A Differential Chaotic Bit-Interleaved Coded Modulation System Over Multipath Rayleigh Channels

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Abstract—In this paper, a novel differential chaotic bit-interleaved coded modulation (DC-BICM) system is proposed for band-limited transmission. This system combines protograph-based low density parity check codes with constellation-based M -ary differential chaos shift keying (DCSK) modulation by one bitwise interleaving. Bit error rate simulation results show that the system has higher coding gain compared with the constellation-based M -ary DCSK modulation system with the same spectral efficiency over multipath Rayleigh fading channels. At the same time, several simulations and P-EXIT analysis are used to analyze the performance of the proposed system. It is found that there is a lot of room for optimization of the system by comparing decoding thresholds and simulation results. Moreover, the system with only partial channel state information has better performance and lower complexity compared with the traditional bit-interleaved coded modulation (BICM) direct-sequence-spread-spectrum system. As a result, the DC-BICM system is a good candidate for band-limited transmission.

Index Terms—Constellation-based M -ary DCSK, BICM, protograph-based LDPC, multipath Rayleigh fading channel, band-limited transmission.

I. INTRODUCTION

CHAOTIC modulation has an excellent anti-interference ability with some unique properties, such as non-periodic signals, low correlation, noise-like, deterministic and wide-band. Differential chaos shift keying (DCSK) [1], [2] is proposed to overcome the synchronization problem in chaotic communication systems, which can offer excellent performance over multipath fading channels [3], [4].

However, DCSK has low data rate and bandwidth efficiency because one half of the bit duration is spent on sending non-information-bearing reference samples. Several different methods have been proposed to solve this problem. M -ary DCSK [5], which employs orthogonal Walsh code sequences

to transmit multi-bits, provides an optimal solution over an AWGN channel with non-coherent detection of chaotic signals by employing the Generalized Maximum Likelihood (GML) decision rule [6], and its improved scheme named MCS-DCSK [7] has lower complexity and higher bandwidth efficiency. Recently, a new M -ary OM-DCSK modulation using both T-R technique and M -ary orthogonal modulation was proposed [8], which has outstanding performances. Differing from these methods, generalized constellation-based M -ary DCSK modulation proposed in [9], as an extension of QCSK [10], the circle-constellation-based M -ary DCSK system is appropriate for band-limited communication systems similarly to the M -ary PSK-DCSK [11], providing better performances than PSK-DCSK. However, its performance decreases with the increasing level M [12]. The square-constellation-based (s-c-b) M -ary DCSK communication system has higher bandwidth-efficiency and peak-to-average-power ratio (PAPR) performance than the circle-constellation-based M -DCSK system [9].

As is well known, the reliability and low power consumption are also very important measures for the performance of wireless communication systems, and the error correction coding can effectively improve reliability and reduce power consumption [13]. It has been proved in [14]–[17] that error control codes can significantly improve the overall performance of chaos-based digital communication systems. However, channel coding can increase system bandwidth by producing redundancy, which is not good for band-limited transmission.

In order to achieve a good trade-off between spectral efficiency and power consumption for transmission systems, coded modulation is the best choice, which can achieve high coding gains without requiring extra bandwidths. BICM, firstly proposed in [19] and later analyzed in [20], is most popular because of its simplicity, good energy efficiency and spectrum efficiency over fading channels [21]. BICM can obtain maximum code-word dispersion degree by combining binary error-control codes and higher-order modulation with a bit-wise interleaving in an appropriate soft-decision bit metric. P-LDPC codes were applied to the BICM system over the Gaussian channel with a mapping scheme called variable degree matched mapping (VDMM) in [22], for which some optimized designs of the scheme were carried out in [23] and [24]. But, very few studies have been performed over multipath fading channels.

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Fig. 1. The block diagram of the DC-BICM scheme.

