J Electr Eng Technol Vol. 8, No. 3: 638-648, 2013
http://dx.doi.org/10.5370/JEET.2013.8.3.638

ISSN(Print) 1975-0102
ISSN(Online) 2093-7423

A Novel Hearability Enhancement Method for Forward-Link Multilateration Using OFDM Signal

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Abstract – Together with the GPS-based approach, geo-location through mobile communication networks is a key technology for location-based service. To save the cost, most geo-location system is implemented on the existed network service, which has a cellular structure. Still, multilateration is limited in cellular structure because it is difficult for the mobile terminal to acquire distance measurements from multiple base stations. This low hearability in the receiver is caused by co-channel interference and multipath environment. Therefore, hearability enhancement is necessary for multilateration under multipath and interference environment. Former time domain based hearability methods were designed for real signals. However, orthogonal frequency division multiplexing (OFDM) signal, which its usage has been increased in digital wireless communication, is a complex signal. Thus, different hearability enhancement method is needed for OFDM signals. This paper proposes a hearability enhancement method for forward-link multilateration using OFDM signals, which employ interference cancellation and multipath mitigation. A novel interference cancellation and multipath mitigation strategy for complex-valued OFDM signals is presented that has an iterative structure. Simulation results show that the proposed multilateration method provides the user's position with an accuracy of less than 80m through the mobile WiMAX cellular network in multipath environment.

Keywords: OFDM, geo-location, Multilateration, Interference cancellation, Hearability

1. Introduction

The global positioning system (GPS) is a widely used satellite navigation system that provides location information with 2DRMS errors of 18 m for L1 C/A code. Based on the GPS modernization plan, location errors are expected to be reduced to 5 to 6 m due to additional usage of L2C code [1-2]. However, the received signal power at the GPS receiver is very weak, which is less than -130 dBm. Therefore, users cannot find their position when signal blocking occurs, especially in urban areas. To resolve the shadow region problem, alternative methods have been attempted, such as integrating GPS with other positioning systems, and GPS sensitivity enhancement technique using mobile cellular networks [3]. The major positioning approaches incorporated with GPS include dead-reckoning (DR) with self-contained inertial/non-inertial sensors, and geo-location using mobile cellular communication networks. In particular, the cell identification (CID) approach based on the cell radius of the serving base station (BS), and multilateration approach based on signals from multiple BSs are commonly used in geo-location. Multilateration provides uniform accuracy for a large cell radius, but it requires the signals from at least three BSs for 2-D positioning [4]. Examples of multilateration used in existing cellular networks include enhanced observed time difference (E-OTD) in GSM, observed time difference of arrival (OTDOA) in universal mobile telecommunications system (UMTS), and advanced forward-link trilateration (AFLT) in IS-95 and cdma2000 [5-6].

Since it is difficult for a mobile station (MS) to acquire signals from multiple BSs in most of the service area except the edge of the cell, it is necessary to improve hearability so that multilateration is possible everywhere in the network. Poor hearability in the forward link is mainly due to path loss, co-channel interference (CCI) or inter-cell interference, and multipath; therefore, improving the signal-to-noise ratio (SNR) and reducing the interference together with multipath are key challenges in enhancing hearability for forward link multilateration. The SNR can be enhanced by long-integration techniques [1, 7, 8], and CCI can be lessen by interference cancellation methods [4, 9] or the inter-cell interference avoidance method [10]. Because the inter-cell interference avoidance method mitigates interference by managing spectral, temporal, or spatial resources, it cannot be directly applied without modifying the existing cellular network system. In the interference cancellation technique, interferences are estimated by minimum mean square error (MMSE) or

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Received: September 18, 2010; Accepted: September 19, 2011

maximum likelihood (ML), and then eliminated from the raw measurements. In dense multipath environment, the performance of interference cancellation technique is degraded, thus multipath mitigation method is needed. Multipath can be mitigated by estimating the channel [26], using a correlator technique in the tracking system [28], or by implementing a cancellation technique similar to interference cancellation [29]. Because most of the existing multipath mitigation and interference cancellation techniques for cellular system have been developed for data decoding, they usually do not estimate propagation time, which is essential for geo-location.

Orthogonal frequency-division multiple access (OFDM) modulation, which was proposed in mid-1960s, had the advantage in mobile digital communication in multipath environment [25]. Due to its high performance when combined with coding modulation, the OFDM became a popular modulation technique. The usage of OFDM has spread to digital television broadcasting, Wireless Local Area Network (WLAN), Long Term evolution (LTE), and mobile Worldwide Interoperability for Microwave Access (WiMAX) and is expected to grow more in its scale. However, most of the researches in OFDM signals were concentrated in enhancing the data communication performance. When the signal is used for positioning purpose, the receiver suffers from hearabilty in time domain due to multipath and CCI.

While interference cancellation methods in [26-29] are applicable to W-CDMA or CDMA2000 cellular systems, they are not suited for OFDM signals, which it is complex-valued. Interference cancellation methods for orthogonal frequency-division multiplex (OFDM) communication systems have been proposed previously. However, most of these works have been focused on frequency-domain inter-carrier or inter-cell interference avoidance for data decoding [12, 13]. Other methods, such as the forward-link hyperbolic geo-location method using pilot sub-carrier in the data symbol were proposed for OFDM communication systems [14]. Still, the complexity of inter-cell interference cancellation proposed in [14] is very high because the time delays of interferences are estimated in the time-domain, and other parameters are obtained in the frequency-domain.

This paper proposes the forward-link hyperbolic geo-location method using OFDM preamble symbols. Because the proposed method uses the preamble signals in the forward-link channel, it is easy to implement geo-location without modifying the existing network equipment. Furthermore, hearability will be enhanced through the interference cancellation technique and the coherent long integration method. In particular, the time-domain interference cancellation method for complex-valued OFDM signal is proposed in Section 2. To apply interference cancellation to geo-location, time delays of interferences are estimated as well as their amplitudes and phases. Considering the computation, an iterative interference cancellation scheme using simplified ML is used in the proposed method. The performance of the proposed method and geo-location using the OFDM signal are analyzed in Section 3 using the mobile WiMAX network.

2. Enhancement of Hearability in Geo-location Using OFDM Signal

When a cellular network is being constructed, cell planning is performed for the suitable allocation of the BSs, considering the traffic and the cost. Accordingly, when the MS is inside the service cell, the signals from the neighbor BSs are weakly received or cannot be detected. Since the main service areas of cellular are urban areas surrounded by tall buildings, the signal strength is dramatically decreased as the distance between the BS and the MS increases. Furthermore, signals from the service cell BS and multipath signals are interference when neighbored signals are needed. It is necessary to increase hearability so that enough BSs are detected to estimate the location by multilateration. Using preamble signals for the multilateration can be a reasonable choice, since it enables the separation of BS signal and its periodic property makes coherent integration possible. Hearability is enhanced by signal integration and multipath concerned interference cancellation.

