Performance of UWB PSK Systems Using Fully Saturated Power Amplifiers

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***Abstract*—This paper studies the performance of ultra-wideband (UWB) radio communications systems employing phase shift keying (PSK) modulation and fully saturated power amplifiers through additive white Gaussian noise (AWGN) channel or Rayleigh fading channel. An impulse response method is provided to analyze the signal-to-noise ratio (SNR) degradation. Both convolutional code and turbo code are employed with the Viterbi decoding and the iterative decoding, respectively. The decoding algorithms are modified for the Rayleigh fading channel. Near optimal bit error rate (BER) performance is achieved when employing either convolutional code or turbo code. The simulation results match the derivation. Therefore, fully saturated power amplifiers can be employed in UWB PSK systems to significantly reduce radio cost, simplify radio circuit, and increase battery life for portable terminals.**

***Index Terms*—Bit error rate (BER), fading channel, modulation, nonlinear radio channel, ultra-wideband (UWB).**

I. Introduction

RECENTLY, the practice in ultra-wideband (UWB) wireless and satellite communications has found challenges to modulation theory. The data rate in broadband communications systems can be very high. In [1], a satellite modem supporting 800-Mbps quadrature phase shift keying (QPSK) in a single 650-MHz Ka-band channel was demonstrated, which remains as the record of demonstrated UWB systems according to the Federal Communications Commission (FCC) Part 15 definition of UWB systems with a bandwidth not less than 500 MHz. The number of terminals in a UWB satellite network can be over 10 million. The traditional modulation methods emphasizing high bandwidth efficiency require either linear power amplifiers (PAs) or complicated receivers to achieve optimal BER performance [2]. These requirements are too hard to meet when the data rate is very high or the number of terminals is very large in UWB networks [1], [3]. When the bandwidth is large, the power amplifier (PA) of the best linearity achieved using the existing technology still caused 2.4 dB signal-to-noise ratio (SNR) degradation at BER $= 10^{-5}$ [1]. At high data rates, the traditional methods for receivers to handle the PA nonlinearity are of impractical complexity. Therefore, studying modulation methods to tolerate PA nonlinearity and have low complexity is very important for the practice of UWB communications [1].

UWB radio communications systems need PAs of higher dc-to-ac power conversion efficiency [3], [4]. In practice, high dc-to-ac power conversion efficiency can be achieved when the PA works in the fully saturated region, i.e., the 3 dB gain compression region [3], [4]. However, the saturation can cause unacceptable degradation of system performance in the traditional PSK communications systems. For this reason, traditional PSK communications systems such as IS-136, IS-95, CDMA 2000, and satellite networks have employed class AB PAs which do not work in the saturation region and provide excellent linearity. Having high cost and low dc-to-ac power conversion efficiency, class AB PAs are not good for UWB radio systems.

In [1], it was shown at the BER $= 10^{-5}$ the SNR degradation was 1 dB after employing the predistortion in the system of 800 Mbps and 650-MHz bandwidth. In other words, the predistortion approach is still far from being able to completely compensate for the PA distortion. Its very high cost may be affordable by ground station or basestation [1], and is not practical for user terminals [6]. The topic to achieve near optimal BER in UWB nonlinear radio channels is open, interesting, and of great value to very large scale UWB radio networks and to the industry [1], [6]–[8].

Fully saturated PAs can achieve the ideal dc-to-ac power conversion efficiency η. Class D PAs can achieve $\eta_D = 100\%$, and class F PAs can achieve $\eta_F = 91\%$ [3], [4]. Circuits for these PAs are much simpler than that for class AB [4]. This means significant reduction of radio cost, simplified radio circuit, and increased terminal reliability. These advantages motivate us to study modulation and coding for UWB radio systems employing PAs of high dc-to-ac power conversion efficiency such as class D and class F. These PAs have the worst nonlinearity. Achieving near optimal BER performance in UWB PSK communications systems using fully saturated PAs will help to answer two open questions: 1) the achievable BER in the presence of the worst nonlinearity caused by PAs [9] and 2) the performance of error correcting codes in the presence of radio nonlinearity [10], [11].

This paper studies the performance of UWB PSK communications systems employing fully saturated power amplifiers for the additive white Gaussian noise (AWGN) channel and for Rayleigh fading channel.