Therefore, a differential chaotic bit-interleaved coded modulation system is proposed in this paper, aiming at improving coding gain without requiring extra bandwidths over multipath Rayleigh fading channels for band-limited transmission. P-LDPC is chosen as the error correction code of this system for its outstanding advantages [18], [29], which has to obtain the channel state information (CSI) because of using belief propagation (BP) iterative decoding algorithm [30]. But this system only needs to know partial CSI by using constellation-based M -ary DCSK modulation, which adopts non-coherent reception. Its design of two-dimensional constellation complies requires for joint coded modulation technique compared with other M -ary DCSK systems. The squared-constellation-based M -ary is selected as an example to verify the feasibility of the proposed system. The proposed system offers good anti-multipath fading performance and high spectrum efficiency, and has lower complexity and better robustness compared with the traditional BICM-DS/SS system [25], [31]. Therefore, it is suitable for band-limited channel scenarios with strong anti-interference ability when it is hard to obtain perfect CSI.

The contributions of this work can be summarized as follows:

- 1) The DC-BICM system is designed. The bit error rate (BER) curves demonstrate that the proposed system has higher coding gains compared with the squared-constellation-based M -ary DCSK system with the same spectral efficiency over three-path fading channel, from which it can be concluded that the system has high spectral efficiency and strong ability to resist multipath interference.
- 2) There exist best spreading factor and best decoding iterations for the system performance according to the simulation results, and the performance gets better with the increasing code length, but, getting worse with increasing M . Also, the gap between simulation results and decoding thresholds signifies

that there is a large room for optimization of the system.

- 3) The comparison between the proposed system and the traditional BICM-DS/SS system over multipath fading channels shows that the proposed system has better performance with partial CSI, which indicates that the proposed system has stronger robustness.

The remainder of this paper is organized as follows. Section II describes the system model and related notions. LLR calculation and P-EXIT analysis are presented in Section III. The simulation results and analysis are given in Section IV. Finally, Section V concludes the investigation.

II. SYSTEM MODEL AND RELATED NOTIONS

In this section, the proposed DC-BICM communication system shown in Fig.1 is described in details. Here, square constellation is adopted as instance.

A. Transmitter

1) *Squared-Constellation-Based M -ary DCSK*: The squared-constellation-based M -ary DCSK is based on the principle of square constellation, where the 4-ary DCSK modulation is same as the QCSK modulation, and the constellation of the squared-constellation-based M -ary DCSK is shown in Fig.2.

2) *Encoder*: Accumulate repeat jagged accumulate (ARJA) codes [18] are applied to the proposed system. There are two ARJA code families constructed by different ways as shown in Fig.3(a) and Fig.3(b), respectively. Here, n denotes the number of variable nodes, and the number of the not-punched variable nodes is $\tilde{n}$, which must equal to $\log_2 M$ according to the variable degree matched mapping (VDMM) rule [22]. Therefore, appropriate code type and code rate both must be selected according to M .

Assumed that M is equal to 16, ARJA with 1/2 rate shown in Fig.3(a) is selected for coding with the corresponding mapping

Fig. 2. The constellation of squared-constellation-based M -ary DCSK.

rules, namely, the number of not-punched variable nodes is just 4. Bit-wise mutual information (MI) $I(x^{e_i}; y)$ [26] in the constellation needs to be computed. Doing this for the case of gray-labeled squared-constellation-based 16-ary DCSK reveals $I(x^{e_0}; y) = I(x^{e_1}; y)$ and $I(x^{e_2}; y) = I(x^{e_3}; y)$ with $I(x^{e_1}; y) > I(x^{e_3}; y)$. The term water-filling refers to the mapping of variable node degrees in proportion to the bit-plane MI, which still performs well because higher-degree nodes can affect more other nodes than lower-degree nodes although it is not optimal [23].

Firstly, encoder generates code bits. Then, the code bits (c_1, c_2, c_3, c_4) corresponding to variable nodes (V_2, V_4, V_1, V_3) in the ARJA protograph are mapped to the signal labeling $(x^{e_0}, x^{e_1}, x^{e_2}, x^{e_3})$ in the s-c-b 16-ary DCSK constellation according to the water-filling, which is shown in Fig.4.

