

Directional Space-Time Waveform Design for Interference-Avoiding MIMO Configurations

Konstantinos Tountas, George Sklivanitis, Dimitris A. Pados

I-SENSE and Department of Computer and Electrical Engineering and Computer Science

Florida Atlantic University

Boca Raton, FL, USA

{ktountas2017, gsklivanitis, dpados}@fau.edu

Abstract—We consider the problem of directional space-time waveform design for proactive interference avoidance in multiple-input multiple-output (MIMO) link configurations. We propose to instruct the transmitter to communicate with the intended receiver at a given code sequence and angle-of-arrival that are jointly optimized to maximize the maximum attainable pre-detection signal-to-interference-plus-noise ratio (SINR). More specifically, directional transmissions to the intended receiver are achieved by designing space-time waveforms that maximize SINR at the output of the max-SINR receiver filter. Simulation of the proposed adaptive interference-avoiding space-time waveform designs in MIMO link configurations show superior performance in terms of post-filtering SINR and bit-error-rate (BER) when compared to interference-suppressive non-adaptive receiver designs. Joint optimization of the code sequence and array response vector at the transmitter demonstrate that the proposed space-time waveform designs perform within 2 dB from the maximum attainable SINR at the intended receiver.

Index Terms—space-time adaptive communications, waveform design, interference-avoidance, MIMO

I. INTRODUCTION

Future resilient networks will be defined by their ability to maintain wireless connectivity in dynamic communication environments and their intelligence to adapt to time-varying traffic loads, RF interference, and frequent network failures. Interference avoidance via waveform design has attracted considerable attention toward the development of spectrally efficient cognitive networks [1]–[6]. Particularly, a finite sequence of repeated square-root-raised cosine (SRRC) pulses that span the entire continuum of the device-accessible spectrum is optimized to maximize the signal-to-interference-plus-noise ratio (SINR) at the output of the max-SINR filter at the intended single-antenna receiver.

Multiple-input multiple-output (MIMO) antenna systems can potentially reduce the interference and improve cognitive spectrum reuse by leveraging appropriate space-time precoding and modulation techniques or employing directional transmissions to exploit the increased spatial degrees-of-freedom and/or diversity and coding gains. Recent works such as [7] found that using directional instead of omni-directional antennas in wireless networks can greatly improve network connectivity by concentrating radio signal power on desired directions. The optimization of pre- and post-filters was considered in [8] for a MIMO system distorted by additive noise,

while a paradigm for the design of transmitter space-time coding, referred to as linear precoding is described in [9]. Linear space-time precoding/decoding designs in [9] target different criteria of optimality and constraints, assuming that channel state information (CSI) is known at both the transmitter and the receiver. CSI can be acquired at the transmitter either over a low data-rate feedback channel or when the transmitter and receiver operate in time-division duplex (TDD) so that the time-invariant MIMO channel transfer function is the same in both ways [10]. In [11], it is shown how conventional baseband modulation can be made directional, when the multi-antenna CSI is available to the transmitter. The transmitter modulates the signal with a space-time varying beamformer that triggers a constant spatial response in the geometrical direction of the intended receiver, while triggering redundant and distinct spatial responses in every other direction. Directional modulation that makes the format of the transmitted signal direction-sensitive is used as a secure transmission technique in [12]. Contrary to [11], modulation in the RF domain is achieved with phase-shifters.

In this paper, we consider the problem of directional space-time waveform design for proactive interference avoidance in MIMO link configurations. We propose to instruct the transmitter to communicate with the intended receiver at a given code sequence and angle-of-arrival that are jointly optimized to maximize the maximum attainable pre-detection SINR. Our objective is to design interference-avoiding space-time waveforms, that once made available to the transmitter through a feedback channel, will be used for directional beamforming to the intended receiver. Low decoding complexity is achieved with space-time matched filtering at the receiver. The performance of the proposed waveforms is evaluated in terms of bit-error-rate (BER) and pre-detection SINR with simulations of a 4×4 MIMO communication system that operates in the presence of multi-user interference and additive noise. Joint optimization of the code sequence and angle-of-arrival for the MIMO link demonstrates superior performance compared to interference-suppressive receiver designs and arbitrarily assigned code sequences and transmission angles.

