

論文 / 著書情報
Article / Book Information

題目(和文)	
Title(English)	Study of Plate-laminated Corporate-feed Waveguide Slot Array Antennas for Sub-millimeter Wave Bands and Dual-polarization Operation
著者(和文)	金東眞
Author(English)	Dongjin Kim
出典(和文)	学位:博士(工学), 学位授与機関:東京工業大学, 報告番号:甲第9292号, 授与年月日:2013年9月25日, 学位の種別:課程博士, 審査員:廣川 二郎,荒木 純道,安藤 真,淺田 雅洋,西方 敦博,Manuel Sierra Castaner
Citation(English)	Degree:Doctor (Engineering), Conferring organization: Tokyo Institute of Technology, Report number:甲第9292号, Conferred date:2013/9/25, Degree Type:Course doctor, Examiner:,,,,,
学位種別(和文)	博士論文
Type(English)	Doctoral Thesis

Doctoral Dissertation

**Study of Plate-laminated Corporate-feed
Waveguide Slot Array Antennas
for Sub-millimeter Wave Bands and
Dual-polarization Operation**

September, 2013

Under the Supervision of
Associate Professor Jiro Hirokawa

Presented by
Dongjin Kim

Department of Electrical and Electronic Engineering

Tokyo Institute of Technology

Contents

1	Introduction.....	1
1.1	High Performance Antennas for Sub-millimeter Wave Bands.....	1
1.2	Antennas for Dual-polarization Operation.....	7
1.3	Research Scope.....	10
2	Antenna Design for the Sub-millimeter Wave Frequency Bands.....	14
2.1	Introductory Remarks.....	14
2.2	Antenna Configurations.....	18
2.3	Arrangement of Limitations of the Diffusion Bonding of Laminated Thin Metal Plates in the High-Frequency Band.....	18
2.3.1	Fabrication Process.....	18
2.3.2	Design Limitations in Diffusion Bonding Technique.....	19
2.4	16x16-element Array Antenna for 120 GHz.....	20
2.4.1	2x2-subarray Element Design.....	20
2.4.2	Full Structure Analysis Results	23
2.4.3	Measurement Configuration.....	25
2.4.4	Fabrication and Measurement Results.....	28
2.5	Data Transmission Experiments in 120 GHz Band.....	31
2.5.1	Data Transmission Experimental System for 120 GHz.....	31
2.5.2	Data Transmission Experiment Results for 120 GHz.....	32
2.6	16x16-element Array Antenna for 350 GHz.....	36
2.6.1	2x2-subarray Element Design.....	36
2.6.2	Full Structure Analysis Results.....	36
2.6.3	Measurement Configuration.....	39
2.6.4	Fabrication and Measurement Results.....	39
2.7	Data Transmission Experiments in 350 GHz Band.....	42
2.7.1	Data Transmission Experimental System for 350 GHz.....	42
2.7.2	Data Transmission Experiment Results for 350 GHz.....	44
2.7.3	Expectable Maximum Data Rate in 350 GHz.....	46
2.8	Concluding Remarks.....	48
3	Antenna Design for the Higher Gain Applications.....	51
3.1	Introductory Remarks.....	51
3.2	Antenna Configuration.....	54

3.3	Fabrication of Diffusion Bonding of Laminated Thin Metal Plates for Large Size Arrays.....	55
3.4	Fabrication Result of 32x32-element Array Antennas with Conventional Feeding Structure.....	58
3.5	New Feeding Structure for Large Size Arrays with Diffusion Bonding of Laminated Thin Metal Plates.....	61
3.6	Fabrication Result of 32x32- and 64x64-element Array Antennas with New Feeding Structure.....	66
3.6.1	32x32-element Array Antenna.....	66
3.6.2	64x64-element Array Antenna.....	70
3.7	Measurement Configuration Description for Large Size Array Antennas.....	74
3.8	Concluding Remarks.....	76
4	Mutual Coupling Control Technique using Cavity for Small Size Array.....	79
4.1	Introductory Remarks.....	79
4.2	Antenna Configuration.....	81
4.3	Mutual Coupling Effects in Small Size Array.....	83
4.3.1	Mutual Coupling Effect for Single Radiating Element.....	83
4.3.2	Mutual Coupling Effect for 4x4-element Array.....	85
4.4	Mutual Coupling Adjustment with Cavity.....	88
4.4.1	Mutual Coupling Adjustment with Cavities for Each Radiating Element.....	88
4.4.2	Effect of Cavity Dimensions for Slot Characteristics.....	90
4.5	Mutual Coupling among Radiating Elements with Cavity.....	90
4.6	Design and Fabrication Results.....	93
4.7	Concluding Remarks.....	95
5	Antenna Design for Dual-polarization Operation in 60 GHz band.....	99
5.1	Introductory Remarks.....	99
5.2	Antenna Configuration.....	101
5.3	2x2-element Subarray Design.....	101
5.3.1	Radiating Part Design for 2x2-element Subarray.....	103
5.3.2	Feeding Part Design for 2x2-element Subarray.....	106
5.3.3	Design Results of the 2x2-element Subarray.....	106

5.4	16x16-element Design Results.....	108
5.4.1	Corporate Feeding Structure for 16x16-element Array for Dual-polarization Operation.....	108
5.4.2	Design Results of the 16x16-element Array for Dual-polarization Operation.....	112
5.5	Grating Lobe Suppression.....	115
5.5.1	Grating Lobe Suppression in $\varphi=0^\circ$ with Cavity Size Reduction.....	115
5.5.2	Grating Lobe Suppression in $\varphi=45^\circ$ with Element Spacing Reduction.....	118
5.6	16x16-element Design Results with Low Grating Lobes.....	119
5.7	Fabrication and Measurement Results.....	123
5.8	Concluding Remarks.....	127
6	Conclusion.....	130
6.1	Summary of Preceding Chapters.....	130
6.2	Remarks for Future Studies.....	132
	List of Publications.....	134
	Acknowledgement.....	138

Chapter 1

Introduction

1-1 High Performance Antennas for Sub-millimeter Wave Bands

Demand has been increasing for higher data rates in wireless access systems because of the spread of broadband networking. Multimedia services based on high-definition digital video are becoming increasingly common for the public [1]-[3]. To meet these needs, multi-gigabit data capacity communication is required not only for wired access networks but also for wireless communications. To do so, the data rates of many current wireless technologies operating at microwave bands, such as the multiple-input multiple-output (MIMO) technique, orthogonal frequency-division multiplexing (OFDM) and ultra-wide-band (UWB), have been improved by increasing the spectral efficiency by means of advanced modulation schemes and signal processing technologies [4]-[6]. However, it appears to be very difficult to achieve the data rate of over 10 Gbit/s using microwave bands because they have the fundamental limitation of a narrow bandwidth. Ultimately, the data rate of wireless communication systems depends on the available bandwidth.

In terms of the bandwidth, the easiest way to achieve a higher data rate is to

increase the carrier frequency to millimeter and sub-millimeter bands over 100 GHz. This is mainly because we can then make use of ultra-large bandwidth for high-speed communications [7]-[8]. So that reason, the millimeter and the sub-millimeter wave bands higher than 100 GHz are very attractive for wireless communications. However, they also suffer from high attenuation in the atmosphere, which is much larger than that in the microwave bands up to 30 GHz [9]-[10]. The large attenuation not only limits the communication distance but also degrades the signal-to-noise ratio (SNR) of the system, which influences the data rate as well. In that sense, an antenna is one of the key components in the data transmission system that employs millimeter and sub-millimeter waves higher than 100 GHz as a carrier frequency. To overcome the large loss even in free space, antennas with high gain and high efficiency are required [11]. Additionally, for the practical data link system, the antenna should meet strict requirements (small size, light weight, wide bandwidth and low production cost) while maintaining a high antenna gain.

The antennas satisfying the conditions of high antenna gain, high efficiency, and broad bandwidth with compact antenna size are very rare in millimeter wave and sub-millimeter wave bands due to the large loss and physical difficulties of the antenna design and fabrication. Several potential candidates have been reported in millimeter wave region [12]-[14].

(1). Microstrip patch array antenna for 77 GHz band automobile radar [12]

Fig. 1 shows the microstrip patch array antenna for 77 GHz automobile radar. It shows very nice antenna performance such as low profile and wide bandwidth. However it shows low antenna efficiency lower than 40% with 20 dBi antenna gain in the 70 GHz region.

(2). Box-horn array antenna for 58 GHz band [13]

Horn array antenna fed with parallel-feed by an H-plane-dividing waveguide shown in Fig. 2 was proposed in the 50 GHz band [13]. The horn antennas are connected to the feed waveguide by the E-corners. Thus the transmission loss in feeding network is substantially low compared with the microstrip antennas. However, it has been structurally impossible to feed each horn antenna without

the spacing being larger than the free-space wavelength. A 16x16-element array had a high gain of 36 dBi, but it shows only about 30% aperture efficiency due to the large spacing among elements.

(3). Lens with slot antenna for 120 GHz band [14]

A silicon based lens is employed for the 120 GHz band antenna with a slot antenna. The lens or reflector could simply increase the antenna directivity and gain. With lens, the directivity of the slot antenna shows over 16 dBi gain while it shows only 5 dBi without lens. Furthermore, it shows a simple structure composed of only antenna and lens as shown in Fig. 3. However, it is difficult to realize thin planar structure because it essentially requires focal length as same as the reflector antennas [15].

Microstrip array antennas are more advantageous than other millimeter wave antennas in terms of its low profile and low cost. However, the feeding loss due to the transmission loss of the microstrip line is a significant problem in the array feeding. Therefore, microstrip array antennas are suitable for relatively low-gain applications with compact size. Horn array antenna shown in [13] shows much potential for millimeter wave region due to its low loss, but it requires large element spacing which results in the low aperture efficiency. Furthermore, its 3-dimensional structure is complicated and the manufacturing cost becomes high.

The reflector or lens antennas are the most promising candidate for high frequency regions including the millimeter wave and the sub-millimeter wave in the view of the antenna performance since they provide a broad bandwidth characteristic, a high gain with a high efficiency. However, they have a very bulky structure due to their focal length. Therefore it seems difficult to be used in the actual data link system.

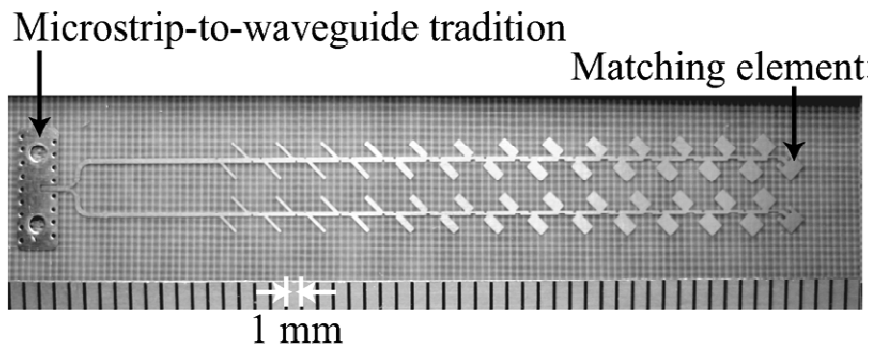


Fig. 1 Microstrip patch array antenna for 77 GHz automobile radar.

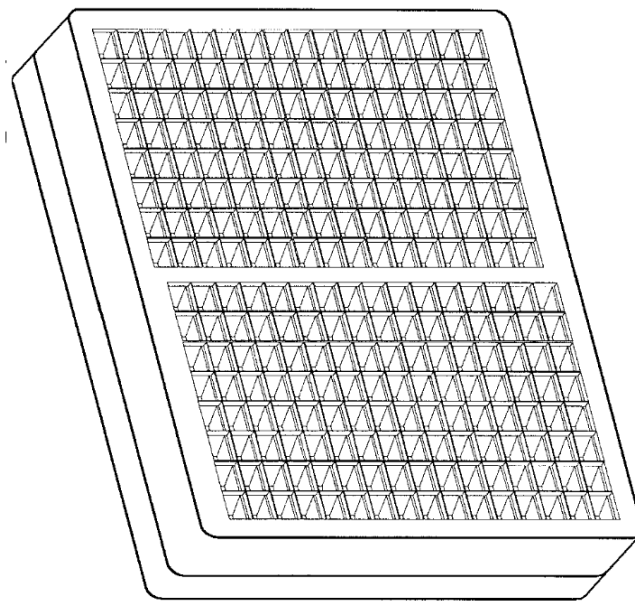


Fig. 2 Box-horn array antenna for 58 GHz

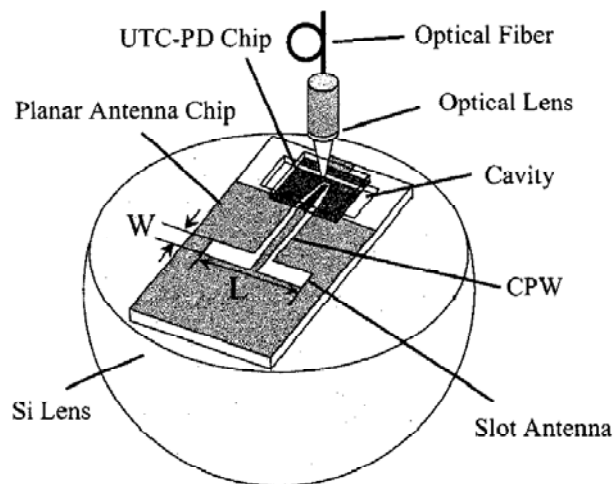


Fig. 3 Lens with slot antenna for 120 GHz band

A hollow-waveguide slot array antenna is a good candidate for the applications for high frequency regions including millimeter wave and sub-millimeter wave bands due to its characteristics such as low loss, low profile, high gain and high antenna efficiency [16]. Since the hollow waveguide slot array antenna neither suffers from dielectric nor radiation loss, high performances can be maintained in such high frequency bands. Fig. 4 shows the typical waveguide flat profile antenna array using slots [17]. As shown in here, it shows quite compact and simple structure. However, conventional waveguide slot array antennas having series-feeding structure have a fundamental problem that the bandwidth is narrowed due to the long-line effect when the array size becomes large [18]-[19]. As a result, it shows the inverse-proportional relationship between antenna gain and bandwidth characteristic in conventional waveguide slot array antennas having series-feeding structure due to the long line effect in feeding network. Therefore, we could not expect a broadband characteristic and a high antenna gain in same time.

By the way, multiple layer structure could be a solution for the problems in the conventional waveguide slot array antennas with series feeding structure, since it has the potential to enhance the design freedom of the feed waveguide to include parallel feed for a wider bandwidth. The double-layer corporate-feed waveguide slot array for the broad bandwidth operation is proposed by author's research group in 60 GHz band [20]. This antenna consists of a corporate-feed waveguide arranged in the lower layer and 2x2-element subarray units in the upper layer, as shown in Fig. 5. The 2x2 elements in the subarray are almost equally spaced from the coupling aperture between the layers at the end of the corporate feed, and the array is substantially fed by a fully corporate feed. The 16x16-element array antenna shows more than 80% antenna efficiency with more than 32 dBi over the broad bandwidth of 4.8 GHz in 60 GHz. Furthermore, the diffusion bonding of laminated thin metal plates is adopted for the fabrication of this antenna, which gives high precision and perfect electrical contact, even at a high frequency, with a relatively low cost compared with a conventional fabrication method, such as die-casting [21]-[22].

In this study, we demonstrate a feasibility of the antenna in [20] for sub-millimeter wave bands with various functions including high frequencies, different array sizes. In the chapter 2, 3 and 4, we show the various antenna design techniques

and performances according to the antenna functions.

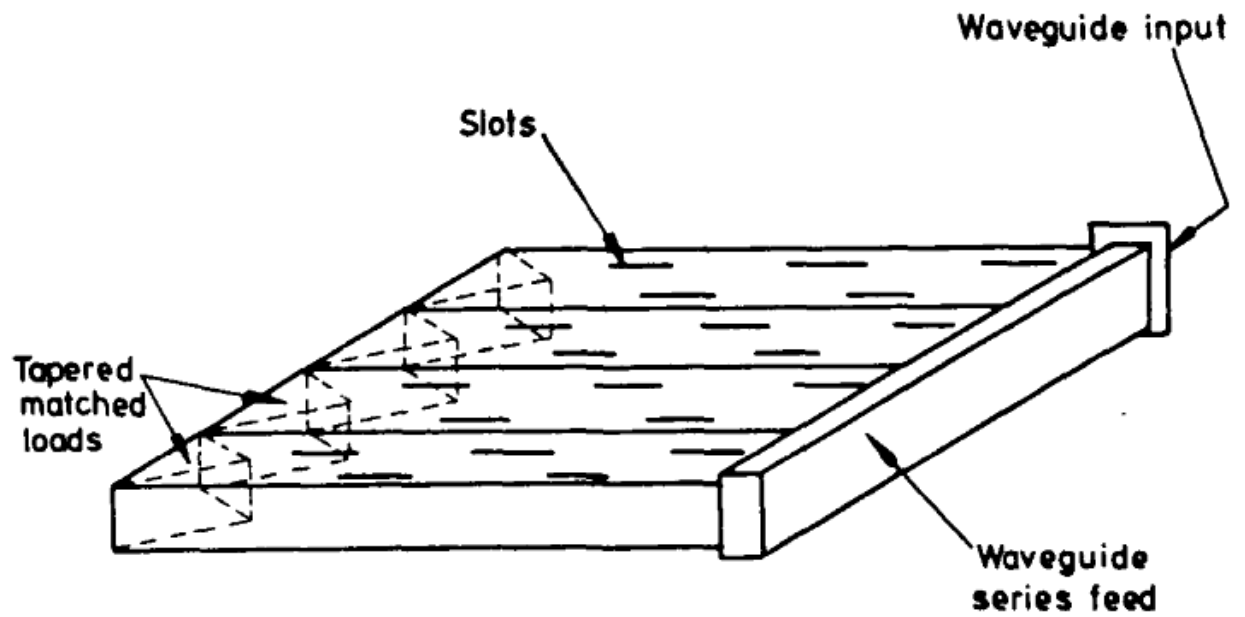


Fig. 4 Typical waveguide flat profile antenna array using slots with series feed

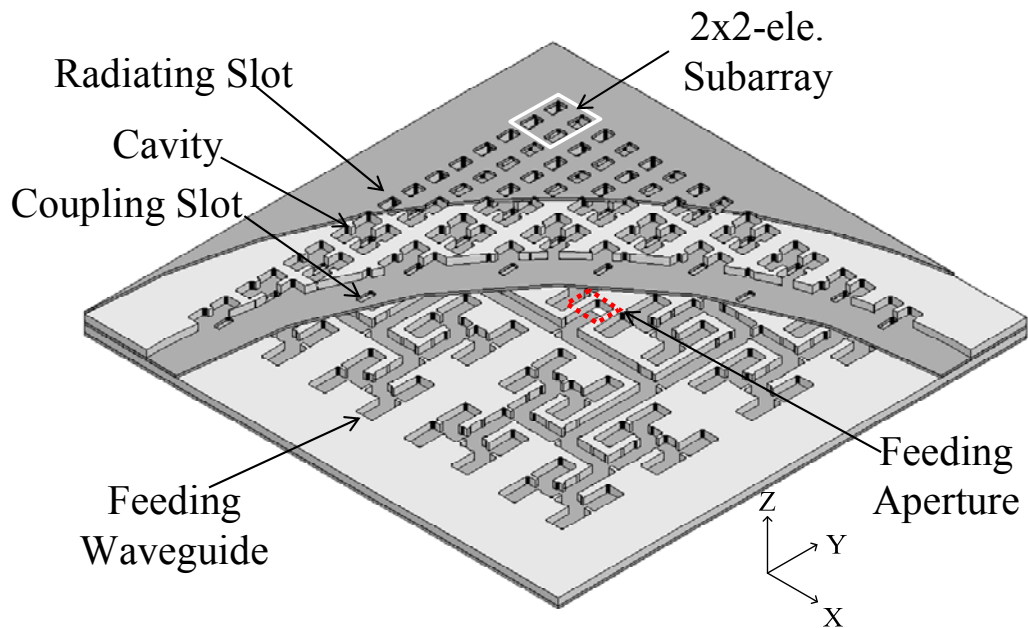


Fig. 5 Double-layer corporate-feed waveguide slot array antenna

1-2 Antennas for Dual-polarization Operation

Meanwhile, the dual polarization technique is also one of the promising methods to increase the frequency reuse efficiency in twice because it can generate two different polarizations in same time. The additional polarization can also be used as a diversity measure to improve the received SNR and/or increase the data rate in communication applications. So that reason, the dual-polarization antennas are widely used for in both of communication systems and radar applications.

Many dual polarized antennas have been presented in the literature. These include dual polarized patch and slot antennas [23]-[27]. A variety of antenna structure to achieve high isolation and good cross-polarization has been demonstrated. However, most of the studies on the dual-polarization antenna arrays use microstrip substrates for the easy realization of the multiple layer feeding network and the common radiating elements which are essential for the dual-polarization operation. Also, their feeding techniques require double-layer or multilayer substrates and the structures are too complicated. Furthermore, the microstrip antennas are not suitable in the high frequency regions including millimeter wave and sub-millimeter wave bands due to the large loss in their feeding networks.

As explained Chapter 1-1, a hollow-waveguide slot array antenna could be a solution in the view of the loss in higher frequency regions, however it also characterized as less design flexibility in dual-polarization operation and beam scanning, etc. Actually, the waveguide slot antennas for the dual-polarization are almost absent from literature. In most previous works [28]-[29], the sub-arrays with orthogonal linear polarizations are interlaced side by side for easy realization of the two different polarizations in one aperture.

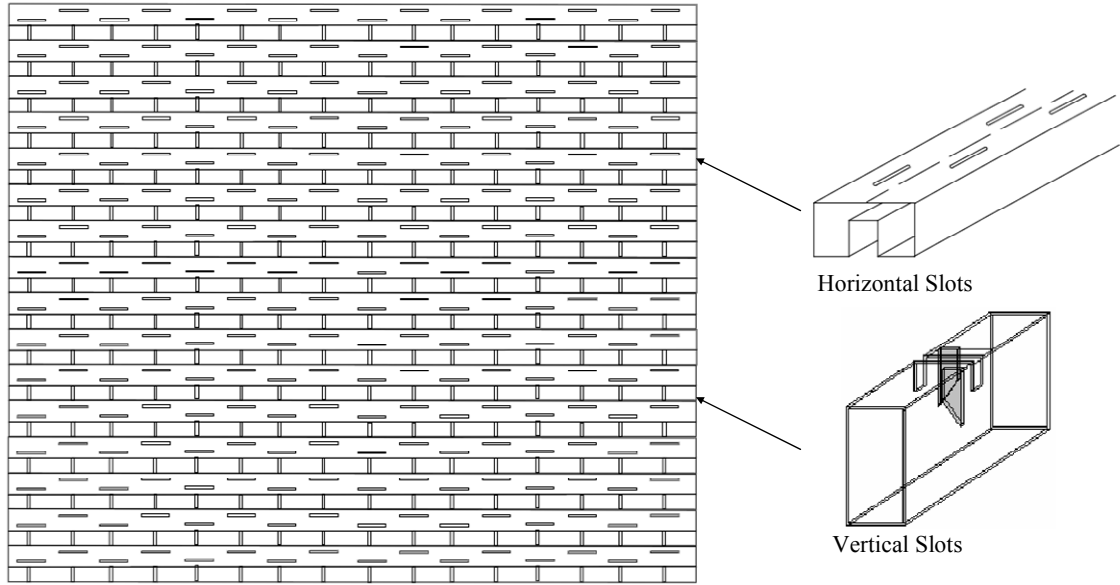


Fig. 6 Waveguide slot array antenna for dual-polarization operation for SAR application

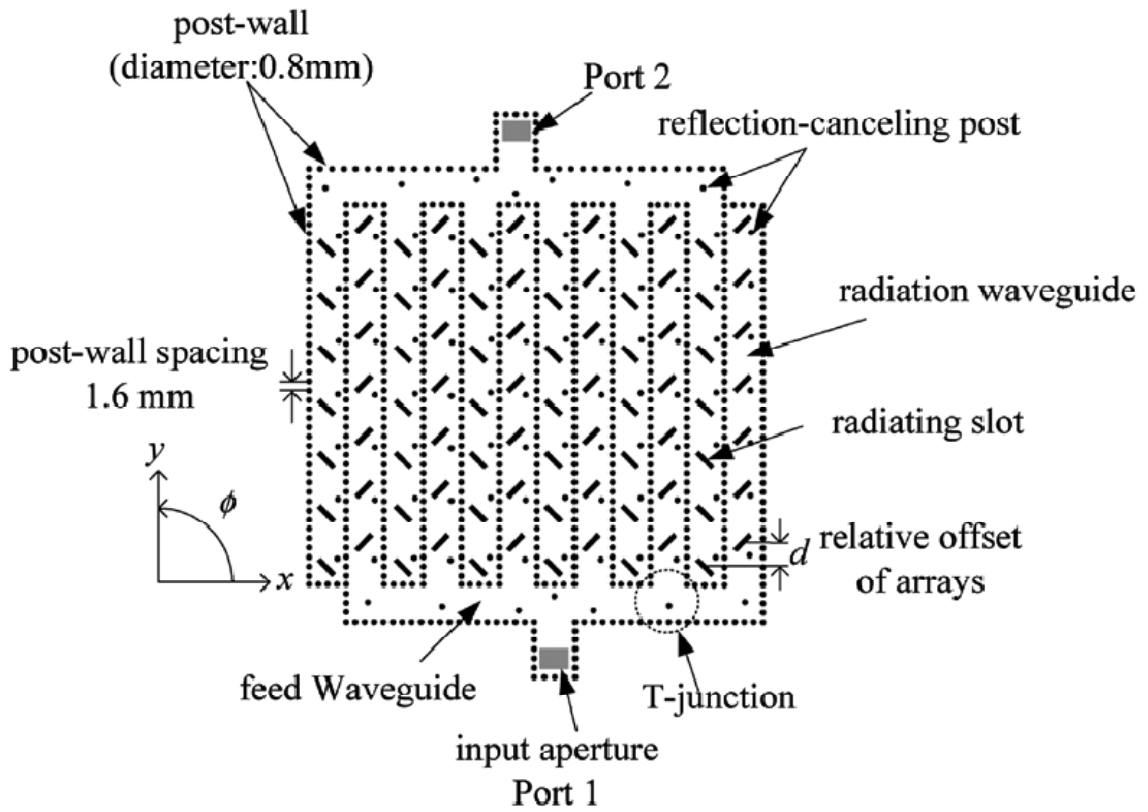


Fig. 7 Parallel-plate slot array antenna with interdigital structure for 45° linear and dual-polarization operation

- (1). Waveguide slot array antenna for dual-polarization operation for SAR application [28].

Fig. 6 shows the waveguide slot array antennas for dual-polarization operation. The vertical polarization is realized with a ridged-waveguide longitudinal slot linear antenna array, while the horizontal polarization is realized with an untilted narrow-wall slot linear array. Two subarrays for each polarization are arrayed alternatively. It shows high isolation over 40 dB in 10 GHz band.

- (2). Parallel-plate slot array antenna with interdigital structure for 45° linear and dual-polarization operation in [29].

In Fig. 7, a slotted post-wall waveguide array antenna for dual-polarization operation is shown. Two post-wall slotted waveguide arrays with orthogonal 45° linearly-polarized waves interdigitally share the aperture on a single layer substrate. It shows isolation more than 30 dB in 25 GHz band.

The antennas in [28]-[29] show the advantages of low profile, high gain with high isolation. However the interlaced structure of two different polarizations increases the antenna aperture size and decreases the aperture efficiency. The large element spacing larger than free space wavelength causes high grating lobes. Furthermore, they still suffer from the narrow bandwidth caused by the long line effects in their feeding structure.

In this study, the multi-layer waveguide slot array antenna for the dual-polarization operation having high gain, high efficiency, and high isolation with low profile is presented. It also based on the structure introduced in Chapter 1-1 [20]. For the broad bandwidth characteristic, the corporate feeding structure is adapted for each polarization. Also, the common radiating slot is employed for the different two polarizations to overcome the low efficiency problems in interdigital structure in [28]-[29]. Furthermore, the fabrication method of diffusion bonding adopted in this research enables mass production of the multi-layer structure. Chapter 5 introduce multi-layer waveguide slot array antenna for the dual-polarization operation.

1-3 Research Scope

The objective of this study is to design the antenna structure described in [20] for the various applications with maintaining its characteristics of high gain, efficiency, and wideband.

Chapter 2 presents an antenna design for the sub-millimeter wave bands including 120 GHz and 350 GHz. The design limitations of diffusion bonding of laminated thin metal plates for the high frequency are described. The antenna design and fabrication results for 120 GHz and 350 GHz band are discussed. Furthermore, to demonstrate the feasibility of the proposed antennas for wireless link systems, we perform data transmission experiments with the wireless link systems in [30]-[31] for multi-gigabit including over 10 Gbit/s.

Chapter 3 presents the large size array antennas for the applications require higher gain with a fabrication method of the diffusion bonding of laminated thin metal plates. Here, we propose a 32x32-element array antenna and a 64x64-element array antenna in the 120 GHz band, respectively. Also, we propose a new feeding structure having a double-layer for the stable fabrication.

Chapter 4 presents a 4x4-element corporate-feed waveguide slot array antenna with cavities for the 120 GHz band. In order to adjust the mutual coupling effects in a small size array antenna, a cavity structure is employed for each radiating slot. The cavity in this antenna structure provides proper mutual coupling effects for each radiating element. As a result, we could realize a high antenna performance in the small size array antenna.

Chapter 5 presents a dual-polarization antenna design in 60 GHz band. The multiple-layer feeding structure is proposed for the dual-polarization operation. It is designed for the dual-polarization excitation with broadband and high isolation. The cross-shaped common radiating slots are employed for the dual-polarization operation.

Finally, Chapter 6 summarizes this study and discusses the future study.

References in Chapter 1

- [1] S. J. Trowbridge “High-speed ethernet transport,” IEEE Comm. Mag., vol. 45, no. 12, pp.120-125, Dec. 2007.
- [2] J. D’Ambrosia, “40 gigabit Ethernet and 100 gigabit Ethernet: the development of a flexible architecture—[Commentary],” IEEE Commun. Mag., vol. 47, no. 3, pp. 8–14, Mar. 2009.
- [3] S. Cherry, “Edholm’s law of bandwidth”, IEEE Spectr., vol. 41, no. 7, pp. 58–60, Jul. 2004.
- [4] G. J. Foschini and M. J. Gans, “On limits of wireless communications in a fading environment when using multiple antennas,” Wireless Pers. Commun., vol. 6, pp. 311–335, Sep. 1998.
- [5] A. J. Paulraj, D.A. Gore, R.U.Nabar and H. Bolcskei, “An overview of MIMO communications—A key to gigabit wireless,” Proc. IEEE, vol. 92, pp. 198–218, Feb. 2004
- [6] V. Dyadyuk, J. Bunton, J. Pathikulangara, R. Kendall, O. Sevimli, L.Stokes and D. Abbott, “A multi-gigabit mm-wave communication system with improved spectral efficiency,” IEEE Trans. Microwave Theory Tech., vol. 55, no. 12, pp. 2813–2821, Dec. 2007.
- [7] R. Piesiewicz, T. Kleine-Ostmann, N. Krumbholz, D. Mittleman, M.Koch, J. Schoebel, and T. Kurner, “Short-range ultra-broadband terahertz communications: Concepts and perspectives,” IEEE Antennas Propag. Mag., vol. 49, no. 6, pp. 24–39, 2007.
- [8] J. Laskar, S. Pinel, D. Dawn, S. Sarkar, B. Perumana, and P. Sen, “The next wireless wave is a millimeter wave,” Microw. J., vol. 50, pp. 22–36, Aug. 2007
- [9] Attenuation by Atmospheric Gases, ITU-R Recommendation P. 676-8, 2009
- [10] P. Scott, “The am Atmospheric model, submillimeter array,” Tech. Memo #152 [Online]. Available: <https://www.cfa.harvard.edu/~spaine/am>
- [11] K. Huang and Z.Wang, “Terahertz terabit wireless communication,” IEEE Microwave Mag., vol. 12, no. 4, pp. 108–116, May. 2011.
- [12] Y. Hayashi, K. Sakakibara, M. Nanjo, S. Sugawa, N. Kikuma, and H. Hirayama,

- “Millimeter-Wave Microstrip Comb-Line Antenna Using Reflection-Canceling Slit Structure,” *IEEE Trans. Antennas Propag.*, vol. 59, no. 2, pp. 398–406, Feb. 2011.
- [13] T. Sehm, A. Lehto, and V. Raisenen, “A High-Gain 58-GHz Box-Horn Array Antenna with Suppressed Grating Lobes,” *IEEE Trans. Antennas Propag.*, vol. 47, no.7, pp. 1125–1130, Jul. 1999.
- [14] A. Hirata, H. Ishii, and T. Nagatsuma, “Design and Characterization of a 120-GHz Millimeter-Wave Antenna for Integrated Photonic Transmitters,” *IEEE Trans. Microwave Theory Tech.*, vol. 49, no.11, pp. 2157–2162, Nov. 2001.
- [15] A. Mizutani, K. Sakakibara, N. Kikuma, and H. Hirayama, “Grating Lobe Suppression of Narrow-Wall Slotted Hollow Waveguide Millimeter-Wave Planar Antenna for Arbitrarily Linear Polarization,” *IEEE Trans. Antenna and Propag.*, vol. 55, no.2, pp.313-320, Feb. 2007.
- [16] S. Oh, J. Lee, M. Song and Y. Kim, “Two-layer slotted-waveguide antenna array with broad reflection/gain bandwidth at millimetre-wave frequencies,” *IEE Proc.-Microw. Antennas Propag.*, Vol. 151, No. 5, pp. 393-398, Oct. 2004.
- [17] P.S. Hall, and J. R. James, “Survey of design techniques for flat profile microwave antennas and arrays,” *Radio and Electronic Engineer*, Vol. 48, Issue_11, pp. 549-565, Nov. 1978
- [18] M. Ando, Y. Tsunemitsu, M. Zhang, J. Hirokawa, and S. Fujii, “Reduction of long line effects in single-layer slotted waveguide arrays with an embedded partially corporate feed,” *IEEE Trans. Antennas and Propag.*, vol. 58, no.7, pp.2275-2280, July, 2010.
- [19] S. Park, J. Hirokawa, and M. Ando, “A planar cross-junction power divider for the center feed in single-layer slotted waveguide arrays,” *IEICE Trans. Commun.*, vol. E85-B, no. 11, pp. 2476–2481, Nov. 2002.
- [20] Y. Miura, J. Hirokawa, M. Ando, Y. Shibuya, and G. Yoshida, “Double-layer full-corporate-feed hollow-waveguide slot array antenna in the 60-GHz band,” *IEEE Trans. Antennas Propag.*, vol. 59, no.8, pp.2844-2851, Aug. 2011.
- [21] R.W. Haas, D. Brest, H. Mueggenburg, L. Lang and D. Heimlich, “Fabrication and performance of MMW and SMMW platelet horn arrays,” *Intl. J. of Infrared and Millimeter Waves*, Vol. 14, Num. 11, pp. 2289-2294, Jan.1993
- [22] J. Hirokawa, M. Zhang, and M. Ando, “94 GHz single-layer slotted waveguide

- array by diffusion bonding of laminated thin plates,” in Proc. Int. Symp. Antenna Propag. TP-C0.-3, Oct. 2008
- [23] S. C. Gao, L. W. Li, M. S. Leong, and T. S. Yeo, “Dual-polarized slotcoupled planar antenna with wide bandwidth,” IEEE Trans. Antennas Propag., vol. 51, no. 3, pp. 441–448, Mar. 2003.
 - [24] B. G. Porter, L. L. Rauth, J. R. Mura, and S. S. Gearhart, “Dual-polarizad slot-coupled patch antennas on Duroid with Teflon lenses for 76.5 GHz automotive radar systems,” IEEE Trans. Antennas Propag., vol. 47, no. 12, pp. 1839–1842, Dec. 1999.
 - [25] S. D. Padhi, N. C. Kamakar Sr., C. L. Law, and S. Aditya Sr., “A dual polarized aperture coupled circular patch antenna using a C-shaped coupling slot,” IEEE Trans. Antennas Propag., vol. 51, no. 12, pp. 3295–3298, Dec. 2003.
 - [26] C.-H. Lee, S.-Y. Chen, and P. Hsu, “Isosceles triangular slot antenna for broadband dual polarization applications,” IEEE Trans. Antennas Propag., vol. 57, no. 10, pp. 3347–3351, Oct. 2009.
 - [27] E. A. Soloman, M. S. Ibrahim, and A. K. Abdelmageed, “Dual-polarized omnidirectional planar slot antenna for WLAN applications,” IEEE Trans. Antennas Propag., vol. 53, no. 9, pp. 3093–3097, Sep. 2005.
 - [28] WangWei, JinJian, and LuJia-Guo, “Waveguide slotted antenna array with broadband, dual-polarization and low cross-polarization for X-Band SAP applications,” IEEE International Radar Conference, pp. 653-656, 2005
 - [29] S. Park, Y. Okajima, J. Hirokawa and M. Ando, “A slotted post-wall waveguide array with interdigital structure for 45 linear and dual polarization,” IEEE Trans. Antenna Propag., vol. 53, pp. 2865-2871, Sep. 2005.
 - [30] A. Hirata, T. Kosugi, H. Takahashi, R. Yamaguchi, F. Nakajima, T. Furuta, H. Ito, H. Sugahara, Y. Sato, and T. Nagatsuma, “120-GHz-Band Millimeter-Wave Photonic Wireless Link for 10-Gbit/s Data Transmission,” IEEE Trans. Microwave Theory and Tech., vol. 54, no.5, pp. 1937-1944, May, 2006.
 - [31] T. Nagatsuma, H.-J Song, Y. Fujimoto, K. Miyake, A. Hirata, K. Ajito, A. Wakatsuki, T.Furuta, N. Kukutsu, and Y. Kado, “Giga-bit wireless link using 300-400 GHz bands”, Tech. Dig. IEEE Intl. Topical Meeting on Microwave Photonics, Th.2.3, Oct. 2009

Chapter 2

Antenna Design for the Sub-millimeter Wave Frequency Bands

2-1 Introductory Remarks

One of the promising techniques to increase the data transmission rate is to use higher frequency band as a carrier signal because the transmission capacity of the wireless link system increases as the carrier frequency becomes higher. For this reason, there has been an increasing interest in the use of the millimeter wave, the sub-millimeter wave and THz bands as the carrier frequency for the wireless communication system [1]-[3]. Furthermore, these high frequency bands over 100 GHz are not allocated for any specific applications and thus we can freely access to wide bandwidth for high-speed communication [4]-[5]. Already it has been reported that communication systems that use 120 GHz and 350 GHz frequency bands as the carrier frequency have achieved over 10Gbps data rate [6]-[7].

By the way, for the wireless link systems mentioned above, the antennas having a high gain and high antenna efficiency are strongly required since the radio signals in the sub-millimeter wave and THz bands are suffered from the high propagation loss even in the free space [8]-[9]. Furthermore, wide bandwidth is also required for

high-speed data communication; the antennas satisfying these conditions are really rare in this high frequency band. A commercial parabolic antenna and a horn antenna with a dielectric lens had been used for the communication system trials [6]-[7]. However they are not suitable for actual wireless link system because they need large volume for only the antennas. To integrate the antenna in practical wireless link system, low profile as well as compact size is highly recommended for the antenna.

We demonstrate the feasibility of a waveguide slot array antenna with a low profile for wireless link systems. A double-layer corporate-feed antenna is a good candidate for the wireless link system for such a high frequency band because of its characteristics, such as low profile, high antenna efficiency and broad bandwidth. We already have proposed a corporate-feed slot array antenna structure for the 60 GHz band [10]. The 16x16-element array antenna shows more than 80% antenna efficiency with more than 32 dBi gain over a broad bandwidth of 4.8 GHz.

In this chapter, we design the antenna in [10] for the 120 GHz and 350 GHz bands to demonstrate the performance. We expect that this antenna could be adaptable in the wireless link system in such high-frequency bands. Because the merits of this antenna such as low profile, high antenna efficiency, high gain and broad bandwidth characteristic are very attractive to the wireless data link system. Basically, the antenna design is based on the previously designed antenna in 60 GHz; it is scaled down for a half size and a one-six size for 120 GHz and 350 GHz bands, respectively. However there are several limitations on the antenna design caused in the fabrication process due to small parameters in these high frequency bands. These limitations on the antenna design significantly affect the antenna performance. We show their effects on the antenna performance and redesign this antenna considering the design limitations. We redesign this antenna considering the design limitations and show the design and measurement results.

Also, to demonstrate the feasibility of the proposed antennas for wireless link systems, we perform data transmission experiments with the wireless link systems in [6]-[7] for various data rates including over 10 Gbit/s.

Firstly, we describe the configuration of the antenna. Next we explain the design limitation in these high frequency bands. We present the antenna design process and results in 120 GHz and 350 GHz bands, respectively. The measurement results of

the antennas with the introduction of the measurement systems will be presented. Finally we discuss the data transmission experimental results with the proposed antennas in 120 GHz and 350 GHz, respectively.

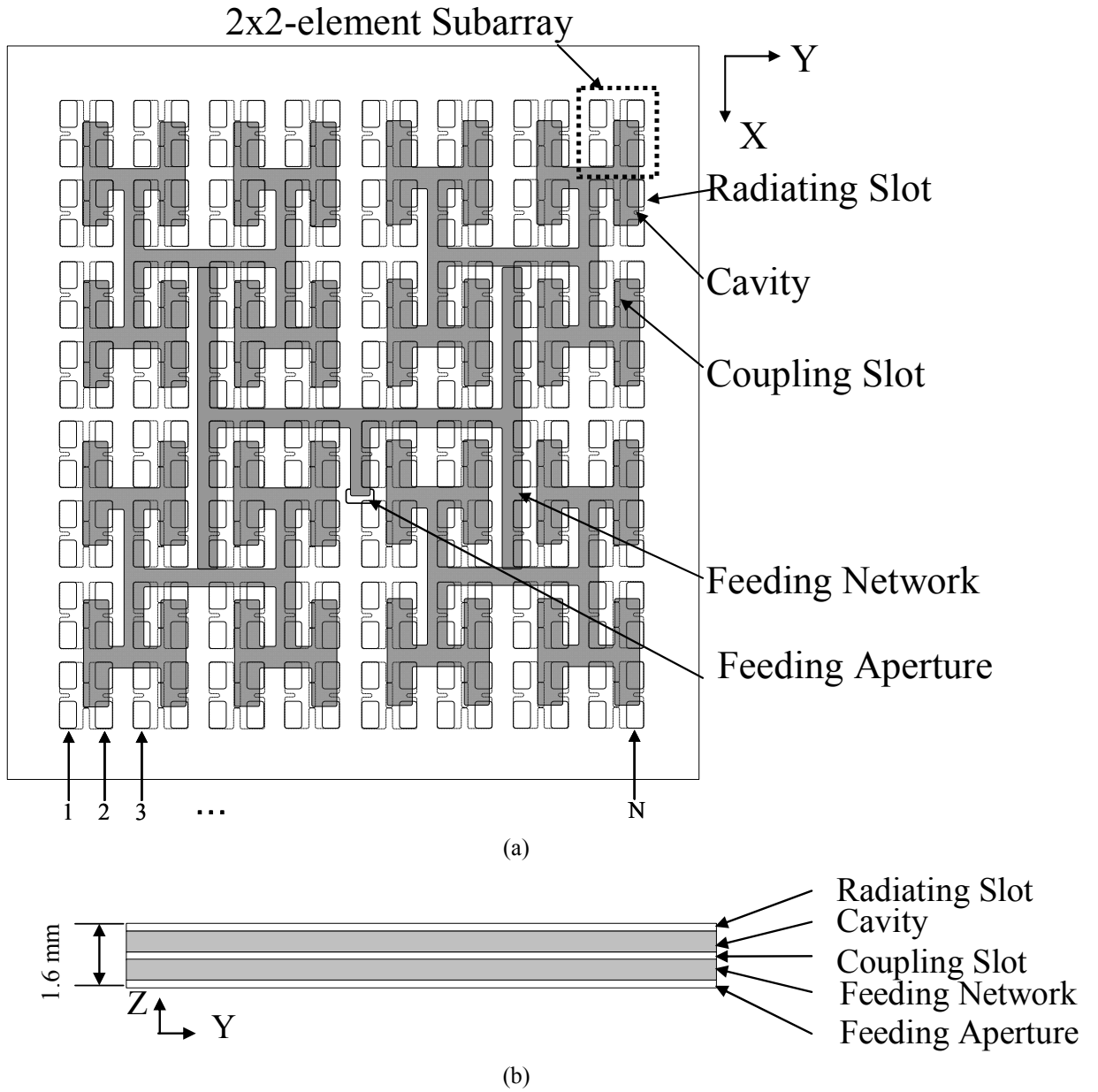


Fig. 1 Antenna configuration of the conventional antenna (a) top-view and (b) side-view.

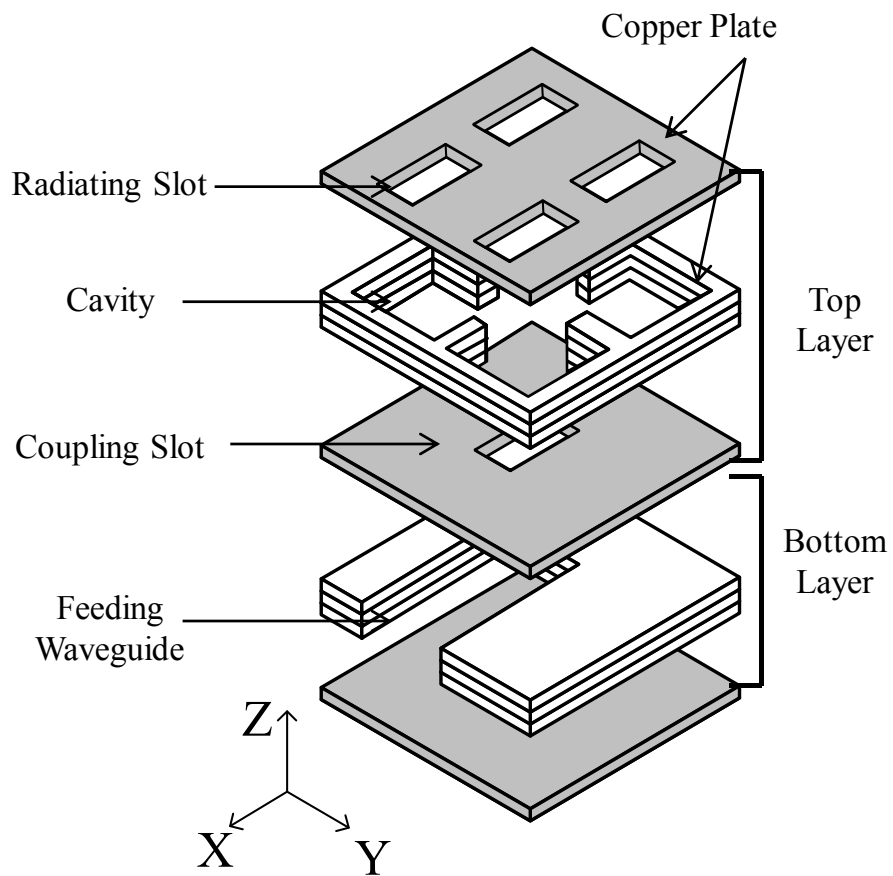
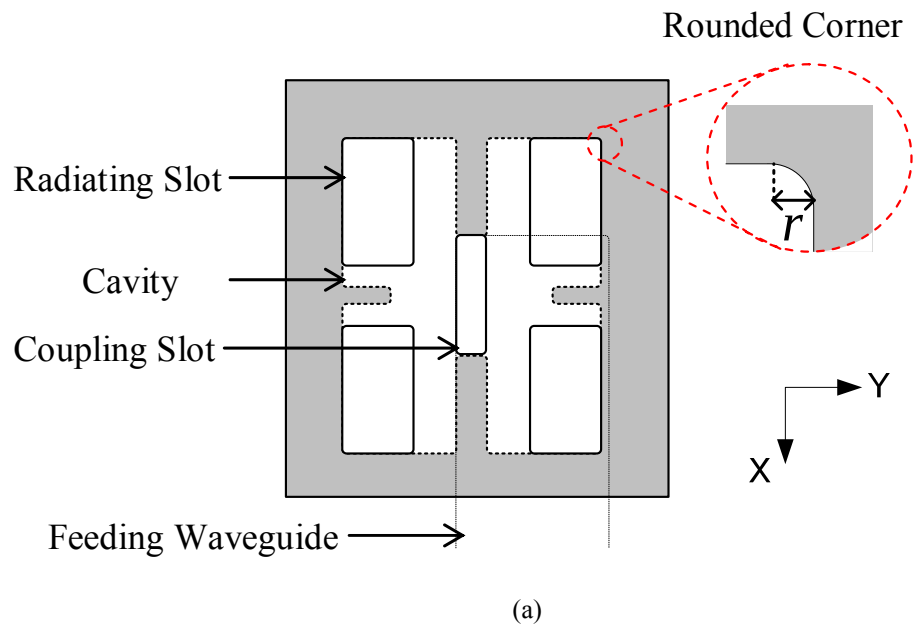


Fig. 2 Configuration of the 2x2-element subarray. (a) top view (b) exploded perspective view

2-2 Antenna Configurations

Fig. 1 shows the configuration of the double-layer corporate-feed hollow-waveguide slot array antenna proposed in [10]. There are 16x16 radiating slots with a 0.86-wavelength interval on the top layer fed by a corporate feed network on the bottom layer. The corporate-feed waveguide consists of several H-plane T-junctions, as marked in gray scale in Fig. 1 (a). The antenna is fed through an input aperture located in the lower layer. All the radiating elements are fed with an equal phase and amplitude by the corporate-feed network. Fig. 2 shows the configurations of the 2x2-element subarray as the design unit to be fed by the full-corporate-feed structure. First, we design this 2x2-element subarray to have good reflection. Then, we try to combine this 2x2-element subarray with the corporate-feed network.

