**QUEEN'S
UNIVERSITY
BELFAST**

Pre-Matched Eigenmode Antenna With Polarization and Pattern Diversity

Wolosinski, G., Fusco, V., Naeem, U., & Rulikowski, P. (2019). Pre-Matched Eigenmode Antenna With Polarization and Pattern Diversity. *IEEE Transactions on Antennas and Propagation*.
<https://doi.org/10.1109/TAP.2019.2912464>

Published in:
IEEE Transactions on Antennas and Propagation

Document Version:
Peer reviewed version

Queen's University Belfast - Research Portal:
[Link to publication record in Queen's University Belfast Research Portal](#)

Publisher rights
© 2019 IEEE. This work is made available online in accordance with the publisher's policies. Please refer to any applicable terms of use of the publisher.

General rights
Copyright for the publications made accessible via the Queen's University Belfast Research Portal is retained by the author(s) and / or other copyright owners and it is a condition of accessing these publications that users recognise and abide by the legal requirements associated with these rights.

Take down policy
The Research Portal is Queen's institutional repository that provides access to Queen's research output. Every effort has been made to ensure that content in the Research Portal does not infringe any person's rights, or applicable UK laws. If you discover content in the Research Portal that you believe breaches copyright or violates any law, please contact openaccess@qub.ac.uk.

Pre-Matched Eigenmode Antenna With Polarization and Pattern Diversity

Grzegorz Wolosinski, *Student Member, IEEE*, Vincent Fusco, *Fellow, IEEE*,
Umair Naeem, and Pawel Rulikowski.

Abstract— In this paper we propose an antenna design procedure based on the eigenmode decomposition technique which avoids the use of the external matching networks that are typically required at the input ports of the mode decomposition network (MDN). We illustrate the technique by examining the properties of a circularly symmetrical 4-port antenna configuration which can achieve four orthogonal radiation patterns or eigenpatterns and that allows both polarization and pattern diversity. We then use the proposed theory to design an actual antenna structure with eigenmodes operating around 2.6 GHz and whose eigenpatterns are directed in one hemisphere. The antenna and an example MDN are manufactured and tested showing that the four eigenmodes can simultaneously operate in 9.5 % of bandwidth (RL > 10 dB) with a worst case total radiation efficiency of 72 %, correlation coefficient of 0.04, and port coupling of –24 dB. Finally, we use the measured results to assess the antenna in a Multiple-Input-Multiple-Output (MIMO) communication system and demonstrate its beam-tilting capability.

Index Terms— Eigenmode Antenna, MIMO, Pattern Diversity, Polarization Diversity, Port Decoupling.

I. INTRODUCTION

MULTIPLE-Input-Multiple-Output (MIMO) technology has received massive interest in wireless communications due to the potential of improving spectral efficiency by exploiting multi-element transmit and receive antennas [1]. In order to achieve the promised MIMO performance, the multi-element antennas (or antenna arrays) need to operate in a complex multipath environment and feature some kind of *antenna diversity* to achieve high antenna port-to-port isolation and uncorrelated radiation patterns. These features can be accomplished, to a degree, with antennas featuring *space diversity*, i.e. spatially separated antennas with the same radiation pattern [1]. When the size of the antenna system becomes constrained, radiating elements with orthogonal polarization states or very different radiation patterns can be exploited leading to *polarization diversity* [2] and *pattern diversity* [3], respectively. In general, a combination of antenna diversity schemes can be used [4]. When the dimensions of the antenna system are further constrained, decoupling techniques

are usually employed to improve the MIMO performance. A robust approach to achieve low port coupling and orthogonal radiation patterns is the *eigenmode decomposition technique* [5]. This technique allows the designer to diagonalize the scattering matrix of a multi-element antenna with circular or linear symmetry by cascading a mode decomposition network (MDN) with the antenna ports. As a result, the input ports of the MDN correspond to orthogonal modes or *eigenmodes* which feature *full port decoupling* and orthogonal radiation patterns or *eigenpatterns* allowing to achieve full antenna diversity. To date, this technique has been employed mainly to decouple closely spaced monopole antennas arranged on a common ground plane. Thereby allowing pattern diversity along the main plane of the array. The implementation of this technique has required matching networks at the MDN input ports leading to limited operational bandwidth, [5]–[11]. More recently, in order to improve the bandwidth of such antenna systems, switchable and tuneable MDN as well as matching networks have been proposed in [12]–[14]. In [15] the concept of mode matching network was introduced in order to implement the matching stage before the interconnection to the MDN.

