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. *Choosing 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 95 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 assignemnt. 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

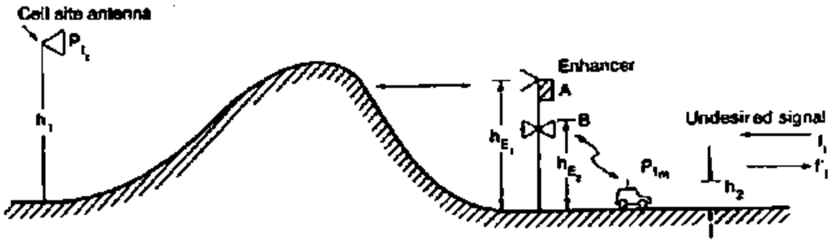


Figure 10.1 Enhancer.

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

$$P_{R_m} = P_{t_c} + g_c - L_a + (G + g_{E_1} + g_{E_2}) - L_b + g_m \qquad (10.1\text{-}9)$$

and

$$P_{R_c} = P_{t_m} + g_m - L_b + (G + g_{E_1} + g_{E_2}) - L_a + g_c \qquad (10.1\text{-}10)$$

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
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

$$d_1 > \frac{2A_T}{\lambda} + \frac{2A_1}{\lambda} \qquad \text{and} \qquad d_2 > \frac{2A_1}{\lambda} + \frac{2A_R}{\lambda} \qquad (10.2\text{-}1)$$

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

$$d_1 > \frac{2L_T^2}{\lambda} + \frac{2A_1}{\lambda} \qquad \text{and} \qquad d_2 > \frac{2A_1}{\lambda} + \frac{2L_R^2}{\lambda} \qquad (10.2\text{-}2)$$

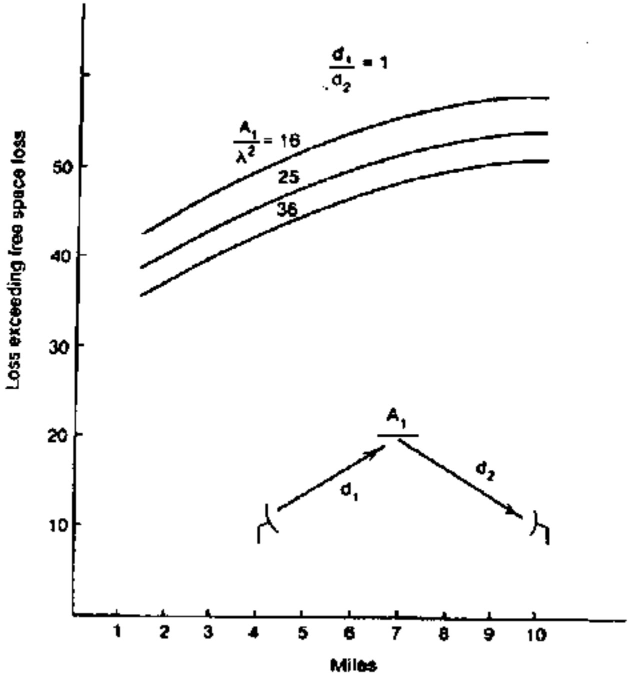


Figure 10.2 Effective use of reflectors $d_1/d_2 = 1$.

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 that 100 percent of the incident power is reflected; then

$$P_R = P_T \frac{A_T A_R A_1^2}{\lambda^4 d_1^2 d_2^2} \qquad (10.2\text{-}3)$$

where P_T, P_R = transmitted and received power, respectively, and

$$A_T = \frac{G_T \lambda^2}{4\pi} \qquad (10.2\text{-}4)$$

$$A_R = \frac{G_R \lambda^2}{4\pi} \qquad (10.2\text{-}5)$$

Then Eq. (10.2-3) becomes

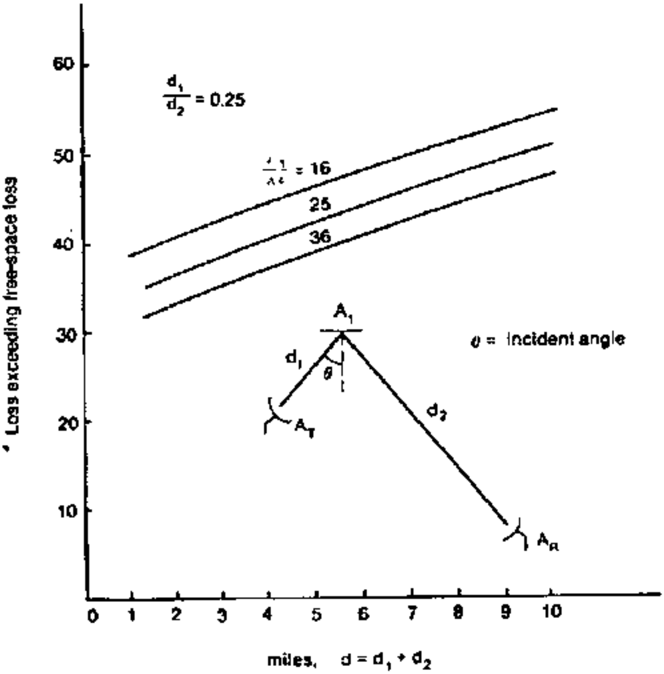


Figure 10.3 Effective use of reflectors $d_1/d_2 = 0.25$.

$$P_R = P_T G_T G_R \frac{A_1^2}{(4\pi)^2 d_1^2 d_2^2}$$

$$= P_T G_T G_R \underbrace{\left[\left(4\pi\,\frac{d}{\lambda}\right)^2\right]^{-1}}_{\text{free-space loss (FSL)}} \cdot \underbrace{\left(\frac{d^2(A_1/\lambda^2)^2}{d_1^2 d_2^2}\right)}_{\text{excessive loss}} \qquad (10.2\text{-}6)$$

where $d = d_1 + d_2$ and

$$P_R = 10\ \log(\text{FSL}) + 10\ \log\left[\left(\frac{d^2\lambda^2}{d_1^2 d_2^2}\right)\left(\frac{A_1}{\lambda^2}\right)^2\right] \qquad (10.2\text{-}7)$$

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

$$P_R = P_T \frac{A_T A_R A_1^2}{\lambda^2 d_1^2 d_2^4}$$

$$= P_T G_T G_R \left[\left(4\pi \frac{d_1}{\lambda} \right)^2 \frac{d_2^4}{\lambda^4} \right]^{-1} \cdot \frac{(A_1/\lambda^2)^2}{1} \qquad (10.2\text{-}8)$$

or

$$P_R = 10 \log(\text{FSL}) + 10 \log \left[\frac{d^2 A_1^2 \lambda^2}{d_1^2 d_2^4} \right] \qquad (10.2\text{-}9)$$

excessive loss

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 sep-

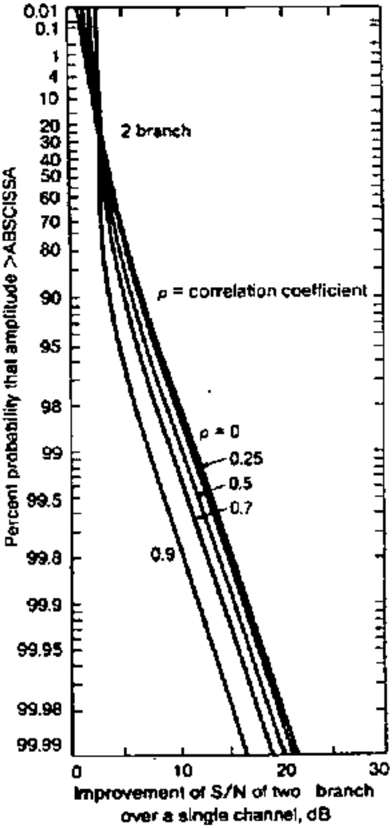
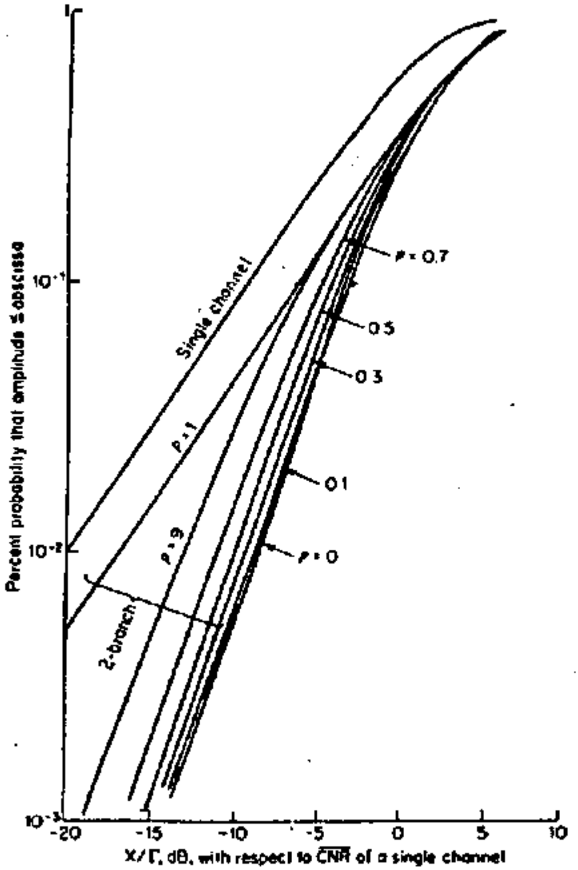


