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A New Ranging Method for OFDMA Systems

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***Abstract*— We present a new initial ranging method for OFDMA systems such as IEEE 802.16a. First, a new orthogonal ranging signal design is proposed by which an efficient, low-complexity multi-user ranging signal detection is developed. Based on the ranging signal detector results, power estimation for the detected ranging subscriber stations is performed. Then, by utilizing the proposed orthogonal ranging signal design together with the redundancy introduced by cyclic prefixes, a new iterative multi-user timing offset estimator is presented. Compared with the existing methods utilizing the CDMA-type ranging codes (in frequency-domain) defined in IEEE 802.16a, the proposed method achieves a better performance and greater robustness against multi-user interference and multi-path fading channels.**

I. Introduction

In orthogonal frequency division multiple access (OFDMA) systems (e.g., IEEE 802.16a [1]), the issues of uplink sub-carrier orthogonality 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) and periodic ranging (to account for the user-movement over time). This paper considers initial ranging process at the base station (BS) which includes multi-user ranging code detection, multi-user timing estimation, and power estimation.

The ranging signals from IEEE 802.16a [1] use CDMA codes on all the ranging sub-carriers. Based on the ranging channel and ranging signals defined in [1], a ranging method based on a correlator bank in frequency-domain was presented in [2] and a time-domain correlator bank based ranging method was proposed in [3]. The method from [2] detects the multi-user ranging codes by a fixed threshold. The method from [3] uses an adaptive detection threshold and achieves an improved detection performance. In [4], a new ranging signal structure is presented which divides all ranging opportunities into several groups. Ranging signals from different groups do not interfere with each other while within each group they interfere with each other. In this scheme, the number of sub-carriers in each sub-channel has to be equal to the length of the CDMA code and the sub-carriers must be assigned adjacently. Hence, it is not applicable in interleaved OFDMA systems. Furthermore, same as the methods in [2] and [3], its ranging performance will depend on multi-path fading channel effect and the correlation properties of CDMA codes.

In this paper, we present a new ranging signal design and a new ranging method for interleaved OFDMA systems. The proposed ranging signal design is based on the orthogonality principle and the best channel identification conditions. This design results in robust, high performance, low-complexity multi-user ranging signal detection and power estimation. A new iterative estimation of timing offsets for all ranging subscriber stations is developed based on the proposed orthogonal ranging signal design and the cyclic prefix redundancy. If compared to existing methods using the existing ranging signals, our ranging signal design and ranging method achieve a better performance and a greater robustness against interference from other subscriber stations and multi-path fading channel effects.

II. System Description and Signal Model

We consider an uplink of an OFDMA system with N sub-carriers in a time-division duplexing setup. During a ranging time-slot of M symbol intervals, the N sub-carriers are grouped into Q_R ranging sub-channels and Q_D data sub-channels. Each ranging sub-channel has N_{1R} sub-carriers and each data sub-channel has N_{1D} sub-carriers where $N_{1R}Q_R + N_{1D}Q_D \leq N$. In general, the sub-carrier assignment of data and ranging sub-channels can be adaptive according to the system and channel conditions.

The indices of all the sub-carriers dedicated for the ranging channel and for the data subscriber stations (DSSs) are denoted by the sets J_R and J_D, and those allocated to i-th RSS and k-th DSS are denoted by $J_{i,R}$ and $J_{k,D}$, respectively. Each RSS uses q sub-channels where $1 \leq q \leq Q_R$. During the m-th symbol interval of the ranging time-slot, i-th RSS transmits $\{C_{i,R}^{(m)}(l) \;:\; l = 0, \ldots, qN_{1R} - 1\}$ on the sub-carriers defined by $J_{i,R}$ while k-th DSS transmits $\{C_{k,D}^{(m)}(l) \;:\; l = 0, \ldots, N_{k,D} - 1\}$ on the sub-carriers defined by $J_{k,D}$ where $N_{k,D}$ denotes the cardinality of the set $J_{k,D}$ which would be an integer multiple of N_{1D}.

In sub-carrier domain at m-th OFDM symbol interval, the length-N ranging code vector for i-th RSS and the length-N data vector for k-th DSS are denoted by $\boldsymbol{X}_{i,R}^{(m)}$ and $\boldsymbol{X}_{k,D}^{(m)}$, respectively. Their n-th elements ($n \in \{0, \ldots, N-1\}$) are given by

$$X_{i,R}^{(m)}(n) = \begin{cases} A_{i,R}C_{i,R}^{(m)}(l), & n = J_{i,R}(l), \\ & l = 0, \ldots, qN_{1R} - 1 \\ 0 & \text{otherwise} \end{cases} \tag{1}$$

$$X_{k,D}^{(m)}(n) = \begin{cases} A_{k,D}C_{k,D}^{(m)}(l), & n = J_{k,D}(l), \\ & l = 0, \ldots, N_{k,D} - 1 \\ 0 & \text{otherwise} \end{cases} \tag{2}$$

