EVALUATING THE EFFECTS OF CO-CHANNEL INTERFERENCE IN WIRELESS NETWORKS

Mahdi Orooji, Erfan Soltanmohammadi, Mort Naraghi-Pour

Division of Electrical and Computer Engineering
School of Electrical Engineering and Computer Science
Louisiana State University, Baton Rouge, LA 70803

ABSTRACT

The growing demand for wireless services has led to the introduction of new paradigms in spectrum sharing such as the unlicensed ISM and U-NII bands and dynamic or opportunistic spectrum access through cognitive radios. Coexistence of users in these technologies leads to increases in co-channel interference (CCI) which needs to be appropriately mitigated. CCI is often modeled as a white Gaussian noise process and assumed to simply reduce the signal-to-noise (plus interference) ratio. In this paper we consider the effect of CCI by a careful examination of the samples at the output of the matched filter receiver. We show that the timing offset between the interference and the desired signals may result in the correlation of errors in adjacent symbols. We evaluate the bit error rate (BER) resulting from CCI as well as the distribution of the total number of errors in a packet.

Index Terms— Interchannel interference, Bit error rate, Wireless communication.

1. INTRODUCTION AND SYSTEM MODEL

The ever growing popularity of wireless services coupled with the limited spectrum have resulted in an increase in the spacial reuse of the radio spectrum where many wireless services, applications or users coexist in the same frequency band. For example in cellular mobile networks frequency reuse enables the users to share the same frequency band as long as they are sufficiently apart. Another example is in the overcrowded 2.4 GHz ISM band in which WLAN, Bluetooth, wireless headsets, and cordless phones may share the same band. Recently FCC has proposed dynamic spectrum access where unlicensed users can share the spectrum with licensed users. One approach is the so-called underlay cognitive radio networks where secondary unlicensed users may coexist with the primary licensed users provided that they adapt their transmission parameters in order to limit their interference to the primary users [1].

Coexistence of users in the same frequency band results in co-channel interference which can severely degrade the performance of wireless transceivers. Co-Channel Interference (CCI) has been the subject of many studies in the literature. Effects of CCI in WLAN with multiple access points is addressed in [2] and [3]. CCI in wireless sensor networks has been studied in [4] and [5]. For cellular networks the radio link performance is usually limited by CCI rather than noise, and the outage probability due to CCI is of primary concern [6], where CCI can meaningfully degrade the performance of users especially near the border of the cells [7, 8].

Several authors have investigated the effects of interference in cognitive radios caused by the secondary users (SUs) on the primary user [9–11]. Others have proposed methods to mitigate the effects of CCI [12–15] or to exploit its effects on the SU receiver statistics in order to detect the emergence of the primary user [16, 17].

In the study of CCI, the interference signal is often modeled as a white Gaussian process. As a result it can be added to the thermal noise and accounted for by a reduction in the signal to noise ratio. However, this is not a good model owing to the fact that the interference signal is generated from a finite set of modulation symbols.

In this paper, we evaluate the effect of CCI on the bit error rate (BER), and the distribution of the number of errors in a packet by a careful examination of the samples at the output of the matched filter receiver. It is demonstrated that, due to the timing offset between the desired and the interference signals, the adjacent samples may be correlated, and that BER and the distribution of the number of errors depend on this timing offset, and identify the best and worst cases for BER. These result can be used for more accurate evaluation of CCI, to reduce CCI in cooperative networks, and for CCI mitigation as in the design of precoders or forward error correction codes.

The remainder of this paper is organized as follows. In Section 2 we evaluate the BER and the distribution of the number of errors in the presence of CCI. Simulation results are compared with those from analysis in Section 3 to demonstrate the accuracy of our modeling assumptions. Conclusions are drawn in Section 4.

