APPENDIX **A**

THE LEVINSON–DURBIN ALGORITHM

The Levinson–Durbin algorithm is an order-recursive method for determining the solution to the set of linear equations

$$\mathbf{\Phi}_p \mathbf{a}_p = \mathbf{\phi}_p \tag{A-1}$$

where $\mathbf{\Phi}_p$ is a $p \times p$ Toeplitz matrix, $\mathbf{a}_p$ is the vector of predictor coefficients expressed as

$$\mathbf{a}_p' = [a_{p1} \quad a_{p2} \quad \ldots \quad a_{pp}]$$

and $\mathbf{\phi}_p$ is a p-dimensional vector with elements

$$\mathbf{\phi}_p' = [\phi(1) \quad \phi(2) \quad \ldots \quad \phi(p)]$$

For a first-order ($p = 1$) predictor, we have the solution

$$\begin{aligned} \phi(0)a_{11} &= \phi(1) \\ a_{11} &= \phi(1)/\phi(0) \end{aligned} \tag{A-2}$$

The residual mean square error (MSE) for the first-order predictor is

$$\begin{aligned} \mathscr{E}_1 &= \phi(0) - a_{11}\phi(1) \\ &= \phi(0) - a_{11}^2\phi(0) \\ &= \phi(0)(1 - a_{11}^2) \end{aligned} \tag{A-3}$$

In general, we may express the solution for the coefficients of an mth-order predictor in terms of the coefficients of the $(m-1)$th-order predictor. Thus, we express $\mathbf{a}_m$ as the sum of two vectors, namely,

$$\mathbf{a}_m = \begin{bmatrix} a_{m1} \\ a_{m2} \\ \vdots \\ a_{mm} \end{bmatrix} = \begin{bmatrix} \mathbf{a}_{m-1} \\ \hdashline 0 \end{bmatrix} + \begin{bmatrix} \mathbf{d}_{m-1} \\ \hdashline k_m \end{bmatrix} \tag{A-4}$$

where the vector $\mathbf{d}_{m-1}$ and the scalar k_m are to be determined. Also, $\mathbf{\Phi}_m$ may be expressed as

$$\mathbf{\Phi}_m = \begin{bmatrix} \mathbf{\Phi}_{m-1} & \boldsymbol{\phi}^r_{m-1} \\ \boldsymbol{\phi}^{r\prime}_{m-1} & \phi(0) \end{bmatrix} \tag{A-5}$$

where $\boldsymbol{\phi}^r_{m-1}$ is just the vector $\boldsymbol{\phi}_{m-1}$ in reverse order.

Now

$$\begin{bmatrix} \mathbf{\Phi}_{m-1} & \boldsymbol{\phi}^r_{m-1} \\ \boldsymbol{\phi}^{r\prime}_{m-1} & \phi(0) \end{bmatrix} \left(\begin{bmatrix} \mathbf{a}_{m-1} \\ 0 \end{bmatrix} + \begin{bmatrix} \mathbf{d}_{m-1} \\ k_m \end{bmatrix} \right) = \begin{bmatrix} \boldsymbol{\phi}_{m-1} \\ \phi(m) \end{bmatrix} \tag{A-6}$$

From (A-6), we obtain two equations. The first is the matrix equation

$$\mathbf{\Phi}_{m-1}\mathbf{a}_{m-1} + \mathbf{\Phi}_{m-1}\mathbf{d}_{m-1} + k_m \boldsymbol{\phi}^r_{m-1} = \boldsymbol{\phi}_{m-1} \tag{A-7}$$

But $\mathbf{\Phi}_{m-1}\mathbf{a}_{m-1} = \boldsymbol{\phi}_{m-1}$. Hence, (A-7) simplifies to

$$\mathbf{\Phi}_{m-1}\mathbf{d}_{m-1} + k_m \boldsymbol{\phi}^r_{m-1} = \mathbf{0} \tag{A-8}$$

This equation has the solution

$$\mathbf{d}_{m-1} = -k_m \mathbf{\Phi}^{-1}_{m-1} \boldsymbol{\phi}^r_{m-1} \tag{A-9}$$

But $\boldsymbol{\phi}^r_{m-1}$ is just $\boldsymbol{\phi}_{m-1}$ in reverse order. Hence, the solution in (A-9) is simply $\mathbf{a}_{m-1}$ in reverse order multiplied by $-k_m$. That is,

$$\mathbf{d}_{m-1} = -k_m \begin{bmatrix} a_{m-1\,m-1} \\ a_{m-1\,m-2} \\ \vdots \\ a_{m-1\,1} \end{bmatrix} \tag{A-10}$$

The second equation obtained from (A-6) is the scalar equation

$$\boldsymbol{\phi}^{r\prime}_{m-1}\mathbf{a}_{m-1} + \boldsymbol{\phi}^{r\prime}_{m-1}\mathbf{d}_{m-1} + \phi(0)k_m = \phi(m) \tag{A-11}$$

We eliminate $\mathbf{d}_{m-1}$ from (A-11) by use of (A-10). The resulting equation gives us k_m. That is,

$$\begin{aligned} k_m &= \frac{\phi(m) - \boldsymbol{\phi}^{r\prime}_{m-1}\mathbf{a}_{m-1}}{\phi(0) - \boldsymbol{\phi}^{r\prime}_{m-1}\mathbf{\Phi}^{-1}_{m-1}\boldsymbol{\phi}^r_{m-1}} \\ &= \frac{\phi(m) - \boldsymbol{\phi}^{r\prime}_{m-1}\mathbf{a}_{m-1}}{\phi(0) - \mathbf{a}'_{m-1}\boldsymbol{\phi}_{m-1}} \\ &= \frac{\phi(m) - \boldsymbol{\phi}^{r\prime}_{m-1}\mathbf{a}_{m-1}}{\mathscr{E}_{m-1}} \end{aligned} \tag{A-12}$$

where $\mathscr{E}_{m-1}$ is the residual MSE given as

$$\mathscr{E}_{m-1} = \phi(0) - \mathbf{a}'_{m-1}\boldsymbol{\phi}_{m-1} \tag{A-13}$$

By substituting (A-10) for $\mathbf{d}_{m-1}$ in (A-4), we obtain the order-recursive relation

$$a_{mk} = a_{m-1\,k} - k_m a_{m-1\,m-k}, \quad k = 1, 2, \ldots, m-1, \quad m = 1, 2, \ldots, p \tag{A-14}$$

and

$$a_{mm} = k_m$$

The minimum MSE may also be computed recursively. We have

$$\mathscr{E}_m = \phi(0) - \sum_{k=1}^{m} a_{mk}\phi(k) \tag{A-15}$$

Using (A-14) in (A-15), we obtain

$$\mathscr{E}_m = \phi(0) - \sum_{k=1}^{m-1} a_{m-1\,k}\phi(k) - a_{mm}\left[\phi(m) - \sum_{k=1}^{m-1} a_{m-1\,k\,m-k}\phi(k)\right] \tag{A-16}$$

But the term in square brackets in (A-16) is just the numerator of k_m in (A-12). Hence,

$$\begin{aligned}\mathscr{E}_m &= \mathscr{E}_{m-1} - a_{mm}^2\mathscr{E}_{m-1} \\ &= \mathscr{E}_{m-1}(1 - a_{mm}^2)\end{aligned} \tag{A-17}$$

APPENDIX **B**

ERROR PROBABILITY FOR MULTICHANNEL BINARY SIGNALS

In multichannel communication systems that employ binary signaling for transmitting information over the AWGN channel, the decision variable at the detector can be expressed as a special case of the general quadratic form

$$D = \sum_{k=1}^{L} (A|X_k|^2 + B|Y_k|^2 + CX_kY_k^* + C^*X_k^*Y_k) \tag{B-1}$$

in complex-valued gaussian random variables. A, B, and C are constants; X_k and Y_k are a pair of correlated complex-valued gaussian random variables. For the channels considered, the L pairs $\{X_k, Y_k\}$ are mutually statistically independent and identically distributed.

