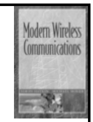
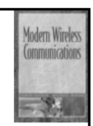


Chapter 3

Modern Wireless Communication



Chapter 3

Modulation and Frequency-Division Multiple Access

Contents

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Contents (Cont.)

3.1 Introduction

- FDMA
  - dividing the bandwidth of a wireless channel equally among a number of users wanting to access the channel.
- Modulation is a process of
  - transforming the frequency content of a particular user's information-bearing signal
  - so as to lie inside the frequency band allotted to that user.

3.2 Modulation

- Modulation:
  – some characteristic of a carrier wave is varied in accordance with an information-bearing signal.
- Benefits of modulation
  – For shift the spectral content of a message signal.
  – Provides a mechanism for putting information into a form that may be less vulnerable to noise or interference.
  – Permits the use of multiple-access techniques.

3.2 Modulation

3.2.1 Linear and Nonlinear Modulation Processes

- Principle of superposition
  - The output of modulator produced by a number of inputs applied simultaneously is equal to the sum of outputs that result when the inputs are applied one at a time.
  - If the input is scaled by a certain factor, the output of the modulator is scaled by exactly same factor
- Satisfy the principle of superposition → linear
- Not satisfy the principle of superposition → nonlinear

3.2 Modulation

3.2.2 Analog and Digital Modulation Techniques

- Distinguish between analog and digital modulation:
  - All analog modulated signals are continuous functions of time.
  - Digital modulated signals can be continuous or discontinuous functions of time.

3.2 Modulation

3.2.3 Amplitude and Angle Modulation Process

- Amplitude modulation: the amplitude of the carrier, Ac is varied linearly with the message signal $m(t)$
- Angle modulation: the angle of the carrier is varied linearly with the message signal $m(t)$
  - Two kinds of angle modulation:
    - Frequency modulation
    - Phase modulation

3.3 Linear Modulation Techniques

3.3 Linear Modulation Techniques

3.3.1 Amplitude Modulation

- Amplitude Modulation

$$s(t) = A_c(1 + k_a m(t))\cos(2pf_c t) \qquad (3.3)$$

where k_a is the sensitivity of amplitude modulator

- Retention of carrier in composition of AM signal represents a loss of transmitted signal power.
- Double sideband-suppressed carrier (DSB-SC) modulation

$$\begin{aligned} s(t) &= c(t)m(t) \\ &= A_c m(t)\cos(2pf_c t) \end{aligned} \qquad (3.4)$$

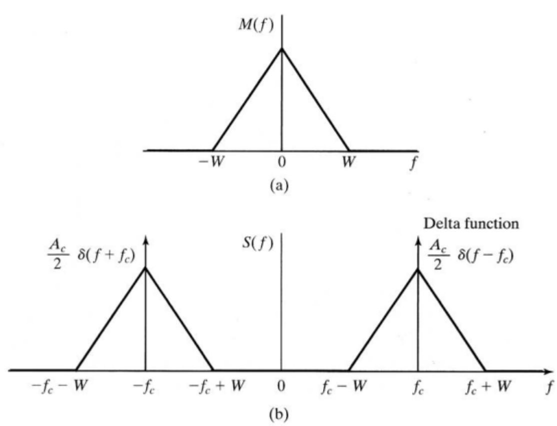

FIGURE 3.4 (a) Message spectrum. (b) Spectrum of corresponding AM signal.

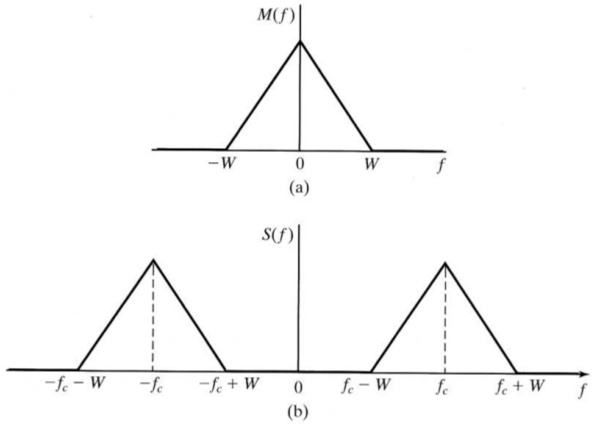

FIGURE 3.5 (a) Message spectrum. (b) Spectrum of corresponding DSB–SC modulated signal.

3.3 Linear Modulation Techniques

3.3.2 Binary Phase-Shift Keying

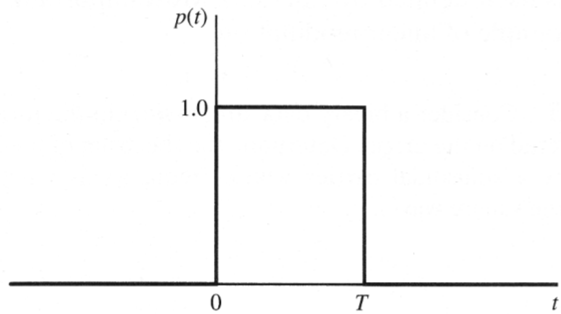

FIGURE 3.6 Rectangular pulse.

- Let $p(t)$ denote the basic pulse and T denote the bit duration. The binary data stream is

$$m(t)=\sum_k b_k p(t-kT) \quad \text{where} \quad b=\begin{cases}+1 \text{ for binary symbol } 1\\ -1 \text{ for binary symbol } 0\end{cases} \qquad (3.5,6)$$

- In binary phase-shift keying (BPSK), the simplest form of digital phase modulation,
  - the binary symbol 1 : setting the carrier phase θ (t)=0 radians
  - binary symbol 0 : setting θ (t)= π
- BPSK can be express as

$$s(t)=c(t)m(t) \qquad \text{where c(t) is the carrier} \qquad (3.10)$$

3.3 Linear Modulation Techniques

3.3.3 Quadriphase-Shift Keying

- Retain the property of BPSK and the transmission bandwidth is equal to the bandwidth of incoming binary data stream.
- QPSK modulator
  - parallel combination of two BPSK modulators that operate in phase quadrature with respect to each other.
- The QPSK signal is obtained by adding two BPSK signals

$$\begin{aligned} s(t) &= s_1(t) + s_2(t) \\ &= A_c m_1(t)\cos(2pf_c t) + A_c m_2(t)\sin(2pf_c t) \end{aligned} \quad (3.16)$$