Secondly, the chaotic generator generates a chaotic reference signal $\mathbf{c}_x = c_{x,1}, c_{x,2}, \dots, c_{x,\beta}$ from the map

$$x_{k+1} = 1 - 2x_k^2. \quad (1)$$

which has a length of β , where β is defined as the spreading factor. For a normalized chaotic signal with zero mean, the variance is equal to one. Then, $\mathbf{c}_x$ is transformed into another quadrature signal $\mathbf{c}_y = c_{y,1}, c_{y,2}, \dots, c_{y,\beta}$ by using Hilbert transform, thus two independent orthogonal chaotic signals are generated. The information-bearing signal can be generated as a linear combination of the quadrature signals $\mathbf{c}_x$ and $\mathbf{c}_y$, in the form of $\mathbf{m}_s = a_s \mathbf{c}_x + b_s \mathbf{c}_y$, where the index s identifies the symbol with a signal labeling X in the signal space, $s = 1, 2, \dots, M$, and a_s and b_s presents the real part and the image part corresponding to this symbols, respectively (see Fig.4 for the squared-constellation-based 16-ary DCSK). During the i -th symbol duration, the output of the transmitter is

$$s(t) = x_r(t) \cos(2\pi f_1 t + \phi_1) + x_i(t) \cos(2\pi f_2 t + \phi_2), \quad (2)$$

$$x_r(t) = \sqrt{\frac{E_s}{2}} \sum_{k=1}^{\beta} c_{x,k} p(t - kT_c),$$

$$x_i(t) = \sqrt{\frac{E_s}{2}} \sum_{k=1}^{\beta} m_{x,k} p(t - kT_c), \quad (3)$$

where $p(t)$ is the square-root-raised-cosine filter, T_c is the chip time, $x_r(t)$ and $x_i(t)$ represent the chaotic reference signal and the information signals, respectively, and ϕ_1 and ϕ_2 denote the phase angles of the carrier modulation in which the two modulated carriers are orthogonal, f_1 and f_2 denote the frequencies of modulated carriers, and E_s denotes the symbol energy equaling 1. In the case of a square constellation with $M = 4, 16, 64, 256, \dots$ and amplitudes of $\pm d, \pm 3d, \pm 5d, \dots, \pm(\sqrt{M} - 1)d$ on both directions, when considering the reference signal, the average symbol energy in the constellation is computed by

$$E_{avg} = \frac{E_s}{2M} \times \frac{2M(M-1)}{3} d^2 + \frac{E_s}{2}. \quad (4)$$

B. The Receiver

The receiver signals is given by

$$r(t) = \sum_{l=1}^L \alpha_l \delta(t - \tau_l) \otimes s(t) + \eta(t), \quad (5)$$

where, $\otimes$ is the convolution operator, $\eta(t)$ is an AWGN with zero mean and power spectral density $N_0/2$, L is the number of paths, and α_l and τ_l are the channel coefficient and the path delay of the l -th path, respectively.

In this channel model, the Rayleigh probability density function (PDF) of the channel coefficient α_l is given by

$$f(x) = \frac{\alpha_l}{\sigma^2} \exp\left(-\frac{\alpha_l^2}{2\sigma^2}\right), \alpha_l > 0. \quad (6)$$

where $\sigma > 0$ is the scale parameter of the distribution. A three-path Rayleigh fading channel with $E[\alpha_0^2] = \frac{1}{2}$, $E[\alpha_1^2] = \frac{2}{5}$, $E[\alpha_2^2] = \frac{1}{10}$, $\tau_0 = 0$, $\tau_1 = 2$, $\tau_3 = 5$ is applied.