2. PROBABILITY OF BIT ERROR

We consider two transmitters, U_1 and U_2, transmitting in the same frequency band. A receiver is interested in detecting the signal from U_1 and experiences interference from U_2. BPSK or QPSK modulation are assumed for U_1 whereas U_2 may employ an arbitrary M-array linear digital modulation scheme. The received signals from U_1 and U_2 are, respectively, given by

$$s_1(t) = \sqrt{\frac{E_1}{T_s}} \sum_{n=-\infty}^{\infty} a_n e^{j\omega_c t} p_1(t - nT_s), \tag{1}$$

$$\begin{aligned} s_2(t) &= \sqrt{\frac{E_2}{T_s}} \sum_{n=-\infty}^{\infty} b_n e^{j\omega_c (t+\tau) + j\zeta} p_2(t + \tau - nT_s) \\ &= \sqrt{\frac{E_2}{T_s}} \sum_{n=-\infty}^{\infty} b_n e^{j(\omega_c t + \theta)} p_1(t + \tau - nT_s), \end{aligned} \tag{2}$$

where, for $i = 1, 2$, E_i, $p_i(t)$ and $\mathcal{S}_i$ denote the energy, pulse shape and set of constellation points of U_i, respectively. Also τ and ζ are the time offset and phase offset between two received signals, and T_s denotes the symbol duration. Finally $a_n \in \mathcal{S}_1$ and $b_n \in \mathcal{S}_2$ are the transmitted symbols by U_1 and U_2, respectively, and $\theta \triangleq \omega_c \tau + \zeta$.

It is assumed that the sequences $\{a_n\}$ and $\{b_n\}$ are independent and identically distributed (iid) and are independent of each other. Furthermore, all the symbols are equally likely. Note that if $\tau \neq 0$, then two adjacent bits of U_1 receive interference from the same symbol of U_2 resulting in the correlation of the error events for the adjacent bits . The output of the receiver matched filter is given by

$$\begin{aligned} r_n =& \sqrt{E_1}\, a_n + \nu_n \\ &+ \frac{\sqrt{E_2}}{T_s} e^{j\theta} \sum_{k=-\infty}^{\infty} b_k \int_{-\infty}^{\infty} p_2(t+\tau-kT_s) p_1(t-nT_s)dt \\ =& \sqrt{E_1}\, a_n + \nu_n + \sqrt{E_2} e^{j\theta} \sum_{k=-\infty}^{\infty} b_k \psi((k-n)T_s-\tau) \end{aligned} \tag{3}$$

where $\{\nu_n\}$ is assumed to be the iid Gaussian noise process and

$$\psi(t) \triangleq \frac{1}{T} p_1(-t) * p_2(t), \tag{4}$$

where $*$ denotes convolution. Using the above notation and the fact that $p_i(t) = 0$ for $t \notin [0, T]$, (3) can be written as

$$r_n = \sqrt{E_1} a_n + \nu_n + \sqrt{E_2} e^{j\theta} \left(b_n \psi(-\tau) + b_{n+1} \psi(T_s - \tau) \right) \tag{5}$$

We first consider BPSK or the in-phase component of QPSK modulation. For the signal in (5) let $x_n = 1$ if the nth bit in the received sequence is in error, and $x_n = 0$ otherwise. Then,

$$\begin{aligned} p_{b,1} \triangleq Pr(x_n = 1) = Pr \Big(&\Re\{ \nu_n + \sqrt{E_2} e^{j\theta} (b_n \psi(-\tau) \\ &+ b_{n+1} \psi(T_s - \tau)) \} > \sqrt{E_1} \Big) \end{aligned} \tag{6}$$

where $\Re\{.\}$ denotes the real part. Here and subsequently, superscripts R and I represent the real and imaginary parts of a signal, respectively. Rewriting (6) we have,

$$\begin{aligned} p_{b,1} = Pr \Big(&\nu_n^R + \sqrt{E_2} \psi(-\tau) \left(b_n^R \cos\theta - b_n^I \sin\theta \right) \\ + \quad &\sqrt{E_2} \psi(T_s - \tau) \left(b_{n+1}^R \cos\theta - b_{n+1}^I \sin\theta \right) > \sqrt{E_1} \Big) \end{aligned} \tag{7}$$

