Gain Design and Power Allocation for Overloaded MIMO-OFDM Systems with Channel State Information and Iterative Multiuser Detection

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***Abstract*—This paper deals with multiuser MIMO-OFDM systems in which single-antenna users transmit encoded data to a common receiver employing iterative multiuser detection. We focus on uplink communications in overloaded scenarios, i.e. the feasibility for the system to support a number of users larger than the number of receive antennas. We assume that perfect channel state information is available at the receiver, and perfect, analog and instantaneous feedback from the receiver to the users. We show how simple power allocation at user location, based on gain design for asymptotic separation at receiver location, makes overloading feasible. Numerical simulations show the benefit of the proposed approach.**

I. INTRODUCTION

Broadband wireless communications for high-data rate multimedia applications receive large interest in modern applied research. The most promising technology is the combination of Multiple-Input Multiple-Output (MIMO) systems with Orthogonal Frequency Division Multiplexing (OFDM) modulation [6], [15]. MIMO-OFDM systems simultaneously mitigate inter-symbol interference and enhance system capacity, currently representing the physical-layer for many wireless communications standards.

Turbo processing, proposed for capacity-achieving coding [2], has been extended to a large number of problems in communications literature, e.g. channel equalization [7], MultiUser Detection (MUD) [17]. Iterative receivers achieve near-optimum performance with limited complexity by decoupling MUD and single-user decoding into separate problems exchanging results via soft information [3]. Advanced receiver architectures include also channel estimation within the iterative loop, e.g. refer to [18] for multicarrier Code-Division Multiple Access (CDMA) systems and to [14] for MIMO-OFDM systems.

Unequal power distribution at the receiver, in the following referred to as gain design, has been shown [4] to be optimum for iterative receivers in overloaded scenarios, i.e. the number of transmit antennas exceeds the number of receive antennas. The gain design problem has been formulated in terms linear programming based on density evolution techniques, and requires the knowledge of the Bit Error Rate (BER) vs. Signal-to-Noise Ratio (SNR) characterization of the users' codes. Simple algorithms for gain design in overloaded CDMA systems were proposed and analyzed in [10], [11]. They only require the knowledge of the users' signatures and are based on Asymptotic Effective Energy (AEE) and user separation [16].

A MIMO-OFDM system with iterative MUD based on Interference Cancellation (IC) and Minimum Mean Square Error (MMSE) filtering was proposed in [12] and tested over real-workd measured channels in [13], however overloaded scenarios were not investigated. Overloading may be interesting in order to increase the number of simultaneous active users but has to be properly designed in order to manage effectively the increased interference. We focus here on the iterative receiver for MIMO-OFDM systems proposed in [12] with the following differences: (i) we assume that the channel is quasi-static; (ii) we assume that perfect Channel State Information (CSI) is available at the receiver; i.e. we ignore the channel estimation problem for sake of simplicity; (iii) we assume perfect, analog and instantaneous feedback from the receiver to the users; (iv) we focus on overloaded scenarios. More specifically, we derive an algorithm along the same lines of [11] suited for the considered MIMO-OFDM system. The goal is to show how CSI at the receiver allows for simple gain design, and power allocation at the transmitter based on it makes overloading feasible. Rather than vector precoding [5], we focus on power allocation (i.e. diagonal precoding) in order to take into account scenarios with multiple independent transmitters each provided with one single antenna.

The paper is organized as follows: the mathematical model for the received signal is described in Sec. II; the structure of the receiver is described in Sec. III; in Sec. IV we describe the algorithm for gain design controlling power allocation; Sec. V shows and compares the performance of systems with and without gain design and power allocation; some concluding remarks are given in Sec, VI.