2.1 Iterative interference cancellation for hearability enhancement in multipath environment

When a signal is transmitted in a cellular network, the signals from other users or BSs sharing the channel act as an interference and this is called CCI. The hearability of the long-distance BS is decreased by CCI; therefore, the development of CCI avoidance techniques has been important in the area of wireless communication. Forward link CCI avoidance methods include the MIMO technique, beam forming, and the interference cancellation method [9], [16]. Both MIMO and beam forming, which use multiple antenna elements, are incongruent for general receivers where single antenna is common.

Forward link interference cancellation commonly exploits the iterative structure as shown in Fig. 1 [9]. The received signal measurement contains the preamble from multiple BSs, and some of the apparent BS signals are detected from the correlator output. Then the coefficients of the detected signals are estimated and eliminated from

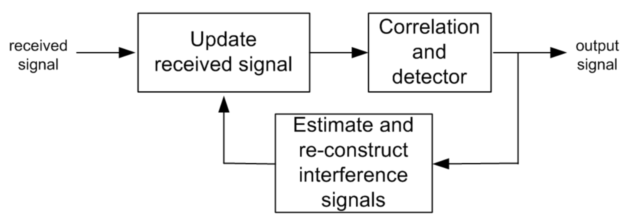


Fig. 1. Structure of iterative interference cancellation.

the measurement. Afterwards, the next visible BS signals are viable for detection, and the same process can be repeated to reduce the influence of interference sequentially.

In the multipath environment, the time shifted multipath components of CCI cause more performance degradation in hearability. In this paper, iterative interference cancellation is performed by considering the CCI and multipath at the same time. Several interference signal estimation methods have been proposed in the past, including ML, zero forcing (ZF), MMSE, and sub-space methods, such as multiple signal classification (MUSIC) [8-9]. The estimations based on ML or sub-space methods provide superior performance, but they have high computational complexity.

In the cellular network based on OFDM signals, sending data are converted by IFFT, thus the transmitted BS signals are complex-valued. Assuming that signal transmitted from each BS has a normalized average power, the baseband measurement equation containing the signals from all neighbor BSs is written as

$$r(t)=\sum_{i=1}^{L}\sum_{q=1}^{M_i}\alpha_{iq}s_i\left(t-\tau_{iq}\right)e^{j\left(\theta_{iq}+2\pi\Delta f_{iq}t\right)}+n(t), \qquad (1)$$

where $s_i(t)$ is the signal transmitted from i^{th} BS, the iq subscript indicate the q^{th} multipath component of the i^{th} BS, α_{iq} is the magnitude of the received signal, τ_{iq} is the delay time, θ_{iq} is the carrier phase, Δf_{iq} indicates the residual frequency in the baseband signal, and L is the number of detected BS signals, M_i is the number of multipath component in the i^{th} BS signal. The clock frequency of all BS is assumed to be perfectly synchronized. Here, $n(t)$ is the complex white Gaussian noise with a variance of σ^2 [17].

When the measurement of (1) is discretized with the sampling frequency of $f_s=1/T_s$, the correlator output for the l^{th} BS preamble signal is denoted by

$$\begin{aligned}c_l\left(nT_S\right)&=\frac{1}{N}\sum_{m=1}^{N}r\left(mT_s\right)s'_l{}^*\left(\left(n+m\right)T_s\right)\\&\cong\frac{1}{N}\sum_{i=1}^{L}\sum_{q=1}^{M_i}\sum_{m=1}^{N}\alpha_{iq}s_i\left(mT_s-\tau_{iq}\right)s'_l{}^*\left(\left(n+m\right)T_s\right)e^{j\theta_{iq}}\\&\quad+\frac{1}{N}\sum_{m=1}^{N}n\left(mT_s\right)s'_l{}^*\left(\left(n+m\right)T_s\right)\\&\equiv\sum_{q=1}^{M_l}\alpha_{lq}\rho_{ll}\left(nT_S-\tau_{lq}\right)e^{j\theta_{lq}}\\&\quad+\sum_{i=1,i\neq l}^{L}\sum_{q=1}^{M_l}\alpha_{iq}\rho_{il}\left(nT_S-\tau_{iq}\right)e^{j\theta_{iq}}+n_l\left(nT_S\right), \qquad (2)\end{aligned}$$

where N is the number of samples for one preamble symbol, $s'_l(\cdot)$ is the replica of l^{th} BS signal generated by the receiver, '*' means the complex conjugate, $\rho_{il}(\cdot)$ is the complex-valued correlation function between the i^{th} BS signal and the l^{th} replica, and $n_l(\cdot)$ represents the complex white Gaussian noise with the variance of $\sigma_c{}^2=\sigma^2/N$. Assuming that residual frequencies are estimated, it was ignored. In (2), the first term of the last row represents the auto-correlation function of l^{th} signal. It is well known that $\rho_{ll}(-\tau)=\rho_{ll}{}^*(\tau)$ and $\rho_{ll}(\tau)$ has the maximum value of 1 at $\tau=0$.

Conventionally, signal detection is determined by the SNR. If the discretized auto-correlation has the maximum value at $\tau_{l1}=nT_S$, and has a sufficiently large sampling rate, then $\left|\rho_{ll}\left(nT_S-\tau_{l1}\right)\right|\cong 1$, where $SNR_l\cong\alpha_{l1}^2/\sigma_c^2$. Assuming that $SNR_1>>SNR_2>\lambda_{Th}>SNR_3>\cdots>SNR_L$ where λ_{Th} is the detection threshold, BS1 and BS2 should be detected. However, if multiple BS signals share a single channel, signal detection in the receiver can be affected by interference signal and by its multipath. To consider interference in signal detection, it is desirable to employ the signal-to-noise-plus-interference ratio (SNIR), which is defined by

$$SNIR_l=\frac{\alpha_{l1}^2}{\sigma_c^2+\sum_{\substack{i=1\\i\neq l}}^{L}\sum_{q=1}^{M_i}\left|\alpha_{iq}\rho_{il}\left(nT_S-\tau_{iq}\right)\right|^2}. \qquad (3)$$

In the equation, the second term of the denominator denotes the interference to the l^{th} BS signal. By selecting an appropriated pseudo random code for the preamble, the cross-correlation suppression ratio (CCSR) of the preamble becomes high. Thus, for BS1, the SNIR can be approximated by $SNIR_1\cong\alpha_{11}^2/\sigma_c^2=SNR_1$. However, if the signal power of BS1 is high so that $\alpha_1^2\rho_{12}^2(\cdot)$ is sufficiently large, $SNIR_2\cong\alpha_{21}^2/\left(\sigma_c^2+\sum_{q=1}^{M_1}\alpha_{1q}^2\rho_{12}^2(\cdot)\right)<SNR_2$ is satisfied for BS2. In this case, SNIR_2 can be smaller than λ_{Th}, which makes it impossible to detect the signal of BS2.

