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Fast Convergence Joint Optimization of PAPR Reduction and Digital Predistortion in the Next-Generation Broadcasting Systems

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Abstract—The vast majority of designs on peak-to-average power ratio (PAPR) reduction and PA linearization schemes in broadcasting systems can be found in literature dealing with both of them in a separate manner on problem formulation, optimization objectives, and implementation issues without considering their mutual influence. Their overall performance might be sub-optimal even if each of them has been optimized independently due to possible conflicts as both techniques are interdependent. This paper proposes an adding signal method that jointly achieves PAPR reduction and PA linearization simultaneously, and no extra processing is required at the receiver. The simulation results show that the proposed scheme offers a good performance/complexity trade-off requiring fewer iterations than recent methods.

Index Terms—digital predistortion (DPD), DVB-T2, DTT, energy efficiency, green communications, power amplifier (PA), joint optimization, non-linear systems, OFDM, peak-to-average power ratio (PAPR), recursive error correction, Tone reservation.

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is widely used in various Digital terrestrial television (DTT) standards such as Digital Video Broadcast-Terrestrial second generation (DVB-T2 [1] and the recent American digital video broadcasting (ATSC 3.0) [2]. DTT networks need to broadcast very high power signals and it is not difficult to find that power amplifiers (PAs) at the base station consume the majority of power. As per green communications obligation, a PA must be operated as close as possible to its saturation point [3]. However, in this saturation region, PAs suffer from inherent non-linear (NL) distortion, causing NL effects such as in-band (IB) and out-of-band (OOB) distortions. Consequently, the input power of the PA is lowered often by a sufficient input back-off (IBO), in order to achieve an amplified signal with minimum distortion. The IBO of a PA, often expressed in dB, can be defined as the ratio between the saturation point to the mean power of the input signal.

Collectively, the compensation techniques can be grouped into two categories: *peak-to-average power ratio (PAPR) reduction schemes* and *PA linearization schemes*. While the first

category aims at reducing the large PAPR of the signal, the latter deals with PA linearization. Most of them are described in [4]. The tone reservation (TR) method has been adopted for PAPR reduction by different broadcasting standards. TR relies on the dedicated usage of a subset of subcarriers [5]. This reserved tones subset is used to generate a kernel signal added to the original signal and resulting in a transmitted signal which has a lower PAPR than the original one. The PAPR reduction technique considered in this work is the clipping control-based tone reservation method, which is commonly used and is adopted in DVB-T2 systems. Among the second category of solutions involving RF linearization of PA, digital predistortion (DPD) is one of the most popular and cost-effective. The DPD technique considered in this work is an adding signal technique that was recently proposed in [6] with improved convergence.

PAPR reduction and PA linearization are complementary. In case even if each of these two solutions is optimized according to its own criteria, their combination can not be optimal because of opposite effects [8]. In [9] where, in a Ping-Pong manner, the correction signal is synthesized accounting for the PAPR reduction and PA linearization. This allows for a better trade-off between PA efficiency and linearity. The main objective of this paper is to propose a novel approach; inspired by the ping-pong joint optimization (P2JO) approach in [9] that jointly combines PAPR reduction and PA linearization with an aim at creating a good synergy between the two techniques in order to provide optimal performance. This joint approach, termed as fast convergence joint optimization, aims at a collaborative exchange between these two processes with a good performance-complexity trade-off requiring few iterations. From the Busgang theorem, the possibility of predistortion via an adding signal will allow a common motive for both PAPR reduction and DPD processing. Thus, we shall be able to get an efficient combination in order to have an optimal compromise between PA's linearity and efficiency. Our proposed solution achieves this objective with fewer iterations.

This paper is organized as follows: Section II reminds the

OFDM system model, PA model, and some metrics related to PAPR reduction and PA linearization. Section III presents a brief overview of the clipping control TR (CC-TR) scheme for PAPR reduction, linearization problems, and the fast convergence DPD (FC-DPD) scheme. Methods adopted in our work for PAPR reduction and PA linearization will be studied. The proposed scheme is presented in Section IV. An analysis of PA linearization performance based on the simulation results is done in section V and then, the paper is concluded.

