Chapter 8
Multiuser Radio Communications

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Chapter 8
Multiuser Radio Communications

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Chapter 8
Multiuser Radio Communications

- Definition:

  simultaneous use of a communication channel by a number of users

- Main Topics:
  - Multiple-access techniques, *which are basic to multiuser communication systems*
  - Satellite communications, *offering global coverage*

Chapter 8
Multiuser Radio Communications

- Main Topics:(cont.)
  - Radio link analysis, *highlighting the roles of transmitting and receiving antennas and freespace propagation*
  - Wireless communications *with emphasis on mobility and the multipath phenomenon*
  - Adaptive antennas *for wireless communications*

8.1 Introduction

- Ideal communication channel
  - definition: *bandwidth limited* + *additive white Gaussian noise (AWGN)*
  - examples:
    - communication channels presented in earlier chapters
    - Satellite communication channel (*discussed first in this chapter*)

8.1 Introduction

- Satellite communication system in geostationary orbit
  - uplink: an earth terminal → transponder
  - downlink: transponder → another earth terminal
  - line-of-sight radio propagation for the operation of its uplink and downlink
  - offering global coverage

8.1 Introduction

- Wireless communication
  - offering mobility (*communicate with anyone, anywhere*)
  - tetherless (*total freedom of location*).

  Advantages in use for local area networks:
    - elimination of wiring and rewiring
    - flexibility of creating new communication services
    - mobility of users

8.1 Introduction

- Wireless communication(cont.)
  - the presence of multipath (*no longer the idealized AWGN channel model*)

    reflections of transmitted signal from fixed and moving objects → *Multipath (a non-Gaussian form of signal-dependent phenomenon,* intrinsic to operation of indoor and outdoor forms of wireless communications)

    - Disadvantages: *raises practical difficulties in the use & complicates its mathematical analysis*

8.2 Multiple-Access Techniques

- Definition:

Multiple access is a technique whereby many subscribers or local stations can share the use of a communication channel at the same time or nearly so.

State in another way, a *multiple-access technique* permits the communication resources of the channel to be shared by a large number of users communicating with each other.

8.2 Multiple-Access Techniques

- Subtle differences between multiple access and multiplexing
  - 1. *multiple access* -- remote sharing of a communication channel(*e.g. satellite or radio channel*) by users in highly dispersed locations

    multiplexing – near sharing of a channel (*e.g. telephone channel*) by users confined to a local site
  - 2. in a *multiple-access system*: user requirements can change dynamically with time, so provisions are necessary for dynamic channel allocation

    in a *multiplexed system*: user requirements are ordinarily fixed

8.2 Multiple-Access Techniques

- Four basic types of multiple access:
  1. *Frequency-division multiple access(FDMA):* The resources of the channel are divided along the frequency coordinate into disjoint frequency bands. *Disjoint subbands of frequency* are allocated to the different users, as shown in figure 8.1a. *Guard bands* are used to act as buffer zones because of the impossibility of achieving ideal filtering for separating the different users.

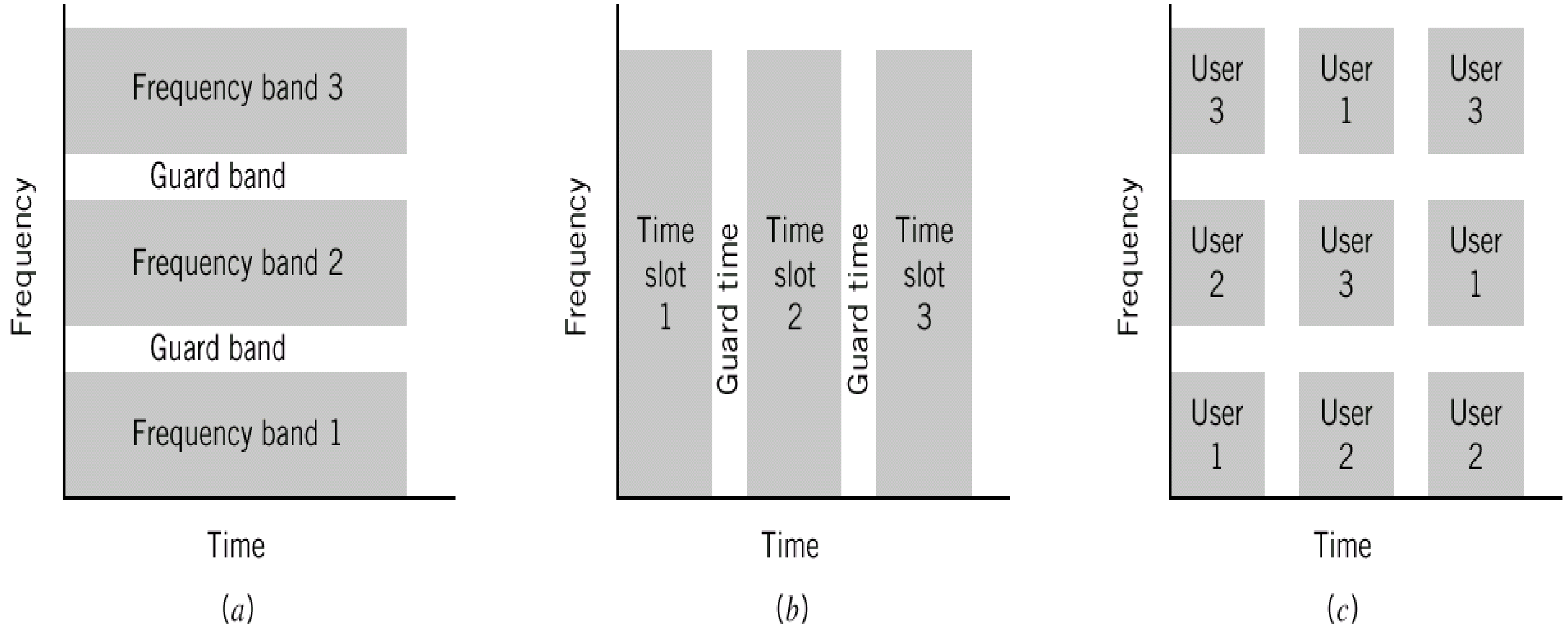


Figure 8.1

Illustrating the ideas behind multiple-access techniques. (*a*) Frequency-division multiple access. (*b*) Time-division multiple access. (*c*) Frequency-hop multiple access.

8.2 Multiple-Access Techniques

- Four basic types of multiple access:

  2. *Time-division multiple access(TDMA)*

  The resources of the channel are divided along the time coordinate into disjoint time slots. *Disjoint time slots* are allocated to different users, as shown in figure 8.1b. Buffer zones in the form of *guard times* are inserted between the assigned time slots for reducing interference.

8.2 Multiple-Access Techniques

- Four basic types of multiple access:
  - 3. *Code-division multiple access(CDMA)*

  The resources are shared by using a hybrid combination of FDMA and TDMA, as shown in figure 8.1c(a specific form of CDMA). *An important advantage* of CDMA over both FDMA and TDMA is that it can provide for *secure* communications.

8.2 Multiple-Access Techniques

- Four basic types of multiple access:
  - 4. *Space-division multiple access(SDMA)* Resource allocation is achieved by exploiting the spatial separation of the individual users (e.g. *Multibeam antennas -- different directions*).
- Common feature of the above multiple access technique: *allocating the communication resources of the channel through the use of disjointedness (or orthogonality in a loose sense) in time, frequency, or space.*

8.3 Satellite Communications

- A geostationary satellite communication system (Figure 8.2)
- The most popular frequency band frequency for satellite communication is 6 GHz (C-band) for the uplink and 4GHz for the downlink.
- Advantages of the use of 6/4GHz:
  - relatively inexpensive microwave equipment
  - low attenuation due to rainfall
  - insignificant sky background noise

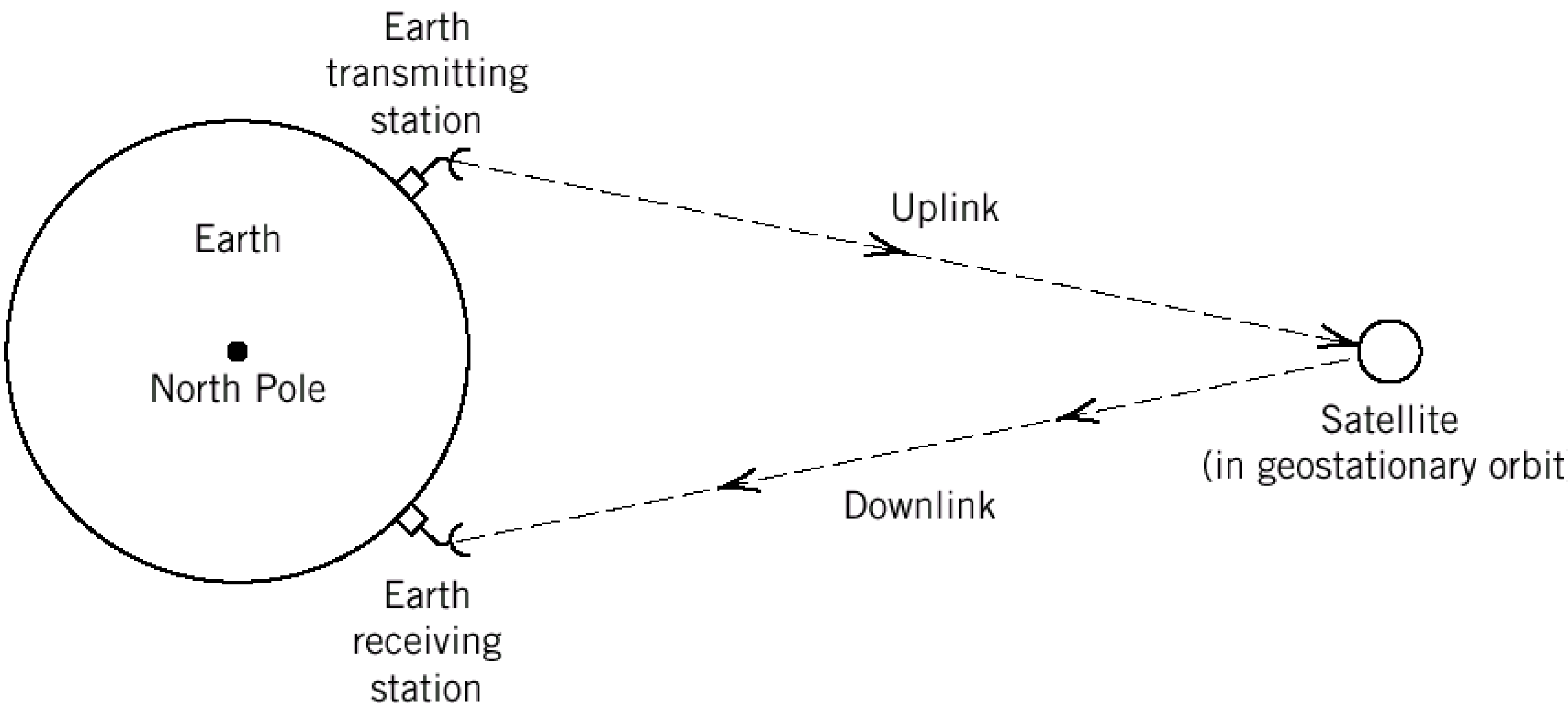


Figure 8.2
Satellite communications system.

8.3 Satellite Communications

- Disadvantage of the use of 6/4GHz:

coinciding with those used for terrestrial microwave system

- Solution:

We develop 14/12GHz frequency band(i.e., Ku-band) to eliminate the problem. Moreover, higher frequencies → smaller and less expensive antennas.

(习题8.5)

8.3 Satellite Communications

- The block diagram of a single transponder channel of a typical communication satellite (Figure 8.3)
- Note:

1. a *single frequency translation* (Figure 8.3).
2. Other channel configurations do the frequency conversion from the uplink to the downlink frequency in *two stages*: *down-conversion to an intermediate frequency*, followed by *amplification*, and the *up-conversion to the desired transmit frequency*.

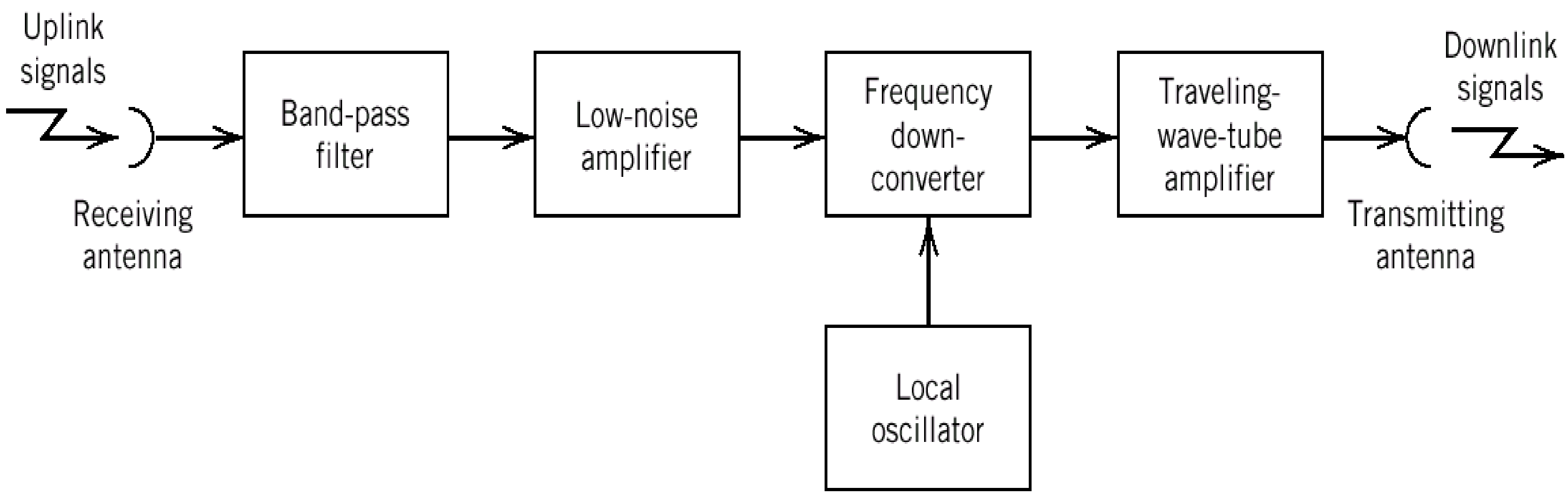


Figure 8.3
Block diagram of transponder.

8.3 Satellite Communications

- In a satellite channel

  large distances → propagation time delay (e.g. *speech signals approximately 270ms*) → echo of the speaker's voice → adaptive echo canceller used

- The satellite channel *(uplink & downlink)* is closely represented by an *additive white Gaussian noise (AWGN) model.* → much of the material presented in chapter 6 is directly applicable to digital satellite communications.

8.3 Satellite Communications

A *satellite transponder* differs from a *conventional microwave line-of-sight repeater* in that many earth stations can access the satellite from *widely different locations* on earth at the same time or nearly so. This capability is made possible by using one of multiple access techniques discussed in Section 8.2.

8.3 Satellite Communications

- Some observations :
  - 1. nonlinearity of the transponder → interference → the traveling-wave tube (TWT) amplifier in the transponder is purposely operated below capacity → power efficiency is reduced in a FDMA system
  - 2. the users access the satellite transponder one at a time in a TDMA system → the satellite transponder can operate close to full power efficiency (permitting TWT amplifier to run into saturation) → TDMA uses transponder more efficiently than FDMA → TDMA used widely in digital satellite communications

8.3 Satellite Communications

- Some observations are as follows:(cont.)
  - 3.SDMA operates by exploiting the spatial locations of earth stations, which is achieved by means of onboard switching. Specifically, the transponder is equipped with multiple antennas, with proper antenna beam being selected for radio transmission to the particular earth station demanding use of the transponder.

8.3 Satellite Communications

- In addition to *multiple access*, another capability of a satellite channel is that of *broadcasting with emphasis on broad area coverage*.
- Broadcasting satellites are characterized by their *high power transmission to inexpensive receivers*. This characteristic is exploited in the use of direct broadcast satellites (DBS), designed for home reception of TV services on a very wide scale.

8.4 Radio Link Analysis

- An important issue in the design of satellite communication systems --- *link (power) budget analysis* --- the totaling of all the gains and losses incurred in operating a communication link
- The essence of radio link analysis also applies to other radio links that rely on *line of sight* for their operation

8.4 Radio Link Analysis

- The *balance sheet* constituting the link budget provides a detailed accounting of three broadly defined items:
  - 1. Apportionment of the resources available to the transmitter and the receiver.
  - 2. Sources responsible for the loss of signal power.
  - 3. Sources of noise.

  Putting all these items together into the link budget, we end up with an *estimation* procedure for evaluating the performance of a radio link.

8.4 Radio Link Analysis

- From the material presented in Chapter 6, we learn that the performance of a digital communication system, in the presence of channel noise modeled as additive white Gaussian noise (AWGN), is defined by a formula having the shape of a “waterfall” curve, as shown in Figure 8.4.

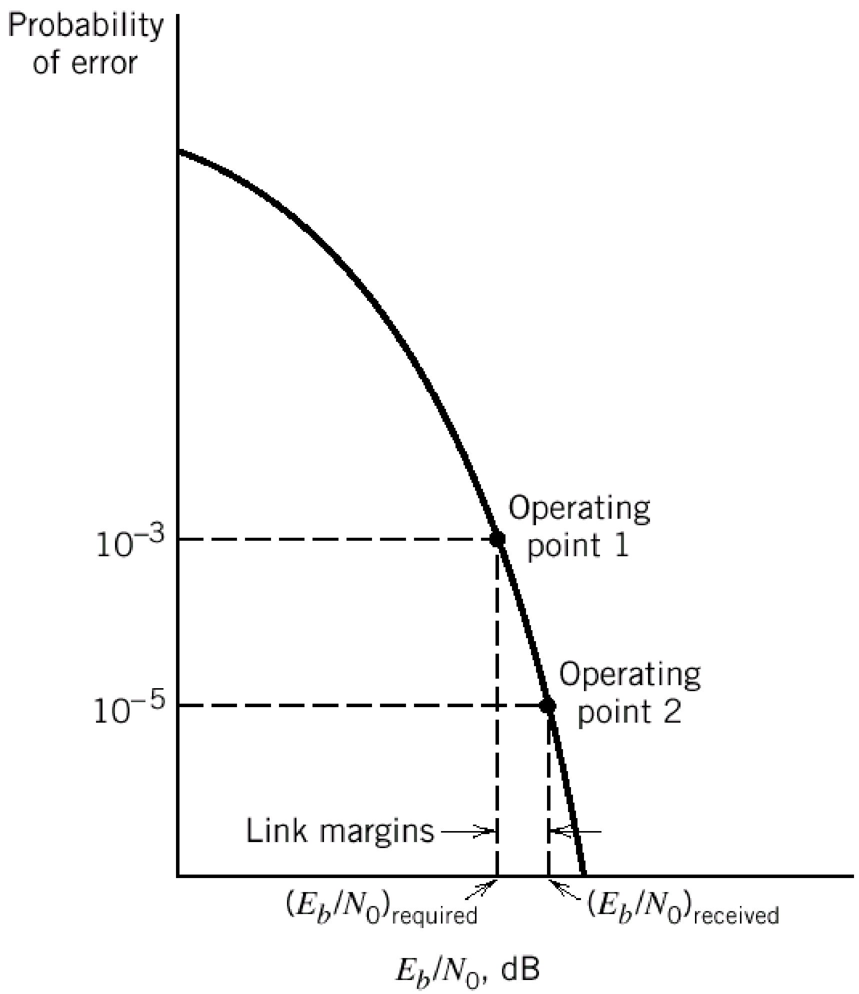


Figure 8.4

“Waterfall” curve relating the probability of error to the E_b/N_0 ratio.

This figure portrays the probability of symbol error, P_e, plotted versus the bit-energy-to-noise spectral density ratio E_b/N_0.

