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Abstract— **Ranging is one of the most important processes in the mobile WiMAX standard, for resolving the uplink synchronization and near/far problems. In this paper, under WiMax OFDMA ranging framework, different methods for carrier frequency offset (CFO) estimation are proposed, and they are compared together with different existing methods for symbol timing offset (STO) estimation. Based on the analysis, a joint timing and frequency offset (STO-CFO) estimation method for WiMAX uplink ranging is proposed, which could be used for multi-user ranging, without interference from synchronized data users and remove the STO impact in CFO estimation. Furthermore, this method is proved to be valid for different kinds of ranging used in WiMAX system. Simulation results show that the proposed method performs better than the other methods and it is more robust to multi-user interference. It is a strong trend to implement the wireless base band in software. Both the complexity analysis on this method and the real performance data got from multicore platform could show, it is a high efficient design which is affordable by pure software implementation.**

I. Introduction

Broadband wireless access (BWA) has attracted much attention recently. Future generation networks will be characterized by variable and high data rates, quality of services and seamless mobility both within a network and between networks of different technologies and service providers. A technology developed to fulfil these characteristics, standardized by IEEE, is 802.16. It is commonly referred to as Worldwide Interoperability for Microwave Access (WiMAX) [1].

One of the most interesting PHY modes supported by WiMAX standard is Orthogonal Frequency Division Multiple Access (OFDMA) PHY mode. In OFDMA, subcarriers are grouped into subchannels which are assigned to multiple users for simultaneous transmissions. If the users are not synchronized with the receiver, they will interfere with each other and the base station (BS) will not be able to recover individual signals of each user. The issues of uplink synchronization and near/far problems are addressed by a process called “ranging”. Generally, a ranging process includes initial ranging (for any ranging subscriber station (RSS) that wants to synchronize to the system for the first time), handover ranging (to support mobility and perform handoff from one access point (AP) to another), periodic ranging (to update and track variations in symbol timing offset (STO) and carrier frequency offset (CFO) and bandwidth request (to request access to the shared spectrum resource).

In ranging, RSS randomly chooses a ranging time-slot and a ranging code which is modulated by Binary Phase Shift Keying (BPSK) modulation. Then the ranging transmission shall be performed during two or more consecutive symbols to increase the probability of code detection [1]. After separating colliding codes and extracting information about timing, frequency and power, the BS will broadcast the identified ranging codes with the needed adjustments information (e.g. timing, frequency and power) and a status notification (e.g. success, continue, and abort). The ranging process at BS mainly includes STO and CFO estimation, and power estimation. This paper mainly considers the STO and CFO estimation for different ranging processes at the BS.

There will be following challenges for STO and CFO estimation in WiMax uplink ranging.

1) STO and CFO estimation needs to handle multi-user problem.
2) The data signal and ranging signal will be transmitted at the same OFDM symbols.
3) The offset in one domain (e.g. time domain) will impact the estimation in the other domain.
4) There are different signal structures employed for different kinds of ranging in WiMax, which may not be suitable for the same algorithm.

Frequency and timing recovery for single-user OFDM has been widely discussed in many literature (e.g. [2], [3]). However, they cannot directly be used in the uplink of a multi-user system which needs to separate different users at the BS before their timing and frequency offsets are estimated respectively. Some synchronization methods for multi-user OFDM systems have the same problem with multiple users colliding in the same ranging channel (e.g. [4]–[6]). In [7], a time-domain correlator bank based ranging method was proposed. In this case, the cross correlation with all possible ranging codes was performed over at least one complete OFDMA symbol. However, the CFO estimation scheme was not addressed. Another disadvantage is that the Pseudo Noise (PN) codes were originally sent over frequency and thus correlating in time domain will impair the cross correlation properties between different codes. In [8], the frequency offset can be acquired by correlating FFT output samples of two consecutive OFDMA symbols at ranging subcarriers. However, this method will be invalid in the case of multiple ranging users as it can not distinguish between them.

For the first two challenges of WiMax ranging in uplink, little work on efficient frequency synchronization has appeared in the literature. And till now, there isn't any discussion about the challenge 3 and 4.

To overcome these limitations and provide an effective solution for STO and CFO estimation in WiMax ranging, we investigate different STO estimation methods for multi-user case, and propose two CFO estimation methods based on cross-correlation in time and frequency domain respectively. It is proven in this paper that, the time offset will heavily impact the CFO estimation. So considering the advantage on handling the challenge 2 and 3, a joint STO-CFO estimation method is proposed, where the STO and CFO estimation both employ the cross-correlation in frequency domain. And it is proven that, this method could be used for different ranging in WiMax.

The rest of the paper is organized as follows. Section II introduces the signal model. Different ranging methods (including STO estimation and our proposed CFO estimations) are analyzed in Section III. Joint STO-CFO estimation method and its implementation architecture is proposed, together with the complexity discussion in Section IV. Performance analysis is provided in Section V, and conclusions are drawn in Section VI.

II. Signal Model

Our system model is mainly based on the IEEE 802.16e standard [1]. We consider an OFDMA uplink system with *N* subcarriers. After assigning DC and guard subcarriers, the remaining subcarriers, N_d, are grouped into *Q* subchannels. Each subchannel has $P = N_d/Q$ subcarriers. The subcarriers assigned to each subchannel are not necessarily adjacent as they are chosen randomly. Each user in uplink is assigned to one or more subchannels. A ranging channel is composed of one or more groups of six adjacent subchannels. WiMAX defines different ranging signal structure among different kinds of ranging. To simplify the discussion, only the signal structure of initial ranging will be considered in Section II and III. The impact of different signal structures will be discussed in Section IV. So for initial ranging, same ranging code with length-*R* is modulated and transmitted in the ranging channel during one ranging time-slot composed of two OFDMA symbol duration.

The *m*-th ranging code randomly selected by the *u*-th RSS is firstly modulated by BPSK modulation, which is denoted by $\mathbf{S}_m = [S_m(0), S_m(1), ..., S_m(R-1)]$. The signal is then mapped into a set of *N* modulation symbols, $X_{R,m}(k), k = 0, 1, ..., N-1$, according to

$$X_{R,m}(k) = \begin{cases} S_m(r), & if \quad k = ind(r) \\ 0, & otherwise, \end{cases} \tag{1}$$

where $ind(r)$ is the index of the *r*-th ranging subcarrier. If the length of the cyclic prefix (CP) is equivalent to N_{CP} samples, and assumed to be longer than the maximum channel delay spread. After *N*-point inverse fast Fourier transform (IFFT) and CP insertion at the transmitter, the *n*-th element of the time-domain ranging signal of the *u*-th ranging user is given by

$$x_{R,m}^{(u)}(n) = \begin{cases} \sum_{k=0}^{N-1} X_{R,m}(k) e^{j2\pi i(n-N_{\text{CP}})/N}, \\ \qquad n = 0, ..., N+N_{\text{CP}}-1 \\ \sum_{k=0}^{N-1} X_{R,m}(k) e^{j2\pi i(n-N-N_{\text{CP}})/N}, \\ \qquad n = N+N_{\text{CP}}, ..., 2N+2N_{\text{CP}}-1. \end{cases} \tag{2}$$

Note that the second symbol has a cyclic postfix instead of a cyclic prefix as shown in Fig. 1. Similarly, the data signal of the *g*-th data subscriber station (DSS) is denoted by

$$x_D^{(g)}(n) = \sum_{k=0}^{N-1} C_g(k) e^{j2\pi i(n-N_{\text{CP}})/N}, \tag{3}$$

where $n = 0, ..., N+N_{\text{CP}}-1$, $C_g(k)$ is the *i*-th subcarrier signal transmitted by the *g*-th DSS. Because different SSs will have different location, the corresponding transmission delays ($d_{R,u}$ for *u*-th RSS and $d_{D,g}$ for *g*-th DSS) in units of OFDM samples are different. The maximum possible relative delay $d_{R,max}$ for the RSS is the round-trip transmission (RTD) at the cell boundary, which can be found from the knowledge of cell radius in practice. The maximum delay $d_{D,max}$ for DSSs is determined by the timing requirement of the ranging process.