II. OVERVIEW OF OFDM SYSTEM MODEL AND PAPR ISSUES

A. OFDM System Model

Considering an OFDM system, the discrete-time baseband signal $\mathbf{x} = \mathcal{F}\{\mathbf{X}\} = [x_0, x_1, \dots, x_{N-1}]$ transmitted over N subcarriers can be expressed as

$$x_k = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} X_n e^{2\pi j \frac{nk}{NT}}, \quad 0 \leq k \leq NT - 1, \quad (1)$$

where \mathbf{X} is the M-ary QAM frequency-domain sequence of complex symbols $[X_0, X_1, \dots, X_{N-1}]$, k stands for a discrete-time index, T is the OFDM symbol duration, \mathcal{F} is the OFDM modulation function and $j = \sqrt{-1}$.

B. PA Model

Let us denote the discrete-time input and output vectors for PA as $\mathbf{x} = [x_0, x_1, \dots, x_{N-1}]$ and $\mathbf{y} = [y_0, y_1, \dots, y_{N-1}]$. We denote the PA amplification and predistortion functions as \mathbb{A} and \mathbb{P} respectively. in a memory-less NL PA model with a complex gain function \mathbb{G} , the output signal \mathbf{y} can be represented as

$$\mathbf{y} = \mathbb{A}(\mathbf{x}) = \mathbb{G}(\mathbf{x}) e^{j\phi_{\mathbf{x}}}, \quad (2)$$

where, $\phi_{\mathbf{x}}$ is the phase vector of the input signal \mathbf{x} , i.e. $[\phi_0, \phi_1, \dots, \phi_{N-1}]$ and ϕ_k is phase of x_k .

In this paper, we consider the well-known Rapp model [10] commonly used to model solid-state PAs in broadcasting systems. It is a memoryless PA model with only AM/AM conversion, i.e. for $0 \leq k \leq N - 1$:

$$y_k = \frac{g_s |x_k|}{\left(1 + \left(\frac{g_s |x_k|}{v_{sat}}\right)^{2p}\right)^{\frac{1}{2p}}}, \quad (3)$$

where g_s is the small signal gain of PA, p is the knee factor and v_{sat} is the input saturation voltage of PA. However, our approach can be extended to any other memoryless PA model.

C. PAPR Definition and Performance metrics

One of the characteristics of any multicarrier signal is the high peak values in the time domain, since, many subcarrier components are added via the Inverse fast Fourier transform (IFFT). The PAPR of a signal waveform is an important metric as a small value implies that the PA used to transmit signals can operate more efficiently. The PAPR of a signal \mathbf{x} measures

the ratio of the signal's maximum instantaneous power to its mean power over a symbol period T_s and is defined by

$$PAPR_{\mathbf{x}} = \frac{\|\mathbf{x}\|_{\infty}^2}{\mathbb{E}[\|\mathbf{x}\|_2^2]}, \quad (4)$$

where \mathbb{E} is the expectation parameter and $\|\cdot\|_2$ and $\|\cdot\|_{\infty}$ denote \mathbf{L}_2 and \mathbf{L}_{∞} norms, respectively.

The main parameters considered in quantifying the performance of the proposed scheme in reducing OOB and IB distortions are discussed in the subsequent subsections.

1) *Power spectral density for the OOB distortion analysis:* The power spectral density (PSD) of any signal is measured in RF to quantify its spectrum in a broadcasting system and is a direct indicator of OOB signal distortion. The PSD of a single subcarrier OFDM signal is given below

$$\text{PSD}(f) = T \left(\frac{\sin(\pi f NT)}{\pi f NT} \right)^2. \quad (5)$$

The overall power spectral density of the modulated data is the sum of the power spectral densities of all the carriers.

2) *Modulation error ratio for the IB distortion analysis:* The modulation error ratio (MER) is the most widely used figure of merit for system performance in the broadcasting community as it indicates the IB signal deterioration even before the BER result turns bad.¹ In frequency-domain it is defined in dB as

$$\text{MER}(\mathbf{X}, \hat{\mathbf{X}}) = 10 \log_{10} \left(\frac{\|\mathbf{X}\|_2^2}{\|\mathbf{X} - \hat{\mathbf{X}}\|_2^2} \right), \quad (6)$$

where \mathbf{X} is the ideal symbol vector measured at the input of the amplifier and $\hat{\mathbf{X}}$ is measured at the output of the PA.

