Frequency-Hopping MIMO Radar-Based Communications: An Overview

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Abstract—Enabled by the advancement in radio frequency technologies, the convergence of radar and communication systems becomes increasingly promising and is envisioned as a key feature of future 6G networks. Recently, the frequency-hopping (FH) MIMO radar is introduced to underlay dual-function radar-communication (DFRC) systems. Superior to many previous radar-centric DFRC designs, the symbol rate of FH-MIMO radar-based DFRC (FH-MIMO DFRC) can exceed the radar pulse repetition frequency. However, many practical issues, particularly those regarding effective data communications, are unexplored/unsolved. To promote the awareness and general understanding of the novel DFRC, this article is devoted to providing a timely introduction of FH-MIMO DFRC. We comprehensively review many essential aspects of the novel DFRC: channel/signal models, signaling strategies, modulation/demodulation processing and channel estimation methods, to name a few. We also highlight major remaining issues in FH-MIMO DFRC and suggest potential solutions to shed light on future research works.

Index Terms—Multi-Functional RF Systems (MFRFS), joint communication and radar/radio sensing (JCAS), dual-function radar-communication (DFRC), frequency hopping (FH), MIMO, timing offset and channel estimation.

I. INTRODUCTION

Enabled by the advancement in radio frequency (RF) technologies and signal processing, the convergence of multi-functional RF systems becomes increasingly promising and is envisioned as a key feature of future 6G networks [1]. Among numerous RF systems, including wireless communications, radio sensing, mobile computing, localization, etc., the former two have achieved significant progress recently on their integration. This is evidenced by several timely overview, survey and tutorial papers [2]–[5].

Driven by the spectrum scarcity cost saving, co-existence between the two RF functions has been investigated in the past few years, with focus on mitigating the interference between the two functions [2]. Thanks to the shared commonalities in terms of signal processing algorithms, hardware and, to some extent, system architecture, joint communication and radar/radio sensing (JCAS), also referred to as dual-function radar-communication (DFRC), is emerging as an effective solution for integrating wireless communication and radio sensing [3]. Substantially different from the co-existence of the two RF functions, JCAS/DFRC aims to design and use a single transmitted signal for both communication and sensing, enabling a majority of the transmitter modules and receiver hardware to be shared.

The design of JCAS/DFRC can be communication-centric or radar-centric [4]. The former performs radar sensing using ubiquitous communication signals, e.g., IEEE 802.11p [6] and IEEE 802.11ad [7], whereas the later embeds information bits into existing radar waveform or specifically optimized dual-function waveforms [4]. Regarding the communication-centric JCAS, it is worth mentioning a recently proposed perceptive mobile network (PMN) [3]. Integrating sensing function into mobile networks, the PMN is envisaged to revolutionize future 5G and beyond networks by offering ubiquitous sensing capabilities for numerous smart applications, such as smart city/factory. Compared with the communication-centric counterpart, radar-centric designs, which has also been typically referred to as DFRC, generally have superior radar sensing performance, given the sensing-dedicated waveform.

Driven by automotive applications, early DFRC tended to employ frequency modulated radar waveforms [8]. The advancement of MIMO radars has made the recent DFRC designs leaning towards this new radar paradigm [4], [5], [9]. Some researchers optimize the beam pattern of a MIMO radar to perform conventional modulations, such as phase shift keying (PSK) and amplitude shift keying, using sidelobes in the MIMO radiation patterns; others optimize radar waveform to perform non-traditional modulations, such as waveform shuffling and code shift keying (see [5], [9] and the references therein). Most previous DFRC works embed one communication symbol within one or multiple radar pulses; hence the communication symbol rate is limited by the pulse repetition frequency (PRF) of the underlying radar.

In some recently reported research [10]–[15], frequency-hopping MIMO (FH-MIMO) radar is introduced to DFRC, which breaks the above limit and substantially increases the communication symbol rate to multiples of (e.g., 15 times) PRF. Pioneering in conceiving the novel DFRC architecture based on FH-MIMO radar (FH-MIMO DFRC), the research reported in [10]–[15] focuses on analyzing the impact of information modulation on the radar ranging performance. Yet, little attention is paid to effective implementation of data communications. Some latest research reported in [16]–[18] develops methods to address the practical issues, e.g., channel estimation and synchronization, in FH-MIMO DFRC.

