OFDM-Based Radar Network Providing Phase Coherent DOA Estimation

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Abstract—Next-generation radar sensors require imaging capabilities with high angular resolution. As for a single sensor, the aperture, and thus the achievable resolution, is limited due to the constraints of the front end, radar networks consisting of multiple sensors are a possible solution. However, their incoherency usually makes joint angle estimation impossible. This article presents a network concept consisting of an orthogonal frequency-division multiplexing (OFDM) radar and repeater elements, which receive the reflections from targets and retransmit them back to the radar. Thereby, any frequency conversion from radio frequency to baseband and vice versa is omitted such that the signal remains coherent to the initial transmit signal. To distinguish the bistatic signal transmitted by the repeater from the monostatic one of the OFDM radar, the orthogonal subcarrier structure of OFDM waveforms is exploited by combining a sparse radar transmit signal with a low-frequency modulation in the repeater. This allows to evaluate the bistatic signals at the radar with standard multiple-input–multiple-output (MIMO)-OFDM signal processing, leading to separate range–Doppler images for each virtual channel. Finally, it is shown that this method offers a coherent angular estimation based on the extended aperture of the network. For this purpose, a method to establish phase coherency by a reconstruction of the modulation phase is presented. The network concept is proved with measurements at 77 GHz.

Index Terms—Angle estimation, bistatic radar, coherency, direction of arrival (DoA), multiple-input–multiple-output (MIMO), orthogonal frequency-division multiplexing (OFDM), OFDM radar, OFDM-MIMO radar, phase reconstruction, phase recovery, radar network, repeater.

I. INTRODUCTION

THE challenge of next-generation radars for automotive, robotic, or industrial applications is to provide high-precision estimation not only for the target range and velocity but also for its angle. While high resolution in range and velocity can be achieved by a large bandwidth and a long observation time, respectively, a precise direction of arrival (DoA) estimation presupposes an extensive spatial sampling due to the Rayleigh criterion [1]. This could be achieved by a multiple-input–multiple-output (MIMO) sensor with a widely spread virtual aperture. However, most sensors are restricted in size due to practical constraints, and hence, the angular resolution is limited. In order to further increase angular resolution and to achieve an additional degree of freedom in the placement of antenna elements, multiple sensors can be combined to form a cooperative network.

Usually, radar networks consist of several radar sensors [2], [3], working cooperatively but incoherently in a network [4]–[7]. While multilateration or image fusion-based processing is still possible in incoherent networks, a phase-based DoA estimation is not. In addition, while the impact of phase noise at near ranges in monostatic radar evaluation is suppressed by range correlation [8], this is not the case for incoherent networks [9]. Phase coherency can be achieved by distributing a joint radio frequency (RF) signal to the different radar nodes [10], which results in a costly RF link, or by synchronizing the signal sources of different radar sensors [11]. However, this synchronization approach does not only result in a more complex topology by means of a distribution of the synchronization signal. For retrieving phase and frequency coherency, additionally, a calibration based on a known signal path must be performed. Furthermore, the phase noise still remains uncorrelated and reduces the performance. In [12], coherency is retrieved without a synchronization signal but with a joint digital signal processing. However, this requires a digital baseband link between the radar nodes and also suffers from uncorrelated phase noise.

In [13], a concept for a cooperative sensor network based on frequency-modulated continuous wave (FMCW) MIMO radars and repeater elements is presented. The repeater receives and retransmits the radar signal reflected at the targets and therefore provides an additional signal, which in the following is called a bistatic signal. In contrast, the radar transmit signal being reflected by the target only is called a monostatic signal. Both signals are received and evaluated at the radar receiver. The separation between these two signals is enabled by a frequency shift in the kilohertz range at the repeater. Due to the low modulation frequency, the additional phase noise introduced is much smaller than the phase noise from the radar sensor and therefore negligible. When using an FMCW radar, the modulation shifts the bistatic targets in range and velocity and separates them from the monostatic ones [13]. However, every range–Doppler (RD)-peak has to be assigned individually to the radar and the different repeater elements,
which is difficult and error-prone. This problem is increased by the fact that FMCW radars only yield a low unambiguous velocity [14], especially as the velocity measured bistatic is doubled due to the double reflection at the target.

In contrast, an orthogonal frequency-division multiplexing (OFDM) signal suits perfectly for the use in networks, as for every participant, a set of subcarriers can be assigned. Due to this subcarrier interleaving, the different signals in the network are orthogonal, which is already exploited in communication networks [15]. As OFDM is suitable for radar imaging [16]–[20], it is an appropriate choice for a radar sensor network [21]. Furthermore, compared to an FMCW radar, the digital signal generation can provide very high unambiguous velocities. However, using individual OFDM radars in a network still suffers from incoherency.

