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Matched Filter Bound of Wireless Systems over Frequency Selective Channels with Receiver Timing Phase Offset

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***Abstract*— The sampler timing (phase) sensitivity of wireless communication systems is discussed in this paper. Based on the matched filter bound technique, a tight error performance bound is derived for systems experiencing frequency selective Rayleigh fading, with the receiver timing offset being quantified in the error performance expressions. With the error performance bound, the timing phase sensitivity of systems with both symbol spaced receivers and fractionally spaced receivers is analyzed. Simulation results show that the new bound can accurately predict the performance of practical communication systems suffering both frequency selective fading and timing phase offset.**

I. Introduction

The sampler timing (phase) of communication receivers is one of the essential factors defining the performance of communication systems. It is pointed out in [1] and [2] that system timing phase sensitivity is introduced by spectrum aliasing of the sampled signals at the receiver, and the relationship between receiver timing phase and system performance is qualitatively discussed in [2]. However, no analytical result is available to quantify the effects of receiver timing phase on system performance. By employing the matched filter bound technique, we are going to derive a tight theoretical performance bound that is able to quantitatively identify the effects of both receiver timing phase and oversampling.

The matched filter bound is a well known technique used to predict the performance for systems experiencing frequency selective fading [3]-[7]. Based on the assumptions that ideal matched filter is available at the receiver and there is no intersymbol interference (ISI), matched filter bound defines the best possible error performance for certain system configurations. The matched filter bounds presented in most previous works are loose performance low bounds. Moreover, conventional matched filter bounds fail to capture the effects of receiver timing phase and oversampling, which may significantly affect the performance of the communication systems.

In this paper, a tight performance low bound for systems with frequency selective fading channels is derived by considering the effects of both receiver oversampling and sampler timing phase. The ISI free assumption used by matched filter bound is adopted in the development of this new bound. With the help of Karhunen-Loève expansion, a unified error probability expression is derived as a tight low bound for various linearly modulated communication systems. The effects of timing phase offset and receiver oversampling are explicitly expressed in the statistical representations of the instantaneous signal to noise ratio (SNR), which is further quantified in the analytical error probability expressions. Simulation results show that the new performance bound can accurately predict the performance of practical communication systems in a wide range of SNR. Moreover, some useful observations of system timing phase sensitivity are obtained via theoretical analysis and numerical examples.

The rest of the paper is organized as follows. Section II presents the system model. In Section III, a tight error performance bound for systems experiencing timing phase offset is derived, and receiver timing phase sensitivity is analyzed based on this new performance bound. Numerical examples are provided in Section IV, and Section V draws the conclusion.

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II. System Model

To adopt the ISI free assumption used by the matched filter bound, we consider a communication system with the information symbol transmitted in isolation. Let $p_T(t)$ and $p_M(t)$ be the impulse response of the transmit filter and receive filter, respectively, then the baseband representation of the signal at the output of the receive filter can be represented by

$$r(t) = x_0 \cdot p_T(t) \otimes g(t) \otimes p_M(t) + z(t), \tag{1}$$

where $\otimes$ denotes the operation of convolution, x_0 is the information symbol with symbol period T_{sym}, $g(t)$ is the impulse response of the frequency selective fading, $z(t) = v(t) \otimes p_M(t)$ is the noise components, and $v(t)$ is the additive white Gaussian noise (AWGN) with variance N_0. The channel is assumed to be varying slowly enough such that the statistical channel impulse response keeps unchanged in the duration of the transmit filter.

If we define the composite impulse response (CIR) $h(t)$ of the frequency selective channel as

$$h(t) = p_T(t) \otimes g(t) \otimes p_M(t), \tag{2}$$

then the sampled output of the receive filter at sampling instant $t = kT_s + \tau_0$ can be expressed by

$$y(k) = x_0 \cdot h(k) + z(k), \tag{3}$$

where $y(k) = y(kT_s + \tau_0)$, $z(k) = z(kT_s + \tau_0)$ are the sampled version of the received signal and noise, respectively, $\tau_0 \in \left[-\frac{T_s}{2}, \frac{T_s}{2}\right]$ is the receiver timing phase offset, and $h(k) = h(kT_s + \tau_0)$ is the discrete-time version of the CIR $h(t)$. The sampling period T_s satisfies $T_s = T_{sym}/\nu$, with the integer ν being the oversampling factor of the system.

