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Capacity and Mutual Information of Soft and Hard Decision Output M-ary PPM over UWB Channels

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***Abstract*—The performance of impulse radio is investigated for frequency-selective ultra-wideband (UWB) channels and orthogonal M-ary pulse position modulation (PPM) in terms of the maximum achievable data rate, where the outputs of the M-ary correlation receiver implemented through the rake structure with conventional maximal ratio combining undergo either hard or soft decisions. The major difference observed between the soft and hard decision output systems is that at low code rates increasing the constellation size, M, requires an increase in the minimum signal-to-noise ratio (SNR) per communicated bit for the system with hard decisions. It is demonstrated that independent of the type of decision making, achievable rates in UWB channels are low compared to those in additive white Gaussian noise channels, as the frequency-selectivity of the UWB channel destroys the orthogonality of the transmitted PPM signals. The interference from the other users and the constellation size put additional constraints on the achievable data rates at high values of SNR.**

I. Introduction

In order to reduce the additional interference imposed on the overlain systems existing in dedicated bands, ultra-wideband (UWB) communications is allowed at a very low average transmit power. Restricted to short range, UWB technology has become an attractive candidate for the physical layer of the IEEE 802.15 Wireless Personal Area Network for high-rate local connectivity [1]. This in turn has rendered the data rates achievable by practical multiple access and modulation schemes with ultra-wide bandwidths a subject of continuing research.

For traditional "impulse radio" UWB communication systems characterized by the transmission of subnanosecond duration pulses, several transmitted pulses comprise an information symbol, and multiple access is obtained through time-hopping (TH). Data modulation can be accomplished through pulse position modulation (PPM) as in [2]. The information-theoretic capacity of M-ary PPM over additive white Gaussian noise (AWGN) channels with soft decision outputs is computed in [3] for single-user UWB communications, and the results are extended to multi-user communications in [4] and [5] using the Gaussian approximation for the multiple access interference (MAI). The inaccuracy of the Gaussian MAI assumption for AWGN channels, which leads to an overestimation of data rates, is demonstrated by developing a more precise statistical analysis of the MAI when evaluating the capacity of M-ary PPM with hard decision outputs [6]. In addition, the single-user high transmission rates foreseen in AWGN channels with soft decision outputs are not carried over to practical UWB systems operating in dense multipath environments: the binary PPM single-user system in [7] has considerably lower rates in non-line-of-sight (NLOS) frequency-selective channels in comparison to the rates anticipated in channels with no fading.

In this paper, the rates achievable by TH-PPM systems over frequency-selective UWB channels with either soft or hard decision outputs are determined in the presence of other UWB users. Since both the loss of orthogonality between the transmitted PPM signals and the interference from the other users are shown to result in a degradation of achievable data transmission rates, the current work substantially differs from [8], which ignores the correlation between the PPM signals and treats impulse radio as a single-user PPM communication over a flat fading channel subject to AWGN.

The rest of the paper is organized as follows: In the next section, the UWB system model is introduced. Section III is devoted to the calculation of the mutual information between the input and the output of a digital communication system with M-ary PPM, which has either soft or hard decisions. The simulation results in Section IV demonstrate the major differences between soft and hard decision making. The last section contains the conclusions.

II. UWB System Model

The signal transmitted by the kth user of the considered impulse radio UWB system with TH for multiple access and M-ary PPM for data modulation is

$$s_{\mathrm{tr}}^{(k)}(t)=\sqrt{\frac{E_s}{N_s}}\sum_{j=-\infty}^{\infty}w_{\mathrm{tr}}(t-jT_f-c_j^{(k)}T_c-\delta d_{\lfloor j/N_s\rfloor}^{(k)}),\quad (1)$$

