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for
Digital Communication

Won Y. Yang, Yong S. Cho, Won G. Jeon, Jeong W. Lee, Jong H. Paik
Jae K. Kim, Mi-Hyun Lee, Kyu I. Lee, Kyung W. Park, Kyung S. Woo
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The MathWorks, Inc.
3 Apple Hill Drive
Natick, MA 01760-2098, USA
☎ 508-647-7000,  Fax: 508-647-7001
E-mail: info@mathworks.com
Web: [www.mathworks.com](http://www.mathworks.com)

Questions about the contents of this book can be mailed to wyyang53@hanmail.net.

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   who love and support us
   and
   to our teachers and students
   who enriched our knowledge
# Table of Contents

**PREFACE** iii

**CHAPTER 1: FOURIER ANALYSIS** 1

1.1 CONTINUOUS-TIME FOURIER SERIES (CTFS) ................................................................. 2
1.2 PROPERTIES OF CTFS ................................................................................................. 6
   1.2.1 Time-Shifting Property ....................................................................................... 6
   1.2.2 Frequency-Shifting Property .............................................................................. 6
   1.2.3 Modulation Property ....................................................................................... 6
1.3 CONTINUOUS-TIME FOURIER TRANSFORM (CTFT) ...................................................... 7
1.4 PROPERTIES OF CTFT ......................................................................................................... 13
   1.4.1 Linearity ................................................................................................................. 13
   1.4.2 Conjugate Symmetry ............................................................................................ 13
   1.4.3 Real Translation (Time Shifting) and Complex Translation (Frequency Shifting) ..... 14
   1.4.4 Real Convolution and Correlation ....................................................................... 14
   1.4.5 Complex Convolution – Modulation/Windowing ................................................ 14
   1.4.6 Duality ................................................................................................................ 17
   1.4.7 Parseval Relation - Power Theorem ...................................................................... 18
1.5 DISCRETE-TIME FOURIER TRANSFORM (DTFT) .......................................................... 18
1.6 DISCRETE-TIME FOURIER SERIES - DFS/DFT ................................................................ 19
1.7 SAMPLING THEOREM ....................................................................................................... 21
   1.7.1 Relationship between CTFS and DFS .................................................................... 21
   1.7.2 Relationship between CTFT and DTFT ................................................................ 27
   1.7.3 Sampling Theorem .............................................................................................. 27
1.8 POWER, ENERGY, AND CORRELATION ........................................................................ 29
1.9 LOWPASS EQUIVALENT OF BANDPASS SIGNALS ....................................................... 30
Problems ........................................................................................................................................ 36

**CHAPTER 2: PROBABILITY AND RANDOM PROCESSES** 39

2.1 PROBABILITY .................................................................................................................. 39
   2.1.1 Definition of Probability ....................................................................................... 39
   2.1.2 Joint Probability and Conditional Probability ....................................................... 40
   2.1.3 Probability Distribution/Density Function ............................................................ 41
   2.1.4 Joint Probability Density Function ....................................................................... 41
   2.1.5 Conditional Probability Density Function ............................................................ 41
   2.1.6 Independence ....................................................................................................... 41
   2.1.7 Function of a Random Variable ............................................................................ 42
   2.1.8 Expectation, Covariance, and Correlation ............................................................ 43
   2.1.9 Conditional Expectation ...................................................................................... 47
   2.1.10 Central Limit Theorem - Normal Convergence Theorem .................................. 47
   2.1.11 Random Processes .............................................................................................. 49
   2.1.12 Stationary Processes and Ergodic Processes ....................................................... 51
   2.1.13 Power Spectral Density (PSD) ............................................................................ 53
   2.1.14 White Noise and Colored Noise ....................................................................... 53
2.2 LINEAR FILTERING AND PSD OF A RANDOM PROCESS ............................................ 57
2.3 FADING EFFECT OF A MULTI-PATH CHANNEL .......................................................... 59
Problems ........................................................................................................................................ 62
<table>
<thead>
<tr>
<th>CHAPTER 3: ANALOG MODULATION</th>
<th>71</th>
</tr>
</thead>
<tbody>
<tr>
<td>3.1 AMPLITUDE MODULATION (AM)</td>
<td>71</td>
</tr>
<tr>
<td>3.1.1 DSB (Double Sideband)-AM (Amplitude Modulation)</td>
<td>71</td>
</tr>
<tr>
<td>3.1.2 Conventional AM (Amplitude Modulation)</td>
<td>75</td>
</tr>
<tr>
<td>3.1.3 SSB (Single Sideband)-AM(Amplitude Modulation)</td>
<td>78</td>
</tr>
<tr>
<td>3.2 ANGLE MODULATION - FREQUENCY/PHASE MODULATIONS</td>
<td>82</td>
</tr>
<tr>
<td>Problems</td>
<td>86</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>CHAPTER 4: ANALOG-TO-DIGITAL CONVERSION</th>
<th>87</th>
</tr>
</thead>
<tbody>
<tr>
<td>4.1 QUANTIZATION</td>
<td>87</td>
</tr>
<tr>
<td>4.1.1 Uniform Quantization</td>
<td>88</td>
</tr>
<tr>
<td>4.1.2 Non-uniform Quantization</td>
<td>89</td>
</tr>
<tr>
<td>4.1.3 Non-uniform Quantization Considering Relative Errors</td>
<td>91</td>
</tr>
<tr>
<td>4.2 Pulse Code Modulation (PCM)</td>
<td>95</td>
</tr>
<tr>
<td>4.3 Differential Pulse Code Modulation (DPCM)</td>
<td>97</td>
</tr>
<tr>
<td>4.4 Delta Modulation (DM)</td>
<td>100</td>
</tr>
<tr>
<td>Problems</td>
<td>103</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>CHAPTER 5: BASEBAND DIGITAL TRANSMISSION</th>
<th>107</th>
</tr>
</thead>
<tbody>
<tr>
<td>5.1 RECEIVER (RCVR) and SNR</td>
<td>107</td>
</tr>
<tr>
<td>5.1.1 Receiver of RC Filter Type</td>
<td>109</td>
</tr>
<tr>
<td>5.1.2 Receiver of Matched Filter Type</td>
<td>110</td>
</tr>
<tr>
<td>5.1.3 Signal Correlator</td>
<td>112</td>
</tr>
<tr>
<td>5.2 SIGNALING AND ERROR PROBABILITY</td>
<td>114</td>
</tr>
<tr>
<td>5.2.1 Antipodal (Bipolar) Signaling</td>
<td>114</td>
</tr>
<tr>
<td>5.2.2 OOK(On-Off Keying)/Unipolar Signaling</td>
<td>118</td>
</tr>
<tr>
<td>5.2.3 Orthogonal Signaling</td>
<td>119</td>
</tr>
<tr>
<td>5.2.4 Signal Constellation Diagram</td>
<td>121</td>
</tr>
<tr>
<td>5.2.5 Simulation of Binary Communication</td>
<td>123</td>
</tr>
<tr>
<td>5.2.6 Multi-level(amplitude) PAM Signaling</td>
<td>127</td>
</tr>
<tr>
<td>5.2.7 Multi-dimensional Signaling</td>
<td>129</td>
</tr>
<tr>
<td>5.2.8 Bi-orthogonal Signaling</td>
<td>133</td>
</tr>
<tr>
<td>Problems</td>
<td>136</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>CHAPTER 6: BANDLIMITED CHANNEL AND EQUALIZER</th>
<th>139</th>
</tr>
</thead>
<tbody>
<tr>
<td>6.1 BANDLIMITED CHANNEL</td>
<td>139</td>
</tr>
<tr>
<td>6.1.1 Nyquist Bandwidth</td>
<td>139</td>
</tr>
<tr>
<td>6.1.2 Raised-Cosine Frequency Response</td>
<td>141</td>
</tr>
<tr>
<td>6.1.3 Partial Respose Signaling - Duobinary Signaling</td>
<td>143</td>
</tr>
<tr>
<td>6.2 EQUALIZER</td>
<td>148</td>
</tr>
<tr>
<td>6.2.1 Zero-Forcing Equalizer (ZFE)</td>
<td>148</td>
</tr>
<tr>
<td>6.2.2 MMSE Equalizer (MMSEE)</td>
<td>151</td>
</tr>
<tr>
<td>6.2.3 Adaptive Equalizer (ADE)</td>
<td>154</td>
</tr>
<tr>
<td>6.2.4 Decision Feedback Equalizer (DFE)</td>
<td>155</td>
</tr>
<tr>
<td>Problems</td>
<td>159</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>CHAPTER 7: BANDPASS DIGITAL TRANSMISSION</th>
<th>169</th>
</tr>
</thead>
<tbody>
<tr>
<td>7.1 AMPLITUDE MODULATION - AMPLITUDE SHIFT KEYING (ASK)</td>
<td>169</td>
</tr>
<tr>
<td>7.2 FREQUENCY MODULATION - FREQUENCY SHIFT KEYING (FSK)</td>
<td>178</td>
</tr>
<tr>
<td>7.3 PHASE MODULATION - PHASE SHIFT KEYING (PSK)</td>
<td>187</td>
</tr>
<tr>
<td>7.4 DIFFERENTIAL PHASE SHIFT KEYING (DPSK)</td>
<td>190</td>
</tr>
<tr>
<td>7.5 QUADRATURE AMPLITUDE MODULATION (QAM) - PAM/PSK</td>
<td>195</td>
</tr>
</tbody>
</table>
7.6 COMPARISON OF VARIOUS SIGNALINGS .............................................................. 200
Problems ........................................................................................................... 205

CHAPTER 8: CARRIER RECOVERY AND SYMBOL SYNCHRONIZATION ............... 225
8.1 INTRODUCTION ..................................................................................................... 225
8.2 PLL (PHASE-LOCKED LOOP) .................................................................................. 226
8.3 ESTIMATION OF CARRIER PHASE USING PLL .................................................... 231
8.4 CARRIER PHASE RECOVERY .................................................................................. 233
  8.4.1 Carrier Phase Recovery Using a Squaring Loop for BPSK Signals .................... 233
  8.4.2 Carrier Phase Recovery Using Costas Loop for PSK Signals ............................ 235
  8.4.3 Carrier Phase Recovery for QAM Signals ....................................................... 238
8.5 SYMBOL SYNCHRONIZATION (TIMING RECOVERY) ......................................... 241
  8.5.1 Early-Late Gate Timing Recovery for BPSK Signals ........................................ 241
  8.5.2 NDA-ELD Synchronizer for PSK Signals ....................................................... 244
Problems ........................................................................................................... 247

CHAPTER 9: INFORMATION AND CODING .............................................................. 255
9.1 MEASURE OF INFORMATION - ENTROPY ............................................................ 255
9.2 SOURCE CODING .................................................................................................. 256
  9.2.1 Huffman Coding ............................................................................................... 256
  9.2.2 Lempel-Zip-Welch Coding ............................................................................. 259
  9.2.3 Source Coding vs. Channel Coding ............................................................... 262
9.3 CHANNEL MODEL AND CHANNEL CAPACITY ...................................................... 263
9.4 CHANNEL CODING ................................................................................................ 268
  9.4.1 Waveform Coding ............................................................................................ 269
  9.4.2 Linear Block Coding ....................................................................................... 270
  9.4.3 Cyclic Coding ................................................................................................... 279
  9.4.4 Convolutional Coding and Viterbi Decoding ................................................... 284
  9.4.5 Trellis-Coded Modulation (TCM) ................................................................. 293
  9.4.6 Turbo Coding ................................................................................................ 297
  9.4.7 Low-Density Parity-Check (LDPC) Coding ................................................... 308
  9.4.8 Differential Space-Time Block Coding (DSTBC) ........................................... 313
9.5 CODING GAIN ....................................................................................................... 316
Problems ........................................................................................................... 318

CHAPTER 10: SPREAD-SPECTRUM SYSTEM ............................................................. 337
10.1 PN (Pseudo Noise) Sequence ................................................................................ 337
10.2 DS-SS (Direct Sequence Spread Spectrum) ....................................................... 345
10.3 FH-SS (Frequency Hopping Spread Spectrum) .................................................. 350
Problems ........................................................................................................... 354

CHAPTER 11: OFDM SYSTEM .................................................................................. 357
11.1 OVERVIEW OF OFDM ....................................................................................... 357
11.2 FREQUENCY BAND AND BANDWIDTH EFFICIENCY OF OFDM ...................... 361
11.3 CARRIER RECOVERY AND SYMBOL SYNCHRONIZATION ............................... 362
11.4 CHANNEL ESTIMATION AND EQUALIZATION ............................................... 379
11.5 INTERLEAVING AND DEINTERLEAVING ....................................................... 382
11.6 PUNCTURING AND DEPUNCTURING ............................................................ 384
11.7 IEEE STANDARD 802.11A - 1999 ................................................................. 386
Problems ........................................................................................................... 393
APPENDICES 407
Appendix A: Fourier Series/Transform ................................................................. 407
Appendix B: Laplace Transform and \( z \)-Transform ........................................... 412
Appendix C: Differentiation w.r.t. a Vector ........................................................ 414
Appendix D: Useful Formulas ........................................................................... 415
Appendix E: MATLAB Introduction .................................................................. 417
Appendix F: Simulink ....................................................................................... 421

REFERENCES 425

INDEX 427
Preface

This book has been designed as a reference book for students or engineers studying communication systems possibly in the curriculum of Electrical Engineering program rather than a text book for any course on communication. Readers are supposed to have taken at least two junior-level courses, one on signals and systems and another one on probability and random processes. In other words, readers should have a basic knowledge about the linear system, Fourier transform, Laplace transform, $z$-transform, probability, and random processes although the first two chapters of this book provide a brief overview of some background topics to minimize the necessity of the prerequisite courses and to refresh their memory if nothing else.

It is not the aim of this book to provide any foundation in the basic theory of digital communication since the authors do not have such a deep knowledge as to do it. The first aim of this book is to help the readers understand the concepts, techniques, terminologies, equations, and block diagrams appearing in the existing books on communication systems while using MATLAB® to simulate the various communication systems most of which are described by block diagrams and equations. Needless to say, the readers are recommended to learn some basic usage of MATLAB® that is available from the MATLAB help function or the on-line documents at the web site <http://www.mathworks.com/matlabcentral/>. However, they are not required to be so good at MATLAB® since most programs in this book have been composed carefully and completely so that they can be understood in connection with related/referred equations and/or block diagrams. The readers are expected to get used to MATLAB software while trying to modify/use the MATLAB® codes and Simulink® models in this book for solving the end-of-chapter problems or their own problems. The second and main aim of this book is to make even a novice at both MATLAB® and communication systems become acquainted, at least comfortable, with MATLAB® as well as communication systems while running the MATLAB programs on his/her computer and trying to understand what is going on in the systems simulated by the programs. Is it too much to expect that a novice will become interested in communications and simultaneously fall in love with MATLAB®, which is a universal language for engineers and scientists after having read this book through? Is it just the authors’ imagination that the readers would think of this book describing and explaining many concepts in MATLAB® rather than in English? In any case, the authors have no intention to hide their hope that this book will be one of the all-the-time-reserved books in most libraries and can be found always on the desks of most communication engineers. The features of this book can be summarized as follows:

1. This book presents more MATLAB programs for the simulation of communication systems than any existent books with the same or similar titles as an approach to explain most things using MATLAB® and figures rather than English and equations.
2. Most MATLAB programs are presented in a complete form so that the readers can run them instantly with no programming skill and focus on understanding the behavior and characteristic of the simulated systems and making interpretations based on the tentative and final simulation results.
3. Many programs have a style of on-line processing rather than batch processing so that the readers can easily understand the whole system and the underlying algorithm in details block by block and operation by operation. Furthermore, the on-line processing style of the programs is expected to let the readers develop their insight into the real system.
4. Authors never think that this book can replace the existent books made by many great authors to whom they are not comparable to. They neither expect that this book can take the place of the MATLAB manual. Instead, this book is designed to play a role of bridge...
between MATLAB® software and the theory, block diagrams, and equations appearing in the field of communications so that the readers can feel free to utilize MATLAB® software for studying communication systems and become much more interested in communications than before reading this book.

The contents of this book are derived from the works of many (known or unknown) great scientists, scholars, and researchers, all of whom are deeply appreciated. We would like to thank the reviewers for their valuable comments and suggestions, which contribute to enriching this book.

We also thank the people of the School of Electronic & Electrical Engineering, Chung-Ang University for giving us an academic environment. Without affections and supports of our families and friends, this book could not be written. Special thanks should be given to Senior Researcher Yong-Suk Park for his invaluable help in correction. We gratefully acknowledge the editorial and production staff of A-Jin Publishing Company for their kind, efficient, and encouraging guide.

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Won Young Yang et al.
The passband $M = 2^b$-ary QAM signaling uses the waveforms which have different amplitudes and phases depending on what data they are carrying and therefore, it can be viewed as a kind of APK (amplitude-phase keying), which combines amplitude modulation and phase modulation. Each of the passband $M = 2^b$-ary QAM signal waveforms can be written as

$$s_m(t) = A_m s_{uc}(t) + A_{ms} s_{us}(t) = \text{Re} \left\{ (A_m + j A_{ms}) \sqrt{\frac{2}{T_s}} e^{j \omega t} \right\} \quad \text{for } m=0, 1, \ldots, M-1$$

$$= A_m \sqrt{\frac{2}{T_s}} \cos(\omega_a t) - A_{ms} \sqrt{\frac{2}{T_s}} \sin(\omega_a t) \quad \text{for } 0 \leq t < T_s$$

$$= A_m \sqrt{\frac{2}{T_s}} \cos(\omega_a t + \theta_m) \quad \text{with } A_m = \sqrt{A_{mc}^2 + A_{ms}^2} \quad \text{and } \theta_m = \tan^{-1} \frac{A_{ms}}{A_{mc}} \quad (7.5.1)$$

where

$$s_{uc}(t) = \sqrt{\frac{2}{T_s}} \cos(\omega_a t), \quad s_{us}(t) = -\sqrt{\frac{2}{T_s}} \sin(\omega_a t) \quad : \text{Basis signal waveforms} \quad (7.5.2)$$

$T_b$: Bit time or bit duration, $T_s = bT_b$: Symbol time or symbol duration

$E_s = bE_b$: Signal energy per symbol
The QAM signal waveforms are illustrated in Fig. 7.1(a4) and each of them can be represented as a vector of length $A_m$

$$s_m = [A_{mc} A_{ms}] = A_m [\cos \theta_m \; \sin \theta_m] \quad (7.5.3)$$

and depicted in the signal space as Fig. 7.1(b4) or Fig. 7.11 where the (orthonormal) bases of the signal space are the unit vectors representing $s_m(t)$ and $s_{m}(t)$ defined by Eq. (7.5.2).

Suppose the amplitudes/phases of the $M = 2^b$-ary QAM signal waveforms are designed in such a way that they can be represented by a rectangular constellation in the signal space as Fig. 7.11(a) and the minimum distance among the signal points is $2A$. Then, the average $E_{s,av}$ of signal powers (represented by the squared distance between signal points and the origin) and the average number $N_b$ of adjacent signal points for a signal point vary with the modulation order $M = 2^b$ or the number $b$ of bits per symbol as

$$M = 2^2 = 4: \quad E_{s,av} = \frac{1}{M} \sum_{m=1}^{M-1} A_m^2 = \frac{4A^2 \times (1^2) \times 2}{4} = 2A^2 = \frac{2(M-1)}{3} \; A^2, \quad N_b = 2 = 4 - \frac{4}{\sqrt{M=2}}$$

$$M = 2^4 = 16: \quad E_{s,av} = \frac{4\times(2-1)^2 \times 4 + (4-2) \times 3 + 2)}{4^2} = 3 = 4 - \frac{4}{\sqrt{M=4}}$$

$$M = 4 \times 2^4 = 64: \quad E_{s,av} = \frac{4A^2 \times (1^2 + 3^2 + 5^2 + 7^2) \times 8}{M=64} = 42A^2 = \frac{2(M-1)}{3} \; A^2$$

$$N_b = \frac{4 \times ((4-1)^2 \times 4 + (8-2) \times 3 + 2)}{8^2} = 7 = 4 - \frac{4}{\sqrt{M=8}}$$

$$E_{s,av} = \frac{2(M-1)}{3} \; A^2: \text{Average signal energy per symbol} \quad (7.5.4a)$$

$$N_b = 4 - \frac{4}{\sqrt{M}}: \text{Average number of adjacent signal points} \quad (7.5.4b)$$

![Figure 7.11 QAM signal constellation diagrams](image)
7.5 Quadrature Amplitude Modulation (QAM)

A half of the minimum distance $d_{\text{min}}$ among the $M=2^b$-ary QAM signal points in the signal space can be expressed in terms of the average signal energy $E_{b,av}$ per bit as

$$\frac{d_{\text{min}}}{2} = A^{(7.5.4a)} = \sqrt{\frac{3/2}{M-1} E_{b,av}} = \sqrt{\frac{3/2}{M-1} b E_{b,av}}$$

(7.5.5)

Note that if we use the circular constellation as in Fig. 7.11(b), we may have a larger minimum distance with the same average energy $E_{b,av}$, but the difference is very small for $M \geq 16$. Besides, an $M=2^b (b=2m: \text{an even number})$-ary QAM signaling with rectangular signal constellation can easily be implemented by two independent $2^{b/2}$-ary PAM signaling, each of which uses one of the quadrature carriers $\cos(\omega t)$ and $\sin(\omega t)$, respectively (see Fig. 7.12). This is why QAM signaling with rectangular signal constellation is widely used.

For an $M=2^b = LN$-ary QAM signaling implemented by combining an $L$-ary PAM signaling and an $N$-ary PAM signaling, the symbol error probability can be found as

$$P_{e,s} (M=LN) = 1 - P(\text{probability of correct detection}) = 1 - (1 - P_{e,s} (L))(1 - P_{e,s} (N))$$

$$= 1 - \left[ 1 - \frac{2(L-1)}{L} Q \left( \sqrt{\frac{3b/2}{L^2 - 1} \text{SNR}_{r,b}} \right) \right] \left[ 1 - \frac{2(N-1)}{N} Q \left( \sqrt{\frac{3b/2}{N^2 - 1} \text{SNR}_{r,b}} \right) \right]$$

$$\leq 4(L-1) \frac{1}{L} Q \left( \sqrt{\frac{3b/2}{M-1} \text{SNR}_{r,b}} \right) \text{ with } L \geq N$$

(7.5.6)

(7.5.7)

where the upperbound on the RHS coincides with what is obtained by substituting Eqs. (7.5.4b) and (7.5.5) into Eq. (5.2.41). How about the bit error probability? Under the assumption that the information symbols are Gray-coded so that the codes for adjacent signal points differ in only one bit, the most frequent symbol errors contain just one of the $b$ bits mistaken and the relationship between the symbol and bit errors can be written as

$$P_{e,b} = \frac{1}{b} P_{e,s}$$

(7.5.8)

Now, let us think about the structure of the $M=2^b = 2^m (b=2m: \text{an even number})$-ary QAM communication system depicted in Fig. 7.12 where the XMTR divides the $b$ bits of a message symbol data into two parts of $m=b/2$ bits, converts them to analog signals, and modulates them with the quadrature carriers that are the basis signal waveforms

$$s_{uc}(t) = \sqrt{\frac{2}{T_s}} \cos(\omega t) \text{ and } s_{us}(t) = -\sqrt{\frac{2}{T_s}} \sin(\omega t)$$

(7.5.10)

respectively. This QAM scheme is basically equivalent to performing two independent quadrature PAMs in parallel. The RCVR has two quadrature correlators, each of which computes a correlation of the received signal $r(t)$ with $s_{uc}(t)$ and $s_{us}(t)$ to make the sampled outputs

$$y_{c,k}^{(5.1.27)} = \int_0^T s_{uc}(t)r(t+(k-1)T_s) \, dt \quad \text{and} \quad y_{s,k}^{(5.1.27)} = \int_0^T s_{us}(t)r(t+(k-1)T_s) \, dt$$

(7.5.11)
respectively. The DTR judges the received signal to be the one represented by the signal point which is the closest to the point \((y_{c,k}, y_{s,k})\) in the signal space as

\[
m_s = \text{Arg Min}_m \| [y_{c,k}, y_{s,k}] - [A_m c, A_m s] \| = \text{Arg Min}_m \{ (y_{c,k} - A_m c)^2 + (y_{s,k} - A_m s)^2 \}
\]  

(7.5.12)

or combines two independent quadrature PAM demodulation results

\[
i_s = \text{Arg Min}_{i=1,...,M} | y_{c,k} - (2i - \sqrt{M} - 1)A_c | \quad \text{(7.5.13a)}
\]

\[
l_s = \text{Arg Min}_{i=1,...,M} | y_{s,k} - (2i - \sqrt{M} - 1)A_s | \quad \text{(7.5.13b)}
\]

to judge the received signal to be the one represented by the \((i_s, l_s)\)th signal point from the left-lower corner in the signal space.

The objective of the following MATLAB program “sim_QAM_passband.m” is to simulate the passband \(M = 2^b = 2^m \) -ary QAM signaling depicted in Fig. 7.12 and plot the bit error probability vs. \(\text{SNR}_{dB} = 10 \log_{10}(E_b/(N_0/2))\) for checking the validity of theoretical derivation results (7.5.8).
%sim_QAM_passband.m
% simulates a digital communication system in Fig.7.13
% with QAM signal waveforms in Fig.7.11
%Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only

clear, clf
b=4; M=2^b; L=2^(b/2); % # of bits per symbol and the modulation order
SNRbdBt=0:0.1:15;   SNRbt=10.^(SNRbdBt/10);
Pm=2*(1-1/L)*Q(sqrt(3/2*b*SNRbt/(M-1))); % Eq.(7.1.5)
pobet= (1-(1-Pm).^2)/b; % Eq.(7.5.8) with (7.5.6)
Tb=1; Ts=b*Tb; % Bit/Symbol time
Nb=16; Ns=b*Nb; % # of sample times in Tb and Ts
T=Ts/Nb; LB=16; LBN1=LB-Ns+1; % Sample time and Buffer size
ssc=[0 0; 0 1; 1 1; 1 0]; sss=ssc;
w=8*pi/Ts; wcT=w*T; t=[0:Ns-1]*T;
su=sqrt(2/Ts)*[cos(wc*t); -sin(wc*t)]; suT=su*T; % Basis signals
Esum= 0;
% 16-QAM signal waveforms corresponding to rectangular constellation
for i=1:L
    for l=1:L
        s(i,l,1)=2*i-L-1; s(i,l,2)=2*l-L-1; %In-phase/quadrature amplitude
        Esum= Esum +s(i,l,1)^2 +s(i,l,2)^2;
        ss(L*(l-1)+i,:)=ssc(i,:); sss(l,:); sw(L*(l-1)+i,:)=s(i,l,1)*su(1,:)+s(i,l,2)*su(2,:);
        if nobe>100; break; end
    end
end
Eav=Esum/M, Es_av=2*(M-1)/3 % Eq.(7.5.4a): Average signal energy (A=1)

for iter=1:length(SNRdBs)
    SNRbdB= SNRdBs(iter);  SNR=10^(SNRbdB/10);
    sigma2=(Es/b)/SNR;  sgmsT=sqrt(sigma2/T);
    yr= zeros(2,LB);  nobe= 0; % Number of bit errors to be accumulated
    for k=1:MaxIter
        im= ceil(rand*L); in= ceil(rand*L);
        s=ss(im,:); % Data bits to transmit
        for n=1:Ns % Operation per symbol time
            wct= wcT*(n-1);  bp_noise= randn*cos(wct)-randn*sin(wct);
            rn= sw(in,n) + sgmsT*bp_noise;
            yr(:,LBN1-LB+1)=yr(:,LBN1-LB+1)+s(i,l,1)*su(1,:)+s(i,l,2)*su(2,:);
        end
    end
end
subplot(222), semilogy(SNRbdBt, pobet, 'k-', SNRdBs, pobe, 'b*')
title('Probability of Bit Error for 16-ary QAM Signaling')
Chapter 7  Passband Digital Communication

Fig. 7.14 shows the BER (bit error rate) curves, i.e. the bit error probabilities versus the average SNR per bit for $M = 2^4 = 16$, $2^6$, $2^8$-ary QAM signalings where the error probability tends to increase as the modulation order $M$ increases. This tendency can be anticipated from the signal constellation diagram where the signal points get denser in the two-dimensional space and consequently, the minimum distance among the signal points gets shorter as $M$ increases.

7.6 COMPARISON OF VARIOUS SIGNALINGS

There are several criteria to consider in deciding the signaling/modulation methods. For example,

◊ BER (bit error rate) performance: How low is the bit error probability for the same SNR?
◊ Data rate: How high is the data transmission rate[bits/sec]?
◊ Power efficiency: How low is the SNR required for keeping the same BER?
◊ Bandwidth efficiency: How narrow is the bandwidth required to keep the same data transmission rate?
◊ PAR (peak-to-average power ratio), interference, and out-of-band radiation
◊ Structural simplicity and cost: How simple is the structure and how cheap is the cost for construction and maintenance?

Table 7.1 shows the BERs for various signalings that have been discussed so far. The following MATLAB routine “prob_error(SNRbdB,signaling,b)” computes the error probability for SNRbdB value(s), signaling method, and number of bits per symbol. The MATLAB built-in function ‘berawgn(EbN0dB,signaling,M)’ is more powerful and convenient to use. The BERTool GUI (graphic user interface) can be invoked by typing ‘bertool’ into the MATLAB Command Window.
### Table 7.1 Bit error probabilities for various signalings \( (SNR_{r,b} = \frac{E_r}{\sigma^2} = \frac{(E_s/b)}{(N_0/2)}) \)

<table>
<thead>
<tr>
<th>( M = 2^b )</th>
<th>Signaling</th>
<th>Coherent (synchronous) detection</th>
<th>Noncoherent detection</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Binary case</strong></td>
<td>OOK</td>
<td>( Q \left( \sqrt{\frac{E_r}{N_0/2}} \right) ) (7.1.24)</td>
<td>( \frac{1}{2} e^{-SNR_{r,b}/4} ) (7.1.22)</td>
</tr>
<tr>
<td>( M = 2^b )</td>
<td>FSK</td>
<td>( Q \left( \sqrt{\frac{E_r}{N_0/2}} \right) ) (7.2.9)</td>
<td>( \frac{1}{2} e^{-SNR_{r,b}/4} ) (7.2.20)</td>
</tr>
<tr>
<td>( (b = 1) )</td>
<td>PSK</td>
<td>( Q \left( \sqrt{SNR_r} \right) ) (7.3.5)</td>
<td>DPSK: ( \frac{1}{2} e^{-SNR_{r,b}/2} ) (7.4.9)</td>
</tr>
<tr>
<td>( M = 2^b )</td>
<td>ASK</td>
<td>( \frac{2(M-1)}{bM} Q \left( \sqrt{\frac{3bSNR_{r,b}}{M^2-1}} \right) ) (7.1.5/6)</td>
<td>-</td>
</tr>
<tr>
<td>( (b &gt; 1) )</td>
<td>FSK</td>
<td>( \frac{M}{2M-1} \left[ 1 - \frac{1}{\sqrt{\pi}} \int_{-\infty}^{\infty} q(y) e^{-y^2} dy \right] ) (7.2.8/9) with ( q(y) = Q^{M-1} \left( -\sqrt{2}y - \sqrt{b SNR_{r,b}} \right) )</td>
<td>( \frac{M}{2M-1} \sum_{m=0}^{M-1} (-1)^{m+1} \left( \frac{M-1}{m+1} \right) e^{-m b SNR_{r,b}/2(m+1)} ) (7.2.19)</td>
</tr>
<tr>
<td>PSK</td>
<td>( \frac{2L}{bL} Q \left( \sqrt{SNR_{r,b}} \sin \frac{\pi}{M} \right) ) (7.3.8)</td>
<td>( \frac{2L}{bL} Q \left( \sqrt{b SNR_{r,b}/2 \sin \frac{\pi}{M}} \right) ) (7.4.8)</td>
<td></td>
</tr>
<tr>
<td>QAM</td>
<td>( \leq \frac{4(L-1)}{bL} Q \left( \sqrt{\frac{3b}{2M-1}SNR_{r,b}} \right) ) (7.5.8) with ( M = LN(L &gt; N) )</td>
<td>-</td>
<td></td>
</tr>
</tbody>
</table>

Now, let us look over the **bandwidth efficiency**, which is defined to be the ratio of the data bit (transmission) rate \( R_b \) [bits/s] over the required system bandwidth \( B \) [Hz], for \( M = 2^b \) -ary ASK/PSK/QAM/FSK signalings. Note the following:

- The rectangular pulse of duration \( D \) [s] in Fig. 1.1(a1) can be viewed as a signal carrying the data with symbol rate \( 1/D \) [symbol/s] and the null-to-null bandwidth of the rectangular pulse is \( 4\pi/D \) [rad/s] = \( 2/D \) [Hz] as can be seen from the spectrum in Fig. 1.1(a1). Thus it may be conjectured that, if a series of data with symbol rate \( R_s \) [symbol/s] is modulated with a carrier frequency \( \omega_c \) and transmitted, the bandwidth is \( 2R_s \) [Hz].

- The interval between adjacent discrete spectra in Fig. 1.1(a1) is \( \pi/D \) [rad/s] = \( 1/2D \) [Hz], which implies that the gap between the carrier frequencies in FSK signaling is \( R_s/2 \) [Hz].

Therefore we can write the bandwidths and bandwidth efficiencies of ASK/PSK/QAM and (coherent) FSK signalings as

\[
B_{ASK/PSK/QAM} = 2R_s = \frac{2}{b} R_b = \frac{2R_b}{b = \log_2 M}; \quad \frac{R_b}{b = \log_2 M} = \frac{b = \log_2 M}{2} \quad (7.6.1)
\]

\[
B_{coh,FSK} = 2R_s + (M-1) \frac{R_s}{2} = \frac{(M+3)R_s}{2 \log_2 M}; \quad \frac{R_s}{B_{coh,FSK}} = \frac{2 \log_2 M}{M+3} \quad (7.6.2)
\]
function p=prob_error(SNRbdB,signaling,b,opt1,opt2)
% Finds the symbol/bit error probability for given SNRbdB (Table 7.1)
% Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
if nargin<5, opt2='coherent'; end % opt2='coherent' or 'noncoherent'
if nargin<4, opt1='SER'; end % opt1='SER' or 'BER'
M=2^b; SNRb=10.^(SNRbdB/10); NSNR=length(SNRb);
if signaling(1:3)=='ASK' % ASK (PAM)
    if lower(opt2(1))=='c' % ASK coherent --> Eq.(7.1.5)
        for i=1:NSNR, p(i)=2*(M-1)/M*Q(sqrt(3*b*SNRb(i)/(M^2-1))); end
        if lower(opt1(1))=='b', p = p/b; end
    else % ASK noncoherent --> Eq.(7.1.22)
        if b==1, for i=1:NSNR, p(i)=exp(-SNRb(i)/4)/2; end; end
    end
elseif signaling(1:3)=='FSK'
    tmpM=2/M/(M-1);
    f5251_=@inline('Q(-sqrt(2)*x-sqrt(b*SNRb)).^(2^b-1)','x','SNRb','b');
    if lower(opt2(1))=='c' % FSK coherent
        if b==1
            for i=1:NSNR, p(i)=Q(sqrt(SNRb(i)/2)); end %Eq.(7.2.9)
        else
            for i=1:NSNR
                p(i) = 1-Gauss_Hermite(f5251_,10,SNRb(i),b)/sqrt(pi);
            end
        end
    else % FSK noncoherent
        for i=1:NSNR
            p(i)=(M-1)/2*exp(-b*SNRb(i)/4); tmp1=M-1;
            for m=2:M-1
                tmp1=-tmp1*(M-m)/m;
                p(i)=p(i)+tmp1/(m+1)*exp(-m*b*SNRb(i)/2/(m+1)); % Eq.(7.2.19)
            end
        end
    end
    if lower(opt1(1))=='b'&b>1, p = p*tmp; end
elseif signaling(1:3)=='PSK' % Eq.(7.3.7)
    for i=1:NSNR, p(i)=(1+(b>1))*Q(sqrt(b*SNRb(i))*sin(pi/M)); end
    if lower(opt1(1))=='b'&b>1, p = p/b; end
elseif signaling(1:3)=='DPS' % DPSK
    if b==1 % Eq.(7.4.8)
        for i=1:NSNR, p(i)=2*Q(sqrt(b*SNRb(i)/2)*sin(pi/M)); end
    else
        for i=1:NSNR, p(i)=exp(-SNRb(i)/2)/2; end %Eq.(7.4.9)
        if lower(opt1(1))=='b', p = p/b; end
    end
elseif signaling(1:3)=='QAM'
    L=2^(ceil(b/2)); N = M/L;
    for i=1:NSNR
        tmpL = 1-2*(L-1)/L*Q(sqrt(3*b/2/(L^2-1)*SNRb(i)));
        tmpN = 1-2*(N-1)/N*Q(sqrt(3*b/2/(N^2-1)*SNRb(i)));
        p(i) = 1-tmpL*tmpN; % Eq.(7.5.6)
    end
    if lower(opt1(1))=='b'&b>1, p = p/b; end
end
### Table 7.2 Power efficiency and bandwidth efficiency with signaling

<table>
<thead>
<tr>
<th>The number of bits per symbol</th>
<th>&lt;Power efficiency&gt; ( SNR_{dB_{r,b}} = 10 \log_{10}(E_b/(N_0/2)) ) [dB]</th>
<th>&lt;Bandwidth efficiency&gt; ( R_b / B ) [bits/Hz]</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>ASK</td>
<td>PSK</td>
</tr>
<tr>
<td>( b = 1 )</td>
<td>12.6</td>
<td>12.6</td>
</tr>
<tr>
<td>( b = 2 )</td>
<td>16.8</td>
<td>12.9</td>
</tr>
<tr>
<td>( b = 3 )</td>
<td>21.3</td>
<td>16.5</td>
</tr>
<tr>
<td>( b = 4 )</td>
<td>26.1</td>
<td>21.1</td>
</tr>
<tr>
<td>( b = 5 )</td>
<td>31.2</td>
<td>26.1</td>
</tr>
<tr>
<td>( b = 6 )</td>
<td>36.5</td>
<td>31.3</td>
</tr>
<tr>
<td>( b = 7 )</td>
<td>41.8</td>
<td>36.7</td>
</tr>
<tr>
<td>( b = 8 )</td>
<td>47.3</td>
<td>42.1</td>
</tr>
</tbody>
</table>

%dc07t02.m
% fills in Table 7.2 with SNRdB for BER=1e-5 and bandwidth efficiency

clear, clf
tol=1e-14; bs=[1:8];
% Define a nonlinear equation to be solved for SNRdB using fsolve()
nonlinear_eq=inline('prob_error(x,signaling,b)-1e-5','x','signaling','b');
% For PSK signaling
for i=1:length(bs)
    b=bs(i); RB_PSK(i)=b/2; % Bandwidth efficiency
    if i==1, x0=10; else x0=SNRbdBs_PSK(i-1); end
    SNRbdBs_PSK(i)=fsolve(nonlinear_eq,x0,optimset('TolFun',tol),'PSK',b);
end
disp('PSK'), [bs; SNRbdBs_PSK; RB_PSK]

% For FSK signaling
for i=1:length(bs)
    b=bs(i); RB_FSK(i)=2*b/(2^b+3);
    if i==1, x0=10; else x0=SNRbdBs_FSK(i-1); end
    SNRbdBs_FSK(i)=fsolve(nonlinear_eq,x0,optimset('TolFun',tol),'FSK',b);
end
disp('FSK'), [bs; SNRbdBs_FSK; RB_FSK]

% For QAM signaling
bs_QAM=[2:2:8];
for i=1:length(bs_QAM)
    b=bs_QAM(i); RB_QAM(i)=b/2;
    if i==1, x0=10; else x0=SNRbdBs_QAM(i-1); end
    SNRbdBs_QAM(i)=fsolve(nonlinear_eq,x0,optimset('TolFun',tol),'QAM',b);
end
disp('QAM'), [bs_QAM; SNRbdBs_QAM; RB_QAM]

% For ASK signaling
for i=1:length(bs)
    b=bs(i); RB_ASK(i)=b/2;
    if i==1, x0=10; else x0=SNRbdBs_ASK(i-1); end
    SNRbdBs_ASK(i)=fsolve(nonlinear_eq,x0,optimset('TolFun',tol),'ASK',b);
end
disp('ASK'), [bs; SNRbdBs_ASK; RB_ASK]
Table 7.2 shows the power efficiency, i.e. the SNR[dB] required to achieve the BER of $10^{-5}$ and the bandwidth efficiency, i.e. the ratio of data transmission rate[bits/s] over bandwidth[Hz], that are also depicted in Fig. 9.7. As the modulation order $M$ or the number $b$ of bits per symbol increases, FSK tends to have higher power efficiency and lower bandwidth efficiency while PSK tends to have lower power efficiency and higher bandwidth efficiency, as can also be seen from Figs. 7.5 and 7.10. The above MATLAB program “dc07t02.m” can be used to compute the values of $SNR_{dB}$, at which PSK/FSK/QAM/ASK signalings achieve the BER of $10^{-5}$.

