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EXACT PERFORMANCE MEASURES FOR DIRECT-SEQUENCE CODE SYNCHRONIZATION IN FADING CHANNELS

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Abstract—This paper presents an exact performance analysis for the evaluating a mean acquisition time in a cellular code division multiple access system. New expressions for the probabilities of detection, miss, and false alarm are derived in a frequency-selective Rician fading channel. From the numerical results, it is shown that our formula gives the accurate results of the mean acquisition time in the performance analysis of the parallel acquisition system with the reference filter.

1. INTRODUCTION

Radio code-division multiple access (CDMA) already plays an important role in the current wireless personal communication system (PCS) such as cellular, microcellular, indoor, and satellite communications [1–6]. Among the cellular systems being considered, IMT-2000 and wideband CDMA (W-CDMA) are both based on direct-sequence CDMA (DS-CDMA) technology [7–10]. Code synchronization is the necessary first step in the receiver of DS-CDMA systems, since data demodulation becomes possible only after code synchronization was established. A wide variety of code acquisition methods have been proposed in various mobile radio channels, analyzed and applied in a wide range of applications, especially for DS-CDMA cellular communications. Some of popular methods are multiple-dwell serial search acquisition, rapid serial search acquisition with matched filters, rapid acquisition with parallel search, and sequential acquisition [11–17].

Among these methods, a parallel acquisition scheme simultaneously tests all possible code phases, therefore, can significantly reduce the mean acquisition time, however, at the expense of hardware complexity. An approach using a bank of N parallel I-Q noncoherent PN matched filters (PNMF) is proposed in [16, 17]. The performance of the acquisition scheme will depend on the setting of the threshold [18, 19]. In practice, a receiver has no prior knowledge about the signal-to-noise ratio (SNR) of any received PN code and it can be estimated after initial synchronization [20], thus the threshold value is often determined numerically to minimize the mean acquisition time in both the search mode and the verification mode [16, 17]. Thus, it is essential to design an acquisition threshold with its value being set according to the SNR of the received PN code. In [11, 12], the running average of the output of a reference filter is estimated and the result multiplied by a weighting factor is used as a decision threshold.

This paper derives and computes new expressions for acquisition probabilities of the DS-CDMA code acquisition system in a frequency-selective Rician fading channel. The acquisition system utilizes a bank of parallel I-Q noncoherent PNMF's similar to [16, 17]. To detect a desired threshold value, this scheme employs a reference filter [11, 12]. From the numerical analysis, it is shown that the exact evaluation of the decision threshold seems more appropriate in the performance analysis of the acquisition system with the reference filter than the approximation of the decision threshold adopted in other previous works [11, 12].

The considered parallel acquisition scheme is briefly revisited in Section 2. In Section 3, the probabilities of detection, miss, and false alarm are derived for the frequency-selective Rician fading channel. Some examples are discussed in Section 4. Finally, conclusions are presented in Section 5.

2. CODE ACQUISITION SYSTEM

Generally, code acquisition schemes have two modes of operation: search mode and verification mode. In the considered parallel acquisition system, the search mode consists of a bank of N parallel detecting I-Q passive noncoherent PNMF's and a reference I-Q PNMF. The number of taps on each delay line is M/Δ with ΔT_c delay between successive taps, where M is the MF length, Δ is 2^{-n} for some $n = 0, 1, 2, \cdots$, and T_c is the chip duration. For our analysis, a typical value for Δ is 1/2. Similar to [16], the full period of the PN code of L chips is divided into N subsequences each of length $M = N/L$. Each of the N detecting I-Q PNMF MF_D has, as a reference input, one of the subsequence of length M and the reference I-Q PNMF MF_R is loaded with a PN code orthogonal to the transmitted PN code [11, 12].

After MT_c seconds, MN/Δ samples are collected from the N parallel detecting MF's, the largest of the resulting MN/Δ samples is chosen, and the sum of M/Δ samples of the reference filter is divided by weighting factor D, where D is an integer which constitutes a fundamental design parameter. If the largest of MN/Δ samples exceeds R_D which is the sum divided by D, the corresponding phase is assumed tentatively to be coarsely aligned with the received PN code signal. There is a miss probability that none of the MN/Δ samples exceed R_D, in this case new MN/Δ samples are collected, and so on. Since one of the MN/Δ samples is the correct phase, it is therefore possible to initially acquire in MT_c seconds. Most code acquisition systems use the verification mode to confirm whether or not the tentative decision is true. However, in the CDMA reverse link, an error handling technique for slotted-mode preamble search, such as a cyclic redundancy check, is used instead of the verification mode. In this paper, we consider a single-dwell system without verification mode.