8.4 Radio Link Analysis

- Once a modulation scheme has been chosen, the *first design task* is to specify two particular values of E_b/N_0:
  - 1. *Required* E_b/N_0

    Let $(E_b/N_0)_{\text{req}}$ denote the value of E_b/N_0 required to realize the prescribed P_e (point 1 in Figure 8.4).
  - 2. *Received* E_b/N_0

    Let $(E_b/N_0)_{\text{rec}}$ denote actually received E_b/N_0 (point 2 in Figure 8.4).

$$(E_b/N_0)_{\text{rec}} > (E_b/N_0)_{\text{req}}$$

8.4 Radio Link Analysis

- To assure *reliable* operation of the communication link, the link budget includes a safety measure called the *link margin* (providing protection against change and the unexpected).
- Definition of link margin

$$(\frac{E_b}{N_0})_{rec} = M(\frac{E_b}{N_0})_{req} \qquad (8.1)$$

or

$$M(dB) = (\frac{E_b}{N_0})_{rec}(dB) - (\frac{E_b}{N_0})_{req}(dB) \qquad (8.2)$$

- The larger is the link margin *M*, the more reliable is the communication link, at the cost of a higher E_b/N_0

8.4.1 Free-Space Propagation Model

- The next step in formulating the link budget --- *calculate the received signal power*

  This calculation accounts for all gains and losses incurred in the transmission and reception of the carrier.
- In a radio communication system, the propagation of the modulated signal is accomplished by means of a transmitting antenna.

8.4.1 Free-Space Propagation Model

- The function of the transmitting antenna:
  - 1. Convert the electrical modulated signal into an electromagnetic field.
  - 2. Radiate the electromagnetic energy in desired direction.
- The function of the receiving antenna:
  - 1. Convert the electromagnetic field into an electrical signal from which the modulated signal is extracted.
  - 2. Suppress radiation originating from directions where it is not wanted.

8.4.1 Free-Space Propagation Model

- Receiver is far from the transmitting antenna → the transmitting antenna may be viewed as a fictitious volumeless emitter or point source → specify the variation of power density for the antenna → simplify matters in that the power density of a point source has only a radial component
- By definition, the Poynting vector or power density is the rate of energy flow per unit area(W/m^2).

8.4.1 Free-Space Propagation Model

- “reference” antenna
  - The performance of the transmitting and receiving antenna can be compared against the reference antenna.
  - In practice, assume reference antenna as an isotropic source (radiating uniformly in all directions).

8.4.1 Free-Space Propagation Model

- Consider an isotropic source radiating a total power power P_t(watts). The radiated power passes uniformly through a sphere of surface area $4\pi d^2$, where d is the distance from the source. Hence, the power density, $\rho(d)$, is

$$\rho(d) = \frac{P_t}{4\pi d^2} \quad \text{watts/m}^2 \qquad (8.3)$$

- *Inverse-square law*: the power density varies inversely as the square of the distance from a point source.

8.4.1 Free-Space Propagation Model

- Radiation intensity Φ

$$\Phi = d^2 \rho(d) \qquad \text{(W/sr)} \qquad (8.4)$$

Here, d is the distance at which it is measured. Φ is measured in watts per unit solid angle (watts per steradian).

- In general, we may express the radiation intensity as $\Phi(\theta,\phi)$ (so speak of a *radiation-intensity pattern*), where θ and ϕ is the spherical coordinates defined in Figure 8.5.

8.4.1 Free-Space Propagation Model

- The power radiated inside an infinitesimal solid angle $d\Omega$ is given by $\Phi(\theta,\phi)d\Omega$, where

$$d\Omega = \sin\theta d\theta d\phi \quad \text{steradians} \qquad (8.5)$$

- The total power radiated is

$$P = \int \Phi(\theta,\phi)d\Omega \quad \text{watts} \qquad (8.6)$$

- The average power radiated per unit solid angle is

$$P_{av} = \frac{1}{4\pi}\int \Phi(\theta,\phi)d\Omega = \frac{P}{4\pi} \quad \text{watts/steradian} \qquad (8.7)$$

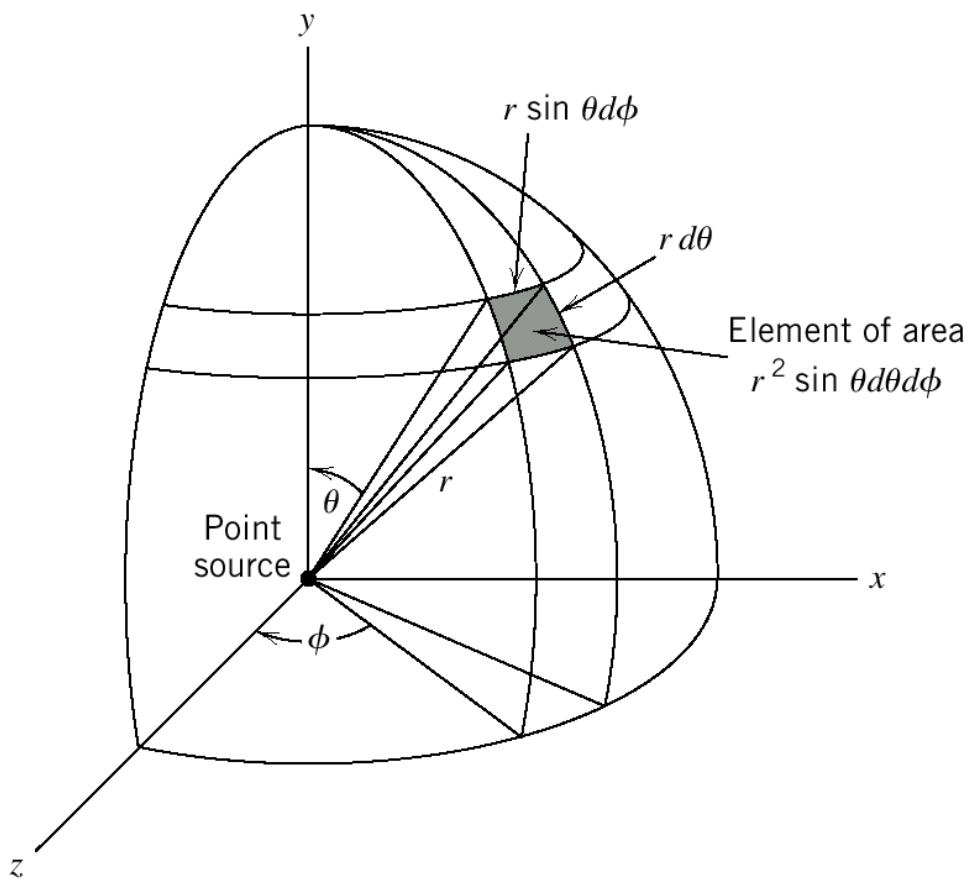


Figure 8.5

Illustrating spherical coordinates of a point source.

8.4.1 Free-Space Propagation Model

- Directive Gain, Directive, and Power Gain

Directive gain or directivity --- the ability of an antenna to concentrate the radiated power in a given direction (*transmitting antenna*), or to absorb the incident power from that direction (*receiving antenna*).

The *directive gain* is defined as

$$g(\theta,\phi)=\frac{\Phi(\theta,\phi)}{P_{av}}=\frac{\Phi(\theta,\phi)}{P/4\pi} \qquad (8.8)$$

8.4.1 Free-Space Propagation Model

- The *directivity* of an antenna -- the maximum value of the directive gain $g(\theta,\phi)$. (*denoted by D*). It is a *constant* that has been maximized for a particular direction.
- The *power gain* of an antenna -- the ratio of the maximum radiation intensity from the antenna to the radiation intensity from a lossless isotropic source , under the constraint that the same input power is applied to both antennas. (*denoted by G*)

8.4.1 Free-Space Propagation Model

- Thus, we have

$$G = \eta_{radiation} D \tag{8.9}$$

  where $\eta_{radiation}$ denote the radiation efficiency factor of the antenna. Obviously, the power gain over a lossless isotropic source equals the directivity if the antenna is 100 percent efficient.

- Henceforth, we assume that the antenna is 100 percent efficient and therefore refer only to the power gain.

8.4.1 Free-Space Propagation Model

- The concept of power gain, which is based on the transmitted power-pattern shape, can be extended to a receiving antenna by virtue of the *reciprocity principle*. For a given antenna structure, the power gains of transmitting and receiving antennas are then identical.
- The power gain of an antenna is the result of concentrating the power density in a restricted region smaller than 4π steradians, as shown in Figure 8.6.

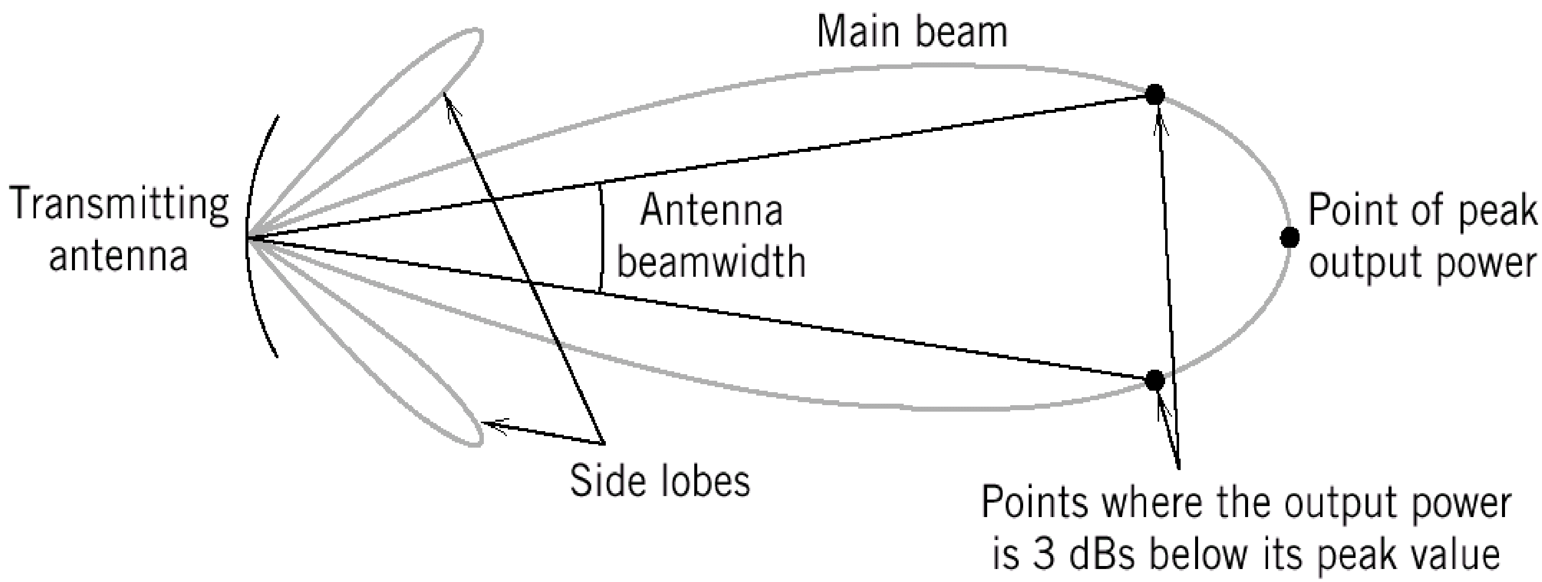


Figure 8.6
Illustrating the concentration of power density of transmitting antenna inside a region smaller than 4π steradians.

8.4.1 Free-Space Propagation Model

- Two parameters in Figure 8.6:
  - 1. *Effective radiated power referenced to an isotropic source* (ERIP) -- the product of P_t (*transmitted power*) and G_t (*power gain of the transmitting antenna*), as shown by

$$ERIP = P_t G_t \quad \text{watts} \qquad (8.10)$$

  - 2. *Antenna beamwidth* -- the angle(in degrees or radians) that subtends the two points on the mainlobe of the field-power pattern at which peak field power is reduced by 3*dBs*. *The higher the power gain of the antenna, the narrower is the antenna beamwidth.*

8.4.1 Free-Space Propagation Model

- Effective Aperture (significant for a receiving antenna) -- *the ratio of the power available at the antenna terminals to the power per unit area of the appropriately polarized incident electromagnetic waves.*

  The effective aperture, denoted by A, is defined as

$$A = \frac{\lambda^2}{4\pi} G \qquad (8.11)$$

where λ is the wavelength of the carrier,

$$\lambda = c/f \qquad (8.12)$$

where c is the speed of light, f is the carrier frequency.

8.4.1 Free-Space Propagation Model

- *The ratio of the antenna's effective aperture to its physical aperture is a direct measure of the antenna's efficiency,* $\eta_{aperture}$, in radiating power to a desired direction or absorbing power from that direction.

- Nominal values for the efficiency $\eta_{aperture}$ of reflector antennas lie in the range of 45 to 75 percent.

8.4.1 Free-Space Propagation Model

- **Friis Free-Space Equation** -- the basic propagation equation for a radio communication link

Consider a *transmitting antenna* with

$$ERIP = P_t G_t \quad \text{watts} \qquad (8.10)$$

and inverse-square law

$$\rho(d) = \frac{P_t}{4\pi d^2} \quad \text{watts/m}^2 \qquad (8.3)$$

we express the power density as

$$EIRP / 4\pi d^2$$

where d is the T-R distance.

8.4.1 Free-Space Propagation Model

- The power P_r absorbed by the receiving antenna is

$$P_r = \left(\frac{EIRP}{4\pi d^2}\right)A_r = \frac{P_t G_t A_r}{4\pi d^2} \qquad (8.13)$$

where A_r is the antenna's effective area of the receiving antenna.

8.4.1 Free-Space Propagation Model

According to reciprocity principle and Equation(8.11), we have

$$A_r = \frac{\lambda^2}{4\pi} G_r$$

where G_r is the power gain of the receiving antenna. Rewrite Equation (8.13), we get the *Friis free-space equation* as

$$\mathrm{P_r} = P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2 \qquad (8.14)$$

8.4.1 Free-Space Propagation Model

- The *path loss*, PL, representing signal “attenuation” in decibels across the entire communication link, is defined as

$$PL = 10\log_{10}\left(\frac{P_t}{\mathrm{P_r}}\right)$$
$$= -10\log_{10}(G_t G_r) - 10\log_{10}\left(\frac{\lambda}{4\pi d}\right)^2 \qquad (8.15)$$

  The minus sign in the first term signifies that it represents a “gain”. The second term is the *free-space loss*, denoted by $L_{free\ space}$.
- Note that the increasing of d will make PL to increase. (→*operate at lower frequency*)

8.4.2 Noise Figure

- To complete the budget link analysis, we need to calculate the average noise power in the received signal. To perform noise analysis at the receiver, we need a convenient measure of noise performance of a linear two-port device. One such measure is furnished by the so-called *noise figure*.
- Consider a linear two-port device connected to a signal source of internal impedance $Z(f) = R(f) + jX(f)$ at the input, as in Figure 8.7.

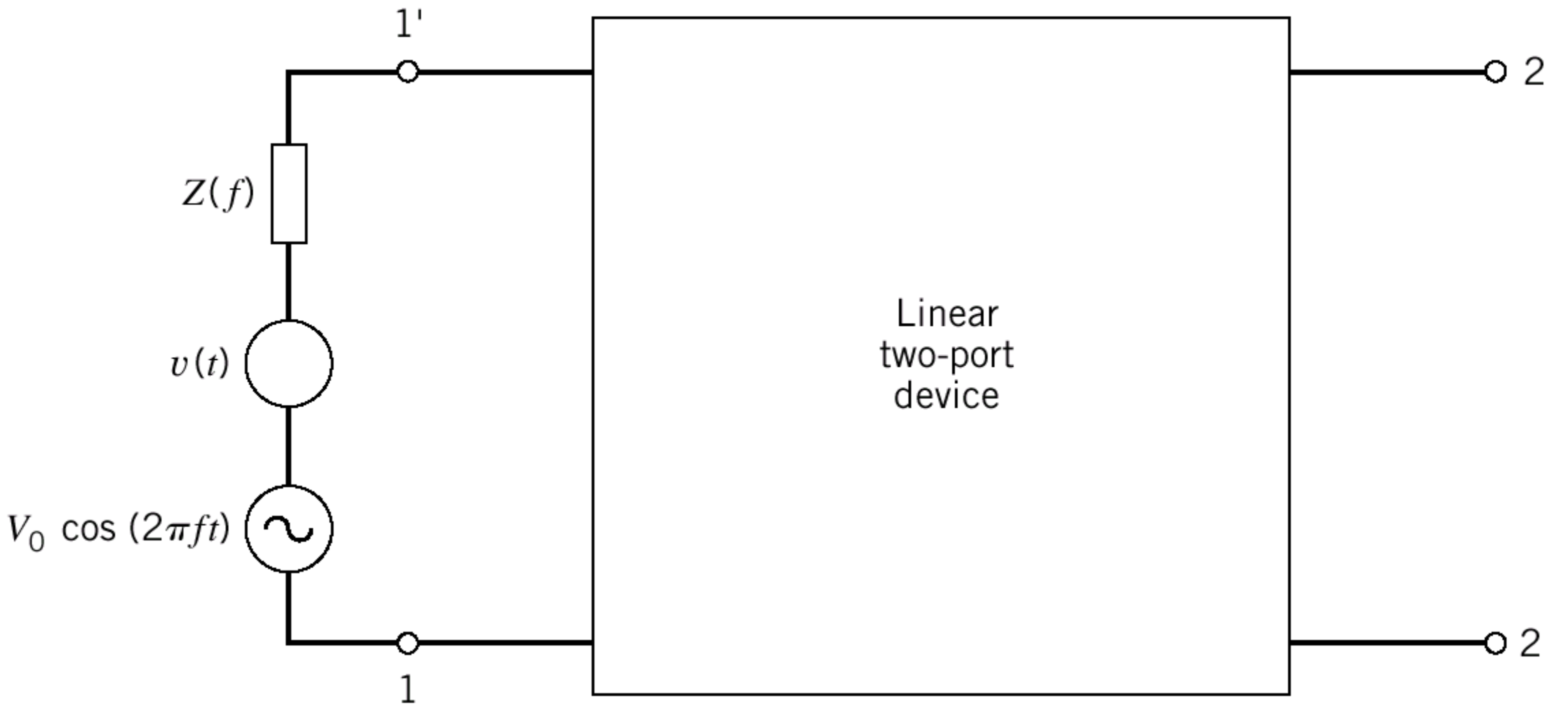


Figure 8.7
Linear two-port device.

8.4.2 Noise Figure

- The output noise of the device consists of two contributions: 1. due to the source, 2. due to the device itself.

- We define the available output noise power in a band of width △f centered at frequency f as the *maximum average noise power* in this band, obtainable at the output of the device.

8.4.2 Noise Figure

- The maximum noise power is obtained when the *load impedance* is the *complex conjugate* of the *output impedance* of the device.
- Definition of the noise figure of the two-port device -- the ratio of the total available output noise power(*due to the device and the source*) per unit bandwidth to the portion thereof due solely to the source.

8.4.2 Noise Figure

- $S_{NO}(f)$ -- the spectral density of the total available noise power of the device output
- $S_{NS}(f)$ - the spectral density of the available noise power due to the source at the device input
- $G(f)$ -- the available power gain of the two-port device -- the ratio of the available signal power at the output of the device to the available signal power of the source when the signal is a sinusoidal wave of frequency f
- We express the noise figure F of the device as

$$F = \frac{S_{NO}(f)}{G(f)S_{NS}(f)} \tag{8.16}$$

8.4.2 Noise Figure

- If the device were noise free, $F = 1$. However, in a physical device, $F > 1$.

- The noise figure may also be expressed in an alternative form.

  $P_S(f)$ -- the available signal power from the source, which is the maximum average power available can be obtained.

8.4.2 Noise Figure

- Assume a source providing a single-frequency signal component with open-circuit voltage $V_0\cos(2\pi ft)$, the load connected to the source is

$$Z^*(f) = R(f) - jX(f)$$

- We have

$$P_S(f) = \left[\frac{V_0}{2R(f)}\right]^2 R(f) = \frac{V_0^2}{4R(f)} \qquad (8.17)$$

8.4.2 Noise Figure

- The available signal power at the output of the device is therefore

$$P_0(f) = G(f)P_S(f) \tag{8.18}$$

- Then we obtain

$$F = \frac{P_S(f)S_{NO}(f)\Delta f}{G(f)P_S(f)S_{NS}(f)\Delta f} = \frac{P_S(f)S_{NO}(f)\Delta f}{P_O(f)S_{NS}(f)\Delta f} = \frac{\rho_S(f)}{\rho_O(f)} \tag{8.19}$$

8.4.2 Noise Figure

- Where

$$\rho_S(f) = \frac{P_S(f)}{S_{NS}(f)\Delta f} \qquad (8.20)$$

$$\rho_o(f) = \frac{P_O(f)}{S_{NO}(f)\Delta f} \qquad (8.21)$$

$\rho_s(f)$ -- the available signal-to-noise ratio of the source

$\rho_o(f)$ -- the available signal-to-noise ratio at the device output

both measured in a narrow band of Δf centered at f

8.4.2 Noise Figure

- *spot noise figure* F -- a function of operating frequency *f*
- *average noise figure* F_0

$$F_0 = \frac{\int_{-\infty}^{\infty} S_{NO}(f)df}{\int_{-\infty}^{\infty} G(f)S_{NS}(f)df} \tag{8.22}$$

- **Disadvantage of noise figure F**

  When it is used to compare low-noise devices, the values of *F* are all close to unity, which makes the comparison rather difficult.

