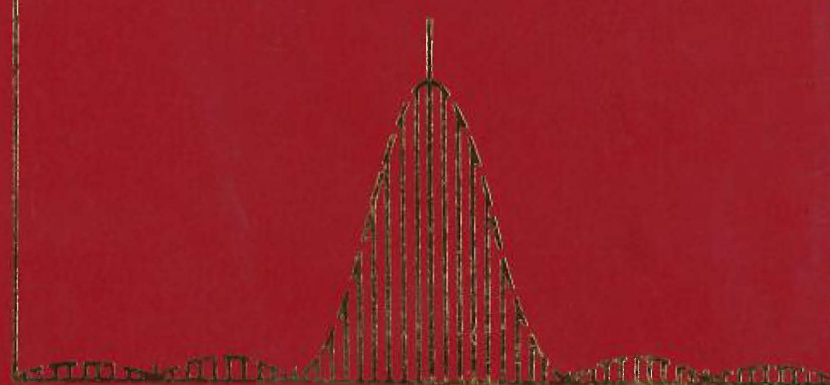
THIRD EDITION

DIGITAL AND ANALOG COMMUNICATION SYSTEMS

Leon W. Couch II



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Abbreviations

ac	alternating current
ADC	analog-to-digital conversion
ADM	adaptive delta modulation
AM	amplitude modulation
ANSI	American National Standards Institute
APLL	analog phase-locked loop
ASCII	American Standard Code for Information Interchange
ATI	American Telephone and Telegraph
AWGN	additive white Gaussian noise
BER	bit error rate
BISYNC	binary synchronous communication protocol
BPSK	binary phase shift keying
CATV	cable antenna television system
CCITI	Consultative Committee for International Telephone and Telegraph
CDMA	code-division multiple access
CFT	continuous Fourier transform
CMOS	complementary metal oxide conductor
c o	central telephone office
CRT	cathode ray tube
dB	decibel
de	direct current
DCE	data communications equipment
OFT	discrete Fourier transform
DM	delta modulation
DPCM	differential pulse code modulation
DTE	data terminal equipment
EIRP	effective isotropic radiated power
EIA	Electronics Industries Association
ERP	effective radiated power
FCC	Fede al Communications Commission (United States)
FDM	frequency-division multiplexing
FET	field effect transistor
FFT	fast Fourier transform
FM	frequency modulation
FSK	frequency shift keying
HDLC	high-level data link control protocol
HF	high frequency
IEEE	Institute of Electrical and Electronics Engineers
IF	intermediate frequency
IMD	intermodulation distortion
ISI	intersymbol interference
ISO	International Organization for Standardization
ITU	International Telecommunications Union of the United Nations
LED	light-emitting diode
LO	local oscillator
LOS	line of sight

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LPF	low-pass filter
LSSB	lower single sideband
MAP	maximum a posteriori (criterion)
MPSK	M-ary phase shift keying
MSK	minimum-shift keying
NBFM	narrowban frequency modulation
NCS	National Communications System of the federal government
NRZ	nonreturn-to-zero
OOK	on-off keying
OQPSK	offset quadrature phase shift keying
OSI	open system interconnection protocol model
PAM	pulse amplitude modulation
PCM	pulse code modulation
PD	phase detection
PDF	probability density function
PEP	peak envelope power
PLL	phase-locked loop
PM	phase modulation
PPM	pulse position modulation
PSD	power spectral density
PSK	phase shift keying
PTM	pulse time modulation
PWM	pulse width modulation
QAM	quadrature amplitude modulation
QPSK	quadrature phase shift
rms	root-mean-square
RT	remote telephone terminal
RZ	return-to-zero
SAW	surface acoustics wave
SDLC	synchronous data link control protocol
SIN	signal-to-noise ratio
SS	spread spectrum (system)
SSB	single sideband
TOM	time-division multiplexing
TOMA	time-division multiplex access
TELCO	telephone company
THO	total harmonic distortion
TTL	transistor transistor logic
TV	television
TYRO	TV receive only terminal
TWT	traveling-wave-tube
UHF	ultra high frequency
USSB	upper single sideband
VCO	voltage-controlled oscillator
VF	voice frequency
VHF	very high frequency
VSB	vestigial sideband
WBFM	wideband frequency modulation
WER	word error rate

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Macmillan Publishing Company
New York

Collier Macmillan Publishers
London

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C5

Bandpass Communication Systems

5-1 Introduction

The goal of this chapter is to obtain a *working knowledge* of bandpass communication *systems.* To do so we will consider examples of data, telephone, cellular radio, television, satellite, spread spectrum, and fiber optic systems. The tools developed in previous chapters will be utilized here.

Theoretical system concepts are also emphasized. As indicated earlier, the goodness of a communication system is measured in terms of the required transmission power or energy, required system bandwidth, system cost, and the quality of information produced at the receiver output. In the first part of this chapter, we will study bandpass digital signals and their associated spectrum and bandwidth. For digital systems, the quality of the recovered information is usually specified by the probability of error on the output digital signal. For analog systems the quality of the recovered information is usually specified by the signal-to-noise ratio associated with the output analog waveform. In the latter part of this chapter, a technique will be developed for evaluating the signal-10-noise ratio at the output of an analog system and the probability of bit error for the output of a digital system. This technique does not require the use of statistics, *provided* that the noise characteristics of the receiver detector circuit are known. If the receiver is purchased from a vendor, these characteristics are included in the receiver specifications. They can also be obtained by measurement or by theoretical evaluation. The important topic of the theoretical evaluation of the detector characteristics is deferred until Chapters 6 through 8. In those chapters we will concentrate on the statistical concepts involved, yet have a practical understanding of their application that was gained in Chapters I through 5.

5-2 Digital Communication Systems

Since digital communication systems are becoming more significant each day, a whole chapter, Chapter 3, was devoted to *baseband* digital signaling. As stated in Chapter I, the source information usually needs to be mapped into a *bandpass* signal in order to communicate over a channel. *Bandpass* digital communication systems consist of modulating a *baseband* digital signal (Chapter 3) onto a carrier using AM, PM, FM, or some type of generalized modulation technique (Chapter 4). In this section we extend our results of baseband digital signaling to bandpass

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digital signaling. From (4-23), the PSD for the bandpass signal, *s(t),* is given by

$$\mathcal{P}_s(f) = \tfrac{1}{4}[\mathcal{P}_g(f - f_c) + \mathcal{P}_g(-f - f_c)] \tag{5-1}$$

where f_c is the carrier frequency and $\mathcal{P}_g(f)$ is the PSD of the complex envelope. [It is recalled that the complex envelope is a mapping of the baseband signal *m(t)* onto the complex plane and that the particular type of mapping function used determines the modulation type (AM, FM, etc.).] In this section we will determine $\mathcal{P}_g(f)$ for various types of digital signaling so that the PSD of the bandpass digital signal can be obtained.

Bandpass digital communication systems may be divided into two main categories: binary digital systems and multilevel (i.e., more than two levels) digital systems.

Binary Signaling

The most common binary bandpass signaling techniques are illustrated in Fig. 5-1 . They are

- *On-off keying* (OOK), also called *amplitude shift keying* (ASK), which consists of keying (switching) a carrier sinusoid on and off with a unipolar binary signal. It is identical to *unipolar* binary modulation on a DSB-SC signal, (4-117). Morse code radio transmission is an example of this technique. Consequently, OOK was one of the first modulation techniques to be used and precedes analog communication systems.
- *Binary-phase shift keying* (BPSK), which consists of shifting the phase of a sinusoidal carrier 0° or 180° with a unipolar binary signal. It is equivalent to PM signaling with a digital waveform and is also equiva)ent to modulating a DSB-SC signal with a polar digital waveform.
- *Frequency-shift keying* (FSK), which consists of shifting the frequency of a s,nusoidal carrier from a mark frequency (corresponding, for example, to sending a binary I) to a space frequency (corresponding to sending a binary 0) according to the baseband digital signal. It is identical to modulating a FM carrier with a binary digital signal.

As indicated in Sec. 3-5, usually the bandwidth of the digital signal needs to be minimized to achieve spectral conservation as one transmits information over a channel. This may be accomplished by using a premodulation raised cosine-rolloff filter to minimize the bandwidth of the digital signal and yet not introduce ISi. The shaping of the baseband digital signal produces an analog baseband waveform that modulates the transmitter. Figure 5-1 f illustrates the resulting DSB-SC signal when a premodulation filter is used. Thus the BPSK signal of Fig. 5- 1d becomes a DSB◦SC signal (Fig. 5- 1f) when premodulation filtering is used.

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In 1980 the AR6A microwave relay system was introduced into service [Markle, 1978, 1983]. Until this time FDM/FM was the modulation method used on microwave links because the intermodulation distortion on klystron microwave power amplifiers was too large for SSB techniques to be used. The development of a low-distortion TWT (traveling-wave tube) amplifier has made FDM/SSB possible in the AR6A. The VF capacity of the AR6A is *three times* that of microwave systems using the FDM/FM technique, so that it will handle 10 mastergroups of 600 VF channels each. The performance of digital microwave systems is also shown.

Referring to Table 5-7 again, we realize that almost all of the open-wire pairs have been replaced by twisted-pair cable, but occasionally, some can still be seen along railroad tracks. Fiber cable is now rapidly overtaking twisted-pair and coaxial cable.

Tropospheric scatter radio is the transmission of UHF signals "over the horizon" by reflecting and refracting the signals in the troposphere. A ground range of 400 mi can be obtained by using high-power and large parabolic antennas. In most applications "tropo" systems are being replaced by satellite systems.

The capacity of satellite communication also depends on the type of modulation used. With FDM/FM the Intelsat IV and V series satellites can handle 300 VF channels on each transponder (receiver-transmitter relay system). If quadrature phase-shift-keyed/single VF channel per carrier (QPSK/SCPC) is used, 800 VF channels can be accommodated. This subject is discussed in more detail in the following section, which describes satellite communication systems.

5-4 Satellite Communication Systems

The number of satellite communication systems has increased tremendously over the last few years. Satellite communications now relay the bulk of transoceanic telephone traffic, having greatly surpassed that sent by submarine cable. In addition, satellites have made transoceanic relaying of television signals possible. Satellite communications thus provide the relaying of data, telephone, and television signals, and it is now feasible to provide national direct-into-the-home television transmission via satellite.

Most communication satellites are placed in *synchronous* orbit. That is, they are located 22,300 mi above the equator so that their orbital period is the same as that of the Earth. Consequently, from Earth these satellites appear to be located at an approximately fixed point in the sky, as shown in Fig. 5-25. This enables the Earth station antennas to be simplified since they are pointed in a fixed direction and do not have to track a moving object. (For communication to the polar regions of the Earth, satellites in polar orbits are used, which require Earth stations with tracking antennas.)

The most desired, as well as the most popular, frequency band for satellite communication systems is 6 GHz on the uplink (Earth-to-satellite transmission)

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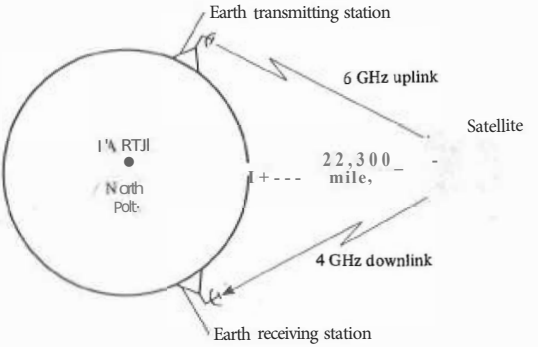


FIGURE 5-25 Communications satellite in synchronous orbit.

and 4 GHz on the downlink (satellite-to-Earth transmission). In this frequency range the equipment is relatively inexpensive, the cosmic noise is small, and the frequency is low enough so that rainfall does not appreciably attenuate the signals. Other losses, such as ionospheric scintillation and atmospheric absorption, are small at these frequencies [Spilker, 1977]. (Absorption occurs in specific frequency bands and is caused by the signal exciting atmospheric gases and water vapor.) However, existing terrestrial microwave radio relay links are already assigned to operate within the 6- and 4-GHz bands (see Table 5-7), so the FCC limits the power density on Earth from the satellite transmitters. One also has to be careful in locating the Earth station satellite receiving antennas so that they do not receive interfering signals from the terrestrial microwave radio links that are using the same frequency assignment. In the 6/4-GHz band, synchronous satellites are assigned an orbital spacing of 2° (United States standard).

Newer satellites have to operate in higher-frequency bands because there are so few vacant spectral assignments in the 6/4-GHz band (C band). The Ku band satellites use 14 GHz on the uplink and 12 GHz on the downlink with an orbital spacing of 3 degrees. For direct-to-the-home broadcasting a 17-GHz uplink and 12-GHz downlink have been assigned [Pritchard and Kase, 1981], but do not appear to be commercially successful.

Each satellite has a number of *transponders* (receiver-to-transmitter) aboard to amplify the received signal from the uplink and to down-convert the signal for transmission on the downlink (see Fig. 5-26). This figure shows a "bent pipe transponder" that does not demodulate the received signal and perform signal processing but acts as a high-power-gain down converter. Most transponders are designed for a bandwidth of 36, 54, or 72 MHz, with 36 MHz being the standard used for C-Band (6/4 GHz) television relay service. As technology permits, processing transponders will come into use since an improvement in error performance (for digital signaling) can be realized.

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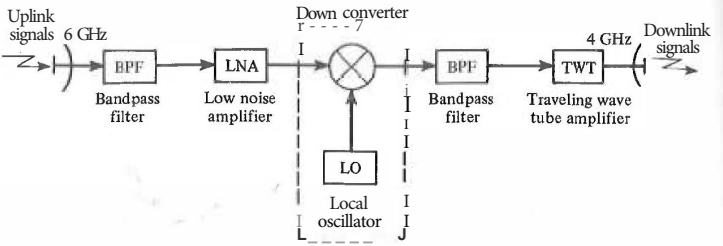
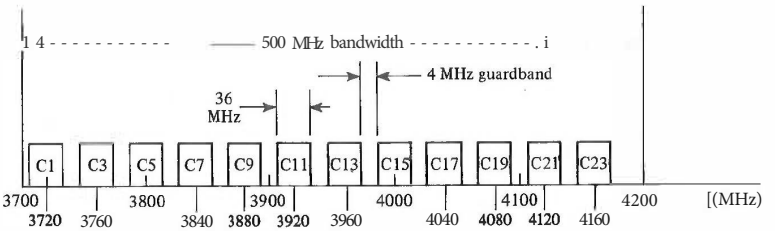


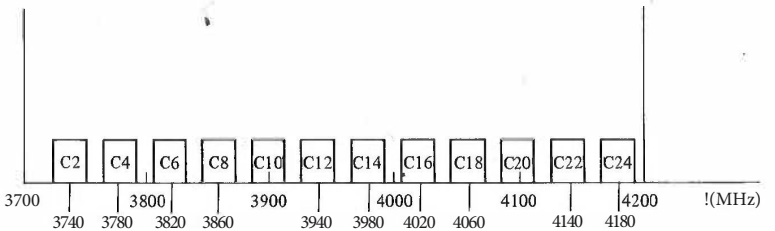
FIGURE 5-26 Simplified block diagram of a communications satellite transponder.

Each satellite is assigned a synchronous orbit position and a frequency band in which to operate. In the 6/4-GHz band, each satellite is permitted to use a 500-MHz-wide spectral assignment, and a typical satellite has 12 transponders aboard, with each transponder using 36 MHz of the 500-MHz bandwidth assignment. Some satellites reuse the same frequency band by having 12 transponders operating with vertically polarized radiated signals and 12 transponders with horizontally polarized signals.† The 6/4-GHz frequency assignment for

†A vertically polarized satellite signal has an E field that is oriented vertically (parallel with the axis of rotation of the Earth). With horizontal polarization the E field is horizontal.