where E_s is the symbol energy, N_s is the number of pulses transmitted per symbol, and each pulse, $w_{\mathrm{tr}}(t)$, with duration T_p, is sent during a frame of T_f seconds. In each frame, the exact pulse position is determined by the TH sequence element $c_j^{(k)}$ and the data symbol, $d_{\lfloor j/N_s\rfloor}^{(k)}\in\{0,\ldots,M-1\}$, both of which are specific to user k. The starting time of the jth pulse is shifted by $c_j^{(k)}T_c + \delta d^{(k)}_{\lfloor j/N_s \rfloor}$, where $c_j^{(k)} \in \{0, 1, \ldots, N_h - 1\}$ is chosen randomly. Here, N_h and T_c are respectively the number and the duration of the bins to which the pulses are allowed to hop. The effect of δ on data modulation is such that $\delta \geq T_p$ is necessary for orthogonal PPM. The frame time and the bin duration are $T_f = (N_h - 1)T_c + M\delta$ and $T_c = T_p$.

Transmitted signals arrive at the receiver through frequency-selective channels due to the wide system bandwidth:

$$h_k(t) = \sum_{\ell=0}^{L_k-1} \alpha_{k,\ell}\delta_D(t - \tau_{k,\ell}), \tag{2}$$

where $\alpha_{k,\ell}$ and $\tau_{k,\ell}$ are the gain and the delay of the ℓth path of the kth user, and $\delta_D(\cdot)$ is the Dirac delta function. The receiver antenna modifies the shape of the transmitted pulse so that the kth received signal is

$$s^{(k)}_{\mathrm{rec}}(t) = \tilde{s}^{(k)}_{\mathrm{tr}}(t) * h_k(t) \tag{3}$$

where $*$ denotes convolution and $\tilde{s}^{(k)}_{\mathrm{tr}}(t)$ is $s^{(k)}_{\mathrm{tr}}(t)$ formed using the received pulse shape, $w_{\mathrm{rec}}(t)$ instead of $w_{\mathrm{tr}}(t)$. The second derivative Gaussian pulse in [9] with the autocorrelation function

$$R_w(\tau) = \left[1 - 4\pi\left(\frac{\tau}{t_n}\right)^2 + \frac{4\pi^2}{3}\left(\frac{\tau}{t_n}\right)^4\right] \exp\left\{-\pi\left(\frac{\tau}{t_n}\right)^2\right\} \tag{4}$$

is the usual choice for $w_{\mathrm{rec}}(t)$. With this model of the received pulse, $T_p = 2t_n$, in which 99.99% of the pulse energy is confined. Representing the asynchronism between the transmitter of user k and the receiver with τ_k, the total received signal is given by

$$r(t) = s^{(0)}_{\mathrm{rec}}(t - \tau_0) + \sum_{k=1}^{K} s^{(k)}_{\mathrm{rec}}(t - \tau_k) + n(t), \tag{5}$$

where $n(t)$ is the AWGN component with two-sided power spectral density $N_0/2$. The signal model in (5) highlights the fact that there are K interfering users and a desired user, which is, without loss of generality, the 0th user. The receiver is assumed to be perfectly synchronized with this user so that it has access to the elements of its TH code and $\tau_0 = 0$.

The optimum detector for orthogonal M-ary PPM signaling in the AWGN channel is the M-ary correlation receiver, which selects the index of the branch corresponding to the largest cross-correlation between the received signal and the M possible transmitted signals when making hard decisions [10]. For the frequency-selective channel given by (2) and $k = 0$, each branch, y_m, of the M-ary correlation receiver, where y_m is the correlator output related to the mth possible transmitted signal, is realized as an all-rake structure

$$y_m = \sum_{j=0}^{N_s-1}\sum_{\ell=0}^{L_0-1} \alpha_{0,\ell} \underbrace{\int_0^{\delta} r(t + jT_f + c_j^{(0)}T_c + \tau_{0,\ell} + m\delta) w_{\mathrm{rec}}(t)\mathrm{d}t}_{y_{j,\ell,m}}. \tag{6}$$

In (6), to detect the 0th symbol of the desired user the correlator outputs of different paths, $y_{j,\ell,m}$, are combined using the conventional maximal ratio combining (MRC) scheme, and the contribution from different pulses are weighted equally. The optimality of the conventional MRC is violated as a result of the interference from the other users, and the degradation of performance with an enlargement of the number of interfering users is accompanied by a lowering of achievable data rates as discussed in the following sections.

