Bit-interleaved Rectangular Parity-Check Coded Modulation with Iterative Demodulation In a Two-Node Distributed Array

Xin Li, Tan F. Wong, and John M. Shea
Wireless Information Networking Group
Department of Electrical & Computer Engineering
University of Florida
Gainesville, Florida 32611-6130, USA
Email: lix@ecel.ufl.edu, {twong, jshea}@ece.ufl.edu

Abstract— **In this paper, we propose a network-based distributed antenna array approach, in which a bit-interleaved rectangular parity-check coded modulation with an iterative demodulation scheme is used. Different from conventional multiple-array systems, this distributed array employs a pair of physically separated identical receiving nodes. These nodes receive the transmitted signal through independent channels. Then each node demodulates and decodes the received signal in an iterative manner. At each iteration, the pair of nodes exchange the reliability measure of a small portion of the symbols to obtain spatial diversity. Simulation results show that significant diversity gain can be achieved at much lower traffic cost than maximal ratio combining.**

I. INTRODUCTION

The growing demand of high bit-rate data transmission in wireless systems continues to propel the research of using antenna arrays to increase the capacity of wireless communication systems. A well-known array technique is to combine the received signals from the array elements optimally, e.g., using maximal ratio combining (MRC) [1], to gain spatial diversity at the receiver. This technique offers capacity gains over single-antenna systems under the assumption of independent fading at different antenna elements. However, fading correlation does exist when the elements are not spaced sufficiently far apart in practice, which can significantly reduce the capacity of a multiple-antenna system [2]. However, too large a size of the array may limit its applicability in many practice scenarios.

In a previous work [3] [4], we proposed a network-based distributed antenna array approach to obtain spatial diversity without the need of physical connections between the antenna elements. In this approach the physically separated antennas form a distributed array via reliable network connections. The array nodes are far apart enough that the assumption of independent and identically distributed (i.i.d.) fading at different nodes holds. Each antenna node may perform the receiving and decoding process independently while it can communicate with the other node. For simplicity, connections between the nodes are assumed to be error free. With such a distributed antenna array, spatial diversity can be obtained through collaboration and communication between the pair of nodes in the cluster.

The conventional MRC technique is not desirable for the proposed distributed antenna array system because of the required excessive amount of traffic between the nodes. Our approach is to employ iterative soft-in/soft-out (SISO) decoder at each node to generate soft outputs for data bits, then exchange a small portion of these soft outputs between the nodes. Each node uses the additional information from the other node as *a priori* information and restart the decoding process. With this iterative decoding procedure, we can obtain a diversity gain close to that provided by MRC for BPSK [3] [4].

In this paper, we consider employing coded modulation (CM) in the distributed array system to explore the possibility of obtaining spatial diversity with higher spectral efficiency for bandwidth-constrained wireless channels. Bit-interleaved coded modulation (BICM) is a CM technique able to increase the diversity order of a code to its minimum Hamming distance, thus leading to a performance improvement over fading channels when high-order signal constellations are used [5]. Although developed primarily for fading channel, by using an iterative demodulation (ID) algorithm, the minimum free Euclidean distance degradation of BICM over additive white Gaussian noise (AWGN) channel can be overcome [6]. This iterative decoding approach of BICM is referred to as BICM-ID. Besides, an advantage of BICM-ID is its flexibility in design. Any code with soft-output decoder can be used. In this paper, we employ it with rectangular parity-check code (RPCC) [9] to construct our transmission system. We also propose a distributed decoding strategy suitable for BICM-ID.

The remainder of this paper is organized as follows. In Section II, we present the two-node distributed array system and channel model. In Section III, the BICM iterative demodulation for RPCCs and the design of distributed decoding are described in detail. Following that, Monte Carlo simulation results for different signal constellations are shown in Section IV. Finally, conclusions are given in Section V.

