Wideband hybrid precoding techniques for THz massive MIMO in 6G indoor network deployment

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Abstract
Terahertz (THz) communication is becoming an up-and-coming technology for the future 6G networks as it offers an ultra-wide bandwidth. Appropriate channel models and precoding techniques are indispensable to support the desired coverage and to resolve the severe path loss in THz signals. Initially, in this work, the Sub-THz channel (140 GHz) response is investigated by using NYUSIM Channel Simulator for 6G indoor office scenario. The major highlight will be on radio propagation mechanisms, which impact the network performance in the form of path-loss, received power, time delays, azimuth AoD, Azimuth AoA, Elevation AoD, Elevation AoA and RMS delay in LOS environments. Recent hybrid precoding techniques depending upon frequency-independent phase-shifters not able to cope up with the beam split effect in THz massive MIMO systems, where the directional beams will split into various physical directions at various sub-carrier frequencies. The beam split effect will result in a serious array gain loss across the entire bandwidth, which has not been well investigated in THz massive MIMO systems. Therefore, to address this challenge, delay-phase precoding is proposed in this work. We then extensively investigate its diverse number of time delayers, varying number of antenna elements, and comparison with frequency—mmWave and Sub-THz have been discussed. Finally, the proposed delay-phase precoding techniques outperforms the other existing narrowband and wideband precoding techniques. Therefore, it is an effective technique to implement the future 6G indoor communication network deployment.

Keywords Terahertz · Delay-phase Precoding · Beam split effect · 6G Indoor office · Channel simulation

1 Introduction
The International Telecommunication Union (ITU) launched the official research investigation over 6G that helps to design pioneer wireless networks and also to attain self-subsisting networks. To quench out the emerging services and the applications like augmented reality, holographic communications, extremely high definition transmission of videos, the Terahertz (THz) communications acts as a backbone for the future 6G wireless networks. 6G also provides the communication with reduced latency for long distance with ultra high reliability. The THz band ranges from 0.1 to 10 THz provides significant bandwidth owing to attain ultra-high data rate. Several interpretations over 6G be has a belief that 6G provides an empowered full-dimensional coverage with unlimited wireless connectivity. For wireless communications, the peak data rate is considered to be an essential indicator to measure its effectiveness and in order to accomplish the above visions of 6G, the peak data rate should be greater than 1 Tbps [1,2]. Nevertheless, this peak data rate
will not be supported by the existing 5G millimeter wave (mmWave) bandwidth. Comparing THz band, which is in
0.1 THz–10 THz range with the mmWave, the THz band pro-
vides the significant bandwidth, for example, the bandwidth
greater than 20 GHz is provided to accomplish the extremely
high data rate. Therefore, it is extensively believed that the
communication through THz band is significant technology
for the emerging 6G wireless networks [2]. The bandwidth
provided should be used effectively in multi-user networks.
An autonomous and large dimensional networks are the key
features of 6G to provide wide coverage and ubiquitous con-
nectivity.

The bandwidth obtained in the sub 6 GHz of band, and the
mmWave band is insubstantial to gratify the exigencies of the
users in sixth generation era. The usage of available spectrum
is started moving towards the THz band in bandwidth hungry
applications [3]. In the applications of cellular, biological,
molecular and vehicular communications, distinct use cases
are encountered [4–6]. Owing to the bottleneck of large path
loss, the THz communication is circumscribed to employ in
short range applications. A good number of advanced appli-
cations like personal area networks, 6G communication, chip
to chip communication uses THz band communication [7].

But the THz signals are often affected by the path loss
for example, at 0.6 THz the path loss of 120 dB/100m will
occur. Due to this path loss issue it becomes challenging to
accomplish the expected coverage. The precoding approach
helps to resolve path loss problem and in this approach there
is no need to increase the power at the transmitter. By using
this precoding methodology, narrow beams can be generated
with large antenna array gain that combats the severe path loss
and also the entire optimization process will be simplified
into sub rate optimization processes and its complexity is
evaluated [8].

