A Novel Approach to MIMO Transmission Using a Single RF Front End

Antonis Kalis, *Member, IEEE,* Athanasios G. Kanatas, *Senior Member, IEEE,* and Constantinos B. Papadias, *Senior Member, IEEE*

***Abstract*—In this paper we introduce a new perspective to the implementation of wireless MIMO transmission systems with increased bandwidth efficiency. Unlike traditional spatial multiplexing techniques in MIMO systems, where additional information can be sent through the wireless channel by feeding uncorrelated antenna elements with diverse bitstreams, we use the idea of mapping diverse bitstreams onto orthogonal bases defined in the beamspace domain of the transmitting array far-field region. Using this approach we show that we can increase the capacity of wireless communication systems using compact parasitic antenna architectures and a single RF front end at the transmitter, thus paving the way for integrating MIMO systems in cost and size sensitive wireless devices such as mobile terminals and mobile personal digital assistants.**

***Index Terms*—Antenna arrays, MIMO systems, Beamspace, Diversity, Spatial multiplexing, ESPAR antennas.**

I. Introduction

OVER the past decade a great deal of effort and literature has been devoted in taking advantage of the properties of the wireless channel to increase the performance of wireless communication systems. These efforts have built upon the ground-breaking work of [1][2] and [3] in defining the achievable capacity through wireless communication channels. Various multiple-input multiple-output (MIMO) architectures as well as diversity techniques have been employed, to achieve greater capacity or more robust error performance in diverse channel conditions. In order for conventional MIMO systems to work, antenna elements are typically driven separately by uncorrelated signals, implying that MIMO transmitters should include as many separate radio frequency (RF) front-ends as the total number of antennas used. Moreover, in order to ensure uncorrelated fading characteristics of the wireless channel, antenna elements should have a substantial spacing of at least half a wavelength, leading to large antenna array implementations [4]. Therefore, the cost of implementing multiple antenna structures with multiple RF chains and the large inter-element distances required to ensure the orthogonality of the system's input signals, makes MIMO systems technically difficult to implement for cost and size sensitive applications such as mobile telephony and mobile computing. As a result, recent research focuses on implementing MIMO transmission schemes only on the base station side of the communication link, where the size and cost limitations are less strict and enable the use of multiple antenna systems.

In this work we propose the implementation of a MIMO transmitter for MIMO wireless communications using a single RF front end at the mobile terminal, by mapping symbols in the wavevector domain of the antenna array, instead of the traditional approach of sending different symbol streams in different antenna elements. The wavevector domain has attracted extended attention in recent literature. In [5] A. Sayeed pioneered the virtual channel representation for uniform linear arrays, using steering vectors on predefined spatial directions for the transmission and reception of energy. In [6], using analysis in the wave-vector domain, the degrees of freedom of multi-element antenna channels were investigated. The capacity of multi-element array (MEA) systems that use the wavevector domain for increasing the spectral efficiency of wireless communications has been introduced in [7] for uniform linear arrays (ULA), and in [8] for switched parasitic arrays (SPA).

This paper builds upon the approach introduced in [7] and [8] to show that MIMO transmission is feasible using a single RF front end and compact arrays. Receiver processing is not the focus of this paper, however we will assume that the receiver is equipped with a uniform linear array (ULA) with multiple RF front ends for conventional MIMO reception. According to this approach, we shift our attention from the inputs of the diverse antenna elements to the use of different radiation patterns in the wavevector domain of the transmitting antenna array. We follow [2] and [3] in the concept that maximum open loop capacity in a MIMO system is achieved when signals that are fed into a set of antenna elements (in the absence of channel knowledge) have an identity covariance matrix in the antenna domain, and we extend these findings by using a set of orthogonal functions in the wavevector domain. By disjoining the system capacity from the inputs of the antenna elements we are able to demonstrate that capacity gains can be achieved in rich scattering environments using a single RF front end and *a parasitic array capable of changing its radiation pattern on each symbol period*. Parasitic arrays have already drawn considerable attention in the literature, since they provide a cost-effective solution for beamforming [9][10], and diversity applications [11]. They have also been used in [12] for evaluating MIMO systems using directional transmissions; however, their use at the transmitter of the communication link for spatial multiplexing has not yet been described, and is presented in this work.

The proposed approach is presented in the rest of the paper as follows: In Section II previous findings on the capacity of

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Antonis Kalis and Constantinos B. Papadias are with Athens Information Technology, 0.8 km Markopoulou Ave., 19002 Paiania, Attika, Greece ({akal,papadias}@ait.edu.gr).

Athanasios G. Kanatas is with Department of Technology Education and Digital Systems, University of Piraeus, 80 Karaoli, Dimitriou Street, GR-185 34, Piraeus, Greece (kanatas@unipi.gr).

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beamspace MIMO systems are reviewed taking into consideration both statistical and virtual channel models, and we investigate how these findings are applied if we employ vectors of signals on the wave-vector domain of the transmitting antenna, while sampling the channel on the antenna domain of the receiving antenna. The performance of the proposed systems is evaluated in MIMO and transmit diversity scenarios. In section III, parasitic array implementations capable of creating multiple channels with a single RF front end are presented. Finally, in Section IV system examples are described that increase the capacity or the robustness of MIMO systems using compact antennas and a single RF front end at the transmitter side of the communication link.

II. Overview of Beamspace MIMO Systems

A. Review of System Model

In order to evaluate the performance of MIMO systems in the beamspace domain a parametric physical model that considers the geometry of the scattering environment is required. Such models have extensively been studied in the literature [13][14]. In these models, each path i connecting the area of the transmitter with the area of the receiver has a single angle-of-departure (AoD) $\theta_{T,i}$ and a single angle-of-arrival (AoA) $\theta_{R,i}$, and a path gain b_i. If K such paths exist, then assuming $\mathbf{a}_R(\theta_{R,i})$, $\mathbf{a}_T(\theta_{T,i})$ the steering vectors of the receiving and transmitting arrays, respectively, the channel response can be written as

$$\begin{aligned}\mathbf{H} &= \sum_{i=1}^{K} b_i \mathbf{a}_R(\theta_{R,i})\mathbf{a}_T^H(\theta_{T,i}) \\ &= \mathbf{A}_R(\widehat{\Theta}_R)\mathbf{H}_b\mathbf{A}_T^H(\widehat{\Theta}_T)\end{aligned} \tag{1}$$

where $\widehat{\Theta}_R$, $\widehat{\Theta}_T$ are the direction vectors of the AoA and AoD respectively, $\mathbf{A}_R(\widehat{\Theta}_R)$, $\mathbf{A}_T(\widehat{\Theta}_T)$ are the $M_{R,T} \times K$ receive and transmit steering matrices, and $\mathbf{H}_b$ is a diagonal $K \times K$ matrix whose entries represent the complex gain of each path. This approach however leads to a representation that is nonlinear in the domain of the direction vectors. We overcome this problem using a virtual representation of the channel model, as proposed in [5]:

$$\begin{aligned}\mathbf{H} &= \sum_{i=-\lfloor M_R/2 \rfloor}^{\lceil M_R/2 \rceil} \sum_{j=-\lfloor M_T/2 \rfloor}^{\lceil M_T/2 \rceil} H_V(i,j)\mathbf{a}_R(\widetilde{\theta}_{R,i})\mathbf{a}_T^H(\widetilde{\theta}_{T,i}) \\ &= \widetilde{\mathbf{A}}_R\mathbf{H}_V\widetilde{\mathbf{A}}_T^H\end{aligned} \tag{2}$$

where, $\widetilde{\theta}_{R,i}$, $\widetilde{\theta}_{T,j}$, are the virtual directions uniformly sampling the spatial channel, $\mathbf{H}_V$ is the virtual channel representation whose entries $\{H_V(i,j)\}$ represent the coupling between the virtual transmit and receive angles, and $\widetilde{\mathbf{A}}_R, \widetilde{\mathbf{A}}_T$ are the steering matrices corresponding to the virtual directions. Iff $\widetilde{\mathbf{A}}_R, \widetilde{\mathbf{A}}_T$ are unitary, the virtual channel matrix can be also written as,

$$\mathbf{H}_V = \widetilde{\mathbf{A}}_R^H\mathbf{A}_R(\widehat{\Theta}_R)\mathbf{H}_b\mathbf{A}_T^H(\widehat{\Theta}_T)\widetilde{\mathbf{A}}_T = \widehat{\mathbf{A}}_R\mathbf{H}_b\widehat{\mathbf{A}}_T^H \tag{3}$$

where $\widehat{\mathbf{A}}_R = \widetilde{\mathbf{A}}_R^H\mathbf{A}_R(\widehat{\Theta}_R)$ and $\widehat{\mathbf{A}}_T = \widetilde{\mathbf{A}}_T^H\mathbf{A}_T(\widehat{\Theta}_T)$ are the projections of $\mathbf{A}_R(\widehat{\Theta}_R)$ and $\mathbf{A}_T(\widehat{\Theta}_T)$ respectively, onto the square steering matrices that correspond to the virtual directions. If we consider that the transmitter maps symbols on the virtual angles of the transmitting array, while the receiver processes the data in the antenna domain using a traditional ULA, then the received signal vector corresponding to one symbol period may be written using the virtual channel model as

$$\begin{aligned}\mathbf{y} &= \mathbf{H}\mathbf{x} + \mathbf{n} \\ &= \widetilde{\mathbf{A}}_R\mathbf{H}_V\widetilde{\mathbf{A}}_T^H\mathbf{x} + \mathbf{n} \\ &= \widetilde{\mathbf{A}}_R\widehat{\mathbf{A}}_R\mathbf{H}_b\widehat{\mathbf{A}}_T^H\widetilde{\mathbf{A}}_T^H\mathbf{x} + \mathbf{n}\end{aligned} \tag{4}$$

Considering the classic beam-space processing approach we define a system where the transmitter is sending diverse symbols streams towards the channel's virtual angles and a ULA based MIMO receiver that performs spatial sampling in the antenna domain. In this case, the input of the system will be equal to

$$\mathbf{x}_{bs} = \widetilde{\mathbf{A}}_T^H\mathbf{x} \tag{5}$$

The $(K \times M_T)$ matrix $\widehat{\mathbf{A}}_T^H$ contains M_T column vectors of length K , representing the vector functions of a basis towards the scatterers, i.e.

$$\widehat{\mathbf{A}}_T^H = \mathbf{B}_T = \left[B_{T,1}(\widehat{\Theta}_T)\ B_{T,2}(\widehat{\Theta}_T)\ \dots\ B_{T,M_T}(\widehat{\Theta}_T)\right] \tag{6}$$

Therefore, the system model may now be written as

$$\begin{aligned}\mathbf{y} &= \mathbf{A}_R(\widehat{\Theta}_R)\mathbf{H}_b\mathbf{B}_T\mathbf{x}_{bs} + \mathbf{n} \\ &= \widetilde{\mathbf{H}}_V\mathbf{x}_{bs} + \mathbf{n}\end{aligned} \tag{7}$$

where the product $\mathbf{B}_T\mathbf{x}_{bs}$ represents the actual radiation pattern created at the transmitter at every symbol period, and the elements of the matrix $\widetilde{\mathbf{H}}_V$ represent the coupling between the virtual transmit angles and the receive antenna elements. One key element of this system representation is that it enables us to disassociate the system performance from the transmit antenna structure, and take into account only the effect of the transmitter radiation pattern on every symbol period according to the input vector $\mathbf{x}_{bs}$. The advantage of this approach is that the set of radiation patterns needed for the implementation of the system described by (7) can be produced using a single RF front end and switched parasitic arrays as described in Section III. In the next paragraphs we evaluate the capacities of the aforementioned system in open loop MIMO and MISO scenarios.

B. Capacity of Beamspace-MIMO Systems

Depending on the time-variation of the channel, there are different quantities that characterize the capacity of the resulting system. Although the capacity of BS-MIMO systems has been already studied in [7], in this paragraph we follow a slightly different approach to elaborate on the capacity of the proposed technique, where symbols are transmitted in the beamspace domain and received in the antenna domain using

ULAs. Deterministic capacity is a meaningful quantity when a static channel model is adopted, which implies that the channel matrix, despite being random, once chosen it is held fixed for the whole transmission. In this case, the Shannon capacity of the system described by (7) is given by

$$C = \log_2 \det \left[\mathbf{I}_{M_R} + \frac{1}{N_0} \widetilde{\mathbf{H}}_V R_{\mathbf{x}_{bs}} \widetilde{\mathbf{H}}_V^H \right] \tag{8}$$

where $tr(R_{\mathbf{x}_{bs}}) = P$, is the power constraint on the transmitter, $\rho = \frac{P}{N_0}$, is the intended average signal-to-noise ratio per antenna at the receiver, $\mathbf{x}_{bs}$, is a zero mean circularly symmetric complex Gaussian vector (ZMCSCGV). Since the transmitter does not know the channel and taking into account the power constraint, it is reasonable to assume that $R_{\mathbf{x}_{bs}} = \frac{P}{M_T}\mathbf{I}_{M_T}$. Therefore, the Shannon capacity is

