

Communication Systems Engineering, 2nd edition, offers comprehensive coverage of the basic principles in the analysis and design of modern communication systems and review of important mathematical foundation topics. It is primarily intended for use as a text for a first course in communications, or as a comprehensive reference for practicing engineers.

This new edition of *Communication Systems Engineering* exposes the reader to relevant topics from digital communication system principles including source coding, channel coding, baseband and carrier modulation, channel distortion, channel equalization, synchronization, and wireless communications.

New content changes for the second edition include:

- Coverage of the JPEG standard for image compression
- Coverage of OFDM and multicarrier modulation
- Coverage of Turbo codes, product codes, the BCJR algorithm, and iterative codes

A chapter on wireless communications (includes new material on GSM and the IS-95 CDMA standard, as well as multiple antenna systems and space-time codes)

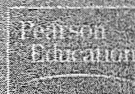
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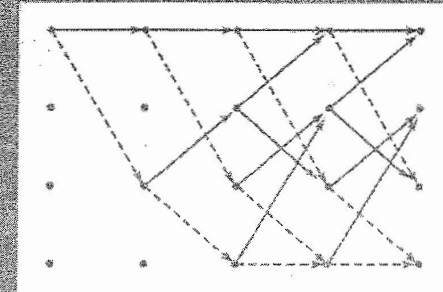
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John G. Proakis
Masoud Salehi

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To Felia, George, and Elena.
—John G. Proakis
To Fariba, Omid, Sina, and my parents.
—Masoud Salehi

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7.5 OPTIMUM RECEIVER FOR DIGITALLY MODULATED SIGNALS IN ADDITIVE WHITE GAUSSIAN NOISE

Let us consider a digital communication system that transmits digital information by use of any one of the M -ary signal waveforms described in the preceding sections. Thus, the input sequence to the modulator is subdivided into k -bit blocks or symbols and each of the $M = 2^k$ symbols is associated with a corresponding baseband signal waveform from the set $\{s_m(t), m = 1, 2, \dots, M\}$. Each signal waveform is transmitted within the symbol (signaling) interval or time slot of duration T . To be specific, we consider the transmission of information over the interval $0 \leq t \leq T$.

The channel is assumed to corrupt the signal by the addition of white Gaussian noise as shown in Figure 7.30. Thus, the received signal in the interval $0 \leq t \leq T$ may be expressed as

$$r(t) = s_m(t) + n(t), \quad 0 \leq t \leq T \quad (7.5.1)$$

where $n(t)$ denotes the sample function of the additive white Gaussian noise (AWGN) process with power-spectral density $S_n(f) = \frac{N_0}{2}$ W/Hz. Based on the observation of $r(t)$ over the signal interval, we wish to design a receiver that is optimum in the sense that it minimizes the probability of making an error.

It is convenient to subdivide the receiver into two parts, the signal demodulator and the detector. The function of the signal demodulator is to convert the received waveform $r(t)$ into an N -dimensional vector $\mathbf{r} = (r_1, r_2, \dots, r_N)$, where N is the dimension of the transmitted signal waveforms. The function of the detector is to decide which of the M possible signal waveforms was transmitted based on observation of the vector $\mathbf{r}$.

Two realizations of the signal demodulator are described in Sections 7.5.1 and 7.5.2. One is based on the use of signal correlators. The second is based on the use of matched filters. The optimum detector that follows the signal demodulator is designed to minimize the probability of error.

7.5.1 Correlation-Type Demodulator

In this section, we describe a correlation-type demodulator that decomposes the received signal and the noise into N -dimensional vectors. In other words, the signal and the noise

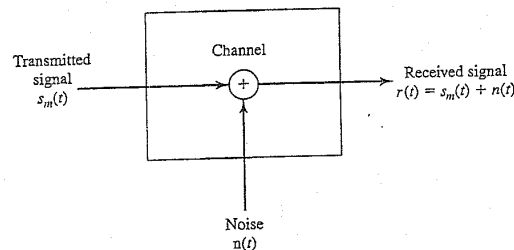


Figure 7.30 Model for received signal passed through an AWGN channel.

are expanded into a series of linearly weighted orthonormal basis functions $\{\psi_n(t)\}$. It is assumed that the N basis functions $\{\psi_n(t)\}$ span the signal space, so that every one of the possible transmitted signals of the set $\{s_m(t), 1 \leq m \leq M\}$ can be represented as a weighted linear combination of $\{\psi_n(t)\}$. In the case of the noise, the functions $\{\psi_n(t)\}$ do not span the noise space. However, we show below that the noise terms that fall outside the signal space are irrelevant to the detection of the signal.