The probability of error is the probability that $D<0$. This probability is evaluated below.

The computation begins with the characteristic function, denoted by $\psi_D(jv)$, of the general quadratic form. The probability that $D<0$, denoted here as the probability of error P_b, is

$$P_b = P(D<0) = \int_{-\infty}^{0} p(D)\,dD \tag{B-2}$$

where $p(D)$, the probability density function of D, is related to $\psi_D(jv)$ by the Fourier transform, i.e.,

$$p(D) = \frac{1}{2\pi}\int_{-\infty}^{\infty} \psi_D(jv)e^{-jvD}\,dv$$

Hence,

$$P_b = \int_{-\infty}^{0} dD \frac{1}{2\pi} \int_{-\infty}^{\infty} \psi_D(jv) e^{-jvD}\, dv \tag{B-3}$$

Let us interchange the order of integration and carry out first the integration with respect to D. The result is

$$P_b = -\frac{1}{2\pi j} \int_{-\infty+j\varepsilon}^{\infty+j\varepsilon} \frac{\psi_D(jv)}{v}\, dv \tag{B-4}$$

where a small positive number ε has been inserted in order to move the path of integration away from the singularity at $v = 0$ and which must be positive in order to allow for the interchange in the order of integration.

Since D is the sum of statistically independent random variables, the characteristic function of D factors into a product of L characteristic functions, with each function corresponding to the individual random variables d_k, where

$$d_k = A\,|X_k|^2 + B\,|Y_k|^2 + CX_kY_k^* + C^*X_k^*Y_k$$

The characteristic function of d_k is

$$\phi_{d_k}(jv) = \frac{v_1 v_2}{(v + jv_1)(v - jv_2)} \exp\left[\frac{v_1 v_2(-v^2\alpha_{1k} + jv\alpha_{2k})}{(v + jv_1)(v - jv_2)}\right] \tag{B-5}$$

where the parameters v_1, v_2, α_{1k}, and α_{2k} depend on the means $\bar{X}_k$ and $\bar{Y}_k$ and the second (central) moments μ_{xx}, μ_{yy}, and μ_{xy} of the complex-vlaued gaussian variables X_k and Y_k through the following definitions ($|C|^2 - AB > 0$):

$$\begin{aligned}
v_1 &= \sqrt{w^2 + \frac{1}{4(\mu_{xx}\mu_{yy} - |\mu_{xy}|^2)(|C|^2 - AB)}} - w \\
v_2 &= \sqrt{w^2 + \frac{1}{4(\mu_{xx}\mu_{yy} - |\mu_{xy}|^2)(|C|^2 - AB)}} + w \\
w &= \frac{A\mu_{xx} + B\mu_{yy} + C\mu_{xy}^* + C^*\mu_{xy}}{4(\mu_{xx}\mu_{yy} - |\mu_{xy}|^2)(|C|^2 - AB)} \\
\alpha_{1k} &= 2(|C|^2 - AB)(|\bar{X}_k|^2\,\mu_{yy} + |\bar{Y}_k|^2\,\mu_{xx} - \bar{X}_k^*\bar{Y}_k\mu_{xy} - \bar{X}_k\bar{Y}_k^*\mu_{xy}^*) \\
\alpha_{2k} &= A\,|\bar{X}_k|^2 + B\,|\bar{Y}_k|^2 + C\bar{X}_k^*\bar{Y}_k + C^*\bar{X}_k\bar{Y}_k^* \\
\mu_{xy} &= \tfrac{1}{2}E[(X_k - \bar{X}_k)(Y_k - \bar{Y}_k)^*]
\end{aligned} \tag{B-6}$$

Now, as a result of the independence of the random variables d_k, the characteristic function of D is

$$\psi_D(jv) = \prod_{k=1}^{L} \psi_{d_k}(jv)$$

$$\psi_D(jv) = \frac{(v_1 v_2)^L}{(v + jv_1)^L(v - jv_2)^L} \exp\left[\frac{v_1 v_2(jv\alpha_2 - v^2\alpha_1)}{(v + jv_1)(v - jv_2)}\right] \tag{B-7}$$

where

$$\alpha_1 = \sum_{k=1}^{L} \alpha_{1k}, \qquad \alpha_2 = \sum_{k=1}^{L} \alpha_{2k} \tag{B-8}$$

The result (B-7) is substituted for $\psi_D(jv)$ in (B-4), and we obtain

$$P_b = -\frac{(v_1 v_2)^L}{2\pi j} \int_{-\infty+j\epsilon}^{\infty+j\epsilon} \frac{dv}{v(v+jv_1)^L(v-jv_2)^L} \exp\left[\frac{v_1 v_2(jv\alpha_2 - v^2\alpha_1)}{(v+jv_1)(v-jv_2)}\right] \tag{B-9}$$

This integral is evaluated as follows.

The first step is to express the exponential function in the form

$$\exp\left(-A_1 + \frac{jA_2}{v+jv_1} - \frac{jA_3}{v-jv_2}\right)$$

where one can easily verify that the constants A_1, A_2, and A_3 are given as

$$\begin{aligned} A_1 &= \alpha_1 v_1 v_2 \\ A_2 &= \frac{v_1^2 v_2}{v_1+v_2}(\alpha_1 v_1 + \alpha_2) \\ A_3 &= \frac{v_1 v_2^2}{v_1+v_2}(\alpha_1 v_2 - \alpha_2) \end{aligned} \tag{B-10}$$

Second, a conformal transformation is made from the v plane onto the p plane via the change in variable

$$p = -\frac{v_1}{v_2}\frac{v-jv_2}{v+jv_1} \tag{B-11}$$

In the p plane, the integral given by (B-9) becomes

$$P_b = \frac{\exp\,[v_1 v_2(-2\alpha v_1 v_2 + \alpha_2 v_1 - \alpha_2 v_2)/(v_1+v_2)^2]}{(1+v_2/v_1)^{2L-1}} \frac{1}{2\pi j}\int_\Gamma f(p)\,dp \tag{B-12}$$

where

$$f(p) = \frac{[1+(v_2/v_1)p]^{2L-1}}{p^L(1-p)} \exp\left[\frac{A_2(v_2/v_1)}{v_1+v_2}p + \frac{A_3(v_1/v_2)}{v_1+v_2}\frac{1}{p}\right] \tag{B-13}$$

and Γ is a circular contour of radius less than unity that encloses the origin.