$$\begin{aligned} C &= \log_2 \det \left[\mathbf{I}_{M_R} + \frac{\rho}{M_T} \widetilde{\mathbf{H}}_V \widetilde{\mathbf{H}}_V^H \right] \\ &= \log_2 \det \left[\mathbf{I}_{M_R} + \frac{\rho}{M_T} \widetilde{\mathbf{A}}_R \mathbf{H}_V \mathbf{H}_V^H \widetilde{\mathbf{A}}_R^H \right] \\ &= \sum_{m=1}^{M_R} \log_2 \left(1 + \frac{\rho}{M_T} \lambda_m (\widetilde{\mathbf{A}}_R \mathbf{H}_V \mathbf{H}_V^H \widetilde{\mathbf{A}}_R^H) \right) \end{aligned} \tag{9}$$

where $\lambda_m(\mathbf{X})$ denotes the m-th eigenvalue of square matrix $\mathbf{X}$ in descending order. Poincare separation theorem states that

$$\lambda_m (\widetilde{\mathbf{A}}_R \mathbf{H}_V \mathbf{H}_V^H \widetilde{\mathbf{A}}_R^H) \leq \lambda_m (\mathbf{H}_V \mathbf{H}_V^H) \tag{10}$$

with equality occurring when the columns of $\widetilde{\mathbf{A}}_R$ are the M_R dominant singular vectors of $\mathbf{H}_V$. Thus,

$$\begin{aligned} C &\leq \sum_{k=1}^{M_R} \log_2 \left(1 + \frac{\rho}{M_T} \lambda_k (\mathbf{H}_V \mathbf{H}_V^H) \right) \\ &= \log_2 \det \left[\mathbf{I}_{M_R} + \frac{\rho}{M_T} \mathbf{H}_V \mathbf{H}_V^H \right] \end{aligned} \tag{11}$$

where equality occurs when

$$\widetilde{\mathbf{A}}_R = \left[\mathbf{u}_1 \, \mathbf{u}_2 \, \mathbf{u}_3 \, \ldots \, \mathbf{u}_{M_R} \right] \tag{12}$$

and $\mathbf{u}_k$ is the k^{th} dominant singular vector of $\mathbf{H}_V$. Therefore, the capacity upper bound of the proposed system corresponds to the standard open loop capacity expression of a MIMO system with channel matrix $\mathbf{H}_V$. In fully scattered environments, the statistics of $\mathbf{H}_V$ and $\mathbf{H}$ are the same, resulting in equivalent capacities [5].

It is demonstrated via numerical results that the actual capacity of our system is comparable to that of an i.i.d. Rayleigh MIMO channel. Namely, we simulate a simple 2 by 2 MIMO system mapping diverse bit streams in the wavevector domain of the transmitting antenna array. We consider an outdoor propagation scenario, described by a 2-D geometrically based circular channel model as that described in [14], where the average number of scatterers is set to 100, the scatterers are uniformly distributed in a circle of radius $R = 100m$ around the transmitter and the distance between the transmitter and the receiver is equal to 1Km. We assume that all signals are transmitted from a single location corresponding to the phase centre of the transmitting array and received by a 2-element ULA. Multiple symbol streams are transmitted simultaneously, being mapped on a set of basis radiation patterns (wavevector functions) that are orthogonal in the domain of virtual transmit angles $[0, \pi]$. The choice of basis functions that are of interest to this work are cardioid functions of the form

$$\begin{aligned} B_{T,1}(\theta) &= 1 + cos(\theta) \\ B_{T,2}(\theta) &= 1 - cos(\theta) \end{aligned} \tag{13}$$

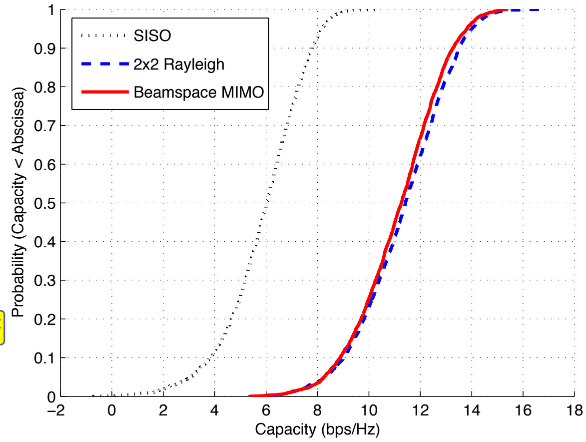


Fig. 1. Empirical capacity CDF of the proposed BS-MIMO architecture for $SNR = 20$ dB

These are representative patterns for parasitic arrays (e.g. [9-12]) considered for implementing the proposed communication link using a single RF front end at the transmitter. Figure 1 shows the cumulative distribution function (CDF) of the system capacity achieved using the aforementioned basis radiation patterns when no knowledge of the channel state information is present at the transmitter, compared to the capacity of a traditional 2 by 2 MIMO system. It is therefore evident that the proposed system achieves spectral efficiencies comparable to these of traditional 2 by 2 MIMO systems in fully scattered environments, as claimed from the theoretical analysis.

C. Beamspace Transmit Diversity

In the previous sub-section we have reviewed the performance of BS-MIMO transmission systems that increase the spectral efficiency of the communication link. Space-time coding schemes can also be used to increase the robustness of wireless communications using transmit diversity algorithms. In this paragraph we examine how transmit diversity systems would perform in the beamspace domain when no channel state information (CSI) is present at the transmitter. Namely, we consider one of the most successful transmit diversity schemes that increases the link robustness with no CSI at the transmitter, which is the one introduced by Alamouti in [15]. This scheme is an orthogonal space-time design, used to provide an efficient means to generate codes that achieve the full diversity gain. According to the Alamouti scheme,

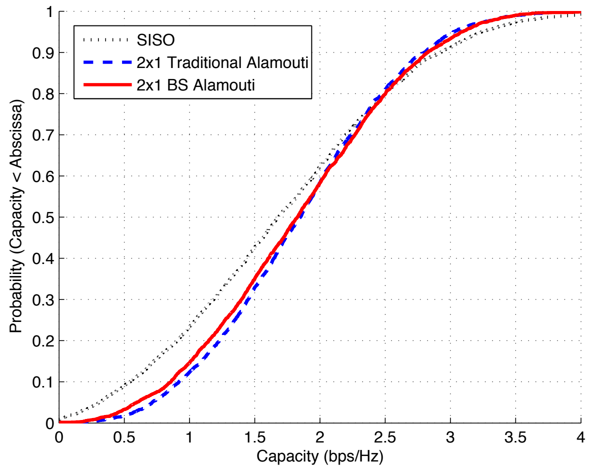


Fig. 2. Empirical capacity CDF of the proposed BS-Alamouti architecture for $SNR = 5\text{dB}$

the 2×2 space-time transmission matrix for the two symbols $\mathbf{x} = [\, x_1 \; x_2]^T$, is

$$\mathbf{X} = \sqrt{\frac{E_s}{2}} \begin{bmatrix} x_1 & x_2 \\ -x_2^* & x_1^* \end{bmatrix} \tag{14}$$

where $(E_s/2)$ is the average transmit energy per symbol period per antenna. The rearranged symbols received over two symbol periods through the wireless channel are

