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2D indoor localization resolution improvement using IR-UWB stepped frequencies signals

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Abstract—This article discusses the ability to improve the resolution of range and azimuth angle measurements by an interferometric localization system used to locate beacons within its line of sight (LoS) by exploiting stepped frequency signals Impulse Radio - Ultra WideBand (IR-UWB) centered on 2 frequencies. This improvement does not require increasing the frequency bandwidth used. This solution uses a phase correlation (PC) method, usually applied to continuous wave (CW) signals, adapted to Ultra WideBand (UWB) pulse signals. The results obtained are compared with those computed by a classical method of energy detection (ED) by exploiting the same frequency band.

Index Terms— Interferometric processing, Localization resolution Improvement, Multistatic IR-UWB radar, Short range localization system

I. INTRODUCTION

Interest in indoor localization is high due to the range of services it can offer in many areas and applications. Global Navigation Satellite Systems (GNSS) are not accurate enough for indoor services and sometimes even not available at all. Complementary systems have been developed to ensure reliability and precision, even in harsh conditions (Multipaths, shadowing, etc.). A solution is to use the propagation time of the IR-UWB signal by exploiting wide frequency bandwidths for better accuracy. Firstly, it might be difficult to use by respecting ITU's (International Telecommunication Union) standards and secondly the accuracy is degraded by the pulse distortions mainly induced by the group speed of the components used in the systems [1] [2] [3]. To reduce the effect of these limitations, several solutions have been proposed but the improvement remains limited. In [4], a two-step ToA estimation based on the Time Delayed Sampling and Correlation method for Transmitted-Reference UWB (TR-UWB) signals is presented. Another solution is based on energy sensing (ED), where energy is collected after some analog front-end processing. In [5], the performances of three approaches based on ToA estimation techniques are analyzed for IR-UWB systems using sub-Nyquist sampling rates. In [6], various ToA estimation algorithms for UWB systems are discussed. These algorithms use a low sampling rate and three variants of the ED method, namely Maximum Energy Selection (MES), Threshold Comparison (TC) and Maximum Energy Selection with Search Back (MES-SB).

These algorithms do not require precise synchronization clock or pulse shapes, but rather require a strong signal at the detector input. This requirement is crucial because the estimation of ToA is considered as a problem of detecting the rising edge of the signal, which can be erroneous when the limit of the SNR (upper than 10 dB) is no longer respected. Therefore, while it is theoretically possible to achieve very accurate ranging, the practical limitations of the above algorithms prevent the tracking system from achieving the desired high resolution. Another constraint imposes a trade-off between required accuracy, resolution and complexity that must be taken into

account when developing a ToA estimation algorithm.

We propose, in this article, an alternative solution of low complexity based on the phase correlation (PC) method fitted for IR-UWB signals, improving the precision and the resolution for the localization of an active tag (AT) by its measurement of distance and azimuth angle relative to an interferometric tracking system. The main advantage of an interferometric method is that the position of AT in a free space room can be obtained using a single fixed anchor instead of uncovering a constellation, thus not requiring the employment of a synchronization clock.

This article is presented as follow. In section II, we present the developed 2D localization system based on a multistatic IR-UWB radar system using signals centered on 2 frequencies for the AT localization. In section II.A, the system and the UWB pulse signal used are presented. In Section II.B, the measurement setup and associated signal processing for the PC method are explained. In section II.C, we present the localization results of AT by angle and distance measurement, obtained by the PC method which are compared with the results obtained by using an ED algorithm. Section III presents the conclusions.

II. 2D IR-UWB LOCALIZATION SYSTEM

A. System principle and Setup

The 2D Localization principle consists to get the relative position of an Active Tag (AT) by estimating the range d and the azimuth angle α from the baseline formed by the 3 antennas A_0 (transmitting antenna) and A_1 & A_2 (two receiving antennas). The 2D localization system, presented in Fig.1, consists of two main parts: the Localization Base Station (LBS) and an Active Tag (AT) to localize.

At the LBS side, the transmitter chain is made by an impulse generator (consisting of a step recovery diode SRD and a short-circuited stub line [7]) delivering a pulse covering a bandwidth of 4 GHz (0-4 GHz).

In order to optimize the useful energy of the pulse in the bandwidth of the transmitter antenna A_0 (3-4 GHz), the UWB pulse is placed in

the 3-4 GHz frequency band via a mixer modulating the pulse by a CW local oscillator LO signal at f_1 (3.5 GHz) or f_2 (3.55 GHz) successively.

