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Constructive Interference as an Information Carrier by Dual-Layered MIMO Transmission

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***Abstract*—We propose a bandwidth-efficient transmission scheme for multiple-input–multiple-output point-to-point and downlink channels. The bandwidth efficiency (BE) of spatial multiplexing (SMX) is improved by implicitly encoding information in the spatial domain based on the existence of constructive interference in the received symbols, which creates a differentiation in the symbol power. Explicitly, the combination of symbols received at a higher power level carries implicit information in the spatial domain in the same manner as that the combination of nonzero elements in the received symbol vector carries information for receive-antenna-based spatial modulation (RSM). The nonzero power throughout the received symbol vector for the proposed technique allows a full SMX underlying transmission, with the BE enhancement brought by the spatial symbol. Our simulation results demonstrate both significant BE gains and error probability reduction for our approach over the conventional SMX and RSM schemes.**

***Index Terms*—Multiple-input multiple-output (MIMO), precoding, spatial modulation (RSM), spatial multiplexing (SMX).**

I. Introduction

Multiple-input multiple-output (MIMO) systems have been shown to improve the capacity of the wireless channel by means of spatial multiplexing (SMX). Transmit precoding (TPC) schemes introduced for multiuser downlink (DL) transmission improve both the power efficiency and cost of mobile stations by shifting the signal processing complexity to the base stations. From the wide range of linear and nonlinear TPC schemes found in the literature, here, we focus our attention on the family of closed-form linear TPC schemes based on channel inversion [1], [2], which pose low computational complexity. More recently, spatial modulation (SM) has been explored as a means of implicitly encoding information in the index of the specific transmit antenna (TA) activated for the transmission of the modulated symbols, which offers a low-complexity design alternative [3]. Its central benefits include the absence of interantenna interference and the fact that, in contrast to SMX, it only requires a subset (down to one) of radio-frequency chains compared with SMX. Early work has focused on the design of receiver algorithms for minimizing the bit error ratio (BER) of SM at low complexity [3]–[5].

In addition to receive processing, recent work has also proposed constellation shaping for SM [6]–[14]. Specifically, the contributions on this topic have focused on three main directions: 1) shaping and optimization of the spatial constellation, i.e., the legitimate sets of activated TAs [6]; 2) modulation constellation shaping [7]–[9] for the SM transmission where the constellation of the classically modulated bits is optimized; and 3) joint spatial and modulation constellation shaping, in the form of optimizing the received constellation [10]–[14]. Closely related treatises have been focused on applying SM to the receive antennas (RAs) of the communication link, forming the RA-based spatial modulation (RSM) regime [15], [16]. By means of precoding at the transmitter, this regime aims at transmitting to a reduced a subset of RAs that receive information symbols, whereas the rest of the antennas receive only noise. A dual-layered transmission (DLT) scheme was proposed in [17], where the spatial symbol is conveyed, not by transmitting a combination of symbols and zeros but by assigning a pair of power levels $\{P_1, P_2\}$ to the received symbols, with the combination of power levels detected at the receiver representing a spatial symbol.

Here, we explore a power-efficient alternative, where the distinction of the power levels in DLT is no longer formed by the aforementioned direct power allocation but rather by allowing the constructive interference to form a subset of received symbols. Indeed, it has been shown that by including simple linear TPC techniques, the aforementioned constructive interference can be exploited to boost the received power of the information symbols in the multiple-input–single-output DL [2], [18]. Here, we selectively apply this concept to a subset of received symbols to enhance their power levels and convey the spatial symbol, thus reusing interfering power in a power-efficient manner.

The remainder of this paper is organized as follows. Section II introduces the proposed transmission scheme. Section III focuses on the calculation of the computational complexity of the proposed scheme, whereas in Section IV, we discuss the error probability of our approach. Finally, Section V presents our numerical results, and our conclusions are offered in Section VI.

II. Dual-Layered Transmission by Constructive Interference

A. System Model

Consider a MIMO system where the transmitter and the receiver are equipped with N_t and N_r antennas, respectively. For simplicity, unless stated otherwise, in this paper, we assume that the transmit power budget is limited to $P = 1$. For the case of the closed-form TPCs in [1] and [2], it is required that $N_t \geq N_r$. The given channel is modeled by

$$\mathbf{r} = \mathbf{H}\mathbf{t} + \mathbf{w} \tag{1}$$

where $\mathbf{r}$ is the vector of received symbols in all RAs, and $\mathbf{H}$ is the MIMO channel vector with elements $h_{m,n}$ representing the complex-valued channel coefficient between the nth TA and the mth RA. Furthermore, $\mathbf{t}$ is the vector of precoded transmit symbols that will be discussed in the following, and $\mathbf{w} \sim \mathcal{CN}(0, \sigma^2\mathbf{I})$ is the additive white Gaussian noise at the receiver, with $\mathcal{CN}(\mu, \sigma^2)$ denoting the circularly symmetric complex Gaussian distribution associated with a mean of μ and a variance of σ^2.

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B. Proposed DLT-CI

The conventional DLT in [17] combines SMX with RSM where the bandwidth efficiency (BE) of conventional SMX MIMO transmission is strictly enhanced. This is achieved by encoding the spatial bits in the RSM fashion in the received power domain, by selecting two distinct nonzero power levels for the transmit supersymbols instead of the conventional 'on–off' RSM transmission. This allows for having nonzero elements throughout the received symbol vector and, therefore, a full SMX transmission in the modulated signal domain. Here, we explore the technique of forming the difference between the received power levels for DLT by actively harvesting the constructive interference at the receiver. This allows for 1) an improved BE of

$$\epsilon = N_r \log_2(M) + \log_2 \binom{N_r}{N_a} \tag{2}$$

for DLT with an M-order modulation by transmission of the spatial symbol, where N_a denotes the number of higher-power received symbols; for 2) enhanced power efficiency where the spatial symbol is formed by the reuse of interference power instead of power allocation; and for 3) an improved average error performance due to the increased power levels of a subset of symbols by means of constructive interference.

1) Transmitter: In [2], Masouros proposed a linear TPC that carefully aligns interference so that it constructively contributes to the desired signal power. In brief, the precoding matrix in [2] is formed as

$$\mathbf{T}_c = \mathbf{T}\mathbf{R}_\phi \tag{3}$$

where $\mathbf{T} = \mathbf{H}^H(\mathbf{H}\mathbf{H}^H)^{-1}$, and $\mathbf{R}_\phi = \mathbf{R} \odot \mathbf{\Phi}$, with $\odot$ denoting element-wise matrix multiplication and $\mathbf{R}_\phi$ representing the correlation rotation (CR) matrix that contains the elements of the channel correlation matrix $\mathbf{R} = \mathbf{H}\mathbf{H}^H$ rotated by the angle-only matrix $\mathbf{\Phi}$ such that the resulting interference constructively aligns to the received signal. To avoid repetition, see [2] for the details of the formation of $\mathbf{R}_\phi$, whereas here, we modify the above operation for our proposed technique as detailed in the following. As an enhancement of the conventional DLT in [17], we employ this concept here by first forming the modulated symbol vector $\mathbf{b}_m = [b_{m_1}, b_{m_2}, \ldots, b_{m_{N_r}}]^T$ where, as opposed to the DLT in [17], all symbols have the same power. Here, $b_{m_i}, m_i \in \{1, \ldots, M\}$ is a symbol taken from an M-order modulation alphabet that represents the transmitted waveform in the baseband domain conveying $\log_2(M)$ bits.