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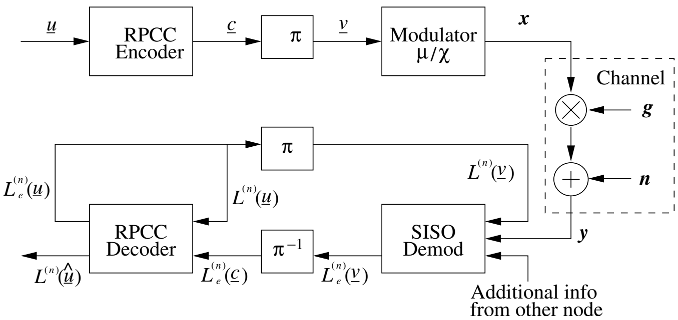

Fig. 1. System model of BICM-ID with RPCC

II. System Model

We consider a simple distributed array system with two identical receiving nodes. A distant transmitter sends a block of modulated signal to the two receiver nodes. The two receiver nodes are physically separated far apart enough that fading at each node is i.i.d.. Each individual node receives and decodes its received signal independently. For simplicity, we assume that the two nodes can communicate with each other reliably. We are only interested in the communication link from the distant transmitter to the two nodes.

The transmitter adopts a typical BICM approach [5], as shown in Fig. 1. A block of data bits $\underline{u}$ to be transmitted are encoded with an RPCC encoder with code rate R_c. Then the coded bit stream $\underline{c}$ are fed into a bit-wise random interleaver π, generating bit stream $\underline{v} = \pi(\underline{c})$. After that, the bit stream $\underline{v}$ is modulated onto a signal sequence $\underline{\boldsymbol{x}}$ over a 2-dimension signal set $\boldsymbol{\chi}$ of size $|\boldsymbol{\chi}| = M = 2^m$ by a M-ary modulator with a one-to-one binary map $\mu : \{0, 1\}^m \rightarrow \boldsymbol{\chi}$. This signal sequence is then sent through the channel. The overall spectral efficiency of this system is mR_c bits/symbol.

Here we use a memoryless fading channel model that includes AWGN channel as a special case. In this model, the received signal $\boldsymbol{y}$ at the two antenna nodes corresponding to the transmitted signal $\boldsymbol{x} \in \boldsymbol{\chi}$ can be expressed as

$$\boldsymbol{y}^{(1)} = \boldsymbol{g}^{(1)}\boldsymbol{x} + \boldsymbol{n}^{(1)},$$
$$\boldsymbol{y}^{(2)} = \boldsymbol{g}^{(2)}\boldsymbol{x} + \boldsymbol{n}^{(2)},$$

where: i) $\boldsymbol{g}^{(1)}$ and $\boldsymbol{g}^{(2)}$ are channel fading gains. For AWGN channels, $\boldsymbol{g}^{(1)} = \boldsymbol{g}^{(2)} = 1$. For Rayleigh fading channels, $\boldsymbol{g}^{(1)}$ and $\boldsymbol{g}^{(2)}$ are independent circular-symmetric complex Gaussian random variables with $\mathrm{E}[\boldsymbol{g}^{(i)}] = 0$ and $\mathrm{E}[|\boldsymbol{g}^{(i)}|^2] = 1$ for $i = 1, 2$; ii) $\boldsymbol{n}^{(1)}$ and $\boldsymbol{n}^{(2)}$ are independent zero-mean, circular-symmetric complex additive Gaussian noise with covariance $\mathrm{E}[|\boldsymbol{n}^{(i)}|^2] = \sigma^2$ for $i = 1, 2$. We normalize the signal energy $\mathrm{E}[|\boldsymbol{x}|^2] = 1$. Thus, the average signal-to-noise ratio (SNR) is $1/\sigma^2$. In this channel model, we assume that perfect channel state information (CSI) $(\boldsymbol{g}^{(1)}, \boldsymbol{g}^{(2)})$ is available at the receiver nodes and hence coherent demodulation is performed at each node. With this model the pdf $p(\boldsymbol{y}^{(i)}|\boldsymbol{x})$, for $i = 1, 2$, with perfect CSI is given by

$$p(\boldsymbol{y}^{(i)}|\boldsymbol{x}) = \frac{1}{\pi\sigma^2} \exp\left(-|\boldsymbol{y}^{(i)} - \boldsymbol{g}^{(i)}\boldsymbol{x}|^2/\sigma^2\right). \qquad (1)$$

At each receiver node, we treat the modulation and code as two components of a concatenated coding system. By employing a maximum *a posteriori* (MAP) demodulator, we feed the extrinsic information from the RPCC decoder back to the demodulator as the *a priori* information to carry out the demodulation and decoding in an iterative manner. After some iterations, we exchange information for a portion of symbols between the two nodes and restart the demodulation and decoding processes.

