On the Error Performance of Wireless Systems with Frequency Selective Fading and Receiver Timing Phase Offset

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***Abstract*— Receiver timing phase is one of the essential factors defining the performance of wireless communication systems. In this paper, we investigate the effects of timing phase offset, which is introduced by the phase difference between the transmitter clock and receiver clock, on the performance of wireless systems over frequency selective fading. With frequency domain analysis, the instantaneous signal-to-noise ratio (SNR) observed by the communication receiver is expressed as an explicit function of system timing phase offset and receiver oversampling factor. A tight performance lower bound, which corresponds to the best possible system performance under particular system configuration, is then derived by examining the statistical properties of the receiver SNR. From the analytical results, it is observed that, if the receiver sampling rate is less than the Nyquist rate of the received signal, then the system performance lower bound is a periodic function of the timing phase offset. On the other hand, the best possible performance of the oversampled system is independent of timing phase offset. Moreover, the oversampled system can use a receive filter matched to the time-invariant transmit filter instead of a statistical filter matched to the joint response of channel and transmit filter without affecting the best possible system performance. Simulation results show that the theoretical bound derived in this paper can accurately predict the performance of practical communication systems suffering from both frequency selective fading and timing phase offset.**

***Index Terms*— Fractionally spaced receiver, frequency selective fading, matched filter bound, timing phase sensitivity.**

I. Introduction

IT is well known that the performance of communication systems with a symbol spaced sampler suffers from extreme sensitivity to receiver timing phase offset, which is introduced by the phase difference between the transmitter clock and receiver clock. It is pointed out in [1] and [2] that the dependence of system performance on timing phase offset is due to the effects of spectrum aliasing of the sampled signals at the receiver. For different system configurations, the overlapped spectral components of the signal samples at the receiver could add up either constructively or destructively based on their phase difference, and this leads to performance enhancement or degradation, accordingly. The phase differences among the overlapped spectral components is a direct result of timing phase offset at the receiver. The relationship between the receiver timing phase and system performance fluctuation is heuristically discussed in [1] and [2]. However, no analytical result is available in the literature to quantify the effects of timing phase offset on system error performance. In this paper, with the help of the matched filter bound technique, we will derive a tight theoretical performance bound that is able to quantitatively identify the effects of both timing phase offset and receiver oversampling.

The matched filter bound is a well known technique used to predict the performance for systems experiencing frequency selective fading [3]-[8]. By assuming that there is no intersymbol interference (ISI) present at the receiver, the matched filter bound defines the best possible error performance under particular system configurations. The matched filter bounds presented in most previous works are loose performance lower bounds, and they are usually far below the actual error performance of practical communication systems. At the first glance, it seems that the performance difference is a result of the ISI free assumption. However, it is shown in [9] that both maximum likelihood sequence estimation (MLSE) and maximum *a posteriori* (MAP) equalizers are asymptotically optimum in the sense of interference cancellation, *i.e.*, the ISI components at the output of MLSE equalizer or MAP equalizer tend to zero provided that the decoding length is long enough. Indeed, conventional matched filter bounds do not capture the effects of timing phase offset and receiver oversampling, both of which have significant impact on communication system performance. We will show in this paper that the performance difference is mainly due to the sampler timing phase and spectrum aliasing at the receiver.

To remove the effects of spectrum aliasing at the receiver, fractionally spaced equalizers are discussed in [1], [10]-[12]. In [10] and [11], the performances of systems with various fractionally spaced receivers are investigated with simulations. The theoretical performance of fractionally spaced equalizer

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is analyzed in [12] with the union bound technique, where the pairwise error probabilities of mutually overlapped error events are added up as an upper bound of system error probability. It is well known that the union bound is rather loose compared to the actual system performance, especially at low signal-to-noise ratio (SNR). Moreover, the union bound technique cannot quantify the effects of timing phase offset and spectrum aliasing.

In this paper, a tight performance lower bound for systems experiencing uncorrelated scattering frequency selective fading is derived by considering the effects of both receiver oversampling and timing phase offset. With the help of Karhunen-Loève expansion, a unified error probability expression is derived as a tight lower bound for the performance of various linearly modulated communication systems. The effects of timing phase offset, receiver oversampling, as well as the power delay profile of the frequency selective fading are explicitly expressed in the statistical representations of the instantaneous SNR observed at the receiver. The conventional matched filter bound can be treated as a special case of the new performance bound. It has been observed from the analytical results that when there is no spectrum aliasing present at the receiver, the best possible performance of a system with receive filter matched to the joint response of transmit filter and frequency selective fading is the same as that of the system with receive filter matched to the transmit filter alone. Simulation results will also demonstrate that the performance bound derived in this paper can accurately predict the performance of communication systems with practical receivers and various modulation schemes in a wide range of SNR.

The rest of the paper is organized as follows. Section II presents the system model used for analysis. In Section III, a tight error performance bound for systems experiencing timing phase offset is derived by analyzing the statistical properties of the instantaneous SNR at the receiver. Based on the new performance bound, case studies of several representative communication systems are carried out in Section IV to investigate the effects of timing phase offset and receiver oversampling on system performance. Numerical examples are provided in Section V, and Section VI concludes the paper.

II. System Model

Let $g(t)$ denote the impulse response of the frequency selective fading channel. The channel is assumed to be quasi static, meaning that the impulse response $g(t)$ remains invariant per transmission burst but may change from burst to burst. Define the composite impulse response (CIR) of the system as

$$h(t) = p_T(t) \otimes g(t) \otimes p_R(t), \tag{1}$$

where $\otimes$ denotes the operation of convolution, $p_T(t)$ and $p_R(t)$ are the normalized impulse response of the transmit filter and receive filter, respectively, and they have unit energy.

The signal at the output of the receive filter can be represented by

$$z(t) = \sum_{i=-\infty}^{+\infty} s_i \cdot h(t - iT_{sym}) + n(t) \otimes p_R(t), \tag{2}$$

where s_i is the M-ary modulated information symbol with symbol period T_{sym} and average energy E_s, and $n(t)$ is the additive white Gaussian noise (AWGN) with variance N_0.