where $|C_{i,R}^{(m)}(l)| = |C_{k,D}^{(m)}(l)| = 1$, and $\{A_{i,R},\; A_{k,D} > 0\}$ are amplitude factors. Denote the N-point unitary inverse fast Fourier transform (IFFT) of $\boldsymbol{X}_{i,R}^{(m)}$ and $\boldsymbol{X}_{k,D}^{(m)}$ by $\boldsymbol{x}_{i,R}^{(m)}$ and $\boldsymbol{x}_{k,D}^{(m)}$, respectively. After cyclic prefix (CP) insertion, the time-domain transmitted signal samples of i-th RSS are denoted by

$$x_{i,R}(n) = \begin{cases} x_{i,R}^{(m)}(l-N_g), & n = m(N+N_g)+l, \\ & l = 0, \ldots, N+N_g-1; \\ & m = 0, \ldots, M-1 \\ 0, & \text{otherwise} \end{cases} \quad (3)$$

where $\{x_{i,R}^{(m)}(-l) : l = 1, \ldots, N_g\}$ represent CP samples and hence, $x_{i,R}^{(m)}(-l) = x_{i,R}^{(m)}(N-l)$. Similarly, those of k-th DSS are denoted by

$$x_{k,D}(n) = \begin{cases} x_{k,D}^{(m)}(l-N_g), & n = m(N+N_g)+l, \\ & l = 0, \ldots, N+N_g-1; \\ & m = 0, 1, \ldots \\ 0, & n < 0. \end{cases} \quad (4)$$

Suppose that there are N_R RSSs and N_D DSSs in the system and the system can support a maximum of N_c simultaneous RSSs (corresponding to N_c ranging opportunities or N_c different ranging signals). The N_R RSSs' ranging signal indices are denoted by the set $\mathcal{I}_R$. For simplicity and without loss of generality, we assume that i-th RSS uses the ranging signal $\{X_{i,R}(k)\}$, and hence $\mathcal{I}_R = \{0, \ldots, N_R-1\}$ from the possible indices $\{0, \ldots, N_c-1\}$. After obtaining the ranging channel information through the control channel UL-MAP, each RSS chooses one of the N_c ranging signals randomly and performs initial ranging process during one ranging time-slot. Since the locations of different SSs are different, the corresponding transmission delays ($d_{i,R}$ for i-th RSS and $d_{k,D}$ for k-th DSS) are different. Hence, at the beginning of the ranging process, their relative delays with respect to the BS's time-slot boundary are different. The maximum possible relative delay $d_{\max,\mathrm{R}}$ for a RSS is the round-trip transmission delay for a RSS at the cell boundary. In practice, we can find this maximum relative delay from the knowledge of cell radius. Note that N_g for RSSs should be designed such that $N_g \geq d_{\max,\mathrm{R}} + L$. For DSSs, since initial ranging processes have already been completed, the maximum possible delay $d_{\max,\mathrm{D}}$ is determined by the timing requirement for the ranging process [1].

We consider a multi-path Rayleigh fading channel with L sample-spaced taps. The channel tap gains for i-th RSS and k-th DSS (denoted by $\{h_{i,R}(l)\}$ and $\{h_{k,D}(l)\}$, $l = 0, \ldots, L-1$, respectively) are assumed to remain constant over one ranging time-slot, and the average total energy of the channel taps is σ_h^2. Let the channel output samples for i-th RSS signal be $\{y_{i,R}(n)\}$ and those for k-th DSS be $\{y_{k,D}(n)\}$, i.e.,[2].

$$y_{i,*}(n) = \sum_{l=0}^{L-1} h_{i,*}(l) x_{i,*}(n-l-d_{i,*}). \quad (5)$$

Then the n-th received signal sample (n = 0, 1, ...) at the BS can be expressed as

$$y(n) = \sum_{i=0}^{N_R-1} y_{i,R}(n) + \sum_{u=0}^{N_D-1} y_{u,D}(n) + w(n) \quad (6)$$

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

[1]The CP interval for data transmission can be designed to be smaller than that of ranging symbols.

[2]In the rest of the paper, the subscript * denotes whether R or D.

At the receiver, we consider an observation window of $M(N_d+N_g)+N_g$ samples to make sure that all received ranging signals reside within this observation window regardless of their possibly different transmission delays. There are M ISI-free windows of N samples each and $(M+1)$ ISI-affected windows of N_g samples each, in the observation window. The sample indices of m-th ISI-free window and m-th ISI-affected window are, respectively, denoted by

$$J_{\mathrm{ISI-free}}^{(m)} = \{m(N+N_g)+N_g, \ldots, (m+1)(N+N_g)-1\}, \quad m = 0, \ldots, M-1 \quad (7)$$

$$J_{\mathrm{ISI}}^{(m)} = \{m(N+N_g), \ldots, m(N+N_g)+N_g-1\} \quad m = 0, \ldots, M. \quad (8)$$

Notation: We use the following notations: $i_M = \lfloor i/M \rfloor$ and $[i]_M = i$ modulo M. Hence $i = i_M M + [i]_M$.

III. Proposed Ranging Signal Design

We consider an interleaved OFDMA system where $N_{1R}(\geq L)$ sub-carriers of each (initial) ranging sub-channel are spread out across the sub-carrier domain with an equal spacing of N/N_{1R} sub-carriers. Each RSS uses N_{1R} sub-carriers (i.e., q=1). The total number of ranging opportunities provided by our design is $N_c = Q_R M$.

