Terahertz Ultra-Massive MIMO-Based Aeronautical Communications in Space-Air-Ground Integrated Networks

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*Abstract*—The emerging space-air-ground integrated network has attracted intensive research and necessitates reliable and efficient aeronautical communications. This paper investigates terahertz Ultra-Massive (UM)-MIMO-based aeronautical com- munications and proposes an effective channel estimation and tracking scheme, which can solve the performance degrada- tion problem caused by the unique *triple delay-beam-Doppler squint effects* of aeronautical terahertz UM-MIMO channels. Specifically, based on the rough angle estimates acquired from navigation information, an initial aeronautical link is established, where the delay-beam squint at transceiver can be significantly mitigated by employing a Grouping True-Time Delay Unit (GTTDU) module (e.g., the designed *Rotman lens*-based GTTDU module). According to the proposed prior-aided iterative angle estimation algorithm, azimuth/elevation angles can be estimated, and these angles are adopted to achieve precise beam-alignment and refine GTTDU module for further eliminating delay-beam squint. Doppler shifts can be subsequently estimated using the proposed prior-aided iterative Doppler shift estimation algorithm. On this basis, path delays and channel gains can be estimated accurately, where the Doppler squint can be effectively attenuated via compensation process. For data transmission, a data-aided decision-directed based channel tracking algorithm is developed to track the beam-aligned effective channels. When the data- aided channel tracking is invalid, angles will be re-estimated at the pilot-aided channel tracking stage with an equivalent sparse digital array, where angle ambiguity can be resolved based on the previously estimated angles. The simulation results and the derived Crame´r-Rao lower bounds verify the effectiveness of our solution.

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*Index Terms*—Terahertz communications, aeronautical com- munications, ultra-massive MIMO, channel estimation and track- ing, space-air-ground integrated network.

The codes and some other associated materials of this work may be available at https://gaozhen16.github.io.

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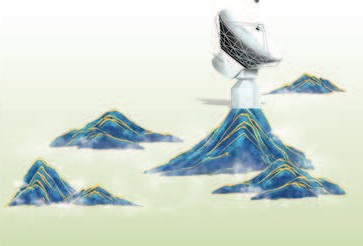
1. INTRODUCTION

Terahertz (THz) communication is expected to play a pivotal role in the future Sixth Generation (6G) wireless systems, which promise to provide ubiquitous connectivity with broader and deeper coverage [1]. THz-band (spectrum ranges from

* 1. to 10 THz) is envisioned to offer significantly larger bandwidths than millimeter-Wave (mmWave) for supporting up to tens of Gigahertz (GHz) ultra-broadband and Terabit per second (Tbps) ultra-high peak data rate [2]–[4]. Meanwhile, THz communications can be conducive to realize the Ultra- Massive Multiple-Input Multiple-Output (UM-MIMO)-based transceivers equipped with tens of thousands of antennas (even the Uniform Planar Array (UPA) with size of 1024 1024 [5]), which can effectively combat the severe path loss of THz signals and further extend the communication range using beamforming techniques [6]–[8]. Therefore, THz UM-MIMO technique has been emerging as a promising candidate for the 6G mobile communication systems [1]. However, due to the severe atmospheric molecular absorption (such as water vapor) and rain attenuation [8], [9], the applications of THz communications are restricted to short-link distance [10]– [12]. Fortunately, those atmospheric molecule absorption and rain attenuation mainly occur in the troposphere, and these negative factors can be largely mitigated due to the negligible absorption in the stratosphere and above [13]–[15].

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On the other hand, the ambitious 6G is poised to seamlessly integrate space-air networks with terrestrial mobile cellular networks. Against this background, the concept of Space-Air- Ground Integrated Network (SAGIN) is conceived and has attracted intensive research [16], [17]. As shown in Fig. 1, a typical SAGIN consists of three layers including spaceborne, airborne, and terrestrial networks [17]. The Geostationary Earth Orbit (GEO), Medium Earth Orbit (MEO), and Low Earth Orbit (LEO) satellites that operate at different altitudes constitute the spaceborne network. In the airborne network, aerial Base Stations (BSs) such as balloons and airships can jointly serve various aircrafts and Unmanned Aerial Vehicles (UAVs). In particular, numerous LEO satellites, aerial BSs, aircrafts, and UAVs can constitute the aeronautical *ad hoc* net- work to achieve the goal of “Internet above the clouds” [18], [19], which necessitates THz UM-MIMO technique to support



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Fig. 1. Typical SAGIN includes spaceborne, airborne, and terrestrial networks, where numerous LEO satellites, aerial BSs, aircrafts, and UAVs together constitute the aeronautical *ad hoc* network [17], [19].

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the reliable and efficient aeronautical communications1.

To guarantee the Quality-of-Service (QoS) for THz UM- MIMO-based aeronautical communications, reliable Channel State Information (CSI) acquisition at the transceiver is in- dispensable [20]. However, due to the high-speed mobility of flying aircrafts/UAVs and the wobbles of aerial BSs, these aerial communication links exhibit the dramatically fast time- varying fading characteristics, which make accurate channel estimation and tracking rather challenging. To acquire the accurate estimate of fast time-varying channel, some channel estimation and tracking schemes [21]–[23] were proposed to reduce the training overhead caused by frequent channel estimation. In [21], a data-aided channel tracking scheme is proposed to estimate and track the partial channel coeffi- cients of angle domain channels using lens antenna array. By exploiting the sparsity of the virtual channel vector in angle domain, the virtual channel parameters based on first order auto regressive model were estimated and tracked using the expectation maximization-based sparse Bayesian learning framework in [22], [23]. Moreover, by acquiring the dominant channel parameters including the Angle of Arrivals/Departures (AoAs/AoDs), Doppler shifts, and channel gains, rather than the complete MIMO channel matrix, some multi-stage channel estimation solutions were proposed in [24], [25] enabling fast channel tracking for narrow-band mmWave MIMO systems. Note that these schemes above just consider the channel estimation and tracking for common mmWave systems. In [26], a priori-aided THz channel tracking scheme with low pilot overhead was proposed to predict and track the physical direction of Line-of-Sight (LoS) component of the time- varying massive MIMO channels in THz beamspace domain. For the dynamic indoor short-range THz communications, the

1In general, civil aircrafts spend most of their flight time at the bottom of the stratosphere, where the relatively stable flight state is convenient for the establishment of THz communication links. Therefore, the aeronautical communications studied in this paper can be mainly aimed at the aircrafts flighted at the stratospheric.

authors in [27] proposed an AoA estimation method based on Markov process and Bayesian inference, where the forward- backward algorithm is implemented to carry out the Bayesian inference.

However, the aforementioned channel estimation solutions are difficult to be applied to the aeronautical THz UM-MIMO systems due to the unprecedentedly ultra-large array aperture, ultra-broad band, and ultra-high velocity. *Compared with the sub-6 GHz or mmWave massive MIMO systems with limited aperture and bandwidth, the aeronautical THz UM-MIMO channels present the unique triple delay-beam-Doppler squint effects*. To be specific, adopting the UPA form, the UM- MIMO arrays mounted on the transceiver of aerial BSs or aircraft can be equipped with up to hundreds of antennas in the single horizontal or vertical dimension, resulting in the ultra-large array aperture even in a small physical size. If the direction of arrival is not perpendicular to the array, we can observe different propagation delays at different antennas for the same received signal filling this array aperture. Moreover, this delay gap can be as large as multiple symbol periods due to the usage of ultra-broadband THz communications. This indicates that the inter-symbol-interference can be non- negligible even for the LoS link, and this phenomenon is termed as the *delay squint effect* of THz UM-MIMO (also named as spatial-frequency wideband effects in [28], [29] and aperture fill time effect in radar systems [30]), which is an inevitable challenge for THz UM-MIMO systems. Meanwhile, this delay squint effect can further introduce the *beam squint effect*, where the beam direction is a function of the operating frequency. This is primarily because radio waves at different frequencies would accumulate different phases given the same transmission distance, while the adjacent antenna spacing is designed according to the central carrier frequency. Hence, beam squint effect would pose undesired beam directions for the signals at marginal carrier frequencies. Furthermore, the high-speed mobility of aeronautical communications causes



Fig. 2. A real-time flight tracking snapshot of civil aircrafts in south China, where the aircrafts generally fly along their fixed routes2.

large Doppler shift and the Doppler shift is also frequency- dependent for aeronautical THz UM-MIMO with very large bandwidth. This phenomenon is called *Doppler squint effect*. Therefore, the aeronautical THz UM-MIMO systems present *triple delay-beam-Doppler squint effects*. However, recent re- searches mainly focus on the impact of beam squint effect on mmWave or THz systems [31]–[35]. To be specific, the impact of beam squint on compressive subspace estimation and the optimality of frequency-flat beamforming was studied in [31]. By projecting all frequencies to the central frequency and constructing the common analog Transmit Precoding (TPC) matrix for all subcarriers, several hybrid TPC schemes were proposed in [32] to design the analog and digital TPC matrices and mitigate the beam squint effect. The channel estima- tion schemes were proposed to exploit the characteristics of mmWave channels affected by beam squint for estimating the wideband mmWave massive MIMO channels [33]–[35], where the beam squint effect is not mitigated. To sum up, the triple squint effects are seldom considered in state-of-the-art channel estimation and hybrid beamforming solutions [21]– [29], [31]–[35] and can dramatically degrade the data trans- mission performance of THz UM-MIMO-based aeronautical communications. Consequently, an efficient signal processing paradigm for channel estimation and data transmission is invoked for enabling aeronautical THz UM-MIMO technique. In this paper, we mainly investigate the THz UM-MIMO- based aeronautical communication links connecting aircraft and aerial BSs in SAGIN3, where the practical triple squint effects of aeronautical THz UM-MIMO channel with LoS link will be considered. Specifically, for the airborne network in

2This real-time snapshot can be found on the website URL link: https://flightadsb.variflight.com/tracker/112.761836,29.084716/6.

3The proposed signal processing solution can also be applied to the space-space/space-air links between the UAVs and multiple aerial BSs, or between aircrafts/UAVs and multiple LEO satellites, etc, and the transmission links between the terrestrial stations built on high-altitude mountains and space-air networks. Furthermore, the research on space-ground or air-ground communications in SAGIN is beyond the scope of this paper, and it may be an important research direction of future work.

Fig. 1, the trajectories of aircrafts are usually regular along their fixed routes, as shown in Fig. 2. Based on this fact, the aerial BSs can be deployed near these trajectories to ensure that multiple aircrafts or UAVs can communicate with multiple aerial BSs for constituting the aeronautical *ad hoc* network. Since there are few other scatterers in the stratosphere except high altitude platforms for THz aeronautical communications, we mainly focus on the THz UM-MIMO channel with only LoS component between the aerial BS and the aircraft in this paper. More specifically, we consider that multiple aerial BSs can jointly serve a high-speed mobile aircraft through respective THz LoS links, and different aerial BSs can be cooperated via THz backbone links connecting different aerial BSs or the air-to-ground backbone links. To combat the mul- tipath effect at the receiver of aircraft caused by multiple THz LoS links, the Orthogonal Frequency-Division Multiplexing (OFDM) technique will be applied to this aeronautical com- munication system4. Among the THz links aforementioned, the THz UM-MIMO-based aeronautical communication links connecting the aircrafts and aerial BSs are the most challeng- ing to be established due to their fast time-varying fading characteristics. On the one hand, by exploiting the prior information (e.g., positioning, flight speed and direction, and posture information) at aerial BSs and aircrafts, some rough channel parameter estimates (e.g., angle and Doppler shift) can be acquired for facilitating the link establishment. On the other hand, these rough channel parameter estimates are not accurate enough for data transmission. Particularly, due to the exceedingly long link distance and extremely narrow beamwidth of aeronautical THz UM-MIMO, a slight deviation of angle parameter resulted from the positioning accuracy error and the posture rotation of antenna arrays mounted on transceiver would lead to the undesired beam pointing. Therefore, how to effectively leverage the prior information above to establish and track the fast time-varying links is vital for THz UM-MIMO-based aeronautical communications.

The proposed channel estimation and tracking solution can be divided into three stages, including the initial channel estimation for link establishment, data-aided channel tracking, and pilot-aided channel tracking. The frame structure is shown in Fig. 3, and the details are presented as follows:

* At the initial channel estimation stage, by utilizing the rough angle estimates acquired according to the positioning

and flight posture information, the rough transmit beamform- ing and receive combining can be achieved to establish the THz UM-MIMO link, where the impact of delay-beam squint effects on both the transmitter and receiver can be significantly mitigated by employing a Grouping True-Time Delay Unit (GTTDU) module with low hardware cost.

* After the link establishment, the fine estimates of az- imuth/elevation angles at both the transmitter and receiver,

4To meet the high quality-of-service requirement for hundreds of people in the aircraft simultaneously, the relatively complicated high-order modu- lation methods, i.e., OFDM and Quadrature Amplitude Modulation (QAM), can be utilized to enhance the data transmission rate and throughput in this paper. Moreover, due to the high Peak-to-Average Power Ratio (PAPR) in OFDM systems, Discrete Fourier Transform-Spread-OFDM (DFT-S-OFDM) technique is also the potential alternative for THz UM-MIMO-based aeronau- tical communication systems.

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| )iQH DQJOH HVWiPDWiRQ DW %6V | )iQH DQJOH HVWiPDWiRQ DW DiUcUDIW | )iQH DRSSOHU VKiIW HVWiPDWiRQ | 3DWK GHOD\ HVWiPDWiRQ |

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Fig. 3. Frame structure of the proposed channel estimation and tracking solution.

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Doppler shifts, and path delays at the receiver are then obtained, where the rough Doppler shift estimates are utilized to compensate the received signals for improved parameter estimation. For the fine azimuth/elevation angle estimation, the UM hybrid array can be equivalently considered as a low-dimensional fully-digital array by employing a reconfig- urable Radio Frequency (RF) selection network with dedicated connection pattern. In this way, the accurate estimates of azimuth/elevation angles at BSs and aircraft can be separately acquired using the proposed prior-aided iterative angle estima- tion algorithm. These fine angle estimates can be used not only to achieve the more precise beam alignment, but also to refine the GTTDU module at the transceiver for further eliminating the delay-beam squint effects. Meanwhile, thanks to the large beam alignment gain and the sufficient receive Signal-to- Noise Ratio (SNR), the Doppler shifts can be accurately estimated based on the proposed prior-aided iterative Doppler shift estimation algorithm, where the Doppler squint effect can be attenuated vastly by compensating the received signals with the rough Doppler shift estimates. On this basis, path delays and channel gains can be estimated subsequently, where Doppler squint effect can be also attenuated vastly via fine compensation process.

* At the data transmission stage, a Data-Aided Decision- Directed (DADD)-based channel tracking algorithm is devel- oped to track the beam-aligned effective channels, where the

correctly decoded data will be regarded as the known signals to estimate channel coefficients.

* The pilot-aided channel tracking is proposed when the data-aided channel tracking is ineffective. At this stage, an equivalent fully-digital sparse array will be formed by recon-

figuring the connection pattern of the RF selection network, where the angle ambiguity issue derived from sparse array can be addressed with the aid of the previously estimated angles at the receiver. Once the precise beam alignment is achieved again, the Doppler shift and path delay estimation can be executed similar to the initial channel estimation stage, and then the transceiver will enter the data transmission stage again.

The main contributions of our proposed scheme are sum- marized as the following aspects:

* + - THz UM-MIMO-based aeronautical communication channels exhibit the huge spatial dimension and very

fast time-variability. To reduce the training overhead, we propose a parametric channel estimation and tracking solution. At the stages of initial channel estimation and pilot-aided channel tracking, by exploiting the proposed prior-aided iterative angle and Doppler shift estimation algorithms, the proposed solution can acquire the fine estimates of channel angles, Doppler shifts, and path delays, whereby some rough channel parameter estimates are leveraged to improve the estimated accuracy and reduce the pilot overhead. At the data transmission stage, to further save the pilot overhead, the proposed DADD- based channel tracking algorithm can reliably track the fast time-varying channel gains of the effective beam- aligned link.

* The proposed scheme can effectively overcome the unique triple delay-beam-Doppler squint effects of aero- nautical THz UM-MIMO communications. Note that this triple squint effects are rarely observed and investigated in the sub-6 GHz or mmWave massive MIMO systems due to the limited aperture and bandwidth. To cope with the delay-beam squint effects, we propose the low-cost GTTDU module at the transceiver, which can compensate the signal transmission delays at different antenna group with the aid of navigation information. In this way, the delay-beam squint effects can be significantly mitigated and the sufficient receive SNR can be guaranteed to establish the THz link. Also, the designed Rotman lens- based GTTDU module in Section VIII provides a feasible implementation architecture of the tunable TTD module based Phase Shift Network (PSN), which would be a potential direction for the future research work. Fur- thermore, by utilizing the proposed prior-aided iterative angle and Doppler shift estimation algorithms to further mitigate the impact of beam and Doppler squint effects, the fine angle and Doppler shift estimates can be acquired for the following data transmission.
* We introduce a reconfigurable RF selection network to obtain the equivalent low-dimensional fully-digital array by designing the dedicated connection pattern. On this basis, the robust array signal processing techniques such as Two-Dimensional Unitary ESPRIT (TDU-ESPRIT) [36], [37] can be utilized to accurately estimate and track the azimuth/elevation angles at the transceiver. Particu-

larly, by reconfiguring the connection pattern of the RF

selection network, the equivalent fully-digital sparse array can be obtained for improved angle estimation accuracy

analog beamforming to serve the assigned aircraft. The specific

configurations of these antenna arrays are as follows. The total number of antennas at BS arrays is *N*BS = *N* h *N* v ,

BS BS

BS

BS

at the pilot-aided channel tracking stage, where angle

where *N* h

and *N* v

are the numbers of antennas in horizontal

ambiguity issue can be addressed well based on the and vertical directions, respectively. Due to the sub-connected

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˜*I*

previously estimated angles.

* + The Crame´r-Rao Lower Bounds (CRLBs) of dominant channel parameters are derived based on the effective received signal models. Particularly, at the pilot-aided channel tracking stage, the CRLBs of angles are derived to theoretically verify the improved estimation accuracy by employing the sparse array. Simulations results have the good tightness with the analytical CRLBs, which testifies the good performance of the proposed scheme.

The remainder of this paper is organized as follows. Sec- tion II introduces the system model, including the signal transmission and channel models with triple squint effects. The initial channel parameter estimation stage, including the estimations of azimuth/elevation angles at BSs and aircraft, Doppler shifts, path delays, and channel gains, is illustrated in Section III. The DADD-based channel tracking and the pilot- aided channel tracking methods are proposed in Sections IV

PSN adopted at aircraft, we define h h ) and v (*M* v ) as the numbers of subarrays (antennas within each subarray) in horizontal and vertical directions, respectively; while *N* h = *I*h h and *N* v = *I*v v are the numbers of antennas in horizontal and vertical directions of array, respectively. Then, the total numbers of antennas in each subarray and the whole antenna array are *M*AC = *M* h v and *N*AC = *N* h v , respectively. Clearly, the aircraft and BS

h v

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AC

AC

*M*

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AC

*M*

AC

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AC

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are equipped with *L* = *I*˜ *I*˜ RF chains and only one RF

AC

AC AC

chain, respectively, and each subarray and the corresponding

RF chain mounted on aircraft are assigned to one BS. According to the frame structure in Fig. 3, the az-

imuth/elevation angles at BSs and aircraft are estimated in the Uplink (UL) and Downlink (DL), respectively, and OFDM with *K* subcarriers is adopted. The UL baseband signal *y*[m] [*k*] received by the *l*th BS at the *k*th subcarrier of the *m*th OFDM symbol can be expressed as

UL,l

and V, respectively. Section VI presents the performance anal-

UL,l

RF

BB

UL

ysis on CRLB and computational complexity. The numerical

*y*[m] [*k*] =√*P q*H

+ *q*

*H*[m] [*k*]*P P* [m][*k*]*s*[m][*k*]

evaluations is given in Section VII. Finally, Section VIII

concludes this paper.

H

RF,l

UL,l

UL,l

l

RF,l

*n*[m] [*k*]*,* (1)

Throughout this paper, boldface lower and upper-case sym- bols denote column vectors and matrices, respectively. ( )∗, ( )T, ( )H, ( )−1, and denote the conjugate, transpose, Hermitian transpose, matrix inversion, and modulus opera- tors, respectively. ǁ*a*ǁ2 and ǁ*A*ǁF are the *ℓ*2-norm of *a*

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where 1 ≤ *l* ≤ *L*, 1 ≤ *k* ≤ *K*, and *P*l is the transmit power.