In this article, a sensor network is presented, which combines a single OFDM radar with repeater elements in a sensor network. By assigning different subcarriers to the monostatic and bistatic signals, it is possible to distinguish among them unambiguously. This is possible by combining frequency shifts of integer subcarrier spacing and a sparse radar transmit signal with unassigned subcarriers. The orthogonal structure of OFDM can be preserved, and a very straightforward signal separation of monostatic and bistatic signals is achieved. Hence, all the information is available at the radar and all the signals can be processed independently due to their orthogonality. In order to exploit the possibilities of such a network in terms of DoA estimation, a fully coherent signal processing is presented. While the bistatic signals are coherent, they suffer from an unknown phase offset caused by the phase of the modulation signal. In this work, a reconstruction of the unknown modulation phase of each repeater is performed. In contrast to the signal processing proposed in [13], this provides the possibility to evaluate the path differences between the repeaters. In this way, a coherent phase progression similar to the steering vector of a MIMO radar can be generated by combining the phases of the bistatic signals. This network steering vector thus is based on the sparse but widespread virtual array of the network. Hereby, a high-resolution DoA estimation is realized.

In order to introduce the OFDM-based radar-repeater network used in this work, its concept is described in Section II. This includes the generation of an appropriate radar transmit signal and its modulation by the repeater. The description of the signal processing and especially its adaptation to the network is discussed in Section III. In Section IV, a new concept for a repeater-based DoA estimation that exploits the path difference between the repeater elements is introduced. Simulation and measurement results are presented in Sections V and VI, respectively.

II. CONCEPT OF THE SENSOR NETWORK

In the radar network used in this work, an OFDM radar is supplemented by repeaters in order to gain additional target information. Each repeater consists of antennas, amplifiers, and a modulator [22] and provides additional signals, as it receives and retransmits the radar signal being reflected by the targets. In the following, these signals are called bistatic, while we name the classical radar measurement with a signal radiated by the radar transmitter, reflected by the target and received by the radar receiver monostatic. Due to the doubled path, the range dependency of the attenuation of the bistatic signal is squared compared with the attenuation of the monostatic signal. This makes the amplification in the repeater crucial. Both the monostatic and the bistatic signals are processed at the radar receiver only. The signal paths occurring for the simple case of one repeater and one target are exemplarily shown in Fig. 1.

Precondition for exploiting the information gained by the bistatic signals is the capability to distinguish between the monostatic and the bistatic signals at the radar receiver, which is done by assigning different subcarriers to the monostatic and the bistatic signal. In Fig. 1, the subcarrier assignment is shown exemplarily for a network with a single repeater, where every second subcarrier is assigned to the bistatic signal.

This subcarrier interleaving is described in Section II-B. It requires the adaption of the transmit (Tx) signal of the OFDM radar presented in Section II-C, considering the modulation of the signal by the repeater described in Section II-D. To begin with, in Section II-A, the basic modulation of an OFDM signal is explained.

A. OFDM Modulation

The basic principle of OFDM modulation is the orthogonality of different subcarriers given by the relation between the length of one OFDM symbol $T$ and the subcarrier spacing $\Delta f = 1/T$ [23]. Each OFDM symbol can be described by the sum over $N$ subcarriers, each at a frequency $f_n = n \Delta f$, $n = 0, 1, \ldots, N − 1$ and modulated by a complex symbol $d_{Tx}(n) \in \mathbb{C}$. In addition, a cyclic prefix of length $T_{CP}$ is added before each OFDM symbol, resulting in a transmission time of

$$T_{OFDM} = T + T_{CP}$$

per OFDM symbol.
Furthermore, an OFDM signal consists of $M$ successively transmitted OFDM symbols denoted by $m = 0, 1, \ldots, M - 1$. This combination of $M$ subsequent OFDM symbols and their prefixes, each with $N$ subcarriers, can be described as

$$x(t) = \sum_{m=0}^{M-1} \sum_{n=0}^{N-1} d_{Tx}(n, m) \exp\left[j2\pi f_n (t - mT_{OFDM} - T_{CP})\right]$$

$$\times \text{rect}\left(\frac{t - (m + \frac{1}{2})T_{OFDM}}{T_{OFDM}}\right)$$

and equals the baseband representation of the OFDM Tx signal [21], [24]. The rectangular function represents the time frame of each OFDM symbol. The modulation symbols $d_{Tx}(n, m)$ can be notated as matrix $D_{Tx} \in \mathbb{C}^{N \times M}$. The modulation scheme described here is also called cyclic-prefix-based OFDM (CP-OFDM).

B. Subcarrier Interleaving

In order to enable the separation of the monostatic and bistatic signals at the radar receiver, the Tx signal of the radar and the modulation in the repeater have to be chosen such that the monostatic and the bistatic receive signals, $y_{\text{mono}}(t)$ and $y_{\text{bi}}(t)$, as well as the bistatic signals from different repeaters are orthogonal. As every element of the symbol matrix represents a time and frequency slot of the OFDM signal, two signals using the same subcarrier spacing and carrier frequency are orthogonal if they do not use the same slots. Therefore, the transmit matrix of the radar $D_{Tx}$ and the symbol matrices of the modulated signals of different repeaters, $D_{bi,p}$ and $D_{bi,q}$ with $p, q = 0, 1, \ldots, N_{Rp} - 1$, have to be chosen in a way that

$$D_{Tx} \odot D_{bi,p} = 0$$

$$D_{bi,p} \odot D_{bi,q} = 0, \quad q \neq p$$

applies. The operator $\odot$ represents an elementwise multiplication, and $N_{Rp}$ is the number of repeaters in the network. In order to satisfy this condition, the different signals are multiplexed to different orthogonal subcarriers, like it is exemplary shown in the signal spectra in Fig. 1. Thus, the modulation frequencies of the repeater and the transmit matrix $D_{Tx}$ have to be chosen appropriately. In the simple case of a network with one repeater with single-sideband (SSB) mixer, as shown in Fig. 1, every second subcarrier of the radar Tx signal has to be unassigned. Thus, the elements of the matrix $D_{Tx}$ are constructed by the complex symbols $s(n, m) \in \mathbb{C}$ by

$$d_{Tx}(n, m) = \begin{cases} s(n, m), & n \text{ even} \\ 0, & n \text{ odd} \end{cases}$$

