

6G Upper Mid-band Channels at 7.5 GHz with Diverse Urban Campaigns and Coverage Analysis via Image-Theory Full-3D Ray-Tracing

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Abstract—We present a measurement-driven characterization of path loss (PL) and delay spread (DS) at 7.5 GHz, in the lower end of the upper mid-band, for Urban Micro (UMi) and Urban Macro (UMa) scenarios, in conjunction with a feasibility assessment of 5G-6G coverage coexistence. Notably, the comparative analysis reveals that the current 3rd Generation Partnership Project (3GPP) model tends to overestimate DS in non-line-of-sight (NLOS) environments for both UMi and UMa scenarios, suggesting a need for refinement for the upper mid-band channels. To rigorously interpret the measurements, we validated the collected urban measurement data using a newly developed image-theory-based ray-tracing (RT) simulator that incorporates full-3D multi-diffraction capability, thereby addressing fundamental limitations in many academic and industrial engines. Finally, RT-based network simulations aligned with actual 5G base station sites show that 7.5 GHz can achieve coverage comparable to that of 3.5 GHz at the 90 % coverage target in the studied UMi layout. Specifically, a gain-learnable digital predistortion (DPD) scheme is employed to increase the average power amplifier (PA) output power by 2.5 dB while preserving linearity and transmission data integrity (i.e., within error vector magnitude limits), thereby confirming the feasibility of 5G-6G coexistence in urban networks.

Index Terms—6G Upper Mid-band, Outdoor Channel Measurement, Urban Micro/Macro, Ray-tracing, Coverage Analysis.

I. INTRODUCTION

IN the emerging 6G era, the demand for high data rates and ubiquitous connectivity is driving the search for new spectral resources. The upper mid-band (7–24 GHz) has emerged as a key candidate because it bridges the gap between the sub-6 GHz 5G allocations and millimeter-wave (mmWave) bands above 24 GHz. At the World Radiocommunication Conference (WRC)-23, the 4.4–4.8 GHz, 7.125–8.4 GHz, and 14.8–15.35 GHz bands were identified as prime candidates for

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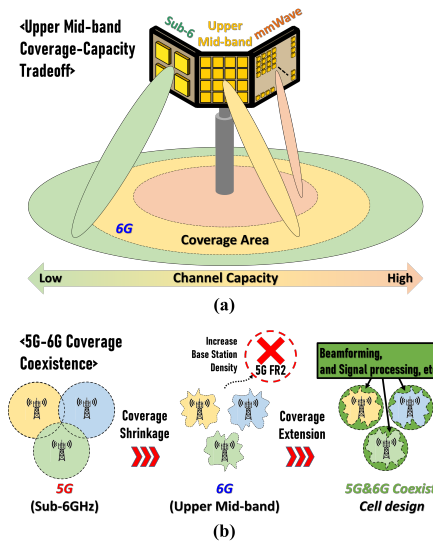


Fig. 1. Conceptual illustration of key considerations for 6G upper mid-band systems: (a) coverage–capacity tradeoff of the upper mid-band between sub-6 GHz and mmWave, and (b) 5G–6G coexistence enabled by coverage extension.

International Mobile Telecommunications (IMT)-2030 spectral assignments. Among the upper mid-band candidates, the 7.125–8.4 GHz range, referred to in this paper as the lower end of the upper mid-band, has attracted considerable interest for cellular deployment because it is expected to offer a favorable balance between propagation characteristics and coverage [1], [2]. In addition, the International Telecommunication Union (ITU) requested that studies on this band be completed in Region 3, which includes South Korea, before WRC-27. These considerations motivated our study that focuses on 7.5 GHz, a practically relevant operating point within this candidate band.

Fig. 1 illustrates the primary considerations for 6G systems operating in the upper mid-band. Although 5G mmWave technology offers high capacity, its limited coverage has spurred active research on solutions such as reconfigurable intelligent surfaces (RISs) [3], [4]. Consequently, the spectral transition for 6G preparation focused on achieving an optimal balance between coverage and capacity, underscoring the importance of precise coverage analysis in the upper mid-band. To address the coverage reduction relative to the 5G sub-6 GHz, cost-effectiveness must be prioritized. This

TABLE I
PATH LOSS AND RMS DELAY SPREAD CHARACTERISTICS FOR 6G UPPER MID-BAND IN OUTDOOR SCENARIOS.

Ref.	Freq. [GHz]	Scenario	Height [m]	T-R Distance [m]	Path Loss (CI, $d_0 = 1$ m)				RMS Delay Spread, $\mathbb{E}(\cdot)$			
					LOS		NLOS		LOS [ns]		NLOS [ns]	
					n	σ [dB]	n	σ [dB]	$\cdot$	3GPP	$\cdot$	3GPP
Our	7.5	UMi	19 (Tx1) 22 (Tx2)	20–290	2.09	4.64	3.11	8.70	45.9	43.4	77.2	88.5
		UMa	15	30–310	2.03	4.71	2.99	6.28	71.7	91.4	131.4	348.0
[5]	6.75	UMi	4	40–1000	1.79	2.57	2.56	6.53	62.8	44.3	75.6	90.5
	16.95	UMi	4	40–1000	1.85	4.05	2.59	8.78	46.5	36.2	65.8	74.0
[6]	6.5	UMi	12.5	< 210	2.07	2.97	2.38	2.54	40.7	44.7	-	-
	15	UMi	12.5	< 210	2.01	0.68	2.59	3.13	21.4	37.2	-	-
[7]	10.1	UMi	10	< 40	2.0	-	3.0	-	20.0	40.7	38.9	83.0
[8]	6	UMa	27.8	-	-	-	-	-	38.9	93.3	72.4	364.5
[9]	6.8	UMa	27	40–135	-	-	-	-	28.9	92.2	110.4	355.0
	13	UMa	27	40–135	-	-	-	-	13.5	86.6	77.6	311.0

objective requires technical solutions such as beamforming (BF) and maximizing the output power of the power amplifier (PA) while maintaining linearity through digital predistortion (DPD), rather than solely increasing the density of the base station (BS). By enhancing the performance at existing sites, these technologies support the sharing of 5G infrastructure, establishing coverage coexistence as a fundamental component of effective 6G deployment strategies.

Accurate coverage analysis first requires measurement-driven channel characterization. While the 3rd Generation Partnership Project (3GPP) channel model provides a comprehensive statistical framework spanning the frequency range of 0.5–100 GHz [10], its parameters have been primarily calibrated using data from sub-6 GHz and mmWave (> 24GHz) bands [11]. Consequently, the upper mid-band remains relatively under-characterized, motivating active 6G research to validate channel characteristics in this range and benchmark them against the 3GPP model [5], [6], [12], [13].

Table I compares recent outdoor upper mid-band studies on path loss (PL) and root-mean-square (RMS) delay spread (DS). In terms of PL, numerous studies [5]–[7], [12]–[14] have investigated the frequency dependence within the upper mid-band. Wideband measurements (3–18 GHz) [12] indicated PL trends comparable to the 3GPP predictions. Notably, band-specific analyses [5]–[7], [14] suggest that PL increases further as a function of frequency, confirming that the lower end of the upper mid-band (<15 GHz) offers superior coverage and is therefore well suited for urban 6G deployment. These findings suggest that the 3GPP PL model is applicable to the upper mid-band. However, significant discrepancies were observed regarding the DS. Multiple studies [6]–[9] have reported that the 3GPP model tends to overestimate DS values, particularly in non-line-of-sight (NLOS) urban scenarios.

Meanwhile, measurement campaigns are costly, time-

consuming, and subject to regulatory restrictions intended to avoid interference with operational commercial BS networks. In particular, the need to conduct channel measurements at sites collocated with actual 5G BSs for accurate 5G–6G coverage coexistence analysis exacerbates these challenges. To address these constraints, the need for high-fidelity ray-tracing (RT) simulators has increased, and these have been extensively developed by both industry and academia [3], [14]–[22]. At the same time, adaptive RT techniques still need to be further developed to address 6G channel characteristics [23] and applications such as integrated sensing and communication (ISAC) [24].

Overall, this study focuses on 7.5 GHz, one of the bands of interest for 6G urban deployment, and presents comparative measurements for both Urban Micro (UMi) and Urban Macro (UMa) scenarios, together with measurement data validation and coverage coexistence analysis using the developed RT framework.

A. Contributions

Our main contributions, summarized and distinguished from recent studies in Table II, are as follows:

- **Development of High-fidelity Full-3D Image Theory (IT) RT Framework:** We developed an IT RT framework that captures full-3D multi-diffraction (MD) paths, addressing limitations of many existing academic and commercial tools restricted to 2D propagation or simpler interaction models [16]–[19], [21], [25]–[30]. This enables precise channel prediction in complex upper mid-band urban environments.
- **Comparative Channel Characterization in Diverse Urban Scenarios:** This study presents a comprehensive channel analysis of PL and DS at 7.5 GHz across UMi and UMa environments, extending the scope beyond the

TABLE II
COMPARISON OF MEASUREMENT SCENARIOS AND ANALYSIS SCOPE WITH RECENT UPPER MID-BAND STUDIES

Ref.	Freq. [GHz]	Measurement Scope		Channel Characterization			System-level Analysis	
		3GPP (Location)	Env. Count	Path Loss	Delay Spread	RT Validation ^a	Coverage Sim. ^b	Real 5G Site
[5]	6.75/16.95	UMi (NYU MetroTech, New York)	Single	✓	✓	✗	✗	✗
[6]	6.5/15	UMi (BUPT Shahe, Beijing)	Single	✓	✓	✗	✓	✗
[7]	10.1	UMi (Oulu Univ. Linnanmaa, Oulu)	Single	✓	✓	✗	✗	✗
[8]	6	UMa (BUPT Shahe, Beijing)	Single	✗	✓	✗	✗	✗
[9]	6.8/13	UMa (Shanghai Univ., Shanghai)	Single	✗	✓	✗	✗	✗
Our	7.5	UMi (KT Gwanak, Seoul) & UMa (SNU Bldg 71, Seoul)	Multi (2)	✓	✓	✓	✓ (Coexistence)	✓

^a IT-RT-based interpretation and validation of measurement data.