The rest of the paper is organized as follows. Section II describes the MIMO system model and shows that the SINR at the output of the max-SINR receiver filter is a function of the code sequence and received angle-of-arrival. Section III offers

an overview of the optimization problem for the design of directional space-time waveforms. A few concluding remarks are drawn in Section V.

II. SYSTEM MODEL

We consider a MIMO link configuration (Fig. 1) with M transmit and receive antennas, and the n -th information symbol $b[n] \in \{\pm 1\}$ is transmitted at rate $1/T$ on a carrier frequency f_c . The transmitted signal at the m -th transmit antenna is given by

$$x_{m_t}(t) = \sqrt{\frac{E}{M}} \sum_{n=0}^{N-1} b[n] s(t - nT) e^{j(2\pi f_c t + \phi)} \cdot w_{m_t} \quad (1)$$

where $m_t = 1, \dots, M$ and $E > 0$ is the transmitted energy-per-bit, N is the number of transmitted symbols, ϕ is the carrier phase relative to the local oscillator of the multi-antenna transmitter and $w_{m_t} \in \mathbb{C}$ is a complex beamforming weight parameter per transmit antenna. The n -th transmitted symbol $b[n]$ is modulated by a digital coded waveform $s(t)$ of duration T that is given by

$$s(t) = \sum_{l=0}^{L-1} s[l] g_{T_c}(t - lT_c) \quad (2)$$

where $s[l] \in \{\pm 1/\sqrt{L}\}$ is the l -th bit of the code sequence, $g_{T_c}(\cdot)$ is a pulse-shaping SRRC with roll-off factor α and duration T_c , so that $T = LT_c$. The bandwidth of the transmitted signal is $B = (1 + \alpha)/T_c$.

The receiver consists of a uniform linear array with M receiving antenna elements, which are spaced half a wavelength apart. All signals are considered to propagate over independent time-varying flat fading channels with impulse response of the following general form

$$h(t) = c \cdot \delta(t - \tau) \quad (3)$$

where $c \in \mathbb{R}^+$ and τ are the channel amplitude and delay, respectively. We consider a tapped delay line channel model that has a single tap at integer multiples of T_c intervals and both c and τ remain time invariant during N symbol transmission periods. Additionally, we assume that the symbol duration T is chosen to be greater than the channel delay spread i.e. $T \gg T_m = \max_{i,j} |\tau_i - \tau_j|$, thus inter-symbol interference (ISI) is not destructive.

The received signal at the input of the M -antenna linear receiver array after carrier demodulation of the transmitted signal from M transmit antennas is given by

$$\mathbf{y}(t) = \sum_{m_t=1}^M \sqrt{\frac{E}{M}} \sum_{n=0}^{N-1} b[n] \tilde{c}_{m_t} s(t - nT - \tau_{m_t}) \cdot \mathbf{a}(\theta) + \mathbf{i}(t) + \mathbf{n}(t) \quad (4)$$

where $\tilde{c}_{m_t} = w_{m_t} \cdot c_{m_t} \cdot e^{-j(2\pi f_c \tau_{m_t} + \phi)}$ with the total carrier phase and transmit beamforming weights absorbed in the baseband complex channel coefficient and c_{m_t} , τ_{m_t} denote the amplitude and delay of the channel from the m -th transmit

