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***Abstract*—In this paper, we develop a new algorithm of joint detection and decoding receiver for the large-scale multiple-input multiple-out (LS-MIMO) coded communication systems with low-resolution ADCs. The new algorithm enables us to perform extensive experiments to address the performance and complexity tradeoff issue for LS-MIMO communication systems. Our study results indicate that 4-bit ADC is the best choice to achieve an excellent balance between the transmit power and complexity. Furthermore, the experiments reveal that the LS-MIMO system with 4-bit ADC can approach the performance of the high-resolution LS-MIMO system in the entire range of signal to noise ratio. In the other extreme, the LS-MIMO system with 2-bit ADC suffers 2.2 dB performance loss in terms of transmit power at bit error rate (BER) 10^{-6}. In all test cases, an extra bit to increase the resolution from 4-bit ADC to 5-bit ADC achieves a tiny fraction of power gain in return.**

I. Introduction

Current mobile networks experience the explosive demand for high-speed traffic due to the rapid growth in the smart device market. Therefore, increasing network capacity is a vital requirement to keep service quality at a high standard. Large-scale multiple-input multiple-out (LS-MIMO) transmission is an enabling technology for current and future wireless communication networks to meet such high traffic demand. The advantage of the LS-MIMO transmission scheme is that the base station with a large number of antennas can focus transmitted signal energy into short-range areas in order to bring significant improvements in terms of system capacity [1].

In traditional wireless communication systems, each radio frequency (RF) port at the receiver is connected two high-resolution ADCs (typically more than 10 bits). Nevertheless, using high-resolution ADCs in the massive MIMO scenario, which has hundreds or thousands of active antenna elements, could lead to immense power consumption and hardware costs. Using the low-resolution ADC with a few quantization bits is a practical solution to keep the power consumption and hardware costs within acceptable limits [1], [2].

Not only facing the ADC power consumption and hardware costs, but the massive MIMO system also encounters challenges in signal detection. The reason is that the detectors, devised for the conventional MIMO systems, are not suitable and scalable to the LS-MIMO systems. In [3], authors introduced a modified minimum mean square error (MMSE) receiver for low-resolution ADC MIMO systems where the quantization noise is taken into account to design the linear receiver. The resulting receiver performs better than the conventional MMSE receiver at the high signal to noise ratio (SNR) regime. Fukada et al. [4] introduced a low complexity detector based on the belief propagation (BP) method for the massive MIMO system. The proposed BP-based MIMO detector requires the second-order calculation and has a reasonably low complexity in comparison to the MMSE detector. When it comes to the performance, the BP-based MIMO detector has a notable bit error rate (BER) improvement in contrast with the MMSE detector.

Even though the BP-based MIMO detector has both performance and complexity advantages over the MMSE detector, the BP-based MIMO detector was designed for high-resolution MIMO systems. There is little knowledge on how to adapt the high-resolution BP-based algorithm to the low-resolution scenarios. Therefore, in this paper, the first contribution is to develop a new algorithm of joint detection and decoding based on message passing method for LS-MIMO communication systems where the quantization noise of the low-resolution ADC is modeled as additive noise and, it is taken into the derivation of the extrinsic message expression.

The next contribution of this paper is that we find the answer to the question that how many bits one should use in the ADCs to avoid significant performance loss when using the BP-based detector in the low-resolution MIMO coded communication systems. The study results in this research work show that 4-bit ADC is the best option to gain an excellent balance between the transmit power and complexity. Moreover, the experiments reveal the interesting fact - the LS-MIMO system with 4-bit ADC can approach the performance of the high-resolution LS-MIMO system in the entire range of signal to noise ratio. In the lowest extreme, the LS-MIMO system with 2-bit ADC encounters 2.2 dB performance loss in terms of transmit power at bit error rate (BER) 10^{-6}. In all test cases, an extra bit to increase the resolution from 4-bit ADC to 5-bit ADC attains a tiny fraction of power gain in return.

The rest of this paper is organized as follows. Section II describes the system model and the scalar uniform quantizer of the Q-bit ADC. Then, Section III derive the joint detection and decoding algorithm based on the message passing method with the help of the double-layered Tanner graph. In Section III, a protograph LDPC code, optimized for low iteration operation, is selected to design an experimental system to test different low-resolution ADCs and MIMO configurations. Section V summaries the paper's contributions.