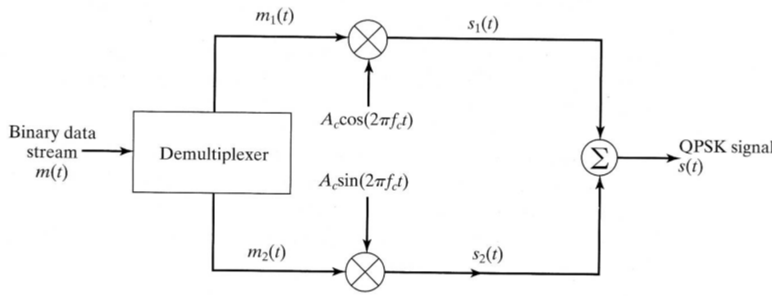

FIGURE 3.7 Block diagram of a QPSK generator, using a phase-quadrature pair of carriers $A_c\cos(2\pi f_c t)$ and $A_c\sin(2\pi f_c t)$.

3.3 Linear Modulation Techniques

3.3.4 Offset Quadriphase-Shift Keying

- For reduction of amplitude fluctuations of conventional QPSK signals
- ±90° phase jump in OQPSK occur twice as frequently, but with a reduced range of amplitude fluctuations, compared with conventional QPSK

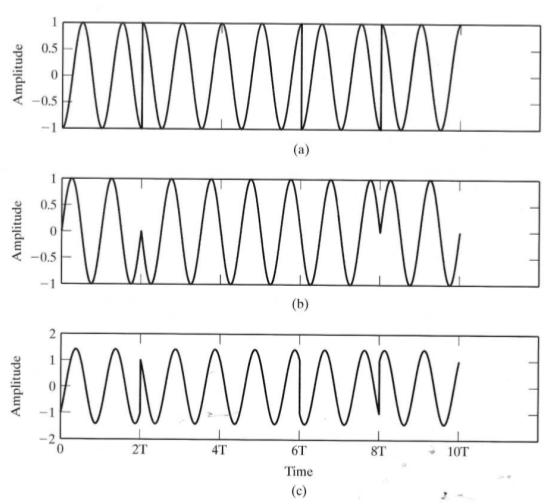

FIGURE 3.8 Waveforms of (a) conventional QPSK (b) offset QPSK, and (c) $\pi/4$-shifted QPSK.

3.3 Linear Modulation Techniques

3.3.5 π/4-Shifted Quadriphase-Shift Keying

- Two setting of carrier phase of QPSK

  $1.\ 0, p/2, p, or\, 3p/2\ radians$

  $2.\ p/4, 3p/4, 5p/4\ or\, 7p/4\ radians$

- π/4-Shifted Quadriphase-Shift Keying:
  - carrier phase alternatively picked from settings 1 and 2.
- Provides higher bandwidth efficiency, with a reduced range of amplitude fluctuations.

3.4 Pulse Shaping

- *Premodulation filter* is used in pulse shaping without *signal distortion* and *intersymbol interference* (ISI) problem.
- Overall frequency response "*raised-cosine (RC) spectrum*", which made up of transmit filter, channel and receiver filter

$$P(f) = \begin{cases} \dfrac{1}{2W} & 0 \le |f| \le f_1 \\ \dfrac{1}{4W}\left[1 + \cos\left(\dfrac{p}{2Wr}\left(|f| - W(1-r)\right)\right)\right] & f_1 \le |f| < 2W - f_1 \\ 0 & |f| \ge 2W - f_1 \end{cases} \quad (3.17)$$

where f_1 is the frequency parameter and W is the bandwidth

- Roll-off factor indicates the excess bandwidth over the ideal solution corresponding to **ρ** =0

$$r = 1 - \frac{f_1}{W} \qquad (3.18)$$

- $p(t)$, the inverse Fourier transform of $p(f)$,
  - has value of unity at current signaling instant
  - *zero crossings* at all other consecutive signaling instants.
- The zero crossing ensure the ISI problem is reduced to zero.

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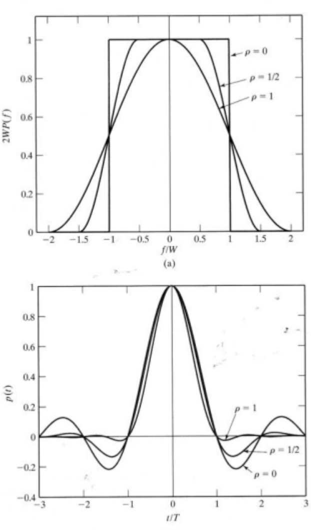

FIGURE 3.9 (a) Frequency response of the raised cosine spectrum for varying roll-off rates. (b) Impulse response of the Nyquist shaping filter (i.e., inverse Fourier transform of the spectrum plotted in part (a) for varying roll-off rates).

3.4 Pulse Shaping

3.4.1 Root Raised-Cosine Pulse Shaping

- Root RC spectrum

$$P(f)=\begin{cases}\frac{1}{\sqrt{2W}} & 0\le |f|\le f_1\\ \frac{1}{\sqrt{2W}}\cos\left(\frac{p}{4Wr}|f|-W(1-r)\right) & f_1\le |f|<2W-f_1\\ 0 & |f|\ge 2W-f_1\end{cases} \quad (3.20)$$

- The roll-off factor is the same as RC spectrum

- The inverse Fourier transform defines the root RC shaping pulse

$$p(t) = \frac{\sqrt{2W}}{\left(1-\left(8rWt\right)^2\right)}\left(\frac{\sin(2pW(1-r)t)}{2pWt} + \frac{4r}{p}\cos\left(2pW(1+r)t\right)\right) \quad (3.21)$$

- The new shaping pulse satisfies an orthogonality constraint under T-shifts (where T is the symbol duration) as shown by

$$\int_{-\infty}^{\infty} p(t)\,p(t-nT)\,dt = 0 \quad \text{for } n = \pm1, \pm2, \ldots \qquad (3.22)$$

- It lacks the zero-crossing property

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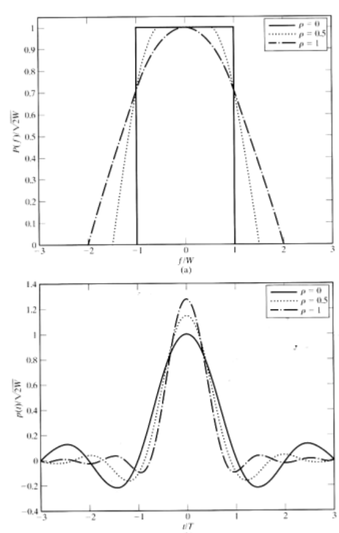

FIGURE 3.10 (a) $P(f)$ for root raised - cosine spectrum. (b) $p(t)$ for root raised - cosine spectrum.

