Hybrid beamforming design for dual-polarised millimetre wave MIMO systems

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Millimetre wave (mmWave) communications provide a favourable solution to the ever-increasing data rate demands because of their abundant bandwidth availability. The conventional wisdom about the mmWave communication channel is that it always exhibits line-of-sight (LOS) propagation with directive antennas. However, in mobile communications, mmWave frequencies suffer from frequent LOS blockages and exhibit multipath propagation, which motivates the study and analysis of the polarimetric properties of the mmWave channel. While the exploitation of polarisation is promising, the cross-polarisation associated with it can be detrimental to the overall performance. Therefore, in this Letter, the authors propose a hybrid precoder design, which mitigates the cross-polarisation by the joint design of a radio-frequency beamformer as well as the precoder and combiner used in the baseband. The authors demonstrate through simulations that the proposed design outperforms eigen-beamforming by 10 dB at a given rate, when considering polarisation.

Introduction: Millimetre wave (mmWave) communication has recently gained attention due to its abundant spectral resources which can be harnessed to accommodate the increasing data rate demands of the mobile users [1]. However, given the high-propagation losses due to the atmospheric absorption, foliage density and rain-induced fading, the signal-to-noise ratio (SNR) at the receiver would be typically low [1]. Therefore, to mitigate the propagation losses, directional transmission in conceived for increasing the SNR at the receiver. Directional transmission at mmWave frequencies would use hybrid beamforming, where the signals are processed both in the analogue and digital domains relying on large $\lambda/2$-spaced antenna arrays. However, owing to the hardware and power requirements, employing large antenna arrays, especially at the mobile station would be challenging [2]. Therefore, a promising solution is to install a compact antenna array relying on a dual-polarisation to leverage diversity in addition to beamforming gain. Contrary to the popular belief, that mmWave links always exhibit line-of-sight (LOS) due to using directive antennas and polarisation matching, mmWave communications suffer from frequent LOS blockages and exhibit sparse multipath propagation, which motivates the study of polarisation effects on the mmWave channel.

Song *et al.* [3] proposed a soft-decision beam-alignment scheme for exploiting the orthogonal polarisation. More recently, Jo *et al.* [4] conducted empirical beamforming investigations by exploiting the dual-polarisation diversity in mmWave MIMO channels. To dynamically exploit the polarisation diversity and antenna array directivity, Wu *et al.* [5] proposed a reconfigurable hybrid beamforming architecture for dual-polarised mmWave channels.

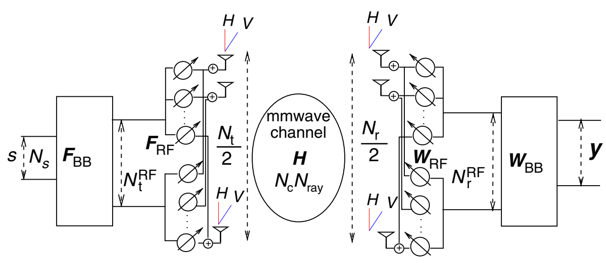

Fig. 1 *Dual-polarised hybrid beamforming architecture*

Against this background, we propose a hybrid precoder design, which mitigates the cross-polarisation by the joint design of a radio-frequency (RF) beamformer as well as a precoder and combiner in the baseband. In this design, we first obtain the fully-digital precoder and combiner using an iterative algorithm, which provides a locally optimum solution. Then, we perform the required matrix decomposition following the approach of [6], where the RF beamformers, as well as precoders and combiners in the baseband, are obtained. We demonstrate through our simulation results that the proposed design outperforms eigen-beamforming by about 10 dB at a given rate.

System model: Consider a single user dual-polarised mmWave MIMO system shown in Fig. 1, where the transmitter and the receiver are equipped with $N_t/2$ and $N_r/2$ antennas, respectively.

