Radio Positioning with EM Processing of the Spherical Wavefront

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Abstract

Next 5G and beyond applications have brought a tremendous interest towards array systems employing an extremely large number of antennas, so that the technology that might be in place for communication can be also exploited for positioning. In particular, in this paper we investigate the possibility to infer the position of an omnidirectional transmitter by retrieving the information from the incident spherical wavefront through its EM processing. Despite such a post-processing of the curvature wavefront has been mainly considered in the past at microwave and acoustic frequencies using extremely large antennas, it is of interest to explore the opportunities offered in the context of next 5G and beyond systems. Thus, differently from the state-of-the-art, here we first introduce a dedicated general model for different EM processing configurations, and successively we investigate the trade-off between the attainable positioning performance and the complexity offered by the different architectures, that might entail or not the use of a lens, that can be either reconfigurable or not. Indeed, we analyze also the effect of the interference, in order to evaluate the robustness of the considered system to the presence of multiple simultaneous transmitting sources. Results, obtained for different number of antennas, i.e., for different array apertures, confirm the possibility to achieve interesting positioning performance using a single antenna array with limited dimensions.

Index Terms

Spherical Wavefront, positioning, massive array, EM lens, Reconfigurable lens, mm-wave

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I. INTRODUCTION

Indoor positioning systems have recently attracted a great interest in a large variety of scenarios where the signal coming from the Global Navigation Satellite System (GNSS) is denied [1]. In fact, even if GNSS is recognized to be one of the most accurate sources of position information, it is often not available in indoors, and alternative positioning systems are required. Currently, there is a large variety of ad-hoc solutions for indoor localization and tracking [2]–[4], spanning from the systems based on ultrasounds or to more recent impulse radio ultrawide bandwidth (UWB) techniques [3], [5], [6]. Unfortunately, most of such solutions require that a mobile node is detected at least from three reference nodes (anchor nodes) located in known positions, with the need to realize ad-hoc and often redundant infrastructures that might become not convenient in many indoor scenarios. It would be of great help if the networks deployed for communication could be exploited also for indoor localization. Unfortunately, such networks are usually designed to guarantee single-anchor coverage.

In this context, fifth generation (5G) mobile wireless networks will introduce new technologies that might be exploited to tackle this problem. In particular, next 5G foresees the joint use of millimeter-waves (mm-wave) and massive arrays to enable the integration of arrays with a large number of antennas into small areas. By enabling such an architecture capable to realize near pencil beam antennas, it becomes feasible not only to boost communication but also single-anchor localization capabilities at an unprecedented scale [7]–[11].

In general, direct positioning approaches applied to single antenna arrays provide better performance than two-step localization algorithms which are based on the estimation of intermediate quantities such as angle-of-arrival (AOA) and time-of-arrival (TOA) as the latter may be suboptimal according to the data processing inequality [12]–[14]. However, two-step localization algorithms are often implemented in practice as they are more pragmatic and less complex [15]–[19].

In any case, such solutions require multiple interactions between transmitter and receiver as well as an extremely precise system synchronization [20], which could reduce the available bandwidth for communication and make the system still costly.

A possible alternative solution is to infer the transmitter position from the spherical wavefront, which is possible in all those situations where the wavefront curvature is significant with respect to the antenna aperture in relation to the wavelength. In fact, while in far-field propagation
regime the wavefront is plane and only the AOA information can be inferred using an antenna array, when operating in near-field regime (Fresnel region) the wavefront tends to be spherical and also the distance information can be inferred from it, and hence the position.

This concept is not new, and it has been widely exploited for acoustic waves [21], [22] or at microwaves only considering very short distances or using very large (often not practical) antennas [23]. In [24], the curvature information has been exploited, with a moving source approaching to the receiver so that, entering in the Fresnel region, the incoming wave cannot be regarded as plane anymore. For instance, in [25], an approach using multi-tone signalling and multi-arrays is described, whereas in [26], a MUSIC-based method is proposed and an extensive analysis on the attainable fundamental localization limits is derived in near-field propagation conditions [27]. In addition, other previous works apply the Fresnel approximation to arrays with special geometries, e.g. uniform linear arrays [28]–[31], and introduce a model mismatch that might jeopardize the achievable positioning precision [32], and thus solutions based on look-up tables have been proposed [33].

With the advent of mm-wave-based solution, direct positioning is in principle possible even with antenna arrays with limited aperture [34]. Several architectures can be of particular interest, exploiting or not the presence of an electromagnetic (EM) lens performing the processing of the incident wavefront. Such EM lenses can be reconfigurable, here namely reconfigurable lens (R-lens), or not, here namely non-reconfigurable lens (NR-lens).

According to NR-lens, recent studies have investigated the possibility to exploit EM lens-based massive arrays operating at mm-wave as a promising solution for drastically reducing the overall system complexity [35]. In fact, by adopting a lens to collimate the beams in precise directions, it is possible to spatially discriminate signals in the analog domain [35]–[37]. Consequently, thanks to the lens, there is a unique relation between the incident and the output angles of the impinging and refracted waves, respectively. This operation allows the reduction of the number of antennas with respect to traditional massive arrays, and to move from discrete beamforming architectures towards continuous-aperture-phased arrays.

While such lenses are passive and thus not reconfigurable, R-lens-based solutions are expected instead to be programmable in real-time so that they are capable to accomplish more advanced tasks. In this sense, beyond 5G solutions are moving towards the realization of reconfigurable intelligent surface (RIS) with a large size, i.e., a wall, for the realization of a smart environment, where walls and objects can be equipped with tailored solutions so that a programmable indoor
wireless environment can be created [38]–[44]. To this purpose, metamaterials represent an important and recent solution for the realization of RIS and R-lens [45] thanks to the possibility to achieve a flexible control of the wavefront while guaranteeing compact size [46], [47]. Other interesting opportunities account for the execution of mathematical operations with layers of metamaterials [48], the realization of electronically reconfigurable transmitarray [49], or reconfigurable reflectarray technology [50].

In this context, starting from our previous analysis in [37], [51], we here investigate the radio positioning capabilities of a mm-wave source by introducing a generic architecture composed of an EM processing section, that can directly operate on the wavefront curvature, and of an array architecture that collects the impinging signal, as shown in Fig. 1. More specifically, differently from the state of the art, we first introduce a general model that entails the presence (if any) of an EM lens performing the processing at EM level before the antenna array, then we investigate the complexity trade-off between different architectures in terms of presence or absence of lens, reconfigurable or not reconfigurable lens, and the number of employed antennas. Such trade-off is then investigated by analyzing the attainable positioning accuracy and interference rejection capability when the aperture is varied.