Denote by $\mathcal{S}_2^\theta$ a new constellation obtained from a rotation of $\mathcal{S}_2$ by θ and let $\mathcal{S}_2^{(\text{eff})}$ be the set of points obtained from projection of $\mathcal{S}_2^\theta$ onto the real line. Then

$$\begin{aligned} p_{b,1} = &\frac{1}{M^2} \times \\ &\sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} Q \left(\frac{\sqrt{E_1} - \sqrt{E_2} \left(\psi(-\tau)\alpha - \psi(T_s - \tau)\beta \right)}{\sqrt{N_0/2}} \right) \end{aligned} \tag{8}$$

where M is the size of the constellation $\mathcal{S}_2$. Define $\Psi(\alpha, \beta) \triangleq \psi(-\tau)\alpha + \psi(T_s - \tau)\beta$ and assume that the constellation $\mathcal{S}_2$ has the symmetry property that if $a \in \mathcal{S}_2$, then $-a$ and $\sqrt{-1}a \in \mathcal{S}_2$[1]. Then the bit error probability can be written as,

$$\begin{aligned} p_{b,1} =& \frac{1}{2M^2} \sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \Big\{ Q\left(\sqrt{2\gamma_1} \sqrt{2\gamma_2} \Psi(\alpha, \beta) \right) \\ &+ Q\left(\sqrt{2\gamma_1} - \sqrt{2\gamma_2} \Psi(\alpha, \beta) \right) \Big\} \end{aligned} \tag{9}$$

[1]Note that all practical constellations such as MPSK and QAM satisfy this property.

where $\gamma_i \triangleq E_i / N_0$, $i = 1, 2$, is the signal to noise ratio. Using Taylor's expansion of the two terms in (9) around $2\gamma_1$ we get

$$\begin{aligned} p_{b,1} = &Q(\sqrt{2\gamma_1}) + Q^{(2)}(\sqrt{2\gamma_1}) \frac{2\gamma_2}{2! M^2} \sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \Psi^2(\alpha, \beta) \\ &+ Q^{(4)}(\sqrt{2\gamma_1}) \frac{4\gamma_2^2}{4! M^2} \sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \Psi^4(\alpha, \beta) + \cdots \end{aligned} \tag{10}$$

where $Q^{(n)}(.)$ denotes the nth derivative of $Q(.)$. For small γ_2, we can approximate Taylor's expansion of $p_{b,1}$ by its first three terms. Using Lemmas 1 and 2, (10) is approximated by,

$$\begin{aligned} p_{b,1} \approx &Q(\sqrt{2\gamma_1}) + \sqrt{\frac{\gamma_1}{\pi}}\, e^{-\gamma_1} \frac{\gamma_2}{M^2} \\ &\times \sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \left(\alpha^2 \psi^2(-\tau) + \beta^2 \psi^2(T_s - \tau) \right) \\ &= Q(\sqrt{2\gamma_1}) + \sqrt{\frac{\gamma_1}{\pi}} e^{-\gamma_1} \frac{\bar{\gamma}_2}{2} \left(\psi^2(-\tau) + \psi^2(T_s - \tau) \right) \end{aligned} \tag{11}$$

where $\bar{\gamma}_2$ is the average signal to noise ratio of U_2 given by,

$$\bar{\gamma}_2 \triangleq \frac{\gamma_2}{M} \sum_{a \in \mathcal{S}_2} |a|^2, \tag{12}$$

One would note that the first term in (11) is the effect of noise and the second term is due to the CCI from U_2. In addition $p_{b,1}$ is independent of θ. From this it follows that for QPSK modulation the errors in the in-phase and quadrature components are iid.

