

Novel Architectures and
an Efficient Calibration Technique
for Low-cost and High-performance
Phased Array Antennas

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Chapter 1

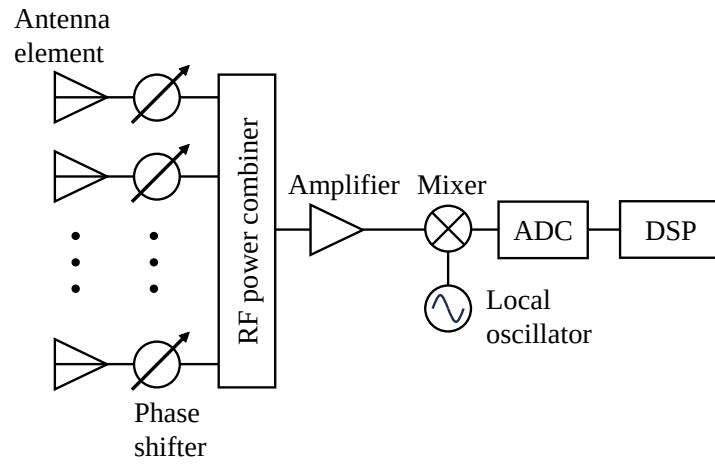
General Introduction

1.1 Background

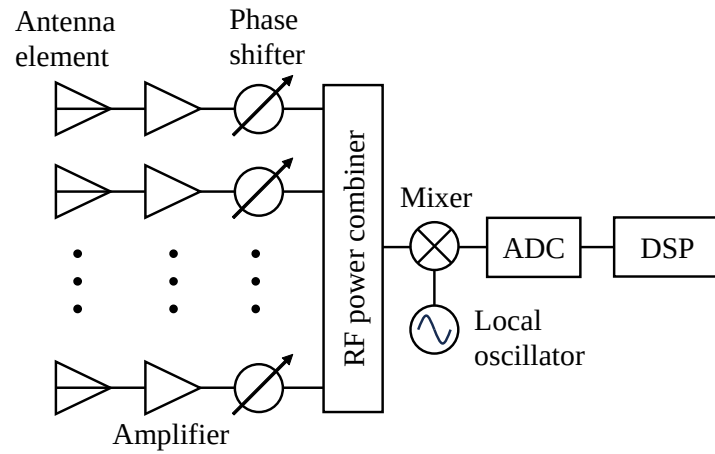
The demand for and sophistication of various wireless systems have been increasing rapidly, as evidenced by the drastic spread of smartphones and Internet of Things (IoT) devices, the increased use of digital data through remote sensing, and the growing popularity of satellite Internet access. As wireless systems become more sophisticated, the antennas used to transmit and receive radio frequency (RF) signals must also become more advanced and sophisticated. Consequently, phased array antennas (PAAs) with flexible beamforming capability are attracting much more attention. This section presents a brief overview of the relationship between PAAs and the aforementioned wireless systems.

1.1.1 Architectures of Phased Array Antennas

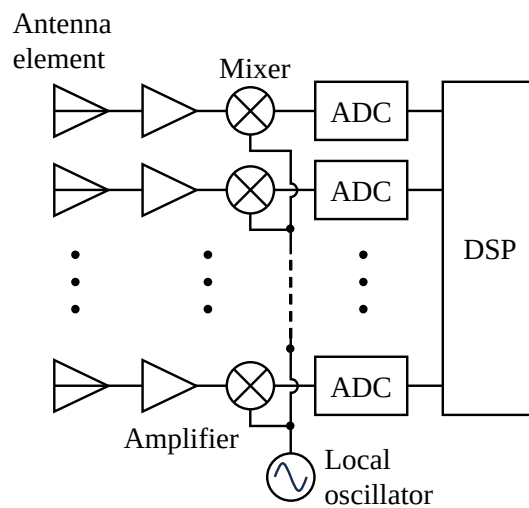
Figure 1.1 shows the block diagrams of typical receiving PAAs. A PAA comprises multiple antenna elements, and the phase of the received RF signals is adjusted in the RF domain by phase shifters connected to each antenna element, as shown in Fig. 1.1 (a) and (b). Figure 1.1(a) depicts a passive PAA (PPAA), in which the RF signal amplification is performed with a single power amplifier unit after the RF signals are combined. In contrast, Fig. 1.1 (b) shows an active PAA (APAA), where the RF signals for each antenna element are amplified separately by their respective power amplifier units and then combined. Figure 1.1 (c) depicts a digital beamforming (DBF) array antenna in which the received signals of each antenna element are digitized by their respective analog-to-digital converters (ADCs) and the amplitude and phase adjustment of the signals, i.e., beamforming, is performed in the digital domain by a digital signal processing (DSP) unit. The PAAs are capable of electronically scanning the beam, and the DBF array antennas possess advanced beamforming capabilities such as multiple simultaneous beamforming and adaptive beamforming.



(a)



(b)



(c)

Figure 1.1: Block diagrams of typical receiving PAAs: (a) passive PAA, (b) active PAA, and (c) DBFarray antenna.

1.1.2 Phased Array Antennas in Various Wireless Systems

In the IoT systems, a multitude of electrical devices, such as home appliances, medical devices, and industrial sensors, are connected to the Internet. Information of each device is collected, shared, and analyzed, and the devices are remotely monitored and controlled [1.1]. The connection between the IoT devices and the network is typically achieved by wireless communication. However, the power for each IoT device is typically supplied by a battery or cables, which requires periodic battery replacements and troublesome wiring work. This is one of the challenges in the implementation and maintenance of the IoT system. To address this challenge, wireless power transfer and RF energy harvesting, which involve the supply or harvesting of power for IoT devices via wireless or ambient radio waves, have been attracted attention and have been actively studied and developed [1.2]–[1.6]. Since microwave power transfer (MPT) can provide higher power more stably than the RF energy harvesting, it is suitable for a wide range of applications. In MPT systems, a microwave transmitter is a crucial component that determines the power transmission capacity and efficiency. In order to efficiently supply power to mobile devices such as wearable devices and drones, which are now considered a new form of the IoT device, PAAs as well as fixed-beam antennas are used for the transmitter, and various MPT systems using PAAs have been studied and developed [1.5], [1.7]–[1.9]. For example, Ossia Inc. has developed an MPT system called Cota, which uses 5.8 GHz band microwave. Cota’s transmitter is a retrodirective PAA, which enables it to track the receivers with beacon signals and to efficiently supply power over the air [1.9]. In recent years, millimeter wave (24 GHz band) phased array transmitters have been developed for improving beam collection efficiency and for miniaturization of the transmitter [1.10], [1.11].

As mentioned above, to enhance the power transmission efficiency of the MPT system, it is essential to improve the efficiency of the transmitter. Therefore, the efficiency of the antenna itself must be improved, as well as the efficiency of the oscillator and power amplifier. In addition, high-gain, high-power phased array transmitters will be needed to power mobile devices that demand high power and are located far from the transmitter, such as drones. Such a PAA becomes physically large and requires many high-power amplifiers, resulting in an expensive MPT system.

The Next topic is remote sensing, with a particular focus on radar systems. Weather radars have been in practical use since the 1950s and are one of the most familiar radar systems in our lives [1.12]. In conventional weather radars, a parabolic antenna with a single linear polarization is elevated to higher elevation angles while being mechanically rotated azimuthally for a volume scan. Since the 2010s, the antennas have been upgraded to have dual polarization for better observational performance such as improved accuracy of precipitation estimation [1.13]. In addition, phased array weather radars have been actively developed in recent years to meet the growing demand for high-precision and rapid

observations, which are necessary for accurate forecasting and timely warning of rapidly evolving severe weather phenomena. The National Weather Radar Testbed Phased Array Radar in the United States is a PPAA and was the first to demonstrate clutter mitigation by adaptive null forming [1.14]. More recently, phased array weather radars with DBF receivers have also been developed for further rapid observation using multi-beam forming technology [1.15], [1.16].

Synthetic aperture radar (SAR) is the most representative system in satellite remote sensing and is used for landscape observation and disaster monitoring. Many satellite SAR systems use not only reflector and planar antennas, which offer lower hardware costs and weight, but also PAAs, which can provide various observation modes of SAR through their electronically beam scanning capability [1.17]–[1.20]. Moreover, dual-polarized antennas are widely used for SAR polarimetry, which enables advanced SAR image analysis. Recently, DBF SAR systems have been developed for both high-resolution and wide-swath observations [1.21], [1.22]. For example, ALOS-4 is an L-band SAR satellite in Japan and will have a swath width four times that of its predecessor due to the use of multi-beam forming technology [1.20].

As described above, phased arrays for remote sensing appear to be maturing. In addition, upgrading from analog phased arrays to DBF arrays is expected to continue to further enhance performance and incorporate additional functionality. However, because DBF arrays require ADCs for each antenna element, as shown in Fig. 1.1(c), the high hardware cost is one of the barriers to widespread adoption of DBF arrays [1.23].

The last topic is satellite communications, with a particular focus on user terminal antennas for satellite Internet access. Since the 2000s, satellite Internet access has expanded considerably in response to the growing demand for Internet communications [1.24], [1.25]. In particular, satellite constellations, which are networks of multiple satellites, have recently been developed to provide low-latency Internet service to users around the world. These constellations include SES’s O3b Medium Earth Orbit satellites and SpaceX’s Starlink Low Earth Orbit (LEO) satellites [1.26]. Reflector and planar antennas have been mainly used as the conventional user terminal antennas for satellite Internet. These antennas have a bulky structure and a slow beam scanning speed due to their mechanical steering by gimbals. Therefore, there are limitations on the platforms on which they can be mounted. Consequently, low-profile, electronically beam-scanned phased-array user terminals have been developed recently for in-flight connectivity and continuous connectivity to LEO satellites. BAE Systems and Viasat have independently developed flat-panel user terminals comprising modular and scalable Ka-band active phased arrays that provide a 360° field of view [1.27], [1.28]. CesiumAstro and Satixfy have developed user terminal antennas of DBF arrays with the multi-beam forming capability for seamless handover between multiple constellation satellites [1.29], [1.30]. PAAs for satellite Inter-

net user terminals appear to be in a growth phase now and are expected to continue to grow in the future.

As described above, the use of phased arrays has been expanded to the various wireless systems and is expected to further increase in the future. This will require the development of low-cost or affordable PAAs that offer high performance and advanced functionalities. While reducing the cost of the components used in PAAs or DBF array antennas is undoubtedly important, this is expected to be a gradual process. Therefore, this study focuses on innovative technologies, particularly antenna architectures, which enable a considerable reduction in the costs of PAAs and DBF array antennas. In addition, this study focuses on PAA calibrations, which are crucial for the accurate and optimal operation of PAAs and DBF array antennas.

