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2.2.3 Biphase (Manchester) Baseband Signaling

Here the two elementary signals are defined by

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

$$s_2(t) = -s_1(t)$$

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

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

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

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

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

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

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

$$\mathbf{P}^{4+i} \mathbf{\Gamma} = -\frac{1}{4} \mathbf{P}^i \mathbf{\Gamma} \quad i \geq 0 \quad (2.70)$$

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

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

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

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

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

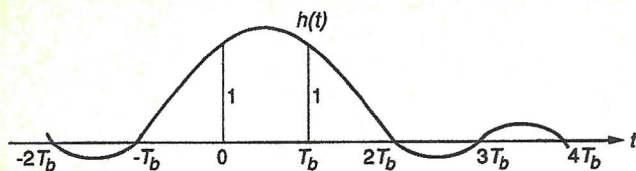


Figure 9.8 Duobinary pulse shape

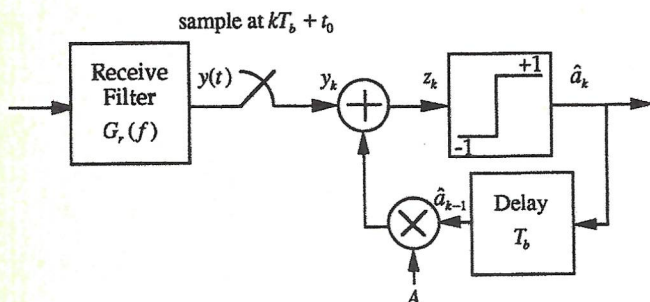


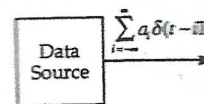
Figure 9.9 Duobinary receiver with perfect decision feedback

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 $\frac{1}{2} \operatorname{erfc}(A/\sqrt{2}\sigma_y)$ 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



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