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IMPACT OF JAMMER NOISE ON BLIND PHASE NOISE COMPENSATION IN OFDM SYSTEMS

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Abstract- This paper presents the performance evaluation of a blind algorithm used for compensating the phase noise effects in orthogonal-frequency-division-multiplexing (OFDM) systems, in which the time-average of phase noise within an OFDM symbol block is calculated at the subblock level. It contributes towards the estimation of common phase error term, which is used for the excision of intercarrier interference. The main focus of presented research work is on the effects of phase noise on the performance of underlying OFDM system for the different number of subcarriers in the fixed/same OFDM symbol block period/duration, for the different M-ary PSK modulation techniques, and it emphasizes on the impact of jammer noise, while using zero-forcing (ZF) and minimum-mean-square-error (MMSE) techniques for detection. Simulation results are demonstrated to verify efficacy of the blind-phase-noise-compensation (BPNC) technique under the fading environment.
Keywords − OFDM, Phase Noise Compensation, Blind Algorithms, M-ary PSK, Jammer Noise

1. INTRODUCTION

Due to the increasing demand for high data rates, the orthogonal-frequency-division-multiplexing (OFDM) has become a technique of choice [1]. Recently, OFDM technology has been extensively utilized in different applications, such as digital-video-broadcasting (DVB) and WiMAX standard communication systems. Unfortunately, the availability of imperfect practical oscillators leads to inevitable phase noise. The phase noise is the random phase fluctuation that causes loss of orthogonality [2-5].

The performance of system can be enhanced by suppressing and compensating the phase noise. Various techniques have been proposed to mitigate the effects of phase noise [6-9]. The blind methods for phase noise compensation are known to improve the bandwidth efficiency, as they do not utilize pilot symbols [7,9]. Lee et al., [9] have proposed a blind algorithm for the phase noise compensation, in which single OFDM symbol is segregated into subblocks, to approximate the time-average of phase noise over each subblock. This time-averaged phase noise and N-point discrete-Fourier-transform (DFT) possess an important relation with the OFDM symbol and channel gain [9]. Based on this relation, the estimated phase noise in this single OFDM symbol block duration is used to compensate for the common-phase-error (CPE) along with the intercarrier interference (ICI) mitigation. However under deleterious conditions, the jammer noise may be present to adversely affect the bit-error-rate (BER) performance of the underlying OFDM system, in addition to additive-white-Gaussian-noise (AWGN).

In this paper, we focus on the impact of jammer noise on the performance of blind-phase-noise-compensation (BPNC) technique for OFDM systems (as proposed by Lee et al., in [9]). The concept of noise bucket effect eventually comes into picture while dealing with jammer noise, in which the jammer noise is considered to be spread out over N number of subcarriers at the receiver during DFT operation [10]. This spreading effect characterizes the jammer noise as Gaussian at the output irrespective of the noise distribution at the receiver input [11], and such Gaussian characteristics get boosted when the number of subcarriers or the OFDM symbol block length N is increased within the fixed/same OFDM symbol block period/duration. Therefore, the presented research work includes the performance evaluation of OFDM systems using BPNC technique in the presence of jammer noise (exhibiting different statistical parameters), for the different number of subcarriers and using different M-ary PSK constant modulus modulation techniques under the multipath Rayleigh fading conditions.

2. OFDM SYSTEM MODEL AFFECTED BY JAMMER NOISE

The input to N-point inverse-discrete-Fourier-transform (IDFT) operation is an OFDM symbol block represented as $\vec{\mathbf{X}} = \left[X_0, X_1, ..., X_{N-1}\right]^T$, where $\left[.\right]^T$ represents the matrix transpose operation, N is number of subcarriers in single OFDM symbol, and X_k denotes the constant modulus modulated data transmitted on the kth subcarrier, where $\left|X_k\right|^2 = E_X$ for all k. Due to the normalized inverse-DFT operation at transmitter, the time-domain sample vector obtained is represented as $\vec{\mathbf{x}} = \left[x_0, x_1, ..., x_{N-1}\right]^T$. The cyclic-prefix (CP) used in OFDM symbol block helps in eliminating the effects of intersymbol interference. This CP-OFDM symbol sample vector is transmitted through the multipath fading channel of length L. It is assumed that the channel state information remains constant during an OFDM symbol block period, and it is perfectly

