REDUCING THE REQUIRED TRANSMITTED POWER FOR PERSONAL COMMUNICATION DEVICES IN INDOOR-VIDEO SPREAD-SPECTRUM TRANSMISSION

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Abstract — **This paper analyzes the use of a coded multicode spread-spectrum technique to reduce radio power for personal communication devices in indoor video transmission. The performance of indoor video transmission is measured by the outage probability, which is the probability that the instantaneous bit error rate (BER) exceeds an acceptable level. We first derive the outage probability. The result is used to characterize the power reduction. It is found that the amount of power reduction depends on the required BER level. For indoor video transmission, which typically requires a BER of 10^{-5}, a substantial power reduction exceeding 3dB can be obtained. If outer coding is used in order that the BER requirement can be relaxed, results show that it is preferred to maintain the required BER to be less than 10^{-4} for more-effective utilization of the power-reduction capability.**

Index terms — **Low power, Video transmission, Indoor communications, Direct sequence spread spectrum, Multicode technique, Reed-Muller code, Bit error rate, Outage probability, Fading, Power control.**

I. INTRODUCTION

Video picture transmission is one of the most important elements in the realization of future personal communication systems. Reducing the transmitted power needed to achieve a given quality-of-service requirement in video transmission has a significant impact on the user satisfaction as the battery lifetime of personal communication devices can be enhanced. In this paper, we are concerned with radio-power reduction for personal communication devices that employ multicode (MC) direct-sequence spread-spectrum (DSSS) techniques for indoor video transmission. Indoor private networks operating in unlicensed spectrum and supporting multimedia services are expected to complement public cellular networks in future mobile communication systems to provide ubiquitous coverage, mobility and personalized services [1]. MC-DSSS techniques [2], because of various advantages such as the support of multirate data transmission, the robustness against interference and the provision of multiple access without the need for time and frequency coordination among users, are attractive for multimedia communications in unlicensed bands.

Recently, the first author [3] has proposed a coded MC-DSSS technique capable of improving the performance and reducing the peak-to-mean envelope power ratio (PMEPR) over the conventional MC-DSSS technique without the need to increase the transmission bandwidth, reduce the data-transmission rate, or sacrifice the processing gain. It has been shown that a high coding gain can be achieved and the PMEPR is reduced. The coding gain that is realized allows a reduction of the transmitted power, and the reduced PMEPR enables transmitters to operate the power amplifiers more efficiently. Both advantages are useful to low-power applications in indoor video communications that employ power-limited personal communication devices.

Applying the coded MC-DSSS technique can lead to a power reduction in indoor video transmission. However, a relevant performance analysis is lacking and the amount of power saving has not been characterized. In this paper, we analyze Q-ary PSK, coded MC-DSSS communications over indoor fading channels. It is shown that a substantial reduction of radio power can be achieved. This power reduction is especially attractive since an expansion of signal bandwidth or a reduction of data-transmission rate (which results in degradation of the visual quality of video pictures) is not required. Personal communication devices can be benefited from this additional power saving.

Since indoor fading channels are lowly dispersive in general [4], we consider the case that the channel fading is frequency-nonselective. Diversity reception is used to improve the performance. The performance of real-time

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video transmission is measured in terms of the outage probability, which is the probability that the instantaneous bit error rate (BER) exceeds an acceptable level. A BER of 10^{-5} is usually required to achieve acceptable quality for reconstructed video pictures [5]-[7], and an outage probability of 1% is widely used as a service-availability requirement [6], [8]. We derive the outage probability for indoor fading channels and the result is used to characterize the power reduction. For indoor wireless channels, measurement results have shown that the fading characteristic can be modeled by Rayleigh, Rician, Nakagami-*m*, Weibull, Suzuki, or lognormal distributions ([9] and the references therein). We consider Rayleigh and Rician distributions in the computation of numerical results for this paper. In low-power video-communication systems, power control is usually implemented in order to avoid transmission with excessive power, which would otherwise be wasted to yield a visual quality better than is required. Power control is considered in the present work.

This paper is organized as follows. Section II details the coded MC-DSSS technique. After describing the system model in Section III, we derive the outage probability in Section IV. The power reduction capability is analytically characterized and numerically evaluated in Sections V and VI, respectively. Finally, conclusions are given in Section VII.

II. CODED MC-DSSS TECHNIQUE

Figs. 1a and 1b plot the block diagrams for signal generation in a conventional MC-DSSS system and a coded system using the technique of [3], respectively. A conventional MC-DSSS system employs M orthogonal spreading sequences to code-division multiplex M symbols so that these symbols can be transmitted in parallel. The technique of [3] is different from the conventional one in that, instead of directly transmitting M uncoded symbols, the M symbols are first encoded into S coded symbols by an error-correcting code and the S coded symbols are then multiplexed by S orthogonal spreading sequences. The power of each coded symbol is adjusted by a factor M/S so that the uncoded and the coded systems have the same transmitted power. Since the chip rate and the spreading-sequence length remain unchanged, it follows that the signal bandwidth, the data-transmission rate and the processing gain are not affected by the addition of coding. The error-correcting capability of the coding can be utilized to improve the performance, thereby offering a possibility of reducing the transmitted power. Note that additional processing power is required for decoding, but it is performed at base stations rather than at personal communication devices.