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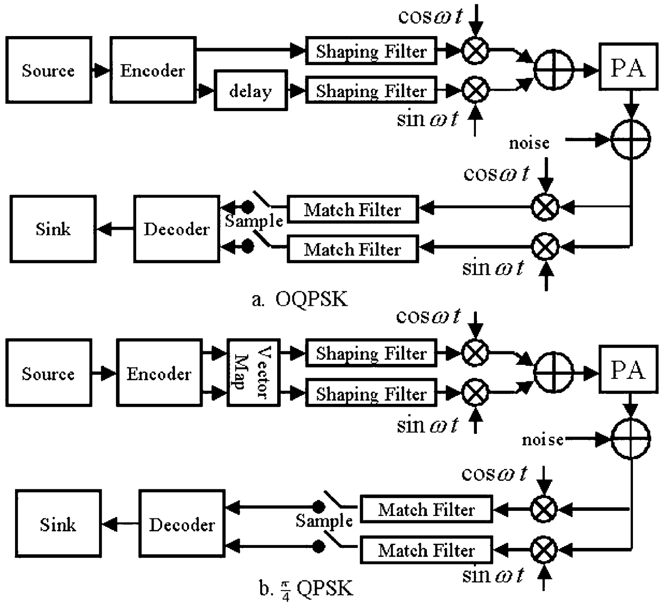

Fig. 1. Block diagram for the communication system employing OQPSK or $\pi/4$ QPSK and fully saturated power amplifiers.

II. System Model

This section describes UWB radio communications systems employing PSK and fully saturated power amplifiers. Fig. 1 is the system block diagram. We consider communications systems employing offset quadrature phase shift keying (OQPSK) or $\pi/4$ QPSK. These modulation methods have served in IS-95, CDMA2000, IS-54, IS-136, and satellite communications. They can serve well in UWB radio communications.

Fig. 1(a) shows a communication system employing OQPSK. The data sequence is fed into the encoder, which can be either a convolutional encoder or a turbo encoder. The in-phase output of the encoder is fed into a shaping filter. The quadrature output is delayed for one bit time before entering a shaping filter. The shaping filter output signals are multiplied with carriers respectively and added. The signal is amplified by a fully saturated power amplifier. The input signal $s_i(t)$ to the power amplifier can be written as

$$s_i(t) = a(t)\cos(\omega_c t + \phi(t)) \tag{1}$$

where $a(t) = \sqrt{s_I^2(t) + s_Q^2(t)}$, $\phi(t) = \arctan(s_Q(t)/s_I(t))$, and ω_c is the carrier frequency. For OQPSK systems, the in-phase baseband signal is $s_I(t) = \sum_{k=-\infty}^{\infty}\cos(\phi_k)h(t - kT_s)$ and the quadrature baseband signal is $s_Q(t) = \sum_{k=-\infty}^{\infty}\sin(\phi_k)h(t - kT_s - T_s/2)$, where $\{\phi_k\}$ is the phase containing the information, $h(t)$ is shaping function and T_s is symbol time.

For $\pi/4$ QPSK systems, as in Fig. 1(b), the in-phase signal is $s_I(t) = \sum_{k=-\infty}^{\infty} A_I[k]\cos(\phi_k)h(t - kT_s)$, and the quadrature signal is $s_Q(t) = \sum_{k=-\infty}^{\infty} A_Q[k]\sin(\phi_k)h(t - kT_s)$, where $\{A_I[k]\}$ and $\{A_Q[k]\}$ are the symbol amplitude in $\{0, 1, \sqrt{2}\}$.

When the PA is fully saturated, the envelope of the output signal is constant, and the phase shift ϕ_d in the output signal is also constant [4], [5], [9], [15], [16]. The PA output signal can be written as

$$s_o(t) = C_1\cos(\omega_c t + \phi(t) + \phi_d) + \sum_{i=2}^{\infty} C_i\cos(i\omega_c t + i\phi(t) + i\phi_d) \tag{2}$$

where $\{C_i\}$ are the cosine transform coefficients of the PA output signal. In practice, a low pass network at the output stage of the power amplifier eliminates high order harmonics [4]. The phase of the output signal is the phase of the input signal plus the constant phase shift ϕ_d, as verified by extensive measurements [6], [15], [17]. The transmitted signal can be written as

$$s(t) = \frac{C_1 s_I(t)}{\sqrt{s_I^2(t) + s_Q^2(t)}}\cos(\omega_c t + \phi_d) - \frac{C_1 s_Q(t)}{\sqrt{s_I^2(t) + s_Q^2(t)}}\sin(\omega_c t + \phi_d). \tag{3}$$

The constant phase distortion ϕ_d can be well taken care by a coherent demodulator with phase estimator and phase lock loop. It is ignorable for BER performance [2], [6], [8].

The equivalent baseband transmitted signal can be written as

$$s_b(t) = s_{bI}(t) + js_{bQ}(t) \tag{4}$$

where $s_{bI}(t) = (C_1 s_I(t)/\sqrt{s_I^2(t) + s_Q^2(t)})$, $s_{bQ}(t) = (C_1 s_Q(t)/\sqrt{s_I^2(t) + s_Q^2(t)})$, and C_1 is the gain of the PA.