III. Mutual Information of M-ary Orthogonal PPM in Frequency-Selective Channels

Evaluation of achievable data rates over frequency-selective channels with impulse radio and M-ary PPM is based on the direct calculation of the mutual information between the input and the output of the system. The input-output relationship of the digital communication system displayed in Fig. 1 indicates that for orthogonal PPM, the output of a k-bit source, $\mathbf{U} = (U_1 \ldots U_k) \leftrightarrow d^{(0)} = i$, is mapped to one of $\mathrm{M} = 2^k$ mutually orthogonal signal waveforms, $s_i(t)$ such that

$$x_m = \int_{-\infty}^{\infty} s_i(t)s_m(t)\mathrm{d}t = E_s\delta_D(i - m) \tag{7}$$

for $i, m \in \{0, \ldots, \mathrm{M}-1\}$. At the receiving end, the waveform affected by the UWB channel, the noise and the MAI is processed by the demodulator to reduce it to a vector representing the estimate of the transmitted symbol [10], which is $\mathbf{Y} = [y_0 \ \ldots \ y_{\mathrm{M}-1}]^T$. In one of the extreme cases of soft decision making the detector following the demodulator performs no quantization on $\mathbf{Y}$ as considered here or it decides on which one of the M symbols is transmitted by choosing the index of the maximum component of $\mathbf{Y}$, in which case it makes a hard decision.

A. Mutual Information of Soft Decision Output M-ary PPM

Since $\mathbf{X} \triangleq [x_0 \ \ldots \ x_{\mathrm{M}-1}]^T$ and $\mathbf{U}$ are related by an invertible transformation, the mutual information between the transmitted symbol, $d^{(0)}$, and $\mathbf{Y}$ is equivalent to the mutual information between $\mathbf{X}$ and $\mathbf{Y}$:

$$I(d^{(0)}; \mathbf{Y}) = I(\mathbf{X}; \mathbf{Y}). \tag{8}$$

The mutual information between the discrete-valued input $\mathbf{X}$ and the continuous-valued output $\mathbf{Y}$ is

$$-\sum_{i=0}^{\mathrm{M}-1} p(\mathbf{x}_i) \int_{-\infty}^{\infty} f(\mathbf{y}/\mathbf{x}_i)\log_2\left(\frac{\sum_{j=0}^{\mathrm{M}-1} p(\mathbf{x}_j) f(\mathbf{y}/\mathbf{x}_j)}{f(\mathbf{y}/\mathbf{x}_i)}\right)\mathrm{d}\mathbf{y}. \tag{9}$$

In (9), $\mathbf{y}$ is the realized vector value of $\mathbf{Y}$ and $\mathbf{x}_i$ is the vector $\mathbf{X}$ for the ith transmitted symbol. For instance,

$$\mathbf{x}_0 = [E_s \underbrace{0 \ldots 0}_{\mathrm{M}-1}]^T. \tag{10}$$

The vector $\mathbf{x}_i$ is disturbed by the UWB channel, the AWGN and the MAI. Therefore, the mth component of $\mathbf{Y}$, is composed of a signal term determined by the value of the symbol transmitted by the desired user, which is s_{mi} for $d^{(0)} = i$,

