HIGH-RATE DIRECT-SEQUENCE SPREAD SPECTRUM

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Abstract

Direct-sequence spread spectrum is popular for commercial wireless local-area networks, because license-free operation is permitted in certain frequency bands. The extensive development of equipment for such applications makes it attractive to investigate the use of the commercial technology for military communication networks. However, because of the small processing gain that is common in the signal designs for commercial direct-sequence spread spectrum, most commercial waveforms are not suitable for channels with interference, such as multiple-access channels. We investigate a class of proposed commercial high-rate spread-spectrum signal designs for which the spreading is derived primarily from nonbinary orthogonal or quasi-orthogonal data modulation. The multiple-access capability is evaluated as a function of the signal-to-interference ratio, and the effects of multipath interference are investigated.

I. Introduction

In most direct-sequence spread-spectrum systems, especially those designed for military applications, the application of a high-rate signature sequence provides the spectral spreading. The signature sequence, usually some type of pseudorandom sequence, is used to generate a continuous-time waveform that consists of a sequence of elemental pulses known as chips. The signature sequence specifies the pattern of chip amplitudes or polarities. The spreading waveform is modulated by the data sequence, and the processing gain that results is typically a linear function of the number of chips per data symbol.

There is widespread application of direct-sequence spread spectrum in commercial wireless local-area networks, because the use of spread spectrum permits license-free transmission in certain frequency bands. The extensive development of commercial equipment for such applications makes it very attractive for use by the military, if the commercial waveforms can provide the necessary protection against the interference that is encountered in military wireless communication networks. As a consequence of striving for high data rates, commercial system designers have focused on signals with a small number of chips per data symbol. Thus, the processing gains are quite small and most previous commercial technology is unsuitable for military networks. An alternative approach has been proposed (e.g., [1] and [2]) in which the spreading is derived from a nonbinary coded waveform rather than from the application of a spreading sequence. In the proposed commercial waveforms, a signature sequence is applied to each signal, but the chip rate of the signature sequence is the same as the chip rate of the coded data waveform.

This general type of waveform is not new; indeed, such a waveform is employed in the Joint Tactical Information Distribution System (JTIDS). JTIDS uses a quasi-orthogonal (i.e., approximately orthogonal) coded waveform whose chip rate is equal to the chip rate of the signature sequence (e.g., see [3] or [4]). However, JTIDS employs time-division multiple access, which requires a degree of synchronism that may not be desirable or achievable in future mobile ad hoc wireless networks. The results in Section VII provide bounds on the performance of certain orthogonal and quasi-orthogonal waveforms, including the JTIDS waveform, when they are employed in an asynchronous spread-spectrum multiple-access communication network.

Because of their relatively low processing gain, high-rate direct-sequence spread-spectrum systems are not as robust as the spread-spectrum systems normally used for military applications. If the chip rates are the same for the signature sequence and the data waveform, then the signature sequence does not spread the signal and does not provide processing gain in the classical sense. As far as multiple-access interference or multipath interference are concerned, about all that is accomplished by the signature sequence is a randomization of the data waveforms in the interference signals. However, this randomization is important, especially if CCSK is used on a multipath channel, because the orthogonal and quasi-orthogonal data waveforms do not have good correlation properties. In this paper, the multiple-access capability of high-rate direct-sequence spread spectrum is evaluated for a range of signal-to-interference ratios, and the effects of multipath interference are investigated as a function of the relative strengths of the multipath components.

II. High-Rate Direct-Sequence Waveforms

For the proposed signal designs, a set of data waveforms is used to convey the information and spread the bandwidth. Each waveform in the set is a sequence of chips, and each chip has amplitude -1 or +1. The waveforms may be orthogonal, but that is not a requirement. However, for the examples considered in this paper, the data waveforms are at least approximately orthogonal.