Furthermore, the antenna arrays of both the transmitter and the receiver of Fig. 1 are dual-polarised with horizontal (H) and vertical (V) polarisation. The transmitter processes the input symbol vector $\boldsymbol{s}$ of size N_s using a digital transmit precoder (TPC) matrix $\boldsymbol{F}_{BB}$ of size $N_t^{RF} \times N_s$ in the baseband and the precoded symbols are then phase shifted using the beamformer matrix $\boldsymbol{F}_{RF}$ of size $N_t \times N_t^{RF}$ in the RF. Then the received signal vector $\boldsymbol{y}$ after RF and baseband combining is given by

$$\boldsymbol{y} = \sqrt{\rho}\mathrm{W}_{BB}^{\dagger}\boldsymbol{W}_{RF}^{\dagger}\boldsymbol{H}\boldsymbol{F}_{RF}\boldsymbol{F}_{BB}\boldsymbol{s} + \boldsymbol{W}_{BB}^{\dagger}\boldsymbol{W}_{RF}^{\dagger}\boldsymbol{n}, \tag{1}$$

where $\boldsymbol{W}_{RF}$ and W_{BB} are the RF and baseband combiner matrices of sizes $N_r \times N_r^{RF}$ and $N_r^{RF} \times N_s$, respectively. To elaborate further, $\boldsymbol{n}$ is the Gaussian noise with distribution $\mathcal{CN}(0, \sigma_n^2)$, while $\boldsymbol{H}$ is the dual-polarised mmWave channel matrix of size $N_t \times N_r$ and $\mathbb{E}[\|\boldsymbol{H}\|_F^2] = N_t N_r$. It is instructive to note that $\boldsymbol{F}_{RF}$ is used for analogue beamforming, which is expressed as $\boldsymbol{F}_{RF} = \mathrm{diag}(\boldsymbol{F}_H, \boldsymbol{F}_V)$, where $\boldsymbol{F}_H$ and $\boldsymbol{F}_V$ are the beamforming matrices of size $N_t/2 \times N_t^{RF}/2$ for horizontal and vertical polarisation, respectively. Similarly, $\boldsymbol{W}_{RF} = \mathrm{diag}(\boldsymbol{W}_H, \boldsymbol{W}_V)$, where $\boldsymbol{W}_H$ and $\boldsymbol{W}_V$ are the beamforming matrices of size $N_r/2 \times N_r^{RF}/2$ for horizontal and vertical polarisation, respectively. Furthermore, the channel matrix $\boldsymbol{H}$ for a dual-polarised antenna array of Fig. 1, is given by [3, 7, 8]

$$\boldsymbol{H} = \left[\boldsymbol{X} \odot \left\{\begin{bmatrix} e^{j\angle\alpha_{hh,n_c}^{n_{ray}}} & e^{j\angle\alpha_{hv,n_c}^{n_{ray}}} \\ e^{j\angle\alpha_{vh,n_c}^{n_{ray}}} & e^{j\angle\alpha_{vv,n_c}^{n_{ray}}} \end{bmatrix}\right\} \otimes \boldsymbol{H}'\right]\boldsymbol{G} = \begin{bmatrix} \boldsymbol{H}_{HH} & \boldsymbol{H}_{HV} \\ \boldsymbol{H}_{VH} & \boldsymbol{H}_{VV} \end{bmatrix},$$

where $\angle\alpha_{xy,n_c}^{n_{ray}}$ is the initial random phase of n_c cluster and n_{ray} ray that departs from polarisation 'x' and arrives in the polarisation direction 'y', while $\boldsymbol{X}$ is the power-imbalance due to the polarisation and $\boldsymbol{H}'$ is the mmWave statistical channel model [3]. Furthermore, $\boldsymbol{G}$ is Givens' rotation that captures the difference in orientation between the transmitter and the receiver [7, 8]. Note that $\odot$ and $\otimes$ represent Hadamard and Kronecker products, respectively.

