Library of Congress Cataloging-in-Publication Data

Simon, Marvin Kenneth.
Digital communication techniques / Marvin K. Simon, Sami M. Hinedi, William C. Lindsey.
p. cm.
Includes bibliographical references and index.
Contents: v. 1. Signal design and detection
ISBN 0-13-200610-3
1. Digital communciations. I. Hinedi, Sami M. II. Lindsey, William C. III. Title.
TK5103.7.S52 1994
621.382—dc20 93-9053
CIP

Editorial/production supervision and interior design: *Harriet Tellem*
Cover design: *Aren Graphics*
Manufacturing buyer: *Alexis Heydt*
Acquisitions editor: *Paul W. Becker*
Editorial assistant: *Maureen Diana*

Prentice-Hall, Inc.
A Paramount Communciations Company
Englewood Cliffs, New Jersey 07632

The publisher offers discounts on this book when ordered in bulk quantities:
Corporate Sales Department
PTR Prentice Hall
113 Sylvan Avenue
Englewood Cliffs, NJ 07632

Phone: 201-592-2863
FAX: 201-592-2249

Printed in the United States of America
10 9 8 7 6 5 4 3 2 1

ISBN: 0-13-200610-3

Prentice-Hall International (UK) Limited, *London*
Prentice-Hall of Australia Pty. Limited, *Sydney*
Prentice-Hall Canada Inc., *Toronto*
Prentice-Hall Hispanoamericana, S.A., *Mexico*
Prentice-Hall of India Private Limited, *New Delhi*
Prentice-Hall of Japan, Inc., *Tokyo*
Simon & Schuster Asia Pte. Ltd., *Singapore*
Editora Prentice-Hall do Brasil, Ltda., *Rio de Janeiro*

Marvin K. Simon
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2.2.3 Biphase (Manchester) Baseband Signaling

Here the two elementary signals are defined by

$$s_1(t) = \begin{cases} A; & 0 \le t \le T/2 \\ -A; & T/2 \le t \le T \end{cases} \tag{2.66}$$
$$s_2(t) = -s_1(t)$$

Substituting the Fourier transform of (2.66) into (2.59) gives

$$\frac{S_m(f)}{E} = \frac{1}{T}(1-2p)^2 \sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} \left(\frac{2}{n\pi}\right)^2 \delta\left(f - \frac{n}{T}\right) + 4p(1-p)\left[\frac{\sin^4(\pi f T/2)}{(\pi f T/2)^2}\right] \tag{2.67}$$

For $p = 1/2$, the line spectrum disappears and

$$\frac{S_m(f)}{E} = \frac{\sin^4(\pi f T/2)}{(\pi f T/2)^2} \tag{2.68}$$

2.2.4 Delay Modulation or Miller Coding

As indicated by Hecht and Guida [3], the Miller coding scheme can be modeled as a Markov source with four states whose stationary probabilities are all equal to 1/4 and whose transition matrix is given by

$$\mathbf{P} = \begin{pmatrix} 0 & 1/2 & 0 & 1/2 \\ 0 & 0 & 1/2 & 1/2 \\ 1/2 & 1/2 & 0 & 0 \\ 1/2 & 0 & 1/2 & 0 \end{pmatrix} \tag{2.69}$$

Another property of the Miller code is that it satsifies the recursion relation

$$\mathbf{P}^{4+i}\mathbf{\Gamma} = -\frac{1}{4}\mathbf{P}^{i}\mathbf{\Gamma} \quad i \ge 0 \tag{2.70}$$

where $\mathbf{\Gamma}$ is the signal correlation matrix whose ik^{th} element is defined by

$$\gamma_{ik} \triangleq \frac{1}{\sqrt{E_i E_j}} \int_0^T s_i(t) s_k(t) dt \qquad i, k = 1, 2, 3, 4 \tag{2.71}$$

For the Miller code, the four elementary signals are defined by

$$s_1(t) = -s_4(t) = A; \quad 0 \le t \le T$$
$$s_2(t) = -s_3(t) = \begin{cases} A; & 0 \le t \le T/2 \\ -A; & T/2 \le t \le T \end{cases} \tag{2.72}$$

and $E_i = A^2 T;\ i = 1, 2, 3, 4$. Substituting (2.72) into (2.71) and arranging the results in

