Assessment of Direct Positioning for IR-UWB in IEEE 802.15.4a Channels

Monica Navarro*, Pau Closas* and Montse Nájar†
*Centre Tecnològic de Telecomunicacions de Catalunya, Castelldefels, Barcelona, Spain
†TSC - Universitat Politècnica de Catalunya, Barcelona, Spain
{monica.navarro,pau.closas}@cttc.es, montse.najar@upc.edu

***Abstract*—This paper assesses the problem of localization in IR-UWB under realistic channel models for Direct Position Estimation (DPE) approaches. DPE schemes have been proposed for positioning and localization for well developed systems like GNSS, where it has been analytically proved that the Maximum-likelihood single-step estimator outperforms two-step procedures. The extension to wideband systems and less favorable scenarios like indoor UWB channels is less explored. We derive a DPE algorithm and analyze its performance against two-step TOA based localization for an IR-UWB system. Numerical results are provided for IEEE 802.15.4a channel model showing positioning performance of the two approaches and highlighting the trade-offs.**

I. Introduction

In general, the positioning problem is divided into two subproblems: ranging to anchor nodes at known locations and trilateration to obtain a position estimate. The former is highly related to the Time Of Arrival (TOA) estimation problem, whereas the latter is typically solved by a simple least squares algorithm. This is also the case in IR-UWB [1, 2]. However, recent works claim that this approach to positioning could be potentially improved, specially in challenging scenarios, if one treats the problem as a whole. That is to convert the two-step approach into a single-step procedure where digitized signals are combined to solve directly for the position coordinates. This positioning approach is known as Direct Position Estimation (DPE) in the literature.

Two different localization problems have been addressed in the literature by DPE. The first one consists on self positioning of a receiver by the signal coming from different synchronized transmitters at known locations. This is the case of the Global Navigation Satellite System (GNSS) where the DPE approach, based on Maximum Likelihood Estimation (MLE), has been considered in [3]. In this scenario, higher accuracy, compared with the two steps approach, has been proved given by the Cramer-Rao Bound (CRB) [4]. On the other hand, the localization of one transmitter from the received signals at different anchor nodes at known locations has been also addressed with DPE. In this case, the DPE requires the transmission of the received signals to a central processing node while the two-stage approach requires only the transmission of the estimated parameters. However, [5] proved that the standard two-step approach cannot overcome the performance of DPE. Dealing with this problem, DPE has been proposed for narrowband emitters positioning in [6] and for passive geolocation [7] applying MLE.

In UWB scenarios, the dense multipath channel poses additional challenges to positioning algorithms. Authors in [8, 9] propose applying MUSIC algorithm for solving DPE for self positioning, which requires knowledge of the channel attenuations for the LOS paths. In this paper, we investigate the feasibility of DPE for IR-UWB considering the localization paradigm where the actual processing is performed at the anchor nodes. Building upon our previous experience in both DPE and IR-UWB positioning, we propose a frequency domain receiver architecture to accomplish DPE in IR-UWB systems for high resolution estimators based on periodogram. We formulate the problem in the frequency domain and relate the parameters of the resulting line-of-sight (LOS) signal to the position coordinates of the receiver. We came out with a non-convex, multidimensional optimization problem that could be solved in a greedy manner or via more sophisticated methods. We present simulation results in realistic scenarios as those channel models provided in the standard IEEE 802.15.4a.

The rest of the paper is organized as follows: Section II introduces the IR-UWB signal model, Section III reviews the two-step positioning scheme based on TOA estimates, the direct position estimator is introduced in Section IV. Performance evaluation is given in Section V and conclusions are drawn in VI.