Proposed transceiver design: In this section, we first present the fully-digital precoder and combiner design for the system model considered. Then after obtaining the fully-digital solution, we decompose the precoder and combiner matrices into the analogue–digital hybrid design. Let us introduce $\boldsymbol{F}_{RF}\boldsymbol{F}_{BB} = \boldsymbol{F}$ and $\boldsymbol{W}_{RF}\boldsymbol{W}_{BB} = \boldsymbol{W}$. Furthermore, the system model in (1) can be decomposed into a sum of co-polarised and cross-polarised terms, which are given as

$$\boldsymbol{y}_H = \underbrace{\rho_{hh}\boldsymbol{W}_H^{\dagger}\boldsymbol{H}_{HH}\boldsymbol{F}_H\boldsymbol{s}_H}_{\text{co-polarisation}} + \underbrace{\rho_{hv}\boldsymbol{W}_H^{\dagger}\boldsymbol{H}_{HV}\boldsymbol{F}_V\boldsymbol{s}_V}_{\text{cross-polarisation}} + \underbrace{\boldsymbol{W}_H^{\dagger}\boldsymbol{n}}_{\text{noise}}, \tag{2}$$

$$\boldsymbol{y}_V = \underbrace{\rho_{vv}\boldsymbol{W}_V^{\dagger}\boldsymbol{H}_{VV}\boldsymbol{F}_V\boldsymbol{s}_V}_{\text{co-polarisation}} + \underbrace{\rho_{vh}\boldsymbol{W}_V^{\dagger}\boldsymbol{H}_{VH}\boldsymbol{F}_H\boldsymbol{s}_H}_{\text{cross-polarisation}} + \underbrace{\boldsymbol{W}_V^{\dagger}\boldsymbol{n}}_{\text{noise}}. \tag{3}$$

The cross-polarisation power leakage from vertical to horizontal polarisation plus noise for (2) is given by $\boldsymbol{C}_H = \mathrm{Tr}(\boldsymbol{W}_H^H\boldsymbol{R}_H\boldsymbol{W}_H)$, where $\boldsymbol{R}_H = \rho_{hv}\boldsymbol{H}_{HV}\boldsymbol{F}_V(\boldsymbol{H}_{HV}\boldsymbol{F}_H)^H + \sigma_n^2\boldsymbol{I}_n$, and $\boldsymbol{I}_n$ is the noise power. Our objective is to design $\boldsymbol{W}_H$ and $\boldsymbol{F}_H$ so that the cross-polarisation leakage power $\boldsymbol{C}_H$ is minimised, while simultaneously preserving the signal dimensions, i.e. rank $\left(\boldsymbol{W}_H^H\boldsymbol{H}_{HH}\boldsymbol{F}_H\right) = N_{s_H}$.

The optimisation problem can be formulated as

$$\begin{aligned} &\min_{\boldsymbol{W}_H}\mathrm{Tr}\left(\boldsymbol{W}_H^{\dagger}\boldsymbol{R}_H\boldsymbol{W}_H\right) \\ &\text{s.t.}\boldsymbol{W}_H^{\dagger}\boldsymbol{H}_{HH}\boldsymbol{F}_H = \alpha\boldsymbol{I}_{N_{s_H}}, \end{aligned} \tag{4}$$

where $\boldsymbol{R}_H$ is a positive definite matrix ($\boldsymbol{R}_H \succ 0$). Then, we continue by forming the Lagrangian function by

$$\mathcal{L}(\boldsymbol{W}_H, z) = \left(\boldsymbol{W}_H^{\dagger}\boldsymbol{R}_H\boldsymbol{W}_H\right) + z\left(\boldsymbol{W}_H^{\dagger}\boldsymbol{H}_{HH}\boldsymbol{F}_H - \boldsymbol{I}_{N_s}\right). \tag{5}$$

Then, the Lagrangian conditions for this problem are

$$\nabla_{(\boldsymbol{W}_H^{opt})^{\dagger}}\mathcal{L} = 0, \tag{6}$$

$$z^*\left((\boldsymbol{W}_H^{opt})^{\dagger}\boldsymbol{H}_{HH}\boldsymbol{F}_H - \alpha\boldsymbol{I}_{N_s}\right) = 0. \tag{7}$$