8.4.2 Noise Figure

- **Equivalent Noise Temperature**
- Consider a linear two-port device as shown in Figure 8.8. The available noise power at the device input is

$$N_1 = kT\Delta f \tag{8.23}$$

- N_d -- the noise power contributed by the device

$$N_d = GkT_e\Delta f \tag{8.24}$$

where G -- the available power gain of the device

T_e -- its equivalent noise temperature

T – room temperature, namely 290K (degree Kelvin)

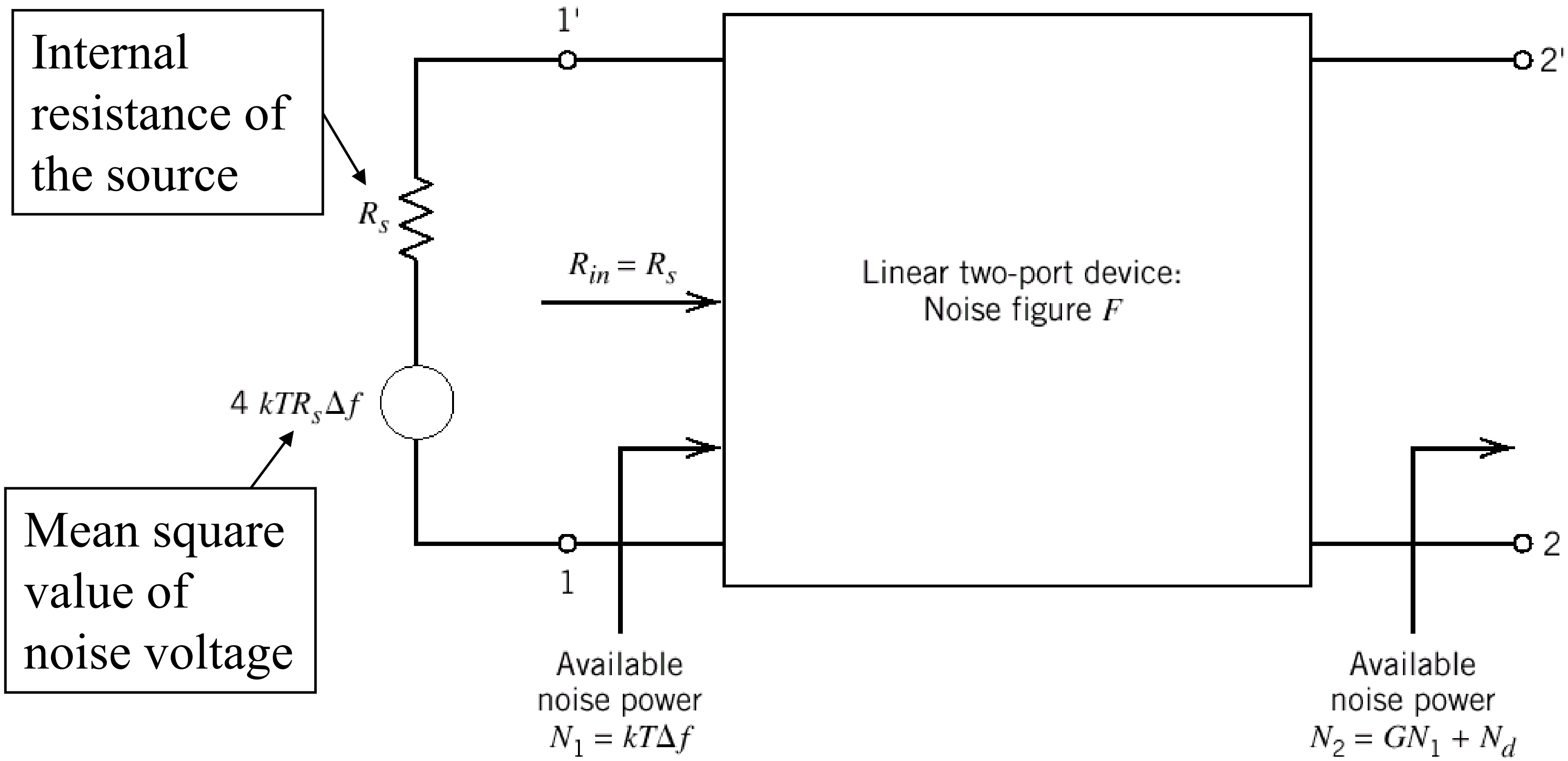


Figure 8.8

Linear two-port device matched to the internal resistance of a source connected to the input.

8.4.2 Noise Figure

- Thus, the total noise power is

$$N_2 = GN_1 + N_d = Gk(T + T_e)\Delta f \qquad (8.25)$$

- The noise figure of the device F is

$$F = \frac{N_2}{N_2 - N_d} = \frac{T + T_e}{T} \qquad (8.26)$$

- Thus, the equivalent noise temperature

$$T_e = T(F - 1) \qquad (8.27)$$

8.4.2 Noise Figure

- **Cascade Connection of Two-port Network**
- How to evaluate the noise figure of a cascade connection of two-port networks whose individual noise figure are known?
- An example is shown in Figure 8.9. It is assumed that the devices are matched, F_2 of the second network is defined assuming an input power N_1.

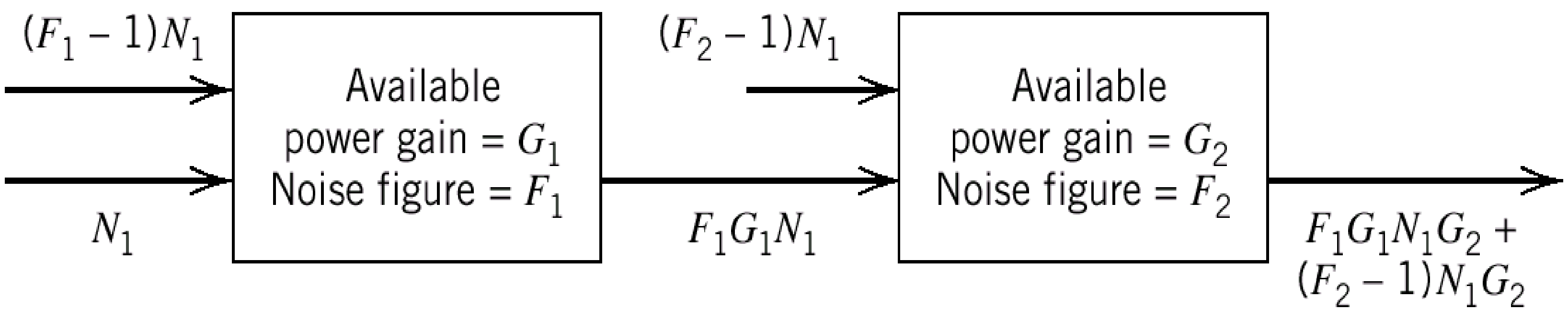


Figure 8.9

A cascade of two noisy tow-port networks.

8.4.2 Noise Figure

- Analyze the networks in Figure 8.9, and according to the definition of F, we may express the overall noise figure of the cascade connection as

$$\begin{aligned} F &= \frac{F_1 G_1 N_1 G_2 + (F_2 - 1) N_1 G_2}{N_1 G_1 G_2} \\ &= F_1 + \frac{F_2 - 1}{G_1} \end{aligned} \qquad (8.28)$$

8.4.2 Noise Figure

- This result may be extended to the cascade connection of any number of two-port networks, as

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} + \cdots \quad (8.29)$$

- It shows that if the first stage of the cascade connection has a high gain, the overall noise figure F is dominated by the noise figure of first stage.

8.4.2 Noise Figure

- Correspondingly, the overall equivalent noise temperature of the cascade connection of any number of two-port network, as

$$T_e = T_1 + \frac{T_2}{G_1} + \frac{T_3}{G_1 G_2} + \frac{T_4}{G_1 G_2 G_3} + \cdots \qquad (8.30)$$

- (8.30) is known as the Friis formula. Again if the gain G_1 of the first stage is high, T_e is dominated by that of the first stage.

8.4.2 Noise Figure

- **Example 8.1 Noise Temperature of Earth-Terminal Receiver** (Figure 8.10)

The equivalent noise temperatures of the components, including the receiving antenna, are

$$T_{antenna} = 50\ \mathrm{K}$$
$$T_{RF} = 50\ \mathrm{K}$$
$$T_{mixer} = 500\ \mathrm{K}$$
$$T_{IF} = 1000\ \mathrm{K}$$

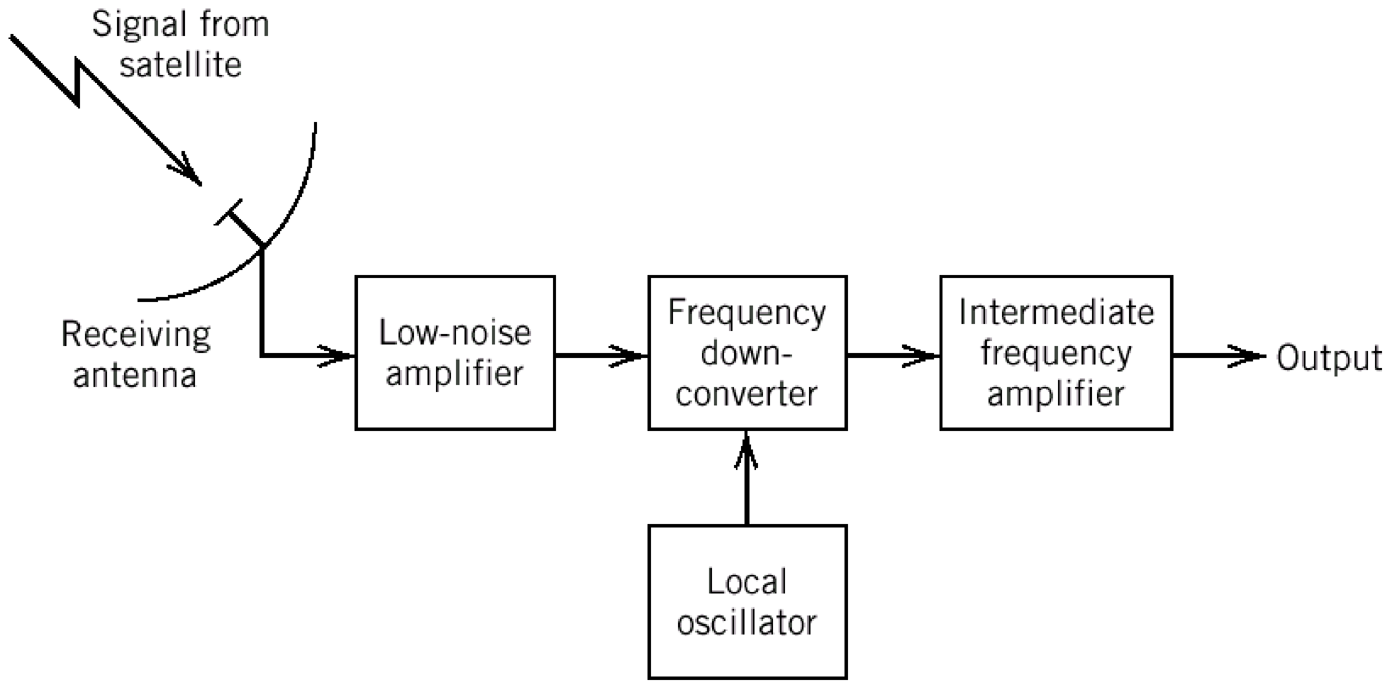


Figure 8.10

Block diagram of earth terminal receiver.

8.4.2 Noise Figure

- The available power gains of the two amplifiers are

$$G_{RF} = 200 = 23 \text{ dB}$$

$$G_{IF} = 1000 = 30 \text{ dB}$$

- Thus, the equivalent noise temperature of the receiver is

$$T_e = T_{antenna} + T_{RF} + \frac{T_{mixer} + T_{IF}}{G_{RF}}$$

$$= 50 + 50 + \frac{500 + 1000}{200} = 107.5K$$

8.4.2 Noise Figure

- **Example 8.2 Downlink Budget Analysis of a Digital Satellite**
- In a digital satellite communication system, the ***downlink power budget*** is usually more critical than the uplink power budget because of the practical constraints imposed on downlink budget and antenna size.
- The example presented here addresses a sample downlink budget analysis, assuming any required uplink power is available for satisfactory operation of the system.

8.4.2 Noise Figure

- The critical parameter to be calculated is the *ratio of received carrier power-to-noise spectral density*, denoted by C/N_0.
- According to Equation (8.14), the average power received at the earth terminal is

$$P_r = P_t G_t G_r \left(\frac{\lambda}{4\pi d}\right)^2$$

- using Equation (1.94), we may express C/N_0 ratio of downlink as

$$\left(\frac{C}{N_0}\right)_{\text{downlink}} = (EIRP)_{satellite}\left(\frac{G_r}{T_e}\right)_{earth\ ter}\left(\frac{\lambda}{4\pi d}\right)^2 \frac{1}{k} \qquad (8.31)$$

8.4.2 Noise Figure

- Express Equation (8.31) in decibels as the sum of gains and loses as itemized here:

  1. $(EIRP)_{satellite}$ measured in dBW;

  2. $(G/T)_{earth\ terminal}$, measured in dB/K, which is short for $(G_r/T_e)_{earth\ terminal}$.

  3. $L_{free\ space}$, denoting $10\log_{10}(4\pi d/\lambda)^2$ in dB.

  4. $-10\log_{10}k$, denoting the gain in dBW/K-Hz due to division by the Boltzmann constant $k = 1.38\times10^{-23}$ joule/K.

8.4.2 Noise Figure

- Table 8.1 shows the values of these four terms for the downlink. Totaling the gains and losses, we thus get

$$(\frac{C}{N_0})_{downlink} = 93.8 dB \cdot HZ$$

- The “received” downlink value of the (C/N_0) ratio may be expressed in terms of $(E_b/N_0)_{\mathrm{req}}$, (see Equation (8.2)),

$$(\frac{C}{N_0})_{downlink} = (\frac{E_b}{N_0})_{req} + 10\log_{10} M + 10\log_{10} R dB \qquad (8.33)$$

8.4.2 Noise Figure

where $10\log_{10} M$ is the link margin in decibels, and R is the data ratio in *b/s*. For operation at the *Ku*-band frequency of 12GHz, we choose a link margin of 6dB. And assume $(E_b/N_0)_{\mathrm{req}} = 12.5$ dB, we get

$$10\log_{10} R = 93.8 - 12.5 - 6$$
$$= 75.3$$

Hence,

$$R = 33.9 \text{ Mb/s}$$

8.4.2 Noise Figure

Assuming the use of coherent 8-PSK for the transmission of digital data via the satellite, and substituting $E_b / N_0 = 12.5dB$ in Equ.(6.47), we find that the probability of symbol error $P_e = 0.6 \times 10^{-3}$

It means data transmission on the downlink at rate R = 33.9 Mb/s and with a probability of symbol error $P_e = 0.6 \times 10^{-3}$ assuming the use of 8-PSK.

8.5 Wireless Communication

- The second type of multiuser radio communication system -- wireless communication -- synonymous with mobile radio (*indoor or outdoor forms of wireless communications 、transmitter or receiver able to move*)

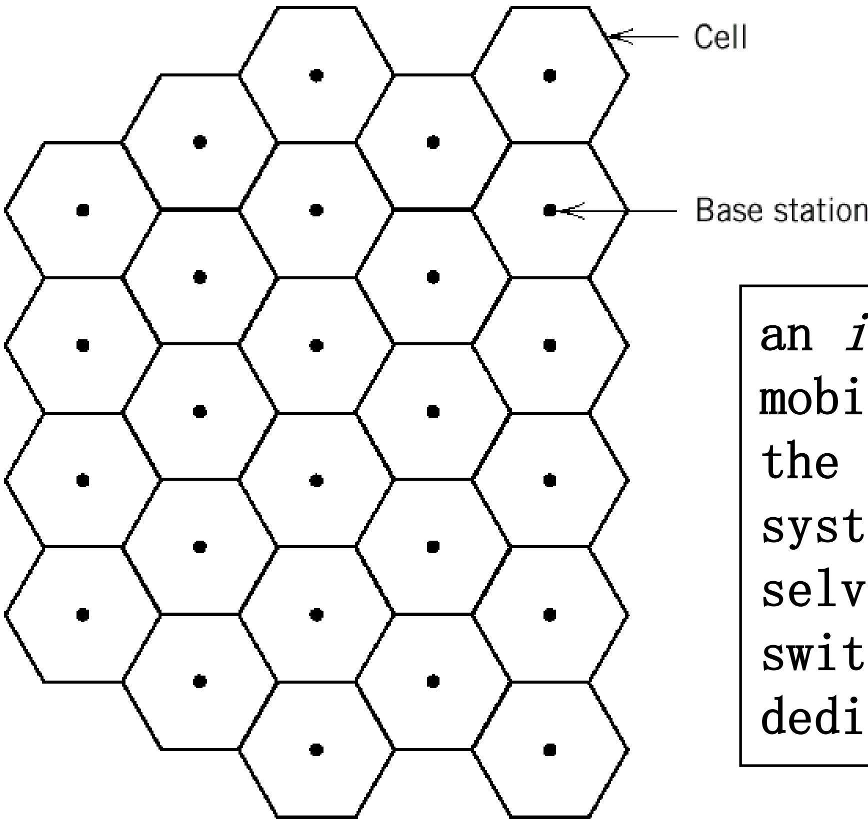
- stochastic nature of the mobile radio channel → practical measurement & statistical analysis

8.5 Wireless Communication

- The aim of the evaluation is to quantify two factors of primary concern:
  - *median signal strength* -- it is used to predict the minimum power needed for the transmitter so as to provide an acceptable quality of coverage
  - *signal variability* -- the fading nature of the channel

8.5 Wireless Communication

- cellular radio -- able to build mobility into the telephone network (user moves freely within a service area & communicate with any telephone subscriber in the world)
- An idealized model of the cellular radio system (Figure 8.11), consists of an array of hexagonal cells with a base station located at the center of each cell.
- A typical cell has a radius of 1 to 12 miles. (1 mile = 1609.3m)



an *interface* between mobile subscribers and the cellular radio system, and are themselves connected to a switching center by dedicated wirelines.

Figure 8.11
Idealized model of cellular radio.

8.5 Wireless Communication

- Two important roles of the mobile switching center:
  - 1. It acts as the interface between the cellular radio system and the public switched telephone network.
  - 2. It performs overall supervision and control of the mobile communications by monitoring the signal-to-noise ratio of a call in progress.
- *handover or handoff* (switching process) -- when the SNR falls below a prescribed threshold, the mobile subscriber is switched to another base station.

8.5 Wireless Communication

- The cellular concept relies on two features:
  - 1. *Frequency reuse* -- the use of the radio channels on the same carrier frequency to cover different areas, which are physically separated from each other sufficiently to ensure that co-channel interference is not objectionable.

    Advantages:

    1. Keep the transmitted power from each base station to a minimum.

    2. Position the antennas of the base stations just high enough to provide for the area coverage of the respective cells.

8.5 Wireless Communication

- 2. *Cell splitting* -- when the demand for service exceeds the number of channels allocated to a particular cell, cell splitting is used to handle the additional growth in traffic within that particular cell. → revision of cell boundaries → smaller cells (*microcells*) → transmitter power & antenna height of the new base stations are reduced + the same set of frequencies are reused according to a new plan

8.5 Wireless Communication

- We may exploit the basic properties of hexagonal cellular geometry to lay out a radio channel assignment plan that determines which channel set should be assigned to which cell.

  We begin with two integers i and j ($i \geqslant j$), called shift parameters, which are predetermined in some manner. Starting with any cell as a reference, we find the nearest *co-channel cell* as follows:

8.5 Wireless Communication

- Move i cells along any chain of hexagons, turn counterclockwise 60°, and move j cells along the chain that lies on this new direction. The jth cells so located and the reference cell constitute the set of co-channel cells. Repeat this procedure for a different reference cell, until all the cells are covered.