The third step is to evaluate the integral

$$\begin{aligned} \frac{1}{2\pi j}\int_\Gamma f(p)\,dp = \frac{1}{2\pi j}\int_\Gamma &\frac{(1+(v_2/v_1)p)^{2L-1}}{p^L(1-p)} \\ &\times \exp\left[\frac{A_2(v_2/v_1)}{v_1+v_2}p + \frac{A_3(v_1/v_2)}{v_1+v_2}\frac{1}{p}\right]dp \end{aligned} \tag{B-14}$$

In order to facilitate subsequent manipulations, the constants $a \geq 0$ and $b \geq 0$ are introduced and defined as follows:

$$\tfrac{1}{2}a^2 = \frac{A_3(v_1/v_2)}{v_1+v_2}, \qquad \tfrac{1}{2}b^2 = \frac{A_2(v_2/v_1)}{v_1+v_2} \tag{B-15}$$

Let us also expand the function $[1+(v_2/v_1)p]^{2L-1}$ as a binomial series. As a result, we obtain

$$\frac{1}{2\pi j}\int_\Gamma f(p)\,dp = \sum_{k=0}^{2L-1}\binom{2L-1}{k}\left(\frac{v_2}{v_1}\right)^k \times \frac{1}{2\pi j}\int_\Gamma \frac{p^k}{p^L(1-p)}\exp\left(\frac{\frac{1}{2}a^2}{p}+\tfrac{1}{2}b^2p\right)dp \tag{B-16}$$

The contour integral given in (B-16) is one representation of the Bessel function. It can be solved by making use of the relations

$$I_n(ab) = \begin{cases} \dfrac{1}{2\pi j}\left(\dfrac{a}{b}\right)^n \displaystyle\int_\Gamma \frac{1}{p^{n+1}}\exp\left(\frac{\frac{1}{2}a^2}{p}+\tfrac{1}{2}b^2p\right)dp \\ \dfrac{1}{2\pi j}\left(\dfrac{b}{a}\right)^n \displaystyle\int_\Gamma p^{n-1}\exp\left(\frac{\frac{1}{2}a^2}{p}+\tfrac{1}{2}b^2p\right)dp \end{cases}$$

where $I_n(x)$ is the nth order modified Bessel function of the first kind and the series representation of Marcum's Q function in terms of Bessel functions, i.e.,

$$Q_1(a,b) = \exp\left[-\tfrac{1}{2}(a^2+b^2)\right] + \sum_{n=0}^{\infty}\left(\frac{a}{b}\right)^n I_n(ab)$$

First, consider the case $0 \le k \le L-2$ in (B-16). In this case, the resulting contour integral can be written in the form†

$$\frac{1}{2\pi j}\int_\Gamma \frac{1}{p^{L-k}(1-p)}\exp\left(\frac{\frac{1}{2}a^2}{p}+\tfrac{1}{2}b^2p\right)dp = Q_1(a,b)\exp\left[\tfrac{1}{2}(a^2+b^2)\right] + \sum_{n=1}^{L-1-k}\left(\frac{b}{a}\right)^n I_n(ab) \tag{B-17}$$

Next, consider the term $k = L-1$. The resulting contour integral can be expressed in terms of the Q function as follows:

$$\frac{1}{2\pi j}\int_\Gamma \frac{1}{p(1-p)}\exp\left(\frac{\frac{1}{2}a^2}{p}+\tfrac{1}{2}b^2p\right)dp = Q_1(a,b)\exp\left[\tfrac{1}{2}(a^2+b^2)\right] \tag{B-18}$$

Finally, consider the case $L \le k \le 2L-1$. We have

$$\begin{aligned}\frac{1}{2\pi j}\int_\Gamma \frac{p^{k-L}}{1-p}\exp\left(\frac{\frac{1}{2}a^2}{p}+\tfrac{1}{2}b^2p\right)dp &= \sum_{n=0}^{\infty}\frac{1}{2\pi j}\int_\Gamma p^{k-L+n}\exp\left(\frac{\frac{1}{2}a^2}{p}+\tfrac{1}{2}b^2p\right)dp \\ &= \sum_{n=k+1-L}^{\infty}\left(\frac{a}{b}\right)^n I_n(ab) = Q_1(a,b)\exp\left[\tfrac{1}{2}(a^2+b^2)\right] - \sum_{n=0}^{k-L}\left(\frac{a}{b}\right)^n I_n(ab)\end{aligned} \tag{B-19}$$

Collecting the terms that are indicated on the right-hand side of (B-16) and using

† This contour integral is related to the generalized Marcum Q function, defined as

$$Q_m(a,b) = \int_b^\infty x(x/a)^{m-1}\exp\left[-\tfrac{1}{2}(x^2+a^2)\right]I_{m-1}(ax)\,dx, \qquad m \ge 1$$

in the following manner:

$$Q_m(a,b)\exp\left[\tfrac{1}{2}(a^2+b^2)\right] = \frac{1}{2\pi j}\int_\Gamma \frac{1}{p^m(1-p)}\exp\left(\frac{\frac{1}{2}a^2}{p}+\tfrac{1}{2}b^2p\right)dp$$

the results given in (B-17)–(B-19), the following expression for the contour integral is obtained after some algebra:

$$\frac{1}{2\pi j}\int_{\Gamma} f(p)\,dp = \left(1+\frac{v_2}{v_1}\right)^{2L-1}\left[\exp\left[\tfrac{1}{2}(a^2+b^2)\right]Q_1(a,b) - I_0(ab)\right] + I_0(ab)\sum_{k=0}^{L-1}\binom{2L-1}{k}\left(\frac{v_2}{v_1}\right)^k + \sum_{n=1}^{L-1} I_n(ab)\sum_{k=0}^{L-1-n}\binom{2L-1}{k}\left[\left(\frac{b}{a}\right)^n\left(\frac{v_2}{v_1}\right)^k - \left(\frac{a}{b}\right)^n\left(\frac{v_2}{v_1}\right)^{2L-1-k}\right] \quad \text{(B-20)}$$

Equation (B-20) in conjunction with (B-12) gives the result for the probability of error. A further simplification results when one uses the following identity, which can easily be proved:

$$\exp\left[\frac{v_1 v_2}{(v_1+v_2)^2}(-2\alpha_1 v_1 v_2 + \alpha_2 v_1 - \alpha_2 v_2)\right] = \exp\left[-\tfrac{1}{2}(a^2+b^2)\right]$$

Therefore, it follows that

$$P_b = Q_1(a,b) - I_0(ab)\exp\left[-\tfrac{1}{2}(a^2+b^2)\right] + \frac{I_0(ab)\exp\left[-\tfrac{1}{2}(a^2+b^2)\right]}{(1+v_2/v_1)^{2L-1}}\sum_{k=0}^{L-1}\binom{2L-1}{k}\left(\frac{v_2}{v_1}\right)^k + \frac{\exp\left[-\tfrac{1}{2}(a^2+b^2)\right]}{(1+v_2/v_1)^{2L-1}} \times \sum_{n=1}^{L-1} I_n(ab)\sum_{k=0}^{L-1-n}\binom{2L-1}{k} \times \left[\left(\frac{b}{a}\right)^n\left(\frac{v_2}{v_1}\right)^k - \left(\frac{a}{b}\right)^n\left(\frac{v_2}{v_1}\right)^{2L-1-k}\right] \quad (L>1) \quad \text{(B-21)}$$

$$P_b = Q_1(a,b) - \frac{v_2/v_1}{1+v_2/v_1} I_0(ab)\exp\left[-\tfrac{1}{2}(a^2+b^2)\right] \quad (L=1)$$

This is the desired expression for the probability of error. It is now a simple matter to relate the parameters a and b to the moments of the pairs $\{X_k, Y_k\}$. Substituting for A_2 and A_3 from (B-10) into (B-15), we obtain

$$a = \left[\frac{2v_1^2 v_2(\alpha_1 v_2 - \alpha_2)}{(v_1+v_2)^2}\right]^{1/2} \qquad b = \left[\frac{2v_1 v_2^2(\alpha_1 v_1 + \alpha_2)}{(v_1+v_2)^2}\right]^{1/2} \quad \text{(B-22)}$$

Since v_1, v_2, α_1, and α_2 have been given in (B-6) and (B-8) directly in terms of the moments of the pairs X_k and Y_k, our task is completed.