(a) Horizontal Polarization•



(b) Vertical Polarization"

•These are the polarizations used for the Galaxy Satellites. Some of the other satellites use opposite polarization assignments.

FIGURE 5-27 The 6/4-GHz satellite transponder frequency plan for the downlink channels. (For the uplink frequency plan, add 2225 MHz to the numbers given above.)

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satellites is shown in Fig. 5-27. These satellites use 24 transponders (12 using vertical polarization and 12 using horizontal polarization). The transponders are denoted by C1 for channel 1, C2 for channel 2, and so on. These satellites are used mainly to relay signals for cable antenna television (CATV) systems.

Figure 5-28 shows an example of a 6/4-GHz satellite. This is the Galaxy series of satellites manufactured by the Hughes Aircraft Company, a member of their HS 376 family of satellites. The satellite is approximately 21 ft 8 in. high after deployment and has a diameter of 7 ft 1 in. The Galaxy mission's lifetime is 9 years and is limited by the amount of propellent stored onboard for firing thrusters that keep the satellite to within 0.1° of its assigned position (in both the north-south and east-west directions). The body of the satellite is spun

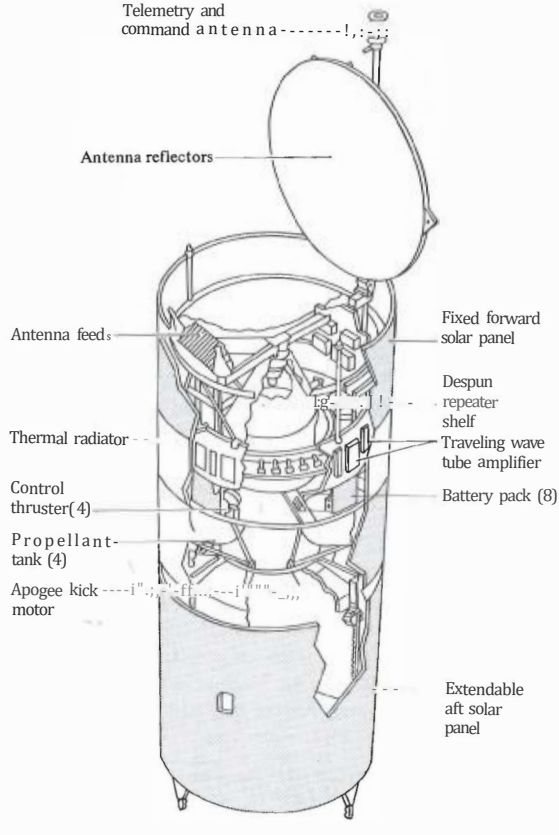


FIGURE 5-28 The Galaxy satellite. [Courtesy of Hughes Communications, Inc., a wholly owned subsidiary of Hughes Aircraft Company.]

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at a rate of 50 rpm to provide gyroscopic stability. The solar cell/battery system supplies a de power of at least 741 W to operate the electronics.

Television Signal Transmission

Broadcast-quality TV signals are relayed via satellite by using a single transponder (36-MHz bandwidth) for ach TV signal. This is accomplished by modulating the composite 4.5-MHz bandwidth visual signal onto a 6-GHz FM carrier, as shown in Fig. 5-29. The composite visual signal consists of the black-and-white video signal, the color subcarrier signals, and the synchronizing pulse signal, as discussed in Sec. 5-9. The aural signal is also relayed over the same transponder by frequency modulating it onto a 6.8-MHz subcarrier that is frequency-division-multiplexed with the composite video signal. The resulting wideband signal frequency modulates the transmitter.

The bandwidth of the 6-GHz FM signal may be evaluated by using Carson's rule. The peak deviation of the composite video is 10.5 MHz and the peak deviation of the subcarrier is 2 MHz, giving an overall peak deviation of ΔF = 12.5 MHz. The baseband bandwidth is approximately 6.8 MHz. The transmission bandwidth is

$$B_T = 2(\Delta F + B) = 2(12.5 + 6.8) = 38.6 \text{ MHz} \qquad (5\text{-}49)$$

which is accepted by the 36-MHz transponder.

This FDM/FM technique is also used by telephone companies for terrestrial relay of TV signals over microwave radio links that were characterized in Table 5-7 (except for the AR6A system, which cannot be used for TV relay). Each microwave link has the capacity for one TV channel (or for the number of VF channels, as indicated in the table). When video signals are relayed, wideband (0- to 15-kHz) aural signals can also be relayed by using FM subcarriers. Some typical subcarrier frequencies that are used are 5.58, 5.76, 6.2, 6.3, 6.48, 6.8, 7.38, and 7.56 MHz.

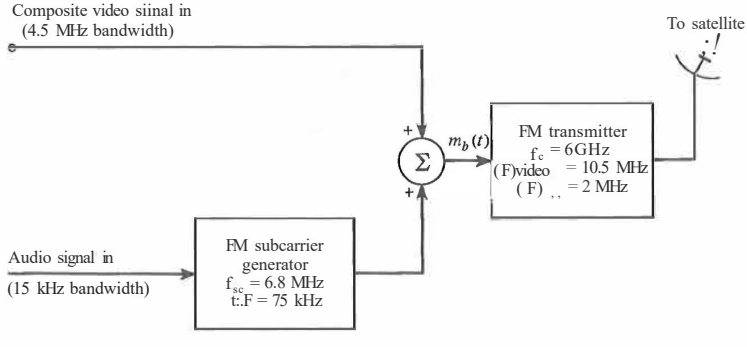


FIGURE 5-29 Transmission of broadcast-quality TV signals from a ground station.

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It is hoped that a second collision will not occur. If it does, the retransmission process is repeated until each party is successful. This technique has the advantage of being relatively inexpensive to implement, but it will not work if there is heavy traffic loading on the satellite. Then the satellite becomes saturated with colliding packets, and the collisions can be avoided only by stopping all new transmissions and increasing the random delay required before retransmission. A more elaborate technique is called *slotted ALOHA.* Here the packets are transmitted at random, but only in certain time slots. This avoids collisions due to partial packet overlap.

Very small aperture terminals (VSAT) have become popular with the availability of Ku band satellites and with the recent advances that have made low-cost terminals possible. "Very small aperture" implies that these systems use Earth-terminal antennas that are relatively small, which is about 1.8 meters in diameter. Solid state power.amplifiers (1 to 2 watts), low-cost frequency converters, digital processing, and VSLI circuits have made VSATs feasible. The objective of VSAT systems is to provide low-cost data and voice transmission directly to users such as auto dealerships, banks (automatic teller machines), brokerage firms, pipeline companies (monitor and control), hotels and airlines (reservations), retail stores (data networks), and corporations (point-to-multipoint transmission). Typically, VSATs offer high-quality transmission (bit error rates of Jess than 10^{-7} for 99.5% of the time) at data rates from 100 b/s to 9.6 kb/s [Chakraborty, 1988]. Many users share a single satellite transponder via TDMA so that the user's cost can be substantially less than that for the same type of service provided by the Jong-distance public telephone network.

In the next section, a procedure will be developed for evaluating the performance of satellite systems in terms of the output probability of bit error for digital systems and the output signal-to-noise ratio for analog systems. More details can also be obtained from books devoted specifically to satellite systems [Ha, 1986, 1989; Pratt and Bostian, 1986; Spilker, 1977).

5-5 Receiver Signal-to-Noise Ratio

In this section, formulas will be developed for the signal-to-noise ratio at the detector input as a function of the transmitted effective isotropic radiated power (EIRP), the free-space loss, the receive antenna gain, and the receive system noise figure. These results will allow us to evaluate the quality of the receiver output (as measured by the probability of bit error for digital systems and output signal-to-noise ratio for analog systems), provided that the noise characteristics of the receiver detector circuit are known.

Signal Power Received

In communication systems the received signal power (as opposed to the voltage or current level) is the important quantity since it is the output signal power that ultimately drives the display device, which, for example, may be a loudspeaker

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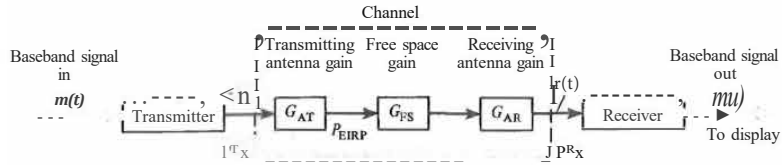


FIGURE 5-34 Block diagram of a communication system with a free-space transmission channel.

or a cathode ray tube (CRT). Consequently, it is the *power* of the received signal that is of prime importance while trying to *minimize* the *noise sources* that feed into the system and get amplified. The voltage gain, current gain, and impedance levels have to be such that the required power gain is achieved. In FET circuits it is accomplished by using relatively large voltage levels and small currents (high-impedance circuitry). With bipolar transistors power gain is achieved with relatively small voltages and large currents (low-impedance circuitry).

A block diagram of a communication system with a free-space transmission channel is shown in Fig. 5-34. The overall power gain (or power transfer function) of the channel is

$$\frac{P_{Rx}}{P_{Tx}} = G_{AT}G_{FS}G_{AR} \tag{5-50}$$

where $P_T x$ is the signal power into the transmitting antenna, G_{AT} is the transmitting antenna power gain, G_{PS} is the free-space power gain† (which is orders of magnitude less than one in typical communication systems), G_{AR} is the receiving antenna power gain, and P_{Rx} is the signal power into the receiver.

In order to use this relationship, these gains should be expressed in terms of useful antenna and free-space parameters. Here GAT and GAR are taken to be the power gains with respect to an isotropic antenna.* The EIRP is

$$P_{EIRP} = G_{AT}P_{Tx} \tag{5-51}$$

The anlenna gain is defined as

$$G_A = \frac{\text{radiation intensity of the actual antenna in the direction of maximum radiation}}{\text{radiation intensity of an isotropic antenna with the same power input}}$$

The gain for some practical antennas is given in Table 5-8.

†A gain transfer function is the output quantity divided by the input quantity, whereas a loss transfer function is the input quantity divided by the output quantity.

*An isotropic antenna is a nonrealizable theoretical antenna that radiates equally well in all directions and is a useful reference for comparing practical antennas.

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TABLE 5-8 Antenna Gains and Effective Areas

Type of Antenna	Power Gain G_A (absolute units)	Effective Area A_e (meters^2)
Isotropic	1	A2/4lr
Infinitesimal dipole o loop	1.5	1.5 „.2/41r
Half-wave dipole	1.64	1.64 A2/4lr
Horn (optimized), mouth area, A	10 A/A2	0.81 A
Parabola or "dish" with area, A	7.0A/A2	0.56 A
Turnstile (two crossed dipoles fed 90° out of phase)	1.15	1.15 A2/4lr

The power density (W/m^2) of an isotropic antenna at a distance d from the antenna is

$$\text{power density at } d = \frac{\text{transmitted power}}{\text{area of a sphere with radius } d} = \frac{\text{pEIRP}}{4\text{ird}^2} \tag{5-52}$$

If the receiving antenna is placed at d meters from the transmitting antenna, it will act like a "catcher's mitt" and intercept the power in an effective area of $(A_e)_{Rx}$ (meters^2), so that the received power will be

$$P_{Rx} = G_{AT}\frac{(PT,)}{4,rd2}(A,)Rx \tag{5-53}$$

where the gain of the transmitting antenna (with respect to an isotropic antenna), G_{AT}, has been included. Table 5-8 also gives the effective area for several types of antennas. The gain and the effective area of an antenna are related by

$$G_A = \frac{41rA}{} \tag{5-54}$$

where $\lambda = c/f$ is the wavelength, c being the speed of light (3 x 10^8 m/sec) and f the operating frequency (hertz). An antenna is a *reciprocal element.* That is, it has the same gain properties whether it is transmitting or receiving. Substituting (5-54) into (5-53), we obtain

$$\frac{P_{Rx}}{P_{Tx}} = G_{AT}\left(\frac{\lambda}{4\pi d}\right)^2 G_{AR} \tag{5-55}$$

where the free-space gain is

$$G_{FS} = (\frac{\lambda}{4,rd})2 = \frac{1}{L_{FS}} \tag{5-56}$$

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and L_{ps} is the free-space path loss (absolute units). Often the channel gain, expressed in decibel units, is obtained from (5-50):

$$(G_{channel})_{dB} = (G_{AT})_{dB} - (L_{ps})_{dB} + (G_{AR})_{dB} \tag{5-57}$$

where the free-space loss is

$$(L_{FS})_{dB} = 20 \log\left(\frac{4\pi d}{\lambda}\right) \quad \text{dB} \tag{5-58}$$

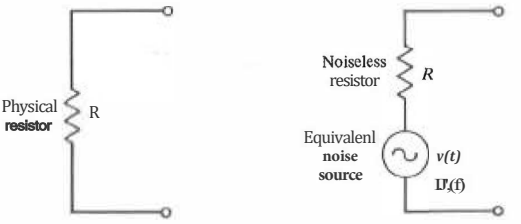
For example, the free-space loss at 4 GHz for the shortest path to a synchronous satellite from Earth (22,300 mi) is 195.6 dB.

Thermal Noise Sources

The noise power generated by a thermal noise source will be studied since the receiver noise is evaluated in terms of this phenomenon. A conductive element with two terminals may be characterized by its resistance, R ohms. This resistive or lossy element contains free electrons that have some random motion if the resistor has a temperature above absolute zero. This random motion causes a noise voltage to be generated at the terminals of the resistor. Although the noise is small, when the noise is amplified by a high-gain receiver it can become a problem. (If no noise were present, we could communicate to the edge of the universe with infinitely small power since the signal could always be amplified without having noise introduced.)

This physical lossy element, or physical resistor, can be modeled by an equivalent circuit that consists of a noiseless resistor in series with a noise voltage source (see Fig. 5-35). From quantum mechanics, it can be shown that the (normalized) power spectrum corresponding to the voltage source is [Van der Ziel, 1986]

$$!!\}\backslash u(f) = \frac{2R[h2]ff}{} + \frac{h\,[\,]+}{\text{chlff}/(kT)} - 1\,] \tag{5-59}$$



(a) A Lossy Circuit Element Characterized by Its Resistance

(b) Thevenin Equivalent Circuit Model of the Lossy Circuit Element

FIGURE 5-35 Thermal noise source.

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where R = value of the physical resistor (ohms)
h = 6.2 x 10^{-34} J-sec is Planck's constant
k = 1.38 x 10^{-23} J/K is Boltzmann's constant, where K is degrees Kelvin
T = (273 + C) is the absolute temperature of the resistor (kelvin)

At room temperature for frequencies below 1000 GHz, [hlfl/(kT)] < 1/5, so that $e^x = 1 + x$ is a good approximation. Then (5-59) reduces to

$$\text{<Tb(f)} = 2RkT \tag{5-60}$$

This equation will be used to develop other formulas in this text since the RF frequencies of interest are usually well below 1000 GHz and we are not dealing with temperatures near absolute zero.

If the open-circuit noise voltage that appears across a physical resistor is read by a true rms voltmeter that has a bandwidth of B hertz, then, using (2-67), the reading would be

$$V_{rms} = Y(v2) = \sqrt{2 \int_0^B \text{<Tb(f)} \, df} = Y4kTBR \tag{5-61}$$

Characterization of Noise Sources

Noise sources may be characterized by the maximum amount of noise power or PSD that can be passed to a load.

