Gain Enhancement of Beam Scanning Substrate Integrated Waveguide Slot Array Antennas Using a Phase-Correcting Grating Cover

Wenquan Cao, *Student Member, IEEE*, Wei Hong, *Fellow, IEEE*, Zhi Ning Chen, *Fellow, IEEE*, Bangning Zhang, and Aijun Liu

***Abstract*—In this paper, planar beam scanning substrate integrated waveguide (SIW) slot leaky-wave antennas (LWAs) are proposed for gain enhancement using a metallic phase correction grating cover. Unlike conventional Fabry-Pérot (FP) cavity antennas, the proposed antenna is fed by the SIW beam scanning LWA instead of a feeding antenna. The beam scanning angle range is enlarged by meandering the entire feeding structure and the gain is enhanced by a metallic grating cover acting as a 1-D lens. Two kinds of the gratings with different metal strip parameters are designed and analyzed. The proposed antennas operating at the center frequency of 25.45 GHz are designed and experimentally verified for an automotive collision avoidance radar with a gain enhancement of about 4∼6 dB. The proposed SIW LWAs have the advantages of high gain, low profile, easy fabrication and beam-scanning capability good for millimeter wave radar applications.**

***Index Terms*—Beam scanning slot array antennas, high gain, metallic grating, substrate integrated waveguide.**

I. Introduction

It is well known that a Fabry-Pérot (FP) cavity comprises of two metallic arrays acting as partially reflective surfaces (PRS). The periodic metallic structures have been utilized as superstrate layers to significantly enhance the directivity of primary radiating sources at boresight [1], [2]. Many artificial electromagnetic structures have been used to enhance the performance of microwave and millimeter-wave antennas [3]–[8]. In particular, low-profile FP cavity antennas were designed using various surfaces [9]–[13]. In [11], Feresidis reduced the thickness of the cavity from $\lambda/2$ to $\lambda/4$ using one artificial magnetic conductor (AMC) layer. The thickness was further reduced to $\lambda/4$ even $\lambda/10$ using a metamaterial-based resonant cavity [12]. Also, Ourir *et al.* used 2-D composite metamaterial with introduced capacitive and inductive grids to further reduce the thickness to $\lambda/60$ [13].

On the other hand, beam-scanning FP antennas have also been shown by utilizing phase-varying surfaces with varying both capacitance and inductance of the PRSs [14]–[16]. In [17]–[19], varactor diode-tuned structures were proposed to realize the electrically scanned antennas. In [20], the composite metamaterial with a capacitive and an inductive grid was used to obtain a $\pm 20^\circ$ deflection of the antenna beam. Then Ghasemi *et al.* proposed a 2-D PRS configuration to realize a beam scanning of a leaky-wave antenna (LWA) [21]. Nevertheless, only a topology of the scanning FP LWA was presented.

Furthermore, majority of the existing FP cavity antennas are fed by the basic radiating elements, such as monopole, dipole, microstrip patch and waveguide aperture and their beam scanning property is achieved based on complex and expensive electronically controlled structures.

In this paper, an FP cavity antenna is excited by a frequency beam scanning SIW slot array antenna instead of basic antennas. Compared to a conventional SIW slot array antenna, the beam scanning angle range of the proposed feed part is enlarged by meandering the entire structure which was firstly proposed in [22]. A 16-element SIW slot array antenna is designed operating at the center frequency of 25.45 GHz for the frequency modulated continuous wave (FMCW) automotive collision avoidance radar applications. The gain is enhanced by introducing a phase-correcting metallic grating cover. Two kinds of gratings are proposed and analyzed for the beam scanning property with less deterioration.

II. Antenna Feeding Geometry

SIW-like structures have been widely used as planar guided-wave structures at the microwave and millimeter-wave bands [23]–[26]. Fig. 1 shows a top view of a conventional SIW LWA and the 3-D and top views of a unit element. The signal input from one end of the antenna, while the other end is short-circuited to maintain longitudinal wave operation. All slots are 45° inclined. The two additional metal vias are used for cancelling reflection from good impedance matching [27], although the impedance matching of the antenna mainly depends on the positions and the size of the slots. The equivalent circuit model and frequency responses of the phase angles have been given in [22].

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W. Cao, B. Zhang, and A. Liu are with the College of Communications Engineering, PLA University of Science and Technology, Nanjing, China (e-mail: cao_wenquan@163.com).

W. Hong is with the State Key Laboratory of Millimeter Waves, School of Information Science and Engineering, Southeast University, Nanjing, 210096, China (e-mail: weihong@seu.edu.cn).

Z. N. Chen is with the Institute for Infocomm Research, A*STAR, Singapore, and also with the National University of Singapore, Singapore 119615, Singapore (e-mail: chenzn@i2r.a-star.edu.sg; eleczn@nus.edu.sg).

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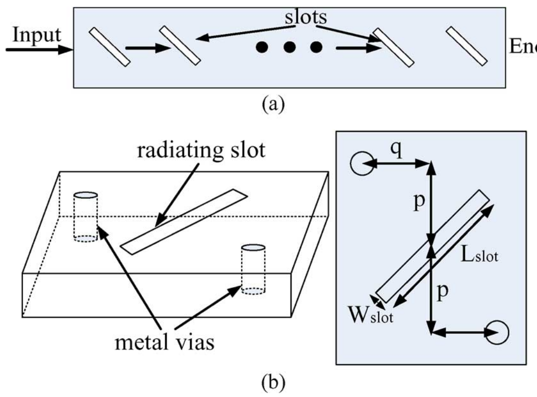


Fig. 1. (a) Top view of the conventional SIW slot array antenna; (b) 3-D and top views of an element of the SIW slot array antenna, where p and q define the position of the metal vias and L slot and W slot are length and width of the radiating slot, respectively.