Figure 10.4 Improvement of signal-to-noise ratio of a two-branch signal over a single channel signal. (*After Lee, Ref. 3.*)

aration 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 $\rho = 0$.

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

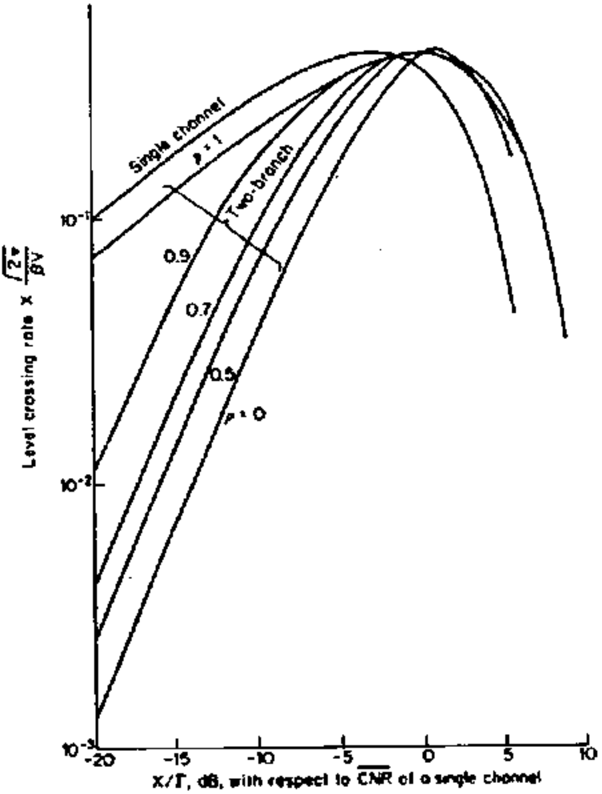


(a)

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.*)

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 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.



(b)

Figure 10.5 *(Continued)*

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.

$$\lambda_g = \frac{V_g}{f} \tag{10.3-1}$$

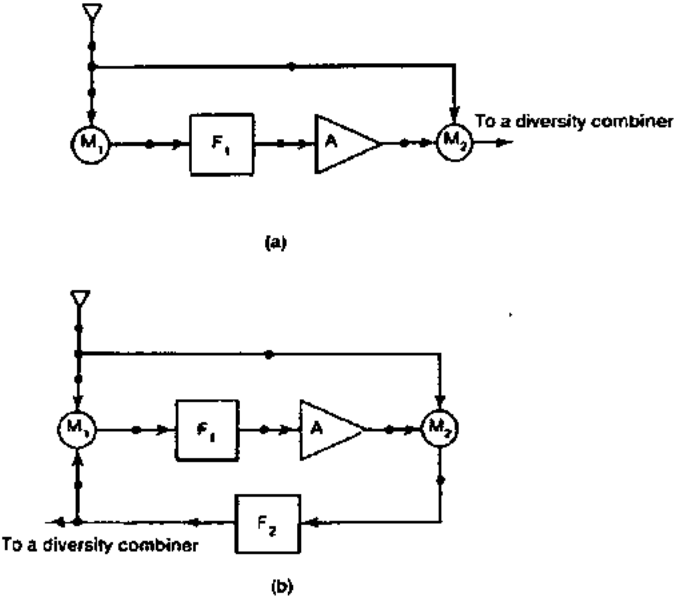


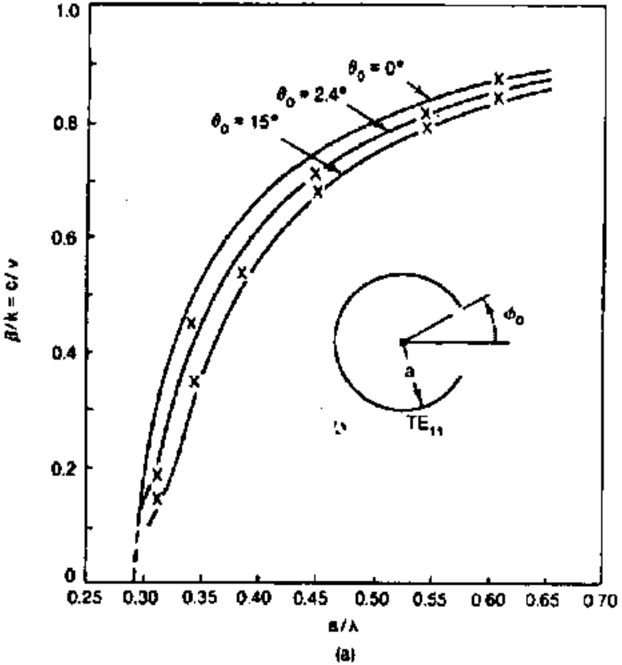
Figure 10.6 Two cophase techniques, feedforward and feedback (F = the filter, M = the mixer, and A = the limiting amplifier, as shown in the figure). (*a*) Feedforward combiner; (*b*) feedback (Granlund) combiner.

Therefore,

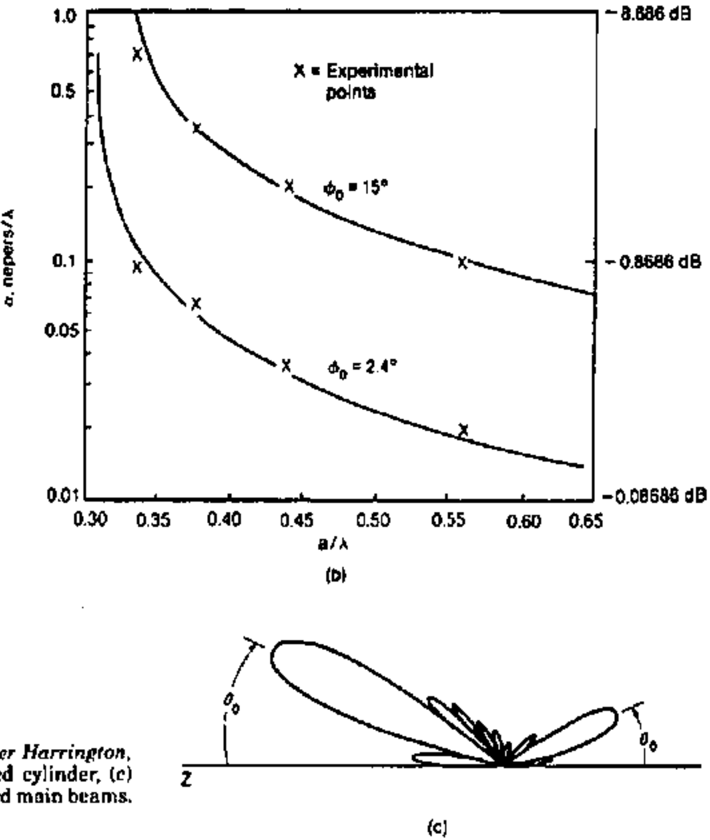
$$\lambda_g > \lambda_c \tag{10.3-2}$$

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.

