# AN EFFICIENT RAKE-CFAR METHOD FOR DOWNLINK MOBILE POSITIONING IN UMTS FDD MODE.

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# ABSTRACT

This paper deals with position location (PL) problem in the context of Wideband Code Division Multiple Access (WCDMA) mode of UMTS systems using the Time Of Arrival (TOA) in the Downlink (DL) scenario. We propose to use a RAKE receiver in conjunction with a Constant False Alarm Rate (CFAR) detector in order to optimally detect the first arriving path. An improved version is also considered using subspace projection and/or high resolution TOA estimation technique. The proposed method allows us to obtain a good estimation of the far-located BS's Time Of Arrival (TOA) and hence, it represents a good solution to the hearability problem. Realistic simulations show the accuracy improvement provided by the proposed detector over a simple RAKE receiver.

*Index Terms*— Mobile positioning, RAKE receiver, CFAR detection.

# 1. INTRODUCTION

Nowadays, the determination of the position of a transmitting mobile station (MS) is becoming a compulsory requirement for mobile operators. Responding to E-911 calls, providing location specific service information, navigation aid and user's tracking, are examples that require an MS location [2]. The frequency with which location requests are made and the desired accuracy vary with the application. The most widely employed location technologies are radio location systems that attempt to locate an MS by measuring radio signals between the MS and a set of base stations (BSs). The main approaches proposed to locate a mobile are based on either signal strength, angle of arrival (AOA), time of arrival (TOA), or time difference of arrival (TDOA) and can be network- or terminal-based. For an overview of various wireless location techniques and technologies, see [2] and [3]. In this paper, we are interested in the MS localization based on TOA techniques in the downlink UMTS-FDD mode [1]. The accuracy of this technique is strongly linked :

- To the existence of a Line Of sight (LOS) path.
- To the accuracy of the estimation of the time of arrival.

In this contribution, we derive new accurate PL methods based on TOA. The first multi-path component will be supposed to represent the LOS path. As far as the accuracy of the estimation of TOA is concerned, a high resolution method has been considered. However, due to the long-code scrambling sequences, it is not conceivable to implement directly subspace based techniques on the received signal. A simple correlation<sup>1</sup> is used to obtain a first estimate of the

TOA at the mobile station. Then, a refined estimation is obtained by projecting the output of the correlation onto the signal subspace resulting in a significant reduction of the noise level.

However, power control algorithms are often used in order to minimize the signal strength emitted at the considered BS. As a result. it is difficult for a MS to "hear", with a good SNR, multiple and far located BS's. To circumvent this near-far problem, it was proposed to insert Idle Period DownLink (IDPL) [5] periods during which the BS's stop transmitting successively. This makes it possible for the MS to 'hear' far-located BS's. However, it is obvious, that the introduction of IPDL reduces a lot the network capacity. In this paper, we propose and compare several solutions to avoid the insertion of IDPL: a consistent and improved estimation of Rayleigh Fading Channel and an optimal first path detection using the CFAR technique. This allows us to estimate with a good accuracy the TOA of far located BS's without using IPDL.

# 2. DOWNLINK RECEIVED SIGNALS

The choice of the downlink scenario rather than the uplink scenario is justified by the fact that the common pilot channel in downlink is transmitted by the base-station continuously with a relatively high power. The signal models are constructed according to the UMTS-FDD standards [7].

A UMTS network is divided into cells, each of them (denoted c) contains a BS c which communicates with  $K^c$  mobile stations. For each slot l, BS c emits simultaneously :

- a QPSK sequence  $d_{l,0}^c$  of  $N_s$  symbols ( $N_s = 10$ ) representing the pilot sequence common for all users of the cell.
- $K^c$  other QPSK sequences  $d_{l,k}^c$ ,  $k = 1 \cdots K^c$  corresponding to the  $K^c$  users' signals of the cell.

