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PREDISTORTER DESIGN EMPLOYING PARALLEL PIECEWISE LINEAR STRUCTURE AND INVERSE COORDINATE MAPPING FOR BROADBAND COMMUNICATIONS

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ABSTRACT

This paper proposes a low-complexity predistorter (PD) for compensation of both the AM/AM and the AM/PM conversions with memory. The nonlinear power amplifier (PA) is modeled as a Wiener type nonlinearity. The quasi-static nonlinearities are modeled using a class of piecewise linear (PWL) functions. The PWL function facilitates an efficient PD identification algorithm. The proposed algorithm involves a novel inverse coordinate mapping (ICM) method that maps the nonlinear characteristics of the PA to that of the PD, and parameter estimations that do not require matrix inversion. The indirect learning architecture is used to provide an on-line compensation of the memory effect of the PA. Simulation results show that the PD that compensates also the AM/PM distortion performs significantly better than one that considers only the AM/AM nonlinearity. The proposed PD is also shown to outperform the orthogonal polynomial PD in both adjacent channel interference suppression and in-band distortion compensation.

1. INTRODUCTION

In communications systems, the power amplifier (PA) in the transmitter is an important component. It compensates the signal attenuation caused by path loss between the transmitter and the receiver. Power efficiency of the PAs in the base stations contributes directly to the network operation cost savings. Due to the intrinsic characteristics of existing PA designs, power efficient PAs are nonlinear. Nonlinear PA causes in-band and out-of-band distortion to the transmitted signals, which lead to bit-error-rate (BER) degradation and adjacent channel interference (ACI), respectively. In broadband systems, the nonlinear PAs also exhibit memory effects.

In practice, the operating point of the PA is chosen to optimize between power efficiency and linearity. For system transmitting constant envelope signals, the optimum operating point is easier to determine. However, the trend for the future mobile communications systems is to employ linear modulation scheme such as M-QAM and multi-carrier systems such as orthogonal-frequency division multiplexing (OFDM) and multi-carrier code division multiple access (MC-CDMA) to gain spectral efficiency and robustness towards interference and multipath effects. While linear modulation produces non-constant modulus constellation, multi-carrier system produces high peak-to-average-power ratio (PAPR) signals. Due to the high PAPR, the optimum operating point for PAs are more difficult to determine without sacrificing power efficiency by means of large power back-off. The spectral mask requirements for the 3G WCDMA system require an adjacent channel power ratio (ACPR) of more than 45 dBc and more than 60 dBc at the mobile terminals and base stations, respectively [1]. To comply with this requirement, linearization techniques are required when power efficient PAs are employed.

Digital baseband predistorter (PD) is a promising linearization technique due to its effectiveness and low-cost implementation. Volterra model based PDs have been proposed for compensation of nonlinear PA with memory in [2] [3] [4]. However, the designs are complex and render computationally demanding implementations. As alternatives to the Volterra model, the Wiener and Hammerstein models have been considered for the modeling and compensation of nonlinear system with memory. The most popular models considered for modeling the static nonlinear block of these systems are polynomial models [3] [5] [6]. The drawback of using a conventional polynomial for nonlinear system modeling is the numerical problem associated with matrix inversions in the parameter estimation. To alleviate the numerical problems, orthogonal polynomials have been proposed for PD designs in [7] [8].

In this paper, the simplicial canonical piecewise linear (SCPWL) function [9] is used to parameterize the static nonlinearities of the PA and PD. The SCPWL function was first proposed for PD design in [10]. The PD identification method in [10] is limited to normalized AM/AM responses of the PA and does not compensate for AM/PM distortion. This paper extends the application of the SCPWL function to model quasi-static nonlinearities, i.e., both the AM/AM and AM/PM functions. A novel inverse coordinate mapping (ICM) method is introduced. The ICM method does not require the AM/AM response to be normalized as was the case in [10]. The PD is also extended to a semi-adaptive algorithm, where the memory of the PA is compensated adaptively using the indirect learning architecture [2]. Finally, the performance of the SCPWL PD is compared to that of the orthogonal PD proposed in [7].

In Section 2 the models for the PA to be linearized and the PD are described. The development of the novel ICM method and the estimation of the SCPWL coefficients are presented in Section 3. The SCPWL function is briefly reviewed in this section. The PD identification procedure is discussed in Section 4 followed by simulation examples in Section 5. Conclusions are drawn in Section 6.