2-3 Arrangement of Limitations of the Diffusion Bonding of Laminated Thin Metal Plates in the High-Frequency Band

2-3-1 Fabrication Process

As mentioned in the previous section, the structure of the antenna for the 120 GHz is based on the previous antenna designed for 60 GHz. However, all parameters cannot be scaled down due to the fabrication method involving diffusion bonding of laminated thin metal plates. In this section, we describe some limitations of the antenna design and fabrication.

The process of fabrication using diffusion bonding is as follows [11-12].

1. Etching desired patterns in copper plates.
2. Stacking the patterns in vacuum with alignment.
3. Applying high heat (approximately 1000°C) with high pressure

The limitations on the antenna design and fabrication are primarily caused by the etching process in Step 1. We describe the reasons of the fabrication errors and limitations in the following section.

2-3-2 Design Limitations in Diffusion Bonding Technique

As shown in Fig. 2(b), we use several thin copper plates to build up the three-dimensional structure using diffusion bonding. We should choose the thickness of the copper plate carefully because it is quite important for the stable fabrication of the antenna. If the thickness of each plate is not thick enough, a deformation might be happen during the fabrication process because we apply high heat and high pressure in the antenna fabrication, especially for a long circuit in the layer of the feed waveguide. In previous research in the 60 GHz-band design, the copper plate has a thickness of 0.3 mm, satisfying both the good performance and the fabrication stability requirements. However, in the case of the 120 GHz band, we cannot use 0.15 mm, which is half the thickness of the antenna in the 60 GHz band due to the stability of the fabrication. Thus, we must redesign the antenna for the 120 GHz band with a plate with a thickness of 0.2 mm. Table 1 shows the specifications of the plate and layer thickness of each antenna for the 60 GHz and 120 GHz bands. Furthermore, the plate thickness much affects to the other parameters in the antenna design. Even though we want to design the corners as a rectangular shape with 90-degree angles, it is impossible to make a rectangular shape for each corner by our etching technology in this moment. Thus, each corner must be rounded, as noted at the upper right of Fig. 2 (a). We call this limitation of design the round corner and the radius (r) of the each corner should be larger than 80% of the plate thickness in our fabrication. The plate thickness and the round corner have not been necessary to consider in the previous studies in the 60 GHz band because the antenna structure is large enough. However, in this research, we must consider these effects because the antenna structure is considerably smaller than that described in our previous research.

Table 1 Layer Thickness of Each Frequency Band

LAYER		60 GHz	120 GHz
Radiating Slot	Thickness	0.3 mm	0.2 mm
	No. of Plates	(0.3 mm x1)	(0.2 mm x1)
Cavity	Thickness	1.2 mm	0.6 mm
	No. of Plates	(0.3 mm x4)	(0.2 mm x3)
Coupling Slot	Thickness	0.3 mm	0.2 mm
	No. of Plates	(0.3 mm x1)	(0.2 mm x1)
Feed Waveguide	Thickness	1.2 mm	0.6 mm
	No. of Plates	(0.3 mm x4)	(0.2 mm x3)

2-4 16x16-element Array Antenna for 120 GHz

2-4-1 2x2-subarray Element Design

As shown in Table 1, the heights of the layers of the antenna are determined to be 0.2 mm for the radiating slots, 0.6 mm for the cavities, 0.2 mm for the coupling slots and 0.6 mm for the feed waveguide. For stable fabrication, the thickness of the copper plate is chosen as 0.2 mm for the 120 GHz band. As a result, the minimum available radius of the round corner is determined 0.16 mm corresponding to 80% of the plate thickness. With these design conditions, the 2x2-element subarray is redesigned for the 120 GHz band.

Basically, this 2x2-element subarray has a wide bandwidth of reflection because of its dual resonance caused by the self-resonance of the radiating slot and the external mutual coupling among the radiating slots. In addition, the ratio between the length and the width of the radiating slot is a very important parameter to determine the bandwidth of the reflection [10]. The round corner strongly affects the bandwidth of the reflection. Because the slot shape is changed according to the radius of the corner (r), the ratio between the width and the length of the slot is also changed. Fig. 3 shows the simulated reflection for various radii of the rounded corner in the 120 GHz band. The bandwidth and the local maximum value of the reflection changes for the radius of the corner.

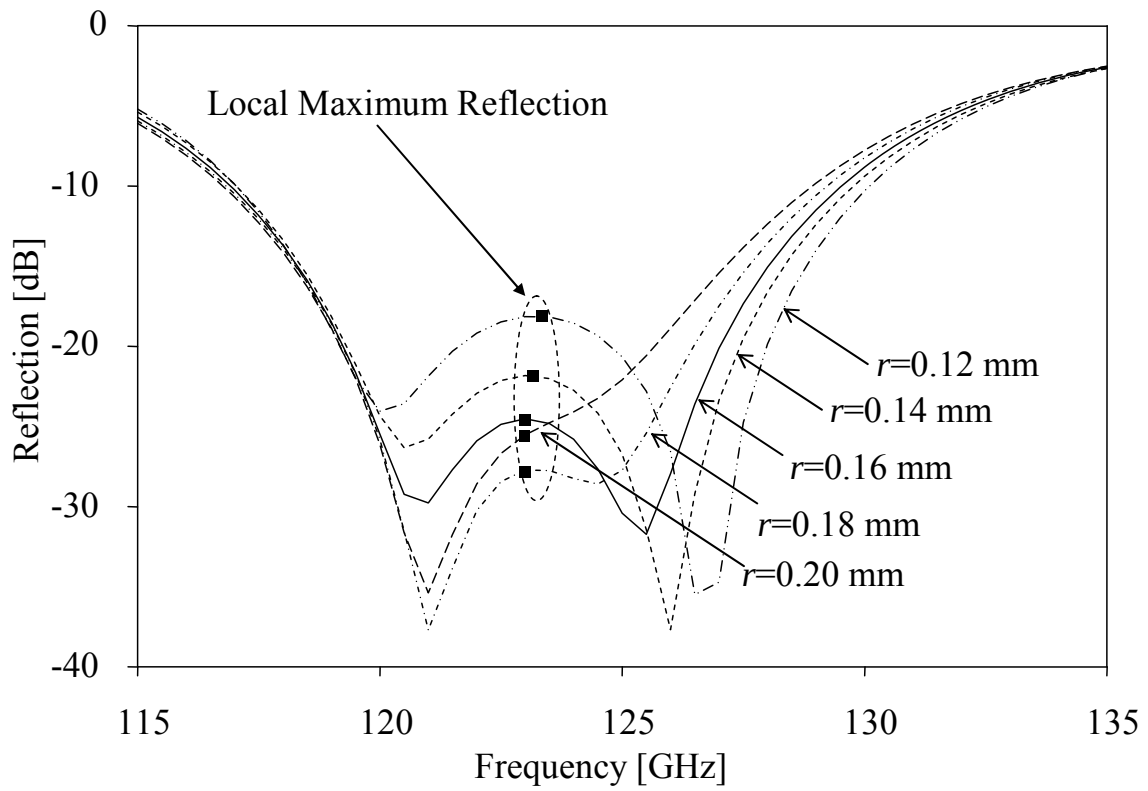


Fig. 3 Simulated reflection of the 2x2-element subarray for various radii of the round corner.

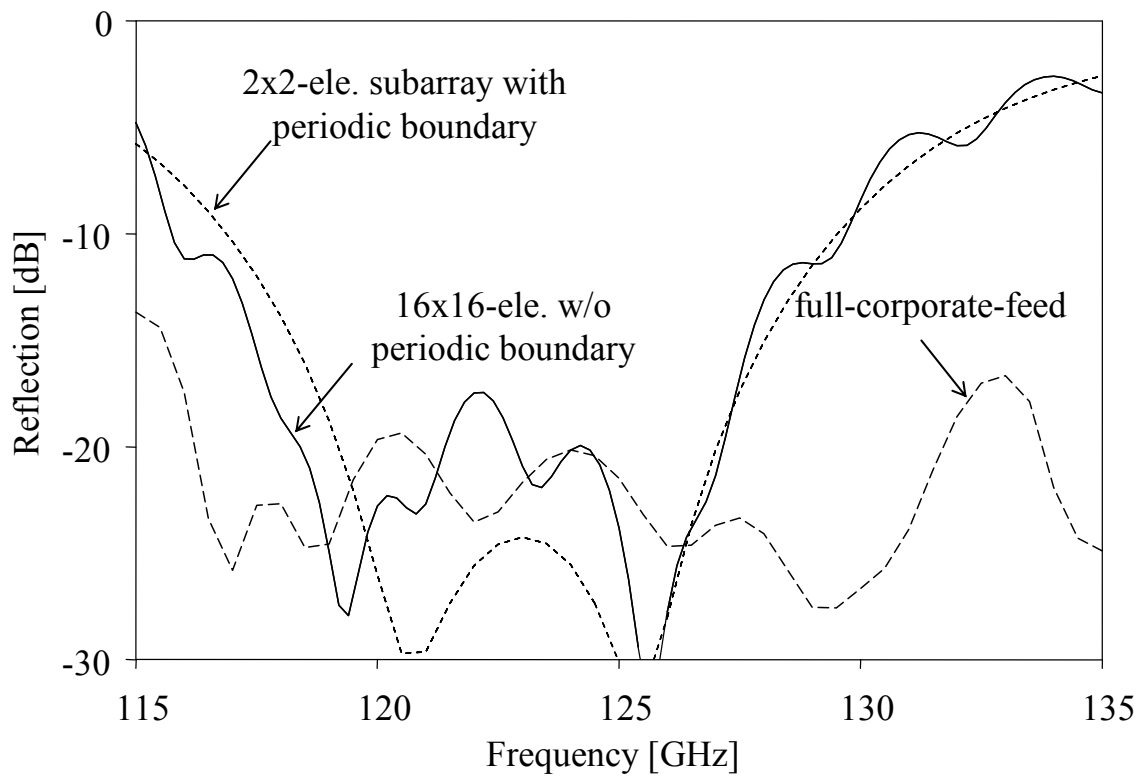


Fig. 4 Calculated reflection of the 16x16-element array.

We could observe a wider bandwidth with a smaller radius of the corner, but the minimum value should be 0.16 mm considering the plate thickness. Thus 0.16 mm is chosen as the radius of the round corner. The specific antenna design parameters are shown in Table 2. The bandwidth of this 2x2-element array is 8.4% for VSWR less than 1.5, and the local maximum reflection is -24 dB as shown in Fig.3. Also, the round corner affects much on the bandwidth of the reflection and the gain of a 16x16-element array; it will be explained in following chapter.

Table 2 Antenna Parameters for the 120 GHz Band

Parameter	Value
Slot spacing in the x and y directions	2.10 mm
Width of the radiating slot	0.88 mm
Length of the radiating slot	1.37 mm
Width of the cavity in the x direction	2.98 mm
Length of the cavity in the y direction	3.47 mm
Width of the cavity wall in the x direction	0.40 mm
Length of the cavity wall in the x direction	1.05 mm
Width of the cavity wall in the y direction	0.38 mm
Length of the cavity wall in the y direction	0.34 mm
Width of the coupling slot	0.50 mm
Length of the coupling slot	1.37 mm
Width of the feed waveguide	1.47 mm
Radius of the corner	0.16 mm

2-4-2 Full Structure Analysis Results

The full structure of the 16x16-element array that combines the 2x2-element subarrays and the corporate-feed waveguide is analyzed by HFSS [13]. The spacing of the radiating slots is 2.1 mm ($=0.86$ wavelengths at the design frequency of 123 GHz) so that the aperture area is 33.6 mm square (2.1 mm x 16 elements). In addition, the total height of the antenna is 1.6 mm. Fig. 4 shows the frequency characteristic of the reflection of the 16x16-element array. The bandwidth is 8.6% for VSWR less than 1.5. This result is well matched with the frequency response of the 2x2-element array model that introduces two sets of periodic boundary walls in the external region to simulate the mutual coupling in an infinite two-dimensional array. However, ripples appear in the reflection by the full simulation, which could come from the corporate-feed network. These ripples are much higher than those of the previous antenna for the 60 GHz band, they are mostly come from the effects of the round corner because it affects much on the bandwidth of each feeding element. Fig.5(a) and (b) show the simulated frequency characteristics of the reflection and the gain of the 16x16-element array depending on the round corner. As seen in these figures, the reflection and the gain characteristics are much changed, depending on the round corner. We can see the widest bandwidth for both of the reflection and the gain with the smallest value (0.12 mm) of the round corner, but it is quite difficult to fabricate by our etching technology at this moment. Thus 0.16 mm is chosen for the radius of round corner because it shows reasonable bandwidths for both of the reflection and the gain.

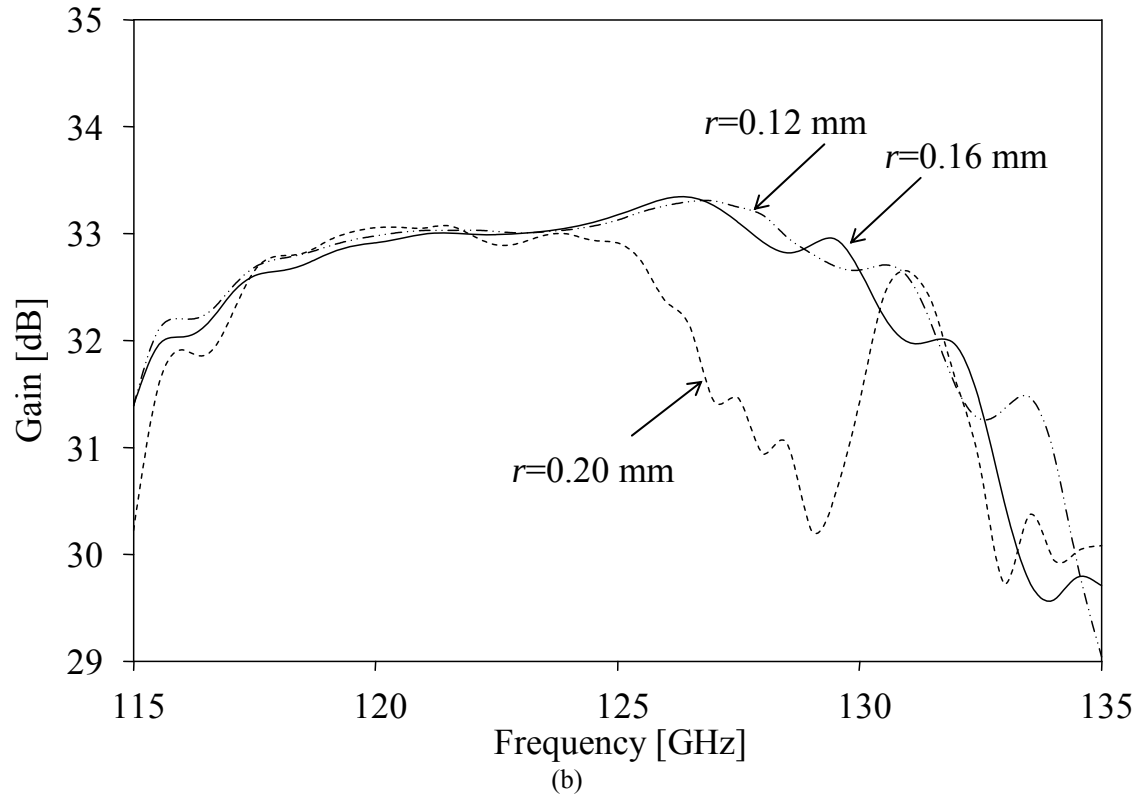
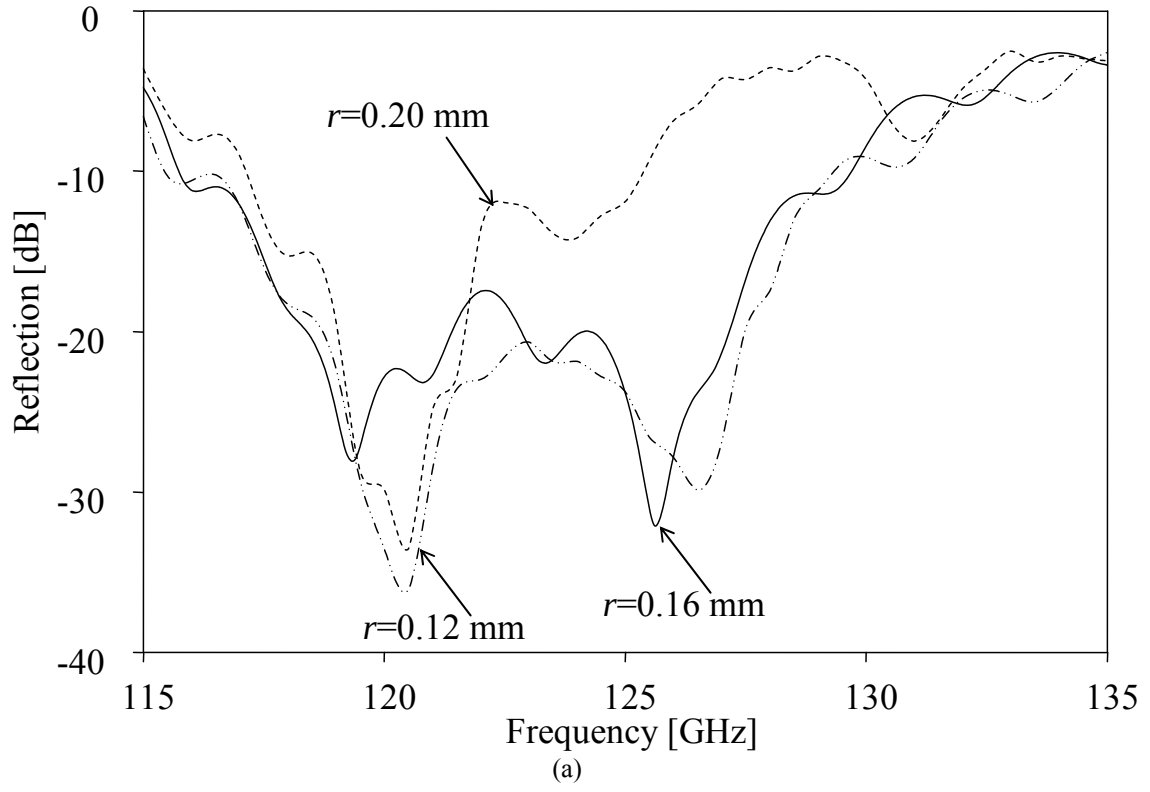


Fig. 5 Simulated reflection and gain of the 16x16-element array for various radii of the round corner.
(a) reflection (b) gain.

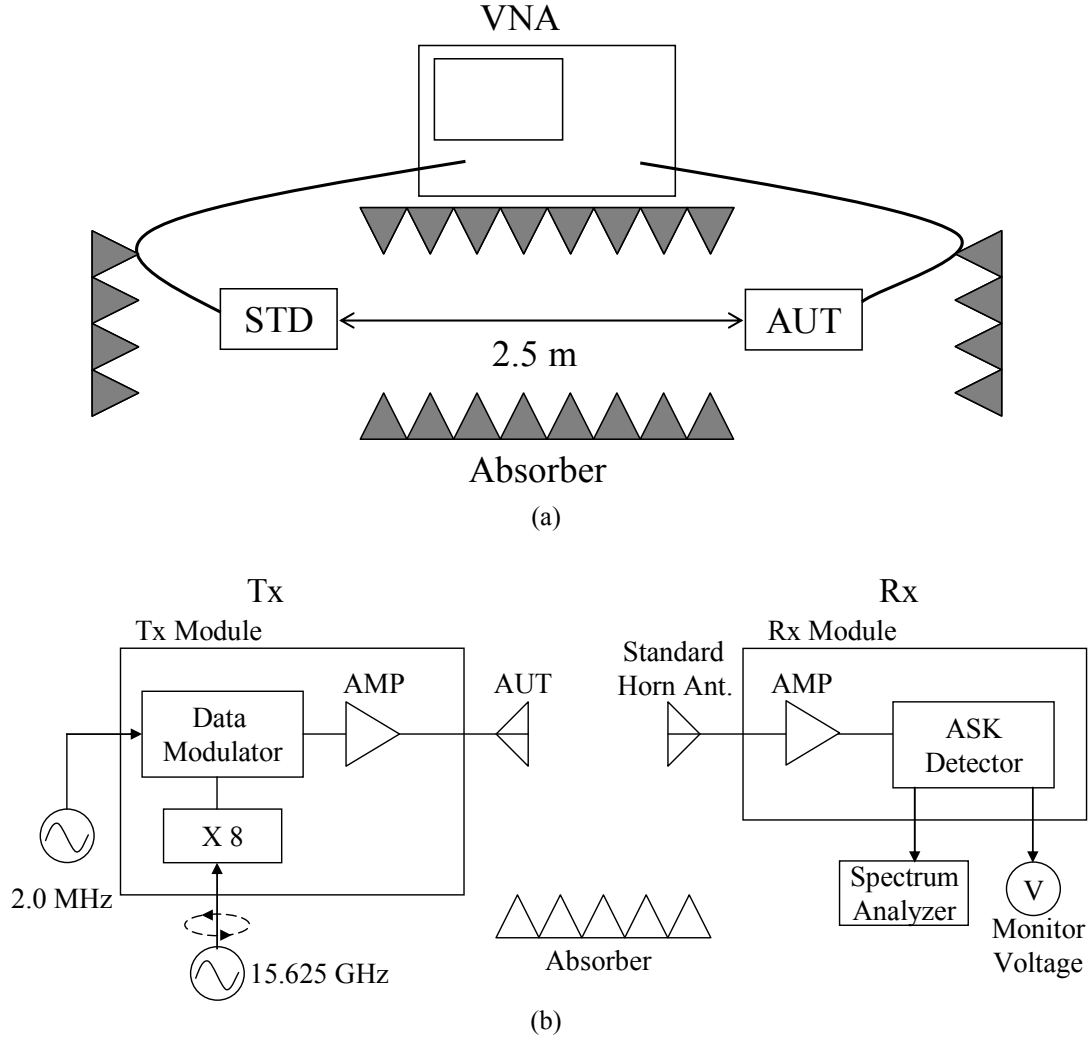


Fig.6 Measurement configurations for 120 GHz band (a) Gain measurement (b) Radiation pattern.

2-4-3 Measurement Configuration

The reflection at the feed point is measured with the VNA (Vector Network Analyzer). Also the antenna gain at the boresite is measured with the VNA at the distance of 2.5 m satisfying the far-field condition. The STD (Standard gain horn antenna) is connected to one port of the VNA. And the STD and the AUT (Antenna Under Test) are alternatively connected to the other port of the VNA. Comparing the S_{21} of the VNA between the STD and AUT, we can determine the gain of the AUT. Fig. 6(a) shows the gain measurement configuration.

The radiation patterns are measured in the measurement system in Fig. 6 (b) which is described in [6]. The Tx module multiplies 15.625 GHz continuous wave (CW) signal from a phase-locked oscillator (PLO) up to 125 GHz and modulates the 125 GHz

CW signal by a 2-MHz CW signal. The output of the TX module is radiated from a horn antenna or AUT. In Rx, the received RF signals are amplified by a low-noise amplifier (LNA) and demodulated by an ASK demodulator. The Rx module outputs the demodulated data signal and monitor voltage, which depends on the intensity of the received MMW signal. The intensity of the demodulated 2 MHz CW signal is measured by a spectrum analyzer. Only an antenna rotator is added in the measurement system in [9] for the radiation pattern measurement. The radiation patterns also measured in the distance of 2.5m. The measurement is performed in a quasi-anechoic environment surrounded by absorbers. However, the measurement results show some ripples due to the imperfect shielding.

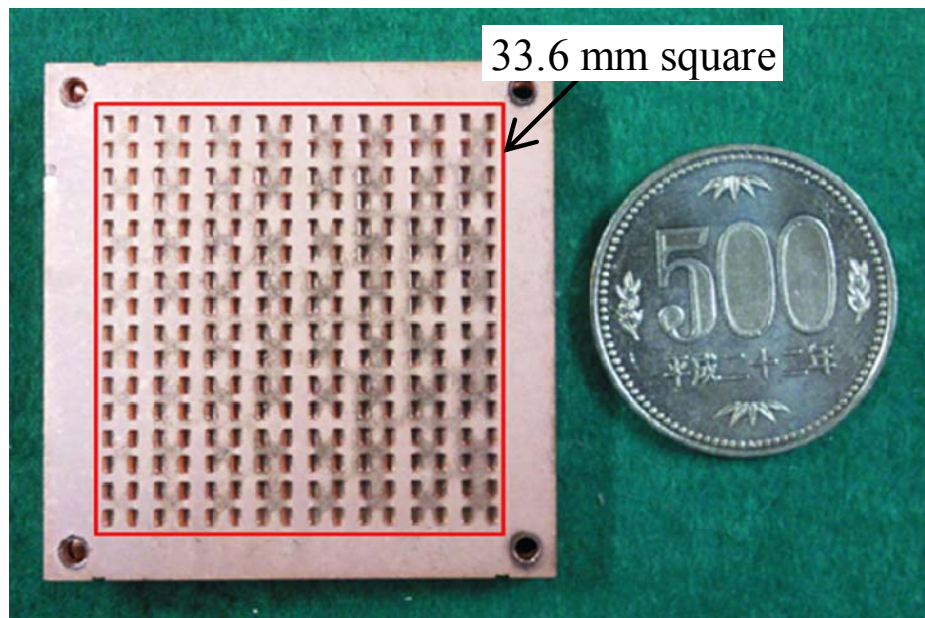


Fig. 7 Picture of the fabricated antenna for 120 GHz

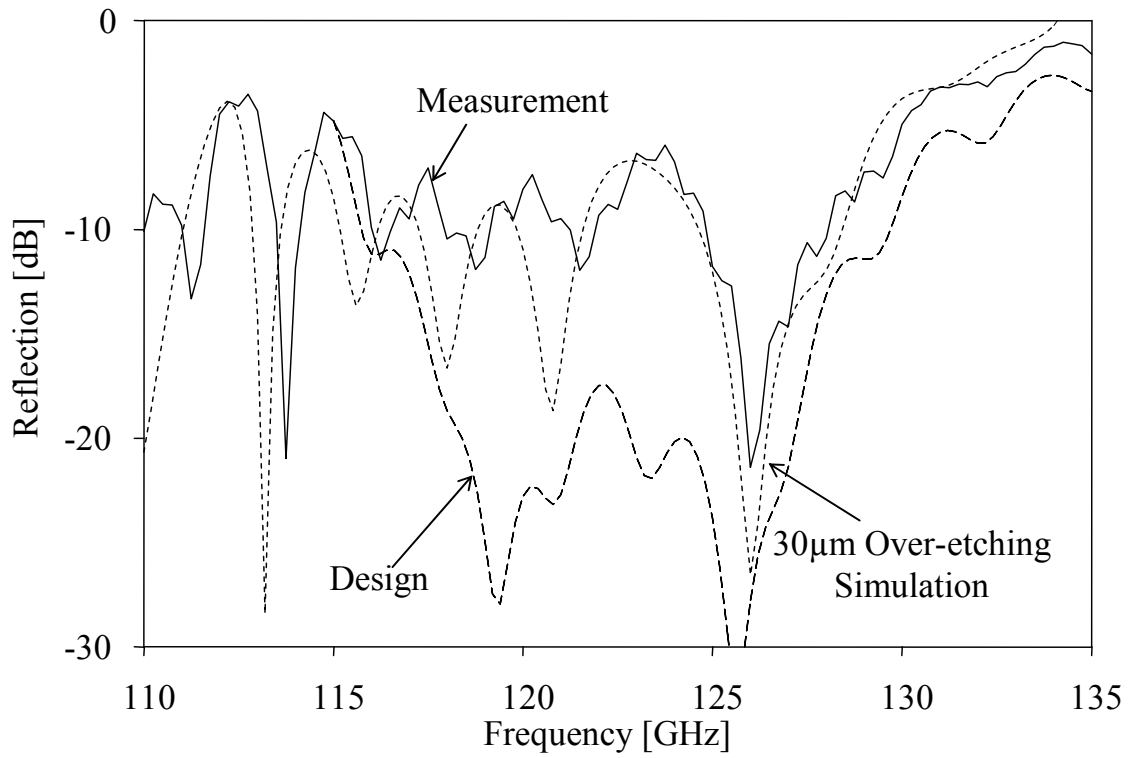


Fig. 8 Frequency behavior of reflection for the 120 GHz antenna.

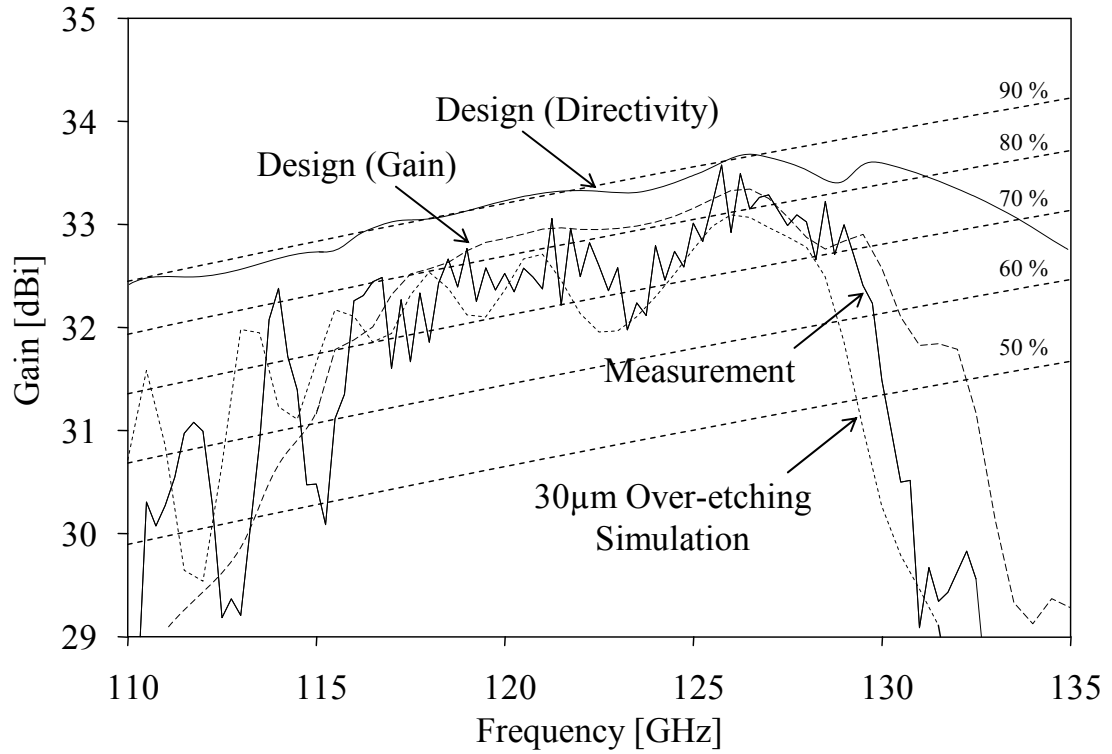


Fig. 9 Frequency behavior of the directivity, gain and efficiency of the 120 GHz antenna.

2-4-4 Fabrication and Measurement Results

Fig. 7 shows the fabricated 120 GHz band antenna. Fig. 8 shows the frequency characteristic of the reflection at the feed aperture. In the simulation, the bandwidth of the reflection for VSWR less than 1.5 is 8.6%. However, the measured reflection is quite different from the designed result. There are various factors to give the discrepancy. Some of them are “over/under etching of the slotted plates themselves”, “the shape of the cross section of the slotted plates after etching” and “misalignment in laminating the slotted plates”. Also there are so many parameters in the design. It is quite difficult to evaluate each of these factors and parameters separately at this moment. However the simulated result assuming 30 μm of over-etching equivalently for all layer patterns shows good agreement with the measured result as shown in Fig.8. This is very valuable from the practical engineering point of view. Fig. 9 shows the frequency characteristic of the gain and the antenna efficiency measured with VNA. The gain is measured at intervals of 0.25 GHz from 110 GHz to 135 GHz. The measured gain is compared with a standard gain horn. In the simulation, the peak of the gain including the conductor loss and reflection is 33.1 dBi with 84% antenna efficiency at 126.5 GHz. In the measurement, we have over 70% antenna efficiency over a broad bandwidth from 117 GHz to 129 GHz. In addition, the 30 μm equivalent over-etching simulation result is added to Fig. 9 which shows a good agreement with the measured result. The ripples in the measured antenna gain came from the instability of the measurement environment, including shielding problems and a short measurement distance. The conductor loss and the insertion loss are included in the gain for the both of the simulation and measurement. In calculation using HFSS, the conductivity loss at 123 GHz is 0.36 dB while it is 0.26 dB in the 60 GHz band when the conductivity of copper is assumed to be $5.8 \times 10^7 \text{ S/m}$ [10].

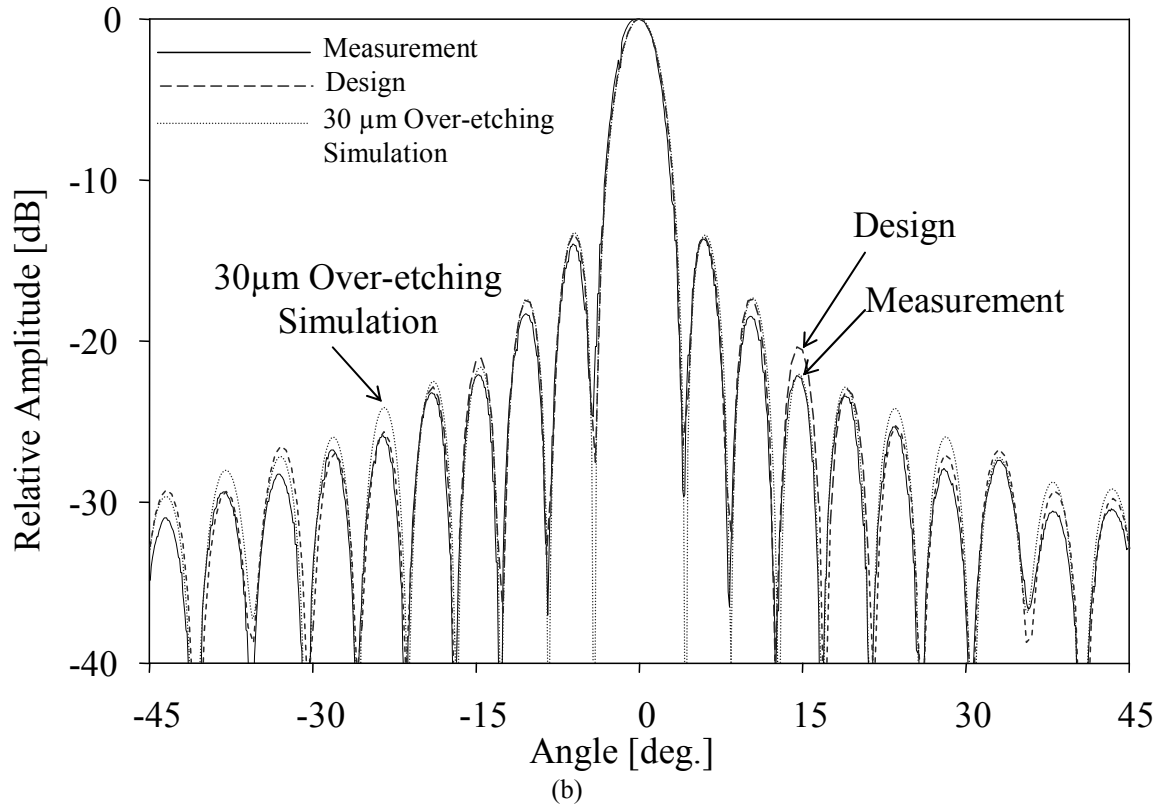
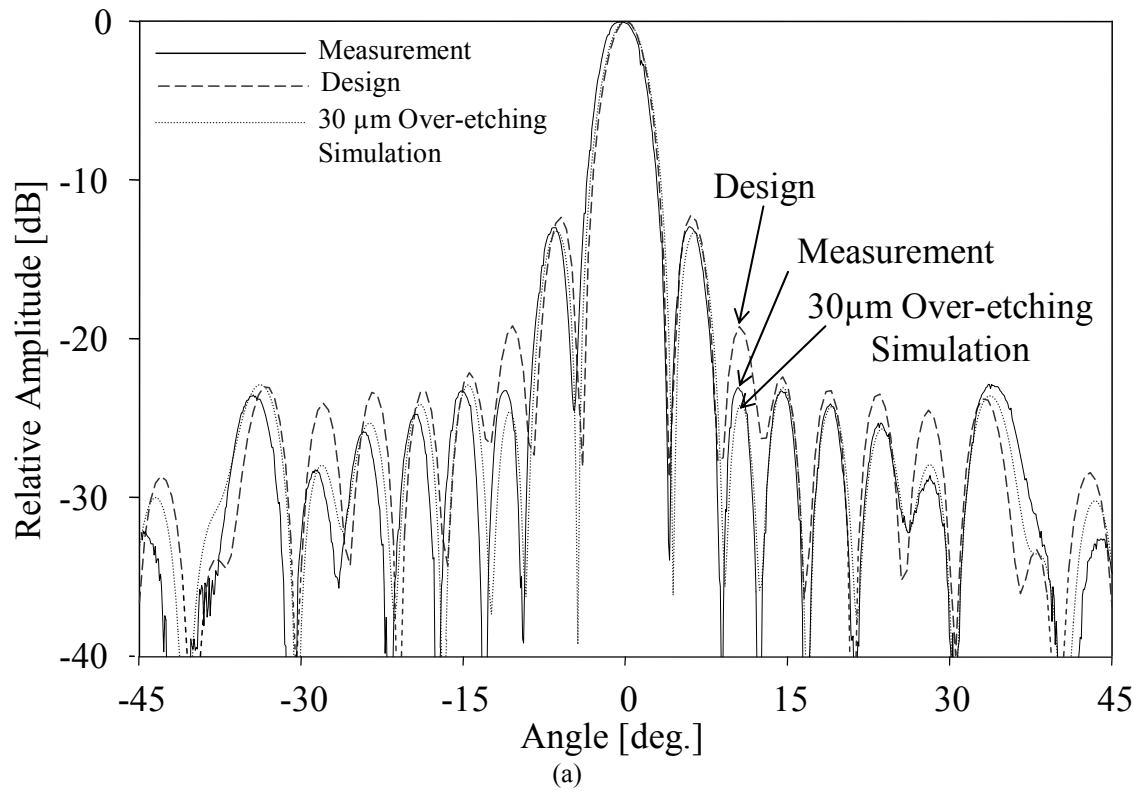


Fig. 10 Radiation patterns. (a) E-plane (b) H-plane at 123 GHz.

The radiation patterns at 123 GHz are measured in the E (yz)- and the H (xz)-planes. Fig. 10 (a) shows the E-plane patterns of the proposed antenna. The results of the design, the measurement and the simulation including 30 μm equivalent over-etching are shown in the same figure. The half-power beam width of 4.0 degrees shows good agreement among the three results; however, the second side lobe level is slightly different. The measurement and simulation including equivalent over-etching shows good agreement at the level of -25 dB, but the design shows a higher level of the side lobes. The radiation patterns in the H-plane also show the same patterns as the E-plane. The measured and simulated half-power beam width is 3.5 degrees in the H-plane.

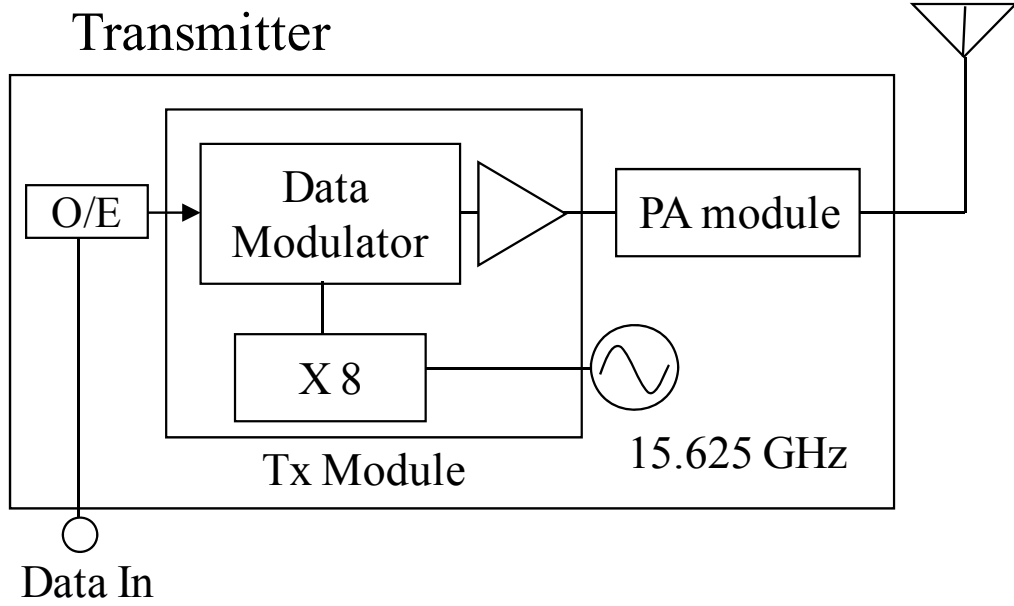
Because fabrication errors, such as over-etching, significantly affect the antenna performance at high frequency, such as 120 GHz, we should consider the fabrication error properly when we design antennas using the diffusion bonding method.

2-5 Data Transmission Experiments in 120 GHz Band

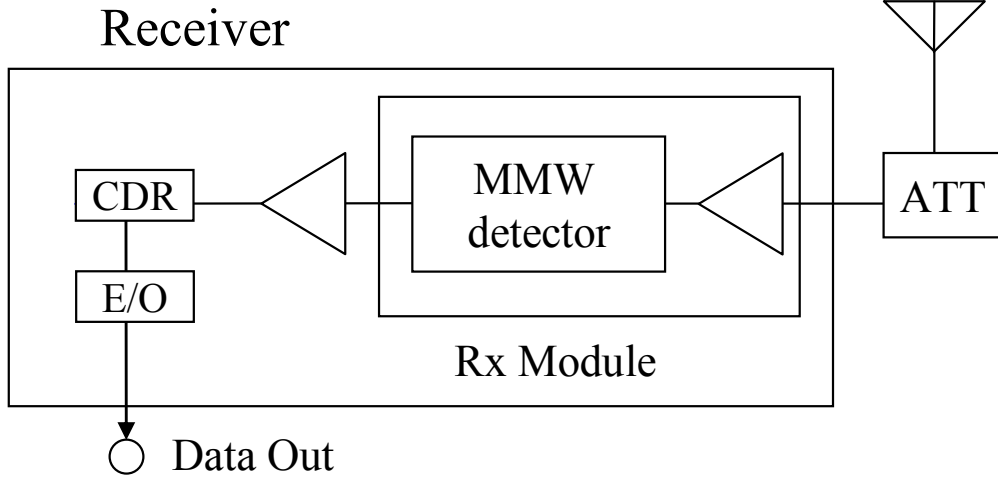
2-5-1 Data Transmission Experimental System for 120 GHz

We perform data transmission experiment with the proposed antenna and the standard horn antenna in the 120 GHz band. The aim of this experiment is to demonstrate the feasibility of the proposed antenna as an alternative of conventional high gain antennas such as a parabolic antenna. In this chapter, we will introduce the experimental configuration and discuss the experimental results.

Fig. 11 shows the configurations of the data transmission tests for the 120 GHz band. The measurement system is basically same with the measurement systems in Fig. 6. Only several modules are added to transmit and to receive the data streams. On the Tx, a power amplifier (PA) module and an optical to electrical converter (O/E) are added. Fig. 11(a) shows the schematic of the Tx system for the data experimental system. Also, on the Rx, a variable attenuator, a clock and data recovery (CDR) and an electrical-to-optical converter (E/O) are inserted. Fig. 11(b) shows the schematic of the Rx system for the data experimental system. For the data signal, we use a $2^{23}-1$ pseudorandom pattern bit stream (PRBS) from a pulse pattern generator (PPG). The data signal from the PPG is converted from O/E and input to the Tx module. The output of the Tx module is input to the PA module. The PA module amplifies the MMW signals up to 10 dBm. The output of the PA module is radiated from a standard horn antenna that has 23.3 to 23.9dBi gain from 110 GHz to 140 GHz. For the Rx system, the proposed antenna and the standard horn antenna are alternatively employed to compare their characteristics. The Rx output from the Rx module is converted from E/O and input to CDR, and the bit error rate (BER) is measured. The received power is controlled by changing the attenuation factor of the attenuator (ATT) in this system. We can evaluate the antenna performance using the attenuator values of ATT because the BER is almost determined by the power which is entered to the MMW detector in this system. The distance between Tx and Rx is fixed at 2.5 m.



(a)



(b)

Fig. 11 Schematics of the 120 GHz experimental system. (a) transmitter and (b) receiver. (PA: power amplifier, O/E: optical to electrical converter, E/O: electrical to optical converter, CDR: clock and data recovery, ATT: attenuator, MMW: millimeter wave, PPG: pulse pattern generator)

2-5-2 Data Transmission Experimental Results for 120 GHz

In this section, we perform the data transmission experiments for the data rates of 5 Gbit/s and 10 Gbit/s, respectively. The carrier frequency of the system is fixed at 125 GHz at first. Fig.12(a) shows the BER dependences as a function of the attenuation factor for each antenna in Rx.

We achieve the BER less than 10^{-12} which ensure the error free data

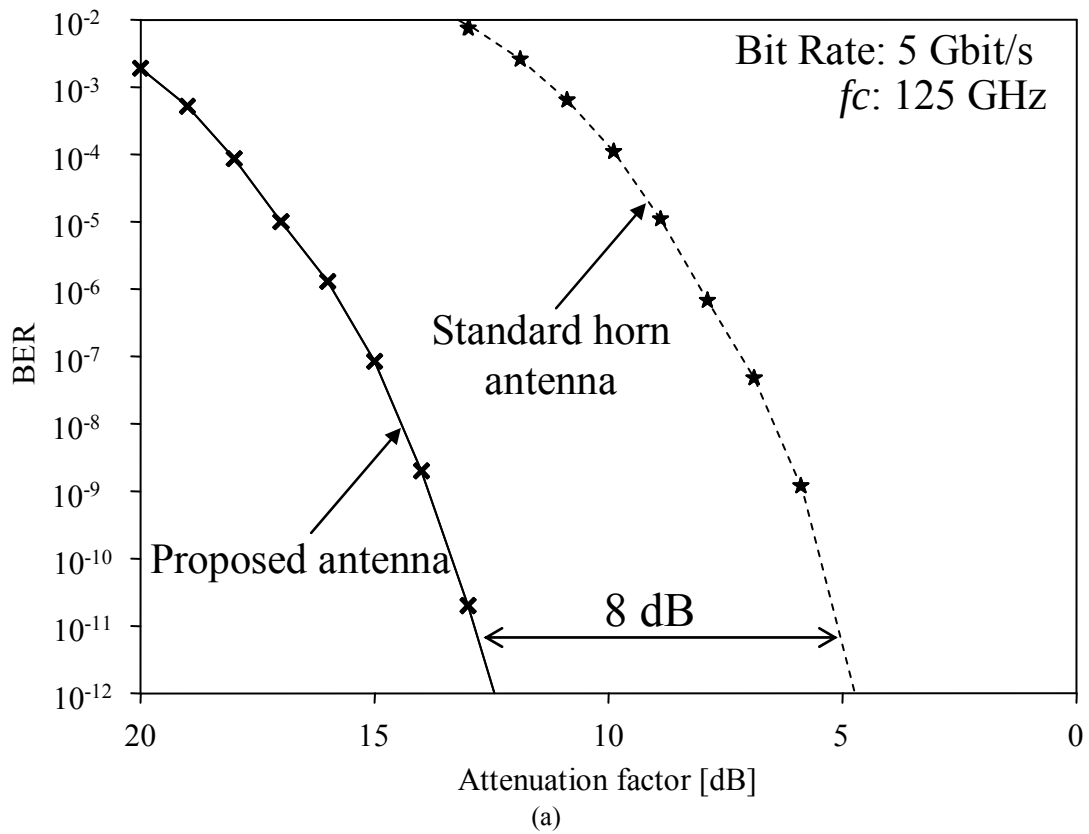
transmission using the both of the proposed antenna and the standard horn antenna at the data rate of 5 Gbit/s. However, the attenuation factors are quite different for each antenna to obtain the same level of BER. The attenuation factors for BER less than 10^{-12} are 12 dB for the proposed antenna and 4 dB for the standard horn antenna, respectively. This result means that we can obtain additional 8 dB transmission margin when we use the proposed antenna for the wireless link system instead of the standard horn antenna. It confirms that the proposed antenna has 8 dB higher gain than the standard horn antenna; The gain of the proposed antenna is about 32 dBi while it is only around 23.6 dBi of a standard horn antenna near the center frequency.

Also we observe the BER characteristics of two antennas for the data rate of 10 Gbit/s. Fig.12(b) shows the BER dependences as a function of the attenuation factor for the data rate of 10 Gbit/s. We also achieve the BER less than 10^{-12} for the both antennas. However, the difference of attenuation factors is only 6 dB for the BER less than 10^{-12} between the two antennas while the difference of gain is more than 8 dB at center frequency. This degradation of 2 dB in attenuation level comes from the lack of uniformity of the gain over the required RF bandwidth of the wireless link system. Since the required RF bandwidth of this wireless link system which uses ASK modulation scheme is about 1.5 times of the data rate, the required RF bandwidth for the 10 Gbit/s is approximately 15 GHz (117.5 to 132.5 GHz). The gain of the proposed antenna is sharply decreased around 130 GHz while the required RF bandwidth of the system is up to 132.5 GHz for the data rate over 10 Gbit/s; the transmitted signal might be distorted in the higher frequency region over 130 GHz due to the antenna gain characteristic. Therefore the difference of attenuation factor is reduced between two antennas because a standard horn antenna has almost flat antenna gain characteristic over the required bandwidth.