The receive filter is followed by a sampler with sampling period $T_s = T_{sym}/\mu$, with the integer μ being the oversampling factor. At sampling instant $t = kT_s + \tau_0$, the sampled output of the receive filter, $z(k) = z(kT_s + \tau_0)$, can be expressed as

$$\begin{aligned} z(k) &= \sum_{i=-\infty}^{+\infty} x_i \cdot h(k-i) + v(k) \\ &= x_k \cdot h(0) + \sum_{\substack{l=-\infty \\ l\neq 0}}^{+\infty} h(l) \cdot x_{k-l} + v(k), \end{aligned} \tag{3}$$

where $\{x_i\}$ is the μ-times oversampled sequence of $\{s_i\}$ with $x_i = s_i$ when $\frac{i}{\mu}$ is integer and $x_i = 0$ when $\frac{i}{\mu}$ is non-integer. Moreover, $v(k)$ is the sample of the noise component $v(t) = n(t) \otimes p_R(t)$, $\tau_0 \in \left[-\frac{T_s}{2}, \frac{T_s}{2}\right]$ is the phase difference between the sampler clock and the transmitter clock, and $h(k) = h(kT_s + \tau_0)$ is the discrete-time version of the CIR $h(t)$. Since the transmit filter usually falls off rapidly with increase of time, it is reasonable to assume that $h(k)$ has finite time domain support [13]. Without loss of generality, it is assumed that the index of the discrete-time CIR $h(l)$ satisfies $l \in [0, L]$.

Stacking up all the received samples related to the T_s-spaced information symbol x_k, we can write the discrete-time system representation of (3) into matrix format as

$$\mathbf{z}_k = x_k \cdot \mathbf{h} + \mathbf{I}_k + \mathbf{v}_k \tag{4}$$

where the vectors $\mathbf{z}_k = [z(k), z(k+1), \cdots, z(k+L)]^T \in \mathcal{C}^{(L+1)\times 1}$, $\mathbf{v}_k = [v(k), v(k+1), \cdots, v(k+L)]^T \in \mathcal{C}^{(L+1)\times 1}$ comprise all the received samples and noise samples related to the T_s-spaced information symbol x_k, with $\mathbf{A}^T$ denoting matrix transpose. The CIR vector $\mathbf{h} = [h(0), h(1), \cdots, h(L)]^T \in \mathcal{C}^{(L+1)\times 1}$ contains all the $(L+1)$ channel taps of the discrete-time CIR that might contribute to the detection of x_k. On the other hand, the vector $\mathbf{I}_k \in \mathcal{C}^{(L+1)\times 1}$ contains all the ISI components relative to the desired symbol x_k, with the ith element of $\mathbf{I}_k$ being $(\mathbf{I}_k)_i = \sum_{\substack{l=0 \\ l\neq(i-1)}}^{L} h(l) \cdot x_{k+i-1-l}$.

The ISI free assumption employed by the matched filter bound is adopted in our performance analysis by discarding the ISI components $\mathbf{I}_k$

$$\mathbf{z}_k = x_k \cdot \mathbf{h} + \mathbf{v}_k. \tag{5}$$

In (5), all the received samples and channel taps related to x_k are collected in an equivalent single-input multiple-output (SIMO) system representation. This model is now equivalent to a frequency nonselective fading channel model (no ISI) with receive diversity. The optimum detection of the equivalent SIMO system with optimum combining performed over all channel taps of the vector $\mathbf{h}$ will lead to the best possible performance for the system described in (3).

The noise sample $v(k)$ is a linear transformation of AWGN $n(t)$. Hence, it is zero-mean Gaussian distributed with the auto-correlation function $r_{vv}(m-n) = \mathbb{E}\left[v(m)v^*(n)\right]$ given by [13]

$$r_{vv}(m-n) = N_0 \cdot r_{p_R p_R}\left[(m-n)T_s\right], \tag{6}$$

where $\mathbb{E}(x)$ is the operation of mathematical expectation, and $r_{p_R p_R}(t) = \int_{-\infty}^{+\infty} p_R(t+\tau)p_R^*(\tau)d\tau$ is the auto-correlation function of the receive filter $p_R(t)$. Due to the time span of the receive filter and the effects of oversampling, the noise component $v(k)$ becomes a colored Gaussian process with auto-correlation function defined in (6). The power spectral density (PSD) $\widehat{R}_{vv}(f)$ of $v(k)$ is

$$\widehat{R}_{vv}(f) = \frac{N_0}{T_s} \sum_{n=-\infty}^{+\infty} R_{p_R p_R}\left[\frac{f-n}{T_s}\right], \quad -f_0 \le f \le f_0. \quad (7)$$

where $f \in [-1/2, 1/2]$ is the digital frequency of discrete-time signals, $f_0 \in (0, 1/2]$ is the digital bandwidth of the receive filter, and $R_{p_R p_R}(F)$ is the Fourier transform (FT) of the continuous-time auto-correlation function $r_{p_R p_R}(t)$, with $F = f/T_s$ being the analog frequency. It should be noted from (7) that the statistical property of the sampled noise component $v(k)$ is independent of the timing phase offset τ_0.

Similarly, based on (1) and the sampling theorem, the frequency domain representation of the discrete-time channel can be obtained through discrete-time Fourier transform (DTFT) of the CIR vector $\mathbf{h}$ as

$$\begin{aligned}\widehat{H}(f) \;=\; & \frac{e^{j2\pi f \frac{\tau_0}{T_s}}}{T_s} \sum_{n=-\infty}^{+\infty} P_T\left(\frac{f-n}{T_s}\right) G\left(\frac{f-n}{T_s}\right) \\ & \cdot P_R\left(\frac{f-n}{T_s}\right) e^{-j2\pi n \frac{\tau_0}{T_s}}, \quad -f_0 \le f \le f_0. \quad (8)\end{aligned}$$

where $j^2 = -1$, $P_T(F)$, $P_R(F)$ and $G(F)$ are the Fourier transforms of $p_T(t)$, $p_R(t)$ and $g(t)$, respectively. It should be noted that the frequency domain support of $\widehat{H}(f)$ is smaller than or equal to that of the noise spectrum due to the limited bandwidth of the receive filter $p_R(t)$.

With the frequency domain representation given in (7) and (8), the instantaneous SNR at the output of an optimum receiver for the ISI free system can be expressed by [14, p.390, eqn. (8)]

$$\gamma = \gamma_0 \cdot \int_{-F_0}^{F_0} |\Psi(F)|^2 dF, \quad (9)$$

with the function $\Psi(F)$ being defined as

$$\Psi(F) = \frac{\sum_{n=-\infty}^{+\infty} R_{P_T P_R}(F-nF_s)\, G(F-nF_s)\, e^{-j2\pi n \frac{\tau_0}{T_s}}}{\sqrt{\sum_{n=-\infty}^{+\infty} R_{p_R p_R}(F-nF_s)}}. \quad (10)$$

In (9) and (10), $F_s = 1/T_s$ is the sampling rate, $F_0 = f_0/T_s \in (0, \frac{1}{2T_s}]$ is the analog bandwidth, $\gamma_0 = E_s/N_0$ with E_s being the symbol energy, $R_{P_T P_R}(F) = P_T(F)P_R(F)$, and the integration variable has been changed to the analog frequency $F = f/T_s$ in (9).

