Energy Efficient Transmitter with Low Resolution DACs for Massive MIMO with Partially Connected Hybrid Architecture

Evangelos Vlachos, Aryan Kaushik and John Thompson
Institute for Digital Communications, The University of Edinburgh, United Kingdom.
Email: {E.Vlachos, A.Kaushik, J.S.Thompson}@ed.ac.uk

***Abstract*—Millimeter wave (mmWave) multiple-input multiple-output (MIMO) systems have recently been proposed to meet the needs of the future wireless communication standards. The efficient use of low resolution digital-to-analog converters (DACs) and hybrid architecture could significantly reduce the high power consumption associated with the mmWave MIMO system components. This paper designs an energy efficient transmitter with low resolution DACs for mmWave massive MIMO systems. An optimization problem is formulated and solved to find the optimal number of radio-frequency (RF) chains to be used at the transmitter to minimize the power consumption. This problem is constrained by the information loss which introduces the reduction of the number of the RF chains, expressed in terms of the system capacity. Rate and energy performance are compared with different beamforming techniques and architectures for various DAC resolutions.**

***Index Terms*—energy efficient transmitter, low resolution digital-to-analog converter (DAC), millimeter wave (mmWave), massive MIMO, hybrid analog/digital beamforming.**

I. Introduction

Millimeter Wave (mmWave) appears to be a promising technology to meet the needs of the fifth generation (5G) wireless communication systems [1][2]. The high bandwidths for mmWave communication must be traded off against high path loss which can be compensated using large antenna arrays. A large number of antenna elements and the high bandwidth make it hard to use many digital-to-analog/analog-to-digital converter (DAC/ADC) which are power hungry components, connecting baseband components to the antennas through radio frequency (RF) chains and phase shifters [3]. ADCs have relatively higher sampling rate in high frequency systems than at the microwave frequencies, and employing high speed converters increases the power consumption and the cost significantly [4], [5]. Implementing low resolution ADCs at receivers such as in 1-bit mmWave MIMO systems improves the power metric of the system [3]. Since amplifiers consume most of the power at the transmitter, employing low resolution DACs at transmitters can help reducing the power consumption, hence enhancing energy efficiency of the system.

Hybrid transceiver architectures have been proposed to enable mmWave massive MIMO systems [6]. The conventional digital beamforming system can provide high rates but the energy consumption becomes unacceptable when the same number of RF chains as the number of antennas are used for the transceiver system. A hybrid beamforming system uses a smaller number of RF chains and can be implemented to provide approximately the same rate performance as a digital beamfoming system but has better energy efficiency [7]. There can be two ways to implement a hybrid architecture: a fully-connected (FC) architecture and a partially-connected (PC) architecture. A FC architecture connects all the transmitter antennas to each RF chain whereas a PC architecture connects only a subset of transmitter antennas requiring only fewer RF chains. The use of a PC architecture at the transceiver can further reduce the power consumption [8]. Reference [9] makes use of the switches and phase shifters to execute analog beamforming for a hybrid model, and then the spectral and energy efficiency are investigated for the system. Reference [10] uses low resolution DACs which we can implement to reduce the power consumption for a hybrid model. Using switches and low-resolution DACs can have significant improvement in lowering the power demands.

This paper proposes an energy efficient design for the transmitter of a mmWave-based massive MIMO system which makes use of low-resolution DACs and the hybrid architecture. We propose a novel formulation for an optimization problem which finds the optimal number of RF chains to be used at the transmitter, so as the system capacity to be lower bounded. The proposed design makes use of switches, which are used to keep a connection between the baseband precoder and a DAC; if that connection is *active* that particular DAC links the basedband precoder with a RF chain.

We follow the following notation throughout this paper: $\mathbf{A}$, $\mathbf{a}$, and a stand for a matrix, a vector, and a scalar, respectively. The complex conjugate transpose, and transpose of $\mathbf{A}$ are denoted as $\mathbf{A}^H$ and $\mathbf{A}^T$; $\|\mathbf{A}\|_F$, tr($\mathbf{A}$), and det($\mathbf{A}$) represent the Frobenius norm, trace, and determinant of $\mathbf{A}$, respectively; diag($\mathbf{A}$) generates a vector by the diagonal elements of $\mathbf{A}$; $\mathbf{I}_N$ represents $N \times N$ identity matrix; $\mathcal{CN}(\mathbf{a}; \mathbf{A})$ denotes a complex Gaussian vector having mean $\mathbf{a}$ and covariance matrix $\mathbf{A}$; the expectation of a complex variable is denoted as $\mathcal{E}\{\cdot\}$; $[\mathbf{A}]_k$ denotes the k-th column of matrix $\mathbf{A}$ while $[\mathbf{A}]_{kl}$ the matrix entry at the k-th row and l-th column.