The scale of the antenna array is directly proportional
to the array gain of the emerging beam. The wavelength
of the signal obtained from the THz band is considerably
very small, therefore in THz communication, antenna arrays
of very large scale is employed. As the optimization of
path loss of THz signals can be used by THz precoding,
which is a requisite methodology for 6G wireless networks.
When compared to the precoding approach used in 5G
mmWave systems, the 6G precoding techniques is facing
new challenges because of its varied characteristics and these
challenges should detected and resolved by an effective 6G
system.

The major bottleneck of current wireless communication
system is the limitation of available spectrum and because of
that the significant quality of service cannot be provided and
this can be alleviated by adopting the THz communication.
The spectrum gridlock can be removed by applying novel
methodologies and new frequency bands. For effective THz
communication, the array-of-sub-array architecture is com-
pared with the fully connected architecture. The comparison
is made in terms of energy efficiency, spectral efficiency,
power consumption, channel estimation etc., Highly com-
plex and power consuming hardware is required for THz
communication and many new communication strategies are
applied owing to the nature of hardware [9]. In order to real-
ize a wireless backhaul with an ultra-high speed, it is vital to
analyze the bandwidth, transmission distance and the physi-
cal properties of the channel. To address these challenges in
THz band, the distance can be enhanced by using distance
aware bandwidth adaptive methodology. This approach cap-
ture all the distinct eccentricities of the channel and uses a full
spectrum of resources by enabling several high-speed links.
Highly advanced communication methodologies are needed
to enhance the distance for transmission and this helps to pro-
vide simultaneous operation of ultra high speed links [10].

2 Related work

Conventionally, various solutions were proposed for hybrid
multi-user mmWave systems [11–15]. A two-stage hybrid
precoding methodology is proposed and it necessitates
explicit channel state information (CSI) feedback from
users [12]. Initially, in the single-user scheme, both MS and
BS collectively choose the better combination of RF beam-
former and RF combiner for maximizing the channel gain
for the particular MS. The interference between the users
can be reduced by using a zero-forcing (ZF) baseband pre-
coding algorithm. Here, this algorithm is applied in the BS by
inverting the effective channel. A digital ZF baseband pre-
coder estimate a non-iterative non-feedback channel [14].
Particularly, the strongest angle of arrivals (AoAs) near the
BS and users are computed and it is used at BS and MSs for
analog beamforming. After that, the orthogonal pilot symbols
are sent by MS to the BS with the strongest AoA directions.
This simplifies the equivalent channel estimation utilised in
the BS digital ZF precoder. The RF combiner is configured
for every independent MS and the RF and baseband precoder
is designed at the BS for all the MSs jointly [15]. The ana-
log/digital precoder reduces the MSE of the data streams that
are received at the MS. Minimum mean square error (MMSE)
is also used as a solution for the part of the methodology.

In the traditional hybrid precoding approach, full array
gain is achieved by aligning the narrow beam from the ana-
log beamformer in the direction of the intended users [16].
Nevertheless, in 5G mmWave massive MIMO systems, the
generated beams from various sub-carrier frequencies con-
centrate on various physical directions because of the usage
of phase shifters which is independent of frequency and this
leads to loss in array gain [16]. Various methods are utilized
to tackle this array gain loss which is encountered by the beam
squint effect [17–21]. An optimized closed form solution is

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proposed in orthogonal frequency division multiplexing to deal with the hybrid precoding problem in the wide band massive MIMO system [18]. To enhance the hybrid precoding approach performance, an optimal solution is proposed to optimize the digital precoder and an analog beamformer iteratively in order to attain the significant performance over the whole bandwidth [1].

Additionally, the design of code books which contains beams of wide bandwidth is made in order to mitigate the beam squint effect which causes loss in array gain [20,21]. The design of wide beams having reduced array gain is attained in each sub-carrier where a semi-definite relaxation methodology is used to enhance the overall antenna array gain over the whole bandwidth [20,21]. The suggested methods show effectiveness to enhance the rate performance because the beams are squinted slightly and in case of mmWave massive MIMO systems, the loss in array gain is not a major consideration [18–21]. But these methods are not highly effective for THz band massive MIMO communication systems. The generated beams obtained at various sub-carrier frequencies decompose into individual physical directions because of the substantial number of antennas and wide bandwidth of THz signal.

