A Non-Coherent Multi-User Large Scale SIMO System Relying on M-ary DPSK and BICM-ID

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***Abstract*—A new constellation is designed for the multi-user non-coherent Large-Scale single input multiple output (LS-SIMO) uplink system based on M-DPSK, which is combined with a Bit Interleaved Coded Modulation-Iterative Decoding (BICM-ID) scheme for attaining improved performance. We conceive a new approach for constructing EXIT charts parameterized by the number of antennas used for optimizing the design. Finally, the system performance is analyzed and compared to previous work. This evaluation shows an overall improvement of two orders of magnitude demonstrating a dramatic reduction in the required number of antennas.**

***Index Terms*—BICM-ID, EXIT Chart, Large-Scale MIMO, Massive MIMO, Non-coherent detection.**

I. Introduction

Large scale multiple input-multiple output (LS-MIMO) systems have attracted substantial interest in the research community because of their impressive spectral- and energy-efficiency. However, one of the main bottlenecks of LS-MIMO is the acquisition of the Channel State Information (CSI) of numerous channels for coherent detection, where the non-orthogonality of the pilot sequences used in neighboring cells seriously compromises the performance of these systems [1].

To circumvent this impediment in this treatise we focus on non-coherent (NC) detection as a design alternative, for avoiding channel estimation. The design dilemma between coherent and NC communication has intrigued the community since 1991 [2]. NC detection has indeed compelling benefits in terms of dispensing with power-thirsty channel estimation, as also argued in [3]-[4]. Indeed, we will demonstrate in this paper that upon taking into account the pilot-overhead in the throughput calculation and the detrimental effects of realistic imperfect estimation, NC systems are potentially capable of outperforming their coherent counterparts.

A comparison between coherent and NC detection for LS-single input multiple output (LS-SIMO) and single-user systems was presented in [5] which emphasizes that NC schemes can outperform coherent schemes. Another recent design used for NC SIMO is focused on uniquely factorable constellation, not developed for LS though [6]. For single-user MIMO systems, Gohary and Davidson [7] proposed a transmitter and a receiver for NC communication based on Grassmanian Constellation, which works well for frequency flat MIMO channel and for high Signal Noise Ratio (SNR), but cannot exploit differential detection. In [8] an Amplitude Shift Keying (ASK) system was proposed that requires an excessive number of antennas for achieving a reasonable performance. A Differential Quaternary Phase Shift Keying (DQPSK) system was presented in [9], relying on a particular channel that resembles an Impulse Radio-Ultra Wide Band (IR-UWB) system, where the users can be spatially separated based on their non-overlapping power-space profiles. However this channel model cannot be exploited in general.

Against this background our novel contribution is that we use differential detection and separate the signals of multiple users merely relying on the advantages of using an increased number of receive antennas. Some preliminary work was presented in [10], where two constellations were proposed, which outperformed the benchmarker of [8]. In this contribution we extend the system design of [10] to include a coding scheme, which drastically improves the system performance hence reducing the number of antennas required for attaining a given performance. The contribution of [8] has also been extended in [11], exploring the impact of random coding for reducing the number of antennas. Here we propose powerful coding construction based on the principle of bit-interleaved coded modulation and iterative decoding (BICM-ID) [12] for NC LS-SIMO aided DPSK. A maximum likelihood approach for NC-LS-SIMO was presented in [13] using energy-based modulation. In [14] a non-coherent multi-user Massive MIMO was introduced based on an iterative decision at the receiver, while neither [13] nor [14] include channel coding.

The EXIT charts concept proposed in [15] constitutes a powerful tool used for designing and analyzing iteratively decoded systems. A comprehensive tutorial is provided in [16] on the design of systems based on EXIT-curves. We will build on these contributions and apply the EXIT charts for designing our BICM-ID MDPSK LS-SIMO system. Therefore, the novel contribution of this work can be summarized as follows:

- A novel constellation design is proposed for the non-coherent LS SIMO uplink, which -in contrast to [10]-, does not rely on strict power control and yet provides the same performance for all users.
- BICM-ID is intrinsically amalgamated with the non-coherent LS-SIMO system based on M-ary DPSK. We present a novel analysis using the EXIT chart tool in order to analyse the effect of the number of antennas on the system performance, demonstrating that the employment of coding considerably reduces the number of receive antennas required. While other contributions rely on high SNR [14], our system is capable of operating at low SNR.

