Distributed TDoA Estimation for Wireless Sensor Networks Based on Frequency-Hopping in Multipath Environment

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***Abstract*—A novel distributed TDoA (Time-Difference-of-Arrival) estimation method for wireless sensor networks (WSN) in multipath environment is proposed in this paper. For the conventional narrow band system usually adopted in the wireless sensors, the frequency-hopping technique can be utilized to perform the estimation of the wide band frequency-domain channel response, by which the super-resolution TDoA estimation is conducted independently at each sensor to detect range difference from the sensor to the two anchors. The proposed method only relies on the conventional narrow band radio transceiver and each sensor conducts range estimation independently without any information exchange or time synchronization requirement, which is a distributed range estimation method more suitable for WSN.**

I. Introduction

Wireless sensor networks (WSNs) is a significant technology attracting considerable research interests, and the self-localization capability is a highly desirable characteristic in WSNs [1].

Different methods try to estimate distances between two nodes using time based measures. The most simple and intuitive is ToA (Time-of-Arrival). In this case the distance between two nodes is directly proportional to the time the signal takes to propagate from one point to another. However, this type of estimation requires precisely synchronized nodes and time at which the signal leaves the node must be in the packet that is sent. On the other hand, the super-resolution ToA estimation, which is based on frequency-domain measurement of channel response, has been studied recently. In the literatures, the time-delay estimation problem has been studied with a variety of super resolution technique, such as minimum-norm [2], root MUSIC [3], TLS-ESPRIT [4], MUSIC with diversity [5]. However, here we should note that, narrow band wireless system is usually adopted in the wireless sensors, and thus, this frequency-domain ToA estimation method can not be directly applied to WSN.

The TDoA (Time-Difference-of-Arrival) is another commonly used method which utilizes the range difference to conduct positioning according to hyperbolic algorithm. TDoA is based on: 1) the difference in the times at which a single signal from a single node arrives at three or more nodes; 2) the difference in the times at which multiple signals from a single node arrive at another node [6]. The first case, which is more common in cellular networks, requires precisely synchronized receiver nodes. In the second case, more common and suitable for WSNs, nodes must be equipped with extra hardware capable of sending two types of signals simultaneously, like radio/ultrasound, that adds to the costs and size of the platform. Moreover, the second case often has the weakness of limited range and directionality constraints. Therefore, it cannot be applied to large-scale networks [7].

For WSN, the authors in [8] has proposed a range estimation method which utilizes two radio transmitter to create an interference signal with a low frequency envelop. The range estimation is based on the multiple measurement of the relative phase offset between two receivers at different carrier frequencies to combat the phase ambiguity effects, which greatly complicates the system operation. Moreover, the multipath effect is not considered in [8]. In order to combate multipath effect, the ultra wide band (UWB) technique can be utilized to conduct TOA or TDOA estimation, with the advantages including low-cost, strong penetrating ability, and easy separation of multipath signals possible, but it still has the shortcoming of short-range [9][10].

In this paper, inspired by frequency-hopping technique, we propose a novel distributed range estimation method for WSN in multipath environment. The key observation is that, for the narrow band system, the frequency-hopping technique can be utilized to perform the estimation of the wide band frequency-domain channel response, by which the super-resolution frequency-domain range estimation can be then conducted. The proposed method only relies on the conventional narrow band radio transceiver and each sensor conducts range estimation independently without any information exchange or time synchronization requirement, which is a distributed range estimation method more suitable for WSN.

This paper is organized as follows. In Section II, the TDoA estimation method is illustrated. Then in Section III, we present simulation results based on our proposed method. Finally in Section IV we present conclusions.

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II. TDoA Estimation

A. Conventional Super-Resolution ToA Estimation

In most applications, the ground around or under the antenna and other nearby objects such as tress or buildings can have significant impact on the shape and strength of the radiated pattern, which makes the transmission environment more complex and multipath effects can not be neglected. Hence, the assumption of existence of only one DLOS (direct line-of-sight) path is no long reasonable. Consequently, a fine anti multipath ability is another important requirement of the localization in WSN.