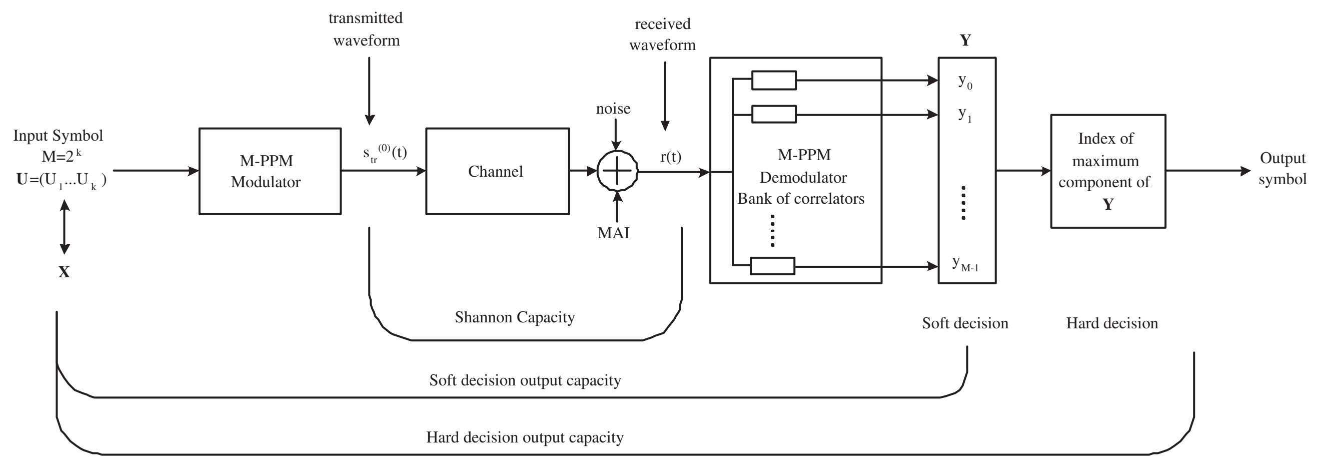

Fig. 1. The model for capacity calculation.

an inter-frame interference (IFI) term, $\eta_m^{(0)}$, a MAI term, $\sum_{k=1}^{K}\eta_m^{(k)}$, and a noise term n_m:

$$y_m = s_{mi} + \eta_m^{(0)} + \sum_{k=1}^{K}\eta_m^{(k)} + n_m \tag{11}$$

For specular channels and orthogonal PPM with $\tau_{k,\ell} = \ell T_p$ and $\delta = T_p$, respectively, the signal term is

$$s_{mi} = \sqrt{E_s N_s}\sum_{\ell=0}^{L_0-1}\alpha_{0,\ell}\alpha_{0,\ell+m-i}, \tag{12}$$

since $R_w(\tau) = 0$ for $|\tau| \geq T_p$. The noise terms n_m for $m \in \{0,\ldots,\mathrm{M}-1\}$ are uncorrelated zero-mean Gaussian random variables with the variance $N_s N_0/2\sum_{\ell=0}^{L_0-1}\alpha_{0,\ell}^2$.

As for the MAI, the term due to the kth interfering user in (11), $\eta_m^{(k)}$, is modeled as a zero-mean white Gaussian random variable with the variance

$$\mathrm{E}\left\{\left(\eta_m^{(k)}\right)^2\right\} = \frac{E_s}{T_f}\sum_{\ell=0}^{L_0-1}\sum_{\ell'=0}^{L_k-1}\left(\alpha_{0,\ell}\alpha_{k,\ell'}\right)^2\int_{-\infty}^{\infty}|R_w(\tau)|^2\,\mathrm{d}\tau. \tag{13}$$

This approximate expression is derived in [11] by assuming an independent random delay for each arriving path of each user, since calculation of the exact variance of $\eta_m^{(k)}$ depends on the involved random variables, which are the TH codes, data symbols, etc., and, hence, is computationally prohibitive. The IFI term in (11), similarly, is a zero-mean Gaussian random variable with the variance

$$\mathrm{E}\left\{\left(\eta_m^{(0)}\right)^2\right\} = \frac{E_s}{T_f}\sum_{\ell=0}^{L_0-1}\sum_{\substack{\ell'=0\\ \ell'\neq\ell}}^{L_0-1}\left(\alpha_{0,\ell}\alpha_{0,\ell'}\right)^2\int_{-\infty}^{\infty}|R_w(\tau)|^2\mathrm{d}\tau. \tag{14}$$