Suppose the received signal $r(t)$ is passed through a parallel bank of N cross correlators which basically compute the projection of $r(t)$ onto the N basis functions $\{\psi_n(t)\}$, as illustrated in Figure 7.31. Thus, we have

$$\int_0^T r(t) \psi_k(t) dt = \int_0^T [s_m(t) + n(t)] \psi_k(t) dt$$

$$r_k = s_{mk} + n_k, \quad k = 1, 2, \dots, N \quad (7.5.2)$$

where

$$s_{mk} = \int_0^T s_m(t) \psi_k(t) dt, \quad k = 1, 2, \dots, N$$

$$n_k = \int_0^T n(t) \psi_k(t) dt, \quad k = 1, 2, \dots, N \quad (7.5.3)$$

Equation (7.5.2) is equivalent to

$$\mathbf{r} = \mathbf{s}_m + \mathbf{n} \quad (7.5.4)$$

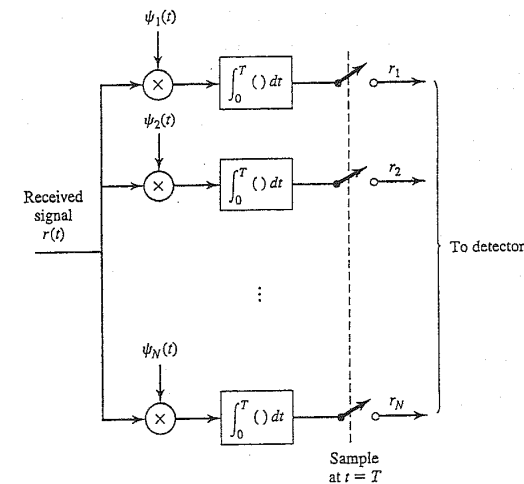
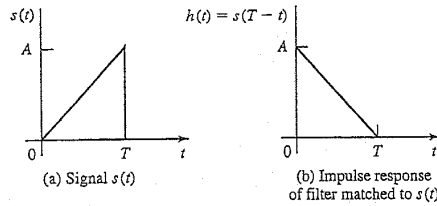
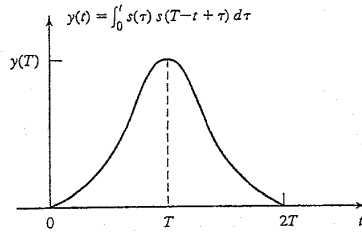


Figure 7.31 Correlation-type demodulator.

Figure 7.33 Signal $s(t)$ and filter matched to $s(t)$.Figure 7.34 Matched filter output is the autocorrelation function of $s(t)$.

In the case of the demodulator described above, the N matched filters are matched to the basis functions $\{\psi_k(t)\}$. Figure 7.35 illustrates the matched filter-type demodulator that generates the observed variables $\{r_k\}$.

Properties of the Matched Filter. A matched filter has some interesting properties. Let us prove the most important property, which may be stated as follows: If a signal $s(t)$ is corrupted by AWGN, the filter with impulse response matched to $s(t)$ maximizes the output SNR.

To prove this property, let us assume that the received signal $r(t)$ consists of the signal $s(t)$ and AWGN $n(t)$ which has zero-mean and power-spectral density $S_n(f) = N_0/2$ W/Hz. Suppose the signal $r(t)$ is passed through a filter with impulse response $h(t)$, $0 \leq t \leq T$, and its output is sampled at time $t = T$. The filter response to the signal and noise components is

$$\begin{aligned} y(t) &= \int_0^t r(\tau) h(t-\tau) d\tau \\ &= \int_0^t s(\tau) h(t-\tau) d\tau + \int_0^t n(\tau) h(t-\tau) d\tau \end{aligned} \quad (7.5.19)$$

At the sampling instant $t = T$, the signal and noise components are

$$\begin{aligned} y(T) &= \int_0^T s(\tau) h(T-\tau) d\tau + \int_0^T n(\tau) h(T-\tau) d\tau \\ &= y_s(T) + y_n(T) \end{aligned} \quad (7.5.20)$$

