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ADAPTIVE SELECTION OF ANTENNAS FOR OPTIMUM TRANSMISSION IN SPATIAL MODULATION FOR MULTI CARRIER SYSTEMS

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Abstract
In this paper, we tend to propose associate degree optimum transmit structure for abstraction modulation (SM), a singular single-stream multiple- input multiple-output (MIMO) transmission technique. As a three-dimensional modulation theme, SM allows a trade-off between the scale of the abstraction constellation diagram and also the size of the signal constellation diagram. Supported this reality, the novel technique, named transmission optimized abstraction modulation (TOSM), selects the most effective transmit structure that minimizes the typical bit error likelihood (ABEP). not like the normal antenna choice strategies, the planned technique depends on applied mathematics channel state data (CSI) rather than instant CSI, and feedback is merely required for the best range of transmit antennas. The overhead for this, however, is negligible. Additionally, TOSM has low machine complexness because the improvement drawback is resolved through an easy closed form objective operate with one variable. Simulation results show that TOSM considerably improves the performance of SM at numerous channel correlations. Forward physicist attenuation channels, TOSM outperforms the first SM by up to nine sound unit. Moreover, we tend to propose one radiofrequency (RF) chain base station (BS) supported TOSM that achieves low hardware complexness and high energy potency. Compared with multi-stream MIMO schemes, TOSM offers associate degree energy saving of a minimum of fifty-six within the continuous transmission mode, and sixty-two within the discontinuous transmission mode.

INTRODUCTION

As foundation Antenna can't be bought, is geared toward coupling the general public sector and can stay operating within the public sphere and domain. Antenna has no growth situation, we tend to believe the networked society, within the network economy, sharing and cooperating with alternative agencies, organizations and services. Antennas financial gain comes for a considerable half by reselling its services via others, however conjointly reselling the services of others via Antenna. Back within the sensible past days, an oversized upper side antenna was seen as a standing image. Today, sensible phones, tablets and GPS units have conditioned shoppers to expect reliable wireless services in terribly tiny packages. These dramatic changes in technology and client preferences, alongside the switch from analog to digital signals, have created a high demand for quality, over-the-air, digital HDTV antennas. The styles for those previous, upper side TV antennas area unit decades previous and contains a configuration in an exceedingly horizontal “fish bone” vogue, with “arms” of varied lengths, giving the reception of a broader vary of frequencies. Although antenna analysis and engineering have seen radical advancement solver the years, manufactures of video equipment have largely cursed theses previous s tyles for economic reasons Since the transition to digital signals, most digital frequencies area unit broadcast

MICROSTRIP ANTENNAS

Intel communication, there square measure many kinds of micro strip antennas (also referred to as written antennas) the most typical of that is that the micro strip patch antenna or patch antenna. A patch antenna may be a narrow band, wide beam Associate in Nursingtenna unreal b yet ching the antenna part pattern in metal trace guaranteed to an insulating stuff substrate, like a computer circuit board, with never-ending metal layer guaranteed to the alternative aspect of the substrate that forms a ground plane. Common micro strip antenna shapes square measure sq. rectangular, circular and elliptical; however any continuous form is feasible. Some patch antennas don't use a stuff substrate and instead square measure manufactured from a metal patch mounted higher than ground plane victimization stuff spacers; the ensuing structure is a smaller amount rugged however contains a wider information measure. as a result of such antennas have a really low profile, square measure automatically rugged and may be formed to evolve to the arched skin of a vehicle, they're typically mounted on the outside of craft and space vehicle, or square measure incorporated into mobile radio communications devices.