We let M denote the number of data waveforms, and we restrict attention to $M = 2^m$ for some integer m. Each data sequence of m bits is represented by a unique binary vector of

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length L, which is referred to as the *data vector*. The data vector that corresponds to the ith data sequence for $0 \leq i \leq M-1$ is denoted by $\mathbf{x}_i = (x_{i,0}, x_{i,1}, \ldots, x_{i,L-1})$. For each choice of n in the range $0 \leq n \leq L-1$, $x_{i,n}$ is either -1 or +1. The elements $x_{i,n}$ are referred to as the chips of the data vector or its corresponding data waveform. Since each data vector of length L represents m bits, the spreading factor for the data modulation alone is L/m. For orthogonal waveforms, $L \geq M$, so the minimum spreading factor is $M/m = M/\log_2(M)$. For a channel in which the only disturbance is thermal noise, referred to as an additive white Gaussian noise (AWGN) channel, it is well known that M-ary orthogonal waveforms and maximum-likelihood demodulation give arbitrarily small bit error probabilities as $M \to \infty$ if the energy-per-bit to noise-density ratio exceeds -1.6 dB.

A signature sequence is applied to the data waveforms with η chips of the spreading sequence per chip of the data waveform. For a fixed bandwidth, the data rate increases as η decreases. Consequently, if η is small we refer to the modulation scheme as *high-rate direct-sequence spread spectrum*. The maximum data rate is achieved for $\eta = 1$ and $L = M$, in which case the bandwidth is not increased by the application of the signature sequence and the data rate is $\log_2(M)/M$ bits per chip. The application of the signature sequence does not affect the system's performance for the AWGN channel.

If the signature sequence is $\mathbf{a} = \ldots, a_{-1}, a_0, a_1, \ldots$, the ith data sequence is represented by the baseband direct-sequence spread-spectrum signal

$$s_i(t) = \sum_{n=0}^{N-1} a_n \, x_{i,\lfloor n/\eta \rfloor} \, \psi(t - nT_c), \quad 0 \leq t \leq T, \tag{1}$$

where $\lfloor u \rfloor$ denotes the integer part of the real number u, T is the duration of the data waveform, ψ is the chip waveform, T_c is the inverse of the chip rate, and $N = \eta L = T/T_c$ is the number of chips per data waveform. Our performance results in Section VII are for $L = M$ with $M = 32$ or $M = 64$. These values of M were chosen to match existing and proposed systems, but our methods extend easily to larger values. Because $L = M$, the ratio of the chip rate of the signature sequence to the chip rate of the data waveform is $\eta = N/M$.

For the JTIDS signals, quasi-orthogonal data waveforms with $M = 32$ are derived from the cyclic shifts of a particular sequence of length 32, as described in [4] and [5]. This method for generating waveforms, often referred to as cyclic code-shift keying (CCSK), was apparently rediscovered in the late 1990s under the name code-phase-shift keying (CPSK). An implementation of CPSK is described in [6]. Recent results in [7] demonstrate that CCSK has a low probability of being intercepted if M is sufficiently large. For the results in Section VII, the data vectors for the orthogonal signals are the rows of an M by M Hadamard matrix. For $M = 64$, the resulting orthogonal signals are referred to as 64-ary Walsh signals in the TIA-95 cellular CDMA system [8]. For the reverse link of the TIA-95 system, $L = M = 64$, $\eta = 4$, and $N = 256$. Previous investigations of direct-sequence spread spectrum with orthogonal modulation include [9]–[12].

In Section VII, performance results are given for 32-ary orthogonal waveforms with $\eta = 1$ and $\eta = 4$, 64-ary orthogonal waveforms with $\eta = 1$, and 32-ary CCSK with $\eta = 1$. The chip waveform is the rectangular pulse that is defined by $\psi(t) = 1$ for $0 \leq t \leq T_c$ and $\psi(t) = 0$ for other values of t. The channel models are the additive white Gaussian noise channel, the multiple-access channel, and the specular multipath channel.