Figure 8.12 shows the procedure for a single reference cell with $i = 2$ and $j = 2$.

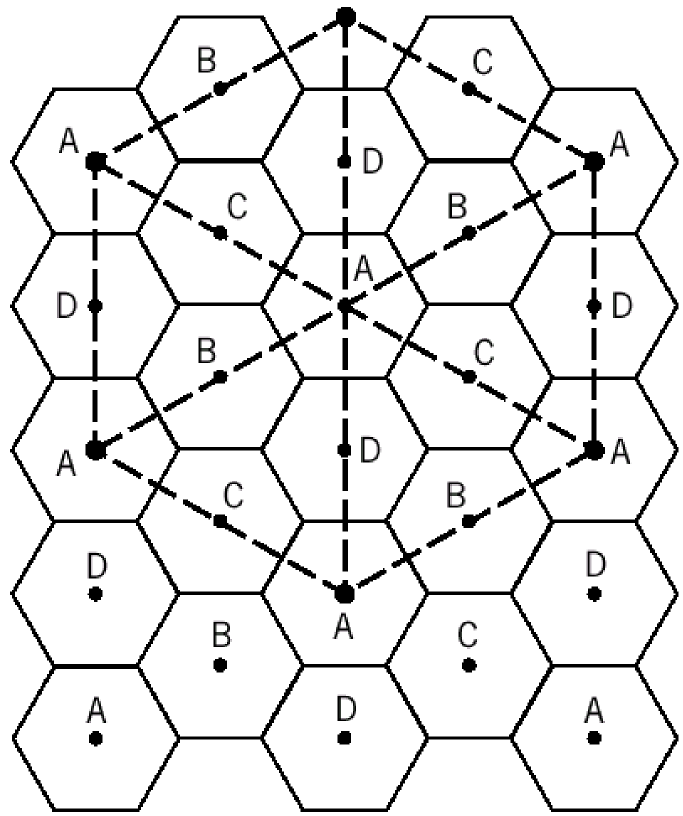


Figure 8.12
Illustrating the determination of co-channel cells.

8.5 Wireless Communication

- In North America

  band assigned to the cellular system :
  800 – 900 MHz

  reverse link(mobile to base station): 824 – 849 MHz

  forward link(base station to mobile): 869 – 894 MHz

- In Europe and elsewhere

  the base-mobile and mobile-base subbands are reversed

8.5 Wireless Communication

- Propagation Effects

The propagation problems encountered in the use of cellular radio

the antenna of a mobile unit may lie well below the surrounding building s in built-up areas → no “line-of-sight” path to the base station → radio propagation takes place mainly by *scattering* from the surfaces of the surrounding buildings and by d*iffraction* over and/or around them, as shown in Figure 8.13.

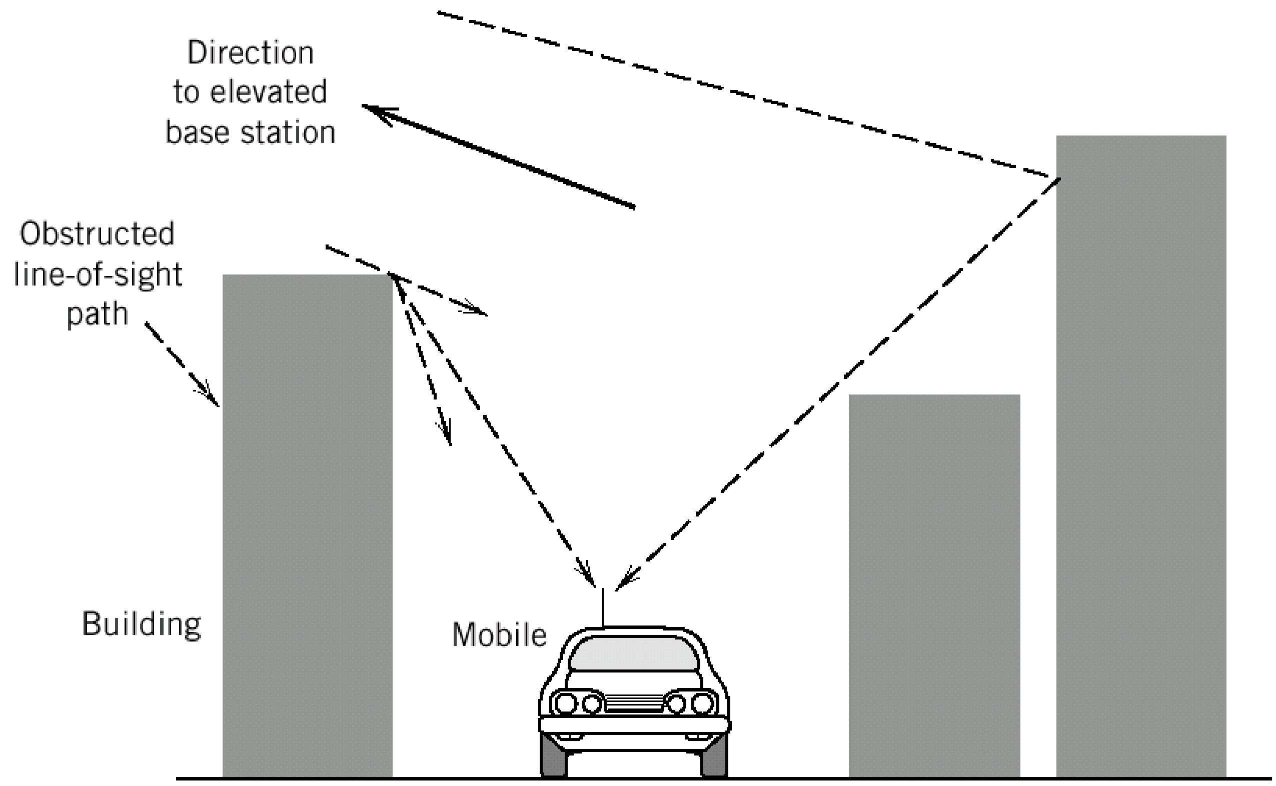


Figure 8.13

Illustrating the mechanism of radio propagation in urban areas. (From Parsons, 1992, with permission.)

8.5 Wireless Communication

- *multipath phenomenon* -- energy reaches the receiving antenna via more than one path (*from different directions and with different time delays*)
- To understand the nature of multipath phenomenon, first consider a “static” multipath environment (*the amplitude of the received signal does not vary with time*) involving a stationary receiver and a narrowband transmitted signal.

8.5 Wireless Communication

- Assume: 1. two attenuated versions of the transmitted signal arrive at the receiver 2. a relative phase shift between the two components of the received signal
- Extreme cases:

  1. The relative phase shift is 0 → *two components add constructively* (Fig. 8.14a)

  2. The relative phase shift is 180 → *two components add destructively* (Fig. 8.14b)

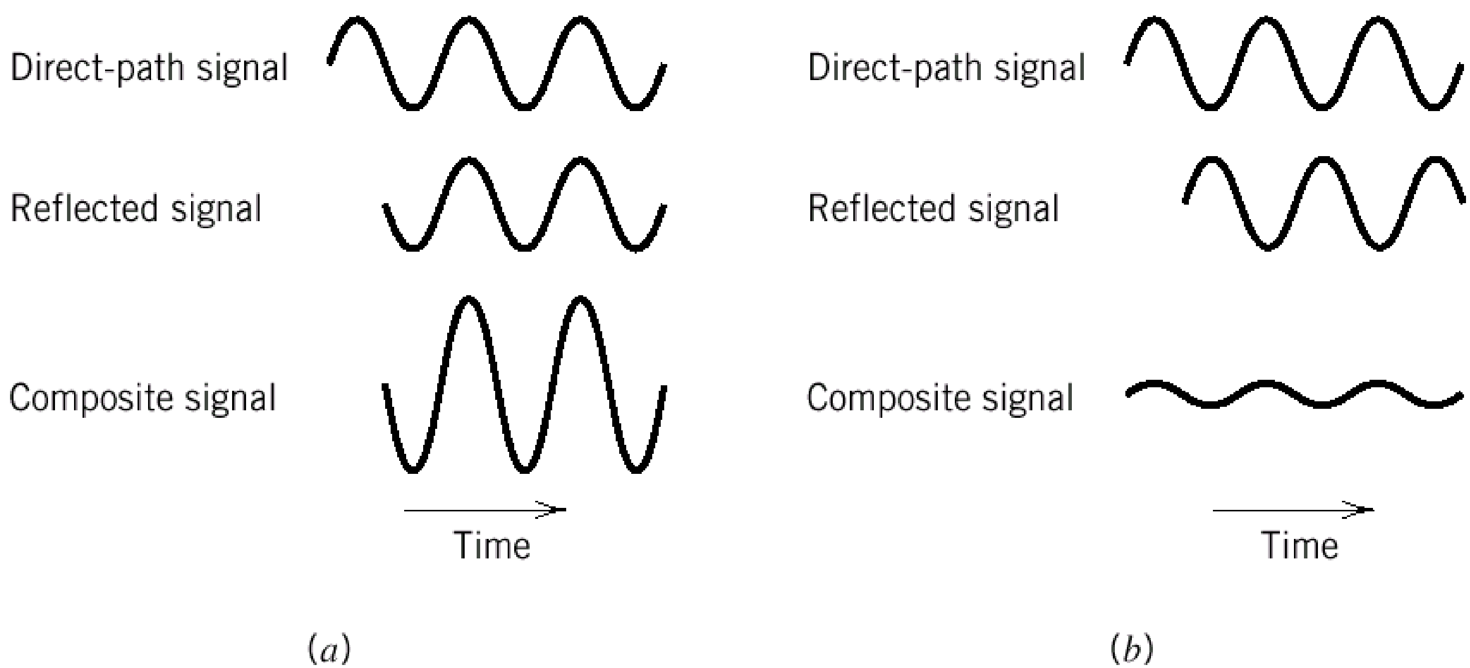


Figure 8.14

(*a*) Constructive and (*b*) destructive forms of the multipath phenomenon for sinusoidal signals.

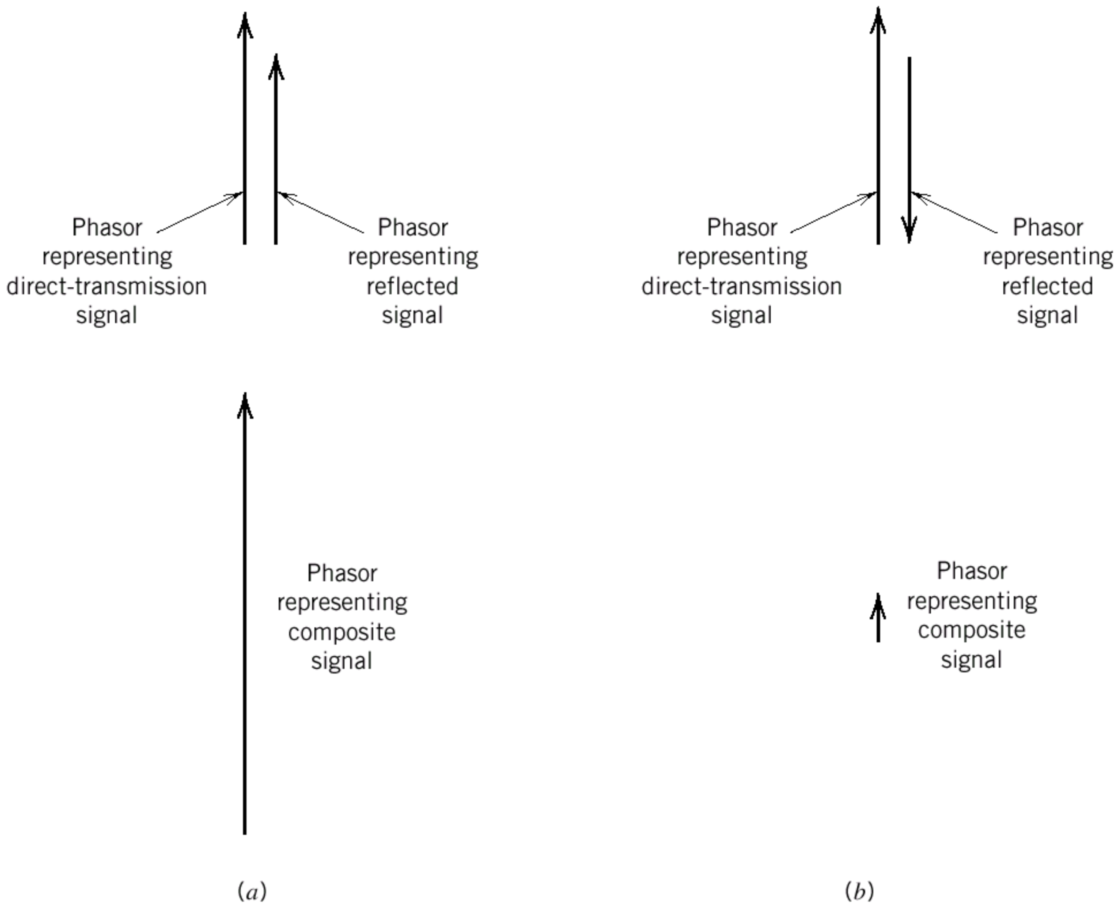


Figure 8.15 Phasor representations of (*a*) constructive and (*b*) destructive forms of multipath.

8.5 Wireless Communication

- Next, consider a “dynamic” multipath environment where the receiver is in motion and two versions of the transmitted narrowband signal reach the receiver.

  Motion of the receiver → continuous change in the length of each propagation path → the relative phase between the two components of the received signal is a function of spatial location of the receiver (Figure 8.16)

the received amplitude varies with distance

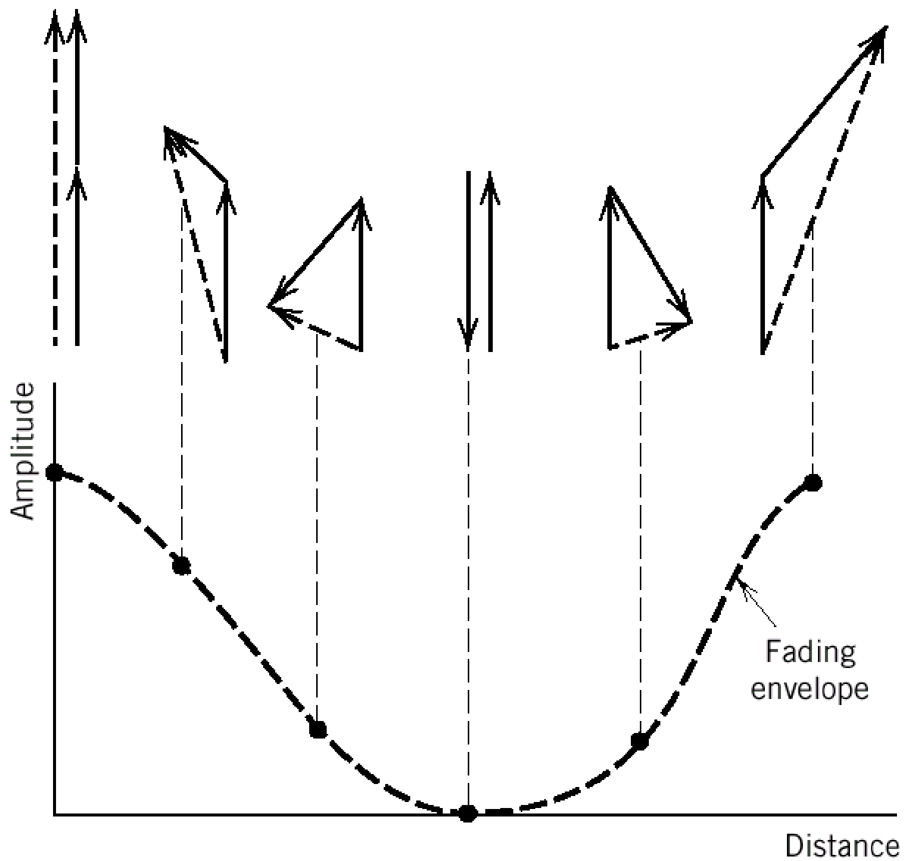


Figure 8.16
Illustrating how the envelope fades as two incoming signals combine with different phases. (From Parsons, 1992, with permission.)

8.5 Wireless Communication

- Figure 8.16 shows there is constructive addition at some locations, and almost complete cancellation at some other locations. This phenomenon is *signal fading*.
- In a practical mobile radio environment, there may be a multitude propagation paths with different lengths, and many components of the received signal could combine in variety of ways → the envelope of the received signal varies with location in a complicated fashion (Figure 8.17)

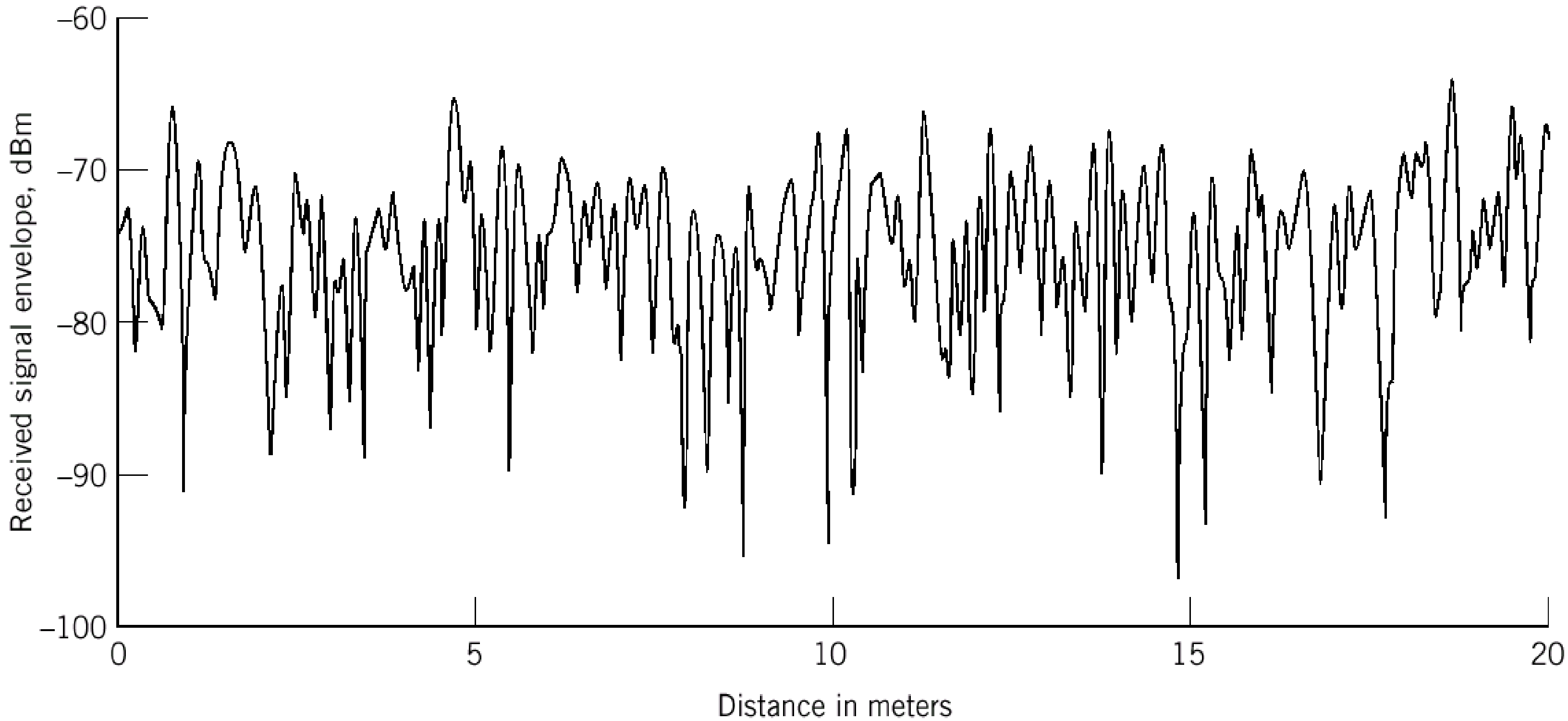


Figure 8.17

Experimental record of received signal envelope in an urban area. (From Parsons, 1992, with permission.)

8.5 Wireless Communication

- Signal fading -- a spatial phenomenon that manifests itself in the time domain as the receiver moves.

