CELLULAR & MOBILE COMMUNICATION

LECTURE NOTES

B.TECH (IV-I SEM)

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UNIT-III (A)

CO-CHANNEL INTERFERENCE

The frequency-reuse method is useful for increasing the efficiency of spectrum usage but results in cochannel interference because the same frequency channel is used repeatedly in different cochannel cells. Application of the cochannel interference reduction factor q= D/R = 4.6 for a seven-cell reuse pattern (K = 7).

In most mobile radio environments, use of a seven-cell reuse pattern is not sufficient to avoid cochannel interference. Increasing K > 7 would reduce the number of channels per cell, and that would also reduce spectrum efficiency. Therefore, it might be advisable to retain the same number of radios as the seven-cell system but to sector the cell radially, as if slicing a pie. This technique would reduce cochannel interference and use channel sharing and channel borrowing schemes to increase spectrum efficiency.

REAL TIME CO-CHANNEL INTERFERENCE MEASURED AT MOBILE RADIO TRANSCEIVER

When the carriers are angularly modulated by the voice signal and the RF frequency difference between them is much higher than the fading frequency, measurement of the signal carrier-to-interference ratio *C/I* reveal that the signal is

$$e_1 = S(t)\sin(\omega t + \phi_1) \qquad (9.3\text{-}1)$$

and the interference is

$$e_2 = I(t)\sin(\omega t + \phi_2) \qquad (9.3\text{-}2)$$

The received signal is

$$e(t) = e_1(t) + e_2(t) = R\sin(\omega t + \psi) \qquad (9.3\text{-}3)$$

where

$$R = \sqrt{[S(t)\cos\phi_1 + I(t)\cos\phi_2]^2 + [S(t)\sin\phi_1 + I(t)\sin\phi_2]^2} \qquad (9.3\text{-}4)$$

and

$$\psi = \tan^{-1}\frac{S(t)\sin\phi_1 + I(t)\sin\phi_2}{S(t)\cos\phi_1 + I(t)\cos\phi_2} \qquad (9.3\text{-}5)$$

The envelope R can be simplified in Eq. (9.3-4), and R^2 becomes

$$R^2 = [S^2(t) + I^2(t) + 2S(t)I(t)\cos(\phi_1 - \phi_2)] \qquad (9.3\text{-}6)$$

Following kozono and sakamoto s analysis Eq (9.3-6),the term $s^2(t)+I^2(t)$ fluctuates closetothe fading frequency V/λ and the term 2S(t)+I(t)cos($\emptyset_1$-$\emptyset_2$)fluctuates to a frequency close to *d/dt*($\emptyset_1$-$\emptyset_2$).which is much higher than the fading frequency. Then the two parts of the squared envelope can be separated as

$$X = S^2(t) + I^2(t) \qquad (9.3\text{-}7)$$

$$Y = 2S(t)I(t)\cos(\phi_1 - \phi_2) \qquad (9.3\text{-}8)$$

assume that the random variables s(t),I(t),$\varnothing_1$,$\varnothing_2$ are independent ;then the average processes on X and Y are

$$\overline{X} = \overline{S^2(t)} + \overline{I^2(t)} \qquad (9.3\text{-}9)$$

$$\overline{Y^2} = 4\overline{S^2(t)I^2(t)}(\tfrac{1}{2}) = 2\overline{S^2(t)I^2(t)} \qquad (9.3\text{-}10)$$

The signal to Interference ratio Γ becomes

$$\Gamma = \frac{\overline{S^2(t)}}{\overline{I^2(t)}} = k + \sqrt{k^2 - 1} \qquad (9.3\text{-}11)$$

$$k = \frac{\overline{X^2}}{\overline{Y^2}} - 1 \qquad (9.3\text{-}12)$$

Because X and Y can be separated in Eq.(9.3-6),the preceding computation of Γ in Eq.(9.3-11)could have been accomplished by means of an envelope detector ,analog to digital converter, and a micro computer. The sampling delay time Δt should be small enough to satisfy

$$S(t) \approx S(t + \Delta t), \qquad I(t) \approx I(t + \Delta t) \qquad (9.3\text{-}13)$$

and

$$\mathrm{E}\left[\cos[\phi_1(t) - \phi_2(t)]\cos[\phi_1(t + \Delta t) - \phi_2(t + \Delta t)]\right] \approx 0 \qquad (9.3\text{-}14)$$

Determining the delay time Δt to meet the requirement of Eq.(9.3-13) for this calculation is difficult and is a drawback to this measurement technique. Therefore, real time cochannel interference measurement is difficult to achieve in practice.

DESIGN OF ANTENNA SYSTEM

DESIGN OF AN OMNIDIRECTIONAL ANTENNA SYSTEM IN THE WORST CASE:

The value of q = 4.6 is valid for a normal interference case in a K=7 cell pattern. In this section we would like to prove that a K=7 cell pattern does not provide a sufficient frequency re-use distance separation eyen when an ideal condition of flat terrain is assumed. The worst case is at the location where the weakest signal from its own cell site but strong interferences from all interfering cell sites. In the worst case the mobile unit is at the cell boundary R, as shown in Fig. **3.** The distances from all six cochannel interfering sites are also shown in the figure: two distances of D - R, two distances of D, and two distances of D + R.

Following the mobile radio propagation rule of 40 dB/dec, we

$$C \propto R^{-4} \qquad I \propto D^{-4}$$

obtain Then the carrier-to-interference ratio is

$$\frac{C}{I} = \frac{R^{-4}}{2(D-R)^{-4} + 2(D)^{-4} + 2(D+R)^{-4}}$$

$$= \frac{1}{2(q-1)^{-4} + 2(q)^{-4} + 2(q+1)^{-4}} \qquad (9.4\text{-}1a)$$

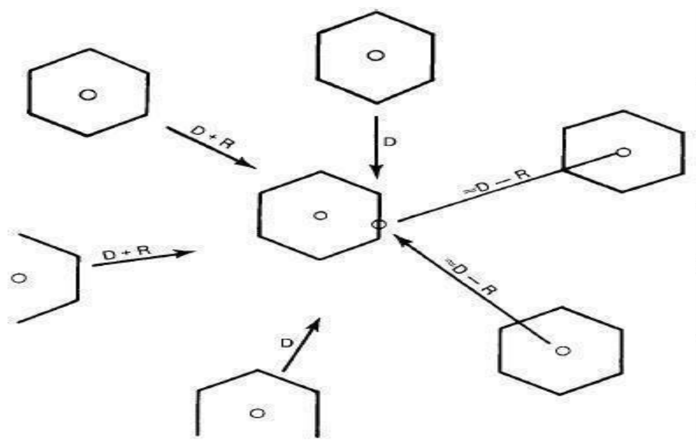


Fig.3.Cochannel interference (worst case)

Where q=4.6 is derived from the normal case. Substituting q=4.6 into above eqn. we obtain C/I =54 or 17 dB, which is lower than 18 dB. To be conservative, we may use the shortest distance D – R for all six interferers as a worst case; then we have

$$\frac{C}{I} = \frac{R^{-4}}{6(D-R)^{-4}} = \frac{1}{6(q-1)^{-4}} = 28 = 14.47 \text{ dB}$$

In reality, because of the imperfect site locations and the rolling nature of the terrain configuration, the C/I received is always worse than 17 dB and could be 14 dB and lower. Such an instance can easily our in a heavy traffic situation; therefore, the system must be designed around the C/I of the worst case. In that case, a cochannel interference reduction factor of q=4.6 is insufficient.

Therefore, in an unidirectional-cell system, K = 9 or K 12 would be a correct choice. Then the values of q are

$$q = \begin{cases} \frac{D}{R} = \sqrt{3K} \\ 5.2 \quad K = 9 \\ 6 \quad K = 12 \end{cases}$$

Substituting these values in Eq. (9.4-1), we obtain

$$\frac{C}{I} = 84.5 \,(=)\, 19.25 \text{ dB} \qquad K = 9$$
$$\frac{C}{I} = 179.33 \,(=)\, 22.54 \text{ dB} \qquad K = 12$$

DESIGN OF ANTENNA SYSTEM

Design of a Directional Antenna System:

When the call traffic begins to increase, we need to use the frequency spectrum efficiently and avoid increasing the number of cells K in a seven-cell frequency reuse pattern. When K increases, the number of frequency channels assigned in a cell must become smaller (assuming a total allocated channel divided by K) and the efficiency of applying the frequency reuse scheme decrease.