In contrast to the above literature, in [16] we exploited the eigenmode decomposition technique to achieve polarization diversity without the use of matching networks allowing reduction implementation complexity and enabling greater operating bandwidth. To generalize this work, in [17] we have proposed an antenna based on a new eigenmode design procedure which can achieve both polarization and pattern diversity without the use of matching networks. In this paper we comprehensively validate the antenna concept sketched out briefly in [17] by providing a theoretical background, analysis, and measurements. The remainder of this paper is organized as follows: Section II outlines the theoretical background and proposes an antenna design procedure based on the eigenmode decoupling technique which avoids the need of matching networks. Section III proposes a 4-port antenna and an example MDN that are able to achieve both polarization and pattern diversity in one hemisphere. Section IV reports the simulations

“This work was supported in part by Marie Curie European Industrial Doctorate (EID) programme ARTISAN (grant no. 316426) and Nokia Bell Labs, Ireland.”

G. Wolosinski, V. F. Fusco, and U. Naeem are with the Institute of Electronics,

Communications and Information Technology (ECIT), Queens University of Belfast, Belfast, BT3 9DT, U.K. (e-mail: gwolosinski01@qub.ac.uk, v.fusco@ecit.qub.ac.uk, u.naeem@qub.ac.uk). P. Rulikowski is with the RF Access Group, Nokia Bell Labs, Dublin, Ireland (e-mail: pawel.rulikowski@nokia-bell-labs.com).

of the antenna structure and MDN while Section V reports the measured results. Finally, Section VI investigates the proposed antenna in terms of MIMO performance and beam-tilting capability.

II. EIGENMODE DECOMPOSITION TECHNIQUE

A. Theoretical Background

Consider an N -port antenna array composed of reciprocal elements which features a symmetric scattering matrix $\mathbf{S}$, i.e. $\mathbf{S} = \mathbf{S}^T$. The diagonalization of the matrix $\mathbf{S}$ can be achieved using the following mathematical condition [5]

$$\mathbf{\Gamma} = \mathbf{PSP}^T \quad (1)$$

where $\mathbf{P}$ is a unitary matrix containing the eigenvectors of $\mathbf{S}$ and $\mathbf{\Gamma}$ is a diagonal matrix containing the eigenvalues of $\mathbf{S}$. In order to implement the diagonalization, the antenna must be connected to a $2N$ -port reciprocal network with N input and N output ports as shown in Fig. 1 (a). The scattering parameters of such a network can be expressed in block matrix notation form as

$$\mathbf{S}_D = \begin{pmatrix} \mathbf{S}_{D,11} & \mathbf{S}_{D,12} \\ \mathbf{S}_{D,12}^T & \mathbf{S}_{D,22} \end{pmatrix} \quad (2)$$

where the subscript 1 and 2 correspond to the input and output ports, which are respectively on the left and right side of the $\mathbf{S}_D$ block in Fig. 1 (a). The antenna array when cascaded with the network creates a new antenna system with the following scattering parameters

$$\mathbf{S}_S = \mathbf{S}_{D,11} + \mathbf{S}_{D,12} \mathbf{S} (\mathbf{I} - \mathbf{S}_{D,22} \mathbf{S})^{-1} \mathbf{S}_{D,12}^T \quad (3)$$

where $\mathbf{I}$ is the identity matrix. Equation (3) reduces to

$$\mathbf{S}_S = \mathbf{S}_{D,12} \mathbf{S} \mathbf{S}_{D,12}^T \quad (4)$$

for a network with matched and decoupled input and output ports. By comparing (1) and (4) we note that if $\mathbf{S}_{D,12} = \mathbf{P}$ we achieve $\mathbf{S}_S = \mathbf{\Gamma}$, thus the network is able to suppress the mutual coupling of the antenna array. The reciprocity of $\mathbf{S}_D$ implies that $\mathbf{P} = \mathbf{P}^T$ must be satisfied for a physically realizable MDN. This, in turn, implies that the antenna array needs to feature a symmetric physical arrangement [5]. The MDN achieves port decoupling by decomposing any arbitrary input excitation into N eigenmodes at the output ports. For example, by feeding the n th input port of the MDN, the antenna ports are excited according to the n th eigenvector. Therefore, each input port corresponds to an eigenmode with an input reflection coefficient according to (4) which takes the name of *modal reflection coefficient*. Moreover, the eigenmodes correspond to far-field *eigenpatterns* which are mutually orthogonal when considered within an *isotropic power angular spectrum (PAS)* [9]. In this case, the spatial correlation coefficient between two arbitrary far-field patterns is defined as [18]

Fig. 1. (a) N -port antenna array cascaded with a $2N$ -port network. (b) 4-port antenna array cascaded with a decomposition network implemented by 180° hybrid couplers.

$$\rho_{mn} = \frac{\int E_{\theta,m}(\Omega) E_{\theta,n}^*(\Omega) + E_{\phi,m}(\Omega) E_{\phi,n}^*(\Omega) d\Omega}{\sqrt{\int |E_{\theta,m}(\Omega)|^2 + |E_{\phi,m}(\Omega)|^2 d\Omega} \sqrt{\int |E_{\theta,n}(\Omega)|^2 + |E_{\phi,n}(\Omega)|^2 d\Omega}} \quad (5)$$

where $E_{\theta,\phi}$ with subscript m and n is the complex electric far-field distribution along the θ and ϕ component, respectively for the m th and n th far-field pattern. The correlation coefficient between any pair of eigenpatterns with $n \neq m$ equals to zero, therefore enabling full antenna diversity. It is worth noting that $\rho_{mn} = 0$ is a general result that holds true regardless of the type of antenna diversity achieved between the eigenpatterns.