II. Hybrid mmWave MIMO System

A. Millimeter wave channel model

The mmWave massive MIMO systems carry high free-space path loss and large tightly-packed antenna arrays which restrict

978-1-5386-6355-4/18/$31.00 ©2018 IEEE

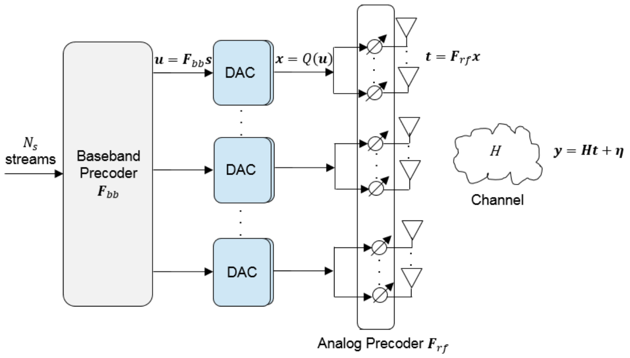

Fig. 1. Block diagram of a mmWave single-user transmitter with the PC architecture.

the use of the fading channel models used in traditional MIMO systems. The mmWave channels can then be modeled by a narrowband clustered channel model with N_{cl} clusters and N_{ray} propagation paths in each cluster. Considering a single-user mmWave system with N_t antennas at the transmitter, transmitting N_s data streams, the mmWave channel matrix can be presented as follows:

$$\mathbf{H} = \gamma \sum_{i=1}^{N_{cl}} \sum_{l=1}^{N_{ray}} \alpha_{il} \mathbf{a}_r(\phi_{il}^r, \theta_{il}^r) \mathbf{a}_t(\phi_{il}^t, \theta_{il}^t)^H \tag{1}$$

α_{il} in (1) is the gain term for the l^{th} ray in the i^{th} cluster and follows i.i.d. $\mathcal{CN}(0, \sigma_{\alpha,i}^2)$. The $\sigma_{\alpha,i}^2$ for the gain represents the average power of the i^{th} cluster where $\sum_{i=1}^{N_{cl}} \sigma_{\alpha,i}^2 = \gamma$, γ being the the normalization factor. Further, $\mathbf{a}_t(\phi_{il}^t, \theta_{il}^t)$ and $\mathbf{a}_r(\phi_{il}^r, \theta_{il}^r)$ represent the normalized transmit and receive array response vectors, where ϕ_{il}^t and θ_{il}^t denote the azimuth and elevation angles of departure, respectively, and ϕ_{il}^r and θ_{il}^r denote the azimuth and elevation angles of arrival, respectively. We require only the transmitter associated terms only for our analysis. We model the antenna elements at the transmitter as ideal sectored elements [11] and evaluate the associated gains over the ideal sectors. We assume the transmit antenna gains to be unity over the sectors defined by $\phi_{il}^t \in [\phi_{min}^t, \phi_{max}^t]$ and $\theta_{il}^t \in [\theta_{min}^t, \theta_{max}^t]$. Although we use uniform linear array antennas for simplification, yet our approach can easily be extended to other array geometries such as rectangular array and circular array. For λ signal wavelength and d inter-element spacing, and considering azimuth angles of departure, the transmit array response vector can be written as follows [12]:

$$\mathbf{a}_t(\phi_{il}^t) = \frac{1}{\sqrt{N_t}} \left[1, e^{j\frac{2\pi}{\lambda} d \sin(\phi_{il}^t)}, ..., e^{j(N_t-1)\frac{2\pi}{\lambda} d \sin(\phi_{il}^t)}\right]^T \tag{2}$$

and we can write similarly for the receive array response vector.

B. System model

Let N_t^{rf} denote the number of RF chains at the transmitter, and we should follow the condition $N_s \leq N_t^{rf} \leq N_t$. $\mathbf{F}_{bb}$ with dimension $N_t^{rf} \times N_s$ and $\mathbf{F}_{rf}$ with dimension $N_t \times N_t^{rf}$ denote the baseband precoder and the RF precoder, respectively. After the RF/ analog precoding, each phase shifter is connected to each antenna element of a sub-array and each data stream is transmitted by that corresponding sub-array. There are number of sub-arrays and the sum of all the antenna elements from each sub-array gives us N_t. Fig. 1 shows the system setup of a mmWave single-user transmitter with the PC architecture. All elements of the analog precoder $\mathbf{F}_{rf}$ are constrained to have equal norm. The transmit power constraint is satisfied through the precoder matrices with condition $\|\mathbf{F}_{rf}\mathbf{F}_{bb}\|_F^2 = N_s$. The vector $\mathbf{x}$ denotes the transmit signal which is equal to $\mathbf{F}_{rf}\mathbf{F}_{bb}\mathbf{s}$. $\mathbf{s}$ with dimension $N_s \times 1$ is the symbol vector such that $\mathcal{E}\{\mathbf{ss}^*\} = \frac{1}{N_s}\mathbf{I}_{N_s}$.