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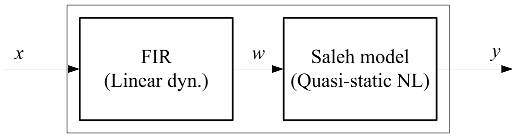

Figure 1: A Wiener model PA

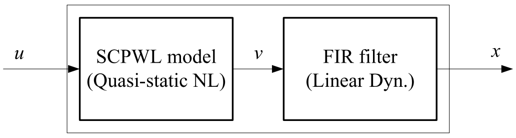

Figure 2: A Hammerstein model PD

2. SYSTEM MODEL

This section introduces the PA model and the PD model used in our simulation examples. The input, output and intermediate signals of these block models are defined in this section.

2.1 Power amplifier model

Fig. 1 illustrates the Wiener model PA to be compensated. It is composed of a linear dynamic block followed by a quasi-static nonlinear block. The input, output and intermediate signals of the PA model are denoted by x, y and w, respectively. The linear dynamic block is modeled with an FIR filter $\mathbf{h} = \begin{bmatrix} h_0 & h_1 & \cdots & h_{L_h-1} \end{bmatrix}^{\mathrm{T}}$ of length L_h. The quasi-static block is modeled with a Saleh model [11] given by

$$y(|w|) = A(|w|)\exp\left[j(\theta + P(|w|)\right], \tag{1}$$

where θ is the phase of w. The functions $A(|w|)$ and $P(|w|)$ are respectively the AM/AM function and AM/PM function of the PA given by

$$A(|w|) = \frac{\alpha_a |w|}{1+\xi_a |w|^2} \tag{2}$$

and

$$P(|w|) = \frac{\alpha_p |w|^2}{1+\xi_p |w|^2}. \tag{3}$$

where α_a is the small signal amplitude gain factor, ξ_a is the amplitude compression factor, α_p is the phase gain factor, and ξ_p is the phase compression factor.

2.2 Predistorter model

The Hammerstein model PD which has the same functional blocks as the PA but cascaded in reverse order, is shown in Fig. 2. The quasi-static block is modeled with two SCPWL functions, one for the AM/AM function and the other for AM/PM function (see Section 3 for details). An FIR filter $\mathbf{g} = \begin{bmatrix} g_0 & g_1 & \cdots & g_{L_g-1} \end{bmatrix}^{\mathrm{T}}$ is used to model the linear dynamic block. The signals of the PD are given by

$$\begin{aligned} v &= f_A^{(\mathrm{PD})}(|u|)\exp\left(j(\phi + f_P^{(\mathrm{PD})}(|u|))\right), \\ x &= \mathbf{g}^{\mathrm{H}}\mathbf{v}, \end{aligned} \tag{4}$$

where ϕ is the phase of the input signal u. The functions $f_A^{(\mathrm{PD})}(|u|)$ and $f_P^{(\mathrm{PD})}(|u|)$ are the SCPWL AM/AM and AM/PM functions, respectively. The quasi-static nonlinearities of the PD are identified using the ICM method that is presented in the following section.

3. THE PD IDENTIFICATION ALGORITHM

This section develops the ICM method and the estimation of the SCPWL coefficients that are used to identify the PD. First, the SCPWL function is briefly reviewed. Thereafter, the ICM method is developed by exploiting the linear affine property of the SCPWL function. Finally, the estimation method for the SCPWL coefficients is discussed.

3.1 SCPWL function

The SCPWL function introduced in [9] uses σ predefined partition points to divide the input space to $(\sigma - 1)$ linear affine regions. The output of the function is given by

$$f\left[s(k)\right] = \Lambda^{\mathrm{T}}\left[s(k)\right]\mathbf{c}, \tag{5}$$

where s is the amplitude of the input signal to the function, and $\Lambda\left[s(k)\right] = \begin{bmatrix} 1 & \lambda_1[s(k)] & \cdots & \lambda_{\sigma-1}[s(k)] \end{bmatrix}^{\mathrm{T}}$ is the basis function vector, and $\mathbf{c} = \begin{bmatrix} c_0 & \cdots & c_{\sigma-1} \end{bmatrix}^{T}$ is the SCPWL coefficient vector. The basis function [9] is given by

