Channel Measurement and Characterization with Modified SAGE Algorithm in an Indoor Corridor at 300 GHz

Yuanbo Li, Yiqin Wang, Yi Chen, Ziming Yu, and Chong Han, Member, IEEE

Abstract—The much higher frequencies in the Terahertz (THz) band prevent the effective utilization of channel models dedicated for microwave or millimeter-wave frequency bands. In this paper, a measurement campaign is conducted in an indoor corridor scenario at 306-321 GHz with a frequency-domain Vector Network Analyzer (VNA)-based sounder. To realize high-resolution multipath component (MPC) extraction for the direction-scan measurement campaigns in the THz band, a novel modified space-alternating generalized expectation-maximization (SAGE) algorithm is further proposed. Moreover, critical channel characteristics, including the path loss, shadow fading, K-factor, delay spread, angular spreads, cluster parameters, and cross correlations are calculated and analyzed in the LoS case. Besides, two contrasted measurement campaigns in the NLoS case are conducted, with and without additional reflective foils on walls to serve as effective scatterers. Comparison results indicate that the reflective foils are useful to improve the channel conditions in the NLoS case by nearly 6 dB, which is potential to be utilized as alternative of intelligent reflecting surfaces (IRS) to enhance the coverage ability of THz communications.

Index Terms—Terahertz communications, Channel measurement, Channel characterization, Reflecting surface.

I. INTRODUCTION

During the last few decades, the communication community has witnessed the giant leap of the communication technologies, from the first generation to the fifth generation mobile communication systems (5G). To support the large amount of intelligent devices and applications, such as metaverse, autonomous driving, etc., it is believed that the data rate in the sixth generation mobile communication system (6G) needs to exceed hundreds of gigabits per second and even Terabits per second to undertake the explosively grown data traffic [2]. As a result, the Terahertz (THz) band, ranging from 0.1 THz to 10 THz, is envisioned as a key technology to enable such high data rate [3], [4]. With abundant spectrum resources and ultra-large bandwidth (more than tens of GHz), the use of THz band can address the spectrum scarcity and capacity limitations of current wireless systems.

However, challenges remain to make THz communications come true, among which a fundamental one lies on the channel modeling in the THz band. As we are moving to a new frequency band, the existing channel models dedicated for lower frequency bands, such as 3GPP 38.901 [5], are not suitable in the THz band. Not only the parameter sets are totally different due to the increase of frequencies, but also new propagation phenomena observed in the THz band, such as significant diffuse reflections, larger diffraction loss, etc., require careful characterization and modeling [6]–[8].

The channel modeling in the THz band can be either studied in a theoretical way, such as deriving geometry-based stochastic channel models (GBSM) [9], [10], or based on real measurement campaigns in typical scenarios [11], [12]. As the latter one is more convincing and thus popular, channel measurement campaigns serve as the basis to offer the realistic data and channel characteristics of the THz channel. Some studies have conducted measurement campaigns in the THz band, in both indoor scenarios [11]–[15] and outdoor scenarios [16]–[19], which are thoroughly summarized in [20]. However, with hardware limitations, these studies are either focused on low-THz band (mainly at 140 GHz) or conducted with directional antenna effects in very short distances (shorter than 1 m).

Another research challenge related to channel modeling is data processing after measurement. After the raw data is recorded, such as the channel impulse responses (CIR) or channel transfer functions (CTF), one key problem is how to extract the parameters of multipath components (MPCs), including their time-of-arrival (ToA), directional-of-departure (DoD), directional-of-arrival (DoA), and power. Currently, there are mainly two kinds of methods in the literature, namely i) noise-eliminating methods, i.e., the temporal samples of the CIR are regarded as MPCs after eliminating the noise in a certain way, such as neglecting the samples below a power threshold or using machine learning based methods [11], [21]; ii) high-resolution parameter estimation (HRPE) algorithms, i.e., the parameters of MPCs are estimated based on the measured CIR or CTF, such as space-alternating generalized expectation-maximization (SAGE) algorithm [22] and RIMAX algorithm [23].

Even though the noise-eliminating methods are easy to implement, they lack enough accuracy and fail to decouple the antenna effects in direction-scan measurement campaigns. In contrast, the HRPE algorithms possess high accuracy for estimation of MPC parameters, which however usually require information including the antenna radiation pattern. Furthermore, among different HRPE algorithms, the SAGE algorithm
is most widely-used, which has been developed and extended by many researchers during the last two decades in [24]–[27].

Existing SAGE algorithms show good performance in MIMO measurement campaigns in cmWave and mmWave bands. However, due to hardware limitations, MIMO measurement in the THz band is not achievable, while directional-scan measurement campaigns are preferred to overcome the high path loss, for which existing SAGE algorithms fail due to two-fold reasons. On one hand, this algorithm requires the complex antenna radiation patterns, which are very hard to obtain in practice for the horn antennas used in direction-scan measurement campaigns. On the other hand, this algorithm is very sensitive to phase errors since it involves computation in the complex domain. As a result, more robust and efficient algorithms are needed to realize high-resolution MPC extraction for the direction-scan measurement campaigns in the THz band, which we try to explore in this work.

As reported in [1], we have conducted a measurement campaign in an indoor corridor around 300 GHz and summarized important channel characteristics, including path loss, delay and angular spreads, cluster parameters, etc. Compared to this preliminary version, a novel modified SAGE algorithm is further proposed to accurately estimate the parameters of MPCs, based on which the measurement data is further processed and the channel statistics are characterized and analyzed, including the path loss, shadow fading, K-factor, etc. Besides, extensive measurement campaigns in the NLoS case are conducted. Inspired by the idea in [19], we investigate the ability of reflective foils to enhance the coverage of THz communications in the NLoS case. Distinctive contributions of our work are summarized as follows.

- We conduct measurement campaigns in an indoor corridor scenario at 306–321 GHz band with a frequency-domain Vector Network Analyzer (VNA)-based channel sounder. 21 Rx positions in the LoS case and 3 Rx positions in the NLoS case are measured to sufficiently characterize the propagation in the corridor scenario.
- We propose a novel modified SAGE algorithm that is dedicated for MPC extraction in direction-scan measurement campaigns. The proposed algorithm only requires real-valued antenna radiation pattern and is immune from phase errors.
- Based on the extracted MPCs, we fully characterize the channel statistics in the LoS case, including the path loss, shadow fading, K-factor, delay spread, angular spreads, cross correlations among them, and cluster parameters. The channel characteristics in the THz channel are compared with existing channel models under 100 GHz to observe the uniqueness of the THz band, such as the weaker multi-path effects.
- To evaluate the effectiveness of reflective foils to improve the channel conditions in the NLoS case, comparative measurement campaigns in the NLoS case are conducted. The path loss in the NLoS cases is reduced by nearly 6 dB and diffuse scattering is much stronger with reflective foils, which indicates its ability to enhance the coverage of THz communications.