1.2 Brief Review of Previous Studies

1.2.1 Technologies for Low-cost PAAs

High gain or large aperture APAAs are typically required for transmitters in MPT systems and satellite Internet user terminal antennas. Antenna elements should be arrayed at an interval of approximately one-half wavelength at the operating frequency to scan the main beam over a wide coverage without grating lobes. Therefore, a considerable number of antenna elements or power amplifiers are required for the APAAs. In particular, APAAs for transmitters in high-power MPT systems require high-power amplifiers, which are typically implemented using expensive monolithic microwave integrated circuit (MMIC) processes such as gallium nitride [1.31], [1.32], resulting in high hardware costs. On the other hand, recent advances in silicon-germanium and CMOS semiconductor technology, which enable highly miniaturized and low-cost RF integrated circuits (RFICs), have made multichannel beamforming RFICs for satellite Internet user terminals commercially available at an affordable price. Many satellite Internet user terminal APAAs employ these RFICs [1.33]–[1.35]. However, the considerable number of RFICs in APAAs results in high power consumption, which hinders the installation of APAAs in mobile platforms where power supply is limited. Therefore, the high hardware cost and the high power consumption remain barriers to the adoption of APAAs in the aforementioned systems.

PPAAs offer the advantage of lower hardware cost and lower power consumption compared to APAAs. However, as shown in Fig. 1.1(a), RF signals pass through the phase shifter before signal amplification at the receiver side or after signal amplification at the transmitter side. Consequently, the power dissipation at the phase shifter (typically insertion losses of 3–10 dB) has a significant effect on antenna performance, such as a decrease in signal-to-noise ratio or transmit power. This appears to be a fatal problem for PPAAs. Therefore, various alternative technologies have been proposed to achieve a low-cost PPAA without the use of conventional phase shifters [1.36]. A brief review of the previous work on these technologies is presented here, with the exception of frequency-scanning and beam-switching PPAAs. The PPAAs may be classified according to their beam scanning approach. Typical examples include reconfigurable reflectarrays and lens antennas, continuous transverse stub (CTS) antennas, holographic beamforming antennas, and arrays using antenna elements with phase-tuning capability. Table 1 shows a comparison of these approaches.

A reflectarray antenna is a hybrid of a reflector antenna and an array antenna, comprising a planar reflector with many reflecting elements and a feeding antenna. Beam scanning is accomplished by adjusting the reflection phase of each reflecting element, which can be controlled by tunable devices integrated into them [1.37]. For example, a reconfigurable reflectarray using resonant patch elements loaded with varactor diodes is

presented in [1.38]. The resonant frequency, and thus the reflection phase, is controlled by tuning the loaded varactor diodes. Reflecting elements with a phase-shifting transmission line loaded with varactor diodes are used in [1.39]. The transmission line functions as a phase shifter, and the phase shift, and thus the reflection phase, is tuned by the varactor diodes. A tunable liquid crystal substrate is used for a reflector in [1.40]. The dielectric constant of the liquid crystal layer is adjusted with a bias voltage to achieve an optimal reflection phase. Spiraphase-type reconfigurable reflecting elements are used in [1.41] and [1.42]. In this type of reflectarray, circularly polarized elements are rotated for beam scanning. The ring-slot elements with switchable radial stubs are rotated electrically by switching the embedded diodes in [1.41], whereas split-ring elements are rotated mechanically by a micromotor in [1.42]. Although reflectarrays have the advantages of ease of fabrication, a simple structure, low cost, and light weight, they are bulky when a feeding antenna is included. In addition, they are fundamentally limited in bandwidth and low efficiencies of less than 70% [1.43].

A lens antenna is a type of aperture antenna comprising a shaped piece of dielectric material or artificial dielectric material and a feeding antenna. Reconfigurable or beam-scanning lens antennas have been achieved through the mechanical alteration of the relative position between the lens and the feeding antenna or through the electronic control of phase shifts through the lens. For example, in [1.44], a feeding antenna is mechanically moved parallel to an inhomogeneous gradient-index dielectric flat lens for beam scanning. A lens antenna with a pair of metasurface lenses is proposed in [1.45]. Each lens has an individual beam tilt angle and is mechanically rotated independently or synchronously around the antenna axis for beam scanning. In [1.46], antennas using lenses of ferroelectric slabs sandwiched between metal plates are presented. The dielectric constants of the ferroelectric slabs can be controlled by changing the bias voltage between each plate to achieve one-dimensional beam scanning. Two-dimensional beam scanning is achieved by placing a 90° polarization rotator between two spatially orthogonal ferroelectric lenses. A metasurface lens comprising multiple parallel substrates on which subwavelength scatterers are periodically printed with embedded varactor diodes is presented in [1.47]. Beam scanning is achieved electronically by adjusting the phase shifts through the lens by individually controlling the varactor diodes for each scatterer. Like reflectarrays, reconfigurable lens antennas have the advantages of ease of fabrication, a simple structure, low cost, and light weight. However, they are bulky when a feed antenna is included, and they have low aperture efficiency and an additional insertion loss of the lens, resulting in a considerably reduced efficiency (e.g., 27% in [1.44] and 41% in [1.47]).

A CTS antenna is a type of leaky wave antenna comprising radiating broad stubs periodically arrayed on a parallel-plate waveguide fed by a line source or a linear array feed antenna [1.48]. Several techniques have been proposed for beam scanning CTS antennas.

In [1.49] and [1.50], a CTS antenna with a parallel-plate waveguide filled with ferroelectric material is presented. The H-plane scanning (parallel to the stubs) is achieved by feeding the parallel-plate waveguide with a linear phased array. The E-plane scanning is achieved by controlling the propagation constant of the guided wave, and thus the radiation phase difference between the stubs, by adjusting the bias voltage on the ferroelectric material. This technique requires a linear phased array with the conventional phase shifters as the feeding antenna. In [1.51] and [1.52], a linear phased array is replaced with tilting or rotating the radiating stubs with respect to the parallel-plate waveguide, which enables the H-plane scanning. The E-plane scanning is achieved by rotating the radiating stubs and parallel-plate waveguide together around the antenna axis. This antenna is called variably inclined CTS (VICTS) antenna and is currently being commercialized as a satellite Internet user terminal antenna [1.53]. Typical CTS antennas are linearly polarized antennas, which require a polarizer over the antenna aperture to effectively handle the circularly polarized wave. In order to overcome this limitation, circularly polarized antennas that replace a radiating broad stub with a truncated patch array or a single-arm spiral element array have been proposed in [1.54] and [1.55], respectively. Although CTS antennas have the advantage of a flat and low-profile structure, they have moderate efficiencies of 30–70% [1.56], [1.57] and are subject to beam squint, resulting in a narrow instantaneous bandwidth. In addition, CTS antennas can only scan the beam and do not have flexible beamforming capabilities such as null forming.

Holographic antennas are antennas in which the holographic technique, as it is employed in the optical field, is applied to the RF pattern synthesis [1.58]. Microwave holograms are typically formed by an artificial impedance surface, or metasurface, in which subwavelength metal patterns are periodically printed on the substrate [1.59], [1.60]. These holographic antennas are a kind of leaky wave antenna, reflectarray, or lens antenna. The metal patterns are determined based on an interference wave of an incident wave (reference wave) and a desired wave (object wave). When the incident wave from a feeding structure illuminates the hologram, the hologram generates the desired scattered field. Reconfigurable or beam scanning holographic antennas are achieved by controlling the scattering characteristics of each unit-cell of the hologram. For example, a microstrip leaky wave antenna with arrayed slot-shaped subwavelength metamaterial elements is presented in [1.61]. Each slot contains a PIN diode, which is used to activate or deactivate the slot element. The slot elements illuminated by the incident wave with approximately correct phase are activated and the others are deactivated to achieve the desired radiation pattern. A metasurface lens antenna, comprising arrayed subwavelength C-shaped split-ring resonators, is shown in [1.62]. Each split-ring resonator is equipped with eight integrated CMOS switches, which control the rotation and a split gap of the split-ring structure. This allows the amplitude and phase of the scattered fields to be

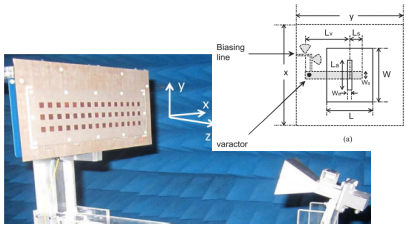
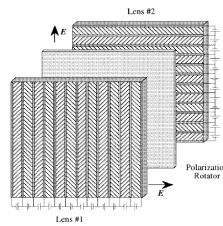
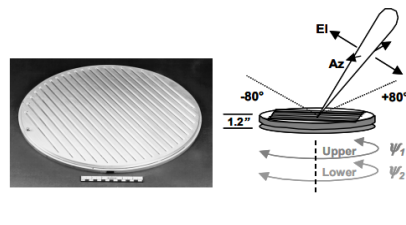
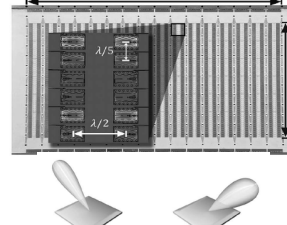
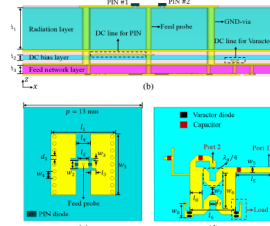
controlled, thus enabling flexible beamforming. Another leaky wave antenna comprising a rectangular waveguide covered with a metasurface is proposed in [1.63] and [1.64]. Each metamaterial element comprises a subwavelength slot coupled to a patch, with a liquid crystal filled between the slot and the patch. The dielectric constant, and thus the resonant frequency of each patch, is controlled by the bias voltage applied to the liquid crystal, thereby allowing the amplitude of the radiated field from each patch to be controlled. The radiation amplitudes of each patch are controlled in accordance with the degree of correspondence between the illuminated and desired radiation phase to achieve the desired radiation pattern. Holographic antennas have the advantages of low-profile and flexible beamforming capabilities. However, they suffer from insertion losses due to the embedded switches or liquid crystal, which results in low efficiencies (e.g., 35% in [1.61]). In addition, high sidelobe level (> -10 dB) due to the inherent imperfections of holograms and the periodicity of the aperture distribution is another challenge, which requires a complex algorithm to prevent [1.63].