The fully saturated power amplifier completely removes the amplitude information [6], [9], [15]–[17]. The distortion in baseband at the PA output is defined as

$$d(t) \triangleq s_b(t) - C_1(s_I(t) + js_Q(t)). \tag{5}$$

The PA nonlinearity is time invariant. The input signal and the PA distortion are dependent. For *M*-ary PSK with $M \geq 4$, the in-phase signal $s_{bI}(t)$ and the quadrature signal $s_{bQ}(t)$ at the PA output are correlated.

The PA output signal is transmitted through the AWGN channel or a fading plus AWGN channel. In the receiver, the signal with noise is down-converted, sampled, filtered by matched filters, and fed into the decoder.

Both the convolutional code and the turbo code are employed to study the coded system performance with fully saturated power amplifiers. The convolutional code is of rate $R = 1/2$, constraint length $K = 7$, and polynomials 133 and 171 [2]. For turbo code, the same encoding scheme as in CDMA2000 is employed [12]. The polynomials are 13 and 15. The constraint length is $K = 4$. The coding rate is $R = 1/2$. The encoder employs two systematic, recursive, convolutional encoders connected in parallel, with an interleaver in front of the second recursive encoder. The encoder output is punctured to achieve the designated data rate. A tail sequence is added so that at the end of each frame these two convolutional encoders go back to all-zero state. If the total number of information bits is N, the output includes $2N$ encoded data symbols followed by 12 tail output symbols. The interleaver in the turbo encoder performs block interleaving for the data. The interleaving algorithm is specified in [12]. The modified BCJR algorithm is implemented in the decoder [13], [14].

III. Demodulation

It is well known that hard limiter makes it extremely difficult to analyze the BER or SNR degradation in demodulation [9], [22]–[24]. For BPSK, Jacobs [9] analyzed the SNR degradation in hard limited spread spectrum systems with large signal-time-duration-noise-bandwidth products and low SNR. For *M*-ary PSK with $M \geq 4$, no work has been reported to analyze BER or SNR degradation. This section provides an impulse response method in baseband to calculate the SNR degradation caused by fully saturated power amplifiers. The method is applicable for *M*-ary PSK system with $M \in \{2, 4, 8, \ldots\}$ and small signal-time-duration-noise-bandwidth product.

The received signal in baseband is

$$r(t) = s_b(t) + n(t) \tag{6}$$

where $s_b(t)$ is the transmitted signal and $n(t)$ is AWGN with zero mean and power σ^2. It can be rewritten as

$$r(t) = C_1(s_{bI}(t) + js_{bQ}(t)) + d(t) + n(t). \tag{7}$$

In [6], [8], and [18], it was shown that the power of the distortion $d(t)$ can be minimized by proper pulse shaping in PSK systems. More precisely, the power of $d(t)$ divided by the power of the linear PSK signal can be less than −21 dB [8]. In practice, PSK modems are qualified at the signal power over the AWGN power not less than 3 dB. Therefore, the power of AWGN is much higher than the power of the PA distortion at the demodulator input. For a coherent receiver, the SNR at the decision instant determines the BER performance [2]. We propose a method to analyze the SNR degradation caused by PA nonlinearity through calculating the signal power degradation at the decision instant. The statistics of the nonlinear PA output is employed.

Assume the source generates a sequence of independent identically distributed (i.i.d.) impulses

$$x(t) = \sum_{i=-\infty}^{\infty} A_i \delta(t - iT_b) \tag{8}$$

where $A_i = \pm 1$ and T_b is the bit time. The autocorrelation is $R_x(t) = \delta(t)$ and the power spectral density is $P_x(f) = 1$. The shaping filter is a finite impulse response low-pass filter with the frequency response $H_s(f)$ and the time response $h(t)$. The baseband OQPSK signal is

$$s_I(t) = \sum_{i=-\infty}^{\infty} A_I[i] h(t - iT_s)$$

and

$$s_Q(t) = \sum_{i=-\infty}^{\infty} A_Q[i] h\left(t - iT_s - \frac{T_s}{2}\right)$$

where $\{A_I[i]\}$ and $\{A_Q[i]\}$ are sequences with elements in $\{+1, -1\}$ and T_s is the symbol time.