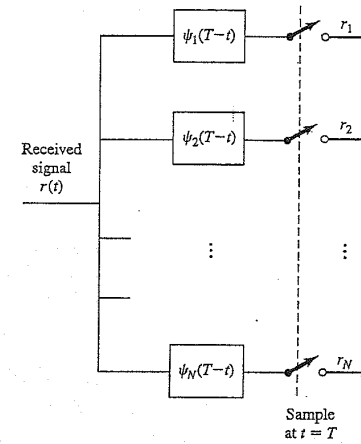


Figure 7.35 Matched filter-type demodulator.

where $y_s(T)$ represents the signal component and $y_n(T)$ represents the noise component. The problem is to select the filter impulse response that maximizes the output SNR defined as

$$\left(\frac{S}{N}\right)_0 = \frac{y_s^2(T)}{E[y_n^2(T)]} \quad (7.5.21)$$

The denominator in Equation (7.5.21) is simply the variance of the noise term at the output of the filter. Let us evaluate $E[y_n^2(T)]$. We have

$$\begin{aligned} E[y_n^2(T)] &= \int_0^T \int_0^T E[n(\tau)n(t)] h(T-\tau) h(T-t) d\tau dt \\ &= \frac{N_0}{2} \int_0^T \int_0^T \delta(t-\tau) h(T-\tau) h(T-t) d\tau dt \\ &= \frac{N_0}{2} \int_0^T h^2(T-t) dt \end{aligned} \quad (7.5.22)$$

Note that the variance depends on the power-spectral density of the noise and the energy in the impulse response $h(t)$.

By substituting for $y_s(T)$ and $E[y_n^2(T)]$ into Equation (7.5.21), we obtain the expression for the output SNR as

$$\left(\frac{S}{N}\right)_0 = \frac{[\int_0^T s(\tau) h(T-\tau) d\tau]^2}{\frac{N_0}{2} \int_0^T h^2(T-t) dt} = \frac{[\int_0^T h(\tau) s(T-\tau) d\tau]^2}{\frac{N_0}{2} \int_0^T h^2(T-t) dt} \quad (7.5.23)$$

Since the denominator of the SNR depends on the energy in $h(t)$, the maximum output SNR over $h(t)$ is obtained by maximizing the numerator of $(S/N)_0$ subject to the constraint that the denominator is held constant. The maximization of the numerator is most easily performed by use of the Cauchy-Schwarz (see Problem 2.9) inequality, which states, in general, that if $g_1(t)$ and $g_2(t)$ are finite energy signals, then

$$\left[\int_{-\infty}^{\infty} g_1(t)g_2(t) dt \right]^2 \leq \int_{-\infty}^{\infty} g_1^2(t) dt \int_{-\infty}^{\infty} g_2^2(t) dt \quad (7.5.24)$$

where equality holds when $g_1(t) = Cg_2(t)$ for any arbitrary constant C . If we set $g_1(t) = h(t)$ and $g_2(t) = s(T-t)$, it is clear that the $(S/N)_0$ is maximized when $h(t) = Cs(T-t)$; i.e., $h(t)$ is matched to the signal $s(t)$. The scale factor C^2 drops out of the expression for $(S/N)_0$ since it appears in both the numerator and the denominator.

The output (maximum) SNR obtained with the matched filter is

$$\begin{aligned} \left(\frac{S}{N} \right)_o &= \frac{2}{N_0} \int_0^T s^2(t) dt \\ &= \frac{2\mathcal{E}_s}{N_0} \end{aligned} \quad (7.5.25)$$

Note that the output SNR from the matched filter depends on the energy of the waveform $s(t)$ but not on the detailed characteristics of $s(t)$. This is another interesting property of the matched filter.

Frequency Domain Interpretation of the Matched Filter. The matched filter has an interesting frequency domain interpretation. Since $h(t) = s(T-t)$, the Fourier transform of this relationship is

$$\begin{aligned} H(f) &= \int_0^T s(T-t)e^{-j2\pi ft} dt \\ &= \left[\int_0^T s(\tau)e^{j2\pi f\tau} d\tau \right] e^{-j2\pi fT} \\ &= S^*(f)e^{-j2\pi fT} \end{aligned} \quad (7.5.26)$$

We observe that the matched filter has a frequency response which is the complex conjugate of the transmitted signal spectrum multiplied by the phase factor $e^{-j2\pi fT}$, which represents the sampling delay of T . In other words, $|H(f)| = |S(f)|$, so that the magnitude response of the matched filter is identical to the transmitted signal spectrum. On the other hand, the phase of $H(f)$ is the negative of the phase of $S(f)$.

