

A Novel Accurate Synthetic Aperture RFID Localization Method with High Radial Accuracy

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Abstract—Internet of Things (IoT) is rather prevalent in many manufacturing and smart city applications, while localization is a premise for many other processes, varying from ordering objects in manufacturing lines to locating books on bookshelves. Radio Frequency Identification (RFID) based localization is of great interest in many IoT applications. Synthetic aperture RFID, due to its anti-noise capability and robustness against multipath distortion, is becoming a rising star in the field of localization. Existing systems achieve finer lateral resolution, whereas their radial accuracy is limited by the narrow bandwidth of RFID signal. In this paper, we present a novel synthetic aperture RFID localization method which combines RFID phase based ranging with synthetic aperture technology, to achieve a higher radial accuracy than the existing systems. With only one reader antenna and one 1-dimensional (1D) trajectory, a synthetic array is constructed to get an accurate localization result both in lateral and radial direction. Its core idea is to make full use of the coherence of all multi-frequency phase data and merge them into a unique ranging based likelihood function. To improve the accuracy, the relative phase is leveraged to eliminate phase offsets caused by the reader antenna, and the phase deviation from the angle-of-arrival response is calibrated by pre-processing. Then a weighted enhancement is fully exploited to further improve the localization performance. We evaluate its performance with commercial-off-the-shelf (COTS) RFID devices and the results show that it achieves median accuracy of $3cm$ in both lateral and radial direction. This novel promising method is suitable for locating tags placed densely in many IoT applications, such as test tubes in hospitals.

I. INTRODUCTION

Wireless location sensing plays a critical role in developing IoT, which has been attracting increasing attention in both the academia and industry in recent years. At the same time, due to its low cost, long range and maintenance-free, RFID is widely used in many fields, such as libraries, postal service, logistic delivery, food supply chain, etc. RFID offers good infrastructure to construct localization system with tags attached to objects.

At first, based on Received Signal Strength Indicator (RSSI), radio propagation model [1], radio fingerprint [2] and device-free technology [3] are utilized for localization. Nevertheless, these state-of-the-art systems cannot achieve the cm -level accuracy due to noise-sensitive RSSI, especially in the complicated indoor propagation environment.

Their accuracy is suitable for some applications but more accurate localization is needed in other scenarios. For example, in the medical field, lots of analyses are made in test tubes, which are densely placed in the tube holder, and their management is usually conducted manually. The importance of positive patient identification (PPI) in reducing medical error is very clear to prevent some serious consequences [4]. At the same time, RFID plays a more and more indispensable role in medical equipment tracking and management [5]. RFID tags are attached to test tubes as information carriers. Once the localization is accurate enough, these tags can be leveraged for test tube localization to improve PPI.

Fortunately, another low level data captured by COTS UHF RFID readers, phase information, brings opportunities to construct more accurate localization systems. Taking Impinj R420 reader for example, theoretically, it can provide the phase resolution of $0.0015rad$, in other words $0.038mm$ (the wavelength is about $32cm$) [6]. At first, Phase Difference of Arrival (PDOA) is utilized for ranging [7] and localization [8]. To achieve a higher accuracy, special devices are necessary [9]–[11].

In complicated propagation environment, noise and multipath distortions change in space [12]. Synthetic Aperture Radar (SAR) [13] which leverages relative motion for accurately imaging has come to the attention of researches due to its good noise-insensitive capability and the robustness against multipath distortions. Then different synthetic aperture RFID localization systems have been proposed with the cm -level accuracy [14]–[16].

Existing synthetic aperture RFID localization systems usually utilize the inverse correlation filter and holographic imaging for localization, which have a good lateral accuracy but a normal radial one [14]–[16]. And most of them use mono-frequency signal for localization without considering the frequency-dependent phase offset and the angle-of-arrival (AoA)-dependent phase offset, which affects the accuracy of the measured phase. Most COTS UHF RFID readers utilize frequency hopping spread spectrum or 4 channel plan signaling with a short maximum channel dwell time for each carrier frequency [17], which means actual phase data is multi-frequency. And on account of the relative motion between the reader antenna and the tag, the AoA changes along the relative motion. So the phase offset caused by its frequency and AoA response must be effectively eliminated. As a very efficient

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method to combine all measurement information, synthetic aperture RFID still has the potential to become more reliable and accurate.

In this paper, we utilize the RFID phase based ranging to realize a novel synthetic aperture RFID localization method, which has a finer radial resolution and a high tolerance of the inevitable multipath distortions and interference. The contributions of this paper are listed below.

After introducing the basic idea of synthetic aperture RFID, this paper proposes a ranging based likelihood reconstruction method, which utilizes the one-to-many relationship between the phase value and the distance from the reader antenna to the tag. This likelihood function merges the phase data sampled at different points and different carrier frequencies. And these coherent received signals construct a virtual aperture that is much longer than the physical reader antenna.

The phase offset of the received signal phase caused by frequency and AoA response is eliminated by normalized relative phase and pre-processing method. Considering the measurement noise, a weighted enhancement method is utilized to improve the localization performance.

Unlike realizing both high lateral and radial resolution by 2D trajectory in other methods, our method only needs one reader antenna and one 1D trajectory. With the synthetic array constructed by continuous motion, the accurate localization results both in lateral and radial direction are achieved. And its better radial resolution is proved theoretically.

Extensive experiments show this method achieves an average error of 3cm in both lateral and radial direction. It is suitable for locating tags placed densely in many IoT applications, such as test tubes localization in hospitals. Furthermore, it can be extended to other wireless communication techniques which can also get signal phase, such as Wi-Fi, Bluetooth and so on. They all can benefit from this method.

The rest of this paper is organized as follows. After reviewing related work in Section II, the basic idea of synthetic aperture RFID is introduced in Section III. The novel synthetic aperture RFID localization method is proposed in Section IV. In Section V, its performance evaluations are compared and analyzed. Finally, Section VI concludes this paper and discusses future work.