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known at the receiver. It is noteworthy that besides AWGN, the transmitted OFDM signal may be affected by the jammer noise from an intended or unintended source. Hence, the data sample y_i received at any time instant i after propagating through a multipath channel and removal of CP, is written as

$$y_i = \exp(j\Phi_i)\sum_{l=0}^{L-1} h_l x_{\langle i-l\rangle_N} + \{w_i + \eta_i\}, \qquad 0 \le i \le N-1 \tag{1}$$

where, Φ_i denotes the phase noise at time instant i, $\{h_l\}_{l=0}^{L-1}$ represent the channel impulse response coefficients of length L, and $\langle . \rangle_N$ indicates the modulo-N operation. Here, w_i is the AWGN component, which is independent and identically distributed complex random variable with zero-mean and variance σ_w^2, and η_i denotes the jammer noise at time instant i. The nature of jammer noise is assumed to be unknown in the time-domain, implying that it may possess any probability density function, while exhibiting zero-mean and variance σ_η^2.

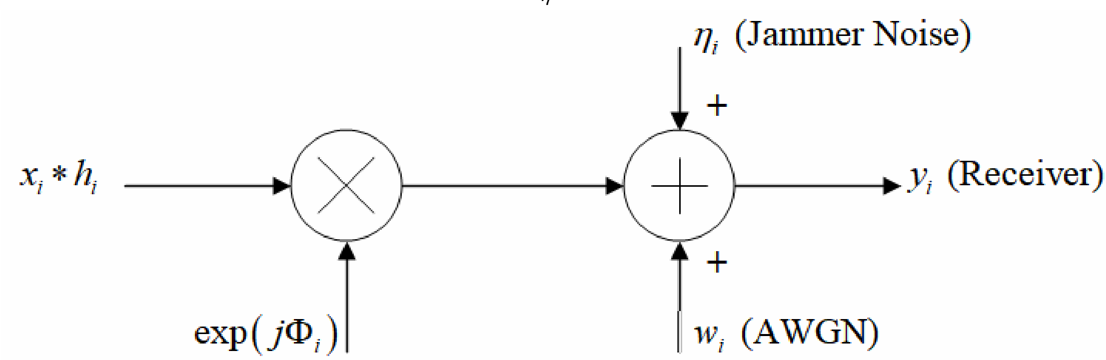

Figure 1. Time-domain OFDM signal reception (CP not shown)

The phase noise Φ_i is modelled as Brownian motion or Wiener process [3],

$$\Phi_i = \Phi_{i-1} + u_i, \qquad 0 \le i \le N-1 \tag{2}$$

where, u_i is the Gaussian distributed random variable with zero-mean and σ_u^2 variance. Φ_{-1} is assumed to be zero considering perfect synchronization at the starting point of the initial OFDM symbol block. Taking N-point DFT of the received signal y_i in Eq. (1), we obtain

$$Y_k = \{H_k X_k J_0\} + \left\{\sum_{n=0, n\neq k}^{N-1} H_n X_n J_{k-n}\right\} + \{W_k + \Lambda_k\}, \qquad 0 \le k \le N-1 \tag{3}$$

where, H_k denotes the channel gain at the kth subcarrier in the frequency-domain, W_k and Λ_k are the kth normalized N-point DFT coefficients of $\{\exp(jw_i)\}_{i=0}^{N-1}$ and $\{\exp(j\eta_i)\}_{i=0}^{N-1}$ respectively. Here, Λ_k is the Gaussian distributed random variable with zero-mean and variance σ_η^2. The Gaussian distribution of Λ_k in the frequency-domain is in close agreement with the noise bucket effect. However, J_0 represents the common phase error [6] and J_k is given as [9],

$$J_k = \frac{1}{N}\sum_{i=0}^{N-1}\exp(j\Phi_i)\exp\left(-j\frac{2\pi ik}{N}\right), \qquad 0 \le k \le N-1 \tag{4}$$