When interference cancellation technique is applied, the SNIR value will approach the SNR. With the iterative interference cancellation shown in Fig. 1, interference can be effectively eliminated by estimating the signal and its multipath coefficients from the strongest BS's signal, and then estimating the next strongest ones sequentially. Therefore, interference cancellation can be considered as a parameter estimation problem, whose evaluation function is denoted by

$$\begin{aligned}\left[\hat{\alpha}_{iq},\hat{\tau}_{iq},\hat{\theta}_{iq}\right]=\min_{\hat{\alpha}_{iq},\hat{\tau}_{iq},\hat{\theta}_{iq}}\sum_{n}^{N}&\Big|\alpha_{iq}\rho_{ii}\left(nT_S-\tau_{iq}\right)e^{j\theta_{iq}}\\&-\hat{\alpha}_{iq}\rho_{ii}\left(nT_S-\hat{\tau}_{iq}\right)e^{j\hat{\theta}_{iq}}\Big|^2 \qquad (4)\end{aligned}$$

For multiple parameter estimation problem, ML or the maximum a posteriori (MAP) method shows excellent performance, but these methods have high computational complexity [19]. In this paper, parameter estimation using a simplified ML method is proposed for complex-valued preamble signal.

Assume that the parameters of signals from BS1 to BS $(l-1)$ are estimated sequentially, and then the estimated signal waveforms are eliminated from (1) using iterative interference cancellation. When the compensated baseband measurement is fed into the correlator for the l^{th} BS signal, its output is then written as

$$c_l(nT_S) = \sum_{q=1}^{M_l} \alpha_{lq}\rho_{ll}(nT_S - \tau_{lq})e^{j\theta_{lq}} + \sum_{k=1}^{l-1} err_k(nT_S) + \sum_{i=l+1}^{L}\sum_{q=1}^{M_i} \alpha_{iq}\rho_{il}(nT_S - \tau_{iq})e^{j\theta_{iq}} + n_l(nT_S), \quad \mathrm{n} = 1,2,\ldots,N, \tag{5}$$

where $err_k(\cdot)$ is the residual error that remains after the elimination of the k^{th} interference. If the interference estimation is successfully done, it can be assumed that $err_k(\cdot) \cong 0$. Also, the third term of the right side of (5) is negligible because $\alpha_{l1} > \alpha_{l+1,1} > \cdots > \alpha_{L1}$, and CCSR is sufficiently large.

2.2 Interference signal estimation method for complex OFDM signal in multipath environment

Under the condition that the SNR of the preamble signal is sufficiently large, the main peak of the autocorrelation is an impulse-like function. If a correlation peak exists at $\tau_{lq} = k_l T_S + \delta\tau_{lq}$, the estimation problem in (4) can be approximated to minimizing the estimation error around the correlation peak, and it is written as

$$\left[\hat{\alpha}_{lq}, \hat{\tau}_{lq}, \hat{\theta}_{lq}\right] = \min_{\hat{\alpha}_{lq},\hat{\tau}_{lq},\hat{\theta}_{lq}} \sum_{n=k-N_p}^{k+N_p} \left| \alpha_{lq}\rho_{ll}(nT_S - k_lT_S - \delta\tau_{lq})e^{j\theta_{lq}} - \hat{\alpha}_{lq}\rho_{ll}(nT_S - k_lT_S - \delta\tau_{lq})e^{j\hat{\theta}_{lq}} \right|^2, \tag{6}$$

where $\delta\tau_{lq}$ is the mismatch error induced by discretization, N_p is the number near the peak that is used in simplified ML. If the second and the third terms are ignored and when only the largest multipath component is considered in (5), the correlator output equation in vector form with $(2N_p+1)$ samples around the peak is expressed as

$$\begin{aligned}\underline{c}_{lq} &= \left[c_{lq}((k_{lq}-N_P)T_S) \;\; c_{lq}((k_{lq}-N_P+1)T_S) \;\cdots\; c_{lq}((k_{lq}+N_P)T_S)\right]^T \\ &\cong \alpha_{lq}\underline{\rho}_{ll}(\delta\tau_{lq})e^{j\theta_{lq}} + \underline{n}_l \\ &= \alpha_{lq}e^{j\theta_{lq}}\left[\rho_{ll}(-N_PT_S - \delta\tau_{lq})\cdots\rho_{ll}((N_PT_S - \delta\tau_{lq}))\right]^T \\ &\quad + \left[n_l((k_l - N_P)T_S)\cdots n_l((k_l+N_P)T_S)\right]^T,\end{aligned} \tag{7}$$

where $\underline{\rho}_{ll}(\delta\tau)$ is the $(2N_p+1)\times 1$ autocorrelation vector, $\underline{c}_{lq}(\cdot)$ is the correlator output of the q^{th} multipath component in l^{th} BS. According to the property of OFDM auto-correlation, the real part of $\underline{\rho}_{ll}(\delta\tau)$ is even, and the imaginary part is odd with respect to the center element. By using the exchange matrix J, $\underline{\rho}_{ll}(\delta\tau)$ can be expressed as $\mathrm{Re}[\underline{\rho}_{ll}(\delta\tau)] = J\cdot\mathrm{Re}[\underline{\rho}_{ll}(\delta\tau)]$, and $\mathrm{Im}[\underline{\rho}_{ll}(\delta\tau)] = -J\cdot\mathrm{Im}[\underline{\rho}_{ll}(\delta\tau)]$.

