

CRLB Analysis for Cognitive IR-UWB with Imperfect Sensing

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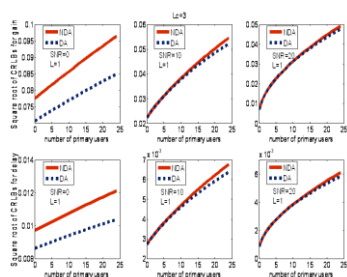
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Graphical abstract



Abstract

Cognitive ultra wideband (UWB) technology has ability to sense the environment, and offers an adaptive system with low transmission power and high throughput for wireless communications. Impulse radio UWB (IR-UWB) supports flexible modulations, with short mono-pulses, in which channel estimation has a critical role. However, imperfect sensing in cognitive IR-UWB introduces the channel estimation errors associated to the interfering primary users. Two main schemes of channel parameters identification are known as pilot-aided and blind methods. In this paper, the effect of primary user interference is considered for channel estimation based on maximum-likelihood (ML) criterion. Cramer-Rao lower band (CRLB) analysis is derived, and performance analysis is simulated to compare both pilot-aided and blind methods in cognitive IR-UWB system. The results verify the better performance of the pilot-aided estimation, and also show the negative effect of increasing the number of primary users and paths on the performance. At SNR=10 dB, for pilot-aided method with 10 users and 3 paths, the CRLB square root of gain is 0.5×10^{-3} and that of delay is 0.1×10^{-3} less than blind method. Besides, when the number of primary users is 20, and the paths are 7, this gap is 4×10^{-3} for gain, and 0.5×10^{-3} for delay.

Keywords: Cognitive radio; ultra wideband communications; Cramer-Rao lower band; channel estimation

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1.0 INTRODUCTION

Ultra wideband (UWB) communication is a promising technology for short or medium-range wireless communication networks with low cost and low power implementation in an extremely large transmission bandwidth. UWB communication is operated as secondary user and coexist primary users which have exclusive right to use the spectrum. For coexistence, cognitive features enable the UWB user to sense the occupied spectrum by primary users, and adapt the opportunistic spectrum and data rate, accordingly [1], [2].

UWB consists of a variety of throughput options including very high data rates, with applications for wireless local area networks (WLAN) and wireless personal area networks (WPAN). Impulse radio based UWB (IR-UWB), known as “carrier-less short pulse”, and OFDM based UWB (OFDM-UWB) are used for UWB implementation. In cognitive IR-UWB, adaptability is performed by varying the pulse duration or waveform shape. When sensing mechanism is not perfect, cognitive IR-UWB needs to mitigate the impairments from the primary users’ interference.

In this paper, the performance of cognitive IR-UWB is investigated in the presence of different number of primary users and channel conditions. We employ Rake receiver with maximum ratio combining (MRC) to exploit multipath diversity. Channel tracking is pursued in both pilot-aided and

blind estimation schemes. Pilot-aided or data-aided (DA) method periodically retransmits the training sequence, or in a decision-directed (DD) manner. On the other hand, blind estimation or non-data-aided (NDA) is helpful when new users enter the network and training sequences impede the transmitter [3].

Herein, the Cramer-Rao lower bound (CRLB) [4], [5], [6] is served as a benchmark for the ML channel estimator in both DA and NDA cases. Using errors predicted by the CRLB, unknown symbols are transmitted over a multipath channel to perform NDA estimation. Then, CRLB of the variance of the delay and gain estimates is derived. In the presence of primary users, we analyze the standard deviation of path gain and delay and compare the system performance of pilot-aided and blind estimation methods.

The rest of the paper is organized as follows. Section 2, describes the system description. Blind channel estimation is discussed in Section 3. Section 4 addresses CRLBs for ML Channel estimation. Simulation results and discussion are presented in Section 5, followed by conclusions in Section 6.

2.0 SYSTEM DESCRIPTION

One of the most significant channel models for UWB systems is the model proposed by Saleh-Valenzuela using a Rayleigh

probability density function for fading channel coefficient [7]. Reflection, diffraction, and scattering are the main reasons for multipath occurrence in UWB channels [8]. Figure 1 represents an L_c -ray multipath channel with independent L_c delays, τ_{c,l_c} , and gains γ_{c,l_c} of each path.