The N_c different ranging signals (corresponding to the N_c ranging opportunities) are divided into Q_R groups. The signals from different groups are transmitted on different sub-carriers. The sub-carrier assignment for i-th RSS is defined by

$$J_{i,R} = \{(\frac{nN}{N_{1R}} + \Delta_{i_M}) : n = 0, \ldots, N_{1R}-1\} \quad (9)$$

where

$$0 \leq \Delta_{i_M} < \frac{N}{N_{1R}} \ \& \ \Delta_{i_M} \neq \Delta_{k_M} \text{if } i_M \neq k_M. \quad (10)$$

A recommended choice for Δ_{i_M} is $i_M N/(Q_R N_{1R})$ which assigns ranging sub-carriers of different gorups equally spread out with equal spacing across the entire bandwidth. Note that $\{J_{i,R}\}$ and $\{J_{k,D}\}$ should be disjoint for all i and k and $\{J_{i,R}\}$ and $\{J_{k,R}\}$ are disjoint for all $i_M \neq k_M$ but they are the same if $i_M = k_M$. Due to the orthogonality among sub-carriers, received signals from different groups would not interfere with each other if perfectly frequency-synchronized. On the other hand, the M ranging signals in the same group are transmitted on the same N_{1R} sub-carriers over M symbol intervals. In order to separate these M ranging signals, we introduce phase-shift-orthogonality over M symbol intervals within each group as follows. The ranging sub-carrier symbols for i-th RSS at m-th symbol interval and 0-th symbol interval are related by

$$C_{i,R}^{(m)}(l) = C_{i,R}^{(0)}(l) e^{j2\pi[i]_M m/M}, \ m = 1, \ldots, M-1. \quad (11)$$

The above orthogonal ranging code design provides orthogonality between any signal pairs of the *received* ranging and data signals and hence, it is optimal for multi-user ranging code detection if perfectly frequency-synchronized.

Note that we set $N_{1R} \geq L$ in our design so that the received ranging signal from any RSS completely characterizes the corresponding channel [3]. Our sub-carrier assignment for ranging sub-channel applies the principle of pilot tone design for MIMO OFDM channel estimation from [6] but we tailor it to the ranging process. Hence, our ranging code design is optimal in terms of characterizing the corresponding channels

[3]At least L pilot tones are necessary for the identification of an L-tap channel.

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(may be useful in adaptive resource allocation) which in turn gives a better representation of average power across the sub-carrier domain for the RSS.

It can be easily observed that all ranging codes are orthogonal to each other, i.e.,

$$\sum_{m=0}^{M-1}\sum_{n=0}^{N-1} X_{i,R}^{(m)}(n)(X_{k,R}^{(m)}(n))^* = \begin{cases} |A_i|^2 N_{1R} M, & \text{if } i = k \\ 0, & \text{if } i \neq k. \end{cases} \quad (12)$$

Due to the orthogonality, all ranging codes can be easily decoupled, rendering an efficient, low-complexity multi-user ranging code detection as will be discussed in the next section.

IV. Proposed Ranging Method

Our proposed method first performs multi-user ranging signal detection based on the BS timing reference. Then utilizing the ranging signal detection results, a power estimation and an iterative timing estimation are carried out for each detected RSS. Power and timing estimates are compared with ranging requirements and necessary control information is broadcast back to the RSSs.

A. Multi-User Ranging Signal Detection

Based on the BS timing reference, we obtain frequency-domain received k-th sub-carrier symbol ($k \in \{0, \dots, N-1\}$) at m-th symbol interval ($m \in \{0, \dots, M-1\}$) by N-point unitary FFT of $\{y(n): \ n \in J_{\text{ISI-free}}^{(m)}\}$ as

$$Y^{(m)}(k) = \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} y(m(N+N_g)+N_g+l)e^{-j2\pi lk/N} \quad (13)$$

$$= \sum_{i=0}^{N_R-1} X_{i,R}^{(m)}(k)H_{i,R}(k) + \sum_{u=0}^{N_D-1} X_{u,D}^{(m)}(k)H_{u,D}(k) + W^{(m)}(k) \quad (14)$$

where

$$H_{i,R}(k) = e^{\frac{-j2\pi k d_{i,R}}{N}} \sum_{l=0}^{L-1} h_{i,R}(l)\, e^{\frac{-j2\pi lk}{N}} \quad (15)$$

$$H_{i,D}(k) = e^{\frac{-j2\pi k d_{i,D}}{N}} \sum_{l=0}^{L-1} h_{i,D}(l)\, e^{\frac{-j2\pi lk}{N}} \quad (16)$$

$$W^{(m)}(k) = \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} w(m(N+N_g)+N_g+l)\, e^{\frac{-j2\pi lk}{N}}. \quad (17)$$

Ranging codes using different ranging sub-carriers are already decoupled in the frequency-domain. Within each group of ranging codes using the same ranging sub-carriers, the sub-carrier symbols of different ranging codes can easily be decoupled by utilizing the orthogonality property (see (12), (1) and (11)) as

$$Z_i(k) = \frac{1}{M} \sum_{m=0}^{M-1} Y^{(m)}(k) e^{-j2\pi [i]_M m/M}, \ k \in J_{i,R} \quad (18)$$

$$= \begin{cases} X_{i,R}^{(0)}(k)H_{i,R}(k) + \bar{W}(k), & i \in \{0, \dots, N_R-1\} \\ \bar{W}(k) & i \in \{N_R, \dots, N_c-1\} \end{cases} \quad (19)$$

where $\{\bar{W}(k)\}$ are iid circularly symmetric complex Gaussian random variables with zero mean and variance σ_w^2/M.

