IRS-Enabled Breath Tracking With Colocated Commodity WiFi Transceivers

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Abstract—Intelligent reflecting surfaces (IRSs) are a key enabler of various new applications in 6G smart radio environments. This article aims to enhance self-interference (SI) cancellation for breath tracking with commodity WiFi devices by utilizing an IRS prototype system. SI suppression is a crucial requirement for breath tracking with a single antenna site, as the SI severely limits the radio sensing range by shadowing the received signal with its own transmit signal. To this end, we propose to assist SI cancellation by exploiting an IRS to form a suitable cancellation signal in the analog domain. Building upon a 256-element IRS prototype, we present results of breath tracking with IRS-assisted SI cancellation from a practical testbed. We use inexpensive WiFi hardware to extract the channel state information (CSI) in the 5-GHz band and analyze the phase change between a colocated transmitter and receiver with added local oscillator (LO) synchronization. As a result, we can track the breath of a test subject regardless of position in an indoor environment with a room-level range. The presented case study achieves promising performance in capturing the breath frequency and breathing patterns.

Index Terms—Breath tracking, channel state information (CSI), IEEE 802.11, intelligent reflecting surface (IRS), joint communication and sensing, reconﬁgurable intelligent surface (RIS), WiFi.

I. INTRODUCTION

Due to society’s demographic change, more and more people need care and have certain health conditions [1]. To enable the affected persons to live independently in their own homes for longer, a wide variety of systems for monitoring vital parameters have been developed [2]. This kind of support is called ambient assisted living (AAL) and can be used to monitor vital parameters, such as respiration and fall detection. In addition, respiration and heartbeat are essential vital signs for physical health monitoring, as the information provides important indications of medical problems [3]. This article focuses on respiration detection as a variant of AAL.

Traditional methods of respiration detection often use body-worn sensor technology that detects the expansion of the chest [2]. However, these sensor technologies are intrusive and often perceived as uncomfortable, so they are unsuitable for long-term respiration monitoring [4].

Due to advances in health monitoring research related to the Internet of Things, some previous works have already presented contactless variants of respiration detection using radio waves [4], [5], [6], [7], [8]. The previous works either use specialized hardware with directional antennas [9] or exploit a separation of multiple nodes in the experiment space, similar to bistatic radar, with the test subject located between transmitter and receiver [4], [5], [6], [7], [8], [10]. However, this particular placement of devices and test subjects is difficult to implement in practice and concepts that require a cabled connection between two nodes as a reference limit application of this technique.

Among many other promising applications and new techniques, research on 6G investigates the use of intelligent reflecting surface (IRS). The IRS is a synthetic surface able to passively manipulate the reflected signal depending on an electrical reconfiguration of the elements. For the first time, this promising concept allows deviation from the paradigm of a determined, passive communication channel and allows optimization of the channel for a specific application or user. These innovative features qualify the IRS as part of an upcoming communication standard [11]. Given this development, IRS will become more readily available as a component of the communication infrastructure in the future.

Therefore, in this article, we propose an IRS for simplifying the user equipment (UE) hardware in full-duplex (FD) communications as well as joint communication and sensing contexts. In in-band FD communication, a frequency resource is used for simultaneous transmission and reception on the same carrier frequency. This simultaneous use of a frequency resource increases spectral efficiency by up to a factor of two: that is, the throughput of a communication link is doubled compared to classic half-duplex systems. Furthermore, joint communication and sensing applications are closely related to FD communication. In this field, a communication system is used to sense a target and transmit data simultaneously. During data transmission, the reflection of the transmitted signal from the environment or the target is received simultaneously. Thus,
additional sensors are omitted, and added value is added on top of the communication.

In such systems, the main challenge is suppressing the self-interference (SI) caused by simultaneous transmission and reception on one frequency. This SI suppression is classically achieved by RF hardware such as circulators or auxiliary transmitters. Addressing these issues, we propose a novel concept that utilizes an IRS to form a cancellation signal that is phase inverted to the sum of the individual SI components caused by multipath reflections and hardware imperfections, such as limited antenna isolation and internal leakage in an FD transceiver. The signal reflected from the IRS aims to cancel the SI at the receiving antenna of a node. Using a greedy algorithm to configure the IRS, our setup adapts to various indoor multipath environments and can generate the desired cancellation signal without prior channel or location knowledge. Furthermore, the implementation costs for a single UE are reduced by taking advantage of the IRS infrastructure that will be deployed in 6G networks. The reduced SI achieved in this way allows us to simultaneously transmit a WiFi packet on a WiFi channel from one antenna location and receive it back at the exact location.

Since the IRS-enabled SIC cancels static reflections, leakage, and other quasi-static signal components, the total signal power at the receiver stemming from the SI is reduced. By reducing the SI power, the signal of interest is amplified by the AGC and thus raised above the level of quantization noise. Thus, much more detailed channel estimation is achieved. For example, if we now consider an indoor respiration detection scenario, the raising and lowering of a subject’s chest causes a time-varying small phase change on some multipath paths. Initially, these tiny changes were overshadowed by the SI and disappeared in the quantization noise of the channel estimation. With the quasi-static SI components now removed, the AGC can amplify the received signal into the ADC range better. Due to this reduced quantization noise, the tiny phase variations caused by reflection from a subject’s chest are detectable from only one antenna location. Additionally, the detection is mainly independent of the subject’s position in space.

II. PRIOR WORK

Several concepts for breath tracking with commodity WiFi hardware have been presented in the past. This section briefly introduces the core concepts of breath tracking and the current state of the art. Then, in Table I, we offer an overview of prior work compared to our proposed approach highlighting important key aspects, which we further elaborate on below.

Zhang et al. [4] demonstrated respiration detection with commodity WiFi hardware in an indoor scenario using a wired reference channel. For this purpose, the transmit signal was split by an RF splitter and passed once to a transmitting antenna and once to the receiver through a coaxial cable. The receiver has three receiving chains: one receives the reference signal, and the other two are connected to two antennas. The reference signal enables a correction of the carrier frequency offset (CFO) and a measurement of the phase information over several WiFi packets. Although this setup shows good performance in the presented scenarios, it requires an additional wired reference channel in the carrier frequency range and at least three receiver chains, as well as an RF splitter. Furthermore, the setup is operated with spatial separation of the transmitter and receiver so that the reflection from the chest is sufficiently large relative to the direct path. Our approach, on the other hand, uses colocated antennas and operates with a single transmitting and a single receiving antenna and associated transceiver chain, hence reducing the cost and complexity of the UE. Another advantage of our approach is the simpler deployability by utilizing only one antenna location, which eliminates the use of expensive coaxial cables in the carrier frequency range.

