# Guided Tour

The book has **three Parts**, namely

I: Introduction to Digital Communication

II: Baseband Systems

III: Bandpass Systems

***Part I***

***Introduction to Digital Communication***

- Introduction
- Signals and Systems: Some Fundamental Concepts

***Part II***

***Baseband Systems***

- Baseband Transmission Techniques: Formatting
- Baseband Transmission Techniques: Coding
- Baseband Reception Techniques

***Part III***

***Bandpass Systems***

- Bandpass Signal Transmission
- Bandpass Signal Reception

**Example 2.22** For the sinusoidal wave with random phase defined in Example 2.15,

(a) Find the time average of $x(\theta, t)$.
(b) Is the process ergodic in mean?

(a)

$$\mu_x(T) = \frac{1}{2T}\int_{-T}^{T} A\cos(\omega_0 t + \theta)\ \mathrm{d}t$$
$$= A\,\frac{\sin\omega_0 T}{\omega_0 T}\ \cos\theta$$
$$\lim_{T\to\infty} \mu_x(T) = 0$$

So, the time average of $x(\theta, t)$ is zero.

(b) The limit of $\mu_x(T)$ is independent of $T$ and constant and (the limit of) its variance, is also zero. Previously we had seen in Example 2.15, that its ensemble mean $\mu_X$ is zero. So, we see that the process $x(\theta, t)$ satisfies Eq. 2.132. Also in Example 2.18, we had seen that the process is WSS. So, the process $x(\theta, t)$ can be said to be ergodic in mean.

**Example 2.23** Check for the ergodicity in mean of the sinusoidal wave with random phase pdf given by Example 2.19.

We had already seen in Example 2.19 that the process is not wide-senses stationary. So, it cannot be ergodic. To underscore this we mention that in Example 2.19 the ensemble mean $\mu_X$ of the process was found to be $\frac{2\sqrt{2}A}{\pi}\cos\omega_0 t$. The time average does not depend on pdf. So, it is same as the time-average of Example 2.22 whose value is zer
ensemble average, we reconfirm our conclusion that t

To elaborate the concept of ergodicity with the hel
the mean and autocorrelation obtained by noting the
particular receiver over sufficiently long time of obs
autocorrelation calculated from the noise voltages sim
identical receivers at any instant of time, if the noise v

It is difficult to test whether a process is ergod
order statistical parameters of the process and prove
parameters. Nevertheless, in practice, we mainly requ
to characterise it. These are mean and autocorrelation
is ergodic in mean and autocorrelation, then our requ
the sample function's time average satisfies Eq. 2.132

Interspersed within the text are **Solved Examples**. These will help the students in better understanding the concepts learned in the preceding sections




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## *Problems*

6.1 A voice-band telephone channel passes the frequencies in the band from 300 Hz to 3.3 kHz. It is desired to design a modem that transmits at a symbol rate of 2.4 ksymbols/sec, with the objective of achieving 9.6 kbps. Select an appropriate QAM signal constellation, carrier frequency and the roll-off factor of a pulse with a raised cosine spectrum that utilises the entire frequency band. Sketch the spectrum of the transmitted signal pulse and indicate the important frequencies.

6.2 (a) In a MSK signal, the initial state for the phase is either 0 or $\pi$ radian. Determine the terminal phase state for the following four input pairs of input data corresponding to each initial state phase:

(a) 00
(b) 01
(c) 10
(d) 11

(b) A partial response MSK has a pulse given by

$$g(t) = \begin{cases} \frac{1}{4T} & 0 \le t \le 2T \\ 0 & \text{otherwise} \end{cases}$$

(a) Draw the phase tree
(b) Draw the state trellis
(c) Draw the state diagram

6.3 (a) Calculate the minimum required bandwidth for a noncoherently detected orthogonal binary FSK system. The symbol duration is 0.1 ms.