Motivated by the rapidly increasing interest in DFRC and...
the lack of awareness and general understanding of FH-MIMO DFRC, this article aims to provide a timely introduction of the novel DFRC architecture. We start by discussing the potential applications and channel scenarios of FH-MIMO DFRC and also the commonly used signal model in the literature. Then, we survey the existing signaling strategies for FH-MIMO DFRC, illustrate their modulation and demodulation methods, and also compare the signaling strategies from various aspects. We further discuss the issue of channel estimation in FH-MIMO DFRC, reporting recently developed solutions. Finally, we highlight the major issues unsolved in FH-MIMO DFRC and suggest potential solutions to shed light on future research directions. It is noteworthy that this article provides a better coverage of the communication aspects, of FH-MIMO DFRC, which are rarely dealt with in prior works.

II. SCENARIOS AND SIGNAL MODEL FOR FH-MIMO DFRC

FH-MIMO radar is pulse-based and hence is likely to be employed in applications requiring a large range coverage, such as long-range air-surveillance [19]. Besides, such radar is generally placed in high altitudes, “looking” ahead and above, so as to reduce ground clutters and co-frequency interference from terrestrial wireless systems. Therefore, FH-MIMO DFRC is promising for providing ground-to-air (G2A) communications for different types of aircraft, as depicted in Fig. 1. The aircraft can be an unmanned aerial vehicle (UAV), a civilian aeroplane (CAP) or a high-altitude platform (HAP). With careful designs, FH-MIMO DFRC can greatly benefit the underlying RF functions, as detailed below.

Benefiting Radar: FH-MIMO DFRC could potentially provide a radar with secure, long-range, low-latency, and high-speed wireless broadband connections to airplanes or warships over very wide areas (hundreds to thousands of kilometers, at least one-way link). The communication links provided by FH-MIMO DFRC are achieved via the careful reuse of the powerful radar infrastructures, without compromising the radar capabilities in detecting, tracking and identifying targets. Compared with typical communication systems, modern/future radar systems have transmitters with much higher power, much larger antenna arrays with very narrow beams and beam steering capability, and receivers with much higher sensitivity. Therefore, FH-MIMO DFRC can provide almost unparalleled communication links that typical communication systems, such as VHF radio and satellite, can hardly provide. Besides, the integrated solution can achieve significant saving on cost, size, and weight, compared to having separate wireless communication and radar devices.

Benefiting Mobile Communications: The G2A link provided by FH-MIMO DFRC may also help support seamless wireless coverage, as expected to be realized by future 6G networks. In particular, FH-MIMO DFRC can contribute to building the integrated space and terrestrial network (ISTN) which is envisioned to be at the core of 6G communication systems [20]. To this end, employing satellites, in combination with FH-MIMO DFRC, may provide a more cost-effective solution to providing wireless connectivity for people and vehicles in remote rural areas and in the air, as well as at sea, as compared with the solutions solely relying on satellites. Against the above potential scenarios, we depict the channel models suitable for FH-MIMO DFRC below.

A. Channel Model Suitable for FH-MIMO DFRC

The channel distribution for FH-MIMO DFRC can vary with the altitude of an airborne user equipment (UE). The typical altitudes of UAVs, CAPs and HAPs are $10^3$ m, $10 \times 10^3$ m and $20 \times 10^3$ m above the sea level, respectively [21]. Therefore, in FH-MIMO DFRC, the following flat-fading channel distributions are likely to be present.

1) Rayleigh channel is likely to exist between the radar and a UAV, particularly when the line-of-sight (LoS) path is blocked by a high-rise building and there exist numerous scattering paths; see Fig. 1.

2) Rician channel can be common between the radar and a CAP, where only few NLoS paths are present and are possibly much weaker than the LoS path;

3) Additive white Gaussian noise (AWGN) channel can prevail between the radar and an HAP where the LoS dominates the channel with negligible NLoS paths.

The geometric Saleh-Valenzuela (SV) channel model [22], with flexible parameters, can be adapted for depicting the three distributions. For ease of illustration, assume that a single antenna is equipped at the UE, which is also widely considered
in the related works [10]–[17]. Accordingly, the SV channel model can be given by
\[ h = \sum_{p=0}^{P-1} \beta_p a_M(u_p) \]  
(1)
where \( a_M(u_p) \) denotes the \( M \)-dimensional steering vector of the radar transmitter array in the direction of \( u_p \). The \( m \)-th element of \( a_M(u_p) \) is \( e^{-j m u_p} \) \((m = 0, 1, \cdots, M-1)\). Here, \( u_p \) is the beamspace-domain angle-of-departure (AoD) [22] which is given by \( \pi \sin \phi_p \) with \( \phi_p \) denoting the physical direction; see Fig. 1 for the case of \( p = 0 \).