Instead of increasing the number K in a set of cells, let us keep K =7 and introduce a directional antenna arrangement. The cochannel interference can be reduced by using directional antenna. This means that each cell is divided into three or six sectors and uses three or six directional antennas at a base station. Each sector is assigned a set of frequencies (channels). The interference between two cochannel cells decreases as shown Fig.4.2 **Directional antennas in K=7 cell patterns:**

Three sector case: The three-sector case is shown in Fig.4.2. To illustrate the worst case situation, two cochannel cells are shown in Fig. 4.3(a). The mobile unit at position E will experience greater interference in the lower shaded cell sector than in the upper shaded cell-sector site. This is because the mobile receiver receives the weakest signal from its own cell but fairly strong interference from the interfering cell. In a three-sector case, the interference is effective in only one direction because the front-to-back ratio of a cell-site directional antenna is at least 10 dB or more in a mobile radio environment. The worst-case cochannel interference in the directional-antenna sectors in which interference occurs may be calculated. Because of the use of directional antennas, the number of principal interferers is reduced from six to two (Fig.4.2). The worst case of C/I occurs when the mobile unit is at position E, at which point the distance between the mobile unit and the two interfering antennas is roughly D + (R/2); however, C/I can be calculated more precisely as follows. The value of C/I can be obtained by the following expression (assuming that the worst case is at position E at which the distances from two interferers are D + 0.7R and D).

$$\frac{C}{I}(\text{worst case}) = \frac{R^{-4}}{(D+0.7R)^{-4}+D^{-4}} = \frac{1}{(q+0.7)^{-4}+q^{-4}}$$

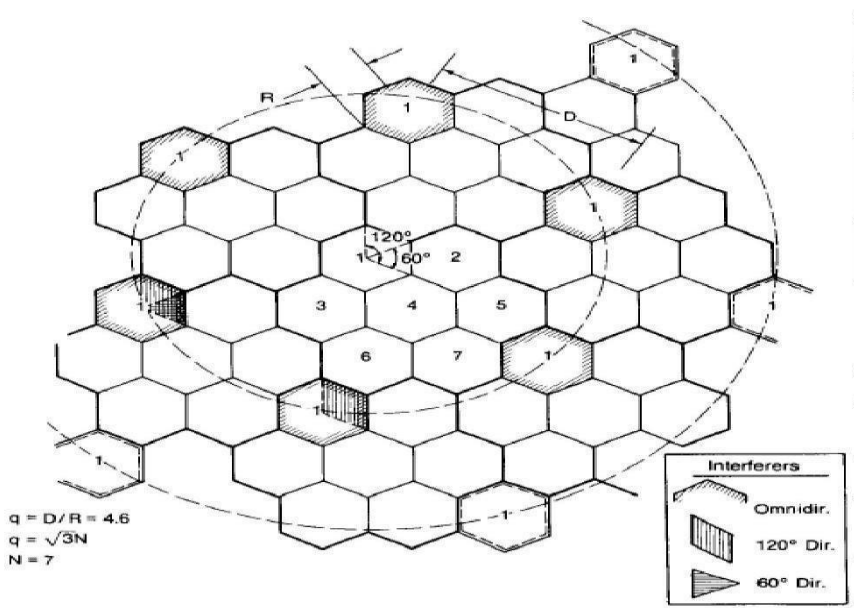
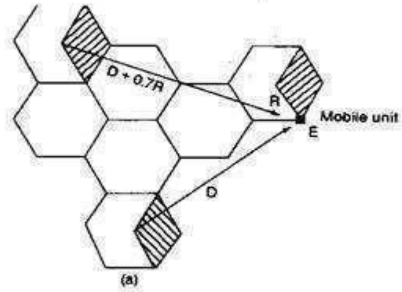


Fig.4.2.Interfering cells shown in a seven cell system (two-tiers)



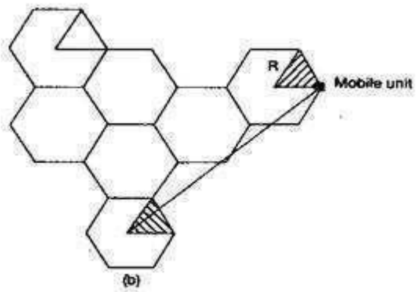


Fig.4.3. Determination of C/I in a directional antenna system. (a)Worst case in a 120 directional Antenna system (N=7); (b) worst case in a 60 directional antenna system(N=7)

Let q=4.6; then we have

$$\frac{C}{I} \text{ (worst case)} = 285 \; (=) \; 24.5 \text{ dB}$$

The C/I received by a mobile unit from the 120° directional antenna sector system expressed in Eq. above greatly exceeds 18 dB in a worst case. Equation above shows that using directional antenna sectors can improve the signal- to-interference ratio, that is, reduce the cochannel interference. However, in reality, the C/I could be 6 dB weaker than in Eq. given above in a heavy traffic area as a result of irregular terrain contour and imperfect site locations. The remaining 18.5 dB is still adequate.

Six-sector case: We may also divide a cell into six sectors by using six 60°-beam directional antennas as shown in Fig.4.2. In this case, only one instance of interference can occur in each sector as shown in Fig, 4.2. Therefore, the carrier-to-interference ratio in this case is which shows a further reduction of cochannel interference. If we use the same argument as we did for Eq. above and subtract 6 dB from the result of Eq. the remaining 23 dB is still more than adequate. When heavy traffic occurs, the 60°-sector configuration can be used to reduce cochannel interference. However, fewer channels are generally allowed in a 60° sector and the trunking efficiency decreases. In certain cases, more available channels could be assigned in a 60° sector.

Directional antenna in K = 4 cell pattern:

Three-sector case: To obtain the carrier-to-interference ratio, we use the same procedure as in the K = 7 cell- pattern system. The 120° directional antennas used in the sectors reduced the interferers to two as in K = 7 systems, as shown in Fig.4.4.

We can apply Eq. here. For K = 4, the value of q = 3.46; therefore, Eq. becomes

$$\frac{C}{I}\text{(worst case)} = \frac{1}{(q+0.7)^{-4} + q^{-4}} = 97 = 20\text{ dB}$$

If, using the same reasoning used with Eq. above, 6 dB is subtracted from the result of Eq. above, the remaining 14 dB is unacceptable.

Six-sector case: There is only one interferer at a distance of D + R shown in Fig.4.4. With q=3.46, we can obtain

$$\frac{C}{I}\text{(worst case)} = \frac{R^{-4}}{(D+R)^{-4}} = \frac{1}{(q+1)^{-4}} = 355 = 26\text{ dB}$$

If 6 dB is subtracted from the above result, the remaining 20 dB is adequate.

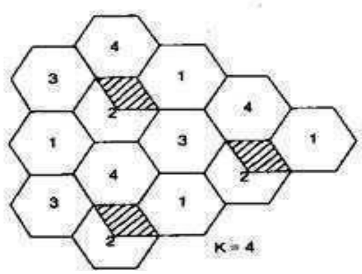


Fig. 4.4 Interference with frequency reuse pattern K=4.

Under heavy traffic conditions, there is still a great deal of concern over using a K =4 cell pattern in a 60° sector.

Comparing K =7 and N =4 systems:

A K =7 cell pattern system is a logical way to begin an omnicell system. The co-channel reuse distance is more or less adequate, according to the designed criterion. When the traffic increases, a three sector system should be implemented, that is, with three 120° directional antennas in place. In certain hot spots, 60° sectors can be used locally to increase the channel utilization.

If a given area is covered by both K=7 and K=4 cell patterns and both patterns have a six-sector configuration, then the K=7 system has a total of 42 sectors, but the K=4 system has a total of only 24 sectors and, of course, the system of K=7 and six sectors has less cochannel interference.

One advantage of 60° sectors with K=4 is that they require fewer cell sites than 120 sectors with K=7. Two disadvantages of 60 deg sectors are that (1) they require more antennas to he mounted on the antenna mast and

(2) they often require more frequent handoffs because of the increased chance that the mobile units will travel across the six sectors of the call. Furthermore, assigning the proper frequency channel to the mobile unit in each sector is more difficult unless the antenna height at the cell site is increased so that the mobile unit can be located more precisely. In reality the terrain is not flat, end coverage is never uniformly distributed; in addition, the

directional antenna front-to-back power ratio in the field is very difficult to predict. In small cells, interference could become uncontrollable; then the use of a K = 4 pattern with 60 deg sectors in small cells needs to be considered only for special implementations such an portable cellular systems or narrow beam applications. For small cells, a better alternative scheme is to use a K =7 pattern with 120° sectors plus the underlay-overlay configuration.

ANTENNA PARAMETERS AND THEIR EFFECTS:

3.7 Antenna Parameters and their Effects

The performance of antenna can be described by various parameters. Every antenna parameter will effect on the performance efficiency of antenna.

The different antenna parameters and their effects discussed in this topic are,

1) Radiation pattern	2) Beamwidth
3) Gain	4) Power density
5) Radiation intensity	6) Directivity
7) Efficiency	8) Effective aperture
9) Antenna bandwidth	10) Front to-Back-ratio
11) Polarization	12) Input impedance.

1) Radiation pattern : An antenna is a fundamental radiating component of an electrical system that links free space with the receiver. The energy radiated by an antenna is not uniform in all the directions. It is strong in one direction and weak or zero in some other direction. The amount of energy being radiated in a direction is measured as the field strength at a point located at distance from the antenna.

The radiation pattern or an antenna pattern is nothing but a mathematical function or it is a graphical representation of radiation properties as a function of the space co-ordinates . The radiation pattern of antenna is usually measured in far-field.

A simple space co-ordinate system of antenna is shown in Fig. 3.24.

In the radiation pattern of antenna the major and minor lobes are shown in Fig. 3.24. The elevation and azimuth planes are mentioned. At a point represented at a distance from the antenna field strength is measured and area be 'dA'.