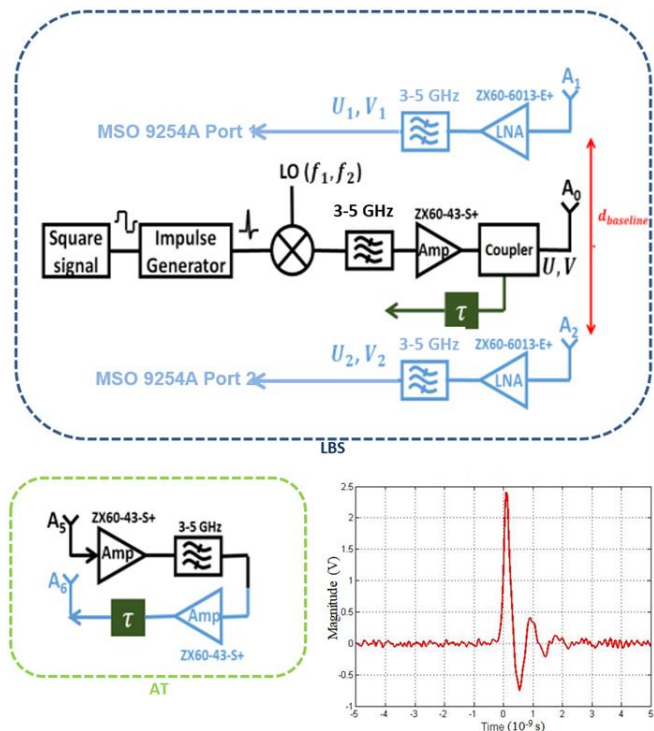


Fig. 1. 2D localization base station (LBS), the active tag (AT) to localize and the UWB pulse at Impulse Generator output.

Next, the output signals (U and V) at the mixer are filtered by a band-pass filter (3-5 GHz) and amplified (Mini-Circuits- ZX60-43-S+). The carrier frequency of U is f_1 and f_2 for V respectively. A 10 dB directional coupler allows to take the delayed (by τ) and required signals for the synchronization at the acquisition system which is an Agilent Infinium MSO 9254A scope (Sampling frequency set at $F_s = 20$ GHz). Next, the signals U and V are radiated, to the AT, by the A0 antenna [7].

The two RF receiver chains, connected to antennas A_1 and A_2 , are identical and provide the amplified (LNA Mini-Circuits- ZX60-6013-E+) and filtered signals coming from AT, in order to deliver the couples U_1, V_1 and U_2, V_2 signals where f_1 and f_2 are the carrier frequencies of U_i and V_i ($i=1..2$) signals respectively. The three antennas are aligned and A_0 is placed in the center of the baseline $d_{baseline} = 30$ cm formed by the antennas A_1 and A_2 .

At the AT side, there are a receiving antenna A_5 and a transmitting antenna A_6 connected via two low noise amplifiers (Mini-Circuits- ZX60-43-S+) and a delay line τ between them. The known delay τ (in AT and LBS) is used mainly to segregate the useful U_i , and V_i ($i=1..2$) signals from the naturally backscattered signals of the indoor environment, taking advantage of the different arrival times between them. Here τ is made by delay lines and fixed depending on the desired maximal range (here $\tau = 30$ ns, insuring a high added delay compared to the time of flight of signal generated by a multipath channel for a 1 to 2 meter range d).

B. Measurement setup & Signal processing

The measurement setup allows to estimate the azimuth angle α and

the range d and for different positions of AT ($p_0 \dots p_{30}$) and ($p_{31} \dots p_{81}$) respectively (Fig.2). The maximum range have been fixed at 2m due to the sensitivity of the MSO9254A used for the proof of concept. To extend this, a dedicated circuit should be developed.