The multipath radio propagation channel is normally modeled as a complex equivalent impulse response given by

$$h(t) = \sum_{k=0}^{L_p-1} \alpha_k \delta(t - \tau_k) \tag{1}$$

where L_p is the number of multipath components, $\alpha_k = |\alpha_k| e^{\mathbf{j}\theta_k}$ and τ_k are the complex attenuation and propagation delay of the kth path, respectively, while the multipath components are indexed so that the propagation delays τ_k, $(k = 0, 1, \cdots, L_p - 1)$, are in ascending order. As a result, τ_0 in the model denotes the propagation delay of the DLOS path. Taking the Fourier transform of (1), the frequency-domain channel response can be expressed as

$$H(f) = \sum_{k=0}^{L_p-1} \alpha_k e^{-\mathbf{j}2\pi f \tau_k} \tag{2}$$

If we exchange the role of the time and frequency variables in (2), we can observe that it becomes a harmonic signal model which is well known in spectral estimation field. Recently, a number of researchers have applied super-resolution spectral estimation techniques for time-domain analysis of different applications. In practice, the discrete samples of frequency-domain channel response can be obtained by using the network analyzer, by using a multicarrier modulation technique such as OFDM, or in direct-sequence spread spectrum (DSSS) system by deconvolving the received signal over the frequency band of high signal-to-noise ratio. In other words, this super-resolution ToA estimation technique requires the support of the wide band system, or even ultra wide band system. However, narrow band wireless system is usually adopted for the wireless sensors, and thus, this frequency-domain ToA estimation method can not be directly applied to WSN.

Fortunately, we have found that some of the existing mature sensor nodes can switch their radio narrow-band channels over a wide frequency band, i.e., the frequency-hopping can be supported [11]. Thus, for the narrow band system, if the wide band frequency-domain channel response can be obtained through frequency-hopping, the super-resolution frequency-domain range estimation can also be conducted.

B. Sampling for the Channel Response

Suppose that the sensor network consists of numerous sensor nodes and three anchors (A, B, C). The sensor nodes with restricted energy supply are distributed randomly in the sensing filed. We assume the anchors are not energy constrained and their communication ranges can cover the whole network. In addition, each anchor has a GPS receiver for self-localization.

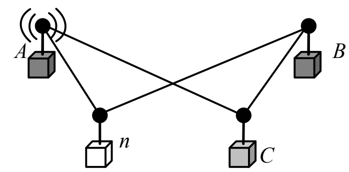

Fig. 1. Illustration of the network structure.

Firstly, let the anchors and all the sensor nodes hop to a common frequency point. Since the frequency offset between the anchors and any sensor node can be directly estimated and compensated, we assume the frequency synchronization is achieved in the network in this paper. Next, for a given sensor node n, we assume there are several reflected paths with small power, as well as the DLOS path with greatest power, between node n and any anchor node. Furthermore, a reasonable assumption is made that the channel amplitude response of the transceiver is invariable during the hopping process in a certain frequency band. Thus, we mainly focus on the affect of the channel phase response.

Assume the anchors A and B emit the time-division sine pilot $e^{\mathbf{j}(2\pi f t + \varphi_A)}$ and $e^{\mathbf{j}(2\pi f t + \varphi_B)}$, respectively, where φ_A and φ_B denote the initial phases of A and B, respectively, and f is the carrier frequency. Then the received invariable baseband signal after down-conversion in the absence of noise at node n can be expressed as

$$\begin{aligned} r_{A,n} &= e^{\mathbf{j}(\varphi_A - \varphi_n)} \sum_{i=0}^{P-1} \alpha_{A,i} e^{-\mathbf{j}2\pi f \tau_{A,i}} \\ r_{B,n} &= e^{\mathbf{j}(\varphi_B - \varphi_n)} \sum_{j=0}^{Q-1} \alpha_{B,j} e^{-\mathbf{j}2\pi f \tau_{B,j}} \end{aligned} \tag{3}$$

