Adaptive Envelope Shaping for Low and Medium Power Amplifiers With Dynamic Supply

Pere L. Gilabert, Senior Member, IEEE, Gabriel Montoro, Member, IEEE, Nieves Ruiz, Student Member, IEEE, and José A. García, Member, IEEE

Abstract—This letter presents an adaptive envelope shaping (AES) method to linearize low and medium power amplifiers (PAs) with dynamic supply. This straightforward linearization solution results very useful for small cell or handset transmitters, where reducing power consumption and computational complexity of the digital part is crucial. With the AES method, there is no need of an a priori characterization of the PA to shape the supply voltage signal targeting maximum linearity. Excellent linearization results are obtained when the PA presents good AM-PM linearity, otherwise, additional phase distortion linearization has to be included to meet the ACLR specifications.

Index Terms—Envelope shaping, linearization, power amplifier.

I. INTRODUCTION

Envelope tracking (ET) and envelope elimination and restoration (EER) are supply modulation techniques suitable to complement existing fixed supply transmitter architectures by enhancing the power amplifier (PA) power added efficiency (PAE), critically degraded when using multicarrier and spectrally efficient modulation schemes presenting high peak to average power ratio (PAPR). The overall power efficiency in ET and EER not only depends on both RF PA and envelope amplifier (EA), but also on the computational load introduced by the linearization algorithm, since the power consumption of the digital part is dominant in small cell transmitters.

By using an envelope shaping function [1], the instantaneous supply voltage can be chosen to either achieve optimum efficiency at the cost of having nonlinear distortion at the PA output or, alternatively, to achieve certain levels of linearity at the cost of a small loss of efficiency (e.g., Nujira-Wilson or $N = 6$ shaping in [2]). To minimize the impact of the linearization subsystem (in terms of computational complexity and energy efficiency) in pico/femto cells or handsets, the envelope shaping function aims to compensate for the PA’s AM-AM distortion. This linearization method assumes the AM-PM distortion to be negligible, otherwise, additional phase distortion compensation (e.g., [3]) has to be included in the in-phase (I) and quadrature-phase (Q) path.

To design the envelope shaping function, a previous characterization of the PA is necessary (e.g., gain versus output power curves for different supply voltages). Normally, this characterization is carried out through CW test signals. However, this measurements are not always feasible or reliable, since thermal and bias network equilibrium are different for CW versus modulated signals. For example, PAs based on GaN HEMT transistors experience a soft-compression effect when the AM-AM measurement is made under static conditions [4]. Moreover, certain topologies of the EA do not allow the use of CW excitation due to its small PAPR and the need of stressing the PA with a high mean output power in order to characterize the peaking range. As a consequence, it is best if the extraction of the different gain characteristic curves are done under a dynamic excitation (e.g., LTE waveforms). This ultimately leads to a lack of precision of the characteristic curves (i.e., blurring of the measured data) that impacts the linearization performance of the envelope shaping function (e.g., unbalanced ACLR compensation).
This letter presents an adaptive envelope shaping (AES) method aimed at compensating for the AM-AM distortion that avoids the need of an \textit{a priori} characterization of the PA (i.e., generating an envelope shaping function by interpolating a set of unprecise measurements). Moreover, unlike Nujira’s isogain shaping described in [2], thanks to its closed-loop nature it can be used for operating in both ET and EER modes and it does not need to be pre-adjusted.

II. ADAPTIVE ENVELOPE SHAPING

The block diagram of a power efficient transmitter with dynamic power supply, including the proposed adaptive envelope shaping (AES) subsystem, is depicted in Fig. 1. Considering the baseband complex signal to be transmitted \( u[n] \), the envelope is defined as \( E[n] = \sqrt{u_r^2[n] + u_Q^2[n]} \). The input-output relationship of the envelope shaping function (SF) is defined (see Fig. 1) as

\[
E_s[n] = f_{SF}(E[n]) = E[n] - \varepsilon[n]
\]

where \( \varepsilon[n] \) is the nonlinear distortion signal that, taking also into account memory terms, can be modeled as

\[
\varepsilon[n] = \sum_{i=0}^{N-1} \sum_{p=0}^{P-1} w_{p,i} \cdot (E[n - \tau_i])^p = \varphi_{n} \cdot w_{n} \quad (2)
\]

where \( \tau_i \) (with \( \tau \in \mathbb{Z} \) and \( \tau_0 = 0 \)) are the most significant sparse delays of the envelope. In matrix notation, \( w_n = (w_{0,0}, \ldots, w_{P-1,0}, \ldots, w_{0,N-1}, \ldots, w_{P-1,N-1})^T \) is a vector of coefficients with dimensions \( O \times 1 \), where \( O = P \cdot N \) is the order of the behavioral model in (2), \( P \) is the polynomial order and \( N \) the number of delays; and \( \varphi_{n} = (E[n], \ldots, E[n]^p, \ldots, E[n - \tau_N - 1], \ldots, (E[n - \tau_N - 1])^p) \) is the 1xO data vector containing \( \tau \) basis waveforms. Following a closed-loop error minimization technique, the coefficients in (2) can be extracted iteratively using a weighted least squares algorithm,