B. Eigenmode Antenna Design

We now use the eigenmode decoupling technique as the basis upon which to propose an antenna design methodology whereby we can avoid the use of matching networks and thus reduce implementation complexity. The proposed design procedure is detailed in the following steps:

- 1) Choose an antenna configuration and identify the set of eigenvectors that allow to diagonalize the scattering matrix by means of a feasible MDN, i.e. $\mathbf{S}_{D,12} = \mathbf{P}$.
- 2) Obtain the closed-form expression of the modal reflection coefficients through (4). If a-priori characterization of a real-world MDN is known, the antenna design procedure can be carried out by considering (3), instead of (4).
- 3) Design the antenna array in such a way to optimize the modal reflection coefficients according to designer specified operational bandwidth criteria.
- 4) If not all of the modal reflection coefficients can be simultaneously optimized, carry out the optimization by disregarding the eigenmode with low radiation efficiency. Reiterate step 3.

The above procedure can be seen as designing the antenna array in the eigenmode domain since the goal is to optimize the eigenvalues of the scattering matrix. Unlike the conventional antenna design procedure which involves optimization of the input reflection coefficients, the design procedure here proposed requires careful optimization of both antenna input reflection coefficients and mutual coupling. Although this results in a more difficult synthesis task for the designer, the interconnection of the antenna array ports with the MDN have the benefit that no further input matching is required at the MDN input ports.

III. ANTENNA DESIGN FOR POLARIZATION AND PATTERN DIVERSITY

In the following we employ the above proposed design procedure to achieve a compact four-element antenna capable of pattern and polarization diversity.

A. Antenna Arrangement

We consider an array of equal elements with rotational symmetry of 90° along its physical center. This leads to a circulant scattering matrix

$$\mathbf{S} = \begin{pmatrix} S_{11} & S_{12} & S_{13} & S_{12} \\ S_{12} & S_{11} & S_{12} & S_{13} \\ S_{13} & S_{12} & S_{11} & S_{12} \\ S_{12} & S_{13} & S_{12} & S_{11} \end{pmatrix} \quad (6)$$

The diagonalization of (6) through (1) leads to

$$\mathbf{P} = \begin{pmatrix} 1 & 1 & 1 & 1 \\ 1 & 1 & -1 & -1 \\ 1 & -1 & 1 & -1 \\ 1 & -1 & -1 & 1 \end{pmatrix} \quad (7)$$

which allows implementation of the MDN with four cross connected 180° hybrid couplers as shown in Fig. 1 (b). As a result, the input ports of the MDN have the modal reflection coefficients

$$\mathbf{S}_S = [S_{11} + 2S_{12} + S_{13}, S_{11} - S_{13}, S_{11} - 2S_{12} + S_{13}, S_{11} - S_{13}] \quad (8)$$

which result from a linear combination of the antenna reflection coefficient, i.e. S_{11} , and mutual coupling terms, i.e. S_{12} and S_{13} . Therefore, the modal reflection coefficients in (8) can be optimized for different frequencies and matching responses by exploiting the mutual coupling terms. In the extreme case of zero coupling the modal reflection coefficients become equal to S_{11} . In the following we optimize (8) to achieve four eigenmodes with modal reflection coefficients better than -10 dB operating within a common bandwidth as wide as possible. In this case we expect the magnitude of S_{12} , and S_{13} to be relatively small to allow common bandwidth operation. The schematic arrangement of the considered array and the main polarization state of each radiating element is shown in Fig. 2 (a). The array factor (AF) of the m th eigenmode is

Fig. 2. (a) 4-element circular antenna array configuration with elements featuring main polarization as indicated by red arrows. Magnitude of (b) eigenpattern 1 and 3 and (c) eigenpattern 2 and 4.