We consider a multipath Rayleigh fading channel with *L* sample-spaced taps. The channel impulse responses (CIRs) for *u*-th RSS and *g*-th DSS (denoted by $h_R^{(u)}(l)$ and $h_D^{(g)}(l)$, $l = 0, ..., L-1$, respectively) are assumed to be constant over one ranging time-slot and nonzero only for $l = 0, .., L-1$. Note that N_{CP} for RSSs should be designed such that $N_{\text{CP}} \geq d_{R,max} + L$. So the channel output samples for *u*-th RSS and those for *g*-th DSS should be presented by

$$y_R^{(u)}(n) = \sum_{l=0}^{N-1} h_R^{(u)}(l) x_R^{(u)}(n - l - d_{R,u}), \tag{4}$$

$$y_D^{(g)}(n) = \sum_{l=0}^{N-1} h_D^{(g)}(l) x_D^{(g)}(n - l - d_{D,g}). \tag{5}$$

Suppose that there are N_R RSSs and N_D DSSs in one ranging time-slot in the system. Then the *n*-th received signal sample at the BS can be expressed as

$$y(n) = \sum_{u=0}^{N_R} y_R^{(u)}(n) + \sum_{g=0}^{N_D} y_D^{(g)}(n) + z(n), \tag{6}$$

where $\{z(n)\}$ are independent and identically distributed (iid), circularly-symmetric complex Gaussian noise samples with zero mean and variance σ_z^2.

III. Ranging Method

This paper is to discuss the STO and CFO estimation method for WiMAX ranging, so we will fully consider the mechanisms provided by WiMAX standard when analyzing different ranging methods. They are cyclic prefix/postfix (CP), PN codes and repeated symbols.

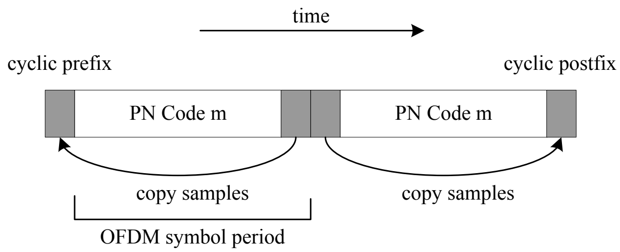

Fig. 1. Initial ranging transmission over two consecutive OFDMA symbols

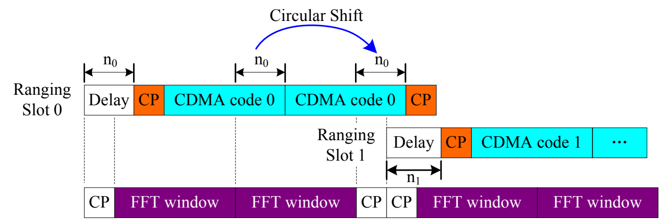

Fig. 2. Frequency domain correlation

In wireless system, usually the purpose for CP is to allow multipath to settle before the main data arrives at the receiver. There will be two properties from CP aiding the normal synchronization.

- CP has the property of duplication, which could be employed for synchronization. But it can't be utilized simply in multiuser environment.
- The length of CP will usually be designed being larger than maximum delay of multipath. So in OFDM system, it could ensure the FFT window to cover the whole OFDM symbol with only a cyclic shifting.

PN codes has the good cross-correlation property, so it's widely used in the wireless system for multiuser detection.

Frequency offset will result in phase rotation across time. By comparing the phase difference of the successive repeated symbols, CFO could be estimated.

Based on these provided mechanisms, in this section, the STO and CFO estimation methods for WiMAX uplink systems are introduced and compared. We assume that N_R is less than the number of initial ranging codes N_c and the ranging code transmitted by all RSSs are different. Ranging code detection is not considered in this paper. The threshold analyzed in [9] is utilized to detect the ranging codes. If the peak correlation value is above the threshold, we consider the corresponding ranging code as the transmitted code.

A. Timing Offset Estimation

1) Cross-correlation in Time Domain: According to the very low cross-correlation property of different ranging codes, a bank of time-domain correlators are used to separate different ranging codes [7]. The output of each correlator can be used to estimate the timing offset of a possible ranging user. Let $x_{R,m}(n)$ represents the *n*-th time-domain signal sample of the *m*-th ranging code, the cross-correlation between the received signal and N_c referential ranging codes can be defined by

$$R_m(d) = \sum_{n=0}^{N-1} x_{R,m}(n) y^*(d+n+N_{\text{CP}}), \tag{7}$$

where $d = 1, 2, ..., d_{max,R}$, $u = 1, 2, ..., N_c$, and $(\cdot)^*$ represents conjugate. Then the corresponding timing offset of the *m*-th correlator is given by

$$\hat{d}_{R,m} = arg \max_{d} |R_m(d)| \,. \tag{8}$$

As the PN codes were originally sent over frequency, correlating in time domain will impair the cross correlation properties between different codes.

2) Cross-correlation in Frequency Domain: The received signal is a combination of data from synchronized users and CDMA code from the ranging subscriber. Until now we considered correlation in time domain, where the correlation is performed over the entire received signal. Timing offset can also be found by performing the correlations in frequency domain. The advantage gained is less interference, as we can separate the data subcarriers from the subcarriers used for ranging. A time offset in time domain corresponds to phase offset in frequency domain.

$$y((n-n_0)_{mod(N)}) \overset{DFT}{\longleftrightarrow} e^{-j2\pi k n_0/N} Y(k), \tag{9}$$

As shown in Fig. 2, the FFT is considered over the received data and since there is a delay involved, only the second FFT would correspond to the circularly shifted code and its output gives us the timing offset value. Then we multiply it with phase component to get the required phase shift corresponding to integer number of samples in time domain, finally we correlate it with the CDMA code. Simulation results prove that this system works very well because of reduced interference from the data subcarriers, but the complexity is very high because of the high number of multiplications required to achieve the phase shifts.

Instead of providing phase shifts by multiplication of phase component, a very simple method is to take an IFFT after multiplication of code in frequency domain. IFFT is equivalent to providing phase shift, hence reduces a lot of complexity.

From the received signal in frequency domain, $Y(k)$, we fetch the signal at the ranging subcarriers $\tilde{X}_R(k)$ which has a phase offset n_0 from the reference ranging code in frequency domain, $X_{R,m}(k)$.

$$\tilde{X}_R(k) = X_{R,m}(k) e^{-j2\pi k n_0/N}. \tag{10}$$

Then we acquire the estimation of n_0 by checking the peak of $P(n)$ as follows:

$$P(n) = |IFFT(\tilde{X}_R(k) X_{R,m}(k)|. \tag{11}$$

$$\hat{n}_0 = \arg\max P(n). \tag{12}$$

B. Carrier Frequency Offset Estimation

Here, two CFO estimation methods are proposed for multi-user ranging.

