Applied Mathematics & Information Sciences
An International Journal

http://dx.doi.org/10.12785/amis/090333

Performance Analysis of $\pi/4$ DQPSK for FrFT-OFDM System with Carrier Frequency Offset

Ankush Kansal[1,*], *Kulbir Singh*[1] *and Rajiv Saxena*[2]

[1] ECED, Thapar University, Patiala (PB), India
[2] Jaypee University, Anoopshahr (UP), India

Received: 3 Aug. 2014, Revised: 4 Nov. 2014, Accepted: 5 Nov. 2014
Published online: 1 May 2015

Abstract: The Orthogonal Frequency Division Multiplexing (OFDM) has widely been used in broadband wireless communication due to its ability to efficiently utilize the spectrum with multicarrier communication. In this paper BER expression for $\pi/4$ DQPSK modulation has been derived from FrFT based OFDM system in the presence of carrier frequency offset (CFO) under frequency selective Rayleigh fading channel. The performance of proposed system has been evaluated and compared with the FFT based OFDM system. It is found that the FrFT based OFDM system outperforms the FFT based OFDM. BER analysis has also been done for various values of α (FrFT angle parameter) on said FrFT based OFDM system, however the system responds best at α = 0.9 as compared with other values of α. Improvement as high as up to 23 dB in SNR has been achieved with the proposed FrFT based OFDM system at α = 0.9 for said modulation with carrier frequency offset of 0.1 under frequency selective Rayleigh fading channel.

Keywords: $\pi/4$ DQPSK, OFDM, FrFT, FFT, CFO

1 Introduction

The current multimedia applications on mobile requires wireless communication systems which can provide high data rates with low delay and low bit error rate (BER). The response of wireless communication is hence determined by the channel environment [1].The higher data rate transmission and due to mobility between source and destination the channel introduces time selective and frequency selective fading [2]. Fractional Fourier Transform (FrFT) had been proposed to analysis this time frequency selective fading in channel with multicarrier communication. Orthogonal Frequency Division Multiplexing (OFDM) is widely used for implementing wireless applications. The major benefit of this parallel bit stream is that each sub-channel has much lower bit rate and is modulated on different carrier. The sub carriers used in OFDM are equally spaced and overlapped; this overlapping provided efficient utilization of bandwidth. The subcarriers are made orthogonal to each other in frequency domain for efficient demodulation at destination [3]. The frequency dispersion of channel introduces the carrier frequency offset at the receiver. This carrier mismatch between transmitter and receiver gives rise to inter carrier interference (ICI) [4]. This ICI will destroy the orthogonality of the data and degrade the system performance by degrading BER [5]. The system performance can be evaluated either by Signal to Noise interference (SIR) or by BER under various wireless fading channels. The Gaussian approximation of ICI for calculating BER under AWGN channel is given in [6]. Their approximation was true for low BER and can be used only for small signal to noise ratio. BER for moments of ICI distribution was proposed by Zhao et al. [7]. The exact symbol to error rate (SER) for characteristic function of ICI had been derived in [8]. All of them had focused the effect of ICI in AWGN channel only, but practically OFDM systems have to cope with various wireless fading channels like Rayleigh, Rician and Nakagami. Since ICI cannot be neglected in practice, the impact of multipath fading should be taken care off. The cyclic prefix has been used to eliminate ISI entirely and therefore only ICI needs to be concerned. Because there are many time-delayed versions of received signals with different gains and different phase offsets, the ICI is more complicated to calculate in various fading channels. Signal processing methods were applied to solve this problem of ICI in OFDM systems. Liu et al. [9] proposed