The noise sample $z(k)$ is a linear transformation of AWGN $v(t)$, hence it is zero-mean Gaussian distributed with the auto-correlation function $r_{zz}(m-n) = \mathbb{E}\left[z(m)z^*(n)\right]$ given by [9]

$$r_{zz}(m-n) = N_0 \cdot r_{p_M p_M}\left[(m-n)T_s\right], \quad (4)$$

where $r_{p_M p_M}(t)$ is the auto-correlation function of the receive filter $p_M(t)$. Due to the time span of the receive filter and the effects of oversampling, the noise samples $z(k)$ becomes a colored Gaussian process, and the power spectral density (PSD) of $z(k)$ is $\widehat{R}_{zz}(f) = N_0 \widehat{R}_{p_M p_M}(f), \quad -f_0 \le f \le f_0, \quad (5)$

where $f \in [\frac{1}{2}, \frac{1}{2}]$ is the digital frequency of discrete-time signals, $f_0 \in (0, 1/2]$ is the digital bandwidth of the receive filter, $\widehat{R}_{zz}(f)$ and $\widehat{R}_{p_M p_M}(f)$ are the discrete-time Fourier transform (DTFT) of $r_{zz}(k)$ and $r_{p_M p_M}(k)$, respectively. Let $R_{p_M p_M}(F)$ be the Fourier transform of the continuous-time auto-correlation function $r_{p_M p_M}(t)$, where $F = f/T_s$ is the analog frequency. According to the sampling theorem, $\widehat{R}_{zz}(f)$ can also be written as

$$\widehat{R}_{zz}(f) = \frac{N_0}{T_s} \sum_{n=-\infty}^{+\infty} R_{p_M p_M}\left[\frac{f-n}{T_s}\right], \quad -f_0 \le f \le f_0. \quad (6)$$

It should be noted from (6) that the statistical property of the noise component $z(k)$ is independent of the timing phase offset τ_0.

With the PSD of the noise component given in (6), the instantaneous signal to noise ratio (SNR) of the ISI free system is

$$\gamma = \gamma_0 T_s \cdot \int_{-f_0}^{f_0} \frac{|\widehat{H}(f)|^2}{\sum_{n=-\infty}^{+\infty} R_{p_M p_M}\left[\frac{f-n}{T_s}\right]} df, \quad (7)$$

where $\gamma_0 = E_s/N_0$ is the SNR without fading, and $\widehat{H}(f)$ is the DTFT of the discrete-time CIR $h(k)$. Based on (2) and the sampling theorem, $\widehat{H}(f)$ can be written by

$$\widehat{H}(f) = \frac{1}{T_s} \sum_{n=-\infty}^{+\infty} P_T\left(\frac{f-n}{T_s}\right) G\left(\frac{f-n}{T_s}\right) P_M\left(\frac{f-n}{T_s}\right) e^{j2\pi(f-n)\frac{\tau_0}{T_s}}, \quad (8)$$

where $j^2 = -1$ is the imaginary part symbol, $P_T(F)$, $P_M(F)$ and $G(F)$ are the Fourier transforms of $p_T(t)$, $p_M(t)$ 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 $\widehat{R}_{p_M p_M}(f)$ because the effect of receive filter $p_M(t)$ is included in the CIR $h(k)$.

Combining (7) and (8), the instantaneous SNR at the output of the sampler can be expressed as

$$\gamma = \gamma_0 \int_{-F_0}^{F_0} \frac{\left| \sum_{n=-\infty}^{+\infty} R_{P_T P_M}(F-nF_s) G(F-nF_s)\, e^{-j2\pi n \frac{\tau_0}{T_s}} \right|^2}{\sum_{n=-\infty}^{+\infty} R_{p_M p_M}(F-nF_s)} dF, \quad (9)$$

where $F_s = 1/T_s$ is the sampling rate, $F_0 = f_0/T_s \in (0, \frac{1}{2T_s}]$ is the analog bandwidth, $R_{P_T P_M}(F) = P_T(F)P_M(F)$, and the integration variable has been changed to the analog frequency $F = f/T_s$ in (9).

III. Performance Analysis

By analyzing the statistical properties of the SNR γ, theoretical performance low bounds for systems with M-ary phase-shift-keying (MPSK), M-ary amplitude-shift-keying (MASK), and M-ary quadrature-amplitude-modulation (MQAM) are derived, and the receiver timing phase sensitivity is analyzed.

