Waveform Design and Hybrid Duplex Exploiting Radar Features for Joint Communication and Sensing

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Abstract—Joint communication and sensing (JCAS) is a very promising 6G technology, which attracts more and more research attention. Unlike communication, radar has many unique features in terms of waveform criteria, self-interference cancellation (SIC), aperture-dependent resolution, and virtual aperture. This paper proposes a waveform design named max-aperture radar slicing (MaRS) to gain a large time-frequency aperture, which reuses the orthogonal frequency division multiplexing (OFDM) hardware and occupies only a tiny fraction of OFDM resources. The proposed MaRS keeps the radar advantages of constant modulus, zero auto-correlation, and simple SIC. Joint space-time processing algorithms are proposed to recover the range-velocity-angle information from strong clutters. Furthermore, this paper proposes a hybrid-duplex JCAS scheme where communication is half-duplex while radar is full-duplex. In this scheme, the half-duplex communication antenna array is reused, and a small sensing-dedicated antenna array is specially designed. Using these two arrays, a large space-domain aperture is virtually formed to greatly improve the angle resolution. The numerical results show that the proposed MaRS and hybrid-duplex schemes achieve a high sensing resolution with less than 0.4% OFDM resources and gain an almost 100% hit rate for both car and UAV detection at a range up to 1 km.

Index Terms—Joint communication and sensing (JCAS), integrated sensing and communication (ISAC), waveform design, max-aperture radar slicing (MaRS), hybrid duplex.

I. INTRODUCTION

As a popular 6G technology, joint communication and sensing (JCAS) [1]-[5] is expected to create considerable add-on values to the current wireless communication system. The widely-deployed communication infrastructures can be enhanced to provide radar services like traffic control and surveillance, drone detection [6], and railway obstacle detection. JCAS can also be realized by the various mobile communication devices in the scenarios of autonomous driving [7][8], smart home, and health care.

The communication and radar are two different ways of utilizing electromagnetic wave. The idea of dual-function design can be traced back to the 1960s according to [2]. It attracts more and more research attention in recent years, and there are many driving factors including (1) the spectrum has been well exploited for two single systems, and the joint spectrum utilization is expected to improve the efficiency and flexibility [2][9]; (2) the hardware designs are becoming similar, as they both have a technology trend of multiple antennas and digital baseband [4][10]; (3) the information fusion and mutual reinforcement of two functions can improve performance, especially for autonomous vehicles [7][8].

From the perspective of system integration, existing JCAS schemes are usually categorized into three levels: coexistence, cooperation, and joint design. In coexistence schemes, the two sub-systems transmit the signals in overlapped resources. Thus, it requires interference management to reduce the performance loss [11]. The cooperation of two functions helps to manage the interference via spectrum sharing [12] and beamforming [13][14]. The cognitive radar [15] can be seen as a special form of cooperation scheme where the secondary function dynamically senses the spectrum of the primary one and finds idle resources to transmit, whose idea is similar to that of cognitive radio [16]. To reduce spectrum sensing overhead, a sub-Nyquist radar can be used to sense the communication resources at low rates [17]. Unlike coexistence and cooperation, the joint design aims to integrate the two sub-systems into one. Most existing joint designs are either radar-centric or communication-centric, and joint optimization methods can be used to get a better trade-off.

The radar-centric system embeds data information in the radar waveform. A popular radar scheme is chirp radar [18] due to its optimal auto-correlation, constant modulus, and simple self-interference cancellation (SIC). Chirp spread spectrum (CSS) modulation has been used in Long Range Radio (LoRa) [19], and similarly, JCAS can utilize chirp to transmit information via phase and frequency modulation [20][21]. Apart from chirp radar, frequency-agile radar [22] was also utilized for JCAS which employs an index modulation of sub-carriers. Furthermore, the beam pattern of radar was proposed for communication via sidelobe control and waveform diversity [23]. A severe problem of radar-centric JCAS schemes is the low data rate.

The communication-centric system uses communication signals to sense. In early times, spread-spectrum communication was used, and JCAS using phase-coded waveform [10] was proposed. Nowadays, orthogonal frequency division multiplexing (OFDM) has been widely used in commercial communication systems. In the OFDM-based passive radar system [24][25], a passive node receives communication signals and realizes bi-static sensing. For the

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more common mono-static sensing scenario, full duplex is usually required to realize an OFDM-based JCAS [26][27]. With full duplex, the base station can get OFDM echo signals for distance and Doppler estimation. Furthermore, joint optimization methods can be used to make a good trade-off. In [28], the power allocation of each sub-system is optimized to maximize a compound mutual information-based function. The optimized data is filled into unused sub-carriers to minimize the estimation variances [29]. Furthermore, the waveform over multiple antennas can be optimized for both two sub-systems [30].

The full-duplex OFDM-based joint design is very promising in terms of spectral efficiency, hardware sharing, and system integration. However, it relies on the assumption of in-band full-duplex communication, which is still very challenging to achieve in commercial systems. Although many full-duplex technologies and prototypes [31][32] have been proposed, the advantage of double throughput is not that attractive as full duplex also requires double antennas. The full duplex of a single antenna requires nonreciprocal circulators using ferrite materials, which is hard to be miniaturized [32]. In the MIMO evolution of 3GPP Release 18 [33], an aggregation of two half-duplex sub-bands named sub-band full duplex, instead of the in-band full duplex, is on the schedule. Therefore, this paper proposes to realize JCAS without the dependence on full-duplex communication.

Apart from radar-centric and communication-centric schemes, some works [34][35] proposed to use brand-new waveforms. Orthogonal chirp division multiplexing (OCDM) [34] was proposed to replace the Fourier transform kernel in OFDM with the Fresnel transform. A DFT-spread-OFDM transceiving scheme [35] is also proposed to generate a similar waveform. Although OCDM-like methods combine OFDM and chirp, they lose the advantages of both waveforms including the multi-path robustness of OFDM and efficient SIC of chirp. To keep the advantages of both signals, the two functions can still use separate waveforms [36][37]. The waiting time of pulsed radar is used for communication in [36], where full-duplex communication is still a must. Our previous JCAS work [37] uses OFDM to generate chirp using partial resources and improves the sensing spectrum efficiency via random sensing sampling. Using OFDM to generate chirp has also been used in MIMO radar [38].

In this paper, mono-static sensing is assumed as the majority of radars are in this mode [18]. Compared with communication systems, radar systems have unique advantages including (1) waveform superiority in terms of zero-auto-correlation sequence and constant modulus; (2) simple SIC in the analog domain; (3) aperture-dependent resolution instead of resource-dependent; (4) virtual aperture (VA) of multiple-input and multiple-output (MIMO) radar [39] to increase angle resolution. The idea of this paper is to fully utilize these excellent radar features by modifying a regular half-duplex communication system. The main contributions of this paper include:

1. This paper proposes max-aperture radar slicing (MaRS) waveform to greatly reduce resource overheads without loss of radar resolution. Different structures of MaRS are proposed and compared, among which comb MaRS performs best. It also keeps the advantages of constant modulus over time, zero auto-correlation when measuring range, and very simple analog SIC.