$$AF_m = \sum_{n=1}^N a_m^{(n)} \exp[jka \sin \theta \cos(\varphi - \varphi_n)] \quad (9)$$

where $a_m^{(n)}$ is the complex modal excitation of the m th eigenmode at the n th antenna element corresponding to the m th row or column of (7), k is the wave number, a is the radius of the array, and φ_n is the position angle of the n th element. In order to predict the achievable eigenpatterns of this configuration, we consider the antenna elements as isotropic sources. By using (9) we can write the m th eigenpattern as

$$E_m = \sum_{n=1}^N E^{(n)} a_m^{(n)} \exp[jka \sin \theta \cos(\varphi - \varphi_n)] \quad (10)$$

where $E^{(n)} = \exp(j\varphi_n)$ is the radiation pattern of the n th element which takes into account its relative polarization orientation with respect to the other elements. As an example, the magnitude of the eigenpatterns for $a = 0.25 \lambda$ is reported in Fig. 2 (b)-(c). We note that E_1 and E_3 (E_2 and E_4) have the same magnitude pattern with minima (maxima) at $\theta = 0^\circ, 180^\circ$ and maxima (minima) at $\theta = 90^\circ$. The complementary shape of these eigenpatterns suggests that full pattern diversity is achieved. The phase of E_1 and E_3 (E_2 and E_4), not shown, are orthogonal via (5) suggesting that a different type of diversity between the eigenpatterns with same magnitude is achieved, i.e. polarization diversity. In the following we use the above design guidelines to design an actual antenna whose eigenpatterns lie in one hemisphere ($0^\circ \leq \theta \leq 90^\circ$ and $0^\circ \leq \varphi \leq 360^\circ$) since this feature is desired in many communication applications.

Fig. 3. (a) Isometric and (b) side view of the four-element antenna. (c) Details of the radiating structure. (d) Details of one feeding section. $L=150$, $H=27.3$, $H_s=1.6$, $L_r=41.2$, $d_f=24.6$, $L_p=29$, $R_1=15$, $R_2=10.7$, $d=5.7$, $d_p=2.1$, $h_f=0.3$. Units are in mm.

B. Antenna Structure

The procedure detailed in the previous subsection has been employed to design a symmetrical four-element antenna as shown in Fig. 3 (a). The antenna consists of four main radiators and a parasitic element etched on a double-sided FR-4 substrate, four coaxial cables, and a planar reflector. The main radiators are etched on the top side of the substrate and are fed by means of the four coaxial cables as detailed in Fig. 3 (b)-(c). Fig. 3 (d) shows the details of one feeding section. Each feed cable connects two consecutive radiators in a dipole-like arrangement, however, the circular symmetry of the antenna implies that each radiator is attached to two coaxial cables. This arrangement allows us to optimize the position of the feeds and therefore provides high flexibility in the optimization of the antenna electrical performance as required by (8). The parasitic element is positioned in the central part of the antenna close to the feeding points allowing to introduce additional degrees of freedom in the optimization process of the antenna. This is etched on the bottom side of the substrate to avoid intersections with the feed lines. The substrate is backed by the reflector to obtain radiation directed mainly along one hemisphere. Both the substrate and reflector are square in shape and of the same size and are spaced by means of four low permittivity dielectric supports placed at the corners of the structure. The four coaxial cables pass through the reflector and are grounded to it. The

Fig. 4. Isometric view of the designed MDN on a substrate with dielectric constant of 3.5 and dissipation factor of 0.0018. $L_g=150$, $h=1.52$, $d=31.6$, $t_1=2.19$, $t_2=3.45$. Units are in mm.

caption of Fig. 3 reports the dimensions of the proposed antenna.

C. MDN

To allow validation of the proposed antenna concept, an example MDN has been designed as shown in Fig 4. The MDN consists of four interconnected 180° hybrid couplers implemented on a single layer PCB in order to be physically attached and interconnected at the ports on the back of the antenna. The employed substrate is a Taconic RF-35 with dielectric constant of 3.5, dissipation factor of 0.0018, and thickness of 1.52 mm. The area of the substrate is chosen to be equal to the footprint of the antenna to allow practical interconnection, i.e. $150 \times 150 \text{ mm}^2$. We can note the symmetry of the network between input (output) port 1, 2, 3, and 4 with respect to the output (input) port 5, 6, 7, and 8, which is expected due to the reciprocity of the network.

IV. SIMULATED RESULTS

This section reports the results of the antenna and MDN obtained with full-wave electromagnetic (EM) simulations, wherein the metallic parts are modeled as perfect conductor while the dielectric materials are modeled with their relative permittivity.

A. Antenna Structure

The S-parameters and eigenvalues of the antenna are reported in Fig. 5. The antenna features four ports with identical input reflection coefficient and fractional bandwidth of 21% around 2.55 GHz and mutual coupling lower than -19 dB . After interconnection of the antenna ports to the MDN, the resulting four modal reflection coefficients feature operations around 2.6 GHz with different bandwidths. In fact, eigenmode 1 and 3 show a fractional bandwidth respectively of 12% and 14%, while that of eigenmode 2 and 4 is 27%. As for the common operation, the four eigenmodes can simultaneously operate in 12% of bandwidth centred at 2.6 GHz. It is worth noting that even though the magnitude of the mutual coupling terms is fairly small, its impact on the modal reflection coefficients is quite significant. The simulated 3D eigenpatterns are reported in Fig. 6. According to the design considerations, the four

Fig. 5. Simulated (a) S-parameters and (b) eigenvalues of the antenna.