From the statistics of the binary impulse sequence, one can calculate the autocorrelation of the power amplifier output signal. The PSK signal is generally considered as a cyclostationary process. Define $R(t)$ as the time-average autocorrelation over the the period T_s [2]. The power spectral density of the PA output signal is

$$P(f) = \int_{-\infty}^{\infty} R(t) e^{-j2\pi f t} dt. \tag{9}$$

Since $R_x(t) = \delta(t)$ and $P_x(f) = 1$, one can find the equivalent impulse response function $h'(t)$ in baseband with the frequency response $|H'_s(f)|^2$ as

$$|H'_s(f)|^2 = \frac{P(f)}{P_x(f)} = P(f)$$

which is a low pass filter. Assume that the filter is symmetric in time domain and its phase is continuous in frequency domain for the i.i.d. input data stream. Let $\{f_i | f_i \geq 0, i = 0, 1, 2, \ldots\}$ be all the frequencies where $P(f_i) = 0, i > 0$ and $f_0 = 0$. Then $H'_s(f)$ can be written as

$$H'_s(f) = A(f)\sqrt{P(f)}$$

where

$$A(f) = \begin{cases} 1 & \text{if } |f| \in [f_i, f_{i+1}), i \text{ is even;} \\ -1 & \text{if } |f| \in [f_i, f_{i+1}), i \text{ is odd.} \end{cases}$$

Assume that the demodulation filter is matched to the shaping filter $h(t)$ [8]. One can assume that

$$H'_s(f) \approx \sqrt{P(f)}. \tag{10}$$

Denote the frequency response of the demodulation filter as $H_m(f)$. The output of the demodulation filter can be written as

$$y'(t) = \int_{-\infty}^{\infty} H'_s(f) \cdot H_m(f) \exp(2\pi j t f) df. \tag{11}$$

The sample value at the decision instant can be written as

$$\begin{aligned} y'(t=0) &= \int_{-\infty}^{\infty} H'_s(f) \cdot H_m(f) \exp(2\pi j f t) df|_{t=0} \\ &= \int_{-\infty}^{\infty} H'_s(f) \cdot H_m(f) df. \end{aligned}$$

Denote the transmitter output in a communication system employing the ideal linear power amplifier as

$$s_l(t) = K_0 \int_{-\infty}^{\infty} H_s(f) \exp(2\pi j f t) df \tag{12}$$

where K_0 is the gain for the linear system, which has the same transmitted power as that in the compatible communication system employing fully saturated power amplifiers. The output of the demodulation filter in the receiver can be written as

$$y(t) = K_0 \int_{-\infty}^{\infty} H_s(f) \cdot H_m(f) \exp(2\pi j f t) df. \tag{13}$$

Then

$$y(t=0) = K_0 \int_{-\infty}^{\infty} H_s(f) \cdot H_m(f) df. \tag{14}$$

Define the SNR degradation caused by the fully saturated power amplifier as

$$D \triangleq -20 \lg \frac{y'(0)}{y(0)}. \tag{15}$$

TABLE I
SNR DEGRADATION IN DECIBEL FOR COMMUNICATION SYSTEMS USING OQPSK OR $\pi/4$ QPSK AND ROOT RAISED COSINE FILTERS WITH THE ROLL-OFF FACTOR β

β	1.0	0.9	0.8	0.7	0.6
OQPSK	0.061	0.081	0.108	0.145	0.204
$\pi/4$ QPSK	0.296	0.285	0.292	0.308	0.339
β	0.5	0.4	0.3	0.2	0.1
OQPSK	0.277	0.382	0.504	0.641	0.763
$\pi/4$ QPSK	0.400	0.467	0.578	0.710	0.825

TABLE II
SNR DEGRADATION IN DECIBEL FOR COMMUNICATION SYSTEMS USING OQPSK OR $\pi/4$ QPSK AND THE SHAPING FILTERS IN CDMA2000

Filter Length	48	108
OQPSK	0.950	1.097
$\pi/4$ QPSK	0.998	1.138

The SNR degradation is observed at the demodulator output and can be written as

$$D = -20\lg \frac{\int_{-\infty}^{\infty} H_s'(f) \cdot H_m(f) df}{K_0 \int_{-\infty}^{\infty} H_s(f) \cdot H_m(f) df}. \qquad (16)$$

Table I shows the SNR degradation caused by the fully saturated power amplifier to the communication systems using OQPSK or $\pi/4$ QPSK. The shaping pulse is the square root raised cosine function with the roll off factor β and the filter length of 8 symbols. It can be seen that the SNR degradation is minimized when β approaches 1.

The two filters in CDMA2000 are used for comparison [12]. For the spreading rate 1, the filter length is 48 with the sampling rate 4. For the spreading rate 3, the filter length is 108 with the sampling rate 4. Employing filters in CDMA2000, the SNR degradations caused by the nonlinear power amplifier to the modulated waveforms using OQPSK or $\pi/4$ QPSK are also calculated using (16) and shown in Table II. The SNR degradation in communication systems employing CDMA2000 filters and fully saturated power amplifiers is 0.9 dB worse than that in a compatible system employing the root raised cosine filter with β close to 1.