Depending on the values of \( P \) and \( \beta_p \) \( \forall p \), the SV channel can be simplified to the three classical channels illustrated above. Specifically, the SV channel model can become
1) Rayleigh channel, given \( P \gg 1 \) and \( \text{var} \{ \beta_p \} \approx \text{var} \{ \beta_{p'} \} \forall p, p' \);  
2) Rician channel, given \( P > 1 \) and \( \text{var} \{ \beta_0 \} \gg \text{var} \{ \beta_p \} \forall p > 0 \); and  
3) AWGN channel, given \( P = 1 \).

So far, most of works only considered AWGN channels for MIMO-DFRC, with exceptions in e.g., [17].

B. Signal Model

The FH-MIMO radar is based on fast frequency hopping. Namely, each pulse is divided into \( H \) sub-pulses, also referred to as hops [14]. The centroid frequency of the transmitted signal changes randomly across hops and antennas, as illustrated in Fig. 2. Denote the radar bandwidth as \( B \). By dividing the frequency band evenly into \( K \) sub-bands, each sub-band has the bandwidth of \( B/K \). Out of the \( K \) centroid frequencies, \( M(< K) \) frequencies are selected to be the hopping frequencies per hop, one for each antenna. Let \( k_{hm} \) denote the index of the sub-band selected by antenna \( m \) at hop \( h \). To ensure the waveform orthogonality of FH-MIMO radar, the following constraints are imposed for the radar [23],
\[ k_{hm} \neq k_{hm'} (\forall m \neq m', \forall h); \Delta \triangleq BT/K \in \mathbb{I}_+, \]  
(2)
where \( T \) is the time duration of a hop.

At hop \( h \), the \( m \)-th antenna of the radar transmitter transmits a single-tone signal, i.e.,
\[ s_{hm}(i) = e^{j2\pi k_{hm} i} \Delta, \quad i = 0, 1, \cdots, L - 1, \]  
(3)
where \( L = \lceil T/T_s \rceil \) is the number of samples per hop and \( T_s \) is the sampling interval. Here, \( \lceil \cdot \rceil \) rounds to the closest integer. Based on the channel model given in (1) and the radar-transmitted signal given in (3), the UE-received signal can be modeled as
\[ y_h(i) = h^H s_h(i) + \xi_h(i), \]  
(4)
s.t. \[ s_h(i) = [s_{h0}(i) F_{h0}(i), \cdots, s_{h(M-1)}(i) F_{h(M-1)}(i)]^T \]
where \( h^H \) denotes the conjugate transpose, and \( \xi_h(i) \) is an AWGN. Note that the above signal model is obtained based on a perfect timing, which may be challenging to realize in practice, as will be discussed in Section III-B. Also note that \( F_{hm}(i) \) is multiplied to the radar signal to indicate the information embedded in the radar waveform. In different signaling schemes, \( F_{hm}(i) \) can vary, as reviewed in the following section.

III. OVERVIEW OF EXISTING SIGNALING STRATEGIES FOR FH-MIMO DFRC

Several signaling strategies have been specifically designed for FH-MIMO DFRC. Overall, the existing strategies can be categorized into two groups, one based on conventional phase modulations [10]–[13] and the other exploiting the diversity in frequency hopping [14]. In this section, the existing signaling strategies are first reviewed with their modulation and demodulation processing illustrated. Then the strategies are compared from various aspects.

A. Information Modulation

1) Phase Modulations: In [10], [11], phase shift keying (PSK) is introduced to FH-MIMO DFRC by embedding PSK phases into radar signals. More specifically, a constant phase \( F_{hm}(i) = e^{j2\pi h_m} \) is taken in (4), where \( h_m \in \Omega_J (J \geq 1) \) and \( \Omega_J = \{ 0, 1, \cdots, 2^{J-1} - 1 \} \) is the \( J \)-bit PSK constellation. In [12], the differential PSK (DPSK) is proposed to replace PSK, so as to reduce range sidelobes and out-of-band transmission. To further improve the spectral shape of the modulated radar signal, the continuous phase modulation (CPM) is introduced in FH-MIMO DFRC [13]. The underlying principle is that signals with smoother modulating phase correspond to lower out-of-band transmission.