In (7), the probability density function (PDF) of the noise $\underline{n}_l$ is written as

$$p_r(\underline{c}_{lq};\alpha,\delta\tau,\theta) = \frac{1}{(2\pi)^{(2N_P+1)/2}|V_n|^{1/2}}\cdot\exp\left\{-\frac{1}{2}\left(\underline{c}_{lq} - \alpha_{lq}\underline{\rho}_{ll}(\delta\tau_{lq})e^{j\theta_{lq}}\right)^H V_n^{-1}\left(\underline{c}_{lq} - \alpha_{lq}\underline{\rho}_{ll}(\delta\tau_{lq})e^{j\theta_{lq}}\right)\right\}, \tag{8}$$

where V_n is $(2N_p+1)\times(2N_p+1)$ complex covariance matrix of $\underline{n}_l$ that is given as in [15] by

$$\begin{aligned}V_n &= \frac{\sigma^2}{N_P}\begin{bmatrix} \rho_{ll}(0) & \rho_{ll}(T_S) & \cdots & \rho_{ll}((2N_P)T_S) \\ \rho_{ll}(-T_S) & \rho_{ll}(0) & \cdots & \rho_{ll}((2N_P-1)T_S) \\ \vdots & \vdots & \ddots & \vdots \\ \rho_{ll}(-(2N_P)T_S) & \rho_{ll}(-(2N_P-1)T_S) & \cdots & \rho_{ll}(0)\end{bmatrix} \\ &= \frac{\sigma^2}{N_P}\begin{bmatrix} 1 & \rho_{ll}(T_S) & \cdots & \rho_{ll}((2N_P)T_S) \\ \rho_{ll}^*(T_S) & 1 & \cdots & \rho_{ll}((2N_P-1)T_S) \\ \vdots & \vdots & \ddots & \vdots \\ \rho_{ll}^*((2N_P)T_S) & \rho_{ll}^*((2N_P-1)T_S) & \cdots & 1\end{bmatrix}.\end{aligned} \tag{9}$$

Because V_n is positive definite and Hermitian Toeplitz matrix for the complex-valued noise, V_n^{-1} is also positive definite and Hermitian per-symmetric matrix [17, 20, 21].

If we define the exponential part of (8) as a likelihood function $F(\alpha_{lq},\delta\tau_{lq},\theta_{lq})$, the estimated parameter $\hat{\alpha}_{lq}$, $\delta\hat{\tau}_{lq}$, $\hat{\theta}_{lq}$ can be obtained by maximizing the likelihood function that satisfies the following equation:

$$\left.\frac{\partial}{\partial\alpha_{lq}}F\right|_{\alpha_{lq}=\hat{\alpha}_{lq}} = \left.\frac{\partial}{\partial\delta\tau_{lq}}F\right|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}} = \left.\frac{\partial}{\partial\theta_{lq}}F\right|_{\theta_{lq}=\hat{\theta}_{lq}} = 0. \tag{10}$$

From (8) and (10), $\delta\hat{\tau}_{lq}$ should satisfy the following equation:

$$0 = \left.\frac{\partial}{\partial\delta\tau_{lq}}F\right|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}} = \alpha_{lq}\left(\underline{c}_{lq}^H V_n^{-1}\frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}(\delta\tau_{lq})\right)e^{-j\theta_{lq}}\right.$$

$$-\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)^{H} V_n^{-1} \frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)\Bigg)\Bigg|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}. \quad (11)$$

As explained above, the autocorrelation peak of a preamble at $\tau_{lq} = k_l T_S + \delta\tau_{lq}$ can be modeled as an impulse-like function denoted by

$$\rho_{ll}\left(k_l T_S - \delta\tau_{lq}\right) = 1 - \beta\left(k_l T_S - \delta\tau_{lq}\right)^2 + j\gamma\left(k_l T_S - \delta\tau_{lq}\right). \quad (12)$$

Because $\mathrm{Im}[\underline{x}^H] = -(\mathrm{Im}[\underline{x}])^T$ for complex-valued vector $\underline{x}$, the real part of the second term on the right side of (11) is written as

$$\mathrm{Re}\left\{\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)^{H} V_n^{-1} \frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)\right\}\Bigg|_{\tau_{lq}=\hat{\tau}_{lq}} =$$
$$\left[\mathrm{Re}\left\{\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right\}\right]^T \left[\mathrm{Re}\left\{V_n^{-1}\right\}\right]\left[\mathrm{Re}\left\{\frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)\right\}\right]\Bigg|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}$$
$$-\left[\mathrm{Re}\left\{\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right\}\right]^T \left[\mathrm{Im}\left\{V_n^{-1}\right\}\right]\left[\mathrm{Im}\left\{\frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)\right\}\right]\Bigg|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}$$
$$+\left[\mathrm{Im}\left\{\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right\}\right]^T \left[\mathrm{Re}\left\{V_n^{-1}\right\}\right]\left[\mathrm{Im}\left\{\frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)\right\}\right]\Bigg|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}$$
$$+\left[\mathrm{Im}\left\{\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right\}\right]^T \left[\mathrm{Im}\left\{V_n^{-1}\right\}\right]\left[\mathrm{Re}\left\{\frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)\right\}\right]\Bigg|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}. \quad (13)$$

From (7) and (12), $\mathrm{Re}\left\{\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right\}\Big|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}$ and $\mathrm{Im}\left\{\frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)\right\}\Big|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}$ are even-symmetric vectors. At the same time, $\mathrm{Im}\left\{\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right\}\Big|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}$ and $\mathrm{Re}\left\{\frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)\right\}\Big|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}$ are odd-symmetric. These even- or odd-symmetric vectors satisfy the following properties.

Theorem 1. *For a skew-symmetric per-symmetric matrix P_1, and even-symmetric (or odd-symmetric) vectors $\underline{x}_1$, $\underline{x}_2$*, the relation $\underline{x}_1^T P_1 \underline{x}_2 = \mathbf{0}$ holds. ■

Proof. From *the* definition of per-symmetric matrix, $P_1 J = J P_1^T$ is satisfied, where J is an exchange matrix. By using the property $P_1^T = -P_1$, and $J = J^{-1}$, the relation $P_1 = -J P_1 J$ can be obtained. Consequently, for even-symmetric vectors $\underline{x}_1$, $\underline{x}_2$, the relation leads to $\underline{x}_1^T P_1 \underline{x}_2 = -\underline{x}_1^T J P_1 J \underline{x}_2 = -\underline{x}_1^T P_1 \underline{x}_2$, which implies $\underline{x}_1^T P_1 \underline{x}_2 = 0$. Similarly, $\underline{x}_1^T P_1 \underline{x}_2 = 0$ is obtained for odd-symmetric vectors $\underline{x}_1$, $\underline{x}_2$. ■

Theorem 2. *For a bi-symmetric (symmetric per-symmetric) matrix P_2, even-symmetric vector $\underline{x}_1$, and odd-symmetric vector $\underline{x}_2$, the relation $\underline{x}_1^T P_2 \underline{x}_2 = \underline{x}_2^T P_2 \underline{x}_1 = 0$ holds.* ■