Each symbol sequence is spread by a BPSK periodic sequence  $c_k^c$  of period N = 256, which represents the spreading factor. For simplicity sake, the spreading factor is supposed lo be the same for all users. Thanks to these spreading codes, it is possible to separate the different users of a same cell. These codes are orthogonal and selected from the Walsh family. Then, the resulting sequences are finally scrambled by the same long aperiodic code  $s_l^c(n)$  which characterizes the considered cell c. They are selected from the GOLD family. For each sequence k, the *i*-th chip of symbol n, emitted during the slot l in the cell c, is expressed as :

$$b_{l,k}^{c}(nN+i) = d_{l,k}^{c}(n)c_{k}^{c}(i)s_{l}^{c}(nN+i)$$

The pulse-shaping filter by which the symbols are modulated, before being emitted, is a square root raised cosine of roll-off 0.22.

 $<sup>^{\</sup>rm l}$  We refer to this receiver as RAKE since it uses the pilot-data correlation eventhough we do not combine the different paths as in the real RAKE receiver.

Through a discrete multi path channel, the signal corresponding to the slot l received by the MS from the BS c is thus given by:

$$x_{l}^{c}(t) = \sum_{k=0}^{K^{c}} G_{k}^{c} \sum_{r=1}^{R^{c}} \sum_{i=0}^{N_{s}N-1} \alpha_{r,l}^{c} b_{l,k}^{c}(i) g(t - iT_{c} - \tau_{r}^{c})$$

where  $G_k^c$  represents the gain factor of the kth user signal of the cth cell. For each cell,  $R^c$  paths are considered characterized by their delays  $\tau_r^c$  and their fading coefficients  $\alpha_{r,l}^c$ . g represents the global filter, (i.e. pulse shaping filter + receiver filter). If the considered MS receives signals from C BSs, the global received signal (corrupted by additive noise  $w_l$ ) is given by :

$$x_l(t) = \sum_{c=1}^{C} x_l^c(t) + w_l(t)$$

At the receiving end, the signal is sampled at the rate  $T_c$  and we denote  $x_l(n) \stackrel{\triangle}{=} x_l(nT_c)$ .

# 3. TOA ESTIMATION

For each cell, under the assumption of a LOS, the considered TOA can be obtained as the position of the first peak of the autocorrelation function of the channel coefficients. In practice, the latter are classically obtained by the RAKE estimator which carries out correlations of the received signal with delayed versions of the pilot sequence. For example, the channel estimate of the *c*th cell, corresponding to the received signal during the slot l, is given by :

$$\hat{h}_{l}^{c}(k) = \frac{1}{N_{s}N} \sum_{i=0}^{N_{s}N-k} x_{l}(i+k) b_{l,0}^{c^{\star}}(i).$$
(1)

To determine the TOA, coherent averaging of the channel coefficient estimates over several slots, supposing the delays constant during this observation period, provides a better accuracy. In this case, it is possible to determine the TOA by estimating the peaks of the following function  $\frac{1}{J} \sum_{l=1}^{J} \left| \hat{h}_{l}^{c}(k) \right|^{2}$ ,  $k = 0 \cdots L - 1$ , where the channel coefficients have been averaged over J snapshots (J denotes the number of slots used for the channel estimation). However, even with this new function, the peak estimation requires the use of a threshold  $\gamma$  to get rid of the noise peaks and to retain only peaks corresponding to effective channel paths. Usually, this thresholding is done in an adhoc way<sup>2</sup> which may affect the performance of mobile localization. Indeed, a high threshold might hide the first path if the latter is of relatively low power and a low threshold would lead to a false (noise) peak detection. In fact, the performance of the thresholding depends essentially on the robustness of the used channel estimation to the Near-Far Effect NFE (i.e. on the effect of the noise terms in the channel estimates). As the RAKE estimator is known to be non robust to the NFE, we propose to use a robust estimation of the channel parameters (the RAKE-SP estimator proposed in [9]: SP stands for subspace projection) with a CA-CFAR based method to reduce the false peak detection probability.