^b Coverage simulation based on link budget analysis and actual operator deployment.

single urban type focus of the previous studies summarized in Table II.

- **RT-based Interpretation and Validation of Urban Measurement Data:** We used the proposed RT framework to interpret and validate the measured data. By decomposing the propagation into reflection (R) and diffraction (D) paths, we validated the reliability of both the simulator and measurements and derived insights relevant to 6G deployment.
- **Assessment of 5G–6G Coverage Coexistence:** Addressing a primary 6G coverage concern, we compared 6G upper mid-band with 5G sub-6 GHz coverage at actual BS sites. By incorporating technical solutions such as gain-learnable DPD to offset coverage loss, our results demonstrate the feasibility of 5G–6G coexistence in urban scenarios.

TABLE III
COMPARISON OF SIMULATION CAPABILITIES FOCUSING ON 3D MULTI-DIFFRACTION FEATURES

Ref.	Alg.	GPU Accel.	Multi-Diffraction Capability		Max. Interactions (D / R)
			Support	Plane Domain	
Our	IT	✓	✓	3D	2 / 6
[25]	IT	–	–	–	1 / 3
[26]	IT	–	–	–	0 / 6
[27]	IT	–	–	–	0 / 4
[21]	SBR	–	✓	2D Vertical	2 / 6
[16]–[18]	SBR	✓	–	–	0 / < 20
[28]	SBR	✓	–	–	0 / *
[29]	Hybrid	✓	*	*	* / 5

* Not reported in the referenced paper.

B. Organization and Notation

The remainder parts of this paper are organized as follows: Section II describes the development and performance of the proposed IT-based RT framework. Section III discusses the channel characteristics, including the PL and DS, using both the measurement data and the RT simulations. Section IV presents the 6G coverage analysis aimed at achieving coverage comparable with that of 5G. Finally, Section V concludes the paper and outlines future work. The symbols $\mathbf{a}$ and $\bar{\mathbf{A}}$ denote the vector and dyadic matrix, respectively.

II. RAY-TRACING DEVELOPMENT FOR 6G CHANNEL INTERPRETATION

In this section, we introduce our proposed IT-based RT framework, which incorporates a robust 3D MD algorithm, a capability that has often been simplified or omitted in previous studies. As the performance of reflection and single diffraction (SD) has already been established in [3], [14], this study focused on the implementation and validation of a newly developed MD engine. This section covers the need for advanced RT (Section II-A), framework architecture (Section II-B), field calculation methods (Section II-C), and a performance evaluation of the developed framework (Section II-D).

A. Necessity of Advanced RT and Diffraction Challenges

RT techniques are broadly categorized into shooting and bouncing rays (SBR) and IT methods [19], [31]. SBR is computationally efficient but often suffers from accuracy trade-offs such as ray-counting errors or cutoff issues [16]–[19]. Conversely, IT guarantees high accuracy by identifying exact propagation paths without missing rays, serving as a ground-truth model, even though it is computationally intensive [16]–[19]. In this study, we adopted an IT-based approach accelerated by the NVIDIA OptiX to ensure maximum physical fidelity for 6G channel analysis, specifically addressing the challenges of modeling complex diffraction mechanisms.

Table III presents representative academic RT studies published within the last few years, selected from a broad range of studies [16]–[19], [21], [25]–[30]. Relative to these studies, our framework provides broader support for full-3D multi-diffraction and deeper interaction modeling. While these studies have advanced RT performance using graphics processing unit (GPU) acceleration or hybrid algorithms, they often exhibit limitations in modeling diffraction. As listed in Table III, many academic simulators either completely omit diffraction [16]–[18], [26]–[28], restrict calculations to 2D planes (e.g., vertical planes only) [21], or limit the interactions to SD [25].

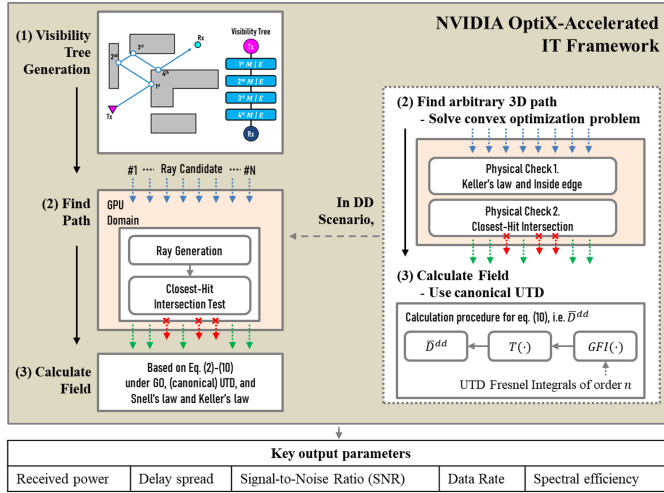


Fig. 2. Full three-dimensional IT-based RT framework.

Furthermore, the established commercial software faces challenges in robustly supporting MD in 3D geometry. Although Wireless InSite's X3D model has been documented to support up to three diffractions [32], benchmarking has been effectively limited to SD paths [33] because of the severe computational slowdown incurred by generating new ray bundles at each diffraction edge. Similarly, the MathWorks Ray-tracing Model supports only first-order diffraction [33], whereas platforms such as the NVIDIA Sionna RT [33] and FEKO WinProp [34] are progressively addressing historical issues related to MD modeling. Consequently, measurement validation using these commercial tools in MD scenarios remains virtually unknown in the literature.

The main novelty of our RT framework lies in its ability to model full-3D MD with arbitrary edge configurations, a capability that is often simplified in both academic and commercial tools. This capability is critical for the 6G upper mid-band in urban environments, where signals frequently traverse complex 3D paths, such as diffraction over rooftops and subsequently around street corners, that 2D-restricted or SD-limited engines fail to capture. The detailed implementation of this 3D MD modeling and its validation are presented in the remainder of Section II.

B. Full-3D IT-based Accelerated RT Framework

Fig. 2 illustrates the overall RT framework, which employs IT-based RT using a GPU-parallelized NVIDIA OptiX engine validated in previous studies [3], [14]. Implemented in C/C++, the framework is benchmarked against the commercial IT-based software WinProp [15] in terms of both accuracy and runtime, using the same maximum interaction order using up to six reflections and two diffractions ($N_{ref} = 6$ and $N_{diff} = 2$).

In a previous study [3], the proposed simulator accurately predicted the indoor mmWave coverage enhancement with RIS, closely matching the actual measurements. However, because highly directive reflection paths dominate in these environments, only a SD was considered. In contrast, in the 6G

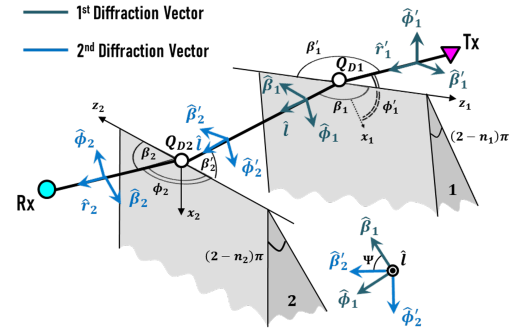


Fig. 3. Ray-fixed coordinate systems for arbitrary wedge geometry. In single diffraction case, $\hat{i}$ denotes the unit vector toward the Rx from the interaction point.

upper mid-band and massive urban channel analysis presented in this study and [14], double diffraction (DD) has a non-negligible impact, making accurate multi-diffraction modeling essential.

C. Field Calculation

1) *Reflection*: By generating a virtual image source that satisfies Snell's law at the reflecting surface, and computing the reflection coefficient using the Fresnel equations, the field at the Rx—located at a distance (l) from the surface, with the Tx at distance (r'_1)—can be calculated using (1).

$$\mathbf{E}^R(s) = \mathbf{E}(Q_R) \cdot \overbrace{\mathbf{R} \frac{r'_1}{r'_1 + l}}^{(\text{Reflection})} e^{-jkl} \quad (1)$$

where $\mathbf{E}(Q_R)$ is the incident field at the point of reflection (Q_R), and $\mathbf{R}$ is the dyadic reflection coefficient, which consists of the perpendicular and parallel reflection components given by (2) and (3):

$$R_{\perp} = \frac{\mu_2(\alpha_1 \cos \zeta_1 + j\beta_1 \cos \xi_1) - \mu_1(\alpha_2 \cos \zeta_2 + j\beta_2 \cos \xi_2)}{\mu_2(\alpha_1 \cos \zeta_1 + j\beta_1 \cos \xi_1) + \mu_1(\alpha_2 \cos \zeta_2 + j\beta_2 \cos \xi_2)} \quad (2)$$

$$R_{\parallel} = \frac{-\varepsilon_2(\alpha_1 \cos \zeta_1 + j\beta_1 \cos \xi_1) + \varepsilon_1(\alpha_2 \cos \zeta_2 + j\beta_2 \cos \xi_2)}{\varepsilon_2(\alpha_1 \cos \zeta_1 + j\beta_1 \cos \xi_1) + \varepsilon_1(\alpha_2 \cos \zeta_2 + j\beta_2 \cos \xi_2)} \quad (3)$$

The parameters ε_1 , ε_2 and μ_1 , μ_2 denote the permittivity and permeability of the two media, respectively, while α_1 , α_2 and β_1 , β_2 represent the effective attenuation and phase constants of the incident and refracted waves, respectively. The parameters ζ_1 , ζ_2 and ξ_1 , ξ_2 are the corresponding angles of incidence and refraction for the attenuation and phase vectors. Detailed derivations and structural illustrations have been provided in [35], [36].