III. OVERVIEW OF TONE RESERVATION AND DIGITAL PREDISTORTION ALGORITHMS

A. PAPR Reduction using the TR-based Algorithm

In TR, a subset of subcarriers is reserved to generate a kernel signal in the time domain, which is then added to the original one in order to lower its PAPR [5]. These reserved subcarriers are called peak reduction tones (PRTs). Let us consider that we reserve R tones for PAPR reduction of the OFDM system of N subcarriers, where $R \ll N$. We define \mathcal{B} as the PRT subset of these R locations in set $\mathcal{N} = [0, 1, \dots, N - 1]$ and \mathbf{C} as the vector of R peak reduction symbols transmitted on these positions and zeros elsewhere. Similarly, let the complement set \mathcal{B}^c be the data tone (DT) subset of the useful data positions also in \mathcal{N} and \mathbf{D} the vector of the $N - R$ associated transmitted data symbols and zeros elsewhere. The DT and PRT sets are disjoint, i.e. $\mathcal{B} \cap \mathcal{B}^c = \emptyset$ and $\mathcal{B} \cup \mathcal{B}^c = \mathcal{N}$. The resulting signal to be transmitted can be represented in frequency and time domains as

$$\mathbf{X} = \mathbf{C} + \mathbf{D} \xrightarrow[\mathcal{F}^{-1}]{\mathcal{F}} \mathbf{x} = \mathbf{c} + \mathbf{d}, \quad (7)$$

¹MER is closely related to error vector magnitude (EVM) [11]. When both are defined in dB, $\text{MER} = -\text{EVM}$.

where, for $0 \leq k \leq N - 1$, k^{th} element of \mathbf{C} is 0 for $k \in \mathcal{B}^c$ and k^{th} element of \mathbf{D} is 0 for $k \in \mathcal{B}$. Also, \mathcal{F}^{-1} is the OFDM demodulation function. On the receiver side, only data tones are considered to recover the transmitted data. Hence, TR requires no side information. Then, the TR optimization

TABLE I
SIZE OF R FOR DIFFERENT MODES IN DVB-T2 AND ATSC 3.0[†].

MODE	1K	2K	4K	8K	16K	32K
N	1024	2048	4096	8192	16384	32768
R	9	18	36	72	144	288

[†] ATSC3.0 has only 8K, 16K and 32K modes.

problem can be stated as below

$$\mathbf{c}^{opt} = \arg \min_{\mathbf{c} \in \mathbb{R}^N} \|\mathbf{d} + \mathbf{c}\|_{\infty}^2. \quad (8)$$

In DVB-T2 and ATSC 3.0 standards, 1% of the FFT size is allocated for TR. The number of reserved tones for PAPR reduction (i.e., PRTs) for different OFDM FFT sizes as specified by the DVB-T2 and ATSC 3.0 standards are given in Table I.

The TR scheme can be achieved using the clipping control method based on tone reservation (CC-TR) [12]. This algorithm generates iteratively peak-canceling signals by clipping in the time domain and filtering in the frequency domain until the desired PAPR reduction is achieved. The predefined threshold A is calculated based on the mean amplitude of the OFDM signal over M symbols as shown below

$$A = \frac{\beta}{M \cdot N} \sum_{m=1}^M \sum_{i=1}^N |d_i^{(m)}|, \quad \beta \in \mathbb{R}_+, \quad (9)$$

where $|d_i^{(m)}|$ is a set of absolute values of all time samples in the m^{th} data symbol $\mathbf{d}^{(m)}$ and β is a multiplicative factor. First we initialize, $\mathbf{x} = \mathbf{d}$. During the r^{th} iteration and for $0 \leq k \leq N - 1$, the clipped signal $\mathbf{x}_r^{clip} = [\tilde{x}_0, \tilde{x}_1, \dots, \tilde{x}_{N-1}]$ can be obtained by

$$\tilde{x}_k = \begin{cases} x_k, & |x_k| \leq A, \\ A \cdot e^{j\phi_k}, & |x_k| > A. \end{cases} \quad (10)$$

