Phase-Only Transmit Beamforming for Spectrum Sharing Microwave Systems

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***Abstract*—This paper deals with the problem of phase-only transmit beamforming in spectrum sharing microwave systems. In contrast to sub-6 GHz schemes, general microwave systems require a large number of antennas due to its huge path loss. As a result, digital beamforming needs a large number of computational resources so that analog beamforming, which only needs a single radio-frequency chain, results the most adequate solution. These schemes are composed by a phase shifter network whose elements transmit at a certain fixed power so that the system designer shall compute the phase values for each element given a set of directions. This approach leads to non-convex quadratic problems whose convex relaxation could not rely to efficient solutions. In order to solve this, we propose a non-smooth method that behaves well in several scenarios. Numerical evaluations in different spectrum sharing scenarios which show the performance of our method are provided. Additionally, the case where the phase shifters have a single control bit is also considered. Finally, robust design considerations are also provided.**

I. INTRODUCTION

Due to the exponential increase of the traffic demands, not only the cellular wireless access shall be reconsidered but also the backhaul schemes. So far, most of the cellular base stations are connected to the backbone through a digital subscriber line (DSL) connection or, eventually, through an optical fibre link. On the contrary, rural or suburban areas base stations are generally connected via a fixed wireless radio link. These current backhaul approaches suffer from certain disadvantages.

Even though optical fibre links offer an ideally unlimited bandwidth connection, their implantation is costly and its average deployment time is large [1]. Consequently, the resulting capital expenditures (CAPEX) are high. On the other hand, wireless backhaul links despite they cannot offer an unlimited bandwidth connection, offer a substantially lower CAPEX and a very short deployment time. As a result, wireless backhaul links are of great interest in future next generation macro and small cell deployments in both high and low populated areas.

As wireless backhaul links will require a very large bandwidth, both academia and industry are proposing to shift the current microwave radio links to millimetre wave carriers such as the E band and the unlicensed 60 GHz band. Although these frequency bands offer larger available bandwidth, their path loss and atmospheric degradation effects convert the communication over these carriers a very challenging problem. This is not the case of microwave links whose reliability has been tested in the recent years in current deployments.

For both, next generation millimetre and micro wave backhaul techniques smart antenna techniques are mandatory. Indeed, in contrast to current fixed wireless links, future deployments are expected to be flexible to traffic demands so that the beam pointing reconfigurations are essential. In addition, in case a spectrum sharing scenario is considered (i.e. several communication links share time and frequency resources) interference mitigation techniques are required. This is the case of the deployments in the Ka band where certain satellite receivers could simultaneously operate [2]. In this context, beamforming techniques play a central role.

In contrast to below 3 GHz beamforming techniques, where the spatial processing is generally done in the digital domain, microwave and millimetre wave beamforming techniques require certain processing in the analog domain [3]. This is due to the large number of required radio frequency chains (dozens in microwave and hundreds in millimetre wave), whose all digital processing becomes a cumbersome task. In order to solve this, the system designer could conceive an hybrid design where an analog subsystem transforms the M transmit signals to N such as

$$M < N, \tag{1}$$

so that the digital processing complexity can be drastically reduced. This paper deals with the case where $M = 1$; this is, the spatial processing is all done in the analog domain. Unfortunately, these beamforming techniques show some additional challenges compared to the all digital case.

Analog beamforming relies on a network of phase shifters and power amplifiers whose transmit power is fixed to a certain value leading to the well-known phased array scheme. The optimization of beamforming techniques in phased array structure is an old problem whose convex approximation approach has been investigated in [4]–[6]. As a general statement, obtaining arbitrary complex array beam patterns is a computationally demanding operation which requires genetic algorithms [7] or requires inefficient approximations [6].

Furthermore, as it happens in all digital schemes, transmit beamforming suffers from a large communication overhead due to its required feedback. A limited feedback analog beamforming scheme can be found in [8] where codebook and access schemes are presented. Indeed, this joint beamforming and access approach is a key challenge also in microwave systems where the acquisition time requires that each beam (transmitter and receiver) scans in all the angle coverage. Another example of this can be found in [9]. Yet another approach is to consider during the scanning period a channel estimation (direction of arrival) based on compress sensing techniques as in [10]. In light of the aforementioned papers, in this preliminary work we consider that the transmitter has access to the interfering direction leading to the so called spatial reference beamforming.