IV. DECODING FOR RAYLEIGH FADING CHANNEL

When Rayleigh fading is present, the received signal can be written as

$$r(t) = \alpha(t)s_b(t) + n(t) \qquad (17)$$

where $\alpha(t)$ is a random process, $s_b(t)$ is the baseband signal, and $n(t)$ is AWGN with zero mean and the variance σ^2. The sample value of $\alpha(t)$ at any time instant has Rayleigh distribution

$$p_\alpha(x) = \frac{x}{a^2} e^{-(x^2/2a^2)}$$

where a is constant and $x \geq 0$ [2], [19].

This section considers the situation where the fading is slow compared with the coded symbol rate. It is assumed the packet length is long enough, and by employing an interleaver, the fading is independent from symbol to symbol. In this case, the Viterbi algorithm and the turbo decoding algorithm should be revised to compensate for the signal level variation.

It is assumed the receiver employs a coherent demodulator and has perfect channel information of the Rayleigh fading channel, i.e., the fading level is known. Consider communication systems using quadriphase modulation and convolutional code with the rate $R = (1/2)$, constraint length $K = 7$. There are 2^K states for the encoder. Let the length of the information block be B. The output of the encoder can be represented as a code tree. There are 2^B branches in the tree. Assume the encoder is in the state $s[i]$ after the first i input bits, with the $(i+1)$th input bit a_{i+1}, the output symbol is $x_{i+1}[s[i]] = \{x_{i+1,1}[s[i]], x_{i+1,2}[s[i]]\}$, and the encoder goes to the state $s[i+1]$.

In the receiver, the received symbol at time $(i+1)$ is $y_{i+1} = \{y_{i+1,1}, y_{i+1,2}\}$. The decoder calculates the likelihood function for each state of the encoder. The likelihood function for the state $s[i] = k$ is defined as

$$\begin{aligned} p(y_{i+1}|s[i]=k) &= p(y_{i+1}|x_{i+1}[s[i]=k]) \\ &= \prod_{j=1}^{2} p(y_{i+1,j}|x_{i+1,j}[s[i]=k]). \end{aligned} \qquad (18)$$

The decoder accumulates the likelihood information for each possible path. After having received the whole information block, it chooses the path with the largest likelihood as the coding path. Let $a^l = \{a_1^l, a_2^l, \ldots, a_B^l\}$ be the input bit sequence to the encoder such that the encoder goes along the lth branch, $1 \leq l \leq 2^B$. Let $s[l_i]$ be the encoder state at the time i, and define $x_i[s[l_i]] = \{x_{i,1}[s[l_i]], x_{i,2}[s[l_i]]\}$ as the output symbol at the time i. The likelihood function for the lth branch of this particular code path can be written as

$$p_l(y) = \prod_{i=1}^{B} p(y_i|x_i[s[l_i]]) = \prod_{i=1}^{B}\prod_{j=1}^{2} p(y_{i,j}|x_{i,j}[s[l_i]]). \qquad (19)$$

Taking the natural logarithm of $p_l(y)$ for the AWGN channel, one has

$$\begin{aligned} \ln p_l(y) &= \sum_{i=1}^{B}\sum_{j=1}^{2} \ln p(y_{i,j}|x_{i,j}[s[l_i]]) \\ &= -c\sum_{i=1}^{B}\sum_{j=1}^{2} (y_{i,j} - x_{i,j}[s[l_i]])^2 + d \end{aligned} \qquad (20)$$

which is

$$\ln p_k(y) = C\sum_{i=1}^{B}\sum_{j=1}^{2} (y_{i,j} \cdot x_{i,j}[s[l_i]]) + D \qquad (21)$$

where c, d, C, and D are constant corresponding to the channel characteristics with both c and C positive.

When fading is present, $p(y_{i,j}|x_{i,j}[s[l_i]])$ becomes much more complicated, and correlation in (21) cannot be applied in the decoding algorithm. With the presence of Rayleigh fading, $p(y_{i,j}|x_{i,j}[s[l_i]])$ can be written as (22), shown at the bottom of the next page. The first factor $(a/(a^2+\sigma^2)^{3/2})e^{-((y_{i,j}^2)/(2(a^2+\sigma^2)))}$ is common for all paths. It can be ignored in the decoding algorithm. We redefine the likelihood function as (23), shown at the bottom of the next page.