II. System Model

We consider an IR-UWB system where transmission of an information symbol is typically implemented by the repetition of N_f pulses of very short duration. The transmitted signal is expressed as,

$$s(t) = \sum_{k=-\infty}^{\infty} \sum_{j=0}^{N_f-1} a_j p(t-(kN_f+j)T_f - c_jT_c - b_kT_\delta) \quad (1)$$

where Pulse Position Modulation (PPM) is assumed with $\{b_k\}$ being the information symbols taking values $\{0, 1\}$ with equal probability. $p(t)$ refers to the single pulse waveform, being typically a Gaussian monocycle or one of its derivatives of duration T_p. $T_{sym} = N_fT_f$ is the symbol duration, where $T_f = N_c \cdot T_c \gg T_p$ is the repetition pulse period also referred to as frame period, and N_f is the number of frames per symbol, T_c is the chip period, T_δ is the PPM modulation

978-1-4799-0969-8/13/$31.00 ©2013 IEEE

interval, N_c is the number of chips per frame and $\{c_j\}$ is the time hopping sequence which takes integer values in $\{0, 1, \ldots, N_c - 1\}$ and $a_j = \pm 1$ denotes a polarization sequence typically used for spectrum shaping. Without loss of generality we assume in the sequel $a_j = 1 \; \forall j$.

The channel model considered is given by the general expression for the multipath fading propagation channel as follows,

$$h(t) = \sum_{m=0}^{M-1} h_m \delta(t - \tau_m) \tag{2}$$

With no loss of generality we assume $\tau_0 < \tau_1 < \ldots < \tau_{M-1}$, being τ_0 the TOA that is to be estimated.

The received signal is then the summation of multiple delayed and attenuated replicas of the received pulse waveform $\tilde{p}(t)$ which includes the antenna and filters distortion,

$$y(t) = \sum_{m=0}^{M-1} \sum_{k=-\infty}^{\infty} \sum_{j=0}^{N_f-1} h_m \tilde{p}(t - \Delta_{k,j} - \tau_m) + v(t) \tag{3}$$

where $\Delta_{k,j} \triangleq (kN_f + j)T_f - c_j T_c - b_k T_\delta$. We assume the received pulse from each m-th path exhibits the same waveform but experiences a different fading coefficient, h_m, and time delay, τ_m. The additive noise $v(t) \sim \mathcal{N}(0, N_0)$ is modeled as Gaussian circularly symmetric. Given the low duty cycle of UWB signals we assume the received signal is free of intersymbol interference (ISI).

The signal associated to the *j*-th transmitted pulse corresponding to the *k*-th symbol in the frequency domain yields:

$$Y_{j,k}(w) = h_0 S_{j,k}(w) e^{-jw\tau_0} + \sum_{m=1}^{M-1} h_m S_{j,k}(w) e^{-jw\tau_m} + V(w) \tag{4}$$

where the LOS contribution is explicitly separated in the first term from the signal replicas associated to multipath. The frequency component associated to the shifted pulse is given by,

$$S_{j,k}(w) = \tilde{P}(w) e^{-jw(\Delta_{k,j})} \tag{5}$$

with $\tilde{P}(w) = \mathcal{F}\{\tilde{p}(t)\}$ and $V(w) = \mathcal{F}\{v(t)\}$ denoting by $\mathcal{F}\{.\}$ the Fourier transform.

Sampling (4) at $w_n = w_0 n$ for $n = 0, 1, \ldots, N-1$ being $w_0 = 2\pi/N$ and rearranging the frequency domain samples $Y_{j,k}[n]$ into the vector $\mathbf{Y}_{j,k} \in \mathbb{C}^{N\times 1}$ yields,

$$\mathbf{Y}_{j,k} = h_0 \mathbf{S}_{j,k} \mathbf{e}_{\tau_0} + \tilde{\mathbf{V}}. \tag{6}$$

with $\tilde{\mathbf{V}} = \mathbf{S}_{j,k} \mathbf{E}_\tau \mathbf{h} + \mathbf{V}$ and $\mathbf{S}_{j,k} \in \mathbb{C}^{N\times N}$ is a diagonal matrix which components are the frequency samples $S_{j,k}(w)$. The matrix $\mathbf{E}_\tau \in \mathbb{C}^{N\times M-1}$ contains the delay-signature vectors (harmonic components) associated to each arriving delayed signal (multipath),

$$\mathbf{E}_\tau = [\; \mathbf{e}_{\tau_1} \;\; \ldots \;\; \mathbf{e}_{\tau_j} \;\; \ldots \;\; \mathbf{e}_{\tau_{M-1}} \;] \tag{7}$$

with $\mathbf{e}_{\tau_j} = [\; 1 \quad e^{-jw_0\tau_j} \quad \ldots \quad e^{-jw_0(N-1)\tau_j} \;]^T$.