- Consider the situation shows in Figure 8.18. Assume: 1. the receiver is moving along the line AA’ with a constant velocity *v*. 2. the received signal is due to a radio wave from a scatter labeled *S*.

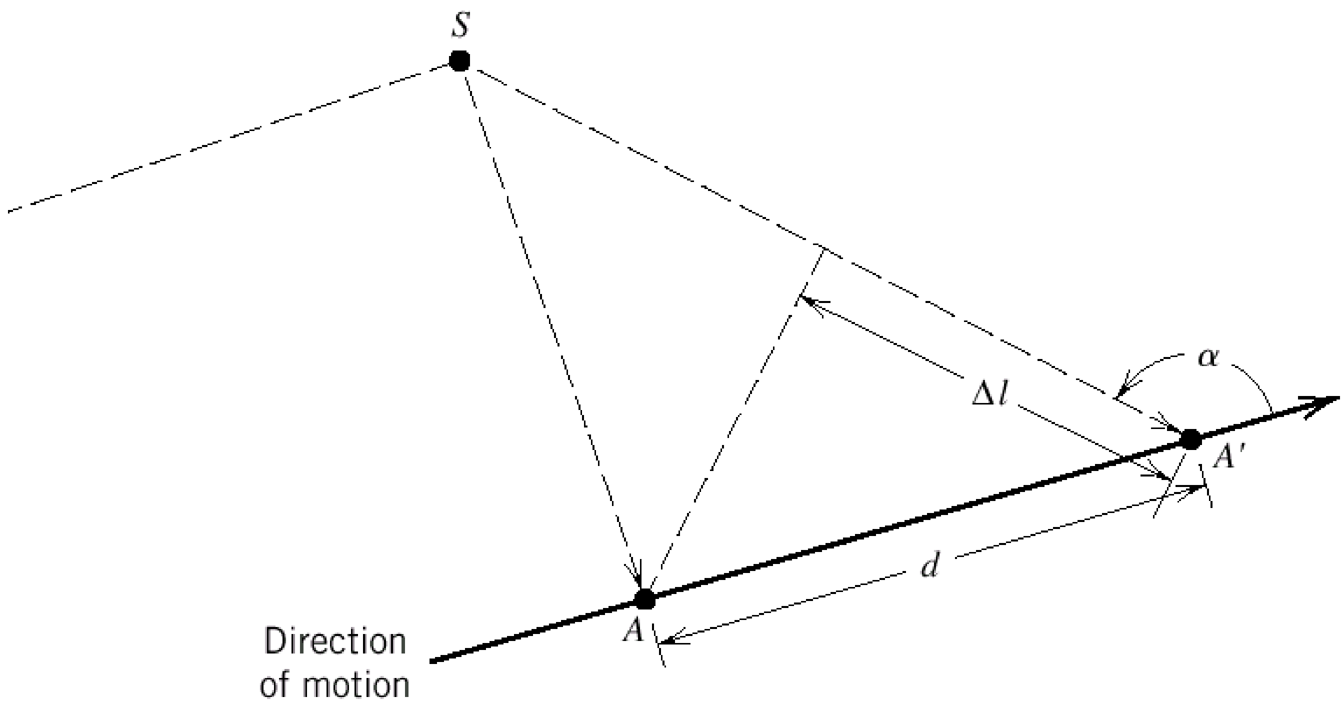


Figure 8.18
Illustrating the calculation of Doppler shift.

8.5 Wireless Communication

- Let $\triangle t$ denote the time taken for the receiver moving from point A to A', then we get incremental change in the path length

$$\triangle l = d\cos a$$
$$= v\triangle t\cos a \qquad (8.34)$$

where a is the spatial angle between the incoming radio wave and the direction of motion of the receiver.

8.5 Wireless Communication

Correspondingly, the change in the phase angle of the received signal at point A’ with respect to that at point A is given by

$$\Delta\phi = \frac{2\pi}{\lambda}\Delta l = \frac{2\pi v\Delta t}{\lambda}\cos\alpha \qquad (8.35)$$

where λ is the radio wavelength.

8.5 Wireless Communication

The apparent change in frequency, or the *Doppler-shift*, is therefore

$$f_d = \frac{1}{2\pi}\frac{\Delta\phi}{\Delta t} = \frac{v}{\lambda}\cos a \qquad (8.36)$$

Thus, the Doppler-shift f_d is positive when the radio waves arrive from ahead of the mobile unit, and negative when the radio waves arrive from behind the mobile unit.

8. 6 Statistical Characterization of Multipath Channels

- In section 8.5, the signal bandwidth $\ll$ the reciprocal of the spread in propagation path delays → *narrowband* characterization of the multipath environment (2 effects: rapid fading of the received signal *envelope* and a spread in Doppler shifts in the received *spectrum*)
- However, real-life signals may occupy a bandwidth wide enough to require more detailed considerations of the effects of multipath propagation on the received signal.

8. 6 Statistical Characterization of Multipath Channels

- According to the complex notation described in Appendix 2, we express the transmitted passband signal as

$$s(t) = \mathrm{Re}[\tilde{s}(t)\exp(j2\pi f_c t)] \qquad (8.37)$$

where $\tilde{s}(t)$ is the complex (low-pass) envelope $s(t)$, and f_c is a nominal carrier frequency.

8.6 Statistical Characterization of Multipath Channels

- multipath effects → the impulse response of the channel is delay dependent and a time-varying function. It is

$$h(\tau;t)=\mathrm{Re}[\tilde{h}(\tau;t)\exp(j2\pi f_c t)] \qquad (8.38)$$

where $\tilde{h}(\tau;t)$ is the (low-pass) complex impulse response of the channel, (called the input delay-spread function of the channel), and τ is a delay variable.

8. 6 Statistical Characterization of Multipath Channels

- Thus, the (low-pass) complex envelope of the channel output is defined by the convolution integral

$$\tilde{s}_o(t) = \frac{1}{2}\int_{-\infty}^{\infty} \tilde{s}(t-\tau)\tilde{h}(\tau;t)d\tau \tag{8.39}$$

where the factor ½ is the result of using complex notation.

8. 6 Statistical Characterization of Multipath Channels

- In general, the behavior of a mobile radio channel can be described only in statistical terms.
- For analytic purpose, consider *Rayleigh fading channel*.

  $\tilde{h}(\tau;t)$ -- a *zero-mean complex-valued Gaussian process*

  envelope $\left|\tilde{h}(\tau;t)\right|$ -- *Rayleigh* distributed

8. 6 Statistical Characterization of Multipath Channels

- The time-varying transfer function of the channel is defined as the Fourier transform of $\tilde{h}(\tau;t)$, as shown by

$$\tilde{H}(f;t)=\int_{-\infty}^{\infty}\tilde{h}(\tau;t)\exp(-j2\pi\tau)d\tau \qquad (8.40)$$

where f denote the frequency variable. This time-varying transfer function may be viewed as frequency transmission characteristic of the channel.

8. 6 Statistical Characterization of Multipath Channels

- For a statistical characterization of the channel, we make two assumptions:
  - The input delay-spread function $\tilde{h}(\tau;t)$ is a *zero-mean, complex-valued Gaussian* process. For short-term fading, assume $\tilde{h}(\tau;t)$ is also *stationary*.
  - The channel is an *uncorrelated scattering channel*.

8. 6 Statistical Characterization of Multipath Channels

- Autocorrelation function definition

$$R_{\tilde{h}}(\tau_1, t_1; \tau_2, t_2) = E[\tilde{h}^*(\tau_1; t_1)\tilde{h}(\tau_2; t_2)] \qquad (8.41)$$

where E is the statistical expectation operator, the asterisk denotes complex conjugation, t_1 and t_2 are the propagation of the two paths involved in the calculation, τ_1 and τ_2 are the times at which the outputs of the two paths are observed.

8. 6 Statistical Characterization of Multipath Channels

- time variable t -- stationary
- time-delay variable τ --uncorrelated scattering

we may rewrite Equation (8. 41) as

$$R_{\tilde{h}}(\tau_1,\tau_2;\Delta t) = E[\tilde{h}^*(\tau_1;t)\tilde{h}(\tau_2;t+\Delta t)]$$

$$= r_{\tilde{h}}(\tau_1;\Delta t)\delta(\tau_1-\tau_2) \qquad (8.42)$$

where Δt is the difference between the observation times, and $\delta(\tau_1-\tau_2)$ is a delta function.

8. 6 Statistical Characterization of Multipath Channels

- Using τ in place of τ_1, the remaining function in Equation (8.42) is defined as

$$r_{\tilde{h}}(\tau;\Delta t) = E[\tilde{h}(\tau;t)\tilde{h}^*(\tau;t+\Delta t)] \tag{8.43}$$

The function $r_{\tilde{h}}(\tau;\Delta t)$ is called the *multipath autocorrelation profile* of the channel.

Back

8. 6 Statistical Characterization of Multipath Channels

- The autocorrelation function of $\tilde{H}(f;t)$ is defined as

$$R_{\tilde{H}}(f_1,t_1;f_2,t_2)=E[\tilde{H}^*(f_1;t_1)\tilde{H}(f_2;t_2)] \qquad (8.44)$$

where f_1 and f_2 represent two frequencies in the spectrum of a transmitted signal. The autocorrelation function $R_{\tilde{H}}(f_1,t_1;f_2,t_2)$ provides a statistical measure of the extent to which the signal is distorted bv transmission through the channel.

8.6 Statistical Characterization of Multipath Channels

- From Equations (8.40), (8.41), (8.44), we find the autocorrelation function $R_{\tilde{H}}(f_1,t_1;f_2,t_2)$ and $R_{\tilde{h}}(\tau_1,t_1;\tau_2,t_2)$ are related by a form of 2-dimensional Fourier transformation as follows

$$R_{\tilde{H}}(f_1,t_1;f_2,t_2)$$

$$=\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} R_{\tilde{h}}(\tau_1,t_1;\tau_2,t_2)\exp[j2\pi(f_1\tau_1-f_2\tau_2)]d\tau_1 d\tau_2 \tag{8.45}$$

8. 6 Statistical Characterization of Multipath Channels

- Invoking stationarity in the time domain, we reformulate Equation (8.44) as

$$R_{\tilde{H}}(f_1, f_2; \Delta t) = E[\tilde{H}^*(f_1; t)\tilde{H}(f_2; t + \Delta t)] \tag{8.46}$$

- This definition means $R_{\tilde{H}}(f_1, f_2; \Delta t)$ may be measured by pairs of spaced tones to carry out cross-correlation measurements on the resulting channel outputs. *Such a measurement presumes stationarity in the time domain.*

8. 6 Statistical Characterization of Multipath Channels

- If we also assume *stationarity in the frequency domain*, we may write

$$R_{\tilde{H}}(f, f+\Delta f; \Delta t) = r_{\tilde{H}}(\Delta f; \Delta t) = E[\tilde{H}^*(f;t)\tilde{H}(f+\Delta f; t+\Delta t)] \quad (8.47)$$

This form of $\tilde{H}(f;t)$ is in fact the Fourier transform of the multipath auto-correlation profile $r_{\tilde{h}}(\tau;\Delta t)$ with respect to τ, as

$$r_{\tilde{H}}(\Delta f; \Delta t) = \int_{-\infty}^{\infty} r_{\tilde{h}}(\tau;\Delta t)\exp(-2j\pi\tau\Delta f)d\tau \quad (8.48)$$

The function $r_{\tilde{H}}(\Delta f;\Delta t)$ is called the *spaced-frequency spaced-time correlation function* of the channel.

8. 6 Statistical Characterization of Multipath Channels

- Finally, we introduce a function $S(\tau;v)$ (called the *scattering function* of the channel) that forms a Fourier-transform pair with the multipath autocorrelation profile $r_{\tilde{h}}(\tau;\Delta t)$ with respect to Δt , as

$$S(\tau;v)=\int_{-\infty}^{\infty} r_{\tilde{h}}(\tau;\Delta t)\exp(-2j\pi v\Delta t)d(\Delta t) \qquad (8.49)$$

and

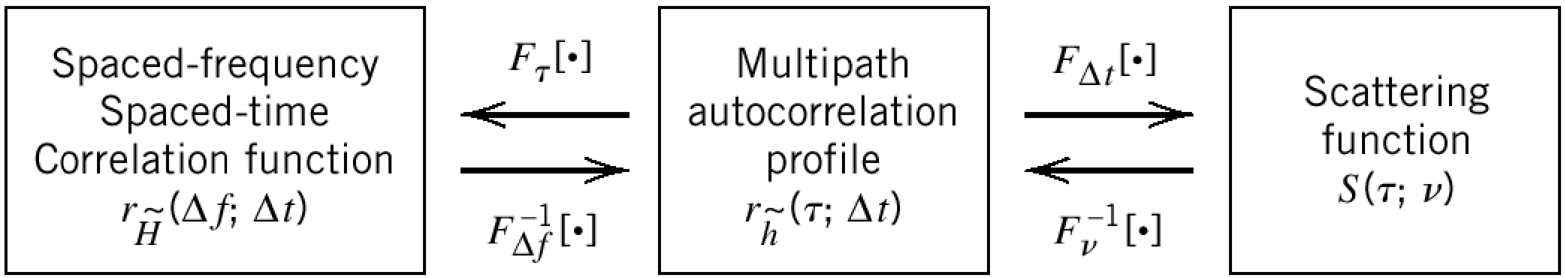
$$r_{\tilde{h}}(\tau;\Delta t)=\int_{-\infty}^{\infty} S(\tau;v)\exp(2j\pi v\Delta t)dv \qquad (8.50)$$

8. 6 Statistical Characterization of Multipath Channels

- The function $S(\tau;v)$ may also be defined in terms of $r_{\tilde{H}}(\Delta f;\Delta t)$ by applying a form of double Fourier transformation: a Fourier transform with respect to the time variable Δt and an inverse Fourier transform with respect to the frequency variable Δf , as

$$S(\tau;v)=\int_{-\infty}^{\infty}\int_{-\infty}^{\infty} r_{\tilde{H}}(\Delta f;\Delta t)\exp(-j2\pi v\Delta t)\exp(-2j\pi\tau\Delta f)d\Delta t d\Delta f \tag{8.51}$$

Figure 8.19 displays these relationships in terms of the Fourier transform and its inverse.



$F_{\tau}[\bullet]$: Fourier transform with respect to delay τ

$F_{\Delta f}^{-1}[\bullet]$: Inverse Fourier transform with respect to frequency increment Δf

$F_{\Delta t}[\bullet]$: Fourier transform with respect to time increment Δt

$F_{\nu}^{-1}[\bullet]$: Inverse Fourier transform with respect to Doppler shift ν

Figure 8.19

Functional relationships between the multipath autocorrelation profile $r_{\tilde{h}}(\tau;\Delta t)$, the spaced-frequency spaced-time correlation function $r_{\tilde{H}}(\Delta f;\Delta t)$, and the scattering function $S(\tau, \nu)$.

8. 6 Statistical Characterization of Multipath Channels

- For a physical interpretation of $S(\tau;v)$, consider the transmission of a single tone of frequency f ’(relative to the carrier). The complex envelope of the resulting filter output is

$$\tilde{s}_o(t) = \exp(j2\pi f' t)\tilde{H}(f';t) \qquad (8.52)$$

8. 6 Statistical Characterization of Multipath Channels

The autocorrelation function of $\tilde{s}_o(t)$ is

$$E[\tilde{s}_O^*(t)\tilde{s}_O(t+\Delta t)] = \exp(j2\pi f^{'}\Delta t)E[\tilde{H}^*(f^{'};t)\tilde{H}(f^{'};t+\Delta t)]$$

$$= \exp(j2\pi f^{'}\Delta t)r_{\tilde{H}}(0;\Delta t) \qquad (8.53)$$

where, in the last line, we made use of Equation (8.47).

8. 6 Statistical Characterization of Multipath Channels

- Putting $\Delta f = 0$ in Equation (8.48), and using Equation (8.50), we may write

$$r_{\tilde{H}}(0;\Delta t) = \int_{-\infty}^{\infty} r_{\tilde{h}}(\tau;\Delta t)d\tau$$

$$= \int_{-\infty}^{\infty}[\int_{-\infty}^{\infty} S(\tau;v)d\tau]\exp(j2\pi v\Delta t)dv \qquad (8.54)$$

8. 6 Statistical Characterization of Multipath Channels

- Hence, we may view the integral

$$\int_{-\infty}^{\infty} S(\tau;v)d\tau$$

as the power spectral density of the channel output relative to the frequency *f* ’ of the transmitted tone, and with the Doppler shift v acting as the frequency variable. So we may state that the scattering function provides a statistical measure of the *output power of the channel*.

8.6 Statistical Characterization of Multipath Channels

- Delay Spread

Putting $\Delta t = 0$ in Equation (8.43), we have

$$
\begin{aligned}
P_{\tilde{h}}(\tau) &= r_{\tilde{h}}(\tau;0) \\
&= E[|\tilde{h}(\tau;t)|^2]
\end{aligned}
\tag{8.55}
$$

The function $P_{\tilde{h}}(\tau)$, which is called the *delay power spectrum* or the *multipath intensity profile* of the channel, describes the intensity (averaged over the fading fluctuations) of the scattering process at propagation delay τ.

8. 6 Statistical Characterization of Multipath Channels

- The function $P_{\tilde{h}}(\tau)$ may also be defined in terms of the scattering function S($\tau; v$) by averaging it over all Doppler shifts. We know

$$r_{\tilde{h}}(\tau;\Delta t)=\int_{-\infty}^{\infty}S(\tau;v)\exp(2j\pi v\Delta t)dv$$

Putting $\Delta t=0$, we get

$$P_{\tilde{h}}(\tau)=\int_{-\infty}^{\infty}S(\tau;v)dv \tag{8.56}$$

8. 6 Statistical Characterization of Multipath Channels

- Figure 8.20 shows an example of a delay power spectrum that depicts a typical plot of the power spectral density versus excess delay, which is measured with respect to the time delay for the shortest echo path. The power is measured in dBm as in Figure 8.17.
- The “threshold level” included in Figure 8.20 defines the power level below which the receiver fails to operate satisfactorily.

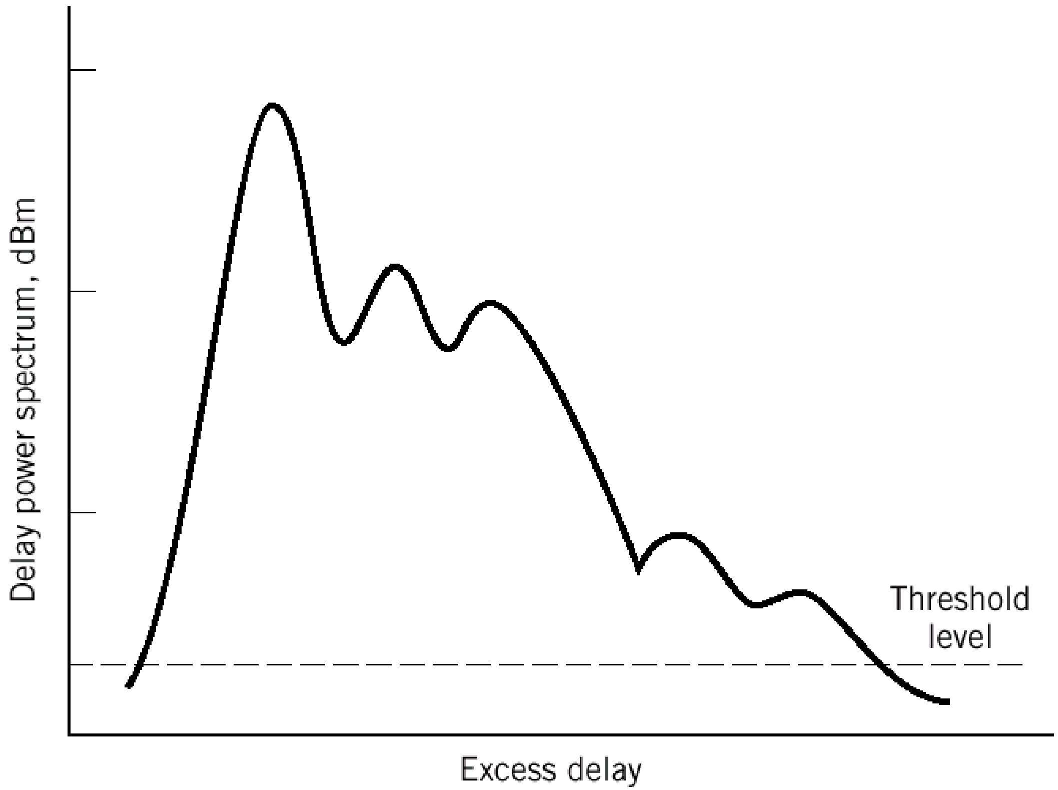


Figure 8.20

Example of a power-delay profile for a mobile radio channel. (From Parsons, 1992, with permission.)