Wang et al. [5] showed a scheme wherein two WiFi devices are spatially separated, with the test subject located between the transmitter and receiver. The phase difference at two receiving antennas is used to obtain stable phase information containing the breath information. This setup, however, requires at least two receiving antennas and is also operated only with spatial separation between transmitter and receiver. Additionally, this setup imposes preliminaries on the position of the test subject, resulting in limited applicability. Our approach allows the use of a single antenna site and

| Key aspect                        | Paper                  | Our approach | BreathTrack [4] | PhaseBeat [5] | Liu, X. et al. [6] | Liu J. et al. [7] | Wang et al. [8] | Aub et al. [9] | MoBreath [10] | TractorBeat [12] | FullBreathe [13] |
|-----------------------------------|------------------------|--------------|-----------------|--------------|-------------------|-----------------|----------------|----------------|--------------|-------------------|-------------------|
| Number of TX/RX antennas          | 1 TX                   | 1 TX         | 1 TX            | 1 TX         | 1 TX              | 1 TX            | 1 TX           | 1 TX           | 1 TX         | 1 TX              | 1 TX              |
| Omnidirectional antennas          | ✓                      | ✓            | ✓               | ✓            | ✓                 | ✓               | ✓              | ✓              | ✓            | ✓                 | ✓                 |
| RF reference channel              | WiFi                   | WiFi         | WiFi            | WiFi         | WiFi              | WiFi            | WiFi           | WiFi           | WiFi         | WiFi              | WiFi              |
| Synchronization                   | LO                     | RF           | x               | x            | x                 | x               | x              | x              | x            | x                 | x                 |
| Colocated transceivers            | ✓                      | x            | x               | x            | x                 | ✓               | ✓              | ✓              | ✓            | ✓                 | ✓                 |
| Free position of test person      | ✓                      | x            | x               | x            | x                 | x               | x              | ✓              | ✓            | ✓                 | ✓                 |
thus removes the preliminaries imposed on the test subject’s position to the greatest extent possible, resulting in a much more convenient application. In addition, the reduction to one antenna location and the use of a single transmitting and one receiving antenna reduce the cost of a UE.

Adib et al. [9] presented an FMCW radar approach to breath tracking. The approach uses a proprietary radar that sweeps from 5.46 to 7.25 GHz by means of two directional antennas aimed at the subject’s chest; the distance to the subject, or the elevation and depression of the chest, can thus be measured. Although providing good measurement results, this approach is rather impractical since the use of directional antennas determines the position of the test subject. In addition, it requires the use of specialized hardware that is only responsible for sensing, thus driving up the cost of a device. In our approach, we utilize commodity WiFi, adding value through reusability as communication hardware or in the joint communication and sensing context. In addition, we use omnidirectional antennas that do not restrict the position of the test subject.

All of these previously outlined concepts use either specialized RF hardware [4], [9] or directional antennas [7], [9] or require spatially separated transmitter and receiver nodes [4], [5], [6], [7], [13]. This is why we propose utilizing IRSs in future 6G networks to assist the SIC and relax the requirements for specialized hardware on the UE side. This can significantly reduce costs and complexity at the UE. By using channel estimation, the IRS is tuned so that the reflected signal cancels the SI at the UE, thus enabling breath tracking from a single antenna site.

III. RELATED WORK ON IRS SENSING

In addition to the work on respiration detection summarized in the previous section, some recent work on sensing with IRS has been published independent of our work. Therefore, we summarize in this section five relevant papers that employ an experimental setup and address some form of sensing, e.g., breath tracking, posture recognition, or localization.

Hu et al. [14] demonstrated an application of an IRS in the frequency range of 3.198 GHz for posture recognition. For this purpose, a 0.69 × 0.69 m IRS is illuminated with a directional antenna in an absorber hall, with the test subject in front of the surface. A separate receiving antenna is located below the surface. The surface supports 3-bit phase shifting, and USRPs are used as transceivers. The authors successfully showed that the reception signal optimized by the IRS could be used to classify the postures “standing,” “sitting,” “bending,” and “lying down” using neural networks. Usman et al. [15] explored an activity recognition application for the NLOS case. Conventional systems do not perform well under NLOS conditions, so the authors use an IRS to bypass the NLOS condition. Here, a 1.08 × 0.78 m IRS at 3.75 GHz is used to detect the activities of a test subject in an angled corridor. The activities investigated were “sitting,” “standing,” and “walking.” In contrast to [14], no static postures were investigated. USRP X300 with spatially separated horn antennas were used as transceivers.

Keykhosravi et al. [16] demonstrated the use of two IRSs in the 60-GHz range for localization application. For the practical implementation, a scenario with one base station, one UE, and two IRS is considered. Localization is to be achieved using AoD estimation. Here, a VNA is used as a transceiver system, where port 1 represents the transmitting BS and port 2 represents the receiving UE. To experimentally verify the effect of a passive IRS, phased array transceivers from SiversIMA are used, which are feedback coupled at the IQ level, allowing phase shift at the IQ level. By cycling the phased array elements, the AoD at the respective IRS is determined and finally the position is determined by the intersection of these two AoDs. This setup has been successfully demonstrated by the authors as a laboratory setup. Li et al. [17] demonstrated full-scene imaging and adaptive recognition of hand signs and vital signs of multiple noncooperative people in the 2.4-GHz range. The authors use two separate receive/transmit antennas or WiFi background signals to illuminate the IRS. Based on the position of the test person or a camera image, the reflection of the surface is then focused on a region of interest to detect breathing. In contrast to our proposed approach, two antennas are required, as well as the test person’s position and the setup’s geometry. In addition, our proposed approach requires no prior knowledge of the geometric arrangement to optimize the IRS state.

Another application in the area of localization or location privacy has been shown in [18]. In this article, the authors show a countermeasure against an attack on the presence detection or localization of a person using IRS. In the scenario considered, a test person is in a room with a WiFi access point, while the Eavesdropper is outside the room. The Eavesdropper estimates the channel state information (CSI) to the access point and determines whether the CSI has been changed by a person in the room based on changes in the CSI over time. This attack scenario requires a channel that is kept static in order to detect the person’s changes. To counter the attack, an IRS was placed in the room and periodically switched to add changes to the channel. It was successfully shown that this prevented the Eavesdropper from detecting the person’s presence in the room.

The remainder of this article is structured as follows. First, in Section IV, we introduce the channel model. Then, in Section V, we propose our approach to IRS-assisted SI cancellation. Next, we detail a greedy algorithm to optimize the IRS state to minimize the SI. In Section VII, we elaborate on the CSI data used in WiFi transceivers, which is used for data acquisition in our experiments. In Section VIII, we first describe the experimental layout and the scenario settings. Then, we detail the IRS prototype and the transceiver hardware used. Finally, in Section IX, the results of our experiments are presented.

IV. CHANNEL MODEL

The acquisition of vital parameters of a test subject rarely takes place under idealized conditions where only the direct path exists. It is, therefore, necessary to establish a suitable channel model for a complete representation of the conditions. In indoor multipath environments, various propagation paths exist that are affected by the environment in distinct ways.
Fig. 1 illustrates the propagation paths in a typical indoor scenario. The present paths can be divided into four types, which are explained in detail in the following and ultimately summarized into an overall channel model.

First, we consider the reflection of interest that contains the relevant change due to breathing. The detection of a test subject’s breathing by means of radio signals uses the premise that breathing is always associated with a periodic raising and lowering of the test subject’s chest. When a radio wave impinges on the subject’s chest, it is reflected and can be evaluated with a suitable receiver. The path length \( d \) that a radio wave travels from a transmitting antenna across the chest to the receiving antenna changes as a function of the deflection of the chest. The effect of breathing on the direct path between the transceiver and the test subject’s chest is one component of the channel model. The direct path can be described by a complex attenuation \( a_0 \) and a phase term as the CSI \( y_d(t) \) as follows:

\[
y_d(t) = a_0 \cdot e^{-j2\pi \frac{d(t)}{\lambda}}.
\]  

Here, \( d(t) \) represents the path length from the transceiver to the chest of the test person, and \( \lambda \) is the wavelength of the carrier frequency used. The path length \( d(t) \) varies according to the elevation and depression of the thorax. The change in chest expansion is expressed as a phase change of the CSI \( y_d(t) \) of the direct path and depends on the expansion of the chest, as well as the carrier wavelength. The typical elevation of the thorax for healthy adults is, on average, approximately 30 mm for deep breathing and 3 mm for shallow breathing, whereas Kaneko and Horie [19] demonstrated dependence on age, sex, and weight of the test subject. Thus, for a realistic scenario using a 5-GHz carrier frequency, a maximum phase change for the direct path of 360° for deep breathing and 36° for shallow breathing is expected.