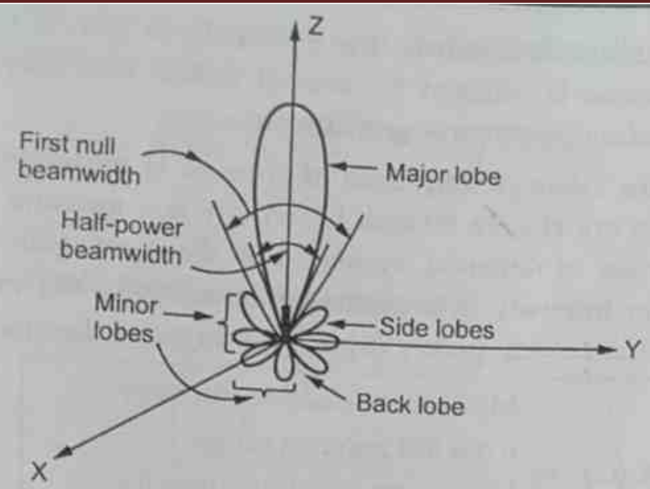


Fig. 3.26 Antenna radiation pattern : Radiation lobes and beamwidth

2) **Antenna beamwidth :** For an antenna it is a measure of directivity. It is an angular width (in degrees) that is measured on the pattern between two points where the power radiated falls to half of it's maximum value. It is known as " half-power beamwidth"

In the Fig. 3.27 in the radiation pattern the angle 'AOB' is the beamwidth of antenna used.

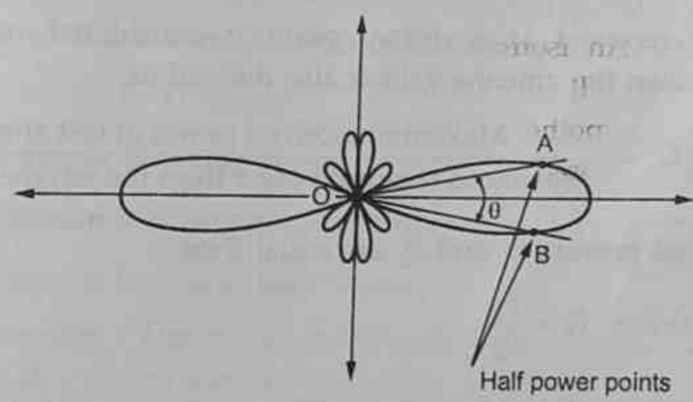


Fig. 3.27 Beam width angle (AOB)

Effects of beamwidth : The direction of signal reception can be determined by the type of beam in the radiation pattern. It is known by a narrow beam.

$$\text{Directivity } D \propto \frac{1}{\text{Beamwidth}}.$$

Thus if the beamwidth is narrow the directivity or gain of the antenna is high. The beamwidth of antenna is affected by several factors like shape of radiation pattern, wavelength (λ), Radius of antenna aperture etc.

3) Antenna gain : The performance of antenna is measured in terms of gain. The directivity and gain are closely related. Directivity is a measure that expresses only the directional properties of antenna system. But the term gain is defined as ratio of maximum radiation intensity in a particular direction to the maximum intensity from the reference antenna having power input level in same direction.

$$\boxed{\text{Gain (G)} = \frac{\text{Maximum radiation intensity from test antenna setup}}{\text{Maximum radiation intensity from a reference antenna having same power input}}}$$

The gain defined above did not include antenna efficiency. If the reference antenna is an isotropic antenna with 100 % efficiency η, the gain will be,

$$G = \frac{\text{Maximum radiation intensity with respect to a test antenna}}{\text{Radiation intensity from an isotropic antenna}}$$

Note : When 100 % η is available the isotropic antenna is a lossless antenna.

If signal power received at a distant point is considered in the direction of maximum radiation then the antenna gain is also defined as,

$$\text{Gain } G = \frac{\text{Maximum received power at test antenna}}{\text{Maximum received power from the reference antenna}}$$

If the two received powers P_1 and P_2 are equal then,

$$\text{Gain} = G = \frac{P_1}{P_2}$$

Directive gain : It is defined as the ratio of antenna radiation intensity in the direction to that of the average radiation power level.

$$\text{Directive gain} = \frac{\text{Radiation intensity in a given direction}}{\text{Average radiated power level}}$$

It depends only on the distribution of the power radiation in space and not on the antenna input power.

If the power distribution is not good it will effect on directive gain and in turn the system performance.

Power gain : In a given direction power gain is defined as the ratio of radiation intensity to the total average input power.

$$\text{Power gain} = G_p = \frac{\text{Radiation intensity in a particular direction}}{\text{Total average power input}}$$

It depends on

i) Volume of the solid radiation pattern of antenna.

ii) Sharpness of the antenna lobe.

$$G_p = 10 \log_{10}\left(\frac{P_1}{P_2}\right)$$

(or)

$$G_p = 20 \log_{10}\left(\frac{V_1}{V_2}\right)$$

Effects of power gain : Consider a given amount of power input to an antenna. The power density at a particular point in space and power gain of antenna are proportional to each other. Thus the available signal in this location to receiving antenna can be increased by increasing the transmitter power level. The power distribution in radiation pattern of antenna influences on gain of the antenna.

4) Power density : The electromagnetic waves are generally used to transport the data through a guiding medium or a wireless medium from one point to another point. The electromagnetic fields are associated with energy and power. The amount of power associated with it is expressed by an instantaneous Poynting vector as,

$$W = E \cdot H \text{ where}$$

W is instantaneous Poynting vector

E is instantaneous electric field intensity

and H is instantaneous magnetic field intensity.

5) Radiation intensity : The radiation intensity in a direction is "The power per unit solid angle " or it is the power radiated from the antenna per unit solid angle. It is denoted as ϕ or U.

6.10 Diversity Techniques

Diversity is a powerful communication receiver technique that provides wireless link improvement at relatively low cost. Unlike equalization, diversity requires no training overhead since a training sequence is not required by the transmitter. Furthermore, there are a wide range of diversity implementations, many which are very practical and provide significant link improvement with little added cost.

Diversity exploits the random nature of radio propagation by finding independent (or at least highly uncorrelated) signal paths for communication. In virtually all applications, diversity decisions are made by the receiver, and are unknown to the transmitter.

The diversity concept can be explained simply. If one radio path undergoes a deep fade, another independent path may have a strong signal. By having more than one path to select from, both the instantaneous and average SNRs at the receiver may be improved, often by as much as 20 dB to 30 dB.

As shown in Chapters 3 and 4, there are two types of fading – small-scale and large-scale fading. Small-scale fades are characterized by deep and rapid amplitude fluctuations which occur as the mobile moves over distances of just a few wavelengths. These fades are caused by multiple reflections from the surroundings in the vicinity of the mobile. Small-scale fading typically results in a Rayleigh fading distribution of signal strength over small distances. In order to prevent deep fades from occurring, *microscopic diversity techniques* can exploit the rapidly changing signal. For example, the small-scale fading shown in Figure 3.1 reveals that if two antennas are separated by a fraction of a meter, one may receive a null while the other receives a strong signal. By selecting the best signal at all times, a receiver can mitigate small-scale fading effects (this is called *antenna diversity* or *space diversity*).

Large-scale fading is caused by shadowing due to variations in both the terrain profile and the nature of the surroundings. In deeply shadowed conditions, the received signal strength at a mobile can drop well below that of free space. In Chapter 3, large-scale fading was shown to be log-normally distributed with a standard deviation of about 10 dB in urban environments. By selecting a base station which is not shadowed when others are, the mobile can improve substantially the average signal-to-noise ratio on the forward link. This is called *macroscopic diversity*, since the mobile is taking advantage of large separations between the serving base stations.

Macroscopic diversity is also useful at the base station receiver. By using base station antennas that are sufficiently separated in space, the base station is able to improve the reverse link by selecting the antenna with the strongest signal from the mobile.

virtually any diversity application, although often at much greater cost and complexity than other diversity techniques.

6.10.3 Practical Space Diversity Considerations

Space diversity, also known as antenna diversity, is one of the most popular forms of diversity used in wireless systems. Conventional cellular radio systems consist of an elevated base station antenna and a mobile antenna close to the ground. The existence of a direct path between the transmitter and the receiver is not guaranteed and the possibility of a number of scatterers in the vicinity of the mobile suggests a Rayleigh fading signal. From this model [Jak70], Jakes deduced that the signals received from spatially separated antennas on the mobile would have essentially uncorrelated envelopes for antenna separations of one half wavelength or more.

The concept of antenna space diversity is also used in base station design. At each cell site, multiple base station receiving antennas are used to provide diversity reception. However, since the important scatterers are generally on the ground in the vicinity of the mobile, the base station antennas must be spaced considerably far apart to achieve decorrelation. Separations on the order of several tens of wavelengths are required at the base station. Space diversity can thus be used at either the mobile or base station, or both. Figure 6.12 shows a general block diagram of a space diversity scheme [Cox83a].