Notation - Column vectors (resp. matrices) are denoted with lower-case (resp. upper-case) bold letters, with a_n denoting the nth element of the vector $\boldsymbol{a}$ and $A_{n,m}$ denoting the (n, m)th element of the matrix $\boldsymbol{A}$; $\mathrm{diag}(\boldsymbol{a})$ denotes a diagonal matrix whose main diagonal is $\boldsymbol{a}$; $\boldsymbol{I}_N$ is the $N \times N$ identity matrix;

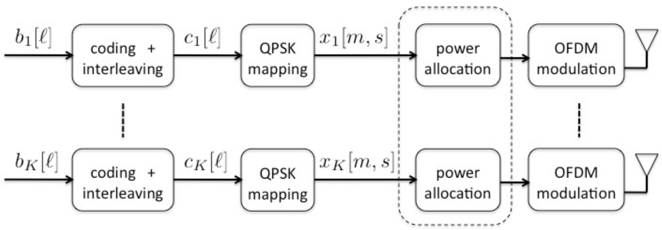

Fig. 1. Block diagram of the transmitter.

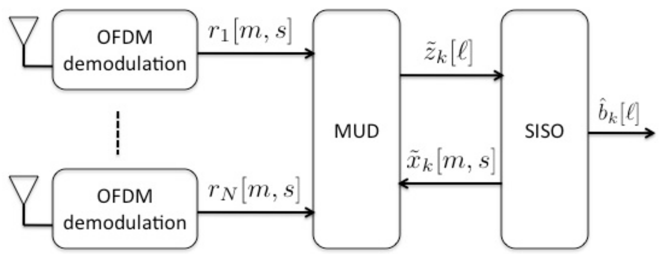

Fig. 2. Block diagram of the receiver.

$\boldsymbol{i}_N^{(n)}$ denotes the nth column of $\boldsymbol{I}_N$; $\mathbf{1}_N$ denotes an N-length column vector whose entries are 1; $(\cdot)^{\mathrm{T}}$ and $(\cdot)^{\mathrm{H}}$ denote transpose and conjugate transpose operators, respectively; $|a|$ denotes the modulus of a; j is the imaginary unit; $\mathcal{N}(\mu, \sigma^2)$ denotes a normal distribution with mean μ and variance σ^2; $\mathcal{N}_{\mathbb{C}}(\boldsymbol{\mu}, \boldsymbol{\Sigma})$ denotes a circular symmetric complex normal distribution with mean vector $\boldsymbol{\mu}$ and covariance matrix $\boldsymbol{\Sigma}$; the symbol $\sim$ means "distributed as".

II. Signal Model

We consider K users, each equipped with one single antenna, transmitting independent data to a common receiver equipped with N antennas, employing OFDM modulation over the same set of M subcarriers. The scenario corresponds to a MIMO-OFDM system with K transmit antennas, N receive antennas, and M subcarriers, in which "transmit antenna" and "user" are equivalent terms. We focus here on system design for overloaded scenarios, i.e. the number of users is larger than the number of receive antennas ($K > N$). We consider slow-fading scenarios modeled via quasi-static channels, i.e. different consecutive OFDM blocks undergo the same set of fading coefficients (block fading).

The transmission is frame oriented and the block diagram at user location is shown in Fig. 1. Each frame, made of S OFDM blocks, contains one single codeword (spanning time and frequency domains). The bit stream is divided in groups of L_b source bits, each producing a frame of L code bits via convolutional coding and interleaving. The bits of the frame are mapped into symbols via Quadrature-Phase Shift Keying (QPSK) modulation [8], thus each frame contains $L_x = L/2$ QPSK symbols. Assuming that L_x is an integer multiple of M, the frame is divided into $S = L_x/M$ blocks and each block gives rise to an OFDM block of M symbols to be transmitted over the wireless channel via the M subcarriers.