$$\lambda_i\left[s(k)\right] = \begin{cases} \frac{1}{2}\left(s(k) - \beta_i + |s(k) - \beta_i|\right), & s(k) \le \beta_\sigma \\ \frac{1}{2}\left(\beta_\sigma - \beta_i + |\beta_\sigma - \beta_i|\right), & s(k) > \beta_\sigma, \end{cases} \tag{6}$$

where β_i, $(i = 1, \ldots, \sigma)$, is the ith predefined partition point. We define the basis function matrix as a collection of basis function vectors evaluated over a block of input data $\mathbf{s} = [s(1)\cdots s(k)]$ in a matrix as

$$\mathbf{L}(\mathbf{s}) = \begin{bmatrix} \Lambda[s(1)] & \cdots & \Lambda[s(k)] \end{bmatrix}^{\mathrm{T}}. \tag{7}$$

Then the block of output data can be expressed in matrix form as

$$f(\mathbf{s}) = \begin{bmatrix} f[s(1)] & \cdots & f[s(k)] \end{bmatrix}^{\mathrm{T}} = \mathbf{L}(\mathbf{s})\mathbf{c}. \tag{8}$$

3.2 Inverse coordinate mapping (ICM)

When a static nonlinearity of a PA is modeled using a PWL function, each subregion of the function is represented by a linear affine function. If the PD's nonlinearity is modeled using another PWL function, the subregions of the PD are also linear affine functions. A linear affine region can be solely represented by the two boundary coordinates that define the region. Based on this property, the ICM method is developed by finding the relationship between the partitions points of the PA and that of the PD. This relationship is represented in the proposed mapping matrix that translates the PWL partitions of the PA to that of the PD.

The ICM matrix that maps the PA's nonlinearity to the PD's nonlinearity for any arbitrary linearized gain, is developed based on the following conditions.

C1 The output of the PD-PA cascade is a linear amplified version of the input signal with linearized gain G.

C2 The output space of the PD must coincide with the input space of the PA.

In order to fulfill C1, let the gain of the PA in an arbitrary PWL region be ρ. Then the gain of the PD, denoted K, resulting in a PD-PA output gain G is

$$K = \frac{G}{\rho}. \tag{9}$$

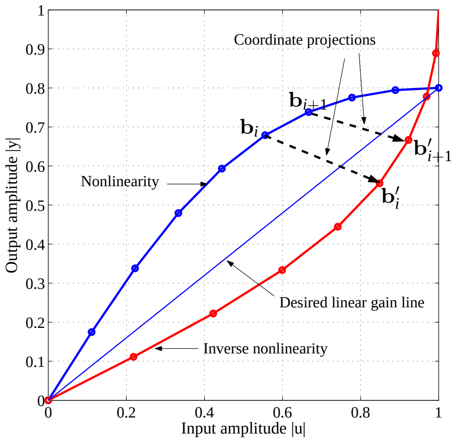

Figure 3: The Inverse-Coordinate Mapping method

To fulfill C2, we search for the input u_i to the PD that results in the output that coincides with the ith partition point β_i of the PA. It follows that

$$\frac{G}{\rho}|u_i| = \beta_i, \qquad |u_i| = \frac{1}{G}(\rho\beta_i). \tag{10}$$

Notice that $\rho\beta_i$ is the PA's response $f(\beta_i) = \rho\beta_i$. Thus, we deduce from (10), that the PD's input partition points can be mapped from that of the PA's output as

$$\beta'_i = \frac{1}{G} f(\beta_i). \tag{11}$$

Condition C2 also indicates that the signal range and the partition points of the PD's output are identical to that of the PA's input. Thus, the mapping is simply a copy of the PA's input partition points. As a result, we can define an ICM matrix for any arbitrary desired linear gain G as

$$\mathbf{Q} = \begin{bmatrix} 0 & \frac{1}{G} \\ 1 & 0 \end{bmatrix}. \tag{12}$$

The ICM matrix maps the PA's AM/AM coordinates to those of the PD's as $\mathbf{b}'_i = \mathbf{Q}\mathbf{b}_i$, as illustrated in Fig. 3.

3.3 Estimation of the SCPWL coefficients

When (8) is evaluated over all the partition points of the SCPWL function, the equation can be rewritten as

$$f(\beta) = \mathbf{L}(\beta)\mathbf{c}, \tag{13}$$

where $\mathbf{L}(\beta)$ is a nonsingular matrix. The coefficient vector can then be estimated as

$$\mathbf{c} = \mathbf{L}^{-1}(\beta) f(\beta). \tag{14}$$

As shown in [10], $\mathbf{L}^{-1}(\beta)$ is a tridiagonal matrix with elements given by the the known user defined partition sizes. Thus, no matrix inversion is needed for the estimation of $\mathbf{c}$ in (14).