We assume a linear model approximation for the quantization noise of the DACs [13]. Given that $Q(\cdot)$ denotes a uniform scalar quantizer then for the i-th RF chain we have,

$$Q(u_i) \approx \sqrt{1 - \rho_{b_i}}\, u + \epsilon_i \tag{3}$$

where $\rho_{b_i} = \frac{\pi\sqrt{3}}{2} 2^{-2b_i}$ is the quantization distortion parameter for bit resolution equal to b_i [13]; ϵ_i is the quantization noise which is uncorrelated with u_i and $\epsilon_i \sim \mathcal{N}(0, \sigma_\epsilon^2)$. Extending our formulation to the MIMO case we have,

$$Q(\mathbf{u}) \approx \mathbf{\Delta u} + \boldsymbol{\epsilon} \tag{4}$$

where $\mathbf{\Delta}$ is a diagonal matrix with values depending on the DAC resolution of each RF chain, while $\boldsymbol{\epsilon} \in \mathcal{CN}(\mathbf{0}, \sigma_\epsilon^2\mathbf{I})$. Let us define the linear approximation of the quantizer as $\tilde{\mathbf{x}} \in \mathcal{C}^{N_t^{rf} \times 1}$, and

$$\tilde{\mathbf{x}} = \mathbf{\Delta u} + \boldsymbol{\epsilon}. \tag{5}$$

This leads us to the following transmitted signal, $\tilde{\mathbf{t}} \in \mathcal{C}^{N_t \times 1}$:

$$\tilde{\mathbf{t}} = \mathbf{F}_{rf}\tilde{\mathbf{x}} = \mathbf{F}_{rf}\Delta\mathbf{u} + \mathbf{F}_{rf}\boldsymbol{\epsilon} = \mathbf{F}_{rf}\Delta\mathbf{F}_{bb}\mathbf{s} + \mathbf{F}_{rf}\boldsymbol{\epsilon}. \tag{6}$$

The received signal can now be expressed as follows:

$$\tilde{\mathbf{y}} = \mathbf{H}\tilde{\mathbf{t}} + \mathbf{n} = \underbrace{\mathbf{HF}_{rf}\Delta}_{\mathbf{H}_e \in \mathbb{C}^{N_r \times N_t^{rf}}} \mathbf{F}_{bb}\mathbf{s} + \underbrace{\mathbf{HF}_{rf}\boldsymbol{\epsilon} + \mathbf{n}}_{\boldsymbol{\eta} \in \mathbb{C}^{N_r \times 1}} \tag{7}$$

where $\mathbf{HF}_{rf}\Delta$ can be denoted as the effective channel $\mathbf{H}_e$ including analog transmit precoding and quantization; $\boldsymbol{\eta}$ as the combined Gaussian and quantization noises with $\boldsymbol{\eta} \sim \mathcal{CN}(\mathbf{0}, \mathbf{C}_\eta)$, where $\mathbf{C}_\eta$ is the noise covariance matrix with $\mathbf{C}_\eta = \mathbf{HF}_{rf}\mathbf{F}_{rf}^H\mathbf{H}^H\sigma_\epsilon^2 + \sigma_n^2\mathbf{I}_{N_r}$. Based on the central limit theorem, when the size of antennas increase the noise covariance matrix can be approximated by $\mathbf{C}_\eta \approx \sigma_\epsilon^2 \mathrm{diag}(\mathbf{HH}^H) + \sigma_n^2\mathbf{I}_{N_r}$.