In the following, for the generic frame, $b_k[\ell]$ and $c_k[\ell]$ denote the ℓth source bit and the ℓth code bit to be transmitted by the kth user; $x_k[m,s]$ denotes the (frequency domain) symbol transmitted by the kth user over the mth subcarrier during the sth OFDM block; $H_{n,k}[m]$ denotes the (frequency domain) channel coefficient between the kth user and the nth receive antenna over the mth subcarrier during the transmission of the whole set of S OFDM blocks in a single frame; $w_n[m,s]$ denotes the (frequency domain) additive noise at the nth receive antenna over the mth subcarrier during the sth OFDM block; $r_n[m,s]$ denotes the (frequency domain) received signal at the nth receive antenna over the mth subcarrier during the sth OFDM block.

We denote the transmitted vector, the channel matrix, the noise vector, and the received vector as

$$\begin{aligned}
\boldsymbol{x}[m,s] &= (x_1[m,s], \ldots, x_K[m,s])^{\mathrm{T}} \ , \\
\boldsymbol{H}[m] &= \begin{pmatrix} H_{1,1}[m] & \ldots & H_{1,K}[m] \\ \vdots & \ddots & \vdots \\ H_{N,1}[m] & \ldots & H_{N,K}[m] \end{pmatrix} , \\
\boldsymbol{w}[m,s] &= (w_1[m,s], \ldots, w_N[m,s])^{\mathrm{T}} \sim \mathcal{N}_{\mathbb{C}}(\mathbf{0}, \sigma_w^2 \boldsymbol{I}_N) \ , \\
\boldsymbol{r}[m,s] &= (r_1[m,s], \ldots, r_N[m,s])^{\mathrm{T}} \ ,
\end{aligned}$$

and assume that the length of the cyclic prefix exceeds the channel delay spread. The discrete-time model for the received signal is then

$$\boldsymbol{r}[m,s] = \boldsymbol{H}[m]\boldsymbol{x}[m,s] + \boldsymbol{w}[m,s] \ , \qquad (1)$$

and the channel vector from the kth user is

$$\boldsymbol{h}_{(k)}[m] = \boldsymbol{H}[m]\boldsymbol{i}_K^{(k)} \ .$$

In presence of power allocation at user location, the discrete-time model for received signal is changed as

$$\boldsymbol{r}[m,s] = \boldsymbol{H}[m]\mathrm{diag}(\boldsymbol{a}[m])\boldsymbol{x}[m,s] + \boldsymbol{w}[m,s] \ , \qquad (2)$$

where $\boldsymbol{a}[m]$ is the gain design solution for the mth subcarrier.

III. Receiver Architecture

Transmissions from the users combine at each receive antenna and are processed according to the receiver model shown in Fig. 2. Perfect synchronization among users is assumed. Due to OFDM robustness to time asynchrony, this assumption is valid as long as synchronization errors do not exceed the length of the cyclic prefix.

To simplify notation, we ignore the dependence on the subcarrier m and on the OFDM block s. Each OFDM block is demodulated and sent to the iterative decoder, composed of a MUD block and a Soft-Input-Soft-Output (SISO) block. The former aims to separate each user's contribution from the aggregate received signal, the latter to decode the information sent by each user. The multiuser detector and the SISO decoders exchange extrinsic information on symbols x_k, denoted $\tilde{x}_k$ when fed to the MUD block and $\tilde{z}_k$ when fed to the SISO block. It is worth noticing that SISO decoders preprocess $\{\tilde{z}_k[1], \ldots, \tilde{z}_k[L_x]\}$ via demapping and deinterleaving, and post-process $\{\tilde{x_k}[1], \ldots, \tilde{x_k}[L_x]\}$ via interleaving and mapping. Secs. III-A and III-B provide processing equations for the signal model (1), the same expressions apply to the signal model (2) replacing $\boldsymbol{H}$ with $\boldsymbol{H}\text{diag}(\boldsymbol{a})$.