* Corresponding author e-mail: akansal@thapar.edu

MUSIC-based and ESPRIT-based algorithms to estimate CFO. Biao Chen [10] proposed Maximum Likelihood for CFO estimation, which was proved to be equivalent to Music-based algorithm. Training sequences for CFO estimation was proposed by Jian Li et al. [11]. Different kinds of windowing schemes were proposed by J.Armstrong [12] to reduce the ICI in OFDM systems due to CFO, such as the Hanning window, the Nyquist window, and the Kaiser window. Muller-Weinfurtner [13] proposed MMSE Nyquist to mitigate the ICI effect. MLSE-type signal detection in ICI with block wise whitening of the residual ICI plus noise was proposed by Hai-wei Wang et al. [2] to mitigate the effect of ICI in OFDM systems. Kun Yi Lin et al. [14] proposed time domain ICI self-cancellation method from the view of equivalent channel time variation mitigation in said system. ICI caused due to phase noise was dealt with an efficient algorithm based on improved tail interpolation and CPE processing in [15]. All of these methods had not focused on the cause of ICI due to changing subcarrier bases in radio channels due to variation in frequency with time. Fractional Fourier transformed (FrFT) based OFDM system has been proposed for reduction of ICI effect, while Chen et al. [16] focused on searching of single optimal FrFT angle. H. Wang et al. [17] focused on reduction of ICI effect with the help of FrFT-OFDM system. In this paper BER of FrFT based OFDM system has been calculated in the presence of ICI. In this paper ICI has been shown as the probability density function (PDF) of a mixture of complex Gaussian random variables. The BER of FrFT based system has been calculated for $\pi/4$ DQPSK modulation under frequency selective fading channels. The results achieved have been compared with FFT based OFDM system under similar conditions.

The present paper is organized as follows. After this introduction, Section 2 reviews the Fractional Fourier Transform (FrFT). In Section 3 the system model of proposed FrFT based OFDM system with its transmitter and receiver have been explained. Furthermore in Section 4 mathematical analysis of $\pi/4$ DQPSK modulation for FrFT based OFDM system has been presented. Section 5 is devoted to BER analysis of proposed model. Finally Section 6 explain and compares the simulation and theoretical results achieved from proposed model.

2 FrFT: Fractional Fourier Transformation

The definition of Fractional Fourier Transform (FrFT) stated it as the chirp basis expansion which is defining rotation in time frequency plane i.e. unified time frequency transformation. FrFT is a powerful tool for analyzing chirp signals i.e. signals which has varying frequency with time. The FrFT is defined for entire time-frequency plane. It has been observed that FrFT is a generalized case of FT. The angle α controls the rotation of the signal to be transformed in time-frequency plan from time-axis [18]. FrFT is defined as

$$F_\alpha\{x(t)\}(u) = X_\alpha(u) = \int_{-\infty}^{\infty} x(t) K_\alpha(t,u)\, dt$$

where $K_\alpha(t,u) = A_\alpha e^{j\pi\left(t^2+u^2\right)cot\alpha - j2\pi tucsc\alpha}$ and is called Transformation Kernel and α is the rotation angle of the transformed signal [19].The kernel of FrFT is converted into identity operator when α becomes zero.At $\alpha = \pi/2$ it becomes Fourier Transform, at $\alpha = \pi$ it becomes reflection operator and it becomes inverse Fourier operator at $\alpha = 3\pi/2$.

3 System Model

The transmitter of FrFT based OFDM system has been shown in Figure 1. It is similar to traditional OFDM system the only difference is that here N-point IFrFT has been used to modulate the subcarriers rather than N-point IFFT.

FrFT based OFDM system is a block modulation scheme where a block of N information symbols is transmitted in parallel on N subcarriers. The time duration of an OFDM symbol is N times larger than that of a single-carrier system. An OFDM modulator can be implemented as an IFrFT on a block of N information symbols. To mitigate the effects of ISI caused by channel time spread, each block of IFrFT coefficients is typically preceded by a cyclic prefix (CP) or a guard interval consisting of N samples, such that the length of the CP is at least equal to the channel length. Under this condition, a linear convolution of the transmitted sequence in the channel is converted to a circular convolution. As a result, the effects of the ISI are easily and completely eliminated. But ICI will still remain there due to frequency dispersion in wireless fading channels, which will degrade the system performance. In this proposed model, the incoming serial data had gone through the channel coding at the rate one half for each bit. Then this coded data had been passed through an inter-leaver for re-sequencing of data bits. This re-sequencing would help in distributing the burst error in to several symbols and ensures the error free reception of bits at receiver with minimum number of bits required for forward error correcting codes. After interleaving these symbols were modulated with $\pi/4$ DQPSK modulation. These serial modulated bits were then converted from serial to parallel and were grouped into x bits each to form a complex number. Next, the IFrFT was used to modulate the complex signal. After this, the parallel bits were converted to a series in time using a so called time-division multiplexer and finally the cyclic prefixes were added. By using the digital to analog converter, the discrete symbols were converted to the actual modulating analog signal. After going through the low-pass filter and Radio Frequency (RF) transmitter, finally the OFDM signal had been transmitted. At receiver side; after the RF receiver and low pass filter, the