Explicitly, (6) can be written as

$$\begin{aligned} &\nabla_{(\boldsymbol{W}_H^{opt})^{\dagger}}\mathrm{Tr}\left((\boldsymbol{W}_H^{opt})^{\dagger}\boldsymbol{R}_H\boldsymbol{W}_H^{opt}\right) + \\ &z^*\nabla_{(\boldsymbol{W}_H^{opt})^{\dagger}}\left((\boldsymbol{W}_H^{opt})^{\dagger}\boldsymbol{H}_{HH}\boldsymbol{F}_H - \boldsymbol{I}_{N_s}\right) = 0, \end{aligned} \tag{8}$$

where ∇ is the gradient operation and z^* is the Lagrangian multiplier. By taking the derivative with the respect to $(\boldsymbol{W}_{\mathrm{H}}^{\mathrm{opt}})^{\dagger}$ in (8), we obtain $\boldsymbol{R}_{\mathrm{H}}\boldsymbol{W}_{\mathrm{H}}^{\mathrm{opt}} + z\boldsymbol{H}_{\mathrm{HH}}\boldsymbol{F}_{\mathrm{H}} = 0$ and $\boldsymbol{W}_{\mathrm{H}}^{\mathrm{opt}} = -\boldsymbol{R}_{\mathrm{H}}^{-1}\boldsymbol{H}_{\mathrm{HH}}\boldsymbol{F}_{\mathrm{H}}z$.

Upon substituting $\boldsymbol{W}_{\mathrm{H}}^{\mathrm{opt}}$ into (7), we obtain

$$\boldsymbol{W}_{\mathrm{H}}^{\mathrm{opt}} = \alpha\boldsymbol{R}_{\mathrm{H}}^{-1}\boldsymbol{H}_{\mathrm{HH}}\boldsymbol{F}_{\mathrm{H}}\left[(\boldsymbol{H}_{\mathrm{HH}}\boldsymbol{F}_{\mathrm{H}})^{\dagger}\boldsymbol{R}_{\mathrm{H}}^{-1}(\boldsymbol{H}_{\mathrm{HH}}\boldsymbol{F}_{\mathrm{H}})\right]^{-1}, \tag{9}$$

where α is the normalisation constant expressed as

$$\alpha = \frac{1}{\sqrt{\mathrm{Tr}\left((\boldsymbol{W}_{\mathrm{H}}^{\mathrm{opt}})^{\mathrm{H}}\boldsymbol{W}_{\mathrm{H}}^{\mathrm{opt}}\right)}}.$$

Having designed the combiner matrix $\boldsymbol{W}_{\mathrm{H}}$, we now aim for designing the precoder matrix that minimises the leakage of its own power into the orthogonal polarisation V.

The cross-polarisation power leakage induced by its own transmission into the vertical polarisation (3) is given by $\boldsymbol{C}_{\mathrm{V}} = \mathrm{Tr}\left(\boldsymbol{F}_{\mathrm{H}}^{\dagger}\boldsymbol{S}_{\mathrm{V}}\boldsymbol{F}_{\mathrm{H}}\right)$, where $\boldsymbol{S}_{\mathrm{V}} = \rho_{\mathrm{vh}}\left(\boldsymbol{W}_{\mathrm{V}}^{\dagger}\boldsymbol{H}_{\mathrm{vh}}\right)^{\dagger}\left(\boldsymbol{W}_{\mathrm{V}}^{\dagger}\boldsymbol{H}_{\mathrm{vh}}\right) + \boldsymbol{I}$.