In addition to the direct paths altered by respiration, there is also a large proportion of static paths in multipath environments. These static paths result from the reflection of the transmitted signal from boundary surfaces, such as ceilings, walls, floors, furniture, and interior environments in general. These objects are stationary for a typical period of several seconds to hours and are not moved. Thus, the static paths can be assumed to be constant during the acquisition period under consideration. However, because boundary surfaces, such as walls, are very large, these paths make up a very large fraction in terms of received power. The effect of all static reflections can be represented as a sum of all static paths. Each static path can be represented by a complex attenuation \( a_l \), which commonly takes the form of a Gaussian distribution. Additionally, a phase term depending on the path length \( d_l \) of the respective path \( l \) is accounted for the respective path. However, because the channel is stationary for the course of our measurement, \( a_l \) drawn from the Gaussian distribution is also constant for the observed period. The CSI of all static paths, \( y_s \), can be written as a sum with respect to the number of paths as

\[
y_s = \sum_{l=1}^{L} a_l \cdot e^{-j2\pi \frac{d_l}{\lambda}}.
\]  

Hereby, \( \lambda \) is the carrier wavelength, and \( L \) denotes the total number of static paths, the number of which is largely determined by the number of scatterers in the near environment.

In addition to the passive objects and their contributions to the reflections discussed above, in future 6G networks, adjustable radio environments will be available, allowing a radio channel to be reprogrammed by the users or network operator’s equipment. A promising candidate for adjustable radio environments is an IRS with binary switching states. These surfaces, which are divided into patches, allow the phase of a signal reflected from that patch to be changed. These adjustable reflection paths must be accounted for in the channel model. The CSI for the IRS-assisted paths \( y_{IRS}(s) \) with an IRS with \( I \) elements can be modeled as a summation of the individual reflections of each element as

\[
y_{IRS}(S) = \sum_{i=1}^{I} a_i \cdot e^{-j2\pi \frac{d_i}{\lambda}} \cdot e^{j\phi_i}.
\]  

where \( a_i \) is a complex attenuation factor, \( s_i \) is the binary switching state of the IRS element, and \( d_i \) is the path length of the respective element \( i \). The binary vector \( S \), therefore, contains the state of all IRS elements. Furthermore, \( \rho \) is the phase shift capability of the IRS, a parameter determined by the design of the surface. Usually, it is designed to be 180° for binary switching surfaces to achieve a maximum effect on the reflected signal.

In addition to the paths affected by the propagation environment described above, there are always paths from the transmitter to the receiver that are independent of the propagation environment. This direct crosstalk of the transmit signal into the receive path is generally referred to as SI. It is caused by, among other things, internal leakage in the transceiver, limited analog isolation of transmitting and receiving antennas, or, in the case of a common transmitting/receiving antenna, limited isolation of circulators, and antenna mismatch. This SI occurs in any FD node and must be specifically addressed. This SI is usually the strongest path in terms of power in colocated
nodes. The path comprises of various parameters, which can usually be assumed to be time invariant over the period under consideration [20], [21]. The SI is additionally temperature dependent, but this is also negligible for many applications due to the slow temperature change of components. The self-interference $y_{SI}$ can be represented accordingly as

$$y_{SI} = a_{iso} \cdot e^{-\theta}$$  \hspace{1cm} (4)

where $a_{iso}$ denotes the attenuation, which results from the antenna isolation, as well as internal leakage of the transceiver and cable attenuation. $\theta$ is the associated phase, which strongly depends on the cable lengths, antenna delays, and the geometrical arrangement of the antennas. For the narrowband case, these parameters can be summarized with the described representation since the determination of the individual components of the SI is often not possible and not necessary.

The model for the entire channel can thus be set up as a sum of the individual paths as

$$h(S, t) = y_s + y_d(t) + y_{IRS}(S) + y_{SI}.$$  \hspace{1cm} (5)

Higher order reflection paths with more than one reflection are not accounted for in the channel model, as they are negligible due to the very low received power [22].

With the channel model from (5), the total received signal at the receiver $y_{recv}(S, t)$ is given as

$$y_{recv}(S, t) = h(S, t) \ast x_{tx}(t) + z(t)$$  \hspace{1cm} (6)

where $x_{tx}(t)$ is the transmit signal and $z(t)$ is the receiver noise.

V. IRS-ASSISTED SI SUPPRESSION

From the presented channel model, it is apparent that the change in $y_d(t)$ caused by the subject’s breathing accounts for only a small part of the received signal. The received power components due to static reflections and SI are orders of magnitude larger than the single reflection of the subject’s chest. As a result, the overshadowing of the phase change caused by respiration by the other signal components occurs. Considering the finite ADC resolutions of communication systems, this change thus falls into the range of quantization noise and can no longer be recovered [23].

Although the transceiver and antenna design can already achieve substantial SI suppression, this is still insufficient for detecting extremely small signals such as those of respiration. In particular, when using commodity WiFi or other low-cost transceivers, the ADC resolution of the systems is limited, and extremely small changes in the overall signal cannot be detected. Therefore, in previous WiFi CSI-based work [4], [5], [6], [7], two spatially separated transceivers have been used to circumvent the problem of SI and to magnify the effect of breathing in the received signal by organizational measures. However, this separation of antennas and specific placement of the test person in prior works is hardly applicable in practice.

Due to the limited SI capability of commodity WiFi devices, we propose to use the IRS available in future 6G networks to reduce the impact of SI and thus simplify the spatial separation of transceivers to a single antenna site. Simultaneous transmission and reception (FD) is required for the detection of a subject’s respiration reflection from a single antenna site. Currently, there are no FD WiFi devices commercially available [24]. Recently, the IEEE 802.11bf Task Group was formed with the goal of standardizing ongoing research on WiFi CSI-based sensing applications [25]. In addition, the IEEE task group working on WiFi-7 (IEEE 802.11be) is considering FD (STR MLMR—simultaneous transmit and receive; multilink multiradio) as part of the upcoming standard [26]. However, since no FD WiFi devices are available yet, we suggest using commercially available half-duplex WiFi devices. Due to the SI cancellation added by the IRS, a suitable FD node can be built from two half-duplex transceivers.

Equation (5) shows that the static reflections and the SI are time invariant and not adjustable. The IRS reflection, on the other hand, is adjustable by the IRS state matrix $S$. The ability of the IRS to change the phase of reflection paths can be used to manipulate a reflected signal to be phase-inverted to the sum of the SI and static reflections. Since it is known that the SI and static reflections are time invariant over the measurement period, cancellation of the static components can be achieved by selecting an appropriate state matrix $S$ of the surface while leaving the time-varying component of the respiratory signal unaffected. The cancellation of the static components results in a much larger phase change, which is caused by respiration and is thus recovered from the quantization noise. If we consider the sum-receive power components in the form of a link-budget calculation, assuming that by a suitable choice of the IRS state $S$, $y_{SI}$ is exactly phase inverted to SI and perfect cancellation results, then it follows that for the residual SI power $P_{res}$:

$$P_{res} = P_{tx} - a_{iso} + P_s - P_{IRS}$$  \hspace{1cm} (7)

where $P_{tx}$ is the transmit power, $a_{iso}$ is the antenna isolation given by the antenna assembly, $P_s$ is the power received from static reflections, and $P_{IRS}$ is the power reflected from the IRS to the receiving antenna (all values except $a_{iso}$ are in dBm). From (7), it can be seen that the received SI power depends only on powers that cannot be changed during runtime, with the exception of $P_{IRS}$. Thus, for maximum SI cancellation, the signal reflected from the IRS must have power exactly equal to the sum of the powers of $P_{tx} - a_{iso}$ and $P_s$ in order for them to cancel exactly.