APPENDIX **C**

ERROR PROBABILITIES FOR ADAPTIVE RECEPTION OF M-PHASE SIGNALS

In this appendix, we derive probabilities of error for two- and four-phase signaling over an L-diversity-branch time-invariant additive guassian noise channel and for M-phase signaling over an L-diversity-branch Rayleigh fading additive gaussian noise channel. Both channels corrupt the signaling waveforms transmitted through them by introducing additive white gaussian noise and an unknown or random multiplicative gain and phase shift in the transmitted signal. The receiver processing consists of cross-correlating the signal plus noise received over each diversity branch by a noisy reference signal, which is derived either from the previously received information-bearing signals or from the transmission and reception of a pilot signal, and adding the outputs from all L-diversity branches to form the decision variable.

C-1 MATHEMATICAL MODEL FOR AN M-PHASE SIGNALING COMMUNICATIONS SYSTEM

In the general case of M-phase signaling, the signaling waveforms at the transmitter are†

$$s_n(t) = \mathrm{Re}\,[s_{ln}(t)e^{j2\pi f_c t}]$$

† The complex representation of real signals is used throughout. Complex conjugation is denoted by an asterisk.

where

$$s_{ln}(t) = g(t)\exp\left[j\frac{2\pi}{M}(n-1)\right], \quad n=1,2,\ldots,M, \quad 0 \le t \le T \tag{C-1}$$

and T is the time duration of the signaling interval.

Consider the case in which one of these M waveforms is transmitted, for the duration of the signaling interval, over L channels. Assume that each of the channels corrupts the signaling waveform transmitted through it by introducing a multiplicative gain and phase shift, represented by the complex-valued number g_k, and an additive noise $z_k(t)$. Thus, when the transmitted waveform is $s_{ln}(t)$, the waveform received over the kth channel is

$$r_{lk}(t) = g_k s_{ln}(t) + z_k(t), \quad 0 \le t \le T, \quad k = 1, 2, \ldots, L \tag{C-2}$$

The noises $\{z_k(t)\}$ are assumed to be sample functions of a stationary white gaussian random process with zero mean and autocorrleation function $\phi_z(\tau) = N_0\delta(\tau)$, where N_0 is the value of the spectral density. These sample functions are assumed to be mutually statistically independent.

At the demodulator, $r_{lk}(t)$ is passed through a filter whose impulse response is matched to the waveform $g(t)$. The output of this filter, sampled at time $t = T$, is denoted as

$$X_k = 2\mathscr{E}g_k\exp\left[j\frac{2\pi}{M}(n-1)\right] + N_k \tag{C-3}$$

where $\mathscr{E}$ is the transmitted signal energy per channel and N_k is the noise sample from the kth filter. In order for the demodulator to decide which of the M phases was transmitted in the signaling interval $0 \le t \le T$, it attempts to undo the phase shift introduced by each channel. In practice, this is accomplished by multiplying the matched filter output X_k by the complex conjugate of an estimate $\hat{g}_k$ of the channel gain and phase shift. The result is a weighted and phase-shifted sampled output from the kth-channel filter, which is then added to the weighted and phase-shifted sampled outputs from the other $L-1$ channel filters.

The estimate $\hat{g}_k$ of the gain and phase shift of the kth channel is assumed to be derived either from the transmission of a pilot signal or by undoing the modulation on the information-bearing signals received in previous signaling intervals. As an example of the former, suppose that a pilot signal, denoted by $s_{pk}(t)$, $0 \le t \le T$, is transmitted over the kth channel for the purpose of measuring the channel gain and phase shift. The received waveform is

$$g_k s_{pk}(t) + z_{pk}(t), \quad 0 \le t \le T$$

where $z_{pk}(t)$ is a sample function of a stationary white gaussian random process with zero mean and autocorrelation function $\phi_p(\tau) = N_0\delta(\tau)$. This signal plus noise is passed through a filter matched to $s_{pk}(t)$. The filter output is sampled at time $t = T$ to yield the random variable $X_{pk} = 2\mathscr{E}_p g_k + N_{pk}$, where $\mathscr{E}_p$ is the energy in the pilot signal, which is assumed to be identical for all channels, and N_{pk} is the additive noise sample. An estimate of g_k is obtained by properly normalizing X_{pk}, i.e., $\hat{g}_k = g_k + N_{pk}/2\mathscr{E}_p$.

On the other hand, an estimate of g_k can be obtained from the information-bearing signal as follows. If one knew the information component contained in the matched filter output then an estimate of g_k could be obtained by properly normalizing this output. For example, the information component in the filter output given by (C-3) is $2\mathscr{E}g_k \exp[j(2\pi/M)(n-1)]$, and hence, the estimate is

$$\hat{g}_k = \frac{X_k}{2\mathscr{E}} \exp\left[-j\frac{2\pi}{M}(n-1)\right] = g_k + \frac{N_k'}{2\mathscr{E}}$$

where $N_k' = N_k \exp[-j(2\pi/M)(n-1)]$ and the pdf of N_k' is identical to the pdf of N_k. An estimate that is obtained from the information-bearing signal in this manner is called a *clairvoyant estimate*. Although a physically realizable receiver does not possess such clairvoyance, it can approximate this estimate by employing a time delay of one signaling interval and by feeding back the estimate of the transmitted phase in the previous signaling interval.