The Viterbi decoding algorithm is revised as in the following. At the beginning, assign the probability of 1 to the all-zero path, and 0 to all other paths. After receiving $i > 0$ symbols, the probability of each path is assigned and the number of paths is 2^7 as the constraint length is $K = 7$. We write the probability density function of the lth path at the symbol i as $p_{l,i}$. For the $(i+1)$th symbol, let

$$p_{l,i+1} = \max_j \{p_{j,i} \cdot p(y_{i+1}|x_{i+1,j})\}$$

where $j \in \{\lfloor 2 \times l \rfloor_{2^K}, \lfloor 2 \times l \rfloor_{2^K} + 1\}$, and $\lfloor x \rfloor_M$ means x modulo M. Then the path $\lfloor 2 \times l \rfloor_{2^K}$ and the path $\lfloor 2 \times l \rfloor_{2^K} + 1$ are merged.

The conventional turbo decoding algorithm for AWGN channel uses Gaussian distribution to calculates the transform probability from state to state forward and backward [13]. When fading is considered, the conditional probability distribution function should also be revised in the same way. More precisely, one should calculate the forward transform probability $p_f(y_{i,j}|x_{i,j}[s[k_i]])$ defined as $p(y_{i,j}|x_{i,j}[s[k_i]])$ above. One needs to calculate the backward transform probability function $p_b(y_{i,j}|x_{i,j}[s[k_i]])$ which is the probability density for the state s_{k_i} to come from its previous state with the ith input.

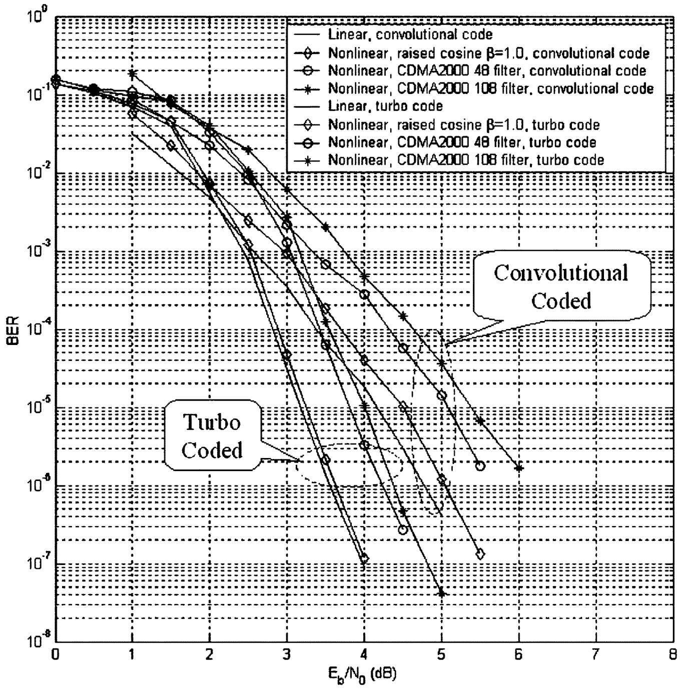

Fig. 2. BER of the communication system using OQPSK and fully saturated power amplifiers through AWGN channel.

V. Numerical Results

The simulated BER for communication systems employing the convolutional coded OQPSK and fully saturated power amplifiers is shown in Fig. 2. The BER in a compatible system employing the ideal linear power amplifier is plotted for comparison. When the system employs the rate 1/2 convolutional code with $K = 7$, the root raised cosine filter with $\beta = 1.0$ and the Viterbi decoding, the SNR degradation is about 0.2 dB at BER $= 10^{-6}$ between the BER for the system using linear power amplifiers and the BER for the system employing fully saturated power amplifiers. It is close to the calculated lower bound 0.061 dB in Table I. The SNR degradation in communications systems employing any of the two CDMA2000 filters is larger. For the 48-tap filter, the SNR degradation is about 1 dB at BER $= 10^{-5}$. Its lower bound is 0.95 dB from Table II. For the 108-tap filter, the SNR degradation is about 1.1 dB. The theoretical lower bound is 1.097 dB in Table II.

The system employing OQPSK and the rate 1/2 turbo code with $K = 4$ is also simulated. The turbo decoding algorithm is the modified BCJR algorithm [13], [14]. The results are shown in Fig. 2 for three iterations. When the root raised cosine filter with $\beta = 1.0$ is used, the BER performance of the communication system employing turbo code is almost completely immune to the power amplifier saturation. At BER $= 10^{-6}$, the SNR degradation is 0.7 dB for the 48-tap filter in CDMA2000, and 0.8 dB for the 108-tap filter in CDMA2000, respectively. The SNR degradation is relative to the BER performance of a compatible system employing ideal linear power amplifiers.