II. System Model

Consider a wireless MIMO coded communication system with M transmit antennas and N receive antennas, as shown in Fig. 1. The information bit block $\mathbf{b}$ of K_c bits is first fed into protograph LDPC code to produce a coded codeword $\mathbf{c}$ of length N_c, and the coding rate is thus $R = K_c/N_c$. Each bit of the coded codeword $\mathbf{c}$ is then modulated by the binary phase shift keying (BPSK) modulator where the modulated symbol s takes values in $\{+1, -1\}$. In one channel use, M modulated symbols are transmitted over M transmit antennas using spatial multiplexing (V-BLAST) scheme [5]. Accordingly, it requires $L = N_c/M$ channel uses to transfer one coded codeword $\mathbf{c}$.

The received signal model is given by

$$\mathbf{r} = \frac{1}{\sqrt{M}}\mathbf{H}\mathbf{x} + \mathbf{w}, \tag{1}$$

where $\mathbf{x} = [x[1], x[2], \cdots, x[M]]^T$ is the transmitted symbol with its elements belong to the BPSK modulation alphabet. $\mathbf{H} \in \mathbb{C}^{N\times M}$ is channel matrix whose entries $h[n][m]$ in the n-th row and m-th column of $\mathbf{H}$. Those elements are modeled as i.i.d complex Gaussian with zero mean and unit variance $\mathcal{CN}(0, 1)$. In this paper, we assume that the perfect channel state information (CSI) is available at the receiver only. The vector $\mathbf{w} = [w[1], w[2], \cdots, w[N]]^T \in C^{N\times 1}$ is complex additive white Gaussian noise vector whose entries follow i.i.d complex Gaussian with zero mean and N_0 variance $\mathcal{CN}(0, N_0)$. Finally, $\mathbf{r} = [r[1], r[2], \cdots, r[N]]^T \in \mathbb{C}^{N\times 1}$ is the received symbol.

In LS-MIMO communication systems, the radio frequency (RF) signal at each receiving antenna is quantized by a pair of low-resolution ADCs - one low-resolution ADC for the real channel and the other for the imaginary channel. In this paper, we consider the low-resolution ADCs in which Q-bit uniform scalar quantizer with $Q = 2, 3, 4, 5$ is used. In prior research works, 1-bit ADC is investigated, see [6] and references therein. But, we defer investigation the 1-bit ADC LS-MIMO systems in the separate research work.

With the assumption of the channel model in (1), the input signal of the Q-bit ADC in Fig. 1, which is the received signal, is a continuous random variable with infinite support. Therefore, the input signal, $r_l[n], l \in \{Re, Im\}$, is first truncated to have finite support in the range $[-R_s, R_s]$ as below:

$$\overline{r_l}[n] = \begin{cases} -R_s, & r_l[n] < -R_s; \\ r_l[n], & -R_s \le r_l[n] \le R_s; \\ R_s, & r_l[n] > R_s. \end{cases} \tag{2}$$

where $\overline{r_l}[n]$ be the truncated version of the received signal $r_l[n]$. The value of R_s is chosen such that the distortion due to the truncation process is negligible. In this work, we adopt the *three-sigma* rule as in [7] to find the truncation limit for the received signal $r[n], n = 1, 2, \cdots, N$ as below:

$$\begin{aligned} R_s &= 3\sigma_{r_{Re}[n]} = 3\sigma_{r_{Im}[n]} \\ &= 3 \times (0.5 + 0.5 \times N_0)^{\frac{1}{2}} \end{aligned} \tag{3}$$

We leave the index n of R_s because we assume the additive noise components and channel gains are i.i.d and the transmit power is equally allocated across transmit antennas. Hence, we have a common truncation limit for all received signals across receive antennas. In case there is a difference in signal strengths across the receiving antennas, the equation (3) is still applicable by normalizing the received signals before sending them to the low-resolution ADC module.