In this case, the repeater has to modulate the signal by $f_{mod} = \Delta f$, which results in a shift by one subcarrier, as it is shown in Fig. 1. Usually, the interleaving scheme is more complex. For this reason, in Sections II-C and II-D the transmit signal and the modulation by the repeater are described for the general case.

C. Modification of the Radar Transmit Signal

In general, every subcarrier, in which the bistatic signal of one of the $N_{Rp}$ repeaters is modulated on, has to be unused by the radar transmitter. For the reconstruction of the modulation phase for DoA estimation, every repeater needs to modulate the signal on two different subcarriers, as shown later. This can be realized, e.g., by a double-sideband (DSB) repeater. In addition, depending on the interleaving scheme, there may be $N_{c}$ carriers per Tx carrier which are neither occupied by the Tx signal nor by a bistatic signal.

Therefore, only every $G = (2N_{Rp} + 1 + N_{c})$th subcarrier is used for the radar Tx signal. The elements of $D_{Tx}$ are created with the complex symbols $s(n, m)$ by

$$d_{Tx}(n, m) = \begin{cases} s(n, m), & \text{mod}(n - n_{0}, G) = 0 \\ 0, & \text{otherwise} \end{cases}$$

with the $n_{0}$th carrier being the first one used in one radar Tx signal. For the additional interleaving of the different radar Tx channels, the subcarriers reserved to the radar Tx signal and occupied in $D_{Tx}$ have to be assigned to the different radar transmitters as described in [21], [25]. Without loss of generality, in this work, the signal processing is described without explicitly considering the different Tx channels.

D. Signal Modulation by the Repeater

To retain the orthogonal structure of the OFDM signals, in general, the modulation frequency

$$f_{mod} = h \Delta f, \quad h \in \mathbb{Z} \setminus \{0\}$$

of the repeater is chosen as a multiple of the carrier spacing $\Delta f$ with the integer factor $h$. This factor is named modulation factor in the following. For the reconstruction of the modulation phase (see Section IV-C), every repeater has to transmit the signal modulated to two different frequencies $f_{mod, I, p}$ and $f_{mod, II, p}$ with the modulation factors $h_{I, p}$ and $h_{II, p}$. The modulation factors of the different repeaters and the transmit signal of the radar have to be chosen in a way that (3) and (4) are fulfilled.

The repeater’s Tx signal $x_{Rp}(t)$ is then the modulated receive (Rx) signal $y_{Rp}(t)$ of the repeater and given by

$$x_{Rp}(t) = y_{Rp}(t) a_{Rp} \left( \exp\left[j(2\pi h_{I, p} \Delta f t + \phi_{mod, I, p})\right] + \exp\left[j(2\pi h_{II, p} \Delta f t + \phi_{mod, II, p})\right] \right).$$

The factor $a_{Rp} \in \mathbb{C}$ considers the static phase modulation and the gain of the repeater and the phases $\phi_{mod, I}$ and $\phi_{mod, II}$ equal the phases of the two modulation signals at $t = 0\,\text{s}$, which is the start time of the OFDM signal at the radar transmitter.

The modified Tx signal and the carrier shift by the repeaters are shown in Fig. 2 exemplarily for a network with $N_{Rp} = 2$ repeaters. Fig. 2(a) shows the subcarrier assignment for repeaters with a DSB mixer, where the lower sideband leads to a second modulation with factor $h_{II, p} = -h_{I, p}$. Fig. 2(b) shows the subcarrier assignment in a network with SSB repeaters, which modulate the signal also to the doubled modulation frequency. This leads to a second modulation with $h_{II, p} = 2h_{I, p}$. Any frequency error in the modulation signal will be interpreted as Doppler shift and thus will cause an
After eliminating the influence of the known transmit symbols \( d_{\text{Tx}}(n, m) \), range and velocity processing can be performed by columnwise and rowwise Fourier transforms, respectively [24].

