Digital Predistortion Concepts for Linearization of mmW Phased Array Transmitters

Nuutti Tervo, Markku Jokinen, Marko E. Leinonen, Janne Aikio, Olli Kursu,
Timo Rahkonen, and Aarno Pärssinen

Faculty of Information Technology and Electrical Engineering (ITEE), University of Oulu, Finland

Abstract—Linearization of millimeter-wave (mmW) phased arrays is one of the key enablers for improving the system performance in terms of power, efficiency and linearity. However, phased array transceiver topologies that have multiple parallel nonlinear components with a shared digital input challenge the standard digital predistortion techniques. In addition, different analogue beamforming techniques complicate the linearization even further due to the fact that the signal nonlinearity has to be observed or modelled over-the-air (OTA) together with the impacts of antennas and even the directive mmW radio channel. The best linearization strategy depends on the system level targets of linearity such as error vector magnitude and adjacent channel power ratio which have slightly different nature when observed in the radiated far-field. In this paper, we present our view and the status of the literature on the topic of phased array digital predistortion. We highlight that the nonlinear distortion have a beam shape which may be different from the linear part of the beam. We also review the antenna array figures of merit describing the nonlinearity. Finally, we show an experimental example of OTA linearization of a 28 GHz phased array transmitter.

Keywords—ACPR, antenna array, beamforming, linearization, over-the-air, 3GPP, 5G.

I. INTRODUCTION

Third generation partnership project / new radio (3GPP/NR) [1] has standardized several millimeter-wave (mmW) frequency bands from 24.25 GHz up to 52.5 GHz to be used in fifth generation (5G) above 6 GHz (FR2) systems. The three mostly referred bands are at 26 GHz (n258), 28 GHz (n257), and 39 GHz (n260) and several variations of these are expected depending on the regional frequency allocations. Antenna arrays with hundreds of elements fed by analogue beamforming networks are used to provide decent spatial coverage and directive beams for multiple spatially separated users [2]. High-order modulations and orthogonal frequency division multiplexing (OFDM) -based waveforms selected by 3GPP guarantee that peak-to-average-power-ratio (PAPR) will remain relatively high also in 5G mmW systems [1]. High PAPR waveforms are known to require highly linear transceivers to provide decent error-vector-magnitude (EVM) and hence bit-error-rate (BER).

System level figures of merit (FOMs) describing the nonlinearity are traditionally expressed in terms of EVM and adjacent channel power ratio (ACPR). The EVM describes the modulation accuracy while the ACPR is the FOM for distortion to the systems and devices operating at the adjacent frequency bands. Transmitter (Tx) end especially power amplifier (PA) efficiency and linearity are known to be inversely proportional

to each other. Linear PAs operating in backoff larger than the PAPR of the waveform can provide decent EVM and ACPR. However, that often results in low efficiency, which means high dissipated power that needs to be conducted away from each element [3]. Hence, efficiency of multiple parallel elements in mmW systems should be improved to provide smaller power consumption, more compact transceiver form-factor and longer device life-time. In addition, limited power delivery capability of integrated PAs in large scale arrays limit the achievable effective isotropic radiated power (EIRP). However, with proper linearization the PA can be driven with higher output power and closer to the saturation that is directly enhancing the coverage of the cell or link range of the mobile device.

In sub-6GHz systems (FR1), the Tx EVM and ACPR are characterized by conducted measurements at each antenna connector [4]. However, in mmW phased arrays where the front-ends are integrated with the antenna array, it is no longer possible to access the radio frequency (RF) output port of each nonlinear branch. That is why 3GPP/NR has specified that EVM and ACPR of mmW 5G transmitter have to be characterized using OTA measurements [4]. 3GPP/NR specifies adjacent channel power (ACP) as total radiated power (TRP) integrated or approximated over the space [4], [5], [6]. This means that ACP direction does not matter for the specifications and hence only the total amount of radiated ACP compared to the total amount of TRP channel power counts.

For 3GPP/NR FR2, the ACPR specifications are from -26 to -28 dBc which already means quite nonlinear behavior. Furthermore, due to the inaccuracies in OTA testing and TRP calculations, the test specifications [4] propose even more relaxed ACPR limit from -23.4 to -25.7 dBc. For such relaxed ACPR targets more nonlinear PA classes can be used especially if some linearization methods can be utilized.