In (1), *q*RF,l ∈ CNBS is the analog combining vector of the *l*th BS, *P*RF CNAC×L and *P* [m][*k*] CL×L are the analog and digital precoding matrices at aircraft, respectively, while *H*[m] [*k*] ∈ CNBS×NAC is the UL effective baseband channel

BB

∈ ∈

UL,l

and the Frobenius norm of *A*, respectively. The Kronecker and Hadamard product operations are denoted by ⊗ and ◦, respectively. ⟨*a, b*⟩ expresses the inner product of vectors *a*

matrix, *s*[m][*k*] ∈ CL is the transmitted signal vector, and

*n*[m] [*k*] ∈ CNBS is the complex Additive White Gaussian

UL

UL,l

Noise (AWGN) vector with the covariance *σ*2 , i.e., *n*[m] [*k*] ∼

and *b*. 0n and *I*n denote the vector of size *n* with all the 2

CN

0NBS *, σ*n*I*NBS

. Similarly, the DL baseband signal vector

n UL,l

|Q|c is the cardinality of the set Q, and {Q}n denotes the *n*th element of the ordered set . [*a*]Q denotes the sub- vector containing the elements of *a* indexed in the ordered set Q. [*a*]m and [*A*]m,n denotes the *m*th element of *a* and

Q

elements being 0 and the *n* × *n* identity matrix, respectively.

the *m*th-row and the *n*th-column element of *A*, respectively.

*y*[n] [*k*] CL received by aircraft at the *k*th subcarrier of the

*n*th OFDM symbol is given by

DL

∈

L

Σ √

*y*[n] [*k*] = (*W* [n][*k*])H*W* H *P*l*H*[n] [*k*]*f*RF,l*s*[n] [*k*]

DL

BB

RF

DL,l

DL,l

l=1

diag(*a*) is the diagonal matrix with the elements of *a* at its diagonal entries. *∂*(·) and *∂*2(·) are the first- and second-order partial derivative operations, respectively. Finally, E(·) and

where *W*

+ *n*[n] [*k*] *,* (2)

∈ CNAC×L and *W* [n][*k*] ∈ CL×L are the ana-

DL

RF

BB

denote the expectation and real part of the argument,

ℜ{·}

respectively.

log and digital combining matrices at aircraft, respectively,

*f*RF,l ∈ CNBS is the analog precoding vector of the *l*th BS,

while *H*[n] [*k*] ∈ CNAC×NBS is the DL effective baseband

DL,l

1. SYSTEM MODEL

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In this section, we will formulate the signal transmission and channel models with LoS link for THz UM-MIMO-based aeronautical communications, where the full-dimensional UM- MIMO channel model using UPAs involves azimuth and ele- vation angles [37], [38]. Fig. 4(a) depicts the specific scenario that *L* aerial BSs jointly serve an aircraft through respective THz LoS links. The aerial BSs and aircraft adopt the hybrid

channel matrix, and *s*[n] [*k*] and *n*[n] [*k*] CNAC are the transmitted pilot signal (or the modulated/coded data) and the AWGN vector (similar to *n*[m] [*k*]), respectively.

To illustrate the delay squint effect of THz UM-MIMO channels, we take the antenna array at BS as an example as shown in Fig. 4(b). Specifically, the first (1*,* 1)th antenna element can be regarded as the reference point, and define *r* = sin(*θ*BS) cos(*ϕ*BS)*,* sin(*ϕ*BS)*,* cos(*θ*BS) cos(*ϕ*BS) as the

DL,l

DL,l

UL,l

l

l

l

l

l

beamforming structure with a sub-connected PSN [4], [9],

unit direction vector, where *θ*BS

and *ϕ*BS

are the azimuth

where the sub-connected PSNs at BSs can be simplified as and elevation angles associated with the *l*th BS, respectively.

l

l

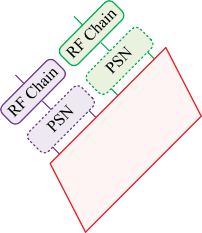
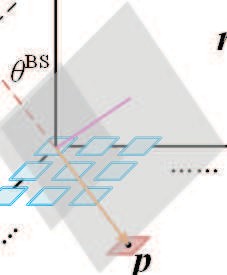
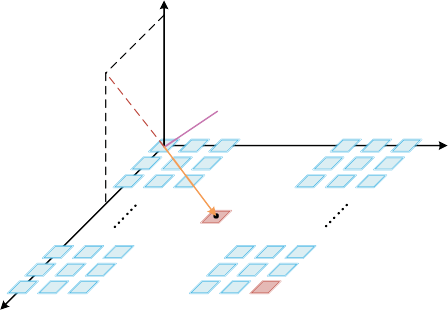
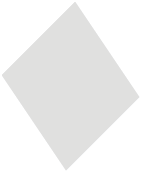
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(*N* K , *N* Y )WK

%6 %6

5) &KDiQ

Fig. 4. The structure diagram of the antenna arrays at transceiver: (a) *L* = 2 BSs that use analog beamforming communicate with aircraft adopting sub-

connected PSN through respective LoS links, and (b) takes the UPA at BS with size of *N* h ×*N* v

as an example to illustrate the delay squint effect of THz

UM-MIMO array.

BS BS

Defining the (*n*h v )th antenna as the *n*BSth antenna with

BS

*, n*

BS

*n*BS =(*n*v −1)*N* h +*n*h , its direction vector relative to the

link and the channel gain6, respectively, *ψ*l = *v*l*/λ*c denotes the Doppler shift with *v*l and *λ*c being the relative radial velocity

BS BS BS

h v

reference antenna is *p* =

(*n*BS −1)*d,* (*n*BS−1)*d,* 0 , where

and carrier wavelength, respectively, *f*c is the corresponding

*d* denotes the adjacent antenna spacing with half-wavelength.

carrier frequency, *τ* [nAC] denotes the transmission delay be-

The wave path-difference between the *n*BSth antenna and the first antenna, denoted by ∆*D*nBS , is equal to the distance between the equiphase surfaces of these two antennas, i.e.,

∆*D*n = *r, p* = (*n*h 1)*d* sin(*θ*BS) cos(*ϕ*BS) + (*n*v

BS

BS

l

l

BS

⟨ ⟩ − −

1)*d* sin(*ϕ*BS). Denoting *τ* [nBS] as the transmission delay from

tween the *n*ACth antenna (*n*AC = (*n*v 1)*N* h h , and it also the (*n*h v )th antenna of UPA at aircraft) and its reference point, and *δ*( ) and *τ*l are the Dirac impulse function and the path delay, respectively. After some algebraic trans-

formations, the DL spatial-frequency channel matrix *H*[n] [*k*]

l

AC

AC

+*n*

AC

*, n*

AC

AC

−

·

l l DL,l

the *n*BSth antenna to the first antenna for the *l*th BS, we in (2) at the *k*th subcarrier of the *n*th OFDM symbol can be

can obtain *τ* [nBS] = ∆*D*n */c* with *c* being the speed of expressed as

l

BS

light. Note that *τ*

[nBS]

is related to the antenna index and

k−1 1

l

the azimuth/elevation angles. When the signal direction is not

perpendicular to the array and *n*BS is large, *τ* [nBS] can be

DL,l

*H*[n] [*k*] =

√*G*l*α*l*e*

j2πψ

*l,k*

(n−1)Tsym *e*

−j2π

K − 2

f*s*τ*l*

5 l × *A*DL,l[*k*]*,* (4)

even larger than the symbol period *T*s , which compels higher

demands on the signal processing at the receiver, especially for the analog or hybrid beamforming architecture. Therefore, the delay squint effect needs to be taken into account for aeronautical THz UM-MIMO systems.

Considering the channel reciprocity in time division duplex systems, we focus on the formulation of DL channel matrix next. According to the channel model in [34], [39], define the DL passband channel matrix in the spatial-delay domain as

where *T*sym and *f*s denote the duration time of an OFDM symbol and system bandwidth, respectively, *ψ*l,k = *ψ*z,l + v*l* ( k−1 − 1 )*f*s is the frequency-dependent Doppler shift at the

*k*th subcarrier with *ψ*z,l being the Doppler shift of the central carrier frequency *f*z (wavelength *λ*z) and v*l* ( k−1 1 )*f*s being the *Doppler squint part* due to the large bandwidth in THz communications, and *A*DL,l[*k*] CNAC×NBS is the DL array response matrix associated with the array response vectors at

∈

c

K

2

c

K

2

−

¯ (t)

*H*

DL,l

(*τ* ) ∈CNAC×NBS at time *t* corresponding to the *l*th BS,

aircraft and the *l*th BS, given by

whose the (*n*AC*, n*BS)th element, i.e., [*H*¯ (t)

DL,l

AC

BS

*A*DL,l[*k*] =

be expressed as

(*τ* )]n

,n , can

AC AC H

BS BS

[*H*¯ (t)

DL,l

(*τ* )]n ,n

AC AC

*a*¯AC(*µ*l *, ν*l *, k*)*a*¯BS(*µ*l *, ν*l *, k*) *,* (5)

H BS BS

*a*AC(*µ*l *, ν*l )*a*BS(*µ*l *, ν*l )

*A*˛D¸L*,l* x

= √*G*l*α*l*e*j2πψ*l*t*δ τ* − *τ*l − (*τ* [nAC] + *τ* [nBS]) *,* (3)

`

AC

BS

◦

l

l

*A*DL*,l*[k] (*Beam squint component*)

` ¯ ˛¸ x

` *Delay*˛¸*squint* x

where *µ*AC = *π* sin(*θ*AC) cos(*ϕ*AC) (*µ*BS =

where 1 ≤ *n*AC ≤ *N*AC, 1 ≤ *n*BS ≤ *N*BS, *G*l and *α*l ∼

*π* sin(*θ*BS) cos(*ϕ*BS)) and *ν*AC

= *π* sin(*ϕ*AC)

CN(0*, σ*2 ) are the large-scale fading gain of communication

l

l

l

l

l l l l

(*ν*BS

l

α

5We consider an extreme scenario that the impinging signal comes from the diagonal direction of UPA of size (*n*BS + 1) (*n*BS + 1), and those (*n*BS +1) diagonal antennas consist of the Uniform Linear Array (ULA) of size (*n*BS +1) with 2*d* antenna spacing. When angle *θ*BS = 60◦, carrier frequency *fc* = 0*.*1 THz, and bandwidth *fs* = 1 GHz for the typical THz UM-MIMO aeronautical communication scenario, *n*BS = 200 antennas will

*l*

√

×

√

[*n* ] 2*n*BS sin(*θ*BS)

l = *π* sin(*ϕ*BS)) are the horizontally and vertically virtual

angles at aircraft (the *l*th BS), respectively, *A*DL,l is the conventional DL array response matrix without beam squint

6Due to the negligible frequency-dependent attenuation of THz commu- nication links (e.g., atmospheric molecular absorption) in the stratosphere and above [13], [14], the channel gain *αl* can be modeled as a frequency flat coefficient, which is different from the frequency-dependent channel

make its filling time satisfy *τl*

5) &KDiQ

BS =

*l*

2*fc*

≈ 1*.*225 *Ts*.

coefficient in [39].

effect at aircraft and BS, and *A*¯

DL,l

[*k*] is the corresponding

array response squint matrix considering beam squint effect.

**. . .**

In (5), *a*AC(*µ*AC*,ν*AC) = *a*v(*ν*AC*,N* v

l

l

l

AC

l

AC

) ⊗ *a*h(*µ*AC*,N* h )

and *a*BS(*µ*BS*,ν*BS) = *a*v(*ν*BS*,N* v

) ⊗ *a*h(*µ*BS*,N* h )

l l

are the

array

l BS l BS

vectors

general

response

at air-

craft and the *l*th BS [37], respectively, and

*a*¯AC(*µ*AC*, ν*AC*, k*) = *a*¯v(*ν*AC*, N* v *, k*) ⊗ *a*¯h(*µ*AC*, N* h

l

l

l

AC

l

AC

*, k*)

and *a*¯BS(*µ*BS*, ν*BS*, k*) = *a*¯v(*ν*BS*, N* v *, k*) ⊗ *a*¯h(*µ*BS*, N* h *, k*)

**.**

l

l

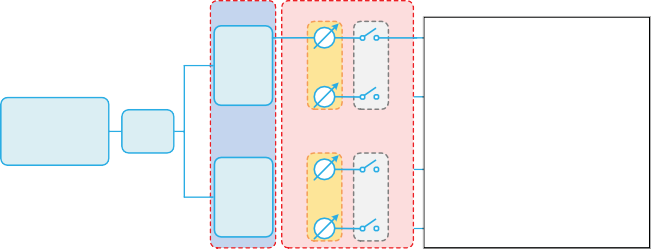
l

BS

l

BS

are the frequency-dependent array response squint vectors,



361 $61

77D8 **.**

**.**

%DVHEDQG G O

SURcHVViQJ

iJiWDO ViJQD

5)

cKDiQ

**.** \*77D8 5HcRQIiJXUDEOH 5)

**.**

PRGXOH VHOHcWiRQ QHWZRUN

77D8 **.**

**.**

**. . .**

**.**

respectively. Moreover, the vectors at aircraft, i.e., the horizontal/vertical steering vectors *a*h(*µ*AC*, N* h ) and

Fig. 5. The transceiver structure corresponding to one RF, where this RF chain connects with the antenna array via the GTTDU module and the reconfigurable RF selection network consisting of a sub-connected PSN and an ASN.

l AC

*a*v(*ν*AC*, N* v

), and the horizontal/vertical steering squint

l AC

vectors *a*¯h(*µ*AC*, N* h

*, k*) and *a*¯v(*ν*AC*, N* v

*, k*) can be further

l AC

written as

AC

l AC

transmission. Therefore, the accurate acquisition of dominant

channel parameters is still indispensable.

*l*

AC

*l*

*a* (*µ*AC*, N* h

h

l

) = h1 *e*jµAC

h

j(N

· · · *e*

−1)µAC iT *,* (6)

To overcome the delay-beam squint effects of THz UM-

MIMO array, the fully-digital array architecture with each

*a* (*ν*AC*, N* v

v

l

) = h1 *e*jνAC

v

· · · *e*

j(N

2

f*z*

AC

*l*

−1)νAC iT *,* (7)

antenna equipping a dedicated RF chain is preferred, but the

involved prohibitive hardware cost and power consumption

*a*¯h(*µ*AC*, N* h *, k*)

AC

*l*

AC

*l*

K

l AC

h j k−1 − 1 f*s* µAC

*l*

=

1 *e*

K

2

f*z*

· · · *e*

j k−1 − 1 f*s* (Nh

−1)µAC iT

make it impracticable. Moreover, the aforementioned hybrid beamforming and channel estimation schemes [31]–[35] uti-

lize some signal processing methods to attenuate the impact of

delay-beam squint effects on the results, rather than eliminate

*a*¯v(*ν*AC*, N* v *, k*)

l AC

h j k−1 − 1 f*s* νAC

*l*

j k−1 − 1 f*s* (Nv

K

−1)νAC iT

*l*

these effects during signal transmission. Therefore, those pro- cessing methods are only suitable for the terrestrial mmWave

or THz cellular networks with abundant scatterers, where

*,*

(8)

= 1 *e* K

2 f*z*

· · · *e*

2 f*z* AC

*.*

(9)

the receiver in short-distance transmission (at most hundreds

of meters) can receive the signals affected by delay-beam

Note that the vectors at BSs, i.e., *a* (*µ*BS*, N* h ),

squint effects. However, for THz UM-MIMO-based aeronau-

*a*v(*ν*BS*, N* v ),

l BS

*a*¯h(*µ*BS*, N* h *, k*), and

h l BS

*a*¯v(*ν*BS*, N* v *, k*),

l BS

tical communication systems that rely on the long-distance

transmission of LoS link (up to hundreds of kilometers)

have the similar definitions and expressions to (6)-(9), and

l BS

their details are omitted for simplicity. The detailed derivation of DL channel matrix *H*[n] [*k*] can be found in Appendix A. Similar to (4), the UL spatial-frequency baseband channel matrix *H*[m] [*k*] in (1) at the *k*th subcarrier of the *m*th OFDM

DL,l

UL,l

symbol corresponding to the *l*th BS can be formulated as

*H*[m] [*k*] = √*G*l*α*l*e*j2πψ*l,k* (m−1)Tsym *A*UL,l[*k*]*,* (10) where the UL array response matrix *A*UL,l[*k*] ∈CNBS×NAC is

UL,l

*A*UL,l[*k*] = *a*BS(*µ*BS*, ν*BS)*a*H (*µ*AC*, ν*AC)

l

l

AC

l

l

without supernumerary scatterers, the receiver will most likely fail to receive the signals at marginal carrier frequencies due to the very narrow pencil beam and (even slight) delay-beam squint effects. Except for the indispensable signal processing, the transceivers of aeronautical communication systems should be elaborately designed to eliminate the delay-beam squint effects and ensure that all carrier frequencies within effective bandwidth can establish a reliable THz communication link. A common treatment of delay-beam squint effects is to design the transceiver based on the TTDU module [40], [41]. The

optimal TTDU module is made up of numerous true-time

` *A*˛U¸L*,l*

BS

l

l

x delay units, and each unit is assigned to its dedicated antenna

AC

l

l

* *a*¯ (*µ*BS*, ν*BS*, k*)*a*¯H

(*µ*AC*, ν*AC*, k*) *.* (11)

[42], where the detailed designs of these tunable TTDUs can

UL*,l*[k] (*Beam s*˛*q*¸*uint component*) x

` *A*

be found in [43], [44]. Nevertheless, the excessively high

¯

1. INITIAL CHANNEL ESTIMATION

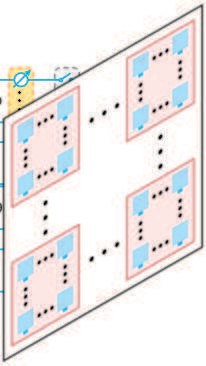
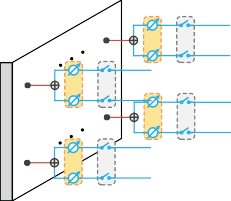
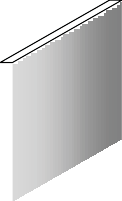
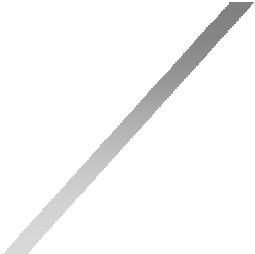
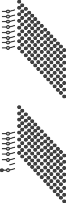
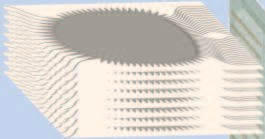
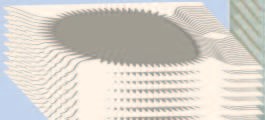
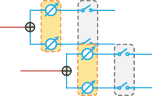
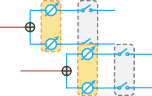
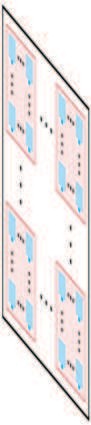
As shown in Fig. 3, at the initial channel estimation stage, the fine azimuth/elevation angles at BSs and aircraft, Doppler shifts, and path delays are estimated successively. At this stage, according to the positioning and flight posture infor- mation acquired in aeronautical systems, some rough channel parameter estimates (e.g., angle and Doppler shift) can be utilized to establish the initial THz UM-MIMO link. Due to the positioning accuracy error and the posture rotations of antenna arrays mounted on aerial BSs and aircraft, these rough channel parameter estimates are not accurate enough for data

hardware complexity and cost of this optimal module prompt

us to design a sub-optimal implementation of TTDU module, i.e., GTTDU module based transceiver structure7 as shown in Fig. 5. From Fig. 5, we observe that except for the antenna array, this transceiver structure contains a GTTDU module and a reconfigurable RF selection network involving a sub- connected PSN and an Antenna Switching Network (ASN) [45], where this ASN can control the active or inactive state of the antenna elements to form different connection patterns

7Since the TTDU/GTTDU module is difficult to tackle multiple path signals in the analog domain simultaneously, the proposed transceiver structure and the subsequent solution for THz aeronautical communications cannot be directly applied in terrestrial vehicular communication scenarios, where the non-LoS components caused by various scatterers are ubiquitous.

[n]



361 $61

1VW-OD\HU 5RWPDQ OHQV 2QG-OD\HU 5RWPDQ OHQV DHOD\ OiQHV

**. .**

**. .**

**. .**

GiJiWDO ViJQDO c5)

%DVHEDQG

**.**

**. .**

SURcHVViQJ

KDiQ

**.**

**.**

**.**

**.**

**.**

**. .**

**.**

**. .**

%HDP SRUWV

**5RWPDQ OHQV EDVHG \*77D8 PRGXOH**

(D)

\*URXSiQJ DQWHQQD SRUWV

**. .**

5HcRQIiJXUDEOH 5) VHOHcWiRQ QHWZRUN

**. .**

**.**

**. .**

{*ν*˜BS}L ({*ν*˜AC}L ), respectively. According to *H* [*k*]

l

l=1

l

l=1

DL,l

in (4), we present the expression of the DL channel matrix after ideal TTDU module processing in the following lemma,

˜denoted by *H* [*k*], which is proved in Appendix B.

