
Peer reviewed version

Link to published version (if available): 10.1109/TVT.2004.827160

Link to publication record in Explore Bristol Research

PDF-document

**University of Bristol - Explore Bristol Research**

**General rights**

This document is made available in accordance with publisher policies. Please cite only the published version using the reference above. Full terms of use are available: http://www.bristol.ac.uk/red/research-policy/pure/user-guides/ebr-terms/
Error-Bound Formulation for Multichannel Reception of $M$-DPSK and Pilot-Aided $M$-PSK Over Rayleigh-Fading Channels With Postdetection Combining

Yuk C. Chow, Associate Member, IEEE, and Joe P. McGeehan

Abstract—Upper bounds for symbol-error probability are developed for multichannel reception of $M$-ary phase-shift keying ($M$-PSK) over frequency-flat Rayleigh-fading channels. Differential coherent demodulation of differentially encoded $M$-PSK ($M$-DPSK), pilot-tone aided coherent demodulation of $M$-PSK (PTA $M$-CPSK) and pilot-symbol aided coherent demodulation of $M$-PSK (PSA $M$-CPSK) are considered in the formulation. The bounds enable the investigation of the effects of fading correlation and unequal average power level between channels using postdetection diversity reception with maximal-ratio combining. Error-performance degradation due to imperfect demodulation effects, such as Doppler spread in $M$-DPSK schemes, noisy reference signals in PTA $M$-CPSK, and PSA $M$-CPSK schemes can be taken into account in the formulation. Exact bit-error probabilities for both binary and quaternary phase-shift keying are also derived.

Index Terms—Correlated fading, differential detection, phase-shift keying (PSK), pilot-aided coherent detection, postdetection diversity combining.

I. INTRODUCTION

In mobile communications, there are several limiting factors for high data rate transmission, notably the availability of suitable radio-frequency (RF) spectrum and multipath propagation. Bandwidth-efficient modulation schemes such as $M$-ary phase-shift keying ($M$-PSK) can help to alleviate such problems, e.g., the global system for mobile communications (GSM) enhanced data rates for GSM evolution (EDGE) system is considering 8-PSK, whereas in wide-band code-division multiple-access (W-CDMA) system quaternary phase-shift keying (Q-PSK) has been selected for radio specifications [1]–[3]. However, it is known that the error performance of phase-shift keying (PSK) degrades with increased modulation levels, since high-level PSK is more sensitive to receiver noise. Furthermore, the hostile multipath-fading environment of the wireless communications further degrades receiver performance [4]–[7]. Therefore, the tradeoff between data rate and robustness of the transmission must always be examined carefully. In this respect, the choice of which demodulation techniques are employed must also be carefully chosen. It has been shown that the error performance for ideal coherent demodulation1 of $M$-PSK ($M$-CPSK) is superior to that of differential coherent demodulation of differentially encoded $M$-PSK ($M$-DPSK) [8], [9]. In wireless communication applications, practical coherent demodulation can be achieved by using pilot-tone-aided (PTA) [10]–[18] and pilot-symbol-aided (PSA) [19]–[23] transmission techniques. In these techniques, a pilot (reference) signal is transmitted along with the data (information) signal. At the receiver, the received pilot is used as a phase and/or amplitude reference for coherent demodulation. However, extra bandwidth and transmit power are required to accommodate the pilot signal together with additional signal-processing algorithms and processing delay for channel estimation in the receiver. Due to these extra requirements, $M$-DPSK can be an attractive option for some low-cost architecture applications, despite the fact that the error performance of $M$-DPSK is inferior to that of coherent $M$-PSK.

It is well known that the performance of the receiver can be improved by using diversity reception techniques. However, diversity gain is reduced by increased correlation between the fading signals and/or unequal average signal power between diversity branches [4]–[7], [24], [25]. The diversity gain reduction may affect the overall performance of the system; hence, it is important for system engineers to predict the modem performance during the design and evaluation phases.

Error-probability expressions for diversity reception of $M$-PSK with unequal average power levels between diversity branches are given in [8], [9], [26], and [27]. Rayleigh-fading channel models are assumed in [26], whereas [8], [9], and [27] provide error expressions for a wider class of fading models, such as Rician and Nakagami models.

Performance analysis of diversity-combining reception for binary-level DPSK (B-DPSK) with correlated fading between diversity branches is given in [28]–[34]. References [28]–[31] assume a predetection diversity-combining technique, whereas [32]–[34] investigated the postdetection diversity-combining performance. Alouini and Simon [35] formulated the $B$- and $Q$-DPSK in Nakagami-fading channels with postdetection

---

1ideal coherent demodulation is referred to the receiver that has perfect carrier phase and frequency references to perform demodulation.

Performance of $B$-CPSK with correlated fading between diversity branches has been analyzed in [28]–[31], [37], and [38]. Error expressions for $M$-CPSK in Rayleigh- [39], [40] and Nakagami-fading [35] channels were also developed. However, all references were assumed to be ideal coherent demodulation only. In the case of pilot-aided $M$-CPSK systems, [36] discussed the extension of the analysis to PTA and PSA systems over Rician-fading channels. However, no analysis has been performed in this paper. Performance analysis of PSA $B$-CPSK over Rayleigh-fading channels with diversity reception has been studied in [41]. The effect of noisy channel estimation was included in the analysis, but only for the case of independent fading and equal average power between diversity branches.

By using the general error-probability formula for a quadratic receiver in a binary hypothesis test, between zero-mean complex Gaussian variables developed by Barrett [42] and the upper bound of symbol-error probability (SEP) derived in [43], upper-bounds SEP for multichannel reception with postdetection combining of $M$-DPSK, PTA, and PSA $M$-CPSK over frequency-flat Rayleigh-fading are formulated in this paper. The effect of correlated fading and unequal average power level between diversity branches are taken into account in the development. In addition, performance degradation due to imperfections in the demodulation process, such as the effects of Doppler spread on the $M$-DPSK scheme and noisy reference signal on PTA $M$-CPSK and PSA $M$-CPSK schemes are considered in the formulation.

II. SYSTEM MODELING

The overall transmission system model with multichannel reception is shown in Fig. 1. The model assumes that there are $D$ diversity channels, each carrying identical information. The fading in each branch is assumed to be frequency flat (or frequency nonselective) with envelope statistics that follow a Rayleigh distribution. Correlated fading between the $D$ channels is assumed. The signal at the output of each channel is further corrupted by an additive white Gaussian noise (AWGN), which is assumed to be statistically independent on each channel. For mathematical convenience, an equivalent baseband (complex envelope) signal representation is used throughout the analysis.