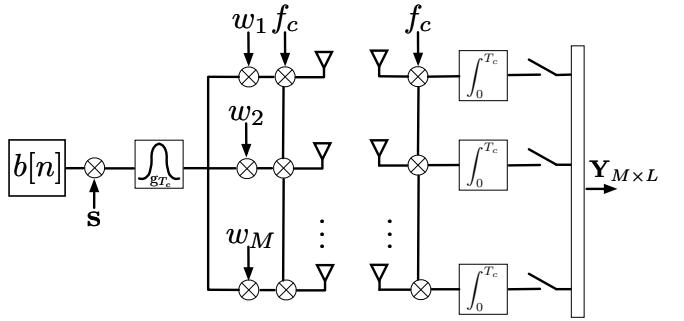


Fig. 1. System model of the M -antenna element MIMO configuration.

antenna, respectively. The m -th component of the $M \times 1$ array response vector $\mathbf{a}(\theta)$ is given by

$$a_m(\theta) = e^{j2\pi(m-1)\sin\theta}, \quad m = 1, \dots, M \quad (5)$$

where θ is the actual incident angle-of-arrival (without loss of generality, we assume that transmitted signals from M antenna elements have the same angle-of-arrival). Finally, $\mathbf{i}(t)$ models environmental disturbance and $\mathbf{n}(t)$ denotes an M -dimensional complex Gaussian noise process that is assumed white both in time and space.

Assuming synchronization at the reference antenna element, matched filtering and sampling of the received carrier-demodulated signal $\mathbf{y}(t)$ at rate $1/T_c$ over the period of LT_c prepares the data for one-shot detection of the n -th information bit of interest $b[n]$. We visualize the collected space-time data in the form of an $M \times (L-1)$ matrix

$$\mathbf{Y}[n] = [\mathbf{y}(0) \quad \mathbf{y}(T_c) \quad \dots \quad \mathbf{y}((L-1)T_c)] \in \mathbb{C}^{M \times L}. \quad (6)$$

We decide to “vectorize” $\mathbf{Y}_{M \times L}$ by sequencing all matrix columns in the form of a vector

$$\mathbf{y}[n] = \text{vec}\{\mathbf{Y}[n]\} \in \mathbb{C}^{ML \times 1} \quad (7)$$

which can be rewritten in the following vector form as

$$\mathbf{y}[n] = \sqrt{\frac{E}{M}} (\mathbf{Hs} \otimes \mathbf{a}(\theta)) \cdot b[n] + \mathbf{i}[n] + \mathbf{n}[n] \quad (8)$$

where \otimes denotes the Kronecker product, $\mathbf{s} \in \{\pm 1/\sqrt{L}\}^L$ is the code sequence, and $\mathbf{H} \in \mathbb{C}^{L \times L}$ is a diagonal channel matrix that is written as

$$\mathbf{H} \triangleq \begin{bmatrix} \sum_{m_t=0}^M \tilde{c}_{m_t} & 0 & \dots & 0 \\ 0 & \sum_{m_t=0}^M \tilde{c}_{m_t} & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & \sum_{m_t=0}^M \tilde{c}_{m_t} \end{bmatrix} \quad (9)$$

where entries $\tilde{c}_{m_t} \in \mathbb{C}$, $m_t = 1, \dots, M$ are considered to be complex Gaussian random variables that model fading phenomena. Finally, $\mathbf{i}[n] \in \mathbb{C}^{ML \times 1}$ models environmental

disturbance (due to ISI and multi-user interference) for the n -th bit and $\mathbf{n}[n] \in \mathbb{C}^{ML \times 1}$ is considered to be zero-mean additive white Gaussian noise with autocorrelation matrix $\sigma^2 \mathbf{I}_{ML}$.