PAAs using antenna elements with a phase-shifting function are proposed in [1.65]–[1.67]. Patch antennas loaded with multiple PIN diodes and varactor diodes are presented in [1.65] and [1.66]. These antennas achieve multi-bit phase control by switching the current path on the patch and its feeding structure with the diodes. In [1.67], spiral antennas are used as a direct radiating spiraphase-type element and are mechanically rotated by stepper motors to achieve phase shifts. Both antennas have the flexible beamforming capabilities because the phase of each antenna element can be controlled individually. However, the former suffers from considerable insertion loss due to the embedded diodes, which results in a low efficiency (33% in [1.66]). The latter also exhibits a low efficiency of less than 20% due to the resistors, absorbing materials, and bulky feeding network.

In conclusion, to the best of the author’s knowledge, there are no low-cost PPAAAs with a simple and flat structure, high efficiency, and individual phase control capability for flexible beamforming. Consequently, the realization of such a PPAA remains a challenge.

1. General Introduction

Table 1.1: Comparison of the PPAA's with different beam scanning approaches.

Antenna	Reflectarray  [1.39]	Lens  [1.44]
Size	Bulky	Bulky
Structure	Simple	Simple
Efficiency	Moderate	Low
Individual phase control	Yes	Yes / No (depends on the type of lens)
Antenna	CTS  [1.54]	Holographic antenna  [1.63]
Size	Low-profile	Low-profile
Structure	Complex feeding structure	Simple
Efficiency	Moderate	Low
Individual phase control	No	Yes / No (depends on the type of hologram)
Antenna	Antenna elements with phase shift function  [1.66]	
Size	Low-profile	
Structure	Simple	
Efficiency	Low	
Individual phase control	Yes	

1.2.2 Technologies for Low-cost DBF Array Antennas

Element-level DBF array antennas have a separate transceiver or ADC/digital-to-analog converter (DAC) at each antenna element of the array, as shown in Fig. 1.1(c). RF signals are digitized at the element level, and amplitude and phase control of each signal and subsequent beamforming are performed in the digital domain. As a result, element-level DBF array antennas are capable of more flexible and precise beamforming and will offer increasing functionality such as simultaneous multi-beam forming, adaptive beamforming, and multiple-input multiple-output (MIMO) processing. Therefore, DBF array antennas have been increasingly implemented in various radar systems and wireless communication systems in recent years [1.23][1.68]. However, element-level DBF array antennas require a separate RF chain (ADC/DAC, amplifier, mixer, etc.) for each antenna element. This results in high power consumption and high hardware costs for large arrays, which is one of the barriers to the adoption of the DBF array antennas in new applications. In order to reduce the hardware cost and power consumption, several approaches have been investigated to reduce the number of ADCs/DACs, which is a major cost factor.

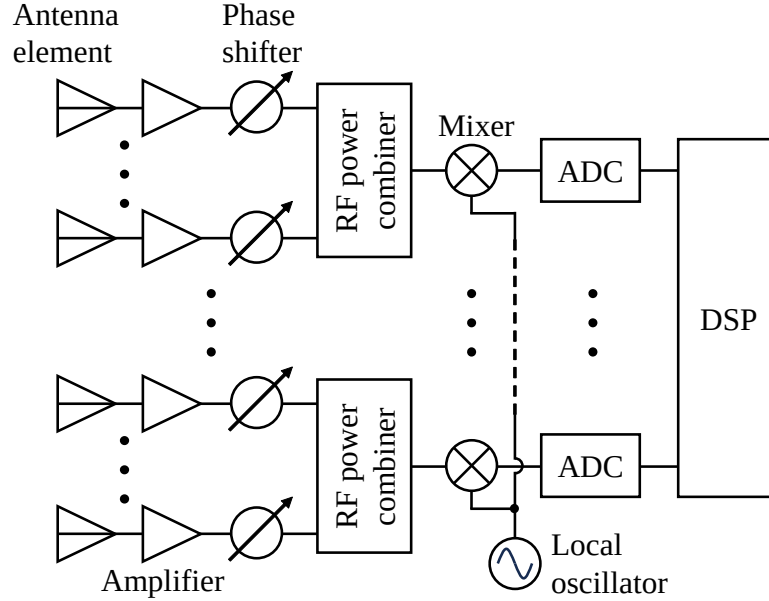


Figure 1.2: Block diagram of typical receiving subarray-level DBF array antenna.

Subarray-level DBF array antennas or hybrid beamforming architectures are frequently employed to obtain the DBF capabilities while reducing the number of ADCs/DACs [1.68], [1.69]. Figure 1.2 depicts a block diagram of a typical subarray-level DBF array antenna. Each antenna element has a separate RF phase shifter, and the analog beamforming is conducted prior to the subarray-level digitization. Therefore, subarray-level DBF array

antennas are limited in their beamforming capabilities, resulting in a diminished performance compared to the element-level DBF [1.69]–[1.71].

Path-sharing architectures using multiplexing techniques have been proposed as a means of reducing the number of ADCs/DACs while maintaining approximately the same beamforming capabilities as the element-level DBF [1.72]–[1.77]. Similar to multiple access technologies in wireless communication systems, multiplexing is performed by three schemes: time-division multiplexing (TDM), frequency-division multiplexing (FDM), and code-division multiplexing (CDM). Figures 1.3, 1.4, and 1.5 show the block diagrams of these schemes [1.77], [1.78]. Their operation is briefly described here using a receiving array as an example.

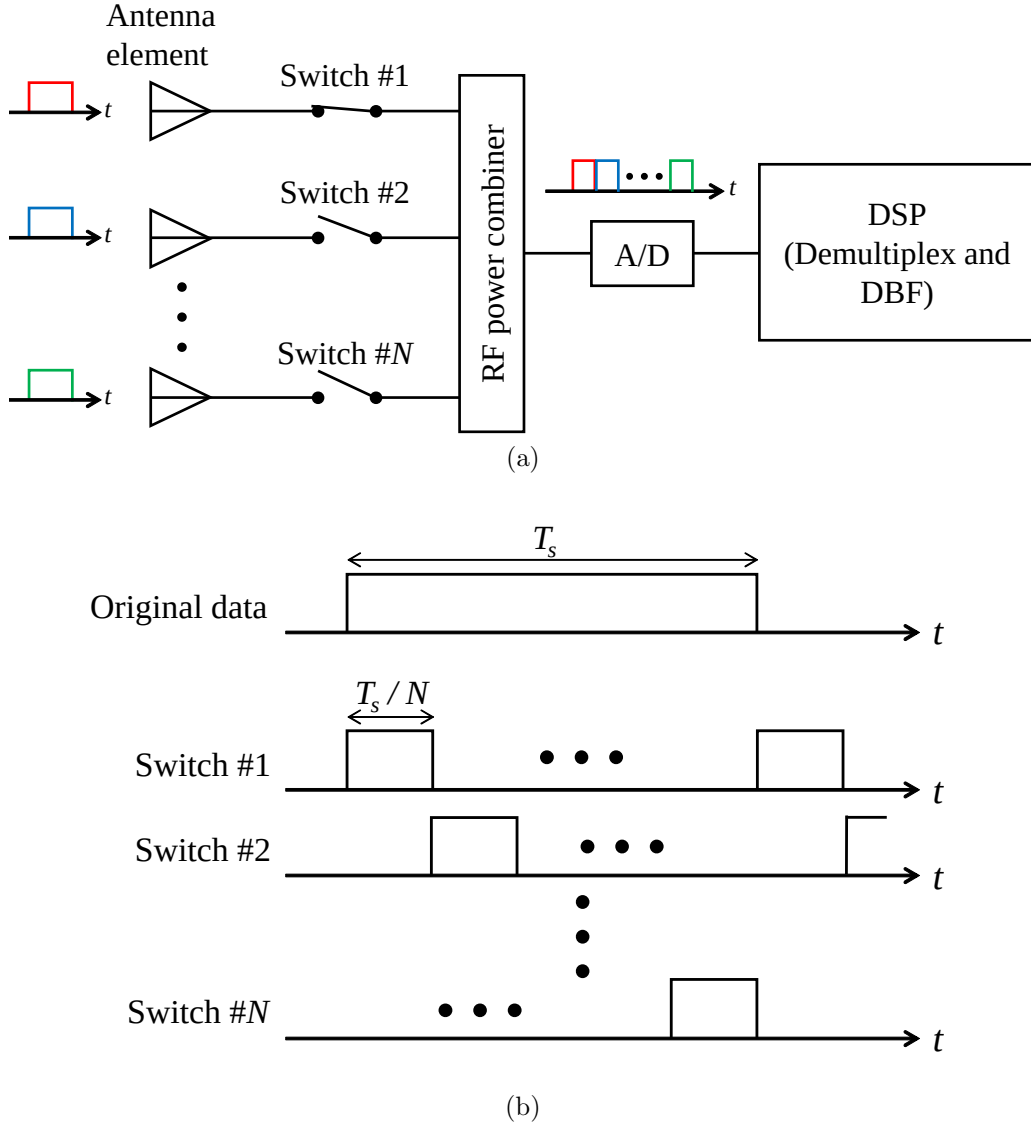


Figure 1.3: TDM-based path-sharing architecture: (a) Block diagram and (b) timing diagram.