Digital predistortion (DPD) is one of the most commonly used linearization techniques in lower-frequency systems to push the PA closer to the nonlinear region where good efficiency can be achieved. However, mmW phased arrays make the use of DPD challenging due to the fact that multiple parallel nonlinear elements share only one or just a few digital signal inputs. Hence, in phased array DPD, multiple parallel elements have to be linearized with one DPD. Array linearization by a single DPD has been presented and studied e.g. in [7]-[14]. Traditional DPD requires Cartesian feedback receiver and several alternative feedback approaches have been presented all the way from switchable feedback [7], [8], [9]

Fig. 1. Sources of nonlinearity in phased array transmitters.

to combined channel-emulating feedback [6], [12] and even over-the-air (OTA) feedback [13], [15]. Some studies e.g. [10] suggest that antenna coupling should be taken into account if the nonlinearity of the phased array is desired to be modelled by using individual PA measurements.

Few papers e.g. [6], [8], [11], [12] present DPD methods for linearizing the array in the intended beamforming direction, but this does not necessary guarantee the total radiated adjacent channel power ratio (TRACPR) performance. On the other hand, EVM describing the modulation accuracy is specified in the intended beamforming direction and it benefits from the directive linearization schemes as shown in [6]. Hence, there is a trade-off between the total radiated ACPR and directive EVM if both are desired to be improved by the DPD.

In this paper, we address the challenges, the benefits and the system level principles of phased array DPD. In Section II the phased array nonlinearity and the OTA characterized array FOMs are revised. In Section III, we discuss on the existing array DPD techniques. Finally, in Section IV we show an experimental example of DPD of a 28 GHz 64-element phased array Tx [16], [17] having 16 parallel Gallium Nitride (GaN) PAs.

II. PHASED ARRAY FIGURES OF MERIT OF LINEARITY

A. Beamformed Distortion

General overview of the dominant sources of nonlinearity of the phased array transmitter is presented in Fig. 1. Roughly, we can divide the nonlinearity to common and parallel parts depicted in the figure. In general, common part of the nonlinearity does not change the beam shape over the space due to the fact that the common part is fully correlated over the antennas (same signal). Similarly, if the waveform is assumed to experience identical nonlinearity in all parallel RF branches, the beam of the nonlinear distortion is similar to the beam of the linear signal [18]. This could be a reasonable assumption, if we assume that the branches are equipped only with identical phase shifters, and the antenna array, PAs, power division network etc. are ideal. However, in practice, the branches are not the same and variations in gain and nonlinearity may occur e.g. due to gain imbalance in power division network, phase shifter control word dependent gains, manufacturing tolerances, impedance matching etc. [6]. Also, in highly compact mmW phased array designs, circulators are not used to isolate the PA outputs from the antennas [19]. The

variations in antenna matching [20] has a direct impact on the PA behavior and may depend even on the phase shifter control [10]. Furthermore, in [21] it is discussed that also strong enough antenna coupling could cause differences in the nonlinearities of individual PAs. Finally, even if all the previously mentioned effects were ignored, the branch-specific gain control, used to shape the beam (beamforming), to create nulls or reduce the sidelobes [8], makes each nonlinear path different from each other.

If the nonlinear characteristics of the individual PAs are not identical, the beam pattern of the nonlinearity differs from the linear beam. Hence, the nonlinearity of an array depends on the direction of observation. Still the maximum of distortion is often in the direction of the main lobe [18], but one cannot say that the beam of the nonlinear distortion is exactly the same as the beam of the linear part of the signal. This has been further discussed in [8] where the authors show that these differences can be actually utilized to compensate each other in the array far-field. Furthermore, measurement campaign presented in [22] and simulations in [6] show that the variations in PA characteristics over the parallel branches cause averaging effect that may slightly improve the linearity at the main lobe even without linearization. Hence, the standard figures of merit describing the nonlinear distortion of an array should be revised in order to have the right meaning from the system perspective. [23]

B. In-band Linearity and EVM

Millimeter-wave systems mostly rely on line-of-sight (LOS) communications and hence it is often assumed that only one dominant propagation path is present. Hence, beamsteering and maximum ratio transmission (MRT) are the most commonly used beamforming methods. In analogue beamforming, these two methods are basically the same in LOS channel. Hence, in the most practical use case, each user communicates to a dedicated analogue beam that the base station (BS) has steered to the direction of the user. As the root mean square (RMS) EVM describes the errors in the actual modulated data, it should be measured in the steering angle and that is how it is specified in 3GPP/NR [4]. EVM measures not only nonlinearity, but also the errors caused by IQ-imbalance, phase noise, quantization, etc. Especially the phase noise in mmW transmitters sets practical limits to the achievable EVM. As the EVM of the nonlinear system is dependent on power, the EVM is traditionally characterized with maximum transmit power i.e. at the cell edge [24]. 3GPP/NR FR2 Tx EVM requirements of different modulations are collected to the Table 1. The specification for 256 - quadrature amplitude modulation (QAM) is taken from the FR1 as the 3GPP/NR EVM specification for 256-QAM does not exist by the time of writing.

