POWER AMPLIFIER MODELING AND POWER AMPLIFIER PREDISTORTION IN OFDM SYSTEM

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(Manuscript Received on August 30th, 2006, Manuscript Revised December 05th, 2007)

ABSTRACT: *This paper presents a baseband predistorter to be used in OFDM systems operating with a nonlinear high power amplifier (HPA). Key features of the predistorter reside in the use of the HPA inverse structure as nonlinear distortion compensator. The performance of the compensated system is analyzed by simulations in an AWGN environment. The receiver also needs furthermore an equalizer in order to combat the distorsion effect.*

Keywords: *OFDM, DAP, HPA, Adaptive Equalizer*

1. INTRODUCTION

Nowadays, the OFDM technology is applied widely in the wireless communication because of many advantages such as robustness to severe multipath channels compared to single carrier (SC) system; effective bandwidth to FDM systems; and transceiver structures simple (based on DFT circuits) [1], [2].

However, in ODFM Systems, we can not ignore a distorsion problem introduced by nonlinear High Power Amplifier (HPA) [3], [4].The main purpose of our paper is to analyze these effects on a high speed OFDM system (WLAN2) [5]. Then, it is focused on designing a Digital Adaptive Pre-distorter (DAP) to overcome the nonlinear effect of HPA.

2. SYSTEM DESCRIPTION

Figure 1 shows the baseband equivalent system of an OFDM system [5]. The input of the system is a serial of binary data, mapped onto the M-ary QAM signal constellation to give a stream of complex symbols which are assumed to be staticscally independent. This complex symbol stream is applied to the OFDM modulation block. In the OFDM block, the stream is serial-to-parallel converted to produce a sequence c_k. c_k is transformed by a inverse fast Fourier transform (IFFT) unit. A guard interval called cyclic prefix (CP) with length T_g is added to this signal, yielding a T-spaced discrete-time representation of the transmitted signal. The nth transmitted OFDM block is given by:

$$s_n(t) = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} c_k \phi_n(t - nT) \tag{1}$$

where

$$\phi_k(t) = \begin{cases} \exp(j \cdot 2\pi \cdot f_k \cdot t), & \forall t \in [-T_g, T] \\ 0 & \text{othersiwe} \end{cases} \tag{2}$$

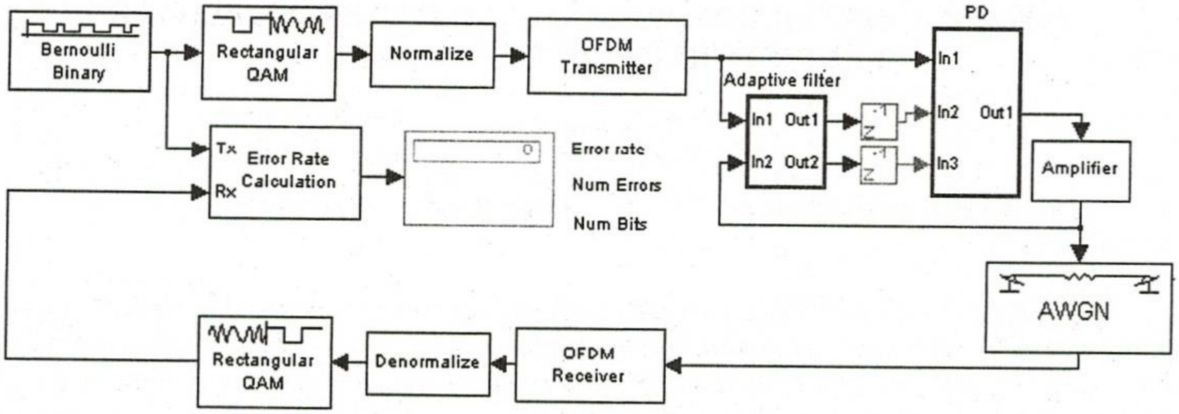

Fig 1. Baseband equivalent of the OFDM system.

where N is the number of the subcarriers. $f_k = f_0 + \frac{k}{T_u}$ and $f_0 = 0$.

The modulated signal $x(t)$ is first pre-distorted and then nonlinearly amplified, and finally propagating over a AWGN channel.