Proof. By the property of a bi-symmetric matrix, $P_2 = J P_2 J$ is satisfied. Then the relation $\underline{x}_1^T P_2 \underline{x}_2 = \underline{x}_1^T J P_2 J \underline{x}_2 = -\underline{x}_1^T P_2 \underline{x}_2$ is obtained for even-symmetric vector $\underline{x}_1$, and odd-symmetric vector $\underline{x}_2$. This leads to $\underline{x}_1^T P_2 \underline{x}_2 = 0$ and $\underline{x}_2^T P_2 \underline{x}_1 = 0$. ■

Because V_n^{-1} is a Hermitian per-symmetric matrix, from Theorem 1 and 2, four terms in (13) become 0, respectively. Consequently, the real part of (11) is written as

$$\mathrm{Re}\left\{\underline{c}_{lq}^H V_n^{-1} \frac{d}{d\delta\tau_{lq}}\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)\Bigg|_{\delta\tau_{lq}=\delta\hat{\tau}_{lq}}\right\} = \mathbf{0}. \quad (14)$$

Applying (12) into (14), the estimate of $\delta\tau_{lq}$ is obtained as

$$\delta\hat{\tau}_{lq} = \frac{\begin{bmatrix} 2\beta\left(\mathrm{Re}[\underline{c}_{lq}^H]\mathrm{Re}[V_n^{-1}] - \mathrm{Im}[\underline{c}_{lq}^H]\mathrm{Im}[V_n^{-1}]\right)\underline{t} \\ -\gamma\left(\mathrm{Re}[\underline{c}_{lq}^H]\mathrm{Im}[V_n^{-1}] - \mathrm{Im}[\underline{c}_{lq}^H]\mathrm{Re}[V_n^{-1}]\right)\underline{1} \end{bmatrix}}{2\beta\left(\mathrm{Re}[\underline{c}_{lq}^H]\mathrm{Re}[V_n^{-1}] - \mathrm{Im}[\underline{c}_{lq}^H]\mathrm{Im}[V_n^{-1}]\right)\underline{1}}, \quad (15)$$

where $\underline{1} = \begin{bmatrix}1 & \cdots & 1\end{bmatrix}^T$ and $\underline{t} = \begin{bmatrix}-N_P T_S & \cdots & N_P T_S\end{bmatrix}^T$. Note that the estimate of delay time is given by $\hat{\tau}_{lq} = kT_S + \delta\hat{\tau}_{lq}$. In a similar manner, the estimate of θ_{lq} is obtained as

$$\hat{\theta}_{lq} = -\tan^{-1}\left(\frac{\mathrm{Im}\left\{\dfrac{\underline{c}_{lq}^H V_n^{-1} \underline{\rho}_{ll}\left(\delta\tau_{lq}\right)}{\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)^H V_n^{-1} \underline{\rho}_{ll}\left(\delta\tau_{lq}\right)}\right\}}{\mathrm{Re}\left\{\dfrac{\underline{c}_{lq}^H V_n^{-1} \underline{\rho}_{ll}\left(\delta\tau_{lq}\right)}{\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)^H V_n^{-1} \underline{\rho}_{ll}\left(\delta\tau_{lq}\right)}\right\}}\right). \quad (16)$$

To resolve $\hat{\theta}_{lq}$ in (16), $\delta\tau_{lq}$ should be determined first. One way is to use the estimate $\delta\hat{\tau}_{lq}$ obtained from (15). Another approach is to reduce the dimension of vector in (7). Assuming that the sampling rate is sufficiently large, $\rho_{ll}\left(\delta\tau_{lq}\right) \cong 1$. When $N_P = 0$, (16) is approximated by

$$\hat{\theta}_{lq} \cong -\tan^{-1}\left(\frac{\mathrm{Im}\left\{ \left(c_{lq}(0) \right)^H \right\}}{\mathrm{Re}\left\{ \left(c_{lq}(0) \right)^H \right\}} \right). \tag{17}$$

The estimate of α_{lq} is given by

$$\hat{\alpha}_{lq} = \frac{\underline{c}_{lq}^H V_n^{-1} \underline{\rho}_{ll}\left(\delta\tau_{lq}\right) e^{-j\theta_{lq}}}{\left(\underline{\rho}_{ll}\left(\delta\tau_{lq}\right)\right)^H V_n^{-1} \underline{\rho}_{ll}\left(\delta\tau_{lq}\right)}. \tag{18}$$

Since $\delta\tau_{lq}$ and θ_{lq} are unknown in (18), the sub-optimal estimate for α_{lq} can be obtain using $\delta\hat{\tau}_{lq}$ and θ_{lq}.

Long integration and interference cancellation can improve the SNR and signal-to-interference ratio (SIR), respectively [7]. When noncoherent integration is used, it is difficult to apply the interference cancellation technique, because the carrier phase is eliminated by a squaring operation. For this reason, only coherent integration and iterative interference cancellation are used, which increase the SNIR effectively and enhance the hearability.

With coherent long integration and the proposed interference cancellation, the SNIR of weak signals in the presence of CCI can be improved efficiently. Consequently, it is expected that the sufficient number of measurements for geo-location is obtained.

3. Simulation

To analyze the hearability enhancement performance using OFDM signal, simulation was performed base on mobile WiMAX network system. Mobile WiMAX is the mobile Internet service of the IEEE 802.16e international standard and has been chosen as the technical standard of 3G mobile communication [11]. Mobile WiMAX has a cellular structure like a traditional mobile communication system. Thus, for geo-location using the mobile WiMAX network, hearability also should be improved as in the traditional cellular network.

The physical profile of mobile WiMAX with a 10 MHz bandwidth has a frame length fixed at 5 ms, and the frame is composed of 42 OFDMA symbols [11]. Each symbol on the subcarriers is converted to time domain by 1024-sized inverse fast Fourier transform (IFFT) procedures. Then the cyclic prefix (CP) of 12.8 μs is added. As a result, the time duration of one symbol becomes 115.2 μs.

The MS can find out the existence of a BS by searching the preamble located at the first symbol of each frame. Also, the preamble is used for time synchronization, frequency-offset compensation, and channel estimation between the BS and MS. A preamble symbol uses one of three segments in the frequency domain as shown in Fig. 2, and one segment consists of 284 subcarriers. To distinguish the preambles transmitted by different BSs or sectors, a 284-sized pseudo-random noise (PN) sequence is assigned to the segment, and 114 sets of PN sequences are given in the standard [11]. The autocorrelation of the transmitted preamble symbol has a main peak and three left and right side peaks, respectively, as shown in Fig. 3. The CCSR [18] of the preamble used in the mobile WiMAX network is 18.23 dB for the same segment group and 27.3 dB for different segment groups.