DEFINITION: The *available noise power* is the *maximum*t actual (i.e., not normalized) power that can be drawn from a source. The *available PSD* is the *maximum* actual (i.e., not normalized) PSD that can be obtained from a source.

For example, the available PSD for a thermal noise source is easily evaluated using Fig. 5-36 and (2-142).

$$\text{<Ta(f)} = \text{<Tb(f)} \, |H(f)|^2 = \tfrac{1}{2}kT \quad \text{W/Hz} \tag{5-62}$$

where $H(f) = \frac{1}{2}$ for the resistor divider network. The available power from a thermal source in a bandwidth of B hertz is

$$P_a = \int_{-B}^{B} \text{<Ta(f)} \, df = \int_{-B}^{B} \frac{1}{2} kT \, df \tag{5-63}$$

or

$$P_a = kTB$$

This equation indicates that the available noise power from a thermal source does *not* depend on the value of R, even though the open-circuit rms voltage does.

tThe maximum power or maximum PSD is obtained when $Z_L(f) = Z;(f)$, where $Z_L(f)$ is the load impedance and $z;(f)$ is the conjugate of the source impedance.

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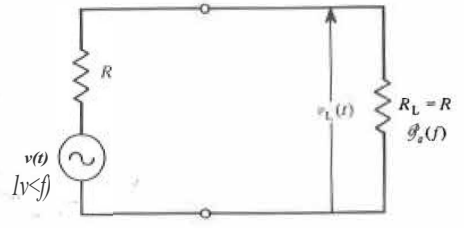


FIGURE 5-36 A thermal source with a matched load.

The available noise power from a source (it does not have to be a thermal source) can be specified by a number called the noise temperature.

DEFINITION: The *noise temperature* of a source is given by

$$T = \frac{P_a}{kB} \tag{5-64}$$

where P_a is the available power from the source in a bandwidth of B hertz.

In using this definition, it is noted that if the source happens to be of thermal origin, T will be the temperature of the device in kelvins, but if the source is not of thermal origin, the number obtained for T may have nothing to do with the physical temperature of the device.

Noise Characterization of Linear Devices

A linear device with internal noise generators may be modeled as shown in Fig. 5-37. Any device that can be built will have some internal noise sources. As shown in the figure, the device could be modeled as a noise-free device having a power gain Ga<f) and an excess noise source at the output to account for the internal n ise of the actual device. Some examples of linear devices that have

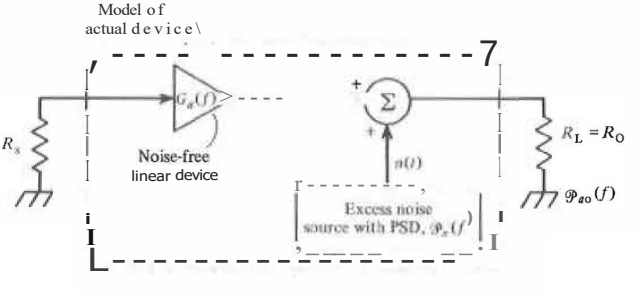


FIGURE 5-37 A noise model for an actual device.

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to be characterized in receiving systems are lossy transmission lines, RF amplifiers, down converters, and IF amplifiers.

The power gain of the devices is the available power gain.

DEFINITION: The *available power gain* of a linear device is

$$G_a(f) = \frac{\text{available PSD out of the device}}{\text{available PSD out of the source}} = \frac{\mathcal{P}_{ao}(f)}{\mathcal{P}_{ai}(f)} \tag{5-65}$$

When $\mathcal{P}_{ao}(f)$ is measured to obtain $G_a(f)$, the source noise power is made large enough so that the amplified *source* noise that appears at the output dominates any other noise. In addition, note that $G_a(f)$ is defined in terms of actual (i.e., not normalized) PSD. In general, $G_a(f)$ will depend on the driving source impedance as well as on elements within the device itself, but it does *not* depend on the load impedance. If the source impedance and the output impedance of the device are equal, $G_a(f) = |H(f)|^2$, where $H(f)$ is the voltage or current transfer function of the linear device.

To characterize the goodness of a device, a figure of merit is needed that compares the actual (noisy) device with an ideal device (i.e., no internal noise sources). Two figures of merit, both of which tell us the same thing-namely, how bad the noise performance of the actual device is-are universally used. They are noise figure and effective input-noise temperature.

DEFINITION: The *spot noise figure* of a linear device is obtained by terminating the device with a thermal noise source of temperature T_0 on the input and a matched load on the output as indicated in Fig. 5-37. The spot noise figure is

$$F_s(f) = \frac{\text{measured available PSD out of the actual device}}{\text{available PSD out of an ideal device with the same available gain}}$$

or

$$F_s(f) = \frac{\mathcal{P}_{ao}(f)}{(kT_0/2)G_a(f)} = \frac{(kT_0/2)G_a(f) + \mathcal{P}_x(f)}{(kT_0/2)G_a(f)} > 1 \tag{5-66}$$

The value of R_s is the same as the source resistance that was used when $G_a(f)$ was evaluated. A standard temperature of $T_0 = 290$ K is used as adopted by the IEEE [Haus, 1963].

$F_s(f)$ is called the *spot noise figure* since it refers to the noise characterization at a particular "spot" or frequency in the spectrum. Note that $F_s(f)$ is always greater than unity for an actual device, but it is nearly unity if the device is almost an ideal device. $F_s(f)$ is a function of the source temperature, T_0. Consequently, when the noise figure is evaluated, a standard temperature of $T_0 = 290$ K is used. This corresponds to a room temperature of 63 °F.

Often an average noise figure instead of a spot noise figure is desired. The average is measured over some bandwidth B.

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DEFINITION: The ***average noise figure*** is

$$F = \frac{P_{ao}}{kT_0 \int_{f_0 - B/2}^{f_0 + B/2} G_a(f)\, df} \tag{5-67}$$

where

$$P_{ao} = 2 \int_{f_0 - B/2}^{f_0 + B/2} g_{ao}(f)\, df$$

is the measured available output in a bandwidth B hertz wide centered on a frequency f_0 and T_0 = 290 K.

If the available gain is constant over the band so that $G_a(f) = G_a$ over the frequency interval $(f_0 - B/2) \le f \le (f_0 + B/2)$, the noise figure becomes[†]

$$F = \frac{P_{ao}}{kT_0 B G_a} \tag{5-68a}$$

The noise figure is often measured by using the Y-factor method. This technique is illustrated by Prob. 5-32 at the end of this chapter. The Hewlett-Packard HP8970B Noise Figure Meter uses the Y-factor method for measuring the noise figure of devices.

The noise figure can also be specified in decibel units,[†]

$$F_{dB} = 10 \log(F) = 10 \log\left(\frac{P_{ao}}{kT_0 B G_a}\right) \tag{5-68b}$$

For example, suppose that the manufacturer of an RF preamplifier specifies that an RF preamp has a 2-dB noise figure. This means that the actual noise power at the output is 1.58 times the power that would occur because of amplification of noise from the input.

The other figure of merit for evaluating the noise performance of a linear device is the effective input-noise temperature, which is illustrated in Figure 5-38.

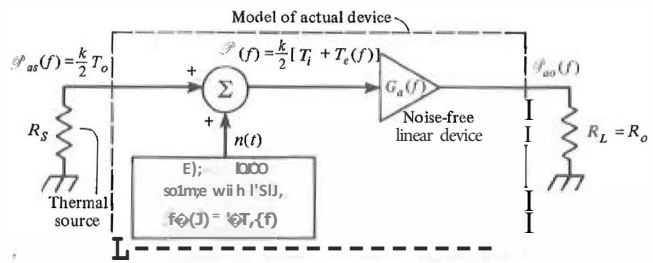


FIGURE 5-38 Another noise model for an actual device.

[†]some authors call F the ***noise factor*** and F_{dB} the ***noise figure.***

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DEFINITION: The *spot effective input-noise temperature,* $T_e(f)$, of a linear device is the *additional* temperature required for an input source, which is driving an ideal (noise-free) device, to produce the same available PSD at the ideal device output as is obtained from the actual device when it is driven by the input source of temperature T; kelvins. That is, $T_e(f)$ is defined by

$$\mathcal{P}_{ao}(f) = G_a(f)\, k\, [T_i + T_e(f)] \tag{5-69}$$

where $\mathcal{P}_{ao}(f)$ is the available PSD out of the actual device when driven by an input source of temperature T_i, and $T_e(f)$ is the spot effective input-noise temperature.

The *average effective input-noise temperature,* T_e, is defined by the equation

$$P_{ao} = k(T_i + T_e) \int_{f_0 - B/2}^{f_0 + B/2} G_a(f)\, df \tag{5-70}$$

where the measured available noise power out of the device in a band B hertz wide is

$$P_{ao} = 2 \int_{f_0 - B/2}^{f_0 + B/2} \mathcal{P}_{ao}(f)\, df \tag{5-71}$$

Since $G_a(f)$ depends on the source impedance as well as on the device parameters, $T_e(f)$ will depend on the source impedance used as well as on the characteristics of the device itself, but it is independent of the value of T; used, In the definition of T_e, note that the IEEE standards do not specify that $T_i = T_0$ since the value of T_e obtained does not depend on the value of T_i that is used. However, $T_i = T_0 = 290$ may be used for convenience. The effective input-noise temperature can also be evaluated by the Y-factor method, as illustrated by Prob. 5-32.

When the gain is flat (constant) over the frequency band, $G_a(f) = G_a$, the effective input-noise temperature is simply

$$T_e = \frac{P_{ao} - kT_i G_a B}{kG_a B} \tag{5-72}$$

Note that $T_e(f)$ and T_e are greater than zero for an actual device, but if the device is nearly ideal (small internal noise sources), they will be very close to zero.

When T_e was evaluated, an input source was used with some convenient value for T_i. However, when the device is used in a system, the available noise power from the source will be different if T_i is different. For example, suppose that the device is an RF preamplifier and the source is an antenna. The available power out of the amplifier when it is connected to the antenna is now†

$$P_{ao} = 2 \int_{f_0 - B/2}^{f_0} \mathcal{P}_a(f) G_a(f)\, df + kT_i \int_{f_0 - B/2}^{f_0} G_a(f)\, df \tag{5-73}$$

†The value of P_{ao} in (5-73) and (5-74) is different from that in (5-70), (5-71), and (5-72);

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where $W_{as}(f)$ is the available PSD out of the source (antenna). T, is the average effective input temperature of the amplifier that was evaluated by using the input source T_i, If the gain of.the amplifier is approximately constant over the band, this reduces to

$$P_{ao} = G_a P_{as} + kT_i B G_a \tag{5-74}$$

where the available power /rom the source (antenna) is

$$P_{as} = 2 \int_{f_c - B/2}^{f_c + B/2} W_{as}(f)\, df \tag{5-75}$$

Furthermore, the available power from the source might be characterized by its noise temperature, T_s, so that, using (5-64),

$$P_{as} = kT_s B \tag{5-76}$$

In satellite Earth station receiving applications, the antenna (source) noise temperature might be $T_s = 32$ K at 4 GHz for a parabolic antenna with a diameter of 5 m, where the noise from the antenna is due to cosmic radiation and to energy received from the ground as the result of the sidelobe beam pattern of the antenna. (The Earth acts as a blackbody noise source with $T = 280$ K.) Note that the $T_s = 32$ K of the antenna is "caused" by *radiation* resistance, which is not the same as a loss resistance (I^2R losses) associated with a thermal source and T_s has no relation to the physical temperature of the antenna.

Two figures of merit have been defined: noise figure and effective input-noise temperature. By combining (5-66) and (5-69) where $T_i = T_0$, a relationship between these two figures of merit for the spot measures is obtained:

$$T_e(f) = T_o[F_s(f) - 1) \tag{5-77a}$$

Here $T_i = T_0$ is required because $T_i = T_0$ is used in the definition for the noise figure that precedes (5-66).

Using (5-68a) and (5-72) where $T_i = T_0$, we obtain the same relationship for the average measures:

$$T_e = To(F - 1) \tag{5-77b}$$

EXAMPLE 5-5 T. AND F FOR A TRANSMISSION LINE

The effective input-noise. temperature, T_e, and the noise figure, F, for a lossy transmission line (a linear device) will now be evaluated. t This can be accomplished by terminating the transmission line with a source and a load resistance (all having the same physical temperature) that are both equal to the characteristic impedance of the line, as shown in Fig. 5-39. The gain of the transmission line is $G_a = 1/L$, where L is the transmission line loss (power in divided by power out) in absolute units (i.e., not dB units). Looking into the output port of the transmission line, one sees an equivalent circuit that is resistive (thermal

tThese results also hold for the T, and F of(impedance) matched attenuators.

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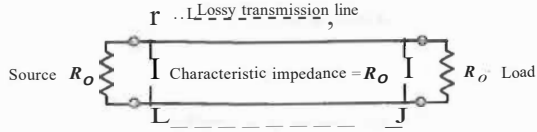


FIGURE 5-39 Noise figure measurement of a lossy transmission line.

source) with a value of R_0 ohms since the input of the line is terminated by a resistor of R_0 ohms (the characteristic impedance). Assume that the physical temperature of the transmission line is T_i, as measured on the Kelvin scale. Since the line acts as a thermal source, the available noise power at its output is $P_{oo} = kT_iB$. Using (5-72) where the source is at the same physical temperature, $T_i = T_i$, we get

$$T_e = \frac{kT_iB - kT_iG_a}{kGJJ} = T_i\left(\frac{1}{G_a} - 1\right)$$

Thus, the effective input-noise temperature for the transmission line is

$$T_e = T_L(L - 1) \tag{5-78a}$$

where T_i is the physical temperature (Kelvin) of the line and L is the line Joss. If the physical temperature of the line happens to be T_0, this becomes

$$T_i = T_0(L - 1) \tag{5-78b}$$

The noise figure for the transmission line is obtained by using (5-77b) to convert T_i to F. Thus, substituting (5-78a) into (5-77b), we get

$$T_i(L - 1) = T_0(F - 1)$$

Solving for F, we obtain the noise figure for the transmission line

$$F = 1 + \frac{T_i}{T_0}(L - 1) \tag{5-79a}$$

where T_i is the physical temperature (Kelvin) of the line, $T_0 = 290$, and L is the line Joss. If the physical temperature of the line is 290 K (63 °F), (5-79a) reduces to

$$F = \frac{1}{G_a} = L \tag{5-79b}$$

In decibel measure, this is $F_{da} = L_{da}$. In other words, if a transmission line has a 3-dB loss, it has a noise figure of 3 dB provided that it has a physical temperature of 63 °F. If the temperature is 32 °F (273 K), the noise figure, using (5-79a), will be 2.87 dB. Thus, F_{da} is approximately L_{da}, if the transmission line is located in an environment (temperature range) that is inhabitable by humans.

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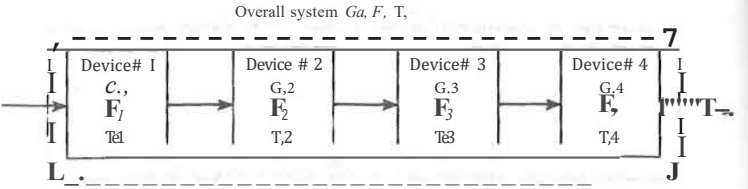


FIGURE 5-40 Cascade of four devices.

Noise Characterization of Cascaded Linear Devices

In a communication system several linear devices, supplied by different vendors, are often cascaded together to form an overall system, as indicated in Fig. 5-40. In a receiving system these devices might be an RF preamplifier connected to a transmission line that feeds a down converter and an IF amplifier. (As discussed in Sec. 4-3, the down converter is a linear device and may be characterized by its conversion power gain and its noise figure.) For system performance calculations, we need to evaluate the overall power gain, G_a, and the overall noise characterization (which is given by the overall noise figure or the overall effective input-noise temperature) from the specifications for the individual devices provided by the vendors.