The main contributions of the manuscript can be summarized as follows:
• We provide a generalized framework for retrieving the information from the wavefront curvature when EM processing is performed prior the processing of signals at each antenna;
• We consider different practical schemes, employing or not the use of an EM lens, that account for a trade-off in terms of required antennas and processing complexity;
• We evaluate the positioning performance, comprising a differential approach that might unburden the computational complexity in practical systems for no EM processing (nEMP)-based architectures;
• We evaluate the impact of the interference by determining the capability of the proposed architectures to spatially discriminate multiple transmitters.

The remainder of the paper is organized as follows. Sec. II contains insights on how to gather position information from the signal wavefront, where Sec. IV reports considerations on the interference. Sec. III shows positioning techniques, and Sec. V describes the achieved results. Conclusions are finally drawn in Sec. VI.

II. Position Information in the Wavefront Curvature

A. Operating Frequency Impact

We now investigate the trade-off between the size of the array and the operating frequency to determine the region where the impact of the wavefront curvature is appreciable, by considering the Fraunhofer distance \( d_F = \frac{2D^2}{\lambda} \), with \( \lambda \) indicating the wavelength, and \( D \) the diameter of the antenna.

The impact of the operating frequency is reported in Fig. 2. More specifically, two scenarios are considered. In the first one, the number of employed antennas is kept constant regardless the frequency or the array size. In the latter, the constraint is on the array dimension: in fact, it is kept fixed since in practical applications it is required the adoption of antennas with a constrained size. Notably, if the constraint is on the number of usable antennas, the curvature effect is more appreciable at lower frequencies as reported in Fig. 2-left. On the other side, if the size of the array is constrained to a certain value, a higher frequency allows to gather a better information on the wavefront curvature (see Fig. 2-right). As an example, for \( D = 50 \) cm, the Fraunhofer region, for which the wavefront is considered planar, starts at \( d \simeq 10 \) m for \( f_0 = 5 \) GHz, and at \( d \simeq 100 \) m for \( f_0 = 60 \) GHz.

Such considerations confirm that the impact of the wavefront curvature is not negligible when large antenna arrays, operating at high frequency, are employed. To this purpose, in the following
Fig. 2: Start of the Fraunhofer region for different frequencies and different array sizes.

we introduce an ad-hoc general model, valid for the presence (if any) of an EM lens, to retrieve the position information from the incident spherical wavefront.

**B. Signal Model**

As previously stated, the EM processing can be realized with different techniques (i.e., EM lens, metamaterials, etc), which allow to retrieve the position from an omnidirectional source by exploiting the wavefront curvature.

To this purpose, consider a transmitting source located at position \( p \), which is at distance \( d \) from the reference point of the RX, and denote with \( \Theta = (\theta, \phi) \) the incident angle.

Suppose there are \( N_A \) receiving antennas in positions \( \{p_n\}, n = 0, 1, \ldots N_A \). Define \( (0, 0, z) \) with \( z \in [0, D] \) the surface of the EM lens (if any).

Then, we denote with \( r = [r_0, \ldots, r_n, \ldots r_{N_A-1}]^T \) the vector containing the equivalent complex baseband signal received at each antenna, defined as

\[
r = s + w = \mathcal{F}(h(y, z, p)) + w
\]

where \( s = [s_0, \ldots, s_n, \ldots s_{N_A-1}]^T \) and \( w = [w_0, \ldots, w_n, \ldots w_{N_A-1}]^T \) are the useful signal component and the additive white Gaussian noise (AWGN) noise, respectively, whereas \( \mathcal{F}(\cdot) \) is the operation performed by the EM processing on the EM signal \( h(z, p) \) observed on the surface as a consequence of a source located in \( p \).
Notably, if we consider the signal in the $z$th position of the flat lens performing the EM processing, $h(z, \mathbf{p})$ contains the information on the extra distance traveled by the EM wave to reach the generic coordinate $z$ of the RX flat lens, that is

$$h(y, z, \mathbf{p}) = A_{pl} e^{-jx} e^{-j2\pi f_0 \tau(y, z, \mathbf{p})} = x_0 e^{-j2\pi f_0 \tau(y, z, \mathbf{p})}$$

(2)

where $A_{pl}$ denotes the received signal amplitude, $\chi \sim U[0, 2\pi]$ is uniformly distributed between 0 and $2\pi$, $f_0$ is the central frequency and

$$\tau(y, z, \mathbf{p}) = \frac{a(y, z, \mathbf{p})}{c}$$

(3)

with $c$ being the speed of light, and

$$a(y, z, \mathbf{p}) = -d + d \sqrt{1 + \frac{d_{0yz}^2}{d^2} + 2 \cdot \frac{d_{0yz}^2}{d} g(\Theta, \Theta_{oyz})}$$

(4)

where $d_{0yz} = \sqrt{y^2 + z^2}$ is the distance of the point with coordinates $y, z$ in the aperture from its reference point located in $(0, 0)$. The term $g(\Theta, \Theta_{oyz})$ is given by

$$g(\Theta, \Theta_{oyz}) = \sin(\theta) \cos(\phi_{0yz} - \phi) + \cos(\theta_{0yz}) \cos(\theta)$$

(5)

with $\Theta = (\theta, \phi)$ and $\Theta_{0yz} = (\theta_{0yz}, \phi_{0yz})$ being the elevation-azimuth pair of the target, and of the generic point of the aperture with coordinates $(y, z)$, respectively. Note that for a planar aperture lying on the $YZ$-plane, it holds $\theta_{oyz} = 90^\circ$.

Thus, $\chi$ includes the complete uncertainty on the received signal phase, since the transmitter and receiver are supposed to be not synchronized and no information can be retrieved from the TOA of the received signal. Differently from classical antenna arrays with planar wavefront, $h$ does not depend only on the AOA $\theta$ but also on the distance $d$, i.e., on the position $\mathbf{p}$.

