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Communication

Design of Zero Clearance SIW Endfire Antenna Array Using Machine Learning-Assisted Optimization

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Abstract-In this paper, a substrate integrated waveguide (SIW) endfire antenna array with zero clearance is proposed for 5th generation 2 3 (5G) mobile applications using machine learning-assisted optimization. In particular, a novel impedance matching architecture that involves 4 three arbitrary pad-loading metallic vias is investigated and adopted 5 for the antenna element. Due to the stringent design requirements, the 6 locations and sizes of the vias and pads are obtained via a state-of-7 the-art machine learning assisted antenna design exploration method, parallel surrogate model-assisted hybrid differential evolution for antenna 9 synthesis (PSADEA). Keeping a very low profile, the array optimized by 10 11 PSADEA covers an operating frequency bandwidth from 36 GHz to 40 GHz. The in-band total efficiency is generally better than 60% and the 12 peak gain is above 5 dBi. The beam scanning range at 39 GHz covers 13 from -20° to 35°. 14

Index Terms—Design exploration, SIW endfire antenna, optimization,
 antenna array, surrogate modelling.

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I. INTRODUCTION

Beam-steerable arrays are widely used for 5G millimeter-wave 18 (mm-wave) applications (e.g., handset devices) to achieve both large 19 spatial coverage and high gain [1]. Due to an increasing demand 20 for thinner profiles and larger screens in the design of present-day 21 mobile devices, there is a requirement to reduce antenna profile and 22 clearance for better integration with the chip set and the printed circuit 23 24 board (PCB) of handset devices. Consequently, the clearance, which is the area to be reserved on the metal ground plane to guarantee the 25 antenna's proper functioning, becomes an important measure for 5G 26 mobile antenna design. Antennas that have zero clearance with the 27 capability of being totally integrated with the PCB board circumvent 28 this bottle neck. 29

Many works have presented the design of 5G mm-wave antennas 30 having small clearance. For example, in [2], a quasi-yagi antenna 31 array operating at 28 GHz with 2.5 mm (0.23 λ_0 , where λ_0 is the 32 wavelength in free space) clearance was proposed, and in [3], the 33 clearance of a proposed dipole array is reduced to 1.2 mm (0.15 λ_0). 34 In [4], a dual-polarized SIW/dipole array is proposed with a clearance 35 of $0.16\lambda_0$. However, it is very challenging to reduce the clearance 36 of dipole antennas to zero by covering them with closely placed 37 metal plates due to the deterioration of the radiation efficiency and 38 impedance matching. 39

In contrast to dipole antennas and some other antenna types, 40 substrate integrated waveguide (SIW) endfire antennas are excellent 41 candidates for mobile applications where antennas need to be highly 42 integrated with PCBs. Even though SIW open-ended antennas are 43 well known as low profile antennas, they suffer from significant 44 reflection at the waveguide-to-air interface when the thickness of the 45 PCB is reduced to approximately $0.1\lambda_0$ [5]. To address this bottle-46 neck, many methods have been proposed for impedance matching and 47

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to broaden the bandwidth. Popular approaches include loading with parallel transition plates [5], [6], dielectric lens [7], metamaterials [8], or the combination of different techniques [9]. For all the above mentioned solutions, enhancing the bandwidth comes with a cost of enlarging the clearance since the matching structures are loaded outside the opening end of the SIW.

SIW antennas with zero clearance and planar structures have been proposed, such as in [10] and [11]. In [10], the H-plane horn antenna employs a tapered metallic ridge inside the waveguide to achieve a wideband impedance matching. In [11], the SIW H-plane horn antenna achieved high efficiency and narrow band matching by using an air-filled waveguide and a tapered microstrip feeding line. A noticeable drawback with these approaches is the large opening angle of the waveguide, which has a great contribution to the impedance matching but also limits its application when considering beamsteering.

In this paper, a zero clearance and low profile SIW endfire array is proposed by implementing a novel impedance matching architecture. The proposed structure includes three arbitrary blind vias inside the SIW, which provides three resonances to cover the required operating band. Due to the sensitive response of the interference between each via (see Section II), the locations and sizes of the vias are critical considering the desired performances for bandwidth, realized gain, and total efficiency. Given that the conventional trialand-error, parameter sweeping, and local optimization methods are infeasible, global optimization methods are needed to find the best set of geometric parameters that meet the desired performance.