Before ending this section, it would be well worth to see the usage of the BER computing function ‘berawgn()’ in the MATLAB Help manual:

```matlab
ber = berawgn(EbNo,'PAM',M)
Ber = berawgn(EbNo,'QAM',M)
Ber = berawgn(EbNo,'PSK',M,dataenc) with dataenc='diff'/'nondiff' for differential/non-differential
ber = berawgn(EbNo,'OQPSK',dataenc) for OQPSK (offset QPSK) (see Problem 7.4)
ber = berawgn(EbNo,'DPSK',M)
ber = berawgn(EbNo,'FSK',M,coh) with coh='coherent'/'noncoherent' for coherent/noncoherent
ber = berawgn(EbNo,'FSK',2,coh,rho) with rho=the complex correlation coefficient
ber = berawgn(EbNo,'MSK',dataenc) with dataenc='diff'/'nondiff' for differential/non-differential
ber = berawgn(EbNo,'MSK',dataenc,coherence)
berlb = berawgn(EbNo,'CPFSK',M,modindex,kmin) % gives the BER lowerbound
```
Problems

7.1 Linear Combination of Gaussian Noises

Suppose we have a zero-mean white Gaussian noise \( n(t) \) of mean and variance as

\[
E\{n(t)\} = 0 \quad \text{(P7.1.1)}
\]

\[
\text{cov}\{n(\tau), n(t)\} = E\{n(\tau)n(t)\} = \sigma^2 \delta(\tau - t) \quad \text{(P7.1.2)}
\]

Also, consider another noise

\[
n_1(t) = \int_0^\tau n(\tau + t)s_u(\tau) \, d\tau
\]

which is obtained from taking the cross-correlation of a zero-mean white noise \( n(t) \) with a unit energy signal \( s_u(t) \) such that

\[
\int_0^\tau s_u^2(\tau) \, d\tau = 1 \quad \text{(P7.1.4)}
\]

It is claimed that this is also a zero-mean Gaussian noise with variance \( \sigma^2 \).

(a) To verify this, check the following derivation of the mean and covariance of \( n_1(t) \). Note that a linear combination of Gaussian processes is also Gaussian.

```matlab
% See if a linear convolution of Gaussian noises with a unit signal waveform is another Gaussian noise clear, clf K=10000; % # of iterations for getting the error probability Ts=1; N=40; T=Ts/N; % Symbol time and Sample time N4=N*4; % Buffer size of correlator wc=10*pi/Ts; t=[0:N-1]*T; wct=wc*t; su=sqrt(2/Ts)*cos(wct); % Unit energy signal signal_power=su*su'*T % Signal energy sigma2=2; sigma=sqrt(sigma2); sqT=sqrt(T); noise= zeros(1,N4); % Noise buffer
for k=1:K
    for n=1:N % Operation per symbol time
        noise0= sigma*randn;
        noise=[noise(2:N4) noise0/sqT]; % Bandpass noise
        noise1= su*noise(3*N+1:N4)';
    end
    noise0s(k)=noise0; noise1s(k)=noise1;
end
phi0=xcorr1(noise0s); phi1=xcorr1(noise1s);
plot(phi0), hold on, pause, plot(phi1,'r')
```
Chapter 7 Passband Digital Communication

(proof)

\[ E\{n_1(t)\} = E\left\{ \int_0^{T_t} n(\tau+t)s_n(\tau) \, d\tau \right\} = \int_0^{T_t} E\{n(\tau+t)\}s_n(\tau) \, d\tau = 0 \quad \text{(P7.1.1)} \]  

\[ E\{n_1^2(t)\} = E\left\{ \int_0^{T_t} n(\tau+t)s_n(\tau) \, d\tau \int_0^{T_t} n(\tau+v)s_n(\tau) \, d\tau \right\} = \int_0^{T_t} s_n^2(\tau) \, d\tau \quad \text{(P7.1.2)} \]

\[ = \sigma^2 \int_0^{T_t} s_n^2(\tau) \, d\tau \quad \text{and} \quad \text{(E1.6.1)} \]

\[ = \sigma^2 \int_0^{T_t} s_n^2(\tau) \, d\tau \quad \text{(P7.1.4)} \]

(b) As an alternative for verification, the above program “dc07p01.m” is composed to generate a noise \( n(t) \) with Gaussian distribution \( N(m=0, \sigma^2=2) \) and take the sampled crosscorrelation between \( n(t) \) and a unit energy signal \( s_n(t) \) to make another noise \( n_1(t) \). Plot the autocorrelation of these two noises (\( \text{noises0} \) and \( \text{noisel} \)) and make a comment on their closeness or similarity of their statistical properties such as the mean and variance.

7.2 Coherent/Noncoherent Detection with Time Difference between XMTR and RCVR

(a) In order to feel how seriously the time difference between XMTR and RCVR affects the BER performance of a communication system, use the following statements to modify the MATLAB program “sim_FSK_passband_coherent.m” so that it can accommodate some delay time or time difference \( (\tau_d) \) between XMTR and RCVR clocks. Then run the modified program to see the changed BER curve. Does it stay away from the theoretical BER curve?

(b) In order to see that noncoherent detection is of relative advantage over coherent detection in facing the time difference between XMTR and RCVR, replace \( \text{nd}=0 \) by \( \text{nd}=2 \) in the 14th line of the program “sim_FSK_passband_noncoherent.m” (Sec. 7.2) and run the modified program to see the changed BER curve. Does it still touch the theoretical BER curve?

(c) Modify the 19th line of the QPSK simulation program “sim_PSK_passband.m” (Sec. 7.3) into \( \text{nd}=1 \) and run the program to see the changed constellation diagram and BER curve. Modify the 16th line of the QDPSK simulation program “sim_DPSK_passband.m” (Sec. 7.4) into \( \text{nd}=1 \) and run the program to see the BER curve. What is implied by the simulation results?

7.3 \( M=2^2 \)-ary QAM (Quadrature AM) and \( \pi/4 \)-QPSK (Quadrature PSK)

From the QAM signal constellation diagrams in Fig. 7.11, it can be seen that \( M=2^2 \)-ary QAM is the same as \( \pi/4 \)-QPSK, whose signal constellation diagram is a rotation of the QPSK signal constellation diagram (Fig. 7.1(b3)) by \( \pi/4=45^\circ \) as depicted in Fig. P7.5(a2). Is it also supported by the conformity of the error probabilities (between the standard QPSK and the \( \pi/4 \)-QPSK) that turned out to be Eq. (7.3.7) and Eq. (7.5.7), respectively? You can substitute \( b=2, L=2^b=2 \), and \( M=2^2 \) into both of the equations and check if they conform with each other.
7.4 OQPSK (Offset Quadrature PSK) or SQPSK (Staggered QPSK)

As depicted in Fig. P7.4, OQPSK is a slight modification of QPSK (Fig. 7.12) in such a way that the Q(quad)-channel bit stream is offset w.r.t. the I(n)-channel by a bit duration (Fig. P7.4(b2)). Compared with QPSK, it allows no simultaneous change of two bits to prevent any state transition accompanying the phase change of 180° (Fig. P7.4(c2)) (in the teeth of making more frequent state transitions possibly every bit time) and consequently, the envelope variation becomes less severe so that the transmitted signal bandwidth can be limited more strictly.
% sim_OQPSK.m
% simulates a digital communication system
% with O(ffset)QPSK signal waveforms in Fig.P7.4
% Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only

clear, clf;
b=2; M=2^b;
SNRbdBt=0:0.1:10; SNRbt=10.^((SNRbdBt/10));
pobet=prob_error(SNRbdBt,'PSK',b,'bit');
Tb=1; Ts=b*Tb; % Bit/symbol time
Nb=16; Ns=Nb*b; % # of sample times in Tb and Ts
T=Ts/Ns; LB=4*Ns; LBN1=LB-Ns+1; % Sample time and Buffer size
Es=2; sqEb=sqrt(Es/b); % Energy of signal waveform
% QPSK signal waveforms
ss=[0 0; 0 1; 1 1; 1 0];
wct=2*pi/Ts; wcT=wct*T; t=[0:Ns-1]*T;
su=sqrt(2/Ts)*[cos(wc*t); ???????????]; suT=su*T;
sw=sqEb*su;
SNRdBs=[1:3:10]; MaxIter=10000; % Range of SNRdB, # of iterations
for iter=1:length(SNRdBs)
    SNRbdB=SNRdBs(iter); SNRb=10^(SNRbdB/10);
    N0=2*(Es/b)/SNRb; sigma2=N0/2; sgmsT=sqrt(sigma2/T);
    yr=zeros(2,LB); yc=zeros(1,2); ys=zeros(1,2);
    iq=0; % Initialize the quadrature bit arbitrarily
    nobe=0; % Number of bit errors to be accumulated
    for k=1:MaxIter
        i=ceil(rand*M); s=ss(i,:); wct=-wctT;
        for n=1:b % Operation per symbol time
            if n==1, ii=2*ss(i,1)-1; % In-phase bit
                else iq=2*ss(i,2)-1; % Quadrature bit
            end
            mn=0;
            for m=1:Nb % Operation per bit time
                wct= wct+wctT; mn=mn+1;
                bp_noise= randn*cos(wct)-randn*sin(wct);
                rn=i*sw(1,mn)+iq*sw(2,mn)+sgmsT*bp_noise;
                yr=[yr(:,2:LB) suT(:,mn)*rn]; % Multiplier
            end
            if n==2 % sampled at t=2k*Tb
                yc=[yc(2) sum(yr(:,LBN1:LB))]; % Correlator output
                else % sampled at t=(2k-1)*Tb
                    ys=[ys(2) sum(yr(:,LBN1:LB))]; % Correlator output
                end
        end
        d=(yc(?) ys(?)>0); % Detector(DTR)
        if k>1, nobe=nobe+sum(s0=-d); end
        if nobe>100, break; end
        s0= s;
    end
    pobe(iter)= nobe/(k*b);
end
pobe
semilogy(SNRbdBt,pobet,'k-', SNRdBs,pobe,'b*')
title('Probability of Bit Error for (4-ary) QPSK Signaling')
The above MATLAB program “sim_OQPSK.m” is a modification of the program “sim_PSK_passband.m” (QPSK) (in Sec. 7.3) to simulate the OQPSK communication system, whose block diagram is depicted in Fig. P7.4, but it is unfinished. Finish it up by replacing the three parts of ?’s with appropriate statements or just indices and run it to see the BER curve.

7.5 \( \pi/4 \)-Shifted QPSK (Quadrature PSK)

\( \pi/4 \)-Shifted QPSK is another way of lowering envelope fluctuations than OQPSK discussed in Problem 7.4. It assigns one of the signal points in the (non-offset) QPSK (Fig. P7.5(a1)) and \( \pi/4 \)-QPSK (Fig. P7.5(a2)) signal constellations to even and odd-number indexed data symbols as depicted in Fig. P7.5(b) and therefore, it virtually uses the dual signal constellation diagram shown in Fig. P7.5(a3) where the modulation causes no state transition path through the origin as in OQPSK. Compared with OQPSK, it has the maximum phase change of \( \pm 3\pi/4 \), which is larger than that (\( \pm \pi/2 \)) of OQPSK; On the other hand, differential encoding/decoding with noncoherent detection can be implemented in \( \pi/4 \)-shifted QPSK signaling, since it has four possible phase-shifts of \( \Delta \theta = \pm \pi/4 \) and \( \pm 3\pi/4 \) (as can be seen from Fig. P7.5(b)) while OQPSK has only two possible phase-shifts of \( \Delta \theta = \pm \pi/2 \).

Table P7.5 Message data dibits and the corresponding phase-shifts in \( \pi/4 \)-shifted QDPSK

<table>
<thead>
<tr>
<th>(Gray-coded) message dibits</th>
<th>Phase-shifts</th>
<th>( s_k = s_0, s_1, s_2, s_3, s_4, s_5, s_6, s_7 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>( d_{2k-1}d_{2k} = 00 )</td>
<td>( \Delta \theta = +\pi/4 )</td>
<td>( s_{(k+1)\text{mod}8} = s_1, s_2, s_3, s_4, s_5, s_6, s_7, s_0 )</td>
</tr>
<tr>
<td>( d_{2k-1}d_{2k} = 01 )</td>
<td>( \Delta \theta = +3\pi/4 )</td>
<td>( s_{(k+3)\text{mod}8} = s_1, s_2, s_3, s_4, s_5, s_6, s_7, s_0 )</td>
</tr>
<tr>
<td>( d_{2k-1}d_{2k} = 11 )</td>
<td>( \Delta \theta = -3\pi/4 )</td>
<td>( s_{(k+5)\text{mod}8} = s_1, s_2, s_3, s_4, s_5, s_6, s_7, s_0 )</td>
</tr>
<tr>
<td>( d_{2k-1}d_{2k} = 10 )</td>
<td>( \Delta \theta = -\pi/4 )</td>
<td>( s_{(k+7)\text{mod}8} = s_1, s_2, s_3, s_4, s_5, s_6, s_7, s_0 )</td>
</tr>
</tbody>
</table>
(a) Modify the (standard) QPSK simulation program “sim_PSK_passband.m” into a $\pi/4$-shifted QPSK simulation program named, say, “sim_S_QPSK.m” and run it to see the BER curve. Does it conform with the theoretical BER curve?

(b) Referring to Table P7.5, modify the (standard) QDPSK (quadrature differential phase shift keying) simulation program “sim_DPSK_passband.m” into a $\pi/4$-shifted QDPSK simulation program named, say, “sim_S_QDPSK.m” and run it to see the BER curve. Does it conform with the theoretical one? If you have no idea, start with the following unfinished program:

```matlab
%sim_S_QDPSK.m
% simulates the pi/4-shifted QDPSK signaling (Table P7.5)
% Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
b=2; M=2^b; M2=M*2;
SNRbdBt=0:0.1:10; SNRbt=10.^(SNRbdBt/10);
pobet=prob_error(SNRbdBt,'DPSK',b,'bit');
Tb=1; Ts=b*Tb; % Bit/symbol time
Nb=16; ns=b*Nb; % Numbers of sample times in Tb and Ts
T=Tb/Nb; LB=4*Nb; LBN1=LB-Ns+1; % Sample time and Buffer size
Es=2; % Energy of signal waveform
% QDPSK signal waveforms
ss=[0 0; 0 1; 1 1; 1 0];
wc=8*pi/Ts; wcT=wc*T; t=[0:Ns-1]*T; nd=1;
for m=1:M2, sw(m,:)=sqrt(2*Es/Ts)*cos(wc*t+(m-1)*pi/M); end
su= sqrt(2/Ts)*[cos(wc*t); -sin(wc*t)]; suT=su*T;
SNRdBs=[1:3:10]; MaxIter=10000; %Range of SNRbdB, # of iterations
for iter=1:length(SNRdBs)
    SNRbdB=SNRdBs(iter); SNRb=10^(SNRbdB/10);
    N0=2*(Es/b)/SNRb; sigma2=N0/2; sgmsT=sqrt(sigma2/T);
    sws=zeros(1,LB); yr=zeros(2,LB);
    nobe=0; % Number of bit errors to be accumulated
    is0=1; % Initial signal index
    th0=1; % Initial guess (possibly wrong)
    for k=1:MaxIter
        i= ceil(rand*M); s=ss(i,:); % Data bits to transmit
        is= mod(is0+?????,M2)+1; % Signal to transmit (Table P7.5)
        for n=1:Ns % Operation per symbol time
            sws=[sws(2:LB) sw(is,n)];
            wct=wcT*(n-1); bp_noise= randn*cos(wct)-randn*sin(wct);
            rn= sws(end-nd) + sgmsT*bp_noise;
            yr=[yr(:,2:LB) suT(:,n)*rn]; % Multiplier
        end
        ycsk=sum(yr(:,LBN1:LB)')'; % Sampled correlator output
        %Detector(DTR)
        th=atan2(ycsk(2),ycsk(1)); dth=th-th0;
        if dth<0, dth=dth+2*pi; end
        [themin,lmin]=min(abs(dth-[?????]*2*pi/M));
        d= ss(lmin,:); % Detected data bits
        nobe= nobe+sum(s~=d); if nobe>100, break; end
        is0=is; th0=th; % update the previous signal and theta
    end
    pobe(iter)= nobe/(k*b); %
end
semilogy(SNRbdBt,pobet,'k', SNRdBs,pobe,'b*')
```
7.6 Minimum-Shift Keying (MSK)

Minimum-shift keying is a type of continuous phase frequency shift keying (CP-FSK). It encodes each data bit into a half sinusoid

\[ s(t) = \sqrt{\frac{2E_b}{T_b}} \cos(\omega_c t + \theta(t)) \]  

(P7.6.1)

with its phase changed continuously as

\[ \theta(t) = \theta(kT_b) + \frac{\pi}{2T_b} (t - kT_b) \quad \text{with } \theta(0) = 0 \]  

(P7.6.2)

where the second term has a plus or minus sign for each bit interval depending on whether the value of message data bit is 1 or 0 (see Fig. P7.6). As a consequence, the phase of the signal waveform at the boundaries between two consecutive bit intervals becomes

\[ \theta(2kT_b) = \pm m\pi \quad \text{or} \quad \theta((2k+1)T_b) = \pm \frac{\pi}{2} \pm m\pi \]  

(P7.6.3)

and the instantaneous frequency, which is obtained by differentiating the angular argument of \( s(t) \) w.r.t. \( t \), becomes

\[ \omega_a = \omega_c + \frac{\pi}{2T_b} \quad \text{or} \quad \omega_b = \omega_c - \frac{\pi}{2T_b} \]  

(P7.6.4)

Note that the difference of these two frequencies is \( \pi / T_b \) and it is the same as the minimum frequency gap required to keep the orthogonality between two frequency components for a bit duration \( T_b \) (see the discussion just below Eq. (7.2.4)).

To find two (orthogonal) basis signal waveforms that can be used to detect the phase of the received signal, we rewrite Eq. (P7.6.1) as

\[ s(t) \quad \text{(P7.6.1)} \Rightarrow \sqrt{\frac{2E_b}{T_b}} \left( \cos \theta(t) \cos \omega_c t - \sin \theta(t) \sin \omega_c t \right) = s_I(t) \cos \omega_c t - s_Q(t) \sin \omega_c t \]  

(P7.6.5)

where

\[ s_I(t) = \left\{ \begin{array}{ll} \sqrt{\frac{2E_b}{T_b}} \cos \theta(t) & \quad \text{with } \theta(t) = \theta(2kT_b) + \frac{\pi}{2T_b} (t - 2kT_b) \\ \sqrt{\frac{2E_b}{T_b}} \cos \theta(2kT_b) \cos \left( \frac{\pi}{2T_b}(t - 2kT_b) \right) - \sin \theta(2kT_b) \sin \left( \frac{\pi}{2T_b}(t - 2kT_b) \right) & \quad \text{for } (2k-1)T_b < t \leq (2k+1)T_b \\ \frac{\sqrt{2E_b}}{T_b} \cos \left( \frac{\pi}{2T_b}(2kT_b) \right) & \quad \text{if } \theta(2kT_b) = 0 \\ \frac{\sqrt{2E_b}}{T_b} \cos \left( \frac{\pi}{2T_b}(2kT_b) \right) & \quad \text{if } \theta(2kT_b) = \pm \pi \end{array} \right. \]  

(P7.6.6a)
\[ s_Q(t) = \sqrt{\frac{2E_b}{T_b}} \sin \theta(t) \quad \text{with} \quad \theta(t) = \theta((2k+1)T_b) \pm \frac{\pi}{2T_b} (t - (2k+1)T_b) \]

\[ = \sqrt{\frac{2E_b}{T_b}} \begin{cases} 
\sin \theta((2k+1)T_b) \cos \left( \pm \frac{\pi}{2T_b} (t - (2k+1)T_b) \right) \\
\pm \cos \theta((2k+1)T_b) \sin \left( \pm \frac{\pi}{2T_b} (t - (2k+1)T_b) \right) 
\end{cases} \]

\[ = \sqrt{\frac{2E_b}{T_b}} \sin \theta((2k+1)T_b) \sin \left( \pm \frac{\pi}{2T_b} t \right) \quad \text{for} \quad 2kT_b < t \leq 2(k+1)T_b \]

\[ = \begin{cases} 
+ \sqrt{\frac{2E_b}{T_b}} \sin \left( \frac{\pi}{2T_b t} \right) & \text{if} \quad \theta((2k+1)T_b) = +\frac{\pi}{2} \\
- \sqrt{\frac{2E_b}{T_b}} \sin \left( \frac{\pi}{2T_b t} \right) & \text{if} \quad \theta((2k+1)T_b) = -\frac{\pi}{2}
\end{cases} \quad \text{(P7.6.6b)} \]

Substituting these two equations (P7.6.6a) and (P7.6.6b) into Eq. (P7.6.5) yields

\[ s(t) = \pm \sqrt{\frac{2E_b}{T_b}} \cos \left( \frac{\pi}{2T_b} t \right) \cos \omega_c t \pm \sqrt{\frac{2E_b}{T_b}} \sin \left( \frac{\pi}{2T_b} t \right) \sin \omega_c t \quad \text{(P7.6.7)} \]

This implies that the basis signal waveforms to be used for detection at RCVR are

\[ s_{we}(t) = \sqrt{\frac{2}{T_b}} \cos \left( \frac{\pi}{2T_b} t \right) \cos \omega_c t \quad \text{(P7.6.8a)} \]

\[ s_{wc}(t) = \sqrt{\frac{2}{T_b}} \sin \left( \frac{\pi}{2T_b} t \right) \sin \omega_c t \quad \text{(P7.6.8b)} \]

Therefore, RCVR computes the correlation between the received signal \( r(t) \) and these two basis signal waveforms and samples it at \( t = 2kT_b \) and \( (2k+1)T_b \) to get

\[ y_c((2k+1)T_b) = \int_{T_k}^{T_k + (2k)T_b} s_{we}(t) r(t + 2kT_b) \, dt = \int_{T_k}^{T_k + (2k+1)T_b} s_{wc}(t) s(t + 2kT_b) \, dt + n_c \]

\[ = \sqrt{\frac{2E_b}{T_b}} \int_{T_k}^{T_k + (2k+1)T_b} \cos \left( \frac{\pi}{2T_b} t \right) \cos \omega_c t \cos \omega_c t + \cos \left( \theta(2kT_b) \pm \theta(2T_b) \right) \, dt + n_c \]

\[ = \sqrt{\frac{2E_b}{T_b}} \cos \left( \frac{\pi}{2T_b} t \right) \cos \left( \theta(2kT_b) + \frac{\pi}{2T_b} t \right) + \cos \left( \theta(2kT_b) - \frac{\pi}{2T_b} t \right) \, dt + n_c \]

\[ = \sqrt{E_b} \cos \left( \theta(2kT_b) \right) + n_c \]
Problems 213

\( y_s(2kT_b) \) \( \overset{(5.1.27)}{=} \sqrt{2T_b} \int_0^{2T_b} s_m(t) r(t + (2k - 1)T_b) \, dt \)
\( = \int_0^{2T_b} s_m(t) s(t + (2k - 1)T_b) \, dt + n_s \)
\( \overset{(P7.6.1)}{=} \sqrt{2E_b \over T_b} \left( \sqrt{2 \over T_b} \right) \int_0^{2T_b} \sin \left( \frac{\pi t}{2T_b} \right) \sin \omega_c t \cos(\omega_c t + \theta((2k - 1)T_b)) \, dt + n_s \)
\( \overset{(D.24)}{=} \sqrt{2E_b \over 2T_b} \int_0^{2T_b} \left[ \cos \left( 2\omega_c t + \theta((2k - 1)T_b) - \frac{\pi}{2T_b} t \right) - \cos \left( 2\omega_c t + \theta((2k - 1)T_b) + \frac{\pi}{2T_b} t \right) \right. \]
\[ + \cos \left( \theta((2k - 1)T_b) + \frac{\pi}{2T_b} t \right) - \cos \left( \theta((2k - 1)T_b) - \frac{\pi}{2T_b} t \right) \] \( dt + n_s \)
\( = \sqrt{E_b \over 2T_b} \sin(\theta((2k - 1)T_b)) + n_s \)

Depending on the sign of these sampled values of correlator outputs, the phase estimates are determined as

\[ \theta(2kT_b) = \begin{cases} 
0 & \text{if } y_c((2k+1)T_b) > 0 \\
\pm \pi & \text{close to } \theta((2k-1)T_b) & \text{if } y_c((2k+1)T_b) < 0 \end{cases} \] \( \overset{(P7.6.10a)}{=} \)
\[ \theta((2k-1)T_b) = \begin{cases} 
-\pi / 2 & \text{if } y_s(2kT_b) > 0 \\
+\pi / 2 & \text{if } y_s(2kT_b) < 0 \end{cases} \] \( \overset{(P7.6.10b)}{=} \)

so that the maximum difference between the phase estimates at successive sampling instants is \( \pm \pi / 2 \). Then the detector judges the value of transmitted data bit to be 1 or 0 depending on whether \( \theta \) turns out to have increased or decreased.

Noting that the sampled values of correlator outputs \( y_c((2k + 1)T_b) \) and \( y_s(2kT_b) \) will be distributed around \((+\sqrt{E_b}, 0)\) and \((-\sqrt{E_b}, 0)\) in the signal space and thus the (minimum) distance between signal points is \( d_{\text{min}} = 2\sqrt{E_b} \), the (symbol or bit) error probability is

\[ P_{e,b} \overset{(5.2.35)}{=} Q \left( \frac{d_{\text{min}} / 2}{\sigma = \sqrt{N_0 / 2}} \right) = Q \left( \sqrt{E_b \over N_0} \right) = Q \left( \sqrt{\text{SNR}_{e,b}} \right) \] \( \overset{(P7.6.11)}{=} \)

---

**Figure P7.6** The phase and waveform of an MSK signal
Chapter 7  Passband Digital Communication

% sim_MSK.m
% simulates a digital communication system with MSK signaling
%Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
b=1; M=2^b; % # of bits per symbol and modulation order
SNRbdBt=0:0.1:12; SNRbt=10.^((SNRbdBt/10));
pbe_MSK = prob_error(SNRbdBt, 'PSK', b, 'bit');
pbe_MSK = prob_error(SNRbdBt, 'FSK', b, 'bit');
SNRdBs=[1 3 10]; % Range of SNRbdB
MaxIter=10000; % Number of iterations
for iter=1:length(SNRdBs)
    SNRbdB=SNRdBs(iter); SNR=10.^((SNRbdB/10));
    N0=2*(Es/b)/SNR; sigma2=N0/2; sgmsT=sqrt(sigma2/T);
    yr=zeros(2,LB); yc=0; ys=0; th0=0; t=0; wct=0; tht=th0; mn=1;
    nobe=0; % Number of bit errors to be accumulated
    for k=1:MaxIter
        i=ceil(rand*M); s(k)=ds(i); sgn=s(k)*2-1;
        for m=1:Nb % Operation per bit time
            bp_noise= randn*cos(wct)-randn*sin(wct);
            rn= sq2EbTb*cos(wct+tht) + sgmsT*bp_noise;
            yr=[yr(:,2:LB) suT(:,mn)*rn]; % Correlator output - DTR input
            t=t+T; wct=wct+wcT; tht=tht+sgn*pihTbT; mn=mn+1;
        end
        if tht<-pi2, tht=tht+pi2; elseif tht>pi2, tht=tht-pi2; end
        thtk(k+1)=tht;
        if mn>size(su,2), mn=1; end
        if mod(k,2)==1
            yc=sum(yr(:,LBN2:LB));
            th_hat(k)=??*(yc<0);
            if k>1&th_hat(k-1)<0, th_hat(k)=-th_hat(k); end
        else
            ys=sum(yr(:,LBN2:LB));
            th_hat(k)=????*(2*(ys<0)-1);
        end
        if k>1 % Detector(DTR)
            d=(dth_hat>0); if abs(dth_hat)>=pi, d=-d; end
            nobe = nobe+(d~=s(???)); if nobe>100; break; end
        end
    end
    pobe(iter)= nobe/((k-1)*b);
end
semilogy(SNRbdBt,pbe_MSK,'k', SNRbdBt,pbe_MSK,'k:', SNRdBs,pobe,'b*')
Compared with Eq. (7.2.9), which is the error probability for BFSK signaling, the SNR has increased by two times, which is attributed to the doubled integration period $2T_b$ of the correlators.

(a) Complete the above incomplete program “sim_MSK.m” so that it can simulate the MSK modulation and demodulation process.

(b) Consider the in-phase/quadrature symbol shaping functions

$$g_c(t) = \begin{cases} \sqrt{\frac{2E_b}{T_b}} \cos \left( \frac{\pi}{2T_b} t \right) & \text{for } -T_b \leq t \leq T_b \\ 0 & \text{elsewhere} \end{cases} \quad (P7.6.12a)$$

$$g_s(t) = \begin{cases} \sqrt{\frac{2E_b}{T_b}} \sin \left( \frac{\pi}{2T_b} t \right) & \text{for } 0 \leq t \leq 2T_b \\ 0 & \text{elsewhere} \end{cases} \quad (P7.6.12b)$$

Verify that their CTFTs and ESDs (energy spectral densities) can be obtained as

$$G_c(\omega) = \mathcal{F}\{g_c(t)\} = \frac{2E_b}{T_b} \int_{-T_b}^{T_b} \cos \left( \frac{\pi}{2T_b} t \right) e^{-j\omega t} dt$$

$$= \sqrt{\frac{E_b}{2T_b}} \left[ e^{j\pi/2T_b} + e^{-j\pi/2T_b} \right] e^{-j\omega t} dt$$

$$= \sqrt{\frac{2E_b}{T_b}} \left( \frac{\sin(\pi/2T_b - \omega)T_b}{\pi/2T_b - \omega} + \frac{\sin(\pi/2T_b + \omega)T_b}{\pi/2T_b + \omega} \right)$$

$$= \frac{2E_b}{T_b} \frac{\sin(\pi/2T_b) \cos \omega T_b - 2\omega \cos(\pi/2) \sin \omega T_b}{(\pi/2T_b)^2 - \omega^2}$$

$$\therefore G_s(\omega) = \mathcal{F}\{g_s(t)\} = \sqrt{\frac{2E_b}{T_b}} \int_{-T_b}^{T_b} \sin \left( \frac{\pi}{2T_b} t \right) e^{-j\omega t} dt$$

$$= \sqrt{\frac{2E_b}{T_b}} \int_{-T_b}^{T_b} \sin \left( \frac{\pi}{2T_b} t \right) e^{-j\omega t} dt$$

$$= \sqrt{\frac{2E_b}{T_b}} \frac{\sin(\pi/2T_b - \omega)T_b}{\pi/2T_b - \omega} - \frac{\sin(\pi/2T_b + \omega)T_b}{\pi/2T_b + \omega}$$

$$= \frac{2E_b}{T_b} \frac{\sin(\pi/2T_b) \cos \omega T_b - 2\omega \cos(\pi/2) \sin \omega T_b}{(\pi/2T_b)^2 - \omega^2} \quad (P7.6.14)$$

\[ \Phi_{g_c}(\omega) = \Phi_{g_s}(\omega) = |G_{g_c}(\omega)|^2 = |G_{g_s}(\omega)|^2 = \frac{2\pi^2 E_b}{T_b} \frac{\cos^2 \omega T_b}{((\pi/2T_b)^2 - \omega^2)} \]
7.7 Simulation of FSK Using Simulink

Fig. P7.7.1 shows a Simulink model ("FSK_passband_sim.mdl") to simulate a BFSK (binary frequency shift keying) passband communication system and Fig. P7.7.2 shows the subsystem for the demodulator and detector where the following parameters are to be set in the workspace:

\[
\begin{align*}
    b &= 1; M = 2^b; \quad \text{% Number of bits per symbol and Modulation order} \\
    N_s &= 40; T_s = 1e-5; T = T_s/N_s; \quad \text{% Symbol time and Sample time} \\
    dw &= 2\pi/T_s; \quad \text{% Frequency spacing and Carrier freq[rad/s]} \\
    EbNoDB &= 10; \quad \text{% Eb/N0[db]} \\
    \text{Target_no_of_error} &= 50; \quad \text{% Simulation stopping criterion}
\end{align*}
\]
Problems 217

%do_FSK_sim.m

clear, clf
b=1; M=2^b; % Number of bits per symbol and Modulation order
Nbsps=b*2^3; % # of samples per symbol in baseband
Nos=5; Ns=Nbsps*Nos; % # of samples per symbol in passband
Ts=1e-5; T=Ts/Ns; % Symbol time and Sample time
dw=2*pi/Ts; % Frequency separation [rad/s]
wc=10*dw; % Carrier Frequency [rad/s] (such that wc*T<pi)
Target_no_of_error=50; SNRdBs=[5 10]; EbN0dBs=SNRdBs-3;
for i=1:length(SNRdBs)
    SNRdB=SNRdBs(i); EbN0dB=SNRdB-3; % Eb/(N0/2)=SNR-> Eb/N0=SNR/2
    sim('FSK_passband_sim',1e5*Ts); SERs(i)=ser(end,1);
end
SNRdBt=0:0.1:13;
poset = prob_error(SNRdBt,'FSK',b,'sym','noncoherent');
semilogy(SNRdBt,poset,'k', SNRdBs,SERs,'*'), xlabel('SNR[dB]')

Received_Signal_Power = 1/Ts;

function d=detector_FSK(z2s)
    [z2max,ind]=max(z2s); d=ind-1; % Fig. 7.4.2

Once the Simulink model named "FSK_passband_sim.mdl" is opened, it can be run by clicking on the Run button on the tool bar or by using the MATLAB command ‘sim( )’ as in the above program “do_FSK_sim.m”. You can disconnect the Display Block for signal power measurement to increase the running speed.
Note that the receive delay in the Error Rate Calculation block is set to 1 due to the sample difference between the transmitted data and detected data observed through the Scope.

(a) After having composed and saved the Simulink model “FSK_passband_sim.mdl” (with no delay) and two MATLAB programs “do_FSK_sim.m” and “detector_FSK.m” (that can be substituted by an Embedded MATLAB Function Block), run the MATLAB program “do_FSK_sim.m” to get the SERs (SER: symbol error rate) for a BFSK communication system with SNRdBs=[5 10]. Does the simulation result conform with that in Fig. 7.5 (for b = 1) or 7.6.2?

(b) Change the 7th statement of the MATLAB program “do_FSK_sim.m” into ‘dw=pi/Ts’ or ‘dw=3*pi/Ts’ and run the program to get the SER. How are the SERs compared with those obtained in (a)? What does the difference come from?

(c) Insert the delay of 1~5 samples and discuss the sensitivity of the system w.r.t. such a delay in terms of the SER performance in connection with coherence or noncoherence of the communication system.

(d) Referring to Fig. 7.4.2, extend the Simulink model into “FSK4_passband_sim.mdl” so that it can simulate an $M=2^4=16$-ary FSK communication system and run it to get the SER curve.

(e) To understand how large the power of noise added by the AWGN Block is, connect the Signal power Subsystem together with the Display Block into the input of the AWGN block, click on the Run button, and read the displayed value of the channel input signal power. Then connect the Signal power Subsystem together with the Display Block into the output of the AWGN Block, click on the Run button, and read the displayed value of the channel output signal power that will be the sum of the input signal power and noise power (variance). Do the channel input and output powers measured through the Simulink simulation conform with those obtained using the last two statements in the above program “do_FSK_sim.m”:

\[
\begin{align*}
\text{Transmitted Signal Power} &= 1/T_s; \\
\text{Received Signal Power} &= \ldots \\
&= \text{Transmitted Signal Power} \times (1 + 10^{-(EbN0dBs(\text{end})+3)/10}T_s/b/T)
\end{align*}
\]

Why is 3 added to $EbN0dBs(\text{end})$ in the above statement?

