A Survey of Delay and Gain Correction Methods for the Indirect Learning of Digital Predistorters

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Abstract—The indirect learning architecture is one of the most common setups for the identification of digital predistorters which are used to linearize radio frequency power amplifiers. To apply the indirect learning architecture in a real transmitter, the unknown delay and gain between the transmitted and the received signal must be corrected. In the present paper we present a survey of delay and gain correction methods and discuss basic objectives for selecting the delay and gain of the linearized system. We demonstrate the equivalence of gain correction before and after identification and conclude with a method for performance comparison at constant output power.

I. INTRODUCTION

The digital enhancement of analog, mixed-signal and radio frequency circuits is a highly active field of research, which is motivated by the good scaling properties of digital circuits, the increased flexibility and the potential for improved power efficiency [1]. Especially for wireless transmitters, the power efficiency is an important design objective, since it directly affects the operation time of battery powered mobile devices, and it is a main factor for the operational costs and the environmental impact of cellular base-stations. Within wireless transmitters, the power amplifier is one of the most critical parts, limiting both the linearity and the efficiency. Since for power amplifiers, there is a trade-off between linearity and efficiency [2], nonlinear transmitter architectures [3]–[5] and the linearization by digital predistortion [6]–[8] have become key technologies for power efficient wireless transmitters.

In digital predistortion, the distortion produced by the power amplifier is pre-compensated in the digital baseband processing of the transmitter. In contrast to post-compensation in a receiver, the desired compensator output signal is not known in the case of a predistorter. Therefore, specific identification architectures are necessary which can be categorized as:

(A1) System identification and inversion. [9]
(A2) Direct learning by an adaptive nonlinear filtered-x algorithm [10] or by stochastic otimization [6].
(A3) Indirect learning by identifying a postdistorter that is used in front of the system as a predistorter. [11]

Allthough the architectures (A1) and (A2) have the potential for improved identification performance [12], the indirect learning architecture is the most popular one because of its simplicity.
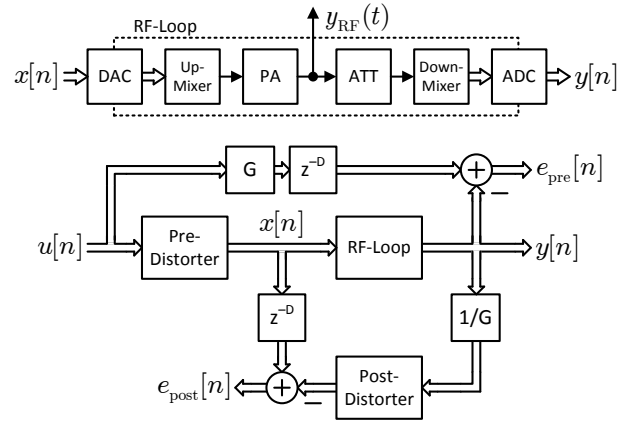
Fig. 1. Digital predistortion using the indirect learning architecture.

A block diagram of digital predistortion using indirect learning is shown in Fig. 1, where single arrows indicate real signals and double arrows indicate complex signals. The top part shows the radio frequency (RF) loop consisting of the transmitter (DAC, up-mixer, power amplifier) followed by the feedback receiver (attenuator, down-mixer, ADC). The bottom part shows the indirect learning architecture, consisting of the RF loop and digital nonlinear baseband systems used as pre- and postdistorters. The aim of digital predistortion is to linearize the path from the source signal $u[n]$ to the output signal $y[n]$, such that the cascade of the predistorter and the RF loop reduces to a desired complex gain G and a delay z^{-D}, minimizing the predistortion error signal $e_{\mathrm{pre}}[n]$. To identify such a predistorter, the indirect learning architecture first identifies a postdistorter using a delayed version of the input signal $x[n]$ and a normalized version of the output signal $y[n]$, which minimizes the postdistortion error signal $e_{\mathrm{post}}[n]$. After identifying the postdistorter, it is used as a predistorter with the expected result that this predistorter also minimizes the predistortion error signal $e_{\mathrm{pre}}[n]$. In practice, the expected result can be achieved if two requirements are fulfilled:

(R1) The postdistorter is able to sufficiently minimize its error signal, so that it is close to a true inverse which is, by definition, both a pre- and a post-inverse.
(R2) The predistorter is operated in a region so that the predistorted signal $x[n]$ does not exceed the magnitude range it had during postdistorter identification.