The TWT Amplifier model given in [3], [6] is used for a nonlinear HPA.

$$z(t) = A(y_\rho)\exp\left[\, j \cdot (y_\theta + B(y_\rho)) \,\right] \tag{3}$$

where y_ρ and y_θ are the amplitude and phase of the complex signal.

The function A(.) and B(.) denote AM/AM conversion (non-linear amplitude) and AM/PM conversion (non-linear phase) respectively, and are given by:

$$A(y_\rho) = \frac{2 \cdot y_\rho}{1 + y_\rho^2} \tag{4}$$

$$B(y_\rho) = \frac{2 \cdot y_\rho^2}{1 + y_\rho^2} \cdot \frac{\pi}{6} \tag{5}$$

The non-linear distortion of a TWT amplifier (TWTA) depends on the back-off. The input back-off (IBO) and the output back-off (IBO) for the amplifier are defined as

$$IBO = 10\log_{10}\left(\frac{P_{sat,i}}{P_{avg,i}}\right)$$

$$OBO = IBO - 10\log_{10}\left[1 - \exp(-\frac{R_{max}^2}{2\sigma^2})\right] \tag{6}$$

where $P_{sat,i}$ is the saturation input power and $P_{avg,i}$ is the average input power of the TWTA.

Figure 2 shows the Saleh model (a typical HPA) written in SIMULINK. Figure 3 gives the AM/AM and AM/PM characteristics of this model.