1) Cross-correlation in Time Domain: As the samples of $y(n)$ are received, the carrier frequency offset of the *u*-th ranging user, $\Delta f^{(u)}$ will result in a phase rotation $\phi(\Delta f^{(u)}, n)$

$$\begin{aligned} \phi(\Delta f^{(u)}, n) &= 2\pi \Delta f^{(u)} n / N \Delta f = 2\pi \xi^{(u)} n / N, \\ \phi(\xi^{(u)}, n) &= \phi(\Delta f^{(u)}, n), \end{aligned} \tag{13}$$

where $\Delta f^{(u)}$ is the CFO between the *u*-th ranging user and the uplink receiver, Δf is the subcarrier spacing, and $\xi^{(u)}$ is defined as the *normalized CFO* of the *u*-th ranging user.

The phase rotation difference between two received samples $y(n)$ and $y(n+1)$ is a function of the normalized CFO and their time delay. As the STOs of all RSSs have been estimated, there is no timing offset between the received samples and the time-domain ranging signal. For the *u*-th ranging user, the phase rotation difference between $y(d_{R,u} + n + N_{\text{CP}})$ and $x_{R,u}(n)$ can be written as

$$\begin{aligned} \Delta\phi_n &= \phi(\xi^{(k)}, d_{R,u} + n + N_{\text{CP}}) - \phi(0, n) \\ &= \phi(\xi^{(k)}, d_{R,u} + n + N_{\text{CP}}) - 0, \end{aligned} \tag{14}$$

where $n = 0, 1, ..., N-1$.

Assuming there are no other significant phase distortion effects, the phase rotation difference can be used to determine the normalized CFO $\xi^{(u)}$.

The phase of $R_u(d_{R,u})$ (Eq. 7) represents the sum of all phase rotation difference between the received samples $y(d_{R,u} + n + N_{\text{CP}})$ and the time-domain ranging signal $x_{R,u}(n)$. We have

$$\begin{aligned} \angle R_u(d_{R,u}) &= \angle \left(\sum_{n=0}^{N-1} x_{R,u} y^*(d_{R,u} + n + N_{\text{CP}}) \right) \\ &= \angle \left(\sum_{n=0}^{N-1} exp(-\Delta\phi_n) \right) \\ &= \angle \left(\sum_{n=0}^{N-1} exp(\frac{-j2\pi(d_{R,u} + n + N_{\text{CP}})\xi^{(u)}}{N}) \right) \\ &= \angle \left(exp(\frac{-j\pi\eta\xi^{(u)}}{N}) \cdot \frac{\sin(\pi\xi^{(u)})}{\sin(\pi\xi^{(u)}/N)} \right), \end{aligned} \tag{15}$$

where $n = 0, 1, ..., N-1$, $\eta = 2(d_{R,u} + N_{\text{CP}}) + N - 1$. Then the normalized CFO of the *u*-th ranging user, $\xi^{(u)}$ can be estimated by the following equation:

$$\xi^{(u)} = \frac{\angle R_u(d_{R,u}) \cdot N}{\eta \cdot \pi}. \tag{16}$$

2) Cross-correlation in Frequency Domain: Because the second ranging symbol is created by repeating the first ranging symbol, as shown in Fig. 1, after the removal of CP, the received ranging signal has the following characteristics:

$$y_R^{(u)}(n+N) = y_R^{(u)}(n) e^{j2\pi\xi^{(u)}}, \tag{17}$$

where $\xi^{(u)}$ is the normalized CFO of the *u*-th ranging user.

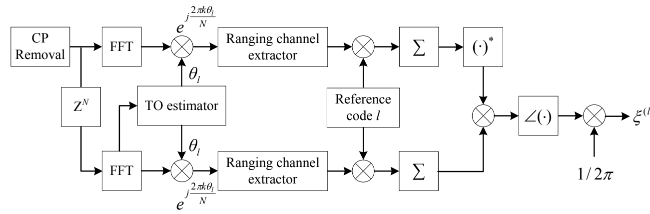

Fig. 3. Uplink frequency offset estimator

Then the received signal at BS can be rewritten as

$$\begin{aligned} y(n+N) = &\sum_{u=0}^{N_R-1} y_R^{(u)}(n) e^{j2\pi\xi^{(u)}} \\ &+ \sum_{g=0}^{N_D-1} y_D^{(g)}(n+N) + z(n+N). \end{aligned} \tag{18}$$

When converted into frequency domain by FFT processing, the first OFDMA symbol will become

$$\begin{aligned} Y_q(k) &= FFT\,(y(n)) \\ &= \sum_{u=0}^{N_R-1} Y_{R,q}^{(u)}(k) + \sum_{g=0}^{N_D-1} Y_{D,q}^{(g)}(k) + Z_q(k). \end{aligned} \tag{19}$$

where $Y_{R,q}^{(u)}(k)$ is the received ranging signal in the first OFDMA symbol from the u-th ranging user in frequency domain, and $Y_{D,q}^{(g)}(k)$ denotes the received signal in the first OFDMA symbol from the g-th data user in frequency domain.

Then the second OFDMA symbol will become

$$\begin{aligned} Y_{q+1}(k) &= FFT\,(y(n+N)) \\ &= \sum_{u=0}^{N_R-1} e^{j2\pi\xi^{(u)}} Y_{R,q}^{(u)}(k) + \sum_{g=0}^{N_D-1} Y_{D,q+1}^{(g)}(k) + Z_{q+1}(k). \end{aligned} \tag{20}$$

Assume $d_{R,u}$ is the timing offset between the received signal from *u*-th user and uplink receiver, and $\tilde{Y}$ is the signal without timing offset to local receiver. According to (10), (19) and (20) could be further described with the consideration of timing offset as

$$\begin{aligned} Y_q(k) = &\sum_{u=0}^{N_R-1} e^{\frac{-j2\pi k d_{R,u}}{N}} \tilde{Y}_{R,q}^{(u)}(k) \\ &+ \sum_{g=0}^{N_D-1} Y_{D,q}^{(g)}(k) + Z_q(k). \end{aligned} \tag{21}$$

$$\begin{aligned} Y_{q+1}(k) = &\sum_{u=0}^{N_R-1} e^{j2\pi\xi^{(u)}} e^{\frac{-j2\pi k d_{R,u}}{N}} \tilde{Y}_{R,q}^{(u)}(k) \\ &+ \sum_{g=0}^{N_D-1} Y_{D,q+1}^{(g)}(k) + Z_{q+1}(k). \end{aligned} \tag{22}$$

To elimate the impact of timing offset on the *i*-th user, we will process (21) and (22) as

$$\begin{aligned} Y_q^{'}(k) &= e^{\frac{j2\pi k d_{R,i}}{N}} Y_q(k) \\ &= \tilde{Y}_{R,q}^{(i)}(k) + \sum_{\substack{u=0\\u\neq i}}^{N_R-1} e^{\frac{j2\pi k(d_{R,i}-d_{R,u})}{N}} \tilde{Y}_{R,q}^{(u)}(k) \\ &+ \sum_{g=0}^{N_D-1} e^{\frac{j2\pi k d_{R,i}}{N}} Y_{D,q}^{(g)}(k) + e^{\frac{j2\pi k d_{R,i}}{N}} Z_q(k). \end{aligned} \tag{23}$$

$$\begin{aligned} Y_{q+1}^{'}(k) &= e^{\frac{j2\pi k d_{R,i}}{N}} Y_{q+1}(k) \\ &= e^{j2\pi\xi^{(i)}} \tilde{Y}_{R,q}^{(i)}(k) + \sum_{\substack{u=0\\u\neq i}}^{N_R-1} e^{j2\pi\xi^{(u)}} e^{\frac{j2\pi k(d_{R,i}-d_{R,u})}{N}} \tilde{Y}_{R,q}^{(u)}(k) \\ &+ \sum_{g=0}^{N_D-1} e^{\frac{j2\pi k d_{R,i}}{N}} Y_{D,q+1}^{(g)}(k) + e^{\frac{j2\pi k d_{R,i}}{N}} Z_{q+1}(k). \end{aligned} \tag{24}$$