III. Proposed method for the transmitter system

In this section, we describe a novel technique which optimally selects the minimum required number of *active RF chains* N_t^{rf} which satisfy a predefined bound on the information loss. Fig. 2 shows the block diagram of the proposed transmitter design for a mmWave single-user massive MIMO system with a hybrid PC precoder architecture. In this architecture, we introduce N_t switches between tha basedband

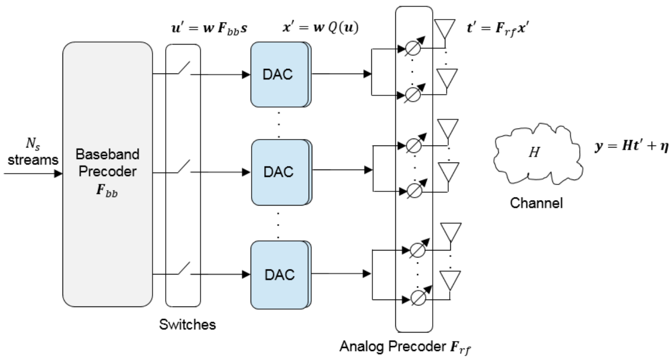

Fig. 2. Block diagram for the proposed design for a mmWave single-user transmitter with the PC architecture.

precoder and the DACs. When a switch is closed then there is a connection between the baseband precoder and a DAC, i.e., the RF chain is *active*. Note that in the proposed design, there are N_t chains in total but only $N_t^{rf} < N_t$ of them are active during the communication phase. Specifically, we consider a two phase procedure, where at the first phase the active RF chains are identified while at the second they are used for communication. In this work, we investigate only the first phase, in order to identify the best N_t^{rf} RF chains from the N_t available.

Let us consider the input of the MIMO communication channel $\tilde{\mathbf{x}}$ and the output $\tilde{\mathbf{y}}$ as discrete random variables with transition probability $p(\tilde{\mathbf{y}}|\tilde{\mathbf{x}})$. The capacity of this channel is given by the maximum of the mutual information $I(\tilde{\mathbf{x}};\tilde{\mathbf{y}})$, i.e., $C = \max_{\tilde{\mathbf{x}}} I(\tilde{\mathbf{x}};\tilde{\mathbf{y}})$. It has been shown that in the case of memoryless channels with very noisy inputs, the capacity is closely related to the Fisher information matrix (FIM) $\mathbf{J} = \mathcal{E}\{\nabla_{\tilde{\mathbf{x}}} p(\tilde{\mathbf{y}}|\tilde{\mathbf{x}}) \nabla_{\tilde{\mathbf{x}}} p(\tilde{\mathbf{y}}|\tilde{\mathbf{x}})\}$ [14]. Specifically, the following approximation holds at low signal-to-noise ratio [15]:

$$I(\tilde{\mathbf{x}};\tilde{\mathbf{y}}) \approx \frac{1}{2}\mathrm{tr}(\mathbf{J}\mathbf{C}_x) \tag{8}$$

where $\mathbf{C}_x = \mathcal{E}\{\tilde{\mathbf{x}}\tilde{\mathbf{x}}^H\} = \sigma_u^2 \boldsymbol{\Delta}\boldsymbol{\Delta}^H + \sigma_\epsilon^2 \mathbf{I}$ is the covariance matrix of the quantized signal $\tilde{\mathbf{x}}$.

To proceed, let us express the FIM considering the system model described by (7). To simplify the problem, we apply a whitening filter $\mathbf{C}_\eta^{-1} = \left(\sigma_\epsilon^2 \mathrm{diag}(\mathbf{H}\mathbf{H}^H) + \sigma_n^2 \mathbf{I}_{N_r}\right)^{-1}$ to the output $\tilde{\mathbf{y}}$, and we get the following system model:

$$\tilde{\mathbf{y}}_w = \mathbf{C}_\eta^{-1}\tilde{\mathbf{y}} = \mathbf{C}_\eta^{-1}\mathbf{H}_e\mathbf{s} + \boldsymbol{\eta}_w \tag{9}$$

where $\boldsymbol{\eta}_w \sim \mathcal{N}(\mathbf{0}, \mathbf{I}_{N_r})$. Note that to obtain (9) we assume perfect knowledge of the channel information state at the transmitter. The FIM can now be given as the sum of K independent measurements:

$$\mathbf{J} = \sum_{k=1}^{K} [\mathbf{C}_\eta^{-1}\mathbf{H}_e]_k [\mathbf{C}_\eta^{-1}\mathbf{H}_e]_k^H \tag{10}$$

where in this case $K = N_t$. This is a key expression for our formulation, since it expresses the contribution of each measurement to the FIM separately. The additive property of the FIM allows us to introduce the selection matrix $\mathbf{W} \in \mathbb{Z}^{N_t \times N_t}$ to obtain

$$\mathbf{J}_{\mathbf{W}} = \sum_{k=1}^{K} [\mathbf{W}]_{kk} [\mathbf{C}_\eta^{-1}\mathbf{H}_e]_k [\mathbf{C}_\eta^{-1}\mathbf{H}_e]_k^H \tag{11}$$

where $[\mathbf{W}]_{kk} \in \{0, 1\}$ and $[\mathbf{W}]_{kl} = 0$ for $k \neq l$. Therefore, a closed switch implies that its measurement will be taken into account via (11), while in the other case, an open switch implies that its measurement will not contribute into the information loss cost function of (11).