Chapter end **Problems** are carefully designed to ensure that the students can practice the theory learnt in the chapters on their own

A comprehensive **Bibliography** at the end of every chapter gives the list of other related titles for those interested. The **Further Reading** section just above it will serve to guide the readers to the Books of their interest.



waveform, then the SNR requirement per bit is 20 unit, whereas if the waveform was 1024-ary, the SNR per bit is 2 unit only. Why should we have to go through such a roundabout computational manipulations to find a metric that represents a figure of merit? Why not describe the metric in terms of an energy related parameter at the bit level? $E_b/\eta$ or $\gamma_b$ foots the bill. It is the ratio of the bit energy of the digital signal waveform to the power spectral density of channel noise. We have already shown in Eq. 5.170 that it is SNR times a conversion factor for analog bandwidth to digital bit transmission rate. But a doubt persists whether $\gamma_b$ is dimensionless like SNR, because its numerator and denominator do not look like same category of physical quantity; one is energy, the other is PSD. To verify the units of $\gamma_b$, we write

$$\gamma_b = \frac{E_b}{\eta} = \frac{[\text{Joule}]}{[\text{Watt}/Hz]} = \frac{[\text{Watt} - s]}{[\text{Watt} - s]}$$

So, $\gamma_b$ also is a dimensionless quantity. Thus we see that $\gamma_b$ clearly satisfies all the conceptual requirements demanded from a figure-of-merit of a digital system. Hence, $\gamma_b$, and not SNR, is used as a figure-of-merit when the detectors of different digital receivers are compared.

### 5.6 FURTHER READING

Many times the students are not clear about the sources of noise in communication systems. We recommend two papers by Johnson [1] and Nyquist [2]. For mathematical treatment of random noise in presence of a deterministic signal [3], [4] and [5] can be read.

Readers interested to go deeper in the theory of matched filters can see the papers [6], [8] and [14].

Various pulse shaping techniques based on the second criterion of Nyquist can be looked into [9], [13], [15] and [19]. For further discussion on Nyquist's third criterion read the material present in [12] and [22]. For advanced discussion on adaptive equalisation techniques, the papers [18], [20] and [21] are highly recommended.

A tutorial on various synchronisation techniques used in digital communication receiver can be ... Franks [17].

... papers by Arthurs et al [7] and Forney [11] for a discussion on the ... on process.

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**Cyclic Codes** Among the error correcting linear block codes, only the single error correcting ones are easy to design and no systematic procedure exists for codes correcting more than one error. Cyclic codes are a subclass of linear block codes, that have got a fair amount of mathematical structure and hence can be systematised for correcting higher number of errors. Also, for cyclic codes, encoding and syndrome calculations are easily implemented by shift registers. So, we impose an additional constraint on the linear code for the purpose of defining cyclic codes.

***Definition 4.9*** *A code C is* ***cyclic*** *if*

(i) *C is a linear code*
(ii) *any cyclic shift of a codeword is also a codeword.*

If $C$ is a $(n, k)$ cyclic code, then $C^{(i)}$ is the code obtained by shifting the codeword cyclically $i$ places to the left, i.e.

$$C^{(i)} = (c_{i+1}, c_{i+2}, \cdots, c_n, c_1, c_2, \cdots, c_i)$$

Cyclic codes can be described in a polynomial form. Thus, a codeword of $C$ can be expressed as a $(n-1)$ degree polynomial,

$$c(x) = c_1 x^{n-1} + c_2 x^{n-2} + \cdots + c_n \tag{4.24}$$

The coefficients of the polynomial are binary valued and obeys modulo-2 arithmetic. Similarly, the code polynomial for the code $C^{(i)}$ is

$$c^{(i)}(x) = c_{i+1} x^{n-1} + c_{i+2} x^{n-2} + \cdots + c_n x^i + c_1 x^{i-1} + \cdots + c_i \tag{4.25}$$

One very interesting property of the cyclic code polynomials is,

***Property V*** *When $x^i c(x)$ is divided by $x^n + 1$, the remainder is $c^{(i)}(x)$.*

We can verify this property as follows:
for $i = 1$, $xc(x) = c_1 x^n + c_2 x^{n-1} + \cdots + c_n x$

$$\begin{array}{r|l} & c_1 \\ x^n + 1 & c_1 x^n + c_2 x^{n-1} + \cdots + c_n x \\ & c_1 x^n + c_1 \\ \hline & c_2 x^{n-1} + c_3 x_{n-2} + \cdots + c_n x + c_1 \leftarrow \text{rem} \end{array}$$

The remainder is clearly $c^{(i)}(x)$. We now introduce an important theo

Other features include **Definitions**, **Laws**, **Properties** and **Relations** that will further aid the readers in their study. These are set off in shaded boxes to provide easy referencing for the students.