Since the requirements (R1) and (R2) are crucial for the success of indirect learning and their fulfillment strongly relies on the choice of the input signal delay z^{-D} and the normalization gain $1/G$ during postdistorter identification, there are many papers on methods for delay correction [13]–[18] and gain correction [19]–[23] in the digital predistortion literature. To our best knowledge, however, there is no survey which presents a concise overview of this important topic.

In the present paper we present such a survey, where we review methods for delay correction, combined with an overview on gain correction. We discuss practical aspects for lab-based measurements and on-chip applications and present methods for delay and gain correction which ensure that the requirements (R1) and (R2) can be fulfilled.

II. Delay Correction

The fulfillment of requirement (R1) depends on the postdistorter model and the input signal delay. The postdistorter model must be able to sufficiently minimize the postdistortion error and the input signal delay must be chosen such that the modeling capabilities of the postdistorter can be exploited.

In the case of narrowband signals, it is typically sufficient to use a quasi-memoryless model which can be implemented by a complex baseband polynomial [7] or a lookup table [14]. Since a quasi-memoryless model is not able to compensate residual loop delay error, accurate alignment of the transmitted and the received signal is very important, requiring either high oversampling ratio or fractional delay correction [16].

With wideband signals, frequency dependent effects arise which cannot be represented by a quasi-memoryless model. In this case, nonlinear models with memory are required such as the memory polynomial [9], the generalized memory polynomial [24] or the baseband Volterra series [25]. Depending on their memory depth, these models are able to compensate a certain amount of loop delay error such that the requirements on the delay correction can be relaxed to integer resolution. An important constraint, however, is that the postdistorter is not able to compensate non-causal misalignment of the training signals or causal misalignment beyond its memory depth [17]. For the indirect learning architecture in Fig. 1 this means that overestimation of the input signal delay up to one sample can be tolerated, but underestimation must be avoided.

Robust delay correction requires an objective function that is insensitive to noise, distortion and complex scaling. In the literature many different objective functions are used like the cross-covariance of magnitudes [16], the cross-correlation of differential magnitudes [13], and the distance between local extrema of the phase characteristic [18]. Although the use of real quantities like magnitude and phase lowers the implementation complexity, the complex cross-correlation [15] may be considered the most general objective function, since it includes both magnitude and phase. In the remaining part of this section, we therefore review the complex cross-correlation, a method for its efficient computation and techniques for integer and fractional delay estimation and correction.

A. Definition of the Complex Cross-Correlation

The complex cross-correlation of two finite-length, discrete-time signals $x[n]$ and $y[n]$, with $0 \leq n \leq N - 1$, is given by

$$r_{xy}[m] = \sum_{n=0}^{N-1} x^*[n - m] y[n] \tag{1}$$

where we can set $x[n]$ and $y[n]$ to zero outside their definition range resulting in the linear cross-correlation, or periodically extend them to get the circular cross-correlation. The linear cross-correlation has a support of $-(N-1) \leq m \leq +(N-1)$. The circular cross-correlation is periodic in m and uniquely defined by the N samples within $0 \leq m \leq +(N - 1)$. The choice whether to use the linear or the circular definition depends on the application. At lab-based measurements, a periodic input signal can be used, which allows the acquisition of the output signal at an arbitrary point in time. In this case, the circular definition should be used. For on-chip applications, typically no periodicity can be assumed and the acquisition of the output signal is synchronized to the generation of the input signal. In this case the linear definition should be used.

B. Computation of the Complex Cross-Correlation

In the case of synchronized acquisition, the cross-correlation has to be computed only for a small number of time-lags, which may be done by direct evaluation of (1). However, if all time-lags need to be calculated, it is more efficient to compute it in the frequency domain. Defining the signal vectors $\boldsymbol{x}$, $\boldsymbol{y}$ and their circular cross-correlation vector $\boldsymbol{r_{xy}}$ by

$$\boldsymbol{x} = [\quad x[0] , \quad x[1] , \quad \ldots , \quad x[N - 1]]^{\mathrm{T}} \tag{2}$$

$$\boldsymbol{y} = [\quad y[0] , \quad y[1] , \quad \ldots , \quad y[N - 1]]^{\mathrm{T}} \tag{3}$$

$$\boldsymbol{r_{xy}} = [r_{xy}[0] , \ r_{xy}[1] , \ \ldots , \ r_{xy}[N - 1]]^{\mathrm{T}} \tag{4}$$

we implement the computation of $\boldsymbol{r_{xy}}$ by [26]

$$\boldsymbol{r_{xy}} = \mathrm{IFFT} \left\{ (\mathrm{FFT}\{\boldsymbol{x}\})^* \circ \mathrm{FFT}\{\boldsymbol{y}\} \right\} \tag{5}$$

where the symbol $\circ$ is used for element-wise multiplication. The equivalence of (2)-(5) with the circular version of (1) follows from two properties of the discrete Fourier transform:

- Multiplication in frequency domain corresponds to circular convolution in time domain.
- Conjugation in frequency domain corresponds to conjugation and time-reversal in time domain.