Then, the clipping noise \mathbf{e}_r^{clip} is defined as $\mathbf{x} - \mathbf{x}_r^{clip}$. Then, we filter out the frequency samples in DT locations in \mathbf{e}_r^{clip} and obtain the clipping error vector \mathbf{e}_r^{tr} . At the end of r^{th} iteration, the clipping signal vector for the next iteration \mathbf{c}_{r+1}^{tr} , is computed through the following recurrence relation

$$\mathbf{c}_{r+1}^{tr} = \mathbf{c}_r^{tr} + \mu_r^{tr} \mathbf{e}_r^{tr} = \mu_r^{tr} \sum_{s=1}^r \mathbf{e}_s^{tr}, \quad 1 \leq r \leq Q, \quad (11)$$

where $\mathbf{c}_1^{tr} = \mathbf{0}$, μ_r^{tr} is the r^{th} iteration TR convergence factor ranging between $[0, 1]$, is the TR convergence factor. This clipping signal is added to the data signal during each iteration until sufficient PAPR reduction performance is achieved.

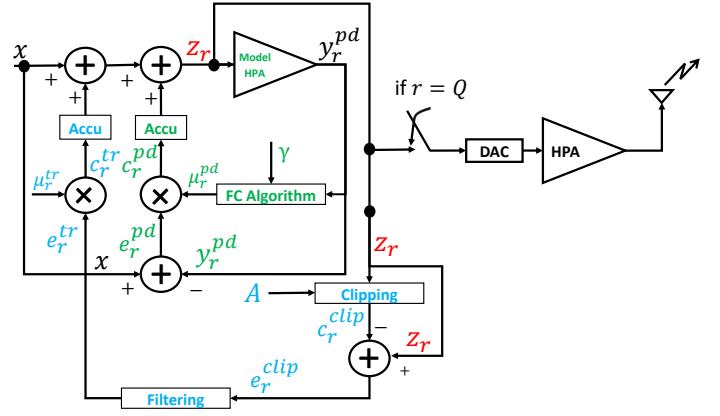


Fig. 1. Block diagram of the proposed scheme, where the correction signal is constructed in an iterative manner in the time domain.

B. PA linearization with FC-DPD

A predistorter approximates conversion characteristics complementary to the conversion characteristics of the PA to ensure that the output of the cascade of the predistorter and the PA has little to no distortion. When predistortion is applied to the baseband symbols before up-conversion, it is referred as DPD. The general objective of any predistorter can be mathematically written as

$$\mathbf{y} = \mathbb{A}(\mathbb{P}\mathbb{D}(\mathbf{x})) = g_s \mathbb{L}(\mathbf{x}), \quad (12)$$

where \mathbb{L} is a linear operator and we recall g_s is the small signal gain of PA.

The predistorter equalizes the PA, which is a non-linear dynamic system. If $\mathbb{L} = \mathbf{I}$, the identity operator, the predistorter $\mathbb{P}\mathbb{D}$ is pre-inverse of the PA, i.e. $\mathbb{P}\mathbb{D} = \mathbb{A}^{-1}$.

The Bussgang theorem states that when a Gaussian stationary process passes through a memoryless NL device, the cross-correlation function of input and output is proportional to the auto-correlation function of input [13]. This theorem facilitates the formulation of DPD as an adding signal technique.

$$\mathbb{P}\mathbb{D}(\mathbf{x}) = \mathbf{z} = \mathbf{x} + \mathbf{c}^{pd}, \quad (13)$$

where \mathbf{z} is the predistorted signal and \mathbf{c}^{pd} is the correction signal vector for DPD. Therefore, the DPD optimization problem can be mathematically formulated as

$$\mathbf{z}^{opt} = \arg \min_{\mathbf{z} \in \mathbb{R}^N} \|g_s \mathbf{x} - \mathbb{A}(\mathbf{z})\|_2^2, \quad (14)$$

We use the FC-DPD technique for DPD [6]. This scheme was originally proposed for a post-OFDM waveform [7] and later extended to broadcasting systems in [6]. In FC-DPD, the correction signal \mathbf{c}^{pd} in (13) is constructed in an iterative manner with error compensation being done in time-domain. At the end of r^{th} iteration, the correction signal vector for the next iteration \mathbf{c}_{r+1}^{pd} , is computed through the following recurrence relation

$$\mathbf{c}_{r+1}^{pd} = \mathbf{c}_r^{pd} + \mu_r^{pd} \mathbf{e}_r^{pd}, \quad 1 \leq r \leq Q, \quad (15)$$

