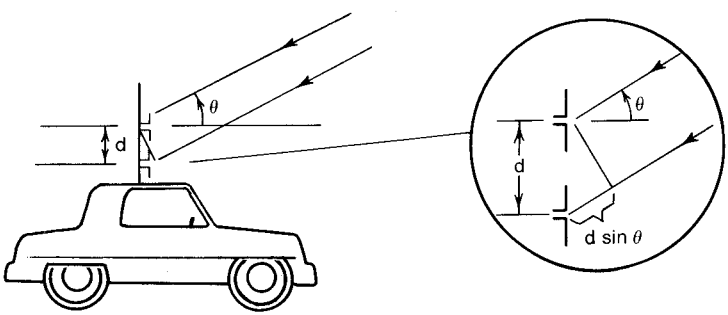


Figure 5.21 Vertical separation between two mobile antennas.

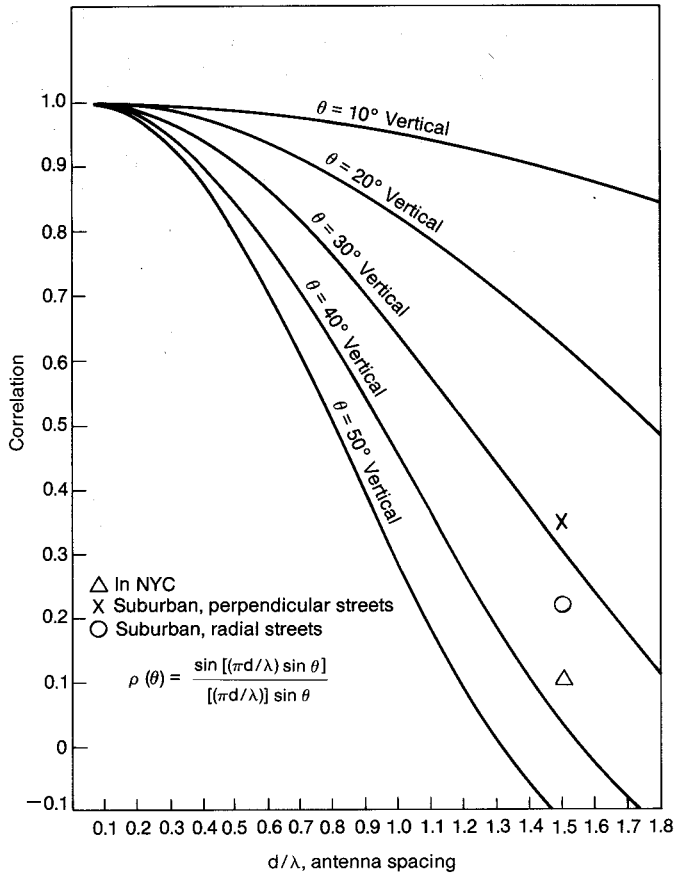


Figure 5.22 Two vertically spaced antennas mounted on a mobile unit.

and for radial streets, 0.225. The following table summarizes the correlation coefficients in different areas and different street orientations.

Area	Correlation coefficient	
	Average	Standard deviation
New York City	0.1	0.06
Suburban New Jersey		
Radial streets	0.226	0.127
Perpendicular streets	0.35	0.182

From Fig. 5.22 we can also see that the signal arrives at an elevation angle of 29° in the suburban radial streets and 33° in the suburban

perpendicular streets. In New York City the angle of arrival approaches 40°.

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Chapter

6

Cochannel Interference Reduction

6.1 Cochannel Interference

The frequency-reuse method is useful for increasing the efficiency of spectrum usage but results in cochannel interference because the same frequency channel is used repeatedly in different cochannel cells. Application of the cochannel interference reduction factor $q = D/R = 4.6$ for a seven-cell reuse pattern ($K = 7$) is described in Sec. 2.4.[1]

In most mobile radio environments, use of a seven-cell reuse pattern is not sufficient to avoid cochannel interference. Increasing $K > 7$ would reduce the number of channels per cell, and that would also reduce spectrum efficiency. Therefore, it might be advisable to retain the same number of radios as the seven-cell system but to sector the cell radially, as if slicing a pie. This technique would reduce cochannel interference and use channel sharing and channel borrowing schemes to increase spectrum efficiency.

6.2 Exploring Cochannel Interference Areas in a System

Problems in mobile telephone coverage (service), particularly holes (weak signal strength*) which result in call drops during the customer's

* Signal strength is measured in dBm, and field strength is measured in dB(μV/m). The conversion between these two units can be found in Sec. 5.1.3.

conversation, have been partially solved by applying the propagation (wave motion) studies discussed in Chap. 4 for the case where no cochannel interference exists.

When customer demand increases, the channels, which are limited in number, have to be repeatedly reused in different areas, which provides many cochannel cells, which increases the system's capacity. But cochannel interference may be the result. In this situation, the received voice quality is affected by both the grade of coverage and the amount of cochannel interference. For detection of serious cochannel interference areas in a cellular system, two tests are suggested.

Test 1—find the cochannel interference area from a mobile receiver. Cochannel interference which occurs in one channel will occur equally in all the other channels in a given area. We can then measure cochannel interference by selecting any one channel (as one channel represents all the channels) and transmitting on that channel at all cochannel sites at night while the mobile receiver is traveling in one of the cochannel cells.

While performing this test we watch for any change detected by a field-strength recorder in the mobile unit and compare the data with the condition of no cochannel sites being transmitted. This test must be repeated as the mobile unit travels in every cochannel cell. To facilitate this test, we can install a channel scanning receiver in one car.

One channel (f_1) records the signal level (no-cochannel condition), another channel (f_2) records the interference level (six-cochannel condition is the maximum), while the third channel receives f_3, which is not in use. Therefore, the noise level is recorded only in f_3 (see Fig. 6.1).

We can obtain, in decibels, the carrier-to-interference ratio C/I by subtracting the result obtained from f_2 from the result obtained from f_1 (carrier minus interference $C - I$) and the carrier-to-noise ratio C/N by subtracting the result obtained from f_3 from the result obtained from f_2 (carrier minus noise $C - N$). Four conditions should be used to compare the results.

1. If the carrier-to-interference ratio C/I is greater than 18 dB throughout most of the cell, the system is properly designed.
2. If C/I is less than 18 dB and C/N is greater than 18 dB in some areas, there is cochannel interference.
3. If both C/N and C/I are less than 18 dB and $C/N \approx C/I$ in a given area, there is a coverage problem.
4. If both C/N and C/I are less than 18 dB and $C/N > C/I$ in a given area, there is a coverage problem *and* cochannel interference.

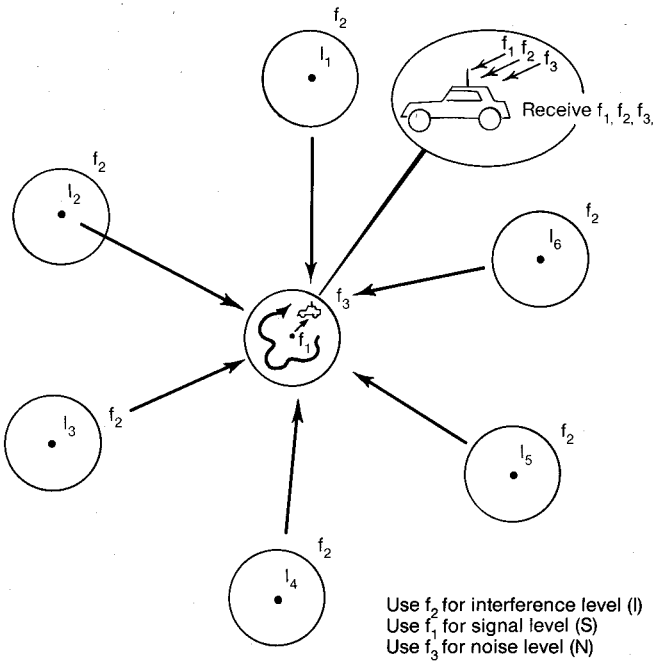

Figure 6.1 Test 1: cochannel interference at the mobile unit.

Test 2—find the cochannel interference area which affects a cell site. The reciprocity theorem can be applied for the coverage problem but not for cochannel interference. Therefore, we cannot assume that the first test result will apply to the second test condition. We must perform the second test as well.

Because it is difficult to use seven cars simultaneously, with each car traveling in each cochannel cell for this test, an alternative approach may be to record the signal strength at every cochannel cell site while a mobile unit is traveling either in its own cell or in one of the cochannel cells shown in Fig. 6.2.

First we find the areas in an interfering cell in which the top 10 percent level of the signal transmitted from the mobile unit in those areas is received at the desired cell site (Jth cell in Fig. 6.1). This top 10 percent level can be distributed in different areas in a cell. The average value of the top 10 percent level of signal strength is used as the interference level from that particular interfering cell. The mobile unit also travels in different interfering cells. Up to six interference levels are obtained from a mobile unit running in six interfering cells. We then calculate the average of the bottom 10 percent level of the signal strength which is transmitted from a mobile unit in the desired

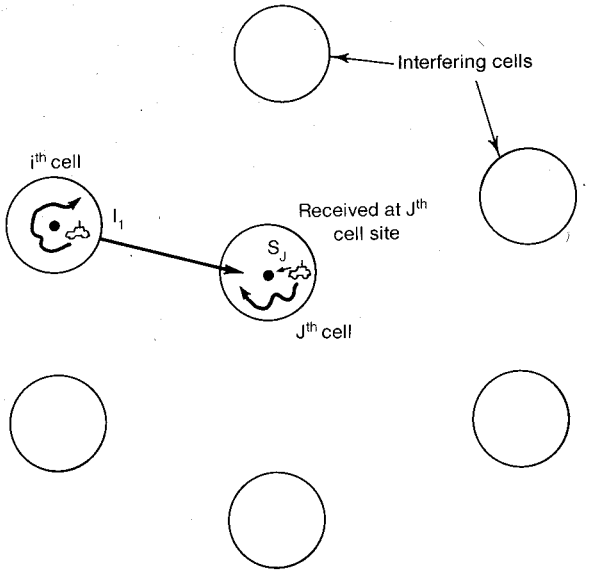

Figure 6.2 Test 2: cochannel interference at the cell site.

cell (Jth cell) and received at the desired cell site as a carrier reception level.

Then we can reestablish the carrier-to-interference ratio received at a desired cell, say, the Jth cell site as follows.

$$\frac{C_J}{I} = \frac{C_J}{\sum_{\substack{i=1 \\ i \neq J}}^{6} I_i}$$

The number of cochannel cells in the system can be less than six. We must be aware that all C_J and I_i were read in decibels. Therefore, a translation from decibels to linear is needed before summing all the interfering sources. The test can be carried out repeatedly for any given cell. We then compare

$$\frac{C_J}{I} \quad \text{and} \quad \frac{C_J}{N_J}$$

and determine the cochannel interference condition, which will be the same as that in test 1. N_J is the noise level in the Jth cell assuming no interference exists.

6.3 Real-Time Cochannel Interference Measurement at Mobile Radio Transceivers

When the carriers are angularly modulated by the voice signal and the RF frequency difference between them is much higher than the fading frequency, measurement of the signal carrier-to-interference ratio C/I reveals that the signal is

$$e_1 = S(t)\sin(\omega t + \phi_1) \tag{6.3-1}$$

and the interference is

$$e_2 = I(t)\sin(\omega t + \phi_2) \tag{6.3-2}$$

The received signal is

$$e(t) = e_1(t) + e_2(t) = R\sin(\omega t + \psi) \tag{6.3-3}$$

where

$$R = \sqrt{[S(t)\cos\phi_1 + I(t)\cos\phi_2]^2 + [S(t)\sin\phi_1 + I(t)\sin\phi_2]^2} \tag{6.3-4}$$

and

$$\psi = \tan^{-1}\frac{S(t)\sin\phi_1 + I(t)\sin\phi_2}{S(t)\cos\phi_1 + I(t)\cos\phi_2} \tag{6.3-5}$$

The envelope R can be simplified in Eq. (6.3-4), and R^2 becomes

$$R^2 = \{S^2(t) + I^2(t) + 2S(t)I(t)\cos(\phi_1 - \phi_2)\} \tag{6.3-6}$$

Following Kozono and Sakamoto's[2] analysis of Eq. (6.3-6) the term $S^2(t) + I^2(t)$ fluctuates close to the fading frequency V/λ and the term $2S(t)I(t)\cos(\phi_1 - \phi_1)$ fluctuates to a frequency close to $d/dt(\phi_1 - \phi_2)$, which is much higher than the fading frequency. Then the two parts of the squared envelope can be separated as

$$X = S^2(t) + I^2(t) \tag{6.3-7}$$

$$Y = 2S(t)I(t)\cos(\phi_1 - \phi_2) \tag{6.3-8}$$

Assume that the random variables $S(t)$, $I(t)$, ϕ_1, and ϕ_2 are independent; then the average processes on X and Y are

$$\overline{X} = \overline{S^2(t)} + \overline{I^2(t)} \tag{6.3-9}$$

$$\overline{Y^2} = 4\overline{S^2(t)}\,\overline{I^2(t)}(1/2) = 2\overline{S^2(t)}\,\overline{I^2(t)} \tag{6.3-10}$$

The signal-to-interference ratio Γ becomes

$$\Gamma = \frac{\overline{S^2(t)}}{\overline{I^2(t)}} = k + \sqrt{k^2 - 1} \tag{6.3-11}$$

where

$$k = \frac{\overline{X}^2}{\overline{Y}^2} - 1 \tag{6.3-12}$$

Since X and Y can be separated in Eq. (6.3-6), the preceding computation of Γ in Eq. (6.3-11) could have been accomplished by means of an envelope detector, an analog-to-digital converter, and a microcomputer. The sampling delay time Δt should be small enough to satisfy

$$S(t) \approx S(t + \Delta t), \qquad I(t) \approx I(t + \Delta t) \tag{6.3-13}$$

and

$$\overline{\cos\,[\phi_1(t) - \phi_2(t)]\cos\,[\phi_1(t + \Delta t) - \phi_2(t + \Delta t)]} \approx 0 \tag{6.3-14}$$

Determining the delay time Δt to meet the requirement of Eq. (6.3-13) for this calculation is difficult and is a drawback to this measurement technique. Therefore, real-time cochannel interference measurement is difficult to achieve in practice.

6.4 Design of an Omnidirectional Antenna System in the Worst Case

In Sec. 2.4 we proved that the value of $q = 4.6$ is valid for a normal interference case in a $K = 7$ cell pattern.[3] In this section we would like to prove that a $K = 7$ cell pattern does not provide a sufficient frequency-reuse distance separation even when an ideal condition of flat terrain is assumed. The worst case is at the location where the mobile unit would receive the weakest signal from its own cell site but strong interferences from all interfering cell sites.

In the worst case the mobile unit is at the cell boundary R, as shown in Fig. 6.3. The distances from all six cochannel interfering sites are also shown in the figure: two distances of $D - R$, two distances of D, and two distances of $D + R$.

Following the mobile radio propagation rule of 40 dB/dec shown in Chap. 4, we obtain

$$C \propto R^{-4} \qquad I \propto D^{-4}$$

Then the carrier-to-interference ratio is

$$\frac{C}{I} = \frac{R^{-4}}{2(D - R)^{-4} + 2(D)^{-4} + 2(D + R)^{-4}}$$

$$= \frac{1}{2(q - 1)^{-4} + 2(q)^{-4} + 2(q + 1)^{-4}} \tag{6.4-1a}$$

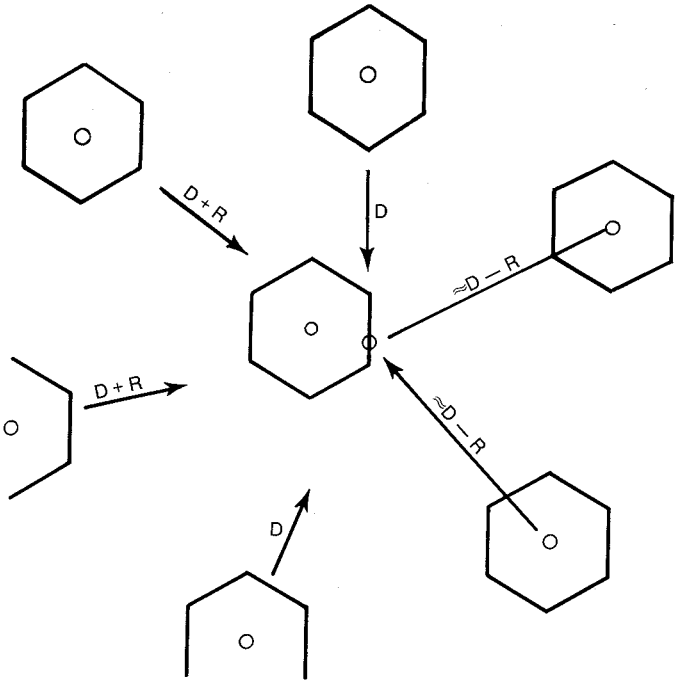

Figure 6.3 Cochannel interference (a worst case).

where $q = 4.6$ is derived from the normal case shown in Eq. (2.4-7). Substituting $q = 4.6$ into Eq. (6.4-1a), we obtain $C/I = 54$ or 17 dB, which is lower than 18 dB. To be conservative, we may use the shortest distance $D - R$ for all six interferers as a worst case; then Eq. (6.4-1a) is replaced by

$$\frac{C}{I} = \frac{R^{-4}}{6(D - R)^{-4}} = \frac{1}{6(q - 1)^{-4}} = 28 = 14.47 \text{ dB} \qquad (6.4\text{-}1b)$$

In reality, because of the imperfect site locations and the rolling nature of the terrain configuration, the C/I received is always worse than 17 dB and could be 14 dB and lower. Such an instance can easily occur in a heavy traffic situation; therefore, the system must be designed around the C/I of the worst case. In that case, a cochannel interference reduction factor of $q = 4.6$ is insufficient.

Therefore, in an omnidirectional-cell system, $K = 9$ or $K = 12$ would be a correct choice. Then the values of q are

$$q = \begin{cases} \dfrac{D}{R} = \sqrt{3K} & \\ 5.2 & K = 9 \\ 6 & K = 12 \end{cases} \qquad (6.4\text{-}2)$$

Substituting these values in Eq. (6.4-1), we obtain

$$\frac{C}{I} = 84.5\ (=)\ 19.25\ \text{dB} \qquad K = 9 \tag{6.4-3}$$

$$\frac{C}{I} = 179.33\ (=)\ 22.54\ \text{dB} \qquad K = 12 \tag{6.4-4}$$

The $K = 9$ and $K = 12$ cell patterns, shown in Fig. 6.4, are used when the traffic is light. Each cell covers an adequate area with adequate numbers of channels to handle the traffic.

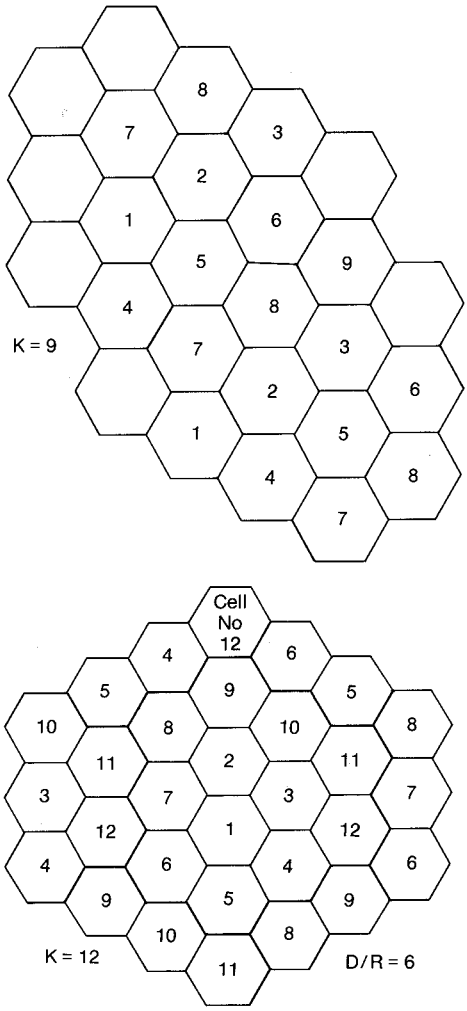

Figure 6.4 Interference with frequency-reuse patterns $K = 9$ and $K = 12$.

6.5 Design of a Directional Antenna System

When the call traffic begins to increase, we need to use the frequency spectrum efficiently and avoid increasing the number of cells K in a seven-cell frequency-reuse pattern. When K increases, the number of frequency channels assigned in a cell must become smaller (assuming a total allocated channel divided by K) and the efficiency of applying the frequency-reuse scheme decreases.

Instead of increasing the number K in a set of cells, let us keep $K = 7$ and introduce a directional-antenna arrangement. The cochannel interference can be reduced by using directional antennas. This means that each cell is divided into three or six sectors and uses three or six directional antennas at a base station. Each sector is assigned a set of frequencies (channels). The interference between two cochannel cells decreases as shown below.

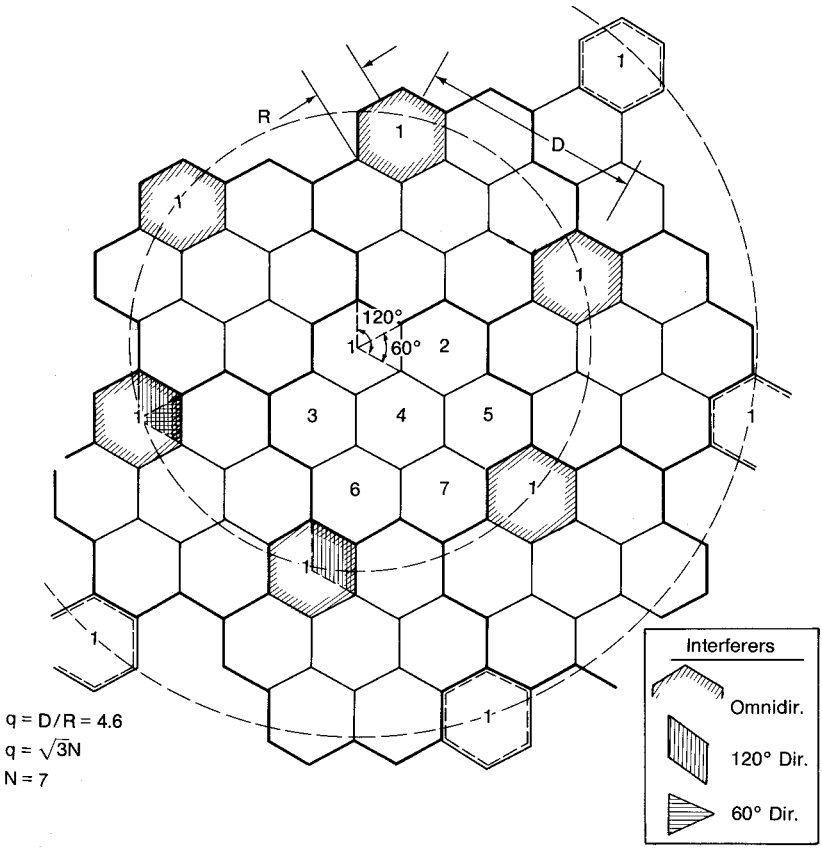

Figure 6.5 Interfering cells shown in a seven-cell system (two tiers).

6.5.1 Directional antennas in $K = 7$ cell patterns

Three-sector case. The three-sector case is shown in Fig. 6.5. To illustrate the worst-case situation, two cochannel cells are shown in Fig. 6.6*a*. The mobile unit at position E will experience greater interference in the lower shaded cell sector than in the upper shaded cell-sector site. This is because the mobile receiver receives the weakest signal from its own cell but fairly strong interference from the interfering cell. In a three-sector case, the interference is effective in only one direction because the front-to-back ratio of a cell-site directional antenna is at least 10 dB or more in a mobile radio environment. The worst-case cochannel interference in the directional-antenna sectors

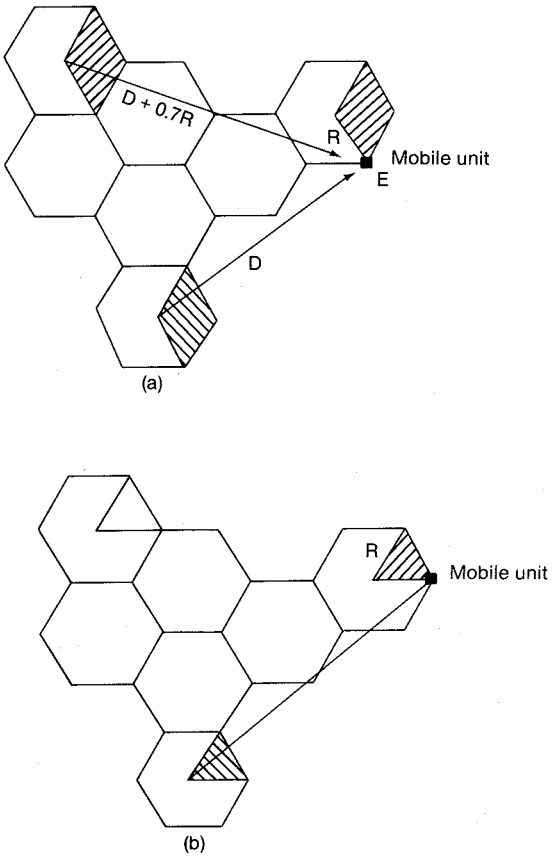

Figure 6.6 Determination of carrier-to-interference ratio C/I in a directional antenna system. (*a*) Worst case in a 120° directional antenna system ($N = 7$); (*b*) worst case in a 60° directional antenna system ($N = 7$).

in which interference occurs may be calculated. Because of the use of directional antennas, the number of principal interferers is reduced from six to two (Fig. 6.5). The worst case of C/I occurs when the mobile unit is at position E, at which point the distance between the mobile unit and the two interfering antennas is roughly $D + (R/2)$; however, C/I can be calculated more precisely as follows.* The value of C/I can be obtained by the following expression (assuming that the worst case is at position E at which the distances from two interferers are $D + 0.7$ and D).

$$\frac{C}{I}\text{ (worst case)} = \frac{R^{-4}}{(D + 0.7R)^{-4} + D^{-4}}$$

$$= \frac{1}{(q + 0.7)^{-4} + q^{-4}} \tag{6.5-1}$$

Let $q = 4.6$; then Eq. (6.5-1) becomes

$$\frac{C}{I}\text{ (worst case)} = 285\ (=)\ 24.5\text{ dB} \tag{6.5-2}$$

The C/I received by a mobile unit from the 120° directional antenna sector system expressed in Eq. (6.5-2) greatly exceeds 18 dB in a worst case. Equation (6.5-2) shows that using directional antenna sectors can improve the signal-to-interference ratio, that is, reduce the cochannel interference. However, in reality, the C/I could be 6 dB weaker than in Eq. (6.5-2) in a heavy traffic area as a result of irregular terrain contour and imperfect site locations. The remaining 18.5 dB is still adequate.

Six-sector case. We may also divide a cell into six sectors by using six 60°-beam directional antennas as shown in Fig. 6.6*b*. In this case, only one instance of interference can occur in each sector as shown in Fig. 6.5. Therefore, the carrier-to-interference ratio in this case is

$$\frac{C}{I} = \frac{R^{-4}}{(D + 0.7R)^{-4}} = (q + 0.7)^4 \tag{6.5-3}$$

For $q = 4.6$, Eq. (6.5-3) becomes

$$\frac{C}{I} = 794\ (=)\ 29\text{ dB} \tag{6.5-4}$$

* The difference in results between using a closed form and an approximate calculation is small.

which shows a further reduction of cochannel interference. If we use the same argument as we did for Eq. (6.5-2) and subtract 6 dB from the result of Eq. (6.5-4), the remaining 23 dB is still more than adequate. When heavy traffic occurs, the 60°-sector configuration can be used to reduce cochannel interference. However, fewer channels are generally allowed in a 60° sector and the trunking efficiency decreases. In certain cases, more available channels could be assigned in a 60° sector.

6.5.2 Directional antenna in $K = 4$ cell pattern

Three-sector case. To obtain the carrier-to-interference ratio, we use the same procedure as in the $K = 7$ cell-pattern system. The 120° directional antennas used in the sectors reduced the interferers to two as in $K = 7$ systems, as shown in Fig. 6.7. We can apply Eq. (6.5-1) here. For $K = 4$, the value of $q = \sqrt{3K} = 3.46$; therefore, Eq. (6.5-1) becomes

$$\frac{C}{I}\text{ (worst case)} = \frac{1}{(q + 0.7)^{-4} + q^{-4}} = 97 = 20\text{ dB} \qquad (6.5\text{-}5)$$

If, using the same reasoning used with Eq. (6.5-4), 6 dB is subtracted from the result of Eq. (6.5-5), the remaining 14 dB is unacceptable.

Six-sector case. There is only one interferer at a distance of $D + R$ shown in Fig. 6.7. With $q = 3.46$, we can obtain

$$\frac{C}{I}\text{ (worst case)} = \frac{R^{-4}}{(D + R)^{-4}} = \frac{1}{(q + 1)^{-4}} = 359.5 = 27\text{ dB} \qquad (6.5\text{-}6)$$

If 6 dB is subtracted from the result of Eq. (6.5-6), the remaining 21 dB is adequate. Under heavy traffic conditions, there is still a great

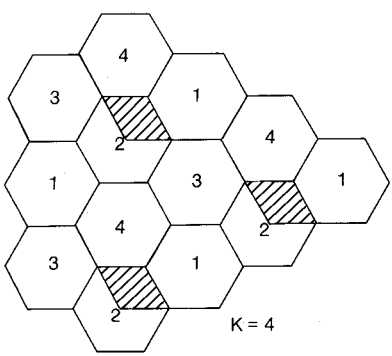

Figure 6.7 Interference with frequency-reuse pattern $K = 4$.

deal of concern over using a $K = 4$ cell pattern in a 60° sector. An explanation of this point is given in the next section.

6.5.3 Comparing $K = 7$ and $K = 4$ systems

A $K = 7$ cell-pattern system is a logical way to begin an omnicell system. The cochannel reuse distance is more or less adequate, according to the designed criterion. When the traffic increases, a three-sector system should be implemented, that is, with three 120° directional antennas in place. In certain hot spots, 60° sectors can be used locally to increase the channel utilization.

If a given area is covered by both $K = 7$ and $K = 4$ cell patterns and both patterns have a six-sector configuration, then the $K = 7$ system has a total of 42 sectors, but the $K = 4$ system has a total of only 26 sectors and, of course, the system of $K = 7$ and six sectors has less cochannel interference.

One advantage of 60° sectors with $K = 4$ is that they require fewer cell sites than 120° sectors with $K = 7$. Two disadvantages of 60° sectors are that (1) they require more antennas to be mounted on the antenna mast and (2) they often require more frequent handoffs because of the increased chance that the mobile units will travel across the six sectors of the cell. Furthermore, assigning the proper frequency channel to the mobile unit in each sector is more difficult unless the antenna height at the cell site is increased so that the mobile unit can be located more precisely. In reality the terrain is not flat, and coverage is never uniformly distributed; in addition, the directional antenna front-to-back power ratio in the field is very difficult to predict (see Sec. 5.4.2). In small cells, interference could become uncontrollable; thus the use of a $K = 4$ pattern with 60° sectors in small cells needs to be considered only for special implementations such as portable cellular systems (Sec. 13.5) or narrowbeam applications (Sec. 10.6). For small cells, a better alternative scheme is to use a $K = 7$ pattern with 120° sectors plus the underlay-overlay configuration described in Sec. 11.4.

6.6 Lowering the Antenna Height

Lowering the antenna height does not always reduce the cochannel interference. In some circumstances, such as on fairly flat ground or in a valley situation, lowering the antenna height will be very effective for reducing the cochannel and adjacent-channel interference. However, there are three cases where lowering the antenna height may or may not effectively help reduce the interference.

On a high hill or a high spot. The effective antenna height, rather than the actual height, is always considered in the system design. Therefore, the effective antenna height varies according to the location of the mobile unit, as described in Chap. 4. When the antenna site is on a hill, as shown in Fig. 6.8*a*, the effective antenna height is $h_1 + H$. If we reduce the actual antenna height to $0.5h_1$, the effective antenna height becomes $0.5h_1 + H$. The reduction in gain resulting from the height reduction is

$$G = \text{gain reduction} = 20 \log_{10} \frac{0.5h_1 + H}{h_1 + H}$$

$$= 20 \log_{10} \left(1 - \frac{0.5h_1}{h_1 + H}\right) \qquad (6.6\text{-}1)$$

If $h_1 \ll H$, then Eq. (6.6-1) becomes

$$G \approx 20 \log_{10} 1 = 0 \text{ dB}$$

This simply proves that lowering antenna height on the hill does not reduce the received power at either the cell site or the mobile unit.

In a valley. The effective antenna height as seen from the mobile unit shown in Fig. 6.8*b* is h_{e1}, which is less than the actual antenna height

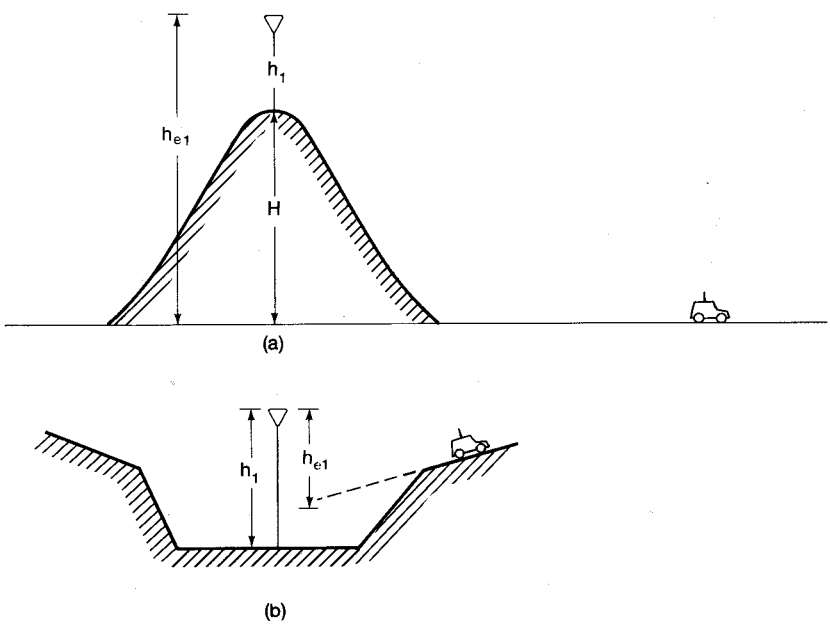

Figure 6.8 Lowering the antenna height (*a*) on a high hill and (*b*) in a valley.

h_1. If $h_{e1} = \frac{2}{3}h_1$ and the antenna is lowered to $\frac{1}{2}h_1$, then the new effective antenna height, determined from Chap. 4, is

$$h_{e1} = \tfrac{1}{2}h_1 - (h_1 - \tfrac{2}{3}h_1) = \tfrac{1}{6}h_1$$

Then the antenna gain is reduced by

$$G = 20 \log \frac{\frac{1}{6}h_1}{\frac{2}{3}h_1} = -12 \text{ dB}$$

This simply proves that the lowered antenna height in a valley is very effective in reducing the radiated power in a distant high elevation area. However, in the area adjacent to the cell-site antenna, the effective antenna height is the same as the actual antenna height. The power reduction caused by decreasing antenna height by half is only

$$20 \log \frac{\frac{1}{2}h_1}{h_1} = -6 \text{ dB}$$

In a forested area. In a forested area, the antenna should clear the tops of any trees in the vicinity, especially when they are very close to the antenna. In this case decreasing the height of the antenna would not be the proper procedure for reducing cochannel interference because excessive attenuation of the desired signal would occur in the vicinity of the antenna and in its cell boundary if the antenna were below the treetop level. This phenomenon is described in Sec. 4.4.

6.7 Reduction of Cochannel Interference by Means of a Notch in the Tilted Antenna Pattern

6.7.1 Introduction

Reduction of cochannel interference in a cellular mobile system is always a challenging problem. A number of methods can be considered, such as (1) increasing the separation between two cochannel cells, (2) using directional antennas at the base station, or (3) lowering the antenna heights at the base station. Method 1 is not advisable because as the number of frequency-reuse cells increases, the system efficiency, which is directly proportional to the number of channels per cell, decreases. Method 3 is not recommended because such an arrangement also weakens the reception level at the mobile unit. However, method 2 is a good approach, especially when the number of frequency-reuse cells is fixed. The use of directional antennas in each cell can serve two purposes: (1) further reduction of cochannel interference if the

interference cannot be eliminated by a fixed separation of cochannel cells and (2) increasing the channel capacity when the traffic increases. In this chapter we try to further reduce the cochannel interference by intelligently setting up the directional antenna.

6.7.2 Theoretical analysis

Under normal circumstances radiation from a cochannel serving site can easily interfere with another cochannel cell as shown in Fig. 6.8. Installation of a 120° directional antenna can reduce the interference in the system by eliminating the radiation to the rest of its 240° sector. However, cochannel interference can exist even when a directional antenna is used, as the serving site can interfere with the cochannel cell that is directly ahead. Let us assume that a seven-cell cellular system ($K = 7$) is used. The cochannel interference reduction factor q becomes

$$q = \sqrt{3N} = 4.6 \tag{6.7-1}$$

and the cochannel cell separation D can be found if the cell radius is known.

$$D = qR = 4.6R \tag{6.7-2}$$

With a separation of $4.6R$, the area of interference at the interference-receiving cell is illuminated by the central 19° sector of the entire (120°) transmitting antenna pattern at the serving cell (see Fig. 6.9). If three identical directional antennas are implemented in every cell, with each antenna covering a 120° sector, then every sector receives interference in the central 19° sector of the entire 120° angle at the interfering cell. Therefore, attempts should be made to reduce the signal strength of the interference in this 19° sector.

To achieve a significant gain of C/I in the interference-receiving cell, we should consider using a notch in the center of the antenna pattern at the interfering cell. An antenna pattern with a notch in the center can be obtained in a number of ways. One relatively simple way is to tilt the high-gain directional antenna downward.[4] A discussion of this method follows.

6.7.3 The effect of tilting antenna on the coverage pattern

Because the shape of the antenna pattern at the base station relates directly to the reception level of signal strength at the mobile unit, the following antenna pattern effect must be analyzed.

When a high-gain directional antenna (the pattern in the horizontal

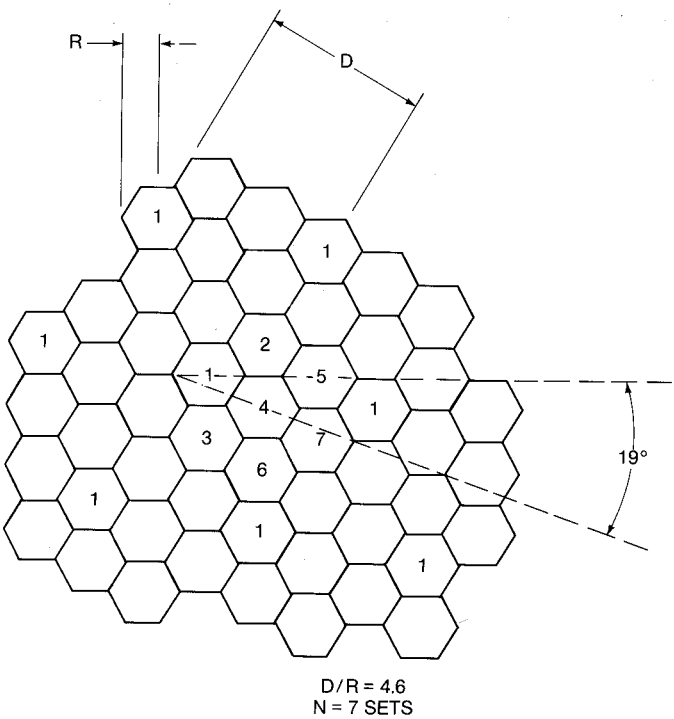

Figure 6.9 A seven-cell cellular configuration.

x-y plane is shown in Fig. 6.10 and in the vertical x-z plane, in Fig. 6.11) is physically tilted at an angle θ in the x-y plane shown in Fig. 6.11, how does the pattern in the x-y plane change? The antenna pattern obtained in the x-y plane after tilting the antenna is shown in Fig. 6.11. When the center beam is tilted downward by an angle θ, the off-center beam is tilted downward by only an angle ψ as shown in Fig. 6.12. The pattern in the x-y plane can be plotted by varying the angle ϕ. From the diagram in Fig. 6.12, we can obtain a derivation which provides the relationship among the angles ψ, θ, and ϕ as

$$\sin\frac{\theta}{2} = \frac{d}{l} \tag{6.7-3}$$

$$\frac{\overline{DB}}{\sin\phi} = \frac{l}{\sin(135^\circ - \phi)} \tag{6.7-4}$$

$$\overline{CD} = l\,\frac{\sin 45^\circ}{\sin(135 - \phi)} \tag{6.7-5}$$

$$\frac{\overline{AD}}{\overline{DF}} = \frac{\overline{AB}}{2d} = \frac{\sqrt{2}l}{2d} \tag{6.7-6}$$

$$\overline{AD} = \overline{AB} - \overline{DB} \tag{6.7-7}$$

$$\cos\psi = \frac{2\overline{CD} - \overline{DF}}{2\overline{CD}} = 1 - \frac{\overline{DF}}{2\overline{CD}} \tag{6.7-8}$$

Substituting Eqs. (6.7-3) to (6.7-7) into Eq. (6.7-8), we obtain

$$\cos\psi = 1 - \cos^2\phi\,(1 - \cos\theta) \tag{6.7-9}$$

or

$$\psi = \cos^{-1}[1 - \cos^2\phi\,(1 - \cos\theta)] \tag{6.7-10}$$

If the physically tilted angle is $\theta = 18°$, then the off-center beam ψ is tilted downward.

$$\phi = \begin{cases} 0° \\ 45° \\ 90° \end{cases} \qquad \psi = \begin{cases} 18° = \theta \\ 12.7° \\ 0° \end{cases}$$

This list tells us that the physically tilted angle ϕ and the angle ψ are not linearly related, and that when $\phi = 90°$, then $\psi = 0°$. When the

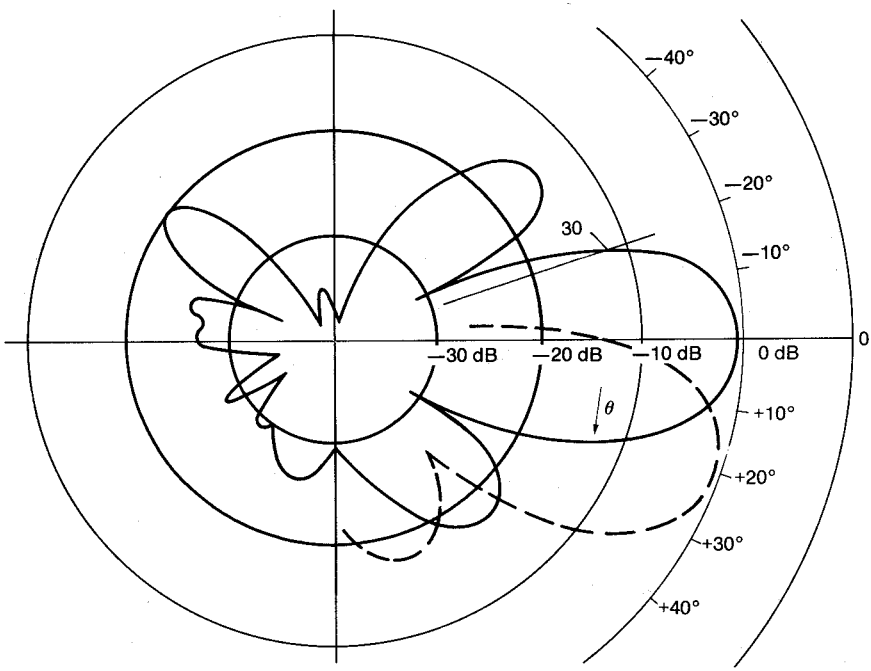

Figure 6.10 Vertical antenna pattern of a 120° directional antenna.

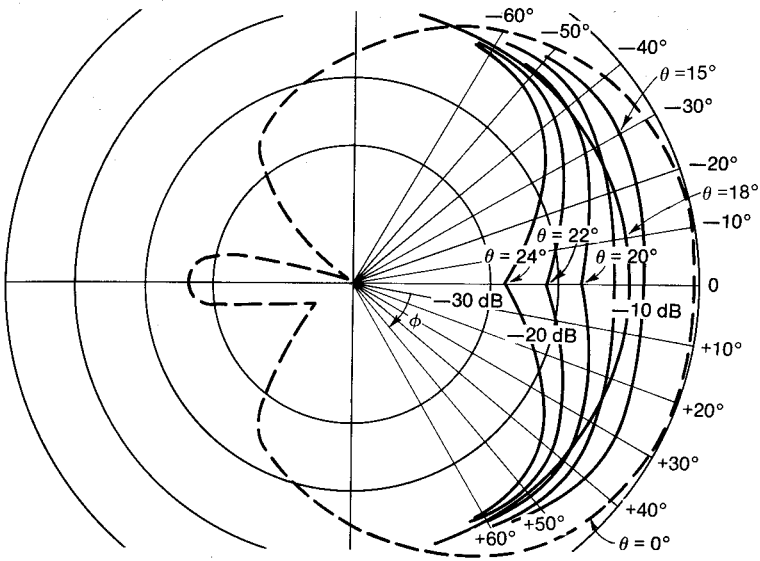

Figure 6.11 Notch appearing in tilted antenna pattern. (*Reprinted after Lee, Ref. 4.*)

angle θ increases beyond 18°, the notch effect of the pattern in the x-y plane becomes evident, as indicated in Fig. 6.11.

6.7.4 Suggested method for reducing interference

Suppose that we would like to take advantage of this notch effect. From Fig. 6.13, we notice that the interfering site could cause interference at those cells within a 19° sector in front of the cell.

In an ideal situation such as that shown in Fig. 6.13, the antenna

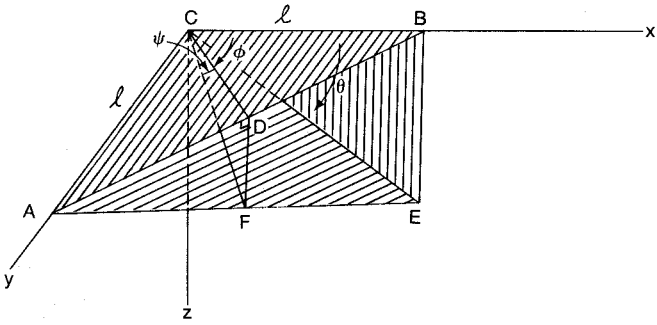

Figure 6.12 Coordinate of the tilting antenna pattern.

pattern of the serving cell must be rotated clockwise by 10° such that the notch can be aimed properly at the interfering cell. The antenna tilting angle θ may be between 22 to 24° in order to increase the carrier-to-interference ratio C/I by an additional 7 to 8 dB in the interfering cell as shown in Fig. 6.11. Now we can reduce cochannel interference by an additional 7 to 8 dB because of the notch in the tilted-antenna pattern. Although signal coverage is rather weak in a small shaded area in the serving cell, as shown in Fig. 6.13, the use of sufficient transmitting power should correct this situation.

6.7.5 Cautions in tilting antennas

When a base-station antenna is tilted down by 10°, the strength of the received signal in the horizontal direction, as shown in Fig. 6.10, is decreased by 4 dB. But the strength of the received signal 1° below the horizontal is decreased by 3.5 dB—only 0.5 dB stronger than in the 0° case. This is a very important observation. For example, the elevation angle at the boundary of a 2-mi serving cell with a 100-ft antenna mast is about 0.5°. This means that the serving cell and the interfering cell are separated by only 0.5° at most. Then by tilting the antenna down by 10°, the interference by the interfering cell is reduced by an additional 0.25 dB. This is an insignificant improvement, yet the total power received is 4 dB less than in the no-tilt case. If the tilt is increased to 20°, the received power drops by 16 dB and the reduction in interference due to tilting the antenna is only 1 dB at the interfering cell (see Fig. 6.10). The justification for implementing the tilting antenna is that the new carrier-to-interference ratio ($\alpha C/\beta I$) after tilting is

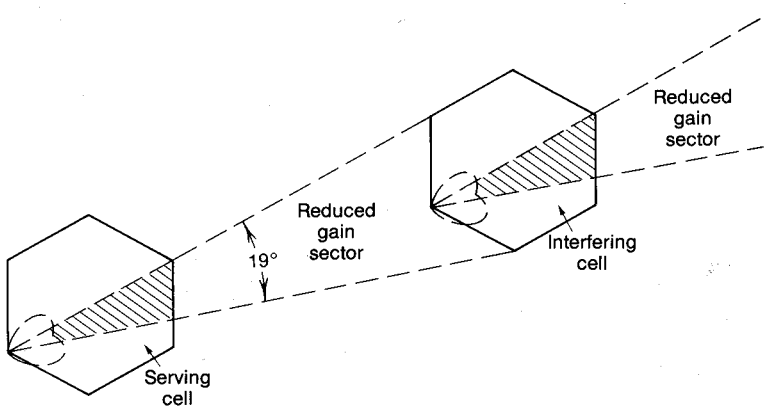

Figure 6.13 Reduced-gain sector of two cochannel cells.

significantly higher than C/I before tilting, where α and β are those constants which can be expressed if the following expression holds.

$$\frac{\alpha C}{\beta I} \text{ (linear scale)} \Longrightarrow \frac{C}{I} + (\alpha - \beta) \text{ (dB scale)}$$

In the above example, at a 10° tilt, $\alpha = 3.75$ dB and $\beta = 4$ dB, and the improved new carrier-to-interference ratio is $(C/I) + 0.25$ dB, which is an insignificant improvement. Therefore, the antenna vertical pattern and the antenna height play a major role in justifying antenna tilting. Some calculations are shown in Sec. 6.8.2. Sometimes, tilting the antenna upward may increase signal coverage if interference is not a problem.

6.8 Umbrella-Pattern Effect

The umbrella pattern can be achieved by use of a staggered discone antenna as discussed in Sec. 5.4.6. The umbrella pattern can be applied to reduce cochannel interference just as the downward tilted directional antenna pattern is. The umbrella pattern can be used for an omnidirectional pattern, but not for a directional antenna pattern. The tilted directional antenna pattern can create a notch after tilting 20° or more in front of the beam, but the umbrella pattern cannot.

Of most concern for future cellular systems is the long-distance interference due to tropospheric propagation as mentioned in Sec. 4.6. In the future, one system may experience long-distance interference resulting from other systems located approximately 320 km (200 mi) away. Cochannel interference, especially cross talk, could be a severe problem. Therefore, the umbrella pattern might be recommended for every cell site where interference prevails.

6.8.1 Elevation angle of long-distance propagation

The elevation of the tropospheric layer is 16 km (10 mi)[5] and the propagation distance is about 320 km (200 mi); thus, the angle of the wave propagating through the tropospheric layers is (see Fig. 6.14) roughly

$$\theta = \tan^{-1} \frac{10 \text{ mi}}{100 \text{ mi}} = 5.7°$$

It indicates that no strong power should be transmitted upward by 5° or more in order to avoid long-distance propagation.

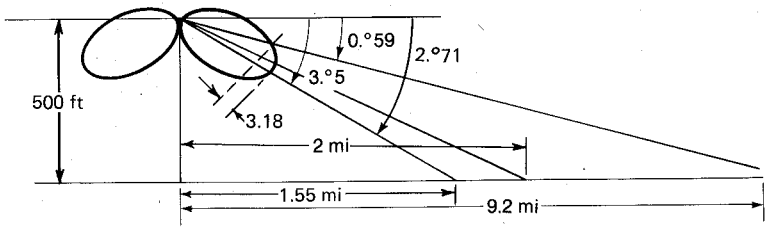

Figure 6.14 Coverage with the tilted-beam pattern.

6.8.2 Benefit of the umbrella pattern

The umbrella pattern, in which energy is confined to the immediate area of the antenna, is effective in reducing both cochannel and long-distance interference. Also, in hilly terrain areas there are many holes (weak signal spots). With a normal antenna pattern, we cannot raise the antenna high enough to cover these holes and decrease cochannel interference at the same time. However, the advantage of the umbrella pattern is that we can increase the antenna height and still decrease cochannel interference.

The frequency-reuse distance can be shortened by use of the umbrella pattern. To demonstrate this fact, we first calculate the two angles, one from the cell-site antenna to the cell boundary and the other from the cell-site antenna to the cochannel cell (the two angles are shown in Fig. 6.14).

Antenna height, ft	Desired maximum beam angle at boundary of a 2-mi cell, degrees	Angle toward cochannel cell at a distance of 4.6R (9.2 mi), degrees
100	0.54	0.12
300	1.63	0.35
500	2.71	0.59

Suppose that we are using an umbrella-pattern antenna with 11-dB gain,* and that the half-power beamwidth is above 5°. A tower of 500 ft is also used to cover a 2-mi cell. Then an approximate 3-dB difference due to the antenna pattern shown in Fig. 6.14 is obtained between the

* Normally antenna gain is measured with respect to a dipole.

area at the maximum beam angle and the area at the angle reaching the cochannel cell.

$$\frac{R^{-4}}{6D^{-4}} = 18 \text{ dB} - 3 \text{ dB} = 15 \text{ dB}$$

$$q^4 = 6 \times 31.6 = 189.74$$

$$q = 3.7$$

where beam strengths in two regions are different by 3 dB. This demonstrates that the required frequency-reuse distance can be reduced. In other words, more protection against cochannel interference is possible with the use of an umbrella pattern than with an omnidirectional beam pattern.

6.9 Use of Parasitic Elements

Interference at the cell site can sometimes be reduced by using parasitic elements, creating a desired pattern in a certain direction. In such instances, the currents appearing in several parasitic antennas are caused by radiation from a nearby drive antenna. A driven antenna and a single parasite can be combined in several ways.

1. *Normal spacing.*[6] We may first generate two separate patterns as shown in Fig. 6.15*a* and *b*. A single parasite spaced approximately one-quarter wavelength from the driven element is shown in Fig. 6.15*a*. Because the current flowing in the parasite is much weaker than that in the driven antenna, the front-to-back ratio is usually high. The two parasites spaced one-half wavelength

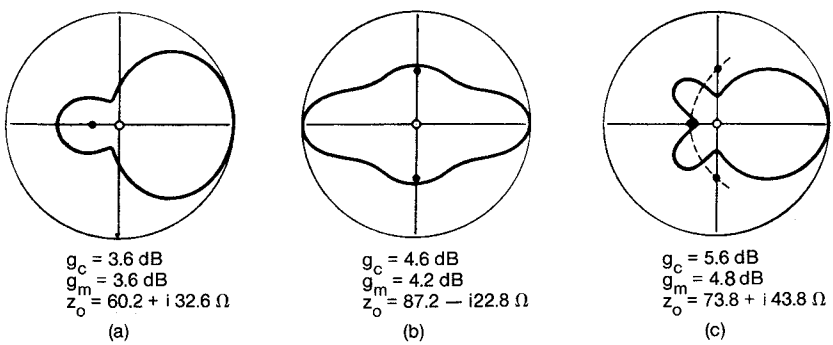

Figure 6.15 Parasitic elements with effective interference reduction. (*a*) One-quarter wavelength spacing; (*b*) one-half wavelength spacing; (*c*) combination of *a* and *b*. (*Reprint after Jasik, Ref. 6.*)

from the driven element are shown in Fig. 6.15*b*. A combination of Fig. 6.15*a* and *b* forming a pattern very similar to that of a parabola dish is shown in Fig. 6.15*c*. This is an effective arrangement for cell-site directional antennas with a non-wind-resistant structure: a four-element structure that has only one active element.

2. *Relatively close spacing.*[7] In relatively close spacing two elements are placed as close as 0.04λ. Three cases can be described here.
   a. The lengths of two elements are identical. Two elements, one active and one parasitic, are separated by only 0.04λ. At this

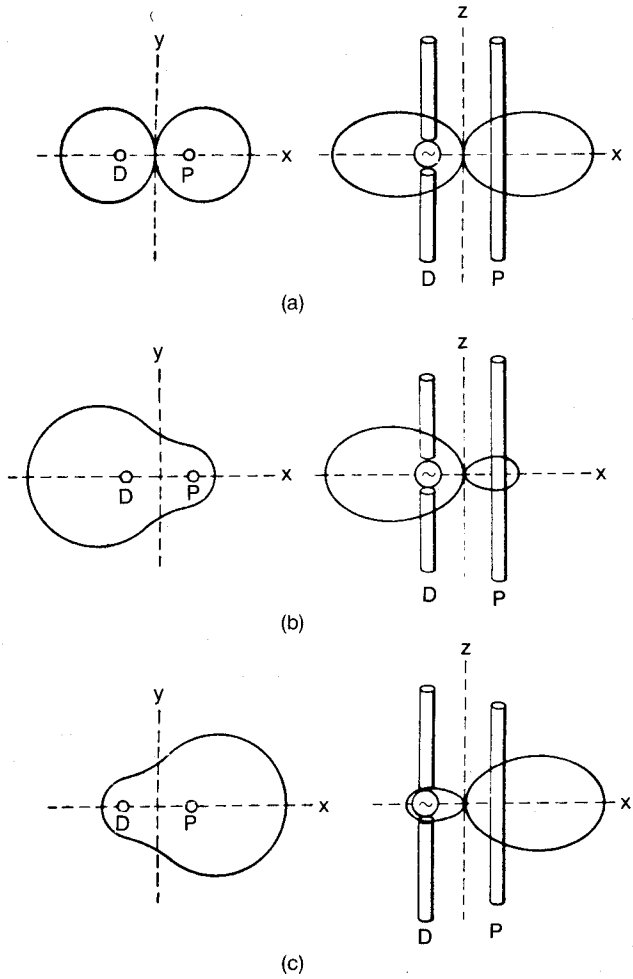

Figure 6.16 A close-in parasitic element with effective interference reduction (D = length of dipole; P = length of parasite). (*a*) $D = P$; (*b*) $D < P$; (*c*) $D > P$. (*Reprint after Jasik, Ref. 6.*)

close spacing, the current flowing in the parasite is very strong. The two elements form a null along the y axis in the horizontal plane and along the z axis in the vertical plane. There is a directive gain of 3 dB relative to a single element. The horizontal pattern and the vertical pattern of the closely spaced arrangement are shown in Fig. 6.16*a*.

b. *The length of the parasite is 5 percent longer than that of the active one.* In this case, the parasite acts as a reflector. The patterns are shown in Fig. 6.16*b* in both the horizontal and vertical planes. A directive gain of 6 dB is obtained.

c. *The length of the parasite is shorter than that of the active one.* In this case, the parasite acts as a director. The patterns are shown in Fig. 6.16*c* in the vertical and horizontal planes. A gain of 8 dB is obtained.

The pattern shown in Fig. 6.16*a* can be used for eliminating the interference to or from a given direction as shown by the null in the y axis. Two elements can be set up such that the y axis is aligned with the direction of interference.

Besides, we should emphasize that a directive antenna can be structured by a single parasite with a single active element (Fig. 6.16*b* and *c*). Therefore, a corner reflector or a ground reflector is not needed.

6.10 Power Control

6.10.1 Who controls the power level

The power level can be controlled only by the mobile transmitting switching office (MTSO), not by the mobile units, and there can be only limited power control by the cell sites as a result of system limitations.

The reasons are as follows. The mobile transmitted power level assignment must be controlled by the MTSO or the cell site, not the mobile unit. Or, alternatively, the mobile unit can lower the power level but cannot arbitrarily increase it. This is because the MTSO is cabable of monitoring the performance of the whole system and can increase or decrease the transmitted power level of those mobile units to render optimum performance. The MTSO will not optimize performance for any particular mobile unit unless a special arrangement is made.

6.10.2 Function of the MTSO

The MTSO controls the transmitted power levels at both the cell sites and the mobile units. The advantages of having the MSTO control the power levels are described in this section.

1. Control of the mobile transmitted power level. When the mobile unit is approaching the cell site, the mobile unit power level should be reduced for the following reasons.
   a. Reducing the chance of generating intermodulation products from a saturated receiving amplifier. This point is discussed in Chap. 7.
   b. Lowering the power level is equivalent to reducing the chance of interfering with other cochannel cell sites.
   c. Reducing the near-end–far-end interference ratio (see Sec. 7.3.1).

   Reducing the power level if possible is always the best strategy.
2. Control of the cell-site transmitted power level. When the signal received from the mobile unit at the cell site is very strong, the MTSO should reduce the transmitted power level of that particular radio at the cell site and, at the same time, lower the transmitted power level at the mobile unit. The advantages are as follows.
   a. For a particular radio channel, the cell size decreases significantly, the cochannel reuse distance increases, and the cochannel interference reduces further. In other words, cell size and cochannel interference are inversely proportional to cochannel reuse distance.
   b. The adjacent channel interference in the system is also reduced.

   However, in most cellular systems, it is not possible to reduce only one or a few channel power levels at the cell site because of the design limitation of the combiner. The channel isolation in the combiner is 18 dB. If the transmitted power level of one channel is lower, the channels having high transmitted power levels will interfere with this low-power channel. (The channel combiner is described in Chap. 7.) The manufacturer should design an unequal-power combiner for the system operator so that the power level of each channel can be controlled at the cell site.
3. The power transmitted from a small cell is always reduced, and so is that from a mobile unit. The MTSO can facilitate adjustment of the transmitted power of the mobile units as soon as they enter the cell boundary.

6.11 Diversity Receiver

The diversity scheme applied at the receiving end of the antenna is an effective technique for reducing interference because any measures taken at the receiving end to improve signal performance will not cause additional interference.

The diversity scheme is one of these approaches. We may use a selective combiner to combine two correlated signals as shown in Fig. 6.17. The performance of other kinds of combiners can be at most 2 dB better than that of selective combiners. However, the selective combining technique is the easiest scheme to use.[8]

Figure 6.17 shows a family of curves representing this selective combination. Each curve has an associated correlation coefficient ρ; when using the diversity scheme, the optimum result is obtained when $\rho = 0$.

We have found that at the cell site the correlation coefficient $\rho \leq 0.7$ should be used[9] for a two-branch space diversity; with this coefficient the separation of two antennas at the cell site meets the requirement of $h/d = 11$, where h is the antenna height and d is the antenna separation.

At the mobile unit we can use $\rho = 0$, which implies that the two roof-mounted antennas of the mobile unit are 0.5λ or more apart. This is verified by the measured data shown in Fig. 6.18.[10]

Now we may estimate the advantage of using diversity. First, let us assume a threshold level of 10 dB below the average power level. Then we compare the percent of signal below the threshold level both with and without a diversity scheme.

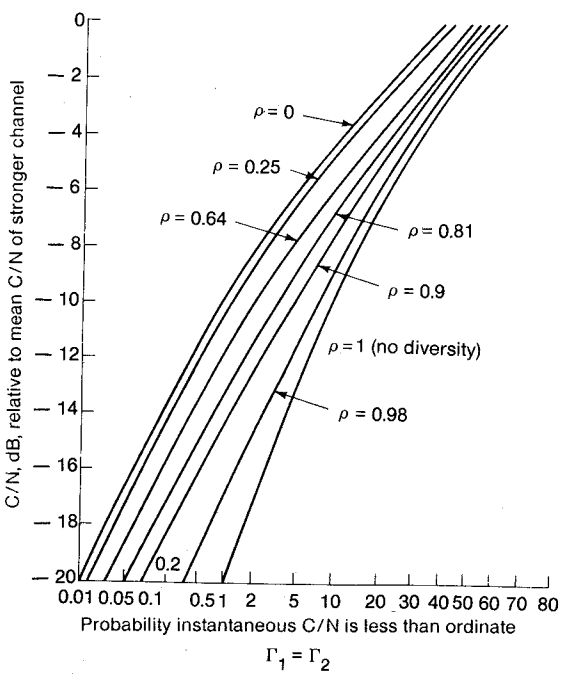

Figure 6.17 Selective combining of two correlated signals.

1. *At the mobile unit.* The comparison is between curves $\rho = 0$ and the $\rho = 1$. The signal below the threshold level is 10 percent for no diversity and 1 percent for diversity. If the signal without diversity were 1 percent below the threshold, the power would be increased by 10 dB (see Fig. 6.17). In other words, if the diversity scheme is used, the power can be reduced by 10 dB and the same performance can be obtained as in the nondiversity scheme. With 10 dB less power transmitted at the cell site, cochannel interference can be drastically reduced.

2. *At the cell site.* The comparison is between curves of $\rho = 0.7$ and $\rho = 1$. We use curve $\rho = 0.64$ for a close approximation as shown in Fig. 6.17. The difference is 10 percent of the signal is below threshold level when a nondiversity scheme is used versus 2 percent signal below threshold level when a diversity scheme is used. If the nondiversity signal were 2 percent below the threshold, the power would have to increase by 7 dB (see Fig. 6.17). Therefore, the mobile transmitter (for a cell-site diversity receiver) could undergo a 7-dB reduction in power and attain the same performance as a nondiversity receiver at the cell site. Thus, interference from the mobile transmitters to the cell-site receivers can be drastically reduced.

6.12 Designing a System to Serve a Predefined Area that Experiences Cochannel Interference

A system for a service area without cochannel interference can be designed by using the propagation prediction model described in Chap. 4. When cochannel interference does exist, the service in the area will deteriorate to one degree or another, depending on the location of the interference (or interferers). First, let us assume that the ground is flat; then two theoretical equations for designing a system in a given service area can be derived for two different interference cases. Then

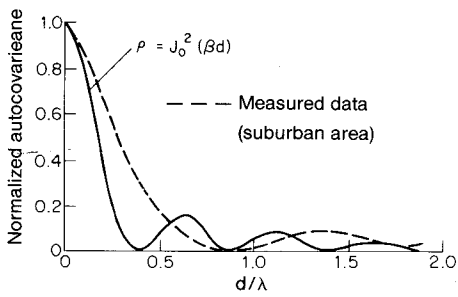

Figure 6.18 Autocorrelation coefficient versus spacing for uniform angular distribution (applied to diversity receiver). (*Reprint after Lee, Ref. 10.*)

the same approach can be used to design the systems for a service area where the ground is not flat.

Flat ground

One-interferer case. An interferer (cochannel site) is a distance d away from the serving cell site, and the mobile unit is traveling along the boundary of the serving-cell coverage. When the interferer is inactive, the coverage boundary is at a distance R from the serving site. When the interferer becomes active, the distance from the serving site to the effective coverage boundary is r, which can be less than R if the interference is either strong or close to the serving site or both. If we use polar coordinates, the serving site is located at (0, 0) with a transmitted power P_0, and the interferer is located at (d, θ_1) with a transmitted power P_1.

The mobile unit is located at $(r, 0)$, where r can be equal to or less than R. Assume the carrier-to-interference ratio requirement is C/I, then

$$\frac{C}{I} \leq \frac{P_0 r^{-4}}{P_1[\sqrt{r^2 + d^2 - 2rd\cos(\theta - \theta_1)}]^{-4}} \qquad r \leq R \qquad (6.12\text{-}1)$$

Let $\theta_1 = 0$ without loss of generality, as shown in Fig. 6.19 as "inter-

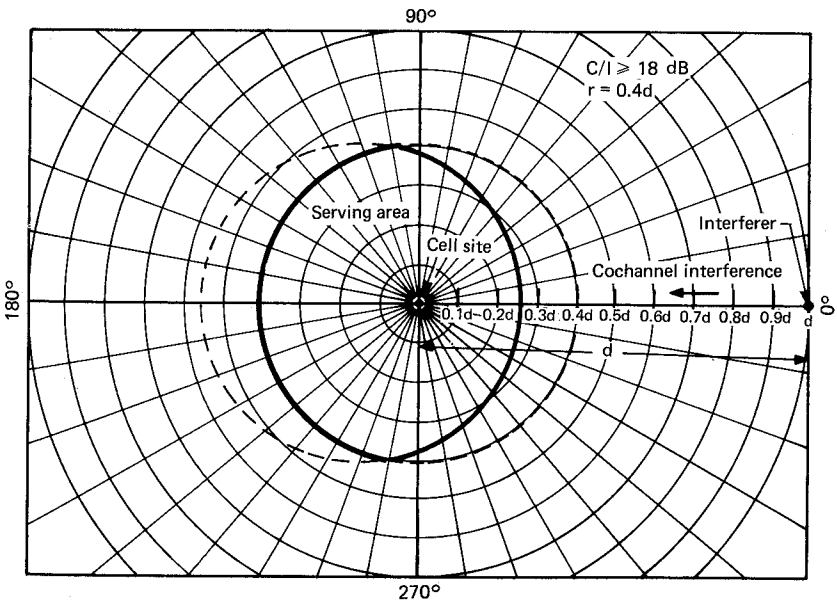

Figure 6.19 Serving area under cochannel interference.

ferer," and $P_0 = P_1$, then

$$\frac{C}{I} \leq \left(1 + \frac{d^2}{r^2} - \frac{2d}{r}\cos\theta\right)^2 \tag{6.12-2}$$

When $r = R$, there is no interference. Then Eq. (6.12-2) becomes

$$\frac{C}{I} \leq \left(1 + \frac{d^2}{R^2} - \frac{2d}{R}\cos\theta\right)^2 \tag{6.12-3}$$

Let $\theta = 0$, the strongest interference condition and $C/I = 18$ dB; then Eq. (6.12-3) becomes

$$63 \leq \left(1 - \frac{d}{R}\right)^4$$

or

$$d \geq 3.82R \tag{6.12-4}$$

When $r < d/3.82$, the serving area starts to decrease. Equation (6.12-2) can be converted to rectangular coordinates.

$$\frac{C}{I} \leq \left(1 + \frac{d^2}{x^2 + y^2} - \frac{2dx}{x^2 + y^2}\right)^2$$

or

$$\left(x + \frac{d}{\sqrt{C/I} - 1}\right)^2 + y^2 = d^2\left(\frac{1}{\sqrt{C/I} - 1} + \frac{1}{(\sqrt{C/I} - 1)^2}\right) \tag{6.12-5}$$

where $x^2 + y^2 \leq R^2$, the serving area has its center at $(-d/\sqrt{C/I} - 1, 0)$, and the radius is $d(C/I)^{1/4}/(\sqrt{C/I} - 1)$. An illustration of Eq. (6.12-5) is shown in Fig. 6.12 as the boundary area.

