Phase-only OFDM Communication for Downlink Massive MIMO Systems

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***Abstract*—The cost and power consumption associated with the large number of linear power amplifiers required at the base station in a massive MIMO-OFDM system present a significant challenge when deploying systems commercially. In this paper, we show that capacities of up to 4 bits per channel use (bpcu) per user can be achieved in the massive MIMO downlink whilst only transmitting the phase component of the precoded OFDM signals. This significantly reduces the peak-to-average power ratio (PAPR) of the transmit signals and relaxes the peak power and linearity requirements of the power amplifiers. Using Bussgang's theorem, it is shown that discarding the envelope information causes each OFDM subcarrier to be corrupted by an uncorrelated additive Gaussian error, which is received non-coherently by the users and asymptotically disappears as the number of transmit antennas is increased. SINR expressions are derived for linear precoders, and Monte Carlo simulations are performed with both Rayleigh and measured massive MIMO propagation channels. Under practical numbers of transmit antennas, we show that this phase-only OFDM scheme gives similar performance to other constant envelope precoding schemes described in the literature, and can achieve capacities of 2 bpcu per user with a 1-2 dB increase in mean transmit power (or 50% more transmit antennas) and 4-5 dB decrease in peak transmit power, compared to standard OFDM.**

Keywords*—*Massive MIMO, OFDM, constant envelope, PAPR, CPM

I. INTRODUCTION

In massive MIMO a large number of base station (BS) antennas are used to spatially multiplex multiple users on the same time-frequency resource. The step-change increase in sum spectral efficiency facilitated by this technology has led to massive MIMO being identified as a key physical layer technology for future 5G cellular networks. However, the OFDM signals used in wideband MIMO systems suffer from a high peak-to-average power ratio (PAPR), requiring expensive linear power amplifiers (PAs) that operate with a large power back-off and poor power efficiency [1]. With massive MIMO systems employing typically 100s of antennas, each driven by a separate RF chain, the cost and power consumption associated with these amplifiers is expected to be significant. In this paper, we analyse a 'constant envelope' transmission scheme—referred to herein as *phase-only OFDM*—that relaxes the PA requirements by reducing the PAPR of the transmit signals.

A. Related Work

PAs in current LTE base stations typically account for 50-80% of the power consumption [2], and much research has addressed PAPR reduction for SISO links—broadly based around either scrambling or distorting the OFDM signals [1]. However, the large number of degrees of freedom available in the massive MIMO channel provide opportunities for new techniques that relax hardware constraints.

The idea of using the excess degrees of freedom present in massive MIMO to reduce signal PAPR was first proposed in [3]. The authors showed that for narrowband channels it is possible to constrain the transmitted signals to be constant envelope (CE) in discrete time and still suppress inter-user interference (IUI). They proposed an iterative gradient descent algorithm for solving the non-linear least squares (NLS) problem of selecting the transmit phase angles and showed that a capacity of 2 bits per channel use (bpcu) per user requires 1.7 dB more transmit power than with standard signalling. This work was extended to frequency selective channels in [4], with an iterative time-domain algorithm that was shown to achieve a capacity of 2 bpcu per user with a 1-2 dB increase in transmit power. As the authors noted, fixing the signal to be CE in discrete time does not produce a signal that is CE (PAPR = 1) in continuous time, but does reduce continuous time PAPR. In [5], PAPR was reduced by constraining the difference between consecutive phase angles, at the expense of increased transmit power. An alternative cross-entropy optimization algorithm for solving the NLS CE problem for narrowband channels was outlined in [6].

The work in [7] looked specifically at OFDM based massive MIMO systems, proposing an iterative algorithm for jointly performing precoding, modulation and PAPR reduction, which reduced (discrete time) PAPR from around 13 dB to 3 dB. In [8] a Bayesian OFDM PAPR reduction scheme was outlined, which achieved similar performance to [7] but with reduced complexity.

B. Contributions

In this paper, we propose and analyse a scheme in which the transmit samples are constrained to be constant envelope by taking the precoded OFDM signal and removing all amplitude variation, retaining only the time varying phase component. This has some advantages over other massive MIMO constant envelope schemes, namely

- The received signal is the original OFDM signal plus an additive Gaussian error, and hence the scheme is compatible with conventional OFDM-based receivers and standards.
- Conventional MIMO/OFDM baseband processing is used, allowing off-the-shelf algorithms and hardware designs, requiring only a scaling of samples prior to digital-to-analogue conversion.

