Performance of Channel Coded Noncoherent Systems: Modulation Choice, Information Rate, and Markov Chain Monte Carlo Detection

Rong-Rong Chen, *Member, IEEE,* Ronghui Peng, *Student Member, IEEE*

Abstract

This paper investigates performance of channel coded *noncoherent* systems over block fading channels. We consider an iterative system where an outer channel code is serially concatenated with an inner modulation code amenable to noncoherent detection. We emphasize that, in order to obtain near-capacity performance, the information rates of modulation codes should be close to the channel capacity. For certain modulation codes, a single-input single-output (SISO) system with only one transmit antenna may outperform a dual-input and single-output (DISO) system with two transmit antennas. This is due to the intrinsic information rate loss of these modulation codes compared to the DISO channel capacity. We also propose a novel noncoherent detector based on Markov Chain Monte Carlo (MCMC). Compared to existing detectors, the MCMC detector achieves comparable or superior performance at reduced complexity. The MCMC detector does not require explicit amplitude or phase estimation of the channel fading coefficient, which makes it an attractive candidate for high rate communication employing quadrature amplitude modulation (QAM) and for multiple antenna channels. At transmission rates of $1 \sim 1.667$ bits/sec/Hz, the proposed SISO systems employing 16QAM and MCMC detection perform within 1.6-2.3 dB of the noncoherent channel capacity achieved by optimal input.

Index Terms

Noncoherent detection, Markov Chain Monte Carlo, fading channel, multiple antenna, transmit diversity, iterative decoding, channel capacity.

This work is supported in part by NSF under grant ECS-0547433.

I. INTRODUCTION

In recent years, noncoherent communication, which assumes that the receiver does not have explicit channel information *a priori* has attracted significant attention. Here, noncoherent communication is interpreted in its broadest sense as joint data and channel estimation. In the noncoherent scenario: (1) The transmitted signal power spent on pilot symbols (if any) is taken into account; (2) Implicit/explicit channel estimations are done using all transmitted symbols rather than using pilots symbols alone.

With the goal of approaching channel capacity, it is important to study channel coded noncoherent systems where powerful channel codes are employed in addition to the modulation codes to strengthen the error correction capability. Capacity-approaching joint channel decoding and noncoherent detection strategies for single-input single-output (SISO) channel at the spectral efficiency of $R = 1/2$ (bits/sec/Hz) with QPSK modulation are developed in [1]. The proposed low-complexity noncoherent detector, which requires separate amplitude estimation and phase quantization of the channel, is designed for the single antenna channel only and requires the signal constellation to have a constant amplitude level. Channel coded SISO systems with $R = 1$ and QPSK modulation are studied in [2] and a noncoherent detector based on linear prediction and per-survivor processing is proposed. For dual-input and single-output (DISO) channels that employ two transmit antennas and single receive antenna, [3] considers turbo coded noncoherent systems employing the unitary space-time modulation (USTM) with $R = 0.875 \sim 1.75$, in which the optimal *a posterior* probability (APP) detector is used. In [4], the performance of channel coded systems employing pilot-symbol assisted modulation (PSAM) based on Alamouti's codes [5] and QPSK/8PSK modulation are studied for $R = 1 \sim 1.5$. The PSAM codes demonstrate better performance than that of the USTM. A low complexity noncoherent detector based on bit-flipping and phase quantization is proposed in [4]. This detector has a complexity that is linear in the coherence length and is shown to obtain near-optimal soft information [6].

Although there is much work on the design of modulation code and noncoherent detection algorithm for both the SISO coded system and the DISO coded system, very little research has been done on the comparison of these two systems. For instance, an interesting question is that, for $R = 1 \sim 2$, is it safe to assume that a system with two transmit antennas, as in a DISO system, automatically performs better than a SISO system with only one transmit antenna? Our results show that adding a second antenna does not necessarily enhance performance. Since the choice of modulation codes directly affects the performance of coded systems, DISO systems with certain choices of modulation codes may perform worse than that of a SISO system.

In this work, we point out an important design criterion for channel coded system. That is, in order to obtain near capacity performance, one should choose modulation codes whose mutual information rates are close to the optimal channel capacity. We provide an explicit comparison of the mutual information rates between a simple 16QAM modulation code for the SISO channel, and USTM code and PSAM code [4] for the DISO channel. It is shown that the mutual information rates of USTM and PSAM codes are much lower than that of the 16QAM code for SISO channel, which implies that they fall well below the DISO channel capacity. This contributes to the fact that such DISO systems perform even worse than the SISO system. To the best of our knowledge, this is the first work to investigate the effect of transmit diversity for channel coded noncoherent systems through an explicit comparison of the SISO system and the DISO system.

