BEAMFORMING DESIGN FOR FULL-DUPLEX CELLULAR AND MIMO RADAR COEXISTENCE: A RATE MAXIMIZATION APPROACH

Sudip Biswas[†], *Keshav Singh*[†], *Omid Taghizadeh*[‡], *T. Ratnarajah*[†], *and Mathini Sellathurai*[*]

[†]Institute for Digital Communications, School of Engineering, University of Edinburgh, UK.
[‡]Institute for Theoretical Information Technology, RWTH Aachen University, Germany
[*]School of Engineering & Physical Sciences, Heriot-Watt University, Edinburgh, UK
E-mail: {sudip.biswas, k.singh, t.ratnarajah}@ed.ac.uk[†], taghizadeh@ti.rwth-aachen.de[‡]

ABSTRACT

We propose a novel transceiver design technique to facilitate flexible spectrum sharing between a multiple-input multiple-output (MIMO) radar and a full-duplex (FD) MIMO cellular system. The optimization problem for maximizing the rate of the cellular system is formulated, subject to the constraints of individual power at the uplink users, total power at the base station, and interference power towards the MIMO radar from the cellular system so that the detection probability of the radar is not hindered. We show that the above problem can be cast as a second-order cone programming problem and the joint design of transceiver matrices can be obtained through an iterative algorithm. Numerical results show that using the spectrum shared by the radar, the FD cellular system can achieve sum rate of up to 25-30 bits/sec/Hz for a reasonable self-interference cancellation of around -70 dB. However, to facilitate this, while also maintaining a detection probability of around 0.9, the radar needs to spend an extra power of around 2-3 dB.

Index Terms— Full-duplex, radar, MIMO, spectrum sharing, beamforming, optimization, licensed shared access.

1. INTRODUCTION

One of the key reasons for the recent spectrum paucity is the highly inefficient spectrum utilization due to fixed static spectrum allocation [1, 2]. In this regard, the S-band (2-4 GHz) and C-band (4-8 GHz) that are occupied by a variety of radar applications are being anticipated to be used for cellular communications in future. As a result, spectrum sharing between radar and communication systems has recently captured the attention of both academia and industry [3]. One of the key techniques for opportunistic spectrum access is licensed shared access (LSA)/authorized shared access (ASA). The motivating factors for LSA/ASA-based spectrum sharing are the reports presented on efficient spectrum utilization by President's Council of Advisers on Science and Technology (PCAST), which focused to share 1.0 GHz of government-held spectrum [4] and the low utilization of huge amounts of spectrum held by the federal incumbents, for example: the $3.55-3.65$ GHz band [5], $5.25-5.35$ GHz, and $5.47-5.725$ GHz [6]. Recently, adequate studies have focused on the subject of the spectrum sharing between radars and communication systems [7–10].

Apart from LSA/ASA-based spectrum sharing, full-duplex (FD) transmission is another promising technology that can significantly improve the spectrum efficiency [11–13]. In particular, a FD transceiver can receive and transmit at the same time and frequency resource. However, the self-interference (SI) caused by the signal leakage from the transmitting antennas to its receiving antennas dominates the performance of FD systems. Nevertheless, recent advances in interference cancellation techniques and transmit/receive antenna isolation such as antenna design, and analog and digital domain SI cancellation techniques [14], have enabled FD transceivers to sufficiently combat the SI. However, due to the non-ideal nature of the transmit and receive chains [11], also known as hardware impairments, the SI cannot be completely eradicated in practice.

Motivated by the aforementioned discussion, in this paper, we consider a hardware impaired FD multiple-input multiple-output (MIMO) cellular system that operates in the spectrum shared by a MIMO radar. Since the cellular system uses the spectrum shared by the radar, it is bound to interfere with the radar. Hence, the interference from the cellular system to the radar must be constrained to meet the requirements of detection probabilities of the radar, while also maximizing the throughput of the cellular system. Accordingly, we formulate a transceiver design problem for the coexistence of the cellular system and the radar.

2. SYSTEM MODEL

We consider the co-existence of a FD MIMO cellular communication system with a MIMO radar as shown in Fig. 1, where the MIMO cellular system operates in the spectrum shared by the MIMO radar[1] over a bandwidth of B Hz. The FD MIMO cellular system comprises of a FD MIMO BS, which consists of M_0 transmit and N_0 receive antennas, and J DL, and K UL users. All DL and UL users operate in half-duplex (HD) mode and each DL and UL user is equipped with N_j receive and M_k transmit antennas, respectively. The k-th UL and the j-th DL channels are represented as $\mathbf{H}_k^{UL} \in \mathbb{C}^{N_0 \times M_k}$ and $\mathbf{H}_j^{DL} \in \mathbb{C}^{N_j \times M_0}$, respectively. The SI channel at the FD BS and the co-channel interference (CCI) channel between the k-th UL and j-th DL users are denoted as $\mathbf{H}_0 \in \mathbb{C}^{N_0 \times M_0}$ and $\mathbf{H}_{jk}^{DU} \in \mathbb{C}^{N_j \times M_k}$, respectively. Let $\mathbf{s}_k^{UL} \in \mathbb{C}^{d_k^{UL} \times 1}$ and $\mathbf{s}_j^{DL} \in \mathbb{C}^{d_j^{DL} \times 1}$ denote the communication symbols for the cellular system at the k-th UL and j-th DL users, respectively. Further, these symbols are assumed to be independent and identically distributed (i.i.d.) with unit power,

[1]In our spectrum sharing model, while we employ beamforming design at the cellular system, null space based projection method [10], in which the radar projects null space towards cellular system is used. Also we assume the availability of global CSI at both systems. This is a reasonable assumption for LSA/ASA scenarios.