The undesirable increase of PMEPR due to multiplexing S signals is minimized by a suitable choice of the error-correcting code. The Reed-Muller (RM) code proposed by Davis and Jedwab [10], therefore, has been adopted in the technique of [3]. It is shown in [3] and [11] that the resultant PMEPR can be as low as 3dB, 6.5dB and 5.2dB for the rectangular, square-root raised cosine, and Gaussian chip pulses, respectively, and is independent of the choice of M and S.[1] The low PMEPR implies that power amplifiers can be operated more efficiently, leading to a lower power consumption for personal communication devices.

The signal construction is detailed as follows. We consider Q-ary PSK transmission. The ith block of the M Q-ary data symbols, $\mathbf{u}_i = [u_{i,0}, u_{i,1}, \cdots, u_{i,M-1}]$, is encoded into a block of S coded symbols, $\tilde{\mathbf{u}}_i = [\tilde{u}_{i,0}, \tilde{u}_{i,1}, \cdots, \tilde{u}_{i,S-1}]$, by the RM code of [10], where $u_{i,m}, \tilde{u}_{i,m} \in \{0, 1, \cdots, Q-1\}$ and S is related to M by

$$S = 2^{M-1}. \tag{1}$$

The generated $\tilde{\mathbf{u}}_i$ is given by

$$\tilde{\mathbf{u}}_i = \frac{Q}{2}\sum_{m=1}^{M-2}\mathbf{x}_{\underline{\pi}(m)}\mathbf{x}_{\underline{\pi}(m+1)} + \sum_{m=1}^{M-1}u_{i,m}\mathbf{x}_m + u_{i,0}\mathbf{1} \mod Q \tag{2}$$

where: $\underline{\pi}$ is a permutation of $\{1, 2, \dots, M-1\}$; $\mathbf{x}_1, \mathbf{x}_2, \dots, \mathbf{x}_{M-1}$ are Boolean functions of length S; $\mathbf{1}$ is an all-one vector of length S; and $\mathbf{x}_{\underline{\pi}(m)}\mathbf{x}_{\underline{\pi}(m+1)}$ is the component-wise product of $\mathbf{x}_{\underline{\pi}(m)}$ and $\mathbf{x}_{\underline{\pi}(m+1)}$. The S coded symbols are multiplexed by S spreading sequences of length N. Let E_s and E_b denote the symbol energy and bit energy, respectively, for the *uncoded* data symbols. It follows that $E_s = E_b \log_2 Q$. The chip rate of the system is $1/T_c$. The complex envelope of the transmitted signal, $s(t)$, is given by

$$s(t) = \sqrt{\frac{2E_sR_c}{T_s}}\sum_{i=-\infty}^{\infty}\sum_{m=0}^{S-1}\tilde{I}_{i,m}\mu_m(t-iT_s) \tag{3}$$

where $R_c = M/S$ is the code rate, $T_s = NT_c$ is the symbol duration, $\tilde{I}_{i,m} = e^{j2\pi\tilde{u}_{i,m}/Q}$ is the ith Q-ary PSK coded symbol, and

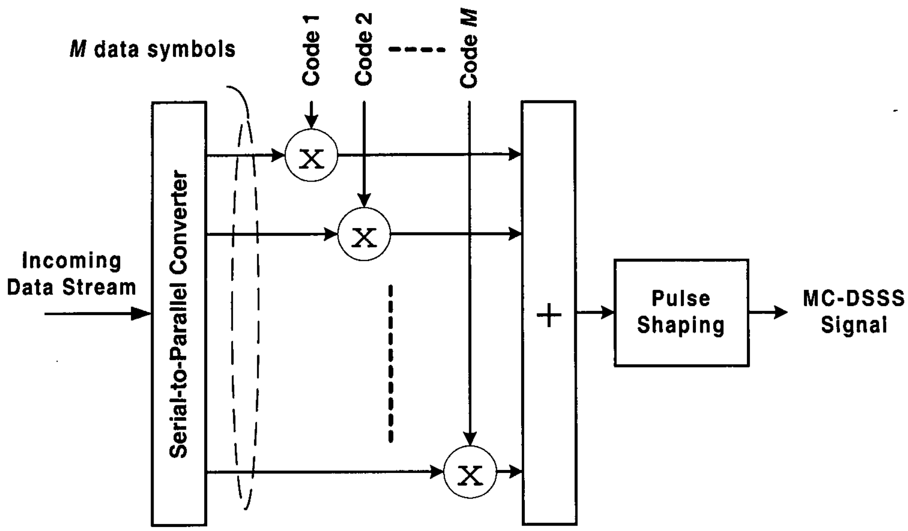
$$\mu_m(t) = \sum_{n=0}^{N-1}a_{n,m}\psi(t-nT_c) \tag{4}$$

is the mth spectral spreading waveform. In (4), $\psi(t)$ is the chip pulse satisfying $\int_{-\infty}^{\infty}|\psi(t)|^2\,dt = T_c$, and $a_{n,m}$, satisfying $|a_{n,m}| = 1$, is the nth chip of the mth spreading sequence. Since the RM code of [10] was originally designed to minimize the PMEPR for OFDM systems, we