### 3.1. Noise mitigation using subspace projection

The channel estimate given in equation (1) suffers from the nearfar problem and several improvements have been proposed in the literature to combat this effect as in [4]. In this paper, we propose to use the projection onto the signal subspace to reduce the noise level at the correlation output as in [9]. Indeed, the correlator output  $\hat{\mathbf{h}}_{l}^{c} \triangleq [h_{l}^{c}(0) \cdots h_{l}^{c}(L)]^{T}$  represents the channel coefficient vector estimate. Given the data model the latter can be written as

$$\hat{\mathbf{h}}_{l}^{c} = \mathbf{U}\mathbf{g}_{l}^{c} + \boldsymbol{\epsilon}_{l}^{c}$$
<sup>(2)</sup>

where

$$\mathbf{U} = [\mathbf{u}(\tau_1^c) \cdots \mathbf{u}(\tau_{R_c}^c)]$$
  
$$\mathbf{u}(\tau) = [g(-\tau) \ g(T_c - \tau) \cdots g((L-1)T_c - \tau)]^T$$
  
$$\mathbf{g}_l^c = G_0^c [\alpha_{1,l}^c \cdots \alpha_{1,R_c}^c]^T$$

and  $\epsilon_l^c$  represents the channel estimation error modeled as a gaussian noise vector.

Clearly, in the noiseless case, the channel coefficient vector lives in a reduced dimensional subspace Range (**U**) referred to as the signal subspace. The dimension of the latter is equal to  $R_c$  (the number of multipaths or eventually an upper bound of it) and can be much smaller than L (the number of considered channel coefficients). Consequently, an orthogonal projection onto the signal subspace, i.e

$$\tilde{\mathbf{h}}_{l}^{c} = P_{\mathbf{U}} \hat{\mathbf{h}}_{c}^{l} \tag{3}$$

would lead to a reduction of the noise power by a factor  $\frac{R_c}{L}$ . In (3),  $P_U$  denotes the orthogonal projection onto the range space of U. We refer to this estimate as the RAKE-SP channel estimate. In practice, the signal subspace is estimated from the sample averaged channel covariance matrix. Efficient algorithms, e.g the FOOja [8], can be used to extract and track the latter in an adaptive scheme for a complexity load equal to  $O(R_cL)$  flops per iteration. Note that other robust estimation procedures of the channel parameters can be considered as alternative solutions, e.g [6].

#### 3.2. RAKE-SP-CFAR

After the channel estimation step, our objective now is to detect the first path of the propagation channel and estimate its corresponding delay  $\tau_1^c$  (we assume that  $0 < \tau_1^c < \tau_2^c \cdots < \tau_{R_c}^c$ ). For that we use the energy estimate function

$$E^{c}(k) = \frac{1}{J} \sum_{l=1}^{J} \left| \tilde{h}_{l}^{c}(k) \right|^{2}, \ k = 0 \cdots L - 1$$

and seek for the location of its peaks (maximal) that correspond to the desired delay parameters.

Due to noise effect, a thresholding procedure is needed to distinguish virtual 'noise peaks' from the actual desired peaks. In [9] a constant ad-hoc threshold value (taken equal to a percentage of the main peak value) has been considered. In this work, we consider an adaptive 'optimized' threshold value that maximizes the correct detection probability of the first channel path.

To this end, we have used statistical hypothesis testing tools. Specifically, we derive the threshold value using the theory of the constant false alarm rate (CFAR) detector.

The latter has been extensively studied by the radar processing community for target detection in presence of clutter signal and hence, for simplicity, we borrow the same terminology here for our considered problem. The test cell<sup>3</sup> represents the energy sample  $E^{c}(k)$  we would like to test to see whether it corresponds to a peak value or

<sup>&</sup>lt;sup>2</sup>It can be taken equal to a percentage of the strength of the received signal

<sup>&</sup>lt;sup>3</sup>In this section, cell means sample.



Fig. 1. CA-CFAR detector

not. The reference cells correspond to the neighboring energy samples (see figure 1) to which the test cell is compared, i.e  $E^{c}(k - k)$  $\frac{m}{2}$ ) · · ·  $E^{c}(k-1), E^{c}(k+1) \cdot \cdot \cdot E^{c}(k+\frac{m}{2})$  where m is a chosen window parameter.

In this paper, we compare the energy level of the test cell to the averaged energy level of the reference cells :

$$E_{average}^{c} \stackrel{\triangle}{=} \sum_{i=k-\frac{m}{2}, i\neq k}^{k+\frac{m}{2}} E^{c}(i)$$

and hence we refer to the method as cell-averaging CFAR (CA-CFAR) method. The probability of false alarm represents the probability of deciding a peak value while it is not and the probability of detection is the probability of detection of a real peak value.