$$\mathbf{y} = \begin{bmatrix} y_1 \\ y_2^* \end{bmatrix} = \sqrt{\frac{E_s}{2}} \begin{bmatrix} h_1 & h_2 \\ h_2^* & -h_1^* \end{bmatrix} \mathbf{x} + \mathbf{n} = \sqrt{\frac{E_s}{2}} \mathbf{H}\mathbf{x} + \mathbf{n} \tag{15}$$

where we have assumed that the channel remains constant over the two symbols period and is frequency flat, and $\mathbf{n} = [n_1 n_2^*]^T$ is the noise vector for the two consecutive time periods, assumed to be ZMCSCG. At the output of the linear combiner of the receiver, the received vector is simply described by the left multiplication with the conjugate transposed matrix $\mathbf{H}^H$:

$$\begin{aligned} \widetilde{\mathbf{y}} &= \sqrt{\frac{E_s}{2}} \mathbf{H}^H \mathbf{H}\mathbf{x} + \mathbf{H}^H \mathbf{n} \\ &= \sqrt{\frac{E_s}{2}} a \mathbf{I}\mathbf{x} + \widetilde{\mathbf{n}} \end{aligned} \tag{16}$$

where $a = (|h_1|^2 + |h_2|^2)$ is the effective attenuation by the channel. Since the columns of the matrix $\mathbf{H}$ are orthogonal, the components of $\widetilde{\mathbf{n}}$ are uncorrelated ZMCSCG, and the received symbols will have a signal-to-noise ratio (SNR), that is equal to the mean SNR of the two diverse channels used, therefore averaging out deep channel fades and high channel gains. In this work we exploit the Alamouti transmit diversity scheme by applying the orthogonal coding scheme onto the beamspace domain. Assume for example that we want to transmit the symbol vector $\mathbf{x}_{bs}$ using a set of orthogonal functions in the beamspace domain $\left\{ B_{T,i}(\widehat{\Theta}_T), i = 1, 2, \ldots, M_T \right\}$. The radiation pattern on each symbol period using this approach will be equal to $\mathbf{B}_T \mathbf{x}_{bs}$ as in (7). Following the analysis in Section II.A, the system will therefore be described by

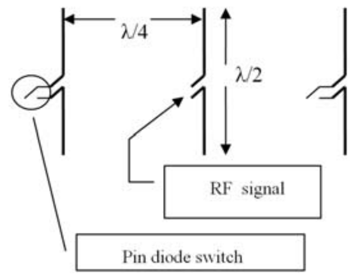


Fig. 3. A 3-element switched parasitic dipole array

$$y = [\, \widetilde{h}_{V1} \; \widetilde{h}_{V2} \,] \mathbf{x}_{bs} + n \tag{17}$$

where $\widetilde{h}_{V,i}$ represents the coupling between the orthogonal function $\left\{ B_{T,i}(\widehat{\Theta}_T), i = 1, 2, \ldots, M_T \right\}$ of the transmitter and the omni-directional antenna at the receiver. We may now define a 2×2 beamspace-time transmission matrix for the input vector $\mathbf{x}_{bs} = [\, s_0 \; s_\pi \,]^T$, according to the Alamouti scheme,

$$\mathbf{X}_{bs} = \sqrt{\frac{E_s}{2}} \begin{bmatrix} s_0 & s_\pi \\ -s_\pi^* & s_0^* \end{bmatrix} \tag{18}$$

where $(E_s/2)$ is the average transmit energy per symbol period per orthogonal basis function. The signal arriving at the receiver will be

$$\mathbf{y}_{bs} = \sqrt{\frac{E_s}{2}} \begin{bmatrix} \widetilde{h}_{V1} & \widetilde{h}_{V2} \\ \widetilde{h}_{V2}^* & -\widetilde{h}_{V1}^* \end{bmatrix} \mathbf{x}_{bs} + \mathbf{n}_{bs} = \sqrt{\frac{E_s}{2}} \widetilde{\mathbf{H}}_V \mathbf{x}_{bs} + \mathbf{n}_{bs} \tag{19}$$

Note that the matrix $\widetilde{\mathbf{H}}_V$ preserves beamspatial and temporal orthogonality, i.e.,

$$\widetilde{\mathbf{H}}_V \widetilde{\mathbf{H}}_V^H = \widetilde{\mathbf{H}}_V^H \widetilde{\mathbf{H}}_V = (|\widetilde{h}_{V1}|^2 + |\widetilde{h}_{V2}|^2) \mathbf{I}_2 \tag{20}$$

Due to the orthogonality of the code, it turns out that the mean SNR of the received signal for the 2 by 1 transmit diversity system implemented in the wavevector domain is equal to the mean SNR of the two diverse channel used, as in the traditional Alamouti case. We have therefore shown that transmit diversity schemes following the Alamouti space-time codes may be implemented by sending diverse symbols simultaneously towards different virtual angles of the wireless channel by using orthogonal bases in the wavevector domain, instead of feeding diverse elements in the antenna domain. This is also shown in Figure 2 through simulation results, where the simulation parameters used are the same as in the BS-MIMO case presented in the previous paragraph.

III. Parasitic Arrays for BS-MIMO Transmission

In this section we present how parasitic arrays that are able to transmit two symbol streams simultaneously towards the channel's virtual angles using a single RF front end are implemented. We consider for this task both SPA and

Electronically Steerable Passive Array Radiator (ESPAR) antenna configurations for On-Off-Keying (OOK) and phase-shift-keying (PSK) modulation formats.

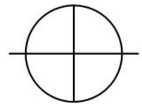
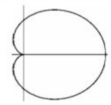
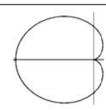
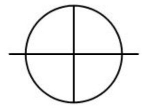
A. SPA for On-Off Keying Modulation

We consider a switched parasitic array of half-wave dipoles as the one described in Figure 3, consisting of one central active element driven by the power amplifier (PA) of the device, and two co-linear parasitic elements that may be either open or short circuited using high speed p-i-n diodes and digital control signals. When both peripheral elements are open-circuited, the radiation pattern is omnidirectional in the azimuth plane. When one peripheral element is open circuited and the other short circuited, then a cardioid pattern is produced with a maximum towards the direction of the short-circuited element, corresponding to the functions described in (13), that are orthogonal towards the virtual angles 0 and π. Assume that we want to transmit simultaneously a symbol vector $\mathbf{x}_{bs} = [\, s_0 \; s_\pi]^T$ digitally modulated by OOK, using the aforementioned orthogonal functions. The set of possible radiation patterns on every symbol period are

$$\begin{aligned} \mathbf{B}_T[\,1 \quad 1\,]^T &= B_{T,1}(\widehat{\Theta}_T) + B_{T,2}(\widehat{\Theta}_T) \\ \mathbf{B}_T[\,1 \quad 0\,]^T &= B_{T,1}(\widehat{\Theta}_T) \\ \mathbf{B}_T[\,0 \quad 1\,]^T &= B_{T,2}(\widehat{\Theta}_T) \\ \mathbf{B}_T[\,0 \quad 0\,]^T &= 0 \end{aligned} \tag{21}$$