2. This paper proposes joint space-time target extraction and space-time adaptive processing (STTE-STAP) algorithm to recover range-velocity-angle information from comb MaRS under strong environmental clutters. Also, space-time hierarchical processing (STHP) is proposed to avoid large-dimension matrix inverse and increase calculation accuracy when the matrix is too ill-conditioned to inverse.

3. A hybrid duplex JCAS scheme is proposed, which comprises half-duplex communication and full-duplex radar as the SIC of the radar signal is much easier. A small antenna array is added to the communication system and specially designed for sensing. It is combined with the communication array to form a large VA, which greatly enhances angle resolution.

4. This paper provides simulations to verify the proposed schemes, which include the practical factors of self interference (SI), noise, and clutters. The simulation results show that high-resolution sensing is achieved with less than 0.4% resource overhead in an OFDM communication system, which means the communication can use almost 100% of resources to ensure throughput.

The rest of the paper is organized as follows. Section II shows the system model in terms of communication, radar, and angle of arrival (AoA). Section III analyzes the sensing aperture and proposes MaRS for resource overhead reduction. Section IV proposes an enhanced MaRS structure named comb MaRS and designs joint space-time algorithms for it. Section V proposes a hybrid duplex for JCAS which does not require full-duplex communication and enables a large VA. Section VI shows the simulation results of the proposed schemes. Section VII briefly concludes the paper. In this paper, (·)*, (·)T, and (·)H denote conjugate, transpose, and Hermitian transpose of a matrix or vector. ⊗, o and mod represent the Kronecker product, Hadamard product, and modulo operation.

II. SYSTEM MODEL

This paper considers a JCAS scenario containing cellular communication and mono-static sensing [18] as shown in Fig. 1. For communication, the uplink and downlink are time-division. For sensing, the station transmits a known signal and receives the echo signal of targets to obtain the sensing information.

A. Communication Sub-system

The communication sub-system is a common half-duplex cellular system, which uses OFDM-based waveforms in the downlink and uplink. The uplink signal is sent by the conventional communication-only user. In the downlink, the base station transmits an OFDM signal of
$x(t) = \frac{1}{\sqrt{N}} \sum_{m=0}^{M-1} \text{rect}(t-mT_{\text{sym}}) \sum_{n=0}^{N-1} X_{x,m,n} e^{j2\pi f(t-mT_{\text{sym}})}$, \hspace{1cm} (1)

where $N$, $M$, and $\Delta f$ denote the total number of sub-carriers, the number of symbols in a JCAS frame, and the sub-carrier spacing, respectively. Function $\text{rect}(t)$ is a time window of

$$\text{rect}(t) = \begin{cases} 1, & 0 \leq t < T_{\text{sym}} \\ 0, & t < 0 \text{ or } t \geq T_{\text{sym}} \end{cases}$$ \hspace{1cm} (2)

The cyclic prefix (CP) of OFDM is omitted here for simplicity. The used bandwidth of $x(t)$ is $B = N\Delta f$.

The transmit power is assumed to be $P_T$. The received power of the communication signal is

$$P_R = \frac{P_T G_T G_R \lambda^2}{(4\pi)^2 R^4},$$ \hspace{1cm} (3)

where $G_T$, $G_R$, $\lambda$, and $R$ denote the transmit gain, the receive gain, the wavelength, and the distance, respectively.

**B. Sensing Sub-system**

This paper considers a radar scenario of detecting multiple targets, or the searching mode. The MIMO radar method [38][39] is used, which transmits omnidirectional beams and receives the echo signals in multiple antennas. It avoids scanning of transmit beams and allows flexible digital receive beamforming. Assume the antenna number is $L$. The phased array radar method uses both transmit and receive beamforming to gain a high echo power, but the cumulative time of one specific beam is $1/L$ of the omnidirectional beam as all $L$ beams require to be scanned in the searching mode.

The sensing sub-system reuses the OFDM baseband to generate radar signals. Take chirp for example, and the chirp signal in one OFDM symbol is

$$x_{\text{chirp}}(t) = e^{j\pi \mu t^2}, 0 \leq t < T_{\text{sym}},$$ \hspace{1cm} (4)

where $\mu$ is the chirp rate. The frequency of this signal rises from 0 to $\mu T_{\text{sym}}$. To utilize the whole bandwidth of $B$, the chirp rate is

$$\mu = \frac{B}{T_{\text{sym}}} = N\Delta f.$$ \hspace{1cm} (5)

The chirp signal is described in the discrete-time form of

$$x_{\text{chirp}}[k] = e^{j2\pi\mu k/N T_{\text{sym}}}, 0 \leq k < N.$$ \hspace{1cm} (6)

It can be generated by the frequency domain OFDM data of

$$X_{\text{chirp},m,n} = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} x_{\text{chirp}}[k] e^{-j2\pi k n N^{-1}}.$$ \hspace{1cm} (7)

That is to say, the OFDM baseband can be reused to generate the chirp signal.

The echo signals experience a channel from the transmitter to the target, a reflection of the target, and a channel from the target to the receiver. The received power of the echo signal is

$$P'_R = \frac{P_T G_T G_R \lambda^2 \sigma}{(4\pi)^2 R^4},$$ \hspace{1cm} (8)

where $\sigma$ is the radar cross section of the target. Compared with the communication receive power, the sensing receive power decreases much faster with the distance rising. Fortunately, sensing can use signal processing methods to accumulate the received energy of sensing targets.

Apart from the echo signal of targets, the receiver also receives the strong SI signal and clutter signal. The power of SI is assumed to be $P_{SI}$. SI is not only from scattering but only from near-field coupling, and the received SI signal in the $l$-th antenna is

$$y_l(t) = \sqrt{P_{SI}} e^{j\Omega t} x(t)$$ \hspace{1cm} (9)

where $\Omega$ is a random variable with a uniform distribution in $[0, 2\pi]$.

This paper also considers the practical environmental clutter. The area reflectivity $\sigma^0$ [18] is used to describe the reflection of area scatterers, which can be calculated by the Georgia Tech Research Institute (GTRI) model given by

$$\sigma^0 = A(\delta + C)\theta^\delta \exp \left( \frac{-D}{1 + 0.1\sigma_k^\delta} \right),$$ \hspace{1cm} (10)

where $\delta$ is the grazing angle, $\sigma_k$ is the RMS surface roughness, and the parameters $A$, $B$, $C$, and $D$ depend on the clutter type and radar frequency [40].

**C. AoA Model**

The steering vector of a uniform linear array (ULA) in Fig. 2(b) is

$$a(\psi, L, d) = \left[ e^{j2\pi \frac{0\psi \sin \theta}{\lambda}} \quad e^{j2\pi \frac{1\psi \sin \theta}{\lambda}} \ldots \quad e^{j2\pi \frac{(L-1)\psi \sin \theta}{\lambda}} \right]^T,$$ \hspace{1cm} (11)

where $\psi$ and $d$ denote the AoA and the spacing of antennas.