The maximum reflected power of the IRS $P_{IRS}$ is determined by the size of the IRS, the reflectivity or gain, and the distance of the transceiver to the IRS. Antenna isolation is largely determined by the antenna design and can be degraded by inefficient design. This results in a tradeoff for the antenna design versus the maximum reflected power $P_{IRS}$ from the surface. On the one hand, a too large distance to the IRS (or a too small IRS) can lower the reflected power $P_{IRS}$ toward the receiving antenna to such an extent that the cancellation signal falls below the leakage in terms of power and thus cancels insufficiently. On the other hand, the antenna isolation can be reduced by a suboptimal antenna design and thus significantly increases the leakage. In this case, the leakage becomes stronger than the cancellation signal, and good cancellation of the SI is no longer possible.
Excessively high power of the IRS reflection would, despite correct phasing, have a negative effect because it would overshoot the target. Since the IRS, as a passive element, can attenuate the reflected signal but not amplify the original incident power, sufficient isolation is required by the antenna design. With the availability of larger IRSs, the problem of distance and antenna isolation is eased since more power can be reflected from a larger IRS.

The goal is now to find a suitable state of the surface for which the condition of 180° phasing relative to the SI is given. We, therefore, consider the 2π-periodic phase on the IRS-affected reflection paths. Since a beamforming effect of IRS occurs only in the far field [27]; near-field or transition-field conditions can be assumed in indoor environments in view of large-dimensional IRS, with the array far-field radius \( r_{FF} \) given by [28] as

\[
r_{FF} = \frac{2D^2}{\lambda}
\]

where \( D \) is the longest dimension of the IRS, \( \lambda \) is the carrier wavelength, and the far-field boundary \( r_{FF} \) of the entire IRS is calculated. For a small surface with an edge length of 40 cm and a carrier frequency of 5 GHz, the far-field begins at a distance of 5.33 m. Since the far-field boundary increases quadratically with the edge length of the surface, a near-field or transition field can always be assumed for large-dimensional IRS in indoor scenarios. We, therefore, use the elementwise representation of the IRS also given in (3) because the antenna array distance in indoor scenarios always exceeds the far-field distance for an individual element of the surface.

Through the summation of all phase changes introduced by the IRS, we obtain the sought cancellation signal. If the distances and geometrical dimensions of antennas and surfaces relative to each other are known, the phase of the respective path can be calculated by a simple geometrical calculation of the path length \( d_i \) to the respective patch of the surface as

\[
d_i = \sqrt{d_{IRS}^2 + (d_h \cdot \frac{w}{N_h})^2 + (d_v \cdot \frac{h}{N_v})^2}
\]

\[
\phi_i = \frac{2\pi}{\lambda} \cdot (d_i \text{ (mod } \lambda)).
\]

Here, \( d_{IRS} \) is the distance between the transmitting/receiving antennas and the surface, \( w \) is the width of the surface, and \( h \) is the height of the surface. The number of elements of the IRS in each row is \( N_h \), and the number in each column is \( N_v \). We assume that the position of the transmitting/receiving antennas is known and that this position is located in front of an element of the surface. Thus, starting from this position, the neighboring elements can be indexed relatively with integer numbers. Here, \( d_h \) is the horizontal index and \( d_v \) the vertical index. The left and right directly adjacent elements would have the index values \( d_h = 1 \) and \( d_v = 0 \). Thus, the path length \( d_i \) can be calculated for each element of the surface. From this, the 2π periodic phase \( \phi_i \) for each surface element \( i \) is calculated by a modulo division by the wavelength \( \lambda \). The antenna spacing for separate transmitting and receiving antennas is usually negligible for the calculation of the phase distribution since the distance of the antennas to each other is much smaller than the distance of the antennas to the surface \( d_{IRS} \).

Fig. 2 provides an example of the 2π-periodic phase distribution for a 16 × 16 IRS with a width of \( w = 0.4 \) m and a height of \( h = 0.32 \) m at a carrier frequency \( f_c = 5.385 \) GHz. The transmitting and receiving antenna is located in the center, at the top of the surface, at a distance \( d_{IRS} = 1 \) m. The resulting pattern of the phase distribution over the surface reflects well the spherical propagation of the wave when it hits the surface. This pattern is a good candidate for achieving the highest possible power of the IRS reflection at the receiving antenna. To achieve the highest possible SI cancellation, the reflections of the individual patches must meet constructively and in phase at the receiving antenna. A good candidate for such an IRS state is thus a projection of the phase distribution onto a binary state of the surface. The adjustment of the surface thus compensates for the different path lengths and minimizes the SI.

VI. Greedy Algorithm for IRS-Assisted SI Cancellation

In this section, we present a greedy algorithm for optimizing SI cancellation. In practical applications, the IRS is part of the infrastructure or space. A known position of the UE can usually not be assumed since its manual measurement is costly, not very user-friendly, and error-prone. Instead, we propose a greedy algorithm that finds an IRS state within a given number of channel estimations that reduces the SI to a very low level. The proposed algorithm is illustrated in the flow graph in Fig. 3. The algorithm is divided into two parts: 1) initialization, which performs initial random measurements and 2) the actual minimization phase. Initially, the two buffers, \( S \) and \( i \), are created. The buffer length \( L \) is an input parameter of the algorithm and determines the size of the applied buffers. In our experiments, a value of 100 was determined to be reasonable for \( L \). Depending on the IRS size, a larger value may be useful here. The binary buffer \( S \) has the dimension \( L \times N_h \times N_v \).
Fig. 3. Flow graph of the proposed greedy algorithm for optimizing the IRS state for SI cancellation.

where \( N_h \) is the number of elements of the IRS in the horizontal direction and \( N_v \) is the number of elements of the IRS in the vertical direction. In \( S, L \) states of the IRS are collected, which are continuously updated as the optimization proceeds. The buffer \( i \) with dimension \( L \times 1 \) contains the measured SI magnitudes corresponding to \( S \).

In the initialization phase of the algorithm, \( S \) and \( i \) are initially populated with readings. In a loop whose number of runs is tracked in the loop counter variable \( t_c \), a buffer entry is filled in each run until \( L \) runs are reached. For this purpose, a random state is first generated for the surface by a uniformly distributed random function, and the surface’s current state matrix, referred to as \( S_t \), is set accordingly. Subsequently, the SI \( i_t \) is quantitatively determined by channel estimation and stored in the buffer \( i \). Different metrics can be used to measure SI, such as the received signal strength indicator (RSSI), the scaled amplitude of the CSI, or an actually received power measurement. Depending on the hardware used, the metric available in the respective case can be chosen. After the completed initialization, \( S \) is filled with \( L \) IRS states and \( i \) with \( L \) associated SI magnitudes.