II. RELATED WORK

Our work is standing on the shoulders of previous wireless localization work and draws inspiration from them.

A. RSSI-based Localization

Some systems (SpotON [1] and EZ [18]), take advantage of the radio propagation model, while other systems like Landmarc [2] utilize RSSI fingerprint map. RSSI changes with the moving target in the region, so device-free systems [3], [19]–[21] leveraging lots of reference tags or sensors to collect these changes are proposed. Unfortunately, it is inconvenient and demanding in many applications. Furthermore, the accuracy of these systems is challenged by the vulnerability of RSSI to multipath distortion and ambient dynamics.

B. CSI-based Localization

Finer-grained Channel state information (CSI) can be captured by some Wi-Fi devices, which contains the signal phase and amplitude of each sub-carrier [22]. On account of the much shorter wavelength than RFID, it is more sensitive to the change of signal propagation distance. Many CSI-based localization systems fuse the AoA approach with additional help, including software radios (Wi-Vi [23], ArrayTrack [24]), natural human mobility (CUPID [12]), multiple devices (SpotFi [25]) and natural rotation behavior of users (D-MUSIC [26]). Fingerprint map based systems, such as FILA [27] and DeepFi [28], harness CSI to construct the radio map. However, for commercial Wi-Fi devices, carrier frequency offset and phase-locked loop phase offset is not negligible due to hardware imperfections, leading to serious phase changes [29], [30]. Furthermore, to get CSI from Wi-Fi devices, their drivers need to be modified, whereas it is not necessary for RFID readers. In contrast, COTS RFID readers can precisely and conveniently obtain the phase.

C. Phase-based RFID Localization

Phase difference of arrival in time, frequency and spatial domain are all used for localization [7], [8]. At first, special devices like software defined radio [9], rotating antenna [10], near field cascaded shelf antennas [10] or multiple antennas [31] are applied for the phase difference measurement. Phase data sampled at different carrier frequencies is utilized for ranging [32], which needs to calibrate all tags and antennas. Tadar [33] leverages the changes of the phase values, but it needs a learning stage and can only track one object.

In 2007, a synthetic aperture secondary radar [34] is proposed for localization. Moreover, it is worth mentioning that RFID can be regarded as a special secondary radar. Then there are more and more synthetic aperture RFID localization systems. A stationary tag is located by a mobile reader antenna with a known trajectory [35], while a mobile tag is located by a static reader in 1D and 2D scenario [36], [37]. This method is expanded to a 3D scenario with a 2D trajectory [14]. PinIt [38] captures the multipath profiles by synthetic aperture RFID with a reference tag array as the position fingerprint. Tagoram [15] estimates the tag position with the cm -level accuracy, which relies on multiple reader antennas. MobiTagbot [16] utilizes multi-frequency to exclude aperture points and achieves high accuracy in lateral direction. All these approaches utilize the inverse correlation filter and holographic imaging for localization. At the same time, the AoA-dependent phase offset is not considered. Furthermore, COTS UHF RFID reader is a multi-frequency system, whereas most of these systems are designed for mono-frequency scenario.

III. BASIC IDEA OF SYNTHETIC APERTURE RFID

The communication between an UHF RFID reader and a tag depends on the backscatter modulation. After IQ coherent demodulation, the reader gets the received signal phase φ . If

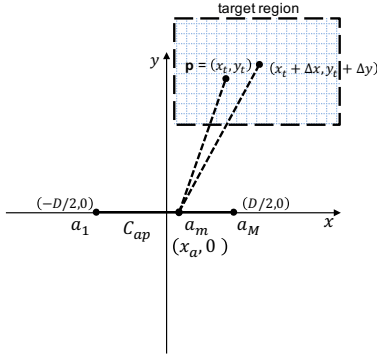


Fig. 1. Geometric condition of synthetic aperture RFID.

the carrier frequency is f , the theoretical received signal phase is [8]:

$$\varphi = (\varphi_{prop} + \varphi_r + \varphi_t) = \left(\frac{4\pi f d}{c} + \varphi_{rt}\right) \mod 2\pi \quad (1)$$

where φ_{prop} is the phase offset caused by the wave propagation in space, while φ_{rt} denotes the phase accumulated by the reader antenna φ_r and tag reflection φ_t . d is the distance between the reader antenna and the tag. For a fix pair of the reader antenna and the tag, φ_t can be seen as a constant, whereas φ_r changes with the carrier frequency. The relationship between φ and d is the base of synthetic aperture RFID. Unfortunately, the RFID reader only gets the result of the modulo operation on φ . Thus φ ranges from 0 to 2π , and it cannot be utilized for distance estimation directly.

The geometric condition of a typical synthetic aperture RFID localization is depicted in Fig. 1. The reader antenna moves along a trajectory C_{ap} , known as the synthetic aperture, which is utilized to construct a synthetic array. It interrogates the tag at different aperture points $\mathbf{a}_m$ with different carrier frequencies. C_{ap} in Fig. 1 is a line segment whose middle point is set as the coordinate origin, and x-direction is the direction of motion.

The actual tag position (x_t, y_t) is unknown, but the scope of its possible positions can be determined by the working range of the antenna, which is called target region. The whole target region is discretized into many grids, and each grid represents a hypothetical tag position $\mathbf{p}$. For each $\mathbf{p}$, if the reader antenna is at $\mathbf{a}_m$, the theoretical received signal phase from $\mathbf{p}$ can be estimated by Equation (1) with the distance between the reader antenna and the tag $d(\mathbf{a}_m, \mathbf{p})$. This phase is supposed to be equal to the actual received one.

