Comparison and Experimental Verification of Two Low-complexity Digital Predistortion Methods

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***Abstract*— Recently, many papers have been published on the designs of predistortion systems for power amplifiers (PA). Most of these studies are based on computer simulations. In this work, we examine our predistorter (PD) designs on a practical microwave PA in a laboratory setup. The objective is to validate two PDs of different design approaches. The first PD is modeled using a Simplicial Canonical Piecewise Linear (SCPWL) function while the second one is modeled using a polynomial function. The former exploits the properties of the SCPWL function to develop a reduced complexity PD identification method. The latter utilizes the Secant method to iteratively search for the signal at the PD output, thus parameter identification is avoided. The implementation complexity and effectiveness of the PDs are compared. The limitations of the measurement system for predistorter design is also discussed briefly.**

I. Introduction

The rapid development of future mobile communications systems has expedited the development of linearization techniques for broadband power amplifiers in the past decade. The fourth generation mobile communications system is likely to employ Orthogonal Frequency Division Multiplexing (OFDM) modulation to achieve bandwidth efficiency and robustness against multi-path effects. However, OFDM signals are known to have high peak-to-average power ratio (PAPR). High PAPR signals are sensitive to nonlinear distortion, which causes adjacent channel interference (ACI). The 3GPP standard [1] recommended for the adjacent channel power ratio (ACPR) to be at least 45 dBc for the WCDMA system. One way to reduce ACI is by backing off the PA but this reduces the power efficiency of the PA significantly.

Power efficiency is an important property for mobile communications systems. Power efficient PAs certainly lend a hand to reduce the operating costs of the base station network and to prolong the battery operation time of the mobile terminals. Power amplifiers are most efficient when operated near saturation but the drawback is their nonlinear responses. Therefore, to meet both complementary requirements, linearization techniques are required to mitigate the nonlinear effects when efficient PAs are used. Among the linearization techniques found in the literatures, digital predistortion (DPD) is most popular due to its good tradeoff between cost and effectiveness. Examples of existing predistortion approaches are [2] [3] [4] [5]. These PDs are implemented using various different approaches in order to reduce implementation complexity. However, most of the proposed PDs are designed and tested via computer simulation with synthetic data. This does not account for many issues and limitations of the measurement system and practical implementation of a real system. The two PD methods considered in this paper are DPDs that are identified from the measurement data of a microwave PA. Then the identified PDs are tested on the PA in a testbed. The objectives of this paper are:

1) to validate the designs of our PDs, namely an SCPWL model PD [6] and an iterative method [7], the Secant-PD, in an experimental setup, and
2) to compare the implementation complexity and the effectiveness of the two PDs.

The organization of the paper is as follows: In Section II, the PD designs to be validated are briefly reviewed. The laboratory setup for PA measurement and PD testing is presented in Section III. The measurement results are presented in Section IV. Finally, conclusions are drawn in Section V.

II. Pre-distortion Methods

Fig. 1 shows the final stage of a transmitter with digital pre-distortion. The amplified signal is transmitted through the antenna and a portion is fed back to the DPD identification algorithm. Given the input signal $u[n]$ and the output signal $z[n]$ from the feedback path, the identification algorithm shall search the DPD filter that linearizes the nonlinear effect of the high-power amplifier (HPA).

In this paper, two low-complexity predistortion methods that linearize only the AM-AM conversion are compared with respect to implementation complexity and effectiveness. The PDs are evaluated on a practical microwave power amplifier from Mini Circuit, ZVE8G. Both methods find the solution for the pre-inverse filter $\mathbb{P}$, whereby $\mathbb{N}$ denotes the static nonlinear AM-AM model of the HPA, see Fig. 2. The following subsections discuss the two DPD methods.