In the optimization phase, \( S \) and \( i \) are sorted in descending order by SI magnitude in \( i \) in the first step. The sorted matrix \( S \) is then linearly weighted so that the state with the highest SI obtains the lowest weight and the state with the lowest SI obtains the highest weight. This results in the normalized ratio \( P_{\text{norm}} \) as

\[
P_{\text{norm}} = \frac{2}{L^2 + L} \sum_{l=1}^{L} i_l \cdot S_l
\]

with \( S_l \) being a subset of \( S \) with the dimension \( N_h \times N_v \) representing a single IRS state and \( L \) being the buffer length. The ratio \( P_{\text{norm}} \) is used as a measure of how often a single patch element was active, with the weighting assigning a higher factor to good IRS states than to less effective IRS states. The ratio obtained is now used as a threshold for a uniformly distributed random function to generate a new IRS state. The state is applied to the surface, and the SI is measured for the new state. If the measured SI is smaller than the largest value of the SI in buffer \( i \), the corresponding state in \( S \) and \( i \) is replaced by the new values. If a new value is found, the loop counter \( t_c \) is also reset. This process is performed in an iterative loop until the loop counter \( t_c \), which is incremented in each run, exceeds a predefined threshold \( t_{\text{lim}} \). The optimization thus proceeds until no new IRS states are found for \( t_c \) runs. As a result, the algorithm returns \( S_{ib} \), which is the IRS state for the lowest SI measured \( i_b \).

VII. WiFi CSI

In our previous work [29], we demonstrated the basic feasibility of SI cancellation using IRS by means of laboratory hardware, more precisely with a vector network analyzer. However, for deployment in realistic scenarios, cheaper and simpler hardware must be used. In addition to the pure ability to measure the channel, other requirements, such as allowed frequency bands and communication, must also be ensured. To address these aspects, we use IEEE 802.11n [24] WiFi transceivers in this work. The communication protocol, as defined in the WiFi standard [24], is not modified in this work, and standard-compliant transmission and reception are guaranteed. This ubiquitous wireless technology is available at low cost, meets all necessary licensing requirements, and is accepted and widespread among the broad population. WiFi uses OFDM as the modulation type, so channel estimation is required at the beginning of each WiFi packet for successful data transmission.

At the beginning of each transmitted IEEE 802.11n packet in WiFi, a preamble is first sent containing a short training field
The STF is used for packet start detection and coarse time synchronization. After the coarse frequency and time synchronization, the LTF is used to obtain a finer estimate of the CFO and sample frequency offset (SFO) and an estimate of the channel. Since the LTF is defined in the IEEE 802.11 WiFi standard, it is known to the receiver. By multiplying the inverse of the known LTF by the received LTF, an estimate of the channel is extracted.

The channel estimate inherent in the WiFi chipset can be extracted with appropriate software tools in certain chipsets. Widely used tools in research are available for Intel network interface card (NIC) 5200 [30], Atheros ATH9k [31], and some Broadcom chipsets [32]. Meanwhile, some chip vendors are starting to make these CSIs available to developers via API interfaces [33]. To capture the phase change of the channel $y_d(t)$ described in Section IV, which is caused by breathing, the channel must be measured under the condition of phase stability between a transmitting and a receiving antenna. Since WiFi is inherently a half-duplex system, we use two colocated NICs, as shown in Fig. 4. One NIC acts as a transmitter with one antenna that injects WiFi packets, while the other NIC acts as a receiver at the other antenna and outputs the channel estimate. The CSI in the receiving device can be written in the frequency domain for devices with a single antenna as a complex-valued vector $h$ of dimension $K$, where $K$ is the number of subcarriers used. The elements of $h$ are given by

$$h_k = |h_k| \exp(j\angle h_k)$$

in which $k$ is the subcarrier index and $\angle h_k$ refers to the angle of the complex channel coefficient $h_k$. In a standard WiFi setup, every transceiver derives its local oscillator (LO) and sampling clocks from its own crystal oscillators. This results in a number of impairments to the CSI, especially with respect to the phase of $h_k$. The estimated phase $\angle h_k$ is given by [31], [34], and [35] as

$$\hat{\angle h_k} = \angle h_k + (\lambda_{PDD} + \lambda_{SFO}) \cdot k + \lambda_{CFO} + \beta + w$$

where $\angle h_k$ is the ground-truth phase, $\lambda_{PDD}$ is the phase slope caused by the packet detection delay (PDD), $\lambda_{SFO}$ is the phase slope caused by the SFO, and $\lambda_{CFO}$ is an additive phase offset caused by the CFO. $\beta$ is a constant system phase offset, and $w$ is the measurement noise. These phase impairments have been extensively studied in various papers [31], [36], [37], [38]. However, all of these studies examined individual unsynchronized crystal oscillators in transceivers.

Note that the advanced phase correction methods applied in commodity WiFi devices, such as carrier phase tracking, are only applied to the trailing OFDM symbols of the payload. The CSI used here is obtained from the preamble only, so these procedures have no effect on the CSI.

As shown in [39], there are four major contributors to $\angle \hat{h}_k$. The PDD and SFO contribute as linear slopes, $\lambda_{PDD}$ and $\lambda_{SFO}$ with respect to the subcarrier index $k$. The CFO $\lambda_{CFO}$ is constant for all subcarriers. The contributors, PDD, SFO, and CFO, are constant for only one packet and are time varying with each packet received. $\beta$ is a time-invariant phase offset as long as the PLL is locked, i.e., the device is not rebooted, and the WiFi channel is not switched.

To estimate the phase between two WiFi NICs in a colocated antenna assembly, we need to remove these phase impairments from the measurement. To remove the $\lambda_{CFO}$ and $\lambda_{SFO}$ from (13), we use the architecture proposed in [39], referred to as wired synchronization for WiFi (WS-WiFi). In more detail, by feeding two colocated transceivers a common clock, the CFO and SFO are reduced significantly. We modify the WiFi NICs to accept the common wired LO clock, which is applicable in practice since they are colocated by definition in the application. The initial phase offset $\beta$ can be removed by a calibration process proposed in ArrayTrack [40]. However, since in our application, we are interested in the breathing pattern of a test person, which is a periodic phase change pattern, the constant offset can be neglected since it is constant over the course of the measurement.

Removing the phase impairments by applying the WS-WiFi LO synchronization and the calibration procedure described in ArrayTrack [40] results in a new improved $\angle \hat{h}_k$ of approx

$$\angle \hat{h}_k = \angle h_k + \lambda_{PDD} \cdot k + w.$$  (14)

From (14), it can be seen that besides the measurement noise, only $\lambda_{PDD}$ disturbs the phase measurement. $\lambda_{PDD}$ is a linear phase slope that depends on the subcarrier index. Thus, the influence of $\lambda_{PDD}$ can be reduced by choosing a low subcarrier index. We, therefore, use the subcarrier index $k = -1$ in this article to keep a residual $\lambda_{PDD}$ as small as possible.

**VIII. Experimental Studies Setup**

In this section, we first describe the measurement setup for the following breath-tracking experiments. Then, we describe the individual hardware components of the setup in detail and explain the parameters used for our investigations.

**A. Breath-Tracking Experiment-Floor Setup**

The presented experiments were conducted in a typical 42 m² office/laboratory environment. The floor plan of the room and the positioning of the equipment are shown in Fig. 5. A photograph of the detailed setup of the IRS and antennas is
Fig. 5. Floorplan of the experimental space shows the dimensions of the room and the relative spacing of the test positions and equipment. The test positions referred to throughout this article are marked with TP1–TP11.

