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BER Performance of OFDM System with Channel Impairments

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Abstract— **This paper analyzes the Bit Error Rate (BER) performance of Orthogonal Frequency Division Multiplexing (OFDM) system. We derive the BER performance of OFDM system, and quantify the effects of channel impairments such as Doppler Shift due to user mobility and High-Power Amplifier (HPA) distortion when amplifying the transmitted OFDM symbol. It is shown that the resulting inter-carrier interference (ICI) generated by these impairments becomes very significant in OFDM system, and severely degrades its BER performance. Simulation results match well and verify the analysis.**

I. Introduction

Wireless communication systems have been all-pervasive lately, with new technologies deployed in various spheres of life. Worldwide Inter-operability for Microwave Access (WiMAX) is one example of these new systems; with Mobile WiMAX (IEEE 802.16e), and Fixed WiMAX (IEEE 802.16-2004 standard) technologies currently being rolled out Worldwide.

Orthogonal frequency-division multiplexing (OFDM) is one of the key broadband technologies deployed in both fixed and mobile WiMAX systems [1]. Broadband wireless access (BWA) based on OFDM has thus gained considerable attentions lately. OFDM provides an efficient broadband data transmission by sending parallel data over a number of subcarriers. To increase data rate for a given OFDM system, the number of subcarriers per OFDM symbol should be increased. As the number of subcarriers increase however, the frequency spacing between the subcarriers in the OFDM symbol reduces. This makes the OFDM system more sensitive to inter-carrier interference (ICI) [3]- [5].

In OFDM systems, because of the high variations in the level of the instantaneous signals transmitted, peak-to-average power ratio (PAPR) is typically large, making operation over linear region of high power amplifier (HPA) difficult. When some parts of the OFDM signals transmitted operate in the nonlinear region of the HPA, nonlinear distortions are introduced in the OFDM signals, resulting in ICI [2]. In addition, relative speed between the transmitter and the receiver in a wireless system introduces Doppler shift in the received frequencies. This destroys the subcarriers' orthogonality, and also contributes to the ICI [3]. These issues cause significant problems in practical systems and should be given due considerations in the design of broadband communication systems.

In this paper, we examine the bit error rate (BER) performance of OFDM system with different quadrature amplitude modulation (*QAM*) schemes (4-*QAM*, 16-*QAM*, and 64-*QAM*), over Rayleigh fading channels, and quantify the effects of Doppler shift and amplifier non-linearity on the BER performance of the system. The results show that the BER degrades as the doppler shift and the amplifier non-linearity increase.

The paper is organized as follows. Section II presents a summary of the previous works in the literature on frequency shift estimations for OFDM systems, Section III provides the OFDM system model with amplifier nonlinear distortion and Doppler shift effects, used in this work, Section IV presents the derivation of the BER expressions for Rayleigh fading channel, while in Section V, we validate the theoretical analysis by computer simulations. Finally, Section VI presents the conclusion of this paper.

II. Related Work

A number of works have considered the estimation of the frequency shift in OFDM systems [3]- [8]. A maximum likelihood estimator (MLE) is proposed in [3]. The symbol timing is assumed to be known, therefore only the subcarrier frequency offsets are to be estimated. The acquisition range is determined to be ± 0.5 of the subcarrier spacing. This range can be increased using a shorter training symbols to determine the subcarrier frequency offset. Therefore, the estimation gets worse since the number of symbols over which the average estimate is computed becomes smaller.

An estimation scheme using a null symbol is introduced in [4]. Where an empty symbol is transmitted to help the receiver to find the start of the symbol by detecting the power drop, and the subcarrier frequency offset is captured after performing the fast fourier transform (FFT) operation. The disadvantage of this scheme is the extra overhead introduced by the transmitted null symbols.

A trial and error scheme to find both the symbol timing and the subcarrier frequency offset is proposed in [5]. This scheme has two disadvantages. First, the frequency offset is increased very fast over a short period of the acquisition range. Second, the overhead is high due to a long search process until the subcarrier frequency offset is found.