LITERATURE SURVEY

Simulating band weakening channels, multiple input multiple-output (MIMO) channels, and diversity-combined weakening channels typically demands the generation of multiple unrelated Lord Rayleigh weakening waveforms. throughout this two applicable parameter computation ways that, specifically technique ology the manoeuvre the strategy of actual scientist unfold (MEDS) and L pnorm method (LPNM), for settled sum-of-sinusoids (S.O.S) channel simulators unit of measurement investigated to confirm the uncorrelatedness between fully totally different simulated Lord Rayleigh weakening processes. Numerical and simulation results show that the following settled S.O.S channel machine can accurately and efficiently reproduce all the desired math properties of the reference model. The generation of multiple unrelated Lord Rayleigh weakening waveforms is often required for simulating band weakening channels, multiple-input multiple-output (MIMO) channels, and diversity combined weakening channels. It's so of nice significance to develop channel simulators capable of accurately and efficiently simulating multiple unrelated Lord Rayleigh weakening processes. Jakes' settled sum-of sinusoids (S.O.S) channel machine has extensively been applied to the simulation of Lord Rayleigh weakening channels. Thus on get multiple unrelated Rayleigh weakening signals, privy and totally different researchers have investigated fully alternative ways to parameterize the underlying settled S.O.S channel simulators.

SYSTEM MODEL

MIMO

The use of multiple-input multiple-output (MIMO) wireless systems has been shown to considerably increase the spectral potency of point-to-point wireless links. The performance limits of MIMO multiple access (MA) and broadcast channels area unit significantly less understood and have recently attracted important interest. Contributions and regard to previous demonstration this project, we tend to specialize in MIMO MA channels with frequency-selective weakening (spatially related to at the receiver) presumptuous good channel state data (CSI) at the multiple-antenna receiver and no channel data at the multiple antenna transmitters. Every of the users employs orthogonal frequency-division multiplexing(OFDM). The ensuing family of MA schemes encompasses the acute cases of frequency-division multiple access (FDMA), wherever every tone is assigned to at the most one user, and code-division multiple access (CDMA), wherever every tone is assigned to any or all the users. Following, we tend to use the term CDMA only to point that each one the users occupy the whole frequency band; the impact of redundancy-introducing spreading won't be thought-about. Besides developing a framework for the analysis of MA schemes realizing a variable quantity of collision in signal house, our main contributions will be summarized as follows.

We show that, regardless of spatial receive weakening correlation and also the range of antennas, the random capability region obtained for a totally collision-based (CDMA) theme is associate degree outer certain to the random capability region for the other MA strategy, wherever users collide solely on subsets of the accessible tones or don't collide in any respect (FDMA). This result generalizes the most end in showing the strict superiority of CDMA over FDMA (two extremes of our MA scheme) in single-antenna frequency-selective weakening channels. Any results examination the performance of CDMA and FDMA within the singleantenna case will be found in. specially, and discuss a TDMA theme implementing a variable quantity of collision and show that within the presence of co channel interference full collision can normally be suboptimal from a capability point-of view. Finally, we tend to note that the capability region for settled MA channels with international intelligence agency (with good CSI each at transmitter and receiver) has been computed in. it's what is more shown in, that FDMA, with optimally elite frequency bands for every user, achieves the overall capability of the Gaussian MA channel with international intelligence agency. Minimizing the quantity of collision in signal house is fascinating as this minimizes the receiver quality incurred by having to separate the meddlesome (colliding) signals. We tend to study the joint decryption performance loss because of suboptimum(i.e., not totally collision based)multiple accessing during systematic fashion. It's initial shown that within the low Signal to Noise quantitative relation (SNR)regime the quantity of signal house collision encompasses a vanishing impact on the random capability. Therefore, we tend to specialize in the high SNR regime associate degreed perform an analysis supported the notion of the multiplexing gain region. Our analysis indicates that for wealthy scattering and a tiny low range of receive antennas, little collision is required to appreciate a big fraction of the accessible add capability. An in depth discussion of this side is provided for the 2-user case. Extending results reported in for the single-antenna case, we tend to any quantify the performance distinction between CDMA (full collision) and