A. Transmitted Signals

1) $M$-DPSK: In this system, the data (information) signals are differentially encoded before transmit. The equivalent baseband transmitted signal $u(t)$ of $M$-DPSK can be written as

$$ u(t) = \sum_{\eta = -\infty}^{\infty} A_p(t - nT_s) e^{j\theta_n} $$

where $\theta_n$ is the transmitted phase at the $n$th time interval and $A_p$ is the amplitude gain of $u(t)$. In $M$-DPSK, the information phase (phase angle of the data symbol $\theta_n$) at the $n$th time interval

$$ \Lambda_{i,n} = \frac{2\pi(i-1)}{M}, \quad i = 1, 2, \ldots, M \quad (2) $$

The index $i$ denotes one of the $M$ possible data symbols. The symbol duration is represented by $T_s$ and $p(t)$ is the impulse response of a pulse-shaping filter with unity energy, i.e., $\int_{-\infty}^{\infty} \vert p(t) \vert^2 dt = 1$. Thus, the transmit energy per symbol $E_{s,Tx} = (A_p^2/2)(\int_{-\infty}^{\infty} \vert p(t) \vert^2 dt) = A^2/2$ and the corresponding transmit power $P_{Tx} = A^2/(2T_s)$. 2) PTA $M$-CPSK: In this system, a tone signal that is known to the receiver is transmitted along with the data signal. Based on pervious analysis given in [14]–[17], this paper assumes that Manchester encoding (also known as biphase line coding) is used to create a spectral null at the direct current (dc) level of data signal spectrum, so that the pilot tone can be inserted without interference to the data signal. For PTA $M$-CPSK, the equivalent baseband transmitted signal $u(t)$ is the summation of the data signal $u_d(t) = \sum_{n=-\infty}^{\infty} A_p(t - nT_s) e^{j\theta_n}$ and pilot signals $u_p(t) = A_p$, i.e.,

$$ u(t) = u_d(t) + u_p(t) $$

In $M$-CPSK, the information phase $(\Lambda_{i,n})$ is encoded into the transmitted phase directly, i.e., $\theta_n = \Lambda_{i,n}$, where $\Lambda_{i,n}$ is given in (2). The amplitudes of the data and pilot signals are denoted as $A_d$ and $A_p$. Similar to the $M$-DPSK discussion, the total average transmit power is $P_{Tx} = P_{Tx,d} + P_{Tx,p}$, where $P_{Tx,d} = A_d^2/(2T_s)$ and $P_{Tx,p} = A_p^2/2$ are the average transmit powers for the data and pilot signals. The ratio between the pilot and data symbol powers is defined as $\eta$ (i.e., $\eta = A_p^2/A_d^2$). Hence, for a given average transmit power $P_{Tx}$, the average transmit power for the data symbol is $P_{Tx,d} = P_{Tx}/(1 + \eta)$. The corresponding transmit data-energy per symbol is $E_{s,d,Tx} = E_{s,Tx}/(1 + \eta)$.

3) PSA $M$-CPSK: In this system, pilot symbols are transmitted along with the data symbols through the use of the

---

Since the bandwidth of the Manchester encoded signal is double of the non-return-to-zero (NRZ) signal, a higher level modulation is required to compensate for the effect of bandwidth expansion. Alternatively, a dual-tone calibration technique [12] or transparent tone-in-band technique [11], [13] can be applied.
time-multiplexing technique. Fig. 2 shows the structure of the
time-multiplexed sequence in which one pilot symbol is perio-
dically time multiplexed with \((N - 1)\) data symbols, where
\(N\) refers to a slot length. In order for pilot symbols to
provide accurate channel estimation at the receiver, the pilot symbol
insertion rate should be at least two times that of the Doppler
spread of the channel [20]–[23] and [41]. This implies that
\(N\) is upper bounded by \(N \leq (2 \times \text{Doppler spread} \times T_{\text{S,FMX}})^{-1}\),
where \(T_{\text{S,FMX}}\) represents the period of time-multiplexed
symbol. The format of equivalent baseband transmitted signal
\(u(t)\) for PSA \(M\)-CPSK is same as the \(M\)-DPSK signal
expressed in (1), with \(T_{\theta}\) replaced by \(T_{\text{S,FMX}}\) and \(\theta_{n}\)
equal to \(\Lambda_{i,n}\) [where \(\Lambda_{i,n}\) is defined in (2)]. The average transmit
power for the data and pilot signals are \(P_{\text{Tx,d}} = A^2/(2T_{\text{S,d}})\)
and \(P_{\text{Tx,p}} = A^2/(2T_{\text{S,p}})\), where \(T_{\text{S,d}} = \kappa^{-1}T_{\text{S,FMX}}\)
and \(T_{\text{S,p}} = NT_{\text{S,FMX}}\) represent the effective data and pilot symbol
periods. The parameter \(\kappa\), which is the ratio between the
number of data symbols and the time multiplexed symbols in
one slot (i.e., \(\kappa = (N - 1)/N\)), is called slot efficiency.
The total average transmit power is \(P_{\text{Tx}} = A^2/(2T_{\text{S,FMX}})\)
and the transmit power per time-multiplexed symbol is
\(E_{\text{S,FMX,TX}} = A^2/2\). The corresponding transmit energy per
data symbol can be expressed as \(E_{\text{S,FMX,TX}} = \kappa^{-1}E_{\text{S,FMX,TX}}\).

### B. Channel Model

For a frequency-flat Rayleigh-fading channel, the equivalent
baseband received signal in the \(k\)th channel of a \(D\)-order
diversity reception is given by

\[
r_k(t) = g_k(t)u(t) + z_k(t), \quad k = 1, 2, \ldots, D
\]

where \(g_k(t)\) represents the zero-mean complex Gaussian-fading
process and \(z_k(t)\) is the zero-mean complex AWGN with power
spectral density of \(2N_o\). The correlation function between \(g_j(t)\)
and \(g_k(t)\) for \(j, k = 1, \ldots, D\) is modeled as [36]

\[
E\{g_j(t)g_k^*(t - \tau)\} = \begin{cases} 
\sigma_j^2\mu(\tau), & \text{for } k = j \\
\sigma_j\sigma_k\rho_{jk}\mu(\tau), & \text{for } k \neq j 
\end{cases}
\]

where \(E\{\cdot\}\) denotes the ensemble average and the \(*\) indicates
the complex conjugate. The parameter \(\sigma_j^2\) is the average power
gain of \(g_j(t)\), \(\mu(\tau)\) is the time-correlation function of the fading
process that measures the similarities between \(g_k(t)\) and \(g_k(t - \tau)\),
and \(\rho_{jk}\) is the complex cross-correlation coefficient between
two fading processes \(g_j(t)\) and \(g_k(t)\). This model assumes that
the time-correlation function is the same for each fading process
and between two fading processes. The analysis also assumes
that the fading variations are slow enough so that the Doppler
spread is much smaller than the carrier and fading
process remains constant over one symbol interval, but varies to
the next symbol. A parameter of interest is the average power
profile for the \(D\)-order diversity channel defined by (in decibel
scale)

\[
H_p(D) = [\sigma_1^2, \sigma_2^2, \ldots, \sigma_D^2]
\]

which is useful to represent the unequal power gain between
diversity channels.