The arguments above designate Gaussian distributed y_m with mean s_{mi} and variance σ_m^2, which is the sum of the variances of n_m, $\eta_m^{(0)}$ and $\eta_m^{(k)}$ for $k = 1,\ldots,K$. Although capacity is achieved by equiprobable symbols, i.e., $p(\mathbf{x}_i) = 1/M$, for orthogonal signals and AWGN channels as indicated in [12], the frequency-selectivity of the UWB channel disturbs the symmetry of the output via s_{mi} except for $\mathrm{M} = 2$. Thus, the expression in (15) is the capacity for $\mathrm{M} = 2$ and the mutual information for higher constellation sizes, where $\mathbf{v} = [v_0 \ \ldots \ {}_{\mathrm{M}-1}]^T$, $v_m = y_m/\sigma_m$ and $\rho_{mi} = s_{mi}^2/\sigma_m^2$. The actual input distribution maximizing the mutual information for $\mathrm{M} > 2$ depends on the channel realizations of the engaged users. Since the expression in (15) requires handling an m-fold integral for the evaluation of $E_{\mathbf{v}/\mathbf{x}_i}\{\cdot\}$, it is computed by creating v_m, which are normally distributed with $\mathcal{N}(\sqrt{\rho_{mi}}, 1)$ and averaging the results.

B. Mutual Information of Hard Decision Output M-ary PPM

The detector makes a hard decision through choosing the maximum component of $\mathbf{Y}$ and assigning the index of this component as the estimate of the transmitted symbol:

$$\hat{d}^{(0)} = \arg\max_m y_m \tag{16}$$

To compute $I(\mathbf{X};\hat{d}^{(0)})$, the conditional probabilities $\Pr\{\hat{d}^{(0)} = m|\mathbf{x}_i\}$, $\forall m, i$,

$$\int_{-\infty}^{\infty}\phi(v_m - \sqrt{\rho_{mi}})\prod_{\substack{p=0\\ p\neq m}}^{\mathrm{M}-1}\Phi(v_m - \sqrt{\rho_{mp}})\mathrm{d}v_m, \tag{17}$$

where ϕ and Φ are the pdf and the cumulative distribution function, respectively, of the standard normal distribution,

$$I(\mathbf{X};\mathbf{Y})\Big|_{p(\mathbf{x}_i)=\frac{1}{\mathrm{M}}} = -\sum_{i=0}^{\mathrm{M}-1}\frac{1}{\mathrm{M}}\mathrm{E}_{\mathbf{v}/\mathbf{x}_i}\left\{\log_2\left(\sum_{j=0}^{\mathrm{M}-1}\frac{1}{\mathrm{M}}\exp\left(\sum_{m=0}^{\mathrm{M}-1}\left[v_m(\sqrt{\rho_{mj}} - \sqrt{\rho_{mi}}) + \frac{\rho_{mi}-\rho_{mj}}{2}\right]\right)\right)\right\}. \tag{15}$$

$\mathcal{N}(0,1)$, are calculated. Then,

$$I(\mathbf{X};\hat{d}^{(0)}) = \mathrm{H}(\hat{d}^{(0)}) - \mathrm{H}(\hat{d}^{(0)}|\mathbf{X}) \tag{18}$$

is the mutual information of impulse radio with M-ary PPM and hard decision outputs, where $\mathrm{H}(\hat{d}^{(0)})$ is the entropy of $\hat{d}^{(0)}$ [13]. As in the soft decision output case, although equally likely symbols achieve the capacity for M = 2, the capacity achieving input distribution for higher M is a function of user channels.

C. Relationship between Shannon Capacity and M-ary PPM

The capacity of the band-limited AWGN waveform channel with a band-limited and average power-limited input originally derived by Shannon as

$$C = W\log_2\left(1+\frac{P_{av}}{WN_0}\right), \tag{19}$$

where C is the capacity in bits/second, W is the bandwidth in Hz and P_{av} is the average power, can be achieved by waveform inputs, whose samples are statistically independent Gaussian random variables [10]. The part of the digital communication system to which this capacity applies is pointed out in Fig. 1 as the "Shannon capacity". Although properties of modulation waveforms do not directly affect the soft and hard decision output M-ary PPM capacities in the single-user case, it is possible to reduce the MAI and improve data rates by choosing received pulse waveforms with desirable autocorrelation properties.