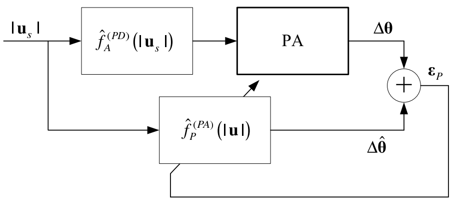

Figure 4: Identification procedure of the AM/PM SCPWL-PD

As can be seen from (4), the AM/AM and AM/PM are functions of the input amplitude only. Thus, the SCPWL functions $f_A^{(\cdot)}(\beta)$ and $f_P^{(\cdot)}(\beta)$ share a same basis function matrix $\mathbf{L}(\beta)$. The simultaneous estimation of the SCPWL coefficient vectors $\mathbf{c}_a$ and $\mathbf{c}_p$ for the AM/AM and AM/PM, respectively, can be written in matrix form as

$$[\mathbf{c}_a \quad \mathbf{c}_p] = \mathbf{L}^{-1}(\beta)\left[f_A^{(\cdot)}(\beta) \quad f_P^{(\cdot)}(\beta)\right]. \tag{15}$$

4. THE HAMMERSTEIN MODEL PD IDENTIFICATION

The Hammerstein PD is identified in two steps. First, the quasi-static model of the PD is identified in a training session. Thereafter, the linear dynamics of the PD is identified adaptively using the indirect learning architecture [2] [7]. The following subsections provide the details of the two steps.

4.1 PD's nonlinear block

The quasi-static block of the PA is identified as two parallel SCPWL functions by exciting the PA with a power-swept single-tone signal $\mathbf{u}_s$. The output amplitude and output phase shift caused by the AM/AM and AM/PM conversions are fitted to the SCPWL functions $f_A^{(\mathrm{PA})}(|\mathbf{u}_s|)$ and $f_P^{(\mathrm{PA})}(|\mathbf{u}_s|)$, respectively. Then the coordinates corresponding to the responses of the AM/AM and AM/PM at the predefined SCPWL partition points $\beta = [\beta_1 \quad \cdots \quad \beta_\sigma]$ are determined. The coefficient vectors are then identified using (15).

Thereafter, the ICM matrix (12) is used to map the PA's AM/AM coordinates to that of the PD's as shown in Fig. 3. The partition points of the PD and the corresponding AM/AM responses $\mathbf{b}'_A = \left[\beta'_A \quad f_A^{\mathrm{PD}}(\beta'_A)\right]^{\mathrm{T}}$ are obtained. The SCPWL coefficient vector κ_a for the AM/AM PD is identified using (14) as

$$\kappa_a = \mathbf{L}^{-1}(\beta') f_A^{(\mathrm{PD})}(\beta'). \tag{16}$$

Note that the PD's AM/PM function has to take into account the total effect of the AM/AM conversion caused by the PA and the PD itself. Therefore, the AM/PM model of the PA $f_P^{(\mathrm{PA})}(\cdot)$ is fitted to the phase shift at the output of the AM/AM predistorted PA as shown in Fig. 4. The coefficient vector $\mathbf{c}_p$ is obtained using (14). Since the desired linearized output phase shift of the PD-PA cascade is zero the AM/PM coefficient of the PD is obtained as

$$\kappa_p = -\mathbf{c}_p. \tag{17}$$

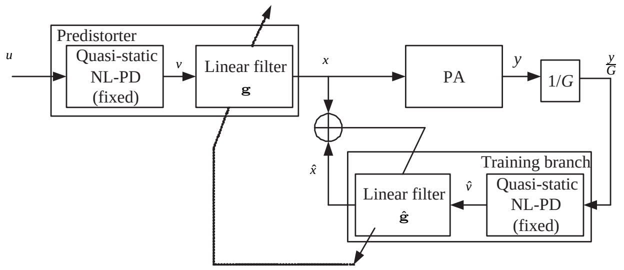

Figure 5: The indirect learning architecture for memory compensation

4.2 PD linear dynamic block

After the training session an indirect learning architecture as shown in Fig. 5 is used to compensate the PA's memory iteratively. A broadband multisine signal is used to emulate a general multicarrier signal transmitted by the system. The static nonlinear PD blocks in the architecture is fixed while the LMS algorithm updates the linear filter coefficients in the training branch. The filter coefficients are then copied to the PD. The LMS update equation for the linear filter is

$$\hat{\mathbf{g}}(k) = \hat{\mathbf{g}}(k-1) + \mu\varepsilon^*\hat{\mathbf{v}}(k), \tag{18}$$

where μ is the step size that controls the convergence of the algorithm, and $\hat{\mathbf{v}}(k)$ is a vector with the L_g most recent output samples.