### C. Receiver Structure

Receiver structures for \(M\)-DPSK and pilot-aided \(M\)-PSK
transmissions are outlined in this section and, more impor-
tantly, this section will show that the format of the decision
variable for \(M\)-DPSK, PTA, and PSA \(M\)-CPSK detectors are equivalent.
Ideal sample timing is assumed in the analysis.

1) \(M\)-DPSK: The \(M\)-DPSK receiver structure for the \(k\)th
diversity branch is shown in Fig. 3(a). The received signal \(r_k(t)\)
is passed through a matched filter and the output of which is
sampled at symbol rate \(R_s\). The sampled outputs at \(n\)th
and \((n - 1)\)th time intervals are \(U_k\) and \(K_k\), which can be written
in the form

\[
U_k = x_{U,k} + w_{U,k} \\
K_k = x_{K,k} + w_{K,k}
\]

where \(x_{U,k} = A g_{k,n} \exp[j\theta_n],\ w_{U,k} = z_{k,n},\ x_{K,k} =
A g_{k,n-1} \exp[j\theta_{n-1}],\ \text{and } w_{K,k} = z_{k,n-1}.\ \text{Variables } g_{k,n}\)
and \(z_{k,n}\) are the sampled outputs at \(n\)th time interval of \(g(t)\)
and \(z(t)\) after the matched filtering. The average received signal
power-to-noise power ratio (SNR) of \(U_k\) and \(K_k\) are

\[
\gamma_{U,k} = \gamma_{K,k} = \frac{\sigma_k^2 A^2}{2N_o}.
\]

Since \(\frac{\sigma_k^2 A^2}{2N_o} = E_{S,TX} = E_{S,RX,k}\) is the average received
energy per symbol for the \(k\)th branch, the SNR of \(U_k\) and \(K_k\) are
equivalent to the average received energy per symbol-to-noise
power spectral density ratio for the \(k\)th branch, \(\Gamma_{s,k}\), i.e.,

\[
\gamma_{U,k} = \gamma_{K,k} = \frac{E_{S,RX,k}}{N_o} = \Gamma_{s,k}.
\]

Therefore, the total average received energy per symbol-to-noise
power spectral density ratio is \(\Gamma_s = \sum_{k=1}^{D} \Gamma_{s,k}\).

2) PTA \(M\)-CPSK: The PTA \(M\)-CPSK receiver structure for the
\(D\)th order diversity branch is shown in Fig. 3(b). The received
signal \(r_k(t)\) is passed into two paths, one for the data signal
processing and the other for the pilot signal recovery. In the
data-processing path, \(r_k(t)\) is passed through a matched filter,
the output of which is sampled at symbol rate \(R_s\).
In the pilot signal recovery path, a pilot filter (PF) is used to extract
the pilot signal from \(r_k(t)\). In this analysis, the PF is assumed to
be an ideal bandpass filter, i.e., the equivalent lowpass form of
PF is an unity gain rectangular filter with bandwidth of \(B_n/2\).
To avoid distortion due to filtering, the value of \(B_n\) must be equal
to or greater than the bandwidth of the power spectrum of the
received pilot signal [14], [16], [17]. In the case of land mobile
radio environments, the bandwidth of the received pilot-tone
signal-power spectrum can be spread up to twice of the maximum Doppler frequency $f_D$ [5]. The sampled outputs at the $n$th time interval of data signal $U_k$ and channel estimate $K_k$ follow the form of (7) with $x_{l,k} = A_l g_{l,k} \exp[j\theta_{l,k}]$, $u_{l,k} = z_{l,k}$, and $A_{l,k} = \sqrt{A_l}$, and $w_{l,k} = z_{l,k}$. Variables $g_{l,k}$ and $z_{l,k}$ are the sampled outputs at the $n$th time interval of $g(t)$ and $z(t)$ after the matched filtering. The noise sample at the output of the PF is represented as $z_{p,k}$, with power of $2N_0B_w$. The analysis assumes that the noise samples $z_{d,k}$ and $z_{p,k}$ are independent. The SNR of $U_k$ and $K_k$ are

$$\gamma_{U_k} = \frac{\sigma_k^2 A_k^2}{(2N_0)}$$

$$\gamma_{K_k} = \frac{\sigma_k^2 A_k^2}{(2N_0B_w)}.$$  

Since $(\sigma_k^2 A_k^2)/2 = \sigma_k^2 E_{s_{d,k},T_{x}} = E_{s_{d,k},R_{x},k}$ is the average received data energy per symbol for the $k$th branch, the SNR of $U_k$ is equivalent to the average received data energy per symbol-to-noise power spectral density ratio for the $k$th branch $\Gamma_{s_{d,k},k}$, i.e.,

$$\gamma_{U_k} = E_{s_{d,k},R_{x,k}} = \Gamma_{s_{d,k},k}.$$  

The SNR of $U_k$ can also be expressed in terms of $\Gamma_{s_{d,k}}$ as

$$\gamma_{U_k} = \Gamma_{s_{d,k}}/(1+\eta).$$

Therefore, the total average received data energy per symbol-to-noise power spectral density ratio is $\Gamma_{s_{d}} = \sum_{k=1}^{D} \Gamma_{s_{d,k}} = \Gamma_{s}/(1+\eta)$, where $\Gamma_{s} = \sum_{k=1}^{D} \Gamma_{s,k}$ is the total average received energy per symbol-to-noise power spectral density ratio.

3) PSA M-CPSK: The PSA M-CPSK receiver structure for the $k$th order diversity branch is shown in Fig. 3(c). The received signal $r_k(t)$ after the matched filtering and sampling is demultiplexed into two paths, one for data samples and the other for pilot samples. The received pilot samples are input to the channel gain detector in which the gain of the fading process is determined. A digital interpolator, which consists of an $N$-fold rate expander and a interpolation filter (INF), is used to estimate the gain of the fading process in the data position by interpolating the samples between the normalized pilot samples. Many types of INF have been considered in the past, such as simple Gaussian filters to more advanced Wiener filters [20], [22], [23]. In this analysis, an ideal INF is assumed, i.e., the INF is an ideal lowpass filter with frequency characteristic of

$$H(f) = \begin{cases} N_s, & \text{if } |f| \leq B_w/2 \\ 0, & \text{if } B_w/2 < |f| \leq R_{s\text{min}}/2 \end{cases}$$