Note also that if $z \ll d$, according to the Taylor-McLaurin series expansion, it is

$$a(y, z, \mathbf{p}) \approx d_{0yz} g(\Theta, \Theta_{oyz}) + \frac{d_{0yz}^2}{2d} \left(1 - g^2(\Theta, \Theta_{oyz})\right)$$

(6)

where the first term refers to the traditional array phase term containing AOA information, whereas the second term includes information on the source distance which becomes negligible for large distances.

In the following, we analyze some possible solutions to realize the architecture herein proposed, by discussing their impact on the EM processing or on the array geometry.
Fig. 3: Left: Top-view of the R-lens scenario where the EM processing, realized with a reconfigurable lens, focuses the entire signal towards a single antenna. Right: equivalent processing performed by the R-lens through $P_1$ and $P_2$.

1) EM Processing with a R-lens: The use of a R-lens that allows to achieve a full control of the phase profile of the signal, represents an asymptotic case, since the computational complexity is put entirely on the EM processing, and the array is reduced to a single antenna, i.e., $N_A = 1$.

The objective of the R-lens herein considered is to focus the impinging wave towards one point, placed in position $p_0 = [-F_p, 0, \frac{D}{2}]$ at distance $F_p$ from the lens surface, as shown in Fig. 3. Consequently, the re-phasing procedure operated by the flat lens can be reproduced by a double-lens system performing the following two processing operations $P_1$ and $P_2$, as depicted in Fig. 3 right. More specifically, through $P_1$, the first lens converts a spherical wavefront into a planar one, with normal direction at its output. On the other side, through the operation $P_2$, the second lens focuses the normal planar incident wavefront into one point, where an antenna is located. Thus, $P_2$ does not depend on the source position $p$, whereas $P_1$ concerns the reconfigurable processing that varies with the source position.

According to such considerations, the term $\kappa(y, z, \hat{p})$, which derives from the EM processing
in the \( z \)th position, is given by
\[
\kappa(y, z, \hat{p}) = \mathcal{P}_1(y, z, \hat{p}) \cdot \mathcal{P}_2(y, z) .
\] (7)

In the following, we explicit the conditions to properly design the phase profiles of the two lenses, that is \( \mathcal{P}_1 \) and \( \mathcal{P}_2 \).

a) Design of \( \mathcal{P}_1 \): At the output of the first lens, it is required to achieve a planar wavefront, thus the phase of the signal should be the same, despite the considered \( z \)th position on the lens. Due to this consideration, \( \mathcal{P}_1(z) \) is chosen so that it holds
\[
\mathcal{P}_1(y, z, \hat{p}) \cdot h(y, z, \hat{p}) = e^{-j\Psi_{01}}
\] (8)
where \( \Psi_{01} \) is a constant term, so that it gives
\[
\mathcal{P}_1(y, z, \hat{p} = p) = e^{j(2\pi f_0 \tau(y, z, p) - \Psi_{01})} .
\] (9)
Operating like that, now we have a planar normal wavefront incident to the second lens.

b) Design of \( \mathcal{P}_2 \): Since now the wavefront is planar, \( \mathcal{P}_2(z) \) is designed so that it holds the condition
\[
\mathcal{P}_2(y, z) \cdot e^{-j\Psi_0(y, z)} = e^{-j\Psi_0}
\] (10)
where \( \Psi_0(y, z) \) accounts for the travelled distance from a generic point in \((x, y)\) on the R-lens to the focal point \( F_p \), that is
\[
\Psi_0(y, z) = \frac{2\pi}{\lambda} \sqrt{F_p^2 + d_{cyz}^2}
\] (11)
where \( d_{cyz} = \sqrt{(y - \frac{D_y}{2})^2 + (z - \frac{D_z}{2})^2} \), and \( \Psi_{02} \) indicates a constant phase term. Thus, \( \mathcal{P}_2 \) indicates the dephasing term introduced by the R-lens such that all rays with normal incidence arrive at \( F_p \) with identical phase for constructive superposition, and can be finally written as
\[
\mathcal{P}_2(h, z) = e^{j\left(\frac{2\pi}{\lambda} \sqrt{F_p^2 + d_{cyz}^2} - \Psi_{02}\right)} .
\] (12)
Without loss of generality, in the rest of the manuscript we will assume \( \Psi_{01} = \Psi_{02} = 0 \).

Thus, the signal received at the antenna, after the EM processing operated by the R-lens response, can be expressed as
\[
r = r_0 = \mathcal{F}(h(y, z, p)) + w = \frac{1}{\lambda \sqrt{D_y D_z}} \int_{D_y} \int_{D_z} \kappa(y, z, \hat{p}) \cdot h(y, z, p) e^{-j\Psi_0(y, z)} dy dz + w
\] (13)
with
\[
\mathcal{F}(h(y, z, p)) = \frac{1}{\lambda \sqrt{D_y D_z}} \int_{D_z} \int_{D_y} \kappa(y, z, \hat{p}) \cdot h(y, z, p) e^{-j\Psi_0(y,z)} dydz .
\] (14)

In case an ideal R-lens is used, it holds \(\kappa(y, z, \hat{p} = p) \cdot h(y, z, p) e^{-j\Psi_0(y,z)} = 1\), and the signal received at the antenna simply reduces to
\[
\mathcal{A} = \sqrt{A} \cdot x_0 + w
\] (15)

where \(\mathcal{A} = \frac{D_y D_z}{\lambda^2}\) represents the normalized aperture of the R-lens exploited for the realization of the EM processing.

2) EM processing with a NR-lens: We now consider the scenario where the EM processing is realized by means of a non-reconfigurable lens and a linear antenna array is employed to collect the processed signal, as shown in Fig. 4. In particular, the lens is introduced to collimate the impinging wave in specific directions in analog domain, without the need of complex ad-hoc digital signal processing or metamaterials. In addition, thanks to the employed lens, the same aperture is achieved with a larger number of antennas when a traditional antenna array is considered. On the other side, the use of only one antenna as before is no more affordable in terms of capability to gather the signal after the lens, as will be detailed in the following.

According to the guidelines given in [35], the array is equipped with \(N_A\) antennas located on the focal arc of the lens, lying on the \(xz\)-plane, with \(\theta_n \in [-\pi/2, \pi/2]\) representing the angle of the \(n\)th generic antenna element. Then, by defining \(\tilde{\theta}(p_n) = \tilde{\theta}_n = \sin \theta_n\), antenna elements are deployed so that \(\tilde{\theta}_n\) results to be equally spaced in the interval \([-1, 1]\) (critical sampling), i.e.,
\[
\tilde{\theta}_n = \frac{n \lambda}{D_z} = \frac{n}{\tilde{D}_z}
\] (16)

where \(\tilde{D}_z = D_z/\lambda\), with \(D_z\) being the lens length along the \(z\)-axis, and the antennas are positioned in \(\{p_n\} = \{(F_p \cos \theta_n, 0, F_p \sin \theta_n)\}\).

