A Novel Digital Predistortion Technique for Class-E PA with Delay Mismatch Estimation

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***Abstract*—This paper proposes a polynomial-based adaptive digital predistortion (DPD) method with delay mismatch estimation for the class-E power amplifier (PA), which can compensate the AM-AM and the strong AM-PM distortion produced by the switching operation of class-E PA. On the other hand, due to the strict requirement of timing alignment in polar transmission, a low-complexity correlation-based scheme for delay mismatch estimation and compensation filter is developed. Simulation results confirmed that the proposed DPD can improve the EVM performance from 5.67% to 0.72%, leading also to 18 dB rejection of spectrum re-growth under the condition of perfect timing alignment. The results also reveal that our proposed method of digital predistortion with delay mismatch estimation, as well as delay mismatch compensation has 1% EVM improvement as compared with applying DPD method directly without timing alignment.**

I. INTRODUCTION

Class-E power amplifier (PA) with polar transmission offers high power efficiency, but suffers from AM-AM and strong AM-PM distortions. The distortion causes spectral re-growth and devastates the constellation diagram. As shown in Fig. 1, especially for the AM-PM distortion, it varies abruptly at low input amplitude. The main cause of this effect is the gate-drain feedthrough of the MOS switch, as depicted in Fig. 2.

In order to obtain high power efficiency while ensuring good linearity to fit high order QAM signal numerous linearization techniques have been proposed for class-E PA. One of the most low-cost techniques is digital pre-distortion (DPD). DPD with look-up table (LUT)-based method was widely used as described in [1] and [2], which simply use the error feedback signal to construct the inverse function and generate LUT for each entity. However, the accuracy of the inverse function depends on the number of entities of the LUT, and a large amount of entities will result in large memory size. On the other hand, polar transmission requires precise timing alignment between amplitude and phase paths [3]. A small timing mismatch (e.g., 2 ns) can degrade the error-vector magnitude (EVM) by 1% and produces obvious spectrum re-growth. This effect will affect the training of DPD and leads to spectrum re-growth even after DPD. Timing mismatch estimation has been proposed in [4], where the complicated least mean squares method makes it quite problematic in terms of implementation.

Here, we propose a DPD scheme using a polar-based polynomial method with recursive least squares algorithm. The advantage of the proposed method is to compute the required coefficients to generate inverse functions of AM-AM and AM-PM distortion of class-E PA. We also propose a piecewise AM-PM predistortion scheme against the strong AM-PM effect. This method saves memory when compared with the LUT-based method.

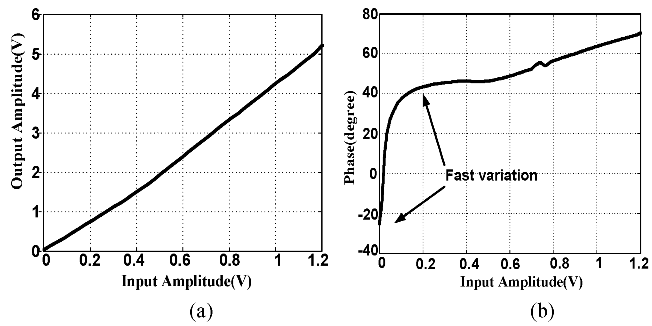

Fig. 1 (a) AM-AM distortion, and (b) AM-PM distortion of Class-E PA.

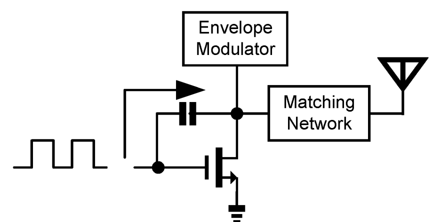

Fig. 2 Simplified Class-E PA.

Section II presents the proposed DPD algorithm. In Section III we propose a simple method for delay mismatch estimation for polar transmission scheme. Simulation results will be shown in Section IV and the conclusions will be drawn in Section V.

