Multiple access techniques

Sometimes a satellite's service is present at a particular location on the earth station and sometimes it is not present. That means, a satellite may have different service stations of its own located at different places on the earth. They send carrier signal for the satellite.

In this situation, we do multiple access to enable satellite to take or give signals from different stations at time without any interference between them. Following are the **three types** of multiple access techniques.

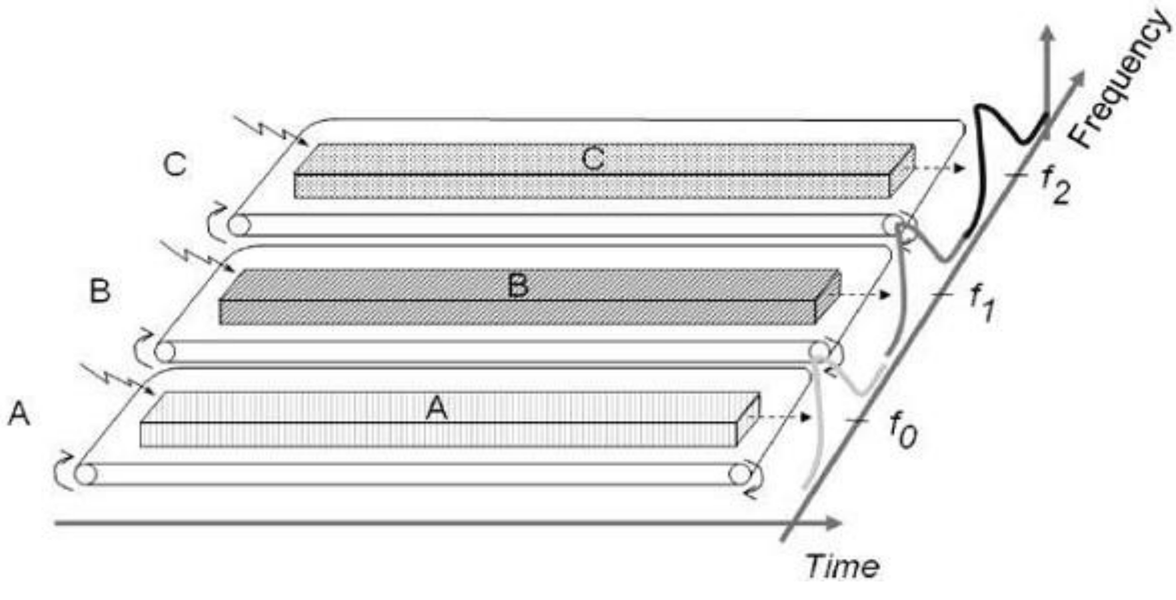
- FDMA (Frequency Division Multiple Access)
- TDMA (Time Division Multiple Access)
- CDMA (Code Division Multiple Access)

Now, let us discuss each technique one by one.

FDMA

In this type of multiple access, we assign each signal a different type of frequency band (range). So, any two signals should not have same type of frequency range. Hence, there won't be any interference between them, even if we send those signals in one channel.

One perfect **example** of this type of access is our radio channels. We can see that each station has been given a different frequency band in order to operate.



Let's take three stations A, B and C. We want to access them through FDMA technique. So we assigned them different frequency bands.