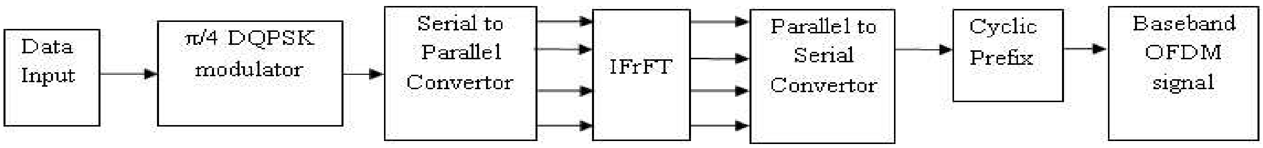

Fig. 1: Transmitter

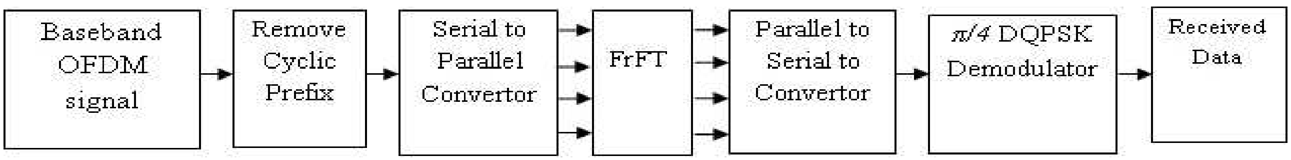

Fig. 2: Receiver

analog to digital converter had converted the analog signal to digital symbols as shown in Figure 2. The cyclic prefix was removed and the symbols were converted from serial to parallel. After this, the FrFT was used to process the symbols and symbols were again serialised. These symbols were then demodulated with $\pi/4$ DQPSK demodulator to obtain originally transmitted data.

4 Mathematical Analysis

The OFDM system considered here for performance analysis is of $\pi/4$ DQPSK modulation. After applying IDFrFT to the sequence $\{X_0, X_1, X_2, , X_(N-1)\}$ which is obtained after encoding with any modulation technique, the expression of the n^{th} transmitted sample is given as

$$x(n) = \sqrt{\frac{N}{N+N_{CP}}} \sum_{k=0}^{N-1} X_{i,n} F_{-\alpha}(n,k)) \tag{1}$$

where $-N_{CP} \leq n < N$
and N_{CP} is the length of the cyclic prefix.
N = Length of FrFT
$X_{i,n}$ = n^{th} symbol transmitted on i^{th} frame

The kernel of IDFrFT, $F_{-\alpha}(n,k)$ is defined as

$$F_{-\alpha}(n,k) = \sqrt{\frac{\sin\alpha + j\cos\alpha}{N}} e^{-\frac{jn^2 T_s{}^2 \cot\alpha}{2}} e^{-\frac{jk^2 u^2 \cot\alpha}{2}} e^{\frac{j2\pi nk}{N}}$$

In case of $\pi/4$ DQPSK modulation

$$X_{i,n} = X_{i,n-1} e^{j\Delta\theta_n[i]} = \sqrt{2E_b} e^{j(\theta_{n-1}[i] + \Delta\theta_n[i])} \tag{2}$$

where, $\theta_{n-1}[i]$ is the phase of the symbol modulated on $(n-1)^{th}$ subcarrier of i^{th} OFDM symbol. $\Delta\theta_n[i]$ is the differential phase carrying information bits. Further Rayleigh fading channel response is given by

$$h_{i,j}(n,l) = h(_j(N+N_g) + n, l + (j-1)(N-N_g))$$

$$E[h(p,l) .* h(p,l)] = \sigma_l^2 J_0(2\pi\Delta t f_d)\,\delta(l-k)$$

where, $h_{i,j}(n,l)$ is n^{th} time channel response of l^{th} tap. Due to the presence of carrier offset at receiver n^{th} received symbol of the j^{th} OFDM frame can be expressed as $y_j(n) = \sum_{i=-\infty}^{\infty} \sum_{m=0}^{N-1} h_{i,j} x_i(m) e^{\frac{j2\pi\varepsilon m}{N}} + w(m)$ $m = 0, 1, \ldots, N-1$ where, $\varepsilon = \Delta f T_u = \Delta f N T_s$ is the normalized CFO and $h(p,l)$ is the channel impulse response.
$T_s = T_u + T_g$ is the duration of OFDM symbol.