where $B_w/2$ denotes the bandwidth of the INF and $R_{s\text{min}} = 1/T_{s\text{min}}$. To avoid distortion due to the filtering, the value of $B_w$ must be equal to or greater than the Doppler spread (fading
bandwidth) of the channel. Sampled outputs at the $n$th time
interval of data signal $U_k$ and channel estimate $K_k$ follow the form
of (7) with $x(t_k) = A_{jk_n} \exp(j \theta_n)$, $w_{U,k} = z_{d,k_n} x_{K,k} = A_{g,k_n}$, and $w_{K,k} = z_{i,k_n}$. Variables $g_{jk_n}$ and $z_{d,k_n}$ are the
sampled outputs at the $n$th time interval of $g(t)$ and $z(t)$ after
the matched filtering. The noise sample at the output of the INF
is represented as $z_{i,k_n}$ with the power of $2N_0N\beta T_{s,n}$. The
analysis assumes that the noise samples $z_{d,k_n}$ and $z_{i,k_n}$ are inden-

tendent. The corresponding SNR of $U_k$ and $K_k$ are
\begin{align}
\gamma_{U,k} &= \frac{\sigma_\theta^2 A^2}{(2N_0)}, \\
\gamma_{K,k} &= \frac{\sigma_\theta^2 A^2}{(2N_0N\beta T_{s,n})}.
\end{align}
\[(14)\]
Since $(\sigma_\theta^2 A^2)/2 = \sigma_\theta^2 E_{s,n}P_{TX} = E_{s,n}P_{RX}$, $\gamma_{K,k}$ is the
average received energy per time multiplexed symbol for the $k$th branch,
the SNR of $U_k$ is equivalent to the average received energy per
\textit{time} multiplexed symbol-to-noise power spectral density ratio
for the $k$th branch, $\Gamma_{s,n}P_{RX}$, i.e.,
\begin{align}
\gamma_{U,k} &= \frac{E_{s,n}P_{RX,k}}{N_0} = \Gamma_{s,n}P_{RX,k}.
\end{align}
\[(15)\]
The SNR of $U_k$ can also be expressed in terms of average received
energy per data symbol-to-noise power spectral density ratio for the $k$th branch $\Gamma_{\text{ds}}$
\begin{align}
\gamma_{U,k} &= \Gamma_{s,n}P_{RX,k} = \kappa E_{s,n}P_{RX,k} = \kappa \Gamma_{\text{ds}}.
\end{align}
\[(16)\]
Therefore, the total average received energy per time multiplexed
symbol-to-noise power spectral density ratio is $\Gamma_{s,n}P_{RX} = \sum_{k=1}^{K} \Gamma_{s,n}P_{RX,k}$, $\gamma_{U,k}$ is the total average received energy per data symbol-to-noise power
spectral density ratio.

4) Diversity Combiner: Based on the analysis given above, it can be shown that if the transmission is ideal (i.e., no fading and AWGN), the phase of the demodulated signal $Z_k = U_k K_k^*$ is simply the phase of the actual transmitted signal, where $U_k$ and $K_k$ are in the forms of (7) (where variables $x(t_k)$, $w_{U,k}$, $x_{K,k}$, and $w_{K,k}$ used in (7) are given in Sections II-C-1, II-C-2, and II-C-3 for $M$-DPSK, PTA, and PSA $M$-CPK). For $D$-order postdetection diversity, the combiner sums all the demodulator outputs and forms a combined vector $Z$, which can be expressed as
\begin{align}
Z = \sum_{k=1}^{K} Z_k = \sum_{k=1}^{K} U_k K_k^*.
\end{align}
\[(17)\]
Therefore, $Z$ can be used as a decision variable for $M$-DPSK,
PTA, and PSA $M$-CPK diversity receivers.

III. ERROR-PROBABILITY ANALYSIS

Section II has shown that the format of the decision variables for $M$-DPSK, PTA, and PSA $M$-CPK are equivalent [see (17)]. Therefore, a common analysis can be applied to formulate the SEP of the three modulation schemes. In the following analysis, intersymbol interference-free transmission and zero-frequency offset between transmitted carrier and receiver oscillator

are assumed. Without a loss of generality, the information phase $A_{i,n} = 0$ can be assumed in the formulations.

1. Upper-Bound Formulation

1) Correlated Fading and Unequal Average Power Level Between Channels: Based on the analysis presented in [43], the upper bound of the SEP for $M$-DPSK, $P_s(M)$, can be expressed as
\begin{align}
P_s(M) < 2\Pr\{V < 0\}
\end{align}
\[(18)\]
where
\begin{align}
\Pr\{V < 0\} = \Pr\left\{ \sum_{k=1}^{K} (X_k Y_k^* + X_k^* Y_k) < 0 \right\}
\end{align}
\[(19)\]
represents the probability of random variable $V$ is less than zero. The two random variables $X_k$ and $Y_k$ are defined as
\begin{align}
X_k &= U_k \\
Y_k &= K_k e^{-j\psi}
\end{align}
\[(20)\]
where $\psi = \pi/2 - \pi/M$. Although the upper bound given in (18) is derived for $M$-DPSK, it is also suitable for the upper-bound development for PTA and PSA $M$-CPK as well, since the format of decision variables are equivalent. The quadratic form $V = \sum_{k=1}^{K} (X_k Y_k^* + X_k^* Y_k)$ in (19) can be expressed into a matrix form $V = \mathbf{w}^H \mathbf{Q} \mathbf{w}$, where $\mathbf{w} = [X_1, X_2, \ldots, X_D, Y_D]^T$ is a complex vector with length of $2D$ and $\mathbf{Q} = \text{diag}(J_1, J_2, \ldots, J_D)$ is the $2D \times 2D$ block diagonal matrix, with $J_k = \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}$ for $k = 1, 2, \ldots, D$. Hence, (19) can be written as
\begin{align}
\Pr\{V < 0\} = \Pr\{\mathbf{w}^H \mathbf{Q} \mathbf{w} < 0\}
\end{align}
\[(21)\]
The complex vector $\mathbf{w}$ has a zero-mean vector, i.e., $E\{\mathbf{w}\} = 0$, since both $g_{jk_n}$ and $z_{d,k_n}$ are zero mean (see Section II-B for definitions). Based on this particular setting, (21) can be solved by the technique proposed by Barrett [42]. The result is given as
\begin{align}
\Pr\{V < 0\} = \Pr\{\mathbf{w}^H \mathbf{Q} \mathbf{w} < 0\} = \sum_{\lambda_i} \prod_{j=1}^{2D} \frac{\lambda_i}{\lambda_i - \lambda_j}
\end{align}
\[(22)\]
where $\lambda_i$ for $i = 1, 2, \ldots, 2D$ are the eigenvalues of the matrix $\mathbf{Q}$. The matrix $\mathbf{C}$ is the covariance matrix of $\mathbf{w}$. Since $\mathbf{w}$ is a zero-mean random vector, $\mathbf{C}$ is reduced to the correlation matrix, i.e., $E\{\mathbf{w}^H \mathbf{w}\}$. Using (7) with $x_{U,k}$, $x_{K,k}$, and $w_{K,k}$ specified in Sections II-C-1, II-C-2, and II-C-3 for $M$-DPSK, PTA, and PSA $M$-CPK covariance matrices for $M$-DPSK, PTA, and PSA $M$-CPK schemes can be constructed as shown in Appendix I. The corresponding upper bound of SEP can be formulated by substituting (22) into (18).