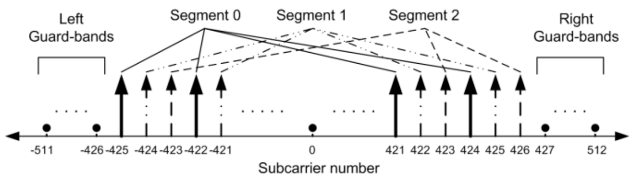


Fig. 2. Sub-carrier allocation structure of mobile WiMAX preamble.

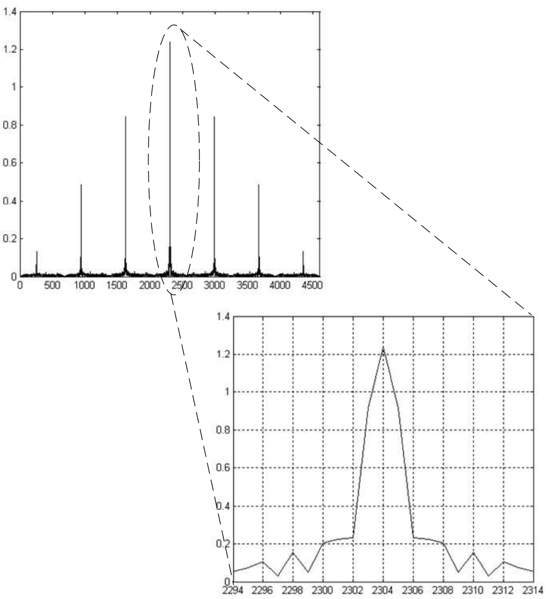


Fig. 3. Autocorrelation of the preamble symbol.

By the mutually orthogonal PN sequence, the preambles of BSs can be discriminated in the mobile WiMAX system, while they share the same physical channel. The MS can easily estimate its location by multilateration if the the range measurements are obtained from the propagation time of the preamble signal. In mobile WiMAX network, one-way ranging can be applied, since the BSs are synchronized with the GPS clock, and every frame is transmitted simultaneously by all the BSs. However, the mobile WiMAX system does not transmit the time information from the BS to the MS, which implies that the time-difference-of-arrival (TDOA) method is more appropriate than the time-of-arrival (TOA) method. Each BS broadcasts a preamble at every frame, which enables the continuous positioning. Thus, long integration method can be easily implemented for the enhancement of sensitivity.

Through computer simulation, the performance of the proposed hearability enhancement algorithm is evaluated and the positioning performance of geo-location based on mobile WiMAX is analyzed. The structure of the geo-location simulator is shown in Fig. 4.

The transmit signal is generated according to the mobile WiMAX standard in [11]. Among the ITU-R IMT-2000 channel models, the 'outdoor to indoor and pedestrian test environment' is chosen for the path loss model [22]. The BSs are placed to establish a cellular structure with a cell radius of 500 m as shown in Fig. 5. It is also assumed that each BS is partitioned into three sectors, and a directional antenna is used with a 10 dB beam width of 120°, and a maximum attenuation of -30 dB [23]. To receive data at the edge of the cell, the transmission power is determined by

$$P_T[dBm] = L_{CE} + P_R, \tag{19}$$

where P_R is the receiver minimum sensitivity level, and L_{CE} is the path loss at the cell edge. By assuming QPSK data modulation, tail-biting convolution coding with a 1/2 coding rate, 114 dBm/MHz noise power, and 3 dB implementation loss, the receiver sensitivity should be -96.86 dBm so that the bit error rate (BER) after forward error correction (FEC) is less than 10^{-6} [11]. At the cell edge with the radius of 500 m, L_{CE} becomes 137.8 dB, and consequently, P_T is equal to 40.95 dBm [22].

The receiver samples the baseband signal at 20 MHz, and interference cancellation is applied after the coherent integration. By detecting the preamble symbol at the correlator output, the receiver can get additional correlation gain of 21 dB compared to the data symbol.

The SNR of the preamble correlator output at the cell boundary is 38.14 dB. Thus, the receiver can detect a preamble symbol at a distance 1.8 km away from the corresponding BS when the detection threshold is 17 dB. At the point (0.5, 0) km, where BS01, BS03, and BS04 intersect in Fig. 5, BS17 is 1.93 km away, and its SNR is 7.67 dB.

Assuming that the frequency residual is within 1 Hz, and 10 preamble symbols are coherently accumulated, the coherent integration gain becomes 11.95 dB. In this case, the SNR of BS17 after the coherent integration exceeds the detection threshold at (0.5, 0) km. In this case, if interference cancellation operates successfully, it is expected that BS17 to BS19 signals can be detected everywhere in cell 1. In the interference cancellation process, (15), (17), and (18) are used in the coefficient estimation with $N_p = \mathbf{1}$, so that three samples in the vicinity of the maximum of the correlator outputs are used in the estimation.

The performance of the interference cancellation at coordinate (x=0m, y=200m) was analyzed by observing the autocorrelation profile of BS03. The path loss for BS03 signal in coordinate (x=0m, y=200m) is 145.6 dB and the SNR is 36.8 dB when 10 times integration is performed. However, the SNIR is 7.1 dB due to interference from BS01 and it is not detectable as shown in Fig. 6 (a). The SNIR of BS01 is 21.6 and is strong enough to be detected.