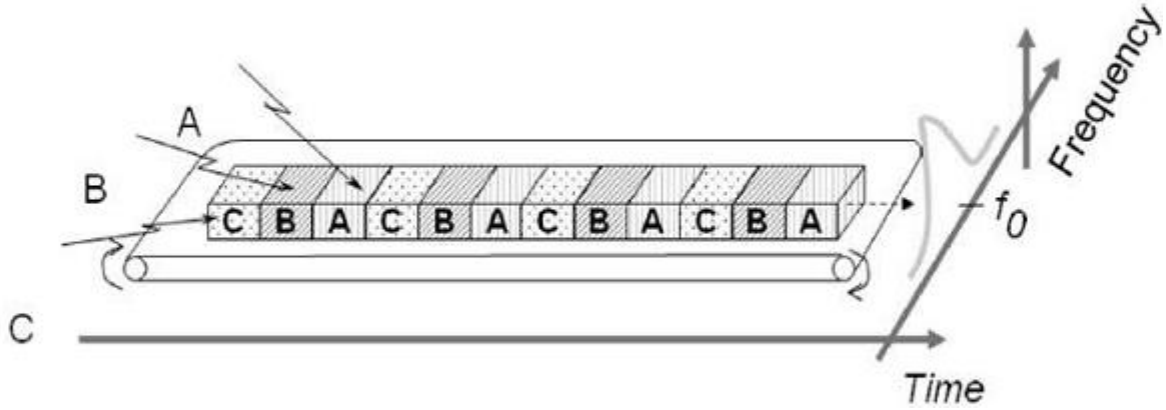
As shown in the figure, satellite station A has been kept under the frequency range of 0 to 20 HZ. Similarly, stations B and C have been assigned the frequency range of 30-60 Hz and 70-90 Hz respectively. There is no interference between them.

The main **disadvantage** of this type of system is that it is very burst. This type of multiple access is not recommended for the channels, which are of dynamic and uneven. Because, it will make their data as inflexible and inefficient.

TDMA

As the name suggests, TDMA is a time based access. Here, we give certain time frame to each channel. Within that time frame, the channel can access the entire spectrum bandwidth

Each station got a fixed length or slot. The slots, which are unused will remain in idle stage.



Suppose, we want to send five packets of data to a particular channel in TDMA technique. So, we should assign them certain time slots or **time frame** within which it can access the entire bandwidth.

In above figure, packets 1, 3 and 4 are active, which transmits data. Whereas, packets 2 and 5 are idle because of their non-participation. This format gets repeated every time we assign bandwidth to that particular channel.

Although, we have assigned certain time slots to a particular channel but it can also be changed depending upon the load bearing capacity. That means, if a channel is transmitting heavier loads, then it can be assigned a bigger time slot than the channel which is transmitting lighter loads. This is the biggest **advantage** of TDMA over FDMA. Another advantage of TDMA is that the power consumption will be very low.

Note − In some applications, we use the **combination** of both **TDMA and FDMA** techniques. In this case, each channel will be operated in a particular frequency band for a particular time frame. In this case, the frequency selection is more robust and it has greater capacity over time compression.

CDMA

In CDMA technique, a unique code has been assigned to each channel to distinguish from each other. A perfect **example** of this type of multiple access is our cellular system. We can see that no two persons' mobile number match with each other although they are same X or Y mobile service providing company's customers using the same bandwidth.

In CDMA process, we do the decoding of inner product of the encoded signal and chipping sequence. Therefore, mathematically it can be written as

Encodedsignal=Orginaldata×chippingsequenceEncodedsignal=Orginaldata×chippingsequence

The basic **advantage** of this type of multiple access is that it allows all users to coexist and use the entire bandwidth at the same time. Since each user has different code, there won't be any interference.

In this technique, a number of stations can have number of channels unlike FDMA and TDMA. The best part of this technique is that each station can use the entire spectrum at all time

Statistical characterisation of multipath channels

Wireless channel is of time-varying nature in which the parameters randomly change with respect to time. Wireless channel is very harsh when compared to AWGN channel model which is often considered for simulation and modeling. Understanding the various characteristics of a wireless channel and understanding their physical significance is of paramount importance. In these series of articles, I seek to expound various statistical characteristics of a multipath wireless channel by giving more importance to the concept than the mathematical derivation.
Complex Baseband Mutipath Channel Model:

In a multipath channel, multiple copies of a signal travel different paths with different propagation delays τ and are received at the receiver at different phase angles and strengths. These rays add constructively or destructively at the receiver front end, thereby giving rise to rapid fluctuations in the channel. The multipath channel can be viewed as a linear time variant system where the parameters change randomly with respect to time. The channel impulse response is a two dimensional random variable – $h(t,\tau)$ that is a function of two parameters – instantaneous time t and the propagation delay τ. The channel is expressed as a set of random complex gains at a given time t and the propagation delay τ. The output of the channel $y(t)$ can be expressed as the convolution of the complex channel impulse response $h(t,\tau)$ and the input $x(t)$
y(t)=∫∞0h(t–τ,τ)x(t−τ)dτ(1)y(t)=∫0∞h(t–τ,τ)x(t−τ)dτ(1)

If the complex channel gains are typically drawn from a complex Gaussian distribution, then at any given time t, the absolute value of the impulse response $|h(t,\tau)|$ is Rayleigh distributed (if the mean of the distribution $E\left(h(t,\tau)\right)=0$ or Rician distributed (if the mean of the distribution $E\left[h(t,\tau)\right)\neq 0]$. These two scenarios model the presence or absence of a Line of Sight (LOS) path between the transmitter and the receiver.
Here, the values for the channel impulse response are samples of a random process that is defined with respect to time t and the multipath delay ***τ***. That is, for each combination of t and ***τ***, a randomly drawn value is assigned for the channel impulse response. As with any other random process, we can calculate the general ***autocorrelation function*** as
Rhh(t1,t2;τ1,τ2)=E[h(t1,τ1)h∗(t2,τ2)](2)Rhh(t1,t2;τ1,τ2)=E[h(t1,τ1)h∗(t2,τ2)](2)

Given the generic autocorrelation function above, following assumptions can be made to restrict the channel model to the following specific set of categories

- Wide Sense Stationary channel model
- Uncorrelated Scattering channel model
- Wide Sense Stationary Uncorrelated Scattering channel model

Wide Sense Stationary (WSS) channel model

In this channel model, the impulse response of the channel is considered Wide Sense Stationary (WSS) , that is the channel impulse response is independent of time t. In other words, the autocorrelation function $R_{hh}(t,\tau)$ is independent of time instant tt and it depends on the difference between the time instants $\Delta t = t_2 - t_1$ where $t_1 = t$ and $t_2 = t + \Delta t$. The autocorrelation function for WSS channel model is expressed as
$Rhh(\Delta t;\tau 1,\tau 2)=E[h(t,\tau 1)h*(t+\Delta t,\tau 2)](3)Rhh(\Delta t;\tau 1,\tau 2)=E[h(t,\tau 1)h*(t+\Delta t,\tau 2)](3)$

Uncorrelated Scattering (US) channel model

Here, the individual scattered components arriving at the receiver front end (at different propagation delays) are assumed to be uncorrelated. Thus the autocorrelation function can be expressed as

$Rhh(t1,t2;\tau 1,\tau 2)=Rhh(t1,t2;\tau 1)\delta(\tau 1-\tau 2)(4)Rhh(t1,t2;\tau 1,\tau 2)=Rhh(t1,t2;\tau 1)\delta(\tau 1-\tau 2)(4)$

Wide Sense Stationary Uncorrelated Scattering (WSSUS) channel model

The WSSUS channel model combines the aspects of WSS and US channel model that are discussed above. Here, the channel is considered as Wide Sense Stationary and the scattering components arriving at the receiver are assumed to be uncorrelated. Combining both the worlds, the autocorrelation function $R_{hh}(\Delta t,\tau)$is

$Rhh(\Delta t,\tau)=E[h(t,\tau)h*(t+\Delta t,\tau)](5)Rhh(\Delta t,\tau)=E[h(t,\tau)h*(t+\Delta t,\tau)](5)$