III. Transmission Protocols

We consider networks in which all transmissions other than broadcast transmissions are receiver directed, some form of reservation protocol is employed for channel access, the terminals cannot simultaneously transmit and receive, and the receivers cannot demodulate multiple signals simultaneously. For receiver-directed direct-sequence spread-spectrum transmissions, a packet that is sent to a terminal must use that terminal's signature sequence. Except for broadcast transmissions, one or more reservation packets must be exchanged before a terminal is permitted to transmit a data packet to another terminal. For example, the transmitting and receiving terminals may exchange request-to-send (RTS) and clear-to-send (CTS) packets (e.g., as in 802.11, but without channel sensing).

Because the transmissions are receiver-directed, it is usually impossible for terminals not involved in an RTS-CTS exchange to extract information from the reservation packets. Also, terminals that are sending or receiving data packets are unable to demodulate reservation packets or sense traffic on the channel. Thus, although the reservation protocol may keep multiple terminals from transmitting simultaneously to the same receiver, simultaneous transmissions between different pairs of transmitters and receivers are likely. To some extent, it may be desirable for the channel-access protocol to permit simultaneous transmissions as a means of increasing network throughput. However, simultaneous transmissions may produce significant multiple-access interference. In the event of an unfavorable near-far ratio, even a single interfering transmission may cause a packet to be received incorrectly. Similarly, multipath propagation may produce interference that prevents demodulation of a packet. The inability of the receiver to demodulate multiple signals simultaneously precludes the use of rake combining for multipath channels.

IV. Demodulation

The received signal is first converted to a baseband signal in a coherent demodulator. The baseband signal $Y(t)$ is the input to the baseband demodulator illustrated in Fig. 1, which consists of a chip-matched filter followed by a discrete-time processor. The impulse response of the filter is $h_c(t) = \psi(T_c - t)$. The discrete-time processor could be implemented as a set of parallel discrete-time correlators or discrete-time sequence-matched filters, one for each waveform in the set, but lower-complexity implementations

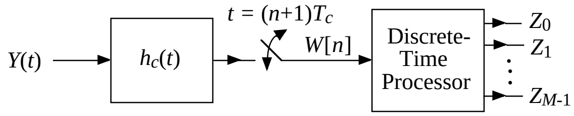

Fig. 1. Chip-matched filter and discrete-time processor.

are common. The baseband demodulator could also be used in the inphase and quadrature branches of a noncoherent receiver.

If the data waveforms are obtained from a Hadamard matrix, then, for each received data symbol, the discrete-time processor could multiply the Hadamard matrix and the vector of samples at the output of the chip-matched filter to produce the vector $\mathbf{Z} = (Z_0, Z_1, \ldots, Z_{M-1})$. The matrix multiplication can be accelerated by use of fast transform techniques, perhaps with commercial products. Even if the waveforms are not orthogonal, the discrete-time processor can be implemented by matrix multiplication, but the matrix is not an orthogonal matrix. The output vector $\mathbf{Z}$ is the input to a decision device that decides in favor of symbol k if $Z_k \geq Z_i$ for $0 \leq i \leq M-1$, which gives the maximum-likelihood coherent demodulator for the AWGN channel.

V. Evaluation of Error Probabilities

For an AWGN channel with two-sided noise spectral density $N_0/2$, the probability of symbol error for maximum-likelihood coherent demodulation of orthogonal waveforms can be computed from the expression

$$P_e = \frac{M-1}{\sqrt{2\pi}} \int_{-\infty}^{\infty} [\Phi(v)]^{M-2} \, \Phi\left(v - \sqrt{2\mathcal{E}/N_0}\right) \times \exp(-v^2/2)\, \mathrm{d}v, \quad (2)$$

where $\mathcal{E}$ is the energy per symbol, and Φ is the standard Gaussian distribution function. A similar expression is available for noncoherent demodulation. The integral in (2) can be written in terms of $\mathcal{E}_b$, the energy per bit, by replacing $\mathcal{E}$ with $m\mathcal{E}_b$. (Recall that $m = \log_2 M$.) Furthermore, it is known that for orthogonal waveforms the probability of bit error is $P_b = MP_e/[2(M-1)]$.