A. MUD

The MUD block implements multiuser detection via IC and MMSE filtering [12], [17]. The received signals are processed separately for each subcarrier and OFDM block. IC is performed using $\tilde{\boldsymbol{x}}$ from the SISO decoders. The residual term from IC for the kth user

$$\tilde{\boldsymbol{r}}_{(k)} = \boldsymbol{r} - \boldsymbol{H}(\tilde{\boldsymbol{x}} - \tilde{x}_k \boldsymbol{i}_K^{(k)}) ,$$

is then MMSE filtered, to reduce noise and multiaccess interference, giving [12] the extrinsic information

$$\tilde{z}_k = \boldsymbol{g}_{(k)}^H \tilde{\boldsymbol{r}}_{(k)} ,$$

where

$$\boldsymbol{g}_{(k)}^H = \frac{\boldsymbol{h}_{(k)}^{\mathrm{H}} \left(\boldsymbol{H}\boldsymbol{V}_{(k)}\boldsymbol{H}^{\mathrm{H}} + \sigma_w^2 \boldsymbol{I}_N\right)^{-1}}{\boldsymbol{h}_{(k)}^{\mathrm{H}} \left(\boldsymbol{H}\boldsymbol{V}_{(k)}\boldsymbol{H}^{\mathrm{H}} + \sigma_w^2 \boldsymbol{I}_N\right)^{-1} \boldsymbol{h}_{(k)}} ,$$

and $\boldsymbol{V}_{(k)} = \text{diag}\left((1 - |\tilde{x}_1|^2, \ldots, 1 - |\tilde{x}_{k-1}|^2, 1, 1 - |\tilde{x}_{k+1}|^2, \ldots, 1 - |\tilde{x}_K|^2)\right)$.

B. SISO Decoding

The SISO block implements single-user soft decoding via the BCJR algorithm [1]. After collecting $\{\tilde{z_k}[1], \ldots, \tilde{z_k}[L_x]\}$, each user can be decoded independently using turbo decoding [2] based on the log-domain BCJR algorithm [9]. The SISO decoder for the kth user is based [17] on the model

$$\tilde{z}_k = \mu_k x_k + v_k ,$$

where $v_k \sim \mathcal{N}_{\mathbb{C}}(0, \eta_k^2)$, $\mu_k = 1$ and

$$\eta_k^2 = \frac{1}{\boldsymbol{i}_K^{(k)\mathrm{T}} \left(\boldsymbol{H}^{\mathrm{H}}\boldsymbol{H} + \sigma_w^2 \boldsymbol{I}_N\right)^{-1} \boldsymbol{H}^{\mathrm{H}} \boldsymbol{h}_{(k)}} .$$

IV. Gain Design and Power Allocation

The design of received-gain distributions allowing for separation and successful decoding of multiple users while minimizing the total power has been studied in [11] using the AEE as a constraint for the quality of service. The AEE is defined as the energy required by the matched-filter detector in a single-user scenario to achieve the same BER that the user experience in the multiuser scenario, i.e.

$$\text{AEE} = \lim_{\sigma_w \to 0} \sigma_w^2 \left(\mathcal{Q}^{-1}(\text{BER}(\sigma_w^2))\right)^2 ,$$

where $\mathcal{Q}(x) = \int_x^\infty \frac{\exp\left(-t^2/2\right)}{\sqrt{2\pi}} dt$. AEE is a measure of multiuser interference in absence of noise. Constraints on AEE allow to scale users constellation in order to avoid error floor in the performance.

We define the kth margin (u_k) as the square root of the positive kth user's AEE, and we define $\tilde{\boldsymbol{g}}_{(k)}$ analogously to $\boldsymbol{g}_{(k)}$ but replacing $\boldsymbol{V}_{(k)}$ with $\boldsymbol{I}_K$. The use of $\tilde{\boldsymbol{g}}_{(k)}$ is motivated as gain design is performed just once for the whole frame and is not dependent on the transmitted symbols and their estimates. From the expression of the kth user's AEE in the case of decision feedback decoders [16], noticing that $\tilde{\boldsymbol{g}}_{(k)}^H \boldsymbol{h}_{(k)} = 1$, we get

$$u_k = \frac{1}{\sqrt{\tilde{\boldsymbol{g}}_{(k)}^H \tilde{\boldsymbol{g}}_{(k)}}} \left(a_k - \sum_{\ell=k+1}^{K} a_\ell \left|\tilde{\boldsymbol{g}}_{(k)}^H \boldsymbol{h}_{(\ell)}\right|\right) , \tag{3}$$