To avoid the degradation of the attenuation level of the proposed antenna, we perform the 10 Gbit/s data transmission tests with the carrier frequency of 123 GHz. Because the required RF bandwidth for the 10 Gbit/s is from 115.5 to 130.5 GHz for the carrier frequency of 123 GHz, it well matches with the antenna gain characteristic of the proposed antenna. Fig.12(c) shows the BER dependences as a function of the attenuation factor for each antenna when the carrier frequency is 123 GHz. Here, we can observe the 8 dB difference of the attenuation level for each antenna which well reflects

the 8 dB higher gain of the proposed antenna than the standard horn antenna. Note that, the overall test results with the carrier frequency of 123 GHz are a little bit degraded compared with the test results with the carrier frequency of 125 GHz since the data transmission system is mainly designed for 125 GHz.

In the data transmission experiments, we confirm the feasibility of the proposed antenna for the wireless link system. Even though, we experience the degradation of the transmission performance due to the lack of bandwidth in the data rate of 10 Gbit/s with the carrier frequency of 125 GHz, this degradation is mainly caused by the frequency shift due to the fabrication error. We could expect the better performance if we fabricate this antenna with considering the fabrication errors such as over-etching because the calculated gain shows the sufficient and smooth gain characteristics for the 10 Gbit/s data transmission.



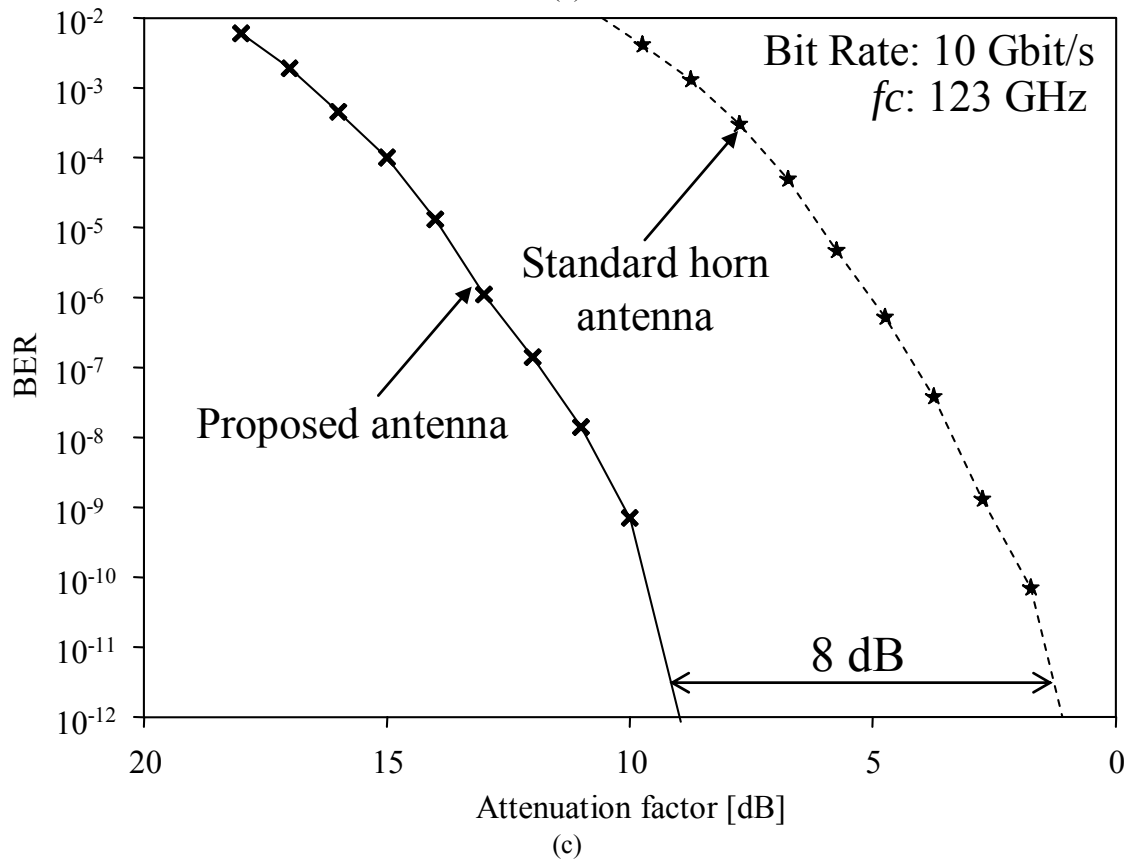
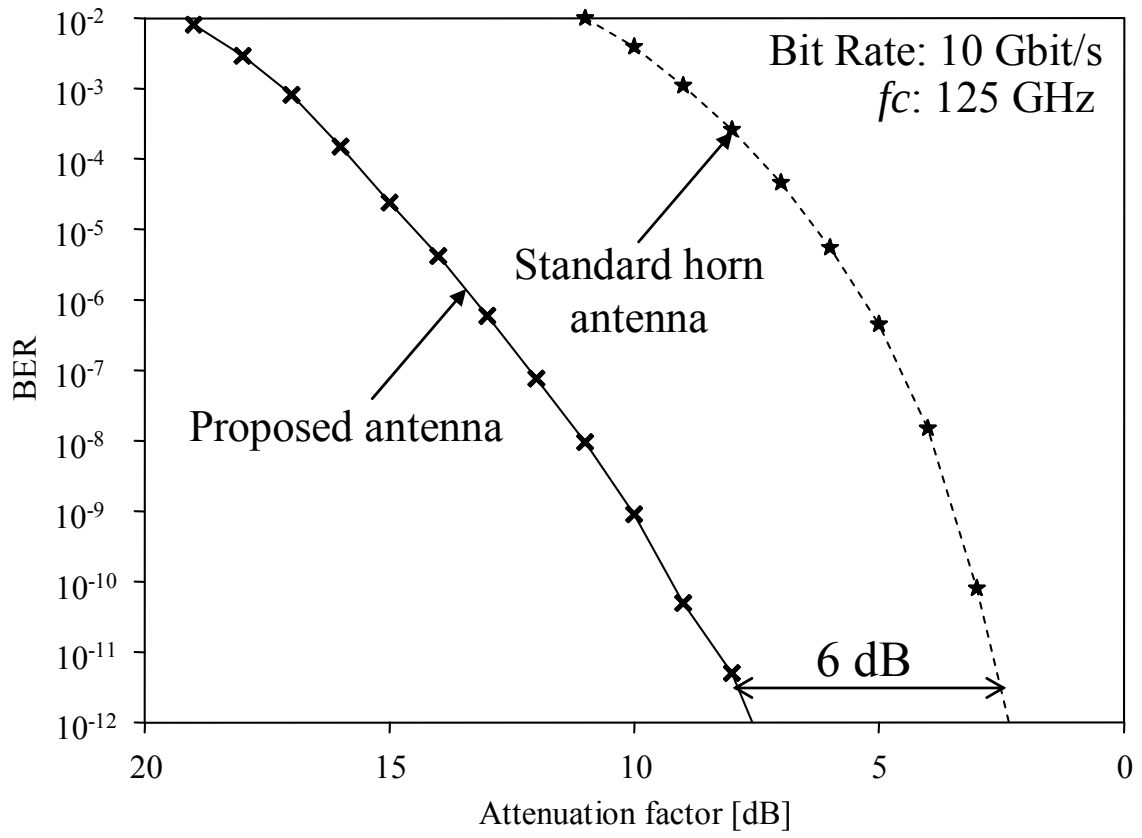


Fig. 12 Measured BERs for various attenuation factors in 120 GHz band

2-6 16x16-element Array Antenna for 350 GHz

2-6-1 2x2-element Subarray Design

The 2x2-subarray element is also designed for the 350 GHz band. 0.2mm-thickness plates are chosen for the cavity and the feeding waveguide. However the plate thickness for the radiating and the coupling slots is chosen 0.1mm not 0.2mm to have wider bandwidth of the reflection by degrade the stable fabrication slightly. Fig. 13 shows the frequency characteristics of the 2x2-element subarray depend on the plate thickness of the radiating slots and the coupling slots. If we use 0.2mm thickness-plate for the all plates, the bandwidth is degraded to 3.4 % while the bandwidth is 8.5 % if we use 0.1 mm thickness-plate for the radiating and coupling slots. The working mechanism and basic characteristic of this 2x2-element is identical with that of the 120 GHz. In this case, the values of the minimum available radius of the rounded corner is also determined as 0.16mm same as the 120 GHz antenna because we use both of 0.1 mm and 0.2 mm thickness-plate in same time. The parameters for 2x2-subarray element are presented in Table 3.

2-6-2 Full Structure Analysis Results

The 16x16-element array also designed for the 350 GHz band. The spacing of the radiating slots are 0.70 mm (= 0.82 wavelengths at the design frequency of 350GHz) so that the aperture area is 11.2 mm square (0.7 mm x 16 elements). And the total height of the antenna is 0.6 mm. Fig.5 shows the frequency characteristic of the reflection of the 16x16-element array of the 350 GHz antenna. The bandwidth of the 16x16-element array is 5.6% while the bandwidth of the 2x2-elements subarray is 8.5 % for VSWR less than 1.5. The cause of the narrow bandwidth of the 16x16-element array is the ripples by the feeding network. While the parameters of the 350GHz band antenna are substantially smaller than those of the 120 GHz band antenna, the radius of the corner and the minimum available value in design is same to the 120 GHz band as 0.16mm due to the plate thickness of 0.2mm. Thus the multi-reflection is occurred in the feeding network because the bandwidth of each feeding component such as T-junction, H-junction and the input aperture is much reduced.

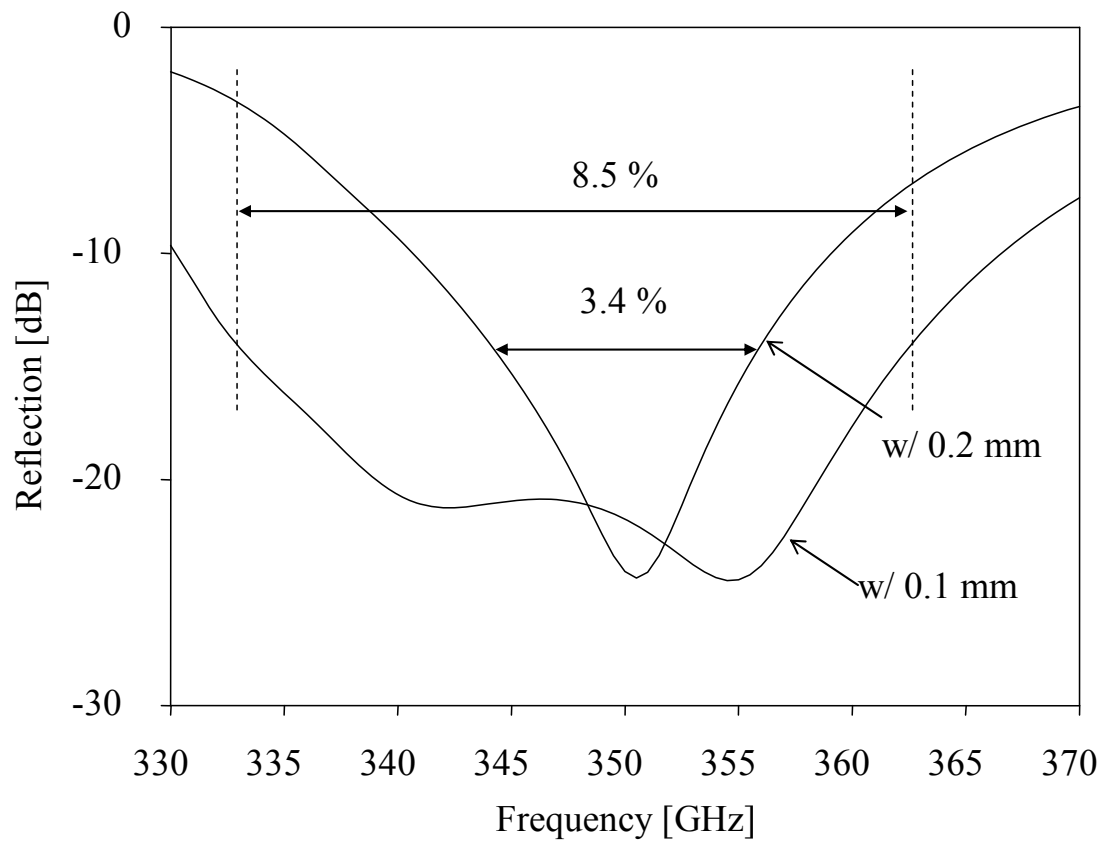


Fig. 13 Calculated reflection coefficients of the 2x2-element subarray in the 350 GHz band.

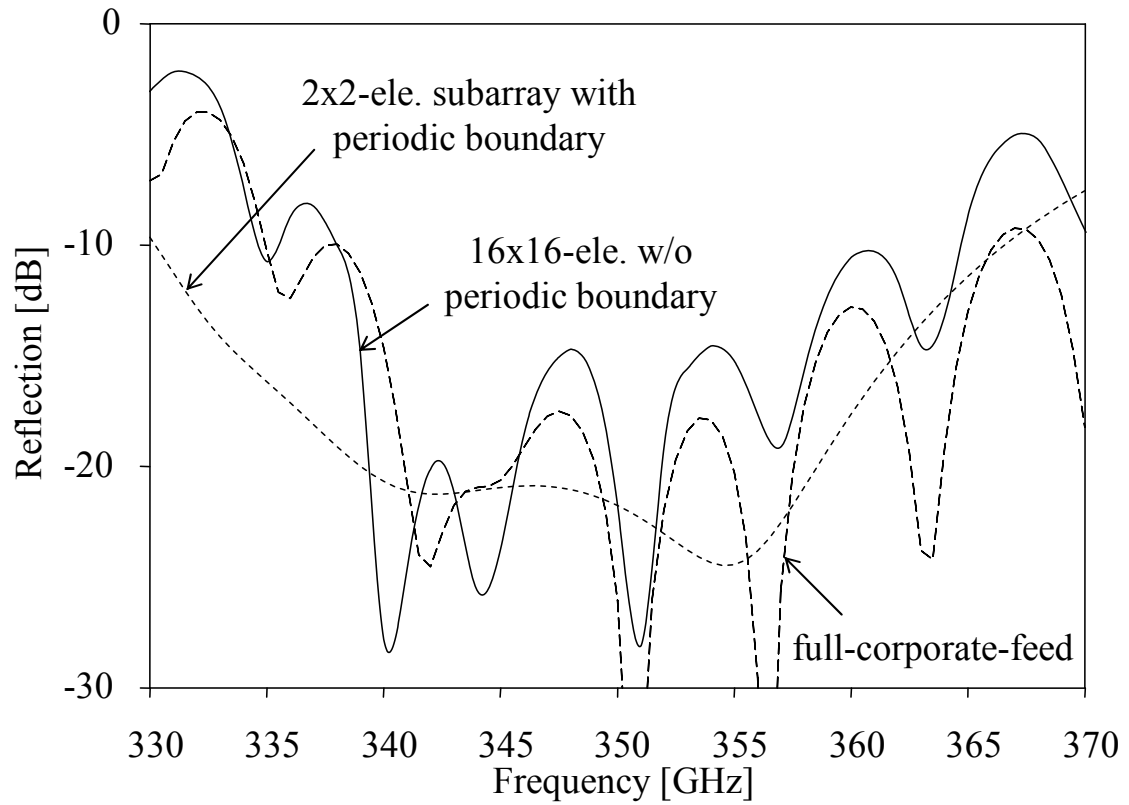


Fig. 14 Calculated reflection coefficients of the 16x16-element array in the 350 GHz band.

Table 3 Antenna Parameters for the 350 GHz Band

Parameter	Value
Slot spacing in the x and y directions	0.70 mm
Width of the radiating slot	0.32 mm
Length of the radiating slot	0.56 mm
Width of the cavity in the x direction	1.02 mm
Length of the cavity in the y direction	1.26 mm
Width of the cavity wall in the x direction	0.16 mm
Length of the cavity wall in the x direction	0.43 mm
Width of the cavity wall in the y direction	0.16 mm
Length of the cavity wall in the y direction	0.16 mm
Width of the coupling slot	0.32 mm
Length of the coupling slot	0.56 mm
Width of the feed waveguide	0.49 mm

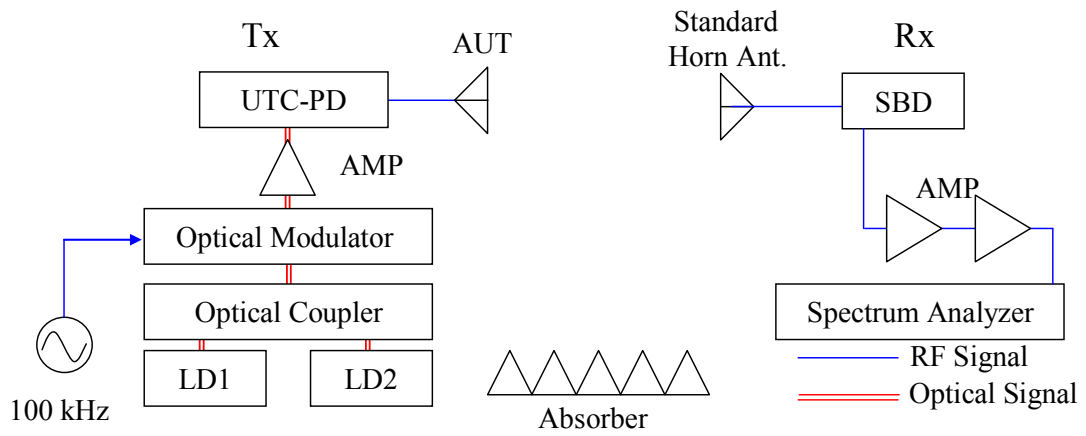


Fig. 15. Measurement configurations using optical devices for 350 GHz band.

2-6-3 Measurement Configuration

The photonic devices are used for the antenna measurement in 350 GHz band. Fig.15 shows the measurement system using photonic devices [14]. First, an optical signal is generated by heterodyning lights with two wavelengths from wavelength-tunable light sources. It is digitally modulated by an optical intensity modulator driven by a continuous wave signal. Finally, it is converted to an electrical signal by a uni-traveling-carrier- Photodiode (UTC-PD) and it is emitted to the free space via a horn antenna or an antenna under test (AUT). The receiver consists of a Schottky barrier diode (SBD) followed by a low-noise pre-amplifier and a limiting amplifier. Using a spectrum analyzer, we can measure the magnitude of the input signal via the SBD. Comparing with a standard horn antenna and AUT, we can find out the antenna gain. The measurement distance is 1m between Tx and Rx and the antenna height is 0.3 m on the ground. The measurement is performed in a quasi-anechoic environment surrounded by absorbers. However, the measurement results show some ripples due to the imperfect shielding.

2-6-4 Fabrication and Measurement Results

The antennas for 350 GHz-band are also fabricated. Fig.16 shows one of the fabricated antennas. In this frequency band, three antennas are fabricated by adding $\pm 10 \mu\text{m}$ into all the designed parameters, assuming over/under etching. Only the antenna gain is measured in the 350 GHz band. Fig.17 shows the frequency characteristics of the gain and the antenna efficiency. The gain is measured at intervals of 3 GHz from 320 GHz to 380 GHz. The frequency dependence is shifted qualitatively lower by increasing all the parameters among the three antennas. Even though we fabricate the three antennas considering with over/under etching, the over-etching is occurred more than we expect. Also various values of over-etching are simulated to find out the over-etching value in the 350 GHz band and the results are added in the Fig. 12. The measured gain value of $-0 \mu\text{m}$ is well matched with over-etching simulation assumed $+30 \mu\text{m}$ as shown in Fig. 12 (b). Similarly, the measured gain values of $-10 \mu\text{m}$ and $+10 \mu\text{m}$ are well matched with over-etching simulation assumed $+20 \mu\text{m}$ and $+40 \mu\text{m}$ as shown in Fig. 12 (a) and 12 (c), respectively.

Thus we confirm that the over-etching value in the 350 GHz band antennas is also 30 μm . In design, the peak of the gain including the conductor loss and reflection is 32.25 dBi with 75% antenna efficiency at 350 GHz. But in measurement, we have 32.0 dBi of the gain with 74% antenna efficiency on the fabricated antenna assumed +10 μm over-etching value at 350 GHz. The measurement antenna efficiency of the antenna assumed +10 μm over-etching value shows over 40% from 335 GHz to 353 GHz.

Since the fabrication errors such as over-etching much effect on the antenna performance in high frequencies such as 120 GHz and 350 GHz, we should consider the fabrication error properly when we design antennas using the diffusion bonding method.

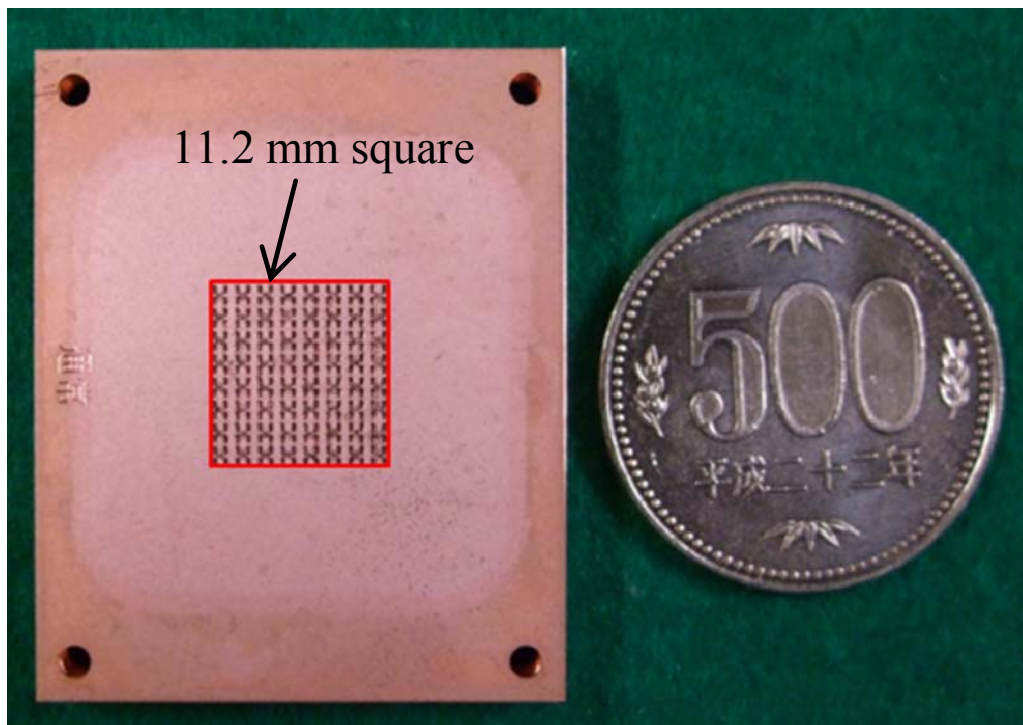
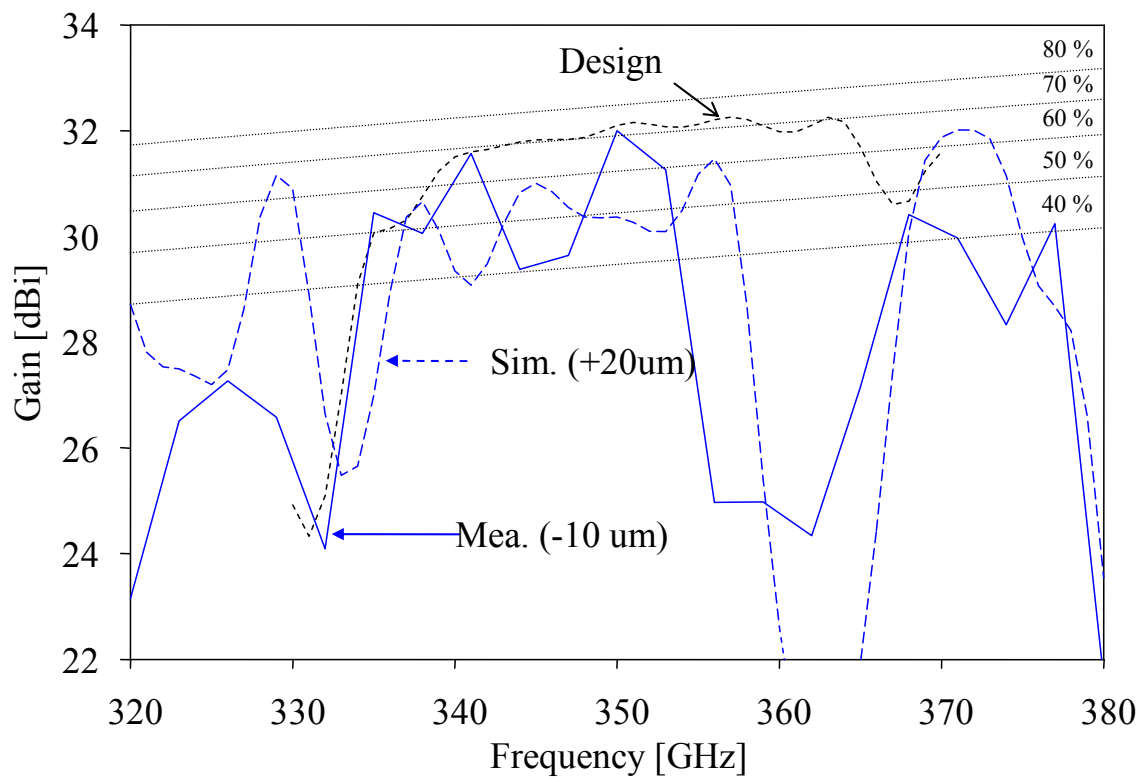
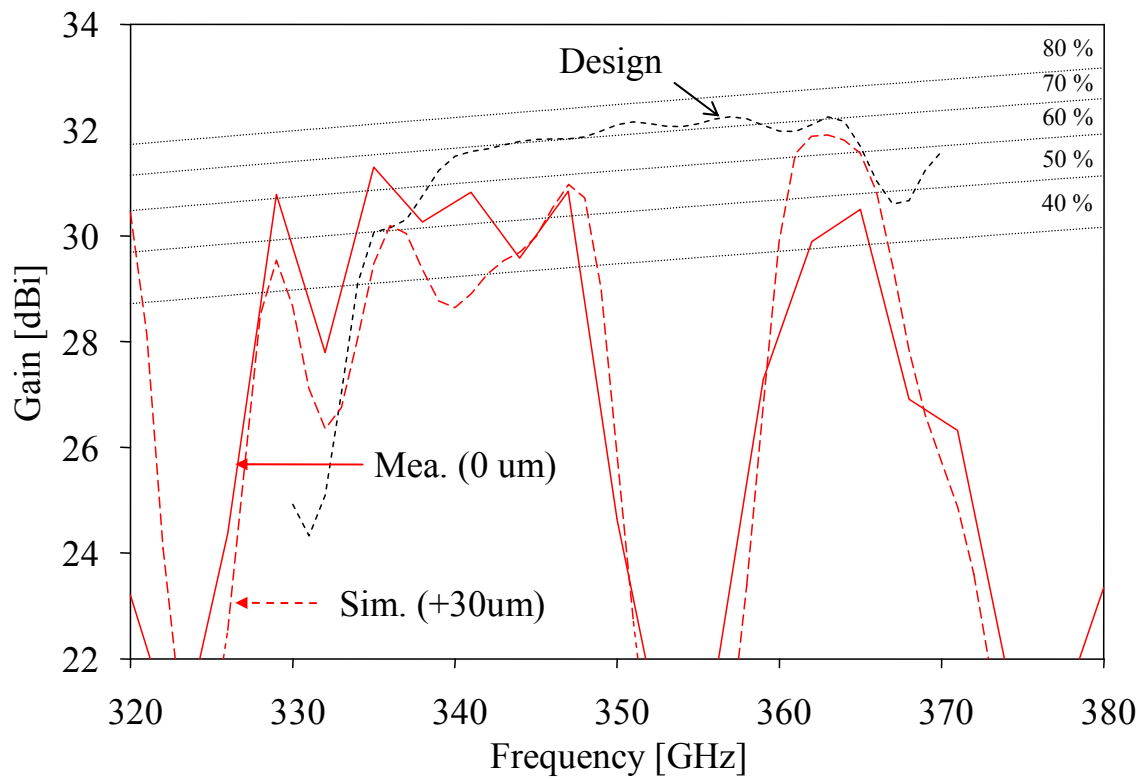


Fig. 16 Picture of the fabricated antenna for the 350 GHz band.



(a)



(b)

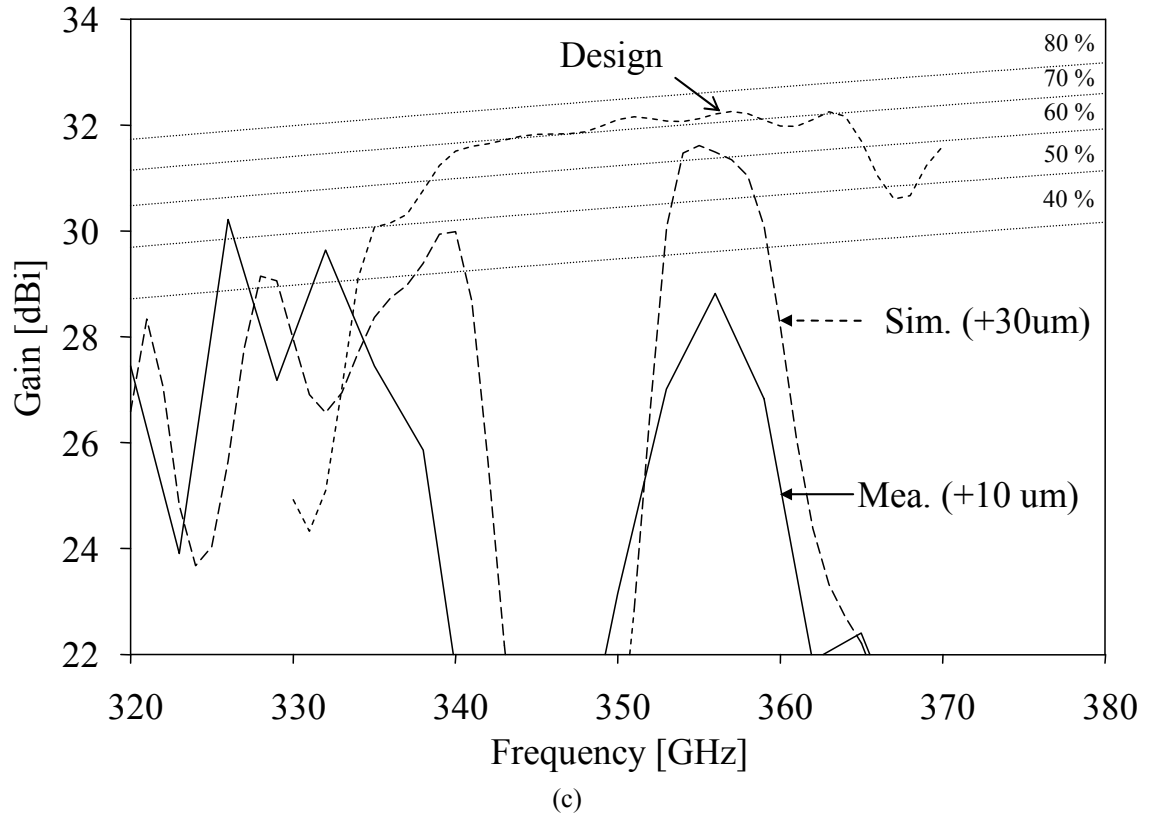


Fig. 17 Frequency behaviors of the gain and the efficiency of 350 GHz antennas

2-7 Data Transmission Experiments in 350 GHz Band

2-7-1 Data Transmission Experimental System for 350 GHz

An experimental system based on optical technology is used for the 350 GHz band. Fig. 18(a) shows the block diagram of the wireless link using a photonics-based THz-wave transmitter [14]. First, an optical THz-wave signal is generated by heterodyning two wavelengths of light from wavelength-tunable light sources. The optical signal is digitally modulated by an optical intensity modulator driven by the pulse pattern generator with ASK modulation. Finally, the optical signal is converted to an electrical signal by the uni-traveling carrier photodiode (UTC-PD) module, which has a WR-3 rectangular waveguide for output and can maximally produce approximately 500 μ W at 300 GHz with 20 mA of photocurrent, and a THz-wave signal is emitted to free space via an antenna. The main features of this photonic approach are that the carrier frequency is widely tunable and that the modulation frequency can be

increased to at least 40 Gbit/s. Fig. 19(b) shows the block diagram of the receiver. The receiver consists of the Schottky barrier diode (SBD) followed by a low-noise preamplifier and a limiting amplifier. The SBD module is a commercially available one (WR2.8ZBD, Virginia Diode Inc.) [15]. The specified range of the operation frequency is 265-400 GHz. For the experiment in the 350 GHz band, pseudorandom data with a sequence length of $2^{21} - 1$ are carried by the intensity of the 350 GHz band signal and are transmitted over a 0.5 m distance to the receiver.

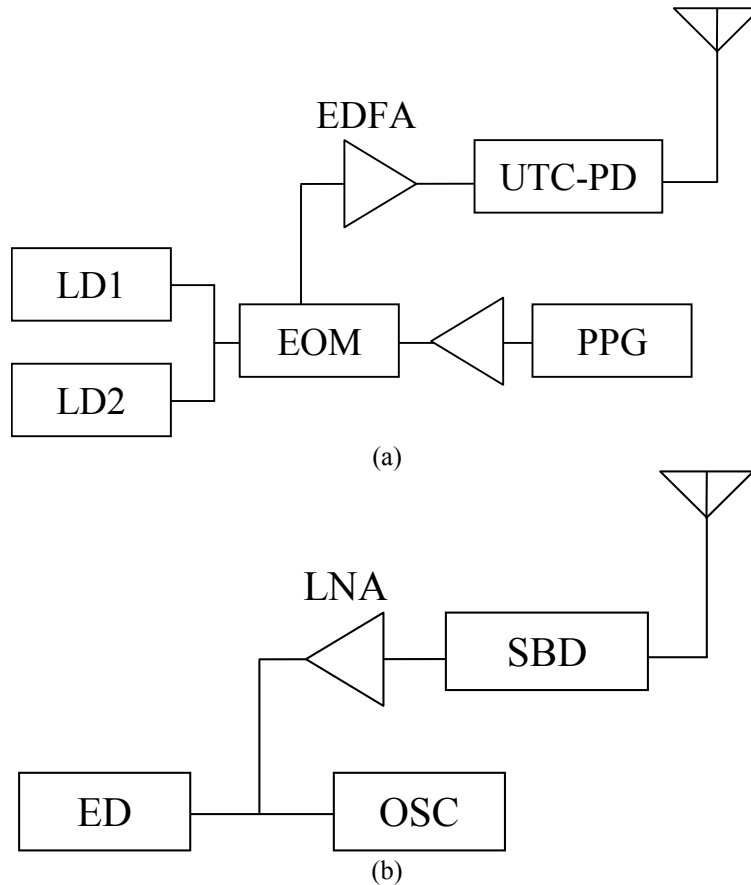


Fig. 18. Schematics of the 350 GHz experimental system. (a) transmitter and (b) receiver. (LD: laser diode, EOM: electrooptic modulator, EDFA: erbium-doped fiber amplifier, UDP-PD: uni-travelling carrier photodiode, SBD: Schottky barrier diode, LNA: low-noise amplifier, ED: error detector, OSC: oscilloscope)

2-7-2 Data Transmission Experimental Results for 350 GHz

In this band, we use the only one experimental scenario. The proposed antenna is set on Rx and the STD is set on the Tx. The antenna described in Fig. 17 (a) is chosen as the Rx considered with the antenna gain characteristics; the antenna in Fig. 17 (a) shows relatively good antenna gain characteristic among three antennas. Additionally, the frequency band of the proposed antenna is shifted lower than we expect, so the experiment is performed around the center frequency of 338 GHz. Fig. 19 shows the measured BER for a 10 Gbit/s data rate with respect to the photocurrent of the UTC-PD, which is directly related to the output power from the transmitter. As can be observed, up to 10 Gbit/s of data are successfully transmitted with a BER less than 10^{-11} . For this transmission, we only require an average power of approximately -14 dBm (corresponding to 40 μ W) at the UTC-PD output or 5.5 mA photocurrent. Fig. 20 shows the eye diagram for a 10 Gbit/s data rate, which ensures a BER less than 10^{-11} . In this experiment, we cannot drive the link at a higher rate and extend the link distance because of the limitations of the instruments, the BER tester, and the experimental space issue. However because the UTC-PD can produce up to approximately -3 dBm (corresponding to 500 μ W) with 20 mA photocurrent at 300 GHz and the optical modulator can handle up to 40-Gbit/s data, it should be possible to transmit data at a higher data rate over a longer link distance. In the following section, we will estimate the maximum available data rate with the antenna for this wireless link system.

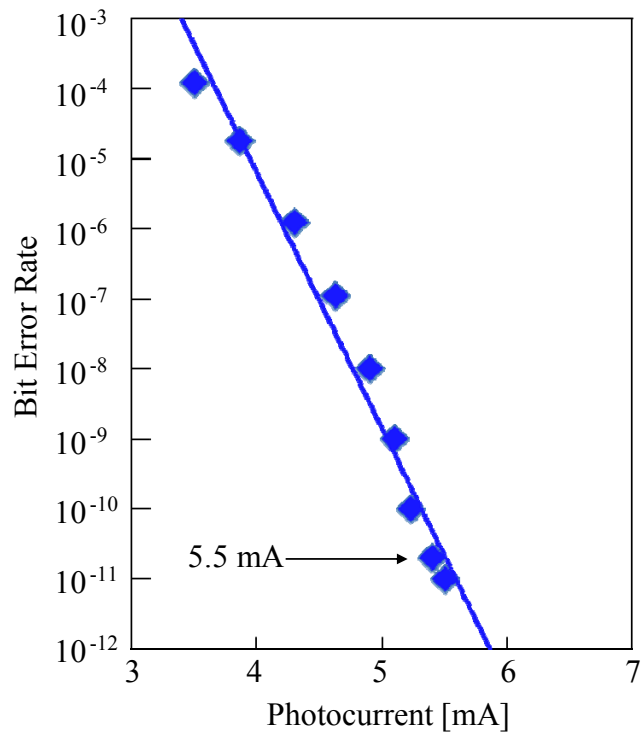


Fig.19 Measured BERs in the 350 GHz band.

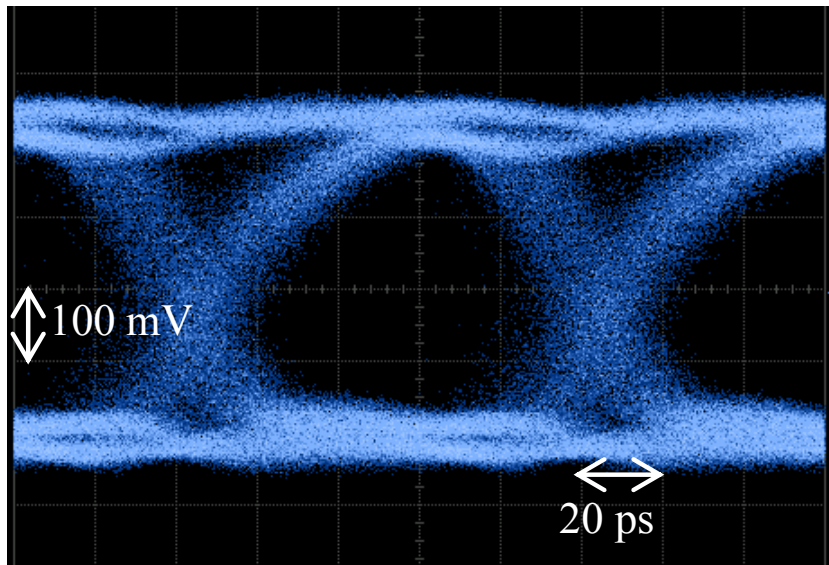


Fig. 20 Eye diagram in the 338 GHz frequency band when the output power of Tx is -14 dBm (corresponding to 40 μ W)

2-7-3 Expectable Maximum Data Rate in 350 GHz

Currently, the upper limits in the bit rate of the PPG and the BER tester of the measurement system for the 350 GHz band are 14 Gbit/s and 12.5 Gbit/s, respectively [16]. Due to those limitations, we cannot measure the maximum possible data rate of the system and the antenna. Thus, we estimate the possible maximum data transmission rate using the relationship between the antenna gain and the BER. Fig. 21 shows the relationship between the antenna gain and the BER as a function of the center frequency. From this relationship, we can estimate the available maximum data transmission rate with this antenna. When we assume that a 7 GHz bandwidth is required for 10 Gbit/s transmission in the base band, the required RF bandwidth becomes 14 GHz for double-sided modulation. The bandwidth for a BER of less than 10^{-10} is 10 GHz, as noted in Fig. 21. In contrast, the bandwidth of gain higher than 20 dBi is approximately 18 GHz, so the available maximum RF bandwidth becomes 24 GHz. Because we achieved 10 Gbit/s with the RF bandwidth of 14 GHz, we can expect that the available maximum data transmission rate is approximately 17 Gbit/s with the RF bandwidth of 24 GHz using the proposed antenna. Note that in this estimation, we assume the quasi-error-free condition ($\text{BER} < 10^{-10}$) to roughly estimate the system performance.

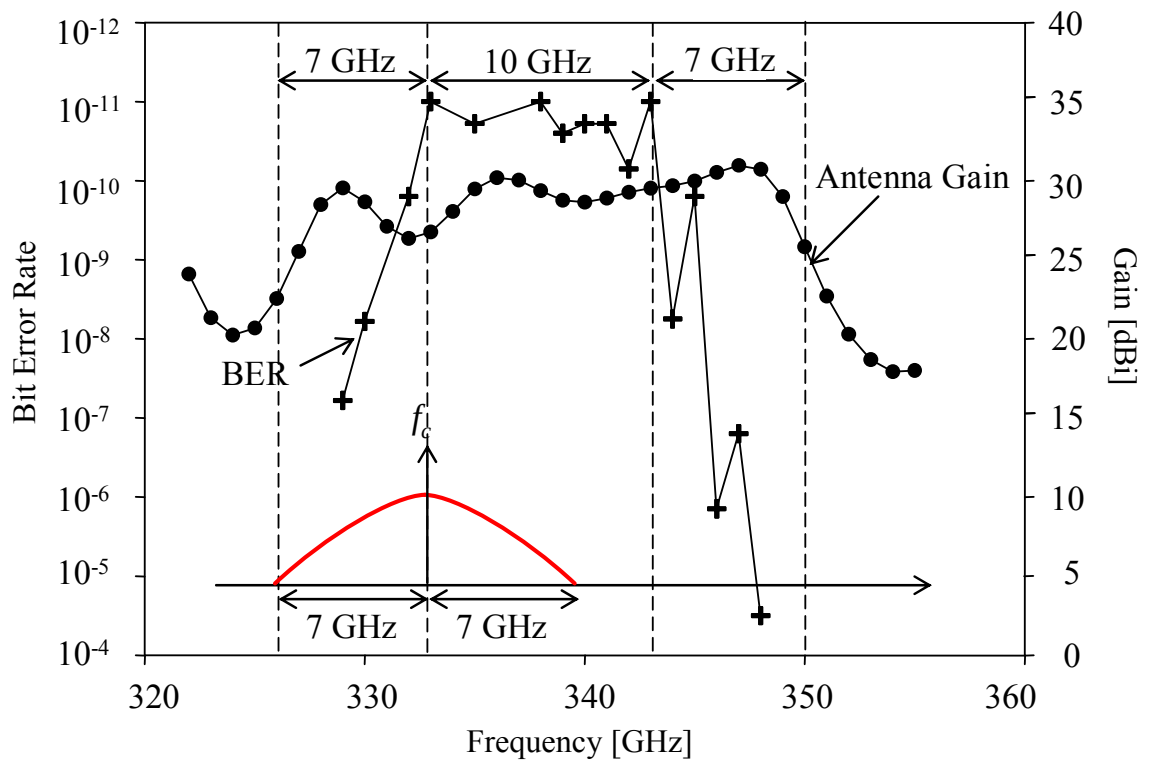


Fig. 21 BER and Antenna Gain.

2-8 Concluding Remarks

A double-layer waveguide slot array antennas with a high gain and a broad bandwidth are designed and fabricated for the 120 GHz band and 350 GHz, respectively. Although there are some disagreements between the simulated and measured results due to fabrication errors, we have confirmed that diffusion bonding technique is suitable for high frequency antennas.

For the 120 GHz band, the high antenna efficiency of 70% has been obtained over a wide bandwidth. A data transmission experiment confirmed that the proposed antenna has feasibility for the wireless link system. We have demonstrated over 10 Gbit/s data transmission with BER less than 10^{-12} using the proposed antenna at 2.5m distance. Of particular importance, the proposed antenna achieved the BERs under 10^{-12} with 8 dB higher attenuation level than the standard horn antenna; this confirms that the proposed antenna has higher antenna gain than the standard horn antenna.

For the 350 GHz band, the antenna efficiency of 40% has been obtained over a wide bandwidth. Even though there are large differences between calculated and measured antenna characteristics due to the fabrication error, the design results show much possibility of this antenna for the sub-millimeter wave and THz regions. Also, we achieve a 10 Gbit/s data transmission rate with a center frequency of 338 GHz for a 0.5 m distance with a BER $<10^{-11}$.

We can expect better transmission performance with improvement of the modulation scheme to reduce the RF bandwidth, or otherwise improvement of the bandwidth of the proposed antenna itself since the data rate and the RF bandwidth have a proportionate relationship.

To our knowledge the proposed antennas are the first planar waveguide slot array antenna prototypes having over 30 dBi antenna gains and over 70% and 40% antenna efficiencies with about 10% bandwidth over 100 GHz frequency bands in the world.

References in Chapter 2

- [1] P. H. Siegel, "Terahertz technology," *IEEE Trans. Microwave Theory Tech.*, vol. 50, no.3, pp.910-928, Mar. 2001.
- [2] J.Laskar, S. Pinel, D. Dawn, S. Sarkar, B. Perumana, and P. Sen, "The next wireless wave is a millimeter wave," *Microw. J.*, vol. 50, pp.22-36, Aug. 2007.
- [3] J. Wells, "Faster than fiber: The future of multi-G/s wireless," *IEEE Microw. Mag.*, vol. 10, no. 3, pp. 104–112, 2009.
- [4] R. Piesiewicz, T. Kleine-Ostmann, N. Krumbholz, D. Mittleman, M.Koch, J. Schoebel, and T. Kurner, "Short-range ultra-broadband terahertz communications: Concepts and perspectives," *IEEE Antennas Propag. Mag.*, vol. 49, no. 6, pp. 24–39, 2007.
- [5] J. Laskar, S. Pinel, D. Dawn, S. Sarkar, B. Perumana, and P. Sen, "The next wireless wave is a millimeter wave," *Microw. J.*, vol. 50, pp. 22–36, Aug. 2007.
- [6] A. Hirata, T. Kosugi, H. Takahashi, R. Yamaguchi, F. Nakajima, T. Furuta, H. Ito, H. Sugahara, Y. Sato, and T. Nagatsuma, "120-GHz-Band Millimeter-Wave Photonic Wireless Link for 10-Gbit/s Data Transmission," *IEEE Trans. Microwave Theory and Tech.*, vol. 54, no.5, pp. 1937-1944, May, 2006.
- [7] T. Nagatsuma, H.-J Song, Y. Fujimoto, K. Miyake, A. Hirata, K. Ajito, A. Wakatsuki, T.Furuta, N. Kukutsu, and Y. Kado, "Giga-bit wireless link using 300-400 GHz bands", *Tech. Dig. IEEE Intl. Topical Meeting on Microwave Photonics*, Th.2.3, Oct. 2009
- [8] Attenuation by Atmospheric Gases, ITU-R Recommendation P. 676-8, 2009
- [9] P. Scott, "The am Atmospheric model, submillimeter array," *Tech. Memo #152* [Online]. Available: <https://www.cfa.harvard.edu/~spaine/am/>
- [10] Y. Miura, J. Hirokawa, M. Ando, Y. Shibuya, and G. Yoshida, "Double-layer full-corporate-feed hollow-waveguide slot array antenna in the 60-GHz band," *IEEE Trans. Antennas and Propag.*, vol. 59, no.8, pp.2844-2851, Aug. 2011.
- [11] M. G. Nicholas and R. M. Crispin, "Diffusion bonding stainless steel to alumina using aluminum interlayers," *Journal of Materials Science*, vol. 17, No. 11, pp.3347-3360, 1982.

- [12] J. Hirokawa, M. Zhang, and M. Ando, “94 GHz single-layer slotted waveguide array by diffusion bonding of laminated thin plates,” in Proc. Int. Symp. Antenna Propag. TP-C0.-3, Oct. 2008.
- [13] Ansys, Inc. Available: <http://ansys.com/>
- [14] T. Nagatsuma, “Extreme Bandwidth Wireless Communications using Terahertz Waves”, in Proc. Int. Conf. Applied Electro. TC21-2, Sept. 2010.
- [15] Virginia Diode Inc. Available: <http://www.virginiadiodes.com/WR2.8DD.php>.
- [16] T. Nagatsuma, “Extreme bandwidth wireless communications using terahertz waves”, IEEJ Electronics, Information and Systems Technical Meeting, TC21-2, Sept. 2010

Chapter 3

Antenna Design for the Higher Gain Applications

3-1 Introductory Remarks

High gain and high efficiency antennas are essential for the high speed wireless link systems for outdoor point-to-point relay. A reflector antenna and a dielectric lens antenna satisfy the high gain and the high efficiency in the same time; however it is difficult to realize thin planar structure because they essentially require focal length [1]. On the other hand, the microstrip line-fed array antennas using dielectric substrates provide a low profile. But they suffer from the large loss; they are not suitable for the high gain because the loss increases with the array size [2].

