COMPENSATION OF SYMBOL CLOCK OFFSET AND CARRIER FREQUENCY OFFSET IN THE MULTI-BAND DFT SPREADING OFDM SYSTEM

Sang-Burm Ryu (Chungbuk National University, KOREA; uglypean@nate.com);
Heung-Gyoon Ryu (Chungbuk National University, KOREA; ecomm@cbu.ac.kr);
Jin-up Kim (ETRI, Dae-jeon KOREA; jukim@etri.re.kr)

ABSTRACT

Spectrum aggregation has been studied widely to extend usable frequency bandwidth. Although bandwidth efficiency is improved by using vacant frequencies, structure of receiver is very complicated. Also, interferences among usable bandwidths can be problem because of receiving data from multi-band. So, in this paper, we simplify the receiver structure and would like to propose a new suitable method for direct conversion receiver demanding of existing receiving method. Proposed system multiplexes many bands by high speed clock sampling after band selection and receives DFT spreading OFDM signal as a TDM (time division multiplexing) method. So, the structure of receiver becomes simple and degradation of system performance by interferences among used bands is reduced. However, throughput is decreased by symbol synchronization error and sampling clock offset because the system uses high speed clock sampling method. Therefore, to solve this problem, we propose a suitable synchronization method and an offset compensation method. We can ensure that frequency efficiency is increased. Also we improve the system performance in a multipath channel environment.

1. INTRODUCTION

For next generation mobile communication system, the bandwidth of 100MHz and transmission speed of 1Gbit/s are demanded [1]. But use of continuous frequency channels is difficult due to shortage of useable frequency bandwidth in practice. Therefore, spectrum aggregation technique gets attention in recent. Spectrum aggregation technique uses distributed frequency channels as a wide bandwidth channel for high speed data communication [2]. Generally, allocated channel of cellular system is always not used. So, useable bandwidth is increased by aggregation of vacancy frequency channels. Such system is spectrum aggregation system. The receiver for spectrum aggregation must be able to receive data from distributed channels simultaneously and to operate in mobile terminal. Also complexity of receiver must be low. To meet those demands, band pass sampling in receiver is proposed. But this structure of receiver is complicated [3]. And also, SDR (software define radio) of direct conversion type and Hilbert transform type are proposed [4][5]. But receiver size is increased as much as useable channels simultaneously in OFDM system.

In OFDM system, there are a lot of studies to remove phase noise and frequency offset. PNS is proposed to eliminate phase noise only in [6] but performance is degraded when multipath channel and frequency offset are considered. In [7], phase noise and frequency offset are suppressed by controlling of CIR (channel impulse response) to remove ICI. But structure of this method is complicated and this does not guarantee the performance by ICI removed.

So, in this paper, we propose a receiver based on TDM that is appropriate structure in DFT spreading OFDM system using a spectrum aggregation technique for transmission of data to multi-band at the same time. Before, proposed direct conversion type receiver using dual band is weak to the interference because this receiver combines two received inputs. Also, block size operating discrete signal is increased according to the number of used band and removing interference among used bands is very difficult in case of that one FFT (fast frequency transform) is exploited. So, band pass filter is used after receive part of antenna such as direct conversion method in receiver using multi-band. Then, outputs of each filter are sampled by sample and hold method by using high speed clock generator. The inputted symbol of DFT spreading OFDM as TDM by multiplexing sampling outputs is allocated and passes through FFT process. So, high speed sampling is possible by one receiver.

In the proposed system, we would like to propose the structure for compensating synchronization of symbol timing offset and sampling clock offset and to improve system performance when different frequency offset and phase noise are compensated in multi-path channel by mobility.

2. DFT SPREADING OFDM FOR MULTI-BAND COMMUNICATION

Figure 1 is a transceiver for multi-band communication. In case of one user, two bands are used like figure 1. For convenience, $x(t)$ is transmitted signal to two bands simultaneously and different frequency offset and phase noise are generated each other in the only transmitter. Transmitted signal passes through channel and band selection process. After that, the signal is converted to TDM signal by multiplexer operated by high speed clock.