Lemma 1. *For the constellation $\mathcal{S}_2$ with the symmetry property that if $a \in \mathcal{S}_u$, then $-a$ and $\sqrt{-1}a \in \mathcal{S}_2$, the following equality holds.*

$$\sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \Psi^2(\alpha, \beta) = M \left(\psi^2(-\tau) + \psi^2(T_s - \tau) \right) \sum_{\alpha \in \mathcal{S}_2} |\alpha|^2 \tag{13}$$

Proof. Regardless of the value of θ, $\alpha \in \mathcal{S}_2^{(\text{eff})}$ implies that $-\alpha \in \mathcal{S}_2^{(\text{eff})}$. Thus it can be shown that

$$\sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \left(\alpha\psi(-\tau) + \beta\psi(T_s - \tau) \right) = 0 \tag{14}$$

which implies that,

$$\begin{aligned} &\sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \Psi^2(\alpha, \beta) \\ &= \sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \left(\alpha^2 \psi^2(-\tau) + \beta^2 \psi^2(T_s - \tau) \right) \\ &= M \left(\psi^2(-\tau) + \psi^2(T_s - \tau) \right) \sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \alpha^2 \end{aligned} \tag{15}$$

$\square$

Lemma 2. *For the constellation $\mathcal{S}_2$ which has the symmetry property in Lemma 1, we have*

$$\sum_{\alpha \in \mathcal{S}_2^{(\mathit{eff})}} \alpha^2 = \sum_{\alpha \in \mathcal{S}_2} |\alpha|^2 \tag{16}$$

Proof. Proof is omitted. □

In the case of non-zero time offsets, the events of two consecutive errors are dependent. Define $P_{2b,1}$ as the probability of two consecutive bits being in error. Then

$$\begin{aligned} p_{2b,1} &\triangleq P(x_n = 1, x_{n+1} = 1) \qquad (17) \\ &= \frac{1}{2}\left[P(x_n = 1, x_{n+1} = 1 | a_n = 1, a_{n+1} = 1)\right. \\ &+ \quad \left. P(x_n = 1, x_{n+1} = 1 | a_n = 1, a_{n+1} = -1)\right], \end{aligned}$$

Note that

$$\begin{aligned} &P(x_n = 1, x_{n+1} = 1 | a_n = i, a_{n+1} = j) \qquad (18) \\ &= P(x_n = 1, x_{n+1} = 1 | a_n = -i, a_{n+1} = -j) \end{aligned}$$

The first term of (17) is evaluated in (19)-(21). From (19) to (20) we use the independence of x_n and x_{n+1} conditioned on b_{n+1}. To simplify (21), we approximate it by substituting the first three terms of Taylor's expansion around $\sqrt{2\gamma_1}$ for each Q-function in (21). Using the same approach for the second term of (17) and after some manipulations, one can show that,

$$\begin{aligned} &P(x_n = 1, x_{n+1} = 1 | a_n = 1, a_{n+1} = \pm 1) \qquad (22) \\ &\approx Q^2(\sqrt{2\gamma_1}) + \frac{2\gamma_2}{M^3} Q(\sqrt{2\gamma_1}) Q^{(2)}(\sqrt{2\gamma_1}) \\ &\times \sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \sum_{\delta \in \mathcal{S}_2^{(\text{eff})}} \left(\frac{\Psi^2(\alpha, \beta) + \Psi^2(\beta, \delta)}{2} \right) \\ &\pm \frac{2\gamma_2 (Q^{(1)}(\sqrt{2\gamma_1}))^2}{M^3} \sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \sum_{\delta \in \mathcal{S}_2^{(\text{eff})}} \Psi(\alpha, \beta) \Psi(\beta, \delta) \end{aligned}$$

Therefore $p_{2b,1}$ is given by,

$$\begin{aligned} p_{2b,1} &\approx Q^2(\sqrt{2\gamma_1}) \\ &+ Q(\sqrt{2\gamma_1}) \sqrt{\frac{\gamma_1}{\pi}} e^{-\gamma_1} \frac{2\gamma_2}{M^2} \sum_{\alpha \in \mathcal{S}_2^{(\text{eff})}} \sum_{\beta \in \mathcal{S}_2^{(\text{eff})}} \Psi^2(\alpha, \beta). \quad (23) \end{aligned}$$