A. Error Performance Bound

Based on the ISI free assumption, the conditional error probability (CEP) for MPSK, MASK, and MQAM systems can be written in a unified form as [8]

$$P(E|\gamma) = \sum_{i=1}^{2} \frac{\beta_i}{\pi} \int_0^{\psi_i} \exp\left\{-\zeta \cdot \frac{\gamma}{\sin^2\theta}\right\} d\theta, \quad (10)$$

where the values of the parameters ζ, β_i and ψ_i for the 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 $P(E) = \mathbb{E}\left[P(E|\gamma)\right]$. The expectation operation can be performed with the help of the characteristic function (CHF) of γ.

The CHF of γ is evaluated with the help of Karhunen-Loève expansion. If we define

$$\Psi(F) = \frac{\sum_{n=-\infty}^{+\infty} R_{P_T P_M}(F-nF_s) G(F-nF_s) e^{-j2\pi n \frac{\tau_0}{T_s}}}{\sqrt{\sum_{n=-\infty}^{+\infty} R_{p_M p_M}(F-nF_s)}}, \quad (11)$$

then the SNR γ given in (9) can be written as

$$\gamma = \gamma_0 \cdot \int_{-F_0}^{F_0} |\Psi(F)|^2 dF. \quad (12)$$

Performing Karhunen-Loève expansion of $\Psi(F)$, we have

$$\Psi(F) = \sum_{l=1}^{L} \sqrt{\lambda_l} \sum_{k=1}^{K_l} c_{l,k} \phi_{l,k}(F), \quad (13)$$

where $\{c_{l,k}\}$ 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, $\{\phi_{l,k}(f)\}$ are the corresponding orthonormal functions with support $[-F_0, F_0]$.

Table 1. CEP Parameters of (10) for Various Modulation Schemes.

Modulation	ζ	β_1	β_2	ψ_1	ψ_2
MPSK	$\sin^2\frac{\pi}{M}$	1	0	$\pi - \frac{\pi}{M}$	0
MASK	$\frac{3}{M^2-1}$	$2(1-\frac{1}{M})$	0	$\frac{\pi}{2}$	0
MQAM	$\frac{3}{2(M-1)}$	$4\left(1-\frac{1}{\sqrt{M}}\right)$	$4\left(1-\frac{1}{\sqrt{M}}\right)^2$	$\frac{\pi}{2}$	$\frac{\pi}{4}$

According to the orthonormality of the functions $\phi_{l,k}(f)$ and (13), the eigenvalues λ_l can be solved with the following eigensystem representation,

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

where $R_\Psi(F_1, F_2) = \mathbb{E}[\Psi(f_1)\Psi^*(f_2)]$ is the frequency domain auto-correlation function of $\Psi(F)$. We are going to derive $R_\Psi(F_1, F_2)$ for both symbol spaced receivers and fractionally spaced receivers in the next subsection.

Substituting (13) into (12), we will have

$$\gamma = \gamma_0 \cdot \sum_{l=1}^{L} \lambda_l \sum_{k=1}^{K_l} |c_{l,k}|^2, \quad (15)$$

where the orthonormality of the functions $\{\phi_{l,k}(F)\}$ are used in the derivation of (15). The value of λ_l can be obtained by solving the eigensystem defined by (14). From (15), the instantaneous SNR γ is the summation of squares of independent zero-mean Gaussian variables, and the CHF of γ is [10]

$$\Phi_\gamma(\omega) = \prod_{l=1}^{L}(1 - j\omega\lambda_l\gamma_0)^{-K_l}. \quad (16)$$

Combining (10) and (16), we have the unconditional error probability $P(E)$ as

$$P(E) = \sum_{i=1}^{2} \frac{\beta_i}{\pi} \int_0^{\psi_i} \prod_{l=1}^{L} \left(1 + \frac{\zeta\gamma_0\lambda_l}{\sin^2\theta}\right)^{-K_l} d\theta. \quad (17)$$

Eqn. (17) gives a unified expression of a performance low bound for MPSK, MASK, and MQAM systems with frequency selective fading channels. The effects of the physical channel fading, timing phase offset τ_0, and receiver oversampling factor ν are quantified in the error probability expression via the help of the eigenvalues λ_l.