As mentioned in chapter 1, a hollow-waveguide slot array antenna is advantageous for a larger size or a higher gain with a high efficiency since it neither suffers from dielectric nor radiation loss in comparison with a microstrip line-fed array antenna. Furthermore, it satisfies a low profile antenna structure as well, it is an attractive candidate for the applications requires a high gain. By the way, the conventional waveguide slot array antennas with series-feed have a fundamental problem that the bandwidth is narrowed due to the long-line effect when the array size

becomes large [3]. To overcome the narrow bandwidth in the conventional waveguide slot array antennas, the waveguide slot array antennas having a multiple layer structure for feeding network have been proposed [4]-[6].

In chapter 2, we already have proposed double layer slot array antennas for the wide bandwidth operation in 120 GHz band with the diffusion bonding technique. The 16x16-element array antenna shows the 70% antenna efficiency with more than 32 dBi over a 13 GHz bandwidth in the 120 GHz band [7].

In this chapter, we design these antennas for the applications require higher gain with a fabrication method of the diffusion bonding of laminated thin metal plates. Here, we propose a 32x32-element array antenna and a 64x64-element array antenna in the 120 GHz band, respectively. Also, we propose a new feeding structure having a double-layer for the stable fabrication in the diffusion bonding technique. These antennas could be an alternative to reflector and lens antennas which have a high antenna gain for the wireless link systems.

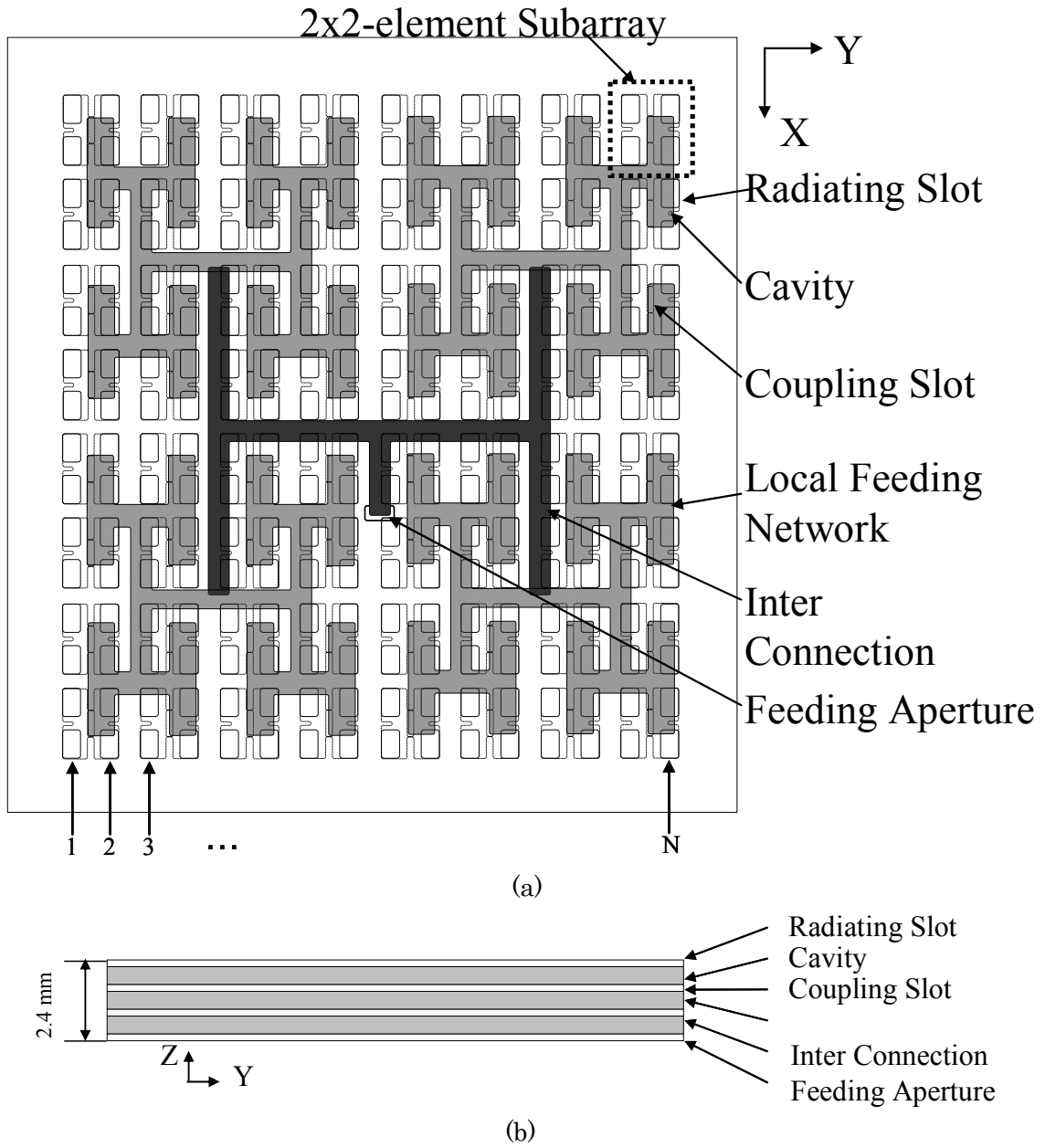


Fig. 1 Antenna configuration of the new feeding structure (a) top-view and (b) side-view.

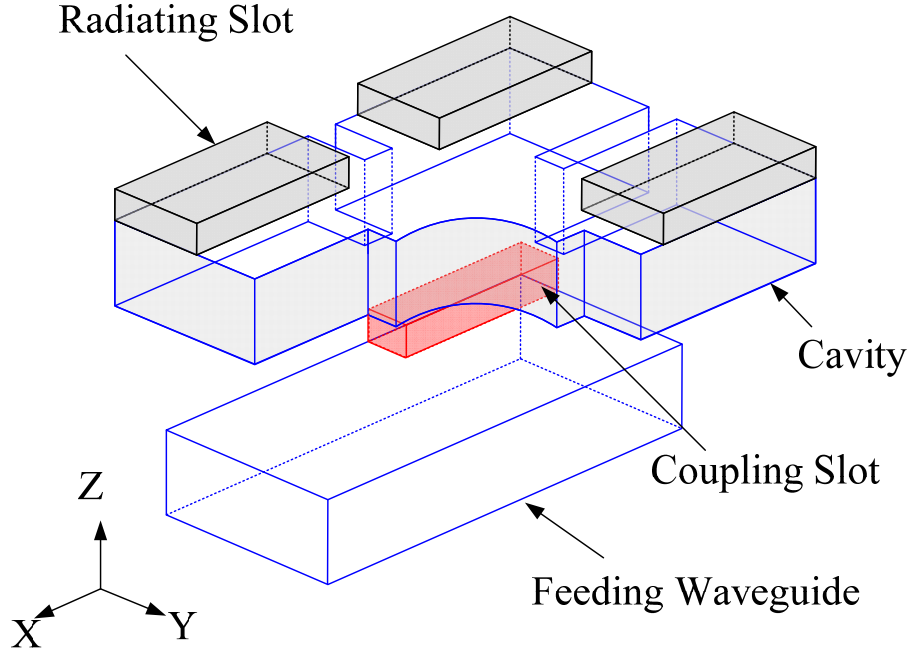


Fig. 2 Exploded perspective view of the 2x2-element subarray.

3-2 Antenna Configurations

Fig. 1 shows the configuration of the slot array antennas for the large size arrays in the 120 GHz band. This structure is basically identical with the antenna structure introduced in chapter 2 and [7] except the feeding network. In this antenna structure the double-layer feeding structure is employed for the stability in fabrication.

There are 2x2-element subarrays on the top layer. All the radiating elements are spaced with a 0.86-wavelength interval and fed with an equal phase and amplitude by the corporate-feed network located in the lower layer. Since this antenna has a repeatable structure composed of 2x2-element subarrays, the large arrays could be easily realized with more subarrays and the expansion of the feeding network. The aperture size is defined with the array sizes and the antenna height is 2.4 mm including all layers. The heights of the layers of the antenna are 0.2 mm for the radiating slots, 0.6 mm for the cavities, 0.2 mm for the coupling slots and 1.4 mm for the double-layer feeding structure. Fig. 2 shows the exploded perspective view of the 2x2-element subarray as the design unit to be fed by the corporate-feed structure. The design parameters of this 2x2-element subarray structure are identical with the structure described in [7].

3-3 Fabrication of Diffusion Bonding of Laminated Thin Metal Plates for Large Size Arrays

The fabrication method of diffusion bonding of laminated thin metal plates is already introduced in section 2-3-1 [8]-[9]. Also its limitations for high frequency antennas are explained in section 2-3-2. For the information of the fabrication process and limitations, please see the section 2-3-1 and 2-3-2.

We have another issue in fabrication using diffusion bonding of laminated thin metal plates for the large size array. In this section, we discuss the problems in fabrication for large size array such as 32x32- and 64x64-element array in 120 GHz band. In section 2-3-2, the limitations are mainly occurred in the etching process in Step 1. In this time, another issue is mainly occurred in the pressing process in Step 3.

As explained in section 2-3-1, the diffusion bonding technique provides a high accuracy ($\pm 20\mu\text{m}$) and a good electric contact among etched metal plates. However, great care must be taken for the fabrication of the large and complex structure using this fabrication method because the metal plates having a large and complex pattern are easy to be deformed due to the high mechanical pressure in fabrication process 3. The etching patterns of the laminated copper plates of the 32x32-element array are shown in Fig. 3 as an example. These etching patterns are designed with the simple expansion of the conventional structure adapted to the 16x16-element array. We need five etching patterns for this antenna structure. All patterns are etched in thin copper plates having 0.2 mm thickness. (Note that, we need same three etching patterns for the cavities and the feeding network because their thickness is 0.6 mm.) The size of each pattern is 75.0 mm square in the xy -plane while the aperture size is 67.2 mm square. Actually, the etching patterns of radiating parts in Fig. 3 (a), (b) and (c) are large, but simple, thus they could be fabricated without any troubles. However the pattern for the feeding network in Fig. 3(d) is large and very complex. Fig. 4 shows the enlarged etching pattern of the feeding network for the 32x32-element array. The white part indicates the cooper which is corresponding to the antenna body and the black part indicates the empty space (air) which is corresponding to the waveguides for the feeding network. As shown here, much portion of the metal plate remains empty space for the waveguide as

well as the etching pattern is complex and long, thus this pattern is weak against the high mechanical pressure. Furthermore, the walls between waveguides are thin and long as marked 'A' in Fig. 4. The wall thickness is only 0.60 mm and it is not enough to withstand the high pressure in fabrication.

Thus the wall may become thicker under the high pressure and the width of the waveguides which are faced the wall may become changed. It seriously affects the antenna performance since the width of the waveguide affects the propagation characteristics of the signal. Small fluctuations in the width of the waveguide can accumulate into large phase differences between the elements due to the long transmission distance from the input port to a single element. Also, each H-junction in the feeding network is surrounded by thin walls as marked 'B' in Fig. 4. This part also may have a similar problem with the part 'A'. With these reasons, it is difficult to guarantee the uniform excitation for all radiating elements in the large size array antenna with this conventional feeding structure. It may cause the high grating lobes and low gain in the array antenna. In the following chapter, we discuss the fabricated results and issues of the 32x32-element array antennas with the conventional structure.

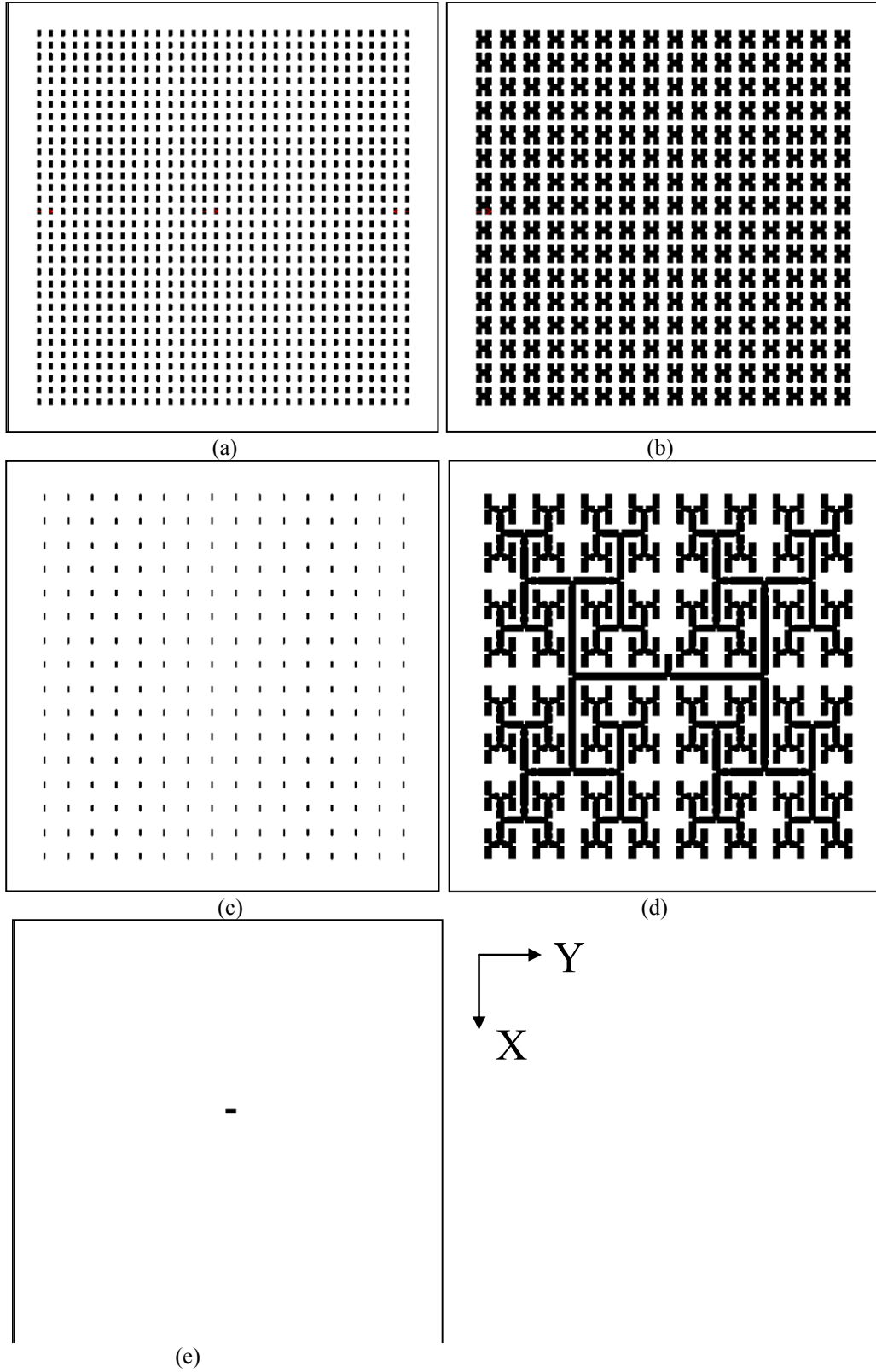


Fig. 3 Etching patterns of copper plates for the fabrication of the 32x32-element array with the conventional feeding structure (a) radiating slots (b) cavities (c) coupling slots (d) feeding network and (e) feeding aperture.

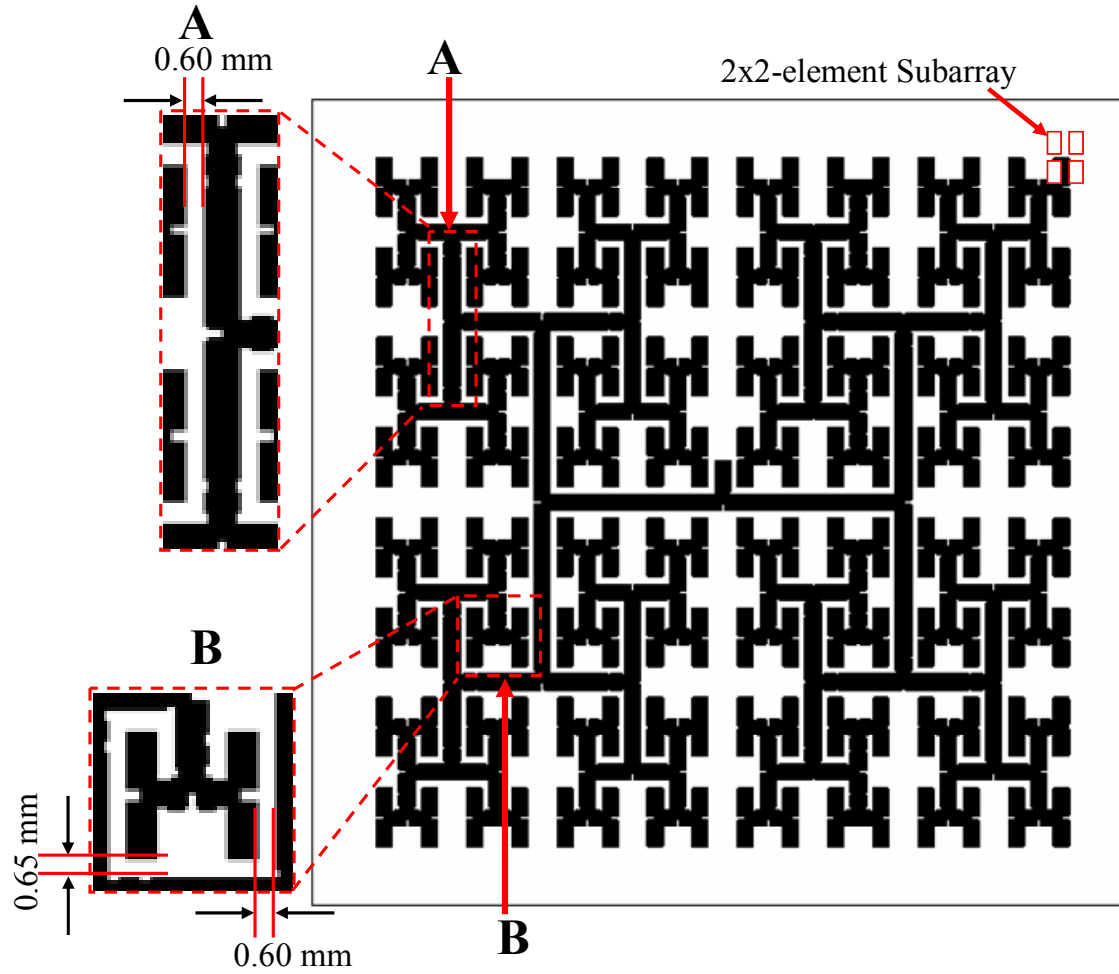


Fig. 4 Etching pattern of the conventional feeding network for the 32x32-element array.

3-4 Fabrication Result of 32x32-element Array

Antennas with the Conventional Feeding Structure

We fabricate three antennas having 32x32-element with the diffusion bonding of laminated thin metal plates. They have the same antenna dimension. The antenna structure is identical with the 16x16-element array antenna described in [10]. Only the array size is increased with more radiating elements and the expansion of the feeding network. The aperture areas are defined as 67.2 mm square (2.1 mm x 32 elements). The height of the antenna is 1.6 mm identical with the 16x16-element array. The calculated and measured reflections are shown in Fig. 5. We measure the reflections using a Vector Network Analyzer (VNA). In the simulation, the bandwidth for VSWR less than 1.5 is

8.6%. Even though there are mismatches among the calculated and the measured results of the three antennas, the measured reflections are suppressed under -10 dB in the design frequency band. Thus we can confirm that the reflection may not much affect the performances such as gain and efficiency. Fig. 6 shows the simulated and measured gain and antenna efficiency. The measured three results are almost identical and show about 50-60% antenna efficiency while the simulated result shows over 80% in the broad bandwidth. Thus the degradation of the antenna efficiency in measurement is quite larger than that of the previous results in the 16x16-element array antenna. The antenna efficiency of the 16x16-element array antenna is higher than 70% in a broad bandwidth [7]. There are various reasons in this degradation of the gain such as frequency shift, and not perfectly suppressed reflection, but we can infer the main reason of the gain degradation from the measured radiation pattern. Fig. 7 shows one of the measured radiation patterns in the yz-plane (*E*-plane) of the three antennas; the three antenna show almost identical radiation pattern. Here, we can observe the mismatches of the first side lobe levels at ± 3 degrees between the calculation and measurement. The first side lobe level in the simulation shows -13.3 dB and it reflects the uniform excitation [10]. However, the first side lobe levels in measurement show quite different. They show -16.0 dB and -10.1 dB at ± 3 degrees, respectively. Also, many grating lobes appear in the measured radiation pattern. So we consider the non-uniform excitation is happened in the feeding network due to its deformation as described in section 3. To prevent the deformation in fabrication, we propose a new feeding network composed of two layers; one is for the local feeding network for 8x8 elements and the other is for the interconnections between the local feeding networks.

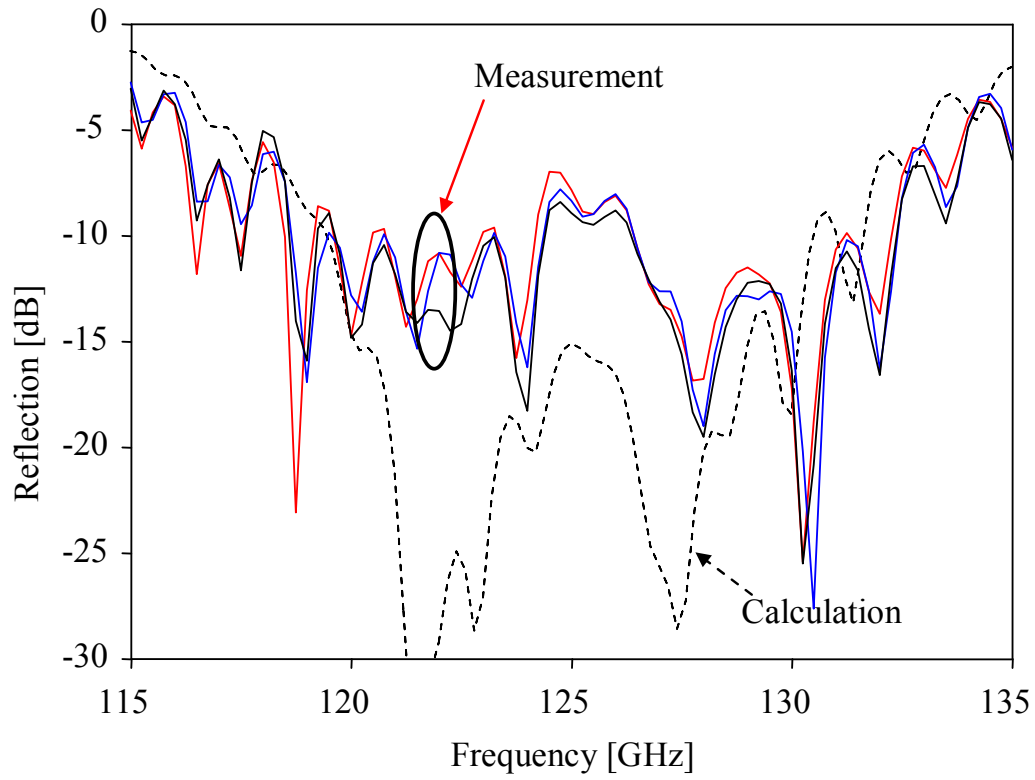


Fig. 5 Calculated and measured reflection of the 32x32-element array antennas with the conventional structure.

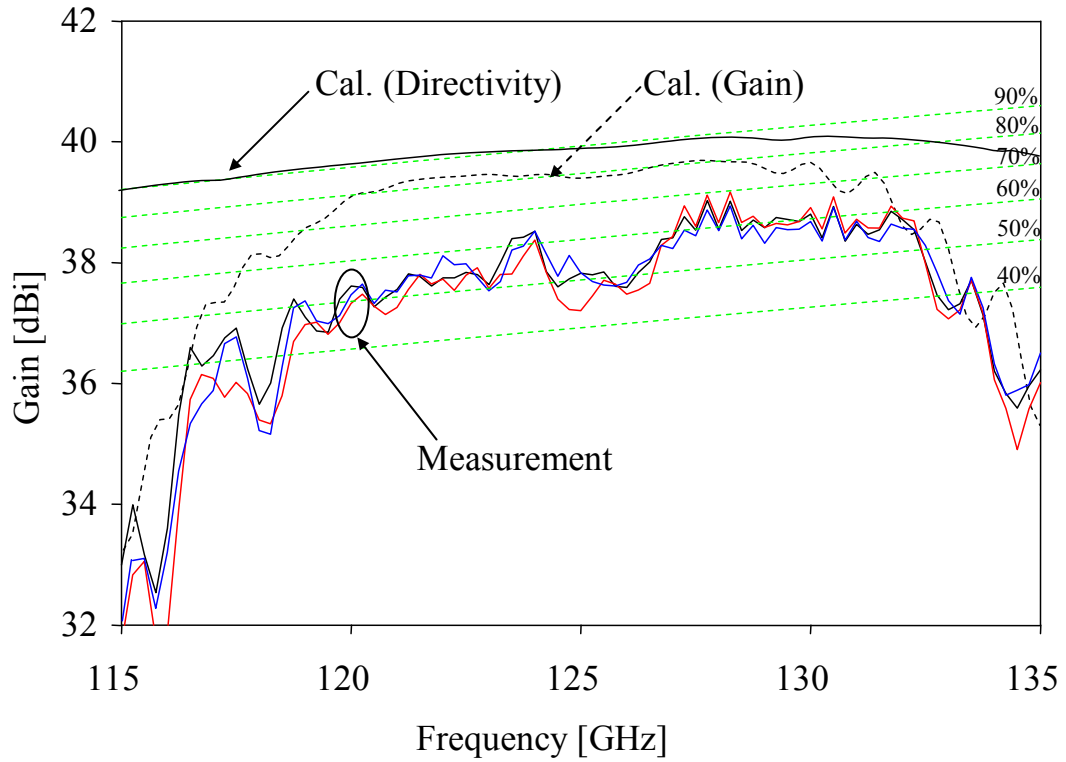


Fig. 6 Calculated and measured gain and efficiency of 32x32-element array antennas with the conventional feeding structures for 120 GHz

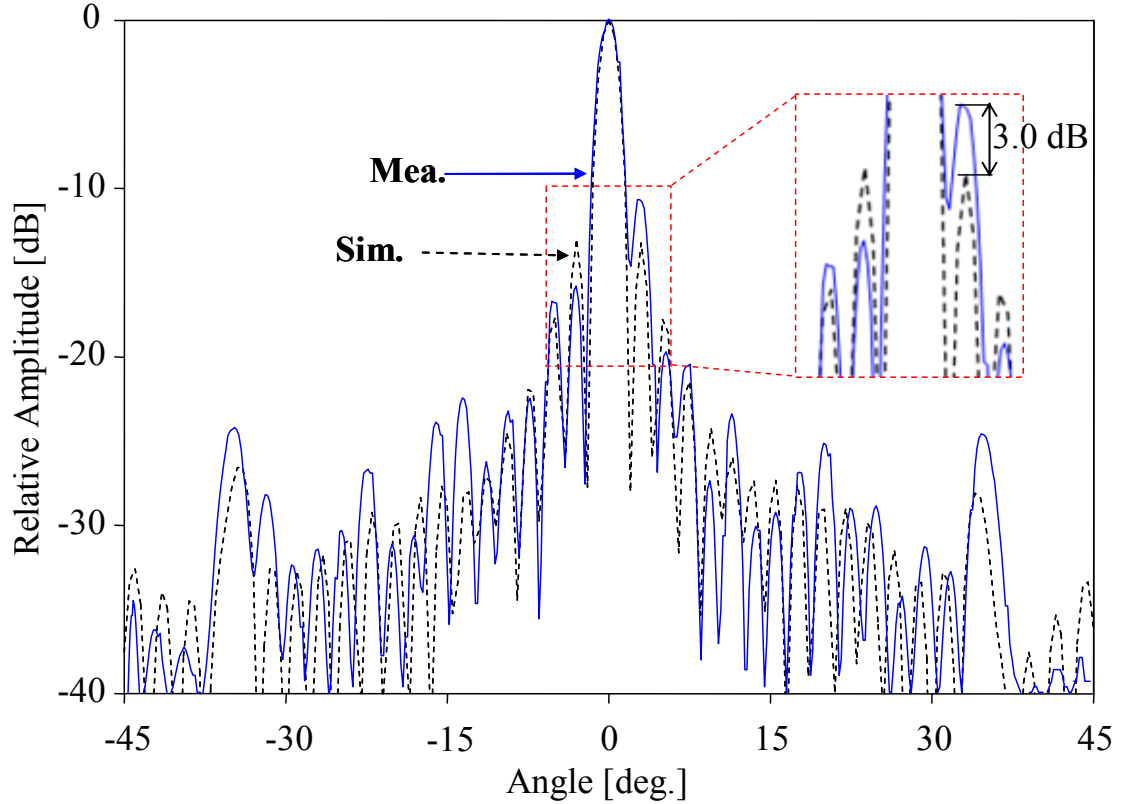
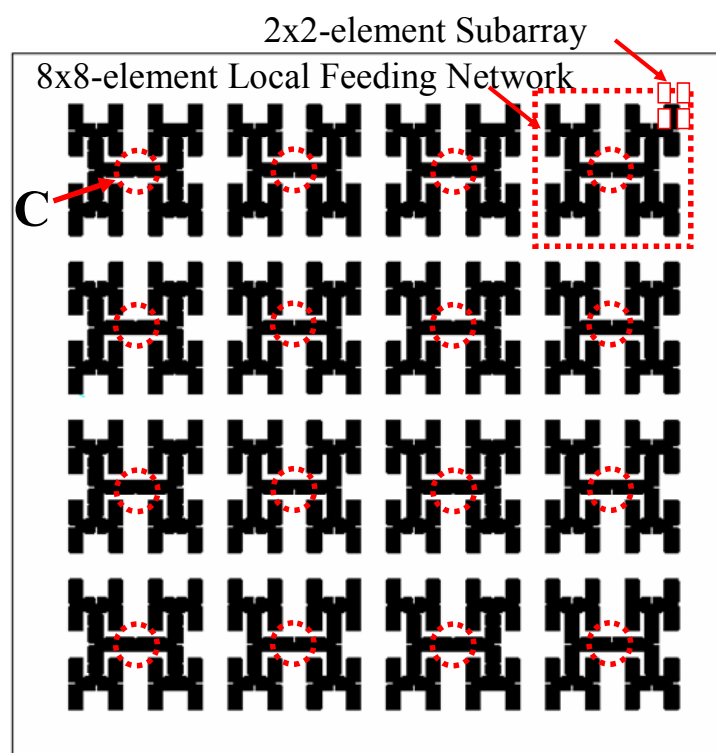


Fig. 7 Calculated and measured E-plane (yz-plane) radiation patterns of 32x32-element array antenna with the conventional feeding structure.

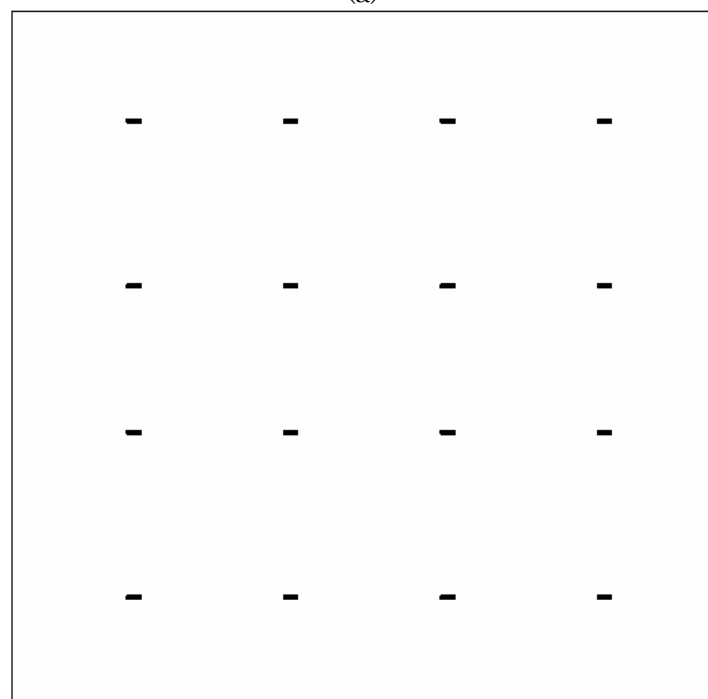
3-5 New Feeding Structure for Large Size Arrays with Diffusion Bonding of Laminated Thin Metal Plates

In this section, we propose a new feeding structure for the stable fabrication of the large size array antennas. Fig. 8 shows a new feeding structure having double-layer for the feeding network. The local feeding networks for 8x8-element arrays are set on the 1st feeding layer and their interconnection is set on the 2nd feeding layer of the new feeding structure. With this separated structure, we can efficiently reduce the complexity and length of the patterns for the feeding network. Especially, the thin and long walls between waveguides marked ‘A’ in Fig. 4 are not required anymore. Also each H-junction in the feeding network is surrounded by thick walls in comparison with that of the conventional one as marked ‘B’ in Fig. 4. Even though the total height of the antenna is increased from 1.6 mm to 2.4 mm due to the additional layers, it still shows

low profile compared with reflector and lens antennas. A connection unit is designed to connect two different feeding layers using a combination of an E-bend and a T-junction. Fig. 9 shows the configuration of the connection unit to connect two different layers. The signal from the input port of the connection unit marked '1' is coupled through the coupling slot and it is divided into two ways with same phase and same magnitude by the splitter. Each end of the 2nd feeding layer (in Fig. 8(c)) is connected to the input port of the connection unit. And the output ports of the connection unit marked '2' and '3' are connected to 'C' of the each local feeding network for 8x8 elements in the 1st feeding layer (in Fig. 8 (a)). Fig.10 shows the calculated frequency characteristics of the reflection of the new and the conventional feeding structures for the 32x32-element array. Also the calculated frequency characteristic of the connection unit in Fig.9 is added on the graph. As shown in Fig. 10, the frequency characteristics of the reflection are almost identical between the new and the conventional feeding structures. They are well suppressed under -20 dB at the wide bandwidth from 118 GHz to 133 GHz except near 125 GHz; the reflection near 125 GHz shows under -18 dB. The reflection of the connection unit is well suppressed under -20 dB from 115 GHz to 135 GHz. Furthermore, the new structure having double-layer for the 32x32-element array is achieved to excite all 2x2-element subarrays uniformly both in amplitude and phase. The amplitude and phase variation over the bandwidth from 115 GHz to 135 GHz are 0.4 dB and 5.2 degrees among the ends of the feeding network, respectively. With the new feeding structure, we design whole antenna structure including a radiating part and a feeding part for the 32x32-element array and the 64x64-element array, respectively. The new antenna configuration with the double-layer feeding structure is introduced in Fig. 1. The design and fabrication results of the 32x32-element and 64x64-element arrays will be presented in the following chapter.



(a)



(b)

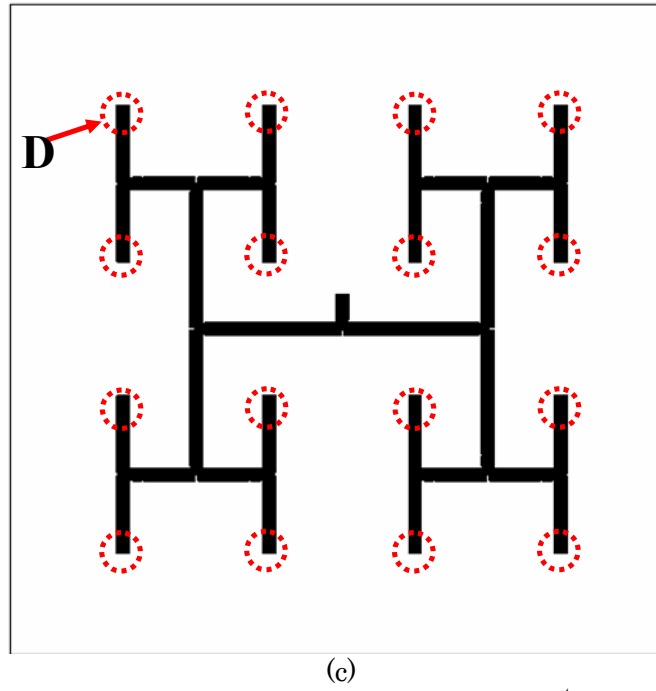


Fig. 8 Double-layer feeding structure for the large size array antenna (a) 1st layer of double layer feeding structure and (b) coupling slot and (c) 2nd layer of double layer feeding structure.

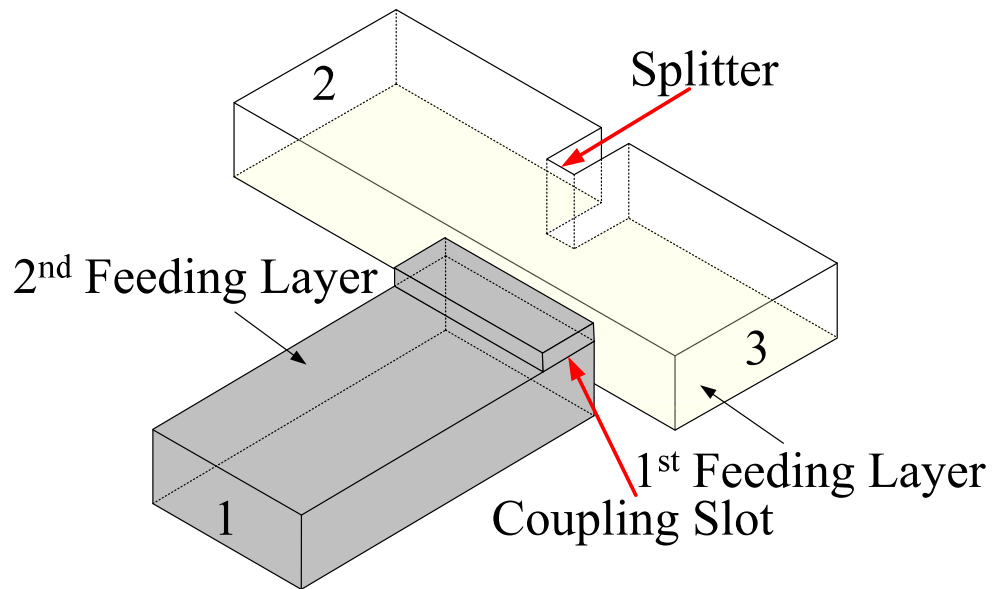


Fig. 9 Connection unit for the new feeding structure.

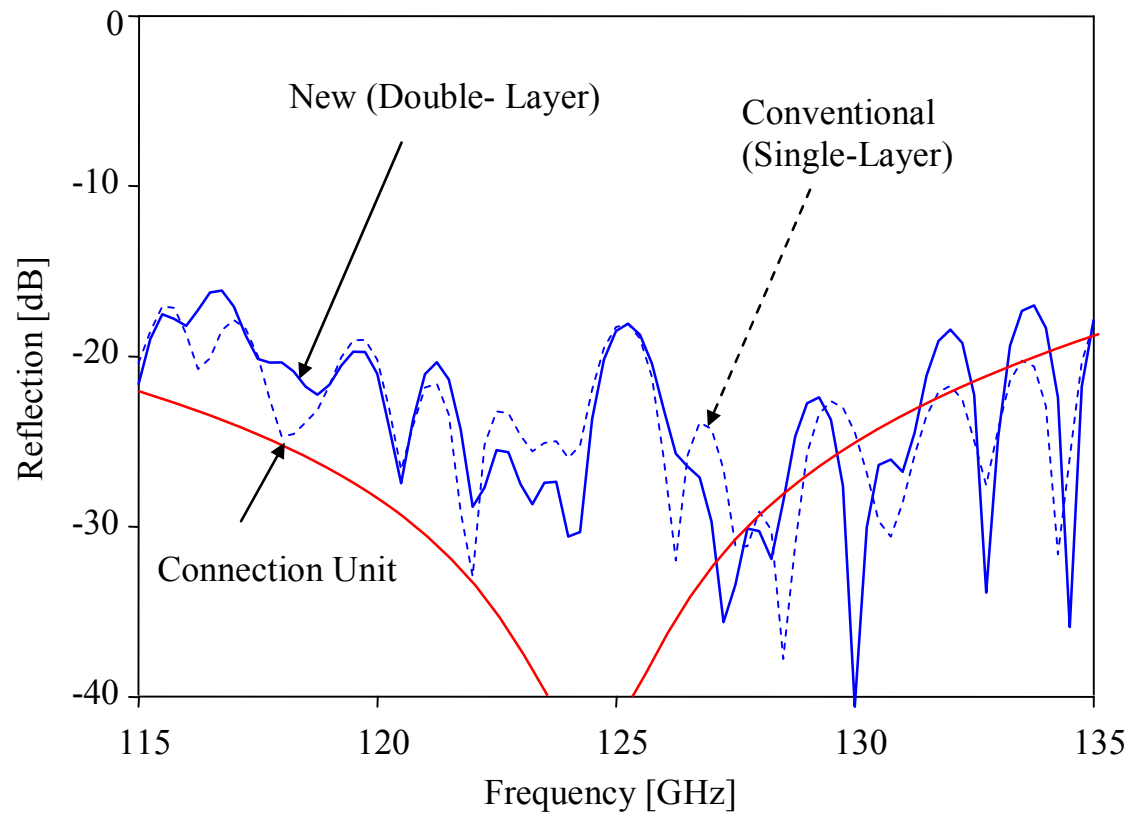


Fig. 10 Calculated reflection coefficients of the full-corporate-feeding networks and the connection unit.

3-6 Fabrication Result of 32x32- and 64x64-element Array Antennas with New Feeding Structure

3-6-1 32x32-element Array Antennas

Fig. 11 shows the 32x32-element array antenna fabricated by diffusion bonding of laminated thin metal plates. Three antennas are fabricated for the 32x32-element array. The new feeding structures are used for each array antenna, but the parameters of the radiating parts are not changed at all from the conventional antenna structure. The aperture area is defined as 67.2 mm square (2.1 mm x 32 elements) for the 32x32-element array. The heights of the antennas are 2.4 mm while the conventional structure was 1.6 mm. Also the number of etching patterns for the antenna fabrication is increased to seven while the conventional antenna requires five patterns.

Fig. 12 shows the frequency characteristics of the reflection at the feed aperture of the 32x32-element arrays. Also the calculated reflection of the 32x32-element array antenna with the new feeding structure is added in Fig. 12. In the simulation, the bandwidth of the reflection for VSWR less than 1.5 is 8.6%. There are mismatches among the calculated and measured results of the both antennas, but the measured reflections are suppressed under -10 dB in the design frequency band.

Fig. 13 shows the frequency characteristics of the gains and the antenna efficiencies of the 32x32-element arrays. The peak of the gain is 39.1 dBi with 84% antenna efficiency at 126.5 GHz in the simulation by HFSS [11]. In the measurement, we have over 60% antenna efficiency over a broad bandwidth from 119.0 GHz to 134.0 GHz. The measured peak gain is 39.1 dBi with 80% antenna efficiency at 128 GHz. Also, the measured gains of the structure with the conventional feeding structure are added in Fig. 13 to confirm the gain difference depends on the feeding structure. Even though the measured antenna gain with the new feeding structure is a little bit lower than the simulation, the overall antenna gain with the new feeding structure is higher than that with the conventional feeding structure in entire frequency band. The average of the gain difference between two types is more than 0.76 dB over the bandwidth from 115 to 135 GHz. This gain difference of 0.76 dB is corresponding to over 15% antenna efficiency difference at 125 GHz. With these results, we confirm that the new feeding

structure is promising to prevent the gain degradation due to the fabrication errors described in chapter 3-3.

The conductor loss and the insertion loss are included in all gain data for the both of the simulation and measurement. In calculation using HFSS, the conductivity losses at 125 GHz is about 0.6 dB for the 32x32-element array when the conductivity of copper is assumed to be 5.8×10^7 S/m. Fig. 14 shows the radiation patterns of the 32x32-element array at 125 GHz. In this time, we could measure within ± 45 degrees due to the physical limitations of the measurement system. However we can clearly observe the difference between the new and conventional structures. In Fig. 14 (a), the measured E-pattern (yz-plane) shows a nice agreement in comparison with the simulated result. This supports our assumption related on the uniform or non-uniform excitation depends on the feeding structure. The measured half power beam widths are 1.6 degrees for the both of the E- and H-patterns. The grating lobes around ± 35 degrees in the E-pattern are caused by the element spacing of 1.72 wavelengths of the 2x2 subarrays.

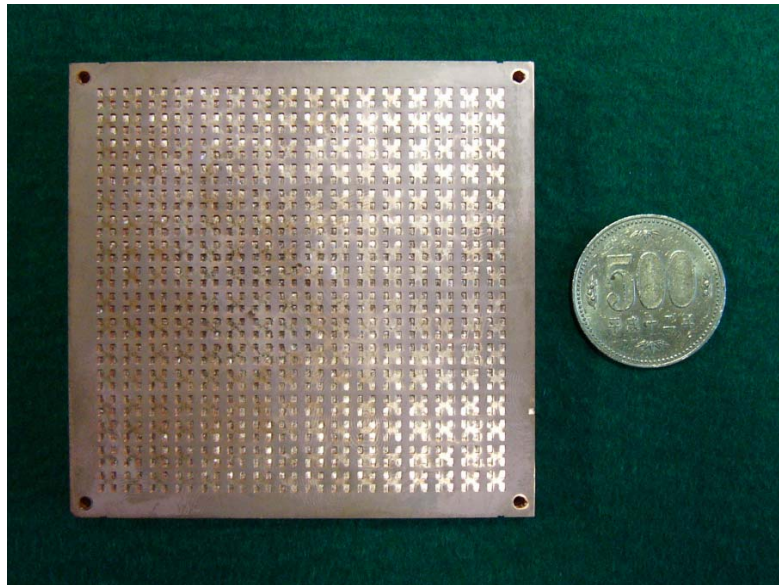


Fig. 11 Picture of the fabricated 32x32-element array antenna with the new feeding structure for the 120 GHz band.