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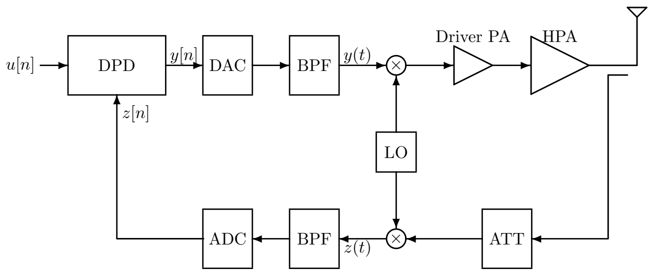

Fig. 1. Radio front-end with digital predistortion (DPD)

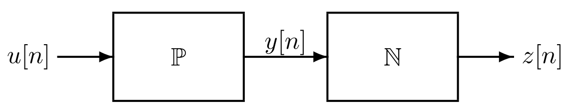

Fig. 2. Predistortion problem

A. The SCPWL Method

The Simplicial [8] Canonical Piecewise Linear (SCPWL) PD proposed in [6] is considered in our experiment. This method models both the PA and PD using the SCPWL function [9] given by

$$\mathbb{N}\Big(\big|y[n]\big|\Big) = c_0 + \sum_{i=1}^{\sigma-1} c_i \Lambda_i\Big(\big|y[n]\big|\Big), \tag{1}$$

where $\Lambda_i(\cdot)$ is the basis function and σ is the number of user-defined partition points, defined by $y - \beta_j = 0$. The basis function is given by

$$\Lambda_i\Big(\big|y[n]\big|\Big) = \begin{cases} \frac{1}{2}\big(y[n] - \beta_i + |y[n] - \beta_i|\big), & y[n] \le \beta_\sigma \\ \frac{1}{2}\big(\beta_\sigma - \beta_i + |\beta_\sigma - \beta_i|\big), & y[n] > \beta_\sigma \end{cases} \tag{2}$$

where the partition points must satisfy the condition that $\beta_1 \le \beta_2 \le \cdots \le \beta_\sigma$.

The SCPWL function divides the input space of a nonlinear function into $\sigma - 1$ segments, each defined by a linear affine function. The basis function imposes a saturation behavior after the last partition point β_σ making the function suitable for describing the PA and PD characteristics. Two properties of the SCPWL function can be exploited to simplify the PD model identification as discussed in [6]:

1) Each SCPWL segment is a linear affine function. Thus, the Inverse-Coordinate Mapping[1] (ICM) method as shown in Fig. 3 can be used to identify the PD's nonlinear characteristic. The mapping matrix is given by

$$\mathbf{Q} = \begin{bmatrix} 0 & \frac{1}{g} \\ 1 & 0 \end{bmatrix}, \tag{3}$$

where g is the desired linear gain. Since each segment of a SCPWL function is a linear affine function, projections of the PA responses at the partition points, $b_j = \big[|\beta_j| \quad |\mathbb{N}(\beta_j)|\big]^{\mathrm{T}}$ where $j = 1, 2, \cdots, \sigma$, are sufficient as shown in Fig. 3.

[1]The mapping matrix given in (3) is simplified and improved. It is different from that proposed in [6]. The development of (3) will be reported elsewhere.

2) Let the basis function matrix of the SCPWL function be defined as

$$\mathbf{\Lambda} = \begin{bmatrix} 1 & 1 & \cdots & 1 \\ \Lambda_1(x_1) & \Lambda_1(x_2) & \cdots & \Lambda_1(x_k) \\ \vdots & \vdots & \ddots & \vdots \\ \Lambda_{\sigma-1}(x_1) & \Lambda_{\sigma-1}(x_2) & \cdots & \Lambda_{\sigma-1}(x_k) \end{bmatrix}. \tag{4}$$