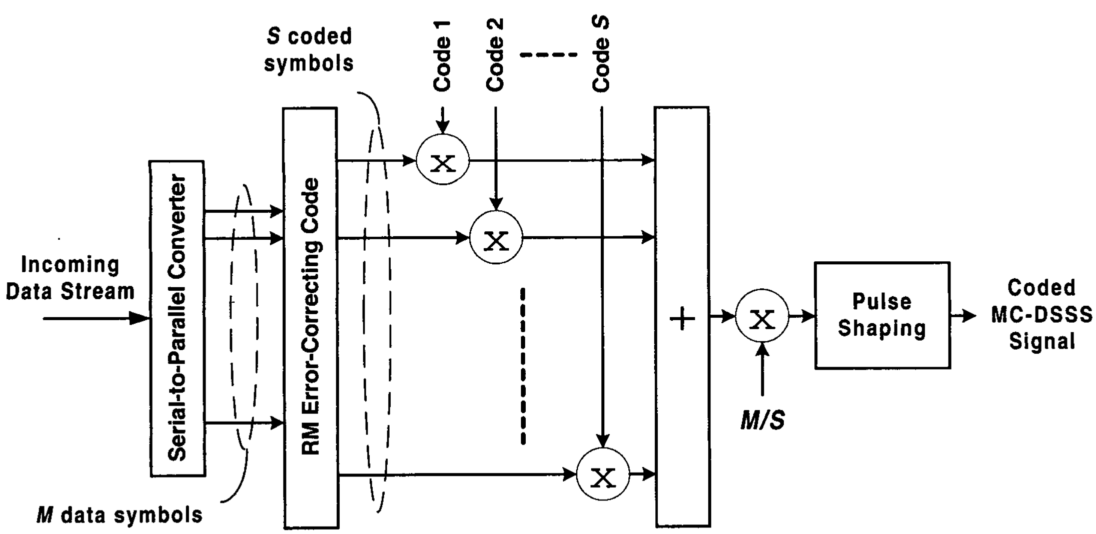
[1] The listed PMEPRs for square-root raised cosine and Gaussian pulses are the minimum values that can be achieved. PMEPRs for these pulses vary with the signal bandwidths [11].

Fig. 1. Signal generation for (a) the conventional MC-DSSS technique, and (b) the coded MC-DSSS technique of [3].

(a)



(b)



design the set of S orthogonal spreading sequences to be OFDM-like. It follows that [3]

$$a_{n,m} = \omega^{m\lfloor n/\kappa \rfloor} c_n, \quad m = 0,1,\cdots,S-1 \text{ and } n = 0,1,\cdots,N-1, \tag{5}$$

where: $\omega = e^{j2\pi/S}$; $\kappa = N/S$; $\lfloor x \rfloor$ is the largest integer less than or equal to x; and $c_0 c_1 \cdots c_{N-1}$ is a pseudo-noise sequence satisfying $|c_n| = 1$ for $n = 0, 1, \cdots, N-1$. Note that that κ is required to be an integer. This requirement is usually satisfied in practice since N is usually chosen to be a power of 2.

III. SYSTEM MODEL

We consider Nyquist pulses so that

$\int_{-\infty}^{\infty} \psi(t)\psi^*(t+iT_c)dt = 0$ for any non-zero integer i.[2] It is assumed that $s(t)$ is transmitted through L independent frequency-nonselective fading channels having identical statistical properties and unity gains. At the receiver, Lth-order diversity reception is used and coherent detection is considered. For the ℓth branch, $\ell \in \{0, 1, \cdots, L-1\}$, the complex envelope of the received signal, $r_\ell(t)$, is given by

$$r_\ell(t) = \alpha_\ell \cdot s(t) + \eta_\ell(t) \tag{6}$$

where $\eta_\ell(t)$ is the complex envelope of the AWGN with a two-sided power spectral density $N_0/2$, and α_ℓ is the channel gain modeled by a random variable with $E\{\alpha_\ell^2\} = 1$. Since the phase shift introduced by the channel is finally compensated at the coherent detector, we do not model its effect in (6).

The received signal $r_\ell(t)$ is processed by S matched filters each of which is matched to a spectral spreading waveform. The matched-filter outputs obtained at the ith sampling instant and for the ℓth branch are given by

$$\begin{aligned}\phi_{i,m,\ell} &= \frac{1}{\sqrt{2E_sR_cT_s}}\int_{-\infty}^{\infty} r_\ell(t)\mu_m^*(t-iT_s)dt \\ &= \alpha_\ell \cdot \tilde{I}_{i,m} + \eta_{i,m,\ell}, \qquad m = 0,1,\cdots,S-1\end{aligned} \tag{7}$$

where $\eta_{i,m,\ell}$ is a zero-mean complex-valued Gaussian random variable with $E\{|\eta_{i,m,\ell}|^2\} = (R_cE_s/N_0)^{-1}$. Note that $\eta_{i,m,\ell}$'s, $m = 0, 1, \cdots, S-1$, and $\ell = 0, 1, \cdots, L-1$, are statistically independent. We consider maximum ratio combining (MRC) and selection combining (SC). The combiner outputs are given by