However due to Eq. 2.22 we need only $2BT_0$ real coefficients to describe $s(t)$. The coefficients $C_n$ actually depend on the shape of $s(t)$ over the time interval $T_0$. The formal statement of this relation is called Dimensionality Theorem.

***Theorem 2.1 (Dimensionality Theorem)***
*A real waveform can be completely specified by $N$ independent pieces of information where $N$ is given by*

$$N = 2BT_0 \tag{2.25}$$

$N$ is called the dimension of the waveform in signal space. Here $B$ is the bandwidth of the signal and $T_0$ is the time over which the signal waveform is being described. So, following the theorem, one can represent the signal $s(t)$ by $N$ finite number of basis functions.

$$s(t) = \sum_{k=1}^{N} s_k \psi_k(t) \quad 0 \le t \le T_0 \tag{2.26}$$

where $\{s_k\}$ and $\{\psi_k\}$ are the digital data set and basis set respectively of the signal $s(t)$.

Note that $N$ sample values do not necessarily have to be periodic samples of the waveform. They only need to be independent basis coefficients. Also, the theorem, though derived by considering Fourier basis, can be proved for any basis set.

If one wants to store a bandlimited digital signal and wants to reconstruct the signal at a later time over a $T_0$ second interval, then at least $N$ points should be stored. So, dimensionality theorem helps to calculate the storage space required to store a digital signal.

Another important application of the dimensionality theorem is to estimate the bandwidth of a digital signal. A digital signal is represented by $N$ points which are transmitted over an interval of $T_0$ second. So, the symbol rate $R_s$ is

$$R_s = \frac{N}{T_0} = \frac{2BT_0}{T_0} = 2B$$

So, the bandwidth of the digital signal is

$$B = \frac{R_s}{2} \tag{2.27}$$

This is actually a lower bound of the bandwidth of digital signal. If the signal is represented in terms of $\sin x/x$ basis set, then only this lower limit is achieved. But $\sin x/x$ is an analog signal. So, the lowest bandwidth of the digital data is achieved only when the basis functions are occupying the least spectrum in frequency (no sharp corners and bends in time domain). For other basis set, the bandwidth is more than this lowest value. Specifically, for rectangular pulse basis, the bandwidth equals the symbol rate. Sometimes, the symbol rate $R_s$ is also represented as $D$ (Baud rate). So, in general, we may state that the bandwidth of digital signal or data is given by,

$$B \ge \frac{D}{2} \tag{2.28}$$

Here we want to emphasise a difference between bandwidth and symbol rate of a digital signal. Bandwidth is the significant spectral width of the signal that contains the most important properties of

Throughout the book various **Theorems** are given in the relevant sections for students' reference.