2) Hopping Frequency-based Modulation: In [14], the combinations of hopping frequencies are used to convey information bits, referred to as frequency hopping code selection (FHCS). As illustrated earlier, only \( M \) out of \( K \) sub-bands are selected each hop as hopping frequencies. Thus, there are \( C_K^M \) combinations of hopping frequencies, and the combinations can be employed to convey up to \( \lceil \log_2 C_K^M \rceil \) number of information bits per radar hop, where \( \lceil \cdot \rceil \) rounds towards the negative infinity.

3) Comparing Two Modulation Methods and Discussions: Both modulation methods will lead to changes to the original FH-MIMO radar system, but in a different manner. The phase modulations change the phases of radar signals, and the phases can severely over antennas and hops, depending on the information bits to be sent. In contrast, FHCS leaves the phases unaltered but changes the hopping frequencies over hops. A windfall of the phase modulations is that the periodic spikes in the sidelobe regions of the range ambiguity function are suppressed, which will be explained shortly. This beneficial feature is not shared by FHCS-modulated waveform which has similar range ambiguity function to the original radar waveform, since both waveforms are essentially based on randomly selected hopping frequencies. Nevertheless, a prominent advantage of FHCS over phase modulations is the reduced complexity in information demodulation and the improved robustness against channel estimation errors, as will be illustrated in Section III-B.

Fig. 3 compares the range ambiguity functions of an FH-MIMO radar under different modulations. We see the aforementioned spikes in both the original radar waveform and the FHCS-modulated one. These spikes are attributed to the reuse of sub-bands as hopping frequencies over hops; refer to [10] for an in-depth analysis. We also see from Fig. 3 that
Pulse spikes are suppressed by the BPSK modulation. This is because BPSK scrambles the phases of the original radar waveform over hops and antennas, hence preventing the coherent accumulation in the sidelobe regions of the range ambiguity function.

B. Information Demodulation

We proceed to describe the methods that demodulate the information embedded in radar signals. As seen from (4), the signals transmitted by different antennas of the radar transmitter are superimposed in the time domain. Nevertheless, based on the waveform orthogonality as depicted in (2), the signals can be separated in the frequency domain. The L-dimensional discrete Fourier transform (DFT) of $y_n(i)$ yields

$$Y_h(l) = \beta_0 e^{-jmu_0} e^{j\pi \mu l} \delta(l - l_{hm}^*), \quad l_{hm}^* = k_{hm} \Delta,$$  

(5)

where $P = 1$ is taken as in prior works [10]–[16] and $\delta(l)$ is the Dirac delta function. The noise term is dropped for illustration convenience.

1) FHCS Demodulation: Based on $Y_h(l)$, an FHCS symbol can be readily demodulated by identifying the peaks in the magnitude of $Y_h(l)$. The indexes of the peaks result in the estimated set of $\{l_{hm}^* \forall m\}$. Based on the relation between $l_{hm}^*$ and $k_{hm}$ given in (5), we can obtain the combination of hopping frequencies used in the radar hop. The estimated combination is then used to demodulate the FHCS symbol.

2) PSK Demodulation: Given the simple signal model, PSK is used to illustrate the demodulation of the phase modulation-based signaling schemes; while DPSK and CPM can be demodulated similarly. Provided $k_{hm}$ is known by the UE, $l_{hm}^*$ is then obtained based on (5). Extract the value of $Y_h(l)$ at $l = l_{hm}^*$, leading to

$$\hat{Y}_{hm} = Y_h(l_{hm}^*) = \beta_0 e^{-jmu_0} e^{j\pi \mu l_{hm}^*}.$$  

(6)

Provided that channel parameters are known, we can suppress $\beta_0 e^{-jmu_0}$ and demodulate the PSK symbol $\phi_{hm}$.

3) Discussions: From the above description, the following information (I1)–(I3) are necessary for demodulating information symbols in FH-MIMO DFRC:

(I1) Accurate timing;

(I2) $k_{hm}$ — the hopping frequency used by antenna $m$ at hop $h$; and

(I3) $\beta_0 e^{-jmu_0}$ — the channel response.

Note that (I1) is applied to obtain (5) from (4) so that inter-hop and inter-antenna interferences are avoided. To this end, both FHCS and PSK demodulations implicitly apply (I1), since the demodulations are performed based on (5). As for (I2) and (I3), they are not required for FHCS demodulation and are explicitly used in the PSK demodulation.