II. ALGORITHM OF DIGITAL PREDISTORTION

A. *Mathematical Models of Power Amplifier and Polar-Based Polynomial Digital Pre-distortion*

Figure 3 shows the block diagram of the pre-distortion structure applied in a transmitter with a class-E PA. M-QAM OFDM baseband signals are transmitted, the i^{th} transmitted sample is expressed as $s_{in}^{(i)} = a_{in}^{(i)} exp(j\theta_{in}^{(i)})$, where $a_{in}^{(i)}$ and $\theta_{in}^{(i)}$ represent the amplitude and phase of the i^{th} transmitted sample respectively. With AM-AM and AM-PM distortion, the i^{th} pre-distorted amplitude $a_{pd}^{(i)}$ and the i^{th} pre-distorted phase $\theta_{in}^{(i)}$ can be expressed as:

$$a_{pd}^{(i)} = \mathbf{P}(a_{in}^{(i)})$$
$$= P_1 a_{in}^{(i)} + P_2 \left(a_{in}^{(i)}\right)^2 + \ldots + P_5 \left(a_{in}^{(i)}\right)^5 \quad (1)$$

and

$$\theta_{pd}^{(i)} = \mathbf{Q}(a_{in}^{(i)})$$
$$= \mathrm{Q}_0 + \mathrm{Q}_1 a_{in}^{(i)} + \mathrm{Q}_2 \left(a_{in}^{(i)}\right)^2 + \ldots \mathrm{Q}_5 \left(a_{in}^{(i)}\right)^5 \quad (2)$$

where pre-distortion functions $\mathbf{P}(.)$ and $\mathbf{Q}(.)$ are the AM-AM and AM-PM inverse function of the class E power amplifier. The i^{th} sample of the received signal from the feedback path will be

$$s_o^{(i)} = a_o^{(i)} \exp(j\theta_o^{(i)})$$
$$= \mathbf{G}(a_{pd}^{(i)}) \exp(j(\theta_{pd}^{(i)} + \theta_{in}^{(i)} + \boldsymbol{\varphi}(a_{pd}^{(i)}))) \quad (3)$$

where $\mathbf{G}(.)$ and $\boldsymbol{\varphi}(.)$ denote the AM-AM and AM-PM distortion functions of the Class E power amplifier, which depend on the pre-distorted signal's amplitude $a_{pd}^{(i)}$.

B. Digital Pre-distortion with Recursive Least Squares

The fundamental idea of the AM-AM pre-distortion algorithm is to obtain the coefficients as defined in (1) to minimize the defined function

$$\mathrm{e}_A = \sum_{i=1}^{M} \left| a_{pd}^{(i)} - \mathbf{P}(a_o^{(i)}) \right|^2$$
$$= \sum_{i=1}^{M} \left| a_{pd}^{(i)} - \mathbf{P}(\mathbf{G}(a_{pd}^{(i)})) \right|^2 \quad (4)$$

if function **P(.)** equals the inverse of the function **G**(.), then $\mathbf{PG}(a_{pd}^{(i)}) = a_{pd}^{(i)}$, which means that the cost function is equal to 0 when the exact inverse function of AM-AM distortion is obtained. To minimize the cost function, we can apply the least squares method directly. However, a large amount of computation to invert the matrix is required; so the recursive least squares (RLS) algorithm is employed to find the AM-AM predistortion vector $[P_1 \quad P_2 \ \ldots . \ P_5]^T$, which can find the least squares solution recursively with small amount of memory requirement and fast convergence [5]. The complexity of this algorithm for multiplications and additions are $O(k^3)$ and $O(k^2)$ (O(.) represents the Big O function), where k is the order of polynomial function. It means that the complexity is dominated by the matrix computation as shown in (6). The algorithm of RLS is organized as below:

The initial conditions are:

$$\mathbf{K}^{(0)} = \beta * \mathbf{I}, \quad (5)$$

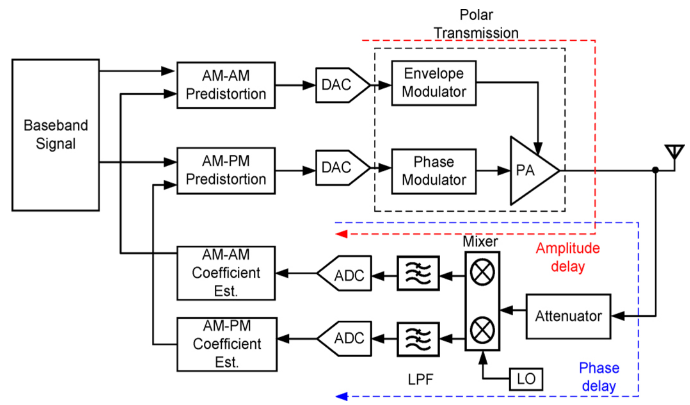

Fig .3 Block diagram of a transmitter with class-E PA, showing amplitude and phase delays.

$$\mathbf{P}^{(0)} = [1 \quad 0 \quad \ldots . \quad 0]^{\mathrm{T}}$$

Where **I** is an identity matrix and β is an arbitrary large number.