A global optimization algorithm which is widely used in antenna design exploration, the particle swarm optimization (PSO), was employed through Computer Simulation Technology Microwave Studio (CST-MWS), but satisfactory design solutions were not obtained (see Section III). Hence, a state-of-the-art machine learningassisted antenna design exploration method, the parallel surrogate model-assisted hybrid differential evolution for antenna synthesis (PSADEA) [12], [13], is then employed. Among related state-ofthe-art methods with various kinds of advantages, e.g., [14]-[16], PSADEA carries out single objective constraint global optimization and has the characteristics of high optimization ability, no need for an initial design, and can handle antenna cases with more than 30 design variables. Compared to standard global optimization methods (e.g., PSO, differential evolution (DE), genetic algorithm), it can obtain design solutions with higher quality and with over 20 times optimization speed improvement [12], [13]. The PSADEA-optimized SIW antenna shows satisfactory performance in terms of bandwidth, realized gain, and total efficiency, for both simulated and measured results.

The remainder of this paper is organized as follows: Section II 94 provides the design guidelines and modus operandi of the proposed 95 SIW endfire antenna, Section III details the optimization procedure, 96 Section IV discusses the proposed antenna's performance (simulated 97 and measured) and the concluding remarks are given in Section V. 98

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Fig. 1: Antenna element configuration.

II. ANTENNA WORKING PRINCIPLE

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100 The SIW endfire antenna was modeled and discretized in CST-MWS using a cell density of 15 cells per wavelength to have about 101 1,850,000 hexahedral mesh cells in total for optimizations. The model 102 is analyzed using time-domain finite integration technique (FIT) with 103 an accuracy of -40 dB, and each full-wave EM simulation costs about 104 20 minutes (from a wall clock) on average on a workstation with an 105 Intel 8-core i9-9900K 3.6 GHz CPU, 64 GB RAM and 8 GB NVIDIA 106 Quadro RTX 4000 GPU. The configuration of the initial model is 107 shown in Fig. 1. The PCB consists of three substrate layers. The top 108 and bottom layers are both RO4350B having a dielectric constant 109 (ϵ_r) of 3.66 and a loss tangent $(\tan \delta)$ of 0.0037. The middle layer 110 is RO4450F with ϵ_r of 3.7 and tan δ of 0.004. The thicknesses of 111 the top, middle, and bottom layers are h_1 , h_2 , and h_3 , respectively. 112 Hence, the total thickness of the PCB is $(h_1+h_2+h_3)$. The ground 113 plane, which are the top and bottom copper metal surfaces of the 114 SIW, are labelled as M1 and M3, respectively. The height of each 115 of the vias is h_1+h_2 . One end of each of the vias is grounded by a 116 connection to M3 and the other end of each of the vias is loaded 117 with a round pad marking as an additional copper layer (labelled as 118 M2 in Fig. 1). 119

The transmission line (TL) model of the SIW antenna, loaded 120 with multiple vias is shown in Fig. 2. Z_{SIW} and R_a represent 121 the characteristic impedance of the SIW and the radiation resistance 122 in the free space, respectively. By introducing vias inside the SIW, 123 multiple resonances can be generated. Each via can be seen as a 124 series of a inductance L_{pn} (n=1, 2, ...) and a capacitance C_{vn} , which 125 determines the resonant frequency. The value of L_{pn} and C_{vn} is 126 directly controlled by the diameter and height of the corresponding 127 via $(r_{vn} \text{ and } h_1+h_2)$, and the size of the pad (p_n) on this via. 128 Besides, the coupling capacitance between the adjacent vias (C_{pn}) 129 has noticeable influences on the resonant frequencies and bandwidth 130 as well. The coupling can be tuned by varying the distance (d_n) 131 between the vias. Another sensitive parameter is the location of the 132 vias relative to the SIW. Hence, all the vias are located close to the 133 center line of the SIW with a small offset (of_n) . 134

The TL model can be verified by the performances of a simple onevia model (Model-A) and a two-via model (Model-B). The structure inside the SIW of Model-A and Model-B are shown in Fig. 3(a) and Fig. 3(b), respectively. In order to further simplify the model for simulation, the substrate is reduced to a single layer of RO4350B



Fig. 2: TL model of multiple vias loaded SIW antenna.