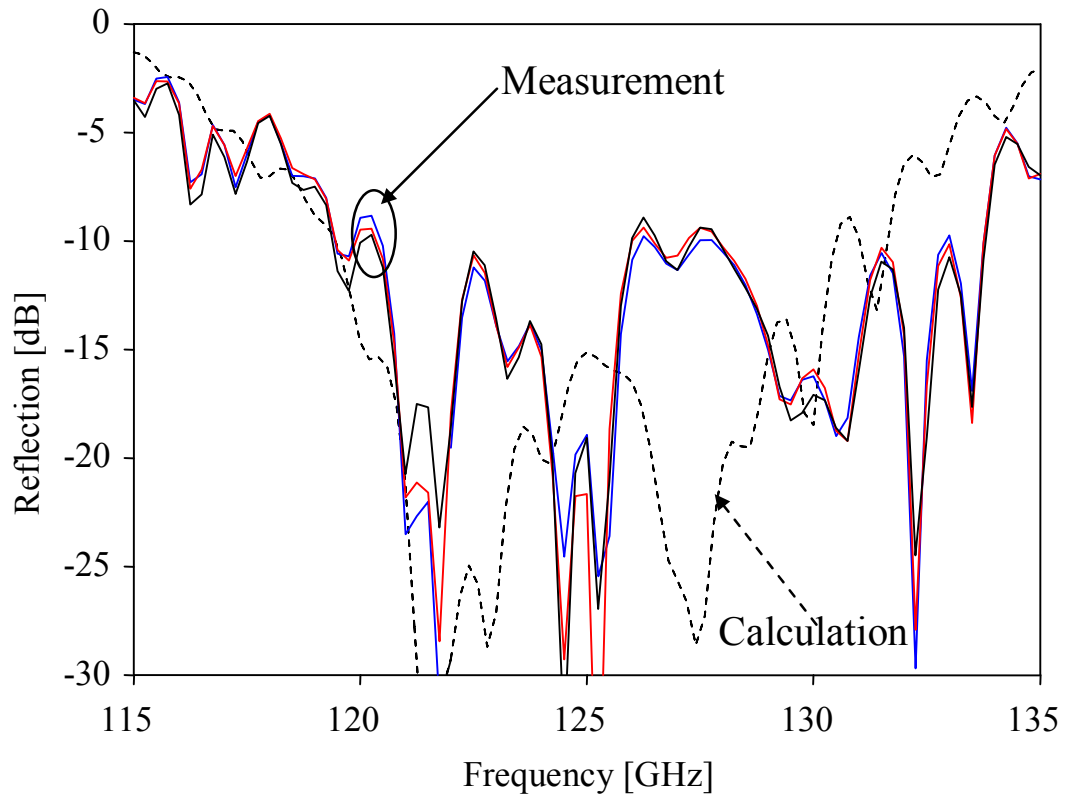


Fig. 12 Calculated and measured reflection coefficients of 32x32-element array antennas with the new feeding structure for 120 GHz

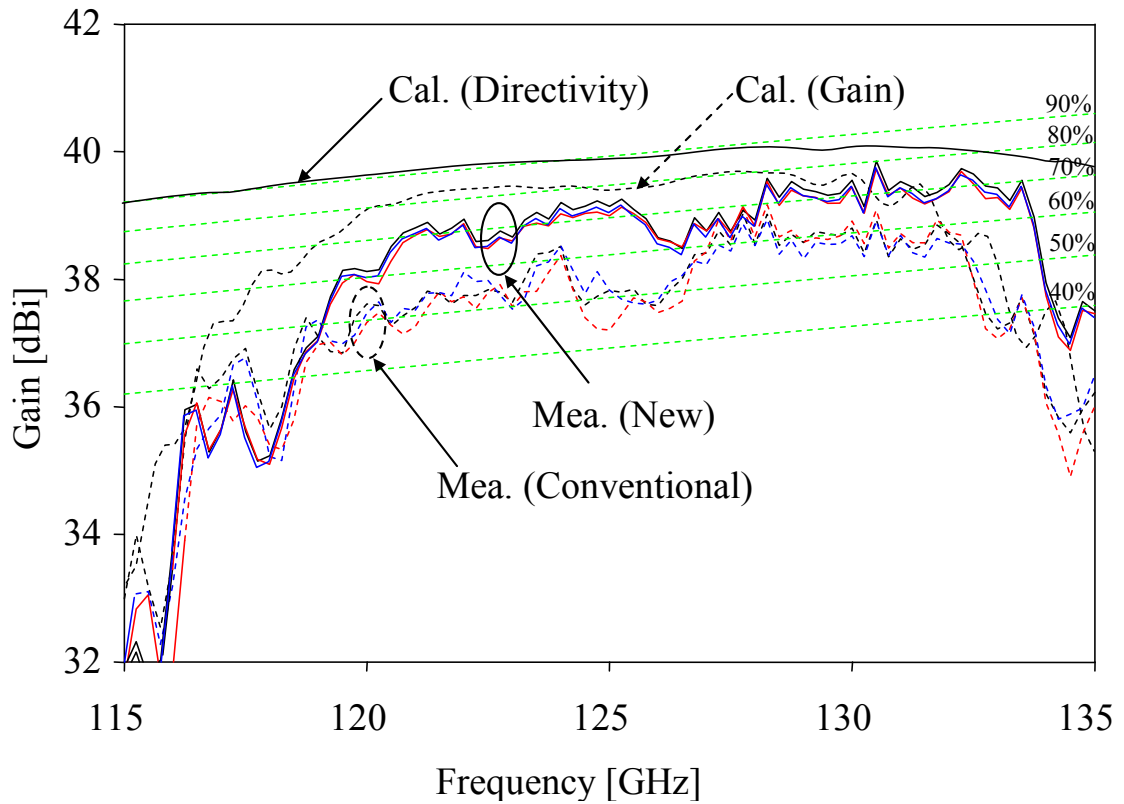


Fig. 13 Calculated and measured gain and efficiency of 32x32-element array antennas with the new and the conventional feeding structures for 120 GHz

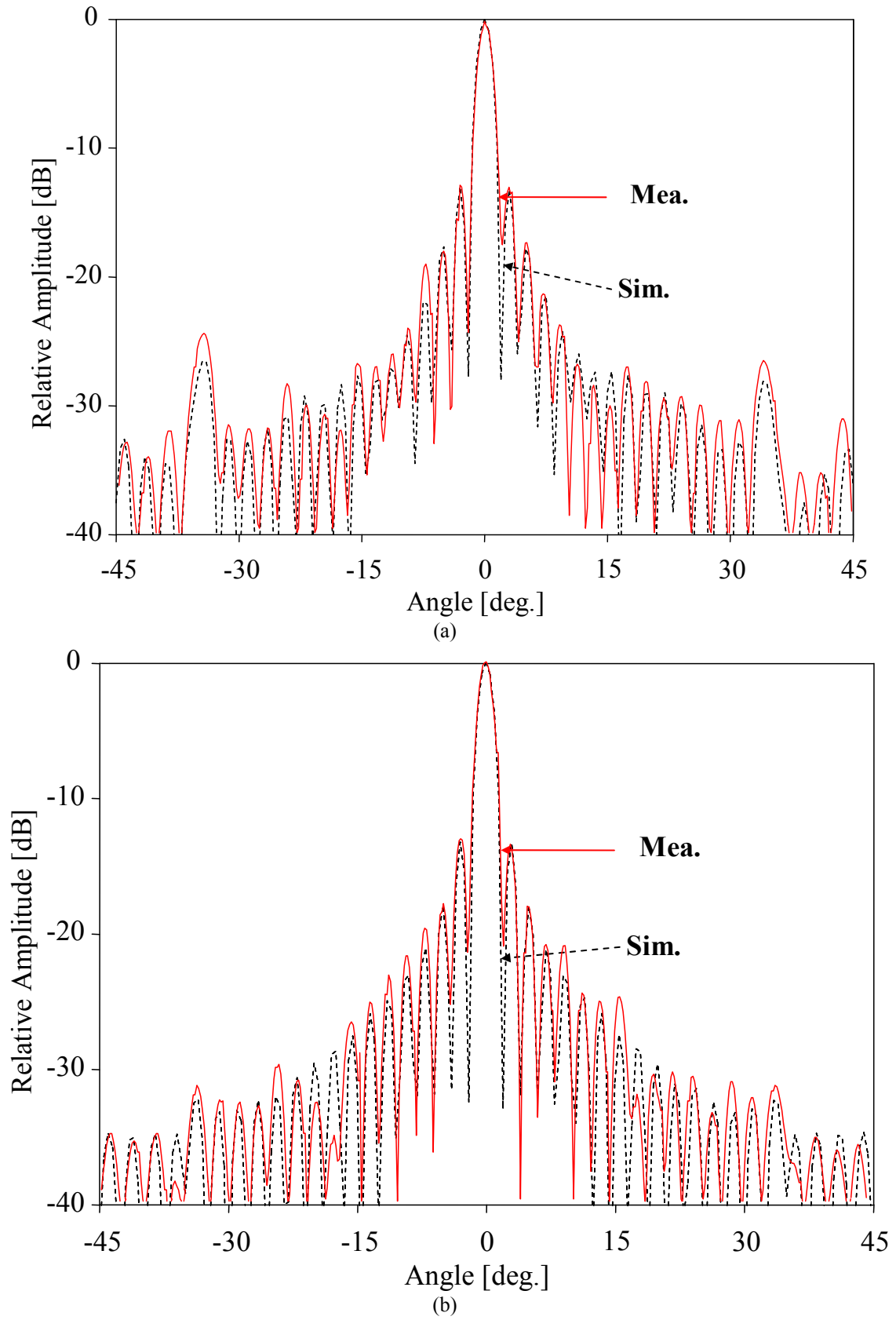


Fig. 14 Calculated and measured radiation patterns of 32x32-element array antenna with the new feeding structure (a)E-plane (yz-plane) and (b)H-plane (xz-plane).

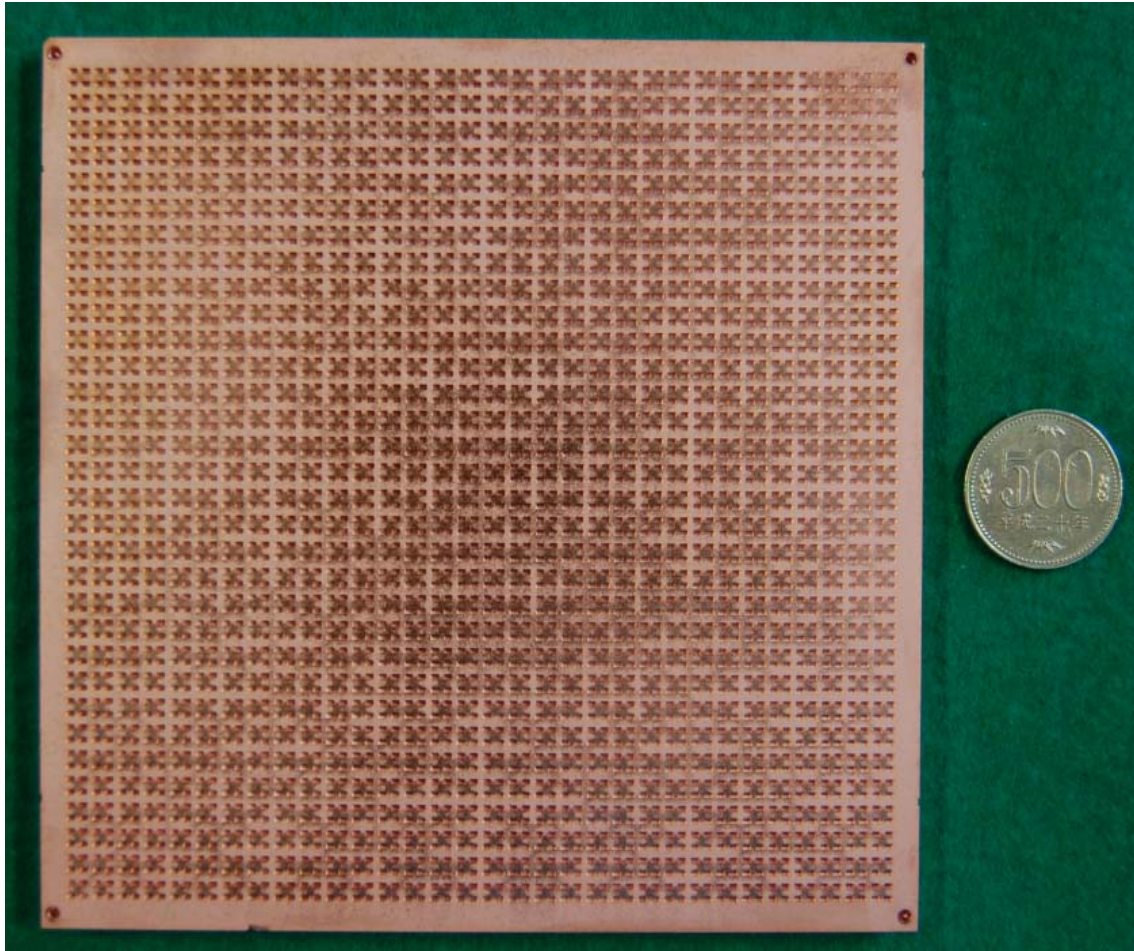


Fig. 15 Picture of the fabricated 64x64-element array antenna with the new feeding structure for the 120 GHz band.

3-6-2 64x64-element Array Antenna

Fig. 15 shows the 64x64-element array antenna fabricated by diffusion bonding of laminated thin metal plates. One antenna is fabricated for the 64x64-element array. Same as the 32x32-element arrays, the new feeding structures are used for 64x64-element array, but the parameters of the radiating parts are not changed at all from the conventional antenna structure. The aperture area is defined as 134.4 mm square (2.1 mm x 64 elements). The antenna height is 2.4 mm.

Fig. 16 shows the frequency characteristic of the reflection at the feed aperture of the 64x64-element array. Also the calculated reflection of the 64x64-element array antenna with the new feeding structure is added in Fig. 16. In the simulation, the bandwidth of the reflection for VSWR less than 1.5 is 8.6%. The measured reflection of the 64x64-element array also shows same manner with the results of 32x32-element

arrays. However, higher ripples than that of the 32x32-element array appear in both of the calculated and measured results. These ripples may come from the multi-reflection in the long feeding network.

Fig. 17 shows the calculated and measured gains of the 64x64-element array antenna. The measured antenna efficiency of the 64x64-element array satisfies 50% antenna efficiency over a broad bandwidth from 118.5 GHz to 133.0 GHz even though there are many ripples on the measurement result. The ripples in the measured antenna gain came from the instability of the measurement environment including shielding problems and a short measurement distance and multi-reflection of antenna reflection. The conductor loss and the insertion loss are included in all gain data for the both of the simulation and measurement. In calculation using HFSS, the conductivity losses at 125 GHz is about 1.2 dB for the 64x64-element array when the conductivity of copper is assumed to be 5.8×10^7 S/m.

Fig. 18 shows the calculated and measured radiation patterns of the 64x64-element array antenna. We could measure within ± 20 degrees for the 64x64-element array antenna in this time. The half-power beam width is 0.8 degrees for the both of E- and H-planes. The dynamic range in the measurement was not sufficient to accurately measure sidelobe levels far from the main lobe; however we can clearly see good agreements of the main lobe and the first side lobe levels between the calculation and the measurement.

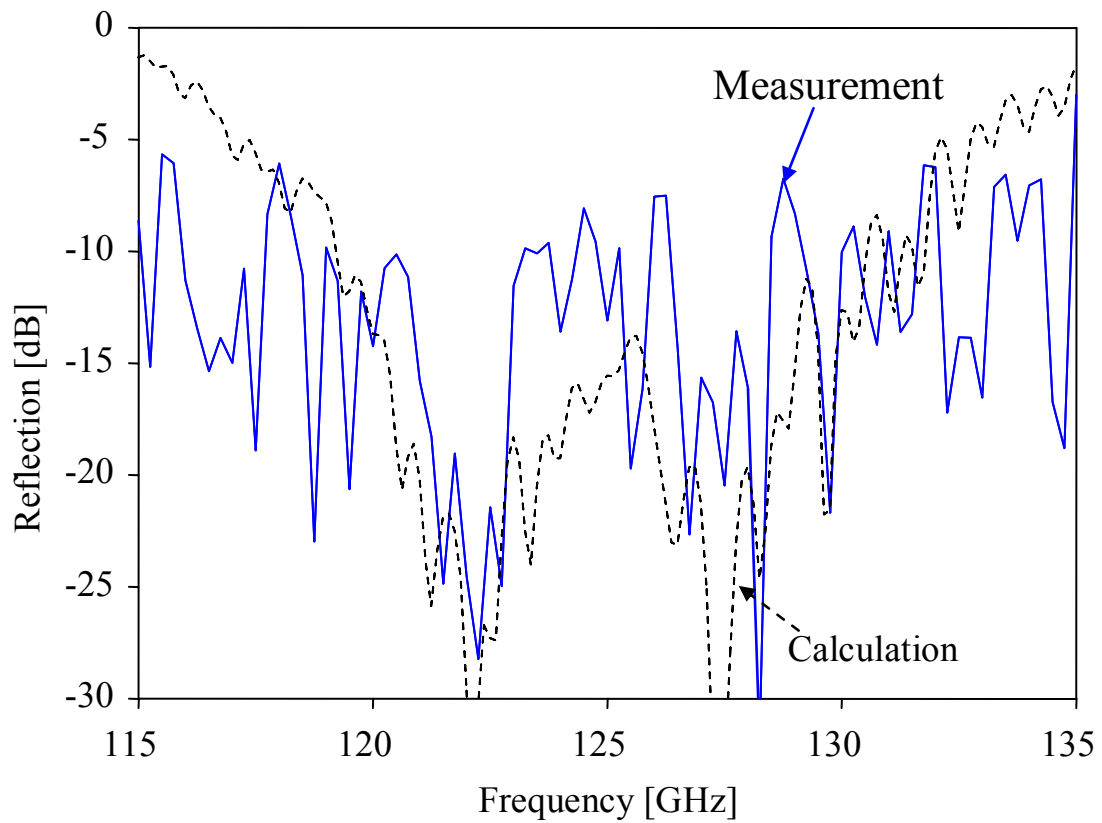


Fig. 16 Calculated and measured reflection coefficients of 64x64-element array antenna with the new feeding structure for 120 GHz

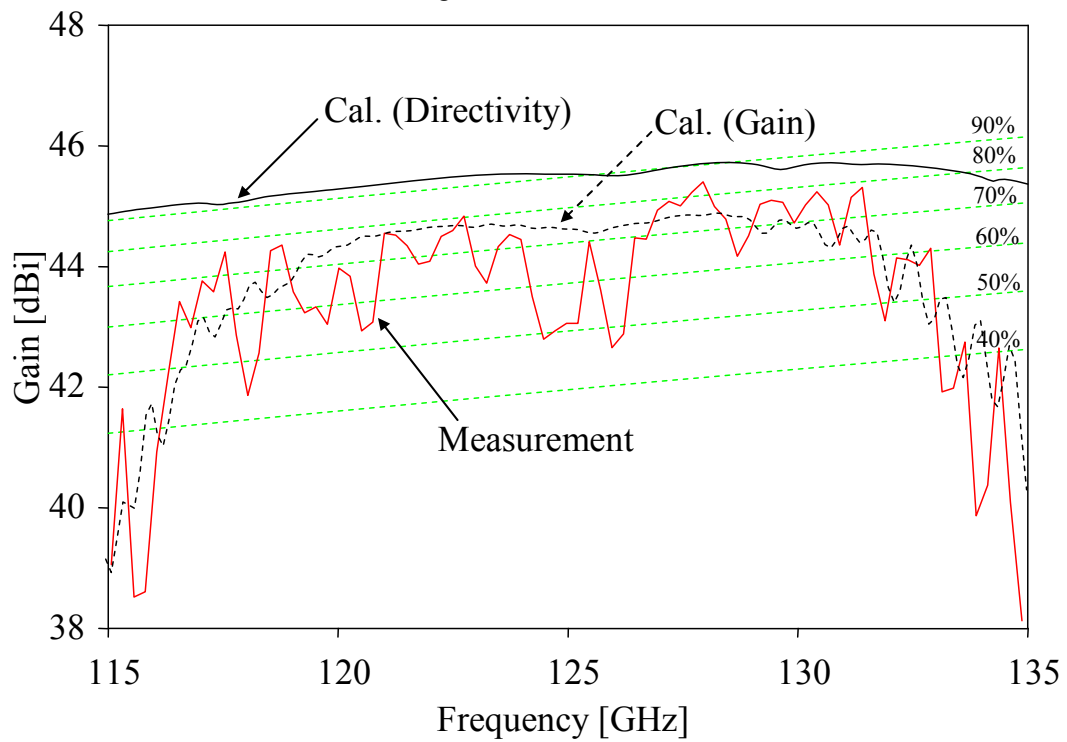


Fig. 17 Calculated and measured gain and efficiency of 64x64-element array antennas with the new and conventional feeding structures for 120 GHz

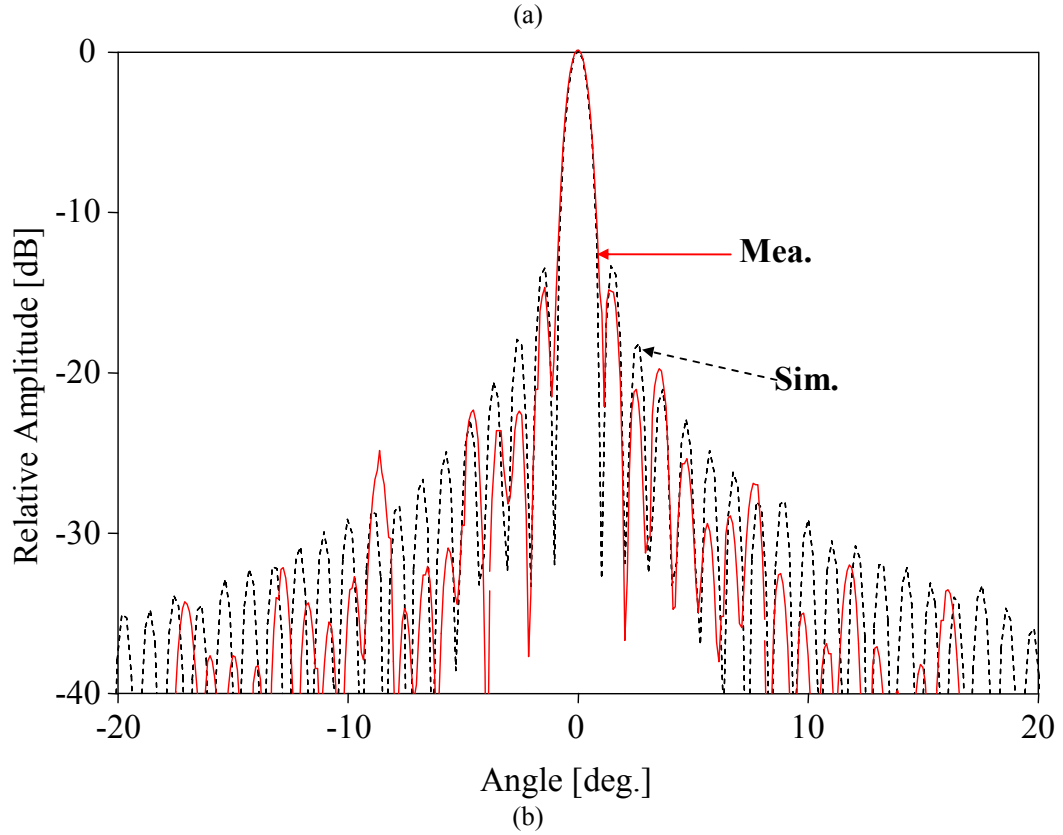
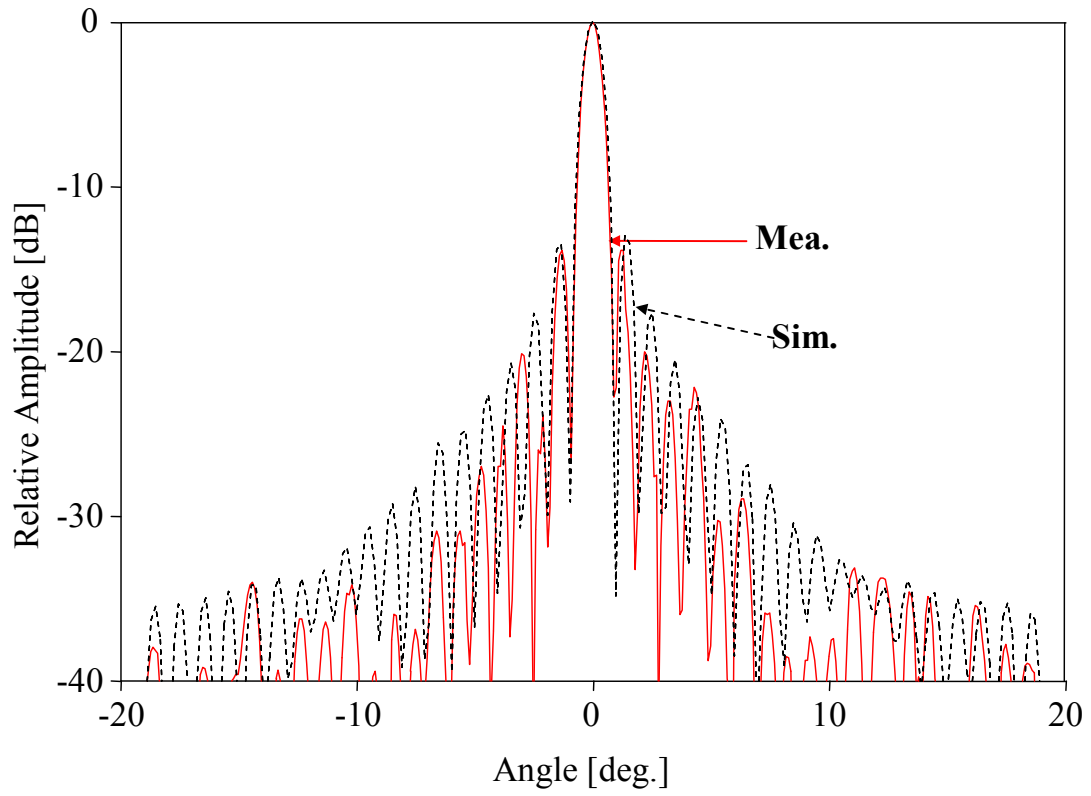


Fig. 18 Measured radiation patterns of 64x64-element array antenna with the new feeding structure
(a) E-plane (yz-plane) and (b) H-plane (xz-plane).

3-7 Measurement Configuration Description for Large Size Array Antennas

All of the reflection and gain are measured with the VNA. The measurement configuration of the gain is described in Fig. 19. The gains are measured with comparison method in the configuration in Fig. 19 (a). The STD (standard gain horn) having 23.5 dBi gain is used for the gain comparison. Basically, the distance of 2.5 m is not satisfying the far field condition ($\text{far field} = 2D^2/\lambda$, where D =antenna maximum diameter and λ = wavelength in free space) for both of the 32x32-element and the 64x64-element arrays. Thus the measured antenna gain is corrected with the calculation with array factor. The radiation patterns also measured in the configuration in Fig. 19 (a) with an antenna rotator. The gains are measured at intervals of 0.25 GHz from 115 GHz to 135 GHz for the all antennas in this manuscript. The antenna heights are 1.5 m on the ground. This configuration is a quasi-anechoic environment surrounded by several absorbers. The ripples on the measurement results may come from the instability of the measurement environment including shielding problems and a short measurement distance. Also, we have measured the antenna gain and radiation pattern patterns outdoors at a distance of 300 m as shown in Fig. 19(b) using the MMIC devices described in [12]. In here, we can measure only 125 GHz. This configuration is set up on the top of two buildings over 20 m height. The results in Fig. 6, 7, 14(a), 14(b), and 17 are measured in configuration in Fig. 19 (a) and corrected. The results in Fig. 18 are measured in configuration in Fig. 19 (b). Also, we compare the antenna gains between the results corrected from the measured results in configuration in Fig. 19(a) and measured in configuration in Fig. 19(b), they show a good agreement.

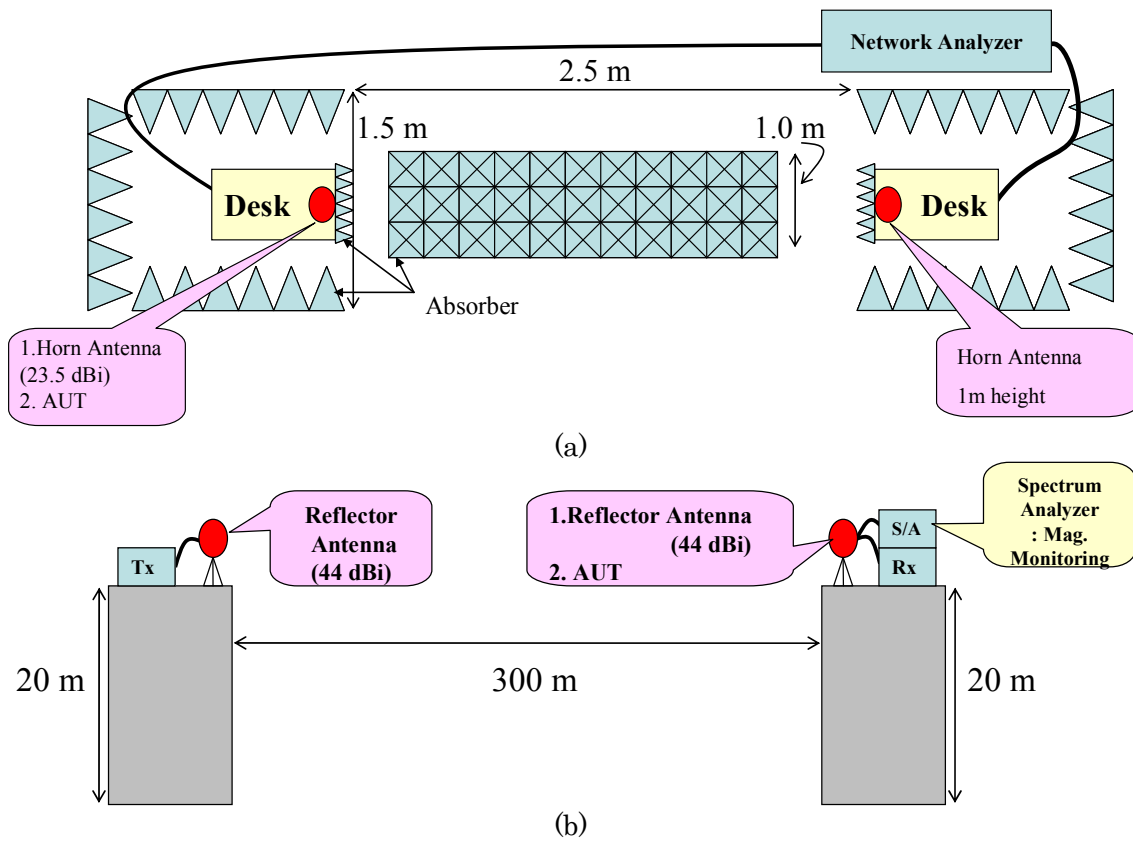


Fig. 19 Configurations of the measurement system. (a) indoor and (b) outdoor

3-8 Concluding Remarks

A 32x32-element slot array antenna and a 64x64-element array antenna are designed for the 120 GHz band. To make a stable structure for the fabrication method of diffusion bonding of laminated thin metal plates, we propose a double-layer feeding structure. With the proposed feeding structure, we successfully fabricate high gain antennas having low profiles. A 32x32-element array antenna shows 38 dBi antenna gain with 60% antenna efficiency over 15 GHz bandwidth and a 64x64-element array shows 43 dBi antenna gain with 50% antenna efficiency over 14.5 GHz bandwidth, respectively. To author's knowledge, these antennas are the first prototypes having high antenna gains such as 38 dBi and 43 dB high antenna efficiency and broad bandwidth of about 10% with planar antenna structure in the world. We hope that these antennas could be alternatives for conventional high gain antennas such as reflector and lens antennas.

References in Chapter 3

- [1] A. Mizutani, K. Sakakibara, N. Kikuma, and H. Hirayama, "Grating Lobe Suppression of Narrow-Wall Slotted Hollow Waveguide Millimeter-Wave Planar Antenna for Arbitrarily Linear Polarization," *IEEE Trans. Antenna and Propag.*, vol. 55, no.2, pp.313-320, Feb. 2007.
- [2] E. Levine, G. Malamud, S. Shtrikman, and D. Treves, "A study of microstrip array antennas with the feed network," *IEEE Trans. Antennas Propag.*, vol. 37, no. 4, pp. 426–434, Apr. 1989.
- [3] L. Pazin and Y. Leviatan, "Effect of amplitude tapering and frequency dependent phase errors on radiation characteristics of radial waveguide fed non-resonant array antennas," *IEE Microw. Antennas Propag.*, vol. 151, no. 4, Aug. 2004.
- [4] S. S. Oh, J. W. Lee, M. S. Song, and Y. S. Kim, "Two-layer slotted waveguide antenna array with broad reflection/gain bandwidth at millimetre- wave frequencies," *IEE Proc.-Microw. Antennas Propag.*, vol. 51, no. 5, pp. 393–398, Oct. 2004.
- [5] K. Jung, H. Lee, G. Kang, S. Han, and B. Lee, "Cavity-backed planar slot array antenna with a single waveguide-fed subarray," in *Proc. IEEE Antennas Propag. Soc. Int. Symp.*, 115.5, Jun. 2004.
- [6] B. Lee, K. Jung, and S. Yang, "High-efficiency planar slot array antenna with a single waveguide-fed cavity-backed subarray," *Microw. Opt. Technol. Lett.*, vol. 43, no. 3, pp. 228–231, Nov. 2004.
- [7] D. Kim, J. Hirokawa, M. Ando, and T. Nagatsuma, "Design and fabrication of a corporate-feed plate-laminated waveguide slot array antenna for 120GHz-band", *Proc. of 2011 IEEE AP-S URSI Intl. Symp.*, 429-6, Jul.2010
- [8] R.W. Haas, D. Brest, H. Muegggenburg, L. Lang and D. Heimlich, "Fabrication and performance of MMW and SMMW platelet horn arrays," *Intl. J. of Infrared and Millimeter Waves*, Vol. 14, Num. 11, pp. 2289-2294, Jan.1993
- [9] M. G. Nicholas and R. M. Crispin, "Diffusion bonding stainless steel to alumina using aluminum interlayers," *J. of Materials Science*, vol. 17, No. 11, pp.3347-3360, 1982.
- [10] S. Silver, *Microwave antenna theory and design*. IET, 1949.
- [11] Ansys, Inc. Available: <http://ansys.com/>

- [12] A. Hirata, R. Yamaguchi, T. Kisugi, H. Takahashi, K. Murata, T. Nagatsuma, N. Kukutsu, Y. Kado, N. Iai, S. Okabe, S. Kimura, H. Ikegawa, H. Nishikawa, T. Nakayama and T. Inada, "10-Gbit/s wireless link using InP HEMT MMICs for generating 120-GHz-band millimeter-wave signal," *IEEE Trans. Microwave Theory Tech.*, vol. 57, no. 5, pp. 1102-1109, May 2009.

Chapter 4

Mutual Coupling Control Technique using Cavity for Small Size Array

4-1 Introductory Remarks

The existence of mutual coupling among radiating elements in an array antenna affects on the antenna performances such as bandwidth, gain and efficiency [1]-[3]. It cannot be neglected in the design of an array antenna because each radiating element is much affected by the mutual coupling from the neighbored radiating elements. In a large size array which has a number of radiating elements, most of the radiating elements have the same electromagnetic environment. Consequently, the majority of the array elements show similar electromagnetic characteristics. Only the outer elements around the edge of the array show different electromagnetic characteristics known as edge-effect due to the different mutual coupling effects, but they are not contributed much on the array antenna characteristics because they are minor in number compared with the total number of elements. Thus it can be neglected their effects on the antenna design for the large size array.

On the other hand, in a small size array, we should consider the different electromagnetic characteristics of the outer elements because the number of the outer

elements is significant compared with the total number of elements. The different mutual coupling effects among the elements in the small size array consequentially result in degradation of the antenna performance. Usually, they increase a main beamwidth with reduction in gain and increased sidelobe levels [4], [5].

Many studies have been proposed to adjust the mutual coupling effects in an array antenna. A dummy radiating element next to the outer radiating element gives a nearly same environment for the outer elements as those elements further way from the edge [6]. Also, a box-shaped metallic wall [7], periodic structures as electromagnetic band gap [8], baffles and the cavities are widely used for the mutual coupling control in array antenna design [9]-[11]. The point of these mutual coupling control techniques is to make each radiating element have same characteristics even in the different environments because the uniform characteristics of the radiating elements guarantee a good radiating performance of the array antenna. Also, the intent of this study is to realize all radiating elements have same characteristic with adjustment of the mutual coupling so that the array antenna have a good radiating performance even in a small size array.

We have already proposed a large size array of hollow-waveguide slot antennas having high gain and antenna efficiency in the previous chapters. The antenna in chapter 2 and 3 are proposed for the millimeter wave wireless link systems as an alternative to reflector and lens antennas which have a high antenna gain for the outdoor applications [12]-[13]. The 16x16-element antennas achieved more than 70% antenna efficiencies with more than 32 dBi over 10% bandwidths in the 60 GHz and 120 GHz bands, respectively. Based on the antenna structure in [12]-[13], we design a small size array having less radiating elements for the data communication systems of indoor applications. To prevent the degradation of the radiation performance in a small size array, we introduce a cavity structure for adjusting the mutual coupling. Fig. 1 shows the proposed antenna structure having 4x4 elements with the cavities. The cavities provide proper mutual coupling effect for each radiating element; as a result, each radiating element could maintain their own radiating characteristics.

In this chapter, we observe the mutual coupling effects in small size array and propose the method of the mutual coupling adjustment using cavities.

4-2 Antenna Configuration

Fig. 1 shows the configuration of a 4x4-element double-layer corporate-feed hollow-waveguide slot array antenna with cavities. This antenna structure is almost identical in comparison with the antenna structures for the large size array described in [12], [13] except the layer for the cavities. There are cavities and radiating slots with a 0.86-wavelength interval on the top layer. The antenna is fed through an input aperture located in the lower layer. All the radiating elements are fed with equal phase and amplitude by the corporate-feed network consisting of an H-plane H-junction. Fig. 2 shows the configurations of the 2x2-element subarray with cavities as the design unit to be fed by the full-corporate-feed structure. Note that, there are two kinds of cavities. The cavities on each radiating slot are for the mutual coupling adjustment and they are not included in the conventional structure in [12], [13]. Also there is a cavity under the radiating slots and now we call this cavity as ‘cavity (backed)’. The four radiating slots are fed in phase and with equal amplitude since the signal from the feeding waveguide is divided into four ways with in phase and with equal amplitude in the cavity (backed).

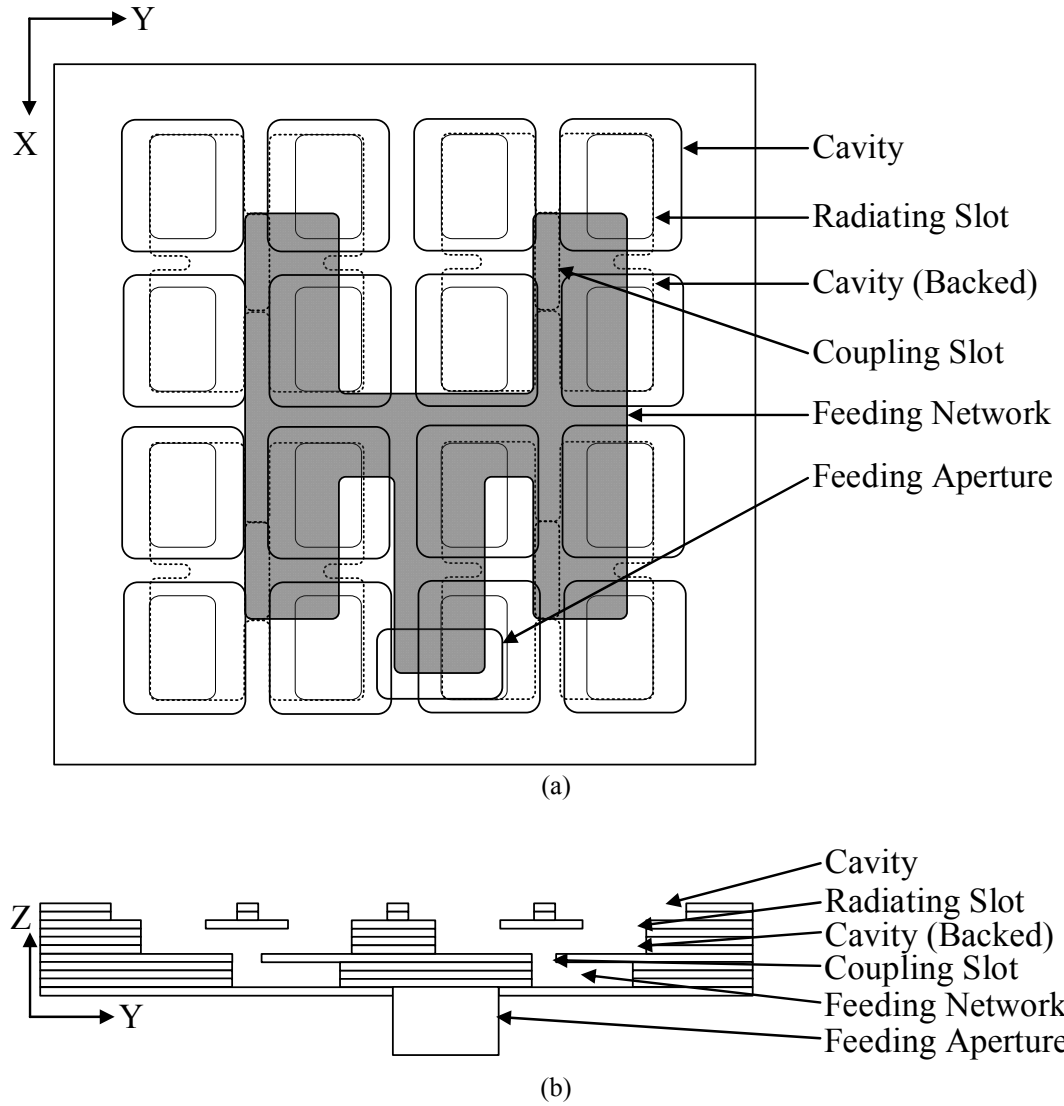


Fig. 1 Antenna configuration of the 4x4-element array antenna with cavities (a) Top-View (xy-plane) and (b) Side-View (yz-plane).

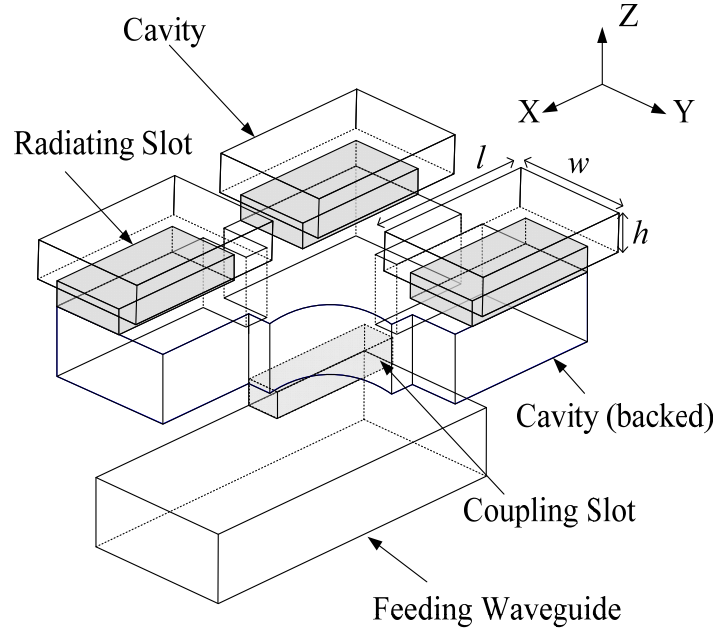


Fig. 2 Exploded perspective view of the 2x2-element subarray with cavities.

4-3 Mutual Coupling Effects in Small Size Array

4-3-1 Mutual Coupling Effect for Single Radiating Element

The mutual coupling effects among the radiating elements are very important in the conventional antenna structure without cavities because it has a wide bandwidth due to its dual-resonance caused by the slot self-resonance and the external mutual coupling [12]. To observe the effect of mutual coupling for a single radiating element in the conventional antenna structure, we compare their characteristics dependent on the existence of mutual coupling effects. We model two unit models having a single radiating element with and without periodic boundaries. To analyze the cavity (backed) and the coupling slot, they are also modeled with PEC and PMC boundaries. Fig. 3 (a) shows the unit model with periodic boundaries (Unit model A) and Fig. 3(b) shows the unit model without periodic boundaries (Unit model B). Note that, the external region of the unit model A is bounded by periodic boundaries whereas the unit model B is bounded by radiation boundaries. Thus the unit model A is equivalent to the single radiating element in an infinite array while the unit model B is equivalent to the isolated radiating element in half free space. Fig. 4 shows frequency responses of the

S-parameters (magnitude and phase) at each port of unit models. As shown in here, their frequency responses are very different. The unit model A shows a broad bandwidth with a dual-resonance and this result is very similar with that of the large size arrays in [12], [13]. However the unit model B shows very different characteristics compared in that of the unit model A due to the absence of the mutual coupling effects even though it has an identical structure with the unit model A.

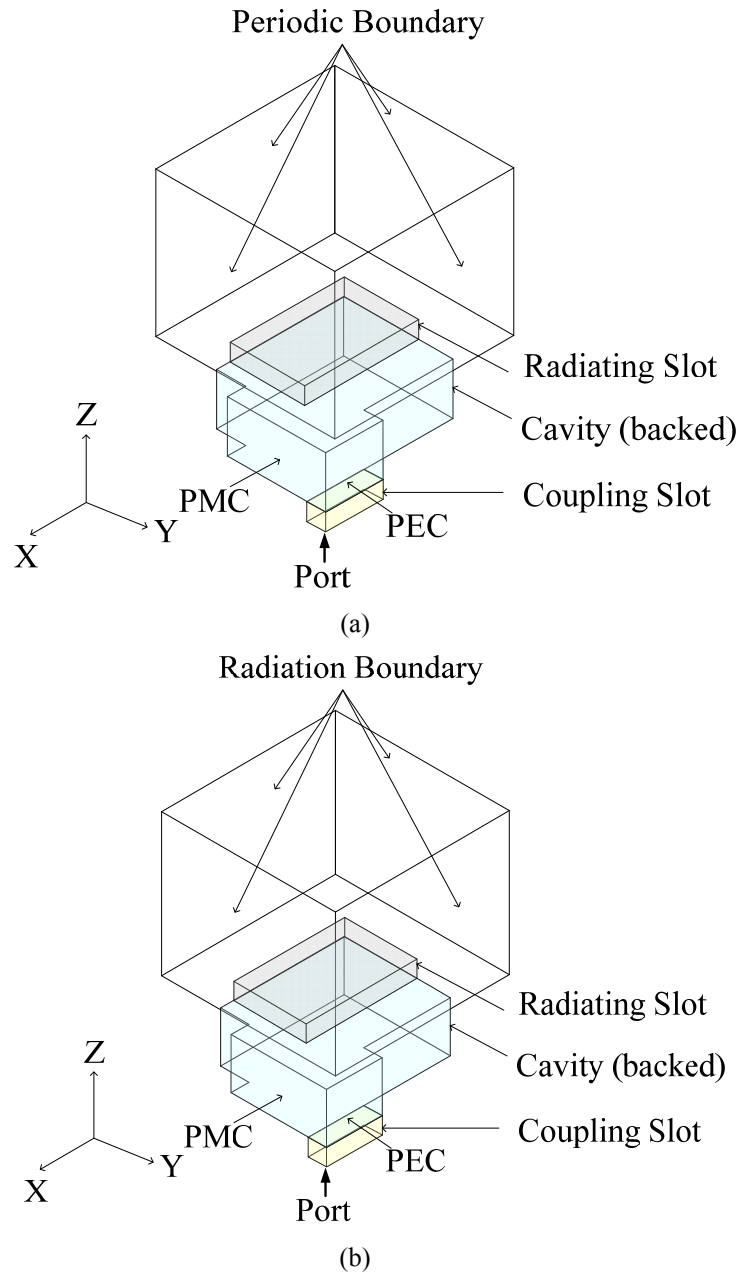


Fig. 3 Unit models to analysis one slot (a) unit model A with periodic boundaries and (b) unit model B without periodic boundaries (isolated slot).

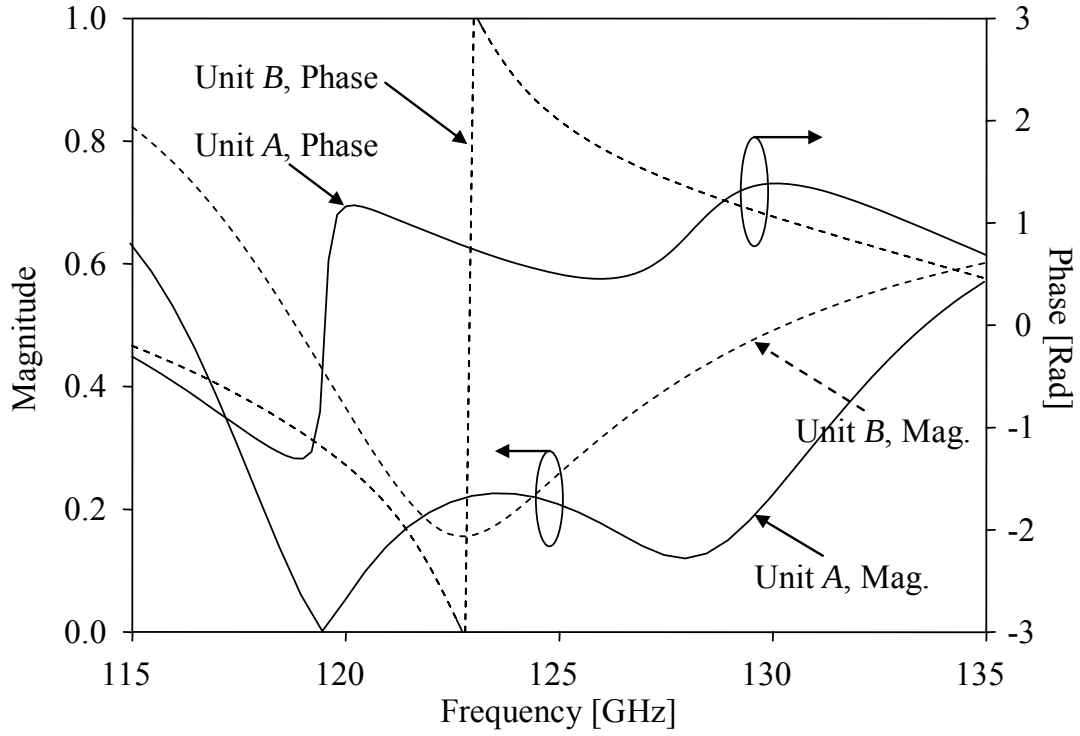


Fig. 4 Calculated S-parameters (magnitude and phase) of the unit model A and B.

4-3-2 Mutual Coupling Effect for the 4x4-element Array

In previous section, we confirm the different characteristic of a single radiating element dependent on the existence of the mutual coupling effects. Their different characteristics dependent on the mutual coupling effects also affect the characteristics of the array antenna. Because the inner radiating elements of the array antenna are fully surrounded by the neighboring radiating elements, they show almost identical characteristics with that of the unit model A. On the other hand, the outer slots may show different characteristics since they are not fully surrounded by the neighboring radiating elements. The different characteristics of the radiating elements depend on their position generally cause a degradation of the array antenna performance. As explained in the introduction, it especially affects to the small size array antenna; thus we consider the small size array having 4x4-element array. Fig. 5 shows the calculated gain and efficiency of a 4x4-element array with and without periodic boundaries. The 4x4-element array with periodic boundaries shows 21 dBi gain with over 80% antenna efficiency at the center frequency while it shows only 19 dBi gain with 57% antenna efficiency without the periodic boundaries at the center frequency. Also it shows poor

bandwidth without the periodic boundaries compared with that with the periodic boundaries. This degradation of the gain and bandwidth without periodic boundaries is mainly caused the different characteristics among the radiating elements; it causes high sidelobes in the radiation pattern. Fig 6 shows the radiation patterns of the 4x4-element array in the xz -plane (E-plane) with and without the periodic boundaries calculated at the center frequency. The sidelobes are highly increased when the periodic boundaries are not adopted in the external region.

We confirm that it is difficult to expect a good antenna performance with the conventional structure due to the different characteristics among radiating elements in a small size array. In the following chapter, we will introduce the cavity structure for the adjusting the mutual coupling to realize the same characteristics among radiating elements in a small size array antenna such as a 4x4-element array.