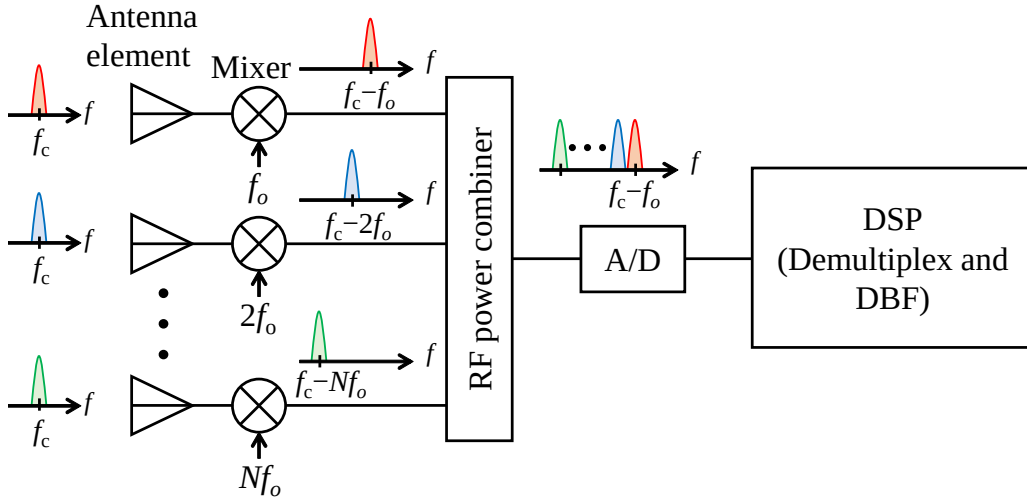


Figure 1.4: FDM-based path-sharing architecture

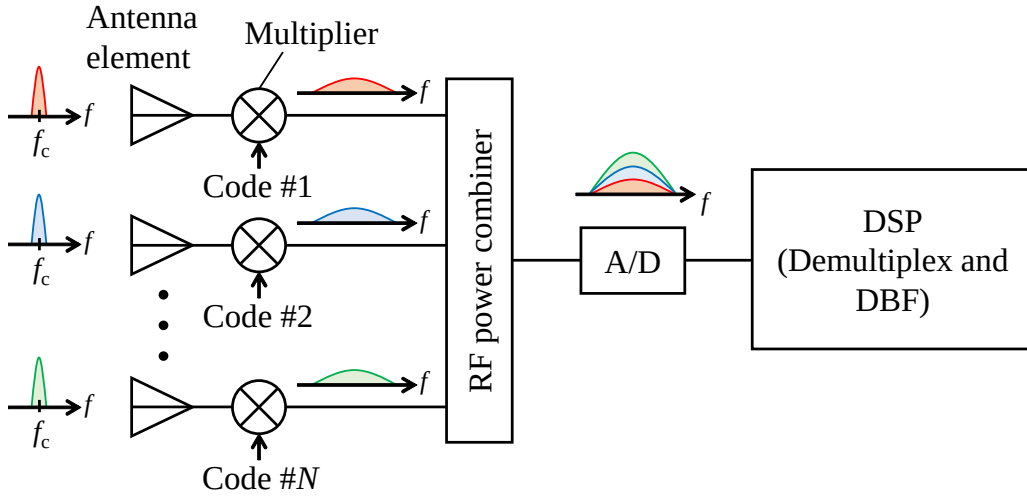


Figure 1.5: CDM-based path-sharing architecture

In array antennas using the TDM scheme, each antenna element typically has a switch [1.63], [1.64], [1.69]–[1.81]. These switches are controlled in such a way that one switch of the array is activated at a time and sequentially activated and deactivated at a rate exceeding the signal bandwidth, as shown in Fig. 1.3(b). The received signals from each antenna element are sequentially fed into a single RF chain and multiplexed in a TDM fashion. The multiplexed signals are then digitized by a single ADC, and the original signals are recovered by signal processing in the digital domain. One of the limitations of the TDM scheme is that the recovered signals are distorted when interference signals are present at adjacent channel frequencies [1.78], [1.82]. In addition, the duration of signal

reception at each antenna element is reduced to $1/N$ of that of ordinary element-level DBF array antennas, resulting in a deterioration of signal-to-noise ratio (SNR) [1.74][1.78]. This SNR reduction is specific to the TDM scheme and does not occur in the FDM and CDM schemes.

In array antennas using the FDM scheme, each antenna element typically has a dedicated frequency mixer [1.72], [1.77]. The received signals from each antenna element are downconverted by each mixer to intermediate frequency (IF) signals with distinct, non-overlapping frequencies. These signals are then combined and fed into a single RF chain, as shown in Fig. 1.4. The original signal can be recovered by filtering out the unwanted signals (signals for the other antenna elements) with a bandpass filter in the digital domain. The FDM scheme also suffers from the interference signals at the adjacent channel frequency. To prevent signal distortion from the interferences, highly selective RF preselect filters are required for each antenna element. These preselect filters should be frequency-reconfigurable for systems that handle frequency-hopping signals or signals with different bandwidths, and they will be considerably larger in size than conventional dedicated filters. In addition, a complex local oscillator (LO) signal generation circuit is necessary to generate multiple closely offset LO frequencies. The FDM scheme still requires many mixers and amplifiers (to compensate for the conversion losses of the mixers), which hinders cost reduction.

In array antennas using the CDM scheme, each antenna element typically has a multiplier or $0^\circ/180^\circ$ phase shifter [1.75], [1.76], [1.78], [1.81], [1.83], [1.84]. In the CDM scheme, the received signals from each antenna element are first multiplied by the spreading codes, whose chip rate is considerably faster than the data symbol rate, as shown in Fig. 1.5. When $0^\circ/180^\circ$ phase shifters are used, the received signals are sequentially phase shifted by the phase shifter in accordance with the spreading codes. The code-modulated signals are then fed into a single RF chain and multiplexed in a CDM fashion. The original signals can be recovered by a despreading process in the digital domain, using the same set of spreading codes that are used in the spreading process. In contrast to the TDM and FDM schemes, the original signals can be accurately recovered without the use of preselect filters, even in the presence of interference signals at adjacent channel frequencies. This is a significant advantage over the other schemes.

In all schemes, the multiplexed signal is fed into the common RF chain with at least N times the bandwidth of the original signals. Consequently, the path-sharing architectures require ADCs with higher sampling rates than those used for conventional element-level DBF array antennas. The hardware cost and power consumption of ADCs typically increase as their sampling rates increase. Therefore, an excessive increase in the number of multiplexed signals and reduction in the number of ADCs will result in an increase in cost and power consumption [1.85].

As mentioned above, the CDM scheme appears to be the most promising path-sharing architecture because it does not suffer from the SNR degradation observed in the TDM scheme and can be implemented with simpler circuitry than the FDM scheme. However, array antennas using the CDM scheme with multipliers still require separate amplifiers for each antenna element to compensate for the conversion loss of the multipliers. Consequently, the number of amplifiers cannot be reduced, which hinders the reduction of cost and power consumption. On the other hand, in the case of the CMD scheme using $0^\circ/180^\circ$ phase shifters, the elimination of amplifiers for each antenna element is feasible if the insertion losses of the phase shifters are acceptably low. In previous works, antennas for the CMD scheme with $0^\circ/180^\circ$ phase shifters were array antennas printed together with phase shifters and combiners on substrates [1.76], [1.84] or reflectarrays/transmitarrays with 1-bit phase reconfigurable elements [1.86]–[1.88]. However, both approaches have limitations. The former is afflicted by the relatively large insertion losses of the phase shifter and combiner at high frequencies, whereas the latter has a low aperture efficiency. Consequently, both have a problem in antenna efficiency.

In conclusion, the path-sharing architecture based on the CDM scheme is considered the most promising approach for maintaining element-level DBF capabilities while reducing costs. In particular, the CMD scheme using $0^\circ/180^\circ$ phase shifters has the potential to significantly reduce the cost of DBF antennas. However, the conventional antennas have a problem in antenna efficiency, which remains a challenge.

1.2.3 Calibration Technologies

In order to achieve the desired radiation patterns or optimum performance of phased array antennas, it is necessary to accurately control the amplitudes and phases of the radiated fields from each antenna element. However, the fields of each antenna element are different and deviate from ideal values due to various error sources. These include manufacturing errors of antenna elements and feeding circuits, individual performance differences of phase shifters and amplifiers, performance variation due to aging, mutual coupling effects between antenna elements, and scattering from surrounding structures. Consequently, a calibration technology that compensates for excitation errors (i.e., deviations of fields of each antenna element from the ideal values) is crucial for the accurate operation of phased array antennas [1.23], [1.89]–[1.91].

Previous studies have primarily focused on the calibration techniques for accurate beamforming of co-polarized fields, including a precise beam scanning and gain improvement [1.92]–[1.95], reduction of the sidelobe level [1.96]–[1.98], precise and deep null-forming [1.96], [1.98], [1.99], and accurate direction of arrival (DOA) estimation [1.91], [1.100], [1.101]. Nevertheless, calibration techniques for reducing cross-polarization have been proposed for systems requiring low cross-polarization performance, such as weather

radar and satellite communication. For example, a calibration technique for a phased array antenna comprising circularly polarized antenna elements with a single feeding port is presented in [1.102]. The co-polarized and cross-polarized fields of each antenna element are measured, and calibration phases are then determined through an optimization process with the objective of improving the co-polarization gain and reducing the cross-polarization level. A calibration technique for reducing the cross-polarization level of a cylindrical polarimetric phased array comprising dual-linearly polarized antenna elements is presented in [1.103]. The full polarization information, including VV, HV, HH, and VH (XY denotes the X-polarized field from the feeding port for the Y-polarization), is measured for each antenna element, and the polarization state is calibrated using this information in the digital domain. These techniques require the individual measurement of each antenna element in the array, which is time-consuming for large arrays. In [1.104]–[1.106], alternative calibration techniques are proposed for phased array antennas comprising dual-polarized antenna elements with a separate phase shifter and variable gain amplifier for each feeding port. In [1.104] and [1.105], following the setting of the phases for beam scanning, the V- and H-polarized fields of the array pattern are measured in the scanning direction. In order to reduce the cross-polarization level, the difference in amplitude and phase of the two fields is corrected using the phase shifters and the variable gain amplifiers. Although this technique does not require the full polarization information and the individual measurement of each antenna element, its effectiveness will be insufficient when the cross-polarization caused by the mutual coupling between the feeding ports of each antenna element is significant. On the other hand, in [1.106], the full polarization information of the array pattern is measured in each scanning direction, and the polarization state is calibrated by the phase shifters and the variable gain amplifiers. Although this technique is more effective in reducing cross-polarization, the full polarization measurement of the array is necessary to implement this technique.