The decomposition of different subcarrier beams in different directions is called beam split, and this is the fundamental difference between THz and mmWave beamforming systems. The beam split effect gives each subcarrier in the signal a different direction, where the subcarriers around the center frequency diverge around the boresight of the beam. It is inferred that the significant array gain will be attained only on the generated beams around the central frequency. And the remaining beams suffer from a high loss in array gain. Consequently, the achievable rate is degraded by the beam split effect [22,23]. This can be overcome by using the proposed Geometric Mean Decomposition (GMD) based delay-phase precoding.

### 2.1 Major contributions

- The key contribution of this work is to apply the channel model to evaluate the radio propagation mechanism at THz frequency range for 6G Massive MIMO system performance using NYUSIM Channel Simulator to verify the channel model parameters and antenna properties.
- Large scale and Small scale fading have been discussed for 6G indoor office scenario for 140 GHz operating frequency under UMi LOS environment at 0 dBm Transmit power with coverage of 100 m.
- The proposed GMD based Delay Phase Precoding with other precoding techniques have been experimented to mitigate the array gain loss due to the beam split effect.
- The proposed GMD based DPP is compared with recent Narrowband and Wideband hybrid precoding techniques in terms of Achievable sum rate per subcarrier (bits/s/Hz).
- Finally, by varying the number of time delayers, frequency (mmWave and Sub-THz), and the number of transmitting antennas in the proposed delay-phase precoding technique, the performances have been compared.

The paper is structured as Sect. 3 explains network deployment model with the small scale fading, large scale fading, power delay profile analysis. Section 4 illustrates proposed GMD based delay-phase precoding technique. The design approach with flowchart is explained under Sect. 5. Section 6 demonstrates simulation results and discussion with its inferences. Finally, Sect. 7 concludes the article.

### 3 Network deployment model

Figure 1 depicts the massive MIMO THz network deployment model. In indoor office scenarios, the major focus will be in the downlink, where the single cell access point (AP) is connected to multiple users. The small scale and large scale fading channel models, the structure of an antenna array, and the power delay profile analysis are discussed in the following subsections using NYUSIM [24].

#### 3.1 Path loss model

With NYUSIM, the expression for Close In (CI) free-space reference distance path loss model having one meter of reference distance with an additional attenuation caused by diversified atmospheric conditions were applied [25–28], and the expression is given as:

\[
PL_{CI}(f, d)[dB] = FSPL(f, 1m)[dB] + 10n\log_{10}(d) + AT[dB] + \chi_{CI}^{\sigma}
\]

(1)

Where ‘d’ represents the three dimensional (3D) receiver-transmitter separation distance, ‘f’ represents carrier frequency in GHz, where ‘n’ denotes the path loss exponent and the attenuation term is denoted by AT which is induced by atmosphere, the path loss in free space (dB) is denoted by FSPL($f$, 1 m) with one meter of separation between transmitter and receiver at $f$ and $\chi_{CI}^{\sigma}$ represents a Gaussian random variable with zero-mean and standard deviation in dB:

\[
FSPL(f, 1m)[dB] = 20\log_{10}\left(\frac{4\pi f \times 10^6}{c}\right)
\]

\[
= 20\log_{10}(f) + 32.4[dB]
\]

(2)
where $c$ represents the speed of light in a vacuum and $f$ is the frequency in GHz. The characterisation of AT is given as:

$$AT[dB] = \alpha[dB/m] \times d[m]$$ (3)

The attenuation factor (dB/m) is denoted by ‘$\alpha$’ at 1 GHz to 100 GHz of frequency, that constitutes the combined effects of attenuation of haze, rain, dry air and water vapor [26]. Here ‘$d$’ represents the 3D transmitter-receiver distance of separation in (1).