2) *Single Diffraction*: Fig. 3 illustrates the ray-fixed coordinate system defined on the wedge geometry where diffraction occurs. The field diffracted from a single wedge is expressed as (4), based on the uniform geometrical theory of diffraction (UTD) for rays satisfying Keller's law.

$$\mathbf{E}^{SD}(s) = \mathbf{E}(Q_{D_1}) \cdot \underbrace{\bar{\mathbf{D}}_1 \sqrt{\frac{r'_1}{l(r'_1 + l)}} e^{-jk l}}_{\text{(Single Diffraction)}} \quad (4)$$

where $\mathbf{E}(Q_{D_1})$ is the incident field at the point of diffraction (Q_{D_1}), and $\bar{\mathbf{D}}_1$ is the dyadic diffraction coefficient. Subscript i denotes the diffraction order.

For single diffraction, we used the UTD coefficient proposed in [37], which is applicable to nonperfectly conducting wedges and is given in (5)–(6). Additionally, R_o and R_n denote the reflection coefficients at the o -face and n -face of the wedge, respectively, where the o -face refers to the front wall surface with an incident angle of ϕ' , and the n -face refers to the backward wall surface.

$$D_1 = D_1(L, n_1; \phi_1, \phi'_1) = R_{o,1} D_1^{(1,1)} + D_1^{(1,2)} + D_1^{(2,1)} + R_{n,1} D_1^{(2,2)} \quad (5)$$

$$D_1^{(p_1, q_1)} = D_1^{(p_1, q_1)}(L, n_1; \phi_1, \phi'_1) = \frac{-e^{-j\pi/4}}{2n_1 \sqrt{2\pi k}} \cot \frac{\Phi_1^{p_1, q_1}}{2n_1} \times F_0 \left(2kLn_1^2 \sin^2 \frac{\Phi_1^{p_1, q_1}}{2n_1} \right) \quad (6)$$

where $L = r'_1 l / (r'_1 + l)$ is a geometric ratio related to the edge geometry, and the angle terms $\Phi_i^{p_i, q_i}$ are given by:

$$\Phi_i^{p_i, q_i} = \pi + (-1)^{p_i} \phi_i + (-1)^{q_i} \phi'_i \quad (7)$$

It is important to note that the Fresnel integral formula $F_0(\cdot)$, as defined in [36], ensures continuity of the single diffraction field at the shadow boundary, thereby overcoming the limitations of the geometrical theory of diffraction.

3) *Double Diffraction*: In general, when considering consecutive multi-diffraction, it is not feasible to simply cascade single diffraction coefficients such as (6); instead, a highly efficient numerical computation of the canonical UTD transition function $T(\cdot)$ is required [38]–[40].¹

In our simulator, we adopted Albani's MD model, which is applicable under general boundary conditions and has been validated for both accuracy and efficiency by comparing it with the numerical integration of $T(\cdot)$. According to the ray-fixed coordinate system defined in Fig. 3, the computation of the DD field is described by (8)–(10). The dyadic DD coefficient ($\bar{\mathbf{D}}^{dd}$) in (8) is given by (10), and $\bar{\mathbf{F}}_i^{p_i, q_i}$ represents the detailed formulation of UTD cotangent multiplicative coefficients, which can be found in [38, Table I].

$$\mathbf{E}^{DD}(s) = \mathbf{E}(Q_{D_1}) \cdot \underbrace{\bar{\mathbf{D}}^{dd} A^{dd} e^{-jk(\ell + r_2)}}_{\text{(Double Diffraction)}} \quad (8)$$

$$A^{dd} = \frac{\sqrt{r'_1}}{\sqrt{\ell r_2}} \left[r'_1 + \ell + r_2 + \frac{r'_1 r_2}{\ell} \sin^2 \psi \right]^{-1/2} \quad (9)$$

¹See [38]–[40] for the detailed formulation of $T(\cdot)$. The internal parameters $a_1^{p_1, q_1}$, $a_2^{p_2, q_2}$, and w are omitted here for brevity.

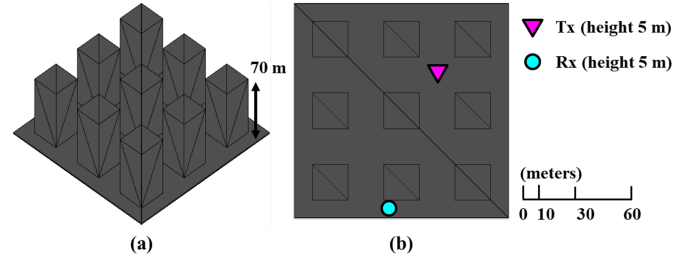


Fig. 4. CAD model of simple buildings. (a) Aerial view, and (b) Bird view.

$$\bar{\mathbf{D}}^{dd} = \frac{1}{8j\pi k n_1 n_2 \sin \beta_1 \sin \beta_2} \sum_{p_2, q_2=1}^2 \sum_{p_1, q_1=1}^2 \bar{\mathbf{F}}_2^{p_2, q_2} \cdot \bar{\mathbf{F}}_1^{p_1, q_1} \cot \left(\frac{\Phi_1^{p_1, q_1}}{2n_1} \right) \cdot \cot \left(\frac{\Phi_2^{p_2, q_2}}{2n_2} \right) T(a_1^{p_1, q_1}, a_2^{p_2, q_2}, (-1)^{p_1+q_2} w) \quad (10)$$

In DD, two distinct types of transitions occur, namely single and double, which differ from those in SD. A single transition refers to the sudden disappearance or appearance of a first-order diffracted ray as it is shadowed by the second wedge. In contrast, a double transition occurs when both geometric optics (GO) rays and first-order diffracted rays are simultaneously shadowed by the wedges. The transition function $T(\cdot)$ ensures continuity across these regions.

However, the computation of the canonical UTD transition function $T(\cdot)$ requires high-precision numerical evaluation [39], including the generalized Fresnel integrals (GFI) and UTD Fresnel integrals of order n [40], creating a major computational bottleneck that many RT tools have not adequately addressed.

D. Performance Evaluation

The simulation followed the DD framework outlined in the bottom-right corner of Fig. 2, which can be described by the algorithmic flow described next. The algorithm first generates candidate diffraction-edge sequences, and the corresponding MD points are then obtained by numerical path-length minimization [41], which is applicable to arbitrary 3D edge arrangements. Each candidate path is then subjected to two physical consistency checks. Physical check 1 verifies that the candidate diffraction points satisfy Keller's law [36, Sec. 13.3] and lie on finite edge segments, whereas physical check 2 uses NVIDIA OptiX's closest-hit program [42] to confirm the visibility along each propagation subpath. The validated paths are then passed to the UTD-based field-calculation stage, as shown in Fig. 2. All simulations were performed on an NVIDIA RTX A6000 GPU using CUDA 11.8 and OptiX 7.2.0, with dual Intel Xeon E5-2687 v4 CPUs (3.00 GHz), 512 GB RAM, and 16 threads.

To verify the accuracy of the proposed framework, we first validated it against WinProp in a benchmark scenario comprising nine buildings (Fig. 4). Given that WinProp exhibits limitations in strictly enforcing Keller's law for arbitrary 3D multi-diffraction paths [34], comparative validation was restricted to 2D-plane paths. Fig. 5 compares the angle of

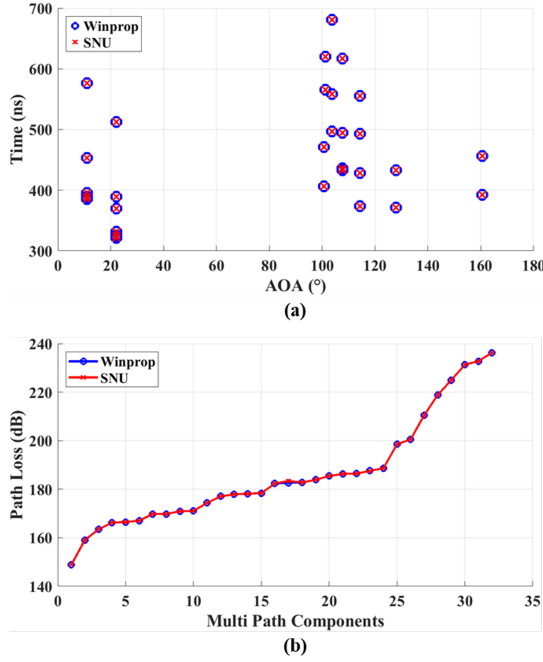


Fig. 5. Comparison between the in-house 3D simulator and commercial tool for double diffraction. (a) AoA and delay distribution, and (b) path loss of each ray.

