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Phase Recovery in Sensor Networks Based on Incoherent Repeater Elements

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***Abstract* — A coherent sensor network without hardware link between the network nodes can be set up based on a single MIMO-radar and several incoherent repeater elements spatially distributed. The repeater sends back the radar signal on the same way it was impinging onto the repeater. All repeater elements have an integrated double-sideband modulator, which allows to separate the signals of different repeater elements at the radar. For angle estimation, the phases of the signals of the repeater elements are crucial, but due to an unknown initial phase of the modulation signal, they cannot be determined directly. Thus, a reconstruction of the modulation phase is necessary for network-based coherent angle estimation. In this paper, a concept for phase reconstruction in such type of networks is proposed, exploiting both sidebands of the modulated signal of the repeater elements.**

***Keywords* — radar network, repeater, MIMO-radar, FMCW, angle estimation, direction of arrival, phase reconstruction.**

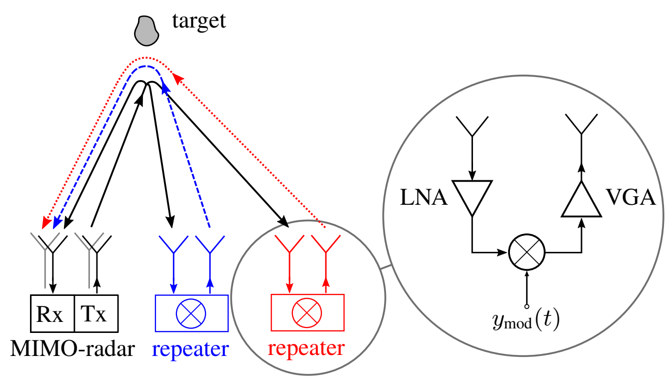

Fig. 1. Concept of the radar network, exemplary with one MIMO-radar and two repeaters.

I. Introduction

Radar networks typically consist of multiple sensor nodes, each of them being a radar on its own. Thus, they usually work incoherently, with each node containing its own signal source. Despite of the incoherence of the signals of different nodes, the information gained by every radar of the network can be combined in a joint evaluation [1], [2], [3], [4]. In such networks bistatic measurements are severely limited by the phase noise, due to incoherent LOs at the receivers and the transmitters.

To overcome the limitations due to incoherent phase noise when evaluating the bistatic signal, in [5] a new concept for a radar network is introduced. This network consists of one frequency modulated continuous wave (FMCW)-radar supplemented by one or more repeater elements. Such a network allows for the measurement of range, Doppler, and angle from all perspectives, the radar ones and the repeater ones. Beneficially, all data is only available at the radar, even the data representing the perspective of the repeater.

Each repeater consists of two amplifiers and a low-frequency double sideband (DSB)-modulator [6]. It receives, amplifies, modulates and re-transmits the signal originating from the radar sensor, as depicted in Fig. 1. Due to the low modulation frequency, the network nodes are coherent in terms of phase noise [5]. In order to further simplify the hardware setup, the modulation signal of each repeater is not connected to the radar. Every repeater has its own independent low frequency signal source. Consequently, each node has its own timebase, which is unknown to the radar. Provided that the frequency deviation of the low-frequency modulation signal is neglectable, this unknown timebase introduces an unknown phase offset. While for bistatic range and velocity processing, this phase offset is static and thus irrelevant [7], for multi-node angle estimation it is crucial.

In [5] based on a network of one multiple-input-multiple-output (MIMO)-FMCW-radar and one repeater node, an angle estimation is performed. With the phase recovery presented in this work, the precision of the angle estimation by a network with at least two repeaters can be improved. This is done by exploiting the phase deviation between the repeater nodes. Instead of estimating it, as proposed in [5], this paper shows how it can be determined analytically. As the phase recovery is based on the two sidebands of the bistatic signal, it works with every repeater containing a simple DSB-modulator.