where P and Q denote the number of multipaths from the two anchors to node n, respectively, φ_n denotes the initial phase of n, $\alpha_{A,i}$ and $\tau_{A,i}$ are the complex attenuation and propagation delay of the ith path from anchor A to node n, $\alpha_{B,j}$ and $\tau_{B,j}$ are the complex attenuation and propagation delay of the jth path from anchor B to node n. Then $\sum_{i=0}^{P-1} \alpha_{A,i} e^{-\mathbf{j}2\pi f \tau_{A,i}}$ and $\sum_{j=0}^{Q-1} \alpha_{B,j} e^{-\mathbf{j}2\pi f \tau_{B,j}}$ can be viewed as the frequency-domain channel response at frequency point f. Thus, we can observe that the received signals $r_{A,n}$ and $r_{B,n}$ will reflect the frequency-domain channel response with the effect of the initial phases, i.e., φ_A, φ_B and φ_n. Here, we should note that, these initial phases will be randomized at different frequent points in the frequency-hopping process, and thus, they cannot keep invariable. Thereby, in order to obtain the effective frequency-domain channel response, the effect of the initial phases should be eliminated.

First, the φ_n can be eliminated through the conjugate multiplication between the two signals $r_{A,n}$ and $r_{B,n}$, given by

$$\begin{aligned} r_n = &\; r_{A,n} (r_{B,n})^* \\ = &\; e^{\mathbf{j}(\varphi_A - \varphi_B)} \sum_{i=0}^{P-1} \sum_{j=0}^{Q-1} \alpha_{A,i} \alpha_{B,j}^* e^{-\mathbf{j}2\pi f (\tau_{A,i} - \tau_{B,j})} \end{aligned} \tag{4}$$

Furthermore, with the assistance of the anchor C, the effective frequency-domain response can be achieved. Specifically, assume that there is only the DLOS existing between C and the two emitting anchors A and B. Here we should note that this assumption is not strict, since the position and posture of anchors can be adjusted. Thus, in above process, anchor C can also obtain the following signal

$$r_C = e^{\mathbf{j}(\varphi_A - \varphi_B)} e^{-\mathbf{j}2\pi f(\tau_{AC} - \tau_{BC})} \tag{5}$$

where τ_{AC} and τ_{BC} are the path delays from A and B to C, respectively. By its known location, the phase component $e^{\mathbf{j}(\varphi_A - \varphi_B)}$ can be estimated by C. Then anchor C broadcasts the information to the network and each sensor will obtain

$$\begin{aligned} H_f = &\ r_n(r_C^* e^{-\mathbf{j}2\pi f(\tau_{AC} - \tau_{BC})}) \\ = &\ \sum_{i=0}^{P-1} \sum_{j=0}^{Q-1} \alpha_{A,i} \alpha_{B,j}^* e^{-\mathbf{j}2\pi f(\tau_{A,i} - \tau_{B,j})} \end{aligned} \tag{6}$$

where H_f can be viewed as a conjugate multiplication between the frequency-domain response from A and B to n at frequency point f, and the affect of the initial phases φ_A, φ_B and φ_n has been eliminated completely. Moreover, we can see that (6) has a similar format to (2). Here, we call the combination of a delay difference $\tau_{A,i} - \tau_{B,j}$ an equivalent time-difference (TD) path, and the $\alpha_{A,i}\alpha_{B,j}^*$ denotes the corresponding equivalent complex attenuation of this TD path. Furthermore, we call H_f as the time-difference (TD) frequency-domain channel response at frequency point f.

C. Range Estimation Based on Frequency-Hopping

From (6), we combine the TD paths with equal time-difference, and assume there are L_p distinct TD paths at the receive node. Denote α_k and τ_k, $(k = 0, 1, \cdots, L_p - 1)$, as the the equivalent complex attenuation and the time-difference of the kth TD path, respectively. As a note, the time-difference $\tau_{A,0} - \tau_{B,0}$ will reflect the range difference from the two emitting anchors to the receive node, which is produced by the two DLOS path $\tau_{A,0}$ and $\tau_{B,0}$, and we can call it direct time-difference (TD) path. Explicitly, the power of this direct TD path will be dominant among the TD paths. Although there is the possibility that the superposition of several reflected paths with equal time-difference has greater power than the direct TD path, in this paper we neglect this situation due to its quite low probability.