### 3.2 Channel model

The double-directional channel impulse response (CIR) $h_{dir}$, in small scale fading, having ‘L’ multi-path components for every transmission link will be provided as follows:

$$h_{dir}(t, \phi) = \sum_{l=1}^{L} P_{RX,l} \cdot e^{j\phi_l} \cdot \delta(t - \tau_l) \cdot G_{TX}(\phi - \phi_{TX,l}) \cdot G_{RX}(\phi - \phi_{RX,l})$$ (4)

Here $G_{TX}$ & $G_{RX}$ represents antenna gain at transmission and reception. $P_{RX,l}$, $\tau_l$ and $\phi_l$ denotes the magnitude of received power, propagation time delay and phase in the multi-path components. $\phi$ represents azimuth angle offset and ‘t’ denotes time. In every multi-path component, $\phi_{TX,l}$ denotes the angle of departure at the access point and $\phi_{RX,l}$ represents the angle of arrival for every mobile users (MUs).

### 3.3 Power delay profile analysis

To create a communication link between the APs and MUs and to maintain the desired data rate of a channel, the received power at the MUs should be modified accordingly. The study of the power delay profile is critical for network deployment. As shown in Table 1, there are a total of 32 input parameters are given as input to the simulator that is categorized into two divisions: antenna properties and channel parameters. The panel antenna properties are made up of 12 input parameters that are linked to the antenna arrays at transmission and reception, while the panel channel parameter is made up of 20 input parameters that provide information on the propagation channel. The proposed network parameters for urban micro-cell (UMi) indoor THz communication systems are described in Table 1. The APs use a carrier frequency of 0.14 THz and transmit power of 0 dBm [29]. The NYUSIM statistical parameters at 0.14THz LOS channel path-loss model is given by free space path loss $PL_0 = 255.32$ dB.

![Network deployment model](image-url)

Table 1  Input parameters settings

| Parameters                      | Value       |
|--------------------------------|-------------|
| Frequency (GHz)                | 140         |
| RF bandwidth (MHz)             | 1000        |
| Scenario                       | UMi         |
| Environment                    | LOS         |
| Lower Bound of T-R Separation Distance (m) | 100         |
| Upper Bound of T-R Separation Distance (m) | 100         |
| TX Power (dBm)                 | 0           |
| Base Station Height (m)        | 35          |
| User Terminal Height (m)       | 1.5         |
| Number of Rx Locations         | 16          |
| Barometric Pressure (mbar)     | 1013.25     |
| Humidity (%)                   | 50          |
| Temperature (°C)               | 20          |
| Rain Rate in mm/hr             | 0           |
| Polarization                   | Co-Pol      |
| Foliage Loss                   | No          |
| Distance within Foliage (m)    | 0           |
| Foliage Attenuation (dB/m)     | 0.4         |
| Outdoor to Indoor (O2I) Penetration Loss | No         |
| O2I Loss Type                  | ULA         |
| TX Array Type                  | ULA         |
| RX Array Type                  | ULA         |
| Number of TX Antenna Elements ($N_t$) | 256         |
| Number of RX Antenna Elements ($N_r$) | 1           |
| TX Antenna Spacing (in wavelength) | 0.5         |
| RX Antenna Spacing (in wavelength) | 0.5         |
| Number of TX Antenna Elements Per Row $W_t$ | 1           |
| Number of RX Antenna Elements Per Row $W_r$ | 1           |
| TX Antenna Azimuth HPBW        | 8°          |
| TX Antenna Elevation HPBW      | 8°          |
| RX Antenna Azimuth HPBW        | 8°          |
| RX Antenna Elevation HPBW      | 8°          |