When $\mathbf{\Lambda}$ is evaluated at the partition points $\big[\beta_1 \quad \beta_2 \quad \cdots \quad \beta_\sigma\big]$, the elements of the matrix can be calculated from partition sizes $s_j = \beta_{j+1} - \beta_j$ of the input space. Likewise for its inverse matrix $\mathbf{\Lambda_I} = \mathbf{\Lambda}^{-1}$, which is given by [6]

$$\mathbf{\Lambda_I} = \begin{pmatrix} 1 & 0 & \cdots & \cdots & 0 \\ -\frac{1}{s_1} & \frac{1}{s_1} & 0 & \cdots & 0 \\ \frac{1}{s_1} & -(\frac{1}{s_1}+\frac{1}{s_2}) & \frac{1}{s_2} & \ddots & \vdots \\ 0 & \frac{1}{s_2} & & \ddots & \vdots \\ \vdots & 0 & \ddots & & \vdots \\ 0 & \cdots & \frac{1}{s_{\sigma-2}} & -(\frac{1}{s_{\sigma-2}}+\frac{1}{s_{\sigma-1}}) & \frac{1}{s_{\sigma-1}} \end{pmatrix}. \tag{5}$$

The SCPWL-PD identification procedure, which involves the ICM method and the construction of $\mathbf{\Lambda_I}$, is carried out as follows. First, the model of the PA is identified by fitting the input-output measurement data to the SCPWL function in the Least-squares (LS) sense. Then, the coordinates $\mathbf{b}$ representing the input-response of the partition points are determined. The PD characteristic is obtained by using the ICM method as $\mathbf{b}' = \mathbf{Q}\mathbf{b}$, as shown in Fig. 3. The PD responses obtained from the mapping fall in the input range of the PA and results in a linearized gain g.

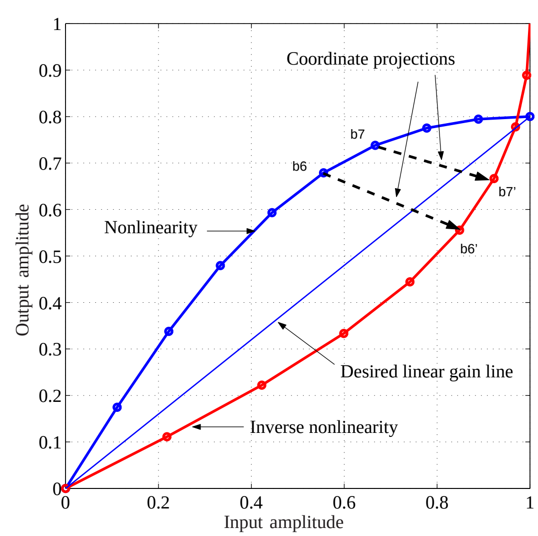

Fig. 3. Identification of PD characteristic using the Inverse-Coordinate Mapping method

The identification of the SCPWL coefficients begins with the construction of $\mathbf{\Lambda_I}(s')$. Then the coefficients can be obtained by LS estimation as

$$\mathbf{c}_{PD} = \mathbf{\Lambda}_{\mathrm{I}}\mathbb{P}(\boldsymbol{\beta}'), \tag{6}$$

where s' is the PD's partition sizes and $\mathbb{P}(\boldsymbol{\beta}')$ is the SCPWL function of the PD evaluated at its partition points.

B. Secant Method

Given the nonlinear AM-AM model $\mathbb{N}$ of the power amplifier, a pre-filter $\mathbb{P}$ that linearizes the signal path is searched, see Fig. 2. Mathematically,

$$\mathbb{N}\Big(\mathbb{P}\big(u[n]\big)\Big) = \mathbb{L}\big(u[n]\big) \; . \tag{7}$$

Here, $\mathbb{L}\big(u[n]\big) = g \cdot u[n]$ with a targeted linear amplification g.

For general dynamic nonlinear models, e.g., Volterra series, or even for static nonlinear models, e.g., a power series, analytic solutions for the pre-inverse are difficult to obtain. Therefore, the solution is approximated with an iterative method. Rewriting (7) gives

$$\mathbb{N}\big(y[n]\big) - g \cdot u[n] = \mathbb{S}_u\big(y[n]\big) = 0 \; , \tag{8}$$

with $\mathbb{P}\big(u[n]\big) = y[n]$. This nonlinear equation must be solved for the signal $y[n]$. Here, an iterative method as follow is used:

$$y_{i+1}[n] = y_i[n] - \mu[n]\mathbb{S}_u\big(y_i[n]\big) \; , i \geq 0 \; , \tag{9}$$