As reported in [1], we have conducted a measurement campaign in an indoor corridor around 300 GHz and summarized important channel characteristics, including path loss, delay and angular spreads, cluster parameters, etc. Compared to this preliminary version, a novel modified SAGE algorithm is further proposed to accurately estimate the parameters of MPCs, based on which the measurement data is further processed and the channel statistics are characterized and analyzed, including the path loss, shadow fading, K-factor, etc. Besides, extensive measurement campaigns in the NLoS case are conducted. Inspired by the idea in [19], we investigate the ability of reflective foils to enhance the coverage of THz communications in the NLoS case. Distinctive contributions of our work are summarized as follows.

- We conduct measurement campaigns in an indoor corridor scenario at 306–321 GHz band with a frequency-domain Vector Network Analyzer (VNA)-based channel sounder. 21 Rx positions in the LoS case and 3 Rx positions in the NLoS case are measured to sufficiently characterize the propagation in the corridor scenario.
- We propose a novel modified SAGE algorithm that is dedicated for MPC extraction in direction-scan measurement campaigns. The proposed algorithm only requires real-valued antenna radiation pattern and is immune from phase errors.
- Based on the extracted MPCs, we fully characterize the channel statistics in the LoS case, including the path loss, shadow fading, K-factor, delay spread, angular spreads, cross correlations among them, and cluster parameters. The channel characteristics in the THz channel are compared with existing channel models under 100 GHz to observe the uniqueness of the THz band, such as the weaker multi-path effects.
- To evaluate the effectiveness of reflective foils to improve the channel conditions in the NLoS case, comparative measurement campaigns in the NLoS case are conducted. The path loss in the NLoS cases is reduced by nearly 6 dB and diffuse scattering is much stronger with reflective foils, which indicates its ability to enhance the coverage of THz communications.

The remainder of the paper is organized as follows. In Sec. II, the VNA-based sounder system, the measurement set-up and the measurement deployment are explained in detail. Furthermore, the modified SAGE algorithm is derived in Sec. III to extract MPC parameters and extracted MPCs are clustered using clustering algorithms. According to the measurement results, the channel characteristics in the LoS case are analyzed in Sec. IV. Moreover, the ability of reflective foils to enhance the coverage of THz communications in the NLoS case is investigated in Sec. IV-E. Finally, Sec. VI concludes the paper.

II. CHANNEL MEASUREMENT CAMPAIGN

In this section, the measurement campaign in the corridor scenario is described, including the sounder system, the measurement set-up and the measurement deployment.

A. Sounder System

The structure of the sounder system is shown in Fig. 1, consisting of a radio frequency (RF) subsystem and a mechanical subsystem. The RF part controls the transmission of the RF signal to measure the THz channels, while the mechanical part serves for changing positions, heights and steering angles of the transceivers. To obtain the CTF of the THz channel,
TABLE I: Measurement parameters

| Parameter          | Values                        |
|--------------------|-------------------------------|
| Frequency band     | 306-321 GHz                   |
| Bandwidth          | 15 GHz                        |
| Sweeping interval  | 2.5 MHz                       |
| Sweeping points    | 6001                          |
| Maximum delay      | 400 ns                        |
| Maximum path length| 120 m                        |
| Time resolution    | 66.7 ps                       |
| Space resolution   | 2 cm                          |
| Tx height          | 2.5 m                         |
| Rx height          | 2 m                           |
| Antenna gain of Tx| 7 dBi                         |
| Antenna gain of Rx | 25 dBi                        |
| HPBW of Rx antenna| 8°                            |
| Rx azimuth rotation range | [0°, 360°]  |
| Rx azimuth rotation step | 10°                       |
| Rx elevation rotation range | [-20°, 20°]  |
| Rx elevation rotation step | 10°                       |

a 9.6926–14.8418 GHz RF signal is firstly generated by the VNA and sent to the transmitter (Tx) module. The RF signal then passes a 27-times frequency multiplier in the Tx module to produce a THz signal with frequencies within 260–400 GHz. Furthermore, the THz RF signal radiates out of the Tx antenna, travels through the THz channel and finally arrives at the receiver (Rx) side. The RF signals before the channel and after the channel are both mixed with 260.279–400.279 GHz local oscillation (LO) signals, which are generated by operating a 24-times frequency multiplication on a 10.8450–16.6783 GHz signal. Thereafter, two 279 MHz intermediate frequency (IF) signals, namely IF₁ at Tx and IF₂ at Rx, are sent back to VNA. Eventually, the ratio between the frequency responses of IF₁ and IF₂ denotes the channel transfer function of the THz channel, i.e., $H = \text{IF}_2 / \text{IF}_1$.

In order to measure the THz channel under different Tx/Rx deployments, the Tx and Rx modules are mounted on mechanical parts, which include a rotator, a lifter and an electrical cart. As highly directional antennas are used in the measurement campaign to overcome the distance limitation of THz transmission, the transceivers need to scan over the spatial domain to capture MPCs from different directions, which is achieved by mechanically rotating the transceivers in both azimuth plane and elevation plane by using the rotators. Furthermore, a lifter is integrated to adjust the heights of the transceivers. Moreover, an electrical cart is used to carry all the above-mentioned equipment and move the locations of Tx/Rx. Last but not least, a laptop is used to control the rotators as well as to record data from the VNA.

B. Our Measurement Setup

The measurement parameters are summarized in Table I, which are explained in detail as follows. The frequency band for this measurement campaign ranges from 306 GHz to 321 GHz, across a 15 GHz wide band. Furthermore, the sweeping interval is 2.5 MHz, which refers to a 400 ns maximum delay and 120 m maximum path length of MPCs. Moreover, the time resolution of the sounder system is 66.7 ps, which allows to separate MPCs whose path length difference is as small as 2 cm. The heights of Tx and Rx are 2.5 m and 2.0 m, respectively. Besides, the transmitter is equipped with a standard waveguide WR2.8, which has 7 dBi antenna gain. As a result, the transmitter has large beamwidth and remains static at 180° azimuth angle (towards Rx) and 0° elevation angle. By contrast, the Rx is equipped with a directional antenna with 25 dBi antenna gain and 8° half-power beamwidth (HPBW). To receive MPCs from different directions, Rx scans the spatial domain with a 10° angle resolution, from 0° to 360° in the azimuth plane and −20° and 20° in the elevation plane. Note that the scanning range of elevation angles of Rx is enough to cover most of the significant paths since the heights of Rx and Tx are close, while the ground reflection path falls out of this range that is omitted.

C. Measurement Deployment

The measurement campaign is conducted in a corridor on the second floor of the Longbin Building in Shanghai Jiao Tong University, as shown in Fig. 2a. Both sides of the corridor are furnished with glass walls and metal pillars are installed on the glass walls. The separation distance between the consecutive metal pillars is around 1.26 m. Laboratory rooms lie on both sides of the corridor, furnished with wooden doors with a width around 0.93 m. Moreover, a 7 m-long segment of the walls on the southern side is covered with wood. Besides, concrete walls are on both ends of the corridor.