The received signal $r(t)$ is first separated by two corresponding modulated carrier frequencies. The separated signals are filtered by two corresponding matched filters. Then, the two filtered signals are sampled every kT_c time unit. This way, the desired signals are demodulated, $\tilde{\mathbf{c}}_x$ and $\tilde{\mathbf{m}}_s$ denote chaotic reference signal and informational signal, respectively. Then, $\tilde{\mathbf{c}}_x$ is transformed into $\tilde{\mathbf{c}}_y$ by the Hilbert transform. Here, the received symbol $\mathbf{z} = (za, zb)$, with

$$za \approx \sum_{l=1}^L \alpha_l^2 a_s E_s / 2 + \sum_{k=1}^{\beta} \sum_{l=1}^L \alpha_l \sqrt{E_s / 2} c_{x,k-\tau_l} \eta_{k+\beta}$$

$$+ \sum_{k=1}^{\beta} \sum_{l=1}^L \alpha_l \sqrt{E_s / 2} (a_s c_{x,k-\tau_l} + b_s c_{y,k-\tau_l}) \eta_k$$

$$+ \sum_{k=1}^{\beta} \eta_k \eta_{\beta+k}, \quad (7)$$

Fig. 3. ARJA codes.

Fig. 4. An example of the VDMM mapper approach for the DC-BICM system with ARJA rate 0.5 and M equal to 16.

$$\begin{aligned}
 zb \approx & \sum_{l=1}^L \alpha_l^2 b_s E_s / 2 + \sum_{k=1}^{\beta} \sum_{l=1}^L \alpha_l \sqrt{E_s / 2} c_{x,k-\tau_l} \eta_{k+\beta} \\
 & + \sum_{k=1}^{\beta} \sum_{l=1}^L \alpha_l \sqrt{E_s / 2} (a_s c_{x,k-\tau_l} + b_s c_{y,k-\tau_l}) \tilde{\eta}_k \\
 & + \sum_{k=1}^{\beta} \tilde{\eta}_k \eta_{\beta+k}.
 \end{aligned} \quad (8)$$

III. LLR COMPUTATION AND P-EXIT ANALYSIS

In this section, LLR computation is carried out, and a P-EXIT analysis is introduced to evaluate the decoding thresholds and to analyze the system.

A. LLR Computation

Assume that the largest multipath delay is much shorter than the symbol duration, i.e., $0 \leq \tau_{l_{max}} \leq \beta$, hence the

inter-symbol interference (ISI) is negligible. It was found [9] that the decision variables z_a and z_b are the Gauss variables with the means $m_1 = \sum_{l=1}^L \alpha_l^2 a_s \frac{E_s}{2}$ and $m_2 = \sum_{l=1}^L \alpha_l^2 b_s \frac{E_s}{2}$, respectively, and the same variance $\sigma^2 = \sum_{l=1}^L \alpha_l^2 \frac{E(a_s^2 + b_s^2 + 1) E_s N_0}{4}$. Here, N_0 is the standard deviation of Gaussian noise. Defining $E(a_s^2 + b_s^2 + 1) \frac{E_s}{2} \approx E_{avg}$. Over the multipath Rayleigh fading channel, the decision variables $z' = z / (h \cdot \frac{E_s}{2})$, $h = \sum_{l=1}^L \alpha_l^2$ denote the partial channel state information. If h is estimated, $h = h'$. Then, the means $\overline{m_1} = a_s$, $\overline{m_2} = b_s$ and, the variance $\sigma^2 = (h \times \frac{E_{avg} N_0}{2} + \frac{\beta N_0^2}{4}) / (h \cdot \frac{E_s}{2})^2$. Therefore, for the given symbol (a_s, b_s) , the probability density function of the observation signal $z' = (z'_a, z'_b)$ is given by

$$p(z'|s) = \frac{1}{2\pi\sigma^2} \exp\left(-\frac{(z'_a - \overline{m_1})^2 + (z'_b - \overline{m_2})^2}{2\sigma^2}\right). \quad (9)$$

Fig. 5. Block diagram of the P-EXIT algorithm.

Here, Gaussian approximation (GA) method is used to compute the LLR for the i -th transmitted bit, via

$$LLR[i] = \log \frac{\sum_{m:c_i(m)=0} P(z'|s_m) p(s_m)}{\sum_{m:c_i(m)=1} P(z'|s_m) p(s_m)}, \quad (10)$$

where $c_i(m)$ returns the i -th bit associated with the m -th modulated symbol s_m according to the VDM mapping rule, $m = 0, 1, \dots, M-1$, and $p(s_m)$ is a prior probability of symbol s_m [15], [16], but, in this paper, $p(s_m)/p(s_{\bar{m}}) = 1$.