We next form the power imbalance at the receiver by allowing constructive interference for the N_a-out-of-N_r RAs by appropriately adapting the TPC in [2]. Explicitly, we modify the precoding matrix of (3) to selectively allow constructive interference imposed only on the N_a "activated" antennas as a means of creating the required data-dependent power difference. First, to ensure uniform power for the desired symbol (excluding interference) across all RAs, we employ a normalized version of the channel correlation matrix formulated as $\mathbf{Q} = \mathbf{R}\mathrm{diag}(\mathbf{R})^{-1}$ with ones along its diagonal. We use the operator $\mathrm{diag}(\mathbf{R})$ to denote the matrix that has the diagonal elements of $\mathbf{R}$ on its diagonal and zeros elsewhere. The normalized CR matrix is then formed as $\mathbf{Q}_\phi = \mathbf{Q} \odot \mathbf{\Phi}$. We then apply the precoding matrix

$$\mathbf{T}^k = \mathbf{T}\mathbf{Q}_\phi^k \tag{4}$$

where $\mathbf{Q}_\phi^k = \{\mathbf{Q}_\phi\}_k$ is the selective CR matrix where the rows in set k are taken from $\mathbf{Q}_\phi$, whereas the remaining rows are taken from the identity matrix with size N_r. Finally, the transmit vector is formed as

$$\mathbf{t} = \beta\mathbf{T}^k\mathbf{b}_m \tag{5}$$

where $\beta = \sqrt{1/\operatorname{tr}(\mathbf{T}^k\mathbf{T}^{kH})}$ is the average power normalization factor. In the given equation, k represents the index of the N_a activated RAs (the index of the high-power elements in the received vector) conveying $\log_2\binom{N_r}{N_a}$ bits in the spatial domain. Matrix $\mathbf{T}^k$ can be thought of as the combined precoding and spatial symbol matrix, which only allows constructive interference to be imposed on the N_a RAs as indicated by the spatial symbol k. From (1)–(5), the received signal is given as

$$\mathbf{r} = \beta\mathbf{Q}_\phi^k\mathbf{b}_m + \mathbf{w} \tag{6}$$

where the dual-layered received supersymbol has been formed as $\mathbf{s}_m^k = \beta\mathbf{Q}_\phi^k\mathbf{b}_m$. It can be seen that for the "inactive" RAs, we have

$$r_i = \beta b_{m_i} + w_i, i \in \mathcal{L} \tag{7}$$

where $\mathcal{L}$ is the set of "inactive" antennas. Clearly, for a normalized modulation constellation, these symbols are received at power levels of $P_L = \beta^2$. For the rest of the symbols, we have

$$\begin{aligned} r_i &= \beta\mathbf{q}_\phi^i\mathbf{b}_m + w_i \\ &= \beta b_{m_i} + \sum_{j\neq i}^{N_r} \mathbf{q}_\phi^{i,j}\mathbf{b}_{m_j} + w_i, i \in \mathcal{L}_c \end{aligned} \tag{8}$$

where $\mathbf{q}_\phi^i = [\mathbf{q}_\phi^{i,1}, \mathbf{q}_\phi^{i,2}, \ldots, \mathbf{q}_\phi^{i,N_r}]$ is the ith row of $\mathbf{Q}_\phi^k$, and $\mathcal{L}_c$ is the complementary set of $\mathcal{L}$, i.e., the set of N_a "active" antennas. The symbols in (8) are received at higher power levels due to constructive interference [2]. Since for CR precoding, all interfering symbols are constructively aligned to the symbol of interest, for the case of constant envelope modulation, it can be seen that the received power levels obey

$$P_i = \beta^2\left(1 + \sum_{j\neq i}^{N_r} \left\|\mathbf{q}_\phi^{i,j}\right\|^2\right) > \beta^2 = P_L. \tag{9}$$

Clearly, this constructive interference is what creates the power level separation between the RAs to form the spatial symbol k.

Remark: Note that a number of alternative precoders such as [18]–[24] can be used in conjunction with the proposed approach to accommodate constructive interference for the formation of the power level separation required for DLT. To constrain the computational complexity, here, we employ the low-complexity approach in [2], as previously detailed.

2) Receiver: At the receiver side, explicit knowledge of the power levels is not required, as long as the detector can distinguish between the power levels. Hence, the receive processing is identical to that for conventional DLT where, first, the N_a "active" antenna indexes are detected based on the N_a highest received power levels among the RAs—formed by constructive interference—according to

$$\hat{k} = \arg\max_{j\in\mathcal{J}} \sum_{i=1}^{N_a} |r_{i,j}|^2 \tag{10}$$

where $\mathcal{J}$ denotes the set of symbols in the spatial domain, and the modulated symbols at all RAs are detected as

$$\hat{\mathbf{b}}_m = \arg\min_{n\in\mathcal{Q}} |\mathbf{r}/\beta - \mathbf{b}_n|^2 \tag{11}$$

where $\mathcal{Q}$ denotes the modulation constellation, and b_n are the symbols in the modulated symbol alphabet.

TABLE I
COMPLEXITY FOR SMX, DLT, AND THE PROPOSED DLT-CI SCHEME

	Operations
Transmitter:	
ZF processing	$N_r^3 + N_t N_r + N_t N_r F$
Selective CR	$N_r^3 + N_t N_r + (N_t N_r + N_t N_r^2) F$
Receiver:	
Spatial detection	$2N_a \binom{N_r}{N_a} F$
Demodulation	$N_r M F$
SMX Total	$N_r^3 + N_t N_r + N_r(N_t + M)F$
DLT Total	$N_r^3 + N_t N_r + [N_r(N_t + M) + 2N_a \binom{N_r}{N_a}]F$
DLT-CI Total	$N_r^3 + N_t N_r +$ $[N_r(N_t + N_t N_r + M) + 2N_a \binom{N_r}{N_a}]F$

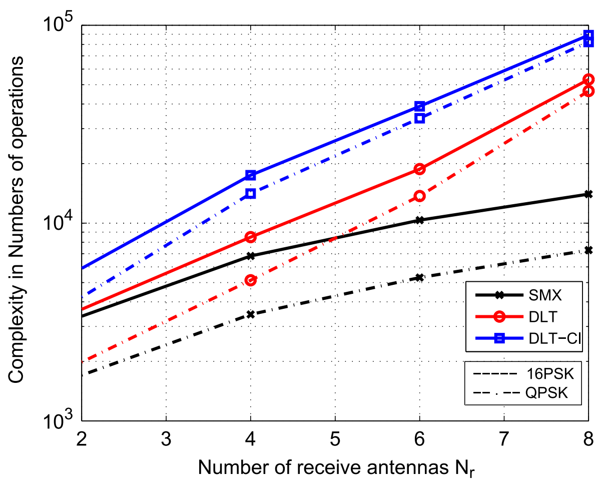

Fig. 1. Complexity versus N_r for $N_t = 8$ and $N_a = N_r/2$ with SMX, DLT, and DLT-CI.