C. Total Radiated ACP

Out-of-band distortion is fundamentally interference to other users and systems that are operating at the adjacent frequency bands. In general, this means that the array ACP is

Table 1. 5G NR Tx EVM specifications of different modulation schemes [5].

Modulation	EVM [%]	EVM [dB]
16-QAM	12.5	-18
64-QAM	8	-22
256-QAM	3.5	-29.4

a severe problem only if the system or user under interference is located in the direction of the radiated ACP. However, in practice the system is not often aware of the direction of the adjacent channel user or system. Thus, a practical way to specify ACP is to integrate it over the PA elements, or equivalently integrate it over the three dimensional space. TRP for the discrete set of directions can be calculated as

$$P_{TRP} = \frac{1}{2N_{az}N_{el}} \sum_{n=0}^{N_{az}-1} \sum_{m=0}^{N_{el}-1} (\text{EIRP}_{\phi}(\phi_n, \theta_m) + \text{EIRP}_{\theta}(\phi_n, \theta_m)) \sin \phi_n, \quad (1)$$

where EIRP_{ϕ} and EIRP_{θ} denotes the EIRPs in ϕ and θ polarizations, ϕ and θ denotes the azimuth and elevation angles with discrete angular grid over N_{az} and N_{el} values, respectively. TRACPR can be calculated as a difference between total radiated ACP (TRACP) and total radiated channel power (TRCP) as

$$\text{TRACPR} = \max\{\text{TRACP}_H, \text{TRACP}_L\} - \text{TRCP}, \quad (2)$$

where the power is expressed in dBm and subindices H and L denote the power of higher and lower adjacent channels, respectively.

The accuracy of the TRP measurement depends on the angular resolution of the measurement as well as the accuracy of the power detectors and reference antennas used. Due to the measurement inaccuracy, a measurement margin of 2.3 – 2.7 dB is subtracted from the TRACP measurement which relaxes the specifications even further [4]. Hence, the system level requirements and test specifications are not the same. The 5G mmW test specifications (spec.) and Tx ACP requirements (req.) of different FR2 frequency bands are collected to Table 2.

Table 2. 5G NR Tx ACPR test specifications and requirements [4], [5].

frequency band	Tx ACPR test spec. [dBc]	Tx ACPR req.
24.25 – 33.4 GHz	-25.7	-28
37 – 52.6 GHz	-23.4	-26

III. PHASED ARRAY LINEARIZATION BY DIGITAL PREDISTORTION

Two main strategies has been proposed in [8] to linearize a phased array with a single DPD. One can either try to linearize the individual PAs e.g. in least squares (LS) sense, or try to linearize the array response to a desired spatial direction. Different feedback strategies can be used to collect the non-linear output of individual PAs. If the nonlinear elements are very similar, single PA feedback [13] may give reasonable

linearization results and can be used for linearization. Due to its simplicity, it may provide a low-complexity solution for linearization. A single feedback path can be also shared in time with multiple PA branches by switching each branch to the feedback path one at a time. This has been proposed and used e.g. in [7], [8], [9] where the linearization object is build based on the single PA outputs by using the well-known array factor principle [25]. However, switchable feedback architecture makes the DPD training time consuming and the phase differences over the branches are challenging to model in different time instants by using limited sampling rates. The individual PA paths can be also combined in analogue domain by applying feedback weights that emulate the radio channel in a certain spatial direction. This has been proposed for example in [6] and [12]. When using such a feedback scheme, traditional DPD techniques can be used to train the array and the DPD algorithm itself does not have to model individual PA outputs or their differences. However, analogue combining may have amplitude and phase imbalance which has a direct impact on the accuracy of the DPD object and hence the DPD performance.