where $y_{-1}[n]$ and $y_0[n]$ are given. The step-size is given by

$$\mu[n] = \frac{y_i[n] - y_{i-1}[n]}{\mathbb{N}\big(y_i[n]\big) - \mathbb{N}\big(y_{i-1}[n]\big)} \; , \tag{10}$$

which corresponds to the Secant-method [7] for root-finding. In contrast to the SCPWL method, this iterative approach does not identify a set of coefficients for a pre-determined PD, but directly identifies the signal after the PD. The algorithm in (9) is generic, i.e., it does not depend on the choice of model $\mathbb{N}$ for the PA. Be it a model that accounts also for AM-PM conversion or dynamic effects such as a Volterra series, the algorithm defined by (9) remains the same.

Here, a very simple static model for the PA which accounts only for the AM-AM conversion is used,

$$z[n] = \mathbb{N}\big(y[n]\big) = \theta_1 y[n] + \theta_3 \big|y[n]\big|^2 y[n], \theta_{1,3} \in \mathbb{R} \; . \tag{11}$$

Only two real-valued coefficients are required for modelling the PA. Three iterations in the algorithm (9) with the starting values $y_{-1}[n] = 0$ and $y_0[n] = u[n]$ are used. With a proper selection of the targeted linear gain g, convergence of (9) could be easily achieved.

C. Comparison of the Complexity

The PD implementation procedure can largely be divided into two sessions:

1) The training session, in which the models of the PA and PD are identified. Note that the Secant method requires only the PA model.
2) The PD operation session, in which the input signal $u[n]$ is predistorted to obtain $y[n]$.

The implementation complexity of the two PD methods is compared in terms of required operations (additions, multiplications, divisions, absolute values, decisions and matrix inversion).

Tables I and II show the operations required during the training session, assuming a data block size of N is used. The complexity of PA identification is higher for the SCPWL method due to the required decision making and absolute value implementation in (2). The high complexity operation involved in the Secant method is the 2×2 matrix inversion for estimating the model coefficients θ_1 and θ_3. This required complexity is comparatively low. Note that the Secant method requires no PD model identification as shown in Tab. II. The number of operations required for the SCPWL method is growing linearly with the number of PWL partitions. Therefore, we conclude that the SCPWL method requires more operations in the training session than the Secant method.

	SCPWL	Secant
# of ADD	$3N + \sigma$	$4N + 2$
# of MUL	$4N + \sigma$	$2N$
# of DIV	0	0
# of ABS	$N + \sigma - 1$	0
# of decisions	$N + \sigma - 1$	0
# of $[\cdot]^{-1}$	0	1

TABLE I

REQUIRED OPERATIONS DURING TRAINING SESSION: PA MODEL IDENTIFICATION

	# of ADD	# of MUL	# of DIV
SCPWL	$\sigma - 2$	$\sigma - 1$	$6\sigma - 8$
Secant	0	0	0

TABLE II

REQUIRED OPERATIONS DURING TRAINING SESSION: PD MODEL IDENTIFICATION

The number of operations required in the PD operation session is shown in Tab. III. Hereby, the Secant method employs three iterations and the SCPWL methods uses $\sigma - 1 = 10$ partitions. The complexity of the SCPWL-PD operation is dependent on the number of partition points σ, while that of the Secant-PD is dependent on the order of PA model used. Although the SCPWL-PD shows lower complexity in the operation session, the overall implementation complexity is higher than that of the Secant method.