3.5 Complex Representation of Linear Modulated Signals and Band-Pass System

- Canonical representation of a band-pass signal

$$s(t) = s_I(t)\cos(2pf_c t) - s_Q(t)\sin(2pf_c t) \qquad (3.23)$$

TABLE 3.1 Special Cases of the Canonical Equation (3.23).

	Type of modulation	In-phase component $s_I(t)$	Quadrature component $s_Q(t)$	Defining equation
Analog	Amplitude modulation	$A_c(1 + k_a m(t))$	0	3.3
	Double sideband–suppressed carrier modulation	$A_c m(t)$	0	3.4
Digital	Binary phase-shift keying	$A_c \sum_k b_k p(t - kT)$	0	3.5
	Quadriphase-shift keying	$A_c \sum_k b_{k,1} p(t - 2kT)$	$-A_c \sum_k b_{k,2} p(t - 2kT)$	3.11

- Complex envelope

$$\widetilde{s}(t) = s_I(t) + js_Q(t) \qquad (3.24)$$

- Invoke Euler's formula

$$\exp(j2pf_c t) = \cos(2pf_c t) + j\sin(2pf_c t) \qquad (3.25)$$

- Single-carrier transmission

$$s(t) = \mathrm{Re}\left\{\widetilde{s}(t)\exp\left(j2pf_c t\right)\right\} \qquad (3.26)$$

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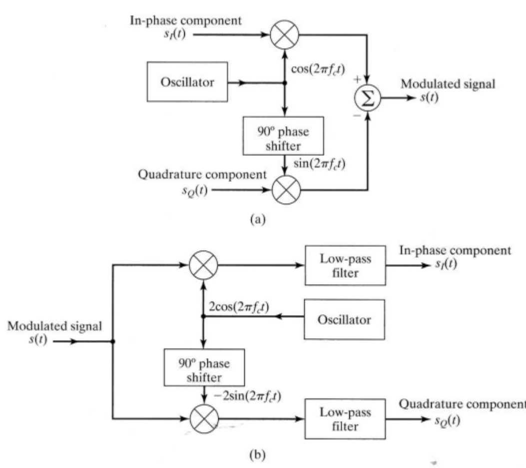

FIGURE 3.12 (a) Synthesizer for constructing a modulated signal from its in-phase and quadrature components. (b) Analyzer for deriving the in-phase and quadrature components of the modulated band-pass signal.

3.5 Complex Representation of Linear Modulated Signals and Band-Pass System

3.5.1 Complex Representation of Linear Band-Pass System

- Two assumptions of linear band-pass system
  - Narrowband
  - The modulated signal whose carrier frequency is the same as the midband frequency of system

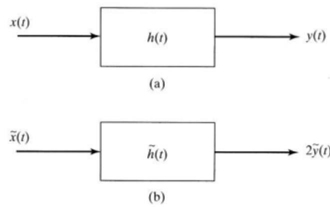

FIGURE 3.13 (a) Block diagram of linear band-pass system driven by a modulated signal $x(t)$ to produce the band-pass signal $y(t)$ as output. (b) Equivalent complex baseband model of the system, where the input signal, the impulse response, and output signal are all complex and in their baseband form.

- Equivalent complex baseband model

$$\tilde{h}(t) = h_I(t) + jh_Q(t)$$

$$\text{where } h_I(t) \text{ is the in-phase component of } \tilde{h}(t) \text{ is the quadrature component} \qquad (3.27)$$

- Impulse response of original band-pass system

$$h(t) = \mathrm{Re}\left\{\tilde{h}(t)\exp(j2pf_c t)\right\} \qquad (3.28)$$

- The output signal of the equivalent model ?(t) is the complex envelope of the original output signal $y(t)$

$$\begin{aligned}\tilde{y}(t) &= \frac{1}{2}\int_{-\infty}^{\infty}\tilde{x}(l)\tilde{h}(t-l)dl \\ &= \frac{1}{2}\int_{-\infty}^{\infty}\tilde{h}(l)\tilde{x}(t-l)dl\end{aligned} \qquad (3.29)$$

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- The important point to take from the equivalence between the two system, real and complex, is that the carrier frequency f_c has been eliminated from the equivalent model.
- In effect, traded complex analysis for the elimination of the carrier frequency
  - simplifying the analysis without any loss of information.
- The actual output signal $y(t)$ is

$$y(t) = \mathrm{Re}\left\{\tilde{y}(t)\exp(j2pf_c t)\right\} \qquad (3.31)$$

3.6 Signal-Space Representation of Digitally Modulated Signals

- Digital modulation:
  - the transmitted signal $s(t)$ assumes one of a discrete set of possible forms
  - mapping of the complex envelope onto a signal space is signal constellation or signal pattern.
- Energy normalized versions is used as the horizontal axis and vertical axis of the two-dimensional signal space.
- Normalized coordinates of unit energy

$$f_1(t) = \sqrt{\frac{2}{T}} \cos(2pf_c t) \quad \text{and} \quad f_2(t) = \sqrt{\frac{2}{T}} \sin(2pf_c t) \quad 0 \le t \le T$$

(3.32,33)

- Two properties:

  1. $\int_0^T f_1(t) f_2(t)\,dt = 0$ (3.34)

  which confirms the orthogonality of $f_1(t)$ and $f_2(t)$ over the interval $0 \le t \le T$

  2. $\int_0^T f_1^2(t)\,dt = \int_0^T f_2^2(t)\,dt = 1$ (3.35)

  which shows that both $f_1(t)$ and $f_2(t)$

- Thus, $f_1(t)$ and $f_2(t)$ are said to constitute an *orthonormal set*.