The relationship between the angle of beam scanning from normal $(\Delta\theta)$ and the phase difference between any two of adjacent slots $(\Delta\varphi)$ is

$$\Delta\theta = \sin^{-1}\frac{\Delta\varphi}{\beta_0 d}, \quad \Delta\varphi = \beta_{SIW} \bullet L \tag{1}$$

where β_0 is the free space propagation constant, d is the physical distance between any two of adjacent slots, $\beta_{\rm SIW}$ is the propagation constant of the SIW and L is the spacing between any two adjacent slots.

For a conventional SIW, $\beta_{\rm SIW}$ is the $\rm TE_{10}$ mode propagation constant of the SIW which is unchanged when the operating frequency and the fabricating material are fixed. Hence, increasing L is the only option to enlarge the value of $\Delta\varphi$ and the beam scanning angle. It is achieved by meandering the entire structure [22]. The phase characteristics along the propagation direction x are shown in Fig. 2(c). It can be found that the meandered structure is longer, the phase slope is shaper. Thus larger beam scanning angle range can be realized with a longer meandered structure. In this design, $\Delta\varphi = n \times 180^\circ$ is chosen, where n is non-negative odd integer to achieve broadside radiation at the center frequency. The steerable angular range can be further enhanced if larger n is chosen at the expense of larger size.

To make a compromise between the maximum steerable range and the antenna size, $\Delta\varphi = 540^\circ$ $(n = 3)$ is chosen. Fig. 2(a) shows the top view of the antenna with the meandered structure. Electric field distribution at the frequency 25.45 GHz is plotted in Fig. 2(b). Nearly three half wavelengths are achieved with three high intensity current distribution centers which confirms $n = 3$. The strength of the field gets weaker in the latter radiating slots. Thus, there is the limitation to enhance the gain of the antenna by increasing the number of the slots.

III. Metallic Grating Design

Usually the FP cavity antennas are analyzed using a ray optics model. The maximum directivity $D_{\max}$ at the boresight direction at a given operating wavelength λ_0 (λ_0 is the free space wavelength) is as follows [1], [28]:

$$D_{\max} = \frac{1 + |Re^{j\phi_{prs}}|}{1 - |Re^{j\phi_{prs}}|} \tag{2}$$

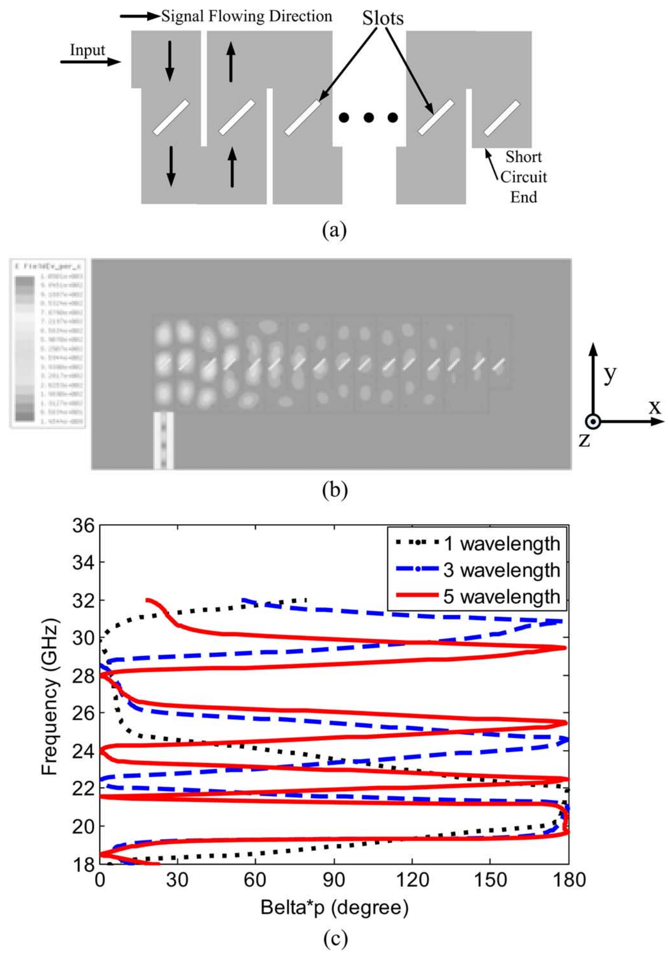


Fig. 2. (a) Top view of the proposed SIW slot array antenna with meandered structure, (b) current distribution of the feeding array antenna at the center frequency 25.45 GHz, and (c) the phase characteristics in direction x.

where $Re^{j\phi_{prs}}$ is the complex reflection coefficient of the PRS. The thickness of the FP cavity, h is determined by

$$-\pi + \varphi_{prs} - 4\pi h/\lambda_0 = 2N\pi, \quad N = 0, \pm 1, \pm 2, \ldots \tag{3}$$

where $-\pi$ is the reflection phase of the ground plane, φ_{prs} is the reflection phase of the PRS. Usually a periodic metallic patch or mesh array is about $-\pi$ and thus the minimum thickness of the cavity is about $\lambda_0/2$. This ray optics model can be adopted as primary design principle. However, it is not accurate enough to optimize the design compared to a full-wave electromagnetic simulation method. Here, the antenna will be designed using the finite element method (FEM) based HFSS software.