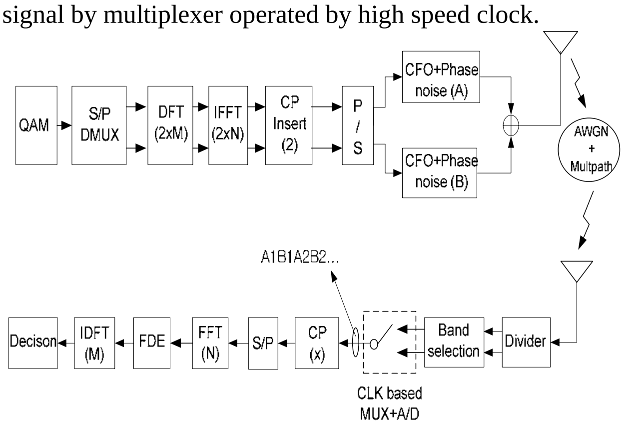

Figure 1. Block diagram of DFT spreading OFDM system for multiband communication.

Received signal is decided after IDFT.

$$x(t)=\begin{cases}\dfrac{1}{\sqrt{N}}\displaystyle\sum_{k=0}^{N-1}X_k^A e^{j(\frac{2\pi k}{N}+f_A)t}, x_A(t)\\ \dfrac{1}{\sqrt{N}}\displaystyle\sum_{k=0}^{N-1}X_k^B e^{j(\frac{2\pi k}{N}+f_B)t}, x_B(t)\end{cases} \tag{1}$$

Equivalent model of phase noise refers to [10]. And the effect of phase noise and frequency offset is below in the receiver.

The effect of phase noise and frequency offset can be indicated to $e^{j(2\pi(L+\varepsilon)n/N+\theta(n))}$. Here, normalized frequency offset is $\varepsilon=\Delta fT$ and T is symbol period. In convenience, $2\pi(L+\varepsilon)n/N+\theta(n)=\Phi(n)$ is replaced. And if $\Phi(n)<<1$, that is expressed as follows.

$$Q_L=\frac{1}{N}\sum_{n=0}^{N-1}e^{j\frac{2\pi}{N}Ln}\cdot(1+j\Phi(n)) \tag{2}$$

$$Q_0=1+\frac{j}{N}\sum_{n=0}^{N-1}\Phi(n)$$

$$Q_{l-k}=\frac{j}{N}\sum_{n=0}^{N-1}e^{j\frac{2\pi}{N}(l-k)n}\cdot\Phi(n)$$

Transmitted signal including phase noise and frequency offset is written as

$$s(t)=x_A(t)\cdot e^{j\Phi_A(t)}+x_B(t)\cdot e^{j\Phi_B(t)}+i(t). \tag{3}$$

Sampled data of received signal which passes through AGWN and multipath channel is expressed as

$$r(n)=s(n)\otimes h(n)+v(n). \tag{4}$$

Here, $s(n)$, $h(n)$, $\upsilon(n)$ and $r(n)$ are transmitted signal, channel impulse response, complex Gaussian noise and received signal.

After band is selected by filtering of received signal, TDM is processed and that signal passes through cyclic prefix process and FFT process. Recovered signal is below and received signal at *k*-th subcarrier is written as below because there is not X^B signal for transmitting time of X^A.

$$\begin{aligned}Y_k^A&=\frac{1}{N}\sum_{n=0}^{N-1}r_A[n]\cdot e^{-j\frac{2\pi}{N}kn}\\&=\frac{1}{N}\sum_{n=0}^{N-1}\{x_A(n)\cdot e^{j\Phi_A(n)}\otimes h(n)+\upsilon_A(n)\}\cdot e^{-j2\pi(n+\delta_A)/N}\cdot e^{-j\frac{2\pi}{N}kn}\\&=\frac{1}{N}\sum_{n=0}^{N-1}\sum_{i=0}^{N-1}X_i^A\cdot H_i\cdot e^{j\left[\frac{2\pi}{N}n\{(i-k)+\Phi_A(n)+\delta_A\}\right]}+N_k\end{aligned} \tag{5}$$

3. SYMBOL SYNCHRONIZATION

Figure 2 is the process about synchronization of symbol or clock for multi-band communication. First, symbol synchronization is done by training symbol. Training symbol is transmitted two times continuously and random PN sequence amount of sub-carriers is send after FFT process. Symbol or clock synchronization process is divided into 3 processes for switching operation using high speed clock for TDM signal.