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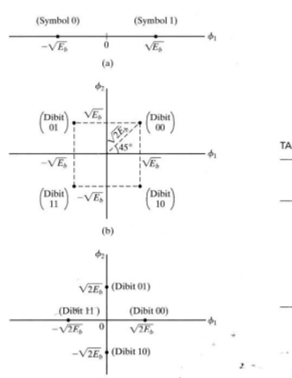

FIGURE 3.14 Signal constellations for (a) BPSK, (b) one version of QPSK, and (c) another version of QPSK.

TABLE 3.2 Signal-space characterization of the QPSK signal constellation described in Fig. 3.14(b).

Input dibit	Gray-encoded phase of QPSK signal (radians)	Coordinates of message points	
		s_{i1}	s_{i2}
10	$7\pi/4$	$+\sqrt{E_b}$	$-\sqrt{E_b}$
11	$5\pi/4$	$-\sqrt{E_b}$	$-\sqrt{E_b}$
01	$3\pi/4$	$-\sqrt{E_b}$	$+\sqrt{E_b}$
00	$\pi/4$	$+\sqrt{E_b}$	$+\sqrt{E_b}$

- For figure 3.14, the energy transmitted remains fixed.
- But in figure 3.15, the energy transmitted is variable, depends on the particular quadbits.

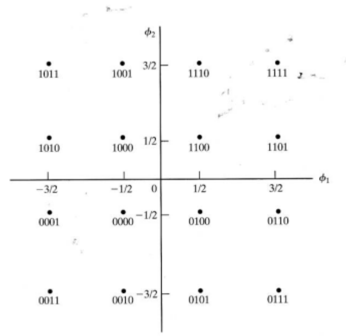

FIGURE 3.15 Signal-space diagram of M-ary QAM for $M = 16$; the message points in each quadrant are identified with Gray-encoded quadbits.

3.7 Nonlinear Modulation Techniques

- Polar form of canonical representation

$$s(t) = a(t)\cos\left[2pf_c t + q(t)\right] \qquad (3.37)$$

- Envelope of the modulated signal s(t)

$$a(t) = \sqrt{s_I^2(t) + s_Q^2(t)} \qquad (3.38)$$

- Phase of s(t)

$$q(t) = \tan^{-1}\left(\frac{s_Q(t)}{s_I(t)}\right) \qquad (3.39)$$

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3.7 Nonlinear Modulation Techniques

3.7.1 Frequency Modulation

- Frequency modulation
  - form of angle modulation
  - the instantaneous frequency of the sinusoidal carrier is varied in accordance with the information-bearing signal.

- Instantaneous frequency of FM signal

$$
\begin{aligned}
f(t) &= \lim_{\Delta t \to 0} f_{\Delta t}(t) \\
&= \lim_{\Delta t \to 0} \left[\frac{y\left(t+\Delta t\right) - \Psi(t)}{2p\Delta t} \right] \\
&= \frac{1}{2p} \frac{d\,y\,(t)}{dt}
\end{aligned}
\qquad (3.41)
$$

- Define instantaneous phase as the integral of the instantaneous frequency with respect to time

$$y\,(t) = \int_0^t f(t\,)dt \qquad (3.42)$$

- The frequency-modulated signal can be described in time domain

$$
\begin{aligned}
s(t) &= A_c \cos\left(y\,(t)\right) \\
&= A_c \cos\left(2pf_c t + 2pk_f \int_0^t m\left(t\,\right) dt \right)
\end{aligned}
\qquad (3.45)
$$

- Frequency modulation is a
  - nonlinear process
  - dependence of the modulated signal on the information-bearing signal violates the principle of superposition.
- The transmission bandwidth of the FM signal $s(t)$ is approximately given by Carson's rule

$$B_T \approx 2\Delta f\left(1+\frac{1}{D}\right) \qquad (3.46)$$

- Parameters of Carson's rule
  - Frequency deviation Δf.
    - The maximum deviation in the instantaneous frequency of FM signal away from the carrier frequency.
  - The deviation ratio D.
    - The ratio of frequency deviation to the highest frequency contained in the modulating signal $m(t)$.

An increase in the FM transmission bandwidth B_T produces a quadratic increase in the signal-noise ratio at the output of the receiver.

3.7 Nonlinear Modulation Techniques

3.7.2 Binary Frequency-Shift Keying

- A typical pair of sinusoidal waves for BFSK is described by

$$s_i(t) = \begin{cases} \sqrt{\frac{2E_b}{T}} \cos(2pf_i t) & 0 \le t \le T_b \\ 0 & otherwise \end{cases} \qquad (3.47)$$

  where $i = 1, 2, ..., T$ is the symbol (bit) duration and E_b is the energy transmitted per bit
- The frequency transmitted is

$$f_i = \frac{n_c + i}{T} \text{ for some fixed integer } n_c \text{ and } i = 1, 2 \qquad (3.48)$$

- continuous-phase signal
  - the phase continuity is maintained everywhere including the interbit switching times.

- Two coordinates of the signal constellation of BFSK defined by

$$f_i(t) = \begin{cases} \sqrt{\dfrac{2}{T}}\cos(2pf_i t) & 0 \le t \le T \\ 0 & otherwise \end{cases}$$

(3.49)

where $i = 1,2$

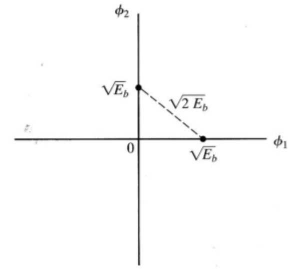


FIGURE 3.16 Signal constellation of binary FSK signal.