Algorithm 1 The FCJO algorithm

Inputs

- 1: \mathbf{x} : input signal
- 2: \mathbb{A} : Amplification function
- 3: \mathcal{B} : PRT set
- 4: Q : maximum number of iterations
- 5: γ : tightness factor
- 6: β : multiplicative factor to vary A

Output

- 1: \mathbf{z} : predistorted signal

Steps

- 1: **procedure** FCJO(\mathbf{x} , \mathbb{A} , \mathcal{B} , Q , γ , β)
 - 2: initialize $\mathbf{c}_1^{pd} = \mathbf{0}$, $\mathbf{c}_1^{tr} = \mathbf{0}$
 - 3: calculate A as per (9)
 - 4: **for** $r = 1; r \leq Q; r++$
 - 5: calculate input signal: $\mathbf{z}_r = \mathbf{x} + \mathbf{c}_r^{pd} + \mathbf{c}_r^{tr}$
 - 6: calculate PA output signal: $\mathbf{y}_r^{pd} = \mathbb{A}(\mathbf{z}_r)$
 - 7: calculate error signal: $\mathbf{e}_r^{pd} = \mathbf{x} - \mathbf{y}_r^{pd}$
 - 8: compute μ_r^{pd} as per (16)
 - 9: accumulate correction signal: $\mathbf{c}_{r+1}^{pd} = \mathbf{c}_r^{pd} + \mu_r^{pd} \mathbf{e}_r^{pd}$
 - 10: calculate clipped signal: $\mathbf{c}_r^{clip} = clip(\mathbf{z}_r, A)$
 - 11: calculate clipping error signal: $\mathbf{e}_r^{clip} = \mathbf{z}_r - \mathbf{c}_r^{clip}$
 - 12: filter clipping error signal: $\mathbf{e}_r^{tr} = filter(\mathbf{e}_r^{clip}, \mathcal{B})$
 - 13: accumulate clipping signal: $\mathbf{c}_{r+1}^{tr} = \mathbf{c}_r^{tr} + \mu_r^{tr} \mathbf{e}_r^{tr}$
 - 14: **if** $r = Q$
 - 15: calculate: $\mathbf{z}_r = \mathbf{x} + \mathbf{c}_r^{pd} + \mathbf{c}_r^{tr}$
 - 16: **end**
 - 17: **end**
 - 18: **end procedure**
-

where $\mathbf{c}_1^{pd} = \mathbf{0}$, μ_r^{pd} ranging between $[0, 1]$, is the DPD convergence factor, $\mathbf{e}_r^{pd} = \mathbf{x} - \mathbb{A}(\mathbf{x} + \mathbf{c}_r^{pd})$ is the error vector during the r^{th} iteration and Q is the number of iterations. The varying convergence factor at each r^{th} iteration is calculated based on another parameter γ referred as tightness factor, also ranging between $[0, 1]$:

$$\mu_r^{pd} = \frac{\gamma}{\min(J_k) + \max(J_k)}, \quad 1 \leq r \leq Q, \quad 0 \leq k \leq N - 1, \quad (16)$$

where J_k is the k^{th} diagonal element in the Jacobian matrix of the memoryless PA model output $\mathbb{A}(\mathbf{z}_r)$ w.r.t. \mathbf{x} and its general expression is given in [6] as

$$J_k = \frac{1}{2} \left(\frac{\partial \mathbb{G}(z_k)}{\partial |z_k|} + \frac{\mathbb{G}(z_k)}{|z_k|} \right), \quad 0 \leq k \leq N - 1. \quad (17)$$

Thus, μ_r^{pd} is a function that varies over γ as well as the Jacobian J_k at each signal sample. FC-DPD has an extra block FC algorithm and includes an additional parameter, i.e., the tightness factor γ in order to optimize the convergence speed. In [6], it was proven that finding μ_r^{pd} is a convex problem and the FC-DPD has good convergence.