Then, the optimisation problem can be formulated as

$$\begin{aligned} &\min_{\boldsymbol{F}_{\mathrm{H}}} \mathrm{Tr}\left(\boldsymbol{F}_{\mathrm{H}}^{\dagger}\boldsymbol{S}_{\mathrm{V}}\boldsymbol{F}_{\mathrm{H}}\right) \\ &\mathrm{s.t.}\, \boldsymbol{W}_{\mathrm{H}}^{\dagger}\boldsymbol{H}_{\mathrm{HH}}\boldsymbol{F}_{\mathrm{H}} = \beta\boldsymbol{I}_{N_{s_{\mathrm{H}}}}. \end{aligned} \tag{10}$$

Following the same analysis for obtaining $\boldsymbol{W}_{\mathrm{H}}^{\mathrm{opt}}$, the locally-optimal solution to $\boldsymbol{F}_{\mathrm{H}}^{\mathrm{opt}}$ is given by

$$\boldsymbol{F}_{\mathrm{H}}^{\mathrm{opt}} = \beta\boldsymbol{S}_{\mathrm{V}}^{-1}\boldsymbol{H}_{\mathrm{HH}}^{\dagger}\boldsymbol{W}_{\mathrm{H}}^{\dagger}\left(\boldsymbol{W}_{\mathrm{H}}^{\dagger}\boldsymbol{H}_{\mathrm{HH}}\boldsymbol{S}_{\mathrm{V}}^{-1}\left(\boldsymbol{W}_{\mathrm{H}}^{\dagger}\boldsymbol{H}_{\mathrm{HH}}\right)^{\dagger}\right), \tag{11}$$

where β is the normalisation constant expressed as

$$\beta = \frac{1}{\sqrt{\mathrm{Tr}\left((\boldsymbol{F}_{\mathrm{H}}^{\mathrm{opt}})^{H}\boldsymbol{F}_{\mathrm{H}}^{\mathrm{opt}}\right)}}.$$

Remark 1: The combiners $\boldsymbol{W}_{\mathrm{H}}$ and $\boldsymbol{W}_{\mathrm{V}}$ are designed to minimise the interference leakage from the opposite polarisation, while the combiners $\boldsymbol{F}_{\mathrm{H}}$ and $\boldsymbol{F}_{\mathrm{V}}$ are designed to minimise the interference caused by its own transmission to the opposite polarisation.

Remark 2: The solution for combiner $\boldsymbol{W}_{\mathrm{v}}$ in (9) and the solution for $\boldsymbol{F}_{\mathrm{H}}$ in (11) are presented for horizontal polarisation. However, the proposed derivation can be readily applied to vertical polarisation to obtain the combiner $\boldsymbol{W}_{\mathrm{V}}$ and precoder $\boldsymbol{F}_{\mathrm{V}}$.

Having obtained the fully-digital solution, we now decompose the digital solution into a hybrid product of $\boldsymbol{F}_{\mathrm{RF}}$ and $\boldsymbol{F}_{\mathrm{BB}}$. However, it is instructive to note that the beamformer matrix followed by the TPC imposes an important constraint to have constant modulus gain. Mathematically, this can be formulated as

$$\min_{\boldsymbol{F}_{\mathrm{H_{RF}}},\, \boldsymbol{F}_{\mathrm{H_{BB}}}} \|\boldsymbol{F}_{\mathrm{H}}^{\mathrm{opt}} - \boldsymbol{F}_{\mathrm{H_{RF}}}\boldsymbol{F}_{\mathrm{H_{BB}}}\|_{\mathrm{F}}^{2}, \tag{12}$$

$$\mathrm{s.t.}\, |\boldsymbol{F}_{\mathrm{H_{RF}}}(m, n)|^2 = 1. \tag{13}$$

Solving (12) is not straightforward, since the problem is non-convex. Therefore, we first obtain the solution using the least squares (LS) algorithm, where we assume that $\boldsymbol{F}_{\mathrm{H_{RF}}}$ is constant to obtain $\boldsymbol{F}_{\mathrm{H_{BB}}}$, while we assume that $\boldsymbol{F}_{\mathrm{H_{BB}}}$ as a constant to obtain $\boldsymbol{F}_{\mathrm{H_{RF}}}$. Then we invoke a specific proposition of [6] to meet the constraints imposed on the RF beamformer matrix. Note that this problem is solved iteratively.