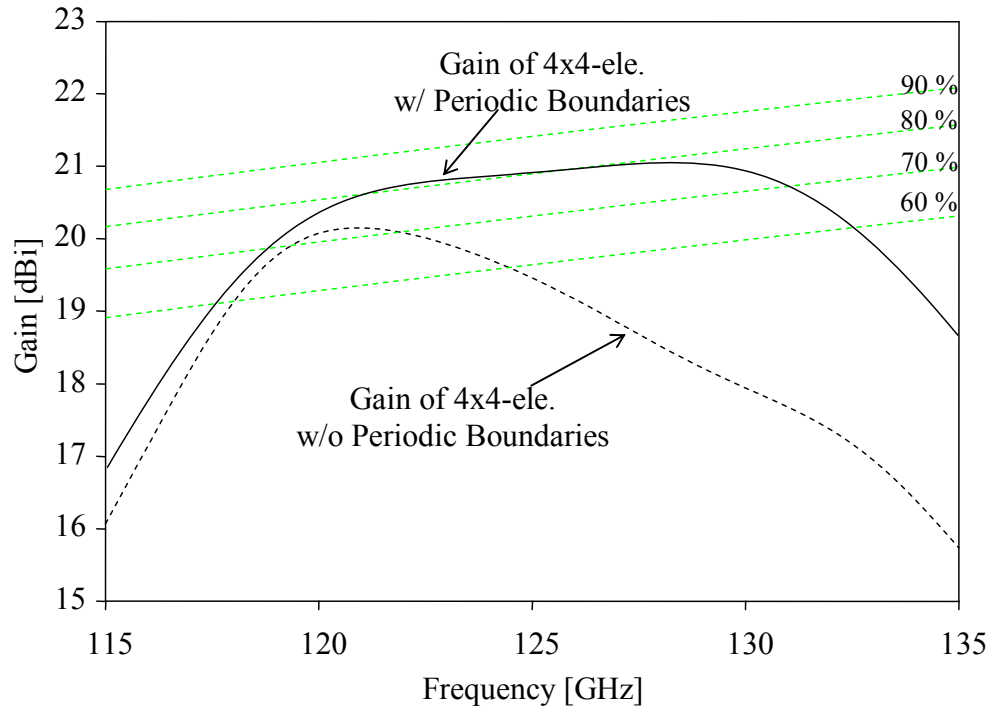


Fig. 5 Calculated gains and efficiencies of the 4x4-element array with and without periodic boundaries.

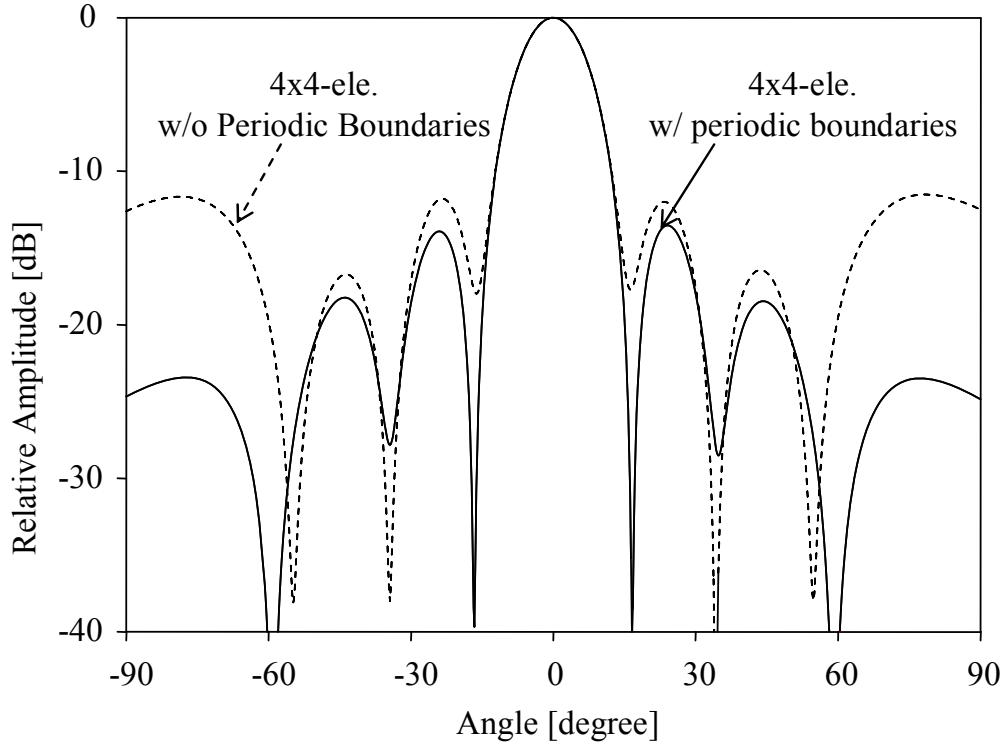


Fig. 6 Calculated radiation patterns of the 4x4-element array with and without periodic boundaries in E-plane (yz-plane).

4-4 Mutual Coupling Adjustment with Cavity

4-4-1 Mutual Coupling Adjustment with Cavities for Each Radiating Element

Based on the results in the previous section, we propose a cavity structure for the mutual coupling adjustment for each radiating element as shown in Fig. 1 and 2. The main purpose of the cavities is to realize the same characteristics among all radiating elements in the array antenna. The cavity walls located near each radiating element are working as perfect electric conductor (PEC) walls which make image magnetic currents flowing in the same direction with the magnetic currents generated by slots themselves [14]. Thus, each radiating element ‘feel’ as there are neighbored radiating elements even though each slot is actually isolated with the cavity walls.

The dimensions of the cavity are carefully determined so that each radiating element with a cavity may have equal characteristics to that of the unit model A assumed for an infinite array. To determine the cavity dimensions, we use another unit model having a cavity on the radiating element. Fig. 7 shows the unit model with a cavity (unit model C). Note that, the external region of the unit model C is bounded by radiation boundaries so that we can analyze only a slot itself without the mutual coupling effects from neighbored radiating elements. All parameters of the unit model C are fixed same as the structure of the unit model A except the cavity dimensions. With comparison of the characteristic of these two unit models, we determine the cavity dimensions. The design procedure is as follows.

- (1). The cavity dimensions of the unit model C are determined to realize an S-parameter (magnitude and phase) equal to that of the unit model A.
- (2). All antenna parameters including the slot, the cavity (backed) and the coupling slot of the unit model C are examined for wide band impedance matching with the cavity dimensions determined in procedure 1.
- (3). With determined antenna parameters in procedures 1 and 2, the new 2x2-element subarrays with cavities are combined for 4x4-element array design with the full-corporate-feeding network.

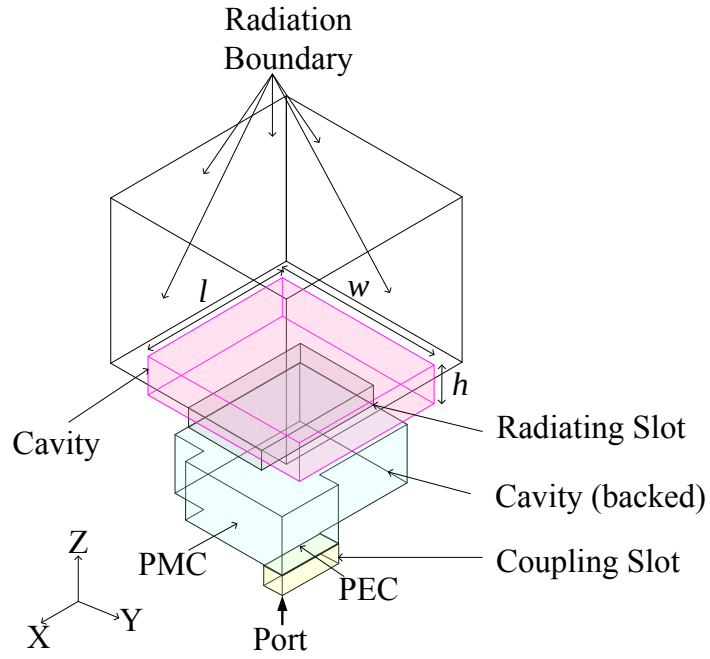


Fig. 7 Unit models to analysis one slot with a cavity. (Unit model C).

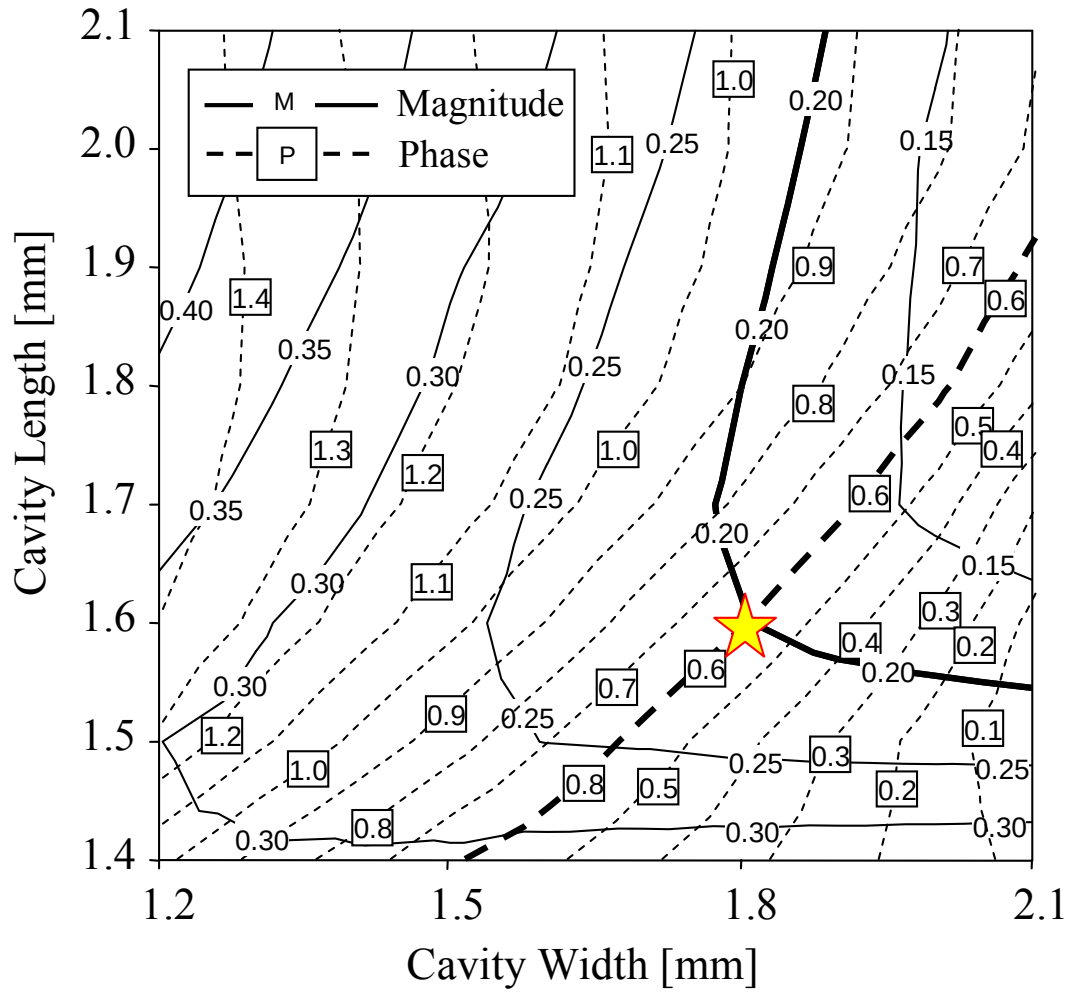


Fig. 8 S-parameter characteristics of unit model C depend on the cavity dimensions.

4-4-2 Effect of Cavity Dimensions for Slot Characteristics

Firstly, we observe the change of the S-parameters (magnitude and phase) of the unit model *C* at the center frequency (125 GHz) with two variables; the cavity width *w* and the cavity length *l*. The cavity height is fixed as 0.6 mm corresponding to a quarter wavelength at first. The effects of cavities are investigated in HFSS [15]. The contour map in Fig. 8 shows the change of the S-parameters of the unit model *C* according to the cavity width and the cavity length. And the star in the contour map indicates the S-parameter of the unit model *A* at the center frequency. Almost identical S-parameters compared with that of the unit model *A* is obtained with the cavity dimensions of $w=1.80$ mm (0.76λ) and $l=1.60$ mm (0.67λ). Also, we observe the change of the S-parameter with various cavity heights and we find that it is almost not changed when the cavity height is higher than 0.4 mm. Considering the antenna height for a low profile, the cavity height is determined as $h=0.40$ mm (0.17λ). Thus the cavity dimensions are determined as $w=1.80$ mm (0.76λ), $l=1.60$ mm (0.67λ) and $h=0.40$ mm (0.17λ). Also, we can observe almost same changes of S-parameter for the different frequencies. Fig. 9 shows the S-parameters of the unit model *A* and *C* in the broad bandwidth and they show good agreement. It implies that the determined cavity dimensions realize almost same environments compared with that of the slot in an infinite array, even though there are no actual neighbored radiating elements.

4-5 Mutual Coupling among Radiating Elements with Cavity

In the previous section, we confirm that the cavities of the proposed antenna provide proper mutual coupling effect and realize almost same characteristic in comparison with that in infinite array for each element. The cavities are also working as a choke and suppress the contribution from the neighbored radiating elements. As a result, the realized characteristics of each radiating element with a cavity could be maintained in their array. To confirm the effect of the cavities for mutual coupling suppression, we compare the actual mutual coupling values (S_{ij}) of two cases with and without the cavities. Fig. 9 shows the configurations to compare the mutual coupling

among elements. There are four sets of the unit model A in Fig. 9 (a) and there are four sets of the unit model C in Fig. 9 (b). Note that, both of them are surrounded by periodic boundaries to assume the infinite array in this time.

They are arranged with 0.86 wavelength intervals same as the actual array structure in [12][13]. Their amounts of the mutual couplings are compared in Fig 10. In the conventional structure without a cavity, we can see strong mutual coupling of -10.2 dB between horizontally located sets in the x direction (S_{21} , same as E-field direction). On the other hand, the mutual coupling between parallel sets in the x direction is well suppressed in the new structure with the cavities; it shows only -26.5 dB at the center frequency. The mutual coupling between vertically located (S_{31}) and diagonally located sets (S_{41}) are also suppressed below -30 dB at entire frequency band in the new structure. This result implies the proposed antenna structure with cavities is not affected by the mutual coupling from the neighbored radiating elements. Consequentially, the characteristics of the single radiating element with a cavity are not changed in any position in the array antenna.

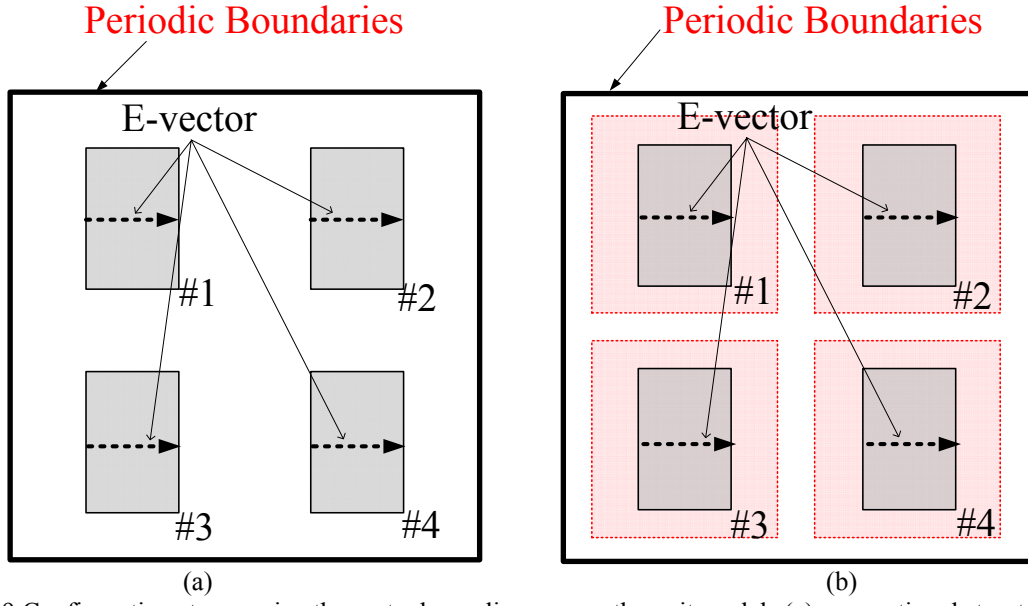


Fig. 9 Configurations to examine the mutual coupling among the unit models (a) conventional structure (b) new structure.

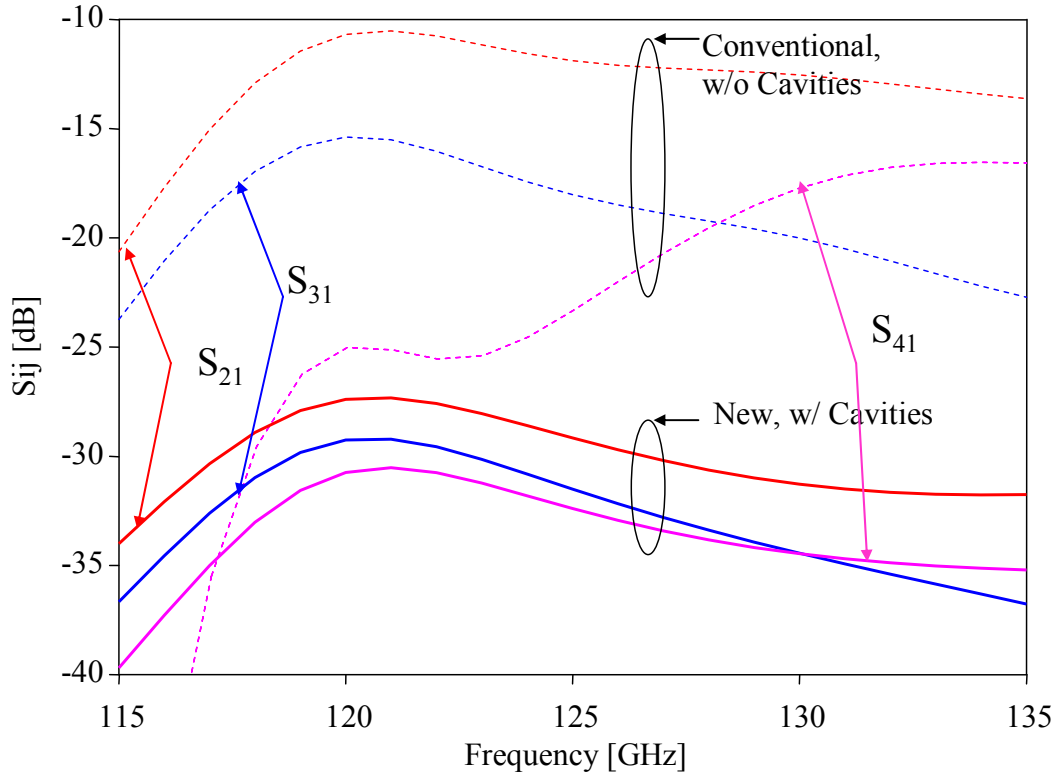


Fig. 10 Mutual Couplings among four elements of unit models A and C

4-6 Design and Fabrication Results with Cavity

Fig. 11 shows the fabricated antenna by diffusion bonding technique [16]. The 4x4-slots and the 4x4-cavities are arranged with 0.86-wavelength interval on the top layer. The aperture size is defined as 8.4 mm square (2.1mm x 4 elements) for the 120 GHz band. This structure is composed of 4-sets of 2x2-element subarrays with cavities and the H-junction feeding network. Each 2x2-element subarray with cavities is connected to each end of the H-junction. The reflection coefficient of the H-junction is well suppressed at less than -30 dB [12][13]. The antenna height is 2.0 mm including all layers while it was 1.6 mm in the previous antenna design.

Fig.12 shows the calculated and measured frequency characteristics of the reflection in the 4x4-element array antenna with the cavities. They show good agreement between the calculation and measurement. The reflection bandwidth of the 4x4-element array with the cavities is 12.0% for VSWR < 1.5. Also, the calculated result of the 4x4-element array with the conventional 2x2-element subarray without cavities is added in Fig. 12 to compare the effect of the cavities. It shows only 5.0% bandwidth for VSWR<1.5. Fig. 13 shows the calculated and measured gains and the efficiencies of the 4x4-element array with the cavities. The gain and efficiency characteristics also show good agreement between the calculation and the measurement. Also we can see a broad bandwidth and high antenna efficiency characteristic with cavities. The peak gain of the 4x4-element array with cavities is 21.2 dBi with 80% antenna efficiency while it is only 20.2 dBi with 71% without the cavities where the conductivity of copper is assumed to be 5.8×10^7 S/m. The gain and radiation pattern measurements are performed with the measurement system described in [17]. There are some ripples on the measured result of gain due to the instabilities of the measurement system such as imperfect shielding. Fig. 14 shows the calculated and measured radiation patterns of the 4x4-element array with the cavities. The radiation patterns are measured in only ± 60 degrees due to the limitation of the measurement setup. The radiation patterns also show good agreement between the calculation and the measurement. The calculated and measured 3-dB beam widths are 15.0 degrees for the E-plane (yz-plane) and H-plane (xz-plane), respectively.

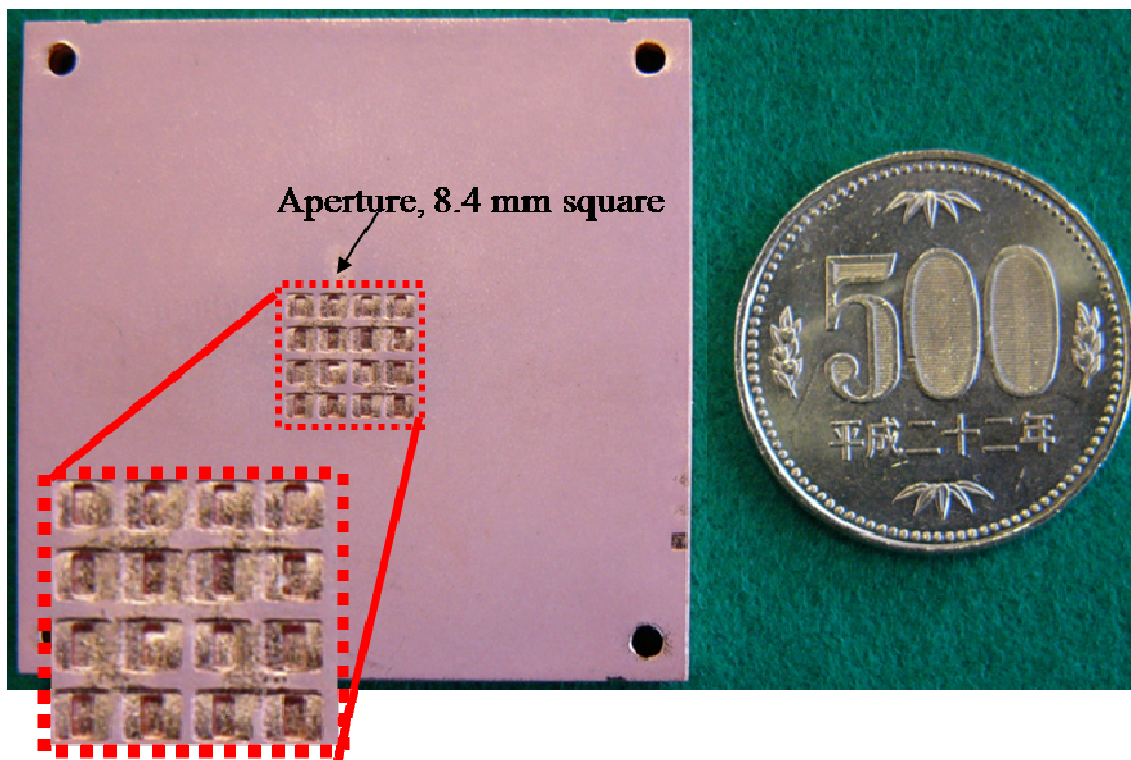


Fig. 11 Picture of the fabricated antenna by diffusion bonding.

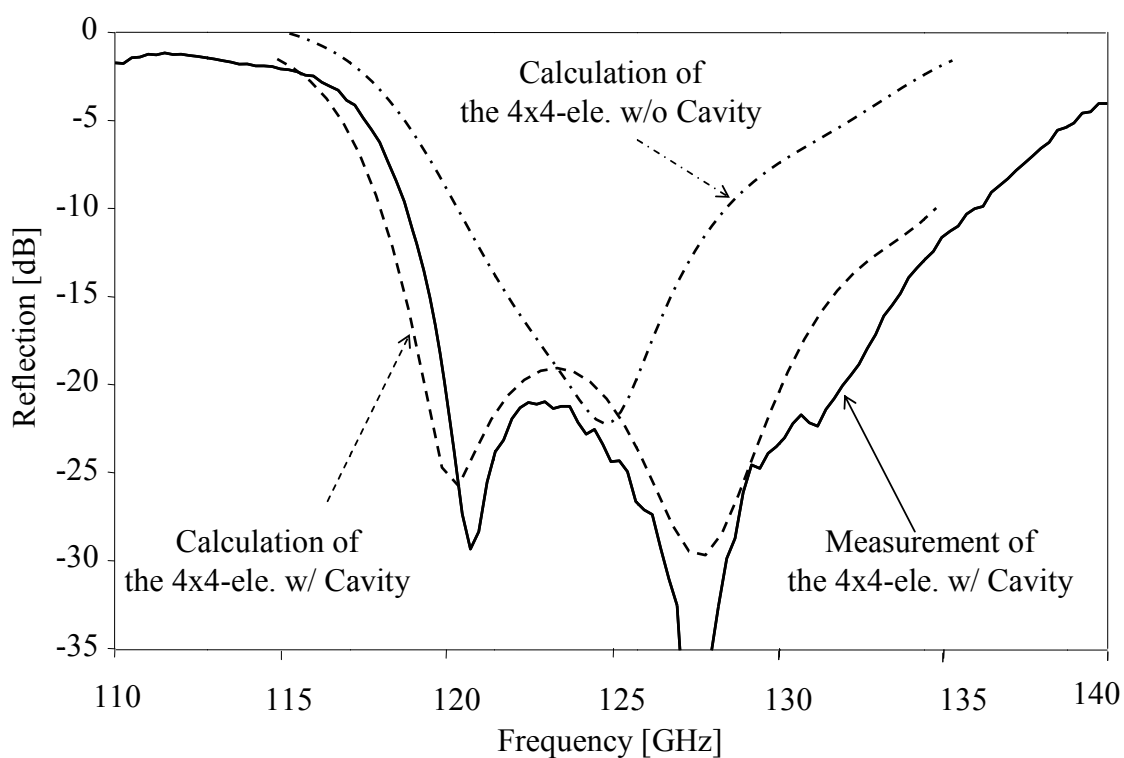


Fig. 12 Calculated and measured antenna reflection of the 4x4-element array with cavities.

4-7 Concluding Remarks

A 4x4-element slot array antenna with the cavities is designed for the 120 GHz band. To adjust the mutual coupling among the radiating elements, the cavity structure is employed for each radiating element. As a result, we have high efficiency over 80% in the broad bandwidth over 18 GHz. Also, the measured 1 dB-down gain bandwidth is 22 GHz (from 118 to 140 GHz) which is corresponding to 17.6% for the center frequency.

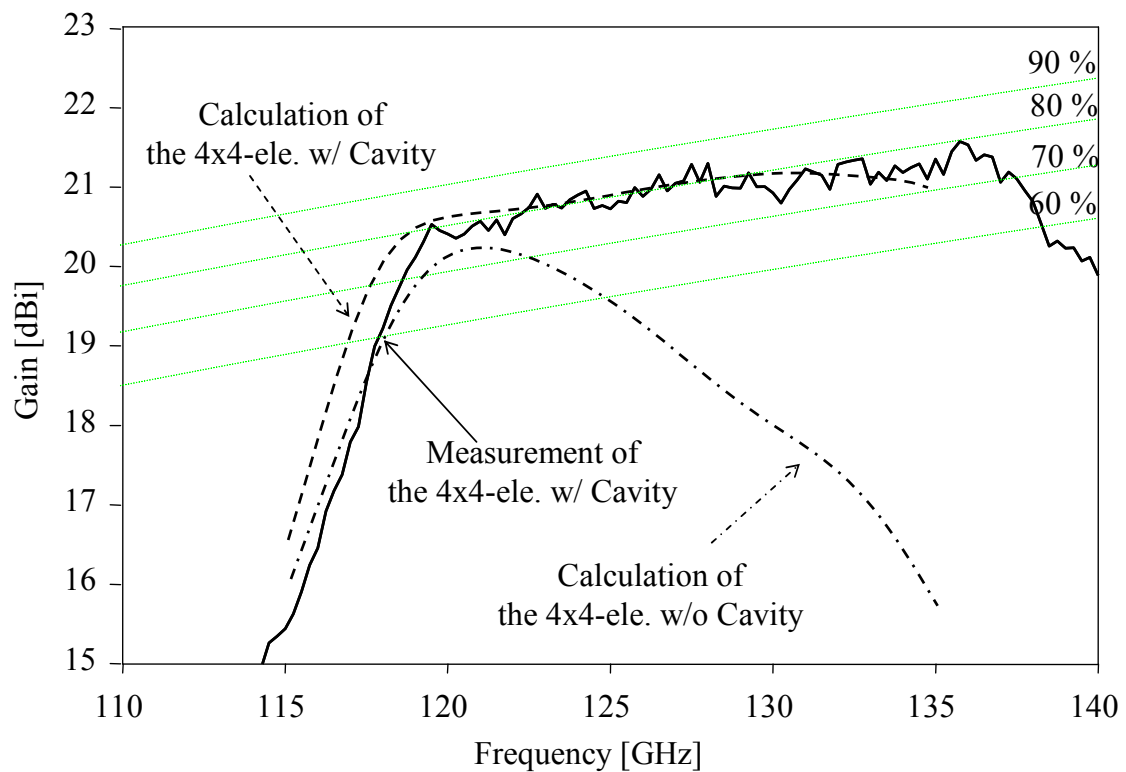


Fig. 13 Calculated and measured antenna gain of the 4x4-element array with cavities.

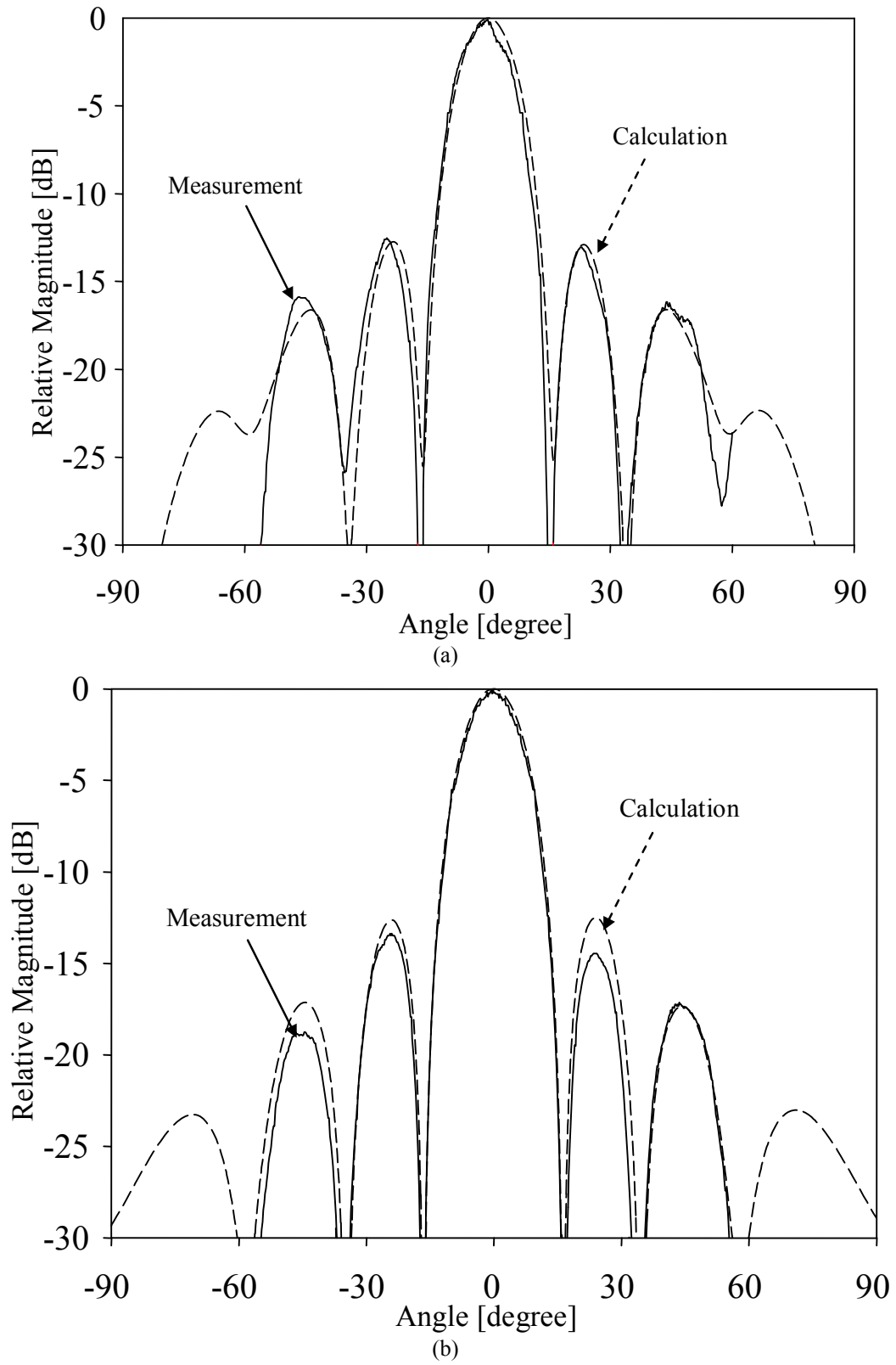


Fig. 14 Calculated and measured radiation patterns of the 4x4-element array with cavities (a) E-plane (yz-plane) (b) H-plane (xz-plane).

References in Chapter 4

- [1] R. S. Elliott and L. A. Kurtz, "The design of small slot arrays," *IEEE Trans. Antennas Propagat.*, vol. AP-26. pp. 216-219, Mar. 1978.
- [2] R. S. Elliott and L. A. Kurtz, "An improved design procedure for small arrays of shunt slots," *IEEE Trans. Antennas Propagat.*, vol. AP-31. pp. 48-53, Jan. 1983.
- [3] H. Steyskal and J. S. Herd, "Mutual coupling compensation in small array antennas," *IEEE Trans. Antennas Propagat.*, vol. AP-38. pp. 1971-1975, Dec. 1990
- [4] D. M. Pozar and D. H. Schaubert, "Scan blindness in infinite phase arrays of printed dipoles," *IEEE Trans. Antennas Propag.*, vol. AP-32, no. 6, pp. 602–610, Jun. 1984.
- [5] R. C. Hansen, *Phased array antennas*. Wiley-Interscience, 1998.
- [6] S. Lundgren. "A study of mutual coupling effects on the direction finding performance of ESPRIT with a linear microstrip patch array using the method of moments," *Proc. of 2011 IEEE AP-S URSI Intl. Symp.*, pp 1372- 1375, July 2011.
- [7] Y. Inoue, H. Sumi and H. Arai, "Pattern calibration free antenna by suppressing mutual coupling between elements," *Proc. of 2004 Intl. Symp. of Ant. and Propag.* pp.17-20, Aug. 2004
- [8] E. Rajo-Iglesias, O. Quevedo-Teruel, and L. Inclan-Sanchez, "Mutual coupling reduction in patch antenna arrays by using a planar EBG structure and a multilayer dielectric substrate," *IEEE Trans. Antennas Propag.*, vol. 56, no. 6, pp. 1648–1655, Jun. 2008
- [9] T. Suzuki, J. Hirokawa, and M. Ando, "Iteration-free design of waveguide slot array with cavities," *IEEE Trans. Antennas Propag.*, vol. 58, no. 12, pp. 3891–3897, Dec. 2010
- [10] S. R. Rengarajan, "Mutual coupling between waveguide-fed longitudinal broad wall slots radiating between baffles," *Electromagnetics*, vol. 16, no. 6, pp. 671–683, Nov.–Dec. 1996.
- [11] K. Forooraghi and P.-S. Kildal, "Admittance of an isolated waveguide-fed slot radiating between baffles using a spectrum of two-dimensional solutions," *IEEE Trans. Antennas Propag.*, vol. 41, no. 4, pp. 422–428, Apr. 1993

- [12] Y. Miura, J. Hirokawa, M. Ando, Y. Shibuya, and G. Yoshida, "Double-layer full-corporate-feed hollow-waveguide slot array antenna in the 60-GHz band," IEEE Trans. Antennas and Propag., vol. 59, no.8, pp.2844-2851, Aug. 2011.
- [13] D. Kim, J. Hirokawa, M. Ando, and T. Nagatsuma, "Design and fabrication of a corporate-feed plate-laminated waveguide slot array antenna for 120GHz-band", Proc. of 2011 IEEE AP-S URSI Intl. Symp., 429-6, Jul.2010
- [14] C. A. Balanis, Advanced engineering electromagnetics. John Wiley & Sons, Inc., 1989.
- [15] Ansys, Inc. Available: <http://ansys.com/>
- [16] R.W. Haas, D. Brest, H. Mueggenburg, L. Lang and D. Heimlich, "Fabrication and performance of MMW and SMMW platelet horn arrays," Intl. J. of Infrared and Millimeter Waves, Vol. 14, Num. 11, pp. 2289-2294, Jan.1993
- [17] A. Hirata, R. Yamaguchi, T. Kisugi, H. Takahashi, K. Murata, T. Nagatsuma, N. Kukutsu, Y. Kado, N. Iai, S. Okabe, S. Kimura, H. Ikegawa, H. Nishikawa, T. Nakayama and T. Inada, "10-Gbit/s wireless link using InP HEMT MMICs for generating 120-GHz-band millimeter-wave signal," IEEE Trans. Microwave Theory Tech., vol. 57, no. 5, pp. 1102-1109, May 2009.

Chapter 5

Antenna Design for Dual - polarization Operation in 60 GHz band

5-1 Introductory Remarks

Dual polarization technique is the easiest way to increase the frequency reuse efficiency in twice because it can generate two different polarizations in same time. The additional polarization can also be used as a diversity measure to improve the received SNR and/or increase the data rate in communication applications [1]-[2]. So that reason, the dual-polarization antennas are widely used for in both of communication systems and radar applications [2].

Most of the studies on the dual-polarized array antennas use microstrip substrates because of the easy realization of the radiating elements and the feeding networks. Many kinds of the multi-layer feeding and radiating structures to achieve high isolation and good cross-polarization have been demonstrated [3]-[7]. But they suffer from the large loss; they are not suitable for the high gain because the loss increases with the array size [8], especially in the higher frequency regions.

On the other hand, the waveguide slot array antenna, which has advantages of low profile, low cross-polarization level and very low transmission loss, is widely employed in the millimeter wave band. However, the conventional waveguide slot array

antennas are series-fed and have a fundamental problem that the bandwidth is narrowed due to the long-line effect when the array size becomes large [9]-[10]. Also, it is also characterized as less design flexibility in dual-polarization operation and beam scanning, etc. due to its complex 3-dimensional structure which is not suitable for the mass productivity. Actually, the dual-polarized waveguide slot antennas are almost absent from literature. In most previous works [11]-[12], the sub-arrays with orthogonal linear polarizations are interlaced side by side. The interlaced structure of two different polarizations increases the antenna aperture size and decreases the aperture efficiency.

The double-layer corporate-feed waveguide slot array antenna [13], where the feeding circuit is located in the bottom layer underneath the radiating waveguides in the top layer has been developed for wideband operation and to avoid the beam squint. Also, the diffusion bonding of laminated thin metal plates provides a high degree of accuracy and mass productivity with low cost. In addition, we can easily realize multi-layer structure as introduced in previous chapter 1 to 4. For the single polarization, a 16x16 element array is designed and fabricated in the 60 GHz band. A high gain of 33 dBi as well as a wide bandwidth of 8 % for the high antenna efficiency of more than 80 % is achieved [13].

Based on the antenna structure in [13], we propose a dual-polarization waveguide slot array antenna. The feeding structure having a multiple-layer and the common radiating elements for the dual-polarization will be introduced in this chapter.

5-2 Antenna Configuration

Fig. 1 shows the antenna configuration for the dual-polarization operation in 60 GHz band. There are 16x16 radiating slots with 0.80 wavelength interval on the top layer fed by two full-cooperate feeding networks on the bottom layers. The aperture area is defined as 62.4 mm square (16 elements x 3.9 mm spacing). The total height of the antenna is 7.5 mm using 25 plates with 0.3mm thickness. Also, this antenna structure is composed of the 2x2-element subarrays as marked in the top of Fig. 1.

The element spacing is 0.80 wavelength in the final design, but it was 0.86 wavelength equal to the single-polarization antenna in [13] in the initial design. Also the antenna height is 7.5 mm in the final design, but it was 6.6 mm in the initial design. These changes of the antenna size will be explained in the following chapters.

5-3 2x2-element Subarray Design

A 2x2-element subarray is designed as a unit to be fed by the full-corporate-feed structure. Fig. 2 shows the configurations of the 2x2-element subarray. To analyze the mutual coupling in an infinite two-dimensional array of the radiating slots, two sets of periodic boundary walls are adopted in the external region in HFSS simulation [14]. Then we can predict the antenna performance such as reflection and radiation for the infinite array. In this chapter, we discuss the design of the 2x2-element subarray.

The arrows with dash-line in Fig.2 indicate the magnetic field directions. Here, we explain overall field behaviors in the view of the magnetic field direction to prevent read's confusing. Thus the fields flowing in x-direction and y-direction are marked as 'longitudinal magnetic field' and 'transverse magnetic field'.

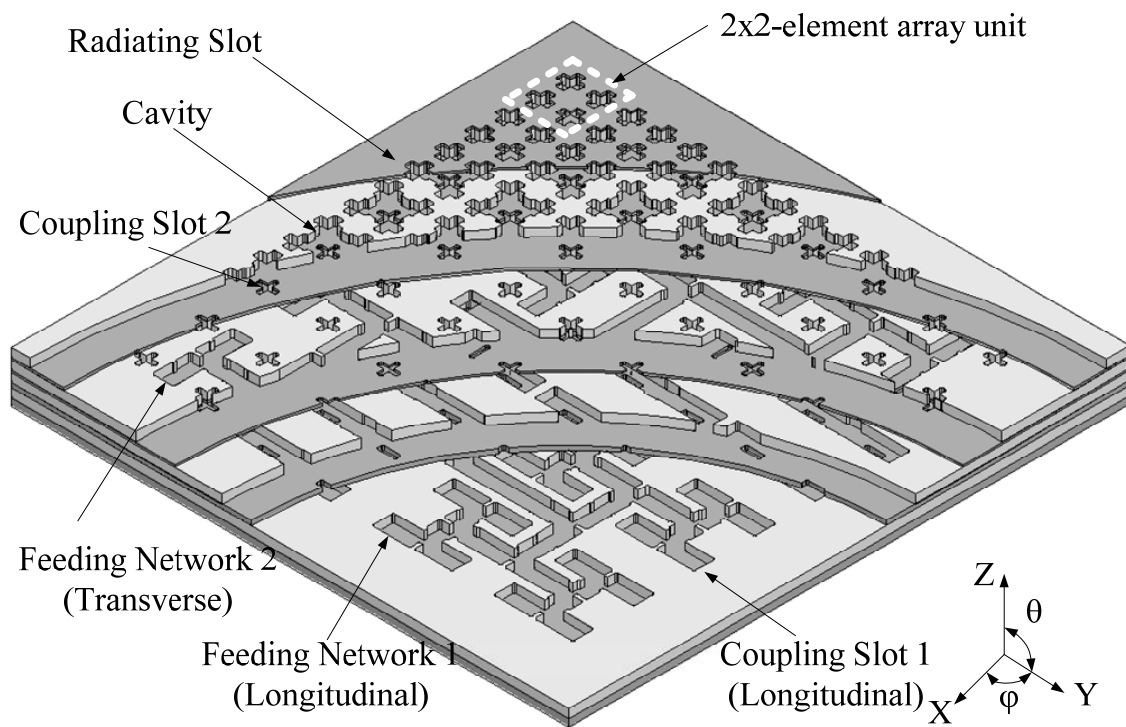


Fig. 1 Antenna configuration for the dual-polarization operation.

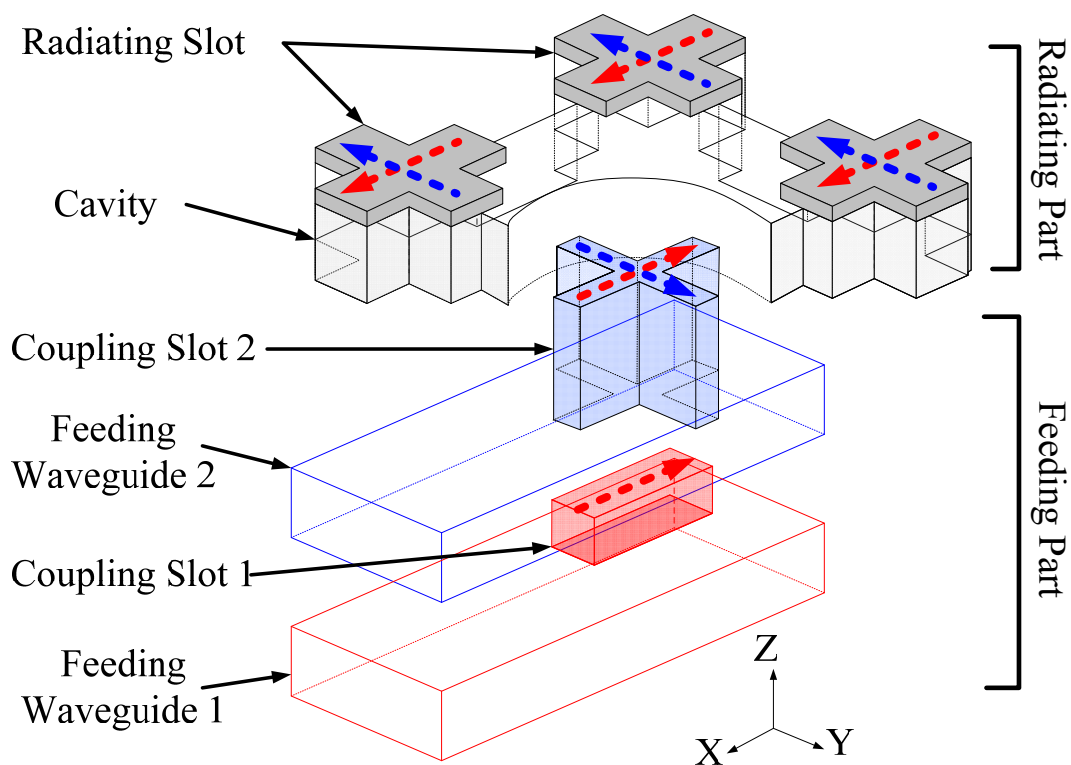


Fig. 2 Exploded perspective view of the 2x2-element subarray to design the large-sized array antenna.

5-3-1 Radiating Part Design for 2x2-element Subarray

Fig. 3 (a) shows the coupling mechanism of the single-polarization antenna proposed in [13]. The details of this structure are described in chapter 2. The arrows lines show the magnetic field distribution inside the waveguide cavity. The upper four radiating slots are excited in same single linear polarization through the coupling slot located underneath the cavity. With this structure the polarization in longitudinal direction is obtained. With the rotation of the structure in Fig. 3(a), the polarization in transverse-direction is also easily obtained as shown in Fig. 3(b). Now, we combine these two structures in Fig. 3(a) and (b) for the dual-polarization operation.

Fig. 3(c) shows the new structure of the radiating part including radiating slots and cavity for the dual-polarization operation. To realize a dual-polarization operation in the radiating unit, all radiating slots are changed to cross-shaped slots and the cavity structure is changed to symmetrical cavity in x- and y-direction. The coupling slot under the cavity is also changed to the cross-shaped slot to excite two different polarizations in same time. The coupling mechanism of the dual-polarization operation is same in principle with that of the single-polarization.

By the way, the unwanted cross-polarization is generated by the field rotation in the structure in Fig. 3 (c). Fig. 3 (d) shows the magnetic field distribution when only one polarization is excited. As shows in Fig. 3 (d), the cross-polarization is generated in the edges near the center of the cavity. To suppress the cross-polarization generated in cavity, we cut the edge parts near the center of the cavity. Finally, the cavity structure is decided for the dual-polarization operation as shows in Fig. 4. Also, the coupling mechanism of this structure is same in principle with that of the single-polarization. Two orthogonal linear polarizations will be realized, if the magnetic fields in both longitudinal and transverse directions can be excited independently with a high XPD in radiation and a high isolation in feed.