The rest of the paper is organized as follows. In Section II the system model is presented. Our constellation design is shown in Section III, while the coding scheme is presented in Section IV. The performance is analyzed in Section V. Finally, Section VI presents our conclusions and future work.

II. System Model

We consider a multi-user SIMO uplink scenario, where a single base station (BS) is equipped with R receive antennas (RA) to receive the signals transmitted from K Mobile Stations (MSs), or users, as shown in Fig. 1.

We denote the signal to be transmitted by user j at time instant n as $x_j[n]$ and the signals from all users are grouped

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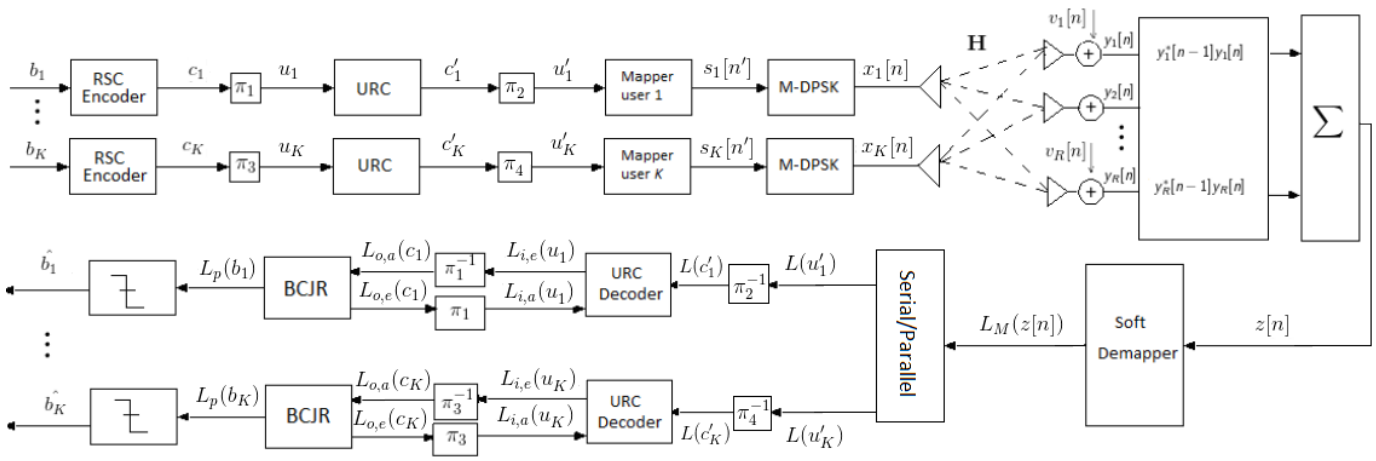


Fig. 1. BICM-ID model for Non Coherent Multi-user Large Scale SIMO system for K users

into the $(K\times 1)$-element vector $\boldsymbol{x}$. Each of the signals $x_j[n]$ is a differentially encoded version of $s_j[n]$ formulated as

$$x_j[n] = x_j[n-1]s_j[n], n>1. \tag{1}$$

As shown in Fig. 1 a block of bits b_j are encoded by a Recursive Systematic Convolutional (RSC) code (outer encoder) generating the bit stream c_j, which is then interleaved by the random bit interleavers π_1 and π_3. The interleaved sequence u_j is then fed through the Unity-Rate Code (URC) encoder (inner encoder), where the coded bits c'_j at the output of the URC are interleaved by the second random bit interleavers π_2 and π_4, producing the bit stream u'_j. The URC encoder is introduced for avoiding the bit error rate (BER) floor because its infinite duration impulse response efficiently randomizes the extrinsic information provided for Bahl-Cocke-Jelinek-Raviv (BCJR) decoder [16]. After bit interleaving, the bits are mapped to symbols $s_j[n]$. These symbols are independent of each other and belong to an M-ary constellation $\mathfrak{M}_j = \{s_{j,m}, m=0,1,...,M-1\}$, which may in fact be different for each user, as we will demonstrate later, where $|s_{j,m}[n]|=1$. Each $x_j[0]$ is a known reference symbol taken from the same constellation.