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Fig. 2  Power spectrum

(a) 3D Power Spectrum of AoD
(b) 3D Power Spectrum of AoA

Fig. 3  Power delay profiles

(a) Omnidirectional Power Delay Profile
(b) Directional Power Delay Profile

Small Scale PDPs - 140 GHz, 1000 MHz, UMi LOS 100.0 m T-R Separation

(a) Small Scale Power Delay Profile
(b) Omnidirectional & Directional Path Loss
The power spectrum of 3D Angle of Departure (AOD) is shown in Fig. 2a, and the power spectrum of 3D Angle of Arrival (AOA) is shown in Fig. 2b which have been simulated in NYUSIM. Whereas, the Fig. 3a depicts the corresponding simulated omni-directional Power delay profile. For the 0.14 THz UMi LOS environment, the separation between transmitter and receiver is held at 100 m. The received power is $-122.8$ dBm and the path delay $\sigma$ is 17.9 ns having a Path Loss Exponent (PLE) of 2.4. The directional power delay profile for a 0.14 THz UMi LOS area with a PLE of 2.7 and transmitter and receiver antenna half power beamwidths (HPBW) of 8° azimuth and 8° elevation as shown in Fig. 3b. Both the transmitter and receiver antennas have a gain of 26.5 dBi and the received power is $-76.4$ dBm with path delay, $\sigma = 0.9$ ns. Figure 4a depicts a small scale PDP for indoor deployments with a Tx-Rx separation distance of 100 m and a frequency of 0.14 THz. For the Tx and Rx gain of 26.5 dBi, the Path Loss for Omnidirectional, Directional and Directional-best for 0.14 Thz, UMi LOS is shown in Fig. 4b with respect to Tx-Rx Separation distance.

Power Delay Profile (PDP) as shown in Fig. 3a, b is represented as a function of time delay which gives the received signal intensity through a multi-path channel. From Fig. 3a, Omnidirectional Power Delay Profile in 140 GHz UMi LOS is presented. This figure shows that in LOS, the received power diminished to zero after about 495 ns. From Fig. 3b, the Directional PDP with Strongest Power in 140 GHz UMi LOS is presented, the received power of PDP diminished to zero after about 495 ns.

From Fig. 4b, i.e. the scatter plot, The omnidirectional and directional path loss values are generated for the 140 GHz UMi LOS, with about 100 m T-R distance for LOS. The path loss scattered between 90–180 dB for LOS.

Throughout this article, the following notations are used: $A$ represents a matrix, ‘a’ denotes a vector and $a$ is a scalar. $A(i)$ illustrates the $i$th column of $A$, $(\cdot)^*$ denotes conjugate transpose, $(\cdot)^T$ denotes transpose and $\text{tr}(A)$ is its trace. $||A||$ is the Frobenius norm of $A$, and $|A|$ is its determinant; $[A|B]$ represents the horizontal concatenation; The $p$-norm of $a$ is represented as $||a||_p$; $\text{diag}(A)$ is a vector generated by the diagonal elements of the matrix $A$; $I_N$ is the $N \times N$ identity matrix; $0_{M \times N}$ is the $M \times N$ all-zeros matrix; $n \sim N(\mu, \sigma^2)$ is the complex Gaussian vector of covariance $\sigma^2$ and mean $0$. $E[\cdot]$ denotes the expectation and $R(\cdot)$ denotes the real part of the variable.

4 Delay phase precoding

As the system model is concerned, the system model of DPP will be similar as hybrid precoding, except the design of analog beamformer. In hybrid precoding technique, the design of analog beamformer includes various phase shifters, but in the delay phase precoding method, the precoding approach
Wideband hybrid precoding techniques for THz massive MIMO communication

4.1 System model

A DPP architecture is proposed, where a TD network has been introduced between the PS network and RF Chains, as shown in Fig. 5, in the existing hybrid precoding architecture [18,21] to overcome the beamsplit effect in wideband THz massive MIMO communication and to increase the array gain performance near to optimal [22,23]. The AP uses
dent beams are generated by phase shifters towards various users which is similar to that of [30]. Time delayers provides time delays and it is linked to certain RF chain that are designed using beam direction compensation methodology depending upon the THz signal bandwidth and the physical direction of user and. Consequently, in the whole ultra-wide bandwidth, the frequency dependent beams are aligned with various users. After that in the second stage, a separate zero-forcing precoding will be done for various sub-carriers for reducing the multiple user interference.