This work was supported by the U.K. Engineering and Physical Sciences Research Council (EPSRC) under Grants EP/N014073/1 and EP/M014126/1.

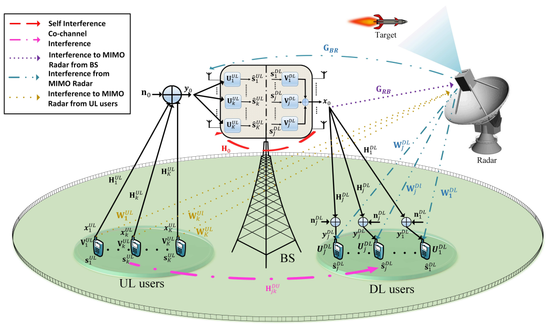

Fig. 1: A FD MIMO cellular system in coexistence with a MIMO radar.

i.e., $\mathbb{E}\left[\mathbf{s}_k^{UL}\left(\mathbf{s}_k^{UL}\right)^H\right] = \mathbf{I}_{d_k^{UL}}$ and $\mathbb{E}\left[\mathbf{s}_j^{DL}\left(\mathbf{s}_j^{DL}\right)^H\right] = \mathbf{I}_{d_j^{DL}}$. The communication symbols $\mathbf{s}_k^{UL}$ and $\mathbf{s}_j^{DL}$ are first precoded by matrices $\mathbf{V}_k^{UL} \in \mathbb{C}^{M_k \times d_k^{UL}}$ and $\mathbf{V}_j^{DL} \in \mathbb{C}^{M_0 \times d_j^{DL}}$, such that $\mathbf{V}_k^{UL}$ denotes the precoder for the data streams of the k-th UL, and $\mathbf{V}_j^{DL}$ indicates the precoder for the data stream of the j-th DL users. Further, the hardware impairment model as given in [11] is adopted, where at each transmit antenna in the cellular system an additive white Gaussian "transmitter distortion" with variance ψ times the energy of the undistorted transmit signal is applied, and at each receive antenna, an additive white Gaussian "receiver noise" with variance υ times the energy of the received signal is applied.

If $\mathbf{x}_k^{UL}$ and $\mathbf{x}_0$ are the signal transmitted from the k-th UL user and the FD cellular BS, defined as $\mathbf{x}_k^{UL} = \mathbf{V}_k^{UL}\mathbf{s}_k^{UL}$ and $\mathbf{x}_0 = \sum_{j=1}^{J}\mathbf{V}_j^{DL}\mathbf{s}_j^{DL}$, then the signal received at the BS and the j-th DL user can be written, respectively, as

$$\mathbf{y}_0 = \sum_{k=1}^{K}\mathbf{H}_k^{UL}\left(\mathbf{x}_k^{UL} + \mathbf{c}_k^{UL}\right) + \mathbf{H}_0\left(\mathbf{x}_0 + \mathbf{c}_0\right) + \mathbf{e}_0 + \mathbf{W}_{BR}^{DL}\mathbf{s}_R + \mathbf{n}_0, \tag{1}$$

$$\mathbf{y}_j^{DL} = \mathbf{H}_j^{DL}\left(\mathbf{x}_0 + \mathbf{c}_0\right) + \sum_{k=1}^{K}\mathbf{H}_{jk}^{DU}\left(\mathbf{x}_k^{UL} + \mathbf{c}_k^{UL}\right) + \mathbf{e}_j^{DL} + \mathbf{W}_j^{DL}\mathbf{s}_R + \mathbf{n}_j^{DL}, \tag{2}$$

where $\mathbf{W}_{BR}^{DL} \in \mathbb{C}^{N_0 \times R_T}$ and $\mathbf{W}_j^{DL} \in \mathbb{C}^{N_j \times R_T}$ denote the interference channels from radar transmitter to BS and j-th DL user, respectively, while $\mathbf{s}_R \in \mathbb{C}^{R_T \times 1}$ represents the transmitted vector by the radar with $\mathbb{E}\left[\| \mathbf{s}_R \|^2\right] = P_R\mathbf{I}$, where P_R indicates the power of the radar signals. The terms $\mathbf{n}_0 \in \mathbb{C}^{N_0}$ and $\mathbf{n}_j^{DL} \in \mathbb{C}^{N_j}$ in (1) and (2) denote the additive white Gaussian noise (AWGN) vector with zero mean and covariance matrix $\mathbf{R}_0 = \sigma_0^2\mathbf{I}_{N_0}$ and $\mathbf{R}_j^{DL} = \sigma_j^2\mathbf{I}_{N_j}$ at the BS and the j-th DL user, respectively. Further, $\mathbf{c}_k^{UL}$ represents the distortion at the transmitter at the k-th UL user and $\mathbf{c}_0$ is the distortion at the BS, which closely approximates the effects of phase noise, non-linearities in the DAC and additive power-amplifier noise. The covariance matrix of $\mathbf{c}_k^{UL}$ is given by

$$\mathbf{c}_k^{UL} \sim \mathcal{CN}\left(\mathbf{0}, \psi \text{ diag}\left(\mathbf{V}_k^{UL}\left(\mathbf{V}_k^{UL}\right)^H\right)\right), \ \mathbf{c}_k^{UL} \perp \mathbf{x}_k^{UL}. \tag{3}$$