The propagation channel is represented by the $(R\times K)$-element channel matrix $\boldsymbol{H}$ with the components $h_{i,j}$ modeling the propagation from user j to the i-th antenna of the BS. The coefficients $h_{i,j}\sim CN(0,1)$ account for Rayleigh fading with zero mean and variance 1. The vector $\boldsymbol{y}$ hosts the signal received at each of the BS antennas during each time instant $y_i[n]$. Then the signal received at the BS is formulated as $\boldsymbol{y} = \boldsymbol{Hx}+\boldsymbol{\nu}$, where the dependence on n is dropped for ease of notation. Here $\boldsymbol{\nu}$ is the $(R\times 1)$-element vector of AWGN components with $\nu_i[n]\sim CN(0,\sigma^2)$. We define the reference SNR as follows

$$\rho = \frac{K}{\sigma^2}. \tag{2}$$

Since the power of each individual user is normalized, the total power is $\sum_{j=1}^{K}|s_{j,m}[n]|^2{=}K$, hence ρ depends on the number of users K. For the sake of performing a fair comparison of the schemes with different channel coding rate η we define also the ratio of bit energy to noise spectral density E_b/N_0 as

$$E_b/N_0 = \frac{\rho}{\eta} = \frac{K}{\eta\sigma^2}. \tag{3}$$

At the receiver, the phase difference of two consecutive symbols received at each antenna is non-coherently detected, assuming that the channel stays time-invariant for these two symbols, which is formulated as $h_{ij}[n-1] = h_{ij}[n] = h_{ij}$, $i=1,...,R, j=1,...,K$, and they are all added to give the received symbol $z[n]$

$$z[n] = \frac{1}{R}\sum_{i=1}^{R} y_i^*[n-1]y_i[n], \tag{4}$$

that contains both the information and interference arriving from all users

$$\begin{aligned} z[n] = &\frac{1}{R}\sum_{j=1}^{K}\sum_{i=1}^{R}|h_{ij}|^2 s_j[n] + \frac{1}{R}\sum_{j=1}^{K}\sum_{\substack{k=1\\k\neq j}}^{K}\sum_{i=1}^{R} h_{ij}^* h_{ik} x_j^*[n-1]x_k[n] \\ &+\frac{1}{R}\sum_{i=1}^{R}\nu_i^*[n-1]\sum_{j=1}^{K}h_{ij}x_j[n] + \frac{1}{R}\sum_{i=1}^{R}\nu_i[n]\sum_{j=1}^{K}h_{ij}^* x_j^*[n-1] \\ &+\frac{1}{R}\sum_{i=1}^{R}\nu_i^*[n-1]\nu_i[n]. \end{aligned} \tag{5}$$

From the Law of Large Numbers we know almost surely that [17]

$$\frac{1}{R}\sum_{i=1}^{R}|h_{ij}|^2 \overset{R\to\infty}{=} 1. \tag{6}$$

Then, if we define the joint received symbol as

$$\varsigma[n] = \sum_{j=1}^{K} s_j[n], \tag{7}$$

we have

$$z[n] \overset{R\to\infty}{=} \varsigma[n] + i[n], \tag{8}$$

where $i[n]$ are the noise terms and the interference imposed by all the antennas and users. As shown in [10], the power of the interference $\mathfrak{I} = E\{|i[n]|^2\}$ depends on R and K as

$$\mathfrak{I} = \frac{K^2+2\sigma^2 K+\sigma^4}{R}. \tag{9}$$

At the receiver of Fig.1, the *soft-demapper block* calculates from the symbols $z[n]$ the soft information for each coded bit of the joint symbol ς, which is expressed in form of the *log-likelihood ratio* (LLR_q) for the q^{th} bit. Hence, from $z[n]$ the set $L_M = \{LLR_q\}_{q=0}^{l}$ is obtained where $l = K\log_2 M$ and M is the order of the individual constellation $\mathfrak{M}_j$. The LLRs are defined in [16] and for the joint constellation $\mathfrak{M}$ each LLR can be written as follows

