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Exact BER of Rectangular-Constellation Quadrature Amplitude Modulation Subjected to Asynchronous Co-Channel Interference and Nakagami-m Fading

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Abstract— **Quadrature Amplitude Modulation (QAM) is a bandwidth-efficient transmission technique. The exact average Bit Error Ratio (BER) of the maximum-minimum-distance rectangular QAM (R-QAM) constellation is studied in the context of asynchronous Co-Channel Interference (CCI) and Nakagami-m fading. A new formula is derived for the Characteristic Function (CF) of the CCI, which requires no knowledge of the CCI distribution. The numerical results obtained from our exact BER expression are verified by our simulation results and are also compared to those of the Gaussian Approximation (GA).**

I. INTRODUCTION

The family of Quadrature Amplitude Modulation (QAM) [1] schemes has found its way into virtually all recent wireless standards, including the third-generation (3G) High-Speed Downlink Packet Access (HSDPA), the 802.11 Wireless Local Area Network (WLAN) family, as well as the Digital Audio Broadcast (DAB) and Video Broadcast (DVB) systems. The maximum-minimum distance rectangular QAM (R-QAM) constellation is popular, since it achieves the best BER in uncoded Gaussian scenarios.

The Symbol Error Rate (SER) performance of R-QAM has been studied using various exact computation techniques in [2]–[6]. By contrast, the novel contribution of this paper is that we evaluate the BER performance of R-QAM, when additionally the Co-Channel Interference (CCI) is taken into account. Conventionally, the R-QAM BER has been estimated by using various approximations or bounds [2], [7]. However, using exact BER calculation is still desirable for verifying the accuracy of various approximation and bounding techniques. The exact BER expressions derived for 16-QAM and 64-QAM constellations were provided in [1]. A general recursive algorithm devised for the exact BER computation of Square QAM (S-QAM) was presented in [8], while an exact and general closed-form BER expression of R-QAM[1] was derived for arbitrary constellation sizes in [9]. Most of these results were obtained for Additive White Gaussian Noise (AWGN) channels. The AWGN result of [9] was later extended to Nakagami-m [10], [11] and Ricean [12] fading channels. A signal-space partitioning method was proposed for calculating the exact SER/BER of arbitrary two-dimensional signaling in the context of various fading channels in [13].

The exact QAM BER calculation becomes even more challenging, when the CCI is taken into account. Conventionally, the CCI is assumed to be Gaussian distributed for the sake of computational simplicity [14]. However, the Gaussian Approximation (GA) is accurate only, when we have a large number of interferers owning to the Central Limit Theorem (CLT) [15]. Moreover, to the best of the authors' knowledge, the exact and general BER expression of R-QAM systems corrupted by CCI has not been derived. In the existing literature, most exact BER analyses procedures were performed for BPSK [16]–[25] and QPSK [24]–[27] systems.

Hence, again, the contribution of this paper is that we derive an exact and general BER expression for general R-QAM systems corrupted by both asynchronous CCI and Nakagami-m fading, while dispensing with the Gaussian distributed CCI assumption. This paper is organized as follows. In Section II a general R-QAM system subject to asynchronous CCI and Nakagami-m fading is described. Its exact BER performance is investigated based on the Characteristic Function (CF) approach in Section III. Our numerical results are presented in Section IV, where we verify the accuracy of our exact BER expression and demonstrate the limited accuracy of the GA method. Finally, we conclude this paper in Section V.

II. SYSTEM MODEL

The R-QAM signal consists of two independent amplitude-modulated signals and can be expressed as [10]:

$$s(t) = d^I b^I(t)\cos(\omega_c t + \theta) + d^Q b^Q(t)\sin(\omega_c t + \theta), \quad (1)$$

where ω_c and θ are the common carrier frequency and the carrier phase shift. As illustrated in [9], $2d^I$ and $2d^Q$ are the minimum distance between signal constellation points along the in-phase and quadrature-phase axes, respectively. Note that d^I and d^Q are not necessarily equal in the general rectangular

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[1]The generic class of R-QAM contains both the specific subclass of square-shaped S-QAM constellations, as well as non-square constellations.