Furthermore, we propose a novel noncoherent detector that is different from existing detectors [1][4] in the sense that it does not require amplitude estimation or phase quantization of the channel fading coefficient. Such a detector is based on the Markov Chain Monte Carlo (MCMC) method. Applications of the MCMC detectors for coherent multiple-input and multiple-output (MIMO) channels have been studied in [7]–[10], which demonstrate significant performance improvement over traditional MIMO detectors such as the sphere decoding detector. In this paper, we extend the application of the MCMC approach to the noncoherent setting. For a SISO system with 16QAM modulation, we are able to achieve near capacity performance at $R = 1 \sim 1.667$. These transmission rates are higher than the rate of $1/2$ in [1] with QPSK modulation, and are comparable with that of the DISO systems considered in [3][4].

Noncoherent MCMC detectors are first studied by X. Wang et. al. [11]–[14] for OFDM systems and multicarrier CDMA systems. The noncoherent MCMC detector proposed in this paper originates from coherent MCMC detectors of [7]–[10]. Such MCMC detectors require neither the burning period nor bit-counting for computing *a posteriori* probabilities [8]. They significantly outperform traditional MIMO detectors such as the sphere decoding detector. Detailed differences between the proposed detector and those of [11]–[14] will be highlighted in Section IV.

The rest of the paper is organized as follows. Section II contains the system model. Section III studies the mutual information rate of modulation codes and its impact on coded performance. Section IV includes a detailed description of the noncoherent MCMC detector. Simulation results of the coded system with MCMC detection are presented in Section V. Conclusions are given in Section VI.

II. System model

We consider a SISO block fading channel where the channel remains constant for each block of T_c symbols (where T_c is called the coherence length), and is independent between blocks. We model the channel by :

$$\mathbf{y} = \sqrt{\rho}\, h\, \mathbf{s} + \mathbf{w}, \tag{1}$$

where $h \sim \mathcal{CN}(0,1)$ is the Rayleigh fading coefficient of a given block and is a circularly symmetric complex Gaussian random variable with zero mean and unit variance; the vectors $\mathbf{y}, \mathbf{s}, \mathbf{w}$ are T_c-dimensional complex vectors representing the received signal, the transmitted signal, and the noise, respectively; the entries of $\mathbf{w}$ are independent and identically distributed with distribution $\mathcal{CN}(0,1)$. The constant ρ represents the signal-to-noise ratio (SNR), assuming that the average power of the transmitted signal $\mathbf{s}$ is normalized such that $E[\mathbf{s}^\dagger \mathbf{s}] = T_c$, and $\dagger$ denotes the Hermitian operator. In the noncoherent scenario, neither the transmitter nor the receiver knows the exact realization of the channel coefficient h. Given $\mathbf{s}$, the noncoherent conditional probability density function (pdf) of $\mathbf{y}$ is given by [15]:

$$p(\mathbf{y}|\mathbf{s}) = \frac{1}{\pi^{T_c}(1+\rho\|\mathbf{s}\|^2)} \exp\Big\{ -\|\mathbf{y}\|^2 + \frac{\rho\|\mathbf{y}^\dagger \mathbf{s}\|^2}{1+\rho\|\mathbf{s}\|^2}\Big\} \tag{2}$$

Fig. 1 shows a block diagram of the channel coded noncoherent system. The transmitter side consists of a serial concatenation of the channel encoder and the modulation coder, with a symbol mapper in between that maps a sequence of binary coded bits to a sequence of complex symbols from a finite constellation of size 2^{M_c} through Gray mapping. In this paper we consider a simple modulation code that maps an input block of $(T_c - 1)$ complex symbols to an output block of T_c symbols by inserting a reference symbol c_0 (from the same constellation) in the front of each input block: $(s_1, \cdots, s_{T_c-1}) \rightarrow \mathbf{s} = (s_0 = c_0, s_1, \cdots, s_{T_c-1})$, where each $s_i, i = 1, \cdots, T_c - 1$ is a complex symbol representing M_c bits. The output of the modulation coder $\mathbf{s}$ is then transmitted through the block fading channel. Assume that the channel code has a rate of R_c, then the overall transmission rate of this system is given by $R = \frac{T_c-1}{T_c} R_c M_c$, where the term $T_c - 1$ is due to the fact that only $T_c - 1$ coded symbols are transmitted out of each block of T_c symbols. At the receiver end, joint channel decoding and noncoherent detection is performed iteratively through soft information exchange between the channel decoder and the noncoherent block detector. After a predetermined number of iterations, decisions are made at the output of the channel decoder to generate the decoded bit sequence.