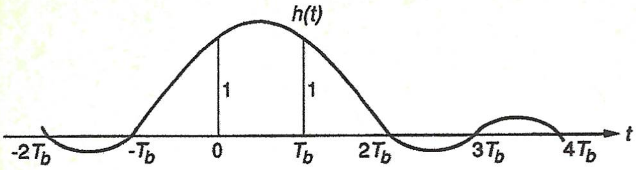

Figure 9.8 Duobinary pulse shape

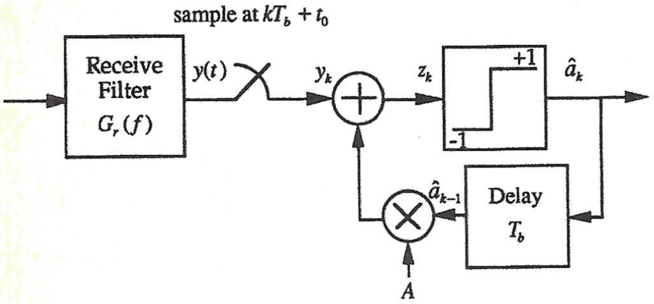

Figure 9.9 Duobinary receiver with perfect decision feedback

If we could always make *perfect* decisions, we could, in principle, subtract out the ISI and use a binary decision scheme with a single threshold. The appropriate receiver for doing this would be as shown in Fig. 9.9. If no error is made on the decision for a_{k-1}, that is, $\hat{a}_{k-1} = a_{k-1}$, then $z_k = y_k - A\hat{a}_{k-1} = Aa_k + \tilde{n}_k$ and the probability of error for a_k would be $\frac{1}{2}\,\mathrm{erfc}\left(A/\sqrt{2}\sigma_y\right)$ which is the identical result as that for binary Nyquist signaling. On the other hand, if an error was made on the decision for a_{k-1}, that is, $\hat{a}_{k-1} = -a_{k-1}$, then $z_k = y_k - A\hat{a}_{k-1} = A(a_k + 2a_{k-1}) + \tilde{n}_k$. Thus, depending on the relative polarities of a_k and a_{k-1}, the double ISI can help or hurt. The problem with this detection scheme then is that once an error occurs, it tends to *propagate* further errors, that is, if a_{k-1} is in error, then its effect is incorrectly compensated for when deciding a_k; consequently, a_k is likely to be in error also.

To get around the problem of error propagation, Lender [6] suggested a means of avoiding the necessity for ISI compensation (subtracting out the single ISI) at the receiver. His scheme suggested that we *precode* the data $\{a_i\}$ at the transmitter into the sequence $\{d_i\}$ as follows. If we think of the binary data as 0's and 1's, then the precoder is described by $d_n = a_n \oplus d_{n-1}$ where the symbol $\oplus$ denotes modulo-2 addition. Note that a_n now indicates the *change* in the output symbol d_n. If $a_n = 0$, then there is no change in d_n. If $a_n = 1$, then there is a change in d_n. This simple encoding scheme is identical to differential encoding described in Chapter 7 in connection with differential detection of BPSK modulation. The precoder can also be reinterpreted in terms of "1"'s and "-1"'s for the input bits $\{a_n\}$. In particular, the modulo-2 summer is replaced by a multiplier resulting in the relation $d_n = a_n \times d_{n-1}$. Note that d_n is still a binary random sequence with equiprobable 1's and -1's.

Using the ± 1 representation for the a_n's, we obtain the implementation for the duobinary system illustrated in Fig. 9.10. The mean of y_k, the sample of $y(t)$ at $t = kT_b + t_0$, is now equal to $A[d_k + d_{k-1}]$. Table 9.2 gives the values for this mean in relation to the possible values of the input (information) binary symbol a_k.

Thus, we see that when the information bit is a "-1", then the detected signal sample is "0" (independent of a_{k-1}). When the information bit is a "$+1$", then the detected signal