3. CODE ACQUISITION ANALYSIS IN FREQUENCY-SELECTIVE RICIAN FADING CHANNEL

The channel to be considered in this section is the frequency-selective Rician fading channel, where the model used for the fading process is the wide-sense-stationary uncorrelated scattering (WSSUS). In deriving the probability expressions, the same assumptions used in [13, 16] are adopted.

3.1. Signal Description

The DS-CDMA transmitted signal with no data modulation can be written as

$$s(t) = \mathcal{R}e\left\{\sum_{l=-\infty}^{\infty} \sqrt{2P}c(t-lT_c)e^{jw_0t}\right\} \tag{1}$$

where $\mathcal{R}e(x)$ denotes the real part of x, w_0 is the carrier frequency, P is the transmitter signal power, and $c(t)$ is a real function that presents the PN spreading code, assumed a square pulse from 0 to T_c seconds that takes values of ±1. The received signal in a general Rician fading channel is given by

$$r(t) = \mathcal{R}e\left\{\alpha\sum_{l=-\infty}^{\infty} \sqrt{2P}c(t-lT_c)e^{j(w_0t+\phi)}\right\} + F(t) + n(t) \tag{2}$$

where α is a real constant, ϕ is a uniformly distributed random phase angle, $n(t)$ represents AWGN with double-sided power spectral density $N_0/2$ and zero mean, and $F(t)$ is the faded component at the output of the considered channel. For the WSSUS channel, $F(t)$ is given by

$$F(t) = \mathcal{R}e\left\{\sqrt{2P}\beta\int_{-\infty}^{\infty} h(\tau;t)c(t-\tau)d\tau e^{j(w_0 t+\phi)}\right\} \tag{3}$$

where β is a real constant and $h(\tau;t)$ is a zero-mean complex Gaussian process. For a WSSUS channel selective only with respect to frequency, the covariance function of $h(\tau;t)$ is written as $\mathrm{E}\{h(\tau_1;t)h^*(\tau_2;t)\}/2 = \rho(\tau_1)\delta(\tau_1-\tau_2)$, where $\rho(\tau)$ is a real function of τ and $\delta(\tau)$ is the Dirac delta function. To obtain specific results, $\rho(\tau)$ used in [21], namely

$$\rho(\tau) = \begin{cases} 0, & |\tau| \geq KT_c \\ \dfrac{1}{KT_c}\left(1-\dfrac{|\tau|}{KT_c}\right), & |\tau| \leq KT_c \end{cases} \tag{4}$$

is arbitrarily chosen. This limits the inter-symbol interference (ISI) to the adjacent KT_c chips.

Referring to the parallel acquisition system in [16], the output of the I and Q branches of each I-Q PNMF is given by $e_{I/Q} = S_{I/Q} + F_{I/Q} + N_{I/Q}$, where X/Y notation denotes the corresponding term-wise pair. S_I and S_Q are outputs due to the specular component, and N_I and N_Q are independent identically distributed zero-mean Gaussian random variables with variance $\sigma_n^2 = N_0 M T_c/2$, respectively. Then, under hypothesis H_0, $S_I|H_0 = S_Q|H_0 = 0$, while under H_1, one gets $S_{I/Q}|H_1 = \alpha\sqrt{S}MT_c\cos\phi/\sin\phi$.

Under hypothesis H_1 : Since we assume that the specular component is exactly matching with one of the I-Q PNMF's, F_I from that I-Q PNMF is given by

$$F_I|H_1 = \beta\sqrt{P}\int_{t=0}^{MT_c}\int_{\tau=-\infty}^{\infty} h(\tau;t)c(t-\tau)c(t)d\tau dt. \tag{5}$$

It can be seen from [16] that F_I follows the Gaussian distribution with zero mean and variance $\sigma_F^2 = \beta^2 SM^2T_c^2 f_0/K^2$ where $f_0 = (4K-1)/6$.