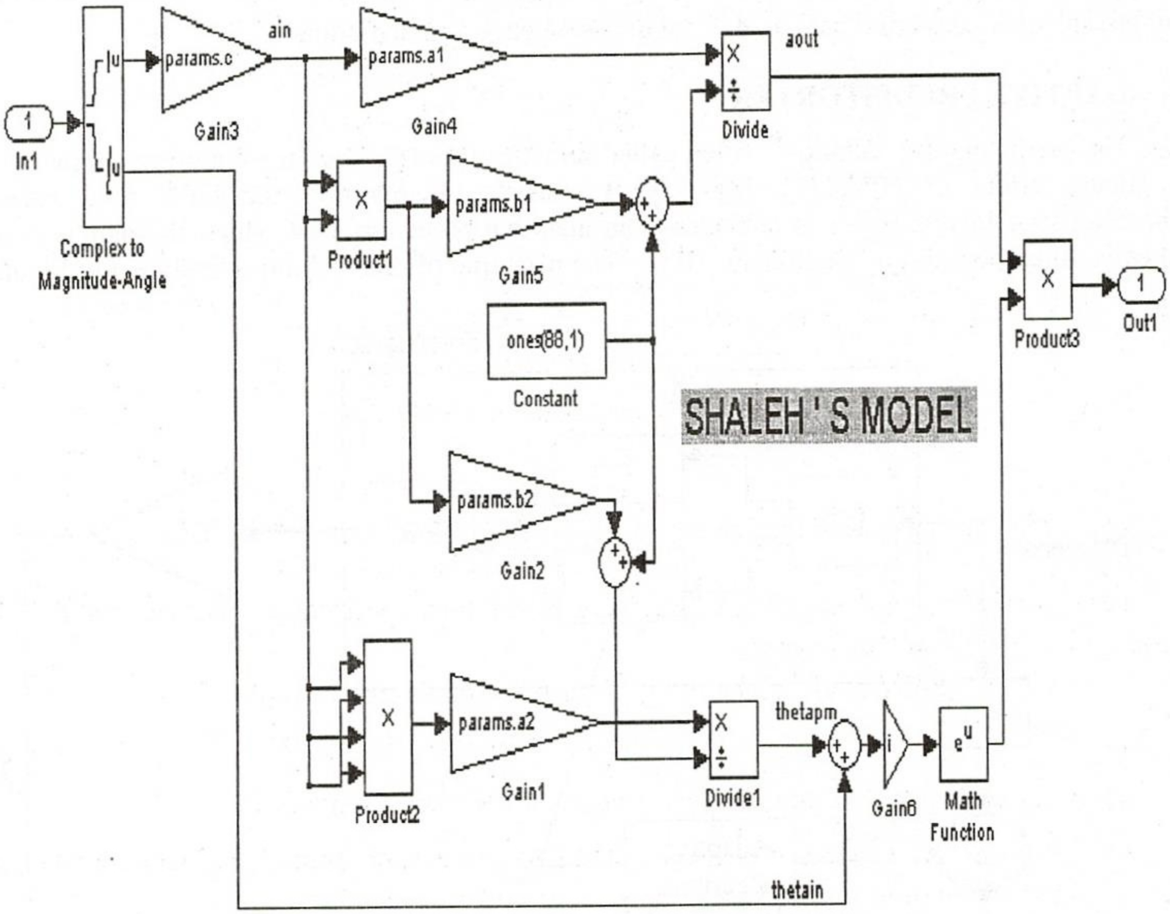

Fig 2. Saleh model

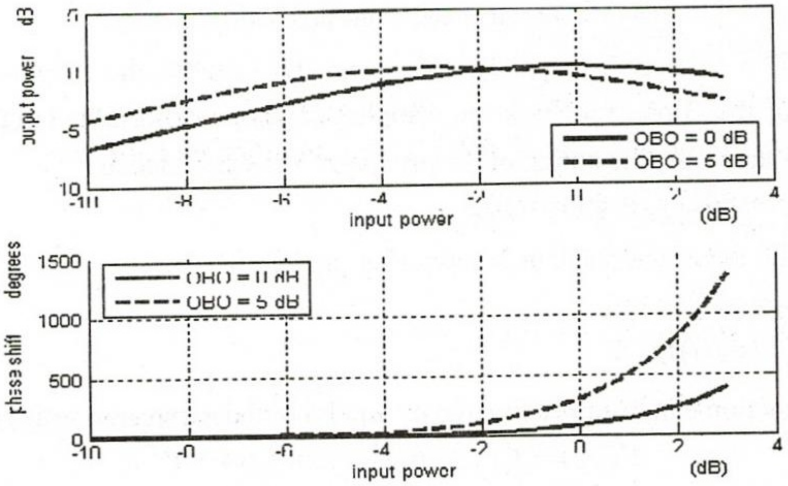

Fig 3. AM/AM and AM/PM characteristics of the Saleh model.

At the receiver, the received signal is passed through receiver filter and then sampled. The data samples are serial to parallel converted, and applied to the remove guard and FFT processor. The guard interval is removed and only the time interval $[0, T]$ is evaluated and the output signal is converted back to a serial data sequence and demodulated.

3. ADAPTIVE PREDISTORTER

The predistorsting technique, often called linearization, is a known solution to combat the nonlinear effect of HPA [7], [8], [9]. It consists of inverting the HPA nonlinearity characteristic. In this paper is considered an adaptive predistorter of which the action is to linearize the operation of a nonlinear HPA. The principle of this technique is shown in Figure 4.

