Chapter

13

System Evaluations

13.1 Performance Evaluation

13.1.1 Blockage

There are two kinds of blockage: set-up channel blockage and voice-channel blockage.

Set-up channel blockage B_1. Information regarding set-up channel blockage cannot be obtained at the cell site because the mobile unit will be searching for the busy/idle bit of a forward set-up channel in order to set up its call. If the busy bit does not change after 10 call attempts in 1 s, a busy tone is generated, and no mobile transmit takes place. In another case the mobile transmit takes place as soon as the idle bit is shown. Several initiating cells can intercollide at the same time. When it occurs, the mobile unit counts it as one seizure attempt. If the number of seizure attempts exceeds 10, then the call is blocked. This kind of blockage can be detected only by mobile phone users. If the occurrence of blockage of the system is in doubt, each of the three specified set-up channels can be assigned in each of the three sectors of a cell, and the total number of incoming calls among the three sectors can be compared with that from a single set-up channel (omni). It should be determined whether there is a difference between two call-completion numbers, one from a single set-up channel and the other from three set-up channels. This is one way to check the blockage if the single set-up channel seems too busy. The set-up channel blockage should be at least less than half of the specified blockage (usually 0.02) in the mobile cellular system.

If all the call-attempt repeats are independent events, then the resultant blocking probability B_1 after n attempts is related to the blocking probability of the single call attempt B, as

$$B_1 = 1 - (1 - B) \sum_{i=0}^{n} B^i = B^n \qquad (13.1\text{-}1)$$

Example 13.1 Assume that the blocking probability of a set-up channel is .005, and the holding time at the set-up channel is 175 ms per call. There is only one channel; then the offered load (from Appendix 1.1) a is .005. Thus the number of set-up calls being handled is

$$C = \frac{.005 \times 3600 \times 1000}{175} = 120 \text{ calls} \qquad \text{(one call attempt)}$$

Example 13.2 All parameters are the same as in Example 13.1, except that the offered load a changes to $a = 0.02$. Then the number of set-up calls is

$$C = \frac{.02 \times 3600 \times 1000}{175} = 480 \text{ calls} \qquad \text{(one call attempt)}$$

Example 13.3 Given the number of set-up calls per hour, find the blocking probability B_1 after 10 call attempts in 1 s.

Consider the following cases.

Case 1. Assume that there are two set-up calls per second or 7200 calls per hour. Since each set-up call takes 175 ms, the offered load a is

$$a = \frac{175 \times 2}{1000} = .35$$

The blocking probability B (see Appendix 1.1) is $B = .25$ (assuming one call attempt).

Since the average interval for each attempt is 100 ms, 10 attempts have to be completed in 1 s. It is a kind of conditional probability problem. In the worst case, a mobile unit has to fail the tenth call attempt before giving up. During this period, because of the failure of all call attempts, the two set-up calls from other mobile units should have been successful with a probability of 1. The length of two set-up calls is 350 ms, which is roughly the time interval required for four attempts; i.e., these four attempts are definitely blocked with a blocking probability of 1 and should not be counted as attempts. Therefore, only six attempts count. Using Eq. (13.1-1), we obtain

$$B_1 = (.25)^6 = .00012$$

which is quite low and, of course, acceptable.

Case 2. Assume that there are three set-up calls per second or 10,800 calls per hour. Then

$$a = \frac{175 \times 3}{1000} = .525 \quad \text{(offered load)}$$

$$B = .342 \qquad \text{(blocking probability; see Appendix 1.1)}$$

Since three set-up calls take 525 ms, roughly six out of ten attempts are definitely blocked following the same argument stated in case 1. Only four attempts count, then the resultant blocking probability is

$$B_t = (.342)^4 = .013$$

which is too high for the set-up channel.

Voice-channel blockage B_2. Voice-channel blockage can be evaluated at the cell site. When all calls come in, some are refused for service because there are no available voice channels. Suppose that we are designing a voice channel blockage to be .02. On this basis, $B_2 = .02$, and after determining the holding time per call[1] and roughly estimating the total number of calls per hour at the site,[2] we can find the number of radios required.

Example 13.4 Assume that 2000 calls per hour are anticipated. The average holding time is 100 s per call, and the blocking probability is .02 (2 percent). Then the offered load is

$$a = \frac{2000 \times 100 \text{ s}}{60 \times 60 \text{ s}} = 55.5 \text{ erlangs}$$

Use $a = 55.5$ and $B_2 = 0.02$ to find $N = 66$ channels required (refer to Appendix 1.1).

The actual blocking probability data must be used to check the outcome from the Erlang B model (Appendix 1.1). Although the difference can be up to 15 percent, the Erlang B model is still considered as a good model for obtaining useful estimates.

End-office trunk blockage B_3. The trunks connecting from the MTSO to the end office can be blocked. This usually occurs when the call traffic starts to build up and the number of trunks connected to the end office becomes inadequate. Unless this corrective action is taken, the blockage during busy periods increases. An additional number of trunks could be provided at the end office when needed.

The total blockage B_t. As the total call blockage is the result of all three kinds of blockage, the total blockage is

$$B_t = B_1 + B_2 (1 - B_1) + B_3 (1 - B_1)(1 - B_2)$$
$$= 1 - (1 - B_1)(1 - B_2)(1 - B_3) \qquad (13.1\text{-}2)$$

Example 13.5 Assume that $B_1 = .01$ and $B_2 = B_3 = .02$. Then the total blockage is

$$B_t = .01 + .0198 + .0194 = .0492 \approx 5\%$$

The result in Example 13.5 indicates that even when each individual blockage (i.e., B_1, B_2, and B_3) is small, the total blockage becomes very large. Therefore, the resultant blockage is what we are determining.

13.1.2 Call drops (dropped-call rate)

Call drops are defined as calls dropped for any reason after the voice channel has been assigned. Sometimes call drops due to weak signals are called *lost calls*. The dropped-call rate is partially based on the handoff-traffic model and partially based on signal coverage. The calculation of dropped call rate is shown in Sec. 9.10 through Sec. 9.12. The evaluation of call drops is stated in this section.

The handoff traffic model. A new handoff cell site treats handoffs the same way as it would an incoming call. Therefore, the blockage for handoff calls is also $B = .02$. Some MTSO systems may give priority to handoff calls rather than to incoming calls. In this case the blocking probability will be less than .02.

A warning feature can be implemented when the call cannot be handed off and may be dropped with high probability, enabling the customer to finish the call before it is dropped. Then the dropped-call rate can be reduced.

The loss of SAT calls. If the mobile unit does not receive a correct SAT in 5 s, the mobile-unit transmitter is shut down. If the mobile unit does not send back a SAT in 5 s, the transmitter at the cell site is shut down. In both cases the call is dropped. If the correct SAT cannot be detected at the cell site, as in cases of strong interference, then (1) the SAT can be offset by more than 15 Hz (see section entitled "The total dropped-call rate," below) or (2) the SAT tone generator in the mobile unit may not produce the desired tone.

Calculation of SAT interference conditions. The desired SAT is $\cos w_1 t$, and the undesired SAT is $\rho \cos w_2 t$. When $\rho \ll 1$, the SAT detector at the cell site can easily detect w_1. When ρ is greater and starts to approach 1, SAT interference occurs. The following analysis shows the degree of the interference due to the value of ρ.

$$\cos w_1 t + \rho \cos w_2 t = A(t) \cos \theta(t) \tag{13.1-3}$$

where

$$A(t) = \sqrt{1 - \rho^2 + 2\rho \cos (w_1 - w_2)t} \tag{13.1-4}$$

$$\psi(t) = w_2 t - w_1 t$$

$$\theta = w_1 t + \tan^{-1} \frac{\rho \sin \psi(t)}{1 + \rho \cos \psi(t)} \tag{13.1-5}$$

$$w = \frac{d\theta}{dt} = w_1 + \frac{w_2 - w_1}{(\{1 + \rho \cos \psi(t)\}/\{\rho[\rho + \cos \psi(t)]\}) + 1} \tag{13.1-6}$$

Let $\cos \psi(t) = 1$ in Eq. (13.1-6), the extreme condition of w which is the offset frequency from a desired SAT.

$$w = w_1 + \left(1 + \frac{1}{\rho}\right)^{-1} (w_2 - w_1) \tag{13.1-7}$$

For

$$\rho = \begin{cases} .3 & w = w_1 + .22(w_2 - w_1) \\ .5 & w = w_1 + .333(w_2 - w_1) \\ .75 & w = w_1 + .45(w_2 - w_1) \end{cases} \tag{13.1-8}$$

If $w_2 - w_1 = 30$ Hz, for two adjacent SATs

$$\rho = \begin{cases} .5 & \omega = \omega_1 \pm 9.95 & \text{(acceptable)} \\ .75 & \omega = \omega_1 \pm 12.9 & \text{(marginal)} \end{cases}$$

If $\omega_2 - \omega_1 = 60$ Hz, for two ends of SATs

$$\rho = \begin{cases} .3 & \omega = \omega_1 \pm 13.8 & \text{(marginal)} \\ .5 & \omega = \omega_1 \pm 19.9 & \text{(unacceptable)} \\ .75 & \omega = \omega_1 \pm 25.8 & \text{(unacceptable)} \end{cases}$$

Adjacent SATs cannot interfere with the desired SAT for an undesired SAT level below $\rho = .75$ (meaning a level of -2.5 dB). However, when two SATs are not adjacent to each other, the undesired SAT level should at least be lower than $\rho = 0.3$ (-10.5 dB) in order for no interference to occur.

Unsuccessful complete handoffs. Because of the limitations in processor capacity, the duration of the handoff process may occasionally be too long, and the mobile unit may not be informed of a new channel to be handed off.

The total dropped-call rate. Assume that the handoff blocking is B_4, the probability of lost SAT calls is B_5, and the probability of an unsuccessful complete handoff is B_6. Then the total drop call rate is

$$B_d = B_4 + B_5 (1 - B_4) + B_6 (1 - B_4)(1 - B_5) \qquad (13.1\text{-}9)$$

Usually, the dropped-call rate should be less than 5 percent.

13.1.3 Voice quality

It is very important that the voice quality of a channel be tested by subjective means. Some engineers try to use the signal-to-noise-plus-distortion ratio (SINAD) to evaluate voice quality. Although SINAD is an objective test, using it to test voice quality may result in misleading conclusions. Worst of all, engineers are always proud of the apparatus they have designed, and they tend to ignore others' opinions. To serve the public interest, we must survey consumers for their opinions.

Evaluation of system performance based on a subjective test can be used to set performance criteria. As was discussed in Chap. 1, 75 percent of cellular phone users report that voice quality is good (CM4; i.e., circuit merit 4) or excellent (CM5) over 90 percent of the service area. These numbers can vary depending on how well the service is performed—in other words, the cost. First we must know what kind of service we are providing to the public. We should then let the customers judge the voice quality.

A typical curve of subjective tests was shown in Fig. 7.1*b* for different conditions. For this particular run, the mobile speed is 56 km/h (35 mi/h).

13.1.4 Performance Evaluation

Often we encounter a situation where two systems are being installed in two different areas; one system is deployed in a flat area where the average measured bit error rate (BER) is low at the cell site on the basis of bit-stream data received from mobile-units, and the other system is deployed in a hilly area where BER is relatively high. Of course, if the same degree of skill was used to install both systems, the system in the hilly area would otherwise be inferior and would provide the poorer performance. But if each system claims to be better than the other, how can we judge these two systems fairly? One way is to compensate for variations in performance due to geographic location.[2] Assume that system A is deployed in a flat terrain and system B, in a hilly area. Both systems use the same kind of antennas and run the measured data with the same kind of mobile units. Then the handicap of building a system in a hilly area can be compensated for before comparing this system's performance with others.

Another way is to set the BER to, say, 10^{-3} and find the percentage of areas where the measured BER is greater than 10^{-3} in relation to the area of the whole system, then compare the percentages from the two different systems in two different areas.

Here the two pieces of raw field-strength recorded data in the area should be received from two mobile units moving around in two systems, respectively. From these data, the two local means (the envelope of the raw field-strength data) can be obtained. Then the two cumulative probability distributions (CPD), $P(x < X)$ of two local means, can be plotted. Let us first normalize the average power at a 50 percent level because the local means have a log-normal distribution, and the differences in transmitted power in the two systems are factored out. There will be two straight lines on log-normal scale paper (see Fig. 13.1). The standard deviations of two log-normal curves can be found from Fig. 13.1 by determining a level X where $P(x \leq X) = 90$ percent.

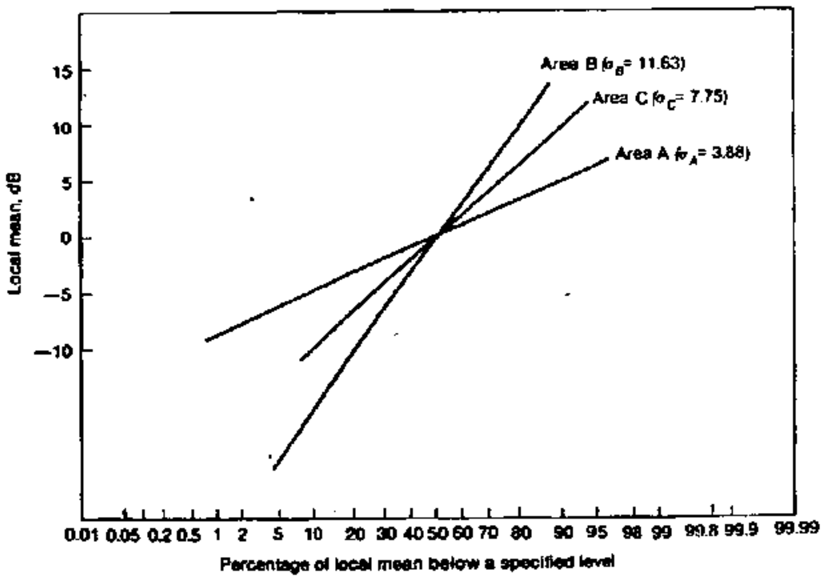


Figure 13.1 Different local mean statistics in different areas.

$$X = \begin{cases} 5\ \text{dB} & \text{for system A} \\ 15\ \text{dB} & \text{for system B} \end{cases}$$

A normal distribution can be found from published mathematical tables indicating that $P(x \leq 1.29) = 90$ percent, where

$$x \leq \frac{X - m}{\sigma} = \frac{X}{\sigma} = 1.29 \tag{13.1-10}$$

where $m = 0$ dB and σ is the standard deviation. For

$$X = \begin{cases} 5\ \text{dB} & \sigma_A = \dfrac{5}{1.29} = 3.88\ \text{dB} \\ 15\ \text{dB} & \sigma_B = \dfrac{15}{1.29} = 11.6\ \text{dB} \end{cases}$$

The standard deviations of A and B indicate that system B covers a relatively hillier area than system A.

Example 13.6 From these measurements we find that the BER in 10 percent of area A in system A is greater than 10^{-3}. The BER in 25 percent of area B in system B is greater than 10^{-3}. The local means of two systems are shown in Fig. 13.1. We would like to judge which system has a better performance. From Fig. 13.1, 10 percent of the area in system A corresponds to 10 percent of the total signal below -5 dB. If system A is deployed in area C ($\sigma_c = 7.75$ dB), 25 percent of area C will have a BER greater than 10^{-3} below the -5-dB level. Now the variance σ_B of area B is greater than the variance σ_C of area C; thus, area B is more difficult to serve than area C. However, the measurement shows that 25 percent of area B in system B is greater than 10^{-3} below the -9-dB level. Therefore, according to the criterion BER $\gtrsim 10^{-3}$, system B proves to be a better system than system A.

13.2 Signaling Evaluation

The signaling protocols of existing systems are evaluated in this section. The signaling format of the forward control channel (FOCC) as deduced from a BCH code (63, 51) becomes a short code of (40, 28) or that of the reverse control channel (RECC) from a BCH becomes a short code of (48, 36), as described in Chap. 3. The 12 parity-check bits always remain unchanged. This BCH code can correct one error and detect two errors.

13.2.1 False-alarm rate

The false-alarm rate is the rate of occurrence of a false recognizable word that would cause a malfunction in a system. The false-alarm rate should be less than 10^{-7}. Now we would like to verify that the BCH code (40, 28) can meet this requirement. The Hamming distance d of BCH (40, 28) is 5. This means that in every different code word at least 5 out of 40 bits are different. Then the false-alarm rate FAR can be calculated as

$$FAR = p_e^d(1 - p_e)^{L-d} \tag{13.2-1}$$

where p_e is the BER and d is the length of a word in bits.

Assume that in a noncoherent frequency-shift-keying (FSK) modulation system, the average BER of a data stream in a Rayleigh fading environment is

$$\langle p_e \rangle = \frac{1}{2 + \Gamma} \qquad \text{(noncoherent FSK)} \tag{13.2-2a}$$

and the average BER of a data stream received by a differential phase-shift-keying (DPSK) modulation system in the same environment is

$$\langle p_e \rangle = ½ \left(\frac{1}{\Gamma + 1}\right) \qquad \text{(DPSK)} \tag{13.2-2b}$$

where Γ is the carrier-to-noise ratio. Let* $\Gamma = 15$ dB; then we obtain BER from Eq. (13.2-2) as

$$\langle p_e \rangle = .03 \qquad \text{(noncoherent FSK)}$$

$$\langle p_e \rangle = .015 \qquad \text{(DPSK)}$$

Substituting $p_e = .03$, which is the higher BER, into Eq. (13.2-1), we obtain

$$FAR \approx (.03)^5 = 2.43 \times 10^{-8} \leq 10^{-7}$$

This meets the requirement that $FAR < 10^{-7}$.

13.2.2 Word error rate consideration

The word error rate (WER) plays an important role in a Rayleigh fading environment. The length of a word of an FOCC is $L = 40$ and the transmission rate is 10 kbps. Then the transmission time for a 40-bit word is

$$T = \frac{40}{10{,}000} = 4 \text{ ms}$$

*The C/N ratio of a data channel can be lower than 18 dB of a voice channel.

From Sec. 1.6.6, the average duration of fades can be obtained from the following assumptions: frequency = 850 MHz, vehicle speed = 15 mi/h, and threshold level = −10 dB (10 dB below the average power level); then the average duration of fades is

$$\bar{t} = 0.33 \times \left(\sqrt{2\pi}\,\frac{V}{\lambda}\right)^{-1} = 7 \text{ ms} \qquad (13.2\text{-}3)$$

Equation (13.2-3) shows that the transmission time of one word is shorter than the average duration of fades while the vehicle speed is 15 mi/h; that is, the whole word can disappear under the fade. Therefore, redundancy schemes are introduced. From Chap. 3, the FOCC format is

200 bits	word A (40 bits × 5 times), 28 information bits	
200 bits	word B (40 bits × 5 times), 28 information bits	
10 bits	bit synchronization	
11 bits	word synchronization	
42 bits	Busy/Idle-status bits	
463 bits		56 information bits

The throughput can be obtained from

$$\frac{56}{463} = \frac{1200 \text{ bps}}{10{,}000 \text{ bps}}$$

Therefore, the throughput is 1200 bps (baseband rate).

13.2.3 Word error rate calculation

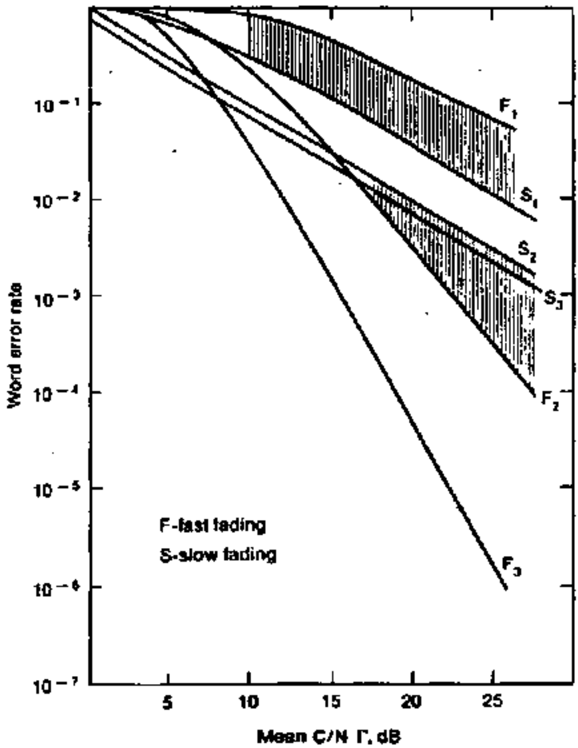
The WER can be calculated as follows. We may use a DPSK system because it has a general but simple analytic formula, more general than Eq. (13.2-2*b*)

$$\langle p_e \rangle = \tfrac{1}{2}\left(\frac{1}{\Gamma + 1}\right)^{M} \qquad (13.2\text{-}4)$$

where M is the number of diversity branches. It is difficult to obtain the WER from the correlation coefficient in the bit stream at a specific vehicle speed because the correlation coefficients of any two bits among all the bits in a word at that particular speed form a correlation coefficient matrix which is difficult to handle. Fortunately, we can find two extreme values, one at the speed $V \to \infty$ and the other at the speed $V - \to 0$. The calculations are described in detail in Ref. 3. Here we are simply illustrating the results.[4]

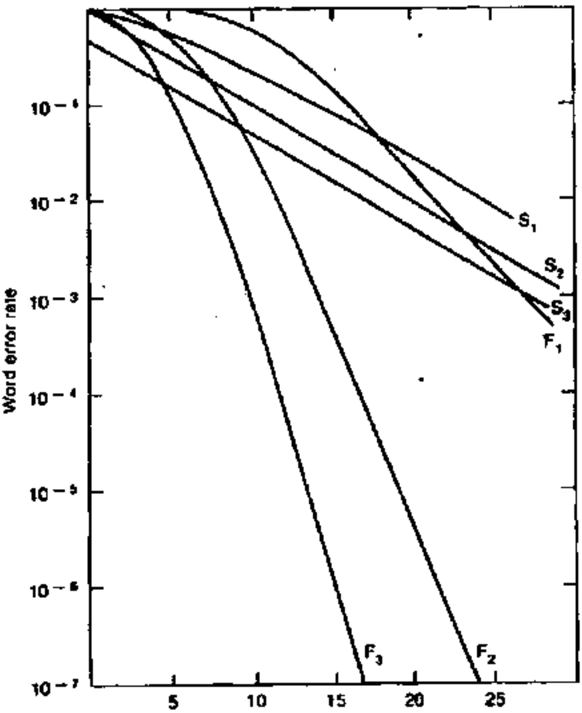
The performance of word error rates is shown in Fig. 13.2 for two cases: (1) no error correction and (2) one error correction. We have noticed that without redundancy (no repeat), the WER of a fast-fading case is worse than that of a slow-fading case. The WER obtained from a finite speed will lie between these two curves.

When a redundancy scheme is applied (Fig.13.2*a*), i.e., repeating K times and making a majority voting on each bit, the WER of a slow-fading case becomes worse than that of a fast-fading case. This change in WER provides a great improvement in performance. Therefore, a redundancy scheme is of value in a mobile radio environment with a variable vehicle speed. This phenomenon is also illustrated in Fig. 13.2*b* for a 1-bit error correction code. Figure 13.3*a* and *b* shows the



(a)

Figure 13.2 Word error rate for $N = 40$ bits. Number of branches $M = 1$; S_1, F_1, no repeat ($K = 1$); S_2, F_2 two-thirds voting ($K = 3$); S_3, F_3 three-fifths voting ($K = 5$). (*a*) Case 1: $M = 1$, $t = 0$. No error correction. (*b*) Case 2: $M = 1$, $t = 1$. One-bit error correction.



(b)

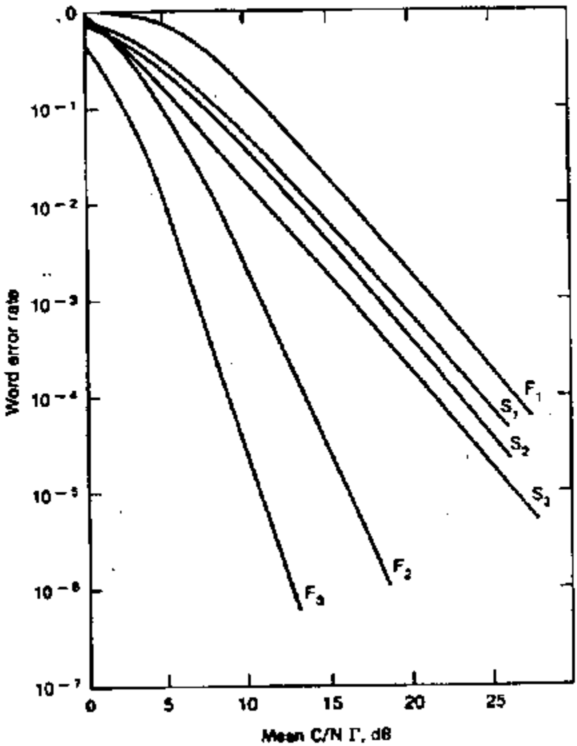
Figure 13.2 (*Continued*)

WER for two-branch diversity. Figure 13.3*a* is WER with no error correction code, and Fig. 13.3*b* is WER with one error correcton code. Further improvements are seen in the figure. The error correction code and the diversity plus the redundancy provide a desired signaling performance.

13.2.4 Parity check bits

In this section, we illustrate the generation of parity check bits in a word. Let a word of (7, 4) be generated with 4 bits of information and a 3-bit parity check. The word matrix $[C]$ can be expressed as

$$[C] = [x_m]\,[G] \tag{13.2-5}$$



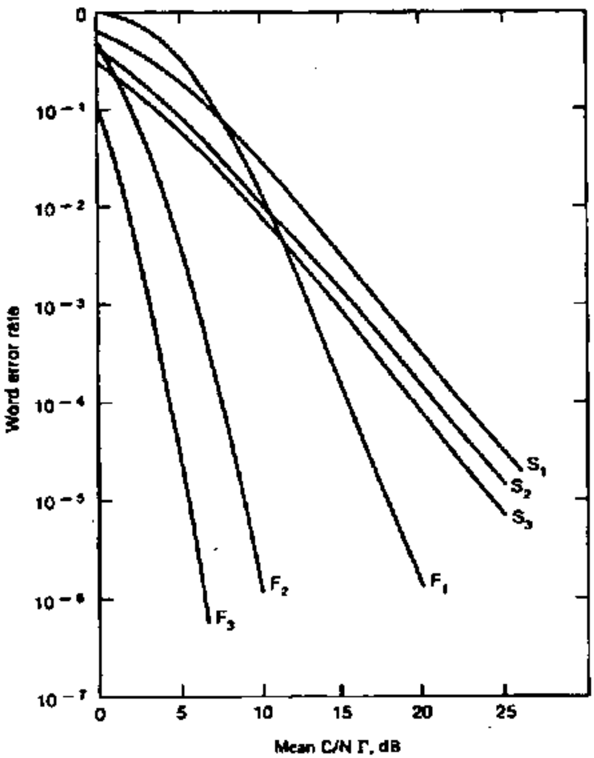
(a)

Figure 13.3 Word error rate for $N = 40$ bits. Number of branches $M = 2$; S_1, F_1, no repeat ($K = 1$); S_2, F_2 two-thirds voting ($K = 3$); S_3, F_3 three-fifths voting ($K = 5$). (*a*) Case 3: $M = 2$, $t = 0$. No error correction. (*b*) Case 4: $M = 2$, $t = 1$. One-bit error correction.

where $[x_m]$ is the information matrix and $[G]$ is the generation matrix. Let $[G]$ have the following form.

$$[G] = \underset{\text{Identity}\qquad\text{Parity}}{\begin{bmatrix} 1000 & 101 \\ 0100 & 111 \\ 0010 & 110 \\ 0001 & 001 \end{bmatrix}} = [IP] \tag{13.2-6}$$

The parity matrix of three bits can be arranged in any order and have



(b)

Figure 13.3 (*Continued*)

any combination of 0s and 1s. For example, let $[x_m] = 1001$, then substituting $[x_m]$ and Eq. (13.2-6) into Eq. (13.2-5) yields

$$[C] = [1001] \begin{array}{c} \begin{array}{cccc} & \text{5th} & \text{6th} & \text{7th} \end{array} \\ \begin{bmatrix} 1000 & 1 & 0 & 1 \\ 0100 & 1 & 1 & 1 \\ 0010 & 1 & 1 & 0 \\ 0001 & 0 & 1 & 1 \end{bmatrix} \end{array} = [1001\ C_5C_6C_7]$$

Then C_5 can be obtained by multiplying the fifth column by $[x_m]$ and applying modulo 2 addition* as

*Modulo 2 additions:
$1 + 1 = 0 \quad 1 + 0 = 1 \quad 0 + 0 = 0 \quad 0 + 1 = 1$

$$C_5 = 1 \cdot 1 + 0 \cdot 1 + 0 \cdot 1 + 1 \cdot 0 = 1 + 0 + 0 + 0 = 1$$

The same process is applied to C_6 and C_7 as

$$C_6 = 0 + 0 + 0 + 1 = 1 \qquad C_7 = 1 + 0 + 0 + 1 = 0$$

Therefore the information fits (1001), along with the three parity check bits, become a word (1001110).

13.3 Measurement of Average Received Level and Level Crossings

13.3.1 Calculating average signal strength[5]

The signal strength can be averaged properly to represent a true local mean $m(x)$ to eliminate the Rayleigh fluctuation and retain the long-term fading information due to the terrain configuration. Let $\hat{m}(x)$ be the estimated local mean. If a length of data L is chosen properly, $\hat{m}(x)$ will approach $m(x)$ as

$$\hat{m}(x) = \frac{1}{2L}\int_{x-L}^{x+L} r(y)\,dy = \frac{1}{2L}\int_{x-L}^{x+L} m(y)r_0(y)\,dy$$

$$= m(x)\left[\frac{1}{2L}\int_{x-L}^{x+L} r_0(y)\,dy\right] = m(x) \qquad (13.3\text{-}1)$$

or

$$\frac{1}{2L}\int_{x-L}^{x+L} r_0(y)\,dy \rightarrow 1 \qquad (13.3\text{-}2)$$

where $r_0(y)$ is a Rayleigh distributed variable. If the value of Eq. (13.3-2) is close to 1, then $\hat{m}(x)$ is close to $m(x)$. The spread of $\hat{m}(x)$, denoted as $\sigma_{\hat{m}}$, can be expressed as

$$1\sigma_{\hat{m}} \text{ spread} = 20 \log \frac{m(x) + \sigma_{\hat{m}}}{m(x) - \sigma_{\hat{m}}} \quad \text{in dB} \qquad (13.3\text{-}3)$$

Equation (13.3-3) is plotted in Fig. 13.4. The $1\sigma_{\hat{m}}$ spread is used to indicate the uncertainty range of a measured mean value from a true mean value if the length of the data record is inadequate.

The proper length 2L. If we are willing to tolerate $1\sigma_{\hat{m}}$ spread in a range of 1.56 dB, then $2L = 20\lambda$. If the tolerated spread is in a range of 1 dB, then $2L = 40\lambda$.

For length $2L$ less than 20 wavelengths, the $1\sigma_{\hat{m}}$ spread begins to

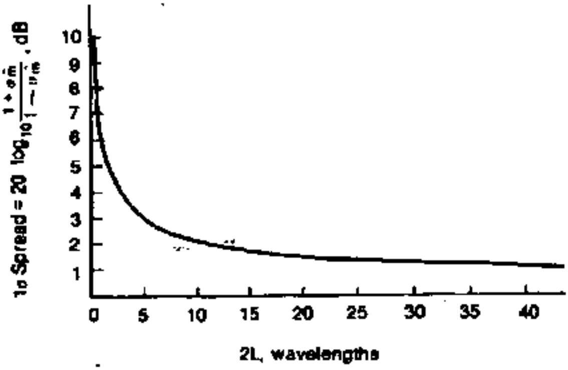


Figure 13.4 The value of $1\sigma_{\hat{m}}$ spread.

increase quickly. When length $2L$ is greater than 40λ, the $1\sigma_{\hat{m}}$ spread decreases very slowly.

In addition, the mobile radio signal contains two kinds of statistical distributions: $m(y)$ and $r_0(y)$. If a piece of signal data $r(y)$ is averaged, we find that if the length is shorter than 40λ, the unwanted $r_0(y)$ may be retained whereas at lengths above 40λ smoothing out of long-term fading $m(y)$ information may result. Therefore, 20 to 40λ is the proper length for averaging the Rayleigh fading signal $r(y)$.

Sampling average.* As mentioned previously, when using the averaging process with a filter, it is difficult to control bandwidth even when the length of the data to be integrated is appropriate. Therefore, the sample values of $r(t)$ are used for sampling averaging instead of analog (continuous waveform) averaging. Then we must determine how many samples need to be digitized across a signal length of $2L$ (see Fig. 13.5). The number of samples taken for averaging should be as small as possible. However, we have to calculate how many sample points are needed for adequate results. We set a confidence level of 90 percent and determine the number of samples required for the sampling average. The general formula is

$$P\left(-1.65 \leq \frac{\bar{r}_j - \hat{m}_j}{\hat{\sigma}_j} \leq 1.65\right) = 90\% \qquad (13.3\text{-}4)$$

*The detailed derivation is shown in W. C. Y. Lee, *Mobile Communications Design Fundamentals*, John Wiley & Sons, 1993, Sec. 2.2.2.

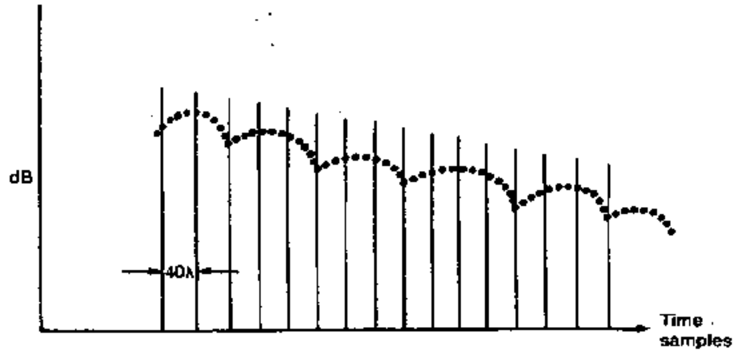


Figure 13.5 Sample average over a $2L = 40\ \lambda$ of data.

Let $\hat{m}_j$ and $\hat{\sigma}_j$ be the mean and the standard deviation of ensemble average* $\bar{r}_j$ of jth interval ($2L$) and r_j be a gaussian variable

$$\hat{\sigma} = \frac{\sigma_r}{N} \qquad \hat{m} = m$$

where m and σ_r are the mean and the standard deviation of a Rayleigh sample r. N is the number of samples. Therefore,[5]

$$m = \frac{\sqrt{\pi}}{2}\sqrt{\overline{r^2}} \tag{13.3-5}$$

$$\sigma_r = \frac{\sqrt{4-\pi}}{2}\sqrt{\overline{r^2}} \tag{13.3-6}$$

$$\frac{\sigma_r}{m} = \sqrt{\frac{4-\pi}{\pi}} \tag{13.3-7}$$

Substitution of Eqs. (13.3-5) to (13.3-7) into Eq. (13.3-4) yields

$$P\left(\left(1 - \frac{0.8625}{\sqrt{N}}\right) m \leq \bar{r}_j \leq \left(1 + \frac{0.8625}{\sqrt{N}}\right) m\right) = 90\% \tag{13.3-8}$$

Then the 90 percent confidence interval CI expressed in decibels is

*Time average in a mobile radio environment is an ergodic process in statistics. Therefore the values from a time average with a proper interval and an ensemble average are the same.

$$90\%\ CI = 20 \log \left(1 + \frac{0.8625}{\sqrt{N}}\right)$$
$$N = \begin{cases} 50 & 90\%\ CI = 1\text{ dB} \\ 36 & 90\%\ CI = 1.17\text{ dB} \end{cases} \qquad (13.3\text{-}9)$$

In an interval of 40 wavelengths using between 36 and 50 samples is adequate for obtaining the local means. For frequencies lower than 850 MHz, we may have to use an interval of 20λ to obtain the local means because the terrain contour may change at distances greater than 20λ when the wavelength increases.

13.3.2 Estimating unbiased average noise levels[8]

Usually the sampled noise in a mobile environment contains high-level impulses that are generated by the ignition noise of the gasoline engine. Although the level of these impulses is high, the pulse width of each impulse generally is very narrow (see Fig. 13.6). As a result, the energy contained in each impulse is very small and should not have any noticeable effect on changing the average power in a 0.5 s interval.

However, in a normal situation averaging a sampled noise is done by adding up the power values of all samples, including the impulse samples, and dividing the sum by the number of samples. This is called the *conventionally averaged noise power* and is denoted n_c. In

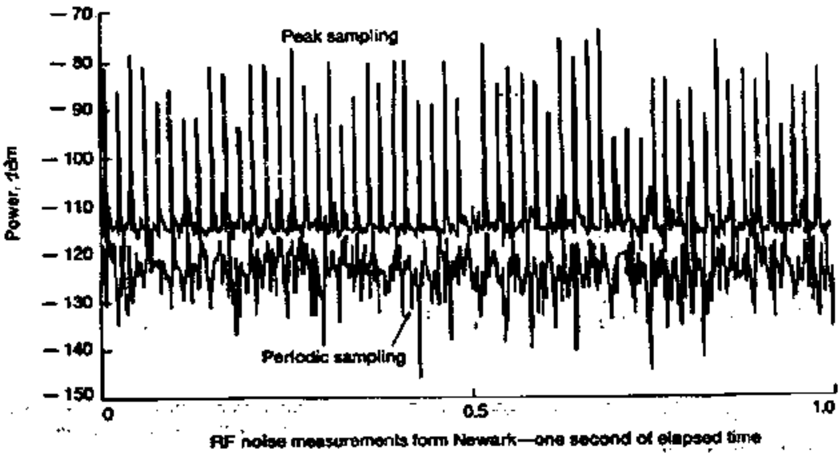


Figure 13.6 Environmental noise traces.

this case, these impulse samples dominate the average noise calculation and result in a mean value which is not representative of the actual noise power. Use of this value affects the design requirements of the system (signaling and voice). Hence, before the following new technique was introduced, it was not known why there was no correlation between BER and signal-to-noise ratio measured in certain geographic areas. In a new statistical method the average noise is estimated by excluding the noise impulses while retaining other forms of interference. This technique is compatible with real-time processing constraints.

Description of the method. A counter in the mobile unit counts the instantaneous noise measurements which fall below a preset threshold level X_t and sends a message containing the number of counts n to the database for recording. From the database data, we can calculate the percentage of noise samples x_i below the present level X_t

$$P(x_i \leq X_t) = \frac{n}{N} \tag{13.3-10}$$

where in our case N is the total number of samples. Once we know the percentage of noise samples below level X_t, we can obtain the average "noise" X_0 exclusive of the noise spikes from the Rayleigh model. Furthermore the level X_t can be appropriately selected for both noise and signal measurements, because both band-limited noise and mobile radio fading follow the same Rayleigh statistics.

Estimating the average noise X_0. For a Rayleigh distribution (band-limited noise), the average noise power exclusive of the noise spikes X_0 can be obtained from

$$X_0 = 10 \log \left\{ - \frac{1}{\ln [1 - P(x_i \leq X_t)]} \right\} + X_t \qquad \text{dBm} \tag{13.3-11}$$

This technique can be illustrated graphically using Rayleigh paper. Since $P(x_i \leq X_t)$ is known for a given X_t, we can find a point P_t on the paper as illustrated in Fig. 13.7. Through that point, we draw a line parallel to the slope of the Rayleigh curve and meet the line of $P =$ 63 percent. This crossing point corresponds to the X_0 level (unbiased average power in decibels over 1 mW, dBm).

Example 13.7 If a total number of samples is 256 and 38 samples are below a level -119 dBm, then the percentage is

$$\frac{38}{256} = 15\%$$

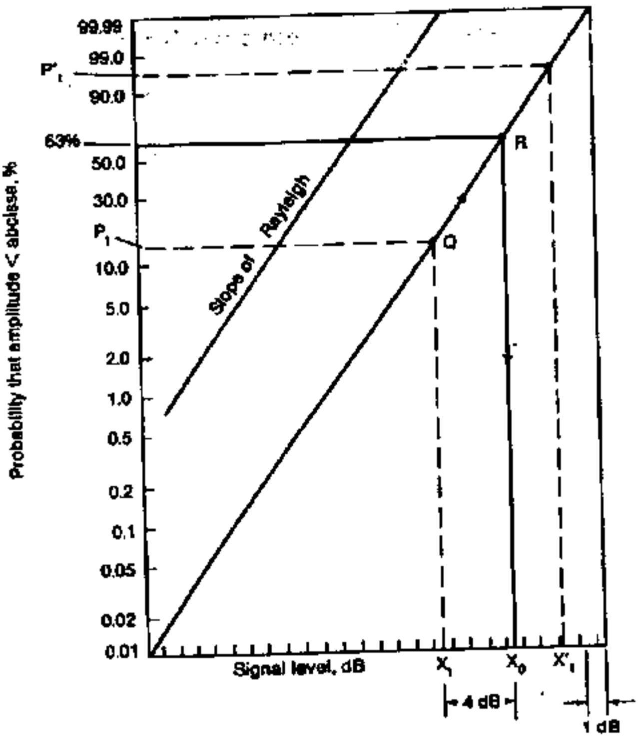


Figure 13.7 Technique of estimating average noise.

Draw a line at 15 percent and meet at Q on the Rayleigh curve (see Fig. 13.7). Assume that the X_1 is -119 dBm. The X_0 is the average power because 63 percent of the sample is below that level. Then

$$X_0 = X_1 + 4 \text{ dB} = -119 + 4 = -115 \text{ dBm}$$

Example 13.8 The total number of samples is 256. Three noise spikes are 20 dB above the normal average. Find the errors, using the following two methods. Compare the results.

Use geometric average method. Let the power value of each sample (of 253 samples) after normalization be 1, i.e., the average is 1. Then the measured average of 256 samples, including three spikes, is

$$\text{Measured average} = \frac{\sum_1^{253} x_i + 100 \sum_1^{3} x_i}{256} = 2.16 \qquad (\text{assume } x_i = 1)$$

$$= 3.3 \text{ dB} \qquad \text{above the true average}$$

Statistical average method

$$63\% \text{ of samples } = 256 \times 0.63 = 161 \text{ samples}$$

This means that 161 samples should be under the average power level. Now three noise spikes added to the 161 samples increases the number of samples to 164.

$$\frac{164}{256} = 64\%$$

The power levels at 63 and 64 percent show almost no change (see Fig. 13.7). Typical data averaging using the geometric and statistical average methods is illustrated in Fig. 13.8. The corrected value is approximately −118 to −119 dBm. The geometric average method biases the average value and causes an unacceptable error as shown in the figure.

13.3.3 Signal-strength conversion

Confusion arises because the field strength (in decibels above 1 μV, dBμ) is measured in free space, and the power level in decibels above 1 mW (dBm) is measured at the terminal impedance of a given receiving antenna. Furthermore, the dimensions of the two units are different. The signal field strength measured on a linear scale is in microvolts per meter (μV/m), and the power level measured on a linear scale is in milliwatts (or watts).

Further confusion arises because of the notation "dBμ." Sometimes dBμ means the number of decibels above 1 μV measured at a given

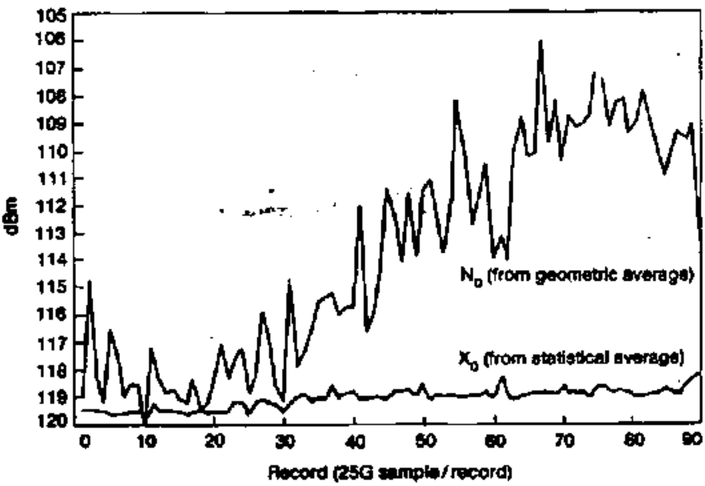


Figure 13.8 An illustration of comparison of n_e with x_e.

voltage. Sometimes, it represents the number of decibels referred to microvolts per meter when field strength is being measured.

The conversion from decibels (microvolts per meter, dBμ) to decibels above 1 mW (dBm) at 850 MHz is shown in Eq. (5.1-13) using the relationship between induced voltage and effective antenna length.[7] The conversion at a frequency other than 850 MHz can be obtained as follows.[8]

$$P_{max}(\text{in dBm at } f_1 \text{ MHz}) = P_{max}(\text{in dBm at 850 MHz}) + 20 \log\left(\frac{850}{f_1}\right) \quad (13.3\text{-}12)$$

where f_1 is in megahertz. The details of this conversion are given in Sec. 5.1.3.

13.3.4 Receiver sensitivity

The sensitivity of a radio receiver is a measure of its ability to receive weak signals. The sensitivity can be expressed in microvolts or in decibels above 1 μV.

$$Y \quad \text{dB}\mu\text{V} = 20 \log (x \quad \mu\text{V}) \qquad (13.3\text{-}13)$$

Also, the sensitivity can be expressed in milliwatts or dBm.

$$y \quad \text{dBm} = 10 \log (x \quad \text{mW}) \qquad (13.3\text{-}14)$$

The conversion from microvolts to decibels above 1 mW, assuming a 50-Ω terminal, has been shown in Eq. (5.1-15) as

$$0 \text{ dB}\mu\text{V} = 10 \log \frac{(1 \times 10^{-6})^2}{50} = -137 \text{ dBW} = -107 \text{ dBm} \qquad (13.3\text{-}15)$$

Example 13.9 A receiver has a sensitivity of 0.7 μV. What is the equivalent level in decibels above 1 mW?

$$20 \log 0.7 = -3$$

Then 0.7 μV equals −107 dBm −3 = −110 dBm.