The tracking stage is operated as:

$$\mathbf{K}^{(n)} = \mathbf{K}^{(n-1)} - \frac{\mathbf{K}^{(n-1)} \mathbf{A}_{n,o} \mathbf{A}_{n,o}^{\mathrm{T}} \mathbf{K}^{(n-1)}}{1 + \mathbf{A}_{n,o}^{\mathrm{T}} \mathbf{K}^{(n-1)} \mathbf{A}_{n,o}}$$
$$\mathbf{P}^{(n)} = \mathbf{P}^{(n-1)} + \mathbf{K}^{(n)} \mathbf{A}_{n,o} (a_{pd}^{(n)} - \mathbf{A}_{n,o}^{\mathrm{T}} \mathbf{P}^{(n-1)}) \quad (6)$$
$$\mathbf{A}_{n,\mathbf{o}}^{\mathrm{T}} = \left[a_o^{(n)} \quad \left(a_o^{(n)}\right)^2 \quad \ldots . \quad \ldots . \quad \left(a_o^{(n)}\right)^5 \right]$$

For AM-PM pre-distortion the RLS algorithm is applied to obtain the coefficients vector $[Q_0 \quad Q_1 \ \ldots . \ Q_5]^T$ which minimizes the cost function e_θ of AM-PM pre-distortion as follows:

$$\mathrm{e}_\theta = \sum_{i=1}^{M} \left| \theta_o^{(i)} - \theta_{in}^{(i)} - \mathbf{Q}(a_{in}^{(i)}) \right|^2 \quad (7)$$

Since $\theta_o^{(i)} = \boldsymbol{\varphi}(a_{pd}^{(i)}) + \theta_{in}^{(i)}$, if the function **Q**(.) equals to $\boldsymbol{\varphi}(a_{pd}^{(i)})$, then the cost function e_θ is 0. The AM-PM with RLS process can be expressed as:

The initial conditions are:

$$\mathbf{D}^{(0)} = \alpha * \mathbf{I}$$
$$\mathbf{Q}^{(0)} = [0 \quad 0 \quad \ldots . \quad 0]^{\mathrm{T}} \quad (8)$$

Tracking stage:

$$\mathbf{D}^{(n)} = \mathbf{D}^{(n-1)} - \frac{\mathbf{D}^{(n-1)} \mathbf{A}_{n,in} \mathbf{A}_{n,in}^{\mathrm{T}} \mathbf{D}^{(n-1)}}{1 + \mathbf{A}_{n,in}^{\mathrm{T}} \mathbf{D}^{(n-1)} \mathbf{A}_{n,in}} \quad (9)$$
$$\theta_{diff}^{(n)} = \theta_o^{(n)} - \theta_{pd}^{(n)}$$
$$\mathbf{Q}^{(n)} = \mathbf{Q}^{(n-1)} - \mathbf{D}^{(n)} \mathbf{A}_{n,in} (\theta_{diff}^{(n)} - \mathbf{A}_{n,in}^{\mathrm{T}} \mathbf{Q}^{(n-1)}) \quad (10)$$

where $\mathbf{A}_{n,in} = [1, a_{in}^{(n)}, (a_{in}^{(n)})^2, \ldots (a_{in}^{(n)})^5]^{\mathrm{T}}$

C. Piece-wise AM-PM Digital Pre-distortion

As shown in Fig. 1(b), AM-PM distortion of the Class E amplifier varies abruptly with low input amplitude. The Least Squares method cannot accurately find the inverse function of AM-PM distortion with limited values of polynomial orders and it may cause the divergence of coefficients tracking. To mitigate this issue, we equally divide the output voltage into three regions and find the inverse AM-PM curve separately with RLS algorithm. Consequently, a better inverse AM-PM curve can be obtained.

III. DELAY MISMATCH ESTIMATION FOR POLAR TRANSMISSION IN CLASS-E AMPLIFIER

The existence of a time difference between the amplitude and phase signals for polar transmission can be exemplified by the diagram of Fig. 3. This mismatch induces spectrum re-growth and degrades the EVM performance, as well as degrades the performance of the DPD. In order to mitigate this effect, amplitude and phase delay mismatches should be estimated and compensated. In addition, during the training process of digital pre-distortion, the timing difference between received signals from the feedback path and the feedforward signal should be aligned, otherwise, the digital pre-distortion will fail (diverge) and spectrum re-grown will become even worse than without applying digital pre-distortion. This path delay can be determined by finding the maximum delay of amplitude and phase delay. Then, the task would be to estimate three delay parameters associated with amplitude, phase and path, and to compensate them. Since the proposed estimation method for amplitude and phase delay is identical, it will only be mentioned the concept of the integer and fractional delay estimation, and afterwards the compensation of these delays with delay filters.