[n]

DL,l

*Lemma 1:* According to the rough angle estimates above, the antenna transmission delays of THz UM-MIMO arrays at BSs and aircraft can be compensated using the ideal TTDU module, and the compensated DL spatial-frequency channel

[n]

matrix *H*˜DL,l[*k*] can then be formulated as

[n] √

**.. ..**

**. .**

**.. ..**

**. .**

*H*˜DL,l

[*k*] =

*G*l*α*l*e*j2πψ*l,k*(n−1)Tsym *e*

K

2

−j2π k−1 − 1 f*s*τ*l*

in which

× *A*˜DL,l[*k*]*,* (12)

l

l

BS

l

l

(E)

Fig. 6. (a) A feasible transceiver structure corresponding to one RF, where the Rotman lens-based GTTDU module can be utilized to implement the practical tunable TTDU module [46]; and (b) the other side elevation drawing of a part of RF front-end that includes the grouping antenna ports, reconfigurable RF selection network, and THz UM-MIMO array. The beam ports of the first-layer and second-layer Rotman lenses steer the horizontal and vertical

directions, respectively. The total number of grouping antenna ports is

¯

˛¸ x˜*A* [k]

*A*˜DL,l[*k*] = *A*DL,l[*k*]◦ *a*¯AC(*µ*˜AC*, ν*˜AC*, k*)*a*¯H (*µ*˜BS*, ν*˜BS*, k*) ∗*.*

`

DL*,l*

(13)

˜

By comparing *A*¯ [*k*] in (13) and *A*¯ [*k*] in (5), we can find that if we can acquire the perfect angle information, the beam squint effect part can be perfectly eliminated, i.e.,

*A*¯

¯

DL,l DL,l

˜DL,l[*k*] = *A* [*k*] and then *A*˜DL,l[*k*] = *A*DL,l when

DL,l

˜

*µ*AC

AC

AC

AC

BS

BS

BS

BS

consistent with that of antenna groups in the previous GTTDU module. This

elaborated cascading two-layer Rotman lenses are equivalent to the wideband

phase shifters of the tunable TTD module, which can be utilized to eliminate

˜l = *µ*l , *ν*˜l = *ν*l , *µ*˜l = *µ*l , and *ν*˜l = *ν*l .

Moreover, according to (10) and (11), the compensated UL

[m]

the beam squint effect.

of the RF selection network at the angle estimation stage. In this GTTDU module, a TTDU can be shared by a group of antennas and this imperfect hardware limitation can be handled by the subsequent signal processing algorithms well. Observe that although the delay squint effect for the whole UM array can be non-negligible, this effect for antennas within a group is mild. Hence, the GTTDU module can mitigate the delay squint effect among the antennas in different groups, and the residual phase deviations of these antennas within each group can be further eliminated using their respective phase shifters. Fur- thermore, to illustrate the feasibility of the transceiver designed in Fig. 5, we propose a potential implementation of transceiver structure involving the Rotman lens-based GTTDU module in Fig. 6, where the cascading two-layer Rotman lenses can be utilized to implement the full-dimensional beamforming [46]. The Rotman lens based GTTDU module is a practical photonic implementation [47], and this design employs the

spatial-frequency channel matrix *H*UL,l[*k*] has the similar

expressions, which are omitted for simplicity.

The ideal TTDU module provides a performance upper- bounds for the parameter estimation or data transmission, and we can design the sub-optimal GTTDU module adopted by our solution and the corresponding signal processing algorithms to approach these upper-bounds. The practical DL/UL spatial- frequency channel matrices compensated by the GTTDU module can be derived from (12) and (13). Specifically, all antenna groups for GTTDU module have the same size, i.e., *M* h *M* v at BSs and *M* h *M* v at aircraft, and the central antenna in each group can be regarded as the benchmark of antenna transmission delay for designing the corresponding TTDU. Moreover, to minimize the beam squint effect caused by antenna grouping as much as possible, the phase deviations of the rest antennas in one group can be compensated using the low-cost PSN, where the phase values at central carrier are treated as the benchmark for calculating these deviations. For convenience, the effective UL and DL channel matrices compensated by the GTTDU module can be also denoted as

˜ ˜

BS

BS

AC

AC

˜ ×˜ ˜ ×˜

[m] [n]

optical properties of electromagnetic waves to achieve the

tunable TTD module [48], [49], which provides a prospective direction for our future research work.

When the acquired angle information is accurate enough, the impact of delay squint effect would be significantly mitigated using this GTTDU module. To be specific, based on the prior information acquired from navigation information, the rough estimates of azimuth and elevation angles at BSs (air-

*H*UL,l[*k*] and *H*DL,l[*k*], respectively.

At the initial channel estimation stage, we adopt the Or- thogonal Frequency Division Multiple Access (OFDMA) to distinguish the pilot signals transmitted from different BSs and improve the accuracy of the estimated channel parameters. Hence, *K* subcarriers can be equally assigned to *L* BSs, where the alternating subcarrier index allocation with equal intervals is adopted and the ordered subcarrier index set assigned to the

craft) can be defined as {*θ*˜BS}L ({*θ*˜AC}L ) and {*ϕ*˜BS}L

l

l=1

l

l=1

l

l=1

{ ˜ }

l

l=1

l

l=1

l

l=1

*l*th BS is Kl with *K*l = |Kl|c. Moreover, the azimuth/elevation

( *ϕ*AC L ), respectively, and the corresponding horizontally

BS L AC L

and vertically virtual angles are {*µ*˜ } ({*µ*˜ } ) and

angles at BSs can be estimated in UL, while the rest of channel

parameters are acquired in DL.

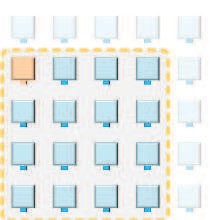
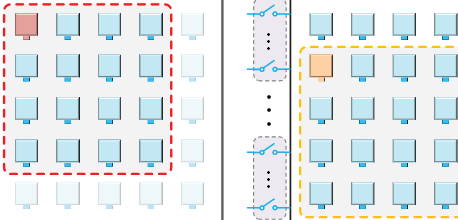
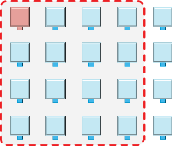
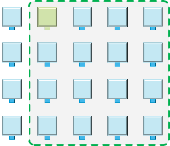
$61 6XEDUUD\ 1

**...**

**..**

**.**

**...**



(TXiYDOHQW

IXOO\-GiJiWDO *d*

DUUD\

**.**

**.**

**.**

**.**

**.**

**.**

**.**

**.**

**.**



$61

6XEDUUD\ 2



$61

6XEDUUD\ 3



$61

6XEDUUD\ 4



H

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| 1VW 2)D0 V\PERO | 2QG 2)D0 V\PERO | 3UG 2)D0 V\PERO | 4WK 2)D0 V\PERO | 7iP |



Fig. 7. The schematic diagram of subarray selection at the initial angle estimation stage, where the different antenna connection patterns can be formed by controlling the ASN of the reconfigurable RF selection network. Taking the UPA of size 5 5 as an example, this UPA can be partitioned into 4 subarrays of size 4 4, and the interval between each subarray is the width of one antenna. The same RF chain sequentially selects the corresponding subarrays in 4 successive OFDM symbols to receive signals, and these received signals will be equivalent to the signals received by a low-dimensional fully-digital array of size 2×2 with the critical antenna spacing *d*.

×

×

*A. Fine Angle Estimation Based on Reconfigurable RF Selec-*

and *i*v being the (*i*h

v )th subarray for 1 ≤ *i*h h

*tion Network*

*, i*

1. *Fine Angle Estimation at BSs:* Due to the insufficient

and 1 *i*v v , respectively, the antenna index of the selected *m*th subarray that corresponds to the *m*th OFDM

symbol can be denoted by I with *M*¯BS = |I |c, so that

BS

BS

BS

≤ *I*

BS

≤ *I*

BS

BS

BS

≤

[m] [m]

valid observation caused by the limited number of RF chains

[m]

BS BS

[m]

at the BSs, it is necessary to accumulate multiple OFDM

I

M¯BS

*q*RF,l can be also initialized as *q*RF,l = 0NBS , and then let

[*q*[m] ]

I

UL,l

symbols in the time domain to estimate the angles. To mitigate

RF,l [*m*] BS

= √ 1 [*a*BS(*µ*˜BS*, ν*˜BS)]

for 1 ≤ *m* ≤ *I*BS.

the inter-carrier interference within one OFDM symbol caused by the large Doppler shifts, the acquired rough Doppler shift estimates are first utilized to compensate the transmitted signals, so that the compensated channels of multiple OFDM symbols can be slow time-varying. By transforming the dif-

l

l

[1]

BS

ferent RF connection pattern of antenna array, we observe

According to the UL transmission model in (1), the received signal *y*[m] [*k*l] at the *k*lth subcarrier of the *m*th OFDM symbol transmitted by the *l*th BS can be expressed as

*y*[m] [*k* ] =√*P* (*q*[m] )H*H*˜′[m] [*k* ]*p s*[m] [*k* ] + *n*[m] [*k* ]*,*

UL,l

l

l

RF,l

UL,l

l

RF,l

UL,l

l

UL,l

l

(14)

a fact that the received signals adopting different selected

where *k*l ∈ Kl, 1 ≤ *m* ≤ *I*BS,

′[m]

*H*UL,l[*k*l] is the channel

˜

subarrays only differ by one envisaged phase value if the

transceiver has the same configuration, and those regular phase

matrix compensated by GTTDU module and rough Doppler

shift estimates, and *s*[m] [*k*l] and *n*[m] [*k*l] are the transmitted

UL,l UL,l

differences can construct the array response vector of low-

pilot signal and noise, respectively. By collecting the received

dimensional fully-digital array. Taking the UPA with size of

5 × 5 in Fig. 7 as an example, we can select 4 subarrays

signals at *K*l subcarriers as *y*[m] ∈ CK*l* and substituting the

of size 4 × 4 in 4 successive OFDM symbols to form the

UL,l

UL channel matrix in (10) into *y*[m] , we have

array response vector of equivalent fully-digital array with

UL,l

*y*UL,l[{Kl}1] · · · *y*UL,l[{Kl}K*l* ]

*y*UL,l =

to estimate the angles at BSs, where each OFDM symbol

RF,l

UL,l

UL,l

size of 2 × 2 by controlling the reconfigurable RF selection

[m]

h [m] [m] iT

[m]

network. Specifically, we intend to use *I*BS OFDM symbols

=√*P G α* (*q*[m] )H*A p*

l

l

l

UL,l

UL,l

*s*[m]

* *y*˜

adopts a dedicated RF connection pattern (i.e., the selected

subarray). By employing the rough angle estimates at aircraft

and BSs, the analog precoding and combining vectors, i.e.,

[m] for 1 ≤ ≤ *L*, 1 ≤ *m* ≤ *I*BS, can be first

*p*RF,l and *q*

*l*

+ *n*[m] *,* (15)

[m] h [m] [m] iT CK , *y*[m]

RF,l

where *s*UL,l

=

*s*

UL,l

[{Kl}1]· · ·*s*

UL,l

[{Kl}K*l* ]

∈

*l*

˜UL,l

is the error vector including the residual beam squint caused by

UL,l

designed. In terms of *p*RF,l, initialize *p*RF,l as *p*RF,l = 0NAC , and then let [*p*RF,l]I = √ 1 [*a*AC(*µ*˜AC*, ν*˜AC)]I . Here

RF,l

AC*,l*

MAC

l

l

AC*,l*

noise vector. Moreover, the same transmitted pilot signals are

subarray assigned to the *l*th BS, since each subarray at aircraft

IAC,l with *M*AC = |IAC,l|c denotes the antenna index of

inaccurate prior information, and *n*[m]

adopted for *I*

OFDM symbol, i.e., *s*

is the corresponding

[*k* ] = *s*[m] [*k* ], and

only communicates with its corresponding BS as shown in Fig. 4(a). To design {*q* } , the UM-MIMO array at BS

accordingly, *s*UL,l = *s*[m]

position of {*y*[m] }IBS

UL,l

BS

UL,l

l

l

UL,l

m=1

for 1 ≤ *m* ≤ *I*BS. Taking the trans-

received from *I*BS OFDM symbols,

UL,l

UL,l

UL,l

UL,l

BS

can be partitioned into *I*BS = *I*h *I*v

RF,l

m=1

BS BS

yield the array response vector of equivalent low-dimensional

*l*

[m] IBS

smaller subarrays to

we can stack them as *Y*

i.e.,

= h*y*[1]

· · · *y*[IBS]iT ∈ CI

×K ,

fully-digital array with size of *I*h v , where the sizes of

BS

×*I*

UL,l

l

l

l

RF,l

UL,l

RF,l

UL,l

UL,l

BS

these smaller subarrays are *M*¯ h ×*M*¯ v (*M*¯ h

BS

BS

BS

BS

BS

BS = *N*BS − *I*BS + 1) and their number of antennas is

and *M*¯ v

v

v

BS

BS

BS

= *N* h −*I*h +1

BS

BS

*Y* =√*P G α*  *Q*H *A p s*T ◦ *Y*˜

*M*¯BS = *M*¯ h *M*¯ v . Defining *m* = (*i*v − 1)*I*h

BS

+*i*h with *i*h

+ *N*UL,l*,* (16)

where *Q*RF,l = h*q*[1]

RF,l

h˜*y*

UL,l

· · ·*q*[IBS]i ∈ CNBS×IBS and

*Y*˜UL,l =

Algorithm 1: Proposed Prior-Aided Iterative Angle

[1]

RF,l

UL,l

· · ·*y*˜[IBS ]i are the analog combining and residual beam

Estimation Algorithm

Input: Rough virtual angle information

BS BS AC AC

*l*

*l*

*l*

*l*

squint matrices, respectively, and *N*UL,l is the noise ma-

trix. By utilizing this analog combining matrix *Q*RF,l, the

array response vector of equivalent low-dimensional fully-

digital array can be formed to estimate the angles at BSs

using array signal processing techniques. To be specific,

*µ , ν , µ , ν* , transmitted pilot signal *s*UL*,l*,

maximum iterations *i*max, and dimensional parameters {*N*AC*, M*AC*, M*¯ BS*, I*BS*, I*h *, I*v *, Kl*}

BS

BS

BS

{˜ ˜ ˜ ˜ }

Output: Estimated azimuth/elevation angles {*θ*^BS *, ϕ*^BS } and

*l*

*l*

*l l*

virtual angles {*µ*^BS *, ν*^BS}

[1]

RF,l l l

compared with (*q*[1] )H*a*BS(*µ*BS*, ν*BS) for *m* = 1 in (15),

1 % Preliminary (subarray selection and signal transmission)

2 Determine antenna indices IAC*,l* and I ;

(*q*[m] )H*a*BS(*µ*BS*, ν*BS) is multiplied by an extra phase shift BS

RF,l l l

*l*

for *m* = (*i*BS −

BS

BS

3 Initialize *p*RF*,l* = 0*N*AC and then let

j((ih

BS v

˜ ˜

*e*

BS

*l*

BS

[*p*RF*,l*]IAC*,l* =

*M*

[*a*AC(*µl , νl* )]IAC*,l* ;

−1)µ +(i

−1)νBS)

v 1)*I*h

+ *i*h

and

√ 1

AC

AC AC

2 *m I*BS. Obviously, these regular phase shifts can

≤ ≤

constitute the effective array response vector of equivalent

l

l

4 for *m* = 1*,*· · ·*, I*BS do

5 Determine antenna index IBS ;

[*m*]

RF*,l*

BS

fully-digital array with size of *I*h ×*I*v

BS

BS

, i.e., *a*¯BS(*µ*BS*, ν*BS) = 6

Initialize *q*[*m*]

= 0*N*

and then let

*a*v(*ν*BS*, I*v

l

BS

l

BS

RF*,l* I[*m*]

[1]

I

signal matrix *Y*UL,l in (16) can be then rewritten as

BS

) ⊗ *a*h(*µ*BS*, I*h

) ∈ CIBS . Thus, the UL received

[*q*[*m*] ]

= √ 1 [*a*

BS

(*µ*˜BS *, ν*˜BS)] ;

*M*¯

BS

*l*

*l*

BS

¯ BS

*Y*UL,l = *γ*UL,l

*a*¯BS(*µ*l *, ν*l )*s*UL,l

UL,l

l

l

l

BS T ˜

RF,l

RF,l

UL,l

7 Transmit pilot signal *s*UL*,l* to obtain received signal

vector *y*[*m*] in (15);

* *Y*UL,l + *N*UL,l*,* (17)

UL*,l*

8 end

UL*,l*

*m*=1

UL*,l*

UL*,l*

UL*,l*

where *γ*

= √*P G α* (*q*[1]

)H*A p*

is the beam-

9 Stack as {*y*[*m*] }*I*BS

as *Y*

=h*y*[1]

· · · *y*[*I*BS]iT in (16)

aligned effective channel gain.

For the received signal model in (17), we propose a prior- aided iterative angle estimation algorithm as follows. At the

first iteration, i.e., *i*BS = 1, the azimuth and elevation angles at

and (17);

10 % Prior-aided iterative angle estimation

max BS

11 for *i* = 1*,*· · ·*, i*  doBS

12 if *i*BS = 1 then

(iBS)

^ ^

(iBS) 13

Apply TDU-ESPRIT algorithm to *Y*UL*,l*;

the *l*th BS can be first estimated as *θ*l and *ϕ*l , and 14

the corresponding horizontally and vertically virtual angles

are *µ*(iBS)

(iBS)

Obtain angle estimates of first iteration as

(*i*BS) (*i*BS) (*i*BS) (*i*BS)

{*θ*^ *, ϕ*^ } and {*µ*^ *, ν*^ };

l and *ν*l for 1 *l L* by applying the TDU- 15 ESPRIT algorithm [36], [37] to the received signal matrix 16 *Y*UL,l. Furthermore, to minimize the impact of *Y*UL,l on (17),

˜

^ ^ ≤ ≤

more accurate angle estimates can be acquired by utilizing 17

the estimated angles above to iteratively compensate *Y*UL,l at

the subsequent iterations (i.e., *i*BS ≥ 2). Specifically, for the 18

*i*BSth iteration, according to the rough virtual angle estimates

else

*l*

*l*

*l*

*l*

Design compensation matrix *Y*˜ (*i*BS−1), whose *kl*th column *y*˜(*i*BS−1)[*kl*] is shown in (18);

Obtain compensated matrix

UL*,l*

UL*,l*

*Y* (*i*BS) = *Y* (*i*BS−1) ∗ ◦ *Y*UL*,l* in (19);

UL*,l*

UL*,l*

˜

Apply TDU-ESPRIT algorithm to *Y* (*i*BS);

UL*,l*

*µ*BS

˜l

and

˜l , and

*µ*(iBS−1)

and

*ν*(iBS−1)

19

estimated at the

Obtain angle estimates of *i*BSth iteration as

(*i*BS) (*i*BS) (*i*BS) (*i*BS)

{*θ*^ *, ϕ*^ } and {*µ*^ *, ν*^ };

(*i*BS − 1)th hiteration, we define the compensatioin matrix as 20

*ν*BS

^l

^l

˜*y*

UL,l

22 Return: *θ*^*l* = *θ*^ BS

UL,l

UL,l

(i −1)BS

(i

−1)

C

*l l l l*

end

BS

(*i*max)

BS

(*i*max)

BS

(*i*max)

, and

*Y*˜ (iBS−1) =

(iBS−1)

[{Kl}1]· · ·*y*˜

[{Kl}K*l* ] , whose the

21 end

*k*lth column *y*˜UL,l [*k*l] ∈

BS

˜*y* [*k* ]lUL,l

IBS

is given by

*l l l*

BS (*i*max)

, *ϕ*^*l* = *ϕ*^

BS

, *µ*^*l* = *µ*^

BS

*ν*^ = *ν* BS

(iBS−1)

= *a*¯v(*ν*˜BS*, I*v

*, k*l) ⊗ *a*¯h(*µ*^l

BS

h BS

*, k*l)

*, k*l) ⊗ *a*¯h(*µ*˜BS*, I*h

*, k*l) ∗

*l* ^*l*

BS

^

*.* (18)

(imax)

^ ≤ ≤

l BS

^v l

*, I*

◦

*a*¯ (*ν*

BS

v BS

(i −1)

l BS

(i

*, I*

−1)

and *ν*l

= *ν*l BS

for 1 *l*

*L*. The proposed prior-aided

After the compensation matrix *Y* processing, the pro-

(i −1)BS

˜UL,l

iterative angle estimation algorithm above is summarized in

Algorithm 1, where the beam squint effect can be addressed

cessed matrix *Y* (iBS) = *Y*˜ (iBS−1) ∗ ◦ *Y*UL,l can be written

UL,l

UL,l

as

well.