	# of ADD	# of MUL	# of DIV
SCPWL	11	11	0
Secant	15	18	3

TABLE III

REQUIRED OPERATIONS DURING PD OPERATION SESSION

III. The Measurement System Setup

The measurement system setup is illustrated in Fig. 4. The system consists of a digital baseband processing part and an RF processing part. The device under test (DUT) is a broadband power amplifier from Minicircuits, type ZVE-8G. In the digital baseband processing part, a random phase multitone excitation signal $u[n]$, covering a bandwidth of

5 MHz is generated using Matlab® as

$$u[n] = \sum_{k=0}^{K-1} A_k[n] \cos\bigl(2\pi\Delta f kn + \phi_k\bigr), \qquad (12)$$

where k and K are the tone number and the total number of tones used to generating the signal, respectively and Δf is the tone spacing. The tone amplitude, $A_k[n] = 1$ for all k, n.The signal is then transferred to the memory of the

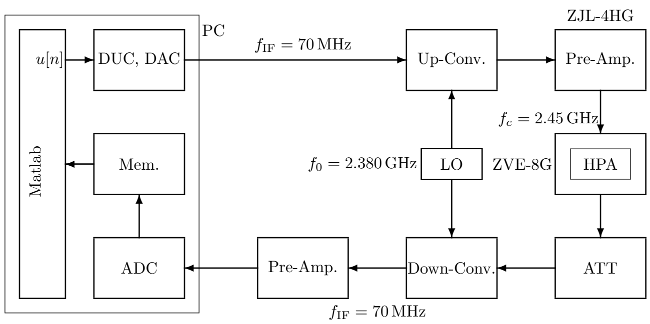

Fig. 4. Measurement setup for evaluation of the presented PD methods

digital-to-analog converter (DAC) module SMT 370 from Sundance [10], carried by a PCI board (SMT 310Q) in the host PC. This excitation signal is digitally modulated ($f_{\mathrm{IF}} = 70\,\mathrm{MHz}$) and converted to analog signal before being fed to the RF part.

The up-converter mixes the signal to a center frequency $f_c = 2.45\,\mathrm{GHz}$ and filters the signal. A pre-amplifier (Mini-circuits ZJL-4HG) and an adjustable attenuator are used to boost the signal before the DUT and to control the input back-off (IBO) of the DUT, respectively. Then the output signal of the DUT is attenuated, down-converted to $f_{\mathrm{IF}} = 70\,\mathrm{MHz}$ and filtered. The filters of both the up- and down-converter are bandlimited to 20 MHz. A common local oscillator is used for the up- and down-conversion to avoid phase imbalance. The analog-to-digital converter (ADC) samples the IF signal at a rate of $f_s = 100\,\mathrm{MHz}$ and a resolution of 14 bit. The output signal is then stored in a memory module (SMT 351G) and is accessible for model identification.

A. Limitation of the measurement system

The measurement system poses a number of limitations to the PD designs. The up- and down-converters may be nonlinear devices. If the input signal level to the converters is too high, the resulting nonlinear distortion will affect the performance of the PD. As a result of limiting the input signal level, a low output signal level is obtained. In order to fully exploit the 14 bits dynamic range of the ADC, pre-amplification before sampling is required. The ADC requires a relatively high amplitude of the analog input signal, i.e., $2.2\,\mathrm{V_{ss}}$. The pre-amplification introduces noise that increases the noise floor of the measured data. This results in a reduced dynamic range of approximately 50 dB, as compared to 60 dB when measurement is done before the down-converter. This is evident in the measurement results presented in the following section. On the other hand, too low a signal level to the ADC results in high quantization noise, which again reduces the dynamic range. Thus, the solution to this problem is to acquire a highly linear down-converter, capable of delivering enough output power, or a low noise pre-amplifier.

The second limitation of the system is caused by the filters of the up- and down-converters. These filters are bandlimited to 20 MHz. The models obtained from the measurement data are able to capture only the intermodulation distortion (IMD) of up to half of the fifth-order band. Therefore, the effectiveness of the PDs identified are also bandlimited to 20 MHz, regardless of sampling rates higher than 20 MHz that the ADC provides. This problem can be circumvented by employing filters with higher bandwidth. In order to include the ℓ-th order IMD in the model, the filter bandwidth should be at least $\ell\times$ 5 MHz, or up to the limit of the ADC's sampling rate.

IV. Measurement Results

Using the input-output (IO) data obtained from the measurement, the PA models for the SCPWL method and the Secant method described in Section II are identified. Subsequently, the SCPWL PA model is used to identify the SCPWL coefficients of the PD. The Secant method uses the PA model directly to calculate the PD output signal as in (9). The predistorted signal of the respective method is then applied to the power amplifier and the output spectrum is measured with a spectrum analyzer.