Since the feeding structure has beam scanning capability in one dimension (x-direction), 1-D gradient structure is used to enhance the radiation characteristics of the antenna in the other dimension (y-direction) with the beam scanning property with less deterioration in the beam scanning dimension. The cover metallic grating with strips is oriented in parallel with the x-direction. Fig. 3 shows that the pattern controlling is realized by changing the gap spacing g between the metallic strips of the grating or the metal strip width w only in the y-direction and keeping all the other geometric parameters unchanged.

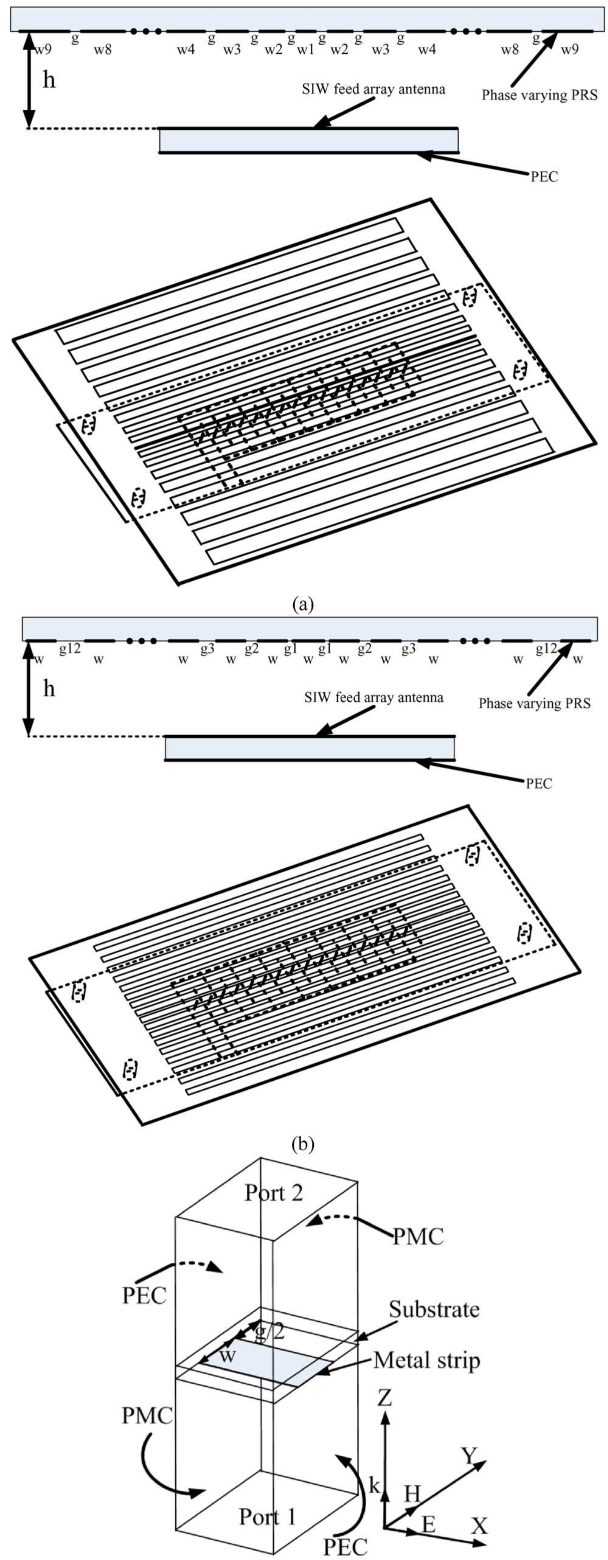


Fig. 3. Schematic view of the proposed antennas composed of an SIW feed array antenna and phase-varying metallic grating, (a) type A, the metal strips with equal gap but gradient widths, (b) type B, the metal strips with equal width but gradient gaps, and (c) characterization model of the grating.

The periodic metallic grating structure can be characterized through its single unit cell. Fig. 3(c) shows that the perfect electric conductor (PEC) and perfect magnetic conductor (PMC) boundary conditions are enforced along both the x- and y-direction to form a waveguide. For simplicity only normal incidence is considered and the incident electric field is polarized along the x-direction.

Since the area of feeding array plane is smaller than that of the grating structure, the metallic grating cover can be considered as a phase-correcting lens. Hence, the transmission characteristic can be investigated instead of reflection property. The transmission characteristic of the grating is simulated at the 15 to 35 GHz frequency bands by changing the size of a unit cell as shown in Fig. 4. So, it concludes as follows:

1) Both the gap g and the strip width w can be adopted to tune the amplitude and phase of the transmission coefficients of the grating unit. It is noted that the variation of g and w account for the shift of the resonant frequency.
2) Concavities exist for the amplitude of the transmission coefficient when the passband of the unit is broadened. The broader the passband is, the stronger the concavity is.
3) An increase in g or a decrease in w causes a shift of the resonance towards lower frequencies. At a fixed frequency, the phase of the grating increases along with a decrease in g or an increase in w. The phase shift for the transmission coefficient is important to control the radiated beam direction of the antenna.