*Remark 1:* Based on the analysis above, by controlling the

connection patterns, the reconfigurable RF selection network

can select the desired subarrays to obtain an equivalent low-

*Y* (iBS) = *γ*UL,l *a*¯BS(*µ*BS*, ν*BS)*s*T

UL,l

l

l

UL,l

UL,l

dimensional fully-digital array, so that the robust array signal

* *Y*˜

UL,l

UL,l

subarray, i.e., *M*¯ h

where *N* (iBS)

* *Y*˜ (iBS−1) ∗ + *N* (iBS)*,* (19)

processing techniques can be utilized to obtain the accurate

angle estimates. On the other hand, the size of each selected

, is large enough. This indicates

UL,l

is the processed noise matrix. By applying

the TDU-ESPRIT algorithm to those matrices *Y* (iBS) L

UL,l

l=1

{ }

again, we can obtain the more accurate angle estimates until

that at the initial angle estimation stage, we can achieve the

sufficient full-dimensional beamforming gain with the aid of rough angle estimates to effectively combat the severe path

BS

× *M*

¯ v

BS

the maximum number of iterations *i*max

BS

is reached, i.e.,

loss of long-distance THz links and improve the receive SNR.

*i*BS = *i*max. Finally, the estimates of azimuth and elevation

BS

angles and the corresponding virtual angles at BSs can be

1. *Fine Angle Estimation at Aircraft:* Due to the channel reciprocity of UL and DL, the acquisition of fine angle esti-

BS (imax) BS

(imax) BS

(imax)

denoted as

*θ*^l = *θ*^l BS

, *ϕ*^l = *ϕ*^l BS

, *µ*^l = *µ*^l BS ,

mates at aircraft in DL is similar to the fine angle estimation at

BSs. At this stage, instead of using the rough angle estimates, the fine angles estimated at BSs in Section III-A1 can be used not only to design the analog precoding vectors at BSs for beam alignment with improved receive SNR, but also to refine the GTTDU modules at BSs. Specifically, we consider

l

l

BS

m=1

BS

BS

BS

low-dimensional fully-digital array at the *l*th subarray of aircraft. For the received signal model in (22), we can also utilize the proposed prior-aided iterative angle estimation algorithm in Algorithm 1 to obtain the more accurate angle estimates. By replacing the input parameters

*I*AC = *I*h *I*v OFDM symbols to estimate the fine azimuth

AC AC

and elevation angles at aircraft, where the size of the equivalent

AC

AC

{*µ*˜BS*, ν*˜BS*,* {I[m]}IBS *, s*UL,l*, M*¯BS*, I*BS*, I*h

for BSs with the corresponding parameters

l

l

AC,l

n=1

*, I*v *, i*BS*, i*max}

AC

AC

low-dimensional fully-digital array is *I*h × *I*v . Based on

{*µ*^BS*, ν*^BS*,* {I[n] }IAC *, s*DL,l*, M*¯AC*, I*AC*, I*h *, I*v *, i*AC*, i*max}

*µ*l *, ν*l }l=1, the analog precoding vector

^ ^

the estimated {

BS

BS

L

can be designed as *f*RF,l = *a*BS(*µ*BS*, ν*BS) for 1 ≤ *l* ≤ *L*.

for aircraft, the estimates of azimuth and elevation angles

and the corresponding virtual angles at aircraft can be

^l ^l

selected antenna index in the *n*th OFDM sysmbol at the *l*th

(imax)

, *ϕ*AC

, *µ*AC

AC )

*ν*AC

(imax l

l

(imax)

l

l

(imax)

AC

l

l

By employing the reconfigurable RF selection network, the

obtained as *θ*^AC = *θ*^ AC

^ = *ϕ*^ AC

^ = *µ*^ AC

, and

aircraft subarray is denoted by I[n] with *M*¯ = |I[n] | .

˜ ˜AC

I

AC,l

AC

AC,l c

Then, initialize the analog combining vector as *w*[n]

= 0N

,

*B. Fine Doppler Shift Estimation under Doppler-Squint Effect*

RF,l

AC

^l = *ν*^l for 1 ≤ *l* ≤ *L*.

and then let [*w*[n]

RF,l I

] [*n*]

= √ 1

AC*,l*

l

l

[1]

AC*,l*

M¯

[*a* (*µ*AC*, ν*AC)]

, for

Based on the fine angle estimates above, the analog com-

1 *n I*AC, 1 *l L*.

≤ ≤ ≤ ≤

According to the DL transmission in (2), at the *l*th RF chain of aircraft, the received signal *y*[n] [*k* ] at the *k* th subcarrier

DL,l

l

l

of the *n*th OFDM symbol corresponding to the *l*th BS can be

expressed as

bining vectors of *L* subarrays at aircraft are designed to

achieve beam alignment, i.e., initialize *w*RF,l as *w*RF,l = 0NAC

and then let [*w*RF,l]I = √ 1 [*a*AC(*µ*^AC*, ν*^AC)]I for

≤ ≤

AC*,l*

MAC

l

l

AC*,l*

1 *l L*. The GTTDU module at aircraft can be also

refined to further mitigate the delay-beam squint effects. Since

*y*[n]

[*k* ] =√*P* (*w*[n]

) *H*˜ [*k* ]*f*

*s*[n]

[*k* ] + *n*[n]

[*k* ]*,*

the rough Doppler shift estimates are not precise enough

for data transmission, we will use *N*do OFDM symbols to

estimate the fine Doppler shifts in DL, where how to solve the

where *k*l ∈Kl, 1 ≤ *n* ≤ *I*AC, *H*˜DL,l[*k*l] is the compensated DL

H ′[n]

AC

DL,l

l

l

RF,l

DL,l

′[n]

l

RF,l

DL,l

l

DL,l

l

(20)

channel matrix, and *s*[n]

[*k*l] and *n*[n]

DL,l

DL,l

[*k*l] are the transmitted

Doppler squint effect is also considered. To ensure the effective

channels within multiple OFDM symbols to be quasi-static

pilot signal and noise, respectively. Considering the received

signals at *K*l subcarriers of *I*AC OFDM symbols, we can obtain the DL received signal matrix *Y*DL,l ∈CIAC×K*l* as

observed at the aircraft, the transmitters at BSs still need to

perform rough Doppler shift pre-compensation on the transmit signals at this stage.

According to the compensated DL channel matrix

*Y*DL,l

′[m¯ ]

[m¯ ]

√ jπf τ H

=

*P*l*G*l*α*l*e*

*s l*

*W*¯ RF,l*A*DL,l*f*RF,l(*a*τ (*µ*l *, K*l) ◦ *s*DL,l)

of the *m*¯ th OFDM symbol observed from the *l*th aircraft RF

τ T

*H*˜DL,l[*k*l], the received signal *y*do,l[*k*l] at the *k*lth subcarrier

` *s*¯D˛L¸*,l*

x chain can be expressed as (23) on the bottom of this page.

* *Y*˜DL,l + *N*DL,l*,* (21)

˜

h*w*

i

In (23), *k*l ∈ Kl, 1 ≤ *m*¯

≤ *N*do, ∆*ψ*˜l,k*l* = *ψ*l,k*l* − *ψ*˜l,k*l* is

where

· · ·*w*

*W*¯ RF,l =

[1] RF,l

[IAC] RF,l

∈ CNAC×IAC

the residual Doppler shift after compensation with *ψ*l,k*l* being

the rough Doppler shift estimates at the *k*lth subcarrier, and

*s* [*k* ] = *s*[m¯ ] [*k* ] for 1 ≤ *m*¯ ≤ *N* and *n*[m¯ ] [*k* ] are the

l,k*l*

is too small to effectively estimate fine Doppler shifts using

do,l

l

do,l

l

do

do,l

l

and

*Y*˜DL,l are the analog combining and residual

DL,l

=

l

1

l

K

*l*

transmitted pilot signal and noise, respectively. Since ∆*ψ*˜

h*s*[n]

beam squint matrices, respectively, *s*DL,l = *s*[n]

DL,l

*e*−j2πψ˜*l,kl* (m¯ −1)Tsym of *y*[m¯ ] [*k*l] in (23) can be removed to

[{K } ]· · ·*s*[n] [{K } ]iT ∈ CK for 1 ≤ *n* ≤ *I* ,

the limited OFDM symbols, the compensated phase difference

do,l

steering vector associated with path delay *τ*l can be defined as obtain

DL,l

*l*

AC

and *N*DL,l is the corresponding noise matrix. In (21), the

τ j({K*l*}1−1)µ*τ* j({K*l*}2−1)µ*τ* j({K*l*}*K* −1)µ*τ* T *ψ*

*a*τ (*µ*l *, K*l) =

*e*

*l e*

*l* · · · *e*

*l*

*l*

*y*¯[m¯ ] [*k* ] = *γ*

*e*j(m¯ −1)ν*l s*¯

[*k* ]

l

with *µ*τ = −2*πf*s*τ*l*/K* being the virtual delay. Similar to

do,l l

do,l do,l l

2πf*s*v*l* k*l*−1 1

j

(17), *Y*DL,l can be rewritten as

¯

AC

* *Y*DL,l + *N*DL,l*,* (22)

× *e* c ( K − 2 )(*m*¯ − 1)*T*sym +*n*¯[m¯ ] [*k*l]*,* (24)

` ˛¸ x

˜

RF,l

l

*y*[m¯ ] [m¯ ]

DL,l

DL,l

l

l

DL,l

do*,l*

*l*

*l*

where *γ*DL,l = √*P*l*G*l*α*l*e*jπf*s*τ*l* (*w*[1]

)H*A*DL,l*f*RF,l, and

where *ν*ψ = 2*πψ*z,l*T*sym denotes the virtual Doppler shift, and

l

AC

do,l

do,l

do,l

*Y* = *γ a*¯ (*µ*AC*, ν*AC)*s*¯T ˜

y[*m*¯ ] [k ](v )

*a*¯AC(*µ*AC*, ν*AC) = *a*v(*ν*AC*, I*v

l

l

l

AC

is the effective array response vector of equivalent

respectively. Considering the signals at *K*l subcarriers of *N*do

) ⊗ *a*h(*µ*AC*, I*h

) ∈ CIAC

˜ [*k*l] and *n*¯ [*k*l] are the Doppler squint value and noise,

*y*[m¯ ] [*k* ] = √*P w*H

do,l

l

l

RF,l

l

l

l

RF,l

DL,l

′[m¯ ]

*H* [*k* ]*f*

˜

DL,l

l

RF,l

l

RF,l

do,l

l

do,l

l

*s*[m¯ ] [*k* ] + *n*[m¯ ] [*k* ]

*l*

*l*

= √*P G α e*jπf*s*τ*l e*j2π∆ψ˜*l,k* (m¯ −1)Tsym *w*H *A*˜

[*k* ]*f*

*e*j(k*l*−1)µ*τ s*

[*k* ] +*n*[m¯ ] [*k* ]*.* (23)

` γ˛do¸*,l* x

do,l

l

do,l

l

` s¯do˛*,l*¸[k*l*] x

Algorithm 2: Proposed Prior-Aided Iterative Doppler

At the *i*do

th iteration, by exploiting the acquired

*v*(ido−1)

Shift Estimation Algorithm

Input: Estimated virtual angles {*µ*^*l , ν*^*l , µ*^*l , ν*^*l* },

}

BS

BS

AC

AC

*Kl*

l

l

l

l

*l,kl*

*kl*=1

*z,l*

be designed as *Y*˜do,l(*v*^l ), and its (*m*¯ *, k*l)th element is

^l

−

at the (*i*do 1)th iteration, the compensation matrix can

(ido−1)

*y*[m¯ ] (ido−1)

l

rough Doppler shift estimates {*ψ*˜ and *ψ*˜ ,

˜ [*k* ](*v*^ ), which can be acquired by replacing *v* of

transmitted pilot signal {*s*do*,l*[*kl*]}*Kl* , maximum

^ ) ◦ *Y*

iterations *i*max

*kl*=1

do,l

˜ [*k* ](*v* ) in (24) with *v*^ . The compensated receive

do , wavelength *λz* at central carrier

*y*[m¯ ] (ido−1)

do,l

l

do,l

frequency, and dimensional parameters

{*N*AC*, M*AC*, N*do*, Kl*}

as

do,l

do,l

l

matrix *Y* (ido) = *Y*˜ ∗

(*v*(ido−1)

can be then rewritten

Output: Doppler shift estimates *ψ*^*z,l* at center frequency and

*Y* (ido) =*γ*do,l *a*ψ(*ν*ψ*, m*¯ )*s*¯T

{*ψ*^*l,k*}*K* at all subcarriers

*k*=1

do,l

l do,l

1 % Preliminary (signal transmission and preprocessing)

2 Determine antenna index IAC*,l* and initialize *w*RF*,l* = 0*N*AC

;

l

do,l

l

do,l

AC*,l*

*l*

*l*

AC*,l*

do,l is the associated noise matrix. According to

◦ *Y*˜do,l(*v* ) ◦ *Y*˜ ∗

(*v*(ido−1)

+ *N* (ido)*,* (26)

AC AC

*M*AC

*l*

*l*

3 Let [*w*RF*,l*]I = √ 1 [*a*AC(*µ*^ *, ν*^ )]I and

where *N* (ido)

*f*RF*,l* = *a*BS(*µ*^BS*, ν*^BS);

· · ·

*Y* (ido)

in (26), we can obtain the Doppler shift estimate

4 for *m*¯ = 1*, , N*do do

^ )

· · ·

5 for *kl* = 1*, , Kl* do

6 Transmit pilot signal *s*do*,l*[*kl*] to obtain received

(ido) (i )do

[*m*¯ ]

do,l

at the center frequency of the *i*doth iteration, denoted by

(ido)

^

*ψ*z,l , using Total Least Squares ESPRIT (TLS-ESPRIT) [50].

(ido)

signal *y*do*,l*[*kl*] in (23);

By employing this estimated *ψ*^z,l to calculate the finely

7 Remove compensated phase *e*−j2*πψ*˜*l,kl* (*m*¯ −1)*T*sym of

*y*[*m*¯ ] [*k* ] to obtain *y*¯[*m*¯ ] [*k* ] in (24);

l

z,l

z

relative radial velocity, i.e., *v*^ = *ψ*^ *λ* , we can design

8

11 % Prior-aided iterative Doppler shift estimation

9 end

end

do*,l l*

[*m*¯ ]

do*,l l*

*Kl N*do

fine compensation matrix to further improve the accuracy of

Doppler estimation. Finally, at the *i*maxth iteration, we can

do

z,l

obtain the fine estimates of Doppler shift corresponding to

10 Gather {{*y*¯do*,l*[*kl*]}*k* =1}*m*¯ =1 into *Y*do*,l* in (25);

(imax)

*l L* BSs, i.e.,

(0)

13 while *i* ≤ *i* dodo

^

˜ ^

[*m*¯ ]

(*i*

*l*

−1)

*ψ*^z,l = *ψ*^ do

, which can be extended to all

12 Initialize: *i* = 0 and *v*^ = *ψ*˜ *λ* ;

subcarriers {*ψ*^l,k}K for 1 ≤ *l* ≤ *L*. The proposed prior-

do

do*,l*

do*,l*

Obtain Doppler shift estimate of *i*doth iteration as

max

do

14 if *i*do = 0 then

*l*

(0)

*l*

(*i*do)

*z,l z*

k=1

aided iterative Doppler shift estimation algorithm above is

˜de,l

will be utilized to estimate the path delays in DL. Recall that

τ

l

l

*l* 1

*l e*

*l* 2

*l*

*e*

*l Kl*

*l*

l

de,l

summarized in Algorithm 2, where the Doppler squint effect

15

l

[m¯ ]

˜

(0)

*Y*do,l = *γ*do,l

*a*ψ(*ν*

ψ

l

*, N*do)*s*¯T

* *Y*˜do,l(*v*l) + *N*do,l*,* (25)

T

*s*¯do,l = [*s*¯do,l[{Kl}1]· · ·*s*¯do,l[{Kl}K*l* ]]

∈ CK*l* , *Y*˜do,l(*v*l) with

l

*l*

do,l

l

and noise matrices, respectively.

on (25), we propose the following prior-aided iterative Doppler

16 else

17

Obtain estimate *ψz,l* (for comparison in simulations)

by applying TLS-ESPRIT algorithm to *Y*do*,l*;

Design compensation matrix *Y*˜do*,l*(*v*^(*i*do −1)), whose

can be addressed well.

*C. Path Delay and Channel Gain Estimation*

de,l

de,l

At the path delay estimation stage, the fine Doppler shift

de,l

de,l

l

τ

1

de,l

[n¯]

l

[n¯]

K*l*

[n¯]

denote the error and noise vector, respectively. Considering

the received signals of *N*

de

the matrix *Y*de,l =

*y*

· · · *y*

de

∈CK*l*×Nde as

de,l

de,l

l

de,l

(*m*¯ *, kl*)th entry is *y*do*,l*[*kl*](*vl* do );

18 Obtain compensated matrix

estimates above can be used to accomplish the fine Doppler

*Y* (*i*do) = *Y*˜ ∗

(*v*^(*i*do−1)) ◦ *Y*do*,l* in (26);

compensation as shown in Fig. 3, and *N*de OFDM symbols

19 Apply TLS-ESPRIT algorithm to *Y*do*,l* ;

20

*a* (*µ*τ *, K* ) = *e*j({K } −1)µ*τ* j({K } −1)µ*τ* · · · j({K } −1)µ*τ* T

*ψ*^*z,l* and calculate *v*^*l* = *ψ*^*z,l λz* ;

do,l

in (21) denotes the steering vector associated with path delay

(*i*do)

21 end

(*i*do)

(*i*do)

*τ*l, and *µ*τ =−2*πf*s*τ*l*/K*. The DL received signal *y*[n¯] [*k*l] at

22 *i*do = *i*do +1

23 end

(*i*max)

the *k*lth subcarrier of the *n*¯th OFDM symbol can be expressed

as (27) on the top of the next page. In (27), *k*l ∈ Kl, 1 ≤

OFDM symbols, we can obtain

24 Return: *ψ*^*z,l* = *ψ*^ do

and extend it to all subcarriers

*n*¯ ≤ *N*

, *s*[n¯]

= *s*[n¯] [*k* ] for *k* ∈ K

is the transmitted pilot

{*ψ*^*l,k* }*K*

*z,l*

de de,l

[n¯]

de,l l l l

*k*=1

signal8, *y* [*k*l] is the error value including the residual beam-

OFDM symbols, we can acquire the received signal matrix

z,l

*Y*do,l ∈CNdo×K*l* as

Doppler squint errors caused by the channel estimation error,

and *n*[n¯] [*k*l] is the noise. By collecting all received signals at

*K*l subcarriers into the vector *y*[n¯] ∈CK*l* , we have

*y*[n¯]

= h*y*[n¯] [{K } ] · · · *y*[n¯] [{K } ]iT

h *ψ ψ* iT

= *γ*de,l*a*τ (*µ*l *, K*l)*s*¯de,l ◦ *y*˜de,l + *n*de,l*,* (28)

where *a*ψ(*ν*ψ*, N*do) =

1 *e*jν*l* · · · *e*j(Ndo−1)ν*l*

∈ CNdo de-

[n¯]

h [n¯]

[n¯]

iT [n¯]

notes the steering vector associated with the Doppler shift *ψ* ,

where

*y*˜de,l =

*y*˜de,l[{Kl}1] · · · *y*˜de,l[{Kl}K*l* ]

and *n*de,l

[*Y*˜do,l(*v* )]m¯ ,k = *y*˜ [*k*l](*v* ) and *N*do,l are the Doppler squint

h [1] [N ]i

To attenuate the impact of Doppler squint matrix *Y*˜do,l(*v*l)

*Y*de,l = *γ*de,l *a*τ (*µ*τ *, K*l)*s*¯T ◦ *Y*˜de,l + *N*de,l*,* (29)

shift estimation algorithm. Define the rough Doppler shift

estimate at the central carrier frequency as *ψ*z,l, and the

initially relative radial velocity is given by *v*^l = *ψ*˜z,l*λ*z.

8Note that we assume the same pilot signals are adopted by *Kl* subcar-

riers, which maybe lead to the high PAPR in OFDM systems. Fortunately, we can utilize a predefined pseudo-random descrambling code spread at all subcarriers [37] to reduce the high PAPR effectively.