Scattering Function

The autocorrelation function of the WSSUS channel model can be represented in frequency domain by taking Fourier transform with respect one or both variables – difference in time Δt and the propagation delay τ. Of the two forms, the Fourier transform on the variable Δt gives specific insight to channel properties in terms of propagation delay τ and the Doppler Frequency f simultaneously. The Fourier transform of the above two-dimensional autocorrelation function on the variable Δt is called ***scattering function*** and is given by
$S(f,\tau)=\int\infty-\infty Rhh(\Delta t,\tau)e-j2\pi f\Delta td\Delta t(6)S(f,\tau)=\int-\infty\infty Rhh(\Delta t,\tau)e-j2\pi f\Delta td\Delta t(6)$

Fourier transform of relative time Δt is Doppler Frequency. Thus, the scattering function is a function of two variables – Dopper Frequency f and the propagation delay τ. It gives the average output power of the channel as a function of Doppler Frequency f and the propagation delay τ.

Two important relationships can be derived from the scattering function – ***Power Delay Profile (PDP)*** and ***Doppler Power Spectrum***. Both of them affect the performance of a given wireless channel. Power Delay Profile is a function of propagation delay and the Doppler Power Spectrum is a function of Doppler Frequency.
Power Delay Profile $p(\tau)$ gives the signal intensity received over a multipath channel as a function of propagation delays. It is obtained as the spatial average of the complex baseband channel impulse response as

$p(\tau)=Rhh(0,\tau)=E[|h(t,\tau)|2](7)p(\tau)=Rhh(0,\tau)=E[|h(t,\tau)|2](7)$

Power Delay Profile can also be obtained from scattering function, by integrating it over the entire frequency range (removing the dependence on Doppler frequency).

$p(\tau)=\int\infty-\infty S(f,\tau)df(8)p(\tau)=\int-\infty\infty S(f,\tau)df(8)$

Similarly, the Doppler Power Spectrum can be obtained by integrating the scattering function over the entire range of propagation delays.

$S(f)=\int\infty-\infty S(f,\tau)d\tau(9)S(f)=\int-\infty\infty S(f,\tau)d\tau(9)$

Fourier Transform of Power Delay Profile and Inverse Fourier Transform of Doppler Power Spectrum:

Power Delay Profile is a function of time which can be transformed to frequency domain by taking Fourier Transform. Fourier Transform of Power Delay Profile is called ***spaced-frequency correlation function***. Spaced-frequency correlation function describes the spreading of a signal in frequency domain. This gives rise to the importance channel parameter – ***Coherence Bandwidth***.
Similarly, the Doppler Power Spectrum describes the output power of the multipath channel in frequency domain. The Doppler Power Spectrum when translated to time-domain by means of inverse Fourier transform is called ***spaced-time correlation function***. Spaced-time correlation function describes the correlation between different scattered signals received at two different times as a function of the difference in the received time. It gives rise to the concept of **Coherence Time**.

- **Diversity techniques**

Forms of diversity The basic idea of diversity reception is to use several replicas of the same information obtained through a number of independently fading channels. If the probability that the channel gain is below a critical level is p, then with L independent channels the probability is L p . Diversity can be reached in different ways:

• Frequency diversity with Ldifferent channels with frequency separation exceeding the coherence bandwidth.

• Time diversity with Ldifferent time slots with time separation exceeding the coherence time.

• Space diversity with L antennas experiencing independently fading channels. In multiantenna systems, diversity can be applied in various different ways.

System model

Here we consider in simplified scenario with

• frequency flat channels

• slow Rayleigh fading, i.e., the channel is constant during each symbol/frame

• independent fading of the diversity branches.

Receiver antenna diversity in rich scattering environment with narrowband signals is one example of such a scenario.

• In this case, different antennas capture energy originating from a single transmission. This leads to additional array gain, i.e., for fixed transmission power, the total average SNR is L times the average SNR in single antenna reception.