In addition, [10] proposes a beamforming design based on the previously channel estimation. A very interesting design can be found in [11] where the channel impairments of small cells backhauls were investigated in case beam pointing algorithms are deployed. Finally, the authors in [12] proposed another beam scanning scheme assuming a transmission through both polarizations, leading to a MIMO system (i.e. a rank two channel) even though the scenario has a strong line-of-sight component.

In contrast to these works, this paper presents a general optimization framework for phase only transmit beamforming designs in spectrum sharing systems. Considering that the desired and interfering directions are known by the transmitter, we propose different efficient optimization methods. Opposed to the popular semidefinite relaxation (SDR) and a posterior Gaussian randomization technique [13], which cannot be applied in certain cases [14], the proposed schemes quickly yield to optimized beamforming solutions. These techniques have been applied before to the multi-relay channel [14] and for general quadratic constraint quadratic programs (QCQP) [15]. Furthermore, we investigate the case where the phase shifters are controlled by a single bit so that the beamforming weights can only take values of 1 or -1 leading to a substantial cost reduction.

Finally, the scheme is evaluated considering a urban and rural channel model and a robust formulation design which is adequate in case uncertainty is also presented. Remarkably, the proposed scheme is able to control the interfere power levels in different directions which is adequate for next generation spectrum sharing systems in contrast to other techniques such as [16].

The rest of the paper is organized as follows. Section II introduces the system. Section III formulates the optimization problem and the relaxation one with ideal phase quantization and one bit quantization. Section IV proposes different optimization techniques for improving the semidefinite relaxation method. Section V investigates the robust designs. Section VI illustrates the performance of the proposed techniques and Section VII concludes.

II. SYSTEM MODEL

Let us consider a base station equipped with N antennas transmitting a unit energy symbol s to a certain receiver. The received signal can be modelled as

$$y_d = \sqrt{PG_d}\mathbf{a}_d(\theta_d)^H\mathbf{w}s + n, \tag{2}$$

where $\mathbf{a}_d \in \mathbb{C}^{N\times 1}$ is the array antenna response which depends on the angle of departure (AoD) between the transmitter and the receiver ($\theta_d \in [0, 2\pi]$) and the array element structure. For the sake of simplicity, we will consider an uniform linear array (ULA) whose array antenna response is

$$[\mathbf{a}(\theta)]_i = \frac{1}{\sqrt{N}}\left(1, e^{j\frac{2\pi}{\lambda}d\sin(\theta)}, \ldots, e^{j\frac{2\pi}{\lambda}(N-1)d\sin(\theta)}\right)^T, \tag{3}$$

where d is the antenna spacing and λ the transmission wavelength. The path-loss is modelled by the real number G_d so that $0 \leq G_d \leq 1$ and P denotes the transmit power.

Vector $\mathbf{w} \in \mathbb{C}^{N\times 1}$ denotes the beamforming to be designed and n the Gaussian distributed zero mean unit variance noise term. Apart from the intended user, this paper considers that the transmission takes place in presence of $K-1$ non-intended users sharing time and spectral resources. These users are modelled by a set of AoDs ($\{\theta_k\}_{k=1}^{K-1} \in [0, 2\pi]$), leading to the following array antenna responses

$$\{\mathbf{a}(\theta_k)\}_{k=1}^{K-1}. \tag{4}$$

The corresponding received signals can be modelled as

$$y_k = \sqrt{PG_k}\mathbf{a}_k(\theta_k)^H\mathbf{w}s + n_k \quad k = 1, \ldots, K-1, \tag{5}$$

where G_k is the path loss for the k non-intended receiver and n_k the Gaussian distributed zero mean unit variance noise term.

This paper focuses on the design on $\mathbf{w}$ so that it maximizes the communication rate with the intended receiver, which maintains certain spectrum sharing constraints with respect to the non-intended receivers, as it will be described in the following sections. Additionally, in contrast to previous works, it is considered that the beamforming is performed in the analog domain by means of a set of N phase shifters. Under this context, the beamforming vector shall be constrained so that

$$|[\mathbf{w}]_i|^2 = \frac{P}{N} \quad i = 1, \ldots, K. \tag{6}$$

The following sections describe different optimizations of the beamforming vector considering different spectrum sharing scenarios so as the presence of AoDs errors. Additionally, since the array cost increases as the number of phase shifters control bits, a low complex one bit phase control scheme is also proposed.