In this paper, we examine the case that the ADC uses the scalar uniform quantizer, [8]. Let $\mathbf{y}$ be the quantized version of the truncated received signal $\bar{\mathbf{r}}$, and we can mathematically write the quantized signal as

$$y_l[n] = I^Q_{n_q} \text{ if } \overline{r_l}[n] \in (g^Q_{n_q-1}, g^Q_{n_q}], l \in \{Re, Im\}, \tag{4}$$

where the representation point $I^Q_{n_q}$, the midpoint of each interval, is expressed

$$I^Q_{n_q} = \frac{g^Q_{n_q} + g^Q_{n_q-1}}{2}, n_q = 1, 2, \cdots, 2^Q. \tag{5}$$

The boundary points are given

$$g^Q_q = -R_s + q\Delta, q = 0, 1, \cdots, 2^Q. \tag{6}$$

In (6), Δ is quantization step size which is given

$$\Delta = \frac{2R_s}{2^Q}. \tag{7}$$

Adopting the additive quantization noise model (AQNM) which is conventionally assumed in the study of the quantized MIMO systems [9], we can model for the quantized signal as below:

$$\mathbf{y} = \varphi\bar{\mathbf{r}} + \mathbf{w}_q \approx \varphi\mathbf{r} + \mathbf{w}_q, \tag{8}$$

where $\varphi = 1 - \rho$, and ρ is the inverse of the signal-to-quantization noise ratio, and $\mathbf{w}_q$ is the additive Gaussian quantization noise vector, which is assumed to be uncorrelated to $\mathbf{y}$. The approximation in (8) is obtained since we assume that the distortion due to the truncation process is negligibly small compared to the quantization noise of the low-resolution ADC. For the uniform scalar quantizer, we approximate the quantization noise as $\Delta^2/12$, [10]. As a result, the signal to quantization noise ratio is approximated by

$$\rho \approx \frac{3}{2^{2b}} \to \varphi \approx 1 - \frac{3}{2^{2b}}. \tag{9}$$

For given channel realization matrix $\mathbf{H}$, the variance of $w_q[n], n = 1, 2, \cdots, N$ is given by

$$\sigma^2_{w_q}[n] = \varphi(1-\varphi)\left(\frac{1}{M}\sum_{m=1}^{M}|h[n][m]|^2 + N_0\right). \tag{10}$$

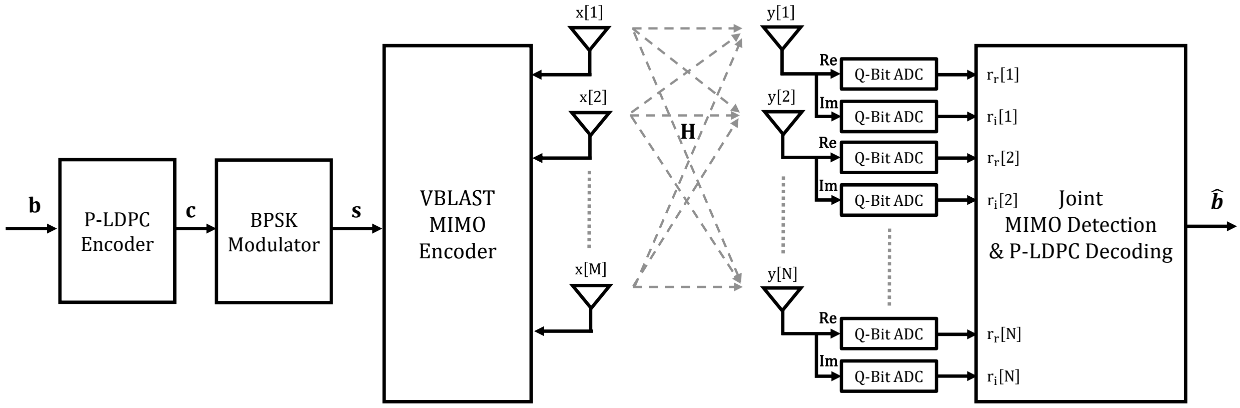

Fig. 1. The MIMO coded communication model.