3.7 Nonlinear Modulation Techniques

3.7.3 Continuous-Phase Modulation: Minimum Shift Keying

- By more constrained choosing of two frequencies (symbol 1 and 0), we can improve the spectral efficiency and improve noise performance.

$$s(t)=\begin{cases}\sqrt{\dfrac{2E_b}{T}}\cos[2pf_1+q(0)] & \text{for symbol 1}\\ \sqrt{\dfrac{2E_b}{T}}\cos[2pf_2+q(0)] & \text{for symbol 0}\end{cases} \qquad (3.50)$$

- CRFSK signal in conventional form of an angle-modulated signal

$$s(t)=\sqrt{\frac{2E_b}{T}}\cos[2pf_c t+q(t)] \qquad (3.51)$$

- The phase $\theta(t)$ of CPFSK signal change linearly with time during each bit duration of T seconds, as shown by the relationship

$$q(t)=q(0)\pm\frac{ph}{T}t \quad 0\le t\le T \qquad (3.52)$$

- The relationship between T, h, f_1, f_2 and f_c

$$f_c=\frac{1}{2}(f_1+f_2) \text{ and } h=T(f_1+f_2) \qquad (3.55,56)$$

- The nominal carrier frequency f_c is the arithmetic mean of the transmit frequencies f_1 and f_2.
- The difference between the transmit frequencies normalized with respect to the bit rate $1/T$.
- h is the deviation ratio of the CPFSK signal.
- When time $t=T$

$$q(T)-q(0)=\begin{cases}ph & \text{for symbol 1}\\ -ph & \text{for symbol 0}\end{cases} \qquad (3.57)$$

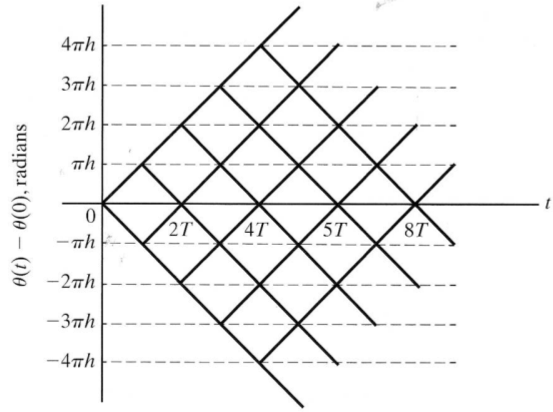


FIGURE 3.17 Phase tree of a CPFSK signal.

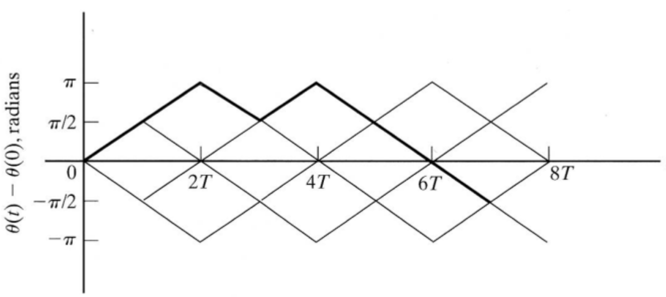


FIGURE 3.18 Phase trellis; boldfaced path represents the sequence 1101000.

- With $h = 1/2$, the frequency deviation equals half the bit rate.
- This is the minimum frequency spacing that allows
  - the two FSK signals representing the symbols 1 and 0 to be coherently orthogonal.
- Two signaling frequencies do not interfere with each other.
- This deviation ratio of one-half is called “minimum shift keying”.

3.7 Nonlinear Modulation Techniques

3.7.4 Power Spectra of MSK Signal

- 1. Depending on the value of the phase state θ (0), the in-phase component equals $+g_I(t)$ or $-g_I(t)$, where

$$g_I(t) = \begin{cases} \sqrt{\dfrac{2E_b}{T}} \cos\left(\dfrac{pt}{2T}\right) & -T \le t \le T \\ 0 & otherwise \end{cases} \qquad (3.63)$$

- 2. Depending on the value of the phase state θ (*T*), the quadrature component equals $+g_Q(t)$ or $-g_Q(t)$, where

$$g_Q(t) = \begin{cases} \sqrt{\dfrac{2E_b}{T}} \sin\left(\dfrac{pt}{2T}\right) & -T \le t \le T \\ 0 & otherwise \end{cases} \qquad (3.64)$$

- 3. The in-phase and quadrature components of the MSK signal are statistically independent, yielding the baseband power spectral density of the MSK signal:

$$S_B(f) = 2\left[\frac{y_g(f)}{2T}\right] = \frac{32E_b}{p^2}\left[\frac{\cos(2pTf)}{16T^2f^2 - 1}\right]^2 \qquad (3.65)$$

- $y_g(f)$ is the common power spectral density of $g_I(t)$ and $g_Q(t)$.
- The baseband power spectral density of the MSK signal falls off as the inverse fourth power of frequency.
- MSK does not produce as much interference outside the signal band of interest as does QPSK.

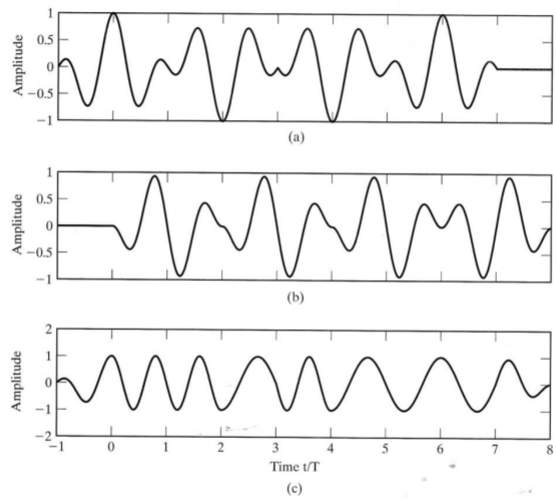
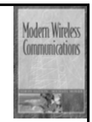


FIGURE 3.19 Waveforms for Problem 3.10: (a) $s_1\phi_1(t)$, (b) $s_2\phi_2(t)$, and (c) MSK signal $s(t)$.