Synthetic aperture RFID utilizes a likelihood function $\chi(\mathbf{p})$ to represent the likelihood that $\mathbf{p}$ matches actual position. This is a coherent reconstruction method, which leverages all received signals to reconstruct $\chi(\mathbf{p})$ for all hypothetical positions. At a point near the actual position, $\chi(\mathbf{p})$ should grow as the hypothetical signals are superimposed constructively. And received signals should be summed up incoherently from false positions, which will cause destructive interference and result in a small $\chi(\mathbf{p})$. The point $\mathbf{p}$ with the maximum $\chi(\mathbf{p})$ is considered as the actual position. Since both the multipath

and noise components are varying from one aperture point to another, they will be weakened as well.

$\chi(\mathbf{p})$ can also be used for imaging, and its value yields an image amplitude for each grid, which is known as the synthetic aperture RFID imaging. Previous methods [14]–[16] use inverse correlation filter to deal with actual received phase with the theoretical one to realize the likelihood function and achieve holograms, so these previous methods are known as the hologram reconstruction. However, their radial resolutions are worse than the lateral ones usually.

Moreover, simple inertial sensors or photoelectric encoders are applied to determine the positions of aperture points. The common error caused by these sensors is much smaller than the aperture length. So the key point for localization is to reconstruct the likelihood function $\chi(\mathbf{p})$. The novel likelihood function for synthetic aperture RFID with a much better radial resolution will be described in the next section.

IV. LOCALIZATION

In this section, a novel accurate synthetic aperture RFID localization method is introduced, which only needs one reader antenna and one 1D trajectory.

A. Normal Ranging Based Reconstruction

According to the relationship shown in Equation (1), for a measured phase φ , the probability that the distance is d can be set as Equation (2):

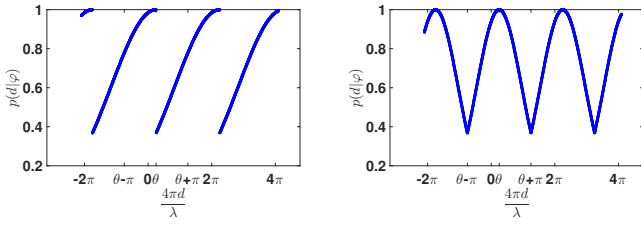
$$p(d|\varphi) = e^{\alpha_{normal}} = e^{-\frac{((\varphi - \varphi_{rt} - \frac{4\pi d}{\lambda}) \mod 2\pi)^2}{(2\pi)^2}} \quad (2)$$

where φ_{rt} denotes the phase offset caused by the reader antenna and the tag as mentioned before, and $4\pi d/\lambda$ is phase offset φ_{prop} in space. The modulo operation on the numerator ensures $-1 < \alpha_{normal} \leq 0$ due to the modulo operation on φ by the reader. Apparently, for a given φ , $p(d|\varphi)$ is a periodic function of d with the cycle $\lambda/2$, and φ cannot be unwrapped without ambiguity, so it is impossible to estimate the actual distance directly.

Along the synthetic aperture, there are M aperture points with the received signal phase φ_m . If the tag is at the hypothetical position $\mathbf{p}$, it has the propagation distance $d(\mathbf{a}_m, \mathbf{p})$ when the reader antenna is at the aperture point $\mathbf{a}_m$. Although $p(d|\varphi)$ is periodic, there will be one point that maximizes all $p(d|\varphi_m)$ with M aperture points, and it is the estimated position. So the likelihood function can be realized by calculating $p(d|\varphi)$ along the synthetic aperture:

$$\begin{aligned} \chi(\mathbf{p}) &= 1 + \frac{1}{M} \ln \left(\prod_{m=1}^M p(d(\mathbf{a}_m, \mathbf{p})|\varphi_m) \right) \\ &= 1 - \frac{1}{M} \sum_{m=1}^M \left(\frac{\Delta\varphi_m \mod 2\pi}{2\pi} \right)^2 \\ \Delta\varphi_m &= \varphi_m - \varphi_{rt} - \frac{4\pi d(\mathbf{a}_m, \mathbf{p})}{\lambda} \end{aligned} \quad (3)$$

Equation (3) utilizes the coherence of the phase signal to eliminate the position ambiguity.



(a) $p(d|\varphi)$ with breakpoints suffers from noise. (b) $p(d|\varphi)$ without breakpoint.

Fig. 2. $p(d|\varphi)$ with and without smoothing (a. Equation (2), b. Equation (6)).

Furthermore, the actual phase data is multi-frequency as mentioned before. And due to the continuous motion of the reader antenna, there is only one effective carrier frequency at each aperture point. So φ_m at different $\mathbf{a}_m$ may have individual carrier frequencies. In other words, they have different φ_{rt} and λ in Equation (3). If the carrier frequency index of $\mathbf{a}_m$ is n_m ($n \in [1, N]$), the likelihood function is:

$$\chi(\mathbf{p}) = 1 - \frac{1}{M} \sum_{m=1}^M \left(\frac{\Delta\varphi_m \bmod 2\pi}{2\pi} \right)^2 \quad (4)$$

$$\Delta\varphi_m = \varphi_m - \varphi_{n_m,rt} - \frac{4\pi d(\mathbf{a}_m, \mathbf{p})}{\lambda_{n_m}}$$

In actual scenarios, a huge number of tags are deployed and their diversity cannot be avoided. It is almost unrealistic to measure φ_{rt} for every reader antenna and tag. Fortunately, φ_{rt} can be canceled out for each other by the relative phase. For each carrier frequency, its first phase value is used as a reference one, and every measured phase at the same carrier frequency subtracts this reference one, while the phase estimated by the hypothetical position at each aperture point also subtracts the first one. For $\mathbf{a}_m$ with carrier frequency index n_m , if the first aperture point with the same carrier frequency is $\mathbf{a}_{\hat{m}}$ ($n_m = n_{\hat{m}}$), Equation (4) can be modified as follow:

$$\chi(\mathbf{p}) = 1 - \frac{1}{M} \sum_{m=1}^M \left(\frac{\Delta\varphi_m \bmod 2\pi}{2\pi} \right)^2 \quad (5)$$

$$\Delta\varphi_m = \varphi_m - \varphi_{\hat{m}} - \frac{4\pi(d(\mathbf{a}_m, \mathbf{p}) - d(\mathbf{a}_{\hat{m}}, \mathbf{p}))}{\lambda_{n_m}}$$