B. Receiver Timing Phase Sensitivity

The timing phase sensitivity of systems with frequency selective fading is discussed in this subsection. Without loss of generality, here we only consider systems with at most 100% excessive bandwidth, and the analysis can be directly extended to systems with arbitrary amount of bandwidth.

■ Symbol Spaced Receiver ($\nu = 1$)

For this system configuration, we are assuming that the receive filter $p_M(t)$ is matched to the time-invariant transmit filter $p_T(t)$, thus $P_M(F) = P_T^*(F)$. Based on the sampling theorem, for systems with at most 100% excessive bandwidth and symbol spaced sampling, there are at most three frequency components overlapped in the frequency range of $[-\frac{1}{2T_s}, \frac{1}{2T_s}]$, and the instantaneous SNR γ can be written by (c.f. (9))

$$\gamma = \gamma_0 \cdot \int_{-\frac{1}{2T_s}}^{\frac{1}{2T_s}} |\Psi(F)|^2 dF. \quad (18a)$$

$$\Psi(F) = \frac{|P_T(F)|^2 G(F) + \sum_{n=\pm 1} |P_T(F - nF_s)|^2 G(F - nF_s) e^{-j2n\pi\frac{\tau_0}{T_s}}}{\sqrt{|P_T(F)|^2 + \sum_{n=\pm 1} |P_T(F - nF_s)|^2}} \quad (18b)$$

The statistical distribution of the instantaneous SNR γ can be evaluated with the help of the eigensystem defined in (14), which is in turn characterized by the frequency domain auto correlation function $R_\Psi(F_1, F_2) = \mathbb{E}[\Psi(F_1)\Psi^*(F_2)]$. Based on the definition of $\Psi(f)$ given in (18b), the frequency domain auto correlation function $R_\Psi(F_1, F_2)$ for symbol spaced receivers is expressed in (19), which is at the top of the next page.

In (19), $R_G(F_1 - F_2) = \mathbb{E}[G(F_1)G^*(F_2)]$ is the frequency domain auto-correlation function of the impulse response of the physical channel fading, and it can be obtained from the Fourier transform of the channel power delay profile [5]. By substituting (19) into (14), we can solve the eigensystem of γ, and the obtained eigenvalues λ_l are used to evaluate the system performance bound described in (17).

In the representation of (18) and (19), the values and statistical properties of the instantaneous SNR γ are explicitly expressed as periodic functions of the timing phase offset τ_0, with period equal to the sampling period T_s. Moreover, it's apparent from (18) that the dependence of γ on τ_0 is introduced by the effect of spectrum aliasing. Since the eigenvalues λ_l and error probability $P(E)$ are uniquely determined by the eigensystem characterized by the periodic function $R_\Psi(F_1, F_2)$ as described by (14), we can conclude that both λ_l and $P(E)$ are also periodic with respect to τ_0. As an example, the eigenvalues and the corresponding error performance bounds of a system with symbol spaced two path equal gain channel profile is plotted in Fig. 1. Root raised cosine filters with roll-off factor $\alpha = 1.0$ are used as both transmit filter and receive filter. It's apparent from this figure that the values of both λ_l and $P(E)$ fluctuates periodically with respect to τ_0 with period being T_s. The performance fluctuation is a result of the τ_0 dependent phase differences among the overlapped spectral components of the signal samples as shown in (19). For different values of τ_0, the overlapped spectrum could add up either constructively or destructively due to the phase difference among the overlapped spectral components, and this will lead to either performance improvement or degradation.

It's worth pointing out that system timing phase sensitivity is qualitatively discussed in [2], but no analytical result was given to describe the relationship between receiver timing phase 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 (18) and (20), and the effects of τ_0 and

$$R_{\Psi}(F_1, F_2) = \frac{\sum_{m=-1}^{+1}\sum_{n=-1}^{+1} |P_T(F_1 - mF_s)\, P_T(F_2 - nF_s)|^2\, R_G[(F_1 - F_2) - (m-n)F_s]\, e^{-j2\pi\frac{(m-n)\tau_0}{T_s}}}{\sqrt{\sum_{m=-1}^{+1}\sum_{n=-1}^{+1} |P_T(F_1 - mF_s)\, P_T(F_2 - nF_s)|^2}}. \tag{19}$$