$$\Phi_{i,m} = \begin{cases} (\sum_{\ell=0}^{L-1}\alpha_\ell\phi_{i,m,\ell})/(\sum_{\ell=0}^{L-1}\alpha_\ell^2) & \text{for MRC} \\ \phi_{i,m,\ell_{SC}}/\alpha_{\ell_{SC}} & \text{for SC} \end{cases} \qquad m = 0,1,\cdots,S-1 \tag{8}$$

where $\ell_{SC} = \arg\max_{\ell\in\{0,1,\cdots,L-1\}} \alpha_\ell$. Let

$$\beta = \begin{cases} \sum_{\ell=0}^{L-1}\alpha_\ell^2 & \text{for MRC} \\ \alpha_{\ell_{SC}}^2 & \text{for SC} \end{cases}. \tag{9}$$

Substituting (7) into (8) yields

$$\Phi_{i,m} = \tilde{I}_{i,m} + \zeta_{i,m}, \quad m = 0,1,\cdots,S-1 \tag{10}$$

where $\zeta_{i,m}$, conditioned on β, is a zero-mean complex-Gaussian random variable with a variance $E\{|\zeta_{i,m}|^2|\beta\} = (\beta R_cE_s/N_0)^{-1}$. Based on minimizing $\sum_{m=0}^{S-1}|\Phi_{i,m} - \tilde{I}_{i,m}|^2$, a maximum-likelihood detector is used to estimate the M transmitted data symbols.

[2] Although Gaussian pulses have an advantage of achieving lower PMEPRs, these pulses are not considered here because their error performances are very close to those for Nyquist pulses [11].

Table 1. Comparison between BER union bounds and simulated results. ($M = 4$; $Q = 4$)

$\beta E_b/N_0$ (dB)	BER union bound	Simulated BER
0	4.95×10^{-1}	1.21×10^{-1}
1	2.12×10^{-1}	7.42×10^{-2}
2	8.05×10^{-2}	3.92×10^{-2}
3	2.70×10^{-2}	1.72×10^{-2}
4	7.74×10^{-3}	5.97×10^{-3}
5	1.78×10^{-3}	1.55×10^{-3}
6	3.03×10^{-4}	2.78×10^{-4}
7	3.42×10^{-5}	3.17×10^{-5}
8	2.28×10^{-6}	2.19×10^{-6}

IV. OUTAGE PROBABILITY

A. *Coded MC-DSSS systems*

We make use of the uniform error property of the RM code in the derivation of the instantaneous BER (conditioned on channel fading). Suppose that an all-zero word, $\mathbf{u}_i^{(0)}$, is transmitted. Let $\tilde{\mathbf{u}}_i^{(0)}$ be the corresponding block of coded symbols. It is apparent that $M\log_2 Q$ data bits are transmitted for each block of M data symbols. The instantaneous BER, P_b, is upper bounded by the union bound:

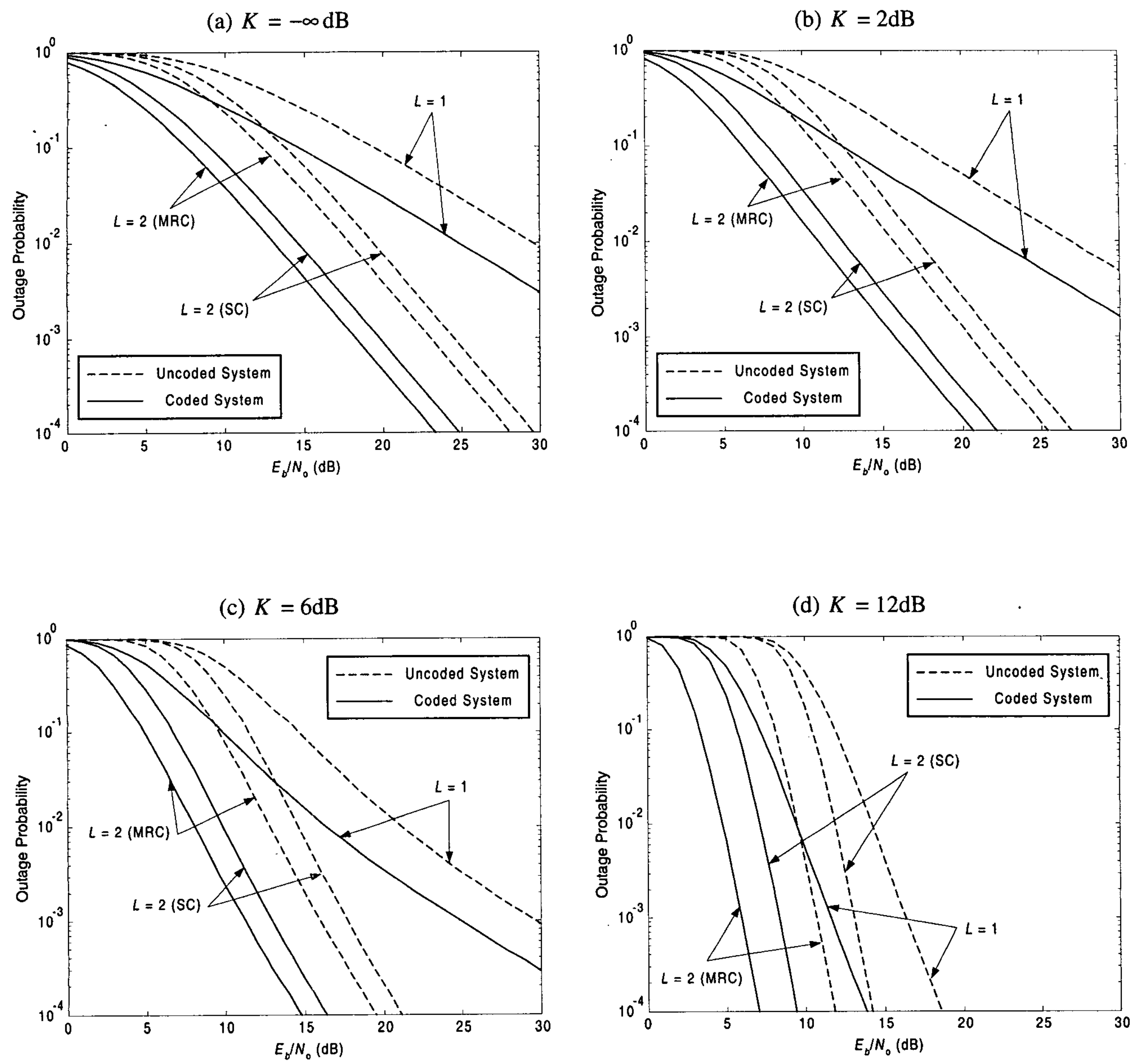
$$P_b \le \sum_{\substack{\mathbf{u}_i\in\{0,1,\cdots,Q-1\}^M \\ \mathbf{u}_i\neq\mathbf{u}_i^{(0)}}} \left\{\frac{1}{M\log_2 Q}\sum_{m=0}^{M\log_2 Q-1}\phi(0,v_{i,m})\right\} \times \Pr(\tilde{\mathbf{u}}_i^{(0)} \to \tilde{\mathbf{u}}_i) \tag{11}$$