In contrast to beamforming of co-polarized fields, the reduction of cross-polarization requires precise amplitude and phase control. Consequently, calibration measurements should be conducted at every scan direction in these techniques. Therefore, in order to achieve low cross-polarization at any scan directions for a wide-angle scanning phased array antenna, calibration measurements should be conducted at fine angular intervals, which is a time-consuming task and increases the measurement cost. When angular variations of a radiation pattern are gradual, it is possible to replace the measurement with approximate values interpolated using adjacent measured values or calculated by a fitting function. This can reduce the number of required measurement samples. For example, polynomial fitting functions are constructed to approximate the radiation patterns of antennas in [1.107]. Although these functions can accurately approximate the radiation patterns, they require many polynomials due to the absence of a physical relationship be-

tween the polynomials and the radiation from the antennas. In [1.108], a fitting function is constructed by applying a two-dimensional discrete Fourier transform (DFT) to the measured values, followed by an inverse DFT of the resulting data, taking into account the 2π periodicity of the radiation patterns. Similarly, a fitting function is constructed by a Fourier tesseral harmonic expansion in [1.109]. In particular, the latter can accurately approximate the radiation patterns with a smaller number of terms because it employs associated Legendre functions, which are the solutions of the scalar Helmholtz equation in spherical coordinates and are related to the radiation of electromagnetic waves. However, in principle, these methods require a full spherical measurement of the radiation pattern in order to determine the unknown coefficients, which typically requires the use of specialized antenna measurement equipment. The active element pattern (AEP) of an array antenna is approximated by a weighted sum of multiple characteristic basis function patterns (CBFPs) in [1.110]. The CBFPs are represented by simple cosine functions with phase patterns that vary in response to the positions of antenna elements. This approach is equivalent to expressing the AEP as a composition of the radiation from the excited antenna element and the re-radiation from the surrounding antenna elements. This method does not require a full spherical measurement of the radiation pattern, thereby eliminating any constraints on the antenna measurement equipment. Furthermore, it can accurately approximate the active element pattern with a relatively small number of CBFPs; in [1.110], accurate approximations of the radiation patterns are obtained with only three CBFPs. However, previous studies have primarily focused on approximating the co-polarized radiation patterns. To the best of the author's knowledge, there has been no evaluation of fitting functions that include the cross-polarized radiation patterns and the performance of calibrations for the cross-polarization reduction using them.

In conclusion, calibration techniques for reducing cross-polarization are essential for systems utilizing dual-polarization, such as weather radar and satellite communication. However, the conventional techniques have limitations in performance and measurement time. Although the calibration technique using the fitting functions based on the CBFPs appears to have the potential to address these issues, there have been no reports on the cross-polarization reduction using them. Consequently, the development of an efficient and effective calibration technique for cross-polarization reduction remains a challenge.

1.3 Objective and Outline

In order to achieve a PAA with high performance and advanced functionality at an affordable cost, and in particular to address the conventional technical challenges described in Section 1.2, the author will focus on the following three technologies and their development will be the objective of this dissertation:

- (1) Low-cost PPAA with flat structure, high efficiency, and individual phase control capability.
- (2) DBF array antenna based on the CDM-based path-sharing architecture with high antenna efficiency.
- (3) Efficient PAA calibration technique for cross-polarization reduction.

Including this introductory chapter, this dissertation comprises five chapters. The following chapters are organized as follows.

In Chapter 2¹, a novel radial line planar phased array, in which low-profile helical antenna elements are individually rotated by their respective connected micromotors, is proposed to realize a PPAA with a flat structure, high efficiency, and flexible beam forming capability [1.111]. To the best of the author's knowledge, this is the first radial line planar array (RLPA) in which the antenna elements are electromechanically rotated by their individual micromotors. A novel design methodology for RLPA is also described, which combines the equivalent circuit theory and electromagnetic simulations of the unit cell element. The proposed procedure is practical for the design of RLPA of antenna elements with arbitrary probe shape without large-scale full-wave analysis of the whole structure of the RLPA. The performance of the proposed antenna and the effectiveness of the design method are verified by simulation and experiment.

Chapter 3² presents a novel waveguide slot array with a CDM-based path-sharing architecture to provide DBF capabilities [1.112]. The proposed antenna is comprised of a rectangular metallic waveguide's bottom part and a multilayer printed circuit board (PCB) with the rectangular waveguide's top wall and slot apertures. Multiple pairs of two symmetrical longitudinal slots are etched on the metal surface of the PCB, and a PIN diode is mounted across each slot. The received signals of each slot pair are multiplexed in a CDM fashion by switching the bias of the diodes according to the spreading code, and the original signals are then recovered by a despreading process in the digital domain.

¹This chapter is based on "Radial Line Planar Phased Array Using Electromechanically Rotated Helical Antennas" [1.111], by the same author, which appeared in IEICE Trans. Commun., Copyright(C)2023 IEICE.

²This chapter is based on "Waveguide slot array with code-division multiplexing function for single RF chain digital beamforming" [1.112], by the same author, which appeared in IEICE Trans. Commun., Copyright(C)2024 IEICE.

To the best of the author's knowledge, this is the first waveguide slot array with a CDM-based path-sharing architecture. The proposed antenna reduces the number of RF chains, amplifiers, and ADCs and has the potential to offer higher efficiency than previous works using microstrip lines or transmittarray/reflectarray architecture. The feasibility of the proposed antenna is tested using an eight slot pair prototype antenna as a proof of concept.

In Chapter 4³, a novel, efficient calibration technique using a fitting function to reduce cross-polarization, or axial ratio (AR), of circularly polarized PAA is presented [1.113]. The proposed technique will be applied to phased arrays of reconfigurable dual-polarized circular patch antennas: each antenna element has two feeding channel, with a separate variable gain amplifier and phase shifter for each channel. In the proposed technique, the compensation weights, excitation amplitudes and phases of each feeding channel, for AR compensation are controlled based on a design formula obtained by a fitting function of an AEP of the array. The fitting function is constructed with CBFPs expressed in cylindrical Bessel functions, which appears to be an efficient approximation of the active element pattern of the circular patch antenna. The effectiveness of the proposed technique is evaluated through simulation and experiment.

Chapter 5 is the conclusions and summarizes the results obtained in this dissertation.

³This chapter is based on "Axial Ratio Improvement for Wide-angle Scanning Circularly-Polarized Phased Array of Circular Patch Antennas" [1.113], by the same author, which was submitted to IEICE Trans. Commun., Copyright(C)2024 IEICE.

Reference

- [1.1] A. Al-Fuqaha, M. Guizani, M. Mohammadi, M. Aledhari and M. Ayyash, “Internet of Things: A Survey on Enabling Technologies, Protocols, and Applications,” *IEEE Commun. Surveys & Tutorials*, vol. 17, no. 4, pp. 2347–2376, Fourthquarter 2015.
- [1.2] W. Ejaz, M. Naeem, A. Shahid, A. Anpalagan and M. Jo, “Efficient Energy Management for the Internet of Things in Smart Cities,” *IEEE Commun. Mag.*, vol. 55, no. 1, pp. 84–91, January 2017.
- [1.3] X. Liu and N. Ansari, “Toward Green IoT: Energy Solutions and Key Challenges,” *IEEE Commun. Mag.*, vol. 57, no. 3, pp. 104–110, March 2019.
- [1.4] D. S. Gurjar, H. H. Nguyen and H. D. Tuan, “Wireless Information and Power Transfer for IoT Applications in Overlay Cognitive Radio Networks,” *IEEE Internet of Things J.*, vol. 6, no. 2, pp. 3257–3270, April 2019.
- [1.5] N. Shinohara, “History and Innovation of Wireless Power Transfer via Microwaves,” *IEEE J. Microw.*, vol. 1, no. 1, pp. 218–228, Jan. 2021.
- [1.6] Powercast Corporation, <https://www.powercastco.com/>, accessed June 1. 2024.
- [1.7] A. Hajimiri, B. Abiri, F. Bohn, M. Gal-Katziri and M. H. Manohara, “Dynamic Focusing of Large Arrays for Wireless Power Transfer and Beyond,” *IEEE J. Solid-State Circuits*, vol. 56, no. 7, pp. 2077–2101, July 2021.
- [1.8] K. W. Choi, A. A. Aziz, D. Setiawan, N. M. Tran, L. Ginting and D. I. Kim, “Distributed Wireless Power Transfer System for Internet of Things Devices,” *IEEE Internet of Things J.*, vol. 5, no. 4, pp. 2657–2671, Aug. 2018.
- [1.9] Ossia Inc, <https://www.ossia.com/>, accessed June 1. 2024.
- [1.10] GuRu Wireless, <https://guru.inc/technology/>, accessed June 1. 2024.
- [1.11] Xiaomi, “Mi Air Charge Technology”, <https://www.mi.com/global/discover/article?id=1653>, accessed June 1. 2024.
- [1.12] National Weather Service, “Radar” ,National Oceanic and Atmospheric Administration, <https://www.weather.gov/about/radar>, accessed June 1. 2024.
- [1.13] Q. Cao, M. B. Yeary and G. Zhang, “Efficient Ways to Learn Weather Radar Polarimetry,” *IEEE Trans. Educ.*, vol. 55, no. 1, pp. 58–68, Feb. 2012.
- [1.14] C. D. Curtis, M. Yeary and J. L. Lake, “Adaptive Nullforming to Mitigate Ground Clutter on the National Weather Radar Testbed Phased Array Radar,” *IEEE Trans. Geosci. Remote Sens.*, vol. 54, no. 3, pp. 1282–1291, March 2016.

- [1.15] H. Kikuchi, E. Yoshikawa, T. Ushio, F. Mizutani and M. Wada, “Application of Adaptive Digital Beamforming to Osaka University Phased Array Weather Radar,” *IEEE Trans. Geosci. Remote Sens.*, vol. 55, no. 7, pp. 3875–3884, July 2017.
- [1.16] R. D. Palmer et al., “Horus—A Fully Digital Polarimetric Phased Array Radar for Next-Generation Weather Observations,” *IEEE Trans. Radar Syst.*, vol. 1, pp. 96–117, 2023.
- [1.17] S. Riendeau and C. Grenier, “RADARSAT-2 Antenna,” 2007 IEEE Aerospace Conf., 2007.
- [1.18] F. Rostan, S. Riegger, W. Pitz, A. Torre and R. Torres, “The C-SAR instrument for the GMES sentinel-1 mission,” 2007 IEEE Int. Geosci. and Remote Sens Symp., pp. 215–218, July 2007.
- [1.19] W. Pitz and D. Miller, “The TerraSAR-X Satellite,” *IEEE Trans. Geosci. Remote Sens.*, vol. 48, no. 2, pp. 615–622, Feb. 2010.
- [1.20] T. Motohka, Y. Kankaku, S. Miura and S. Suzuki, “Overview of ALOS-2 and ALOS-4 L-band SAR,” 2021 IEEE Radar Conf. (RadarConf21), May 2021.
- [1.21] T. Motohka, Y. Kankaku, S. Miura and S. Suzuki, “Alos-4 L-Band SAR Mission and Observation,” *IGARSS 2019 - 2019 IEEE Int. Geosci and Remote Sens. Symp.*, pp. 5271–5273, Aug. 2019.
- [1.22] M. Zimmermanns and C. Roemer, “Copernicus HPCM: ROSE-L SAR Instrument and Performance Overview,” *EUSAR 2022; 14th Eur. Conf. on Synthetic Aperture Radar*, July 2022.
- [1.23] C. Fulton, M. Yearly, D. Thompson, J. Lake and A. Mitchell, “Digital Phased Arrays: Challenges and Opportunities,” *Proc. IEEE*, vol. 104, no. 3, pp. 487–503, March 2016.
- [1.24] Boeing, “Boeing Ships Eutelsat’s e-BIRD Satellite To French Guiana For Late August Launch”, <https://boeing.mediaroom.com/2003-08-05-Boeing-Ships-Eutelsats-e-BIRD-Satellite-To-French-Guiana-For-Late-August-Launch>, accessed June 1. 2024.
- [1.25] G. Peters, “Satellite delivery of next generation broadband access to the UK,” 2010 5th Advanced Satell. Multimedia Syst. Conf. and the 11th Signal Process. for Space Commun. Workshop, pp. 141–146, Sept. 2010.
- [1.26] O. Kodheli et al., “Satellite Communications in the New Space Era: A Survey and Future Challenges,” *IEEE Commun. Surveys & Tutorials*, vol. 23, no. 1, pp. 70–109, Firstquarter 2021.
- [1.27] BAE Systems, “Satellite Communications (SATCOM) Antennas”, <https://www.baesystems.com/en-us/products/satcom-antennas>, accessed June 1. 2024.