IV. Simulation Results

The achievable rates of TH-PPM systems having ultra-wide bandwidths is numerically estimated for equally likely symbols, where the output of the considered digital communication system contains either soft or hard decisions. The frequency-selective channel models are the ones proposed by the IEEE 802.15.3a task group, which are the CM1 and the CM4 models for the 0-4 m line-of-sight (LOS) and the 4-10 m extreme NLOS situations, respectively. The UWB system has $T_f = 100$ ns, $T_p = 1$ ns and $N_s = 1$.

For orthogonal signals in AWGN channels, the probability of error can be made as small as desired by increasing M provided that the bit signal-to-noise ratio (SNR) E_b/N_0, where E_b is the energy per information bit communicated over the channel, is greater than $\log_e 2$, which is called the "Shannon limit" for an AWGN channel [10]. Although the symbol SNR for frequency-selective channels is

$$\mathrm{SNR} = \frac{E_s}{N_0}\mathrm{E}\left\{\sum_{\ell=0}^{L_k-1}\alpha_{k,\ell}^2\right\}, \tag{20}$$

since the average received power of the multipaths is normalized to unity in order to provide a fair comparison with other wideband and narrowband systems [14], the average symbol SNR of frequency-selective and AWGN channels are identical: E_s/N_0. The bit SNR is related to the code rate, R, given in terms of the number of information bits per symbol, through $E_b/N_0 = (E_s/N_0)/R$. Since reliable communication is not possible above the channel capacity, $\xi_b \triangleq (E_b/N_0)_{\min} = (E_s/N_0)/C$.

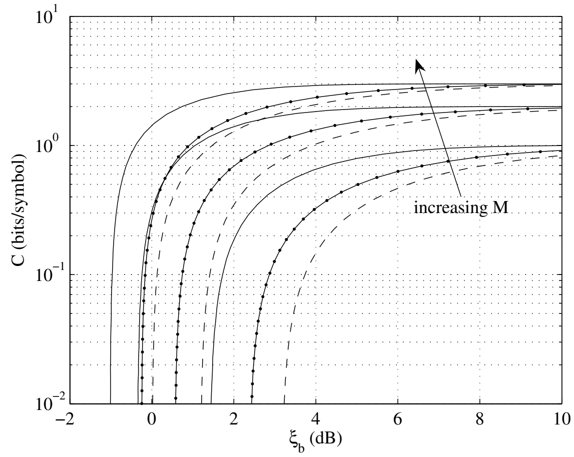

Fig. 2. The single-user mutual information of soft decision output TH-PPM systems against ξ_b in the AWGN (solid), the CM1 (dashed) and the CM4 (dotted) channels for M = 2, 4 and 8.

Obtained by averaging over a large number of channel realizations, Fig. 2 illustrates the capacity (M = 2) and the mutual information (M > 2) of TH-PPM systems in the AWGN and frequency-selective UWB channels with soft decision outputs against ξ_b in the absence of IFI and MAI. For AWGN channels, as M increases and the code rate approaches zero, ξ_b tends toward the Shannon limit. In addition to the associated increase in ξ_b from $\log_e 2$ to $\mathrm{M}\log_e 2/(\mathrm{M}-1)$ with finite M due to the orthogonality between transmitted symbols [12], Fig. 2 demonstrates that ξ_b increases further for the frequency-selective UWB channels considered, which have induced poor cross-correlation values on the previously orthogonal symbols. As compared to the AWGN channel, the rates achievable in terms of the number of information bits per channel symbol in the CM4 channel necessitates a higher minimum bit SNR, which even so is less than that required in the CM1 channel.