Whether the estimate of g_k is obtained from a pilot signal or from the information-bearing signal, the estimate can be improved by extending the time interval over which it is formed to include several prior signaling intervals in a way that has been described by Price (1962a, b). As a result of extending the measurement interval, the signal-to-noise ratio in the estimate of g_k is increased. In the general case where the estimation interval is the infinite past, the normalized *pilot signal estimate* is

$$\hat{g}_k = g_k + \sum_{i=1}^{\infty} c_i N_{pki} \bigg/ 2\mathscr{E}_p \sum_{i=1}^{\infty} c_i \tag{C-4}$$

where c_i is the weighting coefficient on the subestimate of g_k derived from the ith prior signal interval and N_{pki} is the sample of additive gaussian noise at the output of the filter matched to $s_{pk}(t)$ in the ith prior signaling interval. Similarly, the clairvoyant estimate that is obtained from the information-bearing signal by undoing the modulation over the infinite past is

$$\hat{g}_k = g_k + \sum_{i=1}^{\infty} c_i N_{ki} \bigg/ 2\mathscr{E} \sum_{i=1}^{\infty} c_i \tag{C-5}$$

As indicated, the demodulator forms the product between $\hat{g}_k^*$ and X_k and adds this to the products of the other $L-1$ channels. The random variable that results is

$$\begin{aligned} z &= \sum_{k=1}^{L} X_k \hat{g}_k^* = \sum_{k=1}^{L} X_k Y_k^* \\ &= z_r + j z_i \end{aligned} \tag{C-6}$$

where, by definition, $Y_k = \hat{g}_k$, $z_r = \text{Re}(z)$, and $z_i = \text{Im}(z)$. The phase of z is the decision variable. This is simply

$$\theta = \tan^{-1}\left(\frac{z_i}{z_r}\right) = \tan^{-1}\left[\text{Im}\left(\sum_{k=1}^{L} X_k Y_k^*\right) \bigg/ \text{Re}\left(\sum_{k=1}^{L} X_k Y_k^*\right)\right] \tag{C-7}$$

C-2 CHARACTERISTIC FUNCTION AND PROBABILITY DENSITY FUNCTION OF THE PHASE θ

The following derivation is based on the assumption that the transmitted signal phase is zero, i.e., $n = 1$. If desired, the pdf of θ conditional on any other transmitted signal phase can be obtained by translating $p(\theta)$ by the angle $2\pi(n-1)/M$. We also assume that the complex-valued numbers $\{g_k\}$, which characterize the L channels, are mutually statistically independent and identically distributed zero-mean gaussian random variables. This characterization is appropriate for slowly Rayleigh fading channels. As a consequence, the rrandom variables (X_k, Y_k) are correlated, complex-valued, zero-mean, gaussian, and statistically independent, but identically distributed with any other pair (X_i, Y_i).

The method that has been used in evaluating the probability density $p(\theta)$ in the general case of diversity reception is as follows. First, the characteristic function of the joint probability distribution function of z_r and z_i, where z_r and z_i are two components that make up the decision variable θ, is obtained. Second, the double Fourier transform of the characteristic function is performed and yields the density $p(z_r, z_i)$. Then the transformation

$$r = \sqrt{z_r^2 + z_i^2}, \qquad \theta = \tan^{-1}\left(\frac{z_i}{z_r}\right) \tag{C-8}$$

yields the joint pdf of the envelope r and the phase θ. Finally, integration of this joint pdf over the random variable r yields the pdf of θ.

The joint characteristic function of the random variables z_r and z_i can be expressed in the form

$$\psi(jv_1, jv_2) = \left[\frac{\dfrac{4}{m_{xx}m_{yy}(1-|\mu|^2)}}{\left(v_1 - j\dfrac{2|\mu|\cos\varepsilon}{\sqrt{m_{xx}m_{yy}}(1-|\mu|^2)}\right)^2 + \left(v_2 - j\dfrac{2|\mu|\sin\varepsilon}{\sqrt{m_{xx}m_{yy}}(1-|\mu|^2)}\right)^2 + \dfrac{4}{m_{xx}m_{yy}(1-|\mu|^2)^2}} \right]^L \tag{C-9}$$

where, by definition,

$$\begin{aligned} m_{xx} &= E(|X_k|^2) \qquad \text{identical for all } k \\ m_{yy} &= E(|Y_k|^2) \qquad \text{identical for all } k \\ m_{xy} &= E(X_k Y_k^*) \qquad \text{identical for all } k \\ \mu &= \frac{m_{xy}}{\sqrt{m_{xx}m_{yy}}} = |\mu| e^{-j\varepsilon} \end{aligned} \tag{C-10}$$

The result of Fourier-transforming the function $\psi(jv_1, jv_2)$ with respect to the variables v_1 and v_2 is

$$p(z_r, z_i) = \frac{(1-|\mu|^2)^L}{(L-1)!\pi 2^L}\left(\sqrt{z_r^2 + z_i^2}\right)^{L-1} \times \exp\left[|\mu|(z_r\cos\varepsilon + z_i\sin\varepsilon)\right]K_{L-1}\left(\sqrt{z_r^2 + z_i^2}\right) \tag{C-11}$$

where $K_n(x)$ is the modified Hankel function of order n. Then the transformation of random variables, as indicated in (C-8) yields the joint pdf of the envelope r and the phase θ in the form

$$p(r, \theta) = \frac{(1-|\mu|^2)^L}{(L-1)!\pi 2^L} r^L \exp\left[|\mu| r\cos(\theta - \varepsilon)\right]K_{L-1}(r) \tag{C-12}$$

Now, integration over the variable r yields the marginal pdf of the phase θ. We have evaluated the integral to obtain $p(\theta)$ in the form

$$p(\theta) = \frac{(-1)^{L-1}(1-|\mu|^2)^L}{2\pi(L-1)!}\left\{\frac{\partial^{L-1}}{\partial b^{L-1}}\left[\frac{1}{b-|\mu|^2\cos^2(\theta-\varepsilon)} + \frac{|\mu|\cos(\theta-\varepsilon)}{[b-|\mu|^2\cos^2(\theta-\varepsilon)]^{3/2}}\cos^{-1}\left(-\frac{|\mu|\cos(\theta-\varepsilon)}{b^{1/2}}\right)\right]\right\}\Bigg|_{b=1} \tag{C-13}$$

In this equation, the notation

$$\frac{\partial^L}{\partial b^L} f(b,\mu)\Bigg|_{b=1}$$

denotes the Lth partial derivative of the function $f(b,\mu)$ evaluated at $b=1$.

C-3 ERROR PROBABILITIES FOR SLOWLY RAYLEIGH FADING CHANNELS

In this section, the probability of a character error and the probability of a binary digit error are derived for M-phase signaling. The probabilities are evaluated via the probability density function and the probability distribution function of θ.

The Probability Distribution Function of the Phase In order to evaluate the probability of error, we need to evaluate the definite integral

$$P(\theta_1 \leq \theta \leq \theta_2) = \int_{\theta_1}^{\theta_2} p(\theta)\,d\theta$$

where θ_1 and θ_2 are limits of integration and $p(\theta)$ is given by (C-13). All subsequent calculations are made for a real cross-correlation coefficient μ. A real-valued μ implies that the signals have symmetric spectra. This is the usual situation encountered. Since a complex-valued μ causes a shift of ε in the pdf of θ, i.e., ε is simply a bias term, the results that are given for real μ can be altered in a trivial way to cover the more general case of complex-valued μ.