To extract the signal of the *i*-th user in frequency domain, we use the local reference ranging code $\mathbf{X}_{R,i} = [X_{R,i}(0)\ \ X_{R,i}(0)\ \ ...\ \ X_{R,i}(N-1)]^T$ for frequency-domian correlation. Considering $X_{R,i}(k)=0$ at data carriers, the two OFDMA symbols will become

$$\begin{aligned} &\left[\mathbf{E}(d_{R,i})\mathbf{Y}_q\right]^T \mathbf{X}_{R,i} \\ &= \left(\tilde{\mathbf{Y}}_{R,q}^{(i)}\right)^T \mathbf{X}_{R,i} + \sum_{\substack{u=0\\u\neq i}}^{N_R-1} \left[\mathbf{E}(d_{R,i}-d_{R,u})\tilde{\mathbf{Y}}_{R,q}^{(u)}\right]^T \mathbf{X}_{R,i} \\ &+ \sum_{g=0}^{N_D-1} \left[\mathbf{E}(d_{R,i})\mathbf{Y}_{D,q}^{(g)}\right]^T \mathbf{X}_{R,i} + \left[\mathbf{E}(d_{R,i})\mathbf{Z}_q\right]^T \mathbf{X}_{R,i}, \end{aligned} \tag{25}$$

$$\begin{aligned} &\left[\mathbf{E}(d_{R,i})\mathbf{Y}_{q+1}\right]^T \mathbf{X}_{R,i} \\ &= e^{j2\pi\xi^{(i)}} \left(\tilde{\mathbf{Y}}_{R,q}^{(i)}\right)^T \mathbf{X}_{R,i} + \sum_{\substack{u=0\\u\neq i}}^{N_R-1} e^{j2\pi\xi^{(u)}} \left[\mathbf{E}(d_{R,i}-d_{R,u})\tilde{\mathbf{Y}}_{R,q}^{(u)}\right]^T \mathbf{X}_{R,i} \\ &+ \sum_{g=0}^{N_D-1} \left[\mathbf{E}(d_{R,i})\mathbf{Y}_{D,q+1}^{(g)}\right]^T \mathbf{X}_{R,i} + \left[\mathbf{E}(d_{R,i})\mathbf{Z}_{q+1}\right]^T \mathbf{X}_{R,i}, \end{aligned} \tag{26}$$

where $\mathbf{E}(d_{R,i}) = diag\left(1, e^{\frac{j2\pi d_{R,i}}{N}}, ..., e^{\frac{j2\pi(N-1)d_{R,i}}{N}}\right)$.

In (25) and (26), considering the cross-correlation property of different ranging codes, and ignoring the impact of Gaussion noise, the estimated $\xi^{(i)}$ could be derived by

$$\xi^{(i)} = \arg\left(\frac{\left[\mathbf{E}(d_{R,i})\mathbf{Y}_{q+1}\right]^T \mathbf{X}_{R,i}}{\left[\mathbf{E}(d_{R,i})\mathbf{Y}_q\right]^T \mathbf{X}_{R,i}}\right)/2\pi. \tag{27}$$

$\mathbf{E}(d_{R,i})$ could be regarded as the sequence to do the timing offset correction for the *i*-th user in frequency domain. Timing offset will generate a heavy impact on CFO estimation. Here we will provide a rough estimation on its impact. In (25), considering the ratio

$$P_c = \frac{\left(\tilde{\mathbf{Y}}_{R,q}^{(i)}\right)^T \mathbf{X}_{R,i}}{\sum\limits_{u=0,u\neq i}^{N_R-1} \left[\mathbf{E}(d_{R,i}-d_{R,u})\tilde{\mathbf{Y}}_{R,q}^{(u)}\right]^T \mathbf{X}_{R,i}} \tag{28}$$

will be large because of the auto-correlation and cross-correlation of PN sequence, we could remove the second item on the right side of the equation. But if there isn't timing offset correction, the ratio of (28) will become

$$\begin{aligned} P_c^{'} &= \frac{\left(\mathbf{Y}_{R,q}^{(i)}\right)^T \mathbf{X}_{R,i}}{\sum\limits_{u=0,u\neq i}^{N_R-1} \left(\mathbf{Y}_{R,q}^{(u)}\right)^T \mathbf{X}_{R,i}} \\ &= \frac{\left[\mathbf{E}(-d_{R,i})\tilde{\mathbf{Y}}_{R,q}^{(i)}\right]^T \mathbf{X}_{R,i}}{\sum\limits_{u=0,u\neq i}^{N_R-1} \left[\mathbf{E}(-d_{R,u})\tilde{\mathbf{Y}}_{R,q}^{(u)}\right]^T \mathbf{X}_{R,i}} \end{aligned} \tag{29}$$

To give a simple estimation, we ignore the CIR of the i-th user here. So

$$\left(\tilde{\mathbf{Y}}_{R,q}^{(i)}\right)^T \mathbf{X}_{R,i} \approx \sum_{n=0}^{R-1} 1 = R. \tag{30}$$

where *R* is the length of ranging codes. Substituting (30) into (28) and (29) respectively, there will be

$$\begin{aligned} &\frac{P_c}{P_c^{'}} \\ &= R \cdot \frac{1-e^{\frac{-j2\pi d_{R,i}}{N}}}{1-e^{\frac{-j2\pi d_{R,i}R}{N}}} \cdot \frac{\sum\limits_{\substack{u=0\\u\neq i}}^{N_R-1} \left[\mathbf{E}(-d_{R,u})\tilde{\mathbf{Y}}_{R,q}^{(u)}\right]^T \mathbf{X}_{R,i}}{\sum\limits_{\substack{u=0\\u\neq i}}^{N_R-1} \left[\mathbf{E}(d_{R,i}-d_{R,u})\tilde{\mathbf{Y}}_{R,q}^{(u)}\right]^T \mathbf{X}_{R,i}}. \end{aligned} \tag{31}$$

Considering both $(d_{R,i} - d_{R,u})$ and $(-d_{R,i})$ are random number with same distribution, for rough estimation, (31) could be further simplified as

$$\frac{P_c}{P_c^{'}} = R \cdot \frac{1-e^{\frac{-j2\pi d_{R,i}}{N}}}{1-e^{\frac{-j2\pi d_{R,i}R}{N}}}. \tag{32}$$

(32) could be explained as, without the timing offset correction, the signal of ranging user *i* will be decreased to $\left\|\frac{1-e^{-j2\pi d_{R,i}R/N}}{\left(1-e^{-j2\pi d_{R,i}/N}\right)R}\right\|$ of the original signal. For example, when *N*=1024, *R*=144 and $d_{R,u}$ is 20, $\left\|\frac{1-e^{-j2\pi d_{R,i}R/N}}{\left(1-e^{-j2\pi d_{R,i}/N}\right)R}\right\| = 0.0934$. So, timing offset correction is very important in CFO estimation.

In STO and CFO estimation, both cross-correlation methods in time and frequency domain are analyzed. Here, the cross correlation process in frequency domain could provide two advantages. One is that, there will be no interference from synchronized users signal to the ranging channel, because they

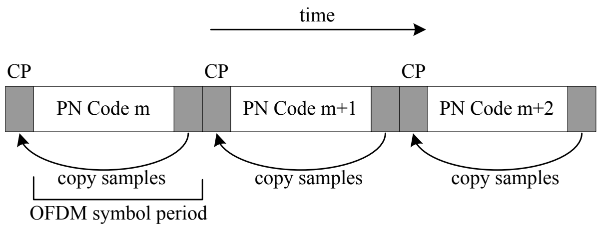

Fig. 4. Periodic ranging transmission over three consecutive OFDMA symbols

are orthogonal in frequency domain. The other one is, PN codes were originally sent over frequency and thus correlating in time domain will impair the cross correlation properties between different codes [10]. Moreover, for CFO estimation, it will be more easy to do the STO compensation in frequency domain than in time domain for the received signal from multi-user.