If required, (5) may also be used for linear cross-correlation. This is done by appending $N-1$ zeros to (2) and (3), resulting in appended values of $r_{xy}[-(N - 1)]$ up to $r_{xy}[-1]$ in (4).

For delay estimation, the cross-correlation as defined in (1) may be used directly, but for comparison it is often useful to normalize the result to get the correlation coefficient

$$\rho_{xy}[m] = \frac{r_{xy}[m]}{\sqrt{(\boldsymbol{x}^{\mathrm{H}}\boldsymbol{x})(\boldsymbol{y}^{\mathrm{H}}\boldsymbol{y})}} \tag{6}$$

where $(\cdot)^{\mathrm{H}}$ represents a Hermitian transpose. The magnitude of $\rho_{xy}[m]$ is in the range between zero and one with $\rho_{xx}[0] = 1$.
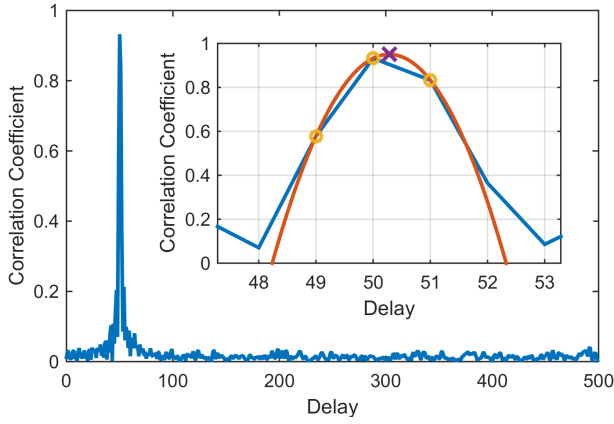
Fig. 2. Integer and fractional delay estimation using circular cross-correlation.

Fig. 3. Gain normalization using peak gain and root-mean-square gain.

C. Delay Estimation

After computing the cross-correlation $r_{xy}[m]$, the integer delay D_{int} is estimated by searching for the lag m which maximizes the magnitude of correlation, expressed by

$$D_{\text{int}} = \arg \max_{m} |r_{xy}[m]|. \tag{7}$$

For fractional delay estimation, a parabola is fitted through the point of maximum correlation and its neighbors [26]. Defining $c_i = |r_{xy}[D_{\text{int}} + i]|$, the maximum of the parabola is located from c_{-1}, c_0, c_{+1}, giving the fractional delay estimate

$$D_{\text{frac}} = \frac{c_{-1} - c_{+1}}{2 (c_{-1} - 2c_0 + c_{+1})} \tag{8}$$

and the overall delay estimate $D = D_{\text{int}} + D_{\text{frac}}$. An example for delay estimation using this technique is shown in Fig. 2.

D. Delay Correction

The integer delay is compensated very easily by shifting either the input or the output signal. For the fractional delay correction, interpolation is needed. This is implemented very efficiently by a Farrow filter [27]. In the case of lab-based measurements with a periodic input signal, the following technique may be used. Transforming the input signal by

$$\boldsymbol{X} = \text{FFTSHIFT}\{\text{FFT}\{\boldsymbol{x}\}\} \tag{9}$$

one gets the spectrum $\boldsymbol{X}$ corresponding to the frequencies

$$\boldsymbol{\omega} = \frac{2\pi}{N} \left[- \left\lfloor \frac{N}{2} \right\rfloor, - \left\lfloor \frac{N}{2} \right\rfloor + 1, \ldots, \left\lceil \frac{N}{2} \right\rceil - 1 \right]^{\text{T}} \tag{10}$$

in radian per sample. Applying a linear phase shift

$$\boldsymbol{X}_{\text{delay}} = \boldsymbol{X} \circ e^{-j\boldsymbol{\omega}D} \tag{11}$$

and transforming back to time domain by

$$\boldsymbol{x}_{\text{delay}} = \text{IFFT}\{\text{IFFTSHIFT}\{\boldsymbol{X}_{\text{delay}}\}\} \tag{12}$$