A set of margins (or equivalently of AEEs) thus provides the set of gains. More specifically, defining a diagonal matrix $\boldsymbol{C}$ and an upper unitriangular matrix $\boldsymbol{D}$, both of size $K \times K$, such that

$$C_{k,\ell} = \begin{cases} \sqrt{\tilde{\boldsymbol{g}}_{(k)}^H \tilde{\boldsymbol{g}}_{(k)}} & \ell = k \\ 0 & \ell \neq k \end{cases} ,$$

and

$$D_{k,\ell} = \begin{cases} 0 & \ell < k \\ 1 & \ell = k \\ -|\tilde{\boldsymbol{g}}_{(k)}^{\mathrm{H}} \hat{\boldsymbol{h}}_{(\ell)}| & \ell > k \end{cases} ,$$

the gain design solution, from (3), with equal and positive margin per user (i.e. assuming $u_k = \beta > 0$) satisfies

$$\boldsymbol{D}\boldsymbol{a} = \beta \boldsymbol{C} \boldsymbol{1}_K .$$

We chose β in order to keep constant the total transmit power of the system, i.e. $\boldsymbol{a}^{\mathrm{T}}\boldsymbol{a} = K$, thus defining $\boldsymbol{b} = \boldsymbol{D}^{-1}\boldsymbol{C}\boldsymbol{1}_K$ we get

$$\boldsymbol{a} = \sqrt{\frac{K}{\boldsymbol{b}^{\mathrm{T}}\boldsymbol{b}}} \boldsymbol{b} .$$

V. Simulation Results

Numerical simulations were run using MATLAB software to obtain BER-vs-SNR curves for various systems. Denoting $P_e(k)$ the BER for the kth user and T_s the OFDM block duration, the throughput (ρ_k) of the kth user is computed as

$$\rho_k = \frac{L_b}{ST_s} \left(1 - P_e(k)\right)^{L_b} .$$

Results shown here refer to systems with $M = 32$ subcarriers and $S = 32$ OFDM blocks per frame (with OFDM-block duration of $T_s = 4\mu$s) thus corresponding to $L = 2048$ code bits per frame. Code bits are generated using a rate-$1/2$ recursive systematic convolutional encoder with generators $(7, 5)_8$ and random interleaving [8]. Two tail bits enforce the final state of the convolutional encoder into 1, thus each frame contains $L_b = 1022$ source bits. The number of iterations at the receiver was set to 5.

Channel coefficients with unitary mean power have been synthetically generated according to Rayleigh fading, assuming statistical independence among antennas and subcarriers. Performance were averaged over 500 realizations. Perfect CSI is available at receiver location in order to run the gain design algorithm. The gain design solution, i.e. the vectors $\boldsymbol{a}[m]$ for $m = 1, \ldots, M$, are assumed perfectly known at the user location, i.e. we assume perfect analog and instantaneous feedback from the receiver to the users. Performance with and without use of gain design at the receiver and corresponding power allocation at the transmitter were compared.

We considered three system setups (although they are not to be considered an exhaustive set of possible scenarios): (i) *fully-loaded* with $K = N$; (ii) *feasible-overloaded* with $K = (3/2)N$; (iii) *unfeasible-overloaded* with $K = 2N$. Results for the unfeasible-loaded setup are not shown: simulations confirmed that overloading (using or not gain design) with $K/N \geq 2$ leads to severe degradation [5]. Results for the feasible-overloaded setup are used to analyze a scenario in which $K = (3/2)N$ users transmit and interfere simultaneously to a receiver with N antennas. The receiver decodes each user only on the basis of the spatial diversity thus according to a space-division mechanism. Systems with and without using of gain design were simulated. Results for the fully-loaded setup are used to build a comparison, i.e. a scenario in which $K = (3/2)N$ users transmit to a receiver with N antennas but not simultaneously. In this case only N users are simultaneously active in order to limit the interference at the receiver, while $K - N$ users periodically stop transmissions according to a time-division mechanism.