Under hypothesis H_0 : F_I from the I-Q PNMF is given by

$$F_I|H_0 = \beta\sqrt{P}\int_{t=qT_c}^{(q+M)T_c}\int_{\tau=-\infty}^{\infty} h(\tau;t)c(t-\tau)c(t-qT_c)d\tau dt \tag{6}$$

where $q = k/2$ for $k = \pm1, \pm2, \pm3, \cdots$. For the considered fading channel, H_0 cells are classified as follows: 1) $2L-1-2(2K+1)$ H_0

cells for which $|q| > K + 1/2$ (H_{no} cells) are non-faded samples, which follow the Gaussian distribution with zero mean and variance σ_n^2; 2) $2(2K+1)$ H_0 cells for which $|q| = 1/2, 1, 3/2, \cdots, K+1/2$ ($H_{0,q}$ cells) are samples with a delay of q chips. These samples follow the zero-mean Gaussian distribution with variance $\sigma_{F,q}^2 = \beta^2 S M^2 T_c^2 f_q / K^2$ [16].

From the above definitions, the probability density function (PDF) of the output $R_k = e_I^2 + e_Q^2$ of the I-Q MF's under hypothesis H_1 is the noncentral χ^2 distribution, which is given by

$$f_R(y|H_1) = \frac{1}{2(\sigma_F^2 + \sigma_n^2)} \exp\left(-\frac{B^2 + y}{2(\sigma_F^2 + \sigma_n^2)}\right) I_0\left(\frac{B\sqrt{y}}{\sigma_F^2 + \sigma_n^2}\right) \tag{7}$$

while under H_0 it follows the χ^2 distribution with two degrees of freedom:

$$f_R(y|H_{0,q}) = \frac{1}{2(\sigma_{F,q}^2 + \sigma_n^2)} \exp\left(-\frac{y}{2(\sigma_{F,q}^2 + \sigma_n^2)}\right); \quad \frac{1}{2} \leq |q| \leq K + \frac{1}{2} \tag{8}$$

$$f_R(y|H_{no}) = \frac{1}{2\sigma_n^2} \exp\left(-\frac{y}{2\sigma_n^2}\right); \quad \text{otherwise} \tag{9}$$

where $B = S_I^2 + S_Q^2 = \alpha^2 M^2 T_c^2 P$ and $I_0(\cdot)$ is the modified Bessel function of first kind and zero order. On the other hand, a threshold value from the reference filter can be expressed as $R_D = D^{-1} \sum_{k=1}^{2M} R_k$. For the case of a frequency-selective Rician fading channel, it can be readily shown that under the assumption that the reference code and the transmitted code in MF_R are purely orthogonal, the variance of the output of each of the MF correlators in MF_R is given by $\sigma_R^2 = \sigma_n^2$. From this assumption, the PDF of R_D and follows the χ^2 distribution with $2M$ degrees of freedom:

$$f_{R_D}(y) = \frac{D^{2M}}{(2\sigma_n^2)^{2M} \Gamma(2M)} y^{2M-1} \exp\left(-\frac{Dy}{2\sigma_n^2}\right) \tag{10}$$

where $\Gamma(p) = (p-1)!$ for any integer $p > 0$.

3.2. Probability Expression

The detection probability of the search mode is the probability that the H_1 cell exceeds all the $2L - 4K - 3$ H_{no} cells, $4K + 2$ $H_{0,q}$ cells, and threshold R_D, which is given by

$$P_D = \int_0^\infty \frac{1}{F_0}\exp\left(-\frac{M\gamma_c}{(1+\Omega)F_0}\right)\exp\left(-\frac{y}{2F_0}\right) I_0\left(\sqrt{\frac{2M\gamma_c y}{(1+\Omega)F_0^2}}\right) \cdot \left[1-\exp\left(-\frac{y}{2}\right)\right]^{2L-4K-3} \prod_{\substack{m=-(2K+1)\\ m\neq 0}}^{2K+1}\left[1-\exp\left(-\frac{y}{2F_{m/2}}\right)\right] \cdot \left[1-\exp\left(-\frac{Dy}{2}\right)\sum_{k=0}^{2M-1}\frac{1}{k!}\left(\frac{Dy}{2}\right)^k\right] dy \tag{11}$$

where $\Omega = 2\beta^2/\alpha^2$ and $F_q = \gamma_c M f_q \Omega/K^2(1+\Omega)+1$. In Eqn. (11), $\gamma_c = \alpha^2 PT_c(1+\Omega)/N_0$ is the total received signal-to-noise ratio (SNR/chip).