Therefore, the LS solution for (12) in the $(k+1)$th iteration is given by

$$\boldsymbol{F}_{\mathrm{H_{BB}}}^{k+1} = \left(\boldsymbol{F}_{\mathrm{H_{RF}}}^{\dagger k}\boldsymbol{F}_{\mathrm{H_{RF}}}^{k}\right)^{-1}\boldsymbol{F}_{\mathrm{H_{RF}}}^{\dagger k}\boldsymbol{F}_{\mathrm{H}}^{\mathrm{opt}}, \tag{14}$$

$$\boldsymbol{F}_{\mathrm{H_{RF}}}^{k+1} = \boldsymbol{F}_{\mathrm{H}}^{\mathrm{opt}}\boldsymbol{F}_{\mathrm{H_{BB}}}^{\dagger k+1}\left(\boldsymbol{F}_{\mathrm{H_{BB}}}^{\dagger k+1}\boldsymbol{F}_{\mathrm{H_{BB}}}^{\dagger k+1}\right)^{-1}. \tag{15}$$

Having obtained the LS solution, we now obtain the constrained RF beamformer solution by following the approach discussed in [6].

Simulation results: In this section, we characterise the performance in terms of rate as well as of BER for both the proposed design and eigen beamforming, where the right and left singular vectors of the corresponding channel matrix are used. To evaluate the performance, we have performed Monte Carlo simulations.

Fig. 2*a* shows the rate of the proposed design and eigen beamforming design for a 128×32 dual-polarised antenna array, where θ and ϕ are uniformly distributed from 0 to 2π and $N_{\mathrm{t}}^{\mathrm{RF}}, N_{\mathrm{r}}^{\mathrm{RF}}, N_{\mathrm{s}}$ are set to 2 whilst $\mathcal{X}$ is varied from 0 to 20 dB, where $\mathcal{X}$ is defined as the ratio of co-polarisation power (ρ_{xx}) to cross polarisation power (ρ_{xy}). In this configuration, two streams are transmitted using two RF chains. It can be seen from Fig. 2*a* that for lower values of $\mathcal{X}$, the eigen beamforming saturates at lower rates, while the proposed design outperforms eigen beamforming with a SNR gain of more than 10 dB, especially when $\mathcal{X}$ is as low as 0 and 10 dB. Furthermore, when $\mathcal{X}$ is increased to 20 dB, the proposed design outperforms the eigen beamforming by about 5 dB.

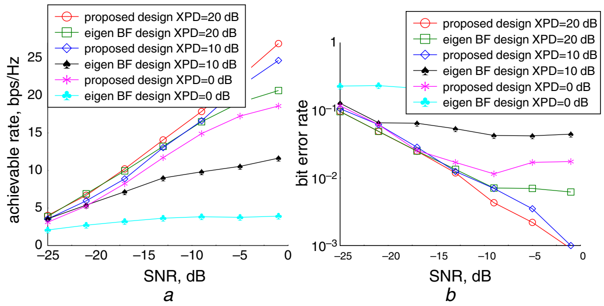

Fig. 2 *Performance of* 128×32 *dual-polarised antenna*
a Achievable rate
b BER

Additionally, to understand the reliability of the system using cross-polarisation, Fig. 2*b* shows the BER of both the proposed and eigen beamforming designs. It is interesting to note that at SNR of −1 dB, the proposed design achieves a BER as low as 10^{-3} while the eigen beamforming produces an error floor for $\mathcal{X} = 10$ dB. In contrast, when the cross-polarisation power leakage is high, i.e. $\mathcal{X} = 0$ dB, both produces error floor. When the cross-polarisation is low, i.e. $\mathcal{X} = 20$ dB, the proposed design outperforms the eigen beamforming by a significant margin.

Conclusion: In this Letter, mmWave hybrid beamforming relying on dual-polarisation is proposed. More explicitly, we proposed a hybrid precoder design, which mitigates the cross-polarisation by the joint design of the RF beamformer as well as the precoder and combiner in the baseband. We demonstrated through our simulation results that the proposed design outperforms eigen-beamforming by more than 10 dB at a given rate.

Submitted: *1 May 2018*
doi: 10.1049/el.2018.5097
One or more of the Figures in this Letter are available in colour online.

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