To minimize the energy consumption of the DACs, the number of closed switches (active RF chains) should be minimum value possible; equivalently the number of non-zero entries of $\mathbf{W}$ has to be minimized. Formally, the optimization problem can be expressed as:

$$\min_{\mathbf{W}} \|\mathrm{vec}(\mathbf{W})\|_0 \text{ subject to } \mathrm{tr}(\mathbf{J}_{\mathbf{W}}\mathbf{C}_x) > \lambda R \tag{12}$$

where $\|\cdot\|_0$ is the ℓ_0-quasi norm and R is the highest reliable rate of communication which is expressed as:

$$R = B \log\det(\mathbf{I} + \mathbf{C}_\eta^{-1}\mathbf{H}_e\mathbf{H}_e^H(\mathbf{C}_\eta^{-1})^H) \tag{13}$$

where B is the transmission bandwidth. The parameter λ takes values in $[0, 1]$, where large values of λ will result in a higher number of active RF chains, while for $\lambda \rightarrow 0$ a sparser solution with fewer RF chains will be obtained. Note that, this formulation is closely related with the problem of best subset selection, where from N_t independent variables we have to select the best $N_t^{rf} < N_t$, using K measurement values [16].

Based on (8), the capacity of the MIMO channel, which is the maximization of the mutual information, can be approximated by the maximization of the trace of the matrix $\mathbf{J}_{\mathbf{W}}\mathbf{C}_x$, which is equal to the sum of its eigenvalues. To simplify the problem, instead of maximizing the sum of the eigenvalues of $\mathbf{J}_{\mathbf{W}}$, we can put a lower bound on each eigenvalue of the matrix $\mathbf{J}_{\mathbf{W}}$. Specifically, based on linear matrix inequalities we have the following:

$$\mathbf{J}_{\mathbf{W}} - \lambda R\mathbf{I} \succeq \mathbf{0}_{N_r}. \tag{14}$$

Moreover, to make the problem (12) computationally tractable, the ℓ_0 norm should be replaced by the ℓ_1 norm. The resulting convex optimization problem can be expressed as:

$$\min_{\mathbf{W}} \|\mathrm{vec}(\mathbf{W})\|_1 \tag{15}$$

$$\text{subject to } \sum_{k=1}^{K} [\mathbf{W}]_{kk} [\mathbf{C}_\eta^{-1}\mathbf{H}_e]_k [\mathbf{C}_\eta^{-1}\mathbf{H}_e]_k^H > \lambda R, \tag{16}$$

$$0 \leq [\mathbf{W}]_{kk} \leq 1, \forall k = 1, 2, \ldots, K. \tag{17}$$

The problem in (15) is a standard semi-definite one which can be efficiently solved in polynomial time using interior-point methods (e.g., CVX [18]).

Remark: In order to ensure that (15) can be solved, the number of the measurements K has to be larger than the number of the system parameter N_t^{rf}, thus FIM $\mathbf{J}$ has full rank. To ensure this property, we replace $\mathbf{H}_e$ with a randomly sub-sampled version of it, i.e., $\tilde{\mathbf{H}}_e = [\mathbf{H}_e]_{|\Omega} \in \mathbb{C}^{|\Omega| \times N_t^{rf}}$ where Ω is the subset of randomly selected indices from the set $\{1, 2, \ldots, N_t^{rf}\}$ with $|\Omega| < N_t^{rf}$.

IV. Simulation Results

In this section, we evaluate the performance of the proposed algorithm through simulation results. We consider the case of a point-to-point mmWave-based massive MIMO system with hybrid analog/digital beamforming. The transmitter and the receiver have $N_t = N_r = 64$ antennas. We assume that the source signal $\mathbf{s}(t) \times \mathbb{C}^{N_s \times 1}, t = 1, \ldots, T$ is normalized with $\mathcal{E}\{\|\mathbf{s}(t)\|^2\} = 1$ and $N_s = 4$. The normalization parameter of the channel in equation (1) is set to $\gamma = \frac{1}{\|\mathbf{H}\|}$. To solve the constrained optimization problem of (15) we have employed CVX [18] routines in MATLAB. The results have been averaged after running 100 Monte-Carlo simulations.