The decision variable D_i to decide the presence of the i-th ranging signal is defined as

$$D_i = \sum_{k \in J_{i,R}} Z_i(k) Z_i^*(k) \quad (20)$$

$$= \begin{cases} \sum_{k \in J_{i,R}} \|X_{i,R}^0(k)H_{i,R}(k) + \bar{W}(k)\|^2, & i \in \mathcal{I}_R \\ \sum_{k \in J_{i,R}} \|\bar{W}(k)\|^2, & i \notin \mathcal{I}_R. \end{cases} \quad (21)$$

The decision variable D_i is not affected by other RSSs and DSSs, resulting in an efficient, low-complexity multi-user ranging signal detection. Our orthogonal ranging signal design is the key to this efficient multi-user ranging signal detection.

The i-th ranging signal detector decides that the i-th ranging signal is detected if $D_i > \eta_i$. The value of the threshold η_i is derived in the following. Since $H_{i,R}(k)$ for $k \in J_{i,R}$ are spread out across the whole bandwidth with equal spacing and $N \gg N_{1R}$ in typical OFDMA systems, $\{H_{i,R}(k): \ k \in J_{i,R}\}$ can be approximately treated as independent random variables. Then D_i is a summation of squares of $2N_{1R}$ approximately-iid real-valued random variables and can be considered as a chi-square random variable with $2N_{1R}$ degrees of freedom. Hence, the probability density function of D_i can be given by

$$p_{D_i}(x) \approx \frac{1}{\sigma_i^{2N_{1R}} 2^{N_{1R}} \Gamma(N_{1R})} x^{N_{1R}-1} e^{\frac{-x}{2\sigma_i^2}}, x \geq 0 \quad (22)$$

where

$$\sigma_i^2 = \begin{cases} \frac{A_i^2\sigma_h^2}{2} + \frac{\sigma_w^2}{2M}, & i \in \mathcal{I}_R \\ \frac{\sigma_w^2}{2M}, & \text{otherwise.} \end{cases} \quad (23)$$

The detection threshold η_i is derived from the following maximum-likelihod criterion

$$p_{D_{i|i \in \mathcal{I}_R}} \gtrless p_{D_{i|i \notin \mathcal{I}_R}} \quad (24)$$

and obtained as

$$\eta_i = \frac{2N_{1R}\sigma^2_{i|i\notin\mathcal{I}_R} \ln\left(\frac{\sigma^2_{i|i\in\mathcal{I}_R}}{\sigma^2_{i|i\notin\mathcal{I}_R}}\right)}{1 - \frac{\sigma^2_{i|i\notin\mathcal{I}_R}}{\sigma^2_{i|i\in\mathcal{I}_R}}} \quad (25)$$

$$= \frac{N_{1R}\sigma_w^2}{M}\left(1 + \frac{N_{1R}}{\text{SNR}_i NM}\right) \ln\left(1 + \frac{\text{SNR}_i NM}{N_{1R}}\right) \quad (26)$$

where SNR_i is the SNR of i-th ranging signal defined as

$$\text{SNR}_i = \frac{N_{1R}A_i^2 E[|H_{i,R}(k)|^2]}{N\sigma_w^2} \approx \frac{N_{1R}A_i^2\sigma_h^2}{N\sigma_w^2} \quad (27)$$

Note that η_i is the same for all i if transmit-powers of all RSSs are the same. Also, under fixed system parameters (N_{1R}, N, M, σ_w^2), the changes in η_i value due to different SNR values (say within [-10dB $\sim$ 20dB]) is very small and can be neglected when compared with the mean of D_i in the first case ($i \in \mathcal{I}_R$) of (21). So we can use a fixed detection threshold as

$$\eta_i = \frac{N_{1R}\sigma_w^2}{M}\left(1 + \frac{N_{1R}}{\text{SNR}_f NM}\right) \ln\left(1 + \frac{\text{SNR}_f NM}{N_{1R}}\right) \quad (28)$$

where SNR_f is a fixed design SNR for the detection threshold and in our simulation we use $\text{SNR}_f = 100$. To estimate the noise power σ_w^2 required in the detection threshold, we can keep one fixed ranging opportunity (say i_0-th ranging code) unused . The corresponding D_{i_0} corresponds to the second case ($i \notin \mathcal{I}_R$) of (21) from which the BS receiver can easily estimate the noise power as

$$\hat{\sigma_w^2} = D_{i_0} M / N_{1R}. \quad (29)$$

It is also possible to pre-measure the noise power by other means at the BS without sacrificing a ranging opportunity. We can alternatively use a fixed design value of σ_w^2 for the detection threshold. For environments with varying noise and interference levels, noise (interference included) power estimator could be used, otherwise, a fixed design value of σ_w^2 would be a preferred choice.