In the bistatic case, the signal passes the targets twice. While in this work, the distance between radar and target is assumed to be approximately the same as the distance between the repeater and the target, the relative velocities may differ between the repeaters’ perspectives and the radar’s one. The relevance of this effect depends on the relation between the target’s distance \( R \) and the distance between radar and repeater and is explained in detail in [26], where the network is exploited for an estimation of the velocity vector. In this article, we consider this by introducing the bistatic Doppler shift \( f_{D,\text{bi},k} \). The received signal then contains a linear combination of all possible propagation paths to and from the repeater. Analogous to (11), this leads to

\[
d_{\text{Rx,bi}}(n + h, m) = d_{\text{Tx}}(n, m) \sum_{k=0}^{N_t-1} \sum_{l=0}^{N_r-1} a_k(n, m) a_l(n + h, m) \times \exp[-j2\pi(\varphi_k(n) + \varphi_l(n + h))] \times \exp[j2\pi(f_{D,\text{bi},k} + f_{D,\text{bi},l})m(T + T_{CP})]
\]

(12)

\( k \) denoting the target, where the signal is reflected on the way to the repeater, and \( l \) the target on the way back. On the way back, the signal is shifted by \( h \) subcarriers. These receive symbols as well as the following steps are described generally for the modulation factor \( h \), which can be either \( h_{\text{ill},p} \) or \( h_{\text{all},p} \) of any repeater, depending on the signal to be processed. Thus, at first, the different monostatic and bistatic signals have to be separated. This is explained in Section III-A. Then, the monostatic signal can be processed with the classical signal processing for OFDM radars. This classical OFDM processing is shown in Fig. 3(b) and is explained in detail in [21] and [24]. For the bistatic signals, this signal processing has to be modified. This adapted processing is shown in Fig. 3(a) as a flowchart and is explained in this section. In Section III-B, the correction of the phase error introduced during the cyclic prefix is explained, while in Section III-C, the range processing used for the bistatic signals is presented. Furthermore, in Section III-D, the interpretation of the bistatic \( R_0 \)-plot is discussed.

### A. Signal Separation

After the transformation of the receive signal into the symbol domain by a fast Fourier transform (FFT), one single symbol matrix \( D_{\text{Rx}} \) is available, containing the information of every monostatic and bistatic signal. For further processing, each signal has to be separated in an individual symbol matrix. This is done when eliminating the corresponding transmit symbols. For the monostatic signal, this can be done by an elementwise multiplication with the complex conjugate of the transmit matrix \( D_{\text{Tx}}^* \). Thus, every element from \( D_{\text{Rx}} \) not assigned to the monostatic signal is multiplied by zero. For the bistatic signals, a matrix \( D_{\text{syn}}(h) \) has to be created, which equals \( D_{\text{Tx}} \) shifted by \( h \) in subcarrier direction, thus considering the modulation by the repeater. For the monostatic

\[
d_{\text{Rx,mono}}(n, m) = d_{\text{Tx}}(n, m) \sum_{k=0}^{N_t-1} a_k(n, m) \exp[-j2\pi \varphi_k(n)] \times \exp[2\pi f_{D,k}m(T + T_{CP})].
\]

(11)
This results in a constant phase shift of \( \Delta \varphi_{\text{mod}}(h) = 2\pi f_{\text{mod}} T_{\text{CP}} \) (14) from one OFDM symbol to another. Thus, without correction, \( d_{\text{Rx},\text{bi}}(n + h, m) \) from (12) will be distracted by another symbol-dependent exponential term, leading to
\[
d_{\text{Rx},\text{bi}}(n + h, m) = d_{\text{Rx},\text{bi}}(n + h, m) e^{i\Delta \varphi_{\text{mod}}(h)(m+1)}. \tag{15}\]

Without correction, this will lead to a velocity offset of
\[
\nu_{\text{rel},\text{err}}(h) = \frac{\Delta \varphi_{\text{mod}}(h)}{2\pi} \nu_{\text{rel,u}} = h \frac{T_{\text{CP}}}{v_{\text{rel,u}}} \tag{16}\]
with \( \nu_{\text{rel,u}} \) being the unambiguous velocity. As \( \Delta \varphi_{\text{mod}}(h) \) is known, this effect can be corrected directly in \( D_{\text{Div}}(h) \) with a correction vector
\[
\kappa(h) = \left( e^{-i\Delta \varphi_{\text{mod}}(h)}, e^{-i2\Delta \varphi_{\text{mod}}(h)}, \ldots, e^{-iM\Delta \varphi_{\text{mod}}(h)} \right) \quad \kappa(h) \in \mathbb{C}^M \tag{17}\]
by multiplying every \( m \)th column of \( D_{\text{Div}}(h) \) with the \( m \)th element of \( \kappa(h) \).

C. Range Processing with Carrier Correction

When the signal is retransmitted by the repeater, the subcarrier has been shifted by \( h \Delta f \). Due to the subcarrier-dependent phase term from (10), this introduces a static phase offset. In the case of bistatic signals being reflected at the same target on the way to the repeater and back \( (k = l) \), the phase term is given by
\[
\varphi_k(n) + \varphi_k(n + h) = \frac{2R_k}{c_0} \left( f_n + f_{n+h} \right) = \frac{2R_k}{c_0} \left( 2f_n + h \Delta f \right) \tag{18}\]
and thus does not equal \( 2\varphi_k(n) \). The phase error depends on the modulation factor \( h \) and amounts to
\[
\phi_{\text{err}} = \varphi_k(n + h) - \varphi_k(n) = \frac{2R_k}{c_0} h \Delta f. \tag{19}\]