Fig. 3: Simplified antenna models and reflection coefficients (a) Model-A with one via loaded. (b) Model-B with two vias loaded. (c) reflection coefficients.

with a thickness of 1 mm and the height of the vias are fixed at 0.9 140 mm. Fig. 3(c) presents the reflection coefficients of Model-A, model-141 B, and a comparison SIW antenna without loading. From Fig. 3(c), 142 it can be seen that without any impedance matching structure, the 143 SIW antenna is totally mismatched at the observed frequency band. 144 By adding one via (Via 1) close to the SIW opening end, Model-145 A obtains one passband at the resonant frequency of 26.3 GHz. 146 When another via (Via 2) is added between Via 1 and the feeding 147 port, Model-B obtains two resonances at 25.8 GHz and 27 GHz, 148 respectively. The antenna quality factor (Q) can be estimated by (1), 149 where f_c is the center frequency, f_1 and f_2 are the cut-off frequencies 150 of the lower and higher bound. 151

$$Q_A = f_c / (f_2 - f_1) \tag{1}$$

If the return loss of 6 dB is taken as a reference for defining the 152 frequency span, the Q value of Model-A is 23.5, whereas the Q values 153 of the two resonances of Model-B are 61.9 and 51.6, respectively. 154 Apparently, the two resonances in Model-B have a higher Q, which 155 results in a narrower bandwidth each. It gives Model-B the potential 156 to serve as a dual-band antenna or a single band antenna by tuning the 157 two resonances closer. Apart from the dimensions and the locations 158 of the vias, the resonant frequencies and Q factors are also influenced 159 by other parameters such as the width of SIW, the PCB thickness, 160 and the substrate characteristics. 161

In Model-A, the resonant frequency can be easily tuned by chang-162 ing the diameter of Via 1 and the size of the pad. In Model-B, Via 163 1 mainly affects lower resonance. As shown in Fig. 4(a), the lower 164 resonant frequency moves higher when the diameter of Via 1 r_{v1} 165 is larger, which corresponds to a lower inductance. Meanwhile, the 166 higher resonant frequency does not move. The influence of Via 2 is 167 more complicated because it affects both resonances. As shown in 168 Fig. 4(b), when the diameter of Via 2 (r_{v2}) is larger, both resonant 169 frequencies move higher. A similar phenomenon is observed from the 170 locations of the vias as well. In Fig. 4(c), the location of Via 1 (d_1) 171



Fig. 4: Reflection coefficients of Model-B with different via diameters and locations (a) Diameter of Via 1. (b) Diameter of Via 2. (c) Distance from Via 1 to the SIW opening edge. (d) Distance from Via 2 to Via 1.

influences the impedance matching but does not shift the resonant 172 frequencies. However, in Fig. 4(d), both impedance matching and 173 resonant frequencies change with the location of Via 2 (d_2) . It is 174 reasonable that Via 2 is more sensitive than Via 1, since it involves 175 both the coupling to Via 1 and the feeding port. However, this causes 176 great difficulties when more vias are applied in order to achieve wider 177 bandwidth by introducing more resonances, since it is very difficult to 178 find a pattern among a lot of sensitive parameters by manual tuning or 179 parameter sweeping. Hence, design exploration via machine learning-180 assisted global optimization is carried out in the next Section. 181

III. MACHINE LEARNING-ASSISTED DESIGN EXPLORATION 182

In order to achieve larger bandwidth for the proposed SIW antenna, 183 one possible solution is to add more vias inside the SIW. However, 184 the influences among multiple vias are complicated. Specifically, each 185 186 resonance is influenced by a large number of design parameters (i.e., more than 20) and each parameter could influence several resonances 187 as well. The high level of interrelation between the design parameters 188 makes the conventional experience-driven trial-and-error method and 189 parameter sweeping method infeasible. 190