For quasi-orthogonal waveforms, such as CCSK, exact expressions for the symbol and bit error probabilities may be complicated and difficult to evaluate numerically, even for an AWGN channel. One exception is for CCSK waveforms that represent the cyclic shifts of a maximal-length linear-feedback shift-register sequence (m-sequence) of period $M-1$. Such a CCSK signal set is a subset of a simplex set of size M. If the energy per symbol for the resulting simplex set of $M-1$ waveforms is $\mathcal{E}_s = (M-1)\mathcal{E}/M$, the symbol error probability is given by (2). The energy saving is very small if M is large, and the simplex CCSK set is not convenient for binary data because $M-1$ is not a power of 2.

The symbol error probability for most CCSK signals, including those used in JTIDS, cannot be evaluated from (2), since the signal sets are neither orthogonal sets nor subsets of simplex sets. Furthermore, there is no simple relationship between the bit and symbol error probabilities for general CCSK. Thus, it is usually more convenient to evaluate the probability of error for CCSK by simulation rather than by analytical methods, even for AWGN channels. Moreover, we are interested in channels with multipath and multiple-access interference, for which analytical evaluation is even more difficult. Thus, we employ a combination of analysis and simulation to determine the performance of high-rate direct-sequence spread spectrum with orthogonal and quasi-orthogonal waveforms for multipath and multiple-access channels. The output of the chip-matched filter is determined by analytical means, and the discrete-time processor and the decision device are simulated. Comparisons of analytical and simulation results for AWGN channels are given in Section VII to verify the accuracy of the simulation.

The signal that the receiver is attempting to demodulate is referred to as the *desired signal*. The *interference model* used in our simulation is the baseband equivalent of RF interference in which the carrier phase for each interference signal is equal to the carrier phase for the desired signal. Furthermore, the chips of the interference signals are aligned with the chips of the desired signal, but the data symbols are not necessarily aligned. Thus, the interference signals and the desired signal are phase coherent and chip synchronous, but not symbol synchronous. This model represents the maximum value of the interference over the full ranges of phase and chip offsets. Thus, the simulation provides a guaranteed performance level that is valid for all phase and chip offsets. Furthermore, the results of the baseband simulation can be used to estimate the performance for interference with random phase angles and chip offsets, as discussed in the next section.

VI. Gaussian Approximation

The Gaussian approximation for systems with random signature sequences, which was suggested in [13] and developed further in [14] and [15], can be combined with (2) to estimate the performance of high-rate direct-sequence spread-spectrum communications over channels with multipath and multiple-access interference. Consider the decision statistic Z_i at the output of the discrete-time processor in Fig. 1, and assume the signature sequences are independent sequences of independent binary random variables. The parameter SNR_i is defined as the conditional mean of Z_i divided by the square root of the conditional variance of Z_i. The mean and variance are conditioned on the ith waveform being sent. The conditional mean of Z_k given that the ith signal is sent is zero for orthogonal waveforms and approximately zero for quasi-orthogonal waveforms.

For systems with orthogonal waveforms and random signature sequences, the symmetry of the model implies that SNR_i does not depend on i. Thus, we remove the subscript and denote the parameter by SNR. From previous results for binary systems (e.g., [14] or [15]), it is easy to show that for orthogonal waveforms

$$\mathrm{SNR} = \left\{ \frac{\beta}{N} + \frac{N_0}{2\mathcal{E}} \right\}^{-\frac{1}{2}}. \quad (3)$$