After the removal of cyclic prefix and taking DFrFT, the data signal in the k^{th} OFDM symbol can be expressed as

$$r_j(k) = \sum_{q=0}^{\mathrm{N}-1} F_{\alpha}(k,q)\, \mathrm{y_j}(q) \tag{3}$$

$q = 0, 1, \ldots, N-1$ where

$$F_{\alpha}(l,m) = \sqrt{\frac{\sin\alpha - j\cos\alpha}{N}} e^{\frac{jm^2 T_s{}^2 \cot\alpha}{2}} e^{\frac{jl^2 u^2 \cot\alpha}{2}} e^{-\frac{j2\pi ml}{N}} \tag{4}$$

On substituting (4) in (3), the signal at the receiver becomes:

$$r_j(k) = \sqrt{\frac{N}{N+N_{CP}}} \sum_{i=-\infty}^{\infty} IS_{i,j}(q,q) X_{p,q} \tag{5}$$

$$+\sum_{i=-\infty}^{\infty}\sum_{\substack{q=0\\q\neq k}}^{N-1} I.S_{i,j}(k,q)X_{i,q}+w(l)$$

Here $I.S_{i,j}(k,q)$ is the ICI coefficient and is given by

$$I.S_{i,j}(k,q)=\sum_{n=0}^{N-1}\sum_{m=-N_{cp}}^{N-1}F_{\alpha}(k,n)F_{-\alpha}(m,q) \quad (6)$$

$$.h_{i,j}(n,n-m)e^{\frac{j2\pi\varepsilon n}{N}}$$

k^{th}symbol on p^{th} frame from equation no. 5 can be rewritten as

$$r_p\left(\mathrm{k}\right)=\sqrt{\frac{N}{\mathrm{N}+N_{CP}}}\sum_{i=-\infty}^{\infty}IS_{i,j}(q,q)X_{p,q} \quad (7)$$

$$+\sum_{i=-\infty}^{\infty}\sum_{\substack{q=0\\q\neq k}}^{N-1} IS_{i,j}(k,q)X_{i,q}+w(l)$$

Defining $X[k]=[X_{i,0},\ X_{i,1},\ \ldots\ldots\ldots\ldots\ldots X_{i,N-1}]$ and
$C_p[k]=\left[IS_{ij}*(k,0),\ IS_{ij}*(k,1)\ldots\ldots S_{ij}(k,\ N-1)\right]$

$r_p\left(\mathrm{k}\right)$ can be calculated and is given by

$$r_p\left(\mathrm{k}\right)=\sqrt{\frac{\mathrm{N}}{\mathrm{N}+\ \mathrm{N_{CP}}}}X[\mathrm{k}]C_p[k]^H+\mathrm{w}(k) \quad (8)$$

5 BER analysis for $\pi/4$ DQPSK modulation

Here the BER evaluation of said system has been done for $\pi/4$ DQPSK modulation. Let $r_p\left(\mathrm{k}\right)$ is the received signal for p^{th} subcarrier on k^{th} frame. Then the symbols are decoded in phase difference of two successive subcarriers for same OFDM symbol. With the employment of differential detection at the receiver

$$X_{k,p}=\sum_{\mathrm{d}=1}^{\mathrm{D}}\mathrm{r_p}\left(\mathrm{k}\right)*\mathrm{r}^*_{\mathrm{p}-1}\left(\mathrm{k}\right) \quad (9)$$

Real and imaginary parts are given by

$$X^I_{k,p}=\sum_{\mathrm{d}=1}^{\mathrm{D}}\mathrm{R}\{\mathrm{r_p}\left(\mathrm{k}\right)*\mathrm{r}^*_{\mathrm{p}-1}\left(\mathrm{k}\right)\} \quad (10)$$

$$X^Q_{k,p}=\sum_{\mathrm{d}=1}^{\mathrm{D}}\mathrm{I}\{\mathrm{r_p}\left(\mathrm{k}\right)*\mathrm{r}^*_{\mathrm{p}-1}\left(\mathrm{k}\right)\} \quad (11)$$

r_p is the sum of several jointly complex Gaussian random variables. It is therefore, a complex random variable with mean zero and variance is given by