**Figures** are used extensively to illustrate the key concepts and methods



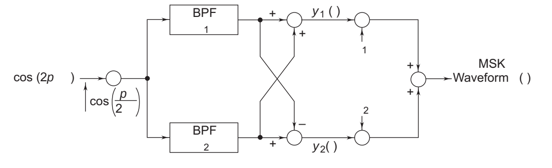


**Fig. 6.24** *MSK Modulator*

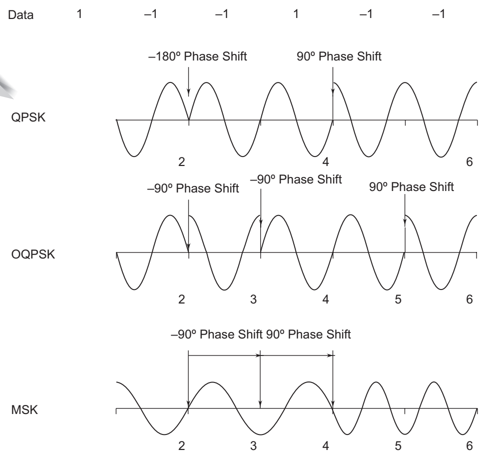


**Fig. 6.25** *Comparison of Waveforms for QPSK, OQPSK and MSK*

...m contains phase jumps of ±90° that may occur as often as every $T_s$ seconds. On ...a QPSK signal contains phase jumps of either ±180° or ±90° every $2T_s$ seconds.

# *List of Abbreviations*

| | |
|---|---|
| ADF | Adaptive Prediction with Forward Estimation |
| ADM | Adaptive Delta Modulator |
| ADPMC | Adaptive Differential Pulse Coded Modulation |
| ADSL | Asynchronous Digital Subscribers Line |
| AGC | Automatic Gain Control |
| AM | Amplitude Modulation |
| APB | Adaptive Prediction with Backward Estimation |
| AQB | Adaptive Quantisation with Backward Estimation |
| AQF | Adaptive Quantisation with Forward Estimation |
| ASBC | Adaptive Suband Coding |
| ASK | Amplitude Shift Keying |
| AWGN | Additive White Gaussian Noise |
| BBC | British Broadcasting Corporation |
| BCH | Bore-Chaudhuri-Hocquenghem |
| BEC | Binary Erasure Channel |
| BER | Bit Error Rate |
| BPSK | Binary Phase Shift Keying |
| BSC | Binary Symmetric Channel |
| CC | Convolutional Code |
| CCITT | Consultative Committee on International Telegraphy and Telephony |
| CD | Compact Disk |
| CDMA | Code Division Multiple Access |
| CELP | Code-Excited Linear P... |
| CPFSK | Continuous Phase Frequency S... Keying |
| CPM | Continuous Phase Modulation |
| CRC | Cyclic Redundancy Check |
| CVSDM | Continuously Variable Slope Delta Modulation |
| CVSDM | Continuously Variable Slope Delta Modulation |
| DAT | Digital Audio Tape |
| DBV | Digital Video Broadcast |
| DCT | Discrete Cosine Transform |
| DFE | Decision Feedback Equaliser |
| DFT | Discrete Fourier Transform |
| DM | Delta Modulator |
| DMS | Discrete Memoryless Source |
| DPCM | Differential Pulse Code Modulation |
| DPSK | Differential Phase Shift Keying |
| DQPSK | Differential Quadrature Phase Shift Keying |
| DSL | Digital Subscribers Line |
| DTE | Data Terminal Equipment |
| ECC | Error Correction Codes |
| FDM | Frequency Division Multiplexing |
| FDMA | Frequency Division Multiple Access |
| FM | Frequency Modulation |
| FSK | Frequency Shift Keying |
| GIF | Graphic Interchange Format |
| GMSK | Gaussian Minimum Shift Keying |
| GRN | Gaussian Random Noise |

A **List of Abbrevations** at the beginning of the book will be helpful to the students for quick referencing

*Appendix*

# A

# *Q and Error Functions*

### A.1 Q FUNCTION

$Q$ function is defined in Eq. 2.69 and reproduced here for convenience,

$$Q(z) = \int_z^{\infty} \frac{1}{\sqrt{2\pi}} e^{-\frac{u^2}{2}} \mathrm{d}u \quad \text{(A.1)}$$

The $Q$-function is bounded by two analytical expressions as follows:

$$\left(1 - \frac{1}{z^2}\right) \frac{1}{z\sqrt{2\pi}} e^{-z^2/2} \leq Q(z) \leq\approx \frac{1}{z\sqrt{2\pi}} \exp(-\frac{z^2}{2}) \quad \text{(A.2)}$$

So, it is apparent that for larger values of $z$, $Q$-function can be approximated by any of these expressions as the error committed becomes insignificant. Usually these analytic approximations are used for $z$ greater than 3.