This paper considers a $L_x \times L_y$ uniform rectangular array (URA), which is more common than the linear array in practical communication systems. As shown in Fig. 2(a), the azimuth angle $\theta \in [0, \pi]$ and elevation angle $\phi \in [-\pi, \pi)$ are
used to decide the target direction. The direction vector from URA to the target is
\[ \hat{a} = [\cos \phi \sin \theta, \sin \phi \sin \theta, \cos \phi], \] (12)
The AoA of URA at the xoy-plane can also be modeled by the AoA of the x-axis ULA and the y-axis ULA, which are denoted as \( \psi_x \) and \( \psi_y \), \([0, \pi]\). Using these two angles, the direction vector is expressed by
\[ \hat{a} = [\sin \psi_x, \sin \psi_y, \sqrt{1-\sin^2 \psi_x - \sin^2 \psi_y}], \] (13)
The two kinds of representation can be converted to each other easily using the equation (12) and (13). This paper uses \( \psi_x \) and \( \psi_y \) as they can be directly estimated using fast Fourier transform (FFT) methods. The 2-dimension (2D) steering vector of a uniform rectangle antenna is
\[ \mathbf{a}_{2D}(\psi_x, \psi_y, L_x, L_y, d_x, d_y) = \mathbf{a}(\psi_x, L_x, d_x) \otimes \mathbf{a}(\psi_y, L_y, d_y), \] (14)
where \( \otimes \) is the Kronecker product. As equation (11) can be seen as a column vector of \( \mathbf{F}_x \), where \( \mathbf{F}_x \in \mathbb{C}^{N \times L} \) is the discrete Fourier transform (DFT) matrix. The 2D vector in equation (14) can be seen as a column vector of a synthesis matrix of
\[ \mathbf{F}_{x, y} = \mathbf{F}_x \otimes \mathbf{F}_y, \] (15)
This synthesis matrix can be viewed as a 1-dimension (1D) space-domain FFT to simplify the explanation, and the 2D-FFT is used in practical computation as FFT reduces the complexity. The complexity of applying 2D-FFT to an \( L_x \times L_y \) matrix is \( L_x \log(L_x) + L_y \log(L_y) = L \log(L_x L_y) \), which is the same as that of the 1D-FFT applying an \( L_x L_y \times 1 \) vector.

III. SENSING APERTURE AND MARS

A. Sensing Aperture Analysis

The sensing function requires a larger aperture in the frequency, time, and space domain to ensure the resolution of range, velocity, and angle, respectively. The range \( R \) of the target is estimated by the round trip delay \( \tau \) using \( R = c\tau/2 \), where \( c \) is the speed of light. Therefore, the range resolution is related to the time resolution of the sensing signal, which is decided by the bandwidth \( B \), or more generally the frequency-domain aperture. The range resolution is
\[ \Delta R = \frac{c}{2B}. \] (16)

The radial velocity of the target is estimated by \( v = cf_x/2f_c \), where \( f_c \) is the carrier frequency, and \( f_d \) is the Doppler frequency of the target. One coherent processing interval (CPI) of \( T_{\text{CPI}} = M T_{\text{sym}} \) is used to measure \( f_d \). \( T_{\text{CPI}} \) can also be seen as the time-domain aperture, which decides the radial velocity resolution as
\[ \Delta v = \frac{c}{2T_{\text{CPI}} f_c}. \] (17)

The space-domain aperture \( D = Ld \) decides the angle resolution. This resolution can be measured by the beamwidth, e.g., the 3 dB beamwidth. It can be calculated by
\[ \Delta \psi \approx a \frac{\lambda}{D} \text{ rad}. \] (18)

The coefficient \( a \) is the beamwidth factor [40], which is related to the value of \( \phi \). To get rid of this impact, this paper uses the resolution of \( \sin \psi \) as
\[ \Delta \sin \psi = \frac{\lambda}{D}. \] (19)

B. The Proposed MaRS

Based on the analysis of sensing aperture, high-resolution radar sensing requires a large aperture, instead of a large number of resources. If a large aperture is realized by a small resource overhead, the sensing resolution can be ensured, and the communication performance is not affected as the majority of time-frequency resources can be used for it. Based on this idea, this paper proposes a sensing waveform design named MaRS.

Fig. 3(a) shows the resource overhead of the conventional chirp signal, which consumes all the resources in the time-frequency-space aperture. Here, continuous OFDM symbols are assumed to be allocated for simplicity. The time-frequency domain resource overhead is measured by the time length and bandwidth. In an OFDM system, they can be represented by the number of sub-carriers and symbols. The space-domain resources can be measured by the number of antennas.

A time-frequency-space MaRS signal is shown in Fig. 3(b). The transmission of such a signal can be divided into three stages: (1) in the first symbol, the JCAS station transmits the chirp signal using \( N \) sub-carriers, and only one pair of transmit antenna and receive antennas is used; (2) in the second symbol, the JCAS station transmits the single-tone sensing signal and all the antennas are used; and (3) in the left \( (M-2) \) symbols, the JCAS station transmits the single-tone sensing signals, and
only one pair of transmit antenna and receive antennas is used. In these three stages, a high-resolution estimation of distance, angle, and velocity can be achieved. The resolution is as high as that of conventional schemes in Fig. 3(a) except that the radial velocity resolution is slightly reduced as the useful aperture is reduced from $MT_{sym}$ to $(M-2)T_{sym}$. This example shows how MaRS keeps the sensing aperture to gain high resolution with much fewer resources. However, the association of estimated range, velocity, and angle cannot be decided, which makes it only work in the ideal single-target and clutter-free scenario. Apart from the association problem, it is hard to utilize the remaining space resources for communication as the time-frequency resource has been used.

To avoid these problems of time-frequency-space MaRS, a time-frequency MaRS signal is proposed as in Fig. 3(c). Using such a signal, all the antennas are used. Two structures of time-frequency MaRS are shown in Fig. 4. The frequency-agile (FA) MaRS in Fig. 4(a) transmits a random sub-carrier in each OFDM symbol, which is similar to the sensing part of [22]. The communication part is different as the left OFDM resources are for data transmission, instead of using low-efficient index modulation. FA MaRS only acquires phase information in each symbol for both distance estimation and speed estimation, and thus the performance is limited.

L-shape MaRS is proposed to use dedicated resources for range and velocity estimation as shown in Fig. 4(b). It is divided into two stages: (1) in the first symbol, the JCAS station transmits the chirp signal using $N$ sub-carriers; (2) in the left $(M-1)$ symbols, the JCAS station transmits the single-tone sensing signals. The resource overhead is greatly reduced from $MN$ to $(M+M-1)$. For example, if $N = 1024$ and $M = 64$, the sensing resource overhead is reduced to only 1.6% of that of the classical chirp scheme. For this MaRS waveform design, both time-space and frequency-space can be jointly processed. In this way, multi-target separation and clutter suppression can be achieved, which is introduced in the next section.