The channel fading coefficients, except for h_0, are arranged in the vector $\mathbf{h} \in \mathbb{R}^{(M-1)\times 1}$ and the noise samples in vector $\mathbf{V} \in \mathbb{C}^{N\times 1}$.

Next, we introduce the standard two-step positioning approach based on TOA measurements. The choice of TOA is well established in IR-UWB based positioning schemes [10] given its accuracy. Without loss of generality the estimator is formulated for a two-dimensional space. The extension to the three dimensional space is straight forward by increasing the number of anchor nodes required for a reliable position estimate.

III. TOA-BASED POSITION ESTIMATION FOR IR-UWB

The most widely adopted positioning scheme in IR-UWB is based on the two-step approach. In particular, when reliable estimates of propagation delays are available TOA-based positioning has shown reliable performance. Next, we review the specific TOA-based positioning scheme considered for baseline performance.

A. Ranging

For the distance measurements, we consider the algorithm introduced in [1] for the estimation of the TOA. The high-resolution TOA estimate is based on calculating the pseudo-periodogram or power delay profile, defined as the signal energy distribution with respect to propagation delays. The power can be obtained by estimating the energy of the signal filtered by the delay-signature vector at each delay. Namely,

$$\hat{\tau}_{0,\ell} = \arg_\tau \max \mathbf{e}_\tau^H \mathbf{R}_\ell \mathbf{e}_\tau \tag{8}$$

where $\ell = 1, \ldots, N_A$ denotes the anchor index, the superscript $(\cdot)^H$ denotes the transpose complex conjugate, and $\mathbf{R}_\ell$ is the sample covariance matrix computed by averaging over properly aligned frequency samples

$$\mathbf{R}_\ell = \frac{1}{N_S} \sum_{n=1}^{N_S} \mathbf{Y}_{j,k}^{(\ell)} \left(\mathbf{Y}_{j,k}^{(\ell)}\right)^H \tag{9}$$

The quadratic form (8) allows for a low complexity implementation applying the FFT algorithm to a set of coefficients computed from $\mathbf{R}_\ell$ and a simple search algorithm. We shall remark that the actual TOA estimate is not the absolute maximum, but the first local maxima above a given threshold γ. We refer the reader to Algorithm 1 in [1] for details.

B. Positioning

The aforementioned TOA estimation procedure generates a set of measures that are used in the trilateration procedure to calculate the receiver position. In the sequel we briefly sketched this procedure. The scenario consists of a network of IR-UWB nodes deployed in a certain geographical area. Anchor nodes are those whose location is known in advanced and to which the target node is computing ranging information via TOA estimates. Let us define their two-dimensional coordinates respectively as

$$\mathbf{p}_\ell = [x_\ell, y_\ell]^T, \quad \ell = 1, \ldots, N_A, \tag{10}$$

$$\mathbf{p} = [x, y]^T, \tag{11}$$

where the set of N_A anchor nodes that provide coverage to the target node is defined as $\mathcal{N}$. A target node positions itself using trilateration, which requires at least three known distances to three known positions, i.e. anchor nodes.

The geometrical distance between the receiver and the ℓ-th anchor node is defined as

$$\varrho_\ell(\mathbf{p}) = \| \mathbf{p} - \mathbf{p}_\ell \| \quad , \tag{12}$$

with $\| \cdot \|$ being the Euclidean norm in $\mathbb{R}$. In general, the trilateration problem is posed as an overdetermined system of equations with the set of N_A noisy range measurements,

$$c\hat{\tau}_{0,\ell} \triangleq \hat{\rho}_\ell \sim \mathcal{N}(\varrho_\ell(\mathbf{p}), \sigma^2_{\rho_\ell}) \tag{13}$$

where c is the speed-of-light and the variance of the observations, $\sigma^2_{\rho_\ell}$, depends on the variance of TOA measurements as [2]. Then, the trilateration problem could be straightforwardly formulated as a Least Squares problem, where the goal is to minimize the squared-error

$$\hat{\mathbf{p}} = \arg\min_{\mathbf{p}} \left\{ \sum_{\ell \in \mathcal{N}} (\hat{\rho}_\ell - \| \mathbf{p} - \mathbf{p}_\ell \|)^2 \right\} \quad , \tag{14}$$

the optimization admits a closed-form solution [11] based on the Moore-Penrose pseudoinverse of the visibility matrix.