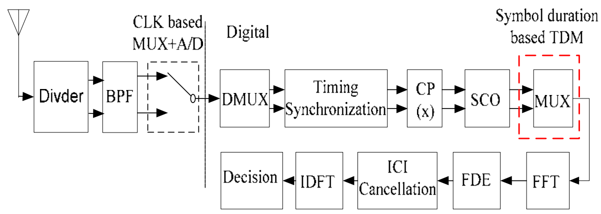

Figure 2. Block diagram of DFT spreading OFDM receiver for multiband communication.

First, estimate an accuracy symbol timing offset by using cyclic prefix of training symbol in time domain. In this process, switching clock of clock based multiplexer is adjusted by estimating one sample of OFDM. For that, symbol synchronization errors of band A and band B are estimated after multiplexing TDM signal. Minimum error method is used as an estimation method.

Second, clock offset is estimated in frequency domain by training symbol because sometimes sampling clock which is generated from FFT between transmitter and receiver is not the same. Estimation method can detect phase rotation value since we can know a specific symbol through pilots of training symbol.

Third, in case that clock frequency is not matched, changed clock offset is compensated by updating to previous process of FFT by using pilot of received signal because estimation through only training symbol is difficult. Also clock offset is compensated continuously by estimated phase value in frequency domain. But system performance is degraded because ICI is not removed. So, if we remove phase noise and frequency offset by proposed ICI cancellation method, we can get satisfied performance although phase noise and frequency offset exist.

Timing offset estimation method in [11] is used in figure 2.

$$\hat{\delta} = \arg\min_{\delta}\{ \sum_{i=\delta}^{N_G-1+\delta} \left| r^A[n+i] - r_l^{A^*}[n+N+i] \right|^2 \} \tag{6}$$

We set synchronization time when difference between window of the same size symbol and window using cyclic prefix is minimum. At this time, good performance is maintained although frequency offset is generated. Cyclic prefix is eliminated after symbol synchronization and process in figure 3 is operated in order to estimate clock offset.

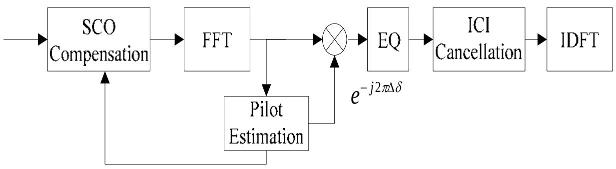

Figure 3. Compensation and estimation of sampling clock offset.

First, oversampling is done to compensate sampling time offset in time domain. That is compensated by controlling sampling clock with phase value estimated after FFT as a criterion and compensated by down-sampling method. And we estimate that by using pilot since effect of clock offset after FFT is shown as phase change.

Equation is assumed after symbol synchronization and FFT and is below.

$$P_k = P(k) \cdot e^{-j2\pi k \Delta\delta / N} \tag{7}$$

A below equation is process to estimate clock offset. Through equation (8) and (9), clock offset is compensated like figure 3 by using received first and third pilots.

$$P_{offset} = \sum P_k P_{k+2}^* \approx \sum |P|^2 \cdot e^{j\pi\Delta d} \tag{8}$$

$$\Delta\delta = \arctan\left(\frac{\mathrm{Im}(P_{offset})}{\mathrm{Re}(P_{offset})} \right) / \pi \tag{9}$$

Effect of phase noise and frequency offset is below after FFT of received signal which is removed about symbol synchronization and clock offset.

$$Y_k = H_k \cdot X_k \cdot Q_0 + \sum_{i=0,\, i\neq k}^{N-1} H_k X_k \cdot Q_{i-k} + N_k \tag{10}$$

Such as below equation, channel estimation process is operated after removing symbol synchronization and clock offset by continuous two training symbol based on simple LS (least square).

$$H_k = Y_k / X_k \tag{11}$$

Factor of channel estimation is below.

$$\tilde{H}_k = H_k Q_0 + H_k \left(\frac{X_k^\#}{X_k} \right) Q_0 + \frac{ICI + N_k}{X_k}$$
$$= H_k \cdot \left(\frac{X_k^\#}{X_k} \right) Q_0 + W_k \tag{12}$$

4. ESTIMATION AND COMPENSATION OF FREQUENCY OFFSET AND PYHASE NOISE

Channel is removed after FFT of received signal after cyclic prefix is eliminated.

$$\tilde{Y}_k = Y_k / H_k = X_k Q_0 + \left(\frac{X_k^\#}{X_k} \right) Q_0 + \frac{ICI + I_k + N_k}{X_k} \tag{13}$$

First part of above equation is CPE part and second part is ICI part. Then, pilot symbol is used in order to estimate CPE after removing channel, estimated signal is below.

$$\tilde{Y}_k = X_k \tilde{Q}_0 + \left(\frac{X_k^\#}{X_k} \right) \tilde{Q}_0 + W_{ICI+AWGN}\,. \tag{14}$$

In equation (14), N_p is the number of pilot symbol, S_p is pilot symbol and W_k is total interference by ICI and AWGN.