Using Lemma 1, Equations (11)-(15), and after some manipulations (23) is given by,

$$p_{2b,1} \approx Q(\sqrt{2\gamma_1}) \left(2p_{b,1} - Q(\sqrt{2\gamma_1}) \right) \qquad (24)$$

Consider a packet transmission system in which users U_1 and U_2 transmit their messages using packets of length N bits. Let $e = \sum_{n=1}^{N} x_n$ denote the number of errors in a received packet of U_1. We would like to find the distribution of e, namely $P(e = \ell)$. Since adjacent errors are dependent, $\{x_n\}$ is not a Bernoulli sequence and thus the distribution of e is not binomial. However, e is the sum of identically distributed random variables which are weekly dependent [18]. More specifically, $\{x_n\}$ is strongly mixing in that x_n and x_m are independent if $|m-n| > 1$. Thus using the central limit theorem for strongly mixing sequences, [18], we conclude that e converges in distribution to a Gaussian distribution $\mathcal{N}(m_1, \sigma_1^2)$, where $m_1 = \sum_{n=1}^{N} \mathbb{E}x_n = Np_{b,1}$, and $\sigma_1^2 = \sum_{i=1}^{N} \sum_{j=1}^{N} \text{Cov}(x_i, x_j)$. It follows that,

$$\sigma_1^2 = N(p_{b,1} - p_{b,1}^2) + 2(N - \mathcal{M}) \left(p_{2b,1} - p_{b,1}^2 \right) \qquad (25)$$

where $\mathcal{M}$ is the number of bits per transmitted symbol of U_1.

2.1. The Worst Case

In section 2 we calculate the probability of bit error and the distribution of the total number of errors in a packet. These quantities depend on the timing offset τ. In particular the average number of errors per packet depends on τ. To determine the worst case for the average number of errors we set $\partial p_{b,1}/\partial\tau = 0$ which results in

$$\frac{\partial}{\partial\tau} \left(\psi^2(-\tau) + \psi^2(T_s - \tau) \right) = 0. \qquad (26)$$

The result depends on the pulse shapes of U_1 and U_2. For the case that they use the same pulse shape, one can verify that

$$\psi(-t) = \psi(T_s - t) \qquad (27)$$

$$\frac{\partial}{\partial t}\psi(-t) = -\frac{\partial}{\partial t}\psi(t). \qquad (28)$$

Using the above, it can be shown that $\psi^2(-\tau)+\psi^2(T_s-\tau)$ is convex for $0 \leq \tau \leq T_s$. Therefore its maximum occurs on the boundaries and its values on $\tau = 0$ and $\tau = T_s$ are equal. Consequently the average number of errors per packet is maximized when U_1 and U_2 are synchronized.

3. SIMULATION RESULTS

In this section we validate our modeling assumptions by comparing the analytical results obtained in the previous section with those from simulation. We assume users U_1 and U_2 employ QPSK and 16-QAM modulation schemes, respectively and both use rectangular pulse shapes. Fig. 1 compares the distribution of the total number of errors derived from analysis with the histogram obtained from simulations for SNR values of γ_1 = 2dB and γ_2 = −3dB and for three different values of the timing offset τ. The figures show a close match between the results from analysis and simulation. As the figure shows, the average number of errors is largest for $\tau = 0$ (maximum $p_{b,1}$) and smallest for $\tau = \frac{T_2}{2}$ (minimum $p_{b,1}$). A significant difference can be observed in the average number of errors as well as the distribution of the number of errors between the best and the worst case. This implies that in cooperative systems where the timing offsets can be adjusted, it is desirable to set $\tau = T_s/2$. On the other hand for system design in non-cooperative networks, one should consider the worst case corresponding to $\tau = 0$.