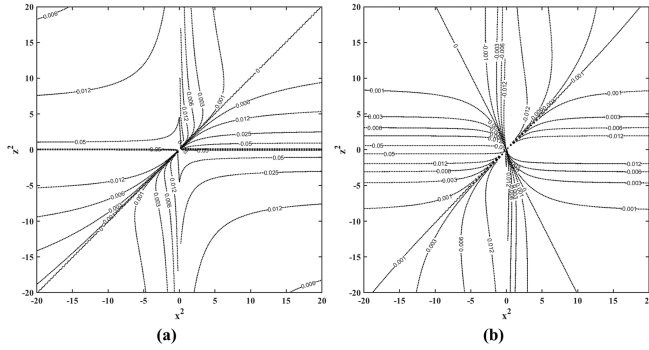


Fig. 6. Contour plots of generalized Fresnel integrals for arbitrary double-diffracted path. (a) Real part of GFI, and (b) Imaginary part of GFI.

arrival (AoA) and delay distributions, together with the PL of each double diffracted ray. Across the 32 paths, the delay and AoA discrepancies were below 0.001 ns and 0.001°, respectively, and the PL difference remained below 0.8 dB, with an average error of 0.03 dB. Furthermore, to validate the field computation capabilities of the 3D MD paths, we utilized the mathematical model and numerical analysis presented in Fig. 6. As shown in Fig. 2, the canonical UTD-based model inherently requires GFI computation; by reproducing the benchmark results of [40, Figs. 6-7], we confirmed the physical validity and accuracy of our 3D multi-diffraction modeling.

As summarized in Table III, our framework distinguishes itself from existing academic and commercial tools by combining full-3D multi-diffraction modeling with deeper interaction support. This high-fidelity capability enables the precise characterization of complex urban 6G channels and supports the

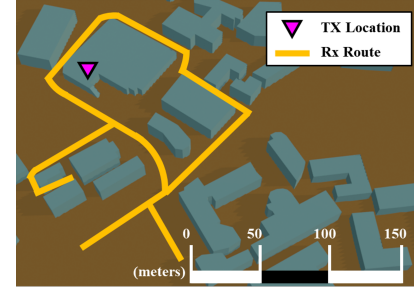


Fig. 7. CAD model of a general urban environment: This geometry corresponds to the SNU Building 71 (UMa) scenario and is used for the RT validation in the following section. The Rx route shown here represents the wideband channel-sounding route.

TABLE IV
COMPUTATION TIME COMPARISON FOR EACH CAD MODEL

CAD model	Scenario			Computation time	
	Range [m]	No. facets	No. Rx	Proposed RT	Winprop
Simple buildings	120 × 120	120	10	12.0 sec	429.0 sec
General urban	345 × 590	727	100	220.3 sec	> 7 days

measurement validation, as explained in the following section.

Meanwhile, as noted in Section II-A, although the IT-RT engine is theoretically regarded as a ground-truth RT model, its prohibitive computational load severely limited its practical utility. To the best of our knowledge, no previous study has reported urban channel validation using image-theory RT at this interaction depth. We addressed this challenge through GPU parallelization [3], achieving a significant efficiency breakthrough compared with the CPU-based WinProp, as summarized in Table IV. The comparison used identical configurations, including a uniformly distributed Rx grid across the entire scenario and maximum ray interactions ($N_{ref} = 6, N_{diff} = 2$), supported by WinProp. Notably, in a general urban environment (Fig. 7), specifically the UMa scenario at the measurement site at Building 71 at the Seoul National University (SNU), our framework completed the simulations within a feasible computation time. By contrast, WinProp cannot process the same large-scale CAD model. This confirms that our engine enables large-scale urban characterization, which is infeasible with conventional IT RT tools. This framework facilitates measurement validation in the following section, which is a dedicated study on acceleration performance, specifically for the UMi measurement scenario at KT Gwanak, featuring a 5617 facets configuration [14, Fig. 5], will be further elaborated in that separate work.

TABLE V
SUMMARY OF CHANNEL SOUNDER SETTINGS AND
MEASUREMENT-CAMPAIGN PARAMETERS

Common sounder settings		
Center frequency [GHz]	7.5	
Bandwidth [MHz]	200	
Multipath resolution [ns]	4	
Sampling rate at Tx/Rx [MHz]	240 / 250	
Sounding sequence duration [μ s]	50	
Measurement time [ms]	5	
Number of CIR snapshots	100	
Sounding waveform	Frank–Zadoff–Chu sequence	
Campaign-specific parameters		
Site	KT Gwanak	SNU Bldg 71
Tx/Rx Antenna type	Omnidirectional half-dipole	
Measurement mode	Quasi-static sounding at stopped Rx locations	
Analyzed PDP samples (LOS/NLOS)	50 / 45	45 / 50
LOS/NLOS rule	LOS: Rx point within the Tx boresight; NLOS: no clear direct Tx–Rx path.	

III. CHANNEL MEASUREMENT AT 7.5 GHZ AND ANALYSIS IN DIVERSE URBAN SCENARIOS

A. Description of the Authors' Channel Measurement Campaign

This study presents a comprehensive analysis of PL, including local transition behavior, and DS at 7.5 GHz, a representative candidate band for 6G [43], in UMi Street-Canyon and UMa scenarios.

1) *Measurement Setup and Verification:* In this study and in [14], 6G channel measurements were conducted using a channel sounder operating at a center frequency of 7.5 GHz with a bandwidth of 200 MHz as summarized in Table V. The system supported long-range transmission and reception via Global Positioning System (GPS) synchronization, where a 10 MHz reference clock, derived from the GPS module's 1 pulse per second (PPS) signal, was applied to both the Tx and Rx. Furthermore, a 35 dB PA and a 40 dB low-noise amplifier (LNA) were utilized to ensure a sufficient link budget.

Unlike directional sounders, which synthesize omnidirectional power delay profiles (PDPs) from sequential angular scans, our omnidirectional system avoids this step and thus relaxes the absolute timing requirements. Detailed system verification methods, specifically used for directional PDP measurements, are provided in [44]. The frequency and phase stability of our channel sounder and the associated measurement procedures were extensively characterized and validated in [45], ensuring reliable channel impulse response (CIR) acquisition. Furthermore, to mitigate potential measurement errors caused by GPS timing jitter, a minimum T-R separation was maintained as a spatial buffer to prevent multipath truncation.

The sounding waveform was a Frank–Zadoff–Chu (FZC) sequence that was extensively used in LTE and 5G NR downlink synchronization because of its good correlation properties. Measurements were obtained over distances up to approximately 300 m. Before data collection, back-to-

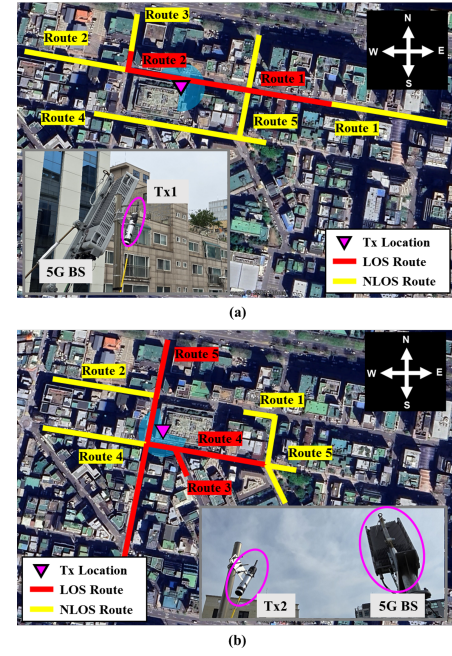


Fig. 8. Measurement route in UMi scenario. (a) Tx1 location, (b) Tx2 location. Route indices denote consistent physical paths (e.g., Route 3 (NLOS) appears only in (a) as it was not measured for Tx2). CAD model is provided in [14, Fig. 5].

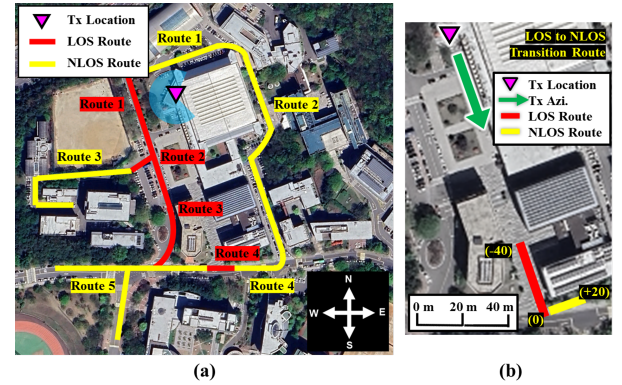


Fig. 9. Measurement routes in UMa scenario. (a) Wideband channel sounding and (b) supplementary LOS-to-NLOS transition route. CAD model is Fig. 7.

back calibration was performed to remove system frequency response noise from the Tx and Rx chains [46].

2) *Measurement Scenario:* In this study, we conducted 6G channel measurements in UMi and UMa scenarios, which were classified based on the intersite distance of 5G BSs at the KT Gwanak Office and at SNU Bldg 71, respectively, as shown in Figs. 8 and 9. The Tx heights for the configurations in Figs. 8(a), 8(b), and 9(a) are 19 m, 22 m, and 15 m, respectively, with a constant Rx height of 2 m. Additional details are provided in [14, Section II-B]. In the UMi scenario, the numbers of collected PDP samples for the LOS and NLOS conditions are 50 and 45, respectively, whereas in the UMa scenario, they are 45 and 50, respectively; these values are distinct from the total counts in [14], which incorporated narrowband continuous wave (CW) data for PL analysis. To ensure data consistency and minimize hardware-induced

variability, an identical channel sounder was used throughout the process. This approach avoids the potential errors that often arise when synthesizing results from different studies, making our dataset a reliable reference for future 6G research. The analyses were conducted near operational 5G BSs with the support of the KT Corporation, a major mobile network operator in South Korea. Unlike the studies listed in Table II, we also incorporated an in-house RT-based analysis, enabling a more comprehensive view of 6G channel characteristics and more refined 5G–6G coverage comparisons.