Assuming that a received OFDM symbol block is divided into B number of subblocks in the time-domain and S is the number of samples in one subblock. The time-average of phase noise within pth subblock is expressed as

$$\bar{\Phi}_p = \frac{1}{S}\sum_{i=pS}^{pS+S-1}\Phi_i, \qquad 0 \le p \le B-1 \tag{5}$$

where, $S = N/B$. The phase noise at pth subblock is approximated with its time-average $\bar{\Phi}_p$ as

$$\Phi_{pS+s} \cong \bar{\Phi}_p, \qquad 0 \le s \le S-1 \tag{6}$$

The received signal at the kth subcarrier, which is now in the frequency-domain, is compensated for the effect of phase noise. It follows that

$$\bar{Y}_k = H_k X_k + \{\bar{W}_k + \bar{\Lambda}_k\}, \qquad 0 \le k \le N-1 \tag{7}$$

$$= \sum_{p=0}^{B-1}\exp(-j\bar{\Phi}_p)\exp\left(-j\frac{2\pi pSk}{N}\right)C_{k,p} \tag{8}$$

where, $C_{k,p} = \frac{1}{\sqrt{N}}\sum_{s=0}^{S-1} y_{pS+s}\exp\left(-j\frac{2\pi sk}{N}\right)$ values are computed by applying the N-point DFT operation on the received samples with zero-padding at pth subblock i.e., $\left[y_{pS}, y_{pS+1}, \ldots, y_{pS+S-1}, 0, 0\ldots, 0\right]^T$. The power of AWGN and jammer noise become almost negligible in the high signal-to-noise-ratio (SNR) regions, and hence it can be neglected i.e., if $\{\bar{W}_k + \bar{\Lambda}_k\} \to 0$, then

$$\bar{Y}_k \approx H_k X_k \tag{9}$$

The next step is to mitigate the effects of ICI and compensate for the common-phase-error (CPE), as in [9]. Blind-ICI-mitigator (BIM) estimates the difference $d_p = \bar{\Phi}_p - \bar{\Phi}_0$ i.e., $\hat{d}_p$, where $\bar{\Phi}_0$ is the time-average. In order to invoke BIM, we utilize this estimated difference to mitigate ICI as

$$z_{pS+s} = \exp\left(-j\hat{d}_p\right) y_{pS+s}, \qquad 0 \le s \le S-1 \tag{10}$$

The residual phase noise (represented by superscript R) is required to obtain the estimate of CPE and residual ICI, and it is therefore calculated as

$$\Phi^R_{pS+s} = \Phi_{pS+s} - \hat{d}_p, \qquad 0 \le s \le S-1 \text{ and } 0 \le p \le B-1 \tag{11}$$

where, $\hat{d}_0 = 0$.

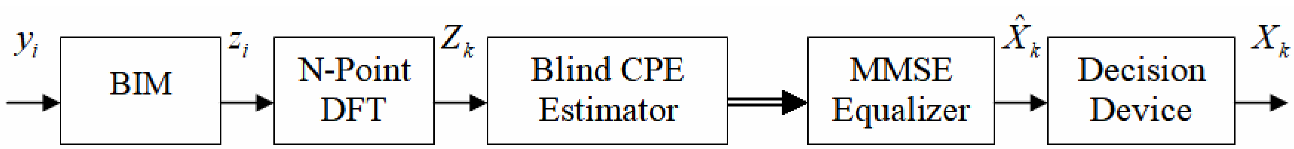

Figure 2. Frequency-domain OFDM symbol detection using blind CPE estimation

3. SYMBOL DETECTION IN THE PRESENCE OF JAMMER NOISE

When the deleterious effects of ICI are tried to be eliminated from the received signal sample, a residual phase noise still exists. Therefore, it is necessary to estimate and rectify this residual CPE present in the following