$$LLR_q = \log\frac{P(q=0|z[n])}{P(q=1|z[n])} = \log\frac{\sum_{\varsigma\in\mathfrak{M}_q^{(0)}} p(z[n]|\varsigma)}{\sum_{\varsigma\in\mathfrak{M}_q^{(1)}} p(z[n]|\varsigma)} \quad (10)$$

where $\mathfrak{M}_q^{(b)}$ is the subset of $\mathfrak{M}$ with the q^{th} bit having the value of 0 or 1. Assuming that $p(z[n]|\varsigma)$ obeys the Gaussian distribution, we can use the max-log approximation as follows

$$\begin{aligned} LLR_q &= \log\frac{\max\limits_{\varsigma\in\mathfrak{M}_q^{(0)}} p(z[n]|\varsigma)}{\max\limits_{\varsigma\in\mathfrak{M}_q^{(1)}} p(z[n]|\varsigma)} \\ &= -\frac{1}{\mathfrak{I}}\left(\min_{\varsigma\in\mathfrak{M}_q^{(0)}}|z[n]-\varsigma|^2 - \min_{\varsigma\in\mathfrak{M}_q^{(1)}}|z[n]-\varsigma|^2\right) \end{aligned} \quad (11)$$

where the channel effects are absent due to (7) and (8). Then the LLRs of each user $L(u'_K)$ are obtained from the set of l values in L_M by a serial-parallel conversion. This is done by grouping l/K consecutive values for each user. This way of separating the individual symbols from the joint symbols is due to the labelling used in the mapping of the joint symbol in connection with the choice of each individual user's labelling at the transmitter. The joint labelling is formed by concatenating the individual labellings. For example, if we have K=2 users and M=4, where user 1 transmits the pair of consecutive bits $\{u_1^{'(1)}, u_1^{'(2)}\}$ and the user 2 transmits $\{u_2^{'(1)}, u_2^{'(2)}\}$, then the label of the joint symbol is the serial concatenation formulated as $\{u_1^{'(1)}, u_1^{'(2)}, u_2^{'(1)}, u_2^{'(2)}\}$. Then, when we have the soft-values $L_M =$ "$\underbrace{LLR_1, LLR_2}_{user1}, \underbrace{LLR_3, LLR_4}_{user2}$", the separation for each individual user gives $L(u_1') =$ "LLR_1, LLR_2" and $L(u_2') =$ "LLR_3, LLR_4".

Regarding the joint mapping scheme, it is widely recognized [16] that Gray Mapping (GM) is not advisable for implementing iterative decoding, since it produces poor BER performance. However, there is a large number of legitimate anti-Gray mappings that can be created from the combination of the individual users constellation. Although, again, their impact on attained performance is not significant. Nonetheless, we have opted for the specific mapping in Fig. 2 that provided the best BER for a given number of antennas among of the ones that we analyzed.

Afterwards, the decoding process is performed independently for each user. The URC decoder processes the information provided by the demapper, deinterleaved by π_2, in conjunction with the *a priori* information $L_{i,a}(u_j)$ passed back to it from the RSC decoder, in order to generate the extrinsic LLRs $L_{i,e}(u_j)$. Then, the LLRs $L_{i,e}(u_j)$ are deinterleaved by a soft-bit deinterleaver π_1, as seen in Fig. 1. Next, the soft-bit values $L_{o,a}(c_j)$ are passed to the RSC decoder in order to compute the extrinsic LLRs $L_{o,e}(c_j)$. The RSC decoder invokes the BCJR algorithm [16]. During the last iteration, only the *a posteriori* LLR values, $L_p(b_j)$, of the original uncoded systematic information bits are required, which are passed to a hard decision decoder in order to determine the estimated transmitted source bits $\hat{b}_j$.