The parameter β is the ratio of the power in the interference to the power in the desired signal. We define the signal-to-interference ratio as $\mathrm{SIR} = 1/\sqrt{\beta}$, which is a voltage ratio rather than a power ratio. In decibels, we have $\mathrm{SIR_{dB}} = 20\log_{10}(\mathrm{SIR}) = -10\log_{10}(\beta)$. Notice that SNR, which is also a voltage ratio, accounts for thermal noise and interference, but SIR accounts for interference only. For the numerical results in Section VII, N is equal to M. Approximations to the probability of symbol error are obtained from (2) if $\sqrt{2\mathcal{E}/N_0}$ is replaced with SNR. In the absence of multipath or multiple-access interference, $\mathrm{SNR} = \sqrt{2\mathcal{E}/\mathrm{N}_0}$ and (2) gives the exact probability of error for orthogonal waveforms. The Gaussian approximation for random interference signals with random phase and chip offsets is obtained by replacing β with $\beta/3$ in (3), reflecting the fact that the conditional variance of the multiple-access interference with no phase or chip offsets is three times the conditional variance for interference with random phase and chip offsets. In the next section, the Gaussian approximation is compared with our simulation results for systems with orthogonal waveforms and channels with multiple-access interference.

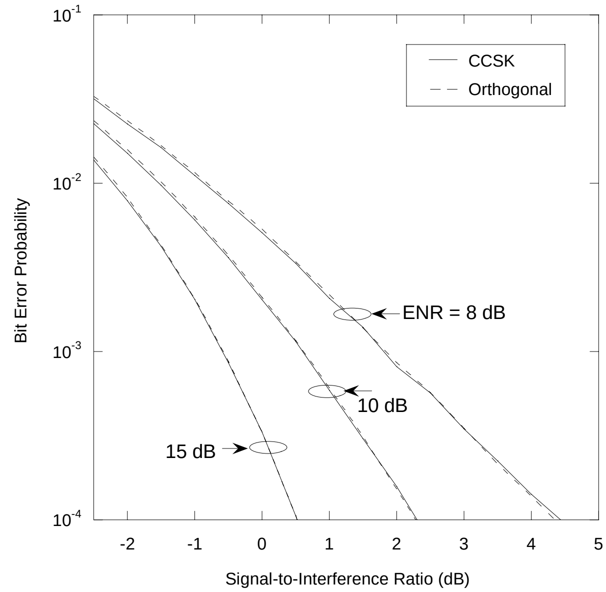

Fig. 2. Bit error probability for CCSK and orthogonal waveforms ($M = 32$, $\eta = 1$).

VII. Performance Results

Performance results are given in this section for high-rate direct-sequence spread-spectrum systems with multiple-access or multipath interference. For $M = 32$ and $M = 64$, the orthogonal waveforms are derived from a Hadamard matrix. The CCSK waveforms, which are for $M = 32$ only, are the JTIDS waveforms [4]. The signature sequences for the desired signal and the interference signals are m-sequences of period 1023.

Our numerical results for multiple-access interference are presented as graphs of the probability of bit error as a function of SIR for different values of $\mathrm{ENR} = \mathcal{E}_b/N_0$, the energy-per-bit to noise-density ratio. The range of bit error probabilities of primary interest is from 10^{-2} to 10^{-4}. Error rates in the lower end of the range may be required for a system with little or no error-control coding, but error rates in the upper end of the range are adequate if a good error-correcting code is employed.

We verified the accuracy of the simulation by comparing the exact bit error probabilities obtained from (2) with those obtained from our simulation of M-ary orthogonal modulation and an AWGN channel. Over the range of bit error probabilities from 10^{-1} to 10^{-4}, the largest percentage difference for $M = 32$ and $\eta = 1$ is 2.7%, the largest difference for $M = 32$ and $\eta = 4$ is 3.3%, and the largest difference for $M = 64$ and $\eta = 1$ is 5.4%. Thus, for example, if the exact probability is 1.0×10^{-4} in a system with 32-ary orthogonal waveforms and an AWGN channel, the probability obtained from simulation is in the range from 1.033×10^{-4} to 0.967×10^{-4} for $\eta = 4$ or $\eta = 1$. The graph of the simulation results is indistinguishable from the graph of the exact results.