QAM constellation. The in-phase and quadrature-phase data signals, $b^I(t)$ and $b^Q(t)$, are given by:

$$b^I(t) = \sum_{n=-\infty}^{\infty} b_n^I p_{T_s}(t - nT_s), \quad (2)$$

$$b^Q(t) = \sum_{n=-\infty}^{\infty} b_n^Q p_{T_s}(t - nT_s), \quad (3)$$

where $\{b_n^I\}_{n=-\infty}^{\infty}$ and $\{b_n^Q\}_{n=-\infty}^{\infty}$ are the in-phase and quadrature-phase data symbols, respectively. The symbol duration is denoted as T_s and $p_T(t)$ is the rectangular pulse having a duration of T, i.e. we have

$$p_T(t) = \begin{cases} 1, t \in [0, T), \\ 0, \text{otherwise}. \end{cases} \quad (4)$$

In the M-ary R-QAM scheme, where we have $M = M^I \times M^Q$, $\log_2 M^I$ and $\log_2 M^Q$ bits are Gray encoded and mapped onto the in-phase and quadrature-phase components [1], [9], respectively. Hence, the in-phase and quadrature data symbols, b_n^I and b_n^Q, are selected from the set of $\mathcal{A}^I = \{\pm 1, \pm 3, ..., \pm(M^I - 1)\}$ and $\mathcal{A}^Q = \{\pm 1, \pm 3, ..., \pm(M^Q - 1)\}$, respectively.

We consider a general R-QAM system subjected to K asynchronous co-channel interferers. The received signal $r(t)$ subjected to fading may be written as:

$$r(t) = \sum_{k=0}^{K} h_k \{d_k^I b_k^I(t - \tau_k) \cos[\omega_c(t - \tau_k) + \theta_k + \varphi_k] \quad (5)$$
$$+ d_k^Q b_k^Q(t - \tau_k) \sin[\omega_c(t - \tau_k) + \theta_k + \varphi_k]\} + \eta(t),$$

where the fading amplitude h_k obeys the Nakagami-m distribution having parameters $\{m_k, \Omega_k\}$ [28], the fading phase φ_k is typically assumed to be uniformly distributed over $[0, 2\pi)$ [28], the time delay τ_k of the kth user is uniformly distributed over $[0, T_s)$, and the Additive White Gaussian Noise (AWGN) $\eta(t)$ has a double-sided power spectral density of $N_0/2$.

Without loss of generality, we assume that the 0th user is the desired one. In the case of coherent demodulation as well as perfect channel estimation, the in-phase and quadrature-phase decision statistics, Z^I and Z^Q, are given by:

$$Z^I = d_0^I h_0 b_{0,0}^I + \sum_{k=1}^{K} h_k \left(X_k^I \cos \Delta_k + X_k^Q \sin \Delta_k \right) + \eta^I, \quad (6)$$

$$Z^Q = d_0^Q h_0 b_{0,0}^Q + \sum_{k=1}^{K} h_k \left(X_k^Q \cos \Delta_k - X_k^I \sin \Delta_k \right) + \eta^Q, \quad (7)$$

where the phase shift difference $\Delta_k = -\omega_c(\tau_k - \tau_0) + (\theta_k - \theta_0) + (\varphi_k - \varphi_0)$ between the kth interferer and the desired user is uniformly distributed over $[0, 2\pi)$. The noise components η^I and η^Q can be shown to be zero-mean Gaussian distributed random variables, both having a variance of N_0/T_s. The random variables X_k^I and X_k^Q are defined as:

$$X_k^I = d_k^I [b_{k,-1}^I \nu_k + b_{k,0}^I (1 - \nu_k)], \quad (8)$$

$$X_k^Q = d_k^Q [b_{k,-1}^Q \nu_k + b_{k,0}^Q (1 - \nu_k)], \quad (9)$$

where $\nu_k = \tau_k / T_s$ is the time delay of the kth interferer normalized by the symbol duration.

III. BER Analysis

Let us now continue by analyzing the error probability of the in-phase component based on the CF approach. The error probability of the quadrature component may be derived in the same way.

