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UPLINK VERSUS DOWNLINK WIRELESS MOBILE POSITIONING IN UMTS CELLULAR RADIO SYSTEMS

Wrya. Muhammad¹, Emmanuèle Grosicki², Karim Abed-Meraim², Jean Pierre Delmas¹,
François Desbouvries¹

¹ Institut National des Télécommunications, Dept. CITI, Evry, France

² Ecole Nationale Supérieure des Télécommunications de Paris, Dept. TSI, France

e-mail: wrya.muhammad@int-evry.fr, grosicki@tsi.enst.fr

ABSTRACT

A performance comparison between radio location positioning (LP) in Wide band Code Division Multiple Access (WCDMA), more precisely in UMTS FDD cellular phones, using the Uplink (UL) and Downlink (DL) scenarios, is carried out. The radio location algorithm using the Time Difference Of Arrival (TDOA) is applied, where the TDOA is the difference between two TOA estimated at two base stations. We propose a practical algorithm to obtain high precision PL under high level interference. It consists in a matched filter followed by high resolution Multiple Signal Classification (MUSIC) algorithm to estimate with a high resolution the TOA. indeed, in practice *near-far* resistant detection algorithms are not easy to implement directly on the received data due to the large spreading factors. The power-up and Idle Slot (IS) techniques together with Interference cancellation algorithms are used to combat the *near-far* problem. Simulations were carried out in the same observation conditions for UL and DL scenarios. The results show an enhancement in the LP using MUSIC algorithm after matched filtering besides, a better performance in the DL scenario.

1 INTRODUCTION

Radio location is a growing practice in cellular mobile phone systems because of its vast useful applications like: localizing the traffic by the operators, emergency interventions, billing tariffs depending on the mobile position and discouraging criminal activities using the cellular systems. Techniques of radio location were developed for the GSM cellular phone and are now standardized. However, these standards are not applicable to the wide band CDMA cellular systems since it uses a different air interface technology. Studies on the radio LP of cellular phones in the IS-95 CDMA systems have been performed [1] [2]. But this system is also different from the Universal Mobile Telecommunications System (UMTS) wide band CDMA standards. Few studies were carried out on the radio LP in UMTS systems [3] [4].

One of the limiting factors of the TDOA radio location algorithms is the Non Line of Sight (NLOS) propagation paths, since it introduces a bias error on the TOA. In [5] an essay is introduced to discriminate the LOS from the NLOS paths. In this work we consider the first multi path component as the LOS path. The TDOA is deduced from the difference between two TOA estimated at 3 or more base stations (BS).

Each two TDOAs are then used to obtain unambiguous hyperbolic curves to position the cellular systems. A matched filter is used to estimate the TOA at the BS or the mobile station (MS) for UL or a DL radio LP scenarios, respectively. However, the matched filter is unable to resolve multi path components that arrive within $\pm T_c/Q$, where T_c is the chip duration and Q is an oversampling factor besides, the multiple users interference will be highly present at the matched filter output. Hence, in the UL scenario, interference cancellation and high resolution estimation of TOA is needed to obtain accurate PL, while in the DL the intracell interference will not highly affect the matched filter output. Nevertheless, the output of the matched filter, after interference cancellation, can be used to obtain high resolution TOA estimations: the idea of using super-resolution method in the context of delay profile estimation of a channel has been proposed in [6]. In this work we study the application of a super-resolution method using MUSIC algorithm together with interference cancellation on the PL.

The Signal to interference and Noise ratio (SINR), as in all other signal processing problems, is another limiting factor because of its direct relation with the variance of the TOA estimation. Besides, since the CDMA systems is interference limited, a power control algorithm must be used to minimize the signal strength at the neighboring stations or mobiles (i.e. to combat the *near-far* problem). As a result, it is difficult to “hear”, with a good SINR, a mobile phone at multiple BS, or to “hear” multiple BS by the mobile phone; in DL and UL location positioning scenarios, respectively. In practice, the SINR is increased using two techniques: (1) the power-up in the UL LP scenario (2) the Idle Slot (IS) in the DL PL scenario [3]. However, these techniques have the disadvantages of reducing network capacity. In fact, theoretically *near-far* resistant estimation methods could be applied directly to the received data to avoid the *near-far* problem. Unfortunately, in practice they are very expensive to implement due to the large spreading factors.