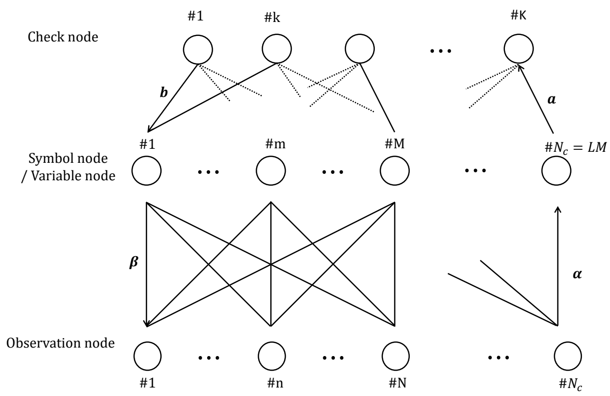

Fig. 2. Double-layered graph for joint detection and decoding.

The quantized received signal $\mathbf{y}$ is fed to the joint MIMO detection and P-LDPC decoding receiver, which is developed in details in the next section.

III. Joint Detection and Decoding Receiver With Soft Interference Cancellation

With a large number of antennas, which is in order of hundred elements, the conventional MIMO detection algorithms such as zero-forcing, minimum mean square error spatial filtering, and sphere decoding are highly complex [4]. The message passing algorithm is a good option to deal with the complexity problem. In the LS-MIMO system with message passing detection and decoding algorithm, there are three operation modes: 1) individual detection and decoding mode, 2) iterative detection and decoding mode, 3) joint detection and decoding mode. The authors in [11] have proved that the joint detection and decoding mode gives the best performance. Hence, we present the message passing algorithm for the joint detection and decoding mode.

The joint detection and decoding algorithm can be presented as the double-layered graph, as shown in Fig. 2. The double-layered graph consists of three types of nodes, namely, $N_c = L \times M$ variable nodes, $K = N_c - K_c$ check nodes, and $N_c = L \times M$ observation nodes. The four types of messages passed over the graph are: $\alpha[n][m]$ is the message passed from the n-th observation node to the m-th variable node, $a[m][k]$ is the message passed from the m-th variable node to the k-th check node, $b[k][m]$ is the message passed from the k-th check node to the m-th variable node, $\beta[m][n]$ is the message passing from the m-th variable node to the n-th observation node. In the following sections, we derive the expressions for those messages.

A. Message Passed From Observation Nodes To Symbol/Variable Nodes

At the n-th observation node, the transmitted symbols are detected, and the extrinsic information (message) is passed to the symbol nodes. In the following, we present the detection algorithm using a message update rule at the n-th observation node.

The quantized received signal on the n-th receive antenna/ observation node can be rewritten as

$$y[n] = \frac{\varphi}{\sqrt{M}} \sum_{m=1}^{M} h[n][m]x[m] + \varphi w[n] + w_q[n] \tag{11}$$

$$= \frac{\varphi}{\sqrt{M}} h[n][m]x[m] + \underbrace{\frac{\varphi}{\sqrt{M}} \sum_{k=1,k\neq m}^{M} h[n][k]x[k]}_{Interference} + \varphi w[n] + \underbrace{w_q[n]}_{QuantizationNoise}$$

where $x[m]$ is the symbol sent on the m-th antenna/ symbol node, $h[n][m]$ is the channel gain from nth transmit antenna to mth antenna, $w[n]$ and $w_q[n]$ the additive white Gaussian noise and the quantization noise, respectively.

As shown in (11), the quantized received signal includes inter-substream interference. The parallel interference cancellation technique is employed to cancel the inter-substream interference. To do so, we first estimate the soft symbol based on the extrinsic LLR transferred from the variable node to the observation node. Denote $\hat{x}[n][m]$ the soft symbol obtained by using the LLR passed from the m-th symbol node to the n-th observation node.

For the BPSK modulation, the soft symbol is given by

$$\hat{x}[n][m] = \tanh\left(\frac{\beta[n][m]}{2}\right), \tag{12}$$

where $\beta[n][m]$ is the extrinsic information passed from the m-th symbol node to the n-th observation node.