(cf) Visit the webpage <http://www.mathworks.com/access/helpdesk/help/helpdesk.html> and refer to the explanation about the function of AWGN Block that can be found in the user manual for Communications Blockset, according to which the SNR is defined to be

\[
\text{SNRdB} = 10\log_{10} \frac{\text{Signal Power}}{\text{Noise Variance}}
\]

\[
= \begin{cases} 
10\log_{10} \frac{E_s/T_s}{N_0/T} = E_sN0dB + 10\log_{10} \frac{T}{T_s} = EbN0dB + 10\log_{10} b \frac{T_s}{T_s} \\
\quad \text{for a complex-valued signal} \\
10\log_{10} \frac{E_s/(N_0/2)T}{T} = E_sN0dB + 10\log_{10} \frac{2T}{T_s} = EbN0dB + 10\log_{10} b \frac{T_s}{T_s} + 3 \\
\quad \text{for a real-valued signal}
\end{cases}
\]  
(P7.7.1)

\[
\begin{align*}
\text{Signal Power} &\quad \text{for a complex-valued signal} \\
\text{Noise Variance} &\quad (T_s/bT)10^{-EbN0dB/10} \\
\text{Signal Power} &\quad (T_s/bT)10^{-EbN0dB+3/10} \\
\text{Noise Variance} &\quad (T_s/bT)10^{-EbN0dB+3/10} \\
\text{for a real-valued signal}
\end{align*}
\]  
(P7.7.2)
7.8 Simulation of PSK Using Simulink

Fig. P7.8.1 shows a Simulink model to simulate a PSK (phase shift keying) passband communication system where the following parameters are to be set in the workspace:

\[
\begin{align*}
    b &= 1; \quad M = 2^b; \quad \% \text{Number of bits per symbol and Modulation order} \\
    N_s &= 40; \quad T_s = 1e-5; \quad T = T_s / N_s; \quad \% \text{Symbol time and Sample time} \\
    w_c &= 2\pi * 10 / T_s; \quad \% \text{Carrier freq [rad/s]} \\
    E_b / N_0 [\text{dB}] &= 7; \quad \% \text{Eb/N0 [dB]} \\
    \text{Target no_of_error} &= 20; \quad \% \text{Simulation stopping criterion}
\end{align*}
\]

![Simulink model](image-url)
%do_PSK_sim.m

clear, clf
b=1; M=2^b; % Number of bits per symbol and Modulation order
Nbsp=b*2^3; % # of samples per symbol in baseband
Nos=5; Ns=Nbsp*Nos; % # of samples per symbol in passband
Ts=1e-5; T=Ts/Ns; % Symbol time and Sample time
wc=2*pi*10/Ts; % Carrier Frequency [rad/s] (such that wc*T<pi)
Target_no_of_error=20;
SNRdBs=[5 10]; EbN0dBs=SNRdBs-3;
for i=1:length(SNRdBs)
    SNRdB=SNRdBs(i); EbN0dB=SNRdB-3; % Eb/(N0/2)=SNR-> Eb/N0=SNR/2
    sim('PSK_passband_sim',1e5*Ts);
    SERs(i) = ser(end,1);
end
SNRdBt=0:0.1:13;
poset = prob_error(SNRdBt,'PSK',b,'sym');
semilogy(SNRdBt,poset,'k', SNRdBs,SERs,'*'), xlabel('SNR [dB]')
Transmitted_Signal_Power = 1/Ts;
Received_Signal_Power = ...
    Transmitted_Signal_Power*(1+10^(-(EbN0dBs(end)+3)/10)*Ts/b/T)

function d=detector_PSK(th,M)
%th=atan2(ycsk(2),ycsk(1));
if th<-pi/M, th=th+2*pi; end
[thmin,ind]=min(abs(th-2*pi/M*[0:M-1])); % Eq.(7.3.9) or Fig. 7.7
D=d=ind-1;

(a) After having composed and saved the Simulink model “PSK_passband_sim.mdl” and two MATLAB programs “do_PSK_sim.m” and “detector_PSK.m”, run the MATLAB program “do_PSK_sim.m” to get the SERs for a BPSK communication system with SNRdBs=[5 10]. Does the simulation result conform with that in Fig. 7.10 (for b=1)?
(b) Modify the MATLAB program “do_PSK_sim.m” (possibly together with) Simulink model “PSK_passband_sim.mdl” to simulate a 224 M==4-ary PSK (QPSK) communication system and run it to get the SER curve.
(c) Rename the Simulink model “PSK_passband_sim.mdl” “DPSK_passband_sim.mdl” and modify it to simulate an M=2^2=4-ary DPSK (QDPSK) communication system. You can refer to the MATLAB program “sim_DPSK_passband.m” and Fig. P7.8.2. Run it to get the SERs with SNRdBs=[5 10].

![Diagram](image)

(a) Extension of the modulator block  
(b) Extension of the detector block

Figure P7.8.2  Modification of the modulator and detector blocks for DPSK
7.9 Simulation of QAM Using Simulink

Fig. P7.9 shows a Simulink model to simulate a QAM (quadrature amplitude modulation) passband communication system where the following parameters are to be set in the workspace:

- $b=4$; $M=2^b$; $L=2^{(b/2)}$; % Number of bits per symbol and Modulation order
- $N_s=40$; $T_s=1e-5$; $T=T_s/N_s$; % Symbol time and Sample time
- $w_c=2\pi*10/T_s$; % Carrier freq[rad/s]
- $E_b/N_0=10$; % $E_b/N_0$[dB]
- Target_no_of_error=100; % Simulation stopping criterion

![Simulink model for simulating a QAM passband communication system](QAM_passband_sim.mdl)
%do_QAM_sim.m
clear, clf
b=4; M=2^b; L=2^(b/2); % # of bits per symbol and Modulation order
Ns=b*10; Ts=1e-5; T=Ts/Ns; % Symbol time and Sample time
wc=2*pi*10/Ts; % Carrier Frequency [rad/s] (such that wc*T<pi)
Target_no_of_error=100;
SNRdBs=[5 10]; EbN0dBs=SNRdBs-3;
for i=1:length(SNRdBs)
    SNRdB=SNRdBs(i); EbN0dB=SNRdB-3; % Eb/(N0/2)=SNR-> Eb/N0=SNR/2
    sim('QAM_passband_sim',1e5*Ts); SERs(i) = ser(end,1);
end
SNRdBt=0:0.1:13; poset = prob_error(SNRdBt,'QAM',b,'sym');
semilogy(SNRdBt,poset,'k', SNRdBs,SERs,'*'), xlabel('SNR [dB]')
Transmitted_Signal_Power = 2*(M-1)/3/Ts;
Received_Signal_Power = ...
    Transmitted_Signal_Power*(1+10^(-(EbN0dBs(end)+3)/10)*Ts/b/T)

function a=slice(x,A)
    [Am,Im]=min(abs(A-x)); a=A(Im);

(a) After having composed and saved the Simulink model “QAM_passband_sim.mdl” and two MATLAB programs “do_QAM_sim.m” and “slice.m” (that can be substituted by an Embedded MATLAB Function Block), run the MATLAB program “do_QAM_sim.m” to get the SERs for SNRdBs=[5 10]. Does the simulation result conform with the theoretical symbol error probability that can be obtained using the MATLAB routine “prob_error()” in Sec. 7.6?

(b) Modify the program “do_QAM_sim.m” and/or Simulink model “QAM_passband_sim.mdl” to simulate an $M=2^6=64$-ary QAM communication system and run it to get the SER curve.

%do_MSK_sim.m
clear, clf
SNRdBs=[0:0.1:10]; SNRdBs= [5 10]; EbN0dBs= SNRdBs-3;
b=1; M=2^b; % Number of bits per symbol and Modulation order
Tb=1; Nb=16; T=Tb/Nb; Ts=b*Tb; % Bit duration and Sample time
wc=4*pi/Tb; % Carrier frequency [rad/s]
t=[0.2*Nb-1]*T; pihTbt=pi/2/Tb*t;
suT=sqrt(2/Tb)*T*[cos(pihTbt-pi/2).*cos(wc*(t-Tb));
    sin(pihTbt).*sin(wc*t)]; % Eq. (7.6.8a,b)
M_filter1=fliplr(suT(1,:)); M_filter2=fliplr(suT(2,:));
Target_no_of_error = 50;
Simulink_mdl='MSK_passband_sim';
for i=1:length(SNRdBs)
    SNRdB=SNRdBs(i); EbN0dB=SNRdB-3;
    sim(Simulink_mdl,1e5*Tb), BERs(i) = ber(end,1);
end
pobet_PSK=prob_error(SNRdBt,'PSK',b,'bit');
pobet_FSK=prob_error(SNRdBt,'FSK',b,'bit');
semilogy(SNRdBt,pobet_PSK,'k', SNRdBt,pobet_FSK,'k:*', ...
    SNRdBs,BERs,'b*')

function d=detector_MSK(dth)
    if dth>0, d=1; else d=0; end
    if abs(dth)>pi, d=1-d; end

%do_MSK_sim.m
7.10 Simulation of MSK Using Simulink

Figure P7.10 Simulink model for simulating an MSK communication system (*MSK_passband_sim.mdl*)
Fig. P7.10 shows a Simulink model to simulate an MSK (minimum shift keying) passband communication system where the following parameters are to be set in the workspace. Note that the demodulator is implemented using matched (FIR) filters (whose coefficient vectors or impulse responses are the reversed and delayed versions of the basis signal waveforms (P7.6.8a,b)) instead of correlators. Having composed and saved the Simulink model “MSK_passband_sim.mdl” and two MATLAB programs “do_MSK_sim.m” and “detector_MSK.m”, run the MATLAB program “do_MSK_sim.m” to get the BERs for SNRdBs=[5 10]. Does the simulation result go into between the theoretical symbol error probabilities for FSK and PSK?

```matlab
b=1; M=2^b; % Number of bits per symbol and Modulation order
Nb=16; Ns=b*Nb;
Tb=1e-5; Ts=b*Tb; T=Ts/Ns; % Symbol/Bit time and Sample time
wc=2*pi*10/Tb; % Carrier freq [rad/s]
t=[0:2*Nb-1]*T; pihTbt=pi/2/Tb*t;
suT=sqrt(2/Tb)*T*[cos(pihTbt-pi/2).*cos(wc*(t-Tb));
                 sin(pihTbt).*sin(wc*t)];
M_filter1=fliplr(suT(1,:)); M_filter2=fliplr(suT(2,:));
EbN0dB=7; % Eb/N0 [dB]
Target_no_of_error=20; % Simulation stopping criterion
```
9.4.3 Cyclic Coding

A cyclic code is a linear block code having the property that a cyclic shift (rotation) of any codeword yields another codeword. Due to this additional property, the encoding and decoding processes can be implemented more efficiently using a feedback shift register. An \((N, K)\) cyclic code can be described by an \((N-K)\)th-degree generator polynomial

\[
g(x) = g_0 + g_1 x + g_2 x^2 + \cdots + g_{N-K} x^{N-K}
\]

(9.4.24)

The procedure of encoding a \(K\)-bit message vector \(m = [m_0 \ m_1 \ \cdots \ m_{K-1}]\) represented by a \((K-1)\)th-degree polynomial

\[
m(x) = m_0 + m_1 x + m_2 x^2 + \cdots + m_{K-1} x^{K-1}
\]

(9.4.25)

into an \(N\)-bit codeword represented by an \((N-1)\)th-degree polynomial is as follows:

1. Divide \(x^{N-K} m(x)\) by the generator polynomial \(g(x)\) to get the remainder polynomial \(r_m(x)\).
2. Subtract the remainder polynomial \(r_m(x)\) from \(x^{N-K} m(x)\) to obtain a codeword polynomial

\[
c(x) = x^{N-K} m(x) \oplus r_m(x) = q(x) g(x)
\]

(9.4.26)

which has the generator polynomial \(g(x)\) as a (multiplying) factor. Then the first \((N-K)\) coefficients constitute the parity vector and the remaining \(K\) coefficients make the message vector. Note that all the operations involved in the polynomial multiplication, division, addition, and subtraction are not the ordinary arithmetic ones, but the modulo-2 operations.

(Example 9.6) A Cyclic Code

With a \((7,4)\) cyclic code represented by the generator matrix

\[
g(x) = g_0 + g_1 x + g_2 x^2 + g_3 x^3 = 1 + 1 \cdot x + 0 \cdot x^2 + 1 \cdot x^3
\]

(E9.6.1)

find the codeword for a message vector \(m = [1 \ 0 \ 1 \ 1]\).

Noting that \(N = 7\), \(K = 4\), and \(N - K = 3\), we divide \(x^{N-K} m(x)\) by \(g(x)\) as

\[
x^{N-K} m(x) = x^3 (1 + 0 \cdot x + 1 \cdot x^2 + 1 \cdot x^3) = x^6 + x^5 + x^3 = q(x) g(x) + r_m(x)
\]

(E9.6.2)

to get the remainder polynomial \(r_m(x) = 1\) and add it to \(x^{N-K} m(x) = x^3 m(x)\) to make the codeword polynomial as

\[
c(x) = r_m(x) + x^3 m(x) = 1 + 0 \cdot x + 0 \cdot x^2 + 1 \cdot x^3 + 0 \cdot x^4 + 1 \cdot x^5 + 1 \cdot x^6
\]

(E9.6.3)

\[
\begin{array}{c|c}
\text{parity} & \text{message} \\
[1 \ 0 \ 0 \ 1 \ 0 \ 1 \ 1] \\
\end{array}
\]

The codeword made in this way has the \(N-K=3\) parity bits and \(K=4\) message bits.
Now, let us consider the procedure of decoding a cyclic coded vector. Suppose the RCVR has received a possibly corrupted code vector \( \mathbf{r} = \mathbf{c} + \mathbf{e} \) where \( \mathbf{c} \) is a codeword and \( \mathbf{e} \) is an error. Just as in the encoder, this received vector, being regarded as a polynomial, is divided by the generator polynomial \( \mathbf{g}(x) \)

\[
\mathbf{r}(x) = \mathbf{c}(x) + \mathbf{e}(x) = \mathbf{q}(x)\mathbf{g}(x) + \mathbf{e}(x) = \mathbf{q}'(x)\mathbf{g}(x) + \mathbf{s}(x)
\]

(9.4.27)

to yield the remainder polynomial \( \mathbf{s}(x) \). This remainder polynomial \( \mathbf{s} \) may not be the same as the error vector \( \mathbf{e} \), but at least it is supposed to have a crucial information about \( \mathbf{e} \) and therefore, may well be called the *syndrome*. The RCVR will find the error pattern \( \mathbf{e} \) corresponding to the syndrome \( \mathbf{s} \), subtract it from the received vector \( \mathbf{r} \) to get hopefully the correct codeword

\[
\mathbf{c} = \mathbf{r} \oplus \mathbf{e}
\]

(9.4.28)

and accept only the last \( K \) bits (in ascending order) of this corrected codeword as a message.

The polynomial operations involved in encoding/decoding every block of message/coded sequence seem to be an unbearable computational load. However, we fortunately have divider circuits which can perform such a modulo-2 polynomial operation. Fig. 9.9 illustrates the two divider circuits (consisting of linear feedback shift registers) each of which carries out the modulo-2 polynomial operations for encoding/decoding with the cyclic code given in Example 9.6. Note that the encoder/decoder circuits process the data sequences in descending order of polynomial.

The encoder/decoder circuits are cast into the MATLAB routines ‘cyclic_encoder()’ and ‘cyclic_decoder0()’, respectively, and we make a program “do_cyclic_code.m” that uses the two routines ‘cyclic_encoder()’ and ‘cyclic_decoder()’ (including ‘cyclic_encoder0()’) to simulate the encoding/decoding process with the cyclic code given in Example 9.6. Note a couple of things about the decoding routine ‘cyclic_decoder()’:

- It uses a table of error patterns in the matrix \( \mathbf{E} \), which has every correctable error pattern in its rows. The table is searched for a suitable error pattern by using an error pattern index vector \( \mathbf{epi} \), which is arranged by the decimal-coded syndrome and therefore, can be addressed efficiently by a syndrome just like a decoding hardware circuit.

- If the error pattern table \( \mathbf{E} \) and error pattern index vector \( \mathbf{epi} \) are not supplied from the calling program, it uses ‘cyclic_decoder0()’ to supply itself with them.

```
% do_cyclic_code.m
% tries with a cyclic code.
clear
N=7; K=4; % N=15; K=7; % Codeword (Block) length and Message size
%N=31; K=16;
g=cyclpoly(N,K); g_=fliplr(g);
lm=5*K; msg= randint(1,lm);
% N=7; K=4; g_=[1 1 0 1]; g=fliplr(g_); msg=[1 0 1 1];
coded = cyclic_encoder(msg,N,K,g_); lc=length(coded);
no_transmitted_bit_errors=ceil(lc*0.05);
errors=randerr(1,lc,no_transmitted_bit_errors);
r = rem(coded+errors,2); % Received sequence
decoded = cyclic_decoder(r,N,K,g_); nobe=sum(decoded==msg)
coded1 = encode(msg,N,K,'cyclic',g); % Use the Communication Toolbox
r1 = rem(coded1+errors,2); % Received sequence
decoded1 = decode(r1,N,K,'cyclic',g); nobel=sum(decoded1==msg)
```
function coded= cyclic_encoder(msg_seq,N,K,g)
% Cyclic (N,K) encoding of input msg_seq m with generator polynomial g
Lmsg=length(msg_seq); Nmsg=ceil(Lmsg/K);
Msg= [msg_seq(:); zeros(Nmsg*K-Lmsg,1)];
Msg= reshape(Msg,K,Nmsg).';
coded= [];
for n=1:Nmsg
  msg= Msg(n,:);
  for i=1:N-K, x(i)=0; end
  for k=1:K
    tmp= rem(msg(K+1-k)+x(N-K),2); % msg(K+1-k)+g(N-K+1)*x(N-K)
    for i=N-K:-1:2,  x(i)= rem(x(i-1)+g(i)*tmp,2); end
    x(1)=g(1)*tmp;
  end
  coded= [coded x msg]; % Eq.(9.4.26)
end
end

function [decodes,E,epi]=cyclic_decoder(code_seq,N,K,g,E,epi)
% Cyclic (N,K) decoding of received code_seq with generator polynomial g
% E:   Error Pattern matrix or syndromes
% epi: error pattern index vector
%Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
if nargin<6
  nceb=ceil((N-K)/log2(N+1)); % Number of correctable error bits
  E = combis(N,nceb); % All error patterns
  for i=1:size(E,1)
    syndrome=cyclic_decoder0(E(i,:),N,K,g);
    synd_decimal=bin2deci(syndrome);
    epi(synd_decimal)=i; % Error pattern indices
  end
end
if (size(code_seq,2)==1)  code_seq=code_seq.'; end
Lcode= length(code_seq);  Ncode= ceil(Lcode/N);
Code_seq= [code_seq(:); zeros(Ncode*N-Lcode,1)];
Code_seq= reshape(Code_seq,N,Ncode).';
decodes=[]; syndromes=[];
for n=1:Ncode
  code= Code_seq(n,:);
  syndrome= cyclic_decoder0(code,N,K,g);
  si= bin2deci(syndrome); % Syndrome index
  if 0<si&si<=length(epi) % Syndrome index to error pattern index
    m=epi(si); if m>0, code=rem(code+E(m,:),2); end % Eq.(9.4.28)
  end
  decodes=[decodes code(N-K+1:N)]; syndromes=[syndromes syndrome];
end
if nargout==2, E=syndromes; end

function x=cyclic_decoder0(r,N,K,g)
% Cyclic (N,K) decoding of an N-bit code r with generator polynomial g
for i=1:N-K, x(i)=r(i+K); end
for n=1:K
  tmp=x(N-K);
  for i=N-K:-1:2,  x(i)=rem(x(i-1)+g(i)*tmp,2); end
  x(1)=rem(g(1)*tmp+r(K+1-n),2);
end
9.4 Channel Coding

(a) A cyclic encoder implemented in a feedback shift register structure

(b) A cyclic decoder implemented in a feedback shift register structure

Figure 9.9 An example of cyclic encoding/decoding processes using divisor circuits for polynomial division
Table 9.1 Communication Toolbox functions for block coding

<table>
<thead>
<tr>
<th>Block coding</th>
<th>Related Communication Toolbox functions and objects</th>
</tr>
</thead>
<tbody>
<tr>
<td>Linear block</td>
<td>encode, decode, gen2par, syndtable</td>
</tr>
<tr>
<td>Cyclic</td>
<td>encode, decode, cyclpoly, cyclgen, gen2par, syndtable</td>
</tr>
<tr>
<td>BCH (Bose-Chaudhuri-Hocquenghem)</td>
<td>bchenc, bchdec, bchgenpoly</td>
</tr>
<tr>
<td>LDPC (Low-Density Parity Check)</td>
<td>fec.ldpcenc, fec.ldpcdec</td>
</tr>
<tr>
<td>Hamming</td>
<td>encode, decode, hammgen, gen2par, syndtable</td>
</tr>
<tr>
<td>Reed-Solomon</td>
<td>rsenc, rsdec, rsgenpoly, rsencof, rsdecof</td>
</tr>
</tbody>
</table>

Table 9.1 shows a list of several Communication Toolbox functions that can be used to simulate various block coding techniques where the block codings are classified as follows (see the MATLAB Help on Communication Toolbox - block coding):

- **Linear Block Codes**
  - LDPC (Low-Density Parity-Check) codes
  - Cyclic codes - BCH (Bose-Chaudhuri-Hocquenghem) code
  - Hamming codes
  - Reed-Solomon codes

Note the following about the functions ‘encode’ and ‘decode’ (use MATLAB Help for details):

- They can be used for any linear block coding by putting a string ‘linear’ and a $K \times N$ generator matrix as the fourth and fifth input arguments, respectively.
- They can be used for Hamming coding by putting a string ‘hamming’ as the fourth input argument or by providing them with only the first three input arguments.

The following example illustrates the usages of ‘encode()’/‘decode()’ for cyclic coding and ‘rsenc()’ and ‘rsdec()’ for Reed-Solomon coding. Note that the $RS$ (Reed-Solomon) codes are nonbinary BCH codes, which has the largest possible minimum distance for any linear code with the same message size $K$ and codeword length $N$, yielding the error correcting capability of $\left\lfloor \frac{N-K}{2} \right\rfloor$. 

```matlab
%test_encode_decode.m  to try using encode()/decode()
N=7; K=4; % Codeword (Block) length and Message size
g=cyclpoly(N,K); % Generator polynomial for a cyclic (N,K) code
Nm=10; % # of K-bit message vectors
msg=randint(Nm,K); % Nm x K message matrix
encoded = encode(msg,N,K,'cyclic',g); % Encoding
% Add bit errors with transmitted BER potbe=0.1
potbe=0.1; received=rem(encoded+randerr(Nm,N,[0 1;1-potbe potbe]),2);
decoded=decode(received,N,K,'cyclic',g); % Decoding
% Probability of message bit errors after decoding/correction
pobe=sum(sum(decoded~=msg))/(Nm*K) % BER
% Usage of rsenc()/rsdec()
M=3; % Galois Field integer corresponding to the # of bits per symbol
N=2^M-1; K=3; dc=(N-K)/2; % Codeword length and Message size
msg=gf(randint(Nm,K,2^M),M); % Nm x K GF(2^M) Galois Field msg matrix
coded = rsenc(msg,N,K); % Encoding
noise = randerr(Nm,N,[1 dc+1]).*randint(Nm,N,2^M);
received = coded+noise; % Add a noise
[decoded,numerr]=rsdec(received,N,K); % Decoding
[msg decoded], numerr, pose=sum(sum(decoded~=msg))/(Nm*K) % SER
```
9.4.4 Convolutional Coding and Viterbi Decoding

In the previous sections, we discussed the block coding that encodes every $K$-bit block of message sequence independently of the previous message block (vector). In this section, we are going to see the convolutional coding that converts a $K$-bit message vector into an $N$-bit channel input sequence dependently of the previous $(L-1)K$-bit message vector ($L$: constraint length). The convolutional encoder has a structure of finite-state machine whose output depends on not only the input but also the state.

Fig. 9.10 shows a binary convolutional encoder with a $K(=2)$-bit input, an $N(=3)$-bit output, and $L-1(=3)$ 2-bit registers that can be described as a finite-state machine having $2^{(L-1)K} = 2^{2^3} = 64$ states. This encoder shifts the previous contents of every stage register except the right-most one into its righthand one and receives a new $K$-bit input to load the left-most register at an iteration, sending an $N$-bit output to the channel for transmission where the values of the output bits depends on the previous inputs stored in the $L-1$ registers as well as the current input.

A binary convolutional code is also characterized by $N$ generator sequences $g_1, g_2, \ldots, g_N$ each of which has a length of $LK$. For example, the convolutional code with the encoder depicted in Fig. 9.10 is represented by the $N(=3)$ generator sequences

\[
\begin{align*}
g_1 &= [0 \ 0 \ 1 \ 0 \ 0 \ 1] \\
g_2 &= [0 \ 0 \ 0 \ 0 \ 0 \ 1] \\
g_3 &= [1 \ 0 \ 0 \ 0 \ 0 \ 0] \\
\end{align*}
\]

(9.4.29a)

which constitutes the generator (polynomial) matrix

\[
G_{NxLK} = \begin{bmatrix}
g_1 \\
g_2 \\
g_3
\end{bmatrix} = \begin{bmatrix}
0 & 0 & 1 & 0 & 0 & 1 \\
0 & 0 & 0 & 0 & 0 & 1 \\
1 & 0 & 0 & 0 & 0 & 0
\end{bmatrix}
\]

(9.4.29b)

where the value of the $j^{th}$ element of $g_i$ is 1 or 0 depending on whether the $j^{th}$ one of the $LK$ bits of the shift register is connected to the $i^{th}$ output combiner or not. The shift register is initialized to all-zero state before the first bit of an input (message) sequence enters the encoder and also finalized to all-zero state by the $(1) LK$ zero-bits padded onto the tail part of each input sequence. Besides, the length of each input sequence (to be processed at a time) is made to be $MK$ (an integer $M$ times $K$) even by zero-padding if necessary. For the input sequence made in this way so that its total length is $(M+L-1)K$ including the zeros padded onto it, the length of the output sequence is $(M+L-1)N$ and consequently, the code rate will be

\[
R_c = \frac{MK}{(M+L-1)N} \rightarrow \frac{K}{N} \quad \text{for} \quad M > L
\]

(9.4.30)
function [output,state]=conv_encoder(G,K,input,state,termmode)
% generates the output sequence of a binary convolutional encoder
% G    : N x LK Generator matrix of a convolutional code
% K    : Number of input bits entering the encoder at each clock cycle.
% input: Binary input sequence
% state: State of the convolutional encoder
% termmode='trunc' for no termination with all-0 state
%Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
if isempty(G), output=input; return; end
tmp= rem(length(input),K);
input=[input zeros(1,(K-tmp)*(tmp>0))];
[N,LK]=size(G);
if rem(LK,K)>0
    error('The number of column of G must be a multiple of K!')
end
%L=LK/K;
if nargin<4|(nargin<5 & isnumeric(state))
    input= [input zeros(1,LK)]; %input= [input zeros(1,LK-K)]; end
end
if nargin<4|~isnumeric(state)
    state=zeros(1,LK-K);
end
input_length= length(input);
N_msgsymbol= input_length/K;
input1= reshape(input,K,N_msgsymbol);
output=[];
for l=1:N_msgsymbol % Convolution output=G*input
    ub= input1(:,l).';
    [state,yb]= state_eq(state,ub,G);
    output= [output yb];
end
function [nxb,yb]=state_eq(xb,u,G)
% To be used as a subroutine for conv_encoder()
K=length(u); LK=size(G,2); L1K=LK-K;
if isempty(xb), xb=zeros(1,L1K);
else
    N=length(xb); %(L-1)K
    if L1K~=N, error('Incompatible Dimension in state_eq()'); end
end
A=[zeros(K,L1K); eye(L1K-K) zeros(L1K-K,K)];
B=[eye(K); zeros(L1K-K,K)];
C=G(:,K+1:end); D=G(:,1:K);
nxb=rem(A*xb'+B*u',2)';
yb=rem(C*xb'+D*u',2)';

Given a generator matrix $G_{N\times LK}$ together with the number $K$ of input bits and a message sequence $m$, the above MATLAB routine `conv_encoder()` pads the input sequence $m$ with zeros as needed and then generates the output sequence of the convolutional encoder. Communication Toolbox has a convolutional encoding function `convenc()` and its usage will be explained together with that of a convolutional decoding function `vitdec()` at the end of this section.
9.4 Channel Coding

Various Representations of a Convolutional Code

There are many ways of representing a convolutional code such as the schematic diagram like Fig. 9.10, the generator matrix like Eq. (9.4.29b), the finite-state machine (state transition diagram), the transfer function, and the trellis diagram. Fig. 9.11 illustrates the various equivalent representations of a simple convolutional code with \( K=1, \ N=2, \) and \( L=3. \) Especially, Fig. 9.11(c1) shows the state diagram that has 4 states corresponding to all possible contents \{a=00, b=01, c=10, d=11\} of the lefthand \((L-1)K\) registers in the schematic diagram (Fig. 9.11(a)) where solid/dashed branches between two states represent the state transitions in the arrow direction in response to input 0/1, respectively. Fig. 9.11(c2) is an augmented state diagram where the all-zero state \( a \) in the original state diagram (Fig. 9.11(c1)) is duplicated to make two all-zero states, one \((a)\) with only out-ward branches and the other \((a')\) with only in-ward branches. Note that the self-loop at the all-zero state is neglected so that the gains of all the paths starting from the all-zero state and coming back to the all-zero state can be obtained in a systematic way as follows. If we assign the gain \( D^\alpha I^\beta J^\gamma \) (\( \alpha \): the Hamming weight (number of 1’s) of the output sequence assigned to the branch, \( \beta \): the Hamming weight of the input sequence causing the output sequence to be generated, and \( \gamma \): is the number of branches contained in the path) to get the overall gain from the input all-zero state \( a \) to the output all-zero state \( a' \) as

\[
T(D, I, J) = \frac{1}{\Delta} \sum_i M_i \Delta_i
\]

\[
= \sum_1 \text{All products of the forward gain } M_i \text{ and } \Delta_i \text{ for the diagram without forward path } i
\]

\[
= \sum_1 \text{All loop gains} + \sum_2 \text{All products of 2 nontouching loop gains} - \cdots
\]

\[
= \frac{D^5 I^3 J^3 (1-DIJ)+D^6 I^2 J^4}{1-DIJ-DIJ^2-D^2 I^2 J^3+D^2 I^2 J^3} = \frac{D^5 IJ^3}{1-(DIJ+DIJ^2)}
\]

This can be expanded into a power series as

\[
T(D, I, J) = D^5 IJ^3 \{1+(DIJ+DIJ^2)+(DIJ+DIJ^2)^2+\cdots\}
\]

\[
= D^5 IJ^3 + D^6 I^2 J^4 + D^6 I^2 J^4 + D^7 I^3 J^5 + \cdots
\]

Each term \( D^\alpha I^\beta J^\gamma \) of this equation denotes the gain of each forward path starting from the all-zero state and going back to the all-zero state where \( \alpha \) is the Hamming weight of the codeword represented by the path, \( \beta \) is the Hamming weight of the input sequence causing the output sequence to be generated, and \( \gamma \) is the number of branches contained in the path. In this context, the first term \( D^5 IJ^3 \) of Eq. (9.4.32) implies that the convolutional encoder in Fig. 9.11(a) may generate a codeword of Hamming weight \( \alpha(5) \) and \( \gamma-N = 3 \cdot 2 = 6 \) bits for the input sequence of Hamming weight \( \beta(1) \):

<table>
<thead>
<tr>
<th>Input:</th>
<th>[1]</th>
<th>[0]</th>
<th>[0]</th>
<th>( \beta=1 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>State:</td>
<td>( a(00) \rightarrow c(10) \rightarrow b(01) \rightarrow e(00) )</td>
<td>( \gamma=3 )</td>
<td>( \alpha=5 )</td>
<td>( D^5 IJ^3 )</td>
</tr>
<tr>
<td>Output:</td>
<td>11 01 11</td>
<td>( \beta=1 )</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

The two codewords corresponding to the terms \( D^6 I^2 J^4 \) and \( D^7 I^3 J^5 \) are

<table>
<thead>
<tr>
<th>Input:</th>
<th>[1]</th>
<th>[0]</th>
<th>[0]</th>
<th>[0]</th>
<th>[1]</th>
<th>[1]</th>
<th>[0]</th>
<th>[0]</th>
<th>[0]</th>
</tr>
</thead>
<tbody>
<tr>
<td>State:</td>
<td>( a(00) \rightarrow c(10) \rightarrow d(11) \rightarrow b(01) \rightarrow e(00) )</td>
<td>( a(00) \rightarrow c(10) \rightarrow d(11) \rightarrow d(11) \rightarrow b(01) \rightarrow e(00) )</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Output:</td>
<td>11 10 10 11</td>
<td>11 10 01 10 11</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
Chapter 9  Information and Coding

(a) Convolutional encoder represented by a schematic diagram and its operation

\[
G_{2\times2} = \begin{bmatrix} 0 & 1 \\ 1 & 1 \end{bmatrix}
\]

\[
g_1(x) = m(x)g_1(x) = (1+0.x+1.x^2)(1+0.x+1.x^2) = 1+0.x+0.x^2+0.x^3+1.x^4
\]

\[
g_2(x) = m(x)g_2(x) = (1+0.x+1.x^2)(1+1.x+1.x^2) = 1+1.x+0.x^2+0.x^3+1.x^4
\]

Output sequence to the input sequence \([0 1]\) = \([1 1 0 0 0 0]\) (Codeword)

(b) Convolutional encoder represented by a generator polynomial matrix and its operation

(c1) State diagram

(c2) Augmented state diagram with the initial and final all-zero state distinguished and \(D^3\) assigned to each branch (\(D^3\): Hamming weights of input/output)

(d) Trellis diagram

Figure 9.11 Various equivalent representations of a convolutional encoder
The minimum degree in $D$ of the transfer function, that is $d_{\text{min}}=5$ in the case of Eq. (9.4.32), is the free distance of the convolutional code, which is the minimum (Hamming) distance among the codewords in the code. Note that the free distance as well as the decoding algorithm heavily affects the BER performance of a convolutional code.

Fig. 9.11(d) shows another representation of the convolutional code called a trellis, which can be viewed as a plot of state diagram being developed along the time. This diagram consists of the columns of $2^{L-1}K$ state nodes per clock cycle and the solid/dashed branches representing the state transition caused by input 0/1 as depicted in the state diagram (Fig. 9.11(c)) where the $N$-bit output of the encoder to the input is attached to each branch. In this trellis diagram, each path starting from the initial all-zero state and ending at another all-zero state represents a codeword. From the trellis shown in Fig. 9.11, we can see the shortest two codewords as

<table>
<thead>
<tr>
<th>Input:</th>
<th>[1]</th>
<th>[0]</th>
<th>[0]</th>
<th>Input:</th>
<th>[1]</th>
<th>[1]</th>
<th>[0]</th>
<th>[0]</th>
</tr>
</thead>
<tbody>
<tr>
<td>State:</td>
<td>a(00) $\rightarrow$ c(10) $\rightarrow$ b(01) $\rightarrow$ a(00)</td>
<td>State:</td>
<td>a(00) $\rightarrow$ c(10) $\rightarrow$ d(11) $\rightarrow$ b(01) $\rightarrow$ a(00)</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Output:</td>
<td>11</td>
<td>01</td>
<td>11</td>
<td>Output:</td>
<td>11</td>
<td>10</td>
<td>10</td>
<td>11</td>
</tr>
</tbody>
</table>

These correspond to the first two terms $D^5I^3J^1$ and $D^6I^2J^4$ of Eq. (9.4.32), respectively.

### <Viterbi Decoding of a Convolutional Coded Sequence>

As a way of decoding a convolutional coded sequence, an ML (maximum-likelihood) decoding called the Viterbi algorithm (VA)\cite{F-1} is most widely used where an ML decoding finds the code sequence that is most likely to have been the input to the encoder at XMTR yielding the received code sequence. Given a received (DTR output or decoder input) sequence $r = m + e$ (\(m\) : a convolutional coded sequence, \(e\) : an error), the Viterbi algorithm with soft-decision searches the trellis diagram for an all-zero-state-to-all-zero-state path whose (encoder) output sequence is closest to $r$ in terms of Euclidean distance and takes the encoder input sequence corresponding to the ‘optimal’ path to be the most likely message sequence, while the Viterbi algorithm with hard-decision first slices $r$ to make $r_q$ consisting of 0 or 1 and then finds an all-zero-state-to-all-zero-state path whose (encoder) output sequence is closest to $r_q$ in terms of Hamming distance.

The procedure of the Viterbi algorithm is as follows (see Fig. 9.12):

0. To each set of nodes (states), assign the depth level index starting from $l=0$ for the leftmost stage. Divide the decoder input sequence $r_q$ (hard, i.e. sliced to 0/1) or $r$ (soft) into $N$-bit subsequences and distribute each $N$-bit subsequence to the stages between the node sets.

1. To each branch, attach the (Hamming or Euclidean) distance between the $N$-bit encoder output and the $N$-bit subsequence of $r$ or $r_q$ (distributable to the stage that the branch belongs to) as the branch cost (metric).

2. With the level index initialized to $l=0$, assign 0 to the initial all-zero state node as its node cost (metric).

3. Move right by one stage by increasing the level index by one. To every node (at the level) that is connected via some branch(es) to node(s) at the previous level, assign the node cost that is computed by adding the branch cost(s) to the left-hand node cost(s) and taking the minimum if there are multiple branches connected to the node. In the case of multiple branches connected to a node, remove other branches than the survivor branch yielding the minimum path cost.

4. If all the $N$-bit subsequence of $r$ or $r_q$ are not processed, go back to step 3 to repeat the procedure; otherwise, go to the next step 5.