B. Channel Estimation

In this paper, the iterative decoding algorithm is adopted. Therefore, it is needed to acquire channel information. According to [9], LS estimator is applied. The pilot symbol (1,0) is inserted into the information symbols every Q symbol intervals. During the transmission time of Q symbols, the channel parameter is estimated as $h' = z_p/(E_s/2)$, where z_p denotes the real component of the received symbol transmitted from the pilot (1,0).

C. P-EXIT Algorithm

The block diagram of P-EXIT algorithm is shown in Fig.5.

Firstly, the extrinsic mutual information (MI) of the demodulator is calculated for a fixed E_b/N_0 [23], [27]. Initialize the priori MI $I_v^{ij} = 0$ and $I_c^{ij} = 0$ for variable node V_i and check node C_j , respectively, $i = 0, 1, 2, \dots, n$, $j = 0, 1, \dots, m$, where m and n denote the number of check nodes and variable nodes in the protograph, respectively. For the i -th code bit c_i , the extrinsic MI is given by

$$I(c_i; LLR) = \frac{1}{2} \times \sum_{g=0,1} \int_{-\infty}^{+\infty} p_{LLR}(\xi|c_i = g) \times \log_2 \frac{2p_{LLR}(\xi|c_i = g)}{p_{LLR}(\xi|c_i = 0) + p_{LLR}(\xi|c_i = 1)} d\xi. \quad (11)$$

The conditional probability density functions (PDF) of extrinsic LLRs values $p_{LLR}(\xi|c_i = g)$ is estimated by using Monte Carlo simulations [28], where c_i is code bit, $g \in 0, 1$. An ARJA code of information length 4096 and rate 1/2 is applied to the system over the multipath Rayleigh fading channel, where $M = 16$, $\beta = 32$, and E_b/N_0 is 8.5 dB (E_b is the average energy per information bit).

Then, it is needed to collect the LLR values generated by simulations within each frame under the condition of

Fig. 6. Distribution of conditional PDFs of the output LLR values from demodulator with $\beta = 32$ when the code bit c_i corresponding to V_2 is equal to 0 and 1, respectively.

Fig. 7. Performance comparison between rate-1/2 ARJA 16-ary DC-BICM and squared-constellation-based 4-ary DCSK systems with same spectrum efficiency 2bit/symbol.

$c_i = 0, 1, i = 0, 1, \dots, len/5$, c_i is the code bit corresponding to V_2 , len is the code length, and plot the conditional probability density functions (PDFs) $p_{LLR}(\xi|c_i = 0)$ and $p_{LLR}(\xi|c_i = 1)$. From Fig.6, it can be seen that $p_{LLR}(\xi|c_i = 0)$ and $p_{LLR}(\xi|c_i = 1)$ curves approximately obey Gaussian distribution with $N(u_0, 2u_0)$ and $N(u_1, 2u_1)$, respectively. $u_0 = E(LLR|c_i = 0)$ and $u_1 = E(LLR|c_i = 1)$.

Simulations have also been performed for c_i corresponding to different variable nodes V_i , $i = 1, 2, 3, 4$, and same results are obtained. Moreover, it was found by performing simulations on demodulator that the LLR values within each coded frame follow Gaussian distributions. However, the PDFs do not obey Gaussian distribution according to simulations when β is not enough large, because the inter-symbol interference (ISI) cannot be ignored at this time. It can be obtained

(a) different code lengths.

(b) different decoding iterations.

Fig. 8. Influences of the main parameters on the performance of the DC-BICM system.

from [16] and [28] that $p_{LLR}(\xi|V_i)$ follows Gaussian distribution when β is very large, with

$$J(\sigma_i) = I(V_i; LLR) = 1 - \int_{-\infty}^{+\infty} \frac{\exp\left(-\frac{(\xi - \sigma_i^2)^2}{2\sigma_i^2}\right)}{\sqrt{2\pi\sigma_i^2}} d\xi, \quad (12)$$

where $\sigma_i^2 = 2u_i$, u_i is the average value of LLRs in the case of $\mathbf{c} = 0$ or $\mathbf{c} = 1$, and the code bit $\mathbf{c}$ is corresponding to V_i .