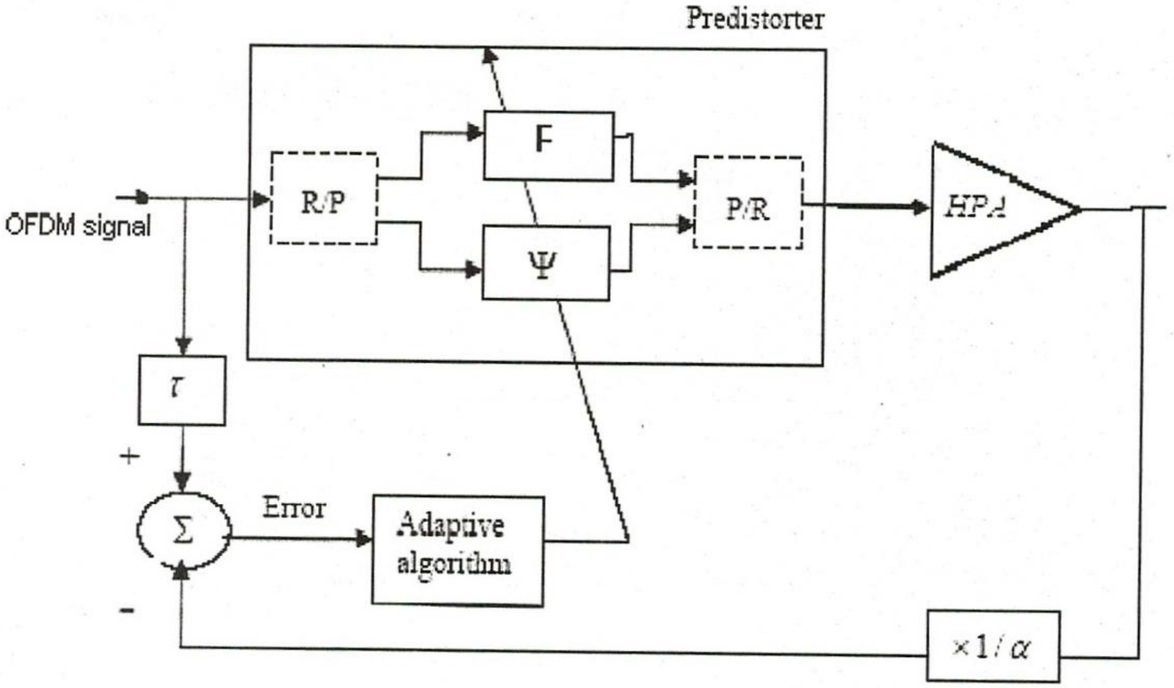

Fig 4. Baseband model of the precompensator

As mentioned in section 2 (Equ. 3), A(.) and B(.) denote the amplitude and phase transfer function of the HPA, $y_p e^{j y_\theta}$ is the complex envelope of the input signal. If we add predistorter before the HPA, the output of the predistorter is expressed as:

$$z_d = F(y_P(t))\exp(j.(y_\theta + \psi(y_P(t)))) \tag{7}$$

Ideally, we want to see the result as belows after predistorter:

$$A(F(y_P(t))) = \alpha.y_P(t)$$
$$\psi(y_P(t)) + B(F(y_P(t))) = 0 \tag{8}$$

The inverse function can be approximated by a polynomial expansion series of yP.

$$F(y_P) = f_1 y_P + f_2 y_P^2 + \dots + f_L y_P^L = V^T R_f$$
$$\psi(y_P) = \psi_0 + \psi_1 y_P + \psi_2 y_P^2 + \dots + \psi_M y_P^M = P^T R_\psi \tag{9}$$

where

$$R_f = [y_P, y_P^2, ..., y_P^L]^T$$

$$R_\psi = [y_P, y_P^2, ..., y_P^M]^T$$

$$V = [f_1, f_2, ..., f_L]^T$$

$$V = [\psi_0, \psi_1, ..., \psi_M]^T$$

The optimal coefficients V and P are determined by using Least Mean Square (LMS) algorithm:

$$J_1(V) = E((\alpha y_P - A(V^T R_f))^2) \tag{10}$$

And the updated values of V and P are:

$$\begin{aligned} V_{k+1} &= V_k + \mu_v R_{f,k} A'(V_k^T R_{f,k})(\alpha y_P - A(V^T R_f)) \\ P_{k+1} &= P_k + \mu_\psi R_{\psi,k} B'(V_k^T R_{\psi,k})(-B(F(y_P)) - P_k^T R_{\psi,k}) \end{aligned} \tag{11}$$

A'(.) and B'(.) are the derivatives of A(.) and B(.) respectively.

At first, it was intended to use one sole device, namely an equalizer, to prevent simultaneously combat the HPA nonlinearity effect and the AWGN effect. But it leads to a very complex structure for the equalizer. Therefore a trade-off is made by introducing a PD to take care of the HPA nonlinearity and an simple equalizer to compensate the AWGN channel effect.

