Additively Manufactured Amplitude Tapered Slotted Waveguide Array with Horn Aperture for 77 GHz

KONSTANTIN LOMAKIN1, SAIF ALHASSON2 AND GERALD GOLD.3

1Institute of Microwaves and Photonics, Friedrich-Alexander University Erlangen-Nuremberg, Erlangen, Bavaria, Germany (e-mail: konstantin.lomakin@fau.de)
2NXP Semiconductors Germany GmbH, Munich, Bavaria, Germany (e-mail: saif.alhasson@nxp.com)
3Institute of Microwaves and Photonics, Friedrich-Alexander University Erlangen-Nuremberg, Erlangen, Bavaria, Germany (e-mail: gerald.gold@fau.de)

Corresponding author: Konstantin Lomakin (e-mail: konstantin.lomakin@fau.de).

The reported results were obtained in the course of the project 3DKonform (AZ-1362-18), funded by the BFS (Bavarian Research Foundation)

ABSTRACT In this work, a combined frontend of horn and slotted waveguide array antenna serving as feeding subsystem is presented. This setup allows to implement an amplitude taper in one plane while using the horn aperture to shape the far field radiation characteristics in the perpendicular plane independently and tailored to the individual demand of the application. Moreover, the feeding system is implemented as a loop resulting in a standing wave distribution for a wide frequency range in order to increase the antenna bandwidth. Two specimens with different horn apertures are additively manufactured in slotted waveguide technology from UV curable photopolymer resin using a digital light processing 3D printer and subsequently metal coated by electroless silver plating. A relatively large bandwidth of $B_{-10\text{dB}}=7.2$ GHz and a $B_{-15\text{dB}}$ of 4.9 GHz is achieved in measurements indicating a relative bandwidth of $9\%$ and $6\%$ respectively. Antenna Gain is characterized as 17.5 and 20.17 dBi in measurements while maintaining a sidelobe level of 20-22.5 dB at 77 GHz and exhibiting a radiation efficiency of 78.9 and 79.4 %.

The proposed antenna architecture provides a flexible approach for beam shaping in applications where azimuth and elevation planes exhibit different or contrary requirements, e.g. broad vs. narrow field of view - especially suitable for automotive MIMO-radar sensors.

INDEX TERMS additive manufacturing, 3D printing, slotted waveguides, electroless plating

I. INTRODUCTION

Technological trends, such as autonomous driving, imaging radar sensors or next generation of MIMO communication standards provide a wide range of applications for mm-Wave components and systems. While antennas play an important role as interface to the free space, their design typically becomes challenging with increasing frequencies and large array setups. In addition to the anyhow nontrivial antenna engineering, limitations of conventional manufacturing technologies become an additional bottleneck with respect to the achievable system performance and design effort. In mass production, printed circuit boards (PCB) are often preferred, though suffering from high dielectric loss, surface roughness, manufacturing tolerances and electromagnetic compatibility (EMI). Moreover, such antennas are limited to a planar layout although the desired antenna geometry may generally require the design freedom of all three dimensions.

While waveguide systems provide an alternative solution, they suffer from high weight, complex manufacturing and assembly involving CNC milling. In addition, with increasing system complexity, the typically utilized split block assembly [1] could turn into a complex puzzle.

Additive manufacturing presents an auspicious alternative as it combines an inherent geometrical freedom with relatively cost efficient machinery and manufacturing. Among a wide range of different 3D printing concepts available on the market, stereolithography (SLA) and digital light processing (DLP) offer the required geometrical precision in the 50-100 μm range while at the same time, yielding a decent surface quality in the order of $R_q < 1$ μm (RMS-surface roughness)[2], which is mandatory for the mm-Wave frequency range [3]. Combining the printed
By making use of the geometrical freedom of 3D-printing, in this work, a modified slotted waveguide array antenna (SWA) with 20 radiating elements is proposed, that allows for relatively independent shaping of the azimuth and elevation beam pattern without requiring a 2D array. While combinations of horn apertures and waveguide feeding subsystems have been proposed, e.g., in [13], [14], the approach proposed in this work provides the unique opportunity to implement an amplitude taper where the individual amplitudes are controlled by geometrical displacements of slots on the SWA subsystem instead of a waveguide distribution network [14] and the distance between the slots can be kept at $\lambda_g/2$ instead of $\lambda_g$ as it is the case in [13]. Moreover, the proposed antenna architecture is additively manufactured as a monolithic part without the need of further assembly. Such antennas could be especially useful in applications with different or contrary requirements towards azimuth and elevation, which is typically the case in automotive radar sensors.