Finally, many factors can be considered in the choice of the LP scenario: cost, accuracy, communication interruptions during localization etc. In this paper we are interested in the accuracy performance of the two mentioned LP scenarios.

2 SIGNAL MODELS

In this paper we tackle Frequency Division Duplex (FDD) UMTS system. The signal models for this system are con-

structured according to the standards proposed by the ETSI [7].

2.1 Uplink Signals

The UL baseband signal referring to the scheme proposed by [7] (see figure (1)) can be written as:

$$x_k(t) = \left\{ \sum_{l=-\infty}^{+\infty} [d_k^{(l)} b_k^I(t - lT) + j c_k^{(l)} b_k^Q(t - lT)] \right\} s_k(t) \quad (1)$$

where T is the data and control bit duration, $s_k(t)$ is the complex scrambling code of the k^{th} user selected from the Large Kasami family. $b^I(t)$ and $b^Q(t)$ are the channelization codes of the dedicated data $d_k^{(l)} \in \{1, -1\}$ and control $c_k^{(l)} \in \{1, -1\}$ channels. These codes are orthogonal and selected from the Walsh family. $s_k(t)$, $b_k^I(t)$ and $b_k^Q(t)$ can be written as:

$$s_k(t) = \sum_{l=0}^{L-1} s_k(l) \Pi(t - lT_c); \quad b_k^I(t) = \sum_{l=0}^{L_c-1} b_k^I(l) \Pi(t - lT_c); \quad b_k^Q(t) = \sum_{l=0}^{L_c-1} b_k^Q(l) \Pi(t - lT_c)$$

where L is the scrambling code length and L_c is the channelization code length or the spreading factor (SF), $s_k(l) \in \{a + jb : a \in \{1, -1\}, b \in \{1, -1\}\}$, $b_k^I(l), b_k^Q(l) \in \{1, -1\}$ and Π is the chip waveform here considered as a rectangular function of width T_c equal to the chip duration.

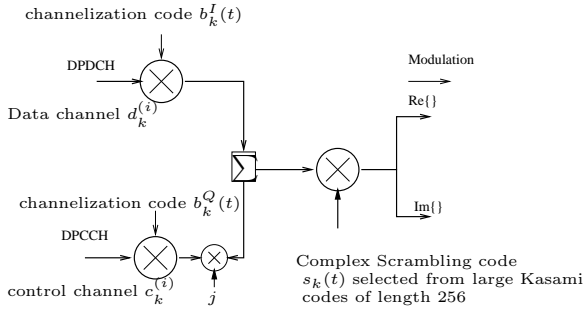


Figure 1: Uplink spreading and modulation of the k^{th} user

The LP can be carried out in the UL scenario either with known or unknown signals in the control channel. Unknown signals can be used provided that the scrambling code length extend over one bit. This is possible in UMTS systems where short scrambling code, such as the large Kasami or Gold code families of length $L = 256$, could be allocated in UL and DL [8]. However, an ambiguity problem will result using these short scrambling codes, if the search radius is greater than $256 * T_c * c \simeq 18$ km, where c is the electromagnetic wave speed. Hence, this is not a drastic problem because normally suburban cell radius have smaller dimensions.