where $v_{i,m}$ is the mth data bit after expressing $\mathbf{u}_i$ by its corresponding bit sequence, $\phi(x,y)$ gives a value of 0 if $x = y$ and 1 otherwise, and $\Pr(\tilde{\mathbf{u}}_i^{(0)} \to \tilde{\mathbf{u}}_i)$ is the pairwise error probability. From (10), it can be shown that

$$\Pr(\tilde{\mathbf{u}}_i^{(0)} \to \tilde{\mathbf{u}}_i) = Q\left(\sqrt{\frac{\beta R_c}{2}\cdot\frac{E_s}{N_0}} \times \left\|\tilde{\mathbf{I}}_i^{(0)} - \tilde{\mathbf{I}}_i\right\|\right) \tag{12}$$

where: $\tilde{\mathbf{I}}_i = [\tilde{I}_{i,0}, \tilde{I}_{i,1}, \cdots, \tilde{I}_{i,S-1}]$; $\tilde{\mathbf{I}}_i^{(0)}$ is the $\tilde{\mathbf{I}}_i$ corresponding to $\mathbf{u}_i^{(0)}$; $\|\cdot\|$ denotes the Euclidean distance; and $Q(x) = (2\pi)^{-1/2}\int_x^\infty e^{-t^2/2}dt$ is the standard Q function.

Fig. 2. Outage probability of the coded and uncoded systems for Rician fading channels with Rician factors K of (a) $-\infty$ dB [Rayleigh fading], (b) 2dB, (c) 6dB, and (d) 12dB. Conditions: $L = 1$, $L = 2$ (SC and MRC), $M = 8$, $Q = 4$ and $P_{b,\text{th}} = 10^{-5}$.



The validity of approximating P_b by the union bound is verified by simulation. Table 1 lists the values of the union bound computed by (11) against the BERs obtained through simulation, wherein $M = 4$ and $Q = 4$. A total of 7.36×10^{7} data bits were involved in the simulation. It is apparent that, although union-bound approximations of P_b fall short at low signal-to-noise conditions, the union bounds are close to the simulated BER values when $P_b \leq 10^{-3}$. Therefore, in the computation of the outage probability, we use the union bound to estimate the required

$\beta E_b/N_0$ that achieves a given P_b only if $P_b \le 10^{-3}$. This limitation does not affect the present work because the P_b values we are interested in are less than or equal to 10^{-3}.