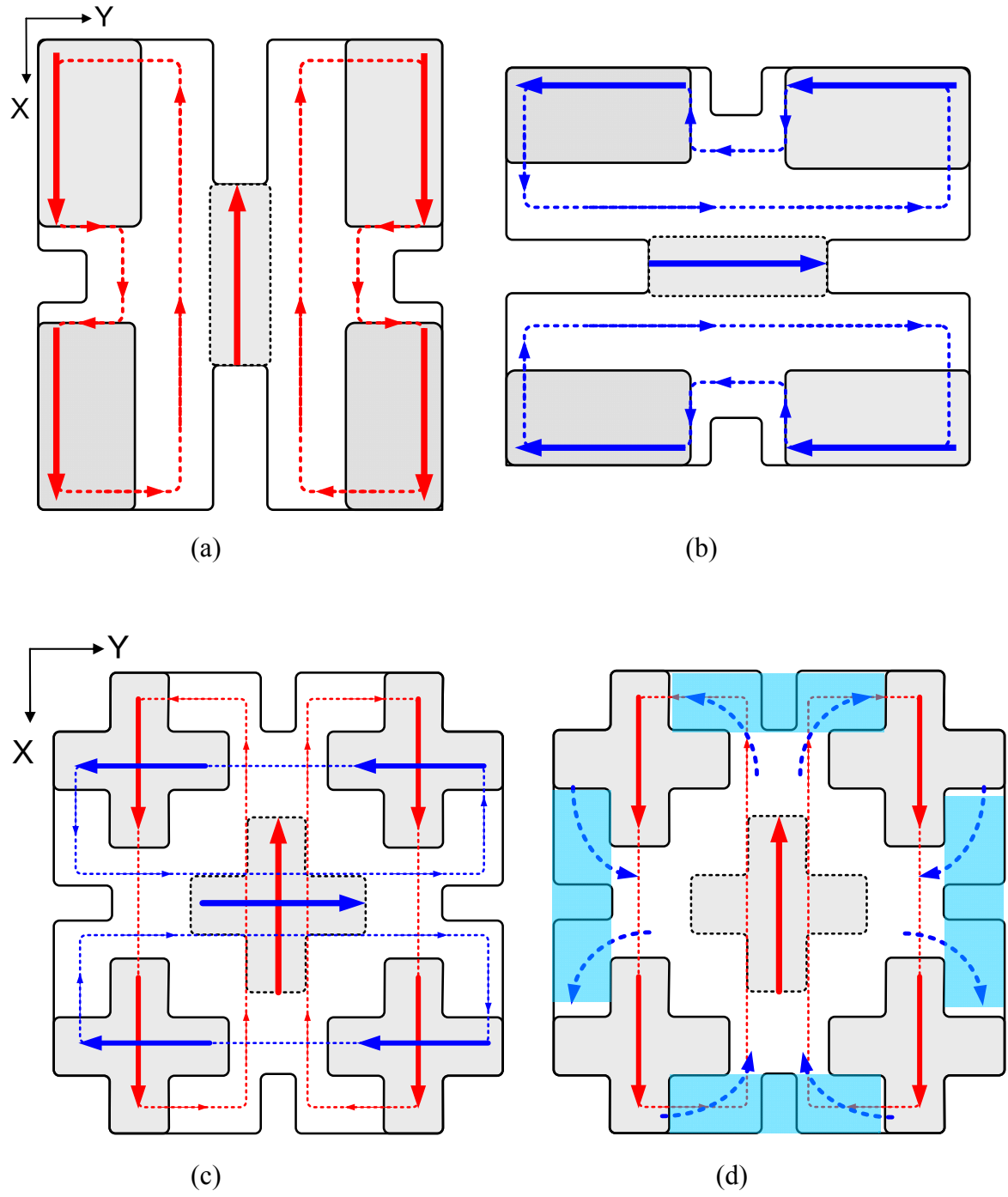


Fig. 3 Coupling mechanism of the radiating part for the dual-polarization in the view of the magnetic field.

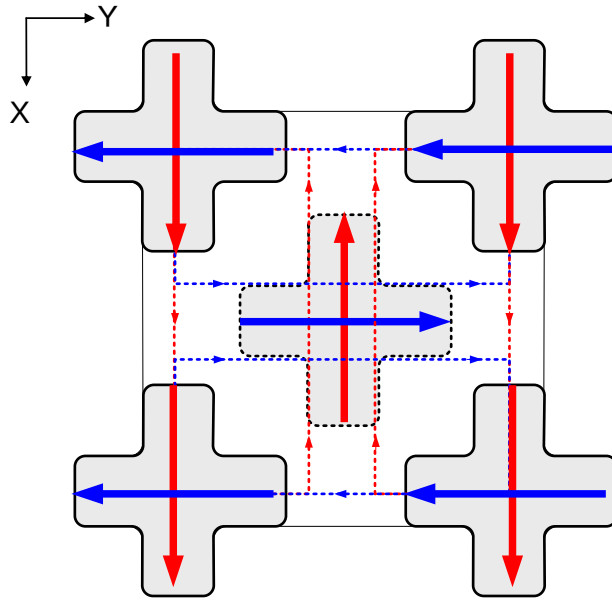


Fig. 4 The structure and the coupling mechanism of the radiating part for the dual-polarization operation

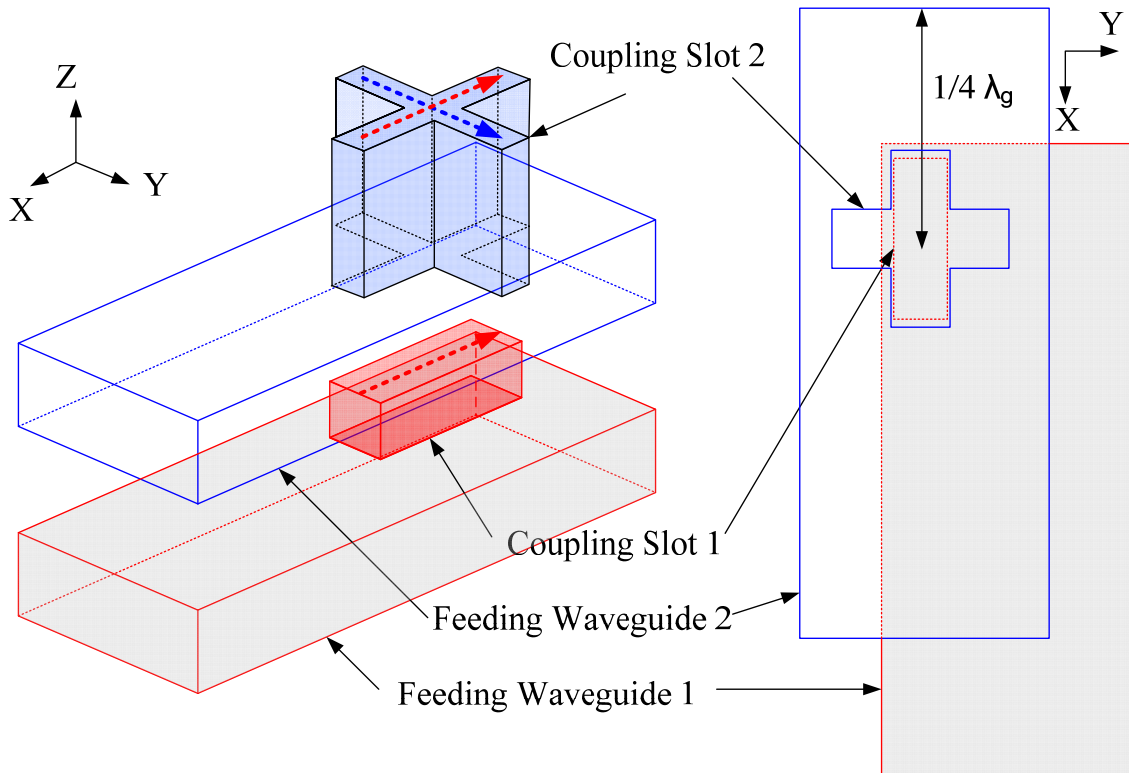


Fig. 5 Feeding structure for the 2x2-element subarray

5-3-2 Feeding Part Design for 2x2-element Subarray

Now, we discuss the design of the feeding structure for the 2x2-element subarray. Fig. 5 shows the feeding part of the 2x2-element subarray. Note that, the radiating part including radiating slot and cavity are not included in here. The longitudinal magnetic field is excited by the feeding waveguide 1 which is located in the bottom of the 2x2-element subarray. Also the longitudinal magnetic field goes to coupling slot 2 through the coupling slot 1 and feeding waveguide 2.

On the other hand, the transverse magnetic field is excited by the feeding waveguide 2. Two waveguides are connected by coupling slot 1. The realization of a high isolation as well as a high XPD is very promising, because the coupling slot 1 locates at the waveguide axis of the feeding waveguide 2 while it locates on the feeding waveguide 1 with offset for the strong coupling with the magnetic field excited by feeding waveguide 1 as illustrated in Fig. 5. Also, the large thickness of the coupling slot 1 contributes on the high isolation between different two polarizations. The higher-order mode excited in the coupling slot 1 attenuates sufficiently by the large thickness of coupling slot 1, and the excitation of the transverse slot as well as that in the feeding waveguide 2 can also be suppressed effectively.

5-3-3 Design Results of the 2x2-element Subarray

Based on the structure in chapter 5-3-1 and 5-3-2, the 2x2-element subarray is designed for the dual-polarization operation. The configuration of the 2x2-element subarray is described in Fig. 2. Fig. 6 shows the calculated frequency characteristics of the reflection and the isolation of the 2x2-element subarray with the periodic boundary walls in external region. The bandwidths for the two polarizations are 5.0% (longitudinal) and 10.5% (transverse) for VSWR less than 1.5. The high isolation over 43 dB is achieved between the 2 ports over a broad frequency bandwidth. Fig. 7 shows the calculated directivity of the 2x2-element subarray with the periodic boundary walls in external region. The calculated maximum aperture efficiencies are 73.0% and 76.1% for the longitudinal and transverse polarization, respectively. Note that, the aperture efficiency of 2x2-element subarray of the single-polarization antenna in [13] is more than 90% in 60 GHz band. The reasons of the degradation in dual-polarization will be explained in chapter 5-4-2.

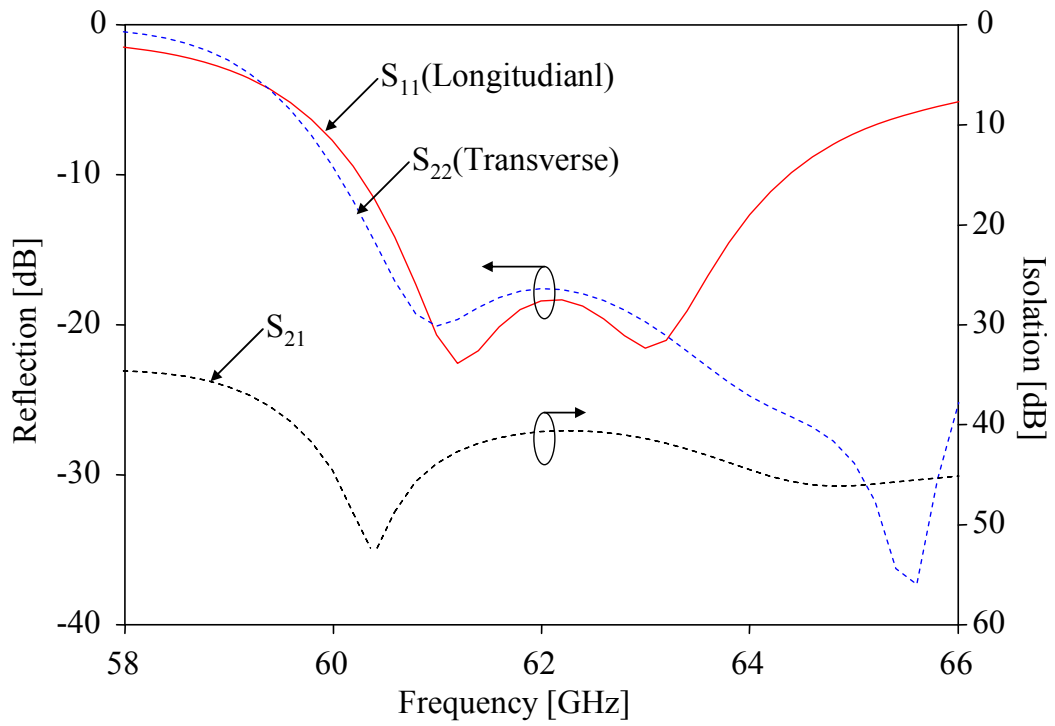


Fig. 6 Calculated frequency characteristics of the reflection and the isolation of the 2x2-element subarray with the periodic boundary walls

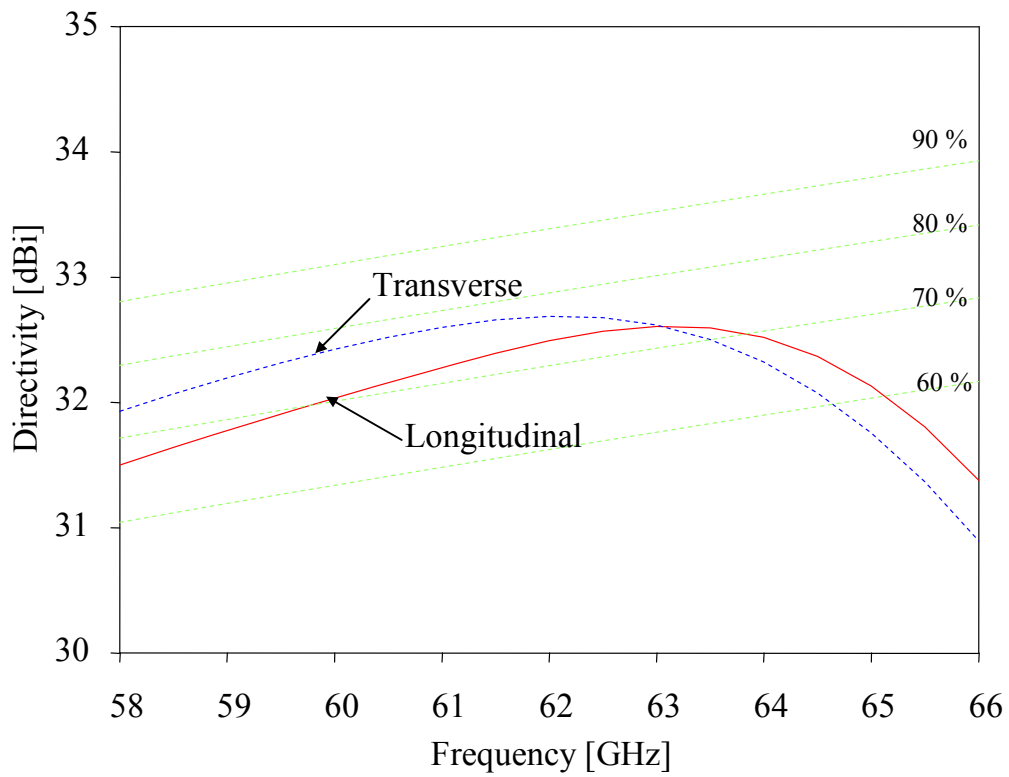


Fig. 7 Calculated frequency characteristics of the directivities of the 2x2-element subarray with the periodic boundary walls

5-4 16x16-element Design Results

5-4-1 Corporate Feeding Structure for 16x16-element Array for Dual-polarization Operation

The feeding structure for the 2x2-element subarray is introduced in chapter 5-3-2. Based on the structure in Fig. 5, the corporate feeding structure for 16x16-element is designed. Firstly, the feeding structure of 4x4-element will be introduced. Fig. 8 shows the configuration of the corporate feeding structure for 4x4-element array. (Note that, the feeding structure for 4x4-, 8x8- and 16x16-element arrays are basically same. In here, the feeding structure of 4x4-element array is shown for the simplified figure, firstly.)

The longitudinal magnetic field is to be excited by the corporate feeding circuit placed in the first layer, which is same with that of the double layer antenna for single-polarization [13]. On the other hand, the transverse magnetic field is to be excited by applying the transverse slot in the second layer. The transverse slots neighbored up and down would be excited in alternating phase, if the third layer was not introduced in Fig. 8 denoted by the dashed lines. The corporate feeding circuit in the third layer is almost identical with that in the first layer except the terminations. Therefore, the design of feeding circuit can also be simplified to some extent by adopting same structural parameters.

By the extension of the feeding structure introduced in Fig. 8, the corporate feeding structure for the 16x16-element array could be designed. Fig. 9 shows the corporate feeding structure for the 16x16-element array. Fig. 9 (a) shows the first layer and (b) shows the second layer and third layer. The third layer is marked in the dash-line in Fig. 9 (b).

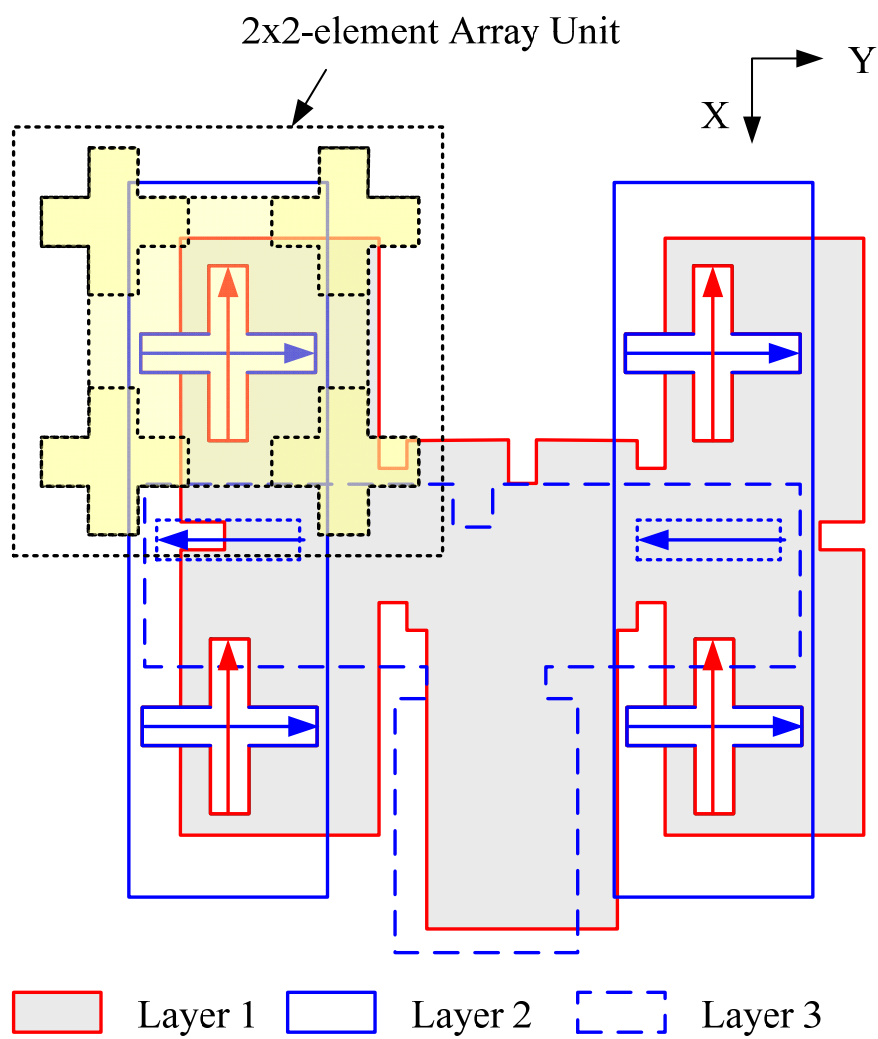
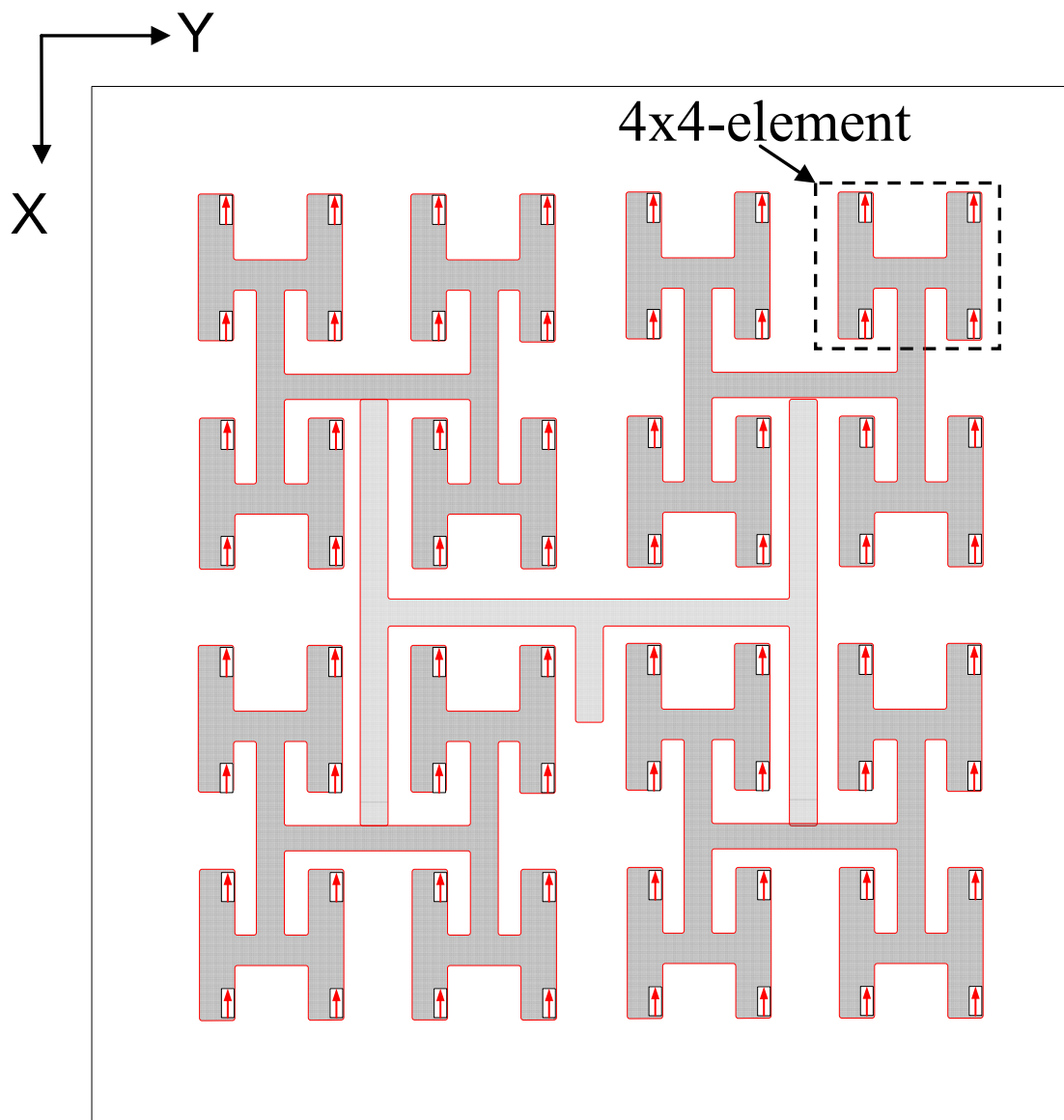
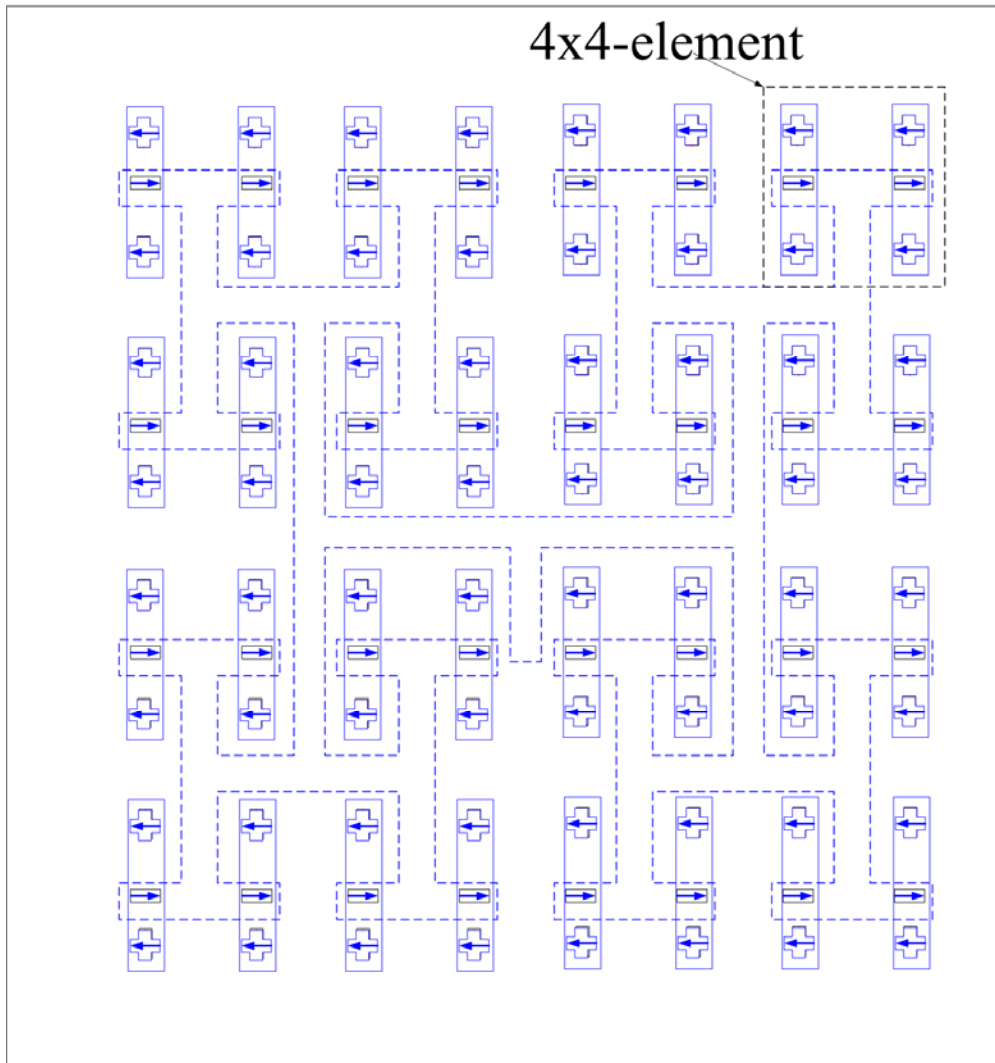


Fig. 8 The corporate feeding structure for the 4x4-element array



(a)



(b)

Fig. 9 The corporate feeding structure for the 16x16-element array

5-4-2 Design Results of the 16x16-element Array for

Dual-polarization Operation

The 16x16-element array for dual-polarization operation is designed. The antenna structure is introduced in Fig. 1 in chapter 5-1. As explained in chapter 5-1, the element spacing among the radiating element is 0.86 wavelength in this initial design. Thus the aperture size is defined as 67.2 mm square (16 elements x 4.2 mm spacing) and the total height of the antenna is 6.6 mm using 22 plates with 0.3mm thickness.

Fig.10 shows the calculated frequency characteristics of the antenna reflection of the 16x16-element array antenna. The bandwidths for the two polarizations are 5.0% (longitudinal) and 8.1% (transverse) for VSWR less than 1.5. The high isolation over 55 dB is achieved between two ports over a broad frequency bandwidth. The isolation is higher than that of the 2x2-element subarray by the feeding layer 3. Fig. 11 shows the frequency characteristic of the antenna gain and efficiency. The peaks of the gain are 32.5 dBi with 75% antenna efficiency at 61.5 GHz and 32.5 dBi with 72% antenna efficiency at 62.8 GHz for the longitudinal and transverse polarizations, respectively, where the conductivity of copper is assumed to be 5.8×10^7 S/m. The 1-dB down gain bandwidths are 7.2% and 8.8%, respectively. The calculated antenna gain and the antenna efficiency are quite lower than those of the single-polarization antenna in [13]. The antenna gain and the antenna efficiency of the single-polarization are more than 33 dBi and 80% in the 60 GHz band. We can find the reason of the antenna gain degradation in the radiation patterns of the dual-polarization antenna. In radiation patterns, we can obtain several unwanted large grating lobes.

Fig. 12 and Fig. 13 show radiation patterns of the x-pol. when the $\varphi=0^\circ$ (Fig. 12) and $\varphi=45^\circ$ (Fig. 13). There are two large grating lobes in Fig. 12 and their levels are -17 dB. Also, there are large grating lobes in Fig. 13 and their levels are -15 dB. The radiation patterns in $\varphi=180^\circ$ and $\varphi=135^\circ$ also show same phenomenon when $\varphi=0^\circ$ and $\varphi=45^\circ$, respectively. The grating lobes in $\varphi=0^\circ$ and $\varphi=180^\circ$ are caused by the co-polarization component and the grating lobes in $\varphi=45^\circ$ and $\varphi=135^\circ$ are caused by the cross-polarization. The radiation patterns of the y-pol. show the identical results with x-pol. To improve the antenna radiating performance, these grating lobes should be suppressed.

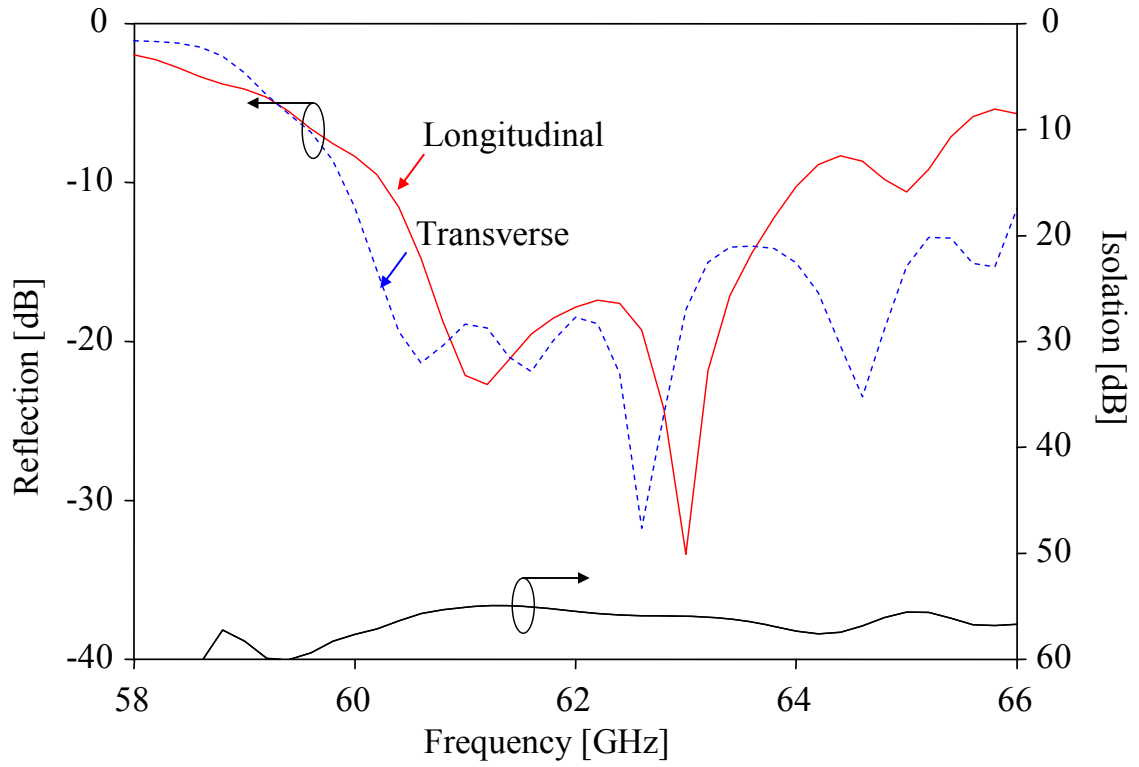


Fig. 10 Calculated frequency characteristics of the reflection and the isolation of the 16x16-element array

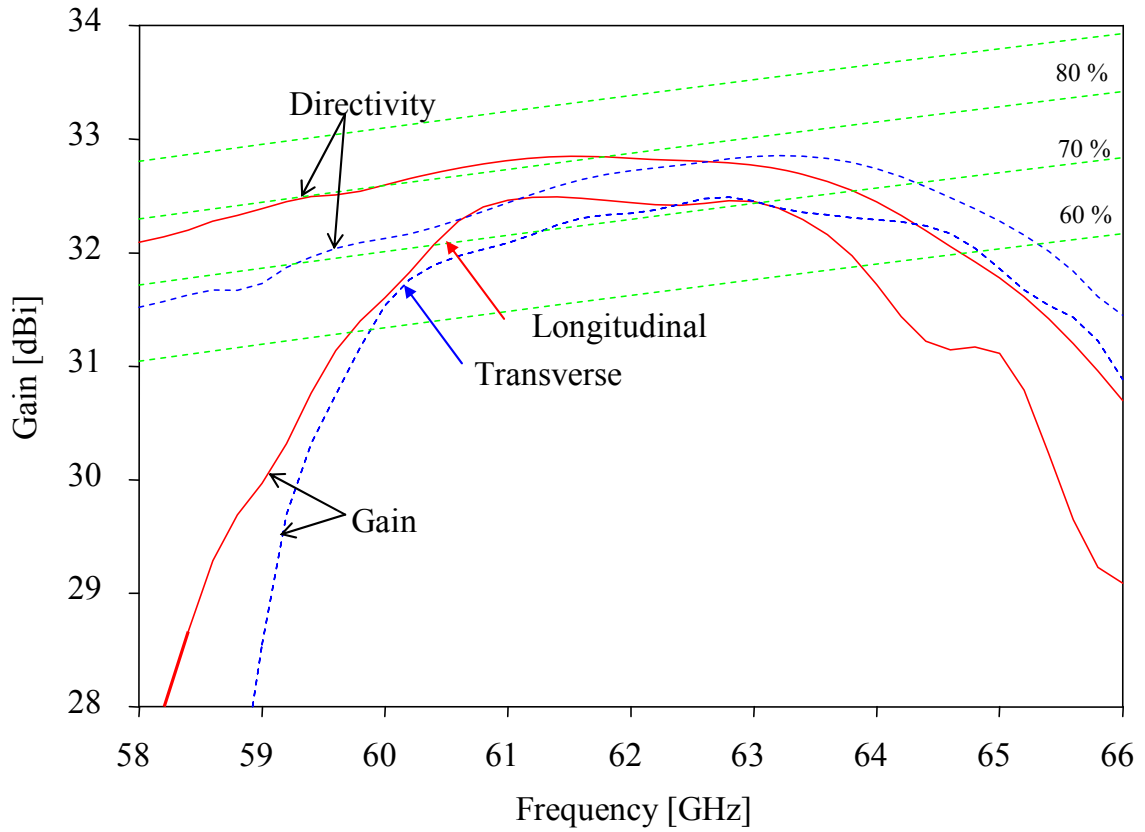


Fig. 11 Calculated frequency characteristics of the gain and the efficiency of the 16x16-element array

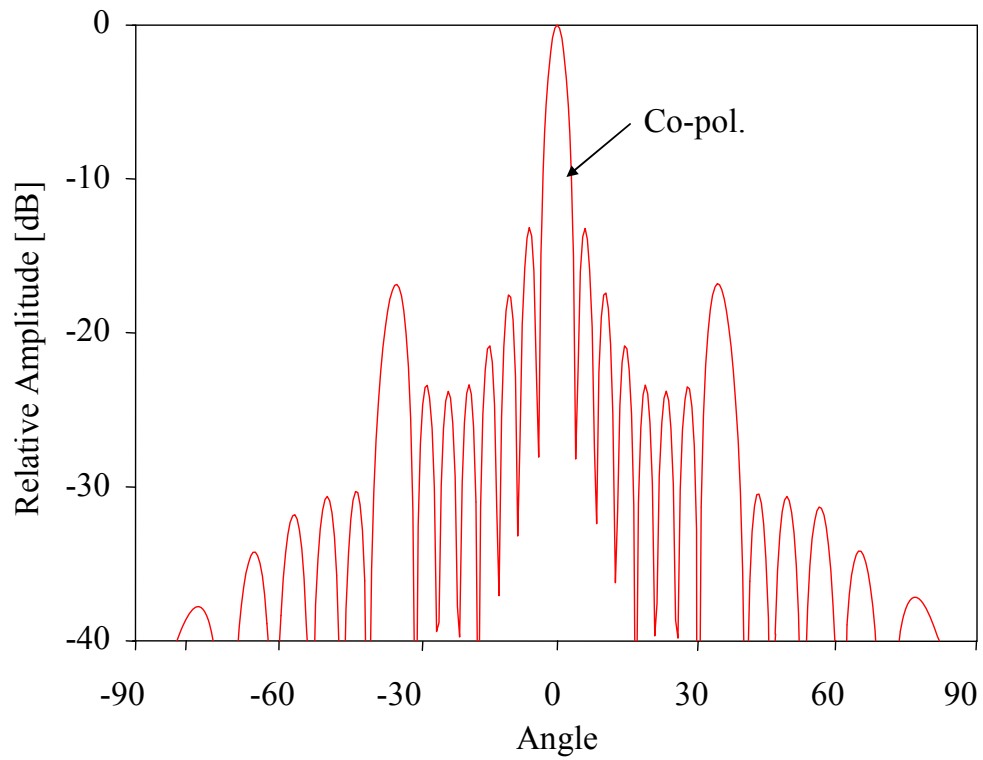


Fig. 12 Radiation pattern(x-pol, $\phi=0^\circ$ and 180°) of the 16x16-element array

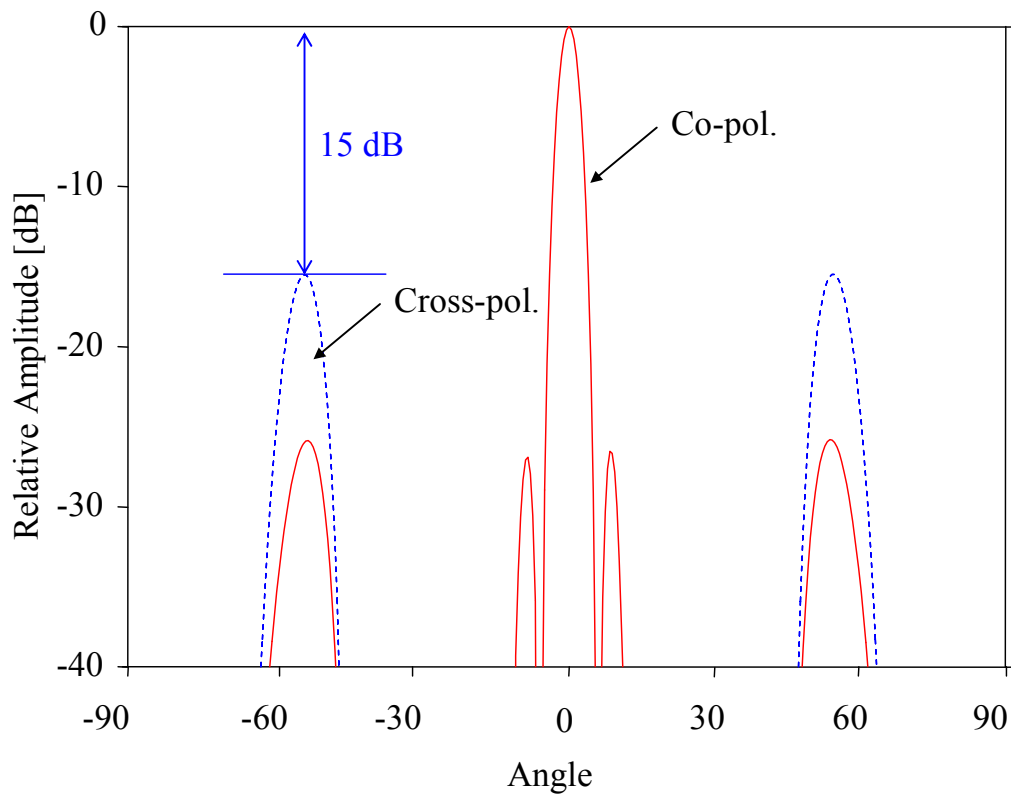


Fig. 13 Radiation pattern(x-pol, $\phi=45^\circ$ and 135°) of the 16x16-element array

5-5 Grating Lobe Suppression

5-5-1 Grating Lobe Suppression in $\varphi=0^\circ$ with Cavity Size

Reduction

The grating lobes in $\varphi=0^\circ$ and 180° are caused by the field unbalance in the cavity region. The unbalanced field distribution in the cavity results amplitude differences of the electric field between the neighbor radiating slots. The dash-line in Fig. 14 shows the relative E-field strength calculated in the center line of the radiating slots. Here, the E-field is very strong near the center between two radiating slots (in the center of the cavity) and it reduced in the edge regions in the cavity.

To realize a uniform excitation in neighbored radiating slots, the electromagnetic field in the cavity has to be symmetric. One way to achieve this is to propagate the dominant mode and cut off of the higher-order modes by the exciting slots [15]. Therefore, the cavity under the radiating slots has to be thick or narrow enough, as the attenuation E and its constant α of TE_{mn} modes through the exciting slots are expressed by (1) and (2)

$$E = E_0 \exp(-\alpha t_c) \quad (1)$$

$$\alpha = \sqrt{\left(\frac{m\pi}{l_c}\right)^2 + \left(\frac{n\pi}{w_c}\right)^2 - k^2} \quad (2)$$

where, t_c , l_c and w_c are the thickness, length, and width of the cavity, respectively. The k is the wave number in free space. Note that, the cavity width and length are actually same because the cavity structure is symmetric in the x- and y-directions.

Fig. 15 shows the grating lobe levels in $\varphi=0^\circ$ and 180° depends on the cavity dimensions. As the exciting slots become thicker or narrower, the grating lobe levels are reduced. Note that, the minimum cavity width and length is 3.8 mm in this design due to the coupling slot under the cavity. Also, the decreasing of the grating lobe level convergences in the cavity thickness of 2.1 mm, thus the cavity dimensions are decided as the cavity thickness and width(= length) are 2.1 mm and 3.8 mm, respectively.

Also, the calculated relative E-field strength in the new cavity is added in Fig. 14, it shows quite better balance compare with that of the initial design.

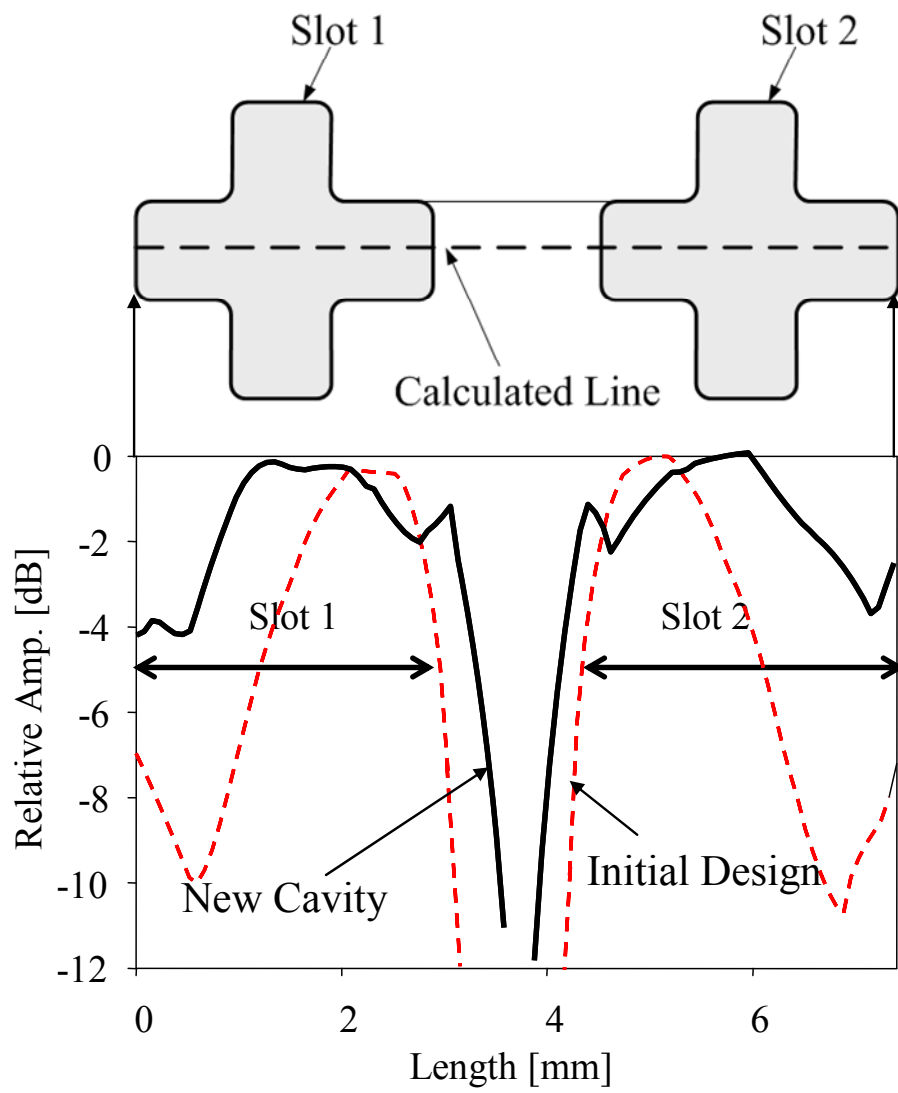


Fig. 14 Relative E-field strength calculated in the center line of the radiating slots.

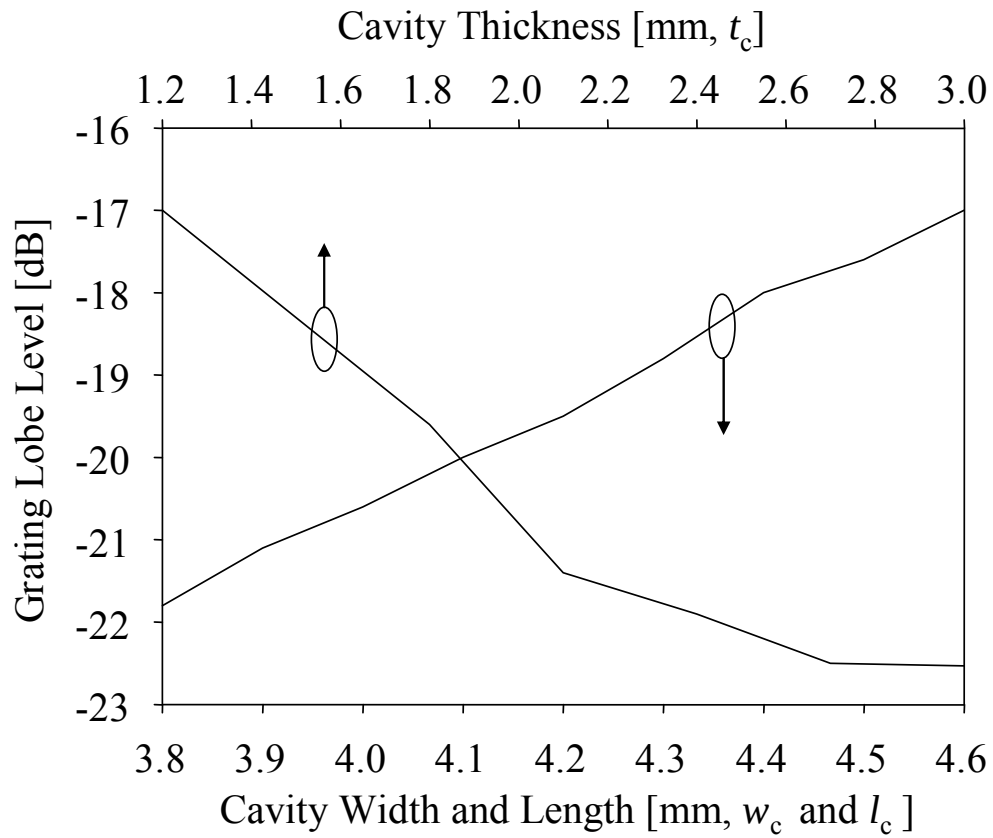


Fig. 15 Grating lobe levels in $\varphi=0^\circ$ and 180° depends on the cavity dimensions

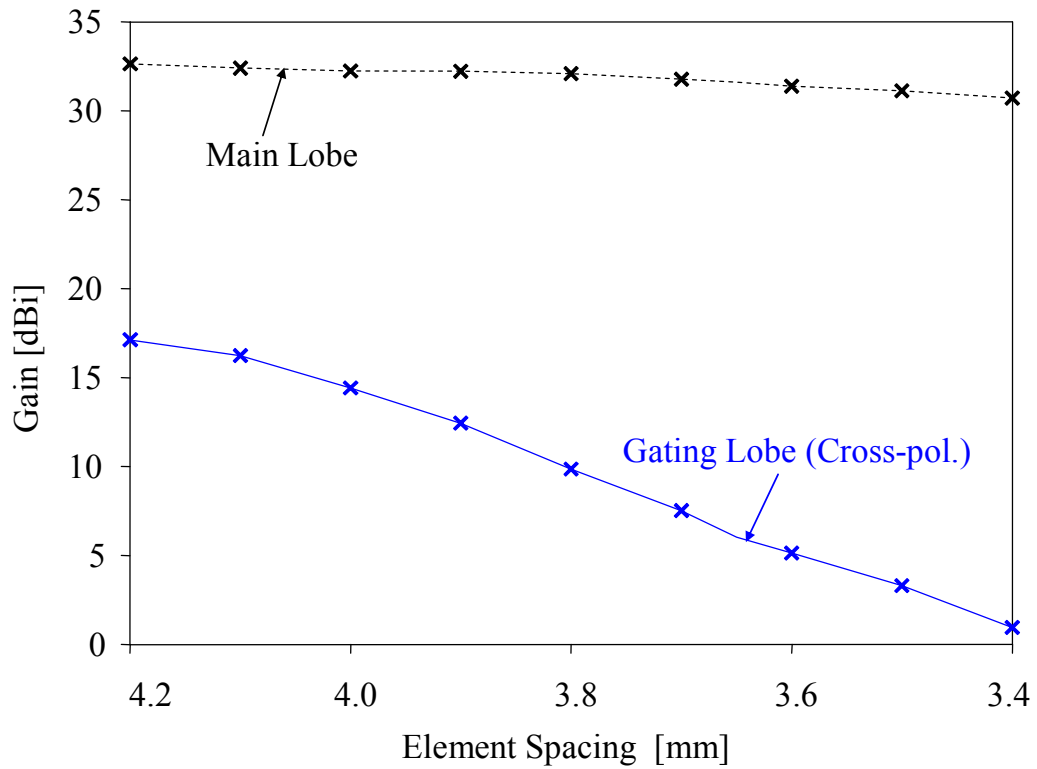


Fig. 16 Main lobe and grating lobe levels in $\varphi=45^\circ$ and 135° depends on the element spacing.