$$\sigma_1^2=\frac{1}{\mathrm{N}^2}\mathrm{X_l}\wedge_{\mathrm{p1,p2}}\mathrm{X_l^H}+\ 2\sigma_1^2 \quad (12)$$

where $\wedge_{\mathrm{p1,p2}}=\mathrm{E}\left[\mathrm{C^H_{p1}C_{p2}}\right]$ and $\wedge_{\mathrm{n1,n2}}=\mathrm{E}\left[\mathrm{IS_{n1,p1}IS^*_{n2,p2}}\right]$

$$\sigma_1^2=\sum_{\mathrm{k}=0}^{\mathrm{N}-1}\sum_{\mathrm{p}=0}^{\mathrm{N}-1}(\mathrm{N}-\mathrm{P})\cos(\cot\alpha\left(\mathrm{T^*_s}\right)\mathrm{p}+\frac{2\pi\,(\mathrm{k}+-\mathrm{k})\,\mathrm{p}}{\mathrm{N}} \quad (13)$$

$$f(x)=\sum_{l=1}^{k}f(x|X_l)Prob(X_l)$$

$$\sum_{l=1}^{K}\frac{1}{\pi\sigma_l^2}e^{\frac{-xHx}{\sigma_l^2}}Prob(X_l) \quad (14)$$

$$E[\mathrm{r_p}\left(\mathrm{k}\right)\mathrm{r_{p-1}}\left(\mathrm{k}\right)|\Delta\theta_1,\Delta\theta_2\ldots\ldots\ldots\Delta\theta_{\mathrm{N}-1}]=0 \quad (15)$$

writing $R\left\{\mathrm{r_p}\left(\mathrm{k}\right)\mathrm{r_{p-1}}\left(\mathrm{k}\right)\right\}$ and $I\left\{\mathrm{r_p}\left(\mathrm{k}\right)\mathrm{r_{p-1}}\left(\mathrm{k}\right)\right\}$ in hermitian quadratic form

$$R\left\{\mathrm{r_p}\left(\mathrm{k}\right)\mathrm{r_{p-1}}\left(\mathrm{k}\right)\right\}=\mathrm{R^HQ_rR} \quad (16)$$

$$I\left\{\mathrm{r_p}\left(\mathrm{k}\right)\mathrm{r_{p-1}}\left(\mathrm{k}\right)\right\}=\mathrm{R^HQ_iR} \quad (17)$$

where the complex Gaussian random vector R is defined as

$$R=\begin{bmatrix}\mathrm{r_p}\left(\mathrm{k}\right)\\ \mathrm{r_{p-1}}\left(\mathrm{k}\right)\end{bmatrix}$$

and Hermitian matrices $Q_r$and Q_i are

$$Q_r=\begin{bmatrix}0 & \frac{1}{2}\\ \frac{1}{2} & 0\end{bmatrix} \text{ and } Q_i=\begin{bmatrix}0 & \frac{-1}{2j}\\ \frac{1}{2j} & 0\end{bmatrix}$$

CHF of $R\left\{\mathrm{r_p}\left(\mathrm{k}\right)\mathrm{r_{p-1}}\left(\mathrm{k}\right)\right\}$ conditioned on

$$g=[\Delta\theta_1,\Delta\theta_2\ldots\ldots\ldots\Delta\theta_{\mathrm{N}-1}]$$

and $\Delta\theta_p$ is given by

$$\theta^{\mathrm{I}}(\omega|\Delta\theta_{\mathrm{P}})=|\mathrm{I_2}-\mathrm{j}\omega\mathrm{MQ_r}|^{-1} \quad (18)$$

where I_2 identity matrix and M is is the covariance matrix defined as

$$M=E\left[RR^H|g,\Delta\theta_P\right]=\begin{bmatrix}\varnothing & \alpha+j\beta\\ \alpha-j\beta & \rho\end{bmatrix} \quad (19)$$

where,ϕ,ρ,α and β are parameters conditioned on differential phase sequence g and $\Delta\theta_p$ and are given by,

$$\varnothing=E[r_P(\hat{k})r_P(\hat{k})g,\ \Delta\theta_P]=\left(\frac{N}{N+N_g}\right)X\wedge_{P,P}X^H+2\sigma^2 \quad (20)$$