It is interesting to note from (10) that the SNR γ is a periodic function of the timing phase offset τ_0 with the fundamental period equal to the sampling period T_s. Thus, it is sufficient for us to examine the system behavior with τ_0 in the range of $[-\frac{T_s}{2}, \frac{T_s}{2}]$.

III. Error Performance of System with Timing Phase Offset

The error performance of a linearly modulated system with timing phase offset is investigated in this section by analyzing the statistical properties of the instantaneous SNR at the receiver.

A. Statistical Properties of SNR

For Rayleigh fading channel, the Fourier transform $G(F)$ of the channel impulse response is zero-mean complex Gaussian distributed. Thus, the function $\Psi(F)$, which is a linear combination of $G(F)$ as in (10), is also a zero-mean Gaussian process in the frequency domain F.

To facilitate the analysis of the statistical properties of the instantaneous SNR γ, Karhunen-Loève expansion is applied to the Gaussian process $\Psi(F)$ in the frequency domain. As a result, we get

$$\Psi(F) = \sum_{l=1}^{L} \sqrt{\lambda_l} \sum_{k=1}^{K_l} w_{l,k}\phi_{l,k}(F), \quad (11)$$

where $\{w_{l,k}\}$ are a set of independent identically distributed (i.i.d.) zero-mean Gaussian random variables with unit variance, $\{\lambda_l\}$ are a set of distinct eigenvalues of the function $\Psi(F)$, K_l is the number of eigenvalues sharing identical values λ_l, $\{\phi_{l,k}(F)\}$ the corresponding orthonormal eigenfunctions with frequency domain support $[-F_0, F_0]$, and they satisfy $\int_{-F_0}^{F_0} \phi_{l,k}(F)\phi_{m,i}^*(F)dF = \delta_{l,m}\delta_{k,i}$, with $\delta_{l,m}$ being the Kronecker delta function.

Given the fact that the set of eigenfunctions $\{\phi_{l,k}(F)\}$ are orthonormal, we can get an alternative representation of the instantaneous SNR by substituting (11) into (9),

$$\gamma = \gamma_0 \cdot \sum_{l=1}^{L} \lambda_l \sum_{k=1}^{K_l} |w_{l,k}|^2, \quad (12)$$

In (12), the instantaneous SNR γ is expressed as the summation of L independent χ^2-distributed random variables $\sum_{k=1}^{K_l} |w_{l,k}|^2$. Thus, the characteristic function (CHF) of γ can be expressed as [15], [16]

$$\begin{aligned}\Phi_\gamma(\omega) &= \mathbb{E}\left(e^{j\omega\gamma}\right) \\ &= \prod_{l=1}^{L} (1 - j\omega\lambda_l\gamma_0)^{-K_l}. \quad (13)\end{aligned}$$

It is apparent from (12) and (13) that the statistical properties of γ is uniquely determined by the eigenvalues λ_l of the random function $\Psi(F)$ as defined in (10).

The analysis of the statistical properties of γ requires the knowledge of the eigenvalues λ_l. To solve the eigenvalues, we formulate the following eigensystem representation from (11) by utilizing the orthonormal properties of the eigenfunctions $\phi_{l,k}(F)$,

$$\int_{-F_0}^{F_0} R_\Psi(F_1, F_2)\phi_{l,k}(F_2)dF_2 = \lambda_l\phi_{l,k}(F_1), \quad (14)$$

where $R_\Psi(F_1, F_2) = \mathbb{E}[\Psi(F_1)\Psi^*(F_2)]$ is the frequency domain auto-correlation function of the random function $\Psi(F)$, and the mathematical expectation operation is performed over the statistical channel response $G(F)$.

For a system with arbitrary power delay profile, the eigensystem described in (14) can be solved with composite Simpson's numerical integration. Divide the integration interval $[-F_0, F_0]$ into $2N$ subintervals, with the length of each subinterval being $\Delta = F_0/N$, and define $F_n = -F_0 + (n-1)\Delta$, for $n = 1, 2, \cdots, 2N+1$, then the integral of (14) can be numerically approximated by [4]

$$\mathbf{R}_{\Psi}\mathbf{D}\cdot\boldsymbol{\phi}_{l,k} = \lambda_l \cdot \boldsymbol{\phi}_{l,k}, \tag{15}$$

where $\mathbf{R}_{\Psi}$ is a $(2N+1)\times(2N+1)$ matrix with the (m,n)th element being $(\mathbf{R}_{\Psi})_{m,n} = R_{\Psi}(F_m, F_n)$, $\boldsymbol{\phi}_{l,k} = [\phi_{l,k}(F_0), \cdots, \phi_{l,k}(F_{2N})] \in \mathcal{C}^{(2N+1)\times 1}$, and $\mathbf{D} \in \mathcal{C}^{(2N+1)\times(2N+1)}$ is a diagonal matrix with the diagonal elements defined as follows,

$$[\mathbf{D}]_{n,n} = \begin{cases} \Delta/3, & n=0 \text{ or } n=2N, \\ 4\Delta/3, & n=1,3,\cdots,2N-1, \\ 2\Delta/3, & n=2,4,\cdots,2N-2. \end{cases} \tag{16}$$

With the representation given in (15), the eigenvalues $\{\lambda_l\}_{l=1}^{L}$ and the eigensystem order $K = \sum_{l=1}^{L} K_l$ are equal to the eigenvalues and rank of the product matrix $\mathbf{R}_{\Psi}\mathbf{D}$. The eigensystem described in (15) can be solved by standard numerical functions given the knowledge of the frequency domain auto-correlation function $R_{\Psi}(F_1, F_2)$.

For systems with fixed receive filter, the function $R_{\Psi}(F_1, F_2)$ can be expressed by (17) given at the top of next page [c.f. (10)], where $R_G(F_1, F_2) = \mathbb{E}[G(F_1)G^*(F_2)]$ is the frequency domain auto-correlation function of the impulse response of the physical channel. For a system with uncorrelated scattering (US) [17] fading, the function $R_G(F_1, F_2)$ can be calculated from

$$\begin{aligned} R_G(F_1, F_2) &= \int_0^{+\infty}\int_0^{+\infty} \mathbb{E}[g(t_1)g^*(t_2)]\, e^{-j2\pi(F_1 t_1 - F_2 t_2)} dt_1 dt_2, \\ &= \int_0^{+\infty} \varphi(t) e^{-j2\pi(F_1-F_2)t} dt, \end{aligned} \tag{18}$$

where $\varphi(t)$ is the power delay profile (PDP) of the frequency selective fading channel. From (18), the function $R_G(F_1, F_2)$ is wide sense stationary (WSS) in the frequency domain F, *i.e.*, $R_G(F_1, F_2) = R_G(F_1 - F_2)$; in addition, $R_G(F)$ can be interpreted as the FT of the PDP $\varphi(t)$.