If this phase error would not be corrected, the path difference between different repeaters could not be extracted, and no network-based DoA estimation is possible. In order to eliminate this phase error, a carrier correction is integrated in the range processing. As shown in Fig. 3, this range processing is performed after the velocity processing, which is done by means of a rowwise FFT [19], [21], [24]. The matrix after velocity processing, \( D_{f,v} \), then has to be shifted by \(-0.5\ h \) in subcarrier direction. For an even \( h \), this is trivial, but for an odd \( h \), a new matrix \( D_{f,v}' \in \mathbb{C}^{2N \times M} \) has to be created with the elements
\[
d_{f,v}'(\tilde{n}, m) = \begin{cases} 
\frac{d_{f,v}(\tilde{n}/2, m)}{2}, & \tilde{n} \text{ even} \\
0, & \text{otherwise} 
\end{cases} \quad \text{with } \tilde{n} = 0, 1, \ldots, 2N - 1. \tag{20}\]

Thus, the size of the matrix is doubled by inserting an empty row between every two rows of \( D_{f,v} \). The new matrix \( D_{f,v}' \) must now be shifted by \(-h \) before doing the inverse fast Fourier transform (IFFT) for range processing. In this way, the modulation factor \( h \) is considered in the range processing, and thus no phase error occurs.
D. Interpretation of the Bistatic Radar Image

By means of the signal processing described in this work, two complex radar images are generated for every repeater: one for $h_{l,p}$ and one for $h_{ll,p}$. The resulting bistatic $R_l$-plots are based on the bistatic receive signals. When interpreting those bistatic $R_l$-plots, the different paths occurring due to the double reflection at the targets have to be considered.

On the one hand, there are paths $s_{k,i}$ resulting from reflections at the same target on the way to the repeater and back to the radar. In the following, we name them single-target paths. On the other hand, there are the paths $s_{k,l}$, $k \neq l$ with the reflections on different targets on the way to the repeater and back, named multitarget paths. Thus, analogous to (12), the bistatic signals are based on a linear combination of all the targets. However, since the paths $s_{k,l}$ and $s_{i,k}$ are resulting in the same range and the same relative velocity, only local maxima appear in the radar image. For further processing, e.g., for DoA estimation, the single-target peaks must be selected.

IV. REPEATER-BASED DOA ESTIMATION

With the extended aperture of the network, a network-based DoA estimation can provide a much higher resolution than the single MIMO-OFDM radar. The challenge is to combine the phase information gained by different network elements in one single steering vector with a consistent phase progression.

In the following, the phase of the bistatic signal will be examined in detail. At first, in Section IV-A the virtual array of the network is explained, followed by a method for phase normalization in Section IV-B and a reconstruction of the modulation phase in Section IV-C, yielding a steering vector that can be used for DoA estimation, as described in Section IV-D.

A. Virtual Network Array

In the case of an MIMO sensor, the virtual array is based on the combination of the phase progression from both the Tx and the Rx antenna positions [27] and can be calculated by a convolution of the Tx with the Rx array. With $z_0 \ldots z_{N_{ant}-1}$ being the positions of the antennas in the virtual array of the MIMO sensor, $\lambda$ the signal wavelength, and $\theta$ the target’s angle with $\theta = 0^\circ$ corresponding to boresight, the steering vector of this single sensor can be calculated as

$$a_\theta = \left( \exp \left[ j 2 \pi \frac{z_0}{\lambda} \sin(\theta) \right], \ldots, \exp \left[ j 2 \pi \frac{z_{N_{ant}-1}}{\lambda} \sin(\theta) \right] \right).$$

(22)

In the network, every bistatic signal contains the signals from the different radar Tx channels and it is received at every radar Rx channel. Thus, for every repeater, the MIMO steering vector will be repeated with an additional phase term, depending on the repeater position $z_{Rp}$. Assuming the far-field condition, the steering vector based on the bistatic signal of the $p$th repeater can be calculated as

$$a_{bi,p} = a \exp \left[ j 2 \pi \frac{z_{Rp,p}}{\lambda} \sin(\theta) \right].$$

(23)

Hence, the bistatic signals give a repetition of the monostatic steering vector $a$ shifted by $\frac{\pi z_{Rp,p}}{\lambda}$ to a different virtual position. It should be noted that the virtual position does not equal the physical position but is determined by doubling the physical distance to the reference antenna. Similar to an MIMO virtual array, the virtual network array can be calculated by a convolution of the MIMO virtual array $r_{\text{virt,radar}}$ and a vector $\bar{r}_{\text{network}}$, indicating the virtual positions of the network elements with respect to the doubled distance $\bar{z}_{Rp,p}$.

Therefore, the virtual network array is given by

$$r_{\text{virt, network}} = r_{\text{virt, radar}} \ast \bar{r}_{\text{network}}$$

(24)

where the operator $\ast$ represents a convolution. The generation of the virtual network array is graphically shown in Fig. 4. In a practical sense, it has to be considered that the placement of the repeater relative to the reference repeater is critical, as any error directly influences $\frac{z_{Rp,p}}{\lambda}$ and, thus, the angle estimation. Due to the doubled virtual distance $\bar{z}_{Rp,p}$, this virtual array usually contains a big gap between the network elements. This has to be considered when choosing repeater positions.