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.



(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 backward main beams.

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

$$V = \frac{\omega}{\beta} = \frac{1}{\sqrt{LC}} \qquad (10.3\text{-}3)$$

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 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 ex-

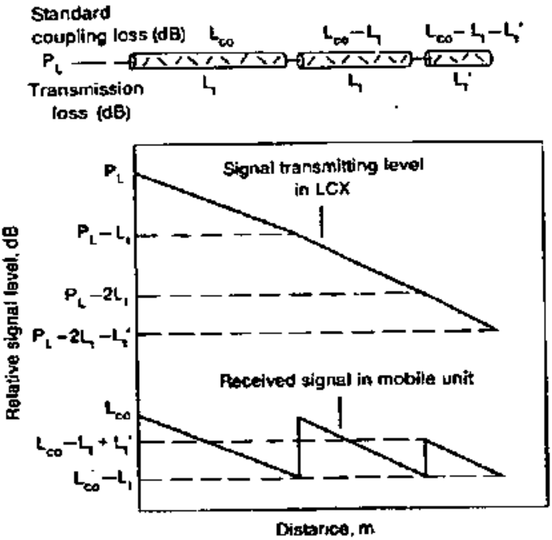


Figure 10.8 Grading technique.

ceed 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 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.

$$L\ (s \text{ at } d) = (\text{coupling loss at 1.5 m}) + 10 \log \left(\frac{d}{1.5}\right) \qquad (10.3\text{-}4)$$

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.

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.

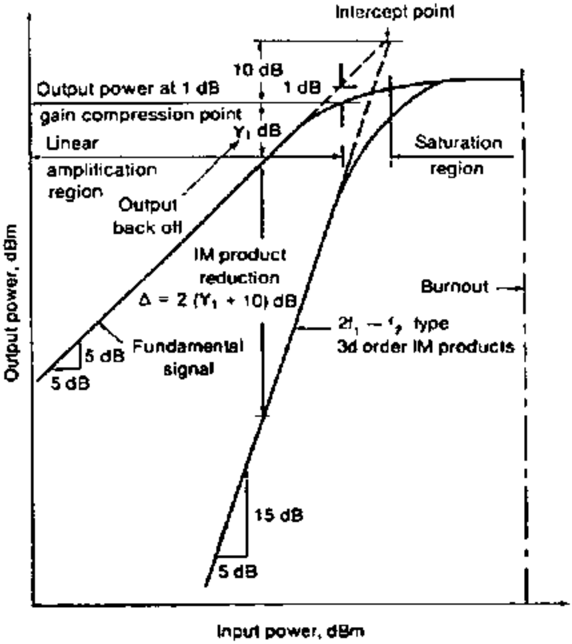


Figure 10.9 Input-output characteristics of a linear amplifier (*After Suzuki et al., Ref. 12*).

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).

$$P_r = \alpha P_{t_0} R_0^{-\gamma} \qquad (10.4\text{-}1)$$

$$P_r = \alpha P_{t_2} \left(\frac{R_1}{2}\right)^{-\gamma} \qquad (10.4\text{-}2)$$

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

$$P_{t_1} = P_{t_0}\left(\frac{1}{2}\right)^{-\gamma} \qquad (10.4\text{-}3)$$

For a typical mobile radio environment, $\gamma = 4$, Eq. (10.4-3) becomes

$$P_{t_1} = \frac{P_{t_0}}{16} \qquad (10.4\text{-}4)$$

or

$$P_{t_1} = P_{t_0} - 12 \qquad \text{dB} \qquad (10.4\text{-}5)$$

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 $R_n = R_0/2^n$.

$$P_{t_n} = P_{t_0} - n(12) \qquad \text{dB} \qquad (10.4\text{-}6)$$

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

$$\text{New traffic load} = (4)^n \times (\text{the traffic load of start-up cell}) \qquad (10.4\text{-}7)$$

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

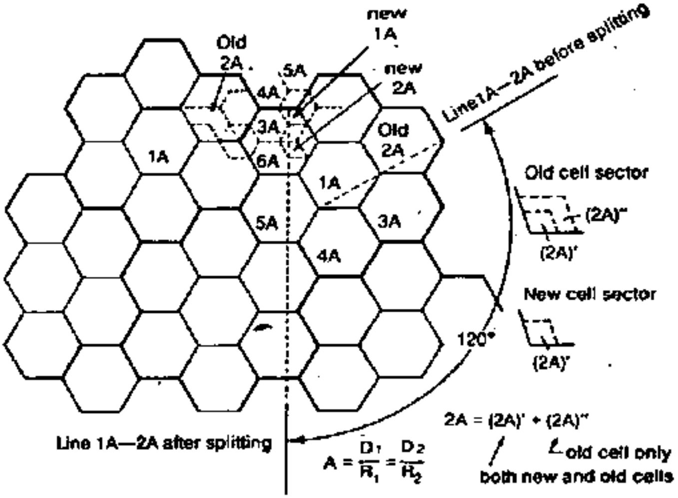


Figure 10.10 Cell-splitting techniques.

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 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.

$$2A = (2A)' + (2A)'' \qquad (10.4\text{-}8)$$

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. In this section, we try to find the means of control, the radiation power. The intelligent cell concept and application to microcells are described in Chap. 16.

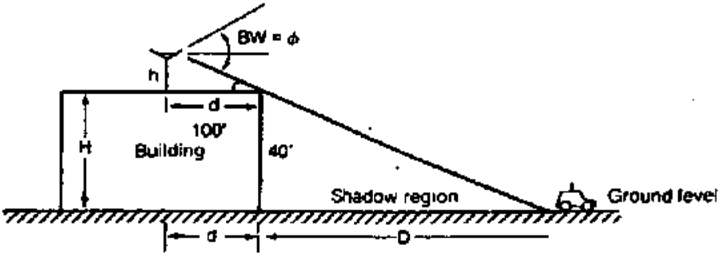


Figure 10.11 Rooftop-mounted antennas.

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

$$h = d \tan\left(\frac{\phi}{2}\right) \tag{10.5-1}$$

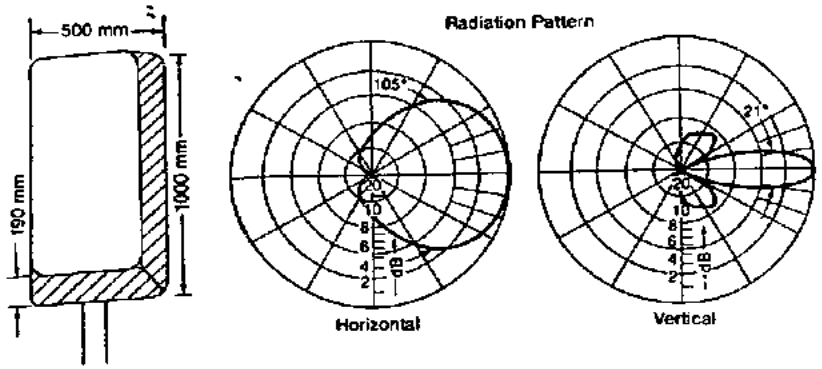
If a 6-dB-gain antenna has a vertical beamwidth of about 28° and the distance from the antenna post to the roof edge is 31 m (100 ft), then the required antenna height is 7.5 m (25 ft). The shaded region around the building depends on the height of the building.

$$D = \frac{H + h}{\tan(\phi/2)} - d \tag{10.5-2}$$

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).

The shadow region is calculated for a single building only. If there are adjacent buildings, multipath scattered waves are generated, and the shadow region is reduced.

Use of the antenna structures. The panel-type antennas[16] are ideal antenna structures for hanging on each side of the wall. For an omnidirectional configuration, the four-panel antennas mounted on the four sides of the building can be combined as in an omnidirectional antenna.



Panel-type antennas

Omnidirectional configuration (top view)

3-sector configuration (top view)

Figure 10.12 Panel-type antennas and their applications (*After Kathrein, Ref. 16*).