The Gaussian approximation depends only on the total power in the interference, so it is independent of the number of interfering signals and the distribution of power among them. The actual probability of error depends on the number of interference signals and the power distribution. In previous investigations, it has been observed that for parameters of practical interest the concentration of the total interference power in a single signal gives the largest probability of error. Our simulations with multiple interfering signals confirm this observation, so we chose to focus on multiple-access channels with a single interference signal.

We begin with a comparison of the performance of the JTIDS 32-ary CCSK waveforms and 32-ary orthogonal waveforms for a system with 32 chips per symbol (i.e., $\eta = 1$) and a channel with multiple-access interference. The results are illustrated in Fig. 2 for three values of ENR, the energy-per-bit to noise-density ratio. Although the JTIDS CCSK modulation has a larger error probability than orthogonal modulation for an AWGN channel, we see that there is no significant performance difference for the multiple-access channel.

Results for 32-ary orthogonal modulation in a system with multiple-access interference are illustrated in Fig. 3. Notice that if ENR is 10 dB or less, a bit error probability of 10^{-3} or smaller is not possible if the power in the interference exceeds the power in the desired signal. If a bit error probability of 10^{-2} is acceptable, the interference power can exceed the power in the desired signal by 1 dB to 2 dB, depending on the energy-per-bit to noise-density ratio.

Comparisons between simulation results and the Gaussian approximation are also illustrated in Fig. 3. If ENR is 8 dB, the Gaussian approximation is fairly accurate, but the accuracy is worse for larger values of ENR. In terms of predicting the required value of SIR, we see that the Gaussian approximation

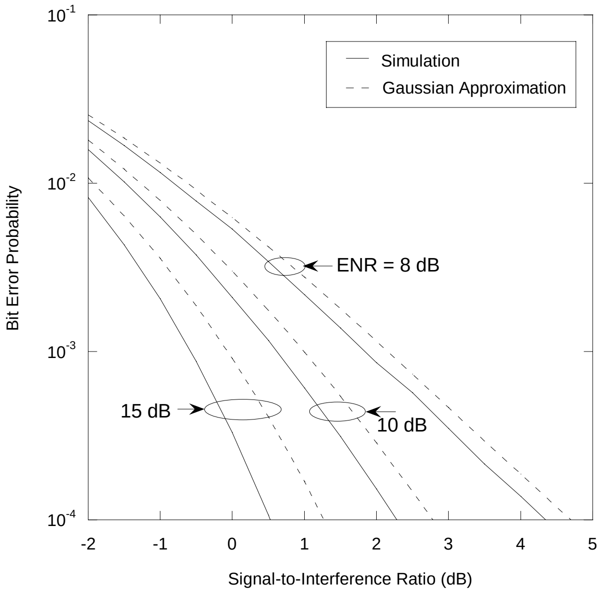

Fig. 3. Bit error probability for orthogonal waveforms ($M = 32$, $\eta = 1$).

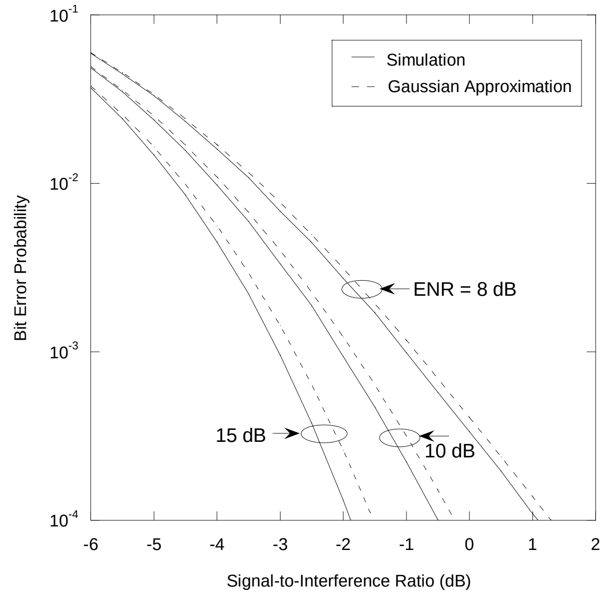

Fig. 4. Bit error probability for orthogonal waveforms ($M = 64$, $\eta = 1$).