Notably, the analysis herein carried out is general and scales according to the operational frequency. Then, according to [35], the relation between \(\tilde{D}_z\) and the required number of antennas \(N_A\) is
\[
N_A = 1 + 2[\tilde{D}_z]
\] (17)

\(^1\)If planar arrays are employed, the analysis can be extended according to [52].
Fig. 4: Top-view of the EM processing, realized with a non-reconfigurable lens. The use of the lens allows to preserve the number of employed antennas affordable \[37\].

with \[\lfloor \cdot \rfloor\] denoting the largest integer no greater than its argument, so that (17) indicates the need of a higher number of antennas when the dimension \(\tilde{D}_z\) of the antenna increases. Since it is \(\theta_n = \arcsin(n/\tilde{D}_z)\), antenna elements are more densely located in the center of the system.

Thus, according to Fig. 4 we consider a 3D lens along the \(yz\)-plane. Differently from the previous case, due to practical limitations, it is not possible to consider a linear non-reconfigurable lens. By accounting for a source located on the \(xz\)-plane, the signal received on the focal arc, in the position \(p_n\), can be expressed as

\[
r_n = r(p_n) = \mathcal{F}(h(y, z, p)) + w_n = \frac{1}{\lambda \sqrt{D_y D_z}} \int_0^{D_y} \int_0^{D_z} s(y, z, p) e^{-j\Psi_n} dy dz + w_n \tag{18}
\]

where \(p_n\) is the coordinate of the point on the focal arc corresponding to angle \(\theta_n\), \(\Psi_n\) is the dephasing term given by the lens towards the antenna in \(p_n\), according to the analysis reported in \[35\] for an incident planar wavefront, and \(s(y, z, p)\) is the signal at the input of the EM processing, with AOA \(\theta\). Note that the normalization term \(1/(\lambda \sqrt{D_y D_z})\) is chosen to guarantee that the overall power intercepted by the lens is proportional to its normalized aperture \(A = (D_y D_z)/\lambda^2\) \[35\].
Notably, in our scenario the term due to the curvature of the wavefront is now present and its distribution at the output of the lens can be used to retrieve the position information. Then, by indicating with \( \tilde{z} = z/\lambda \) and \( \tilde{y} = y/\lambda \), it is possible to write

\[
\begin{align*}
r_n &= \frac{\lambda x_0}{\sqrt{D_y D_z}} \int_0^{D_y} \int_0^{D_z} e^{i2\pi \tilde{a}(y,z,p)} e^{-ij2\pi \tilde{z} \tilde{y}} d\tilde{y} d\tilde{z} + w \\
\end{align*}
\]  

(19)

with

\[
\tilde{a}(y,z,p) = \frac{a(y,z,p)}{\lambda} = -\frac{d}{\lambda} + \frac{d}{\lambda} \sqrt{1 + \frac{d_{0yz}^2}{d^2} + 2 \frac{d_{0yz}}{d} \sin \theta}.
\]

(20)

According to the antenna critical sampling in (16), we can finally write

\[
\begin{align*}
r_n &= \frac{1}{\sqrt{A}} \int_0^{D_y} \int_0^{D_z} e^{i2\pi \tilde{a}(y,z,p)} e^{-i2\pi \tilde{z} \tilde{y}} d\tilde{y} d\tilde{z} \cdot x_0 + w_n.
\end{align*}
\]

(21)

Solving (21) allows to study the impact of wavefront curvature at the receiver antennas. Note that in case the wavefront is planar, the solution of (21) reduces the one given in [35].

3) No use of EM processing (nEMP): For comparison purposes, assume now that the EM processing is avoided. This corresponds to an asymptotic scenario but, differently from the use of R-lens, where the architecture complexity is located in the EM processing, here the complexity is entirely put into the receiving array with \( N_A = N_z \times N_y \) antennas placed at an inter-distance of \( \lambda/2 \), with \( \lambda \) indicating the wavelength.

In this case, antennas are positioned in \( p_n = (0, n_y \lambda/2, n_z \lambda/2) \), with \( n_y = \lfloor \frac{n}{N_z} \rfloor \) and \( n_z = n \mod N_z \) for \( n = 1, \ldots, N_A \), and it holds

\[
\mathcal{F}(h(x,y,p)) = h(p_n,p).
\]

(22)

In AWGN scenario, the RX signal at the \( n \)th antenna element \( r_n \) is simply given by [35]

\[
\begin{align*}
r_n &= r(p_n) = s_n + w_n = x_0 e^{-i2\pi f_0 \tau(p_n,p)} + w_n.
\end{align*}
\]

(23)

Again, here the information on \( p \) is embedded in the received signal and, by properly designing the receiver, it is possible to directly infer the position of the transmitter avoiding a prior synchronization phase to align the source and receiving array clocks.

In the following, we derive possible approaches to directly estimate the source location while using the architectures here introduced.
Fig. 5: Scenario for the nEMP, that is, in absence of EM processing. The information on the position \( p \) is retrieved from the curvature wavefront at each antenna element.

III. Position Estimation Algorithms

As previously stated, the source positioning task can be achieved by exploiting the curvature of the wavefront, by overcoming the need to tightly synchronize the transmitter and the receiver, which is often unfeasible due to the required interactions for estimating the TOA.

For the sake of simplicity, and in order to provide a solid benchmark for more practical estimators for solutions employing NR-lens or without EM processing, we consider only the line-of-sight (LOS) component, as the multi-user scenario is considered in the next section. Here we refer to a likelihood detector, where there is the maximization over the position \( p \) of the transmitter allowed by the curvature of the incoming wavefront and the unknown phase \( \chi \).