The optimal design should meet two requirements: First, the amplitude owns good passband property with no or minor concavity. Second, the phase property is suitable for the beam scanning function which is not only keeping zero phase for a fixed frequency when boresight pattern is needed but also maintaining the frequency beam-scanning characteristic and the gain enhancement.

IV. Parameters Analysis

Although the unit cell modal given above helps us to understand the effect of the gradient parameters on the transmission coefficients, the parameter analysis of the antenna is still needed to design the antenna. The important performance includes the beam scanning angle and the maximum gain of each scanning beam. Two grating structures (Type A and Type B) are introduced as the upper layer of the antenna with an unchanging feeding structure.

A. Type A of Grating

For Type A of grating, the metal strips with equal spacing and gradient width $(w_1, w_2, \ldots, w_9)$ should satisfy the following formula:

$$\mathrm{w_n} = w_1 + (n-1) \bullet dw. \tag{4}$$

The effects of the height of the air-layer h and the gradient width dw are considered in the optimization. The initial metal strip width w_1 is fixed as 0.45 mm. Fig. 5 shows that both the air-layer height and the gradient width have a slight effect on the beam scanning angle range of about $\pm 25^\circ$. That is caused by the fact that the parameters of the loading part are constant and no variation is introduced.

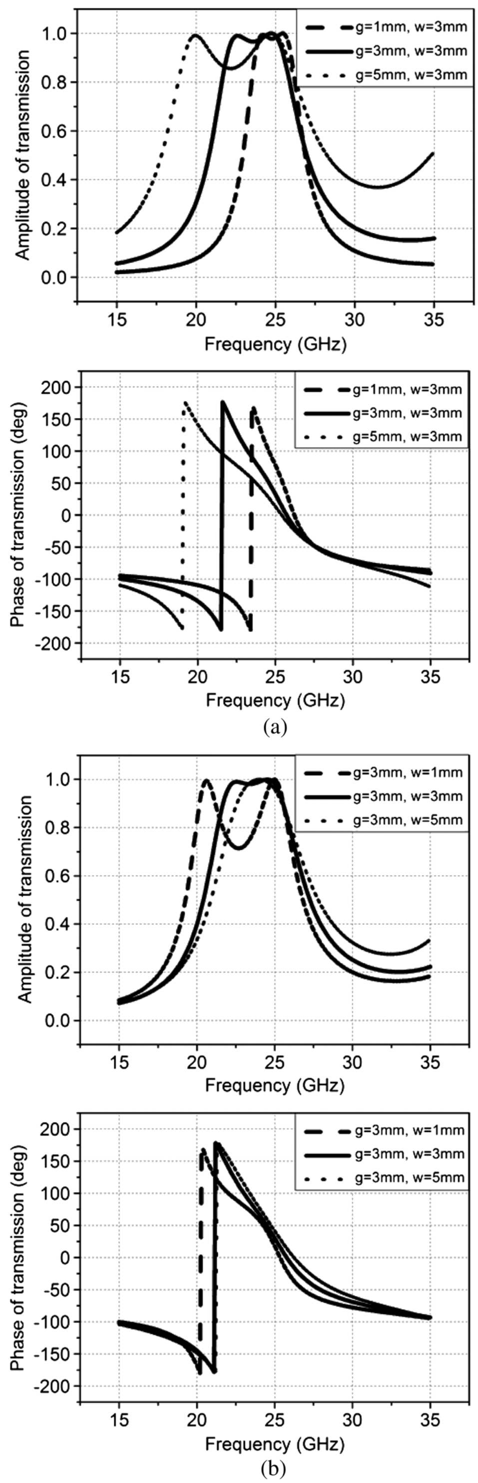


Fig. 4. Transmission coefficients of the grating unit [shown in Fig. 3(c)]: (a) with different gap spacing g; (b) with different strip width w.