A similar analysis performed with the hard decision output systems indicates that the minimum ξ_b for a particular value of the constellation size is attained at a nonzero code rate as opposed to the soft decision output systems. Moreover, the minimum ξ_b with hard decision outputs are always larger than the corresponding values involving soft decisions. Although lower code rates are more efficient with soft decisions, the efficiency related to hard decision making rapidly drops after the minimum ξ_b, especially for $\mathrm{M} > 2$. Operation at the minimum ξ_b involves code rates of approximately 1/3, 1/6 and 1/6 given in terms of the ratio of the number of information bits to the number of channel bits for M = 8 in the AWGN, the CM1 and the CM4 channels, respectively. In Fig. 3, where these results are displayed, another major difference between soft and hard decision output systems is observed: increasing the constellation size for hard decision output systems at low code rates entails larger ξ_b, whereas higher M is advantageous at all code rates for the soft decision output systems.

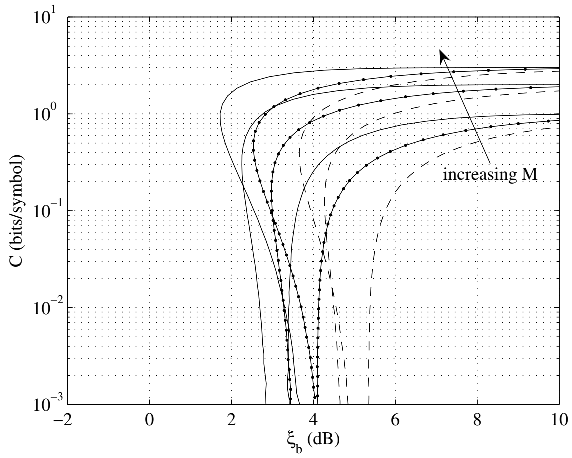

Fig. 3. The single-user mutual information of hard decision output TH-PPM systems against ξ_b in the AWGN (solid), the CM1 (dashed) and the CM4 (dotted) channels for M = 2, 4 and 8.

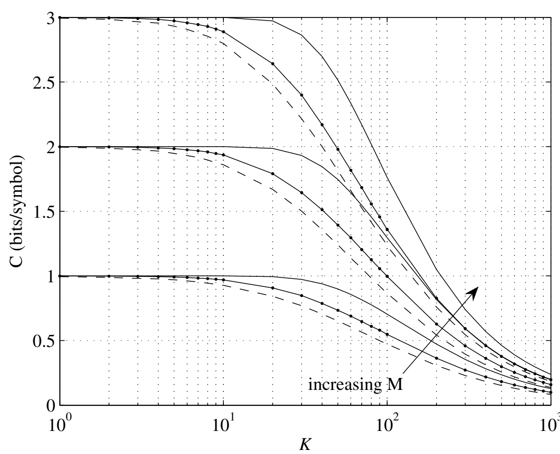

Fig. 4. The mutual information of soft decision output TH-PPM systems against K in the AWGN (solid), the CM1 (dashed) and the CM4 (dotted) channels for M = 2, 4 and 8, where $E_b/N_0 = 20$ dB.

The effect of the interfering users, which propagate over channels that are statistically equivalent to that of the desired user, on the performance of the TH-PPM system with single-user reception and soft decisions is a degradation relative to the single-user case in proportion to the number of interfering users. Therefore, the transmission rates for reliable communication has to decrease with increasing number of users. Fig. 4 serves to demonstrate this idea, where the bit SNR is large enough so that only the effect of the MAI is investigated. Among the UWB channel models, the cross-correlation properties of the CM4 channel are the better of the two in terms of reducing the MAI. When the low SNR region is of interest, it is discovered that the minimum ξ_b values do not change with K, since at low SNR values the system performance is limited by the noise and not the MAI. The same conclusions are also reached for the channels with hard decision outputs.

V. Conclusion

The calculated achievable rates of TH-PPM systems over frequency-selective UWB channels with either soft or hard decision outputs indicate that the correlation induced on the originally orthogonal transmitted signals by the channel limits data rates. The interference from the other UWB users restricts the rates further at high values of SNR, where the MAI dominates the performance of the system with single-user reception. The most important difference between soft and hard decision making is that while increasing the constellation size is advantageous for systems with soft decisions indeed, high M is preferred for hard decision output systems only at large code rates.

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