IV. System Implementation for WiMAX

A. Synchronization mechanism in WiMAX

As introduced in Section I, WiMAX standard defined four kinds of ranging. Initial ranging and HO ranging will share the same ranging symbol structure as shown in Fig.1. But in periodic ranging and BR ranging, the SS can send a transmission in one of the following ways:

a). Modulating one ranging code on the ranging subchannel for a period of one OFDMA symbol.

b). Modulating three consecutive ranging codes (starting code shall always be a multiple of 3) on the ranging subchannel for a period of three OFDMA symbols (one code per symbol).

To perform the frequency offset estimation, we need more than one consecutive symbols in algorithms. So the system will select b) for transmission, as shown in Fig. 4. er.

WiMAX PHY specifies a ranging subchannel and a set of special pseudonoise ranging codes. Subsets of codes shall be allocated for different ranging requests among initial, HO, periodic and BR ranging, so that the BS can determine the purpose of the received code by the subset to which the code belongs.

B. System architecture for joint STO-CFO estimation

In this paper, we propose the joint STO-CFO estimation method, where the STO estimation with cross-correlation in frequency domain and our proposed CFO estimation with cross-correlation in frequency domain will be used jointly.

We hope to use the same ranging processing module for both initial and periodic ranging, but their signal structures are different. So it's necessary to analyze if the algorithms for initial ranging could be reused by periodic ranging. Then based on the analyzed result, the system architecture of the ranging module designed for WiMAX will be introduced.

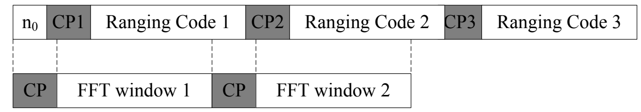

Fig. 5. FFT windows for periodic ranging

1) Algorithm analysis for periodic ranging: In both STO and CFO estimation, FFT is adopted over the received data. With the possible delay on the received data, we will have the FFT windows as shown in Fig. 5. Because CP length $N_{\mathrm{CP}} > d_{max}$, FFT window 1 will not start before CP 1 of the received data. That means, the first FFT window will cover all content of ranging code 1, with a cyclic shifting. Because only one FFT window or one OFDM symbol will be required for STO estimation, the proposed STO estimation method could be used for both initial ranging and periodic ranging.

In CFO estimation, two points should be considered for periodic ranging. One is the ranging codes used in the successive OFDM symbols are different ($X_{R,i,q} \neq X_{R,i,q+1}$). The other one is, there is one CP of the second symbol sitting between the two OFDM symbols which will be used for CFO estimation. Considering these two differences, (25) and (26) could be rewritten as

$$\begin{aligned}&[\mathbf{E}(d_{\mathrm{R},i})\mathbf{Y}_q]^T\mathbf{X}_{\mathrm{R},i,q}\\&=\left(\tilde{\mathbf{Y}}_{\mathrm{R},q}^{(i)}\right)^T\mathbf{X}_{\mathrm{R},i,q}+\sum_{\substack{u=0\\u\neq i}}^{N_R-1}\Big[\mathbf{E}\left(d_{\mathrm{R},i}-d_{\mathrm{R},u}\right)\tilde{\mathbf{Y}}_{\mathrm{R},q}^{(u)}\Big]^T\mathbf{X}_{\mathrm{R},i,q}\\&+\sum_{g=0}^{N_D-1}\Big[\mathbf{E}(d_{R,i})\mathbf{Y}_{D,q}^{(g)}\Big]^T\mathbf{X}_{\mathrm{R},i,q}+\left[\mathbf{E}\left(d_{\mathrm{R},i}\right)\mathbf{Z}_q\right]^T\mathbf{X}_{\mathrm{R},i,q}\end{aligned} \tag{33}$$

and

$$\begin{aligned}&[\mathbf{E}(d_{\mathrm{R},i})\mathbf{Y}_{q+1}]^T\mathbf{X}_{\mathrm{R},i,q+1}\\&=e^{j\frac{2\pi\xi^{(i)}(N+N_{\mathrm{CP}})}{N}}\left(\tilde{\mathbf{Y}}_{\mathrm{R},q+1}^{(i)}\right)^T\mathbf{X}_{\mathrm{R},i,q+1}\\&+\sum_{\substack{u=0\\u\neq i}}^{N_R-1}\Big[\mathbf{E}\left(d_{\mathrm{R},i}-d_{\mathrm{R},u}\right)\tilde{\mathbf{Y}}_{\mathrm{R},q+1}^{(u)}\Big]^T\mathbf{X}_{\mathrm{R},i,q+1}\\&+\sum_{g=0}^{N_D-1}\Big[\mathbf{E}(d_{R,i})\mathbf{Y}_{D,q+1}^{(g)}\Big]^T\mathbf{X}_{\mathrm{R},i,q+1}+[\mathbf{E}(d_{\mathrm{R},i})\mathbf{Z}_{q+1}]^T\mathbf{X}_{\mathrm{R},i,q+1}.\end{aligned} \tag{34}$$

Similarly, the CFO offset could be estimated by

$$\begin{aligned}\xi^{(i)}=\arg&\left(\frac{\left[\mathbf{E}\left(d_{\mathrm{R},i}\right)\mathbf{Y}_{q+1}\right]^T\mathbf{X}_{\mathrm{R},i,q+1}}{\left[\mathbf{E}\left(d_{\mathrm{R},i}\right)\mathbf{Y}_{q}\right]^T\mathbf{X}_{\mathrm{R},i,q}}\cdot\frac{\left(\tilde{\mathbf{Y}}_{\mathrm{R},q}^{(i)}\right)^T\mathbf{X}_{\mathrm{R},i,q}}{\left(\tilde{\mathbf{Y}}_{\mathrm{R},q+1}^{(i)}\right)^T\mathbf{X}_{\mathrm{R},i,q+1}}\right)\\&\cdot\frac{N}{2\pi(N+N_{\mathrm{CP}})}.\end{aligned} \tag{35}$$

Because $\tilde{\mathbf{Y}}_{\mathrm{R},q}^{(i)}$ is the signal without time offset to local receiver, to simplify the analysis, its element could be written as

$$
\begin{aligned}
\tilde{Y}_{\mathrm{R},q}^{(i)}(k) &= \frac{1}{N}\sum_{n=0}^{N-1}\tilde{y}_{\mathrm{R},q}^{(i)}(n)e^{-j2\pi kn/N} \\
&= \frac{1}{N}\sum_{n=0}^{N-1}\tilde{x}_{\mathrm{R},i,q}(n)e^{-j2\pi(k-\xi^{(i)})n/N} \\
&= X_{\mathrm{R},i,q}(k)H_{\mathrm{R},i,q}(k)\frac{1-e^{j2\pi\xi^{(i)}}}{1-e^{j2\pi\xi^{(i)}/N}}.
\end{aligned} \tag{36}
$$

where $H_{\mathrm{R},i,q}(k)$ denotes the channel frequency response on the k-th subcarrier of the i-th ranging user's channel during the q-th OFDMA block.