Fig. 6 Flowchart for proposed delay phase precoding

Table 2 Simulation parameters for precoding techniques

| Parameter                                | Value |
|-------------------------------------------|-------|
| The number of the AP antennas $N_t$       | 256   |
| The number of the user antennas $N_r$     | 1     |
| Number of channel paths $L$               | 4     |
| The central frequency $f_c$               | 0.14 THz |
| The bandwidth $B$                         | 5 GHz |
| The number of the subcarriers $M$         | 128   |
| The number of RF chains $N_{RF}$          | 4     |
| The number of TD elements $K$             | 4,16  |
| Physical directions of the paths $\theta_i$, $\phi_i$ | $\mathcal{U}(-\pi/2, \pi/2)$ |
| The transmission SNR $P_t/\sigma^2$       | $-20 \sim 15$ dB |
The total number of resolvable paths are represented by \( \text{‘L’} \); \( \tau_l \) and \( gl \) denotes the path delay and path gain of the \( l^{th} \) path, \( \theta_{l,m}, \phi_{l,m} \in [-1, 1] \) represent the spatial direction of the transmitter and the receiver of the \( l^{th} \) path and \( m^{th} \) sub-carrier, respectively, and \( f_l (\theta_{l,m}), f_l (\phi_{l,m}) \) denotes the array responses in the transmitter and the receiver.

For example, \( f_l (\theta_{l,m}) \) is represented as follows:

\[
f_l (\theta_{l,m}) = \frac{1}{\sqrt{N_r}} \begin{bmatrix} 1, e^{j \pi \theta_{l,m}}, \ldots, e^{j \pi (N_r - 1) \theta_{l,m}} \end{bmatrix}^T
\]

(10)

\[
f_l (\phi_{l,m}) = \frac{1}{\sqrt{N_r}} \begin{bmatrix} 1, e^{j \phi_{l,m}}, \ldots, e^{j \phi (N_r - 1) \phi_{l,m}} \end{bmatrix}^T
\]

(11)

The direction of the paths in the spatial domain is the spatial direction that is determined by the subcarrier frequency and the physical propagation direction. For example, for the transmitter spatial direction, \( \theta_{l,m} = 2d \sin \gamma_l \), where \( \gamma_l \in [-\pi/2, \pi/2] \) denotes the physical propagation direction of the \( l^{th} \) path, ‘d’ denotes the constant antenna spacing having \( d = \frac{c}{\pi^2} \) and ‘c’ represents the speed of the light.

5 Design approach

To accomplish GMD based proposed DPP precoding techniques, an effective algorithm is proposed. The essence of this proposed algorithm lies in dividing the precoder design into three stages. To maximize the desired signal power, the analog beamformer is designed in stage-1. In the second stage, the design of digital(baseband) precoder using the equivalent channel is done and the time delayers are added to it in stage-3.

In this algorithm, the analog precoder for the \( l^{th} \) beam \( F_{RF_{l,a}} \) is estimated initially in the step 2 for generating the beams in the spatial direction \( \theta_{l,c} \). After that the analog beamformer \( F_{RF_a} \) is generated in step 4. The time delays by ‘K’ TD elements have been generated in subsequent steps 11–13, where the direction of beams are altered from \( \theta_{l,c} \) to \( \theta_{l,m} \) at the frequency \( f_m \). Next in step 16, the analog beamformer with the time delay \( A^{TD}_m \) is generated. Ultimately, in step 5–8, the digital precoder \( F_{BB_m} \) is estimated depending upon the equivalent channel \( H_{m,eq} \) by Geometric Mean Decomposition (GMD) precoding method [31], because the existing Delay Phase Precoding’s requires the complex bit allocation to match the different signal-to-noise ratios (SNRs) of various sub-channels. Therefore, geometric mean decomposition (GMD)-based Digital Precoder is proposed in Delay Phase Precoding to avoid the complex bit allocation. From the Algorithm 1, the GMD based DPP achieves near-optimal achievable rate, as every beam is aligned with the spatial

4.2 THz channel model

A wide band ‘ray-based’ channel model is considered for the THz channel [22]. By denoting ‘\( f \)’ as bandwidth and ‘\( f_c \)’ as central frequency, the \( m^{th} \) sub-carrier frequency is represented as follows:

\[
f_m = f_c + \frac{f}{M} \left( m - 1 - \frac{M - 1}{2} \right), \quad m = 1, 2, \ldots, M.
\]

(8)