III. Distributed Decoding for BICM-ID with Rectangular Parity-Check Code

One important component in our bit-interleaved coded modulation system is the rectangular parity-check code [8]. It consists of single parity-check codes that operate on rows and columns of a square matrix that contains the information bits. RPCCs with large block sizes are very high-rate systematic codes that can be decoded by a low-complexity iterative SISO algorithm. More details of RPCC can be found in [9] [10] [3] and [4]. Another important component in the BICM-ID system is the iterative demodulation module. Based on the idea that performing demodulation and decoding in an iterative manner is a key to improve the performance of BICM [6] [7], we employ the receiver model as illustrated in Fig. 1.

To simplify the iterative decoding process, we first modify the demodulator to work in the log-likelihood ratio (LLR) domain. Suppose that each m-bit vector $\underline{v} = (v_1, v_2, \cdots, v_m)$ from the interleaver are mapped into one signal $\boldsymbol{x}$ out of the 2^m signals in the set $\boldsymbol{\chi}$ by mapping rule μ, i.e., $\boldsymbol{x} = \mu(\underline{v}) \in \boldsymbol{\chi}$, at the modulator, and that the received signal corresponding to $\boldsymbol{x}$ is $\boldsymbol{y}$. Let $\ell^i(\boldsymbol{x})$ denote the ith $(i = 1, 2, \cdots, m)$ bit of the label of $\boldsymbol{x}$. For convenience, we assume that $\ell^i(\boldsymbol{x}) = b$ is in the GF(2) with the elements $\{+1, -1\}$. In our soft demodulator, we will consider the MAP rather than maximum-likelihood (ML) bit metric. It is easy to see that the MAP bit metric of $v_i = b \in \{+1, -1\}$ is given by

$$\begin{aligned}\lambda(v_i = b, \boldsymbol{y}) &= \log P(v_i = b, \boldsymbol{y}) \\ &= \log \sum_{\boldsymbol{z} \in \boldsymbol{\chi}} p(\boldsymbol{y}|\boldsymbol{z}) P(\boldsymbol{z}|v_i = b) P(v_i = b), \qquad (2)\end{aligned}$$

where $p(\boldsymbol{y}|\boldsymbol{z})$ is given in (1) according to our channel model. We assume a perfect bit-interleaver π such that $\{v_1, v_2, \cdots, v_m\}$ are independent to each other. With this assumption, we have

$$P(\boldsymbol{z}) = P(\boldsymbol{z} = \mu(v_1, v_2, \cdots, v_m)) = \prod_{j=1}^{m} P(v_j = \ell^j(\boldsymbol{z})). \quad (3)$$

Hence, the MAP bit metric can be simplified to

$$\begin{aligned}\lambda(v_i = b, \boldsymbol{y}) \approx \max_{\boldsymbol{z} \in \boldsymbol{\chi}_i^b} \Big\{ &\log p(\boldsymbol{y}|\boldsymbol{z}) + \sum_{j \neq i} \log P(v_j = \ell^j(\boldsymbol{z})) \\ &+ \log P(v_i = b) + C \Big\}, \qquad (4)\end{aligned}$$

where $\boldsymbol{\chi}_i^b$ denotes the subset of all signal $\boldsymbol{z} \in \boldsymbol{\chi}$ with $\ell^i(\boldsymbol{z}) = b$, and C is a constant. Above, the approximation

$\log(\sum_i a_i) \approx \max_i(\log a_i)$ is used. For convenience we choose the constant as

$$C = -\frac{1}{2}\sum_{j=1}^{m}\big(\log P(v_j=+1)+\log P(v_j=-1)\big). \quad (5)$$