- SINR expressions are given, enabling performance to be easily analysed for different environments and linear precoders.

II. System Model

In this paper, we consider a downlink MIMO-OFDM system with M transmit antennas, K single-antenna receivers over N subcarriers. The data symbols on each subcarrier are linearly precoded using the matrix $\mathbf{W}_n \in \mathbb{C}^{M\times K}$ as

$$\mathbf{x}_n = \mathbf{W}_n \mathbf{s}_n, \tag{1}$$

where $\mathbf{s}_n = [s_{1,n}, \cdots, s_{K,n}]^T$, $s_{k,n}$ is the data symbol of the kth user transmitted on the nth subcarrier and $\mathbf{x}_n = [x_{n,1}, \cdots, x_{n,M}]^T$ with $x_{n,m}$ being the precoded symbol transmitted from the mth antenna on subcarrier n. In the OFDM system, an inverse fast Fourier transform (IFFT) operation is performed at each antenna element. The time domain signal transmitted from the mth antenna at time instant l is thus given by

$$\tilde{x}_m[l] = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} x_{k,m}.e^{j2\pi\frac{kl}{N}}. \tag{2}$$

In this paper, we propose to use a phase-only OFDM transmission scheme, i.e., the amplitudes of the transmit time domain signals are fixed to a constant value, A. In vector notation, the time domain signal transmitted across the M antennas can be represented as

$$\tilde{\mathbf{x}}'[l] = \left[Ae^{j\phi_1[l]}, \cdots, Ae^{j\phi_M[l]}\right]^T. \tag{3}$$

Note that the frequency domain received signal on the nth subcarrier across the K receivers is

$$\mathbf{y}_n = \sqrt{\rho}\mathbf{H}_n \mathbf{x}'_n + \eta_n, \tag{4}$$

where ρ is transmit power, $\mathbf{H}_n \in \mathbb{C}^{K\times M}$ is the channel matrix between the M transmit antennas and the K users on subcarrier n and η_n is the additive white Gaussian noise (AWGN) vector observed on subcarrier n. In the next section, we present the performance analysis of this scheme.

III. Performance Analysis

A. Phase-only SISO OFDM

Let us first consider a SISO system. The time domain OFDM symbol $x[l]$, with $\mathbb{E}\{|x_n|^2\} = 1$, is

$$x[l] = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} x_n.e^{j2\pi\frac{nl}{N}}. \tag{5}$$

This can be written in polar form as

$$x[l] = a[l]e^{j\phi[l]}. \tag{6}$$

For sufficiently large number of subcarriers N, $x[l]$ can be accurately modelled as a stationary zero-mean complex Gaussian process, the envelope of which, $a[l]$, is Rayleigh distributed [1]. Fixing the envelope at A and retaining just the phase component, the transmit signal is given by

$$x'[l] = Ae^{j\phi[l]} = A\frac{x[l]}{|x[l]|}. \tag{7}$$

This mapping can be represented using a memoryless non-linear transfer function

$$x'[l] = F\left(x[l]\right) = G\left(a[l]\right) e^{j\phi[l]}, \tag{8}$$

where $G(a[l]) = A$. Applying Bussgang's theory for complex signals [9], the time-domain transmit signal can be expressed

$$x'[l] = A\sqrt{\frac{\pi}{4}}x[l] + A\sqrt{1-\frac{\pi}{4}}q[l], \tag{9}$$

where $q[l]$ is an uncorrelated error signal with $\mathbb{E}\{q[l]^* x[l]\} = 0$ and $\mathbb{E}\{|x[l]|^2\} = \mathbb{E}\{|q[l]|^2\} = 1$. Setting $A = \sqrt{\frac{4}{\pi}}$, the phase-only OFDM signal is given by

$$x'[l] = \sqrt{\frac{4}{\pi}}e^{j\phi[l]} = x[l] + \sqrt{\frac{4-\pi}{\pi}}q[l]. \tag{10}$$