5. Starting from the final all-zero state or the best node with minimum cost, find the optimal path consisting of only the survivor paths and take the encoder input sequence supposedly having caused the output sequence (associated with the optimal path) to be the ML decoded message sequence.
Fig. 9.12 illustrates a typical decoding procedure using the trellis based on the Viterbi algorithm (VA) stated above where the detector output (code) sequence is $[11011010101100]$. Note that the removed branches other than the survivor branches are crossed and the optimal path denoted by a series of thick solid/dotted lines (with encoder input 0/1) corresponds to the encoder input sequence $[1011000]$, which has been taken as the decoded result.

Given the generator polynomial matrix $G$ together with the number $K$ of input bits and the channel-DTR output sequence ‘detected’ as its input arguments, the MATLAB routine ‘vit_decoder(G,K,detected)’ constructs the trellis diagram and applies the Viterbi algorithm to find the maximum-likelihood decoded message sequence. The following MATLAB program “do_vitdecoder.m” uses the routine ‘conv_encoder()’ to make a convolutional coded sequence for a message and uses “vit_decoder()” to decode it to recover the original message.

```matlab
%do_vitdecoder.m
% Try using conv_encoder() / vit_decoder()
clear, clf
msg=[1 0 1 1 0 0 0]; % msg=randint(1,100)
lm=length(msg); % Message and its length
G=[1 0 1;1 1 1]; % N x LK Generator polynomial matrix
K=1; % Size of encoder input/output
potbe=0.02; % Probability of transmitted bit error
% Use of conv_encoder() / vit_decoder()
ch_input=conv_encoder(G,K,msg) % Self-made convolutional encoder
notbe=ceil(potbe*length(ch_input));
error_bits=randerr(1,length(ch_input),notbe);
detected= rem(ch_input+error_bits,2); % Received/modulated/detected
decoded= vit_decoder(G,K,detected)
noe_vit_decoder=sum(msg==decoded(1:lm))
```
function decoded_seq=vit_decoder(G,K,detected,opmode,hard_or_soft)
% performs the Viterbi algorithm on detected to get the decoded_seq
% G: N x LK Generator polynomial matrix
% K: Number of encoder input bits
% Copyright: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
% detected = detected(:).';
if nargin<5 | hard_or_soft(1)=='h', detected=(detected>0.5); end
[N,LK]=size(G);
if rem(LK,K)~=0, error('Column size of G must be a multiple of K'); end
tmp= rem(length(detected),N);
if tmp>0, detected=[detected  zeros(1,N-tmp) ]; end
b=LK-K; % Number of bits representing the state
no_of_states=2^b; N_mgsymbol=length(detected)/N;
for m=1:no_of_states
    states(m,n)=0; % inactive in the trellis
    p_state(m,n)=0; n_state(m,n)=0; input(m,n)=0;
end
states(1,1)=1; % make the initial state active
cost(1,1)=0; K2=2^K;
for n=1:N_mgsymbol
    y=detected((n-1)*N+1:n*N); % Received sequence
    n1=n+1;
    for m=1:no_of_states
        if states(m,n)==1 % active
            xb=deci2bin1(m-1,b);
            for m0=1:K2
                u=deci2bin1(m0-1,K);
                [nxb(m0,:),yb(m0,:)]=state_eq(xb,u,G);
                nxm0=bin2deci(nxb(m0,:))+1;
                states(nxm0,n1)=1;
                dif=sum(abs(y-yb(m0,:))); d(m0)=cost(m,n)+dif;
                if p_state(nxm0,n1)==0 % Unchecked state node?
                    cost(nxm0,n1)=d(m0);
                    p_state(nxm0,n1)=m; input(nxm0,n1)=m0-1;
                else
                    [cost(nxm0,n1),i]=min([d(m0) cost(nxm0,n1)]);
                    if i==1, p_state(nxm0,n1)=m; input(nxm0,n1)=m0-1; end
                end
            end
        end
    end
end
decoded_seq=[];
if nargin>3 & ~strcmp(opmode,'term',4)
    [min_dist,m]=min(cost(:,n1)); % Trace back from best-metric state
else m=1; % Trace back from the all-0 state
end
for n=n1:-1:2
    decoded_seq= [deci2bin1(input(m,n),K) decoded_seq];
    m=p_state(m,n);
end
The following program “do_vitdecoder1.m” uses the Communication Toolbox functions ‘convenc()’ and ‘vitdec()’ where ‘vitdec()’ is used several times with different input argument values to show the readers its various usages. Now, it is time to see the usage of the function ‘vitdec()’. 

```matlab
%do_vitdecoder1.m
% shows various uses of Communication Toolbox function convenc()
% with KxN Code generator matrix Gc - octal polynomial representation
%clear, clf
%msg=[1 0 1 1 0 0 0];
msg=randint(1,100)
lm=length(msg); % Message and its length
potbe=0.02; % Probability of transmitted bit error
Gc=[5 7]; % 1 0 1 -> 5, 1 1 1 -> 7 (octal number)
Lc=3; % 1xK constraint length vector for each input stream
[K,N]=size(Gc); % Number of encoder input/output bits
trel=poly2trellis(Lc,Gc); % Trellis structure
ch_input1=convenc(msg,trel); % Convolutional encoder
notbe1=ceil(potbe*length(ch_input1));
error_bits1=randerr(1,length(ch_input1),notbe1);
detected1= rem(ch_input1+error_bits1,2); % Received/modulated/detected
% with hard decision
Tbdepth=3;% Traceback depth
decoded1= vitdec(detected1,trel,Tbdepth,'trunc','hard')
noe_vitdec_trunc_hard=sum(msg~=decoded1(1:lm))
decoded2= vitdec(detected1,trel,Tbdepth,'cont','hard');
noe_vitdec_cont_hard=sum(msg(1:end-Tbdepth)==decoded2(Tbdepth+1:end))
% with soft decision
ncode= [detected1+0.1*randn(1,length(detected1)) zeros(1,Tbdepth*N)];
quant_levels=[0.001,.1,.3,.5,.7,.9,.999];
NSDB=ceil(log2(length(quant_levels)));
% Number of Soft Decision Bits
qcode= quantiz(ncode,quant_levels); % Quantized
decoded3= vitdec(qcode,trel,Tbdepth,'trunc','soft',NSDB);
noe_vitdec_trunc_soft=sum(msg==decoded3(1:lm))
decoded4= vitdec(qcode,trel,Tbdepth,'cont','soft',NSDB);
noe_vitdec_cont_soft=sum(msg==decoded4(Tbdepth+1:end))

% Repetitive use of vitdec() to process the data block by block
delay=Tbdepth*K; % Decoding delay depending on the traceback depth
% Initialize the message sequence, decoded sequence,
% state metric, traceback state/input, and encoder state.
msg_seq=[]; decoded_seq=[];
% for the message sequence
msg=[randint(1,1000); % Generate the message sequence
ms=[msg zeros(1,delay)]; end % Append with zeros
[coded,encoder_state]=convenc(ms,trel,encoder_state);
[decoded,m,s,in]=vitdec(coded,trel,Tbdepth,'cont','hard',m,s,in);
decoded_seq=[decoded_seq decoded];

lm=length(msg_seq);
noe_repeated_use=sum(msg_seq(1:lm)==decoded_seq(delay+[1:lm]))
```
<Usage of the Viterbi Decoding Function ‘vitdec’ with ‘convenc’ and ‘poly2trellis’>

To apply the MATLAB functions ‘convenc’/‘vitdec’, we should first use ‘poly2trellis’ to build the trellis structure with an ‘octal code generator’ describing the connections among the inputs, registers, and outputs. Fig. 9.13 illustrates how the octal code generator matrix \( G_c \) as well as the binary generator matrix \( G \) and the constraint length vector \( L_c \) is constructed for a given convolutional encoder. An example of using ‘poly2trellis’ to build the trellis structure for ‘convenc’/‘vitdec’ is as follows:

\[
\text{trellis=poly2trellis}(L_c,G_c);
\]

Here is a brief introduction of the usages of the Communication Toolbox functions ‘convenc’ and ‘vitdec’. See the MATLAB Help manual or The Mathworks webpage[W-7] for more details.

(1) \( \text{coded=}\text{convenc}(\text{msg},\text{trellis}) \);

\( \text{msg} \): A message sequence to be encoded with a convolutional encoder described by ‘\text{trellis}’.

(2) \( \text{decoded=}\text{vitdec}(\text{coded},\text{trellis},\text{tbdepth},\text{opmode},\text{dectype},\text{NSDB}) \);

\( \text{coded} \): A convolutional coded sequence possibly corrupted by a noise. It should consist of binary numbers \((0/1)\), real numbers between \(1\) (logical zero) and \(-1\) (logical one), or integers between \(0\) and \(2^{\text{NSDB}}-1\) (NSDB: the number of soft-decision bits given as the optional 6th input argument) corresponding to the quantization level depending on which one of \{'hard', 'unquant', 'soft'\} is given as the value of the fifth input argument ‘dectype’ (decision type).

\( \text{trellis} \): A trellis structure built using the MATLAB function ‘poly2trellis’.

\( \text{tbdepth} \): Traceback depth (length), i.e., the number of trellis branches used to construct each traceback path. It should be given as a positive integer, say, about five times the constraint length. In case the fourth input argument ‘opmode’ (operation mode) is ‘cont’ (continuous), it causes the decoding delay, i.e., the number of zero symbols preceding the first decoded symbol in the output ‘decoded’ and as a consequence, the decoded result should be advanced by \( \text{tbdepth} \times K \) where \( K \) is the number of encoder input bits.

\( \text{opmode} \): Operation mode of the decoding process. If it is set to ‘cont’ (continuous mode), the internal state of the decoder will be saved for use with the next frame. If it is set to ‘trunc’ (truncation mode), each frame will be processed independently, and the traceback path starts at the best-metric state and always ends in the all-zero state. If it is set to ‘term’ (termination mode), each frame is treated independently, and the traceback path always starts and ends in the all-zero state. This mode is appropriate when the uncoded message signal has enough zeros, say, \( K \times \text{Max}(L_c)-1 \) zeros at the end of each frame to fill all memory registers of the encoder.

\( \text{dectype} \): Decision type. It should be set to ‘unquant’, ‘hard’, or ‘soft’ depending on the characteristic of the input coded sequence (coded) as follows:
- ‘hard’ (decision) when the coded sequence consists of binary numbers 0 or 1.
- ‘unquant’ when the coded sequence consists of real numbers between -1(logical 1) and +1(logical 0).
- ‘soft’ (decision) when the optional 6th input argument NSDB is given and the coded sequence consists of integers between 0 and \(2^{\text{NSDB}}-1\) corresponding to the quantization level.

\( \text{NSDB} \): Number of software decision bits used to represent the input coded sequence. It is needed and active only when dectype is set to ‘soft’.
(3) \[[\text{decoded}, \text{m}, \text{s}, \text{in}]=\text{vitdec}(\text{code}, \text{trellis}, \text{tbdepth}, \text{opmode}, \text{dectype}, \text{m}, \text{s}, \text{in})\]

This format is used for a repetitive use of \(\text{vitdec()}\) with the continuous operation mode where the state metric \(\text{m}\), traceback state \(\text{s}\), and traceback input \(\text{in}\) are supposed to be initialized to empty sets at first and then handed over successively to the next iteration.

See the above program “do_vitdecoder1.m” or the MATLAB help manual for some examples of using \(\text{convenc()}\) and \(\text{vitdec()}\).

9.4.5 Trellis-Coded Modulation (TCM)

By channel coding discussed in the previous sections, additional redundant bits for error control are transmitted in the code and as a consequence, wider bandwidth is needed to keep the same data rate while lower BER can be obtained with the same SNR or the SNR to obtain the same BER can be decreased. This shows a trade-off between bandwidth efficiency and power efficiency. As an approach to mediate between these two conflicting factors, TCM (trellis-coded modulation) was invented by Gottfried Ungerboeck\[U-1, U-2\]. It is just like a marriage between modulation residing on the constellation and coding living on the trellis toward a more effective utilization of bandwidth and power. A (code) rate-2/3 TCM encoder combining an \((N, K)=(3, 2)\) convolutional coding and an \(2^3=8\)-PSK modulation is depicted in Fig. 9.14(a). In this encoder, the 8-PSK signal mapper generates one of the eight PSK signals depending on the 3-bit symbol \(x_k=[x_{1k}, x_{2k}, x_{3k}]\) that consists of two (uncoded) message bits \(x_{1k}\) and \(x_{2k}\) and an additional coded bit that is the output of the FSM (finite-state machine). Since the encoding scheme can be described by a trellis, the Viterbi algorithm can be used to decode the TCM coded signal. There are a couple of things to mention about TCM:

- The number of signal points in the constellation is larger compared with the uncoded case. For example, the constellation size or modulation order used by the TCM scheme of Fig. 9.14(a) is \(2^{K+1}=8\). This is two times as large as that of QPSK modulation that would be used to send the message by \(K=2\)-bit symbols with no coding. Note that a larger modulation order with the same SNR decreases the minimum distance among the codewords so that the BER may suffer.
- TCM has some measures to combat the possible BER performance degradation problem:
  - The convolutional coding allows only certain sequences (paths) of signal points so that the free distance among different signal paths in the trellis can be increased.
The TCM decoder uses soft-decision decoding to find the path with minimum (squared) Euclidean distance through the trellis. This makes the trellis code design trying to maximize the Euclidean distance among the codewords.

As illustrated in Fig. 9.14(b1) and (b2), the set partitioning let the more significant message bit(s) have a larger Euclidean distance from its complement so that there are less likely errors in the uncoded message bits than in the coded bit(s).

The FSM of TCM encoder has some inputs added to the adders between the shift registers. Therefore it seems that TCM encoding/decoding can not be implemented using the general convolutional encoder/decoder, requiring its own encoder/decoder rountines of every TCM encoder/decoder.

---

Figure 9.14. An example of TCM (Trellis-Coded Modulation) coding
The following program “sim_TCM.m” uses the two subroutines “TCM_encoder()” and “TCM_decoder()” to simulate the TCM encoding (shown in Fig. 9.14(a)) and the corresponding TCM decoding.

```matlab
%sim_TCM.m
% simulates a Trellis-coded Modulation
clear, clf
lm=1e4; msg=randint(1,lm);
K=2; N=K+1; Ns=3; % Size of encoder input/output/state
M=2^N; Constellation=exp(j*2*pi/M*[0:M-1]); % Constellation
ch_input=TCM_encoder('TCM_state_eq0',K,Ns,N,msg,Constellation);
lc=length(ch_input);
SNRbdB=5; SNRdB=SNRbdB+10*log10(K); % SNR per K-bit symbol
sigma=1/sqrt(10^(SNRdB/10));
optimal received = ch_input + noise; var(received)
received = awgn(ch_input,SNRdB); var(received) % Alternative
decoded_seq=TCM_decoder('TCM_state_eq0',K,Ns,received,Constellation);
ber_TCM8PSK = sum(decoded_seq(1:lm)~=msg)/lm
ber_QPSK_theory = prob_error(SNRbdB,'PSK',K,'BER')

function [output,state] = TCM_encoder(state_eq,K,Ns,N,input,state,Constellation,opmode)
% Generates the output sequence of a binary TCM encoder
% Input: state_eq = External function for state eqn saved in an M-file
% K = Number of input bits entering the encoder at each cycle
% Ns = Number of state bits of the TCM encoder
% N = Number of output bits of the TCM encoder
% input= Binary input message seq.
% state = State of the conv_encoder
% Constellation= Signal sets for signal mapper
% Output: output= Sequence of signal points on Constellation
% state = Updated state
% Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
tmp= rem(length(input),K);
input= [input zeros(1,((K-tmp)*(tmp>0))];
if nargin<6, state=zeros(1,Ns);
elseif length(state)==2^N, Constellation=state; state=zeros(1,Ns);
end
input_length= length(input); N_msgsymbol= input_length/K;
input1= reshape(input,K,N_msgsymbol).';
outputs= [];
for l=1:N_msgsymbol
    ub= input1(l,:);
    [state,output]=feval(state_eq,state,ub,Constellation);
    outputs= [outputs output];
end

function [s1,x]=TCM_state_eq0(s,u,Constellation)
% State equation for the TCM_encoder in Fig. 9.14(a)
% Input: s= State, u= Input, Constellation
% Output: s1= Next state, x= Output
s1= [s(3) rem([s(1)+u(1) s(2)+u(2)],2)];
x= Constellation(bin2deci([u(1) u(2) s(3)])+1);
function decoded_seq = TCM_decoder(state_eq,K,Nsb, received, Constellation, opmode)
% Performs the Viterbi algorithm on the PSK demodulated signal
% Input: state_eq = External function for state eqn saved in an M-file
% K = Number of input bits entering the encoder at each cycle.
% Nsb = Number of state bits of the TCM encoder
% received = received sequence
% Constellation: Signal sets for signal mapper

%Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
N_states = 2^Nsb;
N_msgsymbol = length(received);
for m = 1:N_states
    for n = 1:N_msgsymbol+1
        states(m,n) = 0; % inactive in the trellis diagram
        p_state(m,n) = 0; n_state(m,n) = 0; input(m,n) = 0;
    end
end
states(1,1) = 1; % make the initial state active
cost(1,1) = 0; K2 = 2^K;
for n = 1:N_msgsymbol
    y = received(n); % received sequence
    n1 = n+1;
    for m = 1:N_states
        if states(m,n) == 1 % active
            xb = deci2bin1(m-1, Nsb);
            for m0 = 1:K2
                u = deci2bin1(m0-1, K);
                [nxb(m0,:), yb(m0)] = feval(state_eq, xb, u, Constellation);
                nxm0 = bin2deci(nxb(m0,:)) + 1;
                states(nxm0,n1) = 1;
                % Accumulated squared Euclidean distance as path-to-node cost
                difference = y - yb(m0);
                d(m0) = cost(m,n) + difference* conj(difference);
                if p_state(nxm0,n1) == 0
                    cost(nxm0,n1) = d(m0);
                    p_state(nxm0,n1) = m; input(nxm0,n1) = m0-1;
                else
                    [cost(nxm0,n1), i] = min([d(m0) cost(nxm0,n1)]);
                    if i == 1, p_state(nxm0,n1) = m; input(nxm0,n1) = m0-1; end
                end
            end
            decoded_seq = [];
        end
    end
end
decoded_seq = [deci2bin1(input(m,n), K) decoded_seq];
end
9.4.6 Turbo Coding

In order for a linear block code or a convolutional code to approach the theoretical limit imposed by Shannon’s channel capacity (see Eq. (9.3.16) or Fig. 9.7) in terms of bandwidth/power efficiency, its codeword or constraint length should be increased to such an intolerable degree that the maximum likelihood decoding can become unrealizable. Possible solutions to this dilemma are two classes of powerful error correcting codes, each called turbo codes and LDPC (lower-density parity-check) codes, that can achieve a near-capacity (or near-Shannon-limit) performance with a reasonable complexity of decoder. The former is the topic of this section and the latter will be introduced in the next section.

Fig. 9.15(a) shows a turbo encoder consisting of two recursive systematic convolutional (RSC) encoders and an interleaver where the interleaver permutes the message bits in a random way before input to the second encoder. (Note that the modifier ‘systematic’ means that the uncoded message bits are imbedded in the encoder output stream as they are.) The code rate will be 1/2 or 1/3 depending on whether the puncturing is performed or not. (Note that puncturing is to omit transmitting some coded bits for the purpose of increasing the code rate beyond that resulting from the basic structure of the encoder.) Fig. 9.15(b) shows a demultiplexer, which classifies the coded bits into two groups, one from encoder 1 and the other from encoder 2, and applies each of them to the corresponding decoder. Fig. 9.15(c) shows a turbo decoder consisting of two decoders concatenated and separated by an interleaver where one decoder processes the systematic (message) bit sequence \( y^s \) and the parity bit sequence \( y^{1p}/y^{2p} \) together with the extrinsic information \( L_{e} \) (provided by the other decoder) to produce the information \( L_{a} \) and provides it to the other decoder in an iterative manner. The turbo encoder and the demultiplexer are cast into the MATLAB routines ‘encoderm()’ and ‘demultiplex()’, respectively. Now, let us see how the two types of decoder, each implementing the log-MAP (maximum a posteriori probability) algorithm and the SOVA (soft-out Viterbi algorithm), are cast into the MATLAB routines ‘logmap()’ and ‘sova()’, respectively.

<Log-MAP (Maximum a Posteriori Probability) Decoding cast into ‘logmap()’>

To understand the operation of the turbo decoder, let us begin with the definition of priori LLR (log-likelihood ratio), called a priori L-value, which is a soft value measuring how high the probability of a binary random variable \( u \) being +1 is in comparison with that of \( u \) being −1:

\[
L_u(u) = \ln \frac{P_u(u=+1)}{P_u(u=-1)} \quad \text{with } P_u(u) \text{: the probability of } u \text{ being } u \tag{9.4.33}
\]

This is a priori information known before the result \( y \) caused by \( u \) becomes available. While the sign of LLR

\[
\hat{u} = \text{sign}\{L_u(u)\} = \begin{cases} +1 & P_u(u=+1) > P_u(u=-1) \\ -1 & P_u(u=+1) < P_u(u=-1) \end{cases} \tag{9.4.34}
\]

is a hard value denoting whether or not the probability of \( u \) being +1 is higher than that of \( u \) being −1, the magnitude of LLR is a soft value describing the reliability of the decision based on \( \hat{u} \). Conversely, \( P_u(u=+1) \) and \( P_u(u=-1) \) can be derived from \( L_u(u) \):

\[
P_u(u=+1) = e^{L_u(u)} P_u(u=-1) 
\]

\[
P_u(u=-1) \rightarrow P_u(u=+1) = \frac{e^{L_u(u)}}{1+e^{L_u(u)}} \text{ and } P_u(u=-1) = \frac{1}{1+e^{L_u(u)}}
\]
function y = demultiplex(r,map,puncture)
%Copyright 1998, Yufei Wu, MPRG lab, Virginia Tech. for academic use
% map: Interleaver mapping
Nb = 3-puncture; lu = length(r)/Nb;
if puncture==0 % unpunctured
   for i=1:lu, y(:,2*i) = r(3*i-[1 0]).'; end
else % punctured
   for i=1:lu
      i2 = i*2;
      if rem(i,2)>0, y(:,i2)=[r(i2); 0]; else y(:,i2)=[0; r(i2)]; end
   end
sys_bit_seq = r(1,1:Nb:end); % the systematic bits for both decoders
y(:,1:2:lu*2) = [sys_bit_seq; sys_bit_seq(map)];

figure 9.15 A typical turbo coding
function x = rsc_encode(G,m,termination)
% Copyright 1998, Yufei Wu, MPRG lab, Virginia Tech. for academic use
% encodes a binary data block m (0/1) with a RSC (recursive systematic
% convolutional) code defined by generator matrix G, returns the output
% in x (0/1), terminates the trellis with all-0 state if termination>0
if nargin<3, termination = 0; end
[N,L] = size(G); % Number of output bits, Constraint length
M = L-1; % Dimension of the state
lu = length(m)+(termination>0)*M; % Length of the input
lm = lu-M; % Length of the message
state = zeros(1,M); % initialize the state vector
% To generate the codeword
x = [];
for i = 1:lu
    if termination<=0 | (termination>0 & i<=L_info)
        d_k = m(i);
    elseif termination>0 & i>lm
        d_k = rem(G(1,2:L)*state.',2);
    end
    a_k = rem(G(1,:)*[d_k state].',2);
    xp = rem(G(2,:)*[a_k state].',2); % 2nd output (parity) bits
    state = [a_k state(1:M-1)]; % Next state
    x = [x [d_k; xp]]; % since systematic, first output is input bit
end

function x = encoderm(m,G,map,puncture)
% Copyright 1998, Yufei Wu, MPRG lab, Virginia Tech. for academic use
% map: Interleaver mapping
% If puncture=0(unpunctured), it operates with a code rate of 1/3.
% If puncture>0(punctured), it operates with a code rate of 1/2.
% Multiplexer chooses odd/even-numbered parity bits from RSC1/RSC2.
[N,L] = size(G); % Number of output bits, Constraint length
M = L-1; % Dimension of the state
lm = length(m); % Length of the information message block
lu = lm + M; % Length of the input sequence
% 1st RSC coder output
x1 = rsc_encode(G,m,1);
% interleave input to second encoder
mi = x1(1,map);  x2 = rsc_encode(G,mi,0);
% parallel to serial multiplex to get the output vector
x = [];
if puncture==0 % unpunctured, rate = 1/3;
    for i=1:lu
        x = [x x1(1,i) x1(2,i) x2(2,i)];
    end
else % punctured into rate 1/2
    for i=1:lu
        if rem(i,2), x = [x x1(1,i) x1(2,i)]; % odd parity bits from RSC1
        else  x = [x x1(1,i) x2(2,i)]; % even parity bits from RSC2
        end
    end
end
x = 2*x - 1; % into bipolar format (+1/-1)
This can be expressed as

\[
P_{u}(u) = \frac{e^{(u+L(u))/2}}{1+e^{L(u)}} = \begin{cases} 
\frac{e^{L(u)}/(1+e^{L(u)})}{1/(1+e^{L(u)})} & \text{for } u=+1 \\
1/(1+e^{L(u)}) & \text{for } u=-1
\end{cases}
\] (9.4.35)

Also, we define the \textit{conditioned LLR}, which is used to detect the value of \(u\) based on the value of another random variable \(y\) affected by \(u\), as the \textit{LAPP} (Log A Posteriori Probability):

\[
L_{\text{app}}(u \mid y) = \ln \frac{P_{u}(u=+1|y)}{P_{u}(u=-1|y)} = \ln \frac{P(y|u=+1)P_{u}(u=+1) / P(y)}{P(y|u=-1)P_{u}(u=-1) / P(y)} = \ln \frac{P(y|u=+1)}{P(y|u=-1)} + \ln \frac{P_{u}(u=+1)}{P_{u}(u=-1)}
\] (9.4.36)

Now, let \(y\) be the output of a fading AWGN (additive white Gaussian noise) channel (with fading amplitude \(a\) and SNR per bit \(E_b/N_0\)) given \(u\) as the input. Then, this equation for the conditioned LLR can be written as

\[
L_{\text{app}}(u \mid y) = \ln \frac{\exp((-E_b/N_0)(y-a)^2)}{\exp((-E_b/N_0)(y+a)^2)} + \ln \frac{P_{u}(u=+1)}{P_{u}(u=-1)} = 4 a y \frac{E_b}{N_0} + L_{c} y + L_{u}(u)
\] (9.4.37)

with \(L_{c} = 4 a \frac{E_b}{N_0} \): the channel reliability

The objective of \textit{BCJR} (Bahl-Cocke-Jelinek-Raviv) \textit{MAP} (Maximum A posteriori Probability) \textit{algorithm} proposed in [B-1] is to detect the value of the \(k\)th message bit \(u_k\) depending on the sign of the following LAPP function:

\[
L_{\text{A}}(u_k) = \ln \frac{P_{u}(u_k=+1|y)}{P_{u}(u_k=-1|y)} = \ln \frac{\sum_{(s', s) \in S^+} p(s_k = s', s_{k+1} = s, y) / p(y)}{\sum_{(s', s) \in S^-} p(s_k = s', s_{k+1} = s, y) / p(y)}
\]

\[
= \ln \frac{\sum_{(s', s) \in S^+} p(s_k = s', s_{k+1} = s, y)}{\sum_{(s', s) \in S^-} p(s_k = s', s_{k+1} = s, y)}
\] (9.4.38)

with \(S^+ / S^-\): the set of all the encoder state transitions from \(s'\) to \(s\) caused by \(u_k = +1/1\)

Note that \(P\) and \(p\) denote the probability of a discrete-valued random variable and the probability density of a continuous-valued random variable. The numerator/denominator of this LAPP function are the sum of the probabilities that the channel output to \(u_k = +1/1\) will be \(y = \{y_{j<k}, y_k, y_{j>k}\}\) with the encoder state transition from \(s'\) to \(s\) where each joint probability density \(p(s', s, y) = p(s_k = s', s_{k+1} = s, y)\) can be written as

\[
p(s', s, y) = p(s_k = s', s_{k+1} = s, y) = p(s', y_{j<k}) p(s, y_k | s') p(y_{j>k} | s) = \alpha_{k-1}(s') \gamma_k(s', s) \beta_k(s)
\] (9.4.39)

where \(\alpha_{k-1}(s') = p(s', y_{j<k})\) is the probability that the state \(s_k\) at the \(k\)th depth level (stage) in the trellis is \(s'\) with the output sequence \(y_{j<k}\) generated before the \(k\)th level, \(\gamma_k(s', s) = p(s, y_k | s')\) is the probability that the state transition from \(s_k = s'\) to \(s_{k+1} = s\) is made with the output \(y_k\) generated, and \(\beta_k(s) = p(y_{j>k} | s)\) is the probability that the state \(s_{k+1}\) is \(s\) with the output sequence...
generated after the $k$th level. The first and third factors $\alpha_{k-1}(s')/\beta_k(s)$ can be computed in a forward/backward recursive way:

$$\alpha_k(s) = p(s, y_{j<k+1}) = p(s', y_{j<k}, s, y_k) = \sum_{s'\in S} p(s', y_{j<k}) p(s, y_k | s')$$

$$= \sum_{s'\in S} \alpha_{k-1}(s')\gamma_k(s', s) \quad \text{with} \quad \alpha_0(0) = 1, \quad \alpha_0(s) = 0 \quad \text{for} \quad s \neq 0 \quad (9.4.40)$$

$$\beta_{k-1}(s') = p(y_{j<k-1} | s') = p(s', y_k, s, y_{j<k} | s') = \sum_{s'\in S} p(s, y_k | s') p(y_{j<k} | s')$$

$$= \sum_{s'\in S} \gamma_k(s', s)\beta_k(s) \quad (9.4.41a)$$

with

$$\beta_k(s) = \begin{cases} \beta_k(0) = 1 & \text{and} \quad \beta_k(s) = 0 \quad \forall \ s \neq 0 \quad \text{if terminated at all-zero state} \\ \beta_k(s) = 1/N_s \quad \forall \ s \quad \text{otherwise} \end{cases} \quad (9.4.41b)$$

where $N_s = 2^{L-1}$ ($L$: the constraint length) is the number of states and $K$ is the number of decoder input symbols. The second factor $\gamma_k(s', s)$ can be found as

$$\gamma_k(s', s) = p(s, y_k | s') \quad (9.4.40) = \frac{p(s', s, y_k)}{p(s')} = \frac{p(s', s) p(y_k | s', s)}{p(s')}$$

$$= p(s_{k+1} | s_k') p(y_k | s_k', s_{k+1}) = p(u_k) p(y_k | u_k) = p(u_k) p(y_k^p, y_k^p | u_k, x_k^p(u_k))$$

$$\frac{e^{(u+1)L(u)/2}}{1+e^{L(u)}} \exp \left( \frac{E_b}{N_0} \left( y_k^p - a u_k \right)^2 - \frac{E_b}{N_0} \left( y_k^p - a x_k^p(u_k) \right)^2 \right)$$

$$= \frac{e^{(u+1)L(u)/2}}{1+e^{L(u)}} A_k \exp \left( 2a \frac{E_b}{N_0} (y_k^p u_k + y_k^p x_k^p(u_k)) \right) \quad \text{where} \quad u_k^2 = 1$$

$$= \frac{e^{(u+1)L(u)/2}}{1+e^{L(u)}} A_k \exp \left( \frac{1}{2} L_c \left[ y_k^p u_k + x_k^p(u_k) \right] \right) \quad (9.4.42)$$

with

$$A_k = \exp \left( \frac{-E_b}{N_0} \left( y_k^p + a^2 u_k^2 + y_k^p x_k^p(u_k) \right) \left( y_k^p + a^2 x_k^p(u_k) \right) \right)$$

$A_k$ is the exponent of the exponential terms are done by adding their exponents and that is why $A_k$ does not have to be computed since it will be substituted directly or via $\alpha_k$ (Eq. 9.4.40) or $\beta_k$ (Eq. 9.4.41)) into Eq. 9.4.39, then substituted into both the numerator and the denominator of Eq. 9.4.38, and finally cancelled.

The following MATLAB routine ‘logmap()’ corresponding to the block named ‘Log-MAP or SOVA’ in Fig. 9.15(c) uses these equations to compute the LAPP function (9.4.38). Note that in the routine, the multiplications of the exponential terms are done by adding their exponents and that is why $A_k$ is initialized to a large negative number as $-\text{Infty} = -100$ (corresponding to a nearly-zero $e^{-100} = 0$) under the assumption of initial all-zero state and for the termination of decoder 1 in all-zero state, respectively. (Q: Why is Beta initialized to $-\ln N_s (-\log(Ns))$ for non-termination of decoder 2?)
function $L_A = \text{logmap}(L_y, G, L_u, \text{ind\_dec})$
\% Copyright 1998, YuFei Wu, MPRG lab, Virginia Tech. for academic use
\% Log_MAP algorithm using straightforward method to compute branch cost
\% Input: $L_y = \text{scaled received bits}$ $L_y=0.5 \cdot L_c \cdot y = (2 \cdot a \cdot \text{rate} \cdot Eb/N0) \cdot y$
\% $G = \text{code generator for the RSC code a in binary matrix}$
\% $L_u = \text{extrinsic information from the previous decoder.}$
\% ind\_dec= index of decoder=1/2 (assumed to be terminated/open)
\% Output: $L_A = \ln \left( P(x=1|y)/P(x=-1|y) \right)$, i.e., Log-Likelihood Ratio
\% (soft-value) of estimated message input bit at each level
\% $lu=\text{length}(L_y)/2$; Infty=1e2; EPS=1e-50; \% Number of input bits, etc
\% $[N,L] = \text{size}(G)$;
\% $Ns = 2^{(L-1)}$; \% Number of states in the trellis
\% $Le1=-\log(1+\exp(L_u))$; $Le2=L_u+Le1$; \% $\ln(\exp((u+1)/2\cdot L_u)/(1+\exp(L_u)))$
\% Set up the trellis
\% Initialization of Alpha and Beta
\% $\text{Alpha}(1,2:Ns) = -\text{Infty}$; \% Eq.(9.4.40) (the initial all-zero state)
\% if ind\_dec=1 \% for decoder D1 with termination in all-zero state
\% $\text{Beta}(lu+1,2:Ns) = -\text{Infty}$; \% Eq.(9.4.41b) (the final all-zero state)
\% else \% for decoder D2 without termination
\% $\text{Beta}(lu+1,:) = -\log(Ns)\cdot\text{ones}(1,Ns)$;
\% Compute gamma at every depth level (stage)
\% $\text{for } k = 2:lu+1$
\% $Lyk = L_y(k*2-[3 2])$; $\text{gam}(:,:,k) = -\text{Infty}\cdot\text{ones}(Ns,Ns)$;
\% $\text{for } s2 = 1:Ns$ \% Eq.(9.4.42)
\% $\text{gam}(ps(s2,1),s2,k) = Lyk*[-1 pout(s2,2)].' + Le1(k-1);$\n\% $\text{gam}(ps(s2,2),s2,k) = Lyk*[+1 pout(s2,4)].' + Le2(k-1);$\n\% $\text{end}$
\% $\text{end}$
\% Compute Alpha in forward recursion
\% $\text{for } k = 2:lu$
\% $\text{for } s2 = 1:Ns$
\% $\text{alpha} = \text{sum}(\exp(\text{gam}(:,:,k).'+\text{Alpha}(k-1,:)))$; \% Eq.(9.4.40)
\% $\text{if } \text{alpha}<\text{EPS}, \text{Alpha}(k,s2)=-\text{Infty} \text{ else } \text{Alpha}(k,s2)=\log(\text{alpha}); \text{end}$
\% $\text{tempmax}(k) = \max(\text{Alpha}(k,:)); \text{Alpha}(k,:) = \text{Alpha}(k,:)-\text{tempmax}(k);$\n\% $\text{end}$
\% Compute Beta in backward recursion
\% $\text{for } k = lu:-1:2$
\% $\text{for } s1 = 1:Ns$
\% $\text{beta} = \text{sum}(\exp(\text{gam}(s1,:,k+1)+\text{Beta}(k+1,:)))$; \% Eq.(9.4.41)
\% $\text{if } \text{beta}<\text{EPS}, \text{Beta}(k,s1)=-\text{Infty} \text{ else } \text{Beta}(k,s1)=\log(\text{beta}); \text{end}$
\% $\text{Beta}(k,:) = \text{Beta}(k,:) - \text{tempmax}(k);$\n\% $\text{end}$
\% Compute the soft output LLR for the estimated message input
\% $\text{for } k = 1:lu$
\% $\text{for } s2 = 1:Ns$ \% Eq.(9.4.39)
\% $\text{temp1}(s2)=\exp(\text{gam}(ps(s2,1),s2,k+1)+\text{Alpha}(k,ps(s2,1))+\text{Beta}(k+1,s2));$
\% $\text{temp2}(s2)=\exp(\text{gam}(ps(s2,2),s2,k+1)+\text{Alpha}(k,ps(s2,2))+\text{Beta}(k+1,s2));$
\% $L_A(k) = \log(\text{sum}(\text{temp2})+\text{EPS}) - \log(\text{sum}(\text{temp1})+\text{EPS}); \% \text{Eq.(9.4.38)}$
function [nout,nstate,pout,pstate] = trellis(G)
% copyright 1998, Yufei Wu, MPRG lab, Virginia Tech for academic use
% set up the trellis with code generator G in binary matrix form.
% G: Generator matrix with feedback/feedforward connection in row 1/2
% e.g. G=[1 1 1; 1 0 1] for the turbo encoder in Fig. 9.15(a)
% nout(i,1:2): Next output [xs=m xp](-1/+1) for state=i, message in=0
% nout(i,3:4): next output [xs=m xp](-1/+1) for state=i, message in=1
% nstate(i,1): next state(1,...2^M) for state=i, message input=0
% nstate(i,2): next state(1,...2^M) for state=i, message input=1
% pout(i,1:2): previous out [xs=m xp](-1/+1) for state=i, message in=0
% pout(i,3:4): previous out [xs=m xp](-1/+1) for state=i, message in=1
% pstate(i,1): previous state having come to state i with message in=0
% pstate(i,2): previous state having come to state i with message in=1
% See Fig. 9.16 for the meanings of the output arguments.
% [N,L] = size(G); % Number of output bits and Constraint length
% M=L-1; % Number of bits per state and Number of states
% Set up next_out and next_state matrices for RSC code generator G
for state_i=1:Ns
    state_b = deci2bin1(state_i-1,M); % Binary state
    for input_bit=0:1
        d_k = input_bit;
        a_k = rem(G(1,:)*[d_k state_b]',2); % Feedback in Fig.9.15(a)
        out(input_bit+1,:) = [d_k rem(G(2,:)*[a_k state_b]',2)]; % Forward
        state(input_bit+1,:) = [a_k state_b(1:M-1)]; % Shift register
    end
    nout(state_i,:) = 2*[out(1,:) out(2,:)]-1; % bipolarize
    nstate(state_i,:) = [bin2deci(state(1,:)) bin2deci(state(2,:))]+1;
end
% Possible previous states having reached the present state
% with input_bit=0/1
for input_bit=0:1
    bN = input_bit*N; b1 = input_bit+1; % Number of output bits = 2;
    for state_i=1:Ns
        pstate(nstate(state_i,b1),b1) = state_i;
        pout(nstate(state_i,b1),bN+[1:N]) = nout(state_i,bN+[1:N]);
    end
end

Figure 9.16 A trellis built by the routine ‘trellis’ and its output arguments
<SOVA (Soft-In/Soft-Output Viterbi Algorithm) Decoding cast into ‘sova()’>

The objective of the SOVA-MAP decoding algorithm is to find the state sequence \( s^{(i)} \) and the corresponding input sequence \( u^{(i)} \) which maximizes the following MAP (maximum a posteriori probability) function:

\[
P(s^{(i)} | y) \propto p(y | s^{(i)}) \frac{P(s^{(i)})}{p(y)} \sim p(y | s^{(i)}) P(s^{(i)}) \text{ for given } y \quad (9.4.43)
\]

This probability would be found from the multiplications of the branch transition probabilities defined by Eq. (9.4.42). However, as is done in the routine ‘logmap()’, we will compute the path metric by accumulating the logarithm or exponent of only the terms affected by \( u^{(i)}_k \) as follows:

\[
M_k(s^{(i)}) = M_{k-1}(s^{(i)}) + \frac{L(u)}{2} u^{(i)}_k + \frac{1}{2} L_s[y^e_k, y^p_k] \left[ u^{(i)}_k \right] \quad (9.4.44)
\]

The decoding algorithm cast into the routine ‘sova(Ly,G,Lu,ind_dec)’ proceeds as follows:

(Step 0) Find the number of the \([y^e_k, y^p_k] 's\) in \( Ly \) given as the first input argument: \( l_a = \text{length}(Ly)/2 \). Find the number \( N \) of output bits of the two encoders and the constraint length \( L \) from the row and column dimensions of the generator matrix \( G \). Let the number of states be \( N_s = 2^L - 1 \), the SOVA window size \( \delta = 30 \), and the depth level \( k = 0 \). Under the assumption of all-zero state at the initial stage (depth level zero), initialize the path metric to \( M_k(s_0) = 0 = \ln 1 \) (corresponding to probability 1) only for the all-zero state \( s_0 \) and to \( M_k(s_j) = -\infty = \ln 0 \) (corresponding to probability 0) for the other states \( s_j \) (\( j \neq 0 \)).

(Step 1) Increment \( k \) by one and determine which one of the hypothetical encoder input (message) \( u_{k-1} = 0 \) or \( u_{k-1} = 1 \) would result in larger path metric \( M_k(s_i) \) (computed by Eq. (9.4.44)) for every state \( s_i (i = 0 : N_s - 1) \) at level \( k \) and chooses the corresponding path as the survivor path, storing the estimated value of \( u_{k-1} \) into ‘\( pinput(i,k) \)’ and the relative path metric difference ‘\( DM(i,k) \)’ of the survivor path over the other (non-surviving) path for every state at the stage. Repeat this step (in the forward direction) till \( u_{k} = 1 \).

(Step 2) Depending on the value of the fourth input argument ‘\( \text{ind\_dec} \)’, determine the all-zero state \( s_0 \) or any state belonging to the most likely path (with \( \text{Max} M_k(s_i) \)) to be the final state \( \hat{s}(k) \) (\( \text{sh}(k) \)).

(Step 3) Find \( \hat{u}(k) \) (\( \text{uhat}(k) \)) from ‘\( \text{pinput}(i,k) \)’ (constructed at Step 1) and the corresponding previous state \( \hat{s}(k-1) \) (\( \text{shat}(k-1) \)) from the trellis structure. Decrement \( k \) by one. Repeat this step (in the backward direction) till \( k = 0 \).

(Step 4) To find the reliability of \( \hat{u}(k) \), let LLR=\( \Delta M_k(\hat{s}(k)) \). Trace back the non-surviving paths from the optimal states \( \hat{s}(k+i) \) (for \( i = 1 : \delta \) such that \( k+i < l_u \)), find the nearly-optimal input \( \hat{u}_i(k) \). If \( \hat{u}_i(k) \neq \hat{u}(k) \) for some \( i \), let LLR=\( \text{Min}\{ \text{LLR,} \Delta M_{k+i}(\hat{s}(k+i)) \} \). In this way, find the LLR estimate and multiply it with the bipolarized value of \( \hat{u}(k) \) to determine the soft output or L-value:

\[
L_{\delta}(\hat{u}(k)) = (2\hat{u}(k) - 1) \text{LLR} \quad (9.4.46)
\]
% This implements Soft Output Viterbi Algorithm in trace back mode
% Input: Ly : Scaled received bits Ly=0.5*L_c*y=(2*a*rate*Eb/N0)*y
%        G   : Code generator for the RSC code in binary matrix form
%        Lu  : Extrinsic information from the previous decoder.
%        ind_dec: Index of decoder=1/2
%                  (assumed to be terminated in all-zero state/open)
% Output: L_A : Log-Likelihood Ratio (soft-value) of
%                  estimated message input bit u(k) at each stage,
%                  ln (P(u(k)=1|y)/P(u(k)=-1|y))

function L_A = sova(Ly,G,Lu,ind_dec)

lu = length(Ly)/2; % Number of y=[ys yp] in Ly
lu1 = lu+1; Infty = 1e2;

[N,L] = size(G); Ns = 2^(L-1); % Number of states
delta = 30; % SOVA window size

% Make decision after 'delta' delay. Tracing back from (k+delta) to k,
% decide bit k when received bits for bit (k+delta) are processed.
% Set up the trellis defined by G.
[nout,ns,pout,ps] = trellis(G);
% Initialize the path metrics to -Infty
Mk(1:Ns,1:lu1)=-Infty; Mk(1,1)=0; % Only initial all-0 state possible

for k=1:lu
    Lyk = Ly(k*2-[1 0]); k1=k+1;
    for s=1:Ns % Eq.(9.4.44), Eq.(9.4.45)
        Mk0 = Lyk*pout(s,1:2).' -Lu(k)/2 +Mk(ps(s,1),k);
        Mk1 = Lyk*pout(s,3:4).' +Lu(k)/2 +Mk(ps(s,2),k);
        if Mk0>Mk1, Mk(s,k1)=Mk0; DM(s,k1)=Mk0-Mk1; pinput(s,k1)=0;
        else     Mk(s,k1)=Mk1; DM(s,k1)=Mk1-Mk0; pinput(s,k1)=1;
        end
    end
end

% Trace back from all-zero state or the most likely state for D1/D2
% to get input estimates uhat(k), and the most likely path (state) shat
if ind_dec==1, shat(lu1)=1; else [Max,shat(lu1)]=max(Mk(:,lu1)); end

for k=lu:-1:1
    uhat(k)=pinput(shat(k+1),k+1); shat(k)=ps(shat(k+1),uhat(k)+1);
end

% As the soft-output, find the minimum DM over a competing path
% with different information bit estimate.
for k=1:lu
    LLR = min(Infty,DM(shat(k+1),k+1));
    for i=1:delta
        if k+i<lu1
            u_ = 1-uhat(k+i); % the information bit
            tmp_state = ps(shat(k+i+1),u_+1);
            for j=1:i-1:0
                pu=pinput(tmp_state,k+j+1); tmp_state=ps(tmp_state,pu+1);
            end
            if pu==uhat(k), LLR = min(LLR,DM(shat(k+i+1),k+i+1)); end
        end
    end
    L_A(k) = (2*uhat(k)-1)*LLR; % Eq.(9.4.46)
end
%turbo_code_demo.m
% simulates the classical turbo encoding-decoding system.
% 1st encoder is terminated with tails bits. (1m+M) bits are scrambled % and passed to 2nd encoder, which is left open without termination.
clear
dec_alg = 1; % 0/1 for Log-MAP/SOVA
puncture = 1; % puncture or not
rate = 1/(3-puncture); % Code rate
lu = 1000; % Frame size
Nframes = 100; % Number of frames
Niter = 4; % Number of iterations
EbN0dBs = 2.6; %[1 2 3];
N_EbN0dBs = length(EbN0dBs);
G = [1 1 1; 1 0 1]; % Code generator
a = 1; % Fading amplitude; a=1 in AWGN channel
[N,L] = size(G); M=L-1; 1m=lu-M; % Length of message bit sequence
for nENDB = 1:N_EbN0dBs
EbN0 = 10^(EbN0dBs(nENDB)/10); % convert Eb/N0[db] to normal number
L_c = 4*a*EbN0*rate; % reliability value of the channel
sigma = 1/sqrt(2*rate*EbN0); % standard deviation of AWGN noise
noes(nENDB,:) = zeros(1,Niter);
for nframe = 1:Nframes
m = round(rand(1,lm)); % information message bits
[temp,map] = sort(rand(1,lu)); % random interleaver mapping
x = encoderm(m,G,map,puncture); % encoder output x(+1/-1)
noise = sigma*randn(1,lu*(3-puncture));
r = a.*x + noise; % received bits
y = demultiplex(r,map,puncture); % input for decoder 1 and 2
Ly = 0.5*L_c*y; % Scale the received bits
for iter = 1:Niter
% Decoder 1
if iter<2, Lu1=zeros(1,lu); % Initialize extrinsic information
else Lu1(map)=L_e2; % (deinterleaved) a priori information
end
if dec_alg==0, L_A1=logmap(Ly(1,:),G,Lu1,1); % all information
else L_A1=sova(Ly(1,:),G,Lu1,1); % all information
end
L_e1= L_A1-2*Ly(1,1:2:2*lu)-Lu1; % Eq.(9.4.47)
% Decoder 2
Lu2 = L_e1(map); % (interleaved) a priori information
if dec_alg==0, L_A2=logmap(Ly(2,:),G,Lu2,2); % all information
else L_A2=sova(Ly(2,:),G,Lu2,2); % all information
end
L_e2= L_A2-2*Ly(2,1:2:2*lu)-Lu2; % Eq.(9.4.47)
mhat(map)=(sign(L_A2)+1)/2; % Estimate the message bits
noe(iter)=sum(mhat(1:lu-M)==m); % Number of bit errors
end % End of iter loop
% Total number of bit errors for all iterations
noes(nENDB,:)=noes(nENDB,:)+noe;
ber(nENDB,:)=noes(nENDB,:)/nframe/(lu-M); % Bit error rate
fprintf(’\n’);
for i=1:Niter, fprintf(’%14.4e ‘, ber(nENDB,i)); end
end % End of nframe loop
end % End of nENDB loop
Now, it is time to take a look at the main program “turbo_code_demo.m”, which uses the routine ‘logmap()’ or ‘sova()’ (corresponding to the block named ‘Log-MAP or SOVA’ in Fig. 9.15(c)) as well as the routines ‘encoderm()’ (corresponding to Fig. 9.15(a)), ‘rsc_encode()’, ‘demultiplex()’ (corresponding to Fig. 9.15(b)), and ‘trellis()’ to simulate the turbo coding system depicted in Fig. 9.15. All of the programs listed here in connection with turbo coding stem from the routines developed by Yufei Wu in the MPRG (Mobile/Portable Radio Research Group) of Virginia Tech. (Polytechnic Institute and State University). The following should be noted:

- One thing to note is that the extrinsic information \( L_e \) to be presented to one decoder \( i \) by the other decoder \( j \) should contain only the intrinsic information of decoder \( j \) that is obtained from its own parity bits not available to decoder \( i \). Accordingly, one decoder should remove the information about \( y^t \) (available commonly to both decoders) and the priori information \( L(u) \) (provided by the other decoder) from the overall information \( L_A \) to produce the information that will be presented to the other decoder. (Would your friend be glad if you gave his/her present back to him/her or presented him/her what he/she had already got?) To prepare an equation for this information processing job of each encoder, we extract only the terms affected by \( u_k = \pm 1 \) from Eqs. (9.4.44) and (9.4.42) (each providing the basis for the path metric (Eq. (9.4.45)) and LLR (Eq. (9.4.38)), respectively,) to write

\[
\left( \frac{L(u)}{2} u_k^{(i)} + \frac{1}{2} L_c y_k^{(i)} \right)_{u_k^{(i)} = +1} - \left( \frac{L(u)}{2} u_k^{(i)} + \frac{1}{2} L_c y_k^{(i)} \right)_{u_k^{(i)} = -1} = L(u) + L_c y_k^t
\]

which conforms with Eq. (9.4.37) for the conditioned LLR \( L_{uy}(u \mid y) \). To prepare the extrinsic information for the other decoder, this information should be removed from the overall information \( L_A(u) \) produced by the the routine ‘logmap()’ or ‘sova()’ as

\[
L_e(u) = L_A(u) - L(u) - L_c y_k^t
\]  

(9.4.47)

- Another thing to note is that as shown in Fig. 9.15(c), the basis for the final decision about \( u \) is the deinterleaved overall information \( L_{A_2} \) that is attributed to decoder 2. Accordingly, the turbo decoder should know the pseudo-random sequence ‘map’ (that has been used for interleaving by the transmitter) as well as the fading amplitude and SNR of the channel.

- The trellis structure and the output arguments produced by the routine ‘trellis()’ are illustrated in Fig. 9.16.

Interest readers are invited to run the program “turbo_code_demo.m” with the value of the control constant ‘dec_alg’ set to 0/1 for Log-MAP/SOVA decoding algorithm and see the BER becoming lower as the decoding iteration proceeds. How do the turbo codes work? How are the two decoding algorithms, Log-MAP and SOVA, compared? Is there any weakpoint of turbo codes? What is the measure against the weakpoint, if any? Unfortunately, to answer such questions is difficult for the authors and therefore, is beyond the scope of this book. As can be seen from the simulation results, turbo codes have an excellent BER performance close to the Shannon limit at low and medium SNRs. However, the decreasing rate of the BER curve of a turbo code can be very low at high SNR depending on the interleaver and the free distance of the code, which is called the ‘error floor’ phenomenon. Besides, turbo codes needs not only a large interleaver and block size but also many iterations to achieve such a good BER performance, which increases the complexity and latency (delay) of the decoder.
(d) Why don’t we make the code for the above set of four symbols as follows?

<table>
<thead>
<tr>
<th>Symbol</th>
<th>00</th>
<th>01</th>
<th>10</th>
<th>11</th>
</tr>
</thead>
<tbody>
<tr>
<td>Codeword</td>
<td>0</td>
<td>1</td>
<td>10</td>
<td>11</td>
</tr>
</tbody>
</table>

With this code, we could have the average codeword length as

\[ L = 0.81 \times 1 + 0.09 \times 1 + 0.09 \times 2 + 0.01 \times 2 = 1.1 \text{[bits/symbol]} \]

which is shorter than that with the code obtained in (a) or (b). To answer this question, think about the decoding of a coded sequence {1011} based on the above table.

9.4 Channel Capacity and Hamming Codes

Recall that the MATLAB function ‘Hammgen(n)’ can be used to generate the \( n \times N \) parity-check matrix for an \((N,K)\) Hamming code where \( N = 2^n - 1 \) and \( K = N - n \). Suppose the message sequence is coded with the \((7,4)\) or \((15,11)\) Hamming code, BPSK(Binary Phase Keying)-modulated, and then transmitted through a BSC (Binary Symmetric Channel).

(a) Compute the crossover (channel bit transmission error) probability of BPSK signaling with each of these two codings for \( \text{SNRdB}=0\text{--}13 \text{[dB]} \) by

\[ \varepsilon^{(7.3.5)} = Q\left(\sqrt{\frac{2E_b}{N_0}}R_c\right) \text{ with } R_c = \frac{K}{N} \]  \hspace{1cm} (P9.4.1)

and substitute this into Eq. (9.3.13) to get the channel capacity

\[ C_t^{(9.3.13)} = 1 + (1 - \varepsilon) \log_2 (1 - \varepsilon) + \varepsilon \log_2 \varepsilon = 1 - H_\varepsilon(\varepsilon) \text{[bits/symbol]} \]  \hspace{1cm} (P9.4.2)

Plot the channel capacity \( C_t \) vs. \( \text{SNRdB}=0\text{--}13 \) and use the graph to find the rough value of \( \text{SNRdB} \) at which the inequality (9.3.8) is marginally satisfied:

\[ b_c = R_c \cdot b_{\text{BPSK}} \leq C_t \text{[bits/symbol]} \] \hspace{1cm} (P9.4.3)

(b) Referring to the lower part of the MATLAB routine ‘do_Hamming_code74()’ (Sec. 9.4.2), use Eq. (9.4.11) to compute the theoretical (approximate) probabilities of message bit error for \( \text{SNRdB}=0\text{--}13 \text{[dB]} \) with no code, \((7,4)\) Hamming code, and \((15,11)\) Hamming code and plot them versus \( \text{SNRdB} \). Use the theoretical BER curves to find the rough values of \( \text{SNRdB} \) at which the two codings have the same BER as that with no coding.

(c) Modify the MATLAB routine ‘do_Hamming_code74()’ so that it can simulate the BPSK signaling with \((7,4)\) Hamming code and \((15,11)\) Hamming code for \( \text{SNRdB}=5, 7, 9, \text{and} 11 \text{[dB]} \). Run it to plot the BERs versus \( \text{SNRdB} \) on the theoretical BER curves obtained in (b).

Table P9.4  BERs of BPSK signaling with \((7,4)\) or \((15,11)\) Hamming code

<table>
<thead>
<tr>
<th></th>
<th>( \text{SNRdB}=5\text{[dB]} )</th>
<th>( \text{SNRdB}=7\text{[dB]} )</th>
<th>( \text{SNRdB}=9\text{[dB]} )</th>
<th>( \text{SNRdB}=11\text{[dB]} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>No code</td>
<td>0.037679</td>
<td>0.012422</td>
<td>0.002413</td>
<td>0.000118</td>
</tr>
<tr>
<td>((7,4)) Hamming code</td>
<td>0.012422 (0.010996)</td>
<td>0.002413</td>
<td>0.000118 (0.000080)</td>
<td></td>
</tr>
<tr>
<td>((15,11)) Hamming code</td>
<td>0.060606 (0.038549)</td>
<td>0.001196 (0.000830)</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
9.5 Effect of Coding on BER

To understand why the BER of simulation result is much higher than the theoretical value of Eq. (9.4.11) (Sec. 9.4.2), insert the following statements at the appropriate places in the MATLAB routine ‘do_Hamming_code74()’ (Sec. 9.4.2).