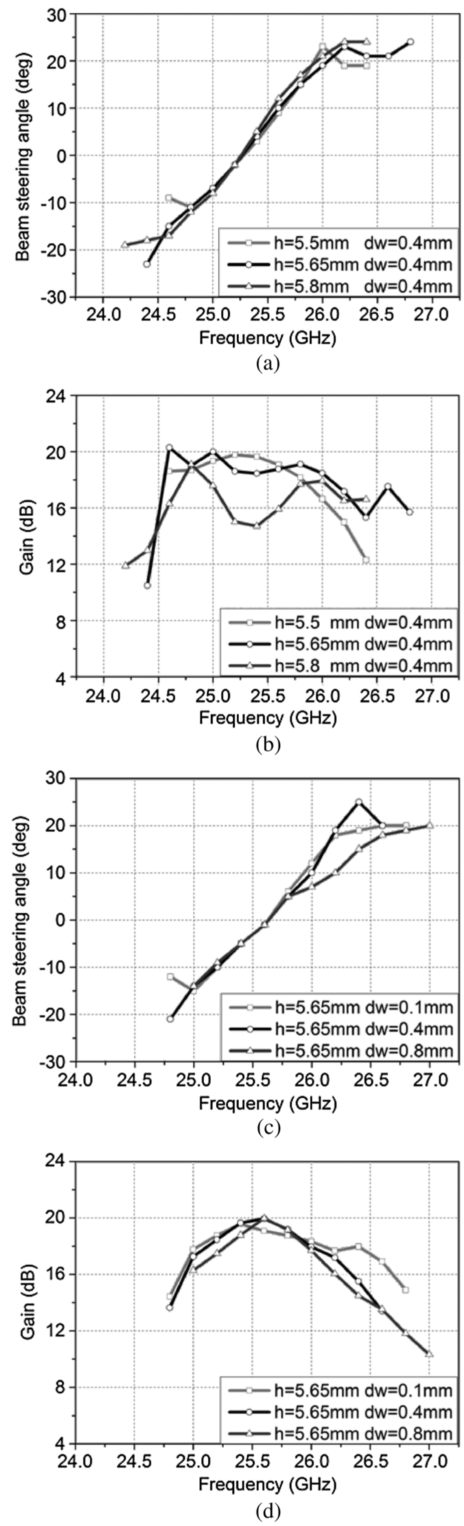


Fig. 5. Simulated antenna performances with varying height of the air-layer h and gradient width dw for Type A: (a), (c) beam scanning angle; (b), (d) the maximum gain of each scanning beam.

Fig. 5(b) shows that using a frequency of 25.45 GHz as a reference, the gain at the other frequencies increases as the height increases. Thus, an optimal value exists for h to obtain higher maximum gain for all scanning beams in the operating band. In this design, 5.65 mm is used. As the height increases, a dip appears in the gain response for the center frequency band. However, varying gradient width dw has limited effect on the gain value. Larger gradient width dw leads to higher maximum gain at the center frequency while the maximum gain of the other frequencies decreases. Thus, tradeoff should be made between the maximum gain of the center frequency and the maximum gains over the operating band.

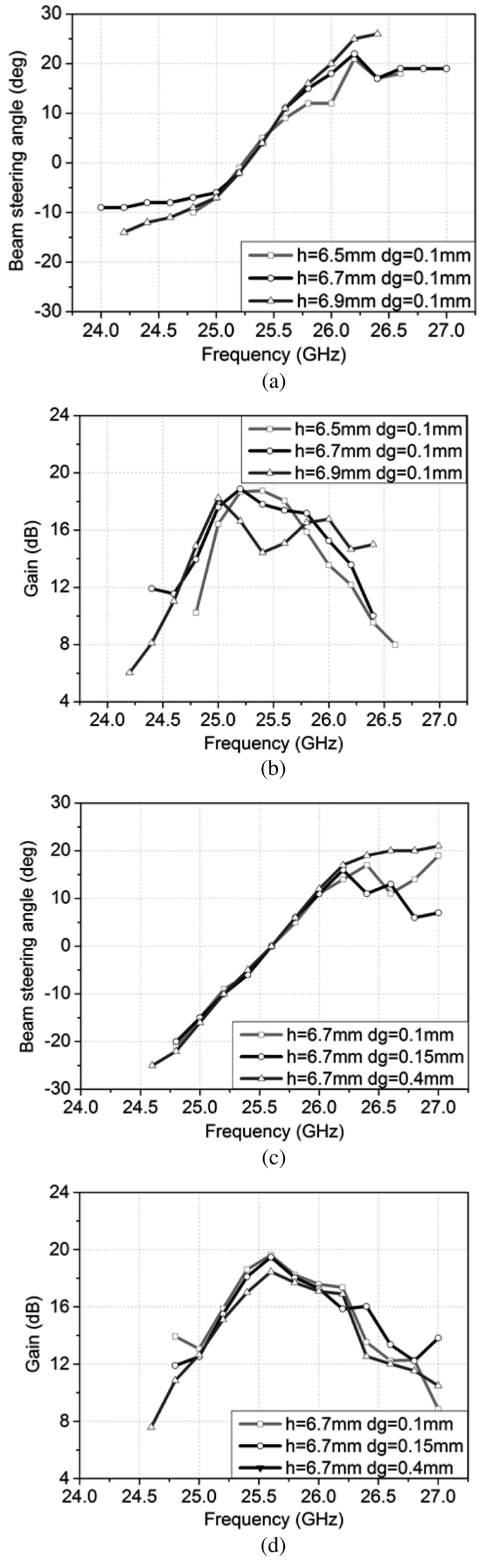


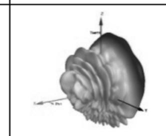
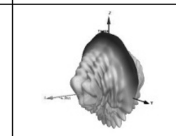
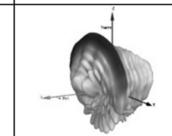
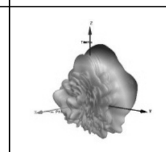
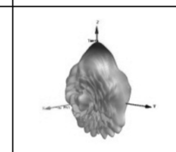
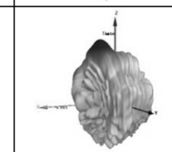
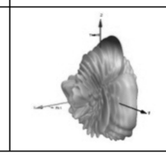
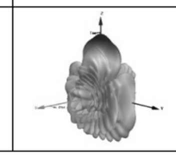
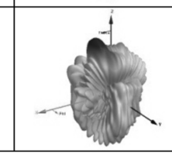
Fig. 6. Simulated antenna performances with varying height of the air-layer h and gradient gap dg for Type B: (a), (c) beam scanning angle; (b), (d) the maximum gain of each scanning beam.