Then the metric becomes

$$\lambda(v_i=b,\boldsymbol{y}) = \max_{\boldsymbol{z}\in\boldsymbol{\chi}_i^b}\big\{\log p(\boldsymbol{y}|\boldsymbol{z}) + \frac{1}{2}\sum_{j=1}^{m}\ell^j(\boldsymbol{z})L(v_j)\big\}, \quad (6)$$

where $L(v_j) = \log\big(P(v_j=+1)/P(v_j=-1)\big)$ is the *a priori* LLR of bit v_j. Thus the soft value of bit v_i in LLR form is computed by

$$\begin{aligned} L(v_i|\boldsymbol{y}) &= L(v_i,\boldsymbol{y}) = \lambda(v_i=+1,\boldsymbol{y}) - \lambda(v_i=-1,\boldsymbol{y}) \\ &= L(v_i) + \max_{\boldsymbol{z}\in\boldsymbol{\chi}_i^{+1}}\big\{\log p(\boldsymbol{y}|\boldsymbol{z}) + \frac{1}{2}\sum_{j\neq i}\ell^j(\boldsymbol{z})L(v_j)\big\} \\ &\quad - \max_{\boldsymbol{z}\in\boldsymbol{\chi}_i^{-1}}\big\{\log p(\boldsymbol{y}|\boldsymbol{z}) + \frac{1}{2}\sum_{j\neq i}\ell^j(\boldsymbol{z})L(v_j)\big\}. \quad (7) \end{aligned}$$

Subtracting the *a priori* LLR of v_i, $L(v_i)$, from (7) we can obtain the extrinsic information of v_i

$$\begin{aligned} L_e(v_i) &= \max_{\boldsymbol{z}\in\boldsymbol{\chi}_i^{+1}}\big\{\log p(\boldsymbol{y}|\boldsymbol{z}) + \frac{1}{2}\sum_{j\neq i}\ell^j(\boldsymbol{z})L(v_j)\big\} \\ &\quad - \max_{\boldsymbol{z}\in\boldsymbol{\chi}_i^{-1}}\big\{\log p(\boldsymbol{y}|\boldsymbol{z}) + \frac{1}{2}\sum_{j\neq i}\ell^j(\boldsymbol{z})L(v_j)\big\}. \quad (8) \end{aligned}$$

We treat this extrinsic information as the output of the soft demodulator. From (8), we can see that in order to obtain the extrinsic LLR of a bit of a signal, we need to use the *a priori* LLRs of the other $m-1$ bits and the channel observation of the signal as input.

With the modification above, the demodulation and decoding procedure can perform in an iterative way conveniently. In Fig. 1 we use $L^{(n)}(\cdot)$ to denote the LLR at the nth iteration. First, we initialize all the *a priori* LLRs $L^{(n)}(\underline{v})$ and $L^{(n)}(\underline{u})$ to zeros for $n=0$. At the nth iteration, when the channel observation $\underline{\boldsymbol{y}}$ of the transmitted signal sequence is received, we demodulate it using (8) to produce $L_e^{(n)}(\underline{v})$. After deinterleaver π^{-1}, $L_e^{(n)}(\underline{c}) = \pi^{-1}(L_e^{(n)}(\underline{v}))$ is fed into the RPCC decoder for decoding. Since the RPCC decoding is an SISO iterative algorithm, we shall use the extrinsic information $L_e^{(n-1)}(\underline{u})$, produced in the $(n-1)$th iteration, as the *a priori* information $L^{(n)}(\underline{u})$ of decoder in the nth iteration. The extrinsic information $L_e^{(n)}(\underline{u})$ generated by the RPCC decoder is then passed thought the interleaver π and fed back as the *a priori* information $L^{(n+1)}(\underline{v})$ for the soft demodulator again. After a number of iterations the estimate of data bits $\hat{u}$ is obtained from the hard decision on $L^{(n)}(\underline{\hat{u}})$.