Secondly, set the MI at the input of variable nodes in line with the permutation pattern of VDM, that is, $I_{ch}^{\pi(i)} = I(V_i; LLR)$, $i = 0, 1, \dots, n$. For V_0 that is punctured, I_{ch}^0 is set to 0.

Thirdly, update the external mutual information from variable nodes to check nodes achieved by

$$I_{vc}^{ij} = J\left(\sqrt{\sum_{t \neq j} b^{i,t} [J^{-1}(I_v^{i,t})]^2 + [J^{-1}(I_{ch}^i)]^2}\right), \quad (13)$$

$$+ (b^{i,j} - 1)[J^{-1}(I_v^{i,j})]^2, \quad (14)$$

where $i = 0, 1, \dots, n$, $j = 0, 1, \dots, m$. If $b^{ij} = 0$, then $I_{vc}^{ij} = 0$, where b^{ij} is the number of edges connecting the check node C_j to the variable node V_i in the protograph. Then, I_{vc}^{ij} is assigned to prior MI of check nodes, namely $I_c^{ij} = I_{vc}^{ij}$.

To this end, update the external mutual information I_{cv} from check nodes to variable nodes given by

$$I_{cv}^{ij} = 1 - J\left(\sqrt{\sum_{t \neq i} b^{t,j} [J^{-1}(1 - I_c^{t,j})]^2 + (b^{i,j} - 1)[J^{-1}(1 - I_c^{i,j})]^2}\right), \quad (15)$$

where $i = 0, 1, \dots, n$, $j = 0, 1, \dots, m$. If $b^{ij} = 0$, then $I_{cv}^{ij} = 0$. Set $I_v^{ij} = I_{cv}^{ij}$.

Finally, evaluate the posteriori probability log-likelihood ratio (APP-LLR) MI according to

$$I_i = J\left(\sqrt{\sum_t b^{i,t} [J^{-1}(I_v^{i,t})]^2 + [J^{-1}(I_{ch}^i)]^2}\right), \quad i = 0, 1, \dots, n. \quad (16)$$

The maximum number of iterations is set to 100, iterating from the second step until $\mathbf{I}$ equal to 1, and the lowest $(E_b/N_0)^*$ satisfying this criterion is the lowest decoding thresholds.

IV. SIMULATION RESULTS AND DISCUSSIONS

To illustrate the effectiveness and advantages of the proposed DC-BICM system, extensive simulations have been conducted under a multipath Rayleigh fading channel with $E[\alpha_0^2] = \frac{1}{2}$, $E[\alpha_1^2] = \frac{2}{5}$, $E[\alpha_2^2] = \frac{1}{10}$ and $\tau_0 = 0$, $\tau_1 = 2$, $\tau_3 = 5$ as follows.

Firstly, Fig.7 compares the performances between the DC-BICM system (code length is 2500 and decoding iterations are 50) and the uncoded squared-constellation-based 4-ary DCSK system. A pair of curves indicate that DC-BICM system obtains almost 12.5dB coding gain better than the uncoded squared-constellation-based 4-ary DCSK system with the same spectrum efficiency of 2 bit/symbol when BER and spreading factor β are equal to 10^{-4} and 64, respectively. It can be concluded from all of the above information that the system can not only get good performance for anti-multipath fading and also has high bandwidth efficiency.

Secondly, a series of BER curves of the DC-BICM system with different code lengths and 100 decoding iterations are displayed in Fig.8(a). Fig.8(b) presents a set of BER curves of the DC-BICM system with different decoding iterations and 2500 code lengths. It can be seen from Fig.8(a) that the performance of the system is better with increasing code length. It can be concluded that the code length has a great impact on the performance. The curves shown in Fig.8(b) also

Fig. 9. The impact of β on the performance of the rate-1/2 ARJA 16-ary DC-BICM system with different values of E_b/N_0 .