5. SIMULATIONS

The PA described in Sec. 2 is the device under test (DUT) in our simulations. The static nonlinear block parameters are $\alpha_a = 2$, $\xi_a = 1$, $\alpha_p = 3.5$ and $\xi_p = 25$. The FIR filter parameters of the linear dynamic part is $\mathbf{h} = [0.7692 \quad 0.1538 \quad 0.0769]^{\mathrm{T}}$ [3].

The Hammerstein model SCPWL-PD, identified as described in Section 4 is used to compensate the DUT. The number of PWL partitions examined for the modeling of the AM/AM and AM/PM functions are $\sigma = 6$, 8, 12, 15 and 18. The length of the FIR filter that models the linear dynamic block is $L_h = 3$ and $L_h = 5$.

The influence of the number of PWL partitions on the accuracy of the model is shown in Fig. 6. The mean-squared error (MSE) between the training branch and the PD output after the FIR filter has converged is used as a performance metric. It is observed that by increasing the number of partition points from 6 to 12, the MSE dropped by almost 2 orders of magnitude. However, increasing the number of partition points beyond 15 does not provide significant improvements. In the following, the SCPWL-PD with $\sigma = 12$ and $\sigma = 15$ are used in the performance evaluation of out-of-band and in-band distortion, respectively.

The ACPR improvement using the SCPWL-PD with $\sigma = 12$ that compensates nonlinearity with memory is compared to an uncompensated PA in Fig. 7. The input back off (IBO) of the multisine excitation signal is 9 dB. The IBO is defined as

$$\mathrm{IBO} = 10\log\left(\frac{P_{i,\mathrm{sat}}}{P_i}\right), \tag{19}$$

where P_i and $P_{i,\mathrm{sat}}$ are the mean input signal power and the input signal power at saturation, respectively. In order to

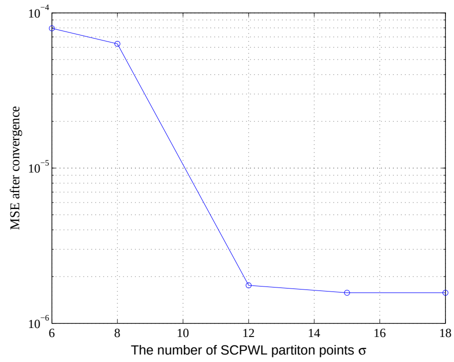

Figure 6: Influence of the number of partition points σ on model accuracy

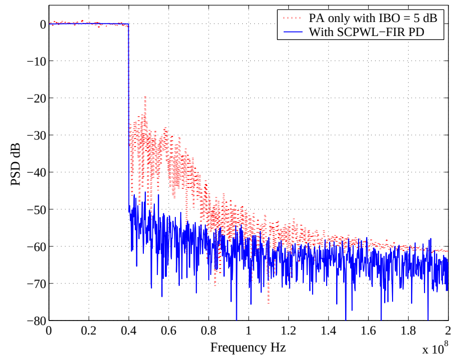

Figure 7: The ACPR improvement using a SCPWL-PD with $\sigma = 12$ compared to an uncompensated PA.

match the in-band signal attenuation caused by the PD, the input signal to the PA is further backed off by 5 dB, i.e., a total of IBO = 14 dB, for a fair comparison of ACPRs. The average ACPR improvement achieved is 20 dB, resulting in an average ACPR of 50 dBc. Fig. 8 shows the ACI suppression obtained by an SCPWL-PD that compensates only the AM/AM nonlinearity and an SCPWL-PD that compensates both AM/AM and AM/PM nonlineairities. It is observed that the quasi-static PD performs 10 dB better than the AM/AM PD. Next, the performance of the SCPWL-PD is compared to that of the orthogonal polynomial PD (OP-PD) proposed in [7]. Fig. 9 shows the average out-of-band power of the PA output compensated by the SCPWL-PD and the OP-PD. The out-of-band power is calculated by averaging the adjacent channel power over the third-order and fifth-order intermodulation distortion (IMD) zones. The results show that the SCPWL-PD outperforms the 3rd-order, 5th-order and 7th-order OP-PD by approximately 1 dB when the corresponding SCPWL-PD uses $\sigma = 6$, 8 and 12 respectively. It is also observed that a 15-partition point SCPWL-PD results in a further ACI suppression of approximately 3 dB.