A. Correlation-Based Integer Delay Estimation

To train coarse / integer timing delay within 1 sample period error training signals with high auto-correlation and small cross-correlation are transmitted and will be received from the RF feedback path. The cross-correlation between the transmitted signals and feedback signals is expressed as

$$C_\tau = \sum_{n=1}^{N} \left(\mathrm{A}_{in}(n) \right) \left(\mathrm{A}_{out}(n-\tau) \right) \quad (11)$$

where τ represents the estimated integer delay (in sample period), A_{out}(n-τ) is the (n-τ)th sample period of feedback path, A_{in}(n) is the nth transmitted signal period and C_τ is the cross-correlation with delay τ. Due to the nature of auto-correlation there will be a maximum value, and the integer delay can be obtained by looking for the value of τ to maximize the function in (11).

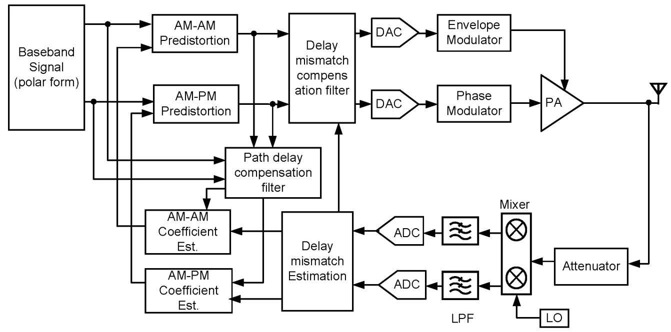

Fig. 4. Block diagram of digital pre-distortion with delay mismatch estimation and compensation.

B. Fractional Delay Estimation

Fractional delay is located in the range of [-0.5 0.5] of the sampling period. This fractional delay signal can be reconstructed by using a Farrow structure interpolation function as:

$$\begin{cases} a_2(\tau_f) = (\tau_f^2 - 1)(\tau_f + 2)\tau_f / 24 \\ a_1(\tau_f) = (\tau_f^2 - 4)(\tau_f + 1)\tau_f / 6 \\ a_0(\tau_f) = (\tau_f^2 - 1)(\tau_f - 4)\tau_f / 4 \\ a_{-1}(\tau_f) = (\tau_f^2 - 4)(\tau_f - 1)\tau_f / 6 \\ a_{-2}(\tau_f) = (\tau_f^2 - 1)(\tau_f - 2)\tau_f / 24 \end{cases} \quad (12)$$

and the reconstructed signal with fractional delay τ_f is expressed as

$$y(n-\tau_f) = \sum_{i=-2}^{2} a_i(\tau_f) A_{out}(n+i) \quad (13)$$

where τ_f is the fractional delay within interval [-0.5 0.5], A_{out}(n-i) is the (n-i)th sample period of received signal from the feedback path. After fractional delay signals are constructed with interval of 0.1 sample period, the correlation method needs to be applied to find the τ_f to maximize the cross-correlation function as

$$C_{\tau f} = \sum_{n=1}^{N} \left(\mathrm{A}_{in}(n) \right) \left(y(n-\tau_f) \right) \quad (14)$$

After finding the amplitude and phase delay, we need to compensate amplitude, phase and path delays before starting the training of the digital pre-distortion. The path delay is equal to the maximum delay between amplitude and phase delays. As shown in Fig. 4, the first step would be to estimate amplitude, phase and path delays from the feedback path, then the delay mismatch compensation filter will compensate the time difference between amplitude and phase delay to align amplitude and phase signals for the polar transmission. On the other hand, the path delay compensation filter will compensate the time delay between feed-forward path and feedback path, which is used for the training process of DPD. The hardware complexity is mainly based on the number of samples for correlation computations in (13) and (14). The

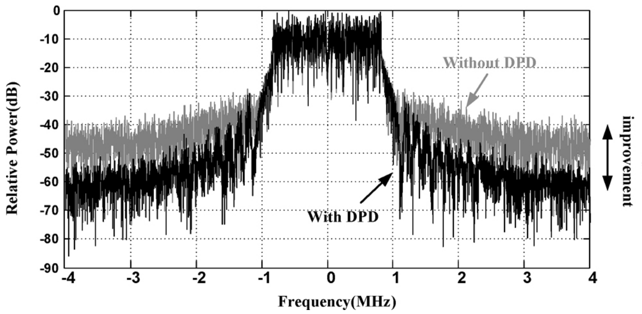

Fig. 5 PA model output power spectrum for OFDM 64-QAM signal.