This set may be produced using the SPA of Figure 3 in the following way: On every symbol period we feed the active element of the antenna with a carrier signal, with power equal to $\sqrt{|s_0|^2 + |s_\pi|^2}$, and control the p-i-n diodes of the parasitic elements using the digital signals s_0 and s_π. The radiation patterns produced in this manner (shown in Table 1) can be described with respect to the orthogonal functions of the transmitter by the product $\mathbf{B}_T\mathbf{x}_{bs}$ as in (21), implying that by using the radiation patterns of Table 1, two OOK modulated symbols may be transmitted simultaneously towards different directions of the wireless channel on each symbol period. One of the advantages of this configuration is that no power dividers/combiners are needed on the feeding network of the antenna, however a reconfigurable matching circuit should be used, common for variable gain antennas as described in [16]. Moreover, in order to mitigate any non-linearities caused by switching between different antenna patterns, pulse-shaping filters may be applied to the RF chain and to the control lines of parasitic elements to facilitate the smooth transition of the antenna patterns.

TABLE I
RADIATION PATTERNS FOR OOK MODULATION.

s_0	s_π	Radiation Pattern
0	0	
0	1	
1	0	
1	1	

B. ESPAR antennas for BPSK and QPSK modulations

ESPAR antennas [17] are smart antenna systems that are able to control their beam patterns as any smart antenna system, while being implemented using a single active antenna element and a number of parasitic elements placed on a circle around the active element. The parasitic elements are short-circuited and loaded with variable reactors (varactors) that control the imaginary part of the parasitic elements' input impedances. By adjusting the varactors' values, the radiation pattern of the ESPAR antenna system can be controlled to direct its beams and nulls towards certain directions in an adaptive or predefined fashion. ESPAR antennas present an attractive solution for small wireless devices due to the simplicity of the antenna feeding network, the small inter-element spacing that can be as small as 0.05λ [18], and to the fact that compact antenna structures can be used as active and parasitic elements [19]. They have already been used at the receiver for adaptive beamforming applications [20] and receive diversity schemes [21], where the antenna is controlled to maximize the received signal-to-noise-ratio (SNR) and the signal-to-noise-and-interference-ratio (SNIR). In this work we focus on using ESPAR antennas on the transmitter side for implementing MIMO/MISO transmitters. We consider 3-element and 5-element ESPAR antenna configurations as shown in Figure 4. The inter-element spacing, d, of the dipoles is used as a system parameter ranging from $\lambda/2$, to $\lambda/16$. The radiation pattern of the ESPAR antenna configuration [17] is given by,

$$A(\theta) = \mathbf{i}^T \mathbf{a}(\theta) \tag{22}$$

where $\mathbf{i}$ is the complex current vector on the antenna elements and $\mathbf{a}(\theta)$ is the steering vector of the array. Following the notation of ESPAR antennas, the first element of the steering vector corresponds to the central active element of the array, while the rest of the elements correspond to the parasitic antennas of the array. When we feed only the central element, the current vector $\mathbf{i}$ is given by

$$\mathbf{i} = \frac{1}{C}\left[\mathbf{I} + \mathbf{YX}\right]^{-1} \mathbf{y}_0 \tag{23}$$

where for the 3-element ESPAR,

$$\mathbf{y}_0 = [\; y_{00} \quad y_{10} \quad y_{10}\;]^T,$$

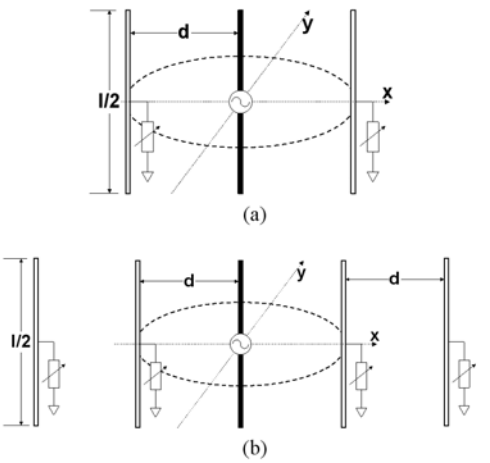


Fig. 4. 3-element (a) and 5-element (b) ESPAR antenna architectures

$$\mathbf{Y} = \begin{pmatrix} y_{00} & y_{10} & y_{10} \\ y_{10} & y_{11} & y_{21} \\ y_{10} & y_{21} & y_{11} \end{pmatrix}$$

and

$$\mathbf{X} = \begin{pmatrix} R_0 & 0 & 0 \\ 0 & jx_1 & 0 \\ 0 & 0 & jx_2 \end{pmatrix} \tag{24}$$

$\mathbf{Y}$ is the admittance matrix that expresses the mutual coupling between the elements, $\mathbf{X}$ is the reactance matrix that adjusts the pattern of the antenna, and $C = 2R_0\sqrt{D}$, $D = 1.64$, $R_0 = 50\Omega$, for dipoles. For the 5-element ESPAR antenna of Figure 4b, $\mathbf{y}_0, \mathbf{X}$ and $\mathbf{Y}$ become

$$\mathbf{y}_0 = [\; y_{00} \;\; y_{10} \;\; y_{20} \;\; y_{10} \;\; y_{20} \;]^T,$$

$$\mathbf{Y} = \begin{pmatrix} y_{00} & y_{10} & y_{20} & y_{10} & y_{20} \\ y_{10} & y_{11} & y_{21} & y_{31} & y_{32} \\ y_{20} & y_{21} & y_{22} & y_{32} & y_{42} \\ y_{10} & y_{31} & y_{32} & y_{11} & y_{21} \\ y_{20} & y_{32} & y_{42} & y_{21} & y_{22} \end{pmatrix}$$

and

$$\mathbf{X} = \begin{pmatrix} R_0 & 0 & 0 & 0 & 0 \\ 0 & jx_1 & 0 & 0 & 0 \\ 0 & 0 & jx_2 & 0 & 0 \\ 0 & 0 & 0 & jx_3 & 0 \\ 0 & 0 & 0 & 0 & jx_4 \end{pmatrix} \tag{25}$$