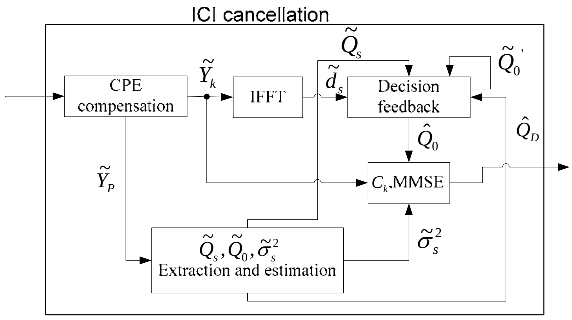

Figure 4. Extraction and compensation of phase noise and frequency offset.

CPE of phase noise is repeated regularly. So, that is compensated by dividing average of pilot symbol to next data symbol. Compensated data is written as

$$CPE_p = \frac{Y_p}{X_p} = Q_0 + \left(\frac{X_p^\#}{X_p} \right) Q_0 + \frac{ICI + N_p}{X_p} = Q_0 + \left(\frac{X_p^\#}{X_p} \right) Q_0 + W_p$$

$$r_{cpe} = \frac{1}{N_p} \sum_{k\in s_p} CPE_k = Q_0 + \left(\frac{X_p^\#}{X_p} \right) Q_0 + 1/N_p \sum_{k\in s_p} W_k\,. \tag{15}$$

Below equations are for removing ICI.

$$\tilde{Q}_s = \left(\frac{1}{N_P} \sum_{k \in s_p} \left| (Y_p - \tilde{Y}_p) \right| \right)^2 \tag{16}$$

$$\min \sum_{Q_0, k \in s_p} \left| \tilde{Y}_P - D_k \tilde{Q}_{i-k} \right|^2 \tag{17}$$

$$\tilde{Q}_0 = \frac{\sum_{k \in s_p} \tilde{Y}_p D_p^*}{\sum_{k \in s_p} \left| D_p \right|^2} \tag{18}$$

$$C_k = \frac{\hat{Q}_0^*}{\left| \hat{Q}_o^* \right|^2 + \frac{\tilde{\sigma}_s^2}{E_x}} \tag{19}$$

$$\hat{Q}_0 = \tilde{Q}_s \tilde{Q}_0 + (1 - \tilde{Q}_s) \hat{Q}_D \tag{20}$$

$$\hat{Q}_D = \frac{\sum_{s \in s_d} \tilde{d}_s d_s^*}{\sum_{s \in s_d} \left| d_s \right|^2} \tag{21}$$

$$\tilde{\sigma}_s^2 = \frac{1}{N_P} \sum_{k \in s_p} \left| \tilde{Y}_k \right|^2 \frac{1}{N_P} \sum_{k \in s_p} \left| Y_p - \tilde{Y}_k \right|^2 \tag{22}$$

To get sample of $\tilde{Q}_o$, (18) satisfying minimum of (17) can be obtained by using pilot symbol. And interference power by ICI and noise is estimated through equation (22) to estimate SNR. By using $\tilde{Q}_0$, initial C_k is estimated through (19) without equalization process. Equation (21) is operated to updating $\hat{Q}_0$.

$(\cdot)^*$, $\tilde{\sigma}_s^2$ and E_x in (19) means conjugate, power of $W_{ICI+AWGN}$ and useful power. Signal without CPE in (15) is extracted as the same number to symbol number for decision feed back in (20). Here, extracted symbol after IFFT is $\tilde{d}_s$ and known data symbol is d_s. By (20) of decision feedback process is operated and $\hat{Q}_D$ is updated and $\hat{Q}_0$ is obtained in frequency domain. And then, received signal in (15) is compensated through (21).

5. SIMULATIONS RESULTS AND DISCUSSION

Simulation parameters are below.