III. COMPUTATIONAL COMPLEXITY

Here, we compare the computational complexity of SMX, DLT, and DLT-CI. First, Table I summarizes the computational complexity of each of the techniques, taking into account the dominant operations at the transmitter and the receiver. We assume a quasi-static channel, which is constant for a frame length of F supersymbols. For SMX and DLT, the zero-forcing precoding at the transmitter involves the inversion of the channel matrix that involves a number of $N_r^3 + N_t N_r$ operations and the multiplication with the supersymbol vector involving an additional $N_t N_r$ operations for the F supersymbols of the transmission frame. The selective CR of DLT-CI involves the additional multiplication of the precoding matrix with $\mathbf{Q}_\phi^k$ at every symbol period, with complexity of $N_t N_r^2$. At the receiver, all techniques require a demodulation stage that involves M comparisons for M-order modulation for each of the N_r RAs. The DLT and DLT-CI require an additional stage for the detection of the spatial symbol, which, from (10), involves N_a complex multiplications and N_a complex additions for each antenna combination out of the $\binom{N_r}{N_a}$ combinations in total.

Fig. 1 shows the complexity of SMX, DLT, and the proposed DLT-CI for a system with $N_t = 8$ TAs and increasing numbers of RAs N_r, with $N_a = N_r/2$. For reference, we have assumed a Long-Term Evolution (LTE) Type-2 time-division duplexing (TDD) frame structure for which $F = 70$, as detailed in [17]. A slow-fading channel is assumed where the channel remains constant for the duration of the frame. It can be seen that the proposed DLT-CI has increased complexity compared with DLT. However, it will be shown in the following results that the improved performance for DLT-CI is worth the added complexity.

IV. ERROR PROBABILITY

The error probability of the proposed scheme can be described by means of the pairwise error probability (PEP) $\mathcal{P}(\mathbf{s}_m^k \to \mathbf{s}_n^l)$. By the use of the union bound, the average bit error probability P_e can be expressed as [13]

$$P_e \leq \frac{1}{b} E \left\{ \sum_{\mathbf{s}_m^k \in \mathcal{B}} \sum_{\mathbf{s}_n^l \in \mathcal{B} \neq \mathbf{s}_m^k} d\left(\mathbf{s}_m^k, \mathbf{s}_n^l\right) \mathcal{P}\left(\mathbf{s}_m^k \to \mathbf{s}_n^l\right) \right\} \quad (12)$$

where $d(\mathbf{s}_m^k, \mathbf{s}_n^l)$ is the Hamming distance between the bit representations of the symbols $\mathbf{s}_m^k, \mathbf{s}_n^l$, and $\mathcal{B}$ is the supersymbol constellation defined as the union of the spatial-domain constellation and of the classic modulation constellation. The PEP can further be decomposed into the PEP for the spatial symbol $\mathcal{P}(s_{m_i}^k \to s_{m_i}^l)$ and the PEP $\mathcal{P}(s_{m_i}^k \to s_{n_i}^k)$ of the modulated symbol. These obey the following lemmas.

Lemma 1: The PEP of the spatial symbol for the DLT-CI transmission obeys

$$\mathcal{P}\left(s_{m_i}^k \to s_{m_i}^l\right) = Q\left(\frac{\beta}{\sqrt{2\sigma^2}} \cdot \frac{\sqrt{P_i} - \sqrt{P_L}}{2}\right) \quad (13)$$

where $Q(.)$ denotes the Gaussian Q-function.

Lemma 2: The PEP for the M-order phase-shift keying (M-PSK) modulated symbol, which is the focus of this work, follows:

$$\mathcal{P}\left(s_{m_i}^k \to s_{n_i}^k\right) = Q\left(\beta\sqrt{\frac{P_i}{2\sigma^2}\log_2(M)\sin\frac{\pi}{M}}\right). \quad (14)$$

Both the above expressions can be straightforwardly derived by adapting the methodology introduced in [17] for the proposed scenario. It is the PEP in (14) that is enhanced for the proposed scheme by allowing constructive interference to increase P_i. The tightness of the above-described bound is validated in Section V.

V. NUMERICAL RESULTS

To evaluate the benefits of the proposed technique, this section presents Monte Carlo simulations of the proposed DLT-CI in comparison to conventional approaches. As the superiority of conventional DLT over the most relevant SM and SMX approaches was thoroughly validated in [17] and to limit the congestion in the following graphs, here, we only use conventional DLT and SMX as a reference for comparison. The channel impulse response is assumed to be perfectly known at the transmitter for all techniques. Without loss of generality, unless stated otherwise, we assume that the transmit power is restricted to $P = 1$. MIMO systems with up to eight TAs employing quaternary phase-shift keying (QPSK), 8-PSK, and 16-PSK modulation are explored, albeit it is plausible that the benefits of the proposed technique extend to larger-scale systems and higher-order modulation. For DLT and DLT-CI, we focus on the case $N_a = N_r/2$, which provides the highest BE [17].

In Figs. 2 and 3, we show the BER with increasing signal-to-noise ratio (SNR) for QPSK and 8-PSK, respectively. To complete our comparisons, for both scenarios in the figure, we also show the cases where the symbol modulation order used for SMX is increased for some of the spatial streams to achieve the same BE values of $\epsilon = 10$ and $\epsilon = 14$ with the proposed DLT, for QPSK and 8-PSK, respectively. The figures also show the theoretical bound of (13) on

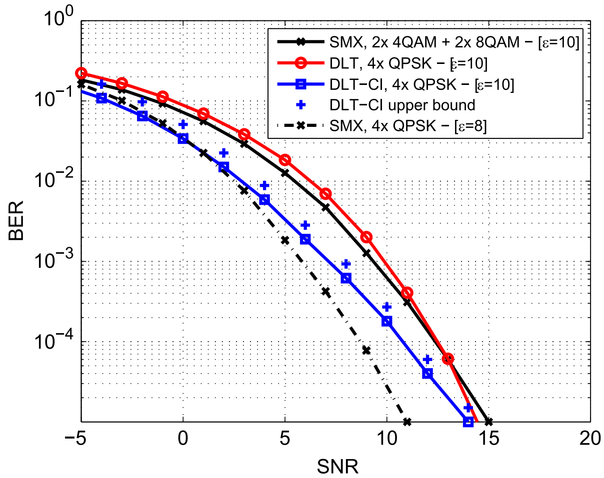

Fig. 2. BER versus SNR for a (8 × 4) MIMO with SMX, DLT, and DLT-CI; QPSK modulation.

