

Design of Pre-coding and Combining in Hybrid Analog-Digital Massive MIMO with Phase Noise

Roberto Corvaja*, Ana García Armada⁺, Miguel Ángel Vázquez[†], Ana Pérez-Neira[‡]

*Department of Information Engineering, University of Padova (e-mail: corvaja@dei.unipd.it)

⁺Department of Signal Theory and Communications, Universidad Carlos III de Madrid (e-mail: agarcia@tsc.uc3m.es)

[†]Centre Tecnologic de Telecomunicacions de Catalunya (CTTC/CERCA) (e-mail: mavazquez@cttc.cat)

[‡]Department of Signal Theory and Communications, Universitat Politècnica de Catalunya (e-mail: aperez@cttc.cat)

Abstract—The design of massive MIMO, especially at millimeter waves, requires a trade-off between cost and power consumption, balancing the complexity and the performance in terms of achievable rate. A recent trend in the design is to split the pre-coding at the transmitter and the combining at the receiver into a digital and analog part, with hybrid analog-digital schemes. In this paper the effect of phase noise is considered in the design of different hybrid analog-digital alternatives to implement massive MIMO, in particular at very high frequencies. Its effect depending on the number of RF chains, oscillators, and groups of antennas is analyzed providing some insights for the system design. In order to limit the penalty introduced by the phase noise to values below 6 dB, with a number of antennas around 64, the value of phase noise increment variance should be limited below 0.005. This limit is slightly lower in a simplified architecture with more blocks driven by independent oscillators.

Index Terms—Massive MIMO, hybrid pre-coding and combining MIMO, phase noise.

I. INTRODUCTION

The trend towards the deployment of massive multiple input – multiple output (MIMO) systems to increase the capacity in the evolution of 5G mobile networks and beyond requires the use of millimeter waves (mmW), where the available bandwidth is much larger and the size of the devices can be reduced. However, the shift towards mmW determines a major complexity and a serious increase of the power consumption, especially of devices such as analog-to-digital (ADC) and digital-to-analog (DAC) converters. In particular, all-digital architectures become unfeasible for the extremely high sampling rate required, so that hybrid analog-digital solutions are a need. In [1] considerations are made on the choice between all-digital and hybrid solutions. Also, to implement a massive number of antennas an important concern is the power consumption of the radio-frequency (RF) chains, consisting of ADC or DAC, up- or down-conversion with RF oscillators, filters and amplifiers. A reduction of the number of RF chains is therefore sought, with a trade-off between spectral efficiency and energy efficiency, as shown in [2]. Also several architectures for the analog stage and for the signal processing have been proposed, see for example [3] for a survey on the signal processing techniques applied to hybrid analog-digital MIMO schemes.

Massive MIMO systems at mmW may be deployed at different elements of the future network architecture [4], namely

at the backhaul/fronthaul and at the radio access. In this last case, they may be deployed at the base station (BS) or even at the user equipment (UE) if the reduction of the dimensions of the antennas facilitated by the migration to mmW is enough (see [5] for other possibilities to deploy a massive number of antennas at the UE). Many previous studies consider a simplified environment with single-antenna receivers [6], [7], [8], which corresponds to a massive antenna deployment only at the BS. In this paper we also consider the other possibilities where a large number of antennas may be possible at both ends of the communication link, transmitter and receiver.

In the recent literature, although the use of mmW implies that the frequency stability of oscillators becomes a challenging issue, the effect of phase noise is not considered. Here we study the design of hybrid analog-digital pre-coding and combining at transmitter and receiver, and we analyze the degradation introduced by the phase noise of the RF chains. Two schemes are considered for the design of the analog stage at the transmitter and receiver (pre-coding and combining), with two possible specifications for the phase shifters which make up the analog matrices. Design considerations are provided for the number of RF chains, number of oscillators and for the schemes where the antennas are grouped in blocks so reduce the complexity. We show that the number of blocks and the design of the RF analog stage have a different impact on the achievable rate depending on the amount of phase noise.