The mapping μ has a significant effect on the performance of BICM-ID. For BICM, Gray code mapping outperforms set partitioning (SP) mapping [5]. However, when associated with the iterative demodulation, SP mapping outperforms Gray mapping at high SNR [6], [7]. This can be seen from (8) that, due to the property of the Gray mapping that the label of a symbol has only one bit different form its nearest neighbors, the effect of *a priori* LLRs can be weakened significantly. However, this is not the case for SP mapping. Thus the MAP demodulator can make a more effective use of the *a priori* information for SP mapping than for Gray mapping.

The presented BICM-ID scheme is readily applicable to a distributed array. As we pointed out in [3], a decoded data bit with a small soft output magnitude from the RPCC decoder is more likely to be in error. However, if the bit-based strategy in [3] is used here to gain diversity from other receiving node, we will lose the advantage against MRC in term of saving information exchange traffic when the modulation order M increases. Hence, we develop a symbol-based strategy for BICM-ID to reduce the information exchanging traffic. At first, we define the symbol reliability measure out of the decoder as

$$L(\hat{\boldsymbol{x}}) = \log\frac{P(\hat{\boldsymbol{x}})}{1-P(\hat{\boldsymbol{x}})} = \log\frac{P(\hat{\boldsymbol{x}})}{\sum_{\boldsymbol{z}\neq\hat{\boldsymbol{x}}}P(\boldsymbol{z})}, \quad (9)$$

where $\hat{\boldsymbol{x}} = \mu(\hat{v}_1,\cdots,\hat{v}_m)\in\boldsymbol{\chi}$ is the estimate of transmitted signal $\boldsymbol{x}$. For convenience, we define symbol metric for each constellation point $\boldsymbol{z}\in\boldsymbol{\chi}$ as

$$\lambda(\boldsymbol{z}) = \frac{1}{2}\sum_{j=1}^{m}\ell^j(\boldsymbol{z})L(\hat{v}_j), \quad (10)$$

where $L(\hat{v}_j)$ is the soft output of the coded bit v_j. This symbol metric reflects the probability $P(\boldsymbol{x}=\boldsymbol{z})$ given the LLRs $L(\hat{v}_j)$ for $j=1,2,\cdots,m$. In fact, $\hat{\boldsymbol{x}}$ should be the constellation point that has the largest reliability, i.e., $\hat{\boldsymbol{x}} = \arg\max_{\boldsymbol{z}\in\boldsymbol{\chi}}\big\{\lambda(\boldsymbol{z})\big\}$. Similar to (3)-(5), (9) can be simplified to

$$L(\hat{\boldsymbol{x}}) \approx \lambda(\hat{\boldsymbol{x}}) - \max_{\boldsymbol{z}\neq\hat{\boldsymbol{x}},\boldsymbol{z}\in\boldsymbol{\chi}}\big\{\lambda(\boldsymbol{z})\big\} = \min_{j=1,\cdots,m}\big\{|L(\hat{v}_j)|\big\}. \quad (11)$$

Since the LLR magnitude of a bit can be used as the measure of its reliability, (11) indicates that the reliability of a decoded symbol is determined by the soft value of its least reliable bit, which is basically in agreement with the bit-based idea in [3].

With this definition, the distributed decoding procedure works as follows. After every I ($I \geq 1$) iterations of demodulation and decoding, each node computes the symbol reliability $L(\underline{\hat{\boldsymbol{x}}})$ and rank the symbols according to their reliability. Then each node requests additional information from the other node for symbol $\boldsymbol{x}$ that $L(\boldsymbol{x})$ ranks in the lowest $a\%$. We denote the additional information for $\boldsymbol{x}$ as $L_a(\boldsymbol{x})$. Suppose that the estimate corresponding to symbol $\boldsymbol{x}$ at the other node is $\tilde{\boldsymbol{x}} = \mu(\tilde{v}_1,\cdots,\tilde{v}_m)$, which may be different from $\hat{\boldsymbol{x}}$ since the assumption of independent channels. Upon receiving the request, a node sends: i) the reliability of the requested symbols generated in its own decoding process as the additional information, i.e., $L_a(\boldsymbol{x}) = L(\tilde{\boldsymbol{x}})$; ii) the hard decision of $\tilde{\boldsymbol{x}}$, $\ell^j(\tilde{\boldsymbol{x}})$ for $j=1,2,\cdots,m$, which is also the hard decision of $(\tilde{v}_1,\tilde{v}_2,\cdots,\tilde{v}_m)$.