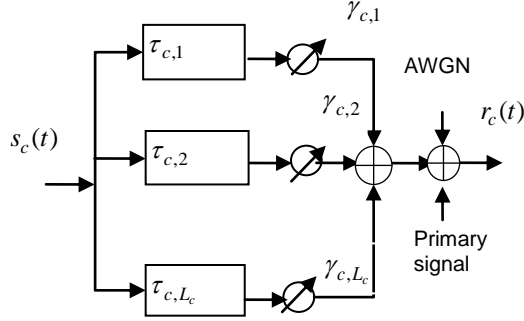


Figure 1 Multipath channel model

A rake receiver is employed to resolve and combine these received copies of the original signal, and take advantage of multipath diversity principle. Figure 2 illustrates a rake receiver comprised of L_c fingers to detect the strongest multipath components. In practice, priori knowledge of the delays and the associated gains is not available, and needs to be estimated as $\hat{\tau}_{c,l_c}$, and $\hat{\gamma}_{c,l_c}$.

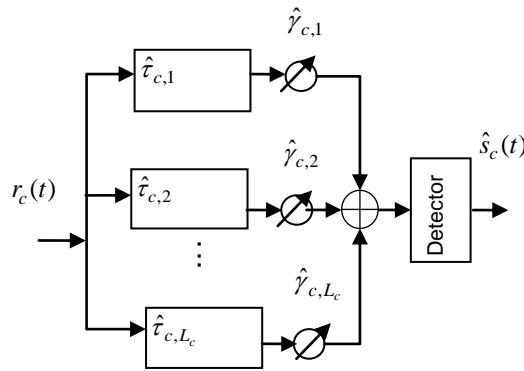


Figure 2 Block diagram of the L_c branch Rake receiver

All-RAKE (A-RAKE) receiver is an optimal design with enough number of fingers, and selective-RAKE (S-RAKE) receiver is suboptimal with the limited number of fingers which are investigated in [9]. A zero forcing (ZF) Rake receiver can be combined with ordered successive interference cancellation (OSIC) as proposed in [10].

IR-UWB serves the same spread spectrum concepts as code division multiple-access (CDMA), with DS-SS transceiver implementation. The information is modulated by pulse position modulation (PPM), in which N positions indicate the N symbols. In order to smooth the energy spikes in the spectrum, caused by periodicity of the pulse repetition, a pseudo-random sequence of delays technique, known as time hopping (TH), is applied.

TH-PPM format of the transmitted cognitive IR-UWB signal in our system is expressed as

$$s_c(t) = \sum_{i=0}^{M-1} b(t - iNT_f - a_i\Delta) \tag{1}$$

where $s_c(t)$ is the cognitive user’s signal, and

$$b(t) = \sum_{j=0}^{N-1} g(t - jT_f - c_jT_c) \tag{2}$$

where $g(t)$ is the monocycle pulse with duration T_p , i is the information bit index, N is the repetition length, T_f is the frame interval, $a_i \in \{0,1\}$ is the i -th information bit with equal a priori probabilities, Δ is the additional time shift introduced when $a_i = 1$, c_j is the time-hopping code and T_c is the chip time. The received signal is observed over an interval of $0 \leq t \leq T_0$, in which T_0 is a multiple M of the symbol period NT_f . We also have $E_{s_c} = ME_b$, and

$$E_b = \int_0^{NT_f} b^2(t) dt.$$

The received signal through L_c paths at the output of the cognitive receiver antenna can be written as

$$r_c(t) = \sum_{l_c=1}^{L_c} \gamma_{c,l_c} s_c(t - \tau_{c,l_c}) + \omega_t(t) \tag{3}$$

where γ_{c,l_c} and τ_{c,l_c} are attenuation, and the delay affecting its replica traveling through the l_c -th path respectively, and ω_t is thermal noise plus the interference caused by the primary users with total power spectral density $\sigma_{\omega_t}^2$, i.e. $\omega_t(t) = I_p(t) + n(t)$, and