To further derive Eqn. (11), the $2(2K+1)$ product term of Eqn. (11) may be rewritten in the form of

$$\prod_{\substack{m=-(2K+1)\\ m\neq 0}}^{2K+1}\left[1-\exp\left(-\frac{y}{2F_{m/2}}\right)\right] = 1+\sum_{k=1}^{2(2K+1)}(-1)^k \sum_{\substack{i_1=-(2K+1)\\ i_1\neq 0}}^{2K+1}\cdots\sum_{\substack{i_k=-(2K+1)\\ i_k\neq 0}}^{2K+1}\prod_{j=1}^{k}\exp\left\{-\frac{y}{2F_{i_j/2}}\right\} = 1+\sum_{k=1}^{2(2K+1)}(-1)^k \sum_{\substack{i_1=-(2K+1)\\ i_1\neq 0}}^{2K+1}\cdots\sum_{\substack{i_k=-(2K+1)\\ i_k\neq 0}}^{2K+1}\exp\left\{-\frac{y}{2F_k}\right\} \tag{12}$$

where $i_1 < i_2 < \cdots < i_k$ and $1/F_k = \sum_{j=1}^{k} 1/F_{i_j/2}$.

Substituting Eqn. (12) into Eqn. (11) and using a binomial theorem gives

$$P_D = \sum_{n=0}^{\mathcal{L}}(-1)^n\binom{\mathcal{L}}{n}\frac{1}{2F_0}\exp\left(-\frac{M\gamma_c}{(1+\Omega)F_0}\right)(P_{D1}-P_{D2}+P_{D3}-P_{D4}) \tag{13}$$

with

$$P_{D1} = \frac{2F_0}{1+nF_0}\exp\left\{-\frac{M\gamma_c}{(1+\Omega)(1+nF_0)F_0}\right\}, \tag{14}$$

$$P_{D2} = \exp\left(\frac{-M\gamma_c}{(1+\Omega)(1+nF_0+DF_0)F_0}\right)\sum_{k=0}^{2M-1}\left(\frac{DF_0}{1+nF_0+DF_0}\right)^{k+1} \cdot\frac{2}{D}F\left(-k,1;-\frac{M\gamma_c}{(1+\Omega)(1+nF_0+DF_0)F_0}\right), \quad (15)$$

$$P_{D3} = \sum_{k=1}^{2(2K+1)}(-1)^k \sum_{\substack{i_1=-(2K+1)\\ i_1\neq 0}}^{2K+1}\cdots\sum_{\substack{i_k=-(2K+1)\\ i_k\neq 0}}^{2K+1}\frac{2F_0F_k}{F_k+nF_0F_k+F_0} \cdot\exp\left\{-\frac{F_kM\gamma_c}{(1+\Omega)(F_k+nF_0F_k+F_0)F_0}\right\}, \quad (16)$$

and

$$P_{D4} = \sum_{l=0}^{2M-1}\sum_{k=1}^{2(2K+1)}(-1)^k \sum_{\substack{i_1=-(2K+1)\\ i_1\neq 0}}^{2K+1}\cdots\sum_{\substack{i_k=-(2K+1)\\ i_k\neq 0}}^{2K+1}\frac{2}{D} \cdot\left(\frac{DF_0F_k}{F_0+F_k(1+nF_0+DF_0)}\right)^{l+1}\exp\left(-\frac{F_kM\gamma_c(1+\Omega)^{-1}F_0^{-1}}{F_0+F_k(1+nF_0+DF_0)}\right) \cdot F\left(-l,1;-\frac{F_kM\gamma_c(1+\Omega)^{-1}F_0^{-1}}{F_0+F_k(1+nF_0+DF_0)}\right) \quad (17)$$

where $\mathcal{L} = 2L-4K-3$.

On the other hand, the miss probability of the search mode is the probability that all samples are less than R_D. This probability can be written as

$$P_M = \int_0^\infty \frac{D^{2M}}{2^{2M}\Gamma(2M)}y^{2M-1}\exp\left(-\frac{Dy}{2}\right)\left[1-\exp\left(-\frac{y}{2}\right)\right]^{\mathcal{L}} \cdot\prod_{\substack{m=-(2K+1)\\ m\neq 0}}^{2K+1}\left[1-\exp\left(-\frac{y}{2F_{m/2}}\right)\right]\left[1-Q\left(\sqrt{\frac{2M\gamma_c}{(1+\Omega)F_0}},\sqrt{\frac{y}{F_0}}\right)\right]dy \quad (18)$$

with

$$Q(a,b) = \exp\left(-\frac{a^2+b^2}{2}\right)\sum_{k=0}^{\infty}\left(\frac{a}{b}\right)^k I_k(ab) \quad (19)$$

where $I_k(\cdot)$ is the kth-order modified Bessel function of first kind. By adopting a similar approach used in deriving the detection probability,