The outage probability is derived as follows. Let

$$\Xi(\Omega) = \sum_{\substack{\mathbf{u}_i \in \{0,1,\cdots,Q-1\}^M \\ \mathbf{u}_i \ne \mathbf{u}_i^{(0)}}} \left[\frac{1}{M \log_2 Q} \sum_{m=0}^{M \log_2 Q - 1} \phi(0, v_{i,m}) \right] \times Q\left(\sqrt{\tfrac{1}{2} R_c \Omega \log_2 Q} \times \left\| \tilde{\mathbf{I}}_i^{(0)} - \tilde{\mathbf{I}}_i \right\| \right). \quad \textbf{(13)}$$

It follows that $\Xi(\beta E_b/N_0)$ is the union bound of P_b. Outage occurs when the instantaneous BER exceeds a given value, $P_{b,\text{th}}$. The outage probability, W, is given by

$$W = F_\beta\left(\Omega_{\text{th}} \cdot (E_b/N_0)^{-1} \right) \quad \textbf{(14)}$$

where Ω_{th} is the solution of $\Xi(\Omega_{\text{th}}) = P_{b,\text{th}}$, and $F_\beta(\cdot)$ is the cumulative distribution function (CDF) of β.

B. Conventional (uncoded) MC-DSSS systems

In the characterization of the power-reduction capability of the coded MC-DSSS technique, it is necessary to compute the outage probability of an uncoded system as a reference. Let $P_{b,\text{uc}}$ and $P_{s,\text{uc}}$ be the BER and the symbol error probability, respectively, of an uncoded system and conditioned on the channel fading. For $Q = 2$ and $Q = 4$, we have $P_{b,\text{uc}} = Q\left(\sqrt{2\beta E_b/N_0}\right)$ [12, ch. 5.2.7]. For $Q > 4$, it is known that $P_{b,\text{uc}} \approx P_{s,\text{uc}}/\log_2 Q$, and $P_{s,\text{uc}}$ is lower-bounded by $Q\left(\sqrt{2\beta E_s/N_0}\sin(\pi/Q)\right)$ [13, ch. 3.3.4]. We employ a lower bound here because it yields a $\beta E_s/N_0$ lower than actually is in achieving a given $P_{s,\text{uc}}$. This result is used to compute the power reduction provided by the coded technique so that the resultant value indicates the lowest power reduction that can be obtained. Let

$$\Xi_{\text{uc}}(\Omega) = \begin{cases} \frac{1}{\log_2 Q} Q\left(\sqrt{2\Omega \log_2 Q}\sin(\pi/Q)\right), & Q > 4 \\ Q\left(\sqrt{2\Omega}\right), & Q \in \{2,4\}. \end{cases} \quad \textbf{(15)}$$

After Ω_{th} that satisfies $\Xi_{\text{uc}}(\Omega_{\text{th}}) = P_{b,\text{th}}$ is calculated, the outage probability can be computed by (14) for a given E_b/N_0.

C. Numerical examples

In indoor communications, fading arises due to the motion of personal communication devices (spatial fading) and also due to the movement of personnel in the surrounding (temporal fading). Measurement results of [4] and [14] have shown that spatial and temporal variations can be characterized by Rician distributions with Rician factors in the ranges of, respectively, $-\infty$dB~2dB and 6dB~12dB. Figs. 2a-2d plot the W against E_b/N_0 for Rician fading channels with Rician factors K = $-\infty$dB (Rayleigh fading), 2dB, 6dB and 12dB, wherein the cases $L = 1$ and $L = 2$ (SC and MRC) are considered and we assume that $M = 8$, $Q = 4$ (i.e., QPSK) and $P_{b,\text{th}} = 10^{-5}$. Expressions of F_β used in the computation are given in Appendix I. As expected, the outage performance is improved for a channel with a higher Rician factor K, or in the presence of diversity reception ($L = 2$). Moreover, MRC gives a better performance than SC. It is also apparent that the coded system outperforms the uncoded one. Notice that the dB difference of E_b/N_0 between the coded and the uncoded systems is nearly the same for any outage probability W, and is independent of the combining scheme, the diversity order L or the Rician factor K. This issue is discussed in greater detail in Section V.

V. POWER-REDUCTION CAPABILITY

We consider two representative cases, fixed- and adaptive-power transmission, in the analysis of power-reduction capabilities. Power control is not employed in the former case. In the latter one, the transmitted power is adjusted according to the channel condition. A perfect channel knowledge is required so that this scheme corresponds to an implementation with perfect power control.

A. Fixed-power transmission

Given a target outage probability W, let $(E_b/N_0)_{\text{re}}$ denote the required E_b/N_0 ratio that achieves this outage probability. It follows from (14) that

$$(E_b/N_0)_{\text{re}} = \Omega_{\text{th}}/\beta_{\text{th}} \quad \textbf{(16)}$$

where β_{th} satisfies

$$F_\beta(\beta_{\text{th}}) = W. \quad \textbf{(17)}$$

The power reduction in dB due to applying the coded MC-DSSS technique, G_{dB}, is given by

$$G_{\text{dB}} = 10\log_{10}(E_b/N_0)_{\text{re}}^{(\text{uc})} - 10\log_{10}(E_b/N_0)_{\text{re}}^{(\text{c})} \quad \textbf{(18)}$$

in dB

where $(E_b/N_0)_{\text{re}}^{(\text{uc})}$ and $(E_b/N_0)_{\text{re}}^{(\text{c})}$ are the $(E_b/N_0)_{\text{re}}$ values computed by (16) for the uncoded and the coded systems, respectively.