- [1.28] F. Klefenz, F. Bongard and M. C. Vigano, “Flat panel mobility user terminals for Ka-band GEO/NGSO broad band satellite access,” 2023 17th Eur. Conf. Antennas Propag. (EuCAP), March 2023.
- [1.29] S. Sabripour, J. Haque, A. Ciszmar and T. Magesacher, “A COTS-based software-defined communication system platform and applications in LEO,” 36th Int. Commun. Satell. Syst. Conf. (ICSSC 2018), Oct. 2018.
- [1.30] D. Sikri and R. M. Jayasuriya, “Multi-Beam phased array with full digital beam-forming for SATCOM and 5G,” *Microw. J.*, vol. 62, pp. 64–79, 2019.
- [1.31] P. Jaffe and J. McSpadden, “Energy Conversion and Transmission Modules for Space Solar Power,” *Proc. IEEE*, vol. 101, no. 6, pp. 1424–1437, June 2013.
- [1.32] C. T. Rodenbeck et al., “Microwave and Millimeter Wave Power Beaming,” *IEEE J. Microw.*, vol. 1, no. 1, pp. 229–259, Jan. 2021.
- [1.33] X. Luo et al., “A Scalable Ka-Band 1024-Element Transmit Dual-Circularly-Polarized Planar Phased Array for SATCOM Application,” *IEEE Access*, vol. 8, pp. 156084–156095, 2020.
- [1.34] J. -C. S. Chieh et al., “Development of Flat Panel Active Phased Array Antennas Using 5G Silicon RFICs at Ku- and Ka-Bands,” *IEEE Access*, vol. 8, pp. 192669–192681, 2020.
- [1.35] K. K. W. Low, S. Zahir, T. Kanar and G. M. Rebeiz, “A 27–31-GHz 1024-Element Ka-Band SATCOM Phased-Array Transmitter With 49.5-dBW Peak EIRP, 1-dB AR, and $\pm 70^\circ$ Beam Scanning,” *IEEE Trans. Microw. Theory Techn.*, vol. 70, no. 3, pp. 1757–1768, March 2022.
- [1.36] F. Ahmed, K. Singh and K. P. Esselle, “State-of-the-Art Passive Beam-Steering Antenna Technologies: Challenges and Capabilities,” *IEEE Access*, vol. 11, pp. 69101–69116, 2023.
- [1.37] S. V. Hum and J. Perruisseau-Carrier, “Reconfigurable Reflectarrays and Array Lenses for Dynamic Antenna Beam Control: A Review,” *IEEE Trans. Antennas Propag.*, vol. 62, no. 1, pp. 183–198, Jan. 2014.
- [1.38] S. V. Hum, M. Okoniewski and R. J. Davies, “Modeling and Design of Electronically Tunable Reflectarrays,” *IEEE Trans. Antennas Propag.*, vol. 55, no. 8, pp. 2200–2210, Aug. 2007.
- [1.39] F. Venneri, S. Costanzo and G. Di Massa, “Design and Validation of a Reconfigurable Single Varactor-Tuned Reflectarray,” *IEEE Trans. Antennas Propag.*, vol. 61, no. 2, pp. 635–645, Feb. 2013.

- [1.40] S. Bildik, S. Dieter, C. Fritzsche, W. Menzel and R. Jakoby, “Reconfigurable Folded Reflectarray Antenna Based Upon Liquid Crystal Technology,” *IEEE Trans. Antennas Propag.*, vol. 63, no. 1, pp. 122–132, Jan. 2015.
- [1.41] J. Rodriguez-Zamudio, J. I. Martinez-Lopez, J. Rodriguez-Cuevas and A. E. Martynyuk, “Reconfigurable Reflectarrays Based on Optimized Spiraphase-Type Elements,” *IEEE Trans. Antennas Propag.*, vol. 60, no. 4, pp. 1821–1830, April 2012.
- [1.42] X. Yang et al., “A Broadband High-Efficiency Reconfigurable Reflectarray Antenna Using Mechanically Rotational Elements,” *IEEE Trans. Antennas Propag.*, vol. 65, no. 8, pp. 3959–3966, Aug. 2017.
- [1.43] M. H. Dahri, M. I. Abbasi, M. H. Jamaluddin and M. R. Kamarudin, “A Review of High Gain and High Efficiency Reflectarrays for 5G Communications,” *IEEE Access*, vol. 6, pp. 5973–5985, 2018.
- [1.44] M. Imbert, A. Papió, F. De Flaviis, L. Jofre and J. Romeu, “Design and Performance Evaluation of a Dielectric Flat Lens Antenna for Millimeter-Wave Applications,” *IEEE Antennas Wireless Propag. Lett.*, vol. 14, pp. 342–345, 2015.
- [1.45] M. U. Afzal and K. P. Esselle, “Steering the Beam of Medium-to-High Gain Antennas Using Near-Field Phase Transformation,” *IEEE Trans. Antennas Propag.*, vol. 65, no. 4, pp. 1680–1690, April 2017.
- [1.46] J. B. L. Rao, D. P. Patel and V. Krichevsky, “Voltage-controlled ferroelectric lens phased arrays,” *IEEE Trans. Antennas Propag.*, vol. 47, no. 3, pp. 458–468, March 1999.
- [1.47] T. Jiang et al., “Low-DC Voltage-Controlled Steering-Antenna Radome Utilizing Tunable Active Metamaterial,” *IEEE Trans. Microw. Theory Techn.*, vol. 60, no. 1, pp. 170–178, Jan. 2012.
- [1.48] W. W. Milroy, “Continuous transverse stub element devices and methods of making same,” U.S. Patent 5 266 961, Nov. 1993.
- [1.49] T. W. Bradley et al., “Development of a voltage-variable dielectric (VVD), electronic scan antenna,” *Radar 97 (Conf. Publ. No. 449)*, pp. 383–385, Oct. 1997.
- [1.50] M. F. Iskander, Z. Yun, Z. Zhang, R. Jensen and S. Redd, “Design of a low-cost 2-D beam-steering antenna using ferroelectric material and CTS technology,” *IEEE Trans. Microw. Theory Techn.*, vol. 49, no. 5, pp. 1000–1003, May 2001.
- [1.51] W. W. Milroy, S. B. Coppedge, and A. C. Lemons, “Variable inclination continuous transverse stub array,” U.S. Patent 6 919 854 B2, Jul. 2005.

- [1.52] T. Sikina, D. McKay, K. Komisarek and B. Porter, “Variably inclined continuous transverse stub-2 antenna,” *IEEE Int. Symp. on Phased Array Syst. Technol.*, pp. 435–440, Oct. 2003.
- [1.53] W. H. Henderson and W. W. Milroy, “Wireless communication applications of the continuous transverse stub (CTS) array at microwave and millimeter wave frequencies,” *IEEE/ACES Int. Conf. Wireless Commun. Applied Comput. Electromagn.*, pp. 253–256, April 2005.
- [1.54] Q. Tang et al., “Flat-Panel Mechanical Beam Steerable Array Antennas With In-Plane Rotations: Theory, Design and Low-Cost Implementation,” *IEEE Open J. Antennas Propag.*, vol. 2, pp. 679–688, 2021.
- [1.55] T. Lou, X. -X. Yang, Q. -D. Cao and S. Gao, “A Low Profile Circularly Polarized Beam Scanning Patch Array Fed by Parallel Plate Waveguide,” *IEEE Trans. Antennas Propag.*, vol. 70, no. 9, pp. 7384–7392, Sept. 2022.
- [1.56] A. Wang, L. Yang, X. Yi, X. Li, J. Zhao, and Y. Zhang, “Wireless communication applications of the variable inclination continuous transverse stub array for Ku-band applications,” *IET Microw., Antennas Propag.*, vol. 15, no. 6, pp. 644–652, Mar. 2021.
- [1.57] S. Y. Liu, C. F. Zhou, J. Liu, J. H. Fu, Q. Wu and X. M. Ding, “Variable Inclination Continuous Transverse Stub Array-Based Beam Steering Antenna System for Vehicle-to-Satellite Communication,” *IEEE Open J. Antennas Propag.*, vol. 3, pp. 1218–1226, 2022.
- [1.58] P. Checcacci, V. Russo and A. Scheggi, “Holographic antennas,” *IEEE Trans. Antennas Propag.*, vol. 18, no. 6, pp. 811–813, Nov. 1970.
- [1.59] B. H. Fong, J. S. Colburn, J. J. Ottusch, J. L. Visher and D. F. Sievenpiper, “Scalar and Tensor Holographic Artificial Impedance Surfaces,” *IEEE Trans. Antennas Propag.*, vol. 58, no. 10, pp. 3212–3221, Oct. 2010.
- [1.60] S. Pandi, C. A. Balanis and C. R. Birtcher, “Design of Scalar Impedance Holographic Metasurfaces for Antenna Beam Formation With Desired Polarization,” *IEEE Trans. Antennas Propag.*, vol. 63, no. 7, pp. 3016–3024, July 2015.
- [1.61] Yurduseven et al., “Dynamically Reconfigurable Holographic Metasurface Aperture for a Mills-Cross Monochromatic Microwave Camera,” *Opt. Express*, vol. 26, no. 5, pp. 5281–91, Feb. 2018
- [1.62] S. Venkatesh, X. Lu, H. Saeidi, and K. Sengupta, “A high-speed programmable and scalable terahertz holographic metasurface based on tiled CMOS chips,” *Nature Electron.*, vol. 3, no. 12, pp. 1–9, Dec. 2020.