The positions of the transmitter and receiver are shown in Fig. 2b. The transmitter and receiver positions are in the middle of the corridor. On the west end of the corridor, the transmitter is static. Towards the east direction, overall 24 receiver positions are selected, including 21 LoS points and
3 NLoS points. In the LoS case, the first 15 positions are uniformly distributed from 5 m to 19 m with a 1 m distance interval. The remaining 6 positions are uniformly distributed from 21 m to 31 m with a 2 m interval. In the NLoS case, three Rx positions are deployed in room f, whose positions are marked clearly in Fig. 2b. For all Rx positions, the 0° azimuth direction is towards the western direction as labeled in Fig. 2b. Moreover, in the bird’s eye view, the Rx module rotates from 0° to 360° in the clockwise direction.

III. MODIFIED SAGE ALGORITHM FOR MPC EXTRACTION AND CLUSTERING PROCESS

In this section, the data processing procedure is introduced, including the MPC extraction process and clustering process. We first estimate the MPC parameters through a novel modified version of the SAGE algorithm. Dedicated for direction-scan measurement campaigns, the modified SAGE algorithm shows great effectiveness and is easier to implement compared to other extensions of SAGE algorithm. Furthermore, in the clustering process, two clustering algorithms are used and compared, namely the KPowerMeans (KPM) method [28] and DBSCAN method [29].

A. Modified SAGE for MPC Extraction

To analyze the behaviors of MPCs, the parameters of MPCs, including their delay, DoA, and DoD, need to be accurately estimated from the measurement data. In this work, we derive a modified SAGE algorithm dedicated for direction-scan measurement campaigns, which only require real-valued antenna radiation patterns rather than complex antenna radiation patterns as required for other MPC extraction algorithms. Furthermore, compared to the SAGE algorithm [26], the proposed modified SAGE algorithm avoids the effects of phase errors. The modified SAGE algorithm is derived in detail as follows.

1) Problem Formulation: In direction-scan measurement campaigns, both Tx and Rx are equipped with a single horn antenna that is mechanically rotated to align to specific directions. To cover the spatial area of interest, we assume that totally $N_t$ and $N_r$ directions are required at the Tx side and Rx side, respectively. For the $(n_t)$th pointing direction at Tx and $(n_r)$th pointing direction at Rx, the CIR is of the form as

$$h_{n_t,n_r}(\tau) = \sum_{l=1}^{L} \alpha_l c_{n_t}(\Omega_{t,l}) c_{n_r}(\Omega_{r,l}) e^{j\phi_{l,n_t,n_r}} \delta(\tau - \tau_l) + w_{n_t,n_r}(\tau)$$

where $\alpha_l$, $\tau_l$, $\Omega_{t,l}$, $\Omega_{r,l}$ denote the real-valued path gain, ToA, DoD, DoA of the $l$th MPC, respectively. Moreover, $\phi_{l,n_t,n_r}$ represents the phase of the $l$th component that attributes to effects of antennas, reflection, etc, which is dependent on the scanning angles due to the antenna effects. Besides, $L$ is the number of MPCs. Moreover, $\delta(\cdot)$ is the Dirac function and $w_{n_t,n_r}(\cdot)$ represents the white Gaussian noise components. Furthermore, $c_{n_t}(\cdot)$ and $c_{n_r}(\cdot)$ stand for the real-valued radiation pattern of Tx and Rx antenna, respectively. Note that the directions $\Omega_{t,l}, \Omega_{r,l}$ are uniquely determined by the azimuth angles $\varphi_{t,l}, \varphi_{r,l}$ and elevation angles $\theta_{t,l}, \theta_{r,l}$, as

$$\Omega_{a,l} = [\cos(\varphi_{a,l}) \cos(\theta_{a,l}), \sin(\varphi_{a,l}) \cos(\theta_{a,l}), \sin(\theta_{a,l})]^T$$

(2)

where $a$ can be either “t” or “r”, representing the DoD or DoA. Besides, $(\cdot)^T$ stands for the transpose operation.

Based on the CIR, the CTF of the channel can be expressed as

$$H_{n_t,n_r}(f) = \sum_{l=1}^{L} \alpha_l c_{n_t}(\Omega_{t,l}) c_{n_r}(\Omega_{r,l}) e^{j\phi_{l,n_t,n_r}} e^{-j2\pi f \tau_l} + W_{n_t,n_r}(f)$$

(3)

During the measurement campaign, the CTF is sampled and recorded, as

$$H_{n_t,n_r}[k] = \sum_{l=1}^{L} \alpha_l c_{n_t}(\Omega_{t,l}) c_{n_r}(\Omega_{r,l}) e^{j\phi_{l,n_t,n_r}} e^{-j2\pi f_k \tau_l} + W_{n_t,n_r}[k]$$

(4)

where $f_k$ denotes the carrier frequency at the $k$th sampling point, i.e., $f_k = f_1 + (k - 1)\Delta f$. $f_1$ and $f_2$ represent the lower and upper frequencies of the measured band. $\Delta f$ is the sampling interval in the frequency domain.

By combining all discrete CTFs from all scanning directions together, we obtain

$$\mathbf{H}[k] = \sum_{l=1}^{L} \alpha_l c_{n_t}(\Omega_{t,l}) c_{n_r}(\Omega_{r,l}) e^{j\phi_{l,n_t,n_r}} e^{-j2\pi f_k \tau_l} + \mathbf{W}[k]$$

(5)

where $\mathbf{H}$ denotes the $N_t \cdot N_r$ CTF matrix. Besides, $c_{n_t}(\cdot)$ and $c_{n_r}(\cdot)$ are column vectors that contain the effects of Tx antenna and Rx antenna, respectively. Furthermore, $\Theta_l = [\tau_l, \Omega_{t,l}, \Omega_{r,l}, \alpha_l, \phi_l]$ contains all the parameters of the $l$th path. Moreover, $s(k; \Theta_l) = \alpha_l c_{n_t}(\Omega_{t,l}) c_{n_r}(\Omega_{r,l}) e^{j\phi_{l,n_t,n_r}} e^{-j2\pi f_k \tau_l}$ denotes the component attributed to the $l$th MPC.

Based on (5), the parameters of MPCs can be estimated iteratively by following the procedure shown in Fig. 3. The procedure includes an inner loop and an outer loop, which are discussed in detail as follows.

2) Inner Loop: In the inner loop, we estimate the parameters of the significant paths that have stronger power than a certain power threshold, based on the following expectation (E) step and maximization (M) step.

a) E-step: During the E-step, we estimate the complete data of the $l$th path by invoking parallel interference cancellation (PIC), as

$$\hat{\mathbf{X}}_l^m[k] = \mathbf{H}[k] - \sum_{l'=1}^{l-1} s(k; \hat{\Theta}_l^{m-l}) - \sum_{l'=l+1}^{L} s(k; \hat{\Theta}_l^{m-l})$$

(6)

where $\hat{\Theta}_l^m$ denotes the estimated values of the parameters of $(l')$th path in the $(m)$th outer loop. As can be observed
where (8b) comes from the fact that $m/K = \Delta f \Delta \tau$. Besides, $\Delta \tau_i = \tau_i - (i - 1) \Delta \tau$.