$\mathbf{e}_j^{DL}$ in (1) and $\mathbf{e}_0$ in (2) are the receiver distortion at the j-th DL user and the BS, respectively, which closely approximates the combined effects of non-linearities in the ADC, additive gain-control noise and phase noise. The covariance matrix of $\mathbf{e}_j^{DL}$ is given by

$$\mathbf{e}_j^{DL} \sim \mathcal{CN}\left(\mathbf{0}, \upsilon \text{ diag}\left(\mathbf{\Phi}_j^{DL}\right)\right), \ \mathbf{e}_j^{DL} \perp \mathbf{u}_j^{DL}, \tag{4}$$

where $\mathbf{\Phi}_j^{DL} = \text{Cov}\{\mathbf{u}_j^{DL}\}$ and $\mathbf{u}_j^{DL}$ is the undistorted received vector at the j-th DL user. Similarly, the above transmitter/receiver distortion model holds for $\mathbf{c}_0$ and $\mathbf{e}_0$, as well.

Now, we apply linear receive filters $\mathbf{U}_k^{UL} \in \mathbb{C}^{N_0 \times d_k^{UL}}$ and $\mathbf{U}_j^{DL} \in \mathbb{C}^{N_j \times d_j^{DL}}$ to $\mathbf{y}_0$ and $\mathbf{y}_j^{DL}$ to obtain the estimated source symbols $\hat{\mathbf{s}}_k^{UL}$ and $\hat{\mathbf{s}}_j^{DL}$, respectively. Using these estimates, the rate of the k-th UL and j-th DL users can be written as

$$R_k^{UL} = \log_2\left|\mathbf{I}_k^{UL} + \mathbf{H}_k^{UL}\mathbf{V}_k^{UL}\left(\mathbf{V}_k^{UL}\right)^H\left(\mathbf{H}_k^{UL}\right)^H\left(\mathbf{\Sigma}_k^{UL}\right)^{-1}\right|, \tag{5}$$

$$R_j^{DL} = \log_2\left|\mathbf{I}_j^{DL} + \mathbf{H}_j^{DL}\mathbf{V}_j^{DL}\left(\mathbf{V}_j^{DL}\right)^H\left(\mathbf{H}_j^{DL}\right)^H\left(\mathbf{\Sigma}_j^{DL}\right)^{-1}\right|, \tag{6}$$

where $\mathbf{\Sigma}_k^{UL}$ and $\mathbf{\Sigma}_j^{DL}$ are the approximated aggregate interference-plus-noise terms[2] given as

$$\begin{aligned}
\mathbf{\Sigma}_k^{UL} \approx & \sum_{j\neq k}^{K}\mathbf{H}_j^{UL}\mathbf{V}_j^{UL}\left(\mathbf{V}_j^{UL}\right)^H\left(\mathbf{H}_j^{UL}\right)^H \\
& + \psi\sum_{j=1}^{K}\mathbf{H}_j^{UL}\text{diag}\left(\mathbf{V}_j^{UL}\left(\mathbf{V}_j^{UL}\right)^H\right)\left(\mathbf{H}_j^{UL}\right)^H \\
& + \sum_{j=1}^{J}\mathbf{H}_0\left(\mathbf{V}_j^{DL}\left(\mathbf{V}_j^{DL}\right)^H + \psi\text{diag}\left(\mathbf{V}_j^{DL}\left(\mathbf{V}_j^{DL}\right)^H\right)\right)\mathbf{H}_0^H \\
& + P_R\left(\mathbf{W}_{BR}^{DL}\left(\mathbf{W}_{BR}^{DL}\right)^H\right) + \sigma_0^2\mathbf{I}_{N_0} \\
& + \upsilon\sum_{j=1}^{J}\text{diag}\left(\mathbf{H}_0\mathbf{V}_j^{DL}\left(\mathbf{V}_j^{DL}\right)^H\mathbf{H}_0^H\right) \\
& + \upsilon\sum_{j=1}^{K}\text{diag}\left(\mathbf{H}_j^{UL}\mathbf{V}_j^{UL}\left(\mathbf{V}_j^{UL}\right)^H\left(\mathbf{H}_j^{UL}\right)^H\right), \\
\mathbf{\Sigma}_j^{DL} \approx & \sum_{i\neq j}^{J}\mathbf{H}_j^{DL}\mathbf{V}_i^{DL}\left(\mathbf{V}_i^{DL}\right)^H\left(\mathbf{H}_j^{DL}\right)^H \\
& + \psi\sum_{i=1}^{J}\mathbf{H}_j^{DL}\text{diag}\left(\mathbf{V}_i^{DL}\left(\mathbf{V}_i^{DL}\right)^H\right)\left(\mathbf{H}_j^{DL}\right)^H \\
& + \sum_{k=1}^{K}\mathbf{H}_{jk}^{DU}\left(\mathbf{V}_k^{UL}\left(\mathbf{V}_k^{UL}\right)^H + \psi\text{diag}\left(\mathbf{V}_k^{UL}\left(\mathbf{V}_k^{UL}\right)^H\right)\right)\left(\mathbf{H}_{jk}^{DU}\right)^H \\
& + P_R\left(\mathbf{W}_j^{DL}\left(\mathbf{W}_j^{DL}\right)^H\right) + \sigma_j^2\mathbf{I}_{N_j} \\
& + \upsilon\sum_{k=1}^{K}\text{diag}\left(\mathbf{H}_{jk}^{DU}\mathbf{V}_k^{UL}\left(\mathbf{V}_k^{UL}\right)^H\left(\mathbf{H}_{jk}^{DU}\right)^H\right) \\
& + \upsilon\sum_{i=1}^{J}\text{diag}\left(\mathbf{H}_j^{DL}\mathbf{V}_i^{DL}\left(\mathbf{V}_i^{DL}\right)^H\left(\mathbf{H}_j^{DL}\right)^H\right).
\end{aligned} \tag{7, 8}$$