- [1.63] M. C. Johnson, S. L. Brunton, N. B. Kundtz and J. N. Kutz, “Sidelobe Canceling for Reconfigurable Holographic Metamaterial Antenna,” *IEEE Trans. Antennas Propag.*, vol. 63, no. 4, pp. 1881–1886, April 2015.
- [1.64] R. Stevenson, M. Sazegar, A. Bily, M. Johnson and N. Kundtz, “Metamaterial surface antenna technology: Commercialization through diffractive metamaterials and liquid crystal display manufacturing,” 2016 10th Int. Congress on Advanced Electromagnetic Materials in Microwaves and Optics (METAMATERIALS), pp. 349–351, Sept. 2016.
- [1.65] P. Liu, Y. Li and Z. Zhang, “Circularly Polarized 2 Bit Reconfigurable Beam-Steering Antenna Array,” *IEEE Trans. Antennas Propag.*, vol. 68, no. 3, pp. 2416–2421, March 2020.
- [1.66] Z. X. Wang, H. Q. Yang, F. Zhai, J. W. Wu, Q. Cheng and T. J. Cui, “A Low-Cost and Low-Profile Electronically Programmable Bit Array Antenna for Two-Dimensional Wide-Angle Beam Steering,” *IEEE Trans. Antennas Propag.*, vol. 71, no. 1, pp. 342–352, Jan. 2023.
- [1.67] Y. -L. Yao, F. -S. Zhang and F. Zhang, “A New Approach to Design Circularly Polarized Beam-Steering Antenna Arrays Without Phase Shift Circuits,” *IEEE Trans. Antennas Propag.*, vol. 66, no. 5, pp. 2354–2364, May 2018.
- [1.68] S. H. Talisa, K. W. O’Haver, T. M. Comberiate, M. D. Sharp and O. F. Somerlock, “Benefits of Digital Phased Array Radars,” *Proc. IEEE*, vol. 104, no. 3, pp. 530–543, March 2016.
- [1.69] K. Roth, H. Pirzadeh, A. L. Swindlehurst and J. A. Nossek, “A Comparison of Hybrid Beamforming and Digital Beamforming With Low-Resolution ADCs for Multiple Users and Imperfect CSI,” *IEEE J. Selected Topics in Signal Process.*, vol. 12, no. 3, pp. 484–498, June 2018.
- [1.70] J. Zhang, X. Yu and K. B. Letaief, “Hybrid Beamforming for 5G and Beyond Millimeter-Wave Systems: A Holistic View,” *IEEE Open J. Commun. Soc.*, vol. 1, pp. 77–91, 2020.
- [1.71] T. Kebede, Y. Wondie, J. Steinbrunn, H. B. Kassa and K. T. Kornegay, “Precoding and Beamforming Techniques in mmWave-Massive MIMO: Performance Assessment,” *IEEE Access*, vol. 10, pp. 16365–16387, 2022.
- [1.72] M. A. Johnson, “Phased-array beam steering by multiplex sampling,” *Proc. IEEE*, vol. 56, no. 11, pp. 1801–1811, Nov. 1968.
- [1.73] J. D. Fredrick, Yuanxun Wang and T. Itoh, “A smart antenna receiver array using a single RF channel and digital beamforming,” *IEEE Trans. Microw. Theory Techn.*, vol. 50, no. 12, pp. 3052–3058, Dec. 2002.

- [1.74] G. Krishnamurthy and K. G. Gard, “Time Division Multiplexing Front-Ends for Multiantenna Integrated Wireless Receivers,” *IEEE Trans. Circuits Syst. I: Regular Papers*, vol. 57, no. 6, pp. 1231–1243, June 2010.
- [1.75] A. Jahanian, F. Tzeng and P. Heydari, “Code-modulated path-sharing multi-antenna receivers: theory and analysis,” *IEEE Trans. Wireless Commun.*, vol. 8, no. 5, pp. 2193–2201, May 2009.
- [1.76] J. Zhang, W. Wu and D. -G. Fang, “Single RF Channel Digital Beamforming Multi-beam Antenna Array Based on Time Sequence Phase Weighting,” *IEEE Antennas Wireless Propag. Lett.*, vol. 10, pp. 514–516, 2011.
- [1.77] N. Akram et al., “Frequency-Multiplexed Array Digitization for MIMO Receivers: 4-Antennas/ADC at 28 GHz on Xilinx ZCU-1285 RF SoC,” *IEEE Access*, vol. 9, pp. 142743–142753, 2021.
- [1.78] F. Tzeng, A. Jahanian, D. Pi and P. Heydari, “A CMOS Code-Modulated Path-Sharing Multi-Antenna Receiver Front-End,” *IEEE J. Solid-State Circuits*, vol. 44, no. 5, pp. 1321–1335, May 2009.
- [1.79] S. Henault, B. R. Jackson and Y. M. M. Antar, “Compensation of Time-Division Multiplexing Distortion in Switched Antenna Arrays With a Single RF Front-End and Digitizer,” *IEEE Trans. Antennas Propag.*, vol. 61, no. 8, pp. 4383–4388, Aug. 2013.
- [1.80] W. Ahmad, H. Zhang, Y. Chen and N. Iqbal, “Low RF-Complexity Digital Transmit Beamforming for Large-Scale Millimeter Wave MIMO Systems,” *IEEE Trans. Wireless Commun.*, vol. 21, no. 10, pp. 8308–8321, Oct. 2022.
- [1.81] S. Bojja Venkatakrishnan, D. K. Papantonis, A. A. Akhiyat, E. A. Alwan and J. L. Volakis, “Experimental Validation of On-Site Coding Digital Beamformer With Ultra-Wideband Antenna Arrays,” *IEEE Trans. Microw. Theory Techn.*, vol. 65, no. 11, pp. 4408–4417, Nov. 2017.
- [1.82] H. Rafati and B. Razavi, “A Receiver Architecture for Dual-Antenna Systems,” *IEEE J. Solid-State Circuits*, vol. 42, no. 6, pp. 1291–1299, June 2007.
- [1.83] K. Ullah, S. B. Venkatakrishnan and J. L. Volakis, “RFSoc-FPGA Realization of a Code-Multiplexed Digital Receiver (CMDR) Using 1-ADC/Quad-Channel,” *IEEE J. Microw.*, vol. 4, no. 1, pp. 123–138, Jan. 2024.
- [1.84] Z. Fang et al., “DOA Estimation Method Based on Space-Time Coding Antennas With Orthogonal Codes,” *IEEE Trans. Antennas Propag.*, vol. 72, no. 2, pp. 1173–1181, Feb. 2024.

- [1.85] E. A. Alwan, A. Akhiyat, M. LaRue, W. Khalil and J. L. Volakis, "Low cost, power efficient, on-site coding receiver (OSCR) for ultra-wideband digital beamforming," 2013 IEEE Int. Symp. on Phased Array Syst. Technol., pp. 202–206, Oct. 2013,
- [1.86] X. Wang and C. Caloz, "Direction-of-Arrival (DOA) Estimation based on Spacetime-Modulated Metasurface," 2019 IEEE Int. Symp. Antennas Propag. USNC-URSI Radio Sci. Meeting, pp. 1613–1614, July 2019
- [1.87] Q. Y. Zhou et al., "Two-dimensional direction-of-arrival estimation based on time-domain-coding digital metasurface," *Appl. Phys. Lett.*, vol. 121, no. 18, Art. no. 181702, Oct. 2022.
- [1.88] X. Wang and C. Caloz, "Spacetime-Modulated Metasurface for Spatial Multiplexing Communication," 2019 Thirteenth Int. Congr. Artificial Materials for Novel Wave Phenomena (Metamaterials), pp. X-465–X-467, Sept. 2019.
- [1.89] T. Takahashi, "Statistical radiation pattern analysis and calibration technologies for phased array antennas," *IEICE Trans. Commun.*, vol. J100-B, no. 9, pp. 748–763, Sept. 2017 (in Japanese).
- [1.90] G. He, X. Gao and R. Zhang, "Impact Analysis and Calibration Methods of Excitation Errors for Phased Array Antennas," *IEEE Access*, vol. 9, pp. 59010–59026, 2021.
- [1.91] H. Yamada, "Array Calibration Technique for High-Resolution Direction of Arrival Estimation," *IEICE Trans. Commun.*, vol. J92-B, no. 9, pp. 1308–1321, Sept. 2009 (in Japanese).
- [1.92] H. M. Aumann, A. J. Fenn and F. G. Willwerth, "Phased array antenna calibration and pattern prediction using mutual coupling measurements," *IEEE Trans. Antennas Propag.*, vol. 37, no. 7, pp. 844–850, July 1989.
- [1.93] T. Takahashi et al., "On-Board Calibration Methods for Mechanical Distortions of Satellite Phased Array Antennas," *IEEE Trans. Antennas Propag.*, vol. 60, no. 3, pp. 1362–1372, March 2012.
- [1.94] H. -T. Chou and D. -Y. Cheng, "Beam-Pattern Calibration in a Realistic System of Phased-Array Antennas via the Implementation of a Genetic Algorithm With a Measurement System," *IEEE Trans. Antennas Propag.*, vol. 65, no. 2, pp. 593–601, Feb. 2017.
- [1.95] Z. Wang, F. Zhang, H. Gao, O. Franek, G. F. Pedersen and W. Fan, "Over-the-Air Array Calibration of mmWave Phased Array in Beam-Steering Mode Based on Measured Complex Signals," *IEEE Trans. Antennas Propag.*, vol. 69, no. 11, pp. 7876–7888, Nov. 2021.