Considering that we are only interested in the MPCs that have much higher power than the noise, the absolute value of the complete data in the time domain can be expressed as

$$|x^m[i]| \approx \alpha_t c_t(\Omega_{t,i}) c_r(\Omega_{r,i})^T \text{sinc}(\Delta f \Delta \tau_i K)$$

b) M-step: After obtaining the complete data, we then estimate the parameters of the $l$th MPC. To reduce the calculation complexity, we proceed by taking two sub-steps, e.g., the rough estimation and the accurate estimation. First, the rough estimation is based on the absolute value of the complete data in time domain, from which we can roughly obtain the delay and directions of the $l$th MPC as

$$[\hat{\tau}'_l, \hat{\Omega}'_{t,l}, \hat{\Omega}'_{r,l}] = \arg \max_{\tau_l, \Omega_{t,l}, \Omega_{r,l}} (|x^m[i]|)$$

where $\tau_i, \Omega_{t,n_t}, \Omega_{r,n_r}$ denote the discrete temporal sampling points, scanning angles of Tx and scanning angels of Rx, respectively. In other words, the delay and directions corresponding to the maximum of the 3D matrix $|x^m[i]|$ are taken as the rough estimation.

Furthermore, we search for the accurate estimations of parameters of the $l$th MPC within the neighborhood of the rough estimated values, by using the complete data in the frequency domain, as

$$\hat{\tau}'_l = \arg \max_{\tau_l \in [\hat{\tau}'_l - \Delta \tau/2, \hat{\tau}'_l + \Delta \tau/2]} \left( \frac{\sum_{k=1}^{K} \hat{X}^m_{n_t,l,n_r,l,k} e^{j2\pi f_k \tau}}{\sum_{k=1}^{K} |\hat{X}^m_{n_t,l,n_r,l,k}|} \right)^2$$

$$\hat{\phi}'_l = \text{angle} \left( \frac{\sum_{k=1}^{K} \hat{X}^m_{n_t,l,n_r,l,k} e^{j2\pi f_k \tau}}{\sum_{k=1}^{K} |\hat{X}^m_{n_t,l,n_r,l,k}|} \right)$$

$$\hat{\Omega}'_{t,l} = \arg \min_{\Omega_{t,l} \in S_{t,l}} \left( \frac{\sum_{k=1}^{K} \hat{X}^m_{n_t,l,n_r,l,k} e^{j2\pi f_k \tau}}{\sum_{k=1}^{K} |\hat{X}^m_{n_t,l,n_r,l,k}|} \right)^2$$

$$\hat{\Omega}'_{r,l} = \arg \min_{\Omega_{r,l} \in S_{r,l}} \left( \frac{\sum_{k=1}^{K} \hat{X}^m_{n_t,l,n_r,l,k} e^{j2\pi f_k \tau}}{\sum_{k=1}^{K} |\hat{X}^m_{n_t,l,n_r,l,k}|} \right)^2$$

where angle($\cdot$) is the function that calculates the phase of the complex number. Moreover, $\tau_l, S_{t,l}, S_{r,l}$ are the local regions around the rough estimations $[\hat{\tau}'_l, \hat{\Omega}'_{t,l}, \hat{\Omega}'_{r,l}]$, which are selected as

$$\tau_l = \{ \tau; \tau \in [\hat{\tau}'_l - \Delta \tau/2, \hat{\tau}'_l + \Delta \tau/2] \}$$

$$S_{t,l} = \{ \Omega_{t,l}; \Omega_{t,l} - \hat{\Omega}'_{t,l} < r \}$$

$$S_{r,l} = \{ \Omega_{r,l}; \Omega_{r,l} - \hat{\Omega}'_{r,l} < r \}$$

where the threshold $r$ is selected as 0.1. Furthermore, the objective function $z$ to estimate directions has the form as

$$z_{n_t,n_r}(\tau, \phi, \Omega_{t,l}, \Omega_{r,l}, X) = \var\left( \frac{\sum_{k=1}^{K} |X|_{n_t,n_r,k} e^{j2\pi f_k \tau} e^{j\phi_{n_t,n_r,k}}|^2}{\sum_{k=1}^{K} |X|_{n_t,n_r,k} e^{j2\pi f_k \tau} e^{j\phi_{n_t,n_r,k}}^T} \right)^2$$

where division in (19) is element-wise division of two same-sized matrices. Furthermore, var($\cdot$) denotes the function that calculates the variance. The subscripts in (13) and (14), namely $n_{t,l}$ and $n_{r,l}$, denote the indices corresponding to scanning angles near the roughly estimated $\Omega'_{t,l}$ and $\Omega'_{r,l}$, respectively. Specifically,

$$n_{t,l} = \{ n_t; |\Omega_{t,n_t} - \hat{\Omega}'_{t,l}| < r' \}$$

$$n_{r,l} = \{ n_r; |\Omega_{r,n_r} - \hat{\Omega}'_{r,l}| < r' \}$$

where $r'$ is selected as 0.3, which is enough to cover the main beam of the antennas, as shown in Fig. 4.
The state of the MPCs are judged by using the multi-path component distance loop that has very similar parameters to it. The similarities of as already estimated if there is one MPC in the last outer components in the last loop and current loop. An MPC is regarded whether the algorithm converges by comparing the MPC components if it is 1 and newly-estimated if it is 0. Besides, the MCD between two MPCs is expressed as

$$\text{MCD}(\Theta_i, \Theta_j) = \sqrt{\xi (\tau_i - \tau_j)^2 + ||\Omega_{t,i} - \Omega_{t,j}||^2_2 + ||\Omega_{r,i} - \Omega_{r,j}||^2_2}$$

where $\xi$ denotes a weighting factor that controls the weight of the delay in the MCD, which is set as 1 here. Besides, $\tau_m$ stands for the maximum delay of the interested MPCs, i.e., in our case, $\tau_m = \max_{l \in [1, L_m]} \{\tau^l_{m}, \tau^l_{m-1}\}$. Furthermore, the threshold in (24) is set as 0.01, which is small enough to distinguish different MPCs.

Based on the indicators of MPCs, the algorithm is regarded as converged if

$$\sum_{l=1}^{L_m} I^m_l \geq a L_m$$

where $a$ is a ratio threshold set as 0.9 in our work. That is, the algorithm terminates until 90 percent of the MPCs remain unchanged in the current outer loop.