The response of the arrays is controlled by setting the reactive weights $x_1, x_2, \ldots, x_n$ so that the resulting radiation pattern corresponds to the product $\mathbf{B}_T\mathbf{x}_{bs}$ of (7), where the set of basis functions $\mathbf{B}_T$ is in our case described by (13). This requires the solution of the aforementioned equations so as to find the unknown weights. Since these are non-linear equations, this is not a trivial task. Pre-coding the transmission vector requires the antenna array to be controlled adaptively. Fortunately, since the constellation of all possible symbol combinations is finite and known at the transmitter, and the transmission scheme does not adapt to channel conditions in open-loop scenarios, we need not rely on numerical methods in [20], where the antenna is used at the receiver. We may use an exhaustive search approach to compute the set of reactive values that produce all the possible products $\mathbf{B}_T\mathbf{x}_{bs}$ for the given modulation scheme and set of basis functions, and hard-wire the results in the device to be used in all channel conditions. According to the exhaustive search algorithm, the array factor for the ESPAR antenna is computed for all possible varactor value combinations within a range of $-j500$ to $j500$ ohm, with a step of 1 ohm. A look-up table is created and the closest match to the table values using the Euclidean distance metric is obtained for each sampled direction of the channel. The result of this process is the set of varactor values that produce the desired radiation pattern. An application of the exhaustive search algorithm is depicted using a simple BPSK example and the 3-element ESPAR antenna of Figure 4a. The possible radiation patterns within a single symbol period are

$$\begin{aligned} \mathbf{B}_T[\;1 \quad 1]^T &= \quad B_{T,1}(\widehat{\Theta}_T) + B_{T,2}(\widehat{\Theta}_T) \\ \mathbf{B}_T[\;1 \;\, -1]^T &= \quad B_{T,1}(\widehat{\Theta}_T) - B_{T,2}(\widehat{\Theta}_T) \\ \mathbf{B}_T[-1 \quad 1]^T &= -B_{T,1}(\widehat{\Theta}_T) + B_{T,2}(\widehat{\Theta}_T) \\ \mathbf{B}_T[-1 \;\, -1]^T &= -B_{T,1}(\widehat{\Theta}_T) - B_{T,2}(\widehat{\Theta}_T) \end{aligned} \tag{26}$$

This set may be produced using ESPAR antennas in the following way: we feed the active element with the symbol s_0, and use the exhaustive search algorithm to calculate the varactors' values for the following radiation patterns

$$\begin{aligned} P_{(1)}(\widehat{\Theta}_T) &= B_{T,1}(\widehat{\Theta}_T) + B_{T,2}(\widehat{\Theta}_T) \\ P_{(-1)}(\widehat{\Theta}_T) &= B_{T,1}(\widehat{\Theta}_T) - B_{T,2}(\widehat{\Theta}_T) \end{aligned} \tag{27}$$

Therefore, a compact representation of the pattern production is

$$\mathbf{B}_T\mathbf{x}_{bs} = P_{(\frac{s_0}{s_\pi})}(\widehat{\Theta}_T)s_\pi \tag{28}$$

Our approach has shown that these combinations are more accurately achieved by ESPAR antennas with inter-element spacing of $\lambda/8$ or $\lambda/16$. The amplitude and phase patterns of the more compact ESPAR antenna with inter-element spacing equal to $\lambda/16$ are drawn in Figure 5 for $s_0/s_\pi = 1$ and $s_0/s_\pi = -1$. We can extend the aforementioned approach to other linear modulation schemes. This is feasible if a set of reactive loads exists such that the ESPAR antenna can create the set of products $\mathbf{B}_T\mathbf{x}_{bs}$ that correspond to the pre-defined modulation scheme. For example, by applying the exhaustive search approach to the ESPAR antennas of Figure 4, we found out that both BPSK and QPSK modulation formats can be supported, to simultaneously transmit two diverse symbols towards angles 0 and π. The corresponding radiation patterns for QPSK modulation when $s_0/s_\pi = 1$ and $s_0/s_\pi = -1$ for the 5-element array with $d = \lambda/16$ are shown in Figure 5, whereas the patterns for $s_0/s_\pi = i$ and $s_0/s_\pi = -i$ are those given in Figure 6.

IV. Performance Evaluation and Simulation Results

We have shown that by using a single active antenna element and compact parasitic arrays, we can transmit two

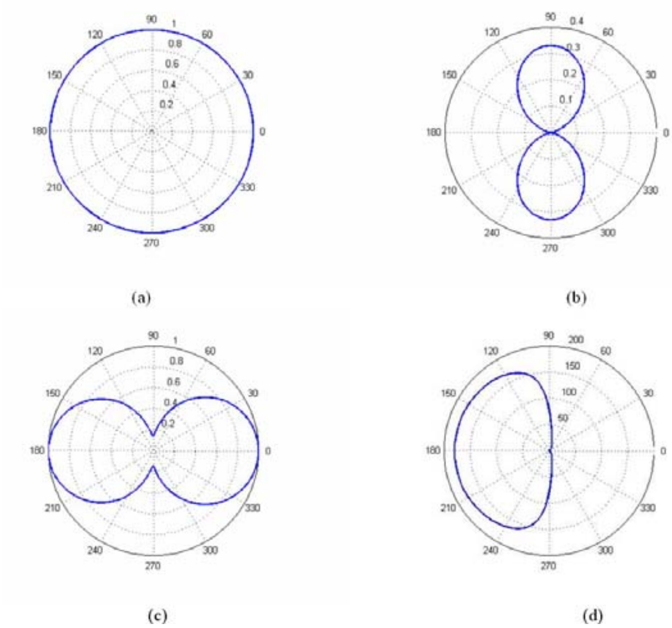


Fig. 5. Amplitude (a) and phase (degrees) (b) patterns for 3-element ESPAR with $d = \lambda/16$, and $s_0/s_\pi = 1$. Amplitude (c) and phase (degrees) (d) patterns for 3-element ESPAR with $d = \lambda/16$, and $s_0/s_\pi = -1$.

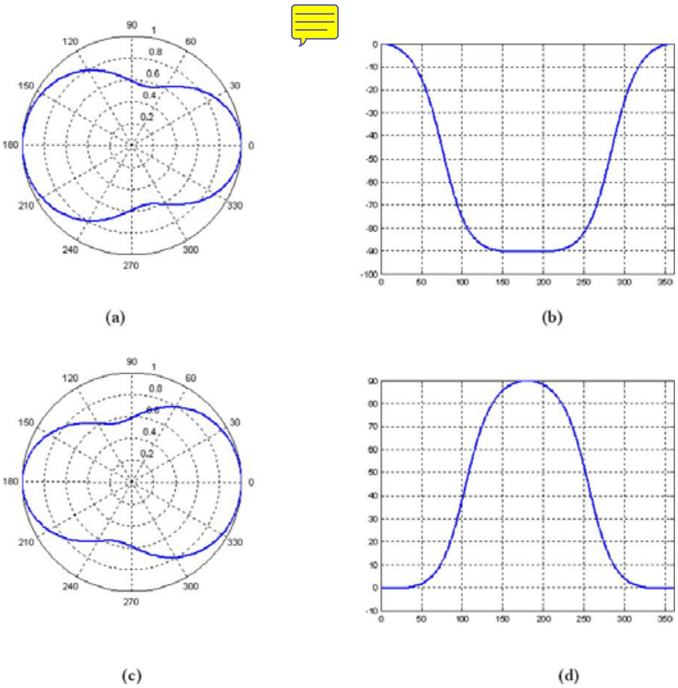


Fig. 6. Amplitude (a) and phase (degrees) (b) patterns for 5-element ESPAR with $d = \lambda/16$, and $s_0/s_\pi = i$. Amplitude (c) and phase (degrees) (d) patterns for 5-element ESPAR with $d = \lambda/16$, $s_0/s_\pi = -i$.