Herein, we adopt following strategy, a node does not request additional information for the symbol if a request has been made for it in all previous exchanges. In this case, the node will request information for the next symbol in the ranking order to make sure that the request for a total of $N \cdot a\%$ symbols will be made for the current exchange, where N is the symbol block size. The advantage of this strategy is that the additional information can cover more symbols for a number of exchanges.

After the exchange, as shown in Fig. 1, each node will use $L_a(\boldsymbol{x})$ and the hard decision $\ell^j(\tilde{\boldsymbol{x}})$ $(j = 1, 2, \cdots, m)$ obtained from the other node to reconstruct an additional symbol metric $\lambda_a(\boldsymbol{z})$ similar to (10) for each possible constellation point $\boldsymbol{z} \in \boldsymbol{\chi}$. Since $\ell^j(\tilde{\boldsymbol{x}})$ is the hard decision of bit $\tilde{v}_j$, we have

$$L(\tilde{v}_j) = \ell^j(\tilde{\boldsymbol{x}})|L(\tilde{v}_j)|. \tag{12}$$

From (11) we can see that $|L(\tilde{v}_j)| \geq L_a(\boldsymbol{x})$. This means each bit in $\tilde{\boldsymbol{x}}$ has as least a reliability of $L_a(\boldsymbol{x})$. Now we replace $|L(\tilde{v}_j)|$ with $L_a(\boldsymbol{x})$ for $j = 1, 2, \cdots, m$ in (12), which is equivalently to set the reliability of all its bits the same as the reliability of a symbol. Thus, we can construct the additional symbol metric as

$$\lambda_a(\boldsymbol{z}) = \frac{1}{2}\sum_{j=1}^{m} \ell^j(\boldsymbol{z})\ell^j(\tilde{\boldsymbol{x}})L_a(\boldsymbol{x}). \tag{13}$$

This additional symbol metric is then used as the *a priori* information for demodulation, and (6) becomes

$$\lambda(v_i{=}b,\boldsymbol{y}){=}\max_{\boldsymbol{z}\in\boldsymbol{\chi}_i^b}\Big\{\log p(\boldsymbol{y}|\boldsymbol{z}){+}\frac{1}{2}\sum_{j=1}^{m}\ell^j(\boldsymbol{z})L(v_j){+}\delta\lambda_a(\boldsymbol{z})\Big\}, \tag{14}$$

where $\delta < 1$ is a scaling factor used to reduce the effect of error propagation. Usually, δ can be set to $0.6 \sim 0.7$.

In the following I iterations, the whole process then repeats with additional exchange of symbol reliability and its hard decision between the two nodes. Note that in this strategy we just need to exchange one real number $L(\tilde{\boldsymbol{x}})$ and m bits $\ell^j(\tilde{\boldsymbol{x}})$ $(j = 1, 2, \cdots, m)$ for each symbol. However, for MRC, one needs to exchange a complex number $\boldsymbol{y}$ (channel observation) and a real number $|\boldsymbol{g}|$ (magnitude of fading gain, for AWGN channel no need to exchange it since $|\boldsymbol{g}| = 1$) for each symbol. This means we just require less than $2/3$ (for Rayleigh fading channel) of or equal (for AWGN channel) to the exchanging traffic of MRC for each symbol, meanwhile we only need to exchange information for a portion of symbols. Hence with this symbol-based strategy, we can reduce the required information exchange traffic significantly.

IV. Simulation Results

In this section, we examine the performance of the proposed distributed array scheme by Monte Carlo simulations. In the simulations, we set the packet size to 1024 data bits , i.e., the data bits are arranged into a 32×32 matrix for the RPCC encoding. With this block size, the RPCC gives a code rate of 0.94. In the decoding procedure, a node requests additional information for 15% of the symbols with the smallest reliability at the beginning of every 10 iterations after the first 10. For instance, 3 exchanges cause an overall traffic approximately equal to 30% (for Rayleigh fading channel) or 45% (for AWGN channel) of what required by MRC.