The soft symbol in (12) is now used to cancel the interference from the quantized received signal at the n-th antenna for the m-th transmit symbol as below

$$\hat{y}[n][m] = y[n] - \frac{\varphi}{\sqrt{M}} \sum_{k=1,k\neq m}^{M} h[n][k]\hat{x}[n][k]. \tag{13}$$

In general, the soft symbol is an imperfect replica of the transmitted symbol. Therefore, the residual interference remains in the signal $\hat{y}[n][m]$. We approximate this residual interference as additive Gaussian noise. Let $\nu[n][m]$ be the residual interference plus noise component which is given by

$$\nu[n][m] = \frac{\varphi}{\sqrt{M}} \sum_{k=1,k\neq m}^{M} h[n][k](x[n][k] - \hat{x}[n][k]) \tag{14}$$
$$+\varphi w[n] + w_q[n],$$

Based on (14), we can rewrite $\hat{y}[n][m]$ in (13) as below:

$$\hat{y}[n][m] = \frac{\varphi}{\sqrt{M}} h[n][m]x[m] + \nu[n][m]. \tag{15}$$

The power of the residual interference plus noise component, $\nu[n][m]$, is calculated by the following equation

$$\Psi[n][m] = \frac{\varphi^2}{M} \sum_{k=1,k\neq m}^{M} |h[n][k]|^2(1 - |\hat{x}[n][k]|^2) \tag{16}$$
$$+\varphi^2 N_0 + \varphi(1-\varphi)\left(\frac{1}{M}\sum_{m=1}^{M}|h[n][m]|^2 + N_0\right),$$

where φ is given in (9). The message passed from the nth observation node to the mth symbol node is the log-likelihood ratio (LLR) and given by

$$\alpha[n][m] = \ln \frac{\Pr(\hat{y}[n][m]|\mathbf{H}, x[m] = +1)}{\Pr(\hat{y}[n][m]|\mathbf{H}, x[m] = -1)} \tag{17}$$
$$= \frac{4\varphi}{\sqrt{M}\Psi[n][m]} \Re(h^*[n][m]\hat{y}[n][m]).$$

The processing at the n-th observation node is finished by passing the message $\alpha[n][m]$ to the m-th symbol node.

B. Message Passed From Variable Nodes To Check Nodes

The next step is to calculate the extrinsic information to transfer from the m-th symbol node to the k-th check node. The extrinsic information should be included the information provided by the observation nodes except the n-th node to avoid message duplication. The message from the m-th variable node to the k-th check node is given by

$$a[m][k] = \sum_{n\in\mathcal{N}_o(m)} \alpha[n][m] + \sum_{n\in\mathcal{N}_c(m)\setminus k} b[n][m], \tag{18}$$

where $\mathcal{N}_o(m)$ is the set of all observation nodes connected to the m-th variable node and $\mathcal{N}_c(m)$ is the set all of check nodes connected to the m-th variable node.

C. Message Passed From Check Nodes To Variable/Symbol Nodes

The message from the k-th check node to the m-th variable node is given by

$$b[k][m] = \ln\left[\frac{1 + \prod_{n\in\mathcal{N}_v(k)\setminus m}(1 - 2\Pr(x[n] = 1))}{1 - \prod_{n\in\mathcal{N}_v(k)\setminus m}(1 - 2\Pr(x[n] = 1))}\right], \tag{19}$$

where $\mathcal{N}_v(k)$ is the set of variable nodes connected to the k-th check node. The $\tanh(\cdot)$ function is often used to compute $b[k][m]$.

D. Message Passed From Symbol Nodes To Observation Nodes

This step calculates the extrinsic information to transfer from the m-th symbol node to the n-th observation node. The extrinsic information should be included the information provided by the check nodes connected to the m-th symbol node and the observation nodes connected to the m-th symbol node except the n-th node to avoid message duplication. The message from the m-th symbol node to the n-th observation node is thus given by

$$\beta[m][n] = \sum_{k\in\mathcal{N}_c(m)} b[k][m] + \sum_{k\in\mathcal{N}_o(m)\setminus n} \alpha[k][m]. \tag{20}$$