3The rotation angle $\Psi$ is used to rotate the decision variable $Z$ [see (17)], so that decision error is equivalent to $V = Z e^{j\psi}$ is less than zero [43, Sec.3.2].

4Superscript $T$ and $\mathbf{H}$ represent conjugate transpose and transpose of a matrix.

5A covariance matrix of a complex vector $\mathbf{x}$ is defined as $E\{\mathbf{x}^H \mathbf{x}\}$, where $\mathbf{x} = [E\{x_1\}, E\{x_2\}, \ldots, E\{x_N\}]^T$ represents the mean vector of $\mathbf{x}$. 
2) Independent Fading and Unequal Average Power Level Between Channels: In this particular case, the matrix \( \mathbf{CQ} \) is reduced to a \( 2D \times 2D \) block diagonal matrix, which can be represented as diag(\( \mathbf{A}_1, \mathbf{A}_2, \ldots, \mathbf{A}_D \)), where

\[
\mathbf{A}_k = \begin{bmatrix}
E \{X_k Y_k^*\} & E \{X_k^2\} \\
E \{Y_k^2\} & E \{X_k Y_k\}
\end{bmatrix}, \quad k = 1, 2, \ldots, D.
\]

(23)

Based on the covariance function between \( X_k \) and \( Y_k \), which is given in Appendix I, the corresponding eigenvalues of \( \mathbf{CQ} \) are \( \lambda_1^{(A)} = \{\lambda_1^{(A_1)}, \lambda_2^{(A_1)} \}, \ldots, \lambda_1^{(A_D)}, \lambda_2^{(A_D)} \}, \) where \( \lambda_1^{(A_k)}, \lambda_2^{(A_k)} \) for \( k = 1, \ldots, D \), (see Appendix II for the derivation)

\[
\begin{align*}
\left\{ \lambda_1^{(A_k)}, \lambda_2^{(A_k)} \right\} &= \\
&= \left\{ \sqrt{\Gamma_{s,d,k}} \sin \left( \frac{\pi}{3} \right), \sqrt{\Gamma_{s,d,k}} \cos \left( \frac{\pi}{3} \right) \right\}, \quad \text{PTA CPSK} \\
&= \left\{ \sqrt{\Gamma_{s,d,k}} \sin \left( \frac{\pi}{3} \right), \sqrt{\Gamma_{s,d,k}} \frac{1}{2} \right\}, \quad \text{DPSK} \\
&= \left\{ \sqrt{\Gamma_{s,ssk}} \sin \left( \frac{\pi}{3} \right), \sqrt{\Gamma_{s,ssk}} \frac{1}{2} \right\}, \quad \text{PSA CPSK}
\end{align*}
\]

(24)

with \( \Xi_{\text{PTA}} = (\Gamma_{s,d,k} + 1)(\gamma_{s,d,k} + \beta_p) \) and \( \Xi_{\text{PSA}} = (\Gamma_{s,ssk} + 1)(\Gamma_{s,ssk} + N) \). The parameter \( \mu_s = \mu(T) \) represents the time-correlation coefficient of the fading process between two successive symbols, \( \Gamma_{s,d,k} = \Gamma_{s}/(1 + \eta) \). \( \beta_p = B_w/R_s \) is the normalized PF bandwidth, \( \Gamma_{s,ssk} = \kappa \Gamma_{s,k} \). \( N \) is the number of time multiplexed symbols per slot, and \( \beta_k = B_w/R_s \) is the normalized IF bandwidth. The upper-bound SEP of the three modulation schemes can be calculated by using (18) together with (22) and (24). It should be noted that in order to obtain meaningful results, all eigenvalues of \( \mathbf{CQ} \) must be distinct; otherwise, (22) will fail. This implies that the above result is only applicable to unequal average power between diversity branches.

3) Single-Channel Reception (\( D = 1 \)): In this case, (22) is simplified to \( \Pr \{V < 0\} = \lambda_2/(\lambda_2 - \lambda_1) \), assuming that \( \lambda_2 < 0 \). The matrix \( \mathbf{CQ} \) is reduced to a \( 2 \times 2 \) matrix in the form of (23) (i.e., \( k = 1 \)). By using \( \lambda_1 \) and \( \lambda_2 \) from (37) (see Appendix II), \( \Pr \{V < 0\} \) can now be expressed as

\[
\Pr \{V < 0\} = \frac{1}{2} \left( \frac{1 - \frac{\text{Re} \{E \{X_1 Y_1^*\}\}}{\sqrt{E \{X_1^2\} E \{Y_1^2\}} - \frac{\text{Im} \{E \{X_1 Y_1^*\}\}}{2}}}{1} \right) = \frac{(1 - F)}{2}.
\]

(25)

Hence, the upper-bound SEP given in (18) is simplified to

\[
P_b(M) < 1 - F.
\]

(26)

Based on the covariance function given in (33)–(35) (see Appendix I), for \( M \)-DPSK, PTA \( M \)-CPSK, and PSA \( M \)-CPSK, the parameter \( F \) in (25) and (26) can be written as

\[
F = \begin{cases} \\
\frac{\mu_s \Gamma_s \sin \left( \frac{\pi}{3} \right)}{\sqrt{\Xi_{\text{PTA}}} - \frac{\Gamma_{s,d}^2}{2}}, & \text{DPSK} \\
\frac{\mu_s \Gamma_s \sin \left( \frac{\pi}{3} \right)}{\sqrt{\Xi_{\text{PSA}}} - \frac{\Gamma_{s,ss}^2}{2}}, & \text{PSA CPSK} \\
\frac{\Gamma_{s,ssk} \sin \left( \frac{\pi}{3} \right)}{\sqrt{\Xi_{\text{PSA}}} - \frac{\Gamma_{s,ss}^2}{2}}, & \text{PTA CPSK}
\end{cases}
\]

(27)

with \( \Xi_{\text{PTA}} = (\Gamma_{s,d} + 1)(\eta \Gamma_{s,d} + \beta_p) \) and \( \Xi_{\text{PSA}} = (\Gamma_{s,ssk} + 1)(\Gamma_{s,ssk} + N) \).

B. Exact Bit-Error Probability (BEP) for Binary and Quaternary Level PSK

Reference [43] has shown that the exact BEP of binary- and quaternary-level PSK with Gray coding can be formulated as

\[
P_b(2) = \Pr \{ V < 0 \}, \quad \text{with } \psi = 0
\]

\[
P_b(4) = \Pr \{ V < 0 \}, \quad \text{with } \psi = \frac{\pi}{4}.
\]

(28)

Exact BEP expressions for \( B \)-, \( Q \)-DPSK, PTA, and PSA \( B \)-, \( Q \)-CPSK can be evaluated by substituting (22) into (28).