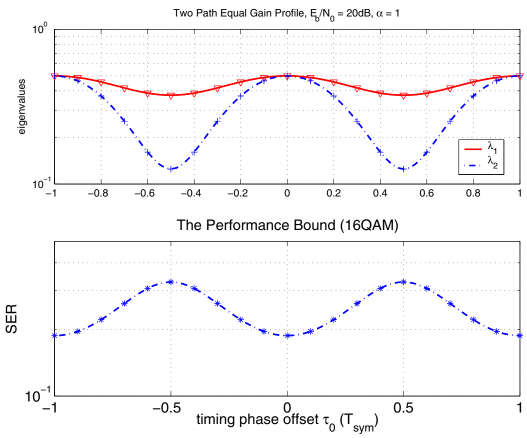

Fig. 1. The variations of eigenvalues and performance bound with respect to timing phase offset for systems with two path equal gain profile.

oversampling factor ν are quantified in the unified error performance bound expression via the eigenvalues λ_l, which clearly describes the dependence of system performance on receiver timing phase.

■ $T_{sym}/2$-spaced Receiver ($\nu = 2$)

For systems with at most 100% excessive bandwidth, two times oversampling ($\nu = 2$) is enough to avoid the phenomenon of spectrum aliasing at the receiver. For systems with receive filter matched to the transmit filter, we have $P_M(f) = P_T^*(F)$, and the instantaneous SNR γ can be simplified to

$$\gamma = \gamma_0 \cdot \int_{-\frac{1}{2T_s}}^{\frac{1}{2T_s}} |\Psi(F)|^2 dF. \tag{20a}$$

$$\Psi(F) = P_T(F)\, G(F). \tag{20b}$$

With the definition of $\Psi(F)$ given in (20b), the frequency auto-correlation function $R_{\Psi}(F_1, F_2)$ can be written as

$$R_{\Psi}(F_1, F_2) = P_T(F_1)\, P_T^*(F_2)\, R_G(F_1 - F_2), \tag{21}$$

Substituting (21) into (14) will lead to the solution of eigenvalues λ_l, for $l = 1, \cdots, L$.

It can be seen from (20) and (21) that the statistical properties of SNR γ are independent of τ_0 thanks to the the removal of spectrum aliasing at the receiver. Since the system performance is uniquely determined by the statistical properties of γ, we can conclude that the system performance for systems without spectrum aliasing is independent of receiver timing phase.

In the analysis above, we are assuming a receive filter matched to the time-invariant transmit filter $p_T(t)$. If we replace the fixed receive filter with a statistical filter matched to the joint response of channel fading and transmit filter, *i.e.*, $P_M(F) = P_T^*(F)G^*(F)$, then the SNR expression for oversampled systems could be expressed as (c.f. (9))

$$\gamma = \gamma_0 \cdot \int_{-\frac{1}{2T_s}}^{\frac{1}{2T_s}} |P_T(F)\, G(F)|^2\, dF. \tag{22}$$

It is interesting to note that the SNR expression given in (22) is exactly the same as the SNR defined in (20). Therefore we can conclude that for systems without spectrum aliasing, the choice of fixed matched filter or statistical matched filter doesn't affect the system performance. Similar observation was made in [11] by the analysis of system with fractionally-spaced receivers. 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 oversampled system without sacrificing the system performance.

It should be noted that the performance bound derived in this paper for receivers without spectrum aliasing coincides with the conventional matched filter bound previously obtained in [3]-[6], where statistical matched filter is assumed at the receiver. Thus the conventional matched filter bound can be viewed as special cases of the performance bound presented in this paper.

IV. Numerical Examples

Simulations are carried out for system with Typical Urban channel profile [12] as shown in Fig. 2. Theoretical performance bounds are compared with the corresponding simulation results in Fig. 3 to verify the accuracy of the analytical expressions derived in this paper. The simulation results are obtained from a system with maximum *a posteriori* (MAP) equalizers. It's apparent from this figure that the new performance bound is very tight compared to the simulation results for various values of τ_0. On the other hand, the conventional matched filter bound is significantly lower than the actual system performance. This performance difference is mainly contributed by the overlook of the effects of spectrum aliasing and receiver timing phase by the conventional matched filter bound. An interesting observation from Fig. 3 is that systems with $\tau_0 = 0$ doesn't yield the best error performance, and this phenomenon can be explained by the fact that the power of the Typical Urban profile is dominated by the delayed paths as shown in Fig. 2.