SIGNAL AND CHANNEL MODELS AND MULTIPLE ACCESS SCHEME

We introduce the MA MIMO channel and signal model and the MA scheme. A. MA MIMO Channel Model. We consider a MA MIMO channel with users, each of which is equipped with transmit antennas; the receiver employs antennas. The individual user's channels are assumed1Eigen values of Hermitan matrices are sorted in descending order with denoting the largest Eigen value frequency-selective with the th user's matrix-valued transfer function given by we restrict ourselves to purely Rayleigh block-fading channels with the elements of being circularly symmetric zero mean complex Gaussian random variables, constant within a block and changing in an ergodic fashion from block to block. Furthermore, the matrices are assumed to be uncorrelated across users (indexed by) and across taps (indexed by). We also assume spatially uncorrelated fading at the transmit arrays. Spatial fading correlation at the receive array is modeled by decomposing the taps according to with denoting a random matrix with i.i.d .entries and is the receive correlation matrix for the th tap of the th user. We note that the power delay profiles of the individual channels are incorporated into the correlation matrices . This channel model corresponds to a non-line-of-sight propagation scenario where the individual users are located in rich scattering environments(accounted for by uncorrelated spatial fading at the transmitters). Finally, we assume that the receiver knows all the channels perfectly whereas the transmitters have no channel state information. B. Signal Model We assume that each of the users employs OFDM with tones and the length of the cyclic prefix (CP) satisfies .The receive signal vector for the th tone is consequently given by where with denoting the data symbol transmitted by the th user from the th antenna on the th tone and is white noise uncorrelated across tones (indexed by). The power allocated to the tone of the th user is denoted as and the total transmit power of user is given by. We next state an important property which will be used frequently in what follows. Under the assumptions using we can conclude that the channel matrices for user are identically distributed for all tones, i.e. In particular, we have where and is a random matrix with entries.

Impact Of Pilot Contamination on Classical Least Squares and Minimum Mean Square Error Algorithms In Multicell Multiuser MIMO Systems

Massive MIMO communication systems, by virtue of utilizing very large number of antennas, have a potential to yield higher spectral and energy efficiency in comparison with the conventional MIMO systems. In this paper, we consider uplink channel estimation in massive MIMO-OFDM systems with frequency selective channels. With increased number of antennas, the channel estimation problem becomes very challenging as exceptionally large number of channel parameters have to be estimated. We propose an efficient distributed linear minimum mean square error (LMMSE) algorithm that can achieve near optimal channel estimates at very low complexity by exploiting the strong spatial correlations and symmetry of large antenna array elements. The proposed method involves solving a (fixed) reduced dimensional LMMSE problem at each antenna followed by a repetitive sharing of information through collaboration among neighboring antenna elements. To further enhance the channel estimates and/or reduce the number of reserved pilot tones, we propose a data-aided estimation technique that relies on finding a set of most reliable data carriers. We also analyse the effect of pilot contamination on the mean square error (MSE) performance of different channel estimation techniques. Unlike the conventional approaches, we use stochastic geometry to obtain analytical expression for interference variance (or power) across OFDM frequency tones and use it to derive the MSE expressions for different algorithms under both noise and pilot contaminated regimes. Simulation results validate our analysis and the near optimal MSE performance of proposed estimation algorithms

Each user communicates with the BS using OFDM and transmits uplink pilots for channel estimation. We assume that all users in a particular cell are assigned orthogonal frequency tones so that there is no intra-cell interference. However, due to necessary reuse of pilots, there are users in the neighbouring cells that transmit pilots at the same frequency tones, resulting in an inter-cell interference or pilot contamination. Since only the user in a particular cell of interest will experience In this section, we present three different techniques for channel estimation in massive MIMO-OFDM based on the well-known LMMSE and LS estimators and discuss their limitations. For now, we assume that estimates are corrupted only by the white noise. Hence, without loss of generality, we consider a single-cell single-user scenario for the approach

$$\mathrm{MSE}^{(\mathrm{O})} = \sum_{j=1}^{R} \sum_{i=1}^{L} \frac{\eta_j \delta_i}{1 + \rho K \eta_j \delta_i},$$

Earlier, we have shown that the MMSE algorithm can obtain optimal performance by using prior information and better suppression to PC. Although the use of SVD of channel correlation matrix is able to reduce the number of multiplications with negligible performance loss, its complexity is still quite high since obtaining the SVD itself has high computational complexity on the order of O(N 3). Here, we introduce the H-inf algorithm, which were proposed Multicell MU-MIMO systems