1) BEP of DPSK With \( D = 1 \): Based on (25) and (28), the BEP for \( B \)- and \( Q \)-DPSK can be derived as

\[
P_b(2) = \left\{ \frac{1 - \mu_s \Gamma_s (\Gamma_s + 1)}{2} \right\} \\
P_b(4) = \left\{ \frac{1 - \mu_s \Gamma_s [2(\Gamma_s + 1)^2 - (\mu_s \Gamma_s)^2]^{-\frac{1}{2}}}{} \right\}
\]

(29)

The above BEP expressions for \( B \)- and \( Q \)-DPSK are equivalent to (79) and (87) of [43], although written in a different form. Both expressions reduce to the results given in (8.173) and (8.174) of [9] if zero Doppler spread is assumed (i.e., \( \mu_s = 1 \)).

2) BEP of PTA CPSK With \( D = 1 \): The BEP for \( B \) and \( Q \)-CPSK can be expressed as

\[
P_b(2) = \left\{ \frac{1 - \Gamma_{s,d}(\Gamma_{s,d} + 1)}{2} \right\} \\
P_b(4) = \left\{ \frac{1 - \Gamma_{s,d} [2(\Gamma_{s,d} + 1)(\Gamma_{s,d} - \beta_p \eta - \frac{1}{2})]}{2} \right\}
\]

(30)

where \( \Gamma_{s,d} = \Gamma_s/(1 + \eta) \). Although written in a different form, the BEP expression for PTA \( B \)-CPSK given above is equivalent to (71) of [15] and (43) of [16]. If the PF output is noise free and all available power is allocated to the data signal (i.e., \( \beta_p = 0 \) and \( \eta \to 0 \)), (30) approaches the expression for ideal coherent detection [8].

6The equations assume that the information phase \( \lambda_{\text{app}} = 0 \) for both binary- and quaternary-level PSK and zero frequency offset between the transmitter and receiver oscillator.
DPSK and \( M \)-DPSK in correlated Rayleigh fading. Correlated fading is assumed in both the triangular and uniform linear array formations. \( \mu_s \) is the time-correlation of the fading process between two successive symbols.

3) **BEP of PSA CPSK With \( D = 1 \)**: BEP for PSA B and \( Q \)-CPSK can be expressed in terms of \( \Gamma_{\text{ds}} \) as

\[
\begin{align*}
P_b(2) &= \frac{1 - \Gamma_{\text{ds}} \left[ \Gamma_{\text{ds}} + N \beta \kappa \right]}{2} \\
P_b(4) &= \frac{1 - \Gamma_{\text{ds}} \left[ \Gamma_{\text{ds}} + N \beta \kappa \right] - \Gamma_{\text{ds}}^2}{2}
\end{align*}
\]

where \( \kappa \) is the slot efficiency that is defined in Section II-A-3. The BEP expression for \( B \)-PSK is equivalent to the one given in [21] and [41]. In the case of the \( Q \)-PSK expression, an equivalent expression has been derived by [21]. If the INF output is noise free and the slot length \( N \gg 1 \), i.e., \( \beta \approx 0 \) and \( \kappa \rightarrow 1 \), (31) approaches the expression for ideal coherent detection [8].

IV. NUMERICAL RESULTS AND DISCUSSION

A. **Exact BEP for \( B \)- and \( Q \)-Level PSK**

Numerical results for the average BEP of \( B \)-DPSK in correlated Rayleigh-fading channels with third-order diversity reception were evaluated in [32]. Both triangular and uniform linear antenna array configurations with specified power cross-correlation values between antenna branches were considered [see (52) and (53) of [32]]. The analysis assumed that the equal power level between all diversity channels and the fading is slow enough so that the channel gain remains constant for two successive symbol periods (i.e., \( \mu_s = 1.0 \)). Using the \( B \)-DPSK expression [see (28)], derived in the last section [with \( \mu_{jk} \) equals to the square root of the power correlation coefficient given in (52) and (53) of [32]], Fig. 4 reproduces the average BEP of \( B \)-DPSK for both triangular and linear antenna array configurations. To approximate the equal diversity channel power and slow-fading conditions, the average power profile for the three diversity channels and the time-correlation of the fading process between two successive symbols are set to \( H_p(3) = [0, -0.1, -0.2] \) and \( \mu_s = 0.99998 \) in the calculation. Results given in Fig. 4 are in good agreement with the numerical results presented in [32, Fig. 4]. When compared to the independent fading, the error performance degradation due to the correlated fading between diversity branches is evident. The graph also shows that the BEP of DPSK degrades further when the time-correlation value between the two symbols decreases.

Fig. 5 shows the average BEP of PTA \( Q \)-CPSK for various diversity order and normalized PF bandwidth \( \beta_p \). Independent fading and unequal power level between diversity channels with \( H_p(2) = [0, -3] \) and \( H_p(4) = [0, -3, -5, -10] \) are considered. The pilot-to-data power ratio \( \eta \) is fixed to \(-7 \) dB for all cases. The BEP curves for ideal coherent demodulated \( Q \)-PSK, derived in [8], are also given in Fig. 5. Results show the performance loss of using pilot-aided technique and the effects of noisy reference signal for channel estimation. Although not shown in the graph, the BEP of PTA \( Q \)-CPSK approaches ideal coherent demodulation of \( Q \)-PSK when \( \beta_p = 0 \) and \( \eta \rightarrow -\infty \) dB.
B. Upper-Bound SEP for M-ary Level PSK

Exact and upper-bound SEP of PTA 4, 8, 16, and 32 CPSK with single-channel reception are illustrated in Fig. 6. The parameters $\eta$ and $\beta_p$ are fixed to $-7$ dB and 0.05 for all cases. The exact SEP are calculated by using (70) in [15] where single integration is required. The upper bounds are calculated by using (26). Results show the improvement of the bound when the modulation level increases. For $M > 8$, the difference between the exact and upper-bound curves is unnoticeable. The average SEP for the ideal coherent demodulations are also presented as references. The result illustrates the performance degradation due to pilot-aided transmission.

Recently, the error performance of ideal coherent demodulated M-PSK with uniform linear antenna array reception has been analyzed in [40]. The effects of correlated fading due to antenna spacing and operating environments, such as angle of arrival (AOA) and beamwidth of incoming signals, were considered. The antenna array’s cross-correlation model used in [40] has a property of $\rho_{kj} = \rho_{jk}^*$, for $j, k = 1, \ldots, D$. By using the same cross-correlation model, Fig. 7 shows the upper-bound SEP of PSA 8-CPSK as a function of beamwidth for two- and four-branch linear antenna array reception. The antenna spacing $x$ is set to half of carrier wavelength and $\Gamma_{\text{half}} = 20$ dB. The relative average power profile (in decibels) is $H_p(2) = [0, -0.1]$ and $H_p(4) = [0, -0.1, -0.15, -0.2]$. The parameter $\beta_p$ is set to 1.0 (i.e., zero Doppler spread).