The graph of the periodic function $p(\varphi|d)$ in Equation (2) has the general appearance of Fig. 2(a). The y -axis is $p(d|\varphi)$, and its x -axis is describing the estimated phase $4\pi d/\lambda$ according to the possible distance d . In addition, the theoretical phase offset due to wave propagation in space is $\theta = \varphi - \varphi_{rt}$. Obviously, it is a periodic function as stated before. However, there are breakpoints at $x = \theta + k2\pi$ ($k \in N$), and the maximum probability is just at the breakpoint of each cycle. On account of the discontinuous function, Equation (2) may not work well in consideration of the interference. Without interference, $\varphi - \varphi_{rt}$ equals to θ and d can be achieved correctly. However, there is a serious risk that d is estimated to right cycle with noise due to the modulo operation.

The solution of this problem can come to mind easily. Equation (2) should be smoothed to a continuous function and in each cycle, it is symmetric with respect to $x = \theta + k \cdot 2\pi$ ($k \in N$). The modulo operation in Equation (2) maps the phase difference to $[0, 2\pi)$, which is why there are breakpoints and $p(\varphi|d)$ is discontinuous. To map the phase difference to $[-\pi, \pi)$, the phase difference adds π before $\bmod 2\pi$, while it subtracts π after modulo operation, which is illustrated in Equation (6) and Fig. 2(b).

$$p(d|\varphi) = e^{-\frac{(((\varphi - \varphi_{rt} - \frac{4\pi d}{\lambda} + \pi) \bmod 2\pi) - \pi)^2}{\pi^2}} \quad (6)$$

As shown in Fig. 2(b), $4\pi d/\lambda$ will be estimated from $\theta - \pi$ to $\theta + \pi$, where θ is the center of the cycle with no ambiguity. Then the likelihood function is also modified to Equation (7):

$$\chi(\mathbf{p}) = 1 - \frac{1}{M} \sum_{m=1}^M \left(\frac{((\Delta\varphi_m + \pi) \bmod 2\pi) - \pi}{\pi} \right)^2 \quad (7)$$

$$\Delta\varphi_m = \varphi_m - \varphi_{\hat{m}} - \frac{4\pi(d(\mathbf{a}_m, \mathbf{p}) - d(\mathbf{a}_{\hat{m}}, \mathbf{p}))}{\lambda_{n_m}}$$

B. Resolution of Imaging

As mentioned in Section III, $\chi(\mathbf{p})$ yields an image amplitude for each grid, and it is an ambiguity function for imaging. If the reader can get the reflected signal at any point along the aperture, the sum in Equation (4) is replaced by an integral with the new $p(d|\varphi)$ in Equation (6):

$$\chi(\mathbf{p}) = 1 - \frac{1}{D} \int_{ap} \left(\frac{((\Delta\varphi_r + \pi) \bmod 2\pi) - \pi}{\pi} \right)^2 dr \quad (8)$$

$$\Delta\varphi_r = \varphi_r - \varphi_{r,rt} - \frac{4\pi d(\mathbf{a}(r), \mathbf{p})}{\lambda(r)}$$

where ap is the aperture curve, D is the aperture length and r is the position of aperture point $\mathbf{a}_m$.

As shown in Fig. 1, the reader antenna $(x_a, 0)$ moves along the line segment from $(-D/2, 0)$ to $(D/2, 0)$. The actual tag position is (x_t, y_t) , and a hypothetical tag position can be expressed by $\mathbf{p} = (x_t + \Delta x, y_t + \Delta y)$. Then $\chi(\mathbf{p})$ is a function of $\Delta = (\Delta x, \Delta y)$. For ease of description, it is assumed that the reader works on a fixed carrier frequency at first and there is no measurement error. So $\varphi_r - \varphi_{r,rt}$ is equal to $(4\pi d_a(0))/\lambda$, and Equation (8) can be written as:

$$\chi(\Delta) = 1 - \frac{1}{D} \int_{-\frac{D}{2}}^{\frac{D}{2}} \left(\frac{((\Delta\varphi_a + \pi) \bmod 2\pi) - \pi}{\pi} \right)^2 dx_a$$

$$\Delta\varphi_a(\Delta) = \frac{4\pi(d_a(0) - d_a(\Delta))}{\lambda(x_a)}$$

$$d_a(\Delta) = \sqrt{(x_t + \Delta x - x_a)^2 + (y_t + \Delta y)^2} \quad (9)$$

For a hypothetical position near the actual one, Δx and Δy are close to zero, and both $(4\pi d_a(0))/\lambda$ and $(4\pi d_a(\Delta))/\lambda$ are in the same cycle. So the range of their difference is from -2π to 2π . Considering that it is the square of the difference used for the integral, the modulo operation can be ignored, and the denominator in the integral should be 2π . Then Taylor series

expansion at $(0,0)$ is utilized to approximate the nonlinear $\Delta\varphi_a(\Delta)$ with a linear one:

$$\Delta\varphi_a(\Delta) \approx \frac{4\pi}{\lambda d}(\Delta x(x_t - x_a) + \Delta y y_t) \quad (10)$$

where the distance between the tag and the aperture points is approximately a constant d . If and only if d is much longer than the aperture length D , it can be seen as a constant.