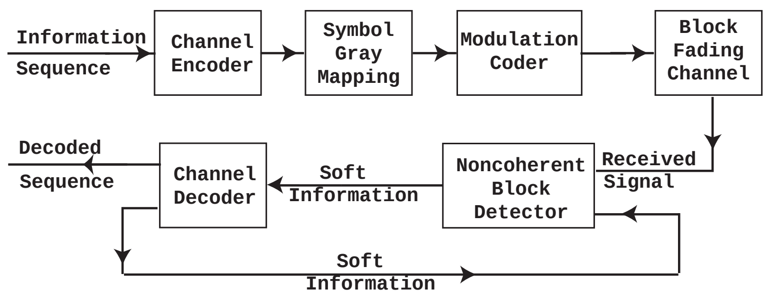

Fig. 1. A schematic block diagram of the channel coded noncoherent system.

III. Information rate of modulation code and its impact on coded performance

In this paper, we emphasize two modules of the channel coded system shown in Fig. 1: the modulation code at the transmitter side, and the noncoherent detector at the receiver side. From a system perspective, the selection of modulation code is important because its mutual information rate determines the maximum information rate that a coded system can achieve for a given average signal-to-noise power ratio E_s/N_0. In other words, for a desired transmission rate R, the information rate of the modulation code determines the minimum E_s/N_0, denoted by $\frac{E_s}{N_0}|_{\min}$, required to achieve R. Note that this is an information-theoretical limit that can be achieved only with optimal detection and a powerful channel code with an arbitrarily long code length and maximum-likelihood decoding. Hence, it is also the performance limit of any practical channel coded system with suboptimal detectors and iterative decoding.

In this section, we examine the information rate of the modulation code defined in Section II for SISO channel and make comparisons with those of certain modulation codes used for DISO channels. This will explain the performance gap between the proposed SISO system and that of the DISO systems in [4], shown in Section V.

A. Information rates of practical modulation codes for SISO channel

We first consider a SISO noncoherent block fading channel with $T_c = 6$. In Fig. 2, we plot the channel capacity achieved by the optimum input and the mutual information rates [1] of the modulation code described in Section II using practical constellations 16QAM, 8QAM, 8PSK, and QPSK (a similar figure for $T_c = 5$ was presented in [1]). From Fig. 2 we make two observations. First, the mutual information rates of modulation codes provide performance benchmarks for channel coded systems. For instance, given the 16QAM constellation, from Fig. 2 we see that when $\frac{E_s}{N_0} = 8$ dB, the mutual information rate of the corresponding modulation code equals $R = 1.667$. This means that, to achieve $R = 1.667$, the $\frac{E_s}{N_0}|_{\min}$ required by *any* channel coded system using this modulation code and 16QAM equals 8 dB. This is independent of the choices of detection algorithms and the channel codes. Second, the information rates of 16QAM are higher than that of the other practical constellations considered here, therefore best approximate the channel capacity achieved by the optimal input. For instance, when $R = 1.667$, we have $\frac{E_s}{N_0}|_{\min} = 7.5,\ 8,\ 8.7$ dB, respectively, for the optimal input, 16QAM, and 8QAM. Hence, 16QAM is better than 8QAM, because the $\frac{E_s}{N_0}|_{\min}$ required is only 0.5 dB away from that of the optimal input. Interestingly, when $R = 1$, even though we have similar values of $\frac{E_s}{N_0}|_{\min}$ for 16QAM and 8QAM (4.2 dB and 4.4 dB, respectively), our simulation results show that actual performance of the coded system is better with 16QAM, due to the use of a lower rate channel code.

From the study of the information rates of modulation codes, we conclude that, in order to obtain capacity-approaching performance, it is important to choose modulation codes whose information rates are close to the optimal channel capacity.

B. Comparisons of information rates for SISO system and DISO system

For similar target transmission rates, recent work ([3], [4]) consider DISO systems with dual transmit antennas and single receive antenna. To facilitate low-complexity noncoherent detection, modulation codes such as USTM are often employed in practical DISO systems. Unfortunately, these codes suffer from intrinsic information rate loss compared to the optimal channel capacity. In [16], it is pointed out that the information rates of USTM achieve only a fraction of channel capacity. In [4] , two modulation codes are considered for a DISO channel with $R = 1$: the 512-ary USTM and the 256-ary QPSK/Alamouti code. The latter is an orthogonal space-time