Two different variations of the proposed architecture are elaborated, manufactured and characterized in order to illustrate the obtained benefit of the proposed design methodology and fabrication process.

**II. PROPOSED ANTENNA DESIGN**

The general idea of the proposed antenna architecture is shown in Figs. 1 and 2 and implements a horn antenna aperture which is fed by 20 radiating elements of an amplitude tapered slotted waveguide array antenna (SWA). Both models present different antenna designs where Design A in Fig. 1 exhibits a narrow horn section with an opening of $w_{Horn,A} = 2.2 \text{ mm}$ aiming for a relatively wide beam pattern in the azimuth plane - which is perpendicular to the array. In contrast thereto, Design B in Fig. 2 exhibits a wide opening of $w_{Horn,B} = 5.5 \text{ mm}$, hence, featuring a relatively narrow beam in azimuth plane.

Fig. 3 provides a cross section view on antenna B showing the parameters referred to in the following. The radiating slots of the feeding SWA (P4) are used to feed the horn aperture in both designs. In contrast to conventional solutions for tapering horn antenna arrays in order to realize a non-homogeneous amplitude configuration, by e.g., implementing complex power distribution networks, in this work, an amplitude taper is applied to the feeding SWA. Consequently, the horn aperture is supplied with an individually shaped power distribution which allows for optimizing the sidelobe level (SLL) in the plane along the feeding SWA (elevation), while leaving the perpendicular plane (azimuth) unchanged. Latter one is in turn mainly defined by the dimensions of the horn aperture (width $w_{Horn}$ and length $l_{Horn}$). Hence, a quasi-horn-antenna-array is achieved, where the radiated power of each horn element
Both antennas are implemented using the slotted waveguide technique, which is described in detail in [2], [6], [16]–[18] where non-radiating slots are introduced into the narrow wall of the waveguide as shown in Fig. 4 including the design parameters. Therein, \( s = 1 \text{ mm} \) represents the periodicity of the slots, while \( d = 400 \mu \text{m} \) depicts the remaining post thickness and \( g = s - d = 600 \mu \text{m} \) being the size of the introduced slots. This way, DLP printing as well as metal plating can be improved even for relatively complex waveguide designs as compared to continuous wall waveguides. For the propagating electromagnetic (EM) wave, the slotted wall behaves almost identically to a continuous wall as long as the design rule established for substrate integrated waveguides (SIW) relating the slot periodicity to the guided wavelength \( s < \lambda_g/4 \) is met.

While the impact of these slots plays merely a negligible role in broadband components, e.g., waveguide distribution networks, horn antennas or couplers, in case of the 20 elements antennas as proposed in this work, their impact is quite significant. Similar to SIW, the guided wavelength decreases with the introduced slots as shown in Fig. 5 for simulated WR12 waveguides with and without slots. Therein, \( \lambda_g \) drops by approximately 100\( \mu \text{m} \) from 5.0 to 4.9\( \text{mm} \) at 77\( \text{GHz} \) when slots of \( g = 600 \text{\( \mu \text{m} \)} \) are introduced. When the radiating slots on the broad wall (P4) of the SWA feeding the horn are separated by \( \lambda_g/2 \) along the propagation direction of the waveguide, a deviation of 50\( \mu \text{m} \) is obtained as a consequence in contrast to the continuous wall waveguide. Although this value remains within typical tolerances of DLP printers, with 20 radiating elements, this deviation is accumulated over the antenna length yielding an error of 1\( \text{mm} \), hence pushing the outer elements far off their intended phase location of the underlying EM field distribution in the feeding waveguide.