C. Imperfect Correlation Between Channel Estimate and Data Signal

This effect is already addressed in the $M$-DPSK analysis, where imperfect correlation between the previous signal (channel estimate) $K_k$ and current signal (data signal) $U_k$ is due to the imperfect time correlation of the fading process $\mu_s$. In the case of pilot-aided $M$-CPSK analysis, perfect correlation between channel estimate and data signal is assumed in Section II-C. However, this assumption is not valid when the bandwidth of PF and INF is narrower than the bandwidth of the fading spectrum [16], [17], [20]–[22]. By following the analysis given in [16], the effect of decorrelation between the channel estimate and data signal on PTA system can be incorporated by modifying the covariance matrix $C$. As in Section II-C.2, assuming that the equivalent lowpass PF is an unity gain rectangular filter with bandwidth of $B_{\text{PF}}/2$, the elements $E[|Y_k|^2]$ are modified to $\varepsilon \eta \Gamma_{\text{half}} G_k R_{\text{PF}} + B_{\text{PF}}$, where $\varepsilon = \int_{-B_{\text{PF}}/2}^{B_{\text{PF}}/2} S_g(f) |f| df$ and has a range of $0 \leq \varepsilon \leq 1$. $S_g(f)$ is the normalized power spectral density of the fading process $g_k(t)$ with fading bandwidth of $B_{\text{PF}}/2$. The parameter $\varepsilon$ is also required in $E[X_k Y_k^*], E[Y_j^* Y_k^*], E[X_j Y_k^*], E[X_j^* Y_k^*]$, as well as in their complex conjugate pairs. Fig. 8 shows the BEP of PTA Q-CPSK with different values of $\nu$ (the ratio of $B_{\text{PF}}$ to $B_{\text{PF}}$). In the calculation, land mobile radio channel model with the “U”-shape power spectral density (i.e., $S_g(f) = [\pi (f^2 - f_0^2)^{\nu/2}]^{-1}$) [7] and normalized maximum Doppler frequency $(f_{\text{DP}} T_s)$ of 0.04 were assumed. For this model, $\nu = B_{\text{PF}}/(2 f_D)$; $\beta_p = 2 \nu f_D T_s$. \(\Gamma_{b,k} = \Gamma_{b}/\log_2 M\) (dB)
and $\varepsilon = (2/\pi) \arcsin(\min(1, \nu))$. Performance degradation due to decorrelation between the channel estimate and data signal on PTA $Q$-CPJK is illustrated in Fig. 8. Results show that the error performance is very sensitive to the band edge distortion (i.e., for $\nu < 1$). This is due to the fact that the "U"-shape spectrum has high proportion of power near $\pm f_D$ [16]. This figure also shows the BEP curves for ideal coherent demodulation $B$-PSK and $B$-DPSK with $h_k = J_0(2\pi f_D T_b)$, where $J_0(.)$ is the zero-order Bessel function of the first kind and $T_b$ represents the bit period. Results given in Fig. 8 provide BEP comparison between the three transmission techniques under common transmit information bit rate and occupied bandwidth conditions.

V. CONCLUSION

In this paper, upper-bound expressions for SEPs of $M$-DPSK, PTA, and PSA $M$-CPJK over frequency-flat Rayleigh-fading channels have been formulated. Postdetection maximal-ratio combining technique is assumed at the receiver. The derived upper bounds are able to investigate the effect of AWGN, correlated fading, and unbalanced power level between diversity branches. The effect of Doppler spread on $M$-DPSK detection, noisy-channel estimation on PTA, and PSA $M$-CPJK detections are also taken into account in the formulations. Expressions to calculate the exact bit-error probability of binary and quaternary DPSK, PTA, and PSA CPJK are also evaluated. To validate the accuracy of the upper bound, comprehensive sets of numerical results produced by the upper bound have been compared with the previous published results.

APPENDIX I

COVARIANCE FUNCTIONS

The $2D \times 2D$ covariance matrix $C$ described in Section III is defined by

$$
C = E \left\{ \begin{bmatrix} X_1^2 & X_1 Y_1^* & \cdots & X_1 X_D^* & X_1 Y_D^* \\
X_1 Y_1 & |Y_1|^2 & & Y_1 X_D^* & Y_1 Y_D^* \\
\vdots & \vdots & \ddots & \vdots & \vdots \\
X_1^* X_D & Y_1^* X_D & \cdots & |X_D|^2 & X_D Y_D^* \\
X_1^* Y_D & Y_1^* Y_D & \cdots & Y_D X_D^* & |Y_D|^2 \end{bmatrix} \right\}
$$

(32)

All matrix elements in $C$ can be derived by using (7) where variables $x_{i,k}$, $w_{i,k}$, $r_{i,k}$, and $w_{i,k}$ used in (7) are given in Sections II-C-1, II-C-2, and II-C-3 for $M$-DPSK, PTA, and PSA $M$-CPJK and along with the channel covariance function specified in (5). The corresponding covariance elements for $M$-DPSK, PTA, and PSA $M$-CPJK schemes are derived below. All expressions below are normalized to $2N_0$ so that results can be expressed in terms of SINR at the $k$th diversity branch. In the following, zero frequency offset between transmitter and receiver oscillators is assumed.

$M$-DPSK: Covariance between $X_j$, $Y_j$, $X_k$, and $Y_k$ for diversity branch $j$ and $k$ is

$$
E \{X_k X_k^*\} = E \{Y_k Y_k^*\} = \Gamma_{s,k} + 1
$$

$$
E \{X_k Y_k^*\} = E \{K_k^2\} = \Gamma_{s,k} + 1
$$

$$
E \{X_k Y_j^*\} = E \{U_k K_j^*\} e^{j\phi} = \Gamma_{s,k} h_k e^{j\phi}
$$

$$
E \{X_k Y_k^*\} = (E \{X_k Y_j^*\})^*$
$$

$$
E \{X_j X_j^*\} = E \{U_j U_j^*\} = \rho_{jk} (\Gamma_{s,j})^{1/2} (\Gamma_{s,k})^{1/2}
$$

$$
E \{X_j Y_j^*\} = (E \{X_j X_j^*\})^*
$$

$$
E \{Y_j Y_j^*\} = E \{K_j^2\} = \rho_{jk} (\Gamma_{s,j})^{1/2} (\Gamma_{s,k})^{1/2}
$$

$$
E \{Y_j Y_k^*\} = (E \{Y_j Y_j^*\})^*
$$

$$
E \{X_j Y_k^*\} = E \{U_j K_j^*\} e^{j\phi} = \rho_{jk} h_k (\Gamma_{s,j})^{1/2} (\Gamma_{s,k})^{1/2} e^{j\phi}
$$

$$
E \{X_j Y_k^*\} = (E \{X_j Y_j^*\})^*$
$$

(33)

All expressions given in (9).