Some modifications to enhance the above two schemes to achieve synchronization for either a continuous stream or

bursty data is introduced in [6]. The computation for capturing the symbol timing is simplified and the acquisition range for subcarrier frequency offset is extended. Instead of using a null symbol, one unique symbol with a repetition within half a symbol period is used to capture the start of the burst. While the symbol timing is found by searching for a symbol which has two identical halves. The subcarrier frequency offset is partially corrected and then found with the second symbol.
The authors in [7], provide an improved scheme for frequency offset estimation. In this scheme the OFDM symbol is divided into P identical parts, where $P > 2$. This scheme achieves better accuracy compared to the scheme proposed in [6]. In addition, it needs only one training symbol instead of two symbols to make the estimation range large enough.
As mentioned above, the subcarriers' orthogonality can be destroyed by Doppler shift as a result of MS's mobility. However, a few algorithms have been proposed for Doppler shift estimation for OFDM systems. In [8], an algorithm was proposed for Doppler shift estimation, but the algorithm can only be applied to signals on the pilot subcarriers. Therefore, to obtain an accurate estimation of Doppler shift, large number of OFDM symbols are required, and this results in a longer estimation delay. In [9], an algorithm is proposed for Doppler shift estimation based on the autocorrelation of time domain channel. However, the receiver still has to know the fading channel coefficient in order to perform the Doppler shift estimation. This results in higher complexity at the receiver. In this work, we employ analytical approach to quantify the effects of channel impairments such as doppler shift and HPA distortion on the BER performance of OFDM systems.

III. SYSTEM MODEL

Consider a broadband wireless network employing OFDM transmission over Rayleigh fading channel, each OFDM symbol consisting of N_c subcarriers. Consider the uplink transmission scenario, where mobile stations (MS) communicate with the base station (BS). We assume that there are L users in the system, uniformly distributed in the coverage area of the BS. An example of this model is illustrated in Fig.1 for a cellular deployment, in which each cell is serviced by a BS, located at the center of the cell.

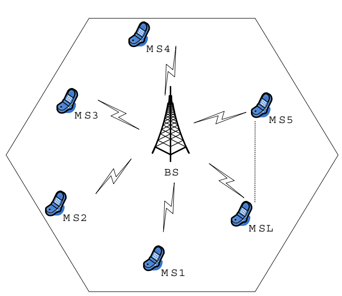

Fig. 1. Mobile WiMAX System

A block diagram of a transceiver (MS and BS) in OFDM system with Inverse Fast Fourier Transform (IFFT) and Fast Fourier Transform (FFT) functions is shown in Fig. 2. This diagram illustrates the different stages each transmitted signal traverse. Each MS's data which in our case generated by *QAM* modulation is modulated with a number of N_c subcarreirs to produce N_c data points in frequency domain. These data points are feed to an IFFT function whose output is a time-domain signal transmitted over Rayleigh fading channel after adding the cyclic prefix (CP) to combat inter symbol interference (ISI) introduced by the multipath channel.

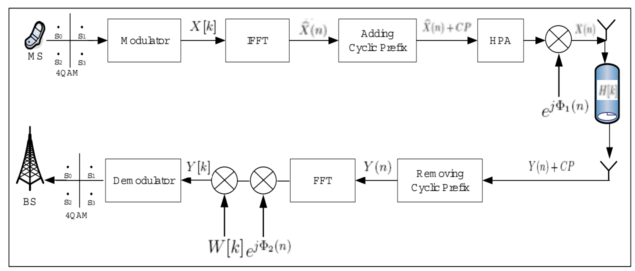

Fig. 2. OFDM transceiver

At the receiver side after the cyclic prefix is removed, the time domain received signal is feed to the FFT function whose output is a frequency-domain signal. Then the demodulation is performed to recover the N_c data points.
Accordingly, consider the analytical representation of an OFDM symbol $\widehat{X}(n)$ generated by performing IFFT operation on QAM data points $X[k]$ modulated onto N_c subcarriers.