2.1. Spectrum Sharing MIMO Radar

In this work, we consider that the MIMO radar operates in the Federal Communications Commission (FCC) proposed $3.55-3.65$ GHz band [15]. Since the cellular system operates in the spectrum shared by the radar, it will create interference to the radar. The interference power from the cellular system with K UL users and a BS towards the radar equipped with R_R receive antennas is given as

[2] Note that approximation of $\mathbf{\Sigma}_k^{UL}$ and $\mathbf{\Sigma}_j^{DL}$ under $\psi << 1$ and $\upsilon << 1$ is a practical assumption [11]. However, although the terms ψ and υ are much smaller than 1, when they are applied on a strong channel alone, i.e., SI channel, they may no longer be negligible [11].

$$I^{RAD} = \sum_{k=1}^{K} \text{tr}\left\{ \mathbf{G}_{RU_k} \left(\mathbf{V}_k^{UL} \left(\mathbf{V}_k^{UL}\right)^H + \psi \text{diag}\left(\mathbf{V}_k^{UL} \left(\mathbf{V}_k^{UL}\right)^H \right) \right) \left(\mathbf{G}_{RU_k}\right)^H \right\} + \sum_{j=1}^{J} \text{tr}\left\{ \mathbf{G}_{RB} \left(\mathbf{V}_j^{DL} \left(\mathbf{V}_j^{DL}\right)^H + \psi \text{diag}\left(\mathbf{V}_j^{DL} \left(\mathbf{V}_j^{DL}\right)^H \right) \right) \left(\mathbf{G}_{RB}\right)^H \right\}, \quad (9)$$

where $\mathbf{G}_{RU_k} \in \mathbb{C}^{R_R \times M_k}$ $\left(\mathbf{G}_{RB} \in \mathbb{C}^{R_R \times M_0}\right)$ is the channel between the radar and k-th UL user (radar and the BS).

By considering the echo wave in a single range-Doppler bin of the radar, the discrete time signal vector received by radar at an angle θ can be expressed as

$$\mathbf{y}_R = \alpha\sqrt{P}_R \mathbf{V}(\theta) \mathbf{S}_R + \mathbf{G}_{RB} \sum_{j=1}^{J} \mathbf{V}_j^{DL} \mathbf{s}_j^{DL} + \sum_{k=1}^{K} \mathbf{G}_{RU_k} \mathbf{V}_k^{UL} \mathbf{s}_k^{UL} + \mathbf{n}_R, \quad (10)$$

where $\mathbf{G}_{RB} \in \mathbb{C}^{R_R \times M_0}$ and $\mathbf{G}_{RU_k} \in \mathbb{C}^{R_R \times M_k}$ are the interference channels matrices from BS to radar receiver and from k-th UL user to radar receiver, respectively. α indicates the complex path loss of the radar-target-radar path including the propagation loss and the coefficient of reflection and $\mathbf{n}_R(n) \sim \mathcal{CN}\left(0, \sigma_R^2 \mathbf{I}_R\right)$, and $\mathbf{V}(\theta)$ denotes the transmit-receive steering matrix expressed as $\mathbf{V}(\theta) \triangleq \mathbf{v}_R(\theta) \mathbf{v}_T^T(\theta)$, where $\mathbf{v}_T \in \mathbb{C}^{R_T \times 1}$ and $\mathbf{v}_R \in \mathbb{C}^{R_R \times 1}$ express transmit and receive steering vectors of radar antenna array. Using the same model defined in [4], we express $\mathbf{V}_{ir}(\theta)$ with assumptions $R_R = R_T = R$, $\mathbf{v}_R(\theta) = \mathbf{v}_T(\theta) = \mathbf{v}(\theta)$, and $\mathbf{V}_{ir}(\theta) = \mathbf{v}_i(\theta) \mathbf{v}_r(\theta) = \exp(-j\omega\tau_{ir}(\theta))$ as $\mathbf{V}_{ir}(\theta) = \exp\left(-j\frac{2\pi}{\lambda}\left[\sin(\theta); \cos(\theta)\right]^T (\mathbf{z}_i + \mathbf{z}_r)\right)$, where $\mathbf{V}_{ir}(\theta)$ denotes the i-th element at the r-th column of the matrix $\mathbf{V}$ and $\mathbf{z}_i = \left[z_i^1; z_i^2\right]$ is the location of the i-th element of the antenna array. ω and λ express the frequency and the wavelength of the carrier.

For determining the detection probability, we use the generalized likelihood ratio test (GLRT) [16], which has the advantage of replacing the unknown parameters with their maximum likelihood (ML) estimates. The asymptotic detection probability of the MIMO radar under the Neyman-Pearson criterion can be given as [10]

$$P_D = 1 - \mathfrak{F}_{\mathcal{X}_2^2(\rho)}\left(\mathfrak{F}_{\mathcal{X}_2^2}^{-1}\left(1 - P_{FA}\right)\right), \quad (11)$$

where P_{FA} denotes the probability of false alarm and $\mathfrak{F}_{\mathcal{X}_2^2(\rho)}$ is the noncentral chi-squared distribution function with two degrees of freedom (DoF) and non-centrality parameter ρ.