$$I_p(t) = \left[\sum_{l_p=1}^{L_p} \sum_{k=1}^{N_p} \gamma_{k,l_p} s_{p,k}(t - \tau_{k,l_p}) \right] \tag{4}$$

where N_p is the number of primary users, L_p denotes the number of primary user paths, and

$$\sigma_{\omega_t}^2 = \sigma_{I_p}^2 + \sigma_n^2 \tag{5}$$

Assume $s_{p,k}(t - \tau_{k,l_p})$ is deterministic, so

$$\sigma_{I_p}^2 = \sum_{l_p=1}^{L_p} \sum_{k=1}^{N_p} [E(\gamma_{k,l_p}^2) - E^2(\gamma_{k,l_p})] E_{s_p} \tag{6}$$

Then

$$\sigma_{\omega_t}^2 = E_{s_p} \sum_{l_p=1}^{L_p} \sum_{k=1}^{N_p} [E(\gamma_{k,l_p}^2) - E^2(\gamma_{k,l_p})] + \sigma_n^2 \tag{7}$$

3.0 BLIND CHANNEL ESTIMATION

In this section, we consider cognitive users, and analyze blind (NDA) channel estimation when the symbols are unknown. As indicated in [7], we can get rid of these unknown information by first computing the Likelihood function, $\Lambda(\hat{a}, \hat{\gamma}_c, \hat{\tau}_c)$, for $a = (a_0, a_1, \dots, a_{M-1})$, $\gamma_c = [\gamma_{c,1}, \gamma_{c,2}, \dots, \gamma_{c,L_c}]$ and $\tau_c = [\tau_{c,1}, \tau_{c,2}, \dots, \tau_{c,L_c}]$, and then averaging over the probability density of \hat{a} . This produces the marginal Likelihood function for (γ_c, τ_c) as

$$\Lambda(\hat{\gamma}_c, \hat{\tau}_c) = \int \Lambda(\hat{a}, \hat{\gamma}_c, \hat{\tau}_c) p(\hat{a}) d(\hat{a}) \tag{8}$$

from which the channel estimates are derived. Since there is no specific knowledge of the data symbols, except for independent zero or one values with the same probability, we model $p(\hat{a})$ as

$$p(\hat{a}) = \prod_{k=0}^{M-1} \frac{[\delta(\hat{a}_k) + \delta(\hat{a}_k - 1)]}{2} \tag{9}$$

where $\delta(\hat{a})$ is the Dirac function. Therefore, (8) and (9) yield

$$\begin{aligned} \log[\Lambda(\hat{\gamma}_c, \hat{\tau}_c)] &= \sum_{k=0}^{M-1} \log\left[\frac{1}{2} \exp\left\{-\frac{1}{\sigma^2} \sum_{l_c=1}^{L_c} \hat{\gamma}_{l_c} z_k(\hat{\tau}_{c,l_c}, 0)\right\}\right. \\ &\quad \left. + \frac{1}{2} \exp\left\{-\frac{1}{\sigma^2} \sum_{l_c=1}^{L_c} \hat{\gamma}_{c,l_c} z_k(\hat{\tau}_{c,l_c}, 1)\right\}\right] - \frac{1}{2\sigma^2} ME_b \sum_{l_c=1}^{L_c} \hat{\gamma}_{c,l_c}^2 \end{aligned} \tag{10}$$

where

$$z_k(\hat{\tau}_{c,l_c}, \hat{a}_k) = [r(t) \otimes b(-t)]_{t=kNT_f + \Delta\hat{a}_k + \hat{\tau}_{c,l_c}} \tag{11}$$

Thus, $\log[\Lambda(\hat{\gamma}_c, \hat{\tau}_c)]$ is maximized as a function of $(\hat{\gamma}_c, \hat{\tau}_c)$.