For many wireless communication systems, the PDP can be represented in the form of a discrete-time function

$$\varphi(t) = \sum_{i=1}^{I} \varphi_i \delta(t - \tau_i), \tag{19}$$

where I is the number of resolvable multipaths of the frequency selective fading channel, φ_i and τ_i are the average power and relative delay of the ith multipath, respectively, and $\sum_{i=1}^{I} \varphi_i = 1$ for normalized PDP. The function $R_G(F)$ of such system configuration can be calculated from the Fourier transform of (19), and the result is

$$R_G(F) = \sum_{i=1}^{I} \varphi_i e^{-j2\pi F \tau_i}. \tag{20}$$

TABLE I

PARAMETERS OF THE UNIFIED ERROR PROBABILITY EXPRESSIONS

Modulation	ζ	β_1	$-\beta_2$	ψ_1	ψ_2
MPSK	$\sin^2\frac{\pi}{M}$	1	0	$\pi-\frac{\pi}{M}$	0
MASK	$\frac{3}{M^2-1}$	$2-\frac{2}{M}$	0	$\frac{\pi}{2}$	0
MQAM	$\frac{3}{2(M-1)}$	$4-\frac{4}{\sqrt{M}}$	$\left(2-\frac{2}{\sqrt{M}}\right)^2$	$\frac{\pi}{2}$	$\frac{\pi}{4}$

Another commonly used PDP is the exponentially decaying profile. The exponential PDP along with its FT $R_G(F)$ can be expressed as

$$\varphi(t) = \frac{\exp\left(-\frac{t-\tau_{\max}}{T_{sym}}\right)}{T_{sym}\left[\exp\left(\frac{\tau_{\max}}{T_{sym}}\right)-1\right]}, \quad 0 \le t \le \tau_{\max}, \tag{21a}$$

$$R_G(F) = \frac{\exp\left(\frac{\tau_{\max}}{T_{sym}}\right) - \exp(-j2\pi F\tau_{\max})}{\left[\exp\left(\frac{\tau_{\max}}{T_{sym}}\right)-1\right]\left[1+j2\pi T_{sym}F\right]}, \tag{21b}$$

where $\tau_{\max}$ is the maximum delay spread of the frequency selective fading channel.

Given transmit filter $p_T(t)$, receive filter $p_R(t)$, and the PDP $\varphi(t)$, we can formulate the frequency domain auto-correlation function $R_\Psi(F_1, F_2)$ by using (20) or (21b). Substituting the resultant function $R_\Psi(F_1, F_2)$ into the eigensystem of (14) or (15) leads to the solution of the eigenvalues λ_l, which are then used in (12) and (13) to represent the statistical properties of the SNR γ.

From (12), (14) and (17), we conclude that the statistical properties of the instantaneous SNR γ are jointly determined by the transmit filter $p_T(t)$, the receive filter $p_R(t)$, the channel power delay profile $\varphi(t)$, the sampling frequency F_s, and the sampler timing phase offset τ_0. Also, it is apparent that the frequency domain autocorrelation function $R_\Psi(F_1, F_2)$, the form of which depends on individual receiver implementations, plays a critical role in determining the properties of γ.

B. Error Performance Bound

Based on the statistical properties of the instantaneous SNR γ, theoretical performance lower bounds of systems with M-ary phase-shift-keying (MPSK), M-ary amplitude-shift-keying (MASK), and M-ary quadrature-amplitude-modulation (MQAM) are derived in this subsection.

Based on the ISI free assumption, the conditional error probability (CEP) $P(E|\gamma)$ for MPSK, MASK, and MQAM systems can be written in a unified form as [9]

$$P(E|\gamma) = \sum_{i=1}^{2} \frac{\beta_i}{\pi} \int_0^{\psi_i} \exp\left\{-\zeta \cdot \frac{\gamma}{\sin^2\theta}\right\} d\theta, \tag{22}$$

where the parameters ζ, β_i and ψ_i for various modulation schemes are listed in Table 1.

The unconditional error probability can be evaluated by averaging over the statistical distribution of the instantaneous SNR as $P(E) = \mathbb{E}[P(E|\gamma)]$. Since the CEP given in (22) is in the form of an exponential function of the instantaneous SNR γ, the expectation operation can be performed with the help of the CHF of γ as defined in (13). By combining (13)

$$R_{\Psi}(F_1,F_2)=\frac{\sum_{m=-\infty}^{+\infty}\sum_{n=-\infty}^{+\infty}R_{p_T p_R}(F_1-mF_s)\,R^*_{p_T p_R}(F_2-nF_s)\,R_G[(F_1-F_2)-(m-n)F_s]\,e^{-j2\pi\frac{(m-n)\tau_0}{T_s}}}{\sqrt{\sum_{m=-\infty}^{+\infty}\sum_{n=-\infty}^{+\infty}R_{p_R p_R}(F_1-mF_s)\,R^*_{p_R p_R}(F_2-nF_s)}}, \tag{17}$$

and (22), we have the unconditional error probability $P(E)$ as

$$P(E)=\sum_{i=1}^{2}\frac{\beta_i}{\pi}\int_0^{\psi_i}\prod_{l=1}^{L}\left(1+\frac{\zeta\gamma_0\lambda_l}{\sin^2\theta}\right)^{-K_l}d\theta. \tag{23}$$

To get the closed-form expression for $P(E)$, we perform partial fraction expansion for the integrand in (23). The result is

$$\prod_{l=1}^{L}\left(1+\frac{\zeta\gamma_0\lambda_l}{\sin^2\theta}\right)^{-K_l}=\sum_{l=1}^{L}\sum_{k=1}^{K_l}c_{l,k}\left(1+\frac{\zeta\gamma_0\lambda_l}{\sin^2\theta}\right)^{-k}, \tag{24}$$

with the partial fraction coefficient $c_{l,k}$ defined as

$$c_{l,k}=\left(\frac{\sin^2\theta}{\zeta\gamma_0}\right)^{K_l-k}\frac{1}{(K_l-k)!}\cdot\frac{\partial^{K_l-k}}{\partial\lambda_l^{K_l-k}}\left[\prod_{\substack{i=1\\i\neq l}}^{L}\left(1+\frac{\zeta\gamma_0}{\sin^2\theta}\lambda_i\right)^{-K_i}\right]\Bigg|_{\lambda_l=-\sin^2\theta/(\zeta\gamma_0)} \tag{25}$$

By substituting (24) into (23), we have the unconditional error probability $P(E)$ represented by

$$P(E)=\sum_{i=1}^{2}\frac{\beta_i}{\pi}\sum_{l=1}^{L}\sum_{k=1}^{K_l}c_{l,k}\int_0^{\psi_i}\left(1+\frac{\zeta\gamma_0\lambda_l}{\sin^2\theta}\right)^{-k}d\theta. \tag{26}$$

The integral in (26) can be solved by employing the definition of the Appell Hypergeometric function $\mathbb{F}_1(\alpha;\beta,\beta';\gamma;x,y)$ [18], and the result is given by (27) at the top of next page.