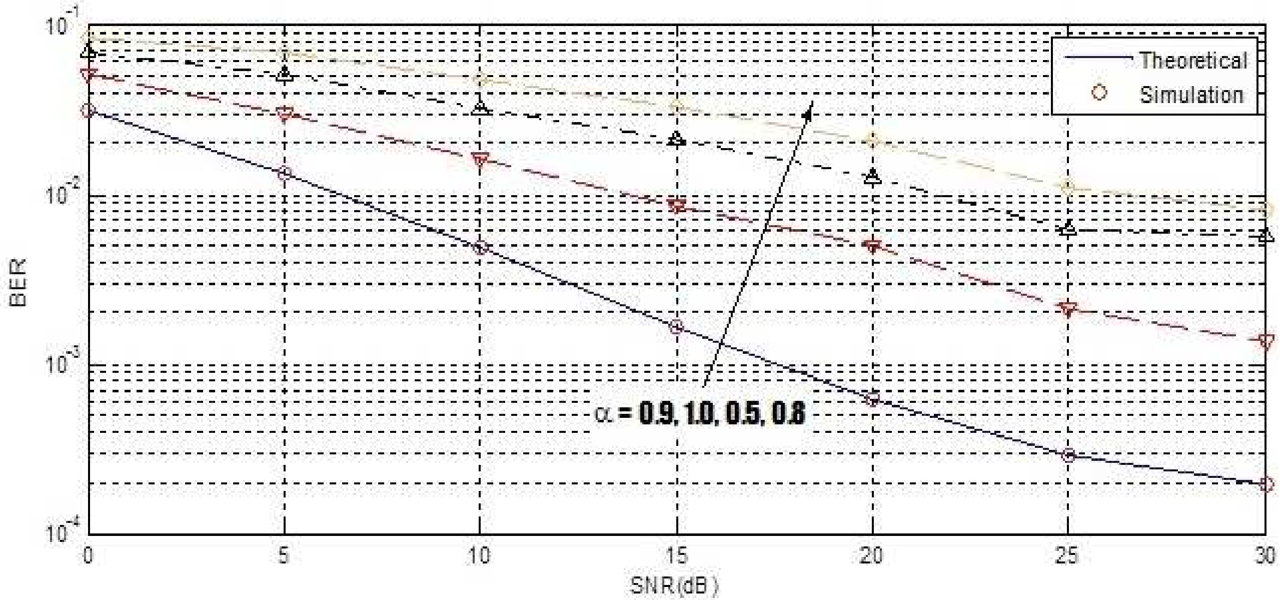

Fig. 3: Theoretical and Simulation results of BER vs. SNR for FrFT-OFDM and FFT-OFDM system

$$\rho = E[\mathrm{r}_{\mathrm{p}-1}\left(\mathrm{k}\right)\mathrm{r}_{\mathrm{p}-1}\left(\mathrm{k}\right)|g,\Delta\theta_P] \tag{21}$$

$$\rho = \left(\frac{N}{N+N_g}\right)X\wedge_{P-1,P-1}X^H + 2\sigma^2$$

$$\alpha + j\beta = E[\mathrm{r}_{\mathrm{p}}\left(\mathrm{k}\right)\mathrm{r}_{\mathrm{p}-1}\left(\mathrm{k}\right)|g,\Delta\theta_P] \tag{22}$$

$$= \left(\frac{N}{N+N_g}\right)X\wedge_{P,P-1}X^H$$

I-bit error probability $P^1_{P,I}\,(g_l)$ of p^{th} subcarrier conditioned on specific phase sequence $g_l = \left[\Delta\theta_{1,l}\ldots\ldots\Delta\theta_{N-1,l}\right]$ is given by

$$P^1_{P,I}\ (g_l) = \tfrac{1}{4}P^1_{P,I}\left(g_l,\Delta\theta_P = \tfrac{\pi}{4}\right) + \tfrac{1}{4}P^1_{P,I}\left(g_l,\Delta\theta_P = \tfrac{3\pi}{4}\right)$$

$$+\tfrac{1}{4}P^1_{P,I}\left(g_l,\Delta\theta_P = -\tfrac{\pi}{4}\right) + \tfrac{1}{4}P^1_{P,I}\left(g_l,\Delta\theta_P = -\tfrac{3\pi}{4}\right)$$

$$P^1_{P,I}\ (g_l) = \tfrac{1}{4}Prob\left\{X^{\,I}_{k,p} < 0|g_l,\Delta\theta_p = \tfrac{\pi}{4}\right\} + \tfrac{1}{4}Prob\left\{X^{\,I}_{k,p} \geq 0|g_l,\Delta\theta_p = \tfrac{3\pi}{4}\right\}$$