The overall available power gain is

$$G_a(f) = G_{a1}(f)G_{a2}(f)G_{a3}(f)G_{a4}(f) \cdots \tag{5-80}$$

since, for example, for a four-stage system,

$$G_a(f) = \frac{\mathcal{P}_{ao4}}{\mathcal{P}_{as}} = \left(\frac{\mathcal{P}_{ao1}}{\mathcal{P}_{as}}\right)\left(\frac{\mathcal{P}_{ao2}}{\mathcal{P}_{ao1}}\right)\left(\frac{\mathcal{P}_{ao3}}{\mathcal{P}_{ao2}}\right)\left(\frac{\mathcal{P}_{ao4}}{\mathcal{P}_{ao3}}\right)$$

THEOREM: *The overall noise figure for cascaded linear devices* is

$$F = F_1 + \frac{F_2 - 1}{G_{a1}} + \frac{F_3 - 1}{G_{a1}G_{a2}} + \frac{F_4 - 1}{G_{a1}G_{a2}G_{a3}} + \cdots \tag{5-81}$$

as shown in Fig. 5-40.

Proof. This result may be obtained by using an excess-noise model of Fig. 5-37 for each stage. We will prove the result for a two-stage system as modeled in Fig. 5-41. The overall noise figure is

$$F = \frac{P_{ao2}}{(P_{ao2})_{\text{ideal}}} = \frac{P_{x2} + p_{ao}\,\text{IG}_{a}2}{G_{a}\,\text{IG}_{a}2P_{as}}$$

which becomes

$$F = \frac{p_{x2} + G_{a2}(P_{x1} + G_{a1}P_{a,})}{G_{a}\,\text{IG}_{a}2P_{as}} \tag{5-82}$$

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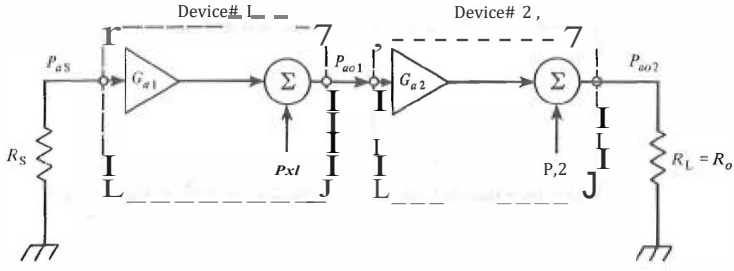


FIGURE 5-41 Noise model for two cascaded devices.

where $P_{as} = kT_0B$ is the available power from the thermal source. P_{x1} and P_{x2} can be obtained from the noise figures of the individual devices by using Fig. 5-37, so that for the ith device

$$F_i = \frac{P_{aoi}}{G_{ai}P_{as}} = \frac{P_{xi} + G_{ai}P_{as}}{G_{ai}P_{as}}$$

or

$$P_{xi} = G_{ai}P_{as}(F_i - 1) \tag{5-83}$$

Substituting this equation into (5-82) for P_{x1} and P_{x2}, we obtain

$$F = F1 + \frac{F_2 - 1}{G_{a1}}$$

which is identical to (5-81) for the case of two cascaded stages. In a similar way (5-81) can be shown to be true for any number of stages.

Looking at (5-81), we see that if the terms $G_{a1} \cdot c_{a}, c_{a}2 \cdot G_{a}P_{a2}G_{a3}$, and so on, are relatively large, F_1 will dominate the overall noise figure. Thus, in n,ceiving system design, it is important that the first stage have a low noise figure and a large available gain so that the noise figure of the overall system will be as small as possible.

The overall effective input-noise temperature of several cascaded stages can also be evaluated.

THEOREM: *The overall effective input-noise temperature for cascaded linear devices is*

$$T_e = T_{e1} + \frac{T_{e2}}{G_{a1}} + \frac{T_{e3}}{G_{a1}G_{a2}} + \frac{T_{e4}}{G_{a1}G_{a2}G_{a3}} + \cdots \tag{5-84}$$

as shown in Fig. 5-40.

A proof of this result is left for the reader as a homework exercise.

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For further study concerning the topics of effective input-noise temperature and noise figure, the reader is referred to an authoritative monograph [Mumford and Scheibe, 1968).

Link Budget Evaluation

The performance of a communication system depends on how large the signa!-to-noise (S/N)'ratio.is at the detector input in the receiver. It is engineering custom to call the signal-to-noise ratio before detection the *carrier-to-noise* ratio (C/N). Thus, in this section, we will use C/N to denote the signal-to-noise ratio before detection (bandpass case) and S/N to denote the signal-to-noise ratio after detection (baseband case). Here we are interested in evaluating the detector input C/N as a function of the communication link parameters, such as transmitted EIRP, space loss, receiver antenna gain, and the effective input-noise temperature of the receiving system. The formula that relates these system link parameters to the C/N at the detector input is called the *link budget.*

The communication system may be described by the block diagram shown in Fig. 5-42. In this model, the receiving system from the receiving-antenna output to the detector input is modeled by one linear block that represents the cascaded stages in the receiving system, such as a transmission line, a low-noise amplifier (LNA), a down converter, and an IF amplifier. This linear block describes the overall available power gain and effective input-noise temperature

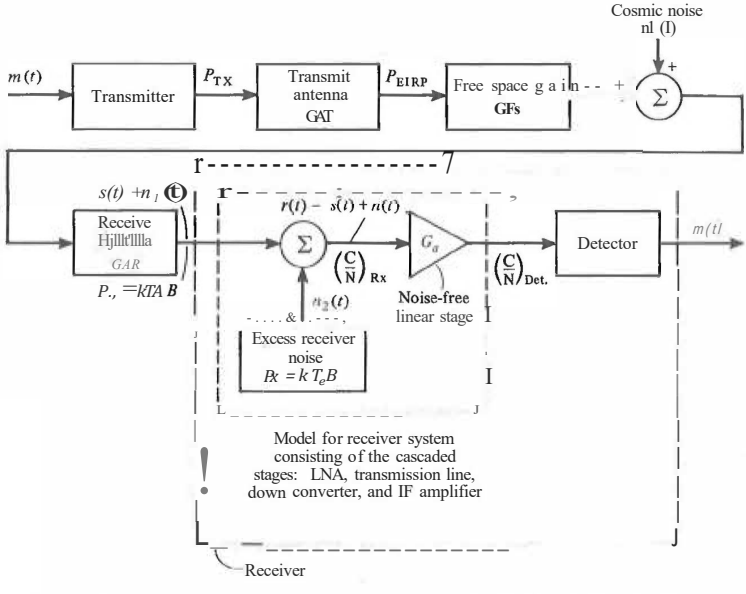


FIGURE 5-42 Communication system model for link budget evaluation.

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of these cascaded devices, as described in the preceding section and modeled in Fig. 5-39.

As shown in Fig. 5-42, the C/N at the ideal amplifier input (with gain G_a) is identical to that at the detector input since the ideal amplifier adds no excess noise and it amplifies the signal and noise equally well in a bandwidth of B hertz (the IF bandwidth). Thus, denoting this simply by C/N,

$$\left(\frac{C}{N}\right) \triangleq \left(\frac{C}{N}\right)_{Rx} = \left(\frac{C}{N}\right)_{det} \tag{5-85}$$

where these carrier-to-noise power ratios are indicated on the figure.

Using (5-50) and (5-51), we obtain the received signal power

$$C_{RX} = (PEJ_{Rp})G_{FS}G_{AR} \tag{5-86}$$

where P_{EIRP} is the EIRP from the transmitter, G_{FS} is the free-space gain, and G_{AR} is the receiving antenna power gain.

When we use (5-64), the available noise power at the input to the ideal amplifier in the model (Fig. 5-42) is

$$N = kT_{syst}B \tag{5-87}$$

where B is the IF bandwidth. The receiving system noise temperature is

$$T_{yst} = T_{AR} + T, \tag{5-88}$$

where T_{AR} is the noise temperature of the antenna (due to received cosmic noise and Earth blackbody radiation) and T_e is the effective input-noise temperature of the overall receiving system.

When (5-86) and (5-87) are combined, the carrier-to-noise ratio at the detector input is

$$\frac{C}{N} = \frac{P_{EIRP}G_{FS}G_{AR}}{kT_{syst}B} \tag{5-89}$$

For engineering applications, this formula is converted to decibel units. Thus the received carrier-to-noise ratio at the detector input in decibels is

$$\left(\frac{C}{N}\right)_{db} = (P_{EIRP})_{dBw} - (L_{FS})_{dB} + \left(\frac{G_{AR}}{T_{syst}}\right)_{dB} - k_{dB} - B_{dB} \tag{5-90}$$

where $(P_{EIRP})_{dBw} = 10\log(P_{EIRP})$ is the EIRP of the transmitter in dB above 1 W

$(L_{FS})_{da} = 20\log[(4T1'd)/X.]$ is the path loss

$k_{da} = 10\log(1.38 \times 10^{-23}) = -228.6$

$B_{da} = 10\log(B)$ (B is the IF bandwidth in hertz)

For analog communication systems, the S/N at the detector output can be related to the C/N at the detector input. The exact relationship depends on the type of detector as well as the modulation used. These relationships will be developed in Chapter 7, where statistical concepts are applied. Example 5-6 uses (5-90) to evaluate the performance of a CATV satellite receiving system.

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E_b/N_0 Link Budget for Digital Systems

In digital communication systems the probability of bit error P_e for the digital signal at the detector output describes the quality of the recovered data. The P_e is a function of the ratio of the energy per bit to the noise PSD (E_b/N_0) as measured at the detector input. The exact relationship between P_e and E_b/N_0 depends on the type of digital signaling used, as will be developed in Chapter 7 using Stl:!tistical concepts. In this section we evaluate the E_b/N_0 obtained at the detect input as a function of the communication link parameters.

The energy per bit is given by $E_b = CT_b$, where C is the signal power and T_b is the time required to send one bit. Using (5-64), we see that the noise PSD (one-sided) is $N_0 = kT_{syst}$. Thus

$$\frac{C}{N} = \frac{E_b/T_b}{N_0 B} = \frac{E_b R}{N_0 B} \tag{5-91}$$

where $R = 1/T_b$ is the data rate (bits/sec). Using (5-91) in (5-89), we have

$$\frac{E_b}{N_0} = \frac{P_{EIRP} G_{FS} G_{AR}}{kT_{sys} R} \tag{5-92}$$

In decibel units, the E_b/N_0 received at the detector input in a digital communications receiver is related to the link parameters by



$$\left(\frac{E_b}{N_0}\right)_{dB} = (EIRP)_{dBW} - (L_{ps})_{dB} + \left(\frac{G_{AR}}{T_{sys}}\right)_{dB} - k_{dB} - R_{dB} \tag{5-93}$$

where $(R)_{dB} = 10 \log(R)$ and R is the data rate (bits/second).

For example, suppose that we have BPSK signaling and an optimum detector is used in the receiver; then $(E_b/N_0)_{dB} = 8.4$ dB is required for $P_e = 10^{-4}$ (see Fig. 7-14).t Using (5-93), the communication link parameters may be selected to give the required $(E_b/N_0)_{dB}$ of 8.4 dB. Note that as the bit rate is increased, the transmitted power has to be increased or the receiving system performance——denoted by $(G_{AR}/T_{sys})_{dB}$- has to be improved to maintain the required $(E_b/N_0)_{dB}$ of 8.4 dB.

EXAMPLE 5-6 LINK BUDGET EVALUATION FOR A TELEVISION RECEIVE-ONLY TERMINAL FOR SATELLITE SIGNALS

The link budget for a television receive-only (TYRO) terminal used for reception of TV signals from a satellite will be evaluated. It is assumed that this receiving terminal is located at Washington, D.C., and is receiving signals from a Hughes Galaxy satellite that is in a geostationary orbit at 134° W longitude above the equator. The specifications on the proposed receiving equipment are given in Table 5-9 together with the receiving site coordinates and satellite parameters. This TYRO terminal is typical of those used at the head end of CATV systems

tlf, in addition, coding were used with a coding gain of 3 dB (see Sec 1-10), then an E_b/N_0 of 5.4 dB would be required for $P_e = 10^{-6}$.

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TABLE 5-9 Typical Satellite Television Relay Parameters

Item	Parameter Value
Hughes Galaxy I satellite	
Orbit	Geostationary
Location	134° W longitude
Uplink frequency band	6 GHz
Downlink frequency band	4GHz
(PEIRP)dBw	36 dBw
TYRO terminal	
Site location	Washington, D.C., 38.8° N latitude, 77° W longitude
Antenna	
Antenna type	5-m parabola
Noise temperature	32 K (at feed output) for 16.8° elevation
Feedline gain	0.98
Low-noise amplifier	
Noise temperature	120 K
Gain	50 dB
Receiver	
Manufacturer	Microdyne Corp., Model 1100-TVR (x 24)
Noise temperature	2610 K
IF bandwidth	30 MHz
FM threshold	8 dB C/N

for reception of TV signals that are relayed via satellite. As discussed in Sec. 5-4, the composite NTSC baseband video signal is relayed via satellite by frequency-modulating this signal on a 6-GHz carrier that is radiated to the satellite. The satellite down-converts the FM signal to 4 GHz and retransmits this FM signal to the TYRO terminal.

The transmit antenna EIRP footprint for the Galaxy I satellite is shown in)Fig. 5-43. From this figure it is seen that the EIRP directed toward Washington, D.C., is approximately 36 dBW, as obtained for inclusion in Table 5-9.