$$Z_k = J_0^R H_k X_k + \bar{D}_k + \overline{W\Lambda_k}, \qquad 0 \le k \le N-1 \tag{12}$$

where, $J_0^R = \frac{1}{N}\sum_{i=0}^{N-1}\exp\left(j\Phi_i^R\right) \cong \exp\left(j\bar{\Phi}_0^R\right)$ is the residual CPE, $\bar{D}_k$ is the ICI due to residual phase noise and $\overline{W\Lambda_k}$ are the normalized N-point DFT coefficients of $\left\{\exp(j(\Phi_i^R - \Phi_i))(w_i + \eta_i)\right\}_{i=0}^{N-1}$, which are approximated as Gaussian distributed random variables with zero-mean and variance $\sigma_w^2 + \sigma_\eta^2$. Based on the observations, the residual CPE J_0^R can be estimated as $\hat{J}_0^R$ by using the phase angle of Z_k w.r.t. H_k (as in [9]) by calculating the phase noise $\hat{\bar{\Phi}}_0^R$. Subsequently, once the residual CPE is estimated, the estimate of the symbol X_k at the kth subcarrier can be obtained by using Eq. (12) as

$$\hat{X}_k = \frac{\left(H_k \hat{J}_0^R\right)^*}{\left|H_k \hat{J}_0^R\right|^2 + \left(\sigma_w^2 + \sigma_\eta^2\right)} Z_k, \qquad 0 \le k \le N-1 \tag{13}$$

where, $\left(.\right)^*$ represents the complex conjugate operation, and $\hat{J}_0^R = \exp\left(j\hat{\bar{\Phi}}_0^R\right)$ is the estimated residual CPE.

4. SIMULATION RESULTS FOR BPNC BASED OFDM SYSTEM

We shall investigate the BER performance of blind phase noise compensation method [9], which is used in OFDM communication system under the multipath Rayleigh fading environment in the presence of jammer noise. The multipath frequency-selective channel is modelled as a tapped-delay-line filter with 15-tap coefficients, which are assumed to be static during an OFDM symbol block period. The frequency-domain Rayleigh fading channel coefficients H_k for $k = 0,\ 1, ..., \text{N-1}$ are considered to be perfectly known at the receiver. The length of cyclic prefix, which is used to combat intersymbol interference is fixed at 16. However, the number of subblocks is set at B = 4 for the appropriate compensation of phase noise at the subblock level, which leads to suppression of intercarrier interference. It is noteworthy that signal-to-noise-ratio is defined as the ratio of signal power to AWGN power σ_w^2. The phase noise variations strictly follow the Brownian motion as per a random-walk model. The Monte-Carlo simulation results are obtained by utilizing the ensemble average of 500 independent trials, while considering different fading channel configurations. Based on this, the impact of different number of subcarriers in the fixed/same OFDM symbol block period/duration, the different M-ary PSK constant modulus modulation techniques and the level of jammer noise power (with parameter σ_u in Eq. (2)) on the performance of aforementioned BPNC technique in OFDM system is explored further, in terms of bit-error-rate. The average BER is calculated as