Assuming that the complex channel coefficients, code sequence, and array response vector are known at the receiver, the space-time matched filter is given by

$$\mathbf{w}_{MF} = \mathbf{Hs} \otimes \mathbf{a}(\theta). \quad (10)$$

Subsequently, assuming perfect knowledge of the input autocorrelation matrix, defined as $\mathbf{R} \triangleq \mathbb{E}\{\mathbf{y}[n]\mathbf{y}[n]^H\}$, the receiver filter that minimizes the output variance subject to the constraint that it remains distortionless in the effective direction \mathbf{w}_{MF} (i.e. equivalent to the maximization of the output SINR) is given by [13]

$$\mathbf{w}_{MVDR} = \frac{\mathbf{R}^{-1}\mathbf{w}_{MF}}{\mathbf{w}_{MF}^H \mathbf{R}^{-1} \mathbf{w}_{MF}}. \quad (11)$$

In practice, however, exact knowledge of the input autocorrelation matrix is not available, therefore minimum-variance distortionless response (MVDR) filtering relies on a sample average estimate of the input autocorrelation matrix

$$\widehat{\mathbf{R}} = \frac{1}{N} \sum_{n=0}^{N-1} \mathbf{y}[n]\mathbf{y}[n]^H \quad (12)$$

where N is the available data record over a received signal duration that does not exceed the channel coherence time. Thus, the sample-matrix-inversion (SMI) filter $\mathbf{w}_{SMI} = \widehat{\mathbf{R}}^{-1}\mathbf{w}_{MF}/\mathbf{w}_{MF}^H \widehat{\mathbf{R}}^{-1} \mathbf{w}_{MF}$ offers an unbiased estimator of the MVDR filter.

Information bit detection is carried out optimally in second-order statistics terms via linear maximum SINR filtering (or, equivalently, minimum mean-square-error filtering) as follows

$$\hat{b}[n] = \text{sgn}\left\{\Re\{\mathbf{w}_{SMI}^H \mathbf{y}[n]\}\right\}, \quad n = 0, \dots, N-1. \quad (13)$$

The output SINR of \mathbf{w}_{SMI} is calculated as

$$\text{SINR}(\mathbf{s}, \theta) \triangleq \frac{\mathbb{E}\left\{\left|\mathbf{w}_{SMI}^H \left(\sqrt{\frac{E}{M}}(\mathbf{Hs} \otimes \mathbf{a}(\theta))b[n]\right)\right|^2\right\}}{\mathbb{E}\left\{\left|\mathbf{w}_{SMI}^H (\mathbf{i}[n] + \mathbf{n}[n])\right|^2\right\}} \quad (14)$$

which is proved to be equal to $\text{SINR}(\mathbf{s}, \theta) = \frac{E}{M} \tilde{\mathbf{s}}^H \widehat{\mathbf{R}}^{-1} \tilde{\mathbf{s}}$, where we define $\tilde{\mathbf{s}} \triangleq \mathbf{Hs} \otimes \mathbf{a}(\theta)$. While the code sequence is assumed to be known to the receiver, the channel coefficients and the angles-of-arrival are, in general, unknown. In this paper we are interested in evaluating the performance of max-SINR optimized space-time waveforms $\tilde{\mathbf{s}}(\mathbf{s}, \theta)$ under the assumption of known channel state information at the receiver.

III. DIRECTIONAL SPACE-TIME WAVEFORM DESIGN

We consider the problem of space-time waveform design for interference-avoiding MIMO configurations. Our objective is to design a binary antipodal code sequence $\mathbf{s} \in \left\{\pm \frac{1}{\sqrt{L}}\right\}^L$ and angle-of-arrival θ so that the corresponding post-filtering

SINR(\mathbf{s}, θ) value of \mathbf{w}_{SMI} is maximized. The problem under consideration can therefore be written as

$$(\mathbf{s}, \theta)^{\text{opt}} = \underset{\mathbf{s}, \theta}{\text{argmax}} \left\{ \tilde{\mathbf{s}}^H \widehat{\mathbf{R}}^{-1} \tilde{\mathbf{s}} \right\} \quad (15)$$

$$\text{subject to } \tilde{\mathbf{s}}^H \tilde{\mathbf{s}} = 1 \quad (16)$$

$$-\pi/2 \leq \theta < \pi/2. \quad (17)$$

Since the input autocorrelation matrix $\widehat{\mathbf{R}} \in \mathbb{C}^{ML \times ML}$ is positive definite