Note that none of the information (I1)–(I3) is easy to acquire in practice. For the first two information (I1) and (I2), a specific synchronization link may be required to establish between the radar and UE. Such a link can make the radar require extra resources, e.g., hardware and frequency etc. More specifically, an extra antenna and correspondingly RF chain can be required for synchronization, since the existing antennas of the radar transmitter are fully occupied for radar detection. Besides, a different frequency band (from the original radar operating frequency) can be required to establish the synchronization link so as to reduce the mutual interference between radar and communications. The acquisition of the third information (I3) is subject to that of the first two, since inter-antenna and inter-hop interference will be incurred without an accurate timing.
C. Comparing Signaling Schemes for FH-MIMO DFRC

The aforementioned signaling schemes are compared in terms of their impact on radar performance, their communication performance and the necessary information for demodulation. The comparison is summarized in Table I. Note that the radar ambiguity functions and spectral containment of the signaling schemes have been comprehensively compared in [15]. The data rates of different schemes are provided in the works [10]–[14]. Yet, the necessary information is overlooked in most of the existing works which consider ideally known channels.

We notice that DPSK demodulation does not require (T2) in quasi-static channels, since the channel response is suppressed in the differential processing [12]. In contrast to phase modulation-based schemes, FHCS does not require (T2) or (T3), since it pursues to estimate (T2) from the magnitude of the frequency-domain signal, as illustrated in Section III-B1. Strictly speaking, FHCS can be demodulated without requiring (T1), provided that the timing offset is small. However, inter-hop and inter-antenna interferences are introduced given a non-zero timing offset, which consequently degrades the demodulation performance of FHCS. As will be shown in Section IV-C, the demodulation performance of FHCS has a significant impact on that of PSK. To this end, accurate timing is still desired for FHCS to attain a high communication performance of FH-MIMO DFRC.

We also notice that, unlike the phase modulation schemes having monotonically increasing data rates with respect to (w.r.t.) M, the data rate of FHCS first increases with M, then saturates as M reaches K/2, and starts decreasing as M keeps increasing, due to the binomial coefficient in its rate expression. To improve the data rate, FHCS can be employed in combination with the phase modulation schemes, as developed in [16], [17]. However, the cost of the rate improvement is that more necessary information will be required for demodulation. This leads to the trade-off between data rate and demodulation performance in FH-MIMO DFRC.

As for the demodulation performance of different singling schemes developed for FH-MIMO DFRC, no analytical result has been published yet. In conventional communications, the asymptotic SER of DPSK and CPM is generally inferior to that of PSK, particularly when the modulation order is high; and the optimal receiver of CPM can be more difficult to achieve compared with the other two phase modulations [24]. These general results, however, may not hold for FH-MIMO DFRC, since the phase demodulation performance is subject to the quality of (T2) acquired in practice. As will be validated in Section IV-C, this coupling will significantly affect the demodulation performance of PSK.

IV. CHANNEL ESTIMATION FOR FH-MIMO DFRC

As illustrated above, most existing signaling schemes for FH-MIMO DFRC require accurate channel information to perform demodulation. Yet, to the best of our knowledge, only a few published work [16], [17] develop channel estimation scheme for FH-MIMO DFRC. In this section, we first review the channel estimation methods designed for DFRC (not based on FH-MIMO radar), and then briefly introduce the solution developed in [16].

A. Overview of Channel Estimations in DFRC

To the best of our knowledge, few published works have explicitly dealt with the channel estimation issue of DFRC. Specifically, sparse recovery-based channel estimation methods are developed in [25], [26] which coordinate radar and communication receiver using probing beams. In a different and yet relevant context (spectrum sharing), interference channel between radar and communication is estimated to achieve co-existence; refer to [4] for a review on those methods. In some recent DFRC works [3], [4] channel estimation methods are developed, which, however, are based on new (future) DFRC waveforms/platforms. These new designs are specifically tailored for dual functions and are non-trivial to be applied in FH-MIMO DFRC. A common feature captured by most of the above methods is the full cooperation between radar and communication. In contrast, the channel estimation for FH-MIMO DFRC is likely to be carried out at a low-profile communication receiver with low computing power and a small number of (or a single) antennas.

B. Recent Advances in Channel Estimation for FH-MIMO DFRC

In [16], a channel estimation scheme is developed for FH-MIMO DFRC to estimate the hopping frequencies and channel parameters, assuming an accurate timing.