8. 6 Statistical Characterization of Multipath Channels

- Two statistical moment of $P_{\tilde{h}}(\tau)$ of interest:
- 1. average delay τ_{av} -- defined as the first central moment of $P_{\tilde{h}}(\tau)$, as

$$\tau_{av} = \frac{\int_0^{\infty} \tau P_{\tilde{h}}(\tau) d\tau}{\int_0^{\infty} P_{\tilde{h}}(\tau) d\tau} \qquad (8.57)$$

8. 6 Statistical Characterization of Multipath Channels

- 2. delay spread σ_τ -- defined as the square root of the second central moment of $P_{\tilde{h}}(\tau)$

$$\sigma_\tau = \left(\frac{\int_0^\infty (\tau - \tau_{av})^2 P_{\tilde{h}}(\tau) d\tau}{\int_0^\infty P_{\tilde{h}}(\tau) d\tau} \right)^{1/2} \qquad (8.58)$$

- The reciprocal of σ_τ is a measure of the *coherent bandwidth* of the channel, denoted by B_c .

8. 6 Statistical Characterization of Multipath Channels

- Doppler Spread (relating the Doppler effects to time variations of the channel)

Set $\Delta f = 0$ (*a single tone*) in Equation (8.48), we have $r_{\tilde{H}}(0;\Delta t)$. Hence, evaluating the Fourier transform of this function w.r.t. Δt as

$$S_{\tilde{H}}(v) = \int_{-\infty}^{\infty} r_{\tilde{H}}(0;\Delta t)\exp(-j2\pi v\Delta t)d(\Delta t) \quad (8.59)$$

8. 6 Statistical Characterization of Multipath Channels

- The function $S_{\tilde{H}}(v)$ defines the power spectrum of the channel output expressed as a function of the Doppler shift v; thus, it is called the *Doppler spectrum* of the channel.

  Using Equation (8.48) & (8.49), it may also be defined in terms of the scattering function by averaging it over all possible propagation delays, as

$$S_{\tilde{H}}(v) = \int_{-\infty}^{\infty} S(\tau; v) d\tau \tag{8.60}$$

8. 6 Statistical Characterization of Multipath Channels

- The square root of the second moment of the Doppler spectrum is thus defined by

$$\sigma_v = \left(\frac{\int_{-\infty}^{\infty} v^2 S_{\tilde{H}}(v)dv}{\int_{-\infty}^{\infty} S_{\tilde{H}}(v)dv} \right)^{1/2} \qquad (8.61)$$

σ_v provides a measure of the width of the Doppler spectrum; thus, it is called the *Doppler spread* of the channel. And the reciprocal of it is called the *coherence time* of the channel, denoted by τ_c.

8. 6 Statistical Characterization of Multipath Channels

- Another useful parameter is the *fade rate* of the channel, providing a measure of the rapidity of fading of the channel. For a Raleigh fading channel, the average fade rate is related to the Doppler spread as

$$f_e = 1.475\sigma_v \text{ crossings per second} \qquad (8.62)$$

8. 6 Statistical Characterization of Multipath Channels

- Some typical values encountered in a mobile radio environment are:
  - The delay spread, σ_τ , amounts to about 20 μs.
  - The Doppler spread, σ_ν , due to the motion of a vehicle may extend up to 40-80Hz.

8. 6 Statistical Characterization of Multipath Channels

- Classification of Multipath Channels

The particular form of the fading experienced by a multipath channel depends on whether the channel characterization is viewed in the frequency domain or the time domain.

8. 6 Statistical Characterization of Multipath Channels

- *In the frequency domain*:

channel’s coherence band-width B_c -- measuring the transmission bandwidth for which signal distortion across the channel becomes noticeable

If B_c < the bandwidth of the transmitted signal, *frequency selective* multipath channel;

If B_c > the bandwidth of the transmitted signal, *frequency nonselective (frequency flat)* multipath channel.

8. 6 Statistical Characterization of Multipath Channels

- *In the time domain*:

  coherence time τ_c -- measuring the transmitted signal duration for which distortion across the channel become noticeable.

  If τ_c < the duration of the received signal, the fading is *time selective*; otherwise, fading is *time nonselective*, or *time flat*.

8. 6 Statistical Characterization of Multipath Channels

- Thus, classify multipath channels as: *Flat-flat* channel, *Frequency-flat* channel, *Time-flat* channel, *Nonflat* channel. They are shown in Figure 8.21.
- The forbidden area, shown shaded in this figure, follows from the inverse relationship that exists between bandwidth and time duration.

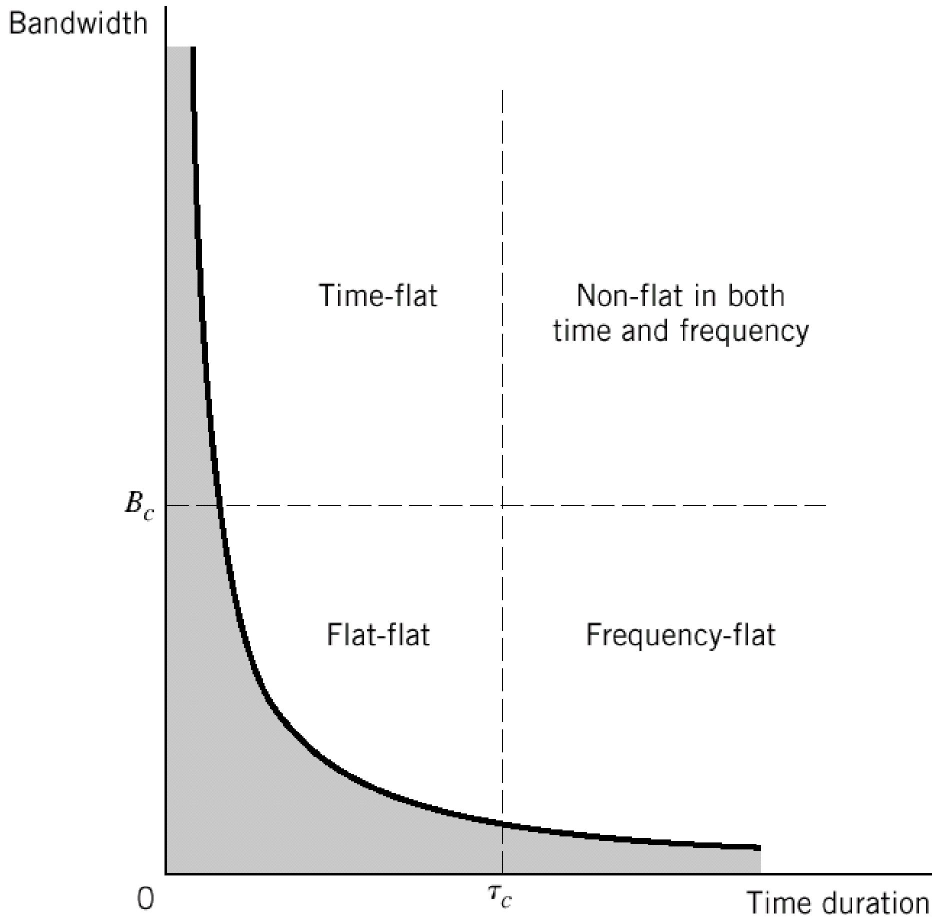


Figure 8.21 Illustrating the four classes of multipath channels: t_c = coherence time, B_c = coherence bandwidth.

8.7 Binary Signaling over a Rayleigh Fading Channel

- In a mobile radio environment, we have to consider the fluctuations in the amplitude and phase of the received signal due to *multipath effects*.
- Consider the transmission of binary data over a Rayleigh fading channel, the (low-pass) complex envelope of the received signal is modeled as

$$\tilde{x}(t) = \alpha \exp(-j\phi)\tilde{s}(t) + \tilde{w}(t) \tag{8.63}$$

8.7 Binary Signaling over a Rayleigh Fading Channel

where $\tilde{s}(t)$ -- the complex envelope of the transmitted (band-pass) signal

α -- the attenuation in transmission; a Rayleigh-distributed random variable

ϕ -- the phase-shift; a uniformly distributed random variable

$\tilde{w}(t)$ -- a complex-valued white Gaussian noise process

8.7 Binary Signaling over a Rayleigh Fading Channel

- Assume:
  - the channel is flat in both time & frequence → the phase-shift ϕ can be estimated without error and α is fixed or constant over a bit interval
  - coherent BPSK is used to do the data transmission

  Under these conditions, we express the average probability of symbol error as follows

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

8.7 Binary Signaling over a Rayleigh Fading Channel

- In the Equation (8.64), γ is an attenuated version of the transmitted E_b/N_0 shown as

$$\gamma = \frac{\alpha^2 E_b}{N_0} \qquad (8.65)$$

- $P_e(\gamma)$ may be viewed as a conditional probability given that α is fixed.Evaluate $P_e(\gamma)$ in the combined presence of fading and noise, namely, average $P_e(\gamma)$ over all values of γ, as

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

where $f(\gamma)$ is the probability density function of γ.

8.7 Binary Signaling over a Rayleigh Fading Channel

- Equation (8.65) → γ depends on the value of α →

  γ has a chi-square distribution with two degrees of freedom.

α is Rayleigh distributed

In particular, we express f (γ) as

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

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

8.7 Binary Signaling over a Rayleigh Fading Channel

$$\gamma_0 = E[\gamma]$$

$$= \frac{E_b}{N_0} E[a^2]$$

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

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

8.7 Binary Signaling over a Rayleigh Fading Channel

- Following a similar approach,we derive the bit error rates for binary signaling over a flat-flat Rayleigh fading channel,as shown in Table 8.2.

- In Figure 8.22, we use exact formulas of Table 8.2 to plot the bit error rate versus γ_0 expressed in decibels. For the sake of comparison, we also plot the figure for a nonfading channel.

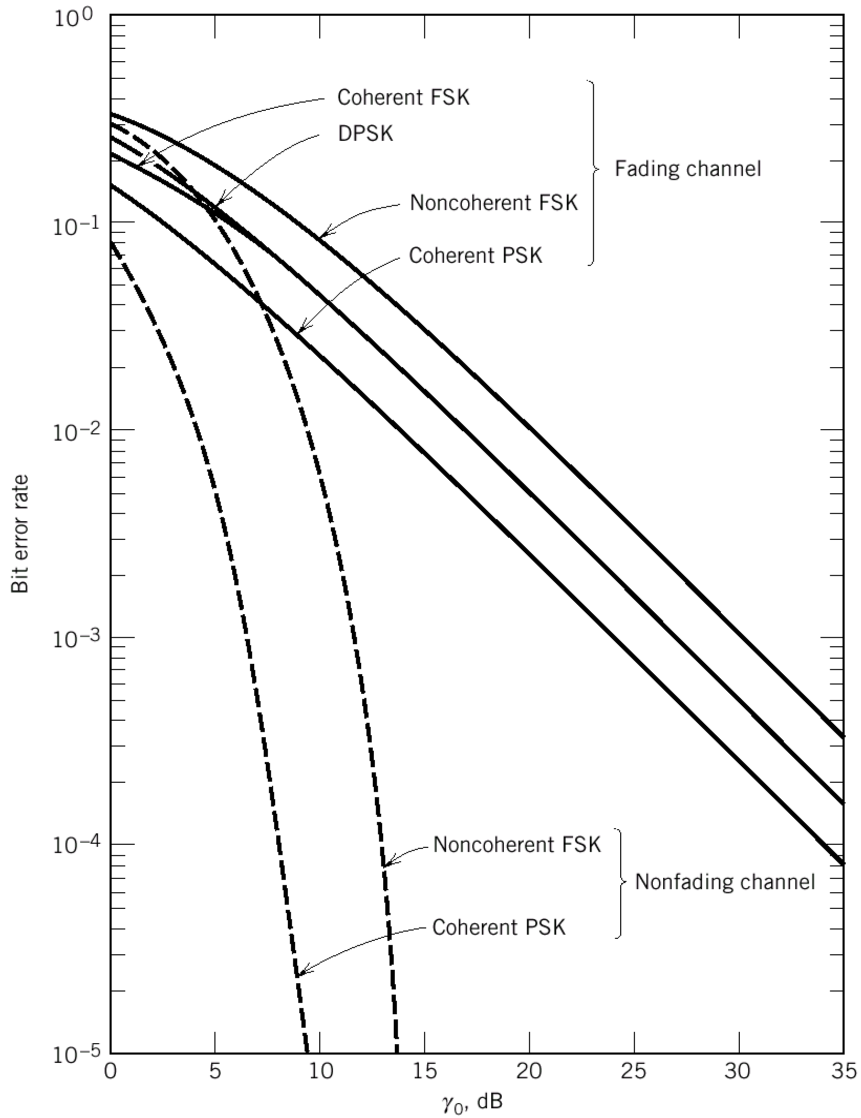


Figure 8.22
Performance of binary signaling schemes over a Rayleigh fading channel, shown as continuous curves; the dashed curves pertain to a nonfading channel.

8.7 Binary Signaling over a Rayleigh Fading Channel

- Rayleigh fading → severe degradation in the noise performance
- For large γ_0

  Fading : BER decrease approximately with γ_0 follows an *inverse law*

  Nonfading: BER decrease approximately with γ_0 follows an *exponential law*
- Thus, in a mobile environment, a large increase in mean SNR must be provided to ensure a BER that is low enough for practical use.

8.7 Binary Signaling over a Rayleigh Fading Channel

- Diversity Techniques

Diversity may be viewed as a form of redundancy. In particular, if *several replicas of the message signal* can be transmitted *simultaneously* over *independently* fading channels, it may be at least one of the received signals will not be severely degraded by fading.

8.7 Binary Signaling over a Rayleigh Fading Channel

- Several methods may make such a provision. The following diversity techniques are of particular interest:

  - Frequency diversity
  - Time (signal-repetition) diversity
  - Space diversity

8.7 Binary Signaling over a Rayleigh Fading Channel

- *Frequency diversity*

  The message signal is transmitted using *several carriers* that are spaced sufficiently apart from each other to provide *independently* fading versions of the signal. This can be done by choosing a frequency spacing equal to or larger than the coherence bandwidth of the channel.

8.7 Binary Signaling over a Rayleigh Fading Channel

- *Time diversity*

  The same message signal is transmitted in *different time slots*, with the spacing between successive time slots being equal to or greater than the coherence time of the channel. This can be done by using a repetition code for error-control coding.

8.7 Binary Signaling over a Rayleigh Fading Channel

- *Space diversity*

  Multiple transmitting or receiving antennas (or both) are used, with the spacing between adjacent antennas being chosen so as to assure the *independence* of fading events; this may be done by spacing the adjacent antennas by at least seven times the radio wavelength.

8.7 Binary Signaling over a Rayleigh Fading Channel

- *Linear diversity combining structure* (Figure 8.23)

  Assume: 1. L independently fading channels are created 2. Noise-free estimate of the channel attenuation factors $\{\alpha_l\}$ and the channel phase-shifts $\{\phi_l\}$ are available. 3. The system is to compensate only for short-term effects of a fading channel.

  Then, we get the optimum linear combiner -- *maximal-ratio combiner* .

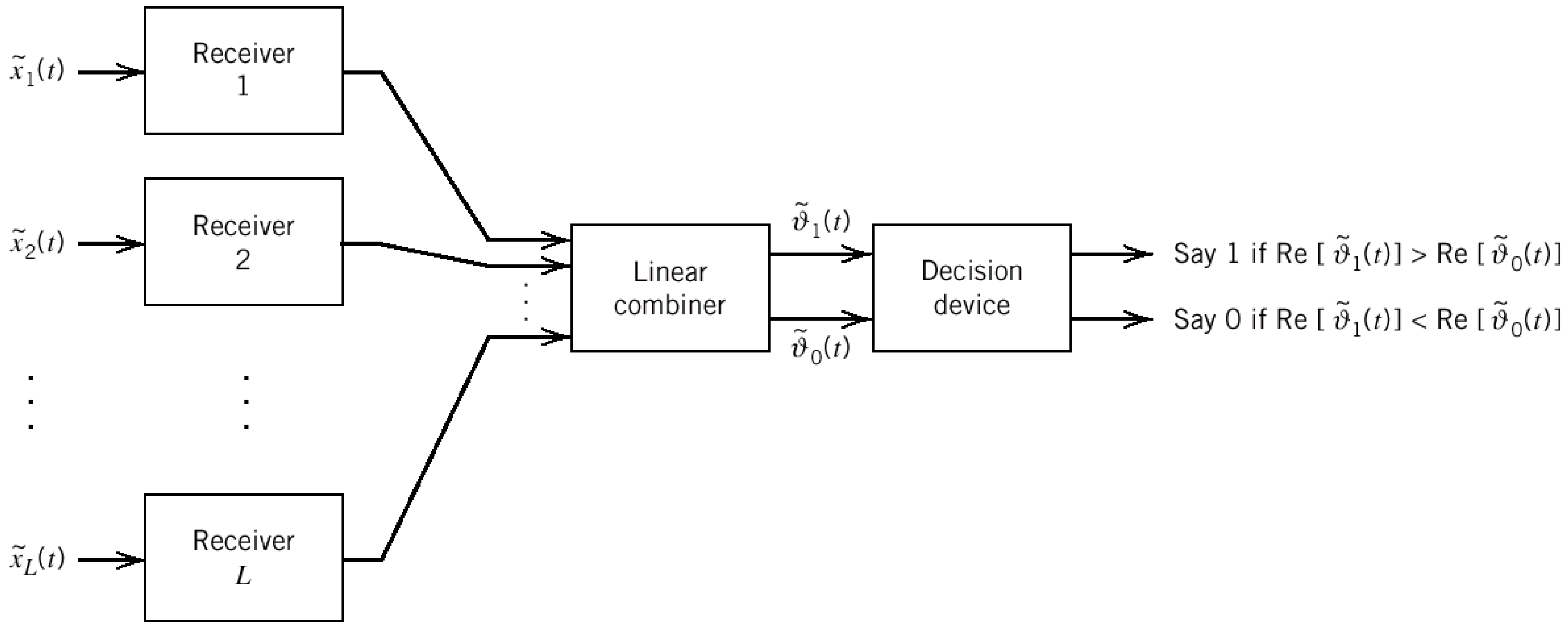


Figure 8.23

Block diagram illustrating the space diversity technique.

8.7 Binary Signaling over a Rayleigh Fading Channel

- The linear combiner results in two output complex envelopes(for BFSK) defined as

$$\tilde{v}_k(t) = \sum_{l=1}^{L} \alpha_l \exp(j\phi_l)\tilde{v}_{lk}(t), \qquad k = 0,1 \qquad (8.70)$$

where $\tilde{v}_{lk}(t)$ is the output of the kth matched filter in the lth receiver, $\alpha_l \exp(j\phi_l)$ is the complex conjugate of the lth channel gain which plays the role of a weighting factor (l = 1, 2, . . . , L, k = 0, 1).

8.7 Binary Signaling over a Rayleigh Fading Channel

- The "instantaneous" SNR of the optimum linear combiner described here is the sum of the instantaneous SNRs on the individual diversity branches (channels).
- Figure 8.24 shows the noise performance of the coherent BPSK, BDPSK, and noncoherent BFSK for L = 2, 4 independently fading channels. It clearly illustrates the effectiveness of diversity as a means of mitigating the short-term effects of Rayleigh fading.