Note that the total power of $x'[l]$ is $\frac{4}{\pi} \approx 1$ dB greater than the original OFDM signal $x[l]$. According to the linearity property of the discrete Fourier transform, the phase-only OFDM signal can be decomposed back into subcarriers, giving

$$x'_n = x_n + \sqrt{\frac{4-\pi}{\pi}}q_n. \tag{11}$$

Hence each subcarrier is corrupted by an additive white error, which according to the central limit theorem is Gaussian, i.e., $q_n \sim \mathcal{CN}(0,1)$. In a phase-only OFDM link, both the data signal and modulation error scale with transmit power, ρ. The received signal on the nth subcarrier is thus

$$\begin{aligned} y_n &= \sqrt{\rho}h_n x'_n + \eta \\ &= \sqrt{\rho}h_n x_n + \sqrt{\rho}\sqrt{\frac{4-\pi}{\pi}}h_n q_n + \eta, \end{aligned} \tag{12}$$

where h_n is the channel coefficient on the nth subcarrier and the second term in the sum may be considered self-interference. It follows that the SINR for a SISO link is

$$SINR = \frac{\rho|h_n|^2}{\rho\frac{4-\pi}{\pi}|h_n|^2 + \sigma_\eta^2}, \tag{13}$$

where σ_η^2 is the variance of the noise. At high transmit power, the system is self-interference limited, leading to

$$SINR \to \frac{\pi}{4-\pi} = 5.6dB. \tag{14}$$

B. MISO Performance

Consider now a MISO OFDM system with M transmit antennas with maximum ratio transmission (MRT) precoding. The precoded vector for the nth subcarrier, $\mathbf{x}_n \in \mathbb{C}^{M\times 1}$, is

$$\mathbf{x}_n = \frac{\mathbf{h}_n^*}{\|\mathbf{h}_n\|^2} s_n, \tag{15}$$

where $E\{|s_n|^2\} = 1$ and $\mathbf{h}_n \in \mathbb{C}^{M\times 1}$ is the downlink channel on the nth subcarrier. Assuming an equal power split between the transmit antennas, the average per-antenna OFDM signal power is

$$\mu = \frac{1}{M}\mathbb{E}\{\|\mathbf{x}[l]\|^2\} = \frac{1}{MN}\sum_{n=0}^{N-1}\frac{1}{\|\mathbf{h}_n\|^2}. \tag{16}$$

Following a similar derivation to Section III-A, the MISO phase-only OFDM transmit vector in the frequency domain is

$$\mathbf{x}'_n = \mathbf{x}_n + \sqrt{\mu}\sqrt{\frac{4-\pi}{\pi}}\mathbf{q}_n \tag{17}$$

where $\mathbf{x}'_n \in \mathbb{C}^{M\times 1}$, $\mathbf{q}_n \in \mathbb{C}^{M\times 1}$ and $\mathbb{E}\{\mathbf{q}_n\mathbf{q}_n^H\} = \mathbf{I}_M$. With transmit power scaling ρ, the received signal is

$$\begin{aligned} y_n &= \sqrt{\rho}\mathbf{h}_n^T\mathbf{x}'_n + \eta_n \\ &= \sqrt{\rho}s_n + \sqrt{\rho\mu}\sqrt{\frac{4-\pi}{\pi}}\mathbf{h}_n^T\mathbf{q}_n + \eta_n. \end{aligned} \tag{18}$$

We observe that it has three components: the data symbol, the modulation error and the receiver noise. The modulation error is uncorrelated with the data symbol and assumed to be uncorrelated with the channel[1]. The modulation error therefore combines non-coherently at the receiver whilst the precoded data symbol combines coherently, giving

$$SINR_n = \frac{\rho}{\rho\mu\frac{4-\pi}{\pi}\|\mathbf{h}_n\|^2 + \sigma_\eta^2}. \tag{19}$$

For large M the 'channel hardening' effect is observed where

$$\|\mathbf{h}_n\|^2 \approx \mathbb{E}\{|h_{n,m}|^2\}M = \beta M, \tag{20}$$

with β accounting for shadow fading. Substitution of (20) into (19) gives

$$SINR \approx \frac{\rho}{\frac{4-\pi}{\pi}\frac{\rho}{M} + \sigma_\eta^2}. \tag{21}$$

In the high SNR regime i.e., $\rho \gg \sigma_\eta^2$, the performance is limited by self-interference, with the SINR given by (22). A diversity gain of M is achieved compared to the SISO link.

$$SINR \to \frac{\pi}{4-\pi}M \tag{22}$$

Figure 1 shows the bit error rate with different numbers of antennas for QPSK modulation.