Upon exploiting the results of [29], we have the CF of the in-phase CCI $I_k^I = h_k \left(X_k^I \cos \Delta_k + X_k^Q \sin \Delta_k \right)$ conditioned on X_k^I and X_k^Q in the following form:

$$\Phi_{I_k^I | X_k^I, X_k^Q}(\omega) = {}_1\mathbb{F}_1 \left(m_k; 1; -\frac{\Omega_k}{4m_k} \left[(X_k^I)^2 + (X_k^Q)^2 \right] \omega^2 \right), \quad (10)$$

where ${}_1\mathbb{F}_1(\alpha; \beta; x)$ is the confluent hypergeometric function [30]. Upon averaging $\Phi_{I_k^I | X_k^I, X_k^Q}(\omega)$ over the kth interferer's data symbols $b_{k,-1}^I$, $b_{k,-1}^Q$, $b_{k,0}^I$, $b_{k,0}^Q$ and the time delay τ_k, we obtain the CF of I_k^I, $\Phi_{I_k^I}(\omega)$, as follows:

$$\Phi_{I_k^I}(\omega) = \frac{1}{M_k^2} \sum_{b_{k,-1}^I, b_{k,0}^I \in \mathcal{A}_k^I} \sum_{b_{k,-1}^Q, b_{k,0}^Q \in \mathcal{A}_k^Q} \Phi_{I_k^I | \lambda_0, \lambda_1, \lambda_2}(\omega), \quad (11)$$

where the coefficients λ_0, λ_1 and λ_2 are defined as:

$$\lambda_0 = (d_k^I)^2 (b_{k,0}^I)^2 + (d_k^Q)^2 (b_{k,0}^Q)^2, \quad (12)$$

$$\lambda_1 = (d_k^I)^2 b_{k,0}^I (b_{k,-1}^I - b_{k,0}^I) + (d_k^Q)^2 b_{k,0}^Q (b_{k,-1}^Q - b_{k,0}^Q), \quad (13)$$

$$\lambda_2 = (d_k^I)^2 (b_{k,-1}^I - b_{k,0}^I)^2 + (d_k^Q)^2 (b_{k,-1}^Q - b_{k,0}^Q)^2. \quad (14)$$

The conditional CF, $\Phi_{I_k^I | \lambda_0, \lambda_1, \lambda_2}(\omega)$, may be shown to be given by Equation 15 seen at the top of the next page. When we have $M_k = 2$, i.e. the kth interferer adopts BPSK modulation and experiences Nakagami-m fading, Equation 15 reduces to Equations 8 and 9 of [25]. By contrast, when $M_k = 4$, $d_k^I = d_k^Q$ and $m = 1$, i.e. the kth interferer adopts QPSK modulation and experiences Rayleigh fading, Equation 15 reduces to Equations 17, 19 and 21 of [25].

The in-phase CCI I_k^I imposed by the different interferers is mutually independent. Upon defining the total in-phase interference plus noise term as $\xi^I = \sum_{k=1}^{K} I_k^I + \eta^I$, it transpires that both its PDF $f_{\xi^I}(x)$ and its CF $\Phi_{\xi^I}(\omega) = \Phi_{\eta^I}(\omega) \prod_{k=1}^{K} \Phi_{I_k^I}(\omega)$ are even. Hence the Cumulative Distribution Function (CDF) $\mathrm{F}_{\xi^I}(x)$ of the total in-phase interference plus noise can be shown to be:

$$\mathrm{F}_{\xi^I}(x) = \frac{1}{2} + \frac{1}{\pi} \int_0^{\infty} \frac{\sin(\omega x)}{\omega} \Phi_{\xi^I}(\omega) \mathrm{d}\omega. \quad (16)$$

Extending the AWGN result of [9] to the scenarios encountered in presence of interference plus noise, the conditional error probability of the uth bit of the in-phase component,

$$\Phi_{I_k^I|\lambda_0,\lambda_1,\lambda_2}(\omega)=\begin{cases} {}_1\mathbb{F}_1\left(m_k;1;-\frac{\Omega_k}{4m_k}\lambda_0\omega^2\right), & \lambda_2=0,\\ \frac{x}{\lambda_2}\,\mathbb{F}^{1:0;1}_{1:0;1}\left(\begin{matrix}[(m_k):1,1]:-;[(\frac{1}{2}):1];\\ [(1):1,1]\ :-;[(\frac{3}{2}):1];\end{matrix}-\frac{\Omega_k}{4m_k}\left(\lambda_0-\frac{\lambda_1^2}{\lambda_2}\right)\omega^2,-\frac{\Omega_k}{4\lambda_2 m_k}\omega^2x^2\right)\Bigg|_{\lambda_1}^{\lambda_1+\lambda_2}, & \lambda_2\neq 0,\end{cases} \tag{15}$$