For a given threshold multiplier  $\gamma_s$ , these two probabilities can be written as

$$P_{fa} = P(E_{test}^c \ge \gamma_s E_{average}^c | H_0)$$
$$P_D = P(E_{test}^c \ge \gamma_s E_{average}^c | H_1)$$

where  $H_0$  (resp  $H_1$ ) denotes the hypothesis of absence of peak value (resp. presence of peak value) at the test cell. In our considered application, the false alarm and the non-detection are equally harmful and result in an estimation bias of the mobile location.

Therefore, our objective is to minimize the probability of 'wrong' detection corresponding to the sum of the probability of false alarm  $P_{fa}$  and the probability of non-detection  $1 - P_D$ . In other words, we minimize the cost function

or equivalently

$$\min_{\gamma_s} P_{fa} + 1 - P_D$$

$$\min_{\gamma_a} P_{fa} - P_D. \tag{4}$$

The following assumptions are considered for the theoretical derivation of the above expression w.r.t  $\gamma_s$ . The channel is of Rayleigh type in the sense that the fading coefficients  $\alpha_{r,l}^c$  are independent random (complex) gaussian variables of power  $\sigma_r^c$  (i.e the multipath amplitudes are iid Rayleigh) and the estimation noise at the correlator output  $(\epsilon_l^c(k))$  is white complex gaussian with power  $\sigma_{\epsilon}^c$ .

We define the (correlator output) SNR value of the cth cell channel first path as

$$SNR^c \stackrel{\scriptscriptstyle \Delta}{=} \frac{\sigma_1^c}{\sigma_\epsilon^c}.$$

The latter is assumed known or adaptively evaluated over past measures.

Under these assumptions, the theoretical expressions of  $P_{fa}$  and  $P_D$ in the mono-slot case are [12]

$$P_{fa} = \left[1 + \frac{\gamma_s}{m}\right]^{-m}$$
$$P_D = \left[1 + \frac{\gamma_s}{m(1 + SNR^c)}\right]^{-m}$$

Given the above expressions, it is easy to verify that the global minimum of (4) is reached for

$$\gamma_{opt} = \frac{m(1 - \sqrt[m+1]{\alpha})}{\sqrt[m+1]{\alpha} - \alpha}$$

where  $\alpha = \frac{1}{1 + SNR^c}$ . **Remark**: several improvements can be considered for the CFAR detector using for example the order statistics to better deal with the case where other peak values exist in the reference cells or by using the guard cells [4] to mitigate the effect of neighboring peaks due to the possibly non integer values of the propagation delays (i.e  $\tau_1^c \neq kT_c \text{ for } k \in \mathbb{N}$ ).

# 3.3. High-resolution TOA estimation

The accuracy of the proposed RAKE-SP-CFAR estimation cannot be lower than  $T_c/2$  (i.e half of the sampling rate) which corresponds approximately to a location error of 37 meters in UMTS-FDD. A better estimation accuracy of the delay parameter can be obtained with an oversampling but this would not help improving the resolution of closely spaced multipaths (i.e if  $\tau_2^c - \tau_1^c \leq T_c$ , one would observe one single peak instead of two peaks).

For this reason, we propose to use instead of the oversampling, a high-resolution TOA estimation technique, namely the MUSIC algorithm in [10].

In other words, we propose a two-step procedure :

- A rough estimation of the first path delay parameter using RAKE-SP-CFAR.
- A refining of the delay parameter estimate using MUSIC algorithm applied to a reduced size window centered at the first peak position.

Indeed, MUSIC is based on an eigendecomposition of the channel covariance matrix. Using a reduced size window of length  $\Delta << L$ reduces the computational cost from  $O(L^3)$  to  $O(\Delta^3)$  in batch processing.

Another drawback of MUSIC resides in the computationally expensive numerical search needed to optimize the MUSIC cost function (see [10] for details). In our case, as we have already a rough estimate of the delay parameter  $\hat{\tau}_1^c$ , one can again limit the search to a small window centered at  $\hat{\tau}_1^c$ . After estimation of the TOA's of at least three BS's, the mobile position is computed by triangulation as done in [11].