IV. Direct Position Estimation for IR-UWB

Direct Position Estimation (DPE) attempts to solve the positioning problem on a single step allowing to perform a multidimensional search directly over the spatial coordinates. The joint optimization problem is possible by relating the propagation delays with the geometric distances between anchors and target node. The motivation for considering DPE is two-fold: on the one hand the theoretical potential gains of the ML estimator [4, 5] with respect two-step approaches; on the second hand overcome the problem of threshold setting required in the TOA ranging problem. As discussed in the previous section, the estimation of delay associated to the LOS $\tau_{0,\ell}$ is carried out by finding the first relative maximum above a given threshold rather than finding the delay associated to the absolute maximum of the power delay profile, since it may correspond to replicas from multipath propagation. The accuracy of ranging based on TOA measurements strongly depends on properly selecting such threshold. The optimum threshold that minimizes ranging errors is scenario dependent and its value may vary considerably. Although there are some proposals for determining the threshold in a more general framework, in practice such selection typically requires a calibration step.

Taking the same set-up as for the two-step TOA-based positioning, the DPE problem can be formulated as follows. Lets define the observation frequency sample vector as the concatenation of signals received from all anchors, $\mathbf{Y} = (\mathbf{Y}^{(1)}, \ldots, \mathbf{Y}^{(\ell)}, \ldots, \mathbf{Y}^{(N_A)})^T$, with $\mathbf{Y}^{(\ell)} = h_{0,\ell}\mathbf{S}\mathbf{e}_{\tau_{0,\ell}} + \tilde{\mathbf{V}}^{(\ell)}$ according to (6) (symbol and pulse indexes are dropped in order to ease notation). The system model can then be written in terms of LOS contribution as,

$$\mathbf{Y} = \bar{\mathbf{S}}\mathbf{e}_{\mathbf{p}} + \tilde{\mathbf{V}} \tag{15}$$

with $\bar{\mathbf{S}} = \mathrm{diag}\,(h_{0,1}\mathbf{S}, \ldots, h_{0,N_A}\mathbf{S})$ being a block diagonal matrix with pulse spectral components weighted by the channel fading coefficients, $\mathbf{e}_{\mathbf{p}} = (\mathbf{e}_{f_1(\mathbf{p})}, \ldots, \mathbf{e}_{f_{N_A}(\mathbf{p})})^T$ the steering vector as a function of the target spatial coordinates and $\tilde{\mathbf{V}} = (\tilde{\mathbf{V}}^{(1)}, \ldots, \tilde{\mathbf{V}}^{(N_A)})^T$. The steering vectors $\mathbf{e}_{f_l(\mathbf{p})}$ are defined as in (7) but in this case the delay is related to the position vector $\mathbf{p}$ by the geometrical relation $f_l(\mathbf{p}) = \varrho_\ell(\mathbf{p})/c$.

The DPE position is solved applying the same principle as for TOA estimation. That is, the algorithm aims to maximize the periodogram by correct alignment with the stearing vector $\mathbf{e}_{\mathbf{p}}$. We shall remark that the proposed solution is not the optimum DPE solution but a suboptimal scheme that aims to extend the reduced complexity super-resolution ranging algorithm based on the pseudo-periodogram to the higher dimensional problem associated to the Euclidean space.

The proposed DPE position estimate maximizes the following cost function

$$\hat{\mathbf{p}} = \arg\max_{\mathbf{p} \in \mathbb{R}^2} \mathbf{e}^H_{\mathbf{p}} \mathbf{R} \mathbf{e}_{\mathbf{p}} \tag{16}$$

where $\mathbf{R} = \mathbf{Y}\mathbf{Y}^H$.

The joint periodogram (cost function) is a non-convex function, which can be solved in a greedy manner by grid search over the two dimensional space.

V. Numerical results

The algorithms have been analyzed by means of Monte-Carlo simulations with the aim of evaluating the potential gains of DPE over standard two-step approach in IR-UWB. The propagation channel for IR-UWB modulation is characterized to exhibit dense multipath components, specially in indoor scenarios [12, 13]. We consider the channel models developed within the framework of the IEEE 802.15.4a [13], focusing on the scenarios under LOS condition.