Multiple-interference case. For K_I interferers, Eq. (6.12-1) can be modified as

$$\frac{C}{I} = \frac{P_0 r^{-4}}{\sum_{i=1}^{K_I} P_i \left[\sqrt{r^2 + d^2 - 2rd\cos(\theta - \theta_1)}\right]^{-4}}$$

$$K_I \leq 6,\ r \leq R \tag{6.12-6}$$

Nonflat ground

The same approach is applied to the propagation model described in Chap. 4 in a real environment for predicting the serving areas.

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Chapter

7

Types of Noncochannel Interference

7.1 Subjective Test versus Objective Test

Voice quality often cannot be measured by objective testing using parameters such as the carrier-to-noise ratio C/N, the carrier-to-interference ratio C/I, the baseband signal-to-noise ratio S/N, and the signal to noise and distortion ratio (SINAD). In a mobile radio environment, multipath fading plus variable vehicular speed are the major factors causing deterioration of voice quality.

Only the following methods can help to correct this imbalance.

1. Let the received carrier level be high to increase the signal level.
2. Let the receiver sensitivity be high to lower the noise level.
3. Maintain a low distortion level in the receiver to increase SINAD.
4. Use a diversity receiver to reduce the fading.
5. Use a good system design in a mobile radio environment and a good adjacent-channel rejection to reduce the interference.

However, when a transceiver is deployed in a mobile radio environment, a subjective test is still the only way to test this receiver, using

different types of modulation, such as single-sideband, double-sideband, amplitude, and frequency modulation (SSB, DSB, AM, FM).

7.1.1 The subjective test

A subjective test can be set up according to the criterion that 75 percent of the customers perceive the voice quality at a given C/N as being "good" or "excellent," the top two levels among the five circuit-merit (CM) grades.[1] The simulator of this test must be adjusted for different mobile speeds. The customers can hear different S/N levels at the baseband on the basis of the carrier-to-noise ratio C/N being changed at the RF transmitter. One typical set of curves from the customers' perception at a mobile speed at 25 km/h (or 16 mi/h) and one at 56 km/h (or 35 mi/h) are shown[2] in Fig. 7.1. Average all the test records for different vehicle speeds and determine a C/N which can satisfy the criterion we have established.

7.1.2 The objective test

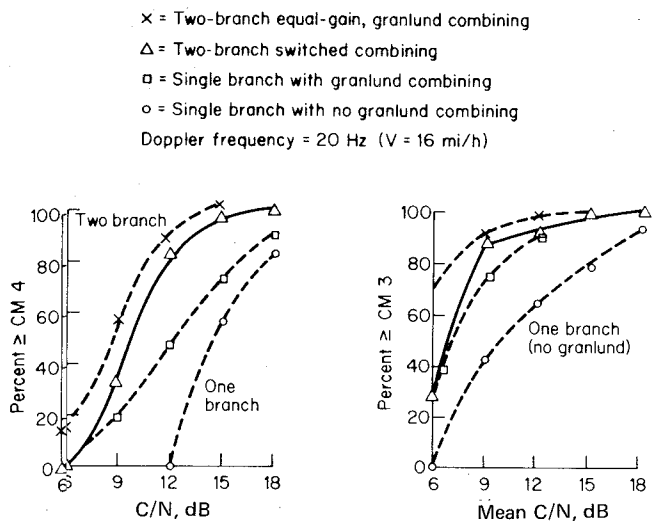
There are many objective tests at the baseband for both voice and data. The characterization of voice quality is very difficult, as mentioned previously, but evaluation of data transmission is easy. There are two major terms, bit-error rates and word error rates. The bit-error rate (BER) is the first-order statistic (independent of time or vehicle speed), and the word-error rate (WER) is the second-order statistic which is affected by the vehicle speed. These rates are discussed in Chap. 12.

7.1.3 Measurement of SINAD

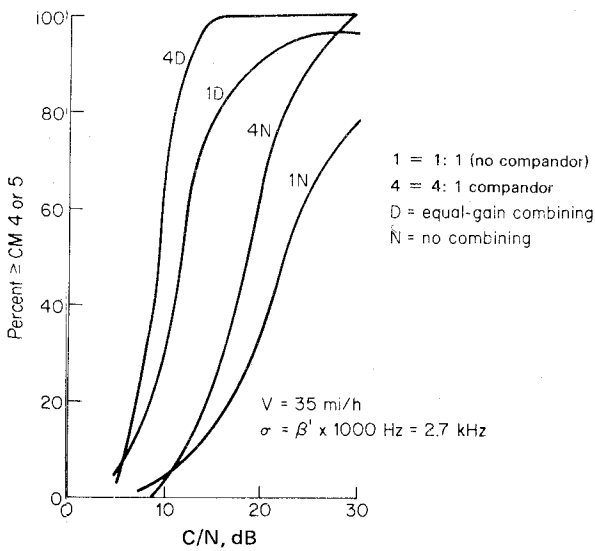
SINAD has been used as a measurement of communication signal quality at the baseband or in the cellular mobile receiver to measure the effective FM receiver sensitivity.[3] Some telephone industries use a "notched noise" measurement, in which a 1000-Hz tone is sent down the telephone line. The line noise is added onto the tone when it is received. By notching out the tone frequency, we can determine the remaining noise. This is a type of SINAD measurement.

1. The SINAD of the baseband output signal is defined as the ratio of the total output power to the power of the noise plus distortion only.

$$\text{SINAD} = \frac{\text{total output power}}{\text{nonsignal portion}} = \frac{\text{signal + noise + distortion}}{\text{noise + distortion}} \tag{7.1-1}$$



(a)



(b)

Figure 7.1 Results from subjective tests. (*Reprinted from W. C. Y. Lee, Mobile Communications Engineering, McGraw-Hill Book Co., 1982, pp. 428–429.*) (*a*) System-versus-performance comparison based on circuit merit CM4 vs. CM3. (*b*) System-versus-performance comparisons based on circuit merit CM4 and CM5.

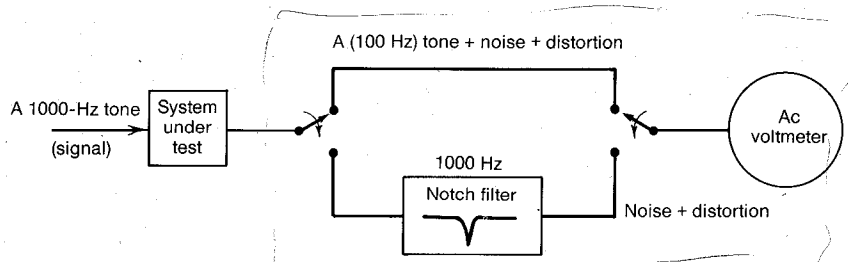


Figure 7.2 A SINAD meter.

The output power can be obtained by measuring the output from a voltmeter and then squaring the voltage, or directly from a power meter. In cellular radio equipment, an input of -116 dBm is equivalent to a SINAD of 12 dB.

2. A high signal level can be measured by

$$\text{SINAD} = \frac{\text{signal} + \text{noise}}{\text{noise}} \approx \frac{\text{signal}}{\text{noise}}$$

The SINAD shown in Fig. 7.2 can be obtained by measuring the signal at the upper position and measuring the noise reading received at the lower position, assuming that the distortion is insignificant.

3. Receiver sensitivity can be measured by modulating with a 1-kHz tone at 3-kHz peak modulation deviation as shown in Fig. 7.3. The signal-generated attenuator should be adjusted until the SINAD meter shows 12 dB. Then the microvolt output is read from the attenuator dial, which reveals the "12 dB" of SINAD "sensitivity" of the receiver. This means that the signal input must be of a certain level for the signal at the output to be 12 dB higher than noise plus distortion. If the receiver noise is higher, the minimum input signal level should also be higher in order to maintain the 12-dB SINAD.

4. Noise voltage can be measured from a c-message weighting filter on any kind of telephone circuit. The frequency response of this c-message weighting filter is based on the human voice. The noise measured at the output of the filter is the noise withholding in the speech frequency spectrum. Therefore telephone line performance is measured by the amount of noise voltage through the c-message-weight filter.

5. The SINAD meter also can be used as a distortion meter if the noise is very low in comparison to the distortion. The SINAD meter

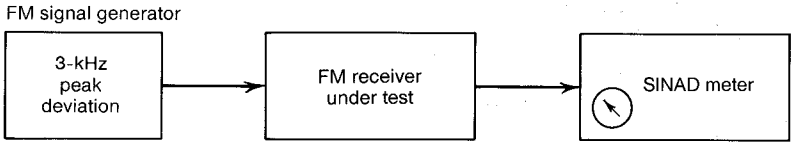


Figure 7.3 Measuring receiver sensitivity.

can be used to check the maximum distortion figures of the receiver. The input signal level is increased until no thermal noise can be heard; the receiver volume meter reads the audio power, and the SINAD meter reads the distortion.

7.2 Adjacent-Channel Interference

The scheme discussed in Chap. 6 for reduction of cochannel interference can be used to reduce adjacent-channel interference. However, the reverse argument is not valid here. In addition, adjacent-channel interference can be eliminated on the basis of the channel assignment, the filter characteristics, and the reduction of near-end–far-end (ratio) interference. "Adjacent-channel interference" is a broad term. It includes next-channel (the channel next to the operating channel) interference and neighboring-channel (more than one channel away from the operating channel) interference. Adjacent-channel interference can be reduced by the frequency assignment.

7.2.1 Next-channel interference

Next-channel interference affecting a particular mobile unit cannot be caused by transmitters in the common cell site, but must originate at several other cell sites. This is because any channel combiner at the cell site must combine the selected channels, normally 21 channels (630 kHz) away, or at least 8 or 10 channels away from the desired one. Therefore, next-channel interference will arrive at the mobile unit from other cell sites if the system is not designed properly. Also, a mobile unit initiating a call on a control channel in a cell may cause interference with the next control channel at another cell site. The methods for reducing this next-channel interference use the receiving end. The channel filter characteristics[4] are a 6 dB/oct slope in the voice band and a 24 dB/oct falloff outside the voice-band region (see Fig. 7.4). If the next-channel signal is stronger than 24 dB, it will interfere with the desired signal. The filter with a sharp falloff slope can help to reduce all the adjacent-channel interference, including the next-channel interference.

7.2.2 Neighboring-channel interference

The channels which are several channels away from the next channel will cause interference with the desired signal. Usually, a fixed set of serving channels is assigned to each cell site. If all the channels are simultaneously transmitted at one cell-site antenna, a sufficient amount of band isolation between channels is required for the multi-

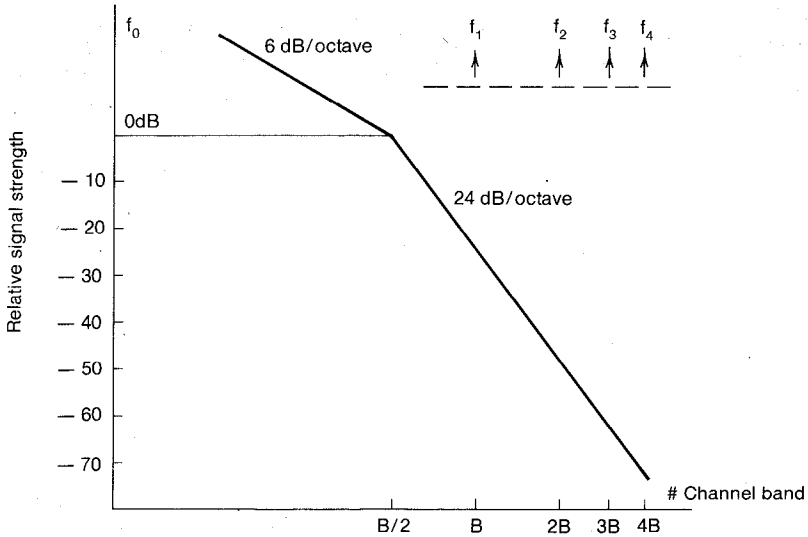


Figure 7.4 Characteristics of channel-band filter.

channel combiner (see Sec. 7.7.1) to reduce intermodulation products. This requirement is no different from other nonmobile radio systems. Assume that band separation requirements can be resolved, for example, by using multiple antennas instead of one antenna at the cell site. What channel separation would be needed to avoid adjacent-channel interference? (See Sec. 8.1.)

Another type of adjacent-channel interference is unique to the mobile radio system. In the mobile radio system, most mobile units are in motion simultaneously. Their relative positions change from time to time. In principle, the optimum channel assignments that avoid adjacent-channel interference must also change from time to time. One unique situation that causes adjacent-channel interference in mobile radio systems is described in the next section.

7.3 Near-End–Far-End Interference

7.3.1 In one cell

Because motor vehicles in a given cell are usually moving, some mobile units are close to the cell site and some are not. The close-in mobile unit has a strong signal which causes adjacent-channel interference (see Fig. 7.5*a*). In this situation, near-end–far-end interference can occur only at the reception point in the cell site.

If a separation of $5B$ (five channel bandwidths) is needed for two adjacent channels in a cell in order to avoid the near-end–far-end

interference, it is then implied that a minimum separation of $5B$ is required between each adjacent channel used with one cell.

Because the total frequency channels are distributed in a set of N cells, each cell only has $1/N$ of the total frequency channels. We denote $\{F_1\}$, $\{F_2\}$, $\{F_3\}$, $\{F_4\}$ for the sets of frequency channels assigned in their corresponding cells C_1, C_2, C_3, C_4.

The issue here is how can we construct a good frequency management chart to assign the N sets of frequency channels properly and thus avoid the problems indicated above. The following section addresses how cellular system engineers solve this problem in two different systems.

7.3.2 In cells of two systems

Adjacent-channel interference can occur between two systems in a duopoly-market system. In this situation, adjacent-channel interference can occur at both the cell site and the mobile unit.

For instance, mobile unit A can be located at the boundary of its own home cell A in system A but very close to cell B of system B as shown in Fig. 7.5*b*. The other situation would occur if mobile unit B were at the boundary of cell B of system B but very close to cell A of system A. Following the definition of near-end–far-end interference given in Sec. 7.3.1, the solid arrow indicates that interference may occur at cell site A and the dotted arrow indicates that interference may occur at mobile unit A. Of course, the same interference will be introduced at cell site B and mobile unit B.

Thus, the frequency channels of both cells of the two systems must be coordinated in the neighborhood of the two-system frequency bands. This phenomenon will be of greater concern in the future, as indicated in the additional frequency-spectrum allocation charts in Fig. 7.6.

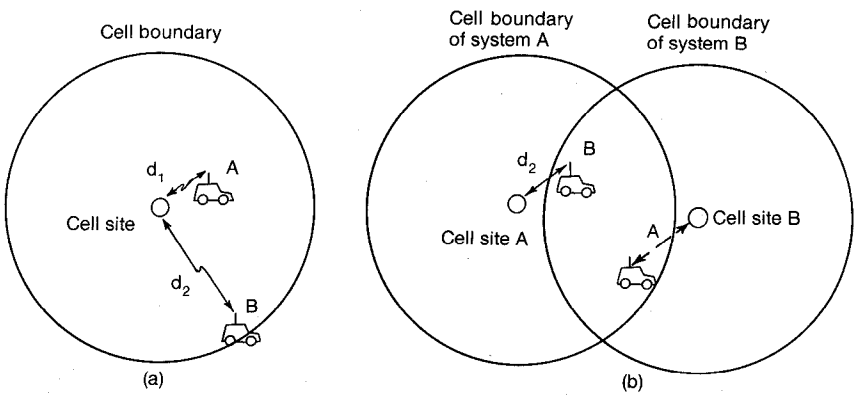


Figure 7.5 Near-end–far-end (ratio) interference. (*a*) In one cell; (*b*) in two-system cells.

The two causes of near-end–far-end interference of concern here are

1. *Interference caused on the set-up channels.* Two systems try to avoid using the neighborhood of the set-up channels as shown in Fig. 7.6.
2. *Interference caused on the voice channels.* There are two clusters of frequency sets as shown in Fig. 7.6 which may cause adjacent-channel interference and should be avoided. The cluster can consist of 4 to 5 channels on each side of each system, that is, 8 to 10 channels in each cluster. The channel separation can be based on two assumptions.
   a. *Received interference at the mobile unit.* The mobile unit is located away from its own cell site but only 0.25 mi away from the cell site of another system.
   b. *Received interference at the cell site.* The cell site is located 10 mi away from its own mobile unit but only 0.25 mi from the mobile unit of another system.

These assumptions are discussed in the next section. If the two system operators do not agree to coordinate their use of frequency channels and some of the cell sites of system B are at the coverage boundaries of the cells of system A, then the two groups of frequencies shown in Fig. 7.6 must not be used if interference has to be avoided. Of course,

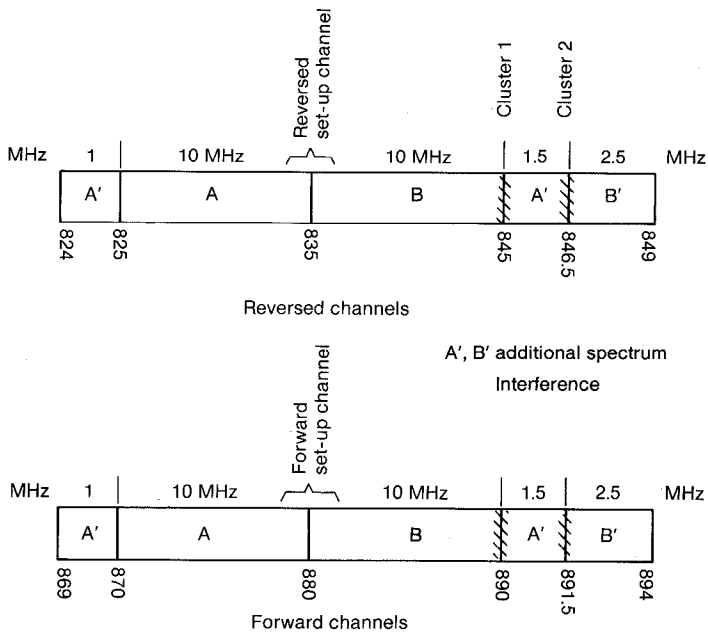


Figure 7.6 Spectrum allocation with new additional spectrum.

if the two systems do coordinate their use of frequency channels, adjacent channels in the two clusters can be used with no interference.

These observations regarding adjacent-channel interference lead the author to conclude that the existence of two systems having all colocation cell sites in a city is desirable since near-end–far-end ratio interference might be easy to control or might not occur if frequency channel use is coordinated.

7.4 Effect on Near-End Mobile Units

7.4.1 Avoidance of near-end–far-end interference

The near-end mobile units are the mobile units which are located very close to the cell site. These mobile units transmit with the same power as the mobile units which are far away from the cell site. The situation described below is illustrated in Fig. 7.7. The distance d_0 between a calling mobile transmitter and a base-station receiver is much larger than the distance d_I between a mobile transmitter causing interference and the same base-station receiver. Therefore, the transmitter of the mobile unit causing interference is close enough to override the desired base-station signal.[5] This interference, which is based on the distance ratio, can be expressed as

$$\frac{C}{I} = \left(\frac{d_0}{d_I}\right)^{-\gamma} \tag{7.4-1}$$

where γ is the path-loss slope. The ratio d_I/d_0 is the near-end–far-end ratio. From Eq. (7.4-1) the effect of the near-end–far-end ratio on the carrier–adjacent-channel interference ratio is dependent on the relative positions of the moving mobile units.

For example, if the calling mobile unit is 10 mi away from the base-station receiver and the mobile unit causing the interference is 0.25 mi away from the base-station receiver, then the carrier-to-interference

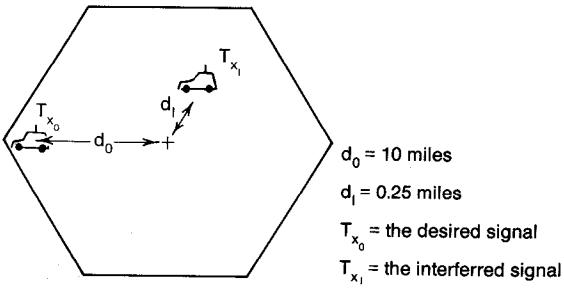


Figure 7.7 Near-end–far-end ratio interference.

ratio for interference received at the base-station receiver with $\gamma = 4$ is

$$\frac{C}{I} = \left(\frac{d_0}{d_I}\right)^{-4} = (40)^{-4} = -64 \text{ dB} \tag{7.4-2}$$

This means that the interference is stronger than the desired signal by 64 dB (see Fig. 7.8).

This kind of interference can be reduced only by frequency separation with narrow filter characteristics. Assume that a filter of channel B has a 24 dB/oct slope;[4] then a 24-dB loss begins at the edge of the channel $B/2$. The increase from $B/2$ to B results in 24-dB loss, the increase from B to $2B$ results in another 24-dB loss, and so forth.

In order to achieve a loss of 64 dB, we may have to double the frequency band more than two times as

$$\frac{64}{L} = \frac{64}{24} = 2.67$$

where L is the filter characteristic. The frequency band separation for 64-dB isolation is

$$2^{-(C/I)/L}\left(\frac{B}{2}\right) = 2^{2.67}\left(\frac{B}{2}\right) = 3.18B \tag{7.4-3}$$

Therefore, a minimum separation of four channels is needed to satisfy the isolation criterion of 64 dB. The general formula for the required

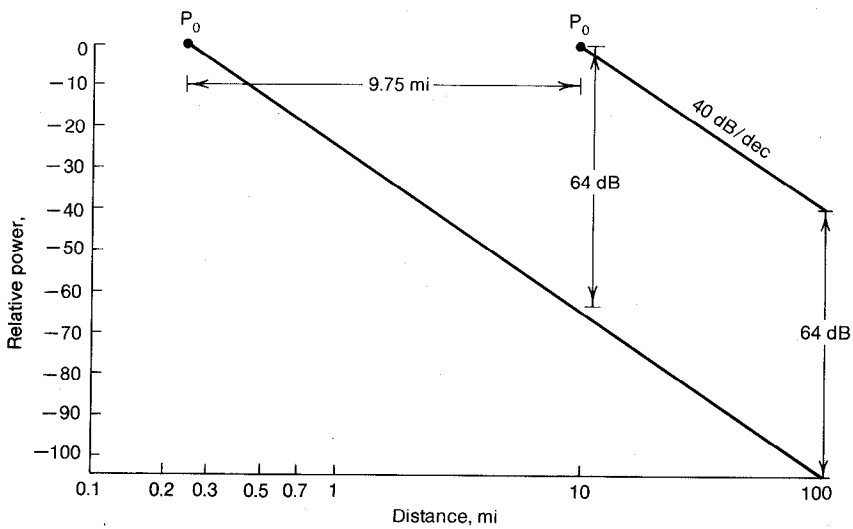


Figure 7.8 Using spacing for cochannel isolation.

channel separation is based on the filter characteristic L, which is expressed as follows.[5]

$$\text{Frequency band separation} = 2^{G-1}B \tag{7.4-4}$$

where

$$G = \frac{\gamma \log_{10}\left(\dfrac{d_0}{d_I}\right)}{L} \tag{7.4-5}$$

7.4.2 Nonlinear amplification

When the near-end mobile unit is close to the cell site, its transmitted power is too strong and saturates the IF log amplifier if the received signal at the cell site exceeds −55 dBm. A typical log IF amplifier characteristic is shown in Fig. 7.9. Assume that the mobile unit transmitted power is 36 dBm and the antenna gain is 2 dBi. The power plus the gain is 38 dBm. The receiver power is −55 dBm at the cell site.

The propagation loss L = 38 dBm − (−55 dBm) = 93 dB. We may calculate the free-space path loss, which is the maximum distance within which the saturation of the IF amplifier will occur. The calculation of free-space loss versus distance at 850 MHz is as follows.

$$\begin{aligned}
-55 \text{ dBm} &= 10 \log \frac{P}{(4\pi)^2(d/\lambda)^2} \\
&= 38 \text{ dBm} - 20 \log 4\pi - 20 \log\left(\frac{d}{\lambda}\right) \\
20 \log_{10}\left(\frac{d}{\lambda}\right) &= 55 + 38 - 22 = 71 \\
\frac{d}{\lambda} &= 10^{71/20} = 3548 \\
d &= 3548\lambda = 4115 \text{ ft} \\
&= 1241 \text{ m} = 1.24 \text{ km}
\end{aligned} \tag{7.4-6}$$

This means that when the mobile unit is within 1.24 km of the cell-site boundary, it is possible to saturate the IF amplifier, and it is likely that intermodulation will be generated because of the nonlinear portion of the characteristics. If the intermodulation (IM) product matches the frequency channel of another mobile unit far away from the cell site where reception is weak, then the IM can interfere with the other frequency received at the cell site.

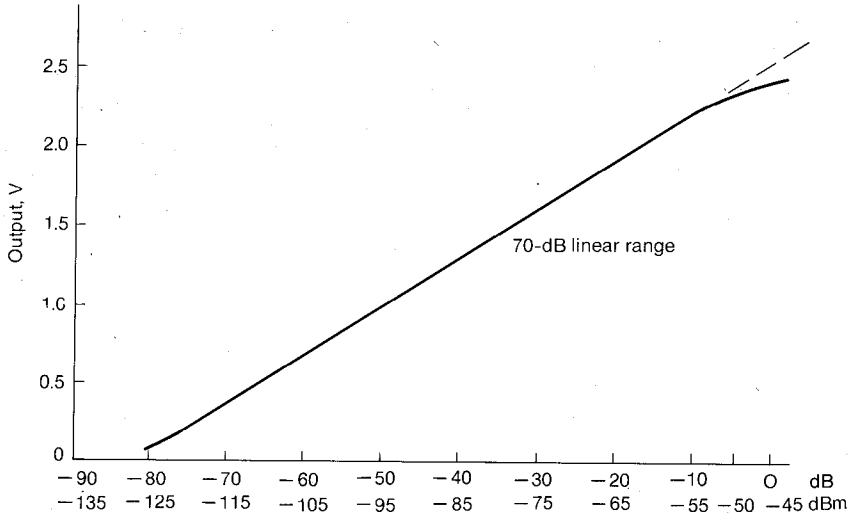


Figure 7.9 A typical intermediate-frequency log amplifier.

Therefore, the near-end mobile unit can cause interference at the cell site with the far-end mobile unit by generating IM at the cell-site amplifier and by leaking into the signal of the far-end mobile unit received at the cell site.

7.5 Cross Talk—a Unique Characteristic of Voice Channels

When the cellular radio system was designed, the system was intended to function like a telephone wire line. A wire pair serves both directions of traffic at the line transmission. In a mobile cellular system there is a pair of frequencies, occupying a bandwidth of 60 kHz, which we simply call a "channel." A frequency of 30 kHz serves a received path, and the other 30 kHz accommodates a transmitted path.

Because of paired-frequency (as a wire pair) coupling through the two-wire–four-wire hybrid circuitry at the telephone central office, it is possible to hear voices in both frequencies (in the frequency pair) simultaneously while scanning on only one frequency in the air. Therefore, just as with a wire telephone line, the full conversation can be heard on a single frequency (either one of the two). This phenomenon does not annoy cellular mobile users; when they talk they also listen to themselves through the phone receiver. They are not even aware that they are listening to their own voices.

This unnoticeable cross-talk phenomenon in frequency pairs has no major impact on both wire telephone line and cellular mobile perform-

ance. But when real cross talk occurs it has a larger impact on the cellular mobile system than on the telephone line, because the amount of cross talk could potentially be doubled since cross talk occurring on one frequency will be heard on the other (paired) frequency. Cross talk occurring on the reverse voice channel can be heard on the forward voice channel, and cross talk occurring on the forward voice channel can be heard on the reverse channel. Therefore, the cross-talk effect is twofold. A number of situations are conducive to cross talk.

Near-end mobile unit. Cross talk can occur when one mobile unit (unit A) is very close to the cell site and the other (unit B) is far from the cell site. Both units are calling to their land-line parties as shown in Fig. 7.10. The near-end mobile unit has a strong signal such that the demultiplexer cannot have an isolation (separation) of more than 30 dB. Then the strong signal can generate strong cross talk while the received signal from mobile unit B is 30 dB weaker than signal A.

Near-end mobile units can belong to one system or to another (foreign) system. If the foreign system units are operating in the new allocated spectrum channels, cross talk can occur. When the mobile unit is close to the cell site and the cell site is capable of reducing the power of the mobile unit, the near-end mobile interference can be reduced.

If the operating frequencies of both home system units and foreign system units are in the new allocated spectrum channels and the isolation of the multicoupler (demultiplexer) could be only 30 dB, cross talk would occur in the two interfering clusters of channels (Fig. 7.6) and could not be controlled by the system operator.

Close-in mobile units. When a mobile unit is very close to the cell site and if the reception at the cell site is greater than −55 dBm, the

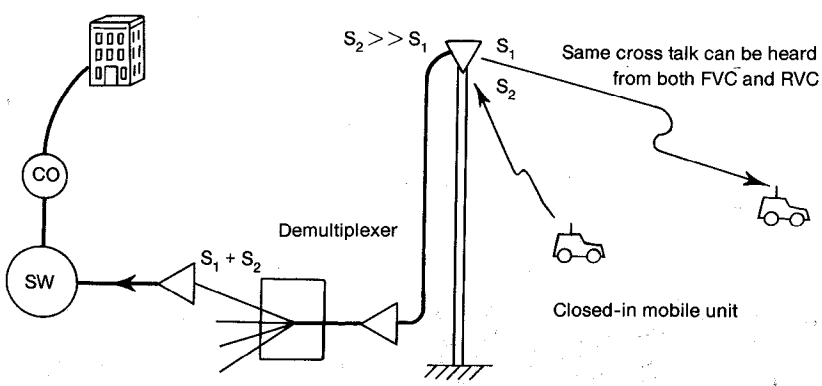


Figure 7.10 Cross-talk phenomenon.

channel preamplifier at the cell site can become saturated and produce IM as a result of the nonlinear portion of the amplification. These IM products are the spurious (unwanted frequency) signal which leaks into the desired signal and produces cross talk. Also, as mentioned previously, the same cross talk can be heard from both the forward and reverse voice channels.

Cochannel cross talk. The cochannel interference reduction ratio q should be as large as possible to compensate for the cost of site construction and the limitation of available channels at each cellular site. There are other ways to increase q, as mentioned in Chap. 6. An adequate system design will help to reduce the cochannel cross talk.

The channel combiner. The signal isolation among the forward voice channels in a channel combiner is 17 dB.[4] The loss resulting from inserting the signal into the combiner is about 3 dB. The requirement of IM product suppression is about 55 dB. If one outlet is not matched well, the signal isolation is less than 17 dB. Therefore, for each channel an isolator is installed to provide an additional 30-dB of isolation with a 0.5-dB insertion loss. This isolator prevents any signal from leaking back to the power amplifier (see Sec. 7.7.1). Spurious signals can be cross-coupled to this weak channel while transmitting. This kind of cross-coupled interference can be eliminated by routinely checking impedance matching at the combiner.

Telephone-line cross talk. Sometimes cross talk can result from cable imbalance or switching error at the central office and be conveyed to the customer through the telephone line. Minimizing this type of cross talk should be given the same priority as reducing the number of call drops, discussed earlier (Chap. 4 and Sec. 6.2).

7.6 Effects on Coverage and Interference by Applying Power Decrease, Antenna Height Decrease, Beam Tilting

Communications engineers sometimes encounter situations where coverage must be reduced to compensate for interference. There are several ways of doing this. Reorienting the directional-antenna patterns, changing the antenna beamwidth, or synthesizing the antenna pattern were discussed in Chap. 6. There are two additional methods, decreasing the power and decreasing the antenna height. Both methods are effective, and engineers often have difficulty choosing between them. Which one is better? The answer is dependent on the situation.

7.6.1 Choosing a proper cell site

Given a fixed transmitted power and a cell-site antenna height, the coverage contours of a cell site for different signal reception levels can be obtained from either the measurement or from the prediction model described in Chap. 4. A typical contour is shown in Fig. 7.11. Because of the irregular terrain contours, contours between different reception levels are not equally spaced.

When a cell site is selected, we must determine whether an ultra-high-frequency (UHF) TV station is nearby (see Sec. 7.9) and whether any future nearby ongoing construction would affect signal coverage from the cell site later. We must check the local noise level and be sure that no spurious signals fall in the cellular frequency band.

Finally, if we are using an existing multiantenna tower, we must ensure that the grounding and shielding are adequate. Otherwise the interference level could become very high and weaken cell-site operation. Sometimes a special isolator may be provided if an AM broadcasting antenna is colocated on the same tower.

7.6.2 Power decrease

As long as the setup of the antenna configuration at the cell site remains the same, and if the cell-site transmitted power is decreased by 3 dB, then the reception at the mobile unit is also decreased by 3 dB. This is a one-on-one (i.e., linear) correspondence and thus is easy to control.

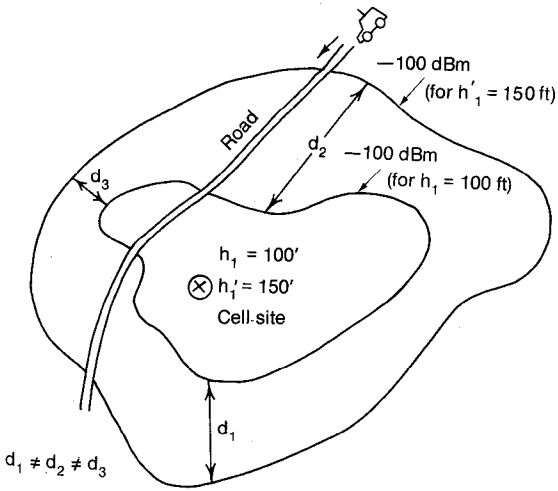


Figure 7.11 Signal-strength contour shape changing as the antenna height changes.

7.6.3 Antenna height decrease

When antenna height is decreased, the reception power is also decreased. However, the formula [see Eq. (4.10-2)]

$$\text{Antenna height gain (or loss)} = 20 \log \frac{h'_{e_1}}{h_{e_1}}$$

is based on the difference between the old and new effective antenna heights and not on the actual antenna heights. Therefore, the effective antenna height is the same as the actual antenna height only when the mobile unit is traveling on flat ground. It is easy to decrease antenna height to control coverage in a flat-terrain area. For decreasing antenna height in a hilly area, the signal-strength contour shown in Fig. 7.12*a* is different from the situation of power decrease shown in Fig. 7.12*b*. Therefore a decrease in antenna height would affect the coverage; thus antenna height become very difficult to control in an overall plan. Some area within the cell may have a high attenuation while another may not.

7.6.4 Antenna patterns

The design of different antenna patterns is discussed and illustrated in Chap. 5. Here we would like to emphasize that the design of the antenna pattern should be based on the terrain contour, the population and building density, and other conditions within a given area. Of course, this is often difficult to do. For instance, implementation of

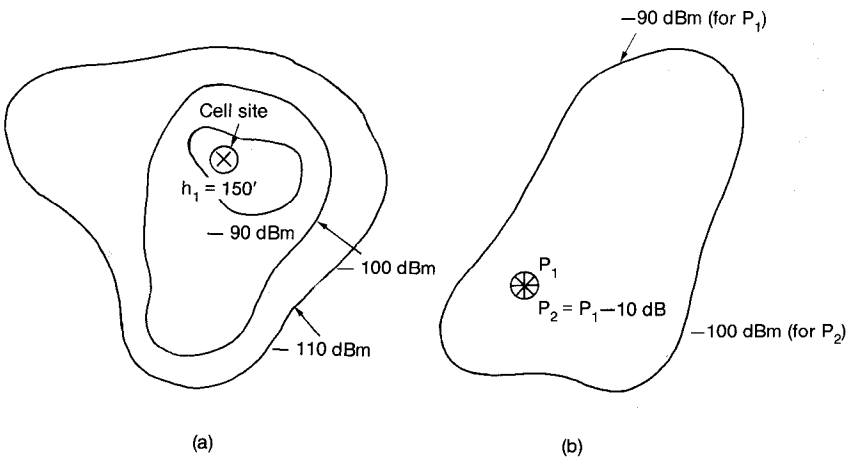


Figure 7.12 The signal-strength effect as measured by different parameters. (*a*) Different signal-strength contours. (*b*) Signal-strength changes with power changes.

antenna tilting or use of an umbrella pattern might be necessary in certain areas in order to reduce interference.

Sidelobe control (i.e., control of secondary lobe formation in an antenna radiation pattern) is also very critical in the implementation of a directional antenna. Coverage can be controlled by means of the following methods.

Using multiple antennas. In a multiple directional antenna pattern, the antennas can have different power outputs and each antenna can form a desired pattern. Two configurations can be mentioned.

1. All the antennas are facing outward (see Fig. 7.13*a*). The resultant pattern is always difficult to control because ripples and deep nulls frequently form.
2. With skewed directional antennas[6] (see Fig. 7.13*b*), the resultant pattern becomes smoother. Therefore, this configuration is more attractive.

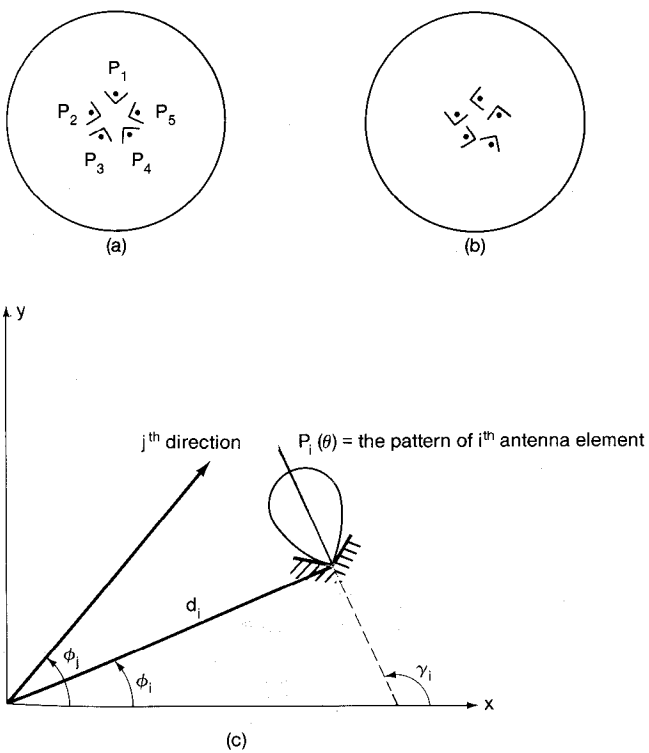


Figure 7.13 Engineering a desired pattern with directional antennas. (*a*) Five directional antennas facing outward; (*b*) a skewed configuration of five directional antennas; (*c*) the coordinate.

Using a synthesis of power pattern. The use of steepest descent techniques for searching the antenna parameters by giving an actual pattern and a desired pattern is introduced here. The signal strength contour obtained from Chap. 4 will be used. The difference between the two patterns, actual and desired, or error ϵ, can be expressed as

$$\epsilon(\phi, d, I, \alpha, \gamma) = \sum_{j=1}^{M} W_j(P_j - Q_j)^2 \tag{7.6-1}$$

The parameters ϕ, d, and γ are shown in Fig. 7.13*c*, where I_i and α_i are the amplitude and phase of ith element, respectively. P_j is the desired field strength at the jth direction, and Q_j is the given (measured) field strength at the jth direction. All cells may be divided into M small angles, and the jth direction is one of these angles. In Eq. (7.6-1), W_j is a weighting function. When a nonuniform pattern is to be synthesized $W_j \neq 1$. The steepest descent technique can be applied to find the five parameters associated with pattern P_j which will yield the minimum ϵ in Eq. (7.6-1).

If we are using L elements, then P_j in Eq. (7.6-1) is the desired radiation field strength.

$$P_j = \sum_{i=1}^{L} P_i(\phi_j - \gamma_i)I_i \times \exp\left\{-j\left[\frac{2\pi d_i}{\lambda}\cos(\phi_j - \phi_i) - \alpha_i\right]\right\} \tag{7.6-2}$$

where $P_i(\phi)$ is the individual pattern of ith element. The magnitude and phase of the ith-element excitation are I_i and α_i, respectively. The remaining variables of Eq. (7.6-2) are shown in Fig. 7.13*c*. Since ϵ is a function of five parameters as indicated in Eq. (7.6-1), we start with an initial guess for the parameters (ϕ_0, d_0, I_0, α_0, γ_0), and then apply the iterative equation

$$\beta_{n+1} = \beta_n - k_{\beta_i}\nabla_{\beta_i}\epsilon_n \tag{7.6-3}$$

where β = one of five parameters
$\nabla_{\beta_i}\epsilon_n$ = component of $\nabla\epsilon$ corresponding to the variable β evaluated at a given point, say, $\beta_n = \phi_n(\beta_n, d_n, I_n, \alpha_n, \gamma_n)$
k_{β_i} = gain constant for the parameter β_i

The value k_{β_i} cannot be small; otherwise the convergent process would be very slow. The iterative process is repeated until $n = N$ is reached,

that is, $\nabla_{\beta_i}\epsilon_N = 0$. Then from Eq. (7.6-3), $\beta_{n+1} = \beta_n = \beta_i$ for any one of five parameters for the ith antenna element.

The same procedures apply for all elements, and all calculations can be performed by computer.

Caution: Because the terrain is not flat, the signal strengths in all directions are not uniformly attenuated at equal distances; thus, we must first obtain an antenna pattern (not desired) corresponding to a cell boundary in the actual field from a set of predetermined parameters (assume that the current distributions of all antenna elements are the same) and then convert the undesired pattern through the use of an iteration process to a desirable pattern that can be used in the field. The propagation model described in Chap. 4 will serve this purpose. Thus we can apply this iterative process to practical problems.

7.6.5 Transmitting and receiving antennas at the cell site

At the base station, the transmitted power of 100 W (+50 dBm) plus an antenna gain of 9 dBi is assumed at one transmitting antenna. The receiving antenna, located at the same site, also has a gain of 9 dBi and receives a mobile signal of −100 dBm. The difference in signal strength is

$$(50 + 9 + 9)\ \text{dBm} - (-100\ \text{dBm}) = +168\ \text{dB}$$

If the space separation between a transmitting antenna and a receiving antenna is 15 m (50 ft) horizontally, the signal isolation obtained from the free-space formula is 56 dB.

The 45-MHz bandpass filter followed by the receiving antenna has at least a 55-dB rejection for signals arriving from the 870- to 890-MHz transmission band. However, the two numbers added together is 111 dB, which is still not sufficient (57 dB short). That is why the transmitting antenna and receiving antenna are not mounted in the same horizontal plane, but rather on the same vertical pole, if they are omnidirectional. This restriction can be moderated for directional antennas because of the directive patterns.

7.6.6 A 39-dBμ and a 32-dBμ boundary

The Federal Communications Commission (FCC) has used a specified received signal strength[9] for the coverage boundary, which is 39 dBμ (dB in μV/m). This value converts to a received power of −93 dBm for dipole or monopole matching on a 50-Ω load at 850 MHz (see Secs.

5.1.3 or 13.3.3). The value of 39 dBμ (i.e., −93 dBm) should be tested to determine if it is too high for use at the cell boundary in the cellular system.

We can calculate an acceptable level as follows. As we know, the accepted carrier-to-noise ratio for good quality (agreed on by most system operators) is 18 dB. The thermal noise level kTB with a bandwidth of 30 kHz and a temperature of 17°C is −129 dBm.

The receiver front-end noise N_f of an average-quality receiver is 9 dB. The noise figure NF usually would add the front-end noise N_f of

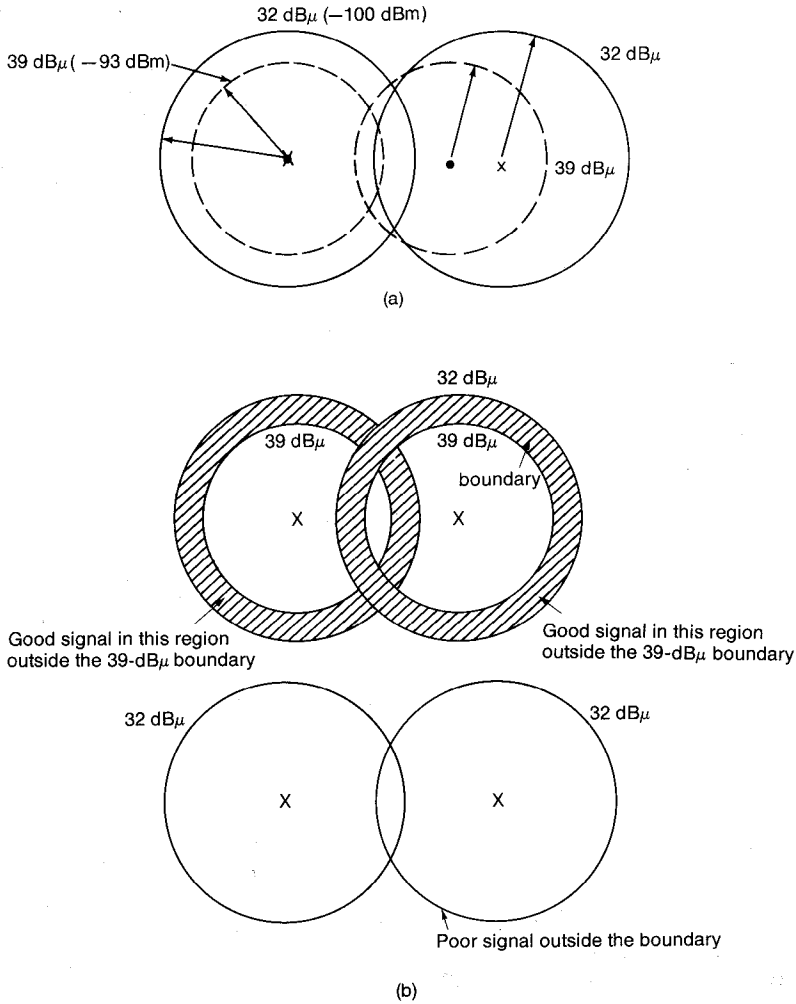


Figure 7.14 (*a*) Using a 32-dB boundary needs fewer cells to cover the area. (*b*) A signal outside its boundary generates noise.

the receiver and the noise N_{cm} introduced from the cellular mobile environment.

$$\mathrm{NF} = \sqrt{N_f^2 + N_{cm}^2} \qquad \mathrm{dB}$$

N_{cm} can either increase or decrease, depending on the system design. The earlier data indicate that N_{cm} can be neglected for 900-MHz curves.[7,8] If we now introduce a safety factor and let $N_{cm} = 6$ dB then

$$\mathrm{NF} = \sqrt{(9)^2 + (6)^2} = 11 \text{ dB}$$

The total noise level is $N = kTB + \mathrm{NF} = -118$ dBm. Because the required C/N is 18 dB, the lowest acceptable signal level is -100 dBm (-32 dBμ), which is 7 dB lower than -93 dBm (39 dBμ). In reality, the cell boundary or the handoff is based on the voice quality, that is, $C/N = 18$ dB or a level of -100 dBm; therefore, the FCC cell boundary of 39 dBμ or -93 dBm is 7 dB higher than the level provided by the system. Thus a cell boundary of 32 dBμ or -100 dBm proved to be sufficient for cellular coverage.

The two main advantages of using a 32-dBμ level (see Fig. 7.14) are that (1) fewer cell sites would be needed to cover a growth area and (2) less interference would be effected at the boundaries. A 32-dB boundary for cells in either boundary of a metropolitan statistical area (MSA) or a rural service area (RSA) is a proper operation, as opposed to a 39-dBμ boundary which is an artificial value.

7.7 Effects on Cell-Site Components

7.7.1 Channel combiner

A fixed-tuned channel combiner at the transmitting side. A channel combiner is installed at each cell site. Then all the transmitted channels can be combined with minimum insertion loss and maximum signal isolation between channels. Of course, we can eliminate the channel combiner by letting each channel feed to its own antenna. Then a 16-channel site will have 16 antennas for operation. It is an economical and a physical constraint.

A conventional combiner has a 16-channel combined capacity based on the frequency subset of 16 channels, and it causes each channel to lose 3 dB from inserting the signal through the combiner. The signal isolation is 17 dB because each channel is 630 kHz or 21 channels apart from neighboring channels (Fig. 7.15*a*). The intermodulation at the multiplexer is controlled by ferrite isolators, which provide a 30-dB reverse loss. The intermodulation (IM) products are at least 55 dB down

from the desired signals. Therefore, the IM will not affect channels within the transmitted band design from this.

Each cable fed into a combiner must be properly shielded. Because it is a nonlinear device, undesired signal leakage into another channel would occur before the combiner can produce the IM products, which would in turn, produce cross-coupled interference. Therefore, proper shielding and impedance match are very important. Fixed-tuned combiners are tuned to match the impedances of a set of fixed frequencies which are assigned to a combiner.

A frequency-agile combiner.[11] This combiner is capable of returning to any frequency by remote control in real time. The remote control device is a microprocessor. The combiner is a waveguide-resonator combiner with a tuning bar in each input waveguide as shown in Fig. 7.15*b*. The bar is mechanically rotated by a motor, and the voltage standing-wave ratio (VSWR) can be measured when the motor starts to turn. The controller received an optimum reading after a full turn and is stopped at that position by the controller. The controller also has a self-adjusting potential. This combiner can be used when a dynamic frequency assignment is applied. In many cases, it is preferable to redistribute the frequency channels to avoid prominent interference in certain areas. To use this kind of combiner, cell-site transceivers should also be able to change their operating frequencies, which are controlled by the MTSO, accordingly. This kind of combiner can also be designed to be tuned electronically.

A ring combiner.[12] A ring combiner is used to combine two groups of channels into a single output. The insertion loss is 3 dB, and the signal

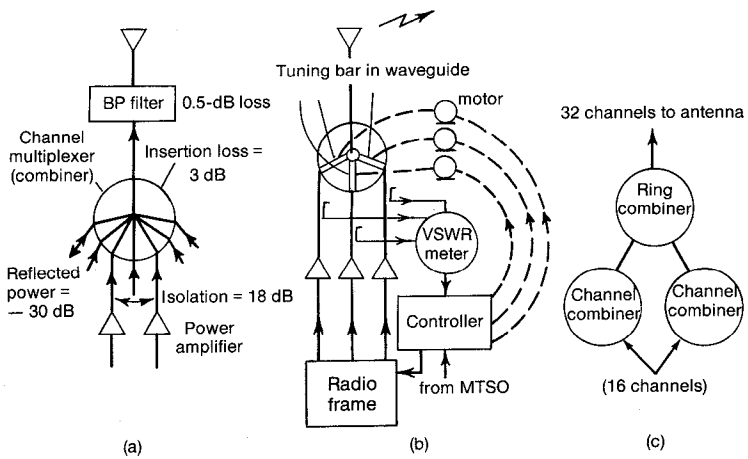


Figure 7.15 Different kinds of channel combiners. (*a*) Fixed-tuned combiner, (*b*) tunable combiner, (*c*) ring combiner.

isolation between channels is 35 to 40 dB. The function of a ring combiner is to combine two 16-channel combiners into one 32-channel output. Therefore, all 32 channels can be used by a single transmitting antenna. If a cell site has two antennas, up to 64 radio channels can be installed in it.

If all the channel-transmitted powers are low, it is possible to combine more than 32 channels by using two or three ring combiners before feeding them into one transmitting antenna. The total allowed transmitted power is a limiting factor. Some ring combiners have a 600-W power limitation. The use of ring combiners reduces adjacent channel separation. If two 16-channel regular combiners are combined with a ring combiner, the adjacent-channel separation at the ring combiner output can be 315 kHz, even though the adjacent-channel separation of each regular combiner is 630 kHz. It is simply a frequency offset of 315 kHz between two regular combiners.

7.7.2 Demultiplexer at the receiving end

A demultiplexer is used to receive 16 channels from one antenna. The demultiplexer is a filter bank as shown in Fig. 7.16. Then each receiving antenna output passes through a 25-dB-gain amplifier to a demultiplexer. The demultiplexer output has a 12-dB loss from the split of 16 channels.

$$\text{Split loss} = 10 \log 16 = 12 \text{ dB}$$

and the IM product at the output of the demultiplexer should be 65 dB down.[4] The two space-diversity antennas each connect to an umbrella filter (block A or B band filter) and have a 55-dB rejection from the other system band. If the undesired mobile unit is close to the cell site, then the preamplifier becomes saturated and generates IM at the output of the amplifier; these IM products (frequencies) could be felt in one of the weak incoming signals. This situation can lead to cross talk (see Sec. 7.4) which can be heard from both ends of the link because of a unique characteristic of cellular channels (see Sec. 7.5).

7.7.3 SAT tone

General description. The major function of a supervisory audio tone (SAT) is to ensure that a SAT tone is sent out at the cell site, is received by the mobile unit on a forward voice channel, is converted on a corresponding reverse voice channel, and is then sent back to the cell site within 5 s. If the time out is more than 5 s, the cell site will terminate the call.

Every cell site has been assigned to one of three SAT tones. The

assignment of three SAT tones in a system is shown in Fig. 7.17. The cells have the same SAT tones, and the same channels are separated by $\sqrt{3}D$, which is farther than the cochannel distance D. Therefore, a receiver located at either the cell site or at the mobile unit and receiving the same frequency with different SAT tones will terminate the call.

Characteristics of SAT. There are three SAT tones, 5970 H, 6000 H, and 6030 Hz, spaced 30 Hz apart. They are narrowband frequency-modulated (FM) with a deviation of $f_\Delta = 2$ kHz. The modulation index is $\beta = 1/3$. Let the SAT tone signal be

$$x(t) = A_m \cos \omega_m t \tag{7.7-1}$$

and the modulated carrier is

$$x_c(t) = A_c \cos(\omega_c t + \beta \sin \omega_m t) \tag{7.7-2}$$

where $\beta = (A_m f_\Delta / f_m)$. Let the amplitude modulation $A_m = 1$; thus,

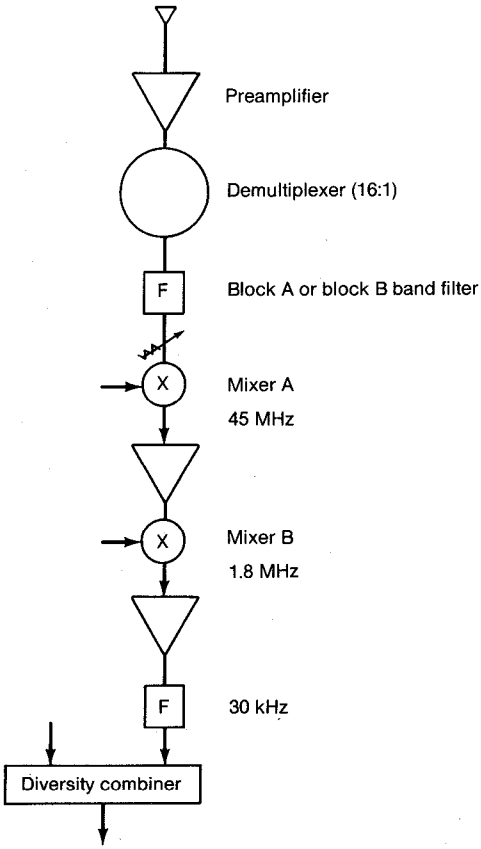


Figure 7.16 A typical cell-site channel receiver.

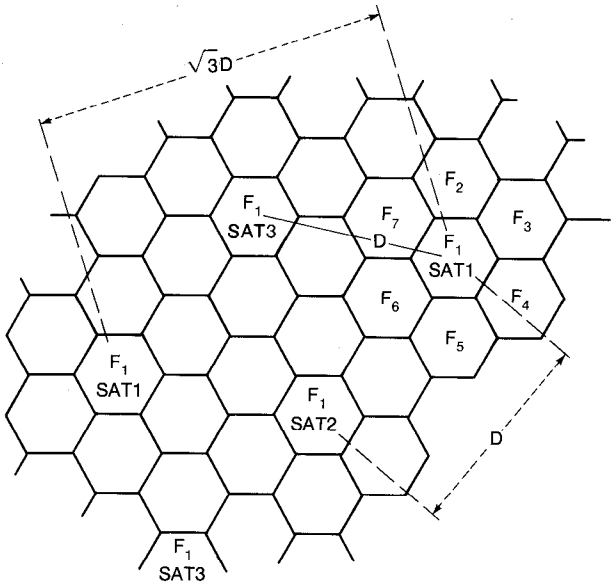


Figure 7.17 SAT spatial allocation.

since β is small, Eq. (7.7-2) becomes

$$x_c(t) \approx A_c \cos(\omega_c t) + \frac{A_c\beta}{2} \times \cos[2\pi(f_c + f_m)t] - \frac{A_c\beta}{2}\cos[2\pi(f_c - f_m)t]$$

$$= R(t)\cos[\omega_c t + \phi(t)] \tag{7.7-3}$$

where[13]

$$R(t) \approx \sqrt{A_c^2 + \left(2\frac{\beta}{2}A_c \sin \omega_m t\right)^2}$$

$$\approx A_c\left[1 + \frac{\beta^2}{4} - \frac{\beta^2}{4}\cos 2\omega_m t\right] \tag{7.7-4}$$

$$\phi(t) \approx \arctan\left[\frac{2(\beta/2)A_c \sin \omega_m t}{A_c}\right] \approx \beta \sin \omega_m t \tag{7.7-5}$$

The FM phasor diagram for $\beta << 1$ is shown in Fig. 7.18. Equation (7.7-4) represents an FM condition in which the amplitude of the carrier always remains constant. This means that the amplitude has no information content. This is a very common consideration in the mobile

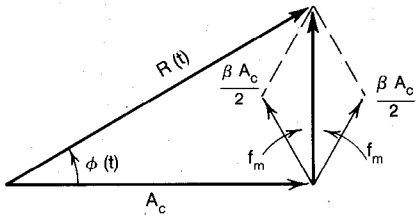


Figure 7.18 Narrowband FM modulation for SAT.

radio environment because of the severe fading which distorts the constant amplitude.

The SAT generator cannot deviate by more than ±15 Hz while receiving the signal. The SAT detector uses this criterion to continuously accept or reject a returned SAT. It has been observed that two SATS with two different audio tone amplitudes can arrive at one cell. If the desired SAT tone is weaker than the undesired one by a certain ratio, then the SAT tone will deviate by ±15 Hz. These conditions are discussed in Sec. 13.1.2. The filter bandwidth of the SAT tone detector relates to call-drop timing, which should be based on the unacceptable voice quality level. In theory, this level is different in different environment. Usually the smaller the filter bandwidth, the lower the call-drop rates. But the voice quality may be very poor before dropping the calls.

7.8 Interference between Systems

7.8.1 In one city

Let us assume that there are two systems operating in one city or one MSA. If a mobile unit of system A is closer to a cell site of system B while a call is being initiated through system A, adjacent channel interference or IM can be produced if the transmitted frequency of mobile unit A is close to the covered band of the received preamplifier at cell site B (see Fig. 7.19*a*). These IM products will then leak into the receiving channel of system B and cross talk will occur. This cross talk can be heard not only at the land-line side but also at the mobile unit because of the unique characteristics described in Sec. 7.5.

This cross-talk situation can be reduced by any of the following measures.

1. All cell sites in the two systems can be located together (*colocated*).
2. Adjacent channels (four or five channels) at each interface (see Fig. 7.6) of the new allocated voice channels between two of the systems should not be used.
3. To prevent a strong mobile signal from saturating the preamplifier at the cell site, a foreign-system signal should be −55 dBm down

from the cell-site reception point. Otherwise IM products can be produced and mixed with the desired system by passage through the system (band) block filter (see Fig. 7.16).

For instance, IM may occur in either of the following cases.

$$(2 \times 838 - 832)\ \text{MHz} = 844\ \text{MHz (system B at the cell site)}$$

[Either signal (838 or 832 MHz) is strong; the IM will leak into the 844-MHz channel.]

$$(2 \times 834 - 836\ \text{MHz} = 832\ \text{MHz (system A at the cell site)}$$

[Either signal (834 or 836 MHz) can be strong; the IM will leak into the 832-MHz channel.]

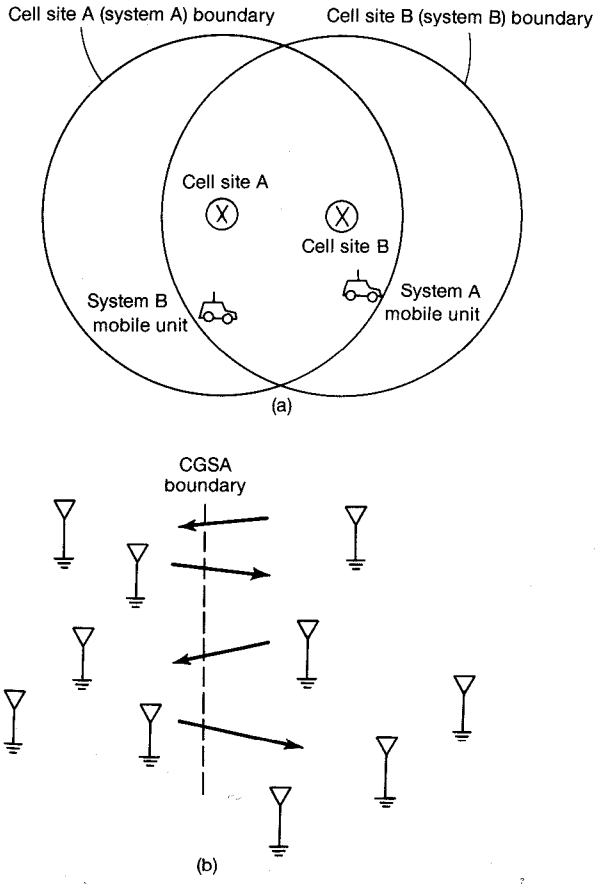


Figure 7.19 Intersystem interference. (*a*) System A cell sites in system B cell coverage; (*b*) interference between two cellular geographic service area (CGSA) systems.

7.8.2 In adjacent cities

Two systems operating at the same frequency band and in two adjacent cities or areas may interfere with each other if they do not coordinate their frequency channel use. Most cases of interference are due to cell sites at high altitudes (see Fig. 7.19*b*). In any start-up system, a high-altitude cell site is always attractive to the designer. Such a system can cover a larger area, and, in turn, fewer cell sites are needed. However, if the neighboring city also uses the same system block, then the result is strong interference, which can be avoided by the following methods.

1. The operating frequencies should be coordinated between two cities. The frequencies used in one city should not be used in the adjacent city. This arrangement is useful only for two low-capacity systems.
2. If both systems are high capacity, then decreasing the antenna heights will result in reduction of the interference not only within each system but also between the two systems.
3. Directional antennas may be used. For example, if one system is high capacity and the other is low capacity, the low-capacity system can use directional antennas but still retain the high tower. In this situation frequency coordination between the two systems has to be worked out at the common boundary because all the allocated frequencies must be used by the high-capacity system in its service area but only some frequencies are used by the low-capacity system.

7.9 UHF TV Interference

Two types of interference can occur between UHF television and 850-MHz cellular mobile phones.

7.9.1 Interference to UHF TV receivers from cellular mobile transmitters

Because of the wide frequency separation between cellular phone systems and the media broadcast services (TV and radio) and the significantly high power levels used by the UHF TV broadcast transmitters, the likelihood of interference from cellular phone transmissions affecting broadcasting is very small.[14,15] There is a slight probability that when the cell-site transmission is 90 MHz above that of a TV channel, it can interfere with the image-response frequency of typical home TV receivers. Interference between TV and cellular mobile channels is illustrated in Fig. 7.20.

Some UHF TV channels overlap cellular mobile channels. These two types of service can interfere with each other only under the following conditions.

1. *Band region with overlapping frequencies.* Two services have been authorized to operate within the same frequency band region.
2. *Image interference region.* This is explained as follows. The TV receiver or the cellular receiver (mobile unit or cell site) can receive two transmitted signals, for instance, one from a TV channel and one from a cellular system, and produce a third-order intermodulation product which falls within the TV or the mobile receive band.

Let

$$f_{Tm} = \text{mobile transmit frequency}$$

$$= f_{Rc} = \text{cell-site receive frequency} = f_{Tc} - 45 \text{ MHz}$$

$$f_{Rm} = \text{mobile receive frequency}$$

$$= f_{Tm} + 45 \text{ MHz} = f_{Tc} = \text{cell-site transmit frequency}$$

$$f_{T,\text{TV}} = \text{TV transmit frequency}$$

$$f_{R,\text{TV}} = \text{TV receive frequency}$$

Third-order intermodulation gives the following results in two cases of interfering UHF TV receivers.

Case 1. Let

$$2f_{Tm} - f_{T,\text{TV}} = f_{Rm} \tag{7.9-1}$$

$$f_{Tm} = f_{Rm} - 45 \tag{7.9-2}$$

then

$$f_{Tm} = f_{T,\text{TV}} + 45 \tag{7.9-3}$$

Since the mobile transmit frequency f_{Tm} lies in the 825- to 845-MHz band, and the TV transmit frequency $f_{T,\text{TV}}$ lies in 780- to 800-MHz band, f_{Tm} will interfere with the TV receiver as seen from Eq. (7.9-3). This interference region is called the *image interference region.*

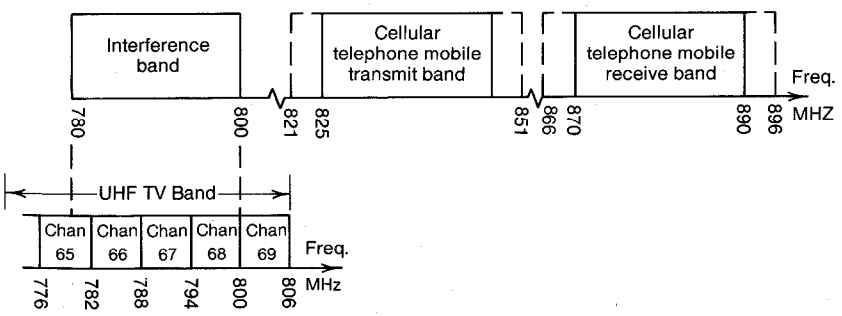


Figure 7.20 Cellular telephone frequency plan.

Case 2. Let

$$2f_{Rc} - f_{T,\mathrm{TV}} = f_{Tc} \tag{7.9-4}$$

then

$$f_{Rc} = f_{Tc} - 45 \tag{7.9-5}$$

and

$$f_{Tc} = f_{T,\mathrm{TV}} + 90 \tag{7.9-6}$$

Because the cell-site transmit frequency f_{Tc} lies in the 870- to 890-MHz band, and $f_{T,\mathrm{TV}}$ lies in 780- to 800-MHz band f_{Tc} will interfere with the TV receiver, as shown in Eq. (7.9-6). This interference region is called the image interference region.

In these two cases an image-interference rejection range of 40 to 50 dB isolation across the UHF TV band is required to prevent this interference. The results from the two cases are as follows.

Case 1: When the mobile transmitter is located near a TV receiver (Eq. 7.9-3). The minimum grade B television service contour of an accepted TV receiver level is −63 dBm with a receiver antenna gain of 6 dB referring to dipole gain. Roughly, this kind of TV station has a coverage of a 56-km (35-mi) radius. Since the cellular telephone mobile unit has an effective radiated power (ERP) of about 37 dBm, the path loss between the TV receiver and the mobile unit must exceed 100 dB (= 63 + 37). The TV antenna height at each residence normally is about $h_2 = 10$ m. The mobile antenna height is about $h_1 = 2$ m. Assume that the cross-modulation loss between two frequency bands is 80 dB and the polarization coupling loss between the bands is 10 dB. Using the formula derived in Eq. (4.2-19), we obtain

$$-63 = 37 - 156 - 40 \log \frac{d_1}{d_0} + 10 \log h_1 + 20 \log h_2 + 6 - (80 + 10) \quad \text{dB} \tag{7.9-7}$$

Substitution of $h_1 = 2$ m (6 ft) and $h_2 = 10$ m (30 ft) into Eq. (7.9-7) yields

$$140 = -40 \log d_1 + 7.78 + 29.54$$

We can solve d_1 as

$$d_1 = 10^{-2.57} = 0.00239 \text{ mi} = 14 \text{ ft}$$

We find that the required distance from a transmitting cellular mobile unit to a TV receiver is only 14 ft. Besides, the mobile unit is always moving while the TV receivers usually are off; thus, the chance of

mobile unit interference occurring within 14 ft of the receiver while TV receivers are operative is very slim. In addition, the chances are that the mobile unit would remain in the area of interference for only 5 to 10 s.