$$\widehat{\mathbf{R}} = \sum_{i=1}^{ML} \lambda_i \mathbf{q}_i \mathbf{q}_i^H, \quad \lambda_1 \geq \lambda_2 \geq \dots \geq \lambda_{ML} > 0, \\ \|\mathbf{q}_i\| = 1, \quad i = 1, 2, \dots, ML \quad (18)$$

represents its eigendecomposition where λ_i and \mathbf{q}_i are the i -th eigenvalue and normalized eigenvector, respectively, of $\widehat{\mathbf{R}}$. Then, the complex-valued space-time waveform $\tilde{\mathbf{s}}(\mathbf{s}, \theta)$ that maximizes SINR(\mathbf{s}, θ) is given by [14], [15]

$$\tilde{\mathbf{s}}^{\text{opt}} = \underset{\tilde{\mathbf{s}} \in \mathbb{C}^{ML}, \|\tilde{\mathbf{s}}\|=1}{\text{argmax}} \left\{ \frac{E}{M} \tilde{\mathbf{s}}^H \widehat{\mathbf{R}}^{-1} \tilde{\mathbf{s}} \right\} = \mathbf{q}_{ML}. \quad (19)$$

Thus, we propose to simplify the optimization problem in (15) by using the approximation of $\widehat{\mathbf{R}}^{-1} \simeq \frac{1}{\lambda_{ML}} \mathbf{q}_{ML} \mathbf{q}_{ML}^H$ and rewrite the problem as

$$(\mathbf{s}, \theta)^{\text{opt}} = \underset{\mathbf{s}, \theta}{\text{argmax}} \left\{ \left| \tilde{\mathbf{s}}^H \mathbf{q}_{ML} \right|^2 \right\} \quad (20)$$

$$\text{subject to } \tilde{\mathbf{s}}^H \tilde{\mathbf{s}} = 1 \quad (21)$$

$$-\pi/2 \leq \theta < \pi/2. \quad (22)$$

An optimized pair of code sequence and angle-of-arrival can be easily found by a two-dimensional (2-D) search over 2^L candidate code sequences and angle $\theta \in [-\pi/2, \pi/2]$. The optimized code and angle pair $(\mathbf{s}, \theta)^{\text{opt}}$ is used for the design of an interference-avoiding space-time waveform at the receiver based on (10). Under the assumption that $(\mathbf{s}, \theta)^{\text{opt}}$ is fed back to the M -antenna transmitter within the channel coherence time, we propose to instruct the transmitter to communicate with the intended receiver with code \mathbf{s}^{opt} and complex beamforming weights defined as $w_{mt} \triangleq e^{j2\pi(m_t-1)\sin\theta^{\text{opt}}}$, for $m_t = 1, \dots, M$ to avoid space-time interference.

IV. SIMULATION STUDIES

We consider the simulation of a MIMO link configuration with $M = 4$ transmit and receive antennas that adaptively reconfigures the code sequence $\mathbf{s} \in \left\{\pm 1/\sqrt{L}\right\}^L$ and complex beamforming weights w_{mt} , $m_t = 1, \dots, M$ at the transmitter to avoid space-time interference. We consider transmission of binary-phase shift keying (BPSK) symbols $b[n] \in \{\pm 1\}$ and K single-antenna interferers that are initialized to arbitrary code sequences of length $L = 8$ and transmit beamforming weights $w_{mt}(\theta)$, where $\theta \in [-\pi/2, \pi/2]$. The signal power of the interferers is uniformly spaced between 8dB and 11dB.