is off by more than 0.7 dB if ENR is 15 dB and the desired bit error probability is 10^{-4}. For 10^{-2}, however, the Gaussian approximation is within 0.3 dB for each of the three values of ENR shown in Fig. 3.

The corresponding simulation results for multiple-access interference in a system with 64-ary orthogonal modulation are shown in Fig. 4. For a bit error probability of 10^{-4}, the SIR requirements for $M = 64$ are more than 2.5 dB less than the requirements for $M = 32$. If ENR is 8 dB, the required value of SIR is more than 3 dB less for $M = 64$ than for $M = 32$. From a comparison of Figs. 3 and 4, we observe that the Gaussian approximation is considerably more accurate for $M = 64$ than for $M = 32$. For $M = 64$, the error in the approximation is only about 0.1 dB if the bit error probability is 10^{-2}.

The accuracy of the Gaussian approximation also improves as η increases and M is held constant, as can be seen by comparing Figs. 3 and 5. If $M = 32$ and $\eta = 4$, the error in the Gaussian approximation is less than 0.1 dB for the entire range of error probabilities. More important is the fact that the performance for a multiple-access channel improves as η increases. For $M = 32$ and $\eta = 4$, a bit error probability of 10^{-4} can be achieved for signal-to-interference ratios less than -1 dB for all three values of ENR shown in Fig. 5.

Simulation results for a high-rate direct-sequence spread-spectrum system that employs CCSK waveforms ($M = 32$) are illustrated in Fig. 6 for two multipath channels. The number of interference components is denoted by J, so J+1 is the total number of paths. For each multipath channel, the receiver demodulates the signal from one path (referred to as the desired signal component), and the signal components from other paths are sources of interference. The two-path channel (J=1) has a desired signal component and a single interference component with a delay of 16 chips compared to the desired signal. The three-path channel (J=2) has a desired signal and two equal-power interference components with delays of 8 chips and 16 chips compared to the desired signal. For the results in Fig. 6, the power in the single interference component of the two-path channel is the same as the total power in the two interference components of the three-path channel. For each multipath channel, the *multipath power ratio* is the ratio of the power in the desired signal to the total power in the interference signals. As expected, for a given amount of power in the interference components, the error probability is larger if the power is concentrated in a single interference component. We also simulated a system with 32 orthogonal waveforms for the same multipath channels, and we found the results differ from those in Fig. 6 by no more than 0.1 dB. Thus, our results do not support the claim in [2] that CCSK is inferior to orthogonal modulation for high-rate direct-sequence spread spectrum in a multipath environment.

VIII. **Conclusions**

If there is no interference in the channel, high-rate direct-sequence spread spectrum has the potential to provide high throughput in mobile ad hoc wireless networks for military applications. The results in this paper indicate that high-rate direct-sequence spread spectrum with either orthogonal or CCSK waveforms can handle only moderate amounts of multipath and multiple-access interference. We found no significant difference in performance between CCSK and orthogonal modulation for multiple-access channels or multipath channels.