The UL baseband signal of equation (1), with synchronous rectangular functions of the scrambling and channelization, can be rewritten for the l^{th} bit and equal data and control rate, as:

$$x_k^{(l)}(t) = \sum_{l=0}^{L_c-1} [d_k^{(l)} b_k^I(t) + j c_k^{(l)} b_k^Q(t)] s_k(t - lT_c) \Pi(t - (l + lL_c)T_c)$$

2.2 Downlink Signals

Figure (2) shows the DL modulation and spreading for K users [7]. It can be noted that the base station uses a com-

mon pilot channel with a known sequence. This pilot channel is used for channel estimation by the mobile phone. It is normally transmitted continuously with a relatively high power and no power control is applied. Estimating the TOA of these signals is a good choice for LP. The channelization codes are Variable Orthogonal SF codes (OVVSF) Walsh codes.

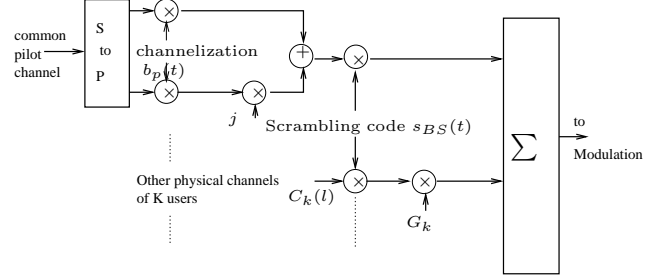


Figure 2: Downlink spreading and modulation for K user

The baseband signal at the l^{th} bit of the pilot channel at any BS is then:

$$x_{BS}^l(t) = \sum_{l=0}^{L_c-1} \{ [p_e^{(l)} + j p_o^{(l)}] b_p(l) + \sum_{k=1}^K G_k C_k(l) \} s_{BS}(l + lL_c) \Pi(t - (l + lL_c)T_c)$$

where the $b_p(l)$ is the Walsh code used for the common pilot channelization, s_{BS} is the scrambling code a base station and $p_e^{(l)}, p_o^{(l)}$ are the l^{th} even and odd bits of the common pilot channel, respectively. $C_k(l)$ represent the sum of all physical channels of a user k and G_k is the gain factor of the user k . In the following we omit the superscript l to simplify. The complex baseband signal is then QPSK modulated and transmitted.

The noise free bandpass signal $y(t)$ received through a static multi path channel with coefficients h_i , where $i = 1, \dots, M$ is the channel length, is:

$$y(t) = \sum_{i=1}^M h_i \sqrt{2P} x(t - \tau_i)$$

where h_i is a complex fading coefficient, τ_i is the time delays of the i^{th} path and P is the carrier transmitted signal power. The QPSK signals at the receiver are down converted integrate-and-dumped with an integration time $T_e = T_c/Q$ with $Q \in \mathbb{N}$. We will assume, without loss of generality, that the τ_i are multiples of T_e (i.e. $\tau_i = \hat{\tau}_i T_e$ where $\hat{\tau}_i \in \mathbb{N}$) this assumption is realistic if $Q \gg 1$. Furthermore, we assume that the integrate-and-dump is synchronous with the received chips. These assumptions simplify to a great extent the equation writings at the receiver without producing big performance changes.

The sampled received signals at the BS for the UL case is:

$$y^{UL}(nT_e) = \sum_{k=1}^K \sum_{i=1}^{M_k} h_{k,i}^{UL} x_k(nT_e - \hat{\tau}_{k,i} T_e) + w(nT_e) \quad (2)$$

where $h_{k,i}^{UL} = h_{k,i}^{ul} \sqrt{P_k} e^{j(\omega_c \tau_{k,i} + \phi_k)}$, $h_{k,i}^{ul}$ is the k^{th} user complex coefficient channel gain, ϕ_k is a phase error between the received and local carrier, M_k is the k^{th} channel length, ω_c is the carrier angular frequency and $w(nT_e)$ is a zero mean complex additive white Gaussian noise with variance σ_0/T_e where σ_0 is the noise power at the receiver input. In matrix

form this is written as:

$$\mathbf{y}^{UL} = \mathbf{X}\mathbf{h}^{UL} + \mathbf{w} \quad (3)$$

where $\mathbf{X} \in \mathbb{C}^{QL \times (\sum_k M_k)}$ and $\mathbf{h}^{UL} \in \mathbb{C}^{(\sum_k M_k)}$. The sampled received signal at the mobile phone for the DL case is:

$$y^{DL}(nT_e) = \sum_{i=1}^M h_i^{DL} x_{BS}(nT_e - \hat{\tau}_i T_e) + w(nT_e) \quad (4)$$

where $h_i^{DL} = h_i^{dl} \sqrt{P_{BS}} e^{j(\omega_c \tau_i + \phi)}$. Or in matrix form:

$$\mathbf{y}^{DL} = \mathbf{X}_{BS} \mathbf{h}^{DL} + \mathbf{w} \quad (5)$$

where $\mathbf{X}_{BS} \in \mathbb{C}^{QL \times M}$ and $\mathbf{h}^{DL} \in \mathbb{C}^M$.

3 ESTIMATION OF TOA

Direct application of MUSIC algorithm to equations (3,5) is very expensive to implement due to the large spreading codes besides the oversampling factor Q . In practice, matched filtering is more convenient, especially since it can be followed by a MUSIC algorithm to obtain higher resolution estimations. In fact, in this case, the MUSIC algorithm is carried out on a data length equal to an order of magnitude of the channel length and not equal to the code length. Nevertheless, matched filtering is not *near-far* resistant. As we mentioned power-up and IS techniques are proposed to combat the problem of hearability in the UL and DL scenarios, respectively. However, the power in the UL scenario cannot be increased infinitely, this is why *near-far* resistant algorithms like Parallel Interference Cancellation (PIC) must be used in the BS to improve the TOA estimation. On the other hand, the IS are periods of silence of BSs positioned in a predictable manner and communicated to the mobiles. This will allow the mobiles, in the silent BS, to hear the far BS.

In the UL scenario a bank of matched filters matched to the control channels of the users is used as the first stage of PIC in the BS will give a first estimation of the channels coefficients:

$$\hat{\mathbf{h}}^{UL(1)} = \mathbf{S}^H \mathbf{X}_1 \mathbf{h}_1^{UL} + \mathbf{S}^H \mathbf{X}_t \mathbf{h}_t^{UL} + \mathbf{S}^H \mathbf{w} \quad (6)$$

where \mathbf{H} represent transconjugate, \mathbf{S} is the matrix of the locally generated replicas of the scrambling and the control channelization codes of all the users, $\mathbf{X}_1 \in \mathbb{C}^{LQ \times M_1}$, $\mathbf{X}_t \in \mathbb{C}^{(LQ \times \sum_{k \neq 1} M_k)}$ and \mathbf{h}_1^{UL} , \mathbf{h}_t^{UL} are the searched user and all other users channel coefficients, respectively. The second term represents interference to the searched users channel estimation. PIC uses first estimation $\hat{\mathbf{h}}^{UL(1)}$ for the interference subtraction. Since the data and control bits are available at the BS, then \mathbf{X} can be reconstructed, and we get:

$$\hat{\mathbf{h}}^{UL(2)} = \mathbf{S}^H \mathbf{X}_1^{UL} \mathbf{h}_1^{UL} + \mathbf{S}^H \mathbf{X}_t^{UL} (\mathbf{h}_t^{UL} - \hat{\mathbf{h}}^{UL(1)}) + \mathbf{S}^H \mathbf{w}$$

The second term now is less important than that of equation (6). $\mathbf{S}^H \mathbf{X}_1 = \mathbf{R}^{UL} \in \mathbb{C}^{L' \times M_1}$ is a cross-correlation matrix between the global signal and the scrambling and control channelization code.