### 4. SIMULATIONS

A macrocell environment has been simulated according to [13]. Each estimate of the channel is obtained from the observations corresponding to a slot (i.e. with 10 symbols). During this period of observation, the coefficients of the channel are supposed to remain constant. We have considered three paths with time-varying and IID complex amplitudes corresponding to a mobile speed of 3 Km/h for each channel. Each cell contains  $K^c = K$  interfering mobiles randomly distributed. A background gaussian noise representing 10% of the farest located base station power has been added. During the observation period chosen equal to 120 slots, the channel fading coefficients  $\alpha_{r,l}^c$  (which are considered to be *IID* gaussian variables) are supposed to vary at each slot with relative powers satisfying  $\sigma_2^c = \sigma_3^c = \sigma_1^c(1 + \alpha)$  where  $\alpha$  is chosen randomly using uniform distribution in  $[-0.7 \ 0.7]$ . The delays  $\tau_r^c$  are supposed to be constant ( $\tau_1^c = \frac{d(MS - BS^c)}{c_0}$ ,  $\tau_2^c = 5\tau_1^c$ ,  $\tau_3^c = 7\tau_1^c$ , where  $c_0$  represents the speed of light). In all our simulations, the results are evaluated over 150 Monte-Carlo runs. For comparison, we consider the approach in [9] where the threshold  $\gamma_s$  used for RAKE and RAKE-SP has been taken equal to 25% of the strength of the received signal. Different scenarios have been simulated to illustrate the impact of the proposed algorithms on the detection of the first arrival path.

In figure 2, we present the configuration of the macrocell environment, we suppose in our simulations that the mobile receives signals from three diffrent BS. We should note that the BS indices are numbered according to their closeness to the MS.

In figure 3, we measure the accuracy of detecting the first arriving path of each BS signal. The result has been obtained from 150 runs corresponding to a random position of the mobile for K = 30 interfering mobiles randomly distributed at each run. As we can see, the RAKE-SP-CFAR detector behaves better than RAKE-SP, similarly to the RAKE-CFAR as compared to RAKE detector. Thus, our proposed algorithm reduces the false peak detection probability. We can also observe, the near-far effect on the detection performance, in fact the detection accuracy for BS 1 is better than that of BS 2 which is in its turn better than BS 3 which is due on the difficulty of the MS to 'hear' far-located BS's.

In figure 4, the same scenario is considered but the corresponding curves represent the circular error cumulative probability density function (CDF) versus the mobile location error. The same scenario is repeated in figure 5 but this time, we have introduced only K = 20 interfering mobiles. As it is shown by simulation results of Figures 4 and 5, our proposed algorithms, RAKE-SP-CFAR and RAKE-CFAR perform better than RAKE-SP and RAKE, respectively. The estimation of the TOA is more accurate by using MUSIC algorithm as compared to the RAKE estimate but it requires more computational cost.

To get more insight on the behavior of the proposed algorithms with respect to the near far effect, we consider two different scenarios. Figure 6 has been obtained from 150 runs corresponding to a position of the mobile very close to the serving BS (d(MS - BS1) = 200 m, d(MS - BS2) = 2.65 Km, d(MS - BS3) = 3.1 Km) for k = 25 interfering mobiles randomly distributed at each run. In Figure 7 we have the same scenario but with different MS position (d(MS - BS1) = 1, 19 Km, d(MS - BS2) = 1.99 Km, d(MS - BS3) = 2.12 Km). As we can observe in all considered scenarios, our algorithms behave much better than RAKE-SP and RAKE which are sensitive to the hearability problem.

#### 5. CONCLUSION

In this paper, a new CA-CFAR based algorithm has been introduced in order to optimally detect the first arrival path and to reduce the NFE problem on mobile positioning. The presented results clearly demonstrate the efficiency of the proposed algorithms with respect to the existing ones. The proposed algorithms have the advantage of requiring no change in the UMTS-FDD standard and do not need a high computational cost which makes their implementation possible at the MS side.



Fig. 2. Configuration of the macrocell



Fig. 3. Accuracy methods for random mobile position : K= 30 users.

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Fig. 4. Random mobile position : K= 30 users per cell.



Fig. 5. Random mobile position : K= 20 users per cell.

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Fig. 6. Fixed mobile position (at 200 m to the serving BS): K= 25 users per cell.



Fig. 7. Fixed mobile position (at 1190 m to the serving BS): K= 25 users per cell.