B. Type B of Grating

For Type B of grating, the metal strips with equal width and gradient gaps $(g_1, g_2, \ldots, g_{12})$ should meet the following formula:

$$\mathrm{g_n} = g_1 + (n-1) \bullet dg \tag{5}$$

TABLE I
RADIATION PATTERNS OF THE ANTENNAS

	24.8GHz	25.4GHz	26GHz
without cover grating			
with Type A grating			
with Type B grating			

The effects of the height of the air-layer h and the gradient space gap dg are considered for the optimization. The initial gap g_1 is fixed as 1.2 mm. The similar conclusions to the previous one can be made as shown in Fig. 6. The parameters slightly affect the beam scanning angle range. The maximum gain is mainly affected by the height of the air-layer. An optimum value exists for h ($h = 6.7\ \mathrm{mm}$ for this type) without a dip in the gain curve. It seems that smaller dg leads to little higher gain value. However, dg of 0.1 mm is used considering the craftwork level.

Besides, two other cases, the case with equal gap and equal strip width and the case with reverse gradient distribution are also considered for comparison. Results show that the proposed two types own better performances. For simplicity, detailed analysis is not given in this work. The 3-D radiation patterns of the unloaded and loaded types are also proposed in Table I. It is found that the introduction of cover gratings enhanced the y-direction patterns maintaining the x-direction beam scanning patterns as predicted.

V. Results and Discussion

The two antennas are fabricated for verification and shown in Fig. 7. Both the phase-correcting gratings and the SIW slot feeding array antenna are etched onto the surfaces of a 0.508-mm thick Rogers 5880 substrate. The optimized parameters of the two antennas are shown in Table II.

A. Feeding Array Antenna

The simulated and measured reflection coefficients of the feeding array antenna are shown in Fig. 8. Good agreement between simulation and experiment is achieved. The measured $|\mathrm{S}_{11}| < -10$ dB bandwidth is about 10% covering the required operating band of 24.25–26.65 GHz. Fig. 9 compares the measured and simulated radiation patterns in the x-direction.

Agreement is obtained for the beam scanning angle with an about $\pm 30^\circ$ range. Over the required operating band, the simulated peak gains of the patterns vary from 11.1 dBi to 14.4 dBi, while the measured antenna gains from 9.4 dBi to 13.1 dBi.

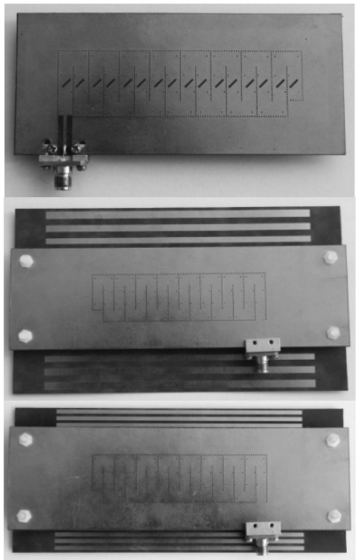


Fig. 7. Photograph of the fabricated antennas.

TABLE II
PARAMETERS OF THE PROPOSED ANTENNAS (UNIT: mm)

p	q	L_{slot}	W_{slot}	r_{via}	a	h_{typeA}
2.9	1.7	3.8	0.6	0.4	5.2	5.65
w_1	dw	g	g_1	dg	w	h_{typeB}
0.45	0.4	3.4	1.2	0.1	2.0	6.7

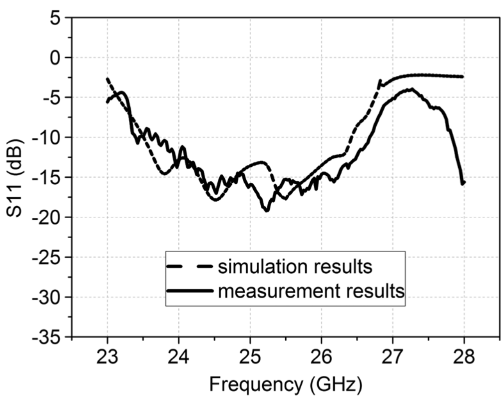


Fig. 8. Simulated and measured reflection coefficients of the feeding array antenna.

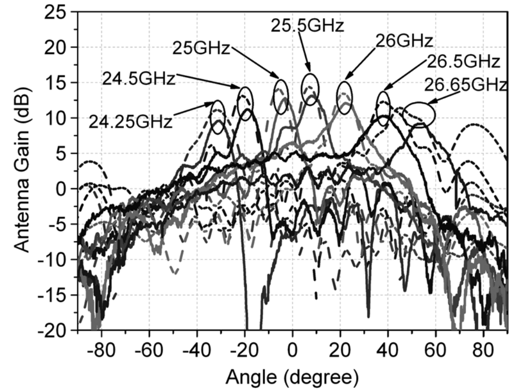


Fig. 9. Simulated (dashed) and measured (solid) radiation patterns of the feeding array antenna.