B. Phase Normalization

Based on the channels of the MIMO radar and the independent processing of the bistatic signals as described in Section III, for every $R_l$-peak, a bistatic phase progression $i_p \in \mathbb{C}^{N_{virt}}$ can be obtained. $N_{virt}$ is the number of elements in the virtual array of the radar. This phase progression contains the steering vector $a_{bi,p}$ from (23) and an additional phase term $\phi_{Rp,p}$ and is given by

$$i_p = e^{j \phi_{Rp,p}} a_{bi,p}.$$  

(25)

The additional phase term is static for one measurement but differs among the repeater elements. It has to be eliminated in
order to extract $a_{bi,p}$. It corresponds to

$$\phi_{R_p,p} = \phi_0 + 2 \left( \frac{2 R}{c_0} f_c + \phi_{refl} \right) + \phi_{HW} + \phi_{mod,p} \quad (26)$$

and includes the phase offset $\phi_0$ between Tx and Rx of the MIMO sensor and the phase of the modulation signal $\phi_{mod,p}$. The phase offset $\phi_{refl}$ due to the hardware and signal delay of the repeater can be calibrated and is therefore neglected. Furthermore, (26) contains a phase term dependent on the range and the carrier frequency $f_c$ as well as the phase jump $\phi_{refl}$ due to the reflection at the target. Both are doubled due to the doubled path to and from the repeater.

In order to extract $a_{bi,p}$, a normalization to a reference element can be performed. As the range-dependent phase term is doubled compared with the monostatic measurement, it is not possible to normalize the bistatic phases to the phases of the monostatic signal. However, as apart from $\phi_{mod,p}$, the static phase $\phi_{R_p,p}$ is the same for every repeater, and the bistatic phases can be normalized to the phase of one reference repeater $p = 0$. When generating the steering vector based on the signal modulated by $h_{1,p}$ for each repeater, it can be extracted by

$$a_{bi,p} = \frac{i_{1,p}}{e^{j\phi_{mod1,p}}} \frac{e^{j\phi_{mod1,0}}}{i_{1,0}(0)}. \quad (27)$$

As $\phi_{mod1,p}$ cannot be eliminated by normalization, each $i_{1,p}$ must be divided by the corresponding $e^{j\phi_{mod1,p}}$ to generate the steering vector. Thus, a proper reconstruction of the modulation phase $\phi_{mod1,p}$ of each repeater is necessary. Then, the steering vectors of every repeater may be combined into a network steering vector according to the network virtual array.

C. Reconstruction of the Modulation Phase

In the case of the radar-repeater network, the signals are coherent, but the modulation signal $y_{mod}(t)$ causes an unknown phase offset. This phase offset

$$\phi_{mod} := \arg(y_{mod}(t = 0 s)) \quad (28)$$

equals the phase of the modulation signal at $t = 0$ s, which is the starting time of the OFDM signal at the radar transmitter. This phase offset can also be described based on $t_{mod0}$, which is the point in time closest to $t = 0$ s where $\arg(y_{mod}(t)) = 0$.

The basic concept of the reconstruction of $\phi_{mod1,p}$ shown in Fig. 5 is to generate two bistatic signals by the same repeater in a way that the phase difference of those signals only depends on the same $\phi_{mod1}$ but with a different factor. Thus, every repeater has to transmit two signals, both based on the same repeater receive signal but modulated with different frequencies. These frequencies can be described by two different modulation factors $h_1$ and $h_{II}$. The ratio of these modulation factors equals the ratio of the modulation phases and can be described as

$$b = \frac{h_{II}}{h_1} = \frac{\phi_{mod,II}}{\phi_{mod,1}}. \quad (29)$$

The equivalence of the ratios is due the same time offset $t_{mod0}$ of $y_{mod,1}(t)$ and $y_{mod,II}(t)$. Therefore, the modulation signal of the $p$th repeater equals

$$y_{mod,p}(t) = \exp[j2\pi h_{1,p} \Delta f (t - t_{mod0,p})] \quad := y_{mod,1,p}$$

$$+ \exp[j2\pi b h_{II,p} \Delta f (t - t_{mod0,p})] \quad := y_{mod,II,p}. \quad (30)$$

Due to the different modulation factors, the two receive signals of the radar $y_{Rx,bi,1,p}(t)$ and $y_{Rx,bi,II,p}(t)$ can be processed independently. Analogous to (26), the repeater phase corresponds to

$$\phi_{R_p,1,p} = \phi_x + \phi_{mod,1,p} \quad (31)$$

$$\phi_{R_p,II,p} = \phi_x + \phi_{mod,II,p}$$
$$= \phi_x + b \phi_{mod,1,p}. \quad (32)$$

By using the differences of these two phases, the modulation phase can be determined by

$$\phi_{mod,1,p} = \frac{\phi_{R_p,II,p} - \phi_{R_p,1,p}}{b - 1}. \quad (33)$$

In this way, the phase can be reconstructed without knowledge of the channel-dependent phase term $\phi_x$. It has to be considered that an unambiguous solution only exists for $|b - 1| = 1$. This problem exists due to the cyclic characteristic of the phase and becomes even more obvious in complex notation with

$$e^{j\phi_{mod1,p}} = \frac{b^{-1}i_{II,p}(u)}{i_{1,p}(u)}. \quad (34)$$

where the $(b - 1)$th complex root has to be calculated. The phase reconstruction can be performed based on the phase of every antenna position in the repeater’s virtual array denoted by $u = 0, 1, \ldots, N_{ant} - 1$, and the results from the different positions may be averaged. By inserting (34) into (27), a consistent steering vector can be calculated.