• In many other forms of diversity transmission, such array gain is not experienced

- **Binary signaling over Rayleigh fading channel**

distributed random variable describing the phase-shift in transmission, and $\tilde{w}(t)$ is a complex-valued white Gaussian noise process. It is assumed that the channel is flat in both time and frequency, so that we can estimate the phase-shift ϕ from the received signal without error. Suppose then that coherent binary phase-shift keying is used to do the data transmission. Under the condition that α is fixed or constant over a bit interval, we may adapt Equation (6.20) of Chapter 6 for the situation at hand by expressing the average probability of symbol error (i.e., bit error rate) due to the additive white Gaussian noise acting alone as follows:

$$P_e(\gamma) = \frac{1}{2}\operatorname{erfc}(\sqrt{\gamma}) \tag{8.64}$$

where γ is an attenuated version of the transmitted signal energy per bit-to-noise spectral density ratio E_b/N_0, as shown by

$$\gamma = \frac{\alpha^2 E_b}{N_0} \tag{8.65}$$

Now, insofar as a mobile radio channel is concerned, we may view $P_e(\gamma)$ as a conditional probability given that α is fixed. Thus, to evaluate the average probability of symbol error in the combined presence of fading and noise, we must average $P_e(\gamma)$ over all possible values of γ, as shown by

$$P_e = \int_0^\infty P_e(\gamma) f(\gamma)\, d\gamma \tag{8.66}$$

where $f(\gamma)$ is the probability density function of γ. From Equation (8.65) we note that γ depends on the squared value of α. Since α is Rayleigh distributed, we find that γ has a *chi-square distribution* with two degrees of freedom.[7] In particular, we may express the probability density function of γ as

$$f(\gamma) = \frac{1}{\gamma_0}\exp\left(-\frac{\gamma}{\gamma_0}\right), \qquad \gamma \geq 0 \tag{8.67}$$

The term γ_0 is the *mean value of the received signal energy per bit-to-noise spectral density ratio*, which is defined by

$$\begin{aligned} \gamma_0 &= E[\gamma] \\ &= \frac{E_b}{N_0} E[\alpha^2] \end{aligned} \tag{8.68}$$

where $E[\alpha^2]$ is the mean-square value of the Rayleigh-distributed random variable α. Substituting Equations (8.64) and (8.67) into (8.66), and carrying out the integration, we get the final result

$$P_e = \frac{1}{2}\left(1 - \sqrt{\frac{\gamma_0}{1+\gamma_0}}\right) \tag{8.69}$$

Equation (8.69) defines the bit error rate for coherent binary phase-shift keying (PSK) over a flat-flat Rayleigh fading channel. Following a similar approach, we may derive the corresponding bit error rates for coherent binary frequency-shift keying (FSK), binary differential phase-shift keying (DPSK), and noncoherent binary FSK. The results of these evaluations are summarized in Table 8.2. In Figure 8.22, we have used the exact formulas of Table 8.2 to plot the bit error rate versus γ_0 expressed in decibels. For the sake of comparison, we have also included in Figure 8.22 plots for the bit error rates of coherent binary PSK and noncoherent binary FSK for a nonfading channel. We see that Rayleigh other users. To overcome the near-far problem, it is customary to use *power control* at the base station, whereby the base station maintains control over the power level of the transmitted signal from every mobile being served by that base station. The use of power control is particularly important in CDMA systems for another reason. A goal of multiple-access systems is to maximize *system capacity*, which is defined as the largest possible number of users that can be reliably served by the system, given prescribed resources. Clearly, system capacity is compromised if each mobile is free to raise its transmitted power level regardless of other users, since that increase in transmitted power will, in turn, raise the level of multiple-access interference in the system. To maximize system capacity, it is therefore essential that each mobile's transmitter be under the control of the serving base station so that the signal-to-interference ratio is maintained at the minimum acceptable level needed for reliable service.

RAKE RECEIVER

A discussion of wireless communications using CDMA would be incomplete without a description of the *RAKE receiver*.[9] The RAKE receiver was originally developed in the 1950s as a "diversity" receiver designed expressly to equalize the effect of multipath. First, and foremost, it is recognized that useful information about the transmitted signal is contained in the multipath component of the received signal. Thus, taking the viewpoint that multipath may be approximated as a linear combination of differently delayed echoes, the RAKE receiver seeks to combat the effect of multipath by using a correlation method to detect the echo signals individually and then adding them algebraically. In this way, intersymbol interference due to multipath is dealt with by reinserting different delays into the detected echoes so that they perform a constructive rather than destructive role.