For \( m^{th} \) sub-carrier, the channel is given as

\[
H_m = \sum_{l=1}^{L} g_l e^{-j 2 \pi \tau_{l,m}} f_l (\theta_{l,m}) f_l (\phi_{l,m})^H
\]

(9)
direction at all the sub-carriers by time delays. Finally, the proposed DPP gives two-dimensional analog beamformer from the traditional one-dimensional analog beamformer, i.e., the only control of the phase shifts is extended to the joint control of time delays and phase shifts as shown in Fig. 6. The DPP approach attains the near optimal array gain in the entire bandwidth whereas the required amount of time delays are reduced significantly. For example, for 1024 antenna elements having four RF chains, the required amount of time delays are minimized from 4096 to 128, which obviously leads to reduced power consumption [30].

6 Simulation results and discussion

We consider the system parameters described in Table 2, with a AP employing an 256 elements of uniform linear array and associated with 128 subcarriers(M) and each having a single antenna. Simulations are performed in MATLAB assuming multi-path channels (L = 4). The performance of the proposed solution is shown in terms of the average achievable sumrate(R) per carrier:

\[
R = \frac{1}{M} \sum_{m=1}^{M} \log_2 \left( I_{N_c} + \frac{P_t}{N_0 \sigma_m^2} H_m F_{RF_m} F_{BB_m} F_{BB_m}^H F_{RF_m}^H H_m^H \right) \]

(12)

Figure 7 compares the achievable rate per subcarrier of different narrowband precoding methodologies along with the Proposed Delay Phase Precoding, the simulation results are provided. Besides Spatially Sparse Precoding [16], Kalman Hybrid Precoding [11], MMSE Hybrid Precoding [15], ZF Hybrid Precoding [14], Analog Beamsteering are included. The proposed Delay Phase Precoding shows the best performance as per Table 3, along with the Spatially Sparse Precoding, and Kalman Hybrid Precoding, while ZF Hybrid Precoding and Analog Beamsteering achievable rate per subcarrier is lower due to the fact that they are dealing with the beam split effect.

Figure 8 compares the achievable rate per subcarrier of different wideband precoding techniques along with the Proposed Delay Phase Precoding and Optimal Precoding. The Proposed GMD based Delay Phase Precoding shows the best performance (yields 79.64% of optimal precoding) compared to that of SVD based DPP (which yields 78.44% of optimal precoding) as shown in Table 4. The Wideband and Widebeam hybrid precoding’s achievable rate per subcarrier are

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**Algorithm 1:** Proposed GMD based Delay Phase Precoding for THz Massive MIMO [22]

**Inputs:** Spatial directions \( \theta_i \), Channel \( H_m \);  
**Outputs:** \( F_{RF_c}, F_{BB_c}, A_m^{TD} \);  
**First stage:** Analog Beamformer  
1. for \( i \in \{1, 2, \ldots, N_{RF}\} \) do  
2. Generate \( F_{RF_i} \) by \([\bar{a}_1, \bar{a}_2, \ldots, \bar{a}_K]^T = f_i(\theta_i)\);  
3. end for  
4. \( F_{RF_c} = \{F_{RF_1}, F_{RF_2}, \ldots, F_{RF_{N_{RF}}}\} \);  

**Second stage:** GMD Based Digital Precoder  
5. \( H_{m, eq} = H_m F_{RF_c} A_m^{TD} \);  
6. \( F_{BB_c} = \mu V_m eq[1:1:N_{RF}] \), \( H_{m, eq} = U_{m, eq} \Sigma_{m, eq} V_{m, eq}^H \);  
7. \( V_{m, eq} \) is fed into GMD function.;  
8. \( Q_meq R_{m, eq} F_{m, eq} V_{m, eq}^H = U_{m, eq} \Sigma_{m, eq} V_{m, eq}^H \) where \( Q, R, P \) are outputs of GMD.;  
9. end for  