II. Signal Model

In a network concept like shown in Fig. 1, there is no direct signal path between the nodes. Instead, the repeaters are receiving the radar signal reflected at all the targets. Vice versa, the sensor receives the bistatic repeater signals only via the target reflection. The modulation of the signal thus is necessary to ensure the ability to distinguish between the monostatic and bistatic signals at the radar.

Since the signal is modulated by a DSB-modulator, the modulation signal $y_{\mathrm{mod}}(t)$ with angular frequency $\omega_{\mathrm{mod}}=2\pi f_{\mathrm{mod}}$ and phase ϕ_{mod} is real-valued and can be described as

$$y_{\text{mod}}(t) = \Bigg(\underbrace{\frac{|y_{\text{mod}}|}{\sqrt{2}} \mathrm{e}^{\mathrm{j}(\omega_{\text{mod}}\, t + \phi_{\text{mod}})}}_{y_{\text{mod,USB}}(t)} + \underbrace{\frac{|y_{\text{mod}}|}{\sqrt{2}} \mathrm{e}^{-\mathrm{j}(\omega_{\text{mod}}\, t + \phi_{\text{mod}})}}_{y_{\text{mod,LSB}}(t)} \Bigg). \tag{1}$$

Mathematically, it may be splitted into the two complex-valued modulation signals for the upper sideband (USB) and the lower sideband (LSB), $y_{\text{mod,USB}}(t)$ and $y_{\text{mod,LSB}}(t)$. In dependence of the receive (Rx)-signal $y_{\text{Rp}}(t)$ and the gain G_{Rp} of the repeater, the repeater transmit (Tx)-signal sums up to

$$x_{\text{Rp}}(t) = G_{\text{Rp}}\; y_{\text{Rp}}(t)\; \big(y_{\text{mod,USB}}\,(t) + y_{\text{mod,LSB}}\,(t)\big). \tag{2}$$

This signal is reflected at the targets and received by the Rx-antennas of the radar.

In a network with two repeaters, both bistatic signals yield a phase difference

$$\Delta\phi_{2,1} \approx 2 \frac{2\pi}{\lambda_c}\, d_{2,1} \sin{(\theta)}\,, \tag{3}$$

where $d_{2,1}$ is the spacing between repeater 1 and repeater 2, λ_c the wavelength of the center frequency f_{c}, and θ the angle of the target, where $0°$ corresponds to boresight direction. The influence of f_{mod} is neglected, as f_{mod} is very small. The factor 2 in (3) stems from the fact that a path difference occurs twice, once by the path from the radar to the repeater and once by the way back. This phase difference $\Delta\phi_{2,1}$ is the basis for angle estimation.

Assuming the two repeaters having the same hardware, this phase difference can be extracted by normalizing the phase of one bistatic signal to the other one. For a single target at distance R, this results in

$$\frac{y_{\text{bi2,USB}}(t)}{y_{\text{bi1,USB}}(t)} = \mathrm{e}^{\mathrm{j}\,\Delta\phi_{2,1}} \frac{\mathrm{e}^{\mathrm{j}(\omega_{\text{mod,2}}\,(t-\tau_{\text{R}})+\phi_{\text{mod,2}})}}{\mathrm{e}^{\mathrm{j}(\omega_{\text{mod,1}}\,(t-\tau_{\text{R}})+\phi_{\text{mod,1}})}}. \tag{4}$$

An additional phase shift is caused by the range-dependent propagation delay τ_{R}, which can be determined with sufficient accuracy after range processing. Then, for evaluating the angle-dependent phase shift $\Delta\phi_{2,1}$ between the bistatic signals, the still unknown modulation phases $\phi_{\text{mod,1}}$ and $\phi_{\text{mod,2}}$ have to be reconstructed.