13.3.5 Level-crossing counter[9]

A signal fading level crossing counter will face a false-count problem as a result of the granular noise as shown in Fig. 13.9. The positive slope crossing count should be 3; but the false counts may be 12. A

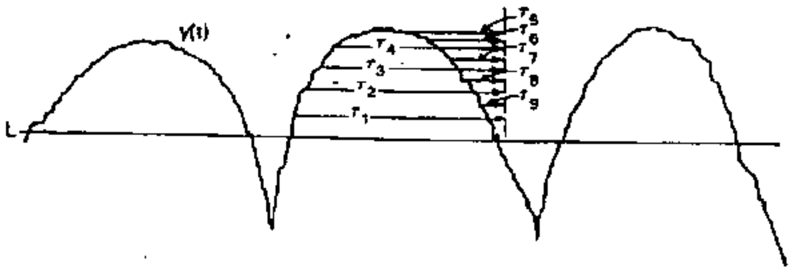


Figure 13.9 An algorithm for a level-crossing counter.

proposed level-crossing counter can eliminate the false counts. First, by sampling the fading signal at an interval of T seconds, we can choose the interval T such that $1/T$ is small in comparison to the fading rate. The duration of stay τ_i is measured for every sample time which is above level L and the time span until the signal drops below level L. We can also use a device for measuring the percentage of time that $y(t)$ is above L. This device may be called a level crossing counter.

Let us define

$$p = P_r[y(t) \geq L] \tag{13.3-16}$$

and

$$q = 1 - p = P[y(t) < L] \tag{13.3-17}$$

We count the number of times M that τ_i is above L and then sum the duration of total stays $T_s(T_s = \Sigma_{i=1}^{M} \tau_i)$ above L. The average duration of upward fading is

$$\tau_p = \frac{2T_s}{M} \tag{13.3-18}$$

The average duration of fades τ_q where $y(t)$ is below L is

$$\tau_q = \tau_p \frac{1-p}{p} \tag{13.3-19}$$

and the level crossing rate n at level L is

$$n = \frac{1}{\tau_p + \tau_q} = \frac{p}{\tau_p} \tag{13.3-20}$$

The advantage of this method is that we stop our time count whenever $y(t)$ crosses L, thus avoiding false counts due to noise when $y(t)$ is close to level L. Obviously noise can give an incorrect measure of the "duration of stay" for a single interval; however, noise shortens many intervals while it lengthens others, and thus, when averaged over many cycles, the "duration of stay" is an accurate number.

13.4 Spectrum Efficiency Evaluation[10]

13.4.1 Spectrum efficiency for cellular systems

Because the frequency spectrum is a limited resource, we should utilize it very effectively. In order to approach this goal, spectrum efficiency should be clearly defined from either a total system point of view or a fixed point-to-point link perspective. For most radio systems, spectrum efficiency is the same as channel efficiency, the maximum number of channels that can be provided in a given frequency band. This is true for a point-to-point system that does not reuse frequency channels such as a cellular mobile radio. An appropriate definition of spectrum efficiency for cellular mobile radio is the number of channels per cell. Therefore, in cellular mobile radio systems:

$$\text{Spectrum efficiency} \neq \text{channel efficiency}$$

The system capacity is directly related to spectrum efficiency but not to channel efficiency.

13.4.2 Advantages and impact of FM

In 1936 E. H. Armstrong published a paper entitled "*A Method of Reducing Disturbance in Radio Signaling by a System of Frequency Modulation*."[11] This paper explored the tradeoffs between noise and bandwidth in FM radio. Since then, engineers have understood the concept of reducing noise by increasing bandwidth in system design.

The parameters for system comparison are *voice quality, transmitted power,* and *cell size.* Satisfactory voice quality is generally accepted as governed by the carrier-to-noise ratio $C/N_{FM} = 18$ dB.* This is the level at which 75 percent of the users state that voice quality is either good or excellent in 90 percent of the service area on a 30-kHz FM channel in a multipath fading environment.[12] For point-to-point radio links, the wider the channel bandwidth, the lower the required level of transmitted power.

To maintain voice quality when channel bandwidth is reduced, it is necessary to increase the signal-to-noise ratio in order to improve the reception. Transmission power is then also increased.

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*If we use values other than 18 dB, the analysis used in this section remains the same.

Every time power is increased, interference problems are created. For point-to-point radio links, these problems are manageable because no frequency reuse is involved. This is not the case, however, for a frequency-reuse system such as a cellular radio. A cellular radio telephone system includes many mobile-unit customers and, depending on demand at any given time in a given system, identical channels will be operating simultaneously in different geographic locations. As the number of cells increases in a given area, interference may appear in one of several forms: cochannel, adjacent-channel, or multichannel at colocations; thus the probability of its occurrence increases. Interference may also result from received power-level differences. In a frequency-reuse system, however, cochannel separation is more critical to the system than adjacent-channel interference because adjacent-channel interference may be eliminated by the use of sharp filters.

13.4.3 Number of frequency-reuse cells K

The formula for determining the number of frequency-reuse cells in a standard cellular configuration is derived by combining Eqs. (2.4-3) and (2.4-5) with $\gamma = 4$ based on the 40 dB/dec path-loss rule.[13]

$$\frac{C}{I} = \frac{1}{6}\left(\frac{D}{R}\right)^4 = \frac{(3K)^2}{6} = \frac{3K^2}{2} \tag{13.4-1}$$

or

$$K = \sqrt{\frac{2}{3}\frac{C}{I}} \tag{13.4-2}$$

The number of frequency-reuse cells is a function of the required carrier-to-interference ratio.

A higher required carrier-to-interference ratio at the boundary of a cell results in the need for more frequency-reuse cells. The pattern of reuse cells can then be determined.

13.4.4 Number of channels per cell m

The next factor to be determined is the number of channels per cell, which is a function of the total number of channels available (amount of available spectrum divided by channel bandwidth) and the required carrier-to-interference ratio. The formula for this factor is[10]

$$m = \frac{B_t}{B_c K} = \frac{B_t}{B_c\sqrt{(2/3)(C/I)}} \tag{13.4-3}$$

for $M = mK$ total number of channels,

where m = number of channels per cell, also called radio capacity by Lee[10]

K = number of frequency-reuse cells (see Eq. 13.4-2)

B_t = total bandwidth (transmitted or received)

B_c = channel bandwidth

13.4.5 Rayleigh fading environment

The Rayleigh fading environment is the mobile radio environment caused by multipath fading, which is the cellular system environment. Therefore, it is more realistic to determine the spectrum efficiency of a cellular mobile radio in a Rayleigh fading environment.

In a multipath fading environment, a simple FM system which may not have either preemphasis-deemphasis or diversity schemes would receive its baseband signal-to-noise ratio (S/N), which is converted from the carrier-to-interference ratio (C/I) but S/N is 3 dB lower than C/I.[14]

System advantages. The FCC has released specifications which result in advantages vis à vis the signal-to-noise ratio for transceivers in the existing FM cellular system. The first advantage is preemphasis-deemphasis, which equalizes the baseband signal-to-noise ratio over the entire voice band (f_1 to f_2). We make the assumption of gaussian noise because the interference obtained from all six cochannel interferers behaves in a noiselike manner. The improvement factor ρ_{FM}, that is, the improvement of FM with preemphasis or deemphasis over FM without them, can be calculated as follows.[15,16]

$$\rho_{FM} = \frac{(f_2/f_1)^2}{3} = \frac{(3000\ \text{hZ}/300\ \text{Hz})^2}{3} = 33.3\ (=)\ 15.2\ \text{dB} \qquad (13.4\text{-}4)$$

Another advantage is the two-branch diversity combining receiver, which is very suitable for FM and reduces multipath fading. The advantage of the two-branch diversity receiver is that the baseband signal-to-noise ratio S/N of a two-branch FM receiver shows a 8-dB improvement over the signal-to-noise ratio of a single FM channel.

$$\left(\frac{S}{N}\right)_{2\text{brFM}} = 8\ \text{dB} + \left(\frac{S}{N}\right)_{\text{FM}} \qquad (13.4\text{-}5)$$

The existence of compandors is assumed for compressing the signal bandwidth and taking advantage of the quieting factor during pauses. However, the voice quality improvement due to the quieting factor cannot be expressed mathematically. It is understood that all the analog modulation systems use compandors.

Present FM system. The subjective required carrier-to-interference ratio for FM is

$$\left(\frac{C}{I}\right)_{\text{FM}} = 18 \text{ dB } (=) \ 63.1 \tag{13.4-6}$$

The baseband signal-to-noise ratio can be obtained from the previous analysis as follows.

$$\left(\frac{S}{N}\right)_{\text{2brFM}} = -3 \text{ dB} + \text{deemphasis gain} + \text{diversity gain} + \left(\frac{C}{I}\right)_{\text{FM}}$$

$$= 38.23 \text{ dB } (=) \ 6652.73 \tag{13.4-7}$$

The signal-to-noise value $S/N = 38$ dB is a reasonable figure for obtaining good quality at the baseband.[17] The notation (=) means a conversion between decibels and a linear ratio.

SSB systems. Single-sideband receivers, best case, have a carrier-to-interference ratio equal to the signal-to-noise ratio at baseband since SSB is a linear modulation.[17] The term "best case" means that the signal fades are completely removed, that is, the environment approaches a gaussian.* There is no advantage in using diversity schemes in a gaussian environment. If the environment is Rayleigh, the carrier-to-interference ratio must always be higher than that in a gaussian environment in order to obtain the same voice quality at the baseband. An explanation is given in Sec. 13.4.7. To obtain a similar voice quality, the signal-to-noise ratio of both FM and SSB systems at the baseband should be the same.[18] The formula used to determine the required C/I of SSB is

$$\left(\frac{C}{I}\right)_{\text{SSB}} = \left(\frac{S}{N}\right)_{\text{SSB}} = \left(\frac{S}{N}\right)_{\text{2brFM}} = 38.23 \text{ dB } (=) \ 6652.73$$

This means that the required 38.23-dB carrier-to-interference ratio of 38.23 dB for an SSB system (see Sec. 13.4.7) is equivalent to the required carrier-to-interference ratio of 18 dB for an FM system for

*SSB systems at 800 MHz have not been commercially available because of technical difficulties. It is assumed here that an ideal SSB at 800 MHz can be built for mobile radios.

TABLE 13.1 Channels per Cell
(Rayleigh Fading Environment)

System	Bandwidth B_c, kHz	Cells per set K	Total number of channels B_t/B_c	Channels per cell m
FM	30.0	7	333	47.57
SSB	7.5	66	1333	20.00
SSB	5.0	66	2000	30.0
SSB	3.0	66	3333	50.05

equivalent voice quality. The number of channels per cell and the number of channels per square mile for the system can then be calculated.

Number of channels per cell *m*. The preceding analysis gives us the information needed to determine the number of channels per cell. Assuming $B_t = 10$ MHz, the formula is [from Eq. (13.4-3)]

$$m = \frac{B_t}{B_c K} = \frac{10 \text{ MHz}}{B_c \sqrt{(2/3)(C/I)}} \tag{13.4-8}$$

Given a total bandwidth (B_t) of 10 MHz, $C/I = 18$ dB for FM, and $C/I = 38.23$ dB for SSB, it is possible to determine the number of channels per cell m by the substitution of the above values in Eq. (13.4-8) and shown in Table 13.1. As this table shows, FM cellular systems need fewer cells than do SSB systems to provide quality voice service.

There is a dispute as to whether at 800 MHz, an SSB system needs a C/N of 38 dB or less to provide an S/N of 38 dB at the baseband.[19–22] Since there is no commercial 800-MHz SSB system, no subjective test can be used for SSB voice quality. Comparing the performance of an existing FM system to that of a nonexisting SSB system is difficult. Also, it is not proper to use the results from a 150-MHz SSB system without making a thorough subjective test in a Rayleigh fading environment* and applying it to a 800-MHz SSB system.

13.4.6 Determination of cell size

It is possible to determine the size of comparable cells for 30-kHz FM, 3-kHz SSB, 5-kHz SSB, and 7.5-kHz SSB once the number of frequency-reuse cells and the number of channels per cell have been calculated. These values are related to the level of carrier power required

*Air-to-ground communications media do not exhibit Rayleigh fading behavior. Also the required C/I of SSB at 800 MHz would be different from that at 150 MHz.

at reception to maintain similar voice quality. Since SSB has a relatively narrow bandwidth, the noise level is also lower. The SSB noise level must be adjusted to the FM noise level in order to determine the power required for SSB.

Required power in each SSB system. The SSB required carrier-to-interference ratio must be 38.23 dB. Therefore, the power required by the SSB system after the noise level (including interference) has been adjusted can then be determined. The voice quality of an SSB system having a carrier-to-interference ratio of 38.23 dB is equivalent to an FM system having a carrier-to-interference ratio of 18 dB. This assumes that in-band pilot tones can smooth out the fading signal and causes no distortion in SSB reception. The noise levels of different SSB bandwidths can be shown as follows.

$$\left(\frac{C}{I}\right)_{\text{SSB}} = 38.23 \text{ dB}$$

$$\left(\frac{C}{I}\right)_{\text{SSB3kHz}} = 10.23 \text{ dB} + 18 \text{ dB} + 10 \text{ dB}$$

$$\left(\frac{C}{I}\right)_{\text{SSB5kHz}} = 12.45 \text{ dB} + 18 \text{ dB} + 7.74 \text{ dB}$$

$$\left(\frac{C}{I}\right)_{\text{SSB7.5kHz}} = 14.21 \text{ dB} + 18 \text{ dB} + 6.02 \text{ dB}$$

These figures are illustrated clearly in Fig. 13.10.

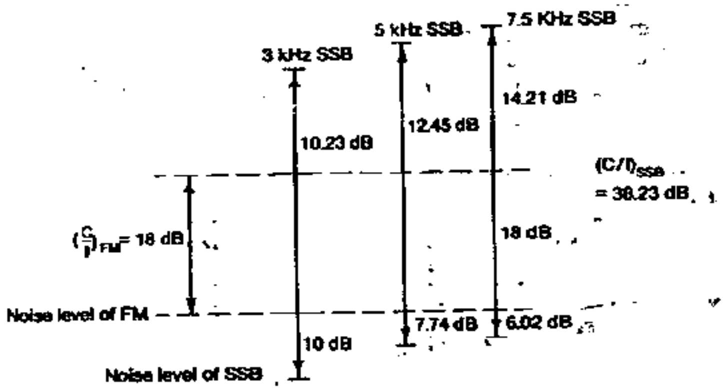

Figure 13.10 The interference-plus-noise levels for 30-kHz FM and different bandwidths of SSB.

SSB cell size determined for the required additional power. For a given transmitted power, the cell size can be determined. Assuming an FM cell radius of 10 mi and applying the 40 dB/dec path-loss rule, the cell sizes for an SSB system may be determined as follows.

For a 3-kHz SSB system

$$10 \log \left(\frac{10^{-4}}{R^{-4}}\right) = -10.23 \text{ dB} \qquad R = 5.55 \text{ mi}$$

For a 5-kHz SSB system

$$10 \log \left(\frac{10^{-4}}{R^{-4}}\right) = -12.45 \text{ dB} \qquad R = 4.88 \text{ mi}$$

For a 7.5-kHz SSB system

$$10 \log \left(\frac{10^{-4}}{R^{-4}}\right) = -14.21 \text{ dB} \qquad R = 4.41 \text{ mi}$$

The higher the carrier-to-interference ratio for a given power, the closer the mobile unit is to the cell. Assuming the same voice quality at the boundary of the cell, a 10-mi-radius FM 30-kHz cell is equivalent to a 5.5-mi 3-kHz SSB cell, a 4.88-mi 5-kHz SSB cell, or a 4.41-mi 7.5-kHz SSB cell. Therefore, an FM system permits larger cells and an SSB system requires smaller cells to provide for the same voice quality in the same area.

Comparison of cochannel cell separation and radius of FM and SSB systems in a Rayleigh fading environment. The cochannel cell separations and the radii of both FM and SSB cells in a Rayleigh fading environment are summarized in Table 13.2 and expressed in Fig. 13.11. Therefore, in a cellular mobile radio environment, a FM system *permits* larger cells with less separation between cochannel cells and a SSB system requires smaller cells with greater separation between cochannel cells.

TABLE 13.2 Comparison of Cochannel Cell Separation and Radius of FM and SSB Systems in a Rayleigh Fading Environment

System	Cells per set K	Radius R, mi	Diameter D, mi	D/R	Bandwidth, kHz
FM	7	10	46	4.6	30
SSB	66.6	5.5	77.77	14.14	3
		4.88	69	14.14	5
		4.41	62.36	14.14	7.5

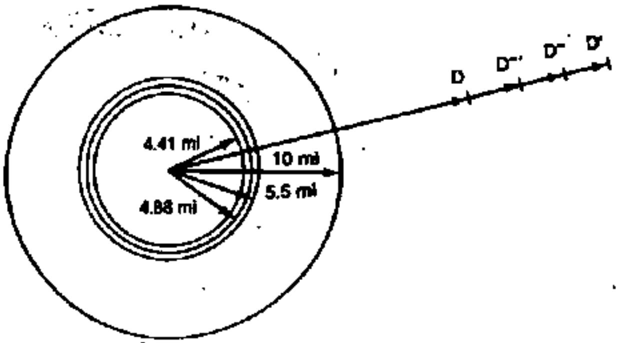

Figure 13.11 Cochannel cell separations and radii of FM and SSB systems in a Rayleigh fading environment.

Channels per square mile. Table 13.3 shows a comparison of the channels per square mile and the spectrum efficiency of each system in the Rayleigh fading environment based on equivalent voice quality and a given transmitted power. This comparison shows that the existing 30-kHz FM cellular system is about as spectrally efficient as the hypothetical 3-kHz SSB system and much more efficient than either of the two SSB (5- or 7.5-kHz) systems offering a voice quality similar to that proposed for commercial service.

13.4.7 Considerations of SSB systems in a Rayleigh fading mobile radio environment

The voice signal requires about a S/N ratio of 40-dB output at the baseband for high quality.[23] Our baseband S/N ratio is calculated to be 38 dB, which is very close to 40 dB.

TABLE 13.3 Comparison of System Efficiency and Spectrum Efficiency in an Area of 10-mi Radius

		System efficiency		Spectrum efficiency	
System	Bandwidth, mHz	Cell radius, mi	No. of cells required	Channels per cell, m	No. of channels per square mile
FM	30	10	1.0	47.5	0.15
SSB	7.5	4.4	$\left(\frac{10}{4.4}\right)^2 = 5.17$	20.0	0.06
SSB	5	4.8	$\left(\frac{10}{4.8}\right)^2 = 4.34$	30.0	0.10
SSB	3	5.5	$\left(\frac{10}{5.5}\right)^2 = 3.31$	50.0	0.16

In a gaussian environment, if the output S/N of a SSB signal at the baseband is 38 dB, then the S/N at the RF is also 38 dB because SSB is linearly modulated.[17] Now, in a Rayleigh fading environment, the received signal can be further degraded as a result of the multipath fading. Therefore, in order to maintain the baseband S/N at 38 dB, we may need to receive a signal much higher than 38 dB if no diversity scheme is implemented. How high the C/I level at RF should be cannot be determined unless the SSB mobile radio at 800 MHz is realized and a subjective test is done. Since a single-branch SSB cannot be used in a mobile radio environment at 800 MHz because of rapid fading,[24] the value of applying a two-branch diversity to the SSB at 800 MHz also must be questioned. Our reasoning is as follows.

Let S_0 be a voice signal, r_1 and r_2 the envelopes of the fading, θ_1 and θ_2 the random phases received by two spaced antennas, and ω_0 the carrier angular frequency. Where an equal-gain combined receiver is considered, in an SSB system the combined signal envelope is

$$\begin{aligned} r &= \left|r_1 S_0 e^{j(\omega_0 t + \theta_1)}\right| + \left|r_2 S_0 e^{j(\omega_0 t + \theta_2)}\right| \\ &= r_1 S_0 + r_2 S_0 = (r_1 + r_2) S_0 \end{aligned} \tag{13.4-9}$$

which is the same as the baseband signal representation. The term $(r_1 + r_2)$ does reduce the fading as compared to either individual r_1 or r_2 to a certain degree, but it also acts as a distortion term to S_0. The distortion of voice S_0 on $(r_1 + r_2)$ received by a two-branch diversity receiver for SSB in a ground mobile radio (Rayleigh) environment is still quite high at 800 MHz. This is because the effect of fading is multiplicative and produces intermodulation products with the signal modulation that cannot be eliminated by filtering.[25] In air-to-ground transmission, the direct-wave path dominates, the fading phenomenon (rician) is not severe, and a two-branch diversity does help in improving the voice quality at the reception.

An amplitude companding single sideband (ACSB) with an in-band pilot tone is considered[26] under the assumption that this kind of SSB can in principle completely remove the Rayleigh fading; therefore, no diversity scheme* is needed. Then we do not require an increase in the received power level at the RF, but rather simply retain the same level as the baseband S/N ratio of 38.23 dB.

Preemphasis and deemphasis are not widely used in SSB systems. The disadvantage of the use for SSB systems in a nonfading environment is discussed by Schwartz[27] and Gregg.[28]

*The diversity scheme is used to eliminate fading.

In a mobile radio environment, in order to transmit a predistorted SSB signal using preemphasis to suppress the noise level, the signal cannot be completely restored because the effect of fading is multiplicative and it produces intermodulation products in the voice band that cannot be eliminated by filtering as mentioned before. Therefore, the use of preemphasis and deemphasis in an 800-MHz SSB mobile radio system is questionable.

Reference 18 shows that an RF signal with required $C/I = 38$ dB received by a SSB system through a gaussian environment results in a baseband signal where $S/N = 38$ dB also. In a Rayleigh fading environment, the C/I must definitely be higher than 38 dB for S/N to be 38 dB. How high is not known since 800-MHz SSB equipment has not been manufactured. We may use the information obtained from a FM system for maintaining the same voice quality in different environments.

$$\begin{array}{ll} \text{Required for FM in a gaussian environment} & \multirow{2}{*}{$\dfrac{C}{I} \geq \begin{cases} 10 \text{ dB} \\ 18 \text{ dB} \end{cases}$} \\ \text{Required for FM in a Rayleigh fading environment} & \end{array}$$

The difference in C/I for the FM system between the two kinds of environment is 8 dB. We may use the 8-dB difference and add it to $C/I = 38$ dB for an SSB system. Then

$$\text{For SSB in a Rayleigh fading environment} \qquad \frac{C}{I} = 46 \text{ dB}$$

Suppose that a diversity scheme is used in SSB as Shivley[21] suggested. Then C/I for SSB in a less-fading (or no-fading) environment is

For SSB in a less-fading (or no-fading) environment

$$\frac{C}{I} - 8 \text{ dB} = 38 \text{ dB}$$

The same result is obtained if we assume that the fading is completely removed, $C/I = S/I$, and that the environment becomes gaussian. In a gaussian environment, we cannot reduce C/I below 38 dB, because S/N is 38 dB. Also, in a gaussian environment, the diversity scheme does not apply and adds no value.

Of course, the analysis shown here remains to be proved if and when an 800-MHz SSB mobile unit is developed in the future. After all, the methodology of solving this problem remains unchanged.

13.4.8 Narrowbanding in FM

Relationship between C/I at IF and S/N at baseband. In Ref. 12 we defined acceptable voice quality as existing when 75 percent of customers say that the voice quality is good or excellent in a 90 percent coverage area. When these numbers change, voice quality changes accordingly. The changes reflect the cost of deploying a cellular system. As the percentages specified above increase, the cost of designing the system to meet these requirements also increases. For now, we will use the numbers specified above as our criteria.

We let the customer listen to the voice quality of a 30-kHz FM two-branch diversity receiver with preemphasis-deemphasis and companding features at the cell site while the mobile transmitter is traveling at speeds ranging from 0 to 60 mi/h in a Rayleigh environment. Judging by the preceding subjective criterion, the C/I level at the input of the receiver is 18 dB.

The baseband signal-to-noise ratio (S/N) has been calculated in Eq. (13.4-7).

$$S/N = \underbrace{18 - 3}_{\text{Rayleigh environment}} + \underbrace{15.23}_{\text{deemphasis advantage}} + \overset{\text{diversity advantage}}{8} = 38.24 \text{ dB}$$

For a 15-kHz FM channel, the bandwidth is half as broad and affects the S/N; the other features, such as diversity and preemphasis-deemphasis, remain the same. We can find from Fig. 13.12 (Ref. 29) that in order to maintain the same voice quality of $(C/I)_{30\text{ kHz}} = 18$ db, then

$$\left(\frac{C}{I}\right)_{15\text{ kHz}} = 24 \text{ dB}$$

We may follow the same steps used in Sec. 13.4.6 along with the diagram shown in Fig. 2.5.

$$\frac{C}{N+1} = \frac{C}{(kTB + NF) + \sum_{i=1}^{6} I_i}$$

$$= 24 \text{ dB (15-kHz FM) or 18 dB (30-kHz FM)} \qquad (13.4\text{-}10)$$

Using the techniques from previous sections, we can perform further calculations.

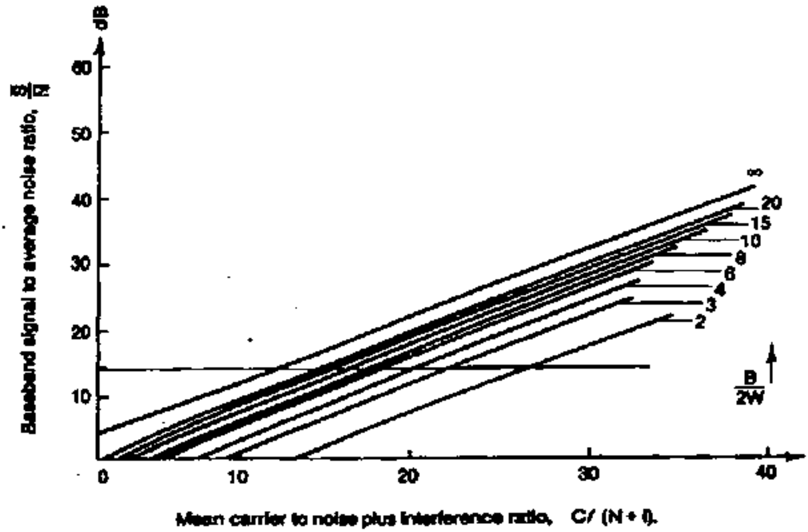

Figure 13.12 Baseband signal-to-noise ratio versus average carrier-to-interference ratio (Rayleigh fading). (*From Ref. 29.*)

The interference-plus-noise levels. From the preceding calculations Eq. (13.4-10), we can obtain the interference-plus-noise levels $(N + I)$ at the boundary of a 10-mi-radius cell as follows.

$$\text{For 30-kHz FM system} \qquad N + I = -117 \text{ dBm}$$
$$\text{For 15-kHz FM system} \qquad N + I = -123 \text{ dBm}$$

Since the received signal strength is -99 dBm for both 30- and 15-kHz, the $C/(N + I)$ ratio must be 18 dB and 24 dB for 30- and 15-kHz FM, respectively, in order to maintain the same voice quality. This relationship is shown in Fig. 13.13.

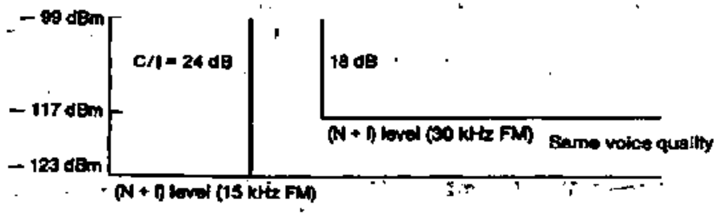

Figure 13.13 The interference-plus-noise levels for 30- and 15-kHz FM channels.

The number of cells in a frequency-reuse pattern. Let K be the number of cells in a frequency-reuse pattern; then in 30-kHz systems,

$$q = \frac{D}{R} \triangleq \sqrt{3K} = 4.6$$

$$K = 7 \qquad \text{(for both noise-neglected and noise-included cases)}$$

For a seven-cell reuse pattern operating in 15-kHz systems,

$$K = \frac{q^2}{3} \begin{cases} = 13 & (q = 6.23) \quad \text{(noise-neglected case)} \\ = 16 & (q = 7.08) \quad \text{(noise-included case)} \end{cases}$$

A 13- to 16-cell reuse pattern is needed.

Intersystem comparison of spectrum efficiency. If both the noise-included and the noise-neglected cases are considered, then

$$K = \begin{cases} 7 & \text{(30-kHz FM system, both noise-included and noise-neglected cases)} \\ 16 & \text{(15-kHz FM system, noise-included case)} \\ 13 & \text{(15-kHz FM system, noise-neglected case)} \end{cases}$$

The spectrum efficiency for both systems can be shown to be about the same as follows:

$$\frac{333}{7} = 47.5 \text{ channels per cell} \qquad \text{(30-kHz FM)}$$

$$\frac{666}{16} = 41.63 \text{ channels per cell} \qquad \text{(15-kHz FM, noise-included case)}$$

$$\frac{666}{13} = 51.2 \text{ channels per cell} \qquad \text{(15 kHz FM, noise-neglected)}$$

Increasing spectrum efficiency by degrading voice quality. If we accept $C/I = 18$ dB for 15-kHz FM, this means that voice quality is degraded by 6 dB. Then the cochannel interference reduction factor q becomes approximately 4.6, corresponding to a frequency-reuse pattern of $K = 7$. Since the frequency channel is doubled, in a 10-MHz system (666 channels)

$$\frac{666}{7} = 2 \times \text{(number of 30-kHz FM channels per cell)}$$

Therefore, voice quality is sacrificed to gain spectrum efficiency.

Another approach is to increase the transmitted power of the 15-kHz FM system by 6 dB. Then the *C/I* remains the same because the interference is also increased by 6 dB. Therefore, no advantage to spectrum efficiency can be obtained by either increasing or reducing power.

13.5 Portable Units

All of today's system design tools are designed to improve the performance of cellular mobile units. Therefore, portable units become a secondary byproduct of the cellular system. Since very few systems in the United States are designed mainly for portable units, it will be necessary to study each existing system individually to determine whether portable units are suitable for it. Portable unit usage can be adopted easily by some existing systems but not by others for many reasons. If we find that portable units become very popular in an existing mobile cellular system, then we have to reconsider some parameters used for the mobile cellular system to adapt to portable unit usage. There are two parts to the calculation.

13.5.1 Loss due to building penetration

The loss (attenuation) when propagating an 800-MHz wave due to building penetration is very high.[31–36] Also, the structure-related attenuation varies, depending on the geographic area. In Tokyo, the path-loss difference inside and outside buildings at first-floor level is about 26 dB. But in Chicago, the path-loss difference under the same conditions is 15 dB as shown in Fig. 13.14.

These variations are attributable to differences in building construction. In Tokyo, many supporting metal frames (mesh configuration) are used in the building structures to allow the buildings to withstand earthquakes. In Chicago, fewer supporting metal frames are needed as there is less risk of earthquakes. Therefore, the *building penetration* is far less severe in Chicago than in Tokyo.

The same would apply to buildings in California. For instance, the loss due to building penetration in Los Angeles would be higher than that in Chicago but lower than that in Tokyo, because Los Angeles has only a few high-rise buildings. In Los Angeles the penetration at the first-floor level would be around −20 dB, as compared with the received signals at mobile units outside at street level.

Signal attenuation at the building basement level in Chicago is 30 dB below the signal received from the street-level mobile unit. This indicates that the signals do not penetrate basement structures easily. There are two ways to solve this problem.

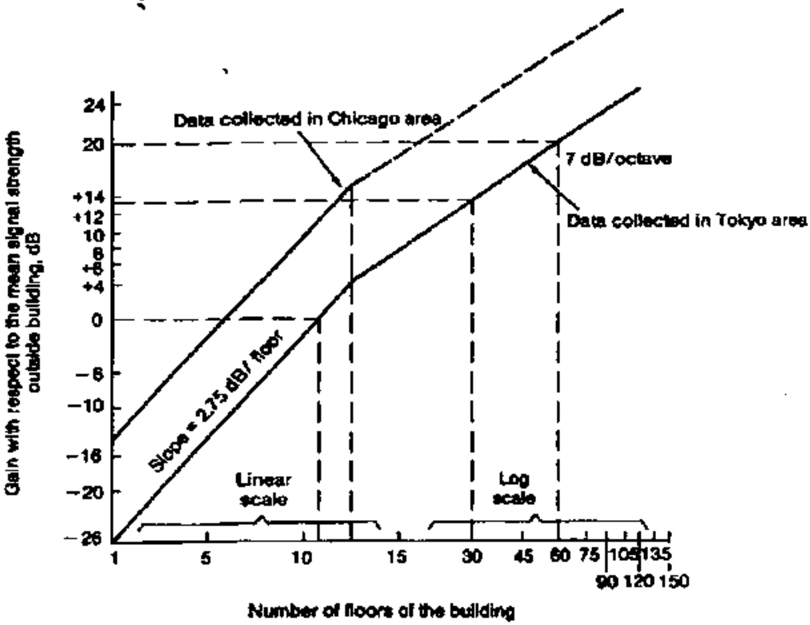

Figure 13.14 The building penetration losses in the Tokyo area and in the Chicago area.

1. Select the cell site closest to those downtown buildings used for conventions. Then calculate the received power for the portable units. Let the cell coverage be at a radial distance R, with the receiving level of $C/I = 18$ dB at the cell boundary as designed. The same site will be used for portable units. The receiving level of portable units is $C/I = 10$ dB because of the slow-motion or no-motion environment. At the basement, the reception level of a portable unit is $-30 + (18 - 10) = -22$ dB, i.e., 22 dB weaker than that received at the cell boundary R. Applying Eq. (4.2-10a), we obtain

$$40 \log \frac{R_1}{R} = -22 \text{ dB} \qquad \text{or} \qquad R_1 = R(10^{-(22/40)}) = 0.28R \qquad (13.5\text{-}1)$$

From this calculation, the region in which the portable unit can be used in the basement is confined to an area of $0.28R$.

2. Install the repeaters (or enhancers) or leaky feeders to enhance the signal strength inside the building.

13.5.2 Building height effect

Usually, the signal reception level increases as the height of the building where the antenna is located increases. Comparing the two measurements from Chicago and Tokyo, we find that the slope (gain increase) of 2.7 dB per floor (i.e., 2.7 dB gain per each floor-level increment) can be a good value on which to base calculations regarding suitability of portable unit usage. However, this value is valid only up to the thirteenth floor. After that, a logarithmic 7 dB/oct scale is used. Now we start at the first floor.

First-floor region. Following the same procedure as when calculating the signal strength in the basement region, we find the signal strength requirements at the first floor inside the building for the portable unit to be

$$-15 + (18 - 10) = -7 \text{ dB} \qquad \text{(Chicago)}$$

$$-26 + (18 - 10) = -18 \text{ dB} \qquad \text{(Tokyo)}$$

$$-20 + (18 - 10) = -12 \text{ dB} \qquad \text{(intermediate value)}$$

Applying these findings to Eq. (13.5-1), we obtain

$$40 \log \frac{R_1}{R} \doteq \begin{cases} -7 \text{ dB} \\ -18 \text{ dB} \\ -12 \text{ dB} \end{cases} \qquad R_1 = \begin{cases} 0.668R & \text{(Chicago)} \\ 0.35R & \text{(Tokyo)} \\ 0.5R & \text{(intermediate value)} \end{cases}$$

Nth-floor region. The area serviced increases as a function of height of 2.7 dB per floor below the thirteenth floor, where we see a gain increase of 2.7 dB per floor, and above the thirteenth floor, 7 dB/oct is used. We can calculate the service area in Chicago as follows.

$$40 \log \frac{R_1}{R} \doteq -7 + N(2.7) \qquad \text{(in Chicago)}$$

where N is the number of floors. The same procedures apply to Los Angeles and Tokyo. In Table 13.4, we see that the service region increases at higher floor levels. However, after the thirteenth floor, the increase in gain is very small. The difference between the two floors from two cities becomes smaller for heights beyond the thirtieth floor (see Fig. 13.14).

TABLE 13.4 Building Penetration Loss

Condition	Building penetration loss	Shadow loss*
Building penetration	+27 dB (Tokyo) +15 dB (Chicago)	27 dB (Chicago) (regardless of floor height)
Window area	+6 dB	
1st–13th floors	2.75 dB/floor (Tokyo) 2.67 dB/floor (Chicago)	
13th–30th floors	7 dB/oct (Tokyo and Chicago)	

**Shadow loss* is defined as the loss due to a building standing in the radio-wave path.

13.5.3 Interference caused by portable units

Interference to the other portable units. The portable unit has a transmitting power of 600 mW (28 dBm). The interference at the cell site from two different portable units can be determined as follows. We now can consider interference at higher floor levels (see Fig. 13.14). We find that reception at the sixth floor is the same as that at street level in Chicago and that reception at the eleventh floor is the same as that at street level in Tokyo. Reception at the thirtieth floor in Tokyo is 13 dB higher than that at street level.

A portable unit transmitter can transmit a signal to a cell site following line-of-sight propagation. A signal from a portable unit on the thirtieth floor that is received by the cell site could interfere with the reception of a signal from a portable unit on the eleventh floor. The interference level for the portable unit on the eleventh floor is 13 dB. The interference range becomes

$$20 \log \frac{R_1}{R} = -13 \text{ dB} \qquad \text{then} \qquad R_1 = 0.22R$$

Since $R_1 = 0.22R$, the portable unit used on the thirtieth floor at the cell boundary R will not interfere with cell-site reception from a portable unit on the eleventh floor at 0.22 R away. If the power of both units can be controlled at the cell site, the near-end to far-end ratio interference of 13 dB can be reduced. We must be aware of this interference and find ways to eliminate it once we know its cause. A method for this was described in Chap. 10. The selection of a method for eliminating interference is based on environmental factors such as building height and density, which vary from area to area.

Interference to the mobile units. Now assume that at the cell boundary (the cell radius is R) the mobile unit received a signal at −100 dBm and the reception level of a portable unit at the thirtieth floor is −87 dBm (−100 + 13 dB). If the cell site has a 10-W transmitter and a 6-dB-gain antenna, then the transmitting site has an effective radiated power (ERP) of 46 dBm. The path loss on a thirtieth floor of the cell boundary becomes 133 dB (46 + 87). Now for calculating a reverse path, the portable unit has a 600-mW (28-dBm) transmitter; and the mobile unit has a 3-W (35-dBm) transmitter. Then the signal received at the 100-ft cell-site antenna is (28 − 133) = −105 dBm for the portable unit and (35 − 133 − 13) = −111 dBm for the mobile unit. The difference in received levels is 6 dB. This near-end to far-end ratio interference can be eliminated by the frequency assignment and power control of the portable units at the cell site.

13.5.4 Difference between mobile cellular and portable cellular systems

It is very interesting to point out the differences in characteristics, coverage charts, and system design aspects for mobile and portable cellular systems.

Different characteristics

Mobile units	Portable units
Two-dimensional system	Three-dimensional system
Needs handoffs	No handoffs
Severe signal fading due to vehicle movement	No fading or mild fading if walking
Gain changes with ground elevation	Gain changes with building height
Loss due to multipath reflection	Loss due to building penetration
Required $C/I \geq 18$ dB	Required $C/I \geq 10$ dB
Power consumption is not an issue	Power consumption is a key issue
Attractive because of various features	Attractive because of their small size and light weight

Different coverage charts. Using the Philadelphia path-loss curve shown in Fig 4.3 with the standard condition parameters listed in Sec. 4.2.1, we can illustrate the differences in the coverage charts of mobile units and portable units in urban areas, as shown in Fig. 13.15. The receiver sensitivity of both units is assumed to be −117 dBm. Also assume that the required C/I of a portable unit is 10 dB, and the required C/I of a mobile unit is 18 dB. The mobile-unit coverage is about 6 mi, while the portable coverages differ with building height, as shown in Fig. 13.15. The higher the building, the greater the coverage range.

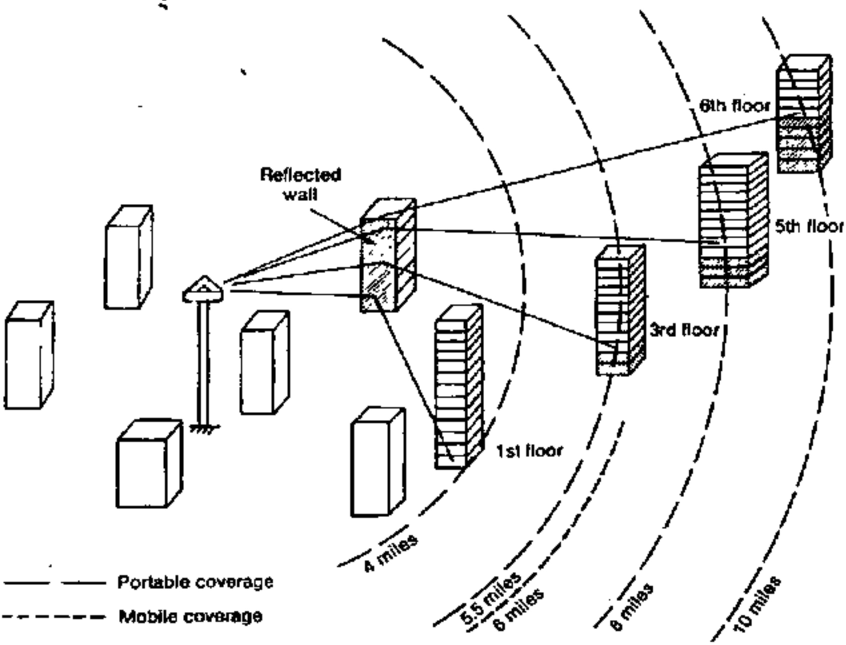

Figure 13.15 Portable and mobile coverage.

Different design concepts. In mobile cellular systems, we try to cover the area with an adequate signal from a cell site; then transmitted power, antenna height and gain, and location are the parameters involved. Reduction of both multipath fading and cochannel interference is described in Chap. 6.

In a portable cellular system, the coverage range increases with the height of the building. Therefore no fixed coverage range can be given for the portable units. Also the cochannel interference reduction ratio q [$q = D/R$ see Eq. (2.3-1)] for the portable cellular systems has no meaning since the cell radius R changes with building height. If we try to apply the design techniques for mobile cellular systems to portable cellular systems, the results are not very good. One way to look at this problem is that each building structure offers less interference inside the building. The lower the building, the greater the protection from interference. Therefore we should not raise the transmitted power and try to penetrate the building; rather we should take advantage of this natural shielding environment. Thus we should link to each repeater (enhancer) mounted at the top of each building. Since the buildings are tall, reception will be good because of the building's height, and only a small amount of transmitted power is needed at the cell site (see Fig. 13.15). If leaky cables or cables with antennas (shown in Fig. 13.16) from the repeater are connected to each floor, the signal in the whole building will be covered. This is the proper arrangement, but it is also a different concept of designing a portable cellular system.

13.6 Evaluation of Data Modem

13.6.1 Requirement

The data modem used in the current analog system must meet the following requirements.

1. Data transmissions have to use 30 kHz voice channels.
2. The SAT tone must be maintained at around 6000 Hz in voice channels. Then the transmission rate has to be either lower or higher, but it must be clear from the 6000-Hz SAT.
3. The transmission rate cannot be lower than a rate[37] which lies in the dominant random-FM region of $f_{rfm} < 2(V/\lambda)$. This specification is based on the vehicle speed V and the wavelength of the operating frequency. For instance, if $V = 104$ km/h (65 mi/h), and $\lambda = 1$ ft (at 850 MHz), then $f_{rfm} = 2(V/\lambda) = 190$ Hz. This means that the data modem transmission rate cannot be below 190 Hz because of

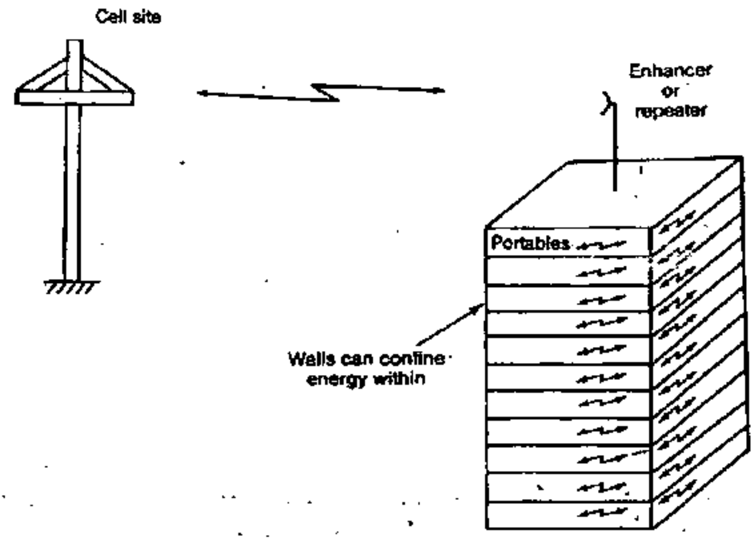

Figure 13.16 Proper arrangement for portable cellular system.

the unique random-FM characteristics in the mobile radio environment. When the mobile unit stops, the random FM disappears.