The first step provides

$$\hat{\gamma}_{c,l_c} = \frac{1}{ME_b} J(\hat{\tau}_{c,l_c}), 1 \leq l_c \leq L_c \tag{12}$$

when

$$J(\hat{\tau}_c) = \sum_{k=0}^{M-1} \frac{z_k(\hat{\tau}_c, 0) + z_k(\hat{\tau}_c, 1)}{2} \tag{13}$$

then, the problem reduces to looking for the location of the extrema of $J(\hat{\tau}_c)$ and specify the value of $\hat{\tau}_c$.

4.0 CRLB FOR ML CHANNEL ESTIMATION

In this section, we calculate the CRLB of the path delay and amplitude estimation errors as a function of the parameters of the transmitted signals. We define channel parameter vector as

$$\theta = [\gamma_{c,1}, \gamma_{c,2}, \dots, \gamma_{c,L_c}, \tau_{c,1}, \tau_{c,2}, \dots, \tau_{c,L_c}]^T \tag{14}$$

Thus, from [10], we have

$$-E\left(\frac{\partial^2 \log[\Lambda(\theta)]}{\partial \gamma_{c,i} \partial \gamma_{c,j}}\right) = \begin{cases} -\frac{1}{\gamma_{c,i} \gamma_{c,j}} \sum_{k=0}^{M-1} F_{ijk}, i \neq j \\ 2M\Gamma - \frac{1}{\gamma_{c,i}^3} \sum_{k=0}^{M-1} G_{ik}, i = j \end{cases} \tag{15a}$$

$$-E\left(\frac{\partial^2 \log[\Lambda(\theta)]}{\partial \tau_{c,i} \partial \tau_{c,j}}\right) = \begin{cases} 0, i \neq j \\ -\sum_{k=0}^{M-1} [\Gamma \gamma_{c,i}^2, I_{ik} + H_{ik}], i = j \end{cases} \tag{15b}$$

$$-E\left(\frac{\partial^2 \log[\Lambda(\theta)]}{\partial \gamma_{c,i} \partial \tau_{c,j}}\right) = -E\left(\frac{\partial^2 \log[\Lambda(\theta)]}{\partial \tau_{c,i} \partial \gamma_{c,j}}\right) = 0 \tag{15c}$$

where $\Gamma = E_b / (2\sigma_{\omega_i}^2)$ is the signal-to-interference-pulse-noise ratio (SINR) and

$$\begin{aligned} F_{ijk} &= \frac{1}{\sqrt{2\pi R\Gamma}} \int_{-\infty}^{+\infty} \text{sech}^2(x) e^{-\frac{(x\bar{a}_k - R\Gamma)^2}{2R\Gamma}} \times \left\{ [\Gamma \gamma_{c,i}^2 + \frac{\gamma_{c,i}^2 (x\bar{a}_k - R\Gamma + \Gamma \gamma_{c,j}^2)}{R - \gamma_{c,j}^2}] \right. \\ &\quad \left. \times [\Gamma \gamma_{c,j}^2 + \frac{\gamma_{c,j}^2 (x\bar{a}_k - R\Gamma)}{R}] \right\} \\ &\quad - \frac{\gamma_{c,i}^2 [\Gamma \gamma_{c,j}^2 + \frac{\gamma_{c,j}^2 (x\bar{a}_k - R\Gamma)}{R}]^2}{R - \gamma_{c,j}^2} - \frac{\Gamma \gamma_{c,i}^2 \gamma_{c,j}^2}{R} \Big\} dx \end{aligned} \tag{16}$$

$$\begin{aligned} G_{ik} &= \frac{1}{\sqrt{2\pi R\Gamma}} \int_{-\infty}^{+\infty} \text{sech}^2(x) e^{-\frac{(x\bar{a}_k - R\Gamma)^2}{2R\Gamma}} \cdot \left\{ [\Gamma \gamma_{c,i}^2 \right. \\ &\quad \left. + \frac{\gamma_{c,i}^2 (x\bar{a}_k - R\Gamma)}{R}]^2 + \frac{\Gamma (R - \gamma_{c,i}^2) \gamma_{c,i}^2}{R} \right\} dx \end{aligned} \tag{17}$$