In the integration of $p(\theta)$, only the range $0 \leq \theta \leq \pi$ is considered, because $p(\theta)$ is an even function. Furthermore, the continuity of the integrand and its derivatives and the fact that the limits θ_1 and θ_2 are independent of b allow for the interchange of integration and differentiation. When this is done, the resulting integral can be evaluated quite readily and can be expressed as follows:

$$\int_{\theta_1}^{\theta_2} p(\theta)\,d\theta = \frac{(-1)^{L-1}(1-\mu^2)^L}{2\pi(L-1)!} \times \frac{\partial^{L-1}}{\partial b^{L-1}}\left\{\frac{1}{b-\mu^2}\left[\frac{\mu\sqrt{1-(b/\mu^2-1)x^2}}{b^{1/2}}\cot^{-1}x - \cot^{-1}\left(\frac{xb^{1/2}\mu}{\sqrt{1-(b/\mu^2-1)x^2}}\right)\right]\right\}\Bigg|_{x_1}^{x_2}\Bigg|_{b=1} \tag{C-14}$$

where, by definition,

$$x_i = \frac{-\mu\cos\theta_i}{\sqrt{b-\mu^2\cos\theta_i}}, \quad i=1,2 \tag{C-15}$$

Probability of a Symbol Error The probability of a symbol error for any M-phase signaling system is

$$P_M = 2\int_{\pi/M}^{\pi} p(\theta)\, d\theta$$

When (C-14) is evaluated at these two limits, the result is

$$P_M = \frac{(-1)^{L-1}(1-\mu^2)^L}{\pi(L-1)!} \frac{\partial^{L-1}}{\partial b^{L-1}} \left\{ \frac{1}{b-\mu^2} \left[\frac{\pi}{M}(M-1) - \frac{\mu \sin(\pi/M)}{\sqrt{b-\mu^2\cos^2(\pi/M)}} \cot^{-1}\left(\frac{-\mu\cos(\pi/M)}{\sqrt{b-\mu^2\cos^2(\pi/M)}} \right) \right] \right\} \Bigg|_{b=1} \tag{C-16}$$

Probability of a Binary Digit Error First, let us consider two-phase signaling. In this case, the probability of a binary digit error is obtained by integrating the pdf $p(\theta)$ over the range $\frac{1}{2}\pi < \theta < 3\pi$. Since $p(\theta)$ is an even function and the signals are a priori equally likely, this probability can be written as

$$P_2 = 2\int_{\pi/2}^{\pi} p(\theta)\, d\theta$$

It is easily verified that $\theta_1 = \frac{1}{2}\pi$ implies $x_1 = 0$ and $\theta_2 = \pi$ implies $x_2 = \mu/\sqrt{b-\mu^2}$. Thus,

$$P_2 = \frac{(-1)^{L-1}(1-\mu^2)^L}{2(L-1)!} \frac{\partial^{L-1}}{\partial b^{L-1}} \left[\frac{1}{b-\mu^2} - \frac{\mu}{b^{1/2}(b-\mu^2)} \right] \Bigg|_{b=1} \tag{C-17}$$

After performing the differentiation indicated in (C-17) and evaluating the resulting function at $b = 1$, the probability of a binary digit error is obtained in the form

$$P_2 = \frac{1}{2}\left[1 - \mu \sum_{k=0}^{L-1} \binom{2k}{k} \left(\frac{1-\mu^2}{4} \right)^k \right] \tag{C-18}$$

Next, we consider the case of four-phase signaling in which a Gray code is used to map pairs of bits into phases. Assuming again that the transmitted signal is $s_{l1}(t)$, it is clear that a single error is committed when the received phase is $\frac{1}{4}\pi < \theta < \frac{3}{4}\pi$, and a double error is committed when the received phase is $\frac{3}{4}\pi < \theta < \pi$. That is, the probability of a binary digit error is

$$P_{4b} = \int_{\pi/4}^{3\pi/4} p(\theta)\, d\theta + 2\int_{3\pi/4}^{\pi} p(\theta)\, d\theta \tag{C-19}$$

It is easily established from (C-14) and (C-19) that

$$P_{4b} = \frac{(-1)^{L-1}(1-\mu^2)^L}{2(L-1)!} \frac{\partial^{L-1}}{\partial b^{L-1}} \left[\frac{1}{b-\mu^2} - \frac{\mu}{(b-\mu^2)(2b-\mu^2)^{1/2}} \right] \Bigg|_{b=1}$$

Hence, the probability of a binary digit error for four-phase signaling is

$$P_{4b} = \frac{1}{2}\left[1 - \frac{\mu}{\sqrt{2-\mu^2}} \sum_{k=0}^{L-1} \binom{2k}{k} \left(\frac{1+\mu^2}{4-2\mu^2} \right)^k \right] \tag{C-20}$$

Note that if one defines the quantity $\rho = \mu/\sqrt{2-\mu^2}$, the expression for P_{4b} in terms of ρ is

$$P_{4b} = \frac{1}{2}\left[1 - \rho \sum_{k=0}^{L-1} \binom{2k}{k} \left(\frac{1-\rho^2}{4} \right)^k \right] \tag{C-21}$$

In other words, P_{4b} has the same form as P_2 given in (C-18). Furthermore, note that ρ, just like μ, can be interpreted as a cross-correlation coefficient, since the range of ρ is $0 \leq \rho \leq 1$ for $0 \leq \mu \leq 1$. This simple fact will be used in Section C-4.

The above procedure for obtaining the bit error probability for an M-phase signal with a Gray code can be used to generate results for $M = 8$, 16, etc., as shown by Proakis (1968).

Evaluation of the Cross-Correlation Coefficient The expressions for the probabilities of error given above depend on a single parameter, namely, the cross-correlation coefficient μ. The clairvoyant estimate is given by (C-5), and the matched filter output, when signal waveform $s_{11}(t)$ is transmitted, is $X_k = 2\mathscr{E}g_k + N_k$. Hence, the cross-correlation coefficient is

$$\mu = \frac{\sqrt{v}}{\sqrt{(\bar{\gamma}_c^{-1} + 1)(\bar{\gamma}_c^{-1} + v)}} \tag{C-22}$$

where, by definition,

$$\begin{aligned} v &= \left|\sum_{i=1}^{\infty} c_i\right|^2 \Big/ \sum_{i=1}^{\infty} |c_i|^2 \\ \bar{\gamma}_c &= \frac{\mathscr{E}}{N_0} E(|g_k|^2), \qquad k = 1, 2, \ldots, L \end{aligned} \tag{C-23}$$

The parameter v represents the effective number of signaling intervals over which the estimate is formed, and $\bar{\gamma}_c$ is the average SNR per channel.

In the case of differential phase signaling, the weighting coefficients are $c_1 = 1$, $c_i = 0$ for $i \neq 1$. Hence, $v = 1$ and $\mu = \bar{\gamma}_c/(1 + \bar{\gamma}_c)$.

When $v = \infty$, the estimate is perfect and

$$\lim_{v \to \infty} \mu = \sqrt{\frac{\bar{\gamma}_c}{\bar{\gamma}_c + 1}}$$

Finally, in the case of a pilot signal estimate, given by (C-4) the cross-correlation coefficient is

$$\mu = \left[\left(1 + \frac{r+1}{r\bar{\gamma}_t}\right)\left(1 + \frac{r+1}{v\bar{\gamma}_t}\right)\right]^{-1/2} \tag{C-24}$$

where, by definition,

$$\begin{aligned} \bar{\gamma}_t &= \frac{\mathscr{E}_t}{N_0} E(|g_k|^2) \\ \mathscr{E}_t &= \mathscr{E} + \mathscr{E}_p \\ r &= \mathscr{E}/\mathscr{E}_p \end{aligned}$$

The values of μ given above are summarized in Table C-1.