P_M can be also divided into four terms as

$$P_M = \sum_{n=0}^{\mathcal{L}} (-1)^n \binom{\mathcal{L}}{n} (P_{M1} - P_{M2} + P_{M3} - P_{M4}) \tag{20}$$

where

$$P_{M1} = \left(\frac{D}{D+n}\right)^{2M}, \tag{21}$$

$$\begin{aligned} P_{M2} = & \left(\frac{DF_0}{1+nF_0+DF_0}\right)^{2M} \\ & \cdot \exp\left(-\frac{M\gamma_c(n+D)}{(1+\Omega)(1+nF_0+DF_0)}\right) \sum_{l=0}^{\infty} \frac{1}{l!} \left\{\frac{M\gamma_c}{(1+\Omega)F_0}\right\}^l \\ & \cdot F\left(l-2M+1, l+1; -\frac{M\gamma_c}{(1+\Omega)(1+nF_0+DF_0)F_0}\right), \end{aligned} \tag{22}$$

$$P_{M3} = \sum_{k=1}^{2(2K+1)} (-1)^k \sum_{\substack{i_1=-(2K+1) \\ i_1 \neq 0}}^{2K+1} \cdots \sum_{\substack{i_k=-(2K+1) \\ i_k \neq 0}}^{2K+1} \left(\frac{DF_k}{1+nF_k+DF_k}\right)^{2M}, \tag{23}$$

and

$$\begin{aligned} P_{M4} = & \exp\left(-\frac{M\gamma_c F_0^{-1}}{1+\Omega}\right) \sum_{l=0}^{\infty} \frac{1}{l!} \left\{\frac{M\gamma_c}{(1+\Omega)F_0}\right\}^l \sum_{k=1}^{2(2K+1)} (-1)^k \\ & \cdot \sum_{\substack{i_1=-(2K+1) \\ i_1 \neq 0}}^{2K+1} \cdots \sum_{\substack{i_k=-(2K+1) \\ i_k \neq 0}}^{2K+1} \left(\frac{DF_0F_k}{F_0+F_k(1+nF_0+DF_0)}\right)^{2M} \\ & \cdot \exp\left(-\frac{F_kM\gamma_c(1+\Omega)^{-1}F_0^{-1}}{F_0+F_k(1+nF_0+DF_0)}\right) \\ & \cdot F\left(l-2M+1, l+1; -\frac{F_kM\gamma_c(1+\Omega)^{-1}F_0^{-1}}{F_0+F_k(1+nF_0+DF_0)}\right). \end{aligned} \tag{24}$$

Finally, the false alarm probability can be obtained from P_D and P_M as $P_F = 1 - P_D - P_M$.

3.3. Mean Acquisition Time

For a single-dwell parallel acquisition scheme, the transfer function is simply obtained by

$$U(z) = \frac{P_D z^{MT_c}}{1 - \left(P_M \cdot z^{MT_c + \overline{T}_r} + P_F \cdot z^{MT_c + \overline{T}_p}\right)} \tag{25}$$

where the average penalty time for a returning false alarm $\overline{T}_p$ and the average reset time for a miss detection $\overline{T}_r$ are assumed to be same, i.e. $\overline{T}_p = \overline{T}_r = JMT_c$. Then, the mean acquisition time $\mathrm{E}\left[T_{acq}\right]$ can be calculated as

$$\begin{aligned} \mathrm{E}\left[T_{acq}\right] &= \left.\frac{dU(z)}{dz}\right|_{z=1} \\ &= \frac{\{1 + J(1 - P_D)\}MT_c}{P_D}. \end{aligned} \tag{26}$$

In this paper, we derive some probability expressions for detection, false alarm, and miss for a considered single-dwell system without verification mode. However, the results can be extended to a double-dwell system with verification mode [16][17] in a straightforward manner.

4. EXAMPLES AND DISCUSSIONS

In simply validating the effectiveness of presented analysis for the considered parallel system, we used the following parameters, namely: 1) PN code length of 1023 chips is considered, 2) matched filter lengths (M) of 93 chips are taken, which correspond to the number of parallel I-Q MF $N = 11$, 3) a penalty factor (J) of 10000 chips is used.