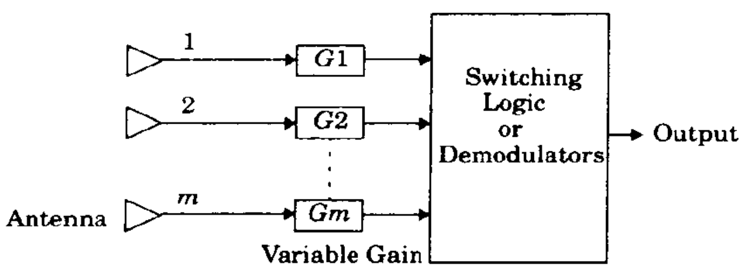


Figure 6.12
Generalized block diagram for space diversity.

Space diversity reception methods can be classified into four categories [Jak71]:

1. Selection diversity
2. Feedback diversity
3. Maximal ratio combining
4. Equal gain diversity

6.10.3.1 Selection Diversity

Selection diversity is the simplest diversity technique analyzed in section 6.10.1. A block diagram of this method is similar to that shown in Figure 6.12, where m demodulators are used to provide m diversity branches whose gains are adjusted to provide the same average SNR for each branch. As derived in section 6.10.1, the receiver branch having the highest instantaneous SNR is connected to the demodulator. The antenna signals themselves could be sampled and the best one sent to a single demodulator. In practice, the branch with the largest $(S + N) / N$ is used, since it is difficult to measure SNR. A practical selection diversity system cannot function on a truly instantaneous basis, but must be designed so that the internal time constants of the selection circuitry are shorter than the reciprocal of the signal fading rate.

6.10.3.2 Feedback or Scanning Diversity

Scanning diversity is very similar to selection diversity except that instead of always using the best of M signals, the M signals are scanned in a fixed sequence until one is found to be above a predetermined threshold. This signal is then received until it falls below threshold and the scanning process is again initiated. The resulting fading statistics are somewhat inferior to those obtained by the other methods but the advantage with this method is that it is very simple to implement — only one receiver is required. A block diagram of this method is shown in Figure 6.13.

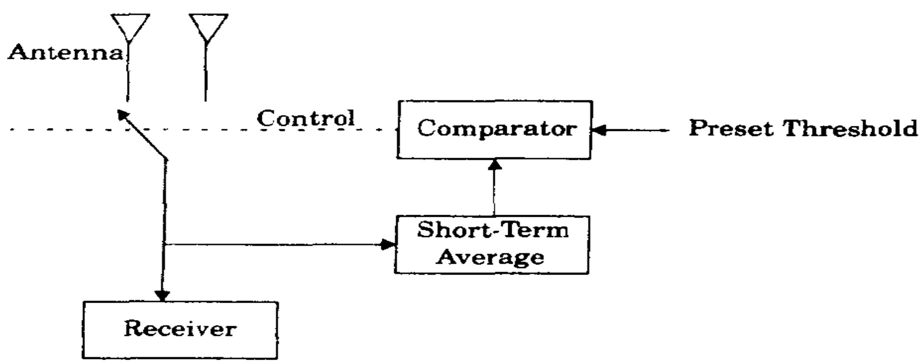


Figure 6.13
Basic form of scanning diversity.

6.10.3.3 Maximal Ratio Combining

In this method first proposed by Kahn [Kah54], the signals from all of the M branches are weighted according to their individual signal voltage to noise power ratios and then summed. Figure 6.14 shows a block diagram of the technique. Here, the individual signals must be co-phased before being summed (unlike selection diversity) which generally requires an individual receiver and phasing circuit for each antenna element. Maximal ratio combining produces an output SNR equal to the sum of the individual SNRs, as explained in section 6.10.2. Thus, it has the advantage of producing an output with an acceptable SNR even when none of the individual signals are themselves acceptable. This technique gives the best statistical reduction of fading of any known linear diversity combiner. Modern DSP techniques and digital receivers are now making this optimal form of diversity practical.

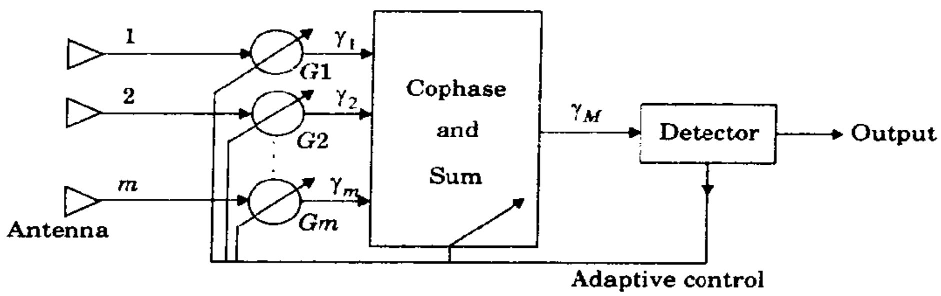


Figure 6.14
Maximal ratio combiner.

6.10.3.4 Equal Gain Combining

In certain cases, it is not convenient to provide for the variable weighting capability required for true maximal ratio combining. In such cases, the branch weights are all set to unity but the signals from each branch are co-phased to provide *equal gain combining* diversity. This allows the receiver to exploit signals that are simultaneously received on each branch. The possibility of producing an acceptable signal from a number of unacceptable inputs is still retained, and performance is only marginally inferior to maximal ratio combining and superior to selection diversity.

6.10.4 Polarization Diversity

At the base station, space diversity is considerably less practical than at the mobile because the narrow angle of incident fields requires large antenna spacings [Vau90]. The comparatively high cost of using space diversity at the base station prompts the consideration of using orthogonal polarization to exploit polarization diversity. While this only provides two diversity branches it does allow the antenna elements to be co-located.

In the early days of cellular radio, all subscriber units were mounted in vehicles and used vertical whip antennas. Today, however, over half of the subscriber units are portable. This means that most subscribers are no longer using vertical polarization due to hand-tilting when the portable cellular phone is

used. This recent phenomenon has sparked interest in polarization diversity at the base station.

Measured horizontal and vertical polarization paths between a mobile and a base station are reported to be uncorrelated by Lee and Yeh [Lee72]. The decorrelation for the signals in each polarization is caused by multiple reflections in the channel between the mobile and base station antennas. Chapter 3 showed that the reflection coefficient for each polarization is different, which results in different amplitudes and phases for each, or at least some, of the reflections. After sufficient random reflections, the polarization state of the signal will be independent of the transmitted polarization. In practice, however, there is some dependence of the received polarization on the transmitted polarization.

Circular and linear polarized antennas have been used to characterize multipath inside buildings [Haw91], [Rap92a], [Ho94]. When the path was obstructed, polarization diversity was found to dramatically reduce the multipath delay spread without significantly decreasing the received power.

. While polarization diversity has been studied in the past, it has primarily been used for fixed radio links which vary slowly in time. Line-of-sight microwave links, for example, typically use polarization diversity to support two simultaneous users on the same radio channel. Since the channel does not change much in such a link, there is little likelihood of cross polarization interference. As portable users proliferate, polarization diversity is likely to become more important for improving link margin and capacity. An outline of a theoretical model for the base station *polarization diversity reception* as suggested by Kozono [Koz85] is given below.

Theoretical Model for Polarization Diversity

It is assumed that the signal is transmitted from a mobile with vertical (or horizontal) polarization. It is received at the base station by a polarization diversity antenna with 2 branches. Figure 6.15 shows the theoretical model and the system coordinates. As seen in the figure, a polarization diversity antenna is composed of two antenna elements V_1 and V_2, which make a $\pm\alpha$ angle (polarization angle) with the Y axis. A mobile station is located in the direction of offset angle β from the main beam direction of the diversity antenna as seen in Figure 6.15(b).

Some of the vertically polarized signals transmitted are converted to the horizontal polarized signal because of multipath propagation. The signal arriving at the base station can be expressed as

$$x = r_1 \cos(\omega t + \phi_1) \tag{6.71.a}$$

$$y = r_2 \cos(\omega t + \phi_2) \tag{6.71.b}$$

where x and y are signal levels which are received when $\beta = 0$. It is assumed that r_1 and r_2 have independent Rayleigh distributions, and ϕ_1 and ϕ_2 have independent uniform distributions.

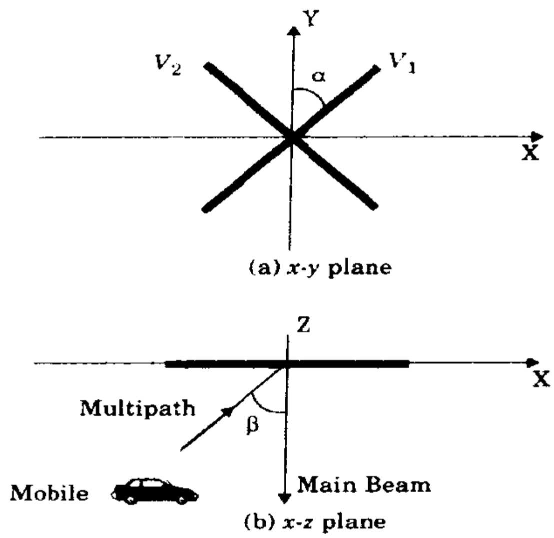


Figure 6.15
Theoretical Model for base station polarization diversity based on [Koz85].