where $\mathbb{F}^{A:B^{(1)};...;B^{(n)}}_{C:D^{(1)};...;D^{(n)}}\left(\begin{matrix}\left[(a):\theta^{(1)},...,\theta^{(n)}\right]:\left[(b^{(1)}):\phi^{(1)}\right];...;\left[(b^{(n)}):\phi^{(n)}\right];\\ \left[(c):\psi^{(1)},...,\psi^{(n)}\right]:\left[(d^{(1)}):\delta^{(1)}\right];...;\left[(d^{(n)}):\delta^{(n)}\right];\end{matrix}x_1,...,x_n\right)$ is the generalized Lauricella function of n variables defined as Equations 21 - 23 of [31] and $f(x)|_{x_1}^{x_2}=f(x_2)-f(x_1)$.

$P^I_{b|h_0}(u)$, can be expressed as follows when $f_{\xi^I}(x)$ is even:

$$P^I_{b|h_0}(u)=\frac{1}{M_k^I}\sum_{l=0}^{(1-2^{-u})M_k^I-1}\left\{(-1)^{\left\lfloor\frac{l2^{u-1}}{M_k^I}\right\rfloor}\right.$$
$$\left.\times\left(2^{u-1}-\left\lfloor\frac{l2^{u-1}}{M_k^I}+\frac{1}{2}\right\rfloor\right)\left[2\left(1-\mathrm{F}_{\xi^I}\left((2l+1)d_0^I h_0\right)\right)\right]\right\}, \tag{17}$$

where $\lfloor x\rfloor$ denotes the largest integer no greater than x. Upon averaging $P^I_{b|h_0}(u)$ over h_0 and applying Parseval's theorem [32], we obtain the error probability $P^I_b(u)$ of the uth bit of the in-phase component in the form of:

$$P^I_b(u)=\frac{1}{2}-\frac{2}{\pi M_k^I}\int_0^\infty\frac{\Phi_{\xi^I}(\omega)}{\omega}\sum_{l=0}^{(1-2^{-u})M_k^I-1}\left\{(-1)^{\left\lfloor\frac{l2^{u-1}}{M_k^I}\right\rfloor}\right.$$
$$\left.\times\left(2^{u-1}-\left\lfloor\frac{l2^{u-1}}{M_k^I}+\frac{1}{2}\right\rfloor\right)\Im\left\{\Phi_{h_0}\left[(2l+1)d_0^I\omega\right]\right\}\right\}\mathrm{d}\omega, \tag{18}$$

where $\Im\{\Phi_{h_0}(\omega)\}$ is the imaginary part of the CF of the desired user's fading amplitude, h_0, which is given by Table II of [32].

Following the same approach, we may derive the error probability $P^Q_b(u)$ of the uth bit of the quadrature-phase component in the form of:

$$P^Q_b(u)=\frac{1}{2}-\frac{2}{\pi M_k^Q}\int_0^\infty\frac{\Phi_{\xi^Q}(\omega)}{\omega}\sum_{l=0}^{(1-2^{-u})M_k^Q-1}\left\{(-1)^{\left\lfloor\frac{l2^{u-1}}{M_k^Q}\right\rfloor}\right.$$
$$\left.\times\left(2^{u-1}-\left\lfloor\frac{l2^{u-1}}{M_k^Q}+\frac{1}{2}\right\rfloor\right)\Im\left\{\Phi_{h_0}\left[(2l+1)d_0^Q\omega\right]\right\}\right\}\mathrm{d}\omega. \tag{19}$$

Finally, the average BER of M-ary general R-QAM can be obtained by averaging the error probabilities given by Equations 18 and 19 [9]:

$$P_b=\frac{1}{\log_2 M_0}\left(\sum_{u=1}^{\log_2 M_0^I}P^I_b(u)+\sum_{u=1}^{\log_2 M_0^Q}P^Q_b(u)\right). \tag{20}$$

When there is no interference, i.e. we have $K=0$, Equations 18 and 19 reduce to the single-user results of [10], [11]. As expected, when only BPSK or QPSK are considered, i.e. we have $d_k^I=d_k^Q$ and $M_k=2,4$ for all users, Equations 18 and 19 reduce to the results of [25].

IV. Numerical Results

In this section, we will verify the accuracy of our exact BER expression provided in Section III and demonstrate the limited accuracy of the GA method by simulations.