As mentioned in section 2 (Equ. 3), we assume that h(n) is the discrete response of the channel. The received sample can be expressed as:

$$y(n) = x(n) \times h(n) + d(n) \tag{12}$$

With the help of the CP, Equ. (12) can be expressed in frequency domain as:

$$Y(z) = X(z)H(z) + D(z) \tag{13}$$

To compensate the channel effects, a FIR linear equalizer with transfer function C(z) is used to estimate the signal X(z):

$$\hat{X}(z) = C(z)Y(z) = C(z)X(z)H(z) + C(z)D(z) \tag{14}$$

This equalizer is added between demapping block and the OFDM demodulator as shown in Figure 5. In this structure, the pilot driven 1-tap LMS algorithm is employed in order to obtain a fast response. In the OFDM modulation block, first of all, a fixed number of pilots is introduced into the data frame. At the ODFM demodulation block, this noisy pilot bit is spitted and fed to LMS block in order to determine the information about the channel characteristics. The errors are calculated by:

$$e(j,k) = \hat{X}(j,k) - \Pi(\hat{X}(j,k)) \tag{15}$$

where $\Pi(X(j,k)$ denotes the decision, k refers to sub-carrier order, j is time index. Error sequence is then used to adjust the equalizer coefficients, based on the LMS algorithm.

$$\frac{\partial \left| e^2(j,k) \right|}{C(j,k)} = 2e(j,k)Y^*(j,k) \tag{16}$$

Thus the tap gain is adjusted according to:

$$C(j+1,k) = C(j,k) - \Delta e(j,k)Y^{*}(j,k) \qquad (17)$$

where Δ is pilot constant.

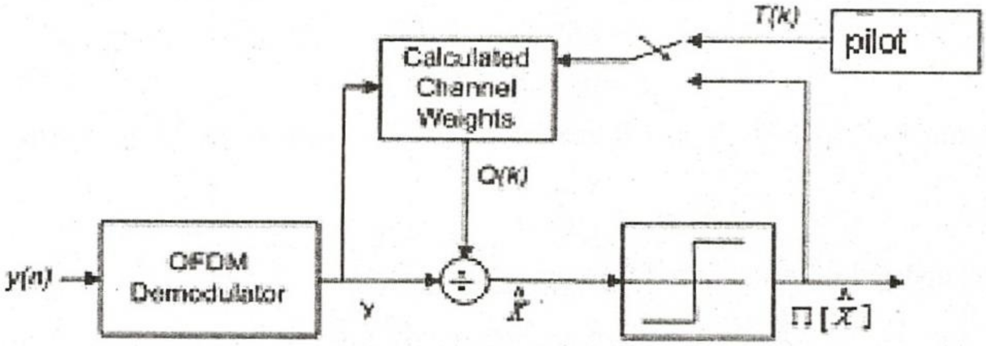

Fig 5. LMS equalizer

4. SIMULATION RESULTS

The effects of nonlinearity on the received 16-ary QAM constellations are shown in Figure 6, Figure 7 and Figure 8 which correspond to the ideal system, AWGN channel system and HPA system, respectively. In the ideal case, there are 16 well defined points. In the unideal cases, the received cloud is characteristic to the AM/AM - PM/AM nonlinearities and the AWGN channel.