Then Δy and Δx in Equation (9) are set to zero respectively for the lateral and radial resolution discussion. When Δy is zero, χ is a quadratic function about Δx :

$$\chi(\Delta x, 0) \approx 1 - \frac{12x_t^2 + D^2}{3(\lambda d)^2} \Delta x^2 \approx 1 - \frac{D^2}{3(\lambda d)^2} \Delta x^2 \quad (11)$$

After setting $\chi(\Delta x, 0) = \sqrt{2}/2$, we can get the -3dB width of the maximum in lateral direction δ_{lat} , which shows the lateral resolution of the imaging. As shown in Equation (12), it is inverse proportional to the aperture length D approximately.

$$\delta_{lat} \approx 1.87 \frac{\lambda d}{D} \quad (12)$$

Similarly, the -3dB width of the maximum in radial direction is:

$$\delta_{rad} \approx 0.54 \frac{\lambda d}{y_t} \approx 0.54 \lambda \quad (13)$$

Then the lateral and radial resolution are extended to the multi-frequency scenario. If the carrier frequency changes from f_c to $f_c + \Delta f$ continuously at any aperture point, Equation (9) is modified to:

$$\chi = 1 - \int_{f_c}^{f_c + \Delta f} \int_{-\frac{D}{2}}^{\frac{D}{2}} \frac{(((\Delta\varphi_a + \pi) \bmod 2\pi) - \pi)^2}{\pi^2 D \Delta f} dx_a df \quad (14)$$

The lateral and radial resolution in multi-frequency scenario can be estimated similar to mono-frequency:

$$\begin{aligned} \delta_{lat} &\approx \frac{1.87cd}{D\sqrt{f_c^2 + f_c\Delta f + \Delta f^2/3}} \\ \delta_{rad} &\approx \frac{0.54c}{\sqrt{f_c^2 + f_c\Delta f + \Delta f^2/3}} \end{aligned} \quad (15)$$

Obviously, the image ambiguity of our method in multi-frequency scenario is smaller than that in mono-frequency scenario, which means multi-frequency improves the accuracy.

C. Phase Calibration

As stated before, relative phase method in Equation (5) is leveraged to deal with φ_{rt} at different carrier frequencies. There may be 16 (China version) or 50 (FCC version) carrier frequencies with a short channel dwell time. In some worse cases, a phase may be measured at individual carrier frequency. So the relative phase method is unfeasible in this situation.

We measure the offset φ_{rt} by directly placing the tag at the center of the reader antenna and adjusting the hopping policy (hopping sequentially from channel 1 to 16) with 3s channel dwell time. As shown in Fig. 3, it is obvious that the phase differences between two adjacent carrier frequencies are almost the same. After changing reader antennas and tags, we

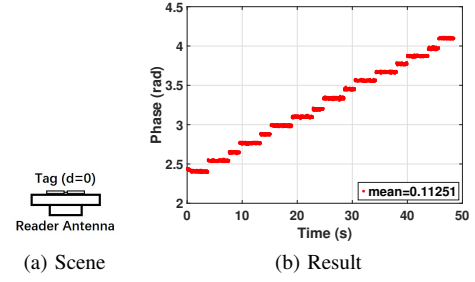


Fig. 3. The phase changes along with carrier frequencies.

repeat this experiment and find that this difference is a constant Δ_f for the same antenna.

Observation 1: The difference between φ_{rt} at two adjacent carrier frequencies is a constant Δ_f for the same antenna.

Therefore, the phase offsets φ_{rt} at different carrier frequencies can be normalized to the same one. If the phase offset at i -th carrier frequency is $\varphi_{i,rt}$, the phase offset at j -th carrier frequency should be $\varphi_{i,rt} + (j - i)\Delta_f$. The measured phase φ_1 at the first aperture point is utilized as the reference one. A normalized relative phase method is proposed to modify Equation (7), and the workload to measure Δ_f is convenient.

$$\begin{aligned} \chi(\mathbf{p}) &= 1 - \frac{1}{M} \sum_{m=1}^M \left(\frac{((\Delta\varphi_m + \pi) \bmod 2\pi) - \pi}{\pi} \right)^2 \\ \Delta\varphi_m &= \varphi_m - \varphi_{1,cal} - 4\pi \left(\frac{d(\mathbf{a}_m, \mathbf{p})}{\lambda_{n_m}} - \frac{d(\mathbf{a}_1, \mathbf{p})}{\lambda_{n_1}} \right) \end{aligned} \quad (16)$$

$$\varphi_{1,cal} = \varphi_1 + (n_m - n_1)\Delta_f$$

In addition, most existing synthetic aperture RFID approaches abstract the antenna as a static point, which is inconsistent with the actual situation. The phase center of the antenna from which radiation emanates to form ideal spherical wave fronts, moves on the evolute of the equiphase wave front [39]. This is known as Angle-of-Arrival response as well.

This AoA response of the antenna can be measured through the method below. If the antenna is at $(0,0)$ and a static tag at $(0,1m)$, the antenna rotates around its vertical axis and AoA changes from 0 to π (This experiment uses the turntable of the linear guide shown in Fig. 5). Red points and blue fitting curve $p(\theta)$ in Fig. 4 depict the relationship between the measured phase and the AoA, and the AoA-dependent phase has a broad range of about $3rad$.

In the 3D scenario, the pitch angle may be different among tags. In Fig. 4(a), the tag and the antenna have the same height. If there is another tag at $(0,1m)$ with the height of $0.12m$, it will have a pitch angle of $0.12rad$. We lift the tag height up to $0.12m$ and repeat the experiment mentioned before, so φ_{rt} will not change in experiments. The new green phase fitting curve is drawn in Fig. 4(b), which has the same shape as the blue one. The phase difference between two curves is about $0.3rad$, and $4\pi(\sqrt{0.12^2 + 1^2} - 1)/0.32 = 0.28rad$. If the pitch angle is in a small range, it will not impact the measured phase.

Observation 2: The measured phase is determined by the distance and its angle-of-arrival to the reader antenna.