Further, as stated before a null space based projection method is used at the radar to attenuate the radar interference towards the cellular system. This enables the cellular system to exist in the spectrum of the MIMO radar by exploiting orthogonal spatial dimensions, that are not in use by the radar, resulting in an interference free environment from the radar towards the cellular system. Let us consider that the MIMO radar shares $\mathcal{L}$ interference channels, denoted as $\mathbf{W}_l \in \mathbb{C}^{N_{BS+UE} \times R_T}$ with the cellular system, where $N_{BS+UE} = 1 + J$ and $l = 1, \ldots, \mathcal{L}$. Accordingly, $\{\mathbf{W}_{BR}^{DL}, \mathbf{W}_j^{DL}\} \subseteq \mathbf{W}_l$, with $j = 1, \ldots, J$. Considering the availability of CSI of all $\mathbf{W}_l$ channels at the radar, singular value decomposition (SVD) can be utilized to find the null space of $\mathbf{W}_l$, which can then be used to create a null space projector matrix.

3. JOINT BEAMFORMING DESIGN

In this section, using (5), (6), and (9), we formulate the joint beamforming problem at the cellular system as

$$\textbf{(P1)} \quad \max_{\mathbf{V}^{UL}, \mathbf{V}^{DL}} \sum_{k=1}^{K} R_k^{UL} + \sum_{j=1}^{J} R_j^{dL} \quad (12)$$

$$\text{subject to} \quad (C.1) \quad \text{tr}\left\{\mathbf{V}_k^{UL}\left(\mathbf{V}_k^{UL}\right)^H\right\} \le P_k, \ \forall k; \quad (13)$$

$$(C.2) \quad \sum_{j=1}^{J} \text{tr}\left\{\mathbf{V}_j^{DL}\left(\mathbf{V}_j^{DL}\right)^H\right\} \le P_0, \quad (14)$$

$$(C.3) \quad I^{RAD} \le \Gamma \quad (15)$$

where $\mathbf{V}^{UL} = \{V_k^{UL}\}$ and $\mathbf{V}^{DL} = \{V_j^{DL}\}$. P_k in (18) is the transmit power constraint at the i-th UL user, P_0 in (19) is the total power constraint at the BS, and Γ in (20) is the upper limit of the interference allowed to be imposed on the MIMO radar. We solve the problem **(P1)** by converting the objective function into a mimimum mean squared error (MSE) and the constraints $(C.1)$-$(C.3)$ into a vector form. The vector form of $I^{RAD} = \|\boldsymbol{\iota}\|_2^2$, can be written as[3]

$$\boldsymbol{\iota} = \begin{bmatrix} \left(\mathbf{V}_k^{UL^T} \otimes \mathbf{I}_R\right) \text{vec}\left(\mathbf{G}_{RU_k}\right) \\ \sqrt{\psi}\left\lfloor \left(\left(\boldsymbol{\Xi}_\ell \mathbf{V}_k\right)^{UL^T} \otimes \mathbf{I}_R\right) \text{vec}\left(\mathbf{G}_{RU_k}\right) \right\rfloor_{\ell \in \mathcal{D}_k^{(\tilde{M})}} \\ \left(\mathbf{V}_j^{DL^T} \otimes \mathbf{I}_R\right) \text{vec}\left(\mathbf{G}_{RB}\right) \\ \sqrt{\psi}\left\lfloor \left(\left(\boldsymbol{\Xi}_\ell \mathbf{V}_j\right)^{DL^T} \otimes \mathbf{I}_R\right) \text{vec}\left(\mathbf{G}_{RB}\right) \right\rfloor_{\ell \in \mathcal{D}_j^{(\tilde{M})}} \end{bmatrix}. \quad (16)$$

Here $\mathcal{D}_k^{(\tilde{M})}(\mathcal{D}_j^{(\tilde{M})})$ represents the set $\{1 \cdots \tilde{M}_k(\tilde{M}_j)\}$ and $\boldsymbol{\Xi}_\ell$ is a square matrix with zero elements, except for the ℓ-th diagonal element, equal to 1. Using (16) and a similar method given in [17, 18], we transform the problem **(P1)** into an equivalent MSE problem as

$$\textbf{(P2)} \max_{\substack{\mathbf{V}^{UL}, \mathbf{V}^{DL}, \mathbf{U}^{UL} \\ \mathbf{U}^{DL}, \mathbf{B}^{UL}, \mathbf{B}^{DL}}} \left[\sum_{k=1}^{K}\left(-\text{tr}\{\mathbf{B}_k^{UL}\mathbf{E}_k^{UL}\} + \log|\mathbf{B}_k^{UL}| + d_k^{UL}\right) + \sum_{j=1}^{J}\left(-\text{tr}\{\mathbf{B}_j^{DL}\mathbf{E}_j^{DL}\} + \log|\mathbf{B}_j^{DL}| + d_j^{DL}\right)\right] \quad (17)$$

$$\text{subject to} \quad (C.1) \quad \|\text{vec}\left(\mathbf{V}_k^{UL}\right)\|_2^2 \le P_k, \ \forall k; \quad (18)$$