III. Phase Recovery

The estimation of the unknown phase ϕ_{mod} of the modulation signal can be considered as the challenge to match the timebases of repeater and radar. When processing the FMCW-signal, the modulation by the repeater adds an alteration to the ramps, which results in a shift in range and velocity [5]. Thus, ω_{mod} does not have any further impact on the angular estimation. However, as there is no reference between the radar and the repeater, the phase

$$\phi_{\text{mod}} = \arg\!\big(y_{\text{mod}}\,(t\!=\!0)\big), \tag{5}$$

at the starting time t=0 of the radar signal can be unequal to zero and is unknown.

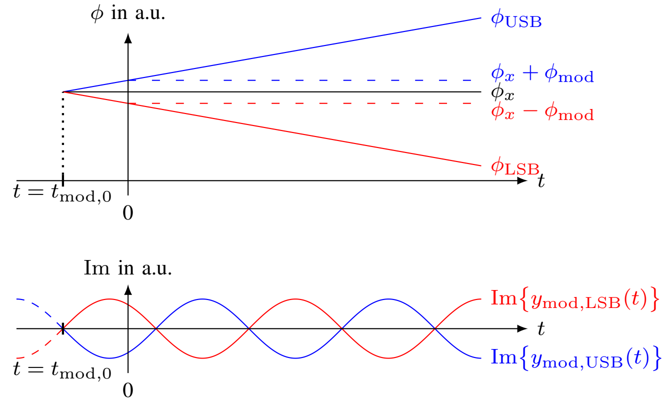

Fig. 2. Relation of the phases of the USB and LSB. At $t_{\text{mod,0}}$ both sidebands are having the same phase equal to the phase offset ϕ_{x}. This represents the common phase offset introduced by the signal path. The modulation phase ϕ_{mod} can be determined based on ϕ_{USB} and ϕ_{LSB} at the starting time of the radar signal, t=0. The influence of τ_R is neglected.

Neglecting the influence of ω_{mod} on the Doppler shift, the path from the repeater back to the radar can be represented by the complex factor $a(t)$. Thus, the USB and LSB can be described as

$$y_{\text{USB}}(t) = \frac{G_{\text{Rp}}}{\sqrt{2}}\, y_{\text{Rp}}(t) a(t)\, \mathrm{e}^{\mathrm{j}(\omega_{\text{mod}}(t-\tau_{\text{R}})+\phi_{\text{mod}})}, \tag{6}$$

$$y_{\text{LSB}}(t) = \frac{G_{\text{Rp}}}{\sqrt{2}}\, y_{\text{Rp}}(t) a(t)\, \mathrm{e}^{-\mathrm{j}(\omega_{\text{mod}}(t-\tau_{\text{R}})+\phi_{\text{mod}})}. \tag{7}$$

As the modulation phase influences both the USB and the LSB but with different signs, the phase can be recovered based on the two signals. This concept is illustrated in Fig. 2. The common phase offset ϕ_{x} of the sidebands stems from the signal path, which is the same for both sidebands. At time $t_{\text{mod,0}}$, they both have the same phase ϕ_{USB}=ϕ_{LSB}=ϕ_{x}. Thus, at this point ϕ_{mod} is equal to zero. From this point the phase of the upper sideband is rotating in positive direction, while the phase of the lower one rotates in negative direction. The phase difference between both signals at t=0 hence equals the doubled modulation phase ϕ_{mod}.

Therefore, ϕ_{mod} can be extracted by dividing the upper by the lower sideband of the same bistatic signal, which sums up to

$$\frac{y_{\text{USB}}(0)}{y_{\text{LSB}}(0)} = \mathrm{e}^{\mathrm{j}2(-\omega_{\text{mod}}\tau_{\text{R}}+\phi_{\text{mod}})} \tag{8}$$

In this way, with correcting $\omega_{\text{mod}} \cdot \tau_R$ by the range processing, the modulation phase ϕ_{mod} can be reconstructed.