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 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, Loran-C transmitters are used by the U.S. Coast Guard. These transmitters operate at 500 kHz. A Loran-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 dimensons, 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 triangulation method to find the location of the mobile unit. The degree of accuracy is very poor because of the multipath phenomenon.

Trangulation 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.

$$M(N) = \frac{\sum_{i=1}^{N} x_i}{N}$$

$$M(N+1) = \frac{\sum_{i=2}^{N+1} x_i}{N}$$

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 afford to 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 antenna 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 con-

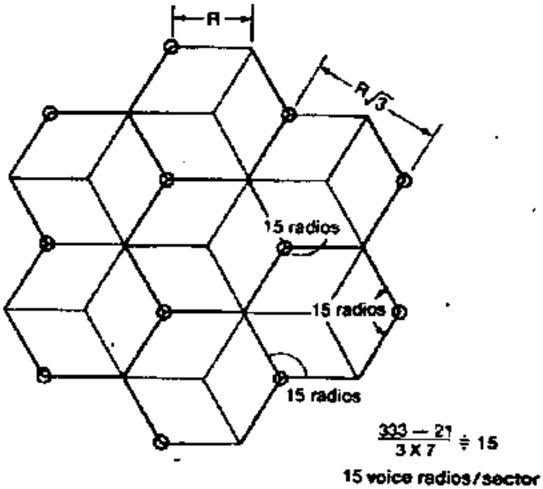


Figure 10.13 Ideally located cell sites over a flat terrain ($K = 7$).

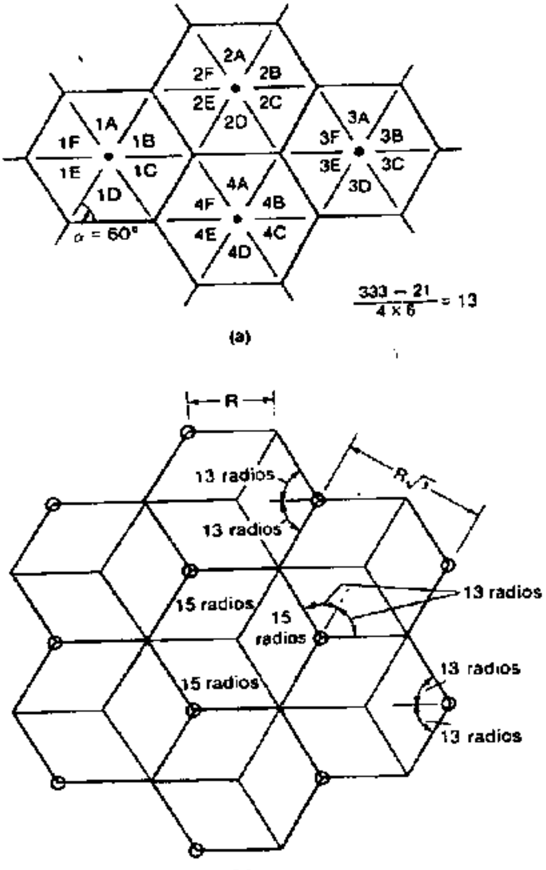
tain approximately 15 voice channels, a number which is derived from the total 312 voice channels

$$\frac{333 - 21}{3 \times 7} = 15 \text{ channels per } 120^\circ \text{ sector}$$

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



(a)

(b)

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).

$K = 7$ 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 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 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 effectively. 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 using an open-area curve instead of the suburban-area curve shown in Fig. 4.3.

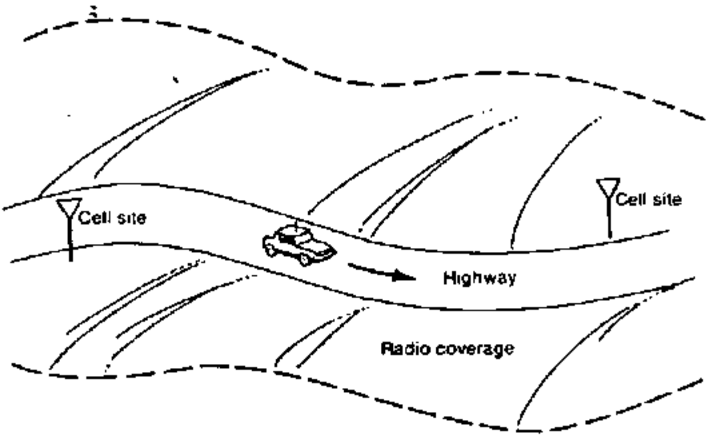


Figure 10.15 Highway cell sites.

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

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

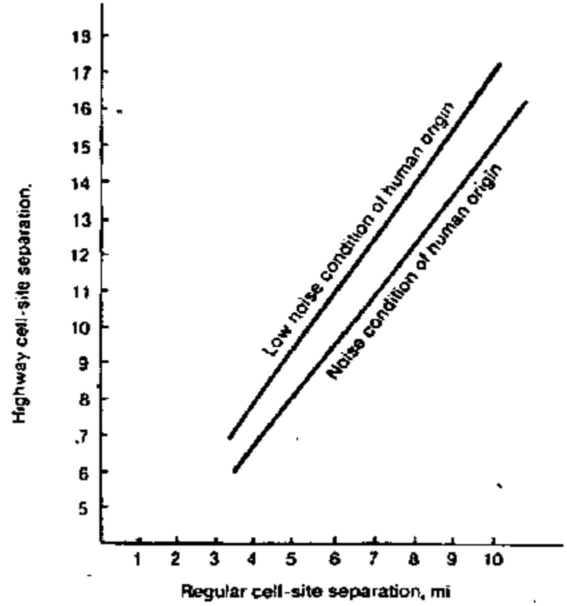


Figure 10.16 Highway cell-site separation.

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.

$$\Delta d = d[10\Delta G/20 - 1] \qquad (10.7\text{-}1)$$

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 subcriber 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.

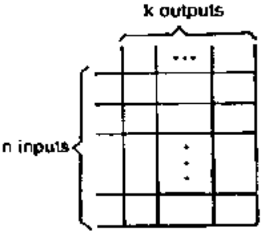


Figure 11.1 Number of crosspoints in space-division switches.

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 is nonblocking.

$$\text{Number of cross-points} = nk \tag{11.1-1}$$

These switches are laid out in space as shown in Fig. 11.1.

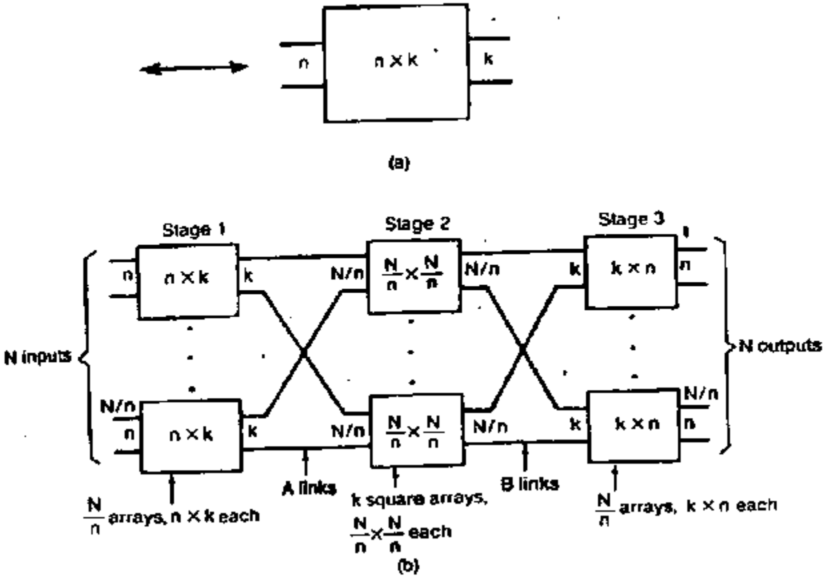
For simplicity in analyzing the following nonblocking system, let the total number of input lines N equal the total number of output lines. An N-input group is fed into N/n switch arrays, and each switching network has an nk switch as shown in Fig. 11.2*a*. The three-stage switching network (S-S-S) is shown in Fig. 11.2*b* 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 supported is



(a)

(b)