Assuming that the signals received by all the antennas are collected and post-processed together in order to estimate the distance of the source, the likelihood function related to the position \( p \) and the unknown phase \( \chi \) can be written as

\[
\Lambda(p) \propto \prod_{n=1}^{N_a} \exp \left\{ -\frac{1}{2\sigma^2} \left\| r_n - s_n(p) \right\|^2 \right\} \tag{24}
\]
where $\sigma^2 = N_0 W$, with $N_0$ representing the noise power spectral density (PSD) at each antenna, and where we have made explicit the dependence of $s_n(p)$ on the position $p$. Taking the logarithm and discarding all the terms that do not bring contribution for maximizing $p$, the log-likelihood function reduces to

$$l(p) = \sum_{n=1}^{N_A} \left\{ \Re \left\{ r_n \cdot s_n^*(p) \right\} - \frac{1}{2} E_{rx}(p) \right\}.$$  \hspace{1cm} (25)

where $E_{rx}(p) = \sum_{n=1}^{N_A} E_n(p)$, with $E_n(p) = |s_n(p)|^2$ being the received energy per antenna. Finally, the maximum likelihood (ML) estimate of the distance can be expressed as

$$\hat{p} = \arg \max_{p, \chi} [l(p)]$$  \hspace{1cm} (26)

that, in accordance with the previous derivation, yields to

$$\hat{p} = \arg \max_{p, \chi} \left\{ \sum_{n=1}^{N_A} \Re \left\{ r_n \cdot s_n^*(p) \right\} - \frac{1}{2} E_{rx}(p) \right\}.$$  \hspace{1cm} (27)

When using a traditional array, $s_n(p)$ is given by (30), whereas with a lens antenna it is expressed by the first term of (21). For the R-lens, the expression is simpler, since only one antenna is used.

A. Differential Approach for nEMP

We now propose a positioning approach that might entail a lower complexity for practical systems, and that can be applied only to the nEMP scenario.

Consider a uniform planar array, with

$$h_n = h_{n_z,n_y} \approx x_0 \exp \left\{ -j2\pi \frac{\left[ n_y^2 + n_z^2 \cos^2(\theta) \right]}{8d} \right\} \exp \left\{ -j \pi n_z \sin \theta \right\}$$  \hspace{1cm} (28)

with $n_z = 1, \ldots, N_z$, $n_y = 1, \ldots, N_y$ and $N_A = N_y N_z$. Then, consider the indices

$$\tilde{n}_y = \begin{cases} 1 & \text{if } n_y - 1 = 0 \\ n_y - 1 & \text{if } n_y - 1 > 1 \end{cases} \hspace{1cm} \tilde{n}_z = \begin{cases} 1 & \text{if } n_z - 1 = 0 \\ n_z - 1 & \text{if } n_z - 1 > 1 \end{cases}$$

so that, for $n > 1$, it is

$$h_n h_{n-1}^* = h_{n_z,n_y} h_{n_z-1,n_y}^*$$

$$= |x_0|^2 \exp \left\{ -j2\pi \frac{\left[ (n_y^2 - \tilde{n}_y^2) + (n_z^2 - \tilde{n}_z^2) \cos^2(\theta) \right]}{8d} \right\} \exp \left\{ -j \pi (n_z - \tilde{n}_z) \sin \theta \right\}$$  \hspace{1cm} (29)
with \(|x_0|^2 = A_{pl}^2\). Thus, now there is only the term with the spherical wavefront information, and by defining

\[ P_n^C = r_n^* r_{n-1} = h_n^* h_{n-1} + W_n^C = S_n^C + W_n^C, \quad n = 2, \ldots, N_A \]  
(30)

now the distance can be estimated as follows

\[ \hat{p} = \arg \max \limits_{\mathbf{p}} \left\{ \sum_{n=2}^{N_A} \left( P_n^C \cdot [S_n^C(\mathbf{p})]^* \right) - \frac{1}{2} E_{rx}(\mathbf{p}) \right\}. \]  
(31)

In this case, the ML search is drastically reduced, since it is performed only along the possible positions \( \mathbf{p} \), while the offset \( \chi \) is neglected.

IV. MULTI-SOURCE INTERFERENCE

While the previous section has highlighted the positioning scheme in a single-user scenario, real environments are usually characterized by the presence of multiple simultaneously transmitting sources, and thus it becomes important to determine the impact of the interference while discriminating two or more different transmitters.

To this purpose, let us now assume that more than one source, which is the intended useful one, is present in the same environment. The signal received from multiple sources can be written as

\[ \mathbf{r} = \mathbf{s}_u + \mathbf{s}_{int} + \mathbf{w} = \mathcal{F}(\mathbf{h}(y, z, \mathbf{p})) + \sum_{i=1}^{N_{int}} \mathcal{F}(\mathbf{h}(y, z, \mathbf{p}^{(i)})) + \mathbf{w} \]  
(32)

where \( \mathbf{s}_u \) and \( \mathbf{s}_{int} \) refer to the contribution from the intended useful source and from the interfering \( i \)th sources, respectively. As previously stated, since we do not consider here a tight synchronization between the transmitter and the receiver, signals are received with a different phase offset \( \chi \), whereas the amplitude \( A_{pl} \) is assumed to be ideally the same due to perfect control approaches. This means that \( h(y, z, \mathbf{p}) \) and \( h(y, z, \mathbf{p}^{(i)}) \) in (32) have independent \( \chi \) and the same amplitude \( A_{pl} = A \).

In the following, we assume that the phase profile of the useful signal is perfectly known, and that power control approaches are exploited so that near-far interference effects are avoided, as typically done in multiple access systems. Then, we write

\[ \gamma = A \cdot \eta(d, \theta) + A \cdot \sum_{i=1}^{N_{int}} e^{j(\chi_u - \chi_i)} \eta(d, \theta, \Delta d_i, \Delta \theta_i) + W \]  
(33)

where \( \eta(d, \theta), \eta(d, \theta, \Delta d_i, \Delta \theta_i) \) and \( W \) represent the output of the matched-filtering operation for the useful, the \( i \)th interference and the noise component, respectively, whereas \( \chi_u \) and \( \chi_i \)
denote the useful and interference offset. Note also that $\Delta d_i = d_i - d$ and $\Delta \theta_i = \theta_i - \theta$, with $d_i$ and $\theta_i$ being the distance and the AOA of the $i$th interference from the receiver, respectively. According to the considered model, the signal-to-interference ratio (SIR) can be then expressed as

$$SIR = \frac{|\eta(d, \theta)|}{\sum_{i=1}^{N_{\text{int}}} e^{j(\chi_u - \chi_i)} \eta(d, \theta, \Delta d_i, \Delta \theta_i)}.$$  \hfill (34)

In the following, considerations are drawn for each architecture.