4) Summary and Discussion: To summarize, the modified SAGE algorithm is different from existing extensions of the SAGE algorithm in several aspects. First, all possible effects on the phase of the received signal, such as the effects of antenna patterns, cables, wave propagation, etc., are integrated together into one variable $\phi_l$, which is accurately estimated during the M-step. In this way, we not only lower the requirement for antenna patterns from complex radiation patterns to real-valued ones, but also avoid the estimation errors caused by the ignorance of possible phase errors in other SAGE algorithms. Second, the estimation in the M-step is separated into the rough estimation and accurate estimation in the modified SAGE algorithm. The rough estimations of MPC parameters are obtained based on the IDFT of the complete data, near which the accurate estimations are further found. This two-step procedure accelerates the estimation process. Third, during the accurate estimation process, only part of the complete data around the roughly estimated DoD and DoA of the underestimated MPC are used. Due to the limited beamwidth of the horn antennas, those data from directions far from the DoD and DoA of MPCs offer no help and even interfere on the estimation of MPC parameters. These modifications make the modified SAGE algorithm much more effective and easier to implement for MPC extractions in direction-scan measurement campaigns.

B. MPC Clustering and Discussions

Methods for clustering of MPCs can be divided into into two categories, namely, partitioning-based algorithms and density-based algorithms [30]. Representative methods for these two categories are KPM method for partitioning-based algorithms...
and DBSCAN for density-based clustering algorithms, respectively [28], [29]. Therefore, taking the first Rx point as an example, we compare the performances of these two methods, where the results are shown in Fig. 5. Note that the parameters of the clusters are determined in the following manner. The delay, AoA and EoA of the clusters are valued as those of the strongest path within the cluster, while the power of clusters are the sum of power of all MPCs within the cluster.

By comparing the results of KPM method and DBSCAN method, we can draw some observations as follows. First, considering MPCs within the azimuth range from $300^\circ$ to $50^\circ$, the KPM method groups the LoS path and reflected paths in one cluster, while the DBSCAN method separates the LoS and reflected clusters. Second, for the MPCs within the azimuth range from $130^\circ$ to $260^\circ$ and delay range from 25 ns to 100 ns, the KPM method separates them into 5 clusters, namely clusters 2-6. On the contrary, the DBSCAN method only separates them as three clusters. The reason behind these two observations is that the KPM method determines the center of the clusters according to the power of the MPCs within the cluster. As a result, centers of cluster are usually near the strong MPCs. Weak MPCs near the strong MPCs make little impact on the position of cluster centers and thus are classified as one cluster with the strong MPCs, e.g., the LoS cluster. However, the DBSCAN method can be easily affected by weak MPCs, even though they are actually not important.

In conclusion, both methods have their advantages and disadvantages. On one hand, by using distance metric without power-weighting, the DBSCAN method can distinguish weak clusters around very strong clusters, e.g., the LoS cluster. However, the DBSCAN method can be easily affected by weak MPCs, even though they are actually not important. On the other hand, the KPM method is able to separate two clusters with similar power, even though they are very close. However, due to the power-weighted distance metric, weak clusters around strong clusters are normally treated as one cluster.

Furthermore, compared with the clustering results using DBSCAN method in our preliminary version at Rx point 1 [1], where the MPCs are extracted by using the noise-elimination method, the advantages of the modified SAGE algorithm are observed in two aspects. On one hand, the MPCs extracted using the noise-elimination method are on spatial/temporal grids, whose resolution is limited by the measurement set-ups, e.g., the scanning angle step, the bandwidth and frequency step, etc. In contrast, the parameters of MPCs obtained by modified SAGE algorithm are of higher resolution. On the other hand, the noise-elimination method fails to separate close clusters due to the antenna effects, while the modified SAGE algorithm shows the different clusters clearly. For instance, the MPCs around the LoS path obtained by the noise elimination method are all mixed together, while there are clear separations in the results by modified SAGE algorithm.

IV. CHANNEL CHARACTERIZATION AND ANALYSIS IN THE LOS CASE

In this section, the channel characteristics in the LoS case are calculated and analyzed, including the path loss, shadow fading, K-factor, delay spread, angular spreads, cluster parameters, as well as cross correlation among characteristics.

A. Path Loss

The path loss of the THz channel can be characterized in the forms of the best direction path loss and the omnidirectional path loss. The best directional path loss evaluates the path
TABLE II: Channel characteristics in the corridor scenario in the THz band.

| Rx points | PL\(_{\text{best}}\) [dB] | PL\(_{\text{omni}}\) [dB] | SF | Log-normal [dB] |
|-----------|----------------|----------------|-----|----------------|
| PL        | 96.10 99.31   | 101.97 103.94 | -1.08 -0.58 | -1.54 -1.77 |
|           | 104.45 105.38| 102.93 104.90| 0.08 -0.12  | 1.06 -0.30 |
|           | 105.56 106.61| 106.64 107.31| -1.74 -0.16 | -0.37 1.11 |
|           | 108.19 108.31| 108.37 108.77| 0.17 0.54   | 0.48 2.10  |
|           | 109.45 109.84| 110.12 110.54| 0.53 0.08   |
| SF        | 5.65 11.18 | 10.72 7.82 | 12.30 8.96 |
|           | 9.40 11.18 | 9.36 7.71 | 26.97 27.95 |
|           | 34.46 48.25 | 48.83 54.45 | 7.43 0.24 |
|           | 19.29 25.99 | 32.18 26.35 | 1.61 1.56 |
|           | 29.46 26.07 | 26.69 27.78 | 21.09 30.48 |
| KF        | 12.93 17.70 | 21.83 24.83 | -7.5870 0.2460 |
|           | 23.16 29.07 | 26.07 34.47 | 17.14 0.1570 |
|           | 32.18 36.35 | 19.15 29.06 | 43.95 59.80 |
|           | 29.06 26.69 | 27.78 27.78 | 43.11 54.63 |
|           | 32.72 32.73 | 27.78 27.78 | 59.80 56.29 |
| DS        | 38.52 42.12 | 46.57 52.65 | 9.17 1.56 |
|           | 43.09 48.43 | 41.75 48.25 | 2.39 2.60 |
|           | 48.83 54.45 | 37.14 41.65 | 5.76 5.60 |
|           | 54.45 59.74 | 37.14 41.65 | 6.31 5.60 |
| ASA       | 4.87 5.02 | 5.47 6.71 | 6.58 6.03 |
|           | 6.81 7.39 | 6.64 6.39 | 3.00 3.01 |
|           | 6.30 6.17 | 5.72 6.06 | 2.89 3.26 |
|           | 5.52 5.60 | 6.31 6.31 |
| ESA       | 4.87 5.02 | 5.47 6.71 | 6.58 6.03 |
|           | 6.81 7.39 | 6.64 6.39 | 3.00 3.01 |
|           | 6.30 6.17 | 5.72 6.06 | 2.89 3.26 |
|           | 5.52 5.60 | 6.31 6.31 |

where \(H_{nt, nr}[k]\) denotes the CTF for \((nt)\)th scanning direction at Tx and \((nr)\)th scanning direction at Rx.