Fig. 2 compares the quality of the directional space-time waveform solution $(\mathbf{s}, \theta)^{\text{opt}}$ in problem (20) in terms of pre-detection SINR with the minimum eigenvalue eigenvector

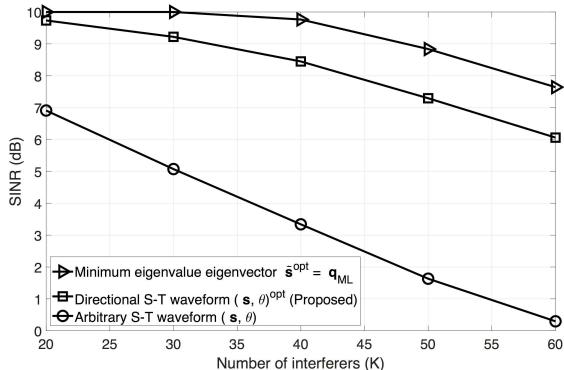


Fig. 2. SINR vs number of (single antenna) interferers for different space-time waveform designs for a 4×4 MIMO configuration for the user of interest.

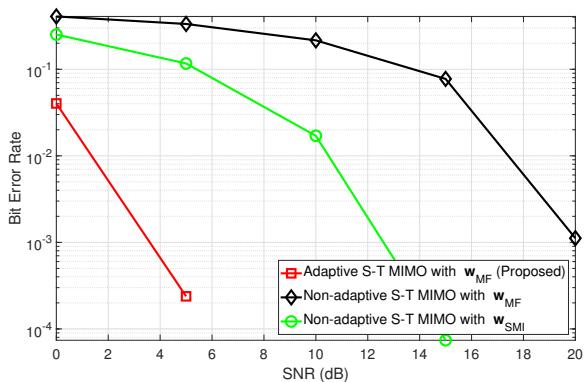


Fig. 3. BER vs SNR for an adaptive and non-adaptive space-time 4×4 MIMO system in the presence of $K = 40$ single antenna interferers.

solution \tilde{s}^{opt} in (19) and arbitrarily designed space-time waveforms used at the transmitter with signal power 10 dB. We observe that the pre-detection SINR at the intended receiver is significantly degraded to 0 dB for $K = 60$ single antenna interferers, while the SINR of the proposed space-time waveform solution is only 2 dB away from the optimal minimum eigenvalue eigenvector waveform solution.

Fig. 3 illustrates the BER versus the signal-to-noise ratio (SNR) for the user of interest for a non-adaptive and an adaptive MIMO system that considers space-time matched filtering with w_{MF} and max-SINR filtering with w_{SMI} . Transmit power is set to SNR/M and the number of interferers is fixed to $K = 40$, while for simplicity in filter design for the adaptive space-time MIMO system, we consider the implementation of w_{MF} only. We observe that matched filtering without adaptation of the code sequence and transmit beamforming weights offers poor BER performance, however the design of an interference-suppressive filter at the receiver (i.e. w_{SMI}) significantly improves the BER performance of the non-adaptive system. The proposed adaptive space-time MIMO system offers additional BER performance improvement by instructing the transmitter to communicate with the intended receiver at an optimized code and angle pair $(s, \theta)^{\text{opt}}$, so that

the corresponding post-filtering SINR(s, θ) is maximized. As a result, proactive interference-avoiding space-time waveform design at the transmitter maximizes the maximum SINR at the intended receiver and allows minimum bit error detection with low decoding complexity (i.e. matched filtering).

V. CONCLUSIONS

We present a waveform design algorithm for directional space-time interference-avoiding MIMO communications. We describe a method to jointly optimize the code sequence and angle-of-arrival in MIMO link configurations to maximize the maximum attainable pre-detection SINR. Simulation studies demonstrate that the proposed space-time waveform designs achieve superior SINR and BER performance with low-complexity receiver designs.