Concerning the quantization model, we assume that all DACs have the same number of bits $b_i = b$ for $i = 1, 2, \ldots, N_t^{rf}$, thus the quantization distortion parameter is the same for all DACs with $\rho_{b_i} = \rho_b$, while the variance of the quantization noise was set to $\sigma_\epsilon^2 = \rho_b^2$. The analog and digital precoding matrices $\mathbf{F}_{rf}, \mathbf{F}_{bb}$ are computed based on the solution of the following

$$(\mathbf{F}_{rf}, \mathbf{F}_{bb}) = \arg \min_{\mathbf{F}_{rf}, \mathbf{F}_{bb}} \|\mathbf{F}_{opt} - \mathbf{F}_{bb}\mathbf{F}_{rf}\|_F \tag{18}$$

$$\text{s.t. } \mathbf{F}_{rf} \in \mathcal{F}_{rf}, \tag{19}$$

$$\|\mathbf{F}_{rf}\mathbf{F}_{bb}\|_F^2 = N_s \tag{20}$$

where $\mathcal{F}_{rf}$ is the set of feasible RF precoders which corresponds to a hybrid architecture based on the phase shifters [3], [7] and $\mathbf{F}_{opt}$ is obtained by the N_s columns of the right singular matrix.

In Figs. 3 and 4 we show the spectral efficiency with respect to (w.r.t.) the signal-to-noise ratio (SNR), which is given by the expression

$$S = \log\det(\mathbf{I} + \mathbf{H}\mathbf{F}\mathbf{F}^H\mathbf{H}^H) \text{ bits/sec/Hz}$$

for the case of a fully-connected (FC) and a partially-connected (PC) system, for both 1 and 3 bits quantization cases. We compare three system setups depending on the number of active RF chains: i) an equal number of the antennas and RF chains, i.e., $N_t = N_t^{rf} = 64$, ii) a smaller number of randomly chosen RF chains compared to the number of the transmitter antennas, i.e., $N_t > N_t^{rf} = 32$ and iii) also a smaller number of RF chains but optimally selected based on the solution of (15).

For the FC case, depicted in Fig. 3, we observe that the proposed technique chooses fewer than 32 RF chains while it achieves better performance in terms of spectral efficiency. This is true for both DAC resolution cases, 1 and 3 bits. For the PC case in Fig. 4, we observe similar behaviour, although the performance range is lower due to the partial connection of the RF chains and the antennas.

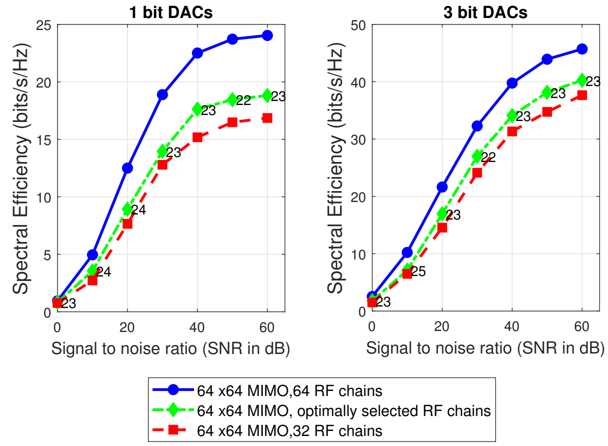

Fig. 3. Spectral efficiency curves w.r.t. the SNR for the FC case with $\lambda = 0.1$.

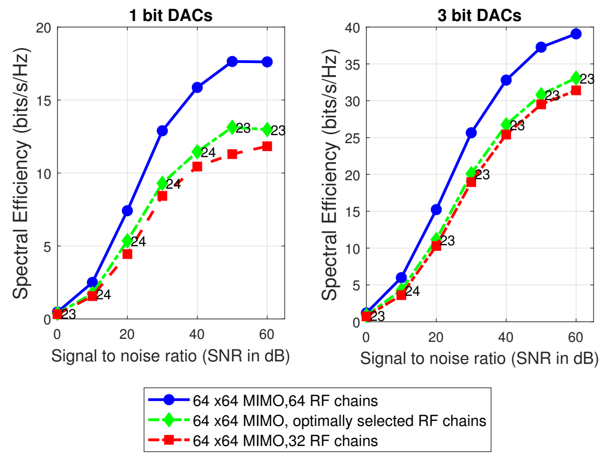

Fig. 4. Spectral efficiency curves w.r.t. the SNR for the PC case with $\lambda = 0.1$.