H-INF CHANNEL ESTIMATION

As an alternative to the classical MMSE estimation, an H-inf filter can achieve an acceptable estimation performance without accurate knowledge of the statistical information of the involved signals. The idea of the H-inf filtering is to construct a filter that guarantees the Hinf norm of the estimation error is less than a prescribed positive value As for multicell MUMIMO systems, the idea of the H-inf is to find an estimation method so that the ratio between the whole channel estimation error (between the jth BS and K users in each cell) and the input noise/interference is less than a prescribed threshold. Given a positive scalar factor s, the H-inf estimator for each received OFDM symbol needs to satisfy the following objective function

$$\sup_{Z_j} = \frac{\|\hat{C}_j - C_j\|_W^2}{\|Z_j\|^2} < s \qquad (12)$$

where _ˆCj − Cj_2 W = (ˆC j − Cj)HW(ˆC j − Cj); ˆC j is a LQK × 1 vector, denoting the channel response vector to be estimated; Cj = [CTj1, . . . ,CTjQ]T ; Cjq = [CTjq1, . . . ,CTjqK]T ; andW > 0 is a weighting matrix.

4.3 PERFORMANCE ANALYSIS

Analysis of Matrix γ: To find a solution for the H-inf, we assume R − s−1W > 0 [22], [23], where R is an identity matrix because QPSK is adopted,3 s is a positive scalar factor, and W is also a diagonal matrix that have equal dimensions. Thus,M1,1,M1,2,M2,1, andM2,2 are all diagonal matrices, respectively. Finally, matrix γ is a real diagonal matrix with equal diagonal elements.

Since the diagonal matrix γ is needed to estimate the performance of the H-inf, we will find the relation between γ and the identity matrix. First, it is assumed that

$$\gamma < I_{LQK}. \tag{22}$$

Note that *R* will not be an identity matrix if 16-QAM, 64-QAM, or other modulations are adopted. However, γ is always a diagonal matrix. The proposed algorithm is valid for the different modulations. To satisfy (22), one has $\varepsilon - \eta > 0$. By applying (14), we can get

$$\begin{aligned}\varepsilon - \eta &= (M_{1,1} + M_{1,2}\xi_j) - (M_{2,1} + M_{2,2}\xi_j)\\ &= R^{-\frac{1}{2}} + s^{\frac{1}{2}}W^{-\frac{1}{2}}\xi_j > 0.\end{aligned} \tag{23}$$

Therefore, our hypothesis is valid. Intuitively, when W is fixed, a smaller s is made, a smaller γ is obtained, and a better performance is achieved, which is the intrinsic characteristic of the H-inf algorithm, as will be discussed in the following

Impact of PC on H-inf: Since the estimation errors in cells are independent of each other, we analyze the channels from the K users in the jth cells. The following assumptions are made: 1) All subcarriers have equal power; 2) phase-shift orthogonal pilot sequences are used for different users within each cell; and 3) the same pilot sequences are reused in other cells.

The channel estimation of the H-inf can be rewritten as

$$\begin{aligned}\hat{C}_{jj}^{\text{H-inf}} &= \gamma T_j^{\dagger} Y_j\\ &= \gamma T_j^{\dagger}\sum_{q\neq j}^{Q} T_q C_{jq} + \gamma C_{jj} + \gamma T_j^{\dagger} Z_j.\end{aligned} \tag{24}$$

The MSE expression of the H-inf algorithm for multicell MU-MIMO systems in the presence of PC is given asfollows:

$$\text{MSE}_{H-\text{inf}} = \underbrace{\frac{1}{L} r_{nn}^2 \sum_{q\neq j}^{Q} d_{jq}}_{\text{PC}} + \underbrace{\frac{1}{L} r_{nn}^2\sigma^2 + \frac{1}{L}(1 - r_{nn})^2}_{\text{noise}}. \tag{25}$$

Complexity Analysis: Considering the number of complex multiplications for each OFDM symbol as a complexity metric, the inversion of an n × n matrix requires n3 operations, the pseudoinverse of an n × r matrix requires 2r2n + r3 operations, and the product of an m × r matrix with an r × n matrix requires mrn operations. Let Kit denote the number of iterations that should not be too large due to the superior convergence property of SAGE [20]. A comparison of complexity between the LS, MMSE, and proposed H-inf algorithms is given in Table I. As expected, the H-inf estimation has less complexity than the MMSE algorithm, and the complexity canbe further reduced by using the SAGE iterative process.