$$H_{ik} = \frac{\bar{a}_k}{\sqrt{2\pi R\Gamma}} \int_{-\infty}^{+\infty} \tanh h(x) e^{-\frac{(x\bar{a}_k - R\Gamma)^2}{2R\Gamma}} \cdot [\Gamma \epsilon \gamma_{c,i}^2 + \frac{\epsilon \gamma_{c,i}^2}{R} (x\bar{a}_k - R\Gamma)] dx \tag{18}$$

$$I_{ik} = \frac{\eta}{\sqrt{2\pi R\Gamma}} \int_{-\infty}^{+\infty} \sec h^2(x) e^{-\frac{(x\bar{a}_k - R\Gamma)^2}{2R\Gamma}} dx + \epsilon \tag{19}$$

$$\text{where } \varepsilon = \frac{\int_0^{NT_f} b(t)b''(t)dt}{E_b}, \eta = \frac{\int_0^{NT_f} b'^2(t)dt}{E_b}, \quad \text{and}$$

$$\bar{a}_k = \text{sgn}(1 - 2a_k), \quad R = \sum_{l_c=1}^{L_c} \gamma_{c,l_c}^2. \quad \text{Based on [5], the}$$

Fisher information matrix is

$$I(\theta) = \begin{bmatrix} B, C \\ C, D \end{bmatrix} \quad (20)$$

where

$$B = \begin{bmatrix} 2M\Gamma - \frac{1}{\gamma_{c,1}^3} \sum_{k=0}^{M-1} G_{1k}, \dots, -\frac{1}{\gamma_{c,i}\gamma_{c,j}} \sum_{k=0}^{M-1} F_{ijk}, \dots, -\frac{1}{\gamma_{c,1}\gamma_{c,L_c}} \sum_{k=0}^{M-1} F_{1L_c,k} \\ \dots \\ \dots \\ 2M\Gamma - \frac{1}{\gamma_{c,i}^3} \sum_{k=0}^{M-1} G_{ik} \dots \\ \dots \\ -\frac{1}{\gamma_{c,L_c}\gamma_{c,1}} \sum_{k=0}^{M-1} F_{L_c1k}, \dots, 2M\Gamma - \frac{1}{\gamma_{c,L_c}^3} \sum_{k=0}^{M-1} G_{L_c,k} \end{bmatrix} \quad (21)$$

$$C = \text{diag}\{0, \dots, 0\}, \quad (22)$$

$$D = \text{diag}\left\{-\sum_{k=0}^{M-1} [\Gamma \gamma_{c,1}^2 I_{1k} + H_{1k}], \dots, -\sum_{k=0}^{M-1} [\Gamma \gamma_{c,i}^2 I_{ik} + H_{ik}], \dots, -\sum_{k=0}^{M-1} [\Gamma \gamma_{c,L_c}^2 I_{L_c,k} + H_{L_c,k}]\right\} \quad (23)$$

The CRLB of the delay estimation and amplitude estimation for l -th path are defined as:

$$\text{var}(\hat{\gamma}_{c,l_c}) \geq [I^{-1}(\theta)]_{l_c,l_c} = [(B - CD^{-1}C)^{-1}]_{l_c,l_c} \quad (24)$$

$$\text{var}(\hat{\tau}_{l_c}) \geq [I^{-1}(\theta)]_{L_c+l_c,L_c+l_c} = [(D - CB^{-1}C)^{-1}]_{l_c,l_c} \quad (25)$$

Then, we can obtain CRLBs of estimates for $l_c = 1, \dots, L_c$.

The CRLB for estimation of γ_{c,l_c} is $[I^{-1}(\theta)]_{l_c,l_c}$, which is the l_c -th diagonal element of the inverse matrix I^{-1} , and the

CRLB for estimation of τ_{c,l_c} is

$$-1 / \sum_{k=0}^{M-1} [\Gamma \gamma_{c,l_c}^2 I_{l_c,k} + H_{l_c,k}]. \quad \text{It is obvious that the CRLBs}$$

for estimations depend on $a_k (k = 0, 1, \dots, M - 1)$,

$\gamma_{l_c}, (l_c = 1, 2, \dots, L_c)$, SINR, and the sample size.