Fig. 5. Concept of the modulation phase reconstruction, exemplary for a repeater, which additionally modulates to the doubled modulation frequency. This results in $b = 2$. 

The opposite phase of the modulation signal of the $p$th repeater equals

$$\phi_{mod0} = \phi_{mod0,p} = \phi_x + 2 \phi_{mod,1,p} + \phi_{refl}.$$
TABLE I
RADAR PARAMETERS

| Parameter                        | Value  |
|----------------------------------|--------|
| Carrier frequency \( f_c \)      | 77 GHz |
| Subcarrier spacing \( \Delta f \)| 200 kHz|
| Bandwidth \( B \)                | 0.4 GHz|
| Number of subcarriers \( N \)    | 2000   |
| Number of OFDM symbols \( M \)   | 1024   |
| Symbol time \( T_{\text{sym}} \)  | 5.0 \( \mu \text{s} \) |
| Prefix time \( T_e \)            | 0.5 \( \mu \text{s} \) |
| MIMO configuration               | \( 4 \times 4 \) |
| Virtual antenna spacing \( \lambda/2 \) |        |
| Range resolution                 | 0.37 m |
| Velocity resolution              | 0.35 m/s|
| Unambiguous range                | 37.5 m |
| Unambiguous velocity             | 177 m/s|
| Repeater interleaving sparsity factor \( G \) | 5 |
| Modulation frequency repeater 0 \( f_{\text{mod},0} \) | 400 kHz|
| Modulation frequency repeater 1 \( f_{\text{mod},1} \) | 200 kHz|

D. DOA Estimation with Compressed Sensing

The distance between the repeaters leads to a gap in the virtual network array. Depending on the DoA approach, this gap results in ambiguities and artifacts. One solution, which is also used for sparse antenna arrays of MIMO radars, is the use of angular estimation based on the Fourier transform \[28\] combined with compressed sensing (CS) for sparsity reconstruction \[29\], \[30\]. Due to the special sampling structure with a repetition of the virtual array at every repeater and thus several rectangular functions, zero padding is crucial. As for an iterative method with adaptive thresholding (IMAT) algorithm \[31\], the number and broadness of peaks in the angular domain may be unknown, this CS algorithm is well-suited for zero-padded signals. Thus, the DoA estimation is performed based on the Fourier transform and the IMAT algorithm.

V. SIMULATIONS

In this work, simulations of a network with a \( 4 \times 4 \) MIMO-OFDM radar with parameters as presented in Table I and two repeaters are shown. The simulations are performed in complex baseband. In Section V-A, the simulation results in the \( R_v \)-domain are presented, whereas the results of the angle estimation are discussed in in Section V-B.

A. Simulation Results in the \( R_v \)-Domain

For the evaluation in the \( R_v \)-domain, a scenario with two targets is simulated. Target 1 is at \( R_1 = 4 \text{ m} \) and \( v_{\text{rel},1} = 0 \text{ m/s} \), and target 2 is at \( R_2 = 7 \text{ m} \) and \( v_{\text{rel},2} = -10 \text{ m/s} \). The results of the simulations using CP-OFDM are shown in Fig. 6. In Fig. 6(a), the \( R_v \)-plot of the monostatic signal is shown, whereas Fig. 6(b) shows the bistatic evaluation based on the signal of one repeater. Every \( R_v \)-plot is normalized to its maximum. The higher noise floor in the bistatic plot is due to the lower Rx power of the bistatic signal. In the bistatic plot, the distance shown at the \( R \)-axis is half of the length of the overall traveled signal path, as it is in the monostatic evaluation. Thus, both targets have the double distance and approximately the double velocity, and an additional third peak occurs due to the multtarget propagation. Furthermore, a leakage of the monostatic peak of the nonstatic target occurs in Fig. 6(b). This is caused by the loss of orthogonality between the subcarriers in the case of a Doppler shift. The shift in \( v_{\text{rel}} \)-direction of this leakage peak exists due to the cyclic prefix correction described in Section III-B. However, since the position of this leakage peak is well known from the monostatic \( R_v \)-evaluation and its known velocity shift, the leakage does not reduce the ambiguity-free zone of the \( R_v \)-plot. A leakage from one bistatic signal to another or to the monostatic signal may also occur but due to the lower Rx power of the bistatic signal it is less relevant.
In order to suppress the leakage, repeat OFDM (RP-OFDM) with Doppler correction can be used [32]. RP-OFDM is an OFDM scheme, where the same OFDM symbol is repeated $M$ times, and no cyclic prefix has to be introduced. It can be adapted to the network analogously to CP-OFDM. With the Doppler correction, the power of the leakage is spread in velocity direction, creating a ridge instead of a peak. This is shown in Fig. 7. The results of a signal processing with Doppler correction are equivalent to the results in Fig. 6 but with suppressed leakage.

**B. DOA Estimation**

For the DOA estimation, simulations with a network consisting of two repeaters with a spacing of $z_{Rp,1} = 32 \lambda / 2$ and a $4 \times 4$ MIMO with 16 virtual elements in $\lambda / 2$-spacing and an aperture size of $15 \lambda / 2$ were performed. This results in a virtual network array with a size of $79 \lambda / 2$, as shown in Fig. 8. As modulation scheme, CP-OFDM is used.