The main results are summarized in Table II. As the distance between Rx and Tx increases, the path loss increases. Generally, the path loss is modeled as a linear function with respect to the logarithm of the Euclidean distance between Tx and Rx. In this work, a close-in (CI) free space reference distance model is used, which is defined as

\[
PL_{\text{CI}}[\text{dB}] = n_{\text{best/omni}} \log_{10} \frac{d}{d_0} + \text{FSPL}(d_0)
\]

where \(n_{\text{best/omni}}\) represents the path loss exponent (PLE) of the CI model. Besides, \(d\) is the Euclidean distance between Tx and Rx and \(d_0\) stands for the reference distance, which is selected as 1 m here. Furthermore, \(\text{FSPL}(d_0)\) is the free space path loss at the reference distance \(d_0\), which is given by the Friis’ law, as

\[
\text{FSPL}(d_0, f) = -20 \times \log_{10} \frac{c}{4\pi f d_0}
\]

where \(f\) denotes the frequency and \(c\) is the velocity of light.
fading effect. The shadow fading is usually characterized by a log-normal random variable, calculated as

$$\chi \text{ [dB]} = PL_{\text{omni}} - PL_{\text{omni}}^{\text{CI}} \sim \mathcal{N}(\mu_{SF}, \sigma_{SF})$$ (31)

where $\mu_{SF}$ and $\sigma_{SF}$ are the mean value and standard deviation of log-normal distribution, respectively.

Based on the measurement data, the shadow fading effect is evaluated and listed in Table II for all Rx positions in the LoS case. The shadow fading is fitted with log-normal distribution with the mean value as $-0.2629 \text{ dB}$ and standard deviation as $1.9338 \text{ dB}$, which is smaller than the typical values (larger than 3 dB) under 100 GHz in indoor hot spot scenarios in 3GPP [5].

C. K-factor

The K-factor is defined as the ratio between the power of the strongest cluster and the power of the remaining clusters, which evaluates how dominant the strongest cluster is. The values of the K-factor in the LoS case are summarized in Table II, which range from 5 to 15 dB. Moreover, it can be observed that the K-factor appears smaller for those Rx positions close to Tx and very far to Tx. This is due to the strong scattering from sides of the corridor when Rx is close to Tx, and the strong reflection from the eastern wall in the end of the corridor when Rx is very far from Tx. In general, the K-factor is fitted with log-normal distribution with the mean value as $9.8502 \text{ dB}$ and standard deviation as $2.0338 \text{ dB}$. These values are slightly larger than the typical values ($\mu=7 \text{ dB}$, $\sigma=4 \text{ dB}$) for frequency bands below 100 GHz in the indoor office scenario [5]. Furthermore, compared with our measurement campaigns in indoor office scenarios at 140 GHz [11], where K-factor values are observed averagely as 20 dB, the K-factor we observed in this work is much smaller. This is because the corridor scenario is narrower than the office scenario and scattering from metal pillars in sides of the corridor is very strong.

D. Delay and Angular Spreads

The power of the MPCs disperses in both the temporal and spatial domains, which is evaluated by the delay and angular spreads. The delay spread, azimuth spread of arrival (ASA) and the elevation spread of arrival (ESA) are summarized in Table II, from which we make several observations. First, the delay and angular spreads at the first three points are relatively small, which is due to the dominant LoS path and scattering from metal pillars in the west direction of the Rx. Second, the ASA grows as the distance between Tx and Rx increases. This is mainly attributed to the reflected path from the concrete wall at the end of the corridor. As the distance between Tx and Rx increases, the propagation distance of LoS path increases while the propagation distance of this reflected path decreases. As a result, the power difference between them decreases and thus the ASA becomes larger. Third, compared to the ASA, the ESA is much smaller, only around 6°. This is due to the fact that heights of the Tx and Rx are similar and the propagation of MPCs mainly disperses in the azimuth plane.

Furthermore, the delay and angular spreads are fitted with log-normal distributions. The fitting parameters are given in Table II. Specifically, the average delay spread is 25.88 ns, while the average values of ASA and ESA are 47.27° and 6.03°, respectively. These spread values are similar to the reported values in the standardization report 3GPP 38.901 in frequency bands lower than 100 GHz [5]. Besides, compared with the 4.7 ns DS and 35° ASA reported in our measurement campaigns in indoor office at 140 GHz [11], the delay and angular spreads observed in this work are larger, due to the strong scattering from metal pillars on sides of the corridor.

E. Cluster Parameters

The parameters of clusters are of great importance for cluster-based channel models, such as geometry-based stochastic channel models (GSCM) [10], [31]. Using the DBSCAN method and KPM method mentioned in Sec. III-B, the cluster parameters are calculated, including the number of clusters, cluster delay spread (CDS), cluster azimuth spread of arrival (CASA) and cluster elevation spread of arrival (CESA), as summarized in Table II. By comparing the results using two different methods, it can be observed that the number of cluster is generally smaller for results using the KPM method, while the intra-cluster delay and angular spreads values are larger for results using the KPM method. Moreover, as can be seen, the number of clusters is larger for Rx points in the middle, while the number of clusters for close or far Rx points reduces. Furthermore, the average CDS is around 6 ns for results using the DBSCAN method and 13 ns for results using the KPM method. Even though the standardized model by 3GPP ignores the CDS in frequency bands lower than 100 GHz [5], such significant CDS values indicate the necessity for modeling them. Besides, the average CASA and CESA are around 4° and 3° for results using the DBSCAN method, while those are 8° and 4° for results using the KPM method, respectively. Compared to those of the standardized indoor office scenario for frequency bands below 100 GHz [5], the observed intra-cluster angular spreads are slightly smaller, which is due to the fact that the THz spectrum exhibits larger reflection and scattering losses.

F. Cross Correlations Among Characteristics

The channel characteristics appear to be correlated to each other [11]. To investigate this, the cross correlations among the channel characteristics, including the SF, KF, DS and AS, are calculated and analyzed. In particular, for two sequences $X$ and $Y$ that contain the values of two kinds of channel characteristics in all the Rx positions, their cross correlation is evaluated with the correlation coefficient, as

$$\rho_{X,Y} = \frac{\text{Cov}(X, Y)}{\sqrt{\text{Var}(X)} \text{Var}(Y)}$$ (32)

where $\text{Cov}(X, Y)$ denotes the covariance of $X$ and $Y$ and $\text{Var}(\cdot)$ stands for the variance, respectively. The correlation coefficient ranges between -1 to 1, where 1 (-1) represents positive (negative) linear correlation and 0 denotes non-correlation, respectively.

The cross correlations among channel characteristics are summarized in Table III, with the following discussions. To
begin with, the shadow fading shows weak correlation with delay and angular spreads, while being slightly positively correlated to the K-factor. This can be understood that the more dominant the LoS cluster is, the closer the path loss is to the free space path loss, which has a PLE as 2. As the shadow fading is the deviation of the path loss to the fitted path loss model with a PLE as 1.7, a larger K-factor indicates larger shadow fading values. Furthermore, the K-factor is negatively correlated to the delay spread and ASA, since a larger K-factor means NLoS paths are less significant compared to the dominant LoS path, resulting in smaller power dispersion in both temporal and spatial domains. Last but not least, the delay and angular spreads are positively correlated to each other, since they are all related to the power dispersion of MPCs.