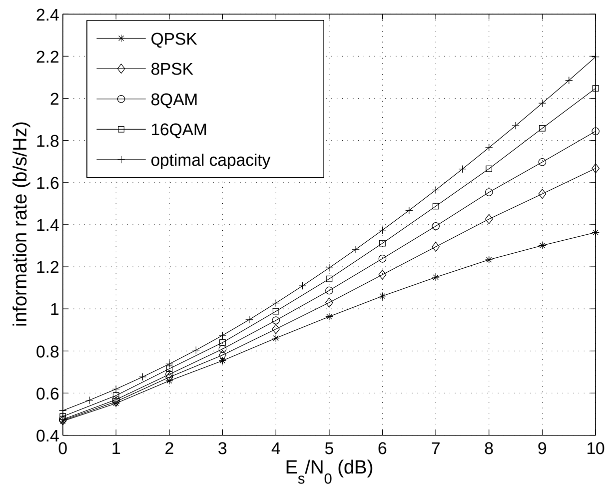

Fig. 2. Comparisons of the optimal capacity with the mutual information rate of a practical modulation code with various constellations. A SISO noncoherent block fading channel with $T_c = 6$ is considered.

code based on the Alamouti's scheme [5] and QPSK modulation. Computation of information rates reveal that $\frac{E_s}{N_0}|_{\min} = 8.15$ dB and 8.45 dB, respectively, for these two codes. These are about 4 dB more than the $\frac{E_s}{N_0}|_{\min} = 4.2$ dB required for a SISO channel with 16 QAM modulation code (see Section III-A). For $R = 1.5$, compared to $\frac{E_s}{N_0}|_{\min} = 7.1$ dB for a SISO channel with 16QAM, [4] shows that a DISO channel with 8PSK/Alamouti code also requires about 4 dB more ($\frac{E_s}{N_0}|_{\min} = 11.76$ dB). Simulation results in Section V will verify that, the proposed SISO system with 16QAM indeed outperforms the DISO systems with the modulation codes above by about 4 dB.

These comparisons clearly show that the information rates of these modulation codes used for DISO systems can be much lower than those of the modulation codes used for SISO systems. Therefore, an important observation is that for such scenarios a SISO system should be chosen over a DISO system, and one should not waste the resource of a second transmit antenna.

We want to emphasize that, for the special scenarios discussed above, a SISO system outperforms a DISO system largely due to the limited information rate of the specific modulation codes used for DISO systems. This by no means suggest that a DISO channel is intrinsically worse than a SISO channel. In fact, the capacity of DISO channel should be no less than that of a SISO channel, because with dual transmit antennas, one can always choose to allocate full power to one of the transmit antennas to realize the single antenna performance. However, the capacity of the DISO channel is achieved only with the optimal input, whose distribution is still unknown. Even if it were known, practical systems in general can not use the optimal input as the modulation code due to prohibitive complexity. Hence, practical DISO systems use modulation codes such as USTM and PSAM that are amenable for low-complexity detection. However, as we point out, the intrinsic information rate loss prevents them from achieving channel capacity. Thus, it remains an open problem to find good modulation codes that can fully utilize the capacity of DISO channels and support low-complexity detection at the same time.

IV. Noncoherent detection based on Markov Chain Monte Carlo (MCMC)

In this section, we propose a novel noncoherent detector based on the MCMC approach. This detector originates from the coherent MCMC detector in [7], [8], [10] where coherent detection is employed assuming perfectly known channel fading coefficient. Here, we extend the basic idea of MCMC detection to the noncoherent scenario where the channel fading coefficient is unknown.

The main function of the MCMC detector is to compute extrinsic log-likelihood ratios (LLR) of the coded bits based on received signal vector $\mathbf{y}$ and prior LLRs $\{\lambda_i\}$ provided by the channel decoder. The MCMC detector operates in two steps. In step 1, it adopts a statistical approach, i.e., the Gibbs sampler, to identify a *small* set of I "important bit vectors", denoted by $\mathcal{A}$. In step 2, the detector computes the output extrinsic LLRs $\{\gamma_i\}$ by applying the max-log algorithm over the set of vectors in $\mathcal{A}$. This greatly reduces the computational complexity compared to that of optimum detection.

Given a modulation codeword $\mathbf{s} = (c_0, s_1, \cdots, s_{T_c-1})$, we denote the bit sequence corresponding to $\{s_1, \cdots, s_{T_c-1}\}$ by $\mathbf{b} = \{b_1, b_2, \cdots, b_K\}$, where $K = (T_c - 1)M_c$. In particular, the M_c bits constituting symbol s_i are $\{b_{(i-1)M_c+1}, \cdots, b_{iM_c}\}$. Each bit b_i equals either 0 or 1.

Step 1: Gibbs Sampler

Initialization $n = 0$;

generate the initial vector $\mathbf{b}^{(0)} = \{b_1^{(0)}, \cdots, b_K^{(0)}\}$ according to (3).