Fig. 3 refers to systems with $N = 2$ and $K = 3$, while Fig. 4 refers to systems with $N = 4$ and $K = 6$. Figs. 3(a) and 4(a) show the BER performance of the single user (within active periods only) in the case of: fully-loaded setup with time-division mechanism (TD), overloaded setup with space-division mechanism and without gain design and power allocation (SD), overloaded setup with space-division mechanism and with gain design and power allocation(GD-SD). The last case is represented by three or six curves because gain design creates unequal power distribution among users, thus each user experiences different performance. Figs. 3(b) and 4(b) show the corresponding system throughput (obtained as the aggregate of the active-users throughput) as well as the single-user throughput for the GD-SD case. It is worth noticing that BER performance cannot be directly compared because they refer to a different number of active users: comparing the aggregate system throughput is more reasonable.

It is apparent how gain design and power allocation affect the system performance. In absence of gain design and power allocation, the system achieves excellent performance if time-division is used in order to undergo a fully-loaded setup, while performance are degraded in overloaded setup because overloading contrasts the capability of the iterative receiver and nulls the turbo effect. In presence of gain design and power allocation, overloading may be feasible depending on K/N and SNR. The presence of the waterfall region in the BER curves, even in the case of overloaded scenarios, confirms the effectiveness of using the AEE as a constraint for the quality-of-service of the users. Comparing the aggregate throughput, it is apparent how in the large SNR range overloading with gain design and power allocation is more convenient than full-loading with time-division from a system performance point of view (in the two examples the aggregate throughput is 8 Mbps and 16 Mbps higher).

Finally, some considerations on the unequal power distribution among users are needed. In the GD-SD simulations, we considered a fixed ordering among the users, i.e. user 1 is always the strongest (regardless of the channel realization) thus experiencing better performance, then user 2 follows, and so on. Fairness issues require a periodic rotation of the ordering among the users, such that each user stay in each generic position $1/K$ of the time, thus undergoing the same performance in the long term. We stress that, at least for this system, the periodic rotation should be constrained such that each user stay in the same position for the entire codeword. Simulation results, not shown for brevity, have pointed out that assigning a different position to the same user within the same codeword (i.e. different user ordering for different subcarrier) leads to performance degradation. Noticing that MUD exploits the soft information from decoding the whole codewords, our explanation is the following. When a fixed ordering is employed for the entire codeword, symbols from user 1 dominate other symbols and are easily decoded. Once decoded, their contribution to the received signal is removed by the iterative decoder. The same concept then applies to symbols from user 2 and so on. Without a fixed ordering in the single codeword, there is no strongest user, thus no guarantee that the iterative receiver decode and remove any user.

VI. Conclusion

Overloading in MIMO-OFDM systems weakens the performance with error floors spanning the whole SNR range. A simple scheme for MIMO-OFDM systems with iterative receivers has been proposed in order to support overloading in uplink communications. The scheme exploits gain design at the receiver, based on AEE, and power allocation at the transmitter. The algorithm for gain design only requires CSI at the receiver. Results from numerical simulations in terms of BER and throughput vs. SNR, assuming perfect analog and instantaneous feedback, show how practical scenarios exist in which overloading is attractive .