Eqn. (27) gives a unified expression of the performance lower bound for MPSK, MASK, and MQAM systems with frequency selective fading, with the values of the parameters ζ, β_i, ψ_i given in Table 1. For the special case that $K_l=1$, for $l=1,2,\cdots,L$, which is true for most practical PDPs, the solutions of the error probability lower bound can be solved without resorting to Hypergeometric functions, and the results are given in [9]. Moreover, the integral in (23) only involves elementary functions and small integration limits, thus it can be easily evaluated with numerical methods.

In (23) and (27), the effects of frequency selective fading, timing phase offset τ_0, and receiver oversampling are quantified in the error probability expressions via the eigenvalues λ_l of the eigensystem defined in (14) or (15).

IV. Case Studies

In this section, we perform case studies of various representative communication systems to further investigate the effects of timing phase offset and receiver oversampling on system performance. In the analysis, we only consider systems with at most 100% excessive bandwidth, *i.e.*, the frequency domain support of the composite impulse response $h(t)$ is in the range of $[-2/T_{sym}, 2/T_{sym}]$, and the analysis can be directly extended to systems with arbitrary amount of excessive bandwidth.

As highlighted in Section III, system error performance lower bound is uniquely determined by the eigensystem defined (14) or (15), which is in turn fully characterized by the frequency domain auto-correlation function $R_\Psi(F_1,F_2)$ as given in (17). In addition, the timing phase offset τ_0 is explicitly expressed in the representation of $R_\Psi(F_1,F_2)$. For this reason, to investigate the effects of τ_0 on system performance, it suffices to examine the properties of $R_\Psi(F_1,F_2)$ for the various representative system configurations.

A. Case 1: T_{sym}-spaced Receiver ($\mu=1$).

For a system with symbol spaced ($T_{sym}=T_s$) receiver and at most 100% excessive bandwidth, there are at most three frequency components overlapped in the frequency range of $[-\frac{1}{2T_s},\frac{1}{2T_s}]$. If the receive filter $p_R(t)$ is matched to the time-invariant transmit filter $p_T(t)$, or $P_R(F)=P_T^*(F)$, then the frequency domain auto-correlation function $R_\Psi(F_1,F_2)$ can be written by (28) given at the top of next page [c.f. (17)]. where $R_G(F_1-F_2)=\mathbb{E}[G(F_1)G^*(F_2)]$ is the FT of the channel PDP $\varphi(t)$. The function $R_\Psi(F_1,F_2)$ completely determines the statistical properties of the SNR γ through the eigensystem defined in (14).

In the representation of (9) and (28), the values and statistical properties of the instantaneous SNR γ is explicitly expressed as periodic functions of the timing phase offset τ_0, and the fundamental period is equal to the sampling period T_s. Moreover, it is apparent from (28) that the dependence of γ on τ_0 is introduced by the effect of spectrum aliasing. Since the eigenvalues λ_l and performance lower bound $P(E)$ are uniquely determined by the eigensystem characterized by the periodic function $R_\Psi(F_1,F_2)$ as described in (14), we can conclude that both λ_l and $P(E)$ are periodic with respect to τ_0. As an example, the eigenvalues and the corresponding error performance bounds of a T_{sym}-sampled system is plotted in Fig. 1. Typical Urban profile [19] as depicted in Fig. 2 is used in the simulation. Root raised cosine (RRC) filters are used as both transmit filter and receive filter. It is apparent from Fig. 1 that the values of both λ_l and $P(E)$ fluctuates periodically with respect to τ_0 with period T_s.

It is worth pointing out that the timing phase sensitivity was qualitatively discussed in [2]. However, no analytical result was available in the literature to quantitatively describe the relationship between the timing phase offset and system performance. In this paper, the timing phase offset τ_0 is explicitly expressed in the representation of the instantaneous SNR γ as described in (9) and (28), and the effects of τ_0 and oversampling factor μ are quantified in the unified error performance bound expression via the eigenvalues λ_l, which

$$P(E) = \sum_{i=1}^{2} \frac{\beta_i}{\pi} \sum_{l=1}^{L} \sum_{k=1}^{K_l} c_{l,k} \left(\zeta\gamma_0\lambda_l\right)^{-k} \mathbb{F}_1 \left[\frac{1}{2}+k; k, 1; \frac{3}{2}+k; -\left(1+\frac{1}{\zeta\gamma_0\lambda_l}\right)\tan^2\psi, -\tan^2\psi\right] \tag{27}$$

$$R_{\Psi}(F_1,F_2) = \frac{\sum_{m=-1}^{+1}\sum_{n=-1}^{+1} \left|P_T(F_1-mF_s)\,P_T(F_2-nF_s)\right|^2 R_G[(F_1-F_2)-(m-n)F_s]\, e^{-j2\pi\frac{(m-n)\tau_0}{T_s}}}{\sqrt{\sum_{m=-1}^{+1}\sum_{n=-1}^{+1} \left|P_T(F_1-mF_s)\,P_T(F_2-nF_s)\right|^2}} \tag{28}$$

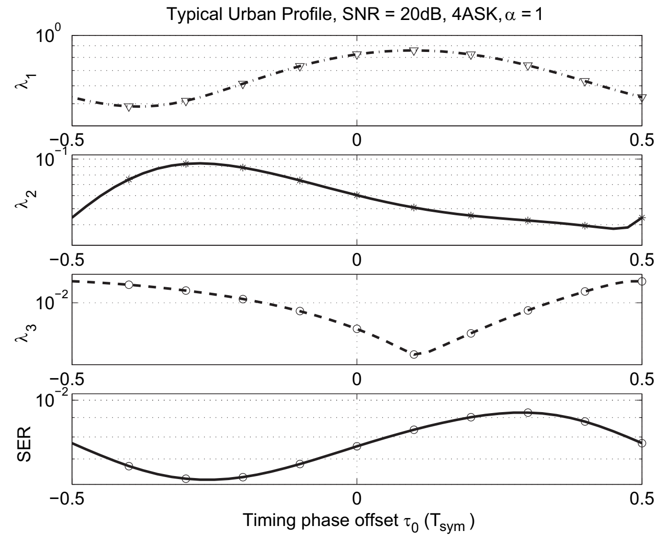

Fig. 1. The variations of eigenvalues and performance bound with respect to timing phase offset for systems with typical urban profile. α: roll-off factor of the RRC filter.