Fig. 6. 3D simulated realized gain at 2.6 GHz of eigenpattern (a) 1, (b) 2, (c) 3, and (d) 4. The arrows indicate the dominant polarization state.

Fig. 7. Simulated (a) total radiation efficiency of the eigenmodes and (b) correlation coefficients between the eigenmodes.

eigenpattern feature radiation mainly into one hemisphere. We can note that eigenpattern 2 and 4 feature a maximum at $\theta = 0^\circ$ which correspond to the radiation from two orthogonal electric

Fig. 8. Simulated (a) input reflection coefficients and (b) port coupling at the input (output) port of the MDN.

dipoles and can therefore achieve polarization diversity as their electric field vectors are rotated by 90° with respect to each other. These eigenpatterns correspond to the radiation modes of a conventional dual linear-polarized antenna. Instead, eigenpattern 1 and 3 feature a null at $\theta = 0^\circ$ and a variable linear polarization state with spatial direction corresponding to the radiation from a magnetic dipole and an electric quadrupole, respectively. Although these eigenpatterns are orthogonal along the full 3D space via (5), their electric field vectors are not always locally orthogonal. Indeed, for the maxima occurring along $\varphi = 0^\circ$ and 90° the electric field vectors are respectively in phase and anti-phase, not allowing polarization diversity. Instead, along $\varphi = 45^\circ$ and 135° the electric field vectors of eigenpattern 1 and 3 are rotated by 90° , allowing polarization diversity. The complementary shape of the eigenpatterns 1 and 3 with respect to eigenpatterns 2 and 4 lead to pattern diversity. The simulated total radiation efficiency of the eigenpatterns and the correlation coefficient between each pair of eigenpatterns considering a rich isotropic multipath environment are reported in Fig. 7. The total radiation efficiency within the common operating bandwidth is better than 85% while the correlation coefficient computed with (5) is better than 0.006, confirming the effective antenna operation without the employment of matching networks and the orthogonality between the four eigenmodes.

B. MDN

The simulated S-parameters at the input (and output) side of the MDN are reported in Fig. 8. The MDN ports are well matched at 2.6 GHz with a reflection coefficient better than -21 dB and port coupling as low as -32 dB. Considering the same simulated bandwidth of the antenna, 12% around 2.6 GHz, the input reflection coefficients and port coupling are better than -6.6 dB and -24 dB, respectively. As far as the transmission coefficients are concerned, not reported, the worst value for the magnitude is -7.7 dB (against the ideal -6 dB) while the maximum phase error between the output ports is 8.6° , within the considered bandwidth.

Fig. 9. (a) Isometric, (b) top, and (c) back view of prototype antenna and MDN interconnected to each other by SMA connectors.

Fig. 10. Measured S-parameters of the MDN cascaded to the antenna ports.

Fig. 11. Measured (a) total radiation efficiency of the antenna system and (b) correlation coefficients between the four eigenmodes.

TABLE I
MEASURED PERFORMANCE OF THE EIGENMODES

Operating eigenmode	Frequency band (GHz)	BW (%)	Total radiation efficiency (%)
Eigenmode 1	2.41 – 2.66	9.8%	72%
Eigenmode 2	2.37 – 2.68	12.3%	77%
Eigenmode 3	2.4 – 2.69	11.4%	73%
Eigenmode 4	2.38 – 2.65	10.7%	77%
Common band	2.41 – 2.65	9.5%	72%

V. MEASUREMENT RESULTS

The antenna and MDN have been manufactured and

Fig. 12. Measured eigenpatterns at 2.41 GHz. From top to bottom $\varphi = 0^\circ$ and $\varphi = 45^\circ$. From left to right the reference electric field component is E_θ and E_φ .

Fig. 13. Measured eigenpatterns at 2.53 GHz. From top to bottom $\varphi = 0^\circ$ and $\varphi = 45^\circ$. From left to right the reference electric field component is E_θ and E_φ .

interconnected with SMA connectors as shown in Fig. 9. The S-parameters of the antenna system are reported in Fig. 10. The central operating frequency is slightly shifted towards lower frequencies with respect to simulations. For a RL > 10 dB, the common operation of the four eigenmodes is 2.41–2.65 GHz which results in 9.5% of bandwidth and central operating frequency of 2.53 GHz. The bandwidth achieved by each eigenmode is reported in Table I. The port coupling within the common bandwidth is better than –24 dB. The antenna system was measured in a reverberation chamber in order to obtain total radiation efficiency and correlation coefficients as reported in

Fig. 14. Measured eigenpatterns at 2.65 GHz. From top to bottom $\varphi = 0^\circ$ and $\varphi = 45^\circ$. From left to right the reference electric field component is E_θ and E_φ .