Since the evaluation of the effects of CCI on the QAM BER is the main objective of our analysis, we assume that the effects of noise are negligible. We assume furthermore that the minimum distances between signal points of the in-phase and quadrature-phase components are the same, i.e. we have $d_k^I=d_k^Q$, which is typical in QAM, although our analysis in Section III also applies to more general cases, where d_k^I and d_k^Q are not necessarily equal. Furthermore, the average interference power imposed by each interferer is common and they experience the same fading statistics as the desired signal, i.e. we have the same m_k value for all users, $k=0,1,...,K$. Nevertheless, our analysis attained in Section III applies to various general cases, where the average power of each interferer is different or where each user experiences different fading distributions. We define the per-bit Signal-to-Interference Ratio (SIR) as:

$$\mathrm{SIR}=\frac{1}{\log_2 M_0}\frac{\left[(d_0^I)^2+(d_0^Q)^2\right]\Omega_0}{\sum_{k=1}^K\left[(d_k^I)^2+(d_k^Q)^2\right]\Omega_k}. \tag{21}$$

Figures 1, and 2 illustrate the average BER performance versus the per-bit SIR expressed in dB in the context of Nakagami-m fading channels associated with the parameters of $m=5$ and $m=10$, respectively. We assume that the number of interferers is $K=6$ in the simulations of both figures. This is typical in the hexagonal cellular model of TDMA cellular networks, where each cell is surrounded by $K=6$ adjacent so-called first-tier interfering cells and usually only the interference from these $K=6$ adjacent cells is considered. As seen in both figures, the results calculated by our exact BER analysis and the simulation results match both for various constellation sizes and for various Nakagami-m fading parameters. On the other hand, the GA slightly over-estimates the average BER. As expected, when the per-bit SIR is high, the fading becomes less severe, i.e. the Nakagami-m parameter increases, and the number of bits/symbol is low, the GA becomes less accurate.

Although there are six adjacent first-tier interfering cells, different interferers may have different levels of influence. It is typical that there are one or two dominant interferers.

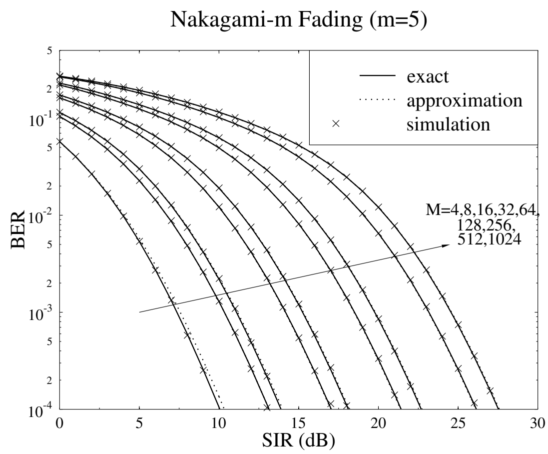

Fig. 1. BER versus the per-bit SIR in a R-QAM system subjected to asynchronous CCI and Nakagami-m fading. All users have the same constellation size, i.e. $M_k = M$. The constellation size is $M = 4$, 8, 16, 32, 64, 128, 256, 512 and 1024, respectively. The minimum distances between signal points of the in-phase and quadrature-phase components are the same, i.e. $d_k^I = d_k^Q$. The number of interferers is $K = 6$. The average power of each interferer is common and they experience the same fading distribution as the desired signal, i.e. $m_k = 5$. The background noise is ignored.

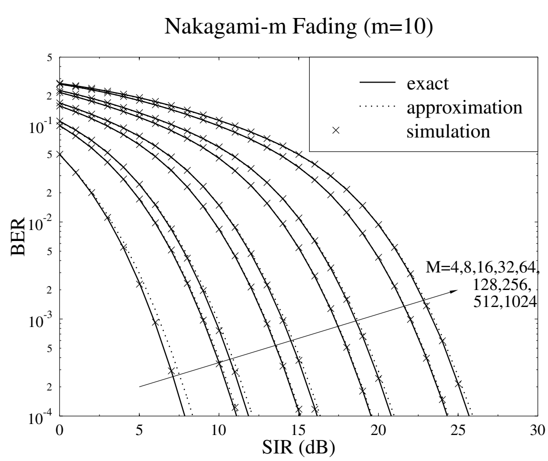

Fig. 2. BER versus the per-bit SIR in a R-QAM system subjected to asynchronous CCI and Nakagami-m fading. All users have the same constellation size, i.e. $M_k = M$. The constellation size is $M = 4$, 8, 16, 32, 64, 128, 256, 512 and 1024, respectively. The minimum distances between signal points of the in-phase and quadrature-phase components are the same, i.e. $d_k^I = d_k^Q$. The number of interferers is $K = 6$. The average power of each interferer is common and they experience the same fading distribution as the desired signal, i.e. $m_k = 10$. The background noise is ignored.