V. NLoS Measurement and Coverage Analysis

Based on our findings in the LoS case that the metal pillars are main scatterers instead of glass or wood, we investigate the feasibility of using reflective foils to improve the channel conditions in the NLoS case.

A. Set-up of the Reflective Foils

Reflective foils are added on the area around the turning corner to see effectiveness to enhance received signal strength in the NLoS case, as shown in Fig. 7. The size of the area with reflective foils are around $1.2\text{m} \times 1.5\text{m}$, which is enough to serve as scatterers, considering that the height difference of Tx and Rx is $0.5\text{m}$ and Rx is $1.65\text{m}$ far from the door of room f in the south direction.

|   | SF | KF | DS | ASA | ESA |
|---|----|----|----|-----|-----|
| SF | 1  | -  | -  | -   | -   |
| KF | 0.3784 | 1  | -  | -   | -   |
| DS | -0.0630 | -0.6495 | 1  | -   | -   |
| ASA | -0.0623 | 0.6213 | 1  | -   | -   |
| ESA | -0.1069 | -0.0101 | 0.5088 | 0.3379 | 1   |

TABLE III: Cross correlations among channel characteristics

To observe the improvement of the channel conditions with reflective foils, we make thorough analysis on the propagation of THz waves based on the measured power-delay-angular profile (PDAP), as shown in Fig. 8. Due to the limitation of 3D images, we omit the variations of power in the elevation plane by adding received power from all elevation angles together. This does not affect our observation since the Tx and Rx are at similar heights and MPCs mostly propagate in the horizontal plane.

First, comparing the PDAPs at Rx point 22-24 without reflective foils, we see that Rx point 22 receives strong scattering from metal pillars but very weak scattering from the walls of room f. On the contrary, walking from Rx point 22 to Rx point 24, even though the scattering from metal pillars is gradually blocked by the northern wall of room f, the scattering from the eastern wall of room f becomes much stronger, resulting in better channel conditions. The reason behind this is that from Rx point 22 to Rx point 24, the scattering from the eastern wall of room f becomes more perpendicular to the wall, for which the scattering loss is much smaller. Second, comparing the PDAPs at these points with/without reflective foils, we see that the path losses of paths 3-5 decrease, due to the smaller scattering loss of reflective foils. Third, interestingly, the amounts of decrease of the path losses of path 3-5 at Rx points 22-24 are not the same. Specifically, for Rx points 23 and 24, the decreases of path losses of paths 3 and 4 are around 10 dB and 5 dB respectively, while those at Rx point 22 are 17 dB and 14 dB. Specifically for path 3, the scattering losses from the eastern wall of room f are 40 dB, 35 dB, 32 dB for Rx points 22-24 without the reflective foils, while those are 23 dB, 25 dB, 23 dB for Rx points 22-24 with the reflective foils, respectively. The difference of scattering loss caused by different scattering angles is smaller with the reflective foils. This indicates that the added reflective foils not only decrease the scattering losses of MPCs, but also change the scattering pattern of the surface, similar to IRS. Based on these observations, we state that it is possible to use reflective foils to improve the channel conditions in the NLoS case in the THz band, as an easy-implemented and cheap alternative to the IRS.

B. Propagation analysis Without/With Reflective Foils

To obtain deep understanding how the reflective foils change the THz channel in the NLoS case, we compare the channel

C. Comparison of Channel Characteristics with/without Reflective Foils

B. Propagation analysis Without/With Reflective Foils

To observe the improvement of the channel conditions with reflective foils, we make thorough analysis on the propagation of THz waves based on the measured power-delay-angular profile (PDAP), as shown in Fig. 8. Due to the limitation of 3D images, we omit the variations of power in the elevation plane by adding received power from all elevation angles together. This does not affect our observation since the Tx and Rx are at similar heights and MPCs mostly propagate in the horizontal plane. Furthermore, the parameters of the significant paths at Rx points 22-24 are extracted using the modified SAGE algorithm, as shown in Fig. 8 (a)-(f), which mainly include four kinds of MPCs, namely i) scattered MPCs from the metal pillars, such as path 1 and 2; ii) scattered MPCs from the eastern wall of room f, such as path 3; iii) MPCs scattered on the metal pillars and then scattered on the eastern wall of room f, such as path 4; iv) MPCs scattered on the eastern wall of room f and then scattered on the western wall of room f, such as path 5. Besides, the propagation paths of these MPCs are shown in Fig. 8 (g)-(i). Based on Fig. 8, several observations are elaborated as follows.

First, comparing the PDAPs at Rx point 22-24 without reflective foils, we see that Rx point 22 receives strong scattering from metal pillars but very weak scattering from the walls of room f. On the contrary, walking from Rx point 22 to Rx point 24, even though the scattering from metal pillars is gradually blocked by the northern wall of room f, the scattering from the eastern wall of room f becomes much stronger, resulting in better channel conditions. The reason behind this is that from Rx point 22 to Rx point 24, the scattering from the eastern wall of room f becomes more perpendicular to the wall, for which the scattering loss is much smaller. Second, comparing the PDAPs at these points with/without reflective foils, we see that the path losses of paths 3-5 decrease, due to the smaller scattering loss of reflective foils. Third, interestingly, the amounts of decrease of the path losses of path 3-5 at Rx points 22-24 are not the same. Specifically, for Rx points 23 and 24, the decreases of path losses of paths 3 and 4 are around 10 dB and 5 dB respectively, while those at Rx point 22 are 17 dB and 14 dB. Specifically for path 3, the scattering losses from the eastern wall of room f are 40 dB, 35 dB, 32 dB for Rx points 22-24 without the reflective foils, while those are 23 dB, 25 dB, 23 dB for Rx points 22-24 with the reflective foils, respectively. The difference of scattering loss caused by different scattering angles is smaller with the reflective foils. This indicates that the added reflective foils not only decrease the scattering losses of MPCs, but also change the scattering pattern of the surface, similar to IRS. Based on these observations, we state that it is possible to use reflective foils to improve the channel conditions in the NLoS case in the THz band, as an easy-implemented and cheap alternative to the IRS.
(a) PDAP at Rx point 22.
(b) PDAP at Rx point 23.
(c) PDAP at Rx point 24.
(d) PDAP at Rx point 22 with reflective foils.
(e) PDAP at Rx point 23 with reflective foils.
(f) PDAP at Rx point 24 with reflective foils.

(g) Main MPCs received at Rx point 22.
(h) Main MPCs received at Rx point 23.
(i) Main MPCs received at Rx point 24.

Fig. 8: The PDAP and propagation analysis in the NLoS case. Significant MPCs are marked with stars, different color and number representing different kinds of MPCs.