$$.[\tfrac{1}{4}Prob\ \left\{X^{\,I}_{k,p} < 0|g_l,\Delta\theta_p = -\tfrac{\pi}{4}\right\}$$

$$+\tfrac{1}{4}Prob\ \left\{X^{\,I}_{k,p} \geq 0|g_l,\Delta\theta_p = -\tfrac{3\pi}{4}\right\}]$$

$$P^1_{P,I}\,(g_l) = \tfrac{1}{2} - \int_0^\infty \frac{I\left\{\theta^I(\omega|g_l,\Delta\theta_P)\right\}}{\pi\omega}\,d\omega$$

$$\text{or } P^1_{P,I}\,(g_l) = \tfrac{1}{2} - \frac{d_{1,l}}{2\sqrt{\varnothing_{1,l}\rho_{1,l} - \beta^2_{1,l}}}$$

Calculating other three probabilities and substituting their values, we get

$$P^1_{P,I}\,(g_l) = \tfrac{1}{2} - \sum_{u=1}^{4}\frac{\sqrt{2}\cos\left(\Delta\theta_{P,u}\right)\alpha_{u,l}}{8\sqrt{\varnothing_{u,l}\rho_{1,l} - \beta^2_{u,l}}}$$

Similarly,

$$P^1_{P,Q}\,(g_l) = \tfrac{1}{2} - \sum_{u=1}^{4}\frac{\sqrt{2}\sin\left(\Delta\theta_{P,u}\right)\beta_{u,l}}{8\sqrt{\varnothing_{u,l}\rho_{1,l} - \alpha^2_{u,l}}}$$

$$P^1_b\,(g_l) = \frac{1}{2(N-1)}\sum_{P=1}^{N-1}[P^1_{P,I}\,(g_l) + P^1_{P,Q}\,(g_l)]$$

Averaging over all possible phase sequences the average BER is given by,

$$P^1_b = \frac{1}{4^{N-2}}\sum_{l=1}^{4^{N-2}}P^1_b\,(g_l)$$

Using Monte Carlo method, final probability of error for large number of subcarriers can be computed as

$$P_e = \frac{1}{M}\sum_{l=1}^{M}P^1_b\,(g_l)$$

6 Simulation and Theoretical Results

In this section, the BER performance of the $\pi/4$-DQPSK FrFT based OFDM system in frequency selective

Rayleigh fading channel had been evaluated. The results achieved had been compared with the BER performance of FFT based OFDM system with same modulation technique under similar conditions. The theoretical and simulation results achieved had been shown in figure 3 for carrier frequency offset of 0.1.

The BER of FFT-OFDM system for $\pi/4$-DQPSK has been evaluated using slandered expression. It had been observed that FrFT based ι OFDM system performance better then FFT based system. For simulation, the number of sub carriers used is 8 with carrier offset of 0.1. Also the theoretical results have been obtained with 100 randomly chosen differential phase sequences for given FrFT based OFDM system. In Figure 3, the solid line shows the theoretical results and the marker shows the simulation results. The simulation has been done to verify the theoretical results and is exactly same with the theoretical results. The FrFT based OFDM system has been simulated for different values of α. The FrFT based OFDM system achieved the best BER of $10^{-3.9}$ with $\alpha = 0.9$ as compared to other values of α as shown in Figure 3. Further, the proposed model have been able to achieve the improvement of 23 dB in SNR for obtaining the $10^{-2.3}$ BER using FrFT based OFDM as compared to FFT based OFDM system.

7 Conclusion

In this paper an expression of BER of FrFT based OFDM system for $\pi/4$ DQPSK modulation has been successfully derived. The response of proposed system for ICI and BER have been derived and simulated for frequency selective Rayleigh fading channel. This study proved that FrFT based OFDM system outperforms the FFT based system under similar conditions. SNR of 30 dB is produced at $10^{-2.3}$ BER whereas the proposed FrFT based OFDM system gives the same BER at 7 dB only. Hence, it could be concluded that the proposed system has achieved the improvement of 23 dB of SNR in the presence of carrier frequency offset of 0.1.

References

[1] S.B.Bulumulla, A.S.Kassam and S.S.Venkatesh, IEEE Transactions on Communications, **48**, 725-728 (2000).