In our experiment, the two PDs are tested on a weakly nonlinear PA and on a PA driven to saturation. An adjustable attenuator was used to back-off the PA for the weakly nonlinear case.

In Fig. 5 the compensation results for the weakly nonlinear PA is shown. The SCPWL-PD employed 10 PWL partitions while the Secant method used a third-order power series (11) to model the PA and three iteration using (9). For comparison, an IBO was applied to the PA so that the inband spectrum of the PA is leveled to that of the PDs'. The spectrum plots show an ACI suppression of approximately 12-15 dB were achieved by the two PDs.

Fig. 6 shows the measured output spectrum in more detail. In the weakly nonlinear case, the Secant-PD is observed to perform slightly better than the SCPWL-PD. On average, the Secant-PD achieved an ACPR performance of 2 dB better compared to the SCPWL-PD.

The ACPR performance of the two PDs in the strong nonlinear case is shown in Fig. 7. Note that in this case, the spectrum measurement was done before the down-converter in order to obtain a larger dynamic range (60 dB). The disadvantage of this setup is that the observation of the fifth-order IMD region is not reliable due to the effects of the missing 20 MHz filter, which was considered in the models.

In the strong nonlinear case, the Secant-PD with power series of order three was not able to model the nonlinearity

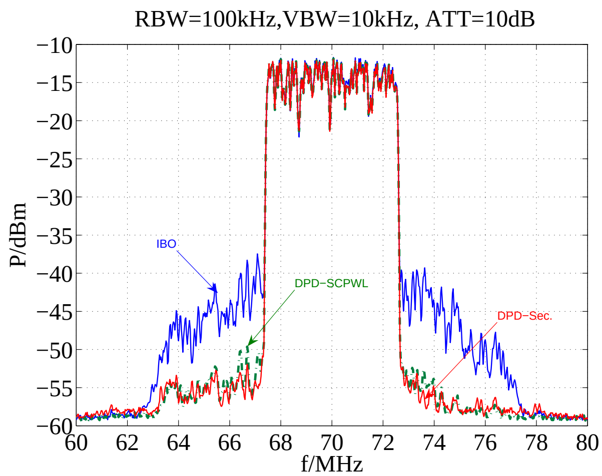

Fig. 5. Measured power spectra – comparison of SCPWL-PD and Secant method with simple IBO

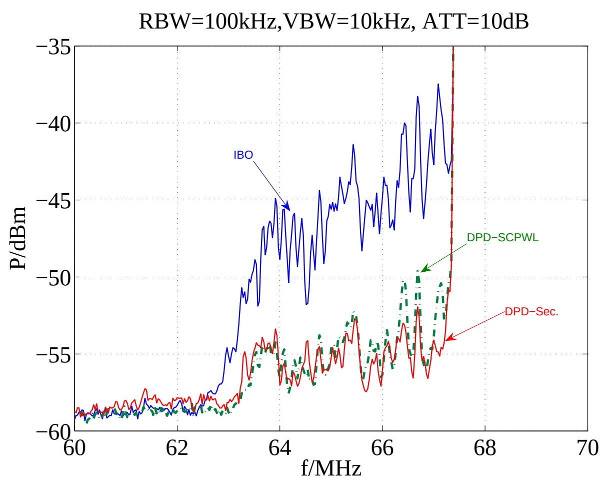

Fig. 6. Measured power spectra – IBO and DPD, based on the Secant method and the SCPWL-method, are compared.

sufficiently, thus unable to compensate the nonlinear effect. Fig. 7 shows the performance of a fifth order power series Secant-PD as compared to the SCPWL-PD and the PA's output with IBO. It is shown that the Secant-PD achieved an ACPR improvement of approximately 10 dB. The SCPWL-PD outperformed the Secant-PD by approximately 5 dB at the best case, resulting in an ACPR improvement of 15 dB.

This result is expected as it is well known that a piecewise linear function is able to model a strong nonlinearity better than a power series.