PROPOSED SYSTEM

The need to curtail the carbon footprint and the operation cost of wireless networks requires an overall energy reduction of base stations (BSs) in the region of two to three orders of magnitude. At the same time, a significant increase in spectrum efficiency from currently about 1.5 bit/s/Hz total least 10 bit/s/Hz is required to cope with the exponentially increasing traffic loads . This challenges the design of multiple-input multiple-output (MIMO) systems associated with the BS. A typical long-term evolution (LTE) BS consists of radio-frequency (RF) chains, baseband interfaces, direct current to direct current (DC-DC) converters, cooling fans, etc. Each RF chain contains a power amplifier (PA), and PA contribute around 65% of the entire energy consumption.

As a three dimensional modulation scheme, SM enables a trade-off between the size of the spatial constellation diagram and the size of the signal constellation diagram, while achieving the same spectrum efficiency. Based on this unique characteristic, transmission optimized spatial modulation (TOSM) aims to select the best combination of these two constellation sizes, which minimizes the average bit error probability (ABEP). To avoid the prohibitive complexity caused by exhaustive search, a two-stage optimization strategy is proposed. The first step is to determine the optimal number of transmit antennas, and this is performed at the receiver. In the second step, the required number of antennas are selected at the transmitter. In addition to low computational complexity, TOSM needs very limited feedback because of two aspects: i) since it is based on statistical CSI, the frequency of updating is relatively low; and ii) feedback is required only to inform the transmitter of the number of selected antennas, instead of the index of each selected antenna. Therefore, the feedback overhead is negligibly low and not considered in this paper. In addition, assuming the SOTA 2010 power model, the overall BS energy consumption is studied for TOSM. The DTX technique is combined with TOSM to further improve the energy efficiency. Compared with our preceding studies the contributions in this paper are four-fold: i) a two-stage optimization method is proposed to balance the spatial modulation order with the signal modulation order in SM systems; ii) a complete derivation of a simplified ABEP bound for SM over generalized fading channels is presented; iii) a direct antenna selection method based on circle packing is proposed; and iv) the energy efficiency of TOSM in terms of the BS energy consumption is evaluated for both the continuous mode and the DTX mode.

CHANNEL DISTRIBUTION

The fading coefficient of the link from the t-th transmit antenna to the r-th receive antenna is denoted by $h_{t,r} = \beta_{t,r} \exp(j\phi_{t,r})$, where $\beta_{t,r}$ and $\phi_{t,r}$ are the amplitude and the phase, receiver. Nakagami-m fading is considered in this paper, i.e. $\beta_{t,r} \sim$ Nakagami($m_{t,r}$,$\Omega_{t,r}$), where respectively. The channel fading distribution as well as the CSI is assumed to be known at the $m_{t,r}$ is the shape parameter (when $m_{t,r} = 1$, the channel is Rayleigh fading) and $\Omega_{t,r}$ is the spread controlling parameter. The phase $\phi_{t,r}$ is uniformly distributed between $(-\pi, \pi]$. 2) Channel Correlation: The correlation coefficient between the amplitudes of the two propagation paths from the transmit antennas t_i and t_j to the r-th receive antenna is denoted by $\rho_{t_i,t_j,r}$. The exponential correlation matrix model in [16] is considered, which is based on the fact that the channel correlation decreases with increasing the distance between antennas. As shown in Fig. 1, the transmit antennas are located in a normalized square area, i.e. the distance between t_1 and t_A is unity. The correlation coefficient between t_1 and t_A with respect to the rth receive antenna is denoted by $\rho_s(r)$.