IV. FAST CONVERGING JOINT OPTIMIZATION (FCJO)

We propose that the objective of PA linearization and PAPR reduction can be combined in a collaborative manner by

constructing the correction signal \mathbf{c} in a ping-pong manner through Q iterations. The block diagram of the proposed FCJO technique is presented in Fig. 1. In this figure, it can be noticed that the CC-TR and FC-DPD chain loops are shown in ‘blue’ and ‘green’ colors, respectively. Since, FCJO is an iterative scheme, in order to reduce the complexity the linearization is performed by a PA model in the baseband. The predistorted signal at r^{th} iteration, denoted as \mathbf{z}_r is calculated as

$$\mathbf{z}_r = \mathbf{x} + \mathbf{c}_r^{pd} + \mathbf{c}_r^{tr}, \quad 1 \leq r \leq Q, \quad (18)$$

where \mathbf{c}_r^{pd} and \mathbf{c}_r^{tr} are correction signals from the TR and DPD chains at the end of each iteration, respectively. It is important to note that these two correction signals are computed as per (11) and (15). Thus, by doing this collaboration, we create a symbiotic effect between the two chains in order to avoid complementary effects leading to a better trade-off between PA efficiency and linearity that we can obtain by using the considered PAPR reduction and DPD techniques. However, one critical necessity of this proposed scheme is that for optimal performance, there must be no timing offset between the TR chain and the DPD chain. The FCJO method’s algorithm is summarized in Algorithm 1.

V. SIMULATION RESULTS

This section deals with the performance of the proposed algorithm in order to optimize the choice of the parameters involved in the FCJO algorithm. A Rapp model with a low value of the knee factor p has been used for the simulations to model the PA. This implies that we test the ability of our proposed scheme under a very high NL scenario. Accordingly, $p = 1$, $v_{sat} = 1$ V, $g_s = 1$. Simulations are done over 10^6 OFDM QPSK complex symbols. When a TR algorithm reaches its optimal performance pertaining to its constraints, then its MER variation with FFT size becomes negligible as shown in [14]. Thus, even though we are confined to 1K mode DVB-T2 system in our analysis, it can be extended to all modes of DVB-T2 and even to ATSC 3.0 systems. In the simulations, MER at r^{th} iteration is calculated as $MER(\mathcal{F}^{-1}(\mathbf{x}), \mathcal{F}^{-1}(\mathbf{z}_r))$. From Table I we can find that PRT size is 9.

A. Optimization of TR and DPD parameters

As done in [9], we keep $\mu_r^{tr} = 1$, in order to keep a maximal performance of power efficiency and push the DPD chain in the next iteration to offer its best performance. Moreover, μ_r^{pd} computation depends on γ . The FCJO algorithm is run with $Q = 10$ even though it is not interesting in terms of complexity in order to visualize with more clarity the joint optimization performance for different values of β and γ .

First, we choose a γ and by keeping it constant, we vary β . The simulation results are plotted in Fig. 2 where we can notice that for a given γ , indeed, there exists a certain value of β where FCJO offers maximum MER. When β is too small, then the TR adding signal \mathbf{c} does not approximate the clipping noise, thereby, leading to severe MER degradation. When β increases and gets close to its optimal value then we see a

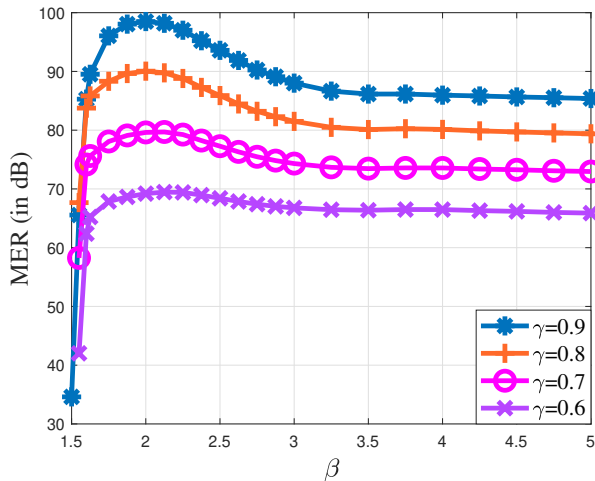


Fig. 2. Optimization of β and γ in FCJO with $Q = 10$ for a QPSK DVB-T2 system in 1K mode for Rapp model PA, $p = 1$ at IBO=6 dB.