$$\begin{aligned} \widehat{X}(n) &= IFFT\{X[k]\} \\ &= \frac{1}{\sqrt{N_c}} \sum_{k=0}^{N_c-1} X[k] e^{j(\frac{2\pi}{N_c})kn} \\ & \quad k, n = 0, 1, \ldots, N_c - 1 \end{aligned} \tag{1}$$

$X[k]$ denotes the *QAM* modulated signal in the k^{th} subcarrier. The generated OFDM symbol $\widehat{X}(n)$ is extended by a cyclic prefix to overcome ISI when transmitted over frequency-selective fading channels $h(n)$ whose FFT is denoted $H[k]$ on subcarrier k, and modelled as flat-fading Rayleigh channel.
The time domain received signal $Y(n)$ can be written as

$$Y(n) = \widehat{X}(n)h(n) + w(n) \tag{2}$$

Where $w(n)$ is the additive white gaussian noise. The transmitted signal $\widehat{X}(n)$ is affected by two deterministic impairments, namely amplifier distortion and Doppler shift. The first impairment can be considered as generated before the fading channel. Therefore, when the transmitted signal is amplified by HPA, it experiences a phase distortion $\Phi_1(n)$.
The well known nonlinear memoryless amplifier [2], is adopted in this paper. Thus the expressions for AM/AM (amplitude-to-amplitude distortions), $A_1(.)$, and AM/PM (amplitude-to-phase distortions), $\Phi_1(.)$, characteristics for a

travelling wave tube amplifier (TWTA) are used

$$A_1(X(n)) = \frac{\nu X(n)}{(1+\eta_a X^2(n))} \quad (3)$$

$$\Phi_1(X(n)) = \frac{\alpha_\phi X^2(n)}{(1+\eta_\phi X^2(n))} \quad (4)$$

where $X(n) = |\widehat{X}(n)|$ and $|.|$ is the magnitude of its argument, ν denotes the small-signal gain, $A_s = 1/\sqrt{\eta_a}$ is the amplifier input saturation voltage, and the HPA maximum output amplitude is $A_0 = max_{X(n)} = \{A[X(n)]\} = \frac{\nu A_s}{2}$. A typical choice for the above parameters is $\nu = 1, \quad \eta_a = 0.25, \quad \eta_\phi = 0.25, \quad and \quad \alpha_\phi = \frac{\pi}{12}$. We assume a receiver with prefect amplitude distortion compensation, consequently, the received signal with HPA phase distortion can be expressed as

$$Y(n) = X(n)h(n)e^{j\Phi_1(n)} + w(n) \quad (5)$$

The transmitted signal is also shifted by Doppler effect $\Phi_2(n)$, which is given by $2\pi f_d T_s$, where f_d denotes the frequency offset between the transmitter and the receiver and can be expressed as

$$f_d = \frac{f_c * V_{S,D}}{c} = \frac{V_{S,D}}{\lambda} \quad (6)$$

where f_c is the transmitted frequency, $V_{S,D}$ is the velocity of the transmitter relative to the receiver (m/s), c is the speed of light $(3\times 10^8(m/s))$, $\lambda = \frac{c}{f_c}$ is the wavelength, and T_s is the OFDM symbol period. Hence, we rewrite the time-domain received signal in (5) after accounting for Doppler effect, as

$$Y(n) = X(n)h(n)e^{j\Phi_1(n)}e^{j\Phi_2(n)} + w(n) \quad (7)$$

After cyclic prefix (CP) removal, and the FFT operation, the frequency domain received signal per subcarrier k, $Y[k]$ can be expressed as

$$Y[k] = \frac{1}{\sqrt{N_c}}\sum_{n=0}^{N_c-1} Y(n)e^{-j(\frac{2\pi}{N_c})kn} + W[k] \quad (8)$$

where $W[k]$ denotes $FFT\{w(n)\}$ on subcarrier k.
Substituting the time domain received signal $Y(n)$ from (6) yields:

$$Y[k] = \frac{1}{\sqrt{N_c}}\sum_{n=0}^{N_c-1} X(n)h(n)e^{-j(\frac{2\pi}{N_c})kn} e^{j\Phi_1(n)}e^{j\Phi_2(n)} + W[k] \quad (9)$$