C. Massive MIMO Performance

We now extend the analysis to a phase-only OFDM massive MIMO system serving K users. For the sake of illustration and clarity, we will drop the subcarrier index in the following analysis. The received signal in this scenario is given by

$$\mathbf{y} = \sqrt{\rho}\mathbf{HWs} + \sqrt{\rho\mu}\sqrt{\frac{4-\pi}{\pi}}\mathbf{Hq} + \eta, \tag{23}$$

where the per-antenna OFDM power is

$$\mu = \frac{1}{M}\mathbb{E}\{\|\mathbf{x}[l]\|^2\} = \frac{1}{MN}\sum_{n=0}^{N-1}\|\mathbf{W}_n\|_F^2. \tag{24}$$

The received symbol at the kth user is given by

$$y_k = \sqrt{\rho}\mathbf{h}_k^T\mathbf{w}_k s_k + \sqrt{\rho}\sum_{j\neq k}\mathbf{h}_k^T\mathbf{w}_j s_j + \sqrt{\rho\mu}\sqrt{\frac{4-\pi}{\pi}}\mathbf{h}_k^T\mathbf{q} + \eta_k \tag{25}$$

where $\mathbf{h}_k \in \mathbb{C}^{M\times 1}$ is the channel vector between the transmitter and kth user and beamforming vectors $\mathbf{w}_k \in \mathbb{C}^{M\times 1}$.

[1]This is supported by simulations using both modelled and measured propagation channels; a rigorous proof is beyond the scope of this paper.

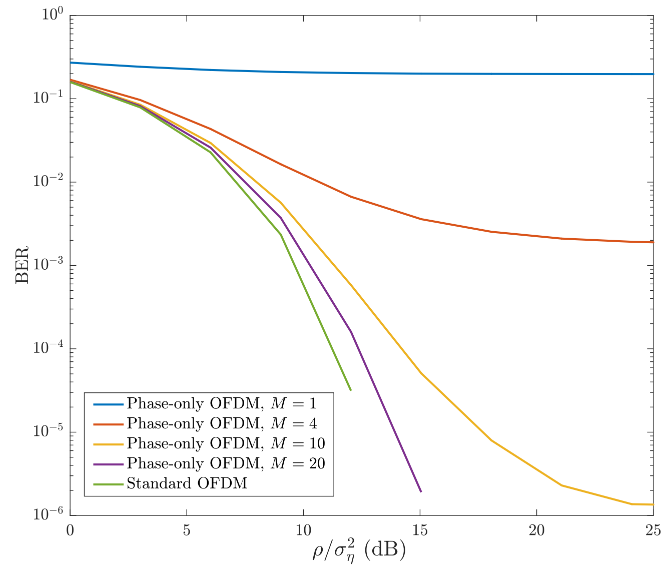

Fig. 1. BER of MISO phase-only OFDM for varying number of transmit antennas, QPSK, $N = 1024$

Note that the expression is similar to (18) with an additional inter-user interference (IUI) term. The resultant SINR is thus

$$SINR_k = \frac{\rho|\mathbf{h}_k^T\mathbf{w}_k|^2}{\rho\sum_{j\neq k}|\mathbf{h}_k^T\mathbf{w}_j|^2 + \rho\mu\frac{4-\pi}{\pi}\|\mathbf{h}_k\|^2 + \sigma_\eta^2}. \tag{26}$$

We next present the SINR expression under specific precoding schemes.