SSB repeaters that modulate the signal also to the double modulation frequency are simulated. This results in a carrier assignment, as shown in Fig. 2(b). One repeater modulates the signal by $1 \cdot \Delta f$ and $2 \cdot \Delta f$, whereas the other repeater modulates by $3 \cdot \Delta f$ and $6 \cdot \Delta f$. In this way, $h_{II,p}$ equals $2h_{I,p}$ and an unambiguous reconstruction of $\phi_{mod}$ is possible. For proper interleaving, the number of subcarriers is increased to 2016.

Fig. 9 shows the results of the DOA estimation for a simulation with two targets, one at $0^\circ$ and $-5.2^\circ$.

**VI. MEASUREMENTS**

For measurements, an MIMO-OFDM radar and two repeater elements were used. The radar has four Tx and four Rx channels, and its virtual array is forming a uniform linear array (ULA) with 16 elements in $\lambda / 2$-spacing. The baseband signals are generated and stored using a Xilinx RFSoC [33], while the design of the Tx signal and the signal processing is done offline for maximum flexibility and accuracy. The radar parameters are presented in Table I.

The repeater elements are each build by two amplifiers and one mixer, thus being similar to the repeater from [22]. As no suitable SSB mixers were available, DSB mixers were used, leading to $b = -1$ and thus an ambiguous phase reconstruction. The modulation frequencies are 200 and 400 kHz, equaling $1 \cdot \Delta f$ and $2 \cdot \Delta f$, respectively. The measurement results in the $R_v$-domain are presented in Section VI-A, followed by the results of the angle estimation in Section VI-B.

**A. Measurement Results in the $R_v$-Domain**

For measurements, in the $R_v$-domain, cylindrical targets were used. The first one, a static target, was placed at a distance of 1.3 m, whereas the second one was moved manually at approximately the same distance.

In Fig. 10, the monostatic and bistatic $R_v$-diagrams are shown. For the evaluation in the $R_v$-domain, the results of the different radar channels were summed up noncoherently. In the monostatic measurement, both targets can be detected at a distance of about 1.1 m and with a speed of 0 and 3 m/s, respectively. Furthermore, as the repeaters used
in this experimental setup have no carrier suppression, also the bistatic targets are visible in the monostatic \( R_v \)-diagram.

In the bistatic \( R_v \)-diagram, both targets can be detected at double the distance of the corresponding monostatic peaks. This is due to the double path length. The velocity of target 1 is 0 m/s in both cases, whereas target 2 is measured at 2.9-m/s monostatic and 6.0-m/s bistatic.

### B. DOA Estimation

For the DoA estimation, the 4 \( \times \) 4 MIMO radar is extended by two repeaters, which are placed 32 \( \cdot \lambda /2 \) apart from each other. The measurement setup is shown in Fig. 11, and the resulting virtual network array is identical to the one used in the simulations and shown in Fig. 8.

Due to the limitations in the experimental setup, especially in terms of repeater gain, no bistatic measurements at larger ranges are possible and the DoA estimation may be affected by effects of near field with respect to the large aperture of the network. Furthermore, due to the operation in near field, cylindrical targets were used. Their diameter is 10 cm.

For DoA estimation, a proper calibration of the radar front end and the repeaters is crucial. In order to be able to calibrate the radar with a point target, a two-step calibration was performed.

First, the MIMO sensor was calibrated based on a corner reflector, which was placed at 0\(^\circ\) at a distance of 2.1 m. For the calibration of the repeater, a second calibration with a cylindrical target placed at 0\(^\circ\) at a distance of 1.4 m (target 1 in Fig. 11) was used to determine the hardware-based path difference between the repeaters by averaging the differences between the bistatic signals of all MIMO channels.

Fig. 12 shows a measurement of target 2 at -5.2° ground truth. In the measurement, the peak is at -6.8°. As a reference, the monostatic DoA estimation based only on the MIMO radar is shown. The monostatic steering vector is zero padded to match the length of the zero-padded bistatic steering vector and evaluated by a Fourier transform. The deviation from ground truth probably is caused by near-field effects and a limited accuracy in the measurement setup.

The results in Fig. 13 show a scenario with both targets from Fig. 11. In the monostatic evaluation based on the MIMO array of the radar, it is not possible to separate the two targets, as the two broad peaks merge. In contrast, the bistatic evaluation with CS shows two clearly distinguishable peaks at -6.7° and -0.6°.

### VII. CONCLUSION

Based on an OFDM radar and repeater elements, a coherent network design not suffering from additional phase noise or intercarrier interference is presented. Using the combination
of a sparse OFDM signal with a frequency shift to unassigned subcarriers by the repeater, the different signals in the network can be separated unambiguously. In this way, each network element can provide an independent radar image. The influence of the modulation from the repeater on the bistatic radar image is analyzed and can be eliminated with the adapted OFDM processing presented in this work.

With a reconstruction of the unknown modulation phase, a fully coherent angle estimation based on the evaluation of the different bistatic signals can be performed. In this way, a high-precision DoA estimation exploiting the extended virtual array of the network is achieved. Signal processing and DoA estimation are verified by measurements, which show significant improvement of the network-based DoA estimation in comparison to the single MIMO radar.

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