From (36) we could get the result that, if CFO $\xi^{(i)} < \Delta f$ which is the subcarrier spacing, the frequency shifting will result in amplitude change on both real and imaginary data. But this amplitude change will be equal for each sampled data in frequency domain. So with the result in (36), the CFO estimated in (35) will become

$$
\begin{aligned}
\xi^{(i)} =& \frac{N}{2\pi(N+N_{\mathrm{CP}})}\cdot\arg\left(\frac{\left[\mathbf{E}\left(d_{\mathrm{R},i}\right)\mathbf{Y}_{q+1}\right]^T\mathbf{X}_{\mathrm{R},i,q+1}}{\left[\mathbf{E}\left(d_{\mathrm{R},i}\right)\mathbf{Y}_{q}\right]^T\mathbf{X}_{\mathrm{R},i,q}}\right. \\
&\left.\cdot\frac{\frac{1-e^{j2\pi\xi^{(i)}}}{1-e^{j2\pi\xi^{(i)}/N}}(\mathbf{H}_{\mathrm{R},i,q}\mathbf{X}_{\mathrm{R},i,q})^T\mathbf{X}_{\mathrm{R},i,q}}{\frac{1-e^{j2\pi\xi^{(i)}}}{1-e^{j2\pi\xi^{(i)}/N}}(\mathbf{H}_{\mathrm{R},i,q+1}\mathbf{X}_{\mathrm{R},i,q+1})^T\mathbf{X}_{\mathrm{R},i,q+1}}\right) \\
=& \frac{N}{2\pi(N+N_{\mathrm{CP}})}\cdot\arg\left(\frac{\left[\mathbf{E}\left(d_{\mathrm{R},i}\right)\mathbf{Y}_{q+1}\right]^T\mathbf{X}_{\mathrm{R},i,q+1}}{\left[\mathbf{E}\left(d_{\mathrm{R},i}\right)\mathbf{Y}_{q}\right]^T\mathbf{X}_{\mathrm{R},i,q}}\right. \\
&\left.\cdot\frac{\sum_{k=0}^{N-1}H_{\mathrm{R},i,q}(k)}{\sum_{k=0}^{N-1}H_{\mathrm{R},i,q+1}(k)}\right)
\end{aligned} \tag{37}
$$

where $\mathbf{H}_{\mathrm{R},i,q} \triangleq diag(H_{\mathrm{R},i,q}(0),\cdots,H_{\mathrm{R},i,q}(N-1))$. Now we consider the case that the neighboring OFDMA symbols have the same channel frequency response for all subcarriers, *i.e*, $H_{\mathrm{R},i,q}(k)=H_{\mathrm{R},i,q+1}(k), k=0,1,\cdots,N-1$. Then (37) can be rewritten as

$$
\xi^{(i)} = \frac{N}{2\pi(N+N_{\mathrm{CP}})}\cdot\arg\left(\frac{\left[\mathbf{E}\left(d_{\mathrm{R},i}\right)\mathbf{Y}_{q+1}\right]^T\mathbf{X}_{\mathrm{R},i,q+1}}{\left[\mathbf{E}\left(d_{\mathrm{R},i}\right)\mathbf{Y}_{q}\right]^T\mathbf{X}_{\mathrm{R},i,q}}\right). \tag{38}
$$

Comparing the method for initial ranging in (27) and that for periodic ranging in (38), there will be only minor differences in processing.

- For periodic ranging channel, the CP between the first and second OFDM symbol should be removed.
- In CFO estimation for periodic ranging, different ranging codes for the two consecutive symbols will be used when correlation in frequency domain.
- In CFO estimation for periodic ranging, to calculate the normalized CFO $\xi^{(i)}$, the factor $\frac{N}{2pi(N+N_{\mathrm{CP}})}$ should be used instead of $1/2\pi$

So, this proposed CFO estimation method with cross-correlation in frequency domain could be used in both initial ranging and periodic ranging scenario.

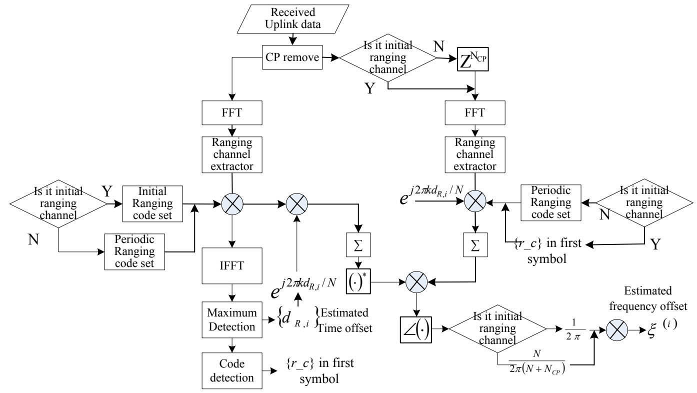

Fig. 6. Ranging module architecture

2) Ranging Module Architecture: The joint STO-CFO estimation method for WiMAX system is proposed. Here, the word "joint" lies in two aspects:

- The result of the estimated STO will be used in CFO estimation. Its importance has been analyzed in session III.
- It has been recognized that, time and frequency offset estimation will require large computation resources. So in our well designed architecture, some components and intermediate results will be used by STO and CFO estimation jointly. Therefore, lots of computation cycles could be saved.

With the design rule of high efficiency and compliance on WiMAX standard framework, a ranging module architecture is proposed for joint STO-CFO estimation in WiMAX base station. The input of the ranging module is the received uplink data from multiple subscribers, which will be the mix of data users and ranging users. And the estimated STO and CFO are the output ranging information which should be broadcasted in downlink for subscribers' adjustment.

C. Implementation Analysis

With the proposed architecture in Fig. 6, the complexity will be analyzed first, then we will introduce its required resources on our Cell-based Software Defined Radio (SDR) Wimax platform [12].

1) Complexity Analysis: Assuming the cycles for multiplication, addition and comparison for the real number are M_{RM}, M_{RA} and M_{RC} respectively, and those for complex number are M_{CM}, M_{CA} and M_{CC}, The required complexity for each sub-component is listed below.

- FFT: $M_{\mathrm{FFT}} = (\frac{N_{\mathrm{FFT}}}{2}\log_2 N_{\mathrm{FFT}})M_{CM} + (N_{\mathrm{FFT}}\log_2 N_{\mathrm{FFT}})M_{CA}$
- Dot product with ranging code: $M_{product\_RC} = 2RM_{RM}$
- Time offset correction: $M_{T_correct} = N_{\mathrm{FFT}}M_{CM}$
- $\Sigma$for correlation: $M_{\Sigma} = (N_{\mathrm{FFT}}-1)M_{CA}$
- Maximum and code detection: $N_{detect} = (N_{\mathrm{FFT}} - 1)M_{CC} + M_{RC}$. Only the simple code detection method is used. For example, the threshold will be set for comparison.

- arc tangent: $M_{\texttt{atan}} \approx 3.3M_{RA}$ (statistic result from real test).