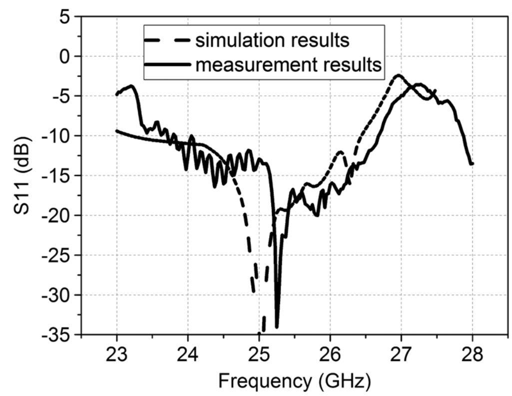


Fig. 10. Simulated and measured reflection coefficients of the proposed antenna with Type A grating.

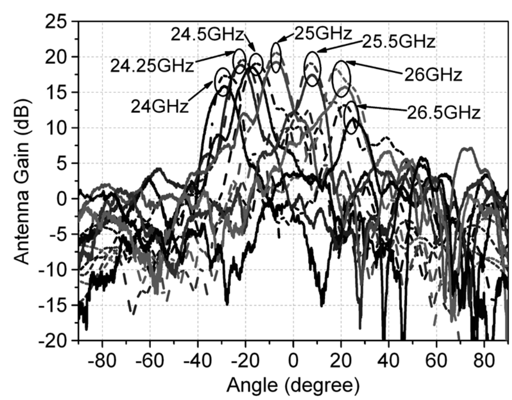


Fig. 11. Simulated (dashed) and measured (solid) radiation patterns of the proposed antenna with Type A grating.

B. Antenna With Type A Grating

Fig. 10 describes the reflection coefficients of the proposed antenna with Type A grating. Minor shift of the resonant frequency between experiment and simulation may be due to the fabrication tolerance. The required operating band is realized. The maximum gains of the scanning beam patterns have been enhanced of about 5–6 dB compared with the antenna without the grating cover as shown in Fig. 11. On the other hand, the beam scanning range has slightly decreased to about $\pm 25^\circ$ which confirms the analysis in Section IV.

C. Antenna With Type B Grating

Figs. 12, 13 show the good agreement between the simulated and measured results of both the reflection coefficients and the radiation patterns for the proposed antenna with Type B grating. Compared with the unloaded one, the gain is enhanced about 4$\sim$5 dB around the operating band with the beam scanning property with less deterioration. It is found that there is a discrepancy around 1.5 dB between the simulated and measured maximum gains of the scanning beams. Measurement errors and fabrication tolerance may contribute to this difference.

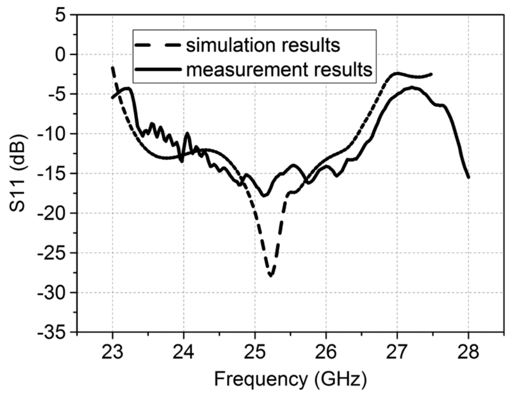


Fig. 12. Simulated and measured reflection coefficients of the proposed antenna with Type B grating.

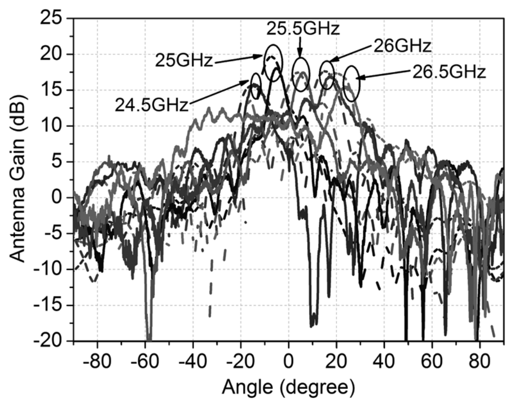


Fig. 13. Simulated (dashed) and measured (solid) radiation patterns of the proposed antenna with Type B grating.

VI. Conclusion

Two types of metallic grating covers together with an SIW LWA as the feeding source have been proposed for 1-D beam steerable antennas. Gain enhancement has been realized by varying the metal strip parameters in the other dimension while keeping the beam steerable dimension unchanged. The unit cell model and the antenna parametric analysis have been introduced. The two types of antennas have been fabricated and measured.

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Wenquan Cao (S'13) received the B.S. degree and Ph.D. degrees from the Institute of Communications Engineering (ICE), Nanjing, China, in 2008 and 2014, respectively.

He is currently an Assistant at the ICE. He has authored and coauthored more than 40 conference and journal papers. His current research interests include microstrip antennas, metamaterials, and their applications to microwave components and antennas.

Mr. Cao has served as a reviewer for the IEEE MICROWAVE AND WIRELESS COMPONENTS LETTERS, IEEE ANTENNAS AND WIRELESS PROPAGATION LETTERS, etc.

Wei Hong (M'92–SM'07–F'12) received the B.S. degree from the University of Information Engineering, Zhengzhou, China, in 1982, and the M.S. and Ph.D. degrees from Southeast University, Nanjing, China, in 1985 and 1988, respectively, all in radio engineering.