III. Design of the input signal constellations

In order to separate the users' signals at the BS, the constellations $\mathfrak{M}_j$ must be specifically designed so that their symbols can still be uniquely and unambiguously distinguished upon superimposing the transmitted signals of all users. More explicitly, the joint constellation $\mathfrak{M} = \{\varsigma_m, m = 0,1,...,\mathfrak{K}-1\}$ of cardinality $\mathfrak{K} = M^K$ is obtained from the superimposed combinations of the individual constellation points $\mathfrak{M}_j$ as $\{s_{1,m^{(1)}} + s_{2,m^{(2)}} + ... + s_{K,m^{(k)}}, m^{(j)} = 0,1,...,M-1\}$. Provided this is accomplished, the individual users' encoded data symbols $s_j[n]$ can be directly obtained from the detected joint symbols $\hat{\varsigma}[n]$ by the soft demapper.

We apply two approaches for designing the constellations and combine them with the channel coding scheme. In the first scheme, we improve the design of the *Unequal Error Protection* (UEP) constellation of [10] with the aid of coding, while in the second scheme, we present a new design to achieve the same performance for all users, hence resulting in *Equal Error Protection* (EEP).

A. Constellation Design for Unequal Error Protection

The design of this constellation was presented in [10]. The symbols of the individual constellations $\mathfrak{M}_j$ are defined as

$$\mathfrak{M}_j^{UEP} = \left\{\frac{2\pi m}{M}L^{1-j}, m = 0,1,...,M-1\right\}, j = 1,...,K, L \geq M. \quad (12)$$

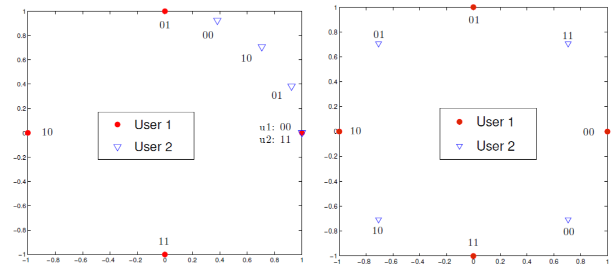
An example of the users' constellation is shown in Fig. 2 (a) for $K = 2$ users and $M = 4$ in conjunction with the mapping for each user.

B. Constellation Design for Equal Error Protection

The constellation for an EEP design is based on placing all symbols of a given user at equal distances and also keeping the same distance between any two symbols of all the users. The role of the minimum constellation distance (MCD) in the error probability was shown in [10]. In order to arrange for these distance properties, the users are intercalated in the unit circle and hence, all users will have the same error performance. Hence, the constellation $\mathfrak{M}_j$ for user j is defined as

$$\mathfrak{M}_j^{EEP} = \left\{\frac{2\pi[(m+1)K-1+j]}{KM}, m = 0,1,...,M-1\right\} j = 1,...,K, \quad (13)$$

In Fig. 2 (b) an example of the user's constellation is shown for $K = 2$ users and $M = 4$ in conjunction with the mapping for each user.



(a) (b)

Fig. 2. Constellations for $K=2$ users and $M=4$ for (a) UEP and (b) EEP.

IV. Coding design for LS Non-Coherent SIMO System based on EXIT Charts

We use the approach of [16] to design and analyse the system using EXIT chart, in order to decide which is the best coding scheme for our system. We use the set of 17 RSC codes presented in [18] as candidates for the outer code in this work. The EXIT chart for the inner decoder (URC and soft demapper) is a function of both the *a priori* mutual information (MI) of the coded bits $\mathbf{u}$, $I_{i,a}(\mathbf{u})$ and the SINR. Based on the interference analysis in [10] and (9), we can express the SINR as a function of ρ and R as

$$SINR = \frac{R}{K}\frac{\rho^2}{\rho^2+2\rho+1}. \tag{14}$$

Then, the demapper's extrinsic information transfer for our particular LS-SIMO system depends on the apriori information and on the SINR, which is formulated as

$$I_{i,e}(\mathbf{u}) = T_i[I_{i,a}(\mathbf{u}), SINR] = T_i[I_{i,a}(\mathbf{u}), R, \rho], \tag{15}$$

We analyse the behavior of the iterative decoder by observing the Monte-Carlo simulations based decoding trajectory. Fig. 3 shows the EXIT chart and trajectory for $K=2$ users and our UEP design. We can verify from Fig. 3 that each user experiences different performance, since they have a different decoding trajectory. Furthermore, the decoding process for user 1 is less complex, since it requires only a single iteration. For our EEP design, we can see from Fig. 4 that the minimum required number of antennas for a 1/2-rate code for the sake of ensuring an open EXIT tunnel facilitating convergence to a vanishingly low BER is $R=90$. We can strike a compelling tradeoff between the number of antennas and the number of iterations (I) necessary so that the probability of error vanishes. For the same number of iterations, increasing R to 350 antennas allows us to increase the coding rate required to 9/10. We decided to use the URC having the octal represented generator and feedback polynomials (G,F)=(2,3) for both designs due to the fact that required value of R is lower for a given coding rate.