To account for this effect, in this work, the longitudinal spacing of the radiating slots is adapted to \( \lambda_g/2 = 4.88 \text{\( \mu \text{m} \)}/2 = 2.44 \text{\( \mu \text{m} \)} \) according to the underlying field distribution with its guided wavelength at 77\( \text{GHz} \).

**III. GUIDED WAVELENGTH IN SLOTTED WAVEGUIDE SYSTEMS**

Both antennas are implemented using the slotted waveguide technique, which is described in detail in [2], [6], [16]–[18] where non-radiating slots are introduced into the narrow wall of the waveguide as shown in Fig. 4 including the design parameters. Therein, \( s = 1 \text{ mm} \) represents the periodicity of the slots, while \( d = 400 \mu \text{m} \) depicts the remaining post thickness and \( g = s - d = 600 \mu \text{m} \) being the size of the introduced slots. This way, DLP printing as well as metal plating can be improved even for relatively complex waveguide designs as compared to continuous wall waveguides. For the propagating electromagnetic (EM) wave, the slotted wall behaves almost identically to a continuous wall as long as the design rule established for substrate integrated waveguides (SIW) relating the slot periodicity to the guided wavelength \( s < \lambda_g/4 \) is met.

While the impact of these slots plays merely a negligible role in broadband components, e.g., waveguide distribution networks, horn antennas or couplers, in case of the 20 elements antennas as proposed in this work, their impact is quite significant. Similar to SIW, the guided wavelength decreases with the introduced slots as shown in Fig. 5 for simulated WR12 waveguides with and without slots. Therein, \( \lambda_g \) drops by approximately 100\( \mu \text{m} \) from 5.0 to 4.9\( \text{mm} \) at 77\( \text{GHz} \) when slots of \( g = 600 \text{\( \mu \text{m} \)} \) are introduced. When the radiating slots on the broad wall (P4) of the SWA feeding the horn are separated by \( \lambda_g/2 \) along the propagation direction of the waveguide, a deviation of 50\( \mu \text{m} \) is obtained as a consequence in contrast to the continuous wall waveguide. Although this value remains within typical tolerances of DLP printers, with 20 radiating elements, this deviation is accumulated over the antenna length yielding an error of 1\( \text{mm} \), hence pushing the outer elements far off their intended phase location of the underlying EM field distribution in the feeding waveguide.

To account for this effect, in this work, the longitudinal spacing of the radiating slots is adapted to \( \lambda_g/2 = 4.88 \text{\( \mu \text{m} \)}/2 = 2.44 \text{\( \mu \text{m} \)} \) according to the underlying field distribution with its guided wavelength at 77\( \text{GHz} \).

**IV. AMPLITUDE TAPER OF FEEDING SWA**

![FIGURE 6: Front view on the 20 radiating elements of the SWA](image)

The elevation plane beam characteristic of the horn aperture is defined by the feeding SWA subsystem below the horn front end. In this work, the radiation amplitudes of the 20 radiating elements are arranged according to a \( \cos^2 \)-amplitude taper in order to achieve a trade-off between relatively high gain > 16\( \text{dB} \), high sidelobe level > 25\( \text{dB} \) and still moderate gain close to the main lobe. The amplitude of each element \( A_i \) with a length of \( L_s = 1.9 \text{\( \text{mm} \)} \) is therefore obtained by:

\[
A_i = A_{\text{min}} + (1 - A_{\text{min}}) \left( \sin \left( \frac{2\pi (i - 1)}{4(N - 1)} \right) \right)^2
\]  

(1)
where $A_{\text{min}} = 0.1$ is added to avoid the amplitude of the first and last element to drop to 0.