\[
w_{n+1} = w_n + \mu (\Phi^H \Phi)^{-1} \Phi^H e
\]

where \( \Phi = (\varphi_0, \varphi_1, \ldots, \varphi_N)^T \) being the \( L \times D \) data matrix, \( L \) is the number of data samples \( (n = 1, 2, \ldots, L) \), and \( \mu (0 \leq \mu \leq 1) \) being the weighting factor. Finally, \( e \) is the \( L \times 1 \) vector of the error defined as

\[
e = \left| \frac{y}{G_0} \right| - E
\]

where \( G_0 \) determines the desired linear gain of the PA, and \( y \) and \( E \) are the \( L \times 1 \) vectors of the PA output and 92 baseband instantaneous envelope, respectively.

Targeting a future FPGA implementation, the AES can be easily carried out using LUTs. Therefore, (1) and (2) can be rewritten as the combination of \( N \) LUTs,

\[
E_s[n] = E[n] - \sum_{i=0}^{N-1} E[n - \tau_i] \cdot G_{LUT, i} (E[n - \tau_i]) + w_{0i}
\]

where \( w_{0i} \) are the offset coefficients that have a big impact on the linearity vs. efficiency trade-off [5].

III. EXPERIMENTAL SETUP AND RESULTS

The experimental test-bench is depicted in Fig. 2. For testing purposes, we used a broadband high efficiency continuous-mode 101 class-J power amplifier at 950 MHz, based on the C03H5030F102 packaged GaN HEMT from Cree Inc. The signal generation and measurement equipment consist of: Texas Instruments 104 boards (TSW1400EVN pattern generator + TSW30H84EVN 105 DACs and I-Q modulator), a Tabor WW2572A arbitrary waveform 106 generator, and a Keysight Infinium DSO9404A oscilloscope 107 for capturing the RF signals. Another oscilloscope is used for 108 capturing the drain voltage and current, and this information 109 is used for calculating the drain power consumption. A PC 110 running Matlab controls all the instrumentation and does all the 111 required digital signal processing. We used as the EA the high- 112 speed (35 MHz bandwidth and 900 V/\mu s slew-rate at Av = 2113 and 10 \Omega load) high-current (1.1 A) Linear Technology IC 114 LT1210. For the sake of simplicity we considered the linear but 115 slightly efficient IC LT1210 as the EA. Thus, the drain biasing 116 voltage could be lowered down to 0 V while the reported 117 PAE values take only into account the consumption at the RF 118 PA drain. The scope of this work is to prove the linearity 119 performance of the proposed AES method. The signal used was an 120 uplink LTE signal of 5 MHz bandwidth and 8.3 dB of PAPR. 121 The targeted ACLR levels to meet the specifications for the 122 LTE uplink channel were -38 dB. Table I shows the ACLR, 123 NMSE and PA’s PAE when using the proposed AES method 124 for different mean output power levels and dynamic supply 125 strategies, namely, ET and EER. To detect the importance of 126 the AM-PM distortion, and thus determine the linearization 127
limits of the AES method, we can compare the NMSE of the modulus \([|y/G_0| - u]\), with \(u\) and \(y\) being the RF PA input and output signals (see Fig. 1), respectively. If the PA shows significant AM-PM distortion (evidenced in the \(\Delta\text{NMSE}\) column in Table I), with only AES linearization may not be enough to meet the ACLR specifications, and additional phase distortion compensation may be required in the I-Q path. However, no mutual optimization of the inputs (supply voltage and RF input amplitude) is done. Figs. 3–6 show the AM-AM, AM-PM and output power spectra for both ET and EER dynamic supply strategies when considering: no shaping (i.e., \(E_s[n] = E[n]\) in (1)), memoryless AES and, AES + Phase DPD (in the case of ET) or AES with memory compensation—4 LUTs—in the case of EER, to compensate for the dynamic dependency on the supply voltage).

**IV. Conclusion**

The proposed AES method is capable of compensating for the AM-AM distortion of PAs with dynamic supply with a single LUT and without any \textit{a priori} characterization of the PA. Experimental results showed that operating in ET mode additional phase distortion compensation was necessary to meet the \(-38\, \text{dB ACLR}\) specifications. Instead, in EER mode, with the class-J PA driven deep into saturation (see the PA gain column in Table I) and operating close to a class-E switched-mode, the AM-PM distortion results less harmful (except for the feedthrough effect at low drain voltage supply levels) and with only AES, in 3 iterations (at least 25 are required in ET mode) the linearization specifications were met. For handsets or low/medium power equipment, the PA has to be carefully designed to show a linear AM-PM characteristic and thus avoiding additional DPD compensation in the I-Q path.