1) Estimating Hopping Frequency: A novel FH-MIMO waveform with minor revisions to conventional ones is designed in [16], which enables UE to estimate k_{hm}, i.e., the hopping frequency used by the m-th radar transmitter at any hop h. In particular, it is proposed that: “the radar first randomly selects available sub-bands as hopping frequencies, as it usually does, then re-orders the frequencies in a deterministic order, ascending for instance, before assigning to each transmitter antenna.” The waveform re-ordering is illustrated in Fig. 2. This simple processing enables k_{hm} to be estimated at the UE (instead of being acquired from the radar).

| TABLE I | COMPARING EXISTING SIGNALING SCHEMES FOR FH-MIMO DFRC |
|---------------------------------|-----------------|------------------|------------------|
| **Modulation** | **Data Rate** | **Range Sidelobe** | **Spectral Containment** |
| | | PHR×HM | PHR×HMJ | PHR×HM log₂ M | PHR×HM log₂ C₀K |
| | | good | better | lowest | high with periodic spikes |
| **Necessary Information for Demodulation** | | (T1), (T2) and (T3) | (T1) and (T2) | (T1), (T2) and (T3) | (T1) (not strictly necessary) |
According to the novel waveform illustrated above, $k_{hm} < k_{hm'}$ ($\forall m, m', m < m'$) is ensured. This relation, applied in the set of sub-band indexes obtained in Section III-B1, leads to an estimate of $k_{hm} \forall m$. Importantly, it is also proved in [27] that the above re-ordering does not incur any change to the range ambiguity function of the FH-MIMO radar. Fig. 4 illustrates the range ambiguity function with or without the range ambiguity function of the FH-MIMO radar. Fig. 4 shows that the above re-ordering does not incur any change to the MSE of $\hat{Y}_0$ and hence has no information embedding, i.e., $\Delta_0$ to update $\hat{Y}_0$ can be regarded as a discrete frequency of a single-tone signal. By taking the DFT of $\hat{Y}_0$ (over $m$), the integer part of $u_0$ can be identified as the index of the strongest peak of $|Z_m|$, where $Z_m$ is calculated as

$$Z_m = \sum_{m=0}^{M-1} \hat{Y}_0 e^{-j\frac{2\pi m}{M} m'}, m' = 0, 1, \cdots, M - 1. \quad (7)$$

Let $\hat{m}$ denote the index of the peak. A coarse estimation of $u_0$ is obtained as $2\pi\hat{m}/M$ whose estimation error is up to half of the frequency bin, i.e., $\pi/M$. To improve the estimation, interpolating DFT coefficients has been suggested in the literature [27], [28]. Next, we summarize the steps for $u_0$ estimation based on the latest $q$-shift estimator (QSE), originally designed in [27] and improved in [28]. The technical details are suppressed here (interested readers are referred to [27], [28]).

To refine the estimate of $u_0$, QSE interpolates the DFT coefficients around $\hat{m}$, yielding

$$Z_\pm = \sum_{m=0}^{M-1} \hat{Y}_0 e^{-j\frac{2\pi m (\hat{m} \mp \epsilon)}{M}}; \quad \epsilon \leq \min\{M^{-\frac{1}{2}}, 0.32\} \quad \text{[28, Eq. (23)]},$$

where $\delta$ is an interpolation factor to be updated iteratively and $\epsilon$ is an intermediate variable satisfying $\epsilon \leq \min\{M^{-\frac{1}{2}}, 0.32\}$ [28, Eq. (23)]. Constructing a ratio based on $Z_\pm$, as given by

$$\gamma = \frac{Z_+ - Z_-}{Z_+ + Z_-}$$

and using the ratio $\gamma$ to update $\delta$ lead to

$$\delta = \frac{\cos^2(\pi \epsilon)}{1 - \pi \epsilon \cot(\pi \epsilon)} \times \Re(\gamma) + \delta, \quad (9)$$

where $\delta$ on the right-hand side (RHS) of the equality is the old value, and $\Re(\cdot)$ takes the real part of a complex number. By updating $\delta$ for three times, the algorithm can generally converge [27]. The final estimate of $u_0$ and $\beta_0$ can be given by

$$\hat{u}_0 = \frac{2\pi}{M} (\hat{m} + \delta); \quad \hat{\beta}_0 = \frac{1}{M} \sum_{m=0}^{M-1} \hat{Y}_0 e^{jm\hat{u}_0}. \quad (10)$$