*y*[n¯] [*k* ] =√*P w*H

˛¸

de,l

l

l

RF,l

`

l

l

l

RF,l

γde*,l*

DL,l

RF,l

x

′[n¯]

*H* [*k* ]*f*

˜

DL,l

l

RF,l

*s*[n¯] [*k* ] + *n*[n¯] [*k* ]

de,l

de,l

l

de,l

l

*l*

[n¯] [n¯]

= √*P G α e*jπf*s*τ*l w*H *A f*

*e*j(k*l*−1)µ*τ e*j2π(ψ*z,l*−ψ^*z,l*)(n¯−1)Tsym *s*[n¯]

de,l

l

de,l

l

[*n*¯ ]

de*,l*

·*y*˜ [*k* ] + *n* [*k* ]*.* (27)

where *s*¯

de

de,l

de,l

= h*s*¯[1] · · ·*s*¯[Nde]iT ∈ CN , and *Y*˜

de,l

= save numerous pilot overhead as the time-varying channels

˛¸s¯

x

`

method utilizes the channel correlation of two adjacent OFDM

h*y*˜[1] · · ·*y*˜[Nde ]i and *N* are the residual beam-Doppler

should be updated frequently. The proposed DADD-based

squint and noise matrices, respectively. By exploiting the TLS-

de,l

de,l

de,l

de,l

ESPRIT algorithm [50], we can obtain the path delay estimates

corresponding to *L* BSs, i.e., *τ*l L . From (29), we observe that the accuracy of path delay estimation depends on the angle and Doppler estimation accuracy, and this conclusion can be further verified by the simulation results in Section VII.

l=1

{^ }

To estimate the channel gains, we need to harness the received signal matrix *Y*de,l in (29). Specifically, this matrix *Y*de,l can be split into the equivalent channel gain *α*¯l and

√

*Y*¯ , i.e., *Y* = *α*¯ *Y*¯ , where *α*¯ = *P G α* . Regardless

de,l

de,l

l

de,l

l

l

l

l

of the residual beam-Doppler squint and noise matrices of *Y*de,l, we can then utilize the previously estimated dominant channel parameters, i.e., the azimuth/elevation angles at BSs and aircraft, Doppler shifts, and path delays, to reestablish the

^

symbols, where the estimated channels in the previous symbol

can be approximately regarded as the real-time channels of

the next symbol to detect the data sequentially. Meanwhile, the powerful error correction capability of the channel coding (e.g., Turbo or LDPC codings) can correct part of the erro- neous detected data to minimize error propagation during the decision-directed process. Note that at the data transmission stage, we consider *L* BSs can simultaneously serve the aircraft using the same time-frequency resource to achieve the high spectrum efficiency, i.e., signals associated with different BSs can be distinguished in the spatial domain, rather than the OFDMA utilized for the initial channel estimation. The pro-

posed DADD-based channel tracking algorithm is summarized in Algorithm 3.

¯ ¯

estimated matrix of *Y*de,l as *Y* de,l. Finally, we can obtain the

estimation of *α*¯l, denoted by *α*^l, as

N

K

*α*

Specifically, considering the *r*th OFDM symbol with *r* =

(*q*−1)*N*C+*p* that corresponds to the *p*th OFDM symbol of the

*q*th TI, the DL channel matrix *H*[n]

[*k*] in (4) can be rewritten

1 Σde Σ*l*

^ *N*

, h ^¯ i

[r]

DL,l

[q]

[q]

l =

de

*K*l n¯=1 k*l*=1

[*Y*de,l]k*l*,n¯

*Y* de,l

*.* (30)

k*l*,n¯

as *H*DL,l[*k*], which contains the channel parameters *G*l , *α*l ,

*ψ*[q], *v*[q], *τ* [q], *θ*AC[*q*], *ϕ*AC[*q*], *θ*BS[*q*], and *ϕ*BS[*q*]. Define the

z,l l l l l l l

l

1. DATA-AIDED CHANNEL TRACKING

In Section III, we have acquired the estimates of dominant

channel parameters, which will be used for the following data

transmission. Although THz UM-MIMO-based aeronautical communication channels exhibit the fast time-varying fading characteristic caused by the large Doppler shifts, the variations

of dominant channel parameters, including angles, delays,

initial data sequence in the *r*th OFDM symbol at the *l*th BS

as *x*[r], and this sequence can be mapped to *K* subcarriers

h i

l

l

l

via channel coding and modulation to obtain the transmitted

signal vector, i.e., *s*[r] = *s*[r][1]· · ·*s*[r][*K*] T ∈ CK. The DL

baseband signal vector *y*[r][*k*] CL received by aircraft at the

∈

*k*th subcarrier of the *r*th OFDM symbol can be expressed as

*y*[r][*k*] = h*y*[r][*k*] · · · *y*[r][*k*]iT

Doppler shifts, and channel gains, can be relatively smooth

within very transitory duration time *T*sym. Hence, we regard

1 L

ΣL √

H

′[r]

[r]

[r] !

the duration time of *N*C

˜

*y*l [*k*] =

*P*l*w*RF,l*H*DL,l[*k*]*f*RF,l *s*l [*k*] + *w*RF,l

OFDM symbols as a Time Interval

= *W*RF

l=1

*P*l*H*˜DL,l[*k*]*f*RF,l*s*l [*k*] + *n*

[*k*] *,*

(TI), and the channel parameters within this TI are assumed

*l*

*l*′=1

*P*l′ *H*DL,l′ [*k*]*f*RF,l′ *s*l′ [*k*] + *n*l [*k*] *.* (32)

*l*

*l*

′

to be stationary. Note that after the rough or fine Doppler

compensation, the channel related to each OFDM symbol

within the same TI is still slowly changing due to the imperfect

Doppler compensation. Hence, after a long period of accu-

mulation, the channels can change obviously, which would

drastically degrade the detection accuracy of received data. To

improve the reliability and efficiency of data transmission, a

(31)

where 1 *k K*, *W*RF =[*w*RF,1 *w*RF,L], and *n*[r][*k*] is the

l

l

≤ ≤ · · ·

˜

*l*

noise vector. In (31), the *l*th received signal *y*[r][*k*] in *y*[r][*k*]

corresponding to the transmitted signal of the *l*th BS is given

by (32) on the bottom of this page. In (32), the second entry is

the interference from other BSs, *n*[r][*k*] is the combining noise,

and *h*[r][*k*] and *z*[r][*k*] are the beam-aligned effective channel

l l

DADD-based channel tracking algorithm is developed to track

the beam-aligned effective channels in real-time, which would

coefficient and interference plus noise, respectively. Note that

the interference entry in (32) is regarded as the additional noise

[r]

√ H ′[r]

[r]

H ΣL √

′[r]

[r]

[r]

` h[˛*r*]¸[k] x `

z[˛*r*]¸[k] x

Algorithm 3: Proposed DADD-Based Channel Track-

ing Algorithm

*s*[r][*k*] =h*s*[r][*k*]· · ·*s*[r][*k*]i∈CL can be estimated as

Input: Estimated channel parameters

^h [k]

1

L

BS

AC

AC

{*θ*^*l , ϕl , θ*^*l , ϕl , ψ*^*z,l, τl, αl*}*l*=1, dimensional

parameters {*K, L, N*C*, K*˜}, and preset threshold ratio

BS

^ ^ ^ ^

*L*

*s*^[r][*k*] = h*s*^[r][*k*] · · · *s*^[r][*k*]iT = *W* [r] [*k*] H *y*[r][*k*]*,* (33)

*l*

1

L

BB

[r]

[*k*] is *s*[r][*k*] =

[*r*]

*l*

[*r*−1]

*s*[r][*k*]+

z[*r*][k]

[*r*−1]

.

[*r*] *L*

*ε*

l

^

^h*l* [k] ^h*l* [k]

Output: Estimated effective channel vector {*hl* }*l*=1 and

[*r*] *L*

detected data sequence {*xl* }*l*=1 for *r* = 1*,* 2*,* 3*,*· · ·

^

1 Initialize: K˜*l* =∅ and

*f τ*

[0]

−j2*π k*−1 − 1

[0]

*s* ^

RF*,l*

By extracting the received signal processed by the *l*th RF chain

and gathering these signals at *K* subcarriers, the estimation of transmitted signal vector *s*l can be denoted by *s*^l ∈

CK . To

^*s*

[r]

[r]

[r]

^l

where the *l*th entry of *s*^

track the effective channel of the current *r*th OFDM symbol,

l

l

*K* 2 *l* H

^*hl* [*k*] = *α*^*le w A*^DL*,lf*RF*,l* for

1 ≤ *k* ≤ *K* and 1 ≤ *l* ≤ *L*;

i.e., *h*[r], this signal vector [r] can be demodulated and

2 for *q* = 1*,* 2*,* 3*,*· · · do

3

for *p* = 1*,*· · ·*, N*C do

decoded as the detected data sequence *x*^l (i.e., the estimate

4 *r* =(*q* 1)*N*C +*p*;

−

[*r* 1]

˜ − ˜if |K | ≤ *K* for 1 ≤ *l* ≤ *L* then*c*

5 *l*

6 Map initial data sequence {*x*[*r*]}*L* to

*l*

*l*=1

7

of initial data sequence *x*l ). This data sequence *x*l can be

then coded and modulated again to yield the transmitted signal vector *s*˜l , which should be more accurate than the estimated

^

[r]

[r]

˜

[r]

[r]

sidering *s*l as the pilot signal, we substitute its *k*th element,

denoted by *s*[r][*k*], into the received signal *y*[r][*k*] in (32) to

[r]

*l*

transmitted signal vector {*s*[*r*]}*L*

{ }

Obtain baseband signal vector *y*[*r*][*k*] *K*

(31), whose *l*th entry is *y*[*r*][*k*] in (32);

i.e., *h*l [*k*] = *y*l [*k*]*/s*l [*k*]. Finally, considering *K* subcarriers,

the estimated effective channel vector of the *r*th OFDM

*l*=1

;

*s*^l thanks to the error correction of channel coding. By con-

*l* ˜l

*k*=1

in

8

Design the digital combining matrix

acquire the estimate of effective channel coefficient *h*l [*k*],

l

[r]

˜

*W* [*r*] [*k*] =diag(^*h*[*r*−1][*k*]· · ·^*h*[*r*−1][*k*]) for

1

BB

≤ *k* ≤

*K*;

[*r*]

1

*K*

*L*

[r]

[r]

[r]

9 Obtain *s*

^

{^

[r] K

[*r*] *L*

[*r*] *L*

{*s*˜ }

as pilot signal;

*l*

*l*=1

[*r*] *L*

[*k*]}*k*=1 in (33) and extract

[*r*] *L*

symbol is *h*^l ∈ C for 1 ≤ *l* ≤ *L*. Accordingly, the digital

{*s*^*l* }*l*=1 to restore data sequence as {*x*^*l* }*l*=1;

^

^

[*r*] *L*

[*r*]

symbol can be designed as *W*BB [*k*] =diag(*h*1 [*k*]· · ·*h*L [*k*]),

[*r*]

which is used to perform the subsequent channel equalization.

[*r*]

[*r*]

[*r*]

Furthermore, by utilizing the previously estimated channel pa-

*l*

*l*=1

*l*

for *k* = 1*,*· · ·*, K* and *l* = 1*,*· · ·*, L* do

combining matrix at the *k*th subcarrier in the (*r*+1)th OFDM

10 Code and modulate {*x*^*l* }*l*=1 again to yield

.

*r* = 1*,* 2*,* 3*,*· · · ;

13

.^*h*[*r*][*k*] − ^*h*[*r*−1][*k*]. *> ε*

*K*

.^*h*[*r*−1][*k*].

DL array response matrix in (5) using the fine angle estimates.

[*r*]

*K*

[*r*]

*k*=1

*l*

[r+1]

[r]

[r]

11 Substitute *s*˜*l* [*k*] into (32) to acquire

[r] [r− *ε*1]

[r−1]

˜ ˜involving erroneous coefficients composes a set K . Let *K*

*l*

l

l

*K*

k=1

l

^*hl* [*k*] = *yl* [*k*]*/s*˜*l* [*k*] for 1 ≤ *k* ≤ *K* and

rameters at the initial channel estimation stage, the estimates of

1 ≤ *l* ≤ *L*;

BB

l

∈

[r−1]

^− ≤ ≤

The *K* channel coefficients {*h*[r][*k*]}K

l

k=1

[0] L

12 Collect {*h*^ }

and initialize K˜ =∅ for

initial beam-aligned effective channel vectors {*h*^l }l=1 can be

1 ≤ *l* ≤ *L*;

— *s* ^*l*

obtained as ^*h* [*k*] = *α*^ *e* K 2 *w*H *A*^ *f*

−j2π[0]

f τ

k−1

1

l

l

RF,l

DL,l

RF,l

14 Satisfy

*l*

. Σ .

. for 1 ≤ *k* ≤ *K* and 1 ≤ *l* ≤ *L*, where *A*^DL,l is the reconstructed

15

16 else

end

in (34) and let K˜*l* = K˜*l* ∪*k*;

As the time goes on, the previously estimated channel

parameters will not match the current effective channels.

^channel vector *h* in the *r*th OFDM symbol, its *k*th channel

[r]

^

off the pilot-aided channel tracking in Section V.

^

Therefore, the quality of the tracked effective channel vectors

[*r*] *L* [*r*] *L*

17

18

19

20

21 end

end

end

Return: {*h*^*l* }*l*=1 and {*x*^*l* }*l*=1 for

Terminate current algorithm and trigger off

pilot-aided channel tracking.

at the data-aided channel tracking stage should be monitored

in real-time by exploiting the temporal correlation of two adja-

cent OFDM symbols. Specifically, for the estimated effective

[r]

l

coefficient *h*l [*k*] can be regarded as a wrong coefficient if

[r]

*h*l [*k*] satisfies

.^*h* [*k*] − ^*h* [*k*]. *>* ΣK .^*h* [*k*].*,* (34)

due to the small interference from other BSs caused by the

large angle differences among different BSs and the extremely

narrow beams formed by THz UM-MIMO array at aircraft.

where *ε* is a preset threshold ratio. The indices of subcarriers

[r]

l

as the acceptable number of erroneous channel coefficients,

the tracked effective channel vectors can be regarded as the

[r]

[r]

[r]

[r]

[r]

Thus, (32) can be rewritten as *y*l [*k*] = *h*l [*k*]*s*l [*k*]+*z*l [*k*].

can form together

invalid estimates if |K˜l |c *> K*˜ for 1 ≤ *l* ≤ *L*, which will trigger

the beam-aligned true effective channel vector *h*[r] CK at

the *k*th subcarrier in the *r*th OFDM symbol.

Based on the estimated effective channel coefficient in the

(*r* 1)th OFDM symbol, denoted by *h*l [*k*] for 1 *l L*,

we can design the digital combining matrix as *W* [r] [*k*] =

1. PILOT-AIDED CHANNEL TRACKING

In this section, the previously estimated channel parameters

in Section III will be exploited as the prior information for

facilitating the pilot-aided channel tracking. This is because

according to the previous analysis, these channel parameters

[r−1] [r−1]

diag(^*h*1 [*k*]· · ·^*h*L [*k*]). According to (31), the signal vector

including angles, Doppler shifts are changing slowly and

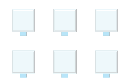
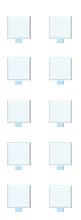
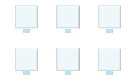
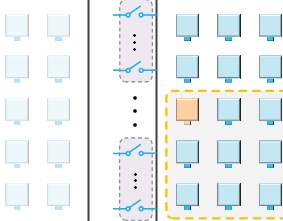
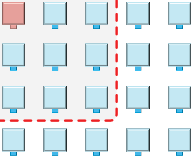
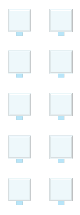
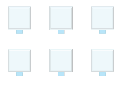
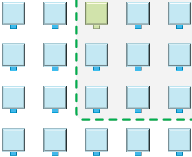
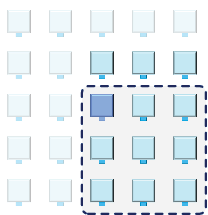
$61 6XEDUUD\ 1

**...**

**..**

**.**

**...**



(TXiYDOHQW

IXOO\-GiJiWDO 2*d*

VSDUVH DUUD\

**.**

**.**

**.**

**.**

**.**

**.**

**.**

**.**

**.**



$61

6XEDUUD\ 2



$61

6XEDUUD\ 3



$61

6XEDUUD\ 4



H

|  |  |  |  |  |
| --- | --- | --- | --- | --- |
| 1VW 2)D0 V\PERO | 2QG 2)D0 V\PERO | 3UG 2)D0 V\PERO | 4WK 2)D0 V\PERO | 7iP |



Fig. 8. The schematic diagram of subarray selection based on different antenna connection patterns of the reconfigurable RF selection network at the angle tracking stage. Taking the UPA of size 5 5 as an example, this UPA can be partitioned into 4 subarrays of size 3 3, and the interval between each subarray is the width of two antennas. The same RF chain sequentially selects the corresponding subarrays in 4 successive OFDM symbols to receive signals, and these received signals will be equivalent to the signals received by a low-dimensional fully-digital sparse array of size 2×2 with the sparse spacing *Ω* = 2.

× ×

BS BS BS BS

^ ^ ^ ^

can usually not vary dramatically. Since previously estimated

channel parameters can be more accurate than the rough estimates based on navigation information, the tracked chan- nel parameters at this stage would be more accurate than

*µ*¯l and *ν*¯l at each iteration. Note that *µ*¯l and *ν*¯l suffer

from the inherent angle ambiguity problem. To further address

this angle ambiguity issue, we define an ordered index set

}

B = −1*,* −1+ 1 *,* −1+ 2 *,*· · ·*,* 1 with |B|c = 2*Ω* +1, and let

BS

BS

Ω

BS

Ω

BS

BS

BS

′BS

those acquired at the initial channel estimation stage. The

main process of the pilot-aided channel tracking is similar

^ ˜ ˜^ ^

to the initial channel estimation in Section III. The difference

*µ*˜¯l = *µ*^¯l */Ω* and *ν*˜¯l = ^*ν*¯l */Ω*. Thus, the estimates of virtual

′BS

′BS

BS

⋆

′BS

BS

⋆

between them lies in that the azimuth and elevation angles

at BSs and aircraft in this section are estimated by forming

*ν*l , should satisfy *µ*l = *µ*¯l +*b*µ*π* and *µ*l = *ν*¯l +*b*ν *π*,

where *b*⋆ ∈ B and *b*⋆ ∈ B are the optimal indices. Due to

B

µ

ν

the array response vector of *equivalent low-dimensional fully- digital sparse array*. By contrast, an equivalent fully-digital

angles corresponding to *µ*˜¯l and ˜*ν*¯l , denoted by *µ*^l and

array with critical antenna spacing (i.e., the half-wavelength

the limited elements in , we adopt the exhaustive method

to search for these optimal indices *b*⋆ and *b*⋆. The previously

µ ν

estimated *µ*BS and *ν*BS in Section III-A1 can be regarded as the prior information, i.e., *µ*BS = *µ*BS and *ν*BS = *ν*BS, and *b*⋆

^ ^

l l

antenna spacing) is considered in Section III. The existing con- ⋆

˜l ^l

˜l ^l µ

clusions indicate that the usage of sparse array can improve the accuracy of angle estimation significantly, but these estimated

angles would suffer from the angle ambiguity issue [51], [52].

and *b*ν can be then obtained as

BS

*b* = arg min .*µ*˜¯ + *b π* − *µ*˜ . *,* (35)

⋆ BS

µ

b*µ* ∈B

l

µ

l

ν

b*ν* ∈B

l

ν

l

⋆ BS BS

Fortunately, this angle ambiguity can be solved with the aid

of the previously estimated angles. Due to space constraints,

*b* = arg min .*ν*˜¯ + *b π* − *ν*˜ . *.* (36)

*µ* BS

*ν* BS

this section focuses on the pilot-aided angle tracking at BSs.

′h

′v

Based on the acquired estimates

′BS

′BS

′

^l and

calculate the updated estimates of azimuth and elevation angles

′

^l , we can

equivalent fully-digital sparse array of size *I*BS ×*I*BS at BSs,

Specifically, *I*B′ S OFDM symbols are used to obtain the

′

′

where *I*B′ S = *I* h *I* v

subarrays can be acquired by reconfig-

at the *l*th BS as *θ*^l and *ϕ*^l , for 1 ≤ *l* ≤ *L*. The remaining

steps are the same as those in Section III-A1 except that

uring the dedicated connection pattern of the RF selection

BS BS

network. Define *Ω* as the sparse antenna spacing relative to the critical antenna spacing *d*. The size of the selected subarray is

¯ ′h

¯ ′v

¯ ′

¯ ′h

¯ ′ v

*M*B′ S ×*M*BS with *M*BS = *M*BS*M*BS ′ antenna elements, where

BS

BS

BS

BS

BS

BS

the exhaustive search in (35) and (36) should be taken into

account. Finally, we can obtain the fine estimates of azimuth and elevation angles at BSs, denoted by {*θ*^BS*, ϕ*^BS}L . In

l

l

l=1

′

′

*I*A′ C = *I* h *I* v

a similar way, the fine estimates of azimuth and elevation

l

l

l=1

′h v v ′v

*M*¯ h = *N* h − *Ω*(*I* − 1) and *M*¯ = *N* − *Ω*(*I* − 1).