Here, $N_{\texttt{FFT}}$ is the length of FFT, and *R* is the length of ranging code. Assuming the sizes of the code set for initial ranging and periodic ranging are N_{IR_C} and N_{PR_C}, and the numbers of detected initial and periodic ranging user are N_{IR_u} and N_{PR_u}, the complexity of joint STO-CFO estimation for initial ranging will be

$$\begin{aligned} & M_{IR}(N_{IR\_C}, N_{IR\_u}) \\ =& 2M_{\texttt{FFT}} + N_{IR\_C}(M_{product\_RC} + M_{\texttt{FFT}} + M_{detect}) \\ & + N_{IR\_u}(M_{T\_correct} + M_{product\_RC} + M_{\Sigma} + M_{\texttt{atan}} + M_{RM}) \\ =& [\frac{1}{2}N_{\texttt{FFT}}(N_{IR\_C}+2)\log_2 N_{\texttt{FFT}} + N_{IR\_u}N_{\texttt{FFT}}]M_{CM} \\ & + [N_{\texttt{FFT}}(N_{IR\_C}+2)\log_2 N_{FFT} + (N_{FFT}-1)N_{IR\_u}]M_{CA} \\ & + N_{IR\_C}(N_{\texttt{FFT}}-1)M_{CC} + [2R(N_{IR\_C}+N_{IR\_u}) + N_{IR\_u}]M_{RM} \\ & + N_{IR\_C}M_{RC} + 3.3N_{IR\_u}M_{RA}. \end{aligned} \quad (39)$$

And the complexity of joint STO-CFO estimation for periodic ranging will be

$$\begin{aligned} & M_{PR}(N_{PR\_C}, N_{PR\_u}) \\ =& 2M_{\texttt{FFT}} + N_{PR\_C}(2M_{product\_RC} + M_{\texttt{FFT}} + M_{detect}) \\ & + N_{PR\_u}(M_{T\_correct} + M_{\Sigma} + M_{\texttt{atan}} + M_{RM}) \\ =& [\frac{1}{2}N_{\texttt{FFT}}(N_{PR\_C}+2)\log_2 N_{\texttt{FFT}} + N_{PR\_u}N_{\texttt{FFT}}]M_{CM} \\ & + [N_{\texttt{FFT}}(N_{PR\_C}+2)\log_2 N_{FFT} + (N_{FFT}-1)N_{PR\_u}]M_{CA} \\ & + N_{PR\_C}(N_{\texttt{FFT}}-1)M_{CC} + (4RN_{PR\_C} + N_{PR\_u})M_{RM} \\ & + N_{PR\_C}M_{RC} + 3.3N_{PR\_u}M_{RA}. \end{aligned} \quad (40)$$

Different processor architectures will need different numbers of instructions for multiplication, addition and comparison for complex and real. For example, in general purpose processor with floating point processing unit (FPU), usually there will be following mapping for instruction number to those scalar processing

Processing	M_{CM}	M_{CA}	M_{CC}	M_{RM}	M_{RA}	M_{RC}
Instruction num	6	2	7	1	1	1

We consider the situation when $N_{\texttt{FFT}} = 1024$ and $R = 144$. The complexity of our proposed architecture for initial ranging will be

$$M_{IR}(N_{IR\_C}, N_{IR\_u}) = 84823N_{IR\_u} + 58650N_{IR\_C} + 102400 \quad (41)$$

and that for periodic ranging will be

$$M_{PR}(N_{PR\_C}, N_{PR\_u}) = 81943N_{PR\_u} + 58939N_{PR\_C} + 102400 \quad (42)$$

In the complexity analysis for STO estimation from [8], to get the best STO estimation, the phase shifting from $[0, N-1]$ for the R ranging subcarriers is used, rather than adopting the IFFT method. So the complexity is much higher than our proposed architecture.

Software radio technology is widely used in wireless system, where the programable platform, e.g. DSP and general purpose processor will be used to carry all the base band processing. The above complexity result could be referenced for system design. But it should be noted that, these numbers only contain the computation part. There will be some overhead for looping, table lookup and data movement etc. Moreover, the factor of Cycle Per Instruction (CPI) of different platform will take an important role in performance evaluation. So when considering how to allocate the computation resources for components in base band processing, we should add some more buffer over these estimated complexity.

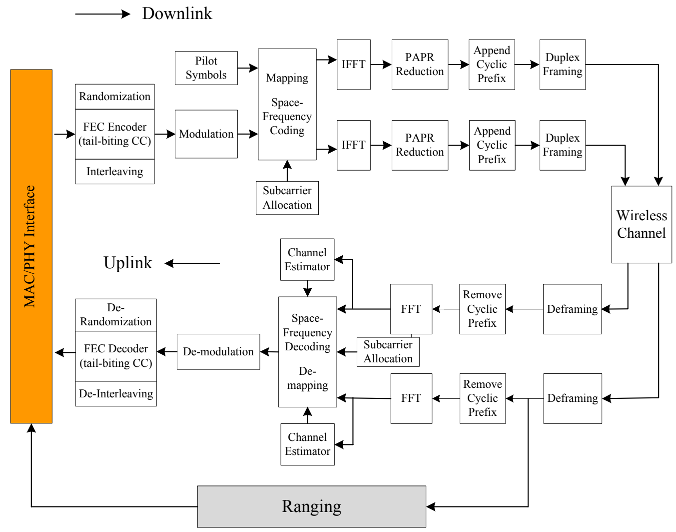

Fig. 7. WiMAX PHY layer architecture with ranging module

2) System implementation on Cell-based SDR WiMAX platform: To explore the capability of general multicore processor on wireless base station system, we developed a general WiMAX base station architecture (IEEE 802.16e OFDMA-mode) on IBM Cell blade server [12]. Here Cell is a multicore processor jointly developed by IBM, SONY and Toshiba. The framework of the physical layer implementation is shown in Fig. 7 [13]. The ranging for timing and frequency offset estimation is part of this framework. The ranging module is optimized on Cell SPU with 128-bit SIMD instruction. We only test the initial ranging as example. When $N_{IR\_C} = 125$ and $N_{IR\_u} = 8$ at one ranging time slot, it will need $4.089 * 10^6$ instructions on Cell SPU. Comparing with the estimated result $7.502 * 10^6$ got from (41), the main reason for the gap between these two numbers is, we use SIMD instructions to parallelize the data processing on Cell. On the other hand, we will need more instructions for data movement, loops, table lookup, etc. which hasn't been taken account in the estimated complexity. But more discussion about it will be out of scope of this paper.

The working frequency of Cell SPU is 3.2GHz. Considering the average CPI for this workload is 1.4 on SPU, it will require $9.75 * 10^6$ cycles to complete the whole ranging module with the maximum size of ranging code 256. There are 2 Cell processors with 16 SPUs on each Cell Blade. If 3 SPUs are allocated for ranging module, they could support the worst case in every 1ms.

V. Simulation Results

A. Simulation Parameters

In the simulation, the OFDMA system parameters are selected from [1]. The uplink bandwidth is 3MHz, the subcarrier frequency spacing Δf is 3.28kHz, N=1024. We use QPSK format for DSS. Within the 1024 subcarriers, there are 92 guard subcarriers on the left-side band and 91 on the right and 1 DC subcarrier residing on index 512. The remaining 840 subcarriers are partitioned into Q=35 subchannels. The ranging channel is composed of six adjacent subchannels and spanning 144 subcarriers per OFDMA symbol. SUI-3 channel model with 3 paths [11] is considered in our simulation. The number of sample-spaced channel taps, L, is set to 4. Channels of different users are generated independently. We consider a cell radius of 5km which gives the maximum transmission delay (round trip) $d_{max,R} \approx 34\mu s = 114$ samples. N_{CP} is set to 128 samples satisfiying the condition $d_{max,R} < N_{\text{CP}} - L$. The timing requirement based on [1] is that all uplink OFDM symbols should arrive at the BS within an accuracy of $\pm 25\%$ of the minimum guard-interval or better. In [1], N_{CP} can be 1/4, 1/8, 1/16, or 1/32 of N, and hence, the timing offset should be within ± 8 samples. So the $d_{max,D}$ equals to 8 samples in our case. Similarly, the frequency offset should be within a tolerance of maximum 2% of the subcarrier spacing.

We assume that $N_c = 128$ ranging codes are assigned for initial ranging and the maximum number of RSSs in one ranging time-slot, K, is set to 15. Because six subchannels are allocated to ranging subchannel, the maximum number of DSSs accessed simultaneously, G, is 29.