We present preliminary results for a simplified signal model where time-hopping and pulse repetition is not explicitly considered since the signal strength is a simulation parameter controlled by means of the SNR. The positioning algorithm is evaluated in a 2D setting with the target placed within a square room of $6 \times 6\ m^2$, with anchor nodes placed at the positions specified in Table I.

Channel realizations are generated according to the standard IEEE 802.15.4a channel models as specified in [13] Matlab implementation and accordingly filtered to match the pulse bandwidth. We consider 100 channel realizations for each scenario. The channels associated to each anchor node are selected randomly out of the 100 realizations. The transmitted pulse is the Gaussian monocycle with pulse duration $T_p = 1$ns. It is worth mentioning that the individual impulse response realizations of the LOS scenarios (CM3 and CM7 in this case) often exhibit stronger multipath components than the first delayed signal replica (LOS component). The specific scenarios are detailed in Table I which include different number of anchor configurations. The table includes the average SNR at each anchor node.

The position error is depicted in Fig. 1-2 in terms of the cumulative density function (CDF) for Industrial and Office

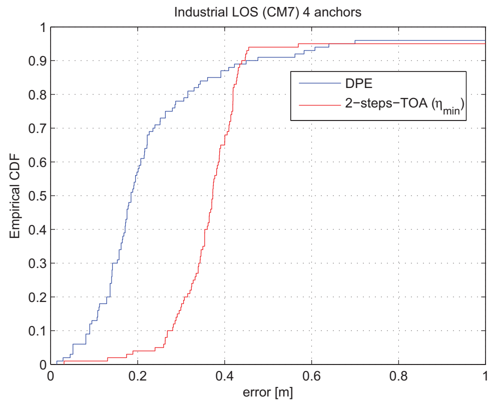
TABLE I
SCENARIO DESCRIPTION

Anchor positions	Average SNR		
Scenario	I.A Industrial LOS	II.A Office LOS	III.A Office LOS/NLOS
N1: (0,0) N2: (0,6) N3: (6,6) N4: (6,0) target: (4,2)	3.51 dB (CM7) 3.59 dB (CM7) 3.63 dB (CM7) 3.65 dB (CM7)	3.93 dB (CM3) 3.92 dB (CM3) 3.98 dB (CM3) 3.94 dB (CM3)	3.96 dB (CM3) 3.98 dB (CM3) 4.03 dB (CM3) 4.12 dB (CM4)
Scenario	I.B Industrial LOS	I.C Industrial LOS	II.B Office LOS
N1: (0,0) N2: (0,6) N3: (6,6) N4: (6,0) N5: (3,3) N6: (3,0) N7: (0,3) N8: (3,6) N9: (6,3) target: (4,2)	3.63 dB (CM7) 3.59 dB (CM7) 3.77 dB (CM7) 3.79 dB (CM7) 3.64 dB (CM7) - - - -	3.57 dB (CM7) 3.62 dB (CM7) 3.74 dB (CM7) 3.63 dB (CM7) 3.60 dB (CM7) 3.80 dB (CM7) 3.77 dB (CM7) 3.65 dB (CM7) 3.85 dB (CM7)	3.91 dB (CM3) 3.99 dB (CM3) 3.95 dB (CM3) 3.97 dB (CM3) 4.04 dB (CM3) 4.07 dB (CM3) 3.99 dB (CM3) 3.98 dB (CM3) 3.94 dB (CM3)

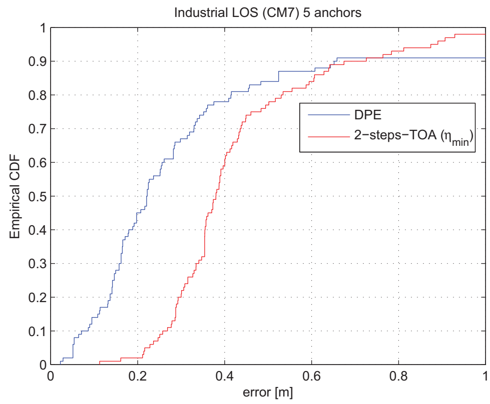
scenarios, respectively. For the two-step approach, the error is depicted for the minimum error obtained with the different threshold values. That is, for each position estimation realization each anchor performs the TOA estimation for threshold values $\gamma = \eta P_{max}$, with $\eta \in 0.1, 0.2, \ldots, 0.9$. For each set of TOA measurements associated to a given threshold (the same threshold is considered for all anchors), we solve the positioning problem and select the one with minimum error. Therefore, we compare DPE with the best two-step solution.