Iteration n

for $n = 1$ to I

for $i = 1$ to K

Sample the i-th bit of $\mathbf{b}^{(n)}$ according to the *a posteriori* probability distribution

$b_i^{(n)} \sim \pi(\cdot|\, b_1^{(n)}, \cdots, b_{i-1}^{(n)}, b_{i+1}^{(n-1)}, \cdots, b_K^{(n-1)}, \lambda_i)$.

end i loop

end n loop

Next, we explain details of the above procedure.

First, the initialization step to find $\mathbf{b}^{(0)}$ is done as follows. For each $i = 1, \cdots, T_c - 1$, we compute the most likely transmitted symbol $\hat{s}_i$ based on the received signals y_0 and y_i by letting

$$\hat{s}_i = \text{argmax}_{s_i}[\ln p(y_0, y_i|c_0, s_i) + \ln P(s_i)], \tag{3}$$

where $\ln P(s_i) = \sum_{j=(i-1)M_c+1}^{iM_c} (\lambda_j/2)(-1)^{b_j}$ is the logarithm of the prior probability of symbol s_i, and λ_j is the prior LLR of the j-th bit. Note that the pdf $p(y_0, y_i|c_0, s_i)$ in (3) is the noncoherent pdf corresponding to $T_c = 2$ because only two signals y_0 and y_i are considered. The symbol $\hat{s}_i$ is then used to define the initial bit vector $\mathbf{b}^{(0)}$ by letting $(\mathbf{b}_{(i-1)M_c+1}^{(0)}, \cdots, \mathbf{b}_{iM_c}^{(0)})$ equal to the bits constituting symbol $\hat{s}_i$.

In the step of sampling $b_i^{(n)}$, we define

$$\begin{aligned} \mathbf{a}_0 &= \{b_1^{(n)}, b_2^{(n)}, \cdots, b_{i-1}^{(n)},\ 0,\ b_{i+1}^{(n-1)}, \cdots, b_K^{(n-1)}\} \\ \mathbf{a}_1 &= \{b_1^{(n)}, b_2^{(n)}, \cdots, b_{i-1}^{(n)},\ 1,\ b_{i+1}^{(n-1)}, \cdots, b_K^{(n-1)}\} \end{aligned} \tag{4}$$

Let

$$x = \ln \frac{p(\mathbf{y}|c_0, \mathbf{a}_0)}{p(\mathbf{y}|c_0, \mathbf{a}_1)} + \lambda_i \quad \text{and} \quad t = e^x/(1 + e^x), \tag{5}$$

where the pdf in (5) is computed based on the noncoherent pdf (2). We then generate a random number $u \in [0, 1]$. If $u < t$, we let $b_i^{(n)} = 0$, otherwise we let $b_i^{(n)} = 1$.

Step 2: Compute the output extrinsic LLR $\{\gamma_i\}$

For each bit vector $\mathbf{b} \in \mathcal{A}$, by replacing its i-th bit by 0 and 1, respectively, and leaving other bits unchanged, we obtain two new bit vectors $\mathbf{b}^{i,0}$ and $\mathbf{b}^{i,1}$. These vectors are used to compute the output extrinsic LLR γ_i for bit i:

$$\gamma_i = \max_{\mathbf{b}\in\mathcal{A}} [\ln p(\mathbf{y}|c_0, \mathbf{b}^{i,0}) + \ln P(\mathbf{b}^{i,0})] - \max_{\mathbf{b}\in\mathcal{A}} [\ln p(\mathbf{y}|c_0, \mathbf{b}^{i,1}) + \ln P(\mathbf{b}^{i,1})] - \lambda_i, \tag{6}$$

where $\ln P[\mathbf{b}^{i,0}] = \sum_{j=1}^{K} (\lambda_j/2)(-1)^{\mathbf{b}_j^{i,0}}$ and $\mathbf{b}_j^{i,0}$ denotes the j-th bit of $\mathbf{b}^{i,0}$. The term $\ln P[\mathbf{b}^{i,1}]$ is computed similarly.

We note that the proposed MCMC detector differs from the MCMC detectors of [11]–[14] in both initialization (3) and computation of output LLRs (6). In [11]–[14], bit-counting (use statistical averaging to estimate the frequency that a particular bit value occurs) is applied to compute the LLRs. In comparison, we use (6) to compute the LLRs based on the *a posteriori* probabilities of the samples generated by the Gibbs sampler. The proposed detector does not require a burning period and only a small number of samples are needed to achieve satisfactory performance. Detailed analysis of the proposed noncoherent MCMC detector resembles those of [8] for coherent MIMO channels.