Fig. 8 depicts an example that the UPA with size of 5 × 5

angles at aircraft can be also acquired as {*θ*^AC*, ϕ*^AC}L , where

AC AC

OFDM symbols are required. Moreover, with

can be divided into 4 subarrays of size 3 3, and these

×

subarrays construct the array response vector of equivalent fully-digital sparse array of size 2 2 with the sparse spacing *Ω* = 2. Similar to the fine angle estimation at BSs in

×

the help of the previously estimated Doppler shifts, the updated

Doppler shift estimates *ψ*z,l L via the pilot-aided channel tracking will be more accurate than those estimated at the initial channel estimation stage, and so do the estimates of

l=1

l=1

{ ^ }

Section III-A1, we can obtain the homologous UL received

I′ ×K

UL,l

path delays {*τ*^l}L

and channel gains {*α*¯l}L

. As shown in

signal matrix *Y*¯

∈ C BS

*l* in (17), where the effective

Fig. 3, the updated beam-aligned effective channels can be

array response vector of the sparse array can be expressed as

′ v

′ h

I′

*a*¯BS(*µ*¯l *, ν*¯l ) = *a*v(*ν*¯l *, I*BS) *a*h(*µ*¯l *, I*BS) C BS with

BS

BS

BS

BS

⊗ ∈

*µ*¯BS = *Ωµ*BS and *ν*¯BS = *Ων*BS. By exploiting the proposed

then used for the following data transmission, and the tracked

channel parameters will be regarded as the prior information for the next pilot-aided channel tracking.

l=1

l l l l

prior-aided iterative angle estimation in Algorithm 1 as before,

the estimates of *µ*¯BS and *ν*¯BS can be respectively obtained as

In order to intuitively describe the relationship among

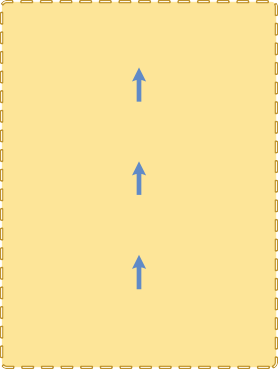
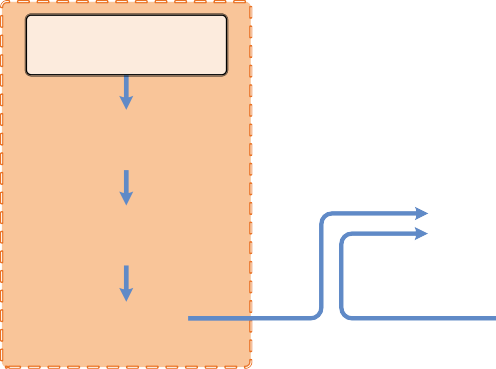
different channel estimation and tracking stages above, the

l l

(VWiPDWH IiQH 5RXJK DQJOH DQG DRSSOHU

**6WDUW**

1DYiJDWiRQ



**3iORW-$iGHG &KDQQHO 7UDFNiQJ**

DW %6V XViQJ $OJRUiWKP 1

VKiIW HVWiPDWHV

iQIRUPDWiRQ

(VWiPDWH DQG

<HV **5HVWDUW**

(VWiPDWH IiQH

DW DiUcUDIW XViQJ $OJRUiWKP 1

(VWiPDWH IiQH

XViQJ $OJRUiWKP 2

(VWiPDWH DQG

**IQiWiDO &KDQQHO (VWiPDWiRQ**

IQYDOiG cKDQQHO WUDcNiQJ

1R

**DDWD-$iGHG &KDQQHO 7UDFNiQJ**

XViQJ $OJRUiWKP 3

1R <HV

(VWiPDWH IiQH

XViQJ $OJRUiWKP 2

7UDcN IiQH

DW DiUcUDIW XViQJ $OJRUiWKP 1

7UDcN IiQH

DW %6V XViQJ $OJRUiWKP 1

Fig. 9. Flow diagram of the proposed channel estimation and tracking solution.

block diagram of the proposed channel estimation and tracking solution is illustrated in Fig. 9.

corresponding the log-likelihood function can be expressed as

(38) on the bottom of this page by defining *η*l =[*α*l*,* (*ξ*BS)T]T

l

with *ξ*BS = [*ν*¯BS*, µ*¯BS]T. Thus, the (*i,j*)th entry of Fisher

l l l

1. PERFORMANCE ANALYSIS

Information Matrix (FIM), denoted by [*G*(*η*l)]i,j, is given by

1. *CRLBs of Channel Parameters*

[*G*(*η*l)]i,j = −E

*∂*2 ln *p*(*Y*¯

; *η*l)

*.* (39)

UL,l

According to the effective received signal models in Sec- tion III, we will investigate the CRLBs of the dominant

*∂*[*η*l]i*∂*[*η*l]j

According to the results in [53], [54], the CRLB

channel parameters, i.e., azimuth/elevation angles at aerial

of *ξ*BS

consisting of the virtual angles

*µ*¯BS

and

BSs and aircraft, Doppler shifts, and path delays. Note that BS l l

practical triple squint effects of aeronautical THz UM-MIMO channels would weaken the accuracy of channel parameter

estimation, and these negative effects are not considered in

*ν*¯l can be expressed as (40) on the bottom of this

hpage. In (40), *B*BS,k*l* = *I*2 ⊗ *s*UL,l[{Kl}k*l* ], *Γ*BS i=

l

*, I*

*l* BS *,*

BS

*l*

*l* BS

*, I*

*a*v(*ν*¯BS

¯v

BS

)⊗

∂*a*h(µ¯BS,I¯h )

∂µ¯*l*

∂*a*v(ν¯BS,I¯v )

∂ν¯BS

⊗*a*h(*µ*¯

BS

l

¯h

BS

) ,

l

l

deriving the CRLBs. So these CRLBs serve as the lower-bound

of parameter estimation.

* 1. *CRLBs of Angle Estimation at BSs and Aircraft:* To investigate the performance at both the initial angle estimation stage and the following angle tracking stage, we consider the

arcsin

and the projection operator *Φ*BS = *a*¯BS(*µ*¯BS*, ν*¯BS)

× *a*¯H (*µ*¯BS*, ν*¯BS)*a*¯ (*µ*¯BS*, ν*¯BS) −1*a*¯H (*µ*¯BS*, ν*¯BS).

BS

l

l

BS

l

l

BS

l

l

To obtain the CRLBs of azimuth and elevation angles, we define the transformation relationship between the virtual angles and the corresponding physical angles as

received signal model corresponding to the equivalent fully-

BS

×*I*

*ϕ*l

digital sparse array with size of *I*¯h

¯v

BS

, where the sparse

" BS # 

BS

*θ*

BS

l

ν¯BS 

*l*

Ωπ

µ¯BS

*l*

*J*(*ξ*l ) =

BS

received signal model without considering the triple squint

spacing is *Ω* ≥ 1. Based on the expression of (17), the effective

Ωπ cos(ϕ*l* )

effects, denoted by *Y*¯ = [*y*¯UL,l[{Kl}1]· · ·*y*¯UL,l[{Kl}K ]] ∈

UL,l

*l*

CI¯BS×K*l*

, can be written as

= 

arcsin

 *.* (41)

Based on the transformation of vector parameter CRLB in [55], defining *∂J*(*ξ*BS)*/∂ξ*BS as the Jacobian matrix, the

l l

l l

¯ ¯ BS BS T ¯

*Y* = *γ a* (*µ*¯ *, ν*¯ )*s* + *N ,* (37)

UL,l UL,l BS l l UL,l UL,l

CRLBs of azimuth angle *θ*BS and elevation angle *ϕ*BS, denoted by CRLBθBS (*Ω*) and CRLBϕBS (*Ω*), can be then formulated

*l l*

Σ

where 1 ≤ *l* ≤ *L*, *I*¯BS =

BS BS

l

l

*I*¯h *I*¯v , *a*¯BS(*µ*¯BS*, ν*¯BS) =

as (42) and (43), respectively, on the top of the next page.

*a*v(*ν*¯BS*, I*¯v ) ⊗ *a*h(*µ*¯BS*, I*¯h

l

BS

l

BS

l

l

*Ωπ* sin(*θ*BS) cos(*ϕ*BS) and *ν*¯BS = *Ων*BS = *Ωπ* sin(*ϕ*BS), and

) ∈ CI¯BS with

*µ*¯BS = *Ωµ*BS =

Finally, the CRLBs of angles at BSs can be obtained as

CRLBθBS (*Ω*) = L

l=1 CRLBθBS (*Ω*) and CRLBϕBS (*Ω*) =

l=1 CRLBϕBS (*Ω*), respectively. Furthermore, the CRLBs

1

L

¯ l l l l

*Y*

UL,l

*N*UL,l is the noise matrix with its entry following CN(0*, σ*n).

L

The likelihood function of

¯

UL,l

is *p*(*Y*¯

l 2 1 ΣL *l*

; *η*l), and the

*l*

of angles at aircraft, i.e., CRLBθAC

(*Ω*) and CRLBϕAC

(*Ω*),

ln *p*(*Y*¯

1 Σ BS BS

; *η* ) = − *I*¯ *K* ln(*πσ*2 ) − *y*¯ [{K } ] − *γ a*¯ (*µ*¯ *, ν*¯ )*s* [{K } ] H

K*l*

*σ*

UL,l

l

BS

l

n

2

n k*l*=1

UL,l

l

k*l*

UL,l

BS

l

l

UL,l

l

k*l*

× *y*¯UL,l[{Kl}k ] − *γ a*¯ (*µ*¯BS*, ν*¯BS)*s*UL,l[{Kl}k ] *.* (38)

*l*

UL,l

BS

l

l

*l*

( Σ

−1 *σ*2

n

*ξ*

K*l*

H H

2

})−1

CRLB BS = *G*

*l*

(*η*l) =

2 |*γ*

UL,l|

ℜ

k*l*=1

*B*BS,k*l Γ*BS

*I*I¯BS − *Φ*BS

*Γ*BS*B*BS,k*l*

*.* (40)

h∂*J*(*ξ*BS)

*l*

∂*J*(*ξ*BS)T i

*l*

CRLB BS

*l* 1*,*1

ξ

CRLBϕBS (*Ω*) =

∂*ξ*BS

CRLB*ξ*BS

∂*ξ*BS

= 2 2 BS 2

*,* (42)

*l l l*

*l* 1,1

Ω (π −(ν*l* ) )

CRLB

h∂*J* (*ξ*BS)

*l*

∂*J* (*ξ*BS)T i

*l*

CRLBξBS

2*,*2

θBS (*Ω*) =

∂*ξ*BS

CRLB*ξ*BS

∂*ξ*BS

= 2 2 2 BS

BS 2

*.* (43)

*l l l*

*l*

2 ( ΣK*l*

*σ*n

ℜ

CRLBν*ψ* =

2

*l* 2,2

Ω (π

|*s*¯do,l[{Kl}k ]|

2

∂*aψ*(ν*ψ* ,Ndo)

*ψ*

cos (ϕ*l* )−(µ*l* ) )

H

*l*

(*I*N

)−1

— *Φ*Do)

∂*aψ* (ν*ψ*,Ndo)

*ψ*

*,* (44)

*l* 2 |*γ*

µ*l*

do,l|

2

. .

k*l*=1

(ΣNde .

n=1

*l*

.2

∂µ*τ*

de,l

∂ν*l*

H

*l*

*l*

do ∂ν*l*

)−1

*l*

*l*

∂µ*τ*

Nde

De

) *l*

*σ*n

CRLB *τ* =

2 |*γ*de,l|

2

ℜ *s*¯[n]

∂*aτ* (µ*τ* ,K*l*) (*I* − *Φ*

∂*aτ* (µ*τ* ,K*l*)

*.* (45)

can be also acquired in a similar way, where the detailed derivations are omitted due to space constraints.

*Remark 2:* According to (40), if the system configuration

parameters of the transceiver are the same except for different sparse spacing *Ω*, the CRLB of *ξ*BS, i.e., CRLB*ξ*BS , is a

O 8*LN* 2 *K*l +8*LK*2*N*De . The second part is the data-aided channel tracking, and its computational complexity consists of the reestablishment of initial beam-aligned effective channel vectors and the tracking of subsequent effective channel vec-

tors, i.e., O (*L*(*N*AC+*N*BS+3*K*)) and O (*LK*), respectively.

Do

l

l *l*

constant. Therefore, we can observe from (42) and (43) that

CRLBθBS (1) and CRLBϕBS (1) for *Ω* = 1 are the *Ω*2 times

It can be seen from the above analysis that although the THz

UM-MIMO arrays employing tens of thousands of antennas

*l l*

as much as CRLBθBS (*Ω*) and CRLBϕBS (*Ω*) for *Ω >* 1,

are equipped at BSs and aircraft, the computational complexity

*l l*

respectively. In other words, compared with the array with

critical antenna spacing, the CRLB of sparse array with sparse spacing *Ω>* 1 can achieve the about 20 lg *Ω* dB Mean Square Error (MSE) performance gain, which theoretically testifies the improved accuracy of angle estimation using sparse array.

* 1. *CRLBs of Doppler Shift and Path Delay Estimation:*

Similar to the CRLB derivations of angle estimation, according

of the proposed solution is in polynomial time, since the

effective low-dimensional signals at the receiver are utilized to estimate and track the aeronautical THz UM-MIMO channels. The state-of-the-art Digital Signal Processing (DSP) hardware devices, such as the latest Field Programmable Gate Array (FPGA), are capable of the operations with the order of trillions of Floating-Point Operations Per Second (FLOPS),

to (25) and (29), the CRLBs of virtual Doppler *ν*ψ

l

and

which can be used for the proposed solution in THz UM-

virtual delay *µ*τ

l

can be obtained directly as (44) and (45),

MIMO-based aeronautical communications with the accept-

respectively, on the top of this page. In (44) and (45), the projection operators *Φ*Do and *Φ*De have the similar form to *Φ*BS. By exploiting the transformation of parameter CRLB [55], the CRLBs of Doppler shift *ψ*z,l and the normalized

CRLB *ψ*

able processing time.

1. NUMERICAL EVALUATION

delay *τ*¯ = *f τ*

can be then expressed as CRLB =

*νl A. Simulation Setup*

l s l

K2CRLB*µτ*

ψ*z,l*

(2πTsym)2

In this section, we evaluate the performance of the proposed

and CRLBτ¯*l* =

(2π)2

*l* , respectively. Finally, the CRLBs

channel estimation and tracking scheme for THz UM-MIMO-

of Doppler shift and the normalized delay for *L* BSs can be

*z*

L

l=1

acquired as CRLBψ

*z,l*

generality, we set the reference altitudes of *L* = 2 suspended

aerial BSs and an aircraft in Fig. 10(a) are 20 kilometer (km)

L

l=1

τ¯*l*

= 1 ΣL

CRLBψ

and CRLBτ¯ =

based aeronautical communications, where the simulation sce-

1 ΣL

CRLB

, respectively.

nario considered can be shown in Fig. 10. Without loss of

1. *Computational Complexity*

The computational complexity of the proposed chan- nel estimation and tracking scheme mainly consists of two portions. The first one is to estimate and track the channel parameters, including the acquisition of az- imuth/elevation angles at BSs and aircraft, Doppler shifts, and path delays using TDU-ESPRIT and TLS-ESPRIT al- gorithms. Since a mass of trivial computations with small computational complexity can be ignored, we focus on the dominant calculation steps involving numerous com- plex multiplications. For the estimation and tracking of an- gles at BSs and aircraft, their total computational complex- ity is O (2*LI*BS*K*l +2*LI*AC*K*l +2*LI*B′ S*K*l+2*LI*A′ C*K*l), where

O(*N* ) stands for “on the order of *N* ”. The computational

complexity of Doppler shift and path delay estimation is

and *D*AC = 10 km (at the top of the troposphere or the bottom of the stratosphere), respectively, and thus, the vertical distance between the aircraft and BSs is *D*AB = 10 km. The distance be- tween two BSs is *D*BS = 200 km. In addition, we can abstract a spatial coordinate system as Fig. 10(b) from this real scenario, where point *O* is the origin of coordinates, and the coordinates of points *A*, *B*, and *C* are (0*,*0*,D*AB), (0*,D*BS*,D*AB), and (*D*BS*/*2*,D*BS*/*2*,* 0), respectively. The position coordinate of the aircraft randomly appears in a horizontal circular plane with *C* as the center and *R*a = 50 km as the radius, and the horizontal direction of aircraft *v*d with flight speed *v*AC = 200 meter per second (*m/s*) falls in the intersection angle OCD. In order to simplify the simulation scenario, we consider that the altitude changes of aerial BSs and aircraft are reflected in the angle change over time.

*z*



*A*

10 NP

*O*

*B*

v 100 NP

*D*

*d*

200 NP

*y*

*C*

*R*D = 50 NP

*D*%6 = 200 NP

*D*$% = 10 NP

6WUDWRVSKHUH $iU

*R*D = 50 NP

7URSRVSKHUH\*URXQG

*D*$& = 10 NP *x*

(D) (E)

Fig. 10. (a) Schematic diagram of simulation scenario, and (b) the corresponding spatial coordinate representation.

In simulations, the central carrier frequency is *f*z = 0*.*1 THz with system bandwidth *f*s = 1 GHz, the horizontal/vertical antenna numbers of all subarrays at BSs and aircraft are

*N*

h BS

= *N*

v BS

= *M*

h AC

= *M*

v AC

= 200, and the horizontal and

which is extremely small, so that the assumption about TI is reasonable. For the data-aided channel tracking, *ε* = 0*.*2

and *K*˜ = *K/*2. Note that the relationship between transmit

power *P*l and large-scale fading gain *G*l is complementary.

vertical numbers of subarrays at aircraft are h = 1 and

˜*I*

AC

Without loss of generality, assume that *P*l*G*l = 1 through the

*I*˜v

AC

equivalent fully-digital (sparse) array are *I*BS = *I*BS = *I*AC =

*I*

= 2, respectively, while the dimensions of the selected

h

v

h

= *I*

= *I*

= *I*

v

transmit power compensation. Therefore, to facilitate the sim-

ulation evaluation, we define *σ*2 */σ*2 with *σ*2 being the noise

v = 5 (*I*′h

BS

AC

′ v ′h

BS AC

′

AC = 5). The numbers of

variance as the transmitted SNR of UL and DL throughout

antennas in each antenna group used for the GTTDU modules

α

n

n

our simulations.

at BSs and aircraft are *M*˜h = *M*˜v = *M*˜h = *M*˜v = 5.

BS

BS

AC

AC

Moreover, the number of OFDM symbols used to estimate and track the Doppler shifts and path delays are *N*do = 6 and *N*de = 10, respectively. The number of subcarriers is set to *K* = 2048 with the length of Cyclic Prefix (CP) being

*N*cp = 128, and perfect frame synchronization and reliable

delay compensation are assumed. The channel parameters

*B. Simulation Results*

First the performance of the initial channel estimation is evaluated using the Root-MSE (RMSE) metric given by RMSE = E 1 ǁ*x*−*x*^ǁ2 , where *x* ∈ RL and *x*^ represent

*x*

L

2

q

the true and the estimated channel parameter vectors, and [*x*]l

comes from the parameters *θ*BS, *ϕ*BS, *θ*AC, *ϕ*AC, *ψ*z,l, or *τ*l.

are listed as follows. The azimuth and elevation angles at

BSs and aircraft {*θ*BS*,ϕ*BS*,θ*AC*,ϕ*AC}L are generated from

l l l l

For the angle estimation at the BSs and aircraft, the state-

l l l

l l=1

of-the-art channel estimation and tracking schemes [21]–[26],

[−*π/*3*, π/*3] randomly. Note that due to the long distance

between the adjacent BSs, *θ*AC*,ϕ*AC L corresponding to different BSs have the large gaps, and these angles can be set based on the position of aircraft in Fig. 10(b). The Doppler shifts {*ψ*z,l}L can be set based on *v*d and the relationship

l=1

l l l=1

{ }

[34] are not suitable for the THz UM-MIMO based aeronau-

tical communication channels with fast time-varying fading characteristics. Hence, we consider the beam sweeping method with severe beam squint effect in IEEE standards 802.11ad

[56] as one of the benchmarks, where its sweeping ranges are

between spatial coordinates of the BSs and aircraft. The path

delay *τ*l follows uniform distribution U[0*, N*cp*T*s] and each of channel gains *α*l is generated according to CN(0*,* 1), i.e.,

*σ*2 = 1, for 1 ≤ *l* ≤ *L*. The rough estimates of azimuth/elevation

5◦ around the corresponding rough angle estimates acquired by BSs and aircraft.

Fig. 11 compares the RMSE performance of the pro-

±

α

˜BS

BS ˜AC

AC L

posed fine angle estimation for {*θ*BS*, ϕ*BS}L

at the initial

angles at BSs and aircraft {*θ*l *, ϕ*˜l *, θ*l *, ϕ*˜l }l=1 can be

l l l=1

randomly selected from the range of these true angles with offset ± 5◦, while the rough Doppler shift estimate *ψ*z,l can be randomly selected from the range of the true *ψ*z,l with offset 0*.*01*ψ*z,l for 1 *l L*. Furthermore, to describe the fast time-varying fading channels, we define the relationship of these channel parameters between the *q*th and (*q* + 1)th TIs as *x*[q+1] = *x*[q] + *s*pm*ρ*x*N*C*T*sym, where *x* represents

˜

± ≤ ≤

AC AC

channel estimation stage, where different processing methods

are investigated. In Fig. 11, the labels “no TTDU module” and “ideal TTDU module” indicate the transceiver adopting ideal TTDU module and without considering TTDU mod- ule, respectively. The label “conventional scheme” indicates directly applying the conventional TDU-ESPRIT algorithm to estimate angles as those used in existing mmWave systems

[37], while *i*max = 1 and *i*max = 2 indicate the maximum

the channel parameter coming from *α*l, *τ*l, *ψ*z,l, *θ*l , *ϕ*l ,

BS BS

*θ*BS, or *ϕ*BS. Here, *s*pm denotes a binary variable selected

iterations in the proposed Algorithm 1. From Fig. 11, it

l l can be seen that the RMSE curves of “proposed algorithm

−

from 1 or 1 randomly, *N*C = 70, *T*sym = (*N*cp + *K*)*T*s =

BS

2*.*176 Microseconds (*µ*s), and the duration time of one TI is

*T*TI = *N*C*T*sym = 152*.*32 *µ*s, while *ρ*x is the rate of change

associated with *x*. We consider *ρ*α = *α*(1)*/*2, *ρ*τ = *τ* (1)*/*2,

1 with *i*max = 2” and “conventional scheme” using “ideal TTDU module” almost overlap, and they are very close to the CRLBs of azimuth and elevation angles at high SNR.