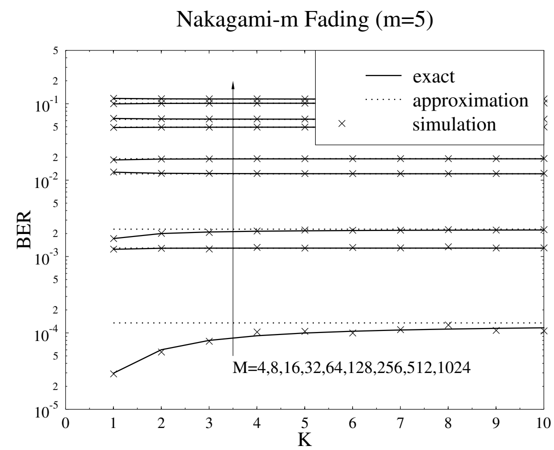

Fig. 3. BER versus the number of interferers in a R-QAM system subjected to asynchronous CCI and Nakagami-m fading. All users have the same constellation size, i.e. $M_k = M$. The constellation size is $M = 4$, 8, 16, 32, 64, 128, 256, 512 and 1024, respectively. The minimum distances between signal points of the in-phase and quadrature-phase components are the same, i.e. $d_k^I = d_k^Q$. The per-bit SIR is 10dB. The average power of each interferer is common and they experience the same fading distribution as the desired signal, i.e. $m_k = 5$. The background noise is ignored.

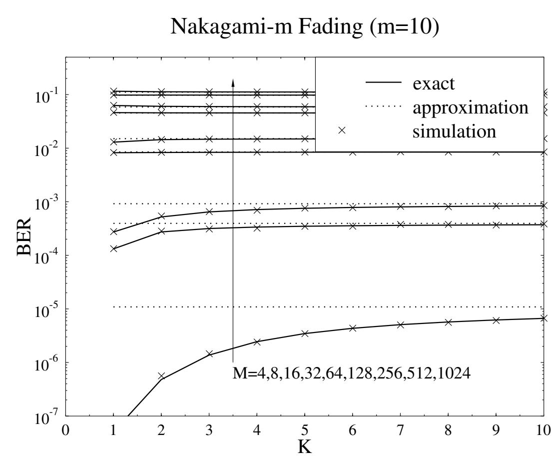

Fig. 4. BER versus the number of interferers in a R-QAM system subjected to asynchronous CCI and Nakagami-m fading. All users have the same constellation size, i.e. $M_k = M$. The constellation size is $M = 4$, 8, 16, 32, 64, 128, 256, 512 and 1024, respectively. The minimum distances between signal points of the in-phase and quadrature-phase components are the same, i.e. $d_k^I = d_k^Q$. The per-bit SIR is 10dB. The average power of each interferer is common and they experience the same fading distribution as the desired signal, i.e. $m_k = 10$. The background noise is ignored.

Figures 3 and 4 illustrate the average BER performance versus the number of interferers. For the sake of simplicity, we assume that the average power of all dominant interferers is the same and the influence of all non-dominant interferers is negligible. As we expected, the results obtained by our exact BER analysis and the simulation results match for various constellation sizes and various Nakagami-m fading parameters. On the other hand, the GA over-estimates the average BER, especially when the constellation size is small, the fading becomes less severe and the number of interferers is small.

V. Conclusion

An exact and general BER expression has been derived for general R-QAM systems subjected to asynchronous CCI and Nakagami-m fading, which requires only two single numerical integrations. A new closed-form formula was provided for the CF of the CCI with the aid of the generalized Lauricella function of n variables [31]. Our simulation results verified the accuracy of our exact BER analysis for different constellation sizes and for various channel statistics. By contrast, the Gaussian model of CCI fails to accurately predict the QAM BER performance. Our future work may consider deriving similar formulae for both dispersive channels and for CDMA systems.

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