- Modulation level: 16 QAM
- OFDM parameter: A band=M(60), B band=M(60), pilots=4, N=512.
- Communication band number: Dual
- Channel: AWGN + Multipath
- Phase noise parameter: -20 dBc~-18dBc, cutoff 10KHz
- Frequency offset parameter: CFO=0.01~0.03
- Cyclic prefix: 32
- Sampling timing offset: 10~4 sample
- Sampling clock offset: 0.1~0.4 sample
- Multipath channel: TS 25.104, No Doppler, No Fading
- Multipath channel 1: Pedestrian A channel
- Multipath channel 2: Vehicular A channel

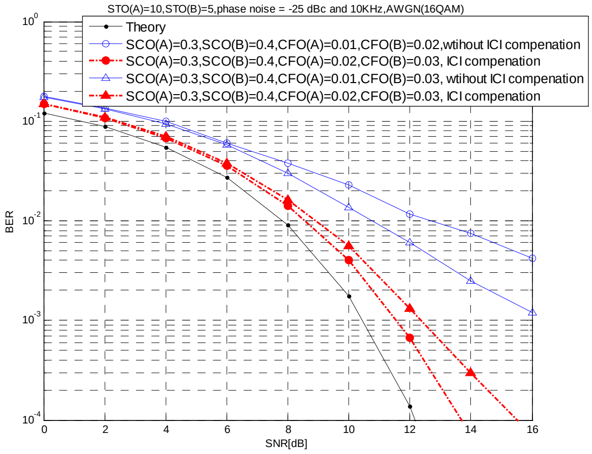

Figure 5. BER comparison according to frequency offset (AWGN, 16QAM).

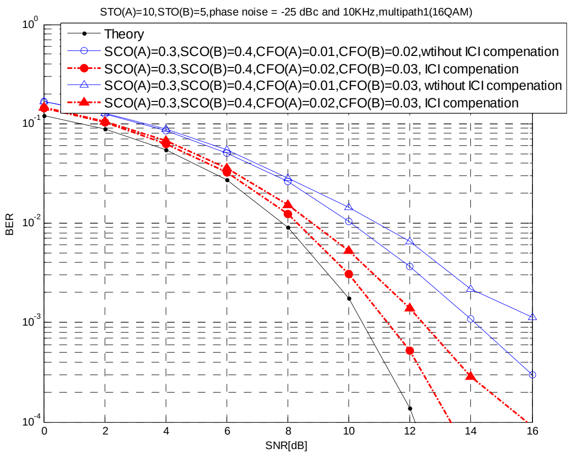

Figure 6. BER comparison according to SCO in multipath 1 (16QAM).

Figure 5 shows BER performance according to the conditions. In A band, STO=10, SCO=0.3 and CFO=0.01~0.02. In B band, STO=5, SCO=0.4 and CFO=0.02~0.03. When CFO(A) and CFO(B) are 0.02 and 0.03, compensation performance about ICI is degraded regarding 1dB than CFO(A) and CFO(B) are 0.01 and 0.02 at BER=10^{-4}. But system performance is improved about 2.5dB by ICI compensation method. Therefore, we know system performance is degraded when frequency offset compensates SCO. And estimation errors increase ICI.

Fig. 6 shows BER performance according to the conditions. In A band, STO=10, SCO=0.3 and CFO=0.01~0.02. In B band, STO=5, SCO=0.4 and CFO=0.02~0.03 in multipath 1 channel. When CFO(A) and CFO(B) are 0.02 and 0.03, compensation performance about ICI is degraded regarding 1.3dB than CFO(A) and CFO(B) are 0.01 and 0.02 at BER=10^{-3}. But system performance is improved about 2.4dB by ICI compensation method. Performance is more degraded in this case because frequency offset influences to estimation of SCO and the amount of estimation error makes system degraded.

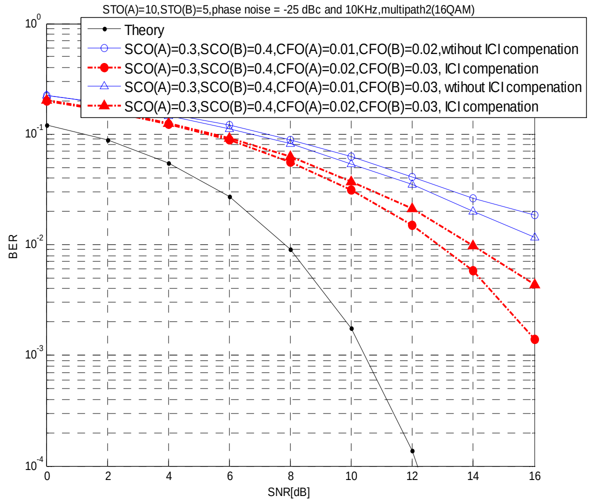

Figure 7. BER comparison according to SCO in Multipath 2 (16QAM).