Figure 11.2 A switching network. (*a*) An *nK* switch; (*b*) a three-stage switching network.

$$N = \frac{t_f}{2\,t_c} \tag{11.1-3}$$

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

$$t_d = \frac{100 \text{ ms} \cdot}{2\,(1 - 0.465)} = 93.46 \text{ ms}$$

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

$$P_{cd} = \frac{\text{number of call drops before completion}}{\text{total number of accepted calls handled by set-up channels}}$$

$$= \frac{\sum_{i=1}^{5} C_{d_i}}{C_t}$$

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 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 in-

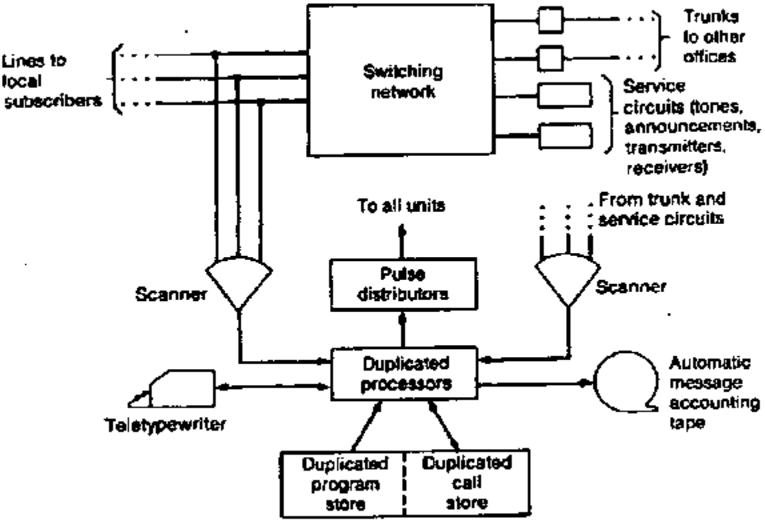


Figure 11.3 A typical analog switching system. (*After Chadha et al., Ref. 6.*)

terconnecting 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 algorithm, billing (air time and wire line), and diagnosis (radio, switching, and other hardware failure).

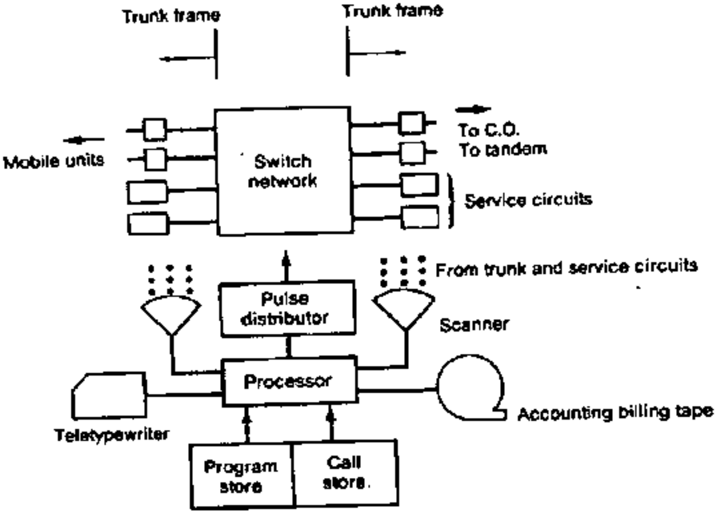


Figure 11.4 A modified analog switching system for cellular mobile systems.

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

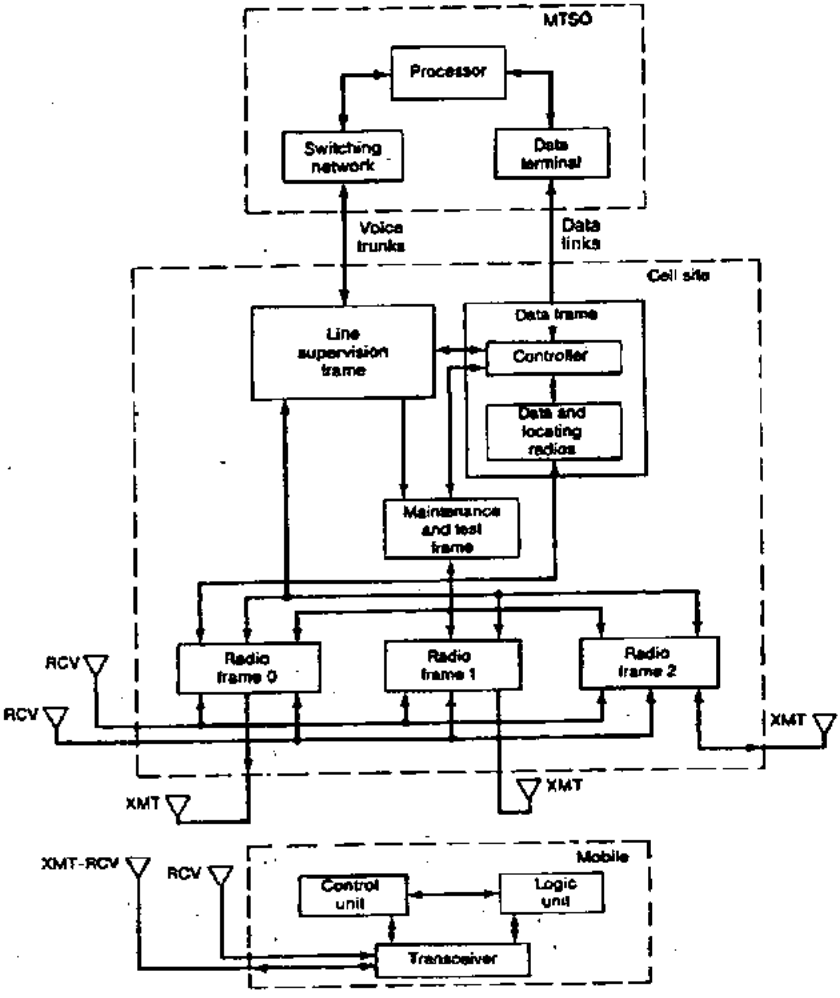


Figure 11.5 Cellular mobile major systems.

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. 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.

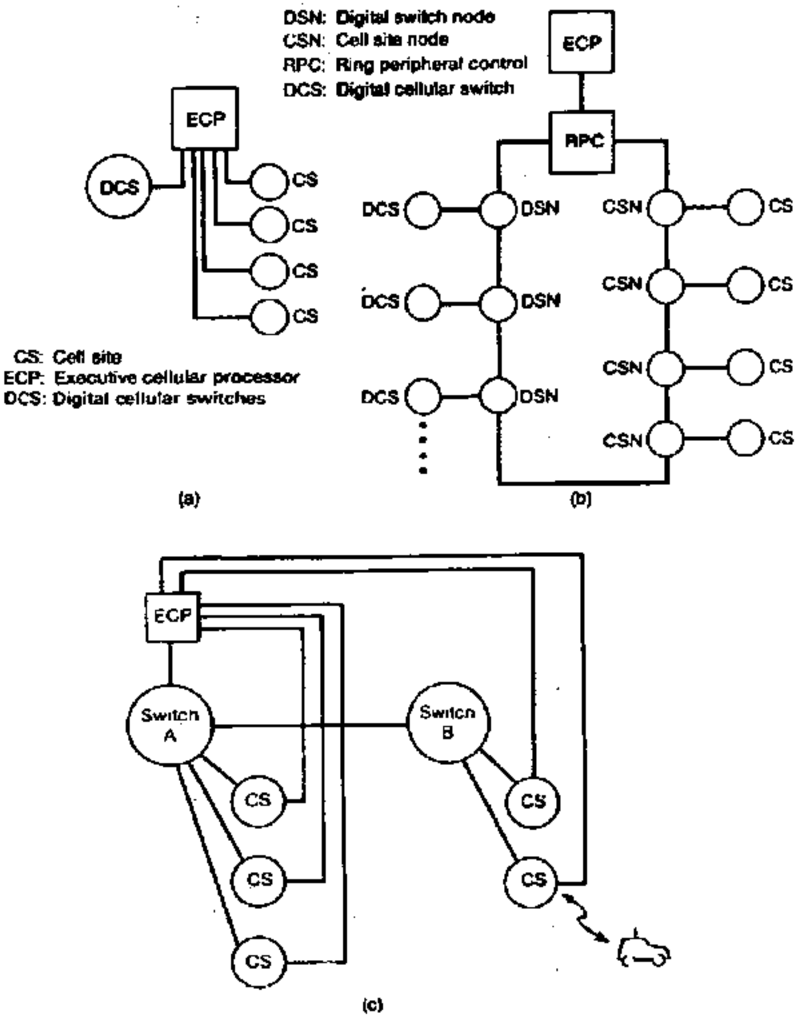


Figure 11.6 Cellular switching equipment. (*a*) A centralized system; (*b*) a decentralized system; (*c*) remote-controlled switching.