IV. ALGORITHMS AND ENHANCED MaRS STRUCTURE

A. Basic MaRS Algorithm

One important step of radar signal processing is to extract the target from the clutters. It is easy to realize via a conventional chirp signal, and the processing steps are shown in Fig. 5. The conventional chirp signal in a JCAS frame is

$$x_c(t) = \sum_{m=0}^{M-1} \text{rect}(t - mT_{sym}) e^{j\omega m [l - mT_{sym}]}, 0 \leq t < MT_{sym}. \quad (20)$$

The echo signal of $x_c(t)$ resulting from the scattering of a point target at the distance $d = ct/2$ is

$$y_{echo}(t) = a_{2D}(\omega_s, \psi_s, L_s, d_s, d_{rev}) h_0 x_c(t - \tau) e^{j2\sigma t (t - \tau)}, \quad \tau \leq t < MT_{sym} + \tau.$$  \quad (21)

where $h_0 \in \mathbb{C}$ is the round trip channel coefficient in the symbol with index 0.

The chirp receiver mix each row of $y_{echo}(t)$ with $x_c(t)$ in the analog domain and the intermediate frequency (IF) signal can be obtained as

$$y_{IF}(t) = a_{2D}(\omega_s, \psi_s, L_s, d_s, d_{rev}) h_0 e^{j2\sigma t (t - \tau)} e^{j2\sigma t (t - \tau - \tau)}. \quad (22)$$

In one symbol, the phase variation is usually omitted as $f_0T_{sym} \ll 1$. The IF SI signal is a direct-currency (DC) signal as $\tau = 0$, and it can be removed via a DC offset cancellation (DCOC) circuit [41]. Further enhancement can be added to DCOC to make it decay faster [42].

The receiving window is set as $[MT_{sym}, (m+1)T_{sym}]$, which is not the same as the definition domain of equation (22). The signal out of the receiving window is omitted in the analysis. The continuous-time IF signal after filtering can be sampled to form a matrix $Y_F \in \mathbb{C}^{N \times MN}$. The processing steps of $Y_F$ is shown in Fig. 5. The space-domain $L$-FFT in this figure can be replaced by the equation (15) when URA is used. The targets and clutters can be separated in the distance-velocity matrix, as the clutter is at the zero-velocity bin [6]. These operations can be seen as a 3D-FFT, and the total complexity is $O(LMN \log(LMN))$.

The time-frequency MaRS signal can be written as

$$x_m(t) = \text{rect}(t)e^{j2\pi t^2} + \sum_{m=1}^{M-1} \text{rect}(t - mT_{sym}) e^{j2\pi t^2}, 0 \leq t < MT_{sym}. \quad (23)$$

By replacing $x_c(t)$ with $x_m(t)$ in equation (20), $Y'_{echo}(t)$ is obtained. We can mix this receiving signal with the

$$x_{IF}(t) = a_{2D}(\omega_s, \psi_s, L_s, d_s, d_{rev}) h_0 e^{j2\sigma t (t - \tau)} e^{j2\sigma t (t - \tau - \tau)}. \quad (24)$$

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By replacing $x_c(t)$ with $x_m(t)$ in equation (20), $Y'_{echo}(t)$ is obtained. We can mix this receiving signal with the
transmitting signal and obtain
\[ y'_1(t) = \begin{cases} y_{f_1}(t), & t \leq T_{sym} + \tau, \\ y_{f_2}(t), & mT_{sym} + \tau \leq t < (m+1)T_{sym} + \tau, m \in [1, M], \end{cases} \]
where \( y'_{f_1}(t) = a_{2D}(\alpha, \beta)g_{0}e^{j\omega_0 t}e^{j2\pi f_1 t}, \) and \( y'_{f_2}(t) = a_{2D}(\alpha, \beta) \)
\[ h_{0}e^{j2\pi f_1 t}e^{j2\pi f_2(t-\tau)}. \]

The IF signal has two parts. The first part \( y'_{IF_1}(t) \) is the same as that in the conventional chirp scheme, and the SI can be easily removed via DCOC. In the second part \( y'_{IF_2}(t) \), the SI of sensing signal as well as the scattering point with \( f_d = 0 \) are becoming DC signal which can also be removed. \( y'_{IF_2}(t) \) also suffers from the SI of the communication signal at other subcarriers, and the receiver can separate them via a lowpass filter. In this paper, the Butterworth filter is used, which has a frequency response of
\[ |H(f)|^2 = \frac{1}{1 + (f/f_{cut})^{2a}}. \]

where \( f_{cut} \) is the cutoff frequency, and \( a \) is the order of Butterworth filter. In practice, the guard band can be further inserted between the single-tone sensing signal and the communications signal if the filter cannot effectively cancel the interference from communication.

These two parts of the IF signal after filtering is sampled as \( Y_{f_1} \in \mathbb{C}^{e \times N} \) and \( Y_{f_2} \in \mathbb{C}^{e \times (M-1) \times N} \). 2D-FFT can be applied to these two matrices. Before 2D-FFT, \( Y'_{f_2} \) is pre-processed to reduce FFT complexity as
\[ Y'_{f_2} = \sum_{n=0}^{N-1} y'_{f_2,n} \sum_{n=0}^{N-1} y'_{f_1,n}. \]

The angle-velocity matrix \( R' \) is used to get the angle and velocity estimation. Using the angle estimation, the distance information can be obtained in the angle-distance matrix \( R' \). The complexity of 2-time 2D-FFT in this proposed method, \( O(LN \log(LN) + LM \log(LM)) \), is much less than that of the 3D-FFT, \( O(LMN \log(LMN)) \), in conventional method.

B. Comb-MaRS and Advanced Algorithm

In the L-shape MaRS, the target distance information cannot be achieved when the clutters are strong and abundant, e.g., the rank of clutters in \( y'_{f_1} \) is already \( L \). To suppress the clutter more effectively, comb MaRS structure and STAP-based method are proposed.

The comb MaRS is shown in Fig. 7(a). It transmits the wideband chirp signal in several OFDM symbols, which enables STAP. The comb MaRS signal is expressed as
\[ x_{cm}(t) = \sum_{m=i_w}^{i_w+M-1} \text{rect}(t - mT_{sym})e^{j\omega t - mT_{sym}} \]
\[ + \sum_{m=0}^{M-1} \text{rect}(t - mT_{sym})e^{j2\pi f_1 t}, 0 \leq t < M T_{sym}. \]

where \( i_w \in [0, M-1] \) with \( w = 0, 1, ..., M-1 \) is the symbol index of the wideband chirp signal. The comb MaRS signal is a generalized version of the basic MaRS signal.