As a diffraction-related supplement, we measured the LOS-to-NLOS transition along the L-shaped building-corner route in the UMa scenario, as shown in Fig. 9(b). This scenario captures the abrupt signal attenuation caused by the disappearance of the direct path and the resulting shift toward diffraction-dominant propagation [47]. The Tx was placed near the actual BS location at a height of 13 m, and the Rx height was 1.5 m. Unlike the main wideband campaign in Table V, this experiment focused on PL variation and therefore used CW sounding. Along the route, the LOS region was sampled every 4 m up to 40 m (starting from the corner), and the NLOS region was sampled every 2 m up to 20 m, with a Tx-to-LOS-end distance of 140 m. Because the building geometry along the measurement route can affect the observed attenuation strength, this transition behavior should be regarded as a local-area characteristic [47]; therefore, we measured the same route at both 7.5 and 15 GHz—the latter being another band of interest [6], [43]—so that the result could be interpreted from the perspective of frequency-dependent diffraction. A 2–18 GHz dual-ridged horn (14.4/14.8 dBi at 7.5/15 GHz) was used at the Tx, while a half-dipole at 7.5 GHz and a $\lambda/4$ monopole at 15 GHz were used at the Rx. Both ends were vertically polarized, and the Tx was pointed toward the building corner.

3) *Data Processing*: The measured PDP was denoised using the double-threshold rule [5], [6], [48]:

$$P_{\text{thre}} = \max(P_{\text{noise}} + 5 \text{ dB}, P_{\text{max}} - 30 \text{ dB}) \quad (11)$$

where P_{noise} is the average noise-floor power and P_{max} is the strongest path power. This rule was adopted to balance the two complementary objectives of suppressing residual noise and limiting excessively weak components that are unrobust within the available dynamic range [44].

The measured PDP was used to compute various channel metrics. Building on the PL-focused analysis in [14], this study is extended to include DS characteristics—specifically, the mean excess delay (τ_{mean}) and the RMS DS (τ_{rms})—providing a comprehensive evaluation of PL and DS across urban scenarios. The calculations were as follows:

The τ_{mean} was calculated as the first moment of the PDP:

$$\tau_{\text{mean}} = \frac{\sum_p |h_{\text{PDP}}[p]| \cdot \tau_p}{\sum_p |h_{\text{PDP}}[p]|} \quad (12)$$

The τ_{rms} was obtained as the square root of the second central moment of the PDP:

$$\tau_{\text{rms}} = \sqrt{\overline{\tau^2} - (\tau_{\text{mean}})^2} \quad (13)$$

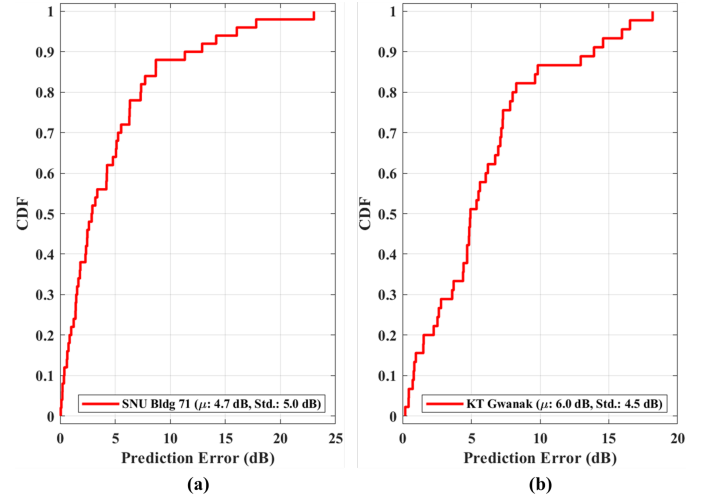


Fig. 10. Absolute prediction error of omnidirectional PL: RT simulation vs. measurements in outdoor scenarios. (a) SNU Bldg 71, (b) KT Gwanak.

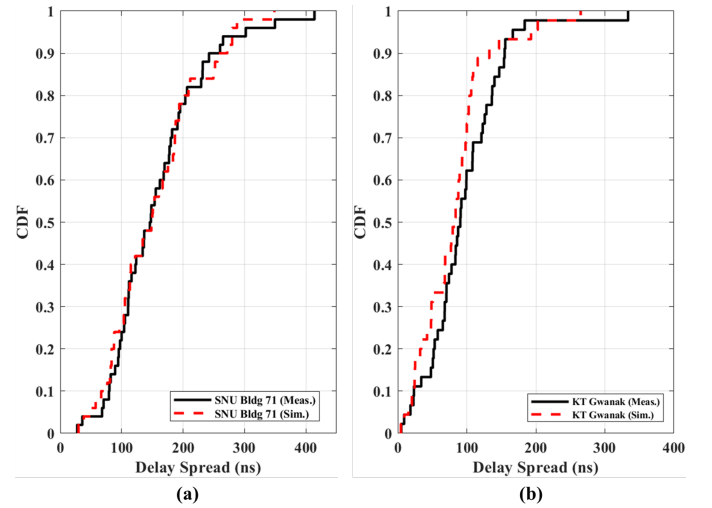


Fig. 11. Comparison between the measured and simulated RMS DS in outdoor scenarios. (a) SNU Bldg 71, (b) KT Gwanak.

where $\overline{\tau^2}$ is the second moment of delay given by:

$$\overline{\tau^2} = \frac{\sum_p |h_{\text{PDP}}[p]| \cdot \tau_p^2}{\sum_p |h_{\text{PDP}}[p]|} \quad (14)$$

B. Validation of Measurement Reliability and RT Engine Fidelity

In this section, we compare simulations with the measured UMi and UMa NLOS data to validate both the measurement reliability and the full-3D IT RT simulator. The validation process was conducted in two stages: verifying the model's generalizability across different urban environments (UMi/UMa) and conducting sensitivity and ablation studies on key simulation parameters.

1) *Validation of IT RT Framework Across Diverse Urban Measurements*: To verify how accurately the IT RT engine

TABLE VI
COMPARISON OF RMS DELAY SPREAD BETWEEN MEASUREMENT AND PREDICTION

Measurement Campaign	RMS Delay Spread					
	μ (ns)			σ (ns)		
	Meas.	Pred.	Diff.	Meas.	Pred.	Diff.
SNU Bldg 71	156.6	153.2	3.4	76.2	72.8	3.4
KT Gwanak	95.9	81.1	14.8	56.8	51.6	5.2

replicates the propagation characteristics of complex urban areas in the upper mid-band, we analyzed and compared the PL and RMS DS data measured at UMi and UMa NLOS locations with simulation results. The corresponding CAD models are presented in [14, Fig. 5] (UMi) and Fig. 7 (UMa), where the building geometries were reconstructed from OpenStreetMap and Korea's V-World platform following the methodology in [3], [14]. The relative permittivity and conductivity were configured to (3, 0.0241) for the ground and (5.31, 0.1666) for buildings, corresponding to the ITU-R standards for dry ground and concrete [49], respectively. In the baseline 7.5 GHz RT validation, all building surfaces were assigned a single default concrete material using the ITU-R values above, while surface roughness was not modeled.

The baseline simulation used a maximum ray depth of six, with $N_{ref} = 6$, $N_{diff} = 2$, fully leveraging the high-fidelity modeling capabilities of our advanced 3D IT-RT engine (Section II). Incoherent phase summation and 30 dB path power culling were applied to align with the measurement campaign specifications [14] and (11).

Fig. 10 and Fig. 11 present the spatial distribution and cumulative distribution function (CDF) comparisons between the measured values and simulation results. As established in [50], analyzing the absolute PL and DS errors is a representative methodology for validating the physical fidelity of RT engines. Specifically, these metrics show how well the simulation captures dominant paths and the delay distribution of measured path power [51].

Quantitatively, the simulation results in Fig. 10 and Table VI show a mean absolute PL error of 4.7 dB and an RMS DS error of 3.4 ns for the SNU Bldg 71 (UMa) scenario. Even in the KT Gwanak (UMi) case, which exhibits denser multipath components, the engine maintains comparable accuracy, with errors of 6.0 dB in PL and 14.8 ns in RMS DS. The slightly larger residual underprediction observed in KT Gwanak suggests that a more detailed facade-specific material assignment and higher-fidelity CAD edge/facet modeling could further improve the agreement in diffraction-heavy NLOS links. Nevertheless, these discrepancies remain within known error margins for NLOS conditions [50], demonstrating the robustness of the proposed IT RT framework despite the well-known tendency of RT to underpredict delay spreads in complex urban environments. To the best of our knowledge, this is the first study to report such rigorous large-scale prediction performance metrics using an IT RT engine in massive urban environments. Given the inherent ray cutoff and overcounting limitations of SBR-based RT, these benchmarks provide a

TABLE VII
SENSITIVITY ANALYSIS PARAMETERS AND CONFIGURATION

Category	Parameter	Status	Baseline	Tuning Range
Geometry	3D Model	Fixed	SNU Bldg. 71 (NLOS)	-
Ray-tracing	Freq. (f_c)	Fixed	7.5 GHz	-
	Ray Depth	Tuned	Ref: 6, Diff: 2	R : 2-6 D : 0-2
	Phase Rule	Tuned	Incoherent	Coherent
	Power Culling	Tuned	Rel. 30 dB	20, 40 dB
Material	Permittivity (ϵ_r)	Tuned	ITU-R (Concrete: 5.31) (Dry ground: 3.0)	$\times 0.75$ $\times 1.25$
	Conductivity (σ)	Tuned	ITU-R (Concrete: 0.17) (Dry ground: 3.0)	$\times 0.5$ $\times 1.5$

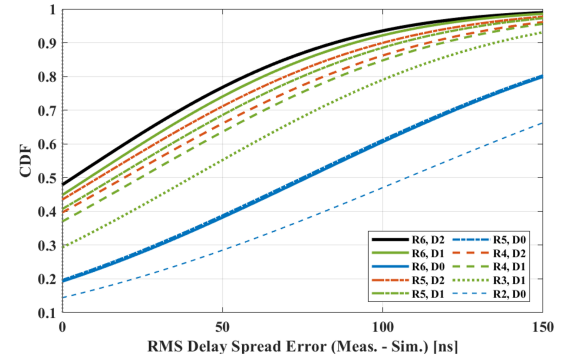


Fig. 12. Impact of ray depth on rms delay spread prediction accuracy: Analysis of reflection and diffraction counts.

meaningful reference for future advances in SBR-based RT. Overall, the results confirm that the proposed RT engine is not overfitted to specific environments, but instead demonstrates strong generalizability across diverse urban scenarios.