1) Maximum Ratio Transmission: The MRT precoding matrix is defined as

$$\mathbf{W}_{MRT} = [\mathbf{w}_1 \quad \mathbf{w}_2 \quad \cdots \quad \mathbf{w}_K]^T, \quad \mathbf{w}_k = \frac{\mathbf{h}_k^*}{\|\mathbf{h}_k\|^2}. \tag{27}$$

For large M channel hardening is observed, with user channel shadowing β_k, yielding the SINR expression

$$SINR_{MRT} \approx \frac{\rho}{\frac{\rho}{M^2}\sum_{j\neq k}|\mathbf{h}_j^H\mathbf{h}_k|^2/\beta_j^2 + \rho\frac{4-\pi}{\pi}\frac{K}{M} + \sigma_\eta^2}. \tag{28}$$

For independent, identically distributed (i.i.d) complex normal channel (Rayleigh) entries with unit variance ($\beta_k = 1$), $|\mathbf{h}_j^H\mathbf{h}_k|^2 \approx M$. Assuming a large number of users, $\frac{K-1}{K} \approx 1$, the SINR can be expressed in terms of the total transmit power, $P_T = \frac{4}{\pi}\rho\mu M$, as

$$SINR_{MRT} \approx \frac{P_T}{(P_T + \sigma_\eta^2)}\frac{\pi}{4}\frac{M}{K}. \tag{29}$$

At high transmit power, performance is limited by both IUI and modulation error, and the SINR is

$$SINR_{MRT} \to \frac{\pi}{4}\frac{M}{K}. \tag{30}$$

In comparison, the SINR expression for MRT precoding in the Rayleigh channel with conventional OFDM is

$$SINR_{MRT}^* \approx \frac{P_T}{P_T + \sigma_\eta^2}\frac{M}{K} \to \frac{M}{K}. \tag{31}$$

The phase-only OFDM system suffers from a constant $\frac{\pi}{4} \approx -1$ dB SINR loss relative to conventional OFDM.

2) Zero Forcing Precoding: The zero-forcing precoder is given by

$$\mathbf{W}_{ZF} = \mathbf{H}^H \left(\mathbf{H}\mathbf{H}^H\right)^{-1}. \tag{32}$$

This precoder removes IUI, leaving only modulation error and noise. With channel hardening and shadowing β, the received SINR is

$$SINR_{ZF} \approx \frac{\rho}{\rho\frac{4-\pi}{\pi}\beta M\mu + \sigma_\eta^2} \tag{33}$$

$$\approx \left(\frac{1}{\frac{4-\pi}{\pi}\frac{P_T}{\sigma_\eta^2}\beta + \frac{4}{\pi}}\right)\frac{P_T}{\sigma_\eta^2}\frac{1}{M\mu} \tag{34}$$

$$\to \frac{\pi}{\beta(4-\pi)}\frac{1}{M\mu}. \tag{35}$$

Relative to ZF with standard OFDM, Eq. 36, at low transmit power it suffers a loss of $\frac{\pi}{4} \approx -1$ dB in SINR, whilst at high power performance is modulation error limited.

$$SINR^*_{ZF} = \frac{P_T}{\sigma_\eta^2}\frac{1}{M\mu} \tag{36}$$

As with MRT, for the i.i.d. Rayleigh channel, the ZF phase-only SINR scales linearly with $\frac{M}{K}$

$$SINR_{ZF} \approx \left(\frac{1}{\frac{4-\pi}{\pi}\frac{P_T}{\sigma_\eta^2} + \frac{4}{\pi}}\right)\frac{P_T}{\sigma_\eta^2}\left(\frac{M}{K}-1\right) \to \frac{\pi}{4-\pi}\left(\frac{M}{K}-1\right). \tag{37}$$

3) Linear MMSE Precoding: From the previous analysis, it is observed that the modulation error scales with the OFDM signal power, μ. With MRT this error is minimised at the expense of significant IUI, whilst ZF eliminates IUI at the expense of increased transmit power and modulation error. Under the assumption that channel statistics are constant across subcarriers, and that, for large M, $\mu = \frac{1}{MN}\sum_{p=0}^{N-1}\|\mathbf{W}_p\|_F^2 \approx \frac{1}{M}\|\mathbf{W}_n\|_F^2$, the minimum mean square error precoder, which balances IUI and modulation error, is given by

$$\mathbf{W}_{MMSE} = \mathbf{H}^H\left(\mathbf{H}\mathbf{H}^H + \frac{4-\pi}{\pi M}\|\mathbf{H}\|_F^2\mathbf{I}_K\right)^{-1}. \tag{38}$$

IV. Simulation Results

In this section, we present the simulation results of the proposed method under two types of channels, namely an analytic Rayleigh fading channel and a measured channel. We further show the PAPR benefit of the proposed method.