```matlab
notbe1 = sum(r_sliced ~= coded); notbec1 = sum(r_c ~= coded);
si = bin2deci(s); % syndrome
if notbe1 > 0 & iter < 30
    fprintf('
 # of transmitted bit errors %d -> ', notbe1);
    fprintf('%d after correction (syndrome= %d)', notbec1, si);
end
```

Then run the routine with SNRbdB=5 to answer the following questions:
(i) Is a single bit error in a codeword always corrected?
(ii) Is there any case where the number of bit errors is increased by a wrong correction?

What do you think is the reason for the big gap between the two BERs that are obtained from simulation and theoretical formula (9.4.11)?

9.6 BCH (Bose-Chaudhuri-Hocquenghem) Codes with Simulink

BCH codes constitute a powerful class of cyclic codes that provides a large selection of block lengths, code rates, alphabet sizes, and error-correcting capability as listed in Table 5.2 of [S-3]. According to the table, the \((N, K) = (15, 7)\) and \((31, 16)\) BCH codes are represented by the generator polynomial coefficient vectors

\[
g_1 = 721 (\text{octal}) = [1 1 1 0 1 0 0 0 1] \quad (P9.6.1)
\]

\[
g_2 = 107657 (\text{octal}) = [1 0 0 0 1 1 1 1 0 1 0 1 1 1 1] \quad (P9.6.2)
\]

and their error correcting capabilities are 2 and 3, respectively. Suppose the message sequence is coded with these two BCH codes, BPSK(binary phase-shift keying)-modulated, and then transmitted through a BSC (binary symmetric channel).

(a) Let us try to make use of the MATLAB built-in function ‘bchgenpoly()’ to make the generator polynomial vectors for the \((31, 16)\) BCH code:

```matlab
N=31; K=16;
gBCH=bchgenpoly(N,K); % Galois row vector in GF(2) representing
% the (N,K) BCH code
g=double(gBCH.x) % Extracting the elements from a Galois array
```

Does the resulting vector conform with \(g_2\)?

(b) It may take a long time to use the MATLAB built-in routine ‘decode()’ with such a big BCH code as \(g_2\). If we prepare the syndrome table using ‘E=syndtable(H)’ beforehand and then use ‘decode(coded,N,K,’cyclic’,g,E)’ with the prepared syndrome table E, it will save a lot of decoding time. Likewise, if we prepare the error pattern matrix E and the corresponding error pattern index vector epi using ‘cyclic_decoder0’ (Sec. 9.4.3) beforehand and then use ‘cyclic_decoder(coded,N,K,g,E,epi)’, it will save the decoding time, too. Complete the following MATLAB program “do_cyclic_codes.m” so that it can simulate a BPSK communication system with each of the above two BCH codes and plot the BERs for SNRdBs=[2 4 10] together with the theoretical BER curves for SNRdBs=[0:0.01:14] obtained from Eq. (9.4.11). Run the completed program to get the BER curves.
%do_cyclic_codes.m
clear, clf
gsymbols=['bo';'r+';'kx';'md';'g*'];
% Theoretical BER curves
SNRdBs=0:0.01:14; SNRs=10.^((SNRdBs/10)); pemb_uncoded=Q(sqrt(SNRs));
semilogy(SNRdBs,pemb_uncoded,'-', hold on
use_decode = 0; % Use encode()/decode or not
MaxIter=1e5; Target_no_of_error=100; SNRdBs1=2:4:10;
for iter=1:2
  if iter==1, N=15; K=7; nceb=2; % (15,7) BCH code generator
    else    N=31; K=16; nceb=3; % (31,16) BCH code generator
  end
  gBCH=bchgenpoly(N,K); % Galois row vector representing (N,K) BCH code
  g=double(gBCH.x);
  clear('S')
  if use_decode>0, H=cyclgen(N,g); E=syndtable(?);
    else
    E= combis(N,nceb); % All error patterns
    for i=1:size(E,1)
        S(i,:) = cyclic_decoder0(E(i,:),N,K,g); % Syndrome
        epi(bin2deci(S(i,:))) = ?; % Error pattern indices
    end
  end
  Rc = K/N; % code rate
  SNRcs=SNRs*Rc; sqrtSNRcs=sqrt(SNRcs);
  ets=Q(sqrt(SNRcs)); % transmitted bit error probability Eq.(7.3.5)
  pemb_t=prob_err_msg_bit(ets,N,nceb);
  semilogy(SNRdBs,pemb_t,gsymbols(iter,1),'Markersize',5)
  for iter1=1:length(SNRdBs1)
    SNRdB = SNRdBs1(iter1); SNR = 10.^((SNRdB/10));
    SNRc = SNR*Rc; sqrtSNRc = sqrt(SNRc);
    et=Q(sqrt(SNRc)); % transmitted bit error probability Eq.(7.3.5)
    K100 = K*100;
    nombe = 0;
    for iter2=1:MaxIter
      msg=randint(1,K100); % Message vector
      if use_decode==0, coded=cyclic_????????(msg,N,K,g);
        else  msg=reshape(msg,length(msg)/K,K);
          coded=????????(msg,N,K,'cyclic',g);
        end
      r= 2*coded-1 + randn(size(coded))/sqrtSNRc; % Received vector
      r_sliced= 1*(r>0); % Sliced
      if use_decode==0
        decoded= cyclic_????????(r_sliced,N,K,g,E,epi);
      else  decoded= ??????(r_sliced,N,K,'cyclic',g,E);
      end
      nombe= nombe + sum(sum(decoded~=msg));
      if nombe>Target_no_of_error, break; end
    end
    lm=iter2*K100; pemb=nombe/lm; % Message bit error probability
    semilogy(SNRdB,pemb,gsymbols(iter,:),'Markersize',5)
  end
end
(c) Referring to the Simulink model “BCH_BPSK_sim.mdl” depicted in Fig. P9.6 and the following program “do_BCH_BPSK_sim.m”, use Simulink to simulate a BPSK communication system with the (31,16) BCH code for error correction and find the BERs for 0/3/7dB. Noting that $\text{SNR}_{bdB}=\text{SNR}_{E/N0}+3$ since $0/(2)$ $\text{SNR}_{E/N0}$, list the BERs together with those obtained using 'encode()'/'decode()' in (b).

```matlab
%do_BCH_BPSK_sim.m
clear, clf
K=16; % Number of input bits to the BCH encoder (message length)
N=31; % Number of output bits from the BCH encoder (codeword length)
Rc=K/N; % Code rate to be multiplied with the SNR in AWGN channel block
b=1; M=2^b; % Number of bits per symbol and modulation order
T=0.001/K; Ts=b*T; % Sample time and Symbol time
SNRbdBs=[2:4:10]; EbN0dBs=SNRbdBs-3;
EbN0dBs_t=0:0.1:10; EbN0s_t=10.^(EbN0dBs_t/10);
SNRbdBs_t=EbN0dBs_t+3;
BER_theory= prob_error(SNRbdBs_t,'PSK',b,'BER');
for i=1:length(EbN0dBs)
    EbN0dB=EbN0dBs(i);
    sim('BCH_BPSK_sim');
    BERs(i)=BER(1); % just ber among {ber, # of errors, total # of bits}
    fprintf(' With EbN0dB=%4.1f, BER=%10.4e=%d/%d
', EbN0dB,BER);
end
semilogy(EbN0dBs,BERs,'r*', EbN0dBs_t,BER_theory,'b')
xlabel('Eb/N0[dB]'); ylabel('BER');
title('BER of BCH code with BPSK');
```
% To practice using convenc() and vitdec() for channel coding

clear, clf

Gc=[4 5 11;1 4 2]; % Octal code generator matrix
K=size(Gc,1); % Number of encoder input bits

% Constraint length vector
Gc_m=max(Gc.);
for i=1:length(Gc_m), Lc(i)=length(deci2bin1(oct2dec(Gc_m(i)))); end
trel=poly2trellis(Lc,Gc);
Tbdepth=sum(Lc)*5; delay=Tbdepth*K;
lm=1e5; msg=randint(1,lm);
transmission_ber=0.02;
notbe=round(transmission_ber*lm); % Number of transmitted bit errors
ch_input=convenc([msg zeros(1,delay)],[trel]);
% Received/modulated/detected signal
ch_output=rem(randerr(1,length(ch_input),notbe)+ch_input,2);
decoded_trunc= vitdec(ch_output,trel,Tbdepth,'trunc','hard');
ber_trunc= sum(msg==decoded_trunc)/lm;
decoded_cont= vitdec(ch_output,trel,Tbdepth,'cont','hard');
ber_cont=sum(msg==decoded_cont)/lm;
% It is indispensable to use the delay for the decoding result
% obtained using vitdec(...,'cont',)
nn=[0:100-1];
subplot(221), stem(nn,msg(nn+1)), title('Message sequence')
subplot(223), stem(nn,decoded_cont(nn+1)), hold on
stem(delay,0,'r')
decoded_term= vitdec(ch_output,trel,Tbdepth,'term','hard');
ber_term=sum(msg==decoded_term)/lm;
fprintf('
 BER_trunc  BER_cont   BER_term')
fprintf('
 %9.2e  %9.2e  %9.2e
', ber_trunc,ber_cont,ber_term)
326  Chapter 9  Information and Coding

9.7 A Convolutional Code and Viterbi Decoding

Fig. P9.7 shows a convolutional encoder (described by two kinds of schematic) and the two corresponding generator matrices where one is a binary generator matrix that is used by the ‘conv_encoder()’/’conv_decoder()’ and the other is an octal generator matrix that is used by the MATLAB built-in functions ‘convenc()’/’vitdec()’. Let the convolutional encoder and the corresponding Viterbi decoder be used for channel coding where there happens a 2%-error in the 100,000 transmitted bits. Complete the above program “dc09p07.m” to simulate this situation and run it to find the BER.

9.8 A Convolutional Code and Viterbi Decoding with Simulink

Fig. P9.8.1 shows a convolutional encoder described by the octal code generator matrix $G_c=[133 171]$. Fig. P9.8.2 shows the Simulink model “Viterbi_QAM_sim.mdl” that can be used to simulate a QAM communication with a given convolutional encoder and the corresponding Viterbi decoder. The following MATLAB function ‘Viterbi_QAM()’ can also be used to simulate a QAM communication with a given convolutional encoder and the corresponding Viterbi decoder. Complete the following MATLAB program “do_Viterbi_QAM”, which uses the MATLAB function ‘Viterbi_QAM()’ and the Simulink model “Viterbi_QAM_sim.mdl” to simulate a 16-QAM communication system with the convolutional encoder of Fig. P9.8.1 and the corresponding Viterbi decoder. Run it to find the BERs for EbN0=3, 6, and 9dB.

<table>
<thead>
<tr>
<th>EbN0dB=3[db]</th>
<th>EbN0dB=6[db]</th>
<th>EbN0dB=9[db]</th>
</tr>
</thead>
<tbody>
<tr>
<td>MATLAB ‘Viterbi_QAM.m’</td>
<td>0.006684</td>
<td>0.000079</td>
</tr>
<tr>
<td>Simulink ‘Viterbi_QAM_sim.mdl’</td>
<td>0.466403</td>
<td>0.000079</td>
</tr>
</tbody>
</table>

function [pemb,nombe,notmb]=Viterbi_QAM(Gc,b,SNRbdB,MaxIter)
if nargin<4, MaxIter=1e5; end
if nargin<3, SNRbdB=5; end
if nargin<2, b=4; end
[K,N]=size(Gc); Rc=K/N; Gc_m=max(Gc.');
% Constraint length vector
for i=1:length(Gc_m), Lc(i)=length(deci2bin1(oct2dec(Gc_m(i)))); end
NF=144; % Number of bits per frame
Nmod=NF*N/K/b; % Number of QAM symbols per modulated frame
SNRb=10.^((SNRbdB/10)); SNRbc=SNRb*Rc; sqrtSNRbc=sqrt(SNRbc);
sqrtSNRc=sqrt(2*b*SNRbc); % Complex noise for b-bit (coded) symbol
trel=poly2trellis(Lc,Gc);
Tbdepth=5; delay=Tbdepth*K;
nombe=0; Target_no_of_error=100;
for iter=1:MaxIter
    msg=randint(1,Nf); % Message vector
coded= convenc(msg,trel); % Convolutional encoding
    modulated= QAM(coded,b); % 2^b-QAM-Modulation
    r= modulated +(randn(1,Nmod)+j*randn(1,Nmod))/sqrtSNRc;
    demodulated= QAM_dem(r,b); % 2^b-QAM-Demodulation
decoded= vitdec(demodulated,trel,Tbdepth,'trunc','hard');
    nombe = nombe + sum(msg==decoded(1:Nf)); % Message bit error probability
    if nombe>Target_no_of_error, break; end
end
notmb=NF*iter; % Number of total message bits
pemb=nombe/notmb; % Message bit error probability
Problems

Figure P.9.8.1 A convolutional encoder and its generator matrix

![Diagram of a convolutional encoder and its generator matrix]

Model Information

NH=144, TH=0.001, Tc=7/Tc, Tc=Tc/Tc, % Sample time and Symbol time
G=[133, 171], Go=[133, 171], Go少了327的8位移位
% Constraint length vector
% Constraint length vector
v=[length(Go, 1), length(Go, 2), length(Go, 3)]

Figure P.9.8.2 Simulink model "Viterbi_QAM_sim.mdl"
clear, clf
Nf=144; Tf=0.001; Tb=Tf/Nf; \% Frame size, Frame and Sample/Bit time
Gc=[133 171];
[K,N]=size(Gc); Rc=K/N; \% Message/Codeword length and Code rate
\% Constraint length vector
Gc_m=max(Gc.');
for i=1:length(Gc_m)
    Lc(i)=length(deci2bin1(oct2dec(Gc_m(i))));
end
Tbdepth=sum(Lc)*5; delay=Tbdepth*K;
b=4; M=2^b; \% Number of bits per symbol and Modulation order
Ts=b*Rc*Tb; \% Symbol time
N_factor=sqrt(2*(M-1)/3); \% Eq.(7.5.4a)
EbN0dBs_t=0:0.1:10; SNRbdBs_t=EbN0dBs_t+3;
BER_theory= prob_error(SNRbdBs_t,'QAM',b,'BER');
EbN0dBs=[3 6]; Target_no_of_error=50;
for i=1:length(EbN0dBs)
    EbN0dB=EbN0dBs(i); SNRbdB=EbN0dB+3;
    randn('state', 0);
    \[pemb,nombe,notmb\]=???????_QAM(Gc,b,SNRbdB,Target_no_of_error);
    pembs(i)=pemb;
    sim('Viterbi_QAM_sim'); pembs_sim(i)=BER(1);
end
[pembs; pembs_sim]
semilogy(EbN0dBs,pembs,'r*', EbN0dBs_t,BER_theory,'b')
xlabel('Eb/N0[dB]'); ylabel('BER');

function qamseq=QAM(bitseq,b)
    bpsym = nextpow2(max(bitseq)); \% no of bits per symbol
    if bpsym>0, bitseq = deci2bin(bitseq,bpsym); end
    if b==1, qamseq=bitseq*2-1; return; end \% BPSK modulation
    \% 2^b-QAM modulation
    N0=length(bitseq); N=ceil(N0/b);
    bitseq=bitseq(:); bitseq=[bitseq zeros(1,N*b-N0)];
    b1=ceil(b/2); b2=b-b1; b21=2^b1; b22=2^b2;
    g_code1=2*gray_code(b1)-b12+1; g_code2=2*gray_code(b2)-b22+1;
    tmp1=sum([1:2:2^b1-1).^2)*b21; tmp2=sum([1:2:2^b2-1).^2)*b22;
    M=2^b; Kmod=sqrt(tmp1+tmp2)/2/(2^b/4) \% Normalization factor
    qamseq=[];
    for i=0:N-1
        bi=b*i; i_real=bin2deci(bitseq(bi+[1:b1]))+1;
        i_imag=bin2deci(bitseq(bi+[b1+1:b]))+1;
        qamseq=[qamseq (g_code1(i_real)+j*g_code2(i_imag))/Kmod];
    end

function [g_code,b_code]=gray_code(b)
    N=2^b; g_code=0:N-1;
    if b>1, g_code=gray_code0(g_code); end
    b_code=deci2bin(g_code);

function g_code=gray_code0(g_code)
    N=length(g_code); N2=N/2;
    if N=4, g_code=fftshift(g_code(N2+1:N)); end
    if N=4, g_code=gray_code0(g_code(1:N2))
    gray_code0(g_code(N2+1:N)); end
%BPSK demodulation
if b==1, bitseq=(qamseq>=0); return; end
%2^b-QAM demodulation
N=length(qamseq);
b1=ceil(b/2); b2=b-b1;
g_code1=2*gray_code(b1)-2^b1+1; g_code2=2*gray_code(b2)-2^b2+1;
tmp1=sum([1:2:2^b1-1].^2)*2^b2;
tmp2=sum([1:2:2^b2-1].^2)*2^b1;
Kmod=sqrt((tmp1+tmp2)/2/(2^b/4)); % Normalization factor
g_code1=g_code1/Kmod; g_code2=g_code2/Kmod;
bitseq=[];
for i=1:N
    [emin1,i1]=min(abs(real(qamseq(i))-g_code1));
    [emin2,i2]=min(abs(imag(qamseq(i))-g_code2));
    bitseq=[bitseq deci2bin1(i1-1,b1) deci2bin1(i2-1,b2)];
end
if (nargin>2)
    N = length(bitseq)/bpsym; bitmatrix = reshape(bitseq,bpsym,N).';
    for i=1:N, intseq(i)=bin2deci(bitmatrix(i,:)); end
    bitseq = intseq;
end

function [pemb,nombe,notmb]=...
    TCM(state_eq,K,Nsb,N,Constellation,SNRdB,TARGET_NO_OF_ERROR)
M=2^N; Rc=K/N;
NF=288; % Number of bits per frame
NMOD=NF/K; % Number of symbols per modulated frame
SNRb=10.^(SNRdB/10); SNRbc=SNRb*Rc;
sqrtSNRc=sqrt(2*N*SNRbc);% Complex noise per K=Rc*N-bit symbol
nombe=0; MaxIter=1e6;
for iter=1:MaxIter
    msg=randint(1,NF); % Message vector
    coded = TCM_encoder(state_eq,K,Nsb,N,msg,Constellation);
    r= coded +(randn(1,NMOD)+j*randn(1,NMOD))/sqrtSNRc;
    decoded= TCM_decoder(state_eq,K,Nsb,r,Constellation);
    nombe = nombe + sum(msg==decoded(1:NF));
    if nombe>TARGET_NO_OF_ERROR, break; end
end
notmb=NF*iter; % Number of total message bits
pemb=nombe/notmb; % Message bit error probability
function [s1,x]=TCM_state_eq1(s,u,Constellation)
    % State equation for the TCM encoder in Fig. P9.9.1(a)
    % Input:  s= State, u= Input
    % Output: s1= Next state, x= Output
    s1 = [u(2) s(1)];
    x= Constellation(bin2deci([u(1) rem(u(2)+s(2),2) s(1)])+1);
(a) Run the following program “do_TCM_8PSK.m” to use the above MATLAB routine “TCM()” and the Simulink model “TCM_sim.mdl” for finding the BERs of the TCM system with the encoder-modulator of Fig. P9.9.1(a) where the SNR is given as $E_b/N_0 = 0.2 \text{ [dB]}$. Set the 8PSK constellation for “TCM_sim.mdl” to 1 or 3 (Gray code).

What difference will be made to the BER? Set the 8PSK constellation for “TCM.m” to 2 or 3. What difference will be made to the BER? Fill in the following table.
Table P9.9  BER of TCM systems

<table>
<thead>
<tr>
<th>$E_b/N_0$</th>
<th>QPSK - no code (Theoretical)</th>
<th>BER</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td></td>
<td>TCM-8PSK (Fig. P9.9.1(a))</td>
</tr>
<tr>
<td>2dB</td>
<td>0.038</td>
<td>0.0018</td>
</tr>
<tr>
<td></td>
<td>TCM with $C_1$</td>
<td>0.059</td>
</tr>
<tr>
<td></td>
<td>TCM with $C_2$</td>
<td>0.018</td>
</tr>
<tr>
<td></td>
<td>TCM with $C_3$</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Simulink with $C_2$</td>
<td>0.020</td>
</tr>
<tr>
<td></td>
<td>Simulink with $C_1$</td>
<td>0.064</td>
</tr>
<tr>
<td></td>
<td>Simulink with $C_3$</td>
<td>0.049</td>
</tr>
</tbody>
</table>

(b) Modify the program “do_TCM_8PSK.m” and make a routine “TCM_state_eq2()” so that the TCM system with the encoder of Fig. P9.9.1(b) can be simulated. Run the modified program “do_TCM_8PSK.m” to find the BERs for $E_b/N_0=2\text{[dB]}$ and fill in the corresponding blanks of Table P9.9.