B. Adaptive-power transmission

Outage occurs when the instantaneous value of β given by (9) is less than β_{th}. When β is greater than β_{th}, the channel is in good condition so that it is possible to

Table 2. Values of $\Omega_{\text{th}}^{(c)}$ and $\Omega_{\text{th}}^{(uc)}$ in dB required to achieve $P_{b,\text{th}}$ of (a) 10^{-3}, (b) 10^{-4}, and (c) 10^{-5}. The dB values of power reduction G_{dB}, computed by (21), are also shown.

(a)

M	Q	$\Omega_{\text{th}}^{(c)}$ (dB)	$\Omega_{\text{th}}^{(uc)}$ (dB)	G_{dB} (dB)
4	2	5.24	6.79	1.55
4	4	5.35	6.79	1.44
4	8	8.83	9.34	0.51
6	2	4.26	6.79	2.53
6	4	4.00	6.79	2.79
6	8	7.40	9.34	1.94
8	2	3.68	6.79	3.11
8	4	3.10	6.79	3.69
8	8	6.36	9.34	2.98

(b)

M	Q	$\Omega_{\text{th}}^{(c)}$ (dB)	$\Omega_{\text{th}}^{(uc)}$ (dB)	G_{dB} (dB)
4	2	6.46	8.79	2.33
4	4	6.53	8.79	2.26
4	8	10.04	11.27	1.23
6	2	5.31	8.79	3.48
6	4	5.05	8.79	3.74
6	8	8.53	11.27	2.74
8	2	4.60	8.79	4.19
8	4	4.00	8.79	4.79
8	8	7.45	11.27	3.82

(c)

M	Q	$\Omega_{\text{th}}^{(c)}$ (dB)	$\Omega_{\text{th}}^{(uc)}$ (dB)	G_{dB} (dB)
4	2	7.42	9.58	2.16
4	4	7.48	9.58	2.10
4	8	11.01	12.63	1.62
6	2	6.17	9.58	3.41
6	4	5.94	9.58	3.64
6	8	9.45	12.63	3.18
8	2	5.36	9.58	4.22
8	4	4.83	9.58	4.75
8	8	8.34	12.63	4.29

reduce the transmitted power as long as the BER requirement is satisfied. The outage probability, on the other hand, is not affected. The power-control policy can therefore be formulated as follows: if $\beta \le \beta_{\text{th}}$, the transmitted power is set to a value such that $E_b/N_0 = (E_b/N_0)_{\text{re}}$; otherwise, the transmitted power is reduced in order that the instantaneous BER just satisfies the BER requirement $P_{b,\text{th}}$, i.e., $E_b/N_0 = \Omega_{\text{th}}/\beta$. The resultant average E_b/N_0 value, $\overline{E_b/N_0}$, is given by

$$\overline{E_b/N_0} = W \cdot (E_b/N_0)_{\text{re}} + \int_{\beta_{\text{th}}}^{\infty} \frac{\Omega_{\text{th}}}{\beta'} p_\beta(\beta') d\beta' \qquad \textbf{(19)}$$

where $p_\beta(\cdot)$ is the probability density function (PDF) of β. Hence, the dB power reduction is

$$G_{\text{dB}} = 10\log_{10}(\overline{E_b/N_0}^{(uc)}) - 10\log_{10}(\overline{E_b/N_0}^{(c)}) \qquad \textbf{(20)}$$

in dB

where $\overline{E_b/N_0}^{(\mathrm{uc})}$ and $\overline{E_b/N_0}^{(\mathrm{c})}$, respectively, are the $\overline{E_b/N_0}$ values computed by (19) for the uncoded and the coded systems.

C. Discussion

For both fixed- and adaptive-power transmission, it is easy to show that (18) and (20) can be reduced to the same expression:

$$G_{\mathrm{dB}} = 10\log_{10}\Omega_{\mathrm{th}}^{(\mathrm{uc})} - 10\log_{10}\Omega_{\mathrm{th}}^{(\mathrm{c})} \quad \text{in dB} \tag{21}$$

where $\Omega_{\mathrm{th}}^{(\mathrm{uc})}$ and $\Omega_{\mathrm{th}}^{(\mathrm{c})}$ are the values of Ω_{th} computed for the uncoded and the coded systems, respectively. That is,

$$\Omega_{\mathrm{th}}^{(\mathrm{uc})} = \Xi_{\mathrm{uc}}^{-1}(P_{b,\mathrm{th}}) \tag{22}$$

and

$$\Omega_{\mathrm{th}}^{(\mathrm{c})} = \Xi^{-1}(P_{b,\mathrm{th}}) \tag{23}$$

where Ξ_{uc}^{-1} and Ξ^{-1} are the inverse functions of Ξ_{uc} and Ξ given by (15) and (13), respectively. Since $\Omega_{\mathrm{th}}^{(\mathrm{uc})}$ and $\Omega_{\mathrm{th}}^{(\mathrm{c})}$ depend only on Q, M, and the BER requirement $P_{b,\mathrm{th}}$, (21) implies that the power reduction realized by the coded MC-DSSS technique is independent of (a) the statistical characteristics of the channel fading, (b) the target outage probability, and (c) the presence or absence of power control. It is also noticed that the power reduction does not depend on the selected diversity-combining method.