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 transmitted 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 received 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.

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 prob-

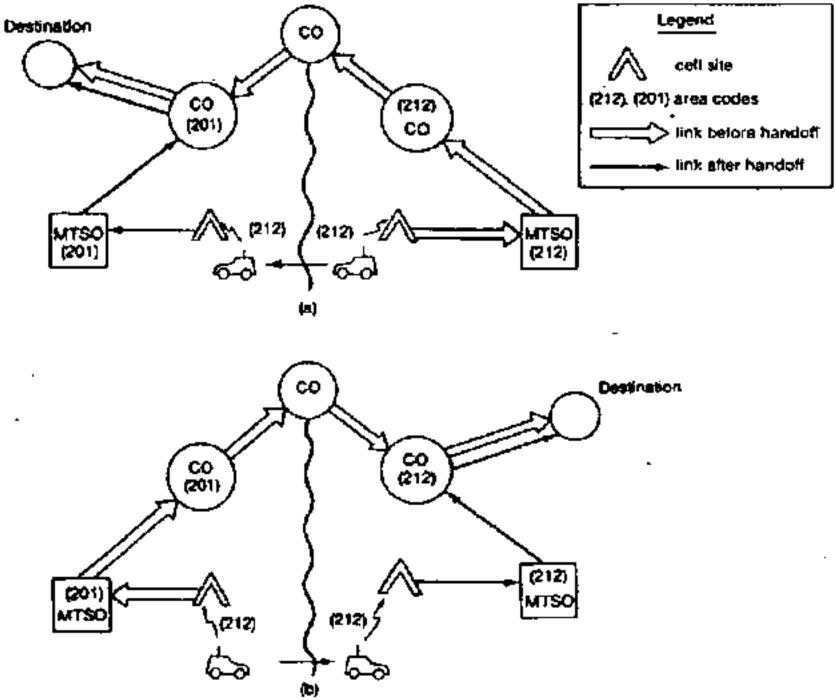


Figure 11.7 Four conditions of intersystem handoffs. (*a*) A toll call becomes a local call and the home mobile unit becomes a roamer; (*b*) a toll call becomes a local call and a roamer becomes a home mobile unit; (*c*) a local call becomes a toll call and the home mobile unit becomes a roamer; (*d*) a local call becomes a toll call and a roamer becomes a home mobile unit.

ability. The improvement in call origination and handoffs as a result 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).

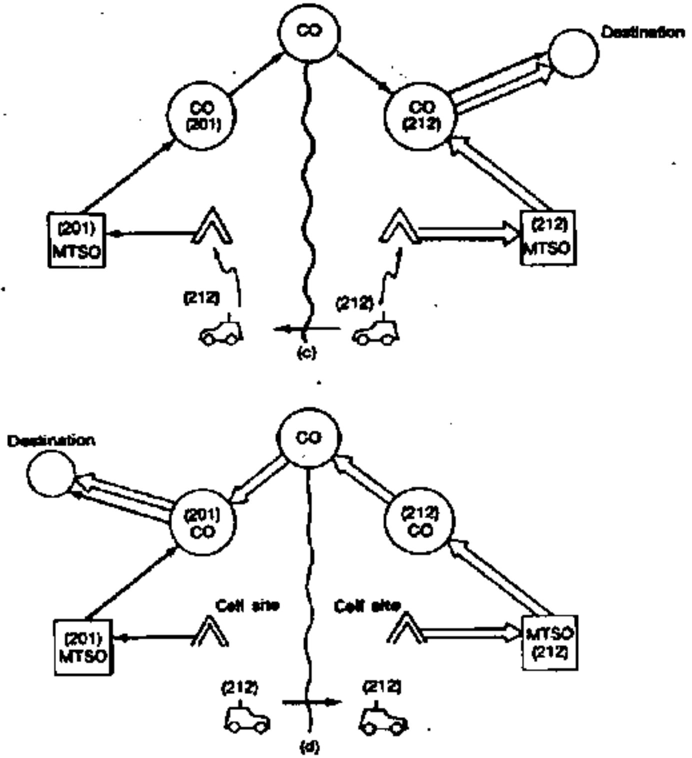
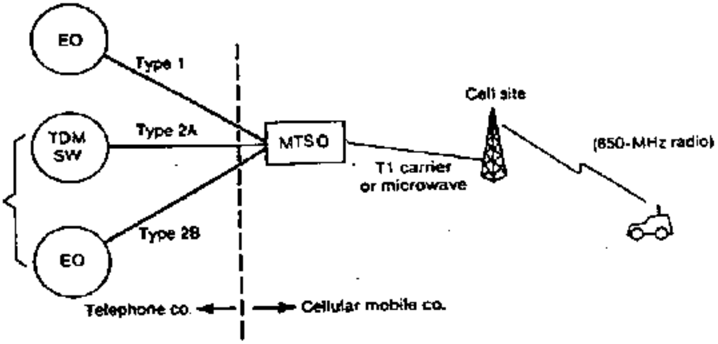


Figure 11.7 *(Continued)*

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 transmitting 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).



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.

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

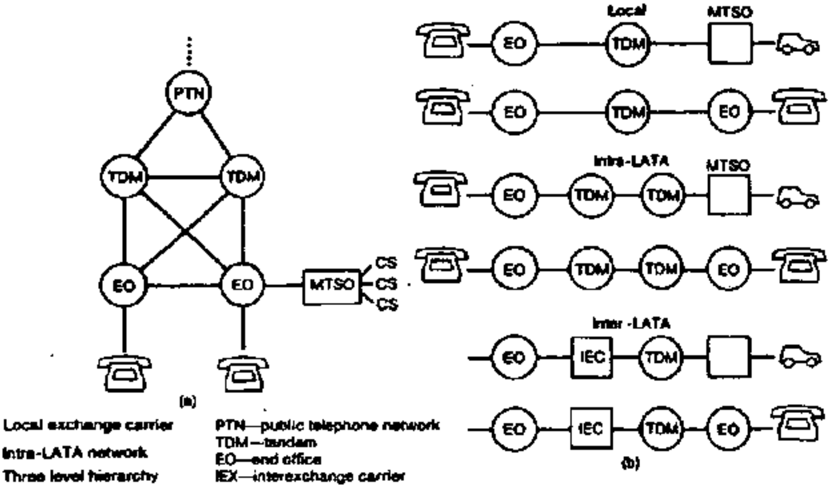


Figure 11.9 Three-level hierarchy. (*a*) Interconnection of MTSO; (*b*) three types of call.

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 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

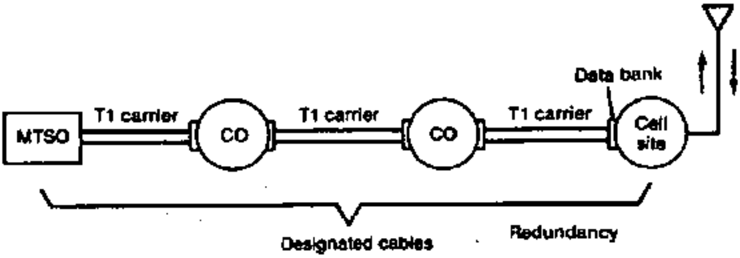


Figure 12.1 Data-link connection through cable.

$$(\underset{\substack{\uparrow\\ \text{one sample}}}{7} \times \underset{\substack{\uparrow\\ \text{signaling bit}}}{1}) \times \underset{\substack{\uparrow\\ \text{samples/channel}}}{24} + \underset{\substack{\uparrow\\ \text{frame bit}}}{1} = 193 \text{ bits/frame}$$

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.