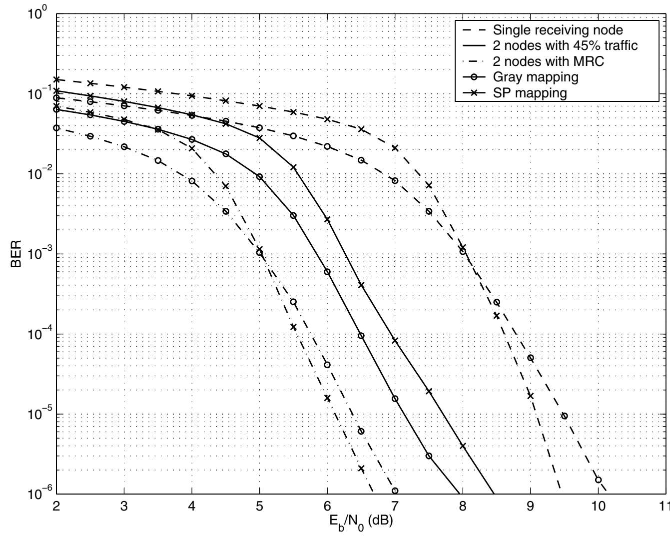

Fig. 2. BER for BICM-ID with 32^2 RPCC and 8PSK in the two-node distributed array over AWGN channel.

In the case of MRC, we assume that each node passes all its channel observations and fading gains to the other node and maximally combines the channel observations before demodulation. Simulations show the bit error rates (BER) at the two nodes are almost the same as each other. So we take the average of them as the performance of the distributed array system.

Fig. 2 shows the BER performance of BICM-ID with 32^2 RPCC in the distributed array over AWGN channels when 8PSK with Gray and SP mapping are used. In the figure, E_b is the received energy per bit per antenna. With MRC, about 3dB spatial diversity gain can be achieved for both mappings. With our distributed array approach, we obtain a 2.4dB and 1.4dB gain for Gray and SP mapping at the traffic cost of 45% (i.e., 3 exchanges in total) of MRC. Fig. 3 shows the BER curves for Rayleigh fading channels. The spatial diversity gain provided by MRC is about 8.5dB for both Gray and SP mapping. By exchanging a total of 20% (i.e., 2 exchanges in total) of the information amount required for MRC, our distributed BICM-ID system obtain a 8.3dB and 7.3dB gain at the BER of 10^{-5} for Gray and SP mapping, respectively.

In Figs. 4 and 5, we show the the average SNR (E_s/N_0) at BER of 10^{-5} versus spectral efficiency for the two-node distributed array system for different constellations with Gray mapping* and SP mapping over AWGN channels and Rayleigh fading channels, respectively. The average SNR can be computed approximately by SNR $= mR_cE_b/N_0$, where m is the number of bits per symbol carrying, and R_c is the code rate of RPCC. We can see that for both AWGN and Rayleigh fading channels the proposed distributed BICM-ID approach can achieve almost the diversity gain provided by MRC, but with only exchanging 20% (2 exchanges in total) and 45%

*For 32QAM, quasi-Gray mapping is used because Gray mapping is impossible in this case.

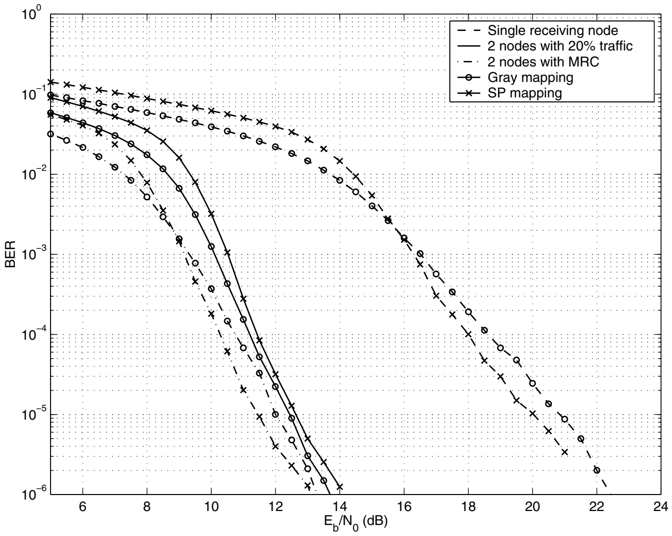

Fig. 3. BER for BICM-ID with 32^2 RPCC and 8PSK in the two-node distributed array over Rayleigh fading channel.