Now, if the signal $s(t)$, with spectrum $S(f)$, is passed through the matched filter, the filter output has a spectrum $Y(f) = |S(f)|^2 e^{-j2\pi fT}$. Hence, the output waveform is

$$\begin{aligned} y_s(t) &= \int_{-\infty}^{\infty} Y(f)e^{j2\pi ft} df \\ &= \int_{-\infty}^{\infty} |S(f)|^2 e^{-j2\pi fT} e^{j2\pi ft} df \end{aligned} \quad (7.5.27)$$

By sampling the output of the matched filter at $t = T$, we obtain

$$y_s(T) = \int_{-\infty}^{\infty} |S(f)|^2 df = \int_0^T s^2(t) dt = \mathcal{E}_s \quad (7.5.28)$$

where the last step follows from Parseval's relation.

The noise of the output of the matched filter has a power-spectral density

$$S_o(f) = |H(f)|^2 N_0/2 \quad (7.5.29)$$

Hence, the total noise power at the output of the matched filter is

$$\begin{aligned} P_n &= \int_{-\infty}^{\infty} S_o(f) df \\ &= \int_{-\infty}^{\infty} \frac{N_0}{2} |H(f)|^2 df = \frac{N_0}{2} \int_{-\infty}^{\infty} |S(f)|^2 df = \frac{\mathcal{E}_s N_0}{2} \end{aligned} \quad (7.5.30)$$

The output SNR is simply the ratio of the signal power P_s , given by

$$P_s = y_s^2(T) \quad (7.5.31)$$

to the noise power P_n . Hence,

$$\left(\frac{S}{N} \right)_o = \frac{P_s}{P_n} = \frac{\mathcal{E}_s^2}{\mathcal{E}_s N_0/2} = \frac{2\mathcal{E}_s}{N_0} \quad (7.5.32)$$

which agrees with the result given by Equation (7.5.25).

Example 7.5.2

Consider the $M = 4$ biorthogonal signals constructed from the binary PPM signals and their negatives in Figure 7.12(b), for transmitting information over an AWGN channel. The noise is assumed to have zero mean and power-spectral density $N_0/2$. Determine the basis functions for this signal set, the impulse response of the matched-filter demodulators, and the output waveforms of the matched filter demodulators when the transmitted signal is $s_1(t)$.

Solution The $M = 4$ biorthogonal signals have dimension $N = 2$. Hence, two basis functions are needed to represent the signals. From Figure 7.12(b), we choose $\psi_1(t)$ and $\psi_2(t)$ as

$$\begin{aligned} \psi_1(t) &= \begin{cases} \sqrt{\frac{2}{T}}, & 0 \leq t \leq \frac{T}{2} \\ 0, & \text{otherwise} \end{cases} \\ \psi_2(t) &= \begin{cases} \sqrt{\frac{2}{T}}, & \frac{T}{2} \leq t \leq T \\ 0, & \text{otherwise} \end{cases} \end{aligned} \quad (7.5.33)$$

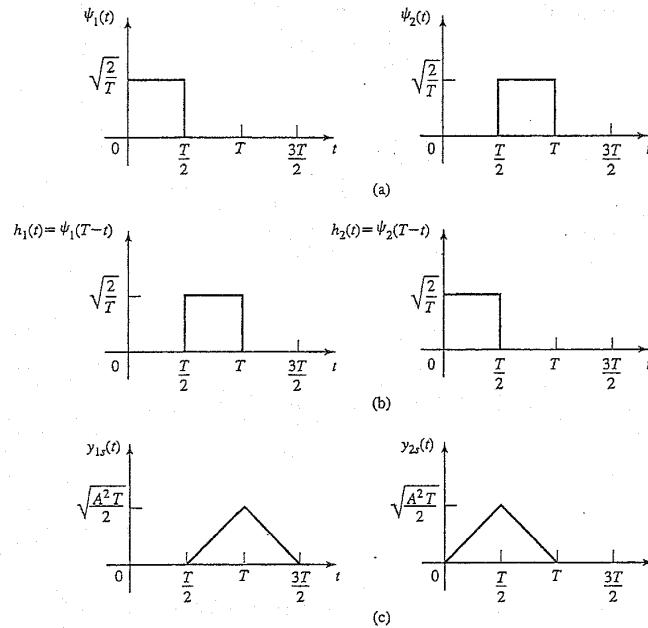


Figure 7.36 Basis functions and matched filter responses for Example 7.5.2.