Considering the energy efficiency of the proposed technique, it can be defined as the ratio between spectral efficiency R and the power consumption P, i.e.,

$$\varepsilon \triangleq \frac{S}{P} \text{bits/Hz/J} \tag{21}$$

where $S \triangleq \frac{R}{B}$bits/s/Hz. We consider that the total power consumption, P, is the sum of the power consumed for transmission and baseband processing and analog processing entities, i.e.,

$$P = \text{tr}(\mathbf{F}\mathbf{F}^*) + N_t^{rf} P_{rf} 2^{2b} + N_t N_t^{rf} P_{ps} \tag{22}$$

where P_{rf} and P_{ps} represent the power per RF chain and the power per phase shifter, respectively, while 2^{2b} represents the DAC resolution power factor. The energy consumed by the RF chains is a major concern leading to a high value of P_{rf} with a substantial increase in energy for each RF chain. In the FC hybrid precoder structure, one can consider that N_{ps} is equal to $N_t^{rf} N_t$ while for the PC hybrid architecture, it should be N_t [19], [20].

In Figs. 5 and 6 we show the energy efficiency ϵ w.r.t. the SNR for FC and PC cases, for both 1 bit and 3 bits cases. Comparing the three curves in each case, we can verify that the

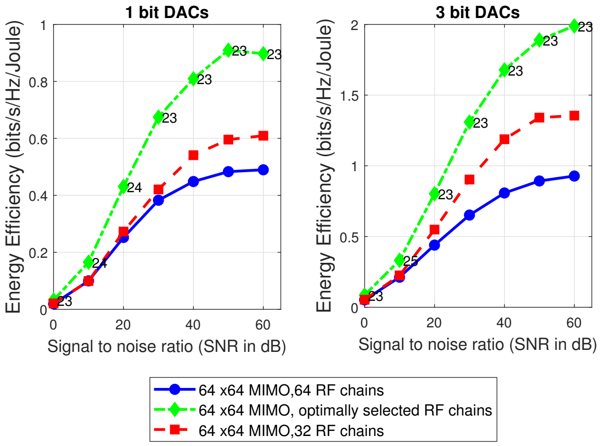

Fig. 5. Energy efficiency curves w.r.t. the SNR for the FC case with $\lambda = 0.1$.

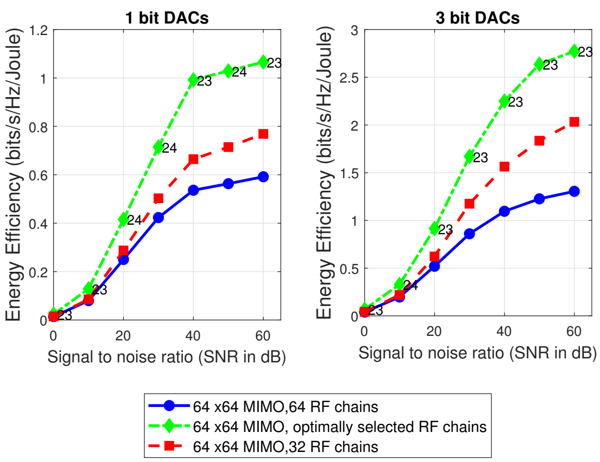

Fig. 6. Energy efficiency curves w.r.t. the SNR for the PC case with $\lambda = 0.1$.

proposed technique provides higher efficiency, minimizing the power consumption. Especially for the case of PC architecture, the proposed technique achieves 30% higher energy efficiency than the case of 32 randomly chosen RF chains.

Table I provides an insight into how the power consumption can be reduced with the use of an optimal number of RF chains at the transmitter. For the case of SNR = 10 dB and 1 bit DAC resolution, the comparison of power consumption for each entity and the total power is shown, where the proposed technique consumes less power.

V. Conclusion

This paper designs an energy efficient transmitter for a mmWave MIMO system with a partially-connected (PC) hybrid architecture and low-resolution DACs. Through a novel optimization problem formulation, the lowest number of RF chains and DACs at the transmitter are obtained in an optimal manner, so that the system capacity meets a minimum desired rate. The proposed transmitter exhibits higher spectral efficiency than the conventional one, where the same number of active RF chains is randomly chosen. The energy efficiency obtained from the proposed technique is also increased by up to 30% compared to the conventional case.

TABLE I
Power consumption (in mWatt) for 10 dB SNR and 1-bit resolution.

Component	Per entity	$N_t^{rf} = 32$	$N_t^{rf} = 23$
RF chains	0.1	3.2	2.3
Phase shifters	0.01	0.32	0.23

Acknowledgment

The authors gratefully acknowledge funding of this work by EPSRC grant EP/P000703/1.