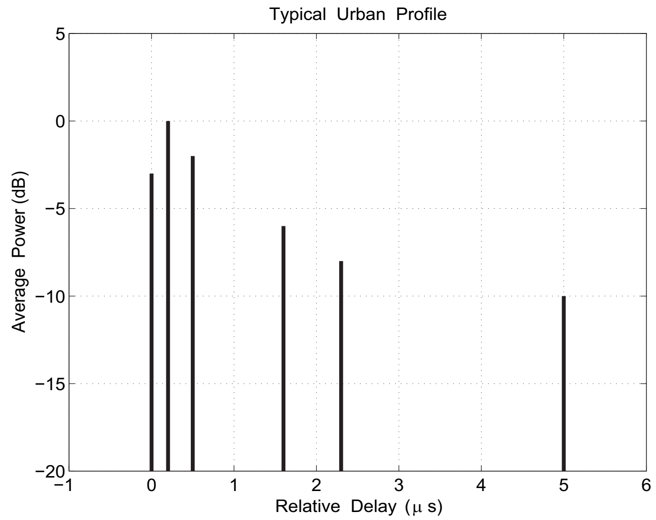

Fig. 2. Typical urban power delay profile.

clearly describes the dependence of system performance on receiver timing phase.

The above discussions are valid for systems with arbitrary PDP. For the special case that the impulse response of the frequency selective fading can be represented as a T_s-spaced tapped delay line filter, *i.e.*, $\varphi(t) = \sum_{l=0}^{L-1} \varphi_l \delta(t - lT_s)$, the representation of $R_\Psi(F)$ can be further simplified, and we have the following proposition for this case.

Proposition 1: If the channel impulse response of the frequency selective channel can be represented as a sampling-interval-spaced tapped delay line filter, then a system with receive filter matched to the joint response of the transmit filter and fading channel has the same error performance lower bound as that of the system with receive filter matched to the transmit filter only.

Proof: For the case that the channel impulse response can be written as a T_s-spaced tapped delay line filter, it can be easily shown that both the frequency impulse response $G(F)$ and the Fourier transform of the channel PDP $R_G(F)$ are periodic functions in the frequency domain with period F_s. Thus, for a system with fixed receive filter, the function $\Psi(F)$ of (10) can be simplified to

$$\Psi(F) = G(F)\frac{\sum_{n=-1}^{+1} \left|P_T(F-nF_s)\right|^2 e^{-j2n\pi\frac{\tau_0}{T_s}}}{\sqrt{\sum_{n=-1}^{+1} \left|P_T(F-nF_s)\right|^2}}, \tag{29}$$

On the other hand, if the receive filter is matched to the joint response of frequency selective fading and transmit filter, then the frequency response of the statistical filter is $P_R(F) = P_T^*(F)G^*(F)$. Replacing $P_R(F)$ into (10), and based on the fact that $G(F)$ is a periodic function of F, we get

$$\Psi(F) = |G(F)|\frac{\sum_{n=-1}^{+1} \left|P_T(F-nF_s)\right|^2 e^{-j2n\pi\frac{\tau_0}{T_s}}}{\sqrt{\sum_{n=-1}^{+1} \left|P_T(F-nF_s)\right|^2}}. \tag{30}$$

Obviously, substituting $\Psi(F)$ of (30) into (9) will lead to exactly the same SNR expression as substituting (29) into (9). Based on the fact that the statistical properties of the SNR γ fully determine the system performance lower bound as expressed by (12) and (23), the proof is complete. ■

A special case of the tapped delay line channel is flat fading, where there is only one channel tap with zero delay. For systems with flat fading, the error probability expressions given in (23) or (27) are exact because there is no ISI present at the system.

B. Case 2: $T_{sym}/2$-spaced Receiver ($\mu = 2$).

For systems with at most 100% excessive bandwidth, two times oversampling ($\mu = 2$) is enough to remove the phenomenon of spectrum aliasing at the receiver. We first consider the performance of a system with receive filter matched to the transmit filter, *i.e.*, $P_R(F) = P_T^*(F)$. The corresponding instantaneous SNR γ and frequency domain autocorrelation function $R_\Psi(F_1, F_2)$ can be written from (9) as

$$\gamma = \gamma_0 \cdot \int_{-\frac{1}{2T_s}}^{\frac{1}{2T_s}} |P_T(F)\, G(F)|^2 dF, \quad (31a)$$

$$R_\Psi(F_1, F_2) = P_T(F_1)\, P_T^*(F_2)\, R_G(F_1 - F_2). \quad (31b)$$

It can be seen from (31) that the statistical properties of SNR γ are independent of the timing phase offset τ_0 thanks to the removal of spectrum aliasing at the receiver. Since the system performance lower bound is uniquely determined by the statistical properties of SNR γ, it can be readily concluded that the performance lower bound for systems without spectrum aliasing is independent of the receiver timing phase [1]. Similar observations were obtained in [11] via simulations. We have the following proposition about the performance of the oversampled system.

Proposition 2: For a system without spectrum aliasing at the receiver, we have

1) the system error performance lower bound, which defines the best possible system performance, is independent of the sampler timing offset;

2) a system with receive filter matched to the transmit filter has the same performance lower bound as that of the system with statistical receive filter matched to the joint impulse response of the transmit filter and the frequency selective fading channel.

Proof: The frequency response of the statistical matched filter is $P_R(F) = P_T^*(F) G^*(F)$. Substituting $P_R(F)$ into (9) yields the SNR expression for oversampled systems with statistical matched filters

$$\gamma = \gamma_0 \cdot \int_{-\frac{1}{2T_s}}^{\frac{1}{2T_s}} |P_T(F)\, G(F)|^2\, dF. \quad (32)$$

The SNR expression given in (32) is exactly the same as the SNR defined in (31), which is the instantaneous SNR for oversampled systems with fixed receive filter. This completes the proof. ■

Since it is much simpler to implement a filter matched to the fixed impulse response of the transmit filter, we can always use simple time-invariant matched filter at the receiver of an oversampled system without sacrificing the best possible system performance. It's worth pointing out that similar observation was made in [20]. In this paper, we not only provide rigorous proof for the statement, but also obtained tight performance lower bounds for such systems.

For systems without spectrum aliasing at the receiver, the performance bounds derived in this paper coincide with the conventional matched filter bound previously obtained in [3]-[6]. Therefore, the conventional matched filter bounds can be viewed as special cases of the performance lower bounds derived in this paper.