V. Performance evaluation

The objective of this analysis is to optimize the value of R required in order to attain a good performance, since R should be a feasible number for a LS-SIMO system. Block fading is assumed, where the channel's envelope remains constant

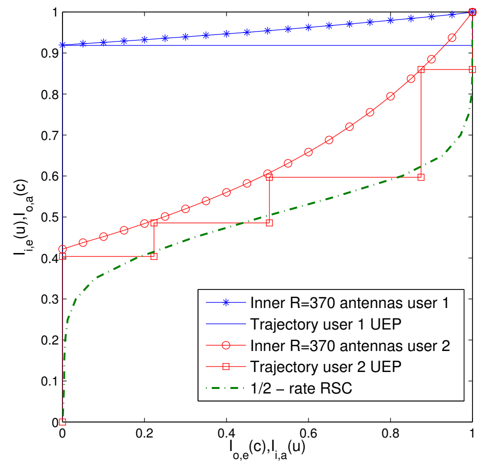


Fig. 3. EXIT Chart for $K=2$ users, $\rho=0$ dB, $M=4$, URC and UEP

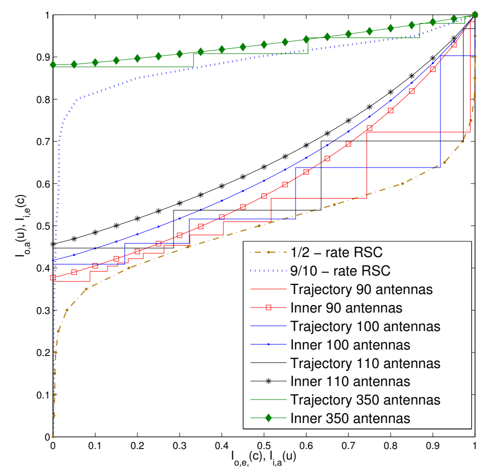


Fig. 4. EXIT Chart for $K=2$ users , $\rho=0$ dB, $M=4$, URC and EEP

during the transmission of a transmission burst *e.g.* 800,000 bits, while they vary randomly between bursts. Explicitly, we use long interleavers for ensuring the Gaussian distribution of the LLRs. Finally, we set $\rho=0$ dB, unless otherwise stated.

Table I provides a comparison of the R required for different outer encoder rates for both the UEP and EEP schemes. For each case, we have chosen the smallest R required for achieving for open EXIT chart tunnel and a low BER. We can see that as expected, for lower coding rate, R is smaller. Note that the results show a considerable reduction in R compared to the previous solutions operating without [10] and with coding [19]. This is important for making LS-SIMO systems practical. For example, at millimeter-waves frequencies a 10x10 array having 100 antennas would require a coding rate of 3/4 at SNR=3dB according to Table I.

For the UEP design we verify the fact that different users have a different performance, as it was shown in Fig. 3. By comparing the UEP and EEP schemes we can see that user 1 now requires a lower value of R while user 2 requires a higher value. Overall, there is an increase in R for the EEP design if we wish to guarantee that both users have the same performance. In any case, R is considerably lower than in [10].

Fig. 5 provides an example of the results shown in Table I, where R = 50 and a code rate of 1/10 are fixed, while varying the E_b/N_0. The difference between EEP and UEP can be explicitly observed. For the second user of the UEP scheme to have the same performance as user relying on the EEP scheme, we have to increase the number of antennas to 250, as shown in Fig. 5. This extra number of antennas compensates for an E_b/N_0 difference of 8 dB, evaluated equivalently for the effective ρ.