Fig. 11. The standard deviation uncertainty of the measurement setup is better than 0.5 dB (12% in linear scale) within the measured bandwidth, [19]. The total radiation efficiency within the common bandwidth is better than 72 % for all the eigenmodes while the correlation coefficient is lower than 0.04. The lowest recorded total radiation efficiency of each eigenmode within the bandwidth is reported in Table I. The eigenpatterns of the antenna have been characterized within a far-field anechoic chamber along $\varphi=0^\circ$ and $\varphi=45^\circ$. The E_θ and E_φ components of the eigenpatterns at 2.41, 2.53, and 2.65 GHz are reported in Fig. 12–14, respectively. The eigenpatterns remain quite stable for the measured frequencies. For $\varphi=0^\circ$, the E_θ and E_φ components correspond to vertical and horizontal polarization, respectively. We notice that for $\varphi=0^\circ$ the pair consisting of eigenpattern 2 and 4 and that of eigenpattern 1 and 3 show very similar radiation patterns (in magnitude) for both polarizations as noted in simulations, see Fig. 6. For vertical polarization, the first pair of eigenmodes show a maximum at $\theta=0^\circ$ while the other pair features a minimum at $\theta=0^\circ$ and maxima at $\theta \approx \pm 40^\circ$ (eigenpattern 1) and $\theta \approx \pm 37^\circ$ (eigenpattern 3) confirming that pattern diversity is achieved. Note that due to the symmetry of the antenna, the same situation occurs at $\varphi=90^\circ$ for horizontal polarization. For $\varphi=45^\circ$, the E_θ and E_φ component correspond to slant $\pm 45^\circ$ polarization. We notice that along this cut the polarization of each pair of eigenpatterns is orthogonal, allowing polarization diversity to be achieved. Again due to the symmetry of the antenna, the same situation occurs for $\varphi=135^\circ$.

VI. ANTENNA CAPABILITIES

A. MIMO Performance

In the following we evaluate the performance of the studied antenna system operating in a 2×2 MIMO communication

system. We consider a narrow band single-user MIMO scenario and choose the spectral efficiency as the performance metric. We consider a rich isotropic multipath environment wherein the transmit end of the link consists of ideal antennas (zero correlation and 100% radiation efficiency) while the receive end is modelled by the studied antenna system, where two ports are simultaneously active at any moment in time while the other ones are terminated with matched loads. This architecture can be practically implemented with a network of switches. The propagation channel $\mathbf{H}$ can be modelled as in [20],

$$\mathbf{H} = \mathbf{C} \mathbf{H}_{\text{iid}} \quad (11)$$

where $\mathbf{H}_{\text{iid}}$ is a 2×2 complex-valued matrix whose entries are independent and identically distributed (iid) random variables with zero mean and unit variance. $\mathbf{C}$ consists of two matrices modelling the power losses and correlation at the receiver

$$\mathbf{C} = \begin{pmatrix} \sqrt{\eta_n} & 0 \\ 0 & \sqrt{\eta_m} \end{pmatrix} \begin{pmatrix} 1 & \rho_{nm} \\ \rho_{nm} & 1 \end{pmatrix} \quad (12)$$

where $\eta_{n,m}$, $n, m = 1, 2, \dots, 4$ is the measured total radiation efficiency of the n th and m th antenna port while ρ_{nm} , $n, m = 1, 2, \dots, 4$ is the measured correlation coefficient between the n th and m th antenna port. The spectral efficiency (SE) of the single-user MIMO system with uniform power control writes

$$SE = \log_2 \left\{ \det \left(\mathbf{I}_{N_r} + \frac{\text{SNR}}{N_t} \mathbf{H} \mathbf{H}^H \right) \right\} \quad (13)$$

where N_t and N_r is the number of transmitting a receiving antennas, respectively, while $\mathbf{I}$ is the identity matrix. Spectral efficiency is computed at each frequency considering 10000 realizations and $\text{SNR} = 10$ dB, which is representative of a realistic value in a wireless communication link, [20]. Fig. 15 reports the spectral efficiency as a function of frequency considering the possible permutations of the ports of the proposed antenna system. As a reference, the spectral efficiency achieved with an ideal 2-port antenna at the receiver is also reported. The results show that regardless of the chosen pair of eigenmodes, the spectral efficiency achieved with the proposed antenna is better than 4.8 bit/s/Hz within the common bandwidth of the four eigenmodes against the ideal case which provides 5.5 bit/s/Hz. This spectral efficiency reduction is mainly due to the power losses of the antenna rather than the correlation coefficients which are very close to zero. We should note that for the chosen system architecture the common bandwidth changes according to the selected pair of eigenmodes and therefore can be greater than 9.5%. Finally, it is interesting to note that, in comparison to a conventional dual-polarized antenna for base station applications, the proposed antenna system is able to provide five additional degrees of freedom (i.e. the additional available port permutations) while maintaining a similar radiating aperture for the same operating frequencies [21]. This makes the proposed antenna a robust solution for spectral efficiency gains in propagation channels

Fig. 15. Spectral efficiency of a 2x2 MIMO system with SNR=10 dB. The transmit antennas are ideal while the receive antennas are modeled with the measured results of the proposed antenna system.