The individual amplitude $A_i$ of each radiating element of the SWA can be controlled by its geometrical displacement $x_i$ on the broad wall from the center line of the waveguide as shown in Fig. 6, since it is related to the transversal current density distribution:

$$J_i(x) = \hat{J} \sin \left( \frac{\pi x}{w} \right)$$

with $\hat{J}$ representing the maximum transversal current density amplitude, $x$ being the displacement and $w$ the waveguide cross section width. Resolving (2) finally allows to obtain the geometrical displacement $x_i$ of each slot according to its desired amplitude $A_i$:

$$x_i = w_{\text{max}} \arcsin \left( A_i \right)$$

with $w_{\text{max}}$ defining the maximum admitted displacement which in turn is used as a parameter for simulation sweeps in order to optimize the antenna return loss. Hence, this value depends on the interaction between the feeding SWA and the horn aperture and is therefore optimized separately for both antennas due to their different horn apertures. Finally the amplitudes and residual displacement values for both antennas considered in this work are provided in tables 1 and 2 respectively.

### V. Optimization Methodology

Both antenna designs, A and B, are optimized in CST Microwave Studio where a complete CAD model of each antenna is created except of the flange geometry in order to keep the simulation space as compact as possible. The radiating slot positions along the waveguide are defined according to the corrected half guided wavelength $\lambda_g / 2$ from Fig. 5 and placed in an alternating manner with respect to the waveguide center line on the broad wall in order to account for the phase difference of $180^\circ$ within $\lambda_g / 2$ and obtain constructive interference of adjacent radiating elements. While the length of each slot $L_n = 1.9 \, \text{mm}$ is identical for both antenna designs and tuned by full-wave simulation, the width $w_{\text{slot}}$ and the maximum displacement $x_{10}$ and $x_{11}$ depend on the dimensions of the respective horn aperture, $l_{\text{Horn}}$ and $w_{\text{Horn}}$. With parameter sweep simulations, the slot widths yielding a return loss $< -15 \, \text{dB}$ within the targeted frequency range are found as $w_{\text{Slot,A}} = 0.4 \, \mu\text{m}$ for design A and $w_{\text{Slot,B}} = 0.5 \, \mu\text{m}$ for design B, with the maximum displacement presented in Tabs. 1 and 2 respectively.

![FIGURE 7: Manufactured narrow horn specimen (Design A)](image7.png)

![FIGURE 8: Manufactured wide horn specimen (Design B)](image8.png)

### VI. Fabricated Specimens and Characterization

The manufactured narrow and wide horn specimens are shown in Figs. 7 and 8 respectively. In order to improve printing quality, the solid parts without relevance to the EM-fields are perforated by drain holes which are oriented along the printing axis to reduce parasitic adhesion effects during the printing process. Additionally, a flange design as already used in [17], [20] is added for measurement purpose. The manufactured specimens were characterized on a single axis turntable setup as shown in Fig. 9. A reference horn antenna [21] with known gain is placed in a distance of 950 mm as demonstrated in Fig. 10. In order to scan the beam pattern in azimuth and elevation, the antenna under test and the reference antenna are rotated by $90^\circ$ around the main lobe axis, hence, azimuth and elevation planes are exchanged while using the same rotation axis of the turntable. Both antennas are attached to E-Band (WR12) frequency converters operated at a VNA. A TOSM waveguide standard algorithm has been performed to calibrate the reference planes towards the interface between the antennas and the respective test port to improve measurement precision.

Fig. 11 shows the simulated and measured return loss $|S_{11}|^2$ of the antenna A. Although a deviation between simulation and measurement is observed in the region below $-15 \, \text{dB}$, still, a $B_{-15\,\text{dB}}$ of 4.9 GHz from 74.8 to 79.7 GHz and a $B_{-10\,\text{dB}}$ of 7.2 GHz is achieved, with latter ranging from 73.0 to 80.2 GHz. While the deviations could originate

### TABLE 1: Offset values for Design A

| $|A_i|$ | x_1 (µm) | 1&20 | 2&19 | 3&18 | 4&17 | 5&16 |
|-------|----------|-----|-----|-----|-----|-----|
| 0.1   | 48       | 61  | 100 | 160 | 230 |
| 0.1   | 86       | 127 | 205 | 325 | 472 |

### TABLE 2: Offset values as for Design B

| $|A_i|$ | x_1 (µm) | 1&20 | 2&19 | 3&18 | 4&17 | 5&16 |
|-------|----------|-----|-----|-----|-----|-----|
| 0.1   | 48       | 61  | 100 | 160 | 230 |
| 0.1   | 86       | 127 | 205 | 325 | 472 |
from manufacturing tolerances of such a complex model and due to the manually performed metal plating process, the broadband characteristics of the loop feed still allow to provide an operational antenna within the band of interest.