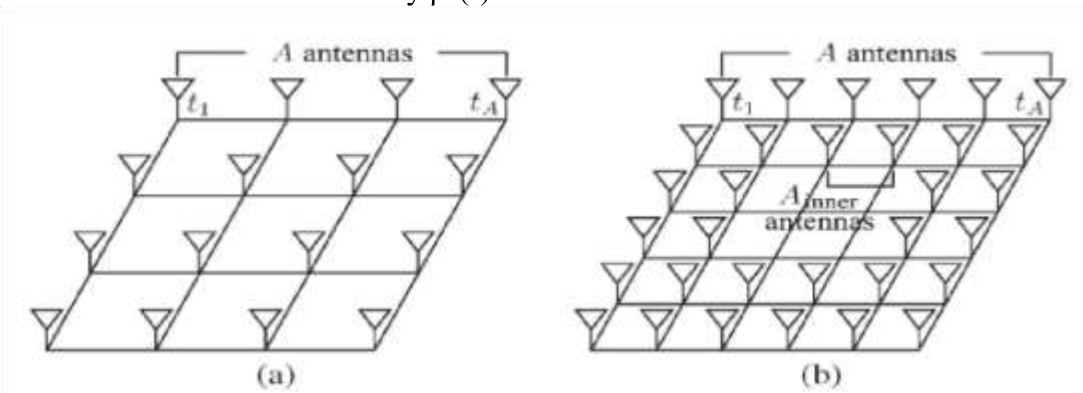


Fig. 5.1. Examples of the transmit antenna array. (a) $N_t = 16$. (b) $N_t = 32$.

The number of antennas on each side of the antenna array is formulated as follows:

$$A = \begin{cases} \sqrt{N_t} & \text{if } \log_2 N_t \text{ is even} \\ 3 \times \sqrt{\frac{N_t}{8}} & \text{if } \log_2 N_t \text{ is odd} \end{cases} \quad (2)$$

When log2 N_t is even, the antennas form a square array with the dimension of $A \times A$. If log2 N_t is odd, the antennas are placed in the shape shown in Fig. 1(b),

Where

The absolute distance between t_i and t_j is denoted by d_{t_i,t_j} , and the correlation between those two antennas is given by [16, Eq. (10)]:

$$\rho_{t_i,t_j,r} = \rho_{s(r)}^{d_{t_i,t_j}}, \quad 0 \leq \rho_{s(r)} \leq 1. \tag{3}$$

The average degree of the channel correlations, denoted by ρav, is calculated by:

$$\rho_{av} = \frac{1}{N_r} \sum_{r=1}^{N_r} \left(\frac{1}{N_t(N_t-1)} \sum_{t_i=1}^{N_t} \sum_{t_j \neq t_i=1}^{N_t} \rho_{t_i,t_j,r} \right). \tag{4}$$

5.2 BASE STATION POWER MODEL

In [17], a linear relationship between the RF output power and the overall consumed power of a multi-sector BS was established. The overall BS power consumption is divided into two parts: the load-dependent portion and the constant portion. The former is dependent on the RF output power, while the latter is invariant. In addition, when no data is transmitted, a sleep mode is enabled to reduce the consumption by switching off unneeded components. In this section, a practical BS power model is introduced for the purpose of evaluating the energy efficiency of the proposed method. 1) Power Model: In [15], based on the above literature, a power model named SOTA 2010 was proposed for a onesector, single-RF-chain BS. Table I specifies the parameters: Pmax is the maximum RF output power; P0 and Ps denote the constant power consumption for the active mode and the sleep mode, respectively; ζ stands for the slope that quantifies the load dependence. The instantaneous BS power consumption, denoted by Pin, is formulated as a function of the RF output power Pout [15, Eq. (1)]:

$$P_{in} = \begin{cases} P_0 + \zeta P_{out} & \text{if } 0 < P_{out} \leq P_{max} \\ P_s & \text{if } P_{out} = 0 \end{cases}. \tag{5}$$

Also, the ratio of the time consumed in the active mode and the total period is referred to as the activation ratio μ. The

TABLE I
BS POWER MODEL PARAMETERS [15]