Methods for the assessment of the multiple-access capability are provided in Sections V and VI. For $M = 32$ and $\eta = 1$, the

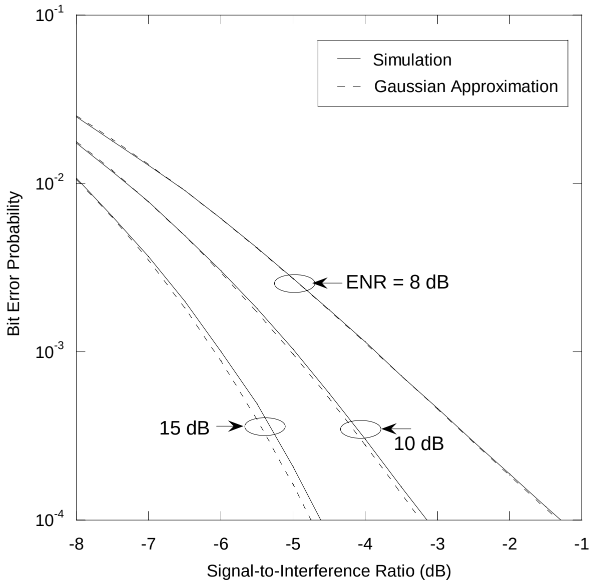

Fig. 5. Bit error probability for orthogonal waveforms ($M = 32$, $\eta = 4$).

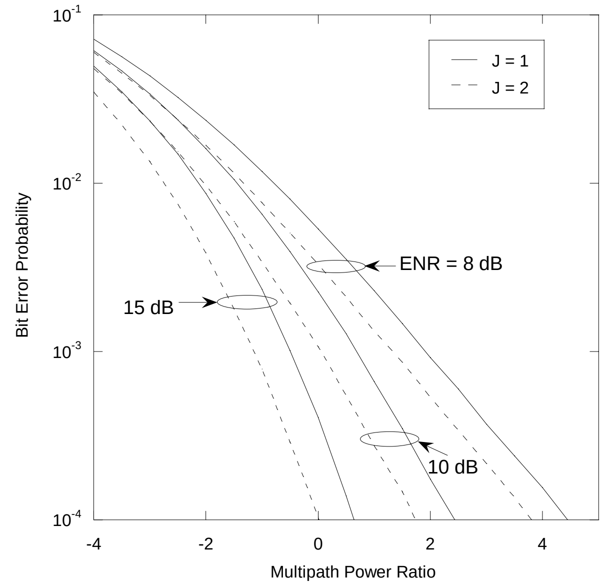

Fig. 6. Bit error probability for CCSK with multipath ($M = 32$, $\eta = 1$).

Gaussian approximation gives error probabilities that are larger than the true values, and they are significantly larger than the true values if $\mathcal{E}_b/N_0$ is large. Both the performance of the system and the accuracy of the Gaussian approximation are better for larger values of M and η.

Signal-to-interference ratios less than 0 dB can be tolerated for $M = 32$ and $\eta = 1$ only if the required bit error rate is in the upper end of the range from 10^{-4} to 10^{-2}. Even for the upper end of the range, $\mathcal{E}_b/N_0$ must be fairly large in order to achieve acceptable performance for SIR less than -1 dB. For SIR less than -3 dB, high-rate direct-sequence spread spectrum with $M = 32$ and $\eta = 1$ cannot achieve $P_b \leq 10^{-2}$ for $\mathcal{E}_b/N_0 \leq 15$ dB. As expected, the performance is better for 64-ary orthogonal modulation and for 32-ary modulation with $\eta = 4$. In either case, $P_b \leq 10^{-3}$ can be achieved for some values of SIR less than -1 dB if $\mathcal{E}_b/N_0 > 8$ dB. For example, if $M = 32$ and $\eta = 4$, then SIR ≈ -5 dB requires that $\mathcal{E}_b/N_0 \approx 10$ dB in order to achieve a bit error probability of 10^{-3}. In general, we find that modest levels of interference can be tolerated if $M = 64$ or if $M = 32$ and $\eta = 4$. However, for $M = 64$ or $M = 32$, future military applications of high-rate direct-sequence spread spectrum in mobile ad hoc networks require adaptive-transmission and channel-access protocols that can ensure the interference power rarely exceeds the power in the desired signal by more than a few dB.

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