Fig. 7 shows BER performance according to the conditions. In A band, STO=10, SCO=0.3 and CFO=0.01~0.02. In B band, STO=5, SCO=0.4 and CFO=0.02~0.03 in multipath 2 channel. The difference between compensation and non-compensation is 4.4dB at BER=10-2. Overall performance is degraded depending on the effect of band which has more poor performance because receive performance get different according to the channel response. Therefore, estimation channel response of whole bandwidth and allocation data to good channel quality band is very important.

6. CONCLUSIONS

We suggest an effective method to high speed data transmission system through compensation of synchronization and clock offset for symbol synchronization by de-multiplexing. Also, we ensure the improved performance by using ICI cancellation method when frequency offset and phase noise exist those are generated in the process of synchronization. To review the effect of frequency offset and phase noise in the process of synchronization of symbol and clock offset, we improve system performance 2.5dB in AWGN, 2.4dB in multipath 1 and 4.4dB in multipath 2 by cancelation of ICI. And we ensure that we can recover the loss by increasing the speed of sampling clock by band width efficiency because system performance is improved when the sub-carrier position allocated to each band is the same. Consequently, TDM is used for reducing complexity of receiver generated by using multi-band simultaneously. And we adapt synchronization method and compensation method of clock offset to know the problem about TDM. And we analyze system through simulation and we overcome the problem by using ICI cancellation method.

7. ACKNOWLEDGEMENT

This work was supported by the IT R&D program of MKE/KEIT. [KI001869, Research on environment-adaptive autonomous technologies for mobile wireless access]

8. REFERENCES

[1] ITU-R M.1645, “Framework and overall objectives of the future development of IMT-2000 and systems beyond IMT-2000,” Jun. 2003.

[2] 3GPP TR 36.913 v8.0.0, “Requirements for Further Advancements for E-UTRA (LTE-Advanced),” June 2008.

[3] Dennis M. Alos and Michael Stockmaster and James B. Y. Tsui and Joe Caschera, “Direct Bandpass Sampling of Multiple Distinct RF Signals,” *IEEE Trans. on Commun.*, Vol. 47, pp.983-988, Jul. 1999

[4] Stephen Wu and Behzad Razavi, “A 900-MHz/1.8-GHz CMOS Receiver for Dual-Band Applications,” *IEEE J.olid-State Circuits*, Vol. 33, pp. 2178-2185, Dec. 1998.

[5] Hossein Hashemi and Ali Hajimiri, “Concurrent Multiband Low-Noise Amplifiers-Theory, Design, and Applications,” *IEEE Trans. Microwave Theory Tech*., pp.288-301, Jan. 2002.

[6] Songping Wu, Bar-Ness, Y, "A phase noise suppression algorithm for OFDM-based WLANs," *IEEE Communications Letters*, vol. 6, Issue, 12, pp. 535-537, Dec. 2002.

[7] Songping Wu and Y. Bar-Ness, "OFDM Channel Estimation in the Presence of Frequency Offset and Phase Noise, " *in Proc. Of IEEE International Conference on Communication (ICC)* vol 5, pp.3366 – 3370, May 2003

[8]R. Corvaja and A.G. Armada, "Joint Channel and Phase Noise Compensation for OFDM in Fast-Fading Multipath Applications," *IEEE Transactions, Vehicular Technology*, vol. 58, issue 2, pp. 636 - 643 Feb. 2009

[9] A. G. Armada, “Estimation and correction of phase noise effects in orthogonal frequency division multiplexing,” *in Proc. of IEEE GLOBECOM, San Francisco*, CA, pp. 1–5, Nov. 2006.

[10] H.G. Ryu and Y.S. Li, "Phase noise analysis of the OFDM communication system by the standard frequency deviation," *IEEE Transactions on Consumer Electronics*, vol. 49, no. 1, pp. 41-47, Feb.2003.

[11] M.Speth.F.Classen, and H. Meyr, " Frame synchronization of OFDM systems in frequency selective fading channels, " *IEEE VTC’97*,pp.1807-1811,May 1997.