**REFERENCES**


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Manuscript received November 6, 2015; revised March 2, 2016; accepted April 12, 2016. This work was partially supported by the Spanish Government (MINECO) and FEDER under projects TEC2014-58341-C4-01-R and 03-R; by the Catalan Government under project 2014-SGR-1103; and by CATRENE under the project named CORTIF CA116.

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Digital Object Identifier 10.1109/LMWC.2016.2574825
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$$w_{n+1} = w_n + \mu (\Phi^H \Phi)^{-1} \Phi^H \varepsilon$$

with $\Phi = (\varphi_0, \varphi_1, \ldots, \varphi_{P-1})^T$ being the $L \times O$ data matrix, where $L$ is the number of data samples ($n = 1, 2, \ldots, L$), and $\mu$ is the weighting factor. Finally, $\varepsilon$ is the $L \times 1$ vector of the error defined as

$$\varepsilon = \left| \frac{y}{G_0} - E \right|$$

where $G_0$ determines the desired linear gain of the PA, and $y$ and $E$ are the $L \times 1$ vectors of the PA output and PA's PAE when using the proposed AES method.

TABLE I

<table>
<thead>
<tr>
<th>Linearization Scheme</th>
<th>PA Gain (dB)</th>
<th>Supp. Voltage Swing (V)</th>
<th>ACLR (dB)</th>
<th>NMSE (dB)</th>
<th>$\Delta$NMSE (dB)</th>
<th>PA PAE (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>ET</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>No Shaping</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Memoryless AES (1-LUT)</td>
<td>27.0</td>
<td>16.3</td>
<td>13-20</td>
<td>-23.1</td>
<td>0.4</td>
<td>39.3</td>
</tr>
<tr>
<td>+ Memory (4-LUTs)</td>
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<td></td>
<td></td>
<td>-31.9</td>
<td>11.9</td>
<td>33.3</td>
</tr>
<tr>
<td>(1-LUT) + Phase DPD</td>
<td></td>
<td></td>
<td></td>
<td>-32.1</td>
<td>9.7</td>
<td>31.1</td>
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<tr>
<td>Memoryless AES (1-LUT)</td>
<td>31.5</td>
<td>18.0</td>
<td>19-28</td>
<td>-45.4</td>
<td>3.0</td>
<td>33.1</td>
</tr>
<tr>
<td>+ Memory (4-LUTs)</td>
<td></td>
<td></td>
<td></td>
<td>-62.7</td>
<td>14.2</td>
<td>47.2</td>
</tr>
<tr>
<td>(1-LUT) + Phase DPD</td>
<td></td>
<td></td>
<td></td>
<td>-54.2</td>
<td>6.4</td>
<td>44.8</td>
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<tr>
<td>EER</td>
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<td>27.1</td>
<td>7.0</td>
<td>0-20</td>
<td>-38.0</td>
<td>8.8</td>
<td>40.5</td>
</tr>
<tr>
<td>+ Memory (4-LUTs)</td>
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<td></td>
<td></td>
<td>-39.8</td>
<td>10.4</td>
<td>39.5</td>
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<tr>
<td>(1-LUT) + Phase DPD</td>
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<td>31.1</td>
<td>8.1</td>
<td>0-28</td>
<td>-43.0</td>
<td>3.9</td>
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where $w_{0i}$ are the offset coefficients that have a big impact on the linearity vs. efficiency trade-off [5].

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limits of the AES method, we can compare the NMSE of the modulus [where the error is defined as in (4)] with the classical NMSE for complex signals. This latter error is defined as 
\[ e = |(y/G_0) - u|, \]
with \( u \) and \( y \) being the RF PA input and output signals (see Fig. 1), respectively. If the PA shows significant AM-PM distortion (evidenced in the \( \Delta \)NMSE column in Table I), with only AES linearization may not be enough to meet the ACLR specifications, and additional phase distortion compensation may be required in the I-Q path. However, no mutual optimization of the inputs (supply voltage and RF input amplitude) is done. Figs. 3–6 show the AM-AM, AM-PM and output power spectra for both ET and EER dynamic supply strategies when considering: no shaping (i.e., \( E_s[n] = E[n] \) in (1)), memoryless AES and, AES + Phase DPD (in the case of ET) or AES with memory compensation—4 LUTs—in the case of EER, to compensate for the dynamic dependency on the supply voltage.

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