It can be shown that the *look angles* of the TYRO antenna can be evaluated using the followirig formulas [Davidoff, 1984, Chap. 9]. The elevation of the TYRO antenna is

$$E = \tan^{-1}\left[\frac{1}{\tan \beta} - \frac{R}{(R + h)\sin \beta}\right] \tag{5-94}$$

where

$$\beta = \cos^{-1}[\cos \phi \cos \lambda] \tag{5-95}$$

and Xis the difference betweenlhe longitude of the TYRO site and the longitude of the *substation point* (i.e., the point directly below the geostationary satellite along the equator). <Pis the latitude of the TYRO site, R = 3963 statute miles is the Earth radius, and h = 22,242 statute miles is the altitude of the synchro-

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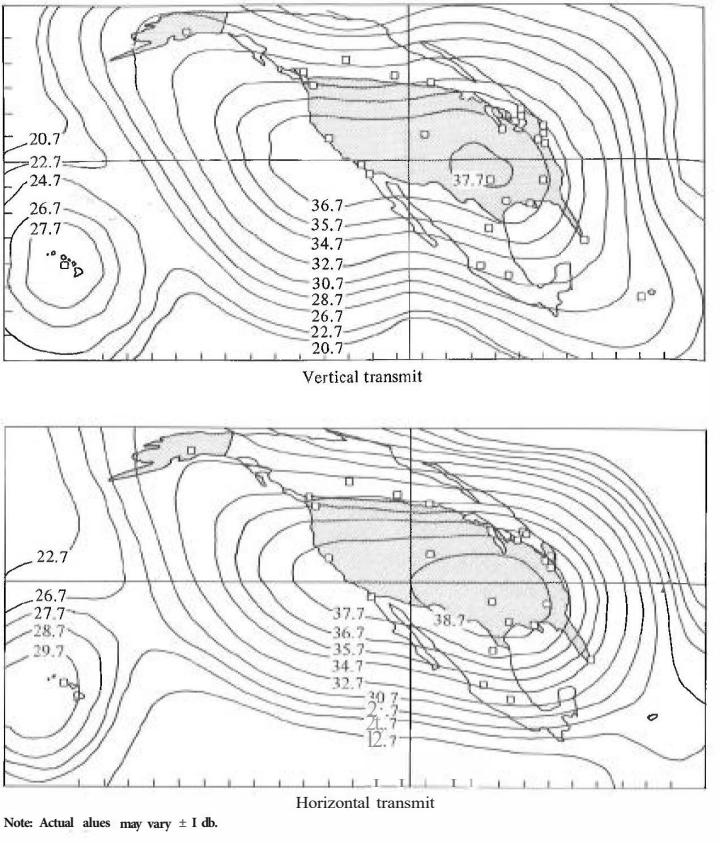


FIGURE 5-43 Galaxy I (134° W longitude) antenna pattern EIRP footprint (contolllrs given in dBW units). [Figure courtesy of Hughes Communications, Inc., a wholly-owned subsidiary of Hughes Aircraft Company.)

nous satellite. The distahce between the TYRO site and the satellite, called the *slant range,* is given by the cosine law:

$$d = \sqrt{(R + h)^2 + R^2 - 2R(R + h) \cos \beta} \tag{5-96}$$

The azimuth of the TYRO antenna is

$$A = \cos^{-1}\left(\frac{\tan \phi}{\tan \beta}\right) \tag{5-97}$$

where the true azimuth as measured clockwise from the north is given by either A or $360^\circ - A$, as appropriate, since a calculator gives a value between O and

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180° for $\cos^{-1}(x)$. Using these formulas for the Washington, D.C., TYRO site, we get $X = 134 - 77 = 57^\circ$ and

$$= \cos^{-1}[\cos(38.8)\cos(57)] = 64.9^\circ \tag{5-98a}$$

The slant range is

$$d = \sqrt{(26{,}205)^2 + (3963)^2 - 2(3963)(26{,}205)\cos(64.9)} \tag{5-98b}$$
$$= 24{,}784 \text{ miles}$$

The elevation angle of the TYRO antenna is

$$E = \tan^{-1}\left[\frac{1}{\tan(64.9) - \dfrac{3963}{(26{,}205)\sin(64.9)}}\right] = 16.80 \tag{5-98c}$$

The azimuth angle of the TYRO antenna is

$$A = \cos^{-1}\left(\frac{\tan(38.8)}{\tan(64.9)}\right) = 247.9^\circ \tag{5-98d}$$

Thus, the look angles of the TYRO antenna at Washington, D.C., to the Galaxy I satellite are $E = 16.8^\circ$ and $A = 247.9^\circ$.

A block diagram describing the receiving system is shown in Fig. 5-44. Using (5-90), the carrier-to-noise ratio at the detector input of the receiver is

$$\left(\frac{C}{N}\right)_{dB} = (EIRP)_{dBW} - (L_{pS})_{dB} + \left(\frac{G_{AR}}{T_{sys}}\right)_{dB} - k_{dB} - B_{dB} \tag{5-99}$$

The corresponding path loss *for* 24,784 mi at a frequency of 4 GHz is

$$(L_{pS})_{dB} = 20 \log\left(\frac{4\pi d}{\lambda}\right) = 196.5 \text{ dB}$$

For a 5-m parabola, using Table 5-8, the receiving antenna gain is

$$(G_{AR})_{dB} = 10 \log[\eta(\pi D/\lambda)^2] = 10 \log(24{,}434) = 44 \text{ dB}$$

Using (5-88) and (5-84), we obtain the system noise temperature

$$T_{syst} = T_{AR} + T_{reed} + \frac{T_{LNA}}{G_{feed}} + \frac{T_{receiver}}{G_{reed}G_{LNA}} \tag{5-100}$$

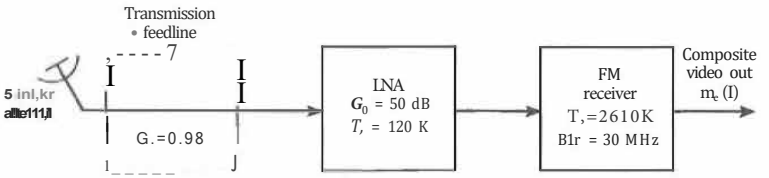


FIGURE 5-44 TVRO terminal.

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where the specified antenna noise temperature, including the feed, is T_A = $(T_{AR} + T_{feed}) = 32$ K.

The T_{AR} is *not* the ambient temperature of the antenna (which is about 290 K) because the effective resistance of the antenna does not consist of a thermal resistor but of a radiation resistor. That is, if RF power is fed into the antenna it will not be dissipated in the antenna but instead will be radiated into space. Consequently, T_{AR} is called the *sky noise temperature* and is a measure of the amount of cosmic noise power that is received from outer space and noise caused by atmospheric attenuation, where this power is $P_{AR} = kT_{ARB}$. Graphs giving the measured sky noise temperature as a function of the RF frequency and the elevation angle of the antenna are available [Jordan, 1985, Chap. 27; Pratt and Bostian, 1986].

Substituting the receive system parameters into (5-100), we have

$$T_{sys} = 32 + \frac{120}{0.98} + \frac{2610}{(0.98)(100{,}000)} = 154.5 \text{ K}$$

Thus†

$$\left(\frac{G_{AR}}{T_{sys}}\right)_{dB} = 10 \log\left(\frac{434}{154.5}\right) = 22 \text{ dB/K}$$

Also,

$$(B)_{dB} = 10 \log(30 \times 10^6) = 74.8$$

Substituting these results into (5-99) yields

$$\left(\frac{C}{N}\right)_{dB} = 36 - 196.5 + 22 - (-228.6) - 74.8$$

Thus the carrier-to-noise ratio at the detector input (inside the FM receiver) is



$$\left(\frac{C}{N}\right)_{dB} = 15.3 \text{ dB} \tag{5-101}$$

A C/N of 15.3 dB does not seem to be a very good signal-to-noise ratio, does it? Actually, the question is: Does a C/N of 15.3 dB at the detector input give a high-quality video signal at the detector output? The theory that answers this question is quite involved. (We will analyze the performance of an FM detector in Chapter 7.) However, this question can be answered easily if we have some detector performance curves as shown in Fig. 5-45. Two curves are plotted in this figure. One is for the output of a receiver that uses an FM discriminator for the FM detector, and the other is for a receiver (Microdyne

†For direct-to-the-home TV broadcast satellites (DBS), $G_{AR}/T_{sys} = 12$ dB/K would be required [Pritchard and Kase, 1981].

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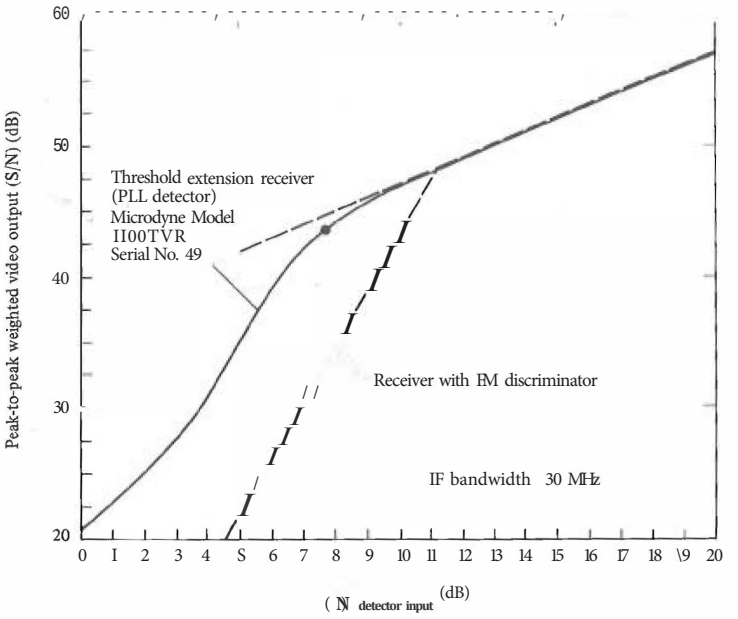


FIGURE 5-45 Output S/N as a function of FM detector input C/N. [Curve for the Microdyne receiver, Model 1100TVR, courtesy of the Microdyne Corporation, Ocala, FL, 1981.]

1100TVR) that uses a PLL FM detector to provide threshold extension.† For a C/N of 15.3 dB on the input, both receivers will give a 52-dB output S/N, which corresponds to a high-quality picture. However, if the receive signal from the satellite fades because of atmospheric conditions, and so on, by, say, 6 dB, the output S/N of the receiver with an FM discriminator will be only about 40 dij, whereas the receiver with threshold extension detection will have an output S/N of about 45 dB.

5-6 Fiber Optic Systems

Since 1980 fiber optic cable transmission systems have become commercially viable. This is indicated in Table 5-7, which shows that common carriers have installed extensive fiber optic systems in the United States, Japan, Britain, and

†The threshold point is the point where the knee occurs in the $(S/N)_{out}$ versus $(C/N)_{in}$ characteristic. In the characteristic for threshold extension receivers, this knee is extended to the left (improved performance) when compared to the knee for the corresponding receiver that uses a discriminator as a detector.

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To place a call the following sequence of events occurs.

1. The cellular subscriber initiates a call by keying in the telephone number of the called party and then presses the *send* key.
2. The **MTSO** verifies that the telephone number is valid and that the user is authorized to place a call.
3. The MTSO issues instructions to the user's cellular phone indicating which radio channel to use.
4. The MTSO sends out a signal to the called party to ring his or her phone. All of these operations occur within 10 sec of initiating the call.
5. When the called party answers, the MTSO connects the trunk lines for the two parties and initiates billing information.
6. When one party hangs up, the MTSO frees the radio channel and completes the billing information.

While a call is in progress, the cellular subscriber may be moving from one cell area to another, so the MTSO

- Monitors the signal strength from the cellular telephone as received at the cell base station. If the signal drops below some designated level, the MTSO initiates a "hand-off" sequence.
- For "hand-off," the MTSO inquires about the signal strength as received at adjacent cell sites.
- When the signal level becomes sufficiently large at an adjacent cell site, the MTSO instructs the cellular radio to switch over to an appropriate channel for communication with that new cell site. This switching process takes less than 250 msec and usually is unnoticed by the subscriber.

Presently, cellular systems and the standards for these systems are in a process of evolution throughout the world. The first commercial cellular service was implemented in the United States in 1983. As described in Table 5-13, systems are also operational in the United Kingdom, Germany, and Japan. The European Conference of Postal and Telecommunications Administration (CEPT) hopes to adopt a single cellular standard for all of Europe, and the MATS-E standard has been proposed. In addition, there is a push for better spectral utilization by using low-bit-rate encoding of VF signals to digitally modulate the carrier instead of using analog FM. Possible speech encoding methods are subband coding at 24 kb/sand adaptive prediction coding at 16 kb/s as described in Sec. 3-7. There is also interest in developing worldwide cellular coverage via spot beams from satellites.

5-9 Television

Television (TV) is a method of reproducing fixed or moving visual images by the use of electronic signals. There are numerous kinds of TV systems and different standards have been adopted around the world.

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In this section, we will concentrate on broadcast TV as it has developed in the United States since the 1930s. Black-and-white television will be described first, and then color television will be examined.

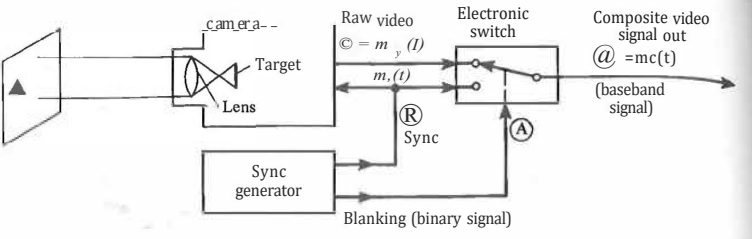
Black-and-White Television

The black-and-white image of a scene is essentially the intensity of the light as a function of the x and y coordinates of the scene and of time. However, an electrical waveform is only a function of time, so that some means of encoding the x and *y* coordinates associated with the intensity of light from a single point in the scene must be used. In broadcast TV this is accomplished by using *raster scanning,* as shown in Fig. 5-51. It is assumed that the whole scene is scanned before any object in the scene has moved appreciably. Thus moving images can be transmitted by sending a series of still images, just as in motion pictures. As the scene is scanned, its intensity is signified by the amplitude of the video signal. A synchronizing pulse is inserted at the end of each scan line by the electronic switch, as shown in Fig. 5-51, to tell the TV receiver to start another scan line. For illustrative purposes, a black triangle is located in the upper left portion of the scene. In Figs. 5-51b and 5-51c, numbers are used to indicate the corresponding times on the video waveform, with the location of the point being scanned in the scene at that instant. The composite video signal is actually a hybrid signal that consists of a digital waveform during the synchronizing interval and an analog waveform during the video interval. Operating personnel monitor the quality of the video by looking at a *waveform monitor,* which displays the composite video waveform as shown in Fig. 5-51c.

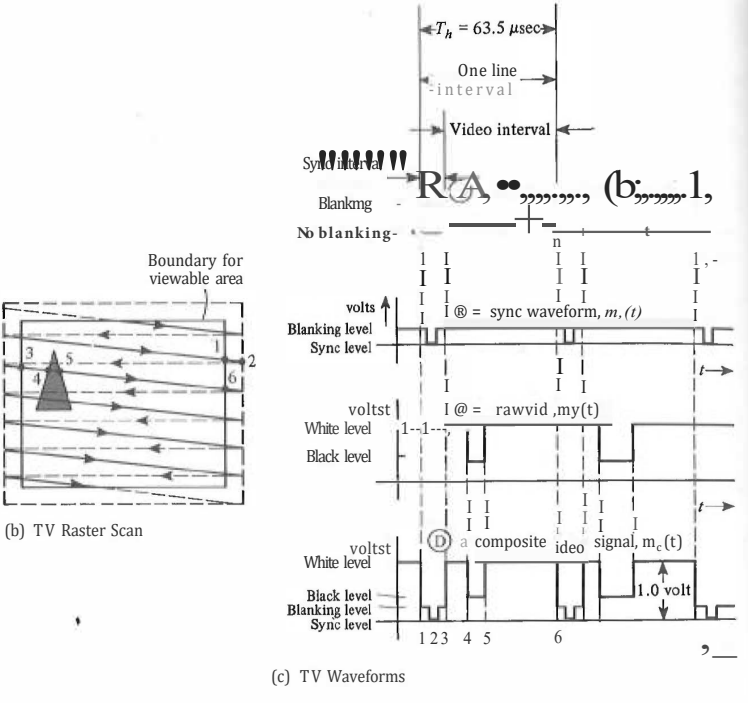
The picture (as viewed) is generated by scanning left to right to create a line, and the lines move from the top to the bottom of the screen. During retrace, indicated by the dashed lines from right to left, the CRT is turned off by the video level that is "blacker than black" during the sync pulse interval.t The scene is actually covered by scanning alternate lines from the top to the bottom and returning to the top to scan the missed lines. This technique is called *interlacing* and reduces the flicker effect on moving pictures, giving two fields (passes doWn the screen) for every frame. In the United States there are approximately 30 frames/sec transmitted, which is equivalent to 60 fields/sec for the 2:1 interlacing that is used. When the scan reaches the bottom of a field, the sync pulse is widened out to cover the whole video interval plus sync interval. This occurs for 21 lines: and, consequently, a black bar is seen across the screen if the picture is "rolled up" by misadjusting the vertical hold control. In the receiver, the vertical sync (wide pulse) can be separated from the horizontal sync (narrow pulse) by using a differentiator circuit to recover the horizontal sync and an integrator to recover the vertical sync. The vertical interval is also used to send special signals, such as testing signals that can be used by TV engineers to test the frequency response and linearity of the equipment on-line

tin addition, most TV sets switch off the electron beam during retrace by using the sync pulse obtained from the horizontal output (flyback) transformer.