**Third stage:** Delay-Phase Precoding  
10: for \( l \in \{1, 2, \ldots, N_{RF}\} \) do  
11: \( s_l = \frac{P_t}{P_{RF}^2} \);  
12: \( t_{l,i} = \left\{ \begin{array}{ll} (K - 1 - i)[s_l]T, \theta_i > 0 \\ i[s_l]T, \theta_i \leq 0 \end{array} \right\} \);  
13: \( t_l = \{t_{l,1}, t_{l,2}, \ldots, t_{l,K}\} \);  
14: end for  
15: for \( m \in \{1, 2, \ldots, M\} \) do  
16: \( A_m^{TD} = \text{blkdiag}(e^{-j2\pi ft_1}, e^{-j2\pi ft_2}, \ldots, e^{-j2\pi ft_N_{RF}}) \);  
17: end for  
18: return \( F_{RF_c}, F_{BB_c} \) and \( A_m^{TD} \).
Table 4  Comparison of various Wideband precoding algorithms and GMD-Proposed DPP (K = 16) with the Optimal precoding at SNR = 15 dB for $N_t = 256$, $N_r = 1$, and $N_{RF} = 4$

| Precoding method                        | SE (bits/s/Hz) | Accuracy in % |
|----------------------------------------|----------------|---------------|
| Optimal Precoding                      | 53.72          | 100%          |
| GMD- Proposed DPP                      | 42.68          | 79.64%        |
| SVD Based DPP                          | 42.14          | 78.44%        |
| Wideband hybrid precoding              | 30.84          | 57.41%        |
| Wide beam hybrid precoding             | 34.17          | 63.61%        |

Fig. 8  Wideband precoding algorithms comparison with the GMD-Proposed DPP (K = 16) for $N_t = 256$, $N_r = 1$ & $N_{RF} = 4$

Fig. 9  The rate performance comparison for different number of time delayers lower compared to that of the Proposed GMD based DPP and the SVD based DPP by around 20% of Accuracy.

The proposed Delay-phase precoding network’s performance with different number of time delayers has been compared in Fig. 9. From Table 5, it is inferred that for the values of K = 4 and K = 8 are higher than the K = 16.

Table 5  Comparison of various DPP precoding algorithms spectral efficiency for different time delayers at SNR = 15 dB

| Precoding method | No. of time delayers (K) | SE (bits/s/Hz) |
|------------------|--------------------------|----------------|
| Proposed DPP     | 4                        | 42.92          |
| Proposed DPP     | 8                        | 43.47          |
| Proposed DPP     | 16                       | 41.76          |

Therefore to implement the proposed network, the optimum value of the number of time delayers used in the circuit should be lesser, like K = 4 or K = 8 than choosing K = 16.

The proposed Delay-phase precoding network’s performance in THz channel and in mmWave channel has been compared in Fig. 10. It can be observed that its performance is much better at frequency, $f = 0.14$ THz (sub-THz channel) than at $f = 28$ GHz (mmWave channel). For THz channels, the DPP network is effectively being able to negate the beam split effect caused by the traditional phase-shifters and improve the performance rate substantially whereas in mmWave channel, the proposed DPP network is not able to effectively cancel out the beam squint effect.

In the Fig. 11, the performance of the proposed DPP is compared for various number of AP antennas (N). There is not much difference observed when the value of N changes.
Optimal performance is observed at $N = 1024$. Therefore, the DPP has the capability for solving the achievable rate degradation received by the beam split effect also to achieve more optimal achievable rate performance.

7 Conclusion

This work verified the downlink single cell AP connected to multiple carrier system for the 6G indoor office network deployment scenario and describes the channel model, the antenna properties, and power delay profile analysis employed using NYUSIM Channel Simulator. After that, we proposed a GMD based DPP technique where a TD network is being introduced to compensate the beam split effect and compared its performance to existing precoding techniques in narrowband and wideband. From the simulation results and the theoretical analysis, it is observed that this DPP technique is capable to eradicate the beam split effect with more optimal (79.64% of optimal) than other DPP methodology’s achievable sum rate per subcarrier performance. Under DPP Technique, the impacts of the carrier frequency of 28 GHz mmWave and 140 GHz, the number of transmitting antennas and the number of time delayers on the achievable sum rate performance for a single cell multi-carrier scenario have been compared. This work will be further broadened for multi-cell scenario in future.

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Declarations

Conflict of interest The authors declare that they have no conflict of interest

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