II. SYSTEM MODEL

We consider a setup where the transmitter and receiver are equipped with N_t and N_r antennas respectively and spatial multiplexing is employed to convey N_s parallel data streams. The hybrid architecture reduces the number of RF chains to L_t at the transmitter (and L_r at the receiver) with respect to the number of antennas, with a corresponding reduction of the number of ADC and DAC converters. We assume that $N_s \leq L_t \leq N_t$ at the transmitter and $N_s \leq L_r \leq N_r$ at the receiver. The scheme is shown in Fig. 1. The N_s streams are pre-coded by a baseband $L_t \times N_s$ matrix \mathbf{F}_{BB} into L_t RF chains at the transmitter and digital-to-analog (DAC) converted, then the signal is precoded with the analog precoder $N_t \times L_t$ matrix \mathbf{F}_{RF} to be transmitted by N_t antennas. The channel is represented by a $N_r \times N_t$ matrix \mathbf{H} and the signal is received by a N_r antennas at the receiver, then combined at RF by the $L_r \times N_r$ matrix \mathbf{W}_{RF}^H

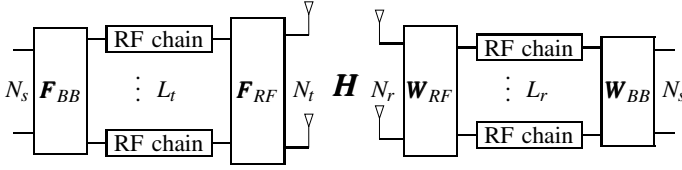


Fig. 1. System model with hybrid digital-analog beamforming and combining with antennas not divided in blocks.

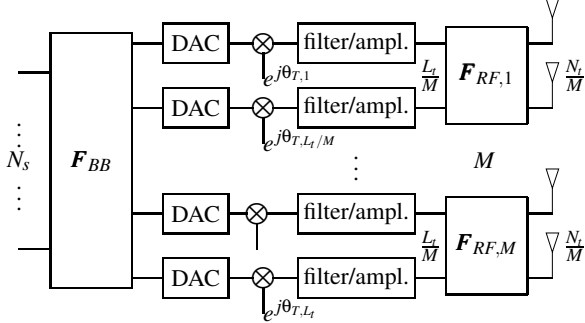


Fig. 2. Localized architecture of the transmitter with M blocks of N_t/M antennas.

into L_r RF chains which are analog-to-digital (ADC) converted at the receiver. Finally, a baseband combining is performed by the $N_s \times L_r$ matrix \mathbf{W}_{BB}^H giving N_s signal streams. As depicted in Fig. 2, each RF chain at the transmitter typically comprises a DAC, a mixer to RF and a filter/amplifier. On the other hand, at the receiver, the inverse operations are performed in each RF chain, with ADCs instead of DACs.

To reduce further the complexity the antennas or the RF stages at the transmitter can be grouped and only some RF outputs are sent to a group of antennas instead of sending all the RF outputs to all the antennas. Then a similar grouping can be applied at the receiver to simplify the combiner. Here we consider both solutions: (i) the case where at the transmitter all the L_t signals are combined by the matrix \mathbf{F}_{RF} and applied to all the N_t antennas, while at the receiver all the N_r antenna outputs are combined by \mathbf{W}_{RF}^H into the L_r RF receiver stages, as in Fig. 1; and (ii) a localized architecture, where the phase shifters have reduced size and the antennas are grouped in blocks. A scheme of the transmitter with M blocks of antennas is shown in Fig. 2.

At the detection point, the sampled signal vector \mathbf{y} , in terms of the vector \mathbf{s} of transmitted symbols, is given by

$$\mathbf{y} = \mathbf{W}^H \mathbf{H} \mathbf{F} \mathbf{s} + \mathbf{W}^H \mathbf{n} \quad (1)$$

where the pre-coding matrix is $\mathbf{F} = \mathbf{F}_{RF} \mathbf{F}_{BB}$, while \mathbf{n} denotes the AWGN vector contribution.