4. The same severe fading requires us to use redundancy, coding, automatic-repeat request (ARQ) scheme, diversity, and so on when transmitting data. The current mobile cellular signaling transmission rate is 10 kbps. Using the BCH code with five repeats, the data throughput is 1200 bits. However, the power spectrum density of a 10-kbps data stream with a Manchester coding (a biphase waveform) is spread out over the 6000-Hz region.[38] If the SAT cannot be detected because of the data modem, then the data modem cannot be used.
5. Mobile unit (vehicular) speed is a significant factor.
   a. Suppose that a car can be driven slowly while approaching a stop but that it never stops. In this case, the average duration of fades is very long. Most of the time, if the word is in the fade, the whole word is undetected.
   b. Suppose that a car can travel at a rate of 104 km/h (65 mi/h). The number of level crossings at −10 dB (10 dB below average power) is 65 crossings per second, and the average duration is 1.54 ms, as mentioned in Sec. 13.2.

   Thus, a word length must be designed to fit in these two cases.
6. Handoff action is another factor in data modem design. Whenever a handoff occurs, a piece of data information is lost. The average would be 200 ms. The ARQ scheme would be useful for this purpose.
7. The data modem must satisfy a specific bit error rate (BER) and word error rate (WER) requirement. The BER is independent of vehicle speed, whereas the WER is not. The higher the throughput rate (baseband transmission rate), the higher the BER and WER. However, there are two data modem markets.
   a. Use of fast data rate—in real-time situations, a customer may need quick access to data but may not need a high degree of accuracy. Examples of such customers are police agencies, real estate agencies, etc.
   b. Use of accurate data—for transmitting figures requiring a high degree of accuracy, a slow data transmission rate is needed. These customers need data accuracy more than fast acquisition, as in banking or computer applications.

13.6.2 Testing

Any data modem operating in a cellular mobile unit must demonstrate that its BER and WER satisfy the following conditions.

1. The data modem should be tested while the mobile unit is parked (stationary) at the side of the highway. Because the mobile radio environment is very noisy, the spike noise resulting from ignition-induced combustion and the sharp fades caused by the noise of passing trucks would also affect data transmission.
2. The data modem should be tested while the mobile unit is driven at different speeds, say 5, 10, 45, and 60 mi/h at the boundary of the cell.
3. The data modem should be tested during handoff conditions.

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Chapter

14

Introduction to Digital Systems

14.1 Why Digital?

14.1.1 Advantages of digital systems

In an analog system, the signals applied to the transmission media are continuous functions of the message waveform. In the analog system, either the amplitude, the phase, or the frequency of a sinosoidal carrier can be continuously varied in accordance with the voice or the message.

In digital transmission systems, the transmitted signals are discrete in time, amplitude, phase, or frequency, or in a combination of any two of these parameters. To convert from analog form to digital form, the quantizing noise due to discrete levels should be controlled by assigning a sufficient number of digits for each sample,and a sufficient number of samples is needed to apply the Nyquist rate for sampling an analog waveform.

One advantage of converting message signals into digital form is the ruggedness of the digital signal. The impairments introduced in the medium in spite of noise and interference can always be corrected. This process, called *regeneration*, provides the primary advantages for digital transmission. However, a disadvantage of this ruggedness is increased bandwidth relative to that required for the original signal.

The increased bandwidth is used to overcome the impairment introduced into the medium. An analogy to this is given in Sec. 14.1.2. In addition to the cost advantage, power consumption is lower and digital equipment is generally lighter in weight and more compact. In mobile cellular systems, there are more advantages in applying digital technology.

14.1.2. Analogy to modulation schemes

We may apply the analogy of moving books to compare analog and digital transmission. We may use a daily event to describe the different modulation schemes in the communications field. Consider the following scenario. On one side of a large hall, 20 books have been piled up (signal to be sent). We want to move these books to the other side of the hall (receiving end). Assume that the floor of the hall is not slippery and is flat (transmission medium). Then we would hire a strong, muscular person who could carry the 20 books with both hands and safely transport them to the other side without dropping a single book. This is analogous to a double sideband. The space which this person is occupying is equal to the width of the person's shoulders, as shown in Fig. 14.1*a*. If the person is very strong, one hand can be used to carry the 20 books while the other hand can be used to carry something else. This is analogous to a single sideband. If the floor in the middle of the hall contains many water puddles and (or) bumps, this is analogous to a rough medium. We might not hire a strong adult to carry the books because we might be afraid that the person would fall and drop all the books in the middle of the hall. Therefore, we might hire 10 children, each one carrying two books as seen in Fig. 14.1*b*. Among 10 children, perhaps only 1 or 2 might fall in the middle of the hall and drop the books. But most of the books will be carried to the other side of the hall. The 10 children, each carrying two books, is analogous to frequency modulation (FM). Spreading the signal into a wide spectrum (children) applies the first principle of using "spread spectrum" in FM.

A similar analogy is shown in Fig. 14.1*c*. Besides puddles and bumps, there are three guard jammers to stop anyone from bringing books over to the other side of the hall. In such a case we would have to hire as many as 20 children, each carrying one book and running against the guards. Some of the children may fall because of the puddles and bumps and some may be blocked by the guards, but most of the books will arrive on the other side of the hall.

If a force of 20 children is not sufficient to carry the books to the other side, then we must hire at least 40 children, each carrying half a book. This is the concept of spread spectrum. The spread-spectrum technique is used for combatting rough media and enemy jamming in military communications.

In each modulation scheme, energy must be confined within a specified bandwidth. In this chapter we will demonstrate the regular mod-

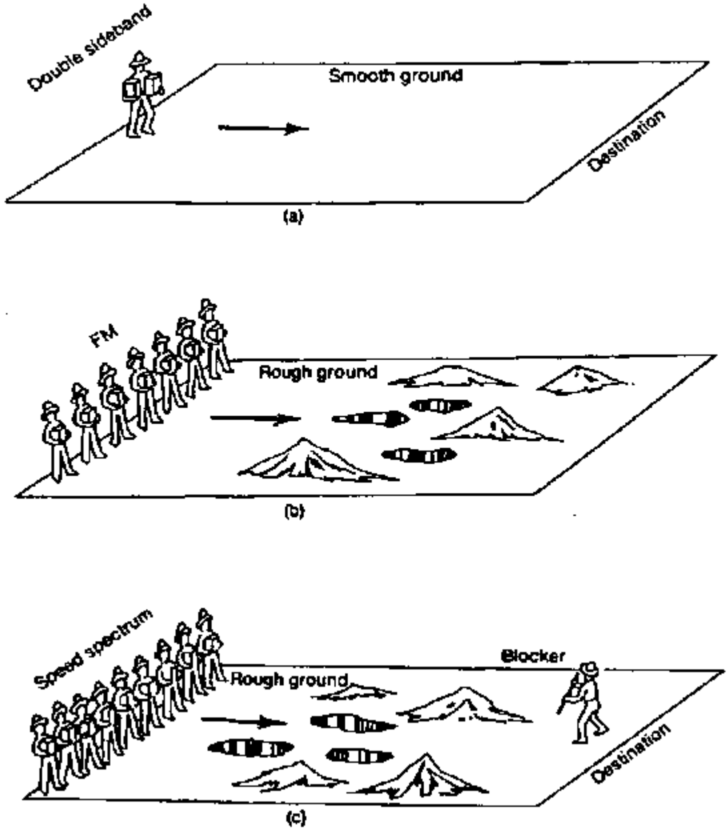

Figure 14.1 Analogy between radio transmission and moving blocks. (*a*) Walking over smooth ground; (*b*) ground with rough conditions; (*c*) ground with rough conditions plus blocks.

ulation schemes, phase shift keying (PSK) and frequency shift keying (FSK). We will also demonstrate how the schemes, MSK, GMSK, and GTFM can be applied to confine energy when transmitting a signal in the medium. The analogy can be extended to the adult or child who saves space while carrying books (Fig. 14.2*a*). The adult carrying the books (Fig. 14.2*b*) is taking too much space, and this method is not recommended. Common sense indicates that if the person wants to carry books through a crowd, the approach illustrated in Fig. 14.2*a* is more suitable than that illustrated in Fig. 14.2*b*. This analogy has been presented in the hopes of stimulating the readers to think about modulation schemes.

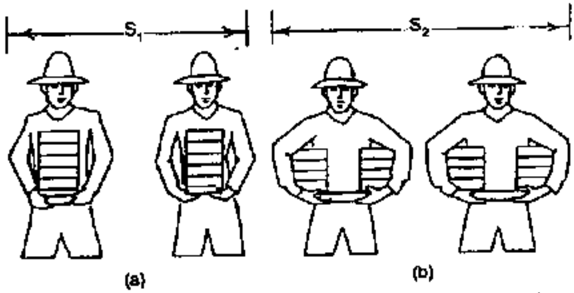

Figure 14.2 Analogy between space utilization and spectrum efficiency. (*a*) A satisfactory space utilization. (MSK, GMSK, or GTFM), (*b*) unsatisfactory space utilization (PSK).

14.2 Introduction to Digital Technology

14.2.1 Digital detection

There are three forms of digital detection: coherent detection, differentially coherent detection, and noncoherent detection. Coherent detection requires a reference waveform accurate in frequency and phase, and the use of a phase-coherent carrier tracking loop for each modulation techique. In a mobile radio environment, noncoherent detection is much easier to implement than coherent detection is. A form of detection that is intermediate in difficulty of implementation is called *differential PSK*. Differential PSK does not need absolute carrier phase information, and it circumvents the synchronization problems of coherent detection. The phase reference is obtained by the signal itself, which is delayed in time by an exact bit of spacing. This system maintains a phase reference between successive symbols and is insensitive to phase fluctuation in the transmission channel as long as these fluctuations are small during each duration of a symbol interval T. In differntial binary-phase shift-keying (DBPSK) a symbol is a bit. The weak point of this scheme is that whenever there is an error in phase generated by the medium, two message error bits will result.

There are several aspects of digital detection.[1]

Carrier recovery. Carrier recovery for the suppressed carrier signal $A(t) \sin(\omega_0 t)$ plus noise $n(t)$ can be obtained by two methods. A squaring or frequency-doubling loop can be used (see Fig. 14.3). The loop contains a phase-locked loop as shown in Fig. 14.4. The phase-locked loop maintains a constant phase ϕ_n of $\cos(2\pi f_0 t + \phi_n)$, which is the recovered carrier. Another carrier-recovery technique uses the Costas loop, which generates a coherent phase reference independent of the

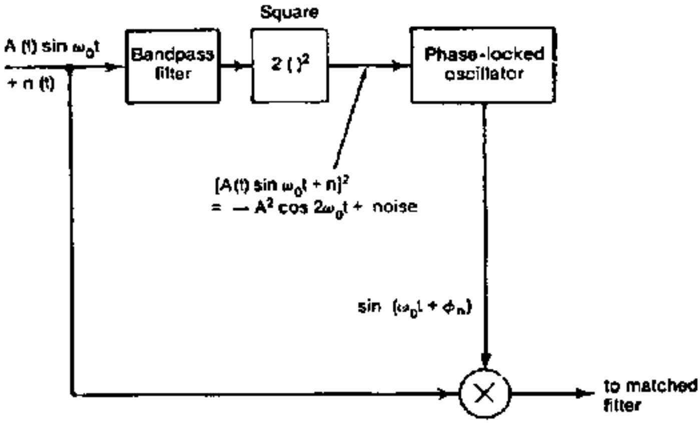

Figure 14.3 Block diagram of the square-law carrier-recovery technique.

binary modulation by using both in-phase and quadrature channels. The Costas loop (Fig. 14.5) is often preferred over the squaring loop because its circuits are less sensitive to center-frequency shifts and are generally capable of wider bandwidth operation. In addition, the Costas loop results in circuit simplicity.

Carrier-phase tracking (phase-locked loop). Carrier-tracking accuracy depends on several system parameters, including the phase noise in the carrier introduced by various oscillator short-term stabilities, carrier-frequency drifts, carrier-tracking-loop dynamics, transient response, the acquisition-performance requirement, and the signal-to-

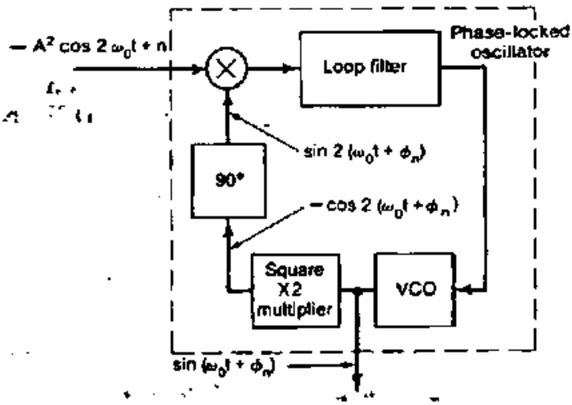

Figure 14.4 Phase-locked oscillator.

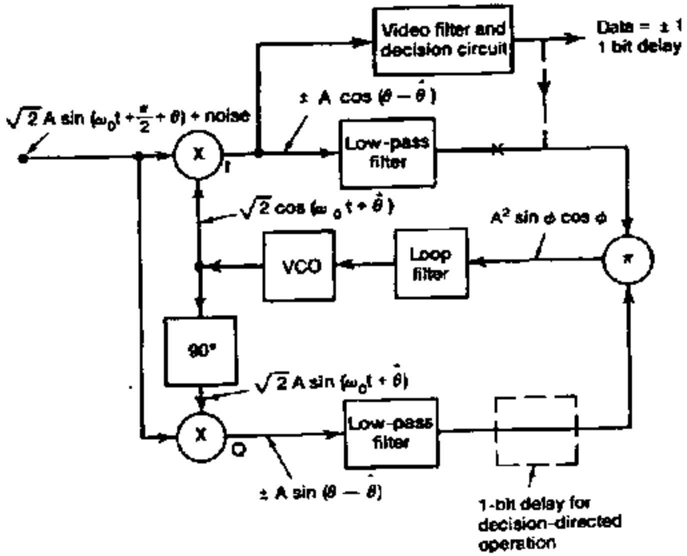

Figure 14.5 Costas loop OSK carrier-recovery circuit and bit detector; the phase error is defined as $\phi \triangleq \theta - \hat{\theta}$. The decision-directed configuration is shown in dashed lines.

noise ratio S/N in the carrier-tracking loop. The phase-locked loop in the carrier-recovery tracking loop must have sufficient noise bandwidth to track the phse noise of the carrier. For a given carrier-phase-noise spectrum, one can compute the phase-locked-loop noise bandwidth B required to track the carrier. Clearly, too large a noise bandwidth can permit the occurrence of the thermal noise effect.

Phase-equalization circuits—for cophase combining

1. *Feedforward.* A circuit using two mixers to cancel random FM can be used (see Fig. 14.6*a*) as a phase-equalization circuit in each branch of an N-branch equal-gain diversity combiner.
2. *Feedback.* A modified circuit from the feedforward circuit is shown in Fig. 14.6*b*. This circuit is also used for each branch of an N-branch equal-gain diversity combiner. The feedback combiner is also called a Granlund combiner.
3. *The total combining circuit.* As shown in Fig. 14.7, either a feedback or a feedforward circuit can be used in the combiner to form a two-branch equal-gain combiner. The circuit connects to a coherent match-filter receiver for BPSK as shown in Fig. 14.8.

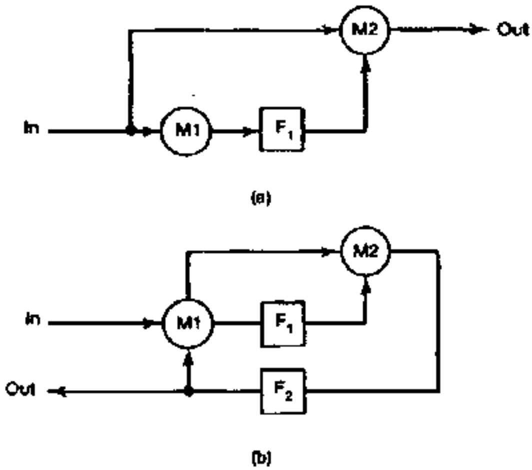

Figure 14.6 Cophase combining techniques. (*a*) Feed-forward cophase combining; (*b*) feedback cophase combining.

Bit synchronization. Power-efficient digital receivers require the installation of a bit synchronizer. Bit synchronization commonly applies self-synchronization techniques, that is, it extracts clock time directly from a noisy bit stream. There are four classes of bit synchronizer.

1. *Nonlinear-filter synchronizer.* This open-loop synchronizer is commonly used in high-bit rate links which normally operate at high signal-to-noise ratios.
2. *The data-transition tracking synchronizer.* This closed-loop synchronizer combines the operations of bit detection and bit synchro-

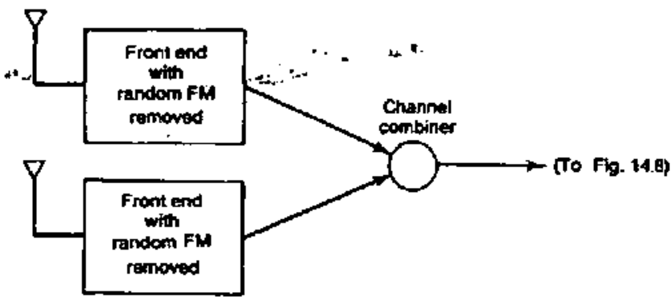

Figure 14.7 A two-branch diversity receiver.

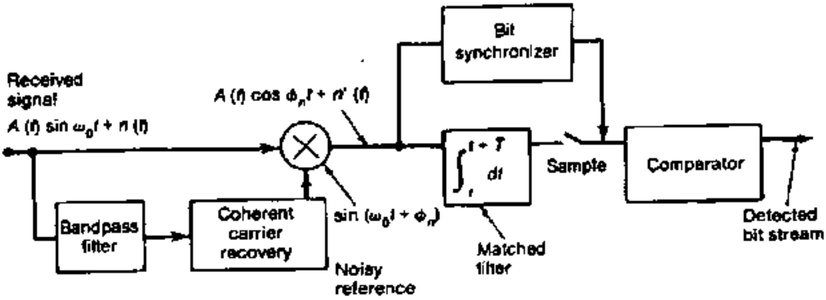

Figure 14.8 Coherent matched-filter receiver for BPSK.

nization. It can be employed at low signal-to-noise ratios and medium data rates.

3. *Early-late synchronizer.* This synchronizer uses early- and late-gate integral and dump channels, which have absolute values. It is simpler to implement than the data-transition tracking synchronizer and less sensitive to dc offsets.

4. *Optimum synchronizer.* This synchronizer provides an optimal means of searching for the correct synchronization time slot during acquisition. However, this approach generally is not practical.

14.2.2 Modulation for digital systems

There are several aspects of digital modulation, and they are described below.

Requirements. Basic digital modulation techniques are amplitude-shift keying (ASK), frequency-shift keying (FSK), phase-shift keying (PSK), and hybrid modulation techniques involving amplitude, frequency, and phase-shift keying.

In mobile cellular systems, the selection of a digital modulation for radio transmission involves satisfaction of the following requirements: (a) narrower bandwidths, (b) more efficient power utilization, and (c) elimination of intermodulation products.

Narrower bandwidths. For all the forms of modulation, it is desirable to have a constant envelope and, therefore, utilize relatively narrower bandwidths. In these cases, FSK and PSK are recommended. For example, multiphase-shift-keying (MPSK) for large values ($M > 4$) has greater bandwidth efficiency than does BPSK or QPSK but power use is less efficient.

More efficient power utilization. It is preferable to provide more channels for a given power level. Therefore, enhanced power utilization is essential. Besides, the FCC has limited the total power (100 W) to be radiated from each base-station antenna. This limitation governs the number of channels which can be served given the power allowed for each channel.

Elimination of intermodulation products. QPSK is commonly used with a transmission efficiency of about 1 to 2 bps/Hz. This value has been found to offer satisfactory trade-off between efficient frequency utilization and transmitter power economy. However, in mobile radio links, when nonlinear class C power amplifiers are used, any spurious radiation should be suppressed. For reducing the spurious signals, we are selecting a constant or low-fluctuation envelope property. There are two types of broadly classified modulations.

Modulation schemes. There are several modulation schemes.

1. *Modified QPSK.* There are two kinds of QPSK besides a regular QPSK with restricted phase-transition rules.
   a. *QPSK.* The conventional QPSK shown in Fig. 14.9*a* has phase ambiguity. The ideal QPSK signal waveform

$$A \sin[\omega_0 t + \theta_m(t)] = \pm\frac{A}{\sqrt{2}} \sin\left(\omega_0 t + \frac{\pi}{4}\right) \pm \frac{A}{\sqrt{2}} \cos\left(\omega_0 t + \frac{\pi}{4}\right) \tag{14.2-1}$$

   where $\theta_m = (0, \pi/2, \pi, 3\pi/2)$ and the value of θ_m should match the sign of Eq. (14.2-1); that is, $\theta_m = 0$ for $(+, +)$, $\theta_m = \pi/2$ for $(+, -)$, $\theta_m = \pi$ for $(-, +)$, and $\theta_m = 3\pi/2$ for $(-, -)$.

Bit pair	Absolute phase
0 0	0
0 1	$\pi/2$
1 1	π
1 0	$3\pi/2$

   b. *Offset QPSK (OQPSK).* This scheme is a QPSK, but the even-bit stream is delayed by a half-bit interval with respect to the odd 1 bit as shown in Fig. 14.9*b*.
   c. *$\pi/4$-shift QPSK.* A phase increment of $\pi/4$ is added to each symbol shown in Fig. 14.9*c*.

   Both OQPSK and $\pi/4$ shift QPSK have no π-phase transition; therefore, no phase ambiguity would occur as in QPSK. However,

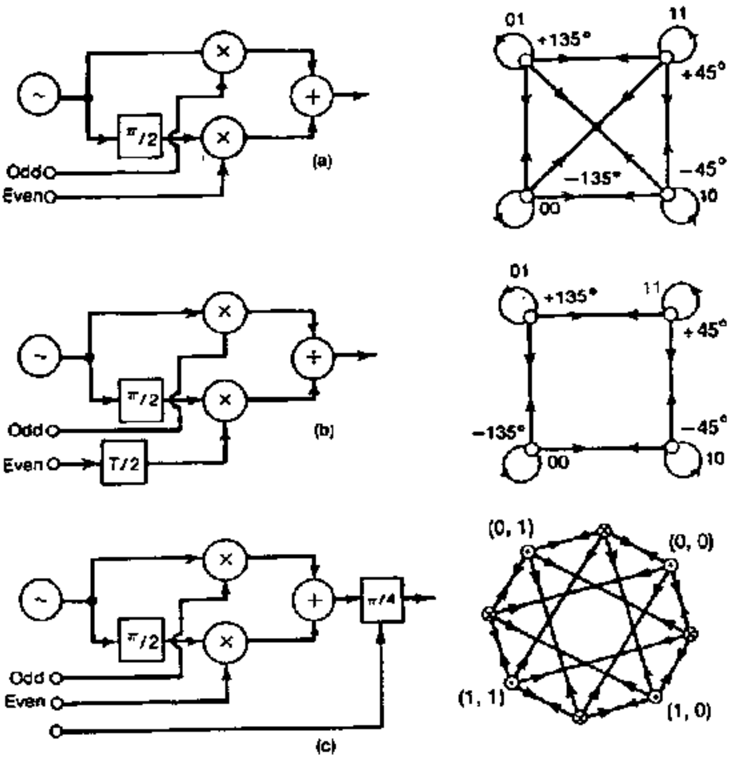

Figure 14.9 Modulator constitution and signal-space diagrams of (*a*) conventional QPSK, (*b*) offset QPSK, and (*c*) $\pi/4$ shift. *(After Hirade et al., Ref. 2, p. 14.)*

intrinsically they produce a certain amount of residual envelope fluctuation. Sometimes, a phase-locked loop (PLL) is inserted at the modulation output to remedy this problem.

2. *The differential encoding of QPSK (DQPSK).* This is the same as in DBPSK, but the differential encoding of the bit pairs selects the phase change rather than the absolute phase. However, DQPSK has phase ambiguity just like QPSK.

Bit pair	Phase changes
0 0	0
0 1	$\pi/2$
1 1	π
1 0	$3\pi/2$

The QPSK carrier recovery would be slightly different than that of BPSK.[2]

3. *Modified FSK—continuous-phase frequency-shift-keying (CP-FSK) with low modulation index*
   a. Minimum-shift-keying (MSK)—also called *fast FSK* (FFSK)
   b. Sinusoidal FSK(SFSK)
   c. Tamed FSK (TFSK) or tamed frequency modulation (TFM)
   d. Gaussian MSK (GMSK)
   e. Gaussian TFM (GTFM)

Since all the schemes listed above are CP-FSK (and have a low modulation index, they intrinsically have constant envelope properties, unless severe bandpass filtering is introduced to the modulator output. In MSK the frequency shift precisely increases or decreases the phase by 90° in each T second. Thus, the signal waveform is

$$s(t) = \sin\left(\omega_0 t + 2\pi\int_0^t s_i\,d\tau + \frac{n\pi}{2}\right) \qquad 0 < t < T$$

where

$$s_i = \begin{cases} s_1 = \dfrac{1}{4T} & \text{for a data bit 1} \\ s_2 = -\dfrac{1}{4T} & \text{for a data bit 0} \end{cases} \tag{14.2-2}$$

or

$$s(t) = \sin\left(\omega_0 t + \frac{n\pi}{2} \pm \frac{\pi t}{2T}\right) \tag{14.2-3}$$

or

$$s(t) = \cos\left(\pm\frac{\pi t}{2T}\right)\sin\left(\omega_0 t + \frac{n\pi}{2}\right) + \sin\left(\pm\frac{\pi t}{2T}\right)\cos\left(\omega_0 t + \frac{n\pi}{2}\right)$$

Comparing Eq. (14.2-1) with Eq. (14.2-3), we find that the two equations are very similar. In fact, the phase-modulation waveforms of the I- and Q-channel modulations of OQPSK are modulated by sine and cosine waveforms, and thus the output will be identical to that of MSK. Note that it is necessary to modulate both the I and Q channels during each bit interval to retain the constant envelope of $s(t)$.

Because the phase is continuous from bit to bit, the spectral sidebands of MSK or OQPSK fall off more rapidly than in BPSK or QPSK (see Fig. 14.10). Although MSK demonstrates a superior property in

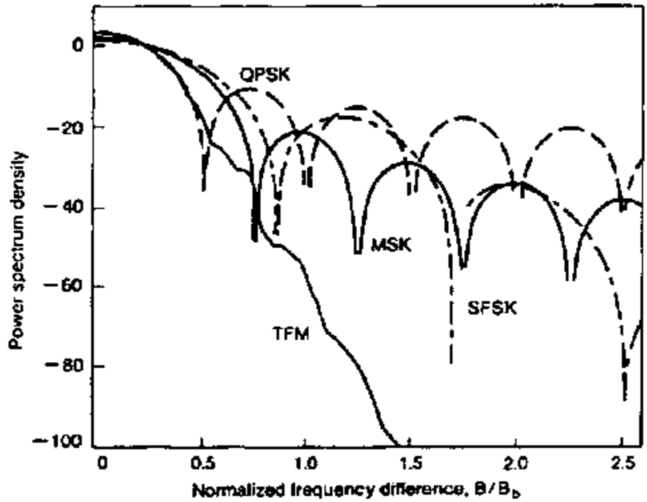

Figure 14.10 Power spectrum density functions of QPSK, MSK, SFSK, and TFM. *(After Hirade et al.*, Ref. 2, p. 15.)

terms of its out-of-band spurious spectrum suppression without any filtering, its out-of-band spurious spectrum suppression will not satisfy the severe requirements in the single carrier per channel (SCPC) communications. The sharp edges in MSK phase-transition trajectories (Fig.14.11) can be smoothed by some premodulation baseband filtering. The SFSK shows a smoother phase transition than does the MSK but little improvement in the suppression of out-of-band spurious spectrum. The TFM is a modified MSK using the partial response encoding rule as the phase-transition rule. The smoothed phase trajectory of TFM is shown in Fig.14.11. The outstanding suppressions for the out-of-band spectrum of TFM is shown in Fig. 14.10. However, this outstanding suppression of the out-of-band spurious spectrum can be achieved by using a suitable premodulation baseband filtering on MSK, such as a baseband gaussian filtering, as shown in Fig. 14.12. The GMSK with $B_bT = 0.2$ has the same power spectrum density curves as does TFM, where B_b is the baseband bandwidth. Furthermore, GMSK is easier to implement than TFM.

The following parameters are defined and are in the figures that follow.

B_i ideal bandpass

B_s channel separation (bandwidth)

f_b bit rate of voice coding = $1/T$

$1/T$ transmission bit rate (16 kbps)

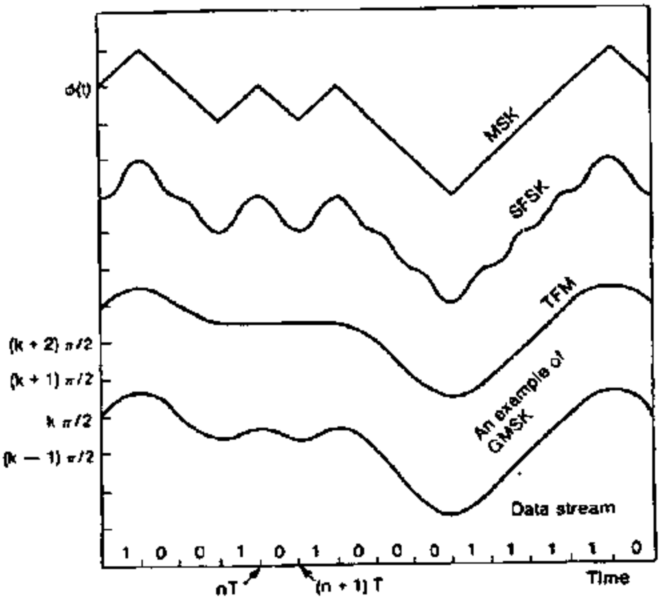

Figure 14.11 Phase-transition trajectories of MSK, SFSK, TFM, and GMSK *(After Hirade et al., Ref. 2, p. 15.)*

B_iT normalized bandwidth of the ideal bandpass filter
b_bT normalized bandwidth of a gaussian filter

Figure 14.13 shows the fractional power in percentage of GMSK signal exceeding the normalized bandwidth B_iT with different values of B_bT ($B_bT = \infty$ means no filter). This becomes conventional MSK. Figure 14.14 shows that the relative power radiated in the adjacent channel for B_sT is equal to 1.5. Let B_s = 30 kHz, then the bit rate $f_b = (1/T) = 20$ kbps. For a normalized filter bandwidth $B_bT = 0.24$ or B_b = 4.8 kHz, and the relative power in the adjacent channel is −60 dB.

4. 16-QAM and code modulation—these two modulations are not constant envelop modulations, but both can achieve better spectrum efficiency. The readers can read Proakis' book[19] if interested.

Demodulation. When the signal is received, we would like to know the performance of the various demodulation schemes. Some demodulation schemes are better than others regardless of what the modulated signal is like. The orthogonal coherent detector proposed by de Buda[3] can be used for both MSK and TFM.

BER performance is always a good criterion for the comparison of different modulation schemes. Measured BER performance is shown

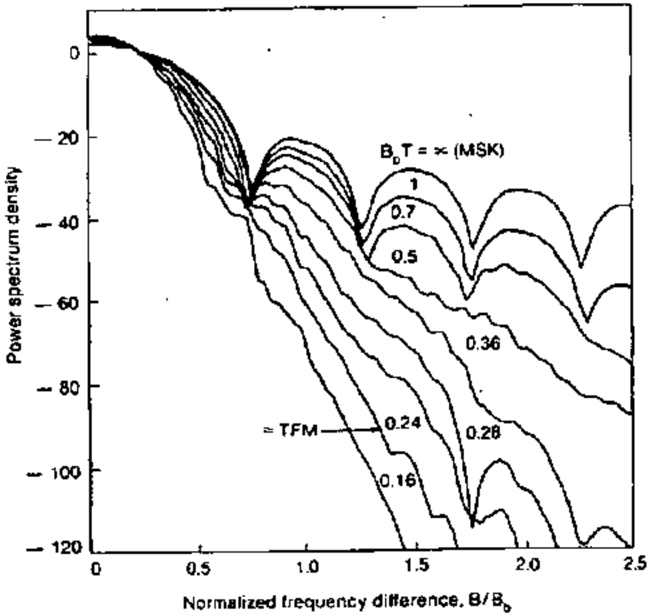

Figure 14.12 Power spectrum density functions of GMSK. *(After Hirade et al., Ref. 2, p. 15.)*

in Fig. 14.15 with a normalized channel bandwidth $B_iT = 0.75$. The family of curves show the BERs for the different filter bandwidths. Also, TFSK (TFM) is plotted for comparison. GMSK with the filter bandwidth of $B_bT = 0.19$ is superior to TFSK. However, one disadvantage of narrowing the channel spectrum is increasing the BER, that is, degrading performance. Sometimes we have to consider whether it is worthwhile to use a gaussian filter.

14.3 ARQ Techniques

14.3.1 Different techniques[4,5]

Automatic-repeat-request (ARQ) techniques include the coding and retransmission request strategy for delivering a message. Preceding the message is a header which contains the source and destination address and useful routing information. Every ARQ message must have a header. There are two principal ARQ techniques.

1. *Stop-and-wait ARQ.* The message originator stops at the end of each transmission to wait for a reply from the receiver (see Fig. 14.16*a*). Then the following steps can be taken.

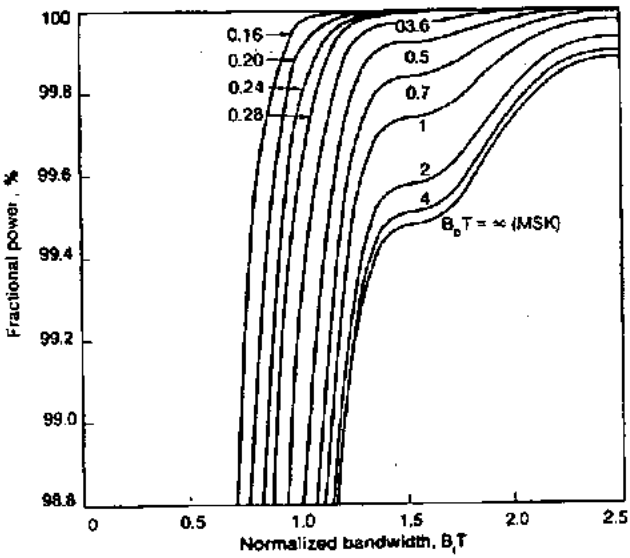

Figure 14.13 Fractional power of GMSK signal. *(After Hirade et al., Ref. 2, p. 15.)*

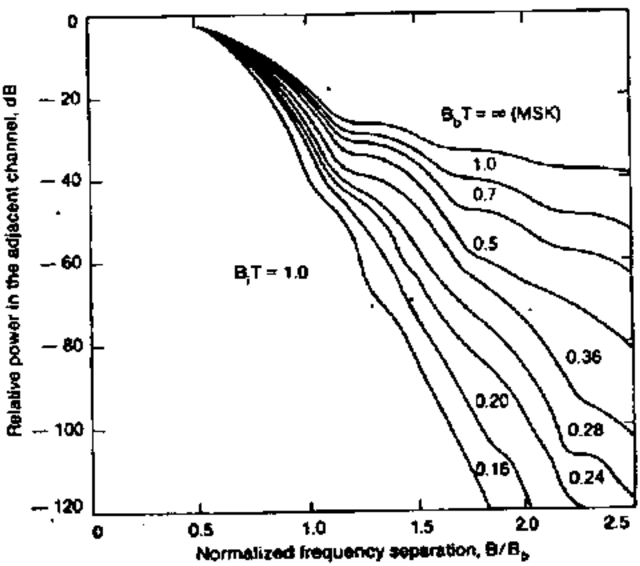

Figure 14.14 Relative power radiated in the adjacent channel. *(After Hirade et al., Ref. 2, p. 16.)*

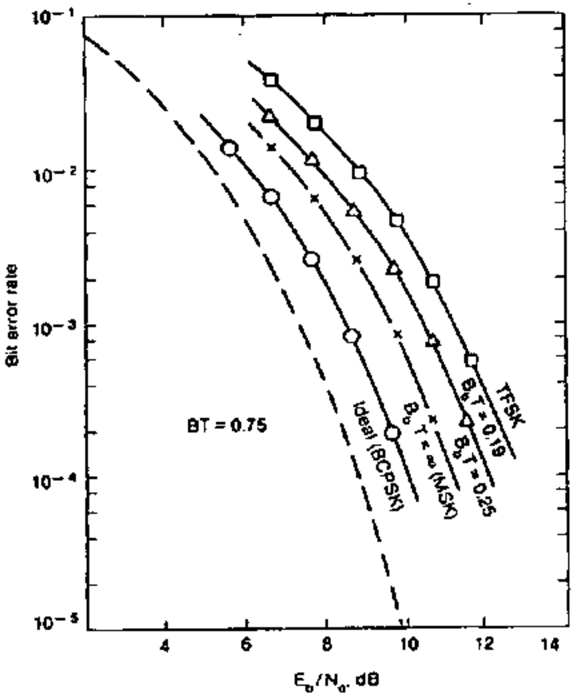

Figure 14.15 Measured BER performance. *(After Hirade et al., Ref. 2, p. 17.)*

a. No forward error correction is used—ARQ(*a*).

b. Both forward error correction and error detection coding are used—ARQ(*b*).

c. Error-detection parity bits are sent, but not the forward error-correction parity bits, which assumes that the probability of an error-free message is great—ARQ(*c*).

2. *Selective retransmission.* When many words are transmitted at once, each word individually can apply error detection, not the message as a whole. Only those words containing detected errors are sent back. This scheme is called *selective retransmission*. Selective retransmission with ARQ(*b*) is shown in Fig. 14.16*b*. Selective retransmission with ARQ(*c*) is shown in Fig. 14.16*c*.

14.3.2 The expected number of transmissions

Stop-and-wait ARQ [apply ARQ(*a*) and ARQ(*b*) only]. Let P_{ew} be a word error rate (WER),and let a message consist of N words. Now the re-

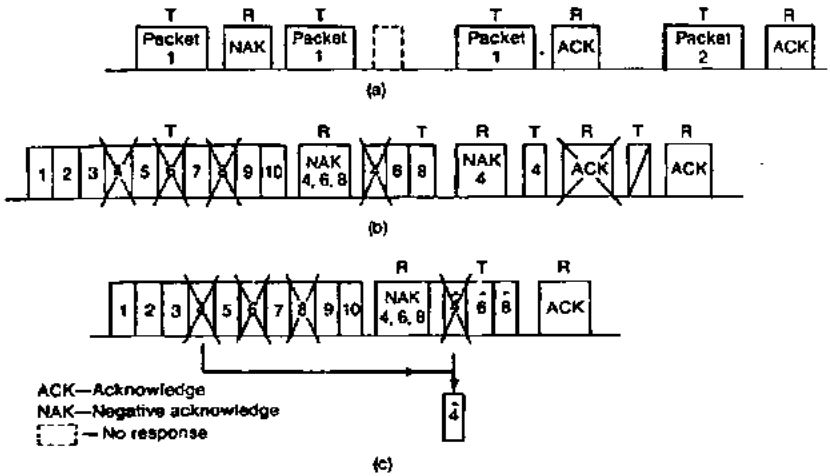

Figure 14.16 ARQ transactions. (*a*) Stop-and wait ARQ; (*b*) selective retransmission with ARQ(*b*); (*c*) selective retransmission with ARQ(*c*).

quired number of transmissions depends on all N words being successfully transmitted. The expected number of transmissions is

$$E_N = \frac{1}{(1 - P_{ew})^N} \tag{14.3-1}$$

assuming independence errors between words and that all words have the same P_{ew}. This assumption can be considered valid for cases where vehicle speed is high. Equation (14.3-1) indicates that the number of transmissions E_N increases more quickly with increasing the message length, that is, as N increases. Equation (14.3-1) is plotted in Fig. 14-17*a*.

Selective retransmission (ST) with ARQ(b). Assume that the number of transmissions of one word is independent of the number of transmissions of any other word. The expected number of transmissions required for sending an N-word message with fewer than i transmissions is

$$E_N = \sum_{i=1}^{\infty} [1 - (1 - P_{ew}^{i-1})^N] \tag{14.3-2}$$

Equation (14.3-2) is plotted in Fig. 14.17*b*. By comparing Fig. 14.17*a* with Fig. 14.17*b*, we see that stop-and-wait ARQ would require a greater number of transmissions to deliver a message than would selective retransmission with the same block error probability.

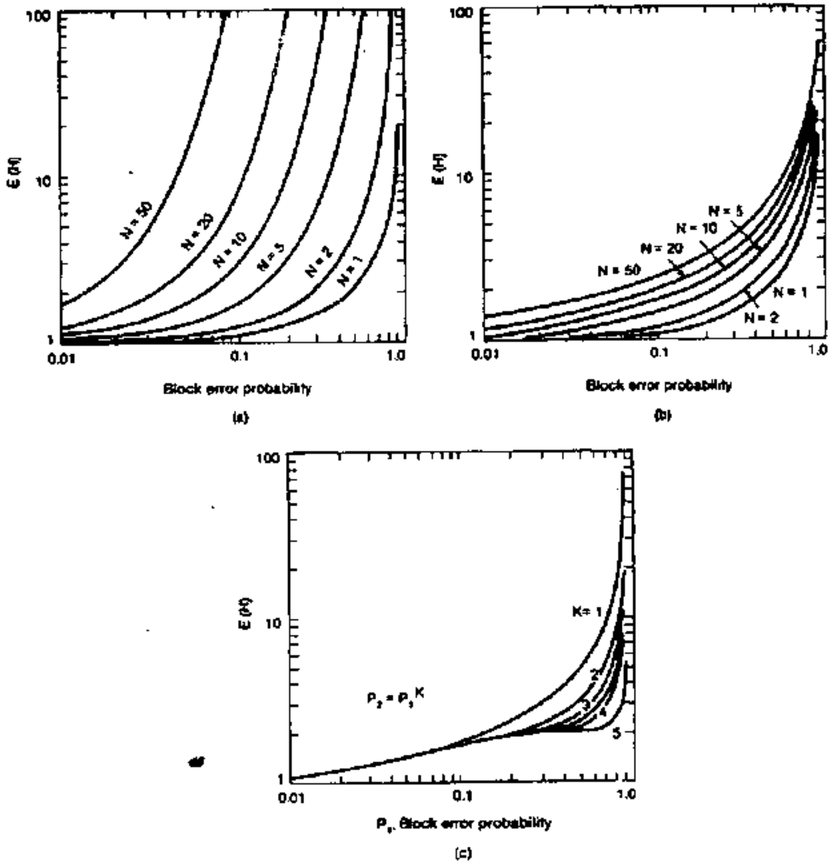

Figure 14.17 Expected number of transmissions for different kinds of ARQ. (*a*) $E(H)$, expected number of transmissions to deliver an N-block message for stop-and-wait ARQ. (*b*) $E(H)$, expected number of transmissions to deliver an N-block message for SRT ARQ(*b*). (*c*) $E(H)$, expected number of transmissions to deliver a 10-block message for SRT ARQ(*c*) for various retransmission-failure probabilities.

Selective retransmission (SRT) with ARQ(*c*). In this scheme, ARQ(*c*) defined in the stop-and-wait ARQ is applied to the selective retransmission scheme. Let the first-transmission probability be P_1 and a retransmission probability be P_2. The expected number of transmissions required for sending an N-word message with fewer than i transmissions is

$$E_N = 1 + \sum_{i=2}^{\infty} [1 - (1 - P_1 P_2^{i-2})^N] \qquad (14.3\text{-}3)$$

Because the forward error correction is added to retransmission, P_2 is smaller than P_1. Let P_2 be a positive integer power of P_1; that is, $P_2 = P_1^k$, where k represents the various powers of the feedforward error correction code. Equation (14.3-3) is plotted in Fig. 14.17*c* for $N = 10$. When $k = 1$, the curve shown in Fig. 14.17*c* is the same as $N = 10$ in Fig. 14.17*b*.

14.3.3 Transmission Efficiency *R*

The transmission efficiency is the ratio of the number of information bits to the total number of transmission bits. Let a message consist of N words, where B is number of bits per word, and

$$B = H + L \tag{14.3-4}$$

where H is the header bits and L is the information bits. Then

Stop-and-wait ARQ technique

$$R = \frac{NL}{(H + NB)E_N} \tag{14.3-5}$$

where E_N is as shown in Eq. (14.3-1).

Selective retransmission

$$R = \frac{NL}{HE_N + NBE_1} \tag{14.3-6}$$

where E_N is as shown in Eq. (14.3-2) and

$$E_1 = \frac{1}{(1 - P_{ew})} \tag{14.3-7}$$

Selective retransmission with ARQ(c)

$$R = \frac{NL}{HE_N + NBE_1} \tag{14.3-8}$$

where E_N is as shown in Eq. (14.3-3), and E_1 is

$$E_1 = 1 + \sum_{i=2}^{\infty} [1 - (1 - P_1 P_2^{i-2})] \tag{14.3-9}$$

14.3.4 Undetected error rates

In the previous sections, we saw that each transmission results in one of two outcomes, success or failure. This is called "hard detection." There is also a "soft detection." In soft detection three probabilities per transmission are denoted: P_c (success), P_d (detected error), and P_u (undetected error) per word. When each of these probabilities is identified for each transmission, then we can perform further calculations.