Substituting the time domain transmitted signal $X(n)$ yields, after some rearrangements:

$$Y[k] = \frac{1}{N_c}\sum_{n=0}^{N_c-1} h(n)\sum_{m=0}^{N_c-1} X[m]e^{j(\frac{2\pi}{N_c})(m-k)n} e^{j\Phi_1(n)}e^{j\Phi_2(n)} + W[k] \quad (10)$$

At this point in the analysis, we write the expression for the received signal as sum of two cases: case $m = k$, and case $m \neq k$. Therefore, (10) can be rewritten as

$$\begin{aligned} Y[k] &= \frac{X[k]}{N_c}\sum_{n=0}^{N_c-1} h(n)e^{j\Phi_1(n)}e^{j\Phi_2(n)} \\ &+ \frac{1}{N_c}\sum_{m=0,m\neq k}^{N_c-1} X[m]\sum_{n=0}^{N_c-1} h(n)e^{j(\frac{2\pi}{N_c})(m-k)n} \\ &\cdot e^{j\Phi_1(n)}e^{j\Phi_2(n)} + W[k] \\ &= \underbrace{X[k]H[k]\Phi_1[0]\Phi_2[0]}_{desired \quad signal \quad \times \quad \Phi[0]} \\ &+ \underbrace{\frac{1}{N_c}\sum_{m=0,m\neq k}^{N_c-1} X[m]\sum_{n=0}^{N_c-1} h(n)e^{j(\frac{2\pi}{N_c})(m-k)n}e^{j\Phi_1(n)}e^{j\Phi_2(n)}}_{\beta[k]} \\ &+ W[k] \end{aligned} \quad (11)$$

From (11) it can be stated that common phase error components $\Phi_1[0]$ and $\Phi_2[0]$ introduce multiplicative distortions on every subcarrier, which cause a constant rotation to the whole OFDM symbol. All subcarriers contribute to the ICI term $\beta[k]$ and is added to the desired signal of each subcarrier. Hence, the frequency domain received signal on subcarrier k can be expressed in compact form as

$$Y[k] = H[k]X[k]\Phi_1[0]\Phi_2[0] + \beta[k] + W[k] \quad (12)$$

where $X[k]$ is the frequency-domain transmitted signal, $H[k]$ is the frequency-domain channel coefficient, $\Phi_1[0]$ and $\Phi_2[0]$ are the common phase error components, $W[k]$ is the additive white Gaussian noise which can be considered as a Gaussian random variable with zero mean and variance σ_W^2, and $\beta[k]$ is the inter-carrier interference (ICI) signal given by

$$\beta[k] = \sum_{m=0,m\neq k}^{N_c-1} X[m]H[m]\Phi_1[m-k]\Phi_2[m-k] \quad (13)$$

IV. BER Performance Analysis

In this section, we derive the BER performance of OFDM systems in the presence of amplifier distortion Φ_1, and Doppler shift Φ_2 impairments. Assume that the constant phase rotation imposed by $\Phi_1[0]$, and $\Phi_2[0]$ can be accurately estimated at the receiver. The frequency-domain received signal on the k^{th} subcarrier obtained in the last section as expressed in (11) can be used to obtain expression for the signal-to-interference-plus-noise-ratio (SINR) for our system as

$$SINR = \frac{E[|X[k]H[k]|^2]}{E[|\beta[k]|^2] + E[|W[k]|^2]} \quad (14)$$

Without loss of generality, we consider the error rate for the zeroth subcarrier ($i.e., k = 0$). We assume that the data symbols $X[k]$ and $X[m]$, $k = 0$, and $m = 1, \ldots, N_c - 1$ are mutually independent random variables (RVs), with average power E_s. $W[k]$ is a complex Gaussian noise with variance σ_W^2, then (14) can be rewritten as