The main complexity of the MCMC detector comes from the *computation of the noncoherent pdf* (CNP) in (5) and (6). Hence, the total number of CNP can fairly accurately represent the complexity of the MCMC detector. After we obtain $\mathbf{b}^{(0)}$, we perform one CNP to obtain $p(\mathbf{y}|c_0, \mathbf{b}^{(0)})$. Subsequently, in the step of generating $b_i^{(n)}$, according to (5), only one CNP is needed for either $p(\mathbf{y}|c_0, \mathbf{a}_0)$, or $p(\mathbf{y}|c_0, \mathbf{a}_1)$, because one of them has been computed in the previous step for generating $b_{i-1}^{(n)}$. Hence, the number of CNPs required to generate the I bit vectors in $\mathcal{A}$ is $1+IK \approx IK$. To compute γ_i, according to (6), we flip the i-th bit of each vector in $\mathcal{A}$, and perform I CNPs for these new vectors. For a total of K bits, this leads to IK CNPs. Hence, the total complexity of the MCMC detector, in terms of CNPs, equals approximately $2IK$. We have also confirmed that the number of CNPs, as a measurement of detection complexity, agrees well with the actual simulation time of the coded systems.

V. Simulation results

In this section, we first examine the effectiveness of the proposed MCMC detector by comparing it with the noncoherent detector in [4]. The latter detector, which we refer to as the bit-flipping (BF) detector, is shown to be near optimal for channels with small or moderate coherence lengths [6]. The complexity of the BF detector, in terms of CNPs, equals QK, where Q is the number of phase quantization, and $K = (T_c - 1)M_c$ is the total number of bits transmitted in each block. From Section IV, we know that the MCMC detector has a complexity of $2IK$ CNPs. We show that, for $R = 1 \sim 1.933$, small values of I such as $I = 3$ give satisfactory performance for both small values of $T_c = 6$, for a fast fading scenario, and $T_c = 30$ for a slower fading scenario. To facilitate fair comparisons, we consider channel coded systems using the same channel code, which is a commonly used regular (3,6) low-density parity-check (LDPC) code with rate $R_c = 1/2$ and code length 10^4, and the same modulation code described in Section III with 16QAM. The overall rate of the system, hence, equals $R = \frac{T_c - 1}{T_c} 4 R_c$, corresponding to $R = 1.667$ for $T_c = 6$ and $R = 1.933$ for $T_c = 30$, respectively.

Using the same channel coded system as described above, we compare the performance of the MCMC detector and the BF detector. Fig. 3 shows the bit-error-rate (BER) of the coded system versus the average energy per information bit to noise ratio E_b/N_0. It relates to E_s/N_0 by $\frac{E_b}{N_0}\big|_{\text{dB}} = \frac{E_s}{N_0}\big|_{\text{dB}} - 10\log_{10} R$. To reduce the overall system complexity, we perform one iteration of noncoherent detection for every 10 inner iterations of LDPC decoding. The maximum number of outer iterations between the detector and the LDPC decoder set to be 60 to ensure the convergence of the decoding process.

Fig. 3 shows that for $T_c = 6$ ($R = 1.667$), the MCMC detector with $I = 3$ performs only slightly better than the BF detector with $Q = 6$. Note that the complexity of these two detectors are roughly the same in terms of CNPs. When Q increases to 10, at the cost of higher complexity, the performance of the BF detector improves slightly and is now virtually the same as the MCMC detector with $I = 3$. For the case of $T_c = 30$ ($R = 1.933$), the performance curves of the MCMC detector ($I = 3$), the BF detector with $Q = 10$, and the BF detector with $Q = 6$ are shown as the three leftmost curves in Fig. 3. Note that with roughly the same complexity, the MCMC detector with $I = 3$ outperforms the BF detector with $Q = 6$ by about 0.5 dB at BER=10^{-4}. Even with an increased complexity by setting $Q = 10$ for the BF detector, the MCMC detector with $I = 3$ still performs slightly better.

TABLE I

OPTIMIZED LDPC CODE PARAMETERS FOR A BLOCK FADING CHANNEL WITH $T_c = 6$.

Modulation	R	R_c	optimzied LDPC degree sequence
16QAM	1.667	0.5	$d_c = 6, d_v = [2, 3, 7, 8], u_v = [0.5843,\ 0.2799,\ 0.0947,\ 0.0411]$
16QAM	1	0.3	$d_c = 5, d_v = [2, 3, 7, 8, 22, 49, 50], u_v = [0.5848,\ 0.2818,\ 0.0386,\ 0.0778,\ 0.009,\ 0.0042,\ 0.0038]$
8QAM	1	0.4	$d_c = 6, d_v = [2, 3, 6, 7, 21, 22], u_v = [0.5562,\ 0.2760,\ 0.0085,\ 0.1252,\ 0.0170,\ 0.0170]$

In Table I, R is the overall transmission rate of the coded system, R_c is the rate of the LDPC code,d_c is the degree of check node, d_v is the degree sequences of variable nodes, $u_v(i)$ is the fraction of variables nodes that has degree $d_v(i)$. The code length is 3×10^4.