In the block architecture the RF matrices are block diagonal, with M diagonal blocks. For \mathbf{F}_{RF} they have size $\frac{N_t}{M} \times \frac{L_t}{M}$

$$\mathbf{F}_{RF} = \begin{bmatrix} \mathbf{F}_{RF,1} & 0 & 0 \\ 0 & \ddots & 0 \\ 0 & 0 & \mathbf{F}_{RF,M} \end{bmatrix}. \quad (2)$$

Similar considerations apply to the receiver side and the combining matrix is $\mathbf{W} = \mathbf{W}_{RF} \mathbf{W}_{BB}$.

A. Channel Model

The channel for massive MIMO at mmW is characterized by high directivity and it is represented in the beamspace [3], [9], [10], where a flat fading channel is described by N_p scatterers, associated to their transmit and receive angles. The channel matrix is then

$$\mathbf{H} = \sum_{p=1}^{N_p} h_p \mathbf{a}_R(\vartheta_{R,p}) \mathbf{a}_T^H(\vartheta_{T,p}), \quad (3)$$

where the vectors $\mathbf{a}_R(\vartheta_{R,p})$ and $\mathbf{a}_T(\vartheta_{T,p})$ denote the array response to the angles $\vartheta_{R,p}$ and $\vartheta_{T,p}$ at the transmitter and receiver, respectively. For a linear array with N elements spaced by d at the wavelength λ the response vector is

$$\mathbf{a}(\vartheta) = [1, e^{j2\pi\phi}, e^{j4\pi\phi} \dots e^{j2\pi\phi(N-1)}]^T \quad \phi = \frac{d}{\lambda} \sin \vartheta. \quad (4)$$

In practice, once the array geometry is fixed, the channel is characterized by N_p triplets of transmit and receiver angles and the corresponding complex gain $(h_p, \vartheta_{T,p}, \vartheta_{R,p})$.

B. Phase noise

The phase noise comprises two independent contributions, one at the transmitter and one at the receiver. It is introduced in the RF chain, as depicted in Fig. 2 for the transmitter. Depending on the design architecture, the number of oscillators, hence of independent phase noises, can vary from one oscillator per block to one oscillator per RF chain. For a flat fading channel, its effect is the rotation of each signal sample by a random phase, which corresponds to the multiplication by a diagonal matrix in the RF chain

$$\mathbf{P}_T = \text{diag} [e^{j\theta_{T,1}}, \dots, e^{j\theta_{T,L_t}}], \quad (5)$$

where $\theta_{T,m}$ are zero-mean Gaussian random variables with independent increments, corresponding to the classic Wiener noise, with variance of the phase increment over the symbol period $\sigma_{\theta_T}^2$. Another contribution is introduced at the receiver, with the corresponding diagonal matrix of size $L_r \times L_r$ and variance of the increments $\sigma_{\theta_R}^2$.

Several choices can be considered for the number of oscillators employed in the RF chains. Here we consider two cases: i) the limiting case where each RF chain is driven by an independent oscillator and, for the block architecture, also ii) the case where each sub-array has only one oscillator, i.e. all the RF chains within each block are fed by the same oscillator. In the latter case, the phase noise is the same for all the RF chains in each block, while different blocks experience independent phase noises.

III. DESIGN OF THE HYBRID SCHEME

We present a design where the analog stage is simplified, with two possible specifications for the analog RF matrices, made by phase shifters. Then the digital stage is optimized to maximize the system achievable sum rate R , given by

$$R = \log_2 \left| \mathbf{I} + \rho \mathbf{R}_n^{-1} (\mathbf{W}^H \mathbf{H} \mathbf{F}) (\mathbf{W}^H \mathbf{H} \mathbf{F})^H \right|, \quad (6)$$

where \mathbf{R}_n is the noise covariance matrix and ρ is the signal-to-noise ratio (SNR), $\rho = \frac{E[|s|^2]}{E[|n|^2]}$.