The simulated BER for communication systems employing $\pi/4$ QPSK and fully saturated power amplifiers with rate 1/2 convolutional code of $K = 7$ is shown in Fig. 3. The BER performance of a compatible system using ideal linear power amplifiers is plotted for comparison. When the root raised cosine filter with $\beta = 1$ is used, at BER $= 10^{-6}$, the SNR degradation is 0.4 dB. This agrees with the theoretical results in Table I. When CDMA2000 filters are employed with fully saturated power amplifiers, the SNR degradation is 1 dB for the 48-tap filter, and 1.2 dB for the 108-tap filter. The SNR degradations also match the theoretical results in Section III. Fig. 3

$$p(y_{i,j}|x_{i,j}[s[l_i]]) = \begin{cases} \frac{a}{(a^2+\sigma^2)^{3/2}} e^{-((y_{i,j}^2)/2(a^2+\sigma^2))} \left[y_{i,j} Q\left(\frac{-a y_{i,j}}{\sigma\sqrt{a^2+\sigma^2}}\right) + \frac{\sigma\sqrt{a^2+\sigma^2}}{a} e^{-((a^2 y_{i,j}^2)/2\sigma^2(a^2+\sigma^2))} \right], & x_{i,j}[s[l_i]] = 1; \\ \frac{a}{(a^2+\sigma^2)^{3/2}} e^{-((y_{i,j}^2)/2(a^2+\sigma^2))} \left[-y_{i,j} Q\left(\frac{a y_{i,j}}{\sigma\sqrt{a^2+\sigma^2}}\right) + \frac{\sigma\sqrt{a^2+\sigma^2}}{a} e^{-((a^2 y_{i,j}^2)/2\sigma^2(a^2+\sigma^2))} \right], & x_{i,j}[s[l_i]] = 0. \end{cases} \quad (22)$$

$$p(y_{i,j}|x_{i,j}[s[l_i]]) = \begin{cases} y_{i,j} Q\left(\frac{-a y_{i,j}}{\sigma\sqrt{a^2+\sigma^2}}\right) + \frac{\sigma\sqrt{a^2+\sigma^2}}{a} e^{-((a^2 y_{i,j}^2)/(2\sigma^2(a^2+\sigma^2)))}, & x_{i,j}[s[l_i]] = 1; \\ -y_{i,j} Q\left(\frac{a y_{i,j}}{\sigma\sqrt{a^2+\sigma^2}}\right) + \frac{\sigma\sqrt{a^2+\sigma^2}}{a} e^{-((a^2 y_{i,j}^2)/(2\sigma^2(a^2+\sigma^2)))}, & x_{i,j}[s[l_i]] = 0. \end{cases} \quad (23)$$

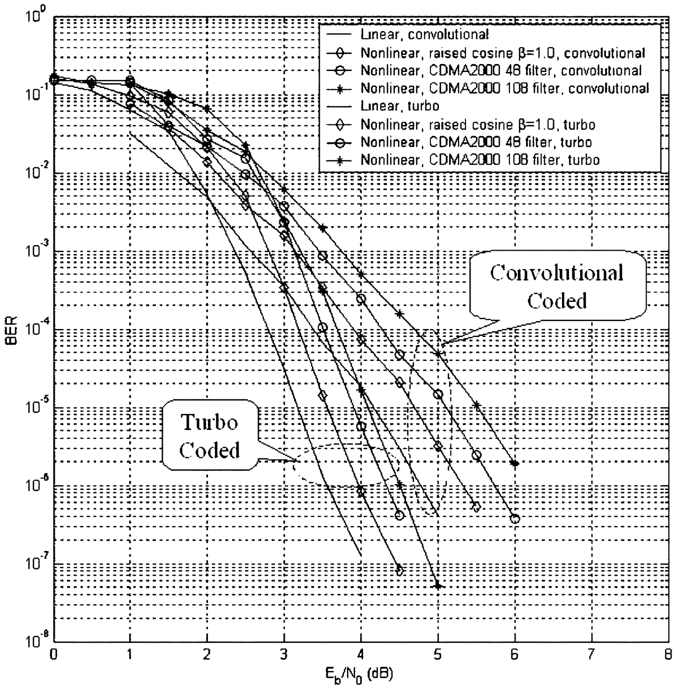

Fig. 3. BER of the communication system using $\pi/4$ QPSK and fully saturated power amplifiers through AWGN channel.