Similar results are obtained for other SNRs (from -1dB to 20 dB). In fact, the SNR does not appear to be the most critical parameter. As mentioned before, the "Industrial LOS" or "Office LOS" scenarios are not classical LOS scenarios where the LOS is present with stronger amplitude than other multipath component like in GNSS positioning. In fact, there are channel realizations where delayed multipath are considerably larger than the LOS. It is in these scenarios where the DPE algorithm experiences larger difficulties to accurate estimate the target position. This fact, can be observed from the graphical representation of the cost function. For instance, Fig. 4 depicts the cost function (16) for a channel realization for the "Industrial LOS" scenario where the multipath effect is more acute. One can appreciate several local maxima of similar amplitude. One way of reducing the ambiguity caused by strong multipath components is to increase the number of nodes, as shown in 4(b) where a clear maximum is now present at the target position. To illustrate the performance difference as a function of the SNR, the cost function is shown in Fig. 5 for a "Office LOS" scenario with 4 anchors at two different SNR values.

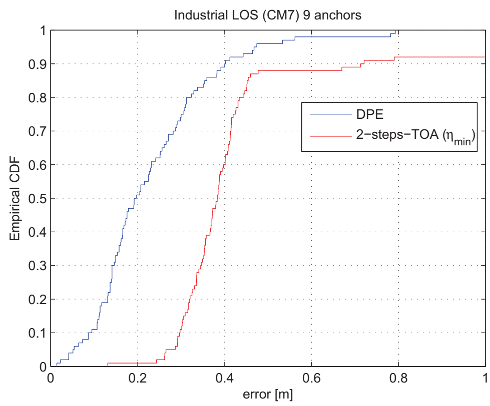
A common feature in the analyzed scenarios is the fact that in a large percentage (about 90%) the DPE algorithm is able to position the target with less error than the two-step approach. However, the estimator exhibits a strong bias in the top region of the CDF, which is mostly explained by the presence of strong multipath components. The preliminary results shown in Fig. V for the scenario III.A when one reference node is in NLOS (CM4), indicates that the potential gains of DPE may also be achieved in more general set-ups.



a) Positioning with 4 anchor nodes



b) Positioning with 5 anchor nodes

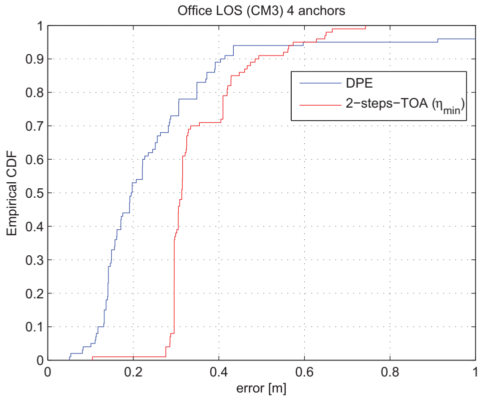


c) Positioning with 9 anchor nodes

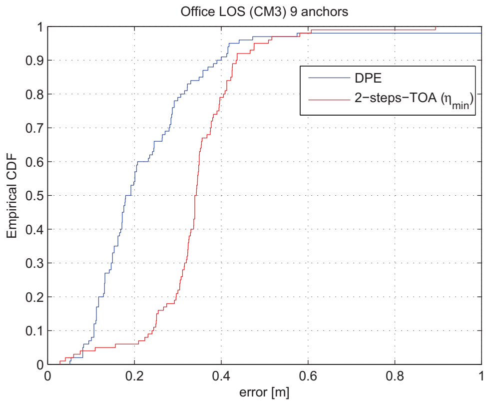
Fig. 1. Empirical CDF of the position error for DPE and Two-step TOA-based positioning. Evaluated in IEEE 802.15.4a Industrial LOS (CM7).