Stop-and-wait ARQ [apply ARQ(*a*) and ARQ(*b*)]. We can find the undetected error probability per single-word message.

$$P_{\mathrm{um}} = \frac{P_u}{P_c + P_u} \tag{14.3-10}$$

where

$$P_c = 1 - P_{\mathrm{ew}} \tag{14.3-11}$$

and

$$P_u \leq P_{\mathrm{ew}}\, 2^{-m} \tag{14.3-12}$$

where m is the number of error-detection parity bits. The undetected error probability per N-word message is

$$P_{\mathrm{um}} = \frac{1 - (1 - P_{\mathrm{ew}})^N}{1 + (1 - P_{\mathrm{ew}})^N(2^m - 1)} \tag{14.3-13}$$

Selective retransmission. In this technique, we consider b parity bits per each word instead of m parity bits for the whole message. The probability of undetected error for a single-word message is

$$P_{\mathrm{um}} = \frac{P_{\mathrm{ew}}}{P_{\mathrm{ew}}(1 - 2^b) + 2^b} \tag{14.3-14}$$

The probability of undetected error for a N-word message is

$$P_{\mathrm{um}} = 1 - \left[1 - \frac{P_{\mathrm{ew}}}{P_{\mathrm{ew}}(1 - 2^b) + 2^b}\right]^N \tag{14.3-15}$$

The word error rate P_{ew} is a function of vehicle speed. Since the word error rate P_{ew} is derived from the bit error rate, each bit in error can be dependent on or independent of the adjacent bit error related to the vehicle speed. Therefore, the word error rate is not easy to obtain. Any oversimplified model may give incorrect answers. We can obtain the word error rate from two extreme cases, one assuming that the speed approaches infinity and the other assuming that the speed approaches zero as described in Chap. 13.

14.4 Digital Speech[6]

Since digital technologies have evolved, an important study focuses on efficient methods for digitally encoding speech. Speech quality implies a measure of fidelity, which is difficult to specify qualitatively because human perception is involved. The two criteria used are

What is being said (low fidelity accepted by military systems)

Who says it (high fidelity important to commercial systems) .

For instance, a military system examiner who comments that someone's speech quality is excellent may be referring to intelligibility and low system noise. It is irrelevant who speaks on the other side (of the examiner) or that the examiner has never spoken to this person before.

14.4.1 Transmission rates in speech coding

These rates are totally dependent on quality characterizations such as toll quality, commentary quality, communications quality, synthetic quality. We may use the mean opinion score (MOS) referred in Sec. 1.5.1 to grade the voice quality.

1. *Toll quality ($4 < MOS < 4.5$).* An analog speech signal is of toll quality when its frequency range is 200 to 3200 Hz; its signal-to-noise ratio is greater than or equal to 30 dB; and its harmonic distortion is less than or equal to 2.3 percent. Digital speech has to have a quality comparable to that of the toll quality of an analog speech signal.

Toll-Quality Transmission

Coder	kbps
Log PCM	56
ADM	40
ADPCM	32
Sub-band	24
Pitch Predictive ADPCM	24
APC, ATC, ΦV, VEV	16

2. *Commentary quality ($MOS > 4.5$).* In general, the signal at bit rates exceeding 64 kbps generates a commentary-quality speech

signal which is better than toll quality, but the input bandwidths are significantly wider than in a noncellular telephone system (up to 7 kHz).

3. *Communications quality* ($3 < MMOS < 4$). At rates below 16 kbps, the signal in the range of 7.2 to 9.6 kbps is a communications-quality speech signal. The signal is highly intelligible but has noticeable reductions in quality and speaker recognition.

Communincations-Quality Transmission

Coder	kbps
Log PCM	36
ADM	24
ADPCM	16
Sub-band	9.6
APC, ATC, ΦV, VEV	7.2

4. *Synthetic quality* ($2.5 < MOS < 3$). At 4.8 kbps and below, the signal provides synthetic quality and speaker recognition is substantially degraded.

Synthetic-Quality Transmission

Coder	kbps
CV, LPC	2.4
Orthogonal	1.2
Formant	0.5

14.4.2 Classes of coder

There are two classes of coder: waveform coders and source coders.

Waveform coders. The speech waveform can be characterized by

1. Amplitude distribution (in time domain)
2. Autocorrelation function (in time domain)
3. Power spectral density (in frequency domain)
4. Spectral flatness measure (removing redundancy in speech waveform)
5. Fidelity criteria for waveforms

$$\text{Coding noise} = \frac{1}{T}\int_0^T (\text{coding error})^2\, dt$$

where the coding error is equal to the amplitude difference (samples of a coded waveform minus the original input waveform).

The signal-to-noise ratio is expressed as

$$\frac{S}{N} = \overline{\left[\frac{(\text{input waveform})^2}{\text{coding noise}}\right]}$$

There are two types of speech waveform coders.

1. *Time-domain coders.* Pulse code modulation (PCM), differential pulse code modulation (DPCM), and delta modulation (DM) are commonly used. Adaptive predictive coding (APC) in time-domain coding systems is limited to linear predictors with changing coefficients based on one of the following three types:
   a. Spectral fine structure—in more periods
   b. Short-time spectral envelope—determined by the frequency response of the vocal tract and by the spectrum of the vocal-cord sound pulses
   c. Combination of types *a* and *b*

   In time-domain coders, speech is treated as a single full-band signal: in time-domain predictive coders, speech redundancy is removed prior to encoding by prediction and inverse filtering so that the information rate can be lower.
2. *Frequency-domain coders.* The speech signal can be divided into a number of separate frequency components, and each of these components can be encoded separately. The bands with little or no energy may not be encoded at all. There are two types of coding:
   a. *Subband coding (SBC).* Each subband can be encoded according to perceptual criteria that are specific to that band.
   b. *Adaptive transform coding (ATC).* An input signal is segmented and each segment is represented by a set of transform coefficients which are separately quantized and transmitted.

Source coders—vocoders. The synthetic quality of source vocoder speech is not appropriate for commercial telephone application. It is designed for very low bit-rate channels. Vocoders use a linear, quasi-stationary model of speech production.

Sound source characteristics. The sound can be generated by voiced sounds, fricatives, or stops. The source for voiced sounds is represented by a periodic pulse generator. The source for unvoiced sounds is represented by a random noise generator. They are mutually exclusive.

System characterization. The acoustic resonances of the vocal tract modulate the spectra of the sources. Different speech sounds correspond uniquely to different spectral shapes. Vocoders depend on a parametric description of the vocal-tract transfer functions.

1. Channel vocoder—speech signal evaluated at specific frequencies
2. LPC (linear prediction code) vocoder—linear prediction coefficients that describe the spectral envelope
3. Formant vocoder—specified frequency values of major spectral resonances
4. Autocorrelation vocoder—specified short-time autocorrelation function of the speech signal
5. Orthogonal function vocoder—specifies a set of orthonormal functions

Frequency-domain vocoders. A single coder is called a *channel vocoder.* Instead of transmitting the telephone signal directly, only the spectrum of each speech signal is transmitted; 16 values along the frequency axis are needed. Each takes 20 ms and requires a bandwidth of $1/(2 \times 20 \text{ ms}) = 25$ Hz and the total frequency requirement is (16×25) or 400 Hz, which is one-tenth of the bandwidth of the speech signal itself.

Time-domain vocoders. Speech samples would have to be spaced $1/(2 \times 4000) = 0.125$ ms apart, which would require 30 samples to ensure a good quality. Then the frequency requirement is $30/(2 \times 0.125 \text{ ms})$ or 120 kHz. For digital transmission, the number of bits per correlation sample used by a time-domain vocoder should be about twice as high for spectral samples in frequency-domain vocoders. Therefore, time-domain vocoders are not desirable. Yet, one of the most successful innovations in speech analysis and synthesis is linear predictive coding (LPC), which is based on autocorrelation analysis.

LPC vocoders. LPC vocoders constitute an APC system in which the prediction residual has been replaced by pulse and noise sources. For the telephone band, the number of predictor coefficients is 8. For low-quality voice, the number can be as small as 4. The RELP (regular-pulse excited LPC) used by GSM, the VSELP (Vector-sum excited LPC) used by TDMA (IS-54) and the modified VSELP used by CDMA all have ten coefficients.

Hybrid waveform coders-vocoders. A hybrid arrangement of SBC, APC, and LPC is coming into vogue where a portion (lower-frequency band) of the transmission is accomplished by waveform techniques and a portion (upper-frequency band) by voice-excited vocoder techniques.

14.4.3 Complexity of coders

A relative count of logic gates is used to judge the complexity of the coders as follows:

Relative complexity		Coder
1	ADM:	adaptive delta modulator
1	ADPCM:	adaptive differential PCM
5	Sub-band:	subband coder (with CCD filters)
5	P-P ADPCM:	pitch-predictive ADPCM
50	APC:	adaptive predictive coder
50	ATC:	adaptive transform coder
50	ΦV:	phase vocoder
50	VEV:	voice-excited vocoder
100	LPC:	linear-predictive coefficient (vocoder)
100	CV:	channel vocoder
200	Orthogonal:	LPC vocoder with orthogonalized coefficients
500	Formant:	formant vocoder
1000	Articulatory:	vocal-tract synthesizer; synthesis from printed English text

Of these coders, LPC is attractive because of its performance and degree of complexity.

14.5 Digital Mobile Telephony

14.5.1 Digital voice in the mobile cellular environment

Since voice communication is the key service in cellular moble systems, when we think of the digital systems, we must think of a digital voice.

In present-day mobile cellular systems, transmission of a digital voice in a multipath fading environment is a challenging job. The major considerations in implementing digital voice in cellular mobile systems are discussed below, along with a tentatively recommended transmission rate for the cellular mobile system.

Digital voice in the mobile radio environment

1. The criterion for judging a good digital voice through a wire line is employed in three existing digital voice schemes.
   a. In a continuously variable step delta (CVSD) modulation scheme, the present transmission rate is 16 kbps. This is not toll-quality voice transmission and is commonly used by the military.

b. In a LPC scheme, the present transmission rate of 2.4 kbps provides a synthetic quality voice, but a rate of 4.8 kbps using vector quantization[8] may provide a communications-quality voice. A rate of 16 kbps can provide a toll-quality voice.[19]
c. In a pulse code modulation (PCM) scheme, the present transmission rates of 32 kbps and 64 kbps are commonly used; 32 kbps is used by the military while 64 kbps is used commercially.

Of the three schemes, LPC seems most attractive because of its low transmission rate. However, LPC is more vulnerable in terms of distortion to the mobile fading environment.
2. Digital voice has to be processed in real time, which imposes constraints on the digital processing time. This adversely affects LPC but not CVSD.
3. When sending a digital stream (voice) through a radio channel in a fading environment, in general, an LPC scheme needs more code protection than CVSD scheme does because LPC is not implemented in a continuous waveform in either the frequency domain or the time domain while CVSD is implemented in a continuous waveform in the time domain.
4. Because the mobile unit is moving, sometimes rapidly, sometimes slowly, insertion of extra synchronization bits is needed in the normal digital stream.

Considerations for a digital voice transmission in cellular mobile system. The following factors are significant.

1. *Digital transmission rate*
   a. *Present cellular signaling rate.* The present signaling format is designed on the assumption that the mobile unit moves at an average of 30 mi/h and that the transmission rate is 10 kbps. The 21 synchronization bits (10 synchronization bits and 11 frame bits) occur in front of every code word of 48 bits to ensure that the bits are not falling out of synchrony before the resynchronization takes place.
   b. *Consideration of LPC scheme.* If a rate of 4.8 kbps using LPC for a communications-quality voice is accepted its rate is almost half of the present transmission rate, and at this transmission rate a 48-bit word would be acceptable in a fading environment. The resynchronization scheme for a mobile receiver should take place in front of every code word of 48 bits [(21 synchronization bits) + (a code word of 48 bits) = 69 bits]. The number of synchronization bits is almost half the number of bits in a code word. Therefore, the transmission rate would be approximately $(4.8 \times 1.5) = 7.2$ kbps.

c. *Redundancy of transmission.* The protection of synchronization in a mobile radio environment is not sufficient. If the digital stream were to occur in a signal fade, partial or whole code words would be lost. In order to prevent fading, redundancy of transmission is often used. We would take a minimum redundancy scheme; for example, we would transmit the same message bits three times and take a "2-out-of-3 majority vote" on each bit to minimize the fading impairment of the message bits. For LPC of 4.8 kbps, an RF transmission rate of (4.8 kbps × 1.5) 3 = 21.6 kbps is needed.* It is reasonable for a 30-kHz channel to carry a transmission rate of 21.6 kbps over a severe fading medium. When an RF transmission rate is given, the channel bandwidth can be narrower with a trade-off of transmitted powers. This point has been described in Sec. 13.4.8.

d. *Modulation, diversity, coding, ARQ, and scrambling.* Diversity and modulation can help in reducing the RF transmission rate for the digital voice. However, ARQ schemes, fancy coding schemes, and complicated scrambling schemes cannot be implemented for voice transmission. This is because the digital voice must be processed in real time, and these three schemes usually require a fair amount of time for processing. These schemes can be used for data transmission.

2. *Word error rate.* In the multipath fading environment, the bit error rate P_e is not the only parameter for voice-quality measurement; the word error rate P_w is also important and varies with vehicle speed. However, information on the word error rate for transmission of digital voice over a mobile radio environment only appears in two extreme cases (see Sec. 13.2.3). Assume that we know the required P_e and P_w. We can convert P_e and P_w to a required carrier-to-noise ratio C/N. If a two-branch diversity scheme is applied after a 2-out-of-3 majority-vote redundancy scheme has been used, the bit error rate of 10^{-3} in a relatively slow fading case requires a C/N level of approximately 15 dB. With the C/N level, a word error rate of a 40-bit word is about 10^{-3} (see Fig.13.3*a*). In general, if the word error rate is the same as or lower than the bit-error rate for a given C/N, the C/N level is acceptable. In our case, P_w and P_e are the same at $C/N = 15$; therefore, the $C/N = 15$ dB is justified.

3. *Relationship between C/N and E_b/N_0.* The relationship between the carrier-to-noise ratio C/N, the energy-per-bit-to-noise-per-hertz ratio E_b/N_0, the transmission rate R, and the bandwidth B can be expressed as

*Applying diversity schemes can reduce this rate.

$$\frac{C}{N} = \frac{E_b}{N_0}\frac{R}{B} \tag{14.5-1}$$

When the number of levels C/N increases, the bandwidth decreases. Keeping E_b/N_0 constant, we see that when the bandwidth decreases, the required carrier-to-noise ratio C/N increases. Previously we calculated that $C/N = 15$ dB works for a two-level (binary) system. If the number of levels increases, the C/N will be higher than 15 dB.

Example 14.1 Let $E_b/N_0 = 15$ dB for a two-level system and R_0 and B_0 be the transmission rate and transmission bandwidth, respectively, of the two-level system. Now if we reduce the bandwidth $B_1 = 0.5B_0$, then

$$\left(\frac{C}{N}\right)_1 = (31.6)\frac{R_0}{0.5B_0} = 2\left(\frac{C}{N}\right)_0$$

$$= \left(\frac{C}{N}\right)_0 + 3\text{ dB}$$

This means that the power increases by 3 dB.If the transmitted power was 50 W, now it is 100 W.

14.5.2 Evaluation of digital voice quality

In general, there are two methods for evaluating digital voice quality.

1. *Listener's opinion.* Use one 16-kbps voice coder and one 8-kbps voice coder in a specified digital system. Then find the two required carrier-to-interference ratios C/I based on the listener's opinion in a Rayleigh fading environment. Then compare the same voice quality with that from an analog FM system at $C/I \geq 18$ dB.
2. *Diagnostic rhyme test (DRT).* The voice quality of a digital format is often tested by DRT. Using the DRT score of 90 as a criterion, above 90 means acceptable for synthetic-quality voice and below 90 means unacceptable. Thus, the bit error should be less than 10^{-3} for an LPC of 2400 bps in a gaussian noise environment. Voice evaluation for an LPC of 4.8 kbps does not appear in the literature. The voice quality in CVSD based on the same DRT criterion requires a bit error rate of only 4 percent or less. The DRT is not designed for toll-quality voice test.

14.6 Practical Multiple-Access Schemes

14.6.1 High-capacity FSK in FDMA

The use of frequency-division multiple access (FDMA) with the present channel bandwidth of 30 kHz is a conventional approach. The high-capacity FSK modulation used in FDMA is based on the rate of digital voice. From the quality transmission of vocoders (see Sec. 14.5.1) we can see that a digital voice must have a 16-kbps coding rate to produce full telephone quality, although the voice quality of a 4.8-kbps LPC with vector quantization may be acceptable for a communications-quality voice (see Sec. 14.5). Now we have to determine the C/I which will provide optimal quality within a given digital system.

Swerup and Uddenfeldt compared a narrowband coherent digital modulation with gaussian MSK to an analog FM system.[11] Two 16-kbps voice coders were used. Residual excited linear predicted codes and subband codes were tested. The digital unit performance can be reduced by 5 dB to obtain the same performance as an analog unit. This 5-dB reduction advantage means a large coverage area and a closed frequency-reuse distance for each cell can be served in a cellular system. This is, in turn, an example of high spectral efficiency usage (described in Sec. 13.4). Consider the following calculations.

1. In an omnidirectional-cell system, assume that $C/I = 13$ dB, i.e.,

$$\frac{C}{I} = \frac{q^4}{6} > 10^{1.3} = 20$$

Solving for q and using Eq. (2.4-5), we obtain

$$q = 3.31 = \sqrt{3K}$$

$$K \approx 4 \qquad \text{(frequency-reuse pattern)}$$

In this case the total number of channels is 333; then

$$m = \frac{333}{4} = 83 \text{ channels/cell}$$

which is higher than the 47 channels per cell for $C/I \geq 18$ dB.

2. In 120° direction cells, we also compare two sets of C/I levels. As shown in Fig. 6.6b, there are only two interference cells. Then we can estimate the distance between the mobile unit and the two interfering sites to be $D + 0.5R$, as mentioned in Sec. 6.5.1.

   a. $C/I \geq 18$ dB.

$$\frac{(q + 0.5)^4}{2} \geq 10^{1.8} = 63$$

or

$$q = 2.85$$

The number of frequency-reuse cells is

$$K = \frac{q^2}{3} = 2.71 \sim 3$$

The number of sectors is $(3 \times 3) = 9$; then

$$m = \frac{333}{9} = 37 \text{ channels/sector}$$

b. $C/I \geq 13$ dB

$$\frac{(q + .05)^4}{2} = 10^{1.3} = 20$$

or $q = 2$. The number of frequency-reuse cells is

$$K = \frac{q^2}{3} = \frac{4}{3} = 1.3 \sim 2$$

The number of sectors $= (2 \times 3) = 6$; then

$$m = \frac{333}{6} = 55.5 \text{ channels/sector}$$

3. In 60° directional cells (see Fig. 6.6c and Sec. 6.5.1)
   a. $C/I \geq 18$ dB

$$\frac{(q + 0.7)^4}{1} \geq 10^{1.8} = 63$$

or

$$q = 2.12$$

$$K = \frac{q^2}{3} = 1.5 \sim 2$$

$$m = \frac{333}{2 \times 6} = 27.75 \text{ channels/sector}$$

   b. $C/I \geq 13$ dB

$$\frac{(q + 0.7)^4}{1} \geq 10^{1.3} = 20$$

or

$$q = 2.11 - 0.7 = 1.41$$

$$K = \frac{q^2}{3} = 0.67 \sim 1$$

$$m = \frac{333}{6} = 55.5 \text{ channels/sector}$$

Apparently, spectrum efficiency is increased by using digital technology.

Discussion. The preceding analysis is based on an ideal situation. It needs to be verified by measurement in a real cellular system. In the future we may achieve a digital system which can narrow the channel bandwidth and increase the transmission rate. The success of such a system would be proved if the reception at the receiving end were the same as that which would be achieved if the medium were nonfading. Modulation schemes, diversity, coding, redundancy, and ARQ can help to achieve this goal.

The signal can be designed using 4-, 8-, or 16-level MFSK or MPSK. Of course, the higher the number of levels, the narrower the channel bandwidth. However, the increase in transmitted power (or the reduction in range) is the key factor.

In an FDMA system, the transmit data rate needs to serve only one voice channel and the rate can be 10 kbps or less. In this circumstance, the equalizer used to reduce the intersymbol interference (ISI) due to high-data-rate transmission is not required in FDMA.

14.6.2 TDMA system[11]

The use of time-division multiple access (TDMA) could be another approach for increasing spectrum channel efficiency. It also has the potential to reduce the cost of both cell-site and mobile terminal equipment. The bulky analog radio equipment can be replaced by very large scale integrated-circuit (VLSI) digital signal processing. This applies to the use of digital equipment in any system; TDMA is no exception. In addition, TDMA will replace the analog duplex filter with a time switch. A zero IF receiver can be used, that is, there can be direct conversion without superheterodyne. The number of radio transceivers can be reduced, and the size of cell site can be much smaller. Assume that a cell site in a TDMA system has a single transceiver and a simple network interface. This is a very attractive feature. The main drawback of TDMA is the accurate clock requirement.

An inaccurate clock results in time jittering due to the instability of the frequency synthesizer at the transmitter and the clock at the receiver. The variable time delay resulting from the change of vehicle positions and from random FM could affect the synchronization and tracking of the bit streams. The most adverse effect of the synchronization problem is the delay spread mentioned in Chap. 1. In TDMA transmission, severe time dispersion occurs. In urban areas of the United States, the mean delay spread can be as much as 3 μs. Then the transmission rate must be limited to

$$R_b \le \frac{1}{2\pi \times 3 \times 10^{-6}} = 50 \text{ kbps}$$

By using diversity schemes,[12] the transmission rate can be increased if the same BER is assumed. Another effective scheme is an adaptive decision feedback equalizer[13] which adaptively sums up the multipaths so that the time-delay spread at the receiver is reduced and the sum of multipaths is in the form of a diversity.

The adaptive equalizer can work for narrowband TDMA also, so TDMA is not based on wideband spread spectrum technology. Since TDMA is very immune to interference in the cellular system (one user is designated in one time slot) and allows a simple handoff procedure, it may be suitable for microcell systems. *Microcells* are loosely defined as cells whose radii are less than 1 mi. However in a heavy traffic area, TDMA performance may degrade faster than FDMA performance. Simulation can be used to decide which transmission method is best.

For economic and spatial reasons, a 300-kHz-bandwidth TDMA carrying 300 kbps can provide 10 voice channels. Since a low-cost base station with one radio transceiver can make the TDMA system seem very attractive, this possibility is worth pursuing. The critical issue is building the adaptive decision feedback equalizer properly.

Since the same digital system working in different areas will result in different levels of performance, we should be aware that the same system might test well in a light traffic area but poorly in a heavy traffic area.

14.6.3 Spread spectrum systems in the '80s

Many suggested systems appeared in the literature; we may mention two of them. The uses of a frequency hopping scheme in FDMA and a direct sequence scheme in both TDMA and CDMA are described as follows.

Frequency-division multiple access (FDMA). There are many spread-spectrum systems proposed for use in mobile radios. Cooper and Nettleton proposed[14] a frequency hopping system. In each hop a binary DPSK-modulated system with error-correction coding is implemented. Comparing the relative spectral efficiencies of this proposed system with today's cellular 30-kHz system, we see that they are almost the same. Goodman et al.[15] proposed using multilevel frequency shift keying (MFSK) plus frequency hopping for land mobile radios. Their theoretical results, based on certain assumptions, show a better spectral efficiency.

Time-division multiple access (TDMA). A digital TDMA system has been proposed by SEL in Germany,[16] which uses a bandwidth of 4 MHz and only one fixed radio frequency for all subscribers in each cell. Therefore it is a spread-spectrum system. In a wideband system, each channel does not need to tune to a particular frequency. All the channels share the same bandwidth and thus have the same type of radio set.

Method of operation. The signal of each channel in TDMA is coded differently to provide identification and proper channel separation. The difference between a digital TDMA system and an analog FDMA system is shown in Fig. 14.18. In a digital TDMA system, each channel has 32 code patterns (5 bits) and one sign bit; thus, a total 6 bits are transmitted.

In each cell, 60 TDMA channels are available in three channel groups. The merit of this system is shown in Fig. 14.19. Each base station covers three cell sectors in succession by means of an electronic scan-and-stay antenna pattern. Twenty channels are allocated to each sector. Once a cell sector has received its 20 channels (time slots), the

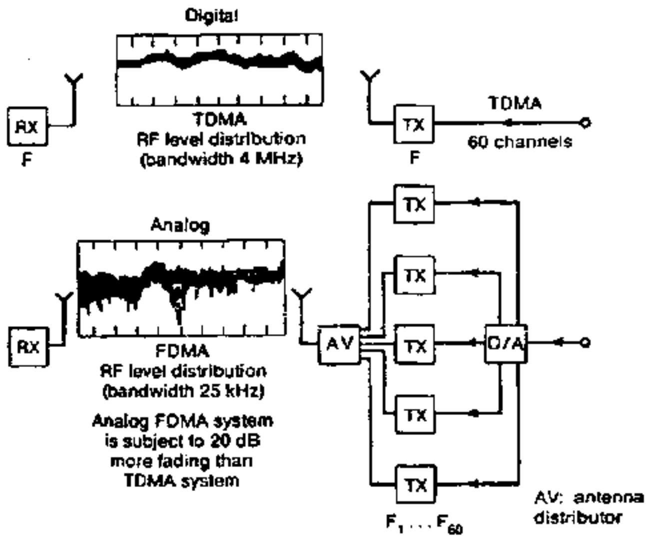

Figure 14.18 Digital TDMA mobile radio and analog FDMA system. *(After Bohm, Ref. 16. © Horizon House—Microwave, Inc.)*

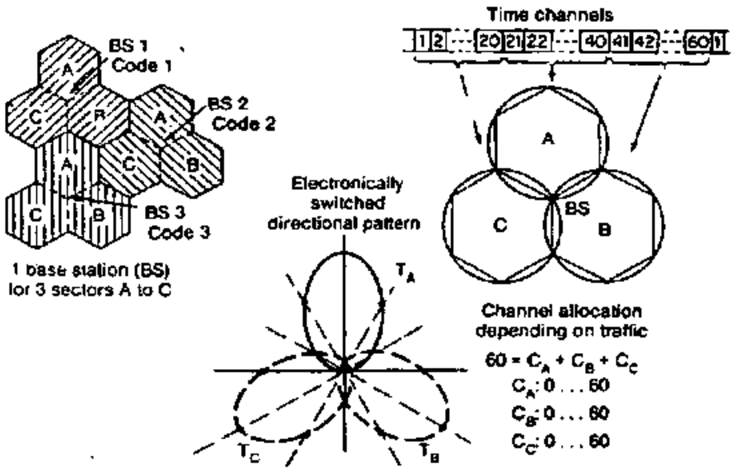

Figure 14.19 Cell structure of wideband TDMA systems. *(After Ref. 16. © Horizon House—Microwave, Inc.)*

antenna pattern of the base station is switched over to the next sector. The 20 channels per sector is an arbitrary number. Any number of channels can be used among the total 60 channels.

The synchronization and the clock accuracy have to be considered. This TDMA scheme does simplify the base station, and there is hopefully less channel interference in this system. However, there is a high risk of developing a spread spectrum system.

14.6.4 Evaluation of a digital cellular system[17,18]

The performance of a digital cellular system should be evaluated by a subjective test. The required C/I is deduced from the subjective test for a given performance. Also the channel bandwidth of a FDMA or the equivalent channel bandwidth of a TDMA should be included in the evaluation of the system's performance. Therefore a nationwide standard subjective test should be established to accomplish this task. In Chap. 15, many standard digital systems will be described.

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Chapter

15

Digital Cellular Systems

Many digital cellular and cordless phone systems have been developed. The cellular systems are GSM, NA-TDMA, CDMA, PDC, and 1800-DCS, and the cordless phone systems are DECT and CT-2 schemes. Although analog cellular systems are limited to using frequency-division multiple-access (FDMA) schemes, digital cellular systems can use FDMA, time-division multiple-access (TDMA), and code-division multiple-access (CDMA). When a multiple-access scheme is chosen for a particular system, all the functions, protocols, and network are associated with that scheme. This chapter covers GSM, NA-TDMA, and CDMA in great detail and briefly describes other systems.

15.1 Global System for Mobile (GSM)[1-5]

CEPT, a European group, began to develop the Global System for Mobile TDMA system in June 1982. GSM has two objectives: pan-European roaming, which offers compatibility throughout the European continent, and interaction with the integrated service digital network (ISDN), which offers the capability to extend the single-subscriber-line system to a multiservice system with various services which are currently offered only through diverse telecommunications networks.

System capacity was not an issue in the initial development of GSM, but due to the unexpected, rapid growth of cellular service, 35 revisions have been made to GSM since the first issued specification. The first commercial GSM system, called D2, was implemented in Germany in 1992.

15.1.1 GSM Architecture

GSM consists of many subsystems, such as the mobile station (MS), the base station subsystem (BSS), the network and switching subsystem (NSS), and the operation subsystem (OSS) (see Fig. 15.1).

The mobile station. The MS may be a stand-alone piece of equipment for certain services or support the connection of external terminals, such as the interface for a personal computer or fax. The MS includes mobile equipment (ME) and a subscriber identity module (SIM). ME does not need to be personally assigned to one subscriber. The SIM is a subscriber module which stores all the subscriber-related information. When a subscriber's SIM is inserted into the ME of an MS, that MS belongs to the subscriber, and the call is delivered to that MS. The ME is not associated with a called number—it is linked to the SIM. In this case, any ME can be used by a subscriber when the SIM is inserted in the ME.

Base station subsystem. The BSS connects to the MS through a radio interface and also connects to the NSS. The BSS consists of a base transceiver station (BTS) located at the antenna site and a base station controller (BSC) which may control several BTSs. The BTS consists of radio transmission and reception equipment similar to the ME in an MS. A transcoder/rate adaption unit (TRAU) carries out encoding and speech decoding and rate adaptation for transmitting data. As a subpart of the BTS, the TRAU may be sited away from the BTS, usually at the MSC. In this case, the low transmission rate of speech code channels allows more compressed transmission between the BTS and the TRAU, which is sited at the MSC.

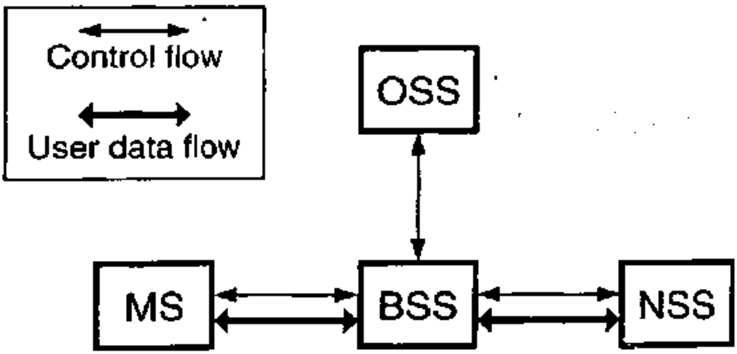


Figure 15.1 The external environment of the BSS.

GSM uses the open system interconnection (OSI). There are three common interfaces based on OSI (Fig. 15.2,): a common radio interface, called *air interface*, between the MS and BTS, an interface A between the MSC and BSC, and an A-bis interface between the BTS and BSC. With these common interfaces, the system operator can purchase the product of manufacturing company A to interface with the product of manufacturing company B. The difference between interface and protocol is that an interface represents the point of contact between two adjacent entities (equipment or systems) and a protocol provides information flows through the interface. For example, the GSM radio interface is the transit point for information flow pertaining to several protocols.

Network and switching subsystem. NSS (see Fig. 15.3) in GSM uses an intelligent network (IN). The IN's attributes will be described later. A signaling NSS includes the main switching functions of GSM. NSS manages the communication between GSM users and other telecommunications users. NSS management consists of:

Mobile service switching center (MSC). Coordinates call set-up to and from GSM users. An MSC controls several BSCs.

Interworking function (IWF). A gateway for MSC to interface with external networks for communication with users outside GSM, such

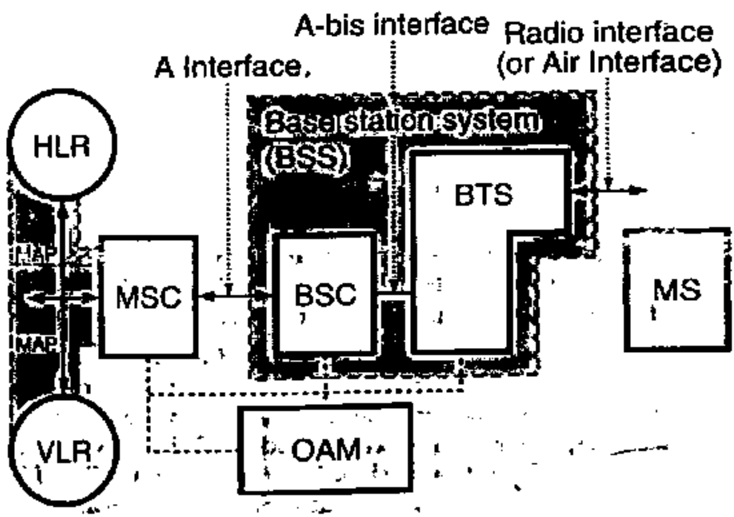


Figure 15.2. Functional architecture and principal interfaces.

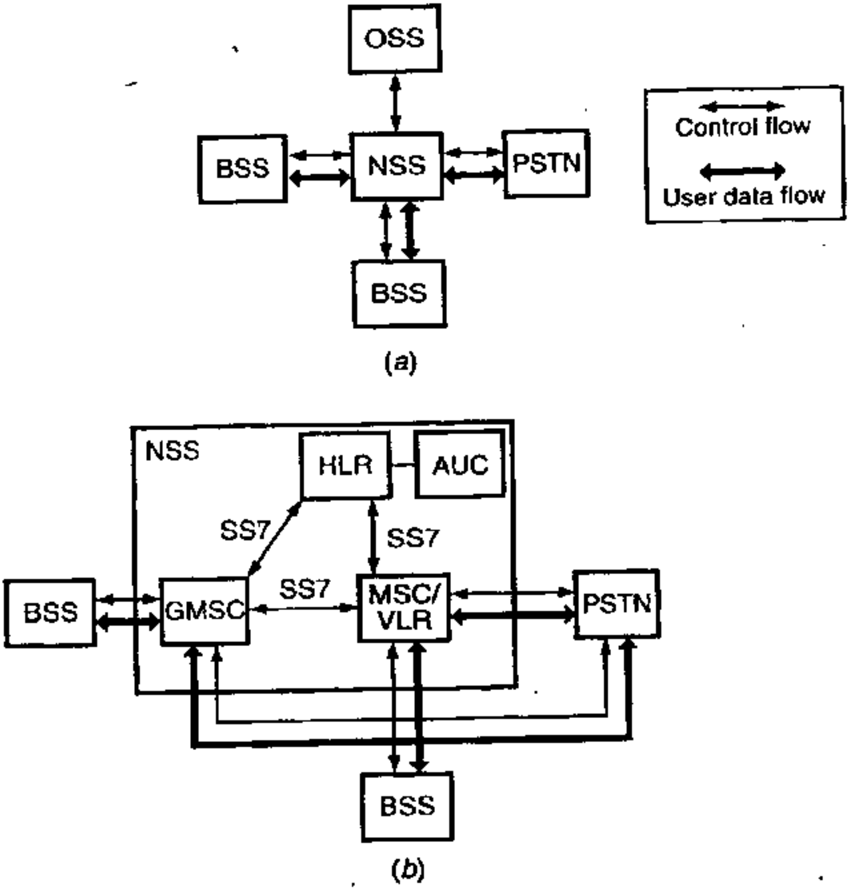


Figure 15.3 NSS and its environment. (*a*) The external environment; (*b*) the internal structure.

as packet-switched public data network (PSPDN) or circuit-switched public data network (CSPDN). The role of the IWF depends on the type of user data and the network to which it interfaces.

Home location register (HLR). Consists of a stand-alone computer without switching capabilities, a database which contains subscriber information, and information related to the subscriber's current location, but not the actual location of the subscriber. A subdivision of HLR is the authentication center (AUC). The AUC manages the security data for subscriber authentication. Another sub-division of HLR is the equipment identity register (EIR) which stores the data of mobile equipment (ME) or ME-related data.

Visitor location register (VLR). Links to one or more MSCs, temporarily storing subscription data currently served by its corresponding MSC, and holding more detailed data than the HLR. For example, the VLR holds more current subscriber location information than the location information at the HLR.

Gateway MSC (GMSC). In order to set up a requested call, the call is initially routed to a gateway MSC, which finds the correct HLR by knowing the directory number of the GSM subscriber. The GMSC has an interface with the external network for gatewaying, and the network also operates the full Signaling System 7 (SS7) signaling between NSS machines.

Signaling transfer point (STP). Is an aspect of the NSS function as a stand-alone node or in the same equipment as the MSC. STP optimizes the cost of the signaling transport among MSC/VLR, GMSC, and HLR.

As mentioned earlier, NSS uses an intelligent network. It separates the central data base (HLR) from the switches (MSC) and uses STP to transport signaling among MSC and HLR.

Operation subsystem. There are three areas of OSS, as shown in Fig. 15.4: (1) network operation and maintenance functions, (2) subscription management, including charging and billing, and (3) mobile

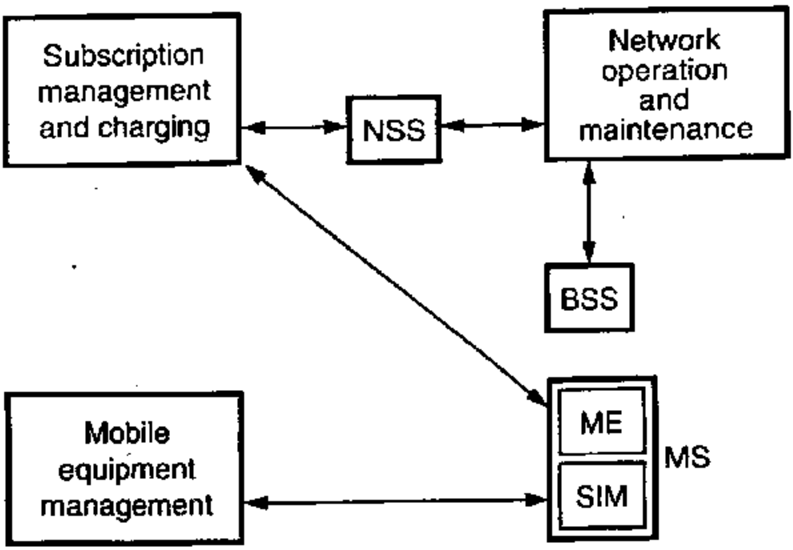


Figure 15.4 OSS organization.

equipment management. These tasks require interaction between some or all of the infrastructure equipment. OSS is implemented in any existing network.

15.1.2 Layer modeling (OSI model)

The Open System Interconnection (OSI) of GSM consists of five layers: transmission (TX), radio resource management (RR), mobility management (MM), communication management (CM), and operation, administration, and maintenance (OAM) (Fig. 15.5). The lower layers correspond to short-time-scale functions, the upper layers are long-time-scale functions.

The TX layer sets up a connection between MS and BTS. The RR layer refers to the protocol for management of the transmission over the radio interface and provides a stable link between the MS and BSC. The BSS performs most of the RR functions. The MM layer (1) manages the subscriber databases, including location data, and (2) manages authentication activities, SIM, HLR, and AUC. The NSS (mainly the MSC) is a significant element in the CM layer. The following functions are parts of the CM layer:

1. *Call control.* The CM layer sets up calls, maintains calls, and releases calls. The CM layer interacts among the MSC/VLR, GMSC,

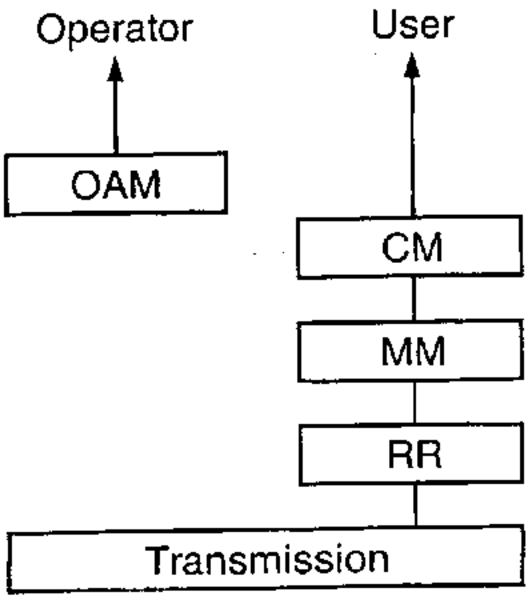


Figure 15.5 The functional planes of GSM.

IWF, and HLR for managing circuit-oriented service, including speech and circuit data.

2. *Supplementary services management.* Allows users to have some control of their calls in the network, and has specific variations from the basic service.
3. *Short message service (SMS).* Related to the point-to-point SMS. A SMS service center (SMS-SC) may connect to several GSM networks. Short message transmission requires setting up a signaling connection between the mobile station and the MSC. The two functions of SMS are
   a. Mobile-originating short message
   b. Mobile-terminating short message

OSS is an integral part of the OAM layer. All the subsystems, such as BSS and NSS, contribute to the OAM operation and maintenance functions.

15.1.3 Transmission

Speech. A 4-kHz analog speech signal converts to a 64-kbps digital signal, then down-converts to 13 kbps before modulation. Using a rate of 13 kbps instead of 64 kbps allows the 13-kbps data rate transmission to occur over a narrowband channel. Since the radio spectrum is a precious and limited resource, using less bandwidth per channel provides more channels within a given radio spectrum.

Digital speech uses:

1. *Regular pulse excitation (RPE).* Generates the impulse noise to simulate the nature of speech.
2. *Linear prediction coding (LPC).* Generates speech waveform by using a filter with eight transmitted coefficients with a speech frame of 20 ms; 260 bits represent a 20-ms speech frame. There are two modes of voice transmission in GSM, continuous (normal mode) and discontinuous.

The discontinuous transmission (DTX) mode decreases effective radio transmission encoding of speech at 13 kbps from a bit rate around 500 bps without speech. In active speech the frame is 260 bits in each 20 ms, and in inactive speech, the frame is 260 bits in 480 ms (24 times longer than normal mode).

A voice activity device (VAD) detects the DTX mode. In the voice protocol, a silence detection (SID) frame precedes the start of DTX. The speech coder provides an additional bit of information indicating whether the speech frame needs to be sent, depending on the VAD algorithm.

An SID starts at every inactivity period and repeats at least twice per second, as long as inactivity lasts. During the inactive speech period determined by VAD, and during every inactive period, artificial noise is generated at the receiver, substituting for background noise.

Data service. The highest data rate is 9600 bps and has two different modes. A forward error correction mechanism is provided in the transparent (T) mode. In the nontransparent (NT) mode, information is repeated when it is not acknowledged by the other end, and may be called an automatic repeat request (ARQ). Three different users' data rates are employed in the T connection: 2400 bps, 4800 bps, and 9600 bps. After insertion of the auxiliary information bits, the intermediate rates bits become 3.6 kbps, 6 kbps, and 12 kbps, corresponding to the user's 2.4 kbps, 4.8 kbps, and 9.6 kbps, respectively.

The basic GSM data rate is also 12 kbps (6 kbps on the half-rate channel) in an NT connection, but the available throughput varies with the quality of basic transmission and the transmission delay. Generally, the NT mode has less transmission error but also less throughput. The NT mode may be considered for a packet data flow application. The user data stream is sliced into blocks of 200 bits, and, with addition of the redundancy and auxiliary information, the user data stream becomes 240 bits per block. These blocks are used in NT while the ARQ scheme is applied.

An adaptation function called interworking function (IWF) at the network side, and terminal adapting function (TAF) at the terminal, is used to accommodate variable transmission rates (Fig. 15.6). The radio link protocol (RLP) is used for transporting signaling messages between the TAF and IWF.

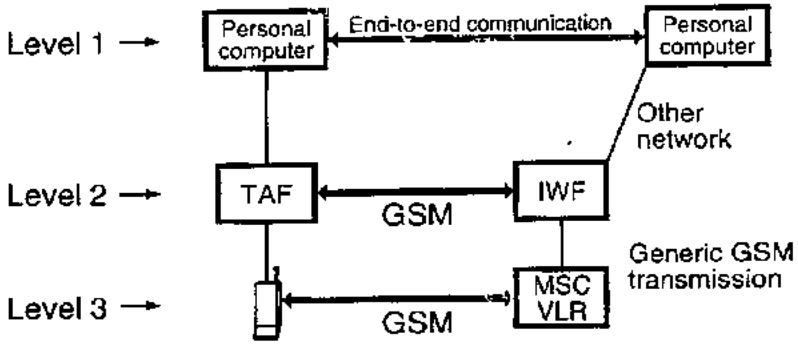


Figure 15.6 Data transmission planes.

Data can transmit over these planes, as shown in Fig. 15.6:

1. End-to-end transmission—direct transmission through hard wire
2. TAF to IWF transmission through subscriber units
3. GSM radio transmission through subscriber units; acts like a voice call in the air.

Although speech interconnection with the ISDN is not a problem, data transmission raises its own problems, as shown in Fig. 15.7. ISDN uses the capacity of a bidirectional 64 kbps/channel, but GSM must use the radio spectrum efficiently, through a bidirectional 13 kbps/channel. Interconnection of data services between GSM and ISDN is not possible without a rate-adapted (RA) box, as shown in Fig. 15.7.

Modulation. Gaussian minimum-shift keying (GMSK), where $BT = 0.3$ is the normalized bandwidth of a gaussian filter, is the modulation scheme of GSM, where B is the baseband bandwidth, and $1/T$ is the transmission rate. $B = 1/T \times 0.3 = 270$ kbps $\times\ 0.3 = 81$ kHz. Minimum means the *minimum tone separation*. GMSK utilizes a small spectrum bandwidth to send a GSM carrier channel. The modulation rate of a GSM carrier channel is 270 kbps.