Fig. 6. Setup of the breath tracking experiment. On the left, the IRS is shown at a distance of 1 m from the antennas in the center of the picture. The test person is positioned on the chair at 1.4-m distance from the antennas to the right at test position 1.

shown in Fig. 6. The IRS is set up in the center of the room on a table, with the front of the IRS facing the west direction of the room. The antennas are positioned at a distance of 1 m from the surface. The center of the antennas is centered in front of the surface in the top quarter of the IRS. The setup of the IRS and antennas is static and is not moved or manipulated during the measurements. The room is furnished with typical office furniture, desks, rolling pedestals, and computer workstations. The east front of the room has a full-length window from halfway up, and the south side is adjacent to a hallway. The northern wall and the western wall bordering the corridor are drywall. The remaining walls are load-bearing reinforced concrete walls, as are the floor and ceiling. A test subject whose breathing was to be detected sat quietly on an office chair with castors and moved to test positions 1–11, one after the other, for the various measurements. To record the ground-truth respiration signal, a NeuLog NUL-236 Chest Strap was used to record the expansion of the chest via a compression sensor.

B. Transceiver System and Antenna Assembly

The transceivers we use in our experiments are commodity WiFi transceivers that comply with the IEEE 802.11n standard. More specifically, we use Atheros ATH9k-based WiFi routers of the TP-Link N750 or WDR4300 type. The routers were patched using the Atheros CSI tool [31] and run OpenWrt, a widely used open-source operating system for WiFi routers. The routers we use each offer three antenna ports, although we only use one for each router. Thus, a variety of alternative, less expensive WiFi chipsets that support only one antenna can be used. We use two routers, one as a transmitter and the other as a receiver. The transmitter is operated as an injector, i.e., it is not integrated into an access point or client network and can thus inject packets into the wireless medium at a constant rate. The receiver is operated in the monitor mode and extracts the CSI of all packets on the set channel that are addressed to its MAC address. The CSI of the received packets is forwarded as UDP packets to a host computer via Ethernet for recording, analysis, and coordination. Matching the design frequency of the intelligent surface, WiFi channel 64, with a center frequency of 5.32 GHz at a channel bandwidth of 20 MHz, is used.

The channel estimate contains 56 complex values for data subcarriers. Unoccupied guard bands and zero subcarriers are not considered. To ensure stable phase measurements between the transmitting and receiving antennas, LO synchronization was added to the transceivers. For this purpose, the internal 40-MHz crystal oscillators of the transceivers were removed and replaced by a coaxial connection, as described in WS-WiFi [39]. An external 40-MHz clock is provided by a DG4162 function generator from the manufacturer Rigol and distributed to both routers using coaxial cables. This ensures that no CFO exists between the transmitter and receiver since both transceivers generate their carrier frequency from the same clock using PLL. The PLL does add phase noise, but this can be effectively removed by a low-pass filter over a longer observation period due to the closed-loop fashion of the PLL [41].

For the reflection of the IRS to significantly lower the SI, initial isolation in the analog domain must be realized to approximately compensate for the path loss to and from the IRS. For this purpose, e.g., RF circulators can be used, which combine transmitter and receiver on a common antenna. However, these analog RF devices are expensive, difficult to integrate, and require a very well-matched antenna or reconfigurable matching circuit [42]. The orders of magnitude of isolation that can be achieved with a circulator are in the range of 20–30 dB, but these can only be achieved under ideal conditions with a perfectly matched antenna. We conducted some experiments for this purpose with a DiTom D3C3060-6 circulator with typical Vert 2450 WiFi antennas from Ettus Research, wherein we were only able to achieve 11 dB for the analog isolation.

We, therefore, use an antenna assembly with separate transmitting and receiving antennas, which we mount at a distance of λ/2, which is realistic for a UE and easy to implement. An image of the antenna assembly is shown in Fig. 7.
The antennas of type Broadspec from Time Domain Inc. are mounted in a polyacetal (POM) plastic bracket using SMA angle adapters and point away from each other. Thus, the 180° elevation angles of the antennas point toward each other. The actual antenna characteristics were measured in an anechoic chamber for the antennas used. The normalized antenna characteristics are given in Figs. 8 and 9.

The antenna attenuation at an elevation angle of 180° can be read as −22 dB. Since both antennas used have this characteristic, analog isolation between the transmitter and receiver of 44 dB could be achieved when the antennas are ideally aligned with each other. In our experiments, we could typically achieve values of 40 dB in an indoor environment, which is sufficient for the intended use with the available IRS. The analog isolation may now seem quite high for embedding in a UE, such as a smartphone; however, this value may be relaxed in the future with the availability of larger IRSs.

In the arrangement of the antennas shown, the test subject to be detected is thus located in the azimuth plane of the antennas. If we examine the antenna pattern in Fig. 8, we observe that the antenna pattern is very good in all directions and almost has an omnidirectional characteristic. The maximum attenuation is 5 dB from the side view of the antennas. This ensures that there are no blind spots and that 360° coverage around the antenna assembly is possible.

C. IRS

For our experiments, we use an IRS with 16 × 16 elements, which can be switched in a binary fashion between two phase states 0° and 180°. The dimensions of the prototype are 0.4-m wide and 0.3-m high. The designed carrier frequency at which the shift of 180° is maintained is 5.3 GHz, which is precisely in the WiFi band at channel 64, with a center frequency of 5.32 GHz. The elements of the IRS can be configured from MATLAB via a USB interface. In our experiments, the IRS is connected to a host computer via USB so that synchronization of the switching operations with the channel measurement is ensured.

Further technical details about the IRS are described in the Appendix.

IX. RESULTS

In this section, we present the results of our investigations. First, we show the phase stability of the applied measurement system with the added LO synchronization over an extended ambient temperature range. Then, we show the performance of the presented greedy algorithm, and finally, we present the results with respect to the application of breath tracking.

A. Phase Stability of the Setup

To verify that the LO synchronization approach presented in [39] also works as expected for our case of synchronized transmitters and receivers, we first check the phase stability of the system used. For this purpose, the transmitting and
receiving antennas of the antenna assembly are detached and replaced by a connection with a coaxial cable and a 30-dB attenuator. This ensures a constant phase over a long period of time, which we examine below.

The transmitter injects WiFi packets at a rate of 400 Hz, which are received by the receiver, and the CSI is evaluated. We consider the subcarrier at subcarrier index $k = -1$ for the phase stability study. The phase of the subcarrier is recorded over a period of 10 s and plotted as a polar plot in Fig. 10. For this initial test at room temperature, we compare two measurements of 10 s each. The measurement shown in blue in the polar plot was performed without LO synchronization. It can be found that the phase of the subcarrier under consideration does not provide stable phase information across several WiFi packets. With LO synchronization applied, we obtain the measurement shown in red. With the applied synchronization, a stable phase measurement is obtained with a maximum scatter of approximately 60°. The scattering, which is significantly added by the phase noise of the PLL [41], is only a short-term effect, which is averaged out over a longer measurement period since long-term drift is not possible due to the closed-loop PLL.

In Fig. 11, the distribution of the phase for the same experiment is depicted. It can be found that for the red curve, with synchronization, the distribution centers at a stable value with small bandwidth. The phase noise can therefore be very well compensated for slow measurements, such as respiration with a typical frequency of below 1 Hz, which can be filtered very well by a moving average filter or a low-pass filter. Since PLLs have significant temperature dependence, this can introduce limitations to the system for certain applications. We perform an analysis of the phase stability of the subcarrier at subcarrier index $k = -1$ over an extended temperature range from $-20^\circ C$ to $40^\circ C$. For this purpose, the transmitter’s SMA port is still connected to the receiver’s SMA port via a coaxial cable to ensure a stable phase in the analog domain. The test is performed in two configurations. In the first, a router is placed in a thermal chamber of type VT4002 from manufacturer Vötsch Industrietechnik, and a router is placed at constant room temperature to obtain a maximum temperature difference. In the second test, both transceivers are placed in the thermal chamber so that they are kept at the same temperature and only experience an ambient temperature difference, but there is no significant temperature difference between the two devices. The ambient temperature is recorded by sensors in the chamber and in the room. In addition, the temperature of the WiFi chipsets is recorded by a DS18B20 temperature sensor from the company Dallas Semiconductor [43], which is glued directly onto the chipset.