C-4 ERROR PROBABILITIES FOR TIME-INVARIANT AND RICEAN FADING CHANNELS

In Section C-2, the complex-valued channel gains $\{g_k\}$ were characterized as zero-mean gaussian random variables, which is appropriate for Rayleigh fading channels. In this section, the channel gains $\{g_k\}$ are assumed to be nonzero-mean gaussian random variables. Estimates of the channel gains are formed by the demodulator and are used

TABLE C-1 RAYLEIGH FADING CHANNEL

Type of estimate	Cross-correlation coefficient μ
Clairvoyant estimate	$\dfrac{\sqrt{v}}{\sqrt{(\bar{\gamma}_c^{-1}+1)(\bar{\gamma}_c^{-1}+v)}}$
Pilot signal estimate	$\dfrac{\sqrt{rv}}{(r+1)\sqrt{\left(\dfrac{1}{\bar{\gamma}_t}+\dfrac{r}{r+1}\right)\left(\dfrac{1}{\bar{\gamma}_t}+\dfrac{v}{r+1}\right)}}$
Differential phase signaling	$\dfrac{\bar{\gamma}_c}{\bar{\gamma}_c+1}$
Perfect estimate	$\sqrt{\dfrac{\bar{\gamma}_c}{\bar{\gamma}_c+1}}$

as described in Section C-1. Moreover, the decision variable θ is defined again by (C-7). However, in this case, the gaussian random variables X_k and Y_k, which denote the matched filter output and the estimate, respectively, for the kth channel, have nonzero means, which are denoted by $\bar{X}_k$ and $\bar{Y}_k$. Furthermore, the second moments are

$$
\begin{aligned}
m_{xx} &= E(|X_k - \bar{X}_k|^2) && \text{identical for all channels}\\
m_{yy} &= E(|Y_k - \bar{Y}_k|^2) && \text{identical for all channels}\\
m_{xy} &= E[(X_k - \bar{X}_k)(Y_k^* - \bar{Y}_k^*)] && \text{identical for all channels}
\end{aligned}
$$

and the normalized covariance is defined as

$$\mu = \frac{m_{xy}}{\sqrt{m_{xx}m_{yy}}}$$

Error probabilities are given below only for two- and four-phase signaling with this channel model. We are interested in the special case in which the fluctuating component of each of the channel gains $\{g_k\}$ is zero, so that the channels are time-invariant. If, in addition to this time invariance, the noises between the estimate and the matched filter output are uncorrelated then $\mu = 0$.

In the general case, the probability of error for two-phase signaling over L statistically independent channels characterized in the manner described above can be obtained from the results in Appendix B. In its most general form, the expresssion for the binary error rate is

$$
\begin{aligned}
P_2 &= Q_1(a,b) - I_0(a)\exp[-\tfrac{1}{2}(a^2+b^2)]\\
&\quad + \frac{I_0(ab)\exp[-\tfrac{1}{2}(a^2+b^2)]}{[2/(1-\mu)]^{2L-1}}\sum_{k=0}^{L-1}\binom{2L-1}{k}\left(\frac{1+\mu}{1-\mu}\right)^k\\
&\quad + \frac{\exp[-\tfrac{1}{2}(a^2+b^2)]}{[2/(1-\mu)]^{2L-1}}\\
&\quad \times \sum_{k=1}^{L-1} I_n(ab)\sum_{k=0}^{L-1-n}\binom{2L-1}{k}\left[\left(\frac{b}{a}\right)^n\left(\frac{1+\mu}{1-\mu}\right)^k - \left(\frac{a}{b}\right)^n\left(\frac{1+\mu}{1-\mu}\right)^{2L-1-k}\right] \quad (L \geqslant 2)
\end{aligned} \tag{C-25}
$$

$$P_2 = Q_1(a,b) - \tfrac{1}{2}(1+\mu)I_0(ab)\exp[-\tfrac{1}{2}(a^2+b^2)] \quad (L=1)$$

where, by definition,

$$
\begin{aligned}
a &= \left(\tfrac{1}{2}\sum_{k=1}^{L}\left|\frac{\bar{X}_k}{\sqrt{m_{xx}}}-\frac{\bar{Y}_k}{\sqrt{m_{yy}}}\right|^2\right)^{1/2} \\
b &= \left(\tfrac{1}{2}\sum_{k=1}^{L}\left|\frac{\bar{X}_k}{\sqrt{m_{xx}}}+\frac{\bar{Y}_k}{\sqrt{m_{yy}}}\right|^2\right)^{1/2} \qquad \text{(C-26)} \\
Q_1(a,b) &= \int_b^{\infty} x \exp\left[-\tfrac{1}{2}(a^2+x^2)\right] I_0(ax)\,dx
\end{aligned}
$$

$I_n(x)$ is the modified Bessel function of the first kind and of order n.

Let us evaluate the constants a and b when the channel is time-invariant, $\mu = 0$, and the channel gain and phase estimates are those given in Section C-1. Recall that when signal $s_1(t)$ is transmitted, the matched filter output is $X_k = 2\mathscr{E}g_k + N_k$. The clairvoyant estimate is given by (C-5). Hence, for this estimate, the moments are $\bar{X}_k = 2\mathscr{E}g_k$, $\bar{Y}_k = g_k$, $m_{xx} = 4\mathscr{E}N_0$, and $m_{yy} = N_0/\mathscr{E}v$, where $\mathscr{E}$ is the signal energy, N_0 is the value of the noise spectral density, and v is defined in (C-23). Substitution of these moments into (C-26) results in the following expressions for a and b:

$$
\begin{aligned}
a &= \sqrt{\tfrac{1}{2}\gamma_b}\,|\sqrt{v}-1| \\
b &= \sqrt{\tfrac{1}{2}\gamma_b}\,|\sqrt{v}+1| \qquad \text{(C-27)} \\
\gamma_b &= \frac{\mathscr{E}}{N_0}\sum_{k=1}^{L}|g_k|^2
\end{aligned}
$$

This is a result originally derived by Price (1962).

The probability of error for differential phase signaling can be obtained by setting $v = 1$ in (C-27).

Next, consider a pilot signal estimate. In this case, the estimate is given by (C-4) and the matched filter output is again $X_k = 2\mathscr{E}g_k + N_k$. When the moments are calculated and these are substituted into (C-26), the following expressions for a and b are obtained:

$$
\begin{aligned}
a &= \sqrt{\frac{\gamma_t}{2}}\left|\sqrt{\frac{v}{r+1}}-\sqrt{\frac{r}{r+1}}\right| \\
b &= \sqrt{\frac{\gamma_t}{2}}\left(\sqrt{\frac{v}{r+1}}+\sqrt{\frac{r}{r+1}}\right)
\end{aligned}
\qquad \text{(C-28)}
$$

where

$$
\begin{aligned}
\gamma_t &= \frac{\mathscr{E}_t}{N_0}\sum_{k=1}^{L}|g_k|^2 \\
\mathscr{E}_t &= \mathscr{E} + \mathscr{E}_p \\
r &= \mathscr{E}/\mathscr{E}_p
\end{aligned}
$$

Finally, we consider the probability of a binary digit error for four-phase signaling over a time-invariant channel for which the condition $\mu = 0$ obtains. One approach that can be used to derive this error probability is to determine the pdf of θ and then to integrate this over the appropriate range of values of θ. Unfortunately, this approach proves to be intractable mathematically. Instead, a simpler, albeit roundabout, method may be used that involves the Laplace transform. In short, the integral in (14-4-14) of the text that relates the error probability $P_2(\gamma_b)$ in an AWGN channel to the error