Then, consider the noise, we can rewrite (6) as

$$H_f = \sum_{k=0}^{L_p-1} \alpha_k e^{-\mathbf{j}2\pi f \tau_k} + w \tag{7}$$

where w is the additive noise with zero-mean. Taking the expectation for H_f, we will obtain

$$\tilde{H}_f = \mathbb{E}(H_f) = \sum_{k=0}^{L_p-1} \alpha_k e^{-\mathbf{j}2\pi f \tau_k} \tag{8}$$

In practice, the expectation can be obtained by the average of multiple response estimations. Afterwards, using frequency-hopping technique at L equally spaced frequencies over a given frequency band, i.e., $(f_l = f_0 + l\Delta f,\ l = 0, 1, \cdots, L-1)$, L samples of the TD frequency-domain channel response will be obtained at each sensor node, given by

$$\mathbf{x} = [x(0) \quad x(1) \quad \cdots \quad x(L-1)]^T \tag{9}$$

where $x(l) = \tilde{H}_{f_l}$. The vector $\mathbf{x}$ is also equivalent to a snapshot for the L_p source waves by a uniform linear array with L antenna elements. Here similar to the spatial smoothing technique in conventional DoA (Direction-of-Arrival) estimation method, we can divide the vector into consecutive segments of length M, given by

$$\mathbf{x}|_l^{l+M-1} = [x(l) \quad x(l+1) \quad \cdots \quad x(l+M-1)]^T \tag{10}$$

we can rewrite (10) into matrix form

$$\mathbf{x}|_l^{l+M-1} = \mathbf{V}\mathbf{\Phi}^l \mathbf{a} \tag{11}$$

where $\mathbf{V}$ is $M \times L_p$ vandermonde matrix, given by

$$\mathbf{V} = [\mathbf{v}(\tau_0) \quad \mathbf{v}(\tau_1) \quad \cdots \quad \mathbf{v}(\tau_{L_p-1})]$$

with

$$\mathbf{v}(\tau_k) = [1 \quad e^{-\mathbf{j}2\pi\Delta f \tau_k} \quad \cdots e^{-\mathbf{j}2\pi(M-1)\Delta f \tau_k}]^T$$

$$\mathbf{a} = [\alpha_0 e^{-\mathbf{j}2\pi f_0 \tau_0} \quad \alpha_1 e^{-\mathbf{j}2\pi f_0 \tau_1} \quad \cdots \quad \alpha_{L_p-1} e^{-\mathbf{j}2\pi f_0 \tau_{L_p-1}}]^T$$

$$\mathbf{\Phi} = \mathrm{diag}(e^{-\mathbf{j}2\pi\Delta f \tau_0} \quad e^{-\mathbf{j}2\pi\Delta f \tau_1} \quad \cdots \quad e^{-\mathbf{j}2\pi\Delta f \tau_{L_p-1}})$$

Furthermore, we construct two matrices consisting of rotation invariance as follows

$$\begin{aligned} \mathbf{X} &= [\mathbf{x}|_0^{M-1} \quad \mathbf{x}|_1^{M} \quad \cdots \quad \mathbf{x}|_{L-M-1}^{L-2}] = \mathbf{VS} \\ \mathbf{Y} &= [\mathbf{x}|_1^{M} \quad \mathbf{x}|_2^{M+1} \quad \cdots \quad \mathbf{x}|_{L-M}^{L-1}] = \mathbf{V\Phi S} \end{aligned} \tag{12}$$

where $\mathbf{S} = [\mathbf{a} \quad \mathbf{\Phi a} \quad \cdots \quad \mathbf{\Phi}^{L-M-1}\mathbf{a}]$. Then the autocorrelation matrix of $\mathbf{X}$ and the crosscorrelation matrix between $\mathbf{Y}$ and $\mathbf{X}$ are defined as

$$\begin{aligned} \mathbf{R}_{XX} &= \mathbf{X}\mathbf{X}^H = \mathbf{V}\mathbf{R}_{ss}\mathbf{V}^H \\ \mathbf{R}_{YX} &= \mathbf{Y}\mathbf{X}^H = \mathbf{V}\mathbf{\Phi}\mathbf{R}_{ss}\mathbf{V}^H \end{aligned} \tag{13}$$