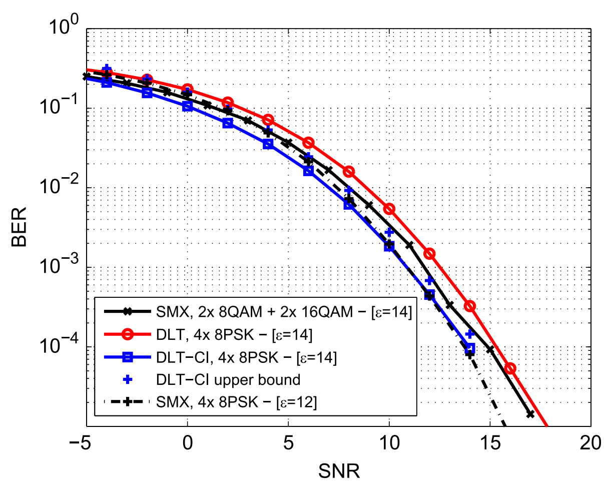

Fig. 3. BER versus SNR for a (8 × 4) MIMO with SMX, DLT, and DLT-CI; 8-PSK modulation.

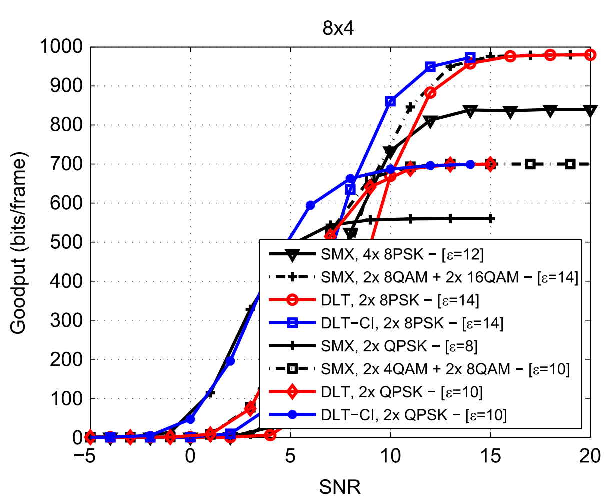

Fig. 4. Throughput versus SNR for a (8 × 4) MIMO with SMX, DLT, and DLT-CI.

the error probability, which closely matches our simulation results in both cases. Clearly, the DLT scheme has an inferior BER performance compared with SMX due to the additional spatial streams, which is the

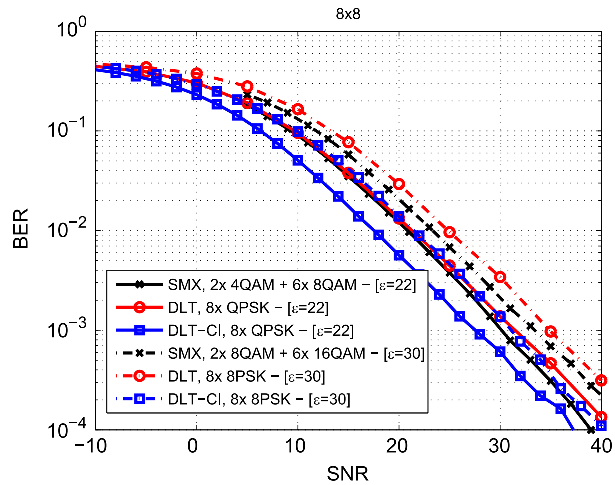

Fig. 5. BER versus SNR for a (8 × 8) MIMO with SMX, DLT, and DLT-CI.

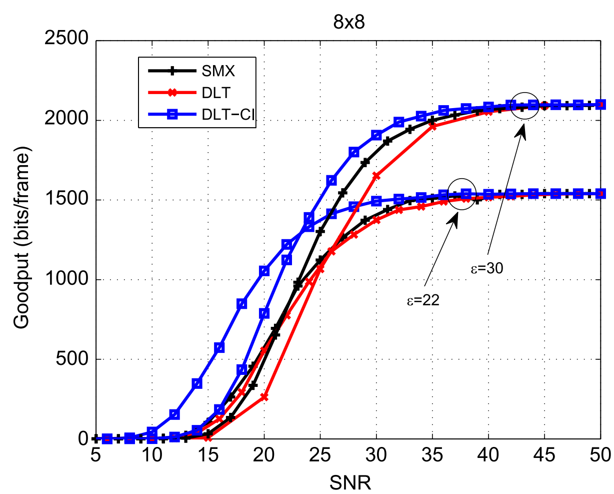

Fig. 6. Throughput versus SNR for a (8 × 8) MIMO with SMX, DLT, and DLT-CI.

price paid for its improved BE. DLT-CI outperforms both SMX and DLT as an explicit benefit of the constructive interference exploited as useful signal power, both in the modulated symbol detection and in the formation of the different power levels employed for the spatial symbol transmission. The improved BE of DLT-CI is demonstrated in Fig. 4,where goodput versus SNR is depicted for the same (8 × 4) AQ2 MIMO scenario. The goodput here is defined as $R = \epsilon F(1 - P_e)^F$, where P_e is the bit error probability [17]. For reference, we have assumed an LTE Type-2 TDD frame structure for which we have $F = 70$, as detailed in [17]. Clearly, DLT-CI provides the best goodput performance among the schemes explored.

Our performance comparison is extended to the (8 × 8) MIMO system in Figs. 5 and 6. The BER performance with increasing SNR is shown in Fig. 5 for the (8 × 8) MIMO system where it can be seen that DLT-CI outperforms both SMX and DLT. Fig. 6 shows the goodput with increasing SNR, where, again, it can be observed that DLT-CI provides the best goodput.

V. Conclusion

An enhanced dual-layered DL transmission scheme has been proposed, which combines traditional MIMO SMX with RSM. The proposed scheme improves upon conventional DLT by allowing constructive interference to carry spatial information, as opposed to the fixed power-level split of the conventional DLT in [17]. Our results show that by allowing constructive interference to separate the power levels and convey the spatial symbol, the proposed DLT-CI improves the BE of SMX while, at the same time, the increased power levels of the subset of symbols improve the average error performance of the system.

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AUTHOR QUERIES

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AQ1 = Please provide the complete current affiliation of author C. Masouros and check if the provided affiliation of author L. Hanzo is correct.
AQ2 = Uncited Figure 4. Figure 6 was changed to Figure 4. Please check if correct.
AQ3 = Please provide publication update in Ref. [13].
AQ4 = Please provide publication update in Ref. [17].