[1]It should be noted that the irrelevance between performance lower bound and timing phase offset doesn't necessarily mean that synchronization is no longer needed. Actually, system performance lower bound corresponds to the best possible performance under certain channel condition and system configuration. For sub-optimum equalizers with large amount of residual interference, system performance might still be affected by timing phase offset.

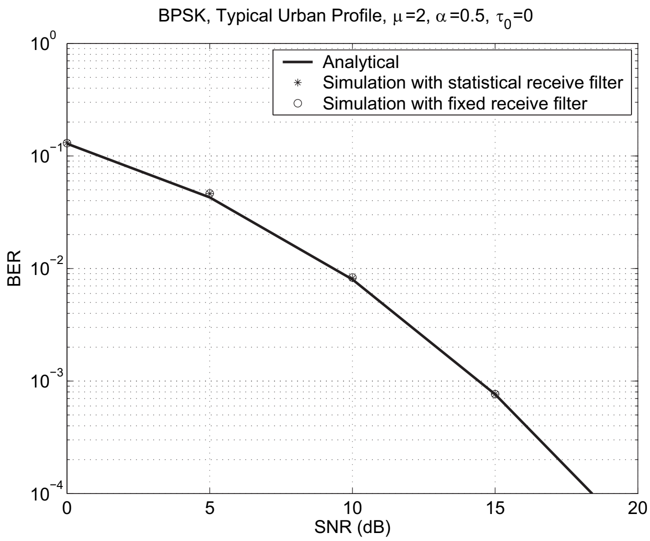

Fig. 3. Comparison of the performance of ISI-removed systems with fixed receive filter and statistical receive filter. $\mu = 2$: oversampling factor. $\tau_0 = 0$: receiver timing phase offset. $\alpha = 0.5$: roll-off factor of the RRC filter.

V. Numerical Examples

In this section, the analytical error performance expressions derived in this paper are verified with Monte-Carlo simulations, and some numerical examples are provided to reveal the effects of receiver timing phase offset on system performances.

To verify the results given in Proposition 2, we perform simulations to compare the symbol error rates of two oversampled systems equipped with fixed receive filter and statistical receive filter, respectively. In the simulation, one information symbol is sent out at each transmission epoch such that it is equivalent to the case that ISI is perfectly removed at the receiver. A RRC filter with roll-off factor $\alpha = 0.5$ (50% excessive bandwidth) is used as the transmit filter. The oversampling factor is $\mu = 2$. Fig. 3 shows the simulation results along with the corresponding theoretical error probability for a system with Typical Urban PDP. As predicted by the theoretical analysis, a perfect match is observed between the symbol error rates of the two oversampled systems with fixed filter and statistical filter.

The analytical SER performance lower bounds along with the corresponding simulation results for systems with various modulation schemes and Typical Urban PDP are shown in Fig. 4. In the simulation, MAP equalizers are employed at the receiver to mitigate the effects of ISI. It can be seen from Fig. 4 that the unified performance bound derived in this paper is very tight compared to the simulation results obtained from systems with ISI present at the receiver. This verifies the claim that the MAP equalizer is asymptotically optimum in the sense of ISI cancellation.

Fig. 5 shows the effects of timing phase offset on the performance of systems with exponentially decaying PDP. For comparison purpose, the conventional matched filter bound [4] is also plotted in the figure. Excellent agreements are observed between our theoretical expressions and simulation results for different values of timing phase offset τ_0. On the other hand, the conventional matched filter bound is significantly lower than the simulation system performance when

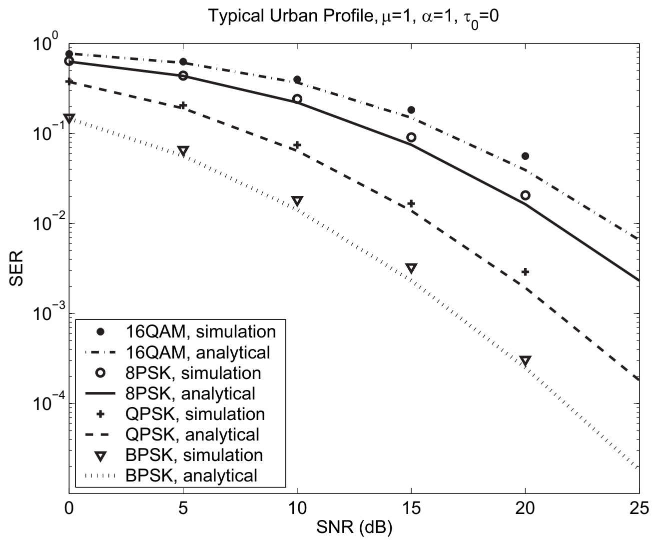

Fig. 4. Comparison of performance bounds with simulation results of systems with Typical Urban profile and various modulation schemes. Block length for the MAP equalizers: 256 symbols. $\mu = 1$: oversampling factor. $\tau_0 = 0$: receiver timing phase offset. $\alpha = 1$: roll-off factor of the RRC filter.

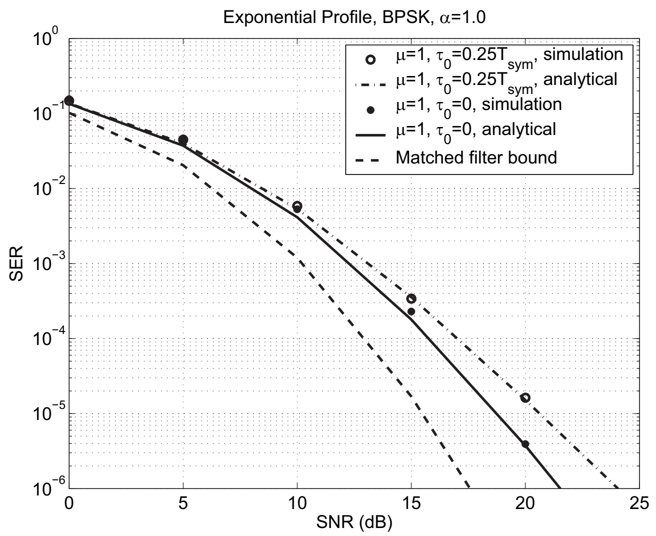

Fig. 5. Comparison of performance bounds with simulation results of systems with Exponentially decaying profile. Block length for the MAP equalizers: 1024 symbols. μ: oversampling factor. τ_0: receiver timing phase offset. $\alpha = 1$: roll-off factor of the RRC filter.