Thus the optimal solution is to obtain \mathbf{W} and \mathbf{F} as the matrices corresponding to the singular value decomposition (SVD) of \mathbf{H} , assuming a perfect channel estimation as in [3]. Unfortunately, this would require a $N_t \times N_t$ precoder \mathbf{F} and a $N_r \times N_r$ combiner \mathbf{W} , with a number of RF chains equal to the number of antennas, not compatible with the hybrid architecture. Also, with hybrid analog-digital techniques, the analog part should be as simple as possible. Common proposals are the use of phase shifters, or even switches, for the implementation of the analog matrices.

A. Achievable rate in the presence of phase noise

Actually, due to the presence of phase noise, the pre-coding and combiner matrices, obtained on the basis of the channel \mathbf{H} , are modified by the phase noise matrices \mathbf{P}_T and \mathbf{P}_R at the transmitter and receiver. Therefore, the overall channel, which includes both pre-coding and combining and determines the rate of the link, is given by the matrix $\tilde{\mathbf{H}}$,

$$\tilde{\mathbf{H}} = (\mathbf{W}_{RF}^H \mathbf{P}_R^H \mathbf{W}_{BB}^H) \mathbf{H} (\mathbf{F}_{RF} \mathbf{P}_T \mathbf{F}_{BB}), \quad (7)$$

and the effect of the phase noise is to introduce an interference from other streams, thus reducing the SINR. Then

$$R = \sum_{i=1}^{N_s} R_i, \quad (8)$$

where the rate R_i of stream i is given by

$$R_i = \log_2 \left(1 + \frac{|\tilde{\mathbf{H}}(i, i)|^2}{\sum_{j=1}^{N_s} |\tilde{\mathbf{H}}(i, j)|^2 + \frac{1}{\rho} \sum_{j=1}^{N_s} |\mathbf{W}^H(i, j)|^2} \right). \quad (9)$$

In other words, the derivation of \mathbf{F} and \mathbf{W} is based on a channel which is not the actual one, due to the phase noise.

B. Analog precoding and combiner

We address the case in which the matrices of the analog stages \mathbf{F}_{RF} and \mathbf{W}_{RF} are made of phase shifters.

$$\mathbf{F}_{RF} = \begin{bmatrix} e^{j\phi_{1,1}} & \dots & e^{j\phi_{1,L_t}} \\ \vdots & & \vdots \\ e^{j\phi_{N_t,1}} & \dots & e^{j\phi_{N_t,L_t}} \end{bmatrix} \quad (10)$$

Here we consider two possible design solutions for the analog matrices to simplify the overall system complexity, where the steering angles of the phase shifter matrices correspond to:

- Equally spaced angles (EQ) $\phi_{T,\ell} = \ell\pi/L_t \quad \ell = 1 \dots L_t$
- Random angles (RN) $\phi_{T,\ell} = u_\ell \quad \ell = 1 \dots L_t$, with u_ℓ independent identically distributed uniform random variables between 0 and π .

Then, according to (4), for a linear array \mathbf{F}_{RF} is given by

$$\mathbf{F}_{RF} = \begin{bmatrix} 1 & \dots & 1 \\ e^{j2\pi \frac{d}{\lambda} \sin \phi_{T,1}} & \dots & e^{j2\pi \frac{d}{\lambda} \sin \phi_{T,L_t}} \\ \vdots & & \vdots \\ e^{j2\pi \frac{d}{\lambda} (N_t-1) \sin \phi_{T,1}} & \dots & e^{j2\pi \frac{d}{\lambda} (N_t-1) \sin \phi_{T,L_t}} \end{bmatrix} \quad (11)$$

In the same way the analog combiner matrix \mathbf{W}_{RF} is obtained, by a suitable substitution of the dimensions L_r and N_r .