III. PHASE ONLY BEAMFORMING OPTIMIZATION

A. *Unicast Transmission*

Let us consider the beamforming optimization of a secondary user in presence of $K-1$ primary users. Whenever the path losses and the AoDs are available, the system designer shall optimize the following problem

$$\begin{aligned}
&\underset{\mathbf{w}}{\text{maximize}} \quad |\mathbf{a}_d^H\mathbf{w}|^2\\
&\text{subject to}\\
&|\mathbf{a}_k^H\mathbf{w}|^2 \leq \epsilon_k \quad k = 1, \ldots, K-1,\\
&|[\mathbf{w}]_i|^2 = P/N \quad i = 1, \ldots, N,
\end{aligned} \tag{7}$$

where ϵ_k for $k = 1, \ldots, K-1$ denote the maximum received interference power level for the k-th primary user. The optimization problem in (7) is non-convex QCQP due to the equality constraint which imposes the constant amplitude. A relaxed optimization of (7) can be written as

$$\begin{aligned}
&\underset{\mathbf{W}}{\text{maximize}} \quad \text{Tr}\left(\mathbf{A}_d\mathbf{W}\right)\\
&\text{subject to}\\
&\text{Tr}\left(\mathbf{A}_k^H\mathbf{W}\right) \leq \epsilon_k \quad k = 1, \ldots, K-1,\\
&\text{diag}\left(\mathbf{W}\right) = P/N\mathbf{1} \quad i = 1, \ldots, N,
\end{aligned} \tag{8}$$

where $\mathbf{A}_d = \mathbf{a}_d\mathbf{a}_d^H$ and $\mathbf{A}_k = \mathbf{a}_k\mathbf{a}_k^H$ for $k = 1, \ldots, K-1$. Note that (8) does not consider the beamforming vector $\mathbf{w}$ but a matrix version $\mathbf{W}$ so that whenever $\mathbf{W}$ yields into a rank one solution, (7) is optimally solved. This relaxation is coined as semidefinite relaxation.

B. Multicast Transmission

Whenever a group of users want to receive the same content, multicast transmissions can substantially increase the spectral efficiency. With this, the transmitter must ensure that a certain symbol is decoded by all users. Under this context, the achievable rate is dictated by the user with lowest SNR, leading to the following optimization problem

$$\begin{aligned} &\underset{\mathbf{w}}{\text{maximize}} \quad \underset{d=1,\ldots,D}{\text{minimum}} \quad |\mathbf{a}_d^H\mathbf{w}|^2 \\ &\text{subject to} \\ &|\mathbf{a}_k^H\mathbf{w}|^2 \leq \epsilon_k \quad k = 1, \ldots, K-1, \\ &|[\mathbf{w}]_i|^2 = P/N \quad i = 1, \ldots, N, \end{aligned} \tag{9}$$

where we consider a set of D intended users whose array antenna responses $\{\mathbf{a}_d\}_{d=1}^D$. It is possible to approximately solve this problem similarly to the previous one by means of the SDR relaxation and adding an auxiliary variable t

$$\begin{aligned} &\underset{\mathbf{w}}{\text{maximize}} \quad t \\ &\text{subject to} \\ &\text{Tr}\left(\mathbf{A}_k^H\mathbf{W}\right) \leq \epsilon_k \quad k = 1, \ldots, K-1, \\ &\text{Tr}\left(\mathbf{A}_d^H\mathbf{W}\right) \geq t \quad d = 1, \ldots, D, \\ &\text{diag}\left(\mathbf{W}\right) = \mathbf{1} \quad i = 1, \ldots, N, \end{aligned} \tag{10}$$

IV. NON-CONVEX QCQP OPTIMIZATION

It will be shown in the simulation section that the Gaussian randomization method for obtaining rank one approximations of the previous SDR problems does not generally yield into efficient methods. This is due to the problem constraints which makes difficult to obtain a feasible point after the randomization. In order to solve this problem we propose two different non-convex optimization techniques. Namely, non-smooth optimization and successive convex approximation methods. These techniques will be described here and posteriorly evaluated and compared in different scenarios.