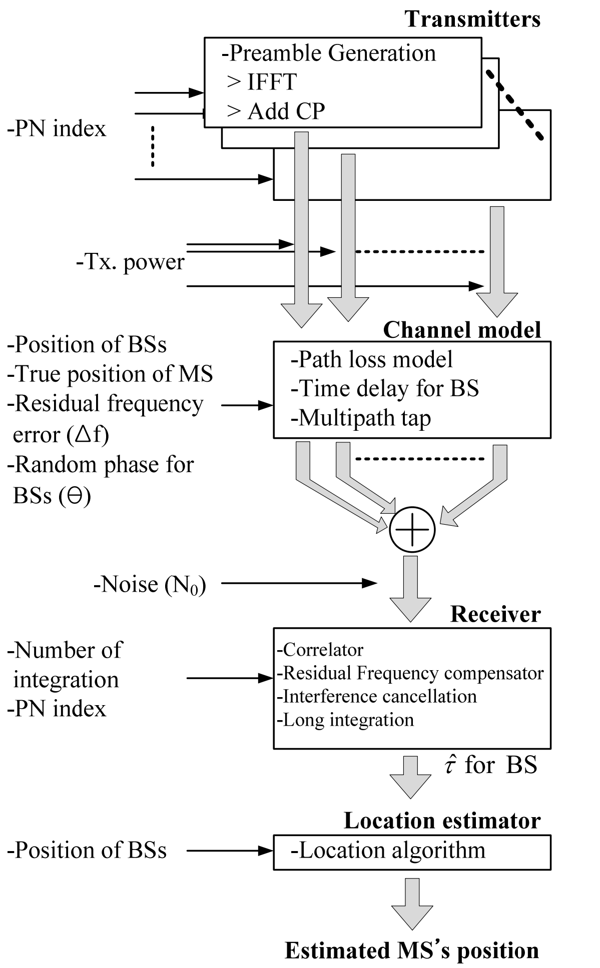


Fig. 4. Structure of mobile WiMAX geo-location simulator.

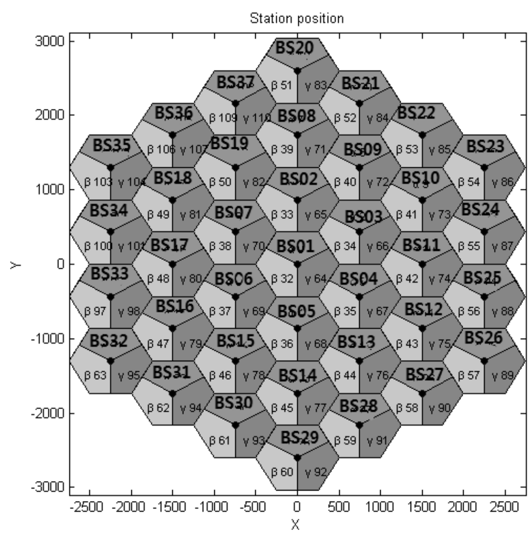
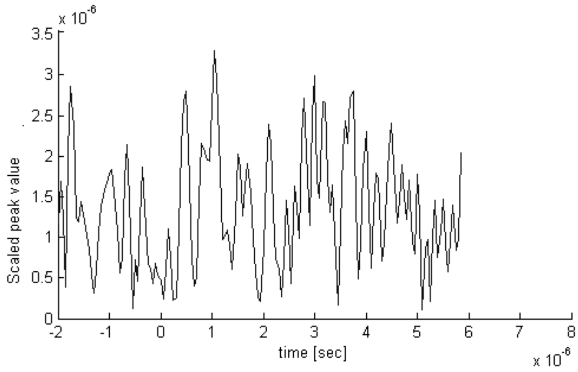
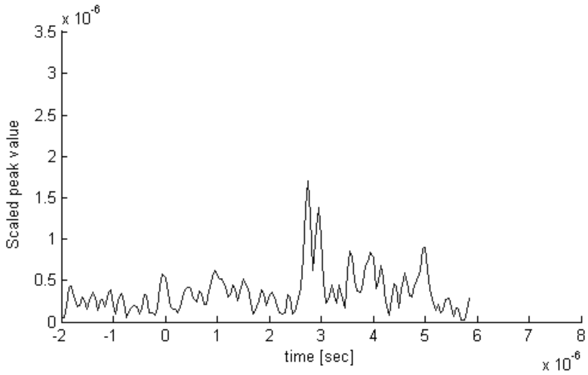


Fig. 5. Allocation of the BSs.



(a) Before interference cancellation.



(b) After interference cancellation.

Fig. 6. The autocorrelation profile of BS03 at coordinate (x=0m, y=200m).

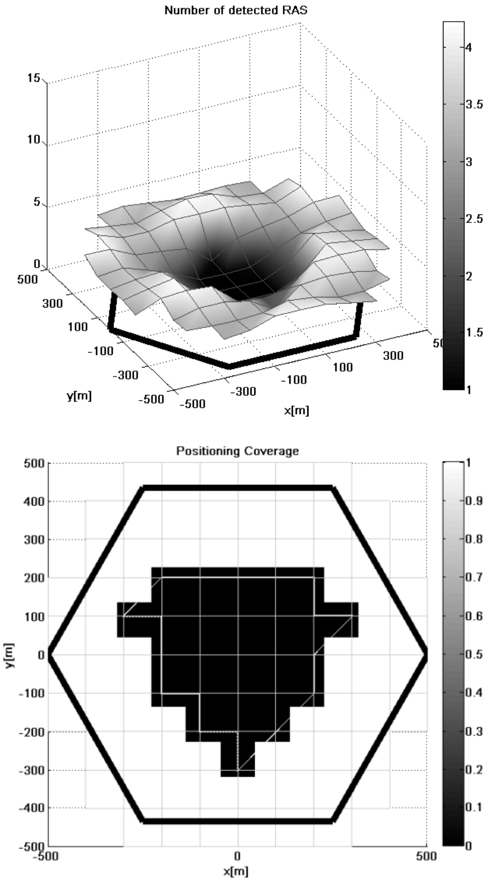


Fig. 7. Number of BSs detected and its positioning coverage when the coherent integration is employed.

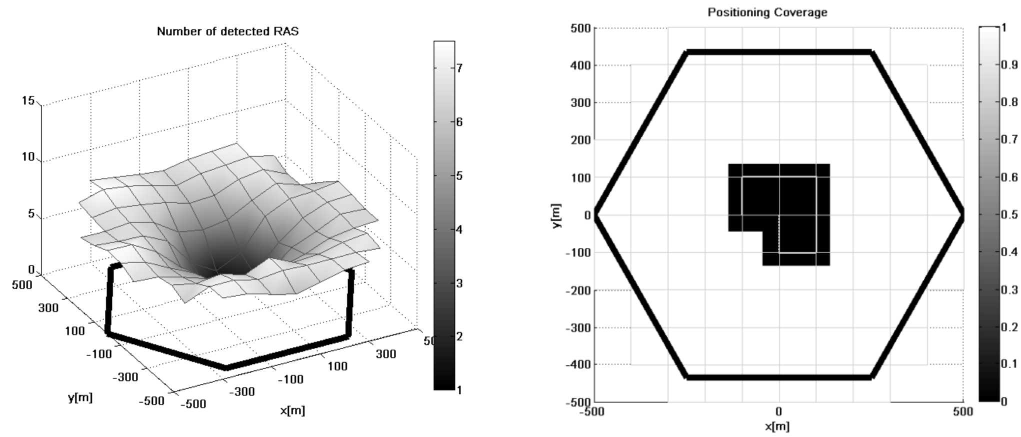
Thus, interference cancellation is performed by estimating the BS01 signal and its multipath component. The Fig. 6 (b) shows that SNIR of BS03 is increased to 17.6 dB after interference cancellation.