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In the presence of residual normalized (by the sub-carrier spacing) frequency offsets, the σ_i^2 required in the detection threshold can be approximated as

$$\sigma_i^2 \approx \begin{cases} \left|\frac{1-e^{j2\pi\bar{v}}}{N(1-e^{j2\pi\bar{v}/N})}\right|^2 \left[\sum_{l=0}^{M-1} G_{l,i} + \sigma_w^2/(2M)\right], & i \in \mathcal{I}_R \\ \left|\frac{1-e^{j2\pi\bar{v}}}{N(1-e^{j2\pi\bar{v}/N})}\right|^2 \left[\sum_{l=0,l\neq i}^{M-1} G_{l,i} + \sigma_w^2/(2M)\right], & i \notin \mathcal{I}_R \end{cases} \tag{30}$$

where $\bar{v}$ is a fixed design value of residual normalized frequency offset (upper end of possible residual normalized frequency offset range) and

$$G_{l,i} = \frac{\sigma_h^2}{2M^2} A_l^2 \left| \frac{1 - e^{j2\pi(M(N+N_g)\bar{v}/N + [l]_M - [i]_M)}}{1 - e^{j2\pi[(N+N_g)\bar{v}/N + ([l]_M - [i]_M)/M]}} \right|^2. \tag{31}$$

The detection threshold in the presence of residual frequency offset can be calculated from (25) where σ_i^2 is given by (30).

B. Power Estimation

After i-th ranging signal is detected, the next step in the ranging process is to estimate the corresponding (normalized) received power P_i defined by

$$P_i = \frac{\sum_{k \in J_{i,R}} |X_{i,R}(k)|^2 |H_{R,i}(k)|^2}{N}. \tag{32}$$

From (21), we obtain

$$P_i = \frac{D_i - \sum_{k\in J_{i,R}} |\bar{W}_i(n)|^2 - \sum_{k \in J_{i,R}} 2\Re[X_{i,R}(k) H_{i,R}(k) \bar{W}_i(k)]\}}{N}. \tag{33}$$

Since the last term in the nominator has a zero mean, we simply estimate the received power of i-th ranging signal as

$$\hat{P}_i = \frac{D_i - \sigma_w^2 N_{1R}/M}{N}. \tag{34}$$

C. Iterative Timing Offset Estimation

The received samples $\{y(n)\}$ are the superposition of signals from N_R RSSs and N_D DSSs with different channels and different timing offsets. Ranging process estimates the timing offsets of all detected RSSs based on the received samples $\{y(n)\}$ within the observation window. We address this multi-user timing offset estimation by means of N_R single RSS timing offset estimators. For each detected RSS, an iterative timing offset estimator is developed. Although multi-user timing offsets are not jointly estimated, each timing offset estimator utilizes the timing offset estimates of the other RSSs obtained in the previous iteration.

The proposed timing offset estimator exploits the proposed orthogonal ranging signal design and the signal redundancy introduced by the CPs. Based on the orthogonality among the RSSs and DSSs signals over M ISI-free windows, we can obtain interference-free clean samples (denoted by $\{\bar{y}_{i,*}^{(m)}(n) : n = 0, \ldots, N-1\}$) of the channel-output m-th symbol (except noise contamination) for i-th SS as (see (19) and (11))

$$\bar{y}_{i,R}^{(m)}(n) = \frac{1}{\sqrt{N}} \sum_{k \in J_{i,R}} Z_i(k) e^{\frac{j2\pi nk}{N}} e^{\frac{j2\pi m[i]_M}{M}} \tag{35}$$

$$\bar{y}_{i,D}^{(m)}(n) = \frac{1}{\sqrt{N}} \sum_{u \in J_{i,D}} Y^{(m)}(u) e^{\frac{j2\pi nu}{N}}. \tag{36}$$

From $\{\bar{y}_{i,*}^{(m)}(n)\}$, we can reconstruct, for i-th RSS (or DSS) with transmission delay d, a clean version of the m-th symbol's CP, $\bar{\boldsymbol{y}}_{i,R,\text{CP}}^{(m)}(d)$ (or $\bar{\boldsymbol{y}}_{i,D,\text{CP}}^{(m)}(d)$) given by

$$\bar{\boldsymbol{y}}_{i,*,\text{CP}}^{(m)}(d) = \begin{cases} \left[\mathbf{0}_d, \bar{y}_{i,*}^{(0)}(N - N_g + d), \ldots, \bar{y}_{i,*}^{(0)}(N-1)\right]^T, & m = 0 \\ \left[\bar{y}_{i,*}^{(m-1)}(0), \ldots, \bar{y}_{i,*}^{(m-1)}(d-1), \right. \\ \left. \bar{y}_{i,*}^{(m)}(N - N_g + d), \ldots, \bar{y}_{i,*}^{(m)}(N-1)\right]^T, & m \neq 0 \end{cases} \tag{37}$$

where $\mathbf{0}_d$ is the all-zero vector of length d.