E. The posteriori LLR

The additional task of the m-th symbol node/variable is to calculate a posteriori probability of the symbol $x[m]$ with the messages received from the observation nodes and the check nodes. Then, the posterior probability is used to detect the transmitted symbol. The posteriori LLR, $\Gamma[m]$, of the symbol $x[m]$ can be obtained by summing up all the extrinsic LLRs received from the observation nodes and the check nodes as below

$$\Gamma[m] = \sum_{n\in\mathcal{N}_o[m]} \alpha[n][m] + \sum_{k\in\mathcal{N}_c[m]} b[k][m], \tag{21}$$

where $\mathcal{N}_o[m]$ is the set of all observation nodes connected to the m-th symbol/variable node, and $\mathcal{N}_c[m]$ is the set of all check nodes connected to the m-th symbol node/variable node. Based on the posterior LLR, the transmitted symbol is estimated as

$$\hat{x}[m] = \text{sign}(\Gamma[m]) \tag{22}$$

The iterative detection and decoding process will stop once all check equations are satisfied, or the maximum number of iterations is reached.

IV. Simulation Results

In this section, the computer simulation is carried out to evaluate the performance of the LS-MIMO communication systems with 10×10, 10×20 MIMO configurations. We perform simulations with low-resolution Q-bit ADCs where $Q = 2, 3, 4, 5$.

A non-punctured P-LDPC code of rate $1/2$ in [12] is selected to construct the coded P-LPDC LS-MIMO communication systems. The reason of this selection is that this code was previously optimized for a limited number of iterations, which is suitable for the joint detection and decoding in the LS-MIMO systems. Furthermore, the P-LDPC codes are structural codes, and thus they have lower encoder/decoder complexity than the irregular LDPC codes. The proto-matrix of the non-punctured code is given below:

$$\mathbf{B}_{1/2} = \begin{pmatrix} 1 & 0 & 0 & 0 & 0 & 1 & 3 & 3 \\ 0 & 1 & 0 & 1 & 0 & 0 & 2 & 3 \\ 0 & 0 & 1 & 2 & 1 & 0 & 1 & 3 \\ 1 & 1 & 1 & 0 & 2 & 2 & 0 & 3 \end{pmatrix}_{8\times 4} . \tag{23}$$

The construction of P-LPDC code from the proto-matrix in (23) is identical to the construction in [12].

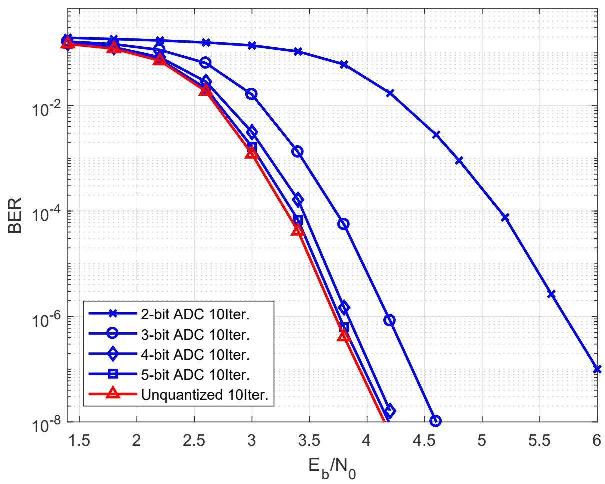

Fig. 3. BER performance: 10×10 LS-MIMO, $R = 1/2$, 10 iterations, information blocklength 1920 bits.

Fig. 3 and Fig. 4 present the simulation results for the 10×10 LS-MIMO configuration. The maximum number of iterations is set 10. One can see that changing the resolution from the 2-bit ADC to the 3-bit ADC leads to huge performance improvement. For instance, if the operation BER = 10^{-6}, the required E_b/N_0 of the 3-bit ADC LS-MIMO system is 4.2 dB while the required E_b/N_0 of the 2-bit ADC system is around 5.7 dB, equivalent to the performance gap of 1.5 dB. If the resolution increase from 3-bit ADC to the 4-bit ADC, the performance gap at $BER = 10^{-6}$ is narrowed to 0.5 dB. The performance gap between 4-bit ADC and 5-bit ADC is very marginal, and they both almost touch the performance curve of the high-resolution LS-MIMO system. This fact suggests that