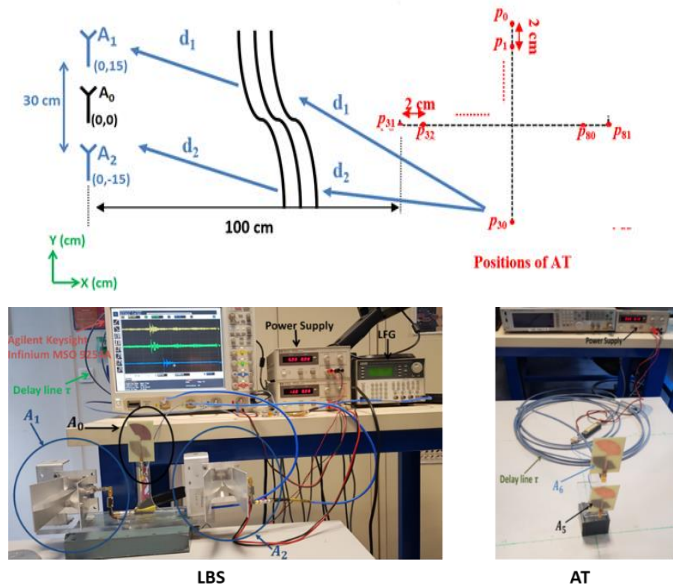


Fig. 2. Measurement setup and Proof of concept Bench

Assuming that the range d is high enough by $d_{baseline}$ (30 cm here) to insure the parallel conditions for the paths from AT to A_1 and A_2 respectively [8], the average value d and the angle α are given by:

$$d = \frac{d_1 + d_2}{2} \tag{1}$$

$$\alpha = \frac{180}{\pi} \arccos\left(\frac{d_1 - d_2}{d_{baseline}}\right) \tag{2}$$

[9] have proved that the deviation angle due to the parallel conditions is equal to 2° for a radial distance d of 1.5 m and a $d_{baseline} = 30$ cm. The signal processing is mainly based to estimate $d_1 = \frac{cT_1}{2}$ and $d_2 = \frac{cT_2}{2}$ (T_i : time-of-flight between A_0 and A_i ($i=1,2$)) by the stepped frequencies radar system using the two LO f_1 & f_2 frequencies and is described below.

Two UWB pulses U and V are transmitted, through the front end antenna A_0 , one by one, by taking into account the time acquisition of signals U_i ($i=1,2$). We assume that the channel remains constant during the transmission of the two pulses (f_1 and f_2 are very close and $d_{baseline} \ll d$). For both transmitted signals, the received signals U_i & V_i reach A_i with the delay T_i estimated by applying the UWB stepped frequencies method [7]. The delay T_i is calculated by estimating the time difference between the emitted signals U or V used at the receiver for the synchronization (explained in section A).

In this way, the added delay τ (Fig.1) allows to let the intended signal independent by the coupling effect between A_0 and A_i and the natural backscattering effect of the environment. By interesting of the intended signals, a better resolution is compulsory to increase the accuracy on the AT position.

With a frequency bandwidth (BW) of 1GHz (i.e. 3-4 GHz here), it is impossible to separate the signals U_i and V_i if the path difference between A_0 and A_1 or A_2 is lower than 15 cm. To improve it, we have developed a method using jointly the pulse signal distortion and a signal processing based on a phase correlation [10] between the

signals U_i or V_i .

In Fig.3, we show an example of the two received signals U_1 and U_2 . These are distorted in the same way by the LBS and AT components and the antennas (assuming that components are the same and the received antennas A_1 and A_2 are closed compared the range involving that channels $A_0 \Rightarrow A_1$ and $A_0 \Rightarrow A_2$ are the same). To describe these pulses, the signals can be assimilated to a sine carrier at f_1 modulated by a Gaussian pulse [10] as:

$$U_1(t) = A \cos(2\pi f_1(t - T_1)) e^{-\left(\frac{t-T_1}{T'}\right)^2} \quad (3)$$

$$U_2(t) = A \cos(2\pi f_1(t - T_2)) e^{-\left(\frac{t-T_2}{T'}\right)^2}$$

where A is the signals magnitude and T' is the distorted pulse duration (Fig.3). The delay between $U_1(t)$ and $U_2(t)$ is defined as $\Delta t = T_2 - T_1$ and shown in the Fig.7.