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-fit 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

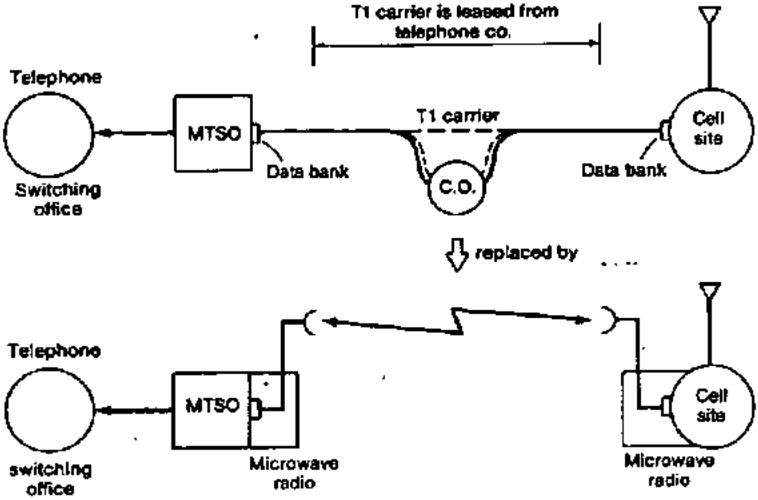


Figure 12.2 Replacement of T1 carrier cables by microwave radios.

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 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.

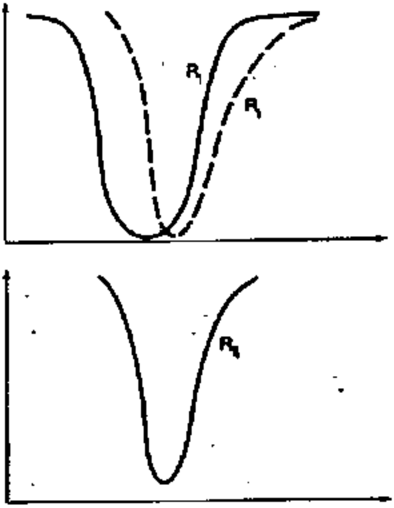


Figure 12.3 Formation of a simultaneous fade from two over-lapped fades.

$$20 \log L < -20 \text{ dB} \qquad \text{or} \qquad L < 0.1$$

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.

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 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

$$F_N = \tfrac{1}{2}qL^{-2} \qquad \text{for} \qquad L < 0.1 \tag{12.3-4}$$

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.

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

$$T_i = N\bar{t} = \begin{cases} (37904.4L/\text{year})\,(490L\ \text{s}) \\ 1857\ \text{s/year} \qquad \text{at} \qquad L = 0.01\ (\text{or } -40\ \text{dB}) \\ 31\ \text{min/year} \qquad \text{at} \qquad L = 0.01\ (\text{or } -40\ \text{dB}) \end{cases}$$

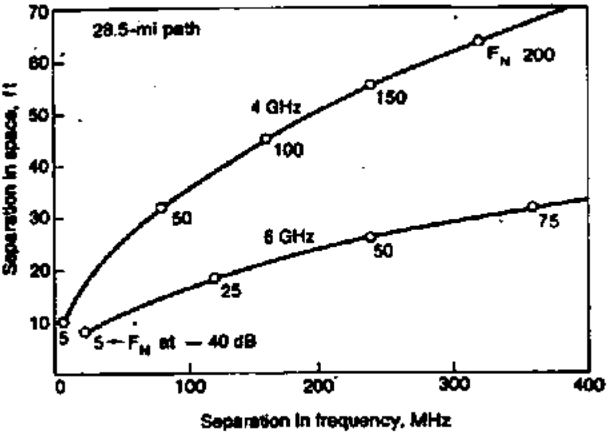


Figure 12.4 Separations in space and in frequency that provide equal values of F_N, the ratio of the number of fades to the number of simultaneous fades. (*After Vigants, Ref. 2.*)

We can use the same step to obtain the total fading time T_t in 4-GHz propagation.

$$T_t = N\bar{t} = \begin{cases} (22002L/\text{year})\ (408L\ \text{s}) & \\ 897.68\ \text{s/year} & \text{at} \quad L = 0.01\ (\text{or } -40\ \text{dB}) \\ 15\ \text{min/year} & \text{at} \quad L = 0.01\ (\text{or } -40\ \text{dB}) \end{cases}$$

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

$$T_{\bar{t}} = \begin{cases} 3.1\ \text{min/year} & (\text{in the 6-GHz band}) \\ 1.5\ \text{min/year} & (\text{in the 4-GHz band})\ . \end{cases}$$

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

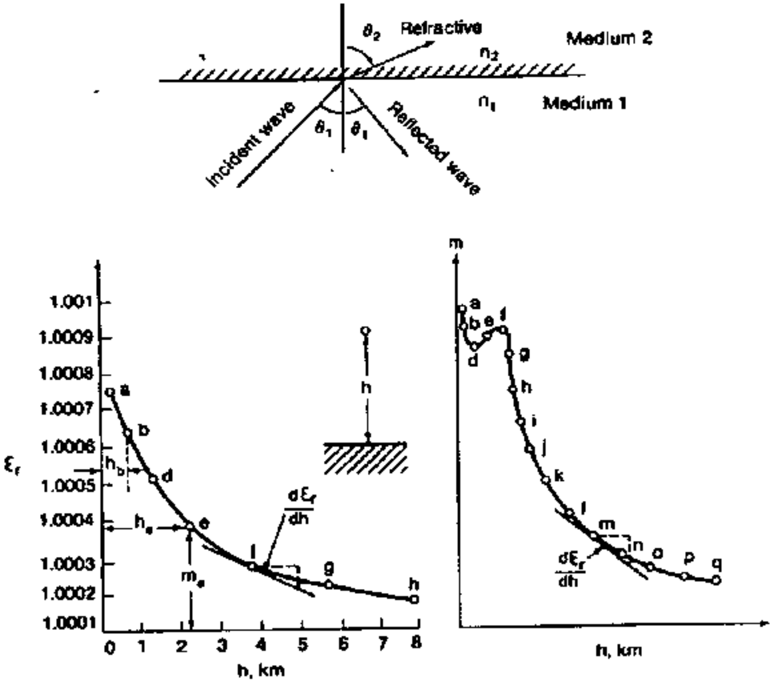


Figure 12.5 Illustration of Snell's law and finding *dn/dh* from a measured piece of data. (*After Hund, Ref. 4.*)

$$\frac{\sin\,\theta_1}{\sin\,\theta_2} = \frac{n_2}{n_1} = \frac{\sqrt{\mu_2\varepsilon_2}}{\sqrt{\mu_1\varepsilon_1}} = \frac{C_1}{C_2} \tag{12.4-1}$$

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.

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 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

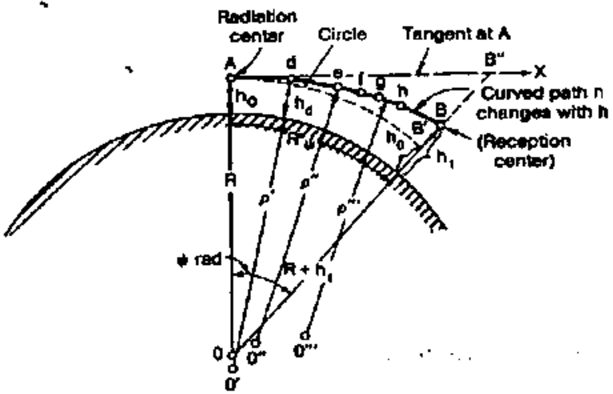


Figure 12.6 The radius of curvature of the wave path. *(After Hund, Ref. 4.)* Curved path *AB* requires the least time to move electromagnetic wave energy from a through angle ψ to line *OB"*.

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.