A. SIR using R-lens

Assuming that the phase-profile of the R-lens is perfectly matched to the received signal phase, we have $\eta(d, \theta) = \mathcal{F}(h(y, z, p))|_{p=p}$ and $\eta(d, \theta, \Delta d_i, \Delta \theta_i) = \mathcal{F}(h(y, z, p^{(i)}))|_{p=p}$, so that

$$|\eta(d, \theta)| = \frac{A_{f}}{\lambda \sqrt{D_y D_z}}$$  \hfill (35)

and $\eta(d, \theta, \Delta d_i, \Delta \theta_i)$ is given by

$$\eta(d, \theta, \Delta d_i, \Delta \theta_i) = \frac{1}{\lambda \sqrt{D_y D_z}} \int_{D_y} \int_{D_z} e^{j2\pi f_0[(\tau(y, z, d, \theta) - \tau(y, z, d + \Delta d, \theta + \Delta \theta))] - \tau(y, z, d + \Delta d, \theta + \Delta \theta))] - \tau(y, z, d, \theta))} dy dz.$$  \hfill (36)

Thus, the SIR for the R-lens can be expressed as

$$SIR_{\text{R-lens}} = \frac{A_{f}}{\sum_{i=1}^{N_{\text{int}}} e^{j(\chi_u - \chi_i)} \eta(d, \theta, \Delta d_i, \Delta \theta_i)}$$  \hfill (37)

with $A_{f} = D_y D_z$.

Then, assume the presence of only one interference user, with a worst case scenario with $\chi_u = \chi_i$. Thus, for a given $p$, i.e., $d$ and $\theta$ of the transmitting source, we can evaluate the impact of the interference according to its deviation from $p$, given by the couple $(\Delta d, \Delta \theta)$. According to such a consideration, we can write

$$\eta(d, \theta, \Delta d, \Delta \theta) \propto \int_{D_y} \int_{D_z} e^{j2\pi f_0[(\tau(y, z, d, \theta) - \tau(y, z, d + \Delta d, \theta + \Delta \theta))] - \tau(y, z, d, \theta))} dy dz.$$  \hfill (38)

To this purpose, for the herein considered scenario, it holds

$$d_{0yz}g(\Theta, \Theta_{0yz}) + \frac{d_{0yz}^2}{2d} (1 - g^2(\Theta, \Theta_{0yz})) = z \sin \theta + \frac{y^2}{2d} + \frac{z^2}{2d} - \frac{z^2 \sin^2(\theta)}{2d} = z \sin \theta + \frac{y^2}{2d} + \frac{z^2 \cos^2(\theta)}{2d}$$  \hfill (39)

\footnote{For the ease of notation, in the following we neglect the factor $1/(\lambda \sqrt{D_y D_z})$.}
so that it is

\[ \eta(d, \theta, \Delta d, \Delta \theta) \propto \int_{D_z} \int_{D_y} \exp \left\{ j \frac{2\pi}{\lambda} \left( z \left[ \sin(\theta + \Delta \theta) - \sin \theta \right] + \frac{y^2}{2d} \left( \frac{1}{1 + \frac{\Delta d}{d}} - 1 \right) + \frac{z^2}{2d} \left[ \cos^2(\theta + \Delta \theta) - \cos^2 \theta \right] \right) \right\} dydz. \]

(40)

The solution of (40) does not allow to easily make considerations. A possibility is to simplify
the model is to consider only distance variations \( \Delta d \), with \( \Delta \theta = 0 \), by assuming \( \cos \theta \approx 1 \),
which is true in the boresight direction. Thus, we can write

\[ \eta(d, \theta, \Delta d) \propto \int_{D_z} \int_{D_y} \exp \left\{ j \frac{2\pi}{\lambda} \left( \frac{y^2}{2d} \left( \frac{1}{1 + \frac{\Delta d}{d}} - 1 \right) \right) \right\} dydz \]

\[ \approx \int_{D_z} \int_{D_y} \exp \left\{ j \frac{2\pi}{\lambda} \left( \frac{y^2 + z^2}{2d} \left( \frac{1}{1 + \frac{\Delta d}{d}} - 1 \right) \right) \right\} dydz \]

(41)

where, by considering polar coordinates with the approximation \( D_z D_y \approx \pi \left( \frac{D_\rho}{2} \right)^2 \) we obtain

\[ \eta(d, \theta, \Delta d) \propto \int_{D_\rho} \int_{0}^{\pi/2} \exp \left\{ j \frac{2\pi}{\lambda} \left( \frac{\rho^2}{2d} \left( \frac{1}{1 + \frac{\Delta d}{d}} - 1 \right) \right) \right\} \rho d\rho d\phi = A_f \left[ e^{-j \frac{\pi}{4} D_\rho^2 \kappa} \operatorname{sinc} \left( \frac{D_\rho^2 \kappa}{\lambda} \right) \right] \]

(42)

where

\[ \kappa = \frac{1}{2d} \left( \frac{1}{1 + \frac{\Delta d}{d}} - 1 \right) \]

(43)

with \( \rho = d_{oz} \). Notably, when \( \Delta d = 0 \), i.e., \( \kappa = 0 \), it is \( \eta(d, \theta, 0) = A_f \).

Finally, coming back to (34), we need to determine when the SIR is above a certain threshold \( \xi^* \), that is

\[ \text{SIR}_{\text{R-lens}} = \left| \frac{\eta(d, \theta, 0)}{\eta(d, \theta, \Delta d)} \right| \frac{A_f}{|\eta(d, \theta, \Delta d)|} > \xi^* \]

(44)

that implies that the following condition is satisfied

\[ \left| \frac{1}{\operatorname{sinc} \left( \frac{D_\rho^2}{2\lambda d} \left( \frac{1}{1 + \frac{\Delta d}{d}} - 1 \right) \right)} \right| > \xi^* . \]

(45)

Notably, (45) puts in evidence that when \( d \gg D_\rho \), it is difficult to discriminate two transmitters
only through the curvature, as it should hold \( \Delta d \gg d \). In addition, note that \( \text{SIR}_{\text{R-lens}} \to \infty \)
when \( \frac{D_\rho^2}{2\lambda d} \left( \frac{1}{1 + \frac{\Delta d}{d}} - 1 \right) = 1 \), that is when

\[ \Delta d \to -\frac{d}{1 + \frac{D_\rho^2}{2\lambda d}} \]

(46)

and thus the interference effect is negligible.
B. SIR using NR-lens

Again, for the NR-lens, it holds (34), where \( |\eta(d, \theta)| \) equals to the normalized aperture only when the transmitter is in the far field, and thus the curvature effect becomes negligible. Since we are interested to evaluate the spherical wavefront effect, this is not true, and should be evaluated according to the distance between the transmitter and the receiver.