Case 2: When the cell site transmitter is located near a TV receiver (Eq. 7.9-6). Usually cell-site antennas are located on high towers, and the vertical antenna pattern usually produces a null under the antenna tower. Therefore, even though Eq. (7.9-6) indicates the possibility of cell-site interference, the TV receivers near the cell site will not be in the area of the main antenna beam and, clearly, the horizontally polarized TV wave will not be distorted by the cellular vertically polarized waves when it reaches the TV receiving antenna on the roof of the house. Because of these differences between antenna beam pattern and waves polarization, no strong interference can be seen in this case. We find that the required distance could be less than 200 m (700 ft). We should also consider the following key points.

1. The polarization coupling loss from vertical (cellular) to horizontal (TV) waves can be 10 dB, according to Lee and Yeh's data.[16]
2. The percentage of active mobile units in that area is small.
3. In the UHF TV fringe areas, cable TV (CATV) usually provides the service.
4. Only four TV channels (Channels 65 to 68) can experience interference. The chance of one TV set tuning to one of these four "interference channels" and the active mobile unit happening to be in that area at the same time is slim.
5. Even if transmission from the mobile unit does interfere with TV reception, the interference time is very short (<15 s). Therefore, no interference should be encountered.

7.9.2 Interference of cellular mobile receivers by UHF TV transmitters

This type of image interference can occur in the following four cases. Here the image-interference region will be the same as that described in Sec. 7.9.1 but in the reversed direction.

Case 1. Let

$$2f_{Tm} - f_{T,\mathrm{TV}} = f_{Rm} \tag{7.9-8}$$

Then

$$2f_{Tm} = 2(f_{Rm} - 45) \tag{7.9-9}$$

and

$$F_{T,\mathrm{TV}} = 2f_{Tm} - f_{Rm} = f_{Rm} - 90 \text{ MHz} \tag{7.9-10}$$

Because the mobile unit receiver frequency f_{Rm} lies in the 870- to 890-MHz band, $f_{T,\mathrm{TV}}$, which lies in 780- to 800-MHz band, will interfere with the mobile unit receiver, as shown in Eq. (7.9-10).

Case 2. Let

$$2f_{Rc} - f_{T,\mathrm{TV}} = f_{Tc} \tag{7.9-11}$$

Then

$$f_{Rc} = f_{Tc} - 45 \tag{7.9-12}$$

and

$$f_{Rc} = 2f_{Rc} - f_{T,\mathrm{TV}} - 45 = f_{T,\mathrm{TV}} + 45 \tag{7.9-13}$$

Since the cell-site receiver frequency f_{Rc} lies in the 825- to 845-MHz band, $f_{T,\mathrm{TV}}$, which lies in 780- to 800-MHz band, will interfere with the cell-site receiver as shown in Eq. (7.9-13). There are two additional, but less important, cases.

Case 3. When a mobile receiver approaches a TV transmitter, it is easy to find that transmission from the TV station will not interfere with the reception at the mobile receiver by following the same analysis shown in Sec. 7.9.1, case 2.

Case 4. When the cell-site receiver is only 1 mi or less away from the TV station, interference may result. However when the cell site is very close to the TV station, the interference decreases as a result of the two vertical narrow beams pointing at different elevation levels. For this reason it is advisable to mount a cell-site antenna in the same vicinity as the TV station antenna if the problems of shielding and grounding can be controlled.

7.10 Long-Distance Interference

7.10.1 Overwater path

This phenomenon is mentioned in several reports.[17,18]

1. A 41-mi overwater path operating at 1.5 GHz in Massachusetts Bay[17]
   a. Low ducts (<50 ft thick); steady signal well above normal level is received
   b. High ducts (≥100 ft thick); a high signal level generally on the average is received but with deep fading
2. A 275-mi overwater path operating at 812 and 857 MHz between Charleston, South Carolina and Daytona Beach, Florida
   a. Charleston—antenna height 500 ft above average terrain antenna pattern, omnidirectional ERP 220 W; receiving sensitivity less than 0.5 μV = −113 dBm (1 μV = −107 dBm) with a 50-Ω terminal

b. Daytona Beach—antenna height 920 ft above average terrain antenna pattern, omnidirectional ERP 440 W; receiving sensitivity 0.7 μV = −110 dBm

Federal Express engineers have discovered the following phenomenon through study of their system.[18] The mobile units in Charleston within 1 to 2 mi of shoreline are capable of clear communication with a repeater station in Daytona Beach. The same situation applies when the mobile unit is in Daytona Beach. These clear path communications occur regardless of weather, time of day, or season. This is a tropospherical propagation, and we should eliminate it in cellular systems to avoid interference among systems in North America. One way of doing this is by use of umbrella antenna patterns.

7.10.2 Overland path

Tropospheric scattering over a land path is not as persistent as that over water and can be varied from time to time. Usually tropospheric propagation is more pronounced in the morning. The distance can be about 200 mi. Federal Express engineers have observed this long-distance propagation throughout their nationwide system.

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Chapter

8

Frequency Management and Channel Assignment

8.1 Frequency Management

The function of frequency management is to divide the total number of available channels into subsets which can be assigned to each cell either in a fixed fashion or dynamically (i.e., in response to any channel among the total available channels).

The terms "frequency management" and "channel assignment" often create some confusion. *Frequency management* refers to designating set-up channels and voice channels (done by the FCC), numbering the channels (done by the FCC), and grouping the voice channels into subsets (done by each system according to its preference). *Channel assignment* refers to the allocation of specific channels to cell sites and mobile units. A fixed channel set consisting of one or more subsets (see Sec. 8.1.2) is assigned to a cell site on a long-term basis. During a call, a particular channel is assigned to a mobile unit on a short-term basis. For a short-term assignment, one channel assignment per call is handled by the mobile telephone switching office (MTSO). Ideally channel assignment should be based on causing the least interference in the system. However, most cellular systems cannot perform this way.

8.1.1 Numbering the channels

The total number of channels at present (January 1988) is 832. But most mobile units and systems are still operating on 666 channels. Therefore we describe the 666 channel numbering first. A channel consists of two frequency channel bandwidths, one in the low band and one in the high band. Two frequencies in channel 1 are 825.030 MHz (mobile transmit) and 870.030 MHz (cell-site transmit). The two frequencies in channel 666 are 844.98 MHz (mobile transmit) and 889.98 MHz (cell-site transmit). The 666 channels are divided into two groups: block A system and block B system. Each market (i.e., each city) has two systems for a duopoly market policy (see Chap. 1). Each block has 333 channels, as shown in Fig. 8.1.

The 42 set-up channels are assigned as follows.

Channels 313–333	block A
Channels 334–354	block B

The voice channels are assigned as follows.

Channels 1–312 (312 voice channels)	block A
Channels 355–666 (312 voice channels)	block B

These 42 set-up channels are assigned in the middle of all the assigned channels to facilitate scanning of those channels by frequency syn-

	1A	2A	3A	4A	5A	6A	7A	1B	2B	3B	4B	5B	6B	7B	1C	2C	3C	4C	5C	6C	7C	
	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	
	22	23	24	25	26	27	28	29	30	31	32	33	34	35	36	37	38	39	40	41	42	
	43	44	45	46	47	48	49	50	51	52	53	54	55	56	57	58	59	60	61	62	63	
	64	65	66	67	68	69	70	71	72	73	74	75	76	77	78	79	80	81	82	83	84	
	85	86	87	88	89	90	91	92	93	94	95	96	97	98	99	100	101	102	103	104	105	
	106	107	108	109	110	111	112	113	114	115	116	117	118	119	120	121	122	123	124	125	126	
	127	128	129	130	131	132	133	134	135	136	137	138	139	140	141	142	143	144	145	146	147	
	148	149	150	151	152	153	154	155	156	157	158	159	160	161	162	163	164	165	166	167	168	
	169	170	171	172	173	174	175	176	177	178	179	180	181	182	183	184	185	186	187	188	189	
	190	191	192	193	194	195	196	197	198	199	200	201	202	203	204	205	206	207	208	209	210	
	211	212	213	214	215	216	217	218	219	220	221	222	223	224	225	226	227	228	229	230	231	
	232	233	234	235	236	237	238	239	240	241	242	243	244	245	246	247	248	249	250	251	252	
Block A system	253	254	255	256	257	258	259	260	261	262	263	264	265	266	267	268	269	270	271	272	273	
	274	275	276	277	278	279	280	281	282	283	284	285	286	287	288	289	290	291	292	293	294	
↑	295	296	297	298	299	300	301	302	303	304	305	306	307	308	309	310	311	312	—	—	—	
	313	314	315	316	317	318	319	320	321	322	323	324	325	326	327	328	329	330	331	332	333	Control channel sets
↓	334	335	336	337	338	339	340	341	342	343	344	345	346	347	348	349	350	351	352	353	354	
Block B system	355	356	357	358	359	360	361	362	363	364	365	366	367	368	369	370	371	372	373	374	375	
	376	377	378	379	380	381	382	383	384	385	386	387	388	389	390	391	392	393	394	395	396	
	397	398	399	400	401	402	403	404	405	406	407	408	409	410	411	412	413	414	415	416	417	
	418	419	420	421	422	423	424	425	426	427	428	429	430	431	432	433	434	435	436	437	438	
	439	440	441	442	443	444	445	446	447	448	449	450	451	452	453	454	455	456	457	458	459	
	460	461	462	463	464	465	466	467	468	469	470	471	472	473	474	475	476	477	478	479	480	
	481	482	483	484	485	486	487	488	489	490	491	492	493	494	495	496	497	498	499	500	501	
	502	503	504	505	506	507	508	509	510	511	512	513	514	515	516	517	518	519	520	521	522	
	523	524	525	526	527	528	529	530	531	532	533	534	535	536	537	538	539	540	541	542	543	
	544	545	546	547	548	549	550	551	552	553	554	555	556	557	558	559	569	561	562	563	564	
	565	566	567	568	569	570	571	572	573	574	575	576	577	578	579	580	581	582	583	584	585	
	586	587	588	589	590	591	592	593	594	595	596	597	598	599	600	601	602	603	604	605	606	
	607	608	609	610	611	612	613	614	615	616	617	618	619	620	621	622	623	624	625	626	627	
	628	629	630	631	632	633	634	635	636	637	638	639	640	641	642	643	644	645	646	647	648	
	649	650	651	652	653	654	655	656	657	658	659	660	661	662	663	664	665	666	—	—	—	

Figure 8.1 Frequency-management chart.

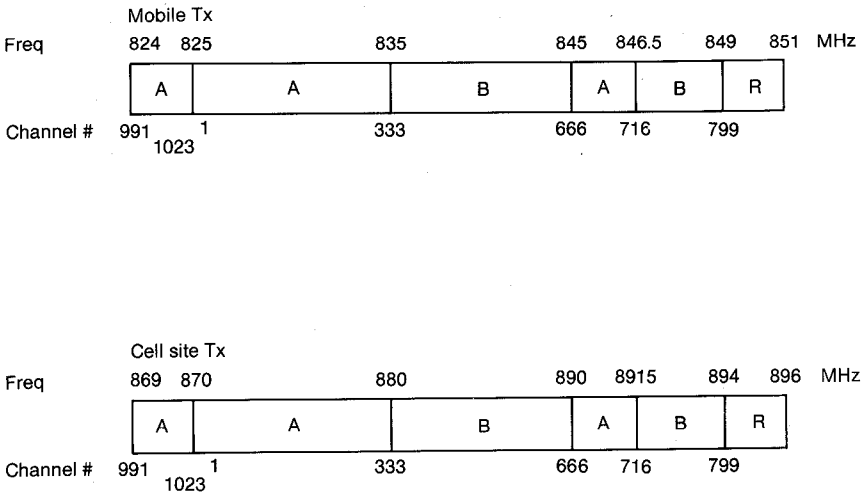


Figure 8.2 New additional spectrum allocation.

thesizers (see Fig. 8.1). In the new additional spectrum allocation of 10 MHz (see Fig. 8.2), an additional 166 channels are assigned. Since a 1 MHz is assigned below 825 MHz (or 870 MHz), in the future, additional channels will be numbered up to 849 MHz (or 894 MHz) and will then circle back. The last channel number is 1023 ($=2^{10}$). There are no channels between channels 799 and 991.

8.1.2 Grouping into subsets

The number of voice channels for each system* is 312. We can group these into any number of subsets. Since there are 21 set-up channels for each system, it is logical to group the 312 channels into 21 subsets. Each subset then consists of 16 channels. In each set, the closest adjacent channel is 21 channels away, as shown in Fig. 8.1. The 16 channels in each subset can be mounted on a frame and connected to a channel combiner. Wide separation between adjacent channels is required for meeting the requirement of minimum isolation. Each 16-channel subset is idealized for each 16-channel combiner. In a seven-cell frequency-reuse cell system each cell contains three subsets, $iA + iB + iC$, where i is an integer from 1 to 7. The total number of voice channels in a cell is about 45. The minimum separation between three subsets is 7 channels. If six subsets are equipped in an omnicell site, the minimum separation between two adjacent channels can be only three ($21/6 > 3$) physical channel bandwidths.

* Not including the new 83 voice channels.

For example,

$$1A + 1B + 1C + 4A + 4B + 4C$$

$$\text{or} \quad 1A + 1B + 1C + 5A + 5B + 5C$$

The antenna arrangement for 90 voice channels was described in Sec. 5.4.1. The requirements for channel separation in a cell are discussed in this chapter.

8.2 Frequency-Spectrum Utilization

Since the radio-frequency spectrum is finite in mobile radio systems, the most significant challenge is to use the radio-frequency spectrum as efficiently as possible. Geographic location is an important factor in the application of the frequency-reuse concept in mobile cellular technology to increase spectrum efficiency. Frequency management involving the assignment of proper channels in different cells can increase spectrum efficiency. Thus, within a cell, the channel assignment for each call is studied. Other factors, such as narrowing of the frequency band, off-air call setup, queuing, and call redirect, are described in different chapters.

The techniques for increasing frequency spectrum can be classified as

1. Increasing the number of radio channels using narrow banding, spread spectrum, or time division (Chap. 13)
2. Improving spatial frequency-spectrum reuse (Chaps. 2, 6, and 7)
3. Frequency management and channel assignment (Chap. 8)
4. Improving spectrum efficiency in time (Chap. 14)
5. Reducing the load of invalid calls (Chap. 11)
   a. Off-air call setup—reducing the load of set-up channels
   b. Voice storage service for No-Answer calls
   c. Call forwarding
   d. Reducing the customers' Keep-Dialing cases
   e. Call waiting for Busy-Call situations
   f. Queuing

In this chapter we concentrate on frequency management and channel assignment (item 3).

8.3 Set-up Channels

Set-up channels, also called *control channels,* are the channels designated to set up calls. We should not be confused by the fact that a call always needs a set-up channel. A system can be operated without set-

up channels. If we are choosing such a system, then all 333 channels in each cellular system (block A or block B) can be voice channels; however, each mobile unit must then scan 333 channels continuously and detect the signaling for its call. A customer who wants to initiate a call must scan all the channels and find an idle (unoccupied) one to use.

In a cellular system, we are implementing frequency-reuse concepts. In this case the set-up channels are acting as control channels. The 21 set-up channels are taken out from the total number of channels. The number 21 is derived from a seven-cell frequency-reuse pattern with three 120° sectors per cell, or a total of 21 sectors, which require 21 set-up channels. However, now only a few of the 21 set-up channels are being used in each system. Theoretically, when cell size decreases, the use of set-up channels should increase.

Set-up channels can be classified by usage into two types: *access channels* and *paging channels*. An access channel is used for the mobile-originating calls and paging channels for the land-originating calls. In a low-traffic system, access channels and paging channels are the same. For this reason, a set-up channel is sometimes called an "access channel" and sometimes called a "paging channel." Every two-way channel contains two 30-kHz bandwidths. Normally one set-up channel is also specified by two operations as a forward set-up channel (using the upper band) and a reverse set-up channel (using the lower band). In the most common types of cellular systems, one set-up channel is used for both paging and access. The forward set-up channel functions as the paging channel for responding to the mobile-originating calls. The reverse set-up channel functions as the access channel for the responder to the paging call. The forward set-up channel is transmitted at the cell site, and the reverse set-up channel is transmitted at the mobile unit. All set-up channels carry data information only.

8.3.1 Access channels

In mobile-originating calls, the mobile unit scans its 21 set-up channels and chooses the strongest one. Because each set-up channel is associated with one cell, the strongest set-up channel indicates which cell is to serve the mobile-originating calls. The mobile unit detects the system information transmitted from the cell site. Also, the mobile unit monitors the Busy/Idle status bits over the desired forward set-up channel. When the Idle bits are received, the mobile unit can use the corresponding reverse set-up channel to initiate a call.

Frequently only one system operates in a given city; for instance, block B system might be operating and the mobile unit could be set to

"preferable A system." When the mobile unit first scans the 21 set-up channels in block A, two conditions can occur.

1. If no set-up channels of block A are operational, the mobile unit automatically switches to block B.
2. If a strong set-up signal strength is received but no message can be detected, then the scanner chooses the second strongest set-up channel. If the message still cannot be detected, the mobile unit switches to block B and scans the block B set-up channels.

The operational functions are described as follows.

1. *Power of a forward set-up channel* [*or forward control channel* (*FOCC*)]. The power of the set-up channel can be varied in order to control the number of incoming calls served by the cell. The number of mobile-originating calls is limited by the number of voice channels in each cell site. When the traffic is heavy, most voice channels are occupied and the power of the set-up channel should be reduced in order to reduce the coverage of the cell for the incoming calls originating from the mobile unit. This will force the mobile units to originate calls from other cell sites, assuming that all cells are adequately overlapped.
2. *The set-up channel received level.* The set-up channel threshold level is determined in order to control the reception at the reverse control channel (RECC). If the received power level is greater than the given set-up threshold level, the call request will be taken.
3. *Change power at the mobile unit.* When the mobile unit monitors the strongest signal strength from all set-up channels and selects that channel to receive the messages, there are three types of message.
   a. *Mobile station control message.* This message is used for paging and consists of one, two, or four words—DCC, MIN, SCC, and VMAX (see Chap. 3).
   b. *System parameter overhead message.* This message contains two words, including DCC, SID, CMAX, or CPA (see Chap. 3).
   c. *Control-filler message.* This message may be sent with a system parameter overhead message, CMAC—a control mobile attenuation code (seven levels).
4. *Direct call retry.* When a cell site has no available voice channels, it can send a direct call-retry message through the set-up channel. The mobile unit will initiate the call from a neighboring cell which is on the list of neighboring cells in the direct call-retry message.

8.3.2 Paging channels

Each cell site has been allocated its own set-up channel (control channel). The assigned forward set-up channel (FOCC) of each cell site is used to page the mobile unit with the same mobile station control message (discussed in Chap. 3 and Sec. 8.3.1).

Because the same message is transmitted by the different set-up channels, no simulcast interference occurs in the system. The algorithm for paging a mobile unit can be performed in different ways. The simplest way is to page from all the cell sites. This can occupy a large amount of the traffic load. The other way is to page in an area corresponding to the mobile unit phone number. If there is no answer, the system tries to page in other areas. The drawback is that response time is sometimes too long.

When the mobile unit responds to the page on the reverse set-up channel, the cell site which receives the response checks the signal reception level and makes a decision regarding the voice channel assignment based on least interference in the selected sector or underlay-overlay region.

8.3.3 Self-location scheme at the mobile unit

In the cellular system, 80 percent of calls originate from the mobile unit but only 20 percent originate from the land line. Thus, it is necessary to keep the reverse set-up channels as open as possible. For this reason, the self-location scheme at the mobile unit is adapted. The mobile unit selects a set-up channel of one cell site and makes a mobile-originating call. It is called a *self-location scheme.*

However, the self-location scheme at the mobile unit prevents the mobile unit from sending the necessary information regarding its location to the cell site. Therefore, the MTSO does not know where the mobile is. When a land-line call is originated, the MTSO must page all the cell sites in order to search for the mobile unit. Fortunately, land-line calls constitute only 20 percent of land-line originating calls, so the cellular system has no problem in handling them. Besides, more than 50 percent of land-line originating calls are no response.

8.3.4 Autonomous registration

If a mobile station is equipped for autonomous registration, then the mobile station stores the value of the last registration number (REGID) received on a forward control channel. Also, a REGINCR (the increment in time between registrations) is received by the mobile station.

The next registration ID should be (see Chap. 3)

$$\text{NXTREG} = \text{REGID} + \text{REGINCR}$$

This tells the mobile unit how long the registration should be repeatedly sent to the cell site, so that the MTSO can track the location of the mobile. This feature is not used in cellular systems at present. However, when the volume of land-line calls begins to increase or the number of cell sites increases, this feature would facilitate paging of the mobile units with less occupancy time on all set-up channels. The trade-off between the self-location scheme and autonomous registration is shown in the following two examples.

Example 8.1 The time spent in the set-up channels for two schemes are compared.

1. Evaluation of a self-location scheme on a land-originating call. Assume that a system has 100 cell sites and a call paging has to reach all 100 cell sites. If every page takes 100 ms and there are 2000 land-originating calls per hour during a busy hour, then the air time spent for the paging during the busy hour is

$$100 \times (100 \text{ ms}) \times 2000 = 20{,}000 \text{ s} = 333 \text{ min/h}$$

This is the time spent on all set-up channels.

2. Evaluation of a registration scheme used on an idle stage for locating mobile units. Assume that the registration for each mobile unit is five times per hour. Each registration takes 100 ms. If 20,000 mobile units are on the road, then

$$(5 \times 100 \text{ ms}) \times 20{,}000 = 1000 \text{ s} = 166.7 \text{ min}$$

This is the time spent on all set-up channels.

In Example 8.1, the time spent on the set-up channels for a self-location scheme is twice as much as that for a registration scheme. In this particular case, the registration scheme is preferable to the self-location scheme.

Example 8.2 Assume that the reverse set-up channels also take the mobile-originating calls, which make up 80 percent of the total number of calls. Assume that 2500 land-originating calls constitute 20 percent of the total number of calls; then the mobile-originating calls represent 10,000 calls per hour handled by the MTSO. Each call initiation takes about 300 ms. Then

$$10{,}000 \times 300 = 3000 \text{ s} = 50 \text{ min}$$

The 50 min is occupied in both schemes. This is because for a mobile-originating call the self-location scheme provides a negligible time for selecting a desired cell site on a reverse set-up channel. The same negligible time is provided by using the registration scheme for selecting the desired cell site.

In a busy (rush) hour, the attempted call originating at a mobile unit is searching for an idle bit sent from the cell site. If an idle bit cannot

be received at the mobile unit after 10 attempts, then a busy tone is heard at the mobile unit.

Therefore, the 50 min calculated above assumes that all 10,000 calls are not blocked. In reality, there is always a certain amount of call blocking during a rush hour. Therefore, even though the MTSO will spend 50 min in a system to process 10,000 calls per hour, the actual attempt calls can be much higher.

8.3.5 Traffic load on a set-up channel and on *N* voice channels

When the traffic of a cell is increasing, more radios will be installed. When a cell has 90 voice channels (radios), one set-up channel must coordinate them in order to set up the calls. On the average, the cell site takes a mobile-originating call on a reverse set-up channel for 100 ms, and the interval between calls is 25 ms (including calls colliding in the air). Thus, in 1 h, if a queuing scheme is applied, the maximum number of calls that a set-up channel can accommodate is

$$\frac{3600 \times 1000}{125 \text{ ms}} = 28{,}800 \text{ calls/h}$$

This equation is based on the assumption that the incoming calls from the mobile units are waiting for the idle bits showing on the forward set-up channel before sending the requests. This is equivalent to a queuing scheme. In general, the waiting period is 1 to 2 s. If the set-up channel is busy during this period, the mobile unit will periodically continue to search for idle channels about every 100 ms. Then the initiating call will be blocked after 10 attempts. An estimate of call blocking can be obtained by using a queuing model. Without queuing schemes, the maximum number of initiating calls that the set-up channel can take during a busy hour, assuming five attempts per call, is 5760 calls per hour.

To calculate the traffic load on 90 voice channels, let us assume a blocking probability of 0.02 and a holding time of 100 s. Now we can check the offered load a from Table 1.1.

$$a = 78.3$$

The number of calls is

$$M = \frac{78.3 \times 3600}{100} = 2818 \text{ calls/h}$$

The carried load of a set-up channel is always greater than the carried load of the 90 voice channels. A load of more than 90 radios in a cell

is not unusual. However, the number of voice channels in a cell rarely exceeds 120. Therefore one set-up channel is used in a cell.

8.3.6 Separation between access and paging

All 21 set-up channels are actually paging channels. The access channel can be assigned by the MTSO as a channel other than the 21 set-up channels in a cell. The mobile unit receives the access channel information from the forward paging channels. In certain cases, as land-originating calls increase, one set-up channel cannot handle all set-up traffic in a cell. In such cases another channel in a group of voice channels is used as an access channel. Now the land-originating calls are using paging channels and the mobile-originating calls are using access channels.

8.3.7 Selecting a voice channel

Assume that a mobile unit calls or responds to a call through a reverse set-up channel which is received from an omnidirectional antenna and the voice channels are assigned from a forward set-up channel at one of three 120°-sector directional antennas.

For mobile-originating calls. The mobile unit selects a cell site based on its received signal-strength indicator (RSSI) reading. When a call of a mobile unit is received by the cell site, the set-up channel receives it through an omnidirectional antenna. The cell-site RSSI scans the incoming signals through three directional antennas and determines which sector is the strongest one. The MTSO then assigns a channel from among those channels designated in that sector. In some systems, a set-up channel is assigned to each sector of a cell.

For paging calls. When any call responds to the cell site, the cell-site RSSI will measure the incoming signal from the three directional antennas and find the strongest sector in which the channel can be assigned to the mobile unit.

8.4 Definition of Channel Assignment

8.4.1 Channel assignment to the cell sites—fixed channel assignment

In a fixed channel assignment, the channels are usually assigned to the cell site for relatively long periods. Two types of channels are assigned: set-up channels and voice channels.

Set-up channels. There are 21 set-up channels assigned each cell in a $K = 4, K = 7$, or $K = 12$ frequency-reuse pattern. If the set-up channel antennas are omnidirectional, then each cell only needs one set-up channel. This leaves many unused set-up channels. However, the set-up channels of blocks A and B are adjacent to each other. In order to avoid interference between two systems, the set-up channels in the neighborhood of Channel 333 (block A) and Channel 334 (block B) are preferably unused.

Voice channels. One way of dividing the total voice channels into 21 sets is exemplified in Sec. 8.1. The assignment of certain sets of voice channels in each cell site is based on causing minimum cochannel and adjacent-channel interference. Cochannel and adjacent channel interference can be calculated from equations in Chaps. 6 and 7.

Supervisory audio tone (SAT). This consists of three SATs. Based on the assignment of each SAT in each cell, we can show the method for further reducing cochannel interference, as mentioned in Sec. 7.7.2.

8.4.2 Channel assignment to traveling mobile units

This situation always occurs in the morning, when cars travel into the city, and at night, when the traffic pattern reverses. If the traffic density is uniform, the unsymmetrical mobile-unit antenna pattern (assuming large backward energy from the motion of the vehicle) does not affect the system operation much. However, when the traffic becomes heavier as more cars approach the city, the traffic pattern becomes nonuniform and the sites closest to the city, or in the city, cannot receive the expected number of calls or handoffs in the morning because of the mobile unit antenna patterns. At night, as the cars move out of the city, the cell sites closest to the city would have a hard time handing off calls to the sites away from the city.

To solve these problems, we have to use less transmitted power for both set-up and voice channels for certain cell sites. We also have to raise the threshold level for reverse set-up channels and voice channels at certain cell sites in order to control the acceptance of incoming calls and handoff calls. Three methods can be used.

Underlay-overlay.[1] The traffic capacity at an omnidirectional cell or a directional cell (see Fig. 8.3) can be increased by using the underlay-overlay arrangement. The underlay is the inner circle, and the overlay is the outer ring. The transmitted powers of the voice channels at the site are adjusted for these two areas. Then different voice frequencies

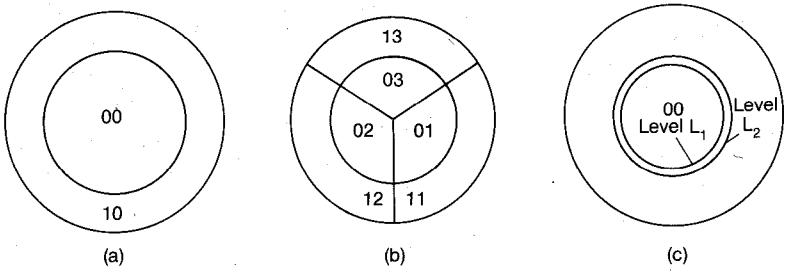


Figure 8.3 Underlaid-overlaid cell arrangements. (*a*) Underlay-overlay in omnicell; (*b*) underlay-overlay in sectorized cells; (*c*) two-level handoff scheme.

are assigned to each area. In an omnidirectional cell, the frequency-reuse distance of a seven-cell reuse pattern is $D = 4.6R$, where R is the radius of the cell. One overlay and one underlay are shown in Fig. 8.3*a*. Because of the sectorization in a directional cell, the channel assignment has a different algorithm in six regions (Fig. 8.3*b*), i.e., three overlay regions and three underlay regions. A detailed description is given in Sec. 8.5.4.

Frequency assignment. We assign the frequencies by a set of channels or any part of a set or more than one set of the total 21 sets. Borrowed-frequency sets are used when needed. On the basis of coverage prediction, we can assign frequencies intelligently at one site or at one sector without interfering with adjacent cochannel sectors or cochannel cells.

Tilted antenna. The tilted directional antenna arrangement can eliminate interference. Sometimes antenna tilting is more effective than decreasing antenna height, especially in areas of tall trees or at high sites. When the tilting angles become 22° or greater, the horizontal pattern creates a notch in the front of the antenna, which can further reduce the interference (see Fig. 6.10).

8.5 Fixed Channel Assignment

8.5.1 Adjacent-channel assignment

Adjacent-channel assignment includes neighboring-channel assignment and next-channel assignment. The near-end–far-end (ratio) interference, as mentioned in Sec. 7.3.1, can occur among the neighboring channels (four channels on each side of the desired channel). Therefore, within a cell we have to be sure to assign neighboring channels in an omnidirectional-cell system and in a directional-antenna-cell system

properly. In an omnidirectional-cell system, if one channel is assigned to the middle cell of seven cells, next channels cannot be assigned in the same cell. Also, no next channel (preferably including neighboring channels) should be assigned in the six neighboring sites in the same cell system area (Fig. 8.4*a*). In a directional-antenna-cell system, if one channel is assigned to a face, next channels cannot be assigned to the same face or to the other two faces in the same cell. Also, next channels cannot be assigned to the other two faces at the same cell site (Fig. 8.4*b*). Sometimes the next channels are assigned in the next sector of the same cell in order to increase capacity. Then performance can still be in the tolerance range if the design is proper.

8.5.2 Channel sharing and borrowing[2,3]

Channel sharing. Channel sharing is a short-term traffic-relief scheme. A scheme used for a seven-cell three-face system is shown in Fig. 8.5. There are 21 channel sets, with each set consisting of about 16 channels. Figure 8.5 shows the channel set numbers. When a cell needs more channels, the channels of another face at the same cell site can be shared to handle the short-term overload. To obey the adjacent-channel assignment algorithm, the sharing is always cyclic. Sharing always increases the trunking efficiency of channels. Since we cannot allow adjacent channels to share with the nominal channels in the same cell, channel sets 4 and 5 cannot both be shared with channel sets 12 and 18, as indicated by the grid mark. Many grid marks are indicated in Fig. 8.5 for the same reason. However, the upper subset of set 4 can be shared with the lower subset of set 5 with no interference.

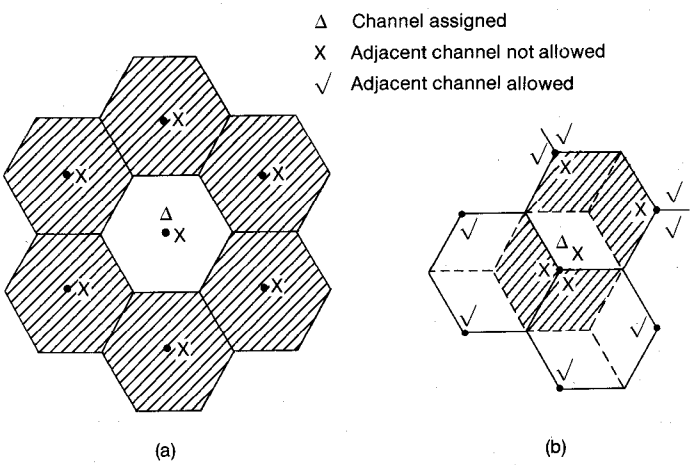


Figure 8.4 Adjacent channel assignment. (*a*) Omnidirectional-antenna cells; (*b*) directional-antenna cells.

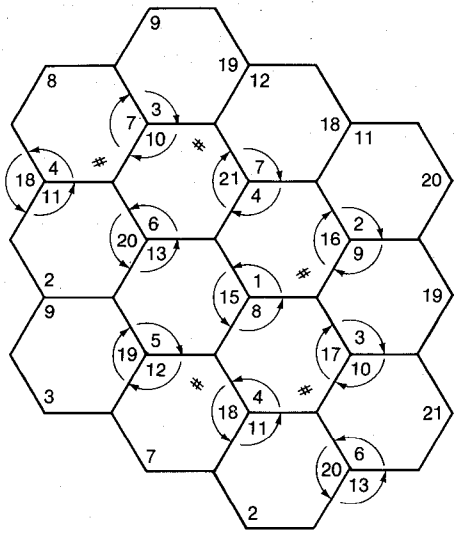


Figure 8.5 Channel-sharing algorithm.

possible interference area

In channel-sharing systems, the channel combiner should be flexible in order to combine up to 32 channels in one face in real time. An alternative method is to install a standby antenna.

Channel borrowing. Channel borrowing is usually handled on a long-term basis. The extent of borrowing more available channels from other cells depends on the traffic density in the area. Channel borrowing can be implemented from one cell-site face to another face at the same cell site.

In addition, the central cell site can borrow channels from neighboring cells. The channel-borrowing scheme is used primarily for slowly-growing systems. It is often helpful in delaying cell splitting in peak traffic areas. Since cell splitting is costly, it should be implemented only as a last resort.

8.5.3 Sectorization

The total number of available channels can be divided into sets (subgroups) depending on the sectorization of the cell configuration: the 120°-sector system, the 60°-sector system, and the 45°-sector system. A seven-cell system usually uses three 120° sectors per cell, with the total number of channel sets being 21. In certain locations and special situations, the sector angle can be reduced (narrowed) in order to assign more channels in one sector without increasing neighboring-channel interference. This point is discussed in Sec. 10.6. Sectorization

serves the same purpose as the channel-borrowing scheme in delaying cell splitting. In addition, channel coordination to avoid cochannel interference is much easier in sectorization than in cell splitting. Given the same number of channels, trunking efficiency decreases in sectorization.

Comparison of omnicells (nonsectorized cells) and sectorized cells

Omnicells. If a $K = 7$ frequency-reuse pattern is used, the frequency sets assigned in each cell can be followed by the frequency-management chart shown in Fig. 8.1. However, terrain is seldom flat; therefore, $K = 12$ is sometimes needed for reducing cochannel interference. For $K = 12$, the channel-reuse distance is $D = 6R$, or the cochannel reduction factor $q = 6$.

Sectorized cells. There are three basic types.

1. The 120°-sector cell is used for both transmitting and receiving sectorization. Each sector has an assigned a number of frequencies. Changing sectors during a call requires handoffs.
2. The 60°-sector cell is used for both transmitting and receiving sectorization. Changing sectors during a call requires handoffs. More handoffs are expected for a 60° sector than a 120° sector in areas close to cell sites (close-in areas).
3. The 120°- or 60°-sector cell is used for receiving sectorization only. In this case, the transmitting antenna is omnidirectional. The number of channels in this cell is not subdivided for each sector. Therefore, no handoffs are required when changing sectors. This receiving-sectorization-only configuration does not decrease interference or increase the D/R ratio; it only allows for a more accurate decision regarding handing off the calls to neighboring cells.

8.5.4 Underlay-overlay arrangement[1]

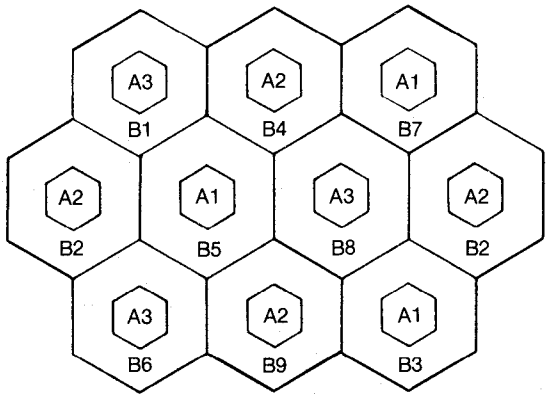
In actual cellular systems cell grids are seldom uniform because of varying traffic conditions in different areas and cell-site locations.

Overlaid cells. To permit the two groups to reuse the channels in two different cell-reuse patterns of the same size, an "underlaid" small cell is sometimes established at the same cell site as the large cell (see Fig. 8.3). The "doughnut" (large) and "hole" (small) cells are treated as two different cells. They are usually considered as "neighboring cells."

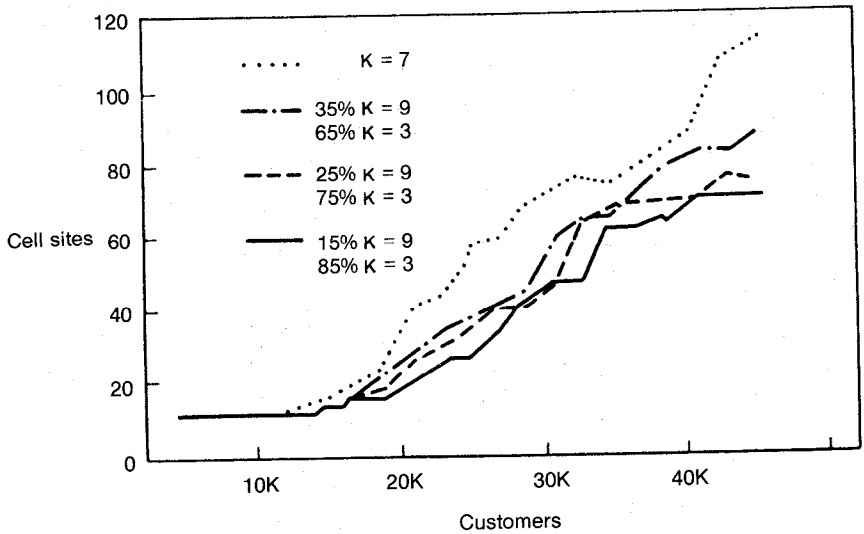
The use of either an omnidirectional antenna at one site to create two subring areas or three directional antennas to create six subareas is illustrated in Fig. 8.3*b*. As seen in Fig. 8.3, a set of frequencies used in an overlay area will differ from a set of frequencies used in an

underlay area in order to avoid adjacent-channel and cochannel interference. The channels assigned to one combiner—say, 16 channels—can be used for overlay, and another combiner can be used for underlay.

Implementation. The antenna of a set-up channel is usually omnidirectional. When an incoming call is received by the set-up channel and its signal strength is higher than a level L, the underlaid cell is



(a)



(b)

Figure 8.6 Reuse-partition scheme. (*After Whitehead, Ref. 1.*) (*a*) Reuse partition $K_A = 3$; $K_B = 9$. (*b*) Reuse-partitioning performance.

assigned; otherwise, the overlaid cell is assigned. The handoffs are implemented between the underlaid and overlaid cells. In order to avoid the unnecessary handoffs, we may choose two levels L_1 and L_2 and $L_1 > L_2$ as shown in Fig. 8.3*c*.

When a mobile signal is higher than a level L_1 the call is handed off to the underlaid cell. When a signal is lower than a level L_2 the call is handed off to the overlaid cell. The channels assigned in the underlaid cell have more protection against cochannel interference.

Reuse partition. Through implementation of the overlaid-cell concept, one possible operation is to apply a multiple-K system operation, where K is the number of frequency-reuse cells. The conventional system uses $K = 7$. But if one K is used for the underlaid cells, then this multiple-K system can have an additional 20 percent more spectrum efficiency than the single K system with an equivalent voice quality. In Fig. 8.6*a*, the $K = 9$ pattern is assigned to overlaid cells and the $K = 3$ pattern is assigned to underlaid cells. Based on this arrangement the number of cell sites can be reduced, while maintaining the same traffic capacity. The decrease in the number of cell sites which results from implementation of the multiple K systems is shown in Fig. 8.6*b*. The advantages of using this partition based on the range of K are

1. The K range is 3 to 9; the operational call quality can be adjusted and more reuse patterns are available if needed.
2. Each channel set of old $K = 9$ systems is the subset of new $K = 3$ systems. Therefore, the amount of radio retuning in each cell in this arrangement is minimal.
3. When cell splitting is implemented, all present channel assignments can be retained.

8.6 Nonfixed Channel Assignment Algorithms[4–9]

8.6.1 Description of different algorithms

Fixed channel algorithm. The fixed channel assignment (FCA) algorithm is the most common algorithm adopted in many cellular systems. In this algorithm, each cell assigns its own radio channels to the vehicles within its cell.

Dynamic channel assignment. In dynamic channel assignment (DCA), no fixed channels are assigned to each cell. Therefore, any channel in a composite of 312 radio channels can be assigned to the mobile unit.

This means that a channel is assigned directly to a mobile unit. On the basis of overall system performance, DCA can also be used during a call.

Hybrid channel assignment. Hybrid channel assignment (HCA) is a combination of FCA and DCA. A portion of the total frequency channels will use FCA and the rest will use DCA.

Borrowing channel assignment. Borrowing channel assignment (BCA) uses FCA as a normal assignment condition. When all the fixed channels are occupied, then the cell borrows channels from the neighboring cells.

Forcible-borrowing channel assignment.[9] In forcible-borrowing channel assignment (FBCA), if a channel is in operation and the situation warrants it, channels must be borrowed from the neighboring cells and at the same time, another voice channel will be assigned to continue the call in the neighboring cell.

There are many different ways of implementing FBCA. In a general sense, FBCA can also be applied while accounting for the forcible borrowing of the channels within a fixed channel set to reduce the chance of cochannel assignment in a reuse cell pattern.

The FBCA algorithm is based on assigning a channel dynamically but obeying the rule of reuse distance. The distance between the two cells is *reuse distance,* which is the minimum distance at which no cochannel interference would occur.

Very infrequently, no channel can be borrowed in the neighboring cells. Even those channels currently in operation can be forcibly borrowed and will be replaced by a new channel in the neighboring cell or the neighboring cell of the neighboring cell. If all the channels in the neighboring cells cannot be borrowed because of interference problems, the FBCA stops.

8.6.2 Simulation process and results

On the basis of the FBCA, FCA, and BCA algorithms, a seven-cell reuse pattern with an average blocking of 3 percent is assumed and the total traffic service in an area is 250 erlangs. The traffic distributions are (1) uniform traffic distribution—11 channels per cell; (2) a nonuniform traffic distribution—the number of channels in each cell is dependent on the vehicle distribution (Fig. 8.7). The simulation model is described as follows:

1. Randomly select the cell (among 41 cells).

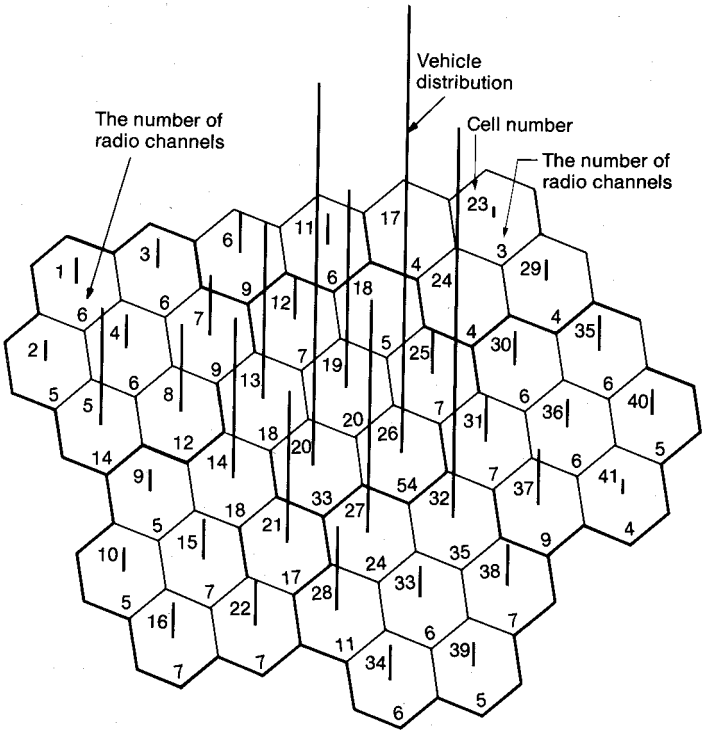


Figure 8.7 Cellular system. Vehicle and radio-channel distribution in the busy rush hour. (*After Sekiguchi et al., Ref. 9.*)

2. Determine the state of the vehicle in the cell (idle, off-hook, on-hook, handoff).
3. In off-hook or handoff states, search for an idle channel. The average number of handoffs is assumed to be 0.2 times per call. However, FBCA will increase the number of handoffs.

Average blocking. Two average blocking cases illustrating this simulation are shown in Fig. 8.8. In a uniform traffic condition (Fig. 8.8*a*), the 3 percent blocking of both BCA and FBCA will result in a load increase of 28 percent, compared to 3 percent blocking of FCA. There is no difference between BCA and FBCA when a uniform traffic condition exists.

In a nonuniform traffic distribution (Fig. 8.8*b*), the load increase in BCA drops to 23 percent and that of FBCA increases to 33 percent, as at an average blocking of 3 percent. The load increase can be utilized in another way by reducing the number of channels. The percent increase in load is the same as the percent reduction in the number of channels.

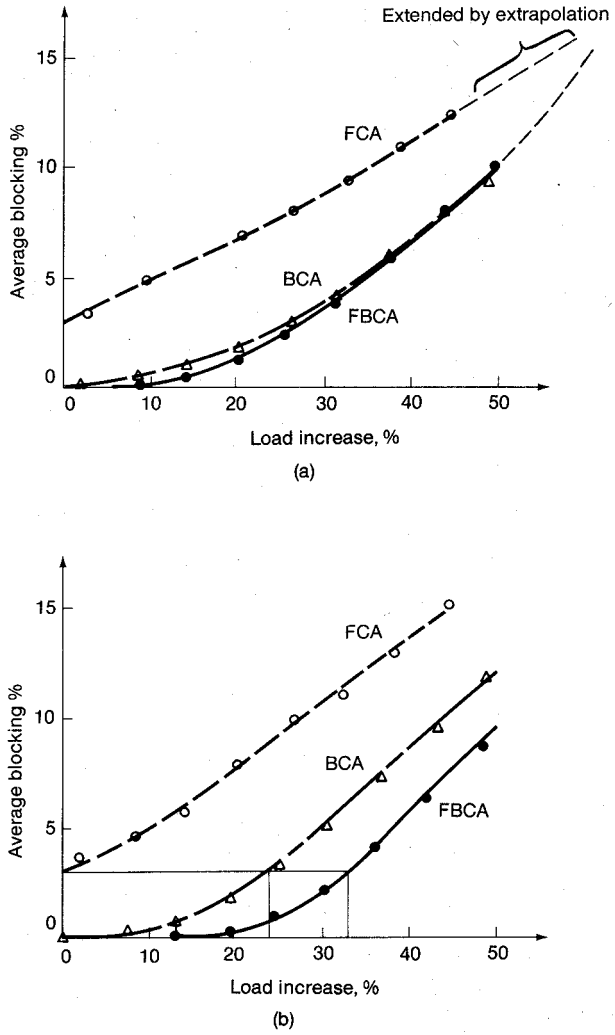


Figure 8.8 Comparison of average blockings from three different schemes. (*After Sekiguchi et al., Ref. 9.*) (*a*) Average blocking in spatially uniform traffic distribution; (*b*) average blocking in spatially nonuniform traffic distribution.

Handoff blocking. Blocking calls from all handoff calls occurring in all cells is shown in Fig. 8.9. In both BCA and FBCA, load is increased almost equally to 30 percent, as compared to FCA at 3 percent handoff blocking in uniform traffic (Fig. 8.9*a*). For a nonuniform traffic distribution, the load increase of both BCA and FBCA at 4 percent blocking is about 50 percent (Fig. 8.9*b*), which is a big improvement, considering

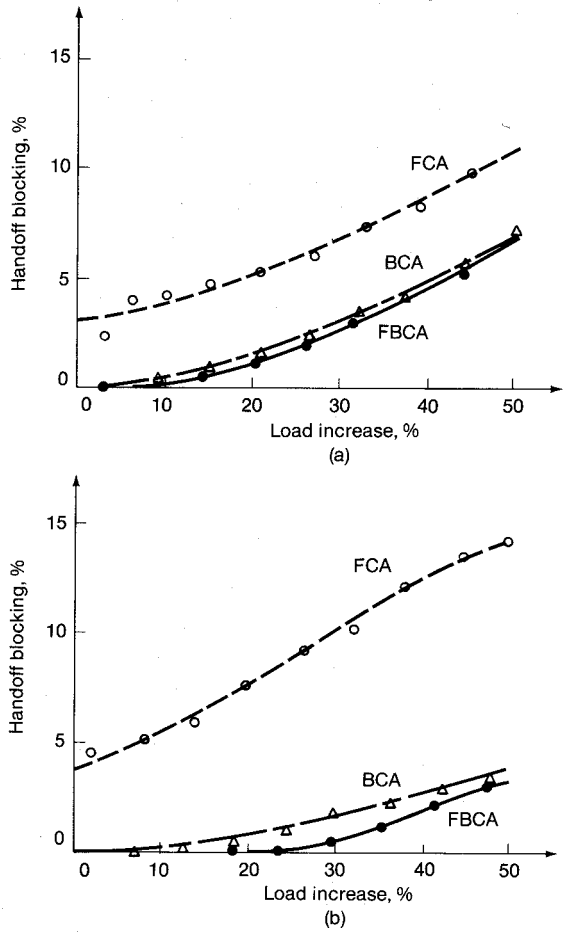


Figure 8.9 Comparison of handoff blocking from three different schemes. (*After Sekiguchi et al., Ref. 9.*) (*a*) Handoff blocking in spatially uniform traffic distribution; (*b*) handoff blocking in spatially nonuniform traffic distribution.

the reduction in interference and blocking. Otherwise, there would be multiple effects from interference in several neighboring cells.

8.7 How to Operate with Additional Spectrum

On July 24, 1986 the FCC announced that a totally new additional spectrum of 10 MHz would be allocated to the cellular mobile industry. This spectrum provides 166 voice channels, with 83 channels for each carrier. The new spectrum allocation is shown in Fig. 8.2.

In the future, cellular systems must serve both the old mobile units, which operate 666 channels, and the new mobile units, which operate 832 channels. The new mobile units will have less blocked calls than the old mobile units when they are used in areas of heavy traffic. However, because the additional spectra for bands A and B are discretely and alternately allocated, the neighboring channels between bands A and B occur at two points (one point between channels 666 and 667, and the other between channels 716 and 717) in the frequency spectrum. At these two points, the tendency for neighboring-channel interference is high.

According to the analysis given in Sec. 7.3.1, the "neighboring channels" can consist of four channels on each side of two systems. Therefore, these eight channels must be used with extreme caution. Unless we know the frequency channel assignments of the other system, or coordinate with the other system, it is not wise to use these channels.

The ratio of the new additional spectrum to the present spectrum is 5/20 MHz = 25 percent, which means that the effective increase in the spectrum is 25 percent if we can fully use it.

The new additional spectrum utilization factor η at any given period of time can be calculated from

$$\eta = \frac{B}{A + B}$$

where A is the number of customers who are using old mobile units and B is the number of customers using new mobile units. If B is increasing very slowly, then η can be very small. This would defeat the purpose of implementing the new additional spectrum. Therefore, the new mobile units should outdate the old mobile units such that A remains the same and B is increasing. Assume that the number of new subscribers per year is

$$\frac{B}{A} = \frac{1}{10}$$

Then the spectrum-utilization factor for the first year that the new system is implemented would be

$$\eta = \frac{0.10A}{A + 0.10A} = 9\%$$

Then for the second year, the B/A ratio would be

$$\frac{B}{A} = \frac{1}{5}$$

and the spectrum-utilization factor η would be

$$\eta = \frac{0.2A}{A + 0.2A} = 17\%$$

These calculations are based on the assumption that new mobile units are assigned only to new additional channels so that the traffic capacity using the old spectrum will not worsen. After η exceeds 20 percent, the new mobile units have to be assigned to all the 395 voice channels. Implementation of the new additional spectrum is discussed further in Chap. 10.

8.8 Traffic and Channel Assignment

The vehicular traffic density of a coverage area is a critical element and must be determined before a system is designed. This traffic pattern in busy hours can be confined to different zones within the service area. This traffic-density information should be converted to the number of cars per 1000- × 1000-ft grid (or 2000- × 2000-ft grid) and stored in the grids of the contour map provided in Sec. 4.7.

If the traffic pattern predominates over the simple signal coverage pattern, cell-site selection will be based on the traffic pattern.

Choice of the initial cell sites should be based on the signal covered in zones of heavy vehicular traffic. This means that the cell site would most likely be located at the center of those zones.

After call traffic data are collected while the system is operating, we can update the call traffic data at each cell site to correlate with the vehicular traffic data. This information will be useful for determining whether new cell splitting is needed. If it is, then we must determine

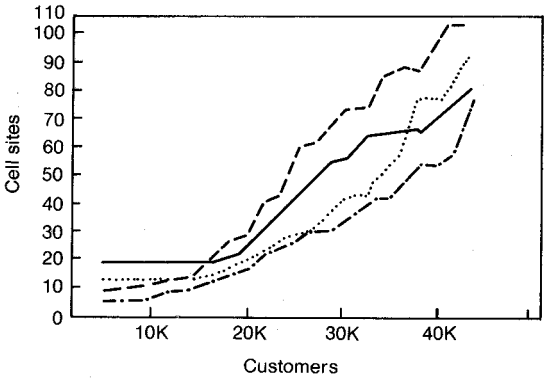


Figure 8.10 City-to-city variation. (*After Whitehead, Ref. 1.*)

how many radios should be installed at the new site and where it is to be located. These decisions are all related to frequency channel assignment. A typical chart illustrating the variation from city to city is shown in Fig. 8.10. A city may have twice as many cell sites to handle the same number of customers in the busy hours. This means that the number of cars per unit area is much higher in one city than that in the other city. Many techniques for implementing the high-capacity cellular systems are discussed in Chap. 10.

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Chapter

9

Handoffs

9.1 Value of Implementing Handoffs

9.1.1 Why handoffs[1–5]

Once a call is established, the set-up channel is not used again during the call period. Therefore, handoff is always implemented on the voice channel. The value of implementing handoffs is dependent on the size of the cell. For example, if the radius of the cell is 32 km (20 mi), the area is 3217 km^2 (1256 mi^2). After a call is initiated in this area, there is little chance that it will be dropped before the call is terminated as a result of a weak signal at the coverage boundary. Then why bother to implement the handoff feature? Even for a 16-km radius cell handoff may not be needed. If a call is dropped in a fringe area, the customer simply redials and reconnects the call.

Handoff is needed in two situations where the cell site receives weak signals from the mobile unit: (1) at the cell boundary, say, −100 dBm, which is the level for requesting a handoff in a noise-limited environment; and (2) when the mobile unit is reaching the signal-strength holes (gaps) within the cell site as shown in Fig. 9.1.

9.1.2 Two types of handoff

There are two types of handoff: (1) that based on signal strength and (2) that based on carrier-to-interference ratio. The handoff criteria are

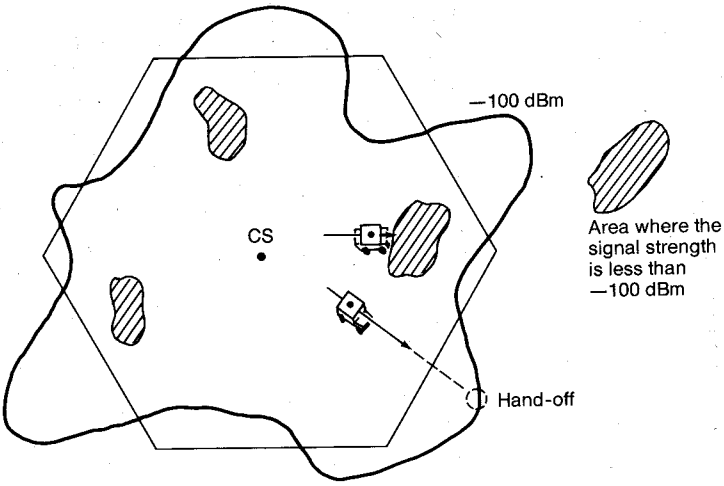

Figure 9.1 Occurrence of handoff.

different for these two types. In type 1, the signal-strength threshold level for handoff is -100 dBm in noise-limited systems and -95 dBm in interference-limited systems. In type 2, the value of C/I at the cell boundary for handoff should be 18 dB in order to have toll quality voice. Sometimes, a low value of C/I may be used for capacity reasons.

Type 1 is easy to implement. The location receiver at each cell site measures all the signal strengths of all receivers at the cell site. However, the received signal strength (RSS) itself includes interference.

$$\text{RSS} = C + I \tag{9.1-1}$$

where C is the carrier signal power and I is the interference. Suppose that we set up a threshold level for RSS; then, because of the I, which is sometimes very strong, the RSS level is higher and far above the handoff threshold level. In this situation handoff should theoretically take place but does not. Another situation is when I is very low but RSS is also low. In this situation, the voice quality usually is good even though the RSS level is low, but since RSS is low, unnecessary handoff takes place. Therefore it is an easy but not very accurate method of determining handoffs. Some systems use SAT information together with the received signal level to determine handoffs (Sec. 13.1.2).

Handoffs can be controlled by using the carrier-to-interference ratio C/I, which can be obtained as described in Sec. 6.3.

$$\frac{C + I}{I} \approx \frac{C}{I} \tag{9.1-2}$$

In Eq. (9.1-2), we can set a level based on C/I, so C drops as a function of distance but I is dependent on the location. If the handoff is dependent

on C/I, and if the C/I drops, it does so in response to increase in (1) propagation distance or (2) interference. In both cases, handoff should take place. In today's cellular systems, it is hard to measure C/I during a call because of analog modulation. Sometimes we measure the level I before the call is connected, and the level $C + I$ during the call. Thus $(C + I)/I$ can be obtained. Another method of measuring C/I is described in Sec. 6.3.

9.1.3 Determining the probability of requirement for handoffs[6]

To find the probability of requiring a handoff, we can carry out the following simulation. Suppose that a mobile unit randomly initiates a call in a 16-km (10-mi) cell. The vehicle speed is also randomly chosen between 8 and 96 km/h (5 to 60 mi/h). The direction is randomly chosen to be between 0 and 360°; then the chance of reaching the boundary is dependent on the call holding time.

Figure 9.2 depicts the probability curve for requiring handoff. Table 9.1 summarizes the results. If the call holding time is 1.76 min, the

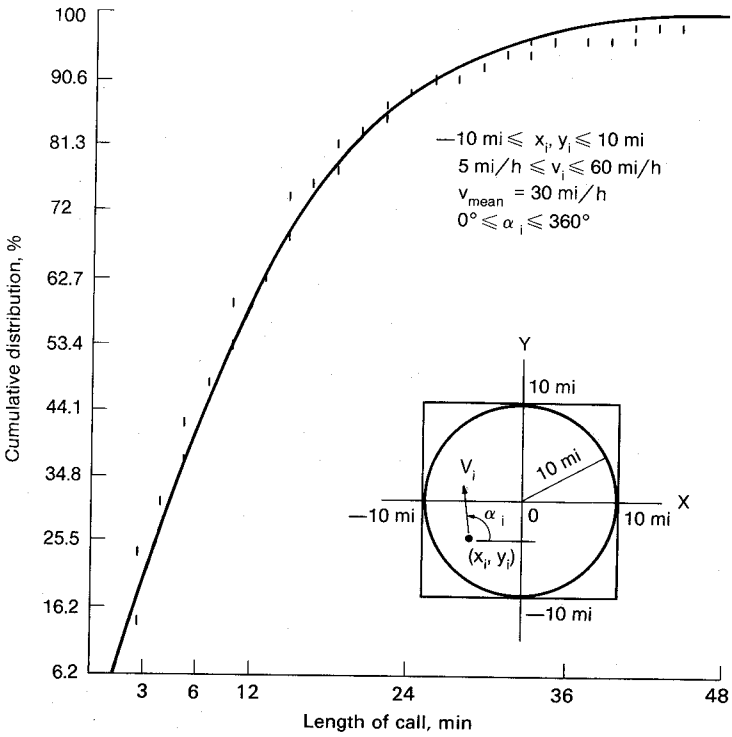

Figure 9.2 The probability of requiring handoff.

TABLE 9.1 Probability of Having a Handoff in a 10-mi Coverage Area

Handoff probability, %	Call length, min
11.3	1.76
18.	3
42.6	6
59.3	9

only chance of reaching the boundary is 11 percent, or the chance that a handoff will occur for the call is 11 percent. If the call holding time is 3 min, the chance of reaching the boundary is 18 percent. Now we may debate whether a handoff is needed or not. In rural areas, handoffs may not be necessary. However, commercial mobile units must meet certain requirements, and handoffs may be necessary at times. Military mobile systems may opt not to use the handoff feature and may apply the savings in cost to implement other security measures.

9.1.4 Number of handoffs per call

The smaller the cell size, the greater the number and the value of implementing handoffs. The number of handoffs per call is relative to cell size. From the simulation, we may find that

0.2 handoff per call in a 16- to 24-km cell

1–2 handoffs per call in a 3.2- to 8-km cell

3–4 handoffs per call in a 1.6- to 3.2-km cell

9.2 Initiation of a Handoff

At the cell site, signal strength is always monitored from a reverse voice channel. When the signal strength reaches the level of a handoff (higher than the threshold level for the minimum required voice quality), then the cell site sends a request to the mobile telephone switching level (MTSO) for a handoff on the call. An intelligent decision can also be made at the cell site as to whether the handoff should have taken place earlier or later. If an unnecessary handoff is requested, then the decision was made too early. If a failure handoff occurs, then a decision was made too late.

The following approaches are used to make handoffs successful and to eliminate all unnecessary handoffs. Suppose that -100 dBm is a threshold level at the cell boundary at which a handoff would be taken. Given this scenario, we must set up a level higher than -100 dBm—say, -100 dBm $+ \Delta$ dB—and when the received signal reaches this

level, a handoff request is initiated. If the value of Δ is fixed and large, then the time it takes to lower -100 dBm $+ \Delta$ to -100 dBm is longer. During this time, many situations, such as the mobile unit turning back toward the cell site or stopping, can occur as a result of the direction and the speed of the moving vehicles. Then the signals will never drop below -100 dBm. Thus, many unnecessary handoffs may occur simply because we have taken the action too early. If Δ is small, then there is not enough time for the call to hand off at the cell site and many calls can be lost while they are handed off. Therefore, Δ should be varied according to the path-loss slope of the received signal strength (Sec. 4.2) and the level-crossing rate (LCR) of the signal strength (Sec. 1.63) as shown in Fig. 9.3.

Let the value of Δ be 10 dB in the example given in the preceding paragraph. This would mean a level of -90 dBm as the threshold level for requesting a handoff. Then we can calculate the velocity V of the mobile unit based on the predicted LCR[7] at a -10-dB level with respect to the root-mean-square (rms) level, which is at -90 dBm; thus

$$V = \begin{cases} \dfrac{n\lambda}{\sqrt{2\pi}\,(0.27)} & \text{ft/s} \\ n\lambda & \text{mi/h} \end{cases} \qquad \text{at } -10\text{-dB level} \qquad (9.2\text{-}1)$$

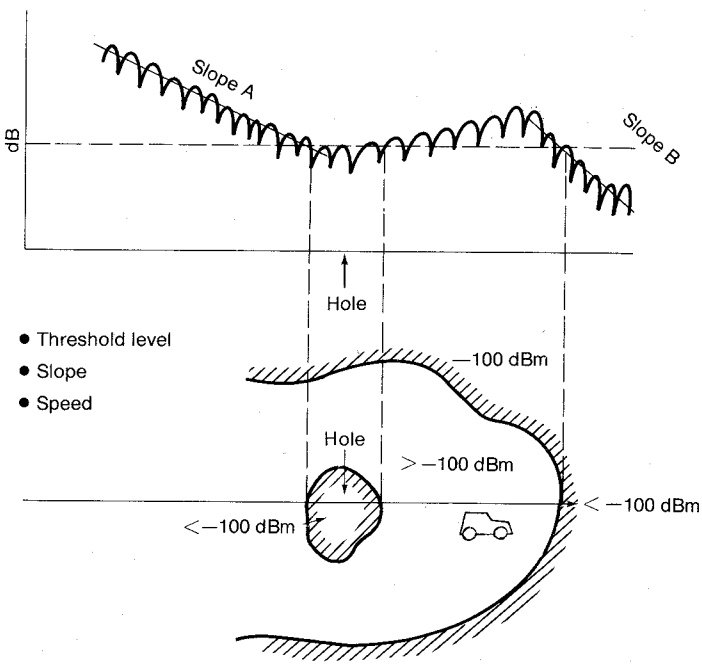

Figure 9.3 Parameters for handling a handoff.

where n is the LCR (crossings per second) counting positive slopes and λ is the wavelength in feet. Equation (9.2-1) can be simplified as

$$V(\text{mi/h}) \approx n(\text{crossings/s}) \text{ at 850 MHz and a } -10\text{-dB level} \quad (9.2\text{-}2)$$

Here, two pieces of information, the velocity of vehicle V and the path-loss slope γ, can be used to determine the value of Δ dynamically so that the number of unnecessary handoffs can be reduced and the required handoffs can be completed successfully.

There are two circumstances where handoffs are necessary but cannot be made: (1) when the mobile unit is located at a signal-strength hole within a cell but not at the boundary (see Fig. 9.3) and (2) when the mobile unit approaches a cell boundary but no channels in the new cell are available.

In case 1, the call must be kept in the old frequency channel until it is dropped as the result of an unacceptable signal level. In case 2, the new cell must reassign one of its frequency channels within a reasonably short period or the call will be dropped.

The MTSO usually controls the frequency assignment in each cell and can rearrange channel assignments or split cells when they are necessary. Cell splitting is described in Sec. 10.4.

9.3 Delaying a Handoff

9.3.1 Two-handoff-level algorithm

In many cases, a two-handoff-level algorithm is used. The purpose of creating two request handoff levels is to provide more opportunity for a successful handoff. A handoff could be delayed if no available cell could take the call.

A plot of signal strength with two request handoff levels and a threshold level is shown in Fig. 9.4. The plot of average signal strength is recorded on the channel received signal-strength indicator (RSSI) which is installed at each channel receiver at the cell site. When the signal strength drops below the first handoff level, a handoff request is initiated. If for some reason the mobile unit is in a hole (a weak spot in a cell) or a neighboring cell is busy, the handoff will be requested periodically every 5 s. At the first handoff level, the handoff takes place if the new signal is stronger (see case I in Fig. 9.4). However, when the second handoff level is reached, the call will be handed off with no condition (see case II in Fig. 9.4).

The MTSO always handles the handoff call first and the originating calls second. If no neighboring cells are available after the second handoff level is reached, the call continues until the signal strength drops below the threshold level; then the call is dropped. If the super-

At cell site

- Each channel receiver has a RSSI (received signal strength indicator)
- Two-level handoff algorithm

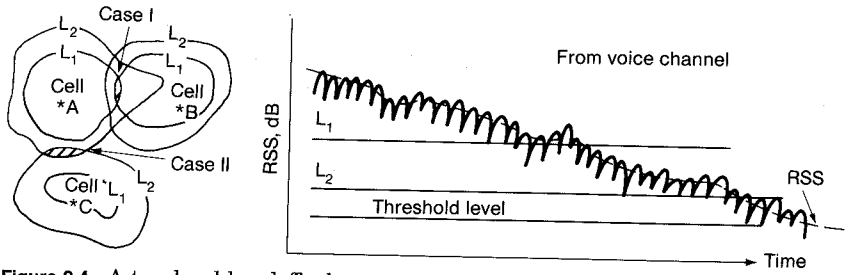

Figure 9.4 A two-level handoff scheme.

visory audio tone (SAT) is not sent back to the cell site by the mobile unit within 5 s, the cell site turns off the transmitter.