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(a) Block Diagram of Camera System



(b) TV Raster Scan

(c) TV Waveforms

FIGURE 5-51 Generation of a composite black-and-white video signal.

[Solomon, 1979]. In addition, some TV programs have closed-caption information that can be decoded on special TV sets used by deaf viewers. These sets insert subtitles across the bottom of the screen. This closed-captioning information is transmitted during line 21 of the vertical interval by inserting a 16-bit code word. Details of the exact sync structure are rigorously specified by the FCC (see Fig. 5-58).

The composite baseband signal consisting of the *luminance* (intensity) video signal $m_y(t)$ and the *synchronizing* signal $m_s(t)$ is described by

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$$m_c(t) = \begin{cases} m_s(t) & \text{during the sync interval} \\ m_v(t) & \text{during the video interval} \end{cases} \tag{5-108}$$

The spectrum of $m_c(t)$ is very wide since $m_c(t)$ contains rectangular synchronizing pulses. In fact, the bandwidth would be infinite if the pulses had perfect rectangular shapes. In order to reduce the bandwidth, the pulses are rounded as specified by FCC standards (see Fig. 5-58). This allows $m_c(t)$ to be filtered to a bandwidth of $B = 4.2$ MHz (U.S. standard). For a still picture, the exact spectrum can be calculated using Fourier series analysis (as shown in Sec. 2-5) since, in this case, $m(t)$ would be periodic with a period of approximately $T_0 = 1/30$ sec corresponding to the frame rate of 30 frames/sec (see Table 5-14 for exact values). Thus, for a still picture, the spectrum would consist of lines spaced at 30-Hz intervals. However, from Fig. 5-51, we can observe that there are dominant intervals of width T_h corresponding to scanned lines of the frame. Furthermore, the adjacent lines usually have similar waveshapes. Consequently, $m(t)$ is quasiperiodic with period T_h, and the spectrum consists of clusters of spectral lines that are centered on harmonics of the scanning frequency, $nf_h = n/T_h$ where the spacing of the lines within these clumps is 30 Hz. For moving pictures, the line structure of the spectrum "fuzzes out" to a continuous spectrum with spectral clusters centered about nf_h. Furthermore, between these clusters the spectrum is nearly empty. As we will see, these "vacant" intervals in the spectrum are used to transmit the color information for a color TV signal. This, of course, is a form of frequency-division multiplexing.

The resolution of TV pictures is often specified in terms of lines of resolution. The number of horizontal lines that can be distinguished from the top to the bottom of a TV screen for a horizontal line test pattern is called the *vertical line resolution.* The maximum number of distinguishable horizontal lines (vertical line resolution) is the total number of scanning lines in the raster less those not used for the image. That is, the vertical resolution is

$$n_v = (N_f - N_v) \quad \text{lines} \tag{5-109a}$$

where N_f is the total number of scanning lines per frame and N_v is the number of lines in the vertical interval (not image lines) for each frame. For U.S. standards (Table 5-14), the maximum vertical resolution is

$$n_v = 525 - 42 = 483 \text{ lines} \tag{5-109b}$$

The resolution in the horizontal direction is limited by the frequency response allowed for $m_c(t)$. For example, if a sine-wave test signal is used during the video interval, the highest sine-wave frequency that can be transmitted through the system will be $B = 4.2$ MHz (U.S. standard) where B is the system video bandwidth. For each peak of the sine wave a dot will appear along the horizontal direction as the CRT beam sweeps from left to right. Thus, the horizontal resolution for 2:1 interlacing is

$$n_h = 2B(T_h - T_b) \quad \text{lines} \tag{5-109c}$$

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where B is the video bandwidth, T_h is the horizontal interval, and T_b is the blanking interval (see Fig. 5-51). For United States standards, $B = 4.2$ MBz, $T_h = 63.5$ μs, and $T_b = 10.5$μs so that the horizontal resolution is

$$n_{ll} = 445 \text{ lines} \tag{5-109d}$$

Furthermore, because of poor video bandwidth and the poor interlacing characteristics of consumer TV sets, this horizontal resolution of 445 is usually not obtained in pi:a<;;tice.t At best, the U.S. standard of 445 x 483 does not provide very good resolution, and this poor resolution is especially noticeable on large-screen TV sets. [For comparison the enhanced graphics adapter (EGA) computer monitor standard provides a resolution of 640 x 480.] Consequently, there is a great incentive to adopt a *high-definition television* (HDTV) standard in which the resolution is at least twice that of the present TV standard so that it approaches the quality of 35 mm film. There are many realities to consider such as (1) the desirability of adopting a HDTV standard that is compatibie with the present system so that standard TV sets may be used to view a HDTV signal, (2) the FCC requirement that HDTV must operate within the present bands allocated for broadcast TV, and (3) the cost, which must be reasonable. Many HDTV standards have been proposed [Jurgen, 1988], but it is not clear which one, if any, will be universally accepted.

For broadcast TV transmission (U.S. standard), the composite video signal of (5-108) is inverted and amplitude modulated onto an RF carrier so that the AM signal is

$$s_v(t) = A_c[1 - 0.875m_c(t)] \cos \omega_c t \tag{5-110}$$

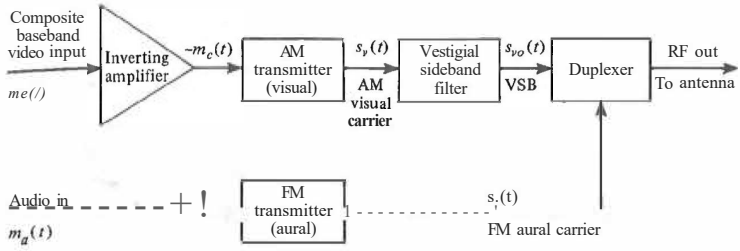
This is shown in Fig. 5-52. The lower sideband of this signal is a(tenuated so that the spectrum will fit into the 6-MHz TV channel bandwidth. This attenuation can be achieved by using a vestigial sideband filter, which is a bandpass filter that attenuates most of the lower sideband. The resulting filtered signal is called a *vestigial sideband* (VSB) signal, $s_{v\theta}(t)$. The discrete carrier term in the VSB signal changes when the picture is switched from one scene to another because the composite baseband signal has a nonzero de level that depends on the particular scene being televised.

The audio signal for the TV program is transmitted via a separate FM carrier located exactly 4.5 MHz above the visual carrier frequency.

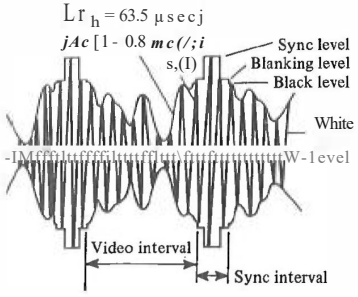
The rated power of a TV visual signal is the effective isotropic radiated peak envelope power (EIRP), i.e., effective sync tip average power, and is usually called simply the *effective radiated power* (ERP). The EIRP is the power that would be required to be fed into an isotropic antenna to get the same field strength as that obtained from an antenna that is actually used as measured in the direction of its maximum radiation. The antenna pattern of an omnidirectional TV station is shaped like a doughnut with the tower passing through the hole. (The antenna structure is actually at the top of the tower for maximum line-of-sight coverage.) The ERP of the TV station (visual or aural) signal is

tTypically, a horizontal resolution of about 300 lines is obtained.

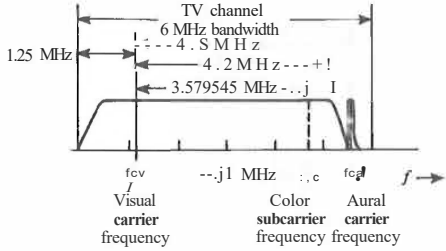
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(a) Equipment Block Diagram



(b) AM Visual Carrier Waveform, s,(I)
(Typical Video Modulation Illustrated)



(c) Magnitude Spectrum of RF Output
(Frequency Response Characteristic)

FIGURE 5-52 TV transmission system.

$$P_{ERP} = P_{PEP} G_A G_L \tag{5-111}$$

where P_{PEP} = peak envelope power out of the transmitter
G_A = power gain of the antenna (absolute units as opposed to decibel units) with respect to an isotropic antenna
G_L = total gain of the transmission line system from the transmitter output to the antenna (including the duplexer gain)

For a VHF TV station operating on channel 5, some typical values are $G_A = 5.7$, $G_L = 0.873$ (850 ft of transmission line plus duplexer), and a visual PEP of 20.1 kW. This gives an overall ERP of 100 kW, which is the maximum power licensed by the FCC for channels 2 to 6 (low-VHF band). For a UHF station operating on channel 20, some typical values are $P_{PEP} = 21.7$ kW, $G_A = 27$, $G_L = 0.854$ (850 ft of transmission line), which gives an ERP of 500 kW.

A simplified block diagram of a black-and-white TV receiver is shown in Fig. 5-53. The composite video, $m_c(t)$, as described earlier, plus a 4.5-MHz FM carrier containing the audio modulation, appears at the output of the envelope detector. The 4.5-MHz FM carrier is present because of the nonlinearity

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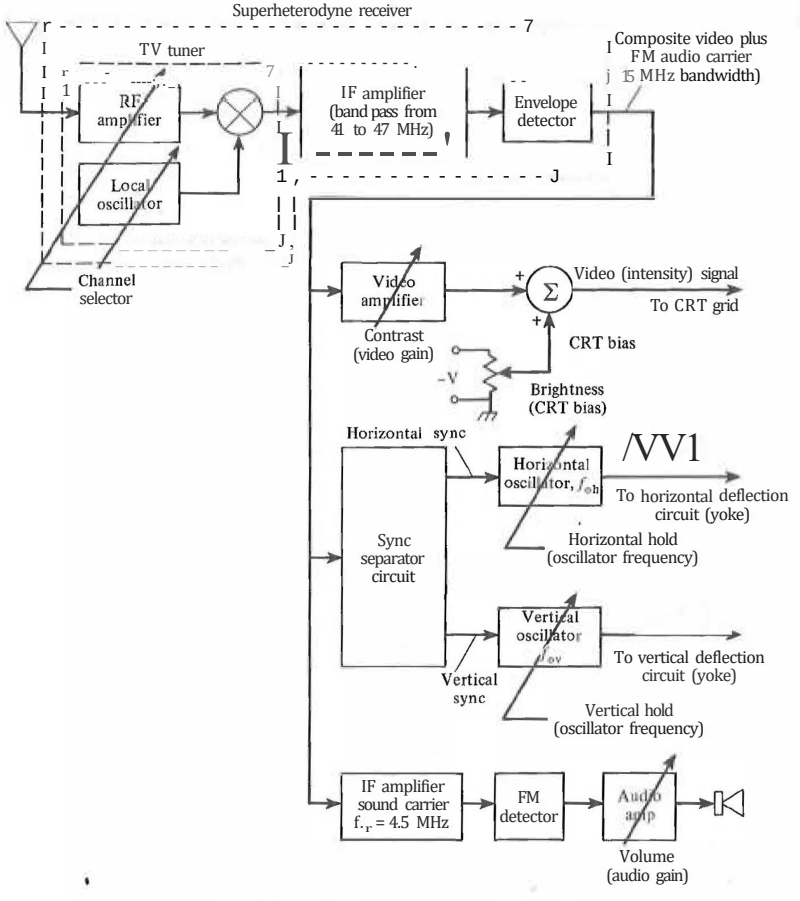


FIGURE 5-53 Black-and-white TV receiver.

of the envelope detector and because the detector input (IF signal) contains, among other terms, the FM aural carrier plus a discrete (i.e., sinusoidal) term, which is the video carrier, located 4.5 MHz away from the FM carrier signal. The intermodulation product of these two signals produces the 4.5-MHz FM signal, which is called the *intercarrier signal,* and contains the aural modulation. Of course, if either of these two input signal components disappears, the 4.5-MHz output signal will also disappear. This will occur if the white level of the picture is allowed to go below, say, 10% of the peak envelope level (sync tip level) of the AM visual signal. This is why the FCC specifies that the white level cannot be lower than (12.5 - 2.5)% of the peak envelope level (see Table

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5-4). When this occurs, a buzz will be heard in the sound signal since the 4.5-MHz FM carrier is disappearing at a 60-Hz (field) rate during the white portion of the TV scene. The sync separator circuit consists of a lossy integrator to recover the (wide) vertical sync pulses and a lossy differentiator to recover the (narrow) horizontal pulses. The remainder of the block diagram is self-explanatory. The user-available adjustments for a black-and-white TV are also shown.

MTS Stereo Sound

Stereo audio, also known as *multichannel television sound* (MTS), is also available and is accomplished using the FDM technique shown in Fig. 5-54. This FDM technique is similar to that used in standard FM broadcasting and studied in Example 5-2. For compatibility with monaural audio television receivers, the left plus right audio, $m_L(t) + m_R(t)$, is frequency modulated directly onto the aural carrier of the television station. Stereo capability is obtained by using a DSB-SC subcarrier with $m_L(t) - m_R(t)$ modulation. The subcarrier frequency is $2f_h$, where $f_h = 15.734$ kHz is the horizontal sync frequency for the video (see Table 5-14). Capability is also provided for a *second audio program* (SAP), such as audio in a second language. This is accomplished by using an FM subcarrier of frequency $5f_h$. dBx audio noise reduction processing is used on the audio on both the DSB-SC stereo subcarrier and the SAP FM subcarrier. (See Sec. 4-7 for a discussion of the dBx noise reduction technique.) Furthermore, a *professional channel* subcarrier at $6.5f_h$ is allowed for transmission of voice or data on a subscription service basis to paying customers. For further details on the FDM aural standards that have been adopted by television broadcasters and for other possible FDM formats, see a paper by Eilers [1985].

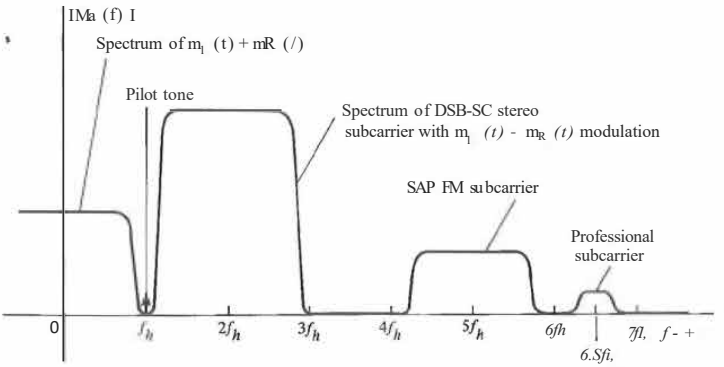


FIGURE 5-54 Spectrum of composite FDM baseband modulation on the FM aural carrier of U.S. TV stations.