(1)

*l* l *l* l

The proposed Algorithm 1 just needs *i*max = 2 iterations to

*ρ*ψ = 0*.*01*ψ*

, *ρ*AC = *ρ*AC = *π/*4, and *ρ*BS = *ρ*BS = *π/*12. BS

*z,l*

z,l θ ϕ

θ ϕ achieve the performance upper-bound that uses ideal TTDU

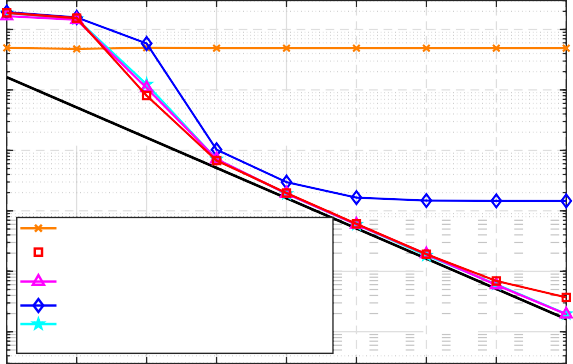
Note that the maximum value of angle changing during one

TI can be approximately calculated as π × *T*TI ≈ 0*.*0069◦,

4

module without beam squint effect. If the beam squint effect is not well handled as “conventional scheme” with “no TTDU

100 100



Beam sweeping method [56]

Proposed Algorithm 1, *i*max = 1

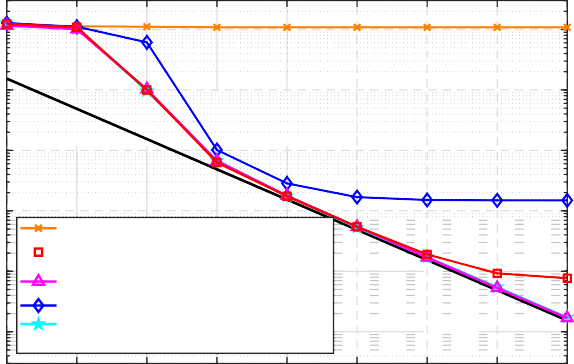
BS

BS

Conventional scheme, no TTDU module

Conventional scheme, ideal TTDU module CRLB

Proposed Algorithm 1, *i*max = 2



Beam sweeping method [56]

Proposed Algorithm 1, *i*max = 1

BS

BS

Conventional scheme, no TTDU module

Conventional scheme, ideal TTDU module CRLB

Proposed Algorithm 1, *i*max = 2

10-1 10-1

10-2 10-2

RMSE

RMSE

10-3 10-3

10-4 10-4

10-5 10-5

-50 -40 -30 -20 -10 0 10 20 30

SNR [dB]

1. azimuth angle BS

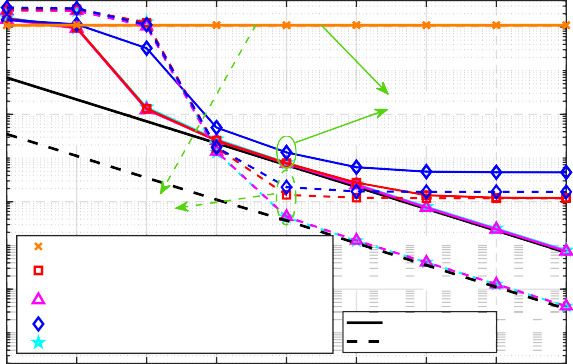
-50 -40 -30 -20 -10 0 10 20 30

SNR [dB]

1. elevation angle BS

Fig. 11. RMSE comparison of {*θ*BS*, ϕ*BS} at the initial angle estimation stage: (a) azimuth angle *θ*BS; and (b) elevation angle *ϕ*BS.

100 100



Method 1

Method 2

Beam sweeping method [56]

Proposed Algorithm 1, *i*max = 1

AC

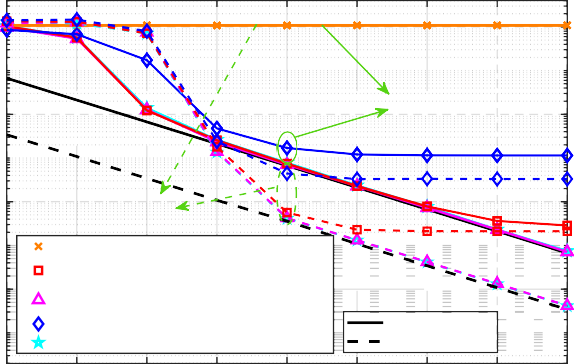
Proposed Algorithm 1, *i*max = 2

AC

Conventional scheme, no TTDU module

CRLB, method 1

Conventional scheme, ideal TTDU module CRLB, method 2



Method 1

Method 2

Beam sweeping method [56]

Proposed Algorithm 1, *i*max = 1

AC

Proposed Algorithm 1, *i*max = 2

AC

Conventional scheme, no TTDU module

CRLB, method 1

Conventional scheme, ideal TTDU module CRLB, method 2

10-2 10-2

10-4 10-4

RMSE

RMSE

10-6 10-6

-50 -40 -30 -20 -10 0 10 20 30

SNR [dB]

* 1. azimuth angle AC

-50 -40 -30 -20 -10 0 10 20 30

SNR [dB]

* 1. elevation angle AC

Fig. 12. RMSE comparison of {*θ*AC*, ϕ*AC} at the initial angle estimation stage: (a) azimuth angle *θ*AC; and (b) elevation angle *ϕ*AC.

module”, its performance of angle estimation will suffer from the obvious RMSE floor at medium-to-high SNR. Note that the angle estimation performance of beam sweeping method is very poor due to the limited training overhead in the fast time-varying channels. Moreover, due to the inaccurately

AC

aircraft9. From Fig. 12, similar conclusions to those observed for Fig. 11 can be obtained. Moreover, it can be observed that the “Method 2” can obtain more accurate angle estimation than that of “Method 1” when SNR is larger than 20 dB. For the curves labeled as “proposed algorithm 1 with *i*max = 1”,

rough angle estimates acquired, “proposed algorithm 1 with

−

“proposed algorithm 1 with *i*max

= 2” and “CRLB”, the

*i*max = 1” only using GTTDU module for compensation at transceiver still suffers from the RMSE floor at high SNR, while “proposed algorithm 1 with *i*max = 2” can further attenuate this beam squint error by finely compensating the received signal matrix *Y*UL,l with the compensation matrix

AC

BS

BS

(1)

*Y*˜UL,l.

Fig. 12 investigates the RMSE performance of the pro- posed fine angle estimation for {*θ*AC*, ϕ*AC}L at the initial

l l l=1

improvement of RMSE performance are more than 12 dB when SNR 10 dB. This is because “Method 2” employs more accurate angles estimated at BSs in high SNR region to obtain the larger beam alignment gain than “Method 1”.

Fig. 13 compares the RMSE performance of the proposed fine Doppler estimation for *ψ*z,l L at the initial channel estimation stage with different processing methods, where the angles at BSs and aircraft are estimated at the fixed SNR = 20 dB. Note that the label “no Doppler squint” denotes the channel model without Doppler squint effect, and the label

l=1

{ }

−

≥ −

“proposed algorithm 2 with *i*max = 0” indicates that the TLS-

channel estimation stage. The accurate angle estimation of do

{*θ*AC*, ϕ*AC}L

l

l

l=1

relies on the fine estimates of {*θ*BS*, ϕ*BS}L

ESPRIT algorithm is applied directly to *Y*do,l for obtaining the

the THz UM-MIMO array can provide a large beam alignment

(0)

in Fig. 11. To investigate the impact of the estimated

l

l

l=1

{*θ*BS*, ϕ*BS}L

on the estimation of {*θ*AC*, ϕ*AC}L

, we

estimate *ψ*^z,l in Algorithm 2. From Fig. 13, we observe that

consider “Method 1” and “Method 2”. “Method 1” adopts

l

l

l=1

l

l

l=1

{*θ*BS*, ϕ*BS}L

estimated at BSs for the fixed SNR=−20 dB,

9It’s worth noting that to ensure the rationality of CRLB at low SNRs

l l l=1

for “Method 2”, the rough angle estimates {*θ*˜BS*, ϕ*˜BS}*L* rather than the

while “Method 2” adopts the {*θ*BS*, ϕ*BS}L estimated at BSs

*l*

l

l

l=1

estimated angle {*θl , ϕ*^

}*l*=1 are considered as the beam-aligned angles

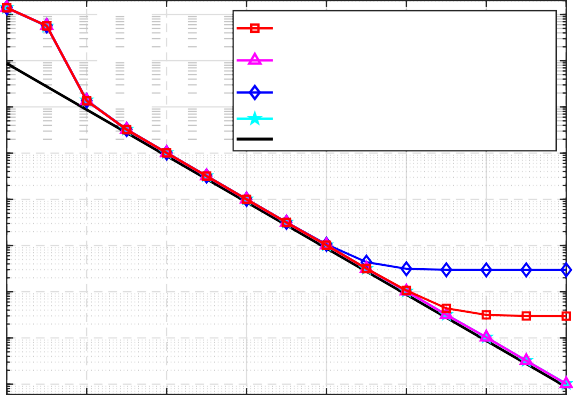
^BS BS *L*

*l l l*=1

for the same SNRs with those of the angle estimation at

at BSs when SNR≤−20 dB.

105



Proposed Algorithm 2, *i*max = 1

do

Proposed Algorithm 2, *i*max = 2

do

Conventional scheme, *i*max = 0

do

Conventional scheme, no Doppler squint CRLB

104

103

102

RMSE

101

100

10-1

10-2

10-3

-120

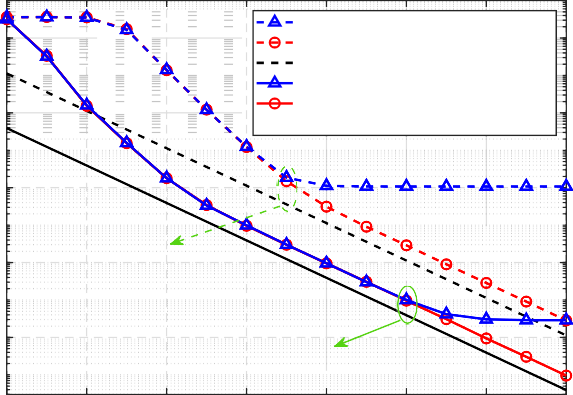
-100 -80

-60

-40 -20 0 20

102

100



Proposed scheme, triple squint Conventional scheme, no triple squint CRLB, SNR = -20 dB

Proposed scheme, triple squint

Conventional scheme, no triple squint CRLB, SNR = 20 dB

SNR = -20 dB

SNR = 20 dB

10-2

RMSE

10-4

10-6

-120 -100 -80 -60

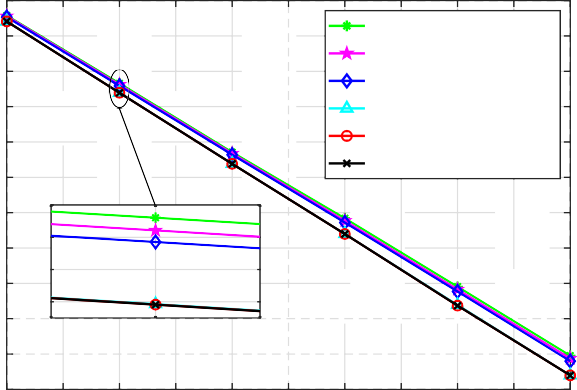
-40 -20 0 20

SNR [dB]

Fig. 13. RMSE comparison of Doppler shift *ψz* estimation.

SNR [dB]

Fig. 14. RMSE comparison of the normalized delay *τ*¯ estimation.

-5

-10

-15

-20

-25

NMSE [dB]

-30

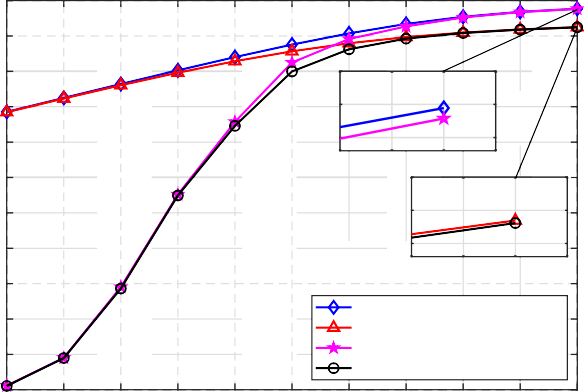
-35

-40

-16.5

-17

55

*fs* = 5 GHz, triple squint 50

Average Spectral Efficiency [bit/s/Hz]

*fs* = 3 GHz, triple squint

*fs* = 1 GHz, triple squint 45

*fs* = 5 GHz, no triple squint 40

*fs* = 3 GHz, no triple squint 35

*fs* = 1 GHz, no triple squint

30

25

20

53.9

53.85

53.8

-5.2 -5.1 -5

51.3

51.25

51.2

-4.9

-45

-50

-17.5

-18

-20.1 -20 -19.9

-5.2 -5.1 -5

15

Perfect CSI, no triple squint

10 Perfect CSI, triple squint

-4.9

-55

-60

-30 -25 -20 -15 -10 -5 0 5 10 15 20

SNR [dB]

Fig. 15. NMSE comparison with different bandwidths.

5

0

-35 -32 -29 -26 -23 -20

SNR [dB]

Fig. 16. ASE comparison with different CSI.

Estimated CSI, no triple squint Estimated CSI, triple squint

-17 -14 -11 -8 -5

gain and greatly improve the receive SNR for Doppler shift estimation, so that the RMSE curves are close to CRLB at

very low SNR, even SNR=−100 dB. Additionally, “proposed

SNR = −20 dB. Note that the errors of the previously es- timated angles {*θ*^BS*, ϕ*^BS*, θ*^AC*, ϕ*^AC}L and Doppler shifts

l=1

impact on the estimation of {*τ*¯l}L

, which leads to

algorithm 2 with *i*max = 0” and “proposed algorithm 2 with

do

*i*max = 1” will encounter the RMSE floors at high SNR, while the curve labeled as “proposed algorithm 2 with *i*max = 2” almost overlap with “conventional scheme” with “no beam squint” when SNR *>* −100 dB.

do

do

the RMSE floors of the normalized delay estimation at high

SNR.

l=1

l

l

l

l

l=1

{*ψ*^z,l}L

According to the estimated channel parameters, the Normalized-MSE (NMSE) metric [37] for the initial channel estimation can be expressed as (46) on the bottom of this

frequency channel matrix at the *k*th subcarrier of the 2nd

[2] [2]

Fig. 14 compares the RMSE performance of the proposed

path delay estimation for the normalized {*τ*¯l}L

at the initial

page. In (46), *H*DL,l[*k*] and *H*^DL,l[*k*] denote the DL spatial-

channel estimation stage, where the angles and Doppler shifts

l=1

are estimated at fixed SNR = 20 dB and SNR = 20 dB, respectively. Note that the labels “triple squint” and “no triple squint” indicate the channel model considering and not considering the practical triple squint effects, respectively. Clearly, when the triple squint effects are considered, the higher angle and Doppler estimation accuracy at SNR= 20 dB will attenuate the impact of triple squint effects to acquire more accurate path delay estimation than that estimated at

−

DL,l

OFDM symbol (considering the impact of Doppler shifts) in

(4) and the reestablished channel matrix based on the estimated channel parameters, respectively. Fig. 15 compares the NMSE performance at the initial channel estimation stage for different system bandwidths *f*s = 1*,* 3*,* 5 GHz. From Fig. 15, we can observe that the channel estimation performance of the proposed solution under triple squint effects is very close to that of the proposed solution without triple squint effects, where the NMSE performance gap between them is about 1 dB

{ }

[2]

*L*

l=1

k=1 ¨

DL,l

¨F

k=1 ¨

DL,l

¨F

NMSE

*H*

[2]

DL

= E 1 ΣL

ΣK

¨*H*[2]

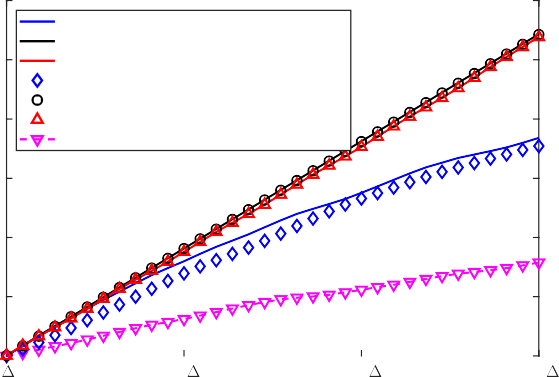
[*k*] − *H*^

[*k*]¨2 , ΣK

¨*H*[2]

[*k*]¨2 *.* (46)

60 300



Perfect CSI, no TTDU module Perfect CSI, ideal TTDU module

Perfect CSI, proposed GTTDU module Estimated CSI, no TTDU module Estimated CSI, ideal TTDU module Estimated CSI, proposed GTTDU module Estimated CSI, beam sweeping method [56]

*f*

684 *f*

1366 *f*

2048

50 250

40 200

Beam sweeping method [56] Proposed scheme, GTTDU module

Conventional scheme, no TTDU module Conventional scheme, ideal TTDU module

Throughput [Gbps]

Throughput [Gbps]

30 150

*fs* = 1 GHz

*fs* = 5 GHz

20 100

10 50

0

2

Bandwidth [Hz] (a)

0

*f* -20 -17 -14 -11 -8 -5 -2 1 4 7 10

SNR [dB]

(b)

Fig. 17. Throughput performance comparison of THz UM-MIMO system adopting different TTDU modules: (a) maximum bandwidth is *fs* = 1 GHz with perfect and the estimated CSI at SNR= 10 dB; and (b) bandwidth *fs* = 1 GHz and 5 GHz with the estimated CSI.

60

Perfect CSI, no TTDU module Perfect CSI, ideal TTDU module

Perfect CSI, proposed GTTDU module Estimated CSI, no TTDU module Estimated CSI, ideal TTDU module Estimated CSI, proposed GTTDU module

50

40

Throughput [Gbps]

30

20

10

Dimension of UPA

Fig. 18. Throughput performance comparison of THz UM-MIMO system adopting different dimensions of UPA at SNR= 10 dB.

at SNR = −20 dB. Furthermore, the results of Fig. 15 show that compared with the system bandwidth *f*s = 1 GHz, the NMSE performance of the proposed solution using the larger bandwidth *f*s = 5 GHz does not deteriorate significantly.

Moreover, we consider the Average Spectral

Efficiency (ASE) performance metric [37], [57] at the data transmission stage, defined as ASE =

2

l

l

Fig. 17 compares the throughput performance of THz UM- MIMO system adopting different TTDU modules, where the transceivers using ideal TTDU module, the proposed GTTDU module, and without TTDU module are considered. Note that ∆*f* denotes the frequency spacing between adjacent subcarriers, typically, ∆*f* ≈ 0*.*488 Megahertz (MHz) for *f*s = 1 GHz and *K* = 2048. In Fig. 17(a), for maximum bandwidth *f*s = 1 GHz, an obvious throughput ceiling can be observed in “beam sweeping method” and conventional scheme with “no TTDU module” as the bandwidth increases, in other words, the severe beam squint effect will restrict the throughput of THz UM-MIMO systems. On the contrary, the throughputs adopting the proposed GTTDU module and ideal TTDU module present a linear growth with the increase of bandwidth. For the estimated CSI at *f*s = 2048∆*f* , the throughput improvements of more than 15 Gigabit per second (Gbps) and 35 Gbps can be acquired by both “ideal TTDU module” and “proposed GTTDU module” compared with the throughput of “no TTDU module” and beam sweeping method in [56], respectively. Furthermore, it can be also observed from Fig. 17(b) that the increase of throughput in the THz UM- MIMO system with severe beam squint effect is extremely limited when the bandwidth is increased to *f*s = 5 GHz.

Fig. 18 compares the throughput performance of THz

ΣL 1 ΣK

k=1

*h*[2][*k*] and *z*[2][*k*] are the beam-aligned effective channel

l=1

K

l

l

log

1 + |*h*[2][*k*]|2*/*|E(*z*[2][*k*])|2 , where

UM-MIMO system adopting different dimensions of UPA

at SNR = 10 dB, where bandwidth *f*s = 1 GHz and the

same transmit power are considered. From Fig. 18, it can be

coefficient and interference plus noise at the *k*th subcarrier

of the 2nd OFDM symbol, respectively. Fig. 16 compares the ASE performance of the proposed solution with different CSI, where the perfect CSI known at both the BSs and aircraft is adopted as the performance upper bound. It can be observed from Fig. 16 that the ASE performance using the estimated CSI almost attains the performance upper bound when SNR 14 dB whether or not the triple squint effects are considered. In addition, since the practicable GTTDU module still has residual beam alignment error caused by beam squint effect, the ASE performance gain achieved by our solution with triple squint effects is 2.5 [bit/s/Hz] lower than the other one at high SNR.