According to the same aforementioned considerations, the evaluation of \( \eta(d, \theta) \) and \( \eta(d, \theta, \Delta d) \) now requires the resolution of the following equations

\[
\eta(d, \theta) \propto \sum_{n=1}^{N_A} \left\{ \left[ \int_{D_z} \int_{D_y} e^{j \frac{2\pi}{\lambda} \frac{y^2 + z^2}{2d^2 + z^2 D_z}} dydz \right] \left[ \int_{D_z} \int_{D_y} e^{j \frac{2\pi}{\lambda} \frac{y^2 + z^2}{2d^2 + z^2 D_z}} dydz \right]' \right\} \\
\eta(d, \theta, \Delta d) \propto \sum_{n=1}^{N_A} \left\{ \left[ \int_{D_z} \int_{D_y} e^{j \frac{2\pi}{\lambda} \frac{y^2 + z^2}{2d^2 + z^2 D_z}} dydz \right] \left[ \int_{D_z} \int_{D_y} e^{j \frac{2\pi}{\lambda} \frac{y^2 + z^2}{2d^2 + z^2 D_z}} dydz \right]' \right\} .
\]

In this case it is not possible to obtain simple expressions but their numerical solution allows to determine the values of the SIR\textsubscript{NR-lens} for different values of \( d, \theta, \Delta d \).

C. SIR using nEMP

In this case, by defining

\[
\eta(d, \theta, \Delta d_i, \Delta \theta_i) = \sum_{i=1}^{N_A} e^{j2\pi f_0 (\tau(p_n, p) - \tau(p_n, p^{(i)}))}
\]

we obtain

\[
\text{SIR}_{nEMP} = \frac{N_A}{\left| \sum_{i=1}^{N_{\text{int}}} e^{j(\chi_u - \chi_i)} \eta(d, \theta, \Delta d_i, \Delta \theta_i) \right|} .
\]

with \( \eta(d, \theta, \Delta d_i, \Delta \theta_i) = \sum_{i=1}^{N_A} e^{j2\pi f_0 (\tau(p_n, p) - \tau(p_n, p^{(i)}))} \). In this case, following the same considerations as before, it holds

\[
\eta(d, \theta, \Delta d) = \sum_{n=1}^{N_A} e^{j \frac{2\pi}{\lambda} \kappa d_{yz}}
\]

\[
= N_A + \sum_{n=1}^{N_A} \left( e^{j \frac{2\pi}{\lambda} \kappa d_{yz}} - 1 \right) = N_A + j 2 \sum_{n=1}^{N_A} e^{j \frac{2\pi}{\lambda} \kappa d_{yz}} \sin \left( \frac{\pi}{\lambda} \kappa d_{yz} \right)
\]

and, by making the same considerations as in (44) to the nEMP scenario, we can write

\[
\text{SIR}_{nEMP} = \frac{N_A}{N_A + j 2 \sum_{n=1}^{N_A} e^{j \frac{2\pi}{\lambda} \kappa d_{yz}} \sin \left( \frac{\pi}{\lambda} \kappa d_{yz} \right)} > \xi^* .
\]
Again, the values of $\Delta d$ satisfying the above relation represent the areas where the system is robust with respect to the interference.

In the following, the performance of the approaches herein described are compared and discussed.

V. RESULTS

In this section, numerical results are reported for the interference and the positioning performance. In particular, for both case studies, we have alternatively taken into account the three aforementioned architectures, and we compared the performance by fixing the aperture $\mathcal{A}$, thus implying the employment of a different number of antennas according to the considered scheme. In fact, as evidenced by Fig. 6 there are two competing effects. From one side, a higher complexity in the processing allows to relax the requirement on $N_A$. In fact, in the extreme scenario of the R-lens, it holds $N_A = 1$, regardless the choice of the aperture. On the other side, without any use of processing complexity, there is the need to employ a large $N_A$, as for the nEMP. In between, there is the trade-off offered by the NR-lens, where the number of antennas employed depends on the geometry, i.e., on the choice of $D_z$.

In the following, if otherwise indicated, we consider a rectangular normalized area of $\mathcal{A} = 100$, $\mathcal{A} = 150$ and $\mathcal{A} = 200$, with $D_y = 2.5$ cm in all scenarios, and $D_z = 10$ cm, $D_z = 15$ cm and $D_z = 20$ cm, respectively.
A. Positioning Results

We now evaluate the position estimation performance. More specifically, in our scenario we account only of the LOS component. Then, we consider a transmitter sending pulses centered at $f_0 = 60$ GHz, with a bandwidth $W = 2$ GHz and an effective radiated isotropic power (EIRP) of 23 dBm. At the receiver, we account for a noise figure $F = 4$ dB, and the parameter $A_{pl}$ is obtained from the link budget.

Results are expressed in terms of the root mean square error (RMSE) of the position estimate, which is evaluated as

$$\text{RMSE}(\hat{p}) = \sqrt{\frac{1}{N_c} \sum_{i=1}^{N_c} \|\hat{p}_i - p\|^2}$$

where $N_c$ is the number of Monte Carlo iterations considered in simulations and $\hat{p}_i$ is the position estimate at the $i$th iteration. For each cycle, a different noise realization is generated according to $\sigma^2$, as well as a different realization of phase $\chi$, which is kept the same for all the antennas. In this way, random phase models a complete clock mismatch between the transmitter and the receiver.

Figure [7-left reports the obtained results for the different schemes and $A$. We initially fixed the AOA to 0° by varying only the TX-RX distance from 5 m to 30 m. As evidenced by the Figure, the larger is $A_k$ the better is the position estimate thanks to the increased physical area. When a larger $N_A$ is privileged, as for the nEMP scenario, performance improves with respect to configurations where only one antenna is employed, where the complexity resides only in the reconfigurable lens. Nevertheless, the use of a R-lens allows to attain an RMSE of about 2 m for $d = 15$ m.