diverse symbols over disjoint directions in the wavevector domain when OOK, BPSK or QPSK modulations are used. In this section, BS-MIMO and BS-MISO system examples will be evaluated in terms of bit error rate, using either switched parasitic arrays or ESPAR antennas. Furthermore, the results will be compared to equivalent traditional MIMO or MISO architectures against the same performance metric. In all MIMO examples (both BS and traditional), a conventional 2-element ULA is used at the receiver to sample the transmitted waveforms and 2x2 linear MMSE detection is employed.

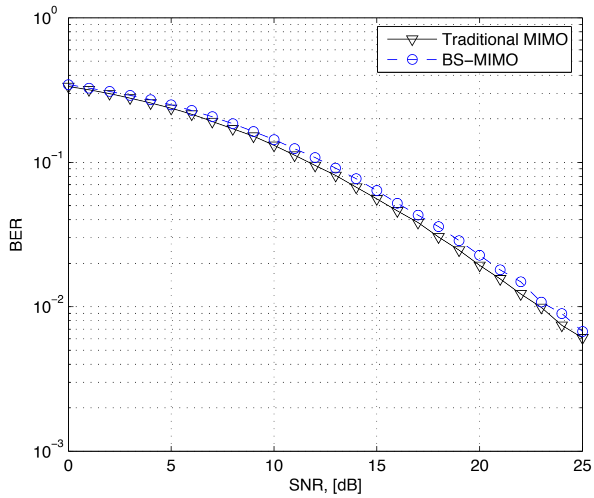


Fig. 7. BER performance of a 2 by 2 beamspace MIMO system for OOK modulation and a 3-element SPA

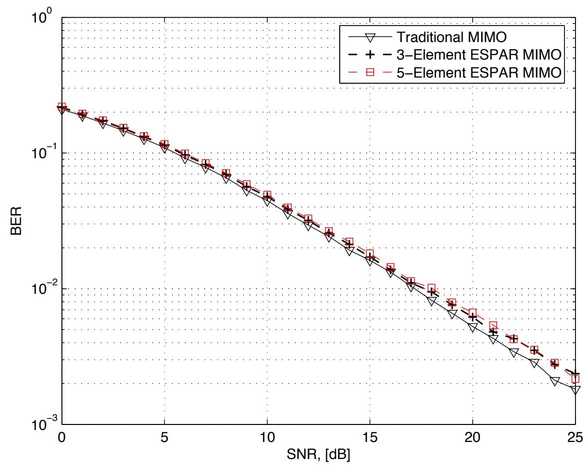


Fig. 8. BER performance of a 2 by 2 beamspace MIMO system for BPSK modulation and a 3 or 5-element ESPAR antenna

The channel used for modeling the system is that described in Section II. The virtual channel angles are defined to be towards the maxima of the basis functions used, namely at directions 0 and π. To ensure fair comparisons, we have set the total radiated power in the beamspace MIMO examples to be equal to the total radiated power of the transmitted symbols in the corresponding traditional MIMO cases for all modulation formats. The results are presented in Figures 7 to 9. Figure 7 depicts the BER vs. SNR curve for the SPA antenna of Figure 3 and OOK modulation, Figure 8 presents the corresponding curves for the 3-element ESPAR antenna shown in Figure 4a and the 5-element ESPAR antenna shown in Figure 4b for BPSK modulation, whereas Figure 9 contains the corresponding curves for QPSK modulation. From the figures we can see that, at the cost of a small degradation in bit error rate, it is indeed possible to transmit at MIMO rates with the use of a single active and several parasitic antenna elements. This finding validates our approach and opens the way for the implementation of MIMO transmitters with a single RF front end.

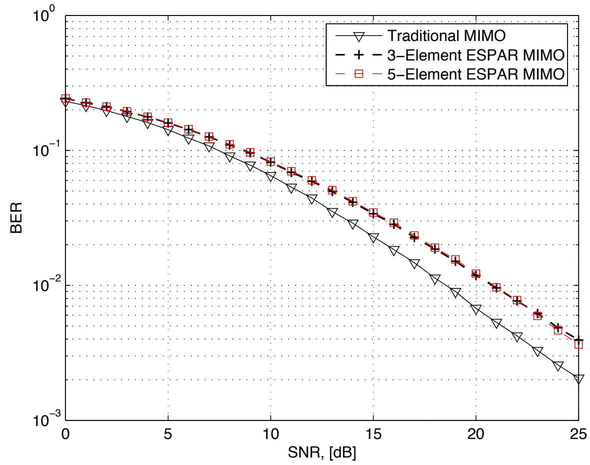


Fig. 9. BER performance of a 2 by 2 beamspace MIMO system for QPSK modulation and 3 or 5-element ESPAR antennas

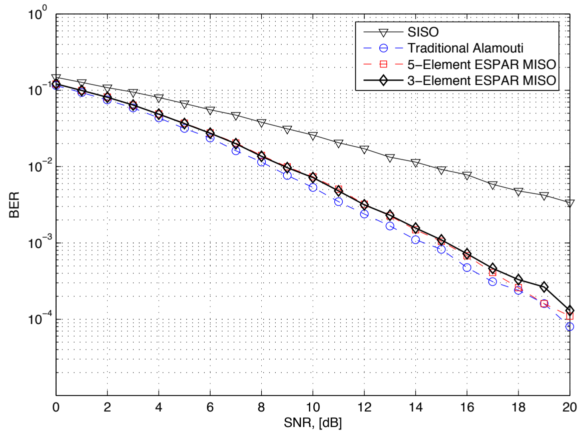


Fig. 10. BER performance of the 3 and 5-element ESPAR proposed system (BS-Alamouti) for BPSK modulation