where $\mathbf{R}_{ss} = \mathbf{S}\mathbf{S}^H = \sum_{l=0}^{L-M-1} \mathbf{\Phi}^l \mathbf{a}\mathbf{a}^H (\mathbf{\Phi}^H)^l$. Assume $\mathbf{V}$ and $\mathbf{S}$ are all column full rank, then $\mathbf{R}_{XX}$ will has rank L_p. Here we should note that in practical implementation, the true correlation matrix $\mathbf{R}_{XX}$ is unavailable in the present of noise, and the number L_p cannot be easily determined. However, classical information theoretic methods for model selection like Akaike's and Rissanen's criteria [12] can be used to estimate L_p. Thus, the Penrose-Moore pseudo-inverse of $\mathbf{R}_{XX}$ is

$$\mathbf{R}_{XX}^\dagger = \sum_{i=1}^{L_p} \frac{1}{\lambda_i} \mathbf{v}_i \mathbf{v}_i^H$$

where λ_i, $i = 1, 2, \cdots, L_p$, are the L_p largest eignvalues of $\mathbf{R}_{XX}$ and $\mathbf{v}_i$ are the corresponding eigenvectors. According to the method in [13], we construct an *auxiliary matrix* as follows

$$\mathbf{R} = \mathbf{R}_{YX}\mathbf{R}_{XX}^\dagger$$

Theorem 1: Given matrix $\mathbf{V}$ is column full-rank, $\mathbf{R}_{ss}$ is non-singular, the diagonal elements of $\mathbf{\Phi}$ are distinct, the *auxiliary* *matrix* $\mathbf{R}$ has its L_p non-zero eigenvalues equal to the L_p diagonal elements of $\mathbf{\Phi}$ and the corresponding eigenvectors equal to the L_p column vectors of $\mathbf{V}$, i.e.,

$$\mathbf{RV} = \mathbf{V\Phi}$$

The detailed proof can be found in [13]. Thus, after the eigenvalue decomposition (EVD) on $\mathbf{R}$, the L_p nonzero eigenvalues directly reflect the time-differences of the multiple TD paths. However, the direct TD path should be picked out from the L_p estimated multiple TD paths. Here we should emphasize that, in the convectional frequency-domain ToA estimation, the minimum path delay can be directly considered as the DLOS. However, in our proposed method, the direct TD path $\tau_{A,0} - \tau_{B,0}$ will conceal in the L_p TD paths estimated by subspace method, which is neither the largest nor the smallest delay. Thus, in this paper, we can only use the power information to identify the hidden direct TD path.

Specifically, after the EVD on $\mathbf{R}$, through the eigenvectors corresponding to the L_p nonzero eigenvalues, we can obtain the estimation for the signal subspace $\tilde{\mathbf{V}}$. Then, the dual space can be estimated given by $\mathbf{V}_{dual} = (\tilde{\mathbf{V}}^H\tilde{\mathbf{V}})^{-1}\tilde{\mathbf{V}}^H$. Afterwards, the power estimation can be achieved through the projection from the correlation matrix $\mathbf{R}_{XX}$ on the signal dual space $\mathbf{V}_{dual}$, given by

$$\frac{\mathrm{diag}(\mathbf{V}_{dual}\mathbf{R}_{XX}\mathbf{V}_{dual}^H)}{L-M} = [\alpha_0^2 \quad \alpha_1^2 \quad \cdots \quad \alpha_{L_p-1}^2] \qquad (14)$$

Then, the TD path with largest power indication can be picked out as the direct TD path.

Identifiability Issue: As pointed out in this section, for *Theorem 1*, the column full-rank of matrix $\mathbf{V}$ and the nonsingular of $\mathbf{S}$ should be satisfied. Since $\mathbf{V}$ and $\mathbf{S}$ are all vandermonde matrix, it is not hard for us to meet the requirements, as long as

$$M \geq L_p \quad \text{and} \quad L - M \geq L_p$$

yields

$$L \geq 2L_p \quad \text{and} \quad L_p \leq M \leq L - L_p$$

For determination of M, we can see that with large values of M, the potential for higher resolution of the subspace algorithm increases, which is similar to that in array signal processing where increasing M means an increase in subarray aperture and, thus, increase in resolution capability. On the other hand, for a given L, from (12), we can see that the increase in M will increases the fluctuations in the matrix $\mathbf{R}_{XX}$ and $\mathbf{R}_{YX}$, resulting in large perturbations of the eigenvalues and eigenvector of *auxiliary matrix* $\mathbf{R}$. Consequently, the value of M needs to be selected so that it provide a balance between resolution and stability of the algorithm.