On the other hand, we can obtain another version of the CP (interference-affected version denoted by $\hat{\boldsymbol{y}}_{i,R,\text{CP}}^{(m)}$) by subtracting the clean version of CPs of the other SSs from the received CP within the ISI-affected window, $\boldsymbol{y}_{\text{CP}}^{(m)}$, as follows:

$$\hat{\boldsymbol{y}}_{i,R,\text{CP}}^{(m)} = \boldsymbol{y}_{\text{CP}}^{(m)} - \sum_{k \in \mathcal{I}_R, k\neq i} \bar{\boldsymbol{y}}_{k,R,\text{CP}}^{(m)}(d_{k,R}) - \sum_{u=0}^{N_D-1} \bar{\boldsymbol{y}}_{u,D,\text{CP}}^{(m)}(\tau^{(D)}) \tag{38}$$

where $\tau^{(D)}$ is the design parameter to replace the actual transmission delays for all DSSs since no timing offset estimation for DSS is performed. The $\hat{\boldsymbol{y}}_{i,R,\text{CP}}^{(m)}$ contains correct timing information $d_{i,R}$ (embedded in $\boldsymbol{y}_{\text{CP}}^{(m)}$) and hence, the timing offset estimate for i-th RSS can be obtained by maximizing a sliding correlation metric between the clean version and the interference-affected version as

$$\hat{d}_{i,R} = \underset{0 \le d \le d_{\max,R}}{\arg\max}\ \Re\left\{ \sum_{m=0}^{M-1} \left(\bar{\boldsymbol{y}}_{i,R,\text{CP}}^{(m)}(d)\right)^H \hat{\boldsymbol{y}}_{i,R,\text{CP}}^{(m)} \right\}. \tag{39}$$

Constructing $\hat{\boldsymbol{y}}_{i,R,\text{CP}}^{(m)}$ requires the knowledge of $\{d_{k,R} : k \neq i\}$ which are unknown and what the timing offset estimator for k-th RSS is estimating. However, we can apply an iterative approach where $\{d_{k,R} : k \neq i\}$ are replaced with $\{\hat{d}_{k,R}^{(0)} : k \neq i\} = \tau^{(R)}$ at the beginning of the first iteration to get the i-th RSS's timing offset estimate $\{\hat{d}_{i,R}^{(1)}\}$ at the end of the first iteration. In general, at θ-th iteration, the interference-affected version $\hat{\boldsymbol{y}}_{i,R,\text{CP}}^{(m,\theta-1)}$ is constructed by using $\{\hat{d}_{k,R}^{(\theta-1)} : k \neq i\}$ and hence, the i-th RSS's timing offset estimate at the θ-th iteration is given by

$$\hat{d}_{i,R}^{(\theta)} = \underset{0 \le d \le d_{\max,R}}{\arg\max}\ \Re\left\{ \sum_{m=0}^{M-1} \left(\bar{\boldsymbol{y}}_{i,R,\text{CP}}^{(m)}(d)\right)^H \hat{\boldsymbol{y}}_{i,R,\text{CP}}^{(m,\theta-1)} \right\}. \tag{40}$$

We assume that the timing offset $d_{k,*}$ is a random variable with a uniform distribution in $\{0, \ldots, \tau_{\max}\}$ where $\tau_{\max}$ is equal to $d_{\max,R}$ for RSS and $d_{\max,D}$ for DSS. Then the best initial timing offset value τ^* which gives the smallest timing-mismatch-interference energy is obtained by

$$\tau^* = \arg\min_{\tau}\{\int_0^{\tau_{\max}} \frac{|x-\tau|}{\tau_{\max}} dx\} = \frac{\tau_{\max}}{2}. \tag{41}$$

Hence, the best initial timing offset estimates used in our timing offset estimator are given by

$$\tau^{(R)} = \frac{d_{\max,R}}{2},\ \tau^{(D)} = \frac{d_{\max,D}}{2}. \tag{42}$$

V. Simulation Results and Discussion

The OFDMA system parameters are selected from [1]. The uplink bandwidth is 3 MHz, the sub-carrier spacing is 1.67 KHz, and $N = 2048$. The ranging channel has 128 sub-carriers over $M = 2$ symbol intervals.

For the proposed ranging signal structure, the parameters used are: $N_{1R} = 8$, $q = 1$, $Q_R = 16$, $\Delta_{i_M} = 16\ i_M$, and