C. Digital precoding and combiner

Once the analog matrices are set, an equivalent channel is obtained by the cascade of the RF matrices and the actual channel

$$\mathbf{H}_{eq} = \mathbf{W}_{RF}^H \mathbf{H} \mathbf{F}_{RF}. \quad (12)$$

In order to maximize the achievable rate the digital part should give the SVD of the equivalent channel \mathbf{H}_{eq} , that is, $\mathbf{H}_{eq} = \mathbf{U} \mathbf{\Lambda} \mathbf{V}^H$. However, since the digital pre-coder and combiner are not square matrices, they are obtained by taking the first N_s columns of \mathbf{U} and \mathbf{V} corresponding to the largest singular values.

$$\mathbf{F}_{BB} = \mathbf{V}_{N_t \times N_s} \quad \mathbf{W}_{BB} = \mathbf{U}_{N_r \times N_s}. \quad (13)$$

IV. NUMERICAL RESULTS

In the following results the channel in (3) is modeled with independent uniform scattering angles and Rayleigh coefficients, with parameters as in [11], and the noise \mathbf{n} has independent components with covariance matrix $\mathbf{R}_n = \sigma^2 \mathbf{I}$. Linear arrays are used at both the transmitter and receiver, with antenna separation d of half a wavelength. For the phase noise, in order to reduce the number of parameters, we assume that the amount of phase noise is the same at the transmitter and receiver, i.e. $\sigma_{\theta}^2 = \sigma_{\theta_T}^2 = \sigma_{\theta_R}^2$.

In Fig. 3 the achievable rate per stream is shown with $N_s = 4$ streams, $L_t = L_r = 8$ RF chains, $N_t = N_r = 16$ antennas at the transmitter and receiver, and different values of the phase noise parameter σ_{θ} . The reference rate in the absence of phase noise is also shown and different design approaches for the analog matrices are compared. We can see the degradation of

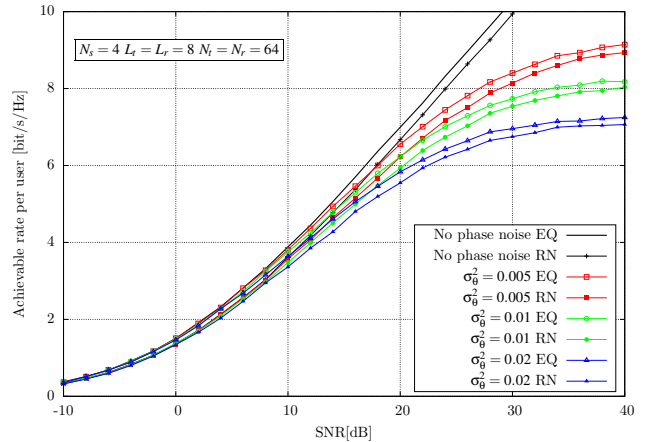


Fig. 3. Achievable rate as a function of the SNR for different design types of the analog stage.

phase noise, where a plateau is reached at high SNR due to the contribution of inter-stream interference. Moreover, a RF analog matrix with equally spaced angles obtains a slightly better rate than the case with random angles.

A. Number of RF chains and antenna blocks

In the localized architecture, where antennas are grouped in blocks to reduce the number of phase shifters, the effect of the number of RF chains is shown in Fig. 4 in the absence of phase noise, using random angles at the analog stages. Note that the curve corresponding to M blocks starts from

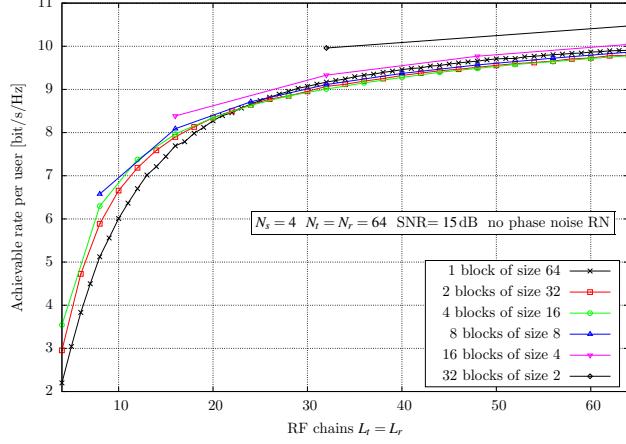


Fig. 4. Achievable rate as a function of the number of RF chains $L_t = L_r$, for a fixed number of antennas $N_t = N_r = 64$ and SNR=15dB and no phase noise, for RN and different number of blocks of antennas M .