The effects of receiver timing phase on system performance are further illustrated in Fig. 4, where the symbol error rate (SER) are plotted against τ_0 for systems with Typical Urban profiles and RRC filters. It can be seen from this figure that

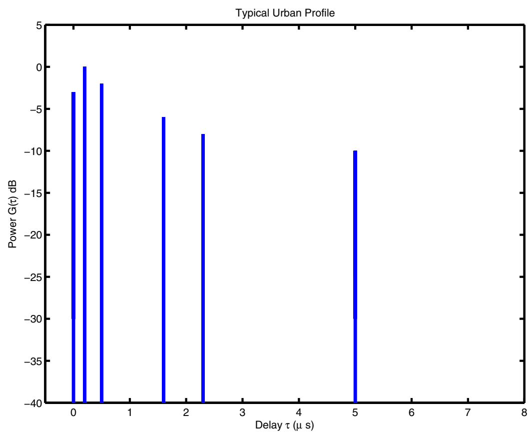

Fig. 2. The Typical Urban Power Delay Profile

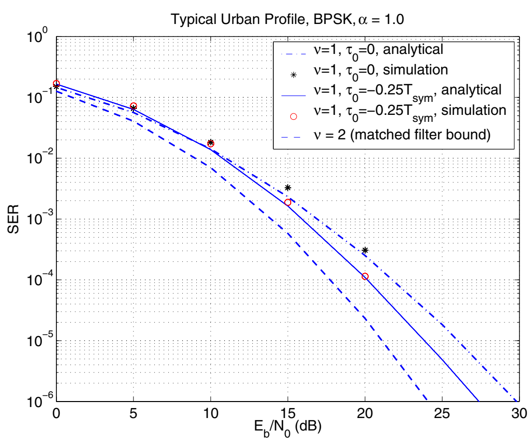

Fig. 3. Comparison of performance bounds with simulation results. Decoding length for the MAP equalizers: 1024 symbols.

the optimum sampling time for Typical Urban profile is $\hat{\tau}_0 = -0.25T_{sym}$. For systems with at most 100% excessive bandwidth ($\alpha \leq 1.0$), two-times oversampling ($\nu = 2$) will completely avoid spectrum aliasing of the received signals, and we can see from Fig. 4 that the performance of systems with $\nu = 2$ and α up to 1 keeps unchanged regardless of the values of τ_0.

The results displayed in Fig. 4 also reveals the effects of signal bandwidth (as represented by the roll-off factor α) on system timing sensitivity. For systems with symbol spaced receivers, performances of systems with larger signal bandwidth is more sensitive to the timing phase offset τ_0, due to the fact that larger excessive bandwidth will result in more spectral components being aliased. On the contrary, for systems without spectrum aliasing, the system performance improves as the increase of α, because more bandwidth is consumed in transmission.

V. Conclusions

The receiver timing phase sensitivity of systems with frequency selective channel fading was investigated based on a tight error performance low bound derived in this paper. Both theoretical analysis and numerical examples showed that for system with spectrum aliasing, the system error performance is a periodic function of the receiver timing phase, with period equal to the sampling period; for system without spectrum aliasing, system error performance is independent of the timing phase offset. An interesting observation from numerical examples is that the optimum sampling time of communication systems depends on the power distribution of channel profiles, and zero timing offset doesn't always yield the best system performance.

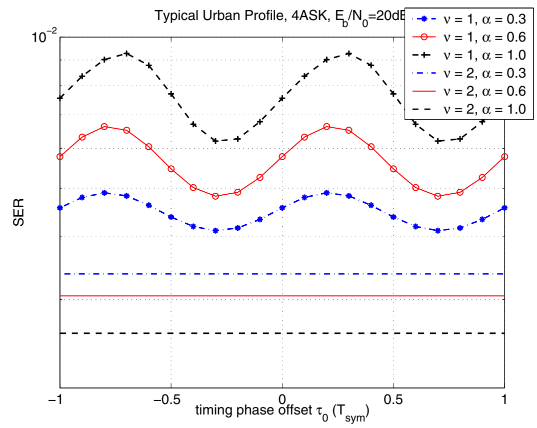

Fig. 4. The effects of receiver timing phase and excessive bandwidth on the error performance of system with Typical Urban profile.

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