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***Abstract*—We propose a bandwidth-efficient transmission scheme for multiple-input–multiple-output point-to-point and downlink channels. The bandwidth efficiency (BE) of spatial multiplexing (SMX) is improved by implicitly encoding information in the spatial domain based on the existence of constructive interference in the received symbols, which creates a differentiation in the symbol power. Explicitly, the combination of symbols received at a higher power level carries implicit information in the spatial domain in the same manner as that the combination of nonzero elements in the received symbol vector carries information for receive-antenna-based spatial modulation (RSM). The nonzero power throughout the received symbol vector for the proposed technique allows a full SMX underlying transmission, with the BE enhancement brought by the spatial symbol. Our simulation results demonstrate both significant BE gains and error probability reduction for our approach over the conventional SMX and RSM schemes.**

***Index Terms*—Multiple-input multiple-output (MIMO), precoding, spatial modulation (RSM), spatial multiplexing (SMX).**

I. Introduction

Multiple-input multiple-output (MIMO) systems have been shown to improve the capacity of the wireless channel by means of spatial multiplexing (SMX). Transmit precoding (TPC) schemes introduced for multiuser downlink (DL) transmission improve both the power efficiency and cost of mobile stations by shifting the signal processing complexity to the base stations. From the wide range of linear and nonlinear TPC schemes found in the literature, here, we focus our attention on the family of closed-form linear TPC schemes based on channel inversion [1], [2], which pose low computational complexity. More recently, spatial modulation (SM) has been explored as a means of implicitly encoding information in the index of the specific transmit antenna (TA) activated for the transmission of the modulated symbols, which offers a low-complexity design alternative [3]. Its central benefits include the absence of interantenna interference and the fact that, in contrast to SMX, it only requires a subset (down to one) of radio-frequency chains compared with SMX. Early work has focused on the design of receiver algorithms for minimizing the bit error ratio (BER) of SM at low complexity [3]–[5].

In addition to receive processing, recent work has also proposed constellation shaping for SM [6]–[14]. Specifically, the contributions on this topic have focused on three main directions: 1) shaping and optimization of the spatial constellation, i.e., the legitimate sets of activated TAs [6]; 2) modulation constellation shaping [7]–[9] for the SM transmission where the constellation of the classically modulated bits is optimized; and 3) joint spatial and modulation constellation shaping, in the form of optimizing the received constellation [10]–[14]. Closely related treatises have been focused on applying SM to the receive antennas (RAs) of the communication link, forming the RA-based spatial modulation (RSM) regime [15], [16]. By means of precoding at the transmitter, this regime aims at transmitting to a reduced a subset of RAs that receive information symbols, whereas the rest of the antennas receive only noise. A dual-layered transmission (DLT) scheme was proposed in [17], where the spatial symbol is conveyed, not by transmitting a combination of symbols and zeros but by assigning a pair of power levels $\{P_1, P_2\}$ to the received symbols, with the combination of power levels detected at the receiver representing a spatial symbol.

Here, we explore a power-efficient alternative, where the distinction of the power levels in DLT is no longer formed by the aforementioned direct power allocation but rather by allowing the constructive interference to form a subset of received symbols. Indeed, it has been shown that by including simple linear TPC techniques, the aforementioned constructive interference can be exploited to boost the received power of the information symbols in the multiple-input–single-output DL [2], [18]. Here, we selectively apply this concept to a subset of received symbols to enhance their power levels and convey the spatial symbol, thus reusing interfering power in a power-efficient manner.

The remainder of this paper is organized as follows. Section II introduces the proposed transmission scheme. Section III focuses on the calculation of the computational complexity of the proposed scheme, whereas in Section IV, we discuss the error probability of our approach. Finally, Section V presents our numerical results, and our conclusions are offered in Section VI.

II. Dual-Layered Transmission by Constructive Interference

A. System Model

Consider a MIMO system where the transmitter and the receiver are equipped with N_t and N_r antennas, respectively. For simplicity, unless stated otherwise, in this paper, we assume that the transmit power budget is limited to $P = 1$. For the case of the closed-form TPCs in [1] and [2], it is required that $N_t \geq N_r$. The given channel is modeled by

$$\mathbf{r} = \mathbf{H}\mathbf{t} + \mathbf{w} \tag{1}$$

where $\mathbf{r}$ is the vector of received symbols in all RAs, and $\mathbf{H}$ is the MIMO channel vector with elements $h_{m,n}$ representing the complex-valued channel coefficient between the nth TA and the mth RA. Furthermore, $\mathbf{t}$ is the vector of precoded transmit symbols that will be discussed in the following, and $\mathbf{w} \sim \mathcal{CN}(0, \sigma^2\mathbf{I})$ is the additive white Gaussian noise at the receiver, with $\mathcal{CN}(\mu, \sigma^2)$ denoting the circularly symmetric complex Gaussian distribution associated with a mean of μ and a variance of σ^2.

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B. Proposed DLT-CI

The conventional DLT in [17] combines SMX with RSM where the bandwidth efficiency (BE) of conventional SMX MIMO transmission is strictly enhanced. This is achieved by encoding the spatial bits in the RSM fashion in the received power domain, by selecting two distinct nonzero power levels for the transmit supersymbols instead of the conventional 'on–off' RSM transmission. This allows for having nonzero elements throughout the received symbol vector and, therefore, a full SMX transmission in the modulated signal domain. Here, we explore the technique of forming the difference between the received power levels for DLT by actively harvesting the constructive interference at the receiver. This allows for 1) an improved BE of

$$\epsilon = N_r \log_2(M) + \log_2 \binom{N_r}{N_a} \tag{2}$$

for DLT with an M-order modulation by transmission of the spatial symbol, where N_a denotes the number of higher-power received symbols; for 2) enhanced power efficiency where the spatial symbol is formed by the reuse of interference power instead of power allocation; and for 3) an improved average error performance due to the increased power levels of a subset of symbols by means of constructive interference.

1) Transmitter: In [2], Masouros proposed a linear TPC that carefully aligns interference so that it constructively contributes to the desired signal power. In brief, the precoding matrix in [2] is formed as

$$\mathbf{T}_c = \mathbf{T}\mathbf{R}_\phi \tag{3}$$

where $\mathbf{T} = \mathbf{H}^H(\mathbf{H}\mathbf{H}^H)^{-1}$, and $\mathbf{R}_\phi = \mathbf{R} \odot \mathbf{\Phi}$, with $\odot$ denoting element-wise matrix multiplication and $\mathbf{R}_\phi$ representing the correlation rotation (CR) matrix that contains the elements of the channel correlation matrix $\mathbf{R} = \mathbf{H}\mathbf{H}^H$ rotated by the angle-only matrix $\mathbf{\Phi}$ such that the resulting interference constructively aligns to the received signal. To avoid repetition, see [2] for the details of the formation of $\mathbf{R}_\phi$, whereas here, we modify the above operation for our proposed technique as detailed in the following. As an enhancement of the conventional DLT in [17], we employ this concept here by first forming the modulated symbol vector $\mathbf{b}_m = [b_{m_1}, b_{m_2}, \ldots, b_{m_{N_r}}]^T$ where, as opposed to the DLT in [17], all symbols have the same power. Here, $b_{m_i}, m_i \in \{1, \ldots, M\}$ is a symbol taken from an M-order modulation alphabet that represents the transmitted waveform in the baseband domain conveying $\log_2(M)$ bits.