15.1.4 GSM channels and channel modes

Channel structure. The services offered to users have four radio transmission modes, three data modes, and a speech mode. The radio transmission modes use the physical channels.

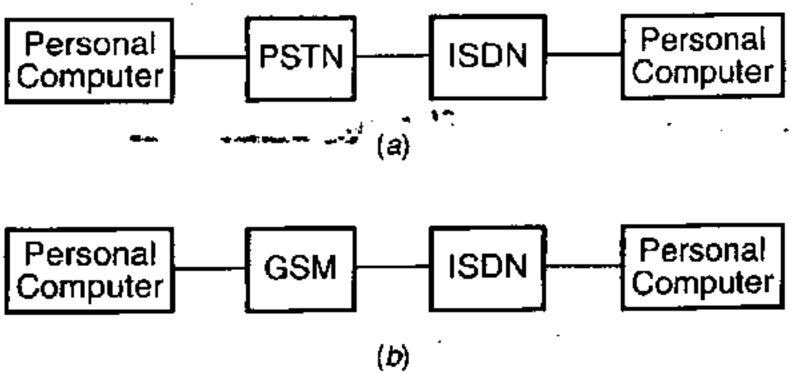


Figure 15.7 Interconnection with ISDN. (*a*) PSTN user to ISDN user; (*b*) GSM user to ISDN user.

Physical channels. There are three kinds of physical channels, also called traffic channels (TCHs):

1. *TCH / F (full rate).* Transmits a speech code of 13 kbps or three data-mode rates, 12, 6, and 3.6 kbps.
2. *TCH / H (half rate).* Transmits a speech code of 7 kbps or two data modes, 6 and 3.6 kbps.
3. *TCH / 8 (one-eighth rate).* Used for low-rate signaling channels, common channels, and data channels.

Logical Channels

Common channels. All the common channels are embedded in different traffic channels. They are grouped by the same cycle (51×8 BP), where BP stands for burst period (i.e. time slot), which is 577 μs.

Downlink common channels. There are five downlink unidirectional channels, shared or grouped by a TCH.

- Frequency correction channel (FCCH) repeats once every 51×8 BPs; used to identify a beacon frequency.
- Synchronization channel (SCH) follows each FCCH slot by 8 BPs.
- Broadcast control channel (BCCH) is broadcast regularly in each cell and received by all the mobile stations in the idle mode.
- Paging and access grant channel (PAGCH). Used for the incoming call received at the mobile station. The access grant channel is answered from the base station and allocates a channel during the access procedure of setting up a call.
- Call broadcast channel (CBCH). Each cell broadcasts a short message for 2 s from the network to the mobile station in idle mode. Half a downlink TCH/8 is used, and special CBCH design constraints exist because of the need for sending two channels (CBCH and BCCH) in parallel.

The mobile station (MS) finds the FCCH burst, then looks for an SCH burst on the same frequency to achieve synchronization. The MS then receives BCCH on several time slots and selects a proper cell, remaining for a period in the idle mode.

Uplink common channel. The random-access channel (RACH) is the only common uplink channel. RACH is the channel that the mobile station chooses to access the calls. There are two rates: RACH/F—full rate, one time slot every 8 BP, and RACH/H—half rate, using 23 time slots in the 51×8 BP cycle, where 8 BP cycle (i.e. a frame) is 4.615 ms.

Signaling channels. All the signaling channels have chosen one of the physical channels, and the logical channels names are based on their logical functions:

Slow associated control channel (SACCH). A slow-rate TCH used for signaling transport and used for nonurgent procedures, mainly handover decisions. It uses one-eighth rate. The TCH/F is always allocated with SACCH. This combined TCH and SACCH is denoted TACH/F. SACCH occupies 1 time slot (0.577 ms) in every 26 frames (4.615 ms × 26). The time organization of a TACH/F is shown in Fig. 15.8.

Fast associated control channel (FACCH). Indicates cell establishment, authenticates subscribers, or commands a handover.

Stand-alone dedicated control channel (SDCCH). Occasionally the connection between a mobile station and the network is used solely for passing signaling information and not for calls. This connection may be at the user's demand or for other management operations such as updating the unit's location. It operates at a very low rate and uses a TCH/8 channel.

Radio slots are allocated to users only when call penetration is needed. There are two modes, dedicated and idle. The mode used depends on the uplink and the downlink. In GSM terminology, the downlink is the signal transmitted from the base station to the mobile station, and the uplink is the signal transmitted in the opposite direction. (*Note:* The terrestrial communication terms *uplink* and *downlink* are not to be confused with the same terms used in satellite communications. In many instances the position of the mobile station can be higher than the base station antenna because of the terrain contour. Using the terms in this kind of situation may cause confusion, therefore it is more important for the reader to remember the *definitions* of these terms as used in terrestrial communications. This approach is analogous to using the term *handoff* instead of *handover* in discussions about European cellular telecommunications—as long as the definition itself is clear, the terms will be understood.)

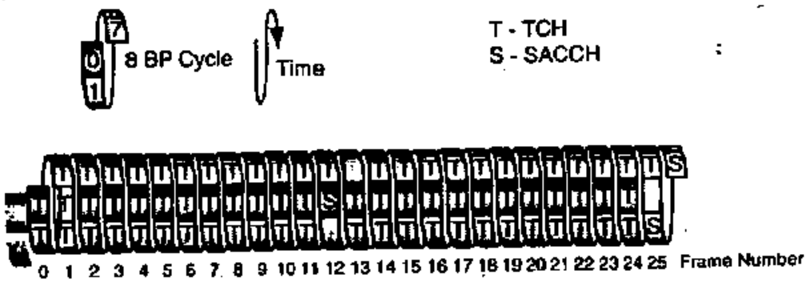


Figure 15.8 Time organization of a TACH/F.

Voice / data channels. Each time slot of a voice channel contains 260 bits per block. The entire block contains 316 bits. Each time slot of a data channel contains 120 or 240 bits per block.

Channel modes. Because of the precious value of the radio spectrum, individual users cannot have their own TCH at all times.

Dedicated mode. Uses TCH during call establishment and uses SACCH to perform location updating in the dedicated mode. TCH and SACCH are dedicated channels for both uplink and downlink channels.

Idle mode. During noncall activities, the five downlink channels are in the idle mode: FCCH; SCH; BCCH, which is broadcasting regularly; PAGCH and CBCH, which sends one message every 2 s. During idle mode, the mobile station listens to the common downlink channels, and also uses SDCCH to register a mobile location associated with a particular base station to the network.

15.1.5 Multiple-access scheme

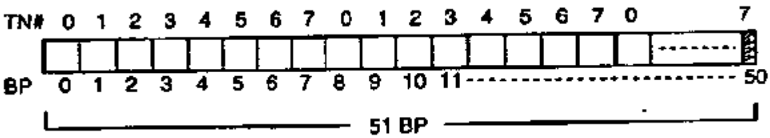
General description. GSM is a combination of FDMA and TDMA. The total number of channels in FDMA is 124, and each channel is 200 kHz. Both the 935–960 MHz uplink and 890–915 MHz downlink have been allocated 25 MHz, for a total of 50 MHz. Duplex separation is 45 MHz. If TDMA is used within a 200-kHz channel, 8 time slots are required to form a frame, frame duration is 4.615 ms, and the time slot duration burst period is 0.577 ms. There is a DCS-1800 system, which has the same architecture as the GSM, but it is upconverted to 1800 MHz. The downlink is 1805–1880 MHz (base TX) and the uplink is 1700–1785 MHz.

Constant time delay between uplink and downlink. The numbering of the uplink slots is derived from the downlink slots by a delay of 3 time slots. This allows the slots of one channel to bear the same time slot number in both directions. In this case, the mobile station will not transmit and receive simultaneously because the two time slots are physically separated. Propagation delay when the mobile station is far from the BTS is a major consideration. For example, the round trip propagation delay between an MS and BTS which are 35 km apart is 233 μs. As a result, the assigned time slot numbers of the uplink and downlink channels may not be the same (less than 3 time slots apart). The solution is to let BTS compute a time advance value. The key is to allow significant guard time by taking into account that BCCH is using only even time slots. This avoids the uncertainty of numbering the wrong time slot. Once a dedicated connection is established, the BTS continuously measures the time offset between its own burst schedule and the reception schedule of mobile station bursts on the bidirectional SACCH channel. The time compensation for the propagation delay (sending to the mobile station via SACCH) is 3 time slots minus the time advance.

Frequency hopping. GSM has a slow frequency-hopping radio interface. The slow hopping is defined in bits per hop. Its regular rate is 217 hops/s, therefore, with a transmission rate of 270 kbps, the result is approximately 1200 bits/hop.

If the PAGCH and the RACH were hopping channels, then hopping sequences could be broadcast on the BCCH. The common channel is forbidden from hopping and using the same frequency.

Different types of time slots. Each cell provides a reference clock from which the time slots are defined. Each time slot is given a number (TN) which is known by the base station and the mobile station. The time slot numbering is cyclic. TN0 is a single set broadcast in any given call and repeated every 8 BPs for the confirmation of all common channels. The organization of TN0 (first of eight time slots) in sequence is as follows: FCCH (1), SCH (1), BCCH (4), PAGCH (4), FCCH (1), SCH (1), PAGCH (8), FCCH (1), SCH (1), PAGCH (8), FCCH (1), SCH (1), PAGCH (8), FCCH (1), SCH (1), PAGCH (8).



The symbol PAGCH (4) means that the PAGCH channel information appears in consecutive ones of every 8 BP cycle 4 times. Each of the remaining seven TNs (TN1 to TN7) is assigned to one TACH/F channel.

Bursts and training sequences. In TDMA, the signal transmits in bursts. The time interval of the burst brings the amplitude of a transmitted signal up from a starting value of 0 to its normal value. Then a packet of bits is transmitted by a modulated signal. Afterward, the amplitude decreases to zero. These bursts occur only at the mobile station transmission or at the base station if the adjacent burst is not transmitted.

There are tail bits and training sequence bits within a burst. The tail bits are three 0 bits added at the beginning and at the end of each burst which provide the guard time.

The training sequence is a sequence known by the receiver which trains an equalizer, a device which reduces intersymbol interference. The training sequence bits are inserted in the middle of a time slot sometimes called a *midamble,* for the same purpose as a preamble, so that the equalizer can minimize its maximum distance with any useful bit. There are eight different training sequences, with little between any two sequences to distinguish the received signal from the interference signal.

There are several kinds of bursts:

1. The normal burst used in TCH:

		Flag		Flag		
3	57	1	26	1	57	3
Tail	Information		Training Sequence		Information	Tail

   The 1-bit binary information indicating data or signaling is called the *stealing flag.*

2. The access burst used on the RACH in the uplink direction:

		Flag		
7	41	1	35	3
Tail	Training Sequence		Information	Tail

3. The F and S bursts. The F burst is used on the FCCH and has the simplest format. All of the 148 bits are zero, producing a pure sine wave. Five S bursts in each 51 × 8 BP cycle are used on the SCH. One S burst is shown below:

		Flag		Flag		
3	38	1	64	1	38	3
Tail	Information		Training Sequence		Information	Tail

15.1.6 Channel coding and interleaving

Channel coding. Channel coding improves transmission quality when interference, multipath fading, and Doppler shift are encountered. As a result, the bit error rate and frame error rate or word error rate are reduced, but throughput is also reduced. Four kinds of channel codings are used in GSM:

1. Convolutional codes (L, k) are used to correct random errors: k is the input block bits, and L is the output block bits. Convolutional

codes have three different rates in GSM: (1) the one-half rate ($L/k = 2$), (2) the one-third rate ($L/k = 3$), and (3) the one-sixth rate ($L/k = 6$).

2. Fire codes (L, k) are used as a block code to detect and correct a single burst of errors, where k is the information bits and L is the coded bits.
3. Parity check codes (L, k) are used for error detection. L is the bits of a block, k is the information bits, $L - k$ is the parity check bits.
4. Concatenation codes use convolutional code as an inner code and fire code as an outer code. Both the inner code and the outer code reduce the probability of error and correct most of the channel code. The advantage of using concatenation code is a reduction of the implementation complexity as compared with a single coding operation.

GSM's speech code is sent at a rate of 13 kbps, which represents 260 bits in each 20-ms speech block. After channel coding, each block contains 456 bits and the transmission rate is 22.8 kbps, or 114 bits for time slots. Adding the overhead bits such as tail bits (6), training bits (26), flag bits (2), and guard time bits (8.25), the total bits of a traffic channel is 156 bits in one time slot of 0.577 ms, as shown in Fig. 15.9.

Interleaving. Interleaving scrambles and/or spreads a sequence of bits prior to transmitting them. The sequence of bits is put back in order at the receiving end. Bursts of errors occur during transmission because of signal fading. After being received, these bursts of errors are then converted to random errors and put back in the correct sequence. Interleaving's major drawback is the corresponding delay at the receiving end.

Interleaving schemes are relatively simple in GSM. A code word of 456 bits could be spread into the following format:

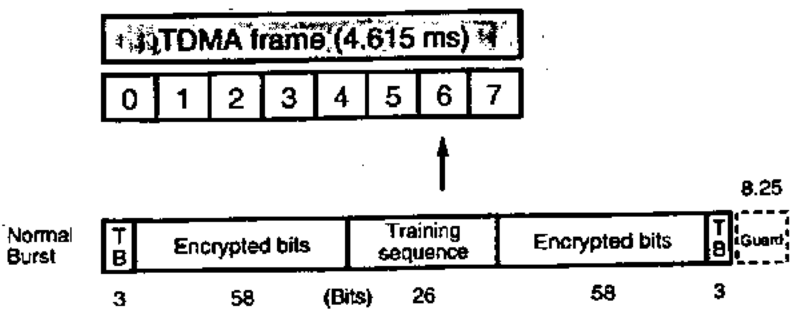


Figure 15.9 TDMA frame and normal burst.

1. Four full bursts—divide 456 bits into 4 parts, each one filling up a whole burst. This interleaving format takes 4.615 ms × 4 = 18.46 ms.
2. Eight half bursts—divide 456 bits into 8 parts, each one filling up half a burst. This interleaving format takes 4.615 ms × 8 = 36.92 ms. Four parts share with the previous and four parts with the new partial code word.

The interleaving and coding for different transmission modes are shown in Table 15.1. Interleaving is a powerful scheme which converts burst errors into random errors, and although it is very effective for data transmission, it is not effective for voice transmission. Voice

TABLE 15.1 Interleaving and Coding for the Different Transmission Modes *(Reprinted from Ref. 3 p. 246)*

Channel and transmission mode		Input rate (kbit/s)	Input block (in bits)	Coding	Output block (in bits)	Interleaving
TCH/FS	Ia		50	Parity (3 bits) convolutional 1/2		
	Ib	13	132	Convolutional 1/2	456	On 8 half-bursts
	II		78	None		
TCH/F9.6 TCH/H4.8		12 6	240	Convolutional 1/2 punctured 1 bit out of 15	456	Complex, on 22 unequal burst portions
TCH/F4.8		6	120	Addition of 32 null bits Convolutional 1/3	456	Complex, on 22 unequal burst portions
TCH/F2.4		3.6	72	Convolutional 1/6	456	On 8 half bursts
TCH/H2.4		3.6	144	Convolutional 1/3	456	Complex, on 22 unequal burst portions
SCH			25	Parity (10 bits) Convolutional 1/2	78	On 1 S surst
RACH (+ Handover Access)			8	Parity (6 bits) Convolutional 1/2	36	On 1 access burst
Fast associated signalling on TCH/F and /H			184	Fire code 224/184	456	On 8 half bursts
TCH/8, SACCH; BCCH, PAGCH				Convolutional 1/2		On 4 full bursts

transmission operates in real time, and a long delay in response cannot be tolerated.

Without interleaving and overhead bits, the transmit rate for a speech channel is 22.8 kbps, 114 bits per time slot, and 456 bits per four time slots.

15.1.7 Radio resource (RR) management

In a mobile network, radio channels must allocate for call setup, handover and release, on a call bias. This management is additional to the conventional fixed network call handling procedures. There are three management functions; location, handover, and roaming. The implementation of the RR functions require some kind of protocol between the mobile station and the network.

Link protocol. We studied the means of transporting user information in previous sections. But in addition to the user's information, the signaling transfer information exchanges must be sent and understood by every piece of signaling transport equipment. Most information exchange functions are distributed to different kinds of equipment. There are three link protocols to provide information exchanges.

Radio link protocol (RLP), specified in GSM link access protocol over the radio link called LAPDm.

LAPD, the link access protocol (LAP) adapted from ISDN D channel.

Message transfer part (MTP), the protocols used for signaling transport on an SS7 network.

The radio link protocol's signaling message rate is 22.8 kbps. The signaling message rate on the other link protocol is 64 kbps.

Interfaces associated with link protocols

Interface	Link protocol
MS-BTS	LAPDm (GSM spec)
BTS-BSC	LAPD (adopted from ISDN)
BSC-MSC	MTP (SS7 protocol)
MSC/VLR/HLR—SS7 network	MTP (SS7 protocol)
MSC-MSC (call-related signaling)	TUP (telephone user part)
BSC-relay MSC (non-call-related signaling)	ISUP (ISDN user part) BSS MAP (MAP/B)
MSC-MSC (non-call-related signaling)	MAP (mobility application part)

Non-call-related signals correspond to protocols in the MSC that are different from those in other MSCs or other HLRs and are grouped together in the MAP. We can distinguish them by MAP/X, where X can be B, C, D, and so forth.

MAP/B Protocol between BSC and relay MSC

MAP/C Protocol between GMSC and an HLR

MAP/D Protocol between another MSC/VLR and HLR

MAP/E Protocol between MSCs

Figure 15.10 shows the relationships of MAP/*X* protocols.

15.1.8 Mobility management (MM)

The mobility of cellular system users requires mobility management for location updates, handovers, and roaming. A handover occurs when a voice channel changes as the mobile station enters another cell during a call. Roaming is the ability to initiate a call in one network system and deliver it to another network system by using MM and location update management.

Location update management. The subscription is always associated with its home public land mobile network (PLMN). The roaming customer is associated with visited PLMNs. We may identify whether the call is from PLMN or visited PLMN from the location of the MS.

In the PLMN selection process, the MM normally looks for cells only in the home (serving) PLMN. If no service is available, the user can choose either the automatic mode (the network searches) or the manual mode (the user searches) to search for the desired PLMN. In the limited-service case, the MM continuously monitors only the 30 strongest carriers. Limited service usually takes care of coverage at the border areas of a foreign country.

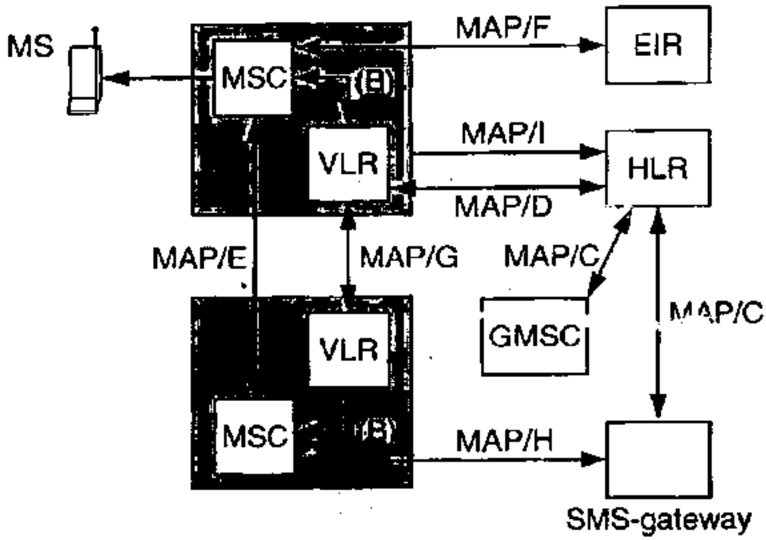


Figure 15.10 MAP/C to MAP/I protocols.

Cell selection. Choosing the best cell from an MS depends on three factors: (1) the level of the signal received by the mobile station, (2) the maximum transmission power of the mobile station, and (3) two parameters p_1 and p_2 specified by the cell. This is called the C_1 criterion.

$$C_1 = A - \max(B, O)$$

$$A = \text{received level average} - p_1$$

$$B = p_2 - \text{maximum RF power of the MS}$$

$$p_1 = \text{a value between } -110 \text{ and } -48 \text{ dBm}$$

$$p_2 = \text{a value between } 13 \text{ and } 43 \text{ dBm}$$

Both values of p_1 and p_2 are broadcast from the cells.

MS maximum power = 29 to 43 dBm

The cell selection algorithm is as follows:

- A SIM must be inserted.
- The strongest C_1 is chosen by obtaining C_1 from candidate cells; the C_1 has to be higher than 0.
- All cells must not be barred from service.

Authentication. Authentication protects the network against unauthorized access.

First Phase. A PIN (personal identification number) code protects the SIM. The PIN is checked by the SIM locally, so the SIM is not sent out over the radio link.

Second phase. The GSM network makes an inquiry by sending a random number (RAND). The 128-bit RAND is sent from the network to the MS, and mixes with the MS's secret parameter, K_i, in an A3 processing algorithm, which produces a 32-bit-long SRES (signed result) number. The SRES is then sent to the network from the MS for verification (see Fig. 15.11).

Encryption. Encryption protects against unauthorized listening. The MS uses the RAND received from the network and mixes K_i through a different algorithm, called A8, and generates K_c (64 bits). The ciphering sequences are generated from the K_c (see Fig. 15.12). The frame number and K_c move to a ciphering algorithm, A5, and generate

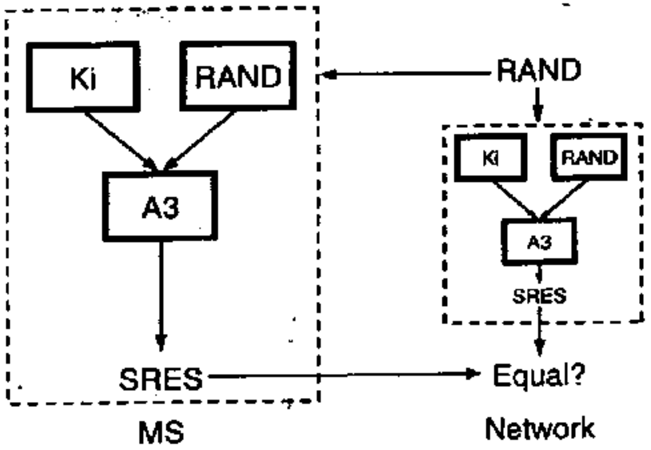


Figure 15.11 The authentication computation.

S_2 (114 bits), which plays an exclusive-or operation between the 114 bits of plain text and ciphering sequence S_2, as shown in Fig. 15.12.

User identity protection—security management. SIM (MS side) and AUC (network side) are the repositories of the subscriber's key K_i. Key K_i never transmits over the air. Both sides perform A3 and A8 computations.

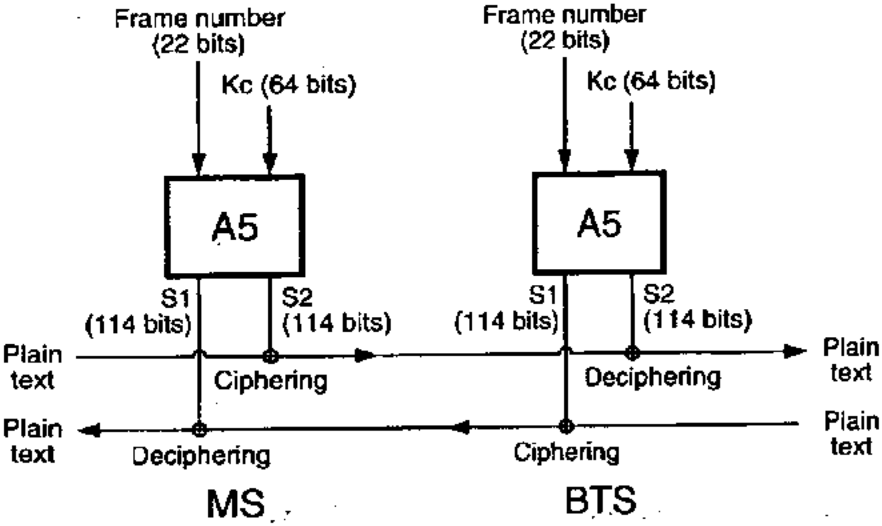


Figure 15.12 Ciphering and deciphering.

15.1.9 Communication management

The CM layer provides telecommunications services such as speech, fax, and data to users via RR and MM layers, as shown in Fig. 15.13. The users include GSM calling party, GSM called party, and the users in both GSM calling and called parties. The management functions of CM are call control, service management, and short message service.

Call control. CC manages the most circuit-oriented services (speech, circuit data) through the MSC/VLR, GMSC, IWF, and HLR. CC functions set up calls (mobile-originating or base-originating), maintain calls, and release calls. To establish calls, the MS number has to be assigned. MS/ISDN is a mobile station ISDN number, part of the same numbering plan as ISDN numbers. Mobile station roaming number (MSRN) is the routing number, another number which can be a GSM subscriber or third party international mobile subscriber identity (IMSI) and provided by MS to access a foreign network. Figure 15.14 illustrates a domestic call through a GMSC. Figure 15.15 illustrates an international call.

Handover. The GSM handover algorithm is not specified as a standard. It is a feature of mobile assistance handover (MAHO) and is carried out within the unit. The MS scans for another radio carrier under direction from a base station. It monitors those time slots which are not its own assigned time slots for receiving the signal. In this case, on the request of a base station, the signal strength of a specified

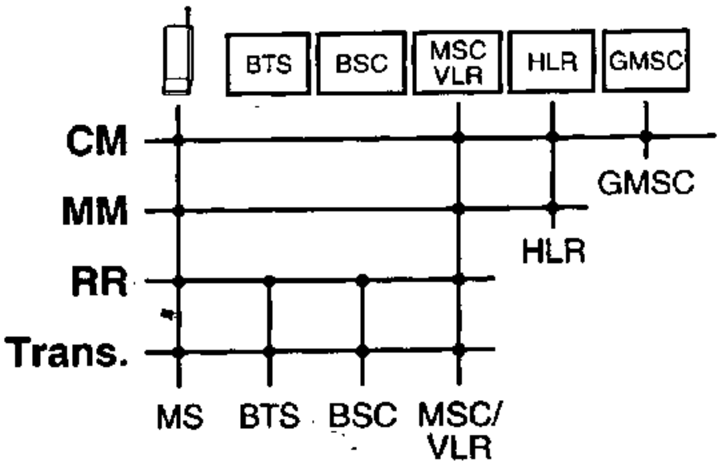


Figure 15.13 General protocol architecture of GSM.

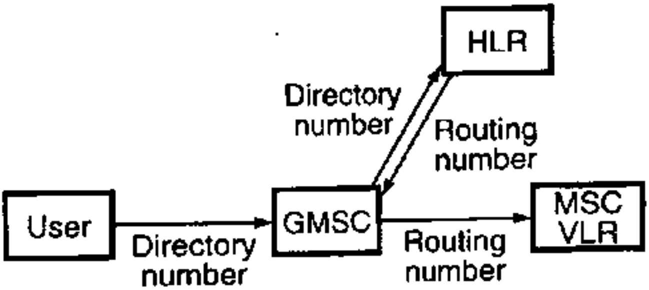


Figure 15.14 The key role of the GMSC for a domestic call.

radio carrier is measured in one time frame, and, on request, the measurements are forwarded to the base station to assist in the handover process. This is called MAHO. The MSC uses two sets of information to decide whether a handover should be initiated and which BTS is the candidate BTS for the handover. The two sets are (1) the signal strengths of the MS as received at the neighboring BTSs and (2) the signal strengths of neighboring BTSs received at the MS. The latter information is from MAHO.

Supplementary services management (SSM). CC provides supplementary services such as call waiting, call forwarding, and automatic answering. SSM is a point-to-point management service. An SSM service center (SSM-SC) may connect to several GSM networks. SSM consists of two functions:

1. Mobile terminating short message

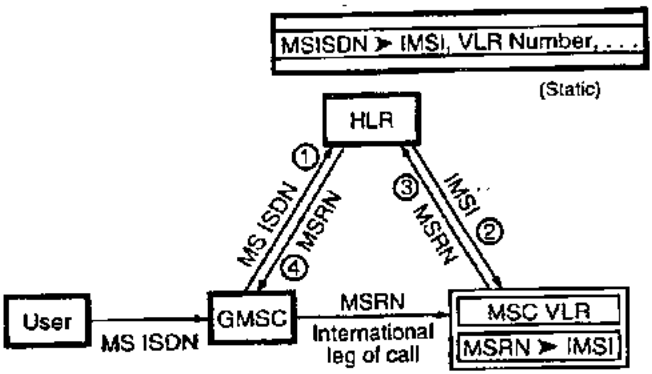


Figure 15.15 The provision of the MSRN for an international call.

2. Mobile originating short message

Short message services. CC provides point-to-point short message services (SMS-PP). GSM is connected to the short message service center. The signaling transmission uses digital audio tones [digital tone multifrequency (DTMF)] to control voice mailbox, answering machine, conferencing, etc.

15.1.10 Network management (NM)

An NM center oversees the following administration tasks:

1. Subscriber management—subscription administration
2. Billing and accounting
3. Maintenance
   a. Minimizing failures
   b. Monitoring operations and indicating by alarm improper operation situations
4. Subscriber administration tasks provide the selected approval code for ME (mobile equipment) within the international mobile equipment identity (IMEI) number. The code totals 15 digits and consists of type approval code (TAC) + final assembly code (FAC) + serial number which stores in EIR.
5. In the GSM telecommunication management network (TMN), all operation and maintenance machines compose a network which is linked to all traffic handling machines. The GSM Q3 is a network management protocol for operation systems functions and traffic handling machines. Two aspects of using GSM Q3 protocol are important. (a) Standardizing data communication protocols on the application level, such as file transfer. (b) Embodying the network modelling in GSM Q3 protocol.

15.1.11 Overview of GSM

Summary of physical layer parameters

TDMA structure	8 time slots per radio carrier
Time slot	0.577 ms
Frame interval	8 time slots = 4.615 ms
Radio carrier number	124 radio carriers (935–960 MHz downlink, 890–915 MHz uplink)
Modulation scheme	Gaussian minimum shift keying with $BT = 0.3$
Frequency Hopping	Slow frequency hopping (217 hops/s)
Equalizer	Equalization up to 16 μs time dispersion

GSM's strength. GSM is the first to apply the TDMA scheme developed for mobile radio systems. It has several distinguishing features:

1. Roaming in European countries
2. Connection to ISDN through RA box
3. Use of SIM cards
4. Control of transmission power
5. Frequency hopping
6. Discontinuous transmission
7. Mobile-assisted handover

15.2 North American TDMA[6–8]

15.2.1 History

North American TDMA (NA-TDMA) is a digital cellular system sometimes called American digital cellular (ADC) or digital AMPS (DAMPS), or North American digital cellular (NADC) or IS-54 system. This TDMA system was approved and design on it was started in 1987 by a group named TR45-3 after the industry debated between frequency-division multiple access and time-division multiple access. The reason those members voted for TDMA was the big influence of European GSM, which is the TDMA system. However, the requirements of designing a digital cellular system in Europe and in North America are different. In Europe, there is a virgin band (935–960 MHz downlink and 890–915 MHz uplink) for the digital cellular system. In North America, there is no new allocated band for the digital cellular system. The digital cellular system has to share the same allocated band with the analog system (AMPS, described in Chap. 3). Also, the digital and the analog systems have to be coexistent. In this circumstance, the low-risk approach is to use the same signal signature as the analog system, i.e., FDMA. Besides, because of the urgent need for large system capacity, the time for designing a new North American system had to be very short. The North American digital system was needed to be available in 1990, in only 3 years. To design a digital FDMA system would be a straightforward task. Since the analog system is a FDMA system, all the physical data gathered for the analog system in the past 20 years could be used for designing the FDMA digital system, and design time would be shortened. On the other hand, to design a TDMA digital system in the same band shared with an FDMA analog system, much more physical data would have to be developed and time would be needed to understand them. Without a good understanding of the limitation of coexistence between two different signal signatures, FDMA and TDMA, it would be very difficult to complete a digital system with good performance in a very short time. If GSM had taken 8 years to develop, NA-TDMA might also need as much time to be revised in order for it to be mature.

Because of the requirement of coexistence, a dual-mode mobile unit was decided on; i.e., the unit can work on both analog and digital systems. In a dual-mode mobile unit, the 21 call set-up channels for the analog system are available in the unit. Why not share the same call set-up channels (analog) for both the analog voice channels and digital voice channels? In this case, no additional spectrum is needed for the digital set-up channels.[9] The spectrum is saved for adding more digital voice. Furthermore, for the sake of speeding up the completion of North American digital systems, the call set-up channels of the digital system could be shared with the analog system to make the call processing the same between the two systems. Thus, the first phase of the NA-TDMA system could be completed earlier.

15.2.2 NA-TDMA architecture

The NA-TDMA architecture is similar to GSM architecture. The only difference is that in NA-TDMA, there is only one common interface, which is the radio interface as shown in Fig. 15.16. The NA-TDMA uses the intelligent network. All the components such as HLR, VLR, AUC, and EIR are the same as used in GSM (see Sec. 15.1.1). In developing the NA-TDMA system, there were two phases:

First phase—to commonly share the 21 set-up channels which are used for the analog system. The first-phase system is only for voice transmission. Both modes, AMPS and digital, are built in the same unit. The handoff procedure has to take care of the following four features:

1. AMPS cell to AMPS cell
2. TDMA cell to TDMA cell
3. AMPS cell to TDMA cell
4. TDMA cell to AMPS cell

Second phase—(1) generate new digital set-up channels (they were in the voice band) to access to TDMA voice channels so that a digital stand-alone unit can be provided and (2) specify a data-service signal protocol for transmitting data.

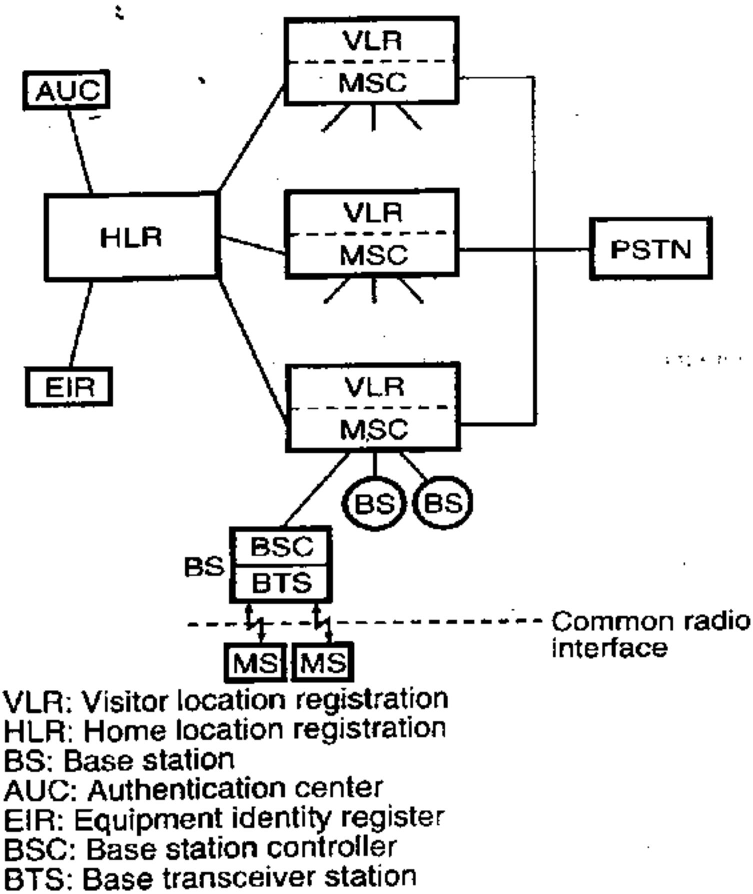


Figure 15.16 NA-TDMA system architecture.

15.2.3 Transmission and modulation

TDMA structure (digital channels) In NA-TDMA, the set-up channels are analog channels shared with the AMPS system. One digital channel (a 30 kHz TDMA channel) contains 25 frames per second. Each frame is 40 ms long and has 6 time slots. Each time slot is 6.66 ms long. One frame contains 1944 bits (972 symbols), as shown in Fig. 15.17.

Each slot contains 324 bits (162 symbols) and the duration between bits is 20.57 μs. Therefore, one radio channel is transmitted at 48.6 kbps but only 24,000 symbols per second over the radio path. Each frame consists of 6 time slots. The maximum effect on the signal for

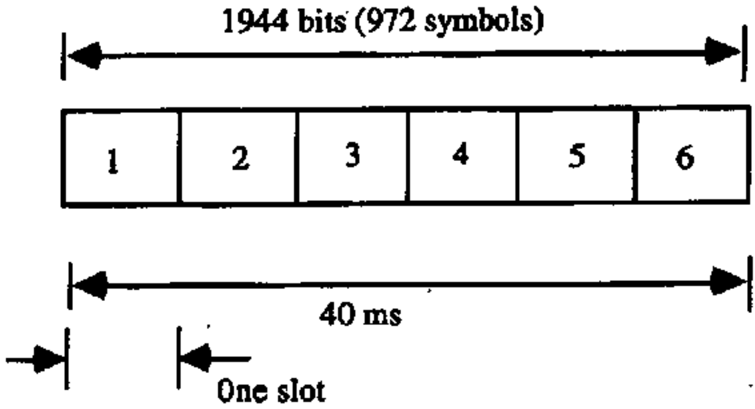


Figure 15.17 TDMA frame structure.

a forward time slot is one-half full symbol period and for a reverse time slot is 6 symbol periods (Fig. 15.18).

Frame length. There are two frame lengths, full rate and half rate. Each full-rate traffic channel shall utilize two equally spaced time slots of the frame. The overall length in each slot is shown in Fig. 15.18.

Channel 1 uses time slots 1 and 4

Channel 2 uses time slots 2 and 5

Channel 3 uses time slots 3 and 6

Each half-rate traffic channel shall utilize one time slot of the frame:

Channel 1 uses time slot 1

Channel 2 uses time slot 2

Channel 3 uses time slot 3

Channel 4 uses time slot 4

Channel 5 uses time slot 5

Channel 6 uses time slot 6

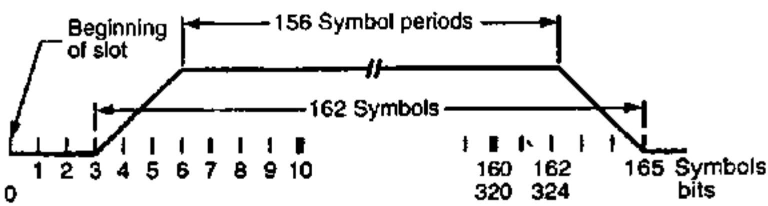


Figure 15.18 Overall length in each slot.

Frame offset. At the mobile station, the offset between the reverse and forward frame timing (without time advanced applied), is

$$\text{Forward frame} = \text{reverse frame} + (1 \text{ time slot} + 44 \text{ symbols})$$

$$= \text{reverse frame} + 206 \text{ symbols}$$

The time slot (TS) 1 of frame N (in forward link) occurs 206 symbol periods after TS 1 of frame N in the reverse link.

Modulation timing

Modulation timing within a forward time slot. The first modulated symbol (the first symbol of the sync word) used by the mobile unit shall have maximum effect on the signal (156 symbols) transmitted from the base antenna, one-half symbol (1 bit) period after beginning the time slot.

Modulation timing within a reverse time slot. The first modulated symbol has a maximum effect on the signal transmitted at the mobile unit 6 symbol periods after the beginning of the reverse time slot.

Power level. In the AMPS system, there are eight power levels. In TDMA there are an additional three levels. Therefore, in total, TDMA has 11 power levels, as shown in Table 15.2. In the carrier-off condition, the output power of the transmitting antenna must fall to −60 dBm within 2 ms. In the carrier-on condition, the output power of the transmitting antenna must come to within 3 dB of the specified level.

TABLE 15.2 Mobile Station Nominal Power Levels

Mobile station power level (PL)	Mobile attenuation code (MAC)	Nominal effective radiated power, dBW, for mobile station power class							
		I	II	III	IV	V	VI	VII	VIII
0	000	6	2	−2	−2	•	•	•	•
1	001	2	2	−2	−2	•	•	•	•
2	010	−2	−2	−2	−2	•	•	•	•
3	011	−6	−6	−6	−6	•	•	•	•
4	100	−10	−10	−10	−10	•	•	•	•
5	101	−14	−14	−14	−14	•	•	•	•
6	110	−18	−18	−18	−18	•	•	•	•
7	111	−22	−22	−22	−22	•	•	•	•
Dual-mode only									
8					−26 ± 3 dB	•	•	•	•
9					−30 ± 6 dB	•	•	•	•
10					−34 ± 9 dB	•	•	•	•

Speech coding (full rate). The NA-TDMA speech coding is a class of speech coding known as code excited linear predictive (CELP) coding. The code is called vector-sum excited linear predictive (VSELP) coding. It uses a codebook to vector-quantize the excitation (residual) signal such that the computation required for the codebook search process at the sender can be significantly reduced. The speech coder sampling rate is 7950 bps. Speech is broken into frames; each frame is 20 ms long and contains 160 symbols. Each frame is further divided into subframes 40 samples (5 ms) long. At the mobile station, the analog speech is converted to uniform pulse-code modulation (PCM) format. The speech coder is preceded by the following voice processing stages: (1) level adjustment, (2) bandpass filter, and (3) analog-to-digital conversion. The VSELP speech decoder is shown in Figure 15.19. The first part is the generation of pulse excitation and the second part is the speech waveform synthesis. Adding the two parts results in quality speech. All the values of parameters H, I, γ, β, L, $\alpha_1 \ldots \alpha_{10}$ for a 20-ms frame of speech are received with a low rate of transmission. Then those parameters are inserted into proper places in the speech decoder, and the speech is recovered at the receiving end.

The delays due to the air interface between the base station and the mobile station may exceed 100 ms, and echo control measures therefore necessary. In a half-rate speech coder, the speech frame of 20 msec may contain 80 symbols.

Modulation. NA-TDMA uses a constant envelope modulation with $\pi/4$-shifted differential quadrature phase shift keying (DQPSK). The modulation scheme uses the phase constellation shown in Fig. 15.20.

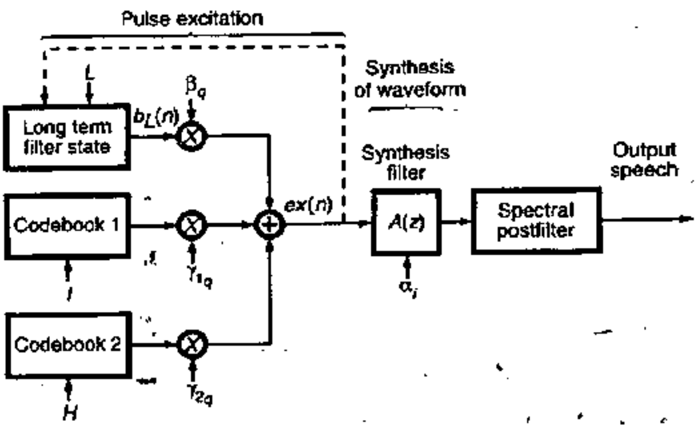


Figure 15.19 VSELP speech decoder.

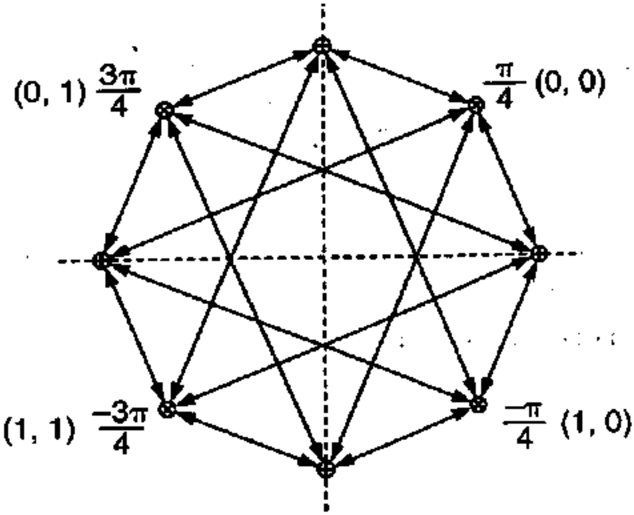


Figure 15.20 Phase constellation of a $\pi/4$-shifted DQPSK.

The rotation of $\pi/4$ occurs alternately at the odd state ⊕ and even state ⊗. The Gray code is used in the mapping. Every signal phase represents a di-bit symbol as shown in Fig. 15.20. Any two adjacent signal phases differ only in a single bit. The information is encoded differentially; symbols do not correspond to absolute phases, but to the phase difference between two adjacent symbols. A binary data stream b_m is separated into two streams: X_k is the even-numbered bit stream, and Y_k is the odd-numbered bit stream (Fig. 15.21). The streams $\{X_k\}$ and $\{Y_k\}$ are encoded onto $\{I_k\}$ and $\{Q_k\}$ by:

$$I_k = I_{k-1} \cos [\Delta\phi(X_k, Y_k)] - Q_{k-1} \sin [\Delta\phi(X_k, Y_k)] \quad (15.2\text{-}1)$$

$$Q_k = I_{k-1} \sin [\Delta\phi(X_k, Y_k)] + Q_{k-1} \cos [\Delta\phi(X_k, Y_k)] \quad (15.2\text{-}2)$$

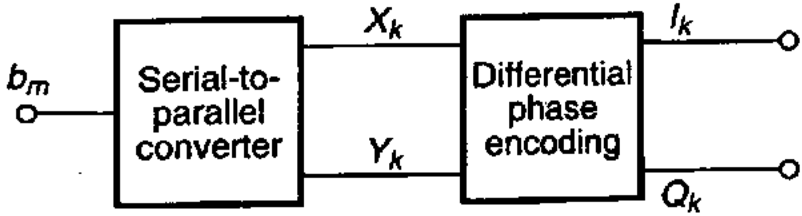


Figure 15.21 A binary data stream conversion.

where I_{k-1}, Q_{k-1} are the amplitudes at the previous pulse time. The phase change $\Delta\phi$ is determined by the following table:

X_k	Y_k	$\Delta\phi$ (even state)	$\Delta\phi$ (odd state)
1	1	$-3\pi/4$	π
0	1	$3\pi/4$	$\pi/2$
0	0	$\pi/4$	0
1	0	$-\pi/4$	$-\pi/2$

The signals I_k and Q_k at the output of the differential phase encoding device can take one of the four values 0, ± 1, $1/\sqrt{2}$.