$$SINR = \frac{E_s E[|H[k]|^2]}{\sigma_\beta^2 + \sigma_W^2} \quad (15)$$

Let z be the multiplicative effect imposed by the channel on the received signal of the k^{th} subcarrier given by

$$z = \frac{1}{N_c} \sum_{n=0}^{N_c-1} |h(n)| \quad (16)$$

and $\beta[k]$ is the ICI term caused by the loss of orthogonality between subcarriers. The PDF's of z and $\beta[k]$ are required for BER performance analysis. However, it is not straightforward task to derive the exact PDF's of z and $\beta[k]$. We observe that z can be expressed as [10]

$$\begin{aligned} z &= \frac{1}{N_c}[|h(0)| + \ldots |h(N_c-1)|] \\ &= \frac{1}{N_c}[|h(0)| + (\frac{|h(N_c-1)| - |h(0)|}{N_c - 1}) + |h(0)| \\ &+ \ldots + (\frac{(N_c-1)|h(N_c-1)| - |h(0)|}{N_c-1}) + |h(0)|] \end{aligned}$$

which simplifies to

$$z = \frac{1}{2}[|h(0)| + |h(N_c-1)|] \quad = \frac{1}{2}\Big(\alpha_{(0)} + \alpha_{(N_c-1)}\Big) \quad (17)$$

where $\alpha_{(0)}$ and $\alpha_{(N_c-1)}$ are two random variables representing the multiplicative distortion $|h(0)|$ and $|h(N_c-1)|$, respectively. The joint probability density function of $\alpha_{(0)}$ and $\alpha_{(N_c-1)}$ can be expressed as

$$f(\alpha_{(0)}, \alpha_{(N_c-1)}) = \frac{\alpha_{(0)}\alpha_{(N_c-1)}}{\sigma^4(1-\lambda_1^2)} e^{\frac{-(\alpha_{(0)}^2 + \alpha_{(N_c-1)}^2)}{2\sigma^2(1-\lambda_1^2)}} \quad (18)$$

where $\alpha_{(0)}, \alpha_{(N_c-1)} \geq 0$, and λ_1^2 is the correlation coefficient between $\alpha_{(0)}$ and $\alpha_{(N_c-1)}$, which can be expressed as $J_0(2\pi f_d|n-k|)$, where $J_0(.)$ is the zero order Bessel function of the first kind. For simplicity we replace $\alpha_{(N_c-1)}$ with y, then we write Eq. (18) as: $2z = \alpha_{(0)} + y$, $\alpha_{(0)} = 2z - y$. The pdf of z can then be expressed as

$$f_z(z) = 2\int_0^{2z} f(2z-y, y)dy$$

$$f_z(z) = 2\int_0^{2z} \frac{(2z-y, y)}{\sigma^4(1-\lambda_1^2)} e^{\frac{-((2z-y)^2+y^2)}{2\sigma^2(1-\lambda_1^2)}} dy \quad (19)$$

For large number of N_c, we invoke the central limit theory to assume that the ICI term $\beta[k]$ is approximated as Gaussian Random Variables with zero mean. The desired signal power of the k^{th} subcarrier is given by $E[|zX[k]|^2] = E_s E[z^2]$, where E_s is the energy for the transmitted symbol on subcarrier k, and the expectation $E[z^2]$ can be derived as

$$\begin{aligned} E[z^2] &= E\left[\left(\frac{1}{N_c}\sum_{n=0}^{N_c-1}|h(n)|\right)\left(\frac{1}{N_c}\sum_{k=0}^{N_c-1}|h(k)|\right)\right] \\ &= \frac{1}{N_c^2}\sum_{n=0}^{N_c-1}\sum_{k=0}^{N_c-1}\int_0^\infty\int_0^\infty \alpha_{(n)}\alpha_{(k)} \\ &\quad f(\alpha_{(n)}, \alpha_{(k)})d\alpha_{(n)}d\alpha_{(k)} \end{aligned}$$