$$BER_{AVG} = \left.\frac{\sum_{j=1}^{500} BER(j)}{500}\right|_{SNR} \tag{14}$$

4.1 BER performance without jammer noise

In this case, the phase noise is varied by changing the parameter σ_u from 0 to 0.9°. The number of subcarriers is considered to be $\text{N} \rightarrow \left\{128,\ 256,\ 512,\ 1024,\ 2048\right\}$. The digital signal constellation used for symbol mapping is 8-PSK at +30dB SNR. It may be inferred from Fig. 3 that the BER increases with the increasing value of σ_u for all the values of N. But the value of BER decreases, as the number of subcarriers increases under similar fading conditions in case of MMSE criterion based data symbol detection. For $\sigma_u = 0.75^\circ$, when N changes from 1024 to 2048, the BER gets reduced from 0.0056 to 0.0028 (approximate values). It is due to the alleviation of Gaussian noise power per subcarrier at the receiver output, as the number of subcarriers is increased within the fixed/same OFDM symbol block period/duration.

Further, we reduce SNR to +22dB at N = 512 for analysing the impact of different constant modulus modulation techniques, while comparing the performance of ZF and MMSE detectors. For $\sigma_u = 0.75^\circ$ and 8-PSK constellation, the BER is observed to be approximately 0.04 in case of MMSE detector and 0.05 in case of ZF detector. As M increases in M-ary PSK from 4 to 16, the Euclidean distance between adjacent constellation points decreases, which leads to elevation in BER value i.e., success rate deteriorates (as shown in Fig. 4).

Therefore, it may be observed from the results presented in Fig. 3 and Fig. 4 that the BER performance of underlying OFDM system gets degraded with the increasing value of M for the fixed value of transmitter power and AWGN power. Moreover, the MMSE detector performs marginally better than ZF detector under similar scenarios.