C. Performance Illustration

Numerical results are provided here to demonstrate the high estimation performance of the method elaborated on above. Fig. 5 plots the mean squared error (MSE) of the estimated channel parameters against the SNR, as measured based on (4) and denoted by $\gamma_0$ below. To evaluate the MSE of $\hat{u}_0$, $N = 2 \times 10^4$ independent trials are performed, and the MSE is calculated by $\sum_{n=0}^{N-1} (\hat{u}_0(n) - u_0)^2/N$, where $u_0(n)$ is an estimate of $u_0$ and is obtained in the $n$-th independent trial. The MSE of $\hat{\beta}_0$ is calculated similarly. Based on [27], the Cramér-Rao lower bound (CRLB) of $u_0$ estimation is given by

$$\text{CRLB}(u_0) = \frac{3}{(\pi^2 M L \gamma_0^2)}.$$
A. New Signaling Strategies Enabling Higher Data Rates

As shown in Section IV-C, the overall data rate achieved by FH-MIMO DFRC is 27 Mbps, under the condition that BPSK and FHCS is combined, the radar has 10 antennas and the bandwidth of 100 MHz, and the UE has a single antenna. To provide an effective high-speed G2A communication link using FH-MIMO DFRC, new signaling strategies are required.

We remark that the information conveyed in the hopping frequencies has not been fully explored yet. In light of FHCS as illustrated in Section III-A, the permutations of hopping frequencies can be employed to convey information bits [18]. Using permutations is expected to increase the data rate dramatically, since the number of permutations can be huge. Take the parameters used in Fig. 6 for an example. The number of permutations that permutes \( M(=10) \) out of \( K(=20) \) sub-bands can be larger than 670 billion, which, in theory, is able to convey up to 39 bits per radar hop. As a result, an increase of 21 Mbps in data rate can be achieved through using permutations. This improvement of data rate, however, requires a proper demodulation method to be developed at the UE. One challenge is to deal with the reduced Euclidean distances between the permutations. Another challenge is to develop effective receiving schemes for UE to reliably demodulate the permutation-based information symbol, particularly when the issues discussed sequentially are present.

B. Timing Offset

The prior works on FH-MIMO DFRC [10]–[16] assume perfect timing which is also the case in our elaboration provided in Section IV. However, there can be a non-zero timing offset in practice, particularly when a synchronization link between the radar and the UE is unavailable; see Section III-B for details. In a packet-based communication, the UE can attain a coarse timing by performing conventional methods like energy or correlation-based detection [17]. Provided a non-zero timing offset, a sampled hop spans over two actual hops, as illustrated in Fig. 7. As a consequence, there will be inter-hop interference introduced to the received signals. Moreover, due to the incomplete sampling of a hop, the waveform orthogonality of the FH-MIMO radar, see (2), is destroyed, hence introducing inter-antenna interference [17].

It is noteworthy that the impact of timing offset is frequency-varying, which can result in great difficulty for channel estimation and information demodulation. Based on the baseband radar signal given in (3), we can model the received signal with a non-zero timing offset, as given by \( e^{2\pi i (L_{\eta}/K) k_{hm}} \), where \( L_{\eta} \) is the number of samples in the non-zero timing offset denoted by \( \eta \). Since \( k_{hm} \) varies across hop \( h \) and antenna \( m \), we notice that the phase nuisance incurred by \( L_{\eta} \) also varies over hops and antennas. To this end, the timing offset can lead to the following severe consequences:

1) The antenna-varying phase disturbance will invalidate many angle estimation methods, including the one elaborated on in Section IV-B, that rely on the linear phase relation across the antennas of radar transmitter;
2) The DFRC channel presents a fast fading from hop to hop, which dramatically burdens the channel estimation and communication overhead;

A seemingly straightforward solution to this issue is to perform additional timing offset estimation, which helps recover a complete hop and revive the methods for channel estimation illustrated in Section IV. However, the estimation of the timing offset is challenging. As mentioned earlier, inter-antenna and inter-hop interference exists for any given non-zero timing offset. On the other hand, the antenna- and hop-varying phase resulted from the timing offset is coupled with the phase of channel response in a point-wise manner. This makes the estimation of either part challenging. To address the above challenges, we may resort to the degrees of freedom (DoF) available in the FH-MIMO radar waveform. For example, if \( k_{hm} = 0 \), then the phase caused by the timing offset is removed according to \( e^{j2\pi(l-L)k_{hm}\Delta} \). More works are demanded to address the issue of timing offset effectively.