To illustrate the compensation of in-band distortion, Fig. 10 shows the bit-error rate (BER) performance generated with 2000 OFDM symbols, each with 128 subcarriers employing 16-QAM modulation received over an AWGN channel. The IBO of the OFDM signal used in the BER evalution is 8 dB. The quasi-static SCPWL-PD and the OP-PD achieved similar performance in compensation of in-band distortion. Both achieved BER in the range of $10^{-}3$ and 10^{-4} for SNR above 15 dB and 18 dB, respectively. Further improvement in BER performance is observed when

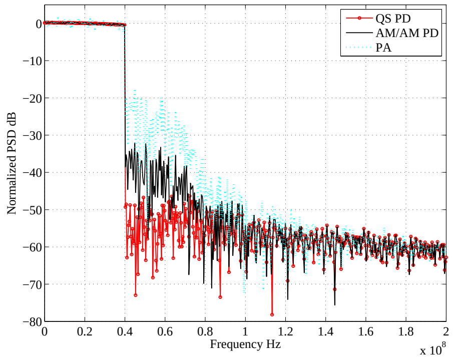

Figure 8: Performance comparison of an SCPWL AM/AM PD and an SCPWL quasi-static PD.

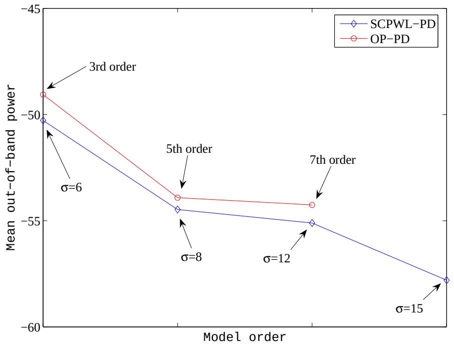

Figure 9: Performance comparison of the SCPWL-PD and the OP-PD. [Note: The x-axis scale does not indicate equal complexity model order of the two PDs.]

the SCPWL-PD that compensates nonlinearity with memory is employed. Approximately 1 dB gain is obtained at SNR $= 13$ dB, and gradually increase to a gain of 2 dB at SNR $= 18$ dB. The BER performance of the dynamic SCPWL-PD at IBO $= 8$ dB attained the performance of the system with a linear PA in an AWGN channel.

6. CONCLUSIONS

A low-complexity predistorter (PD) designed using the simplicial canonical piecewise linear (SCPWL) function that compensates for nonlinear power amplifiers (PA) with memory is proposed. The properties of the SCPWL function is exploited to develop a simple and efficient PD identification algorithm. The low-complexity PD identification algorithm includes a novel inverse coordinate mapping method, and an efficient method for the estimation of the SCPWL coefficients that does not require matrix inversions. Unlike polynomial models, the SCPWL parameter identification does not impose numerical problems. In addition, the modeling capability of the SCPWL function is highly accurate even with reasonably few number of partitions. Simulations verified that the SCPWL-PD outperforms the OP-PD in adjacent channel interference suppression as well as in-band distortion compensation. We also shown that compensation of memory effects improves both ACPR as well as BER performance more significantly as compared to PDs that compensates only static nonlinearities.

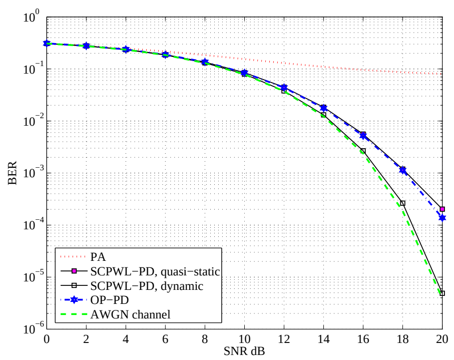

Figure 10: Bit-error rate performance comparison of the SCPWL-PD and the OP-PD.

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