Using local optimization (e.g., CST TRF) from an initial design 191 is a popular method [17], [18]. However, this is also infeasible for 192 this SIW antenna. Because the amount of inductance and capacitance 193 introduced by a metallic via is difficult to ascertain, an initial design 194 with reasonably good quality is difficult to be derived or estimated. 195 Without a good initial design, local optimization methods are less 196 likely to obtain a satisfactory design. To verify this, TRF in CST-197 MWS is used from three initial designs obtained by Latin hypercube 198 199 sampling [19] of the design space. Although a sigma value of unity 200 is used to avoid being trapped in a local optimum, the obtained designs are still trapped in local optima, whose performances are 201 far from satisfactory (Table I shows the best design obtained in three 202 runs). This shows the necessity of employing artificial intelligence 203 (AI) techniques (i.e., machine learning) for design exploration. 204

205 A. Antenna element design exploration

Global optimization is applied on the three vias model shown 206 in Fig. 1. The critical design parameters and their ranges and the 207 geometric constraints described in Table II have been considered. 208

The optimization goal is to achieve the maximum bandwidth in the frequency spectrum of 24 GHz to 40 GHz, where the maximum inband reflection coefficient is less than or equal to -10 dB, subject to a minimum in-band realized gain better than 0 dBi and a minimum in-band total efficiency better than or equal to 60%. The optimization has been implemented on the workstation described in Section II. The time reported is from a wall clock. 215



Fig. 5: Antenna element configuration and performances. (a) Antenna configuration. (b) S_{11} and realized gain. (c) Radiation pattern at 39 GHz.

PSO [20] is a metaheuristic algorithm based on the concept 216 of swarm intelligence, which is widely used in antenna design 217 exploration (e.g., [21]). Here, PSO in CST-MWS is used with a swarm 218 size of 50, a computing budget of 2,000 EM simulations; all other 219 algorithmic settings are the default settings in CST-MWS. Since each 220 run of PSO costs about one month, three runs are carried out. All 221 of them failed to obtain satisfactory designs, as shown in Table I for 222 their typical results (the best design over 3 runs). This shows the 223 necessity of introducing advanced machine learning-assisted global 224 optimization methods. 225

PSADEA [12], [13] is one of the state-of-the-art machine learning-226 assisted global optimization methods considering the design land-227 scape characteristics of antennas and arrays. It employs Gaussian 228 process supervised learning and differential evolution (DE) to predict 229 antenna performances and conduct global optimization, respectively. 230 Compared to other state-of-the-art methods in the SADEA series [22], 231 it aims at satisfying stringent design specifications for antennas with 232 less than 30 design parameters. To achieve this, the cooperative use 233 of complementary DE mutation operators, which is controlled by a 234 reinforcement learning scheme, is the main driving force; an adaptive 235 Gaussian process supervised learning scheme supporting the above 236 search scheme is also essential. More details about PSADEA can be 237 found in [12], [13]. 238

In PSADEA-based design exploration, a population size of 50 is 239 used (same as CST-MWS PSO) and the computing budget is 500 EM 240 simulations. All other PSADEA settings are the default settings in 241 [12], [13]. The convergence criterion is an improvement of less than 242 1e-3 in the objective value after 100 EM simulations. After 5 days' 243 optimization, PSADEA obtains the satisfactory design in Table II 244 using 312 EM simulations. In comparison to the best designs obtained 245 by the CST-MWS optimizers, the PSADEA-optimized design offers 246 2.1 GHz and 2.2 GHz more bandwidth compared to the best designs 247 obtained by CST TRF and CST PSO, respectively. The satisfactory
performance of the PSADEA-optimized design in terms of bandwidth, realized gain and radiation efficiency ensured the viability of
the proposed SIW endfire antenna for physical implementation (i.e.,
prototyping and fabrication).