In Fig. 12, the measured phase of the subcarrier at subcarrier index $k = -1$ is plotted versus temperature. A complete temperature cycle from $40^\circ C$ to $-20^\circ C$ ambient temperature was performed over a time of 20 min at a packet rate of 1 Hz to ensure sufficient acclimation to the continuous slow temperature drop in the chamber. Temperature alignment was ensured by comparing the on-chip temperature sensors and the ambient temperature sensors. These had a constant relation to each other throughout the measurements with a maximum deviation of $2.3^\circ C$. In the plot, the phase change over the chamber ambient temperature is shown in blue, where the transmitting router
was operated outside the chamber at a constant ambient temperature of 22 °C. The temperature throughout the experiment is shown in red. Temperature dependence of the measured phase is observed.

For the case of breath detection in indoor environments considered here, a rather slow ambient temperature change within a moderate temperature range can be assumed. Furthermore, since the two transceivers are always placed spatially close to each other by the colocated antennas, a very small differential temperature can generally be assumed. Considering the slow change in temperature over several seconds to minutes, the change in respiration is much faster. Since only a relative observation of the phase change over time is relevant for the detection of respiration, the influence of a temperature difference change can be neglected. Furthermore, if the observation period is very long, the slow increase or decrease in the phase difference due to the temperature can be removed by a simple high-pass filter with a cutoff frequency below the respiration frequency.

B. Reduction of SI and Convergence of the Greedy Algorithm

To ensure the effective operation of SI cancellation by the IRS and the convergence to good SI values with the presented greedy algorithm and Atheros ATH9k-based WiFi transceivers, we analyze the measurement of SI using the CSI. The greedy algorithm proposed in Section VI can use different metrics as a measure of SI. In addition to a relatively coarse receive level measurement using the RSSI, WiFi CSI allows for a subcarrier-based analysis. For this, we use the amplitude of the CSI extracted using the Atheros CSI tool [31]. Since the amplitude is scaled to 10 bits and does not take the AGC gain into account, it is scaled with normalization using the RSSI, and thus, a more accurate receive level estimate is achieved than with the pure RSSI. The scaled amplitude of the CSI is now averaged over all 56 subcarriers to obtain the broadest possible SI suppression in the WiFi band and used as the SI estimate for the greedy algorithm.

For verification, we run the algorithm ten times in the setup described in Section VIII-A and track the cumulative minimum during the optimization iterations. Fig. 13 shows the convergence averaged over the ten runs. The diagram shows the convergence (yellow line) of the algorithm presented in Section VI. With the applied linear weighting, the algorithm reduces the averaged amplitude as a measure of the SI to −70 after the initialization phase within 500 iterations. If the linear weighting is neglected within the greedy algorithm and all collected values in the buffer S are considered with the same weighting, the value of −70 is also reached after 1600 steps, but 1100 more iterations are required than with the linear weighting. As a baseline, an additional comparison is drawn with a random generation of IRS states as an optimization method. The “random” case shown in blue shows a barely noticeable improvement in SI over the 2500 steps considered. A single iteration of the greedy algorithm’s optimization loop requires 0.1 s for 2500 optimization iterations; hence, 250 s is needed, which is mainly limited by the execution times of the different interfaces and the maximum update rate of the IRS. With optimized control of the surface, this value can be reduced substantially. The maximum update rate of the IRS limited by the pin diodes specified in Section VIII-C is significantly higher than the packet rate achievable with WiFi.

Note that since we are working with noncalibrated commodity instruments, the convergence, or SI, cannot be given in absolute power levels but only as a relative estimate. For an quantitative absolute investigation of the SI, refer to our previous paper [29].

After successful execution and convergence of the greedy algorithm, the IRS states buffer $S$ can be analyzed for a better understanding of the found IRS state. The buffer $S$ contains the $L$ best states found during all runs at the end of the greedy algorithm. Averaging $S$ over the $L$ runs and normalizing it to 1 results in the pattern shown in Fig. 14 in the form of a heatmap. Here, the values shown represent the ratio of how many times an element was activated in the $L$ states. It is easy to observe a strong similarity with the phase distribution in Fig. 2.

The main reason for the similarity between the two figures is the spherical phase distribution over the surface. To overcome the considerable power loss due to the path loss and reflection loss of the IRS and to obtain a constructive addition of the individual paths at the receiver, an equally spherical pattern is necessary to equalize the phases of the individual paths. Therefore, the similarity between Figs. 2 and 14 makes sense, as the pattern equalizes the geometric path lengths as much as possible and results in a constructive overlap at the receiving antenna.

The obtained values for SI cancellation show a significant improvement over the nonoptimized case. The effect of buffer size $L$ on convergence will be examined in more detail in the following section.

C. Effect of the Greedy Algorithm Buffer Size

The parameter $L$ defines the buffer size of the greedy algorithm. The buffer is used to store the best $L$ states and to iteratively generate new IRS states until the greedy algorithm converges. Therefore, the parameter $L$ has a crucial impact on the functionality and performance of the algorithm. In this section, the greedy algorithm is studied for different sized $L$ in the setup described in Section VIII.
Fig. 14. Converged, averaged, and normalized buffer $S$, after running the greedy algorithm.

SI cancelation using the weighted greedy algorithm is measured for $L$ ranging from 10 to 1000. A step size of 10 was used in the interval 10–100, and a step size of 100 was used in the interval 100–1000. Thus, measurements for 19 different settings of $L$ were examined. For each setting of $L$, ten runs of the weighted greedy algorithm were conducted. The results of these microbenchmarks are shown in Figs. 15 and 16. In each case, the average of the ten runs is used for the plot. Fig. 15 shows the convergence for small $L$ in the range 10–100. Fig. 16 shows the convergence for large $L$ in the range of 100–1000.

From the microbenchmarks performed, it is found that the weighted greedy algorithm converges very quickly for $L$ smaller than 40 and reaches a stationary point within the first 500 iterations. However, the achieved minimum value of SI for these small $L$ is up to 14 dB higher than the lowest achieved SI for larger $L$. For $L$ in the range 70–100, the lowest SI is reached within the first 1500 iterations. Compared to smaller $L$, the smallest SI amplitude is thus reached later. However, the smallest SI achievable in our experiments is always reached for these sizes of $L$. For $L$ larger than 100, the convergence of the algorithm takes significantly longer than for smaller $L$. Here, the minimum possible SI is not reached in our observation period of a maximum of 2500 iterations.

It can thus be concluded that for fast convergence of the greedy algorithm, the smallest possible size for $L$ should be chosen. However, the size for $L$ is lower bounded so that it still reaches the smallest possible SI. This empirical analysis leaves 100 for $L$ as a reasonable tradeoff between minimum SI and convergence time for our application. We have therefore chosen a value of 100 for $L$ for the experiments presented throughout this article unless otherwise stated.