VI. Conclusion

A Direct Position Estimation (DPE) for IR-UWB localization based on frequency domain signal has been introduced and assessed under realistic channel models developed

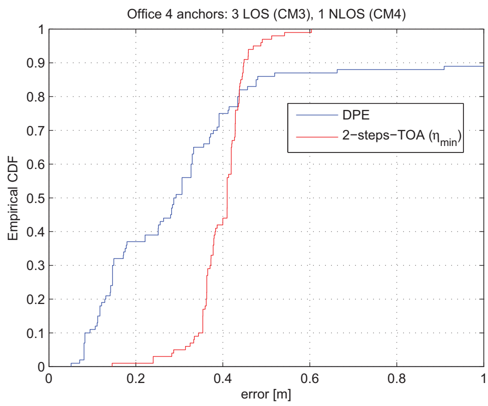


a) Positioning with 4 anchor nodes



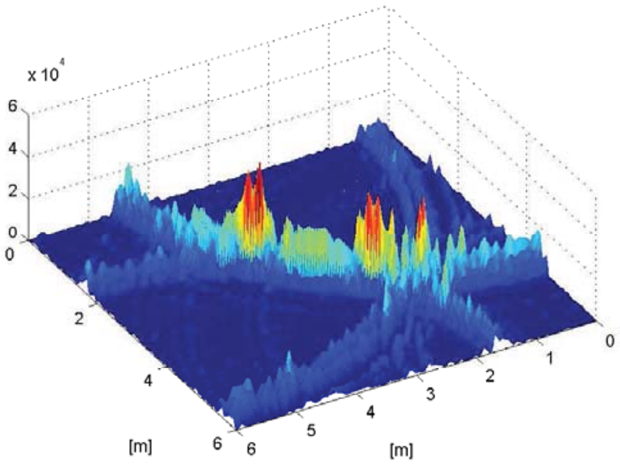
b) Positioning with 9 anchor nodes

Fig. 2. Empirical CDF of the position error for DPE and Two-step TOA-based positioning. Evaluated in IEEE 802.15.4a Office LOS (CM3).

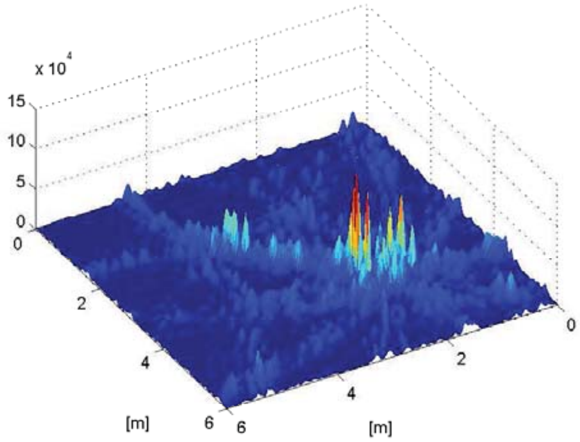


a) Positioning with 4 anchor nodes

Fig. 3. Empirical CDF of the position error for DPE and Two-step TOA-based positioning. Evaluated in IEEE 802.15.4a Office (CM3,CM4).



a) 4 anchor nodes



b) 9 anchor nodes

Fig. 4. Periodogram realisation for DPE. Evaluated in IEEE 802.15.4a Industrial LOS at SNR = 4.5 dB.

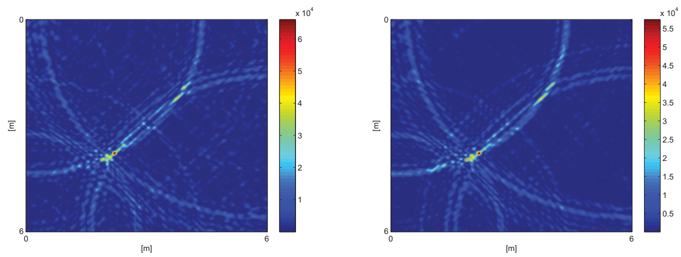


Fig. 5. Periodogram realisation for DPE with 4 anchors. Evaluated in IEEE 802.15.4a Office LOS, SNR = -1.3 dB (left) and 3.6 dB (right).