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

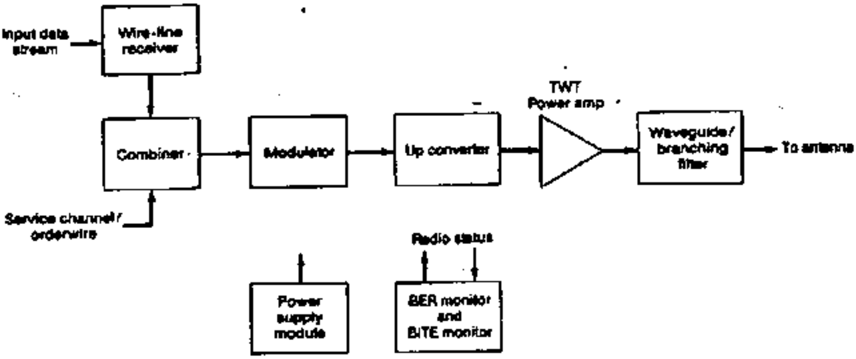


Figure 12.7 Microwave radio transmit block diagram.

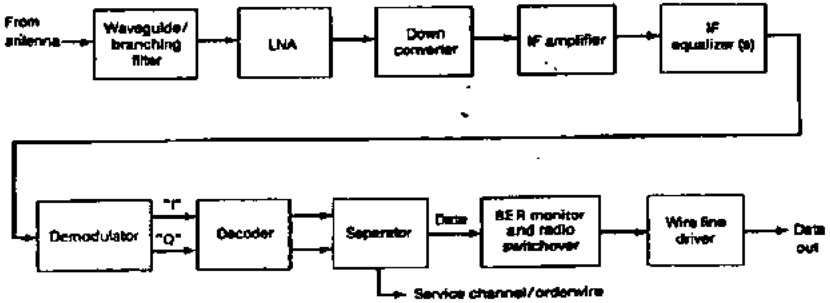


Figure 12.8 Microwave radio receive block diagram.

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.

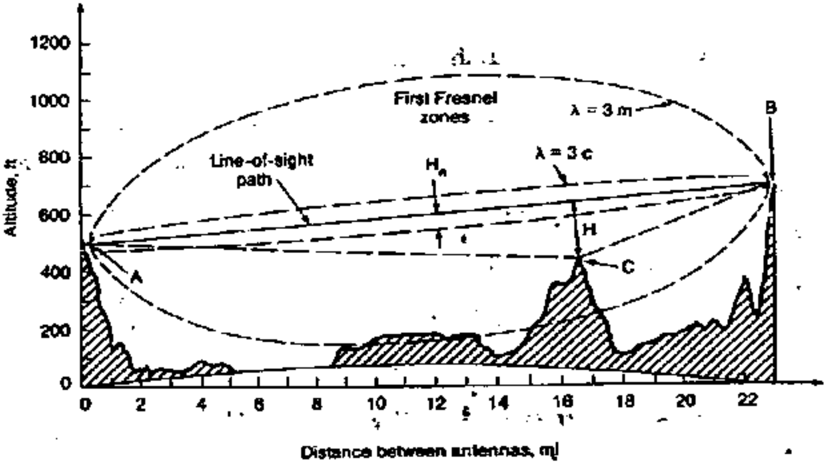


Figure 12.9 The clearance distance from the closest objects. (*From Ref. 8, p. 439.*) Typical profile plot showing first Fresnel zones for 100 MHz and 10 GHz.

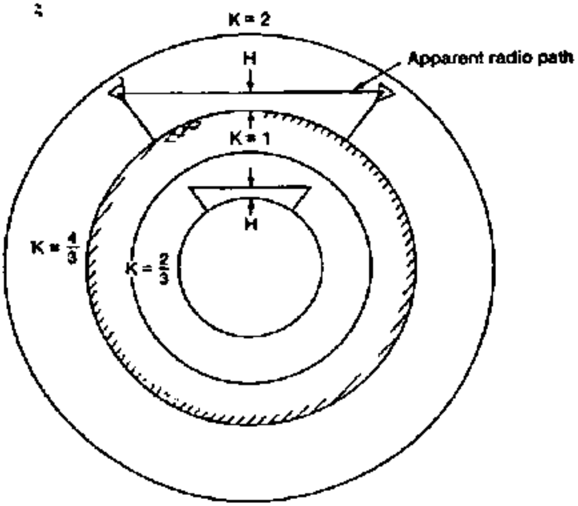


Figure 12.10 Illustration of earth bulge conditions.

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 value other than $K = \frac{4}{3}$ 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 $K = \frac{4}{3}$, 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 $K = \frac{2}{3}$ will tilt upward and may cause interference over a long distance. An earth bulge factor of $\frac{2}{3}$ or $\frac{4}{3}$ is used to provide a clear area.

$$H \geqslant \frac{d_1 d_2}{2} \qquad \text{(for a factor of } \tfrac{4}{3}\text{)} \tag{12.5-2}$$

$$H \geqslant d_1 d_2 \qquad \text{(for a factor of } \tfrac{2}{3}\text{)} \tag{12.5-3}$$

where d_1 and d_2 are in miles and H is in feet.

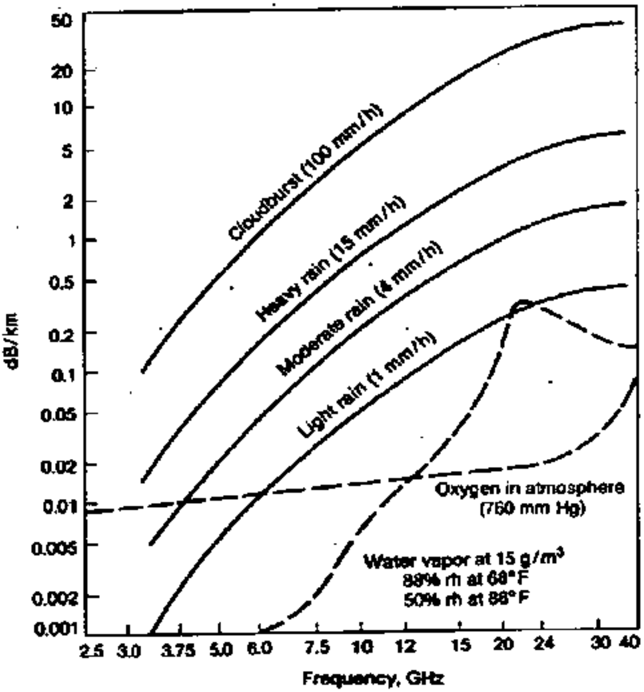


Figure 12.11 Estimated atmospheric absorption. (*From Ref. 9, p. 443.*)

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 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.

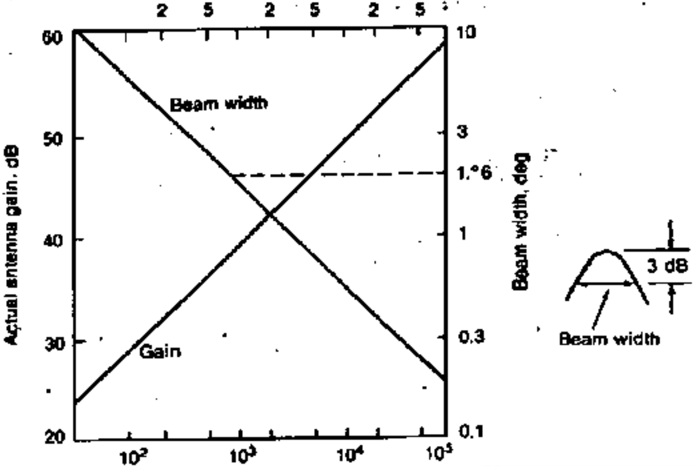
Microwave antenna location. Sometimes the reception is poor after the microwave antenna has been mounted on the antenna tower. A quick way to check the installation before making any other changes is to move the microwave antenna around within a 2 to 4 ft radius of the previous position and check the 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.



Note: Abscissa is actual antenna area, and "actual antenna gain" is taken to be 3 dB below theoretical.

Antenna	$\frac{A}{\lambda^2}$		
	4 GHz	8 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

Figure 12.12 Approximate antenna gain and beamwidth. (*From Ref. 8, p. 445.*)

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 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.* 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 re-

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

ception 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.
   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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