A. Rayleigh Channel

In Figure 2, we show the capacity of an i.i.d Rayleigh channel, per subcarrier, as a function of P_T/σ_η^2, for $M/K = 10$ and $K = 12$. The markers show the capacity calculated using the SINR in (29) and (37). At low user capacity phase-only OFDM requires an additional 1 dB transmit power over standard OFDM. At higher transmit powers the performance is limited and the difference between phase-only and standard OFDM increases. For ZF precoding an additional 1.5 dB is required for 2 bpcu, 2 dB for 3 bpcu and 3.7 dB for 4 bpcu. MMSE precoding marginally outperforms ZF, requiring 0.3 dB less power to achieve 4 bpcu, and both outperform MRT precoding over the investigated range of transmit powers. In the low transmit power region, this performance loss can be compensated for by using 50% more BS transmit antennas.

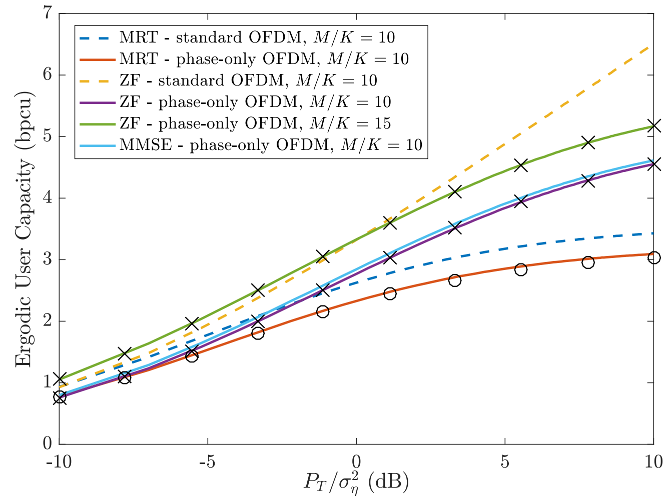

Fig. 2. Variation of ergodic user capacity with transmit power for i.i.d Rayleigh fading channel with $K = 12$. Capacity according to (29) and (37) shown by markers.

B. Measured Propagation Channel

Figure 3 shows the average per-user capacity (per subcarrier) as a function of P_T/σ_η^2, for $K = 6$, $M = 60$ and $N = 1200$ in a measured 20 MHz indoor channel captured using Bristol's Massive MIMO testbed. The channel was measured using frequency orthogonal uplink pilots - one per UE per 12-subcarrier resource block, with the user channel frequency responses assumed constant over each resource block. For further information about the architecture of the testbed, the interested reader is referred to [10]. The

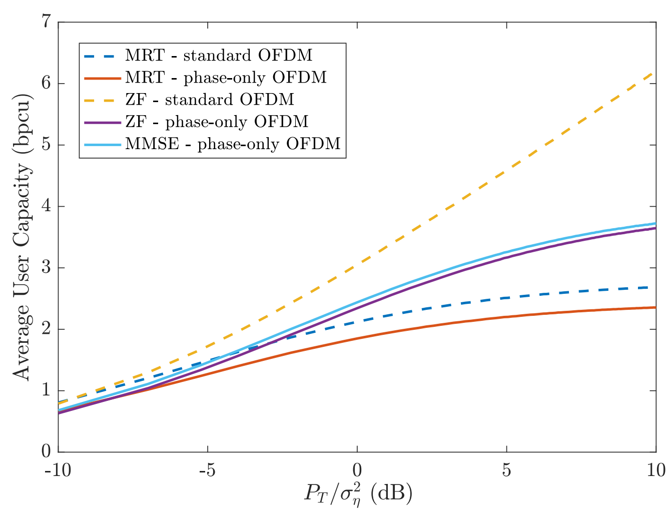

Fig. 3. Variation of mean user capacity with transmit power for measured indoor propagation channel with $M = 60$ and $K = 6$.

channel in Figure 3 was captured in an indoor conference hall, 30m×20m, with three UEs spaced along each side of the hall. Processing was performed offline using the captured channel matrices, with perfect transmitter channel knowledge assumed. Monte Carlo simulation was used to estimate the user SINRs and capacities, with the captured channel normalised to give $\sum_{n=0}^{N-1}\|\mathbf{H}_n\|^2 = MKN$.