by the IEEE 802.15.4a standardization group. The proposed DPE algorithm is based on a generalization of the pseudo-periodogram approach proposed for TOA estimation. This scheme is of lower complexity than the MUSIC algorithm proposed in [9] and does not require the estimation of the channel coefficients. One advantage of DPE versus two-step positioning is that DPE does not require any calibration, in the sense that two-step scheme requires the proper setting of the threshold value. Which in practice varies with the propagation conditions. Although the numerical results show improvements with respect the two-step approach, further work on multipath effect mitigation shall be considered in the future. As well as more detailed comparison with two-step approaches considering threshold calibration.

ACKNOWLEDGEMENTS

This work has been partially supported by the Catalan government under 2009SGR1046, 2009SGR891, the Spanish Ministry of Economy and Competitiveness project TEC2012-39143(SOSRAD), TEC2011-29006-C03-02 (GRE3N-LINK-MAC), the European Commission in the framework of the FP7 Network of Excellence in Wireless COMmunications NEWCOM♯ (Grant agreement no. 318306) and European Cooperation in Science and Technology under project COST Action IC0902.

REFERENCES

[1] M. Navarro and M. Nájar, “Frequency domain joint TOA and DOA estimation in IR-UWB,” *IEEE Trans. Wireless Commun.*, vol. 10, no. 10, pp. 3174–3184, Oct. 2011.

[2] S. Gezici, Z. Tian, G. Giannakis, H. Kobayashi, A. Molisch, H. Poor, and Z. Sahinoglu, “Localization via ultra-wideband radios: a look at positioning aspects for future sensor networks,” *IEEE Signal Processing Magazine*, vol. 22, no. 4, pp. 70–84, July 2005.

[3] P. Closas, C. Fernández-Prades, and J. A. Fernández-Rubio, “Maximum Likelihood Estimation of Position in GNSS,” *IEEE Signal Processing Lett.*, vol. 14, no. 5, pp. 359–362, May 2007.

[4] ——, “Cramér-Rao Bound Analysis of Positioning Approaches in GNSS Receivers,” *IEEE Trans. on Signal Processing*, vol. 57, no. 10, pp. 3775–3786, October 2009.

[5] A. Amar and A. J. Weiss, “New Asymptotic Results on two Fundamental Approaches to Mobile Terminal Location,” in *Proc. of the 3rd International Symposium on Communications, Control and Signal Processing (ISCCSP)*, March 2008, pp. 1320–1323.

[6] A. J. Weiss, “Direct position determination of narrowband radio frequency transmitters,” *IEEE Signal Process. Lett.*, vol. 11, no. 5, pp. 513 – 516, May 2004.

[7] ——, “Direct geolocation of wideband emitters based on delay and doppler,” *IEEE Trans. Signal Process.*, vol. 59, no. 6, pp. 2513 – 2521, June 2011.

[8] M. Erić and D. Vučic, “Direct position estimation of UWB transmitters in multipath conditions,” in *IEEE International Conference on Ultra-Wideband, 2008. ICUWB 2008.*, vol. 1, 2008, pp. 241–244.

[9] M. Erić, M. Dukić, and D. Vučic, “Method for direct self-localization of IR UWB node(s) in indoor scenario,” in *IEEE International Conference on Ultra-Wideband, 2011. ICUWB 2011.*, 2011.

[10] Y. Shen and M. Z. Win, “Fundamental limits of wideband localization Part I: A general framework,” *IEEE Trans. Inf. Theory*, vol. 56, no. 10, pp. 4956–4980, Oct. 2010.

[11] S. M. Kay, *Fundamentals of Statistical Signal Processing. Estimation Theory*. Prentice Hall, 1993.

[12] A. F. Molisch, “Ultrawideband propagation channels- theory, measurement, and modeling,” *IEEE Trans. Veh. Technol.*, vol. 54, no. 5, pp. 1528 – 1545, Sept. 2005.

[13] A. F. Molisch, K. Balakrishnan, C. Chong, S. Emami, A. Fort, J. Karedal, J. Kunisch, H. Schantz, U. Schuster, and K. Siwiak, “P802.15-04-0662-00-004a, ieee 802.15.4a channel model - final report,” 2004.