Figure 8.26 shows the basic idea behind the RAKE receiver. The receiver consists of a number of *correlators* connected in parallel and operating in a synchronous fashion. Each correlator has two inputs: (1) a delayed version of the received signal and (2) a replica of the pseudo-noise (PN) sequence used as the spreading code to generate the spread-

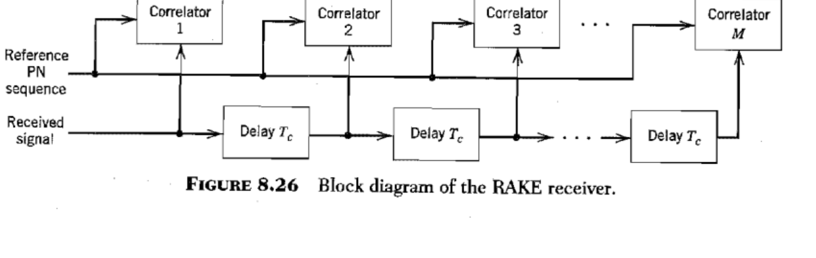


FIGURE 8.26 Block diagram of the RAKE receiver.

spectrum modulated signal at the transmitter. In effect, the PN sequence acts as a "reference signal." Let the nominal bandwidth of the PN sequence be denoted as $W = 1/T_c$, where T_c is the chip duration. From the discussion of spread-spectrum modulation presented in Chapter 7, we recall that the autocorrelation function of a PN sequence has a single peak of width $1/W$, and it disappears toward zero elsewhere inside one period of the PN sequence (i.e., one symbol period). Thus we need only make the bandwidth W of the PN sequence sufficiently large to "identify" the significant echoes in the received signal. To be sure that the correlator outputs all add constructively, two other operations are performed in the receiver by the functional blocks labeled "phase and gain adjustors":

1. An appropriate delay is introduced into each correlator output so that the phase angles of the correlator outputs are in agreement with each other.
2. The correlator outputs are weighted so that the correlators responding to strong paths in the multipath environment have their contributions accentuated, while the correlators not synchronizing with any significant path are correspondingly suppressed.

The weighting coefficients, α_k, are computed in accordance with the *maximal ratio combining principle*:[10]

> The signal-to-noise ratio of a weighted sum, where each element of the sum consists of a signal plus additive noise of fixed power, is maximized when the amplitude weighting is performed in proportion to the pertinent signal strength.

The linear combiner output is

$$y(t) = \sum_{k=1}^{M} \alpha_k z_k(t) \tag{8.71}$$

where $z_k(t)$ is the phase-compensated output of the kth correlator, and M is the number of correlators in the receiver. Provided we use enough correlators in the receiver to span a region of delays sufficiently wide to encompass all the significant echoes that are likely to occur in the multipath environment, the output $y(t)$ behaves essentially as though there was a single propagation path between the transmitter and receiver rather than a series of multiple paths spread in time.

To simplify the presentation, the receiver of Figure 8.26 assumes the use of binary phase-shift keying in performing spread-spectrum modulation at the transmitter. Thus the final operation performed in Figure 8.26 is that of integrating the linear combiner output $y(t)$ over the bit interval T_b and then determining whether binary symbol 1 or 0 was transmitted in that bit interval.

The RAKE receiver derives its name from the fact that the bank of parallel correlators has an appearance similar to the fingers of a rake. Because spread spectrum modulation is basic to the operation of CDMA wireless communications, it is natural for the RAKE receiver to be central to the design of the receiver used in this type of multiuser radio communication.[11]

8.9 *Source Coding of Speech for Wireless Communications*

For the efficient use of channel bandwidth, digital wireless communication systems, be they of the TDMA or CDMA type, rely on the use of *speech coding* to remove almost all