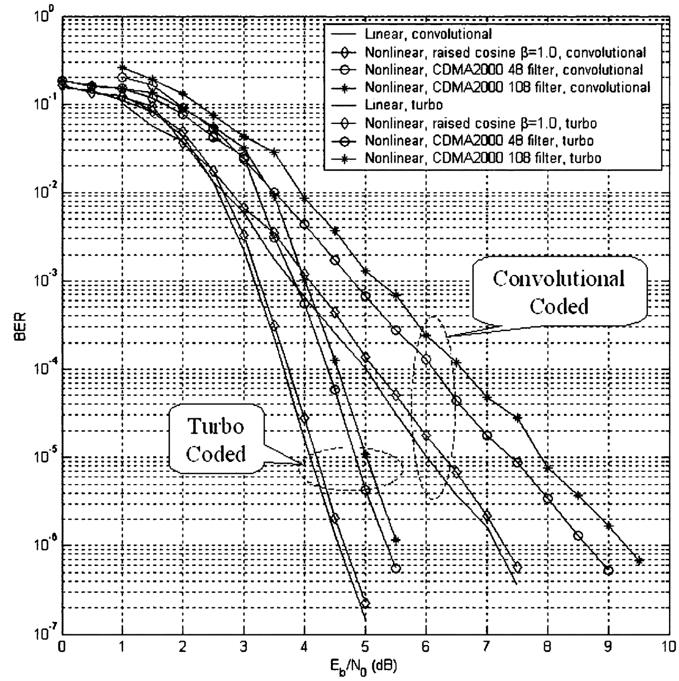

Fig. 4. BER of the communication system using OQPSK and fully saturated power amplifiers through Rayleigh fading channel.

also shows the simulated BER for communication systems employing $\pi/4$ QPSK and fully saturated power amplifiers with rate 1/2 turbo code of $K = 4$, and BER for a compatible communication system employing ideal linear power amplifiers. At BER $= 10^{-6}$, the SNR degradation caused by the fully saturated power amplifiers is 0.4 dB for the root raised cosine filter with $\beta = 1.0$, 0.7 dB for the 48-tap filter in CDMA2000, and 1.0 dB for the 108-tap filter in CDMA2000, respectively.

We have simulated the coded BER performance for communications systems employing fully saturated power amplifier through Rayleigh fading channel. The BER for OQPSK communication systems employing the rate 1/2 convolutional code with $K = 7$ and fully saturated power amplifier through the Rayleigh fading channel is shown in Fig. 4. It also shows the results of turbo coded OQPSK through the fading channel. As it can be seen, the SNR degradations caused by the fully saturated power amplifiers are almost the same for both the Rayleigh fading channel and the AWGN channel.

Fig. 5 shows the results for the coded communications systems employing $\pi/4$ QPSK and fully saturated power amplifiers through Rayleigh fading channels. The power amplifier nonlinearity does not cause the SNR degradation to increase when Rayleigh fading is present.

When turbo code is employed, the SNR degradation decreases to the value D given in Section III. This is because the turbo decoder estimates the parameters for decoding to achieve the best performance. Communications systems employing OQPSK can achieve a better BER performance than $\pi/4$ QPSK.

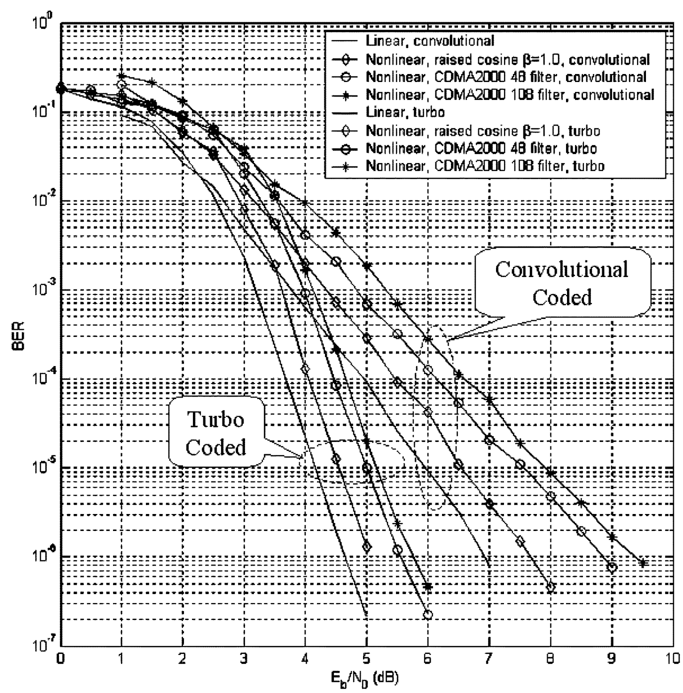

Fig. 5. BER of the communication system using $\pi/4$ QPSK and fully saturated power amplifiers through Rayleigh fading channel.

VI. Conclusion

A baseband equivalent model for communications systems employing fully saturated power amplifiers is provided. The SNR degradation is derived for demodulation performance. The BER is simulated for communication systems employing OQPSK or $\pi/4$ QPSK with convolutional code or turbo code. The simulation results match the derived results.

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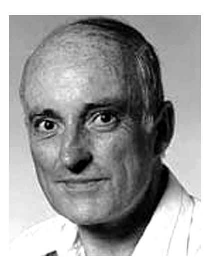

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