The received signal values at elements V_1 and V_2 can be written as:

$$v_1 = (ar_1\cos\phi_1 + r_2 b\cos\phi_2)\cos\omega t - (ar_1\sin\phi_1 + r_2 b\sin\phi_2)\sin\omega t \qquad (6.72)$$

$$v_2 = (-ar_1\cos\phi_1 + r_2 b\cos\phi_2)\cos\omega t - (-ar_1\sin\phi_1 + r_2 b\sin\phi_2)\sin\omega t \qquad (6.73)$$

where $a = \sin\alpha\cos\beta$ and $b = \cos\alpha$.

The correlation coefficient ρ can be written as

$$\rho = \left(\frac{\tan^2(\alpha)\cos^2(\beta) - \Gamma}{\tan^2(\alpha)\cos^2(\beta) + \Gamma}\right)^2 \qquad (6.74)$$

where

$$\Gamma = \frac{\langle R_2^2\rangle}{\langle R_1^2\rangle} \qquad (6.75)$$

where

$$R_1 = \sqrt{r_1^2 a^2 + r_2^2 b^2 + 2r_1 r_2 ab\cos(\phi_1 + \phi_2)} \qquad (6.76)$$

$$R_2 = \sqrt{r_1^2 a^2 + r_2^2 b^2 - 2r_1 r_2 ab\cos(\phi_1 + \phi_2)} \qquad (6.77)$$

Here, Γ is the cross polarization discrimination of the propagation path between a mobile and a base station.

The correlation coefficient is determined by three factors: polarization angle; offset angle from the main beam direction of the diversity antenna and the cross polarization discrimination. The correlation coefficient generally becomes higher as offset angle β becomes larger. Also, ρ generally becomes lower as polarization angle α increases. This is because the horizontal polarization component becomes larger as α increases.

Because antenna elements V_1 and V_2 are polarized at $\pm\alpha$ to the vertical, the received signal level is lower than that received by a vertically polarized antenna. The average value of signal loss L, relative to that received using vertical polarization is given by

$$L = a^2/\Gamma + b^2 \tag{6.78}$$

The results of practical experiments carried out using polarization diversity [Koz85] show that polarization diversity is a viable diversity reception technique.

6.10.5 Frequency Diversity

Frequency diversity transmits information on more than one carrier frequency. The rationale behind this technique is that frequencies separated by more than the coherence bandwidth of the channel will not experience the same fades [Lem91]. Theoretically, if the channels are uncorrelated, the probability of simultaneous fading will be the product of the individual fading probabilities (see equation (6.58)).

Frequency diversity is often employed in microwave line-of-sight links which carry several channels in a frequency division multiplex mode (FDM). Due to tropospheric propagation and resulting refraction, deep fading sometimes occurs. In practice, *1:N protection switching* is provided by a radio licensee, wherein one frequency is nominally idle but is available on a stand-by basis to provide frequency diversity switching for any one of the N other carriers (frequencies) being used on the same link, each carrying independent traffic. When diversity is needed, the appropriate traffic is simply switched to the backup frequency. This technique has the disadvantage that it not only requires spare bandwidth but also requires that there be as many receivers as there are channels used for the frequency diversity. However, for critical traffic, the expense may be justified.

6.10.6 Time Diversity

Time diversity repeatedly transmits information at time spacings that exceed the coherence time of the channel, so that multiple repetitions of the signal will be received with independent fading conditions, thereby providing for diversity. One modern implementation of time diversity involves the use of the

UNIT-III(B)
NON CO-CHANNEL INTERFERENCE

ADJUSCENT CHANNEL INTERFERENCE

Adjacent channel interference can be eliminated on the basis of the channel assignment, the filter characteristics and be reduction of near-end-far-end interference. Adjacent channel interference is a board term. It includes next channel interference (the channel next to the operating channel) and neighboring channel interference (more than one channel away from the operating channel). Adjacent channel interference can be reduced by the frequency assignment.

NEXT CHANNEL INTERFERENCE

Next channel interference in an AMPS system affecting a particular mobile unit cannot be caused by transmitters in the common cell site but must originate at several other cell sites. This is because any channel combiner at the cell site must combine the selected channels. Normally 21 channels(630 kHz) away, or at least 8 or 10 channels away from the desired one.therefore,next channel interference will arrive at the mobile unit from other cell sites if the system is not designed property .also a mobile unit initiating a call on a control channel in cell may cause interference with the next control channel at another cell site. The methods for reducing this next channel interference use the receiving end. The channel filter characteristics are a 6dB/oct slope in the voice band and a 24 dB/oct falloff outside the voice band region. If Next channel signal is stringer than 23 Db, it Will interfere with the desired signal. The filter with a sharp falloff slope can help to reduce all the adjacent channel interference, including the next channel interference. The same consideration is applied to digital systems.

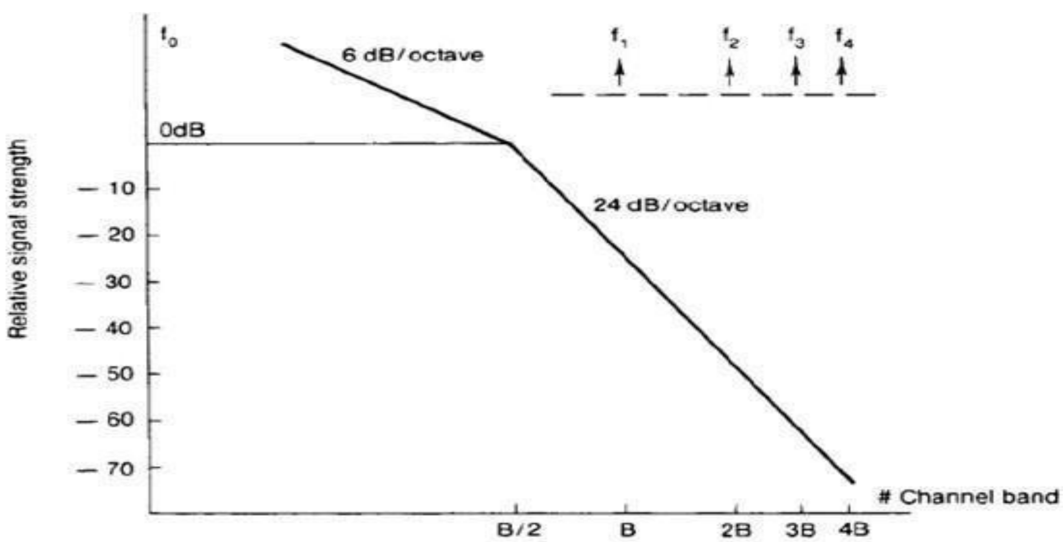


FIGURE 10.3 Characteristics of channel-band filter.

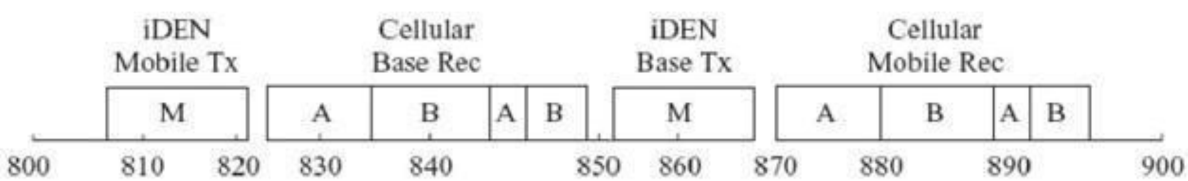
NEIGHBORING CHANNEL INTERFERENCE

The channels that are several channels away from next channel will cause interference with the desired signal. Usually, a fixed set of serving channels is assigned to each cell site. If all the channels are simultaneously transmitted at one cell site antenna; a sufficient amount of band isolation between channels is required for a multichannel combiner to reduce products. This requirement is no different from other non mobile radio systems. Assume that band separation requirements can be resolved, for example, by using multiple antennas instead of one antenna at the cell site. There will be no intermodulation products. A truly linear broadband amplifier can realize this idea. However, it is a new evolving technology.

Another type of adjacent channel interference is unique to the mobile radio system. In the mobile radio system, most mobile units are in motion simultaneously. Their relative positions change from time to time. In principle, the optimum channel assignments that avoid adjacent channel interference must also change from time to time .one unique station that causes adjacent channel interference in mobile radio system.

TRANSMITTING AND RECEIVING CHANNELS INTERFERENCE

In FDMA and TDMA systems, the transmitting channels and receiving channels have to be separated by a guard band mostly 29MHz.it is because the transmitting channels are so strong that they can mask the weak signals received from receiving channels. The duplexer can only provide 30 dB to 40dB isolation. The band isolation is the other means to reduce the interference.



		Mobile Tx	Base Tx
A	Cellular band A	824-835, 845-846.5 MHz	869-880, 890-891.5 MHz
B	Cellular band B	835-845, 846.5-849 MHz	880-890, 891.5-894 MHz
	iDEN	806-821 MHz	851-866 MHz

FIGURE 10.4 Cellular and iDEN spectrum in 800 MHz.

INTERFERENCE FROM ADJACENT SYSTEMS

The frequency bands allocated between AMPS and iDEN in 800 MHz systems are shown in Fig 10.4 in 1993. IDEN transmitted in the band 851-866 MHz , using several broad band amplifiers to cover this band. The IM(2A- B)generated from the nonlinear amplifier interfered with the cellular base received signals. Then , the broadband amplifiers were removed.