3.7 Nonlinear Modulation Techniques

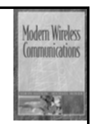
3.7.5 Gaussian-Filtered MSK



- *Gaussian-filtered* MSK:
  - a binary frequency modulation that a polar nonreturn-to-zero (NRZ) binary data stream through a baseband pulse-shaping filter whose impulse response is defined by a Gaussian function. In the line code, binary symbols 0 and 1 are represented by -1 and +1.
- Let W denote 3-dB *baseband bandwidth*. The transfer function $H(f)$ and impulse response $h(t)$ of pulse-shaping filter

$$H(f) = \exp\left(-\frac{\log_e 2}{2}\left(\frac{f}{W}\right)^2\right) \text{ and } h(t) = \sqrt{\frac{2p}{\log_2 2}}W \exp\left(-\frac{2p^2}{\log_e 2}W^2 t^2\right)$$

(3.66,67)



- The response of Gaussian filter to a rectangular pulse of unit amplitude and duration T is given by

$$g(t) = \int_{-T/2}^{T/2} h(t-t)dt = \sqrt{\frac{2p}{\log_e 2}}W\int_{-T/2}^{T/2} \exp\left(-\frac{2p^2}{\log_e 2}W^2(t-t)^2\right)dt \quad (3.68)$$

- It also can be expressed as the difference between two complementary error functions

$$g(t) = \frac{1}{2}\left[erfc\left(p\sqrt{\frac{2p}{\log_e 2}}WT\left(\frac{t}{T}-\frac{1}{2}\right)\right) - erfc\left(p\sqrt{\frac{2p}{\log_e 2}}WT\left(\frac{t}{T}+\frac{1}{2}\right)\right)\right] \quad (3.69)$$

- The pulse response, $g(t)$ constitutes the *frequency-shaping pulse* of the GMSK modulator, with the dimensionless *time-bandwidth product* WT playing the role of a design parameter.

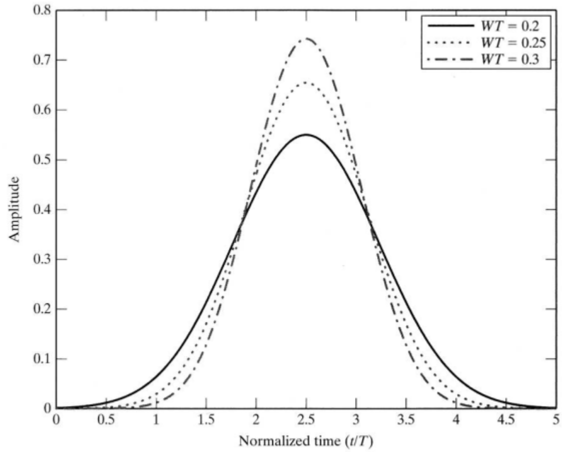


FIGURE 3.20 Frequency-shaping pulse $g(t)$ of Eq. (3.63), shifted in time by $2.5T$ and truncated at $\pm 2.5T$ for varying time–bandwidth product WT.

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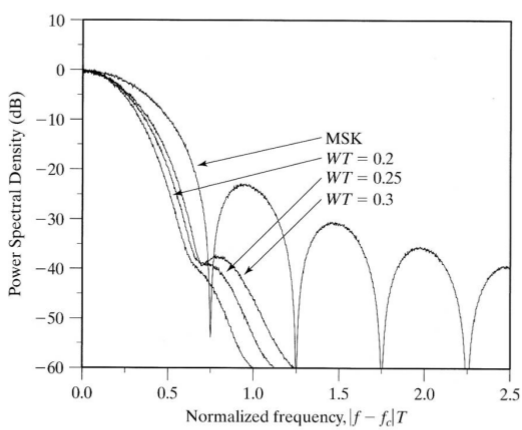

FIGURE 3.21 Power spectra of MSK and GMSK signals for varying time–bandwidth product.

3.8 Frequency-Division Multiple Access

- Available spectrum is divided into
  - a set of continuous frequency bands labeled 1 to N.
  - The bands are assigned to individual users for communication on continuous-time basis.
- Precautionary measures of adjacent channel interference
  - The power spectral density of modulated signal is carefully controlled
  - Guard bands are inserted as buffer zones in the channel assignment
- Duplexing allows mobile user to send and receive information-bearing signals on the same antenna

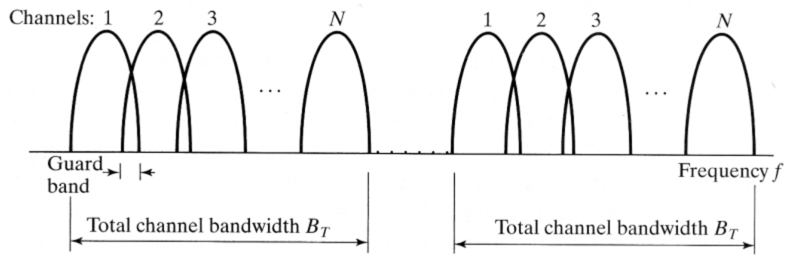

FIGURE 3.22 Channel allocations in FDD/FDMA system.

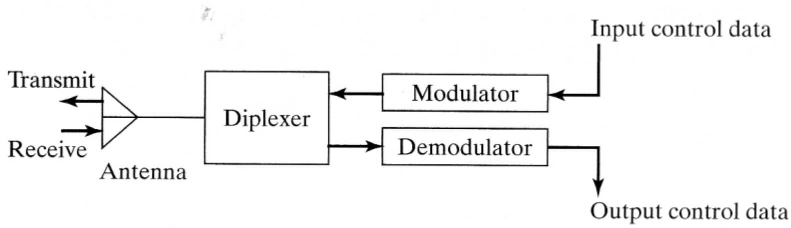

FIGURE 3.23 Frequency-division diplexer.

3.9 Two Practical Issues of concern

3.9 Two Practical Issues of concern

3.9.1 Adjacent Channel Interference

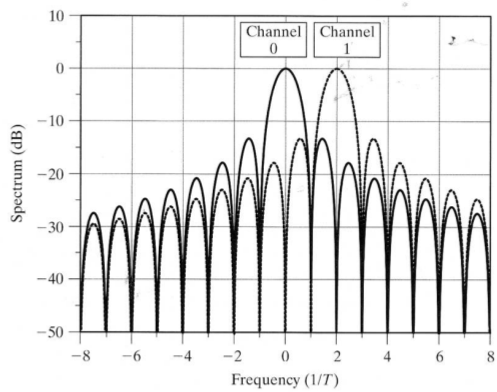

FIGURE 3.24 Adjacent-channel interference problem, illustrated with the spectrum of a rectangular pulse of duration T.