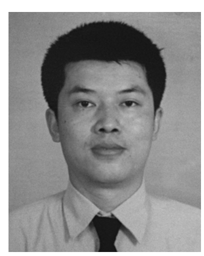
Since 1988, he has been with the State Key Laboratory of Millimeter Waves, Nanjing, China, where he has served as Director of the Laboratory since 2003, and where he is currently a Professor and Vice-Dean of the School of Information Science and Engineering, Southeast University. In 1993, 1995, 1996, 1997, and 1998, he was a short-term Visiting Scholar with the University of California, Berkeley, CA, USA, and the University of California, Santa Cruz, CA, USA, respectively. He has been engaged in numerical methods for electromagnetic problems, millimeter wave theory and technology, antennas, RF technology for wireless communications, etc. He has authored and coauthored over 300 technical publications and authored two books.

Dr. Hong was thrice awarded the first-class Science and Technology Progress Prizes issued by the Ministry of Education of China and Jiangsu Province Government. He also received the Foundations for China Distinguished Young Investigators and for the "Innovation Group" issued by the NSF of China. He is a Fellow of CIE, Vice-President of the Microwave Society and Antenna Society of CIE, Chairperson of the IEEE MTT-S/AP-S/EMC-S Joint Nanjing Chapter, served as an Associate Editor for the IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES from 2007–2010, and is an editorial board member of *IJAP, China Communications, Chinese Science Bulletin*, etc.

Zhi Ning Chen (M'99–SM'05–F'07) received the B.Eng., M.Eng., and Ph.D. degrees in electrical engineering from the Institute of Communications Engineering (ICE), China, and a second Ph.D. degree from the University of Tsukuba, Japan.

During 1988–1995, he was a Lecturer at ICE and later an Associate Professor, as well as a Postdoctoral Fellow at Southeast University, China, and later an Associate Professor. From 1995–1997, he was with the City University of Hong Kong as a Research Assistant and later as a Research Fellow. In 1997, he conducted his research at the University of Tsukuba, Japan, supported by the Japan Society for the Promotion of Science (JSPS) Fellowship. In 2001 and 2004, he visited the University of Tsukuba under a JSPS Fellowship Program (senior level). In 2004, he was with the IBM T. J. Watson Research Center, Yorktown Heights, NY, USA, as an academic visitor. During 1999–2012, he was with the Institute for Infocomm Research (I2R) (formerly, the Centre for Wireless Communications and Institute for Communications Research) as Member of Technical Staff (MTS), Senior MTS, Principal MTS, Senior Scientist, Lead Scientist, and Principal Scientist, as well as Head for RF & Optical Department. In 2012, he joined the Department of Electrical and Computer Engineering, National University of Singapore, as a Professor and is concurrently holding a joint appointment in I2R as Advisor as well as Visiting/Adjunct/Guest Professor positions at Southeast University, Nanjing University, Shanghai Jiaotong University, Tsinghua University, Tongji University, Fudan University, Shanghai University, University of Science and Technology, China, Dalian Maritime University, Chiba University, and City University of Hong Kong. He visited the "Laboratoire des Signaux et Systèmes," UMR8506 CNRS-SUPELEC-Univ. Paris Sud, Gif-sur-Yvette, France, as a senior DIGITEO Guest Scientist in 2013. He is currently interested in engineering electromagnetics and antennas for microwaves, mmW, submmW, and THz systems. He has published 480 technical papers and authored/edited the books entitled *Broadband Planar Antennas, UWB Wireless Communication, Antennas for Portable Devices,* and *Antennas for Base Stations in Wireless Communications*. He has also contributed to the books entitled *UWB Antennas and Propagation for Communications, Radar, and Imaging*, the *Antenna Engineering Handbook*, as well as *Microstrip and Printed Antennas*. He is currently working on the *Handbook of Antenna Technologies* (Springer Reference) as Chief Editor. He holds 38 granted/filed patents with 35 licensed deals.

Prof. Chen has been the founding General Chair of the International Workshop on Antenna Technology (iWAT), the International Symposium on InfoComm & Media Technology in Bio-Medical & Healthcare Applications (IS 3T-in-3A), the International Microwave Forum (IMWF), as well as the Asia-Pacific Conference on Antennas and Propagation (APCAP). He is the recipient of the ASEAN Outstanding Engineering Achievement Award 2013, the International Symposium on Antennas and Propagation Best Paper Award 2010, the CST University Publication Award 2008, the IEEE AP-S Honorable Mention Student Paper Contest 2008, the IES Prestigious Engineering Achievement Award 2006 and 2013, the I2R Quarterly Best Paper Award 2004, and the IEEE iWAT 2005 Best Poster Award. He is an Associate Editor for the IEEE TRANSACTIONS ON ANTENNAS AND PROPAGATION and has served the IEEE Antennas and Propagation Society as a Distinguished Lecturer.



Bangning Zhang received the B.S. and M.S. degrees from the Institute of Communications Engineering (ICE), Nanjing, China.

He is currently a Professor at the ICE. He has authored and coauthored more than 80 conference and journal papers. He has served as a reviewer for several journals in the communications field. His current research interests include communication anti-jamming technologies, microwave technologies, and satellite communications systems.

Aijun Liu received the B.S., M.S., and Ph.D. degrees from the Institute of Communications Engineering (ICE), Nanjing, China, in 1990, 1994, and 1997, respectively.

He is currently a Professor at the ICE. He has authored and coauthored more than 60 conference and journal papers. His current research interests include antenna technologies, wireless communications, and signal processing.