$L_t = L_r = M$ and the values of RF-chains considered are multiples of M . It can be clearly seen that a small gain in terms of rate is obtained by a block design of the RF stage. This occurs in the absence of phase noise due to the fact that the diagonalization, by successive SVD decompositions as described previously, works slightly better when the RF analog matrices are block diagonal. Similar considerations apply to the EQ option. Considering now the effect of phase noise, the increasing rate with the number of RF chains is shown in Fig. 5, where different number of blocks are compared and the RF analog matrices are set by equally spaced angles (EQ). Note that in this case, although the phase shifters and the antennas are grouped in blocks, each RF chain is driven by an independent oscillator. A first conclusion is that, to have a negligible degradation in terms of achievable rate, the number of RF chains should be around half the number of final antennas, which is not always easy to achieve. Similar conclusions apply to the RN case and the difference between EQ and RN becomes less noticeable in the presence of phase noise.

B. One oscillator per block

However, the block architecture is normally employed to reduce the number of oscillators and to group some RF chains in order to make a single frequency conversion per group, i.e. the same oscillator is used in all the RF chains of the block. For equally spaced angles, the comparison between the use of one oscillator per block or one oscillator per RF chain is shown in Fig. 6 in terms of achievable rate, again as a function of the number of RF chains. We can see that with one oscillator per block the advantage of dividing the antennas into blocks

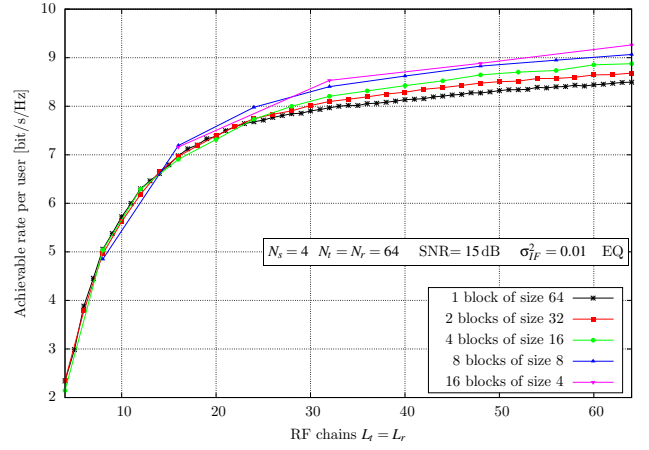


Fig. 5. Achievable rate for EQ as a function of the number of RF chains $L_t = L_r$, for a fixed number of antennas $N_t = N_r = 64$ and SNR=15dB and phase noise with $\sigma_\theta^2 = 0.01$, for different number of blocks of antennas.

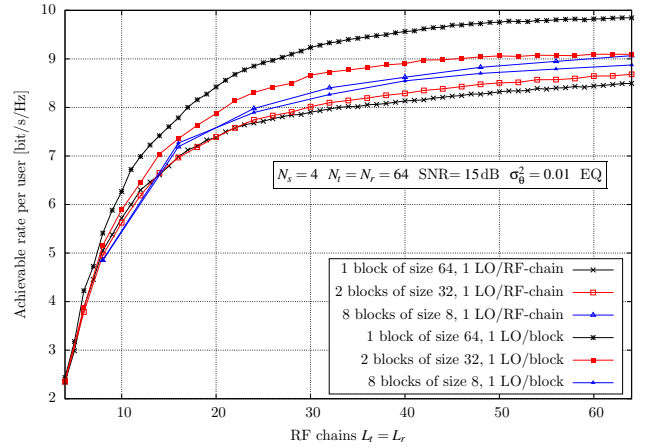


Fig. 6. Achievable rate with EQ as a function of the number of RF chains $L_t = L_r$, for a fixed number of antennas $N_t = N_r = 64$ and SNR=15dB and phase noise with $\sigma_\theta^2 = 0.01$, for different number of blocks of antennas.

is partially lost. This is due to the number of independent contributions of phase noise. So the design should tend to minimize the number of oscillators employed.