Fig. 12 shows the measured and simulated return loss of antenna B. A $B_{-15\, \text{dB}}$ of 3.5 GHz and a $B_{-10\, \text{dB}}$ of 5.6 GHz is achieved. Even a $B_{-20\, \text{dB}}$ of 1.5 GHz at 75.3-76.8 GHz can be specified, highlighting the auspicious possibilities of additive manufacturing in the mmWave range, when keeping in mind the relatively high level of geometrical complexity.

Due to the broadband and symmetric loop feed, the gain maintains a relatively constant magnitude over a wide frequency range, dropping by 3 dB from the peak within a bandwidth of $B_{3\, \text{dB}} \approx 8.8$ and 9.5 GHz within 71.5-80.3 GHz and 71-80.5 GHz, respectively for both antenna designs A and B as shown in Fig. 13. The measured cross-polarization gain over frequency in main-lobe direction for both antennas is plotted in Fig. 14, which is approximately 25 dB below the measured co-polarization gain and even 30 dB below latter between 77 and 79 GHz.

Figs. 15 and 16 show the measured and simulated beam patterns with absolute gain levels for both antennas in azimuth and elevation planes respectively at 77 GHz. For design A, a measured main lobe gain of 17.5 dBi is achieved, while back lobes beyond $\pm 120^\circ$ do not exceed 1.9 dBi in simulation which is not shown here. In elevation plane - where the amplitude taper is aimed to achieve a SLL of >40 dB in simulation - 20.5-22.5 dB were achieved, probably due to limited resolution of the printer and therefore, discretization of the position values in Tab. 1. In azimuth, a HPBW of 71° is achieved in contrast to simulated 67°. In contrast to the gain, the simulated directivity yields $D_{\text{sim,A}} = 18.5 \, \text{dBi}$. Hence, the achieved radiation efficiency of the manufactured design A reaches $\eta_A \approx 79.4\%$ from the relation between measured gain and $D_{\text{sim}}$ - both in linear magnitude - as proposed in [22].

The measured beam characteristics of antenna B yield a main lobe gain of 20.17 dBi at 77 GHz with a deviation from simulation by as little as 0.13 dBi while achieving
an SLL of 20-22.2 dB, hence, similar to design A. Back lobes beyond ±120° do not exceed 2.2 dB in simulation. The simulated directivity yields $D_{\text{sim},B} = 21.2$ dBi hence, a radiation efficiency $\eta_B \approx 78.9\%$ is achieved in design B.

The frequency shift and increased loss figure due to surface roughness are considered early in simulation by making use of the Gradient Model [3], [23], [24] which takes into account the impact of surface roughness on both, attenuation and phase of propagating EM waves. Based on experience from previous work, the surface roughness was set to a rather conservative value of $R_q = 1$ µm and bulk conductivity $\sigma_{\text{DC}} = 10$ MS/m in simulation. Furthermore, the main intention to achieve different azimuth radiation characteristics by adapting the horn aperture geometry while remaining the SLL in elevation in a similar magnitude - i.e., 20-22.5 dB in both designs A and B - is successfully demonstrated.

The radiation efficiency of different array antennas in literature spans a wide range. Up to 88.6 dB are reported in [25] with an $8 \times 8$ elements patch antenna array at 35 GHz in simulation. A simulation study of a substrate integrated waveguide slot array antenna exhibiting $3 \times 5$ sub-arrays each implementing 8 radiating slots yields approximately 80 % at 79 GHz. In contrast thereto, the efficiency of a manufactured $8 \times 8$ slot array based on gap technology in [26] achieves 65 % at 60 GHz and the additively manufactured $10 \times 10$ slotted waveguide array in [10] reaches about 66 %. Although the designs proposed in this work achieve less gain than both
latter manufactured antennas with a gain of 26 and 27.3 dBi, respectively, the antenna design B still yields 20.17 dBi while implementing an amplitude taper - hence, reducing the gain - and a single-column design only. Both of the proposed and manufactured antennas exhibit a radiation efficiency above 78%.