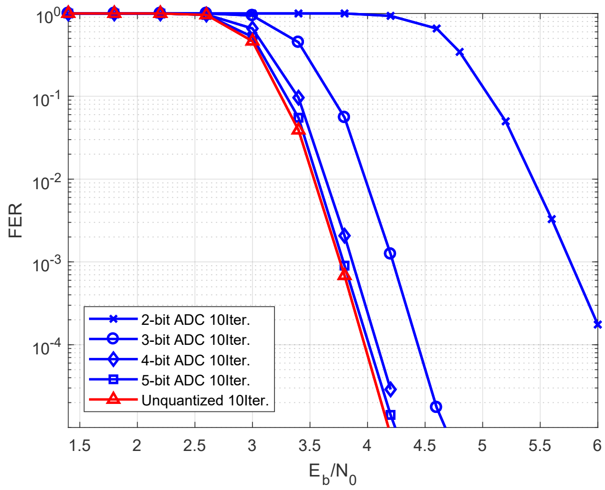

Fig. 4. BER performance: 10×10 LS-MIMO, $R = 1/2$, 10 iterations, information blocklength 1920 bits.

the 4-bit ADC is a sensible choice to achieve both excellent BER performance and cut down the power consumption and complexity of the RF processing block.

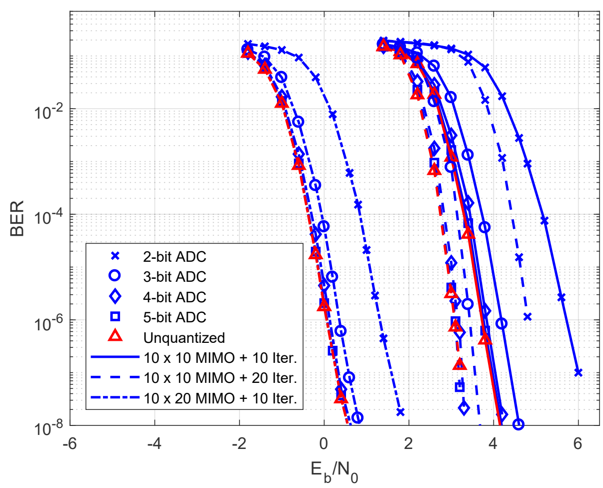

Fig. 5. BER performance: Impact of MIMO configuration and maximum number of iterations.

We examine the impact of the MIMO configuration and the maximum number of iterations on the performance of the low-resolution ADCs, and the results are shown in Fig. 5 and Fig. 6. Increase either the maximum number of iterations or the receive antennas cannot help to lessen the performance gap between 2-bit ADC and 4-bit ADC. Further, the experiment results strengthen the claim that 4-bit ADC is the best resolution for LS-MIMO systems to decrease the complexity. As we can see, there is a very tiny performance difference between 4-bit ADC system and the high-resolution/unquantized system.

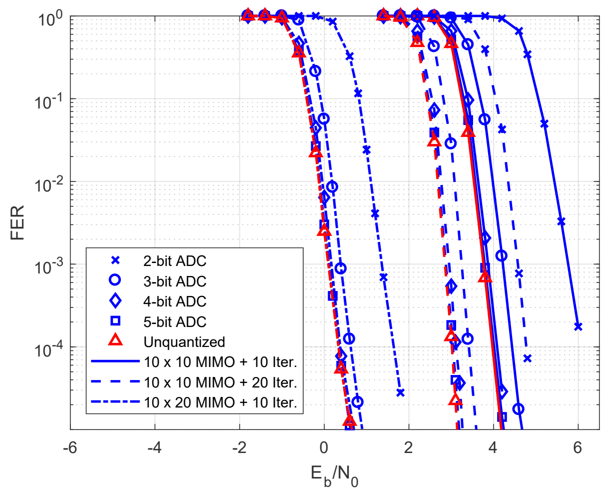

Fig. 6. FER performance: Impact of MIMO configuration and maximum number of iterations.

V. Conclusions

We develop a new algorithm of the joint detection and decoding receiver for LS-MIMO communication systems with low-resolution ADC. The new algorithm allows us to investigate the performance of different low-resolution ADC. Our results indicate that 4-bit ADC is the best choice to achieve complexity and performance.

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