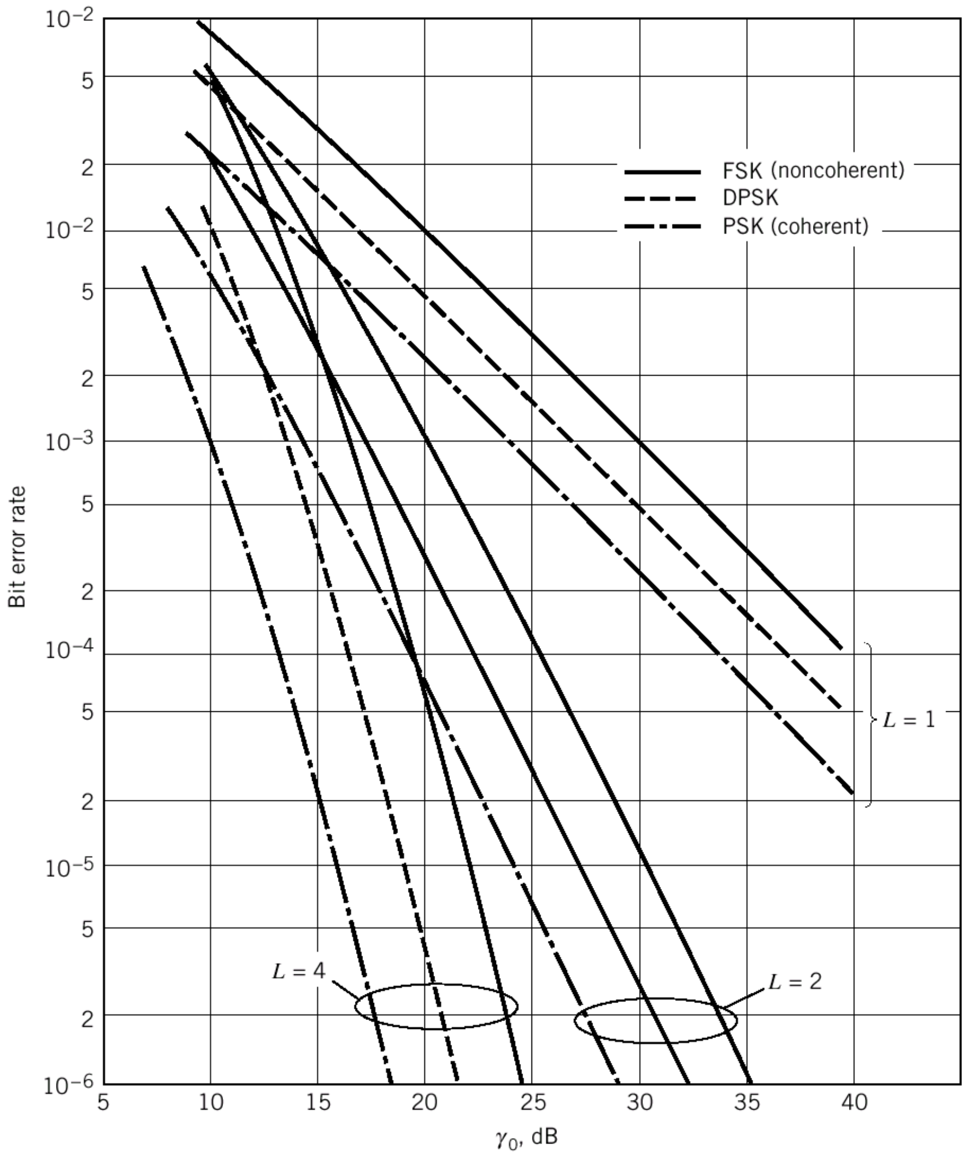


Figure 8.24
Performance of binary signaling schemes with diversity. (From Proakis, 1995, with permission of McGraw-Hill.)

8.8 TDMA and CDMA Wireless Communication Systems

- A user would like to talk and listen simultaneously in wireless communications → some form of duplexing is required (e.g. FDD -- *frequency division duplexing* as in North America).
- Indeed, FDD is an integral part of the two widely used wireless communication system summarized in Table 8.3.

8.8 TDMA and CDMA Wireless Communication Systems

- GSM : uses TDMA, data are transmitted over the channel in bursts, as shown in the frame structure of Figure 8.25.
- IS-95: uses CDMA.

deviation of the speeding codes from perfect orthogonality → multiple-access interference (MAI) → near-far problem → overcome by power control or multiuser detection → maximize system capacity

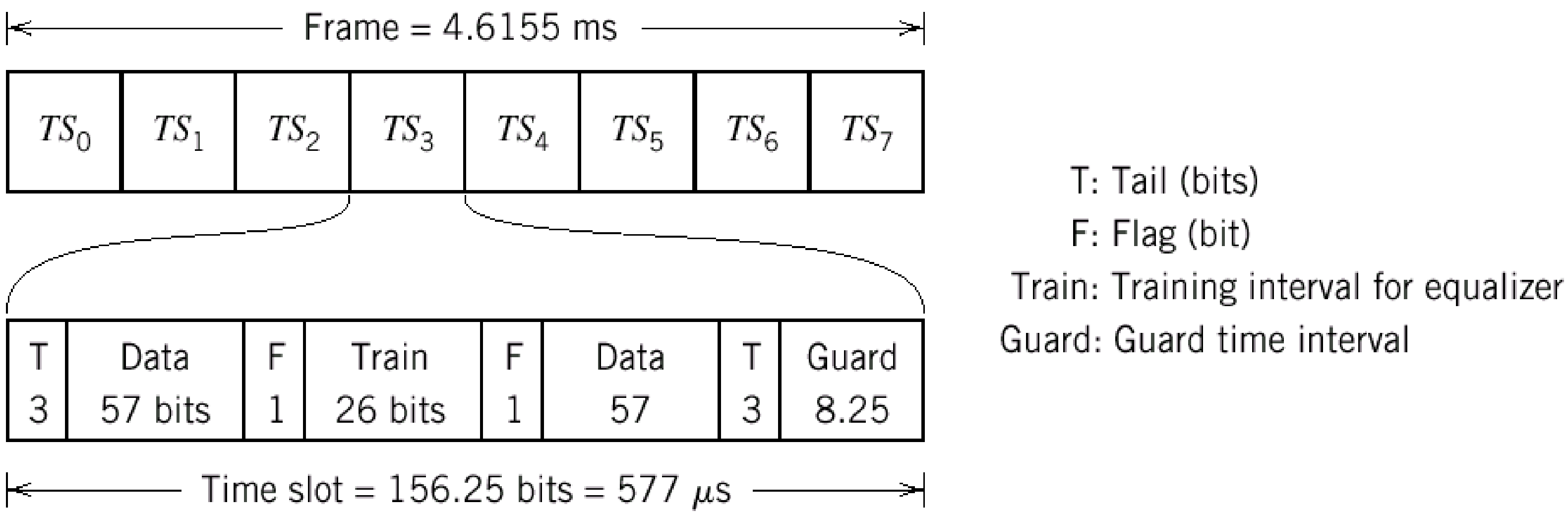


Figure 8.25

Frame structure of the GSM wireless communication system.

8.8 TDMA and CDMA Wireless Communication Systems

- RAKE Receiver

  RAKE receiver is originally developed as a ‘diversity’ receiver to equalize the effect of multipath.

- Two basic ideas

  1. Useful information about the transmitted signal is contained in the multipath component of the received signal.

  2. Multipath is approximated as a linear combination of differently delayed echoes.

8.8 TDMA and CDMA Wireless Communication Systems

- Method

Combating the effect of multipath by using a correlation method to detect the echo signals individually and adding them algebraically. In this way, ISI due to multipath is 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.

BPSK modulation at the transmitter

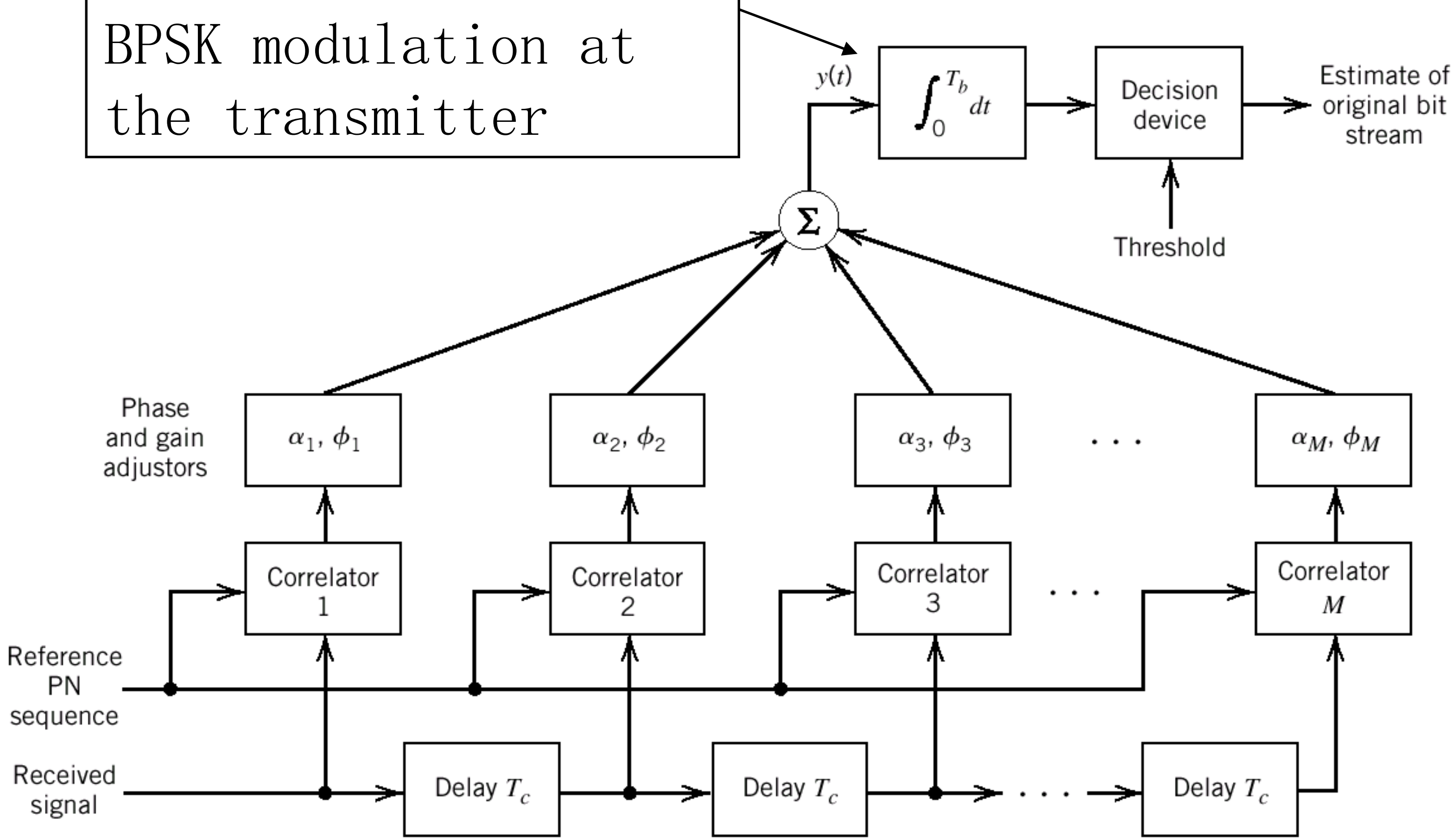


Figure 8.26 Block diagram of the RAKE receiver. (*the bank of parallel correlators has an appearance similar to the fingers of a rake → RAKE receiver*)

8.8 TDMA and CDMA Wireless Communication Systems

- As discussed in Chapter 7, we recall that the autocorrelation function of a PN sequence has a single peak of width T_c. Thus we need only make the bandwidth W$(=1/T_c)$ of the PN sequence sufficiently large to identify the significant echoes in the received signal.
- “phase and gain adjustors” -- Introduce appropriate correlator outputs delay and weighting to sure that the correlator outputs all add constructively.

8.8 TDMA and CDMA Wireless Communication Systems

- The weighting coefficients, α_k, are computed in accordance with the maximal ratio combining principle:

  The SNR 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 weighing is performed in proportion to the pertinent signal strength.

8.8 TDMA and CDMA Wireless Communication Systems

- The linear combiner output is

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

where $z_k(t)$ is the phase-compensated output of the kth correlator, and M is the number of correlators in the receiver.

8.9 Source Coding of Speech for Wireless Communications

- For the efficient use of channel bandwidth, digital wireless communication systems rely on the use of *speech coding* to remove almost all of the natural *redundancy* in speech, while maintaining a high-quality speech on decoding. The common approach is to use *source coding*, namely, *the linear predictive coding (LPC) of speech*.

8.9 Source Coding of Speech for Wireless Communications

- In this section, we discuss two types of speech coding:
    1. multi-pulse excited LPC (used in GSM)
    2. code-excited LPC (used in IS-95)

Our treatment of both of these speech coding techniques is in conceptual terms.

8.9.1 Multi-pulse Excited LPC

- This form of speech coding exploits the *principle of analysis by synthesis*, which means that the encoder includes a replica of the decoder in its design.
- The encoder consists of three main parts as indicated in Figure 8.27a.

Closed-loop optimization procedure

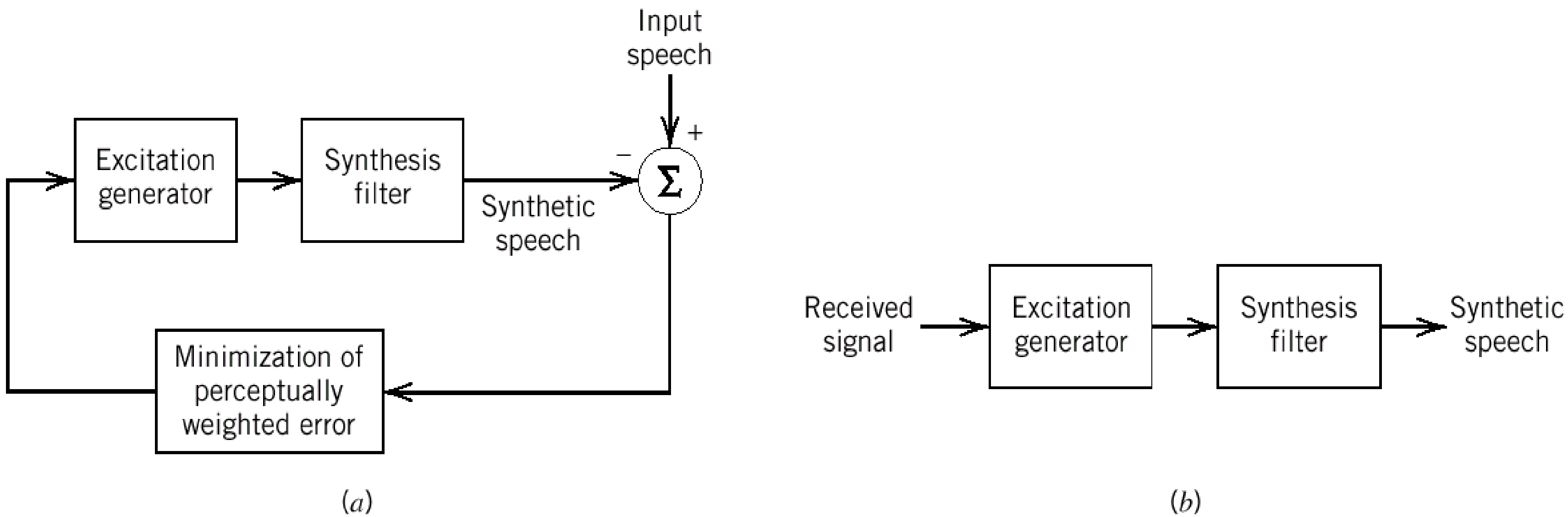


Figure 8.27

Multi-pulse excited linear predictive codec. (*a*) Encoder. (*b*) Decoder whose input (the received signal) consists of quantized filter parameters and quantized excitation as produced by the encoder.

8.9.1 Multi-pulse Excited LPC

- Three parts of the encoder:

  1. *Synthesis filter* for the predictive modeling of speech. It is to produce a synthetic version of the original speech that is of high quality.

  2. *Excitation generator* for producing the excitation applied to the synthesis filter.

  3. *Error minimization* for optimizing the perceptually weighted error between the original speech and synthesized speech.

8.9.1 Multi-pulse Excited LPC

- The encoding procedure itself has two main steps:

  1. The free parameters of the synthesis filter are computed using the actual speech samples as input.

  2. The optimum excitation for the synthesis filter is computed by minimizing the perceptually weighted error with the loop closed as in Figure 8.27a.

8.9.1 Multi-pulse Excited LPC

- The quantized filter parameters and quantized excitation constitute the transmitted signal.
- The decoder, located in the receiver, consists simply of two parts: *excitation generator and synthesis filter*, which are identical to the corresponding ones in the encoder, as shown in Figure 8.27b.
- The function of the decoder is to use the received signal to produce a synthetic version of the original speech signal.

8.9.2 Code-excited LPC

- Figure 8.28 shows the block diagram of the *code-excited LPC*, commonly referred to as CELP.
- The distinguishing feature of CELP is the use of a predetermined *codebook* of stochastic (zero-mean white Gaussian) vectors as the source of excitation for the synthesis filter. The synthesis filter consists of two all-pole filters, which perform short-term prediction and long-term prediction.

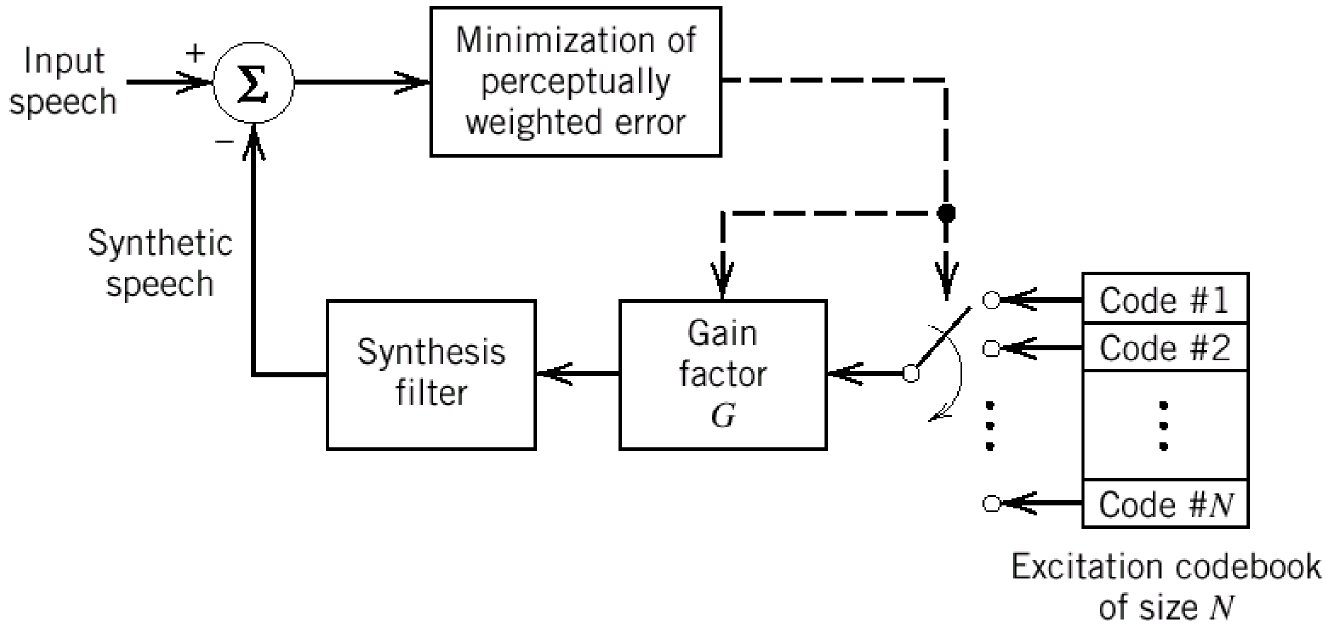


Figure 8.28

Encoder of the code-excited linear predictive codec (CELP): the transmitted signal consists of the address of the code selected from the codebook, quantized G, and quantized filter parameters.

8.9.2 Code-excited LPC

- The encoding procedure:

  1. The free parameters of the synthesis filter are computed, using the actual speech samples as input.

  2. The choice of a particular vector(code) stored in the excitation codebook and the gain factor G in Figure 8.28 is optimized by minimizing the average power of the perceptually weighted error between the original speech and synthesized speech.

8.9.2 Code-excited LPC

- The transmitted signal: the address of the stochastic vector selected from the codebook, the corresponding quantized gain factor and the quantized filter parameters.
- Advantage:

  CELP is capable of producing good-quality speech at bit rates below 8kb/s.

  Disadvantage:

  exhaustive search of the excitation codebook in the encoder → intensive computational complexity

8.10 Adaptive Antenna Arrays for Wireless Communications

- The goal of wireless communications is to allow as many users as possible to communicate reliably regardless of location and mobility.

- From the discussion presented in Sections 8.5 and 8.6, we find this goal is seriously impeded by three major channel impairments: *multipath, delay spread and co-channel interference*.