Fig. 5(a) demonstrates the proposed antenna structure obtained 253 from the optimization. Fig. 5(b) shows the S_{11} and realized gain 254 of the proposed model. The best bandwidth is obtained from 37.5 255 GHz to 40.1 GHz, where the realized gain within the bandwidth is 256 above 3.3 dBi. The radiation patterns are stable over the operating 257 band. As an example, the realized gain at 39 GHz is 4.1 dBi and 258 the radiation pattern is shown in Fig. 5(c). The cross-polarization is 259 below -20 dB in the radiation direction for all frequencies. 260



Fig. 6: 4-element array configuration.



Fig. 7: Array performances. (a) Reflection coefficients and mutual coupling. (i,j = $\{1, 2, 3, 4\}$.) (b) Beam scanning pattern at 39 GHz.

It is worth noting that further increasing the number of vias 261 can stimulate more resonances but the benefit on the bandwidth is 262 limited. For one thing, it is because the Q value of each resonance 263 264 increases and for another thing, the interference between the vias prevents some of the resonances to form a wide band. As an example, 265 the best bandwidth of a model with five vias is 7.2%, which has 266 little advantage compared to the proposed model with three vias 267 considering the complexity of the structure. 268

269 B. Array performances

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The proposed antenna is applied in a 4-element array. The inside 270 configuration is shown in Fig. 6. The ground plane size of the array 271 is 21.2 mm \times 37.8 mm. Fig. 7(a) presents the reflection coefficients 272 of all the four array elements and the mutual coupling between the 273 adjacent elements. The array elements maintains the same bandwidth 274 as the single antenna from 37.5 GHz to 40 GHz and the mutual 275 coupling is below -18 dB within the bandwidth. Fig. 7(b) shows the 276 beam scanning pattern at 39 GHz. The scanning range is from -20° 277 to 35° and the realized gain ranges from 5.62 dBi to 9.25 dBi. 278

IV. MEASUREMENTS AND RESULTS

The proposed antenna and the 4-element array are fabricated and measured. The front and back photographs of the fabricated single antenna and array are shown in Fig. 8. The antennas are fed with MMPX connectors mounted on the back side of the PCB. The antenna

TABLE I: (Previous Table II) Optimization results of the synthesized SIW Endfire Antenna (PSADEA, TRF, PSO).

Results	Penalty Weight	PSADEA	TRF	PSO
Bandwidth (GHz)	1	37.5 - 40.0	27.3 - 27.7	35.2 - 35.5
Minimum In-Band Realized Gain (dBi)	50	3.3	2.7	2.6
Minimum In-Band Total Efficiency	50	60%	72%	55%

TABLE II: (Previous Table I) Search space for the structure parameters and optimal values obtained by PSADEA (units in mm). All parameters are continuous variables, with the exceptions of h_1 and h_3 , which take values in $\{0.422, 0.508, 0.762\}$ and $\{0.101, 0.168, 0.254, 0.338, 0.422, 0.508\}$, respectively. h_1 and h_3 must satisfy the constraint $h_1 + h_2 + h_3 \leq 1.2$. r_i and r_f must satisfy the constraint $r_f \leq r_i - 0.1$.

DESCRIPTION	PARAM.	VALUES	Optimum
Spacing: via 1 / opening	d_1	[0.0, 1.5]	0.27
Spacing: via 2 / via 1	d_2	[0.0, 3.0]	0.97
Spacing: via 3 / via 2	d_3	[0.0, 7.0]	0.36
Spacing: feed / end	f_d	[1.5, 3.0]	1.58
Spacing: via 3 / feed	l_{ex}	[0.0, 4.0]	1.16
OFFSET: VIA 1 FROM CENTER	of_1	[-0.5, 0.5]	0.00
OFFSET: VIA 2 FROM CENTER	of_2	[-0.5, 0.5]	0.45
OFFSET: VIA 3 FROM CENTER	of_3	[-0.5, 0.5]	-0.35
PAD WIDTH OF VIA 1	p_1	[0.0, 1.0]	0.12
PAD WIDTH OF VIA 2	p_2	[0.0, 2.0]	1.51
PAD WIDTH OF VIA 3	p_3	[0.0, 2.0]	0.90
DIAMETER: SOLDER PAD	r_i	[0.3, 0.5]	0.45
DIAMETER: FEED PIN	r_{f}	[0.1, 0.3]	0.17
DIAMETER: FEED PAD	r_{fin}	[0.0, 0.5]	0.37
DIAMETER: VIA 1	r_{v1}	[0.1, 0.5]	0.25
DIAMETER: VIA 2	r_{v2}	[0.1, 0.5]	0.25
DIAMETER: VIA 3	r_{v3}	[0.1, 0.5]	0.23
SIW WIDTH	w	[5.4, 6.6]	5.94
THICKNESS: BOTTOM SUB.	h_1	$\{0.422, \ldots, 0.762\}$	0.508
THICKNESS: MIDDLE SUB.	h_2	$\{0.200\}$	0.200
THICKNESS: TOP SUB.	h_3	$\{0.101, \ldots, 0.508\}$	0.338