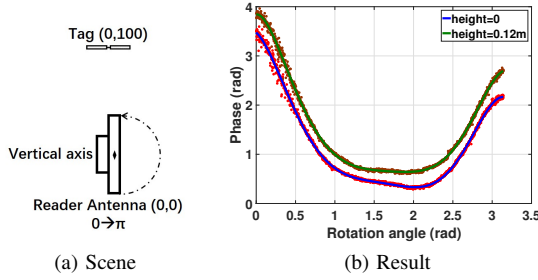


Fig. 4. The phase changes along with angle-of-arrival.

So for the likelihood $\chi(\mathbf{p})$ estimation such as Equation (16), this AoA-dependent phase offset needs compensation. The phase value at $\text{AoA}=\pi/2$ in $p(\theta)$ is used as a reference. Then all ideal phase values are corrected with $p(\theta)$ by adding $p(\theta) - p(\pi/2)$, where θ is estimated based on the hypothetical tag position $\mathbf{p}$ and the reader antenna position $\mathbf{a}_m$. So $\Delta\varphi_m$ in Equation (16) is modified, and $\varphi_{\mathbf{p}}(\mathbf{a}_m)$ is the theoretical phase value from $\mathbf{p}$ at aperture point $\mathbf{a}_m$.

$$\begin{aligned} \Delta\varphi_m &= \varphi_m - \varphi_{1,cal} - (\varphi_{\mathbf{p}}(\mathbf{a}_m) - \varphi_{\mathbf{p}}(\mathbf{a}_1)) \\ \varphi_{\mathbf{p}}(\mathbf{a}_m) &= \frac{4\pi d(\mathbf{a}_m, \mathbf{p})}{\lambda_{n_m}} + (p(\theta(\mathbf{a}_m, \mathbf{p})) - p(\pi/2)) \end{aligned} \quad (17)$$

It seems difficult to get $p(\theta)$ without the turntable. Nevertheless, $p(\theta)$ can still be estimated without it. If there is a tag whose position is known, its AoA and distance can be estimated when the reader antenna interrogates it at different aperture points. Then all phase offsets at different points due to the distance difference are normalized with the reference distance $\text{AoA}=\pi/2$. For example, if the distance between the tag and the aperture point $\mathbf{a}_{\pi/2}$ with AoA $\pi/2$ is $d(\mathbf{a}_{\pi/2})$, and the tag is interrogated at $\mathbf{a}_\theta$ with AoA θ and distance $d(\mathbf{a}_\theta)$, $p(\theta)$ can be estimated by: $p(\theta) = \varphi - 4\pi(d(\mathbf{a}_\theta) - d(\mathbf{a}_{\pi/2}))/\lambda$, where φ is the measured phase at $\mathbf{a}_\theta$.

After phase calibration, the estimated phase values are more consistent with the measured ones, and the localization result will be more accurate.

D. Weighted Enhancement

Different measured phase values may be subject to different interference like multipath distortion and measurement noise. This diversity is a guarantee for localization accuracy owing to the coherent reconstruction. Correspondingly, different measured phase values have dissimilar reliability, so the localization performance can be further improved with weighted method.

According to official Impinj documentation [6], the received signal phase follows a low amplitude Gaussian distribution with the standard deviation 0.1rad . So we assume that the measured phase φ_m at $\mathbf{a}_m$ with carrier frequency f_n follows a Gaussian distribution $\varphi_m \sim N(\mu_{m_n}, \delta^2)$. After Kolmogorov-Smirnov test, δ is 0.1, which matches with [6]. So if the hypothetical position $\mathbf{p}$ is the actual one, the phase difference between the measured phase and the theoretical one according to $\mathbf{p}$ should follow Gaussian distribution: $(\varphi_m - \varphi_{rt} -$

$4\pi d(\mathbf{a}_m, \mathbf{p})/\lambda) \sim N(0, \delta^2)$, and this difference is $\Delta\varphi_m$ in Equation (3) and (4).

In Equation (7) and (16), the relative phase method or normalized one is utilized, where $\Delta\varphi_m$ is the relative phase difference, and it follows the difference of two identical Gaussian distributions: $\Delta\varphi_m \sim N(0, 2\delta^2)$. So for every φ_m with the hypothetical tag position $\mathbf{p}$ in Equation (16), it can be weighted by $w_{\mathbf{p}}(m)$:

$$\begin{aligned} \chi(\mathbf{p}) &= K \left(1 + \frac{1}{M} \ln \left(\prod_{m=1}^M w_{\mathbf{p}}(m) p(d(\mathbf{a}_m, \mathbf{p}) | \varphi_m) \right) \right) \\ w_{\mathbf{p}}(m) &= f(\Delta\varphi_m; 0, \sqrt{0.02}) = \frac{1}{\sqrt{0.04\pi}} e^{-\frac{\Delta\varphi_m^2}{0.04}} \end{aligned} \quad (18)$$

where K is a normalization factor and f is the probability density function of the Gaussian distribution $N(0, 0.02)$. Due to limited space, normalized relative phase and AoA calibration are omitted in Equation (18).

With weighted enhancement, the main lobe in the imaging will be much narrower, and there will be a rapid drop of the likelihood around the actual position. The main lobe expresses the high likelihood region around the actual position, so a larger one means the estimation result is more vulnerable to the interference. The weight $w_{\mathbf{p}}(m)$ enhances the hypothetical tag position with high likelihood, and weakens other positions to filter out them.

V. EVALUATION AND ANALYSIS

Extensive experiments are designed to demonstrate the key characteristics of this novel synthetic aperture RFID method.

A. Experiment Description

A linear guide shown in Fig. 5 is used for performance evaluation. A Laird S9028PCR antenna connected with a COTS RFID reader Impinj 420 is mounted on this experimental equipment with the maximum horizontal moving range 1m . This equipment reports the antenna position every 10ms . The reader antenna is under the lab table and moves along the linear guide, and there are 8 RFID-tagged test tubes which contain some water. The distance between the tags and the aperture is about 1m . In addition, a reflector is leveraged to enhance multipath distortion while extra readers and tags are utilized to produce interference. The RFID reader operates from 920.625MHz to 924.375MHz with 16 different carrier frequencies. The communication between the reader and the computer is based on LLRP protocol [40]. The computer is equipped with an Intel Core i7-4790 CPU 3.6GHz and a 16GByte memory.