5-5-2 Grating Lobe Suppression in $\phi=45^\circ$ with Element

Spacing Reduction

The grating lobes in $\phi=45^\circ$ and 135° are caused by the large element spacing and the cross-polarization components. In the previous design for the single-polarization, the element spacing of 0.86 wavelength not cause the grating lobes in $\phi=45^\circ$ and 135° , because it shows very low cross-polarization levels. However, the antenna structure for the dual-polarization should have relatively large cross-polarization levels because two different polarizations share the one common radiating slot as well as the common cavity.

To suppress the grating lobes in $\phi=45^\circ$ and 135° , we observe the change of the grating lobe levels according to the element spacing. Fig. 16 shows the grating lobe levels in $\phi=45^\circ$ and 135° depends on the element spacing. Here, the levels of the lobe are also marked in Fig. 16 since the main lobe levels also changed depend on the element spacing. As shown in here, the grating lobe levels in $\phi=45^\circ$ and 135° remarkably decrease in the small element spacing. Considering the physical issues such as feeding structure, 0.39 mm ($=0.80$ wavelength) is chosen for the new element spacing to suppress the grating lobes.

5-6 16x16-element Design Results with Low Grating

Lobes

Base on the results discussed in chapter 5-4, the new antenna dimensions are determined. The element spacing is changed from 4.2 mm (0.86 wavelength) to 0.39 mm (0.80 wavelength) and the thickness of the cavity is changed from 1.2mm to 2.1 mm to suppress the unwanted grating lobes. With these conditions, the antenna structure is totally redesigned. Now, the aperture size is defined as 62.4 mm square (16 elements x 3.9 mm spacing) and the total height of the antenna is 7.5 mm using 25 plates with 0.3mm thickness.

Fig.17 shows the calculated frequency characteristics of the antenna reflection of the 16x16-element array antenna. The bandwidths for the two polarizations are 5.0% (Longitudinal) and 6.8% (Transverse) for VSWR less than 1.5. The bandwidths are slightly changed due to the change of the antenna parameters.

The high isolation over 50 dB is achieved between two ports over a broad frequency bandwidth. Fig. 18 shows the frequency characteristic of the antenna gain and efficiency. The peaks of the gain are 32.5 dBi with 82.0% antenna efficiency at 63.3 GHz and 32.5 dBi with 82.0% antenna efficiency at 63.3 GHz for the longitudinal and transverse polarizations, respectively, where the conductivity of copper is assumed to be 5.8×10^7 S/m. The 1-dB down gain bandwidths are 6.5% and 7.7%, respectively.

Because the aperture sizes are different between two antennas which have 0.86 element spacing and 0.80 element spacing, we can not compare their gains, directly. Instead, their efficiencies are compared in Fig. 19. The antenna efficiencies satisfy more than 80% for the both of the polarization in the new design with the small/thick cavity and small element spacing while the antenna efficiencies show only around 65% in the initial design. This is mainly caused by the different grating lobe levels in the radiation patterns. Fig. 20 and 21 show the radiation patterns of the x-pol. when the $\varphi=0^\circ$ (Fig. 19) and $\varphi=45^\circ$ (Fig. 20). The grating lobes are remarkably reduced in the both radiation patterns.

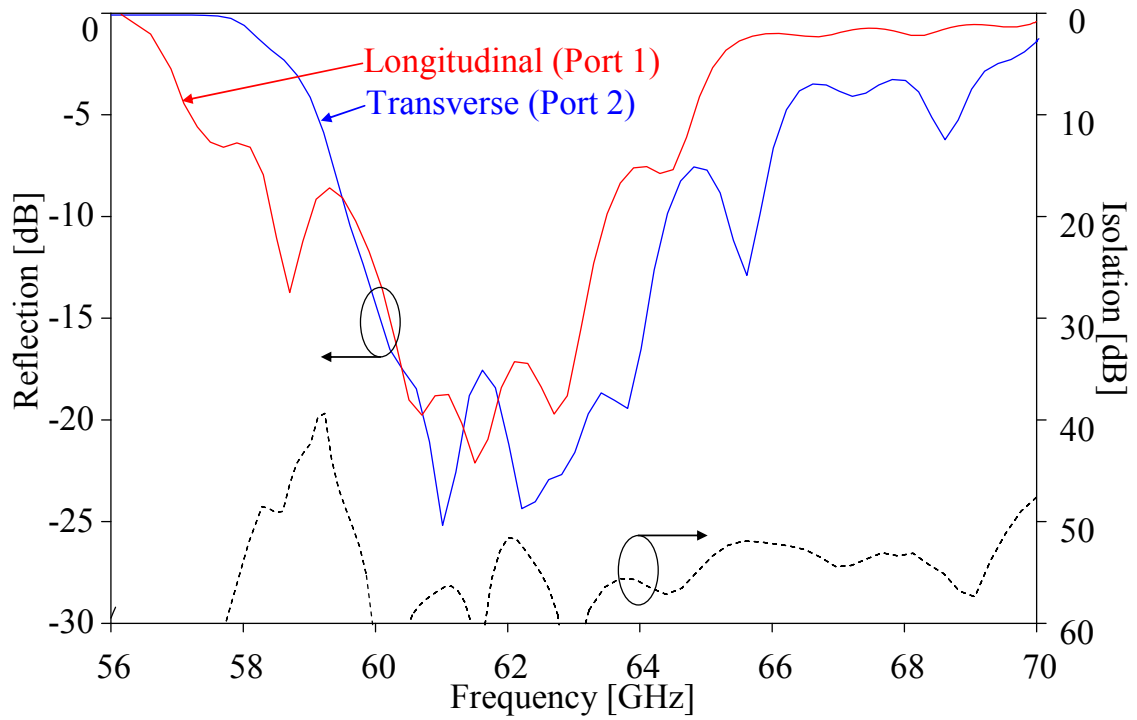


Fig. 17 Calculated frequency characteristics of the reflection and the isolation of the 16x16-element array.

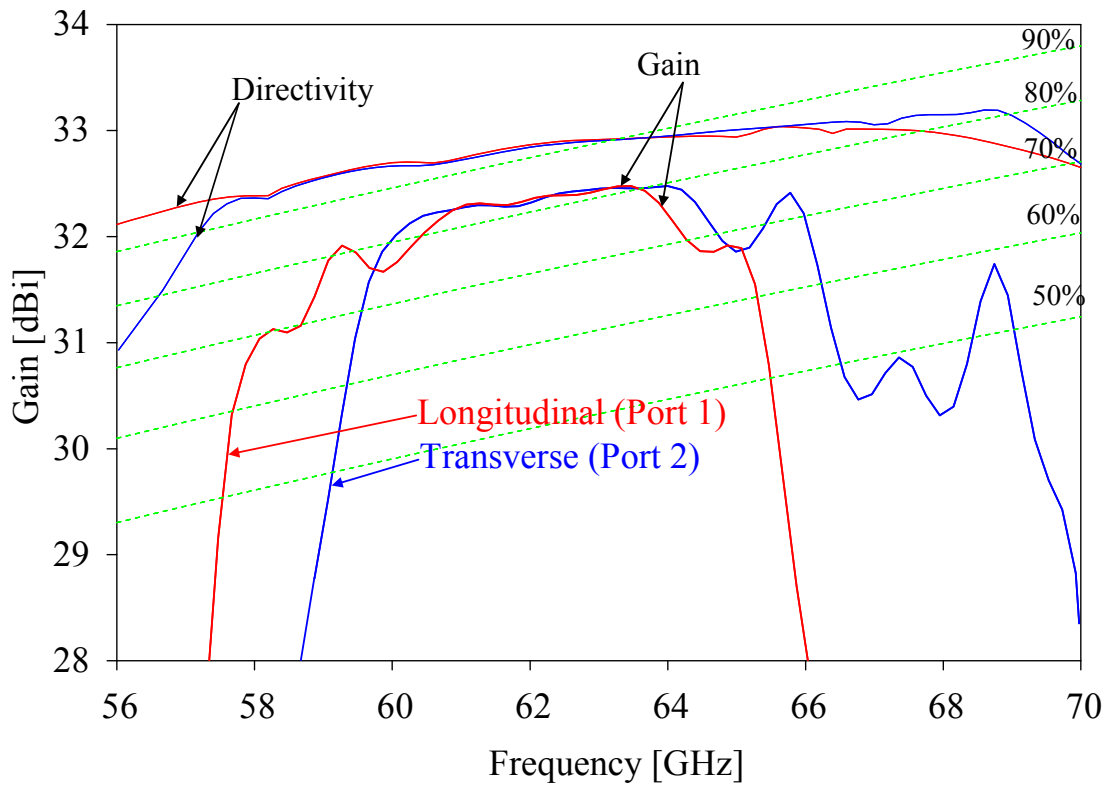


Fig. 18 Calculated frequency characteristics of the gain and the efficiency of the 16x16-element array.

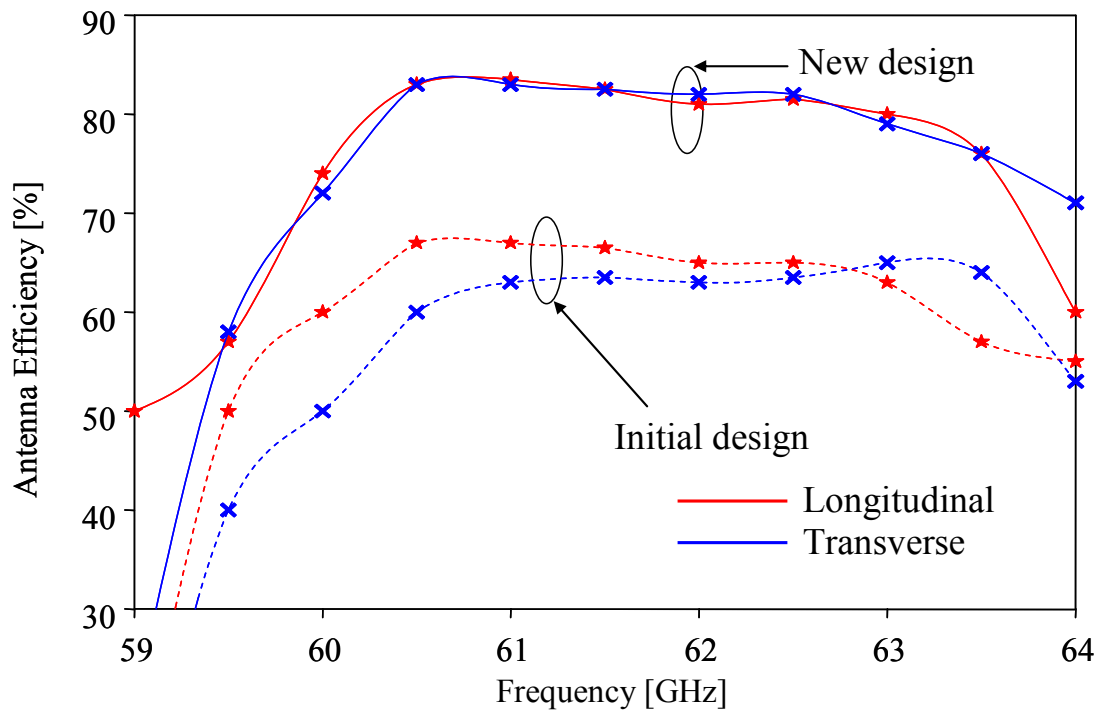


Fig. 19 Efficiencies of the initial design and the new design of the 16x16-element array.

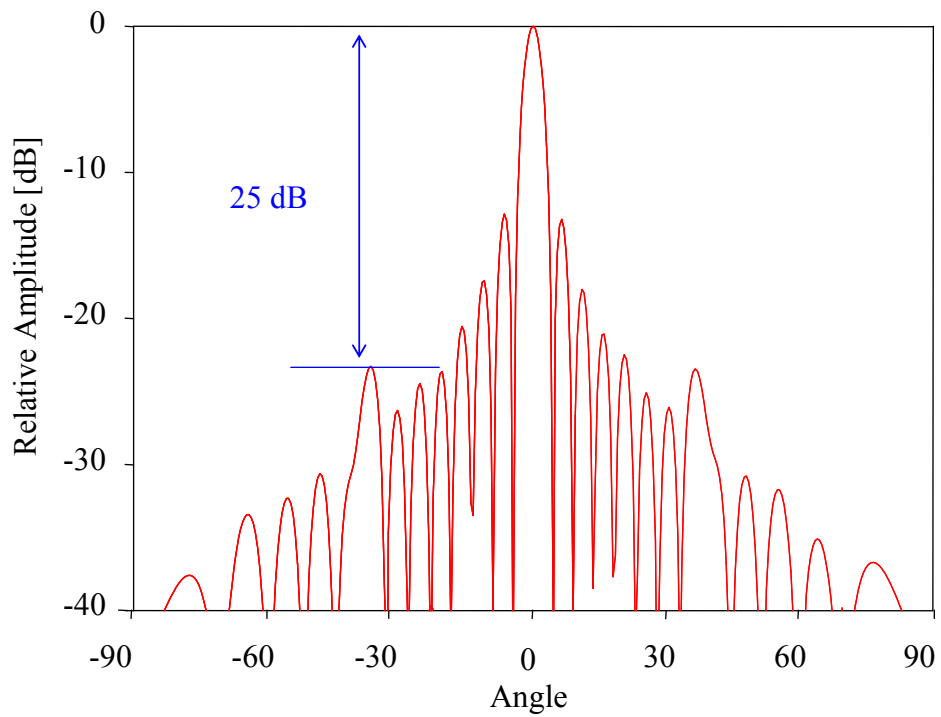


Fig. 20 Radiation pattern(x-pol, $\phi=0^\circ$ and 180°) of the 16x16-element array

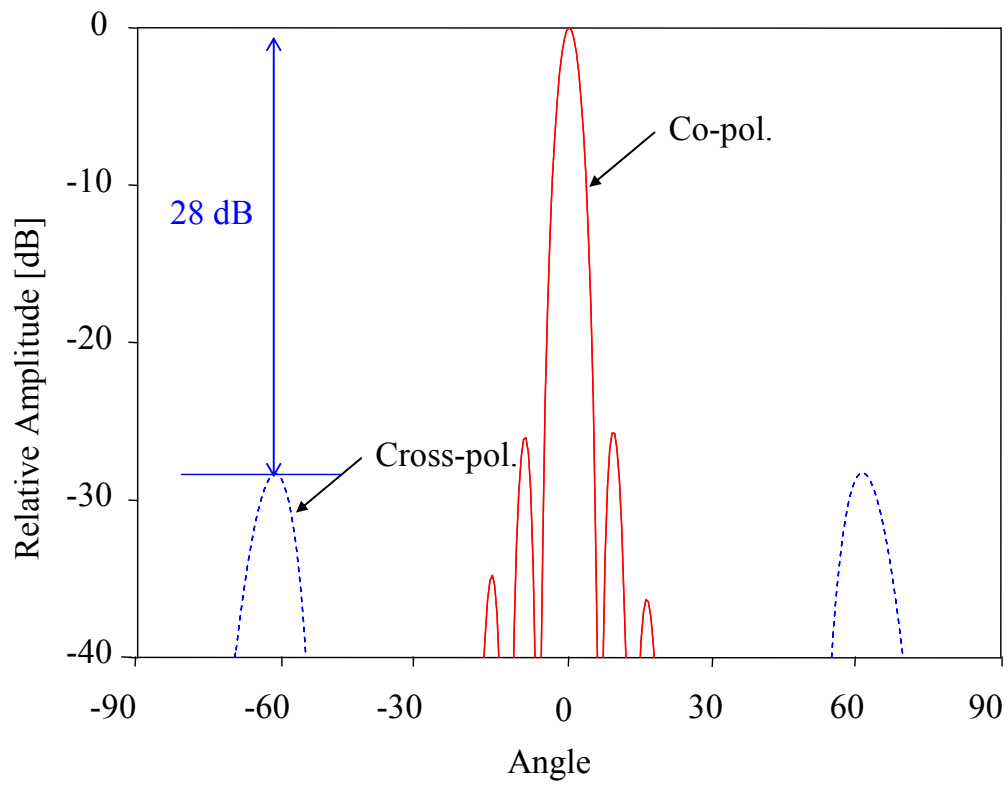


Fig. 21 Radiation pattern(x-pol, $\varphi=45^\circ$ and 135°) of the 16x16-element array

5-7 Fabrication and Measurement Results

Fig. 22 shows the fabricated antenna by diffusion bonding technique. The 16x16 radiating elements are arranged with 0.8-wavelength interval on the top layer. The aperture size is defined as 62.4 mm square (3.9mm x 16 elements) for the 60 GHz band. The antenna height is 7.5 mm including all layers.

Fig.23 shows the calculated and measured frequency characteristics of the reflection in the 16x16-element array antenna for dual-polarization. Even though some ripples are on the measurement results, they show good agreement between the calculation and measurement. The reflection bandwidths of each polarization are 5.0% and 9.2% for VSWR < 2.0. The gain and efficiency characteristics also show good agreement between the calculation and the measurement. Fig. 24 shows the calculated and measured antenna gains and efficiencies for each polarization. The peak gains of each polarization are 32.5 dBi with 82.0% antenna efficiency at 63.3 GHz, respectively. 1-dB down gain bandwidths are 10.4% and 10.9% for longitudinal and transverse polarization, respectively.

Fig. 25 shows the measured radiation patterns of each polarization. The radiation patterns are measured in ± 90 degrees. The side lobes are well suppressed under -25 dB for the both of co- and cross-polarization.

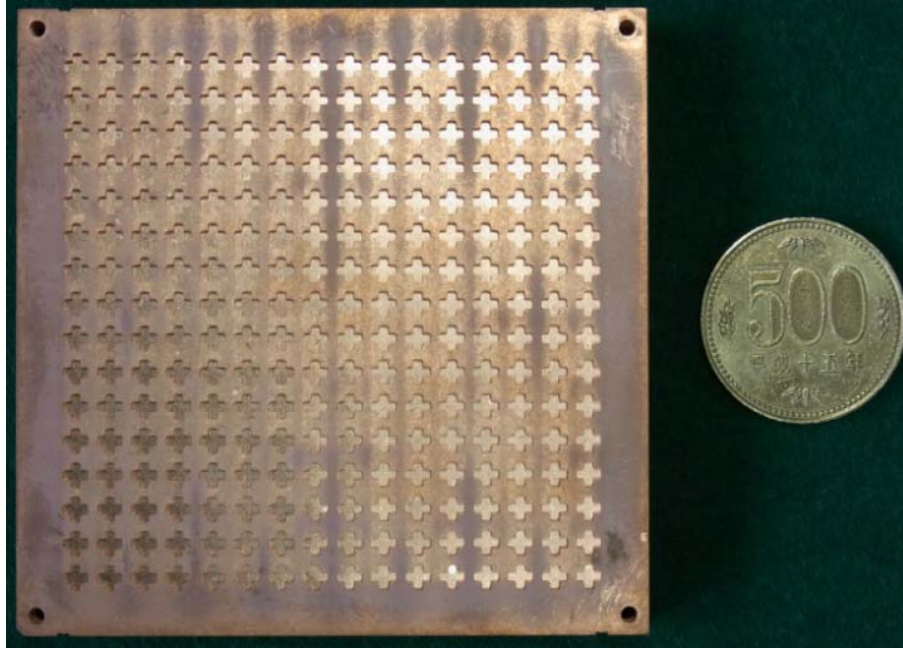


Fig. 22 Fabricated antenna for dual-polarization.

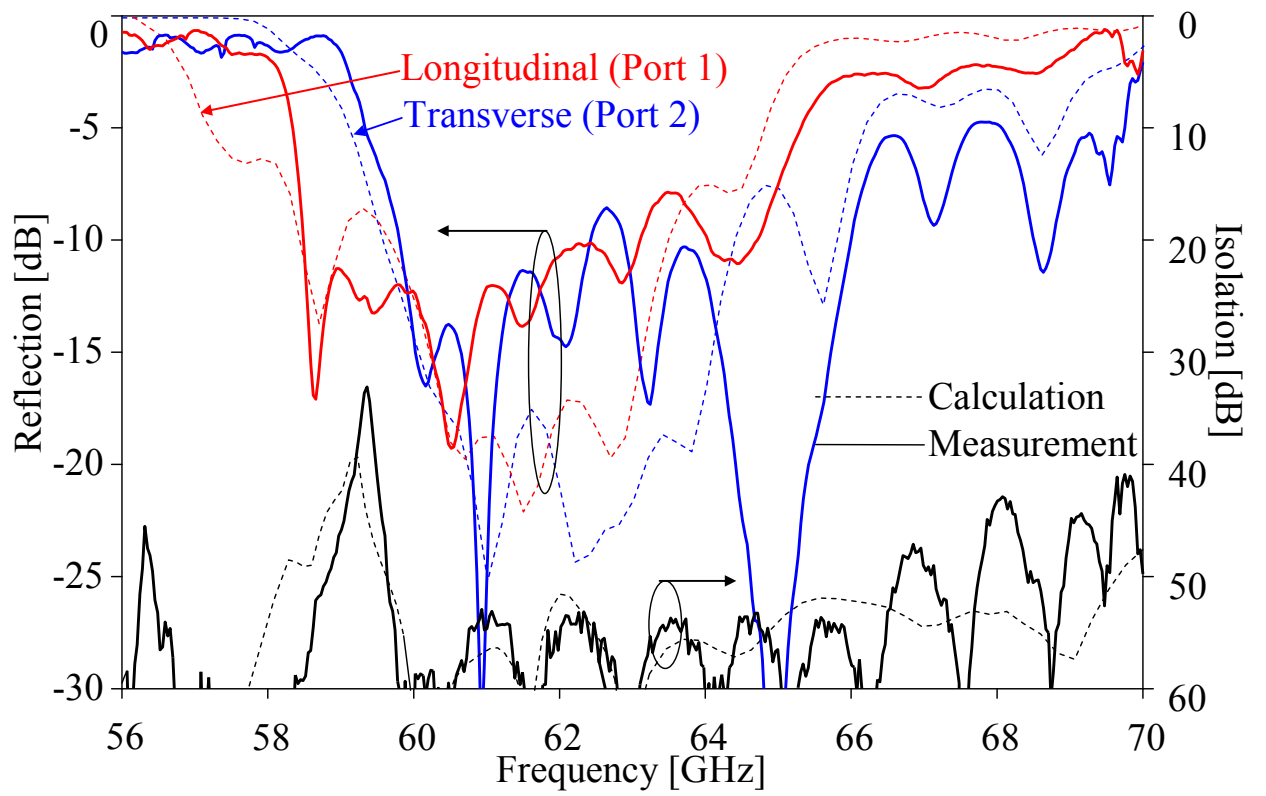


Fig. 23 Calculated and measured frequency characteristics of the reflection and the isolation of the 16x16-element array.

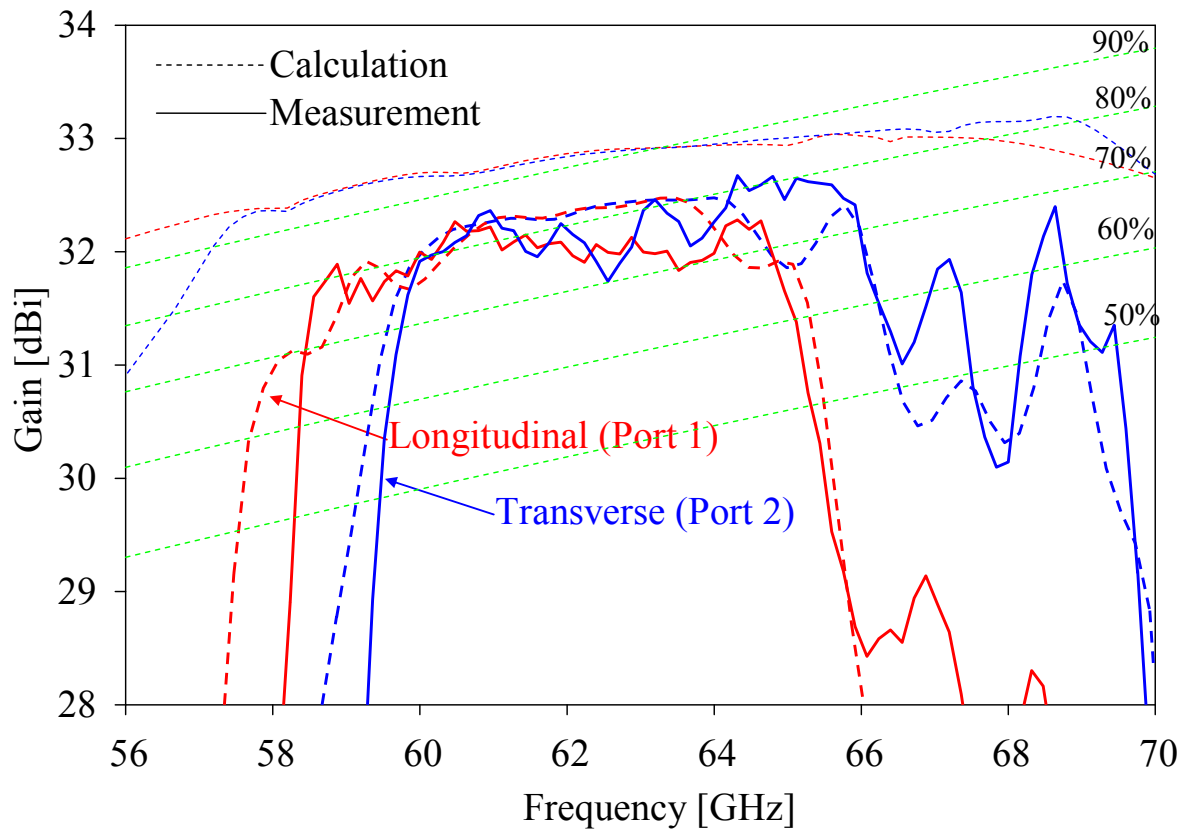


Fig. 24 Calculated and measured frequency characteristics of the gain and the antenna efficiency of the 16x16-element array.

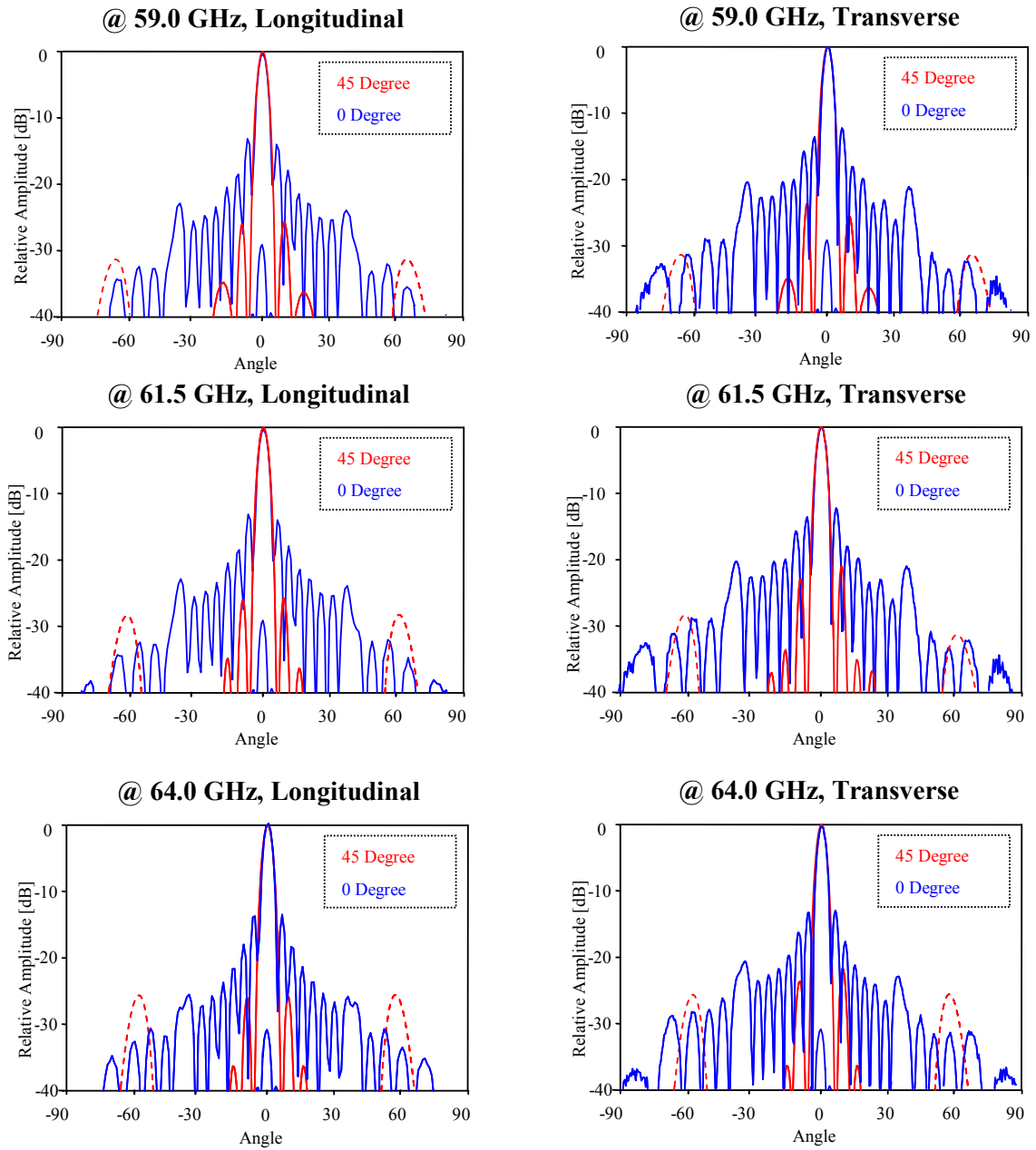


Fig. 24 Measured radiation patterns of each polarization at 3 frequencies of the 16x16-element array.

5-7 Concluding Remarks

A dual-polarization waveguide slot array antenna for 60 GHz band is proposed. In order to realize the dual-polarization operation, the cross-shaped radiating slots which are sharing the common aperture and the multi-layer feeding structure are employed. A 16x16-element array for the dual-polarization is fabricated by diffusion bonding of laminated thin copper plates which has the advantages of high precision and low loss characteristic. The antenna gain of 32.5 dBi and the antenna efficiency of 82.0 % are measured at center frequency for two different polarizations. The 1-dB-down gain bandwidths are no less than 10.4% and 10.9% and a wideband characteristic is achieved. High isolation over 50 dB is achieved in the broad bandwidth between two different polarizations.

References in Chapter 5

- [1] Elsherbini and K. Sarabandi, "Ultra-wideband dual polarized cavity backed antenna," Proc. of 2011 IEEE AP-S URSI Intl. Symp., pp 2043, July 2011
- [2] R. G. Vaughan, "Polarization diversity for mobile communications," IEEE Trans. Veh. Technol., vol. 39, no. 3, pp. 177–186, Aug. 1990.
- [3] S. C. Gao, L. W. Li, M. S. Leong, and T. S. Yeo, "Dual-polarized slotcoupled planar antenna with wide bandwidth," IEEE Trans. Antennas Propag., vol. 51, no. 3, pp. 441–448, Mar. 2003.
- [4] B. G. Porter, L. L. Rauth, J. R. Mura, and S. S. Gearhart, "Dual-polarized slot-coupled patch antennas on Duroid with Teflon lenses for 76.5 GHz automotive radar systems," IEEE Trans. Antennas Propag., vol. 47, no. 12, pp. 1839–1842, Dec. 1999.
- [5] S. D. Padhi, N. C. Kamakar Sr., C. L. Law, and S. Aditya Sr., "A dual polarized aperture coupled circular patch antenna using a C-shaped coupling slot," IEEE Trans. Antennas Propag., vol. 51, no. 12, pp. 3295–3298, Dec. 2003.
- [6] C.-H. Lee, S.-Y. Chen, and P. Hsu, "Isosceles triangular slot antenna for broadband dual polarization applications," IEEE Trans. Antennas Propag., vol. 57, no. 10, pp. 3347–3351, Oct. 2009.
- [7] E. A. Soloman, M. S. Ibrahim, and A. K. Abdelmageed, "Dual-polarized omnidirectional planar slot antenna for WLAN applications," IEEE Trans. Antennas Propag., vol. 53, no. 9, pp. 3093–3097, Sep. 2005.
- [8] E. Levine, G. Malamud, S. Shtrikman, and D. Treves, "A study of microstrip array antennas with the feed network," IEEE Trans. Antennas Propag., vol. 37, no. 4, pp. 426–434, Apr. 1989.
- [9] M. Ando, Y. Tsunemitsu, M. Zhang, J. Hirokawa, and S. Fujii, "Reduction of long line effects in single-layer slotted waveguide arrays with an embedded partially corporate feed," IEEE Trans. Antennas and Propag., vol. 58, no.7, pp.2275-2280, July, 2010.

- [10] S. Park, J. Hirokawa, and M. Ando, "A planar cross-junction power divider for the center feed in single-layer slotted waveguide arrays," *IEICE Trans. Commun.*, vol. E85-B, no. 11, pp. 2476–2481, Nov. 2002.
- [11] S. Park, Y. Okajima, J. Hirokawa, and M. Ando, "A slotted post-wall waveguide array with interdigital structure for 45 linear and dual polarization," *IEEE Trans. Antennas Propag.*, vol. 53, pp. 2865–2871, Sep. 2005.
- [12] WangWei, JinJian, LuJia-Guo, et al. "Waveguide slotted antenna array with broadband, dualpolarization and low cross-polarization for X-Band SAR applications," *IEEE International Radar Conference*, pp. 653-656, 2005.
- [13] Y. Miura, J. Hirokawa, M. Ando, Y. Shibuya, and G. Yoshida, "Double-layer full-corporate-feed hollow-waveguide slot array antenna in the 60-GHz band," *IEEE Trans. Antennas and Propag.*, vol. 59, no.8, pp.2844-2851, Aug. 2011.
- [14] Ansys, Inc. Available: <http://ansys.com/>
- [15] T. Tomura, Y. Miura, M. Zhang, J. Hirokawa and M. Ando, "A 45 Linearly Polarized Hollow-Waveguide Corporate-Feed Slot Array Antenna in the 60-GHz Band," *IEEE Trans. Antennas and Propag.*, vol. 60, no.8, pp.3640-3646, Aug. 2012.

Chapter 6

Conclusion

6-1 Summary of Preceding Chapter

The plate-laminated corporate-feed waveguide slot array antennas are proposed for the sub-millimeter wave bands and dual-polarization operation. The antennas are fabricated by diffusion bonding of laminated thin metal plates for high precision and perfect electrical contact.

For the sub-millimeter wave band applications, the 4x4-, 16x16-, 32x32- and 64x64-element array antennas are designed and fabricated. They show more than 21 dBi, 32 dBi, 38 dBi and 43 dBi gains with more than 60% antenna efficiencies over 13 GHz in the 120 GHz band, respectively.

The antenna for dual-polarization with a multiple feeding layer is proposed in the 60 GHz band. It shows over 32 dBi gain with over 80% antenna efficiency and high isolation over 50 dB between two different polarizations.

Chapter 1 presents the background of this research.

Chapter 2 presents the design of a double-layer corporate-feed waveguide antenna for sub-millimeter wave bands including 120 GHz and 350 GHz bands. The design limitations with the diffusion bonding of plate laminated thin metal plate in high frequency are discussed. With the considerations of the design limitations, the 16x16-element array antennas are designed for 120 GHz and 350 GHz bands, respectively. For the 120 GHz band, the high antenna efficiency of 70% has been obtained over a wide bandwidth. For the 350 GHz band, the antenna efficiency of 40% has been obtained over a wide bandwidth. Furthermore, the data transmission experiments are performed with the proposed antennas to demonstrate their feasibility for the wireless link systems in sub-millimeter wave bands. The error-free data transmission is achieved with the proposed antennas at up to 10Gbit/s in 120 GHz and 350 GHz bands, respectively.

Chapter 3 presents the design of the large size array antennas for 120 GHz band. To make a stable structure for the fabrication method of diffusion bonding of laminated thin metal plates, we propose a double-layer feeding structure. With the proposed feeding structure, we successfully fabricate high gain antennas having low profiles. A 32x32-element array antenna shows 38 dBi antenna gain with 60% antenna efficiency over 15 GHz bandwidth and a 64x64-element array shows 43 dBi antenna gain with 50% antenna efficiency over 14.5 GHz bandwidth, respectively.

Chapter 4 presents the design of the 4x4-element corporate-feed waveguide slot array antenna with cavities for the 120 GHz band. In order to adjust the mutual coupling effects in a small size array antenna, a cavity structure is employed for each radiating slot. The cavity in this antenna structure provides proper mutual coupling effects for each radiating element. As a result, high antenna efficiency is achieved in a broad bandwidth. The antenna gain of 21.1 dBi and the antenna efficiency of 80.0 % are measured at the center frequency. The measured 1 dB-down gain bandwidth is 22 GHz (from 118 to 140 GHz) which is corresponding to 17.6%.

Chapter 5 presents the design of the dual-polarization slot array antennas for 60 GHz band. In order to realize the dual-polarization operation, the cross-shaped radiating slots which are sharing the common aperture and the multi-layer feeding structure are employed. A 16x16-element array for the dual-polarization is fabricated by diffusion bonding of laminated thin copper plates which has the advantages of high precision and low loss characteristic. The antenna gain of 32.4 dBi and the antenna efficiency of 84.0 % are measured at center frequency for two different polarizations. The 1-dB-down gain bandwidths are no less than 6.5% and 7.0% and a wideband characteristic is achieved. High isolation over 50 dB is achieved in the broad bandwidth between two different polarizations.

6-2 Remarks for the Future Studies

The remarks for further investigation are listed below:

(1). Evaluation of effective loss for the antennas made by diffusion bonding

Strictly, the antennas fabricated by diffusion bonding of laminated thin metal plates have higher loss than the calculated result.

It may come from

(a) Change of conductivity after diffusion bonding

Because the atom structure of copper could be changed during diffusion bonding. Thus, actual conductivity of copper also could be changed in fabrication.

(b) Surface quality of etching

In many case, we could find a rough surface in the etched antenna pattern as shown in Fig. 1. This rough surface may result scattering of the electromagnetic field. Ultimately, it may cause a loss in antenna structure.

Thus, the evaluation of effective loss for the structure fabricated by diffusion bonding of laminated metal plates is needed.

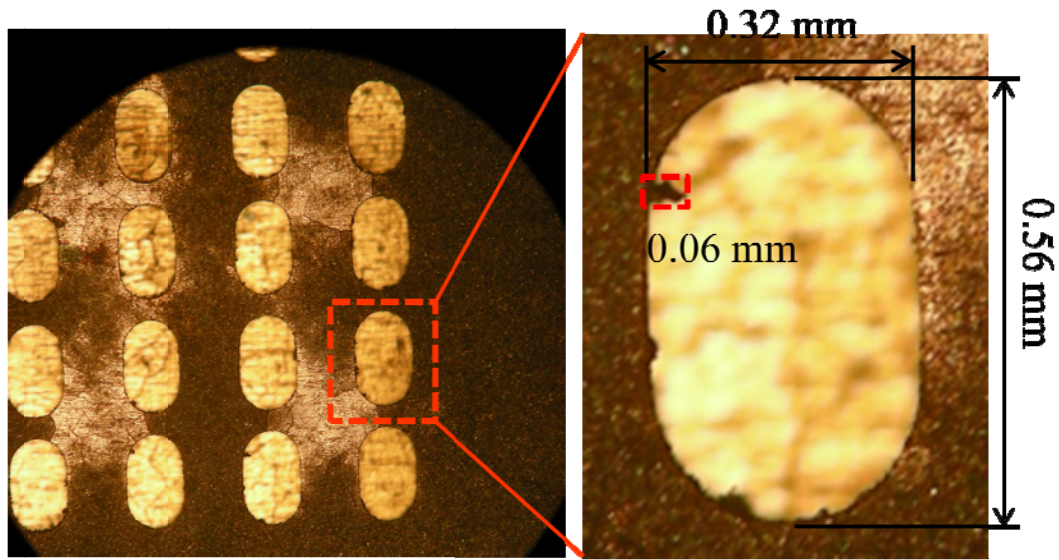


Fig. 1 Microscope picture of slot fabricated by diffusion bonding of laminated thin metal plates,
16x16-element array for 350 GHz

(2). Improvement of the Bandwidth

The 2x2-element subarray structure in Chapter 2, 3, and 4 should be improved to obtain the wider gain bandwidth. Actually, the full-corporate feeding networks in 120 GHz satisfy the reflection bandwidth more than 20% at the VSWR 1.2. Thus, if we improve the reflection bandwidth of the 2x2-element subarray structure, we could have a wider gain bandwidth in the proposed antenna structures. Also, we could expect the better data transmission performance with the improved bandwidth if it is possible.

(3). Improvement of the Measurement System

The antennas proposed for the 120 GHz band in Chapter 2, 3, and 4 are measured with the Vector Network Analyzer. However the measurement is performed in very short distance with a quasi-anechoic environment surrounded by absorbers. It results many ripples and errors in measurement results. The potential error in antenna gain is more than ± 0.5 dB at this moment. To increase the reliability of the measurement results, the measurement system should be improved. Also, the near field measurement using Vector Network Analyzer will be a good method to evaluate the antenna performances.

List of Publications

1. Journal Paper

- [1] **Dongjin Kim**, Jiro Hirokawa, Kimio Sakurai, Makoto Ando, Takada Takumi, Tadao Nagatsuma, Jun Takeuchi and Akihiko Hirata, "Design and measurement of the plate laminated waveguide slot array antenna and its data transmission experimental results for the 120 GHz band", IEICE Trans. Communications., vol. E96-B, no.08, pp. 2102-2111, Aug. 2013.

- [2] **Dongjin Kim**, Jiro Hirokawa, Makoto Ando, Jun Takeuchi and Akihiko Hirata, "4x4-element Corporate-feed Waveguide Slot Array Antenna with Cavities for the 120 GHz-band", – Conditional Acceptance to IEEE Transaction on Antenna and Propagation

- [3] **Dongjin Kim**, Jiro Hirokawa, Makoto Ando, Jun Takeuchi and Akihiko Hirata, "64x64-element and 32x32-element slot array antennas using double-layer hollow-waveguide corporate-feed in the 120 GHz band", –Conditional Acceptance to IEEE Transaction on Antenna and Propagation

- [4] **Dongjin Kim**, Miao Zhang, Jiro Hirokawa and Makoto Ando, "Design and Fabrication of Dual-polarization Waveguide Slot Array Antenna with High Isolation and High Antenna Efficiency for 60 GHz-band", –Submitted to IEEE Transaction on Antenna and Propagation

2. International Conference

- [1] **Dong-Jin Kim**, Jiro Hirokawa, Makoto Ando, Makoto Ando, Jun Takeuchi and Akihiko Hirata, "4x4-element Corporate-feed Plate-laminated Waveguide Slot Array Antenna with Cavities for the 120 GHz-band", Session 124.1 *IEEE APS/URSI 2013*, July 2013, Florida, USA

- [2] **Dong-Jin Kim**, Miao Zhang, Jiro Hirokawa and Makoto Ando, "Dual-polarized Corporate-feed Plate-laminated Waveguide Slot Array Antenna for 60 GHz-band", Session IB3-4, *ISAP 2012*, Oct. 2012, Nagoya, Japan

- [3] **Dongjin Kim**, Miao Zhang, Jiro Hirokawa and Makoto Ando, "Design of Dual-polarization Waveguide Slot Array Antenna using Diffusion Bonding of Laminated Thin Plates for the 60 GHz-band", Session 463.10 *IEEE APS/URSI 2012*, July 2012, Chicago, USA

- [4] **Dongjin Kim**, Jiro Hirokawa, Makoto Ando, Takada Takumi, Tadao Nagatsuma, Jun Takeuchi and Akihiko Hirata, "Design and Measurement of Plate Laminated Waveguide Slot Array Antennas in Sub-millimeter wave frequency bands", Session WD, *APMP 2012*, April 2012, Kyoto, Japan

- [5] Jiro Hirokawa, **Dongjin Kim**, Takuma Takada, Tadao Nagatsuma, Jun Takeuchi, Akihiko Hirata "Designs and Measurements of Plate-laminated Waveguide Slot Array Antennas for 120GHz Band and 350GHz Band", Session TU6E-01, *APMC 2011*, Dec. 2011, Melbourne, Australia

- [6] **Dong-Jin Kim**, Miao Zhang, Jiro Hirokawa and Makoto Ando, "Feeding Structure to Widen Bandwidth for Dual-Polarization Corporate-feed Waveguide Slot Array Antenna ", Session PB2, *ISAP 2011*, Oct. 2011, Jeju, Korea

- [7] **Dong-Jin Kim**, Jiro Hirokawa, Tadao Nagatsuma and Makoto Ando, "Design and Fabrication of a Corporate-feed Plate-laminated Waveguide Slot Array Antenna for 120GHz-band", Session 429.7, *IEEE APS/URSI 2011*, July 2011, Spokane, USA

3. Domestic Conference

- [1] **Dongjin Kim**, Jiro Hirokawa and Makoto Ando, " Design of 4x4-element Corporate-feed Waveguide Slot Array Antenna with Cavity for 120 GHz-band", Session B-1, *IEICE General Conference 2013*, March 2013, Gifu

- [2] **Dongjin Kim**, Jiro Hirokawa, Makoto Ando, Makoto Ando, Jun Takeuchi and Akihiko Hirata, "64x64- and 32x32-slot array antennas using double-layer hollow-waveguide corporate-feed in the 120 GHz band", AP2013-05, *IEICE Technical Meeting*, May 2013, Beppu

- [3] **Dongjin Kim**, Jiro Hirokawa, Kimio Sakurai, Makoto Ando, Takuma Takada and Tadao Nagatsuma, "Double-layer Corporate-feed Hollow-waveguide Slot Array Antenna and Its Data Transmission Test Results in millimeter wave band", ED2012-112, *IEICE Technical Meeting*, January 2012, Tokyo

- [4] **Dongjin Kim**, Jiro Hirokawa, Kimio Sakurai, Takuma Takada and Tadao Nagatsuma, "Design of New Feeding Structure for Large-size Corporate-feed Waveguide Slot Array Antenna using Diffusion Bonding of Laminated Thin Plate for 120 GHz-band", Session B-1, *IEICE Communication Society Conference 2012*, September 2012, Toyama

- [5] **Dongjin Kim**, Miao Zhang, Jiro Hirokawa and Makoto Ando, "Design of Dual-polarization Corporate-feed Waveguide Slot Array Antenna using Diffusion Bonding of Laminated Thin Plate for 60 GHz-band", Session B-1, *IEICE General Conference 2012*, March 2012, Okayama

- [6] Jiro Hirokawa and **○Dongjin Kim**, "Plate-laminated Waveguide Slot Array Antennas for 120GHz Band and 350GHz Band", AP2011-132, *IEICE Technical Meeting*, January 2012, Naha.

- [7] **Dongjin Kim**, Jiro Hirokawa, Kimio Sakurai, Takuma Takada and Tadao Nagatsuma, "Design and fabrication of double-layer waveguide slot array

antennas using diffusion bonding of laminated thin plate for 350 GHz-band ",
Session B-1, *IEICE Communication Society Conference 2011*, Oct. 2011,
Hokkaido

- [8] **Dongjin Kim**, Jiro Hirokawa, Kimio Sakurai, Makoto Ando, Takuma Takada and Tadao Nagatsuma, "Design and Measurement of a corporate-feed plate-laminated waveguide slot array antenna for 120 GHz-band", AP2011-44, *IEICE Technical Meeting*, July 2011, Shinsyu

- [9] **Dongjin Kim**, Jiro Hirokawa, Tadao Nagatsuma and Makoto Ando, "120GHz-band Design of a Corporate-feed Plate-laminated Waveguide Slot Array Antenna", Session B-1, *IEICE General Conference 2011*, March 2011, Tokyo

4. Award

- [1] APMC 2011 Award, APMC 2011, December 2011
- [2] Young Scientist Award, IEEE photonic society, April 2012
- [3] Winner, The 1st Global Talent Forum, Hyundai Motor Company, August 2012

5. Patent

- [1] 広川二郎、金東眞、“平面スロットアレイアンテナ”, February 2013

Acknowledgement

I would like to express my gratitude to Associate Professor Jiro Hirokawa for his continuous guidance and encouragement. I also wish to express my gratitude to Professor Makoto Ando for his fruitful advice.

I am very grateful to Professor Kyeong-sik Min in Korea Maritime University. He was my advisor during my master degree course and also he is a lifelong advisor. I wish to express my gratitude to my great mentor, Dr. Se-hyun Park in Samsung Electronics.

I am indebted to Professor Tadao Nagatsuma in Osaka University, Dr. Akihiko Hirata and Mr. Takeuchi Jun in NTT Microsystem Integration Laboratory for their helpful advices during my doctor degree course.

I would like to thank Mr. Kimio Sakurai, Dr. Takuichi Hirano for their technical assistance. I wish to thank Ms. Keiko Nagao, Ms. Moto Sugano and Ms. Akane Saito for their helps in my daily life. Many thanks are due to Dr. Miao Zhang, Dr. She Yuanfeng, Dr. Maifuz Ali, Mr. Takashi Tomura, Mr. Lu Pengfei, Mr. Sano Makoto, Mr. Le Vu Hung, Mr. Nguyen Xuan Tung, Mr. Takayuki Kohama and the other members of Ando and Hirokawa Laboratories for their fruitful helps.

My wife, Ye-Ri Lee makes me a better person in everyday. Without her, I never started and even finished this study and successful life in Japan. She is my good partner, best friend and a great adviser in my life. I deeply appreciate for her devoted support for every moment in my life. I am also very thanking to my two daughters, Daeun and Yueun. They are the biggest pleasure in my life.

Finally, I express my respects to my parents and my brother for their support.