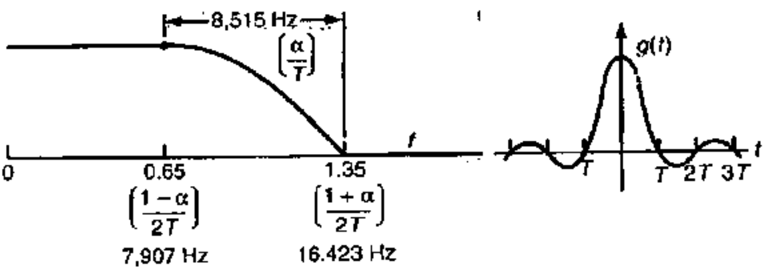
The baseband filters. The baseband filter shall have (1) linear phase and (2) square-root-raised cosine frequency response as shown in Fig. 15.22 where T is the period that equals 41.1 μs. The QPSK modulation with two components I_k and Q_k is differentially encoded as shown in Fig. 15.21.

The transmitted signal. The resultant transmitted signal $S(t)$ is given by:

$$S(t) = \sum_n g(t - nT) \cos \phi_n \cdot \cos w_c t + \sum_n g(t - nt) \sin \phi_n(t) \sin w_c t \qquad (15.2\text{-}3)$$

where $g(t)$ is the pulse shaping with a time response of $H(f)$:

$$g(t) = \frac{1}{2\pi} \int_{-\infty}^{\infty} H(f) e^{j w_c t} \, dt \qquad (15.2\text{-}4)$$



$$|H(f)| = \sqrt{\frac{1}{2}\left\{1 - \sin\left[\frac{\pi(2fT - 1)}{0.70}\right]\right\}} \qquad 7{,}907 \text{ Hz} < f < 16{,}423 \text{ Hz}$$

Figure 15.22 Baseband filter characteristics.

and ω_c is the radian carrier frequency. ϕ_n is from the differential encoding

$$\phi_n = \phi_{n-1} + \Delta\phi_n \tag{15.2-5}$$

15.2.4 Time alignment and limitation of emission

Time alignment. It is necessary to control the TDMA time slot burst (advancing or retarding) transmission from the mobile unit, so that it arrives at the base station receiver in the proper time relationship with respect to other time slot burst transmissions. An error in time alignment causes errors in two signals in the overlap at the head or tail of a time slot.

System access. The mobile station receives an initial traffic channel designation (ITCD) message (order code 01110) which is contained in word 2 (extended address word), then moves to a traffic channel. The mobile station first synchronizes to the forward traffic channel. The time alignment is sent by a physical layer control message over a shortened burst transmission. The mobile station, while operating on a digital traffic channel, is transmitting over a slot interval 324 bits long at certain times. The mobile station continues to transmit a shortened burst at the standard offset reference position until a time alignment message is received from the base station. The mobile station adjusts its transmission time during the next available slot.

Time alignment in handoff message. A mobile handoff message contains estimated time alignment information. Analog-to-digital and digital-to-digital handoff messages contain a shortened burst indicator (SBI) field:

SBI = 00	A handoff to a small-diameter cell
SBI = 01	A handoff from sector to sector
SBI = 10	A handoff to a large-diameter cell.

The shortest burst format

G1	R S D S D V S D W S D X S D Y S	G2

3 symb. 22 symb.

The shortened burst contains:

G1	3 symbol length guard time
R	3 symbol length ramp time

S 14 symbol length sync work

D 6 symbol length coded digital verification color code (CDVCC) (on reverse channel)

G2 22 symbol length guard time

The fields V, W, X, Y consist of

V = 4 zero bits (2 symbols)

W = 8 zero bits (4 symbols)

X = 12 zeros (6 symbols)

Y = 16 zeros (8 symbols)

In the shortened burst format, the symbol interval between any two sync words (total 6 sync words) is a unique interval. After detection of any two or more sync words, the timing alignment is determined at the base station.

Limitations on emissions from digital transmission The total emission power is shown in Fig. 15.23. This limitation is for the suppression of the energy within the cellular band. In addition, the transmitter emissions in each 30-kHz band anywhere in the mobile station receive band must not exceed −80 dBm at the transmit antenna connector.

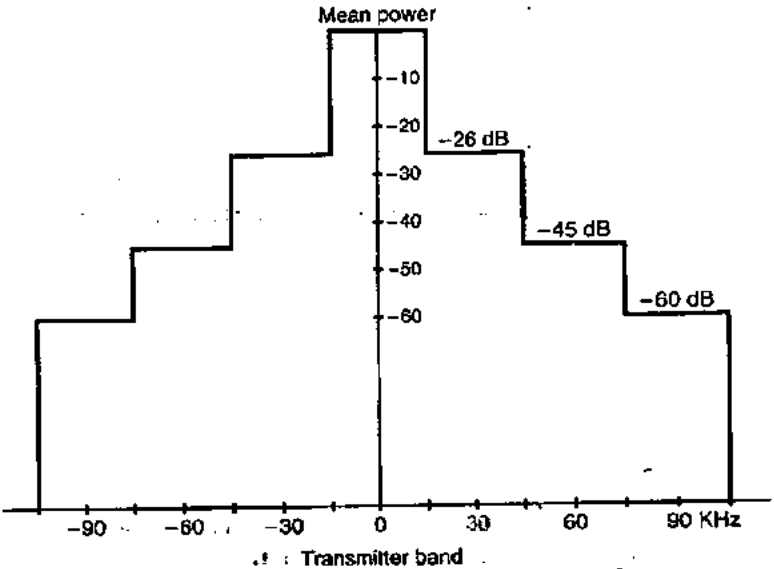


Figure 15.23 Suppression inside cellular band.

15.2.5 Error corrections

Speech data classes. Channel error correcting for the speech code employs three mechanisms:

1. A rate one-half convolutional code to protect the more vulnerable bits of the speech code.
2. Interleaving the transmitted data for each speech coder frame over two time slots to reduce the burst error due to Rayleigh fading.
3. Use of a cyclic redundancy check (CRC). After the error correction is applied at the receiver, these CRC bits are checked to see if most perceptually significant bits were received properly.

The 159-bit speech coder frame is separated into two classes:

Class 1	77 bits
Class 2	82 bits

Class 1 bits are the important bits to which the convolutional coding is applied. Among the 77 bits, there are 12 most perceptually significant bits in which a 7-bit CRC is used for error detection purposes. Class 2 bits are unimportant bits and are transmitted without any error protection.

Cyclic redundancy check. The 12 most perceptually significant bits of the 77 bits are coded in CRC. The generator polynomial is

$$g(X) = 1 + X + X^2 + X^4 + X^5 + X^7 \tag{15.2-6}$$

A 7-bit CRC is used for error detection if one or more of the 12 significant bits are in error.

The 12 most significant bits form an input polynomial:

$$a(X) = \sum_{k=0}^{11} B_k X^k \tag{15.2-7}$$

The polynomial $b(X)$ is the remainder of the division of $a(x)$ and $g(x)$ obtained from

$$\frac{a(X)X^7}{g(X)} = q(X) + \frac{b(X)}{g(X)} \tag{15.2-8}$$

where $q(X)$ is the quotient of the division which is discarded.

The remainder $b(X)$ can be generated from Eq. (15.2-6) and Eq. (15.2-7)

$$b(X) = \sum_{k=1}^{7} C_k X^{k-1} \tag{15.2-9}$$

The input 77 bits $B_1 \cdots B_{77}$ (including 12 significant bits) adding $C_1 \cdots C_7$ and 5 zero bits become 89 important bits.

0-3	4 - 80	81 - 82	84 - 88
4 bits CRC	77 bits	3 bits CRC	5 bits (five 0's) tail bits

A cyclic redundancy check is performed at the receiving end. After decoding of the class 1 bits, the received CRC $b'(x)$ bits are checked to determine if any errors have been detected. The process of checking the error in CRC uses the received 12 most perceptually significant bits $a'(x)$ in each frame divided by the generator polynomial in Eq. (15.2-6):

$$\frac{a'(X) \cdot X^7}{g(X)} = q'(X) + \frac{b''(X)}{g(X)} \tag{15.2-10}$$

The received CRC $b'(x)$ are compared with the CRC bits $b''(x)$ generated by Eq. (15.2-10); if $b'(x) \neq b''(x)$, an error has occurred. The causes of error are (1) the data was corrupted by channel errors or (2) an FACCH message was transmitted in place of the speech data. As a result, the speech quality is degraded. A bad frame masking strategy is taking place. There are six states. The state 0 means no error is detected. When each successive speech frame is found to be in error, the state machine moves to the next higher state. Moving to a higher state means more repeats. If two successive frames occur with no detected errors, the state machine is returned to state 0.

15.2.6 Interleaving and coding

Convolutional encoding. The 89 important bits are input to convolutional coder and 176 bits are at the output of the coder. Then adding 82 unimportant bits become 260 bits total for a 20 ms speech frame. Convolutional encoding uses a code rate of 1/2 and memory order 5. The five memory elements generate 32 states in this code. Since the code is 1/2 rate, two outputs alternately come out and are in a sequential order. CC0 is the convolutional code at one output and the other CC1 is the other output.

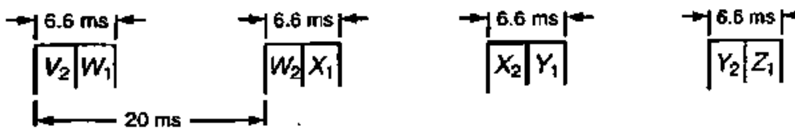
Interleaving and deinterleaving. The encoded speech after the convolutional code is interleaved over two time slots (Fig. 15.24). Each time slot contains two frames. The encoded speech is placed into a rectangular interleaving array columnwise. The two encoded speech frames are referred to as X and Y.

0X	26X · · ·	234X
1Y	27Y	235Y
2X		
3Y		
24X	50X	258X
25Y	51Y	259Y

The speech code consists of 88 class 1 bits (after CRC coding) and 80 class 2 bits. The class 2 bits are intermixed with the convolutionally coded class 1 bits. The bits in the above array are then transmitted row-wise. The place of the coded class 1 bits and class 2 bits are in a certain mixed order.

Deinterleaving. At the receiving end, each time slot contains the interleaved data from two speech coder frames, X_1 and X_2, which are 20 ms apart. The received data are placed row-wise into a 26 × 10 deinterleaving array. Once the data from the two time slots are used to fill the deinterleaving array, all the data for frames X are available and can be decoded. After deinterleaving one entire speech coder frame is available.

Delay interval requirement. The mobile station and the base station shall have a delay interval compensation of up to one symbol length.



$X = X_1 + X_2$ encoded speech frame

Figure 15.24 Interleaving slots arrangement.

15.2.7 SCM and SID

Station class mark (SCM) must be stored in a mobile station. It formerly used 4 bits for identifying the maximum powers of three different kinds of mobile stations. Now SCM uses 5 bits and can identify eight different power levels.

Power class	Max. power, dBm	Min. power, dBm	Number of power levels	SCM	Transmission	SCM	Bandwidth	SCM
I	6	−22	0–7	0XX00	Continuous	XX0XX	20 MHz	X0XXX
II	2	−22	1–7	0XX01	Discontinuous	XX1XX	25 MHz	X1XXX
III	−2	−22	2–7	0XX10				
IV	−2	−34 ± 9 dB	2–10	0XX11				
V				1XX00				
VI				1XX01				
VII				1XX10				
VIII				1XX11				

Home system identification (SID) is a 15-bit system identification indicator

14 — 13	12 — 0
2 bits	System number

00 USA

01 Other countries

10 Canada

11 Mexico

The first two bits indicate the country of origin:

00 United States

01 Other countries

10 Canada

11 Mexico

15.2.8 NA-TDMA channels

In NA-TDMA, there are no common channels such as those used in GSM. The digital call set-up uses the 21 set-up channels which are shared with the analog system.

Supervision of the digital voice channel. The supervision channels in NADC are similar to those in GSM:

- *Fast associated control channel.* FACCH is a blank and burst channel equivalent to a signaling channel for the transmission of control and supervision messages between the base station and the mobile station. It consists of 260 bits. Mostly FACCH is used for handoff messages.
- *Slot associated control channel.* SACCH is a signaling channel including twelve code bits present in every time slot transmitted over the traffic channel whether these contain voice or FACCH information.

Mobile-assisted handoffs. The mobile station performs signal quality measurements on two types of channels:

1. Measures the RSSI (received signal strength indicator) and the BER (bit error rate) information of the current forward traffic channel during a call.
2. Measures the RSSI of any RF channel which is identified from the measurement order message from the base station.

MAHO consists of three messages:

1. Start measurement order
   - Measurement order message—sent from the base station to the mobile station.
   - Measurement order acknowledge message—sent from the mobile station to the base station.
2. Stop measurement order
   - Stop measurement order—sent from the base station to the mobile station.
   - Mobile acknowledge—sent from the mobile station to the base station.
3. Channel quality message (mobile to base only)

The mobile transmits the signal quality information over either the SACCH or FACCH. In the case of discontinuous transmission (DTX):

- Whenever the mobile is in the DTX high state, the mobile transmits channel quality information over the SACCH
- When the mobile is in the DTX low state, the mobile transmits the channel quality information over the FACCH

Handoff action. When a handoff order is received, the mobile station is at DTX high state and stays at that state. If the mobile station is at DTX low state it must enter the DTX high state and wait for 200 ms before taking the handoff action. Handoff to a digital traffic channel is described as follows:

1. Turn on signaling tone for 50 ms, turn off signaling tone, turn off transmitter which was operating on the old frequency.
2. Adjust power, tune to new channel, set stored DVCCs to the DVCC field of the received message.
3. Set the transmitter and receiver to digital mode, set the transmit and receive rate based on the message-type field.
4. Set time slot based on the message-type field.
5. Set the time alignment offset to the value based on the TA field.
6. Once the transmitter is synchronized, enter the conversation task of the digital traffic channel.

15.2.9 Discontinuous transmission on a digital traffic channel

In DTX, certain mobile stations can switch autonomously between two transmitter power-level states: DTX high and DTX low. In the DTX high state, the power level of transmitter at the mobile station is indicated by the most recent power-controlling order. In this state, the CDVCC (coded digital verification color code) is sent at all times. CDVCC is used to distinguish the current traffic channel from traffic co-channels. Decoding CDVCC (12, 8) becomes $DVCC_r$. $DVCC_r$ is received and checked with DVCCs for identification. There are 255 codes (2^8, but 0 is not used). In the DTX low state, the transmitter remains off and CDVCC is not sent except for the transmission of FACCH messages. All the SACCH messages will be sent as a FACCH message. After sending all the messages, the transmitter will return to the off state.

15.2.10 Authentication

A secret number PIN (personal identification number) is assigned to each subscriber. The mobile station, on receipt of a random challenge global action message, updates its internal stored RAND variable used as input to the authentication algorithm AUTH1. The mobile uses its PIN, its ESN and the MIN to compute a response to RAND in accordance with AUTH1. The mobile then responds to the page by transmitting its MIN, the output of AUTH1, RANDC (random confirmation), and COUNT (call history parameter). See also Sec. 15.34 for similarity.

15.2.11 Signaling format

A reverse digital traffic channel (RDTC) is used to transport user information and signaling. Two control channels are used: the FACCH is a blank and burst channel, the SACCH is a continuous channel, and interleaving is on the SACCH. The signaling formats of these two channels are shown in Fig. 15.25.

Interleaving on the FACCH bits from 0 to 259 is

Row number	FACCH bits interleaving									
0	215	256	223	258	230	219	257	227	259	189
1	0	25	50	75	231	89	114	139	164	190
2	1	26	51	76	232	90	115	140	165	191
⋮	⋮									
14	13	38	63	88	244	102	127	152	177	203
15	14	39	64	216	245	103	128	153	178	204
⋮	⋮									
25	24	49	74	229	255	113	138	163	118	214

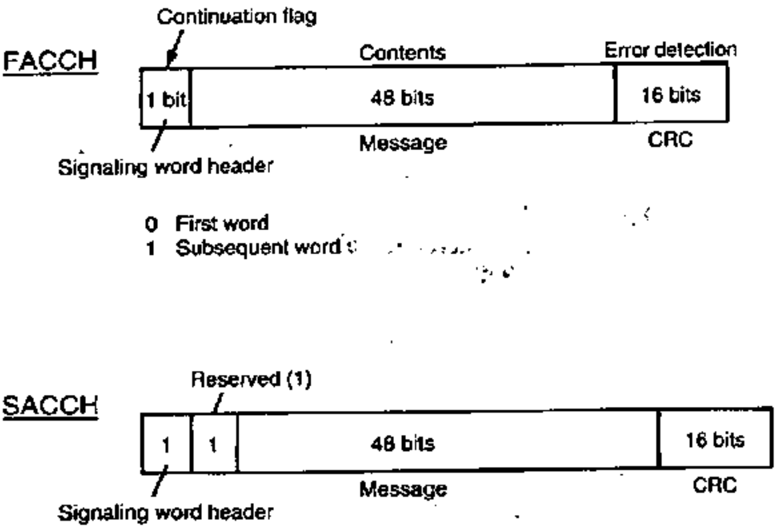
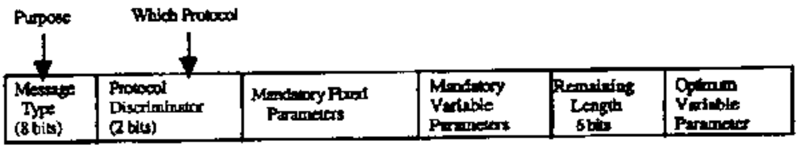


Figure 15.25 Signaling formats of FACCH and SACCH.

The Y row (odd) of this frame combines alternately with the X row (even) of the previous frame to form a FACCH block. In SACCH code, the output of the convolutional coder is diagonally interleaved as the 12 coded bits are transmitted over 12 time slots.

Message structure. All messages contain:

1. An application message header
2. Mandatory fixed parameters
3. Mandatory variable parameters
4. Remaining length
5. Optional variable parameters



A forward digital traffic channel (FDTC) has same format as the RDTC (reverse digital traffic channel).

15.2.12 Word format

The same word format is used for FACCH and SACCH.

15.3 CDMA[10–12]

CDMA development started in early 1989 after the NA-TDMA standard (IS-54) was established. A CDMA demonstration to test its feasibility for digital cellular systems was held in November 1989. The CDMA "Mobile Station-Base Station Compatibility Standard for Dual Mode Wideband Spread Spectrum Cellular System," was issued as IS-95 (PN-3118, Dec. 9, 1992). CDMA uses the idea of tolerating interference by spread-spectrum modulation. The power control scheme in a CDMA system is a requirement for digital cellular application. However, it was a challenging task and has been solved. Before describing the structure of the system, we list the key terms of CDMA systems.

15.3.1 Terms of CDMA systems

Active set: The set of pilots associated with the CDMA channels containing forward traffic channels assigned to a particular mobile station (MS).

CDMA channel number An 11-bit number corresponding to the center of the CDMA frequency assignment.

Code channel A subchannel of a forward CDMA channel. A forward CDMA channel contains 64 code channels. Certain code channels are assigned to different logic channels.

Code channel zero: Pilot channel.

Code channels 1 through 7: Either paging channels or traffic channels.

Code channel 32: A sync channel or a traffic channel.

The remaining code channels are traffic channels.

Code symbol The output of an error-correcting encoder.

Dim-and-burst A frame in which the primary traffic is multiplexed with either secondary traffic or signal traffic. It is equivalent to the blank-and-burst function in AMPS.

Forward CDMA channel Contains one or more code channels.

Frame A basic timing interval in the system. For the access channel, paging channel, and traffic channel, a frame is 20 ms long. For the sync channel, a frame is 26.666 ms long.

Frame offset A time skewing of traffic channel frames from system time in integer multiples of 1.25 ms. The maximum frame offset is 18.75 ms.

GPS (Global Position System) System used for providing location and time information to the CDMA system.

Handoff (HO) The act of transferring communication with a mobile station from one base station to another.

Hard HO Occurs when (1) the MS is transferred between disjoint active sets, (2) the CDMA frequency assignment changes, (3) the frame offset changes, and (4) the MS is directed from a CDMA traffic channel to an analog voice channel.

Soft HO HO from CDMA cell to CDMA cell at the same CDMA frequency.

Idle HO Occurs when the paging channel is transferred from one base station (BS) to another.

Layering A method of organization for communication protocols. A layer is defined in terms of its communication protocol to a peer layer.

Layer 1: Physical layer presents a frame by the multiplex sublayer and transforms it into an over-the-air waveform.

Layer 2: Provides for the correct transmission and reception of signaling messages.

Layer 3: Provides the control of the cellular telephone system. The signaling messages orginate and terminate at layer 3.

Long code A PN (pseudonoise) sequence with period 2^{42}-1 using a tapped n-bit shift register.

Modulation symbol The output of the data modulator before spreading. There are 64 modulation symbols on the reverse traffic channel, 64-ary orthogonal modulation is used, and six code symbols are associated with one modulation symbol. On the forward traffic channel, each code symbol (data rate is 9600 bps) or each repeated code symbol (data rate is less than 9600 bps) is 1 modulation symbol.

Reverse: $\overset{53}{\overline{110101}}\ \overset{48}{\overline{101110}} \longrightarrow$

$\text{(Walsh } \overset{64\text{ bits}}{\text{function}} \text{ 53) (Walsh } \overset{64\text{ bits}}{\text{function}} \text{ 48)}$

6 code symbols $\longrightarrow$ 1 modulation symbol

Forward 1 code symbol = 1 modulation symbol

Multiplex option The ability of the multiplex sublayer and lower layers to be tailored to provide special capabilities. A multiplex option defines the frame format and the rate decision rules.

Multiplex sublayer One of the conceptual layers of the system that multiplexes and demultiplexes primary traffic, secondary traffic, and signaling traffic.

Nonslotted mode An operating mode of an MS in which the MS continuously monitors the paging channel.

Null traffic data A frame of sixteen 1's followed by eight 0's sent at the 1200 bps rate. Null traffic channel data serve to maintain the connectivity between MS and BS when no service is active and no signaling message is being sent.

Paging channel A code channel in a forward CDMA channel used for transmission of (1) control information and (2) pages from BS to MS. The paging channel slot has a 200-ms interval.

Power control bit A bit sent in every 1.25 ms intervals on the forward traffic channel to the MS that increases or decreases its transmit power.

Primary CDMA channel A preassigned frequency used by the mobile station for initial acquisition.

Primary paging channel The default code channel (code channel 1) assigned for paging.

Primary traffic The main traffic stream between MS and BS on the traffic channel.

Reverse traffic channel Used to transport user and signaling traffic from a single MS to one or more BSs.

Shared secret data (SSD) A 128-bit pattern stored in the MS.

> SSD is a concatenation of two 64-bit subsets.
>
> SSD-A is used to support the authentication
>
> SSD-B serves as one of the inputs to generate the encryption mask and private long code.

Secondary CDMA channel A preassigned frequency (one of two) used by the mobile station for initial acquisition.

Secondary traffic An additional traffic stream carried between the MS and the BS on the traffic channel.

Slotted mode An operation mode of MS in which the MS monitors only selected slots on the paging channel.

Sync channel Code channel 32 in the forward CDMA channel which transports the synchronization message to the MS.

Pilot channel An unmodulated, direct-sequence (DS) signal transmitted continuously by each CDMA BS. The pilot channel allows a mobile station to acquire the timing of the forward CDMA channel, provides a phase reference for coherent demodulation, and provides a means for signal strength comparisons between base stations for determining when to hand off.

System time The time reference used by the system. System time is synchronous to universal time coordination (UTC) time and uses the same time origin as GPS time. All BSs use the same system time. MSs use the same system time, offset by the propagation delay from the BS to the MS.

Time reference A reference established by the MS that is synchronous with the earliest arriving multipath component which is used for demodulation. The time reference establishes transmit time and the location of zero in PN space.

Walsh chip The shortest identifiable component of a 64-walsh function. On the forward CDMA channel, one chip equals 1/1.2288 MHz or 813.802 ns. On the reverse CDMA channel, one chip equals 4/1.2288 MHz or 3255 ns.

15.3.2 Output power limits and control

Output power. The mean output power of the mobile station shall be less than −50 dBm/1.23 MHz (−111 dBm/Hz) for all frequencies within ±615 kHz of the center frequency.

Gated output power. The MS shall transmit at nominal controlled levels during gated-on periods. A typical output power in a gated-on period is shown in Figure 15.26. The transmitter noise floor should be less than −60 dBm/1.23 MHz.

Controlled output power. Implementing CDMA power control is a must in the cellular CDMA system for the reverse link transmission in order to eliminate the near-far interference. If all the mobile transmitters

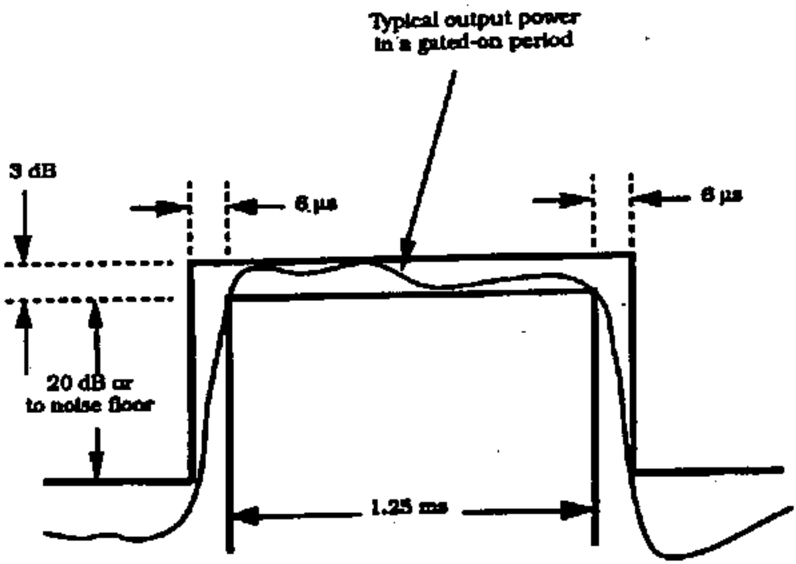


Figure 15.26 Transmission envelope mask (single gated-on power control group).

within a cell site's area of coverage are so controlled, then the total signal power received at the cell site from all mobiles will be equal to the nominal received power times the number of mobiles.

CDMA reverse-link open-loop power control. The mobile station receives a signal suffering both the log-normal and Rayleigh fadings from the forward link, as shown in Fig. 15.27*a*. The average path loss is obtained as shown in the figure. If the transmitting and receiving ends are sharing the same frequency channel, then reversing the received signal strength as shown in Fig. 15.27*b*, indicated as the transmit power without smoothing filter, would eliminate the power variation at the cell site. Since CDMA uses duplexing channels, the Rayleigh fading on the forward channel and the reverse channel are not the same. Therefore, the desired average transmit power is sent back on the reverse channel.

At the cell site, the available information on instantaneous value versus the expected value of frame error rate (FER) of the received signal is examined to determine whether to command a particular mobile to increase or decrease its transmit power. This mechanism is called CDMA closed-loop power control. The mobile power received at a cell site after close-loop control is shown in Fig. 15.27c.

In transmission mode, the MS has two independent means for output power adjustment:

1. Open-loop output power
   - The mobile station shall transmit the first probe on the access channel:

     $$\overline{P_A} = \text{mean output power, dBm} = -\text{mean input power, dBm} - 73 + \text{NOMPWR, dB} + \text{INITPWR, dB}$$

     where NOMPWR = the correction of received power at the base station and INITPWP = adjustment of the received power less than the required signal power. When INITPWR = 0, $\overline{P_A} = \pm 6$ dB.
   - For initial transmission on the reverse channel,

     $$\overline{P_I} = \text{mean output power, dBm} = \overline{P_A} + \text{the sum of all access probe correction, dBm}$$
   - For normal reverse traffic channel,

     $$\overline{P_R} = \text{mean output power, dBm} = \overline{P_I} + \text{the sum of all closed-loop power control correction, dB}$$

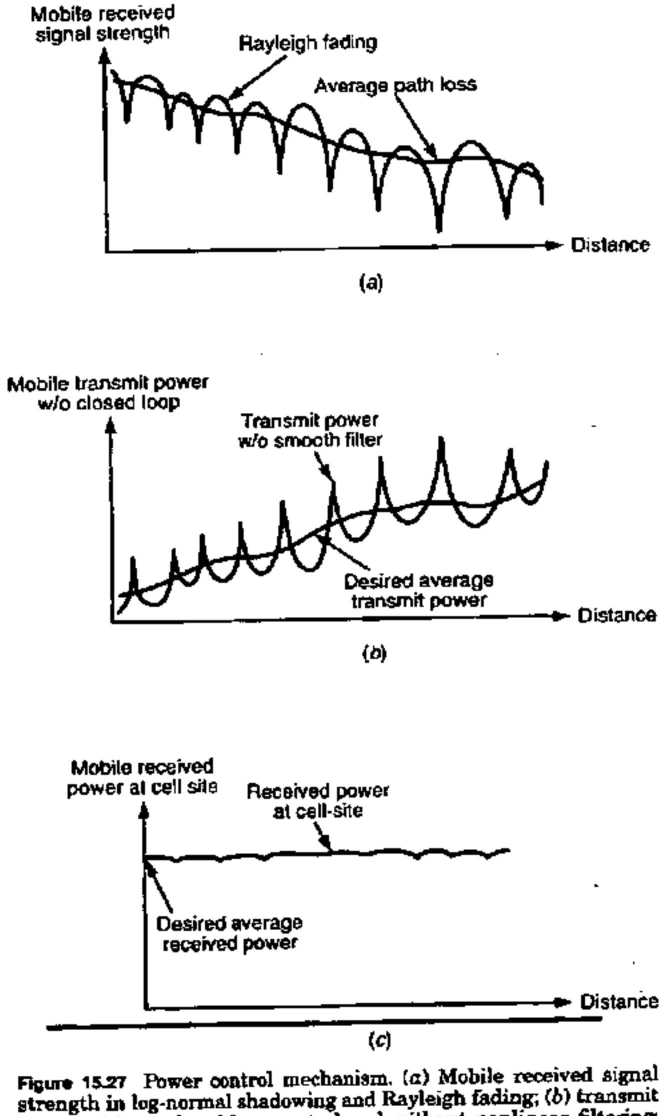


Figure 15.27 Power control mechanism. (*a*) Mobile received signal strength in log-normal shadowing and Rayleigh fading; (*b*) transmit power without closed-loop control and without nonlinear filtering; (*c*) mobile power received at cell site.

- For example, without any correction or adjustment,

$$\begin{aligned}\text{Mean output power} &= -\text{ mean input power} - 73\\ &= -(-90\text{ dBm}) - 73\\ &= +17\text{ dBm}\end{aligned}$$

2. Closed-loop output power (involving both the mobile station and base station). The mobile station shall adjust its mean output power level in response to each valid power control bit received on the forward traffic channel. The change in mean output power per single power control bit shall be 1 dB nominal, within ±0.5 dB of the nominal change.

15.3.3 Modulation characteristics

Reverse CDMA channel signals. The reverse CDMA channel is composed of access channels and reverse traffic channels. Since the MS does not establish a system time as at the BS, the reverse channel signal received at the BS cannot use coherent detection. Thus the modulation characteristics for the forward channel and reverse channel are different. The modulation of the reverse channel is 64-ary orthogonal modulation at a data rate of 9600, 4800, 2400, or 1200 bps, as shown in Fig. 15.28 at point A. The actual burst transmission rate is fixed at 28,800 code symbols per second. This results in a fixed Walsh chip rate of 307.2 thousand chips per second (kcps). Each Walsh chip is spread by four PN chips. The rate of the spreading PN sequence is fixed at 1.2288 million chips per second (Mcps). The reverse traffic channel modulation parameters and the access channel modulation parameters are listed in Tables 15.3 and 15.4, respectively.

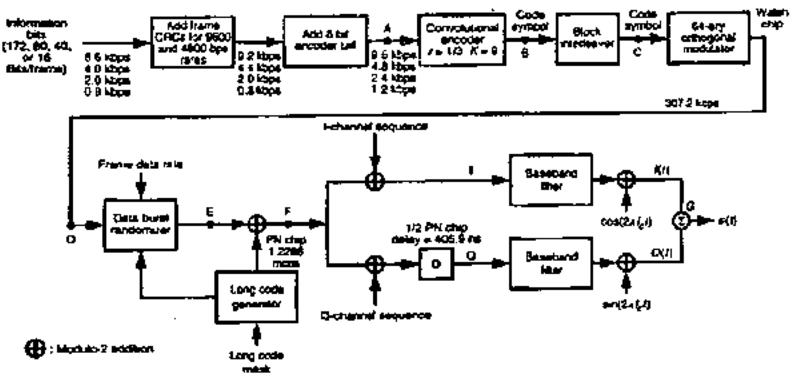


Figure 15.28 Reverse CDMA channel modulation process.

TABLE 15.3 Reverse Traffic Channel Modulation Parameters

Parameter	Data rate, bps 9600	4800	2400	1200	Units
PN chip rate	1.2288	1.2288	1.2288	1.2288	Mcps
Code rate	⅓	⅓	⅓	⅓	bits/code symbol
Transmit duty cycle	100.0	50.0	25.0	12.5	%
Code symbol rate	28,800	28,800	28,800	28,800	sps
Modulation	6	6	6	6	code symbol/ mod symbol
Modulation symbol rate	4800	4800	4800	4800	sps
Walsh chip rate	307.20	307.20	307.20	307.20	kcps
Mod. symbol duration	208.33	208.33	208.33	208.33	μs
PN chips/code symbol	42.67	42.67	42.67	42.67	PN chip/code symbol
PN chips/mod. symbol	256	256	256	256	PN chip/mod symbol
PN chips/Walsh chip	4	4	4	4	PN chips/Walsh chip

Convolutional encoding. At point B in Fig. 15.28, with a $K = 9$ (9 register) and rate 1/3 convolutional encoder:

1. On the access channel, each code symbol has a fixed data rate of 4800 bps, and each symbol repeats one time consecutively.
2. On the reverse traffic channel, the full data rate is 9600 kbps. For the data rate of 4800 kbps, each symbol repeats one time consecutively. For the data rate of 2400 kbps, each symbol repeats three times consecutively. For the data rate of 1200 kbps, each symbol repeats seven times consecutively.

TABLE 15.4 Access Channel Modulation Parameters

Parameter	Data rate, bps 4800	Units
PN chip rate	1.2288	Mcps
Code rate	⅓	bits/code symbol
Code symbol repetition	2	symbols/code symbol
Transmit duty cycle	100.0	%
Code symbol rate	28,800	sps
Modulation	6	code sym/mod. symbol
Modulation symbol rate	4800	sps
Walsh chip rate	307.20	kcps
Mod. symbol duration	208.33	μs
PN chips/code symbol	42.67	PN chip/code symbol
PN chips/mod. symbol	256	PN chip/mod. symbol
PN chips/Walsh chip	4	PN chips/Walsh chip

Interleaving. At point C in Fig. 15.28, the interleaving algorithm will form an array with 32 rows and 18 columns. At 9600 kbps, the interleaver forms a 32 × 18 matrix as in Table 15.5.

At 9600 bps, the transmission sequence is to send row by row in a sequence order up to row 32. At 4800 bps, the transmission sequence is to send by the unique order of rows as follows:

Row Number →
1 3 2 4 5 7 6 8 9 11 10 12 13 15 14 16 17 19 18 20 21 23 22 24 25 27 26 28 29 31 30 32,

Expressed in a formula, the transmission sequence is

$$J, J + 2, J + 1, J + 3$$

for $J = 1 + 4i$ and $i = 0, 1, 2, 3, \ldots, (32/4 - 1)$.

At 2400 bps, the transmission sequence is by a unique order of rows as follows:

$$J, J + 4, J + 1, J + 5, J + 2, J + 6, J + 3, J + 7$$

for $J = 1 + 8i$ and $i = 0, 1, 2 \ldots, (32/8 - 1)$.

At 1200 bps,

$$J, J + 8, J + 1, J + 9, J + 2, J + 10, J + 3, J + 11, J + 4,$$
$$J + 12, J + 5, J + 13, J + 6, J + 14, J + 7, J + 15$$

for $J = 1 + 16i$ and $i = 1, 2$.

For access channel code symbols, the interleaver rows follow this order:

TABLE 15.5 Interleaving Algorithm

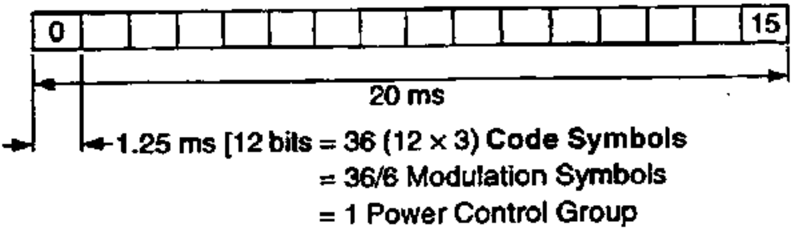
Row \ Column	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18
1	1	33	65	97	129	161	193	225	257	289	321	353	385	417	449	481	513	545
2	2	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
3	3	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
4	4	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
5	5	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
6	6	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮	⋮
32	32	64	96	128	160	192	224	256	288	320	352	384	416	448	480	512	544	576

$$J, J + 16, J + 8, J + 24, J + 4, J + 20, J + 12, J + 28, J + 2,$$
$$J + 18, J + 10, J + 26, J + 6, J + 22, J + 14, J + 30$$

for $J = 1, 2$.

Orthogonal modulation for reverse channel. As at point D in Fig. 15.28, the 64-ary Walsh codes consist of 64 codes each 64 bits long. They are orthogonal to each other as shown in Table 15.6. Every sixth symbol interpretating each Walsh code of 64 chips is sent out. For example,

Each 20-ms reverse traffic channel frame shall be divided into 16 equal-length (i.e., 1.25 ms) power control groups numbered from 0 to 15.



The reverse traffic channel and the access channel shall be direct-sequence spread by the long code prior to transmission. The long code shall be periodic with period $2^{42} - 1$ chips and shall satisfy the linear recursion specified by the polynomial

$$p(x) = x^{42} + x^{35} + x^{33} + x^{31} + x^{27} + x^{26} + x^{25} + x^{22} + x^{21} + x^{19} + x^{18} + x^{17} + x^{16} + x^{10} + x^{7} + x^{6} + x^{5} + x^{3} + x^{2} + x^{1} + 1$$

Each PN chip of the long code shall be generated by a 42-shift-register long-code generator.

Data burst randomizing. At point E in Fig. 15.28, the data burst randomizer has generated a masking pattern of 0s and 1s that randomly masks out the redundant data generated by the code repetition. The mask pattern is determined by the data rate of the frame and by a block of 14 bits taken from the long code. These 14 bits shall be the last 14 bits of the long code used for spreading.

Direct sequence spreading. At point F in Fig. 15.28, prior to transmission, the reverse traffic channel and the access channel are direct-sequence spread by the long code. This spreading operation involves modulo-2 addition of the data burst randomizer output stream and

TABLE 15.6 64-ary Orthogonal Symbol Set

	0123	4567	11 8901	1111 2345	1111 6789	2222 0123	2222 4567	2233 8901	3333 2345	3333 6789	4444 0123	4444 4567	4455 8901	5555 2345	5555 6789	6666 0123
0	0000	0000	0000	0000	0000	0000	0000	0000	0000	0000	0000	0000	0000	0000	0000	0000
1	0101	0101	0101	0101	0101	0101	0101	0101	0101	0101	0101	0101	0101	0101	0101	0101
2	0011	0011	0011	0011	0011	0011	0011	0011	0011	0011	0011	0011	0011	0011	0011	0011
3	0110	0110	0110	0110	0110	0110	0110	0110	0110	0110	0110	0110	0110	0110	0110	0110
4	0000	1111	0000	1111	0000	1111	0000	1111	0000	1111	0000	1111	0000	1111	0000	1111
5	0101	1010	0101	1010	0101	1010	0101	1010	0101	1010	0101	1010	0101	1010	0101	1010
6	0011	1100	0011	1100	0011	1100	0011	1100	0011	1100	0011	1100	0011	1100	0011	1100
7	0110	1001	0110	1001	0110	1001	0110	1001	0110	1001	0110	1001	0110	1001	0110	1001
8	0000	0000	1111	1111	0000	0000	1111	1111	0000	0000	1111	1111	0000	0000	1111	1111
9	0101	0101	1010	1010	0101	0101	1010	1010	0101	0101	1010	1010	0101	0101	1010	1010
10	0011	0011	1100	1100	0011	0011	1100	1100	0011	0011	1100	1100	0011	0011	1100	1100
11	0110	0110	1001	1001	0110	0110	1001	1001	0110	0110	1001	1001	0110	0110	1001	1001
12	0000	1111	1111	0000	0000	1111	1111	0000	0000	1111	1111	0000	0000	1111	1111	0000
13	0101	1010	1010	0101	0101	1010	1010	0101	0101	1010	1010	0101	0101	1010	1010	0101
14	0011	1100	1100	0011	0011	1100	1100	0011	0011	1100	1100	0011	0011	1100	1100	0011
15	0110	1001	1001	0110	0110	1001	1001	0110	0110	1001	1001	0110	0110	1001	1001	0110
16	0000	0000	0000	0000	1111	1111	1111	1111	0000	0000	0000	0000	1111	1111	1111	1111
17	0101	0101	0101	0101	1010	1010	1010	1010	0101	0101	0101	0101	1010	1010	1010	1010
18	0011	0011	0011	0011	1100	1100	1100	1100	0011	0011	0011	0011	1100	1100	1100	1100
19	0110	0110	0110	0110	1001	1001	1001	1001	0110	0110	0110	0110	1001	1001	1001	1001
20	0000	1111	0000	1111	1111	0000	1111	0000	0000	1111	0000	1111	1111	0000	1111	0000
21	0101	1010	0101	1010	1010	0101	1010	0101	0101	1010	0101	1010	1010	0101	1010	0101
22	0011	1100	0011	1100	1100	0011	1100	0011	0011	1100	0011	1100	1100	0011	1100	0011
23	0110	1001	0110	1001	1001	0110	1001	0110	0110	1001	0110	1001	1001	0110	1001	0110
24	0000	0000	1111	1111	1111	1111	0000	0000	0000	0000	1111	1111	1111	1111	0000	0000
25	0101	0101	1010	1010	1010	1010	0101	0101	0101	0101	1010	1010	1010	1010	0101	0101
26	0011	0011	1100	1100	1100	1100	0011	0011	0011	0011	1100	1100	1100	1100	0011	0011
27	0110	0110	1001	1001	1001	1001	0110	0110	0110	0110	1001	1001	1001	1001	0110	0110
28	0000	1111	1111	0000	1111	0000	0000	1111	0000	1111	1111	0000	1111	0000	0000	1111
29	0101	1010	1010	0101	1010	0101	0101	1010	0101	1010	1010	0101	1010	0101	0101	1010
30	0011	1100	1100	0011	1100	0011	0011	1100	0011	1100	1100	0011	1100	0011	0011	1100
31	0110	1001	1001	0110	1001	0110	0110	1001	0110	1001	1001	0110	1001	0110	0110	1001
32	0000	0000	0000	0000	0000	0000	0000	0000	1111	1111	1111	1111	1111	1111	1111	1111
33	0101	0101	0101	0101	0101	0101	0101	0101	1010	1010	1010	1010	1010	1010	1010	1010
34	0011	0011	0011	0011	0011	0011	0011	0011	1100	1100	1100	1100	1100	1100	1100	1100
35	0110	0110	0110	0110	0110	0110	0110	0110	1001	1001	1001	1001	1001	1001	1001	1001
36	0000	1111	0000	1111	0000	1111	0000	1111	1111	0000	1111	0000	1111	0000	1111	0000
37	0101	1010	0101	1010	0101	1010	0101	1010	1010	0101	1010	0101	1010	0101	1010	0101
38	0011	1100	0011	1100	0011	1100	0011	1100	1100	0011	1100	0011	1100	0011	1100	0011
39	0110	1001	0110	1001	0110	1001	0110	1001	1001	0110	1001	0110	1001	0110	1001	0110
40	0000	0000	1111	1111	0000	0000	1111	1111	1111	1111	0000	0000	1111	1111	0000	0000
41	0101	0101	1010	1010	0101	0101	1010	1010	1010	1010	0101	0101	1010	1010	0101	0101
42	0011	0011	1100	1100	0011	0011	1100	1100	1100	1100	0011	0011	1100	1100	0011	0011
43	0110	0110	1001	1001	0110	0110	1001	1001	1001	1001	0110	0110	1001	1001	0110	0110
44	0000	1111	1111	0000	0000	1111	1111	0000	1111	0000	0000	1111	1111	0000	0000	1111
45	0101	1010	1010	0101	0101	1010	1010	0101	1010	0101	0101	1010	1010	0101	0101	1010
46	0011	1100	1100	0011	0011	1100	1100	0011	1100	0011	0011	1100	1100	0011	0011	1100
47	0110	1001	1001	0110	0110	1001	1001	0110	1001	0110	0110	1001	1001	0110	0110	1001
48	0000	0000	0000	0000	1111	1111	1111	1111	1111	1111	1111	1111	0000	0000	0000	0000
49	0101	0101	0101	0101	1010	1010	1010	1010	1010	1010	1010	1010	0101	0101	0101	0101
50	0011	0011	0011	0011	1100	1100	1100	1100	1100	1100	1100	1100	0011	0011	0011	0011
51	0110	0110	0110	0110	1001	1001	1001	1001	1001	1001	1001	1001	0110	0110	0110	0110
52	0000	1111	0000	1111	1111	0000	1111	0000	1111	0000	1111	0000	0000	1111	0000	1111
53	0101	1010	0101	1010	1010	0101	1010	0101	1010	0101	1010	0101	0101	1010	0101	1010
54	0011	1100	0011	1100	1100	0011	1100	0011	1100	0011	1100	0011	0011	1100	0011	1100
55	0110	1001	0110	1001	1001	0110	1001	0110	1001	0110	1001	0110	0110	1001	0110	1001
56	0000	0000	1111	1111	1111	1111	0000	0000	1111	1111	0000	0000	0000	0000	1111	1111
57	0101	0101	1010	1010	1010	1010	0101	0101	1010	1010	0101	0101	0101	0101	1010	1010
58	0011	0011	1100	1100	1100	1100	0011	0011	1100	1100	0011	0011	0011	0011	1100	1100
59	0110	0110	1001	1001	1001	1001	0110	0110	1001	1001	0110	0110	0110	0110	1001	1001
60	0000	1111	1111	0000	1111	0000	0000	1111	1111	0000	0000	1111	0000	1111	1111	0000
61	0101	1010	1010	0101	1010	0101	0101	1010	1010	0101	0101	1010	0101	1010	1010	0101
62	0011	1100	1100	0011	1100	0011	0011	1100	1100	0011	0011	1100	0011	1100	1100	0011
63	0110	1001	1001	0110	1001	0110	0110	1001	1001	0110	0110	1001	0110	1001	1001	0110

the long code. This long code shall be periodic with period $2^{42} - 1$ chips.