D. Experimental Results of IRS-Assisted Breath Tracking

The system described previously was evaluated for application to the acquisition of respiration from a test person. To capture respiration, the phase of subcarrier $k = -1$ of the received CSI data is examined and presented below. In addition, for comparison with the actual respiration pattern, a NeuLog NUL-236 chest strap sensor is used. The chest strap measures respiration using a rubber balloon that is compressed as the chest expands and is evaluated by a pressure sensor. However, since the measured absolute pressure values of the chest belt strongly depend on the pressure on the chest belt and the tightness of the belt, the amplitude is normalized to 1 over the measurement period in the following. Since the envelope of the respiration signal is essentially relevant for the recording of respiration, no information is lost as a result, but a more consistent, comparable ground-truth reading is obtained.

As an initial measurement, a measurement without optimization of the IRS was performed. For this purpose, the surface was set to the “all elements off” state, and a measurement was recorded over the period of 180 s at test position 1. During the measurement period, the subject sat quietly on an office chair and breathed regularly at the beginning, held his breath after a few breaths, and then continued to breathe regularly. The results are shown in Fig. 17. The bottom plot of the diagram shows the ground-truth breath signal measured by the chest strap. The upper plot shows the raw measurement of the phase on subcarrier $k = -1$ in blue, and the phase filtered using a low-pass filter at 0.5-Hz cutoff frequency is shown in red. The packet rate for the phase measurement is 400 Hz.
long-term stability is achieved, so by applying the 0.5-Hz low-pass filter to limit the noise to the range of a typical respiratory signal, the short-term phase jitter is also removed. The resulting signal contains little phase change and cannot be used to detect respiration.

We now apply the optimization using the greedy algorithm. The algorithm is executed once at the beginning, and the converged IRS state is retained for the measurements. The same breathing pattern of regular breathing, holding breath, and regular breathing again was measured. The results are shown in Fig. 18. The lower plot again shows the actual breathing pattern detected using the chest strap. The top plot shows the raw phase signal and the low-pass filtered phase of subcarrier $k = -1$.

The phase plot shows a much larger change of up to 400° and traces the respiration signal very clearly. It is readily apparent that by means of IRS-assisted reduction of the SI, detection of respiration is made possible. It also becomes possible to detect not only the frequency of respiration but also the pattern of respiration. The plot demonstrates, for the first time, respiration detection with commodity WiFi devices and a single colocated transceiver site in an indoor environment.

The phase pattern shown in Figs. 17 and 18 is exemplary for test position 1, and the remaining test positions 2–7 in the 1.40-m radius of the antenna setup show a comparable pattern and are not broken down in detail. We can confirm that there is no significant directionality in respiration detection. 360° detection is possible except for the angles physically obstructed by the IRS.

Fig. 19 shows the extracted phase for test position 8. It can be seen that the phase plot still contains changes that are due to the breath signal and can be easily identified. However, the absolute maximum phase change of 110° has a smaller maximum value than test positions 1–7. Test positions 8 and 9 show comparable performance.

Fig. 20 shows the extracted phase for test position 11, which is located outside the office room in the corridor. It can be seen from the figure that a respiratory signal can no longer be identified. Test positions 10 and 11 in the corridor no longer provide meaningful results. This observation is expected due to the high attenuation of the drywall and cabinets that separate the corridor from the office space.

Our experiments demonstrate that the respiration detection of a single person sitting quietly at room level range is possible with commodity WiFi hardware and a supporting IRS with only a single colocated antenna site.

X. Conclusion

In this article, we have demonstrated for the first time that IRS-assisted SI cancellation with commodity WiFi devices for respiration detection is possible with a single colocated
antenna site. We have built a prototype system that has delivered very promising results in the 5-GHz WiFi band in a typical indoor environment. We first investigated the reduction of SI using a greedy algorithm to tune the IRS and ensured the phase stability of the measurement system using LO synchronization. This made it possible, with only a single antenna site with two antennas separated from each other by λ/2, to track a phase change in the CSI data over a period of several minutes. It was possible to detect not only the breathing rate of a test subject but also the pattern of breathing. The presented approach has achieved a very good performance, which can significantly increase the quality of life in AAL scenarios since body-worn devices for monitoring vital parameters can be omitted.

**APPENDIX**

**TECHNICAL DETAILS OF THE IRS PROTOTYPE**

The IRS used throughout the experiments is an array of 16 × 16 identically structured unit cells embedded on a printed circuit board (PCB). Each unit cell offers a binary switchable resonance frequency, which results in two switchable reflection coefficients with different phases. The unit cell consists of a vertically polarized rectangular patch reflector on the top side of the PCB and a contiguous ground plane on the bottom side. The length of the patch is on the order of half of one wavelength and therefore defines the resonance frequency of the patch reflector in the unloaded state. A copper-plated through-hole connection (via) connects the patch on the top side to the anode of a PIN diode located on the bottom side of the IRS. The cathode of the PIN diode is connected to the ground plane. An eccentric position of the via affects the electrical length of the patch when the PIN diode is biased. Thus, the resonance frequency of the patch reflector can be altered by shorting the via to the ground. This causes a binary-switchable resonance frequency of the unit cell. The resulting phase response of the IRS is shown in Fig. 21, which is the phase difference of the surface’s reflection coefficient between the IRS configured as all-ON and all-OFF states. It was measured with the reflected wave vector and the incident wave vector perpendicular to the surface.

Fig. 22 shows the physical dimensions and driver circuitry of the unit cell. Standard FR4 material with a dielectric constant of εr ≈ 4.5 and a dissipation factor of tan δ ≈ 0.02 is used as the substrate. A PIN diode from the manufacturer Skyworks of type SMP1320-079LF is used. The forward current through the PIN diode of I = 1 mA is set by a biasing resistor of value R = 4.7 kΩ. A capacitor of value C = 100 pF is used for additional decoupling of unwanted ac signals. In the ON state or the OFF state, a biasing voltage of Vb = 5 V or Vb = 0 V is applied, respectively. This unit cell is an evolution of our prior work as described by [44].

The IRS can be controlled as follows: ASCII-coded text commands containing the surface state to be configured are sent to a microcontroller located on the back side of the IRS via a universal asynchronous receive–transmit (UART) interface. The microcontroller then generates a binary sequence, which is shifted into 32 daisy-chained 8-bit shift registers. Each unit cell is connected to one of the 256 outputs of the shift registers, which supply the biasing voltages to the array elements. We use the following components: 74HC565D shift registers from NXP Semiconductors, an STM32F446RET6 microcontroller from STMicroelectronics running at 180 MHz, and an MCP2221A from Microchip Technology as a USB-to-UART bridge.

With the selected PIN diode and the implemented biasing circuit, we measured a minimum switching time of approx. 4 μs to switch a unit cell from the ON state to the OFF state. This is the time required for the PIN diode to become nonconductive for RF signals. Thus, a maximum of approx. 250k IRS states can be configured in one second. With the shift registers used, one can attain a maximum of approx. 390k configurations per second. Thus, the PIN diodes limit the achievable configuration rate, which affects the time required to find a suitable IRS state for the cancellation of self-interference. To provide a preview of our future work, we measured the configuration rate of a unit cell with a modified biasing network. By omitting the decoupling capacitor C and adding a parallel resonance circuit across the biasing resistor R, which decouples the control circuit from the unit cell, a switching time of approx. 40 ns can be achieved. This leads to configuration rates of 25M configurations per second if the proposed biasing network is modified and the digital control circuitry is improved for fast switching. In this case, a control interface offering much higher data rates than UART must be used.

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