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TABLE 5-14 U.S. Television Broadcasting Standards

Item	FCC Standard
Channel bandwidth	6 MHz
Visual carrier frequency	1.25 MHz ± 1000 Hz above the lower boundary of the channel
Aural carrier frequency	4.5 MHz ± 1000 Hz above the visual carrier frequency
Chrominance subcarrier frequency	3.579545 MHz ± 10 Hz
Aspect (width-to-height) ratio	Four units horizontally for every three units vertically
Modulation-type visual carrier	AM with negative polarity (i.e., a decrease in light level causes an increased real envelope level)
Aural carrier	FM with 100% modulation being l:iF = 25 kHz with a frequency response of 50 to 15,000 Hz using 75-μsec preemphasis
Visual modulation levels	
Blanking level	75% ± 2.5% of peak real envelope level (sync tip level)
Reference black level	7.5% ± 2.5% of the video range between blanking and reference white level below blanking level; this is called the ***setup*** level by TV engineers"
Reference white level	12.5% ± 2.5% of sync tip level•
Scanning	
Number of lines	525 lines/frame, interlaced 2:1
Scanning sequence	Horizontally: left to right; vertically: top to bottom
Horizontal scanning frequency	15,734.264 ± 0.044 Hz (2/455 of chrominance frequency); 15,750 Hz may be used during monochrome transmissions
Vertical scanning frequency	59.94 Hz (2/525 of horizontal scanning frequency); 60 Hz may be used during monochrome transmissions; 21 equivalent horizontal lines occur during the vertical sync interval of each field

"See Fig. 5-52b.

Color Television

Color television pictures are synthesized by combining red, green, and blue light in the proper proportion to produce the desired color. This is done by using a color CRT that has three types of phosphors-one type that emits red light, one that emits green, and a third type that emits blue light when they are struck by electrons. Therefore, three electronic video signals-$m_R(t)$, $m_0(t)$, and $m_8(t)$-

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are needed to drive the circuits that create the red, green, and blue light. Furthermore, a compatible color TV transmission system is needed such that black-and-white TV sets will receive the luminance video signal, $m_y(t)$, which corresponds to the gray scale of the scene. This compatibility between color and black-and-white transmission is accomplished by using a multiplexing technique analogous to that of FM stereo, except that three signals are involved in the problem instead of two.

The National Television System Committee (NTSC) compatible color TV system that was developed in the United States and adopted for U.S. transmissions in 1954 will be discussed. This technique is illustrated in Fig. 5-55. For compatible transmission, the waveforms that correspond to the intensity of the red, green, and blue components of the scene-$m_R(t)$, $m_G(t)$, and $m_B(t)$- are linearly combined to synthesize the equivalent black-and-white signal, $m_p(t)$, and two other independent video signals, called the *in-phase* and *quadrature* components, $m_i(t)$ and $m_q(t)$. (The names of these components arise from the quadrature modulation technique that is used to modulate them onto a subcar-

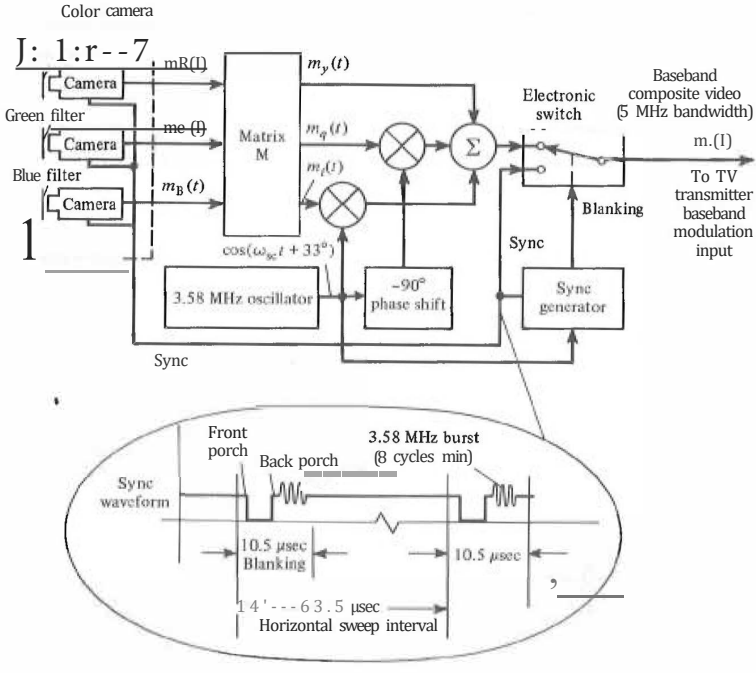


FIGURE 5-55 Generation of the composite NTSC baseband video signal. [Technically, $m_q(t)$ is low-pass filtered to an 0.5-MHz bandwidth and $m_i(t)$ is low-pass filtered to a 1.5-MHz bandwidth.]

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rier.) The exact equations used for these signals can be expressed in matrix form:

$$\begin{bmatrix} m_y(t) \\ m_i(t) \\ m_q(t) \end{bmatrix} = \underbrace{\begin{bmatrix} 0.3 & 0.59 & 0.11 \\ 0.6 & -0.28 & -0.32 \\ 0.21 & -0.52 & 0.31 \end{bmatrix}}_{M} \begin{bmatrix} m_R(t) \\ m_G(t) \\ m_B(t) \end{bmatrix} \tag{5-112}$$

where M is a 3 x 3 matrix that translates the red, green, and blue signals to the brightness, in-phase, and quadrature phase video (baseband) signals. For example, from (5-112), the equation for the *luminance* (black and white) signal is

$$m_y(t) = 0.3m_R(t) + 0.59m_G(t) + 0.11m_B(t) \tag{5-113}$$

Similarly, the equations for *m;(t)* and *m/t),* which are called the *chrominance* signals, are easy to obtain. In addition, note that if $m_R(t) = m_o(t) = m_8(t) = 1$, which corresponds to maximum red, green, and blue, then $m_y(t) = 1$, which corresponds to the white level in the black-and-white picture.

The chrominance components are two other linearly independent components, and if they are transmitted to the color TV receiver together with the luminance signal, these three signals can be used to recover the red, green, and blue signals using an inverse matrix operation. The bandwidth of the luminance signal needs to be maintained at 4.2 MHz to preserve sharp (high-frequency) transition in the intensity of the light that occurs along edges of objects in the scene. However, the bandwidth of the chrominance signals does not need to be this large since the eye is not as sensitive to color transitions in a scene. In NTSC standards the bandwidth of the m;(t) signal is 1.5 MHz and the bandwidth of m/t) is 0.5 MHz. When they are quadrature-modulated onto a subcarrier, the resulting composite baseband signal will then have a bandwidth of 4.2 MHz. (The upper sideband of the in-phase subcarrier signal is attenuated to keep a 4.2-MHz bandwidth.)

The composite NTSC baseband video signal is

$$m_c(t) \quad \begin{cases} m.(t) & \text{during the sync interval} \\ m_y(t) + \text{Re}\{g.c(t)ej'''sc'\} & \text{during the video interval} \end{cases} \tag{5-114}$$

where

$$gsc(t) = [m;(t) - jm_q(t)]e^{F} \tag{5-115}$$

Equation (5-115) indicates that the chrominance information is quadrature modulated onto a subcarrier (as described in Table 4-1). m,(t) is the sync signal, *m/t)* is the luminance signal, and g,c(t) is the complex envelope of the subcarrier signal. The subcarrier frequency is $f_{sc} = 3.579545$ MHz ± 10 Hz, which, as we will see later, is chosen so that the ubcarrier signal will not interfere with the luminance signal even though both signals fall within the 4.2-MHz passband.

In broadcast TV, this composite NTSC baseband color signal is amplitude-modulated onto the visual carrier as described by (5-110) and shown in Fig. 5-52. In microwave relay applications and satellite relay applications, this NTSC

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baseband signal is frequency-modulated onto a carrier, as described in Secs. 5-3 and 5-4. These color baseband signals can be transmitted over communication systems that are used for black-and-white TV, although the frequency response and linearity tolerances that are specified are tighter for color TV transmission.

As indicated before, the complex envelope for the subcarrier, as described by (5-115), contains the color information. The magnitude of the complex envelope, $|g(t)|$, is the *saturation* or amount of color in the scene as it is scanned with time. The angle of the complex envelope, $\angle g(t)$, indicates hue or tint as a function of time. Thus $g_{sc}(t)$ is a vector that moves about in the complex plane as a function of time as the lines of the picture are scanned. This vector may be viewed on an instrument called a *vectorscope,* which has a CRT that displays the vector $g_{sc}(t)$, as shown in Fig. 5-56. The vectorscope is a common piece of test equipment that is used to calibrate color TV equipment and to provide an on-line measure of the quality of the color signal being transmitted. The usual vectorscope presentation has the x axis located in the vertical direction, as in-

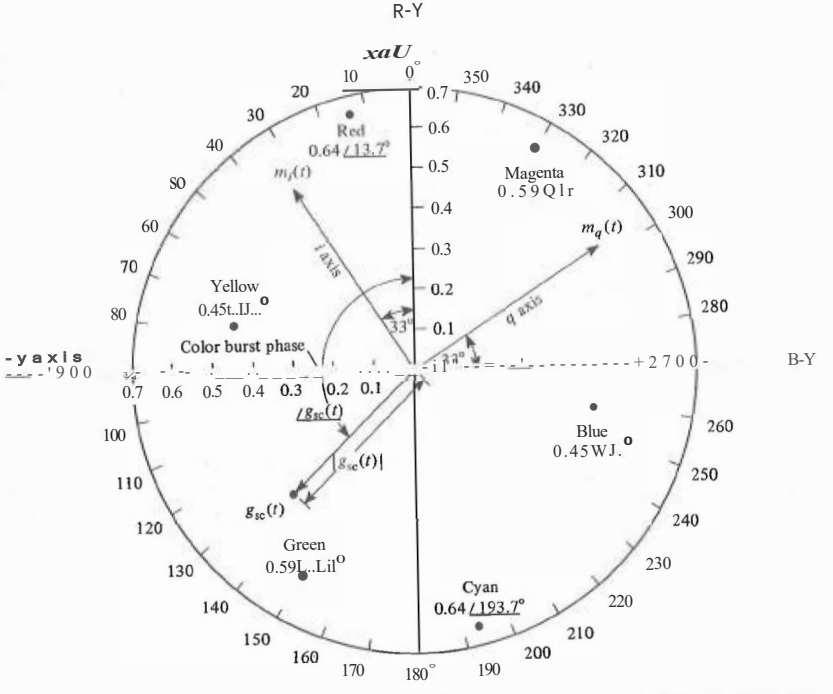


FIGURE 5-56 Vectorscope presentation of the complex envelope for the 3.58-MHz chrominance subcarrier.

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dicated on the figure. Using (5-115), the positive axis directions for the $m_i(t)$ and $m_q(t)$ signals are also indicated on the vectorscope presentation, together with the vectors for the saturated red, green, and blue colors. For example, if saturated red is present, then $m_R = 1$ and $m_0 = m_8 = 0$. Using (5-112), we get $m_i = 0.6$ and $m_q = 0.21$. Thus, from (5-115), the saturated red vector is

$$(g_{sc})_{red} = (0.60 - J0.21)e^{P3^\circ} = 0.64\angle 13.7^\circ$$

Similarly, the saturated green vector is $(g_{sc})_{green} = 0.59\angle 151^\circ$, and the saturated blue vector is $(g_{sc})_{blue} = 0.45\angle 257^\circ$. These vectors are shown in Fig. 5-56. It is interesting to note that the complementary colors have the opposite polarity on the vectorscope, where cyan is the complement of red, magenta is the complement of green, and yellow is the complement of blue.

For color subcarrier demodulation at the receiver, a phase reference is needed. This is provided by keying in an eight-cycle (or more) burst of the 3.58-MHz subcarrier sinusoid on the "back porch" of the horizontal sync pulse, as shown in Fig. 5-55. The phase of the reference sinusoid is $+90^\circ$ with respect to the x axis, as indicated on the vectorscope display (Fig. 5-56).

A color receiver is similar to a black-and-white TV receiver except that it has color demodulator circuits and a color CRT. This is shown in Fig. 5-57, where the color demodulator circuitry is indicated. The in-phase and quadrature signals, $m_i(t)$ and $m_q(t)$, are recovered (demodulated) by using product detectors where the phase reference signal is obtained from a PLL circuit that is locked onto the color burst (which was transmitted on the back porch of the sync pulse). M^{-1} is the inverse matrix of that given in (5-112) so that the three video waveforms corresponding to the red, green, and blue intensity signals are recovered. It is realized that the hue control on the TV set adjusts the phase of the reference that sets the tint of the picture on the TV set. The color control sets the gain of the chominance subcarrier signal. If the gain is reduced to zero, no subcarrier will be present at the input to the product detectors. Consequently, a black-and-white picture will be reproduced on the color TV CRT.

Of course, it is also possible to use different reference phases on the balanced detector inputs *provided* that the decoding matrix is changed to produce m_R, m_O, and m_B at its output. For example, one popular technique uses the R-Y and B-Y phases on the input to the balanced modulators [Grob, 1975]. Here the reference phase out of the hue control circuit is 0° instead of 33°, as shown in Fig. 5-57. This corresponds to detecting the subcarrier along the R-Y and B-Y axes, as indicated on the vectorscope presentation (Fig. 5-56). In this case $m_R(t) - m_y(t)$ is obtained at the output of the upper product detector, and if a gain of two is used on the lower product detector, $m_B(t) - m_y(t)$ is obtained at its output. These two utputs may be added in the proper proportions to obtain $m_O(t) - m_y(t)$. These three signals, $m_R(t) - m_y(t)$, $m_O(t) - m_y(t)$, and $m_B(t) - m_y(t)$, may be applied to the three control grids of a three-gun color CRT (one gun activating the red phosphor, one gun for the green, and one for the red). The $-m_y(t)$ signal is applied to the common cathode of the three guns so that the effective signals on the guns are $m_R(t)$, $m_O(t)$, and $m_B(t)$, respectively, which are the desired control signals. Furthermore, for black-and-white

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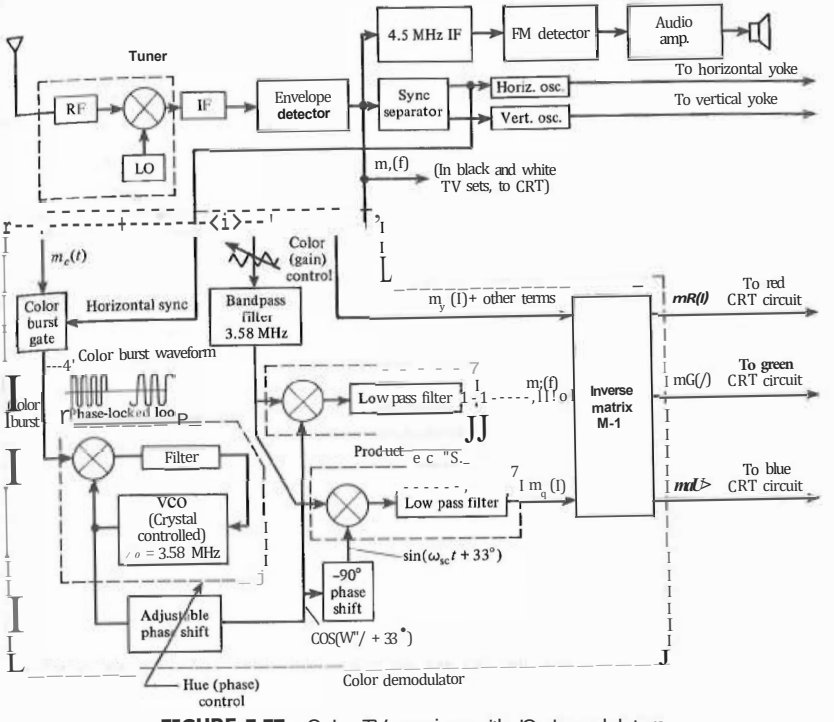


FIGURE 5-57 Color TV receiver with IQ demodulators.

operatio.n, it is only necessary to remove the three signals from the control grid since the luminance drive signal is applied at the cathode.