2) *Sensitivity Characterization in IT RT Modeling*: To further validate the physical modeling plausibility of the RT engine, we analyzed the sensitivity of RMS DS to changes in key parameters based on the UMa NLOS environment. Table VII summarizes the baseline conditions and the ranges of the tuning parameters (ray depth, phase rule, culling threshold, and material properties) varied from the baseline. The baseline settings were identical to those mentioned previously: $N_{ref} = 6$, $N_{diff} = 2$, incoherent sum, 30 dB culling, and the ITU-R material model.

Fig. 12 illustrates that restricting ray depth progressively intensifies underprediction relative to the measurements; for instance, limiting the diffraction count to 1 exacerbates the RMS DS error from the baseline's 3.4 ns to 8.4 ns, underscoring the critical role of higher-order diffractions and reflections. Complementing this, the sensitivity analysis (Fig. 13) confirmed the baseline robustness. Regarding the phase rule, coherent summation resulted in a marginal mean DS change (+5.2 ns) but yielded a significant increase in standard deviation (+12.4 ns), indicating that incoherent summation offers stability against fluctuations in wideband channels. For path power culling, the simulation results at a 30 dB threshold exhibited the highest correlation with the measurements, as

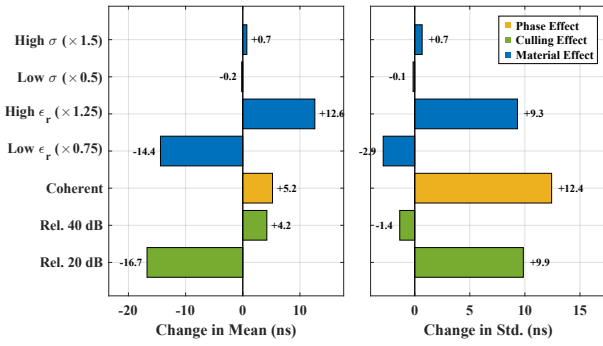


Fig. 13. Impact of material properties, phase rules, and path culling on rms delay spread. Deviations in mean and std. relative to baseline configuration.

expected from the double-threshold settings defined in (11). Finally, regarding the material properties based on the proposed settings, the conductivity had a negligible impact, whereas the permittivity exhibited a strong positive correlation with both the mean and standard deviation of DS. Collectively, this study secured the reliability of the measured data by refining calibration via multi-diffraction mechanisms and validated the RT engine through comprehensive quantitative evaluations across diverse NLOS environments.

C. Statistical Analysis of Measurements and Comparison with 3GPP Models

To mitigate the impact of uneven sample sizes and varying measurement conditions across routes, we used the inverse variance-weighted mean as the representative value for each scenario, as listed in Table VIII. In addition, to rigorously test the significance of the deviation from the 3GPP TR 38.901 reference models, we employed a linear mixed model (LMM) with the route as a random intercept. The comparative analysis yields the following insights.

- **LOS Scenarios (Consistent Agreement):** In the UMa LOS scenario, the weighted pooled mean (71.7 ns) was numerically lower than that of the 3GPP reference (91.4 ns). However, this lower value is largely driven by specific routes with low variance (e.g., Route 4) receiving higher weights. LMM analysis, which accounted for random route effects, estimated the mean at 90.5 ns, which was not statistical difference ($p = 0.96$). Thus, we conclude that the UMa LOS measurements agree well with the 3GPP model. Similarly, the UMi LOS results (45.9 ns) do not show a significant deviation from the reference (43.4 ns), confirming the validity of the standard model under LOS conditions.
- **NLOS Scenarios (Overestimation by 3GPP):** The analysis revealed that the 3GPP model tended to overestimate the RMS delay spread in NLOS environments.
 - For the UMa NLOS, both the weighted mean (131.4 ns) and the LMM-estimated mean (156.6 ns) were significantly lower than those of the 3GPP reference (348.0 ns), a finding supported by the highly significant outcome ($p < 0.001$).
 - For the UMi NLOS, the weighted mean was 77.2 ns and the LMM-estimated mean was 78.0 ns, both of

TABLE VIII
ROUTE-WISE STATISTICAL ANALYSIS OF RMS DELAY SPREAD
COMPARING SIMPLE VS. WEIGHTED POOLING ACROSS ALL SCENARIOS

Scenario / Route	N	μ	Std (σ)	95% CI
UMa LOS (Ref. 3GPP: 91.4 ns)				
Route 1	11	146.3	60.5	[105.7, 186.9]
Route 2	12	78.7	32.6	[58.0, 99.4]
Route 3	9	79.2	63.3	[30.5, 127.8]
Route 4	13	58.5	21.4	[45.5, 71.4]
Simple Pooled	45	89.5	55.6	[72.8, 106.2]
Weighted Pooled	–	71.7	–	[62.4, 81.0]
UMa NLOS (Ref. 3GPP: 348.0 ns)				
Route 1	11	152.2	51.0	[118.0, 186.4]
Route 2	8	204.0	63.6	[150.9, 257.2]
Route 3	10	102.3	24.7	[84.6, 119.9]
Route 4	7	160.2	39.2	[124.0, 196.4]
Route 5	14	169.8	112.6	[104.8, 234.9]
Simple Pooled	50	156.6	–	[135.5, 177.7]
Weighted Pooled	–	131.4	–	[119.6, 143.2]
UMi LOS (Ref. 3GPP: 43.4 ns)				
Route 1	16	49.2	24.4	[36.2, 62.2]
Route 2	8	48.6	19.9	[32.0, 65.2]
Route 3	4	44.3	10.1	[28.3, 60.3]
Route 4	10	53.0	44.3	[21.3, 84.6]
Route 5	12	45.4	24.5	[29.8, 60.9]
Simple Pooled	50	48.5	27.3	[40.8, 56.3]
Weighted Pooled	–	45.9	–	[39.2, 52.5]
UMi NLOS (Ref. 3GPP: 88.5 ns)				
Route 1	5	76.8	29.5	[40.1, 113.5]
Route 2	9	60.9	46.0	[25.6, 96.3]
Route 3	15	75.7	21.0	[64.1, 87.3]
Route 4	7	106.0	49.8	[59.9, 152.0]
Route 5 ^a	9	167.1	66.5	[116.0, 218.2]
Simple Pooled	36	78.0	–	[65.3, 90.8]
Weighted Pooled	–	77.2	–	[67.1, 87.3]

Note: N denotes sample size. **The weighted pooled mean uses inverse variance weighting ($w_i = 1/\sigma^2$), giving less weight to routes with higher standard deviation (σ).** ^aRoute 5 in UMi NLOS is excluded from the pooled estimates due to its atypical steep slope topology. All values are in nanoseconds (ns).

which were lower than that of the 3GPP reference (88.5 ns). The LMM analysis further supported a consistent trend of overestimation ($p = 0.10$), with the reference value lying near the upper bound of the 95% confidence interval.

Consequently, our findings suggest that recalibration of 3GPP models is essential for NLOS environments to prevent systematic overestimation [1].² By evaluating diverse urban scenarios using a consistent measurement setup, we have strengthened and generalized the recent findings from single-scenario studies [5], [7]–[9], which also reported 3GPP NLOS overestimations. However, the 3GPP DS parameters are jointly

²As the threshold choice can affect channel analysis results, we also checked the sensitivity of the DS mean to the threshold relative to the maximum multipath peak. For SNU Bldg 71 (NLOS), the DS means were 149.2 ns and 158.7 ns for relative thresholds of 25 dB and 35 dB, respectively. Within this threshold range, the interpretation of the DS values remains essentially unchanged.

fitted over 0.5–100 GHz and various deployment scenarios, so any update should carefully preserve this broadband validity [11]. In this sense, our results provide additional measurement evidence for validating and refining the DS model in the upper mid-band, rather than for retuning the model to a single frequency point.

D. Delay Spread Characteristics

Fig. 14 compares the measured DS with the 3GPP prediction model. The measured DS values are plotted as CDFs of τ_{rms} , revealing an overall increasing trend in urban scenarios in the order of KT Gwanak (UMi)-LOS, SNU Bldg 71 (UMa)-LOS, KT Gwanak (UMi)-NLOS, and SNU Bldg 71 (UMa)-NLOS. To provide further quantitative details, Table IX presents the OFDM system parameters calculated from the measured DS data in the corresponding UMi and UMa scenarios.

1) *Coherence Bandwidth*: The coherence bandwidth (B_{coh}) commonly used to evaluate a channel's frequency selectivity, is inversely proportional to τ_{rms} and is computed as

$$B_{coh} \approx \frac{1}{2\pi \cdot \tau_{rms}}. \quad (15)$$

As shown in Fig. 14 and Table IX, when multipath propagation becomes stronger, the resulting increase in τ_{rms} leads to a decrease in B_{coh} . For both urban types, the LOS conditions yielded higher B_{coh} values than the NLOS conditions. Specifically, the coherence bandwidths in the UMi scenario were 3.5 MHz for LOS and 2.1 MHz for NLOS. In contrast, the UMa scenario yielded the lower values of 2.2 MHz under LOS and 1.2 MHz under NLOS conditions, respectively. This confirms that UMi environments typically maintain a greater frequency coherence than UMa environments. Given that the maximum subcarrier spacing (SCS) in 5G systems is standardized at 240 kHz, the measured channels can be regarded as frequency-flat within each subcarrier. Consequently, the channel gain remains relatively constant across subcarriers in a resource block, enabling simpler equalization and more robust system performance.