REFERENCES

- [1] G. Sklivanitis, E. Demirors, A. M. Gannon, S. N. Batalama, D. A. Pados, and T. Melodia, "All-spectrum cognitive channelization around narrowband and wideband primary stations," in *2015 IEEE Global Communications Conference (GLOBECOM)*, Dec. 2015, pp. 1–7.
- [2] G. Sklivanitis, A. Gannon, K. Tountas, D. A. Pados, S. N. Batalama, S. Reichhart, M. Medley, N. Thawdar, U. Lee, J. D. Matyjas, S. Pudlewski, A. Drozd, A. Amanna, F. Latus, Z. Goldsmith, and D. Diaz, "Airborne cognitive networking: Design, development, and deployment," *IEEE Access*, vol. 6, pp. 47217–47239, Jul. 2018.
- [3] K. Tountas, G. Sklivanitis, D. A. Pados, and S. N. Batalama, "All-spectrum digital waveform design via bit flipping," in *2018 IEEE Global Communications Conference (GLOBECOM)*, Dec. 2018, pp. 1–6.
- [4] L. Ding, K. Gao, T. Melodia, S. N. Batalama, D. A. Pados, and J. D. Matyjas, "All-spectrum cognitive networking through joint distributed channelization and routing," *IEEE Transactions on Wireless Communications*, vol. 12, no. 11, pp. 5394–5405, Nov. 2013.
- [5] K. Gao, S. N. Batalama, D. A. Pados, and J. D. Matyjas, "Cognitive code-division channelization," *IEEE Transactions on Wireless Communications*, vol. 10, no. 4, pp. 1090–1097, Apr. 2011.
- [6] G. Sklivanitis, P. P. Markopoulos, S. N. Batalama, and D. A. Pados, "Sparse waveform design for all-spectrum channelization," in *2017 IEEE International Conference on Acoustics, Speech and Signal Processing (ICASSP)*, March 2017, pp. 3764–3768.
- [7] H. Dai, K. Ng, R. Wong, and M. Wu, "On the capacity of multi-channel wireless networks using directional antennas," in *IEEE INFOCOM 2008*, Apr. 2008, pp. 628–636.
- [8] H. S. Malvar and D. H. Staelin, "Optimal pre- and postfilters for multichannel signal processing," *IEEE Transactions on Acoustics, Speech, and Signal Processing*, vol. 36, no. 2, pp. 287–289, Feb. 1988.
- [9] A. Scaglione, P. Stoica, S. Barbarossa, G. B. Giannakis, and H. Sampath, "Optimal designs for space-time linear precoders and decoders," *IEEE Transactions on Signal Processing*, vol. 50, no. 5, pp. 1051–1064, May 2002.
- [10] and G. B. Giannakis, "Optimal transmitter eigen-beamforming and space-time block coding based on channel correlations," *IEEE Transactions on Information Theory*, vol. 49, no. 7, pp. 1673–1690, Jul. 2003.
- [11] O. N. Alrabadi and G. F. Pedersen, "Directional space-time modulation: A novel approach for secured wireless communication," in *2012 IEEE International Conference on Communications (ICC)*, Jun. 2012, pp. 3554–3558.
- [12] M. P. Daly, E. L. Daly, and J. T. Bernhard, "Demonstration of directional modulation using a phased array," *IEEE Transactions on Antennas and Propagation*, vol. 58, no. 5, pp. 1545–1550, May 2010.
- [13] D. A. Pados and S. N. Batalama, "Joint space-time auxiliary-vector filtering for ds/cdma systems with antenna arrays," *IEEE Transactions on Communications*, vol. 47, no. 9, pp. 1406–1415, Sep. 1999.
- [14] C. Rose, S. Ulukus, and R. D. Yates, "Wireless systems and interference avoidance," *IEEE Transactions on Wireless Communications*, vol. 1, no. 3, pp. 415–428, July 2002.
- [15] G. N. Karystinos and D. A. Pados, "Rank-2-optimal adaptive design of binary spreading codes," *IEEE Transactions on Information Theory*, vol. 53, no. 9, pp. 3075–3080, Sep. 2007.