≥ −

observed that the usage of regular UPA with size of 16 16

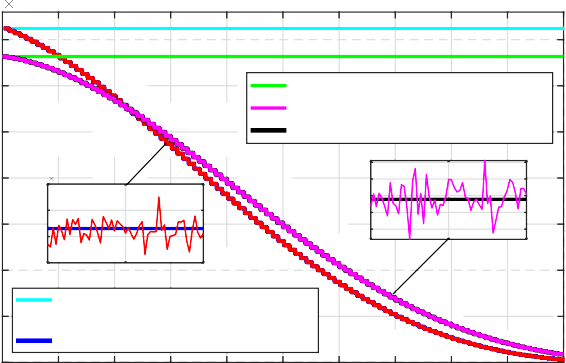
in the ultra-long-distance THz aeronautical communications cannot establish an efficient communication link, which causes the degraded throughput performance. Due to the pencil-like beams and less interference, the system throughput will be improved significantly as the dimension of UPA equipped at the transceiver increases. However, the increase of array dimension leads to more obvious beam squint effect, which inhibits the improvement of throughput performance in turn (observed from the curves labeled as “no TTDU module”). For the transceiver equipped with UM-MIMO array of size

×

256 256, the throughput adopting the proposed GTTDU module is closed to the throughput of “ideal TTDU mod-

×

30



104

Initial channel estimation, BS #2 DADD-based channel tracking, BS #2 Real-time effective channel, BS #2

2.37235 10

4

2.3723

2.37225

7056.2

7056

7055.8

7055.6

7055.4

69.1

69.5

69.9

2.3722

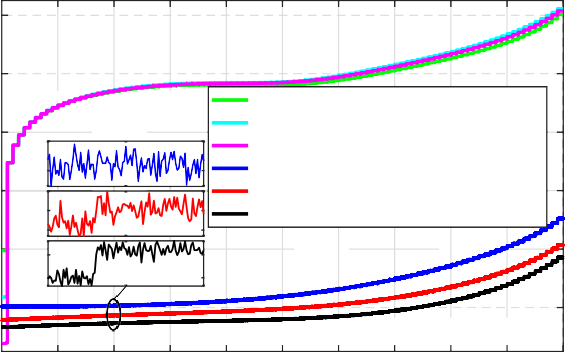
29.1

29.5

29.9

Initial channel estimation, BS #1

DADD-based channel tracking, BS #1 Real-time effective channel, BS #1



Initial channel estimation, SNR = -20 dB Initial channel estimation, SNR = -10 dB Initial channel estimation, SNR = 0 dB DADD-based channel tracking, SNR = -20 dB DADD-based channel tracking, SNR = -10 dB

DADD-based channel tracking, SNR = 0 dB

-75.7

-75.75

19

21

20

-73

-73.1

-73.2

-69.4

-69.6

3.5

Amplitude of Effective Channels

NMSE of Effective Channels [dB]

3

2.5

2

1.5

1

10

-10

-30

-50

0.5 -70

0

10 20 30 40 50 60 70 80 90 100

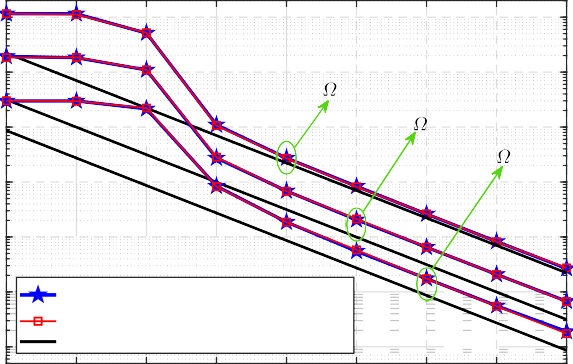
Number of TI (a)

10 20 30 40 50 60 70 80 90 100

Number of TI (b)

Fig. 19. Performance comparison of the proposed DADD-based channel tracking: (a) amplitude of effective channels; and (b) NMSE of effective channels.

100 100



= 1

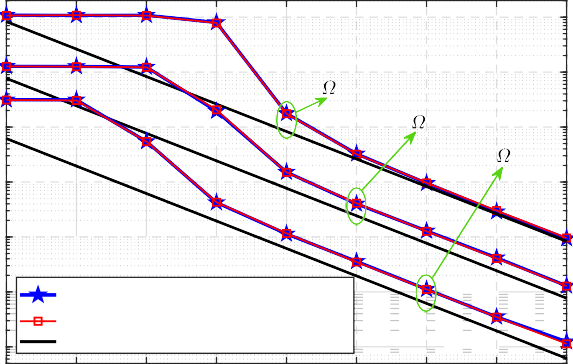
= 4

= 16

Proposed Algorithm 1, *i*max = 2

BS

Conventional scheme, ideal TTDU module CRLB



= 1

= 4

= 16

Proposed Algorithm 1, *i*max = 2

AC

Conventional scheme, ideal TTDU module CRLB

10-1 10-1

10-2 10-2

10-3 10-3

RMSE

RMSE

10-4 10-4

10-5 10-5

10-6 10-6

(a) azimuth angle at BS: BS (b) azimuth angle at aircraft: AC

|  |  |  |  |  |  |  |  |  |  |  |  |  |  |  |
| --- | --- | --- | --- | --- | --- | --- | --- | --- | --- | --- | --- | --- | --- | --- |
| -100 | -90 | -80 | -70 -60 | -50 | -40 | -30 | -20 | -120 | -110 | -100 | -90 -80 -70 | -60 | -50 | -40 |
|  |  |  | SNR [dB] |  |  |  |  |  |  |  | SNR [dB] |  |  |  |

Fig. 20. RMSE performance at pilot-aided angle tracking stage: (a) azimuth angle *θ*BS at BS; and (b) azimuth angle *θ*AC at aircraft.

ule”, and it can achieve throughput improvement more than

55 Gbps compared with that of transceiver using UPA of size

16 16. Therefore, it is necessary to use UM-MIMO array in aeronautical communications to cater for the high data rate requirements of hundreds of users in the cabin.

×

Next, the performance of the proposed DADD-based chan- nel tracking algorithm is evaluated according to the metrics of effective channels’ amplitude and NMSE, where the NMSE of effective channels for the *r*th OFDM symbol is given by NMSE = E 1 L ǁ*h*[r] −*h*[r]ǁ2*/*ǁ*h*[r]ǁ2 . For the data-aided channel tracking scheme, Fig. 19 compares the effective channels’ amplitude performance (at SNR= 20 dB) and NMSE performance (at SNR= 20*,* 10*,* and 0 dB) for the different numbers of TI. Here, the Turbo coding and QPSK modulation are considered during the data transmission. From Fig. 19(a), we can observe that the amplitude of effective channels decreases rapidly as time goes by, where the proposed DADD-based channel tracking method can track the amplitude changes of true effective channels in real-time. This decreasing amplitudes mean that the gains of beam alignment becomes small. Also observe in Fig. 19(b) that the NMSE performance of the proposed DADD-based channel tracking method slowly worsens as the number of TI increases, while the NMSE of the initial channel estimation without tracking will deteriorate

— −

−

*h*[*r*]

L

l=1

l

l

2

l

2

Σ ^

rapidly after several TIs.

Fig. 20 investigates the RMSE performance of the proposed pilot-aided angle tracking scheme against different sparse spacing *Ω* = 1, *Ω* = 4, and *Ω* = 16. Here the angle tracking at aircraft adopts the angles *θ*BS*, ϕ*BS L estimated at BSs using the fixed SNR = 60 dB. Note that the RMSE curves of the elevation angles *ϕ*BS and *ϕ*AC are omitted due to the similar performance to the azimuth angles. From Fig. 20, it can be observed that the usage of sparse array can significantly improve the accuracy of angle estimation, and these results testify that the improved RMSE performance is consistent with the conclusion in *Remark* 2, i.e., the proposed solution using the sparse array with sparse spacing *Ω* can achieve about 20 lg *Ω* dB performance gain.

−

l

l

l=1

{^ ^ }

1. CONCLUSIONS

We have proposed an effective channel estimation and tracking scheme for THz UM-MIMO-based aeronautical com- munications in SAGIN, which can solve the unique triple delay-beam-Doppler squint effects not considered in the sub- 6 GHz or mmWave systems. The proposed solution includes the initial channel estimation, data-aided channel tracking, and pilot-aided channel tracking. Specifically, based on the rough angle estimates acquired from navigation information,

the initial THz UM-MIMO link can be established, where the delay-beam squint effects at transceiver can be significantly mitigated by employing the proposed GTTDU module. By exploiting the proposed prior-aided iterative angle estimation algorithm, the fine azimuth/elevation angles can be estimated based on the equivalent low-dimensional fully-digital array. These estimated angles can be used not only to achieve a highly accurate beam alignment, but also to refine the GTTDU module at the transceiver for further eliminating the delay- beam squint effects. The Doppler shifts can be subsequently estimated using the proposed prior-aided iterative Doppler shift estimation algorithm. On this basis, path delays and channel gains can be estimated accurately, where Doppler squint effect can be attenuated vastly via fine compensation process. At the data transmission stage, a DADD-based chan- nel tracking algorithm is developed to track the beam-aligned effective channels. When the data-aided channel tracking is

invalid, the pilot-aided channel tracking is proposed to re-

research directions in the THz UM-MIMO-based aeronau- tical communications include more specific and universal THz UM-MIMO channel modeling under LoS path scenario [9], long-distance air-ground communication scheme design, low-complexity signal transmission and tracking methods for the large bandwidth and high dynamic environment, THz transceiver design using more practical hardware components (e.g., TTDU module, high-frequency switch [58], and low- energy antenna array [2]), advanced DSP module design sup- porting ultra-high data rate with the order of Tbps, modulation and coding design at the physical layer [10], as well as the deployment and power optimization of aerial BSs at the network and transport layer.

APPENDIX A

DERIVATION OF DL CHANNEL MATRIX *H*[n] [*k*]

DL,l

By taking the Fourier transform of (3) with respect to *τ* ,

DL,l

estimate the angles at transceiver using an equivalent fully- digital sparse array, where the angle ambiguity issue derived

¯

c

nAC

,nBS

l

l

*l*

*c l*

*c l*

the frequency response of [*H*¯ (t)

(*τ* )]nAC

,nBS

is given by

[*H*(t)

from sparse array can be addressed based on the previously

DL,l

(*f* )]

=√*G α e*j2πψ t*e*−j2πf τ *e*−j2πf τ[*n*AC]

*c l*

estimated angles. Finally, the CRLBs of dominant channel

parameters and the simulation results evaluate the effectiveness

of the proposed solution for THz UM-MIMO-based aeronau- tical communications.

It is worth mentioning that the proposed solution in this paper still has some improvements in the following aspects. First, the proposed Rotman lens-based GTTDU module of transceiver in Fig. 6 can be further researched. Second, the signal frame structure (e.g., the length of OFDM symbols, CP length) in THz communications can be also optimized based on the parameter configurations of specific scenarios. Third, some new data-aided channel tracking methods with the lower

× *e*−j2πf τ[*n*BS] *,* (47)

where the large-scale fading gain *G*l can be modeled as *G*l = *λ*2*/*(4*πD*l)2 based on the free-space path loss of Friis’ formula with *D*l being the communication distance between the aircraft and the *l*th BS. Considering the large system bandwidth *f*s, the carrier frequency can be expressed as *f*c = *f*z+*f* , where *f* denotes the baseband frequency satisfying

c

−*f*s*/*2 ≤ *f* ≤ *f*s*/*2 and the wavelength corresponding to the central carrier frequency *f*z is *λ*z. After the down-conversion and focussing on the baseband frequency, we can obtain the (*n*AC*, n*BS)th element of the DL baseband channel matrix

*H*

computational complexity can be considered in Section IV,

(t)

DL,l

(*f* ) in the spatial-frequency domain [29], [34], [59], i.e.,

such as uniformly-spaced pilot interpolation in the frequency

domain. Fourth, according to the specific communication sce- narios, the transmit power at the transceiver can be also further optimized to improve the spectrum efficiency of systems and reduce the bit error rate.

For future work, our proposed THz UM-MIMO-based aeronautical communication solution can be also suitable for the long distance communications or backhaul in SAGIN such as the space information network consisting of air- crafts/UAVs, aerial BSs, and LEO/MEO/GEO satellites, or

1. on the bottom of this page.

Due to the large bandwidth in THz UM-MIMO, the car- rier frequencies and wavelengths at different subcarriers are different, so the frequency-dependent Doppler shift at the *k*th subcarrier is given by *ψ*l,k = *ψ*z,l + v*l* ( k−1 − 1 )*f*s with *ψ*z,l = *v*l*/λ*z. Let the antenna spacing *d* = *λ*z*/*2, the baseband frequency response in (48) can be further expressed as the

c

K

2

spatial-frequency channel coefficient at the *k*th subcarrier, i.e.,

1. on the bottom of this page. In (49), *a*AC(*µ*AC*, ν*AC*, k*) ∈

l

l

CNAC and *a*BS(*µ*BS*, ν*BS*, k*) ∈ CNBS are the array response

l

l

the air-ground communication links between the high-altitude vectors associated with the *k*th subcarrier at aircraft and

terrestrial stations and the LEO satellite systems. Potential the *l*th BS, respectively, and *a*AC(*µ*AC*, ν*AC*, k*) and

(t)

[*H*

√ j2πψ t

−j2πfτ

j 2d ((nh

−1)µ +(n

AC v

*l*

AC

−1)νAC)

−j ((n

*e*

λ*c*

l

−1)µ +(n

2d h

BS v

BS

*l*

BS

l

−1)νBS)

*l*

nAC

(t)

DL,l

(*f* )]nAC

,nBS

=

*G*l*α*l*e*

*l e*

*l e* λ*c*

AC

*l*

*.* (48)

[*H*

[*k*]]n

=

*G*l*α*l*e*j2πψ*l,k*t*e*

DL,l

BS

(*µ*AC*, ν*AC*, k*)

√ −j2π k−1 − 1 f*s*τ*l*

K

2

*a*AC(*µ*AC*, ν*AC*, k*)

*a*∗

(*µ*BS*, ν*BS*, k*)

*.* (49)

AC

,n

l

l

nAC

BS

l

l

nBS

AC v

AC

h AC v AC j k−1 − 1 f*s* ((nh

−1)µ +(n

−1)ν )

AC l l

*a*

AC

*l*

AC

*l*

= *e*j((nBS−1)µ*l* +(nBS−1)ν*l* )*e*

K

2

f*z*

BS

*l*

*.* (51)

*a*

BS

l

nAC

K

2

f*z*

(*µ*BS*, ν*BS*, k*)

l

nBS

h BS v BS j k−1 − 1 f*s* ((nh

−1)µ +(n

BS v

= *e*j((nAC−1)µ*l* +(nAC−1)ν*l* )*e*

*l*

BS

−1)νBS)

*,* (50)

*A*DL,l[*k*] = *a*AC(*µ*AC*, ν*AC*, k*)*a*H (*µ*BS*, ν*BS*, k*)

l l BS l l

= *a*

=

(*µ*AC*, ν*AC) ◦ *a*¯ (*µ*AC*, ν*AC*, k*) *a*

(*µ*BS*, ν*BS) ◦ *a*¯

(*µ*BS*, ν*BS*, k*) H

*.* (53)

AC l l

1. AC AC

*a*AC(*µ*l *, ν*l )*a*BS(*µ*l *, ν*l )

◦

*a*¯AC(*µ*l *, ν*l *, k*)*a*¯BS(*µ*l *, ν*l *, k*)

l

AC

AC l

H BS

l

BS

l

BS l l

AC AC

BS l l

H BS BS

*τ*˜[nAC] = (*n*h

— 1)*d* sin(*θ*˜AC) cos(*ϕ*˜AC) + (*n*v

— 1)*d* sin(*ϕ*˜AC) */c,* (55)

*τ*˜[nBS ] = − (*n*h − 1)*d* sin(*θ*˜BS) cos(*ϕ*˜BS) + (*n*v

l

AC

l

(t)

DL,l

AC

BS

AC

l

l

DL,l

AC

— 1)*d* sin(*ϕ*˜BS) */c.* (56)

BS

l

l

[*H*˜

l BS l

[*k*]]n ,n = *a*¯∗ (*µ*˜AC*, ν*˜AC*, k*)

nAC

l

l

l

[*H*(t)

[*k*]]n

l

BS l

,n *a*¯BS(*µ*˜BS*, ν*˜BS*, k*) *.* (58)

nBS

*A*˜DL,l[*k*] = *a*AC(*µ*AC*, ν*AC) ◦ *a*¯AC(*µ*AC*, ν*AC*, k*) ◦ *a*¯∗ (*µ*˜AC*, ν*˜AC*, k*)

l

AC

l

l

BS

l

l

BS

l

l

BS

l

l

× *a* (*µ*BS*, ν*BS) ◦ *a*¯ (*µ*BS*, ν*BS*, k*) ◦ *a*¯∗ (*µ*˜BS*, ν*˜BS*, k*) H

l

l

BS

l

l

= *A*DL,l[*k*] ◦ *a*¯ C(*µ*˜AC*, ν*˜AC*, k*)*a*¯T (*µ*˜BS*, ν*˜BS*, k*) *.* (60)

*a*BS(*µ*BS*, ν*BS*, k*) can be expressed as (50) and (51), respectively, on the bottom of the previous page.

l

l

nBS

Taking all *N*AC and *N*BS antennas of THz UM-MIMO ar- rays at aircraft and the *l*th BS into consideration, the complete DL spatial-frequency channel matrix at the *k*th subcarrier of the *n*th OFDM symbol, i.e., *H*[n] [*k*] in (2), can be then formulated as

DL,l

The spatial-frequency channel coefficient at the *k*th subcarrier [*H*DL,l[*k*]]nAC,nBS can be then written as (58) on the top of this page. Finally, by collecting all *N*AC and *N*BS antennas

of THz UM-MIMO arrays at aircraft and the *l*th BS, the compensated DL spatial-frequency channel matrix at the *k*th

˜

[n]

˜

(t)

subcarrier of the *n*th OFDM symbol, i.e., *H*DL,l[*k*] in (12), can be formulated as

[n]

*G*l*α*l*e*j2πψ*l,k*(n−1)Tsym *e*

K

2

*H*˜DL,l[*k*] =

−j2π

K − 2

f*s*τ*l*

[n] √

*H*

[*k*] =

DL,l

−j2π k−1 − 1 f*s*τ*l* √

*G*l*α*l*e*

k−1 1

*e*

× *A*DL,l[*k*]*,* (52)

j2πψ*l,k*(n−1)Tsym

where the DL array response matrix *A*DL,l[*k*] CNAC×NBS associated with the array response vectors at aircraft and the *l*th BS is given by (53) on the top of this page. In (53), we have used the identity (*a b*)(*c d*)H = (*ac*H) (*bd*H) [60] in equation (*a*).

∈

◦ ◦ ◦

APPENDIX B PROOF OF LEMMA 1

After compensating the antenna transmission delay via the ideal TTDU module, the compensated (*n*AC*, n*BS)th el- ement of DL spatial-delay domain passband channel matrix

˜

× *A*DL,l[*k*]*,* (59)

where the compensated DL array response matrix *A*DL,l[*k*] is given by (60) on the top of this page.

˜

˜

The proof of Lemma 1 is completed.

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¯ (t)

¯ (t)

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*H*DL,l(*τ* ) in (3), denoted by [*H*DL,l(*τ* )]nAC,nBS , can be ex-

pressed as

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¯ (t)

˜[*H*

DL,l

(*τ* )]n ,n

AC

= *δ*(*τ* − *τ*˜[nAC]) ⊛ [*H*¯ (t)

(*τ* )]n ,n

AC

BS

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[nBS]

⊛ *δ*(*τ* − *τ*˜ )*,* (54)

BS

l

DL,l

l

where ⊛ represent the linear convolution operation, and *τ* [nAC]

˜

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and *τ*

˜

[nBS]

l

are the compensated transmission delays yielded

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l

by TTDUs at aircraft and BSs, respectively, denoted by

(55) and (56), respectively, on the top of this page. Similar to (48), by taking the Fourier transform of (54) and the down-conversion, the baseband frequency domain response of [*H*˜DL,l(*τ* )]nAC,nBS is given by

¯ (t)

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(t)

DL,l

nAC,nBS

˜*l*

−j2πfτ[*n*BS]

DL,l

nAC,nBS

[*H*˜

(*f* )]

= *e*−j2πfτ[*n*AC] [*H*(t)

(*f* )]

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× *e* ˜*l .* (57)

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