Comb MaRS enables STAP after space-time, or angle-velocity, estimation, and STTE-STAP is proposed. Similar to equation (21) to (24), we can get the echo signal of \( x_{cm}(t) \), \( y''_{echo}(t) \), and mix it with the transmitting MaRS to
\[ y''_1(t) = \begin{cases} y'_{f_1}(t)h_{0}e^{j\omega t}e^{j2\pi f_1 t}, & t \leq (i_w + 1)T_{sym}, \end{cases} \]
\[ \text{for } t \leq T_{sym} + \tau \leq (i_w + 1)T_{sym} + \tau, \]
\[ y''_2(t) = a_{2D}(\alpha, \beta)h_{0}e^{j\omega t}e^{j2\pi f_2(t-\tau)}, \]
\[ \text{for } mT_{sym} + \tau \leq t < (m+1)T_{sym} + \tau, m \in [0, M] & m \neq i_w. \]
These two parts of the IF signal after filtering are then sampled using a frequency of $1/B$ as $Y''_{F1} \in \mathbb{C}^{L' \times W'}$ and $Y''_{F2} \in \mathbb{C}^{M' \times W' \times N'}$. The second part $Y''_{F2}$ is transformed into $Y''_{S2} \in \mathbb{C}^{(M'W') \times W}$ using the similar operation of equation (26). Unlike $Y''_{S2}$ in equation (26), $Y''_{S2}$ is discontinuous at arbitrary $i_w$ in $[0, M-1]$. This discontinuous matrix is expanded to $Y''_{E2} \in \mathbb{C}^{L' \times M}$ via the simple nearest-neighbour interpolation method. The $m$-th column vector of $Y''_{E2}$ is

$$y''_{E2,m}^* = \begin{cases} y''_{S2,m,g(a_n)+1,m} = i_w, \\ y''_{S2,m,g(a_n)+m} = i_w, \end{cases}$$

where function $g(m)$ returns the number of $i_w$ values satisfying $i_w < m$. This equation requires $i_w$ to be non-zero. With $Y''_{E2}$,

$$R^*_2 = F_{2,0}^\top (L_2, L_3, F, y''_{E2}) = F_{2,0}^\top (L_2, L_3) Y''_{E2} F_M$$

In $R''_2$, STTE can be done as $K$ targets can be extracted with their velocity and angle estimations $\hat{v}_k, \hat{\psi}_v, k$. FFT in this estimation can be replaced by super-resolution estimation methods like MUSIC, which will be applied in simulations.

With the estimation from STTE, STAP can be realized. $Y''_{F1}$ is reshaped to a space-time signal of

$$Y_{ST} = \begin{bmatrix} y_{F1,0} & y_{F1,1} & \cdots & y_{F1,N-1} \\ y_{F1,0}^* & y_{F1,1}^* & \cdots & y_{F1,N-1}^* \\ \vdots & \vdots & \ddots & \vdots \\ y_{F1,(W'-1)N} & y_{F1,(W'-1)N+1} & \cdots & y_{F1,(W'-1)N+1}^* \end{bmatrix} \in \mathbb{C}^{W' \times N}.$$ (33)

The rank of clutter correlation [18] in STAP is estimated by

$$\text{rank} \{S_c\} = L + (M-1)\beta.$$ (34)

where $\beta$ is clutter ridge slope, which is proportional to the speed of the radar platform. As the radar platform is static in this paper, we can have $\beta = 0$. The space-time steering vector is constructed using the estimated target information as

$$\hat{a}_{ST,k} = \hat{a}_{n,k} \otimes a_{ST}(\hat{v}_k, \hat{\psi}_v, L_1, L_2, d_1, d_2) \in \mathbb{C}^{W' \times 1}.$$ (35)

where $\hat{a}_{n,k} = \left[ e^{j\frac{2\pi f_c}{c} d_1} e^{j\frac{2\pi f_c}{c} d_2} \cdots e^{j\frac{2\pi f_c}{c} d_{W'1}} \right]^\top \in \mathbb{C}^{W' \times 1}.$

The interference correlation matrix is used to suppress the clutter and multi-target interference, which is estimated by

$$\hat{S}_I = (Y_{ST}^* Y_{ST})/N.$$ (36)

The distance spectrum can be obtained using STAP together with a distance-dimension FFT as

$$d_k = \begin{bmatrix} \hat{a}_{HST,k}^\top (Y_{ST}^* F_N) \end{bmatrix}^\top \in \mathbb{C}^{N \times 1},$$ (37)

where $d_k$ is a vector representing $N$ distance bins, and the bin of the largest value is the estimation of target range $R$. Thus, the distance, velocity and angle information of multiple targets can be obtained, and the procedure is also shown in Fig. 7(b). $\hat{S}_I$ is assumed to be full-rank with the presence of noise as it is easy to have $WL > N$. When $WL > N$, this method also works by replacing the inverse with the pseudo inverse.

An eigen decomposition can be done to $\hat{S}_I$ as

$$\hat{S}_I = X \begin{bmatrix} \lambda_1 \\ \lambda_2 \\ \vdots \\ \lambda_{WL} \end{bmatrix} X^\top.$$ (38)

Without loss of generality, the eigenvalues are assumed to be descending. The eigenvalues of $\lambda_1$ to $\lambda_{KL-1}$ are from the target and clutters which is proportional to the signal power, while the eigenvalues of $\lambda_{KL-1}$ to $\lambda_{WL}$ are from the noise which is proportional to the noise power.

One problem is that when the signal-to-noise ratio (SNR) is higher, the condition number of $\hat{S}_I$, $\lambda_1/\lambda_{WL}$, is larger, which means $\hat{S}_I$ is more ill-conditioned. Note that $\hat{a}_{ST,k}$ contains two parts, the true value and estimation error. The true value matches with the eigenvectors from the signal, while the estimation error part cannot. It makes the inverse of $\hat{S}_I$ amplifies the estimation errors of $\hat{a}_{ST,k}$. Such a noise amplification won’t happen if $\hat{a}_{ST,k}$ is used to construct $\hat{S}_I$ as the eigenvectors with large eigenvalues match $\hat{a}_{ST,k}$. However, such calculation cannot be done as the space-time information of environmental clutters is hard to be obtained. Therefore, there can be a mismatch of $\hat{a}_{ST,k}$ and $\hat{S}_I$ in STTE-STAP as $\hat{a}_{ST,k}$ is not used in $\hat{S}_I$.

To avoid this problem, a noise-like item with a power of $\varepsilon$ is added to improve the small eigenvalues as

$$d_k = \begin{bmatrix} \hat{a}_{HST,k}^\top (R_1 + \varepsilon I_{WL})^{-1} (Y_{ST}^* F_N) \end{bmatrix}^\top \in \mathbb{C}^{N \times 1},$$ (39)

where $I_N$ is an $N \times N$ unit matrix. To make $\varepsilon$ effective to affect $\hat{S}_I$, it can be set to

$$\varepsilon_{st} = \frac{1}{WL} tr(R_1).$$ (40)

where $tr(\cdot)$ is the trace of a matrix. The complexity of STTE-STAP is $O(LM \log(LM) + W^2L^2N^2 + W^2L^2 + W^4L^2 + kWLN)$. When $\varepsilon \rightarrow 0$, the matrix inverse can be removed and (39) degrades a matched filtering of

$$d_k = \begin{bmatrix} \hat{a}_{HST,k}^\top (Y_{ST}^* F_N) \end{bmatrix}^\top \in \mathbb{C}^{N \times 1}.$$ (41)

The complexity is reduced to $O(LM \log(LM) + kWLN)$.