In each simulation run, the timing offset for RSSs and DSSs are generated randomly from the interval $[0, d_{max,R}]$ and $[0, d_{max,D}]$, respectively. The normalized CFOs of RSSs and DSSs uniformly distribute in the range of [-0.2, 0.2] and [-0.02, 0.02]. In order to evaluate the effect of data users on the performance of proposed algorithms, we consider the following conditions of 0 DSS, 15 DSSs, and 29 DSSs in one ranging time-slot.

B. Timing Estimation Performance

First, we compare the effectiveness of the two discussed STO-estimation algorithm. Simulations are performed in the presence of SNR = 20dB, a normalized CFO of 0.2. The numbers of RSSs and DSSs are K=2 and G=0, respectively.

Fig. 8 shows the root mean square error (RMSE) performance of the timing offset estimated as a function of SNR. The number of RSSs is K=1. The performances of all timing estimators improve as the SNR increases and degrade with the increasing number of DSSs. It is also presented that the cross-correlation method in frequency domain performs better than the time-domain correlation method.

Fig. 9 depicts the RMSE performance of the timing offset estimate versus the number of RSSs. We assume that all RSSs and DSSs exhibit the same SNR of 20dB. The performances of all methods degrade as the numbers of RSSs and DSSs increase but the proposed method gives a much better performance and satisfies the timing requirement in [1], especially in multiple RSSs condition, such as 15 simultaneous RSSs in one ranging time-slot. However, the time-domain correlation method satisfies the requirement only when the numbers of RSSs and DSSs are small, such as $K \leq 2$ and $G < 15$.

C. Frequency Estimation Performance

Fig. 10 shows the MSE performance of the frequency offset estimated as a function of SNR. The number of RSSs is K=1. We observe that although the performances of all CFO estimators degrade as the number of DSSs increases, the proposed method sill maintain good MSE performance and is slightly better than the method in [8].

Fig. 11 depicts the MSE performance of the frequency offset estimate versus the number of RSSs. Since the method in [8] can not work in the case of multiple users simultaneously existing in one ranging time-slot, only the proposed method's performances in the absence/presence of timing offset are plotted. It is assumed that all SSs have the same SNR of 20dB. The performance degrades as the numbers of RSSs and DSSs increase. But with the increasing number of DSSs, the performance loss is negligible in the condition of multiple RSSs. In the absence of timing offset, the MSE is less than 10^{-4} even in multiple RSSs and DSSs condition, such as $K = 15$ and $G = 29$, which satisfies the frequency offset requirement (2% of the subcarrier spacing) in [1]. By comparing (a) and (b) in Fig. 11, it could be observed that the proposed joint STO-CFO estimation method could provide similar CFO estimation performance under the environment with STO.

VI. Conclusion

We have presented a ranging method of joint STO-CFO estimation in the uplink of WiMAX OFDMA mode. Both the STO and CFO estimation are carried out by cross-correlation in frequency domain. It avoids the interference from synchronized users and keep the cross-correlation properties between different ranging codes. It is an effective multi-user ranging method, which could be used for different kinds for ranging signal structure in WiMAX. Through the joint design, STO could be removed when CFO estimation, and more computation components could be shared for saving the resources. The simulation result shows that the proposed method has better performance than the existing methods. And it is a high efficient design which could be carried on normal multicore platform in software. The observation and some conclusions in this paper will not be limited in WiMax standard. It could be easily extended to other widely used OFDM-based wireless system.

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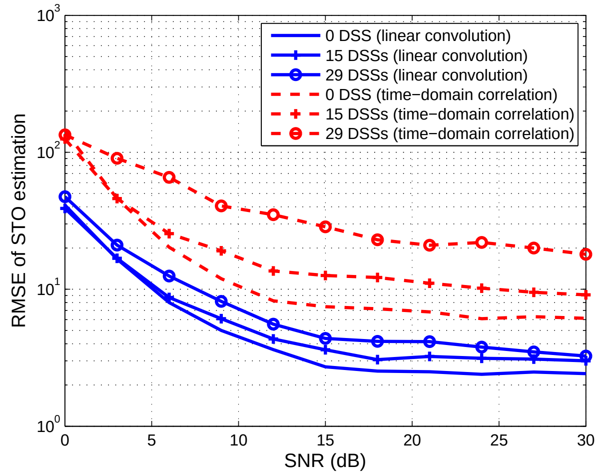

Fig. 8. The RMSE of the STO estimation versus SNR. The number of RSSs is K=1.

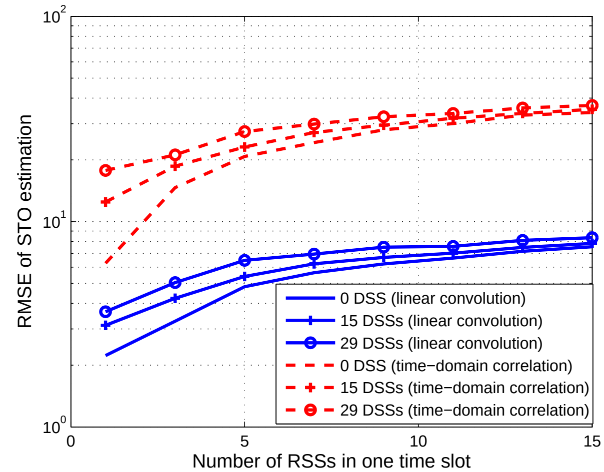

Fig. 9. The RMSE of the STO estimation versus number of RSSs. SNR = 20dB.

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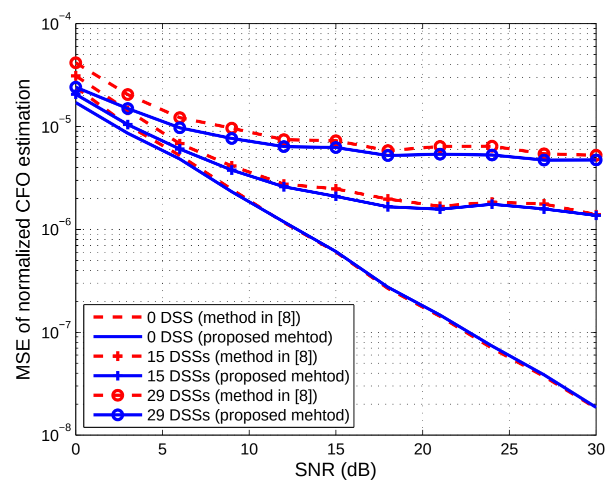

Fig. 10. The MSE of the CFO estimation versus SNR. The number of RSSs is K=1.

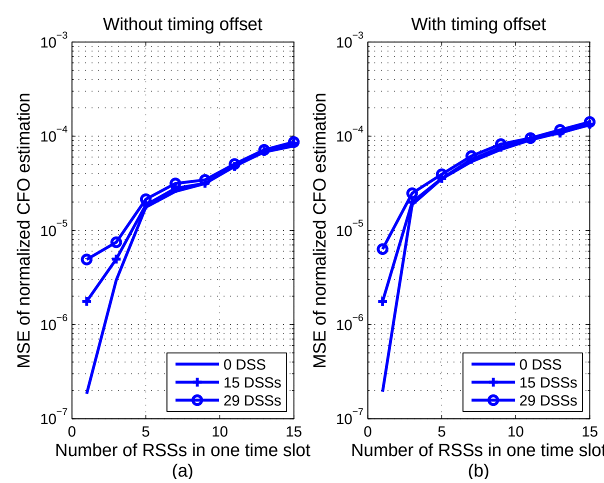

Fig. 11. The MSE of the CFO estimation versus number of RSSs. SNR = 20dB. (a) Without timing offset. (b) With timing offset