As a note, another work in [14] has proposed a model-based ToA estimation method using frequency-hopping, and the multi-band signals are coherently combined in order to improve the resolution of the estimation. However, the phase randomness in the frequency-hopping process is not considered in [14].

D. TDoA Estimation with Frequency Diversity

If the sensor node support frequency hopping at several frequency band, then the frequency diversity can be utilized to improve the TDoA estimation performance. Assume over another frequency band, P equally spaced frequencies can be utilized, i.e., $f = \tilde{f}_0 + p\Delta f$, $(p = 0, 1, \cdots, P-1)$, and we denote $\tilde{\alpha}_k$, $k = 0, 1, \cdots, L_p - 1$, as the equivalent complex attenuation of the kth TD path over this frequency band. Then the corresponding P samples of the TD frequency-domain channel response can be obtained, given by

$$\tilde{\mathbf{x}} = [\tilde{x}(0) \quad \tilde{x}(1) \quad \cdots \quad \tilde{x}(P-1)]^T \qquad (15)$$

Similar to (10), we will obtain

$$\tilde{\mathbf{x}}|_p^{p+M-1} = [\tilde{x}(p) \quad \tilde{x}(p+1) \quad \cdots \quad \tilde{x}(p+M-1)]^T \quad (16)$$

which can be rewritten in matrix form

$$\tilde{\mathbf{x}}|_p^{p+M-1} = \mathbf{V\Phi}^p\tilde{\mathbf{a}} \qquad (17)$$

where

$$\tilde{\mathbf{a}} = [\tilde{\alpha}_0 e^{\mathbf{j}2\pi\tilde{f}_0\tau_0} \quad \tilde{\alpha}_1 e^{\mathbf{j}2\pi\tilde{f}_0\tau_1} \quad \cdots \quad \tilde{\alpha}_{L_p-1} e^{\mathbf{j}2\pi\tilde{f}_0\tau_{L_p-1}}]^T$$

Then, we can append new column vectors to matrix $\mathbf{X}$ and $\mathbf{Y}$ without changing the rotation invariance property between $\mathbf{X}$ and $\mathbf{Y}$. Specifically, for MATLAB notation, we will obtain

$$\begin{aligned}\mathbf{X} &= [\mathbf{X} \quad \tilde{\mathbf{x}}|_0^{M-1} \quad \tilde{\mathbf{x}}|_1^{M} \quad \cdots \quad \tilde{\mathbf{x}}|_{P-M-1}^{P-2}] = \mathbf{V}[\mathbf{S} \quad \tilde{\mathbf{S}}] \\ \mathbf{Y} &= [\mathbf{Y} \quad \tilde{\mathbf{x}}|_1^{M} \quad \tilde{\mathbf{x}}|_2^{M+1} \quad \cdots \quad \tilde{\mathbf{x}}|_{P-M}^{P-1}] = \mathbf{V\Phi}[\mathbf{S} \quad \tilde{\mathbf{S}}]\end{aligned} \qquad (18)$$

where $\tilde{\mathbf{S}} = [\tilde{\mathbf{a}} \quad \mathbf{\Phi}\tilde{\mathbf{a}} \quad \cdots \quad \mathbf{\Phi}^{P-M-1}\tilde{\mathbf{a}}]$. Afterwards, following the steps presented in the preview subsection, the improved estimation performance with frequency diversity will be achieved.

III. SIMULATION

A. TD frequency-domain Channel Response Estimation Performance

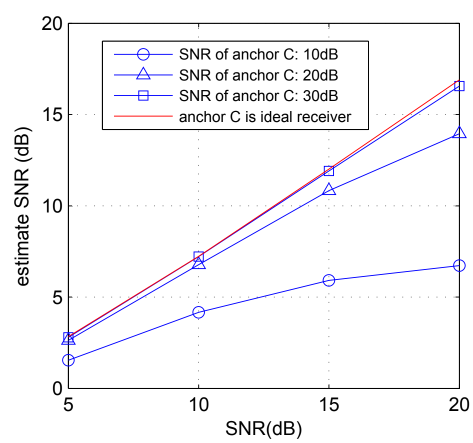

Fig. 2. Frequency-domain channel response estimation performance versus SNR with different SNR of anchor C.