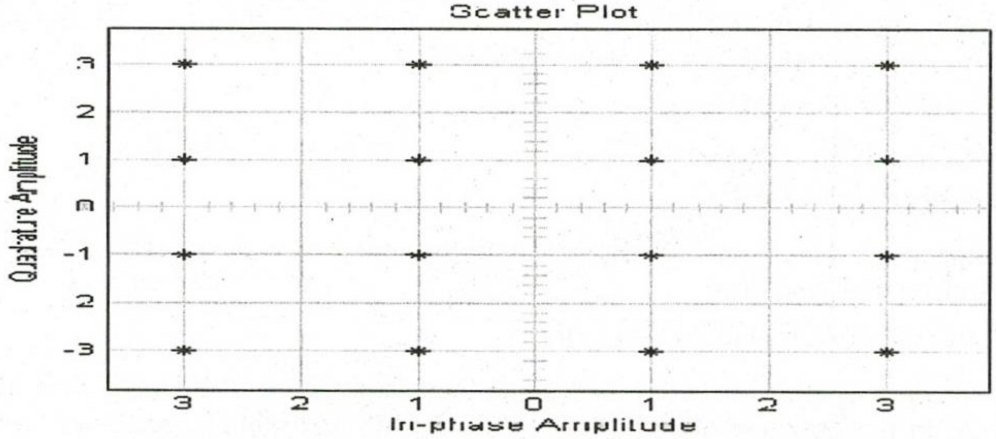

Fig 6. Received 16-ary QAM constellation with the ideal system

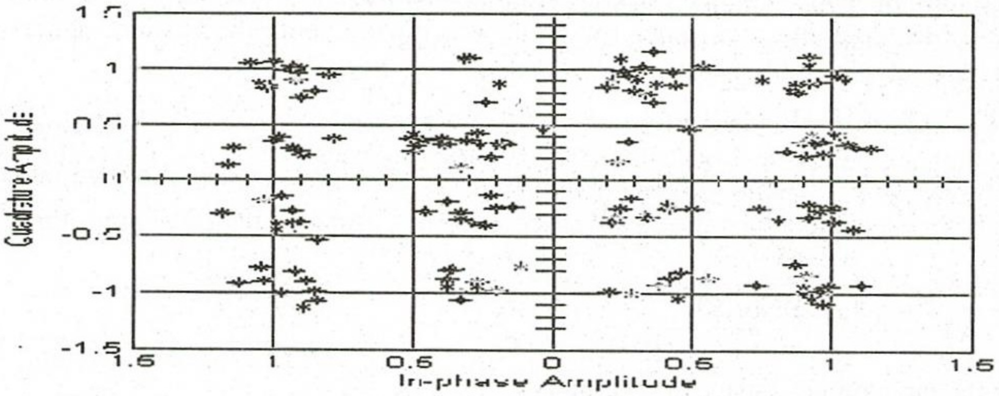

Fig 7. Received 16-ary QAM constellation with the AWGN channel

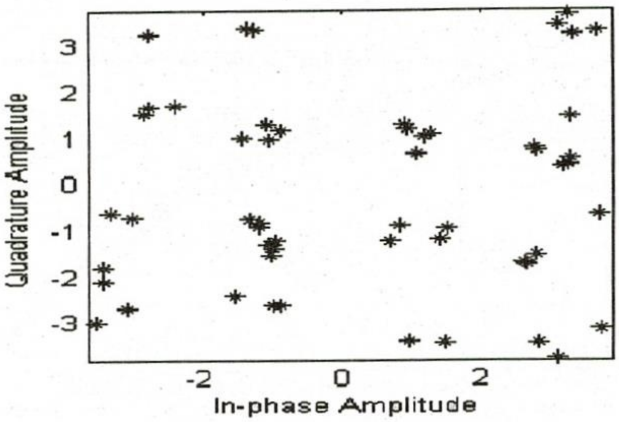

Fig 8. Received 16-ary QAM constellation with the Saleh model (HPA at OBO = 4.6 dB)

5th order polynomials are adopted to approximate the AM/AM conversion characteristic of the Saleh model. Figure 9 illustrates the convergence of coefficients of the predistorter.