We next form the power imbalance at the receiver by allowing constructive interference for the N_a-out-of-N_r RAs by appropriately adapting the TPC in [2]. Explicitly, we modify the precoding matrix of (3) to selectively allow constructive interference imposed only on the N_a "activated" antennas as a means of creating the required data-dependent power difference. First, to ensure uniform power for the desired symbol (excluding interference) across all RAs, we employ a normalized version of the channel correlation matrix formulated as $\mathbf{Q} = \mathbf{R}\mathrm{diag}(\mathbf{R})^{-1}$ with ones along its diagonal. We use the operator $\mathrm{diag}(\mathbf{R})$ to denote the matrix that has the diagonal elements of $\mathbf{R}$ on its diagonal and zeros elsewhere. The normalized CR matrix is then formed as $\mathbf{Q}_\phi = \mathbf{Q} \odot \mathbf{\Phi}$. We then apply the precoding matrix

$$\mathbf{T}^k = \mathbf{T}\mathbf{Q}_\phi^k \tag{4}$$

where $\mathbf{Q}_\phi^k = \{\mathbf{Q}_\phi\}_k$ is the selective CR matrix where the rows in set k are taken from $\mathbf{Q}_\phi$, whereas the remaining rows are taken from the identity matrix with size N_r. Finally, the transmit vector is formed as

$$\mathbf{t} = \beta\mathbf{T}^k\mathbf{b}_m \tag{5}$$

where $\beta = \sqrt{1/\operatorname{tr}(\mathbf{T}^k\mathbf{T}^{kH})}$ is the average power normalization factor. In the given equation, k represents the index of the N_a activated RAs (the index of the high-power elements in the received vector) conveying $\log_2\binom{N_r}{N_a}$ bits in the spatial domain. Matrix $\mathbf{T}^k$ can be thought of as the combined precoding and spatial symbol matrix, which only allows constructive interference to be imposed on the N_a RAs as indicated by the spatial symbol k. From (1)–(5), the received signal is given as

$$\mathbf{r} = \beta\mathbf{Q}_\phi^k\mathbf{b}_m + \mathbf{w} \tag{6}$$

where the dual-layered received supersymbol has been formed as $\mathbf{s}_m^k = \beta\mathbf{Q}_\phi^k\mathbf{b}_m$. It can be seen that for the "inactive" RAs, we have

$$r_i = \beta b_{m_i} + w_i, i \in \mathcal{L} \tag{7}$$

where $\mathcal{L}$ is the set of "inactive" antennas. Clearly, for a normalized modulation constellation, these symbols are received at power levels of $P_L = \beta^2$. For the rest of the symbols, we have

$$\begin{aligned} r_i &= \beta\mathbf{q}_\phi^i\mathbf{b}_m + w_i \\ &= \beta b_{m_i} + \sum_{j \neq i}^{N_r} \mathbf{q}_\phi^{i,j}\mathbf{b}_{m_j} + w_i, i \in \mathcal{L}_c \end{aligned} \tag{8}$$

where $\mathbf{q}_\phi^i = [\mathbf{q}_\phi^{i,1}, \mathbf{q}_\phi^{i,2}, \ldots, \mathbf{q}_\phi^{i,N_r}]$ is the ith row of $\mathbf{Q}_\phi^k$, and $\mathcal{L}_c$ is the complementary set of $\mathcal{L}$, i.e., the set of N_a "active" antennas. The symbols in (8) are received at higher power levels due to constructive interference [2]. Since for CR precoding, all interfering symbols are constructively aligned to the symbol of interest, for the case of constant envelope modulation, it can be seen that the received power levels obey

$$P_i = \beta^2 \left(1 + \sum_{j \neq i}^{N_r} \left\|\mathbf{q}_\phi^{i,j}\right\|^2\right) > \beta^2 = P_L. \tag{9}$$

Clearly, this constructive interference is what creates the power level separation between the RAs to form the spatial symbol k.

Remark: Note that a number of alternative precoders such as [18]–[24] can be used in conjunction with the proposed approach to accommodate constructive interference for the formation of the power level separation required for DLT. To constrain the computational complexity, here, we employ the low-complexity approach in [2], as previously detailed.

2) Receiver: At the receiver side, explicit knowledge of the power levels is not required, as long as the detector can distinguish between the power levels. Hence, the receive processing is identical to that for conventional DLT where, first, the N_a "active" antenna indexes are detected based on the N_a highest received power levels among the RAs—formed by constructive interference—according to

$$\hat{k} = \arg\max_{j \in \mathcal{J}} \sum_{i=1}^{N_a} |r_{i,j}|^2 \tag{10}$$

where $\mathcal{J}$ denotes the set of symbols in the spatial domain, and the modulated symbols at all RAs are detected as

$$\hat{\mathbf{b}}_m = \arg\min_{n \in \mathcal{Q}} |\mathbf{r}/\beta - \mathbf{b}_n|^2 \tag{11}$$

where $\mathcal{Q}$ denotes the modulation constellation, and b_n are the symbols in the modulated symbol alphabet.

TABLE I
COMPLEXITY FOR SMX, DLT, AND THE PROPOSED DLT-CI SCHEME

	Operations
Transmitter:	
ZF processing	$N_r^3 + N_tN_r + N_tN_rF$
Selective CR	$N_r^3 + N_tN_r + (N_tN_r + N_tN_r^2)F$
Receiver:	
Spatial detection	$2N_a\binom{N_r}{N_a}F$
Demodulation	N_rMF
SMX Total	$N_r^3 + N_tN_r + N_r(N_t + M)F$
DLT Total	$N_r^3 + N_tN_r + [N_r(N_t + M) + 2N_a\binom{N_r}{N_a}]F$
DLT-CI Total	$N_r^3 + N_tN_r +$ $[N_r(N_t + N_tN_r + M) + 2N_a\binom{N_r}{N_a}]F$

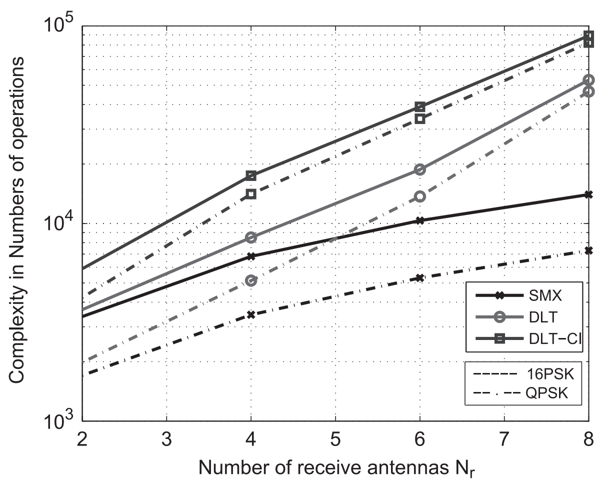

Fig. 1. Complexity versus N_r for $N_t = 8$ and $N_a = N_r/2$ with SMX, DLT, and DLT-CI.