These waveforms are illustrated in Figure 7.36(a). The impulse responses of the two matched filters are

$$h_1(t) = \psi_1(T-t) = \begin{cases} \sqrt{\frac{2}{T}}, & \frac{T}{2} \leq t \leq T \\ 0, & \text{otherwise} \end{cases}$$

$$h_2(t) = \psi_2(T-t) = \begin{cases} \sqrt{\frac{2}{T}}, & 0 \leq t \leq T/2 \\ 0, & \text{otherwise} \end{cases} \quad (7.5.34)$$

and are illustrated in Figure 7.36(b).

If $s_1(t)$ is transmitted, the (noise-free) responses of the two matched filters are shown in Figure 7.36(c). Since $y_1(t)$ and $y_2(t)$ are sampled at $t = T$, we observe that $y_{1s}(T) = \sqrt{\frac{A^2T}{2}}$ and $y_{2s}(T) = 0$. Note that $A^2T/2 = \mathcal{E}_s$, the signal energy. Hence, the received vector formed from the two matched filter outputs at the sampling instant

$t = T$ is

$$\mathbf{r} = (r_1, r_2) = (\sqrt{\mathcal{E}_s} + n_1, n_2) \quad (7.5.35)$$

where $n_1 = y_{1n}(T)$ and $n_2 = y_{2n}(T)$ are the noise components at the outputs of the matched filters, given by

$$y_{kn}(T) = \int_0^T n(t) \psi_k(t) dt, \quad k = 1, 2 \quad (7.5.36)$$

Clearly, $E[n_k] = E[y_{kn}(T)] = 0$. Their variance is

$$\begin{aligned} \sigma_n^2 &= E[y_{kn}^2(T)] = \int_0^T \int_0^T E[n(t)n(\tau)] \psi_k(t) \psi_k(\tau) dt d\tau \\ &= \frac{N_0}{2} \int_0^T \int_0^T \delta(t-\tau) \psi_k(t) \psi_k(\tau) dt d\tau \\ &= \frac{N_0}{2} \int_0^T \psi_k^2(t) dt = \frac{N_0}{2} \end{aligned} \quad (7.5.37)$$

Observe that the $(S/N)_o$ for the first matched filter is

$$\left(\frac{S}{N}\right)_o = \frac{(\sqrt{\mathcal{E}_s})^2}{N_0/2} = \frac{2\mathcal{E}_s}{N_0}$$

which agrees with our previous result. Also note that the four possible outputs of the two matched filters, corresponding to the four possible transmitted signals are $(r_1, r_2) = (\sqrt{\mathcal{E}_s} + n_1, n_2)$, $(n_1, \sqrt{\mathcal{E}_s} + n_2)$, $(-\sqrt{\mathcal{E}_s} + n_1, n_2)$, and $(n_1, -\sqrt{\mathcal{E}_s} + n_2)$.

7.5.3 The Optimum Detector

In Sections 7.5.1 and 7.5.2, we demonstrated that, for a signal transmitted over an AWGN channel, either a correlation-type demodulator or a matched-filter-type demodulator produces the vector $\mathbf{r} = (r_1, r_2, \dots, r_N)$ which contains all the relevant information in the received signal waveform. In this section, we describe the optimum decision rule based on the observation vector $\mathbf{r}$. For this development, we assume that there is no memory in signals transmitted in successive signal intervals.

As it is seen from Equation (7.5.4), the received vector $\mathbf{r}$ is the sum of two vectors, $\mathbf{s}_m$, the vector equivalent of the transmitted signal, and $\mathbf{n}$, the noise vector. The vector $\mathbf{s}_m$ is a point in the signal constellation and the vector $\mathbf{n}$ is an N -dimensional random vector with i.i.d. components each being a Gaussian random variable with mean 0 and variance $N_0/2$. Since components of the noise are independent and have the same mean and variance, the distribution of the noise vector $\mathbf{n}$ in the N -dimensional space has spherical symmetry. When $\mathbf{s}_m$ is transmitted, the received vector $\mathbf{r}$, which represents the transmitted signal $\mathbf{s}_m$ plus the spherically symmetric noise $\mathbf{n}$, can be represented by a spherical cloud centered at $\mathbf{s}_m$. The density of this cloud is higher at the center and becomes less as we depart from $\mathbf{s}_m$; i.e., these points become less likely to be received. The variance of the noise $N_0/2$ determines the density of the noise