Directive linearization schemes introduced in [6], [8], [12] are known to provide linearization in the beamforming angle. In [9], a method for widening the linearization angle is proposed and the concept is verified by experimental results. Hence, the linearization performance observed over the space depends significantly on the feedback strategy and hence the DPD method. An example of the experimental OTA linearization performance is given in the following chapter.

IV. EXPERIMENTAL EXAMPLE WITH 28 GHz PHASED ARRAY TRANSMITTER

A. OTA DPD Training and Measurement Platform

One way to show the meaning of OTA DPD is to place the receiver antenna to the array far-field and use it as an observation receiver. Experimental example of OTA DPD was carried out in an anechoic chamber. The phased array under test (AUT) [16], [17] is a 64-element, 16-chain phased array equipped with Qorvo TGA2595 PAs operating at 28.1 GHz center frequency. Each branch is equipped with a 5-bit phase shifter and each branch drives a 2x2 element subarray [26]. The details of the AUT are given in [16]-[17] and thus not repeated here. A photograph of the measurement arrangement in the chamber is shown in Fig. 2. In the Tx side, the measurement setup consist of Keysight M8190A arbitrary waveform generator (ARB), E8257B PSG signal generator. A-info LB-28-15 standard gain horn antenna, followed by a CA2630-141 pre-amplifier and N9040B UXA signal analyzer are used as the observation Rx. The AUT is placed on top of a rotary table controlled by stepper motors. All measurement equipment and AUT are controlled by MATLAB-based control software. The example DPD performance is measured using 100 MHz wide 256-QAM cyclic prefix (CP)-OFDM downlink waveform. The waveform is generated based on the 3GPP/NR standard.

Fig. 2. OTA DPD measurement setup placed in an anechoic chamber.

(a) ACPR at the steering angle w/o DPD

(b) ACPR at the steering angle w DPD

Fig. 3. OTA measured spectrums at steering angle (a) w/o and (b) w DPD, respectively.

The DPD model is a memory polynomial with memory length of 4 and polynomial order of 9 (in total $4 \times 5 = 20$ coefficients). Only the odd-order nonlinear coefficients are used. The coefficients are trained over 10^4 samples of the four times oversampled input waveform by using LS estimation. In the measurements, the phase shifters are set to zero-phase and the Rx antenna is placed in the main lobe direction at the distance of 2 meters. Figs. 3 (a) and (b) show the ACPR measured at the steering angle (zero-direction) without and with DPD, respectively. We can observe that even relatively simple memory polynomial improves the main lobe ACPR significantly. The measured ACPR is improved from around -30 dB to lower than -40 dB with DPD.

B. Linearized Beam Pattern

The ACP beams were measured by physically rotating the array over the $\phi \in [-45^\circ, 45^\circ]$ with 1° angular resolution. The measured channel powers and ACPs are given in Fig 4. The figure shows that the DPD has only a minor impact on the beam shape of the channel power (P_{CH}). However, the figure also shows that the ACP beam does not fully follow the beam

Fig. 4. OTA measured channel power and ACP beams w and w/o DPD.

Fig. 5. OTA measured ACPRs w and w/o DPD.

of the channel power. This is due to the fact that the parallel nonlinear branches are different. The ACP is reduced in the linearization angle (steering angle), but some of the directions may even suffer from the DPD.

Fig. 5 presents the ACPR beams measured over the azimuth quarter plane. Compared to the Fig. 4, this presents the difference between the channel power and ACP. As we can see, the ACPR is not identical over the plane and it depends naturally on both ACP and channel power beams. The ACPR is improved in the main lobe, but all directions do not experience the same benefit. This is why linearization concepts for antenna arrays have to be rethought if ACPR improvement is desired in all directions.

V. CONCLUSION

Phased array linearization by DPD can improve the linearity and efficiency of the mmW phased array transmitter. However, linearizing multiple elements with a single digital input complicates the DPD design. In this paper, we discussed about the FOMs describing the nonlinearity of the phased array which are the key goals for improving linearity. Highly

integrated and compact mmW phased arrays are preferably characterized by OTA measurements, and therefore the FOMs have to be thought differently than at lower frequencies. It was concluded that EVM and ACPR have totally different meanings when specified by OTA measurements. We also gave a measurement based example of OTA linearization of a 28 GHz phased array transmitter. The measurement results indicate that linearizing the array in one direction does not guarantee that the other directions are improved. This should be carefully thought when designing linearization concepts for mmW phased arrays.

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