V. ROBUST DESIGN

Let us consider that the AoD are corrupted by either calibration errors or location inaccuracies. Mathematically, this can be modelled as

$$\hat{\mathbf{a}}_d = \boldsymbol{\delta}_d \circ \mathbf{a}_d \quad d = 1, \ldots, D, \tag{11}$$

$$\hat{\mathbf{a}}_k = \boldsymbol{\delta}_k \circ \mathbf{a}_k \quad k = 1, \ldots, K-1, \tag{12}$$

where

$$[\boldsymbol{\delta}_d]_i = u_{d,i}e^{j\phi_{d,i}}, \tag{13}$$

$$[\boldsymbol{\delta}_k]_i = u_{k,i}e^{j\phi_{k,i}}, \tag{14}$$

for $i = 1, \ldots, N$ where $u_{d,:}$ and $\phi_{k,:}$ are uniform random variable in $[0, U_d]$ and $[0, 2\pi_d]$ respectively. The operator $\circ$ denotes the Hadamard product. Under this context, the worst case robust optimization of the multicast problem can be written as

$$\begin{aligned} &\underset{\mathbf{w}}{\text{maximize}} \quad \underset{U_d, U_k, \phi_d, \phi_k}{\text{minimize}} \quad t \\ &\text{subject to} \\ &\text{Tr}\left(\left(\boldsymbol{\Phi}_k \circ \mathbf{A}_k^H\right)\mathbf{W}\right) \leq \epsilon_k \quad k = 1, \ldots, K-1, \\ &\text{Tr}\left(\left(\boldsymbol{\Phi}_d \circ \mathbf{A}_d^H\right)\mathbf{W}\right) \geq t \quad d = 1, \ldots, D, \\ &\text{diag}\left(\mathbf{W}\right) = \mathbf{1} \quad i = 1, \ldots, N, \\ &|U_d|^2 \leq \alpha_d \quad d = 1, \ldots, D, \\ &|U_k|^2 \leq \alpha_k \quad k = 1, \ldots, K-1, \\ &|\phi_d|^2 \leq \beta_d \quad d = 1, \ldots, D, \\ &|\phi_k|^2 \leq \beta_k \quad k = 1, \ldots, K-1, \end{aligned} \tag{15}$$

where

$$\boldsymbol{\Phi}_k = \boldsymbol{\delta}_k\boldsymbol{\delta}_k^H \tag{16}$$

$$\boldsymbol{\Phi}_d = \boldsymbol{\delta}_d\boldsymbol{\delta}_d^H. \tag{17}$$

The robust design only depends on the trace values of $\text{Tr}\left(\boldsymbol{\Phi}_d\right)$ and $\text{Tr}\left(\boldsymbol{\Phi}_k\right)$. An upper bound of this problem will be presented in the final paper.

VI. NUMERICAL RESULTS

The proposed phase only beamforming solutions will be evaluated and compared. The scenarios will be defined considering different backhaul current topologies and link budgets. As a preliminary result, we show the array gain of a phase only beamformer obtained via the non-smooth optimization method. We consider a ULA with 20 antenna elements with an ideal phase shifter network. Assuming a multicast transmission, the desired AoDs are located at 90 and 110 degrees whereas the interfering signals are in 60, 70, 80 and 130 degrees. The power levels for the interfering directions shall be lower than -20 dBs.

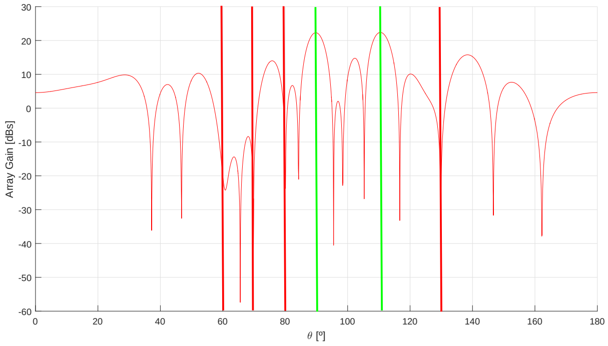

Fig. 1. Array factor obtained via non-smooth optimization. It consists of a multicast transmission whose desired AoD are in 90 and 110 degrees. The interfering directions are in 60, 70, 80 and 130 degrees. The power levels for the interfering directions shall be lower than -20 dBs.

After 6 iterations the method is able to obtain a phase only beamformer able to meet these requirements. Remarkably, SDR after 10^4 randomizations was unable to obtain a feasible solution.

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