Fig. 16. Measured E_ϕ at $\phi=45^\circ$ obtained for (a) anti-phase and (b) in-phase superposition with equal power between eigenpattern 1 and 4.

with different multipath scenarios.

B. Beam-Tilting

In the following we show how judicious superposition of the eigenpatterns can lead to some beam-tilting capability. Fig. 16 shows the superposition of E_ϕ at $\phi = 45^\circ$ of eigenpattern 1 and 4. The in-phase and anti-phase superposition with equal power ratio result in tilting of the maximum of eigenpattern 4 from $\theta = 0^\circ$ towards $\theta = -30^\circ$ and $\theta = 30^\circ$, respectively. Therefore, the pick gain of eigenpattern 4 can be tilted in any position within $-30^\circ \leq \theta \leq 30^\circ$ by changing the power ratio between eigenmode 1 and eigenmode 4 and setting the relative phase shift between the eigenmodes at 0° or 180° . For symmetry considerations, the same capability can be achieved for the orthogonal polarization and considering eigenmode 1 and 2 at $\phi=135^\circ$. Therefore, proper antenna selection can lead to beam-tilting capability along two orthogonal directions making the antenna an appealing candidate for applications where adaptive beams are required. Finally, it is worth noting that this capability is unattainable with a conventional dual-polarized antenna.

VII. CONCLUSION

In this paper we have proposed a new antenna design procedure based on the eigenmode decomposition technique which allows to avoid the use of matching networks at the MDN input ports. Based on this theory, we have proposed and validated with measurements a 4-port antenna and an example

MDN which allow polarization and pattern diversity in one hemisphere. Measured results have shown that the four antenna eigenmodes feature a worst case total radiation efficiency of 72%, correlation coefficients of 0.04, and port coupling of -24 dB within 9.5% of common bandwidth ($RL > 10$ dB). We have used these results to investigate the performance of the proposed antenna in a 2x2 MIMO communication system showing that, for any chosen pair of eigenmodes, the antenna system can achieve a spectral efficiency of 4.8 bit/s/Hz for SNR=10 dB against the theoretical limit of 5.5 bit/s/Hz. Finally, we have shown that a suitable combination of two eigenmodes can lead to beam-tilting capability along two orthogonal directions. These results show that the proposed antenna can be efficiently exploited for spectral efficiency gains in different propagation channels and for applications that require beam-tilting capabilities from a physical compact footprint.

REFERENCES

- [1] G. J. Foschini and M. J. Gans, "On limits of wireless communications in a fading environment when using multiple antennas," *Wireless Personal Commun.*, vol. 6, pp. 311-335, Mar. 1998.
- [2] H. Huang, Y. Liu and S. Gong, "A novel dual-broadband and dual-polarized antenna for 2G/3G/LTE base stations," *IEEE Trans. Antennas Propag.*, vol. 64, no. 9, pp. 4113-4118, Sept. 2016.
- [3] R. Li, H. Yang, B. Liu, Y. Qin and Y. Cui, "Theory and realization of a pattern reconfigurable antenna based on two dipoles," *IEEE Antennas Wireless Propag. Lett.*, 2018.
- [4] B. Wu and K. M. Luk, "A 4-port diversity antenna with high isolation for mobile communications," *IEEE Trans. Antennas Propag.*, vol. 59, no. 5, pp. 1660-1667, May 2011.
- [5] C. Volmer, J. Weber, R. Stephan, K. Blau and M. A. Hein, "An eigen-analysis of compact antenna arrays and its application to port decoupling," *IEEE Trans. Antennas Propag.*, vol. 56, no. 2, pp. 360-370, Feb. 2008.
- [6] J. C. Coetzee and Y. Yu, "New modal feed network for a compact monopole array with isolated ports," *IEEE Trans. Antennas Propag.*, vol. 56, no. 12, pp. 3872-3875, Dec. 2008.
- [7] C. Volmer, M. Sengul, J. Weber, R. Stephan and M. A. Hein, "Broadband decoupling and matching of a superdirective two-port antenna array," *IEEE Antennas Wireless Propag. Lett.*, vol. 7, pp. 613-616, 2008.
- [8] V. C. Papamichael, "Eigen-analysis of lossy compact multielement antenna systems," *IEEE Antennas Wireless Propag. Lett.*, vol. 8, pp. 1334-1336, 2009.
- [9] L. K. Yeung and Y. E. Wang, "Mode-based beamforming arrays for miniaturized platforms," *IEEE Trans. Microw. Theory Techn.*, vol. 57, no. 1, pp. 45-52, Jan. 2009.
- [10] T. I. Lee and Y. E. Wang, "Mode-based beamforming with closely spaced antennas," *2007 IEEE/MTT-S Int. Microwave Symp.*, Honolulu, HI, 2007, pp. 1723-1726.
- [11] R. Shavit and E. Rivkin, "An efficient and practical decoupling feeding network for antenna phased arrays," *IEEE Antennas Wireless Propag. Lett.*, vol. 9, pp. 966-969, 2010.
- [12] Y. Cai and Y. Jay Guo, "A reconfigurable decoupling and matching network for a frequency agile compact array," *Proc. 5th European Conf. Antennas Propag.*, Rome, 2011, pp. 896-899.
- [13] M. Bouezzeddine and W. L. Schroeder, "On the design of wideband and tunable matching and decoupling networks for N-port antennas," *Proc. 8th European Conf. Antennas Propag.*, The Hague, 2014, pp. 2727-2731.
- [14] Y. Chen and D. Manteuffel, "A tunable decoupling and matching concept for compact mobile terminal antennas," *IEEE Trans. Antennas Propag.*, vol. 65, no. 4, pp. 1570-1578, April 2017.
- [15] Krewski and W. L. Schroeder, "N-port DL-MIMO antenna system realization using systematically designed mode matching and mode decomposition network," *2012 42nd Eur. Microw. Conf.*, Amsterdam, 2012, pp. 156-159.
- [16] G. Wolosinski, V. Fusco and O. Malyuskin, "2-bit polarisation agile antenna with high port decoupling," *IET Electronics Letters*, vol. 52, no. 4, pp. 255-256, 2 18 2016.