Substituting the joint PDF of $\alpha_{(n)}$ and $\alpha_{(k)}$, and after some manipulations, the expectation of $E[z^2]$ can be simplified as

$$\begin{aligned} E[z^2] &= \frac{1}{N_c^2}\Bigg[2N_c + \pi\sum_{i=1}^{N_c-1}(N_c - i) \\ &\quad {}_2F_1\Big(-\frac{1}{2}, -\frac{1}{2}; 1; \lambda_1^2\Big)\Bigg] \end{aligned} \quad (20)$$

where $i = 0, 1, \ldots, N_c - 1$, and $F(.,.;.;.)$ is the hypergeometric function. Therefore, the variance of the ICI of the k^{th} subcarrier can be derived by subtracting the desired signal power from the total power of ICI as

$$\sigma_\beta^2 = E_s\left(E\left[\frac{1}{N_c}\sum_{n=0}^{N_c-1}|h(n)|^2\right] - E[z^2]\right)$$

Substituting $E[z^2]$, and after some re-arrangements the variance of ICI, σ_β^2, can be expressed as

$$\begin{aligned} \sigma_\beta^2 &= E_s\Bigg(1 - \frac{1}{N_c^2}\Bigg[2N_c + \pi\sum_{i=1}^{N_c-1}(N_c - i) \\ &\quad {}_2F_1\Big(-\frac{1}{2}, -\frac{1}{2}; 1; \lambda_1^2\Big)\Bigg]\Bigg) \end{aligned} \quad (21)$$

Using (15), (20), and (21), the SINR can be expressed as

$$SINR = \frac{E_s E[z^2]}{\sigma_\beta^2 + \sigma_{W[k]}^2} = \frac{E[z^2]}{\frac{\sigma_\beta^2}{E_s} + \frac{1}{\gamma}} \quad (22)$$

Where $\gamma = \frac{E_s}{\sigma_W^2}$. The probability of error per subcarrier k in the case of *M-QAM* modulation (M = 4, 16, 64) can be expressed as [11]

$$\begin{aligned} P_e(E,k) &= \int_0^\infty \Bigg\{1 - \Bigg[1 - \frac{2(\sqrt{M}-1)}{\sqrt{M}} \\ &\quad Q\Big(\sqrt{2gSINR}\Big)\Bigg]^2\Bigg\} f_z(z)dz \end{aligned} \quad (23)$$

where $Q(.)$ is the Gaussian tail function and $g = \frac{3}{(2(M-1))}$. Substituting SINR from (22) and the joint PDF from (19) into (23), gives the probability of error per subcarrier k. Using the probability of error per subcarrier above, the total average OFDM symbol error rate (SER_{OFDM}) can be expressed as

$$SER_{OFDM} = \frac{1}{N_c} \sum_{k=0}^{N_c-1} P_e(E,k) \qquad (24)$$

Finally, the bit error rate for the system (assuming uniform bit loading across subcarriers) can be estimated as

$$BER = \frac{SER_{OFDM}}{log_2(M)} \qquad (25)$$

V. Simulation Results

In this section, we present some simulation results to verify the performance of an *M-QAM*-OFDM system with 256 subcarriers. The system BER performance is evaluated in the presence of both the HPA distortions and Doppler shift. In the simulation model, three signal constellations are used 4-*QAM*, 16-*QAM*, and 64-*QAM*. The CP is added to each symbol to overcome ISI effects, the CP can take the values 25%, 12.5%, 6.25%, and 3.125% of the OFDM symbol length. However, in this work only CP = 25% is used. The Doppler shift values used in the simulation are 0Hz, and 100Hz, while the HPA distortion values are $\pi/12$, and $\pi/6$, respectively. The transmission of 20000 OFDM symbols are simulated, then the BER performance is measured using (25).

Fig 3 presents the BER performance of OFDM-256 system, using different *QAM* constellations (4, 16, and 64-*QAM*). For the illustration in this figure, the frequency shift is set to $f_d = 0$, and $\alpha_\phi = \pi/12$ which means the MS is stationary, and the HPA is applying the ideal linear characteristics. A perfect agreement is observed between the simulation and the theoretical results as can be noticed from the BER curves. Therefore the theoretical results are verified by the simulations.