9.3.2 Advantage of delayed handoffs

Consider the following example. The mobile units are moving randomly and the terrain contour is uneven. The received signal strength at the mobile unit fluctuates up and down. If the mobile unit is in a hole for less than 5 s (a driven distance of 140 m for 5 s, assuming a vehicle speed of 100 km/h, the delay (in handoff) can even circumvent the need for a handoff.

If the neighboring cells are busy, delayed handoff may take place. In principle, when call traffic is heavy, the switching processor is loaded, and thus a lower number of handoffs would help the processor handle call processing more adequately. Of course, it is very likely that after the second handoff level is reached, the call may be dropped with great probability.

The other advantage of having a two-handoff-level algorithm is that it makes the handoff occur at the proper location and eliminates possible interference in the system. Figure 9.4, case I, shows the area where the first-level handoff occurs between cell A and cell B. If we only use the second-level handoff boundary of cell A, the area of handoff is too close to cell B. Figure 9.4, case II, also shows where the second-level handoff occurs between cell B and cell C. This is because the first-level handoff cannot be implemented.

9.4 Forced Handoffs

A *forced handoff* is defined as a handoff which would normally occur but is prevented from happening, or a handoff that should not occur but is forced to happen.

9.4.1 Controlling a handoff

The cell site can assign a low handoff threshold in a cell to keep a mobile unit in a cell longer or assign a high handoff threshold level to request a handoff earlier. The MTSO also can control a handoff by making either a handoff earlier or later, after receiving a handoff request from a cell site.

9.4.2 Creating a handoff

In this case, the cell site does not request a handoff but the MTSO finds that some cells are too congested while others are not. Then the MTSO can request cell sites to create early handoffs for those congested cells. In other words, a cell site has to follow the MTSO's order and increase the handoff threshold to push the mobile units at the new boundary and to hand off earlier.

9.5 Queuing of Handoffs

Queuing of handoffs is more effective than two-threshold-level handoffs. The MTSO will queue the requests of handoff calls instead of rejecting them if the new cell sites are busy. A queuing scheme becomes effective only when the requests for handoffs arrive at the MTSO in batches or bundles. If handoff requests arrive at the MTSO uniformly, then the queuing scheme is not needed. Before showing the equations, let us define the parameters as follows.

$1/\mu$	average calling time in seconds, including *new calls* and *handoff calls* in each cell
λ_1	arrival rate (λ_1 calls per second) for originating calls
λ_2	arrival rate (λ_2 handoff calls per second) for handoff calls
M_1	size of queue for originating calls
N	number of voice channels
a	$(\lambda_1 + \lambda_2)/\mu$
b_1	λ_1/μ
b_2	λ_2/μ

The following analysis can be used to see the improvement. We are analyzing three cases.[8]

1. *No queuing on either the originating calls or the handoff calls.* The blocking for either a originating call or a handoff call is

$$B_o = \frac{a^N}{N!} P(0) \tag{9.5-1}$$

where

$$P(0) = \left(\sum_{n=0}^{N} \frac{a^N}{n!} \right)^{-1} \tag{9.5-2}$$

2. *Queuing the originating calls but not the handoff calls.* The blocking probability for originating calls is

$$B_{oq} = \left(\frac{b_1}{N} \right)^{M_1} P_q(0) \tag{9.5-3}$$

where

$$P_q(0) = \left[N! \sum_{n=0}^{N-1} \frac{a^{n-N}}{n!} + \frac{1 - (b_1/N)^{M_1+1}}{1 - (b_1/N)} \right]^{-1} \tag{9.5-4}$$

The blocking probability for handoff calls is

$$B_{oh} = \frac{1 - (b_1/N)^{M_1+1}}{1 - (b_1/N)} P_q(0) \tag{9.5-5}$$

3. *Queuing the handoff calls but not the originating calls.* The blocking

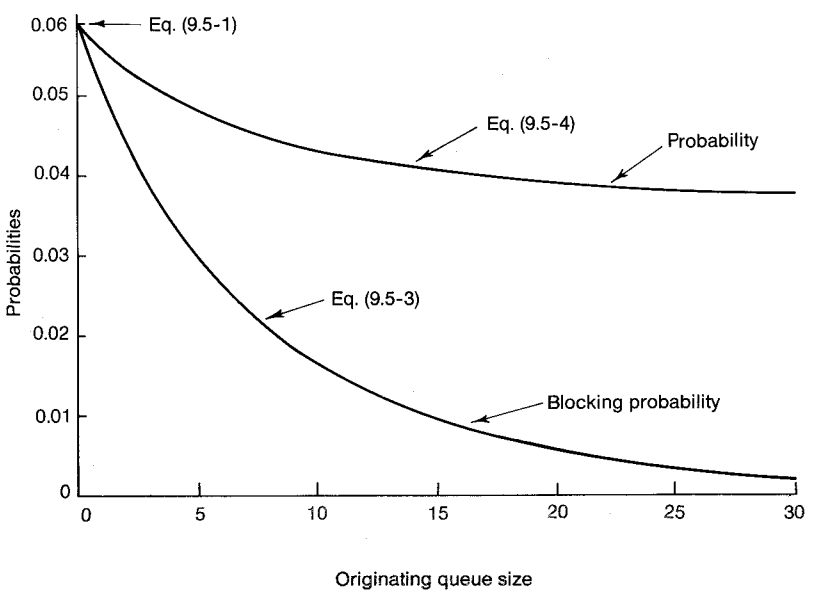

Figure 9.5 Probability and blocking probability graph showing blocking probability for originating calls queuing for originating calls ($N = 70$).

probability for handoff calls is

$$B_{hq} = \left(\frac{b_2}{N}\right)^{M_2} P_q(0) \tag{9.5-6}$$

where $P_q(0)$ is as shown in Eq. (9.5-4). The blocking probability for originating calls is

$$B_{ho} = \frac{1 - (b_2/N)^{M_2+1}}{1 - (b_2/N)} P_q(0) \tag{9.5-7}$$

Example 9.1 The following parameters are given. The number of channels at the cell site $N = 70$. The call holding time is 101 s = 0.028 h. The number of originating calls attempted per hour is expressed as $\lambda_1 = 2270$. The number of handoff calls attempted per hour is expressed as $\lambda_2 = 80$. Then

$$A = \frac{\lambda + \lambda_2}{\mu} = (2270 + 80)\,0.028 \doteq 65.80$$

$$b_1 = \frac{\lambda_1}{\mu} = 2270 \times 0.028 = 63.60$$

$$b_2 = \frac{\lambda_2}{\mu} = 2.24$$

Given these parameters, Eqs. (9.5-1), (9.5-3), (9.5-5), (9.5-6), and (9.5-7) have been plotted in Figs. 9.5, 9.6, 9.7, and 9.8 respectively.

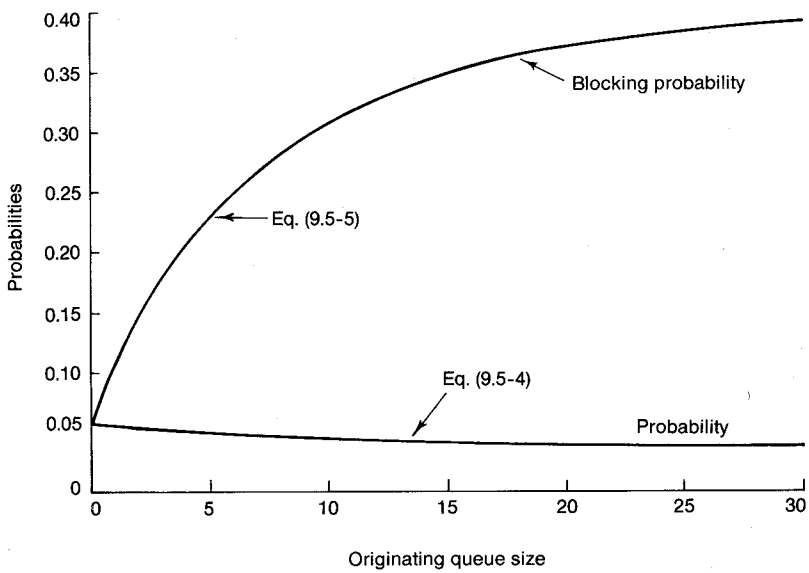

Figure 9.6 Probability and blocking probability graph showing blocking probability for handoff calls queuing for originating calls ($N = 70$).

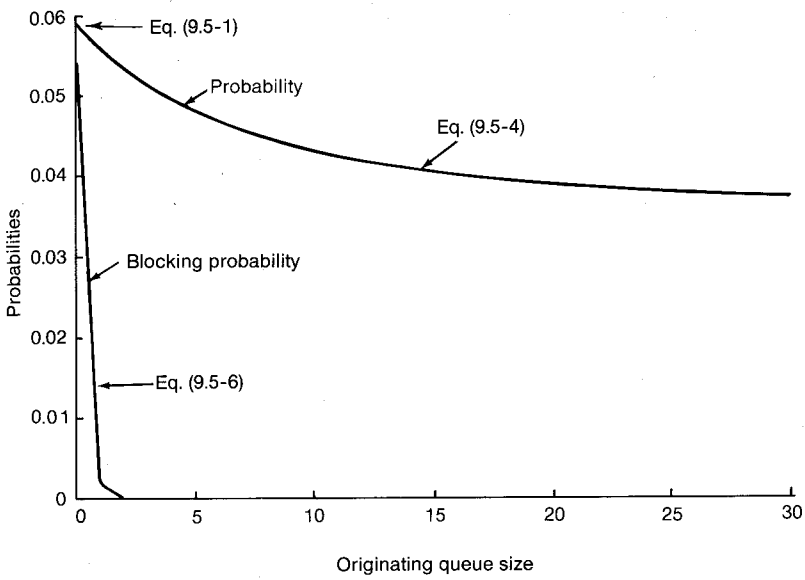

Figure 9.7 Probability and blocking probability graph showing blocking probability for handoff calls (queuing for handoff calls) ($N = 70$).

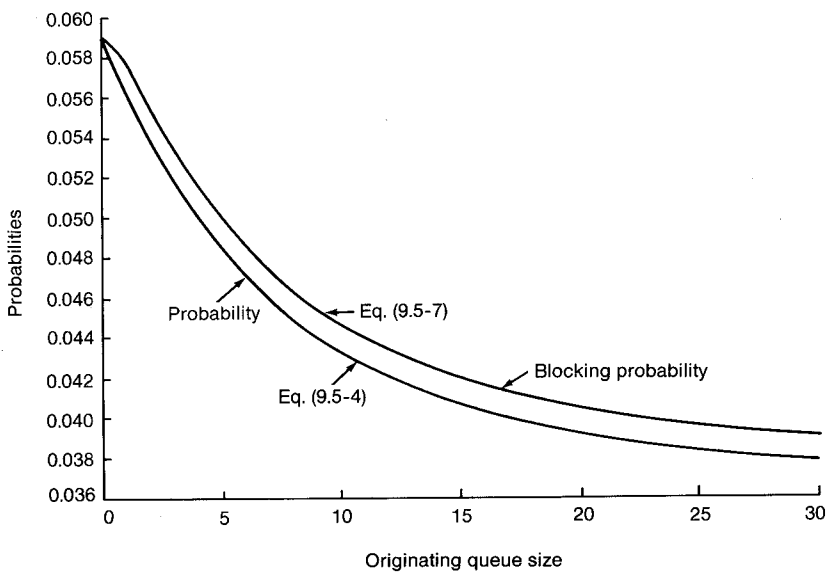

Figure 9.8 Probability and blocking probability graph showing blocking probability for originating calls (queuing for handoff calls) ($N = 70$).

We have seen (Figs. 9.5 and 9.6) with queuing of originating calls only, the probability of blocking is reduced. However, queuing of originating calls results in increased blocking probability on handoff calls, and this is a drawback. With queuing of handoff calls only, blocking probability is reduced from 5.9 to 0.1 percent by using one queue space (see Fig. 9.7). Therefore it is very worthwhile to implement a simple queue (one space) for handoff calls. Adding queues in handoff calls does not affect the blocking probability of originating calls in this particular example (see Fig. 9.8). However, we should always be aware that queuing for the handoff is more important than queuing for those initiating calls on assigned voice channels because call drops upset customers more than call blockings.

9.6 Power-Difference Handoffs

A better algorithm is based on the power difference (Δ) of a mobile signal received by two cell sites, home and handoff. Δ can be positive or negative. The handoff occurs depending on a preset value of Δ.

$$\Delta = \text{the mobile signal measured at the candidate handoff site} - \text{the mobile signal measured at the home site} \quad (9.6\text{-}1)$$

For example, the following cases can occur.

$\Delta > 3$ dB	request a handoff
$1 \text{ dB} < \Delta < 3$ dB	prepare a handoff
$-3 \text{ dB} < \Delta < 0$ dB	monitoring the signal strength
$\Delta < -3$ dB	no handoff

Those numbers can be changed to fit the switch processor capacity. This algorithm is not based on the received signal strength level, but on a relative (power difference) measurement. Therefore, when this algorithm is used, all the call handoffs for different vehicles can occur at the same general location in spite of different mobile antenna gains or heights.

9.7 Cell-Site Handoff Only

This scheme can be used in a noncellular system. The mobile unit has been assigned a frequency and talks to its home cell site while it travels. When the mobile unit leaves its home cell and enters a new cell, its frequency does not change; rather, the new cell must tune into the frequency of the mobile unit (see Fig. 9.9). In this case only the cell sites need the frequency information of the mobile unit. Then the as-

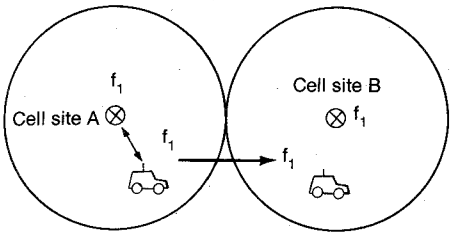

Figure 9.9 Cell-site handoff-only scheme.

pects of mobile unit control can be greatly simplified, and there will be no need to provide handoff capability at the mobile unit. The cost will also be lower.

This scheme can be recommended only in areas of very low traffic. When the traffic is dense, frequency coordination is necessary for the cellular system. Then if a mobile unit does not change frequency on travel from cell to cell, other mobile units then must change frequency to avoid interference.

Therefore, if a system handles only low volumes of traffic, that is, if the channels assigned to one cell will not reuse frequency in other cells, then it is possible to implement the cell-site handoff feature as it is applied in military systems.

9.8 Intersystem Handoff

Occasionally a call may be initiated in one cellular system (controlled by one MTSO) and enter another system (controlled by another MTSO) before terminating. In some instances, *intersystem handoff* can take place; this means that a call handoff can be transferred from one system to a second system so that the call can be continued while the mobile unit enters the second system.

The software in the MTSO must be modified to apply this situation. Consider the simple diagram shown in Fig. 9.10. The car travels on a highway and the driver originates a call in system A. Then the car leaves cell site A of system A and enters cell site B of system B. Cell sites A and B are controlled by two different MTSOs. When the mobile unit signal becomes weak in cell site A, MTSO A searches for a candidate cell site in its system and cannot find one. Then MTSO A sends the handoff request to MTSO B through a dedicated line between MTSO A and MTSO B, and MTSO B makes a complete handoff during the call conversation. This is just a one-point connection case. There are many ways of implementing intersystem handoffs, depending on the actual circumstances. For instance, if two MTSOs are manufactured by different companies, then compatibility must be determined before

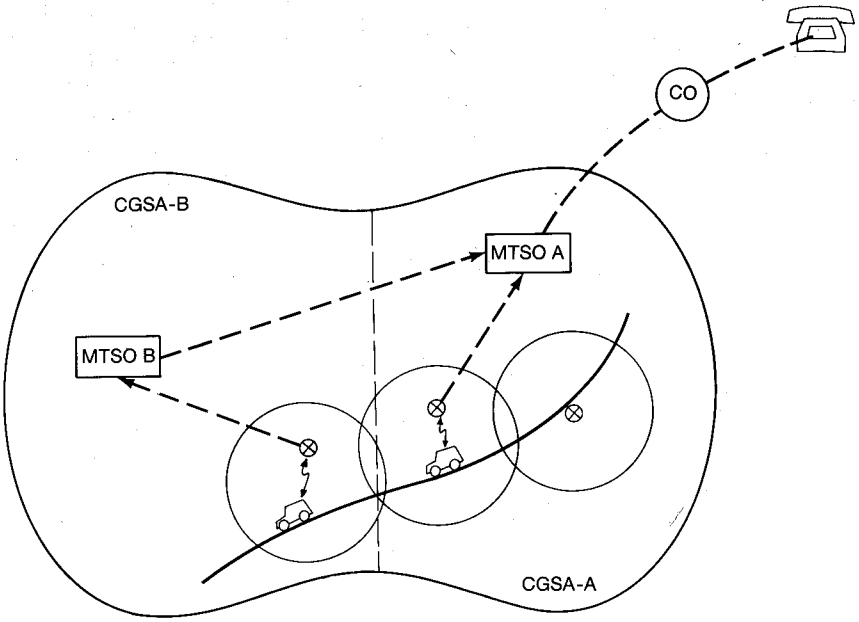

Figure 9.10 Intersystem handoffs.

implementation of intersystem handoff can be considered. A detailed discussion of this topic appears in Sec. 11.4.

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Chapter

10

Operational Techniques and Technologies

10.1 Adjusting the Parameters of a System

10.1.1 Increasing the coverage for a noise-limited system

In a noise-limited system, there is no cochannel interference or adjacent-channel interference. This means that either (1) no cochannels and adjacent channels are used in the system or (2) channel reuse distance is so large that the interference would be negligible. The following approaches are used at the cell site to increase the coverage.

Increasing the transmitted power. Usually, increasing the transmitted power of each channel results in coverage of a larger area. When the power level is doubled, the gain increases by 3 dB. Increase in covered area can be found as follows. The received power P_r can be obtained from the transmitted power P_t (see Chap. 4), where P_r is a function of the cell radius. Let the received power P_r be the power received in an original cell of a radius of r_1

$$P_{r_1} = \alpha P_{t_1} r_1^{-4} \tag{10.1-1}$$

Area covered then is

$$A_1 = \pi r_1^2$$

where α is a constant and P_{r_1} can be obtained from P_{t_1}.

Case 1. The transmitted power remains unchanged but the received power changes. If the received power is to be strong, the cell radius should be smaller. The relation is

$$\frac{P_{r_1}}{P_{r_2}} = \frac{r_1^{-4}}{r_2^{-4}} = \frac{r_2^4}{r_1^4} \tag{10.1-2}$$

or

$$r_2 = \left(\frac{P_{r_1}}{P_{r_2}}\right)^{1/4} r_1 \tag{10.1-3}$$

If $P_{r_2} = 2P_{r_1}$, and the transmitted power remains the same, the radius reduces to

$$r_2 = (0.5)^{1/4} \qquad r_1 = 0.84r_1$$

and the area reduces to

$$\frac{A_2}{A_1} = \frac{\pi r_2^2}{\pi r_1^2} = \frac{r_2^2}{r_1^2} = \frac{(0.84r_1)^2}{r_1^2} = 0.71 \tag{10.1-4}$$

Case 2. The transmitted power changes but the received power doesn't; then the 1-mi reception level changes if the transmitted power changes. From Eq. (10.1-1) we obtain

$$P_{r_1} = \alpha P_{t_1} r_1^{-4} \qquad P_{r_2} = \alpha P_{t_2} r_2^{-4}$$

In this case, since $P_{r_1} = P_{r_2}$, it follows that

$$r_2 = \left(\frac{P_{t_2}}{P_{t_1}}\right)^{1/4} r_1 \tag{10.1-5}$$

If the transmitted power P_{t_2} is 3 dB higher than P_{t_1}, then

$$r_2 = (2)^{1/4} \qquad r_1 = 1.19r_1$$

and the area increase is

$$\frac{A_2}{A_1} = \frac{r_2^2}{r_1^2} = (1.19)^2 = 1.42 \tag{10.1-6}$$

A general equation should be expressed as

$$r_2 = \left(\frac{P_{r_1}P_{t_2}}{P_{r_2}P_{t_1}}\right)^{1/4} r_1 \tag{10.1-7}$$

or

$$A_2 = \left(\frac{P_{r_1}P_{t_2}}{P_{r_2}P_{t_1}}\right)^{1/2} A_1 \tag{10.1-8}$$

Increasing cell-site antenna height. In general, the 6 dB/oct rule applies to the cell-site antenna height in a flat terrain, that is, doubling the antenna height causes a gain increase of 6 dB. If the terrain contour is hilly, then an effective antenna height should be used, depending on the location of the mobile unit. Sometimes doubling the actual antenna height results in a gain increase of less than 6 dB and sometimes more. This phenomenon was described in Chap. 4.

Using a high-gain or a directional antenna at the cell site. The gain and directivity of an antenna increase with the received level—the same effect seen with an increase of transmitted power.

Lowering the threshold level of a received signal. When the threshold level is lowered, the acceptable received power is lower and the radius of the cell increases [Eq. (10.1-3) applies]. The increase in service area due to a lower received level can be obtained from Eq. (10.1-8). Let $P_{t_2} = P_{t_1}$, and $P_{r_2} = 0.25P_{r_1}$ (i.e., -6 dB). Then $A_2 = 2A_1$. The received level is reduced by 6 dB, and the service area is doubled.

A low-noise receiver. The thermal noise kTB level (see Sec. 7.6.6) is -129 dBm. In a noise-limited environment, if the front-end noise of the receiver is low and the received power level remains the same, the carrier-to-noise ratio becomes large in comparison to a receiver with a high front-end noise. This low-noise receiver can receive a signal from a farther distance than can a high-noise receiver.

Diversity receiver. A diversity receiver is very useful in reducing the multipath fading. When the fading reduces, the reception level can be increased. Diversity receiver performance is discussed in further detail in Sec. 10.2.3.

Selecting cell-site locations. With a given actual antenna height and a given transmitted power, coverage area can be increased if we can select a proper site. Of course, in principle, for coverage purposes, we always select a high site if there is no risk of interference. However,

sometimes we need to cover an important area within the coverage area; in such cases it is necessary to move around the site location.

Using repeaters and enhancers to enlarge the coverage area or to fill in holes. This is discussed in Sec. 10.2.

Engineering the antenna patterns. The technique of engineering the antenna patterns mentioned in Sec. 7.6.4 can be used to cover a desired service area.

10.1.2 Reducing the interference

In most situations, the methods mentioned in Sec. 10.1.1 for increasing the coverage area would cause interference if cochannels or adjacent channels were used in the system. Methods for reducing the interference are as follows.

1. *A good frequency-management chart.* As shown in Fig. 8.1, there are 21 sets of channels in the chart. In each channel set, the neighboring frequency is 21 channels away. No interference can be caused within a set of 16 channels.
2. *An intelligent frequency assignment.* In order to assign the 21 sets in a $K = 7$ frequency-reuse pattern and to avoid the interference problems from adjacent-channel or cochannel interference, an intelligent frequency assignment in real time is needed.
3. *A proper frequency among a set assigned to a particular mobile unit.* Depending on the current situation, some idle channels may be noisy, some may be quiet, and some may be vulnerable to channel interference. These factors should be considered in assignment of frequency channels.
4. *Design of an antenna pattern on the basis of direction.* In some directions a strong signal may be needed; in other directions no signal may be needed. The design tool should include the findings of signal requirements on the basis of antenna direction.
5. *Tilting-antenna patterns.* To confine the energy within a small area, we may use an umbrella-pattern omnidirectional antenna or downward tilting directional antenna.
6. *Reducing the antenna height.* We can use this method because reducing interference is more important than radio coverage.
7. *Reducing the transmitted power.* In certain circumstances, reducing transmitted power can be more effective in eliminating interference than reducing the height of the antenna.

8. *Chosing the cell-site location.* The propagation prediction model described in Chap. 4 can be used to select cell-site locations for eliminating interference.

10.1.3 Increasing the traffic capacity

Small cell size. If we can control the radiation pattern, we can reduce the size of the cell and increase the traffic capacity. This approach is based on the assumption that all the mobile units are identical, including the mobile antennas and their mounting.

Increasing the number of radio channels in each cell. Either omnidirectional or directional antennas can be used in each cell. Sometimes the channel combiner can process only 16 channels. Thus, if we need 96 channels, we need six transmitted antennas. Also, if 6 frequency sets are used, then the total of 21 sets is divided by 6. The closest neighboring channels would be only four channels away. A good channel assignment method is needed (see Chap. 8).

Enhanced frequency spectrum. Cellular mobile industries have been allocated an additional 166 voice channels. With an enhanced frequency spectrum, traffic capacity is increased.

Queuing. Queuing of handoff calls can increase traffic capacity, as discussed in Chap. 9.

Dynamic channel assignment. Dynamic, rather than fixed, channel assignment is another means of increasing traffic capacity. As mentioned in Chap. 8, external environmental factors, such as traffic volume, are considered in dynamic channel assignment.

10.2 Coverage-Hole Filler

Because the ground is not flat, many water puddles form during a rainstorm; for the same reason, many holes (weak spots) are created in a general area during antenna radiation. There are several methods for filling these holes.

10.2.1 Enhancers (repeaters)[1]

An enhancer is used in an area which is a hole (weak spot) in the serving cell site. There are two types of enhancer: wideband and channelized enhancers.

The wideband enhancer is a repeater. It is designed for either block A or block B channel implementation. All the signals received will be amplified. Sometimes it can create intermodulation products; therefore, implementation of an enhancer in an appropriate place to fill the hole without creating interference is a challenging job. One application is shown in Fig. 10.1. The amplifier requires only low amplification. The signal is transmitted from the cell site and received at the enhancer site by a higher directional antenna which is mounted at a high altitude. The signal received in the forward channel will be radiated by the lower antenna, which is either an omnidirectional or a directional antenna at the enhancer. The mobile units in the vicinity of the enhancer site will receive the signal. The mobile unit uses the reverse channel to respond to calls (or originate calls) through the enhancer to the cell site.

However, the amplifier amplifies both the signal and the noise, as discussed in Sec. 7.4.2. Therefore, the enhancer cannot improve the signal-to-noise (S/N) ratio. The function of enhancers is actually a relay, receiving at a lower height h_2 and transmitting to a higher height h_1 or vice versa. The gain of the enhancer can be adjusted from 10 to 70 dB, and the range is from 0.5 to 3 km.[1] The received signal at the mobile units and at the cell site with an enhancer placed in the middle can be expressed as

$$P_{Rm} = P_{t_r} + g_c - L_a + (G + g_{E_1} + g_{E_2}) - L_b + g_m \qquad (10.1\text{-}9)$$

and

$$P_{Rc} = P_{t_m} + g_m - L_b + (G + g_{E_1} + g_{E_2}) - L_a + g_c \qquad (10.1\text{-}10)$$

where
P_{t_c} = transmitted power at cell site
P_{t_m} = transmitted power at mobile unit
g_c = antenna gain at cell site
g_m = antenna gain at mobile unit
g_{E_1}, g_{E_2} = antenna gain at enhancer
G = amplification gain at enhancer

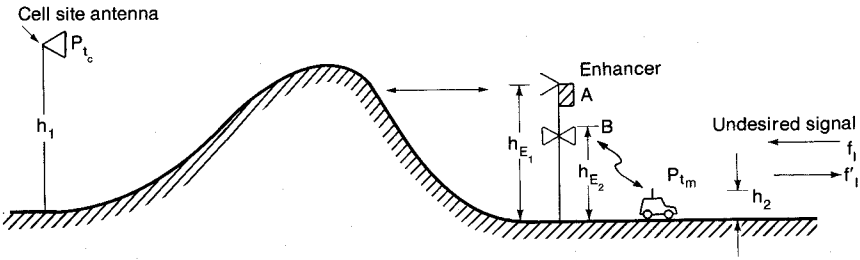


Figure 10.1 Enhancer.

P_{R_c}, P_{R_m} = received power at cell site and at mobile unit, respectively
h_1 = antenna height at cell site
h_2 = antenna height at mobile unit
h_{E_1}, h_{E_2} = antenna heights at enhancer
L_a = path loss between cell site and enhancer
L_b = path loss between enhancer and mobile unit

The general formula of path loss in a mobile radio environment [see Eq. (4.2-18)] can be used to calculate both L_a and L_b. Equation (4.2-18) contains an expression of a function of antenna height which would vary in different situations.

If the undesired signal received by the antenna at height h_{E_1} is transmitted back to the cell site, cochannel or adjacent-channel interference may result. This could also occur when an undesired signal is received by the antenna at height h_{E_1} because of poor design and is repeatedly transmitted by the antenna at height h_{E_2}, causing interference in a region in which undesired signal enhancement should not occur.

The channelized enhancer should amplify only the channels that it selected previously with a good design. Therefore, it is a useful apparatus for filling the holes.

Caution: Three points should be noted in the installation of an enhancer.

1. Ring oscillation might easily occur. The separation between two (upper and lower) antennas at the enhancer is very critical. If this separation is inadequate, the signal from the lower antenna can be received by the upper antenna or vice versa and create a ring oscillation, thus jamming the system instead of filling the hole.
2. The distance between the enhancer and the serving cell site should be as small as possible to avoid spread of power into a large area in the vicinity of the serving site and beyond.
3. Geographic (terrain) contour should be considered in enhancer installation.

10.2.2 Passive reflector

In order to redirect the incident energy, the reflector system should be installed in a field far from both the transmitting antenna and the receiving antenna.[2] The approximate separation between the antenna and the reflector is

$$d_1 > \frac{2A_T}{\lambda} + \frac{2A_1}{\lambda} \quad \text{and} \quad d_2 > \frac{2A_1}{\lambda} + \frac{2A_R}{\lambda} \qquad (10.2\text{-}1)$$

where A_T, A_R = effective aperture of transmitting antenna and receiving antennas, respectively
d_1, d_2 = distance from reflector to transmitting antenna and receiving antenna, respectively
λ = wavelength

If the transmitting and receiving antennas are linear elements, then

$$d_1 > \frac{2L_T^2}{\lambda} + \frac{2A_1}{\lambda} \quad \text{and} \quad d_2 > \frac{2A_1}{\lambda} + \frac{2L_R^2}{\lambda} \tag{10.2-2}$$

where L_T and L_R are, respectively, the transmitted and received lengths of the elements. The incident angle in this case would be less than 70° in order to deflect the energy in another direction (see Fig. 10.2).

The dimension of the reflector should be many wavelengths. Assume

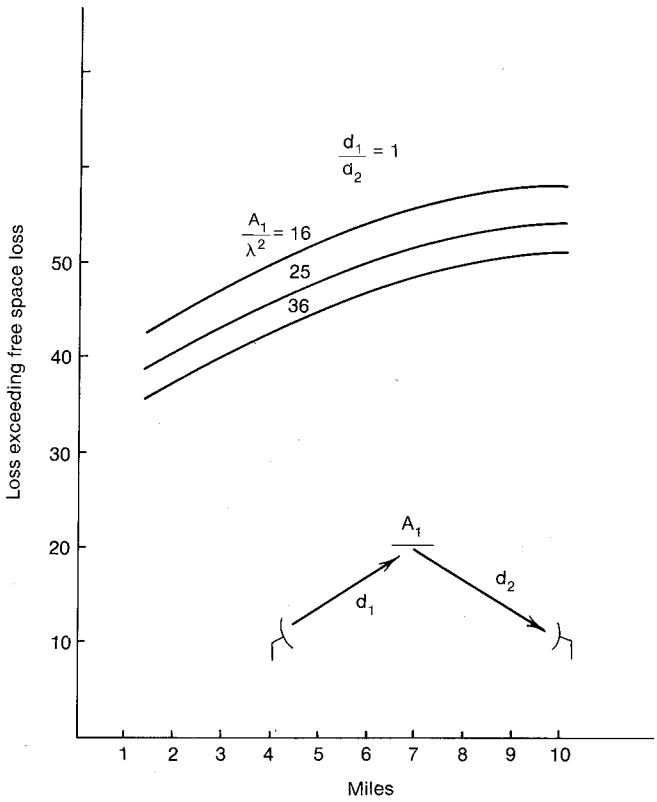


Figure 10.2 Effective use of reflectors $d_1/d_2 = 1$.

that 100 percent of the incident power is reflected; then

$$P_R = P_T \frac{A_T A_R A_1^2}{\lambda^4 d_1^2 d_2^2} \tag{10.2-3}$$

where P_T, P_R = transmitted and received power, respectively, and

$$A_T = \frac{G_T \lambda^2}{4\pi} \tag{10.2-4}$$

$$A_R = \frac{G_R \lambda^2}{4\pi} \tag{10.2-5}$$

Then Eq. (10.2-3) becomes

$$\begin{aligned} P_R &= P_T G_T G_R \frac{A_1^2}{(4\pi)^2 d_1^2 d_2^2} \\ &= P_T G_T G_R \underbrace{\left[\left(4\pi \frac{d}{\lambda}\right)^2\right]^{-1}}_{\text{free-space loss (FSL)}} \cdot \underbrace{\left(\frac{d^2 (A_1/\lambda^2)^2}{d_1^2 d_2^2}\right)}_{\text{excessive loss}} \end{aligned} \tag{10.2-6}$$

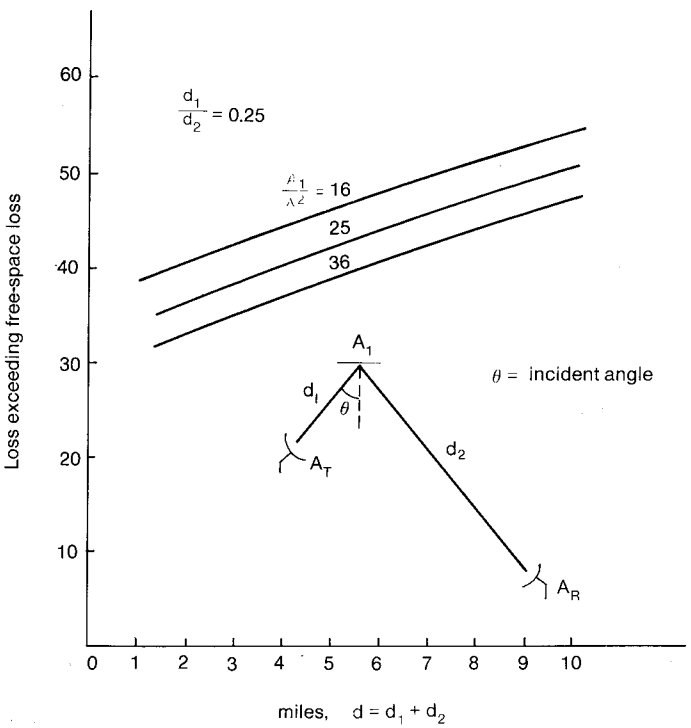


Figure 10.3 Effective use of reflectors $d_1/d_2 = 0.25$.

where $d = d_1 + d_2$ and

$$P_R = 10 \log(\text{FSL}) + 10 \log \left[\left(\frac{d^2\lambda^2}{d_1^2 d_2^2} \right) \left(\frac{A_1}{\lambda^2} \right)^2 \right] \tag{10.2-7}$$

The excessive loss in Eq. (10.2-7) is plotted in Fig. 10.2 for the case of $d_1/d_2 = 1.0$ and in Fig. 10.3 for $d_1/d_2 = 0.25$ at 850 MHz. In a mobile radio environment, d_1 can be considered to be in a free space and d_2 to be a mobile radio path from the reflector to the mobile unit.

Then Eq. (10.2-3) can be modified as

$$\begin{aligned} P_R &= P_T \frac{A_T A_R A_1^2}{\lambda^2 d_1^2 d_2^4} \\ &= P_T G_T G_R \left[\left(4\pi \frac{d_1}{\lambda} \right)^2 \frac{d_2^4}{\lambda^4} \right]^{-1} \cdot \frac{(A_1/\lambda^2)^2}{1} \end{aligned} \tag{10.2-8}$$

or

$$P_R = 10 \log(\text{FSL}) + 10 \log \underbrace{\left[\frac{d^2 A_1^2 \lambda^2}{d_1^2 d_2^4} \right]}_{\text{excessive loss}} \tag{10.2-9}$$

Comparing Eq. (10.2-9) with Eq. (10.2-7), we realize that the excessive loss from a reflector is greater in a mobile radio environment.

10.2.3 Diversity

The diversity receiver can be used to fill the holes. Because the diversity receiver can receive a lower signal level, the hole that existed in a normal receiver reception case now becomes a no-hole (or lesser hole) situation with the use of the diversity receiver. An improvement in the signal-to-noise ratio of a two-branch diversity receiver[3] is shown in Fig. 10.4. The diversity schemes can be classified as[4] (1) polarization diversity, (2) field component–energy density,[5] (3) space diversity, (4) frequency diversity, (5) time diversity, and (6) angle diversity.

For any two independent branches the performance obtained from any of the diversity schemes listed above is the same; that is, the correlation coefficient of the two received signals becomes zero. The performance can be degraded if the two signals obtained from the two branches are dependent on a correlation coefficient, as shown in Fig. 10.4. The performance can also vary with different diversity-combiner techniques.[6] The maximal-ratio combiner is the best performance combiner. The equal-gain combiner has a 0.5-dB degradation as compared with the maximal-ratio combiner. The selective combiner has a 2-dB degradation as compared with the maximal-ratio combiner.

The performance increase based on a diversity scheme for a two-branch equal-gain diversity combiner is shown in Fig. 10.5*a* for the cumulative probability distribution (CPD) and in Fig. 10.5*b* for the level-crossing rate (LCR). The average duration of fades $\bar{t}$ can be obtained by calculating $\bar{t}$ = CPD/LCR as shown in Eq. (1.6-9). Also, we can plot the performance of the diversity combined signal with different correlation coefficients between two branches. For example, at the cell site, the correlation coefficient ρ between branches is set to be 0.7 for the reality of physical antenna separation. At the mobile unit, however, the signal correlation of two branches is almost zero with a separation of $d = 0.5\lambda$. Reductions in fading and in level-crossing rate are shown in Fig. 10.5*a* and *b*, respectively. The improvement in the signal-to-noise ratio of a two-branch signal over a single branch with different values of correlation coefficients between channel signals is shown in Fig. 10.4. The maximum improvement occurs when $\rho = 0$.

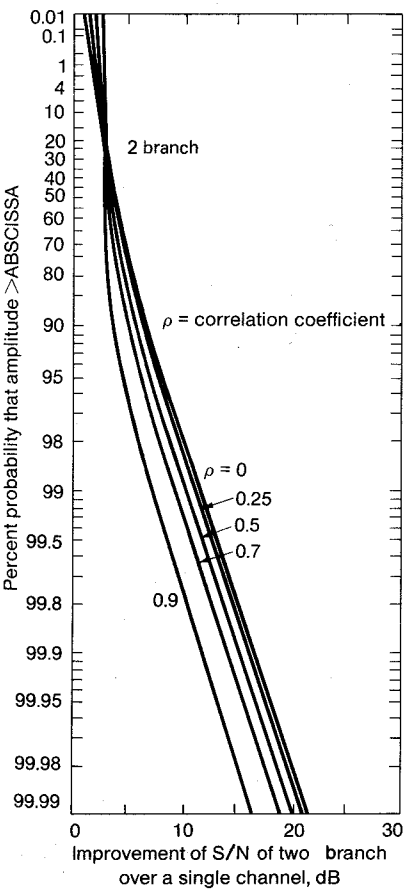
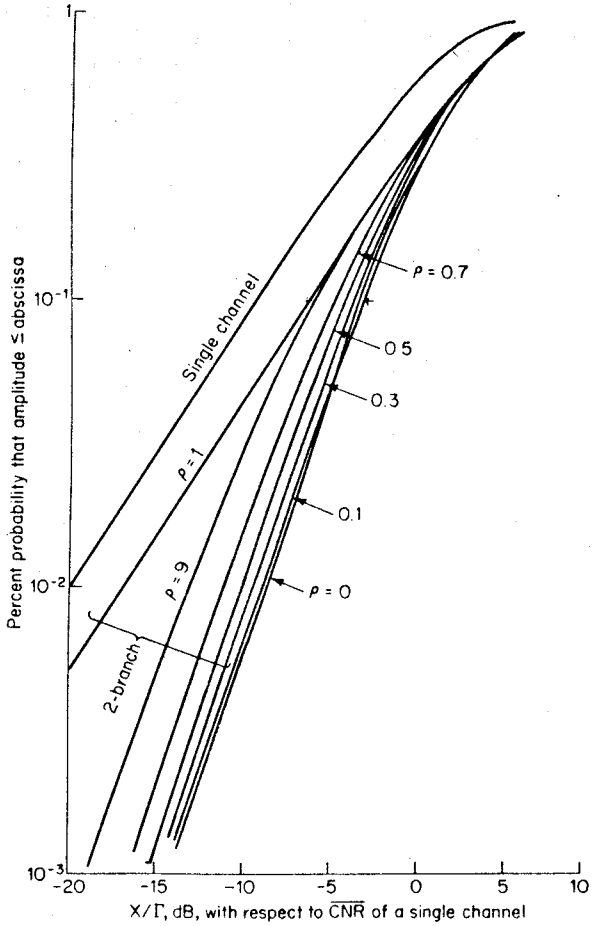


Figure 10.4 Improvement of signal-to-noise ratio of a two-branch signal over a single channel signal. (*After Lee, Ref. 3.*)



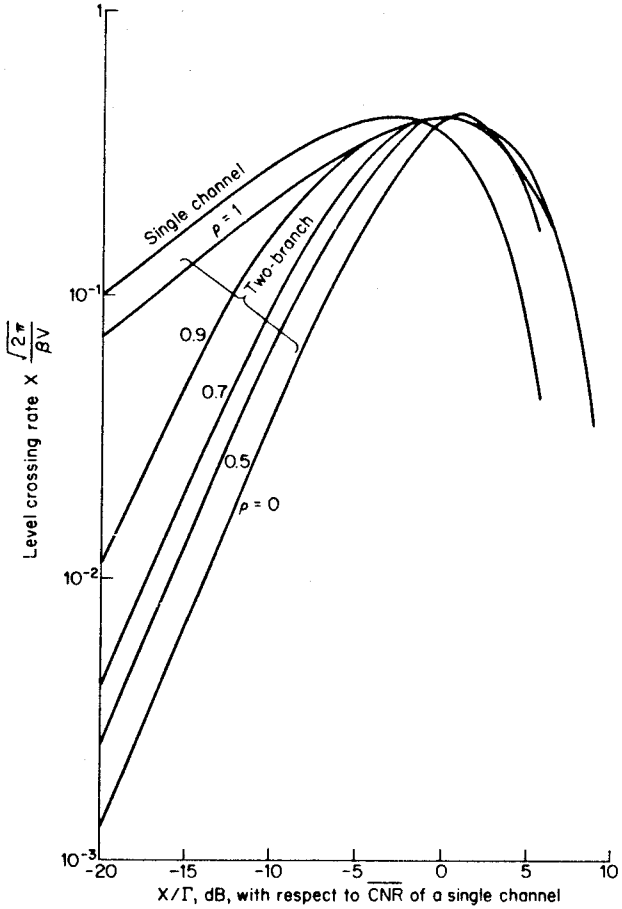
(a)

Figure 10.5 (*a*) Cumulative probability distribution of a two-branch correlated equal-gain-combining signal. (*After Lee, Ref. 7.*) (*b*) level-crossing rate of a two-branch equal-gain-combining signal. (*After Lee, Ref. 7.*)

10.2.4 Cophase technique

The cophase technique is used to bring all signal phases from different branches to a common phase point. Here, the common phase point is the point at which the random phase in each branch is reduced. There are two kinds of cophase techniques: feedforward and feedback[7] (these circuits are shown in Fig. 10.6*a* and *b*, respectively).

The feedforward cophase technique has been used for satellite communication applications. It is simpler than the phase-locked loop. The



(b)

Figure 10.5 (*Continued*)

latter is also called the *Granlund combiner*. The outcome of the feedback technique is always better than that of the feedforward technique provided the two filters in the circuit have been properly designed to avoid any significant time delay.

10.3 Leaky Feeder

10.3.1 Leaky waveguides

Typically, the velocity of propagation of an electromagnetic wave V_g in the waveguide is greater than the speed of light V_c. However, the

carrier frequency in hertz should be the same as in the waveguide and in free space. Thus, if two waves have the same frequency, their wavelengths will be longer in the waveguide than in free space, as seen from the following equation.

$$\lambda_g = \frac{V_g}{f} \tag{10.3-1}$$

Therefore,

$$\lambda_g > \lambda_c \tag{10.3-2}$$

If the waveguide structure supporting this mode is properly opened up, then the energy will leak into the exterior region.[8] The opening slots (apertures) will usually be placed along the waveguide periodically. This leaky waveguide is different from a slot antenna. The slot antenna is designed to radiate all the energy into the space at the slot, whereas in the leaky waveguide, fractional energy will be leaking constantly. Because V_g is greater than V_c, the leaky waveguide may sometimes be categorized as a fast-wave antenna.

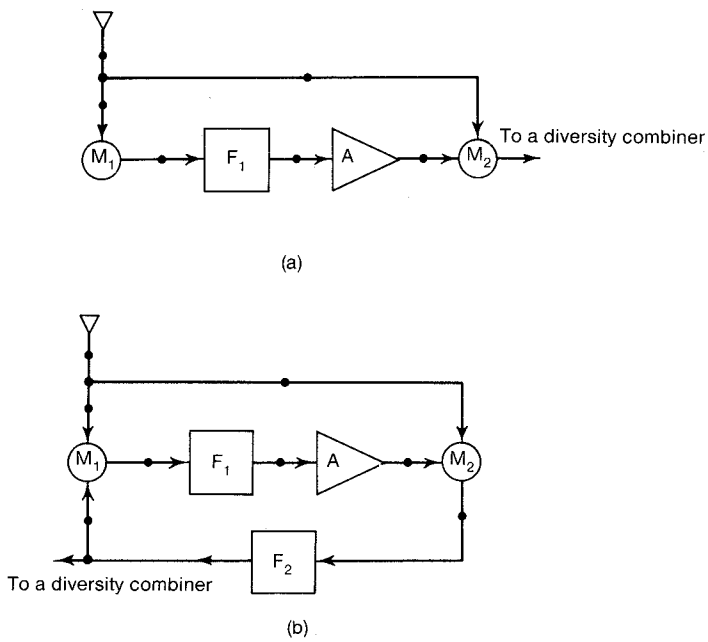


Figure 10.6 Two cophase techniques, feedforward and feedback (F = the filter, M = the mixer, and A = the limiting amplifier, as shown in the figure). (a) Feedforward combiner; (b) feedback (Granlund) combiner.

The general field expression can be written for the interior and exterior regions of the waveguide and matched across the slot boundary. For a circular-shaped waveguide,[9] the internal field is TE_{11}. The attenuation, or the leakage energy, is shown in Fig. 10.7*a*. Figure 10.7*b* shows the dimensions of the circular waveguide. The leakage rate is a function of position in the waveguide, where $\alpha = 0.1$ is the fraction of the input power absorbed in the load, that is, the amount of energy that leaks out.

The leaky waveguide pattern. Pattern is also a very important factor in the application of leaky waveguides. We would like the pattern to be similar to that in Fig. 10.7*c*, which can serve a larger area along the waveguide.

The coaxial cable or leaky coaxial. The phase velocity V of a wave traveling on the line is given by

$$V = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}} \tag{10.3-3}$$

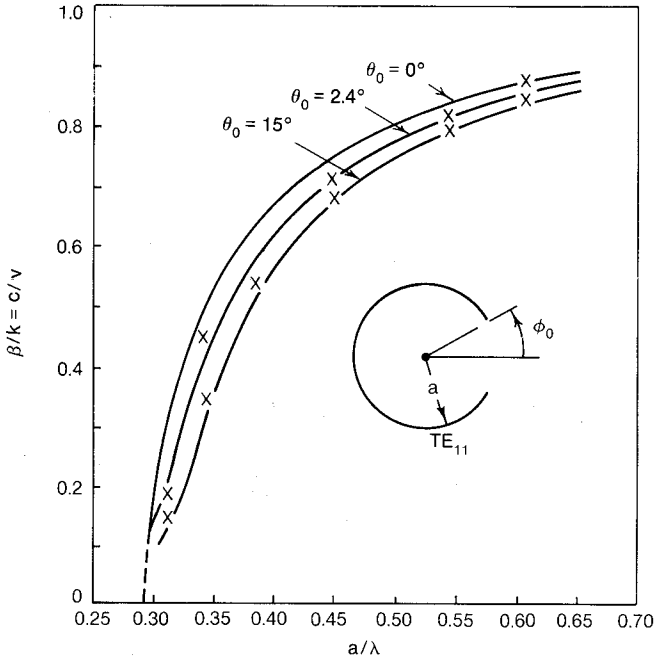
At frequencies below 1000 MHz, the use of coaxial cable is universal because the attenuation per unit length is reasonable and the dimensions are practical for passing the principal modes of leaky waves. At higher frequencies, since the dimensions of coaxial cable cannot be physically reduced in order to suppress the high modes of leaky waves in the coaxial cable, excessively high attenuation might occur. Consequently, for frequencies above 3000 MHz, waveguides are generally used.

10.3.2 Leaky-feeder radio communication

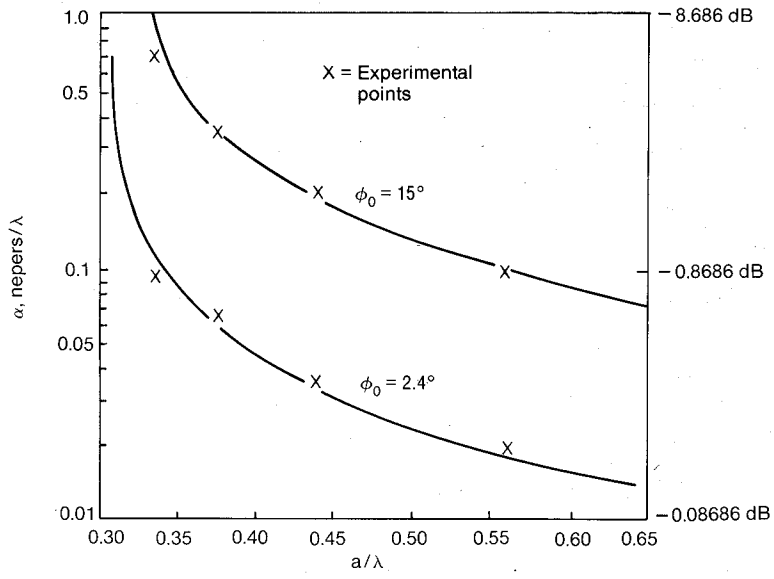
In some areas, such as in tunnels or in other confined spaces such as underground garages or a cell of less than 1-mi radius, leaky-feeder techniques become increasingly important to provide adequate coverage and reduce interference.

In 1956 a "guided radio" was introduced.[10] This "radio" is actually a low-frequency inductive communication device. The proposal included utilization of existing conductors such as power cables and telephone lines to transmit the signal.

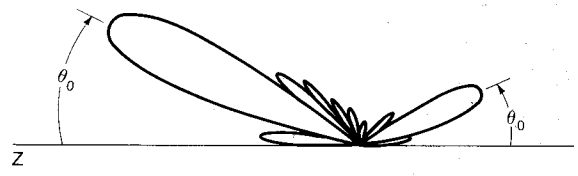
Also in 1956, the leaky-feeder principle for propagation of VHF and UHF signals through a tunnel or a confined area was presented.[11] The



(a)



(b)



(c)

Figure 10.7 Complex propagation constant for a TE_{11} slotted cylinder. (*After Harrington, Ref. 9.*). (*a*) Leaky energy due to slot size; (*b*) wave velocity in a slotted cylinder; (*c*) principal plane pattern of a leaky wave slot; showing forward and rearward main beams.

open-braided type (i.e., containing zigzag slots) of coaxial cable is used in most applications for suppressing any resulting surface-wave interference (Fig. 10.8). However, in this design, if the cable slots are all the same size, then there is nonuniform energy leakage along the cable. A great deal of energy may leak out at the slots which are arrived at first. For instance, a leaky cable can have a loss of 2 dB per 100 ft at 1000 MHz. The "daisy chain" system patented in 1971 avoids the complications and shortcomings of two-way signal boosters along the cable.[12] Therefore the radiation signal level can be within a specified range.

Because of "intrinsic safety" considerations, in order to prevent any incendiary sparks (e.g., as in a coal mine), the RF powers cannot exceed a maximum of 500 mW, and any line-fed power passed over leaky feeders used for boosters (power amplification along the cable) should be limited to a few watts. In urban applications, a 0.25-mi-long leaky cable will be used without the power amplifying stage.

The leaky feeder is characterized by transmission and coupling losses. Transmission loss is expressed in decibels per unit length. Coupling loss is defined by the ratio of power received by a dipole antenna at a distance s equal to 1.5 m away from the cable to the transmitted power in the cable at a given point. The smaller the ratio, the greater the

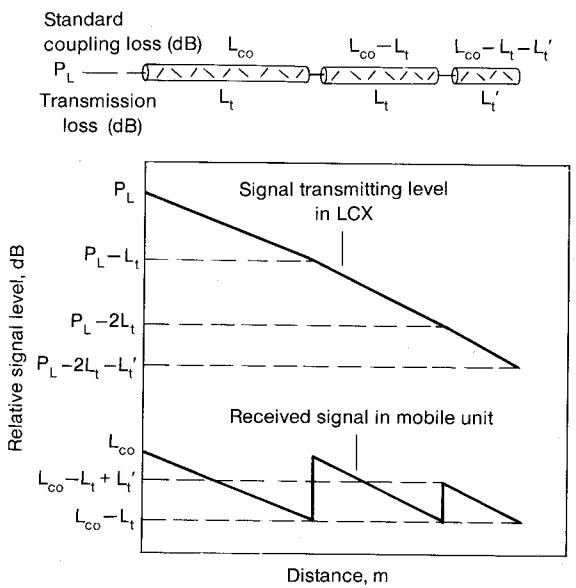


Figure 10.8 Grading technique.

loss. If the distance is other than 1.5 m, the coupling loss (or free-space loss from a leaky cable) L increases as d increases.

$$L\ (s \text{ at } d) = (\text{coupling loss at 1.5 m}) + 10 \log \left(\frac{d}{1.5}\right) \quad (10.3\text{-}4)$$

The free-space loss from a leaky cable is described later. The coupling loss can be controlled by size and slot angle, whereas the transmission loss varies with the coupling loss and cannot be chosen independently of the coupling loss. The principal of leaky-cable operation is

1. Use high-coupling-loss (little energy will leak out) cables near the transmitter end. Usually high-coupling-loss cables have a low-transmission loss and are of greater length in use. We can arrange the lengths of cables due to different coupling losses as shown in Fig. 10.8.
2. The intensive radiation pointing to a specific direction is caused by periodic spacing of slots along the cable. Radiation can be distributed through joint points or boosters and by adjusting the signal phases around boosters as needed.

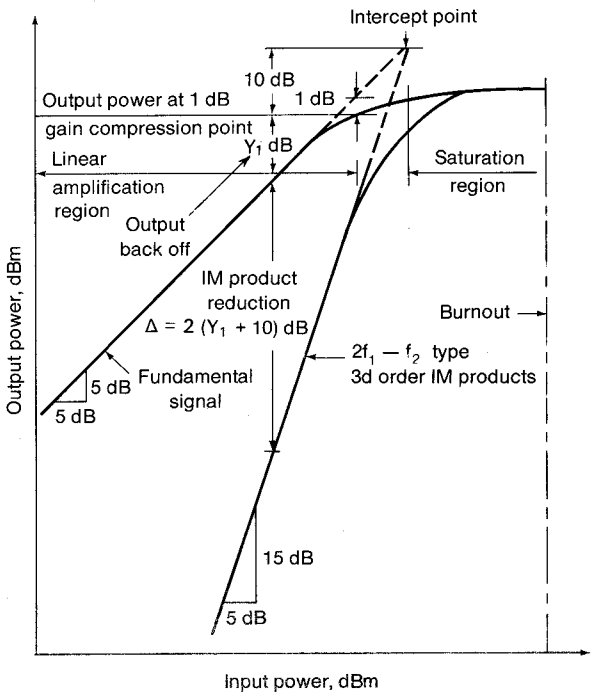


Figure 10.9 Input-output characteristics of a linear amplifier (*After Suzuki et al., Ref. 12*).

3. Leaky cables are open fields. Leaky cables in the tunnels are easily implemented because their energy is confined to the tunnel. However, in an open field, if no obstacle blocks the path between the cable and the mobile receiver, the signal should be less varied. The electric field leaking out from the leaky cable is reciprocally proportional to the square of the distance from the leaky cable.

$$L_r = 20 \log s \qquad \text{dB} \tag{10.3-5}$$

4. Low temperature affects leaky cable. Transmission-loss levels change with change in temperature. The lower the temperature, the less the transmission loss.

5. Snow accumulation around slots causes an increase in transmission loss. Reflection and path loss due to snow on leaky cable cause an increase of coupling loss.

6. The boosters are power amplifiers. Therefore, many narrowband-modulated carriers passing through common broadband amplifiers generate intermodulation (IM) product power. In order to reduce the IM product to a specified level, the linear amplifiers should be operated at a reduced output level by backing them off from the 1-dB-gain compression point.

The amplification of a fundamental signal and its most dominant IM (i.e., third-order IM) is illustrated in Fig. 10.9. Because the slopes of curves for fundamental signal and third-order IM are always fixed, the higher the intercept point, the lower the IM product interference. Also, we can find the output backoff level from a given IM product suppression, Y_1, as

$$2\,(Y_1 + Y_0) = \Delta$$

$$Y_1 = \frac{\Delta}{2} - Y_0 \tag{10.3-6}$$

where Y_0 is the power difference between the intercept point and the 1-dB-gain compression point. The IM product levels and numbers are given in Ref. 13. Other literature references can be found in Refs. 22 and 23.

10.4 Cell Splitting

When the call traffic in an area increases, we must split the cell so that we can reuse frequency more often, as we have mentioned in Chap. 2. This involves reducing the radius of a cell by half and splitting an old cell into four new small cells. The traffic is then increased fourfold.[14]

10.4.1 Transmitted power after splitting

The transmitted power P_{t_2} for a new cell, because of its reduced size, can be determined from the transmitted power P_{t_1} of the old cell.

If we assume that the received power at the cell boundary is P_r, then the following equations (where α is a constant) can be deduced from Eq. (10.1-1).

$$P_r = \alpha P_{t_0} R_0^{-\gamma} \tag{10.4-1}$$

$$P_r = \alpha P_{t_2} \left(\frac{R_1}{2}\right)^{-\gamma} \tag{10.4-2}$$

Equation (10.4-1) expresses the received power at the boundary of the old cell and Eq. (10.4-2), the received power at the boundary of the new cell $R_1 = (R_0/2)$. To set up an identical received power P_r at the boundaries of two different-sized cells, and dropping the parameter P_r by combining Eqs. (10.4-1) and (10.4-2), we find

$$P_{t_1} = P_{t_0} \left(\frac{1}{2}\right)^{-\gamma} \tag{10.4-3}$$

For a typical mobile radio environment, $\gamma = 4$, Eq. (10.4-3) becomes

$$P_{t_1} = \frac{P_{t_0}}{16} \tag{10.4-4}$$

or

$$P_{t_1} = P_{t_0} - 12 \quad \text{dB} \tag{10.4-5}$$

The new transmitted power must be 12 dB less than the old transmitted power. The new cochannel interference reduction factor q_2 after cell splitting is still equal to the value of q (see Eq. 2.3-1) since both D and R were split in half. A general formula is for a new cell which is split repeatedly n times, and every time the new radius is one-half of the old one; then $R_n = R_0/2^n$.

$$P_{t_n} = P_{t_0} - n(12) \quad \text{dB} \tag{10.4-6}$$

When cell splitting occurs, the value of the frequency-reuse distance q is always held constant. The traffic load can increase four times in the same area after the original cell is split into four subcells. Each subcell can again be split into four subcells, which would allow traffic to increase 16 times. As the cell splitting continues, the general formula can be expressed as

$$\text{New traffic load} = (4)^n \times (\text{the traffic load of start-up cell}) \tag{10.4-7}$$

where n is the number of splittings. For $n = 4$, this means that an original start-up cell has split four times. The traffic load is 256 times larger than the traffic load of the start-up cell.

10.4.2 Cell-splitting technique

The two techniques of cell splitting are described below.

Permanent splitting. Selecting small cell sites is a tough job. The antenna can be mounted on a monopole or erected by a mastless arrangement which will be described later. However, these splittings can be easy to handle as long as the cutover from large cells to small cells takes place during a low traffic period. The frequency assignment should follow the rule (see Sec. 2.3) based on the frequency-reuse distance ratio q with the power adjusted.

Real-time splitting (dynamic splitting). In many situations, such as traffic jams at football stadiums after a game, in traffic jams resulting from automobile accidents, and so on, the idle small cell sites (inactive ones) may be rendered operative in order to increase the cell's traffic capacity.

Cell splitting should proceed gradually over a cellular operating system to prevent dropped calls. Suppose that the area exactly midway between two old $2A$ sectors requires increased traffic capacity as indicated in Fig. 10.10. We can take the midpoint between two old $2A$ sectors and name it "new $2A$." The new $1A$ sector can be found by

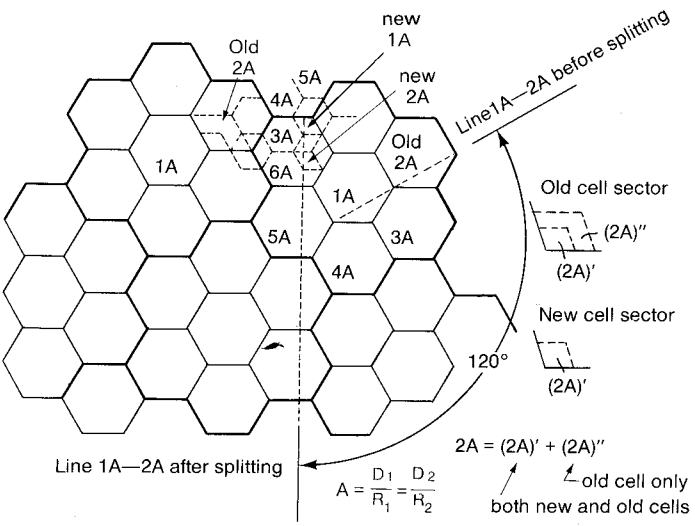


Figure 10.10 Cell-splitting techniques.

rotating the old $1A$–$2A$ line (shown in Fig. 10.10) clockwise[15] 120°. Then the orientation of the new set of seven split cells is determined. To maintain service for ongoing calls while doing the cell splitting, we let the channels assigned in the old $2A$ sector separate into two groups.

$$2A = (2A)' + (2A)'' \tag{10.4-8}$$

where $(2A)'$ represents the frequency channels used in both new and old cells, but in the small sectors, and $(2A)''$ represents the frequency channels used only in the old cells.

At the early splitting stage, only a few channels are in $(2A)'$. Gradually, more channels will be transferred from $(2A)''$ to $(2A)'$. When no channels remain in $(2A)''$, the cell-splitting procedure will be completed. With a software algorithm program, the cell-splitting procedure should be easy to handle.

10.4.3 Splitting size limitations and traffic handling

The size of splitting cells is dependent on two factors.

The radio aspect. The size of a small cell is dependent on how well the coverage pattern can be controlled and how accurately vehicle locations would be known.

The capacity of the switching processor. The smaller the cells, the more handoffs will occur, and the more the cell-splitting process is needed. This factor, the capacity of a switching processor, is a larger factor than the handling of coverage areas of small cells.

10.4.4 Effect on splitting

When the cell splitting is occurring, in order to maintain the frequency-reuse distance ratio q in a system, there are two considerations.

1. Cells splitting affects the neighboring cells. Splitting cells causes an unbalanced situation in power and frequency-reuse distance and makes it necessary to split small cells in the neighboring cells. This phenomenon is the same as a ripple effect.
2. Certain channels should be used as barriers. To the same extent, large and small cells can be isolated by selecting a group of frequencies which will be used only in the cells located between the large cells on one side and the small cells on the other side, in order to eliminate the interference being transmitted from the large cells to the small cells.

10.5 Small Cells (Microcells)

As we mentioned in Sec. 10.4.3, the limitation of a small cell is based on the accuracy of vehicle locations and control of the radiation patterns of the antennas.

10.5.1 Installation of a mastless antenna

Use of existing building structures. Building structures can be used to mount cell-site antennas. In such cases the rooftop usually is flat. There should be enough clearance around the antenna post mounted in the middle of the building to avoid blockage of the beam pattern from the edge of the roof (see Fig. 10.11). A formula may be applied for this situation. Given the vertical beamwidth of antenna ϕ and the distance from the antenna post to the edge of the roof d, the height of the post can be determined by

$$h = d \tan\left(\frac{\phi}{2}\right) \tag{10.5-1}$$

If a 6-dB-gain antenna has a vertical beamwidth of about 28° and the distance from the antenna post to the roof edge is 31 m (100 ft), then the required antenna height is 7.5 m (25 ft). The shaded region around the building depends on the height of the building.

$$D = \frac{H + h}{\tan(\phi/2)} - d \tag{10.5-2}$$

If $H = 12$ m (40 ft), $h = 7.5$ m (25 ft), $\phi/2 = 14°$, then Eq. (10.5-2) becomes $D = 49$ m (160 ft).

Use of the antenna structures. The panel-type antennas[16] are ideal antenna structures for hanging on each side of the wall. For an omni-

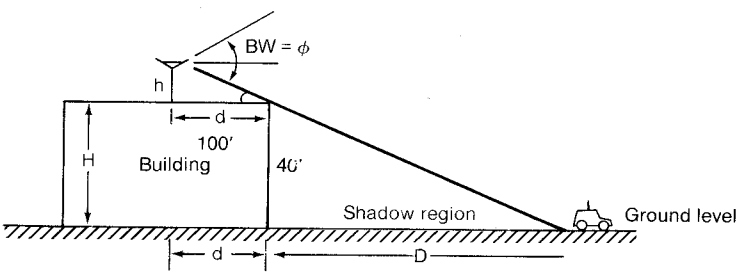


Figure 10.11 Rooftop-mounted antennas.

directional configuration, the four-panel antennas mounted on the four sides of the building can be combined as in an omnidirectional antenna.

For a sectorized configuration, each antenna occupies one sector. If a three-sector configuration is used, two panel antennas should be mounted close to the two corners of the building and one panel mounted on a flat wall of the building as shown in Fig. 10.12.

10.5.2 Tailoring a uniform-coverage cell

We will develop a lightweight, fold-up, portable directional antenna with adjustable beamwidth capability. It can be in the form of a corner reflector or an n-element array. This kind of antenna can be attached to the outer walls of the building in different directions. Perhaps we can attach such antennas to different buildings and form a desired coverage. The transmitted power of the antenna in each direction is also adjusted so that the coverage becomes uniformly distributed around the cell boundary. These antennas may be called *coverage sectored antennas*. The "coverage sectors" and the "frequency sectors" are not necessarily the same. Usually, several coverage sectors represent a frequency sector when the cell size becomes small. Since the power coverage is based on the coverage sectors and the frequency assignment

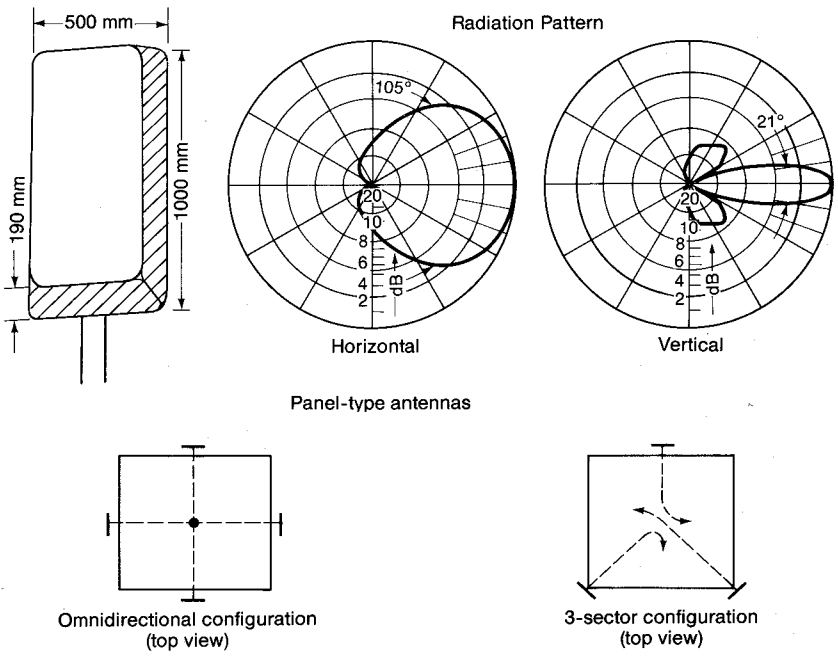


Figure 10.12 Panel-type antennas and their applications (*After Kathrein, Ref. 16*).

is based on the frequency sectors, the existing software should be modified to incorporate this feature.

10.5.3 Vehicle-locating methods

By locating the vehicle and calculating the distance to it, we can obtain information useful for assigning proper frequency channels.

There are many vehicle-locating methods. In general, we can divide these into two categories: installation of equipment (1) in the vehicles and (2) at the cell site.