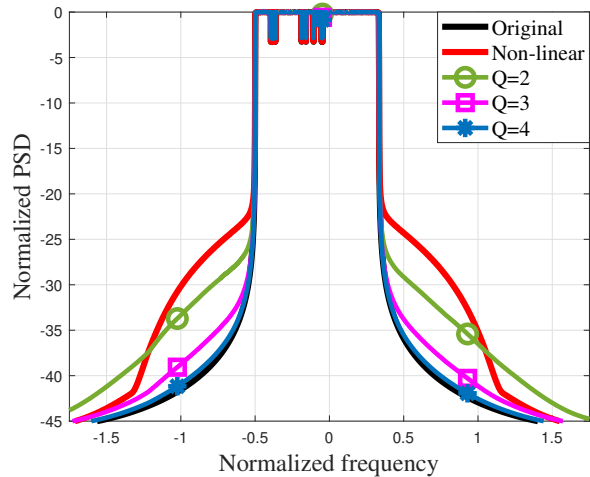


Fig. 4. Impact of Q on the PSD with FCJO ($\beta = 2$, $\gamma = 0.9$) for a QPSK DVB-T2 system in 1K mode for Rapp model PA, $p = 1$ at IBO=6 dB.

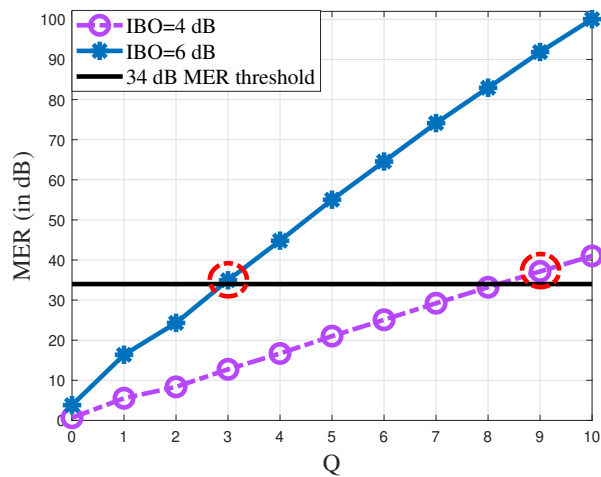


Fig. 3. Impact of IBO on the MER with FCJO ($\beta = 2$, $\gamma = 0.9$) for a QPSK DVB-T2 system in 1K mode for Rapp model PA, $p = 1$.

step increase in MER. This indicates that at this juncture, there is good synergy between TR and DPD chains. On the contrary, when β is too large, the peak-canceling signal \mathbf{c} is close to zero, leading to very little PAPR reduction during each iteration. In such cases, improvement in MER is solely due to DPD. That phenomenon can be observed from the figure where the MER seems to stabilize for large values of β . Though not plotted in the Fig. 2, during the simulation it was found that for $\gamma > 0.9$ the FCJO algorithm is under-performing w.r.t. than i.e. $\gamma = 0.9$. Therefore, it has been found heuristically that $\beta = 2$ and $\gamma = 0.9$ yields optimal performance for the considered IBO and p of the PA model.

B. Impact of IBO on MER Optimization

The Bussgang theorem can be applied only when IBO is sufficiently large. We consider two IBO values 4 and 6 dB and evaluate the FCJO w.r.t. different number of iterations, i.e. $Q \in [1, 10]$. The MER performance for different Q is plotted in Fig. 3. In DVB-T transmission, an MER less than 34 dB is deemed to induce transmission failure, and the same can be inferred in the case of DVB-T2 systems [14]. Moreover, a target value of MER above 34 dB also assures that the transmitter RF coverage is almost similar to that of the theoretically achievable RF coverage limit. Then, the performance of FCJO at different IBO values is presented in Fig. 3 where the proposed scheme is compared with the NL amplified signal without any PAPR reduction and PA linearization. We can notice in Fig. 3 that with 6 dB IBO we need only 3 iterations to achieve the targeted MER, whereas by lowering it further to 4 dB we see that FCJO performance is severely compromised, requiring 9 iterations to achieve the same target.