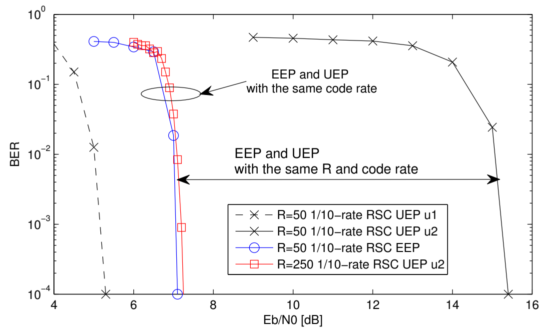


Fig. 5. NC performance comparison for $K = 2$ users, EEP and UEP designs.

Fig. 6 shows the attainable performance for K=2 users relying on the EEP design. As it can be inferred by carefully studying Fig. 4 the BER of user 1 becomes vanishingly low for 90 antennas after a number of decoding iterations $I = 12$. Reducing the number of iterations to $I = 4$, upon using a 1/2-rate RSC as the outer code we have reduced the value of R from 1,000 to 100 antennas compared to the uncoded case of [10]. The number of iterations I is obtained with the aid of the EXIT chart analysis. It indicates the number of iterations which are needed so that BER tends to an infinitesimally low value. We chose for each case the specific number of iterations which achieve the best BER. Alternatively, observe in Fig. 6 that we are able to reduce R to 350 antennas with respect to $R =$ 1,000 for the uncoded case in conjunction with $I = 5$ using a 9/10-rate RSC as the outer code. Furthermore, we can compare both schemes for the same number of iterations, such as $I = 4$. In this case, the performance of the 9/10 rate code is slightly worse, as seen in Fig. 4. Additionally, it is explicitly observed that $I = 4$ is sufficient for the 1/2-rate RSC to converge, while 9/10 requires one additional iteration.

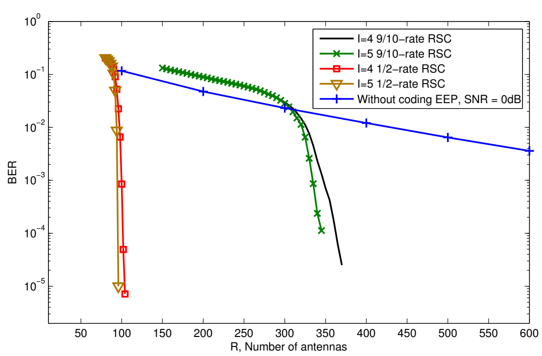


Fig. 6. BER for $K = 2$ users EEP, $\rho = 0$ dB, URC as inner encoder and 1/2, 9/10-rate RSC as outer encoder

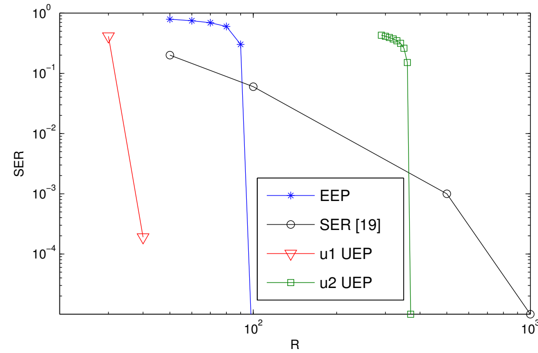


Fig. 7. SER comparison with [19] for $K = 2$ users EEP, $\rho = 0$ dB, URC , 1/2-rate RSC

In order to compare the performance to previous work [19] we use now the symbol error rate (SER). Fig. 7 shows the SER for both the UEP and EEP schemes as well as for [19]. For the same performance, our system needs a lower number of antennas than [19] for SER=10^{-3} and more than 3 orders of magnitude lower for SER=10^{-5}, which is an explicit benefit of iterative decoding.