3.9 Two Practical Issues of concern

3.9.2 Power Amplifier Nonlinearity

- Transmit power amplifier: significant consumer of power in mobile radio
- The gain is not constant over all input levels means that
  - the amplifier introduces amplitude distortion in the form of AM.
  - This distortion is called AM-to-AM distortion.
- The phase characteristic is not constant over all input levels means
  - the amplifier introduces phase distortion in PM.
  - This distortion is called AM-to-PM distortion

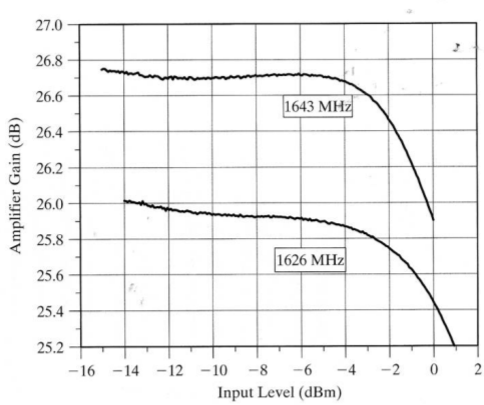

FIGURE 3.25 Gain characteristic of a solid-state amplifier at two different operating frequencies: 1626 MHz and 1643 MHz.

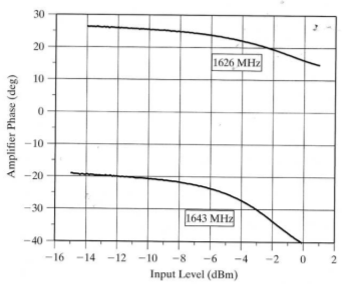

FIGURE 3.26 Phase characteristic of a nonlinear amplifier at two different operating frequencies: 1626 MHz and 1643 MHz.

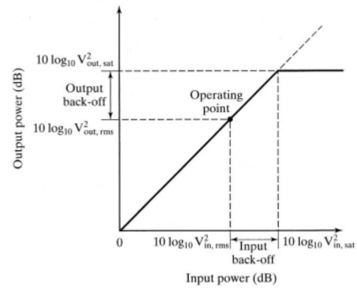

FIGURE 3.27 AM/AM characteristic of an "ideal" form of amplifier nonlinearity.

- Input back-off is the operating point the amplifier.

$$\text{Input back-off} = 10\log_{10}\left(\frac{V_{in,rms}}{V_{out,sat}}\right)^2 \qquad (3.71)$$

- The operating point can be expressed in terms of output back-off.

$$\text{Output back-off} = 10\log_{10}\left(\frac{V_{out,rms}}{V_{out,sat}}\right)^2 \qquad (3.72)$$

3.10 Comparison of Modulation Strategies for Wireless Communications

- 3.10.1 Linear Channels
- 3.10.2 Nonlinear Channels

3.10 Comparison of Modulation Strategies for Wireless Communications

3.10.1 Linear Channels

- Transmit spectrum important for selecting modulation strategies, which determined by
  - Pulse
  - Other Filtering
  - Presence of nonlinearities
- Advantages of former linear method of modulation:
  - The main lobe is the narrowest
  - Negligibly small sidelobes
- In a linear channel, the QPSK with root RC pulse shaping is clearly the superior strategy in terms of minimizing adjacent channel interference.

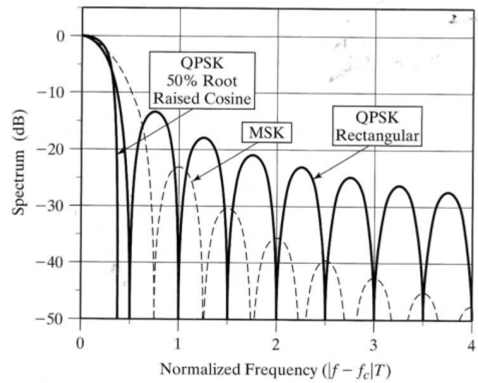

FIGURE 3.28 Comparison of the spectra of QPSK with rectangular pulse shaping, MSK, and QPSK with 50% root-raised-cosine pulse shaping.

3.10 Comparison of Modulation Strategies for Wireless Communications

3.10.2 Nonlinear Channels

- For nonlinear power amplifier, system performance depends on
  - how close to saturation the amplifier is operated.
- The effect of the amplifier also depends on
  - the type of modulation employed.
- For rectangular QPSK,MSK and GMSK, since the nonlinear effects depend upon envelope variation,
  - the spectra of modulations are unaffected by a nonlinear amplifier.

- For QPSK with root RC filtering, it relies on
  - envelope variations to produce a compact, the effect of the nonlinear amplifier on the respective modulated signals will differ.
- The choice of modulation depends on
  - the simplicity of detection, transmit spectrum and error-rate performance.
- Just as with AM-to-AM distortion, AM-to-PM distortion varies with the signal level at power amplifier input

3.11 Channel Estimation and Tracking

3.11 Channel Estimation and Tracking

3.11.1 Differential Detection

- Transmitted band-pass signal

$$s(t) = A\,\mathrm{Re}\left\{m(t)e^{j2pf_c t}\right\} \qquad (3.73)$$

- Received complex baseband signal

$$\tilde{x}(t) = A' m(t) e^{j(2p\Delta f t + f)} + w(t) \qquad (3.74)$$

- Differential detection

$$m(t) = \sum_k b_k\, p(t - kT) \qquad (3.75)$$

- Practical significance of differential detection
  - Can be applied linear and nonlinear phase modulation.
  - Can be applied to the wireless system that the fading is slow relative to the data rate.