Installing equipment in the vehicles

Triangulation. Three or more transmitting antennas are used at different cell sites. Since the locations of the sites are known, the vehicle's location can be based on identification of three or more sites. However, the accuracy is limited by the multipath phenomenon.

In certain areas of the United States, especially near the coasts, Logran-C transmitters are used by the U.S. Coast Guard. These transmitters operate at 500 kHz. A Logran-C receiver installed in a vehicle can facilitate locating the vehicle.

Fifth-wheel and gyroscope equipment. A gyroscope and a fifth wheel are used for determining the direction and distance a vehicle has traveled from a predetermined point at any given time.

The globe-position satellite (GPS). There are seven active GPSs, each of them circling the earth roughly twice a day at an altitude of 1840 km (11,500 mi) and transmitting at a frequency of approximately 1.7 GHz. There are two codes, the C code and the P code. The C code is the coarse code which can be used by the civilian services. The P code is the precision code, which is used only by the military forces. At least three or more GPS satellites should be seen in space at any time, so that a GPS receiver can locate its position according to the known positions of the GPS satellites. Under this condition, we need at least 18 GPSs but only 7 GPSs are in space today. The GPS location is very accurate, generally within 6 m (20 ft). When four GPSs are in space, we can measure three dimensions, i.e., latitude, longitude, and altitude. The cost of a GPS receiver is very high and is not affordable at present for commercial cellular mobile systems.

Installing equipment at the cell site. In general, either of the following three methods alone cannot provide sufficient accuracy for locating vehicles; a combination of two or all three methods is recommended.

Triangulation based on signal strength. Record the signal strength received from the mobile unit at each cell site and then apply the trian-

gulation method to find the location of the mobile unit. The degree of accuracy is very poor because of the multipath phenomenon.

Triangulation based on angular arrival. Record the direction of signal arrival at each cell site and then apply the triangulation method to find the location of the mobile unit.

Triangulation based on response-time arrival. Send a signal to the mobile unit. It will return with a time delay or a phase change. Measurement of the time delay or the phase change at each site can indicate the distance from that site.

Two or more distances from different sites can help us determine the location of the vehicle. However, the delay spread in the mobile environment can be 0.5 μs in suburban areas to 3 μs in urban areas. We may need ingenuity to solve this problem if the locating method is based on the response-time arrival.

Present cellular locating receiver. Each cell site is equipped with a locating receiver which can both scan and measure the signal strengths of all channels. This receiver can be used to continuously scan the frequencies, or to scan on request.

Continuous-scanning scheme. In continuous scanning of all 333 channels, assume that scanning each channel takes 20 ms. Thus, the time interval between two consecutive measurements of any single channel is 333 × 20 ms = 6.6 s.

If a car is driven at 30 mi/h (=44 ft/s), the interval between two different measurements on one frequency will be 6.6 s or 290 ft, so we would not expect a drastic change in this interval. However, the time interval of 20 ms for measuring the signal strength on a frequency is too short—only about 1 ft in distance (about 1 wavelength). As discussed in Chap. 1, we need 40 wavelengths to obtain good measurement data.[17] Therefore, if we are using the continuous-scanning scheme, the running mean $M(N)$ at the Nth sample based on the average of N samples should be tracked.

$$M(N) = \frac{\sum_{i=1}^{N} x_i}{N}$$

$$M(N+1) = \frac{\sum_{i=2}^{N+1} x_i}{N}$$

$$\vdots$$

The advantage of this scheme is that

1. Each cell site "knows" the signal-plus-interference levels ($S + I$) of all active channels. The cell site can respond without delay to the mobile telephone switching office (MTSO) regarding the $S + I$ level of any one channel for measuring.
2. Each cell site knows the interference (or noise) levels of all idle channels. The cell site can choose a prospective (candidate) channel on the basis of its low interference level.

The selected channel mentioned in item 2 must not only generate a better signal-to-interference ratio but also less interference that affects the other cells. The argument for this is that if no interference is received by a channel, this channel will not cause interference in others. Use of a high-interference channel will not only cause deterioration in voice quality but also generate more interference in other cells.

Scan-under-request scheme. When measurement of a channel's signal strength is requested, the cell site must have enough time to measure it with a locating receiver; there is usually one locating receiver per cell site. Each locating receiver is capable of tuning and measuring all channels. Therefore, actual amount of time spent measuring the signal strength of each individual channel upon request can be relative long for high accuracy. The disadvantages of the continuous-scanning scheme are compensated for by its advantages and vice versa.

10.5.4 Portable cell sites

For rapid addition of new cell sites to an existing system, portable cell sites are used to serve the traffic temporarily while the permanent site is under construction. In other situations, when it has not been determined whether the prospective site will be appropriate, the portable cell site can be used for a short period of time so that real operational data can be collected to determine whether this site will be suitable for a permanent site installation. Construction of a cell site normally requires three primary activities: (1) site acquisition, (2) building and tower construction, and (3) equipment installation and testing. A fair amount of time is needed for each activity. The portable cell site consists of buildings, equipment, and antennas, and all three of these items should be transportable.[18]

10.5.5 Different antenna mountings on the mobile unit

The different antenna mountings used on the mobile units affect system performance. The rooftop-mounted antenna, because of the great an-

tenna height above the ground and the roughly uniform coverage, provides maximum coverage.

On the other hand, antennas mounted on windows, car bumpers, or trunks provide less coverage than do those mounted on roofs. However, more than 70 percent of the cars in cellular systems use glass-mounted antennas. Now the system operator must decide whether the available cellular system has to tune for glass-mounted mobile units or rooftop-mounted mobile units.

If the system is tuned for rooftop-mounted mobile antennas, that is, if the $q = D/R$ ratio is based on the reception of $C/I \geq 18$ dB of the rooftop-mounted antennas, then the cell coverages for rooftop-mounted units provides for no gaps. However, the cell coverage for glass-mounted units does allow gaps to form because of the weaker signals around all cell boundaries, which in turn results in excessive call drops.

If the system is tuned for all mobile units with glass-mounted antennas, then coverage of all cells will be suitable for these units. But the mobile units with rooftop-mounted antennas will travel deeply into the neighboring cells because of a still adequate signal at the cell boundaries and the delay of handoffs taking place. Then the units with rooftop-mounted antennas will experience channel interference such as cross talk and dropped calls due to distorted SAT tones.

Since the system tuning mentioned above cannot satisfy both kinds of mobile antenna usage, the author recommends that if the handoff is based on signal strength level a system should be tuned for those mobile units which have glass-mounted antennas. Then for the rooftop-mounted antenna units, attenuation can be added or the antenna can be tilted at some appropriate angle (loss due to off-vertical position).

If the handoff is based on power-difference schemes (Sec. 9.6), the effect of this scheme on different mobile antenna mountings becomes much less. The handoff occurrence areas for the mobile units with different antennas and different mountings are very much the same.

10.6 Narrowbeam Concept

The narrowbeam-sector concept is another method for increasing the traffic capacity (see Fig. 10.13). For a $K = 7$ frequency-reuse pattern with 120° sectors as a conventional configuration, each sector will contain approximately 15 voice channels, a number which is derived from the total 312 voice channels

$$\frac{333 - 21}{3 \times 7} = 15 \text{ channels per } 120° \text{ sector}$$

For a $K = 4$ frequency-reuse pattern[19] with 60° sectors (Fig. 10.14*a*), the number of channels in each 60° sector is

$$\frac{333 - 21}{4 \times 6} = 13 \text{ channels per } 60^\circ \text{ sector}$$

In the $K = 7$ pattern there is a total of 21 sectors with 15 channels in each sector; in the $K = 4$ pattern there is a total of 24 sectors with 13 channels in each sector. The spectrum efficiency of using these two patterns can be calculated using the Erlang B table in Appendix 1.1. With a blocking probability of 2 percent, the results are: an offer load of 189 erlangs for $K = 7$ and 177 erlangs for $K = 4$. This means that the $K = 7$ pattern offers a 7 percent higher spectrum efficiency than the $K = 4$ pattern does. As seen in Fig. 10.14*a* a number of cell sites have been eliminated for $K = 4$ as compared with $K = 7$, assuming the same coverage area. However, the $K = 4$ arrangement results in increased handoff processing. Also the antennas erected in each site with a $K = 4$ pattern should be relatively higher than those with a $K = 7$ pattern. Otherwise, channel interference among channels will be increased because the wrong frequency channels will be assigned to the mobile units due to the low antenna height in the system. As a result, the actual location of the mobile units in smaller sectors may be incorrect.

Here we could use the scheme shown in Fig. 10.14*b* for customizing channel distribution; that is, usage of the 120° and 60° sectors can be mixed. Some 120° sectors can be replaced by two 60° sectors in a $K = 7$ pattern. The number of channels can then be increased from 15 to 26 as shown in Fig. 10.14*b*. This scheme would be suitable for small-cell systems. The antenna-height requirement for 60° sectors in small cells

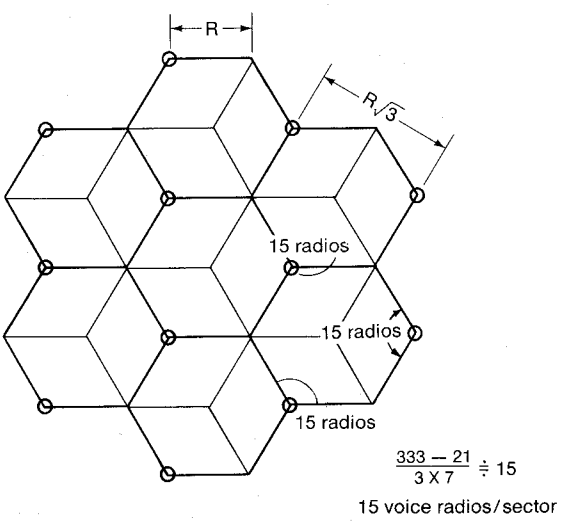


Figure 10.13 Ideally located cell sites over a flat terrain ($K = 7$).

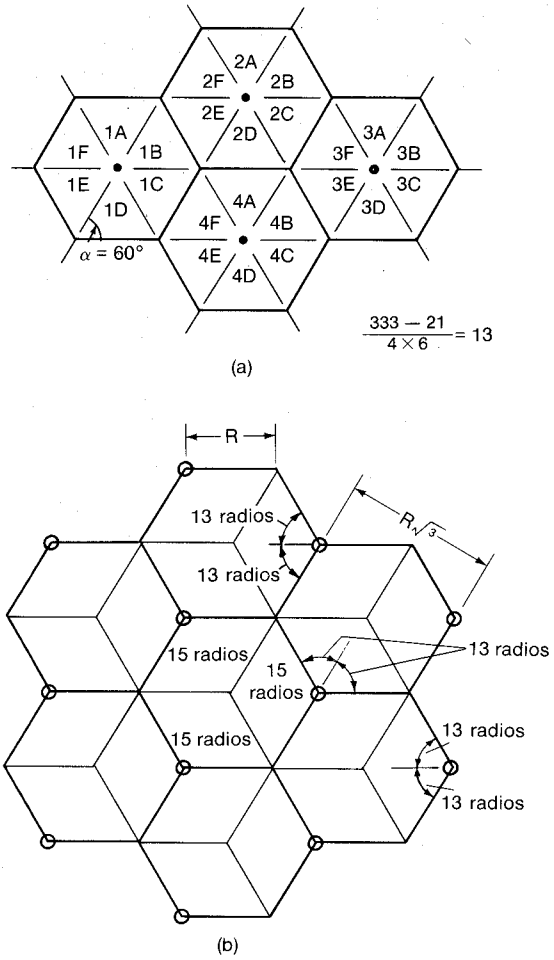


Figure 10.14 60°-sector cell sites. (*a*) Motorola's plan ($K = 4$): thirteen voice radios per sector in every 60° sector; (*b*) ideally located cell sites ($K = 7$) (mixed 120° and 60° sectors as needed).

is relatively higher than that for 120° sectors. Besides, the 24 subgroups (each containing 13 channels) are used as needed in certain areas. These sector-mixed systems follow a $K = 7$ frequency-reuse pattern, and the traffic capacity is dramatically increased as a result of customizing the channel distribution according to the real traffic condition.

Comparison of narrowbeam sectors with underlay-overlay arrangement. In certain situations the narrowbeam sector scheme is better in a small cell than the underlay-overlay scheme. In a small cell, it is very difficult to control power in order to make underlay-overlay schemes work ef-

fectively. For 60° sectors, the 60° narrowbeam antennas would easily delineate the area for operation of the assigned radio channels. However, choosing the correct narrowbeam sector where the mobile unit is located is hard. As a result, many unnecessary handoffs may take place.

In a 1-mi cell, if the traffic density is not uniformly distributed throughout the cell, the choice of using narrowbeam sectors or an underlay-overlay scheme is as follows; use the former for the angularly nonuniform cells, and use the latter for the radially nonuniform cells.

10.7 Separation between Highway Cell Sites

In generally light traffic areas, signal-strength coverage is a major concern, especially the coverage along the highways. There are two potential conditions to be considered in highway coverage: (1) relatively heavy traffic and (2) light traffic. In condition 1 there would be a high to average human-made noise level and in condition 2, a relatively to very low human-made noise level. Under these conditions, we recommend that the new highway cell-site separation should be much greater than that used for a normal cell site.

Omnidirectional antenna. As it is necessary to cover not only the highway but also areas in the vicinity of the highway, the omnidirectional antenna is used. When the cell sites are chosen and put up along the highway (see Fig. 10.15), the line-of-sight situation is usually assumed.

Although the general area around the highway could be suburban, because of the line-of-sight situation the path loss should be calculated

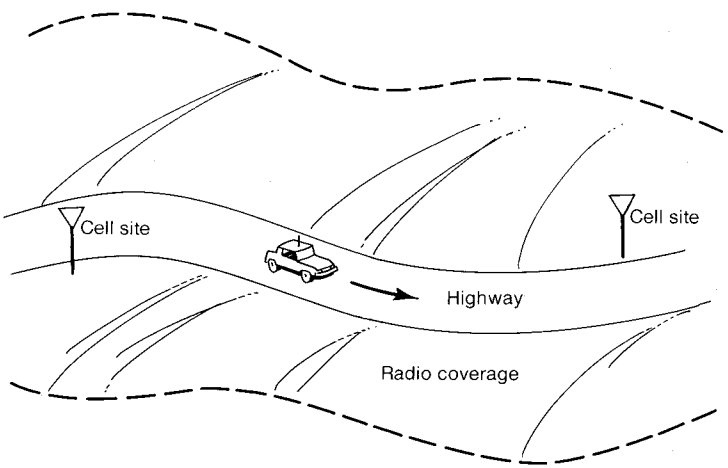


Figure 10.15 Highway cell sites.

using an open-area curve instead of the suburban-area curve shown in Fig. 4.3.

The differences between highway cell-site separation and normal cell-site separation using path-loss values from the suburban and the open-area curves (see Fig. 4.3), respectively, are plotted in Fig. 10.16. The curve is labeled "Noise condition of human origin."

Traffic along highways away from densely populated areas is usually light and the level of automotive noise is low, perhaps 2 dB lower than the average human-made noise-level condition. Based on the 2-dB noise quieting assumption, another curve labeled "Low noise condition of human origin" is also shown in Fig. 10.16.

From Fig. 10.16 we can obtain the following data.

Average cell-site separation, mi; normal human-made noise condition	Highway cell-site separation, mi	
	Normal human-made noise condition	Low human-made noise condition
6	9.5	11
10	15	17

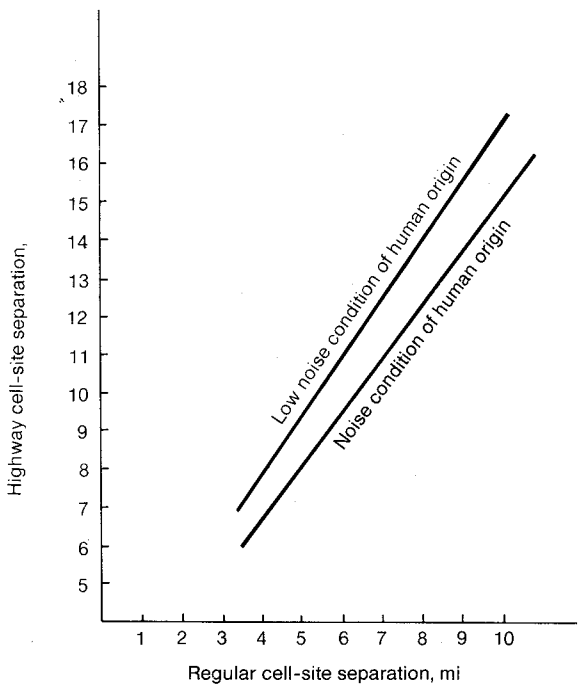


Figure 10.16 Highway cell-site separation.

The purpose of using the omnidirectional antenna at the cell sites along the highway is to cover the area in the vicinity of the highway where residential areas are usually located.

Two-directional antennas. In certain areas where only highway coverage is needed, two-directional antennas, such as horns or a pair of yagi antennas placed back to back at the cell site along the highway, could be installed. The directivity can result in a further increase in separation between the sites. Equation (10.7-1) shows the relation between an increase in directivity ΔG in decibels and the increase in the additional separation Δd.

$$\Delta d = d[10\Delta G/20 - 1] \tag{10.7-1}$$

This equation can easily be derived from a free-space path-loss condition.

10.8 Low-Density Small-Market Design

In a small market (city) one of the primary concerns is cost, since the low-density subscriber environment is basically suburban. Here, antennas can be lower but cover the same area as would a higher antenna in an urban area. In a noise-limited environment such as a suburban one, there are no problems in frequency assignment. The antenna tower can be constructed according to the following four considerations.

1. Use an existing high tower if available to obtain the maximum coverage in a given area. Because no frequency-reuse scheme will be applied, cochannel interference is of no concern.
2. Use a low-cost antenna tower. If there is no existing tower, then construct a low-cost tower in a farm area, using chicken wire to fasten the structure at a height of 15 to 24 m (50 to 80 ft).
3. Use portable sites. Portable sites can be moved around to serve the best interests of the system. Downtown call traffic and highway call traffic usually occur at different times of the day. One or two portable sites can be moved around to cover two traffic patterns at two different times if needed.
4. Apply enhancers (repeaters). This is an economical way to extend coverage to peripheral fringe areas.

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Chapter

11

Switching and Traffic

11.1 General Description

11.1.1 General introduction

Switching equipment is the brain of the cellular system. It consists of two parts, the switch and the processor. The switch is no different from that used in the telephone central office. The processor used in cellular systems is a special-purpose computer. It controls all the functions which are specific for cellular systems, such as frequency assignment, decisions regarding handoff (including decisions regarding new cells for handoff), and monitoring of traffic. The smaller the cell, the more handoffs involved, and the greater the traffic load required. The processor can be programmed to correct its own errors and to optimize system performance. General (noncellular) telephone switching equipment is described first, and then cellular switching equipment is discussed in detail.

11.1.2 Basic switching[1]

In circuit switching (analog switching), a dedicated connection is made between input and output lines or trunks at the switching office and physical switching begins. Space-division switches have been generally used in circuit switching, but they can also be used for digital switching.

However, time-division switches can be used only for digital switching. In large digital switches, both time- and space-division switching are used.

Space-division switching. A rectangular coordinate switch interconnects n inputs from n lines to k outputs of other lines. When interconnection is possible at every cross-point, the switch in nonblocking.

$$\text{Number of cross-points} = nk \tag{11.1-1}$$

These switches are laid out in space as shown in Fig. 11.1.

For simplicity in analyzing the following nonblocking system, let the total number of input lines N equal the total number of output lines. An N-input group is fed into N/n switch arrays, and each switching network has an nk switch as shown in Fig. 11.2a. The three-stage switching network (S-S-S) is shown in Fig. 12.2b with N outputs. The total number of cross-points is[2]

$$C = 2\left(\frac{N}{n}\right)(nk) + k\left(\frac{N}{n}\right)^2 = 2\,Nk + k\left(\frac{N}{n}\right)^2 \tag{11.1-2}$$

If $k = 2n - 1$, the nonblocking rule, then the number of cross-points C can be obtained from Eq. (11.1-2).

Time-division switching. For time switching to be carried out, all call messages must be first slotted into time samples, such as in pulse-code modulation (PCM) for the digital form of voice transmission. Voice quality is sampled at 8000 samples per second. Each sample (125-μs frame) must have eight levels (2^3); then each voice channel requires 64 kbps (kilobits per second) of transmission.

The time-slot switchings limit the number of channels per frame that may be multiplexed. Multiplexing of more channels requires more memory storage. Take a t_f-microsecond frame and let t_c be the memory cycle time in microseconds required for both write-in and readout of a memory sample. Then the maximum number of channels that can be

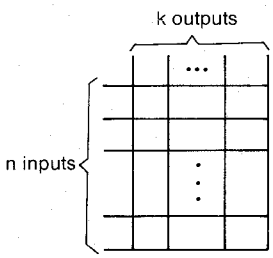


Figure 11.1 Number of cross-points in space-division switches.

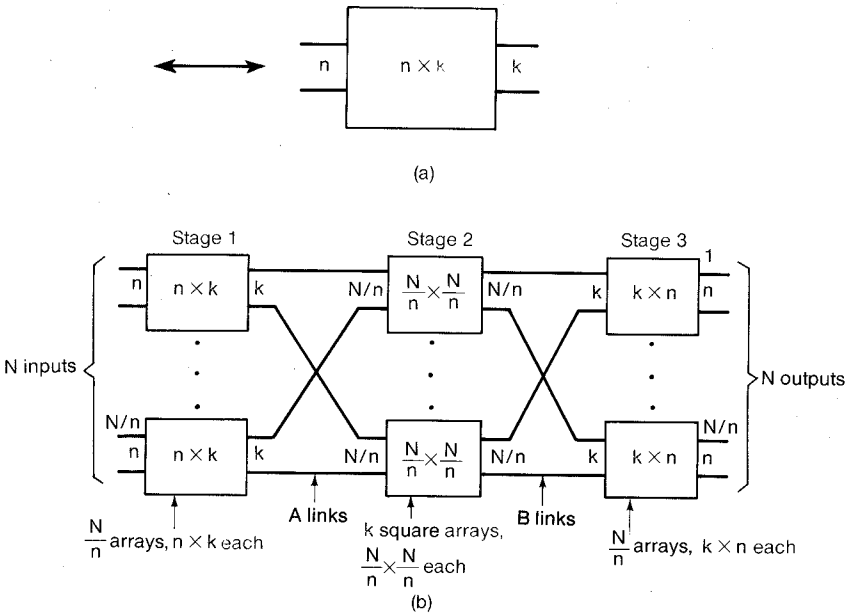


Figure 11.2 A switching network. (*a*) An *nK* switch; (*b*) a three-stage switching network.

supported is

$$N = \frac{t_f}{2\,t_c} \tag{11.1-3}$$

For t_c = 50-nsec logic (write-in–readout time) and t_f = 125 μs, the number N is 1250.

Blocking probability analysis of multistage switching.[3] In three-stage switching, the assumption of an S-S-S case (traversing two links) is the same as that of a T-S-T case. In the latter case, time division and space division are combined. The blocking probability that no free path from input channel to output channel is available is

$$P_B = [1 - (1 - p)^2]^k \tag{11.1-4}$$

where

$$p = \frac{An}{k} \tag{11.1-5}$$

where k = number of slot frames or number of outputs of each stage
n = number of input time slots or input lines of each stage
$A = \lambda/(\lambda + \mu)$ is the inlet channel utilization

We may also treat A as the probability that a typical channel is busy. The parameters λ and μ are explained in Sec. 9.5.

If $n = 120$, $k = 128$, and

$$A = \begin{cases} 0.7, \\ 0.9, \\ 0.95, \end{cases} \quad \text{then} \quad P_b = \begin{cases} 10^{-7} \\ 0.042 \\ 0.214 \end{cases}$$

This indicates the sensitivity of the blocking probability to the input utilization. However, when the input utilization reaches a certain level, the blocking probability must increase very rapidly.

Dimensions of switching equipment. Switching equipment must be designed to meet the projected growth rate of the system. The probability of loss P_b is

$$P_b \text{ (calls lost)} = \frac{(a^N/N!)}{\sum_{n=1}^{N} (a^n/n!)} \tag{11.1-6}$$

$$P_b \text{ (lost calls held)} = \frac{(a^N/N!)\,[N/(N-a)]}{\sum_{n=1}^{N} (a^n/n!)} \tag{11.1-7}$$

where N is number of trunks and a represents the traffic load, as in Sec. 9.6. Using either Eq. (11.1-6) or Eq. (11.1-7), we can find the number of trunks needed in a given demanding situation.

11.1.3 System congestion[4,5]

Time congestion. Consider a generic model in a circuit-switched exchange with a simple output trunk group. Each M input either is idle for an exponential length of time $1/\lambda$ or generates a call with a holding time of $1/\mu$. Each arriving call will be assigned to one of the outgoing trunks. For the probability of the number of calls in progress p_n, we obtain

$$p_n = \frac{(\lambda/\mu)^n \binom{M}{n}}{\sum_{n=0}^{N} (\lambda/\mu)^n \binom{M}{n}} \qquad 0 \leq n \leq N \tag{11.1-8}$$

If the system is fully occupied, then

$$P_B = P_N \tag{11.1-9}$$

This is called "time congestion."

Call congestion. The other way to measure congestion is to count the total number of calls arriving during a long time interval and record those calls that are lost because of a lack of resources, such as busy trunks.

The probability of call loss is P_L. Let $p(a)$ be the unconditional probability of arrival of a call, P_B be the probability that the system is blocked, and $p_N(a)$ be the probability that a call arrives when the system is blocked. Then

$$P_L = \frac{p_N(a)}{p(a)} P_B \tag{11.1-10}$$

If the conditional probability $p_N(a)$ is independent of the state of system blocking, then $p_N(a) = p(a)$, Eq. (11.1-10) becomes $P_L = P_B$, and the two measurements—time congestion and call congestion—are the same.

11.1.4 Ultimate system capacity

There are two limits on system capacity: (1) the amount of traffic that the switches can carry and (2) the amount of control that the processor can exercise without the occurrence of unacceptable losses. Limit 2, the amount of control that the processor can exercise without excessive losses, can be broken down as follows.

1. *Ultimate capacity due to traffic load.* Traffic capacity consists of two parameters, the number of calls per hour λ and their duration $1/\mu$. The average call duration is about 100 s (i.e., $1/\mu = 100$ s). The physical limits of switching capacity are reflected in the number of trunk interfaces N and the traffic load a.

2. *Ultimate capacity due to control.* Processor control operates on a delayed basis when requests are queued or scanned at regular intervals. There are two levels of control. At level 1, processor control is involved in scanning and interfacing with customers. At level 2, central processors are involved when all the data for a call request are received. The delay on the processor can be calculated as

$$\text{Average call delay (s)} = \frac{1/\mu}{2\,(1 - a')} \tag{11.1-11}$$

where a' is the traffic load on the processor and $1/\mu$ is the holding time that the processor takes to handle a call.

Ultimate system capacity limitations are reflected in limits on control, such as the number of calls that the system can handle, including handoffs, scanning and locating, paging, and assigning a voice channel. Therefore, the processing capacity for cellular mobile systems is much greater than that for noncellular telephone systems. In noncellular telephone switching the duration of the call is irrelevant, but in a cellular system it is a function of frequency management and the number of handoffs.

Assigning a value to the processor traffic

Level 2 control only (centralized system). It is extremely difficult to estimate accurately how a system will perform under real traffic conditions. A traffic simulation can be used. For total capacity, let

	P, %	$1 - P$, %
For eventualities	5	95
For false traffic (i.e., call abandoned before completion)	30	70
For peak traffic	30	70

Thus, $(1 - 0.05)(1 - 0.3)(1 - 0.3) = 0.465$ or a total capacity of 46.5 percent (or in this case 0.465 erlangs) for call processing.

It is assumed that no level 1 control is operating in the processor. Assume that the processor holding time is 100 ms; then applying this to Eq. (11.1-11), we find that the average call delay is

$$t_d = \frac{100 \text{ ms}}{2\,(1 - 0.465)} = 93.46 \text{ ms}$$

The average delay on call processing during the busy call equals

$$93.46 \times 0.465 = 43.46 \text{ ms}$$

Level 1 and level 2 control (decentralized system). Assume that level 1 control in a system reduces the load on level 2 control by absorbing most of the false traffic at level 1 control. Then at level 2 control the call-processing occupancy rate is $(1 - 0.05)(1 - 0.3) = 0.665$ (or 66.5 percent).

Assume that the total processor holding time is less than 100 ms for

the average call (the total time taken to handle a call is the sum of the total number of instructions required during the call); then

$$t_d = \text{average call delays} = \frac{100 \text{ ms}}{2\,(1 - 0.665)} = 150 \text{ ms}$$

The average delay on call processing during the busy call equals 150 ms × 0.665 = 100 ms.

When comparing the centralized system with the decentralized system, we find that in the centralized system the average call delay time is much shorter but its processing capacity is much less. However, the decentralized system is more flexible, is easier to install, and has a greater potential for expansion.

11.1.5 Call drops

Call drops are caused by factors such as (1) unsuccessful completion from set-up channel to voice channel, (2) blocking of handoffs (switching capacity), (3) unsuccessful handoffs (processor delay), (4) interference (foreign source), and (5) improper setting of system parameters.

The percentage of call drops is expressed as

$$P_{\text{cd}} = \frac{\text{number of call drops before completion}}{\text{total number of accepted calls handled by set-up channels}}$$

$$= \frac{\sum_{i=1}^{5} C_{d_i}}{C_t}$$

Because this percentage is based on many parameters, there is no analytic equation. But when the number of call drops increases, we have to find out why and take corrective action. The general rule is that unless an abnormal situation prevails call drops usually should be less than 5 percent.

11.2 Cellular Analog Switching Equipment

11.2.1 Description of analog switching equipment

Most analog switching equipment consists of processors, memory, switching network, trunk circuitry and miscellaneous service circuitry. The control is usually centralized, and there is always some degree of

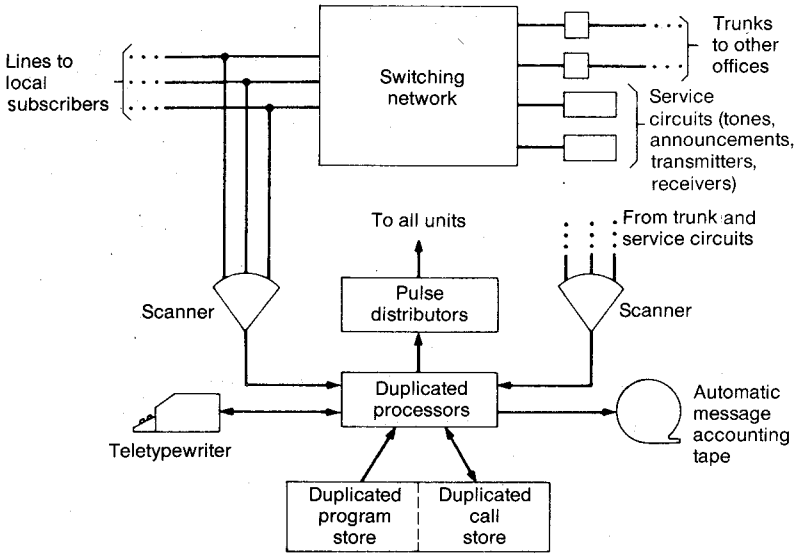


Figure 11.3 A typical analog switching system. (*After Chadha et al., Ref. 6.*)

redundancy. A common control system is shown in Fig. 11.3. The programs are stored in the memory that provides the logic for controlling telephone calls. The processor and the memory for programming and calls are duplicated. The switching network provides a means for interconnecting the local lines and trunks. The scanners are read under the control of the central processor. The changes in every connection at the line side and at the trunk side are also controlled by the processor. The central processor sends the order to all the units (switching network, trunk, service circuits) through pulse distributions. The automatic message accounting (AMA) tapes are used for recording the call usage. Three programs are stored in most switching equipment: (1) call processing (set up, hand off, or disconnect a call), (2) hardware maintenance (diagnose failed or suspected failed units), and (3) administration (collect customer records, trunk records, billing data, and traffic count).

11.2.2 Modification of analog switching equipment

The local line side has to change to the trunk side as shown in Fig. 11.4 because the mobile unit does not have a fixed frequency channel. Therefore, the mobile unit itself acts as a trunk line. In addition, the processors have to be modified to handle cellular call processing, the locating algorithm, the handoff algorithm, the special disconnect al-

gorithm, billing (air time and wire line), and diagnosis (radio, switching, and other hardware failure).

11.2.3 Cell-site controllers and hardware

Mobile telephone switching office (MTSO) system manufacturers designed their own cell-site controllers and transceivers (radios). Cell-site equipment is shown in Fig. 11.5. The cell site can be rendered "smarter," that is, programmed to handle many semiautonomous functions under the direction of the MTSO. Cell-site equipment consists of two basic frames.

1. Data frame—consists of controller and both data and locating radios
   a. Providing RF radiation, reception, and distribution
   b. Providing data communication with MTSO and with the mobile units
   c. Locating mobile units
   d. Data communication over voice channels
2. Maintenance and test frame
   a. Testing each transmitting channel for
      i. Incident and reflected power to and from the antenna
      ii. Transmitter frequency and its deviation
      iii. Modulation quality
   b. Testing each receiving channel for
      i. Sensitivity
      ii. Audio quality

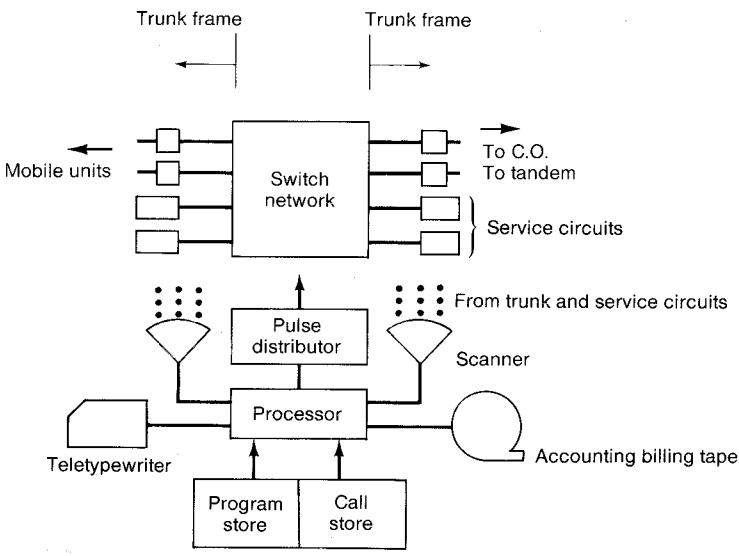


Figure 11.4 A modified analog switching system for cellular mobile systems.

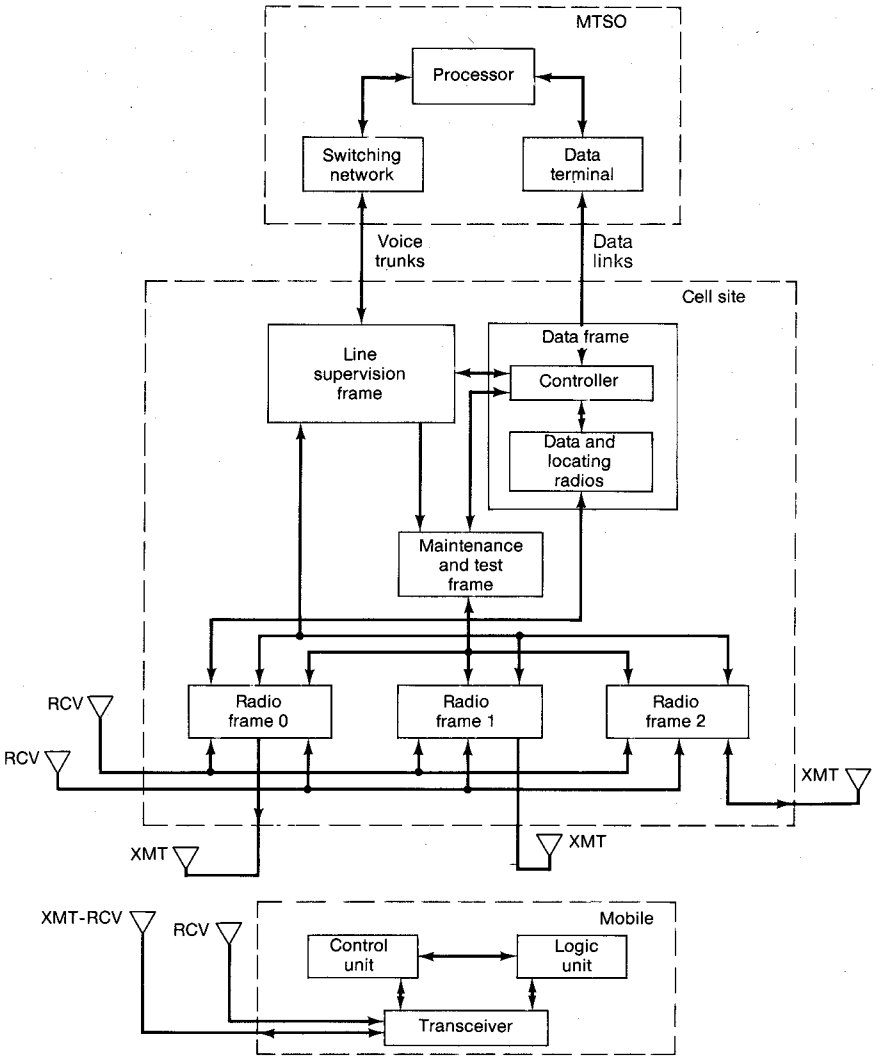


Figure 11.5 Cellular mobile major systems.

11.3 Cellular Digital Switching Equipment

11.3.1 General concept

The digital switch, which is usually the message switch, handles the digitized message. The analog switch, which is the circuit switch, must hold a call throughout the entire duration of the call. The digital (message) switch can send the message or transmit the voice in digital form; therefore, it can break a message into small pieces and send it at a

fast rate. Thus, the digital switch can alternate between ON and OFF modes periodically during a call. During the OFF mode, the switch can handle other calls. Hence, the call-processing efficiency of digital switching is higher than that of analog switching.

The future digital switch could be a switching packet which would send digital information in a nonperiodic fashion on request. There are other advantages to digital switching besides its greater efficiency.

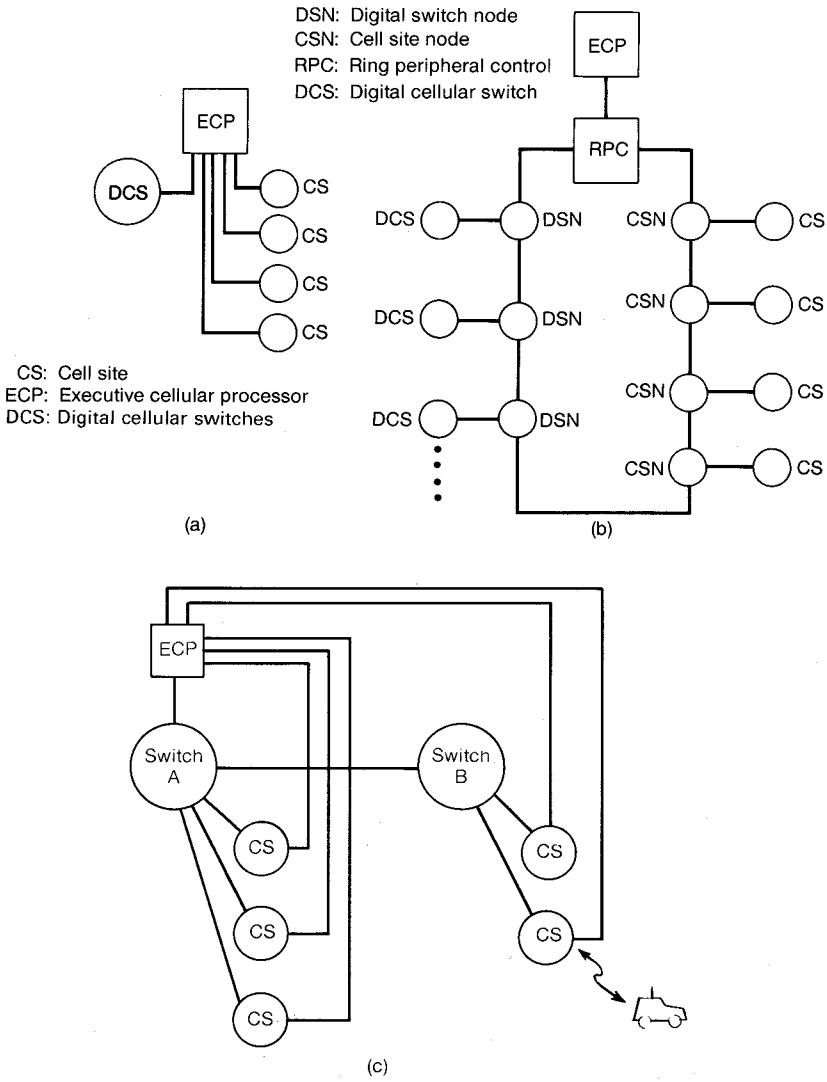


Figure 11.6 Cellular switching equipment. (*a*) A centralized system; (*b*) a decentralized system; (*c*) remote-controlled switching.

Digital switches are always small, consume less power, require less human effort to operate and are easier to maintain. Digital switching is flexible and can grow modularly. Digital switching equipment can be either centralized[7,8] (Fig. 11.6*a*) or decentralized[9,10] (Fig. 11.6*b*). A centralized system of a digital system has an architecture similar to that of an analog system. Motorola's EMX2500, Ericsson's AXE-10, and Northern Telecom International's DMS-MTXM are large centralized digital systems. A decentralized system is described here.

A decentralized system is slightly different from a remote-control switching system. In the remote-control switching system, a main switch is used to control a remote secondary switch as shown in Fig. 11.6*c*. In a decentralized system, all the switches are treated equally; there is no main switch.

11.3.2 Elements of switching

One decentralized switching system is introduced here for illustration. It is the American Telephone & Telegraph (AT&T) Autoplex 1000,[9] which consists of an executive cellular processor (ECP), digital cellular switches (DCS), an interprocess message switch (IMS), RPC (ring peripheral control), and nodes.

1. ECP transports messages from one processor to another.
2. IMS attached to a token ring (IMS uses a token ring technology) provides interfaces between ECP, DCS, cell sites, and other networks. The RPC attached to the ring permits direct communication among all the elements through the ring.
3. DCS, which are digital cellular switches, function as modules to allow the system to grow.
4. RPC forms a ring that connects two types of nodes: CSN (cell-site node) and DSN (digital switch nodes).
5. Nodes are: RPC nodes for connecting the ECP to the ring, CSN nodes for connecting cell-site data to the ring, and DSN nodes for connecting data links to the ring.

11.3.3 Comparison between centralized and decentralized systems

The analysis of overall system capacity given in Sec. 11.1.3 can be used here for comparison. In general, a centralized system has only one control level, whereas the decentralized system has more than one. In a one-level control system, the utilization of call processing is less than that in a multilevel control system. Moreover, the delay time for a central control system is always shorter than that for a decentralized

system; thus, the more levels of control, the greater the call-processing utilization and the longer the delay time. However, in principle, decentralized systems always have room to grow and are flexible in dealing with increasing capacity. Centralized systems deal with large traffic loads.

11.4 Special Features for Handling Traffic

The switching equipment of each cellular system has different features associated with the radios (transceivers) installed at the cell sites.

11.4.1 Underlay-overlay arrangement

The switching equipment treats two areas as two cells, but at a cocell site (i.e., two cells sharing the same cell site). Therefore, the algorithms have to be worked out for this configuration. In Sec. 8.5.4 we discussed the underlay-overlay arrangement in terms of channel assignment. Here we discuss this arrangement in terms of MTSO control.

To initiate a call, the MTSO must know whether the mobile unit is in an overlay or an underlay area. The MTSO obtains this information from the received signal strength tramsmitted by the mobile unit. To hand off a call, the MTSO must know whether this is a case of handoff from (1) an overlay area to an underlay area or (2) vice versa. In case 1 the signal strengthens and exceeds a specific level, and then the unconventional handoff takes place. In case 2 the signal strength weakens and falls below a specific level, and then the conventional handoff takes place.

11.4.2 Direct call retry

Direct call retry is applied only at the set-up channel. When all the voice channels of a cell are occupied, the set-up channel at that cell can redirect the mobile unit to a neighboring-cell set-up channel. This is the order used by the original set-up channel to override the "pick the strongest signal" algorithm. In this scheme, the MTSO receives all the call traffic information from all the cells and thus can distribute the call capacity evenly to all the cell sites.

11.4.3 Hybrid systems utilizing high sites and low sites

The high site is always used for coverage, and it can also be used to fill many holes which may be created by the low site. Therefore, if in some areas the mobile unit cannot communicate through the low site, the high site will take over, and as soon as the signal reception gets

better at the low site, the call will hand off to the low site. The algorithms include computation of the following configuration.

1. When the signal strength received at the cell site weakens, the handoff is requested and the high site picks it up.
2. The MTSO continues to check the signal of this particular mobile unit from all the neighboring low sites. If one site receives an acceptable signal from the mobile unit, the handoff will be forwarded to that site.

11.4.4 Intersystem handoffs

Intersystem handoffs were described in Chap. 9. The processor requires particular software to utilize this feature. There are four conditions of intersystem handoff, as shown in Fig. 11.7.

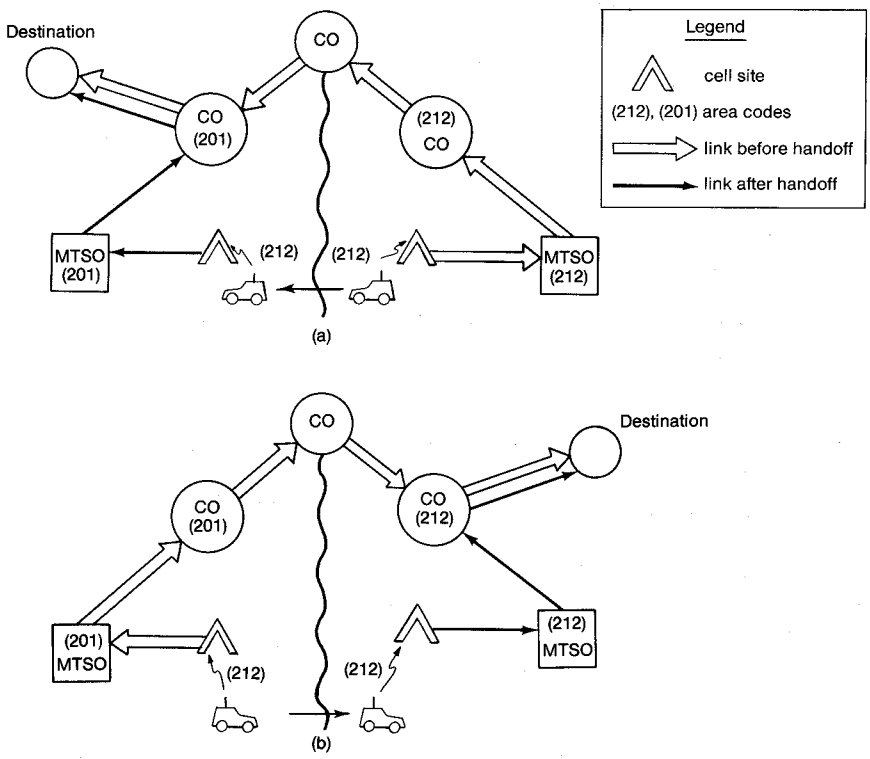


Figure 11.7 Four conditions of intersystem handoffs. (*a*) A toll call becomes a local call and the home mobile unit becomes a roamer; (*b*) a toll call becomes a local call and a roamer becomes a home mobile unit; (*c*) a local call becomes a toll call and the home mobile unit becomes a roamer; (*d*) a local call becomes a toll call and a roamer becomes a home mobile unit.

1. A long-distance call becomes a local call while a home mobile unit becomes a roamer (Fig. 11.7*a*).
2. A long-distance call becomes a local call while a roamer becomes a home mobile unit (Fig. 11.7*b*).
3. A local call becomes a long-distance call while a home mobile unit becomes a roamer (Fig. 11.7*c*).
4. A local call becomes a long-distance call while a roamer becomes a home mobile unit (Fig. 11.7*d*).

All four cases have to be implemented.

11.4.5 Queuing feature

When a nonuniform traffic pattern prevails and the call volume is moderate, the queuing feature can help to reduce the blocking probability. The improvement in call origination and handoffs as a result

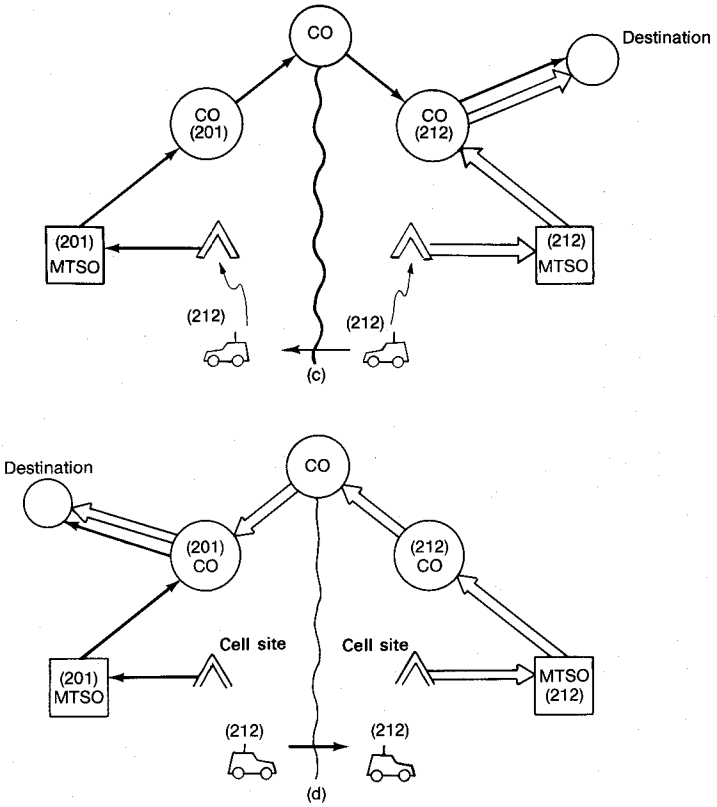


Figure 11.7 (*Continued*)

of queuing is described in Chap. 9. The switching system has to provide memory or buffers to queue the incoming calls if the channels are busy. The number of queue spaces does not need to be large. There is a finite number beyond which the improvement due to queuing is diminished, as described in Chap. 9.

11.4.6 Roamers

Initiating the call. If two adjacent cellular systems are compatible, a home mobile unit in system A can travel into system B and become a "roamer." The switching MTSO can identify a valid roamer and offer the required service. The validation can be the mobile unit's (MIN) or (ESN) (see Chap. 3).

Handing off the call. The feature of intersystem handoffs can be applied in order to continue the call. Intersystem handoffs are described in Sec. 11.4.4.

Clearinghouse concept. Because of the increase of roamers in each system, checking the validation of each roamer in the roamer's own system becomes a complex problem for an automatic roaming system. The cellular system "clearinghouses" (several nationwide companies) provide a central file of the validation of all users' MIN and ESN in every system. There are two files, positive and negative validation. Positive validation is done by checking whether the user's number is on the active customer list. The negative validation file lists the numbers of users whose calls should be rejected from the automatic roaming system. The payment for transmiting validation data to and from the clearinghouse plus the service charge has to be justified against the revenue lost through delinquent users (those who do not pay on time).

11.5 MTSO Interconnection

11.5.1 Connection to wire-line network

The MTSO operates on a trunk-to-trunk basis. The MTSO interconnection arrangement is similar to a private-branch exchange (PBX) or a class 5 central office (a tandem connection) (see Fig. 11.8). The MTSO has three types of interconnection links.

Type 1—interconnects a MTSO to a local-exchange carrier (LEC) end office

Type 2A—interconnects an LEC tandem office

Type 2B—interconnects to an LEC end office in conjunction with type 2A on a high-usage alternate-routing basis

The three-level hierarchy of a public telephone network is shown in Fig. 11.9. With this diagram, we can illustrate the three types of calls: (1) a local call, (2) an intra-LATA (local access and transport area) call, and (3) an inter-LATA call.

11.5.2 Connection to a cell site

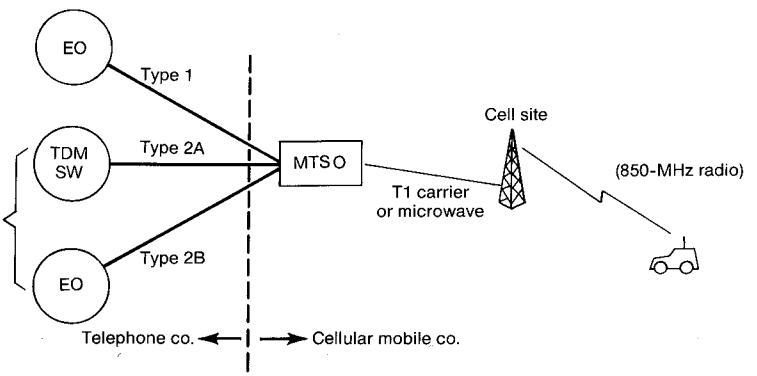
Two types of facility are used.

1. Cell-site *trunks* provide a voice communication path. Each trunk is physically connected to a cell-site voice radio. The number of trunks is decided on the basis of the traffic and the desired blocking probability (grade of service).
2. The cell site acts as a *traffic concentrator* for the MTSO. For instance, we may design an average busy-hour radio channel occupancy of at least 60 to 70 percent for high-traffic cells.

Both T1-carrier cables and microwave links are used. The duplication is needed for reliability.

11.6 Small Switching Systems[11–13]

Small switching equipment can be used in a small market (city). This switching equipment can usually be developed modularly. It consists



Type 1 interconnects an MTSO to an LEC end office

Type 2A interconnects an MTSO to an LEC tandem office

Type 2B interconnects to an LEC end office in conjunction with type 2A on a high-usage alternate-routing basis.

Figure 11.8 Three types of interconnection linkage.

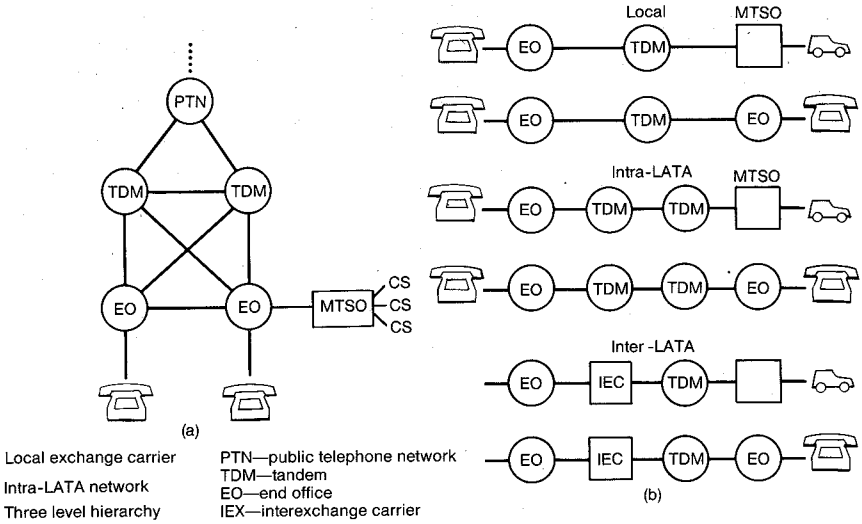


Figure 11.9 Three-level hierarchy. (*a*) Interconnection of MTSO; (*b*) three types of call.

of (1) a transmitter and a receiver, (2) a cell-site controller, (3) a local switch (a modified PABX should be used), (4) a channel combiner, and (5) a demultiplexer.

Small switches should be low-cost items. A high existing tower can be used for a cell-site antenna to cover a large area.

11.7 System Enhancement

Consider the following scenarios.

1. Each trunk is now physically connected to each voice radio as mentioned previously. But if the trunk can be switched to different radios, then the dynamic frequency assignment scheme can be accomplished.

2. Let a cell site pick up a switch in a decentralized multiple switching equipment system. This is a different concept than usual. Normally the switching equipment controls a cell site. For the land-initiated calls, the switching equipment picks up a cell site through paging. For mobile-originated calls, the cell site handling the call can select the appropriate switching equipment (DCS) from among the two or three units of switching equipment, assuming the cell site can detect the traffic conditions.

We can construct an analogy here. In a supermarket, everyone is waiting in line to pay the cashier for their merchandise. In order to reduce the waiting time, the store manager (central switching office)

can direct the customer to the cashier (switching equipment) with the shortest line or the customer (cell site) can select the cashier with the shortest line. Both methods can work equally well assuming both the manager and the customer have the ability to choose well.

This system enhancement may be made in the future to all systems when artificial intelligence techniques become fully developed and can be used to implement the enhancements.

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Chapter

12

Data Links and Microwaves

12.1 Data Links

Implementation of data links is an integral part of cellular mobile system design, and the performance of data links significantly affects overall cellular system performance.

The cell site receives the data from the MTSO to control the call process of mobile units. It also collects data from the reverse set-up channel from active mobile units and attempts to send it to the MTSO. There are three types of data links available: (1) wire line, (2) 800-MHz radios, and (3) microwaves. The following discussion describes each alternative and its advantages and disadvantages. The wireline connection[1] uses the telephone company's T1 carrier. Regular telephone wire can transmit only at a low rate (2.4 kbps); therefore, a high-data-rate cable must be leased. The T1 carrier has a wideband transmission (1.5 Mbps) that consists of 24 channels, and each channel can transmit at a rate of 64 kbps. For handling the data, a digital terminal converts the incoming analog signals to a digital form suitable for application in a digital transmission facility. Many digital terminals are multiplexed to form a single digital line called a *digital channel bank*.

Digital channel banks multiplex many voice-frequency signals and code them into digital form. The sampling rate is 8 kHz. Each channel is coded into 7-bit words. A signaling bit indicates the end of each 7-

bit sample. After 24 samples, one sample for each channel, a frame bit is sent again. The total number of bits per frame is

$$(7 \times 1) \times 24 + 1 = 193 \text{ bits/frame}$$

one sample

signaling bit

samples/channel

frame bit

Since 8000 frames per second and 193 bits per frame are specified, a digital capacity of 1.54 Mbps is required.

The data link has to have a data bank at each end of the T1 carrier cable. The data bank has to convert all the information into the 1.54 Mbps before sending it out through the cable. The number of T1 carriers required is determined by the number of radios installed at the cell site, e.g., if 60 radios are installed, then three T1-carrier cables are needed. The T1 carrier cables are installed in duplicate to provide redundancy (see Fig. 12.1). A major disadvantage to using wire-line data links is that the T1-carrier route may be rearranged by the telephone company at any time without notice. Therefore, it is not totally under the user's control. In addition, the leasing cost should be compared to the long-run cost of using the microwave link if owned by the cellular operator.

The data could also be sent by 800-MHz radios. However, this would cause interference among all the channels, and since every radio channel can handle a signaling rate of only 10 kbps, we would need an additional 666 channels just to handle this data link from the cell sites to the MTSO. This can be a good idea for low-capacity systems.

Microwave links seem to be most economical and least problematical. Details of their installation are given below. However in a rural area, capacity is not a problem. We can use half of the cellular channels for data-link use.

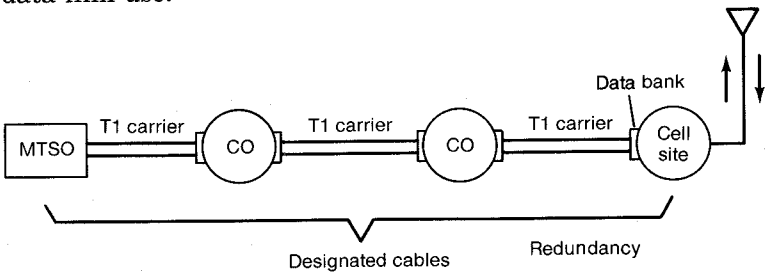


Figure 12.1 Data-link connection through cable.

12.2 Available Frequencies for Microwave Links

The microwave system is used to cover a large area; it should also be used as the "backbone." Before designing it, we must consider (1) system reliability, (2) economical design, (3) present and future frequency selection, (4) minimization of the number of new microwave sites, and (5) flexible and multilevel systems. The microwave frequencies can be grouped as follows.

Frequency, GHz	Allowed bandwidth, MHz	5-year channel loading	Minimum path length, km
2	3.5	None	5
4	20	900	17
6	30	900	17
11	40	900	5
18	220	None	None
23	100	None	None

As can be seen from this tabular analysis, for the higher frequencies there are fewer restrictions, thus allowing greater flexibility in system design.

The 2-GHz band. The minimum path length of 5 km (3.1 mi) and the limited 3.5-MHz radio-frequency (RF) bandwidth place several restrictions on the use of the 2-GHz band. Capacity is probably limited to eight T1 span lines. Installation of a 6- or 8-ft dish is required. Because of the limited path length, the limited traffic capacity, larger antennas at cell sites, and the difficulty in obtaining frequency coordination, 2 GHz is not desirable.

The 4- and 6-GHz bands. The minimum path length of 17 km (10.5 mi) and the minimum channel load of 900 channels for 6 GHz along with the 4-GHz frequency present a restriction.

The 11-GHz band. The minimum path length of 5 km (3.1 mi) and the minimum channel loadings of 900 voice channels would make this band a poor choice for the final path to the cell sites. However, the greater bandwidth availability and lower frequency congestion would make this an ideal band for high-density routes between the collection points and the MTSO.

The 18- and 23-GHz bands. The lack of FCC restrictions on minimum path length and the minimum channel loadings would appear to make these two frequency bands ideal for paths to the cell site. These frequency bands are not characterized by the presence of the RF congestion at lower frequencies. Cell sites can be implemented with 2- or 4-ft dishes, compared to the larger 6- or 8-ft dishes needed at lower frequencies.

12.3 Microwave Link Design and Diversity Requirement

There are three basic considerations here. First, the microwave propagation path length is always longer than the cellular propagation path, say, 25 mi or longer. Second, the path is always 100 or 200 ft above the ground. Third, the microwave transmission is a line-of-sight radio-relay link. Figure 12.2 shows the replacement of T1 carrier cable by microwave radios.

However, microwave links will render the system susceptible to one kind of multipath fading, in which the microwave transmission is affected by changes in the lower atmosphere, where atmospheric conditions permit multipath propagation.

Although deep fades are rare, they are sufficient to cause outage problems in high-performance communications systems.[2,3] A signal

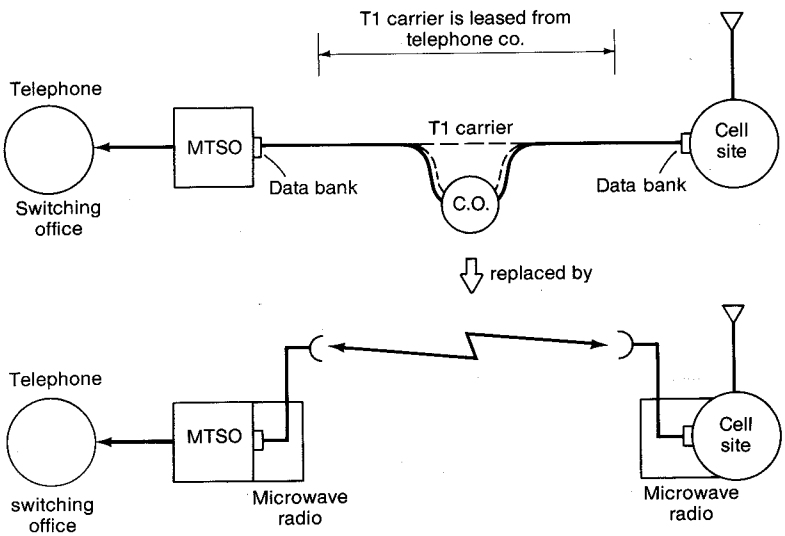


Figure 12.2 Replacement of T1 carrier cables by microwave radios.

is said to be in a fade of depth 20 log L dB; that is, the envelope (20 log l) of the signal is below the level L.

$$20 \log l \leqslant 20 \log L$$

Usually, we are interested only in fades deeper than -20 dB.

$$20 \log L < -20 \text{ dB} \qquad \text{or} \qquad L < 0.1$$

From the experimental data, the number of fades can be formulated as

$$N = \begin{cases} 6410L/60.88 \text{ days} = 105.29L/\text{day} & \text{(in the 6-GHz band)} \\ 3670L/60.88 \text{ days} = 60.28L/\text{day} & \text{(in the 4-GHz band)} \end{cases} \tag{12.3-1}$$

The average deviation of fades is

$$\bar{t} = \begin{cases} 490L \text{ s} & \text{(in the 6-GHz band)} \\ 408L \text{ s} & \text{(in the 4-GHz band)} \end{cases} \tag{12.3-2}$$

In order to reduce the fades, two methods can be used: a spaced diversity and a frequency diversity. In order to use diversity schemes, we have to gather some additional data. The two signals obtained individually from two channels can be used to measure the number of simultaneous fades, as shown in Fig. 12.3.

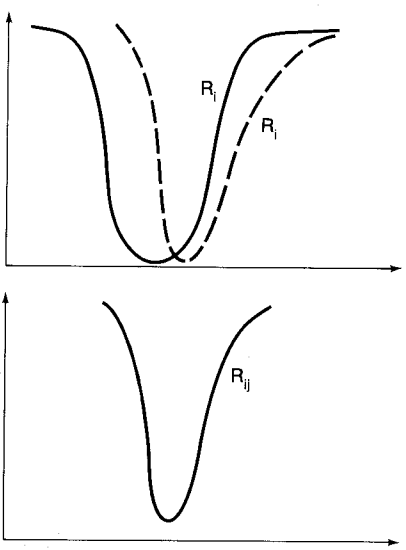


Figure 12.3 Formation of a simultaneous fade from two overlapped fades.

If the frequency separation or the spaced separation is very large, the number of simultaneous fades can be drastically reduced. Therefore, it follows that the number of simultaneous fades of two signals must be small to obtain good diversity reception.

A parameter F_N is defined as the ratio of N_i to N_{ij}, where N_i is the number of fades from a single channel and N_{ij} is the number of simultaneous fades from two individual channels.

$$F_N = \frac{N_i}{N_{ij}} \tag{12.3-3}$$

The ratio F_N should be large. For deep fades, F_N is

$$F_N = \tfrac{1}{2} q L^{-2} \qquad \text{for} \qquad L < 0.1 \tag{12.3-4}$$

The q is a parameter defined by the following equations.

1. For separations in frequency,

$$q = \frac{1}{4}\left(\frac{\Delta f}{f}\right) \qquad \text{(in the 6-GHz band)} \tag{12.3-5}$$

$$q = \frac{1}{2}\left(\frac{\Delta f}{f}\right) \qquad \text{(in the 4-GHz band)} \tag{12.3-6}$$

where Δf is the frequency separation and f is the operational frequency.

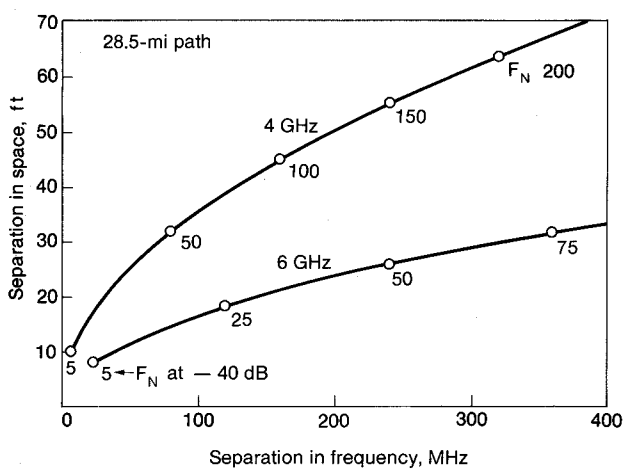


Figure 12.4 Separations in space and in frequency that provide equal values of F_N, the ratio of the number of fades to the number of simultaneous fades. (*After Vigants, Ref. 2.*)

2. For separation in space,

$$q = (2.75)^{-1}\left(\frac{s^2}{\lambda d}\right) \tag{12.3-7}$$

where s is the vertical antenna separation, λ is the wavelength, and d is the path length. All values are measured in the same units.