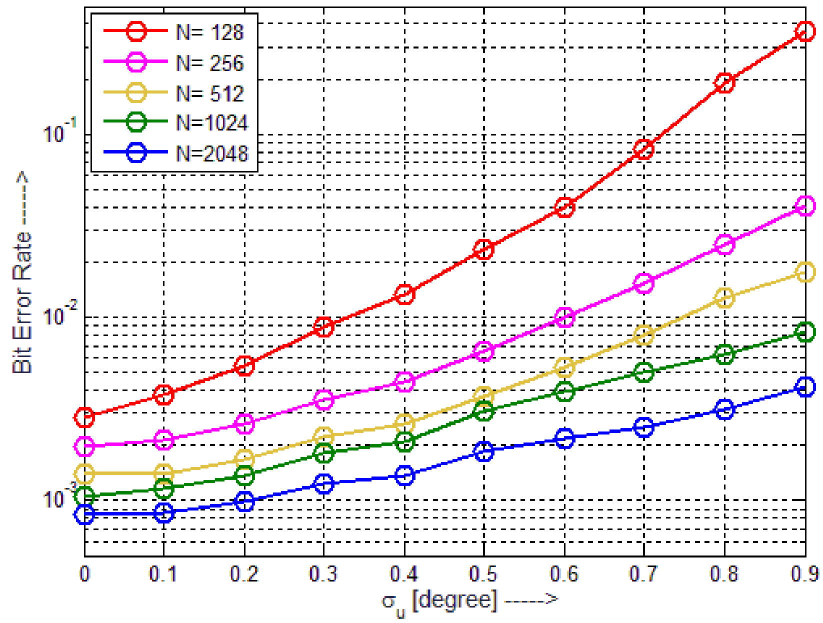

Figure 3. BER vs. σ_u [degree] at SNR=+30dB for different number of subcarriers

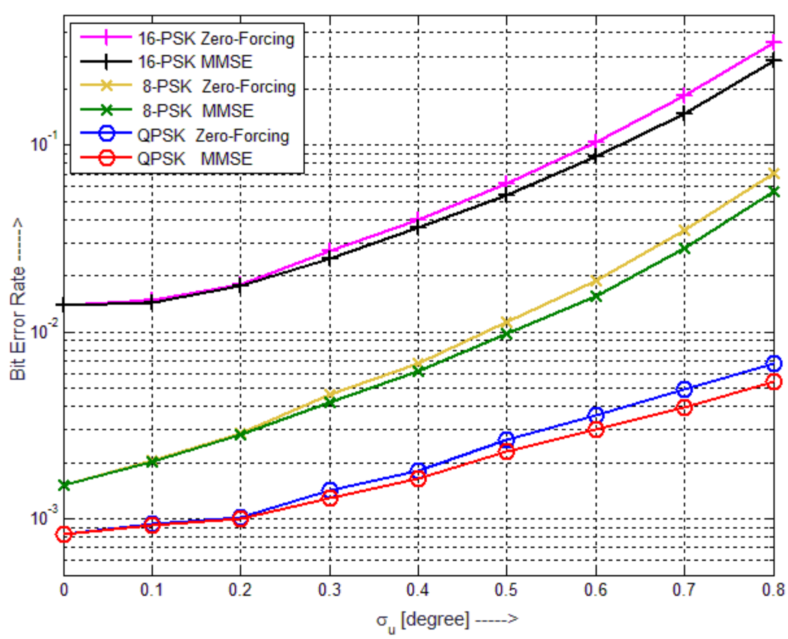

Figure 4. BER vs. σ_u [degree] at SNR=+22dB for M-ary PSK digital modulation technique

4.2 BER performance with jammer noise

In this case, we consider the presence of jammer noise with zero-mean and variance $10\log_{10}\sigma_\eta^2$ dB in the time-domain. The value of parameter σ_u is set at 0.25° for the generation of phase noise. We incorporate 8-PSK digital modulation

technique, while employing MMSE criterion based detector. In the presence of jammer noise with variance -30dB, the BER value is observed to be approximately 0.01 for N = 256 and 0.0066 for N = 512 (as shown in Fig. 5). However, the BER is found to be approximately 0.051 for N = 1024 and 0.0272 for N = 2048, in case the jammer noise variance is -15dB. Moreover, the BER performance deteriorates with the increasing value of jammer noise variance and with the decreasing number of subcarriers for the constant value of transmitted power and total noise (jammer noise plus AWGN) power.
Further, the SNR value is reduced to +22dB for the different types of M-ary PSK digital modulation techniques, at N = 512 and $\sigma_u = 0.25^\circ$. It is apparent from the simulation results presented in Fig. 6 that the BER performance of BPNC technique based OFDM system gets degraded with the increasing value of M from 4 to 16, but the MMSE detector performs marginally better than ZF detector in the presence of jammer noise.
Although, the jammer noise may possess any probability distribution function in the time-domain, yet the DFT operation at the receiver causes the jammer noise power to spread over all the subcarriers leading to Gaussian distribution in the frequency-domain. This exclusive feature helps BPNC technique based OFDM system to combat jammer noise, while communicating over the multipath fading channels.