Next, to further approach channel capacity, we optimize the channel code to best match the

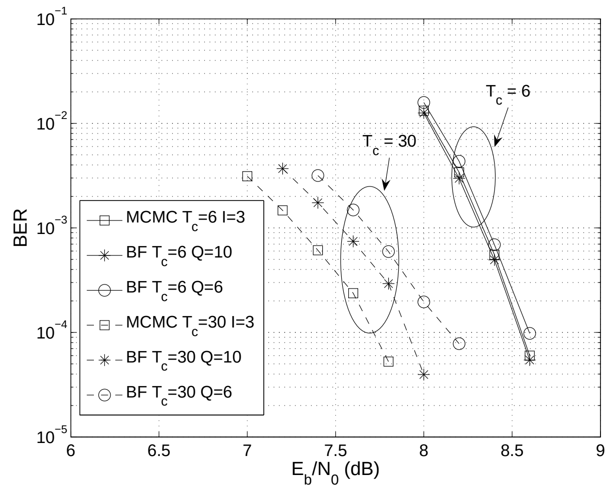

Fig. 3. Performance comparisons of the same SISO system with the MCMC detector and the BF detector in [4]

characteristics of the MCMC detector. The degree sequence of the LDPC code is optimized by following the extrinsic information transfer (EXIT) chart approach in [17] and the optimized code parameters are given in Table I. Performance of the optimized channel coded system using MCMC detection is shown in Fig. 4. For $R = 1.667$, with 16QAM and an optimized LDPC code of rate $R_c = 1/2$, the channel coded system achieves within 1.8 dB of the capacity limit of 16QAM ($\frac{E_b}{N_0}|_{\min} = 5.78$ dB) at BER $= 10^{-4}$, and is 2.3 dB away from the capacity limit of the optimal input. For $R = 1$, with 16QAM and an optimized LDPC code of rate $R_c = 0.3$, we achieve within 1.2 dB of the capacity limit of 16QAM ($\frac{E_b}{N_0}|_{\min} = 4.2$ dB), and is 1.6 dB away from the capacity of the optimal input. When compared to the performance of the DISO systems [4] at $R = 1$ and $R = 1.5$, we note that the proposed SISO system achieves a roughly 4 dB performance gain. This is consistent with our observation in Section III that, at these transmission rates, the $\frac{E_b}{N_0}|_{\min}$ required by the 16QAM code here is about 4 dB less than that of the modulation codes used in [4]. In Section III we also note that the $\frac{E_s}{N_0}|_{\min}$ for 16QAM and 8QAM have similar values (differ by only 0.2 dB). For the 8QAM coded system, an optimized LDPC code of higher rate of $R_c = 0.4$ is used to achieve $R = 1$. Fig. 4 shows that the 8QAM system performs about 0.6 dB worse than the 16QAM system.

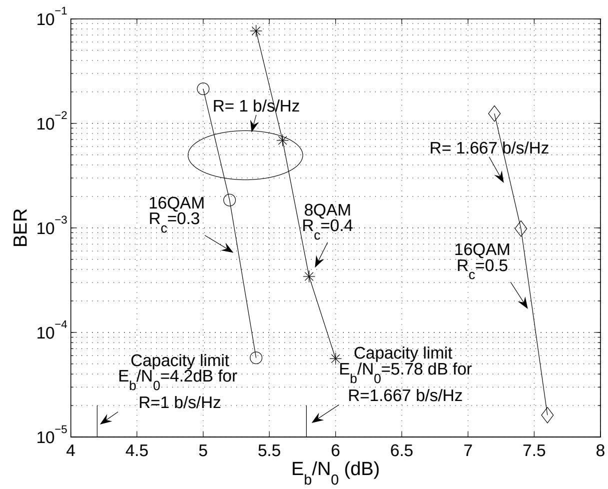

Fig. 4. Performance of the SISO system with optimized LDPC codes and MCMC detection for $T_c = 6$.

VI. Conclusion

In this paper, we examine the performance of noncoherent channel coded systems. We show that transmit diversity does not necessarily enhance performance when there is a large gap between the mutual information rates of the modulation codes used in such systems and the optimal channel capacity. This explains the interesting but somewhat surprising fact that the proposed SISO systems significantly outperform certain DISO systems by as much as 4 dB. While in this work we focus on systems with single receive antenna, we note that the basic principles presented here are applicable to general scenarios with multiple receive antennas. For instance, our preliminary results show that even with dual receive antennas, systems with one transmit antenna can still outperform dual transmit antenna systems employing similar modulation codes discussed here for the DISO channel. Hence, an interesting direction for future work is to design better modulation codes for multiple transmit antenna channels that can fully exploit the channel capacity and also allow for low-complexity detection. The proposed MCMC detector achieves excellent performance for the SISO channel without explicit channel amplitude or phase estimation. We believe that it will be instrumental in facilitating efficient implementation of the capacity-approaching noncoherent systems for multiple antenna channels.