8.10 Adaptive Antenna Arrays for Wireless Communications

- *Multipath* : phase cancellation → severe fading → reduction in available signal power → a degraded noise performance
- *Delay spread* : different propagation delay(over 10% of the symbol duration) → ISI → a reduction in the attainable data rate
- *Co-channel interference(CCI)*: the same set of frequencies is used by several cells → *Co-channel interference* → limits the system capacity

8.10 Adaptive Antenna Arrays for Wireless Communications

- Cellular system use 120° sectorization at each base station, only one user accesses a sector of a base station at a given frequency.
- For combating the effects of multipath fading and co-channel interference, we use *three identical but separate antenna arrays*, one for each section of base station. Delay spread will be considered later.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Figure 8.29 shows the block diagram of an *array signal processor*.

  Assume:

  N users : One user is of interest

  $(N-1)$ users give rise to CCI

  M identical antenna elements

  each received signal corrupted by AWGN

  analysis for baseband signals(complex valued)

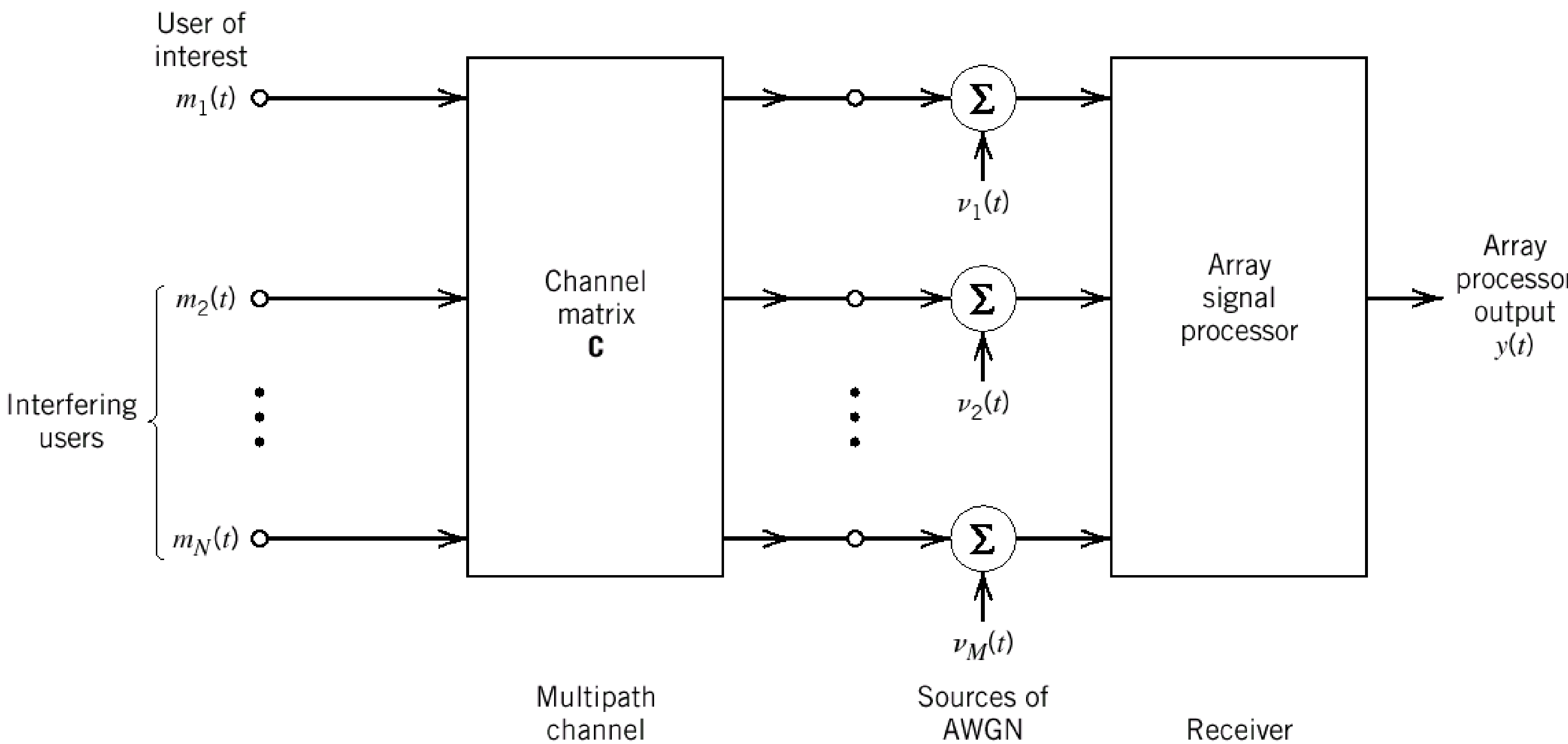


Figure 8.29

Block diagram of array signal processor that involves M antenna elements, and that is being driven by a multipath channel.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Multipath channel -- channel matrix, denoted by C

  The matrix C has dimensions M-by-N and may be expanded into N column vectors, as

  $$C = [c1, c2, ..., c_N] \qquad (8.72)$$

  where each column vector is of dimension M.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Goal:

  How to design a linear array signal processor for the receiver?

- Two requirements:

  Requirement 1: The co-channel interference is cancelled.

  Requirement 2: The output SNR for the user of interest is maximized.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Assume:

  1. flat Rayleigh fading channel

  2. large spacing between antenna elements → c1, c2, ..., c_N *linearly independent*

- Key design issue:

  How to find the *weight vector* denoted by w, which characterizes the array signal processor.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Two-step subspace procedure: (*zero-forcing kind*)

1. To fulfill requirement 1

Choose the M-dimensional weight vector w to be orthogonal to the vectors $c_2, \ldots, c_N$

2. To satisfy requirement 2

(i) Construct a subspace denoted by *W*. *W* has dimension M-(N − 1) = M-N + 1.

(ii) Project the complex conjugate of the channel vector c_1 onto the subspace W.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Given a vector space, formed by a set of linearly independent vectors, a *subspace* of the space is a subset satisfying two condition:

  (i) Given two vectors z_1 and z_2 in the subspace, $(z_1 + z_2)$ is still in the subspace.

  (ii) Given any vector z in the subspace & any scalar a, the multiple az is still in the subspace.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Example 8.3

Objective: To illustrate the 2-step subspace method for determining the weight vector w.

Assume:

- 2 users in the system (with vectors c_1 and c_2)
- antenna array consisting of three elements

Answer:

N = 2 and M = 3, Thus, $M-N+1 = 3-2+1 = 2$.

The subspace W is two-dimension.

8.10 Adaptive Antenna Arrays for Wireless Communications

- View user 1 as the user of interest, and user 2 as the interferer, we construct the signal-space diagram as shown in Figure 8.30.
- Conclusion: A linear receiver using optimum combining with M antenna elements and involving N–1 interfering users has the same performance as a linear receiver with M–N+1 (M>N–1) antenna elements without interference, independent of the multipath environment.

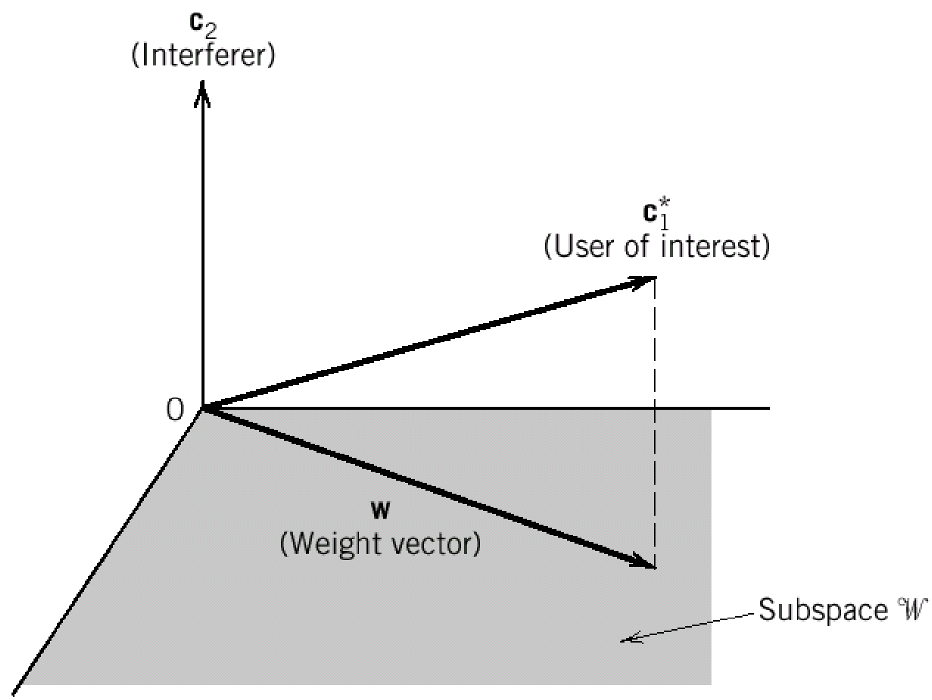


Figure 8.30

Signal-space diagram for Example 8.3, involving a user of interest, a single interferer, and an antenna array of 3 elements. The subspace $\mathcal{W}$, shown shaded, is two-dimensional in this example.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Compensation of the delay spread (significant compared to the symbol duration)

  Why? Causing intersymbol interference.

  How?

  Incorporating a *linear equalizer* in each antenna branch of the array. Figure 8.31 shows a space-time processor, which combines temporal and spatial processing.

a bank of FIR filters antenna array

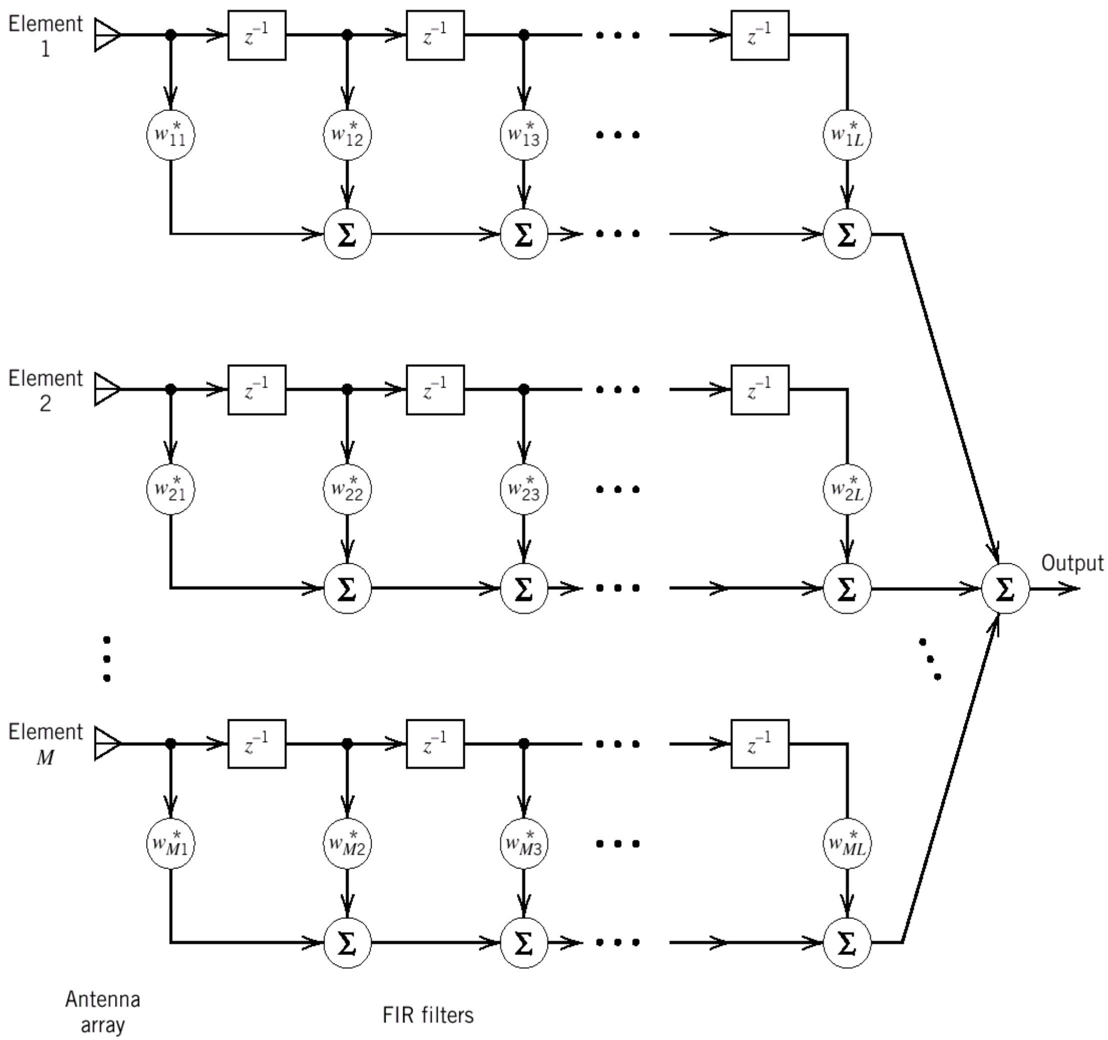


Figure 8.31

Baseband space-time processor. The blocks labeled z^{-1} are unit-delay elements with each delay being equal to the symbol period. The filter coefficients are complex valued. The FIR filters are all assumed to be of length *L*.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Adaptive Antenna Array

  Why adaptive?

  Figure 8.29 channel impairment are stationary

  However, *multipath fading, delay spread, co-channel interference* are all *nonstationary*, and the channel characterization may be *unknown*.

8.10 Adaptive Antenna Arrays for Wireless Communications

- *Note:* adaptive spatial processing
  least-mean-square(LMS) algorithm

  Assume: The delay spread is negligible and the multipath fading phenomenon is slow enough.
- Figure 8.32(baseband processing) shows the structure of an adaptive antenna array.

  A feedback system in the Figure 8.32 is built to make the antenna array adaptive to changes in the environment.

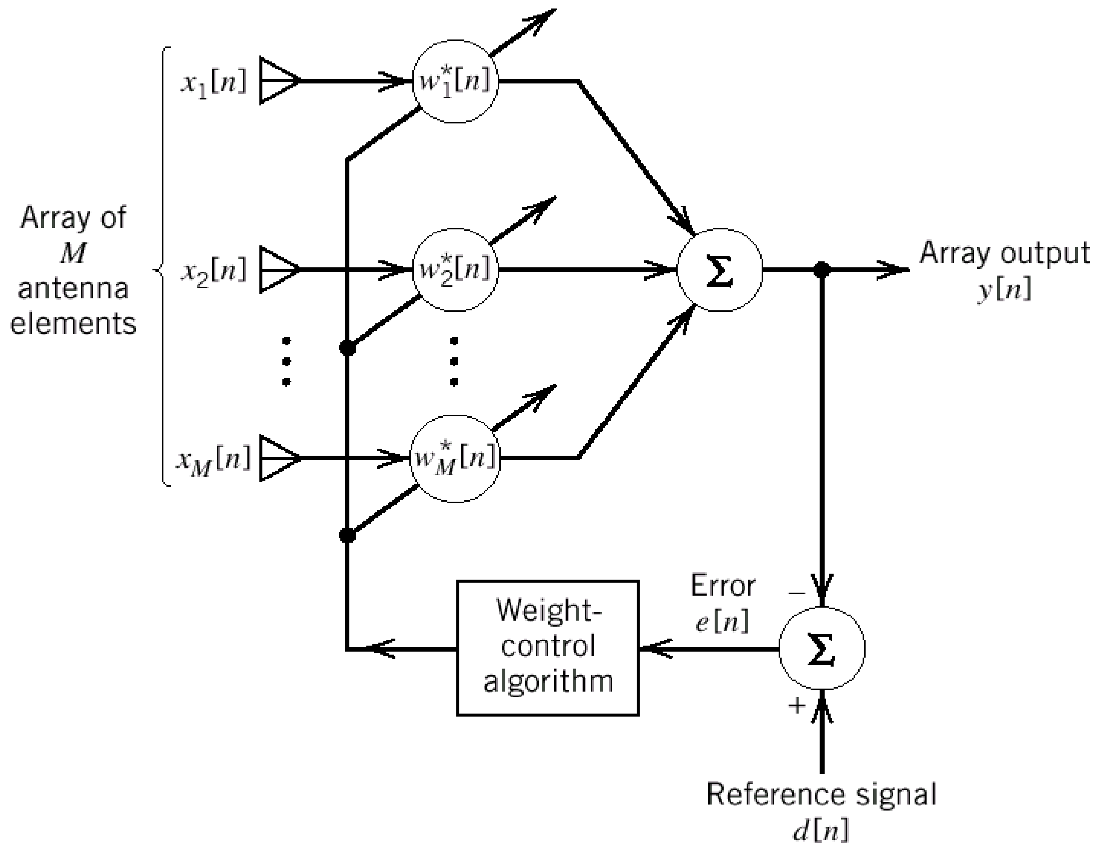


Figure 8.32

Block diagram of adaptive antenna array.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Optimization:

  Criterion: minimum mean-square error (MMSE)

  Cost function

$$J = E[|e[n]|^2] \tag{8.73}$$

where $e[n]$ -- the error signal at time $t = nT$

T -- the symbol period

n -- an integer serving as discrete time

8.10 Adaptive Antenna Arrays for Wireless Communications

- Complex LMS algorithm

Array output
$$y[n] = \sum_{k=1}^{M} w_k^*[n] x_k[n] \qquad (8.74)$$

Error signal
$$e[n] = d[n] - y[n] \qquad (8.75)$$

Weight adjustment
$$\Delta w_k[n] = u e_k^*[n] x_k[n], \quad k = 1,2,\ldots,M \qquad (8.76)$$

Updated weight
$$w_k[n+1] = w_k[n] + \Delta w_k[n], \quad k = 1, 2, \ldots, M \qquad (8.77)$$

8.10 Adaptive Antenna Arrays for Wireless Communications

where

- $x_k[n]$-- the kth array input
- $w_k[n]$-- the weight connected to kth element
- $d[n]$ -- the reference signal
- μ -- the step-size parameter
- The algorithm is initiated by setting $w_k[0] = 0$ for all k.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Advantages of using the complex LMS algorithm:
- Simplicity of implementation.
- Linear growth in complexity with the number of antenna elements.
- Robust performance with respect to disturbances.

8.10 Adaptive Antenna Arrays for Wireless Communications

- Drawbacks of the system:
- Slow rate of convergence. This limits the use of the complex LMS algorithm to a slow-fading environment.
- Sensitivity of the convergence behavior to variations in the reference signal and co-channel interference powers.

8.10 Adaptive Antenna Arrays for Wireless Communications

- *Direct matrix inversion* (DMI) algorithm
- Overcome the limitations of the complex LMS algorithm.
- Operate in the batch mode in that the computation of the elemental weight is based on a batch of K snapshots.

↓

RLS algorithm

8.10 Adaptive Antenna Arrays for Wireless Communications

- How to choose the batch size K ?
- K should be small enough for the batch of snapshots used in the computation to be justifiably treated as pseudo-stationary.
- K should be large enough for the computed values of the elemental weights to approach the MMSE solution.
- K is chosen as a compromise between the two conflicting requirements.

8.10 Adaptive Antenna Arrays for Wireless Communications

- When the teletraffic is high, the base stations are ordinarily figured as microcells. In such configuration, there are many inexpensive base station in close proximity to each other.
- Adaptive antenna arrays provides the means for an alternative configuration where there are fewer base stations and further apart from each other.

8.11 Summary and Discussion

- Two multiuser communications: (Both rely on radio propagation to link the receiver to the transmitter.)

  1. satellite communication (global coverage)

     uplink: earth terminal → satellite

     downlink: satellite → earth terminal

     line-of-sight path

     additive white Gaussian noise channel

8.11 Summary and Discussion

2. wireless communication (mobility)

reverse link(uplink): mobile → base station

forward link(downlink):base station → mobile

co-channel interference

multipath → fading

↘ delay spread

signal-dependent phenomena

specialize techniques used, such as:

diversity

adaptive array antennas

RAKE receiver

8.11 Summary and Discussion

- Remarks contrasting wireless communications to wired communications

  1. in wired communication system

     major source of concern -- noise

     sufficient channel bandwidth -- PCM(64 kb/s) as standard method for converting speech

     → almost noise-free performance

8.11 Summary and Discussion

2. In wireless communication

channel bandwidth -- precious resource

→ spectral efficient speech coding (e.g. CELP)

high-quality

rate(<16 kb/s) << PCM rate

channel coding used (Chapter 10)