VI. NUMERICAL RESULTS AND DISCUSSION

We consider BPSK, QPSK and 8-PSK, corresponding to Q = 2, 4 and 8, respectively. The M values are selected to be M = 4, 6 and 8 for illustration purposes. In practical systems, outer error-correcting coding is often used to make the systems more robust. Outer coding can be applied to data symbols before the application of the RM coding. Here, the performance improvement due to outer coding is taken into account by considering different cases of $P_{b,\mathrm{th}}$. Since the target BER of 10^{-5} is required for video transmission, we consider $P_{b,\mathrm{th}} = 10^{-3}$, 10^{-4} and 10^{-5}.

Table 2 lists the dB values of $\Omega_{\mathrm{th}}^{(\mathrm{uc})}$ and $\Omega_{\mathrm{th}}^{(\mathrm{c})}$ computed by (22) and (23), respectively, and the G_{dB} values calculated by (21). We first consider the case of $P_{b,\mathrm{th}} = 10^{-5}$ (Table 2c). It is apparent that a higher M results in a higher power reduction G_{dB}, a result that is consistent with the result of [3] for AWGN channels. This result indicates that it is desirable to use a higher M to realize a greater transmit-power reduction for personal communication devices in indoor video transmission. Notice that greater signal-processing power is required to decode the coded symbols in the case of a higher M but it is done at base stations. For a given M, it is found that the G_{dB} values are comparable for Q = 2, 4 and 8, although a slightly lower G_{dB} is obtained for a higher modulation of Q = 8 (the difference being around 0.5dB). Despite this, all cases of Q yield significant power reduction exceeding 3dB for M = 6 and 8. In particular, a large reduction of 4.22~4.75dB can be realized for M = 8. Next, we consider all cases of $P_{b,\mathrm{th}}$ (Tables 2a-2c). It is apparent that the G_{dB} values for $P_{b,\mathrm{th}} = 10^{-4}$ are comparable to those for $P_{b,\mathrm{th}} = 10^{-5}$ (within 0.2dB in difference for Q = 2 and 4, and 0.5dB for Q = 8). However, larger differences of ~1dB are observed between the cases of $P_{b,\mathrm{th}} = 10^{-3}$ and $P_{b,\mathrm{th}} = 10^{-5}$. The results indicate that it is preferred to maintain the required BER to be less than 10^{-4} in order to more-effectively utilize the power-reduction capability. This finding enables system designers to select appropriate outer-coding methods in designing personal communication systems involving indoor video transmission.

VII. CONCLUSIONS

We have characterized the power-reduction capability of the coded MC-DSSS technique for indoor video transmission. Analytical results have revealed that the amount of power reduction that can be realized is independent of

(a) the fading statistics of indoor wireless channels,
(b) the outage-probability (or service-availability) requirement for video transmission,
(c) the presence or absence of power control in personal communication devices, and
(d) whether diversity reception is used, and which combining scheme is employed.

The realizable power reduction, however, does depend on the target BER requirement. For indoor video transmission, which typically requires a BER of 10^{-5}, numerical results have shown that a large reduction over 3dB in the transmitted power can be obtained. In case outer coding is also used, the BER requirement can be relaxed. Results have indicated that it is preferred to maintain $P_{b,\mathrm{th}}$ to be less than 10^{-4} for more-effective utilization of the power-reduction capability offered by the coded MC-DSSS technique.

APPENDIX I. CDFS OF β FOR RICIAN-FADING CHANNELS

In the following, we consider that $x \geq 0$. Since $E\{\alpha_\ell^2\} = 1$, the PDF of α_ℓ, denoted by $p_{\alpha_\ell}(x)$, is given by [12, eqn. 2-1-141]

$$p_{\alpha_\ell}(x) = 2x(K+1)I_0(2\sqrt{K(K+1)}x)e^{-(K+1)x^2-K} \qquad \textbf{(24)}$$

where K is the Rician factor, and $I_0(\cdot)$ is the modified Bessel function of the first kind and order zero. Note that $\alpha_0, \alpha_1, \cdots, \alpha_{L-1}$ are statistically independent.

MRC) The random variable β given by (9) has a chi-square distribution. It yields [12, eqns. 2-1-138, 2-1-145]

$$F_\beta(x) = \begin{cases} 1 - Q_L\left(\sqrt{2LK}, \sqrt{2(K+1)x}\right), & K > 0 \\ 1 - e^{-x}\sum_{k=0}^{L-1}\frac{1}{k!}x^k, & K = 0 \end{cases} \qquad \textbf{(25)}$$

where $Q_m(\cdot,\cdot)$ is the generalized Marcum's Q function [12, eqn. 2-1-122].

SC) Since $F_\beta(x) = \Pr(\max \alpha_\ell^2 \le x) = \Pr(\alpha_0 \le \sqrt{x}, \alpha_1 \le \sqrt{x}, \cdots, \alpha_{L-1} \le \sqrt{x})$, it follows that [12, eqn. 2-1-142]

$$F_\beta(x) = \begin{cases} \left[1 - Q_1\left(\sqrt{2K}, \sqrt{2(K+1)x}\right)\right]^L, & K > 0 \\ (1-e^{-x})^L, & K = 0. \end{cases} \qquad \textbf{(26)}$$

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