$$(C.2) \quad \sum_{j=1}^{J} \|\text{vec}\left(\mathbf{V}_j^{DL}\right)\|_2^2 \le P_0, \quad (19)$$

$$(C.3) \quad \|\boldsymbol{\iota}\|_2^2 \le \Gamma, \quad (20)$$

where $\mathbf{U}^{UL} = \{\mathbf{U}_k^{UL}, \ k = 1, \ldots, K\}$ and $\mathbf{U}^{DL} = \{\mathbf{U}_j^{DL}, \ j = 1, \ldots, J\}$ are the receive beamforming matrices, respectively. Here, $\mathbf{B}_k^{UL}$ and $\mathbf{B}_j^{UL}$ denote the weight matrix for the k-th UL and j-th DL users and $\mathbf{E}_k^{UL}$ is defined as

$$\mathbf{E}_k^{UL}(\{\mathbf{U}\}, \{\mathbf{V}\}) = \left(\left(\mathbf{U}_k^{UL}\right)^H \mathbf{H}_k^{UL}\mathbf{V}_k^{UL} - \mathbf{I}_{d_k^{UL}}\right) \times \left(\left(\mathbf{U}_k^{UL}\right)^H \mathbf{H}_k^{UL}\mathbf{V}_k^{UL} - \mathbf{I}_{d_k^{UL}}\right)^H + \left(\mathbf{U}_k^{UL}\right)^H \boldsymbol{\Sigma}_k^{UL}\mathbf{U}_k^{UL}. \quad (21)$$

Similar to $\mathbf{E}_k^{UL}$, $\mathbf{E}_j^{DL}$ can be also defined. Though the optimization problem **(P2)** is not jointly convex in $\mathbf{V} = \{\mathbf{V}^{UL}, \mathbf{V}^{DL}\}$, $\mathbf{U} = \{\mathbf{U}^{UL}, \mathbf{U}^{DL}\}$, and $\mathbf{B} = \{\mathbf{B}^{UL}, \mathbf{B}^{DL}\}$ due to coupling of optimization variables, it is convex for each of the individual variables. Therefore, we apply an alternating approach to solve the problem using standard second-order cone programming solvers [19] using interior point methods [20] with polynomial complexity. In particular,

[3] To simplify the presentation, we assume $\tilde{M} = M_0 = M_i, \ i \in \mathcal{S}^{UL}$.

for fixed $\mathbf{V}$ and $\mathbf{U}$, we solve the problem **(P2)** to find $\mathbf{B}$. Similarly, when $\mathbf{B}$ and $\mathbf{V}$ ($\mathbf{B}$ and $\mathbf{U}$) are fixed, $\mathbf{U}$($\mathbf{V}$) can be obtained.

4. SIMULATION RESULTS

In this section, we numerically investigate the coexistence of a FD MIMO cellular system and a MIMO radar based on the proposed algorithm. To model the path loss, we consider the close-in (CI) free space reference distance path loss model [21], which is a generic model that describes the large-scale propagation path loss at all relevant frequencies ($>$ 2GHz) and is given as $PL(f,d) = PL_F(f,d_0) + 10\alpha_c \log_{10}(d/d_0) + \mathcal{X}_\sigma,\ d > d_0$. Here, d_0 is a reference distance at which or closer to, the path loss inherits the characteristics of free-space path loss $PL_F(f,d_0)$, f is the carrier frequency, α_c is the path loss exponent, d is the distance between the transmitter and receiver and $\mathcal{X}_\sigma$ is the shadow fading standard deviation. Further, we consider small cell deployments under the 3GPP LTE specifications [22]. In particular, we consider a single hexagonal cell of radius 40m consisting of a BS in the center with M_0 transmit and N_0 receive antennas. $K = 2$ UL and $J = 2$ DL users (UEs) equipped with N antennas randomly distributed in the cell[4]. For simplicity, we assume $M_0 = N_0 = N = \tilde{N}$. The MIMO radar is located at the circumference of the hexagonal cell. We consider the carrier frequency of 3.6 GHz with a bandwidth of 500 MHz and $d_0 = 1$m. The thermal noise density is set at -174 dBm/Hz and the noise figures at BS and UEs are set at 13 dB and 9 dB respectively.

The estimated channel gain between the BS to kth UL UE is given by $\tilde{\mathbf{H}}_k^{UL} = \sqrt{\wp_k^{UL}}\hat{\mathbf{H}}_k^{UL}$, where $\hat{\mathbf{H}}_k^{UL}$ denotes the small scale fading following a complex Gaussian distribution with zero mean and unit variance, and $\wp_k^{UL} = 10^{(-A/10)}$, $A \in \{\mathrm{LOS}, \mathrm{NLOS}\}$ represents the large scale fading consisting of path loss and shadowing, where LOS and NLOS are calculated based on a street canyon scenario [23]. In particular α_c for LOS and NLOS are 2.0 and 3.1, respectively and the shadow fading standard deviation σ for LOS and NLOS are 2.9 dB and 8.1 dB, respectively. The channels between BS and DL UEs, between UL UEs and DL UEs, between BS and radar, and between UL UEs and radar are defined similarly. To model the SI channel, we adopt the Rician model in [14], in which the SI channel is distributed as $\tilde{\mathbf{H}}_0 \sim \mathcal{CN}\left(\sqrt{\frac{K_R}{1+K_R}}\hat{\mathbf{H}}_0, \frac{1}{1+K_R}\mathbf{I}_{N_0} \otimes \mathbf{I}_{M_0}\right)$, where K_R is the Rician factor, and $\hat{\mathbf{H}}_0$ is a deterministic matrix[5]. Unless stated otherwise, we consider, $\tilde{N} = 2$, $\psi = \upsilon = -70$ dB, CCI cancellation factor $= 0.5$, $R = 8$, $P_{FA} = 10^{-3}$, and $\Gamma = 0$ dB.