Firstly, we will show some simulation results for our proposed estimation of the time-difference frequency-domain channel response. In simulation, we adopt the estimation

SNR to evaluate the estimation performance, which is defined by $10\log_{10}(H^2/\sigma^2)$, where H denotes the ideal values and σ denote the estimation standard deviation, given by $\sigma = \sqrt{\frac{\sum_{\rho=1}^{\Pi}|H_\rho - H|^2}{\Pi-1}}$ where H_ρ is the ρth sample of the estimation, Π is the number of the total samples. The estimation SNR versus SNR is shown in Fig.2. The three curves correspond to different SNR condition at anchor C, and the curve identified by red line represents that ideal receiver is adopted at anchor C. It is shown that the SNR increase at sensor node and anchor C will both improve the estimation performance. Since the anchor servers for all network, the increase of its receiver performance will introduce the estimation accuracy improvement of whole network. As a note, when SNR at anchor reaches 30dB, it is shown that the estimation performance is basically close to the performance when ideal receiver is equipped for anchor.

B. TDoA Estimation Performance

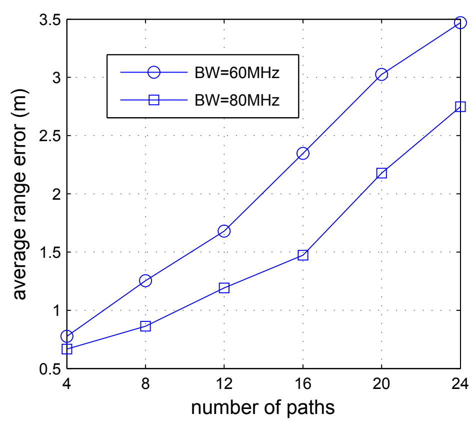

Fig. 3. Average range error versus number of TD paths with different frequency-hopping bandwidth.

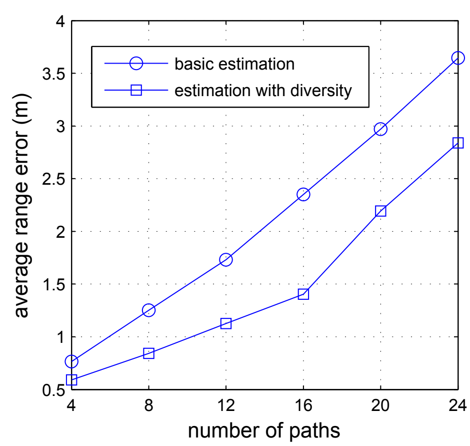

Fig. 4. Average range error versus number of TD paths.

In this example, the range difference from the two anchors to the receive node obeys uniform distribution: $\mathrm{U}(-100\mathrm{m}, 100\mathrm{m})$. We set the amplitude of the direct TD path is 1, and the amplitude of other reflected TD paths obeys uniform distribution: $\mathrm{U}(0.1, 0.5)$. Assume the reflected TD paths are distributed in the both sides of direct TD path with equal spacing 10ns. In simulation, we assume the estimation SNR for the TD frequency-domain channel response is 5dB, and 64 samples are averaged to regard as the expectation of the sample. We set the frequency hopping interval $\Delta f = 1$MHz and $M = 30$.

Fig.3 shows the average range error versus number of TD paths L_p with different frequency-hopping bandwidth. It is shown that, the increase of TD paths L_p will introduce more variables to estimate, which will inevitably deteriorate the estimation performance. On the other hand, the increase of the hopping bandwidth will improve the performance.

In Fig.4, we shows the performance comparison between the basic method and the improved method with frequency diversity, where another 60MHz hopping band is utilized in latter. As expected, when the frequency diversity is utilized, the estimation performance is improved.

IV. Conclusions

This paper presents a novel TDoA estimation method for wireless sensor networks in multipath environment. The frequency hopping technique is utilized and the super-resolution TDoA estimation is conducted independently at each sensor to detect range difference from the sensor to the two anchors. The analysis and simulations demonstrate the effectiveness of the proposed method.

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