To achieve 2 bpcu, phase-only OFDM requires an additional 1.8 dB of power relative to standard OFDM, and 3.2 dB more for 3 bpcu. With ZF/MMSE precoding, the capacity is lower than for the Rayleigh channel model, as the measured channel requires more power (μ) to cancel IUI, giving higher modulation error. These results indicate that phase-only transmission would be most appropriate for noise-limited and low user data rate applications, such as large cells or M2M.

C. PAPR Reduction

Constraining the signal to have a constant envelope at the sample points does not produce a signal with a constant envelope after reconstruction in continuous time. Figure 4 shows empirical probability density functions for the continuous time PAPR of an OFDM symbol for both standard and phase-only OFDM, with $N = 1024$ and QPSK modulation, measured by resampling the discrete time signals at $10\times$ the Nyquist rate (with ideal brickwall filter). The PAPR for phase-only OFDM shows a significant reduction in both mean and variance relative to standard OFDM, reducing the back-off required at the PA. The PAPR of the phase-only OFDM symbol is below 6.7 dB with 99.99% probability, compared to 12.4 dB for standard OFDM. This indicates, for example, that under Rayleigh fading conditions a capacity of 2 bpcu per user could be achieved with a 4.2 dB reduction in peak transmit power compared to standard OFDM. Comparison of the relative merits of phase-only transmission and conventional OFDM clipping schemes is left as future work.

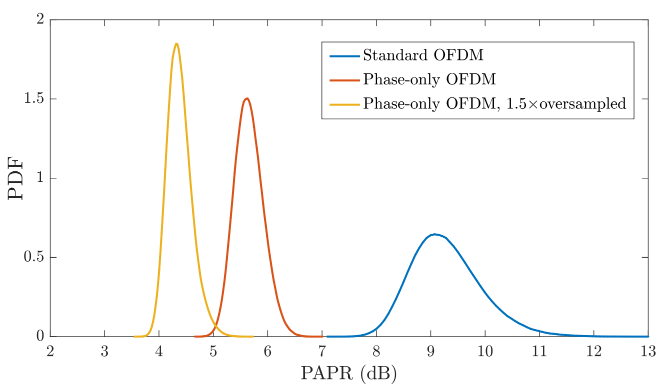

Fig. 4. Probability density of phase-only OFDM continuous time PAPR compared to standard OFDM

Further PAPR reduction can be achieved by oversampling $\mathbf{x}[n]$ prior to fixing the signal envelope, at the expense of increased signal bandwidth. If the signal is constrained to be phase-only for all time a continuous phase modulated (CPM) signal is produced, given according to Bussgang's theorem by (39), where $x(t)$ is the OFDM signal in continuous time and $q(t)$ the uncorrelated error. Combined with upcoming technologies such as direct antenna modulation [11], in which the carrier phase is directly modulated at the antenna, this could form the basis of a low cost, low complexity massive MIMO transmitter. However, the error signal $q(t)$ is not bandlimited, and hence $x'(t)$ has high spectral leakage. Further work on pulse shaping and filtering methods to reduce out-of-band emissions is therefore required.

$$x'(t) = x(t) + \frac{4-\pi}{\pi} q(t). \tag{39}$$

V. Conclusion

We have analysed a massive MIMO constant envelope precoding scheme in which only the phase component of OFDM signals is transmitted. This has been shown to corrupt each subcarrier with an uncorrelated additive Gaussian error, which decreases in power at the receivers as the number of transmit antennas is increased. Using phase-only OFDM transmission, performance is limited by this modulation error, with the capacity limit set by the antenna ratio M/K, precoder type and propagation channel characteristics. Analysis of Rayleigh and measured propagation channels with $M/K = 10$ show that at low transmit power phase-only OFDM can match the performance of standard OFDM with a 1 dB increase in mean transmit power, whilst capacities of 3-4 bpcu per user can be achieved with an increase of 3-4 dB, or 50% more transmit antennas. Simulations show that the scheme reduces the continuous time PAPR of the OFDM signals by 6 dB, enabling peak transmit power to be reduced by 2-5 dB.

VI. Acknowledgements

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