C. SNR Penalty

In order to show the sensitivity to the phase noise, we present the SNR penalty. The SNR penalty is defined as the increase of SNR necessary to achieve the same achievable rate obtained in the absence of phase noise (at a reference SNR). Fixing the number of antennas to $N_t = N_r = 64$ and the number of RF chains $L_t = L_r = 16$, we show the SNR penalty introduced by the phase noise at the reference SNR of 15dB in Fig. 7, with one oscillator per block. The case with one block $M = 1$ is not significant in terms of SNR penalty, since, considering a power penalty and flat fading, the penalty introduced by the same phase shift on all the elements of the signal vector is zero. Again, it is evident from Fig. 7 the increased sensitivity as the number of blocks increases, due to the larger number of independent oscillators. Moreover, when

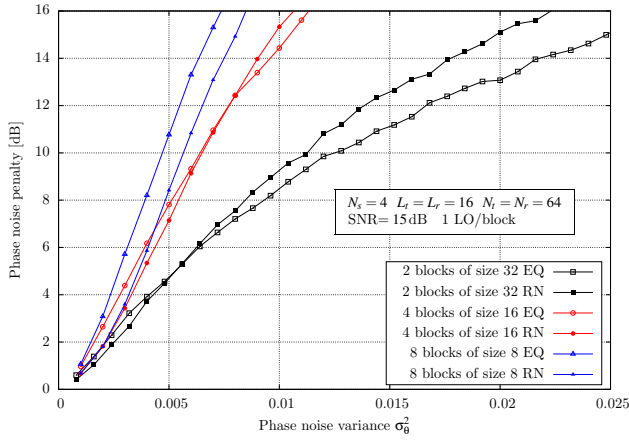


Fig. 7. Phase noise penalty at the reference SNR of 15 dB, as a function of σ_ϕ^2 for $N_t = N_r = 64$, $L_t = L_r = 16$, for different number of blocks of antennas and one oscillator per block.

the number of blocks increases, the SNR penalty is slightly lower for the RN scheme. For small phase noise the penalty is lower for EQ, while increasing the phase noise variance or the number of independent noises, in this case the number of blocks, the RN scheme becomes less sensitive. This is shown by the crossing between the EQ and RN lines in Fig. 7. Note that for $M = 8$ the crossing point occurs at a very high penalty.

If we consider a non-symmetric scenario in terms of number of antennas at the transmitter and receiver, for example a case in which the receiver has much less antennas than the transmitter, we can see the comparison in terms of penalty in Fig. 8. In particular we show the case of $N_t = 64$, $L_t = 16$ at the transmitter and at the receiver either the symmetric case ($N_r = 64$, $L_r = 16$) or a case with much less antennas and correspondingly RF-chains ($N_r = 8$, $L_r = 6$). The penalty is shown for 2 blocks of antennas, with one oscillator per block or one oscillator per RF-chain. It can be seen clearly that in

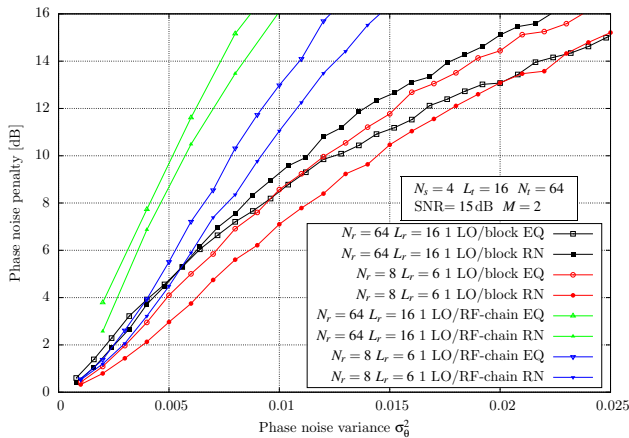


Fig. 8. Phase noise penalty at the reference SNR of 15 dB, as a function of σ_ϕ^2 for $N_t = 64$, $L_t = 16$, and different number of antennas at the receiver.

the case of one oscillator per block, being similar the total number of independent oscillators in the system, the penalty

is quite similar. On the other hand, when one oscillator per RF-chain is used, then the penalty is somehow larger for the setting with more antennas and RF chains.