Power model	P_{max} (W)	P_0 (W)	P_s (W)	ζ
SOTA 2010	40	119	63	2.4

Average power consumption of a single-RF-chain BS is then computed by:

$$P_{BS} = \mu(P_0 + \zeta P_{out}) + (1-\mu)P_s. \tag{6}$$

Unlike SM using a single RF chain, multi-stream MIMO schemes require multiple RF chains to activate the transmit antennas simultaneously. For *N*act *activated* antennas, the RF output power of each antenna is 1/*N*act of *P*out. As a result, the overall power consumption of a BS with multiple RF chains is calculated by:

$$P_{BS} = N_{act}\left[\mu\left(P_0 + \zeta \frac{P_{out}}{N_{act}}\right) + (1-\mu)P_s\right]. \tag{7}$$

5.3 CONTINUOUS MODE AND DTX MODE

Two modes are considered to operate the RF chains: the continuous mode and the DTX mode [7]. In the continuous mode, the RF chains are always delivering output power of the same level. As a result, Pout is equal to the average RF output power ¯ Pout, and μ = 1. Substituting those conditions into (7), the overall BS power consumption in the continuous mode is obtained by:

$$P_{BS\cdot cont.} = N_{act}P_0 + \zeta \bar{P}_{out}. \tag{8}$$

The data rate of the continuous mode is equal to the average data rate, which is denoted by ¯*Rb* = ¯ *P*out/*Eb*. Conversely, the DTX mode conveys data with full load, and the instantaneous data rate is *Rb*max = *P*max/*Eb* that is higher than ¯*R b*. Then the system is enabled into sleep mode during the saved time to maintain the same average data rate. The parameter *μ* of the DTX mode is computed by:

$$\mu = \frac{\bar{R}_b}{R_{b\,max}} = \frac{\bar{P}_{out}}{P_{max}}. \tag{9}$$

Substituting (9) and Pout = Pmax into (7), the overall BS power consumption in the DTX mode is expressed as:

$$P_{\mathrm{BS\cdot DTX}} = N_{\mathrm{act}} P_s + \left(\zeta + \frac{N_{\mathrm{act}}(P_0 - P_s)}{P_{\max}} \right) \bar{P}_{\mathrm{out}}. \quad (10)$$

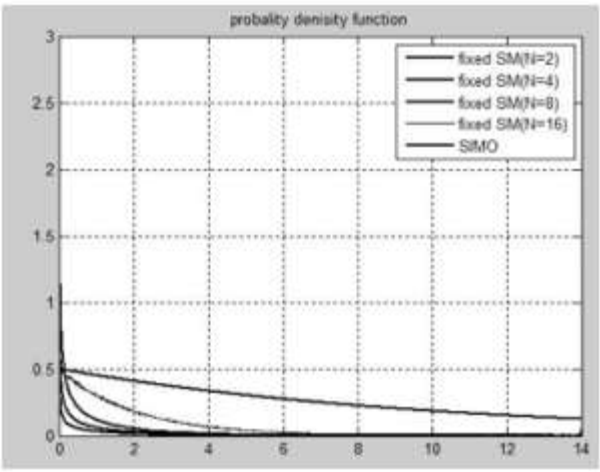
TOSM OVER C.I.D. RAYLEIGH FADING CHANNELS

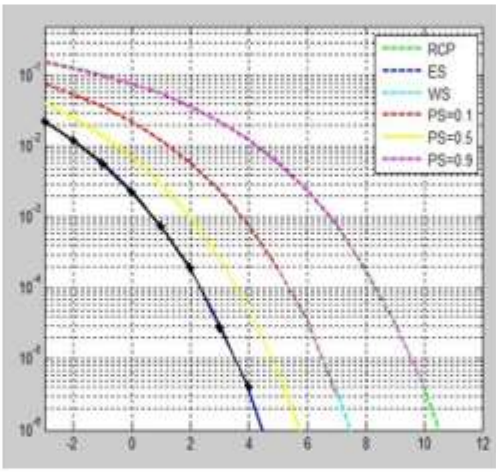
In this section, we study TOSM in the special case of channel fading, Rayleigh fading. We present that TOSM is independent