B. Localization Accuracy

The mean error in both lateral direction and radial direction is utilized for localization performance evaluation, which is the absolute distance between the estimated position and the ground truth in two directions respectively. With 1m aperture length along the x-axis, 21 aperture points and 16 carrier frequencies, the mean error of different methods is shown in Table I. The accuracy of the holographic imaging methods

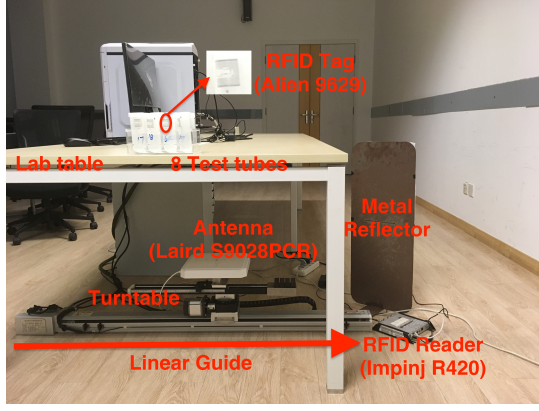


Fig. 5. The experiment equipment.

TABLE I
THE MEAN ERROR OF DIFFERENT METHODS.

method	lateral (m)	radial (m)	likelihood
this paper	0.031	0.035	ranging
Tagoram [15]	0.042	0.133	holographic
MobiTagbot [16]	0.026	0.108	holographic

(Tagoram [15] and MobiTagbot [16]) in lateral direction is one order of magnitude higher than that in radial direction. These three methods have similar accuracy in lateral direction whereas our method is more accurate in radial direction.

Tagoram [15] leverages multiple antennas for good accuracy but its performance is poor in the single antenna scenario. MobiTagbot [16] uses weight based on entropy to enhance its performance, but its antenna moves intermittently to measure phase on different carrier frequencies at the same aperture point, which is inconvenient for many applications.

Similar to the lateral and radial resolution of the ranging based reconstruction method in Equation (15), the resolution of holographic imaging can also be estimated. For the scenario shown in Fig. 1, the lateral and radial resolution of holographic imaging in multi-frequency scenario are shown in Equation (19), which are also estimated by Taylor series expansion [41].

$$\delta_{lat} \approx \frac{0.44cd}{Df_c}, \quad \delta_{rad} \approx \frac{0.44c}{\Delta f} \quad (19)$$

Apparently, the lateral resolution of ranging based and holographic imaging based methods are in the same level, but the radial resolution of ranging based one is much better due to the sharp difference between the carrier frequency f_c and the bandwidth Δf . And the analyses in theory accord well with the experiment results.

As stated before, the likelihood function is also an amplitude function for imaging. And images are intuitive to show the performance of these methods. These images based on holographic imaging and ranging methods are depicted in Fig. 6 with the same data collected from one typical experiment, in which the actual tag position is $(0.46m, 0.89m)$. Although the ranging method utilizes the received phase for distance

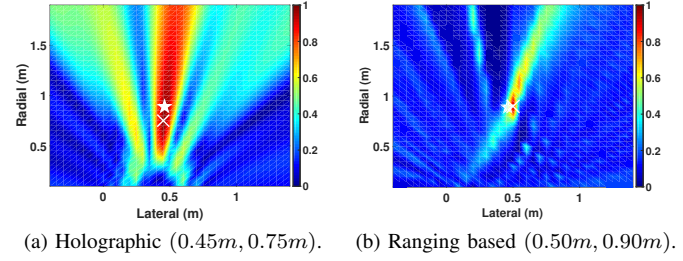


Fig. 6. The likelihood reconstruction images (the white star is the actual position, and the white cross is the estimated position).

estimation which leads to a $\lambda/2$ ambiguity, it has a satisfactory accuracy.

The proposed ranging based method differs from previous synthetic aperture RFID systems in four aspects. Firstly, it is a ranging based method and not a holographic imaging one, and it has better theoretical accuracy in radial direction. Secondly, the frequency response on different carrier frequencies is eliminated by normalized relative phase, and the pre-processing on every tag is not needed. Thirdly, the AoA response of the received signal phase is normalized by pre-processing, which greatly improves the accuracy. Previous work does not consider the received signal phase with respect to the AoA, and a longer aperture even cannot improve accuracy. Because in this situation, AoA can vary in a wider range and the measured phase is inaccurate. Multi-frequency provides more diversity when signals are faded at some carrier frequencies. With multi-frequency and weighted enhancement, the phase noise and multipath distortion are efficiently suppressed. So the accuracy is further improved with a robust position estimation in complicated indoor propagation environment.

C. Effectiveness of Calibration and Weighted Enhancement

The calibration and weighted enhancement play important roles in the ranging reconstruction method. As illustrated in Fig. 7(a), without phase calibration, the mean error in lateral and radial direction increases by $0.006m$ and $0.014m$ respectively. Moreover, without the weighted enhancement, the likelihood reconstruction image is shown in Fig. 7(b), which uses the same data with Fig. 6. Obviously, the side lobes may seriously damage the position estimation in the complicated indoor propagation environment. The calibration and weighted enhancement are highly effective and can significantly improve the localization accuracy.

D. Benchmarks

To deeply analyze the performances of this novel ranging based synthetic aperture RFID method, several detailed benchmarks are performed.