As stated earlier, both the luminance and chrominance subcarrier signals are contained in the 4.2-MHz composite NTSC baseband signal. This gives rise to interfering signals in the receiver video circuits. For example, in Fig. 5-57 there is an interfering signal on the input to the inverse matrix circuit. Here $m_y(t)$ plus interference is averaged out on a line-by-line scanning basis by the viewer's vision if the chrominance subcarrier frequency is an odd multiple of half of the horizontal scanning frequency. For example, from Table 5-14, f_{sc} = 3,579,545 Hz and f_r = 15,734 Hz. Consequently, there are 227.5 cycles of the 3.58-MHz chrominance interference across one scanned line of the picture. Since there is a half cycle left over, the interference on the *next* scanned line will be *180° out of phase* with that of the present line and the interference will be *canceled out* by the eye when one views the picture.

As we have seen, the NTSC color system is a very ingenious application of basic engineering principles, and this ingenuity is certainly not appreciated by the average color TV viewer.

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Standards for TV and CATV Systems

A summary of some of the U.S. TV transmission standards as adopted by the FCC is shown in Table 5-14, and details of the synchronizing waveform are given in Fig. 5-58. A listing of the frequencies for the TV channels is given in

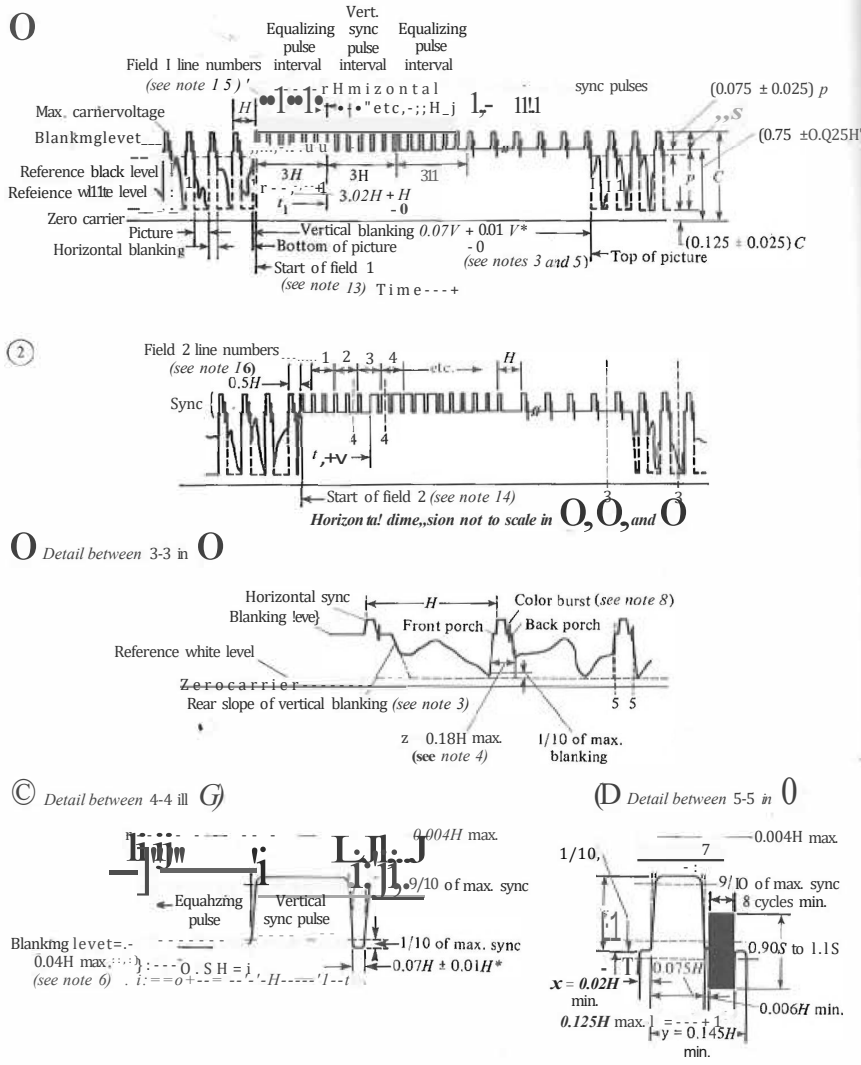


FIGURE 5-58 U.S. television synchronizing waveform standards for color transmission. (From Part 73 of FCC Rules and Regulations)

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TABLE 5-15 U.S. Television Channel Frequency Assignments

	Channel Number	Band (MHz)	UHF Channel Number	Band	UHF Channel Number	Band
Radio ast_ronomy	2	54-60	29	560-566	57	728-734
{ Aeronautical }	3	60-66	30	566-572	58	734-740
Two-way radfo	4	66-72	31	572-578	59	740-746
	5	76-82	32	578-584	60	746-752
FM broa casting	6	82-88	33	584-590	61	752-758
{ Aeronautical }	7	174-180	34	590-596	62	758-764
Two-way radio	8	180-186	35	596-602	63	764-770
	9	186-192	36	602-608	64	770-776
	10	192-198	37	608-614	65	776-782
	II	198-204	38	614-620	66	782-788
VHF	12	204-210	39	620-626	67	788-794
channels	13	210-216	40	626-632	68	794-800
UHF channels	14	470-476	41	632-638	69	800-806
	15	476-482	42	638-644	70"	806-812
	16	482-488	43	644-650	71Å	812-818
	17	488-494	44	650-656	72Å	818-824
	18	494-500	45	656-662	73Å	824-830
	19	500-506	46	662-668	74Å	830-836
	20	506-512	47	668-674	75Å	836-842
	21	512-518	48	674-680	76"	842-848
	22	518-524	49	680-686	77Å	848-854
	23	524-530	50	686-692	78"	854-860
	24	530-536	51	692-698	79Å	860-866
	25	536-542	52	698-704	SO•	866-872
	26	542-548	53	704-710	81Å	872-878
	27	548-554	54	710-716	82"	878-884
	28	554-560	55	716-722	83L	884-890
			56	722-728		

•These frequencies are now allocated to two-way radio (land mobile services), but were once allocated to UHF TV broadcasting.

NOTES TO FIGURE 5-58

1 H = time from star! of one line to start of next line = $1/f_H$ (see Table 5-14).
2 V = time from start of one field to start of next field = !//, (see Table 5-14).
3 Leading and trailing edges of vertical blanking should be complete in less than O ***IH.***
4 Leading and trailing slopes of horizontal blanking must be steep enough to preserve minimum and maximum values of $(x + y)$ and (z) under all conditions of picture content.
*5 Dimensions marked with asterisk indicate that tolerances given are permitted only for long-time variations and not for successive cycles.
6 Equalizing pulse area shall be between 0.45 and 0.5 of area of a horizontal sync pulse.
7 Color burst follows each horizontal pulse, but is omitted following the equalizing pulses and during the broad vertical pulses.
8 Color bursts to be omitted during monochrome transmission.
9 The burst frequency shall be 3.579545 MHz. The tolerance on the frequency shall be ± 10 Hz with a maximum rate of change of frequency not to exceed 1/10 Hz/sec.

(continues)

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NOTES TO FIGURE 5-58 ***(continued)***

10 The horizontal scanning frequency shall be 2/455 times the burst frequency.
11 The dimensions specified for the burst determine the times of starting and stopping the burst but not its phase. The color burst consists of amplitude modulation of a continuous sine wave.
12 Dimension ***P*** represents the peak excursion of the luminance signal from blanking level, but does not include the chrominance signal. Dimension S is the sync amplitude above blanking level. Dimension C is the peak carrier amplitude.
13 Start of field I is defined by a whole line between first equalizing pulse and preceding *H* sync pulses.
14 Start of field 2 ts defined by a half line between first equalizing pulse and preceding *H* sync pulses.
15 Field I line numbers start with first equalizing pulse in field I.
16 Field 2 line numbers start with second equalizing pulse in field 2.
17 Refer to Table 5-14 for further explanations and tolerances.

TABLE 5-16 Comparison of Some Broadcast Television Standards

	Where Used				
Item	North America, South America, Japan	Spain, Italy, England, Germany, CCIR System aa	France		USSR
Lines/frame	525	625	625	819	625
Lines/second	15,750	15,625	15,625	20,475	15,625
Frames/second	30	25	25	25	25
Baseband video bandwidth (MHz)	4.2	5	6	10	6
Channel bandwidth (MHz)	6	7	8	14	8
Polarity of AM video modulation	Negative	Negative	Positive	Positive	Negative
Type of aural carrier	FM	FM	AM	AM	FM
Color system	NTSC[b]	PAL[c]	SECAM[d]	SECAM[d]	SECAM[d]
Color subcarrier frequency (MHz)	3.58	4.43	4.43	4.43	4.43

"The 625-lines/frame, 25-frames/sec system is the CCIR (International Radio Consultative Committee) system B standard and is the dominant TV standard used on the European continent.

bNTSC, National Television System Committee (United States).

cPAL, Phase Alternation Line (Europe).

"SECAM, Sequential Couleur a Memoire (French).

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TABLE 5-17 CATV Channel Frequencies

	CATV Channel Number	Band (MHz)	Letter Channel Designator	CATV Channel Number	UHF Band (MHz)
		5.75-11.75	T-7	37	300-306
		11.75-17.75	T-8	38	306-312
		17.75-23.75	T-9	39	312-318
Suh-VHF		23.75-29.75	T-I0	40	318-324
		29.75-35.75	T-11	41	324-330
		35.75-41.75	T-12	42	330-336
		41.75-47.55	T-13	43	336-342
				44	342-348
	2	54-60		45	348-354
	3	60-66		46	354-360
Low-VHF	4	66-72		47	360-366
	5	76-82		48	366-372
	6	82-88		49	372-378
				50	378-384
	FM Broadcasting	88-I08		51	384-390
				52	390-396
	7	174-180		53	396-402
	8	180-186		54	402-408
	9	186-192		55	408-414
High-VHF	JO	192-198		56	414-420
	lJ	198-204		57	420-426
	12	204-210		58	426-432
	13	210-216		59	432-438
	14	120-126	A	60	438-444
	15	126-132	B	61	444-450
	16	132-138	C		
	17	138-144	D		
Mid-band	18	144-150	E		
	19	150-156	F		
	20	156-162	G	89	612-618
	21	162-168	H	90	618-624
	22	168-174	I	91	624-630
				92	630-636
	23	216-222	J	93	636-642
	24	222-228	K	94	642-648
	25	228-234	L	95	90-96
	26	234-240	M	96	96-102
	27	240-246	N	97	102-108
	28	246-252	0	98	108-114
Super band	29	252-258	p	99	114-120
	30	258-264	Q		
	31	264-270	R		
	32	270-276	S		
	33	276-282	T		
	34	282-288	u		
	35	288-294	V		
	36	294-300	W		

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Table 5-15. In addition, a comparison of some broadcast TV standards that are used by various countries is shown in Table 5-16. The CCIR system B standard which is a 625-lines/frame, 25-frames/sec, negative modulation system is th dominant European TV standard [Dhake, 1980]. In addition to these standards for conventional TV as shown in this table, there are many proposed standards for HDTV [Jurgen, 1988]. When TV signals are relayed from one country to another via satellite, the standards of the originating country are used, and, if needed, the signal is converted to the standards of the receiving country at the receiving ground station using Digital Intercontinental Conversion Equipment (DICE).

United States CATV channel standards are presented in Table 5-17. CATV systems use coaxial cable with amplifiers to distribute TV and other signals from the "head end" to the subscriber. Since the cable signals are not radiated, the frequencies that are normally assigned to two-way radio and other on-the-air services can be used for additional TV channels on the cable. Consequently, the CATV channel assignments have been standardized as shown in this table. To receive these added channels, the subscriber has to have a TV set with CATV channel capability or use a down converter to convert the CATV channel to a standard channel such as channel 3 or channel 4. CATV systems can also be *bidirectional,* allowing some subscribers to transmit signals to the head end. In low-band split systems, frequencies below 50 MHz (T-band channels) are used for transmission to the head end. In mid-band, split system frequencies below 150 mHz are used for transmission to the head end.

Premium TV programs are often available to subscribers for an additional cost. The premium channels are usually scrambled by modifying the video sync or possibly by using digital encryption techniques. Subscribers to the premium services are provided with appropriate decoders that, in some systems, are addressable and remotely programmable by the CATV company.

5-10 Summary

In this cl\apter, a wide range of practical communication systems were presented based on the theory developed in Chapters 1 through 4.

Basic bandpass digital signaling techniques such as OOK, BPSK, and FSK were developed from classical AM, PM, and FM theory. The spectra of these digital signals were evaluated in terms of the bit rate of the digital information source. Multilevel digital signaling techniques such as QPSK, MPSK, QAM, and MSK were also studied, and their spectra were evaluated.

Telephone and common carrier communication systems were examined. Specifications for voice frequency and data transmission were given for transmission over twisted-pair wire, coaxial cable, microwave radio relay, satellites, cellular radio, and fiber optic links. Examples of satellite frequency-division multiple access and time-division multiple access were given. Television signal relay via satellites was studied.

A technique was developed for evaluating communication system performance. This required the calculation of the signal-to-no.ise ratio at the detector

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input of a receiver for cases of both digital and analog signaling. In developing this result the concepts of received signal power from a free-space channel, thermal noise sources, noise figure, and noise temperature were established. In Example 5-6 a link budget for the received signal-to-noise ratio at the output of a TV receive-only terminal for reception of satellite signals was presented. An example of a link budget for a fiber optic common carrier system was also given.

The antijam and low probability of intercept properties of spread spectrum systems were studied. This included the calculation of the spectrum for a spread spectrum signal and the examination of transmitter and receiver block diagrams for both direct sequence and frequency hopped spread spectrum systems.

Television transmission, both black-and-white and NTSC color, was discussed in detail. Standards for television and CATV systems were given.

The link budget analysis in this chapter pointed out the need for the use of statistics to evaluate the performance of digital and analog detector circuits. These techniques will be developed in succeeding chapters. In Chapter 6 the concepts of random processes are developed from probability theory (which is established in Appendix B), and in Chapter 7 the results are used to evaluate the performance of digital and analog receiving circuits for the case of noisy input signals.

PROBLEMS

5-1 A digital baseband signal consisting of rectangular binary pulses occurring at a rate of 2400 bits/sec is to be transmitted over a bandpass channel.
- **(a)** Evaluate the magnitude spectrum for OOK signali'1g that is keyed by a baseband digital test pattern consisting of alternating 1s and Os.
- **(b)** Sketch the magnitude spectrum and indicate the value of the first null-to-null bandwidth. Assume a carrier frequency of 50 kHz.
- **(c)** For a random data pattern, find the PSD and plot the result. Compare this result with that obtained in parts (a) and (b) for alternating data.

5-2' Repeat Prob. 5-1 for the case of BPSK signaling.

5-3 Evaluate the magnitude spectrum for an FSK signal with alternating 1 and 0 data. Assume that the mark frequency is 50 kHz, the space frequency is 55 kHz, and the bit rate is 2400 bits/sec. Find the first null-to-null bandwidth.

5-4 Assume that 4800-bits/sec random data are sent over a bandpass channel by BPSK signaling. Find the transmission bandwidth B_T such that the spectral envelope is down at least 35 dB outside of this band.

5-5 As indicated in Sec. 5-2, a BPSK signal can be demodulated by using a synchronous detector where the carrier reference is provided by a Costas loop for the case of $h = 1.0$. Alternatively, the carrier reference can be provided by a squaring loop that uses a x 2 frequency multiplier. A block diagram for a squaring loop is shown in Fig. P5-5.

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