A matched filter in the mobile phone (DL case) with the pilot channel data known will give:

$$\begin{aligned} \hat{\mathbf{h}}^{DL} &= \mathbf{S}_p^H \mathbf{y}^{DL} = \mathbf{S}_p^H \mathbf{X}_{BS} \mathbf{h}^{DL} + \mathbf{S}_p^H \mathbf{w} \\ &= \mathbf{R}_p \mathbf{h}^{DL} + \mathbf{R}_t \mathbf{h}^{DL} + \mathbf{y}_w^{DL} = \mathbf{R}^{DL} \mathbf{h}^{DL} + \mathbf{y}_w^{DL} \end{aligned}$$

where $\mathbf{R}_p^{DL} \in \mathbb{C}^{LQ \times M}$. The second term is unknown for the mobile which represents the intra-cell interference and it is not negligible and depends on the orthogonality factor of the link level.

In the two cases, an averaging of these matched filter outputs over a number J of independent snapshots is necessary to get rid of the noise and interference components. A threshold γ then should be chosen and the values of the arguments of $\hat{\mathbf{h}}$ where $\hat{\mathbf{h}} > \gamma$ are relative to TOA estimations. It is clear that if the paths i are close to each other (to less than one chip period) then the matched filter is unable to resolve them and a bias could be introduced. In order to have more accurate results a high resolution method must be applied.

The MUSIC algorithm can be applied to the output of the delay profile vector $\hat{\mathbf{h}}$ after an interpolation of the latter by N times to obtain higher resolution TOA estimations [6]. The correlation matrix of the delay profile vector can be estimated using the J independent snapshots:

$$\hat{\mathbf{\Lambda}} = \frac{1}{J} \sum_{j=1}^J \hat{\mathbf{h}}_j \hat{\mathbf{h}}_j^H = \mathbf{RGR}^H + \frac{\sigma_w^2}{T_e} \mathbf{R}_0$$

where \mathbf{R}_0 is the Hermitian correlation matrix of the de-scrambling code and \mathbf{G} is the correlation matrix of the channel coefficients. Now, the generalized eigenvalue problem $\hat{\mathbf{\Lambda}} \mathbf{e}_i = \lambda_i \mathbf{R}_0 \mathbf{e}_i$ for $i = 1, \dots, \sum_k M_k$ for the UL and $i = 1, \dots, M$ for DL, must be solved. The TOA estimations are then the maximums of the criterion:

$$v(\tau) = \frac{\mathbf{r}_\tau \mathbf{r}_\tau^H}{\mathbf{r}_\tau \mathbf{E} \mathbf{E}^H \mathbf{r}_\tau^H} \quad (7)$$

where \mathbf{E} is the matrix of the column eigenvectors from $i = K, \dots, \sum_k M_k$ or M and \mathbf{r}_τ are the inter-correlation vector, at shift τ , between the baseband signal and the De-scrambling and channelization sequences.

4 SIMULATION RESULTS

Simulations are carried out for the two scenarios using Walsh codes for channelization and Large Kasami set of length 256 as scrambling code in the UL and DL. Propagation in micro-cell environment is simulated using the channel delay profile proposed by the ETSI [9] with an average delay spread of 750 ns. The first path is taken always as LOS path. The path loss is simulated using the ITU recommendations for Manhattan micro-cell deployment; the effect of shadowing with a standard deviation of $\sigma = 4$ dB is also added to the path loss. A bit rate of 16 kHz is chosen, $Q = 3$. A three point interpolation $N = 3$ is carried out on the matched filter output before the application of MUSIC algorithm. A required $E_b/N_0 = 3$ dB, at the BS in the UL case is considered, with a transmission power of 14 dBm at the MS. A pilot channel power of 33 dBm and a power of 20 dBm [9] per user channel are simulated in the DL case. Other parameters such as frequency are chosen according to the WCDMA FDD features. The mobile movement is taken to be 3km/h, then according to the channel simulations the coherence time was about 30 ms so the channel can be considered static over the observation period $T=1/16000=62.5 \mu s$, 25 observation periods were combined noncoherently to combat interference and fast fading. Having one observation each 35 ms will give a LP in 0.875 s. Hence, the mobile phone can be considered static for 0.875s at a 3 km/h speed. The radio location algorithm used is the one presented in [10] using 3 and 4 BS. A power-up

of 20dB is simulated to combat the near-far problem in the UL scenario; a constraint on the maximum UL transmitted power of 24dBm is imposed. While, for the DL scenario silent neighboring BS are assumed when TOA estimation from a given BS is carried out.