[2] H.Wang, W.L.David and T.H.Sang, IEEE Journal on Selected Areas of Communications, **30**, 684-694 (2012).

[3] U.S.Jha and R.Prasad, OFDM towards fixed and mobile broadband wireless access, Artech House, London, 2007.

[4] M.Russell and G.L.Stuber, IEEE 42th Vehicular Technology Conference, 820-824, Chicam, Illinois, USA (1995).

[5] P.Dharmawansa, N.Rajatheva and H.Minn, IEEE Transactions on Communications, **57**, 26-31 (2000).

[6] T.Keller and L.Hanzo, Proceedings of IEEE, **88**, 611-640 (2000).

[7] L.Xue, S.Hong, V.D.Chakravarthy, M.Temple and Z.Zhiqiang, IEEE Transactions on Communications, **61**, 658-668 (2013).

[8] K.Sathananthan and C.Tellambura, IEEE Transactions on Communications, **49**, 1884-1888 (2001).

[9] H.Liu and U.Tureli, IEEE Communication Letters, **2**, 104-106 (1998).

[10] B.Chen, IEEE Signal Processing Letters, **9**, 123-126 (2002).

[11] L.Jian, L.Guoqing and G.B.Giannakis, IEEE Signal Processing Letters, **8**, 80-82 (2001).

[12] J.Armstrong, IEEE Transactions on Communications, **47**, 365-369 (1999).

[13] S.H.K.Weinfurtner, IEEE Transactions on Communications, **49**, 417-420 (2001).

[14] K.Y.Lin, H.P.Lin and M.C.Tseng, IEEE Transactions on Broadcasting, **58**, 472-479 (2012).

[15] N.N.Tchamoy, J.Rinne, A.Hazmi, M.Valkama, V.Syrjala and M.Renfors, IEEE Wireless Communication Letters, **2**, 6-9 (2013).

[16] E.Chen, R.Tao and X.Meng, IEEE International Conference on Innovative Computing, Information and Control, 14-17, Beijing, China (2006).

[17] H.Wang, L.Wangyong, B.Wensimi and M.Hong, Wireless Personal Communications, **66**, 707-716 (2012).

[18] A.K.Singh and R.Saxena, Wireless Personal Communications, **65**, 189-201 (2011).

[19] R.Saxena and K.Singh, Journal of Indian Institute of Science, **85**, 11-26 (2006).

Ankush Kansal received his B. Tech. and M. Tech. Degree in Electronics and Communication Engineering from PTU, Jalandhar. He is currently pursuing Ph.D. from Thapar University, Patiala in the area of Wireless Communication. He is currently working as Assistant Professor in Thapar University, Patiala. He has published 15 research articles in referred international journals, international conference and national conference. He is a life time member of ISTE. His research interest includes Networking, Wireless Communication and Embedded Systems.

Kulbir Singh was born in Batala (Pb) India. He received his B.Tech degree in 1997 from PTU, Jalandhar. He obtained his ME and Ph.D. degree from Thapar Institute of Engineering and Technology, Patiala in 2000 and 2006 respectively. He worked as lecturer from 2000 to 2007 and as Assistant Professor from 2007 to 2011 in Electronics and Communication Engineering Department, Thapar University, Patiala. Presently he is working as Associate Professor in Thapar University, Patiala since July 2011. He has published about 75 research articles in refereed international journals, international conference and national conference. He is life time member of IETE and ISTE. His research interests include Digital Signal Processing, Image Processing, Fractional Fourier Transform and Communication systems.

Rajiv Saxena was born at Gwalior in Madhya Pradesh in 1961 obtained B.E. (Electronics & Telecommunication Engineering) in the year 1982 from Jabalpur University, Jabalpur. Subsequently, Dr. Saxena joined the Reliance Industries, Ahmedabad, as Graduate Trainee. In 1984, Dr. Saxena joined Madhav Institute of Technology & Science, Gwalior as Lecturer in Electronics Engineering. He obtained his M.E. (Digital Techniques & Data Processing) from Jiwaji University, Gwalior in 1990. The Ph.D. degree was conferred on him in 19961997 in Electronics & Computer Engineering from IIT, Roorkee (erstwhile UOR, Roorkee). Dr. Saxena was head and professor in ECE department at JUET, Guna. Currently he is working as Director, Jaypee University, Anoopshahr (UP).