TABLE IV: Comparison of channel characteristics in the NLoS case.

| Point | 22 | 23 | 24 |
|-------|----|----|----|
|       | w̄ | w   | w̄ | w | w̄ | w |
| PL    | 135.20 | 126.33 | 131.29 | 124.83 | 129.55 | 124.88 |
| PL_{best} [dB] | 129.26 | 122.60 | 127.24 | 122.82 | 127.58 | 123.13 |
| DS    | 4.56 | 7.54 | 5.81 | 5.65 | 19.73 | 14.19 |
| AS    | 13.96 | 22.81 | 10.78 | 16.52 | 48.19 | 45.71 |
| ESA[^\circ] | 3.90 | 7.18 | 6.00 | 4.40 | 3.34 | 6.00 |

*: wo means without reflective foils, w means with reflecting foils

characters, including the path loss, delay spread, and angular spreads. The channel characteristics in the NLoS case are summarized in Table IV for the three Rx locations 22-24, from which we elaborate several key observations as follows. First, a significant improvement lies on the decrease of path loss values with the aid of reflective foils. The best direction path loss reduces around 6 dB and the omnidirectional path loss reduces around 5 dB. This improvement would be particularly useful to enhance the link performance of THz communications in the NLoS case. Second, pertaining the delay spread and angular spreads, the variations are not the same at the three measurement positions. For instance, the delay spread at Rx 22 increases while those at Rx 23 and 24 decrease. Hence, the variations of these spread values are position-related, depending on particular propagation environment.

VI. CONCLUSION

In this paper, we have conducted a measurement campaign in an indoor corridor scenario at 306–321 GHz using a VNA-based channel sounder. A novel modified SAGE algorithm is proposed, which enables accurate estimations of MPC parameters in direction-scan measurement campaigns. The proposed algorithm only requires knowledge to real-valued antenna radiation patterns rather than complex ones and avoids possible phase errors, which is much easier to implement than existing parameter extraction algorithms. Furthermore, the extracted MPCs are clustered by using the DBSCAN method and the KPM method, whose advantages and disadvantages are thoroughly discussed.

According to the measurement results, the channel characteristics, including the path loss, shadow fading, K-factor, delay spread, angular spreads, cluster parameters, and cross correlations among channel statistics, are calculated and analyzed. The PLE values equal to 1.7213 for omnidirectional path loss
and 1.9596 for best direction path loss indicate the strong waveguide effect in the corridor. Compared to frequency bands lower than 100 GHz, the conducted measurement campaigns produce similar delay spread and angular spreads yet smaller cluster spread parameters.

Last but not least, two comparative measurement campaigns in the NLoS case are conducted, with and without additional reflective foils on walls. It has been observed that the reflective foils not only decrease the scattering loss, but also changes the scattering pattern of the surface, working similarly to IRS. The channel characterization results demonstrate that the help of reflective foils decrease the path loss by around 6 dB, which is useful to enhance the coverage of THz communications. Reflective foils might serve as a cheap and convenient alternate to intelligent reflecting surfaces (IRS).

REFERENCES

[1] Y. Li, Y. Wang, Y. Chen, Z. Yu, and C. Han, “Channel Measurement and Analysis in an Indoor Corridor Scenario at 300 GHz,” in Proc. of IEEE International Conference on Communications (ICC), 2022.

[2] Z. Chen, C. Han, Y. Wu, L. Li, C. Huang, Z. Zhang, G. Wang, and W. Tong, “Terahertz wireless communications for 2030 and beyond: A cutting-edge frontiers,” IEEE Communications Magazine, vol. 59, no. 1, pp. 66–72, 2021.

[3] I. F. Akyildiz, C. Han, and S. Nie, “Combating the distance problem in the millimeter wave and terahertz frequency bands,” IEEE Communications Magazine, vol. 56, no. 6, pp. 102–108, 2018.

[4] T. S. Rappaport, Y. Xing, O. Kanhere, S. Ju, A. Madanayake, S. Mandal, A. Alkhateeb, and G. C. Trichopoulos, “Wireless communications and applications above 100 GHz: Opportunities and challenges for 6G and beyond,” IEEE access, vol. 7, pp. 78729–78757, 2019.

[5] 3GPP, “Study on channel model for frequencies from 0.5 to 100 GHz,” 3rd Generation Partnership Project (3GPP), Technical Report (TR) 38.901, Dec. 2019, version 16.1.0.

[6] M. Jacob, S. Priebe, R. Dickhoff, T. Kleine-Ostmann, T. Schrader, and T. Kürner, “Diffraction in mm and Sub-mm Wave Indoor Propagation Channels,” IEEE Transactions on Microwave Theory and Techniques, vol. 60, no. 3, pp. 833–844, Jan. 2012.

[7] C. Jansen, S. Priebe, C. Moller, M. Jacob, H. Dierke, M. Koch, and T. Kürner, “Diffuse Scattering From Rough Surfaces in THz Communication Channels,” IEEE Transactions on Terahertz Science and Technology, vol. 1, no. 2, pp. 462–472, 2011.

[8] S. Ju, S. H. A. Shah, M. A. Javed, J. Li, G. Palteru, J. Robin, Y. Xing, O. Kanhere, and T. S. Rappaport, “Scattering mechanisms and modeling for terahertz wireless communications,” in Proc. of IEEE International Conference on Communications (ICC), pp. 1–7, 2019.

[9] J. Wang, C.-X. Wang, J. Huang, H. Wang, X. Gao, X. You, and Y. Hao, “A Novel 3D Non-Stationary GBSM for 6G THz Ultra-Massive MIMO Wireless Systems,” IEEE Transactions on Vehicular Technology, vol. 70, no. 12, pp. 12312–12324, 2021.

[10] J. Wang, C.-X. Wang, J. Huang, H. Wang, and X. Gao, “A General 3D Space-Time-Frequency Non-Stationary THz Channel Model for 6G Ultra-Massive MIMO Wireless Communication Systems,” IEEE Journal on Selected Areas in Communications, vol. 39, no. 6, pp. 1576–1589, 2021.

[11] Y. Chen, Y. Li, C. Han, Z. Yu, and G. Wang, “Channel Measurement and Ray-Tracing-Statistical Hybrid Modeling for Low-Terahertz Indoor Communications,” IEEE Transactions on Wireless Communications, vol. 20, no. 12, pp. 8163–8176, 2021.

[12] S. Ju, Y. Xing, O. Kanhere, and T. S. Rappaport, “Millimeter wave and sub-terahertz spatial statistical channel model for an indoor office building,” IEEE Journal on Selected Areas in Communications, vol. 39, no. 6, pp. 1561–1575, 2021.

[13] C.-L. Cheng and A. Zajic, “Characterization of propagation phenomena relevant for 300 GHz wireless data center links,” IEEE Transactions on Antennas and Propagation, vol. 68, no. 2, pp. 1074–1087, 2019.

[14] A. Madanayake, M. Jacob, T. Kleine-Ostmann, T. Schrader, and T. Kürner, “Channel and propagation measurements at 300 GHz,” IEEE Transactions on Antennas and Propagation, vol. 59, no. 5, pp. 1688–1698, 2011.