During each observation a random deployment of the MS interferers together with the static searched mobile in the cell are simulated. Zero intercell interference is assumed and background noise is added. Figure (3) depicts the circular error cumulative probability density function (CDF) of the estimated locations using 3 BS with 5 and 30 users at each cell. The results of the two TOA estimation methods and the two scenarios are plotted together. Figure (4) shows the same results using 4 BS. Figure (5) shows a scatter plot of the position location using 4 BS in the DL scenario with matched filter and MUSIC TOA estimation methods and having 30 users.

5 DISCUSSIONS

The results shows more accurate location positioning in the DL scenario. This is due to the synchronous sum of orthogonal users before convolution by a single channel (see equation (4)). While, signals in the UL are transmitted from different locations, which means asynchronous reception of the users (see equation (2)). As seen from the equation, we have more parameters to estimate from the same given number of observations as the DL case. Although the DL scenario give better PL estimations, it requires modifications in the mobile phone to make the required computations especially the implementation of the MUSIC algorithm. Robust *near-far* resistant methods are not easy to implement in practice due to the large spreading factors. However, the use of high resolution algorithms on the delay profile improves the location positioning provided that the number of observations are sufficiently high to give full rank channel coefficients correlation matrix. This is why the LP time in UL scenario may increase drastically with the number of users. The LP is well improved by increasing the number of involved BS.

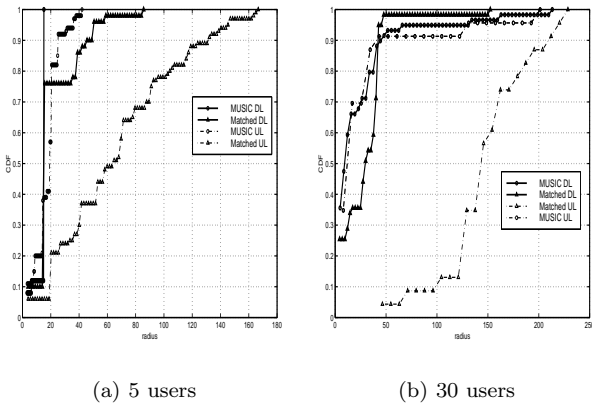


Figure 3: Circular CDF of the location positioning error using 3 BS

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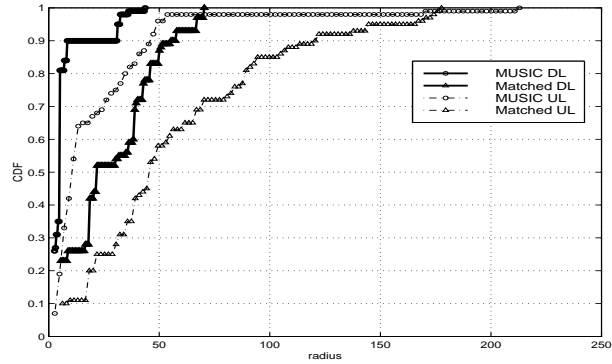


Figure 4: Circular CDF of the location positioning error using 4 BS and having 5 users

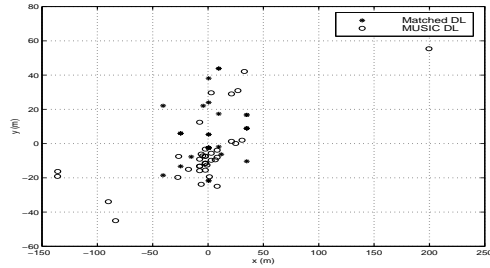


Figure 5: Scatter plot of the location positioning using 4 BS and 30 users in the DL scenario, (0,0) is the exact position

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