2) *Cyclic Prefix Length*: In OFDM systems, a cyclic prefix (CP) is used to mitigate inter-symbol interference caused by multipath delays and must be longer than the maximum expected channel delay. Common methods for determining CP length use either τ_{rms} alone or in combination with τ_{mean} . For example, [6] sets the CP to four times τ_{rms} , while [52] defines it as $\tau_{rms} + \tau_{mean}$ (τ_{sds}). The computed values for both methods are summarized in Table IX.

In both definitions, the required guard intervals follow the same increasing order, which is consistent with the DS trends shown in Fig. 14. Consequently, because the UMa scenario exhibits larger overall delays, it is more suitable for designing systems with larger cell radii.

E. Local-Area Transition Channel Characteristic

Fig. 15 shows that at both 7.5 and 15 GHz, the measured route exhibits a distinct and abrupt LOS-to-NLOS transition after the corner. At 10 m beyond the corner, the excess loss

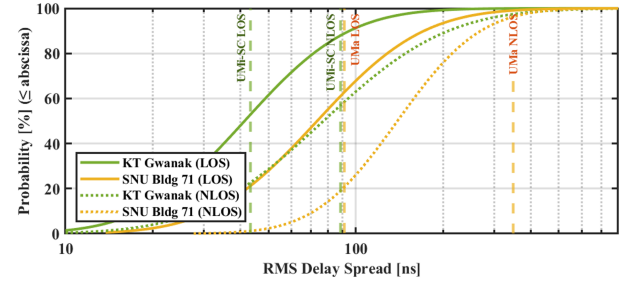


Fig. 14. CDF plot of measured DS in diverse urban scenarios.

TABLE IX
7.5 GHz OFDM SYSTEM PARAMETERS

Channel Type	τ_{mean} [ns]	τ_{rms} [ns]	τ_{sds} [ns]	B_{coh} [MHz]
UMi - LOS	34.4	45.9	80.3	3.5
UMi - NLOS	92.1	77.2	169.3	2.1
UMa - LOS	32.1	71.7	103.8	2.2
UMa - NLOS	187.1	131.4	318.5	1.2

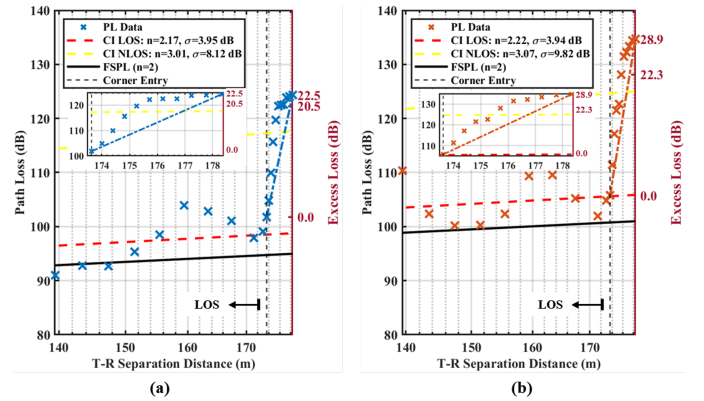


Fig. 15. Measured PL versus Tx-Rx separation distance along the LOS-to-NLOS transition route in Fig. 9(b): (a) 7.5 GHz and (b) 15 GHz.

relative to the corner point was 20.5 dB at 7.5 GHz and 22.3 dB at 15 GHz. At 16 m, these values become 22.0 dB and 27.5 dB, and at the end of the measured NLOS route (20 m), they reach 22.5 dB and 28.9 dB, respectively. This trend—more severe attenuation at higher frequencies and deeper in the shadow region—is supported by diffraction theories [36, Sec. 13] and [47].³ This indicates that achieving the desired coverage in the upper mid-band requires accounting not only for free-space path loss (FSPL), but also for additional margin to accommodate NLOS losses [43].

³Route-specific CI fitting also supported this interpretation. The fitted exponents were $n_{LOS}/n_{NLOS} = 2.17/3.01$ at 7.5 GHz and $2.22/3.07$ at 15 GHz, with corresponding σ values of 3.95/8.12 dB and 3.94/9.82 dB, respectively. This indicates a steeper and more variable postcorner NLOS behavior at both frequencies, with a slightly stronger attenuation at 15 GHz. However, these fitted values should be regarded as local descriptors of the measured corner transition route rather than as generalized UMa PL parameters.

TABLE X
DOWNLINK LINK BUDGET SIMULATION SETUP IN UMi SCENARIO [14,
TABLE II]

Etc. Parameter	Center frequency [GHz]	3.5	7.5
	Bandwidth [MHz]	100	
	SNR [dB] at edge	5	
	Edge coverage probability [%]	86	
	σ [dB]	7.82	
	Interference margin [dB]	5	
	Shadowing margin [dB]	8.45	
	Active BS locations	7 actual 5G BSs	
Transmitter end	BS HPBW in Azi., Ele. [deg]	65, 6	
	BS antenna gain [dBi]	16.00	
	BS output power [dBm]	33.00	
	BS EIRP [dBm]	49.00	
Receiver end	UE antenna gain [dBi]	0	
	UE noise figure [dB]	10	
	Receiver sensitivity [dBm]	-79.00	
Maximum allowable path loss (MAPL) [dB]		114.55	114.55

IV. 7.5 GHz COVERAGE EVALUATION

Building on this channel-level implication, we performed an RT-based coverage coexistence analysis comparing 3.5 and 7.5 GHz at actual 5G BS sites in the studied UMi environment (Fig. 8), following the measurement-to-coverage perspective adopted in prior mmWave/upper mid-band studies [6], [14], [53]. Section IV-A quantifies the additional gain required for the 7.5 GHz system to recover the reference 3.5 GHz coverage, where the additional gain is defined as an effective isotropic radiated power (EIRP) increase from array gain and higher PA output power. Section IV-B examines how the residual coverage gap after array expansion can be closed through a higher PA output power while preserving signal integrity.

A. Coverage Analysis with 5G sub-6 GHz [14]

In [14], we estimated the additional gain required for 6G to match 5G coverage in the studied UMi scenario shown in Fig. 8. The downlink link budget was calculated using the parameters in [14, Table II], from which the maximum allowable path loss (MAPL) was obtained. The BS configuration followed the synchronization signal block (SSB) beam pattern of KT Corporation's commercial 5G BSs. Since the target cell-edge SNR, edge coverage probability, and UE noise figure are typically vendor-specific and not publicly disclosed, we adopted the assumptions in [6]. The shadowing margin was determined from the selected edge coverage probability and the σ of the 3GPP UMi model using the complementary error function $\text{erfc}(\cdot)$, as discussed in [14].

In this study, a cell was formally defined as a region where the PL did not exceed the MAPL. To strictly evaluate the feasibility of 5G-6G coexistence, the coverage of the 7.5 GHz band was analyzed within the footprint of the reference 3.5 GHz cell. In the 5G (3.5 GHz) baseline, the system utilized 32 transceiver (TRX) and 192 antenna elements to form a common beam with a half-power beamwidth (HPBW) of

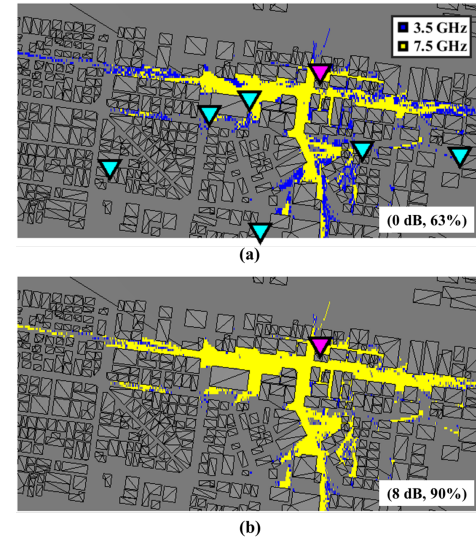


Fig. 16. Coverage comparison with the 3.5 GHz reference footprint. The pink triangle indicates the activated actual BS location. (a) Initial coverage using the legacy 5G configuration (0 dB relative gain), resulting in a limited coverage ratio of 63%. (b) Extended coverage achieving the 90% target, realized by a total gain of approximately 8 dB (comprising 6 dB from array gain and 2 dB from PA output power boosting).

$65^\circ \times 6^\circ$. For the 6G (7.5 GHz) evaluation, we modeled an antenna configuration that quadruples the number of elements compared to the 5G setup while maintaining a similar physical aperture size. Due to the increased element density at the higher frequency, the resulting HPBW narrows to approximately 16° in the horizontal direction.

The coverage performance was analyzed by comparing the reference footprint of a single 5G common beam with the spatial envelope formed by four static 6G SSB beams (i.e., multiple SSB indices). This approach synthesized the target angular service area of the 6G system to match the service area of the 3.5 GHz reference, ensuring a fair assessment of coverage coexistence. For site-specific fidelity, the Tx locations were aligned with seven actual 5G BS sites (triangles in Fig. 16(a)), including measurement sites Tx1 and Tx2. To mitigate geometric bias, the final coverage ratio was derived by averaging the results across these seven sites.