We also compare the performance of K=2 users relying on our EEP design to that achieved by a coherent Maximum Ratio Combining (MRC) receiver. For this comparison we assume that the CSI is estimated and hence it is subject to a realistic estimation error, which is assumed to be Gaussian. Moreover, for a fair comparison we should take into account the effective throughput reduction due to the insertion of pilots for channel estimation. We will assume a rate-loss of 33% as detailed in [10]. Therefore, in Fig. 8 for $R = 90$, the NC scheme using a 1/5-rate is compared to its coherent counterpart associated with a 3/10-rate, so they have the same effective rate. We can see that for the target of BER$= 10^{-4}$ and for $R = 90$ the E_b/N_0 difference is 2 dB, the same that was found without coding [10]. This difference can be overcome by increasing R in the NC scheme. As shown in the figure, for $R = 110$ antennas de difference is approximately 0.5 dB. If we target a lower BER, the difference between the NC and the coherent scheme is reduced to 1 dB, which is compensated by having only 20 additional antennas. For even lower BER, the two schemes are equivalent. The authors of [5] and [19] have also found the same asymptotic relationship between the coherent and NC schemes. Moreover, the fact that for low E_b/N_0 the use of CSI can result in a waste of resources has already been pointed out in [5].

VI. Conclusions

We have proposed a multi-user BICM-ID aided NC-LS SIMO system based on M-ary DPSK. We presented a constellation design for EEP, which allows all users to have the same performance. We parameterized the EXIT curves by the number of antennas that were used to optimize the design. We demonstrated that a feasible number of $R =$ 100 antennas are sufficient for several situations. We have seen that R may be reduced from 1,000 to 100 with respect to the system operating without coding [10]. Moreover, using iterative decoding also

TABLE I
PERFORMANCE COMPARISON FOR $K = 2$ USERS, EEP AND UEP CONSTELLATION FOR DIFFERENT RSC ENCODERS AND URC .

	Outer RSC	Required number of antennas R for each SNR: users 1,2 EEP					Required R for each SNR: user 1 UEP			Required R for each SNR: user 2 UEP		
		0 dB	3 dB	6 dB	-3 dB	-6 dB	0 dB	3 dB	6 dB	0 dB	3 dB	6 dB
Coding rate	1/10	20	20	20	50	120	20	20	20	100	60	40
	3/20	30	20	20	70	180	20	20	20	130	80	55
	1/5	40	20	20	90	230	20	20	20	170	95	65
	1/4	50	25	20	110	280	20	20	20	200	115	80
	3/10	55	30	20	125	310	30	20	20	230	130	90
	7/20	60	35	25	140	370	30	20	20	260	150	115
	2/5	70	40	30	110	440	30	20	20	290	170	130
	9/20	80	45	30	170	500	35	20	20	320	190	140
	1/2	90	50	35	200	550	40	25	20	360	210	155
	11/20	100	60	40	250	650	45	25	20	400	230	165
	3/5	120	65	45	270	750	50	30	20	440	260	175
	13/20	130	75	55	300	850	55	35	25	500	290	200
	7/10	150	90	60	350	1000	65	40	25	550	320	220
	3/4	180	100	70	400	1150	75	45	35	600	360	250
	4/5	210	120	85	500	1300	100	55	40	640	420	280
	17/20	260	150	100	600	1600	130	80	55	690	480	330
	9/10	350	180	130	750	1900	190	120	85	740	530	380

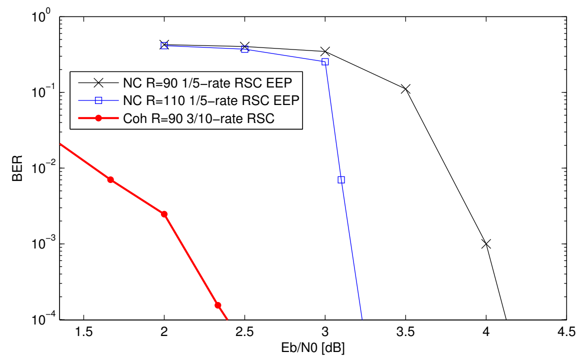


Fig. 8. Performance comparison between the non-coherent and coherent scheme for $K = 2$ users for the EEP design.

improves the performance compared to other coded schemes [19]. Increasing the number of users that can be supported by the BS was enhanced at the cost of increasing R. The number of users that can be multiplexed may be further increased with the aid of more powerful coding schemes chosen as the outer code, which is the subject of our future work. Moreover, an optimized mapping for the joint constellation is also left for further research.

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