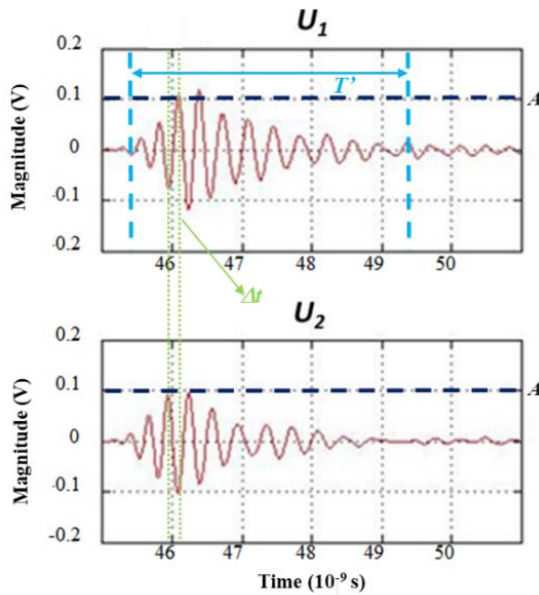


Fig. 3. Δt and T' estimation

As presented in [10], the terms $e^{-\left(\frac{t-T_1}{T'}\right)^2} \approx e^{-\left(\frac{t-T_2}{T'}\right)^2} \approx 1$ and the computation of the cross correlation:

$$C_{U_1 U_2}(\Delta t) \approx \frac{A^2}{2} \cos(2\pi f_1 \Delta t) \quad (4)$$

which is considered as the Phase Correlation (PC) method directly related to the azimuth α through $\Delta t = T_2 - T_1$ according to Equation (2). The magnitude A is determined by the recorded maximal magnitude of $U_1(t)$ or $U_2(t)$ chronograms. Δt is obtained as: $\Delta t = \frac{1}{2\pi f_1} \arccos\left(\frac{2 C_{U_1 U_2}(\Delta t)}{A^2}\right)$. So the azimuth is calculated as:

$$\alpha = \frac{180}{\pi} \arccos\left(\frac{c \Delta t}{d_{baseline}}\right) \quad (5)$$

Next by setting the LO frequency at f_2 after a fixed and known delay t_s (from the initial time of the transmitted pulse) upper than the delay to record the signals $U_1(t)$ and $U_2(t)$, the signal received at A_1 and A_2 are expressed here as:

$$V_1(t) = A \cos(2\pi f_2(t - T_1)) e^{-\left(\frac{t-T_1}{T'}\right)^2}$$

$$V_2(t) = A \cos(2\pi f_2(t - T_2)) e^{-\left(\frac{t-T_2}{T'}\right)^2}$$

Assuming than f_1 (3.5 GHz) and f_2 (3.55 GHz) are closed enough compared to the frequency behavior of the propagation channel and the distortions generated by the system components, the magnitude A and the distorted pulse duration T' are the same for $U_{1,2}(t)$ and $V_{1,2}(t)$. After recording, a data shift (by t_s) is compulsory to process the cross correlation $C_{U_1 V_1}(T_1)$. After a digital low pass filtering to cancel high frequency components ($f_2 + f_1$), $C_{U_1 V_1}(T_1)$ is given by:

$$C_{U_1 V_1}(T_1) \approx \frac{A^2}{2} \cos(2\pi(f_2 - f_1)T_1)$$

So T_1 is calculated as:

$$T_1 = \frac{1}{2\pi(f_2 - f_1)} \arccos\left(\frac{2 C_{U_1 V_1}(T_1)}{A^2}\right)$$

According the Equation (1), T_2 is determined from $C_{U_2 V_2}(T_2)$ and the range d is obtain from the mean of (T_1, T_2) to improve the accuracy as:

$$d = \frac{c(T_1 + T_2)}{2}$$

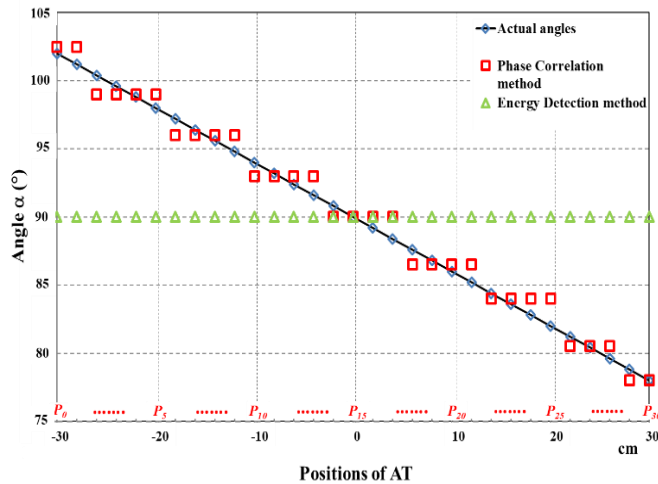
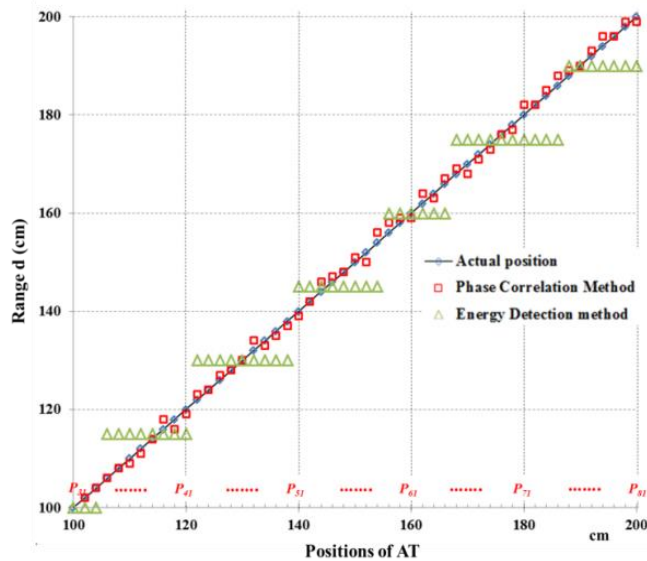
The range measurement is also limited by the frequency difference $f_2 - f_1$, here equal to 50 MHz. In that case, the maximum range, determined without ambiguity as in step CW radar systems, is equal to $d_{max} = \frac{c}{f_2 - f_1} = 3$ m.

C. Measurement results

In a first part, angle measurement is discussed through the positions of AT from P_0 to P_{30} (Fig. 2) on an axis which is parallel and placed at 1.5m from the receiver baseline. The step of the positions is 2 cm. The center of the AT displacement is fixed at P_{15} . So the range of displacement is 60 cm [-30 cm ; 30 cm].

In Fig.4, blue diamond points are the actual angle direction of the AT position relatively to P_i ($i = 0..30$). The red square points are the measured angle using the Phase Correlation (PC) algorithm [11]. The green triangle points depict the measured angle using a classical Energy Detection (ED) algorithm [6], used for UWB signals. Due to the closed path difference ($d_2 - d_1$), it becomes impossible to estimate it by classical time methods due to the too low frequency bandwidth (resolution for 1GHz = 15cm so it becomes impossible to estimate the path difference $d_1 - d_2$). For PC algorithm, the presented results show that the angular measurement error is stepped (with the sampling frequency $F_s = 20$ GHz [11]) and lower than $\pm 2^\circ$. A higher frequency sampling provides a linear evolution versus the positions of AT with a lower error [10].

In Fig.5, blue diamond points are the actual range of the AT position relatively to P_i ($i = 31..81$). The red square points are the measured range using the PC algorithm. The green triangle points depict the measured range using a classical ED algorithm, used for UWB signals. Due to the used Frequency bandwidth (1 GHz), the range is stepped by 15 cm as a classical pulse radar measurement. It is not a continuous measurement as the range given by the PC algorithm; the presented results show that the range measurement error is lower than ± 2 cm.

Fig. 4. Measured Angle α ($^{\circ}$) Vs AT positions.Fig. 5. Measured Range d Vs AT positions

III. CONCLUSION

In this paper, we have presented a proof of concept of a 2D UWB Localization system allowing estimating the position of an active tag, in polar coordinates, by a localization base station. The used UWB pulse signal, covering a frequency bandwidth of 1 GHz) modulates a LO sine signal whose carrier frequency can be switched over 2 frequencies (3.5 GHz and 3.55 GHz). The frequency shift of the pulse allows, on the one hand, to optimize the power transfer in the antennas matching frequency bandwidth and, on the other and, to use an algorithm based on an interferometric method. Originally, this algorithm is matched for CW sine signals. It becomes possible to exploit it for the distorted pulses due to the components and antennas effects. In this case, the distorted signals look like sine signals, whose each carrier frequency is one of the LO frequencies $f_1 = 3.5$ GHz and $f_2 = 3.55$ GHz, over the pulse time spreading T' . Therefore, it processes the phase correlation estimation of UWB pulse signal for improving the measurement resolution in terms of range and azimuth angle. This method allows to obtain an improve resolution according to an error of ± 2 cm for the range and $\pm 2^{\circ}$ for angles measurement for a radial distance inside [1-2 m].

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