TABLE C-2 TIME-INVARIANT CHANNEL

Type of estimate	a	b
	Two-phase signaling	
Clairvoyant estimate	$\sqrt{\tfrac{1}{2}\gamma_b}\,\lvert\sqrt{v}-1\rvert$	$\sqrt{\tfrac{1}{2}\gamma_b}(\sqrt{v}+1)$
Differential phase signaling	0	$\sqrt{2\gamma_b}$
Pilot signal estimate	$\sqrt{\frac{\gamma_t}{2}}\left\lvert\sqrt{\frac{v}{r+1}}-\sqrt{\frac{r}{r+1}}\right\rvert$	$\sqrt{\frac{\gamma_t}{2}}\left(\sqrt{\frac{v}{r+1}}+\sqrt{\frac{r}{r+1}}\right)$
	Four-phase signaling	
Clairvoyant estimate	$\sqrt{\tfrac{1}{2}\gamma_b}\,\left\lvert\sqrt{v+1+\sqrt{v^2+1}}-\sqrt{v+1-\sqrt{v^2+1}}\right\rvert$	$\sqrt{\tfrac{1}{2}\gamma_b}\left(\sqrt{v+1+\sqrt{v^2+1}}+\sqrt{v+1-\sqrt{v^2+1}}\right)$
Differential phase signaling	$\sqrt{\tfrac{1}{2}\gamma_b}\left(\sqrt{2+\sqrt{2}}-\sqrt{2-\sqrt{2}}\right)$	$\sqrt{\tfrac{1}{2}\gamma_b}\left(\sqrt{2+\sqrt{2}}+\sqrt{2-\sqrt{2}}\right)$
Pilot signal estimate	$\sqrt{\frac{\gamma_t}{4(r+1)}}\left\lvert\sqrt{v+r+\sqrt{v^2+r^2}}-\sqrt{v+r-\sqrt{v^2+r^2}}\right\rvert$	$\sqrt{\frac{\gamma_t}{4(r+1)}}\left(\sqrt{v+r+\sqrt{v^2+r^2}}+\sqrt{v+r-\sqrt{v^2+r^2}}\right)$

probability P_2 in a Rayleigh fading channel is a Laplace transform. Since the bit error probabilities P_2 and P_{4b} for a Rayleigh fading channel, given by (C-18) and (C-21), respectively, have the same form but differ only in the correlation coefficient, it follows that the bit error probabilities for the time-invariant channel also have the same form. That is, (C-25) with $\mu = 0$ is also the expression for the bit error probability of a four-phase signaling system with the parameters a and b modified to reflect the difference in the correlation coefficient. The detailed derivation may be found in the paper by Proakis (1968). The expressions for a and b are given in Table C-2.

APPENDIX **D**

SQUARE-ROOT FACTORIZATION

Consider the solution of the set of linear equations

$$\mathbf{R}_N \mathbf{C}_N = \mathbf{U}_N \tag{D-1}$$

where $\mathbf{R}_N$ is an $N \times N$ positive-definite symmetric matrix, $\mathbf{C}_N$ is an N-dimensional vector of coefficients to be determined, and $\mathbf{U}_N$ is an arbitrary N-dimensional vector. The equations in (D-1) can be solved efficiently by expressing $\mathbf{R}_N$ in the factored form

$$\mathbf{R}_N = \mathbf{S}_N \mathbf{D}_N \mathbf{S}'_N \tag{D-2}$$

where $\mathbf{S}_N$ is a lower triangular matrix with elements $\{s_{ik}\}$ and $\mathbf{D}_N$ is a diagonal matrix with diagonal elements $\{d_k\}$. The diagonal elements of $\mathbf{S}_N$ are set to unity, i.e., $s_{ii} = 1$. Then we have

$$\begin{aligned} r_{ij} &= \sum_{k=1}^{j} s_{ik} d_k s_{jk}, \quad 1 \le j \le i-1, \quad i \ge 2 \\ r_{11} &= d_1 \end{aligned} \tag{D-3}$$

where $\{r_{ij}\}$ arte the elements of $\mathbf{R}_N$. Consequently, the elements $\{s_{ik}\}$ and $\{d_k\}$ are determined from (D-3) according to the equations

$$\begin{aligned} d_1 &= r_{11} \\ s_{ij} d_j &= r_{ij} - \sum_{k=1}^{j-1} s_{ik} d_k s_{jk}, \quad 1 \le j \le i-1, \quad 2 \le i \le N \\ d_i &= r_{ii} - \sum_{k=1}^{i-1} s_{ik}^2 d_k, \quad 2 \le i \le N \end{aligned} \tag{D-4}$$

Thus, (D-4) define $\mathbf{S}_N$ and $\mathbf{D}_N$ in terms of the elements of $\mathbf{R}_N$.

The solution to (D-1) is performed in two steps. With (D-2) substituted into (D-1) we have

$$\mathbf{S}_N \mathbf{D}_N \mathbf{S}_N' \mathbf{C}_N = \mathbf{U}_N$$

Let

$$\mathbf{Y}_N = \mathbf{D}_N \mathbf{S}_N' \mathbf{C}_N \tag{D-5}$$

Then

$$\mathbf{S}_N \mathbf{Y}_N = \mathbf{U}_N \tag{D-6}$$

First we solve (D-6) for $\mathbf{Y}_N$. Because of the triangular form of $\mathbf{S}_N$, we have

$$\begin{aligned} y_1 &= u_1 \\ y_i &= u_i - \sum_{j=1}^{i-1} s_{ij} y_j, \qquad 2 \le i \le N \end{aligned} \tag{D-7}$$

Having obtained $\mathbf{Y}_N$, the second step is to compute $\mathbf{C}_N$. That is,

$$\begin{aligned} \mathbf{D}_N \mathbf{S}_N' \mathbf{C}_N &= \mathbf{Y}_N \\ \mathbf{S}_N' \mathbf{C}_N &= \mathbf{D}_N^{-1} \mathbf{Y}_N \end{aligned}$$

Beginning with

$$c_N = y_N / d_N \tag{D-8}$$

the remaining coefficients of $\mathbf{C}_N$ are obtained recursively as follows:

$$c_i = \frac{y_i}{d_i} - \sum_{j=i+1}^{N} s_{ji} c_j, \qquad 1 \le i \le N-1 \tag{D-9}$$

The number of multiplications and divisions required to perform the factorization of $\mathbf{R}_N$ is proportional to N^3. The number of multiplications and divisions required to compute $\mathbf{C}_N$, once $\mathbf{S}_N$ is determined, is proportional to N^2. In contrast, when $\mathbf{R}_N$ is Toeplitz the Levinson–Durbin algorithm should be used to determine the solution of (D-1), since the number of multiplications and divisions is proportional to N^2. On the other hand, in a recursive least-squares formulation, $\mathbf{S}_N$ and $\mathbf{D}_N$ are not computed as in (D-3), but they are updated recursively. The update is accomplished with N^2 operations (multiplications and divisions). Then the solution for the vector $\mathbf{C}_N$ follows the steps (D-5)–(D-9). Consequently, the computational burden of the recursive least-squares formulation is proportional to N^2.

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