NEAR-END–FAR-END INTERFERENCE

In One Cell

Because motor vehicles in a given cell are usually moving, some mobile units are close to the cell site and some are not. The close-in mobile unit has a strong signal that causes adjacent-channel interference (see Fig. 10.5*a*). In this situation, near-end–far-end interference can occur only at the reception point in the cell site.

If a separation of $5B$ (five channel bandwidths) is needed for two adjacent channels in a cell in order to avoid the near-end–far-end interference, it is then implied that a minimum separation of $5B$ is required between each adjacent channel used with one cell.

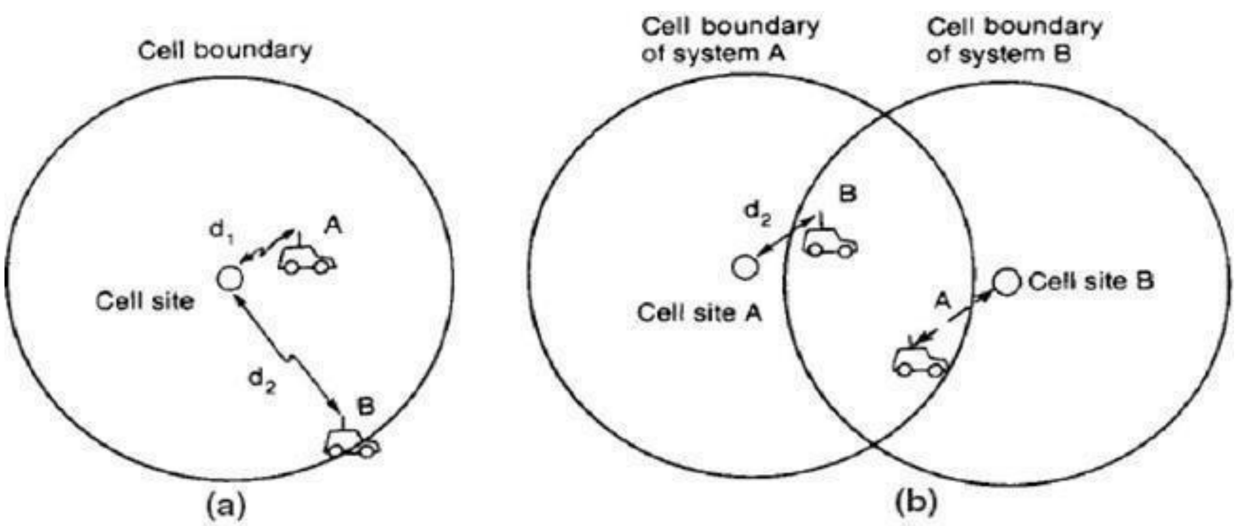


FIGURE 10.5 Near-end–far-end (ratio) interference. (*a*) In one cell; (*b*) in two-system cells.

Because the total frequency channels are distributed in a set of N cells, each cell only has $1/N$ of the total frequency channels. We denote $\{F_1\}$, $\{F_2\}$, $\{F_3\}$, $\{F_4\}$ for the sets of frequency channels assigned in their corresponding cells C_1, C_2, C_3, C_4.

The issue here is how can we construct a good frequency management chart to assign the N sets of frequency channels properly and thus avoid the problems indicated above. The following section addresses how cellular system engineers solve this problem in two different systems.

In Cells of Two Systems

Adjacent-channel interference can occur between two systems in a duopoly-market system. In this situation, adjacent-channel interference can occur at both the cell site and the mobile unit.

For instance, mobile unit A can be located at the boundary of its own home cell A in system A but very close to cell B of system B as shown in Fig 10.5*b*. The other situation would occur if mobile unit B were at the boundary of cell B of system B but very close to cell A of system A. Following the definition of near-end–far-end interference

, the solid arrow indicates that interference may occur at cell site A and the dotted arrow indicates that interference may occur at mobile unit A. Of course, the same interference will be introduced at cell site B and mobile unit B.

Thus, the frequency channels of both cells of the two systems must be coordinated in the neighborhood of the two-system frequency bands. This phenomenon causes a great concern as indicated in the additional frequency-spectrum allocation charts in Fig. 10.6 as an example.

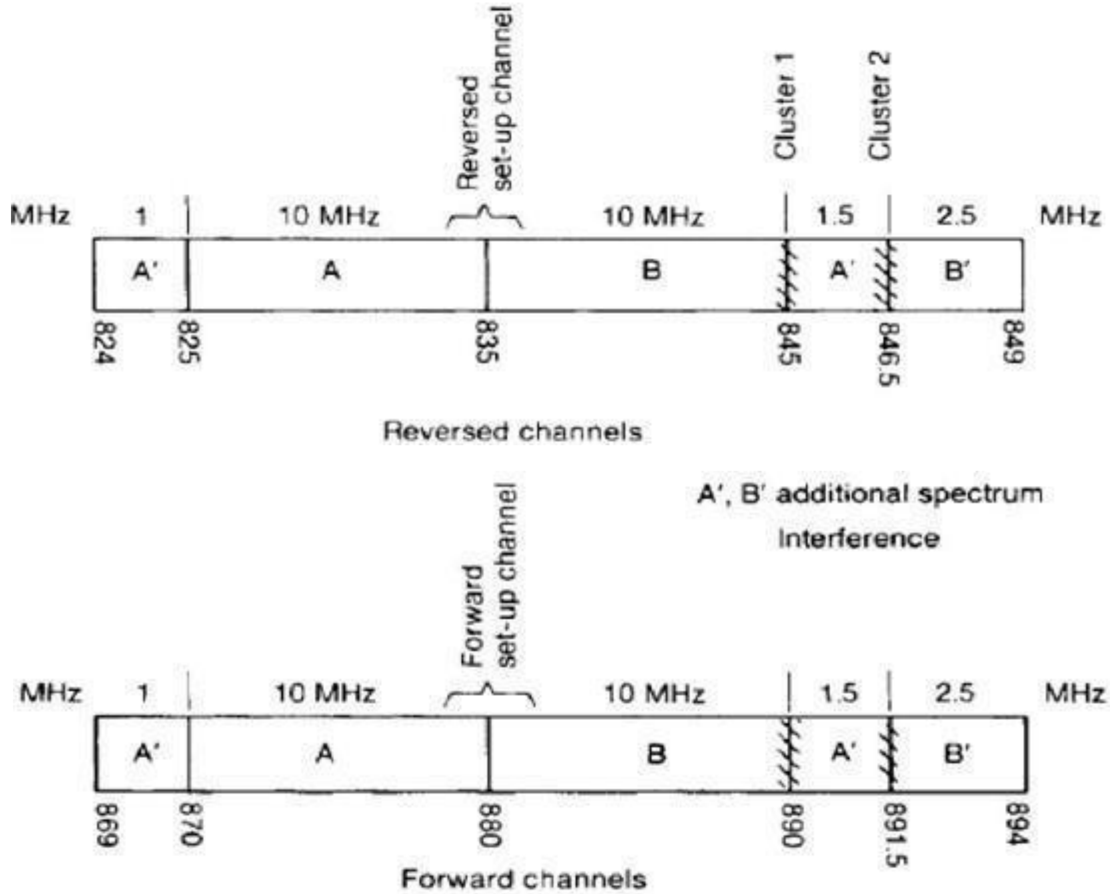


FIGURE 10.6 Spectrum allocation with new additional spectrum.

The two causes of near-end–far-end interference of concern here are

1. *Interference caused on the set-up channels.* Two systems try to avoid using the neighborhood of the set-up channels as shown in Fig. 10.6.
2. *Interference caused on the voice channels.* There are two clusters of frequency sets as shown in Fig. 10.6 that may cause adjacent-channel interference and should be avoided. The cluster can consist of 4 to 5 channels on each side of each system, that is, 8 to 10 channels in each cluster. The channel separation can be based on two assumptions.

a. *Received interference at the mobile unit.* The mobile unit is located away from its own cell site but only 0.25 mi away from the cell site of another system.

b. *Received interference at the cell site.* The cell site is located 10 mi away from its own mobile unit but only 0.25 mi from the mobile unit of another system.

CROSS TALK-

When the cellular radio system was designed, the system was intended to function like a telephone wire line. A wire pair serves both directions of traffic at the line transmission. In a mobile cellular system there is a pair of frequencies, occupying a bandwidth of 60 kHz, which we simply call a "channel." A frequency of 30 kHz serves a received path, and the other 30 kHz accommodates a transmitted path.

Because of paired-frequency (as a wire pair) coupling through the two-wire–four-wire hybrid circuitry at the telephone central office, it is possible to hear voices in both frequencies (in the frequency pair) simultaneously while scanning on only one frequency in the air. Therefore, just as with a wire telephone line, the full conversation can be heard on a single frequency (either one of the two). This phenomenon does not annoy cellular mobile users; when they talk they also listen to themselves through the phone receiver. They are not even aware that they are listening to their own voices.