We begin by showing the detection probability of the MIMO radar with respect to radar transmit power in Fig. 2. Here, $\tilde{N} = 2$, $P_{FA} = 10^{-3}$ and $\Gamma = 0$ dB. We consider two cases here: 1) $R = 4$ (straight lines) and $R = 8$ (dashed lines). It can be seen that for fixed P_{FA}, in order to achieve a particular P_D the radar needs more power (to create the NSP waveforms to enable spectrum coexistence) than the case without spectrum sharing scenario. Also it can be seen that the radar needs more power when $R = 4$ than $R = 8$ to achieve similar performance. This is because, while the number of antennas at the cellular system (BS and UEs) are fixed, increasing the radar antennas, increases the dimension of the null space of the radar interference channel, which ensures the operability of the cellular system

[4]Although the BS has $N_0 + M_0$ antennas in total, we assume that only M_0 (N_0) antennas can be used for transmission (reception) in HD mode.

[5]Similar to [24], without loss of generality, we set $K_R = 1$ and $\tilde{\mathbf{H}}_0$ to be the matrix of all ones for all experiments.

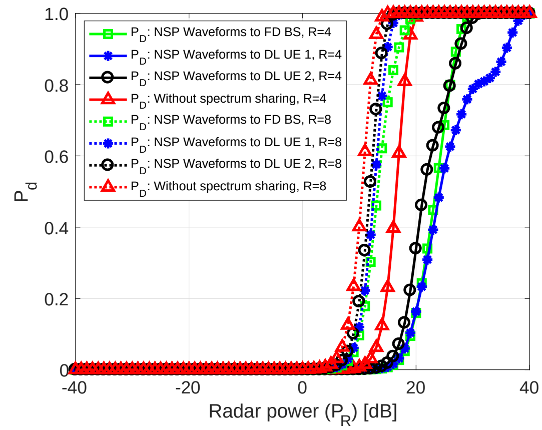

Fig. 2: Detection probability (P_D) vs radar transmit power (P_R).

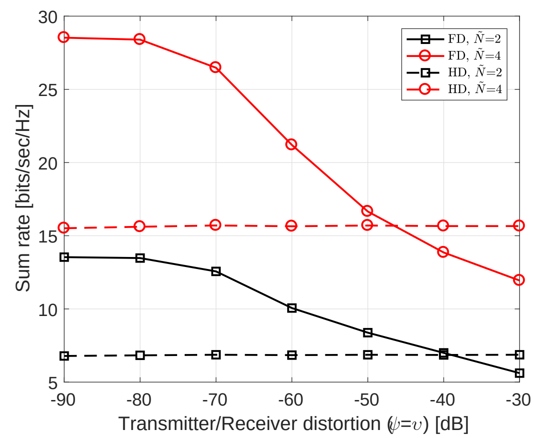

Fig. 3: Sum rate of FD cellular system vs hardware impairments.

while also yielding better detection performance of radar even with spectrum sharing. Now, we quantify the performance of the cellular system, which uses the spectrum shared by the radar in terms of sum rate (in bits/sec/Hz). In particular, Fig. 3 shows the sum rate of the cellular system as a function of transmitter/receiver (ψ/υ) distortion values for different number of antennas. It is seen from the figure that as the SI cancellation capability of the system increases, the throughput achieved by the FD system also increases. However, the performance of the HD system is invariant to ψ and υ values. In particular FD achieves around $40 - 50\%$ improvement in throughput over HD at reasonable SI cancellation of -70 dB. However, at low self-interference cancellation levels (i.e., ≤ -50 dB), the distortion is magnified with the increasing number of antennas and the HD system starts outperforming the FD system.

5. CONCLUSION

The optimization problem for joint beamforming design at a FD cellular system suffering from transmit/receive distortions was formulated to facilitate the coexistence of a cellular system and a MIMO radar under the same spectrum. Numerical results show the feasibility of spectrum sharing, albeit with certain tradeoffs in radar transmit power, detection probability and sum rate of the cellular system. In a nutshell, the designed framework provides a cornerstone and importantly, the essential understanding for successful development of future cellular systems in-conjunction with MIMO radar that can operate under same spectrum resources.

6. REFERENCES

[1] Federal Communications Commission (FCC), *Spectrum policy task force*, [Online] Available: https://transition.fcc.gov/sptf/files/IPWGFinalReport.pdf, Nov. 2002.

[2] S. Haykin, "Cognitive radio: Brain-empowered wireless communications," *IEEE J. Sel. Areas Commun.*, vol. 23, no. 2, pp. 201-220, Feb. 2005.

[3] H. Griffiths, L. Cohen, S. Watts, E. Mokole, C. Baker, M. Wicks, and S. Blunt, "Radar spectrum engineering and management: Technical and regulatory issues," *Proc. IEEE*, vol. 103, no. 1, pp. 85-102, Jan. 2015.

[4] President's Council of Advisors on Science and Technology, *Realizing the full potential of government-held spectrum to spur economic growth*, [Online] Available: https://obamawhitehouse.archives.gov/sites/default/files/microsites/ostp/pcast_spectrum_report_final_july_20_2012.pdf, Jul. 2012.