C. OOB Analysis

The FCJO algorithm was run with $Q = \{2, 3, 4\}$ iterations, and respective PSD plots are shown in Fig. 4. In the legend of figures, “Linear”, and “NL” indicate normalized PSD at the input and output of the Rapp model of PA, without any predistortion. Then, all remaining ones shown in the legend are jointly optimized and then non-linearly amplified. In PSD simulations, we have considered oversampling factor $L = 4$ in order to visualize the OOB spectrum. As observed from Table II with just 4 iterations, we can see that regrowth of the OFDM signal with FCJO at a normalized frequency of 1 is 0.63 dB only. This implies that FCJO can linearize the OFDM signal quickly.

TABLE II
OOB ANALYSIS FOR DVB-T2 SYSTEM WITH 1K MODE WITH FCJO AT A
NORMALIZED FREQUENCY OF 1 FOR DIFFERENT Q.

Q	Normalized PSD with FCJO [†]	Spectral regrowth w.r.t. linear case	Spectral degrowth w.r.t. NL case
2	-33.33 dB	8.20 dB	3.14 dB
3	-38.74 dB	2.79 dB	8.55 dB
4	-41.53 dB	0.63 dB	10.71 dB

[†] Normalized PSDs in linear and NL cases at a normalized frequency of 1 are -41.53 dB and -30.19 dB respectively.

D. Complexity Aspects

The proposed scheme's complexity is almost the same as that of the CC-TR and FC-DPD schemes except for the fact that both these chains need to be in good synchronization for correct calculation of predistortion signal \mathbf{z}_r at every iteration.

E. Comparison of P2JO and FCJO schemes

We recall that the PAPR reduction chain involving TR is identical for both P2JO and FCJO. However, the DPD chain in P2JO involves an artificial neural network (ANN) while FCJO uses an adaptive filtering technique involving a recursive error correction mechanism. Therefore, a direct comparison of both P2JO and FCJO schemes in terms of performance, complexity, and latency is not straightforward.

P2JO was analyzed using the output back-off (OBO) which is defined as the ratio between the PA output saturation power to the mean power of the output signal. We found that 6 dB IBO in our simulations roughly corresponds to 4.57 dB OBO and therefore it is fair to have a rough comparison of linearization performance comparison of P2JO at 4.5 dB OBO with FCJO at 6 dB IBO as summarized in Table III. We can notice that FCJO outperforms P2JO due to varying convergence factor at each iteration and also achieves the threshold MER in just 3 iterations. Moreover, it is important to note that P2JO and FCJO have the PRT size 32 out of 256 sub-carriers and 9 out of 1024 sub-carriers respectively. Therefore, the former is not compatible with the standard due to 12.5% PRT size while the latter obeys the standard with 1% PRT size.

TABLE III
PERFORMANCE COMPARISON OF THE P2JO AND FCJO SCHEMES.

P2JO ($\mu_r^{tr} = 1, \mu_r^{pd} = 0.02$) [‡] 4.5 dB OBO		FCJO ($\mu_r^{tr} = 1, \text{varying } \mu_r^{pd}$) [§] 6 dB IBO (≈ 4.57 dB OBO)	
MER	Iterations	MER	Iterations
15.91 dB	10	16.37 dB	1
18.06 dB	≥ 20	34.95 dB	3

[‡] 16-QAM, 256 FFT, Rapp ($p = 1.1$) and TR ($R = 32$) [9].

[§] QPSK, 1K FFT, Rapp ($p = 1$) and TR ($R = 9$).

VI. CONCLUSION

In this paper, we proposed an adding signal scheme for joint optimization of PAPR reduction and digital predistortion for next-generation broadcasting systems. This scheme offers

good linearization performance with fast convergence and outperforms the recently proposed P2JO scheme. The simulation results infer that a varying convergence factor can be beneficial to maintain ANN-based DPD level performance, with lower joint optimization latency. Moreover, the proposed approach accords with broadcasting standards-compliant and requires no special RF circuits. Having shown some promising results, we will explore ML-based tightness factor estimates over heuristic search in our future works.

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