Quadrature spreading. The sequences used for the spread in quadrature are shown in Fig. 15.28 at point F. These sequences are periodic with period 2^{15} chips, and the spread polynomials of channel I and Q pilot PN sequences are

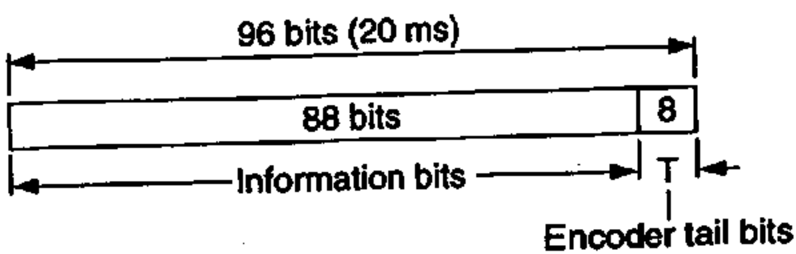
$$P_I(x) = x^{15} + x^{13} + x^9 + x^8 + x^7 + x^5 + 1$$

$$P_Q(x) = x^{15} + x^{12} + x^{11} + x^{10} + x^6 + x^5 + x^4 + x^3 + 1$$

which are of period $2^{15} - 1$. The pilot PN sequence repeats every 26.66 ms (2^{15}/1228800 s). There are exactly 75 repetitions in every 2 s. Reverse CDMA channel I and Q mapping for an offset QPSK modulation is shown in Fig. 15.29.

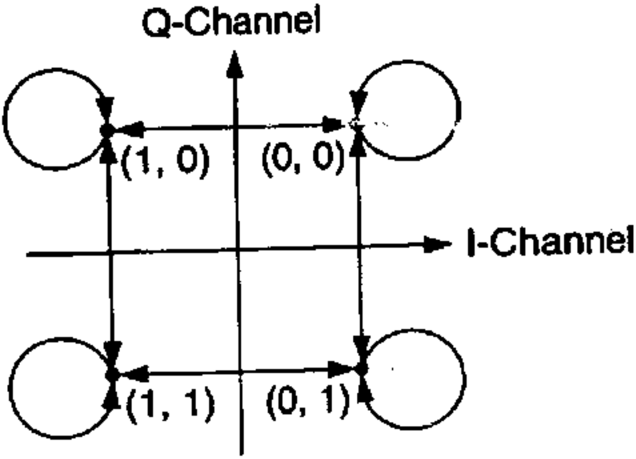
Access channel and reverse traffic channel

1. Access channel
   - Time alignment—an access channel frame shall begin only when system time is an integral multiple of 20 ms.
   - Modulation rate—a fixed rate of 4800 bps.
   - The reverse CDMA channel may contain up to 32 access channel numbers, 0 through 31, per supported paging channel (Fig. 15.30*a*). Each access channel is associated with a single paging channel on the corresponding forward CDMA channel (Fig. 15.30*b*). The forward CDMA channel structure will be described later.
   - Frame structure:



I	Q	Phase
0	0	$\pi/4$
1	0	$3\pi/4$
1	1	$-3\pi/4$
0	1	$-\pi/4$

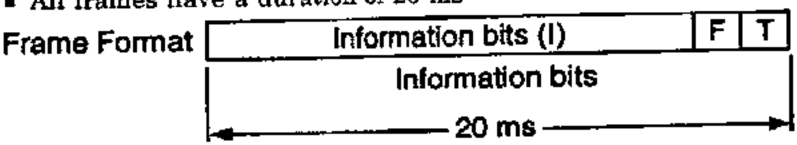
(*a*)



(*b*)

Figure 15.29 Reverse CDMA channel quadrature spreading. (*a*) Reverse CDMA channel I and Q mapping; (*b*) offset QPSK constellation and phase transition.

2. Reverse traffic channel
   - A variable data rate of 9600, 4800, 2400, or 1200 bps
   - All frames have a duration of 20 ms



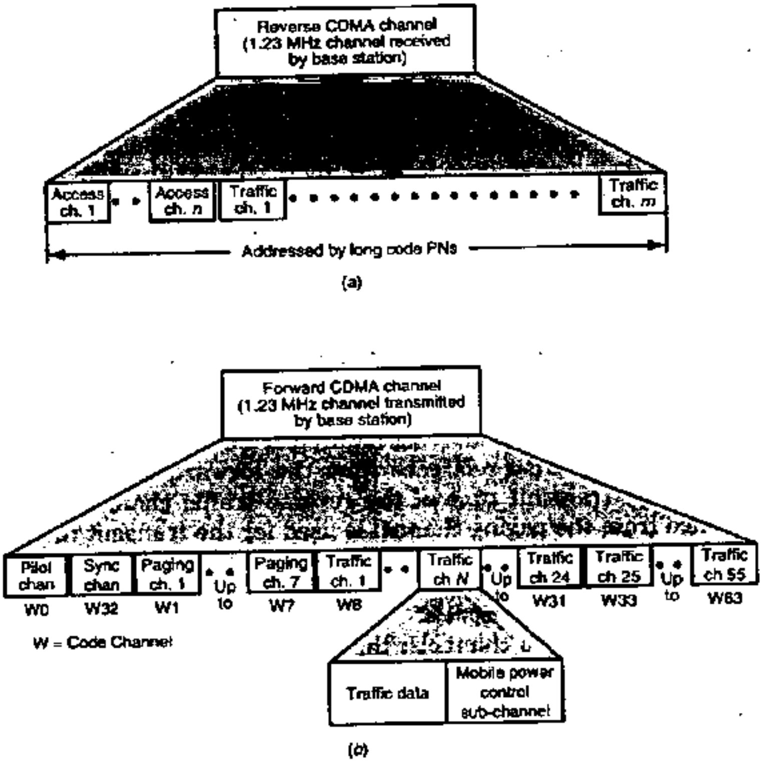


Figure 15.30 CDMA channel structure. (*a*) Example of logical reverse CDMA channels received at a base station; (*b*) example of a forward CDMA channel transmitted by a base station.

Information bits (I)	172 bits (for 9600 bps)
	80 bits (for 4800 bps)
	40 bits (for 2400 bps)
	16 bits (for 1200 bps)
Frame quality indicator (F)	12 bits (for 9600 bps)
(detect errors by CRC)	8 bits (for 4800 bps)
	0 bits (for 2400 bps)
	0 bits (for 1200 bps)
Tail bits (T)	8 bits for all data rates

where the generator polynomials for frame quality indicators are

$$g(x) = x^{12} + x^{11} + x^{10} + x^9 + x^8 + x^4 + x + 1 \qquad \text{(for 9600 bps)}$$

$$g(x) = x^8 + x^7 + x^4 + x^3 + x + 1 \qquad \text{(for 4800 bps)}$$

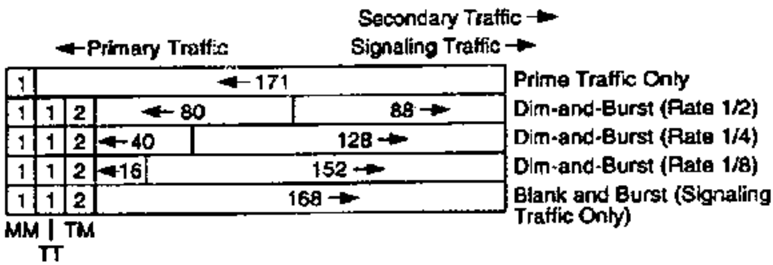
Reverse traffic channel preamble. Used to aid the BS in performing initial acquisition of the reverse traffic channel. The preamble shall consist of frames of 192 zeros at the 9600-bps rate.

Null reverse traffic channel. Used when no service option is active. It is a keep-alive operation. The null traffic channel data shall consist of a frame of 16 ones followed by 8 zeros at the 1200-bps rate.

Information bits and time reference. The information bits (172 bits) can be used to provide for the transmission of primary traffic and signaling or secondary traffic. Signaling traffic may be transmitted via blank-and-burst with the primary traffic and signaling traffic sharing the frame. Five different information bit structures described in Fig. 15.31 are for the mobile station use.

The time reference will be established at the MS. The time of occurrence of the earliest arriving multipath component is used for demodulation. The time reference from the forward traffic channel is used for the transmit time of the reverse traffic channel. The time reference from the paging channel is used for the transmit time of the access.

Forward CDMA channel signals. The forward CDMA channel consists of the following code channels: the pilot channel, the sync channel, paging channels (1 to 7), and forward traffic channels. They are code channels. Each is orthogonally spread by one of 64 Walsh function codes, and is then spread by a quadrature pair of PN sequences at a fixed chip rate of 1.2288 Mcps. The example of a forward CDMA channel transmitted by a BS is shown in Fig. 15.32. Each traffic channel consists of traffic data and mobile power control subchannels.



MM - mixed mode bit

TT - traffic type bit

TM - traffic mode bits

Figure 15.31 Information bits for primary traffic and secondary traffic.

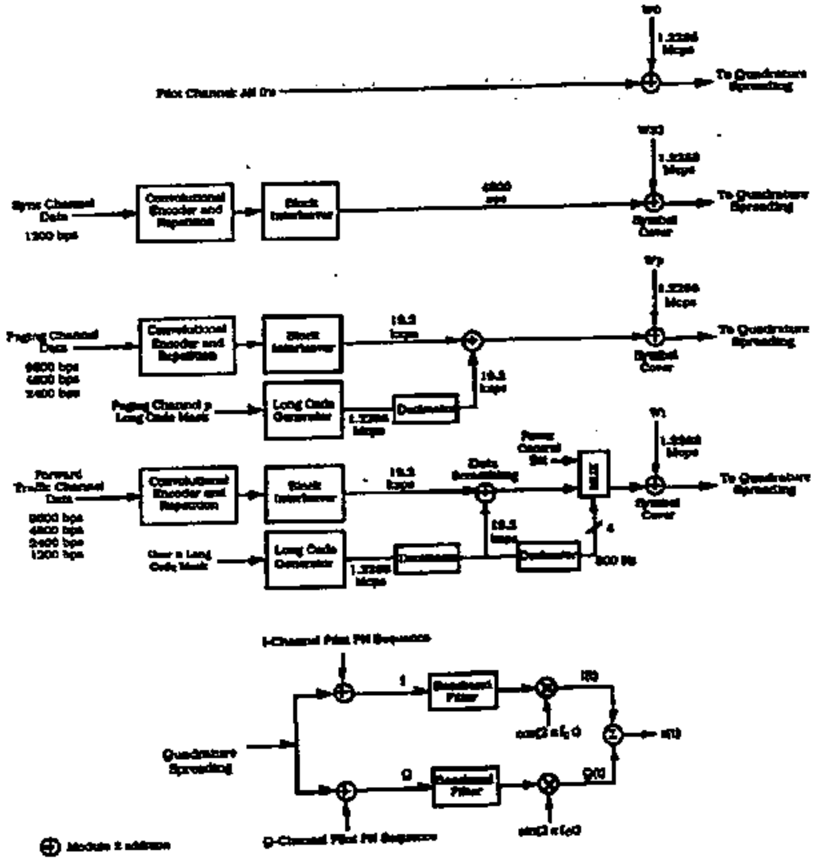


Figure 15.32 Forward CDMA channel structure. (*a*) Modulation; (*b*) quadrature spreading.

Forward CDMA channel structure. The structures of pilot channels, sync channel, paging channel, and forward traffic channel data are shown in Fig. 15.32. There are two parts, modulation and quadrature spreading. In the modulation part, the data rate is at the input.

Data rates at the input

1. The pilot channel sends all 0s at 19.2 kbps rate.
2. The sync channel operates at a fixed rate of 1200 bps.
3. The paging channel supports the fixed data rate at 9600, 4800, or 2400 bps.
4. The forward traffic channel supports variable data rate operation at 9600, 4800, 2400, or 1200 bps.

Modulation. The modulation of the pilot channel has not used the error-correction prior to transmission. The channel takes each bit and spreads it into a 64-bit Walsh code. The data rate of 19.2 kbps becomes 1.2288 Mcps. The modulation parameters of sync channel, paging channel, and forward traffic channel are listed in Tables 15.7, 15.8, and 15.9, respectively. The sync channel, paging channel, and forward traffic channel are encoded prior to transmission. The rate of convolutional code is 1/2 with constraint length of 9 (9 registers).

Code symbol repetition. For paging and forward traffic channels, repetition depends on the data rate of each channel. A low data rate needs more repeats in order to make up the modulation symbol rate of 19.2 kbps.

For a sync channel, each encoded symbol is repeated two times and the modulation symbol rate is 4800 sps. The 4800-sps data are modulated with Walsh function code W32 which has been multiplied by 4. In other words, each symbol becomes $4 \times 64 = 256$ cps.

Block interleaving. The purpose of using block interleaving is to try to avoid burst errors while sending the data through a multipath fading environment. The input of the sync channel interleaver is shown in Table 15.10 and the output is shown in Table 15.11, where the arrows in each column show the sequential order of the data flow. For the forward traffic channel and paging channel, the input of the interleaver is shown in Table 15.12 and the output of the interleaver is shown in Table 15.13.

Data scrambling. Data scrambling shall be accomplished by performing modulo-2 addition of the interleaver output symbol with the binary value of the long-code PN chip ($2^{42} - 1$); the long-code mask is for privacy. Also, the long-code data rate after passing through two decimators is reduced to 800 Hz, which is used for multiplexer (MUX) timing control. The circuit is shown in Fig. 15.33.

TABLE 15.7 Sync Channel Modulation Parameters

Parameter	Data rate, bps 1200	Units
PN chip rate	1.2288	Mcps
Code rate	½	bits/code symbol
Code repetition	2	mod. symbol/code symbol*
Modulation symbol rate	4800	sps
PN chips/modulation symbol	256	PN chips/mod. symbol
PN chips/bit	1024	PN chips/bit

*Each repetition of a code symbol is a modulation symbol.

TABLE 15.8 Paging Channel Modulation Parameters

Parameter	Data rate, bps			Units
	9600	4800	2400	
PN chip rate	1.2288	1.2288	1.2288	Mcps
Code rate	½	½	½	bits/code symbol
Code repetition	1	2	4	mod. symbol/ code symbol*
Modulation symbol rate	19,200	19,200	19,200	sps
PN chips/modulation symbol	64	64	64	PN chips/mod. symbol
PN chips/bit	128	256	512	PN chips/bit

*Each repetition of a code symbol is a modulation symbol.

Power control subchannel. At the rate of one bit every 1.25 ms (i.e., 800 bps), a 0 bit indicator is sent to the MS to increase the mean output power level or a 1 bit is sent to decrease it. There are 16 possible starting positions. Each position corresponds to one of the first 16 modulation symbols. Figure 15.34 indicates the randomization of power control bit positions. The reverse traffic channel sends a bit with 6 Walsh symbols in 1.25 ms. The base station measures signal strength, converts the measured signal strength to a power control bit, and transmit with a 4-bit binary number (levels 0 to 15) by scrambling bits 23, 22, 21, and 20. In Fig. 15.34, the value of bits 23, 22, 21, and 20 is 1011 binary (11 decimal). The power control bit starting position is the eleventh position within 1.25 ms of the seventh slot.

Orthogonal spreading. In the forward channel, each code channel transmits one of 64 Walsh functions at a fixed chip rate of 1.2288 Mcps to provide orthogonal channelization among all code channels on a given forward CDMA channel.

TABLE 15.9 Forward Traffic Channel Modulation Parameters

Parameter	Data rate, bps				Units
	9600	4800	2400	1200	
PN chip rate	1.2288	1.2288	1.2288	1.2288	Mcps
Code rate	½	½	½	½	bits/code symbol
Code repetition	1	2	4	8	mod symbol/code symbol*
Modulation symbol rate	19,200	19,200	19,200	19,200	sps
PN chips/modulation symbol	64	64	64	64	PN chips/mod. symbol
PN chips/bit	128	256	512	1024	PN chips/bit

*Each repetition of a code symbol is a modulation symbol.

TABLE 15.10 Sync Channel Interleaver Input (Array Write Operation)

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

TABLE 15.11 Sync Channel Interleaver Output (Array Read Operation)

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

TABLE 15.12 Forward Traffic and Paging Channel Interleaver Input (Array Write Operation at 9600 bps)

1	25	49	73	97	121	145	169	193	217	241	265	289	313	337	361
2	26	50	74	98	122	146	170	194	218	242	266	290	314	338	362
3	27	51	75	99	123	147	171	195	219	243	267	291	315	339	363
4	28	52	76	100	124	148	172	196	220	244	268	292	316	340	364
5	29	53	77	101	125	149	173	197	221	245	269	293	317	341	365
6	30	54	78	102	126	150	174	198	222	246	270	294	318	342	366
7	31	55	79	103	127	151	175	199	223	247	271	295	319	343	367
8	32	56	80	104	128	152	176	200	224	248	272	296	320	344	368
9	33	57	81	105	129	153	177	201	225	249	273	297	321	345	369
10	34	58	82	106	130	154	178	202	226	250	274	298	322	346	370
11	35	59	83	107	131	155	179	203	227	251	275	299	323	347	371
12	36	60	84	108	132	156	180	204	228	252	276	300	324	348	372
13	37	61	85	109	133	157	181	205	229	253	277	301	325	349	373
14	38	62	86	110	134	158	182	206	230	254	278	302	326	350	374
15	39	63	87	111	135	159	183	207	231	255	279	303	327	351	375
16	40	64	88	112	136	160	184	208	232	256	280	304	328	352	376
17	41	65	89	113	137	161	185	209	233	257	281	305	329	353	377
18	42	66	90	114	138	162	186	210	234	258	282	306	330	354	378
19	43	67	91	115	139	163	187	211	235	259	283	307	331	355	379
20	44	68	92	116	140	164	188	212	236	260	284	308	332	356	380
21	45	69	93	117	141	165	189	213	237	261	285	309	333	357	381
22	46	70	94	118	142	166	190	214	238	262	286	310	334	358	382
23	47	71	95	119	143	167	191	215	239	263	287	311	335	359	383
24	48	72	96	120	144	168	192	216	240	264	288	312	336	360	384

TABLE 15.13 Forward Traffic and Paging Channel Interleaver Output (Array Read Operation at 9600 bps)

1	9	5	13	3	11	7	15	2	10	6	14	4	12	8	16
65	73	69	77	67	75	71	79	66	74	70	78	68	76	72	80
129	137	133	141	131	139	135	143	130	138	134	142	132	140	136	144
193	201	197	205	195	203	199	207	194	202	198	206	196	204	200	208
257	265	261	269	259	267	263	271	258	266	262	270	260	268	264	272
321	329	325	333	323	331	327	335	322	330	326	334	324	332	328	336
33	41	37	45	35	43	39	47	34	42	38	46	36	44	40	48
97	105	101	109	99	107	103	111	98	106	102	110	100	108	104	112
161	169	165	173	163	171	167	175	162	170	166	174	164	172	168	176
225	233	229	237	227	235	231	239	226	234	230	238	228	236	232	240
289	297	293	301	291	299	295	303	290	298	294	302	292	300	296	304
353	361	357	365	355	363	359	367	354	362	358	366	356	364	360	368
17	25	21	29	19	27	23	31	18	26	22	30	20	28	24	32
81	89	85	93	83	91	87	95	82	90	86	94	84	92	88	96
145	153	149	157	147	155	151	159	146	154	150	158	148	156	152	160
209	217	213	221	211	219	215	223	210	218	214	222	212	220	216	224
273	281	277	285	275	283	279	287	274	282	278	286	276	284	280	288
337	345	341	349	339	347	343	351	338	346	342	350	340	348	344	352
49	57	53	61	51	59	55	63	50	58	54	62	52	60	56	64
113	121	117	125	115	123	119	127	114	122	118	126	116	124	120	128
177	185	181	189	179	187	183	191	178	186	182	190	180	188	184	192
241	249	245	253	243	251	247	255	242	250	246	254	244	252	248	256
305	313	309	317	307	315	311	319	306	314	310	318	308	316	312	320
369	377	373	381	371	379	375	383	370	378	374	382	372	380	376	384

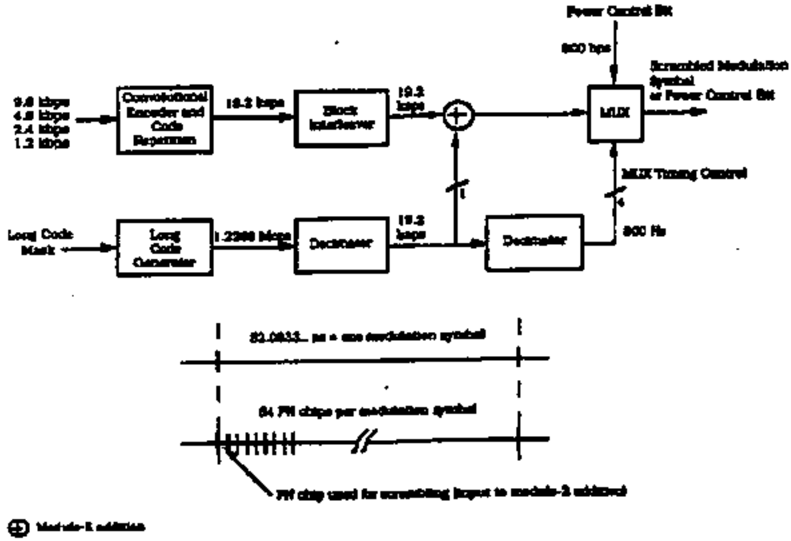


Figure 15.33 Data scrambler function and timing.

PN sequence offset

Pilot channel. A pilot channel is transmitted all times on Walsh function W0 by the base station. Pilot PN sequence offset is used for identifying each base station. Time offset may be revised within a CDMA cellular system.

Sync channel. The sync channel is an encoded, interleaved, spread, and modulated spread signal. The sync channel uses the same pilot PN sequence offset as the pilot channel for a given base station.

Receiver at MS. The MS demodulation process shall perform complimentary operations to the BS modulation process. The MS shall provide a minimum of four processing elements. Three of them are capable of tracking and demodulating multipath components of the forward CDM channel. At least one element shall be a searcher element capable of scanning and estimating the signal strength at each pilot PN sequence offset. The signal strength of the pilot is used to select the desired BS during the idle or initialization stage. Also, the signal strength of the pilot is used for the MS to determine when the handoff shall be requested and which new BS is the candidate. The information on handoff will be sent to the BS via the reverse signaling traffic channel (see Table 15.14). The multiplex option is the same on both the forward traffic channel and the reverse traffic channel.

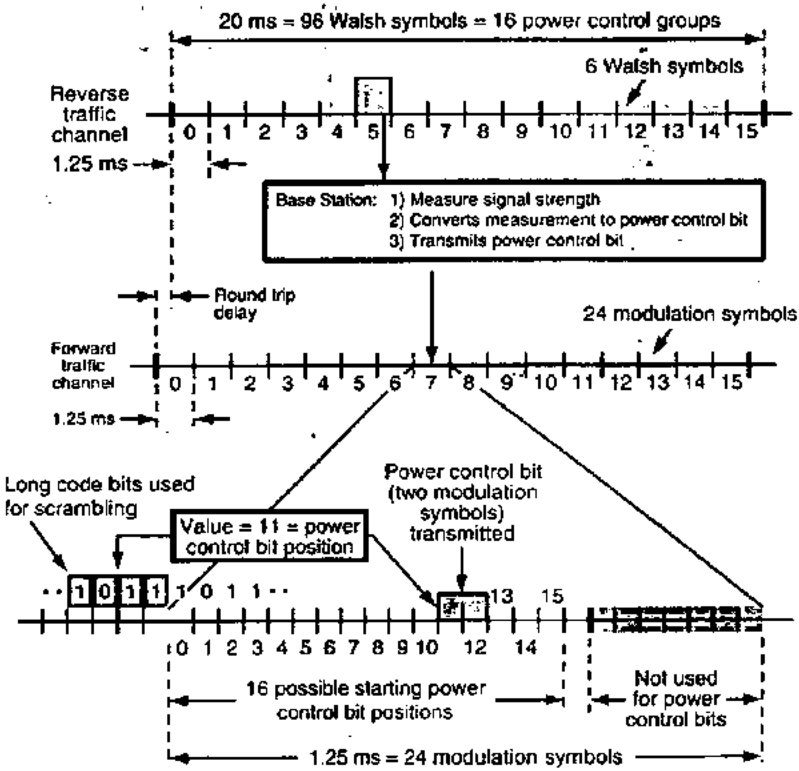


Figure 15.34 Randomization of power control bit positions.

Forward traffic channel frames. There are 14 categories for multiplex option 1. Among those categories, 12 are listed in Table 15.14 and are considered good frames. Categories 9 and 10 are bad frames:

Category 9: 9600-bps frame, primary traffic only, with bit errors

Category 10: Insufficient frame quality

15.3.4 Authentication, encryption, and privacy

Authentication refers to the process by which the base station confirms the identity of the mobile station, i.e., the identical sets of shared secret data. SSD is a 128-bit pattern in the MS. SSD-A consists of 64 bits and SSD-B consists of 64 bits. SSD-A supports the authentication procedure initialized with mobile station specific information, random

TABLE 15.14 Reverse and Forward Traffic Channel Information Bits for Multiplex Option 1

	Format bits			Primary traffic	Signaling traffic	Secondary traffic	Categories of received traffic channel frame
Transmit rate (bits/sec)	Mixed mode (MM)	Traffic type (TT)	Traffic mode (TM)	bits/frame	bits/frame	bits/frame	
9600	'0'	-	-	171	0	0	1
	'1'	'0'	'00'	80	88	0	2
	'1'	'0'	'01'	40	128	0	3
	'1'	'0'	'10'	16	152	0	4
	'1'	'0'	'11'	0	168	0	5
	'1'	'1'	'00'	80	0	88	11
	'1'	'1'	'01'	40	0	128	12
	'1'	'1'	'10'	16	0	152	13
	'1'	'1'	'11'	0	0	168	14
4800	-	-	-	80	0	0	6
2400	-	-	-	40	0	0	7
1200	-	-	-	16	0	0	8

data and the mobile station's A key (64 bits long). A key may be also called PIN. SSD-B supports CDMA voice privacy and message confidentiality.

The purposes for using authentication are (1) MS registration, (2) MS origination, and (3) MS termination. When the information element AUTH in the system parameters overhead message is set to 1 and the MS attempts to register, originate, or terminate, then the auth-signature procedure is executed, and AUTHR is obtained (see Fig. 15.35) and sent with RANDC (eight most significant bits of RAND confirmation) and COUNT to the base station for validation.

Authentication of MS data bursts

1. The BS sends an SSD update message on either the paging channel or the forward traffic channel. In the SSD update message, there is a RANDSSD field which is used for the computation of SSD at the home location register/authentication center (HLR/AUC). (The A key is stored at the MS and the HLR/AUC.) The MS shall then execute the SSD generation procedure by using RANDSSD, ESN, and A key to produce SSD-A-New and SSD-B-New.

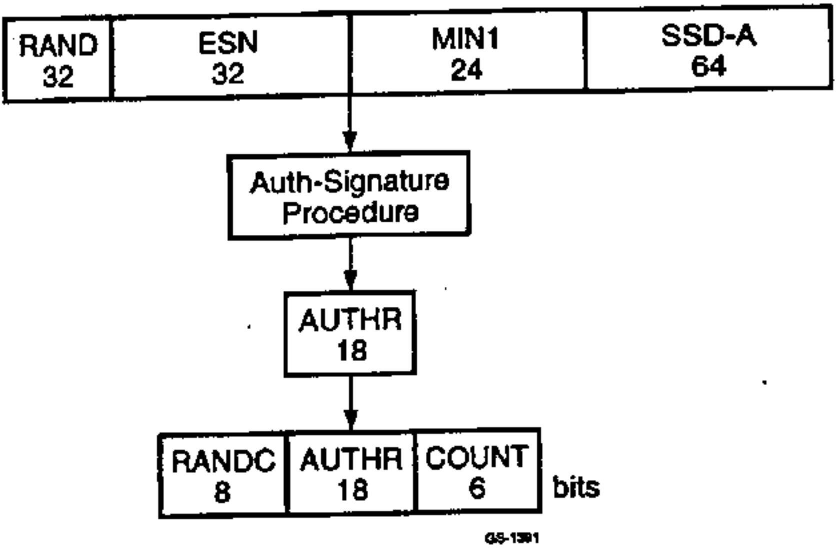


Figure 15.35 Authentication parameters for base station validation.

2. The MS shall select a 32-bit random number RANDBS to BS in a base station challenge order via the access channel or reverse traffic channel.

3. Both BS and MS shall execute an auth-signature procedure by using SSD-A-New and RANDBS, and both obtain an 18-bit AUTHBS.

4. The BS sends the AUTHBS in the base station challenge order confirmation on the paging channel or the forward traffic channel.

5. The MS compares the two AUTHBS, one from its own MS and one from the BS. If the comparison is successful, it sets SSD-A and SSD-B to SSD-A-New and SSD-B-New, respectively. Also, the MS shall send an SSD update confirmation order to the BS indicating the successful comparison. If the comparison is not successful, it discards the two new SSDs and sends an SSD update rejection order to the BS, indicating unsuccessful comparison.

On receipt of the SSD update confirmation order, the BS sets SSD-A and SSD-B to the values received from the HLR/AUC. The SSD update message flow is shown in Fig. 15.36.

Signaling message encryption. In an effort to protect sensitive subscriber information (such as PIN), the availability of encryption algorithm information is governed under the U.S. International Traffic and Arms Regulation (ITAR) and export administration regulations.

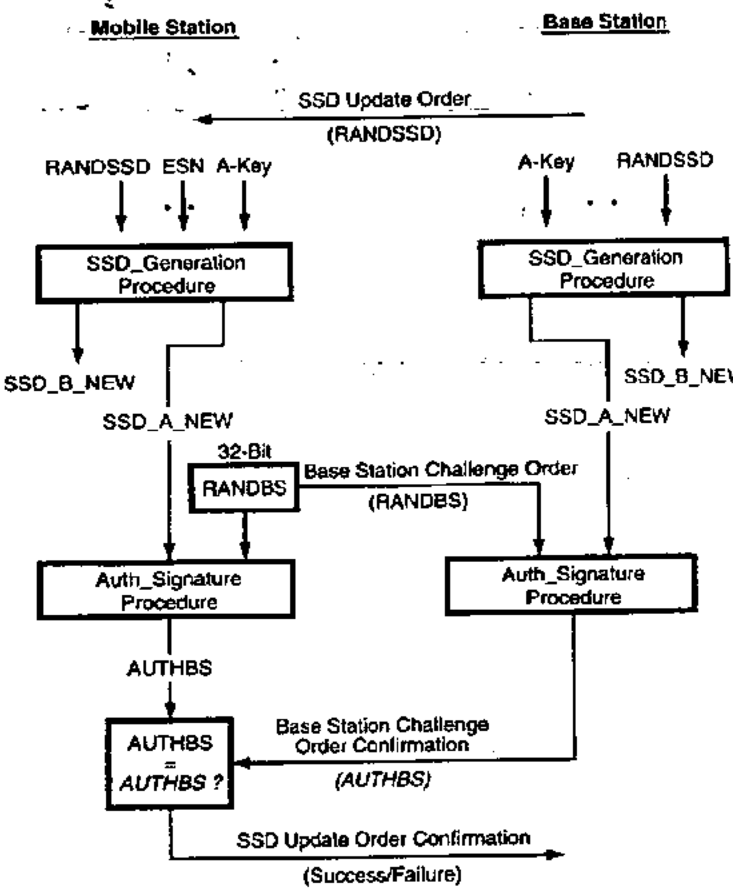


Figure 15.36 SSD update message flow.

Messages shall not be encrypted if authentication is not performed. Signaling message encryption is controlled for each call individually.

Voice privacy. Voice privacy is provided in the CDMA system by means of the private long-code mask used for PN spreading. Voice privacy control is provided on the traffic channels only. All calls are initiated by using the public long-code mask for PN spreading. To initiate a transition to the private or public long-code mask, either the BS or the MS sends a long-code transition request order on the traffic channel.

15.3.5 Malfunction detection

The BS detects the malfunction of an MS by asking the MS to respond to the lock order, lock until power-cycled order, and maintenance required order. This feature identifies the malfunctional MS and prevents the MS from contaminating the CDMA system by sending a signal to disconnect the MS transmit power.

15.3.6 Call processing

MS call processing consists of the following states:

MS initialization state

- The MS selects which system to use.
- It acquires the pilot channel of a CDMA system within 20 ms.
- It obtains system configuration and timing information for a CDMA system.
- It synchronizes its timing to that of a CDMA system.

MS idle state

- The MS shall perform paging channel monitoring procedures. The paging channel is divided into 200-ms slots called *paging channel slots*. Paging and control messages for an MS operating in the nonslotted mode can be received in an array of the paging channel slots. Therefore, the nonslotted mode of operation requires the MS to monitor all slots. An MS operating in the slotted mode generally monitors the paging channel for one or two slots per slot cycle. The MS can control the length of the slot cycle.
- Unless otherwise specified in the requirements for processing a specific message, the MS shall transmit an acknowledgement in response to any message received that is addressed to the MS.
- The MS shall maintain all active registration timers.

  The CDMA system supports nine different forms of registration:
  Autonomous registrations:

1. *Power-up registration.* The mobile station registers when it powers on, switches from using the alternate serving system, or switches from using the analog system.
2. *Power-down registration.* The mobile station registers when it powers off if previously registered in the current serving system.

3. *Timer-based registration.* The mobile station registers when a timer expires.
4. *Distance-based registration.* The mobile station registers when the distance between the current base station and the base station in which it last registered exceeds a threshold.
5. *Zone-based registration.* The mobile station registers when it enters a new zone.

Registrations under different requests:

6. *Parameter-change registration.* The mobile station registers when certain of its stored parameters change.
7. *Ordered registration.* The mobile station registers when the base station requests it.
8. *Implicit registration.* When a mobile station successfully sends an origination message or page response message, the base station can infer the mobile station's location, causing an implicit registration.
9. *Traffic channel registration.* Whenever the base station has registration information for a mobile station that has been assigned to a traffic channel, the base station can notify the mobile station that it is registered.

System access state. The MS sends messages to the BS on the access channel and receives messages from the base station on the paging channel. The entire process of sending one message and receiving acknowledgment for that message is called an *access attempt.* Each transmission in the access attempt is called an *access probe.* The mobile station transmits the same message in each access probe in an access attempt. Each access probe consists of an access channel preamble and an access channel message capsule. There are two types of messages sent on the access channel: a response message and a request message. The access attempt ends after an acknowledgment is received.

MS control on the traffic channel state. The mobile station communicates with the BS using the forward and reverse traffic channels. There are five functions:

1. The MS verifies that it can receive the forward traffic channel and begins transmitting on the reverse traffic channel.
2. The MS waits for an order on an alert with information message.
3. The MS waits for the user to answer the call.

4. The MS's primary service option application exchanges primary traffic packets with the base station.
5. The MS disconnects the call.

15.3.7 Handoff procedures

Types of handoffs. The MS supports four handoff procedures:

1. *Soft handoff.* The MS commences communication with a new base station without interrupting communication with the old base station. Soft handoff means an identical frequency assignment between the old BS and new BS. Soft handoff provides different-site selection diversity to enhance the signal.
2. *CDMA-to-CDMA hard handoff.* The MS transmits between two base stations with different frequency assignments.
3. *CDMA-to-analog handoff.* The MS is directed from a forward traffic channel to an analog voice channel with a different frequency assignment.
4. *Softer handoff.* Handoffs between sectors within a cell.

Pilot sets. The information obtained from the pilot channel is used for the handoff. A pilot is associated with the forward traffic channels in the same forward CDMA channel. A pilot channel is identified by a pilot sequence offset. Each pilot channel is assigned to a particular BS. The MS can obtain four sets of pilot channels:

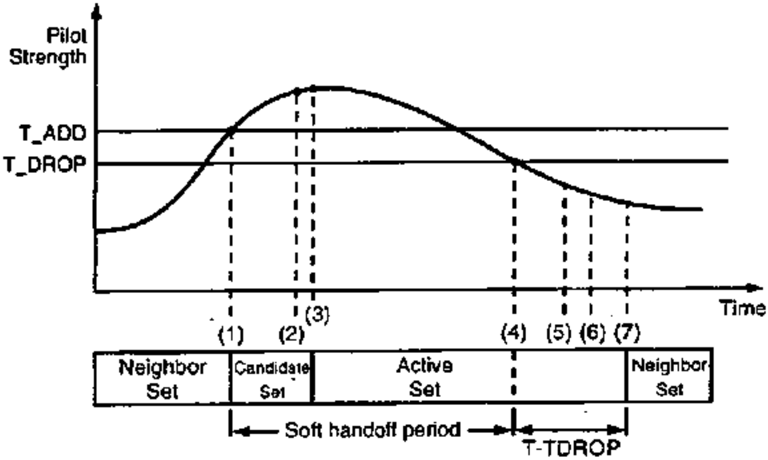
1. *Active set.* The pilot associated with the forward traffic channels assigned to the MS.
2. *Candidate set.* The pilots that are not in the active set but are received by MS with sufficient strength.
3. *Neighbor set.* The pilots that are not in the active set or the candidate set and are likely candidates for handoff.
4. *Remaining set.* The set in the current system on the current CDMA frequency assignment, excluding the above three sets.

Pilot requirement

1. The base station specified for each of the above pilots sets the search window in which the mobile station is to search for usable multipath components of the pilots in the set.
2. The MS assists the base station in the handoff process by measuring and reporting the strengths of received pilots.

3. A handoff drop timer shall be maintained for each pilot in the active set and candidate set. When the signal strength level is below TDROP (also called T-DROP), T-TDROP is set to zero, i.e., it expires within 100 ms. There are 15 T-TDROP values. The highest value of T-TDROP is 319 s. When the MS receives a signal strength level from the neighboring cell exceeding a given TADD (also called T_ADD) level in decibels, the soft handoff starts. When the MS receives a signal strength level from the home cell below TDROP, the soft handoff ends. The handoff action would take place after the received level from the home cell is below the TDROP. If the time between TADD and TDROP is very short, the T-TDROP time has to be longer. Also, in certain circumstances, it is preferable to reduce the call drops and sacrifice voice quality.
4. The MS shall measure the arrival time for each pilot reported to the base station. The time of the earliest arriving usable multipath component of the pilot is used to measure relative to the MS's time reference in units of PN chips.
5. Soft handoff
   a. All forward traffic channels associated with pilots in the active set of the MS carry modulation symbols identical to those of the power control subchannel. When the active set contains more than one pilot, the MS should provide diversity combining of the associated forward traffic channels. The MS shall provide for differential propagation delays from zero to at least 150 μs.
   b. For reverse traffic channel power control during soft handoff, the handoff direction message identifies sets of forward traffic channels that carry identical closed-loop power control subchannels. A set consists of one or more forward traffic channel transmissions with identical power control information. The MS will obtain at most one power control bit from each set of identical closed-loop power control subchannels. If the power control bits obtained from all sets are equal to 0, the MS shall increase its power; if they are equal to 1, the MS shall decrease its power.
   c. The typical message exchanges between the MS and the BS during handoff are shown in Fig. 15.37. There are seven messages during the soft handoffs. The first message that the MS sends is a pilot strength measurement message when the neighboring pilot strength exceeds TADD. The soft handoff starts. The seventh message is that the MS should move pilot from the active set to the neighboring set and send a handoff completion message. The soft handoff is then completed.

The pilot strength measurement triggered by a candidate pilot is shown in Fig. 15.38. During the soft handoff starts, the two pilots P_1



(1) Pilot strength exceeds T_ADD. Mobile station sends a *pilot strength measurement message* and transfers pilot to the candidate set.

(2) Base station sends a *handoff direction message.*

(3) Mobile station transfers pilot to the active set and sends a *Handoff completion message.*

(4) Pilot strength drops below T_DROP. Mobile station starts the handoff drop timer.

(5) Handoff drop timer expires. Mobile station sends a *pilot strength measurement message.*

(6) Base station sends a *handoff direction message.*

(7) Mobile station moves pilot from the active set to the neighbor set and sends a *handoff completion message.*

Figure 15.37 Handoff threshold example.

and P_2 are indicated in the active set. There is a P_0 in the candidate set that is stronger than a pilot in the active set only if the difference between their respective strengths is at least T-Comp (level in decibels) as shown in Fig. 15.38.

15.4 Miscellaneous Mobile Systems

In the previous sections, GSM, NA-TDMA, and CDMA were introduced. This setion will briefly introduce other systems such as PHP, PDC, CT-2, DECT, CDPD, PCN, and PCS.

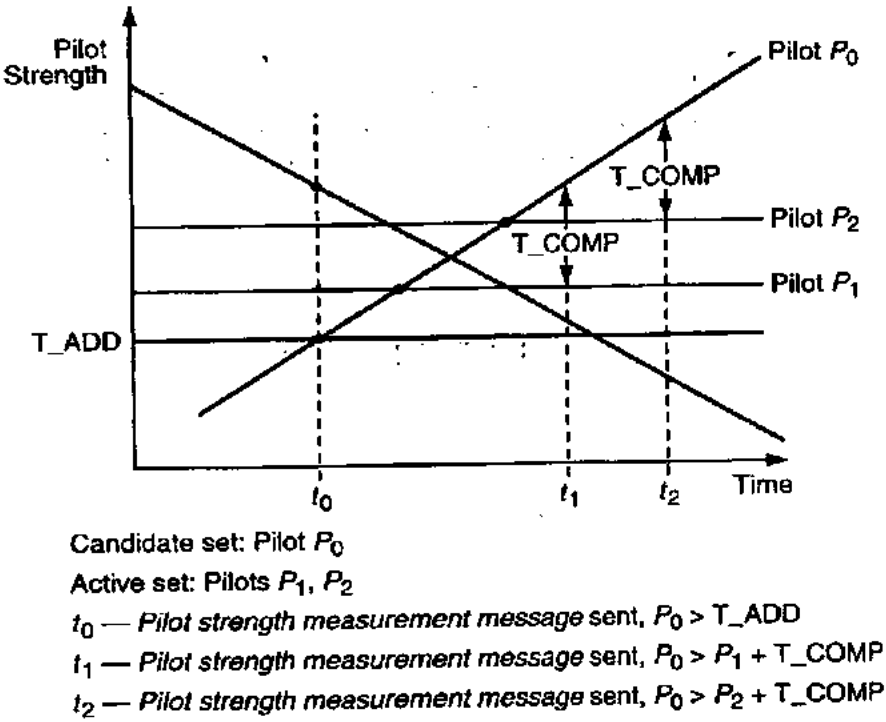


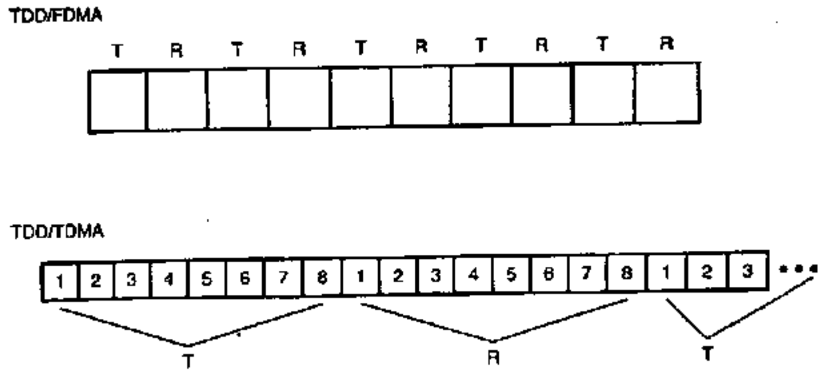
Figure 15.38 Pilot strength measurements triggered by a candidate pilot.