References

[1] Z. Pi, and F. Khan, “An introduction to millimeter-wave mobile broadband systems”, *IEEE Commun. Mag.*, vol. 49, no. 6, pp. 101-107, June 2011.
[2] J. G. Andrews et al., “What will 5G be”, *IEEE Journ. Sel. Areas Commun.*, vol. 32, no. 6, pp. 1065-1082, June 2014.
[3] R. W. Heath et al.,“An overview of signal processing techniques for millimeter wave MIMO systems”, *IEEE J. Sel. Topics Signal Process.*, vol. 10, no. 3, pp. 436-453, Apr. 2016.
[4] B. Le et al.,“Analog-to-digital converters”, *IEEE Signal Process. Mag.*, vol. 22, no. 6, pp. 69-77, Nov. 2005.
[5] R. Walden,“Analog-to-digital converter survey and analysis”, *IEEE J. Sel. Areas Commun.*, vol. 17, no. 4, pp. 539-550, Apr. 1999.
[6] R. W. Heath,“Millimeter wave: the future of commercial wireless systems”, in *Proc. 2016 IEEE Compound Integrated Circuit Symp. (CSICS)*, Austin, TX, USA, pp. 1-4, Oct. 2016.
[7] A. Kaushik et al.,“Sparse hybrid precoding and combining in millimeter wave MIMO systems”, in *Proc. IET Radio Prop. Tech. 5G*, Durham, UK, pp. 1-7, Oct. 2016.
[8] C. G. Tsinos, S. Maleki, S. Chatzinotas and B. Ottersten, “On the energy-efficiency of hybrid analog-digital transceivers for single- and multi-carrier large antenna array systems”, in *IEEE J. Sel. Areas Commun.*, vol. 35, no. 9, pp. 1980-1995, Sept. 2017.
[9] R. Mendez-Rial et al., “Hybrid MIMO architectures for millimeter wave communications: Phase shifters or switches?”, *IEEE Access*, vol. 4, pp. 247-267, Jan. 2016.
[10] A. Mezghani et al., “Transmit processing with low resolution D/A-converters”, in *Proc. Int. Conf. Electronics, Circuits, and Systems*, Tunisia, pp. 683-686, Dec. 2009.
[11] S. Singh, R. Mudumbai, and U. Madhow, “Interference analysis for highly directional 60-GHz mesh networks: The case for rethinking medium access control”, *IEEE/ACM Trans. Netw.*, vol. 19, no. 5, pp. 1513-1527, Oct. 2011.
[12] C. Balanis, *Antenna Theory*, Wiley, 1997.
[13] O. Orhan, E. Erkip and S. Rangan, “Low power analog-to-digital conversion in millimeter wave systems: Impact of resolution and bandwidth on performance”, in *Inform. Theory Appl. Workshop, ITA*, pp. 191-198, Oct. 2015.
[14] V. V. Prelov and E. C. van der Meulen, “An asymptotic expression for the information and capacity of a multidimensional channel with weak input signals”, in *IEEE Trans. Inform. Theory*, vol. 39, no. 5, pp. 1728-1735, Sept. 1993.
[15] Lubomir Kostal, “Information capacity in the weak-signal approximation”, in *Physical Review*, ed. 82, Aug. 2010.
[16] S. P. Chepuri and G. Leus, “Sparsity-promoting sensor selection for non-linear measurement models”, in *IEEE Trans. Signal Process.*, vol. 63, no. 3, pp. 684-698, Feb. 2015.
[17] M. Grant and S. Boyd, “Graph implementations for nonsmooth convex programs”, in *Recent Adv. Learning and Control*, Springer-Verlag Ltd., pp. 95-110, 2008.
[18] X. Gao et al., “Energy-efficient hybrid analog and digital precoding for mmWave MIMO systems with large antenna arrays”, *IEEE J. Sel. Areas Commun.*, vol. 34, no. 4, pp. 1-12, Mar. 2016.
[19] X. Yu et al., “Alternating minimization algorithms for hybrid precoding in millimeter wave MIMO systems”, *IEEE J. Sel. Topics Signal Process.*, vol. 10, no. 3, pp. 485-500, Feb. 2016.
[20] A. K. Fletcher, S. Rangan, V. K. Goyal, and K. Ramchandran, “Robust predictive quantization: Analysis and design via convex optimization”, *IEEE Journal of Selected Topics in Signal Processing*, vol. 1, no. 4, pp. 618632, 2007