As the MS position moves by 100 m towards the x-axis and y-axis directions, respectively in cell BS01, the numbers of detected preamble signals were analyzed. At every point, 100 trials were repeated to obtain the statistical data.

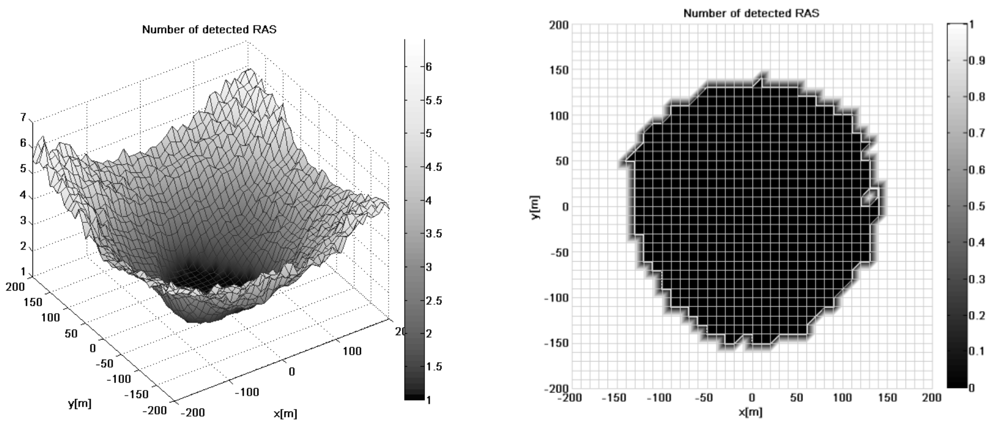
The average number of BSs detected over the whole cell area if neither long integration nor interference cancellation is 3.15. Only 1 to 2 signals were detected in the vicinity of the BS, and about 4 BSs were detected at the boundaries of the service cell. The area where two-dimensional positioning is possible is 65.7% of the cell area. The average number of the detected BSs after the long integration of 10 preamble symbols is shown in Fig. 7 with positioning coverage which indicates the area where geo-location is possible. There is almost no improvement compared to the case of not using coherent integration. The average number of BSs detected in the whole cell area is 3.15, and the area where positioning is available is 65.7%. This shows that hearability cannot be improved by long integration alone.

Fig. 8(a) shows the number of BSs detected when both interference cancellation and long integration are applied. The average number of the BSs detected over the whole cell area is 5.96, and 6 to 8 BSs were detected near the boundaries of the cell, which is close to the theoretical value previously described. To analyze the detection performance within 200 m of BS01 in more detail, an experiment was carried out at every 10 m as shown in Fig. 8(b). In the area where the MS is located 150 m to 200 m away from BS01, 3 to 7 BSs were detected, and the average number of BS signals detected was 3.32. However, within 150 m of BS01, only 1 to 2 signals were detected. From the results, if we employ long integration and the proposed interference cancellation method, two-dimensional positioning by multilateration is possible everywhere in the mobile WiMAX cellular network except where the distance between the BS and the MS is less than 150 m. Even when the MS is located within 150 m of the BS, using the signal power and the CID method, we can acquire the MS location with error of about 80 m.

The TDOA measurement can be obtained from the detected preamble signals, and two-dimensional positioning is possible using hyperbolic positioning if at least two TDOA measurements are available. Fig. 9 shows the root mean-square error (RMSE) of the position estimate as the



(a) Whole area of cell 1 and its positioning coverage



(b) The area within 100 m of the origin of cell 1 and its positioning coverage

Fig. 8. Number of BSs detected when both coherent long integration and interference cancellation are employed.

MS moves in cell 1. Both interference cancellation and long integration are used in obtaining TDOA measurements, and the Gauss-Newton method is employed as the location algorithm [24]. In the middle of the cell, MS location cannot be estimated by multilateration due to the scarcity of measurements. Excluding the center of the cell, the RMSE of the location measurements is 14.31m on average.

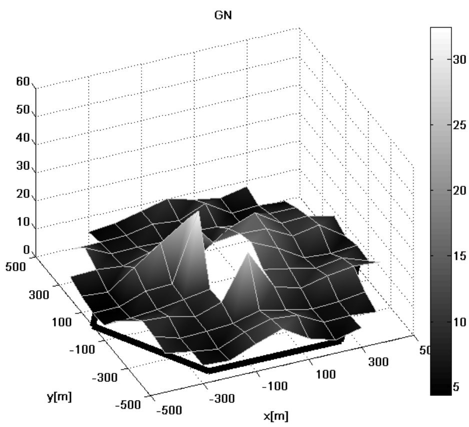


Fig. 9. Location error in cell 1 of the proposed geo-location.

The location accuracy described above is only valid when there is no non-line of sight (NLOS) situation. In the conventional cellular network, it is known that 500 to 700 m NLOS error may be produced in a macro-cell with a radius of 2 km [4]. It is expected that in the mobile WiMAX network, in which the cell size is less than 500 m, the NLOS error is less than 100 m. Therefore, in practice, the geo-location error using mobile WiMAX may be 150m to 200 m assuming that dilution of precision (DOP) is less than 2.

4. Conclusion

This paper presented a hearability enhancement method that can be applied to geo-location system using the OFDM preamble symbol. Since the OFDM signals are not real-valued signal in the time domain, iterative interference cancellation for complex-valued signal was proposed using simplified ML. The multipath environment was also considered in the proposed method so that CCI and multipath effect decrease together. Coherent long integration and the proposed interference cancellation were used to maximize the hearability of the cellular network. If enough preambles are detected, TDOA multilateration will be performed. For areas where multilateration is impossible, signal strength and CID information can be used for substitute.

To see the effect of the hearability enhancement, geo-location simulation was performed for mobile WiMAX system. As shown in the simulation, MS position was obtained with the accuracy of less than 80 m assuming no NLOS. If the proposed geo-location scheme using preamble symbols is applied in mobile WiMAX, location-based service (LBS) capability can be easily employed in the MS without additional hardware. This hearability enhancement method is also applicable to any cellular network system that uses the OFDM preamble signal.

Acknowledgements

This work has been supported by National GNSS Research Center program of Defense Acquisition Program Administration and Agency for Defense Development.

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