15.4.1 TDD Systems

Time-division duplexing (TDD) systems are digital systems and use only one carrier to transmit and receive information. There are two kinds of TDD systems (see Fig. 15.39):

- TDD/FDMA—each carrier serves only one user.
- TDD/TDMA—each carrier can have many time slots and each slot can serve one user. Then N transmit time slots can serve N users.

A TDD system is used when only one chunk of spectrum is allocated. In cellular systems, there are two chunks of spectrum, separated by 20 MHz. In each cellular channel, the base transmit frequency and the mobile transmit frequency are 45 MHz apart. Therefore, the separation in frequency between transmitting and receiving is adequate to avoid interference. In TDD there is no separation in frequency between transmitting and receiving, but a separation in time interval.



■ **Remarks**

- All cell sites have to be synchronized in order to eliminate the near-far interference from nieghboring sites
- The guard time in TDD slot would be longer than in regular TDMA slots
- The diversity scheme can be applied at one end to serve both ends
- Do not increase spectrum efficiency from a traffic/capacity point of view

Figure 15.39 Time-division duplexing.

The advantages of TDD are as follows:

1. When only one chunk of spectrum is available, TDD is the best utilization of spectrum.
2. Diversity can be applied at one end (terminal) to serve both ends, since the fading characteristics of one carrier are the same when received at both ends. At the base station, the information on selecting antennas for the space-diversity selective combining receiver can be used to switch to one of two transmitting antennas. Thus the mobile unit (or portable unit) can achieve the same diversity gain with a nondiversity receiver.

One of the concerns is that the TDD system has to be a synchronized system with a master clock. Otherwise, when one BS transmits and another BS receives, equivalent cochannel (co-time slot) interference occurs. Another concern is that the signal structures of TDD and of a frequency duplexing division (FDD) are different. Therefore, the two systems should not coexist in the same area because of their mutual interference.

In this section, several TDD systems will be briefly described.

Personal handy phone (PHP) system[14]. PHP is a wireless communication TDD system which supports personal communication services (PCS). It uses small, low-complexity, lightweight terminals called personal stations (PSs). PHP can be used for public telepoint (a portable phone booth), wireless PBX, home cordless telephone, and walkie-talkie (PS-to-PS) communication.

PHP features wider coverage per cell; operation in a mobile outdoor environment; faster and distributed control of handoff; enhanced authentication, encryption, and privacy; and circuit- and packet-oriented data services. PHP system is also called PHS (personal handy phone system).

Transmission parameters

- Full duplex system
- Voice coder—32 kbps adaptive differential pulse-code modulation (ADPCM).
- Duplexing—TDD. The portable and base units transmit and receive on the same frequency but different time slots. The slot arrangement is shown in Fig. 15.40.
- Multiple access—TDMA-TDD, up to four multiplexed circuits.
- Modulation—$\pi/4$ DQPSK, roll-off rate = 0.5.
- Data rate—192 ksps (or 384 kbps).
- Spectrum allocation—1895 to 1918.1 MHz. This spectrum has been allocated for private and public use.

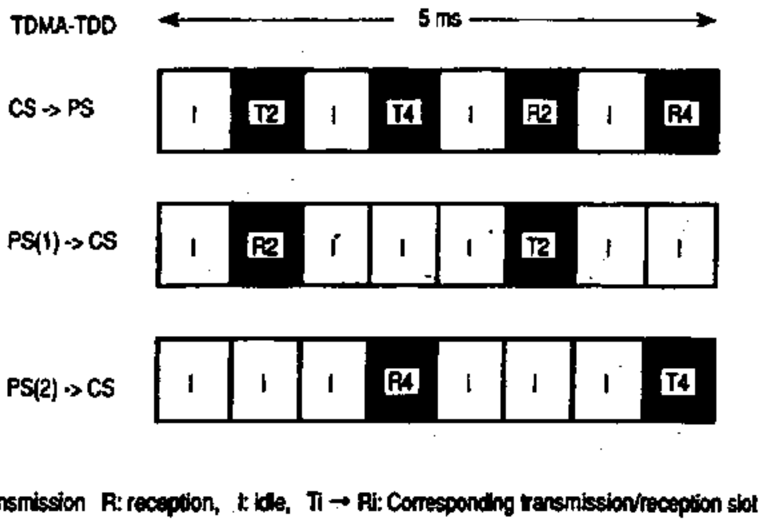


Figure 15.40 Slot arrangement (corresponding to 32 kbps).

- Carrier frequency spacing—300 kHz.
- Carrier frequency—1895.15 MHz or 1895.15 + N·300 kHz where N is an integer.

Function channel structure The control channel consists of the following logical channels:

1. *Broadcast control channel*. BCCH is a one-way downlink channel for broadcasting control information from CS to PS.
2. *Common control channel*. CCCH sends out the control information for call connection.
   a. *Paging channel*. PCH is a one-way downlink channel.
   b. *Signaling control channel*. SCCH is a bidirectional point-to-point channel.
3. *User packet channel*. UPCH is a bidirectional point-multipoint channel that sends control signal information and user packet data.
4. Associated control channel. ACCH is a bidirectional channel that is associated with the TCH. It carries out control information and user packet data. There is a SACCH and a FACCH.

The traffic channel is a point-to-point bidirectional channel and is used for transmitting user information.

Carrier Structure

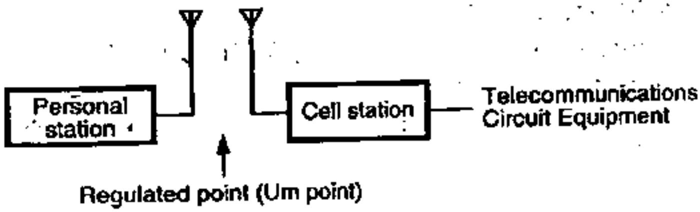
Control carrier. A carrier in which only common usage slots can be assigned to study intermittent transmission in a cordless station (CS).

Communications carrier. A carrier in which the user can perform communication through the individual assigned slot. It also can allocate common usage slots for a communications carrier.

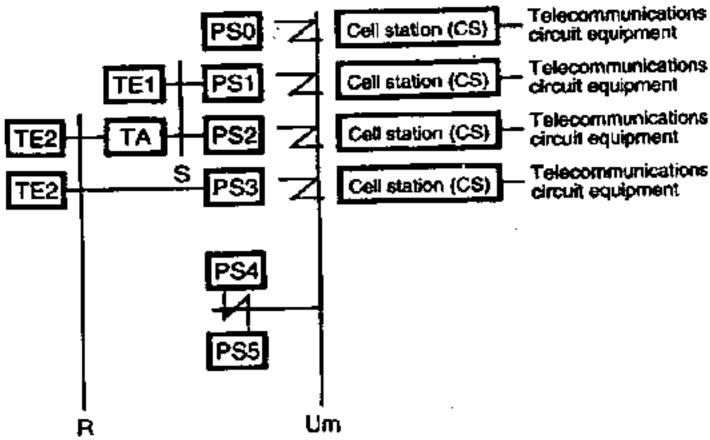
Carrier for direct communication between personal stations. A carrier providing direct communication without going through a CS. The connection control and conversation can be carried out on the same slot.

Structure and interfaces of PHP system. The structure of PHP is shown in Fig. 15.41*a*. The CS is connected to the telecommunications circuit equipment. The interfaces of PHP are shown in Fig. 15.41*b*. There are three interface points:

Um is the interface point between personal station and cell station or between personal station and personal station. Conforms to the standard.



(a)



(b)

Figure 15.41 System structure and interface of PHP. (*a*) Structure of personal handy phone system; (*b*) interface points.

> R is the interface point between I interface nonconforming terminal and mobile terminal equipment or terminal adapter. Outside scope of the standard.
>
> S is the interface point between I interface conforming terminal or terminal adapter TA and mobile terminal equipment TE.

The five different classes of PS in Fig. 15.41 are defined as follows:

PSO, PS4, PS5	Personal station, including integrated man/machine interface of terminals, etc.
PS1, PS2	Personal station with I interface.
PS3	Personal station without I interface.

Terminal equipment TE1 is with I interface, and TE2 is without I interface. TA is the interface conversion equipment for non–I interface and I interface.

Cordless phone 2 (CT-2).[15] CT-2 was developed by GPT Ltd. in the United Kingdom and was the first TDD system for mobile radio communications. All the other TDD systems such as PHP and DECT adopted CT-2's structure. CT-2 is a portable payphone booth. Calls can be dialed out but not dialed in, and there is no handoff.

System structure. The structure is the same as PHP in Fig. 15.41.

- The CS is called the telepoint or phone zone
- Carrier frequency—864.1 to 868.1 MHz
- Total spectrum—4 MHz
- Channel access—FDMA/TDD
- Number of channels—40
- Channel bandwidth—100-kHz spacing
- Handset output power—1 to 10 mW

Characteristics

- Overall data rate—72 kbps
- Data rate per speech channel—32 kbps
- Speech coding—ADPCM
- Modulation—GMSK

Cordless phone 3 (CT-3). Originally called DCT900, CT-3 was developed by Ericsson as an upgrade from the CT-2 version.

System structure

- Time slots—64
- Two-way call system—call send and call delivery
- Slow-speed handoff
- Caller authentication
- Encryption
- Roaming

Characteristics

- Handset effective radiated power (ERP)—80 mW
- Modulation—filtered minimum-shift keying (MSK)

- Spectrum range—8 MHz
- Frequency—800 to 1000 MHz
- Bandwidth—1 MHz
- RF output power—80 mW peak, 5 mW average
- Number of slots per frame—8
- Overall data rate—640 kbps
- Data rate per speech code—32 kbps
- Speech coding—ADPCM

Digital European cordless telecommunication system (DECT).[16,17] DECT is a European standard system. It is a CT-2–like system, and the applications are slightly different from the cellular system.

Applications

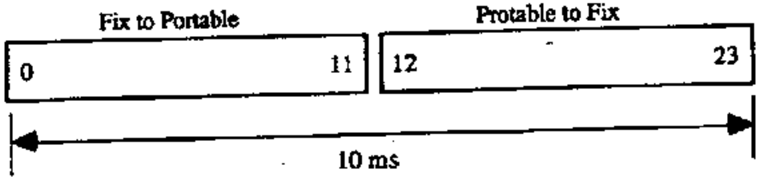
1. Use the public network to mobile communications within the home and the immediate vicinity.
2. Provide business communications locally. In this case, the PSTN has been replaced by the PBX. The base station has been renamed the radio fixed port (RFP), also called a *cluster controller*. The RFP can be treated as the microcell site.
3. Provide mobile public access. It is a PCS application. The system monitors the location of the active handsets and provides call delivery capability, like a cellular system.
4. Local loop. Using DECT to provide wireless local loops is a cost-effective alternative to running copper wires to residential premises.

System structure

- Duplex method—TDD
- Access method—TDMA
- RF power of handset—10 mW
- Channel bandwidth—1.728 MHz/channel
- Number of carriers—5 (a multiple-carrier system)
- Frequency—1800 to 1900 MHz

Characteristics

- Frame—10 ms
- Time slots—12

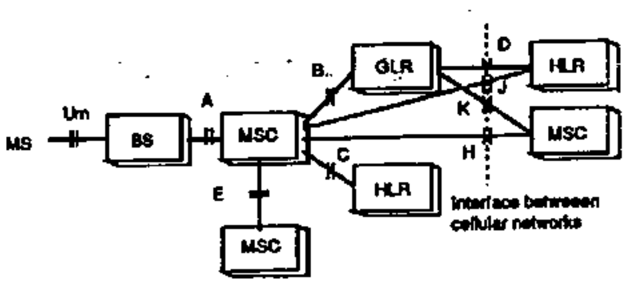


- Bit rate—38.8 kb/slot
- Modulation—GFSK (Gausian FSK)
- Handoff—Yes

15.4.2 Other full-duplexed systems

We have addressed the full-duplexed system such as GSM, NA-TDMA, and CDMA in the previous sections. In this setion PDC, PCN, and PCS will be briefly described.

Personal digital cellular (PDC).[18,19] PDC is a standard system in Japan. The system is a TDMA cellular system operating at 800 MHz and 1.5 GHz which used to be called Japanese digital system (JDC). 1.5 GHz PDC was in service publicly in Osaka in 1994. The PDC network reference model is shown in Fig. 15.42. This system provides nine interfaces among the cellular network. Um is the air interface, which was standardized. Interfaces B, C, D, E, J, K, and H were defined by cellular carriers in Japan. Interface A is an option for operators.



Um: Mobile user-network interface
MS: Mobile Station
BS: Base Station
MSC: Mobile service Switching Center
HLR: Home Location Register
GLR: Gate Location Register

Figure 15.42 PDC network reference model.

Signaling structure. The signaling structure of PDC has three layers (Fig. 15.43). The third layer consists of three functional entities: call control, mobility management, and radio transmission management. In general, the structure of PDC is very similar to that of NA-TDMA.

System structure

- Multiplex access—TDMA
- Number of time slots—3

Characteristics

- 800 MHz

    810—826 MHz
    940—956 MHz

- 1500 MHz

    1429—1441 MHz
    1453—1465 MHz
    1477—1489 MHz
    1561—1513 MHz

- Modulation—$\pi/4$, DQPSK
- Speech coder—VSELP

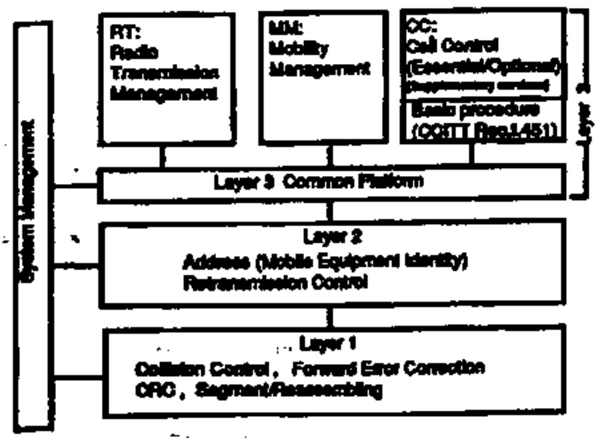


Figure 15.43 Subscriber line signaling structure model.

Personal communication network (PCN).[20] The PCN system was first initiated by Lord Young in 1988. The characteristics of PCN are as follows:

1. Operational frequency—1.7 to 1.88 GHz (1710–1785 MHz and 1805–1880 MHz)
2. Uses 30 GHz or up for microwave backbone system
3. Covers both small cells and large cells (rural areas)
4. Coverage inside and outside buildings
5. Handover
6. Call delivery
7. Portable handset
8. Uses intelligent network

PCN uses the DCS-1800 system, which is similar to GSM, but upconverts the frequency to 1.7–1.88 GHz. Therefore, the network structure, the signaling structure, and the transmission characteristics are similar between PCN and GSM, but the operational frequencies are different.

Personal communication service.[21] See Appendix 15.1 for (2) broadband PCS designated in major trading areas (MTA), (1) MTA in the U.S. and (3) basic trading areas (BTA) in the U.S.

Ever since cellular systems were deployed in 1983 in the United States, the rapidly growing wireless communication systems have faced limitations of spectrum utilization. However, the trend is toward personal communications and includes wireless communications for pedestrians and for in-building communications. In 1989 Lord Young of the United Kingdom was promoting a PCN system operating at 1.8 GHz. PCN uses a GSM version of cellular communication. The spectrum range is 1710–1785 MHz and 1805—1880 MHz.

In 1991 the FCC issued a Notice of Proposed Rule Making, Docket No. 90-134, considering allocating spectrum between 1.85 and 2.2 GHz. The FCC has now allocated 120 MHz of spectrum into seven bands as shown in Fig. 15.44 for wideband PCS and narrowband PCS allocated spectrum.

There are now many systems, such as GSM, NADC, CDMA, PCN, and DECT, as described in this chapter, that are being considered as candidates to be adapted as the future PCS system. How to improve the equalizer for the TDMA[28] and reduce the intererence for CDMA are the major concerns. However, users want to carry one unit with

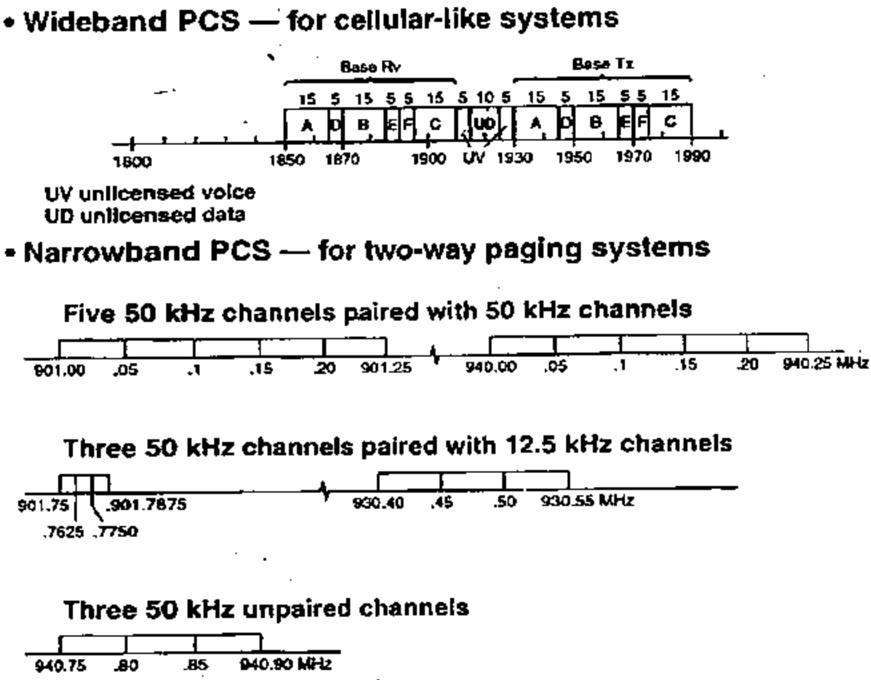


Figure 15.44 FCC's PCS allocation.

which they can place a call to wherever they want and receive calls wherever they may be. Of course, the unit has to be small in size, light in weight, and provide long talking time.

15.4.3 Noncellular systems

There are three systems that can be mentioned in this section: CDPD, MIRS, and satellite mobile systems.

Cellular digital packet data (CDPD) system.[22,27] CDPD is a packet switching system which uses idle voice channels from the cellular system band to carry out traffic. This system can be assigned a dedicated channel or it can hop to idle channels.

In the network reference model of CDPD in Fig. 15.45, acronyms are defined as follows:

SIM	Subscriber identity module
SU	Subscriber unit [also can be indicated as mobile-end station (M-ES)]

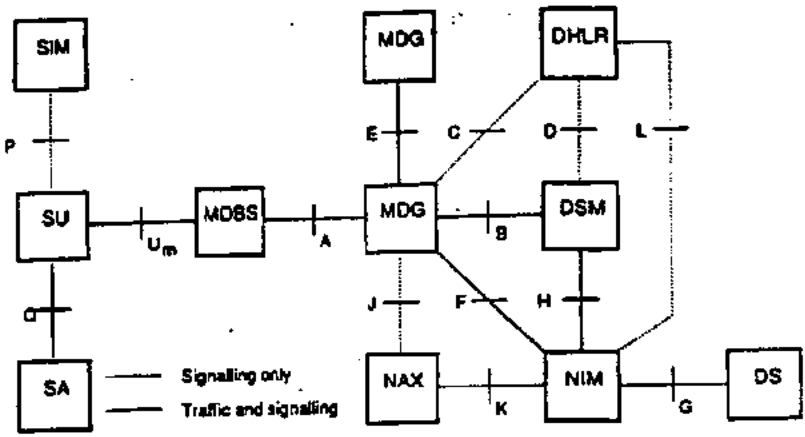


Figure 15.45 CDPD network reference model.

EIR	Equipment identity register
MDBS	Mobile data base station
MDG	Mobile data gateway
NIM	Network interface manager
DHLR	Data home location register
NAX	Network address translator
DSM	Data service manager [also can be indicated as mobile data information station (MD, IS)]
DS	Data service
NG	Network gateway

The network in Fig. 15.45 is self-explanatory. There are two communication interfaces, called *reference points*, identified with a letter (e.g., A or Um). On a given reference point there may be many physical devices to transport the signal such as routes, switches, multiplexers, demultiplexers, and modems.

System operation. The MDBS is collocated at the cellular cell site. DSM is a control center for CDPD. DSM operates independently from the MTSO of cellular systems. In the MDBS, a forward-link logical channel is always on to send the overhead message or send the data to the user. There are two set-ups:

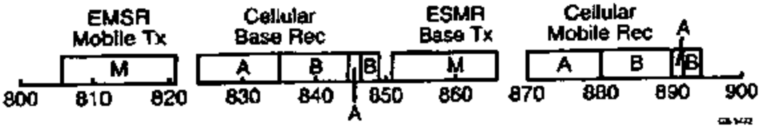
1. *For a dedicated channel setup:* The MS, at the initialization stage, scans the assigned *N* CDPD channels to lock on a strong CDPD channel. The CDPD channel will be changed while the MS is moving from one cell to another cell.

2. *For a frequency-hopping channel setup:* There is a device called a *sniffer* installed in the MDBS. The sniffer monitors the cellular control channels on both the forward and reverse links and chooses an idle cellular channel for CDPD.

Transmission structure

- Roaming support
- Security and authentication across the airlink
- Forward channel block: Reed-Solomon (63,47) data symbols
- Reverse channel block: 8-bit delay maximum plus dotting sequence (38 bits) plus reverse sync work (22 bits); Reed-Solomon (63,47) data symbols
- Frequency-agile with in-band control
- Channel hopping and dedicated channel
- Cell transfer controlled by M-ES (SU in Fig. 15.45).
- Power control
- AMPS-compatible
- Essentially transparent to AMPS
- Modulation—GMSK (same as GSM)
- Data rate—19.2 kbps
- Link protocol—LAPD
- Point-to-point, broadcast, and multicast delivery

Mobile integrated radio system (MIRS). MIRS is a cellular-like system developed by Motorola to operate at the special mobile radio (SMR) band shown in Fig. 15.46. MIRS system is called by the manufacturers. The system providers call this system ESMR (enhanced SMR). The SMR band is in the Part 90 of FCC CFR in the private sector.



		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 Mhz
	ESMR	806-821 MHz	851-866 MHz

Figure 15.46 Cellular and ESMR spectrum.

Public information on this system is limited. However, the operational function is the same as the cellular system. Since many existing SMR systems are still in operation, the MIRS frequency channels in the SMR band are restricted depending on the area in which they operate. Therefore, a special frequency assignment arrangement will be imposed in the MIRS system.

A top-level system specification is as follows:

- Full-duplex communication system
- Frequency—806 to 824 MHz (mobile transmitter), 851 to 869 MHz (base transmitter)
- Channel bandwidth—25 KHz
- Multiple access—TDMA
- Number of time slots—6
- Rate of speech coder—4.2 kbps/slot
- Modulation—16 QAM (quadrature AM)
- Speech coder—VSELP
- No equalizer is implemented
- Handoff
- Transmission rate—6.5 kbps/slot
- Forward error correction—3 kbps
- Interleaving

Mobile satellite communications. In the future, mobile satellite communications will enhance terrestrial radio communication, either in rural areas or in global communication to become a broad sense PCS system. Mobile satellite communications is classified by three satellite altitude positions. One uses stationary satellites which are 22,000 mi above the earth. Another uses low-earth-orbit (LEO) satellites which are 400 to 800 mi above the earth. The third uses medium-earth-orbit satellites (MEO) anywhere between the stationary satellite altitude and LEO satellite altitude.

Stationary satellites. To cover the entire earth, only four stationary satellites are required. The required number of stationary satellites is low and the life span is 10 to 15 years. Thus the cost is low. Immarsat[23] is introduced to PCS. However, the time delay due to the long range can be 0.25 s (two ways) and does not include the other delays such as signal processing and network routing. Also, at high latitudes, the stationary satellite has a low elevation angle as seen from the earth.

TABLE 15.15 Comparative Low-Earth-Orbiting Mobile Satellite Service Applications*

System characteristics	Loral/QUALCOMM GLOBALSTAR	Motorola IRIDIUM	TRW ODYSSEY	Constellation ARIES (b)	Ellipsat ELLIPSO (c)
Number of satellites	48	66	12	48	24
Constellation altitude (NM)	750	421	5600	550	1767 x 230
Unique feature	Transponder	Onboard Processing	Transponder	Transponder	Transponder
Circuit capacity (U.S.)	6500	3835	4600	100	1210
Signal modulation	CDMA	TDMA	CDMA	FDMA/CDMA	CDMA
Gateways in U.S.	6	2	2	5	6
Gateway	C-band existing	New Ka-Band	New Ka-band	Unknown	Unknown
Coverage	Global	Global	Global	Global	Northern hemisphere

For instance, in Chicago, the elevation angle is 19°. Under this condition, the buildings and the hills block the direct path between the satellite and the mobile. This causes multipath fading and shadow loss. Also, the transmit power and size of antenna to send the signal up to the satellite from a mobile station is limited.

LEO satellites. Many LEO satellites are required to cover the earth. The Iridium[24] system (Motorola) needs 66 satellites, and Globalstar[25] needs 48 satellites. The average satellite life is 5 years. The cost of utilizing LEO mobile satellite systems is high. The satellite will circle the earth with a period between 1 to 2 h, depending on the satellite altitude. Handoffs between satellite antenna spot beams and between satellites may create difficulties for switching signals in space. However, there is no noticeable delay in talk time, and the mobile transmit power is not an issue. A comparison of LEO mobile satellite service applications among different systems is shown in Table 15.15.

MEO Satellites. MEO satellites may share the advantages and disadvantages of both stationary satellites and LEO satellites. Odyssey Personal Communication Satellite System[26] developed by TRW belongs to MEO satellites.

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APPENDIX 15.1 Major Trading Areas in the United States ***(from Rand McNally & Company)***

No.	Major trading area	Abbrev.	Number of basic areas
01	Atlanta	ATL	14
03	Birmingham	BIR	10
05	Boston-Providence	BOS-PRO	14
07	Buffalo-Rochester	BUF-ROC	4
09	Charlotte-Greensboro-Greenville-Raleigh	C-G-G-R	23
11	Chicago	CHI	18
13	Cincinnati-Dayton	CIN-DAY	9
15	Cleveland	CLEV	10
17	Columbus	COL	6
19	Dallas-Fort Worth	DAL-F.W.	22
21	Denver	DEN	12
23	Des Moines-Quad Cities	DES-Q.C.	13
25	Detroit	DET	18
27	El Paso-Albuquerque	ELP-AL	8
29	Honolulu	HON	4
31	Houston	HOU	6
33	Indianapolis	IND	11
35	Jacksonville	JAX	7
37	Kansas City	K.C.	9
39	Knoxville	KNOX	3
41	Little Rock	L.R.	9
43	Los Angeles-San Diego	L.A.-S.D.	7
45	Louisville-Lexington-Evansville	L-L-E	9
47	Memphis-Jackson	MEM-JAK	11
49	Miami-Fort Lauderdale	MIA-F.L.	5
51	Milwaukee	MILW	16
53	Minneapolis-St. Paul	MPLS-S.P.	23
55	Nashville	NASH	3
57	New Orleans-Baton Rouge	N.O.-B.R.	13
59	New York	N.Y.	20
61	Oklahoma City	O.C.	8
63	Omaha	OMA	7
65	Philadelphia	PHIL	11
67	Phoenix	PHOE	7
69	Pittsburgh	PGH	12
71	Portland	POR	9
73	Richmond-Norfolk	RICH-NOR	7
75	St. Louis	ST.L.	12
77	Salt Lake City	S.L.C.	8
79	San Antonio	SANT	6
81	San Francisco-Oakland-San Jose	SF-O-SJ	13
83	Seattle	SEAT	11
85	Spokane-Billings	SPOK-BIL	11
87	Tampa-St. Petersburg-Orlando	T-SP-O	7
89	Tulsa	TUL	4
91	Washington-Baltimore	WASH-BAL	9
93	Wichita	WICH	8
	U. S. Total		487

Broadband PCS Major Trading Area Designations *(from FCC)*

Market no.	Major trading area
M 001	New York
M 002	Los Angeles-San Diego
M 003	Chicago
M 004	San Francisco-Oakland-San Jose
M 005	Detroit
M 006	Charlotte-Greensboro-Greenville-Raleigh
M 007	Dallas-Fort Worth
M 008	Boston-Providence
M 009	Philadelphia
M 010	Washington-Baltimore
M 011	Atlanta
M 012	Minneapolis-St. Paul
M 013	Tampa-St Petersburg-Orlando
M 014	Houston
M 015	Miami-For Lauderdale
M 016	Cleveland
M 017	New Orleans-Baton Rouge
M 018	Cincinnati-Dayton
M 019	St. Louis
M 020	Milwaukee
M 021	Pittsburgh
M 022	Denver
M 023	Richmond-Norfolk
M 024	Seattle (Excluding Alaska)
M 025	Puerto Rico-U.S. Virgin Islands
M 026	Louisville-Lexington-Evansville
M 027	Phoenix
M 028	Memphis-Jackson
M 029	Birmingham
M 030	Portland
M 031	Indianapolis
M 032	Des Moines-Quad Cities
M 033	San Antonio
M 034	Kansas City
M 035	Buffalo-Rochester
M 036	Salt Lake City
M 037	Jacksonville
M 038	Columbus
M 039	El Paso-Albuquerque
M 040	Little Rock
M 041	Oklahoma City
M 042	Spokane-Billings
M 043	Nashville
M 044	Knoxville
M 045	Omaha
M 046	Wichita
M 047	Honolulu
M 048	Tulsa
M 049	Alaska
M 050	Guam-Northern Mariana Islands
M 051	American Samoa

Basic Trading Areas in the United States

No.	Basic trading area
001	Aberdeen, SD
002	Aberdeen, WA
003	Abilene, TX
004	Ada, OK
005	Adrian, MI
006	Albany-Tifton, GA
007	Albany-Schenectady, NY
008	Albuquerque, NM
009	Alexandria, LA
010	Allentown-Bethlehem-Easton, PA
011	Alpena, MI
012	Altoona, PA
013	Amarillo, TX
014	Anchorage, AK
015	Anderson, IN
016	Anderson, SC
017	Anniston, AL
018	Appleton-Oshkosh, WI
019	Ardmore, OK
020	Asheville-Hendersonville, NC
021	Ashtabula, OH
022	Athens, GA
023	Athens, OH
024	Atlanta, GA
025	Atlantic City, NJ
026	Augusta, GA
027	Austin, TX
028	Bakersfield, CA
029	Baltimore, MD
030	Bangor, ME
031	Bartlesville, OK
032	Baton Rouge, LA
033	Battle Creek, MI
034	Beaumont-Port Arthur, TX
035	Beckley, WV
036	Bellingham, WA
037	Bemidji, MN
038	Bend, OR
039	Benton Harbor, MI
040	Big Spring, TX
041	Billings, MT
042	Biloxi-Gulfport-Pascagoula, MS
043	Binghamton, NY
044	Birmingham, AL
045	Bismarck, ND
046	Bloomington, IL
047	Bloomington-Bedford, IN
048	Bluefield, WV
049	Blytheville, AR
050	Boise-Nampa, ID

Basic Trading Areas in the United States

No.	Basic trading area
051	Boston, MA
052	Bowling Green-Glasgow, KY
053	Bozeman, MT
054	Brainerd, MN
055	Bremerton, WA
056	Brownsville-Harlingen, TX
057	Brownwood, TX
058	Brunswick, GA
059	Bryan-College Station, TX
060	Buffalo-Niagara Falls, NY
061	Burlington, IA
062	Burlington, NC
063	Burlington, VT
064	Butte, MT
065	Canton-New Philadelphia, OH
066	Cape Girardeau-Sikeston, MO
067	Carbondale-Marion, IL
068	Carlsbad, NM
069	Casper-Gillette, WY
070	Cedar Rapids, IA
071	Champaign-Urbana, IL
072	Charleston, SC
073	Charleston, WV
074	Charlotte-Gastonia, NC
075	Charlottesville, VA
076	Chattanooga, TN
077	Cheyenne, WY
078	Chicago, IL
079	Chico-Oroville, CA
080	Chillicothe, OH
081	Cincinnati, OH
082	Clarksburg-Elkins, WV
083	Clarksville, TN-Hopkinsville, KY
084	Cleveland-Akron, OH
085	Cleveland, TN
086	Clinton, IA-Sterling, IL
087	Clovis, NM
088	Coffeyville, KS
089	Colorado Springs, CO
090	Columbia, MO
091	Columbia, SC
092	Columbus, GA
093	Columbus, IN
094	Columbus-Starkville, MS
095	Columbus, OH
096	Cookeville, TN
097	Coos Bay-North Bend, OR
098	Corbin, KY
099	Corpus Christi, TX
100	Cumberland, MD

Basic Trading Areas in the United States

No.	Basic trading area
101	Dallas-Fort Worth, TX
102	Dalton, GA
103	Danville, IL
104	Danville, VA
105	Davenport, IA-Moline, IL
106	Dayton-Springfield, OH
107	Daytona Beach, FL
108	Decatur, AL
109	Decatur-Effingham, IL
110	Denver, CO
111	Des Moines, IA
112	Detroit, MI
113	Dickinson, ND
114	Dodge City, KS
115	Dothan-Enterprise, AL
116	Dover, DE
117	Du Bois-Clearfield, PA
118	Dubuque, IA
119	Duluth, MN
120	Dyersburg-Union City, TN
121	Eagle Pass-Del Rio, TX
122	East Liverpool-Salem, OH
123	Eau Claire, WI
124	El Centro-Calexico, CA
125	El Dorado-Magnolia-Camden, AR
126	Elkhart, IN
127	Elmira-Corning-Hornell, NY
128	El Paso, TX
129	Emporia, KS
130	Enid, OK
131	Erie, PA
132	Escanaba, MI
133	Eugene-Springfield, OR
134	Eureka, CA
135	Evansville, IN
136	Fairbanks, AK
137	Fairmont, WV
138	Fargo, ND
139	Farmington, NM-Durango, CO
140	Fayetteville-Springdale-Rogers, AR
141	Fayetteville-Lumberton, NC
142	Fergus Falls, MN
143	Findlay-Tiffin, OH
144	Flagstaff, AZ
145	Flint, MI
146	Florence, AL
147	Florence, SC
148	Fond du Lac, WI
149	Fort Collins-Loveland, CO
150	Fort Dodge, IA

Basic Trading Areas in the United States

No.	Basic trading area
151	Fort Myers, FL
152	Fort Pierce-Vero Beach-Stuart, FL
153	Fort Smith, AR
154	Fort Walton Beach, FL
155	Fort Wayne, IN
156	Fredericksburg, VA
157	Fresno, CA
158	Gadsden, AL
159	Gainesville, FL
160	Gainesville, GA
161	Galesburg, IL
162	Gallup, NM
163	Garden City, KS
164	Glens Falls, NY
165	Goldsboro-Kinston, NC
166	Grand Forks, ND
167	Grand Island-Kearney, NE
168	Grand Junction, CO
169	Grand Rapids, MI
170	Great Bend, KS
171	Great Falls, MT
172	Greeley, CO
173	Green Bay, WI
174	Greensboro-Winston-Salem-High Point, NC
175	Greenville-Greenwood, MS
176	Greenville-Washington, NC
177	Greenville-Spartanburg, SC
178	Greenwood, SC
179	Hagerstown, MD-Chambersburg, PA-Martinsburg, WV
180	Hammond, LA
181	Harrisburg, PA
182	Harrison, AR
183	Harrisonburg, VA
184	Hartford, CT
185	Hastings, NE
186	Hattiesburg, MS
187	Hays, KS
188	Helena, MT
189	Hickory-Lenoir-Morganton, NC
190	Hilo, HI
191	Hobbs, NM
192	Honolulu, HI
193	Hot Springs, AR
194	Houghton, MI
195	Houma-Thibodaux, LA
196	Houston, TX
197	Huntington, WV-Ashland, KY
198	Huntsville, AL
199	Huron, SD
200	Hutchinson, KS

Basic Trading Areas in the United States

No.	Basic trading area
201	Hyannis, MA
202	Idaho Falls, ID
203	Indiana, PA
204	Indianapolis, IN
205	Iowa City, IA
206	Iron Mountain, MI
207	Ironwood, MI
208	Ithaca, NY
209	Jackson, MI
210	Jackson, MS
211	Jackson, TN
212	Jacksonville, FL
213	Jacksonville, IL
214	Jacksonville, NC
215	Jamestown, NY-Warren, PA-Dunkirk, NY
216	Janesville-Beloit, WI
217	Jefferson City, MO
218	Johnstown, PA
219	Jonesboro-Paragould, AR
220	Joplin, MO-Miami, OK
221	Juneau-Ketchikan, AK
222	Kahului-Wailuku-Lahaina, HI
223	Kalamazoo, MI
224	Kalispell, MT
225	Kankakee, IL
226	Kansas City, MO
227	Keene, NH
228	Kennewick-Pasco-Richland, WA
229	Kingsport, TN-Johnson City, TN-Bristol, VA-TN
230	Kirksville, MO
231	Klamath Falls, OR
232	Knoxville, TN
233	Kokomo-Logansport, IN
234	La Crosse, WI-Winona, MN
235	Lafayette, IN
236	Lafayette-New Iberia, LA
237	La Grange, GA
238	Lake Charles, LA
239	Lakeland-Winter Haven, FL
240	Lancaster, PA
241	Lansing, MI
242	Laredo, TX
243	La Salle-Peru-Ottawa-Streator, IL
244	Las Cruces, NM
245	Las Vegas, NV
246	Laurel, MS
247	Lawrence, KS
248	Lawton-Duncan, OK
249	Lebanon-Claremont, NH
250	Lewiston-Moscow, ID

Basic Trading Areas in the United States

No.	Basic trading area
251	Lewiston-Auburn, ME
252	Lexington, KY
253	Liberal, KS
254	Lihue, HI
255	Lima, OH
256	Lincoln, NE
257	Little Rock, AR
258	Logan, UT
259	Logan, WV
260	Longview-Marshall, TX
261	Longview, WA
262	Los Angeles, CA
263	Louisville, KY
264	Lubbock, TX
265	Lufkin-Nacogdoches, TX
266	Lynchburg, VA
267	McAlester, OK
268	McAllen, TX
269	McComb-Brookhaven, MS
270	McCook, NE
271	Macon-Warner Robins, GA
272	Madison, WI
273	Madisonville, KY
274	Manchester-Nashua-Concord, NH
275	Manhattan-Junction City, KS
276	Manitowoc, WI
277	Mankato-Fairmont, MN
278	Mansfield, OH
279	Marinette, WI-Menominee, MI
280	Marion, IN
281	Marion, OH
282	Marquette, MI
283	Marshalltown, IA
284	Martinsville, VA
285	Mason City, IA
286	Mattoon, IL
287	Meadville, PA
288	Medford-Grants Pass, OR
289	Melbourne-Titusville, FL
290	Memphis, TN
291	Merced, CA
292	Meridian, MS
293	Miami-Fort Lauderdale, FL
294	Michigan City-La Porte, IN
295	Middlesboro-Harlan, KY
296	Midland, TX
297	Milwaukee, WI
298	Minneapolis-St. Paul, MN
299	Minot, ID
300	Missoula, MT

Basic Trading Areas in the United States

No.	Basic trading area
301	Mitchell, SD
302	Mobile, AL
303	Modesto, CA
304	Monroe, LA
305	Montgomery, AL
306	Morgantown, WV
307	Mount Pleasant, MI
308	Mount Vernon-Centralia, IL
309	Muncie, IN
310	Muskegon, MI
311	Muskogee, OK
312	Myrtle Beach, SC
313	Naples, FL
314	Nashville, TN
315	Natchez, MS
316	New Bern, NC
317	New Castle, PA
318	New Haven-Waterbury-Meriden, CT
319	New London-Norwich, CT
320	New Orleans, LA
321	New York, NY
322	Nogales, AZ
323	Norfolk, NE
324	Norfolk-Virginia Beach-Newport News-Hampton, VA
325	North Platte, NE
326	Ocala, FL
327	Odessa, TX
328	Oil City-Franklin, PA
329	Oklahoma City, OK
330	Olean, NY-Bradford, PA
331	Olympia-Centralia, WA
332	Omaha, NE
333	Oneonta, NY
334	Opelika-Auburn, AL
335	Orangeburg, SC
336	Orlando, FL
337	Ottumwa, IA
338	Owensboro, KY
339	Paducah-Murray-Mayfield, KY
340	Panama City, FL
341	Paris, TX
342	Parkersburg, WV-Marietta, OH
343	Pensacola, FL
344	Peoria, IL
345	Petoskey, MI
346	Philadelphia, PA-Wilmington, DE-Trenton, NJ
347	Phoenix, AZ
348	Pine Bluff, AR
349	Pittsburgh-Parsons, KS
350	Pittsburgh, PA

Basic Trading Areas in the United States

No.	Basic trading area
351	Pittsfield, MA
352	Plattsburgh, NY
353	Pocatello, ID
354	Ponca City, OK
355	Poplar Bluff, MO
356	Port Angeles, WA
357	Portland-Brunswick, ME
358	Portland, OR
359	Portsmouth, OH
360	Pottsville, PA
361	Poughkeepsie-Kingston, NY
362	Prescott, AZ
363	Presque Isle, ME
364	Providence-Pawtucket, RI-New Bedford-Fall River, MA
365	Provo-Orem, UT
366	Pueblo, CO
367	Quincy, IL-Hannibal, MO
368	Raleigh-Durham, NC
369	Rapid City, SD
370	Reading, PA
371	Redding, CA
372	Reno, NV
373	Richmond, IN
374	Richmond-Petersburg, VA
375	Riverton, WY
376	Roanoke, VA
377	Roanoke Rapids, NC
378	Rochester-Austin-Albert Lea, MN
379	Rochester, NY
380	Rockford, IL
381	Rock Springs, WY
382	Rocky Mount-Wilson, NC
383	Rolla, MO
384	Rome, GA
385	Roseburg, OR
386	Roswell, NM
387	Russellville, AR
388	Rutland-Bennington, VT
389	Sacramento, CA
390	Saginaw-Bay City, MI
391	St. Cloud, MN
392	St. George, UT
393	St. Joseph, MO
394	St. Louis, MO
395	Salem-Albany-Corvallis, OR
396	Salina, KS
397	Salinas-Monterey, CA
398	Salisbury, MD
399	Salt Lake City-Ogden, UT
400	San Angelo, TX

Basic Trading Areas in the United States

No.	Basic trading area
401	San Antonio, TX
402	San Diego, CA
403	Sandusky, OH
404	San Francisco-Oakland-San Jose, CA
405	San Luis Obispo, CA
406	Santa Barbara-Santa Maria, CA
407	Santa Fe, NM
408	Sarasota-Bradenton, FL
409	Sault Ste. Marie, MI
410	Savannah, GA
411	Scottsbluff, NE
412	Scranton-Wilkes-Barre-Hazleton, PA
413	Seattle-Tacoma, WA
414	Sedalia, MO
415	Selma, AL
416	Sharon, PA
417	Sheboygan, WI
418	Sherman-Denison, TX
419	Shreveport, LA
420	Sierra Vista-Douglas, AZ
421	Sioux City, IA
422	Sioux Falls, SD
423	Somerset, KY
424	South Bend-Mishawaka, IN
425	Spokane, WA
426	Springfield, IL
427	Springfield-Holyoke, MA
428	Springfield, MO
429	State College, PA
430	Staunton-Waynesboro, VA
431	Steubenville, OH-Weirton, WV
432	Stevens Point-Marshfield-Wisconsin Rapids, WI
433	Stillwater, OK
434	Stockton, CA
435	Stroudsburg, PA
436	Sumter, SC
437	Sunbury-Shamokin, PA
438	Syracuse, NY
439	Tallahassee, FL
440	Tampa-St. Petersburg-Clearwater, FL
441	Temple-Killeen, TX
442	Terre Haute, IN
443	Texarkana, TX-AR
444	Toledo, OH
445	Topeka, KS
446	Traverse City, MI
447	Tucson, AZ
448	Tulsa, OK
449	Tupelo-Corinth, MS
450	Tuscaloosa, AL

Basic Trading Areas in the United States

No.	Basic trading area
451	Twin Falls, ID
452	Tyler, TX
453	Utica-Rome, NY
454	Valdosta, GA
455	Vicksburg, MS
456	Victoria, TX
457	Vincennes-Washington, IN
458	Visalia-Porterville-Hanford, CA
459	Waco, TX
460	Walla Walla, WA-Pendleton, OR
461	Washington, DC
462	Waterloo-Cedar Falls, IA
463	Watertown, NY
464	Watertown, SD
465	Waterville-Augusta, ME
466	Wausau-Rhinelander, WI
467	Waycross, GA
468	Wenatchee, WA
469	West Palm Beach-Boca Raton, FL
470	West Plains, MO
471	Wheeling, WV
472	Wichita, KS
473	Wichita Falls, TX
474	Williamson, WV-Pikeville, KY
475	Williamsport, PA
476	Williston, MD
477	Willmar-Marshall, MN
478	Wilmington, NC
479	Winchester, VA
480	Worcester-Fitchburgh-Leominster, MA
481	Worthington, MN
482	Yakima, WA
483	York-Hanover, PA
484	Youngstown-Warren, OH
485	Yuba City-Marysville, CA
486	Yuma, AZ
487	Zanesville-Cambridge, OH