In between, the NR-lens represents the trade-off between the two architectures, i.e., between $N_A$ and the complexity of the lens. This effect is more pronounced for larger distances, where the path loss increases. In fact, as an example, for NR-lens with $A = 250$ and $N_A = 81$, the positioning error is kept at about 1 m for $d = 20$ m. Instead, for its counterpart without EM processing, the positioning error is lower than 1 m at $d = 20$ m, and it is about 1 m for $d = 30$ m. Finally, it is important to remark that the herein described localization performance is obtained for $A_f = (10 \times 2.5)$ cm², $A_f = (15 \times 2.5)$ cm² and $A_f = (20 \times 2.5)$ cm², which are extremely compact and thus suitable for real scenarios, e.g., for an integration in future generation of access points.
Fig. 7: RMSE as a function of the TX-RX distance \( d \), fixed AOA = 0° and different architectures (left), or for nEMP (right) when the differential approach is applied or not.

If from one side the attained performance is interesting in terms of RMSE per transmitter distance, the application of the ML here described might entail a high complexity, since it requires the search along two dimensions: the phase offset and the position. In this sense, according to the analysis reported in Sec. III-A, the differential approach allows to avoid a bi-dimensional search (i.e., over \( \chi \) and \( p \)), by keeping the same simulation scenario.

Results are reported in Fig. 7 and they are encouraging, since performance is still reliable (i.e., \( \text{RMSE}(\hat{p}) \approx 3 \text{ m at } d = 20 \text{ m for } A = 200 \)) with the advantage that the computational complexity of the positioning algorithm can be reduced.

Finally, the aforementioned considerations for the different architectures are corroborated by the coverage maps reported in Fig. 8 for different \( A \) and for the three architectures. They are obtained by placing the receiver (i.e., the green square marker) in \((0, 0)\) with a rotation of 45° towards the area, i.e., towards the red marker, and by alternatively placing the transmitting source in all the grid points of the environment. For the considered scenarios, in accordance with the results reported in Fig. 7-left, the nEMP allows to attain the best performance thanks to the larger number of antennas involved in the curvature processing.
Fig. 8: Coverage maps (errors in meters) for nEMP (left), NR-lens (middle) and R-lens (right), for $A = 100$ (top), $A = 150$ (middle) and $A = 200$ (bottom).

B. SIR Results

1) Single Interference Scenario: We first evaluate the SIR when there is another interference source located in the environment. To that purpose, we considered a room with size $(40 \times 40) \text{m}^2$, represented with a grid of points with dimension $(1 \times 1) \text{m}^2$. Then, we fixed the RX (green square marker), rotated of $45^\circ$ towards the room, in $(0, 0)$ and we located the useful TX (red square marker) in $(15, 15)$, while alternatively placing the interference in each point for computing the SIR.

In particular, we considered $\xi^* = 10 \text{dB}$, which is a typical value in multiple-access schemes, and we discriminate yellow points from the blue points if the SIR is above the threshold.
Fig. 9: SIR for a transmitter placed at $d \simeq 20$ m and one interference place in different position. Blue and yellow points denote SIR below or above $10$ dB, respectively.

In this sense, results reported in Fig. 9 show the impact of the antenna architecture, when the intended useful source is at a distance of about $20$ m (red marker) from the receiver (green marker). In particular, the interference effect is slightly more evident for the NR-lens, and it is mainly affected by the AOA, as intuitively predictable. This effect is better highlighted by Fig. 10-left, where $d$ is fixed to $20$ m, and $\Delta d$ is varied to evaluate the SIR. Indeed, despite the approximations done in (45) and (51), it is evident that when $d \gg D_z, D_y$, it is not possible to discriminate close transmitting sources. On the contrary, increasing $A_f$ while maintaining the same conditions translates into a performance improvement, as evidenced in Fig. 10-right for the nEMP.

Finally, Fig. 11 reports an example of the spatial discrimination for different TX-RX distances (diamond markers) for the nEMP (left) and the R-lens (right), when $A = 40000$, i.e., $A_f = 1 \text{ m}^2$. Notably, the nEMP is extremely robust to interference till $20$ m, whereas more oscillations are present for larger $d$. For the R-lens, only two distances are considered for the sake of clarity. Nevertheless, similar considerations as before hold.

2) Multi-Interference Scenario: We then consider a scenario with multiple interference, and the intended useful source alternatively placed in each grid point of the environment. Notably, for each TX position and for each Monte Carlo cycle, we generate interference whose position is distributed according to a Poisson point process (PPP), with intensity $\lambda_p = 5$.

To this purpose, Fig. 12 shows maps for the three architectures (from the left to the right) and for different $A$ (from top to bottom). The value of the points of the map refers to the average normalized number of times that SIR $> \xi^*$. 
Fig. 10: Interference discrimination for \( d = 20 \text{ m}, \theta = 0^\circ \) and different \( A \). Left: different architectures. Right: nEMP cas.

Fig. 11: Interference discrimination for \( A = 40000 \), i.e., \( A_f = 1 \text{ m}^2 \), when the transmitter is in different positions (diamonds), for nEMP (left) and R-lens (right).

Indeed, now the impact of \( A \) is much more evident as before, and again there is a slight improvement when solutions based on R-lens or nEMP are employed, with respect to the NR-lens. Solution with \( A = 10000 \), i.e., \( A_f = 1 \text{ m}^2 \) denote a robust interference rejection, even for distances up to 50 m.

VI. CONCLUSIONS

In this paper, the possibility to infer the transmitter position from the impinging spherical wavefront curvature at mm-wave has been investigated. This is of crucial importance whenever
the TX and the RX are not in far-field, with the advantage that they do not need any ad-hoc synchronization procedure. To this purpose, we proposed a general model that accounts for the EM processing of the wavefront curvature through a lens (if any), followed by an antenna array scheme.

Then, we analyzed the possibility to employ an EM lens that can be either reconfigurable or
not, showing how performance varies when complexity is put in the processing of the curvature or in the number of employed antennas. Additionally, we evaluated also the impact of the multiple-source interference, showing what happens both in a single interference scenario, and in a multiple-interference scenario generated through a PPP. The outcomes of this paper highlight that the considered solutions allow robust source localization using a single receiver, while guaranteeing robustness against multiple-source interference, with a certain degree of flexibility in managing the complexity either in the EM lens or in the antenna array configuration.

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