$N_c = 32$. We use QPSK format for DSS. The combined transmit and receive filter is a raised-cosine filter $g_T(t)$ with a roll-off factor of 0.5. The SUI-3 channel model with 3 paths [5] is used and the (statistical) average total energy of all channel taps σ_h^2 is set to unity. The channel impulse response for the i-th user is given by $h_i(l) = \sum_{i=0}^{2} \gamma_i\, g_T(lT_s - \tau_i - t_0)$, $l = 0, \ldots, L-1$; where $\{\gamma_i\}$ and $\{\tau_i\}$ are the gains and delays of the channel paths, t_0 is a time shift for causality, and $1/T_s$ is N times the sub-carrier spacing. The number of sample-spaced channel taps, L, is set to 7. Channels of different users are assumed to be independent. We consider a cell radius of 5km which gives the maximum transmission delay (round trip) $d_{\max,R} = 34\mu s$ = 102 samples. N_g is set to 128 samples satisfying the condition $d_{\max,R} < N_g - L$. 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_g can be $1/4$, $1/8$, $1/16$, or $1/32$ of N, and hence, the timing offset should be within ± 16 samples. We set $d_{\max,D} = 32$. For comparison, we include the performance of the methods from [2] and [4] which use the CDMA/GCL ranging codes. In the simulation, the maximum number of RSSs in one time slot, $N_{R,\max}$, is set to 15. We also include residual normalized frequency offsets in the range of $[-0.02,\ 0.02]$ which are assumed to be iid for different SSs. The value of noise power required in setting the detection threshold for the proposed method is obtained by (29).

A. Multi-User Detection Performance

Fig. 1 shows the probability of correct detection ($P_{\rm CD}$) versus the number of RSSs for the conditions of 0 DSS, 15 DSSs, and 30 DSSs in one ranging time-slot. The $P_{\rm CD}$ is defined as $E[\frac{D_c}{N_R}]$ where D_c is the number of correct detection in one ranging time-slot. All three methods are robust to the data users' interference and residual frequency offsets. However, the reference methods' $P_{\rm CD}$ performances degrade significantly as the number of RSSs increases (from 1.00 for $N_R = 1$ down to 0.65 for $N_R = 15$) while the proposed method's performance remains the same at $P_{\rm CD} = 1.00$.

Fig. 2 shows the probability of detection false-alarm ($P_{\rm FA}$) versus the number of RSSs for the conditions of 0 DSS, 15 DSSs, and 30 DSSs in one ranging time-slot. The $P_{\rm FA}$ is defined as $E[\frac{D_a}{N_c - N_R}]$ where D_a is the number of ranging codes within one ranging time-slot which are detected at the BS but are not transmitted from any RSSs. All methods are robust to frequency offsets. Proposed method and the method from [2] are robust to RSS interference and DSS interference while $P_{\rm FA}$ performance of the method from [4] degrades as the number of RSS increases. Proposed method has a better $P_{\rm FA}$ performance than both reference methods under all conditions considered.

Fig. 3 shows the probability of missed detection ($P_{\rm MD}$) versus the number of RSSs for the conditions of 0 DSS, 15 DSSs, and 30 DSSs in one ranging time-slot. The $P_{\rm MD}$ is defined as $E[\frac{D_m}{N_R}]$ where D_m is the number of RSSs which are transmitted by RSSs but are not detected at the BS. The proposed method's $P_{\rm MD}$ performance is robust to the DSS interference, RSS interference and frequency offsets while the reference methods' performances degrade as the number of RSSs increases (from $P_{\rm MD} = 0$ at $N_R = 1$ to $P_{\rm MD} = 0.35$ at $N_R = 15$). Hence, the proposed method has a much better performance than the two reference methods, especially in multiple RSSs conditions.

B. Power Estimation Performance

Fig. 4 shows the normalized power estimation MSE defined as $E[(1 - \frac{\hat{P}_i}{P_i})^2]$ versus the number of RSSs for the conditions of 0 DSS, 15 DSSs, and 30 DSSs in one ranging time-slot. Since there is no power estimator provided in [2] and [4], only the proposed method's performances in the absence/presence of frequency offsets are plotted. Power estimation performance is robust to DSS interference and the number of RSSs in the absence of frequency offsets. We observe that although the power estimation performance degrades as the number of RSSs increases in the presence of frequency offsets, the performance is still very good.

C. Timing Estimation Performance

Fig. 5 shows the standard deviation of the timing offset estimate versus the number of RSSs for the conditions of 0 DSS, 15 DSSs, and 30 DSSs in one ranging time-slot. In each simulation run, the true timing offsets for RSSs and DSSs are taken randomly from the interval $[0, d_{\rm max,R}]$ and $[0, d_{\rm max,D}]$, respectively. All the three methods are robust to the DSS interference and residual frequency offsets. The performances of all timing estimators degrade as the numbers of RSSs and DSSs increase but the proposed method gives a much better performance, especially in multiple RSSs condition. Furthermore, the proposed method satisfies the timing requirement provided in [1] all the time while the two reference methods satisfy the requirement only when the number of RSSs is small. We also evaluated the proposed timing estimator with different number of iterations in the presence of frequency offsets (the plot is omitted due to space limitation). Two-iteration is the best choice since more iterations do not bring in a noticeable improvement.

VI. Conclusions

We have presented a new ranging signal design and a new initial ranging method for OFDMA systems in multipath fading channel environments. The proposed ranging signal design is based on the orthogonality principle and the best channel identification conditions. This orthogonal ranging signal design results in robust, high performance, low-complexity multi-user ranging signal detection and power estimation. A new iterative estimation of timing offsets for all ranging subscriber stations is developed based on the proposed orthogonal ranging signal design and the cyclic prefix redundancy. The simulation results show that the new approach is more robust to interference from other ranging/data subscriber stations and multi-path fading channel effects and also has better performance than the existing methods which use the CDMA-type ranging codes in frequency-domain.