A tabulation of the $Q$-function upto $z = 3.9$ is given in Table A.1 an
in Fig. A.1.

### A.2 THE ERROR FUNCTIONS

The error function *erf* is defined as,

$$erf(z) = \frac{2}{\sqrt{\pi}} \int_0^z e^{-\frac{u^2}{2}} \mathrm{d}u$$

and the complementary error function *erfc* is defined as,

$$erfc(z) = \frac{2}{\sqrt{\pi}} \int_z^{\infty} e^{-\frac{u^2}{2}} \mathrm{d}u$$

Three **Appendices** at the end of the book, namely
A: Q and error Functions
B: Continuous Phase Modulation (CPM)
C: Decision-Directed Carrier Recovery provide additional information to the readers of the book

*Appendix*

# B

# *Continuous Phase Modulation (CPM)*

### B.1 REPRESENTATION

CPFSK signals defined by Eqs 6.60 and 6.61 is a special case of a more general signal set called CPM signal in which the phase of the modulated waveform is given by [14] (Chapter 6).

$$\phi(t; a) = 2\pi \sum_{k=0}^{n} a_k h_k q(t - kT_s), \quad nT_s \leq t \leq (n+1)T_s \quad \text{(B.1)}$$

where $\{a_k\}$ is the sequence of M-ary information symbols, selected from the alphabet $\pm1, \pm3, \cdots, \pm(M-1)$; $\{h_k\}$ is a sequence of modulation indices and $q(t)$ is some normalised waveform shape obtained by integrating the basic

When $h_k = h$ for all $k$, the modulation index is
varies from one symbol to another, the CPM sig
made to vary in a cyclic manner through a set of i

The waveform $q(t)$ is the integral of the pulse

$$q(t) = \int_0$$

If $f(t) = 0$ for $t > T_s$, the CPM waveform is
the modulated waveform is called *partial response*
pulse shapes are shown in Figs B.1 and B.2. It is
can be generated by choosing different pulse sha
and the alphabet size $M$.

Unlike other modulated waveforms like ASK
continuous-phase waveform cannot be represente
phase of the continuous-phase modulated wave g

*Appendix*

# C

# *Decision-Directed Carrier Recovery*

Let us begin by estimating an unknown parameter $\beta$ associated with a received noisy signal,

$$r(t) = g(t, \beta) + n(t) \quad \text{(C.1)}$$

where $n(t)$ is the AWGN noise having a PSD $\eta/2$. We can represent the signal $r(t)$ in terms of its N-dimensional basis set and write,

$$r(t) = \sum_k r_k \Psi_k(t) = \sum_k [g_k(\beta) + n_k] \Psi_k(t) \quad \text{(C.2)}$$

So, the sample values of $r(t)$ are $r_k$ and given by $r_k = g_k + n_k$. The unconditional pdf of the random vector $R$ is,

$$p_R = \lim_{N\to\infty} \left(\frac{1}{\sigma\sqrt{2\pi}}\right)^N \exp\left[-\frac{1}{2\sigma^2} \sum_{k=1}^{N} |r_k|^2\right] \quad \text{(C.3)}$$

Also, the expectation of the conditional event $r_k|\beta$ is $E[r_k|\beta] = g_k(\beta)$. With this statistics, we can write the joint pdf of the conditional event $r_k|\beta$

$$p_{R|\beta} = \lim_{N\to\infty} \left(\frac{1}{\sigma\sqrt{2\pi}}\right)^N \exp\left[-\frac{1}{2\sigma^2} \sum_{k=1}^{N} |r_k - g_k(\beta)|^2\right] \quad \text{(C.4)}$$

We are interested in finding the value of $\beta$ that maximises the conditional probability. To find this value, we could differentiate the conditional probability directly or, equivalently, we could form the likelihood ratio,

$$\frac{p_{R|\beta}}{p_R} = \lim_{N\to\infty} \exp\left[\frac{1}{2\sigma^2} \sum_{k=1}^{N} [r_k^* g_k(\beta) + r_k g_k^* - |g_k(\beta)|^2]\right] \quad \text{(C.5)}$$