The phase recovery can be performed based on the complex range-velocity-matrix. While ω_{mod} results in a shift of the range-velocity (rv)-bin, ϕ_{x} as well as ϕ_{mod} remain as static phase offset. Precondition for this is a correct sampling of f_{mod}, as any sampling error results in a phase error after the fast Fourier transform (FFT). Alternatively, the phase error may be corrected according to [8] or zero padding can be applied in velocity direction.

The correct sampled rv-bin of the USB and LSB, i_{USB} and i_{LSB}, are having the same phase as $y_{\text{USB}}(0)$ and

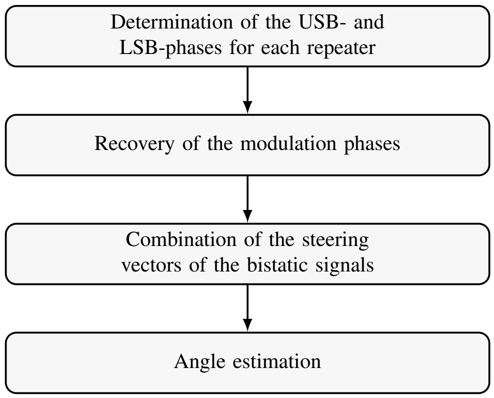

Fig. 3. Flowchart for angle estimation with phase recovery based on a radar network with repeaters.

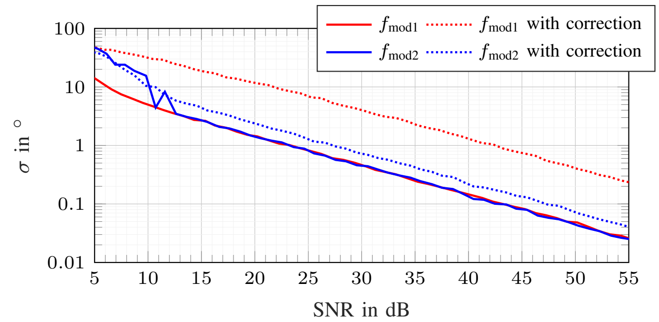

Fig. 4. Simulation results: Standard deviation σ of the phase recovery error vs. range for two different modulation frequencies. The phase recovery was performed with sampling correction as well as with no correction.

$y_{\text{LSB}}(0)$, respectively. Thus, the modulation phase can now be reconstructed by calculating

$$\begin{aligned}\tilde{\phi}_{\text{mod}} &= \arg\left(\sqrt{\frac{i_{\text{USB}}}{i_{\text{LSB}}}}\mathrm{e}^{\mathrm{j}\,\omega_{\text{mod}}\tau_{\text{R}}}\right) \\ &= \arg\left(\pm\mathrm{e}^{\mathrm{j}\phi_{\text{mod}}}\right) = \begin{cases}\phi_{\text{mod}}+180^\circ \\ \phi_{\text{mod}}\end{cases} .\end{aligned} \tag{9}$$

Thus, an ambiguous determination of the modulation phase is achieved. For a radar with N_{ch} channels, the phase recovery can be improved by averaging the results of all channels. Thus, the standard deviation σ of the estimation error can be reduced by $1/\sqrt{N_{\text{ch}}}$. Now, the recovered $\tilde{\phi}_{\text{mod}}$ as well as $\tilde{\phi}_{\text{mod}} + 180^\circ$ can be inserted into (4). Therefore, two solutions for the phase offset between two repeaters are gained.

The correct solution can be chosen after performing an angle estimation with each of them. This is done by combining the steering vectors of the bistatic signals as aqcuired by the radar with respect to the phase difference between the repeaters. Then, an angle estimation, e.g. with compressed sensing (CS) [5], [9], can be performed. The whole process is depicted as flowchart in Fig. 3. In this way, the coherency of the network nodes is established and exploited for the estimation of the target angle.