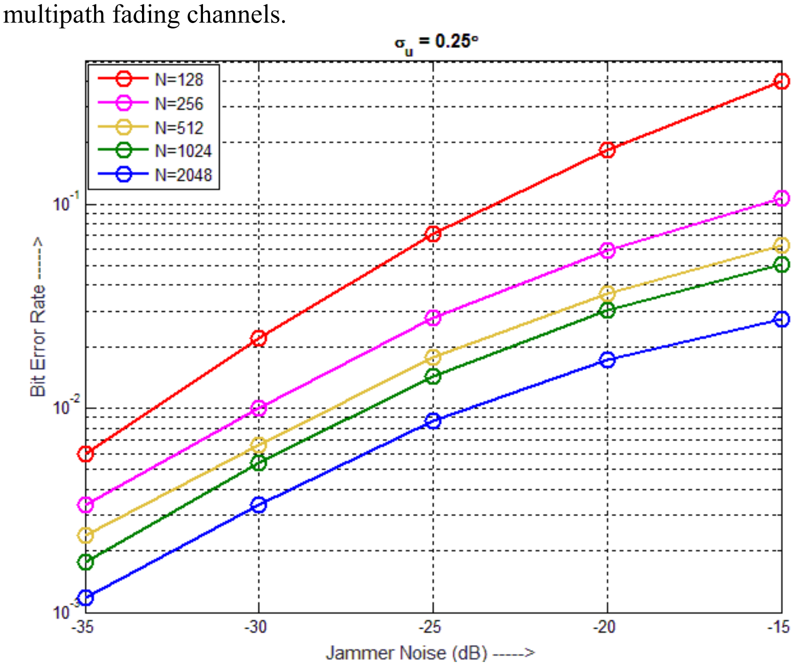

Figure 5. BER vs. jammer noise [dB] at SNR=+30dB for different number of subcarriers

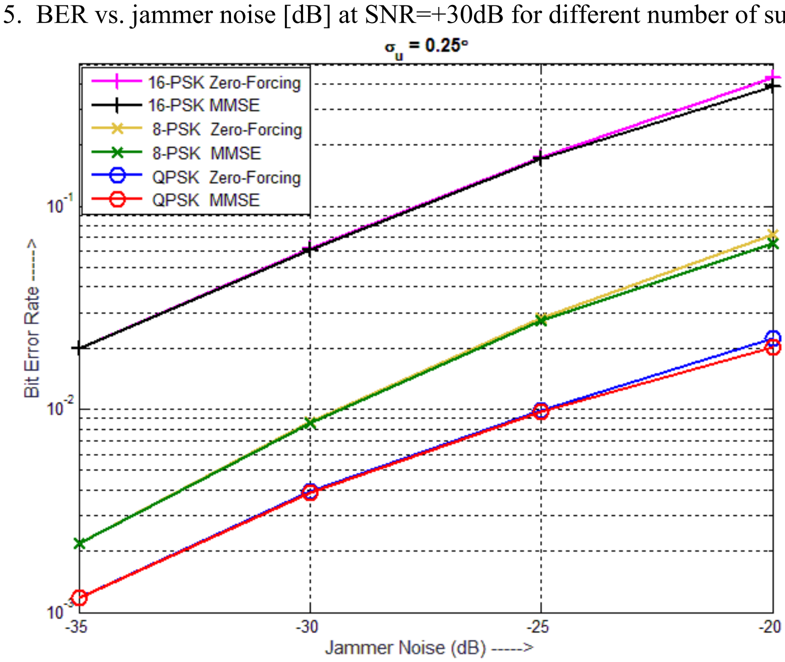

Figure 6. BER vs. jammer noise [dB] at SNR=+22dB for M-ary PSK digital modulation technique

5. CONCLUDING REMARKS

In this correspondence, the BER performance of blind phase noise compensation method based OFDM system is investigated further for the different number of subcarriers or length of OFDM symbol block, for the different M-ary PSK digital modulation techniques and also for the different levels of jammer noise power. Under the multipath frequency-selective Rayleigh fading environment, the phase noise is estimated at the subblock level, which is further utilized to compensate the adverse effects of intercarrier interference by using the blindly estimated common phase error. The spreading of jammer noise with any probability density function over all subcarriers takes place due to the noise bucket effect, which reduces the impact of jammer noise power in the frequency-domain due to the DFT operation at the receiving end. It is evident from simulation results that the long OFDM symbol block length or the large number of subcarriers is beneficial in tackling the jammer noise plus AWGN (total noise), at the constant level of transmitted signal power and total noise, in the fixed/same OFDM symbol block period/duration. As the number of constellation points increases in the higher-order M-ary PSK modulation techniques, the BER performance gets deteriorated and the underlying OFDM system appears to be vulnerable to the adverse effects of jammer noise.

As the backbone of BPNC technique is the availability of perfect channel state information at the receiver, therefore future work includes the analysis of the impact of imperfectly estimated channel state information [13,14,15] on the BER performance of BPNC technique based OFDM systems.

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