V. Conclusions

Two low-complexity PDs for microwave PA are evaluated in an experimental setup for a weakly nonlinear PA and a PA driven to saturation. The computational complexity of the two PDs is also compared.

The Secant-PD, which employs a power series to model the PA, is found to be more effective in compensating a

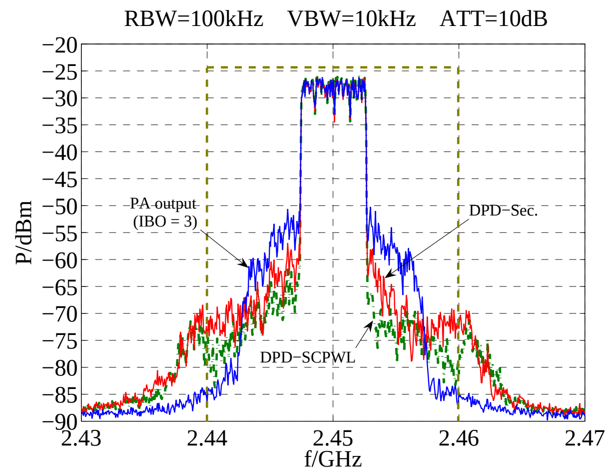

Fig. 7. Measured power spectra of PA driven to saturation – IBO and DPD the Secant method and the SCPWL method, are compared.

weakly nonlinear PA, being 2 dB better than the SCPWL-PD on average. However, in the case of a strong nonlinear PA, the SCPWL-PD is superior by 5 dB on average compared to the Secant-PD. This is due to the limitation of power series in modeling strong nonlinearities. In terms of complexity, the SCPWL-PD implementation requires more operations as compared to the Secant-PD.

In order to obtain more accurate measurement data, improvements can be made on the measurement system as discussed in Sec. III.

References

[1] 3GPP, "TS 25.141 V7.0.0 6.5.2.2," Specifications, 2005.

[2] Changsoo Eun and Edward J. Powers, "A new Volterra predistorter based on the indirect learning architecture," *IEEE Trans. Signal Processing*, vol. 45, no. 1, pp. 223–227, Jan. 1997.

[3] K. J. Muhonen, M. Kavehrad, and R. Krishnamoorthy, "Look-up table techniques for adaptive digital predistortion: A development and comparison," *IEEE Trans. Veh. Technol.*, vol. 49, no. 5, pp. 1995–2002, Sept. 2000.

[4] Yuanming Ding, Lianming Sun, and Akira Sano, "Adaptive nonlinear predistortion schemes with application to OFDM system," in *Proc. of IEEE 2003 Conference on Control Applications*, Istanbul, Turkey, June 2003, vol. 1, pp. 1130–1135.

[5] In-Seung Park and Edward J. Powers, "A new predistorter design technique for nonlinear digital communications channels," in *Proc. The International Symposium on Signal Processing and Its Application ISSPA*, Gold Coast, Australia, 1996, pp. 618–621.

[6] Mei Yen Cheong, Stefan Werner, Juan Cousseau, and Timo I. Laakso, "Predistorter identification using the Simplicial Canonical Piecewise Linear function," in *Proc. 12th International Conference on Telecommunications, ICT 2005,*, Cape Town, South Africa, May 2005.

[7] Ernst Aschbacher, Holger Arthaber, and Markus Rupp, "A fast algorithm for digiatl pre-distorter of nonlinear power amplifiers," in *Proc. The 13th European Signal Processing Conference (EUSIPCO '05)*, Antalya, Turkey, Sept. 2005.

[8] A. Hatcher, *Algebraic topology*, Cambridge University Press, 2002.

[9] J. Figueroa, Juan Coussau, and R. J. P. de Figueiredo, "A Simplicial Canonical Piecewise Linear adaptive filter," *Circ. Syst. and Signal Proc.*, vol. 5, no. 23, pp. 365–386, 2004.

[10] Sundance Multiprocessor Technology Ltd., Chesham, UK, *SMT370 User Manual*.