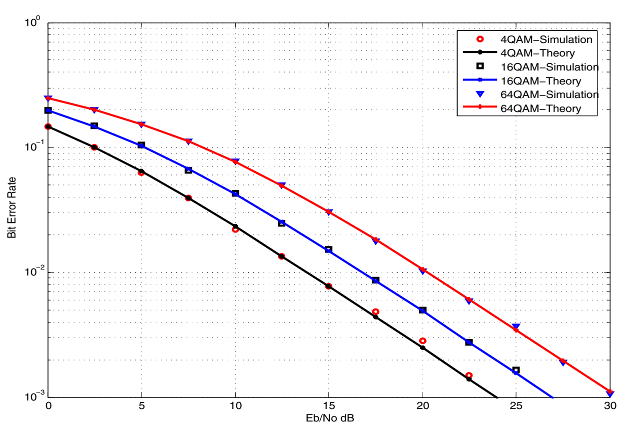

Fig. 3. BER versus E_b/N_0 of OFDM-256 system over Rayleigh Fading channel, $f_d = 0$, and $\alpha_\phi = \pi/12$.

Fig.4 demonstrates the BER performance of OFDM-256 system in the presence of Doppler shift and HPA nonlinearity distortion. For this illustration we used the following values $f_d = 100Hz$, and $\alpha_\phi = \pi/6$. It can be observed that, for 16-*QAM* constellation, the loss in power efficiency at BER = 10^{-3} due to Doppler shift and HPA is approximately 2dB. This loss becomes more severe for 64-*QAM* constellation, and is approximately 3dB, while in the case of 4-*QAM* constellation, better performance is observed. It can be noticed that the error floor starts to appear at higher E_b/N_0 due to ICI component as a result of doppler shift and HPA distortion. Furthermore, as the constellation order *M* increases, the ICI distortion increases. In this case, ICI distortion controls the system performance improvement.

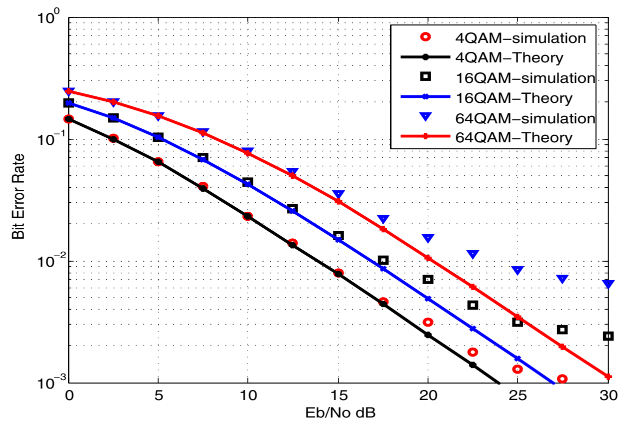

Fig. 4. BER versus E_b/N_0 of OFDM-256 system over Rayleigh Fading channel, $f_d = 100Hz$, and $\alpha_\phi = \pi/6$.

VI. Conclusion

A theoretical analysis of the effect of two impairments on the OFDM system has been presented in this paper. The first impairment is due to the amplifier non-linearity. The HPA shifts the transmitted signal before transmission over the channel, resulting in ICI. The second impairment is due to Doppler shift. The relative speed between the transmitter and the receiver shifts the transmitted frequency at the receiver side causing also ICI. The analysis in this paper presents a quantification of the ICI in OFDM system caused by these two impairments. Analytical BER performance of the system is evaluated and simulation results are presented to verify the analysis. A good match is observed between the analytical and simulation results. Based on some observations from the BER curves, the following conclusions can be drawn. (i) OFDM system is sensitive to ICI and this sensitivity is increased with the constellation order M. (ii) Amplifier distortion is one cause of the ICI, which depends on the amplifier characteristics. (iii) The Doppler shift is another cause of the ICI, which increases with the speed of the MS. (iv) The aforementioned issues should be given due consideration in the design of broadband wireless system.

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