III. COMPUTATIONAL COMPLEXITY

Here, we compare the computational complexity of SMX, DLT, and DLT-CI. First, Table I summarizes the computational complexity of each of the techniques, taking into account the dominant operations at the transmitter and the receiver. We assume a quasi-static channel, which is constant for a frame length of F supersymbols. For SMX and DLT, the zero-forcing precoding at the transmitter involves the inversion of the channel matrix that involves a number of $N_r^3 + N_tN_r$ operations and the multiplication with the supersymbol vector involving an additional N_tN_r operations for the F supersymbols of the transmission frame. The selective CR of DLT-CI involves the additional multiplication of the precoding matrix with $\mathbf{Q}_\phi^k$ at every symbol period, with complexity of $N_tN_r^2$. At the receiver, all techniques require a demodulation stage that involves M comparisons for M-order modulation for each of the N_r RAs. The DLT and DLT-CI require an additional stage for the detection of the spatial symbol, which, from (10), involves N_a complex multiplications and N_a complex additions for each antenna combination out of the $\binom{N_r}{N_a}$ combinations in total.

Fig. 1 shows the complexity of SMX, DLT, and the proposed DLT-CI for a system with $N_t = 8$ TAs and increasing numbers of RAs N_r, with $N_a = N_r/2$. For reference, we have assumed a Long-Term Evolution (LTE) Type-2 time-division duplexing (TDD) frame structure for which $F = 70$, as detailed in [17]. A slow-fading channel is assumed where the channel remains constant for the duration of the frame. It can be seen that the proposed DLT-CI has increased complexity compared with DLT. However, it will be shown in the following results that the improved performance for DLT-CI is worth the added complexity.

IV. ERROR PROBABILITY

The error probability of the proposed scheme can be described by means of the pairwise error probability (PEP) $\mathcal{P}(\mathbf{s}_m^k \to \mathbf{s}_n^l)$. By the use of the union bound, the average bit error probability P_e can be expressed as [13]

$$P_e \leq \frac{1}{b}E\left\{\sum_{\mathbf{s}_m^k\in\mathcal{B}}\sum_{\mathbf{s}_n^l\in\mathcal{B}\neq\mathbf{s}_m^k} d\left(\mathbf{s}_m^k, \mathbf{s}_n^l\right)\mathcal{P}\left(\mathbf{s}_m^k \to \mathbf{s}_n^l\right)\right\} \qquad (12)$$

where $d(\mathbf{s}_m^k, \mathbf{s}_n^l)$ is the Hamming distance between the bit representations of the symbols $\mathbf{s}_m^k, \mathbf{s}_n^l$, and $\mathcal{B}$ is the supersymbol constellation defined as the union of the spatial-domain constellation and of the classic modulation constellation. The PEP can further be decomposed into the PEP for the spatial symbol $\mathcal{P}(s_{m_i}^k \to s_{m_i}^l)$ and the PEP $\mathcal{P}(s_{m_i}^k \to s_{n_i}^k)$ of the modulated symbol. These obey the following lemmas.

Lemma 1: The PEP of the spatial symbol for the DLT-CI transmission obeys

$$\mathcal{P}\left(s_{m_i}^k \to s_{m_i}^l\right) = Q\left(\frac{\beta}{\sqrt{2\sigma^2}}\cdot\frac{\sqrt{P_i}-\sqrt{P_L}}{2}\right) \qquad (13)$$

where $Q(.)$ denotes the Gaussian Q-function.

Lemma 2: The PEP for the M-order phase-shift keying (M-PSK) modulated symbol, which is the focus of this work, follows:

$$\mathcal{P}\left(s_{m_i}^k \to s_{n_i}^k\right) = Q\left(\beta\sqrt{\frac{P_i}{2\sigma^2}\log_2(M)\sin\frac{\pi}{M}}\right). \qquad (14)$$

Both the above expressions can be straightforwardly derived by adapting the methodology introduced in [17] for the proposed scenario. It is the PEP in (14) that is enhanced for the proposed scheme by allowing constructive interference to increase P_i. The tightness of the above-described bound is validated in Section V.

V. NUMERICAL RESULTS

To evaluate the benefits of the proposed technique, this section presents Monte Carlo simulations of the proposed DLT-CI in comparison to conventional approaches. As the superiority of conventional DLT over the most relevant SM and SMX approaches was thoroughly validated in [17] and to limit the congestion in the following graphs, here, we only use conventional DLT and SMX as a reference for comparison. The channel impulse response is assumed to be perfectly known at the transmitter for all techniques. Without loss of generality, unless stated otherwise, we assume that the transmit power is restricted to $P = 1$. MIMO systems with up to eight TAs employing quaternary phase-shift keying (QPSK), 8-PSK, and 16-PSK modulation are explored, albeit it is plausible that the benefits of the proposed technique extend to larger-scale systems and higher-order modulation. For DLT and DLT-CI, we focus on the case $N_a = N_r/2$, which provides the highest BE [17].

In Figs. 2 and 3, we show the BER with increasing signal-to-noise ratio (SNR) for QPSK and 8-PSK, respectively. To complete our comparisons, for both scenarios in the figure, we also show the cases where the symbol modulation order used for SMX is increased for some of the spatial streams to achieve the same BE values of $\epsilon = 10$ and $\epsilon = 14$ with the proposed DLT, for QPSK and 8-PSK, respectively. The figures also show the theoretical bound of (13) on

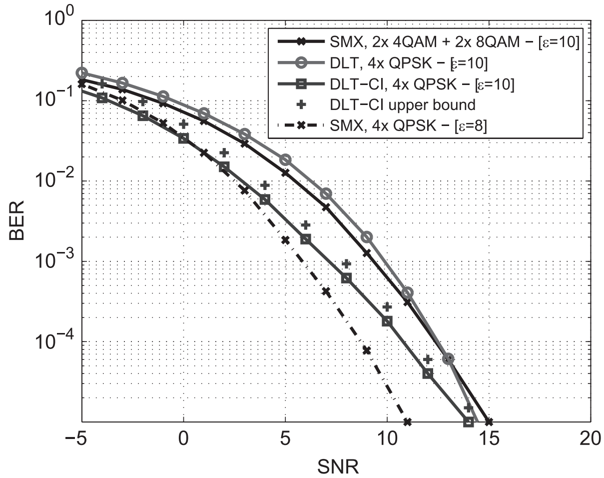

Fig. 2. BER versus SNR for a (8 × 4) MIMO with SMX, DLT, and DLT-CI; QPSK modulation.