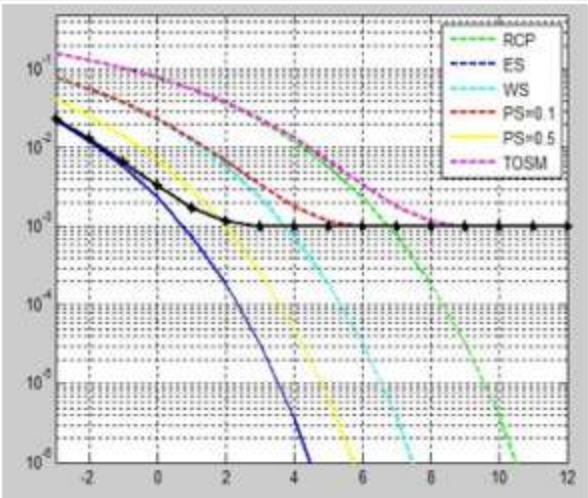
$$M_{\mathrm{opt}} = \arg\min_{M} \quad \frac{\overline{B}_{N_t}}{\gamma^{\overline{m}_r N_r}} M^{2\overline{m}_r N_r} + \frac{\overline{C}_{N_t}}{\gamma^{N_r}} \left(2^{\eta_s}\eta_s - M\log_2(M)\right), \quad \text{subject to}: \quad 1 \le M \le 2^{\eta_s} \quad (32)$$

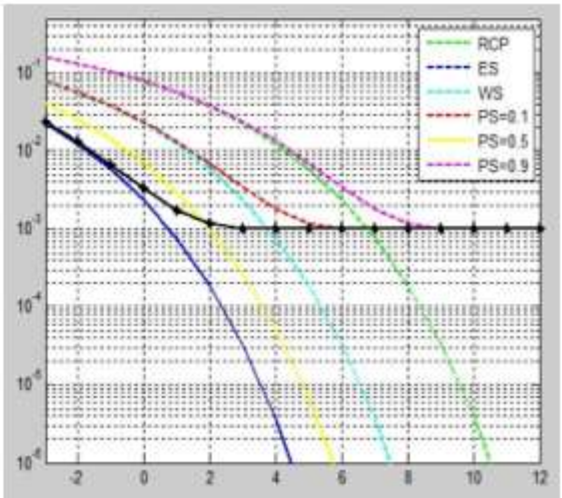
of SNR in this particular scenario, and a look-up table can be built to quickly determine the best choice of (M,N). In addition, given a target bit error rate (BER) and transmission rate, a closedform expression of the BS energy consumption is derived for TOSM. This allows us to evaluate the proposed scheme analytically.

Results









CONCLUSION

In this paper, we proposed an optimum transmit structure for SM, which balances the size of the spatial constellation diagram and the size of the signal constellation diagram. Instead of using exhaustive search, a novel two-stage TAS method has been proposed for reducing the computational complexity, where the optimal number of transmit antennas and the specific antenna positions are determined separately. The first step is to obtain the optimal number of transmit antennas by minimizing a simplified ABEP bound for SM. In the second step, a direct antenna selection method, named RCP, was developed to select the required number of transmit antennas from an antenna array. In addition, a look-up table was built in the case of c.i.d. Rayleigh fading, which can readily be used to determine the optimum transmit structure. Results show that TOSM improves the BER performance of the original SM significantly, and outperforms V-BLAST and STBC greatly in terms of the overall BS energy consumption. A further study shows that TOSM is more energy efficient when combined with the DTX mode than the continuous mode. Furthermore, the issue with respect to the maximum transmission rate in the SM systems has been addressed, which is caused by the limited output power of a single RF chain. It was shown that TOSM uplifts the maximum transmission rate of the original SM greatly, and diminishes the gap between SM and STBC significantly. All these merits make TOSM a highly energy-efficient, low-complexity scheme to satisfy the requirement of high data rate transmission, and an ideal candidate for massive MIMO. Further research will extend the optimum transmit structure to generalized SM, where several antennas are activated simultaneously.

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