The term improvement F has been used to describe the ratio of the total time T_i spent in fades to the total time T_{ij} spent in simultaneous fades. For deep fades

$$F = \frac{T_i}{T_{ij}} \approx 2F_N \qquad L < 0.1 \tag{12.3-8}$$

The total fading time T_i in a year in a 6-GHz propagation can be obtained from Eqs. (12.3-1) and (12.3-2) as

$$T_i = N\bar{t} = \begin{cases} (37904.4L/\text{year})\ (490L\ \text{s}) & \\ 1857\ \text{s/year} & \text{at} \quad L = 0.01\ (\text{or } -40\ \text{dB}) \\ 31\ \text{min/year} & \text{at} \quad L = 0.01\ (\text{or } -40\ \text{dB}) \end{cases}$$

We can use the same step to obtain the total fading time T_i in 4-GHz propagation.

$$T_i = N\bar{t} = \begin{cases} (22002\ L/\text{year})\ (408L\ \text{s}) & \\ 897.68\ \text{s/year} & \text{at} \quad L = 0.01\ (\text{or } -40\ \text{dB}) \\ 15\ \text{min/year} & \text{at} \quad L = 0.01\ (\text{or } -40\ \text{dB}) \end{cases}$$

The values of F_N at -40 dB are shown along the curves in Fig. 12.4 for 4 and 6 GHz, respectively. To achieve $F_N = 5$ in 4 GHz, we must use a separation in vertical spacing of *10 ft,* or the frequency separation should be *8 MHz.*

To achieve $F_N = 5$ in 6 GHz, we must use a vertical antenna separation of 9 ft, or the frequency separation should be 25 MHz. The improvement F can be obtained from Eq. (12.3-8) as $F = 2 \times 5 = 10$, or the total fading time after a diversity scheme at $L = 0.01$ is reduced to

$$T_{ij} = \begin{cases} 3.1\ \text{min/year} & (\text{in the 6-GHz band}) \\ 1.5\ \text{min/year} & (\text{in the 4-GHz band}) \end{cases}$$

12.4 Ray-Bending Phenomenon[4]

This phenomenon occurs because air is denser at lower levels than at higher levels. Starting with Snell's law for two layers with different

refractive indices n_1 and n_2, we obtain

$$\frac{\sin\theta_1}{\sin\theta_2} = \frac{n_2}{n_1} = \frac{\sqrt{\mu_2\varepsilon_2}}{\sqrt{\mu_1\varepsilon_1}} = \frac{C_1}{C_2} \tag{12.4-1}$$

The other parameters are explained as follows.

1. If θ_2 is the reflection angle (Fig. 12.5) and $n_1 = n_2$, then

$$\sin\theta_1 = \sin\theta_2 \tag{12.4-2}$$

Snell's law indicates that the incident angle θ_1 is equal to the reflected angle θ_2. Also assume that there are no conductivity effects in the atmosphere and that the troposphere is not magnetic: $\mu_1 = \mu_2 = \mu_0$. For layer 1, the dielectric constant is ε_1 and the corresponding velocity C_1 of wave propagation is

$$C_1 = \frac{1}{\sqrt{\mu_0\varepsilon_1}}$$

and for layer 2, the dielectric constant is ε_2 and the velocity of wave propagation is C_2.

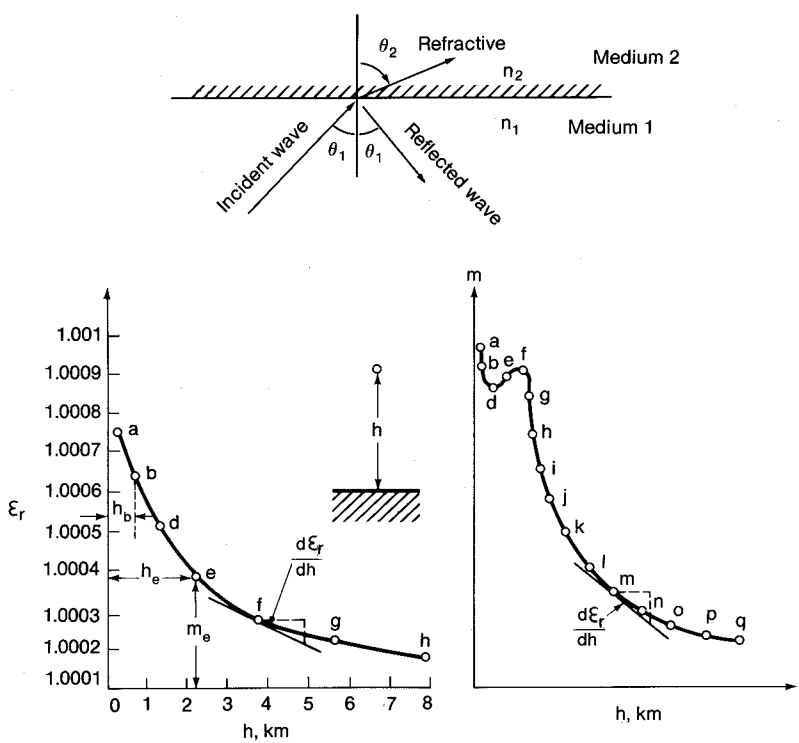


Figure 12.5 Illustration of Snell's law and finding dn/dh from a measured piece of data. (*After Hund, Ref. 4.*)

2. If the θ_2 is the refraction angle (see Fig. 12.5) for the wave transmission into medium 2, then

$$n_1 \sin \theta_1 = n_2 \sin \theta_2 \tag{12.4-3}$$

Assume that the refraction indexes n_1, n_2, n_3 decrease as the altitude h_1, h_2, h_3 increases. Then

$$\frac{\sin \theta_1}{\sin \theta_2} = \frac{n_2}{n_1} \tag{12.4-4}$$

$$\frac{\sin \theta_2}{\sin \theta_3} = \frac{n_2}{n_3} \tag{12.4-5}$$

for a gradient dn/dh of n, and $n \sin \theta$ = constant. Then at a certain altitude h, the index of refraction is

$$n = \sqrt{\varepsilon_r} \tag{12.4-6}$$

At the altitude $h + dh$

$$n + \left(\frac{dn}{dh}\right) dh = \sqrt{\varepsilon_r + \left(\frac{d\varepsilon_r}{dh}\right) dh} \tag{12.4-7}$$

The $d\varepsilon_r/dh$ can be found from a measured (or statistically predicted) curve at a given location (see Fig. 12.5). Then dn/dh can be found. The equation for ray bending is expressed as

$$\frac{1}{\rho} = -\frac{1}{n}\frac{dn}{dh} \tag{12.4-8}$$

Now ρ is the radius of curvature which can be calculated as dn/dh

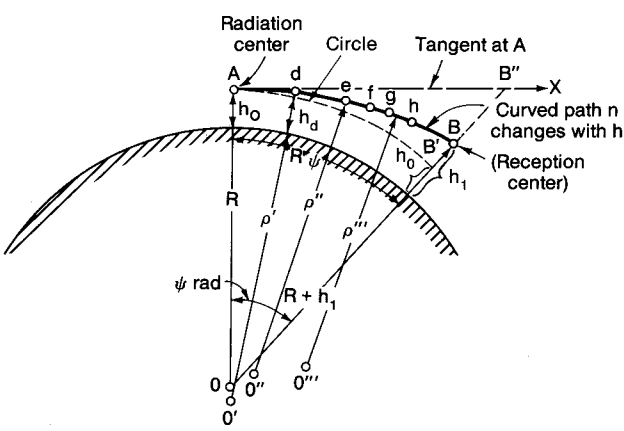


Figure 12.6 The radius of curvature of the wave path. (*After Hund, Ref. 4.*) Curved path *AB* requires the least time to move electromagnetic wave energy from a through angle ψ to line *OB"*.

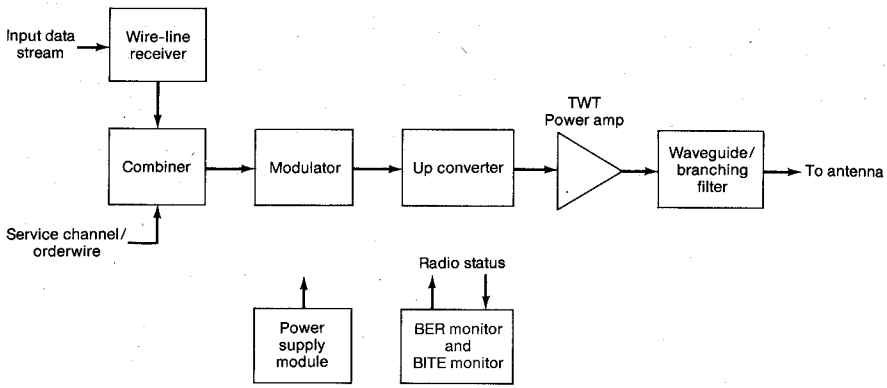


Figure 12.7 Microwave radio transmit block diagram.

changes as a result of $d\varepsilon_r/dh$. The radius ρ can be plotted by computer as shown in Fig. 12.6.

12.5 System Reliability

The microwave radio link is a stand-alone system. A typical system layout is shown in Fig. 12.7 for a transmitter and in Fig. 12.8, for a receiver.

12.5.1 Equipment reliability

All radio equipment should be redundant (i.e., equipped with duplicates) with a standby and automatic switchover in case of failure. An alarm system is available for reporting an emergency condition at any microwave site to the central alarm station at the MTSO. Redundant power converters have been included at each cell site. A space diversity can be implemented for further increasing system reliability.

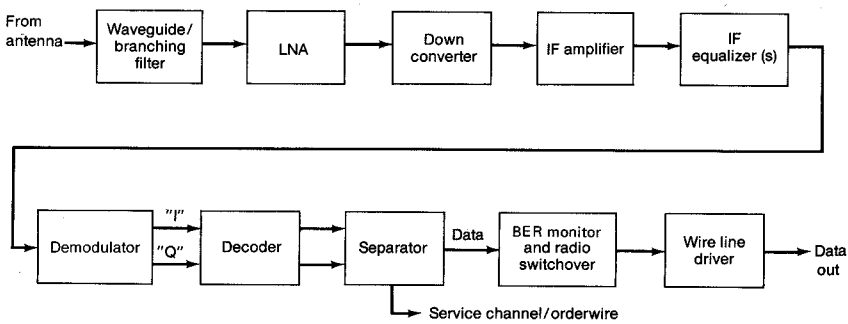


Figure 12.8 Microwave radio receive block diagram.

12.5.2 Path reliability

The microwave path should be a clear line-of-sight path between two points. Each path should be calculated and studied by field survey. Sometimes larger antenna size, higher tower, shorter distance, more diversity, or greater capacity are required to increase the path reliability. An important consideration is elimination or reduction of the multipath reception at the receiver as a result of the reflection along the path. The reflected energy would be negligible if the reflector were out of the first Fresnel zone, which is H

$$H \geqslant \sqrt{\frac{\lambda d_1(d - d_1)}{d}} \tag{12.5-1}$$

where d_1 and d_2 are as shown in Fig. 12.9 and λ is the operating wavelength. Five special cases are as follows.

Hyperreflectivity. Hyperreflectivity may occur, such as in wave propagation over water, metal objects, and large flat surfaces. In these cases, additional path clearance is recommended.

Bending. Since the earth is curved and because dielective permittivity varies with height, bending occurs. On the average, the radio wave is bent downward, i.e., the earth radio wave acts as if the earth's radius were four-thirds of its real value (see Fig. 12.10). The effective radius for K (ratio of effective earth radius to true earth radius) can be any

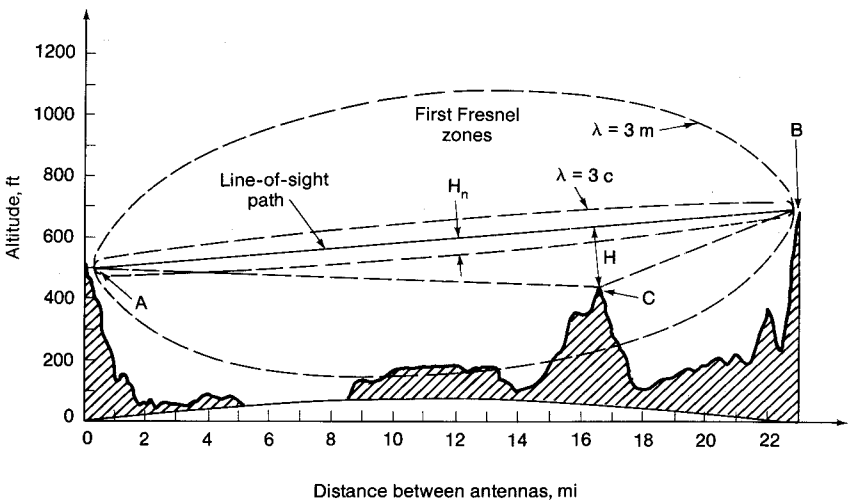


Figure 12.9 The clearance distance from the closest objects. (*From Ref. 8, p. 439.*) Typical profile plot showing first Fresnel zones for 100 MHz and 10 GHz.

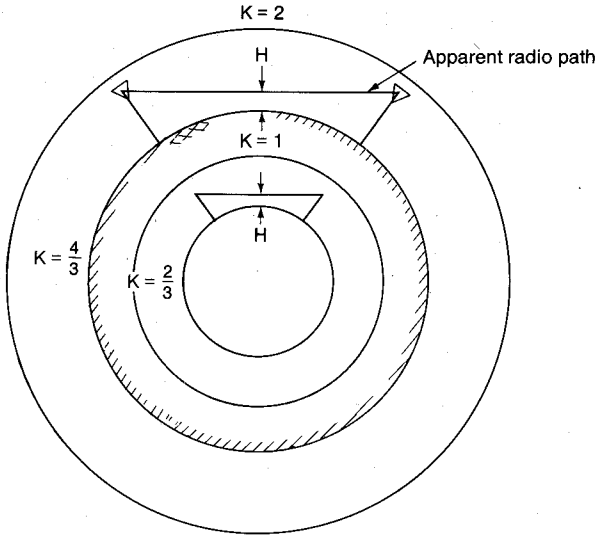


Figure 12.10 Illustration of earth bulge conditions.

value other than $K = 4/3$ and can be treated as a function of atmospheric conditions. Sometimes it can be as low as one-half for a small percentage of time. If we base our calculations on $K = 4/3$, then the ray is curved downward. This is called an "earth bulge" condition. It will cause the path loss to increase over a wide range of frequencies unless adequate path clearance is provided. A value of $K = 1$ would indicate that the earth is completely flat or the ray travels in a straight line. On the other hand, the wave based on $K = 2/3$ will tilt upward and may cause interference over a long distance. An earth bulge factor of $2/3$ or $4/3$ is used to provide a clear area.

$$H \geqslant \frac{d_1 d_2}{2} \qquad \text{(for a factor of } 4/3\text{)} \tag{12.5-2}$$

$$H \geqslant d_1 d_2 \qquad \text{(for a factor of } 2/3\text{)} \tag{12.5-3}$$

where d_1 and d_2 are in miles and H is in feet.

High microwave frequencies. For a microwave frequency above 10 GHz, the oxygen, water vapor, and rain attenuate (or scatter) the microwave beam. To determine the total path attenuation, we must add the free-space loss (FSL) to the rain loss while considering the anticipated rain rates. The rain rate is measured by millimeters per hour. Usually, a

rain rate of 15 mm/h or greater will be considered as heavy rain. Some areas, such as Florida, may have a great deal of precipitation. Some areas, such as southern California, are arid. The history of rain-rate data can be obtained from the U.S. Weather Bureau's annually accumulated rain statistics collected since 1953 in 263 cities. The author was the first to suggest this method at Bell Laboratories in 1972. (Several rain-rate models are given in Refs. 5, 6, and 7.) Once the rain statistics of each city are known, the decibels per kilometer for different rain rates can be found at each operating frequency as shown in Fig. 12.11. The rain rate is governed by the size and shape of the raindrops. The path loss varies with both the raindrop size and rain rate.

The effects of haze, fog, snow, and dust are insignificant. The size of the rain cell (a rain-occupied area) will be considered along the microwave link. The heavier the rain, the smaller the rain cell. Also, the rain-rate profile will be nonuniform. To calculate a microwave link, we need to know the (1) link gain—power, antenna size, antenna height, and receiver sensitivity; (2) free-space loss; (3) attenuation due to a

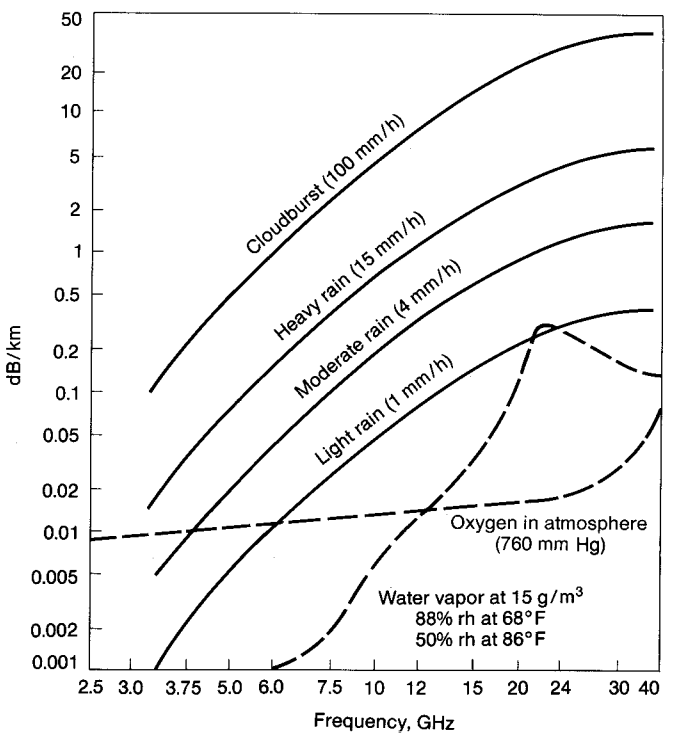


Figure 12.11 Estimated atmospheric absorption. (*From Ref. 8, p. 443.*)

predicted rain rate; and (4) given availability—allowable downtime, such as 1 h in a year or 10^{-4}.

The transmission rate of a signal over a microwave link is limited to the time-delay spread, and the time-delay spread is based on the distance. Usually we design the link primarily on the basis of the rain effect; therefore, the link is usually short because of the rain attenuation, and the time-delay spread at the shorter distance is not considered.

Power system reliability. Battery systems or power generators are needed for the microwave systems in case of power failure. Usually a 24-V dc battery system will be installed with 8 to 10 h of reserve capacity.

Microwave antenna location. Sometimes the reception is poor after the microwave antenna has been mounted on the antenna tower. A quick way to check the installation before making any other changes is to move the microwave antenna around within a 2 to 4 ft radius of the previous position and check the reception level. Surprisingly favorable results can be obtained immediately because multipath cancellation is avoided as a result of changing reflected paths at the receiving antenna.

Also, at any fixed microwave antenna location, the received signal level over a 24-h time period varies.

12.6 Microwave Antennas[8]

12.6.1 Characteristics of microwave antennas

Microwave antennas can afford to concentrate their radiated power in a narrowbeam because of the size of the antenna in comparison to the wavelength of the operating frequency; thus, high antenna gain is obviously desirable. Some of the more significant characteristics are discussed in the following paragraphs.

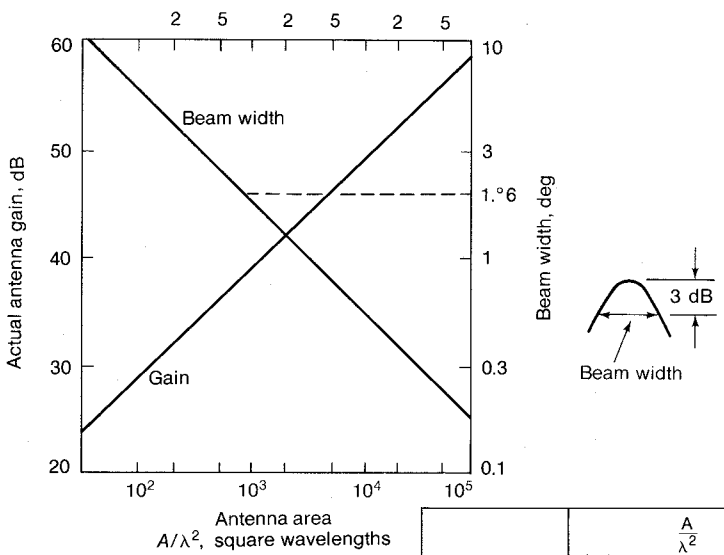
Beamwidth. The greater the size of the antenna, the narrower the beamwidth. Usually the beamwidth is specified by a half-power (3-dB) beamwidth and is less than 10° at higher microwave frequencies. The beamwidth sometimes can be less than 1°. The narrowbeam can reduce the chances of interference from adjacent sources or objects such as adjacent antennas. However, a narrowbeam antenna requires a fair amount of mechanical stability for the beam to be aimed at a particular direction. Also, the problem of antenna alignment due to the ray-bending problems discussed earlier restricts the narrowbeam antenna to a certain degree. The relationship between gain and beamwidth is depicted in Fig. 12.12.

Sidelobes. The sidelobes of an antenna pattern would be the potential source of interference to other microwave paths or would render the antenna vulnerable to receiving interference from other microwave paths.

Front-to-back ratio. This is defined as the ratio of the maximum gain in the forward direction to the maximum gain in the backward direction. The front-to-back ratio is usually in the range of 20 to 30 dB because of the requirement for isolating or protecting the main transmission beam from interference.

Repeater requirement. The front-to-back ratio is very critical in repeaters because the same signal frequencies are used in both directions at one site. An improper design can cause a ping-pong ringing type of oscillation from a low front-to-back ratio or from poor isolation between the transmitting port and receiving port of the repeater.

Side-to-side coupling loss. The coupling loss, in decibels, should be designed to be high as a result of the transmitting antenna carrying only the output signal and the receiving antenna receiving only the



Antenna	$\frac{A}{\lambda^2}$ 4 GHz	6 GHz	11 GHz
8-ft paraboloid	8×10^2	2×10^3	6×10^3
4-ft paraboloid	5×10^2	5×10^2	1.5×10^3
8-ft square	1×10^3	2.5×10^3	8×10^3

Note: Abscissa is actual antenna area, and "actual antenna gain" is taken to be 3 dB below theoretical.

Figure 12.12 Approximate antenna gain and beamwidth. (*From Ref. 8, p. 445.*)

incoming signal. If the transmitting and receiving antennas are installed side by side, the typical transmitter outputs are usually 60 dB higher than the receiver input level. Longer link distance results in increased values. Therefore, the coupling losses must be high in order to avoid internal system interference. The space separation between two antennas and the filter characteristics in the receiver can be combined with a given antenna pattern to achieve the high coupling loss.

Back-to-back coupling. The back-to-back coupling loss also should be high (e.g., 60 dB) between two antennas. Two antennas are installed back to back, one transmitting and one receiving. However, it is much easier to reach a high back-to-back coupling loss than a side-to-side coupling loss.

12.6.2 Polarization and space diversity in microwave antennas

Polarization. To reduce adjacent-channel interference, microwave relay systems can interleave alternate radio-channel frequencies from a horizontal polarized wave to a vertical polarized wave.

The same approach can be applied to the left- and right-handed circularly polarized waves, but the beamwidths of antennas for this orthogonal system are relatively large and therefore are not attractive.

In the polarization system, the *cross-coupling loss* is specified. This loss is defined as the ratio of the power received in the desired polarization to the power coupling into other polarization. The cross-coupling loss (isolation) should be as high as possible. Usually 25 to 30 dB is required for one hop.

Space diversity.[9] The two antennas separated vertically or horizontally as described in Sec. 12.3 can be used for a two-branch space-diversity arrangement. In a space-diversity receiver, the required reception level is relatively low so that the transmitted power on the other end of the link can be reduced. This is also an effective method for increasing the coupling loss between the transmitting antenna and receiving antenna.

12.6.3 Types of microwave-link antenna

Two kinds of antenna are used for microwave links.

1. A parabolic dish, used for short-haul systems. Antennas sizes range from 1.5 m (5 ft) to 3 m (10 ft) in diameter.
2. A horn-reflector antenna, to trap the energy outward from the focal point. The advantages of using this antenna are
   a. Good match—return loss 40–50 dB.

TABLE 12.1 Horn-Reflector Antenna Characteristics

	4 GHz		6 GHz		11 GHz	
Frequency polarization	Vertical	Horizontal	Vertical	Horizontal	Vertical	Horizontal
Midband gain, dB	39.6	39.4	43.2	43.0	48.0	47.4
Front-to-back ratio, dB	71	77	71	71	78	71
Beamwidth (azimuth), degrees	2.5	1.6	1.5	1.25	1.0	0.8
Beamwidth (elevation), degrees	2.0	2.13	1.25	1.38	0.75	0.88
Sidelobes, dB below main beam	49	54	49	57	54	61
Side-to-side coupling, dB	81	89	120	122	94	112
Back-to-back coupling, dB	140	122	140	127	139	140

b. Broadband—a horn antenna can work at 4, 6, and 11 GHz.
c. One horn can be used for two polarizations with high cross-coupling loss.
d. Small sidelobes—high back-to-back coupling loss.

The gains, coupling losses, and beamwidths are listed in Table 12.1 for different frequencies and different polarizations.

12.6.4 Installation of microwave antennas

A microwave antenna cannot be installed at any arbitrary location. Selection of an optimum position is very important. In many situations if we cannot move horizontally, we can move vertically. In a microwave-link setup, there are two fixed effective antenna heights, one at each end based on each reflection plane where the reflection point is incident on it. The gain of the received signal also relates to the two effective antenna heights if they are low. The antenna location can be moved around to find the best reception level. Sometimes it is worthwhile to take time to search for the location that gives the best reception.

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Chapter

13

System Evaluations

13.1 Performance Evaluation

13.1.1 Blockage

There are two kinds of blockage: set-up channel blockage and voice-channel blockage.

Set-up channel blockage B_1. Information regarding set-up channel blockage cannot be obtained at the cell site because the mobile unit will be searching for the busy/idle bit of a forward set-up channel in order to set up its call. If the busy bit does not change after 10 call attempts in 1 s, a busy tone is generated, and no mobile transmit takes place. In another case the mobile transmit takes place as soon as the idle bit is shown. Several initiating cells can intercollide at the same time. When it occurs, the mobile unit counts it as one seizure attempt. If the number of seizure attempts exceeds 10, then the call is blocked. This kind of blockage can be detected only by mobile phone users. If the occurrence of blockage of the system is in doubt, each of the three specified set-up channels can be assigned in each of the three sectors of a cell, and the total number of incoming calls among the three sectors can be compared with that from a single set-up channel (omni). It should be determined whether there is a difference between two call-completion numbers, one from a single set-up channel and the other from three set-up channels. This is one way to check the blockage if

the single set-up channel seems too busy. The set-up channel blockage should be at least less than half of the specified blockage (usually 0.02) in the mobile cellular system.

If all the call-attempt repeats are independent events, then the resultant blocking probability B_1 after n attempts is related to the blocking probability of the single call attempt B, as

$$B_1 = 1 - (1 - B) \sum_{i=0}^{n} B^i = B^n \tag{13.1-1}$$

Example 13.1 Assume that the blocking probability of a set-up channel is .005, and the holding time at the set-up channel is 175 ms per call. There is only one channel; then the offered load (from Appendix 1.1) a is .005. Thus the number of set-up calls being handled is

$$C = \frac{.005 \times 3600 \times 1000}{175} = 120 \text{ calls} \qquad \text{(one call attempt)}$$

Example 13.2 All parameters are the same as in Example 13.1, except that the offered load a changes to $a = 0.02$. Then the number of set-up calls is

$$C = \frac{.02 \times 3600 \times 1000}{175} = 480 \text{ calls} \qquad \text{(one call attempt)}$$

Example 13.3 Given the number of set-up calls per hour, find the blocking probability B_t after 10 call attempts in 1 s.

Consider the following cases.

Case 1. Assume that there are two set-up calls per second or 7200 calls per hour. Since each set-up call takes 175 ms, the offered load a is

$$a = \frac{175 \times 2}{1000} = .35$$

The blocking probability B (see Appendix 1.1) is $B = .25$ (assuming one call attempt).

Since the average interval for each attempt is 100 ms, 10 attempts have to be completed in 1 s. It is a kind of conditional probability problem. In the worst case, a mobile unit has to fail the tenth call attempt before giving up. During this period, because of the failure of all call attempts, the two set-up calls from other mobile units should have been successful with a probability of 1. The length of two set-up calls is 350 ms, which is roughly the time interval required for four attempts; i.e., these four attempts are definitely blocked with a blocking probability of 1 and should not be counted as attempts. Therefore, only six attempts count. Using Eq. (13.1-1), we obtain

$$B_1 = (.25)^6 = .00012$$

which is quite low and, of course, acceptable.

Case 2. Assume that there are three set-up calls per second or 10,800 calls per hour. Then

$$a = \frac{175 \times 3}{1000} = .525 \qquad \text{(offered load)}$$

$$B = .342 \qquad \text{(blocking probability; see Appendix 1.1)}$$

Since three set-up calls take 525 ms, roughly six out of ten attempts are definitely blocked following the same argument stated in case 1. Only four attempts count, then the resultant blocking probability is

$$B_t = (.342)^4 = .013$$

which is too high for the set-up channel.

Voice-channel blockage B_2. Voice-channel blockage can be evaluated at the cell site. When all calls come in, some are refused for service because there are no available voice channels. Suppose that we are designing a voice channel blockage to be .02. On this basis, $B_2 = .02$, and after determining the holding time per call[1] and roughly estimating the total number of calls per hour at the site,[2] we can find the number of radios required.

Example 13.4 Assume that 2000 calls per hour are anticipated. The average holding time is 100 s per call, and the blocking probability is .02 (2 percent). Then the offered load is

$$a = \frac{2000 \times 100 \text{ s}}{60 \times 60 \text{ s}} = 55.5 \text{ erlangs}$$

Use $a = 55.5$ and $B_2 = 0.02$ to find $N = 66$ channels required (refer to Appendix 1.1).

The actual blocking probability data must be used to check the outcome from the Erlang B model (Appendix 1.1). Although the difference can be up to 15 percent, the Erlang B model is still considered as a good model for obtaining useful estimates.

End-office trunk blockage B_3. The trunks connecting from the MTSO to the end office can be blocked. This usually occurs when the call traffic starts to build up and the number of trunks connected to the end office becomes inadequate. Unless this corrective action is taken, the blockage during busy periods increases. An additional number of trunks could be provided at the end office when needed.

The total blockage B_t. As the total call blockage is the result of all three kinds of blockage, the total blockage is

$$B_t = B_1 + B_2(1 - B_1) + B_3(1 - B_1)(1 - B_2)$$
$$= 1 - (1 - B_1)(1 - B_2)(1 - B_3) \qquad (13.1\text{-}2)$$

Example 13.5 Assume that $B_1 = .01$ and $B_2 = B_3 = .02$. Then the total blockage is

$$B_t = .01 + .0198 + .0194 = .0492 \approx 5\%$$

The result in Example 13.5 indicates that even when each individual blockage (i.e., B_1, B_2, and B_3) is small, the total blockage becomes very large. Therefore, the resultant blockage is what we are determining.

13.1.2 Call drops (dropped-call rate)

Call drops are defined as calls dropped for any reason after the voice channel has been assigned. Sometimes call drops due to weak signals are called *lost calls*. The dropped-call rate is partially based on the handoff-traffic model and partially based on signal coverage.

The handoff traffic model. A new handoff cell site treats handoffs the same way as it would an incoming call. Therefore, the blockage for handoff calls is also $B = .02$. Some MTSO systems may give priority to handoff calls rather than to incoming calls. In this case the blocking probability will be less than .02.

A warning feature can be implemented when the call cannot be handed off and may be dropped with high probability, enabling the customer to finish the call before it is dropped. Then the dropped-call rate can be reduced.

The loss of SAT calls. If the mobile unit does not receive a correct SAT in 5 s, the mobile-unit transmitter is shut down. If the mobile unit does not send back a SAT in 5 s, the transmitter at the cell site is shut down. In both cases the call is dropped. If the correct SAT cannot be detected at the cell site, as in cases of strong interference, then (1) the SAT can be offset by more than 15 Hz (see section entitled "The total dropped-call rate," below) or (2) the SAT tone generator in the mobile unit may not produce the desired tone.

Calculation of SAT interference conditions. The desired SAT is $\cos w_1 t$, and the undesired SAT is $\rho \cos w_2 t$. When $\rho \ll 1$, the SAT detector at the cell site can easily detect w_1. When ρ is greater and starts to approach 1, SAT interference occurs. The following analysis shows the

degree of the interference due to the value of ρ.

$$\cos w_1 t + \rho \cos w_2 t = A(t) \cos \theta(t) \tag{13.1-3}$$

where

$$A(t) = \sqrt{1 - \rho^2 + 2\rho \cos (w_1 - w_2)t} \tag{13.1-4}$$

$$\psi(t) = w_2 t - w_1 t$$

$$\theta = w_1 t + \tan^{-1} \frac{\rho \sin \psi(t)}{1 + \rho \cos \psi(t)} \tag{13.1-5}$$

$$w = \frac{d\theta}{dt} = w_1 + \frac{w_2 - w_1}{([1 + \rho \cos \psi(t)]/\{\rho[\rho + \cos \psi(t)]\}) + 1} \tag{13.1-6}$$

Let $\cos \psi(t) = 1$ in Eq. (13.1-6), the extreme condition of w which is the offset frequency from a desired SAT.

$$w = w_1 + \left(1 + \frac{1}{\rho}\right)^{-1} (w_2 - w_1) \tag{13.1-7}$$

For

$$\rho = \begin{cases} .3 & w = w_1 + .22(w_2 - w_1) \\ .5 & w = w_1 + .333(w_2 - w_1) \\ .75 & w = w_1 + .45(w_2 - w_1) \end{cases} \tag{13.1-8}$$

If $w_2 - w_1 = 30$ Hz, for two adjacent SATs

$$\rho = \begin{cases} .5 & \omega = \omega_1 \pm 9.95 & \text{(acceptable)} \\ .75 & \omega = \omega_1 \pm 12.9 & \text{(marginal)} \end{cases}$$

If $\omega_2 - \omega_1 = 60$ Hz, for two ends of SATs

$$\rho = \begin{cases} .3 & \omega = \omega_1 \pm 13.8 & \text{(marginal)} \\ .5 & \omega = \omega_1 \pm 19.9 & \text{(unacceptable)} \\ .75 & \omega = \omega_1 \pm 25.8 & \text{(unacceptable)} \end{cases}$$

Adjacent SATs cannot interfere with the desired SAT for an undesired SAT level below $\rho = .75$ (meaning a level of -2.5 dB). However, when two SATs which are not adjacent to each other, the undesired SAT level should at least be lower than $\rho = 0.3$ (-10.5 dB) in order for no interference to occur.

Unsuccessful complete handoffs. Because of the limitations in processor capacity, the duration of the handoff process may occasionally be too long, and the mobile unit may not be informed of a new channel to be handed off.

The total dropped-call rate. Assume that the handoff blocking is B_4, the probability of lost SAT calls is B_5, and the probability of an unsuccessful complete handoff is B_6. Then the total drop call rate is

$$B_d = B_4 + B_5 (1 - B_4) + B_6 (1 - B_4) (1 - B_5) \qquad (13.1\text{-}9)$$

Usually, the dropped-call rate should be less than 5 percent.

13.1.3 Voice quality

It is very important that the voice quality of a channel be tested by subjective means. Some engineers try to use the signal-to-noise-plus-distortion ratio (SINAD) to evaluate voice quality. Although SINAD is an objective test, using it to test voice quality may result in misleading conclusions. Worst of all, engineers are always proud of the apparatus they have designed, and they tend to ignore others' opinions. To serve the public interest, we must survey consumers for their opinions.

Evaluation of system performance based on a subjective test can be used to set performance criteria. As was discussed in Chap. 1, 75 percent of cellular phone users report that voice quality is good (CM4; i.e., circuit merit 4) or excellent (CM5) over 90 percent of the service area. These numbers can vary depending on how well the service is performed—in other words, the cost. First we must know what kind of service we are providing to the public. We should then let the customers judge the voice quality.

A typical curve of subjective tests was shown in Fig. 7.1*b* for different conditions. For this particular run, the mobile speed is 56 km/h (35 mi/h).

13.1.4 Performance Evaluation

Often we encounter a situation where two systems are being installed in two different areas; one system is deployed in a flat area where the average measured bit error rate (BER) is low at the cell site on the basis of bit-stream data received from mobile-units, and the other system is deployed in a hilly area where BER is relatively high. Of course, if the same degree of skill was used to install both systems, the system in the hilly area would otherwise be inferior and would provide the

poorer performance. But if each system claims to be better than the other, how can we judge these two systems fairly? One way is to compensate for variations in performance due to geographic location.[2] Assume that system A is deployed in a flat terrain and system B, in a hilly area. Both systems use the same kind of antennas and run the measured data with the same kind of mobile units. Then the handicap of building a system in a hilly area can be compensated for before comparing this system's performance with others.

Another way is to set the BER to, say, 10^{-3} and find the percentage of areas where the measured BER is greater than 10^{-3} in relation to the area of the whole system, then compare the percentages from the two different systems in two different areas.

Here the two pieces of raw field-strength recorded data in the area should be received from two mobile units moving around in two systems, respectively. From these data, the two local means (the envelope of the raw field-strength data) can be obtained. Then the two cumulative probability distributions (CPD), $P(x < X)$ of two local means, can be plotted. Let us first normalize the average power at a 50 percent level because the local means have a log-normal distribution, and the differences in transmitted power in the two systems are factored out. There will be two straight lines on log-normal scale paper (see Fig. 13.1). The standard deviations of two log-normal curves can be found

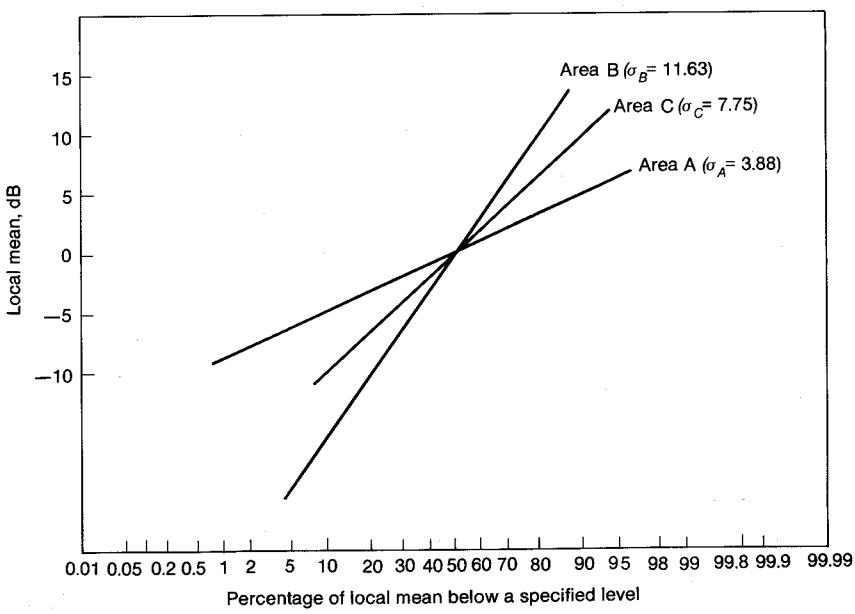


Figure 13.1 Different local mean statistics in different areas.

from Fig. 13.1 by determining a level X where $P(x \leq X) = 90$ percent.

$$X = \begin{cases} 5 \text{ dB} & \text{for system A} \\ 15 \text{ dB} & \text{for system B} \end{cases}$$

A normal distribution can be found from published mathematical tables indicating that $P(x \leq 1.29) = 90$ percent, where

$$x \leq \frac{X - m}{\sigma} = \frac{X}{\sigma} = 1.29 \tag{13.1-10}$$

where $m = 0$ dB and σ is the standard deviation. For

$$X = \begin{cases} 5 \text{ dB} & \sigma_A = \dfrac{5}{1.29} = 3.88 \text{ dB} \\ 15 \text{ dB} & \sigma_B = \dfrac{15}{1.29} = 11.6 \text{ dB} \end{cases}$$

The standard deviations of A and B indicate that system B covers a relatively hillier area than system A.

Example 13.6 From these measurements we find that the BER in 10 percent of area A in system A is greater than 10^{-3}. The BER in 25 percent of area B in system B is greater than 10^{-3}. The local means of two systems are shown in Fig. 13.1. We would like to judge which system has a better performance. From Fig. 13.1, 10 percent of the area in system A corresponds to 10 percent of the total signal below -5 dB. If system A is deployed in area C ($\sigma_c = 7.75$ dB), 25 percent of area C will have a BER greater than 10^{-3} below the -5-dB level. Now the variance σ_B of area B is greater than the variance σ_C of area C; thus, area B is more difficult to serve than area C. However, the measurement shows that 25 percent of area B in system B is greater than 10^{-3} below the -9-dB level. Therefore, according to the criterion BER $\geq 10^{-3}$, system B proves to be a better system than system A.

13.2 Signaling Evaluation

The signaling protocols of existing systems are evaluated in this section. The signaling format of the forward control channel (FOCC) as deduced from a BCH code (63, 51) becomes a short code of (40, 28) or that of the reverse control channel (RECC) from a BCH becomes a short code of (48, 36), as described in Chap. 3. The 12 parity-check bits always remain unchanged. This BCH code can correct one error and detect two errors.

13.2.1 False-alarm rate

The false-alarm rate is the rate of occurrence of a false recognizable word that would cause a malfunction in a system. The false-alarm rate

should be less than 10^{-7}. Now we would like to verify that the BCH code (40, 28) can meet this requirement. The Hamming distance d of BCH (40, 28) is 5. This means that in every different code word at least 5 out of 40 bits are different. Then the false-alarm rate FAR can be calculated as

$$FAR = p_e^d(1 - p_e)^{L-d} \tag{13.2-1}$$

where p_e is the BER and d is the length of a word in bits.

Assume that in a noncoherent frequency-shift-keying (FSK) modulation system, the average BER of a data stream in a Rayleigh fading environment is

$$\langle p_e \rangle = \frac{1}{2 + \Gamma} \qquad \text{(noncoherent FSK)} \tag{13.2-2a}$$

and the average BER of a data stream received by a differential phase-shift-keying (DPSK) modulation system in the same environment is

$$\langle p_e \rangle = \tfrac{1}{2}\left(\frac{1}{\Gamma + 1}\right) \qquad \text{(DPSK)} \tag{13.2-2b}$$

where Γ is the carrier-to-noise ratio. Let* $\Gamma = 15$ dB; then we obtain BER from Eq. (13.2-2) as

$$\langle p_e \rangle = .03 \qquad \text{(noncoherent FSK)}$$

$$\langle p_e \rangle = .015 \qquad \text{(DPSK)}$$

Substituting $p_e = .03$, which is the higher BER, into Eq. (13.2-1), we obtain

$$FAR \approx (.03)^5 = 2.43 \times 10^{-8} \leqslant 10^{-7}$$

This meets the requirement that $FAR < 10^{-7}$.

13.2.2 Word error rate consideration

The word error rate (WER) plays an important role in a Rayleigh fading environment. The length of a word of an FOCC is $L = 40$ and the transmission rate is 10 kbps. Then the transmission time for a 40-bit word is

$$T = \frac{40}{10{,}000} = 4 \text{ ms}$$

* The C/N ratio of a data channel can be lower than 18 dB of a voice channel.

From Sec. 1.6.6, the average duration of fades can be obtained from the following assumptions: frequency = 850 MHz, vehicle speed = 15 mi/h, and threshold level = −10 dB (10 dB below the average power level); then the average duration of fades is

$$\bar{t} = 0.33 \times \left(\sqrt{2\pi}\,\frac{V}{\lambda}\right)^{-1} = 7 \text{ ms} \tag{13.2-3}$$

Equation (13.2-3) shows that the transmission time of one word is shorter than the average duration of fades while the vehicle speed is 15 mi/h; that is, the whole word can disappear under the fade. Therefore, redundancy schemes are introduced. From Chap. 3, the FOCC format is

200 bits	word A (40 bits × 5 times),	28 information bits
200 bits	word B (40 bits × 5 times),	28 information bits
10 bits	bit synchronization	
11 bits	word synchronization	
42 bits	Busy/Idle-status bits	
463 bits		56 information bits

The throughput can be obtained from

$$\frac{56}{463} = \frac{1200 \text{ bps}}{10{,}000 \text{ bps}}$$

Therefore, the throughput is 1200 bps (baseband rate).

13.2.3 Word error rate calculation

The WER can be calculated as follows. We may use a DPSK system because it has a general but simple analytic formula, more general than Eq. (13.2-2*b*)

$$\langle p_e \rangle = \tfrac{1}{2}\left(\frac{1}{\Gamma + 1}\right)^{M} \tag{13.2-4}$$

where M is the number of diversity branches. It is difficult to obtain the WER from the correlation coefficient in the bit stream at a specific vehicle speed because the correlation coefficients of any two bits among all the bits in a word at that particular speed form a correlation coefficient matrix which is difficult to handle. Fortunately, we can find two extreme values, one at the speed $V \to \infty$ and the other at the speed $V - \to 0$. The calculations are described in detail in Ref. 3. Here we are simply illustrating the results.[4]

The performance of word error rates is shown in Fig. 13.2 for two cases: (1) no error correction and (2) one error correction. We have noticed that without redundancy (no repeat), the WER of a fast-fading case is worse than that of a slow-fading case. The WER obtained from a finite speed will lie between these two curves.

When a redundancy scheme is applied (Fig. 13.2*a*), i.e., repeating K times and making a majority voting on each bit, the WER of a slow-fading case becomes worse than that of a fast-fading case. This change in WER provides a great improvement in performance. Therefore, a redundancy scheme is of value in a mobile radio environment with a variable vehicle speed. This phenomenon is also illustrated in Fig. 13.2*b* for a 1-bit error correction code. Figure 13.3*a* and *b* shows the WER for two-branch diversity. Figure 13.3*a* is WER with no error correction code, and Fig. 13.3*b* is WER with one error correction code. Further improvements are seen in the figure. The error correction code and the diversity plus the redundancy provide a desired signaling performance.

13.2.4 Parity check bits

In this section, we illustrate the generation of parity check bits in a word. Let a word of (7, 4) be generated with 4 bits of information and a 3-bit parity check. The word matrix $[C]$ can be expressed as

$$[C] = [x_m]\,[G] \tag{13.2-5}$$

where $[x_m]$ is the information matrix and $[G]$ is the generation matrix. Let $[G]$ have the following form.

$$[G] = \begin{bmatrix} 1000 & 101 \\ 0100 & 111 \\ 0010 & 110 \\ \underbrace{0001}_{\text{Identity}} & \underbrace{001}_{\text{Parity}} \end{bmatrix} = [IP] \tag{13.2-6}$$

The parity matrix of three bits can be arranged in any order and have any combination of 0s and 1s. For example, let $[x_m] = 1001$, then substituting $[x_m]$ and Eq. (13.2-6) into Eq. (13.2-5) yields

$$[C] = [1001] \begin{array}{c} \begin{array}{cccc} & 5\text{th} & 6\text{th} & 7\text{th} \end{array} \\ \begin{bmatrix} 1000 & 1 & 0 & 1 \\ 0100 & 1 & 1 & 1 \\ 0010 & 1 & 1 & 0 \\ 0001 & 0 & 1 & 1 \end{bmatrix} \end{array} = [1001\ C_5C_6C_7]$$

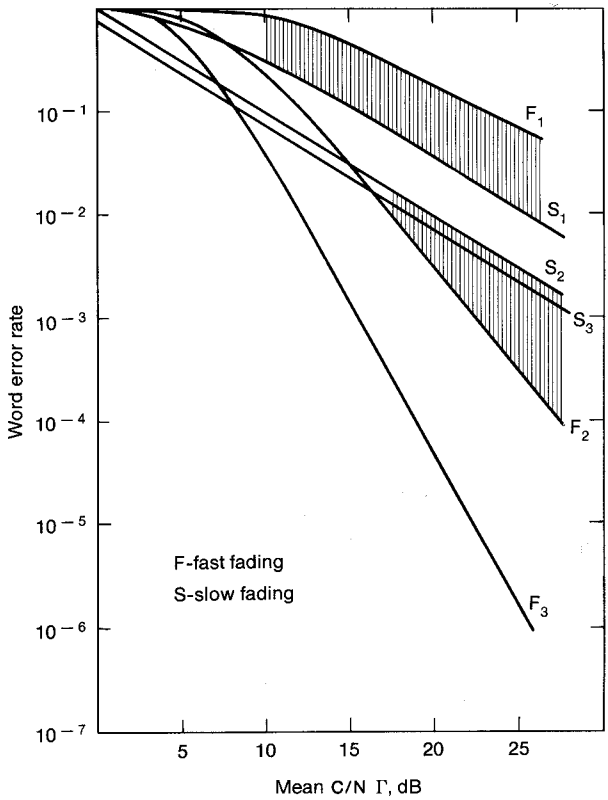
Then C_5 can be obtained by multiplying the fifth column by $[x_m]$ and applying modulo 2 addition* as

$$C_5 = 1 \cdot 1 + 0 \cdot 1 + 0 \cdot 1 + 1 \cdot 0 = 1 + 0 + 0 + 0 = 1$$

The same process is applied to C_6 and C_7 as

$$C_6 = 0 + 0 + 0 + 1 = 1 \qquad C_7 = 1 + 0 + 0 + 1 = 0$$

Therefore the information fits (1001), along with the three parity check bits, become a word (1001110).



(a)

Figure 13.2 Word error rate for $N = 40$ bits. Number of branches $M = 1$; S_1, F_1, no repeat ($K = 1$); S_2, F_2 two-thirds voting ($K = 3$); S_3, F_3 three-fifths voting ($K = 5$). (*a*) Case 1: $M = 1$, $t = 0$. No error correction. (*b*) Case 2: $M = 1$, $t = 1$. One-bit error correction.

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* Modulo 2 additions:

$$1 + 1 = 0 \qquad 1 + 0 = 1 \qquad 0 + 0 = 0 \qquad 0 + 1 = 1$$

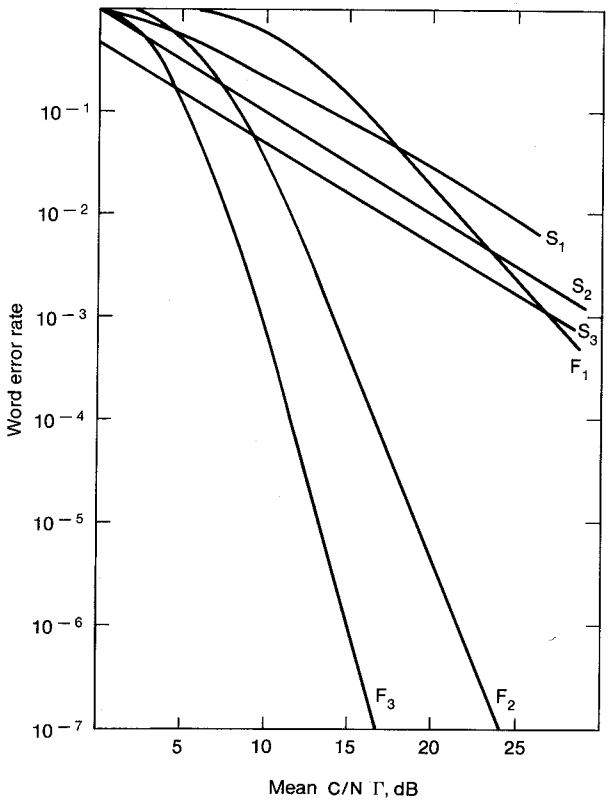
13.3 Measurement of Average Received Level and Level Crossings

13.3.1 Calculating average signal strength[5]

The signal strength can be averaged properly to represent a true local mean $m(x)$ to eliminate the Rayleigh fluctuation and retain the long-term fading information due to the terrain configuration. Let $\hat{m}(x)$ be the estimated local mean. If a length of data L is chosen properly, $\hat{m}(x)$ will approach $m(x)$ as

$$\hat{m}(x) = \frac{1}{2L}\int_{x-L}^{x+L} r(y)\,dy = \frac{1}{2L}\int_{x-L}^{x+L} m(y)r_0(y)\,dy$$

$$= m(x)\left[\frac{1}{2L}\int_{x-L}^{x+L} r_0(y)\,dy\right] = m(x) \qquad (13.3\text{-}1)$$



(b)

Figure 13.2 (*Continued*)

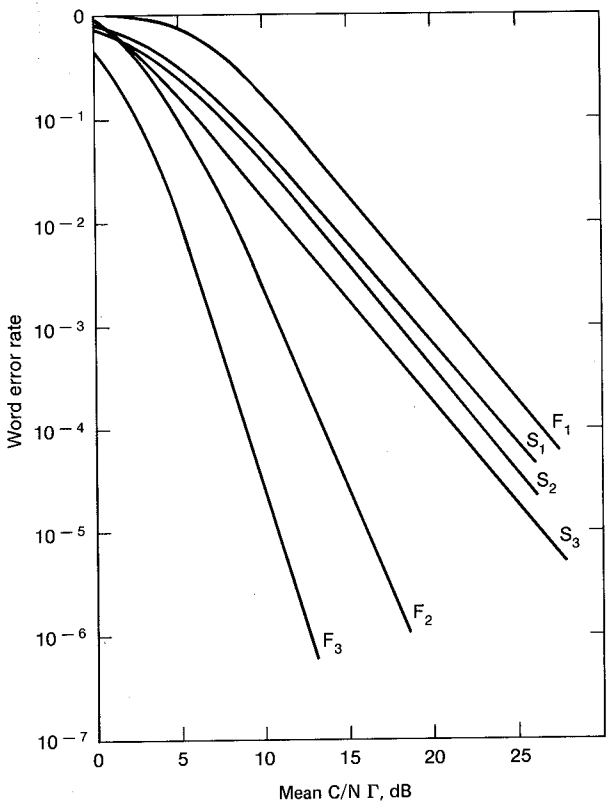
or

$$\frac{1}{2L}\int_{x-L}^{x+L} r_0(y)\,dy \to 1 \tag{13.3-2}$$

where $r_0(y)$ is a Rayleigh distributed variable. If the value of Eq. (13.3-2) is close to 1, then $\hat{m}(x)$ is close to $m(x)$. The spread of $\hat{m}(x)$, denoted as $\sigma_{\hat{m}}$, can be expressed as

$$1\sigma_{\hat{m}} \text{ spread} = 20 \log \frac{m(x) + \sigma_{\hat{m}}}{m(x) - \sigma_{\hat{m}}} \quad \text{in dB} \tag{13.3-3}$$

Equation (13.3-3) is plotted in Fig. 13.4. The $1\sigma_{\hat{m}}$ spread is used to indicate the uncertainty range of a measured mean value from a true mean value if the length of the data record is inadequate.



(a)

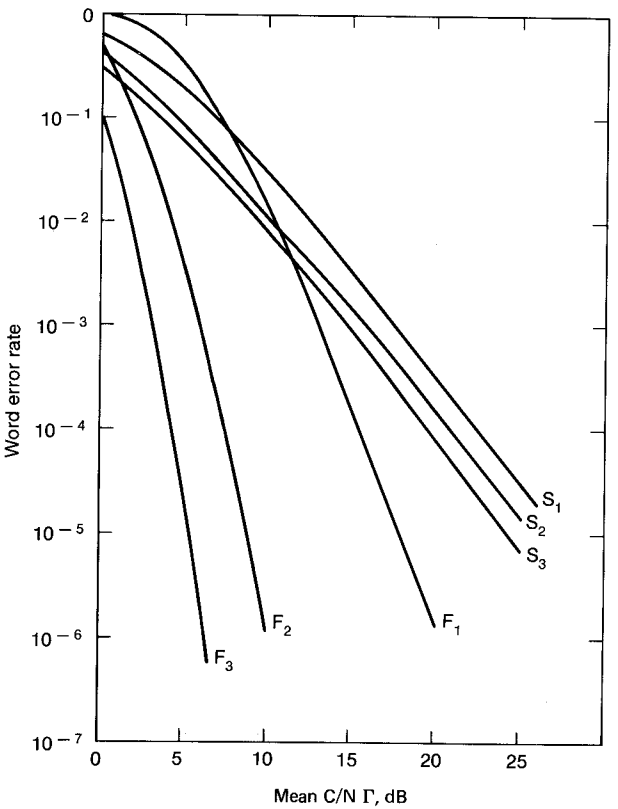
Figure 13.3 Word error rate for $N = 40$ bits. Number of branches $M = 2$; S_1, F_1, no repeat ($K = 1$); S_2, F_2 two-thirds voting ($K = 3$); S_3, F_3 three-fifths voting ($K = 5$). (*a*) Case 3: $M = 2$, $t = 0$. No error correction. (*b*) Case 4: $M = 2$, $t = 1$. One-bit error correction.

The proper length 2*L*. If we are willing to tolerate $1\sigma_{\hat{m}}$ spread in a range of 1.56 dB, then $2L = 20\lambda$. If the tolerated spread is in a range of 1 dB, then $2L = 40\lambda$.

For length $2L$ less than 20 wavelengths, the $1\sigma_{\hat{m}}$ spread begins to increase quickly. When length $2L$ is greater than 40λ, the $1\sigma_{\hat{m}}$ spread decreases very slowly.

In addition, the mobile radio signal contains two kinds of statistical distributions: $m(y)$ and $r_0(y)$. If a piece of signal data $r(y)$ is averaged, we find that if the length is shorter than 40λ, the unwanted $r_0(y)$ may be retained whereas at lengths above 40λ smoothing out of long-term fading $m(y)$ information may result. Therefore, 20 to 40λ is the proper length for averaging the Rayleigh fading signal $r(y)$.

Sampling average.* As mentioned previously, when using the averaging process with a filter, it is difficult to control bandwidth even when the length of the data to be integrated is appropriate. Therefore,



(b)

Figure 13.3 (*Continued*)

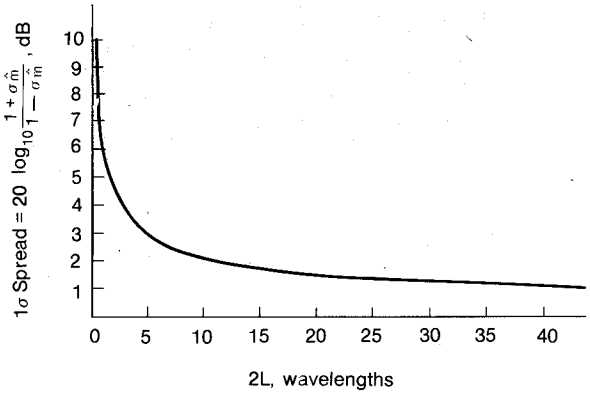


Figure 13.4 The value of $1\sigma_{\hat{m}}$ spread.

the sample values of $r(t)$ are used for sampling averaging instead of analog (continuous waveform) averaging. Then we must determine how many samples need to be digitized across a signal length of $2L$ (see Fig. 13.5). The number of samples taken for averaging should be as small as possible. However, we have to calculate how many sample points are needed for adequate results. We set a confidence level of 90 percent and determine the number of samples required for the sampling average. The general formula is

$$P\left(-1.65 \leqslant \frac{\bar{r}_j - \hat{m}_j}{\hat{\sigma}_j} \leqslant 1.65\right) = 90\% \tag{13.3-4}$$

Figure 13.5 Sample average over a $2L = 40\ \lambda$ of data.

* The detailed derivation is shown in W. C. Y. Lee, *Mobile Communications Design Fundamentals,* Howard W. Sams & Co., 1986, Sec. 2.2.2.

Let $\hat{m}_j$ and $\hat{\sigma}_j$ be the mean and the standard deviation of ensemble average* $\bar{r}_j$ of jth interval ($2L$) and r_j be a gaussian variable

$$\hat{\sigma} = \frac{\sigma_r}{N} \qquad \hat{m} = m$$

where m and σ_r are the mean and the standard deviation of a Rayleigh sample r. N is the number of samples. Therefore,[5]

$$m = \frac{\sqrt{\pi}}{2}\sqrt{\overline{r^2}} \tag{13.3-5}$$

$$\sigma_r = \frac{\sqrt{4-\pi}}{2}\sqrt{\overline{r^2}} \tag{13.3-6}$$

$$\frac{\sigma_r}{m} = \sqrt{\frac{4-\pi}{\pi}} \tag{13.3-7}$$

Substitution of Eqs. (13.3-5) to (13.3-7) into Eq. (13.3-4) yields

$$P\left(\left(1 - \frac{0.8625}{\sqrt{N}}\right) m \leq \bar{r}_j \leq \left(1 + \frac{0.8625}{\sqrt{N}}\right) m\right) = 90\% \tag{13.3-8}$$

Then the 90 percent confidence interval CI expressed in decibels is

$$90\%\ CI = 20 \log\left(1 + \frac{0.8625}{\sqrt{N}}\right)$$

$$N = \begin{cases} 50 & 90\%\ CI = 1\ \text{dB} \\ 36 & 90\%\ CI = 1.17\ \text{dB} \end{cases} \tag{13.3-9}$$

In an interval of 40 wavelengths using between 36 and 50 samples is adequate for obtaining the local means. For frequencies lower than 850 MHz, we may have to use an interval of 20λ to obtain the local means because the terrain contour may change at distances greater than 20λ when the wavelength increases.

13.3.2 Estimating unbiased average noise levels[6]

Usually the sampled noise in a mobile environment contains high-level impulses that are generated by the ignition noise of the gasoline

* Time average in a mobile radio environment is an ergodic process in statistics. Therefore the values from a time average with a proper interval and an ensemble average are the same.

engine. Although the level of these impulses is high, the pulse width of each impulse generally is very narrow (see Fig. 13.6). As a result, the energy contained in each impulse is very small and should not have any noticeable effect on changing the average power in a 0.5 s interval.

However, in a normal situation averaging a sampled noise is done by adding up the power values of all samples, including the impulse samples, and dividing the sum by the number of samples. This is called the *conventionally averaged noise power* and is denoted n_c. In this case, these impulse samples dominate the average noise calculation and result in a mean value which is not representative of the actual noise power. Use of this value affects the design requirements of the system (signaling and voice). Hence, before the following new technique was introduced, it was not known why there was no correlation between BER and signal-to-noise ratio measured in certain geographic areas. In a new statistical method the average noise is estimated by excluding the noise impulses while retaining other forms of interference. This technique is compatible with real-time processing constraints.

Description of the method. A counter in the mobile unit counts the instantaneous noise measurements which fall below a preset threshold level X_t and sends a message containing the number of counts n to the database for recording. From the database data, we can calculate the percentage of noise samples x_i below the present level X_t

$$P(x_i \leq X_t) = \frac{n}{N} \tag{13.3-10}$$

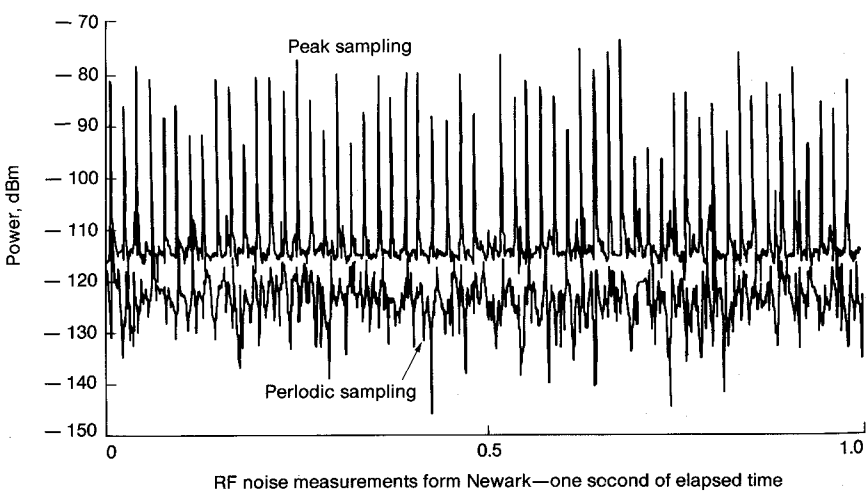


Figure 13.6 Environmental noise traces.

where in our case N is the total number of samples. Once we know the percentage of noise samples below level X_t, we can obtain the average "noise" X_0 exclusive of the noise spikes from the Rayleigh model. Furthermore the level X_t can be appropriately selected for both noise and signal measurements, because both band-limited noise and mobile radio fading follow the same Rayleigh statistics.

Estimating the average noise X_0. For a Rayleigh distribution (band-limited noise), the average noise power exclusive of the noise spikes X_0 can be obtained from

$$X_0 = 10 \log \left\{ -\frac{1}{\ln [1 - P(x_i \leq X_t)]} \right\} + X_t \quad \text{dBm} \qquad (13.3\text{-}11)$$

This technique can be illustrated graphically using Rayleigh paper. Since $P(x_i \leq X_t)$ is known for a given X_t, we can find a point P_t on the paper as illustrated in Fig. 13.7. Through that point, we draw a line

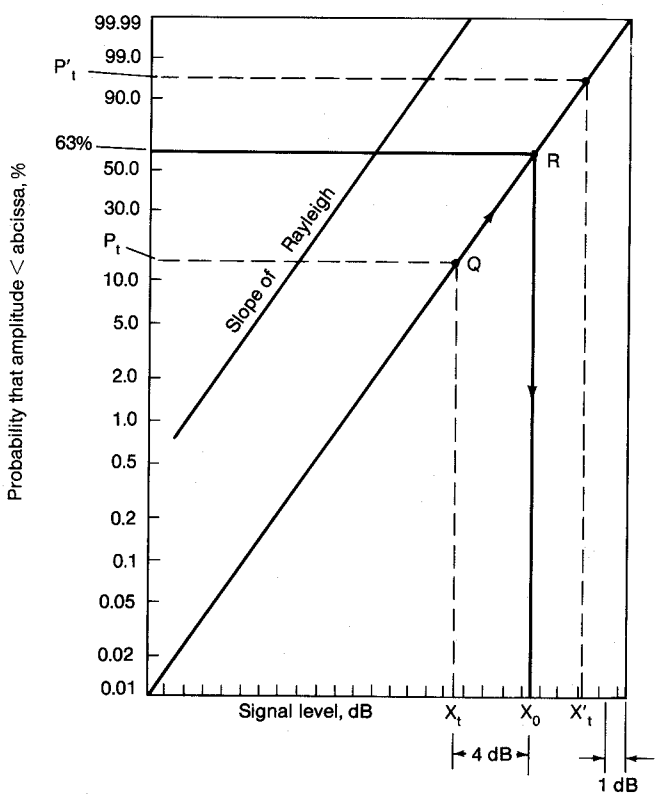


Figure 13.7 Technique of estimating average noise.

parallel to the slope of the Rayleigh curve and meet the line of P = 63 percent. This crossing point corresponds to the X_0 level (unbiased average power in decibels over 1 mW, dBm).

Example 13.7 If a total number of samples is 256 and 38 samples are below a level −119 dBm, then the percentage is

$$\frac{38}{256} = 15\%$$

Draw a line at 15 percent and meet at Q on the Rayleigh curve (see Fig. 13.7). Assume that the X_t is −119 dBm. The X_0 is the average power because 63 percent of the sample is below that level. Then

$$X_0 = X_t + 4 \text{ dB} = -119 + 4 = -115 \text{ dBm}$$

Example 13.8 The total number of samples is 256. Three noise spikes are 20 dB above the normal average. Find the errors, using the following two methods. Compare the results.

Use geometric average method. Let the power value of each sample (of 253 samples) after normalization be 1, i.e., the average is 1. Then the measured average of 256 samples, including three spikes, is

$$\text{Measured average} = \frac{\sum_{1}^{253} x_i + 100 \sum_{1}^{3} x_i}{256} = 2.16 \qquad (\text{assume } x_i = 1)$$

$$= 3.3 \text{ dB} \qquad \text{above the true average}$$

Statistical average method

$$63\% \text{ of samples} = 256 \times 0.63 = 161 \text{ samples}$$

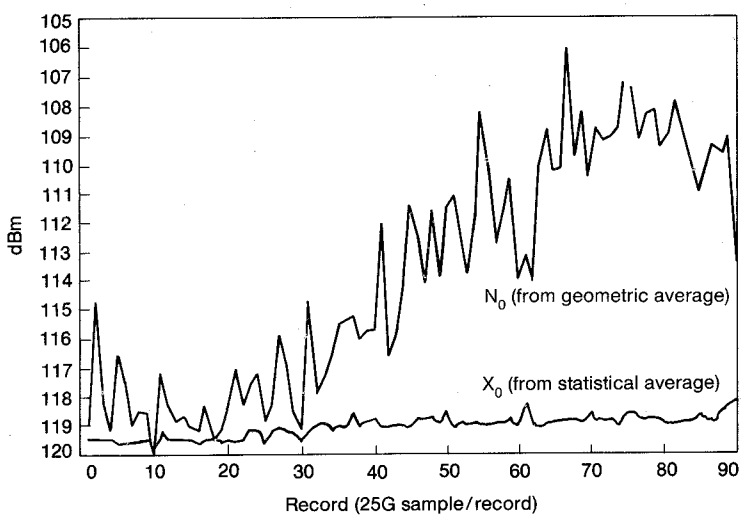


Figure 13.8 An illustration of comparison of n_c with X_e.

This means that 161 samples should be under the average power level. Now three noise spikes added to the 161 samples increases the number of samples to 164.

$$\frac{164}{256} = 64\%$$

The power levels at 63 and 64 percent show almost no change (see Fig. 13.7). Typical data averaging using the geometric and statistical average methods is illustrated in Fig. 13.8. The corrected value is approximately −118 to −119 dBm. The geometric average method biases the average value and causes an unacceptable error as shown in the figure.