Cramer-von Mises criterion, [19, 20], provides a metric to test the goodness of fit of a distribution compared to the empirical distribution. For the distribution of the number of errors in a packet of length N this metric is given by

$$\mathcal{D}_f \triangleq \frac{1}{N+1} \sum_{n=0}^{N} [\mathfrak{F}_y(n) - F_y(n)]^2 \mathfrak{p}_y(n) \qquad (29)$$

where $\mathfrak{F}_y$, $\mathfrak{p}_y$, and F_y are the empirical cumulative distribution function (CDF), the empirical probability density function (PDF), and the CDF of number of errors from analysis in a received packet, respectively. Fig. 2 shows the value of $\mathcal{D}_f$ versus γ_2 for different values of τ when $N = 1024$, γ_1 = 3dB. Fig. 2 demonstrates that $\mathcal{D}_f$ is quite small but increases with γ_2 and decreases from $\tau = 0$ to $\tau = T_s/2$. This is due to the fact that the approximation of $p_{b,1}$ in (11) is less accurate for larger values of $p_{b,1}$.

In the following example we consider the problem of code design for user U_1. Suppose U_1 and U_2 employ QPSK and 64-QAM modulation schemes, respectively and that U_1 uses a $(1023, \mathcal{K})$ BCH code for forward error correction. We are interested in the

$$
\begin{aligned}
&P(x_n = 1, x_{n+1} = 1 | a_n = 1, a_{n+1} = 1) \\
&= P\left(\nu_n^R + \sqrt{E_2}\left(\psi(-\tau)b_n^{(\text{eff})} + \psi(T_s-\tau)b_{n+1}^{(\text{eff})}\right) > \sqrt{E_1}, \nu_{n+1}^R + \sqrt{E_2}\left(\psi(-\tau)b_{n+1}^{(\text{eff})} + \psi(T_s-\tau)b_{n+2}^{(\text{eff})}\right) > \sqrt{E_1}\right) \quad (19)\\
&= \frac{1}{M}\sum_{\beta\in\mathcal{S}_2^{(\text{eff})}}\left[P\left(\nu_n^R + \sqrt{E_2}\left(\psi(-\tau)b_n^{(\text{eff})} + \psi(T_s-\tau)\beta\right) > \sqrt{E_1} \;\, | b_{n+1}^{(\text{eff})} = \beta\right)\right. \\
&\qquad \left.\times P\left(\nu_{n+1}^R + \sqrt{E_2}\left(\psi(-\tau)\beta + \psi(T_s-\tau)b_{n+2}^{(\text{eff})}\right) > \sqrt{E_1}\,\middle|\, b_{n+1}^{(\text{eff})} = \beta\right)\right] \quad (20)\\
&= \frac{1}{2M^3}\sum_{\alpha\in\mathcal{S}_2^{(\text{eff})}}\sum_{\beta\in\mathcal{S}_2^{(\text{eff})}}\sum_{\delta\in\mathcal{S}_2^{(\text{eff})}} Q\left(\sqrt{2\gamma_1} - \sqrt{2\gamma_2}\Psi(\alpha,\beta)\right)\times Q\left(\sqrt{2\gamma_1} - \sqrt{2\gamma_2}\Psi(\beta,\delta)\right) \\
&+ \frac{1}{2M^3}\sum_{\alpha\in\mathcal{S}_2^{(\text{eff})}}\sum_{\beta\in\mathcal{S}_2^{(\text{eff})}}\sum_{\delta\in\mathcal{S}_2^{(\text{eff})}} Q\left(\sqrt{2\gamma_1} + \sqrt{2\gamma_2}\Psi(\alpha,\beta)\right)\times Q\left(\sqrt{2\gamma_1} + \sqrt{2\gamma_2}\Psi(\beta,\delta)\right) \quad (21)
\end{aligned}
$$