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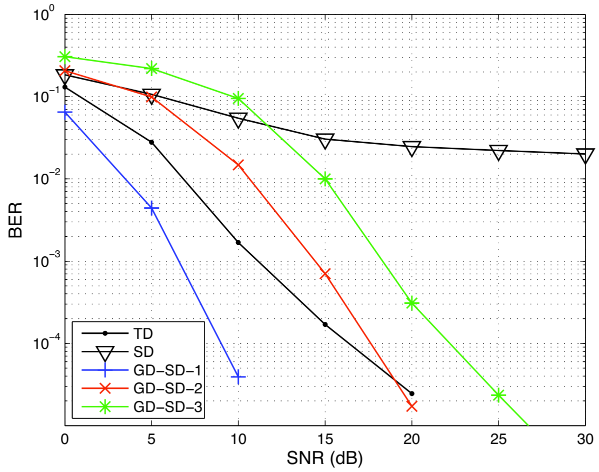
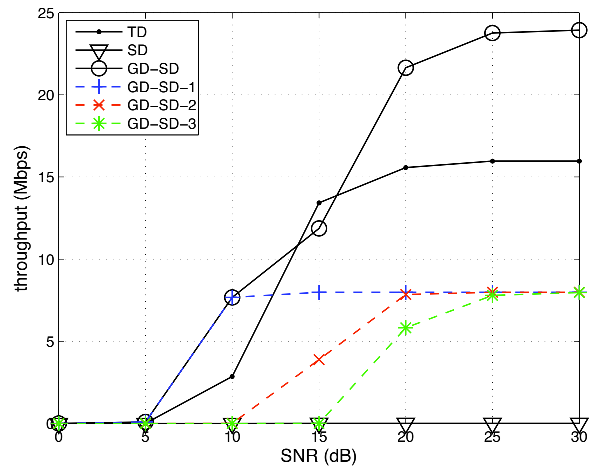
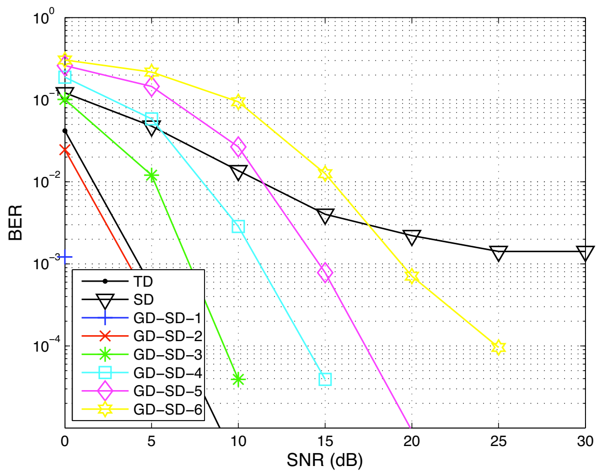


(a) BER vs. SNR.

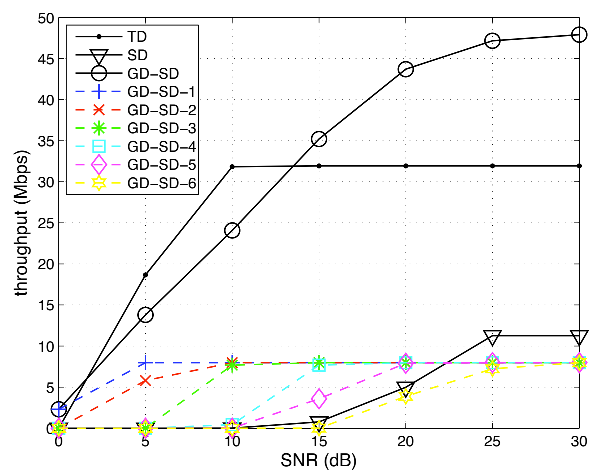


(b) Throughput vs. SNR.

Fig. 3. Systems with $N = 2$ receive antennas and $K = 3$ users (each with $M = 32$ subcarriers per OFDM block and $S = 32$ OFDM blocks per frame).



(a) BER vs. SNR.



(b) Throughput vs. SNR.

Fig. 4. Systems with $N = 4$ receive antennas and $K = 6$ users (each with $M = 32$ subcarriers per OFDM block and $S = 32$ OFDM blocks per frame).

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