VII. CONCLUSION

In this work, a combination of horn aperture and a SWA feed is proposed, which allows to independently shape the radiation characteristics in azimuth and elevation plane respectively. The loop feed design increases the bandwidth of the proposed architecture up to 9% of relative bandwidth with a measured -10 dB-Bandwidth of \( B_{-10\,\text{dB}} = 7.2\,\text{GHz} \) and a -15 dB-Bandwidth of \( B_{-15\,\text{dB}} = 4.9\,\text{GHz} \) at a center frequency of \( f = 77\,\text{GHz} \), which is quite extraordinary for such a kind of antenna which typically exhibits relative \( B_{-10\,\text{dB}} \leq 5\% \).

Two specimens with different horn apertures are additively manufactured in slotted waveguide technology from UV curable photopolymer resin using a digital light processing 3D printer and subsequently metal coated by electroless silver plating. The measured gain is characterized as 17.5 and 20.17 (dBi) for design A and B respectively while both designs maintain an SLL of 20–22.5 dB at 77 GHz and realizing a radiation efficiency of 79.4 and 78.9 %. The proposed antenna architecture provides a flexible approach for beam shaping in applications where azimuth and elevation planes exhibit different or contrary requirements, especially in automotive MIMO–radar sensors. Future effort will be dedicated to improve the radiation pattern of the proposed individual horn design.

ACKNOWLEDGMENT

The reported results were obtained in the course of the project 3DKonform (AZ-1362-18), funded by the BFS (Bavarian Research Foundation).

REFERENCES

[1] P. J. Bruneau, H. D. Janzen, and J. S. Ward, “Machining of terahertz split-block waveguides with micrometer precision,” in 2008 33rd International Conference on Infrared, Millimeter and Terahertz Waves, 2008, pp. 1–2. DOI: 10.1109/ICMWT.2008.4665435.

[2] K. Lomakin, S. Herold, D. Simon, M. Sippel, A. Sion, M. Vossiek, K. Helmreich, and G. Gold, “3D Printed Slotted Waveguide Array Antenna for Automotive Radar Applications in W-Band,” in Proceedings of the 48th European Microwave Conference (EuMW 2018), 2018.

[3] K. Lomakin, D. Simon, M. Sippel, G. Gold, K. Helmreich, E. Seler, Z. Tong, and R. Reuter, “3D Printed Slotted Waveguide Array Antenna for Automotive Radar Applications in W-Band,” in Proceedings of the 48th European Microwave Conference (EuMW 2018), 2018.

[4] A. Kantemur, Y. Sharma, J. Tak, and H. Xin, “A w-band slotted waveguide antenna based on 3d printing technology,” in 12th European Conference on Antennas and Propagation (EuCAP 2018), 2018, pp. 1–3. DOI: 10.1049/cp.2018.0974.

[5] L. Klein, K. Lomakin, M. Sippel, K. Helmreich, and G. Gold, “Additively manufactured six-port for mm-wave applications,” in 2020 50th European Microwave Conference (EuMC), 2021, pp. 384–387. DOI: 10.23919/EuMC48046.2021.9338044.

[6] K. Lomakin, M. Sippel, I. Ullmann, K. Helmreich, and G. Gold, “3D Printed Helix Antenna for 77 GHz,” in European Conference on Antennas and Propagation, 2020.

[7] N. M. Ridler, and S. Lucyszyn, “3-D Printed Metal-Pipe Waveguides with micrometer precision,” in 2008 33rd International Conference on Infrared, Millimeter and Terahertz Waves, 2008, pp. 1–2. DOI: 10.1109/ICMWT.2008.4665435.