Fig. 16 compares the coverage of 5G (blue) and 6G (yellow), showing the percentage of the 6G service area within each 5G cell footprint. Fig. 16(a) illustrates the initial coverage using the legacy 5G configuration (0 dB relative gain), which resulted in a limited coverage ratio of 63%. In contrast, Fig. 16(b) presents an extended coverage scenario achieving the typical 90% requirement, which necessitates a total additional gain of approximately 8 dB. Although the physical array expansion contributes approximately 6 dB, it is insufficient to close the link budget gap, leaving a residual deficit of about 2 dB that would need to be compensated by a higher PA output power. This is consistent with the upper mid-band coverage perspective in [43]. The PA linearity implications of closing this residual gap are addressed in Section IV-B. Table XI summarizes the step-by-step roadmap for this 5G-6G coverage comparison from physical array expansion to power boosting.

TABLE XI
DETAILED GAIN DECOMPOSITION AND 5G-6G COVERAGE COMPARISON

Step (Scenario & Strategy)	HPBW (Azi. $\times$ Ele.)	Avg. Output Power	Max. EIRP	Tx Linearity Status	Coverage Coexistence
	[deg]	[dBm]	[dBm]	64 QAM	
5G Reference (3.5 GHz)	$65^\circ \times 6^\circ$	33.0	49.0	-	Reference
1. 6G Baseline (Legacy Config. Reuse)	$65^\circ \times 6^\circ$	33.0	49.0	-	n.a. (High PL)
2. Array Gain Increase (Max. Target Aperture)	$16^\circ \times 6^\circ$	33.0	55.0	-	n.a. (Gap $\approx$ 2 dB)
3. (2) + Power Boost (Conv. DPD Application)	$16^\circ \times 6^\circ$	35.5	57.5	Fail (EVM $>$ 4%)	n.a.
4. (2) + Power Boost (Proposed DPD Application)	$16^\circ \times 6^\circ$	35.5	57.5	Pass (EVM $\approx$ 1.1%)	Satisfied

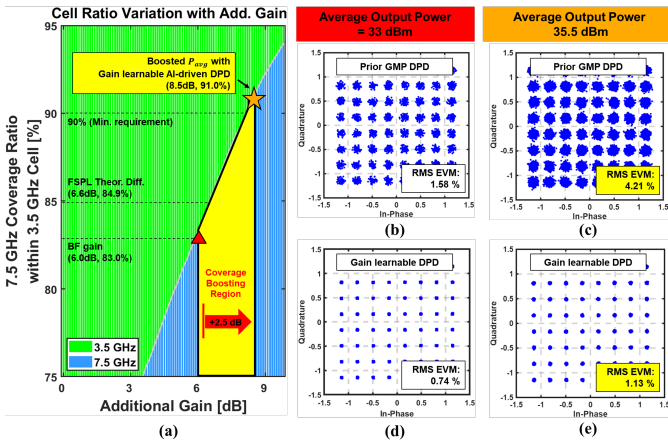


Fig. 17. Coverage variation with additional gain and constellation diagrams: (a) histogram of coverage ratio; (b), (c) GMP baseline at 33 and 35.5 dBm; (d), (e) proposed gain-learnable DPD at 33 and 35.5 dBm.

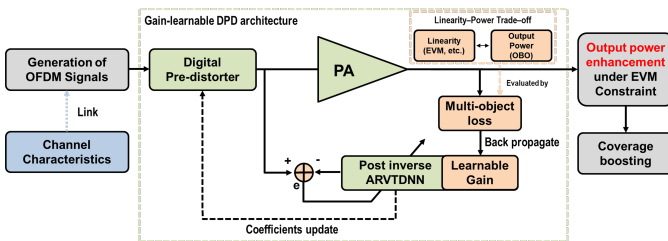


Fig. 18. Block diagram of the proposed gain-learnable DPD architecture.

B. Coverage Coexistence with 5G and 6G

Section IV-A identifies a residual coverage deficit of about 2 dB after the array expansion. Closing this gap requires increasing the average PA output power; however, this is fundamentally constrained by the PA back-off required to maintain linearity. Reducing the back-off to increase the average output power drives the PA to saturation and severely degrades the signal quality.

Two approaches are considered to resolve this conflict: deploying high-power PAs and boosting the output of existing

PAs. The former is cost prohibitive and lacks flexibility. The latter offers a cost-effective alternative by employing transmit linearization techniques, such as DPD. The DPD restores the output margin of the PA, enabling an increased average output power while preserving link quality.

To select the linearization scheme, we benchmarked the generalized memory polynomial (GMP) [54], a widely deployed baseline. However, GMP exhibits fundamental limitations when the PA is driven into deep saturation to maximize output power. Its fixed polynomial bases fail to model abrupt saturation-induced distortion, and fixed normalization creates a structural tradeoff between linearity (e.g., EVM) and output power. To overcome these limitations, we adopted a gain-learnable DPD based on an augmented real valued time delay neural network (ARVTDNN) [55], as depicted in Fig. 18. Unlike conventional DPDs with static parameters, our model jointly learns the normalization gain and network weights and adaptively optimizes its nonlinear basis. A unified loss function that jointly accounts for the normalized mean-square error and average output power guides training. This approach preserves link quality while enabling a higher average output power, yielding superior EVM performance compared with GMP. Furthermore, as a digital baseband technique, this architecture is not inherently limited to the 7.5 GHz band. The fundamental challenge of compensating for deep saturation is a common issue in PAs operating at other 6G upper mid-band frequencies (e.g., 14–15 GHz). While specific model weights require retraining to capture the unique PA characteristics, the proposed gain-learnable framework remains highly generalizable.

To evaluate transmission signal integrity, we used a 100 MHz 64-QAM OFDM signal as the test stimulus. The waveform parameters in Table XII were selected jointly from the link budget setup in Table X, the UMi DS-derived OFDM constraints in Table IX, and 5G NR FR1-compliant waveform settings while intentionally preserving a rigorous wideband linearity challenge. It should be noted that the ARVTDNN does not learn the channel inverse, as illustrated in Fig. 18. Memory depth is governed primarily by the intrinsic memory

TABLE XII
OFDM SIGNAL PROPERTIES.

Property	Value
Modulation	64-QAM
Multiplexing	OFDM
Oversampling Factor	5
FFT Length	4096
Number of Subcarriers	3276
Cyclic Prefix Length	288
Bandwidth	100 MHz
Number of Frames	100
Sample Rate	614.4 MHz

effects of the PA rather than by the channel delay spread itself, whereas the RT-derived channel statistics indirectly influence the DPD setup by informing the OFDM waveform constraints. Specifically, the high peak-to-average power ratio (PAPR ≈ 9 dB) and wide bandwidth of the selected OFDM stimulus provide a stringent PA-linearization stress condition, thereby validating that the achievable power gain is hardware-grounded rather than derived from relaxed signal conditions. For comparative analysis, the GMP model served as the conventional baseline. The PA behavioral model used in this study was derived from physical measurements and fitted with a GMP structure.

Fig. 17 shows the efficacy of the proposed solution using constellation diagrams. Fig. 17(b)–(e) compares the performance as the BS output power increased from the baseline 33 dBm. At 33 dBm, the GMP-DPD yielded a reasonable RMS EVM of 1.58%, whereas the proposed DPD achieved the superior outcome of 0.74% (Figs. 17(b) and (d)). A critical difference emerged when the average output power was boosted to 35.5 dBm. Under the conventional GMP scheme, a substantial output back-off (OBO) of approximately 7 dB was required to handle a high PAPR, limiting the feasible output to 33 dBm. Consequently, boosting the power to 35.5 dBm caused the GMP model to fail, resulting in an RMS EVM of 4.21% (Fig. 17(c)), which significantly exceeds the 2–3% limit required for high-fidelity 64-QAM transmission [56].

In contrast, the proposed DPD maintained an RMS EVM of 1.13% even at this boosted power level (Fig. 17(e)). This validates that the additional 2.5 dB gain is physically based on the hardware capability, specifically achieved by reducing the operational output back-off (OBO) to 5 dB via superior linearization. This ensures that the extra power directly translates into an improved SNR at the receiver, rather than into noise or distortion. Consequently, the system achieved 91% of the reference coverage, showing that the proposed DPD enables higher PA output power levels while preserving the signal integrity essential for tangible coverage enhancement. Thus, the DPD analysis effectively completes the RT-derived coverage evaluation by demonstrating how the remaining 7.5 GHz link budget deficit can be practically closed through transmitter-side linearization.

V. CONCLUSION AND FUTURE WORK

Focusing on 7.5 GHz within the 6G upper mid-band, this study presented a comprehensive assessment of UMi and

UMa scenarios, including measurement-driven PL and DS characterization and a coverage study on 5G–6G coexistence. The measurements showed that while the 3GPP DS model remained reasonable in terms of LOS, it tended to overestimate NLOS DS in both urban scenarios, indicating the need for further validation and refinement in upper mid-band urban channels. To interpret the measurements, we developed and validated a full-3D IT-based RT framework with multi-diffraction capability, which reproduced the measured urban channels with high fidelity. RT-based coverage analysis aligned with actual 5G BS sites further showed that 7.5 GHz can achieve coverage comparable to 3.5 GHz at the 90% target in the studied UMi layout when combined with array expansion and a gain-learnable DPD that boosted the PA output power by 2.5 dB while preserving signal integrity. Consequently, closing the residual gain gap through transmitter-side linearization demonstrates coverage coexistence and supports the practical viability of 6G deployment at the lower end of the upper mid-band. Future work will address multiple-input multiple-output channel capacity under various rank conditions, evaluate the system-level gains enabled by the enhanced spatial selectivity of larger arrays, and pursue further GPU acceleration and AI-assisted RT informed by measurement data.

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