This paper further proposes another method named space-time hierarchical processing (STHP) to alleviate $\hat{a}_{ST,k}/\hat{S}_I$ mismatch and matrix inverse complexity. The space-time vector $\hat{a}_{ST,k}$ is divided into $\hat{a}_{ST} \otimes \hat{S}_{ST,ST}$ in the $W$-dimension time domain, zero-forcing (ZF) is used to suppress the static clutter using the combining weight vector of

$$w_{ST,k} = \tau_{ST,k} = \begin{bmatrix} \tau_{ST,k} \\ \tau_{ST,k} \end{bmatrix} = \left[ a_{ST,k} \ 1_{WL} \right],$$ (42)

where $1_{WL}$ is a $W \times 1$ all-one vector, and (·) is the pseudo-inverse. It combines $Y_{ST}$ into $Y_{S} \in \mathbb{C}^{W \times N}$. After the time-domain combination, the static clutter is forced to zero power. Then, a matched filtering or adaptive processing can be done, which can be represented by

$$w_{ST,k} = \hat{a}_{ST,k}^\top (R_1 + \varepsilon I_{WL})^{-1},$$ (43)
When $\varepsilon = 0$, it is adaptive processing, while when $\varepsilon \to \infty$, it is a matched filter. A trade-off can also be made for $\varepsilon = \varepsilon_m$. The distance detection result is $d_i = w^H S d_i$. Compared to the STAP, the proposed STHP filters clutters by $W$-dimension ZF and then separates multiple targets by $L$-dimension processing like STAP. The clutter suppression performance can be ensured in STHP, and the matrix inverse complexity is reduced from $W^3L^3$ to $L^3$. The total complexity of STTE-STHP is $O(LM\log(LM) + L^2N + KL^2 + L^3 + KLN)$, and it reduces to $O(LM\log(LM) + KLN)$ when $\varepsilon \to \infty$.

V. Hybrid Duplex and Virtual Aperture

A. Hybrid Duplex Scheme

Full duplex, or SIC, is a must for sensing, while it may not be required by communication. The communication usually employs half-duplex schemes to avoid the complex full-duplex hardware including 2-fold antennas and sophisticated radio frequency front-end. In SoftNull full-duplex system [43], the performance gain degrades greatly with the path loss gain increasing as SI becomes harder for the low-SNR users. Also, the low latency advantage of full duplex can be realized by sub-band full duplex using two non-overlap half-duplex sub-bands, which will be discussed in 3GPP release 18 [33].

Unlike SIC of the random data signal in communication, the sensing signal SI is fixed and can be designed to obtain a simple SIC, and thus radar can have very simple SIC processing. Pulse radar can use a shared antenna to transmit and receive via switching the circulator, while chirp radar usually filters the SI after mixing the received signal with the local transmitted signal. It is easy for radar to achieve full duplex, and this paper proposes a hybrid-duplex JCAS scheme of half-duplex communication and full-duplex radar as shown in Fig. 8.

The proposed hybrid-duplex JCAS adds a small antenna array and the corresponding RF chain to realize full-duplex radar, instead of requiring two-fold antennas. Filters are required to realize the analog SIC for radar sensing. Also, a STTE-STAP/STHP baseband is required, which can be implemented in a shared processing unit with the half-duplex OFDM baseband. The mixer in the RF chain of the large communication antenna array switches to the mixing of the local chirp signal in the $i$-th OFDM symbol.

Note that in the proposed scheme, the antenna cannot simultaneously transmit and receive. When the large antenna array is receiving the single-tone sensing signal, the downlink communication service cannot be supported. Therefore, a 100% duty ratio MaRS signal cannot be used in practice. A MaRS example of a 1/3 duty ratio is shown in Fig. 9. In the communication-only symbols, both uplink and downlink can be supported, while in the symbols containing single-tone sensing signals, only uplink is supported.

B. Virtual Aperture in Hybrid Duplex

The proposed hybrid structure can be further utilized to realize VA as the small sensing-dedicated array can be specially designed. VA is a crucial technology to increase the angle resolution of MIMO radar [44]. The spacing of the small array can be designed to fit the large antenna array, and a uniform VA can be formed as in Fig. 10. The antenna spacing of the small array is $L_d i_x$ in the column direction and $L_d i_y$ in the row direction.

VA is a unique feature of the sensing function as the sensing signal experiences the round-trip channel. The sensing transmitting steering vector is

$$a_x = a(\psi_x, \eta_x, S, d_x) \otimes a(\psi_x, \eta_x, S, d_x), \quad (44)$$

while the sensing receiving steering vector is

$$a_R = a(\psi_x, L_x, \lambda/2) \otimes a(\psi_x, L_x, \lambda/2). \quad (45)$$

The channel information of total $\eta_x \eta_y L_x L_y$ MIMO channels forms a steering vector of large VA as

$$a_{\psi x} = (a(\psi_x, \eta_x, L_x, d_x) \otimes a(\psi_x, \eta_x, L_x, d_x)) \otimes (a(\psi_y, \eta_y, L_y, d_y) \otimes a(\psi_y, \eta_y, L_y, d_y)) \quad (46)$$

$$= a(\psi_x, \eta_x, L_x, d_x) \otimes a(\psi_y, \eta_y, L_y, d_y)$$

which is equivalent to a URA of $\eta_x \eta_y L_x L_y$ antennas with the
array spacing \( d_a \) and \( d_b \) in Fig. 10. Using the VA, array size can be greatly improved to \( n_a n_b \) times.

The construction of VA requires every antenna in transmit array sends orthogonal signals. There are three ways to send orthogonal signals, including: time division [44], orthogonal code spreading [38][44], and cyclic shift [34][35]. These methods are shown in Fig. 11 using the wideband chirp signal as an example. The first two methods can also be used for the single-tone signal, while the last one only works for the wideband chirp signal. Further, these methods can be used within an OFDM symbol or inter OFDM symbols. This paper employs the inter-symbol time-division method. The received sensing signal in these symbols is

\[
Y_0 = \left[ Y_{1,1}, \ldots, Y_{n_a,1}, Y_{1,2}, \ldots, Y_{n_a,2}, \ldots, Y_{1,n_b}, \ldots, Y_{n_a,n_b} \right] \in \mathbb{C}^{n_a n_b \times N}, \tag{47}
\]

where \( Y_{a,b} \) denotes the OFDM symbol transmitted by the antenna \((a, b)\) in the small antenna array and received by the large antenna array. The \( \eta_a \eta_b \) symbols in the \( Y_0 \) can be converted to a VA receiving signal in one symbol as

\[
Y_{VA} = \left[ Y_{1,1}^T, \ldots, Y_{n_a,1}^T, Y_{1,2}^T, \ldots, Y_{n_a,2}^T, \ldots, Y_{1,n_b}^T, \ldots, Y_{n_a,n_b}^T \right] \in \mathbb{C}^{n_a n_b \times N}. \tag{48}
\]

Using this signal conversion, the VA can be easily applied to the methods in the previous sections.