This unnoticeable cross-talk phenomenon in frequency pairs has no major impact on both wire telephone line and cellular mobile performance. But when real cross talk occurs it has a larger impact on the cellular mobile system than on the telephone line, because the amount of cross talk could potentially be doubled since cross talk occurring on one frequency will be heard on the other (paired) frequency. Cross talk occurring on the reverse voice channel (RVC) can be heard on the forward voice channel (FVC), and cross talk occurring on the forward voice channel can be heard on the reverse channel. Therefore, the cross-talk effect is twofold. A number of situations are conducive to cross talk.

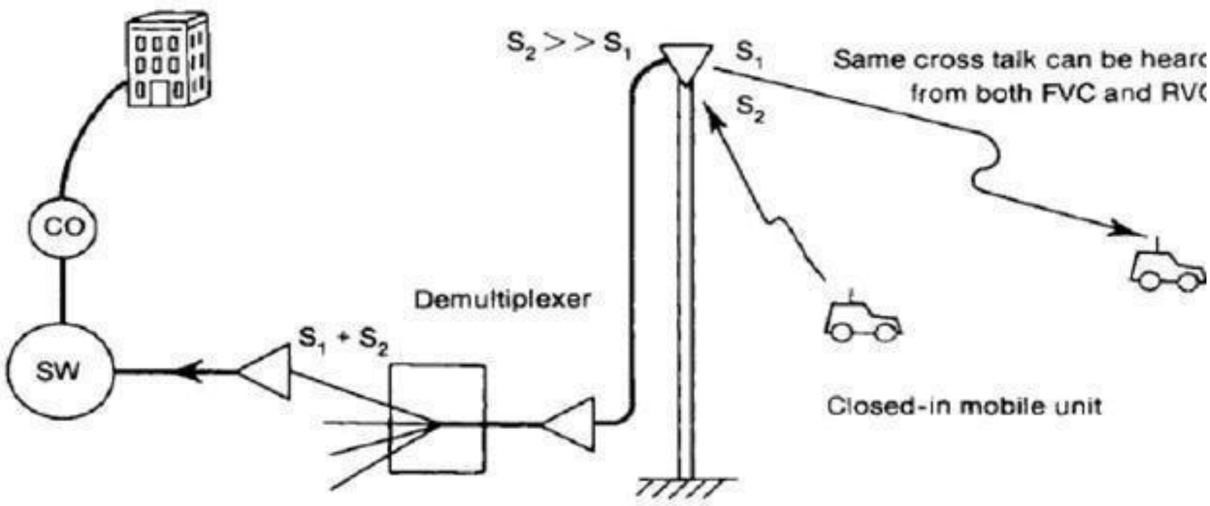


FIGURE 10.10 Cross-talk phenomenon.

Near-end mobile unit. Cross talk can occur when one mobile unit (unit A) is very close to the cell site and the other (unit B) is far from the cell site. Both units are calling to their land-line parties as shown in Fig. 10.10. The near-end mobile unit has a strong signal such that the demultiplexer cannot have an isolation (separation) of more than 30 dB. Then the strong signal can generate strong cross talk while the received signal from mobile unit B is 30 dB weaker than signal A.

Near-end mobile units can belong to one system or to another (foreign) system. If the foreign system units are operating in the new allocated spectrum channels, cross talk can occur. When the mobile unit is close to the cell site and the cell site is capable of reducing the power of the mobile unit, the near-end mobile interference can be reduced.

If the operating frequencies of both home system units and foreign system units are in the new allocated spectrum channels and the isolation of the multicoupler (demultiplexer) could be only 30 dB, cross talk would occur in the two interfering clusters of channels (Fig. 10.10) and could not be controlled by the system operator.

Close-in mobile units. When a mobile unit is very close to the cell site and if the reception at the cell site is greater than -55 dBm, the channel preamplifier at the cell site can become saturated and produce IM as a result of the nonlinear portion of the amplification. These IM products are the spurious (unwanted frequency) signal that leaks into the desired signal and produces cross talk. Also, as mentioned previously, the same cross talk can be heard from both the forward and reverse voice channels.

Cochannel cross talk. The cochannel interference reduction ratio q should be as large as possible to compensate for the cost of site construction and the limitation of available channels at each cellular site. There are other ways to increase q, as mentioned in Chap. 9. An adequate system design will help to reduce the cochannel cross talk.

The channel combiner. The signal isolation among the forward voice channels in a channel combiner is 17 dB.[4] The loss resulting from inserting the signal into the combiner is about 3 dB. The requirement of IM product suppression is about 55 dB. If one outlet is not matched well, the signal isolation is less than 17 dB. Therefore, for each channel an isolator is installed to provide an additional 30-dB of isolation with a 0.5-dB insertion loss. This isolator prevents any signal from leaking back to the power amplifier (see Sec. 10.7.1). Spurious signals can be cross-coupled to this weak channel while transmitting. This kind of cross-coupled interference can be eliminated by routinely checking impedance matching at the combiner.

Telephone-line cross talk. Sometimes cross talk can result from cable imbalance or switching error at the central office and be conveyed to the customer through the telephone line. Minimizing this type of cross talk should be given the same priority as reducing the number of call drops,

Effects on coverage and interference by power decrease and Antenna height decrease

Power Decrease

As long as the setup of the antenna configuration at the cell site remains the same, and if the cell-site transmitted power is decreased by 3 dB, then the reception at the mobile unit is also decreased by 3 dB. This is a one-on-one (i.e., linear) correspondence and thus is easy to control.

Antenna Height Decrease

When antenna height is decreased, the reception power is also decreased. However, the formula

$$\text{Antenna height gain (or loss)} = 20 \log \frac{h'_{e_1}}{h_{e_1}}$$

is based on the difference between the old and new effective antenna heights and not on the actual antenna heights. Therefore, the effective antenna height is the same as the actual antenna height only when the mobile unit is traveling on flat ground. It is easy to decrease antenna height to control coverage in a flat-terrain area. For decreasing antenna height in a hilly area, the signal-strength contour shown in Fig. 10.12*a* is different from the situation of power decrease shown in Fig. 10.12*b*. Therefore, a decrease in antenna height would affect the coverage; thus, antenna height becomes very difficult to control in an overall plan. Some area within the cell may have a high attenuation while another may not.

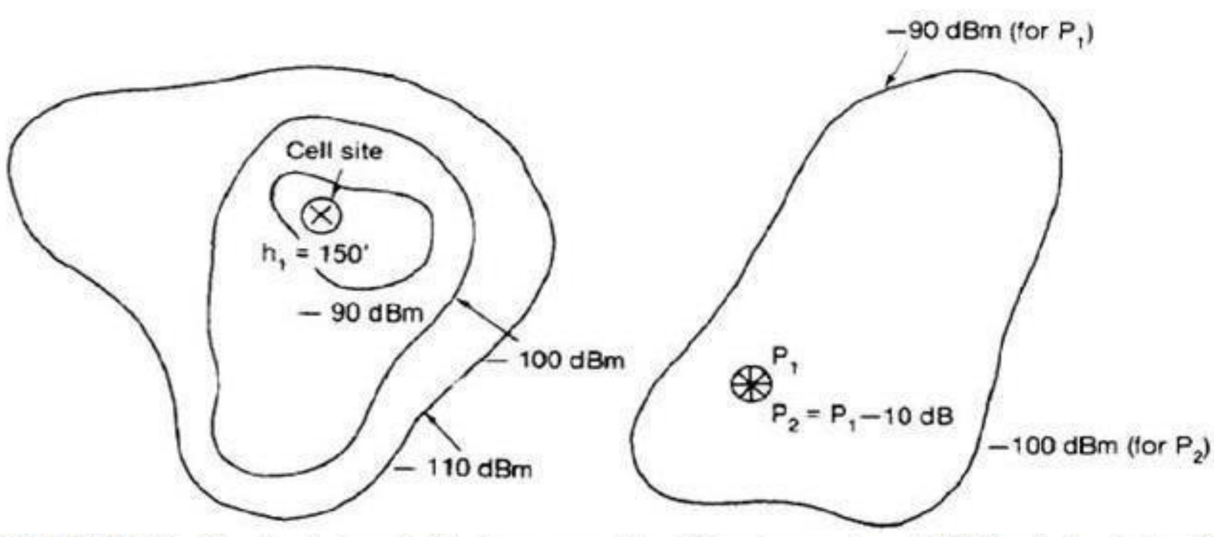


FIGURE 10.12 The signal-strength effect as measured by different parameters. (*a*) Different signal-strength contours. (*b*) Signal-strength changes with power changes.

the receiver and the noise N_{cm} introduced from the cellular mobile environment.

$$NF = \sqrt{N_f^2 + N_{cm}^2} \qquad dB$$

N_{cm} can either increase or decrease, depending on the system design. The earlier data indicate that N_{cm} can be neglected for 900-MHz curves.[7,8] If we now introduce a safety factor and let $N_{cm} = 6$ dB then

$$NF = \sqrt{(9)^2 + (6)^2} = 11 \text{ dB}$$

The total noise level is $N = kTB + NF = -118$ dBm. Because the required C/N is 18 dB, the lowest acceptable signal level is -100 dBm (-32 dBμ), which is 7 dB lower than -93 dBm (39 dBμ). In reality, the cell boundary or the handoff is based on the voice quality, that is, $C/N = 18$ dB or a level of -100 dBm; therefore, the FCC cell boundary of 39 dBμ or -93 dBm is 7 dB higher than the level provided by the system. Thus a cell boundary of 32 dBμ or -100 dBm proved to be sufficient for cellular coverage.