C. Multi-Path Channel

Thus far, only the LoS-dominated AWGN channel is considered for FH-MIMO DFRC. As discussed in Section II-A, it can be practical to consider the LoS channel for some scenarios, such as the G2A communications between a ground-based radar and an HAP; see Fig. 1. However, as also illustrated in the figure, multi-path fading can happen in FH-MIMO DFRC, when the airborne UE, e.g., UAV, flies at a low altitude. Under the multi-path fading, the signal transmitted by some antennas of the radar transmitter can be severely attenuated.

Fig. 8 illustrates the signal attenuation caused by multi-path fading by plotting \( |Y_n(l)| \), namely, the magnitude of the DFT of the UE-received signal within a radar hop, where \( Y_n(l) \) is given in (5) and an accurate timing is assumed here. Based on the waveform design illustrated in Section IV-B1, the \( m \)-th DFT peak corresponds to the signal from antenna \( m = 0,1,\ldots,M-1 \). We see from the figure that a severe attenuation of the signal transmitted by antenna three is observed on the first UE antenna. Performing the method elaborated in Section IV-B1 to detect the hopping frequency, we see from the left subfigure in Fig. 8 that the hopping frequency of the third antenna is incorrectly detected. This incorrect detection will lead to incorrect demodulation of not only hopping frequency-based modulations, e.g., FHCS, but also phase-based modulations, e.g., PSK. For the latter modulation, the signal used for demodulation is extracted based on the identified indexes of the DFT peaks.

A viable solution to combating the multi-path fading in FH-MIMO DFRC is equipping the UE with multiple antennas and hence exploit the antenna diversity. This is demonstrated in Fig. 8. The left subfigure also plots the magnitude of the DFT of the signal received by another UE antenna. Clearly, the multi-fading attenuation of the signal transmitted by the third antenna is greatly relieved. By combining the signals of the two antennas incoherently, the sum signal can be used to detect the hopping frequency more reliably; see the right subfigure in Fig. 8. It is worth mentioning that improving the detecting performance of the hopping frequencies also substantially benefits to the SER performance of demodulating phase-based modulations, as demonstrated in Fig. 6. With the benefits of using multi-antenna receivers at the UE, recognized, we should also bear in mind that effective signal reception also relies on the accurate estimation of timing offset and channel parameters. These practical issues have not been properly addressed for a single-antenna receiver yet, not to mention multi-antenna configurations.

D. Security Issues

FH-MIMO DFRC can have a serious security issue, since the underlying FH-MIMO radar has an omnidirectional radiation pattern in power [18]. Provided that the signal framework used between the radar and the legitimate UE is known by an eavesdropper, the radar-transmitted signal can be readily demodulated by an eavesdropper using the method provided in Section IV. As an effective supplement to upper-layer security techniques [29], physical layer security has attracted increasing attention recently in 5G and beyond communications, and can be a promising solution to achieving a secure FH-MIMO DFRC.

Specifically, the secrecy enhancement method developed in [30] may be tailored for FH-MIMO DFRC. In short, the method randomly selects a subset of antennas in a transmitter array and half of the antennas in the subset have the signs of their beamforming weights reversed. Through the above processing, AWGNs are injected to the whole spatial region except for the direction of a targeted receiver. The method may be tailored for FH-MIMO DFRC, however, at the cost of reducing the data rate. This is a typical trade off between the secrecy rate and data rate of a communication system. To this end, the security issue can be treated in combination with designing new signaling strategies for high data rates, as discussed in Section V-A.

VI. Conclusions

In this article, a comprehensive review on FH-MIMO DFRC is provided. The potential applications and channel scenarios are presented. The existing signaling strategies for FH-MIMO DFRC are surveyed with their performance compared from various aspects. A channel estimation scheme specifically designed for FH-MIMO DFRC is presented with numerical results to demonstrate estimation and communication performance. Major issues remained in FH-MIMO DFRC are discussed with potential solutions suggested. We envision that, with the highlighted challenges properly addressed in the future, FH-MIMO DFRC is poised to become a viable solution to long-range radar and communications, and forthcoming ISTN and 6G mobile networks.

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Fig. 8. Illustration of the severe signal attenuation caused by multi-path fading, where the Rician channel is considered with the Rician factor of $49$ dB and the LoS path takes the unit power. Two antennas with the spacing of half a wavelength are considered at the UE, each with an independent AWGN of $-10$ dB noise power. The radar parameters are set the same as those in Fig. 3.

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