3.11 Channel Estimation and Tracking

3.11.2 Pilot Symbol Transmission

- If channel exhibits flat fading, complex baseband received signal

$$\tilde{x}(t) = \tilde{a}(t)m(t) + \tilde{w}(t) \qquad (3.78)$$

- The samples at output of matched filter in discrete time

$$y_k = a_k b_k + w_k \qquad (3.79)$$

- Linear estimator

$$\tilde{a}_{Ki} = \sum_{m=-L}^{L} a_m h_{K(i+m)} \quad \text{where} \quad h_{Ki} = b_{Ki}^* y_{Ki} \qquad (3.81,80)$$

$$= a_{Ki}|b_{Ki}|^2 + b_{Ki}^* w_{Ki}$$

$$= a_{Ki} + w_{Ki}$$

- Cost Function $J = E\left[|a_{Ki}|^2 + \sum_{m=-L}^{L}\sum_{n=-L}^{L} a_m h_{K(i+m)} a_n^* h_{K(i+n)}^* - \sum_{m=-L}^{L} a_m h_{K(i+m)} a_{Ki}^* - \sum_{n=-L}^{L} a_n^* h_{K(i+n)}^* a_{Ki}\right]$ (3.83)
- Average power $P = E|a_{Ki}|^2$ (3.3)
- Autocorrelation of fading-plus-noise samples

$$E\left[h_{K(i+m)} h_{K(i+n)}^*\right] = E\left[(a_{K(i+m)} + w_{K(i+m)})(a_{K(i+n)}^* + w_{K(i+n)}^*)\right] \qquad (3.84)$$

$$= E\left[a_{K(i+m)} a_{K(i+n)}^*\right] + E\left[w_{K(i+m)} w_{K(i+n)}^*\right]$$

$$= \begin{cases} r_{K(m-n)} & m \neq n \\ r_0 + s^2 & m = n \end{cases}$$

- Similarly, we find that

$$E\left[h_{K(i+m)} a_{Ki}^*\right] = E\left[(a_{K(i+m)} + w_{K(i+m)}) a_{Ki}^*\right] \qquad (3.86)$$

$$= E\left[a_{K(i+m)} a_{Ki}^*\right] + E\left[w_{K(i+m)} w_{K(i+n)}^*\right]$$

$$= r_{Km}$$

- Rewrite the cost function in compact form

$$j = P + a^{+} Ra - a^{+} r - r^{+} a \qquad (3.87)$$

- By completing the square, we get

$$J = P - r^{+} R^{-1} r + (a - R^{-1} r)^{+} R(a - R^{-1} r) \qquad (3.88)$$

- Wiener-Hopf equation

$$a = R^{-1} r \qquad (3.89)$$

- It provides the solution for weighting what is the optimal estimate of fading in mean-square error sense.

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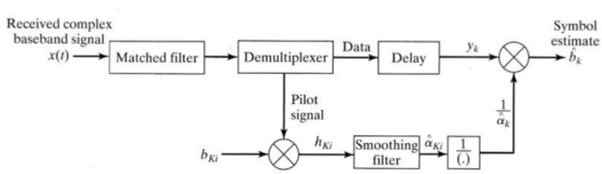

FIGURE 3.31 Pilot scheme for tracking and compensating for channel variations.
Note: The b_{Ki} are pilot symbols that, by definition, are known at the receiver.

- Energy loss: devoted to pilot symbol
- Expanding the bandwidth to accommodate the inclusion of the pilot symbols and data.
- Minimum pilot symbol rate: several times of fading process bandwidth.

3.12 Receiver Performance: Bit Error Rate

- Bit error rate
  - average probability of symbol error
- Receiver performance is considered under:
  - Presence of additive channel noise
  - Presence of multiplicative noise exemplified by frequency-flat, slowly fading channel

3.12 Receiver Performance: Bit Error Rate

3.12.1 Channel Noise

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TABLE 3.4 Summary of Formulas for the bit error rate (BER) of coherent and noncoherent digital communication receivers.

Signaling Scheme	BER (Additive white Gaussian noise channel)	BER (Slow Rayleigh fading channel)
(a) Coherent BPSK Coherent QPSK Coherent MSK	$\frac{1}{2}\text{erfc}\left(\sqrt{\frac{E_b}{N_0}}\right)$	$\frac{1}{2}\left(1-\sqrt{\frac{\gamma_0}{1+\gamma_0}}\right)$
(b) Coherent BFSK	$\frac{1}{2}\text{erfc}\left(\sqrt{\frac{E_b}{2N_0}}\right)$	$\frac{1}{2}\left(1-\sqrt{\frac{\gamma_0}{2+\gamma_0}}\right)$
(c) Binary DPSK	$\frac{1}{2}\exp\left(-\frac{E_b}{N_0}\right)$	$\frac{1}{2(1+\gamma_0)}$
(d) Noncoherent BFSK	$\frac{1}{2}\exp\left(-\frac{E_b}{2N_0}\right)$	$\frac{1}{2+\gamma_0}$

Definitions:
E_b = transmitted energy per bit
N_0 = one-sided power spectral density of channel noise
γ_0 = mean value of the received energy per bit-to-noise spectral density ratio

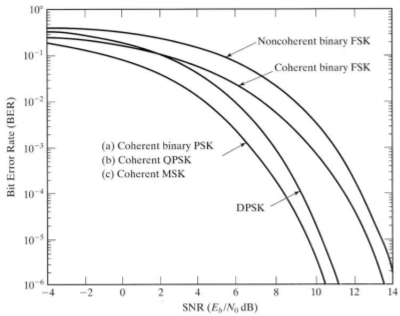

FIGURE 3.32 Comparison of the noise performance of different PSK and FSK schemes.

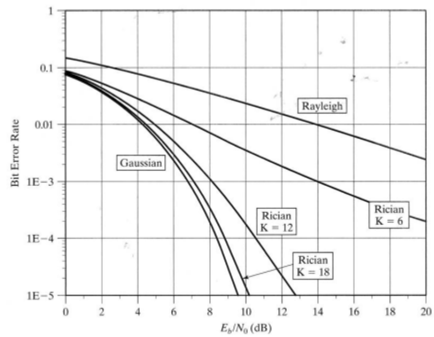

FIGURE 3.33 Comparison of performance of coherently-detected BPSK over different fading channels.

- BER of GMSK

$$P_e \approx \frac{1}{2} erfc\left(\sqrt{\frac{bE_b}{2N_0}}\right) \qquad (3.90)$$

- BER for slow Rayleigh fading channel

$$g_0 = \frac{E_b}{N_0} E\left[a^2\right] \qquad (3.91)$$