```matlab
%do_TCM_8PSK.m
clear
Nf=144; Tf=0.001; Tb=Tf/Nf; % Frame size, Frame and Sample/Bit times
Gc=[1 0 0; 0 5 2]; state_eq='TCM_state_eq1'; Ns=2; % Fig. P9.9.1(a)
%Gc=[1 2 0; 4 1 2]; state_eq='TCM_state_eq2'; Ns=3; % Fig. P9.9.1(b)
[K,N]=size(Gc); Rc=K/N; % Message/Codeword length and Code rate
% Constellation for TCM Simulink block
Constellation2=exp(j*2*pi/8*[0 4 2 6 1 5 3 7]);
% Constellation good for TCM() MATLAB function
Constellation1=exp(j*2*pi/8*[0:7]);
% Constraint length vector
Gc_m=max(Gc.');
for i=1:length(Gc_m), Lc(i)=length(deci2bin(oct2dec(Gc_m(i)))); end
%trell=poly2trellis(Lc,Gc);
Tbdepth=sum(Lc)*5; delay=Tbdepth*K;
Ts=K*Tb; % or N*Rc*Tb % Symbol time
EbN0dBs_t=0:0.1:10; SNRbdBs_t=EbN0dBs_t+3;
BER_theory= prob_error(SNRbdBs_t,'PSK',K,'BER');
EbN0dBs=2; %[3 6];
Target_no_of_error=50;
for i=1:length(EbN0dBs)
    EbN0dB=EbN0dBs(i); SNRbdB=EbN0dBs+i;
    pemb=TCM(state_eq,K,Ns,N,Constellation1,SNRbdB,Target_no_of_error);
    pembs_TCM8PSK(i)=pemb
    sim('TCM_sim'); pembs_TCM8PSK_sim(i)=BER(1);
end
figure(1), clf
semilogy(EbN0dBs_t,BER_theory,'b'), hold on
semilogy(EbN0dBs,pembs_TCM8PSK_sim,'r*', EbN0dBs,pembs_TCM8PSK_sim,'m*')
```
(c) For the TCM encoder with no input bit inserted into between the flip-flops (registers), the routines ‘TCM()’, ‘TCM_encoder()’, and ‘TCM_decoder()’ can be modified as ‘TCM1()’, ‘TCM_encoder1()’, and ‘TCM_decoder1()’ (listed below) so that they can accommodate any (convolutional) encoder using the (octal) code generator instead of a function representing the state equation for the TCM encoder. Run the program “do_TCM_8PSK.m” with ‘TCM()’ replaced by ‘TCM1()’ and check if the routines work properly. Fill in the corresponding blanks of Table P9.9.

(d) With reference to the set partitioning shown in Fig. 9.14 (Sec. 9.4.5), why do you think the BER is affected by different constellations? Why is the TCM system of Fig. P9.9.1(b) less affected by different constellations than that of Fig. P9.9.1(a)?

```matlab
function [outputs,state]=TCM_encoder1(Gc,input,state,Constellation)
% generates the output sequence of a binary TCM encoder
% Input:  Gc    = Code generator matrix consisting of octal numbers
%         K     = Number of input bits entering the encoder
%         Nsb   = Number of state bits of the TCM encoder
%         N     = Number of output bits of the TCM encoder
%         input = Binary input message sequence
%         state = State of the TCM encoder
%         Constellation = Signal sets for signal mapper
% Output: output = a sequence of signal points on Constellation
%Copyyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
[K,N]=size(Gc);
for k=1:K, nsb(k)=length(de2bi(oct2dec(max(Gc(k,:)))))-1; end
Nsb=sum(nsb);
tmp= rem(length(input),K);
if nargin<3, state=zeros(1,Nsb);
elseif length(state)==2^N
Constellation=state; state=zeros(1,Nsb);
end
input_length= length(input);
N_msgsymbol= input_length/K;
input1= reshape(input,K,N_msgsymbol).';
outputs= [];
for l=1:N_msgsymbol
ub= input1(l,:);
is=1; nstate=[];
output=zeros(1,N);
for k=1:K
    tmp = [ub(k) state(is:is+nsb(k)-1)];
    nstate = [nstate tmp(1:nsb(k))]; is=is+nsb(k);
    for i=1:N
        output(i) = output(i) + ...
            deci2bin1(oct2dec(Gc(k,i)),nsb(k)+1)*tmp';
    end
end
state = nstate;
output = Constellation(bin2deci(rem(output,2))+1);
outputs = [outputs output];
end
```
function decoded_seq=TCM_decoder1(Gc,demod,Constellation,opmode)
% performs the Viterbi algorithm on the PSK demodulated signal
% Input: state_eq = External function for state equation saved
%        in an M-file
% K = Number of input bits entering encoder at each cycle.
% Nsb = Number of state bits of TCM encoder
% Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
[K,N]=size(Gc);
for k=1:K, nsb(k)=length(de2bi(oct2dec(max(Gc(k,:)))))\-1; end
Nsb=sum(nsb);
N_states=2^Nsb;
N_msgsymbol=length(demod);
for m=1:N_states
    for n=1:N_msgsymbol+1
        states(m,n)=0; % inactive in the trellis diagram
        p_state(m,n)=0; n_state(m,n)=0; input(m,n)=0;
    end
end
states(1,1)=1; % make the initial state active
K2=2^K;
for n=1:N_msgsymbol
    y=demod(n); % received sequence
    n1=n+1;
    for m=1:N_states
        if states(m,n)==1 %active
            xb=deci2bin1(m-1,Nsb);
            for m0=1:K2
                u=deci2bin1(m0-1,K);
                is=1; nstate=[]; output=zeros(1,N);
                for k=1:K
                    tmp = [u(k) xb(is:is+nsb(k)-1)];
                    nstate = [nstate tmp(1:nsb(k))]; is=is+nsb(k);
                    for i=1:N
                        output(i) = output(i) +
                            deci2bin1(oct2dec(Gc(k,i)),nsb(k)+1)*tmp';
                    end
                end
                nxb(m0,:) = nstate;
                yb(m0) = Constellation(bin2deci(rem(output,2))+1);
                nxm0=bin2deci(nxb(m0,:))+1;
                states(nxm0,n1)=1;
                difference=y-yb(m0);
                d(m0)=cost(m,n)+difference*conj(difference);
                if p_state(nxm0,n1)==0
                    cost(nxm0,n1)=d(m0);
                    p_state(nxm0,n1)=m; input(nxm0,n1)=m0-1;
                else
                    [cost(nxm0,n1),i]=min([d(m0) cost(nxm0,n1)]);
                    if i==1, p_state(nxm0,n1)=m; input(nxm0,n1)=m0-1; end
                end
            end
        end
end
end
decoded_seq=[];
if nargin<4 | ~strncmp(opmode,'term',4)
    % trace back from the best-metric state (default)
    [min_cost,m]=min(cost(:,n1));
else m=1; %trace back from the all-0 state
end
for n=n1:-1:2
    decoded_seq= [deci2bin1(input(m,n),K) decoded_seq];
m=p_state(m,n);
end

function [pemb,nombe,notmb]=
    TCM1(Gc,Constellation,SNRdB Target_no_of_error)
    if nargin<4, Target_no_of_error=50; end
    [K,N]=size(Gc); M=2^N; Rc=K/N;
    Nf=288; % Number of bits per frame
    Nmod=Nf/K; % Number of symbols per modulated frame
    SNRb=10.^(SNRdB/10); SNRbc=SNRb*Rc;
    sqrtSNRc=sqrt(2*N*SNRbc); % Complex noise per K=Rc*N-bit symbol
    nombe=0; MaxIter=1e6;
    for iter=1:MaxIter
        msg=randint(1,Nf); % Message vector
        coded = TCM_encoder1(Gc,msg,Constellation); % TCM encoding
        r= coded + (randn(1,Nmod)+j*randn(1,Nmod))/sqrtSNRc;
        decoded= TCM_decoder1(Gc,r,Constellation);
        nombe = nombe + sum(msg~=decoded(1:Nf)); %
        if nombe>Target_no_of_error, break; end
    end
    notmb=Nombe*iter; % Number of total message bits
    pemb=nombe/notmb; % Message bit error probability

9.10 DSTBC (Differential Space-Time Block Coding) with Four Transmit Antennas

The DSTBC with two transmit antennas (Sec. 9.4.8) can be extended into the DSTBC with four
transmit antennas as follows:

The transmission matrix for sending $[x_0 x_1 x_2 x_{3+0} x_{3+1} x_{3+2} \cdots x_{3n} x_{3n+1} x_{3n+2} \cdots]$ to the receiver
is constructed as

$$S_{4(n+1)} = S_{4n} \cdot X_{3n} \text{ with } S_0 = I_{4x4}$$ \hspace{1cm} (P9.10.1)

where

$$X_{3n} = \begin{bmatrix}
    x_3n & -x_{3n+1} & x_{3n+2} / \sqrt{2} & x_{3n+2} / \sqrt{2} \\
    x_{3n+1} & x_3n & x_{3n+2} / \sqrt{2} & -x_{3n+2} / \sqrt{2} \\
    x_{3n+2} / \sqrt{2} & -x_{3n+2} / \sqrt{2} & x_{3n} + jx_{3n+1} & jx_{3n+1} \\
    x_{3n+2} / \sqrt{2} & -x_{3n+2} / \sqrt{2} & jx_{3n+1} & -x_{3n} + jx_{3n+1}
\end{bmatrix} \frac{1}{\sqrt{3}}$$ \hspace{1cm} (P9.10.2)

Each row of this transmission matrix is transmitted through each antenna and each column is sent
during each one of four consecutive periods. Then the signal received through a noise-free channel
having frequency response $h=[h_1 h_2 h_3 h_4]$ can be described as

$$R_{4(n+1)} = \begin{bmatrix}
    r_{4(n+1)} & r_{4(n+1)+1} & r_{4(n+1)+2} & r_{4(n+1)+3}
\end{bmatrix} = h S_{4(n+1)} X_{4n} = R_{4n} X_{4n}$$ \hspace{1cm} (P9.10.1)

$$h S_{4n} X_{4n} = R_{4n} X_{4n}$$ \hspace{1cm} (P9.10.3)
The receiver uses the following decoding rule\(^{(W-10)}\) to estimate the message sequence \(\{x_n\}\):

\[
\hat{x}_n = \begin{bmatrix}
\hat{x}_n \\
\hat{x}_{n+1} \\
\hat{x}_{n+2}
\end{bmatrix} = \frac{\sqrt{3}}{|r_{4n}^*|^2 + |r_{4n+1}^*|^2 + |r_{4n+2}^*|^2 + |r_{4n+3}^*|^2}
\left[
\begin{array}{c}
r_{4n}r_{4(n+1)}^* + r_{4n}r_{4(n+1)+1}^* + \ldots \\
r_{4n+1}r_{4(n+2)}^* - r_{4n+1}r_{4(n+2)+1}^* - \ldots \\
r_{4n+2}r_{4(n+3)}^* - r_{4n+2}r_{4(n+3)+1}^* + \ldots \\
r_{4n+3}r_{4(n+4)}^* + r_{4n+3}r_{4(n+4)+1}^* + \ldots
\end{array}
\right]
\] (P9.10.4)

Complete the following program “test_DSTBC_H4_PSK.m” and run it to see if this scheme works.

```matlab
%test_DSTBC_H4_PSK.m
clear, clf
ITR=10; MaxIter=100;
sq2= sqrt(2); sq3= sqrt(3); sq6=sqrt(6);
M=4; phs= 2*pi*[0:M-1]/M; MPSKs= exp(j*phs); % 4-PSK symbols
h=[0.95*exp(j*0.5) exp(-j*0.4) 1.02*exp(-j*0.2) 0.98*exp(j*0.1)]; % Channel
dh= 0.01*[exp(-j*0.1) -exp(j*0.2) exp(j*0.1) -exp(-j*0.1)]; % Channel variation
Amp_noise = 0.2; % Amplitude of additive Gaussian noise
ner=0;
for itr1=1:ITR
  S= eye(4); % S0=I;
  r= h*S + Amp_noise*[randn(1,4)+j*randn(1,4)]; % Eq.(P9.10.3)
  x=[]; xhn=[]; % Initialize the sequence of transmitted signal
  nn=[-3:0];
  for n=1:MaxIter
    nn= nn+4;
    xn= MPSKs(randint(1,3,M)+1); % -1+-j
    x= [x xn]; % Sequence of transmitted symbols
    x1=xn(1)/sq3; x2=xn(2)/sq3; x3=xn(3)/sq6; x3c=x3';
    x1R=real(x1); x1I=imag(x1); x2R=real(x2); x2I=imag(x2);
    X=[x1 -x2' x3c; x2  x1' x3c -x3c;
        x3 x3 -x1R+j*x2I x2R+j*x1I; x3 -x3 -x2R+j*x1I -x1R-j*x2I]; %Eq.(P9.10.2)
    S = S*X; % Encoded signal Eq.(P9.10.1)
    r0=r; % Previously received signal
    r= (h+dh*n)*S; % Received signal Eq.(P9.10.3)
    noise = Amp_noise*[randn(1,4)+j*randn(1,4)]; % Noise components
    r = r + noise; % Received signal
    r034=[r(0) r(1) r(2) r(3)]; r034_r=r034.*r(4);
    xhn1= r0(0)'*r(2)+r0(1)'*r(0)+r0(2)'*r(3)+r0(3)'*r(4); %Decoded signal Eq.(P9.10.4)
    xhn2= (r0(3)'*r(0)+r0(1)'*r(2)+r0(2)'*r(3)+r0(4)'*r(1)+r0(0)'*r(3)+r0(1)'*r(1)) + (r0(1)'*r(3)+r0(3)'*r(1)+r0(2)'*r(2)+r0(0)'*r(0)+r0(0)'*r(0)+r0(3)'*r(3)+r0(4)'*r(2))+r0(2)'*r(4); %Decoded signal Eq.(P9.10.4)
    xhn3= [xhn1 xhn2 xhn3]*(sq3/(sqrt(2)));
    ner= ner + sum(abs(x-xh)>0.1);
  end
  SER = ner/(MaxIter*3*ITR); % Symbol error rate
end
```
Accordingly, we can use the complex convolution or modulation property (1.4.11) of CTFT to find the spectrum of a sinusoidal pulse \( r_D(t) \cos(\omega_k t) \) of frequency \( \omega_k = 2\pi k / T_B \) and duration \( D = T_B \) as

\[
\begin{align*}
r_D(t) \cos(\omega_k t) & \quad (11.2.1),(11.2.2) \\
& = \frac{T_B}{2} \left\{ \frac{\sin((\omega - \omega_k)T_B / 2)}{(\omega - \omega_k)T_B / 2} + \frac{\sin((\omega + \omega_k)T_B / 2)}{(\omega + \omega_k)T_B / 2} \right\} = \frac{T_B}{2} \left\{ \text{sinc}((f - f_k)T_B) + \text{sinc}((f + f_k)T_B) \right\}, \quad f_k = k / T_B 
\end{align*}
\]

(11.2.3)

Fig. 11.3(b) based on these facts shows that the bandwidth of an \( N \)-subcarrier DFT-based OFDM signal is \( (N+1)/T_B \) [Hz] where \( N \) is the DFT size. If each carrier is loaded with \( b \) bits of data, the data transmission bit rate is

\[
\frac{N b}{T_B + T_g} \text{[bits/s]} (T_B: \text{block interval}, \ T_g: \text{guard interval , i.e., cyclic prefix duration}) \quad (11.2.4)
\]

so that the bandwidth efficiency is

\[
\frac{N b/(T_B + T_g)}{(N+1)/T_B} = \frac{N b}{N+1} \frac{T_B}{T_B + T_g} \text{[bits/Hz]} \quad (11.2.5)
\]

which becomes closer to the bandwidth efficiency \( b \) of the single-carrier communication as the number \( N \) of OFDM subcarriers increases and the guard interval \( T_g \) gets shorter.

### 11.3 CARRIER RECOVERY AND SYMBOL SYNCHRONIZATION

This topic was discussed in Chapter 8. To address the issue of carrier recovery and symbol timing on OFDM systems, let us begin with observing the effects of \( \text{STO} \) (symbol time offset) and \( \text{CFO} \) (carrier frequency offset). With this objective, we modify the MATLAB program “do_OFDM0.m” (in Section 11.1) by increasing the value of SNRbdB (in the 19th line) to 20 and activating the 11th line so that the STO amounting to one sample time is introduced. Running the modified program will yield a received signal constellation diagram like Fig. 11.4(a) and a much worse BER than that with neither STO nor CFO. Also, to take a look at the effect of CFO, activate the 14th line from the bottom (with the 11th line inactivated and SNRbdB=20) so that the CFO of 0.1 is introduced. This will yield a received signal constellation diagram like Fig. 11.4(b) and a worse BER. These effects of STO and CFO, called the \( \text{ISI} \) (inter-symbol interference) and \( \text{ICI} \) (inter-carrier interference), on the received signal can be observed and analyzed through the following expression of a time-domain received OFDM symbol:

\[
y^{(l)}[n] = \frac{1}{N} \sum_{k=0}^{N-1} G_k X^{(l)}_k e^{j2\pi(k+\varepsilon)(n+\delta)/N} + \text{noise} \quad (11.3.1)
\]

where \( y^{(l)}[n] \) denotes a (time-domain) received sequence for the \( l \)th OFDM symbol, \( G_k \) the channel frequency response (at the \( k \)th frequency index), \( X^{(l)}_k \) the \( l \)th (frequency-domain) transmitted OFDM symbol, \( \varepsilon \) the CFO, \( \delta \) the STO, and \( N \) the DFT size.
11.3 Carrier Recovery and Symbol Synchronization

Figure 11.4 The signal constellation diagrams showing the effects of STO and CFO

(a) With the STO of one sample
(b) With the CFO of 0.1

Figure 11.5 The time-domain preambles in the IEEE Standard 802.11a

(a) The structure of preamble in the IEEE Standard 802.11a

(b1) The time-domain short preamble (the real part)
(b2) The time-domain long preamble (the real part)
function [short_preamble,Xs]=short_train_seq(Nfft)
if nargin<1, Nfft=64; end
sqrt136 = sqrt((13/6)*(1+i)); Xs=zeros(1,53);
Xs([3 11 23 39 43 47 51]) = sqrt136; % Frequency domain for k=-26:26
Xs([7 15 19 31 35]) = -sqrt136;
xs = ifft([Xs(27:53) zeros(1,11) Xs(1:26)]); % Time domain
short_preamble = [xs xs xs(1:Nfft/2)];

function [long_preamble,Xl]=long_train_seq(Nfft)
if nargin<1, Nfft=64; end
Xl([1:26 28:53]) = 1; % Frequency domain for k=-26:26
Xl([3 4 7 9 16 17 20 22 29 30 33 35 37 41 44 45 47 49])=-1;
xl = ifft([Xl(27:53) zeros(1,11) Xl(1:26)]); % Time domain
long_preamble= [xl(Nfft/2+1:Nfft) xl xl];

(a) Frequency-domain short preamble displayed in the negative-positive/positive-negative frequency axes

(b) Frequency-domain long preamble displayed in the negative-positive/positive-negative frequency axes

(c) Pilot symbol displayed in the negative-positive/positive-negative frequency axes

Figure 11.6: Short preamble, long preamble, and pilot symbol in the IEEE 802.11a standard
11.3 Carrier Recovery and Symbol Synchronization

function phase_estimate=phase_from_pilot(outofFFT,pm)
% To estimate the phase offset based on the received pilot symbol
% outofFFT: OFDM symbol obtained from FFT at the RCVR
% pm : the sign of pilot signals for each OFDM symbol
determined by the PN sequence
phase_estimate = angle(outofFFT([8 22 44 58])*[pm;-pm;pm;pm]);

function y=compensate_phase(x,phase)
% To compensate the received OFDM symbol x by phase deviation
y = x*exp(-j*phase);

Many carrier recovery schemes to overcome the carrier phase distortion and frequency offset and many symbol synchronization schemes to overcome the STO effect are implemented with the correlator, PLL, narrow-band filter, and/or tuner using the pilot symbol, phase reference symbol, training symbol (called a preamble), cyclic prefix (CP), and/or null symbol. For example, Fig. 11.5(a) shows the structure of the short and long OFDM training symbols (preambles) that is used in the IEEE Standard 802.11a [W-8, Sec. 17.3.3] where the short preamble consists of 10 repetitions of a (16-sample) ‘short training sequence’ and the long one consists of 2.5 repetitions of a (64-sample) ‘long training sequence’. The above MATLAB routines ‘short_train_seq()’ and ‘long_train_seq()’ can be used to generate the short and long preambles in the time/frequency domain, respectively. Figs. 11.5(b1) and (b2) show the real parts of the short and long preambles that have been generated using each of the two routines, respectively. Figs. 11.6(a) and (b) show the frequency-domain short and long preambles, respectively, from which it can be seen that the energy of the frequency-domain long preamble $X_l(k)$ is 52 and the frequency-domain short preamble $X_s(k)$ has nonzero values of $\sqrt{13/6}(1+j)$ only for 12 subcarriers so that its energy is also $12\times(\sqrt{13/6}|1+j|^2)=52$. On the other hand, Fig. 11.6(c) shows that the pilot symbols (in the IEEE Standard 802.11a) are inserted in the $\{8^{th}$, $22^{th}$, $44^{th}$, $58^{th}\}$ subcarriers (each corresponding to frequency indices $k=\{7, 21, -21, -7\}$, respectively,) of OFDM symbols. The above routines ‘phase_from_pilot()’ and ‘compensate_phase()’ estimate the phase offset based on the received pilot symbols and compensate the phase of the received signal by the phase estimate, respectively.

Not only the phase compensation but also the CFO (carrier frequency offset) compensation is needed for successful carrier recovery. In the IEEE Standard 802.11a, the fixed-lag correlation of the short training sequence is used for the coarse estimation of CFO and the fixed-lag correlation of the long training sequence is used for the fine estimation of CFO as follows:

Coarse CFO estimate: $\hat{\epsilon}_c = \frac{N}{2\pi N_g} \angle \left\{ \sum_{n=1}^{N_g} x_s[n+N_g] x_i^* [n] \right\}$  \hspace{1cm} (11.3.2a)

Fine CFO estimate: $\hat{\epsilon}_f = \frac{1}{2\pi} \angle \left\{ \sum_{n=1}^{N} x_l[n+N] x_i^* [n] \right\}$ \hspace{1cm} (11.3.2b)

where $x_s$ is the (time-domain) short preamble, $x_l$ the (time-domain) long preamble, $N$ the FFT size and also the period of $x_i$, and $N_g$ the length of the guard interval or cyclic prefix and also the period of $x_s$. Note that the estimable CFO ranges of the coarse and fine CFO estimations are

$$\frac{N}{2\pi N_g} [-\pi, +\pi] = \frac{64}{2\pi \times 16} [-\pi, +\pi] = [-2, +2] \quad \text{and} \quad \frac{1}{2\pi} [-\pi, +\pi] = [-0.5, +0.5]$$  \hspace{1cm} (11.3.3)

respectively, where the latter estimate can be represented with less error by the same number of bits.
A question about the phase estimator (11.3.2) may arise in your mind. How can the information about the frequency offset be extracted from the time-domain correlation of the signal? To find the answer to this question, let us use Eq. (11.1.3) to write the values of, say, the short preamble \( x_s[n] \) at two points apart from each other by \( N_f = 16 \) samples as

\[
x_s[n] \overset{\text{(11.1.3)}}{=} \frac{1}{N} \sum_{m=0}^{16-1} X_s(4m) e^{j2\pi(4m+\varepsilon)n/64}
\]

and \( x_s[n+16] \overset{\text{(11.1.3)}}{=} \frac{1}{N} \sum_{m=0}^{16-1} X_s(4m) e^{j2\pi(4m+\varepsilon)(n+16)/64} \) \hspace{1cm} (11.3.4)

where we have taken into consideration that the spectrum \( X_s(k) \) of the short preamble \( x_s[n] \) has nonzero values only for \( k = 4m \) (see Fig. 11.6(a)). Multiplying one with the other’s conjugate yields

\[
x_s[n+16]x_s^*[n] = \frac{1}{N^2} \sum_{m=0}^{16-1} X_s(4m) X_s^*(4m) e^{j2\pi(4m+\varepsilon)(n+16)/64} - j2\pi(4m+\varepsilon)n/64
\]

(The cross multiplication terms for \( m \neq i \) disappears due to their orthogonality.)

\[
= \frac{1}{16\times13} (1+j)(1-j) e^{j2\pi(4m+\varepsilon)16/64} = \frac{13}{128} \angle(2m\pi + \frac{\pi}{2} \varepsilon) = K \angle\frac{\pi}{2} \varepsilon \quad \text{(11.3.5)}
\]

which presents the basis of the coarse CFO estimator Eq. (11.3.2a). Similarly, the correlation of the long preamble yields

\[
x_i[n+N]x_i^*[n] = \frac{1}{N} \sum_{m=0}^{N-1} X_i(m) e^{j2\pi(m+\varepsilon)(n+N)/N} - j2\pi(m+\varepsilon)n/N
\]

(The cross multiplication terms for \( m \neq i \) disappears due to their orthogonality.)

\[
= \frac{1}{N} e^{j2\pi(m+\varepsilon)} = \frac{1}{N} \angle(2m\pi + 2\pi \varepsilon) = \frac{1}{N} \angle 2\pi \varepsilon \quad \text{(11.3.6)}
\]

which presents the basis of the fine CFO estimator Eq. (11.3.2b). Note from Eqs. (11.3.2a) and (11.3.2b) that the phase is taken for the sum of the multiplication terms instead of the average. Why? Because the phase of a complex number does not depend on its magnitude. That is also why the average is not taken for the sum of the four pilot symbol terms in the routine ‘phase_from_pilot()’.

The following routines ‘coarse_CFO_estimate()’ and ‘fine_CFO_estimate()’ can be used to get the coarse and fine CFO estimates via Eqs. (11.3.2a) and (11.3.2b), respectively. Note that the two routines differ in the frequency range to estimate, but not in the accuracy and therefore we can get a good CFO estimate using only the first one. However, to emphasize the quantitative difference between the two routines, it is assumed that an 8-bit number representation with relative resolution \( 2^{-8} \) is used for storing the CFO estimation result. The next MATLAB program “do_CFO.m” uses the routines ‘short_train_seq()’/’long_train_seq()’ to generate the short/long preambles, uses the routine ‘set_CFO()’ to set up a CFO, uses ‘coarse_CFO_estimate()’ and ‘fine_CFO_estimate()’ to estimate the CFO, and then uses ‘compensate_CFO()’ to compensate the CFO.
function coarse_CFO_est = coarse_CFO_estimate(tx,NB,Nw,STO)
    % Input:
    %    tx = Received signal
    %    NB = Which block of size Nw to start computing correlation with?
    %    Nw = Correlation window size
    %    STO= Symbol Time Offset
    % Output: coarse_CFO_est = Estimated carrier frequency offset
    if nargin<4, STO = 0; end
    if nargin<3, Nw = 16; end
    if nargin<2, NB = 6; end
    for i=1:2
        nn = STO + Nw*(NB+i) + [1:Nw];
        CFO_est(i) = angle(tx(nn+Nw)*tx(nn)'); % Eq.(11.3.2a)
    end
    CFO_est = sum(CFO_est)/pi; % Average
    coarse_CFO_est = CFO_est - mod(CFO_est,4/128); % Stored with 8 bits
end

function fine_CFO_est = fine_CFO_estimate(tx,coarse_CFO_est,Nfft,STO)
    % Input : tx = Received signal
    %         coarse_CFO_est = coarse CFO estimate
    %         Nfft = FFT size
    %         STO  = Symbol Time Offset
    % Output: fine_CFO_est = Estimated carrier frequency offset
    if nargin<4, STO = 0; end
    if nargin<3, Nfft = 64; end
    if nargin<2, coarse_CFO_est = 0; end
    tx1 = CFO_compensation(tx,coarse_CFO_est,Nfft,STO);
    nn = STO + Nfft/2 + [1:Nfft];
    cfo_est = angle(tx1(nn+Nfft)*tx1(nn)')/(2*pi); % Eq.(11.3.2b)
    fine_CFO_est = CFO_est - mod(CFO_est,1/128); % Stored with 8 bits
end

%do_CFO.m
clear, clf
Nfft = 64; Ng = 16;
CFO = 1.7; phase = 0; STO = -1; % CFO/Phase Offset/STO
NB = 6; % Which block to start computing the correlation with?
Nw = Ng; % Correlation window size
[short_preamble,S] = short_train_seq(Nfft);
[long_preamble,L] = long_train_seq(Nfft);
% Time-domain training symbol
tx = [short_preamble  long_preamble];
% Set up a pseudo CFO
tx_offset = set_CFO(tx,CFO,phase,Nfft);
% Coarse CFO estimation
coarse_CFO = coarse_CFO_estimate(tx_offset(1:160),NB,Nw,STO);
% Fine CFO estimation
fine_CFO = fine_CFO_estimate(tx_offset(161:320),coarse_CFO,Nfft);
% Overall CFO estimate
CFO_estimate = coarse_CFO + fine_CFO;
form1 = '\n For CFO=%10.8f, CFO estimate(%10.8f) = ';
form2 = 'coarse estimate(%10.8f)+fine estimate(%10.8f)\n';
fprintf([form1 form2],CFO,CFO_estimate,coarse_CFO,fine_CFO);
% CFO compensated symbols
tx_compensated = compensate_CFO(tx_offset,CFO_estimate,Nfft,STO);
discrepancy = norm(tx-tx_compensated)/length(tx)
function tx_with_CFO = set_CFO(tx,CFO,phase,Nfft,STO)
    % CFO/phase = pseudo Carrier frequency/phase offset (pretended)
    % STO = Symbol Time Offset
    if nargin<5, STO = 0; end % Symbol Time Offset
    if nargin<4, Nfft = 64; end
    if nargin<3, phase = 0; end
    n = -STO + [0:length(tx)-1];
    tx_with_CFO = tx.*exp(j*(2*pi*CFO/Nfft*n+phase));
end

function CFO_compensated = compensate_CFO(tx,CFO_est,Nfft,STO)
    % CFO_est = Estimated carrier frequency offset
    % STO = Symbol Time Offset
    if nargin<4, STO = 0; end
    if nargin<3, Nfft = 64; end
    n = -STO + [0:length(tx)-1];
    CFO_compensated = tx.*exp(-j*2*pi*CFO_est/Nfft*n);
end

(a) A subsystem for setting or compensating CFO (carrier frequency offset)

(b) A subsystem for coarse CFO estimation based on the short preamble

(c) A subsystem for fine CFO estimation based on the long preamble

Figure 11.7 Subsystems to be used for CFO estimation and compensation
11.3 Carrier Recovery and Symbol Synchronization

Fig. 11.7 shows the three Simulink subsystems that can be used to estimate and compensate the CFO in a Simulink model for OFDM system simulation. Note the following about the subsystems:

- The subsystem of Fig. 11.7(a) is equivalent to the MATLAB routine `set_CFO()`, but it can be made to work like `compensate_CFO()` by negating the CFO input as can be seen in Fig. 11.8.
- The subsystem of Fig. 11.7(b) is equivalent to the MATLAB routine `coarse_CFO_estimate()`, which uses Eq. (11.3.2a) to make a coarse CFO estimation based on the two correlation values, one between the 8th (16-sample) block and the 9th block and the other between the 9th block and the 10th block of the received sequence corresponding to the (160-sample) short preamble.
- The subsystem of Fig. 11.7(c) is equivalent to `fine_CFO_estimate()`, which uses Eq. (11.3.2b) to make a fine CFO estimation based on the correlation value between the 1st (64-sample) block and the 2nd block of the received sequence corresponding to the (160-sample) long preamble.

Fig. 11.8 shows a Simulink model named “CFO_sim.mdl”, which demonstrates the CFO estimation and compensation using the short and long preambles and the three subsystems depicted in Fig. 11.7. About this Simulink model, note the following:

- The coarse CFO estimate is computed by the (subsystem) block ‘coarse CFO estimate’ and then is used to compensate the CFO of the long preamble through the block ‘compensate_CFO’.
- The fine CFO estimate (obtained from the correlation value of the coarse-CFO-compensated long preamble by the block ‘fine CFO estimate’) and the coarse CFO estimate are negated, fed into the block ‘compensate_CFO1’, and then used to compensate the CFO of a general signal.

Interested readers are invited to compose the Simulink model “CFO_sim.mdl” and run it to see if the CFO is well tracked for different values of CFO and consequently, the mean square error between the original signal and the CFO-inserted-and-compensated one (computed by the block ‘MSE’) is very small.
%do_STO_estimation.m

% To estimate the STO (Symbol Time Offset)
% Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only
% clear, clf
Cor_thd = 0.93; % Threshold of correlation for peak detection
Nfft=64; Ng=16; % FFT/Guard interval size
Nsym=Nfft+Ng; Nw=Ng; % Symbol/Window size
Nd=65; % Remaining period of the last symbol in the previous frame

% Make a pseudo frame to transmit
% Evaluate the number of OFDM symbols to be generated for simulation
for i=1:N_Symbols
    symbol = rand(1,Nfft)-0.5+j*(rand(1,Nfft)-0.5); % An arbitrary data
    symbol_cp = [symbol(end-Ng+1:end) symbol]; % OFDM Symbol with CP
    t_frame = [t_frame symbol_cp]; % Append a frame by a symbol with CP
end
L_frame = length(t_frame); nn=0:L_frame-1;
noise = 0.2*(rand(1,L_frame)-0.5 +j*(rand(1,L_frame)-0.5));
r = t_frame + noise;
sig_w = zeros(2,Nw); % Initialize the two sliding window buffers
STOs = [0]; % Initialize the STO buffer
for n=1:L_frame
    sig_w(1,:) = [sig_w(1,2:end) r(n)]; % Update signal window 1
    m = n-Nfft;
    if m>0
        sig_w(2,:) = [sig_w(2,2:end) r(m)]; % Update signal window 2
        den = norm(sig_w(1,:))*norm(sig_w(2,:));
        corr(n) = abs(sig_w(1,:)*sig_w(2,:)/den;
        if corr(n)>Cor_thd & m>STOs(end)+Nsym-15
            STOs=[STOs m]; % List the estimated STO
        end
    end
end
Estimated_STOs = STOs(2:end)
True_STOs = Nd+Ng + [0:N_Symbols-1]*Nsym
subplot(311)
stem(nn,real(r)), ylim([-0.6 1.1])
hold on, stem(True_STOs,0.8*ones(size(True_STOs)),'k*')
stem(Estimated_STOs,0.6*ones(size(Estimated_STOs)),'rx')
title('Estimated Starting Times of OFDM Symbols

subplot(312)
plot(nn+1,corr), ylim([0 1.2]), hold on,
stem(Estimated_STOs+Nfft,corr(Estimated_STOs+Nfft),'r:^')
% The points at which the correlation is presumably maximized,
% yielding the STO estimates.
stem(Estimated_STOs,0.9*ones(size(Estimated_STOs)),'rx')
title('Correlation between two sliding windows across Nfft samples')
set(gca,'XTick',sort([Estimated_STOs Estimated_STOs+Nfft])))
11.3 Carrier Recovery and Symbol Synchronization

The above program "do_STO_estimation.m" demonstrates how the correlation value can be used to estimate the STO of an OFDM symbol consisting of $N_g$ (guard interval size) cyclic prefix (CP) samples and $N$ (FFT size) signal samples. As shown in Fig. 11.9(a), it keeps updating a $2 \times N_g$ matrix \( \text{sig}_w \) that contains the $N_g$ samples of two sliding windows in each of its two rows, finds the (local) peak times of the crosscorrelation value between two row vectors and sets the times \((m=n-N_{\text{fft}})\) of $N$ samples before the local peak times to the STO estimates (to be stored in the vectors named 'STOs' and 'Estimated_STOs'). In Fig. 11.9(b), the measured peak times of the correlation value and the corresponding STO estimates are denoted by dotted lines and solid lines, respectively.

![Diagram](image.png)

(a) True starting times and estimated ones of OFDM symbols each of which consists of \((N_g+N)\) samples

(b) Correlation between two sliding windows (of size \(N_g\)) apart from each other by \(N_{\text{fft}}\) samples

Figure 11.9 STO (symbol time offset) estimation

The correlation value is divided by the product of the norms of the two vectors for normalization as

\[
\phi[n] = \frac{\sum_{m=-N_g+1}^{n} x[m] \bar{x}[m-N]}{\sqrt{\sum_{m=-N_g+1}^{n} |x[m]|^2} \sqrt{\sum_{m=-N_g+1}^{n} |x[m-N]|^2}} = \frac{\text{sig}_w(1,:)'\text{sig}_w(2,:)}{|\text{sig}_w(1,:)| |\text{sig}_w(2,:)|}
\]

(\(\bar{}\)): the MATLAB operator representing the conjugate transpose, i.e., Hermitian)

and a threshold of, say, Cor_thd=0.93 is used to determine if the presumable peak time has been reached or not. The additional condition \((m>\text{STOs(end)}+15)\) to determine peak times has been required not to let another (most probably wrong) peak reported within $N_{\text{sym}}-15$ \((N_{\text{sym}}=N_g+N: \text{OFDM symbol duration})\) samples after a peak is reported.
Let us consider the synchronization scheme that is used in the EUREKA-147 standard for DAB (digital audio broadcasting). It consists of the four processes working in coordination, i.e., the coarse frame synchronization, the IFO (integral frequency offset) estimation, the fine frame synchronization, and the FFO (fractional frequency offset) estimation, as depicted by the block diagram of Fig. 11.10(a). Fig. 11.10(b) shows the EUREKA-147 transmission frame structure of 96ms for transmission mode I where the null symbol and PRS (phase reference symbol) symbol preceding the frame can be used for the frame synchronization.

<table>
<thead>
<tr>
<th>Transmission mode</th>
<th>I (Large area)</th>
<th>II (Topographical situation)</th>
<th>III (Satellite transmission)</th>
<th>IV (Limited area and direct line of sight)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum carrier frequency</td>
<td>375 MHz</td>
<td>1.5 GHz</td>
<td>3 GHz</td>
<td>1.5 GHz</td>
</tr>
<tr>
<td>Number of subcarriers</td>
<td>1.536</td>
<td>384</td>
<td>192</td>
<td>768</td>
</tr>
<tr>
<td>Subcarrier interval frequency</td>
<td>1 kHz</td>
<td>4 kHz</td>
<td>8 kHz</td>
<td>2 kHz</td>
</tr>
<tr>
<td>Guard interval time</td>
<td>245 us</td>
<td>62 us</td>
<td>31 us</td>
<td>123 us</td>
</tr>
<tr>
<td>OFDM symbol period</td>
<td>1 ms</td>
<td>250 us</td>
<td>125 us</td>
<td>500 us</td>
</tr>
<tr>
<td>Frame period</td>
<td>96 ms</td>
<td>24 ms</td>
<td>24 ms</td>
<td>40 ms</td>
</tr>
</tbody>
</table>

Now, let us consider the synchronization scheme that is used in the EUREKA-147 standard for DAB (digital audio broadcasting). It consists of the four processes working in coordination, i.e., the coarse frame synchronization, the IFO (integral frequency offset) estimation, the fine frame synchronization, and the FFO (fractional frequency offset) estimation, as depicted by the block diagram of Fig. 11.10(a). Fig. 11.10(b) shows the EUREKA-147 transmission frame structure of 96ms for transmission mode I where the null symbol and PRS (phase reference symbol) symbol preceding the frame can be used for the frame synchronization.
11.3 Carrier Recovery and Symbol Synchronization

1. Coarse Frame Synchronization

The tactics to estimate the starting points of a null symbol and a frame based on the energy ratio between the two adjacent windows of size \( N_{e} \) can be expressed by the following equations:

Null time estimate:
\[
\hat{n}_{N} = \text{Arg Min}_{n-N_{e}} \frac{\sum_{m=n-N_{e}+1}^{n} T_{m}^* T_{m}}{\sum_{m=n-N_{e}+1}^{n} T_{m}^* T_{m-N_{e}}} \quad (11.3.8a)
\]

Frame time estimate:
\[
\hat{n}_{F} = \text{Arg Max}_{n-N_{e}} \frac{\sum_{m=n-N_{e}+1}^{n} T_{m}^* T_{m}}{\sum_{m=n-N_{e}+1}^{n} T_{m}^* T_{m-N_{e}}} \quad (11.3.8b)
\]

The estimated null symbol period \((\hat{n}_{F} - \hat{n}_{N})\) can be used to detect the transmission mode since it depends on the transmission mode (see Table 11.1).

The following program "do_sync_w_double_window.m" simulates the process of catching the starting points of a null symbol and the frame (preceded by the null symbol) based on the energy ratio between two successive blocks of size \( N_{null} \). It also simulates the process of estimating the starting points of each OFDM symbol based on the correlation value between two blocks of size \( N_{g} \) spaced \( N_{fft} \) samples apart (\( N_{fft} \) : FFT size). To minimize the number of computations as well as to save the memory, we maintain several windows (buffers) to store some duration of signal samples, powers, energies, and correlations as follows: (see Fig. 11.11)

1. \( \text{power}_{w}[n] = r[n]^{*} r[n] \) with size of \( N_{null} + 1 \)
2. \( \text{energy}_{w1}[n] = \sum_{m=n-N_{null}+1}^{n} \text{power}_{w}[m] = \text{energy}_{w1}[n-1] + \text{power}_{w}[n] - \text{power}_{w}[n-N_{null}] \)
   \( = \text{energy}_{w1}[n-1] + \text{power}_{w[end]} - \text{power}_{w}[1] \) with size of \( N_{null} + 1 \)
3. \( \text{sig}_{w}[n] = r[n] \) with size of \( N_{fft} + 1 \)
4. \( \text{corr}_{w}[n] = r[n]^{*} r[n-N_{fft}] = \text{sig}_{w[end]}^{*} \text{sig}_{w}[1] \) with size of \( N_{g} + 1 \)
5. \( \text{energy}_{w2}[n] = \sum_{m=n-N_{g}+1}^{n} \text{power}_{w}[m] = \text{energy}_{w2}[n-1] + \text{power}_{w}[n] - \text{power}_{w}[n-N_{g}] \)
   with size of \( N_{fft} + 1 \)

\[\text{correlation}[n] = \frac{\text{corr}[n]}{\sqrt{\text{energy}_{w2}[n] \text{energy}_{w2}[n-N_{fft}]}} \]

Figure 11.11 Window buffers maintained to catch the starting points of the null, frame, and OFDM symbols
%do_sync_w_double_window.m
% Copyleft: Won Y. Yang, wyyang53@hanmail.net, CAU for academic use only

clear, clf
Cor_thd=0.988; % The threshold to determine the peak of correlation
Nfft=64; Ng=16; Nsym=Nfft+ Ng; Nsym1=Nsym+1;
Nnull=Nsym; Nw=Nnull; Nw1=Nw+1; Nw2=Nw*2; Ng1=Ng+1; Nfft1=Nfft+1;
Nd=90; % Remaining period of the last symbol in the previous frame
N_OFDM=3; % One Null + N_OFDM symbols
Max_energy_ratio=0; Min_energy_ratio=1e10;

t = [rand(1,Nd)-0.5+j*(rand(1,Nd)-0.5) zeros(1,Nnull)];
for i=1:N_OFDM
    symbol=rand(1,Nfft)-0.5 +j*(rand(1,Nfft)-0.5);
t = [t symbol(end-Ng+1:end) symbol];
end
L_frame = length(r);
t = t + 0.1*(rand(1,L_frame)-0.5+j*(rand(1,L_frame)-0.5));
energy_wl=zeros(1,Nw1); power_w=zeros(1,Nw1);
sig_w=zeros(1,Nfft1); energy_w2=zeros(1,Nfft1); corr_w=zeros(1,Ng1);
OFDM_start_points = [0]; corr=0;
for n=1:L_frame
    sig_w = [sig_w(2:end) t(n)]; % Signal window
    power_n = t(n)'*t(n); % Current signal power
    power_w = [power_w(2:end) power_n]; % Power window
    energy_wl=[energy_wl(2:end) energy_wl(end)+power_n]; % Energy window
    if n>Nw, energy_wl(end)=energy_wl(end)-power_w(1); end % of size Nw
    energy_w2=[energy_w2(2:end) energy_w2(end)+power_n]; % Energy window
    if n>Ng, energy_w2(end)=energy_w2(end)-power_w(end-Ng); end
    corr_w(1:end-1) = corr_w(2:end);
    if n>Nfft
        % Correlation between signals at 2 points spaced Nfft samples apart
        corr_w(end)=abs(sig_w(end)'*sig_w(1)); corr=corr+corr_w(end);
    end
    if n>Nsym, corr=corr-corr_w(1); end % Correlation window of Ng pts
end
if n=Nw2
    energy_ratio = energy_wl(end)/energy_wl(1);
    energy_ratios(n) = energy_ratio;
    if energy_ratio<Min_energy_ratio % Eq.(11.3.8a)
        Min_energy_ratio = energy_ratio; Null_start_point = n-Nw+1;
    end
    if energy_ratio>Max_energy_ratio % Eq.(11.3.8b)
        Max_energy_ratio = energy_ratio; F_start = n-Nw+1;
    end
end
% CP-based Symbol Time estimation
if n>Nsym
    % Normalized, windowed correlation across Nfft samples for Ng pts
    correlation=corr/sqrt(energy_w2(end)*energy_w2(1)); % Eq.(11.3.9)
    corrations(n) = correlation;
    if correlation>Cor_thd&n-Nsym>OFDM_start_points(end)+Nfft
        OFDM_start_points = [OFDM_start_points n-Nsym+1];
    end
end
11.3 Carrier Recovery and Symbol Synchronization

Estimated_start_points =
[Null_start_point F_start OFDM_start_points(2:end)]

True_start_points = [Nd+1:Nsym:L_frame]

N_True_start_points = length(True_start_points);

subplot(311),stem(real(r)),set(gca,'XTick',True_start_points)
hold on, stem(True_start_points,0.9*ones(1,N_TRUE_start_points),'k*')

N_Sym = length(Estimated_start_points);

stem(Estimated_start_points,1.1*ones(1,N_Sym),'rx')
title('Estimated Starting Points of Symbols')

subplot(312),semilogy(energy_ratios),set(gca,'XTick',True_start_points)
title('Ratio of 2 Successive Windowed Energies for Nw samples')

subplot(313), plot(correlations)
hold on, title('Correlation across Nfft samples')

At every instant when a sampled signal arrives, the normalized and windowed correlation value

\[ \text{correlation}[n] = \frac{\sum_{m=n-N_{m}+1}^{n} \text{corr}_w[m]}{\sqrt{\text{energy}_w2[n] \text{energy}_w2[n-N_{m}]} > \text{Threshold}} \]  (11.3.9)

is computed to determine if it is the correlation value between the CP and the last \( N_{m} \) samples of an OFDM symbol, i.e. if the current sample is the end of an OFDM symbol or not. If the normalized correlation value is found to exceed some threshold, say, 0.988 at \( n \), the starting point of the detected OFDM symbol is determined to be \( n - N_{sym} + 1 \) (one OFDM symbol duration before the detection time) where \( N_{sym} = N_{d} \text{sym} + N_{fft} \) is the OFDM symbol duration. Compared with this symbol timing process, catching the starting points of a null symbol and a frame based on the energy ratio between two successive blocks is very misty since there can be no normalization of energy ratio and accordingly, the appropriate threshold values to determine the maximum and minimum of the energy ratio are difficult to fix. Let us run the MATLAB program “do_sync_w_double_window.m” to get Fig. 11.12 together with the following result:

![Diagram](image)

(a) Estimated starting points of a null, the frame preceded by the null, and successive OFDM symbols

(b) Energy ratio between adjacent blocks (windows) of size Nnull

(c) Correlation values between two blocks (windows) of size Nfft spaced Nfft samples apart

Figure 11.12 An example of simulation results obtained by running "do_sync_w_double_window.m"
Chapter 11 OFDM System

>> do_sync_w_double_window
    Estimated_start_points = 91 171 171 248 330
    True_start_points = 91 171 251 331

The detection of the starting points of a null, the frame preceded by the null symbol, and successive OFDM symbols seems to be successful. That is right as far as the OFDM symbol starting points are concerned. However, the maximum/minimum points of the energy ratio for estimating the starting points of a null/frame (shown in Fig. 11.12(b)) have not been detected on-line since they could be recognized only after a considerable time span. In this aspect, the above program has a room for improvement towards more practical detection of the starting points of a null and the frame preceded by the null symbol, possibly in concert with the OFDM symbol starting point detection.

2. Integral Frequency Offset Estimation

The CFO (carrier frequency offset) that may be caused by the mismatch of the oscillators in the transmitter and receiver can be regarded as the sum of the IFO (integral frequency offset) and the FFO (fractional frequency offset) where the IFO and the FFO are an integer and a fraction times the subcarrier frequency interval \( \Omega = \frac{2\pi}{N_{fb}} \), respectively. The IFO can be estimated based on the correlation between the received PRS (phase reference symbol)-FFT signal \( Y_k \) and the local (frequency-domain) PRS signal \( X_{PRS}(k) \) at the receiver by applying one of the following three methods:

APRS (Algorithm using PRS):

\[
\hat{f}_i = \text{ArgMax}_d \left| \sum_{k=0}^{N_{fb}-1} Y_{PRS}(k)X_{PRS}^*(k-d) \right| 
\]  
(11.3.10a)

ACIR (Algorithm using the channel impulse response):

\[
\hat{f}_i = \text{ArgMax}_d \left\{ \text{Max}_n z[n] = \text{IFFT}\{Y_{PRS}(k)X_{PRS}^*(k-d)\} \right\} 
\]  
(11.3.10b)

AIDC (Algorithm using the intercarrier differential correlation):

\[
\hat{f}_i = \text{ArgMax}_d \left| \sum_{k=0}^{N_{fb}-1} Y_{PRS}(k)X_{PRS}^*(k-d) \right| 
\]  
(11.3.10c)

where

\[
\bar{Y}(k) = Y(k)\hat{Y}^*(k-1), \quad \bar{X}(k) = X(k)\hat{X}^*(k-1) 
\]  
(11.3.10d)

3. Fine Frame Synchronization

The fine frame synchronization can be performed by taking the peak point of the correlation between the received signal \( y_{PRS}[n] \) and the time-domain PRS \( x_{PRS}[n] \) where the (time-domain) correlation can equivalently be computed from the IFFT of the (frequency-domain) product of the received FFT signal \( \bar{Y}(k) \) and the IFO-compensated PRS \( X_{PRS}^*(k-\hat{f}_i) \):

\[
\hat{n}_f = \text{ArgMax}_n \left\{ z[n] = \text{IFFT}\{\bar{Y}(k)\bar{X}^*(k-\hat{f}_i)\} \right\} 
\]  
(11.3.11)
4. Fractional Frequency Offset Estimation

With the same idea as Eq. (11.3.2b), the fractional frequency offset (FFO) estimate can be obtained from $1/2\pi$ times the phase difference between the two (fine frame-synchronized) blocks of size $N_w$ spaced $N_{fr}$ samples apart:

$$\hat{\Delta f} = \frac{1}{2\pi} \arg \left\{ \sum_{n=1}^{N_w} y[n + \hat{r}_f + N_{fr}] y^*[n + \hat{r}_f] \right\}$$

(11.3.12)

where the two blocks are supposed to agree with each other without CFO and noise and the window size $N_w$ is often set equal to the guard interval or cyclic prefix (CP) size $N_g$. Note that a large window size will help increasing the accuracy of the FFO estimation.

---