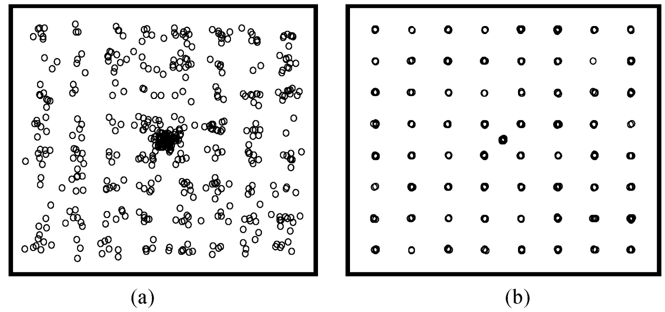

Fig 6. 64-QAM constellation diagram (a) before digital pre-distortion (b)after digital pre-distoriton.

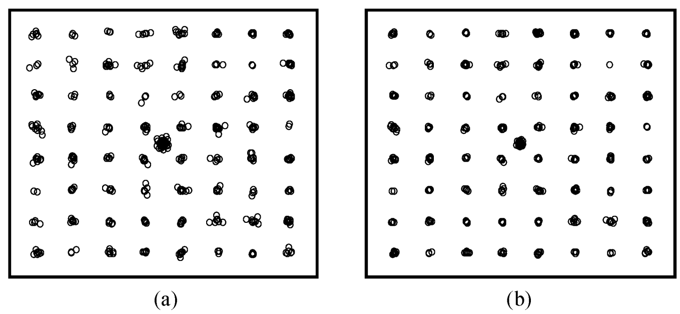

Fig 7. 64-QAM constellation diagram with and without delay mismatch compensation for 1.5 ns while applying digital predistortion for class E PA.

Big O functions for multiplications and additions are $O(N^2)$ and O(N) respectively, where N is the number of samples for calculating correlation functions.

IV. Simulation Results

The MATLAB/Simulink is employed as the simulator to verify the proposed method and compare the performances with the requirement of the IEEE 802.11a standard. Fig. 5 shows the PA output spectrum of OFDM 64-QAM signals with and without digital pre-distortion under the condition of perfect timing alignment. Spectrum re-growth has an overall 18 dB improvement after applying digital pre-distortion. Comparing with the IEEE 802.11a standard, the spectrum mask requires -20/-28/-40 dBc at 11/20/28 MHz. It can be observed that at 30 MHz, the spectrum re-growth is close to -40 dBc without digital predistortion. As a result, the Class E PA will be suitable for the 802.11a system after DPD. Fig. 6 shows the simulated constellation diagram of decoded OFDM 64-QAM signals. The constellation diagram without any DPD is shown in Fig. 6(a), the EVM output in this case is 5.67%. The constellation diagram with DPD is shown in Fig. 6(b). The EVM performance in this case is improved to 0.72%. This EVM can fulfill the requirement of 802.11a standard, which specifics 3% EVM.

Fig. 7 shows the constellation diagram of decoded OFDM 64-QAM with and without delay mismatch compensation for 1.5 ns mismatch while applying DPD. As shown in Fig. 7(a), the presence of distortions produced from class-E PA is compensated by digital predistortion, however, delay mismatch degrades the EVM performance which is 2.52%. In Fig. 7(b), after estimated and compensated delay mismatch, as well as applying DPD, the EVM performance is improved from 2.52% to 1.49%.

V. Conclusions

A novel DPD with delay mismatch estimation for Class-E PA has been proposed. It allows class-E PA to reach high power efficiency while ensuring linearity suitable for high-tier wireless applications such as IEEE 802.11a Wireless LAN. The proposed piecewise AM-PM predistortion method can effectively compensate the abrupt variation of AM-PM distortion generated by the Class E PA. The polynomial-based RLS method saves memory when compared to the look-up table method. In addition, a simple method for delay mismatch estimation has been proposed. It mitigates the effect of timing mismatch in the polar transmission scheme, and provides timing alignment for the training process of the DPD. EVM and constellation diagram results verified the feasibility of the entire DPD scheme.

Acknowledgments

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