IV. Simulations

To evaluate the phase recovery, simulations of a network with a MIMO-radar and one repeater were performed. The simulations contain a single target at a distance of $R = 8\,\text{m}$. The signal-to-noise ratio (SNR) is directly controlled by adding white Gaussian noise to the signal. As radar, a four channel sensor with a bandwidth of 2 GHz, an up-ramp duration of T_{up}=256 µs and a chirp repetition time of T_{r}=328 µs is simulated. Every ramp is sampled with 1024 samples, and N_{r}=128 ramps are evaluated.

For the repeater, two different modulation frequencies are chosen. The first frequency, $f_{\text{mod1}} = 31\,250\,\text{Hz}$, can be sampled correctly by the FFTs of the rv-processing, while the second, $f_{\text{mod2}} = 31\,260\,\text{Hz}$, can not. The simulations are done for SNRs of the bistatic signal between 5 dB and 55 dB in steps of 1 dB. For a proper statistic evaluation, 500 simulations per setting are performed, each with the modulation phase ϕ_{mod} chosen randomly. The reconstructed phase $\tilde{\phi}_{\text{mod}}$ is then compared to ϕ_{mod} and the error $\Delta\phi_{\text{error}}$ is calculated.

In Fig. 4, the standard deviation σ of $\Delta\phi_{\text{error}}$ is shown, which decreases inversely proportional to the SNR. With small σ even for small SNR, the phase recovery shows a very good performance.

For $f_{\text{mod2}} = 31\,260\,\text{Hz}$, due to the incorrect sampling, a bias gets introduced. This bias can be corrected according to [8]. But, with correction, the standard deviation is degraded, see Fig. 4. This is caused by the influence of the neighboring FFT-bin.

V. Measurements

Measurements were performed with two setups, one for statistical evaluation of the phase recovery and one for angle estimations. The parameters of the radar with $f_{\text{c}} = 78\,\text{GHz}$ are chosen equivalently to the simulations. Due to the low gain of the repeater from [6], measurements with the signal path as shown in Fig. 1 yield only a very low SNR. Therefore, in the first measurement setup the radar directly faces the repeater. The modulation signal as well as the trigger of the radar are each supplied by a signal source, and both are also connected to an oscilloscope for recording the signals. In this way, a reference for the reconstructed phase can be determined.

In Fig. 5a the standard deviation of the phase recovery error is plotted vs. range. As reference in Fig. 5b the SNRs of the peaks related to the USB and LSB are plotted. The difference between the sidebands is due to the bandpass characteristic of the radar Rx filter. Taking the minimum SNR of the two sidebands into account, the standard deviation σ in Fig. 5a is close to the values expected from the simulations.

For angle estimation, measurements with a network consisting of one 16-channel radar and two repeaters are performed. The linear array of the radar is uniformly spaced by 0.5 λ_c, while the distance between the two repeaters amounts to 23 λ_c. The measurement setup is equivalent to the setup shown in Fig. 1. The target is placed at $\theta \approx 3.4°$. In Fig. 6 (top), the phases of the first measurement were