```matlab
function IFO_est=IFO_estimate(y,X,IFO_range)
% To estimate the IFO (Integral Frequency Offset)
% y: A received time-domain signal, supposedly containing ifft(PRS)
% X: (frequency-domain) Phase Reference Symbol
% IFO_range : Range of possible IFOs to be searched
M=3; Max=zeros(1,M); IFO_est=zeros(1,M); % 3 methods to estimate IFO
Nfft=length(X); Y = fft(y,Nfft);
Ybar=Y.*conj(Y([Nfft 1:Nfft-1])); % Eq.(11.3.10)
Xbar=X.*conj(X([Nfft 1:Nfft-1]));
for i=1:length(IFO_range)
d=IFO_range(i); YX = Y.*conj(rotate_r(X,d));
Mag(1) = abs(sum(YX)); % Eq.(11.3.10a) APRS
Mag(2) = max(abs(ifft(YX))); % Eq.(11.3.10b) ACIR
Mag(3) = abs(Ybar*rotate_r(Xbar,d)'); % Eq.(11.3.10c) AIDC
for m=1:M
    if Mag(m)>=Max(m), Max(m)=Mag(m); IFO_est(m)=d; end
end
end

function PRS=phase_ref_symbol()
% Nfft=Ns+Nvc=1536+512=2048;
Nfft=1536+512;%Nsd(# of data subcarriers)+Nvc(# of virtual carrier)
h= ...
0 2 0 0 0 0 1 1 2 0 0 0 2 2 1 1 0 2 0 0 0 0 1 1 2 0 0 0 2 2 1 1;
0 3 2 3 0 1 3 0 2 1 2 3 2 3 3 0 0 3 2 3 0 1 3 0 2 1 2 3 2 3 3 0;
0 0 0 2 0 2 1 3 2 2 0 2 2 0 1 3 0 0 0 2 0 2 1 3 2 2 0 2 2 0 1 3;
0 1 2 1 0 3 3 2 2 3 2 1 2 1 3 2 0 1 2 1 0 3 3 2 2 3 2 1 2 1 3 2;
0 2 0 0 0 0 1 1 2 0 0 0 2 2 1 1 0 2 0 0 0 0 1 1 2 0 0 0 2 2 1 1;
0 1 2 0 1 3 2 3 2 1 2 3 1 2 3 3 2 2 2 1 1 3 1 2 ...
3 1 1 1 2 2 1 0 2 2 3 3 0 2 1 3 3 3 0 3 0 1 1];
jp12 = j*pi/2;
for p=1:12 %12*4*32=1536
    for i=1:4
        if p<=6, il=i; else il=6-i; end
        for k=1:32
            temp_seq((p-1)*128+(i-1)*32+k)=exp(jp12*(h(il,k)+n((p-1)*4+i)));
        end
    end
end
PRS = [zeros(1,256) temp_seq(1:768) 0 temp_seq(769:end) zeros(1,255)];
```
%do_sync_for_DMB.m

clear, clf
Nfft=2048; Ng=504; Nnull=2656;
CFO = -1.7; phase = 0; % A pseudo Carrier Frequency/Phase Offset
nF = 1 % A pseudo Frame Time Offset (delay)
PRS = phase_ref_symbol;
prs = ifft(PRS,Nfft); % Time-domain PRS
tx = [prs(Nfft-Ng+1:Nfft) prs]; % Add Cyclic Prefix
L_tx=length(tx);
% Set up the (pseudo) CFO (pretended)
Nd = 100; % Delay tolerance
y = set_CFO(tx,CFO,phase,Nfft); A=0.05;
if nF>0
    y=[A*(rand(1,nF)-0.5+j*(rand(1,nF)-0.5)) y(1:end-nF)]; %Delayed
elseif nF<0
    y=[y(1-nF:end) A*(rand(1,-nF)-0.5+j*(rand(1,-nF)-0.5))]; %Advanced
end
% IFO (Integral Frequency Offset) estimation
IFO_range = [-3:3];
y_ifft = y(Nd+Ng+[1:Nfft]);
IFO_est = IFO_estimate(y_ifft,PRS,IFO_range); % Eq.(11.3.10a,b,c)
%In order to realize how critical the IFO estimate is for FFS,
% activate the following statement even with a very short delay nF=1;
% IFO_est = zeros(1,3);
% FFS (fine frame synchronization)
Y = fft(y_ifft,Nfft);
YX = Y.*conj(rotate_r(PRS,IFO_est));
[Max,nF_] = max(abs(ifft(YX)));% Eq.(11.3.11)
if nF_>Nfft/2, nF_=nF_-Nfft; end % Periodicity of FFT/IFFT
nF_h = (nF_-1)
% FFO (Fractional Frequency Offset) estimation
nn = Nd+[1:Ng]; nn1 = Nfft + nn;
% To realize the importance of reflecting the FFS into FFO estimation,
% activate the following statement with a delay nF=2;
% nF_h = 0;
FFO_est = angle(y(nF_h+nn1)*y(nF_h+nn)')/(2*pi) % Eq.(11.3.12)
% FFO estimate without incorporating the FFS result
FFO_est0 = angle(y(nn1)*y(nn)')/(2*pi) % Eq.(11.3.12) without nF_h
% Overall CFO estimate
CFO_est = IFO_est + FFO_est;
fprintf('
 For CFO=%10.8f,
', CFO);
form = '  CFO_estimate(%10.8f) = IFO(%10.8f)+FFO(%10.8f)
';
for i=1:3
    fprintf(form, CFO_est(i), IFO_est(i), FFO_est);
end
% CFO compensated symbols
y_compensated = compensate_CFO(y,CFO_est(3),Nfft);
% Discrepancy between the CFO compensated symbols and original ones
discrepancy = norm(tx-y_compensated(Nd+[1:L_tx]))/L_tx
The above program “do_sync_for_DMB.m” simulates the synchronization scheme depicted in Fig. 11.10(a) where the routines ‘IFO_estimate()’ and ‘phase_ref_symbol()’ are used to compute the IFO estimates by Eqs. (11.3.10a–c) and to generate the frequency-domain PRS (phase reference symbol), respectively.

### 11.4 CHANNEL ESTIMATION AND EQUALIZATION

A frequency-domain training sequence $X(k) = X_R(k) + jX_I(k)$ with the corresponding channel output $Y(k) = Y_R(k) + jY_I(k)$ can be used for channel estimation as

$$
\hat{H}(k) = \frac{Y(k)}{X(k)} = \frac{X^*(k)Y(k)}{X^*(k)X(k)} = \frac{(X_R(k)Y_R(k)+X_I(k)Y_I(k))+j(X_R(k)Y_I(k)-X_I(k)Y_R(k))}{X_R^2(k)+X_I^2(k)}
$$

(11.4.1)

where $X(k)$ and $Y(k)$ are the FFTs of the (long preamble) input and the corresponding output of the channel, respectively, $X_R(k)$ and $Y_R(k)$ are the real parts, and $X_I(k)$ and $Y_I(k)$ are the imaginary parts of $X(k)$ and $Y(k)$, respectively. Noting that in the IEEE Standard 802.11a, the long preamble used for channel estimation is $X(k) = X_R(k) + jX_I(k) = 1$ or $-1$ (with $X_I(k) = 0$) as shown in Fig. 11.6(b), the channel estimator (11.4.1) can be simplified as

$$
\hat{H}(k) = \frac{Y(k)}{X(k)} = \frac{X_R(k)Y_R(k)+X_I(k)Y_I(k)}{X_R^2(k)+X_I^2(k)} = X(k)Y(k)
$$

(11.4.2)

Since the long preamble contains two repeated training sequences, the average of the FFTs ($Y_I(k)$ and $Y_2(k)$) of the channel outputs to the two consecutive training sequences can be taken for better channel estimation:

$$
\hat{H}(k) = \frac{1}{2} X(k)(Y_1(k)+Y_2(k))
$$

(11.4.3)

This estimation scheme is cast into the MATLAB routine ‘channel_estimate().’ The following program “do_channel_estimation.m” uses this routine to estimate a channel based on the frequency-domain long preamble $X(k)$ and the FFT $Y(k)$ of the output of the channel to the (time-domain) long preamble. It also uses another routine ‘equalizer_in_freq()’ to equalize the output to the unknown input by compensating the channel effect:

$$
\hat{X}(k) = \frac{Y(k)}{\hat{H}(k)}
$$

(11.4.4)