REFERENCES

[1] R. R. Chen, R. Koetter, U. Madhow, and D. Agrawal, "Joint noncoherent demodulation and decoding for the block fading channel: a practical framework for approaching shannon capacity," *IEEE Trans. Commun.*, vol. 51, no. 12, pp. 1676–1689, Oct. 2003.

[2] P. Hoeher and J. Lodge, ""Turbo DPSK": iterative differential PSK demodulation and channel decoding," *IEEE Trans. on Communication*, vol. 47, pp. 837–843, Jun. 1999.

[3] I. Bahceci and T. M. Duman, "Combined turbo coding and unitary space-time modulation," *IEEE Trans. on Commun.*, vol. 50, no. 8, pp. 1244–1249, Aug. 2002.

[4] A. Krishnamoorthy and A. Anastasopoulos, "Code and receiver design for the noncoherent fast-fading channel," *IEEE J. Select. Areas Commun.*, vol. 23, no. 9, pp. 1769–1778, Sep. 2005.

[5] S. Alamouti, "A simple transmitter diversity technique for wireless communications," *IEEE J. Select. Areas Commun.*, vol. 16, no. 8, pp. 1451–1458, Oct. 1998.

[6] I. Motedayen-Aval, A. Krishnamoorthy, and A. Anastasopoulos, "Optimal joint detection/estimation in fading channels with polynomial complexity," *IEEE Trans. on Inform. Theory*, vol. 53, pp. 209–223, Jan. 2007.

[7] H. Zhu, Z. Shi, and B. Farhang-Boroujeny, "MIMO detection using Markov Chain Monte Carlo techniques for near capacity performance," in *Proc. IEEE Int. Conf. on Acoustic, Speech, and Signal Processing (ICASSP'05)*, (Philadelphia, PA), pp. 1017–1020, Mar. 2005.

[8] B. Farhang-Boroujeny, H. Zhu, and Z. Shi, "Monte Carlo Markov chain techniques for CDMA and MIMO communication systems," *IEEE Trans. Signal. Process.*, vol. 54, no. 5, pp. 1896–1909, May 2006.

[9] H. Zhu, B. Farhang-Beroujeny, and R. R. Chen, "On performance of sphere decoding and Markov Chain Monte Carlo methods," *IEEE Signal Processing Letters*, vol. 12, no. 10, pp. 669–672, Oct. 2005.

[10] Z. Shi, H. Zhu, and B. Farhang-Boroujeny, "Markov chain Monte Carlo techniques in iterative detectors," in *Proc. IEEE Globecom'04*, (Dallas, TX), pp. 325–329, Nov./Dec. 2004.

[11] B. Lu and X. Wang, "Bayesian Blind Turbo Receiver for Coded OFDM Systems with Frequency Offset and Frequency-Selective Fading," *IEEE Journal on Selected Areas on Communications*, vol. 19, pp. 2516–2527, Dec. 2001.

[12] Z. Yang, B. Lu, and X. Wang, "Bayesian Monte Carlo Multiuser Receiver for Space-Time Coded Multicarrier CDMA Systems," *IEEE Journal on Selected Areas on Communications*, vol. 19, pp. 1625–1637, Aug. 2001.

[13] X. Wang and R. Chen, "Blind turbo equalization in Gaussian and Impulsive Noise," *IEEE Trans. on Vehicular Technology*, vol. 50, pp. 1092–1105, July 2001.

[14] A. Doucet and X. Wang, "Monte Carlo methods for signal processing," *IEEE Signal Processing Magazine*, vol. 22, pp. 152–170, Nov. 2005.

[15] T. Marzetta and B. Hochwald, "Capacity of a mobile multiple-antenna communication link in Rayleigh flat fading," *IEEE Trans. Inform. Theory*, vol. 45, no. 1, pp. 139–157, Jan. 1999.

[16] B. M. Hochwald, T. L. Marzetta, T. J. Richardson, W. Sweldens, and R. Urbanke, "Systematic design of unitary space-time constellations," *IEEE Trans. Inform. Theory*, vol. 46, no. 6, pp. 1962–1973, Sep. 2000.

[17] S. ten Brink, G. Kramer, and A. Ashikhmin, "Design of low-density parity-check codes for modulation and detection," *IEEE Trans. on Commun.*, vol. 52, no. 4, pp. 670–678, Apr. 2004.