[8] K. Lomakin, D. Simon, M. Sippel, G. Gold, K. Helmreich, and G. Gold, “3D Printed E-Band Hybrid Coupler,” IEEE Microwave and Wireless Components Letters, pp. 1–1, 2019.

[9] K. Lomakin, M. Sippel, K. Helmreich, and G. Gold, “Design and analysis of 3d printed slotted waveguides for d-band using stereolithography and electroless silver plating,” in 2020 IEEE/MTT-S International Microwave Symposium (IMS), 2020, pp. 177–180. DOI: 10.1109/IMS30576.2020.9223819.

[10] K. Lomakin, M. Sippel, and G. Gold, “Design of pyramidal horn antennas for 3d printing in the mm-wave range,” in 2022 16th European Conference on Antennas and Propagation (EuCAP), 2022.

[11] Y. Huang, “Radiation efficiency measurements of small antennas,” in Handbook of Antenna Technologies, Z. N. Chen, D. Liu, H. Nakano, X. Qing, and T. Zwick, Eds. Singapore: Springer Singapore, 2016, pp. 2165–2189, ISBN: 978-981-4560-44-3. DOI: 10.1007/978-981-4560-44-3_71. [Online]. Available: https://doi.org/10.1007/978-981-4560-44-3_71.

[12] G. Gold and K. Helmreich, “A physical model for skin effect in rough surfaces,” Microwave Integrated Circuits Conference (EuMIC), 2012 7th European, 2012.

[13] G. Gold, Modellierung rauher Oberflächen und Materialcharakterisierung für den Entwurf von Leiterplatten für Hochfrequenzanwendungen, Dissertation, 2015.
[25] M. Zhao, S. Chai, K. Xiao, and X. Zhou, “Design of millimeter-wave compact waveguide slot coupled microstrip array antenna,” in 2018 IEEE 3rd International Conference on Signal and Image Processing (ICSIP), 2018, pp. 516–519. DOI: 10.1109/ SIPROCESS.2018.8600430.

[26] A. Vosough and P.-S. Kildal, “V-band high efficiency corporate-fed 8×8 slot array antenna with etsi class ii radiation pattern based on gap technology,” in 2016 IEEE International Symposium on Antennas and Propagation (APSURSI), 2016, pp. 803–804. DOI: 10.1109/APSURSI.2016.7596110.

KONSTANTIN LOMAKIN received the B.Sc. Degree in electrical engineering from Friedrich-Alexander University Erlangen-Nürnberg (FAU) in 2013, and the M.Sc. degree in information and communication technology from FAU in 2015. He is a member of the Microwave Assembly and Interconnects Group with the Institute of Microwaves and Photonics, FAU. His research interests include signal integrity, EM-modeling and 3-D-printed mm-Wave components.

SAIF ALHASSEN was born in Baghdad, Iraq, in 1981. He received the B.Sc. and M.Sc. degrees in electronic and communication engineering from the University of Baghdad, Baghdad, Iraq, in 2002 and 2005, respectively. He was awarded a full DAAD doctoral scholarship to pursue the Ph.D. degree in conformal microwave components. He received the Ph.D. degree from Technische Universität Braunschweig, Braunschweig, Germany, in 2014. He worked as a Research Assistant with Microwave Group, Technische Universität Braunschweig. Currently, he works with Pro-micron Company, Kaufbeuren, Germany, among engineering team specialized in wireless sensor technologies for measuring temperature and strain in harsh environment, e.g., very high temperature or within objects moving at high speed. His research interests include design of passive RF circuit components, antennas, and electromagnetic sensors, as well as digital signal processing for FMCW radar technology used in Pro-micron Readers for passive SAW delay line and resonator sensors.

GERALD GOLD received the Diploma degree in mechatronics and the Dr. Eng. Degree from Friedrich-Alexander-University Erlangen-Nürnberg (FAU) in 2009 and 2016, respectively. Since 2010, he has been a Research Assistant with the Institute of Microwaves and Photonics, FAU. He became the Group Leader of the Microwave Assembly and Interconnects Group in 2018. His current research interests include 3-D-printed microwave components, electromagnetic interaction with non-ideal surfaces, and automation of RF measurements.