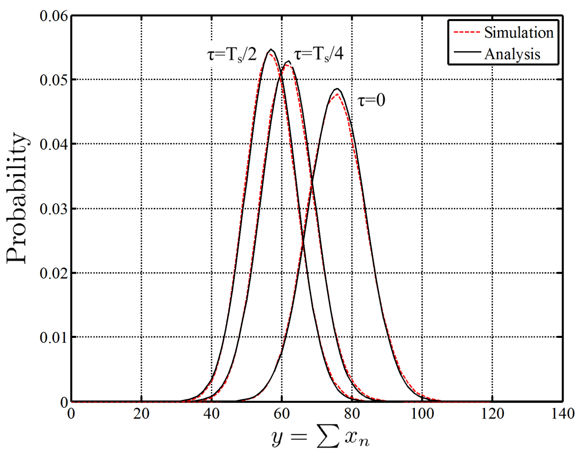

Fig. 1. Distribution of the number of errors in a packet of length $N = 1024$, when $\gamma_1 = 2$dB, $\gamma_2 = -3$dB and U_1 and U_2 use QPSK and 16-QAM, respectively.

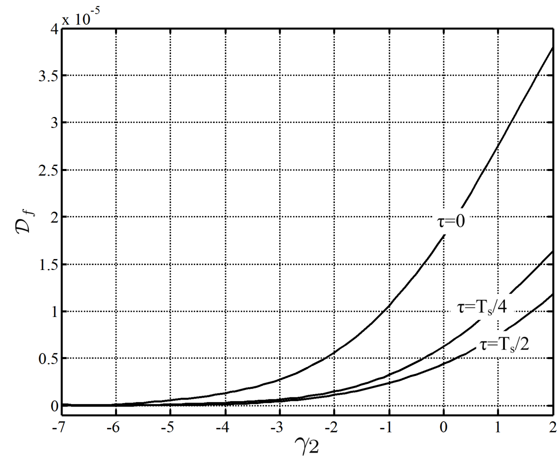

Fig. 2. Cramer-von Mises test of accuracy of the estimated distribution of the number of errors in a packet when $N = 1024$, $\gamma_1 = 3$dB and U_1 and U_2 use QPSK and 16-QAM, respectively.

largest value of $\mathcal{K}$ (highest code rate) which guarantees an average packet error rate below $\eta\%$ in the presence of CCI from U_2. TABLE 1 demonstrates these values for the parameters $\gamma_1 = 5$dB, $\gamma_2 = 0$dB and $\gamma_1 = 4$dB, $\gamma_2 = -1$dB, and the average packet error probability $\mathcal{P}_{\text{th}}$. We would like to point out the more than 20% increase in code rate from the worst case to the best case.

Table 1. Minimum $\mathcal{K}$ and required coding rate for $\gamma_1 = 5$dB, $\gamma_2 = 0$dB and $\gamma_1 = 4$dB, $\gamma_2 = -1$dB.

		$\gamma_1 = 5, \gamma_2 = 0$		$\gamma_1 = 4, \gamma_2 = -1$	
	$\mathcal{P}_{\text{th}}$	$\mathcal{K}$	Rate	$\mathcal{K}$	Rate
Worst Case	0.1%	618	0.60411	473	0.462366
	1%	648	0.63343	513	0.501466
$(\tau = 0)$	10%	708	0.69208	573	0.560117
Best Case	0.1%	738	0.72141	628	0.613881
	1%	768	0.75073	658	0.643206
$(\tau = T_s/2)$	10%	808	0.78983	708	0.692082

4. CONCLUSION

In this paper we consider the effect of co-channel interference on a desired signal by a careful examination of the samples at the output of the matched filter receiver. We show that the timing offset between the interference and the desired signals may results in the correlation of adjacent sample. We evaluate the bit error probability resulting from CCI as well as the distribution of the total number of errors in a packet. It is shown that the bit error probability is largest when the interference and the desired signals are synchronized. Our results can be employed for more accurate evaluation of CCI effects and in developing techniques for CCI mitigation such as designing precoders or forward error correction codes.

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