SNR $>$ 10dB. For example, at the SER level of 10^{-5}, there is a 5dB performance difference between the conventional matched filter bound and the simulation results for system with $\tau_0 = 0.25T_{sym}$. This performance gap is mainly due to the overlook of the effects of spectrum aliasing by conventional matched filter bound. For exponentially decaying PDP, no symbol spaced receiver will be able to achieve the matched filter bound due to the destructive effects of spectrum aliasing. The optimum τ_0 for the exponential PDP is further illustrated in Fig. 6, where system performance is plotted as a function of τ_0.

It can be seen from Fig. 6 that the optimum sampling time for Exponentially decaying profile is $\hat{\tau}_0 = -0.12T_{sym}$, or, equivalently, $\hat{\tau}_0 = 0.88T_{sym}$. For systems with at most 100% excessive bandwidth, two-times oversampling ($\mu = 2$) will

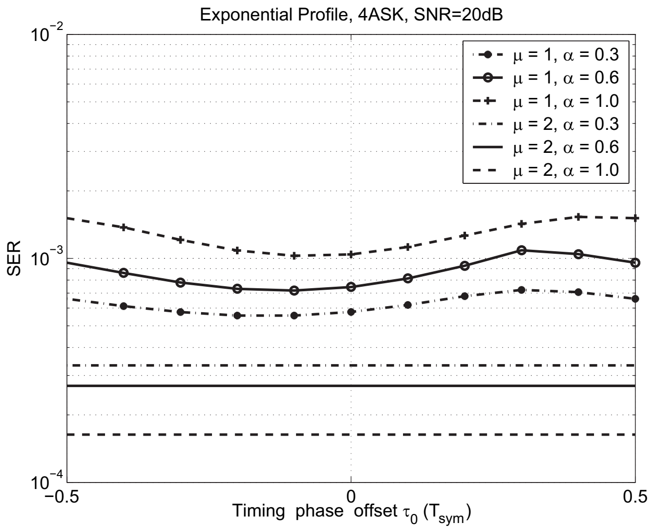

Fig. 6. The effects of receiver timing phase and excessive bandwidth on the error performance of system with Exponentially Decaying profile with $\tau_{\max} = 3T_{sym}$. μ: oversampling factor. α: roll-off factor of the RRC filter.

completely avoid spectrum aliasing in the received signals. Fig. 6 shows that the performance of systems with $\mu = 2$ and α up to 1 keeps unchanged regardless of the values of τ_0.

A close observation of Fig. 6 also reveals the effects of signal bandwidth (as represented by the roll-off factor α) on the timing sensitivity of the system performance. For systems with symbol spaced receivers, the numerical results show that the performance of systems with larger signal bandwidth (or larger value of α) is more sensitive to the timing phase offset τ_0. This phenomenon can be explained by the fact that larger excessive bandwidths will result in more spectral components being aliased. On the other hand, for systems without spectrum aliasing, the system performance improves with the increase of α, because more bandwidth is consumed during transmission.

The results presented in the previous examples show that the optimum sampling time $\hat{\tau}_0$ is a function of the power distribution of the channel profile. To investigate the relationship between the PDP and optimum receiver sampling time, we use a simple two-path equal power channel profile $\varphi(t) = 0.5\delta(t) + 0.5\delta(t - \tau_{\max})$ in this example. The SER performance lower bounds of systems with various values of τ_0 are plotted against the maximum delay spread $\tau_{\max}$ in Fig. 7. From this figure, we have the following observations: 1) For systems without spectrum aliasing ($\mu = 2$), the SER decreases monotonically with the increase of $\tau_{\max}$ when $\tau_{\max} \leq T_{sym}$, and it keeps constant after $\tau_{\max} > T_{sym}$ since no extra diversity gain can be achieved; 2) For systems with symbol spaced sampling, the SER performance fluctuates with respect to the maximum delay spread $\tau_{\max}$; 3) Systems without spectrum aliasing always outperform systems with symbol spaced receivers, thus, confirming the fact that the conventional matched filter bound is a theoretical lower bound for systems with frequency selective fading; and 4) For systems with $\tau_0 = 0$ and $\tau_{\max} = T_{sym}$, the performance of symbol spaced receiver is the same as that of the system without spectrum aliasing. For this special case, all the overlapped spectral components have the same phase and are added up constructively. Therefore, no information is lost due to spectrum aliasing.

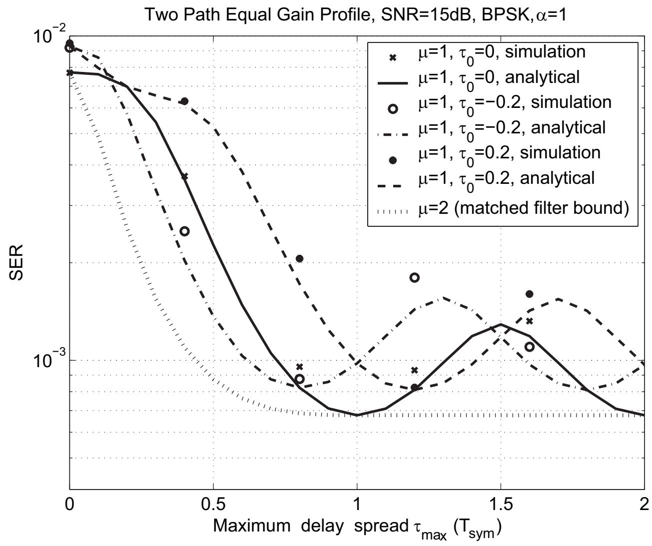

Fig. 7. The SER performance of systems with two-path equal gain profile with respect to the maximum delay spread of the channel (or the relative delay between the two channel paths). μ: oversampling factor. τ_0: receiver timing phase offset. $\alpha = 1$: roll-off factor of the RRC filter.

VI. Conclusions

The effects of timing phase offset and receiver oversampling on the performance of systems with frequency selective fading were investigated based on a tight error performance lower bound derived in this paper. The effects of timing phase offset and receiver oversampling were explicitly expressed in the statistical representation of the receiver SNR, which was further quantified in the error probability bound expressions. The conventional matched filter bound can be viewed as a special case of the performance bound derived in this paper. Simulation results showed that the new error probability bound can accurately predict the performance of practical communication systems by taking into account the effects of sampler timing phase offset and receiver oversampling.

Both theoretical analysis and numerical examples showed that for a system with spectrum aliasing, the performance lower bound, which defines the best possible system performance, is a periodic function of the receiver timing phase offset. For a system without spectrum aliasing, the performance lower bound, however, is independent of the timing phase offset. Moreover, if the effect of spectrum aliasing is completely removed at the receiver, then the choice between a fixed receive filter or statistical receive filter does not affect system performance lower bound. Finally, an interesting observation from the numerical examples is that the optimum sampling time of communication systems depends on the power distribution of the channel profiles, and that zero timing offset does not always yield the best system performance.

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