5.0 SIMULATION RESULTS AND DISCUSSION

The following pulse wave form is adopted in these simulations:

$$g(t) = [1 - 16\pi(\frac{t - \frac{T_p}{2}}{T_p})^2] \exp[-8\pi(\frac{t - \frac{T_p}{2}}{T_p})^2] \quad (26)$$

Parameters of (1), and (2) are substituted as $\Delta = T_p$, $T_f = 40T_p$, and $T_c = T_f / 20$. The N_h is 20 and random symbols are in the interval of $[0, 1, \dots, 19]$. N_p asynchronous primary users are considered, the number of paths $L_c = 3$, $R = 1.1$, and the path delay of all users are $5lT_p$. The path gains are, $\gamma_1 = 0.73$, $\gamma_2 = 0.67$, and $\gamma_3 = 0.35$, while the other gains vary from user to user with the Rayleigh distribution and their expectations are proportional to $\exp(-l/4)$. Observation length of 100 symbols ($M=100$) and SNR of 0 dB to 60 dB ($SNR = E_b / 2\sigma_n^2$) are assumed.

Figure 3 shows the CRLBs of blind and pilot-aided methods for (γ_1, τ_1) as the number of primary users varies at different values of SNR, with $L_c = 3$. In this figure, L is the effective number of the paths that has been used for the processing. It has been shown that blind estimator is poorer than pilot-aided method for both gain and delay.

In Figure 4 the value of SNR is equal to 10 dB and it compares the derived CRLBs for blind and pilot-aided of the standard deviations as the number of primary users varies for L -th path of total paths, $L_c = 3$. Figure 5 shows obvious gaps between blind and pilot-aided curves in the derived CRLBs at SNR=10 dB, which increases with large number of primary users. Figure 6 and Figure 7 are the CRLBs of blind and pilot-aided methods for γ_1 and τ_1 as the number of primary users varies at different numbers of $L_c = [3, 5, 7]$ which the path gains are $\gamma_1 = 0.73$, $\gamma_2 = 0.67$, $\gamma_3 = 0.35$, $\gamma_4 = 0.23$, $\gamma_5 = 0.12$, $\gamma_6 = 0.1$, and $\gamma_7 = 0.05$. These figures show that the CRLBs increases with increasing of the number of total paths. In the previous work, [4], the standard deviation curves of gain and delay are derived only for data-aided method. Besides, the authors did not compare the pilot-aided and blind methods while their system is a non-cognitive UWB.

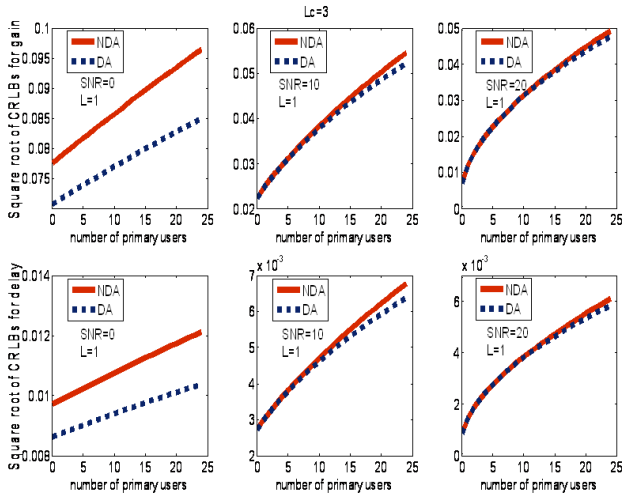


Figure 3 The CRLBs of NDA and DA methods for (γ_1, τ_1) at different values of SNR, with $L_c = 3$