13.3.3 Signal-strength conversion

Confusion arises because the field strength (in decibels above 1 μV, dBμ) is measured in free space, and the power level in decibels above 1 mW (dBm) is measured at the terminal impedance of a given receiving antenna. Furthermore, the dimensions of the two units are different. The signal field strength measured on a linear scale is in microvolts per meter (μV/m), and the power level measured on a linear scale is in milliwatts (or watts).

Further confusion arises because of the notation "dBμ." Sometimes dBμ means the number of decibels above 1 μV measured at a given voltage. Sometimes, it represents the number of decibels referred to microvolts per meter when field strength is being measured.

The conversion from decibels (microvolts per meter, dBμ) to decibels above 1 mW (dBm) at 850 MHz is shown in Eq. (5.1-13) using the relationship between induced voltage and effective antenna length.[7] The conversion at a frequency other than 850 MHz can be obtained as follows.[8]

$$P_{\max}(\text{in dBm at } f_1 \text{ MHz}) = P_{\max}(\text{in dBm at 850 MHz}) + 20 \log \left(\frac{850}{f_1}\right) \quad (13.3\text{-}12)$$

where f_1 is in megahertz. The details of this conversion are given in Sec. 5.1.3.

13.3.4 Receiver sensitivity

The sensitivity of a radio receiver is a measure of its ability to receive weak signals. The sensitivity can be expressed in microvolts or in decibels above 1 μV.

$$Y \quad \text{dB}\mu\text{V} = 20 \log (x \quad \mu\text{V}) \quad (13.3\text{-}13)$$

Also, the sensitivity can be expressed in milliwatts or dBm.

$$y \quad \text{dBm} = 10 \log (x \quad \text{mW}) \tag{13.3-14}$$

The conversion from microvolts to decibels above 1 mW, assuming a 50-Ω terminal, has been shown in Eq. (5.1-15) as

$$\begin{aligned} 0 \text{ dB}\mu\text{V} &= 10 \log \frac{(1 \times 10^{-6})^2}{50} \\ &= -137 \text{ dBW} = -107 \text{ dBm} \end{aligned} \tag{13.3-15}$$

Example 13.9 A receiver has a sensitivity of 0.7 μV. What is the equivalent level in decibels above 1 mW?

$$20 \log 0.7 = -3$$

Then 0.7 μV equals -107 dBm $-3 = -110$ dBm.

13.3.5 Level-crossing counter[9]

A signal fading level crossing counter will face a false-count problem as a result of the granular noise as shown in Fig. 13.9. The positive slope crossing count should be 3, but the false counts may be 12. A proposed level-crossing counter can eliminate the false counts. First, by sampling the fading signal at an interval of T seconds, we can choose the interval T such that $1/T$ is small in comparison to the fading rate. The duration of stay τ_i is measured for every sample time which is above level L and the time span until the signal drops below level L. We can also use a device for measuring the percentage of time that $y(t)$ is above L. This device may be called a level crossing counter.

Let us define

$$p = P_r[y(t) \geq L] \tag{13.3-16}$$

and

$$q = 1 - p = P[y(t) < L] \tag{13.3-17}$$

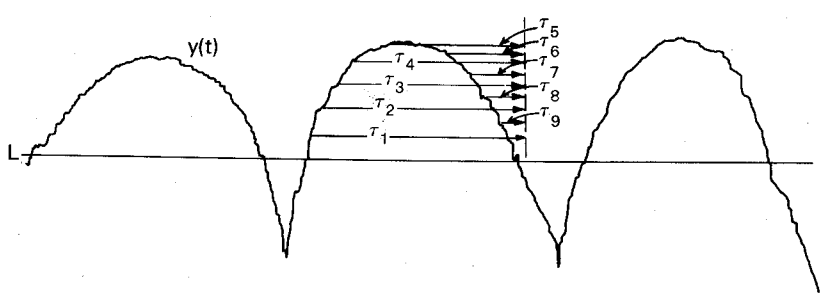


Figure 13.9 An algorithm for a level-crossing counter.

We count the number of times M that τ_i is above L and then sum the duration of total stays $T_s (T_s = \Sigma_{i=1}^{M} \tau_i)$ above L. The average duration of upward fading is

$$\tau_p = \frac{2T_s}{M} \tag{13.3-18}$$

The average duration of fades τ_q where $y(t)$ is below L is

$$\tau_q = \tau_p \frac{1 - p}{p} \tag{13.3-19}$$

and the level crossing rate n at level L is

$$n = \frac{1}{\tau_p + \tau_q} = \frac{p}{\tau_p} \tag{13.3-20}$$

The advantage of this method is that we stop our time count whenever $y(t)$ crosses L, thus avoiding false counts due to noise when $y(t)$ is close to level L. Obviously noise can give an incorrect measure of the "duration of stay" for a single interval; however, noise shortens many intervals while it lengthens others, and thus, when averaged over many cycles, the "duration of stay" is an accurate number.

13.4 Spectrum Efficiency Evaluation[10]

13.4.1 Spectrum efficiency for cellular systems

Because the frequency spectrum is a limited resource, we should utilize it very effectively. In order to approach this goal, spectrum efficiency should be clearly defined from either a total system point of view or a fixed point-to-point link perspective. For most radio systems, spectrum efficiency is the same as channel efficiency, the maximum number of channels that can be provided in a given frequency band. This is true for a point-to-point system that does not reuse frequency channels such as a cellular mobile radio. An appropriate definition of spectrum efficiency for cellular mobile radio is the number of channels per cell. Therefore, in cellular mobile radio systems:

$$\text{Spectrum efficiency} \neq \text{channel efficiency}$$

The system capacity is directly related to spectrum efficiency but not to channel efficiency.

13.4.2 Advantages and impact of FM

In 1936 E. H. Armstrong published a paper entitled "*A Method of Reducing Disturbance in Radio Signaling by a System of Frequency Modulation.*"[11] This paper explored the tradeoffs between noise and bandwidth in FM radio. Since then, engineers have understood the concept of reducing noise by increasing bandwidth in system design.

The parameters for system comparison are *voice quality, transmitted power,* and *cell size.* Satisfactory voice quality is generally accepted as governed by the carrier-to-noise ratio $C/N_{FM} = 18$ dB.* This is the level at which 75 percent of the users state that voice quality is either good or excellent in 90 percent of the service area on a 30-kHz FM channel in a multipath fading environment.[12] For point-to-point radio links, the wider the channel bandwidth, the lower the required level of transmitted power.

To maintain voice quality when channel bandwidth is reduced, it is necessary to increase the signal-to-noise ratio in order to improve the reception. Transmission power is then also increased.

Every time power is increased, interference problems are created. For point-to-point radio links, these problems are manageable because no frequency reuse is involved. This is not the case, however, for a frequency-reuse system such as a cellular radio. A cellular radio telephone system includes many mobile-unit customers and, depending on demand at any given time in a given system, identical channels will be operating simultaneously in different geographic locations. As the number of cells increases in a given area, interference may appear in one of several forms: cochannel, adjacent-channel, or multichannel at colocations; thus the probability of its occurrence increases. Interference may also result from received power-level differences. In a frequency-reuse system, however, cochannel separation is more critical to the system than adjacent-channel interference because adjacent channel interference may be eliminated by the use of sharp filters.

13.4.3 Number of frequency-reuse cells *K*

The formula for determining the number of frequency-reuse cells in a standard cellular configuration is derived by combining Eqs. (2.4-3) and (2.4-5) with $\gamma = 4$ based on the 40 dB/dec path-loss rule.[13]

$$\frac{C}{I} = \frac{1}{6}\left(\frac{D}{R}\right)^4 = \frac{(3K)^2}{6} = \frac{3K^2}{2} \tag{13.4-1}$$

or

$$K = \sqrt{\frac{2}{3}\frac{C}{I}} \tag{13.4-2}$$

* If we use values other than 18 dB, the analysis used in this section remains the same.

The number of frequency-reuse cells is a function of the required carrier-to-interference ratio.

A higher required carrier-to-interference ratio at the boundary of a cell results in the need for more frequency-reuse cells. The pattern of reuse cells can then be determined.

13.4-4 Number of channels per cell *m*

The next factor to be determined is the number of channels per cell, which is a function of the total number of channels available (amount of available spectrum divided by channel bandwidth) and the required carrier-to-interference ratio. The formula for this factor is[10]

$$m = \frac{B_t}{B_c K} = \frac{B_t}{B_c\sqrt{(2/3)(C/I)}} \tag{13.4-3}$$

$M = mK$ total number of channels

where m = number of channels per cell, also called radio capacity by Lee[10]
K = number of frequency reuse cells (see Eq. 13.4-2)
B_t = total bandwidth (transmitted or received)
B_c = channel bandwidth

13.4.5 Rayleigh fading environment

The Rayleigh fading environment is the mobile radio environment caused by multipath fading, which is the cellular system environment. Therefore, it is more realistic to determine the spectrum efficiency of a cellular mobile radio in a Rayleigh fading environment.

In a multipath fading environment, a simple FM system which may not have either preemphasis-deemphasis or diversity schemes would receive its baseband signal-to-noise ratio (S/N), which is converted from the carrier-to-interference ratio (C/I) but S/N is 3 dB lower than C/I.[14]

System advantages. The FCC has released specifications which result in advantages vis à vis the signal-to-noise ratio for transceivers in the existing FM cellular system. The first advantage is preemphasis-deemphasis, which equalizes the baseband signal-to-noise ratio over the entire voice band (f_1 to f_2). We make the assumption of Gaussian noise because the interference obtained from all six cochannel interferers behaves in a noiselike manner. The improvement factor ρ_{FM}, that is, the improvement of FM with preemphasis or deemphasis over FM without them, can be calculated as follows.[15,16]

$$\rho_{FM} = \frac{(f_2/f_1)^2}{3} = \frac{(3000\text{ Hz}/300\text{ Hz})^2}{3} = 33.3\ (=)\ 15.2\text{ dB} \tag{13.4-4}$$

Another advantage is the two-branch diversity combining receiver, which is very suitable for FM and reduces multipath fading. The advantage of the two-branch diversity receiver is that the baseband signal-to-noise ratio S/N of a two-branch FM receiver shows a 8-dB improvement over the signal-to-noise ratio of a single FM channel.

$$\left(\frac{S}{N}\right)_{2\text{br}\,\text{FM}} = 8 \text{ dB} + \left(\frac{S}{N}\right)_{\text{FM}} \tag{13.4-5}$$

The existence of compandors is assumed for compressing the signal bandwidth and taking advantage of the quieting factor during pauses. However, the voice quality improvement due to the quieting factor cannot be expressed mathematically. It is understood that all the analog modulation systems use compandors.

Present FM system. The subjective required carrier-to-interference ratio for FM is

$$\left(\frac{C}{I}\right)_{\text{FM}} = 18 \text{ dB } (=)\ 63.1 \tag{13.4-6}$$

The baseband signal-to-noise ratio can be obtained from the previous analysis as follows.

$$\begin{aligned}\left(\frac{S}{N}\right)_{2\text{br}\,\text{FM}} &= -3 \text{ dB} + \text{deemphasis gain} + \text{diversity gain} + \left(\frac{C}{I}\right)_{\text{FM}} \\ &= 38.23 \text{ dB } (=)\ 6652.73 \end{aligned} \tag{13.4.7}$$

The signal-to-noise value $S/N = 38$ dB is a reasonable figure for obtaining good quality at the baseband.[17] The notation (=) means a conversion between decibels and a linear ratio.

SSB systems. Single-sideband receivers, best case, have a carrier-to-interference ratio equal to the signal-to-noise ratio at baseband since SSB is a linear modulation.[17] The term "best case" means that the signal fades are completely removed, that is, the environment approaches a gaussian.* There is no advantage in using diversity schemes in a gaussian environment. If the environment is Rayleigh, the carrier-to-interference ratio must always be higher than that in a gaussian environment in order to obtain the same voice quality at the baseband. An explanation is given in Sec. 13.4.7. To obtain a similar voice quality,

* SSB systems at 800 MHz have not been commercially available because of technical difficulties. It is assumed here that an ideal SSB at 800 MHz can be built for mobile radios.

the signal-to-noise ratio of both FM and SSB systems at the baseband should be the same.[18] The formula used to determine the required C/I of SSB is

$$\left(\frac{C}{I}\right)_{\text{SSB}} = \left(\frac{S}{N}\right)_{\text{SSB}} = \left(\frac{S}{N}\right)_{\text{2brFM}} = 38.23 \text{ dB } (=) \ 6652.73$$

This means that the required 38.23-dB carrier-to-interference ratio of 38.23 dB for an SSB system (see Sec. 13.4.7) is equivalent to the required carrier-to-interference ratio of 18 dB for an FM system for equivalent voice quality. The number of channels per cell and the number of channels per square mile for the system can then be calculated.

Number of channels per cell *m*. The preceding analysis gives us the information needed to determine the number of channels per cell. Assuming $B_t = 10$ MHz, the formula is [from Eq. (13.4-3)]

$$m = \frac{B_t}{B_c K} = \frac{10 \text{ MHz}}{B_c\sqrt{(2/3)(C/I)}} \tag{13.4-8}$$

Given a total bandwidth (B_t) of 10 MHz, $C/I = 18$ dB for FM, and $C/I = 38.23$ dB for SSB, it is possible to determine the number of channels per cell m by the substitution of the above values in Eq. (13.4-8) and shown in Table 13.1. As this table shows, FM cellular systems need fewer cells than do SSB systems to provide quality voice service.

There is a dispute as to whether at 800 MHz, an SSB system needs a C/N of 38 dB or less to provide an S/N of 38 dB at the baseband.[19–22] Since there is no commercial 800-MHz SSB system, no subjective test can be used for SSB voice quality. Comparing the performance of an existing FM system to that of a nonexisting SSB system is difficult. Also, it is not proper to use the results from a 150-MHz SSB system without making a thorough subjective test in a Rayleigh fading environment* and applying it to a 800-MHz SSB system.

* Air-to-ground communications media do not exhibit Rayleigh fading behavior. Also the required C/I of SSB at 800 MHz would be different from that at 150 MHz.

TABLE 13.1 Channels per Cell
(Rayleigh Fading Environment)

System	Bandwidth B_c, kHz	Cells per set K	Total number of channels B_t/B_c	Channels per cell m
FM	30.0	7	333	47.57
SSB	7.5	66	1333	20.00
SSB	5.0	66	2000	30.0
SSB	3.0	66	3333	50.05

13.4.6 Determination of cell size

It is possible to determine the size of comparable cells for 30-kHz FM, 3-kHz SSB, 5-kHz SSB, and 7.5-kHz SSB once the number of frequency-reuse cells and the number of channels per cell have been calculated. These values are related to the level of carrier power required at reception to maintain similar voice quality. Since SSB has a relatively narrow bandwidth, the noise level is also lower. The SSB noise level must be adjusted to the FM noise level in order to determine the power required for SSB.

Required power in each SSB system. The SSB required carrier-to-interference ratio must be 38.23 dB. Therefore, the power required by the SSB system after the noise level (including interference) has been adjusted can then be determined. The voice quality of an SSB system having a carrier-to-interference ratio of 38.23 dB is equivalent to an FM system having a carrier-to-interference ratio of 18 dB. This assumes that in-band pilot tones can smooth out the fading signal and causes no distortion in SSB reception. The noise levels of different SSB bandwidths can be shown as follows.

$$\left(\frac{C}{I}\right)_{\text{SSB}} = 38.23 \text{ dB}$$

$$\left(\frac{C}{I}\right)_{\text{SSB3kHz}} = 10.23 \text{ dB} + 18 \text{ dB} + 10 \text{ dB}$$

$$\left(\frac{C}{I}\right)_{\text{SSB5kHz}} = 12.45 \text{ dB} + 18 \text{ dB} + 7.74 \text{ dB}$$

$$\left(\frac{C}{I}\right)_{\text{SSB7.5kHz}} = 14.21 \text{ dB} + 18 \text{ dB} + 6.02 \text{ dB}$$

These figures are illustrated clearly in Fig. 13.10.

SSB cell size determined for the required additional power. For a given transmitted power, the cell size can be determined. Assuming an FM cell radius of 10 mi and applying the 40 dB/dec path-loss rule, the cell sizes for an SSB system may be determined as follows.

For a 3-kHz SSB system

$$10 \log \left(\frac{10^{-4}}{R^{-4}}\right) = -10.23 \text{ dB} \qquad R = 5.55 \text{ mi}$$

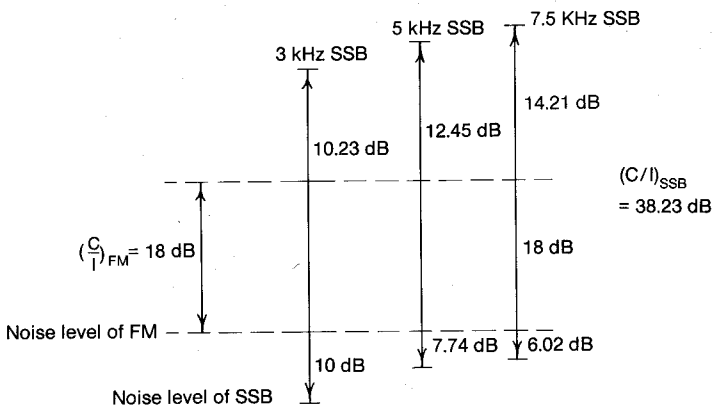


Figure 13.10 The interference-plus-noise levels for 30-kHz FM and different bandwidths of SSB.

For a 5-kHz SSB system

$$10 \log \left(\frac{10^{-4}}{R^{-4}}\right) = -12.45 \text{ dB} \qquad R = 4.88 \text{ mi}$$

For a 7.5-kHz SSB system

$$10 \log \left(\frac{10^{-4}}{R^{-4}}\right) = -14.21 \text{ dB} \qquad R = 4.41 \text{ mi}$$

The higher the carrier-to-interference ratio for a given power, the closer the mobile unit is to the cell. Assuming the same voice quality at the boundary of the cell, a 10-mi radius FM 30-kHz cell is equivalent to a 5.5-mi 3-kHz SSB cell, a 4.88-mi 5-kHz SSB cell, or a 4.41-mi 7.5-kHz SSB cell. Therefore, an FM system permits larger cells and an SSB system requires smaller cells to provide for the same voice quality in the same area.

Comparison of cochannel cell separation and radius of FM and SSB systems in a Rayleigh fading environment. The cochannel cell separations and the radii of both FM and SSB cells in a Rayleigh fading environment are summarized in Table 13.2 and expressed in Fig. 13.11. Therefore,

TABLE 13.2 Comparison of Cochannel Cell Separation and Radius of FM and SSB Systems in a Rayleigh Fading Environment

System	Cells per set K	Radius R, mi	Diameter D, mi	D/R	Bandwidth, kHz
FM	7	10	46	4.6	30
SSB	66.6	5.5	77.77	14.14	3
		4.88	69	14.14	5
		4.41	62.36	14.14	7.5

in a cellular mobile radio environment, a FM system *permits* larger cells with less separation between cochannel cells and a SSB system requires smaller cells with greater separation between cochannel cells.

Channels per square mile. Table 13.3 shows a comparison of the channels per square mile and the spectrum efficiency of each system in the Rayleigh fading environment based on equivalent voice quality and a given transmitted power. This comparison shows that the existing 30-kHz FM cellular system is about as spectrally efficient as the hypothetical 3-kHz SSB system and much more efficient than either of the two SSB (5- or 7.5-kHz) systems offering a voice quality similar to that proposed for commercial service.

13.4.7 Considerations of SSB systems in a Rayleigh fading mobile radio environment

The voice signal requires about a S/N ratio of 40-dB output at the baseband for high quality.[23] Our baseband S/N ratio is calculated to be 38 dB, which is very close to 40 dB.

In a gaussian environment, if the output S/N of a SSB signal at the baseband is 38 dB, then the S/N at the RF is also 38 dB because SSB is linearly modulated.[17] Now, in a Rayleigh fading environment, the received signal can be further degraded as a result of the multipath fading. Therefore, in order to maintain the baseband S/N at 38 dB, we may need to receive a signal much higher than 38 dB if no diversity scheme is implemented. How high the C/I level at RF should be cannot be determined unless the SSB mobile radio at 800 MHz is realized and a subjective test is done. Since a single-branch SSB cannot be used in a mobile radio environment at 800 MHz because of rapid fading,[24] the

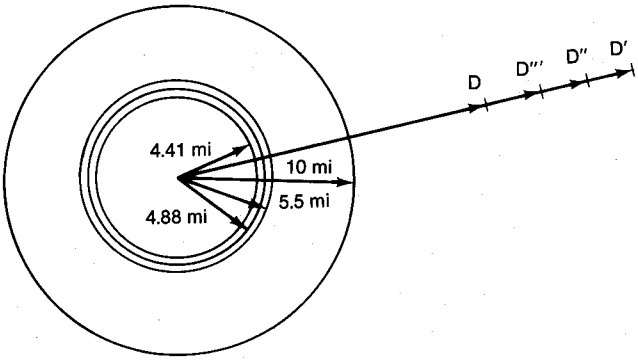


Figure 13.11 Cochannel cell separations and radii of FM and SSB systems in a Rayleigh fading environment.

TABLE 13.3 Comparison of System Efficiency and Spectrum Efficiency in an Area of 10-mi Radius

System	Bandwidth, mHz	System efficiency: Cell radius, mi	System efficiency: No. of cells required	Spectrum efficiency: Channels per cell, m	Spectrum efficiency: No. of channels per square mile
FM	30	10	1.0	47.5	0.15
SSB	7.5	4.4	$\left(\frac{10}{4.4}\right)^2 = 5.17$	20.0	0.06
SSB	5	4.8	$\left(\frac{10}{4.8}\right)^2 = 4.34$	30.0	0.10
SSB	3	5.5	$\left(\frac{10}{5.5}\right)^2 = 3.31$	50.0	0.16

value of applying a two-branch diversity to the SSB at 800 MHz also must be questioned. Our reasoning is as follows.

Let S_0 be a voice signal, r_1 and r_2 the envelopes of the fading, θ_1 and θ_2 the random phases received by two spaced antennas, and ω_0 the carrier angular frequency. Where an equal-gain combined receiver is considered, in an SSB system the combined signal envelope is

$$\begin{aligned} r &= \left|r_1 S_0 e^{j(\omega_0 t + \theta_1)}\right| + \left|r_2 S_0 e^{j(\omega_0 t + \theta_2)}\right| \\ &= r_1 S_0 + r_2 S_0 = (r_1 + r_2) S_0 \end{aligned} \tag{13.4-19}$$

which is the same as the baseband signal representation. The term $(r_1 + r_2)$ does reduce the fading as compared to either individual r_1 or r_2 to a certain degree, but it also acts as a distortion term to S_0. The distortion of voice S_0 on $(r_1 + r_2)$ received by a two-branch diversity receiver for SSB in a ground mobile radio (Rayleigh) environment is still quite high at 800 MHz. This is because the effect of fading is multiplicative and produces intermodulation products with the signal modulation that cannot be eliminated by filtering.[25] In air-to-ground transmission, the direct-wave path dominates, the fading phenomenon (rician) is not severe, and a two-branch diversity does help in improving the voice quality at the reception.

An amplitude companding single sideband (ACSB) with an in-band pilot tone is considered[26] under the assumption that this kind of SSB can in principle completely remove the Rayleigh fading; therefore, no diversity scheme* is needed. Then we do not require an increase in the received power level at the RF, but rather simply retain the same level as the baseband S/N ratio of 38.23 dB.

* The diversity scheme is used to eliminate fading.

Preemphasis and deemphasis are not widely used in SSB systems. The disadvantage of the use for SSB systems in a nonfading environment is discussed by Schwartz[27] and Gregg.[28]

In a mobile radio environment, in order to transmit a predistorted SSB signal using preemphasis to suppress the noise level, the signal cannot be completely restored because the effect of fading is multiplicative and it produces intermodulation products in the voice band that cannot be eliminated by filtering as mentioned before. Therefore, the use of preemphasis and deemphasis in an 800-MHz SSB mobile radio system is questionable.

Reference 18 shows that an RF signal with required $C/I = 38$ dB received by a SSB system through a gaussian environment results in a baseband signal where $S/N = 38$ dB also. In a Rayleigh fading environment, the C/I must definitely be higher than 38 dB for S/N to be 38 dB. How high is not known since 800-MHz SSB equipment has not been manufactured. We may use the information obtained from a FM system for maintaining the same voice quality in different environments.

$$\begin{array}{ll} \text{Required for FM in a gaussian environment} & \\ \text{Required for FM in a Rayleigh fading environment} & \end{array} \quad \frac{C}{I} \geq \begin{cases} 10 \text{ dB} \\ 18 \text{ dB} \end{cases}$$

The difference in C/I for the FM system between the two kinds of environment is 8 dB. We may use the 8-dB difference and add it to $C/I = 38$ dB for an SSB system. Then

$$\text{For SSB in a Rayleigh fading environment} \quad \frac{C}{I} = 46 \text{ dB}$$

Suppose that a diversity scheme is used in SSB as Shivley[21] suggested. Then C/I for SSB in a less-fading (or no-fading) environment is

For SSB in a less-fading (or no-fading) environment

$$\frac{C}{I} - 8 \text{ dB} = 38 \text{ dB}$$

The same result is obtained if we assume that the fading is completely removed, $C/I = S/I$, and that the environment becomes gaussian. In a gaussian environment, we cannot reduce C/I below 38 dB, because S/N is 38 dB. Also, in a gaussian environment, the diversity scheme does not apply and adds no value.

Of course, the analysis shown here remains to be proved if and when an 800-MHz SSB mobile unit is developed in the future. After all, the methodology of solving this problem remains unchanged.

13.4.8 Narrowbanding in FM

Relationship between *C/I* at IF and *S/N* at baseband. In Ref. 12 we defined acceptable voice quality as existing when 75 percent of customers say that the voice quality is good or excellent in a 90 percent coverage area. When these numbers change, voice quality changes accordingly. The changes reflect the cost of deploying a cellular system. As the percentages specified above increase, the cost of designing the system to meet these requirements also increases. For now, we will use the numbers specified above as our criteria.

We let the customer listen to the voice quality of a 30-kHz FM two-branch diversity receiver with preemphasis-deemphasis and companding features at the cell site while the mobile transmitter is traveling at speeds ranging from 0 to 60 mi/h in a Rayleigh environment. Judging by the preceding subjective criterion, the C/I level at the input of the receiver is 18 dB.

The baseband signal-to-noise ratio (S/N) has been calculated in Eq. (13.4-7).

$$S/N = \underbrace{18 - 3}_{\text{Rayleigh environment}} + \underbrace{15.23}_{\text{deemphasis advantage}} + \overset{\text{diversity advantage}}{8} = 38.24 \text{ dB}$$

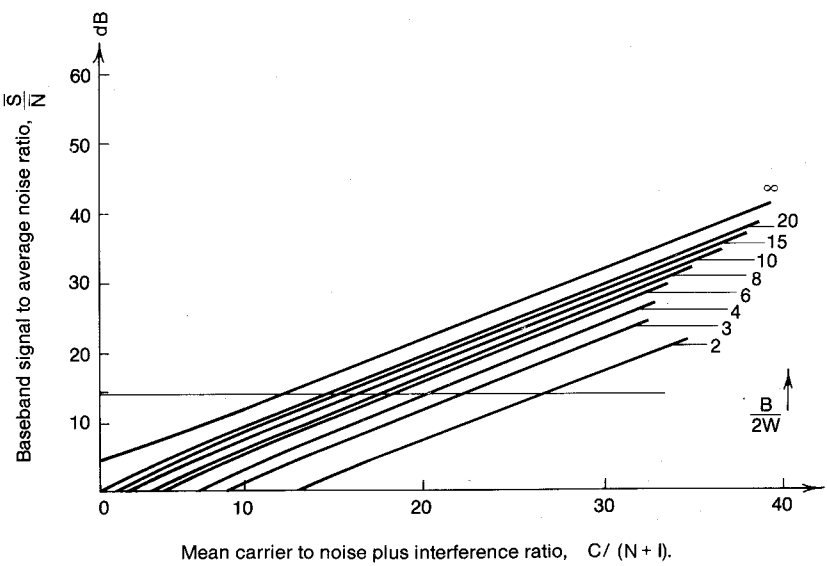


Figure 13.12 Baseband signal-to-noise ratio versus average carrier-to-interference ratio (Rayleigh fading). (*From Ref. 29.*)

For a 15-kHz FM channel, the bandwidth is half as broad and affects the S/N; the other features, such as diversity and preemphasis-deemphasis, remain the same. We can find from Fig. 13.12 (Ref. 29) that in order to maintain the same voice quality of $(C/I)_{30\text{ kHz}} = 18$ dB, then

$$\left(\frac{C}{I}\right)_{15\text{ kHz}} = 24 \text{ dB}$$

We may follow the same steps used in Sec. 13.4.6 along with the diagram shown in Fig. 2.5.

$$\frac{C}{N+I} = \frac{C}{(kTB + NF) + \sum_{i=1}^{6} I_i}$$

$$= 24 \text{ dB (15-kHz FM) or } 18 \text{ dB (30-kHz FM)} \qquad (13.4\text{-}10)$$

Using the techniques from previous sections, we can perform further calculations.

The interference-plus-noise levels. From the preceding calculations Eq. (13.4-10), we can obtain the interference-plus-noise levels $(N + I)$ at the boundary of a 10-mi-radius cell as follows.

For 30-kHz FM system $\quad N + I = -117$ dBm.

For 15-kHz FM system $\quad N + I = -123$ dBm.

Since the received signal strength is -99 dBm for both 30- and 15-kHz, the $C/(N + I)$ ratio must be 18 dB and 24 dB for 30- and 15-kHz FM, respectively, in order to maintain the same voice quality. This relationship is shown in Fig. 13.13.

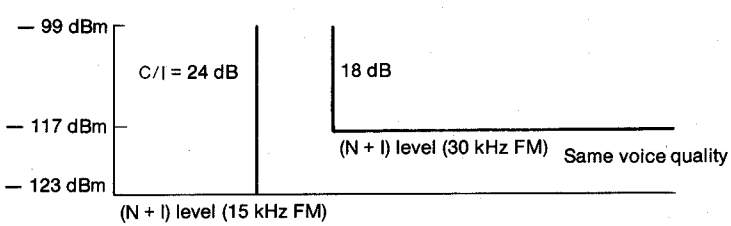


Figure 13.13 The interference-plus-noise levels for 30- and 15-kHz FM channels.

The number of cells in a frequency-reuse pattern. Let K be the number of cells in a frequency-reuse pattern; then in 30-kHz systems,

$$q = \frac{D}{R} \triangleq \sqrt{3K} = 4.6$$

$$K = 7 \qquad \text{(for both noise-neglected and noise-included cases)}$$

For a seven-cell reuse pattern operating in 15-kHz systems,

$$K = \frac{q^2}{3} \begin{cases} = 13 & (q = 6.23) \quad \text{(noise-neglected case)} \\ = 16 & (q = 7.08) \quad \text{(noise-included case)} \end{cases}$$

A 13- to 16-cell reuse pattern is needed.

Intersystem comparison of spectrum efficiency. If both the noise-included and the noise-neglected cases are considered, then

$$K = \begin{cases} 7 & \text{(30-kHz FM system, both noise-included and noise-neglected cases)} \\ 16 & \text{(15-kHz FM system, noise-included case)} \\ 13 & \text{(15-kHz FM system, noise-neglected case)} \end{cases}$$

The spectrum efficiency for both systems can be shown to be about the same as follows:

$$\frac{333}{7} = 47.5 \text{ channels per cell} \qquad \text{(30-kHz FM)}$$

$$\frac{666}{16} = 41.63 \text{ channels per cell} \qquad \text{(15-kHz FM, noise-included case)}$$

$$\frac{666}{13} = 51.2 \text{ channels per cell} \qquad \text{(15 kHz FM, noise-neglected)}$$

Increasing spectrum efficiency by degrading voice quality. If we accept $C/I = 18$ dB for 15-kHz FM, this means that voice quality is degraded by 6 dB. Then the cochannel interference reduction factor q becomes approximately 4.6, corresponding to a frequency-reuse pattern of $K = 7$. Since the frequency channel is doubled, in a 10-MHz system (666 channels)

$$\frac{666}{7} = 2 \times \text{(number of 30-kHz FM channels per cell)}$$

Therefore, voice qualtiy is sacrificed to gain spectrum efficiency.

Another approach is to increase the transmitted power of the 15-kHz FM system by 6 dB. Then the C/I remains the same because the interference is also increased by 6 dB. Therefore, no advantage to spectrum efficiency can be obtained by either increasing or reducing power.

13.5 Portable Units

All of today's system design tools are designed to improve the performance of cellular mobile units. Therefore, portable units become a secondary byproduct of the cellular system. Since very few systems in the United States are designed mainly for portable units, it will be necessary to study each existing system individually to determine whether portable units are suitable for it. Portable unit usage can be adopted easily by some existing systems but not by others for many reasons. If we find that portable units become very popular in an existing mobile cellular system, then we have to reconsider some parameters used for the mobile cellular system to adapt to portable unit usage. There are two parts to the calculation.

13.5.1 Loss due to building penetration

The loss (attenuation) when propagating an 800-MHz wave due to building penetration is very high.[31–34] Also, the structure-related attenuation varies, depending on the geographic area. In Tokyo, the path-loss difference inside and outside buildings at first-floor level is about 26 dB. But in Chicago, the path-loss difference under the same conditions is 15 dB as shown in Fig. 13.14.

These variations are attributable to differences in building construction. In Tokyo, many supporting metal frames (mesh configuration) are used in the building structures to allow the buildings to withstand earthquakes. In Chicago, fewer supporting metal frames are needed as there is less risk of earthquakes. Therefore, the *building penetration* is far less severe in Chicago than in Tokyo.

The same would apply to buildings in California. For instance, the loss due to building penetration in Los Angeles would be higher than that in Chicago but lower than that in Tokyo, because Los Angeles has only a few high-rise buildings. In Los Angeles the penetration at the first-floor level would be around -20 dB, as compared with the received signals at mobile units outside at street level.

Signal attenuation at the building basement level in Chicago is 30 dB below the signal received from the street-level mobile unit. This indicates that the signals do not penetrate basement structures easily. There are two ways to solve this problem.

1. Select the cell site closest to those downtown buildings used for conventions. Then calculate the received power for the portable units.

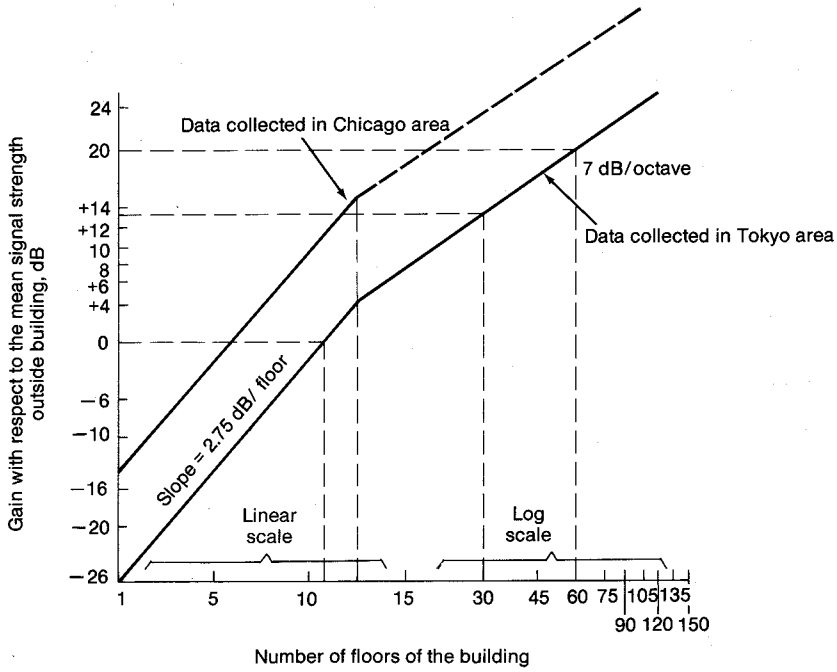


Figure 13.14 The building penetration losses in the Tokyo area and in the Chicago area.

Let the cell coverage be at a radial distance R, with the receiving level of $C/I = 18$ dB at the cell boundary as designed. The same site will be used for portable units. The receiving level of portable units is $C/I = 10$ dB because of the slow-motion or no-motion environment. At the basement, the reception level of a portable unit is $-30 + (18 - 10) = -22$ dB, i.e., 22 dB weaker than that received at the cell boundary R. Applying Eq. (4.2-10a), we obtain

$$40 \log \frac{R_1}{R} = -22 \text{ dB} \qquad \text{or} \qquad R_1 = R(10^{-(22/40)}) = 0.28R \qquad (13.5\text{-}1)$$

From this calculation, the region in which the portable unit can be used in the basement is confined to an area of $0.28R$.

2. Install the repeaters (or enhancers) or leaky feeders to enhance the signal strength inside the building.

13.5.2 Building height effect

Usually, the signal reception level increases as the height of the building where the antenna is located increases. Comparing the two measurements from Chicago and Tokyo, we find that the slope (gain in-

crease) of 2.7 dB per floor (i.e., 2.7 dB gain per each floor-level increment) can be a good value on which to base calculations regarding suitability of portable unit usage. However, this value is valid only up to the thirteenth floor. After that, a logarithmic 7 dB/oct scale is used. Now we start at the first floor.

First-floor region. Following the same procedure as when calculating the signal strength in the basement region, we find the signal strength requirements at the first floor inside the building for the portable unit to be

$$-15 + (18 - 10) = -7 \text{ dB} \qquad \text{(Chicago)}$$

$$-26 + (18 - 10) = -18 \text{ dB} \qquad \text{(Tokyo)}$$

$$-20 + (18 - 10) = -12 \text{ dB} \qquad \text{(intermediate value)}$$

Applying these findings to Eq. (13.5-1), we obtain

$$40 \log \frac{R_1}{R} = \begin{cases} -7 \text{ dB} \\ -18 \text{ dB} \\ -12 \text{ dB} \end{cases} \qquad R_1 = \begin{cases} 0.668R & \text{(Chicago)} \\ 0.35R & \text{(Tokyo)} \\ 0.5R & \text{(intermediate value)} \end{cases}$$

***N*th-floor region.** The area serviced increases as a function of height of 2.7 dB per floor below the thirteenth floor, where we see a gain increase of 2.7 dB per floor, and above the thirteenth floor, 7 dB/oct is used. We can calculate the service area in Chicago as follows.

$$40 \log \frac{R_1}{R} = -7 + N(2.7) \qquad \text{(in Chicago)}$$

where N is the number of floors. The same procedures apply to Los Angeles and Tokyo. In Table 13.4, we see that the service region increases at higher floor levels. However, after the thirteenth floor, the increase in gain is very small. The difference between the two floors from two cities becomes smaller for heights beyond the thirtieth floor (see Fig. 13.14).

TABLE 13.4 Building Penetration Loss

Condition	Building penetration loss	Shadow loss*
Building penetration	+27 dB (Tokyo) +15 dB (Chicago)	27 dB (Chicago) (regardless of floor height)
Window area	+6 dB	
1st–13th floors	2.75 dB/floor (Tokyo) 2.67 dB/floor (Chicago)	
13th–30th floors	7 dB/oct (Tokyo and Chicago)	

* *Shadow loss* is defined as the loss due to a building standing in the radio-wave path.

13.5.3 Interference caused by portable units

Interference to the other portable units. The portable unit has a transmitting power of 600 mW (28 dBm). The interference at the cell site from two different portable units can be determined as follows. We now can consider interference at higher floor levels (see Fig. 13.14). We find that reception at the sixth floor is the same as that at street level in Chicago and that reception at the eleventh floor is the same as that at street level in Tokyo. Reception at the thirtieth floor in Tokyo is 13 dB higher than that at street level.

A portable unit transmitter can transmit a signal to a cell site following line-of-sight propagation. A signal from a portable unit on the thirtieth floor that is received by the cell site could interfere with the reception of a signal from a portable unit on the eleventh floor. The interference level for the portable unit on the eleventh floor is 13 dB. The interference range becomes

$$20 \log \frac{R_1}{R} = -13 \text{ dB} \qquad \text{then} \qquad R_1 = 0.22R$$

Since $R_1 = 0.22R$, the portable unit used on the thirtieth floor at the cell boundary R will not interfere with cell-site reception from a portable unit on the eleventh floor at $0.22R$ away. If the power of both units can be controlled at the cell site, the near-end to far-end ratio interference of 13 dB can be reduced. We must be aware of this interference and find ways to eliminate it once we know its cause. A method for this was described in Chap. 10. The selection of a method for eliminating interference is based on environmental factors such as building height and density, which vary from area to area.

Interference to the mobile units. Now assume that at the cell boundary (the cell radius is R) the mobile unit received a signal at -100 dBm and the reception level of a portable unit at the thirtieth floor is -87 dBm ($-100 + 13$ dB). If the cell site has a 10-W transmitter and a 6-dB-gain antenna, then the transmitting site has an effective radiated power (ERP) of 46 dBm. The path loss on a thirtieth floor of the cell boundary becomes 133 dB (46 + 87). Now for calculating a reverse path, the portable unit has a 600-mW (28-dBm) transmitter; and the mobile unit has a 3-W (35-dBm) transmitter. Then the signal received at the 100-ft cell-site antenna is $(28 - 133) = -105$ dBm for the portable unit and $(35 - 133 - 13) = -111$ dBm for the mobile unit. The difference in received levels is 6 dB. This near-end to far-end ratio interference can be eliminated by the frequency assignment and power control of the portable units at the cell site.

13.5.4 Difference between mobile cellular and portable cellular systems

It is very interesting to point out the differences in characteristics, coverage charts, and system design aspects for mobile and portable cellular systems.

Different Characteristics.

Mobile units	Portable units
Two-dimensional system	Three-dimensional system
Needs handoffs	No handoffs
Severe signal fading due to vehicle movement	No fading or mild fading if walking
Gain changes with ground elevation	Gain changes with building height
Loss due to multipath reflection	Loss due to building penetration
Required $C/I \geq 18$ dB	Required $C/I \geq 10$ dB
Power consumption is not an issue	Power consumption is a key issue
Attractive because of various features	Attractive because of their small size and light weight

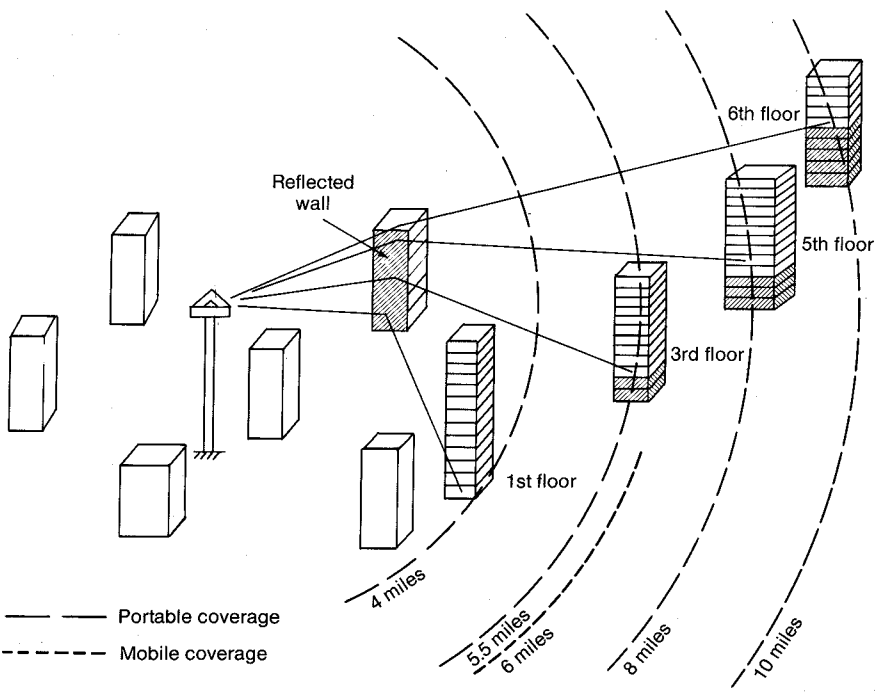


Figure 13.15 Portable and mobile coverage.

Different coverage charts Using the Philadelphia path-loss curve shown in Fig. 4.3 with the standard condition parameters listed in Sec. 4.2.1, we can illustrate the differences in the coverage charts of mobile units and portable units in urban areas, as shown in Fig. 13.15. The receiver sensitivity of both units is assumed to be −117 dBm. Also assume that the required C/I of a portable unit is 10 dB, and the required C/I of a mobile unit is 18 dB. The mobile-unit coverage is about 6 mi, while the portable coverages differ with building height, as shown in Fig. 13.15. The higher the building, the greater the coverage range.

Different design concepts. In mobile cellular systems, we try to cover the area with an adequate signal from a cell site; then transmitted power, antenna height and gain, and location are the parameters involved. Reduction of both multipath fading and cochannel interference is described in Chap. 6.

In a portable cellular system, the coverage range increases with the height of the building. Therefore no fixed coverage range can be given for the portable units. Also the cochannel interference reduction ratio q [$q = D/R$ see Eq. (2.3-1)] for the portable cellular systems has no meaning since the cell radius R changes with building height. If we try to apply the design techniques for mobile cellular systems to portable cellular systems, the results are not very good. One way to look at this problem is that each building structure offers less interference

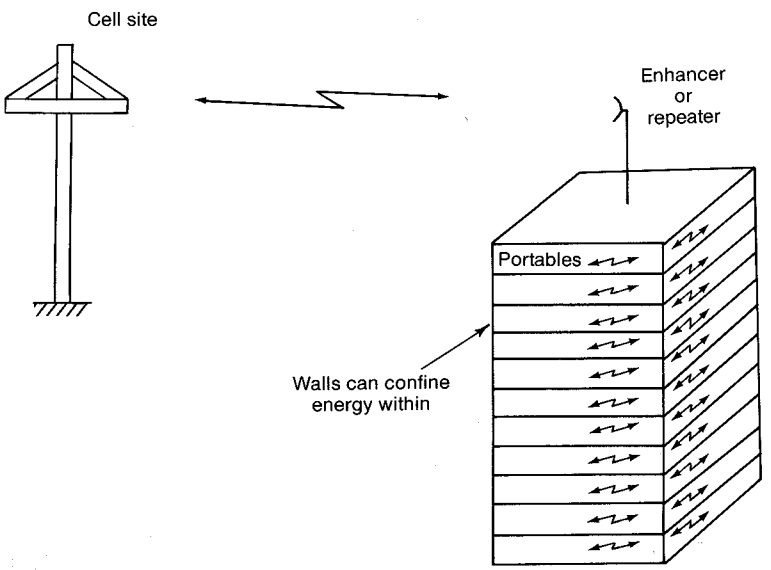


Figure 13.16 Proper arrangement for portable cellular system.

inside the building. The lower the building, the greater the protection from interference. Therefore we should not raise the transmitted power and try to penetrate the building; rather we should take advantage of this natural shielding environment. Thus we should link to each repeater (enhancer) mounted at the top of each building. Since the buildings are tall, reception will be good because of the building's height, and only a small amount of transmitted power is needed at the cell site (see Fig. 13.15). If leaky cables or cables with antennas (shown in Fig. 13.16) from the repeater are connected to each floor, the signal in the whole building will be covered. This is the proper arrangement, but it is also a different concept of designing a portable cellular system.

13.6 Evaluation of Data Modem

13.6.1 Requirement

The data modem used in the current analog system must meet the following requirements.

1. Data transmissions have to use 30 kHz voice channels.
2. The SAT tone must be maintained at around 6000 Hz in voice channels. Then the transmission rate has to be either lower or higher, but it must be clear from the 6000-Hz SAT.
3. The transmission rate cannot be lower than a rate[35] which lies in the dominant random-FM region of $f_{\text{rfm}} < 2(V/\lambda)$. This specification is based on the vehicle speed V and the wavelength of the operating frequency. For instance, if $V = 104$ km/h (65 mi/h), and $\lambda = 1$ ft (at 850 MHz), then $f_{\text{rfm}} = 2(V/\lambda) = 190$ Hz. This means that the data modem transmission rate cannot be below 190 Hz because of the unique random-FM characteristics in the mobile radio environment. When the mobile unit stops, the random FM disappears.
4. The same severe fading requires us to use redundancy, coding, automatic-repeat request (ARQ) scheme, diversity, and so on when transmitting data. The current mobile cellular signaling transmission rate is 10 kbps. Using the BCH code with five repeats, the data throughput is 1200 bits. However, the power spectrum density of a 10-kbps data stream with a Manchester coding (a biphase waveform) is spread out over the 6000-Hz region.[36] If the SAT cannot be detected because of the data modem, then the data modem cannot be used.
5. Mobile unit (vehicular) speed is a significant factor.
   a. Suppose that a car can be driven slowly while approaching a stop but that it never stops. In this case, the average duration

of fades is very long. Most of the time, if the word is in the fade, the whole word is undetected.

b. Suppose that a car can travel at a rate of 104 km/h (65 mi/h). The number of level crossings at −10 dB (10 dB below average power) is 65 crossings per second, and the average duration is 1.54 ms, as mentioned in Sec. 13.2.

Thus, a word length must be designed to fit in these two cases.

6. Handoff action is another factor in data modem design. Whenever a handoff occurs, a piece of data information is lost. The average would be 200 ms. The ARQ scheme would be useful for this purpose.
7. The data modem must satisfy a specific bit error rate (BER) and word error rate (WER) requirement. The BER is independent of vehicle speed, whereas the WER is not. The higher the throughput rate (baseband transmission rate), the higher the BER and WER. However, there are two data modem markets.
   a. Use of fast data rate—in real-time situations, a customer may need quick access to data but may not need a high degree of accuracy. Examples of such customers are police agencies, real estate agencies, etc.
   b. Use of accurate data—for transmitting figures requiring a high degree of accuracy, a slow data transmission rate is needed. These customers need data accuracy more than fast acquisition, as in banking or computer applications.

13.6.2 Testing

Any data modem operating in a cellular mobile unit must demonstrate that its BER and WER satisfy the following conditions.

1. The data modem should be tested while the mobile unit is parked (stationary) at the side of the highway. Because the mobile radio environment is very noisy, the spike noise resulting from ignition-induced combustion and the sharp fades caused by the noise of passing trucks would also affect data transmission.
2. The data modem should be tested while the mobile unit is driven at different speeds, say 5, 10, 45, and 60 mi/h at the boundary of the cell.
3. The data modem should be tested during handoff conditions.

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Chapter

14

Digital Systems

14.1 Why Digital?

14.1.1 Advantages of digital systems

In an analog system, the signals applied to the transmission media are continuous functions of the message waveform. In the analog system, either the amplitude, the phase, or the frequency of a sinosoidal carrier can be continuously varied in accordance with the voice or the message.

In digital transmission systems, the transmitted signals are discrete in time, amplitude, phase, or frequency, or in a combination of any two of these parameters. To convert from analog form to digital form, the quantizing noise due to discrete levels should be controlled by assigning a sufficient number of digits for each sample, and a sufficient number of samples is needed to apply the Nyquist rate for sampling an analog waveform.

One advantage of converting message signals into digital form is the ruggedness of the digital signal. The impairments introduced in the medium in spite of noise and interference can always be corrected. This process, called *regeneration,* provides the primary advantages for digital transmission. However, a disadvantage of this ruggedness is increased bandwidth relative to that required for the original signal.

The increased bandwidth is used to overcome the impairment intro-

duced into the medium. An analogy to this is given in Sec. 14.1.2. In addition to the cost advantage, power consumption is lower and digital equipment is generally lighter in weight and more compact. In mobile cellular systems, there are more advantages in applying digital technology.

14.1.2. Analogy to modulation schemes

We may apply the analogy of moving books to compare analog and digital transmission. We may use a daily event to describe the different modulation schemes in the communications field. Consider the following scenario. On one side of a large hall, 20 books have been piled up (signal to be sent). We want to move these books to the other side of the hall (receiving end). Assume that the floor of the hall is not slippery and is flat (transmission medium). Then we would hire a strong, muscular person who could carry the 20 books with both hands and safely transport them to the other side without dropping a single book. This is analogous to a double sideband. The space which this person is occupying is equal to the width of the person's shoulders, as shown in Fig. 14.1*a*. If the person is very strong, one hand can be used to carry the 20 books while the other hand can be used to carry something else. This is analogous to a single sideband. If the floor in the middle of the hall contains many water puddles and (or) bumps, this is analogous to a rough medium. We might not hire a strong adult to carry the books because we might be afraid that the person would fall and drop all the books in the middle of the hall. Therefore, we might hire 10 children, each one carrying two books as seen in Fig. 14.1*b*. Among 10 children, perhaps only 1 or 2 might fall in the middle of the hall and drop the books. But most of the books will be carried to the other side of the hall. The 10 children, each carrying two books, is analogous to frequency modulation (FM). Spreading the signal into a wide spectrum (children) applies the first principle of using "spread spectrum" in FM.

A similar analogy is shown in Fig. 14.1*c*. Besides puddles and bumps, there are three guard jammers to stop anyone from bringing books over to the other side of the hall. In such a case we would have to hire as many as 20 children, each carrying one book and running against the guards. Some of the children may fall because of the puddles and bumps and some may be blocked by the guards, but most of the books will arrive on the other side of the hall.

If a force of 20 children is not sufficient to carry the books to the other side, then we must hire at least 40 children, each carrying half a book. This is the concept of spread spectrum. The spread-spectrum technique is used for combatting rough media and enemy jamming in military communications.

In each modulation scheme, energy must be confined within a specified bandwidth. In this chapter we will demonstrate the regular modulation schemes, phase shift keying (PSK) and frequency shift keying (FSK). We will also demonstrate how the schemes, MSK, GMSK, and GTFM can be applied to confine energy when transmitting a signal in the medium. The analogy can be extended to the adult or child who saves space while carrying books (Fig. 14.2*a*). The adult carrying the books (Fig. 14.2*b*) is taking too much space, and this method is not recommended. Common sense indicates that if the person wants to carry books through a crowd, the approach illustrated in Fig. 14.2*a* is more suitable than that illustrated in Fig. 14.2*b*. This analogy has been presented in the hopes of stimulating the readers to think about modulation schemes.

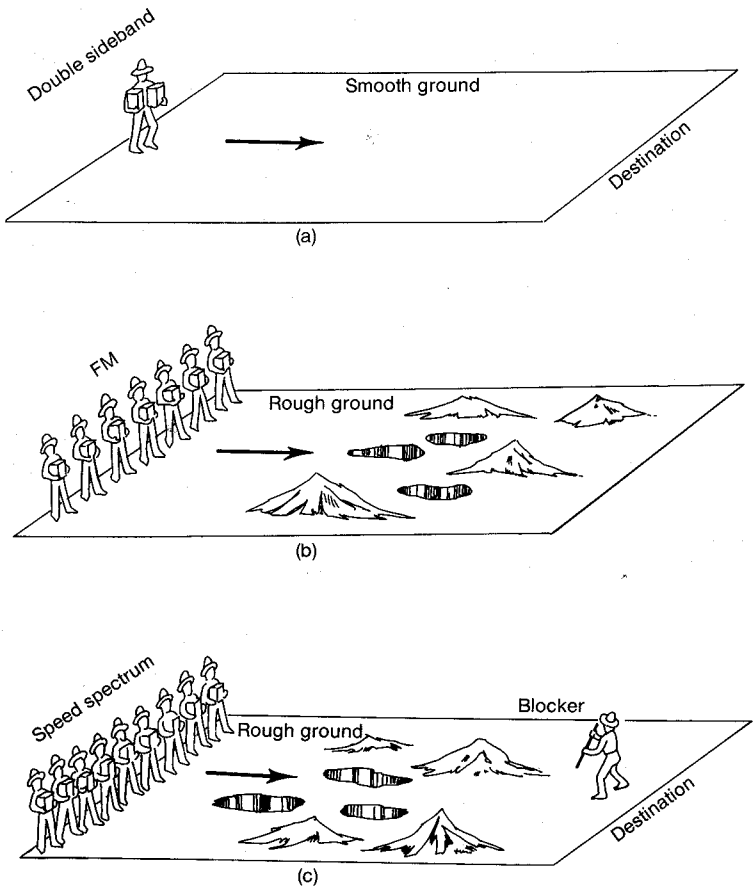

Figure 14.1 Analogy between radio transmission and moving blocks. (*a*) Walking over smooth ground; (*b*) ground with rough conditions; (*c*) ground with rough conditions plus blocks.

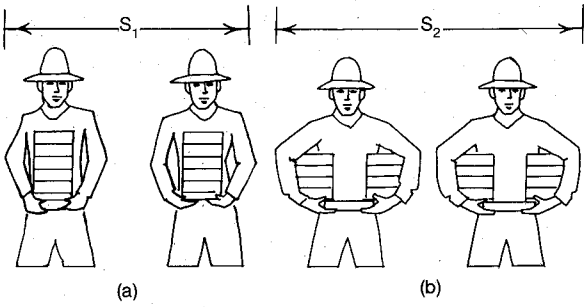

Figure 14.2 Analogy between space utilization and spectrum efficiency. (*a*) A satisfactory space utilization. (MSK, GMSK, or GTFM), (*b*) unsatisfactory space utilization (PSK).

14.2 Introduction to Digital Technology

14.2.1 Digital detection

There are three forms of digital detection: coherent detection, differentially coherent detection, and noncoherent detection. Coherent detection requires a reference waveform accurate in frequency and phase, and the use of a phase-coherent carrier tracking loop for each modulation technique. In a mobile radio environment, noncoherent detection is much easier to implement than coherent detection is. A form of detection that is intermediate in difficulty of implementation is called *differential PSK*. Differential PSK does not need absolute carrier phase information, and it circumvents the synchronization problems of coherent detection. The phase reference is obtained by the signal itself, which is delayed in time by an exact bit of spacing. This system maintains a phase reference between successive symbols and is insensitive to phase fluctuation in the transmission channel as long as these fluctuations are small during each duration of a symbol interval T. In differential binary-phase shift-keying (DBPSK) a symbol is a bit. The weak point of this scheme is that whenever there is an error in phase generated by the medium, two message error bits will result.

There are several aspects of digital detection.[1]

Carrier recovery. Carrier recovery for the suppressed carrier signal $A(t)\sin(\omega_0 t)$ plus noise $n(t)$ can be obtained by two methods. A squaring or frequency-doubling loop can be used (see Fig. 14.3). The loop contains a phase-locked loop as shown in Fig. 14.4. The phase-locked loop maintains a constant phase ϕ_n of $\cos(2\pi f_0 t + \phi_n)$, which is the recovered carrier. Another carrier-recovery technique uses the Costas loop, which generates a coherent phase reference independent of the binary mod-

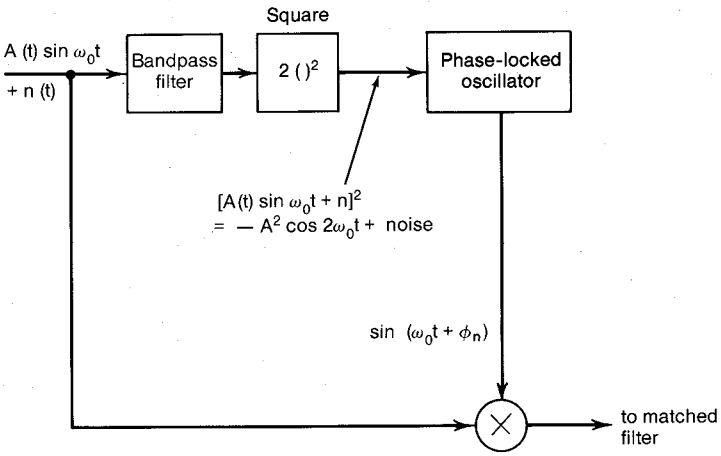

Figure 14.3 Block diagram of the square-law carrier-recovery technique.

ulation by using both in-phase and quadrature channels. The Costas loop (Fig. 14.5) is often preferred over the squaring loop because its circuits are less sensitive to center-frequency shifts and are generally capable of wider bandwidth operation. In addition, the Costas loop results in circuit simplicity.

Carrier-phase tracking (phase-locked loop). Carrier-tracking accuracy depends on several system parameters, including the phase noise in the carrier introduced by various oscillator short-term stabilities, carrier-frequency drifts, carrier-tracking-loop dynamics, transient response, the acquisition-performance requirement, and the signal-to-noise ratio S/N in the carrier-tracking loop. The phase-locked loop in

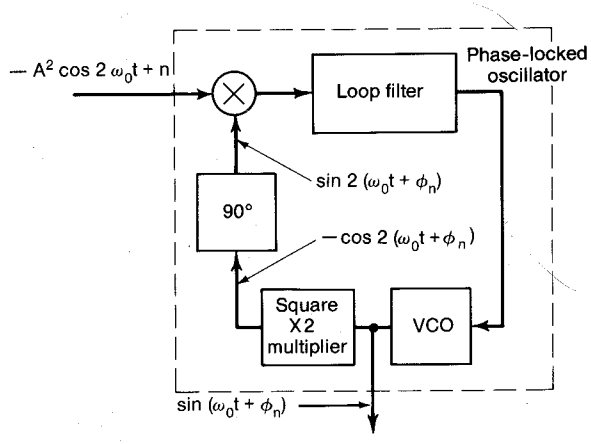

Figure 14.4 Phase-locked oscillator.

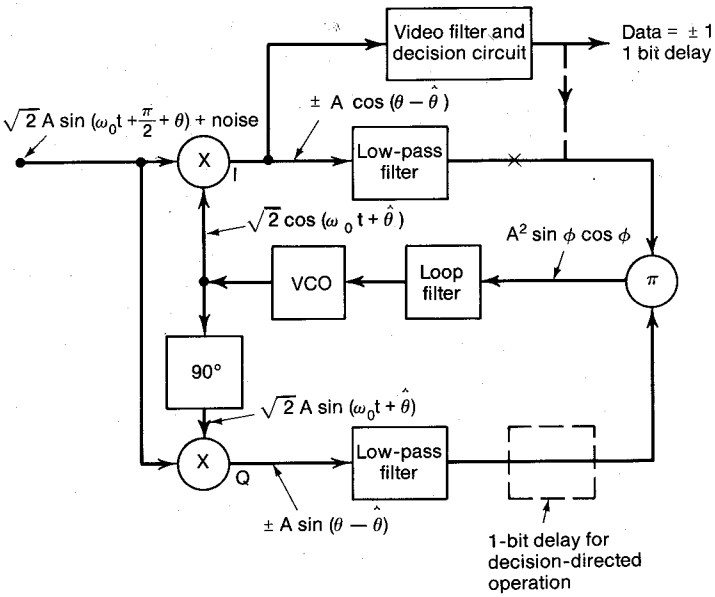

Figure 14.5 Costas loop OSK carrier-recovery circuit and bit detector; the phase error is defined as $\phi \triangleq \theta - \hat{\theta}$. The decision-directed configuration is shown in dashed lines.

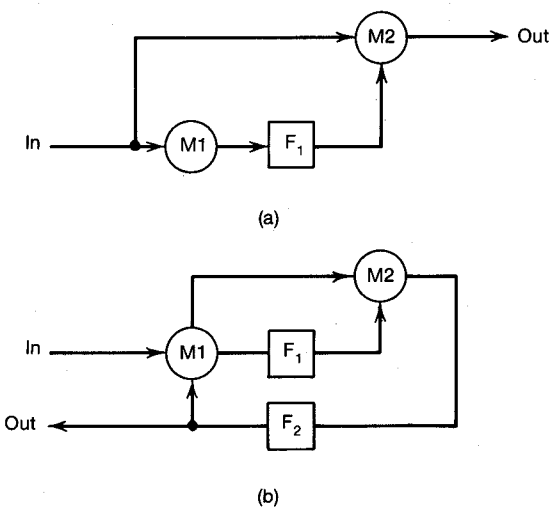

Figure 14.6 Cophase combining techniques. (*a*) Feedforward cophase combining; (*b*) feedback cophase combining.

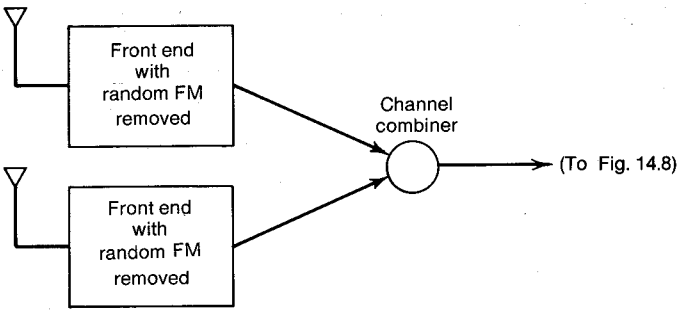

Figure 14.7 A two-branch diversity receiver.

the carrier-recovery tracking loop must have sufficient noise bandwidth to track the phase noise of the carrier. For a given carrier-phase–noise spectrum, one can compute the phase-locked-loop noise bandwidth B required to track the carrier. Clearly, too large a noise bandwidth can permit the occurrence of the thermal noise effect.

Phase-equalization circuits—for cophase combining

1. *Feedforward.* A circuit using two mixers to cancel random FM can be used (see Fig. 14.6*a*) as a phase-equalization circuit in each branch of an N-branch equal-gain diversity combiner.
2. *Feedback.* A modified circuit from the feedforward circuit is shown in Fig. 14.6*b*. This circuit is also used for each branch of an N-branch equal-gain diversity combiner. The feedback combiner is also called a Granlund combiner.
3. *The total combining circuit.* As shown in Fig. 14.7, either a feedback or a feedforward circuit can be used in the combiner to form a two-branch equal-gain combiner. The circuit connects to a coherent match-filter receiver for BPSK as shown in Fig. 14.8.

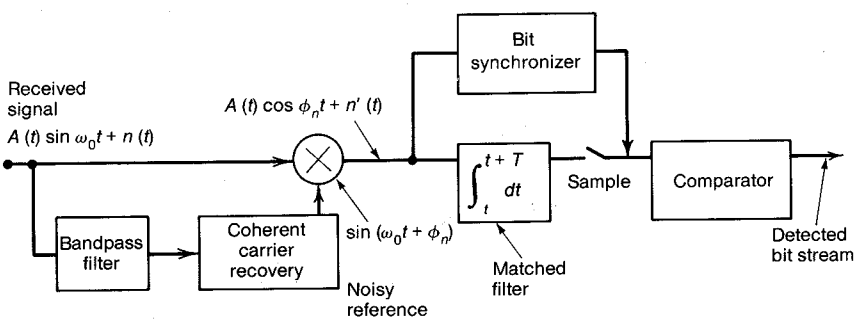

Figure 14.8 Coherent matched-filter receiver for BPSK.

Bit synchronization. Power-efficient digital receivers require the installation of a bit synchronizer. Bit synchronization commonly applies self-synchronization techniques, that is, it extracts clock time directly from a noisy bit stream. There are four classes of bit synchronizer.

1. *Nonlinear-filter synchronizer.* This open-loop synchronizer is commonly used in high-bit rate links which normally operate at high signal-to-noise ratios.
2. *The data-transition tracking synchronizer.* This closed-loop synchronizer combines the operations of bit detection and bit synchronization. It can be employed at low signal-to-noise ratios and medium data rates.
3. *Early-late synchronizer* This synchronizer uses early- and late-gate integral and dump channels, which have absolute values. It is simpler to implement than the data-transition tracking synchronizer and less sensitive to dc offsets.
4. *Optimum synchronizer.* This synchronizer provides an optimal means of searching for the correct synchronization time slot during acquisition. However, this approach generally is not practical.

14.2.2 Modulation for digital systems

There are several aspects of digital modulation, and they are described below.

Requirements. Basic digital modulation techniques are amplitude-shift keying (ASK), frequency-shift keying (FSK), phase-shift keying (PSK), and hybrid modulation techniques involving amplitude, frequency, and phase-shift keying.

In mobile cellular systems, the selection of a digital modulation for radio transmission involves satisfaction of the following requirements: (a) narrower bandwidths, (b) more efficient power utilization, and (c) elimination of intermodulation products.

Narrower bandwidths. For all the forms of modulation, it is desirable to have a constant envelope and, therefore, utilize relatively narrower bandwidths. In these cases, FSK and PSK are recommended. For example, multiphase-shift-keying (MPSK) for large values ($M > 4$) has greater bandwidth efficiency than does BPSK or QPSK but power use is less efficient.

More efficient power utilization. It is preferable to provide more channels for a given power level. Therefore, enhanced power utilization is essential. Besides, the FCC has limited the total power (100 W) to be radiated from each base-station antenna. This limitation

governs the number of channels which can be served given the power allowed for each channel.

Elimination of intermodulation products. QPSK is commonly used with a transmission efficiency of about 1 to 2 bps/Hz. This value has been found to offer satisfactory trade-off between efficient frequency utilization and transmitter power economy. However, in mobile radio links, when nonlinear class C power amplifiers are used, any spurious radiation should be suppressed. For reducing the spurious signals, we are selecting a constant or low-fluctuation envelope property. There are two types of broadly classified modulations.

Modulation schemes. There are several modulation schemes.