```matlab
function H_est=channel_estimate(X,y)
% X = Known frequency-domain training symbol
% y = Time-domain output of the channel
% H_est = Estimate of the channel (frequency) response
if length(X)>52, X = X([1:26 28:53]); end % for k=[-26:-1 1:26]
y1 = y(32+[1:64]); y2 = y(96+[1:64]);
Y1 = fft(y1); Y1 = Y1([39:64 2:27]); % Arranged in the -/+ frequency
Y2 = fft(y2); Y2 = Y2([39:64 2:27]); % Arranged in the -/+ frequency
H_est = X.*(Y1+Y2)/2; % Eq.(11.4.3)
```
%do_channel_estimation.m

clker, clf
% Read the time-domain channel response stored in a 64x2 matrix
load 'ch_complex.dat';
h=(ch_complex(:,1)+j*ch_complex(:,2)).'; % Time-domain channel response
Nfft=64; Ng=16; Nsym=Nfft+Ng; Nnull=Nsym; Nw=2*Ng; Tsym=4e-6;
A_sig=0.5; A_noise=0.01; % Amplitudes of signal and noise
Nd=120; % Remaining period of the last symbol in the previous frame
[s_preamble,Short] = short_train_seq(Nfft);
[l_preamble,Long] = long_train_seq(Nfft);
% Make a pseudo received sequence
t_frame=[A_sig*(rand(1,Nd)-0.5) zeros(1,Nnull) s_preamble l_preamble];
symbol=A_sig*(rand(1,Nd)-0.5); symbol=[symbol(end-Ng+1:end) symbol];
t_frame=[t_frame symbol]; L_frame=length(t_frame);
noise=A_noise*(rand(1,L_frame)-0.5 +j*(rand(1,L_frame)-0.5));
r_frame=channel(t_frame,h) + noise;
True_STO = Nd+Nnull+Nsym*2+1 % Starting point of the long preamble
STO = True_STO
% STO estimation is critical to the performance of channel estimation
% To realize this, change the above line into STO=True_STO+1 or -1
H est = channel_estimate(Long,r_frame(STO+[0:159]));
% The true frequency-domain channel response is obtained from the FFT
% of the time-domain channel (impulse) response.
H = fft(h,Nfft); % k=0(1):26(27)  27(28):37(38)  38(39):63(64)
H_true = H([39:64 2:27]); % Arranged in +/- frequency k=-26:-1 1:26
 discrepancy_H_est_and_H_true = norm(H_est-H_true)/norm(H_true)
% Let's see how the channel equalizer with the estimated channel
% response (H_est) works.
X = rand(1,52)-0.5+j*(rand(1,52)-0.5); % for k=[-26:-1 1:26]
X_arranged = [0 X(27:end) zeros(1,11) X(1:26)]; % in +/- frequency
x = ifft(X_arranged); x_CP = [x(49:64) x]; % IFFT and add CP
y = channel(x_CP,h); % Channel output to an arbitrary input with CP
Y = fft(y(17:80)); % Remove CP and FFT
Ye=Y([39:64 2:27]); % Arranged in +/- frequency for k=[-26:-1 1:26]
Yeq = equalizer_in_freq(Y,H_est);
 discrepancy_X_and_Y = norm(X-Y)/norm(X) % With no channel compensation
discrepancy_X_and_Yeq = norm(X-Yeq)/norm(X) % With channel equalizer

function [y, ch_buf] = channel(x,h, ch_buf)
L x = length(x); L h = length(h); h=h(:);
if nargin<3, ch_buf = zeros(1,L_h);
else L ch_buf = length(ch_buf);
    if L ch_buf<L_h, ch_buf = [ch_buf zeros(1,L_h-L_h)];
    else ch_buf = ch_buf(1:L_h);
end
for n=1:L x, ch_buf=[x(n) ch_buf(1:end-1)]; y(n)=ch_buf*h; end
function Y eq = equalizer_in_freq(Y,H_est)
H_est(find(abs(H_est)<1e-6))=1; Y eq = Y ./H_est; % Eq.(11.4.4)

%ch_complex.dat
0.4923667628304478349754 -0.3721824742433228472294
-0.5175114648023440817804
-0.02495648028624753096 -0.3043313718301440817804
.................
11.4 Channel Estimation and Equalization

Figure 11.13 Simulink model for channel estimation and compensation ("channel_estimation_sim.mdl")
INDEX

A
adaptive equalizer (ADE), 154-155
address decoding circuit, 277
Alamouti, 313
A-law, 92
analog frequency, 20
analytic signal, 30, 33
angle modulation, 82
antipodal (bipolar) signaling, 114, 122
ASK (amplitude shift keying), 169
asynchronous, 78
autocorrelation, 29, 30, 50, 55, 58
autocorrelation function, 50
autocovariance function, 50
average signal energy of QAM symbol, 196
AWGN (additive white Gaussian noise), 108, 136

B
balance property (of PN), 340
bandpass Gaussian noise, 49, 62
bandwidth efficiency, 200, 203, 268
Bayes’ rule, 40
BCJR (Bahl-Cocke-Jelinek-Raviv), 301
BCH (Bose-Chaudhuri-Hocquenghem) code, 284, 322
belief propagation algorithm, see BPA
Bessel function, 24, 26, 50
BFSK (binary phase shift keying) signal, 22
binary symmetric channel (BSC), 264, 321
bi-orthogonal signaling, 133-134
bit error probability, 129, 131
bit error rate (BER), see bit error probability
block coding, 271, 284
Blockset, 423
Communication ~, 424
Signal Processing ~, 423
BPA (belief propagation algorithm), 310
branch cost (metric), 289
BSC, see binary symmetric channel

C
capacity boundary, 268
carrier frequency offset (CFO), 362-363, 376
carrier phase recovery, 225, 233, 248-252
~ for BPSK, 248
~ for PSK, 251-252
~ for QAM, 238-242, 250
~ for QPSK, 249
carrier recovery, 225, 236, 239-240
carrier timing recovery, 252-253
Carson’s (bandwidth) rule, 36, 83
CDMA (code division multiple access), 354
central limit theorem (CLT), 47, 48
centroid, 87
CFO (carrier frequency offset), 362-363, 376
~ estimation, 365-369, 398-405
channel capacity, 265-269, 319, 321
channel coding, 263, 269
channel estimation, 379, 381, 405
channel reliability, 300, 309
Chebyshev inequality, 43
class node (c-node), 310-313
chip interval, 347
circular correlation, 69
coarse CFO estimate, 365-366, 404
code division multiple access (CDMA), 354
code efficiency, 257
code rate, 265, 285
code vector, 273, 309
codeword, 257, 265, 270-276, 279-281, 287
codeword matrix, 270, 273
coding gain, 317-318
co-error function, see complementary error function
coherent, 75, 82, 171, 180-183, 190, 206
colored noise, 53
Communication Blockset, 424
Communication Toolbox, 284
compansion, 93
complementary error (co-error) function, 43, 118
complex envelope, 33
conditional entropy, 264
conditional expectation, 47
conditional probability, 40
conditional probability density function, 41
conjugate symmetry, 13
constraint length, 285, 286
constraint length vector, 292, 293, 325, 326, 328
conventional ISI signaling, 143
conventional AM, 75
convolution, 14
convolutional code, 285, 287
convolutional coding, 285
convolutional encoder, 285, 289, 325
correlation, 14, 29, 44, 375
  circular ~, 69
correlation coefficient, 44
correlation property (of PN), 340
correlator, 112-113
Costas loop, 236-237, 244, 249
covariance, 44
covariance matrix, 45
CP (cyclic prefix), 358, 365, 371, 377, 393
crosscorrelation, 50, 52, 58, 111, 270
crosscovariance function, 50
crosscovariance, 45
crosscovariance function, 50
crossover probability, 271, 272, 277
CTFS (continuous-time Fourier series), 1, 407, 408
CTFT (continuous-time Fourier transform), 7, 27, 407, 409-411
cycle, 310
cyclic code, 280, 284
cyclic decoder, 282, 283
cyclic encoder, 282, 283
cyclic prefix (CP), 358, 365, 371, 377, 393

d
debugging (MATLAB), 419
decision-feedback equalizer (DFE), 155-158
deinterleaving, 382
delta modulation (DM), 100-101
delta-sigma modulation, 105
depuncturing, 384
despreading, 346, 347
development, 43
deviation, 43
defe (decision-feedback equalizer), 155-158
DFS (discrete Fourier series), 19
df (discrete Fourier transform), 19
differential PSK (DPSK), 190-195
differential PCM (DPCM), 97-99
differential space-time block code, see DSTBC
differentiation w.r.t. a vector, 414
digital frequency, 20
discrete memoryless channel (DMC), 263
DM (delta modulation), 100-101
DMC (discrete memoryless channel), 264
Doppler effect, 61
DPCM (differential pulse code modulation), 97-99
DS (direct sequence)-SS, 345-349
DSB (double sideband)-AM, 71
DSB-AMSC, 71-72
dsm (delta-sigma modulation), 105
DSTBC (differential space-time block code), 314, 333
DTFS (discrete-time Fourier series), 19
DTFT (discrete-time Fourier transform), 18, 27
duality, 11, 17
duobinary precoding, 146
duobinary signaling, 143-144, 146-147, 165-167
d
early-late gate timing recovery, 241-243
energy-type signal, 29
entropy, 256, 319-320
envelope detector, 76, 173, 182
equalizer, 148-158, 167
adaptive ~, 154
decision-feedback ~, 155
minimum mean-square error (MMSE) ~, 151
zero-forcing ~ (ZFE), 148
equalization 379
ergodic, 52
error correcting capability, 272, 284, 322
error detecting capability, 272
error floor, 308
error function, 43
error pattern, 274-275, 281-282, 322-323
error probability, 122, 123, 201, 202
  ~ for antipodal signaling, 117
  ~ for BASK with coherent detection, 178
  ~ for BASK with non-coherent detection, 174
  ~ for bi-orthogonal signaling, 134-135
  ~ for BPSK, 187
  ~ for DPSK (differential PSK), 193
  ~ for FSK with coherent detection, 179
  ~ for FSK with non-coherent detection, 182-183
  ~ for multidimensional signaling, 130-131
  ~ for multi-level signaling, 128-129
  ~ for OOK signaling, 118
  ~ for orthogonal signaling, 121
  ~ for PAM, 171
  ~ for PSK, 188
  ~ for QAM, 197
ESD (energy spectral density), 29
EUREKA-147 DAB, 372
expectation, 43
extended Golay code, 279
extrinsic information, 308
f
fading, 60
  ~ amplitude, 301
  fast ~, 61
  (frequency-)flat ~, 61
  (frequency-)selective ~, 61
  Raleigh ~, 60
  Rician ~, 60
  slow ~, 61
fast FH-SS, 350-352
feedback shift register, 280, 281, 283, 337-341
FFO (fractional frequency offset), 376, 377
FH (frequency hopping)-SS, 350-353
   fast ~, 350-352
   slow ~, 350-351, 355
filtering, 57
fine CFO estimate, 365-366, 405
fine frame synchronization, 376
finite pulsewidth sampler, 15
flat fading, 61
FM (frequency modulation), 82-83
fractional frequency offset (FFO), 376, 377
frame synchronization, 373, 376
free distance, 289
code, 57
frequency aliasing, 28
frequency deviation ratio, 36
frequency modulation (FM), 82-83
frequency offset, 376-377
frequency response, 7, 14, 18, 58, 63
   ~ of a discrete-time LTI system, 63
frequency-selective fading, 61
frequency shifting, 6, 14
FSK (frequency shift keying), 170, 178-186
function of a random variable, 42
fundamental frequency, 20

G
Gallager, 309
Gaussian, 43-46, 49, 205, 235
Gaussian noise, 49, 205
   sampled ~, 57
generator matrix, 272-274, 278, 285, 325, 327
generator polynomial, 280, 285
generator sequence, 285
Golay code, 279
Gold code, 340-343
Gold sequence, see Gold code
gradient, 154, 414
granular (idling) noise, 100-101
guard interval, 358, 362, 365, 387, 398

H
Hadamard matrix, 270
Hamming code, 278, 284, 321
Hamming distance, 272
Hamming weight, 272, 287
hard value, 298
hard decision, 289, 292
Hermitian symmetry, 13
Hilbert transform, 32, 37, 78
Huffman code, 257-259, 320

I
ideal LPF (lowpass filter), 140
IEEE Standard 802.11a, 386-388
timing-related parameters, 387
IFO (integral frequency offset), 376
impulse function, 5
impulse sequence, 6
impulse train, 5
independent, 41-42, 45
inner (or dot or scalar) product, 121
in-phase component, 33, 49
instant sampler, 15
integral frequency offset (IFO), 376
interleaving, 382
ISI (intersymbol interference), 139
ISI free condition, 141-142

J
Jacobian, 42, 43, 49, 412
jamming 347
jointly normal distribution, 44
joint probability, 40
joint probability density function, 41

K
Kasami sequence, 341-342

L
L-value, 298, 305
Laplace transform, 412
LAPP (log a posteriori probability), 301
LDPC (low-density parity-check)
   ~ code, 284, 309
   ~ decoding, 310
   ~ encoding, 309
least mean square (LMS), 154
Lempel-Ziv-Welch (LZW) coding, 260
Lempel-Ziv-Welch (LZW) decoding, 261-262
LF (loop filter), 226-230
LFSR, see linear feedback shift register
line code, 65
linear block coding, 271, 284
linear feedback shift register (LFSR),
   281, 283, 337-338, 340-341
Lloyd-Max, 89-90
LLR (log-likelihood ratio), 298, 301, 305
   conditioned ~, 301, 308, 310
LMS (least mean square), 154
log-MAP decoding, 298
long preamble, 363-365
long training sequence, see long preamble
loop filter (LF), 226-230
lowpass equivalent, 33
LSSB (lower single sideband), 80
LZW (Lempel-Ziv-Welch) coding, 260
LZW (Lempel-Ziv-Welch) decoding, 262

M
m-sequence (maximal length sequence), 339
preferred pairs of ~s, 340
MAP (maximum a posteriori probability), 298, 301
marginal probability density function, 41
Mason’s gain formula, 287
matched filter, 110-111
MATLAB Command Window, 418
MATLAB Editor Window, 418
MATLAB introduction, 417
maximal length sequence (m-sequence), 339
maximum a posteriori probability (MAP), 298
maximum likelihood estimate (MLE), 231
mean, 43, 55
mean function, 50
mean square quantization error (MSQE), 87
message passing algorithm (MPA), 310
minimum distance, 123, 271, 272, 289
minimum mean-square error equalizer, see MMSEE
minimum-shift keying (MSK), 211-215
minimum tone spacing, 179, 181
MLE (maximum likelihood estimate), 231
MMSEE (minimum mean-square error equalizer), 151, 153
modified duobinary signaling, 147-148
modulation, 6, 14, 71
modulation order, 127, 204
modulation efficiency, 76
MPA (message passing algorithm), 310
MSK (minimum-shift keying), 211-215
MSQE (mean square quantization error), 87
μ-law, 92
multi-amplitude, 127, 132
multi-dimensional (orthogonal), 129, 132
multi-level (multi-amplitude), 127, 132
multi-path channel, 59
mutual information, 265

N
NDA-ELD (nondata-aided early-late delay), 244-246
node cost (metric), 289
noise power (or variance), 123
noise PSD, 55
noncoherent, 78, 171-173, 181-183, 190, 193, 206
nondata-aided early-late delay (NDA-ELD), 244-246
nonuniform quantization, 89-91, 94
normal convergence theorem, 47
normal distribution, 43, 44
Nyquist band, 143, 144
Nyquist bandwidth constraint, 140
Nyquist frequency, 28

O
OFDM, 357-358
OFDM receiver, 392
OFDM symbol, 358, 362, 388, 394
OFDM symbol duration, 371, 375, 388
OFDM transmitter, 392
offset QPSK (OQPSK), 207-208
on-off keying (OOK), 118, 122
OOK (on-off keying), 118
OQPSK (offset QPSK), 207-208
orthogonal, 45, 119, 121, 122, 129, 179
orthogonal signaling, 119-121, 126, 132
orthogonality, 397, 398

P
PAM (pulse amplitude modulation), 15, 17, 127
parity check matrix, 274, 278, 279, 309
Parseval’s relation, 18
partial response signaling, 143
path metric, 305, 308
PCM (pulse code modulation), 95-97
perfect (code), 279
phase-locked loop, see PLL
phase modulation (PM), 82-83
phase offset, 73
phase shift keying (PSK), 170, 187-190
phase tracker, 403
physically realizable, 141
π/4-shifted QPSK (quadrature PSK), 209-210
pilot, 365, 387, 388
pilot polarity sequence, 387-388
PLCP (physical layer convergence procedure), 386
PLL (phase-locked loop), 226-232, 247
PM (phase modulation), 82-83
PN (pseudo or pseudo-random noise), 337-338
posterior probability, 40, 231
power efficiency, 200, 203
power spectral density (PSD), 29, 53, 57-58
power theorem, 18
power-type signal, 29
PPDU (PHY packet data unit), 386
PRBS (pseudo-random binary sequence), 337
preamble, 365
precoding, 146
predictor, 98
pre-envelope signal, 30
prefix, see CP
principal frequency range, 21
priori information, 298, 307, 308, 310
priori probability, 40
probability, 39
probability density function, 41
probability distribution function, 41
probability transition matrix, 264
processing gain, 347, 350
PSD (power spectral density), 29, 53, 57-58
pseudo noise (PN), 337-338, 387
PSK (phase shift keying), 170, 187-190
pulse amplitude modulation (PAM), 15, 17, 127
pulse code modulation (PCM), 95-97
puncturing, 298, 384
Q
QAM (quadrature amplitude modulation), 195-200
QPSK (quadrature PSK), 189-190
$\pi/4$-shifted $\sim$, 209-210
offset $\sim$ (OQPSK), 207-208
staggered $\sim$, see OQPSK
quadrature amplitude modulation (QAM), 195-200
quadrature carrier, 197
quadrature component, 33, 49
quadrature correlator, 197
quantization, 87-91, 94
R
raised-cosine (RC) filter, 160-164
raised-cosine frequency response, 141-142, 159
raised-cosine impulse response, 142, 160
random process, 49
Rayleigh fading, 60
Rayleigh probability density function, 50, 174, 182
$RC$ filter, 109-110
real convolution, 14
rectangular pulse, 8
rectangular wave, 2
recursive systematic convolutional encoder, 298
Reed-Solomon (RS) code, 284
resolution frequency, 20
Rice probability density function, 50, 62, 174, 182
Rician fading, 60
roll-off factor, 141
RSC encoder, 298
run property (of PN), 340
S
sampling frequency, 27
sampling theorem, 27, 28
scrambler, 387, 388
Shannon-Hartley channel capacity theorem, 267
Shannon limit, 268
short preamble, 363-365
short training sequence, see short preamble
sigma-delta modulation, see delta-sigma modulation
signal constellation diagram, 122
signal correlator, 112
Signal Processing Blockset, 423
signal space, 121, 122
signal-to-quantization noise ratio (SQNR), 87
Simulink Library Browser Window, 422
sinusoidal FM (frequency-modulation) signal, 24
slope overload distortion, 100-101
slow fading, 61
slow FH-SS, 355
soft value, 297
soft-decision, 289, 293
soft-in/soft-output Viterbi algorithm, see SOVA
source coding, 257, 263
SOVA (soft-in/soft-output Viterbi algorithm), 305
SPA (sum-product algorithm), 310, 311
space-time block code (STBC), 313
spread-spectrum (SS), 337
spreading, 346, 347
SQNR (signal-to-quantization noise ratio), 87
square-root raised-cosine (SRRC) filter, 162-164
square-root raised-cosine impulse response, 163
squaring loop, 233-235, 248
squaring timing recovery, 252
SS (spread-spectrum), 337
DS (direct sequence) $\sim$, 345-349
FH (frequency hopping) $\sim$, 350-353, 355
SSB (single-sideband)-AM, 78
staggered QPSK, see OQPSK
state diagram for convolution encoder, 288
stationary, 51-52
STBC (space-time block code), 313
steepest descent method, 154
STO (symbol time offset), 362-363
$\sim$ estimation, 370-372, 382, 393, 399
stochastic process, 49
strict-sense stationary, 51
subcarrier, 358
sum-product algorithm, see SPA
suppressed carrier, 71
symbol error probability, 128, 130, 131
symbol error rate (SER).
see symbol error probability
symbol synchronization, 225, 241
symbol time offset (STO), 362-363
synchronous, 75, 82
synchronization, 241, 365, 372, 393
syndrome, 274-277, 281-282, 309, 322
systematic, 274, 298
Tanner graph, 310
TCM, 294-297, 329-334
time shifting, 6, 14
timing recovery, 241-245
traceback depth, 292, 293
transmission bandwidth, 75, 80
transmission mode, 372
transmitted carrier, 75
trellis, 288-290, 293, 304
trellis-coded modulation, see TCM
triangular pulse, 8
triangular wave, 2
turbo code, 298-308
turbo decoding, see log-MAP or SOVA

Uncorrelated, 44, 45
Ungerboeck TCM encoder, 329-330
uniform distribution, 43, 46
uniform number, 46
uniform quantization, 88
unit signal waveform, 108

USSB (upper single sideband), 78

Variable node (v-node), 310, 311, 312
variance, 43, 55
VCO, 226-232, 247
vector quantization, 103-104
Viterbi
~ (decoding) algorithm, 289-293
~ decoding, 325
voltage-controlled oscillator, see VCO

Waveform coding, 270
weight, 271
white noise, 53, 55, 62, 108, 136, 205
wide-sense stationary, 51

Z

z-transform, 412-413
zero-forcing equalizer (ZFE), 148, 151

Index for MATLAB Routines (*: MATLAB built-in function)

<table>
<thead>
<tr>
<th>MATLAB routine name</th>
<th>Description</th>
<th>Page number</th>
</tr>
</thead>
<tbody>
<tr>
<td>ADC()</td>
<td>Analog-to-digital conversion</td>
<td>93</td>
</tr>
<tr>
<td>ade()</td>
<td>tunes the adaptive equalizer</td>
<td>154</td>
</tr>
<tr>
<td>Alaw()</td>
<td>$A$ -law (Eq. (4.1.8a))</td>
<td>92</td>
</tr>
<tr>
<td>Alaw_inv()</td>
<td>$A^{-1}$-law (Eq. (4.1.8b))</td>
<td>92</td>
</tr>
<tr>
<td>awgn(*)</td>
<td>Additive white Gaussian noise</td>
<td>136</td>
</tr>
<tr>
<td>awg(<em>*)</em></td>
<td>Additive white Gaussian noise</td>
<td>136</td>
</tr>
<tr>
<td>bchgenpoly()</td>
<td>makes the BCH code generator polynomial for given (N,K)</td>
<td>321</td>
</tr>
<tr>
<td>berawgn()</td>
<td>Probability of bit error for various modulation schemes</td>
<td>202</td>
</tr>
<tr>
<td>bercoding()</td>
<td>Probability of bit error for various coding schemes</td>
<td>279</td>
</tr>
<tr>
<td>bilinear()</td>
<td>bilinear transformation (optionally with prewarping)</td>
<td>228-229</td>
</tr>
<tr>
<td>channel()</td>
<td>simulates the channel effect using convolution</td>
<td>380</td>
</tr>
<tr>
<td>channel_estimate()</td>
<td>finds the estimate of channel (frequency) response</td>
<td>379</td>
</tr>
<tr>
<td>coarse_CFO_estimate()</td>
<td>finds the coarse estimate of CFO</td>
<td>367</td>
</tr>
<tr>
<td>combis()</td>
<td>makes an error pattern matrix</td>
<td>275</td>
</tr>
<tr>
<td>compensate_CFO()</td>
<td>compensate the CFO</td>
<td>368</td>
</tr>
<tr>
<td>compensate_phase()</td>
<td>compensate the carrier phase</td>
<td>365</td>
</tr>
<tr>
<td>conv_encoder()</td>
<td>convolutional encoding</td>
<td>286</td>
</tr>
<tr>
<td>convenc()</td>
<td>convolutional encoding</td>
<td>293, 325</td>
</tr>
<tr>
<td>corr_circular()</td>
<td>Circular (or cyclic or periodic) correlation</td>
<td>69, 343</td>
</tr>
<tr>
<td>CTFS()</td>
<td>CTFS coefficients together with the reconstruction</td>
<td>4, 5</td>
</tr>
<tr>
<td>CFTFT()</td>
<td>CFTFT spectrum together with the inverse CFTFT</td>
<td>9</td>
</tr>
<tr>
<td>cyclic_decoder()</td>
<td>Cyclic decoder</td>
<td>282</td>
</tr>
<tr>
<td>cyclic_encoder()</td>
<td>Cyclic encoder</td>
<td>282</td>
</tr>
<tr>
<td>cyclpoly()</td>
<td>makes the cyclic code generator polynomial for given (N,K)</td>
<td>281, 284</td>
</tr>
<tr>
<td>MATLAB routine name</td>
<td>Description</td>
<td>Page number</td>
</tr>
<tr>
<td>---------------------</td>
<td>-------------</td>
<td>-------------</td>
</tr>
<tr>
<td>de01e01</td>
<td>plots the CTFS spectra of rectangular/triangular waves</td>
<td>4</td>
</tr>
<tr>
<td>de01e03</td>
<td>plots the CTFT spectra of rectangular/triangular pulses</td>
<td>9</td>
</tr>
<tr>
<td>de01e16</td>
<td>plots the FFT spectrum of a FSK signal</td>
<td>23</td>
</tr>
<tr>
<td>de01e17</td>
<td>plots the spectrum of a sinusoidal-modulated FM signal</td>
<td>25</td>
</tr>
<tr>
<td>de01p02</td>
<td>Lowpass equivalent of a bandpass signal</td>
<td>38</td>
</tr>
<tr>
<td>de0109_1</td>
<td>plots the Hilbert transform of a sinusoidal wave</td>
<td>31</td>
</tr>
<tr>
<td>de0109_2</td>
<td>Lowpass equivalent of a bandpass signal</td>
<td>35</td>
</tr>
<tr>
<td>de02e02a</td>
<td>plots the distribution of a uniform noise</td>
<td>46</td>
</tr>
<tr>
<td>de02e02b</td>
<td>plots the distribution of a Gaussian noise</td>
<td>47</td>
</tr>
<tr>
<td>de02e03</td>
<td>checks the validity of central limit theorem (CLT)</td>
<td>48</td>
</tr>
<tr>
<td>de02e05</td>
<td>plots the autocorrelation and histogram of a white noise</td>
<td>56</td>
</tr>
<tr>
<td>de02p01</td>
<td>plots a white noise together with its correlation and PSD</td>
<td>54</td>
</tr>
<tr>
<td>de02p01</td>
<td>plots the distribution of a bandpass noise with Rice pdf</td>
<td>62</td>
</tr>
<tr>
<td>de04e03</td>
<td>Nonuniform quantization considering relative errors</td>
<td>93</td>
</tr>
<tr>
<td>de0501</td>
<td>simulates binary communications with RC/matched filters</td>
<td>113</td>
</tr>
<tr>
<td>de05f17</td>
<td>plots the theoretical BER curve for orthogonal signaling</td>
<td>131</td>
</tr>
<tr>
<td>de07p01</td>
<td>simulates a linear combination of Gaussian noises</td>
<td>205</td>
</tr>
<tr>
<td>de07f02</td>
<td>Table 7.2 (SNRdB for each signaling to achieve BER=10^{-5})</td>
<td>203</td>
</tr>
<tr>
<td>de09e05</td>
<td>constructs the codeword matrix for a linear block code</td>
<td>273</td>
</tr>
<tr>
<td>de09p07</td>
<td>shows some examples of using convenc() and vitdec()</td>
<td>324</td>
</tr>
<tr>
<td>deci2bin1()</td>
<td>Decimal-to-binary conversion</td>
<td>272</td>
</tr>
<tr>
<td>decode()</td>
<td>Decoding</td>
<td>284</td>
</tr>
<tr>
<td>decoder()</td>
<td>Decoding of a signal value by the given code table</td>
<td>96</td>
</tr>
<tr>
<td>dem_PSK_or_QAM()</td>
<td>performs the PSK or QAM demodulation</td>
<td>390</td>
</tr>
<tr>
<td>depuncture()</td>
<td>Depuncturing</td>
<td>384</td>
</tr>
<tr>
<td>detector_FSK()</td>
<td>Detection of FSK signals</td>
<td>217</td>
</tr>
<tr>
<td>detector_MSK()</td>
<td>Detection of MSK signals</td>
<td>222</td>
</tr>
<tr>
<td>detector_PSK()</td>
<td>Detection of PSK signals</td>
<td>220</td>
</tr>
<tr>
<td>dfe()</td>
<td>tunes the decision feedback equalizer (DFE)</td>
<td>156</td>
</tr>
<tr>
<td>do_ade</td>
<td>simulates the adaptive-equalizer (ADE)</td>
<td>155</td>
</tr>
<tr>
<td>do_BCH_BPSK_sim</td>
<td>runs the Simulink model &quot;BCH_BPSK_sim&quot;</td>
<td>323</td>
</tr>
<tr>
<td>do_CFO</td>
<td>simulates the CFO estimation and compensation</td>
<td>367</td>
</tr>
<tr>
<td>do_CFO_PHO_STO</td>
<td>simulates the CFO/PHO/STO estimation and compensation</td>
<td>399-401</td>
</tr>
<tr>
<td>do_channel_estimation</td>
<td>tries using channel_estimate() for channel estimation</td>
<td>380</td>
</tr>
<tr>
<td>do_cyclic_code</td>
<td>tries with a cyclic code</td>
<td>281</td>
</tr>
<tr>
<td>do_cyclic_codes</td>
<td>tries with cyclic codes</td>
<td>323</td>
</tr>
<tr>
<td>do_dfe</td>
<td>simulates the decision feedback equalizer (DFE)</td>
<td>156</td>
</tr>
<tr>
<td>do_FSK_sim</td>
<td>runs the Simulink model “FSK_passband_sim”</td>
<td>217</td>
</tr>
<tr>
<td>do_Hamming_code74</td>
<td>finds the BER performance of the Hamming (7,4) code</td>
<td>276</td>
</tr>
<tr>
<td>do_interleaving</td>
<td>tests the MATLAB routines interleaving() and deinterleaving()</td>
<td>383</td>
</tr>
<tr>
<td>do_mmsee</td>
<td>simulates the minimum mean-square error equalizer</td>
<td>152</td>
</tr>
<tr>
<td>do_MSK_sim</td>
<td>runs the Simulink model “MSK_passband_sim”</td>
<td>222</td>
</tr>
<tr>
<td>do_OFDM0</td>
<td>simulates a basic OFDM system</td>
<td>359</td>
</tr>
<tr>
<td>do_OFDM1</td>
<td>simulates an OFDM system (IEEE Std 802.11a)</td>
<td>389</td>
</tr>
<tr>
<td>do_PLL</td>
<td>simulates a PLL (phase-locked loop) system</td>
<td>229</td>
</tr>
<tr>
<td>do_PNG</td>
<td>uses MATLAB program ‘png()’ and runs Simulink model “PN_generator_sim.mdl” to generate a PN sequence</td>
<td>343</td>
</tr>
<tr>
<td>MATLAB routine name</td>
<td>Description</td>
<td>Page number</td>
</tr>
<tr>
<td>-----------------------------------</td>
<td>--------------------------------------------------------------------------------------------------------------------------------------------</td>
<td>-------------</td>
</tr>
<tr>
<td>do_PSK_sim</td>
<td>runs the Simulink model “PSK_passband_sim”</td>
<td>220</td>
</tr>
<tr>
<td>do_puncture</td>
<td>tests the MATLAB functions puncture() and depuncture()</td>
<td>385</td>
</tr>
<tr>
<td>do_QAM_carrier_recovery</td>
<td>simulates QAM with decision-feedback carrier recovery</td>
<td>239</td>
</tr>
<tr>
<td>do_QAM_sim</td>
<td>runs the Simulink model “QAM_passband_sim”</td>
<td>222</td>
</tr>
<tr>
<td>do_QPSK_Costas</td>
<td>simulates QPSK using Costas loop for carrier phase recovery</td>
<td>237</td>
</tr>
<tr>
<td>do_QPSK_Costas_earlylate</td>
<td>simulates QPSK using Costas loop and Early-Late algorithm for carrier phase recovery and timing recovery, respectively</td>
<td>244-245</td>
</tr>
<tr>
<td>do_rcos1</td>
<td>draws the impulse/frequency responses of a raised cosine filter</td>
<td>161</td>
</tr>
<tr>
<td>do_rcos2</td>
<td>performs the two cascaded square-root raised-cosine filtering</td>
<td>162</td>
</tr>
<tr>
<td>do_square_filter_clock</td>
<td>Square-law bit synchronizer</td>
<td>68</td>
</tr>
<tr>
<td>do_squaring_loop</td>
<td>detects a squaring loop to get a reference signal from BPSK</td>
<td>234</td>
</tr>
<tr>
<td>do_STO</td>
<td>simulates the STO estimation/compensation</td>
<td>393-395</td>
</tr>
<tr>
<td>do_SYM_sync_earlylate</td>
<td>simulates the early late sampling time control for symbol sync</td>
<td>243</td>
</tr>
<tr>
<td>do_sync_for_DMB</td>
<td>simulates the frame synchronization and CFO estimation</td>
<td>378</td>
</tr>
<tr>
<td>do_sync_w_double_window</td>
<td>simulates the estimation of OFDM frame and symbol start times using the double sliding window</td>
<td>374</td>
</tr>
<tr>
<td>do_TCM_8PSK</td>
<td>runs MATLAB routine ‘TCM’ and Simulink model</td>
<td>330</td>
</tr>
<tr>
<td>do_vector_quantization</td>
<td>simulates the vector quantization</td>
<td>103</td>
</tr>
<tr>
<td>do_vitdecoder</td>
<td>tries using vit_decoder() and vitdec()</td>
<td>290</td>
</tr>
<tr>
<td>do_vitdecoder1</td>
<td>tries the various usages of vitdec()</td>
<td>292</td>
</tr>
<tr>
<td>do_Viterbi_QAM</td>
<td>runs MATLAB routine ‘Viterbi_QAM.m’ together with Simulink model ‘Viterbi_QAM_sim.mdl’</td>
<td>327</td>
</tr>
<tr>
<td>do_zfe</td>
<td>simulates the zero forcing equalizer (ZFE)</td>
<td>150</td>
</tr>
<tr>
<td>DS_SS</td>
<td>simulates BPSK with DS-SS(spread spectrum)</td>
<td>348</td>
</tr>
<tr>
<td>encode()</td>
<td>Encoding</td>
<td>284</td>
</tr>
<tr>
<td>equalizer_in_freq()</td>
<td>simulates the frequency-domain equalizer</td>
<td>380</td>
</tr>
<tr>
<td>FH_SS</td>
<td>simulates BFSK with fast FH-SS(spread spectrum)</td>
<td>352</td>
</tr>
<tr>
<td>FH_SS2</td>
<td>simulates BFSK with slow FH-SS(spread spectrum)</td>
<td>355</td>
</tr>
<tr>
<td>fine_CFO_estimate()</td>
<td>finds the fine estimate of CFO</td>
<td>367</td>
</tr>
<tr>
<td>freqz()</td>
<td>Frequency response of a discrete-time system</td>
<td>63</td>
</tr>
<tr>
<td>Gauss_Hermite()</td>
<td>Integration using the Gauss-Hermite quadrature</td>
<td>132</td>
</tr>
<tr>
<td>GI_and_orthogonality</td>
<td>tests the orthogonality of an OFDM signal</td>
<td>398</td>
</tr>
<tr>
<td>gm2gM()</td>
<td>PNG polynomial into one used by Simulink</td>
<td>344</td>
</tr>
<tr>
<td>gray_code()</td>
<td>Gray coding</td>
<td>327</td>
</tr>
<tr>
<td>Hamm_gen()</td>
<td>makes the parity-check/generator matrices for Hamming code</td>
<td>278</td>
</tr>
<tr>
<td>Hammgen()</td>
<td>makes the parity-check/generator matrices for Hamming code</td>
<td>278</td>
</tr>
<tr>
<td>hilbert()</td>
<td>analytic signal with Hilbert transform as the imaginary part holds a histogram</td>
<td>76-77, 83</td>
</tr>
<tr>
<td>hist()</td>
<td>plots a histogram</td>
<td>46-48, 56</td>
</tr>
<tr>
<td>Huffman_code()</td>
<td>Huffman code generator for a given symbol probability vector</td>
<td>258</td>
</tr>
<tr>
<td>IFO_estimate()</td>
<td>finds the estimate of IFO (integral frequency offset)</td>
<td>377</td>
</tr>
<tr>
<td>interleaving()</td>
<td>Interleaving</td>
<td>382</td>
</tr>
<tr>
<td>Jkb()</td>
<td>The first kind of kth-order Bessel function</td>
<td>25</td>
</tr>
<tr>
<td>LDPC_decoder()</td>
<td>Low-density parity-check (LDPC) decoder</td>
<td>311</td>
</tr>
<tr>
<td>LDPC_demo</td>
<td>simulates an LDPC coding and decoding</td>
<td>312</td>
</tr>
<tr>
<td>logmap()</td>
<td>Log-MAP algorithm for turbo coding</td>
<td>302</td>
</tr>
<tr>
<td>long_train_seq()</td>
<td>generates the long training sequence for 802.11a</td>
<td>364</td>
</tr>
<tr>
<td>MATLAB routine name</td>
<td>Description</td>
<td>Page number</td>
</tr>
<tr>
<td>---------------------</td>
<td>-------------</td>
<td>-------------</td>
</tr>
<tr>
<td>LZW_coding()</td>
<td>performs the LZW coding on an input symbol sequence</td>
<td>260</td>
</tr>
<tr>
<td>LZW decoding()</td>
<td>performs the LZW decoding on an input coded sequence</td>
<td>262</td>
</tr>
<tr>
<td>mmsee()</td>
<td>tunes the minimum mean-square error equalizer (MMSEE)</td>
<td>152</td>
</tr>
<tr>
<td>mod_PSK_or_QAM()</td>
<td>performs the PSK or QAM modulation</td>
<td>390</td>
</tr>
<tr>
<td>mulaw()</td>
<td>$\mu$-law (Eq. (4.1.7a))</td>
<td>92</td>
</tr>
<tr>
<td>mulaw_inv()</td>
<td>$\mu^{-1}$-law (Eq. (4.1.7b))</td>
<td>92</td>
</tr>
<tr>
<td>nextpow2()</td>
<td>nextpow2(N) returns the first P such that $2^P \geq</td>
<td>N</td>
</tr>
<tr>
<td>OFDM_parameters()</td>
<td>set the OFDM parameters for IEEE Std 802.11a</td>
<td>388</td>
</tr>
<tr>
<td>phase_from_pilot()</td>
<td>finds the OFDM carrier phase estimate based on the pilot</td>
<td>365</td>
</tr>
<tr>
<td>phase_ref_symbol()</td>
<td>generates the PRS for EUREKA-147 DAB</td>
<td>377</td>
</tr>
<tr>
<td>plot_ds_ss</td>
<td>plots the signals in a DS-SS (spread spectrum) system</td>
<td>345</td>
</tr>
<tr>
<td>plot_MOD()</td>
<td>plots the signals appearing in a specified modulation process</td>
<td>74</td>
</tr>
<tr>
<td>PNG()</td>
<td>Pseudo-noise generator</td>
<td>338</td>
</tr>
<tr>
<td>poly2trellis()</td>
<td>makes a trellis structure for a given generator polynomial</td>
<td>292-293</td>
</tr>
<tr>
<td>prdctr()</td>
<td>One-step-ahead predictor based on RLSE</td>
<td>99</td>
</tr>
<tr>
<td>prob_err_msg_bit()</td>
<td>Theoretical message bit error probability by Eq. (9.4.11)</td>
<td>276</td>
</tr>
<tr>
<td>prob_error()</td>
<td>Probability of error for various signaling</td>
<td>202</td>
</tr>
<tr>
<td>PSK_slicer()</td>
<td>slices each of PSK signals into the regular constellation points</td>
<td>516</td>
</tr>
<tr>
<td>puncture()</td>
<td>Puncturing</td>
<td>385</td>
</tr>
<tr>
<td>Q()</td>
<td>Complementary error (co-error) function</td>
<td>125</td>
</tr>
<tr>
<td>QAM()</td>
<td>QAM (quadrature amplitude modulation) modulation</td>
<td>327</td>
</tr>
<tr>
<td>QAM_dem()</td>
<td>QAM demodulation</td>
<td>328</td>
</tr>
<tr>
<td>quantize_nonuniform</td>
<td>Lloyd-Max nonuniform quantization</td>
<td>90</td>
</tr>
<tr>
<td>quantize_uniform</td>
<td>Uniform quantization</td>
<td>89</td>
</tr>
<tr>
<td>rand()</td>
<td>generates an array of uniformly-distributed random numbers</td>
<td>47</td>
</tr>
<tr>
<td>randerr()</td>
<td>generates a binary matrix having 0 or 1 at random positions</td>
<td>281, 284</td>
</tr>
<tr>
<td>randint()</td>
<td>generates a matrix consisting of random integers</td>
<td>284</td>
</tr>
<tr>
<td>randn()</td>
<td>generates an array of Gaussian-distributed random numbers</td>
<td>47</td>
</tr>
<tr>
<td>rcosflt()</td>
<td>designs and implements a raised-cosine filter</td>
<td>160-163</td>
</tr>
<tr>
<td>rcosine()</td>
<td>designs a raised-cosine filter</td>
<td>161-163</td>
</tr>
<tr>
<td>rD()</td>
<td>Rectangular pulse</td>
<td>9</td>
</tr>
<tr>
<td>rD_wave()</td>
<td>Rectangular wave</td>
<td>4</td>
</tr>
<tr>
<td>Rice_pdf()</td>
<td>Rice probability density function</td>
<td>62</td>
</tr>
<tr>
<td>rsdec()</td>
<td>RS (Reed-Solomon) decoding</td>
<td>284</td>
</tr>
<tr>
<td>rsenc()</td>
<td>RS (Reed-Solomon) encoding</td>
<td>284</td>
</tr>
<tr>
<td>set_CFO()</td>
<td>set up the CFO</td>
<td>368</td>
</tr>
<tr>
<td>set_parameter_11a</td>
<td>set the parameters for running “ieee_11a_CFO_est.mdl”</td>
<td>402</td>
</tr>
<tr>
<td>short_train_seq()</td>
<td>generates the short training sequence for 802.11a</td>
<td>364</td>
</tr>
<tr>
<td>sim_antipodal</td>
<td>simulates an antipodal (bipolar) signaling (in baseband)</td>
<td>125</td>
</tr>
<tr>
<td>sim_ASK_passband_coherent</td>
<td>simulates the coherent BASK in passband</td>
<td>176</td>
</tr>
<tr>
<td>sim_ASK_passband_noncoherent</td>
<td>simulates the non-coherent BASK in passband</td>
<td>177</td>
</tr>
<tr>
<td>sim_biological</td>
<td>simulates a bi-orthogonal signaling (in baseband)</td>
<td>135</td>
</tr>
<tr>
<td>sim_Delta_Sigma</td>
<td>simulates the delta-sigma ((\Delta\Sigma)) modulation</td>
<td>105</td>
</tr>
<tr>
<td>sim_DM</td>
<td>Delta modulation</td>
<td>101</td>
</tr>
<tr>
<td>sim_DPCM</td>
<td>Differential pulse code modulation</td>
<td>99</td>
</tr>
<tr>
<td>sim_DPSK_passband</td>
<td>simulates the QDPSK (differential QPSK) in passband</td>
<td>194</td>
</tr>
<tr>
<td>MATLAB routine name</td>
<td>Description</td>
<td>Page number</td>
</tr>
<tr>
<td>-------------------------------------</td>
<td>-----------------------------------------------------------------------------</td>
<td>-------------</td>
</tr>
<tr>
<td>sim_DSB_AMSC</td>
<td>simulates the DSB-AM</td>
<td>74</td>
</tr>
<tr>
<td>sim_DSB_AMTC</td>
<td>simulates the conventional AM (DSB-AMTC)</td>
<td>77</td>
</tr>
<tr>
<td>sim_FM</td>
<td>simulates the FM (frequency modulation)</td>
<td>84</td>
</tr>
<tr>
<td>sim_FSK_passband_coherent</td>
<td>simulates the coherent FSK in passband</td>
<td>184</td>
</tr>
<tr>
<td>sim_FSK_passband_noncoherent</td>
<td>simulates the noncoherent FSK in passband</td>
<td>185</td>
</tr>
<tr>
<td>sim_MSK</td>
<td>simulates the passband MSK communication</td>
<td>214</td>
</tr>
<tr>
<td>sim_QOQPSK</td>
<td>simulates the offset QPSK (OQPSK)</td>
<td>208</td>
</tr>
<tr>
<td>sim_orthogonal</td>
<td>simulates an orthogonal signaling (in baseband)</td>
<td>126</td>
</tr>
<tr>
<td>sim_PCM</td>
<td>Pulse code modulation</td>
<td>96</td>
</tr>
<tr>
<td>sim_PSK_passband</td>
<td>simulates the QPSK in passband</td>
<td>189</td>
</tr>
<tr>
<td>sim_QAM_passband</td>
<td>simulates the QAM in passband</td>
<td>199</td>
</tr>
<tr>
<td>sim_S_QDPSK</td>
<td>simulates the ( \pi/4 )-shifted QPSK</td>
<td>210</td>
</tr>
<tr>
<td>sim_SSB_AM</td>
<td>simulates the USSB/LSSB (upper/lower single sideband)-AM</td>
<td>82</td>
</tr>
<tr>
<td>sim_TCM</td>
<td>simulates 8PSK with the trellis-coded modulation</td>
<td>296</td>
</tr>
<tr>
<td>slice()</td>
<td>slices a received signal according to the given constellation</td>
<td>222</td>
</tr>
<tr>
<td>source_coding()</td>
<td>performs the source coding on an input symbol sequence</td>
<td>259</td>
</tr>
<tr>
<td>source_decoding()</td>
<td>performs the source coding on an input coded sequence</td>
<td>259</td>
</tr>
<tr>
<td>sova()</td>
<td>Soft-output Viterbi algorithm for turbo coding</td>
<td>306</td>
</tr>
<tr>
<td>state_eq()</td>
<td>State equation used in conv_encoder()</td>
<td>286</td>
</tr>
<tr>
<td>TCM()</td>
<td>Trellis-coded modulation</td>
<td>329</td>
</tr>
<tr>
<td>TCM()</td>
<td>TCM using TCM_encoder1() and TCM_decoder1()</td>
<td>334</td>
</tr>
<tr>
<td>TCM_decoder()</td>
<td>TCM decoder with a convolutional encoder touched by input</td>
<td>297</td>
</tr>
<tr>
<td>TCM_decoder1()</td>
<td>TCM decoder w. a convolutional encoder untouched by input</td>
<td>333</td>
</tr>
<tr>
<td>TCM_encoder()</td>
<td>TCM encoder with a convolutional encoder touched by input</td>
<td>296</td>
</tr>
<tr>
<td>TCM_encoder1()</td>
<td>TCM encoder w. a convolutional encoder untouched by input</td>
<td>332</td>
</tr>
<tr>
<td>TCM_state_eq1()</td>
<td>State equation used in TCM()</td>
<td>329</td>
</tr>
<tr>
<td>test_corr_circular</td>
<td>tests the MATLAB routine 'corr_circular()'</td>
<td>69</td>
</tr>
<tr>
<td>test_DSTBC_G2_PSK</td>
<td>applies the DSTBC with 3 transmit antennas for 4PSK</td>
<td>316</td>
</tr>
<tr>
<td>test_DSTBC_H4_PSK</td>
<td>applies the DSTBC with 4 transmit antennas for 4PSK</td>
<td>335</td>
</tr>
<tr>
<td>test_Rayleigh_fading</td>
<td>Rayleigh fading effect</td>
<td>60</td>
</tr>
<tr>
<td>test_unwrap()</td>
<td>tests the function of unwrap()</td>
<td>84</td>
</tr>
<tr>
<td>trellis()</td>
<td>Trellis structure</td>
<td>304</td>
</tr>
<tr>
<td>tri()</td>
<td>Triangular pulse</td>
<td>9</td>
</tr>
<tr>
<td>tri_wave()</td>
<td>Triangular wave</td>
<td>4</td>
</tr>
<tr>
<td>turbo_code_demo</td>
<td>demonstrates turbo coding/decoding with logmap() or sova()</td>
<td>307</td>
</tr>
<tr>
<td>unwrap()</td>
<td>undo the radian phase wrapped into ((-\pi, \pi])</td>
<td>83-84</td>
</tr>
<tr>
<td>vector_quantization()</td>
<td>Vector quantization</td>
<td>104</td>
</tr>
<tr>
<td>vit_decoder()</td>
<td>Viterbi decoding</td>
<td>290, 291</td>
</tr>
<tr>
<td>vitdec()</td>
<td>Viterbi decoding</td>
<td>292, 293</td>
</tr>
<tr>
<td>Viterbi_QAM</td>
<td>QAM with convolution encoding and Viterbi decoding</td>
<td>326</td>
</tr>
<tr>
<td>xcorr()</td>
<td>correlation</td>
<td>53</td>
</tr>
<tr>
<td>xcorr_my()</td>
<td>correlation</td>
<td>52</td>
</tr>
<tr>
<td>zfe()</td>
<td>zero-forcing equalizer</td>
<td>149</td>
</tr>
</tbody>
</table>
## Index for Simulink Models

<table>
<thead>
<tr>
<th>Simulink model name</th>
<th>Description</th>
<th>Page number</th>
</tr>
</thead>
<tbody>
<tr>
<td>BCH_BPSK_sim</td>
<td>simulates BPSK with BCH coding</td>
<td>324</td>
</tr>
<tr>
<td>BPSK_squaring</td>
<td>simulates BPSK with squaring loop for carrier phase recovery</td>
<td>248</td>
</tr>
<tr>
<td>CFO_sim</td>
<td>simulates the CFO estimation and compensation</td>
<td>369</td>
</tr>
<tr>
<td>channel_estimation_sim</td>
<td>simulates the channel estimation for OFDM system</td>
<td>381</td>
</tr>
<tr>
<td>Delta_Sigma_sim</td>
<td>simulates the delta-sigma (ΔΣ) modulation</td>
<td>105</td>
</tr>
<tr>
<td>do_OFDM0_sim</td>
<td>simulates a basic OFDM system</td>
<td>360</td>
</tr>
<tr>
<td>do_OFDM1_sim</td>
<td>simulates an OFDM system (IEEE Std 802.11a)</td>
<td>392</td>
</tr>
<tr>
<td>DS_SS_sim</td>
<td>simulates QAM with DS-SS (spread spectrum)</td>
<td>349</td>
</tr>
<tr>
<td>DS_SS2_sim</td>
<td>simulates QAM with 2-user DS-SS (CDMA) system</td>
<td>354</td>
</tr>
<tr>
<td>FSK_passband_sim</td>
<td>simulates the passband FSK (frequency-shift keying)</td>
<td>216</td>
</tr>
<tr>
<td>ieee_11a_CFO_est</td>
<td>simulates 802.11a system with CFO estimation/compensation</td>
<td>403-404</td>
</tr>
<tr>
<td>interleaving_sim</td>
<td>tries the interleaving and deinterleaving</td>
<td>384</td>
</tr>
<tr>
<td>MSK_passband_sim</td>
<td>simulates the passband MSK (minimum-shift keying)</td>
<td>223</td>
</tr>
<tr>
<td>PLL_sim</td>
<td>simulates a PLL (phase-locked loop) system</td>
<td>247</td>
</tr>
<tr>
<td>PN_generator_sim</td>
<td>tries generating the PN/Gold/Kasami sequences</td>
<td>342</td>
</tr>
<tr>
<td>PSK_carrier_phase_timing_recovery</td>
<td>simulates PSK with carrier phase and timing recovery</td>
<td>251</td>
</tr>
<tr>
<td>PSK_passband_sim</td>
<td>simulates the passband PSK (phase-shift keying)</td>
<td>219</td>
</tr>
<tr>
<td>puncture_sim</td>
<td>tries puncturing and depuncturing</td>
<td>385</td>
</tr>
<tr>
<td>QAM_carrier_recovery</td>
<td>simulates QAM with decision-feedback carrier phase recovery</td>
<td>250</td>
</tr>
<tr>
<td>QAM_passband_sim</td>
<td>simulates the passband QAM</td>
<td>221</td>
</tr>
<tr>
<td>QPSK_Costas_sim</td>
<td>simulates BPSK with Costas loop for carrier phase recovery</td>
<td>249</td>
</tr>
<tr>
<td>scrambling_sim</td>
<td>tries using Scrambler and Descrambler blocks</td>
<td>388</td>
</tr>
<tr>
<td>SRRC_filter</td>
<td>performs the cascaded square-root raised-cosine filtering</td>
<td>164</td>
</tr>
<tr>
<td>TCM_sim</td>
<td>simulates a Ungerboeck TCM (trellis-coded modulation)</td>
<td>330</td>
</tr>
<tr>
<td>Viterbi_QAM_sim</td>
<td>QAM with convolution encoding and Viterbi decoding</td>
<td>327</td>
</tr>
</tbody>
</table>