In the BS-MISO case the proposed antenna array transmission schemes are used at the transmitter, whereas a single omni-directional antenna is assumed at the receiver. The performance was evaluated again in terms of bit error rate, using the same channel model, for the cases of BPSK and QPSK modulation formats and ESPAR antennas. These curves are compared to the corresponding performance of traditional Alamouti MISO systems using the same modulation schemes. The results are presented in Figures 10 and 11. As can be seen from the figures, the Beam-Space approach yields results that are very close to that of traditional MISO transmit diversity, despite the availability of a single active antenna at the transmitter side. The fact that in the MISO case the performance difference between BS and traditional MISO is even smaller than in the MIMO case is related to the corresponding degree of closeness between the corresponding CDFs shown in Figures 1 and 2. Our results suggest that the implementation of transmit diversity or spatial multiplexing on the side of wireless terminals (e.g. for high-speed data uplink transmission) via a combination of ESPAR antennas and BS modulation is an attractive option for future wireless systems

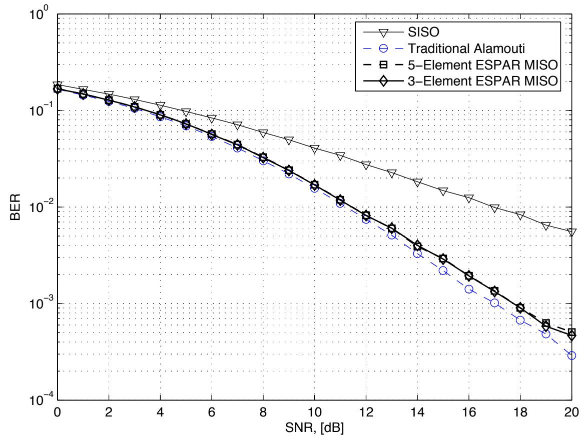


Fig. 11. BER performance of the 3 and 5-element ESPAR proposed system (BS-Alamouti) for BPSK modulation

V. Conclusion

In this paper, we have shown that it is possible to achieve spatial multiplexing gains using wireless transmitters that are equipped with a single RF front end and an array of parasitic antenna elements that are spaced as closely as $\lambda/16$ from each other. In our transmission technique, information symbols are mapped onto the radiation patterns of parasitic antenna arrays, by modulating orthogonal functions directly in the wave-vector domain of the antennas; this results in diverse symbol streams being simultaneously transmitted towards different angles of departure. In rich scattering environments, these symbol streams experience multipath fading in a manner similar to conventional MIMO transmission. Using a traditional MIMO receiver it is then feasible to retrieve the transmitted information at a rate that is equal to that of comparable conventional MIMO systems. Overall, we suggest that, unlike conventional wisdom, spatial multiplexing is indeed possible with wireless transmitters that are equipped with a single RF front end and have critical size and power constraints. We believe that this newly shown capability can help MIMO transmission technology make its way into small wireless devices, thus enabling and benefiting a large variety of future wireless networking paradigms and applications.

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Antonis Kalis is an associate professor of AIT and an adjunct professor of Carnegie Mellon University. He received his Electrical Engineering Diploma degree in 1997 from the EE Department of the University of Patras, Greece, and joined the Lab of Electromagnetics at the University of Patras, participating in various R&D projects for the Greek Government and the European Union, as research staff. In 2000 he worked as a Research Engineer and an Assistant Research Unit Manager at the Computer Technology Institute. He received his PhD with honors from the University of Patras, in June 2002. His research interests are in the areas of radio communications, antenna design and wireless networks. He has numerous journal and conference publications, a US patent, and the 2000 Chester Sall award of the Consumer Electronics Society in the above areas. Antonis Kalis is a member of the Technical Chamber of Greece, and the IEEE.

Athanasios G. Kanatas received the Diploma in Electrical Engineering from the National Technical University of Athens, Greece, in 1991, the M.Sc. degree in Satellite Communication Engineering from the University of Surrey, Surrey, UK in 1992, and the Ph.D. degree in Mobile Satellite Communications from the National Technical University of Athens, Greece in February 1997. From 1993 to 1994 he was with National Documentation Center of National Research Institute. In 1995 he joined SPACETEC Ltd. where he was Technical Project Manager for VISA/EMEA VSAT Project in Greece. In 1996 he joined the Mobile Radio-Communications Laboratory as a research associate. From 1999 to 2002 he was with the Institute of Communication & Computer Systems. In 2000 he became a member of the Board of Directors of OTESAT S.A. In 2002 he joined the University of Piraeus where he is an Associate Professor in the Department of Technology Education and Digital Systems. His current research interests include channel characterization, estimation, simulation and modelling for mobile, mobile satellite, and future wireless communication systems, antenna selection techniques and new transmission schemes for MIMO systems, and energy efficient techniques for Wireless Sensor Networks. He has been a Senior Member of IEEE since 2002, and is also a member of the Technical Chamber of Greece. In 1999 he was elected Chairman of the Communications Society of the Greek Section of IEEE.

Constantinos B. Papadias (S88, M96, SM03) was born in Athens, Greece, in 1969. He received the Diploma of Electrical Engineering from the National Technical University of Athens (NTUA) in 1991 and the Doctorate degree in Signal Processing (highest honors) from the Ecole Nationale Supérieure des Télécommunications (ENST), Paris, France, in 1995. From 1992 to 1995, he was Teaching and Research Assistant at the Mobile Communications Department, Eurécom, France. In 1995, he joined the Information Systems Laboratory, Stanford University, Stanford, CA, as Post-Doctoral Researcher, working in the Smart Antennas Research Group. In November 1997 he joined the Wireless Research Laboratory of Bell Labs, Lucent Technologies, Holmdel, NJ, as Member of Technical Staff and was later promoted to Technical Manager. From 2004 to 2005 he was an adjunct Associate Professor at Columbia University. In 2006 he joined Athens Information Technology (AIT) in Athens Greece, as an Associate Professor and was later promoted to Professor. He is also currently an Adjunct Professor at Carnegie Mellon University’s Information Networking Institute (INI). His research interests range from baseband wireless communications and smart antennas to scheduling and system-level optimization of wireless systems to cognitive radio and multihop wireless sensor networks. He has authored over 100 papers and book chapters on these topics. His distinctions include the 2002 Bell Labs President’s Award, the 2003 IEEE Signal Processing Society’s Young Author Best Paper Award and ESI’s “most cited paper of the decade” citation in the area of wireless networks. He has also made standards contributions (most notably as the co-inventor of the Space-Time Spreading (STS) technique that was adopted by the cdma2000 wireless standard for voice transmission) and holds 9 patents.

He has participated in several research projects funded by the European Commission and DARPA and has served on the steering board of the Wireless World Research Forum (WWRF) from 2002-2006, acting as its R&D program manager. He is involved in the organization of several technical conferences and journal special issues and is the co-editor of a recently appeared book on MIMO systems. He has served as Associate Editor for the IEEE Transactions on Signal Processing and is currently an Editor of the Journal of Communications and Networks. He is the industrial liaison of the Signal Processing for Communications Technical Committee of the IEEE. Most recently, he was named a National Representative / Expert of Greece in the FP7 program IDEAS. Dr. Papadias is a Senior Member of the IEEE and a member of the Technical Chamber of Greece.