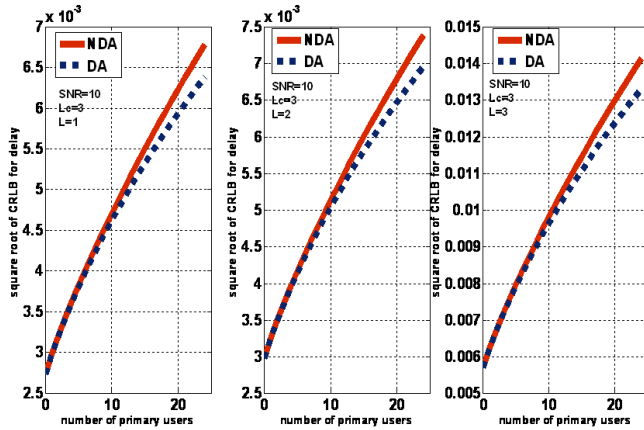


Figure 4 The square root of CRLBs for ML estimation of τ_1, τ_2 , and τ_3 fo L -th path, when $L_c = 3$ and SNR=10 dB

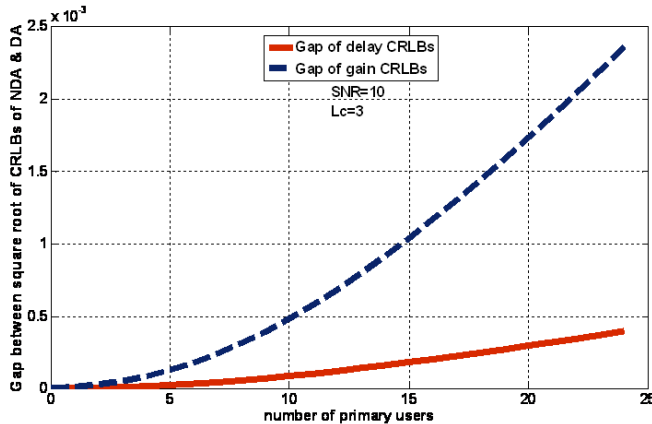


Figure 5 The gap between CRLBs of NDA and DA methods for (γ_1, τ_1) at SNR=10 dB

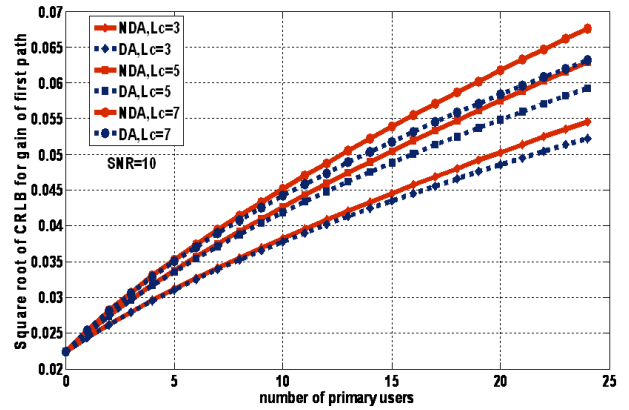


Figure 6 The CRLBs of NDA and DA methods for gain of γ_1 versus number of primary users

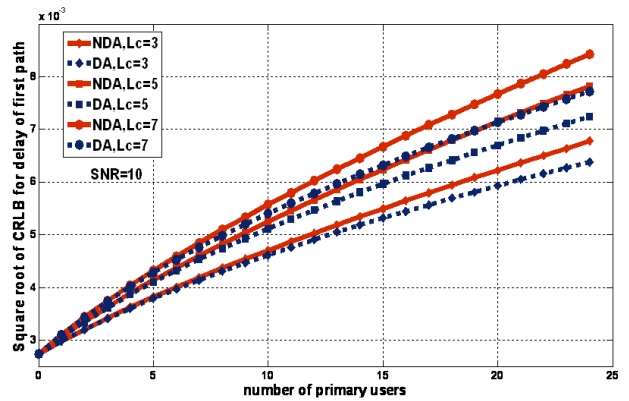


Figure 7 The CRLBs of NDA and DA methods for delay of τ_1 versus number of primary users.

6.0 CONCLUSION

In this article, CRLBs of pilot-aided and blind ML channel identification are provided for cognitive IR-UWB communication. Numerical investigation has verified that non-data-aided method is poorer than data-aided method, and there exist obvious gaps between NDA and DA curves in the derived CRLBs. The curves illustrate that the CRLBs increase with the total number of primary users and paths. As a future work, the imperfect sensing errors can be investigated for CM1-CM4 standard of UWB channel models.

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