- [17] G. Wolosinski, V. Fusco and P. Rulikowski "Mode-based MIMO antenna with polarization and pattern diversity for base station applications," *2018 IEEE Int. Symposium Antennas Propag.*, Boston, 2018.
- [18] R. Tian, B. K. Lau, and Z. Ying "Multiplexing efficiency of MIMO antennas in arbitrary propagation scenarios," *Proc. 6th European Conf. Antennas Propag.*, Prague, 2012, pp. 373-377.
- [19] C. S. Lötbäck Patané, M. Franzén and C. Orlenius, "Analysis and improvement of reverberation chamber method for characterization of small and terminal antennas," *Proc. 5th European Conf. Antennas Propag.*, Rome, 2011, pp. 1832-1836.
- [20] A. Paulraj, R. Nabar, and D. Gore, "Introduction to Space-Time Wireless Communications," 1st ed., Cambridge, Cambridge University Press, 2003.
- [21] G. Wolosinski, V. Fusco and P. Rulikowski, "High-performance balun for a dual-polarized dipole antenna," *IET Microwaves, Antennas & Propagation* (accepted for publication) DOI: 10.1049/iet-map.2018.5294.

multi-functional RF front-ends and electromagnetic band gap structures.

Grzegorz Wolosinski (S'15) was born in Rome, Italy in 1989. He received the B.S. and M.S. (*summa cum laude*) degrees in electronic engineering from Roma Tre University, Rome, in 2012 and 2014, respectively. He is currently pursuing the Ph.D. degree in electronic engineering at the Queens University of Belfast ECIT Research Institute, Belfast, UK.

From 2014 to 2015, he was an Intern in the Antenna Design Division of Elettronica Group. From 2015 to 2018, he was a Marie Curie Early Stage Researcher at the Queens University of Belfast ECIT Research Institute. From 2015 to 2017, he was a Visiting Research Student at Nokia Bell Labs, Dublin. Since 2018 he has been a Senior Research Engineer at the Huawei German Research Center, Munich, Germany. His research interests include MIMO antennas for cellular communications, mutual coupling effect, antenna diversity, and OTA measurement techniques.

Vincent Fusco PhD, DSc, FIET, IEEE Fellow, FREng. He is current direct of Research at the Queens University of Belfast ECIT Research Institute. In 2012 he was awarded the IET senior achievement award the Mountbatten Medal for seminal contributions in the field of microwave electronics and its impact on UK industry. His fundamental

work on active antenna front-end techniques has provided generic advances in low cost phased and self-tracking antenna array architectures. He has written over 500 papers and two books and holds a number of antenna related patents.

Umair Naeem received his BE degree in electrical engineering from NED University of Engineering and Technology, Karachi, Pakistan, in 2005. He completed his Masters and Doctorate education, with specialization in high frequency electronics and optoelectronics, from University of Limoges, Limoges,

France, in 2007 and 2010 respectively. He is currently with Centre for Wireless Innovation, ECIT, Queen's University Belfast, UK. His research interests include synthesis of microwave and millimeter wave components and circuits,