The two main advantages of using a 32-dBμ level (see Fig. 7.14) are that (1) fewer cell sites would be needed to cover a growth area and (2) less interference would be effected at the boundaries. A 32-dB boundary for cells in either boundary of a metropolitan statistical area (MSA) or a rural service area (RSA) is a proper operation, as opposed to a 39-dBμ boundary which is an artificial value.

7.7 Effects on Cell-Site Components

7.7.1 Channel combiner

A fixed-tuned channel combiner at the transmitting side.[4] A channel combiner is installed at each cell site. Then all the transmitted channels can be combined with minimum insertion loss and maximum signal isolation between channels. Of course, we can eliminate the channel combiner by letting each channel feed to its own antenna. Then a 16-channel site will have 16 antennas for operation. It is an economical and a physical constraint.

A conventional combiner has a 16-channel combined capacity based on the frequency subset of 16 channels, and it causes each channel to lose 3 dB from inserting the signal through the combiner. The signal isolation is 17 dB because each channel is 630 kHz or 21 channels apart from neighboring channels (Fig. 7.15*a*). The intermodulation at the multiplexer is controlled by ferrite isolators, which provide a 30-dB reverse loss. The intermodulation (IM) products are at least 55 dB down from the desired signals. Therefore, the IM will not affect channels within the transmitted band design from this.

Each cable fed into a combiner must be properly shielded. Because it is a nonlinear device, undesired signal leakage into another channel would occur before the combiner can produce the IM products, which would in turn, produce cross-coupled interference. Therefore, proper shielding and impedance match are very important. Fixed-tuned combiners are tuned to match the impedances of a set of fixed frequencies which are assigned to a combiner.

A frequency-agile combiner.[11] This combiner is capable of returning to any frequency by remote control in real time. The remote control device is a microprocessor. The combiner is a waveguide-resonator combiner with a tuning bar in each input waveguide as shown in Fig. 7.15*b*. The bar is mechanically rotated by a motor, and the voltage standing-wave ratio (VSWR) can be measured when the motor starts to turn. The controller received an optimum reading after a full turn and is stopped at that position by the controller. The controller also has a self-adjusting potential. This combiner can be used when a dynamic frequency assignment is applied. In many cases, it is preferable to redistribute the frequency channels to avoid prominent interference in certain areas. To use this kind of combiner, cell-site transceivers should also be able to change their operating frequencies, which are controlled by the MTSO, accordingly. This kind of combiner can also be designed to be tuned electronically.

A ring combiner.[12] A ring combiner is used to combine two groups of channels into a single output. The insertion loss is 3 dB, and the signal

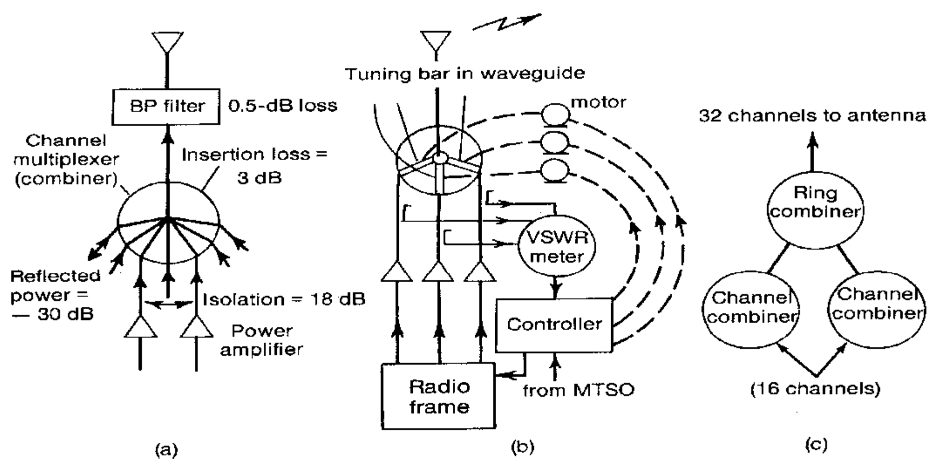


Figure 7.15 Different kinds of channel combiners. (*a*) Fixed-tuned combiner, (*b*) tunable combiner, (*c*) ring combiner.

isolation between channels is 35 to 40 dB. The function of a ring combiner is to combine two 16-channel combiners into one 32-channel output. Therefore, all 32 channels can be used by a single transmitting antenna. If a cell site has two antennas, up to 64 radio channels can be installed in it.

If all the channel-transmitted powers are low, it is possible to combine more than 32 channels by using two or three ring combiners before feeding them into one transmitting antenna. The total allowed transmitted power is a limiting factor. Some ring combiners have a 600-W power limitation. The use of ring combiners reduces adjacent channel separation. If two 16-channel regular combiners are combined with a ring combiner, the adjacent-channel separation at the ring combiner output can be 315 kHz, even though the adjacent-channel separation of each regular combiner is 630 kHz. It is simply a frequency offset of 315 kHz between two regular combiners.

7.7.2 Demultiplexer at the receiving end

A demultiplexer is used to receive 16 channels from one antenna. The demultiplexer is a filter bank as shown in Fig. 7.16. Then each receiving antenna output passes through a 25-dB-gain amplifier to a demultiplexer. The demultiplexer output has a 12-dB loss from the split of 16 channels.

$$\text{Split loss} = 10 \log 16 = 12 \text{ dB}$$

and the IM product at the output of the demultiplexer should be 65 dB down.[4] The two space-diversity antennas each connect to an umbrella filter (block A or B band filter) and have a 55-dB rejection from the other system band. If the undesired mobile unit is close to the cell site, then the preamplifier becomes saturated and generates IM at the output of the amplifier; these IM products (frequencies) could be felt in one of the weak incoming signals. This situation can lead to cross talk (see Sec. 7.4) which can be heard from both ends of the link because of a unique characteristic of cellular channels (see Sec. 7.5).

7.7.3 SAT tone

General description. The major function of a supervisory audio tone (SAT) is to ensure that a SAT tone is sent out at the cell site, is received by the mobile unit on a forward voice channel, is converted on a corresponding reverse voice channel, and is then sent back to the cell site within 5 s. If the time out is more than 5 s, the cell site will terminate the call.

Every cell site has been assigned to one of three SAT tones. The assignment of three SAT tones in a system is shown in Fig. 7.17. The cells have the same SAT tones, and the same channels are separated by $\sqrt{3}D$, which is farther than the cochannel distance D. Therefore, a receiver located at either the cell site or at the mobile unit and receiving the same frequency with different SAT tones will terminate the call.

Characteristics of SAT. There are three SAT tones, 5970 H, 6000 H, and 6030 Hz, spaced 30 Hz apart. They are narrowband frequency-modulated (FM) with a deviation of $f_\Delta = 2$ kHz. The modulation index is $\beta = 1/3$. Let the SAT tone signal be

$$x(t) = A_m \cos \omega_m t \tag{7.7-1}$$

and the modulated carrier is

$$x_c(t) = A_c \cos(\omega_c t + \beta \sin \omega_m t) \tag{7.7-2}$$

where $\beta = (A_m f_\Delta / f_m)$. Let the amplitude modulation $A_m = 1$; thus,

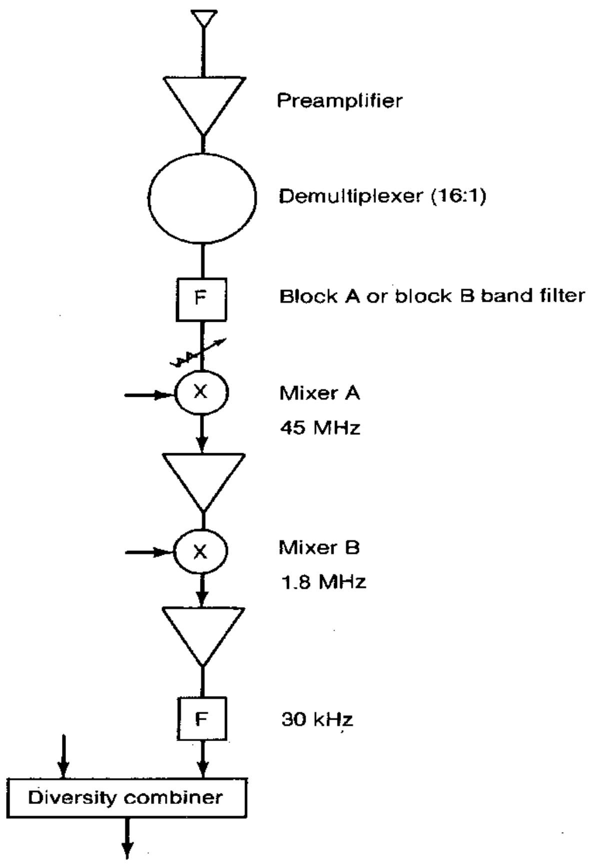


Figure 7.16 A typical cell-site channel receiver.