1) *Aperture Length*: As illustrated in Equation (15), a longer aperture can improve the localization accuracy in lateral direction. A longer aperture length means more significant difference in multipath distortion and noise among aperture points, which can be weakened by each other in the coherent reconstruction. As shown in Fig. 7(c), a shorter aperture brings

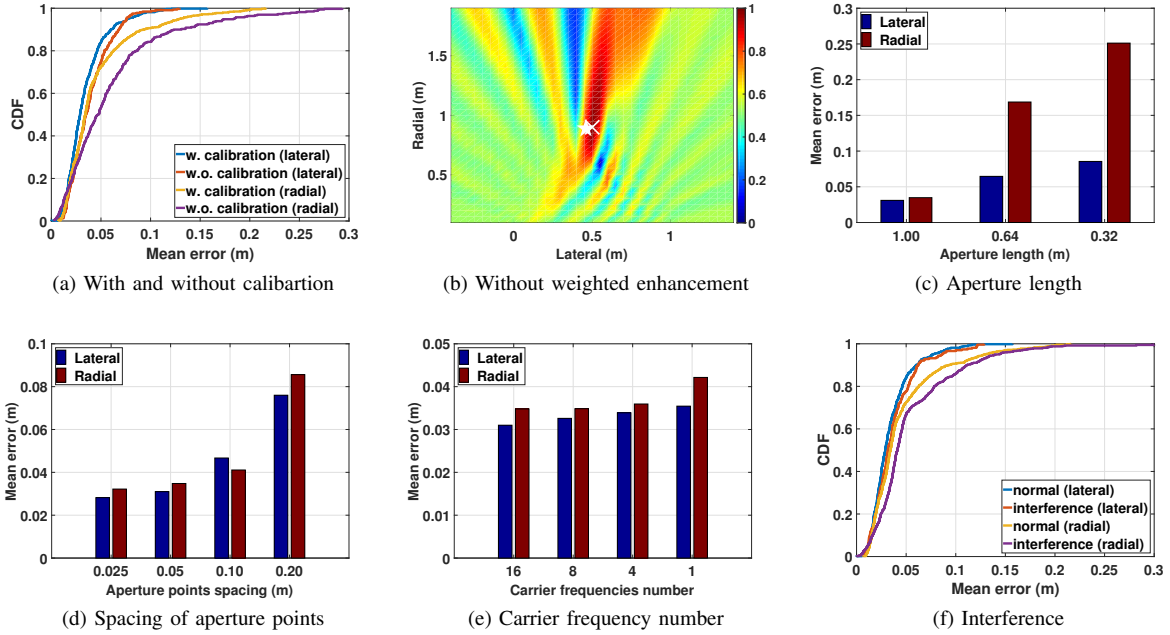


Fig. 7. The evaluation of the novel ranging based synthetic aperture RFID localization. (baseline with $1m$ aperture length, $0.05m$ spacing and 16 carrier frequencies.)

larger error. When the aperture length is decreased from $1m$ to $0.32m$ (one wavelength), the mean error in lateral and radial direction increases by $0.05m$ and $0.21m$ respectively. So the aperture length is set to $1m$ which can achieve satisfactory accuracy.

2) *Spacing of Aperture Points*: For a fixed length aperture, smaller spacing of aperture points means more number of aperture points, which is similar to repeated measurements to reduce error. And the localization error caused by measured error and multipath distortion, can be suppressed effectively with more sample points. As illustrated in Fig. 7(d), when the spacing of aperture points increases from $5cm$ to $10cm$, the mean error in radial direction increases by less than $0.01m$. Usually, down-sampling along the aperture is leveraged to reduce the computational burden. There is a trade-off between the accuracy and the computational burden. And in this paper, the spacing of aperture points is $0.05m$.

3) *Carrier Frequencies Number*: As shown in Equation (19), a wider bandwidth has much more improvement in the radial resolution for holographic imaging. The impact of the carrier frequencies number on the ranging based reconstruction is illustrated in Fig. 7(e). It is evident that multi-frequency can improve the precision only a little, which is about $0.01m$ in radial direction. This result matches Equation (15). In respect that the bandwidth of our RFID system is less than $5MHz$, it is still too narrow, especially compared with the RFID signal frequency $920MHz$. Our method achieves similar accuracy with 4 to 16 carrier frequencies, so it is not necessary to modify the reader with special carrier frequencies numbers, whose default number is 16.

4) *Interference (noise and Multipath Distortion)*: The received signal consists of noise and the signals reflected from

a tag and other reflectors known as multipath signals. Only the signal directly reflected by the tag includes the available information. In the ranging based method, interference like noise and multipath distortion in measured phase can be weakened by each other in coherent reconstruction. It is difficult to carry out a quantitative analysis on them. So we utilize an extra reader and extra tags to produce interference and a metal reflector to enhance multipath distortion. As depicted in Fig. 7(f), the mean error increases by about $0.008m$ in both lateral and radial direction. We can conclude that this method is not so sensitive to the multipath and noise in that the multipath and noise profiles are different from one aperture point to another resulting in being effectively canceled out for each other. And the interference by noise and multipath distortion can be further suppressed by the weighted enhancement.

VI. CONCLUSION

In this paper, we present a novel ranging based reconstruction method for the synthetic aperture RFID localization. The RFID phase based ranging method is leveraged to construct the likelihood function for synthetic aperture RFID. Both the frequency and AoA response of the received signal phase are eliminated by normalized relative phase and pre-processing method. Additionally, the localization performance is enhanced by a weighted method. Our method utilizes only one reader antenna and one trajectory to construct the synthetic array with accurate localization results both in lateral and radial direction. We conduct extensive experiments to validate that this method can be robust enough to work in the complicated indoor propagation environment with commercial-off-the-shelf RFID tags and readers. This method achieves a median accuracy of $3cm$ in both lateral and radial direction,

which opens up new possibilities and opportunities for many applications. In the future work, we will focus on the denser test tubes localization, a more convenient calibration based on AoA response, the challenges to adaptively compensate for drift of sensors and to reduce the computational burden of this method.

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