C. Filter Design

For the wideband chirp signal in MaRS, the filter design is the same as the conventional chirp radar. However, filtering the single-tone signal in MaRS is more challenging than that of Doppler radar. The single-tone signal in MaRS is discontinuous as the MaRS signal has a duty ratio lower than 100% to support downlink communication service. Also, the receive array receives the single-tone sensing signal together with the uplink communication signal.

This paper proposes a 2-step filtering method to cancel both the uplink communication signal and the sensing SI as shown in Fig. 12. There are \( M \) pulse repetition intervals (PRI) in one JCAS frame. One PRI contains two parts including the sensing-communication-shared signal and the communication-only signal. The shared signal is passed to a low-pass filter (LPF), which removes the uplink communication signal. Then, a sample and hold (S/H) module is used to integrate the received sensing signal and hold the integral value. During the analog integral, the sum in Equation (25) is done, which greatly saves storage. Every value is held for one PRI, and these values form a DC-biased staircase single-tone signal. In the last step, a band-pass filter (BPF) is used to remove the DC bias and shape edges of staircases, and the sing-tone signal with target Doppler information can be obtained.

VI. NUMERICAL RESULTS

A. Parameter Settings

This paper considers a simulation scenario of Fig. 1. The simulation parameters are listed in Table 1. The received SI power \( P_{SI} \) is usually lower than the transmit power by 30–72 dB according to a survey of full duplex [45]. Here, \( P_{SI} \) is assumed to be 50dB lower than the transmit power. A CP length of 8.33 us is assumed in the simulation to avoid interference of consecutive chirp symbols for VA, which covers a maximum sensing range of 1.25 km. If the maximum sensing range is 625 m, the 3GPP standard defined 60 kHz extend CP length of 4.17 us can also be used. The GTRI empirical model [40] of grass is used for environmental clutters. Since the GTRI model does not provide the distance information, this paper divides a 2 km x 1 km area into massive 1m x 1m grids and uses the grid center to calculate the distance, which can be generated with unit transmit power.

| Parameter | Value |
|-----------|-------|
| Carrier frequency, \( f_c \) | 5 GHz |
| Sub-carrier spacing, \( \Delta f \) | 60 kHz |
| Bandwidth, \( B \) | 122.88 MHz |
| Total sub-carrier number, \( N \) | 2048 |
| Average sensing TX power | 10 dBm (Car) 45 dBm (UAV) |
| Noise power density | –174 dBm/Hz |
| PRI | 0.25 ms |
| CPI | 125 ms |
| Sensing OFDM symbol number, \( M \) | 500 |
| Sensing symbol length | 16.67 us |
| Large URA size, \( (L_x, L_y) \) | (8, 8) |
| Small URA size, \( (\eta_x, \eta_y) \) | (2, 2) |
| Radar cross section | 100 m² (Car) 0.02 m² (UAV) |
| Sensing range | 100–1000 m |
| Speed range | 5–300 km/h (Car) 5–80 km/h (UAV) |
| Clutter model | GTRI Model @ 5GHz |
| Butterworth filter order | 5 |
| Uplink transmit power | 23 dBm |
| Uplink Node Distance | 100 m |
for one time and multiplied by true transmit power in different simulation cases.

The power allocation strategy of L-shape MaRS and comb MaRS is simply set as allocating the chip part with 10-fold average transmit power and the single-tone part with \((M - 10W)/(M-W)\) of the average transmit power. The parameters of comb MaRS are \(W = 4, [i_0, i_1, i_2, i_3] = [1, 3, 6, 10]\) for car detection and \([i_0, i_1, i_2, i_3] = [1, 6, 16, 30]\) for UAV detection. The height of the UAV is 0~100 m. Using the parameters in this table, the sensing range resolution is 1.22 m. The angular resolution \(\Delta\sin\psi\) is 1/4 without VA and 1/8 with VA, which corresponds to a resolution of 14.5° and 7.18° at \(\psi = 0°\). The speed resolution is 0.24 m/s or 0.864 km/h, while the maximum ambiguous speed is \(\pm 120\) m/s or \(\pm 432\) km/h.

B. Sensing Performance Simulation

The detection hit rate is defined as the percentage of targets in different drops which are successfully detected in the optimal resolution unit or the neighbour resolution unit. The neighbour resolution unit is allowed as the ground truth may lie between two resolution units and the round-trip distance of every pair of antennas for one target can locate in two resolution units. Fig. 13 shows the detection hit rate comparison of different types of MaRS. In this comparison, VA is enabled while the super resolution is not used for fair comparison. FA MaRS performs badly, and it can only be used for car detection at a very close range. This is because that FA MaRS gets only the phase information in every symbol for both range and Doppler estimation. The strong static clutter signals are not removed or suppressed before detection, which greatly limits the performance. Comb MaRS performs best and achieves an over 90% detection hit rate up to 1000 m for both the two scenarios. L-shape MaRS performs worse than comb MaRS. The gap is larger for UAV detection as it is difficult to separate small-RCS targets from strong clutters. L-shape MaRS uses STTE-STAP \((W = 1, \varepsilon = 0)\) for car detection and STTE-STAP \((W = 1, \varepsilon \to \infty)\) for UAV detection. Comb MaRS uses STTE-STHP \((\varepsilon \to \infty)\) for both two scenarios.

Different methods are compared in terms of distance. In Fig. 14(a), the detection hit rate reduces to under 90% at around 600 m. Also, a performance degradation at 100m is found for STTE-STAP \((\varepsilon = 0)\), which results from \(\hat{\mathbf{S}}_{\text{ST},L}\) mismatch problem. In Fig. 14(b), the detection hit rate decreases to under 90% around 500m. The \(\hat{\mathbf{S}}_{\text{ST},L}\) mismatch problem still exists for STTE-STAP \((\varepsilon = 0)\). In this UAV detection scenario, STTE-STAP with \(\varepsilon = \varepsilon_M\) and \(\varepsilon \to \infty\) performs much worse due to weaker clutter suppression. Note the clutter cannot be suppressed by higher transmit power, as the clutter power also increases with transmit power. The simulation results of Fig. 14(a) and (b) show that STTE-STAP suffers from \(\hat{\mathbf{S}}_{\text{ST},L}\) mismatch problem for \(\varepsilon = 0\) and clutter interference for \(\varepsilon = \varepsilon_M\) and \(\varepsilon \to \infty\), while STTE-STHP avoids these problems.