[5] National Telecommunications and Information Administration (NTIA), *An assessment of the near-term viability of accommodating wireless broadband systems in the 1675-1710 MHz, 1755-1780 MHz, 3500-3650 MHz, 4200-4220 MHz, and 4380-4400 MHz bands (Fast track report)*, [Online] Available: https://www.ntia.doc.gov/files/ntia/publications/fasttrack evaluation_11152010.pdf, Nov. 2010.

[6] Federal Communications Commission (FCC), *In the master of revision of parts 2 and 15 of the commission's rules to permit unlicensed national information infrastructure devices in the 5 GHz band*, ET Docket No. 03-122, 2006.

[7] F. Hessar and S. Roy, "Spectrum sharing between a surveillance radar and secondary Wi-Fi networks," *IEEE Trans. Aerosp. Electron. Syst.*, vol. 52, no. 3, pp. 1434-1447, Jul. 2016.

[8] S.-S. Raymond, A. Abubakari, and H.-S. Jo, "Coexistence of power-controlled cellular networks with rotating radar," *IEEE J. Sel. Areas Commun.*, vol. 34, no. 10, pp. 2605-2616, Oct. 2016.

[9] R. Saruthirathanaworakun, J. M. Peha, and L. M. Correia, "Opportunistic sharing between rotating radar and cellular," *IEEE J. Sel. Areas Commun.*, vol. 30, no. 10, pp. 1900-1910, Nov. 2012. "Synthesis of whole-body behaviors through hierarchical control of behavioral primitives," *Int. J. Humanoid Robot.*, vol. 2, no. 4, pp. 505-518, Dec. 2005.

[10] A. Khawar, A. Abdelhadi, and T. C. Clancy, "Target detection performance of spectrum sharing MIMO radar," *IEEE Sensors J.*, vol. 15, no. 9, pp. 4928-4940, Sep. 2015.

[11] B. P. Day, A. R. Margetts, D. W. Bliss, and P. Schniter, "Full-duplex bidirectional MIMO: Achievable rates under limited dynamic range," *IEEE Trans. Signal Process.*, vol. 60, no. 7, pp. 3702-3713, Jul. 2012.

[12] A. C. Cirik, S. Biswas, S. Vuppala and T. Ratnarajah, "Robust transceiver design for full duplex multiuser MIMO systems," *IEEE Wireless Commun. Lett.*, vol. 5, no. 3, pp. 260-263, Jun. 2016.

[13] A. C. Cirik, S. Biswas, S. Vuppala and T. Ratnarajah, "Beamforming design for full-duplex MIMO interference channels-QoS and energy-efficiency considerations," *IEEE Trans. Commun.*, vol. 64, no. 11, pp. 4635-4651, Nov. 2016.

[14] M. Duarte, C. Dick, and A. Sabharwal, "Experiment-driven characterization of full-duplex wireless systems," *IEEE Trans. Wireless Commun.*, vol. 11, no. 12, pp. 4296-4307, Dec. 2012.

[15] Federal Communications Commission (FCC), *Facilitating opportunities for flexible, efficient, and reliable spectrum use employing cognitive radio technologies*, [Online] Available: https://apps.fcc.gov/edocs_public/attachmatch/FCC-10-12A1_Rcd.pdf, Jan. 2010.

[16] L. Xu and J. Li, "Iterative generalized-likelihood ratio test for MIMO radar," *IEEE Trans. Signal Process.*, vol. 55, no. 6, pp. 2375-2385, Jun. 2007.

[17] Q. Shi, M. Razaviyayn, Z. Q. Luo, and C. He, "An iteratively weighted MMSE approach to distributed sum-utility maximization for a MIMO interfering broadcast channel," *IEEE Trans. Signal Process.*, vol. 59, no. 9, pp. 4331-4340, Sep. 2011.

[18] J. Shin and J. Moon, "Weighted-sum-rate-maximizing linear transceiver filters for the K-user MIMO interference channel," *IEEE Trans. Commun.*, vol. 60, no. 10, pp. 2776-2783, Oct. 2012.

[19] M. Lobo. L. Vandenberghe, S. Boyd, and H. Lebret, "Applications of second-order cone programming," *Linear algebra APP.*, vol. 284, no. 1-3, pp. 193-228, Nov. 1998.

[20] J. Sturm, "Using SeDuMi 1.02, A Matlab toolbox for optimization over symmetric cones," *Optim. Methods Softw.*, vol. 11-12, pp. 625-653, 1999.

[21] J. B. Andersen, T. S. Rappaport and S. Yoshida, "Propagation measurements and models for wireless communications channels," *IEEE Commun. Mag.*, vol. 33, no. 1, pp. 42-49, Jan 1995.

[22] 3GPP, TR 36.828, "Further enhancements to LTE time division duplex (TDD) for downlink-uplink (DL-UL) interference management and traffic adaptation (Release 11)," Jun. 2012.

[23] S. Sun et al., "Propagation Path Loss Models for 5G Urban Micro- and Macro-Cellular Scenarios," in *Proc. IEEE 83rd Vehicular Technology Conference (VTC Spring)*, May 2016, pp. 1-6.

[24] D. Nguyen, L. Tran, P. Pirinen, and M. Latva-aho, "On the spectral efficiency of full-duplex small cell wireless systems," *IEEE Trans. Wireless Commun.*, vol. 13, no. 9, pp. 4896-4910, Sep. 2014.