Fig. 10. The performance of the ARJA coded DC-BICM system with different M , where the curves labeled by the circle symbol denote β is 128 and the curves labeled by plus symbol denote β is 64.

indicate that the performance of the system is better with the increasing decoding iterations. However, the performances have not much differences between 50 and 100 iterations. Therefore, a conclusion can be drawn that there exists best iterations.

Thirdly, the impact of the spreading factor β on the performance of the proposed system can be seen from Fig.9. When β is smaller than a certain threshold value, the performance gets better and better as its increases, This is because inter-symbol interference can not be ignored when the spreading factor is not very large. It is just opposite when β is larger than this threshold value. This is in accordance with the performance analysis of s-c-b M-ary DCSK in [9]. Therefore, in the proposed system, there exists its own best spreading factor with different values of E_b/N_0 . Moreover, It can be seen

Fig. 11. Comparison between simulation results and threshold values for the DC-BICM system with different spreading factors.

Fig. 12. Block diagram of the BICM DS/SS system.

TABLE I
THRESHOLDS CALCULATED BY THE P-EXIT, THE MAXIMUM
NUMBER OF ITERATION IS SET TO 100

β	32	64	128
$(E_b/N_0)^*$	10.543	11.334	12.302

from Fig.10 that when M is larger, the system performance becomes worse.

Fourthly, the BER versus E_b/N_0 performances for the rate-1/2 ARJA 16-ary DC-BICM system with different spreading factors are depicted in Fig.11, where the maximum number of decoding iterations is set to 50. And, accordingly, the calculated decoding thresholds with different spreading factors according to the P-EXIT analysis are shown in Table.1, and are also plotted in Fig.11. It can be seen that there is a big gap between the simulation results and the analytical values, from which it is found that the system still has a large room for optimization.

Finally, a comparison between DC-BICM and traditional BICM DS/SS systems is given. The block diagram of the BICM DS/SS system is shown in Fig.12, where the

(a) BICM DS/SS system.

(b) DC-BICM system.

Fig. 13. Performance comparison between the BICM DS/SS system and the DC-BICM system with 50 decoding iterations and 2500 code length.

M sequence is selected as the spreading sequence. The BICM DS/SS system chooses the same code and mapping scheme as DC-BICM, the modulation is the 16-ary QAM.

As can be seen from Fig.13, the BICM DS/SS system has better performance with a perfect CSI, but cannot work with partial CSI compared to the proposed system. From system implementation, the BICM DS/SS system has high demands on synchronization, it is necessary to obtain every channel coefficient. However, the DC-BICM system not only can achieve the channel power value, but also has a simple structure, and its performance is not bad with the optimal spreading factor. Therefore, for those scenarios without very accurate CSI and complex equipment, the proposed system is a better choice.

V. CONCLUSION

In this paper, a simple generalized DC-BICM system is designed. The comparisons between the DC-BICM and the squared-constellation-based M -ary DCSK systems with same spectral efficiency verify that the proposed system has high coding gain without requiring additional bandwidths and has excellent performance of anti-multipath fading. At the same time, BER simulations show that there exist the best spreading factor and best iterative decoding value for performance of the proposed system, and the performance becomes better with the increasing code length, but being worse with increasing M . Furthermore, simulation results comparing with the threshold values calculated by P-EXIT analysis demonstrate that the system still has a large room for optimization. Finally, a comparison between the DC-BICM and the BICM DS/SS systems reveals that the proposed system has lower complexity and better performance with partial CSI.

Hence, it is concluded that the proposed system offers a good anti-multipath fading performance, high spectrum efficiency, simplicity and robustness thanks to the inheriting advantages of chaotic modulation and BICM technology, and is

suitable for band-limited transmission scenarios with ambitious performance of anti multi-path attenuation, low complexity and without requiring accurate CSI.

Future work will focus on optimization design of this proposed system and finding more practical applications.

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