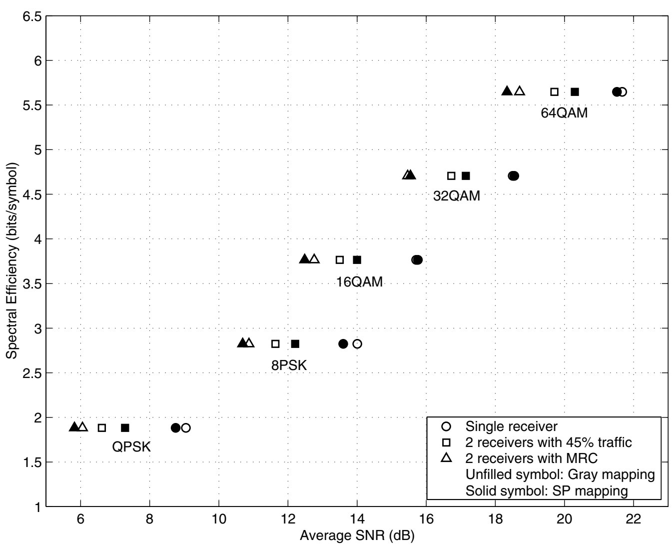

Fig. 4. Average SNR at 10^{-5} BER versus spectral efficiency for BICM-ID with 32^2 RPCC in a two-node distributed array over AWGN channels.

(3 exchanges in total) of the amount of information required by MRC between the two nodes for AWGN channel and Rayleigh fading channel, respectively. This especially shows the advantage of our approach for fading channels.

V. Conclusion

In this paper, we have investigated a network-based distributed array approach, in which high-order constellations with iterative demodulation and RPCCs are used. A significant diversity can be obtained with a relatively small amount of information exchange between the independent and physically separated receiving nodes. This investigation motivates us to consider employing other more powerful codes to replace the RPCC in the distributed array. Another consideration is to

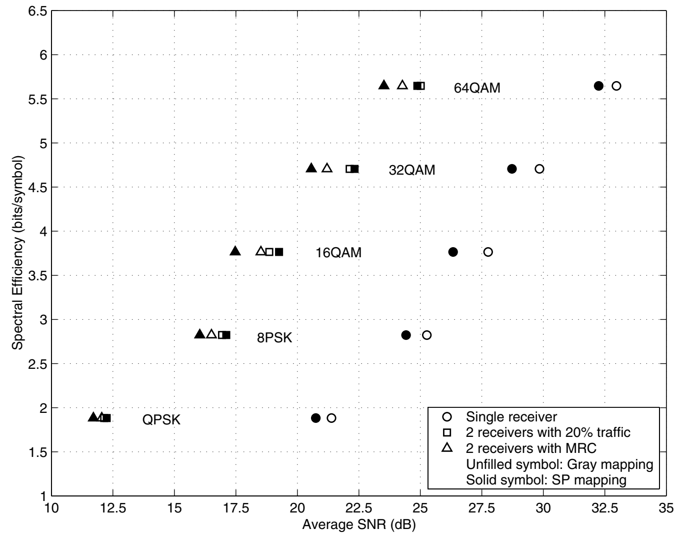

Fig. 5. Average SNR at 10^{-5} BER versus spectral efficiency for BICM-ID with 32^2 RPCC in a two-node distributed array over Rayleigh fading channels.

extend the two-node array to the case of more than two receiving nodes. This will rise many joint network coordination and physical layer decoding issues. These issues will be our future research directions.

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