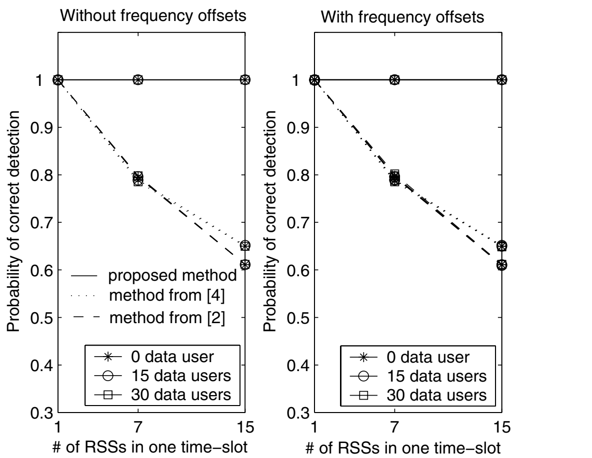

Fig. 1. The probability of correct detection for several ranging code detectors

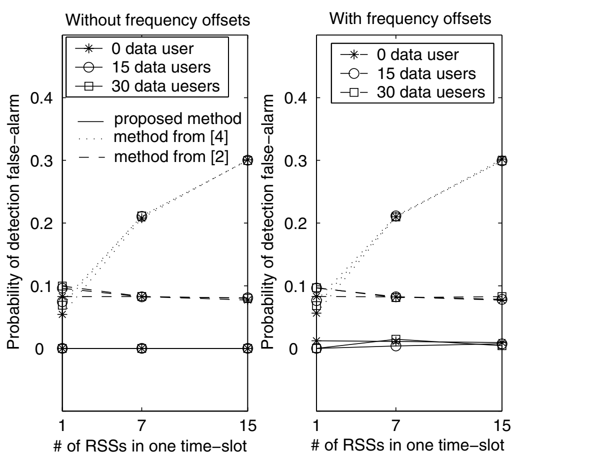

Fig. 2. The probability of detection false-alarm for several ranging code detectors

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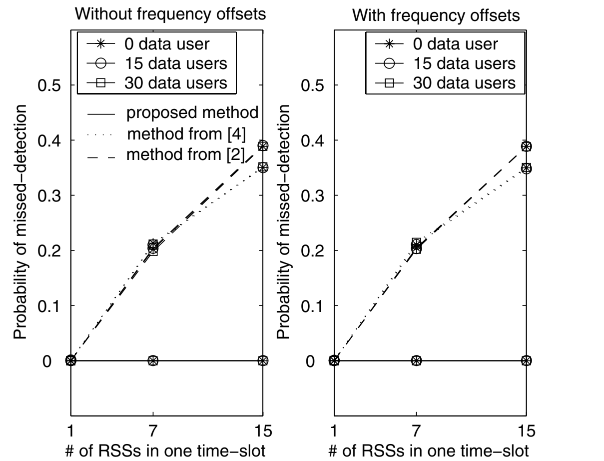

Fig. 3. The probability of missed-detection for several ranging code detectors

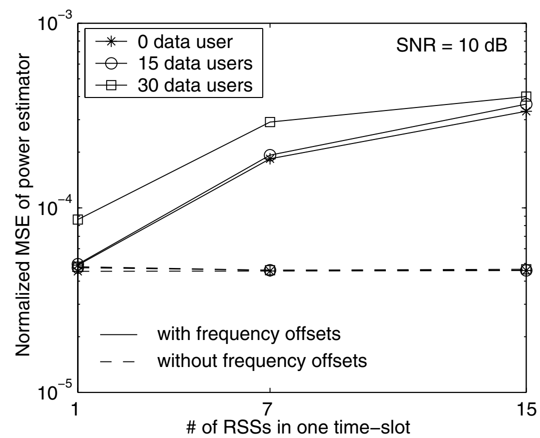

Fig. 4. The normalized MSE of the power estimator at SNR = 10 dB

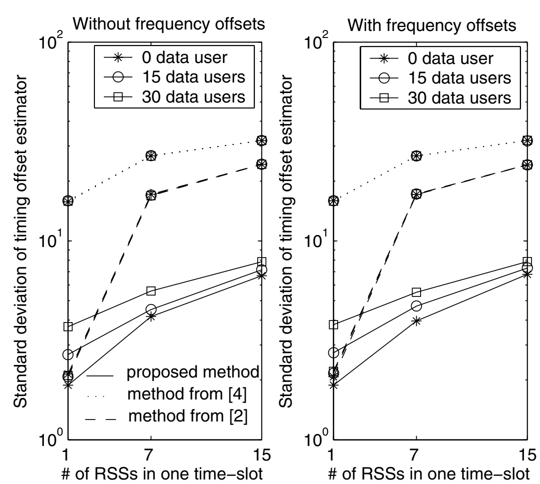

Fig. 5. The standard deviation (in samples) of the timing offset estimators at SNR = 10 dB