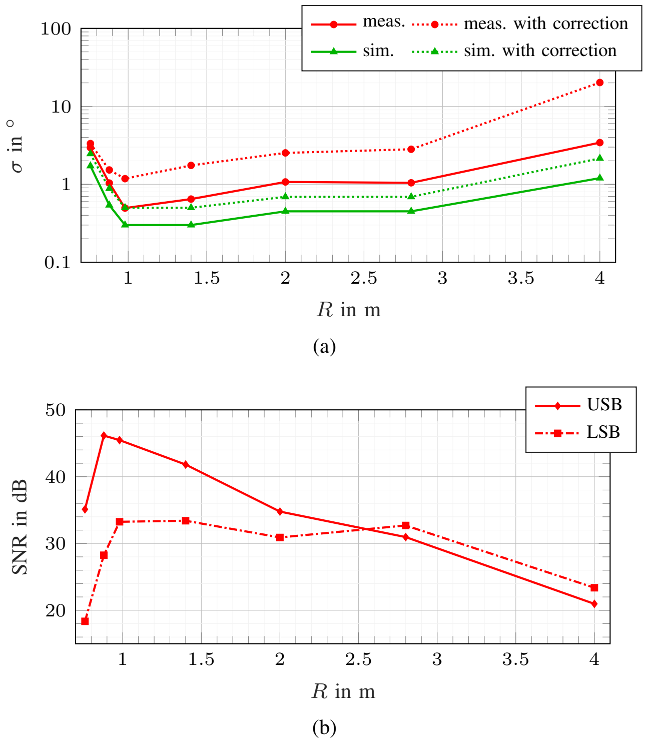

Fig. 5. Measurement results of the phase recovery, with correction of the sampling error and with no correction: (a) Standard deviation σ of the phase recovery error vs. range. As reference, also σ from the simulations is plotted; (b) SNR of the USB and the LSB.

evaluated with an FFT-based angle estimation [10]. The results for both solutions of ϕ_{mod} are plotted. The correct one has one single peak, while the wrong solution has two main peaks with equal power. Due to the large gap in the total aperture of repeaters and radar, a $\sin(x)/x$ envelope occurs. With an angle estimation based on CS this can be solved.

Additionally, in Fig. 6 (bottom) the angle estimation is plotted over 100 measurements, each evaluated with an iterative hard thresholding algorithm for CS, like in [5]. For all measurements, the results are close to the ground truth and the standard deviation of the estimated angle amounts to $0.09°$.

VI. Conclusion

In this work, a method for phase recovery in a radar-repeater network is presented. With exploiting the phases of the USB and LSB, the unknown modulation phase of the repeater can be reconstructed, which enables new possibilities for better angle estimation. The phase recovery was evaluated using simulations and measurements. In both cases, it provides good results with a low standard deviation. Furthermore, an angle estimation was performed based on measurements. It shows a good performance with a very low deviation despite the arbitrary modulation phase in every measurement.

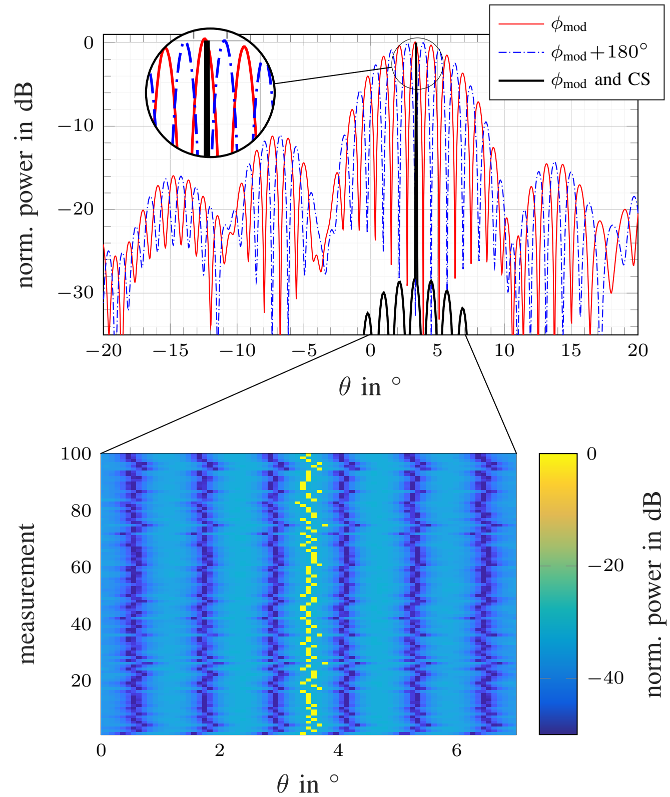

Fig. 6. Measurement results: Top: Angle estimation for a single measurement. The results for an angle estimation with FFT and zero padding as well as the result with compressed sensing are plotted; Bottom: Angle estimation with compressed sensing, plotted for 100 measurements.

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