Figure 1 shows the mean acquisition time versus the weighting factor D according to various values of SNR/chip when $K = 50$ and $\Omega = 1.5$ are used. In this figure, the mean acquisition time has been normalized by T_c. An immediate conclusion is that optimal values of weighting factor which minimize the mean acquisition time exist for each value of SNR/chip, but for a wide range of SNR/chip the mean acquisition time is less sensitive to changes in D about the optimum. It is also shown from this figure that for properly selected value of D, the mean acquisition time is less affected by the variation of SNR/chip. In the considered fading channel, therefore, there is no need to determine a threshold value for each value of SNR/chip independently if the parallel acquisition system with the reference filter is used instead of the conventional parallel acquisition system.

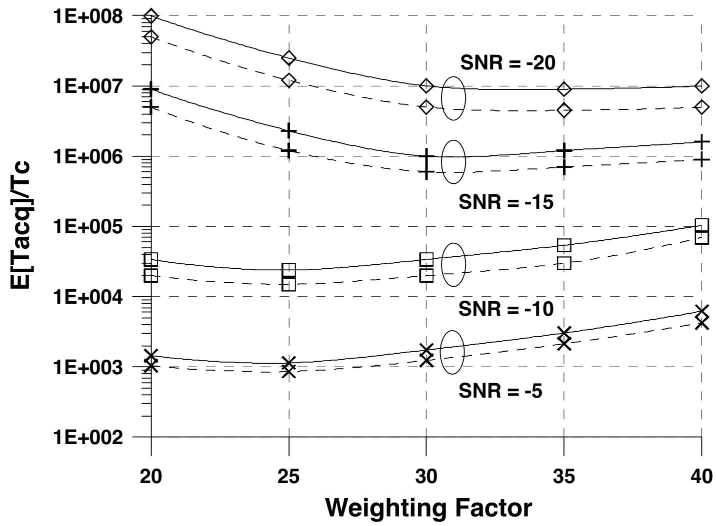

Figure 1. Mean acquisition time versus D for various values of SNR when $K = 50$ and $\Omega = 1.5$ are used: (1) Solid lines - exact mean acquisition time (2) Dash lines - approximated mean acquisition time.

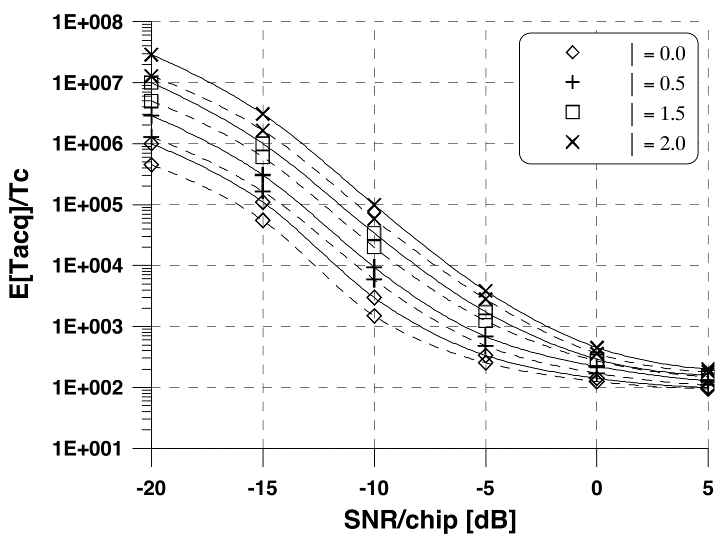

Figure 2. Mean acquisition time with Ω as a parameter when $K = 50$ and $D = 30$ are used: (1) Solid lines - exact mean acquisition time (2) Dash lines - approximated mean acquisition time.

The comparison between the approximated and the exact mean acquisition time is given in Fig. 2 for a selected value of $D = 30$. It can be seen from this figure that the approximation of the threshold value is reasonable especially for a high SNR/chip. However, there are noticeable differences in mean acquisition time performance between the exact expression when SNR/chip is low. In particular, numerical examples show that the approximation gives significantly optimistic mean acquisition time values at low SNR/chip.

5. CONCLUSION

In this paper, an exact formulation for obtaining the acquisition probabilities for the DS-CDMA parallel code acquisition system with the reference filter has been developed in a frequency-selective Rician fading channel. To evaluate the mean acquisition time in a single-dwell acquisition system without verification mode, some expressions of acquisition probabilities for the parallel acquisition system were obtained and is readily evaluated numerically. We find that our formula provides the accurate results of the mean acquisition time in the performance analysis of the parallel acquisition system with the reference filter.

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