S-parameters are measured with the PNA vector network analyzer and 284 the radiation patterns are measured in the far-field anechoic chamber. 285 Fig. 9 shows the radiation pattern measurement setup. The under-test 286 antennas are mounted on a rotational plate, which covers 240° of the 287 Θ plane from -120° to 120° . The remaining 120° area cannot be 288 measured due to the blockage of the absorber behind the antennas. 289 Each antenna is measured with two setups. As shown in Fig. 9(a), 290 the measurement catches the radiation pattern in YoZ plane (rotation 291 in Θ direction) and in Fig. 9(b), the measurement catches the XoY 292 plane. In the array measurements, the non-tested ports are terminated 293 with 50 Ω loads. 294

Fig. 10(a) shows the simulated and measured S-parameters and 295 realized gain of the single antenna. The three resonances in the 296 simulation are observed in the measurements with a slightly fre-297 quency shift to the lower band. The measured bandwidth is from 298 36 GHz to 40.6 GHz with reflection coefficients below -8.8 dB. The 299 discrepancy between the simulation and measurement is mainly due 300 to the fabrication errors at the locations of the vias. The via location 301 uncertainty is ± 0.075 mm, which is large enough to cause an obvious 302 frequency shift. By applying the location errors on the blind vias 303 and the SIW, the resonant frequencies become more similar to the 304



Fig. 8: Fabricated antenna and 4-element array. (a) Single antenna. (b) 4-element array. (c) Radiation pattern at 39 GHz.



Fig. 9: Radiation pattern measurement setup. (a) measurement in YoZ plane. (b) measurement in XoY plane.

measurement but with a worse impedance matching (that may caused
by other minor location errors during the fabrication). The measured
realized gain ranges from 1 dBi to 5.3 dBi within the operating band.
Correspondingly, the gain curve of the simulation considering errors
is also more similar to the measurement.

The simulated and measured results of the array are shown in 310 Fig. 10(b). Since the four array elements have very similar perfor-311 mances in the simulations, only S_{11} and S_{21} are demonstrated for 312 comparing with the measurements. For the measurement results, the 313 S_{22} of array element P2 has a slightly higher reflection compared 314 to S_{11} of array element P1. The bandwidth of the array elements 315 is from 36 GHz to 40.7 GHz with reflection coefficients below -8 316 dB, which is similar to the result of the single antenna. The mutual 317 coupling between array elements P1 and P2 is below -22 dB in the 318 measurement. The measured realized gain of array element P1 and 319 P2 ranges from 1.4 dBi to 4.9 dBi and from 0.5 dBi to 5.2 dBi, 320 respectively. The results of array elements P3 and P4 are similar to 321 the results of the array elements P1 and P2 due to the symmetry of 322 the array. Note that these are not shown to ensure brevity. 323

The radiation patterns of the single antenna at 39 GHz are shown in Fig. 11(a) and Fig. 11(b). The measured realized gain is 4.27 dBi and the radiation pattern is in good agreement with the simulations. The simulated and measured cross-polarization levels are both below -20 dB. Fig. 11(c) and Fig. 11(d) are the radiation patterns of array element P2 at 39 GHz. The measured realized gain is 3.7 dBi and the cross-polarization is below -20 dB.