V. CONCLUSIONS

The effect of phase noise in massive MIMO with hybrid analog-digital schemes cannot be neglected in general, since the degradation in terms of achievable rate can be large, especially if the number of independent oscillators is large. A first design consideration is that the number of RF chains to achieve almost the same rate as in the full architecture should be about half the number of antennas, then the further enhancement with more RF chains is small.

The localized architecture with reduced phase shifters gives an advantage when the phase noise is negligible, but in the presence of phase noise the penalty increases with the number of blocks, due to the increased number of oscillators. In general we can say that the penalty increases with the number of independent phase noise sources.

In order to limit the penalty introduced by the phase noise to less than 6 dB (with a number of antennas around 64) the value of phase noise increment variance should be below 0.005. This limit is slightly lower if a block architecture is employed, where each block is driven by a single oscillator.

ACKNOWLEDGMENTS

This work has received funding from the Spanish Ministry of Economy and Competitiveness (Ministerio de Economía y Competitividad) under projects TEC2014-59255-C3-1-R and TEC2014-59255-C3-3-R (ELISA).

REFERENCES

- [1] T. E. Bogale and L. B. Le, "Beamforming for multiuser massive MIMO systems: Digital versus hybrid analog-digital," *IEEE Globecom 2014*, Austin, TX, pp. 4066–4071, 2014.
- [2] S. Han, C. I. I. Z. Xu, and C. Rowell, "Large-scale antenna systems with hybrid analog and digital beamforming for millimeter wave 5G," *IEEE Commun. Mag.*, Vol. 53, No. 1, pp. 186–194, Jan. 2015.
- [3] R. W. Heath Jr., N. Gonzalez-Prelcic, S. Rangan, W. Roh, A. Sayeed, "An overview of signal processing techniques for millimeter wave MIMO systems", *IEEE J. Sel. Topics Signal Process.*, Vol. 10, No. 3, pp. 436–453, April 2016.
- [4] T. E. Bogale and L. B. Le, "Massive MIMO and mmWave for 5G Wireless HetNet: Potential Benefits and Challenges," *IEEE Veh. Technol. Mag.*, Vol. 11, No. 1, pp. 64–75, March 2016.
- [5] M. Sanchez-Fernandez et al., "Blended antenna wearables for an unconstrained mobile experience," submitted to *IEEE Commun. Mag.*, 2016.
- [6] Y. Wu, C. K. Wen, D. W. K. Ng, R. Schober, and A. Lozano, "Low-complexity MIMO precoding with discrete signals and statistical CSI," *IEEE Int. Conf. on Communications (ICC)*, Kuala Lumpur, 2016.
- [7] W. B. Sun, Q. Y. Yu, W. X. Meng, and C. Li, "Random beamforming for multiuser multiplexing in downlink correlated Rician channel," *IEEE Int. Conf. on Communications (ICC)*, Kuala Lumpur, 2016.
- [8] Z. Ding, R. Schober and H. V. Poor, "On the design of MIMO-NOMA downlink and uplink transmission," *IEEE Int. Conf. on Communications (ICC)*, Kuala Lumpur, 2016.
- [9] M. R. Akdeniz, Y. Liu, M. K. Samimi, S. Sun, S. Rangan, T. S. Rappaport, and E. Erkip, "Millimeter Wave Channel Modeling and Cellular Capacity Evaluation," *IEEE J. Sel. Areas Commun.*, Vol. 32, No. 6, pp. 1164–1179, June 2014.
- [10] A. M. Sayeed, "Deconstructing multiantenna fading channels," *IEEE Trans. Signal Process.*, Vol. 50, No. 10, pp. 2563–2579, Oct 2002.
- [11] A. Maltsev, A. Puduev, A. Lomayev, and I. Bolotin, "Channel modeling in the next generation mmWave Wi-Fi: IEEE 802.11ay standard," *European Wireless 2016*, Oulu, Finland, 2016.