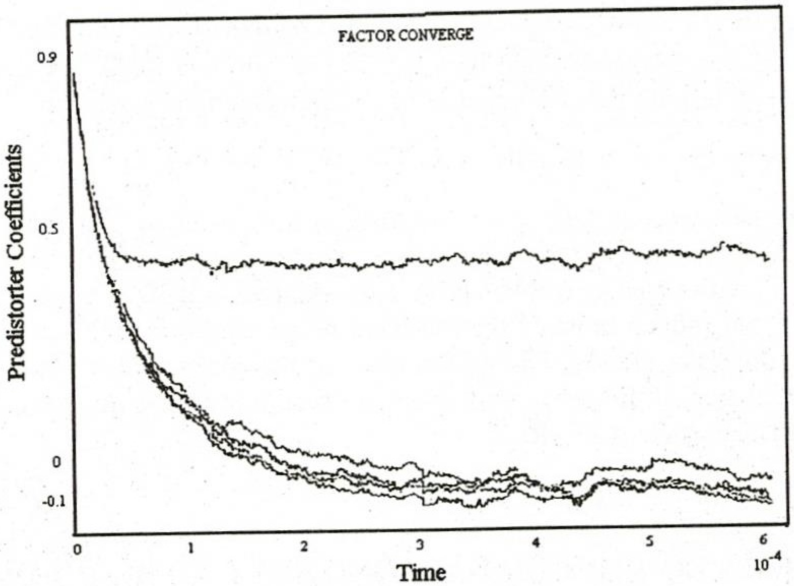

Fig 9. Convergence of coefficients in amplitude predistorter

To demonstrate the performance of the proposed linearization system, we evaluated the Bit Error Rate (BER) using Monte Carlo simulation for systems with the PD and Adaptive Equalizer. For comparison purpose, we also show the performance for systems without linearizers and system with ideal channels.

The simulations are carried out for a OFDM system with 192 subscribers and 16-ary QAM signaling on each subcarrier for 3 different scenarios listed in Table 1. Figure 10 shows the BER in term of SNR, varying between 0 and 18 dB.

Table 1. Five schemes in the proposed OFDM system

Scheme No.	Nonlinearity	Performance
1	HPA & AWGN noise	Un-PD & un-EQ
2	AWGN noise	EQ
3	HPA & AWGN noise	PD & EQ

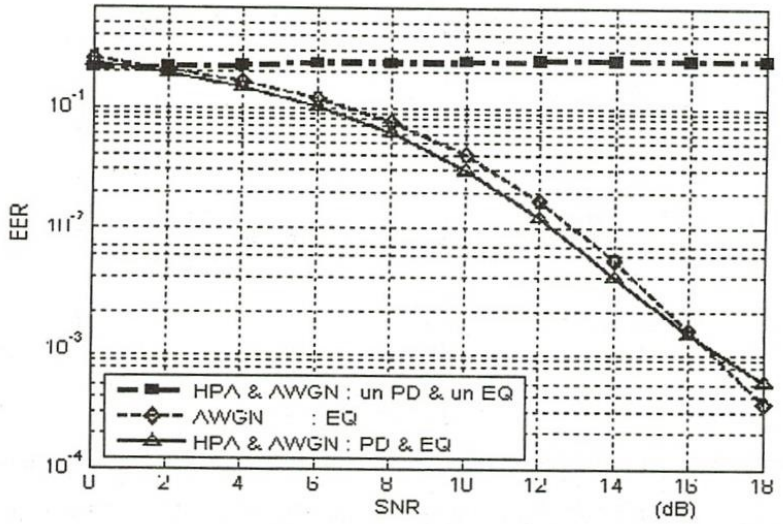

Fig 10. BER vs. SNR for the OFDM system OBO = 5 dB

The results of the 1st shows the severe impact of AWGN noise channel and HPA effects.

It's very interesting to observe that both the 2nd scenario and the 3rd have nearly the same BER. This result agrees with the convergence of coefficients in the predistorter.

5.CONCLUSION

Due to large envelope variations, the distortion introduced by nonlinear HPA is more obvious in OFDM systems. In this paper two compensation methods are combined and studied: adaptive predistorter to combat HPA and adaptive equalizer to combat the AWGN channel. The PD can reduce most of the out-band noise caused by HPA, while the eualizer LMS algorithm converges slowly. The performance of the compensated system tremendously enhanced. The next step of this work will consider other approaches to accelerate he adaptive equalizer convergence, such as ZF, RLS.

MÔ HÌNH HOÁ BỘ KHUẾCH ĐẠI CÔNG SUẤT VÀ BỘ DỰ ĐOÁN MÉO TRONG HỆ THỐNG OFDM

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TÓM TẮT: *Công nghệ điều chế số đa sóng mang trực giao (OFDM) đang được ứng dụng ngày càng rộng rãi trong lĩnh vực truyền thông không dây. Tuy nhiên các hệ thống OFDM lại chịu tác động rất lớn bởi hiện tượng phi tuyến gây ra bởi các bộ khuếch đại công suất cao. Hệ thống được mô phỏng trên kênh truyền có nhiễu trắng cộng tính (AWGN). Thuật toán thích nghi đã được sử dụng để thiết kế một bộ dự đoán méo và một bộ cân bằng nhằm loại bỏ các yếu tố phi tuyến này.*

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