one gets the delayed time-domain vector $\boldsymbol{x}_{\text{delay}}$. The delay D applied in (11) can be any real number of samples and corresponds to circular sinc interpolation in the time-domain.
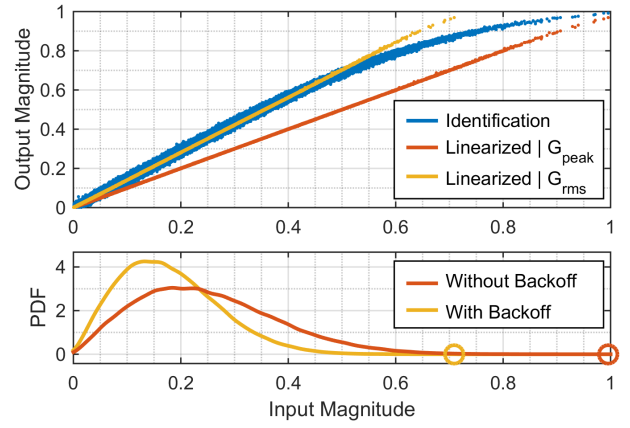
III. GAIN CORRECTION

The fulfillment of requirement (R2) depends on the operation point of the power amplifier, the statistics of the source signal during postdistorter identification and predistorter operation and the normalization gain during postdistorter identification. For our present discussion we assume that the gain between predistorter output and power amplifier input is fixed. At lab-based measurements using a vector signal generator, a fixed gain requires that the automatic-level-control is deactivated. For identification, we use a signal with peaks close to digital full scale, leading to a compressive characteristic like in Fig. 3. To linearize this compression, the predistorter has to apply an expansion on the source signal which increases its crest factor. A method to ensure that this expansion does not increase the signal peaks is to normalize by the peak gain [22]

$$G_{\text{peak}} = \frac{\max |\boldsymbol{y}|}{\max |\boldsymbol{x}|} \tag{13}$$

during postdistorter identification. This ensures that clipping is avoided and requirement (R2) is fulfilled, but it reduces the gain after linearization and the output power. An alternative is to normalize by the root-mean-square gain [23]

$$G_{\text{rms}} = \frac{\text{rms}\{\boldsymbol{y}\}}{\text{rms}\{\boldsymbol{x}\}} \tag{14}$$

during postdistorter identification. This ensures that the output power with and without predistorter is nearly identical, but to avoid clipping and to fulfill requirement (R2), a digital backoff must be applied to the source signal, which is given by

$$G_{\text{backoff}} = \frac{G_{\text{peak}}}{G_{\text{rms}}} = \frac{\text{crestfactor}\{\boldsymbol{y}\}}{\text{crestfactor}\{\boldsymbol{x}\}}. \tag{15}$$

This backoff is equal to the gain reduction if normalization by the peak gain is used. Therefore, both methods are equivalent and reduce the output power, compared to the training with no digital backoff and no predistorter. For a fair performance comparison with and without predistorter, the digital backoff has to be applied also in the evaluation without the predistorter. To increase the output power, the digital backoff may also be replaced by crest factor reduction of the source signal. [28]
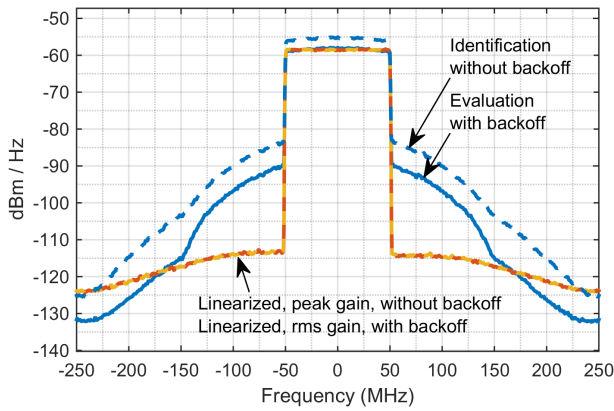
Fig. 4. Comparison of power spectral densities of the output signal.

As an example, Fig. 4 shows the power spectral densities of the output signals with and without predistorter. The results for the normalization by G_{peak} without digital backoff and the normalization by G_{rms} with digital backoff are identical.

IV. CONCLUSION

In the present paper, we presented a survey of delay and gain correction methods for the indirect learning of digital predistorters. We identified two requirements for the success of indirect learning and related them to objectives for the delay and gain correction. For delay correction, we presented a review of the complex cross-correlation, an efficient method for its computation and simple methods for integer and fractional delay estimation and correction. For gain correction, we compared the normalization by the peak gain and the root-mean-square gain and concluded that both methods are equivalent if the required digital backoff is included. To get the same output power without the predistorter, the digital backoff must be applied also during evaluation without predistorter.

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