Fig. 14(c) and (d) show the detection hit rates when VA is enabled. The finer angular resolution unit brought by VA provides a more accurate time-space estimation, which improves the space-time processing performance a lot. In the car detection scenario, STTE-STAP \((\varepsilon = \varepsilon_M/\epsilon \to \infty)\) performs best, which is close to 100% in all range bins. In UAV detection, STTE-STHP \((\varepsilon = \varepsilon_M/\epsilon \to \infty)\) performs best which is close to 100% in 100~900m. In both two scenarios, STTE-STAP \((\varepsilon = 0)\) still suffers from the \(\hat{\mathbf{S}}_{\text{ST},L}\) mismatch problem in the near range, and STTE-STHP \((\varepsilon = 0)\) also has a little bit of degradation due to this problem. The strong clutter interference still limits the UAV detection performance of STTE-STAP \((\varepsilon = \varepsilon_M/\epsilon \to \infty)\). Fig. 14(e) and (f) further employs the MUSIC method to get super-resolution space-time vector estimation to improve the performance. The super-resolution ratio is 10-fold. In both two scenarios, all methods gain an almost 100% detection hit rate, apart from STTE-
Fig. 15. The normalized RMSE performance comparison. (a)–(d) are the RMSE of \( \sin(\psi_x) \), \( \sin(\psi_y) \), \( v \) and \( R \) in car detection scenario, and (e)–(h) are the RMSE of \( \sin(\psi_x) \), \( \sin(\psi_y) \), \( v \) and \( R \) in UAV detection scenario. STTE-STHP \((\epsilon \rightarrow \infty)\) is used in this simulation.

STAP in UAV detection \((\epsilon =\epsilon_M / \epsilon \rightarrow \infty)\). The \( \hat{\psi}_t \) mismatch problem is solved as the estimation becomes more accurate.

The simulation case in Fig. 14 cannot show the root mean square errors (RMSE) as the ground-truth distance is fixed which makes the estimation error biased due to the resolution unit. A sensing case of setting random distance between 100m and 1000 m is assumed, and multiple targets \((K = 5)\) are also considered. Fig. 15 shows the normalized RMSE performance comparison, which normalizes RMSE by the resolution unit without VA. Only the errors of the successful detection are taken into account as the unsuccessful detection may have large random errors, the successful detection rates are in Fig. 16. As mentioned before, \( \Delta \sin(\psi_x) \) and \( \Delta \sin(\psi_y) \) are \( 1/4 \), which is equivalent to \( 14.5^\circ \) at \( 0^\circ \) angle, \( \Delta v \) is 0.24 m/s, and \( \Delta R \) is 1.22 m. The RMSE caused by the resolution unit is as

\[
\int_{-\frac{1}{2}}^{\frac{1}{2}} x^2 dx = 6/\sqrt{3} \approx 0.2887.
\]  (49)

Fig. 16. The random-distance and multi-target detection hit rates of different receiving algorithms. (a) and (b) are the performance of FFT-based method without VA for car and UAV detection, (c) and (d) are the performance of FFT-based method with VA for car and UAV detection, and (e) and (f) are the performance of MUSIC-based method with VA for car and UAV detection.

This value can be observed in different parameters for the normalized RMSE of FFT-based methods without VA.

In the car detection scenario of Fig. 15(a)–(d), VA can decrease RMSE of both two angle estimations to 1/2, and the super-resolution method further increases the RMSE accuracy by around 4-fold. For the velocity estimation, the MUSIC-based method reduces RMSE to lower than 1/3. For the range estimation, different schemes achieve similar RMSE. MUSIC-based methods use MUSIC for angles and velocity, while cannot use it for distance as there are only \( W \) distance measurement samples. Similar results can be found for UAV detection in Fig. 15(e)–(h). The results of lower transmit power fluctuate as the corresponding hit rate is low as shown in Fig. 16. One main difference is that the MUSIC-based increase the RMSE performance by more times. Using MUSIC, the RMSE accuracy of \( \sin(\psi_x) \) and \( v \) are improved to around 7 and 9 times. The reason is that UAV has a more free space distribution as the height is random. The RMSE performance of \( \sin(\psi_y) \) is not increased as the values of \( \sin(\psi_y) \) are still mainly focused on angles around \( 0^\circ \) with the simulation setting of a large distance and limited height values.

For the multiple target detection with random distances in Fig. 16, the detection hit rate increases with the transmit power \( P_x \), and it reaches a saturated value as there are non-SNR contributory factors like environmental clutters and
The random-distance and single-target detection hit rates of MUSIC-based methods with VA for (a) car and (b) UAV detection.

Fig. 17. The random-distance and single-target detection hit rates of MUSIC-based methods with VA for (a) car and (b) UAV detection.

Fig. 18. The sensing resource overhead comparison of different sensing waveform.

The proposed MaRS observes the large aperture while consuming less communication overhead. In this way, MaRS still gains the radar advantages of constant-modulus, zero-correlation and simple SIC. Time-frequency MaRS is suggested to be used, including FA MaRS, L-shape MaRS, and comb MaRS. Comb MaRS performs best as it enables joint space-time processing, and STTE-STAP and STTE-STHP algorithms are proposed. The latter avoids large-dimension matrix inverse and increases the calculation accuracy when SNR is too high. Apart from the waveform and algorithm, the hardware structure of the hybrid duplex is proposed. The hybrid duplex adds a small sensing-dedicated antenna array to the existing half-duplex communication array. The small sensing-dedicated antenna array is designed with large antenna spacing and forms a large VA in the space-domain together with the communication array. Only the full duplex of sensing signal is required, which can be realized by analog filtering. The method of filtering discontinuous single-tone signals is also proposed, and sensing-used OFDM symbols can be either in the uplink or downlink. The numerical results verify these proposed schemes and show that the range, velocity, and angle resolution of a large time-frequency-space aperture is gained with the resource-limited MaRS and hardware-efficient hybrid duplex.

VII. CONCLUSIONS

This paper proposes the MaRS waveform design for JCAS which allows sensing to gain a large aperture and communication to gain a large number of resources. MaRS is made up of classical radar waveforms, and OFDM hardware is reused to generate it. In this way, MaRS still gains the radar advantages of constant-modulus, zero-correlation and simple SIC. Time-frequency MaRS is suggested to be used, including FA MaRS, L-shape MaRS, and comb MaRS. Comb MaRS performs best as it enables joint space-time processing, and STTE-STAP and STTE-STHP algorithms are proposed. The latter avoids large-dimension matrix inverse and increases the calculation accuracy when SNR is too high. Apart from the waveform and algorithm, the hardware structure of the hybrid duplex is proposed. The hybrid duplex adds a small sensing-dedicated antenna array to the existing half-duplex communication array. The small sensing-dedicated antenna array is designed with large antenna spacing and forms a large VA in the space-domain together with the communication array. Only the full duplex of sensing signal is required, which can be realized by analog filtering. The method of filtering discontinuous single-tone signals is also proposed, and sensing-used OFDM symbols can be either in the uplink or downlink. The numerical results verify these proposed schemes and show that the range, velocity, and angle resolution of a large time-frequency-space aperture is gained with the resource-limited MaRS and hardware-efficient hybrid duplex.

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