Table III gives the comparison of the proposed antenna with the other mm-wave endfire antennas. [2] proposed a design of a quasiyagi antenna with a driven dipole and a director. [3] proposed a dipole



Fig. 10: Simulated and measured S-parameters and realized gain. (a) Reflection coefficients and realized gain of the single antenna. (b) Reflection coefficients, mutual coupling, and realized gain of array element P1 and P2.

antenna with the arms tilted to the ground plane in order to reduce 334 the clearance and enhance the bandwidth (BW). References [4], [6], 335 [10], [11] proposed SIW open-ended antennas on a thin PCB. In 336 [4], the impedance matching is achieved by adding two symmetric 337 metallic vias close to the SIW aperture. In [6], two rows of transition 338 plates are adopted to improve the impedance matching. In [10], the 339 SIW antenna has a tapered ridge inside the SIW for achieving a 340 wideband impedance matching. In [11], the SIW open-ended antenna 341 is air-filled fed with a tapered microstrip line obtaining a narrow 342 band impedance matching. The antennas in the references [2]-[4], 343 [6] and the proposed work have an inter-element distance of nearly 344 half wavelength, which is capable of being used in a beam-steerable 345 array for 5G mobile devices. The antennas in [10] and [11] have an 346 wide opening width of more than $3\lambda_0$. Note that the wide opening is 347 also an important factor in improving the impedance matching and 348 the bandwidth, but it strongly limits the scanning range. The proposed 349 antenna has achieved zero clearance, which is the lowest compared 350 to the other works. Besides, the proposed design has a reasonable 351 bandwidth, which is comparable to the low-profile SIW open-ended 352 antennas in [6] and the Yagi antenna in [2]. The proposed antenna 353 also achieves a reasonable peak gain of 5 dBi and a small thickness 354 of nearly $0.1\lambda_0$. 355

V. CONCLUSIONS

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A novel design of a zero clearance SIW endfire antenna array for 357 5G mobile applications is proposed. The array element offers 12% 358 bandwidth and a peak gain of 5.3 dBi with zero clearance and very 359 low profile of 1 mm (0.13 λ_0 at 38 GHz). The beam steering of 360 the proposed array covers from -20° to 35° at 39 GHz. The very 361 compact size makes the proposed antenna very suitable for mobile 362 applications. Moreover, the design is achieved by using a state-363 of-the-art machine learning-assisted antenna optimization method, 364

TABLE III: Comparison of the proposed antenna against other mm-wave endfire antennas.

References	Antenna Type	Frequency (GHz)	BANDWIDTH	Peak Gain (dBi)	THICKNESS	CLEARANCE	BEAM SCANNING
Hwang, et al. (2019) [2]	YAGI	28	12.3%	5.5	$0.07\lambda_0$	$0.23\lambda_0$	YES
Syrytsin, et al. (2018) [3]	DIPOLE	29	27.6%	5	$0.15\lambda_0$	$0.15\lambda_0$	YES
Li, et al. (2021) [4]	SIW	27	18.9%	3.4	$0.19\lambda_0$	$0.16\lambda_0$	YES
Lu, et al. (2020) [6]	SIW	26	15.4%	9	$0.09\lambda_0$	$0.61\lambda_0$	YES
Mallahzadeh, et al. (2012) [10]	SIW	29	75.9%	10	$0.25\lambda_0$	0	No
Mateo, et al. (2016) [11]	SIW	15	1.4%	8	NOT GIVEN	0	No
This work	SIW	38	12.0%	5	$0.13\lambda_0$	0	YES



Fig. 11: Simulated and measured radiation pattern of at 39 GHz. (a) Single antenna in XoY plane. (b) Single antenna in YoZ plane. (c) Array element P2 in XoY plane. (d) Array element P2 in YoZ plane.

PSADEA. In comparison to conventional and widely used antenna optimization methods, it shows clear advantages in terms of both efficiency and design solution quality, thanks to the machine learningassisted optimization techniques. The optimization results also reveal the potential of the SIW endfire antenna in terms of the balance between the largest achievable bandwidth and the complexity of the topological profile of the structure.

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