PTA $M$-CPJK: Covariance between $X_j$, $Y_j$, $X_k$, and $Y_k$ for diversity branch $j$ and $k$ is

$$
E \{X_k X_k^*\} = E \{Y_k Y_k^*\} = \Gamma_{s,d,k} + 1
$$

$$
E \{X_k Y_k^*\} = E \{K_k^2\} = \Gamma_{s,d,k} + B_w
$$

$$
E \{X_k Y_j^*\} = E \{U_k K_j^*\} e^{j\phi} = \Gamma_{s,d,k} \sqrt{\eta} R_s e^{j\phi}
$$

$$
E \{X_k Y_j^*\} = (E \{X_k Y_j^*\})^*
$$

$$
E \{X_j X_j^*\} = E \{U_j U_j^*\} = \rho_{jk} \sqrt{\Gamma_{s,d,j}} \sqrt{\Gamma_{s,d,k}}
$$

$$
E \{X_j Y_j^*\} = (E \{X_j X_j^*\})^*
$$

$$
E \{Y_j Y_j^*\} = E \{K_j^2\} = \rho_{jk} \sqrt{\Gamma_{s,d,j}} \sqrt{\Gamma_{s,d,k}}
$$

$$
E \{Y_j Y_k^*\} = (E \{Y_j Y_j^*\})^*
$$

$$
E \{X_j Y_k^*\} = E \{U_j K_j^*\} e^{j\phi} = \rho_{jk} \sqrt{\Gamma_{s,d,j}} \sqrt{\eta} R_s e^{j\phi}
$$

$$
E \{X_j Y_k^*\} = (E \{X_j Y_j^*\})^*$

(34)

where $\Gamma_{s,d,k}$ is given in (11).
PSA M-CPSK: Covariance between $X_{j}$, $Y_{j}$, $X_{k}$, and $Y_{k}$ for diversity branch $j$ and $k$ is

$$
E \{ X_{k}^{*} \} = E \{ U_{k}^{*} \} \\
= \Gamma_{\text{splice}}K_{k} + 1 \\
E \{ Y_{k}^{*} \} = E \{ U_{k}^{*} \} \\
= \Gamma_{\text{splice}}K_{k}e^{j\psi} \\
E \{ X_{j}Y_{k}^{*} \} = E \{ X_{j}K_{k}^{*} \}e^{j\psi} \\
E \{ X_{j}X_{k}^{*} \} = E \{ X_{j}K_{k}^{*} \} \\
= \rho_{jk}(\Gamma_{\text{splice}}K_{j})^{1/2}(\Gamma_{\text{splice}}K_{k})^{1/2} \\
E \{ Y_{j}Y_{k}^{*} \} = E \{ U_{j}K_{k}^{*} \}e^{j\psi} \\
E \{ Y_{j}X_{k}^{*} \} = E \{ U_{j}K_{k}^{*} \} \\
= \rho_{jk}(\Gamma_{\text{splice}}K_{j})^{1/2}(\Gamma_{\text{splice}}K_{k})^{1/2}e^{j\psi} \\
E \{ Y_{j}X_{j}^{*} \} = E \{ X_{j}Y_{j}^{*} \}^{*} \\
= \rho_{kk}(\Gamma_{\text{splice}}K_{j})^{1/2}e^{j\psi} \\
(35)
$$

where $\Gamma_{\text{splice}}K_{k}$ is given in (15).

**APPENDIX II**

**EIGENVALUES OF BLOCK DIAGONAL MATRICES**

A $2D \times 2D$ block diagonal matrix $C = \text{diag}(A_{1}, A_{2}, \ldots, A_{D})$ with $A_{k} = [a_{k} \quad x_{k}^{*} \quad y_{k} \quad a_{k}^{*}]^{T}$ for $k = 1, 2, \ldots, D$, where $a_{k}$ is a complex variable and $x_{k}$ and $y_{k}$ are real variables. The corresponding eigenvalues of $C$ are

$$
\lambda(C) = \left\{ \lambda_{1}(A_{1}), \lambda_{2}(A_{1}), \ldots, \lambda_{1}(A_{D}), \lambda_{2}(A_{D}) \right\} \\
(36)
$$

where

$$
\begin{align*}
\lambda_{1}(A_{k}) &= \text{Re}\{a_{k}\} \pm \sqrt{3|a_{k}y_{k}| - (\text{Im}\{a_{k}\})^{2}}. \\
\lambda_{2}(A_{k}) &= \text{Re}\{a_{k}\} \pm \sqrt{3|a_{k}y_{k}| - (\text{Im}\{a_{k}\})^{2}}.
\end{align*}
\tag{37}
$$

**ACKNOWLEDGMENT**

The authors are grateful to M. Fitton, M. Ismail, and K. Rizvi, Toshiba Research Europe Limited, Bristol, U.K., for their valuable comments and suggestions in relation to the preparation of this paper. The authors also would like to thank the reviewers for their valuable and constructive comments, which improved the accuracy and presentation of this paper.

**REFERENCES**


Yuk C. Chow (S’91–A’97) was born in Hong Kong, in 1969. He received the B.Eng. degree from the University of Glasgow, Glasgow, U.K., in 1990, the M.Sc. and D.I.C. degrees from Imperial College, University of London, London, U.K., in 1991, and the Ph.D. degree from the University of Bristol, Bristol, U.K., in 1997.

From 1995 to 1997, he was a Research Assistant in the Department of Electrical and Electronic Engineering, University of Bristol. He was with Motorola Semiconductors Hong Kong Ltd., from 1997 to 1999. He now is a Principal Research Engineer with Toshiba Research Europe Ltd., Bristol, U.K. His research interests include digital modulations, receiver-processing techniques, and transmitter/receiver system modeling.

Joe P. McGeehan received the B.Eng. and Ph.D. degrees in electrical and electronic engineering from the University of Liverpool, Liverpool, U.K., in 1967 and 1971, respectively.

He now is a Professor of Communications Engineering and Dean of the University of Bristol. He also is Managing Director of the Telecommunications Research Laboratory, Toshiba Research Europe Ltd., Bristol, U.K. He has been actively researching spectrum-efficient mobile radio communication systems since 1973 and has pioneered work in many areas, including linear modulation, linearized power amplifiers, smart antennas, propagation modeling/prediction using ray tracing, and phase-locked loops.

Dr. McGeehan is a Fellow of the Royal Academy of Engineering and of the Institution of Electrical Engineers (IEE), U.K. He has served on numerous international committees and standards bodies and was advisor to the U.K.’s first DTI/MOD “Defence Spectrum Review Committee” in the late 1970s. He was the Joint Recipient of the IEEE TRANSACTIONS ON VEHICULAR TECHNOLOGY Award—Neal Shepherd Memorial Award (for work on smart antennas) and of the IEE Proceedings Mountbatten Premium for work on satellite tracking and frequency-control systems.