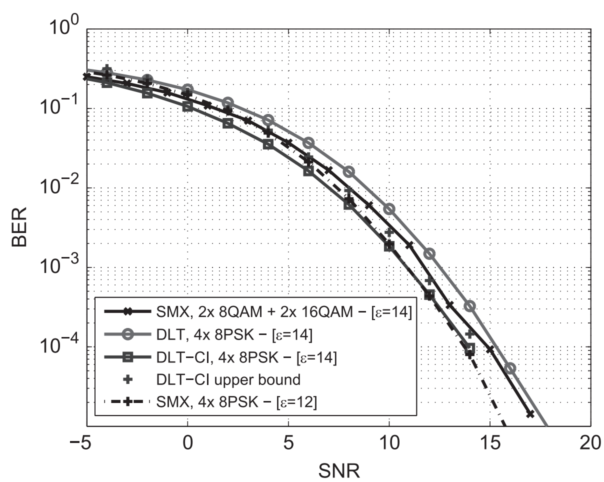

Fig. 3. BER versus SNR for a (8 × 4) MIMO with SMX, DLT, and DLT-CI; 8-PSK modulation.

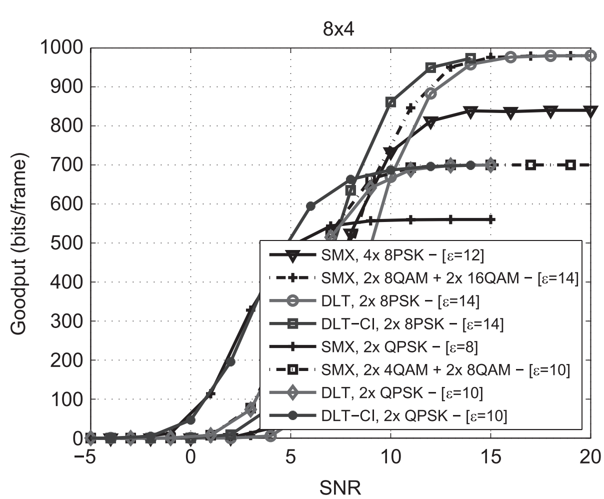

Fig. 4. Throughput versus SNR for a (8 × 4) MIMO with SMX, DLT, and DLT-CI.

the error probability, which closely matches our simulation results in both cases. Clearly, the DLT scheme has an inferior BER performance compared with SMX due to the additional spatial streams, which is the price paid for its improved BE. DLT-CI outperforms both SMX and DLT as an explicit benefit of the constructive interference exploited as useful signal power, both in the modulated symbol detection and in the formation of the different power levels employed for the spatial symbol transmission. The improved BE of DLT-CI is demonstrated in Fig. 4,where goodput versus SNR is depicted for the same (8 × 4) MIMO scenario. The goodput here is defined as $R = \epsilon F(1 - P_e)^F$, where P_e is the bit error probability [17]. For reference, we have assumed an LTE Type-2 TDD frame structure for which we have $F = 70$, as detailed in [17]. Clearly, DLT-CI provides the best goodput performance among the schemes explored. AQ2

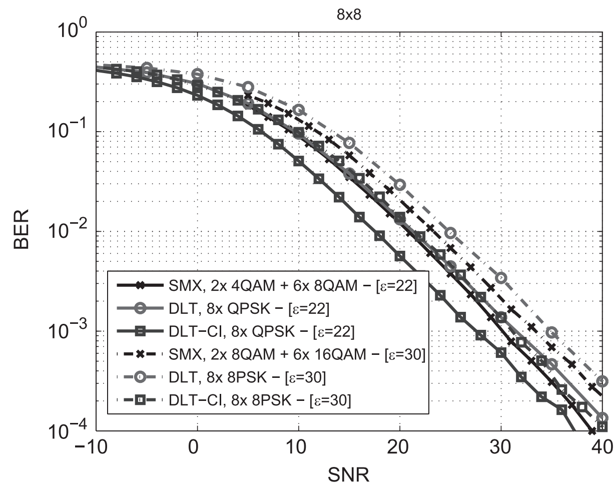

Fig. 5. BER versus SNR for a (8 × 8) MIMO with SMX, DLT, and DLT-CI.

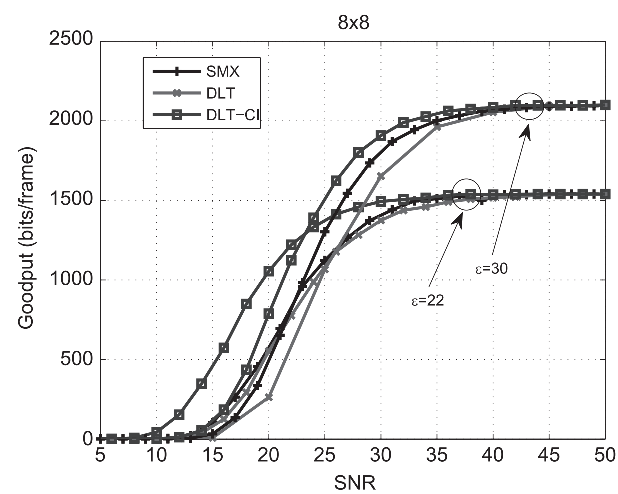

Fig. 6. Throughput versus SNR for a (8 × 8) MIMO with SMX, DLT, and DLT-CI.

Our performance comparison is extended to the (8 × 8) MIMO system in Figs. 5 and 6. The BER performance with increasing SNR is shown in Fig. 5 for the (8 × 8) MIMO system where it can be seen that DLT-CI outperforms both SMX and DLT. Fig. 6 shows the goodput with increasing SNR, where, again, it can be observed that DLT-CI provides the best goodput.

V. Conclusion

An enhanced dual-layered DL transmission scheme has been proposed, which combines traditional MIMO SMX with RSM. The proposed scheme improves upon conventional DLT by allowing constructive interference to carry spatial information, as opposed to the fixed power-level split of the conventional DLT in [17]. Our results show that by allowing constructive interference to separate the power levels and convey the spatial symbol, the proposed DLT-CI improves the BE of SMX while, at the same time, the increased power levels of the subset of symbols improve the average error performance of the system.

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AUTHOR QUERIES

AUTHOR PLEASE ANSWER ALL QUERIES

AQ1 = Please provide the complete current affiliation of author C. Masouros and check if the provided affiliation of author L. Hanzo is correct.
AQ2 = Uncited Figure 4. Figure 6 was changed to Figure 4. Please check if correct.
AQ3 = Please provide publication update in Ref. [13].
AQ4 = Please provide publication update in Ref. [17].

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