1. *Modified QPSK.* There are two kinds of QPSK besides a regular QPSK with restricted phase-transition rules.
   a. QPSK. The conventional QPSK shown in Fig. 14.9*a* has phase

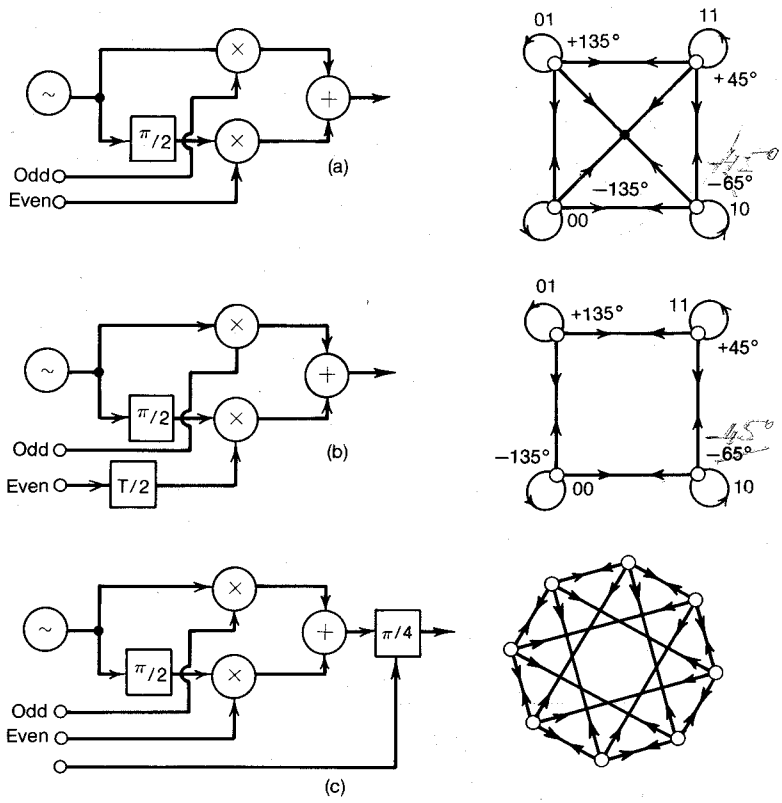

Figure 14.9 Modulator constitution and signal-space diagrams of (*a*) conventional QPSK, (*b*) offset QPSK, and (*c*) $\pi/4$ shift. (*After Hirade et al., Ref. 2, p. 14.*)

ambiguity. The ideal QPSK signal waveform

$$A \sin[\omega_0 t + \theta_m(t)] = \pm\frac{A}{\sqrt{2}} \sin\left(\omega_0 t + \frac{\pi}{4}\right) \pm \frac{A}{\sqrt{2}} \cos\left(\omega_0 t + \frac{\pi}{4}\right) \tag{14.2-1}$$

where $\theta_m = (0, \pi/2, \pi, 3\pi/2)$ and the value of θ_m should match the sign of Eq. (14.2-1); that is, $\theta_m = 0$ for $(+, +)$, $\theta_m = \pi/2$ for $(+, -)$, $\theta_m = \pi$ for $(-, +)$, and $\theta_m = 3\pi/2$ for $(-, -)$.

Bit pair	Absolute phase
0 0	0
0 1	$\pi/2$
1 1	π
1 0	$3\pi/2$

b. *Offset QPSK (OQPSK).* This scheme is a QPSK, but the even-bit stream is delayed by a half-bit interval with respect to the odd 1 bit as shown in Fig. 14.9*b*.

c. *$\pi/4$-shift QPSK.* A phase increment of $\pi/4$ is added to each symbol shown in Fig. 14.9*c*.

Both OQPSK and $\pi/4$-shift QPSK have no π-phase transition; therefore, no phase ambiguity would occur as in QPSK. However, intrinsically they produce a certain amount of residual envelope fluctuation. Sometimes, a phase-locked loop (PLL) is inserted at the modulation output to remedy this problem.

2. *The differential encoding of QPSK (DQPSK).* This is the same as in DBPSK, but the differential encoding of the bit pairs selects the phase change rather than the absolute phase. However, DQPSK has phase ambiguity just like QPSK.

Bit pair	Phase changes
0 0	0
0 1	$\pi/2$
1 1	π
1 0	$3\pi/2$

The QPSK carrier recovery would be slightly different than that of BPSK.[2]

3. *Modified FSK—continuous-phase frequency-shift-keying (CP-FSK) with low modulation index*
   a. Minimum-shift-keying (MSK)—also called *fast FSK* (FFSK)
   b. Sinusodial FSK(SFSK)
   c. Tamed FSK (TFSK) or tamed frequency modulation (TFM)
   d. Gaussian MSK (GMSK)
   e. Gaussian TFM (GTFM)

Since all the schemes listed above are CP-FSK (and have a low modulation index, they intrinsically have constant envelope properties, unless severe bandpass filtering is introduced to the modulator output. In MSK the frequency shift precisely increases or decreases the phase by 90° in each T second. Thus, the signal waveform is

$$s(t) = \sin\left(\omega_0 t + 2\pi \int_0^t s_i \, d\tau + \frac{n\pi}{2}\right) \qquad 0 < t < T$$

where

$$s_i = \begin{cases} s_1 = \dfrac{1}{4T} & \text{for a data bit 1} \\ s_2 = -\dfrac{1}{4T} & \text{for a data bit 0} \end{cases} \tag{14.2-2}$$

or

$$s(t) = \sin\left(\omega_0 t + \frac{n\pi}{2} \pm \frac{\pi t}{2T}\right) \tag{14.2-3}$$

or

$$s(t) = \cos\left(\pm \frac{\pi t}{2T}\right) \sin\left(\omega_0 t + \frac{n\pi}{2}\right) + \sin\left(\pm \frac{\pi t}{2T}\right) \cos\left(\omega_0 t + \frac{n\pi}{2}\right)$$

Comparing Eq. (14.2-1) with Eq. (14.2-3), we find that the two equations are very similar. In fact, the phase-modulation waveforms of the I- and Q-channel modulations of OQPSK are modulated by sine and cosine waveforms, and thus the output will be identical to that of MSK. Note that it is necessary to modulate both the I and Q channels during each bit interval to retain the constant envelope of $s(t)$.

Because the phase is continuous from bit to bit, the spectral sidebands of MSK or OQPSK fall off more rapidly than in BPSK or QPSK (see Fig. 14.10). Although MSK demonstrates a superior property in terms of its out-of-band spurious spectrum suppression without any filtering, its out-of-band spurious spectrum suppression will not satisfy

the severe requirements in the single carrier per channel (SCPC) communications. The sharp edges in MSK phase-transition trajectories (Fig. 14.11) can be smoothed by some premodulation baseband filtering. The SFSK shows a smoother phase transition than does the MSK but little improvement in the suppression of out-of-band spurious spectrum. The TFM is a modified MSK using the partial response encoding rule as the phase-transition rule. The smoothed phase trajectory of TFM is shown in Fig. 14.11. The outstanding suppressions for the out-of-band spectrum of TFM is shown in Fig. 14.10. However, this outstanding suppression of the out-of-band spurious spectrum can be achieved by using a suitable premodulation baseband filtering on MSK, such as a baseband gaussian filtering, as shown in Fig. 14.12. The GMSK with $B_bT = 0.2$ has the same power spectrum density curves as does TFM, where B_b is the baseband bandwidth. Furthermore, GMSK is easier to implement than TFM.

The following parameters are defined and are in the figures that follow.

B_i	ideal bandpass
B_s	channel separation (bandwidth)
f_b	bit rate of voice coding $= 1/T$
$1/T$	transmission bit rate (16 kbps)
B_iT	normalized bandwidth of the ideal bandpass filter
B_bT	normalized bandwidth of a gaussian filter

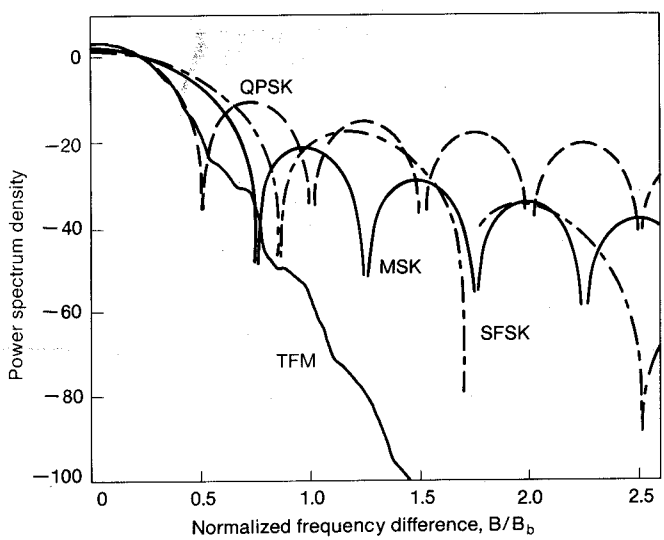

Figure 14.10 Power spectrum density functions of QPSK, MSK, SFSK, and TFM. (*After Hirade et al.*, Ref. 2, p. 15.)

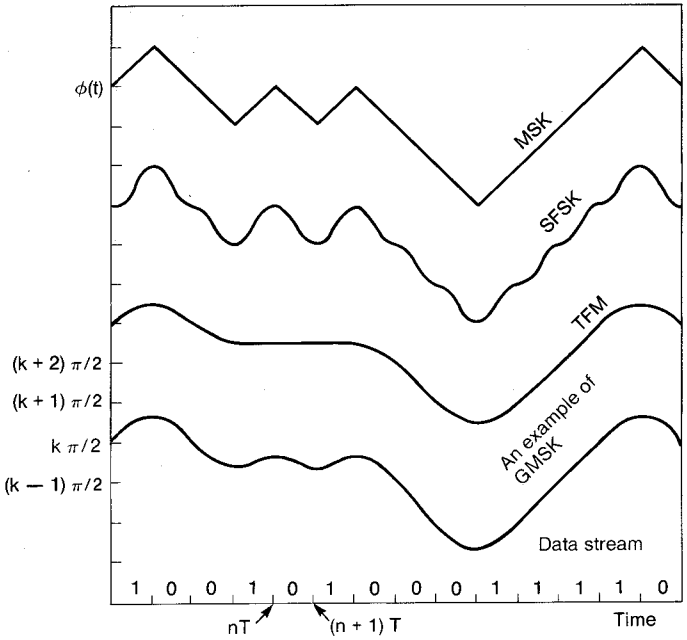

Figure 14.11 Phase-transition trajectories of MSK, SFSK, TFM, and GMSK (*After Hirade et al., Ref. 2, p. 15.*)

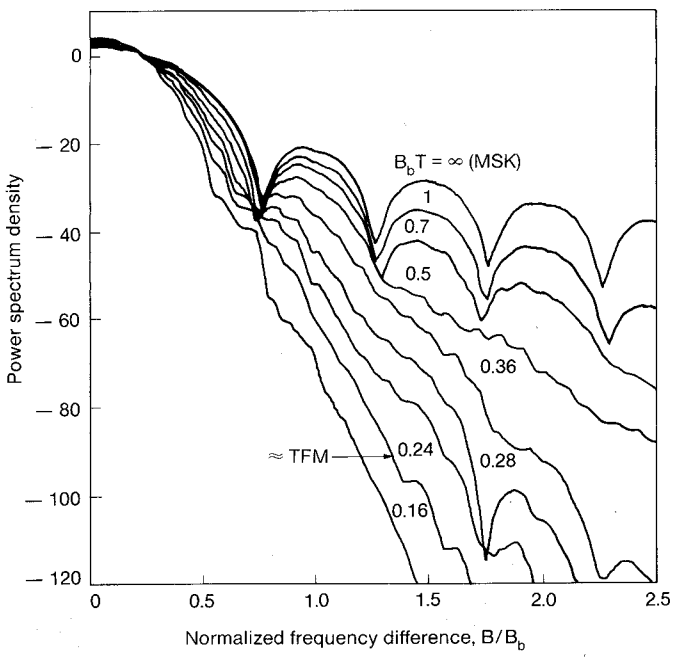

Figure 14.12 Power spectrum density functions of GMSK. (*After Hirade et al., Ref. 2, p. 15.*)

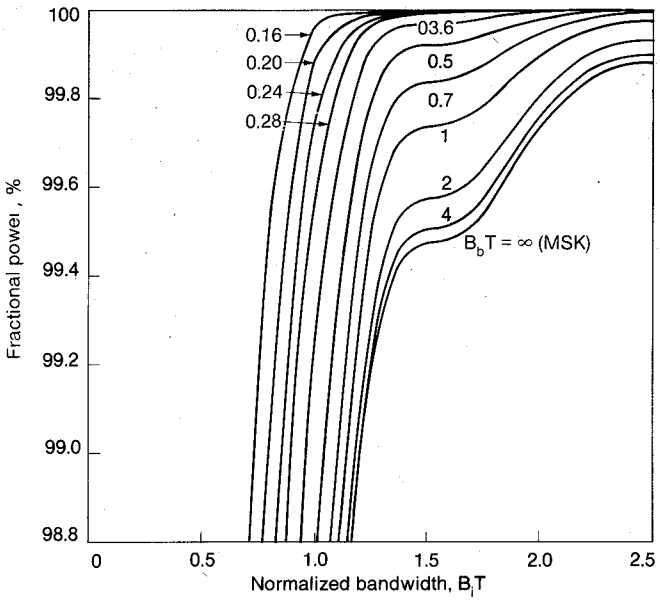

Figure 14.13 Fractional power of GMSK signal. (*After Hirade et al., Ref. 2, p. 15.*)

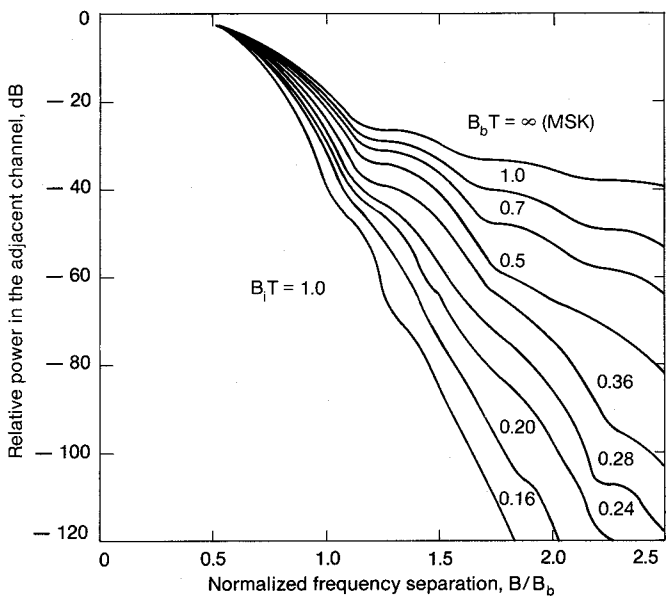

Figure 14.14 Relative power radiated in the adjacent channel. (*After Hirade et al., Ref. 2, p. 16.*)

Figure 14.13 shows the fractional power in percentage of GMSK signal exceeding the normalized bandwidth B_iT with different values of B_bT ($B_bT = \infty$ means no filter). This becomes conventional MSK. Figure 14.14 shows that the relative power radiated in the adjacent channel for B_sT is equal to 1.5. Let $B_s = 30$ kHz, then the bit rate $f_b = (1/T) = 20$ kbps. For a normalized filter bandwidth $B_bT = 0.24$ or $B_b = 4.8$ kHz, and the relative power in the adjacent channel is -60 dB.

Demodulation. When the signal is received, we would like to know the performance of the various demodulation schemes. Some demodulation schemes are better than others regardless of what the modulated signal is like. The orthogonal coherent detector proposed by de Buda[3] can be used for both MSK and TFM.

BER performance is always a good criterion for the comparison of different modulation schemes. Measured BER performance is shown in Fig. 14.15 with a normalized channel bandwidth $B_iT = 0.75$. The family of curves show the BERs for the different filter bandwidths. Also, TFSK (TFM) is plotted for comparison. GMSK with the filter

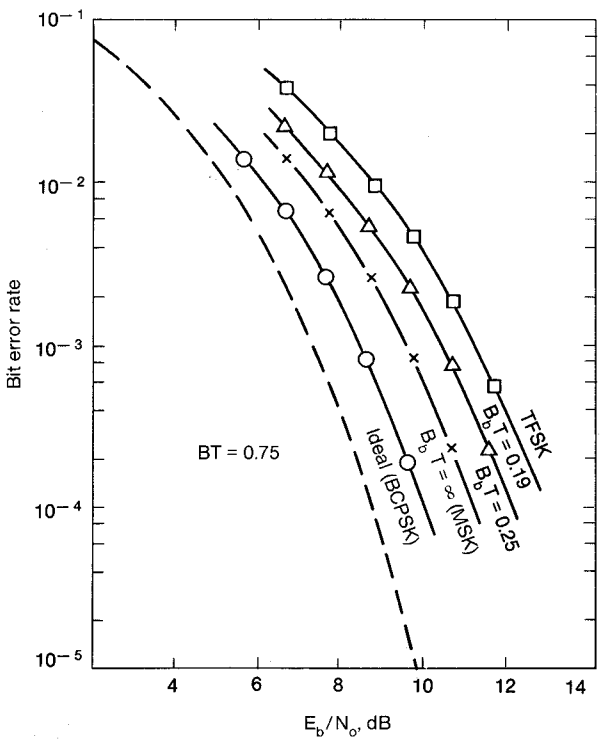

Figure 14.15 Measured BER performance. (*After Hirade et al., Ref. 2, p. 17.*)

bandwidth of $B_bT = 0.19$ is superior to TFSK. However, one disadvantage of narrowing the channel spectrum is increasing the BER, that is, degrading performance. Sometimes we have to consider whether it is worthwhile to use a gaussian filter.

14.3 ARQ Techniques

14.3.1 Different techniques[4,5]

Automatic-repeat-request (ARQ) techniques include the coding and retransmission request strategy for delivering a message. Preceding the message is a header which contains the source and destination address and useful routing information. Every ARQ message must have a header. There are two principal ARQ techniques.

1. *Stop-and-wait ARQ*. The message originator stops at the end of each transmission to wait for a reply from the receiver (see Fig. 14.16*a*). Then the following steps can be taken.
   a. No forward error correction is used—ARQ(*a*).
   b. Both forward error correction and error detection coding are used—ARQ(*b*).
   c. Error-detection parity bits are sent, but not the forward error-correction parity bits, which assumes that the probability of an error-free message is great—ARQ(*c*).
2. *Selective retransmission*. When many words are transmitted at once, each word individually can apply error detection, not the

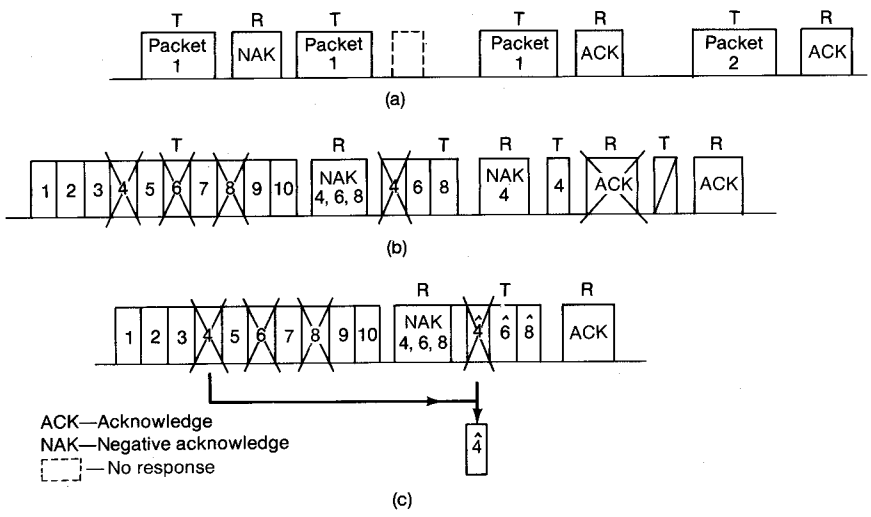

Figure 14.16 ARQ transactions. (*a*) Stop-and wait ARQ; (*b*) selective retransmission with ARQ(*b*); (*c*) selective retransmission with ARQ(*c*).

message as a whole. Only those words containing detected errors are sent back. This scheme is called *selective retransmission*. Selective retransmission with ARQ(*b*) is shown in Fig. 14.16*b*. Selective retransmission with ARQ(*c*) is shown in Fig. 14.16*c*.

14.3.2 The expected number of transmissions

Stop-and-wait ARQ [apply ARQ(*a*) and ARQ(*b*) only]. Let P_{ew} be a word error rate (WER), and let a message consist of N words. Now the required number of transmissions depends on all N words being successfully transmitted. The expected number of transmissions is

$$E_N = \frac{1}{(1 - P_{ew})^N} \tag{14.3-1}$$

assuming independence errors between words and that all words have the same P_{ew}. This assumption can be considered valid for cases where vehicle speed is high. Equation (14.3-1) indicates that the number of transmissions E_N increases more quickly with increasing the message length, that is, as N increases. Equation (14.3-1) is plotted in Fig. 14-17*a*.

Selective retransmission (SRT) with ARQ(*b*). Assume that the number of transmissions of one word is independent of the number of transmissions of any other word. The expected number of transmissions required for sending an N-word message with fewer than i transmissions is

$$E_N = \sum_{i=1}^{\infty} [1 - (1 - P_{ew}^{i-1})^N] \tag{14.3-2}$$

Equation (14.3-2) is plotted in Fig. 14.17*b*. By comparing Fig. 14.17*a* with Fig. 14.17*b*, we see that stop-and-wait ARQ would require a greater number of transmissions to deliver a message than would selective retransmission with the same block error probability.

Selective retransmission (SRT) with ARQ(c). In this scheme, ARQ(*c*) defined in the stop-and-wait ARQ is applied to the selective retransmission scheme. Let the first-transmission probability be P_1 and a retransmission probability be P_2. The expected number of transmissions required for sending an N-word message with fewer than i transmissions is

$$E_N = 1 + \sum_{i=2}^{\infty} [1 - (1 - P_1 P_2^{i-2})^N] \tag{14.3-3}$$

Because the forward error correction is added to retransmission, P_2 is smaller than P_1. Let P_2 be a positive integer power of P_1; that is, $P_2 = P_1^k$, where k represents the various powers of the feedforward error correction code. Equation (14.3-3) is plotted in Fig. 14.17*c* for $N = 10$. When $k = 1$, the curve shown in Fig. 14.17*c* is the same as $N = 10$ in Fig. 14.17*b*.

14.3.3 Transmission Efficiency *R*

The transmission efficiency is the ratio of the number of information bits to the total number of transmission bits. Let a message consist of

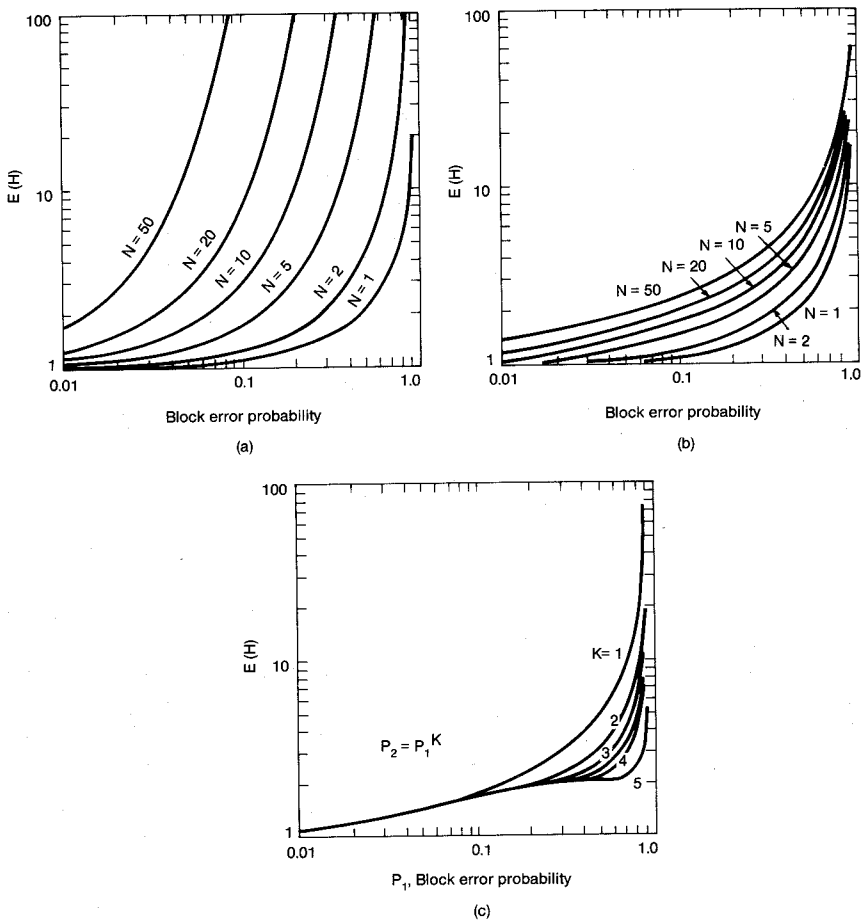

Figure 14.17 Expected number of transmissions for different kinds of ARQ. (*a*) $E(H)$, expected number of transmissions to deliver an N-block message for stop-and-wait ARQ. (*b*) $E(H)$, expected number of transmissions to deliver a 10-block message for SRT ARQ(*b*). (*c*) $E(H)$, expected number of transmissions to deliver a 10-block message for SRT ARQ(*c*) for various retransmission-failure probabilities.

N words, where B is number of bits per word, and

$$B = H + L \tag{14.3-4}$$

where H is the header bits and L is the information bits. Then

Stop-and-wait ARQ technique

$$R = \frac{NL}{(H + NB)E_N} \tag{14.3-5}$$

where E_N is as shown in Eq. (14.3-1).

Selective retransmission

$$R = \frac{NL}{HE_N + NBE_1} \tag{14.3-6}$$

where E_N is as shown in Eq. (14.3-2) and

$$E_1 = \frac{1}{(1 - P_{\text{ew}})} \tag{14.3-7}$$

Selective retransmission with ARQ(c)

$$R = \frac{NL}{HE_N + NBE_1} \tag{14.3-8}$$

where E_N is as shown in Eq. (14.3-3), and E_1 is

$$E_1 = 1 + \sum_{i=2}^{\infty} [1 - (1 - P_1 P_2^{i-2})] \tag{14.3-9}$$

14.3.4 Undetected error rates

In the previous sections, we saw that each transmission results in one of two outcomes, success or failure. This is called "hard detection." There is also a "soft detection." In soft detection three probabilities per transmission are denoted: P_c (success), P_d (detected error), and P_u (undetected error) per word. When each of these probabilities is identified for each transmission, then we can perform further calculations.

Stop-and-wait ARQ [apply ARQ(*a*) and ARQ(*b*)]. We can find the undetected error probability per single-word message.

$$P_{\text{um}} = \frac{P_u}{P_c + P_u} \tag{14.3-10}$$

where

$$P_c = 1 - P_{ew} \tag{14.3-11}$$

and

$$P_u \leq P_{ew}\, 2^{-m} \tag{14.3-12}$$

where m is the number of error-detection parity bits. The undetected error probability per N-word message is

$$P_{um} = \frac{1 - (1 - P_{ew})^N}{1 + (1 - P_{ew})^N(2^m - 1)} \tag{14.3-13}$$

Selective retransmission. In this technique, we consider b parity bits per each word instead of m parity bits for the whole message. The probability of undetected error for a single-word message is

$$P_{um} = \frac{P_{ew}}{P_{ew}(1 - 2^b) + 2^b} \tag{14.3-14}$$

The probability of undetected error for a N-word message is

$$P_{um} = 1 - \left[1 - \frac{P_{ew}}{P_{ew}(1 - 2^b) + 2^b}\right]^N \tag{14.3-15}$$

The word error rate P_{ew} is a function of vehicle speed. Since the word error rate P_{ew} is derived from the bit error rate, each bit in error can be dependent on or independent of the adjacent bit error related to the vehicle speed. Therefore, the word error rate is not easy to obtain. Any oversimplified model may give incorrect answers. We can obtain the word error rate from two extreme cases, one assuming that the speed approaches infinity and the other assuming that the speed approaches zero as described in Chap. 13.

14.4 Digital Speech[6]

Since digital technologies have evolved, an important study focuses on efficient methods for digitally encoding speech. Speech quality implies a measure of fidelity, which is difficult to specify qualitatively because human perception is involved. The two criteria used are

What is being said (low fidelity accepted by military systems)

Who says it (high fidelity important to commercial systems)

For instance, a military system examiner who comments that someone's speech quality is excellent may be referring to intellegibility and low system noise. It is irrelevant who speaks on the other side (of the examiner) or that the examiner has never spoken to this person before.

14.4.1 Transmission rates in speech coding

These rates are totally dependent on quality characterizations such as toll quality, commentary quality, communications quality, synthetic quality.

1. *Toll quality.* An analog speech signal is of toll quality when its frequency range is 200 to 3200 Hz; its signal-to-noise ratio is greater than or equal to 30 dB; and its harmonic distortion is less than or equal to 2.3 percent. Digital speech has to have a quality comparable to that of the toll quality of an analog speech signal.

Toll-Quality Transmission

Coder	kbps
Log PCM	56
ADM	40
ADPCM	32
SUB-BAND	24
Pitch Predictive ADPCM	24
APC, ATC, ΦV, VEV	16

2. *Commentary quality.* In general, the signal at bit rates exceeding 64 kbps generates a commentary-quality speech signal which is better than toll quality, but the input bandwidths are significantly wider than in a noncellular telephone system (up to 7 kHz).
3. *Communications quality.* At rates below 16 kbps, the signal in the range of 7.2 to 9.6 kbps is a communications-quality speech signal. The signal is highly intelligible but has noticeable reductions in quality and speaker recognition.

Communications-Quality Transmission

Coder	kbps
Log PCM	36
ADM	24
ADPCM	16
SUB-BAND	9.6
APC, ATC, ΦV, VEV	7.2

4. *Synthetic quality.* At 4.8 kbps and below, the signal provides synthetic quality and speaker recognition is substantially degraded.

Synthetic-Quality Transmission

Coder	kbps
CV, LPC	2.4
ORTHOG	1.2
FORMANT	0.5

14.4.2 Classes of coder

There are two classes of coder: waveform coders and source coders.

Waveform coders. The speech waveform can be characterized by

1. Amplitude distribution (in time domain)
2. Autocorrelation function (in time domain)
3. Power spectral density (in frequency domain)
4. Spectral flatness measure (removing redundancy in speech waveform)
5. Fidelity criteria for waveforms

$$\text{Coding noise} = \frac{1}{T}\int_0^T (\text{coding error})^2\, dt$$

where the coding error is equal to the amplitude difference (samples of a coded waveform minus the original input waveform).

The signal-to-noise ratio is expressed as

$$\frac{S}{N} = \overline{\left[\frac{(\text{input waveform})^2}{\text{coding noise}}\right]}$$

There are two types of speech waveform coders.

1. *Time-domain coders.* Pulse code modulation (PCM), differential pulse code modulation (DPCM), and delta modulation (DM) are commonly used. Adaptive predictive coding (APC) in time-domain coding systems is limited to linear predictors with changing coefficients based on one of the following three types:
   a. Spectral fine structure—in more periods
   b. Short-time spectral envelope—determined by the frequency

response of the vocal tract and by the spectrum of the vocal-cord sound pulses

c. Combination of types *a* and *b*

In time-domain coders, speech is treated as a single full-band signal; in time-domain predictive coders, speech redundancy is removed prior to encoding by prediction and inverse filtering so that the information rate can be lower.

2. *Frequency-domain coders.* The speech signal can be divided into a number of separate frequency components, and each of these components can be encoded separately. The bands with little or no energy may not be encoded at all. There are two types of coding:
   a. *Subband coding* (*SBC*). Each subband can be encoded according to perceptual criteria that are specific to that band.
   b. *Adaptive transform coding* (*ATC*). An input signal is segmented and each segment is represented by a set of transform coefficients which are separately quantized and transmitted.

Source coders—vocoders. The synthetic quality of source vocoder speech is not appropriate for commercial telephone application. It is designed for very low bit-rate channels. Vocoders use a linear, quasi-stationary model of speech production.

Sound source characteristics. The sound can be generated by voiced sounds, fricatives, or stops. The source for voiced sounds is represented by a periodic pulse generator. The source for unvoiced sounds is represented by a random noise generator. They are mutually exclusive.

System characterization. The acoustic resonances of the vocal tract modulate the spectra of the sources. Different speech sounds correspond uniquely to different spectral shapes. Vocoders depend on a parametric description of the vocal-tract transfer functions.

1. Channel vocoder—speech signal evaluated at specific frequencies
2. LPC (linear prediction code) vocoder—linear prediction coefficients that describe the spectral envelope
3. Format vocoder—specified frequency values of major spectral resonances
4. Autocorrelation vocoder—specified short-time autocorrelation function of the speech signal
5. Orthogonal function vocoder—specifies a set of orthonormal functions

Frequency-domain vocoders. A single coder is called a *channel vocoder.* Instead of transmitting the telephone signal directly, only the spectrum of each speech signal is transmitted; 16 values along the frequency

axis are needed. Each takes 20 ms and requires a bandwidth of $1/(2 \times 20 \text{ ms}) = 25$ Hz and the total frequency requirement is (16×25) or 400 Hz, which is one-tenth of the bandwidth of the speech signal itself.

Time-domain vocoders. Speech samples would have to be spaced $1/(2 \times 4000) = 0.125$ ms apart, which would require 30 samples to ensure a good quality. Then the frequency requirement is $30/(2 \times 0.125 \text{ ms})$ or 120 kHz. For digital transmission, the number of bits per correlation sample used by a time-domain vocoder should be about twice as high for spectral samples in frequency-domain vocoders. Therefore, time-domain vocoders are not desirable. Yet, one of the most successful innovations in speech analysis and synthesis is linear predictive coding (LPC), which is based on autocorrelation analysis.

LPC vocoders. LPC vocoders constitute an APC system in which the prediction residual has been replaced by pulse and noise sources. For the telephone band, the number of predictor coefficients is 8. For low-quality voice, the number can be as small as 4.

Hybrid waveform coders-vocoders. A hybrid arrangement of SBC, APC, and LPC is coming into vogue where a portion (lower-frequency band) of the transmission is accomplished by waveform techniques and a portion (upper-frequency band) by voice-excited vocoder techniques.

14.4.3 Complexity of coders

A relative count of logic gates is used to judge the complexity of the coders as follows:

Relative complexity	Coder	
1	ADM:	adaptive delta modulator
1	ADPCM:	adaptive differential PCM
5	SUB-BAND:	subband coder (with CCD filters)
5	P-P ADPCM:	pitch-predictive ADPCM
50	APC:	adaptive predictive coder
50	ATC:	adaptive transform coder
50	ΦV:	phase vocoder
50	VEV:	voice-excited vocoder
100	LPC:	linear-predictive coefficient (vocoder)
100	CV:	channel vocoder
200	ORTHOG:	LPC vocoder with orthogonalized coefficients
500	FORMANT:	formant vocoder
1000	ARTICULATORY:	vocal-tract synthesizer; synthesis from printed English text

Of these coders, LPC is attractive because of its performance and degree of complexity.

14.5 Digital Mobile Telephony

14.5.1 Digital voice in the mobile cellular environment

Since voice communication is the key service in cellular mobile systems, when we think of the digital systems, we must think of a digital voice.

In present-day mobile cellular systems, transmission of a digital voice in a multipath fading environment is a challenging job. The major considerations in implementing digital voice in cellular mobile systems are discussed below, along with a tentatively recommended transmission rate for the cellular mobile system.

Digital voice in the mobile radio environment

1. The criterion for judging a good digital voice through a wire line is employed in three existing digital voice schemes.
   a. In a continuously variable step delta (CVSD) modulation scheme, the present transmission rate is 16 kbps. This is not toll-quality voice transmission and is commonly used by the military.
   b. In a LPC scheme, the present transmission rate of 2.4 kbps provides a synthetic quality voice, but a rate of 4.8 kbps using vector quantization[8] may provide a communications-quality voice. A rate of 16 kbps can provide a toll-quality voice.[19]
   c. In a pulse code modulation (PCM) scheme, the present transmission rates of 32 kbps and 64 kbps are commonly used; 32 kbps is used by the military while 64 kbps is used commercially.

   Of the three schemes, LPC seems most attractive because of its low transmission rate. However, LPC is more vulnerable in terms of distortion to the mobile fading environment.
2. Digital voice has to be processed in real time, which imposes constraints on the digital processing time. This adversely affects LPC but not CVSD.
3. When sending a digital stream (voice) through a radio channel in a fading environment, in general, an LPC scheme needs more code protection than CVSD scheme does because LPC is not implemented in a continuous waveform in either the frequency domain or the time domain while CVSD is implemented in a continuous waveform in the time domain.

4. Because the mobile unit is moving, sometimes rapidly, sometimes slowly, insertion of extra synchronization bits is needed in the normal digital stream.

Considerations for a digital voice transmission in cellular mobile system. The following factors are significant.

1. *Digital transmission rate*
   a. *Present cellular signaling rate.* The present signaling format is designed on the assumption that the mobile unit moves at an average of 30 mi/h and that the transmission rate is 10 kbps. The 21 synchronization bits (10 synchronization bits and 11 frame bits) occur in front of every code word of 48 bits to ensure that the bits are not falling out of synchrony before the resynchronization takes place.
   b. *Consideration of LPC scheme.* If a rate of 4.8 kbps using LPC for a communications-quality voice is accepted its rate is almost half of the present transmission rate, and at this transmission rate a 48-bit word would be acceptable in a fading environment. The resynchronization scheme for a mobile receiver should take place in front of every code word of 48 bits [(21 synchronization bits) + (a code word of 48 bits) = 69 bits]. The number of synchronization bits is almost half the number of bits in a code word. Therefore, the transmission rate would be approximately $(4.8 \times 1.5) = 7.2$ kbps.
   c. *Redundancy of transmission.* The protection of synchronization in a mobile radio environment is not sufficient. If the digital stream were to occur in a signal fade, partial or whole code words would be lost. In order to prevent fading, redundancy of transmission is often used. We would take a minimum redundancy scheme; for example, we would transmit the same message bits three times and take a "2-out-of-3 majority vote" on each bit to minimize the fading impairment of the message bits. For LPC of 4.8 kbps, an RF transmission rate of (4.8 kbps $\times$ 1.5) 3 = 21.6 kbps is needed.* It is reasonable for a 30-kHz channel to carry a transmission rate of 21.6 kbps over a severe fading medium. When an RF transmission rate is given, the channel bandwidth can be narrower with a tradeoff of transmitted powers. This point has been described in Sec. 13.4.8.
   d. *Modulation, diversity, coding, ARQ, and scrambling.* Diversity and modulation can help in reducing the RF transmission rate for the digital voice. However, ARQ schemes, fancy coding schemes, and complicated scrambling schemes

* Applying diversity schemes can reduce this rate.

cannot be implemented for voice transmission. This is because the digital voice must be processed in real time, and these three schemes usually require a fair amount of time for processing. These schemes can be used for data transmission.

2. *Word error rate.* In the multipath fading environment, the bit error rate P_e is not the only parameter for voice-quality measurement; the word error rate P_w is also important and varies with vehicle speed. However, information on the word error rate for transmission of digital voice over a mobile radio environment only appears in two extreme cases (see Sec. 13.2.3). Assume that we know the required P_e and P_w. We can convert P_e and P_w to a required carrier-to-noise ratio C/N. If a two-branch diversity scheme is applied after a 2-out-of-3 majority-vote redundancy scheme has been used, the bit error rate of 10^{-3} in a relatively slow fading case requires a C/N level of approximately 15 dB. With the C/N level, a word error rate of a 40-bit word is about 10^{-3} (see Fig. 13.3*a*). In general, if the word error rate is the same as or lower than the bit error rate for a given C/N, the C/N level is acceptable. In our case, P_w and P_e are the same at $C/N = 15$; therefore, the $C/N = 15$ dB is justified.
3. *Relationship between C/N and E_b/N_0.* The relationship between the carrier-to-noise ratio C/N, the energy-per-bit-to-noise-per-hertz ratio E_b/N_0, the transmission rate R, and the bandwidth B can be expressed as

$$\frac{C}{N} = \frac{E_b}{N_0}\frac{R}{B} \tag{14.5-1}$$

When the number of levels C/N increases, the bandwidth decreases. Keeping E_b/N_0 constant, we see that when the bandwidth decreases, the required carrier-to-noise ratio C/N increases. Previously we calculated that $C/N = 15$ dB works for a two-level (binary) system. If the number of levels increases, the C/N will be higher than 15 dB.

Example 14.1 Let $E_b/N_0 = 15$ dB for a two-level system and R_0 and B_0 be the transmission rate and transmission bandwidth, respectively, of the two-level system. Now if we reduce the bandwidth $B_1 = 0.5B_0$, then

$$\left(\frac{C}{N}\right)_1 = (31.6)\frac{R_0}{0.5B_0} = 2\left(\frac{C}{N}\right)_0$$

$$= \left(\frac{C}{N}\right)_0 + 3 \text{ dB}$$

This means that the power increases by 3 dB. If the transmitted power was 50 W, now it is 100 W.

14.5.2 Evaluation of digital voice quality

In general, there are two methods for evaluating digital voice quality.

1. *Listener's opinion.* Use one 16-kbps voice coder and one 8-kbps voice coder in a specified digital system. Then find the two required carrier-to-interference ratios C/I based on the listener's opinion in a Rayleigh fading environment. Then compare the same voice quality with that from an analog FM system at $C/I \geq 18$ dB.

2. *Diagnostic rhyme test (DRT).* The voice quality of a digital format is often tested by DRT. Using the DRT score of 90 as a criterion, above 90 means acceptable for synthetic-quality voice and below 90 means unacceptable. Thus, the bit error should be less than 10^{-3} for an LPC of 2400 bps in a gaussian noise environment. Voice evaluation for an LPC of 4.8 kbps does not appear in the literature. The voice quality in CVSD based on the same DRT criterion requires a bit error rate of only 4 percent or less. The DRT is not designed for toll-quality voice test.

14.6 Practical Digital Cellular Systems

14.6.1 High-capacity FSK in FDMA

The use of frequency division multiple access (FDMA) with the present channel bandwidth of 30 kHz is a conventional approach. The high-capacity FSK modulation used in FDMA is based on the rate of digital voice. From the quality transmission of vocoders (see Sec. 14.5.1) we can see that a digital voice must have a 16-kbps coding rate to produce full telephone quality, although the voice quality of a 4.8-kbps LPC with vector quantization may be acceptable for a communications-quality voice (see Sec. 14.5). Now we have to determine the C/I which will provide optimal quality within a given digital system.

Swerup and Uddenfeldt compared a narrowband coherent digital modulation with gaussian MSK to an analog FM system.[11] Two 16-kbps voice coders were used. Residual excited linear predicted codes and subband codes were tested. The digital unit performance can be reduced by 5 dB to obtain the same performance as an analog unit. This 5-dB reduction advantage means a large coverage area and a closed frequency-reuse distance for each cell can be served in a cellular system. This is, in turn, an example of high spectral efficiency usage (described in Sec. 13.4). Consider the following calculations.

1. In an omnidirectional-cell system, assume that $C/I = 13$ dB, i.e.,

$$\frac{C}{I} = \frac{q^4}{6} > 10^{1.3} = 20$$

Solving for q and using Eq. (2.4-5), we obtain

$$q = 3.31 = \sqrt{3K}$$

$$K \approx 4 \qquad \text{(frequency-reuse pattern)}$$

In this case the total number of channels is 333; then

$$m = \frac{333}{4} = 83 \text{ channels/cell}$$

which is higher than the 47 channels per cell for $C/I \geq 18$ dB.

2. In 120° direction cells, we also compare two sets of C/I levels. As shown in Fig. 6.6*b*, there are only two interference cells. Then we can estimate the distance between the mobile unit and the two interfering sites to be $D + 0.5R$, as mentioned in Sec. 6.5.1.

   a. $C/I \geq 18$ dB

$$\frac{(q + 0.5)^4}{2} \geq 10^{1.8} = 63$$

   or $$q = 2.85$$

   The number of frequency-reuse cells is

$$K = \frac{q^2}{3} = 2.71 \sim 3$$

   The number of sectors is $(3 \times 3) = 9$; then

$$m = \frac{333}{9} = 37 \text{ channels/sector}$$

   b. $C/I \geq 13$ dB

$$\frac{(q + .05)^4}{2} = 10^{1.3} = 20$$

   or $q = 2$. The number of frequency-reuse cells is

$$K = \frac{q^2}{3} = \frac{4}{3} = 1.3 \sim 2$$

   The number of sectors $= (2 \times 3) = 6$; then

$$m = \frac{333}{6} = 55.5 \text{ channels/sector}$$

3. In 60° directional cells (see Fig. 6.6*c* and Sec. 6.5.1)
   a. $C/I \geq 18$ dB

$$\frac{(q + 0.7)^4}{1} \geq 10^{1.8} = 63$$

or

$$q = 2.12$$

$$K = \frac{q^2}{3} = 1.5 \sim 2$$

$$m = \frac{333}{2 \times 6} = 27.75 \text{ channels/sector}$$

   b. $C/I \geq 13$ dB

$$\frac{(q + 0.7)^4}{1} \geq 10^{1.3} = 20$$

or

$$q = 2.11 - 0.7 = 1.41$$

$$K = \frac{q^2}{3} = 0.67 \sim 1$$

$$m = \frac{333}{6} = 55.5 \text{ channels/sector}$$

Apparently, spectrum efficiency is increased by using digital technology.

Discussion. The preceding analysis is based on an ideal situation. It needs to be verified by measurement in a real cellular system. In the future we may achieve a digital system which can narrow the channel bandwidth and increase the transmission rate. The success of such a system would be proved if the reception at the receiving end were the same as that which would be achieved if the medium were nonfading. Modulation schemes, diversity, coding, redundancy, and ARQ can help to achieve this goal.

The signal can be designed using 4-, 8-, or 16-level MFSK or MPSK. Of course, the higher the number of levels, the narrower the channel bandwidth. However, the increase in transmitted power (or the reduction in range) is the key factor.

14.6.2 TDMA system[11]

The use of time-division multiple access (TDMA) could be another approach for increasing spectrum channel efficiency. It also has the potential to reduce the cost of both cell-site and mobile terminal equipment. The bulky analog radio equipment can be replaced by very large scale integrated-circuit (VLSI) digital signal processing. This applies to the use of digital equipment in any system; TDMA is no exception. In addition, TDMA will replace the analog duplex filter with a time switch. A zero IF receiver can be used, that is, there can be direct conversion without superheterodyne. The number of radio transceivers can be reduced, and the size of cell site can be much smaller. Assume that a cell site in a TDMA system has a single transceiver and a simple network interface. This is a very attractive feature. The main drawback of TDMA is the accurate clock requirement.

An inaccurate clock results in time jittering due to the instability of the frequency synthesizer at the transmitter and the clock at the receiver. The variable time delay resulting from the change of vehicle positions and from random FM could affect the synchronization and tracking of the bit streams. The most adverse effect of the synchronization problem is the delay spread mentioned in Chap. 1. In TDMA transmission, severe time dispersion occurs. In urban areas of the United States, the mean delay spread can be as much as 3 μs. Then the transmission rate must be limited to

$$R_b \leq \frac{1}{2\pi \times 3 \times 10^{-6}} = 50 \text{ kHz}$$

By using diversity schemes,[12] the transmission rate can be increased if the same BER is assumed. Another effective scheme is an adaptive decision feedback equalizer[13] which adaptively sums up the multipaths so that the time-delay spread at the receiver is reduced and the sum of multipaths is in the form of a diversity.

The adaptive equalizer can work for narrowband TDMA also, so TDMA is not based on wideband spread spectrum technology. Since TDMA is very immune to interference in the cellular system (one user is designated in one time slot) and allows a simple handoff procedure, it may be suitable for microcell systems. *Microcells* are loosely defined as cells whose radii are less than 1 mi. However in a heavy traffic area, TDMA performance may degrade faster than FDMA performance. Simulation can be used to decide which transmission method is best.

For economic and spatial reasons, a 300-kHz-bandwidth TDMA carrying 300 kbps can provide 10 voice channels. Since a low-cost base station with one radio transceiver can make the TDMA system seem

very attractive, this possibility is worth pursuing. The critical issue is building the adaptive decision feedback equalizer properly.

Since the same digital system working in different areas will result in different levels of performance, we should be aware that the same system might test well in a light traffic area but poorly in a heavy traffic area.

14.6.3 Spread spectrum

The uses of a frequency hopping scheme in FDMA and a direct sequence scheme in TDMA are described as follows.

Frequency division multiple access (FDMA). There are many spread-spectrum systems proposed for use in mobile radios. Cooper and Nettleton proposed[14] a frequency hopping system. In each hop a binary DPSK-modulated system with error-correction coding is implemented. Comparing the relative spectral efficiencies of this proposed system with today's cellular 30-kHz system, we see that they are almost the same. Goodman et al.[15] proposed using multilevel frequency shift keying (MFSK) plus frequency hopping for land mobile radios. Their theoretical results, based on certain assumptions, show a better spectral efficiency.

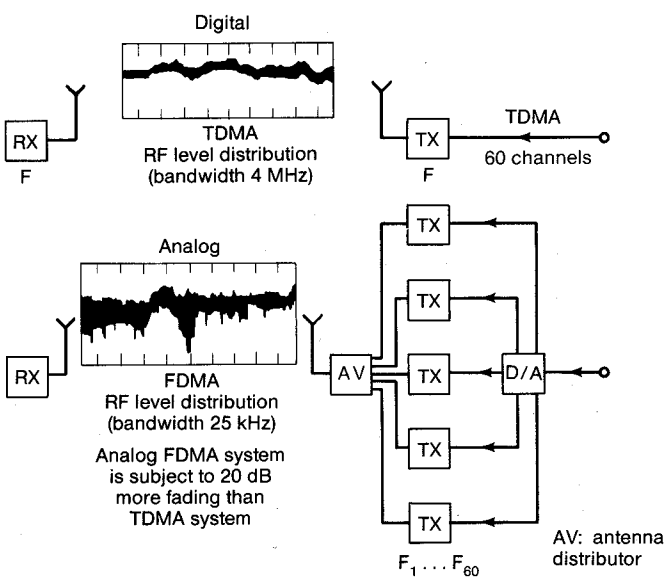

Figure 14.18 Digital TDMA mobile radio and analog FDMA system. (*After Bohm, Ref. 16. ©Horizon House—Microwave, Inc.*)

Time-division multiple access (TDMA). A digital TDMA system has been proposed by SEL in Germany,[16] which uses a bandwidth of 4 MHz and only one fixed radio frequency for all subscribers in each cell. Therefore it is a spread-spectrum system. In a wideband system, each channel does not need to tune to a particular frequency. All the channels share the same bandwidth and thus have the same type of radio set.

Method of operation. The signal of each channel in TDMA is coded differently to provide identification and proper channel separation. The difference between a digital TDMA system and an analog FDMA system is shown in Fig. 14.18. In a digital TDMA system, each channel has 32 code patterns (5 bits) and one sign bit; thus, a total 6 bits are transmitted.

In each cell, 60 TDMA channels are available in three channel groups. The merit of this system is shown in Fig. 14.19. Each base station covers three cell sectors in succession by means of an electronic scan-and-stay antenna pattern. Twenty channels are allocated to each sector. Once a cell sector has received its 20 channels (time slots), the antenna pattern of the base station is switched over to the next sector. The 20 channels per sector is an arbitrary number. Any number of channels can be used among the total 60 channels.

The synchronization and the clock accuracy have to be considered. This TDMA scheme does simplify the base station, and there is hopefully less channel interference in this system. However, there is a high risk of developing a spread spectrum system.

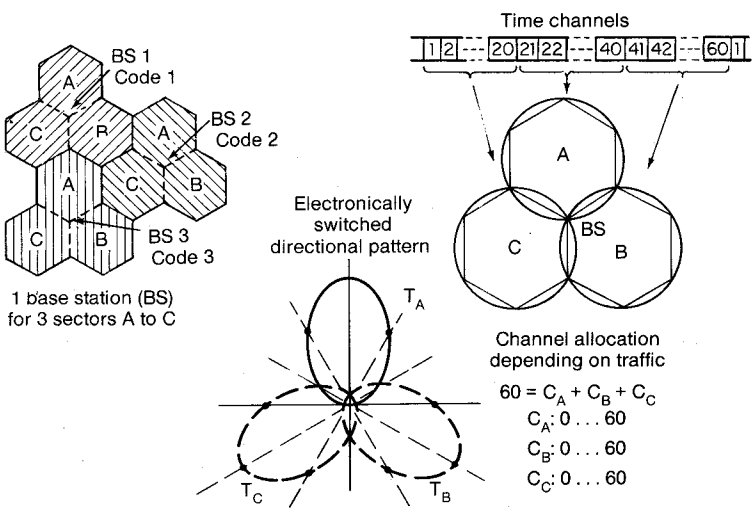

Figure 14.19 Cell structure of wideband TDMA systems. (*After Ref. 16. ©Horizon House—Microwave, Inc.*).

14.6.4 Evaluation of a digital cellular system

The performance of a digital cellular system should be evaluated by a subjective test. The required C/I is deduced from the subjective test for a given performance. Also the channel bandwidth of a FDMA or the equivalent channel bandwidth of a TDMA should be included in the evaluation of the system's performance. Therefore a nationwide standard subjective test should be established to accomplish this task.

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Chapter

15

Cellular Related Topics

15.1 Study of a 60-GHz Cellular System

A series of studies[1,2] have explored the implementation of a 60-GHz mobile telephone system using direct line-of-sight transmission along urban streets. Use of the 60-GHz band is encouraged for the following reasons.

1. The 60-GHz band is in the oxygen-absorption range. In this range the attenuation of radio signals in the oxygen-absorption band is two orders of magnitude greater than the attenuation outside the oxygen-absorption band over a 1-km distance. Therefore, the 60-GHz band cannot be used by many applications; however, if it is properly implemented, it can be allocated to a cellular mobile radio system. The FCC may be glad to give this band away.
2. The characteristics of this high-attenuation signal over the propagation path create a natural barrier to cochannel or adjacent-channel interference in the cellular mobile system.

15.1.1 Propagation in the scattered environment

In the UHF range or at X band, multipath signal scattering from vehicles and buildings in the mobile radio environment has frequently

caused deep and rapid fading of the received signals. Sometimes, the signal can be received by propagation through scattered or reflected signals, a natural phenomenon. However, when a 60-GHz band is used, line-of-sight propagation is required. A series of experiments at 59.5 GHz were carried out in 1972 in urban areas.[2]

The average output power at the mobile antenna was 40 mW (16 dBm), and the parabola antenna beamwidths were 3° at both the base station and mobile ends. The data were collected on three streets in Red Bank, New Jersey. The streets were chosen because they were representative of the urban mobile environment and were at least 0.5 mi long.

15.1.2 Fixed terminals

The mobile transmitter and the mobile receiver were parked about 0.56 km (0.35 mi) apart on opposite sides of a street. The results for each street are plotted in Fig. 15.1. Fades exceeding 3 dB occurred 5 percent of the time on Monmouth Street, 20 percent of the time on Bridge Street, and 34 percent of the time on Broad Street. The measuring limit was at −52 dBm, which presents a signal fade of 15 dB. Fades exceeding this limit occurred 1 percent of the time on Bridge Street, 2 percent of the time on Monmouth Street, and 3 percent of the time on Broad Street. Fades exceeding 15 dB were caused by large trucks or buses completely blocking the path within 30 m (100 ft) of the transmitter or receiver.

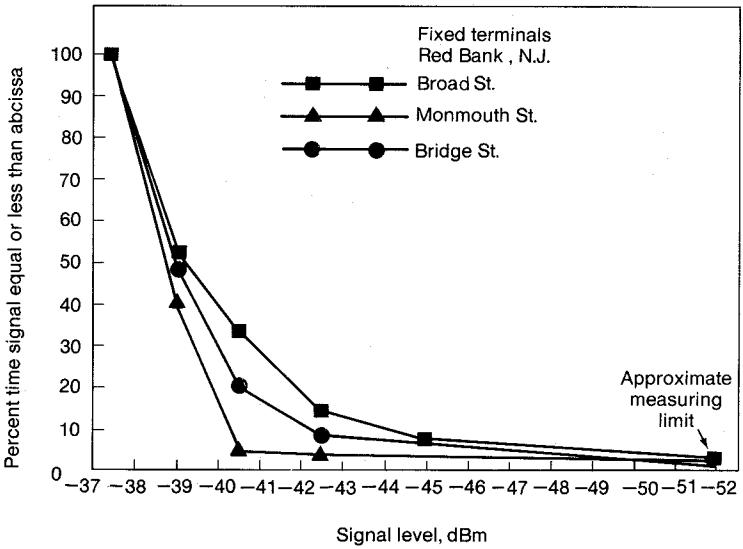

Figure 15.1 Fading statistics from fixed terminals, Red Bank, N.J. (*From Ref. 2.*)

15.1.3 Moving terminal

The mobile receiver was parked at the side of the street, and the mobile transmitter started at least 0.8 km (0.5 mi) from the receiver and proceeded toward and past the receiver at a constant speed of 24 km/h (15 mi/h). The results are plotted in Fig. 15.2. The fading statistics of all three streets are roughly the same. The fades exceeded 3 dB about 70 percent of the time. The fades exceeding the −52-dBm measuring limit (15 dB fades) occurred 9 percent of the time on Broad and Monmouth Streets but only 5 percent on Bridge Street because the line-of-sight signal was blocked by large trucks or buses. The Doppler frequencies ranged from a few hertz to 170 Hz. When vehicle speed was 15 mi/h the fading rate was 25 Hz. Fades exceeding 10 dB occurred 20 percent of the time. A typical run measured on Monmouth Street is shown in Fig. 15.3.

The experiments at 60 GHz reveal that the amplitude of the scattered signal was small in comparison to the direct signal. The effect of the scattered signal is small. The fading observed was neither deep nor rapid.

15.1.4 System consideration

From the preceding data, we may note the following problems in a 60-GHz mobile radio system: (1) the excessive propagation loss, (2) the pointing mechanism of the antennas, (3) the small separation between

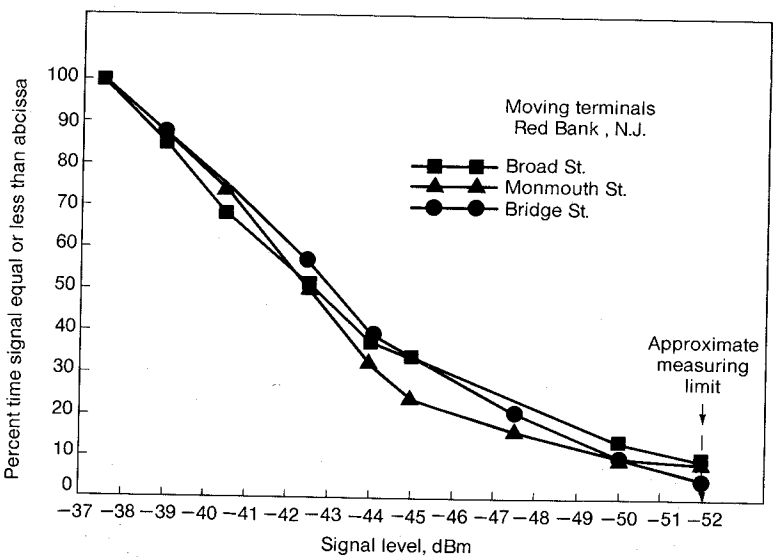

Figure 15.2 Fading statistics from moving terminals, Red Bank, N.J. (*From Ref. 2.*)

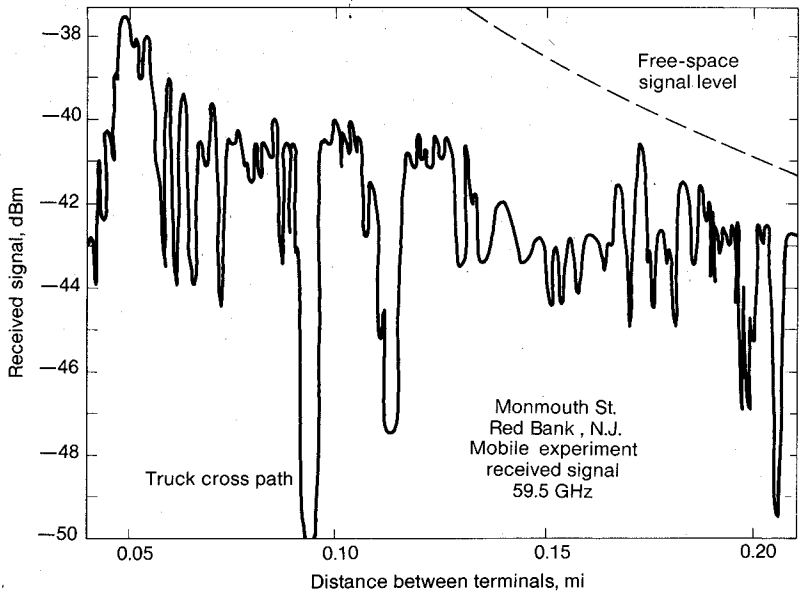

Figure 15.3 Received signal at Monmouth Street, Red Bank, N.J. (*From Ref. 2.*)

repeaters, and (4) the ways of avoiding line-of-sight blockage from large trucks and buses.

15.2 Cellular Fixed Stations

15.2.1 Radios for fixed-station telecommunications

Many alternatives can be used in the installation of radios in fixed-station telecommunications.[3] The cellular fixed station, one of these alternatives, is installed according to certain specifications.

1. To ensure a sufficient number of radio channels, each radio channel is assigned to a customer on a permanent basis.
2. When the number of available (unoccupied) radio channels is limited, two approaches can be used. The trunking approach is used to scan for an available channel and a set-up channel approach is used to initiate a call, and the call is then transferred to and carried on an assigned voice channel.
3. For cellular fixed radios, the limited radio channels should be reused. Entire control functions are the same for fixed cellular systems as for other cellular mobile systems.

15.2.2 Two approaches for fixed cellular systems

The fixed cellular concept is not a new one. The implementation of 800-MHz technology has gradually reduced the cost of cellular components, and the fixed cellular system could provide an economic means of providing telephone coverage in a rural area.

There are two approaches to implementation of fixed cellular systems.

1. Use cellular mobile equipment for the fixed cellular stations. It may save the cost of manufacturing a new fixed cellular radio. No new spectrum would be needed.
2. Design a cellular fixed radio.
   a. Operation of this radio could be integrated with local telephone company switching offices (see Fig. 15.4*a*). However, a new spectrum would be needed.
   b. Operation of this radio could be integrated with cellular mobile switching; in such a case, no new spectrum would be needed.

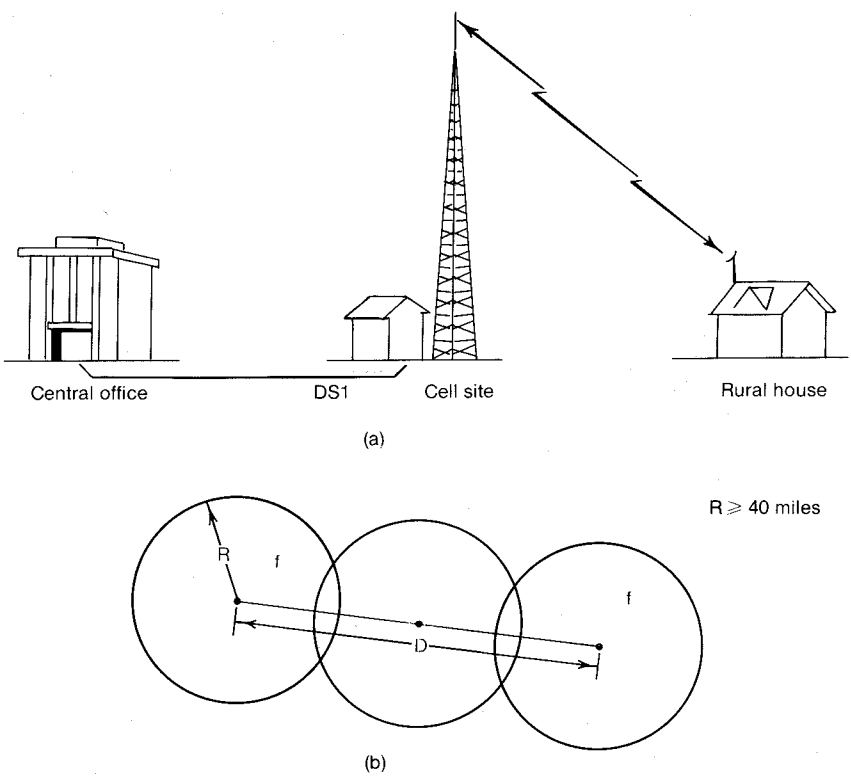

Figure 15.4 Fixed cellular system.

15.2.3 Special features of fixed cellular systems

There are several advantages.

1. The cost of the radio is low because the logic unit is simpler.
2. Either regular switching or MTSO switching can be used.
3. No handoff is needed.
4. Each link can be carefully engineered.
5. A cell with a large area can be served.
   a. Both antenna heights at two ends are high, and most links are line-of-sight links. The propagation loss approaches the free-space loss.
   b. There is no multipath fading. The required C/I for fixed cellular systems is $C/I \geqslant 10$ dB.
   c. The surroundings are quiet.
   d. Directional antennas can be used at the customer site.
6. A low-cost controller can be provided at either the cell site or the telephone company relay site.
7. The fixed cellular system will expect a longer holding time, and no calls will be dropped.
8. The busy periods (rush hours) in fixed cellular systems usage are different from those in mobile cellular system usage.
9. It is possible to use radio channels for data links between cell site and MTSO.

15.2.4 Design of a cellular fixed system

Because fixed cellular stations cover larger areas, the system can cover a 40-mi-radius cell ($R = 40$ mi) or larger. The frequencies used in the fixed cellular system can be identified as a subset $\{f_i\}$ of the total cellular mobile frequencies, say, 80 channels. The 80 channels need to be divided into only two groups, each with 40 channels. Each cell is assigned one group, and the other group is assigned to an adjacent cell. The separation between cochannel cells is $3R$ (see Fig. 15.1*b*). The reason for a closed cochannel cell separation is the natural isolation caused by the radio horizon phenomenon. The maximum area of each cell is

$$A = \pi(40)^2 = 5026 \text{ mi}^2$$

With 40 channels per cell, we can serve roughly 800 customers. If the area has to serve more customers, the size of the cells can be decreased.

15.3 Cellular Systems in Rural Service Areas[4]

Cellular mobile systems in the United States, roughly 733 market areas, have been designated as *cellular geographic service areas* (CGSA).

Each CGSA can have block A systems and block B systems. The largest 305 cities of the 733 markets are called *metropolitan statistical areas* (MSAs) and the rest (428) are called *rural service areas* (RSAs). RSAs are usually adjacent to MSAs.

How to achieve a flexible and cost-effective cellular system design that will provide adequate RSA coverage but that will not interfere with MSA coverage is a challenging problem. In fact, there are three common obstacles.

1. RSA operators generally do not want to limit their design parameters.
2. MSA operators are afraid of the interference that might result from RSA service.
3. Any cell-site distance-separation specification might be adequate for one system (RSA or MSA) but not for the other.

Two approaches are recommended for controlling cochannel interference and adjacent-channel interference between RSA and MSA systems.

Approach 1—create a buffer zone between the MSA and the RSA. The RSA will be mandated to limit its transmitted power and the antenna height at the RSA cell site so that they are compatible with the MSA specifications. (See Fig. 15.5.)

Approach 2—let the RSA system designer know both the MSA boundary and the carrier-to-interference ratio at the boundary before planning the system. This way the RSA system designer will know the received signal level at the RSA boundary adjacent to the MSA.

Approach 2 may be more suitable in real situations. Its methodology is as follows.

1. The MSA provides the boundary signal level requirement (level A) to the RSA (see Fig. 15.5).
2. The RSA system designer considers and designs the RSA system in accordance with MSA compatibility requirements.
3. The RSA can implement a signal level of -110 dBm at the boundary of the CGSA if there is no interference with the MSA.
4. The MSA and RSA system designers can consider what specifications will be compatible for both parties.

At the cell site, the RSA operator can install any kind of directional antenna and raise the antenna to maximize the transmitted power, provided the above requirements are satisfied. The directional antenna can also be used to reduce the interference to or from the other site.

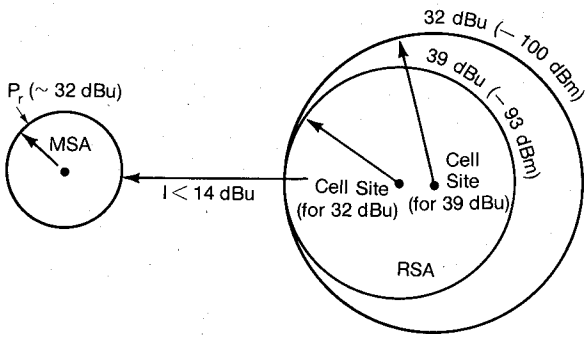

Figure 15.5 Received signal-strength power of an RSA.

There are a few additional considerations.

1. A low-cost antenna mast or an existing mast should be used.
2. Half of the unused voice channels could be used for microwave transmission.
3. The installation of small switching systems should be considered, or these systems could be shared with neighboring MSAs or RSAs.

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Dedicated to my mother

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