- [1.96] H. M. Aumann and F. G. Willwerth, “Phased-array calibration by adaptive nulling,” MIT Lincoln Lab., Cambridge, MA, USA, Tech. Rep. 915, 1991.
- [1.97] H. M. Aumann and F. G. Willwerth, “Phased array calibrations using measured element patterns,” IEEE Antennas Propag. Soc. Int. Symp. 1995 Digest, vol. 2, pp. 918–921, June 1995.
- [1.98] N. Nakamoto, T. Takahashi, M. Otsuka, and H. Miyashita, “Phase-Only Synthesis for Phased Array Antenna with Element Field Errors at Every Phase Shift,” IEICE Trans. Commun., vol. J98-B, no. 12, pp. 1255–1265, Dec. 2015. (in Japanese).
- [1.99] Y. -S. Chen and I. -L. Tsai, “Detection and Correction of Element Failures Using a Cumulative Sum Scheme for Active Phased Arrays,” IEEE Access, vol. 6, pp. 8797–8809, 2018.
- [1.100] B. P. Ng, J. P. Lie, M. H. Er and A. Feng, “A Practical Simple Geometry and Gain/Phase Calibration Technique for Antenna Array Processing,” IEEE Trans. Antennas Propag., vol. 57, no. 7, pp. 1963–1972, July 2009.
- [1.101] P. Heidenreich, A. M. Zoubir and M. Rubsamén, “Joint 2-D DOA Estimation and Phase Calibration for Uniform Rectangular Arrays,” IEEE Trans. Signal Process., vol. 60, no. 9, pp. 4683–4693, Sept. 2012.
- [1.102] I. Chiba, M. Ohtsuka, S. Matsumoto, and T. Katagi, “Phase controlling for improving the gain or providing the low cross-polarization level in circularly polarized phased array antennas,” IEICE Trans. Commun., vol. J75-B-II, no.7, pp.456–463, July 1992 (in Japanese).
- [1.103] C. Fulton et al., “Cylindrical polarimetric phased array radar: Beamforming and calibration for weather applications,” IEEE Trans. Geosci. Remote Sens., vol. 55, no. 5, pp. 2827–2841, May 2017.
- [1.104] A. H. Aljuhani, T. Kanar, S. Zahir and G. M. Rebeiz, “A 256-Element Ku-Band Polarization Agile SATCOM Receive Phased Array With Wide-Angle Scanning and High Polarization Purity,” IEEE Trans. Microw. Theory Techn., vol. 69, no. 5, pp. 2609–2628, May 2021.
- [1.105] J. -W. Kim, S. -C. Chae, H. -W. Jo, T. -D. Yeo and J. -W. Yu, “Wideband Circularly Polarized Phased Array Antenna System for Wide Axial Ratio Scanning,” IEEE Trans. Antennas Propag., vol. 70, no. 2, pp. 1523–1528, Feb. 2022.
- [1.106] T. Jaschke and A. F. Jacob, “A dual-polarized SIW lens antenna array for Rx-/Tx-integration at K/Ka-band,” IEEE Trans. Antennas Propag., vol. 71, no. 3, pp. 2443–2453, March 2023.
- [1.107] D. H. Werner and R. J. Allard, “The simultaneous interpolation of antenna radiation patterns in both the spatial and frequency domains using model-based parameter

- estimation,” *IEEE Trans. Antennas Propag.*, vol. 48, no. 3, pp. 383–392, March 2000.
- [1.108] M. Landmann and G. Del Galdo, “Efficient antenna description for MIMO channel modelling and estimation,” *7th Eur. Conf. on Wireless Technol.*, pp. 217–220, Oct. 2004.
- [1.109] R. J. Allard and D. H. Werner, “The model-based parameter estimation of antenna radiation patterns using windowed interpolation and spherical harmonics,” *IEEE Trans. Antennas Propag.*, vol. 51, no. 8, pp. 1891–1906, Aug. 2003.
- [1.110] R. Maaskant, M. V. Ivashina, S. J. Wijnholds and K. F. Warnick, “Efficient Prediction of Array Element Patterns Using Physics-Based Expansions and a Single Far-Field Measurement,” *IEEE Trans. Antennas Propag.*, vol. 60, no. 8, pp. 3614–3621, Aug. 2012.
- [1.111] N. Nakamoto et al. , “Radial Line Planar Phased Array Using Electromechanically Rotated Helical Antennas,” *IEICE Trans. Commun.*, vol. E106-B, no.2, pp.174–183, Feb. 2023.
- [1.112] N. Nakamoto, K. Kihira, T. Fukasawa, Y. Inasawa, and N. Shinohara, “Waveguide slot array with code-division multiplexing function for single RF chain digital beamforming,” *IEICE Trans. Commun.*, vol. E107-B, no. 8, pp. 541–551, Aug. 2024.
- [1.113] N. Nakamoto, H. Watanabe, T. Takahashi, Y. Inasawa, and N. Shinohara, “Axial Ratio Improvement for Wide-angle Scanning Circularly-Polarized Phased Array of Circular Patch Antennas,” *IEICE Trans. Commun.*, submitted for publication.

Chapter 2

Radial Line Planar Phased Array Using Electromechanically Rotated Helical Antennas

2.1 Introduction

Radial line planar arrays (RLPAs) with probe-fed antenna elements [2.1]–[2.4] have been extensively used in various applications such as direct broadcast satellite [2.3]–[2.5], satellite communications [2.6], radar system [2.7], and high-power microwave systems [2.8] because of their high gain with high efficiency, low profile, lightweight, and high power capacity. RLPAs comprise two parallel metallic plates, antenna elements on the top plate, and probes inserted into the waveguide. Thus, RLPAs have very low losses in their waveguides and typically offer high antenna efficiencies of 60%–90%. Recently, the demand for beam-scanning antennas or phased array antennas has been increasing with the rapid development of wireless communication systems. In an RLPA with probe-fed circularly polarized antenna elements, beam-steering can be usually accomplished by axially rotating each antenna element [2.1], [2.3], [2.5], [2.9]. In previous works, antenna elements were manually rotated and beam-tilting in fixed directions was demonstrated. In [2.5], the entire array structure was rotated by a motor to scan the beam, but only conical beam-scanning was achieved. To the best of author’s knowledge, an RLPA for dynamic and electrically controlled two-dimensional beam-scanning has not been shown in previous works.

Moreover, with the advancement of mechanical controlling devices such as miniaturized micromotors, phased arrays using electromechanically rotated circularly polarized antenna elements have been reported mainly for reflectarray antennas [2.10]–[2.12]. Recently, a beam-scanning helical array fed by rectangular waveguides, in which helices are axially rotated by their individual micromotors, was reported [2.13], [2.14]. In this study, a similar idea is applied to RLPA, and a phased array using electromechanically controlled

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helical antenna elements fed from a radial line is proposed. The proposed antenna has a micromotor connected to each individual helix, which is axially rotated individually for two-dimensional beam-scanning. To our knowledge, this is the first radial line planar phased array that uses antenna elements electromechanically rotated by micromotors. The proposed antenna can be less costly than conventional active phased array antennas (APAAs) and can offer higher efficiency than conventional passive PAAs and individual phase control capabilities. The helix and probe are usually fabricated by bending a conducting wire [2.3] or wrapping a flexible substrate with a printed stripline [2.15]. When these conventional antennas are connected to a motor shaft, additional plastic shafts are required between them, which makes it harder to manufacture the antenna. In the proposed antenna, to facilitate fabrication, the helix and probe are directly metallized on the plastic shaft using molded interconnect device (MID) technology [2.16], and the motor shaft is directly press-fitted into the plastic shaft.

The probe design is critical for the RLPA design, and researchers have reported several design methodologies. In [2.1]–[2.4], the probe lengths were experimentally determined, which is time-consuming. In [2.7], a design procedure based on an analytical expression for the coupling between the feeding probe and the probe of each antenna element, in which the mutual couplings between probes in the radial line are neglected, was presented. Although this assumption was removed in [2.17]–[2.19] and more accurate analysis based on the method of moments was also presented in [2.20], the probe shape was limited to a straight monopole in these works. In [2.21], a probe with a circular disk at its end was introduced, and an improved analysis approach using an equivalent circuit model for the radial line, in which the load admittance of the probe was calculated by a field matching method, was also presented; however, the detailed design procedure of the probes was not presented. The proposed antenna uses probes fabricated via MID technology, which include dielectrics and have various shapes. Therefore, a design procedure is presented to design an RLPA with such probes. This procedure is based on an equivalent circuit model of the radial line loaded with the probes whose load admittance is calculated by full-wave analysis of the unit cell. In the proposed design procedure, the admittances for the desired aperture distribution are first determined using the equivalent circuit model and then the probe structures are designed using the full-wave analysis to realize the determined admittances. Subsequently, the excitation phases are calculated using the equivalent circuit model with the obtained admittances, and the rotation angles of the antenna elements for beam-scanning are also calculated. Thus, the proposed procedure is practical to design an RLPA using antenna elements with arbitrary probe shape without large-scale full-wave analysis of the whole structure of the RLPA.

The rest of this chapter is organized as follows. Section 2.2 explains the detailed antenna configuration; Section 2.3 describes the design procedure and design results for a

prototype antenna; Section 2.4 presents the experimental and simulated performance of the prototype antenna; finally, Section 2.5 describes the conclusions of this chapter.

2.2 Antenna Configuration

Figure 2.1 shows the configuration of the proposed antenna. The antenna comprises a radial line fed with a coaxial probe, probe-fed helical antennas, micromotors, and a motor driver/controller. The radial line has a metallic side-wall and is short-circuited at its end. The height of the radial line is selected so that the transverse electromagnetic (TEM) mode is the only propagating mode. The coaxial probe is an extension of the inner conductor of the coaxial line and is located at the center of the radial line. The helical antennas are arranged in M concentric circles. Each circle is equally spaced, and the number of helical antennas arranged in the m th circle is $N_m = 6m$ to achieve better aperture distribution [2.1], [2.9]. Thus, the radius of the m th circle R_m and the element spacings are given by

$$R_m = md_\rho \quad (2.1)$$

$$d_\phi = \frac{2\pi R_m}{N_m} = \frac{\pi d_\rho}{3} \quad (2.2)$$

where d_ρ and d_ϕ are the element spacings in the radial and circumference directions, respectively.

Each helical antenna is fed by a probe, which couples with the traveling wave in the radial line. The probe lengths are determined to realize the appropriate excitation coefficients of the array. In this study, the array is assumed to have an axially symmetric amplitude distribution, and the probe lengths are identical in each circle but are different for different circles. The helices and the probes are metallized on the surface of plastic shafts via MID technology. All plastic shafts and helical antennas in the array have identical shapes and conducting patterns, respectively; however, the probe lengths are different depending on the circle in the array. Each plastic shaft passes through the holes on the top and bottom plates of the radial line, and the motor shaft is press-fitted into each plastic shaft at its end. The holes on the bottom face have small radii to prevent the electromagnetic wave in the radial line from leaking through these holes. The motor driver/controller controls the rotation angle of each helical antenna individually to realize an appropriate rotation angle for beam-scanning.

The proposed antenna does not require many amplifiers and typical digital phase shifters, which are the primary factor of the hardware cost of the conventional APAAs. Therefore, it will offer a lower-cost alternative to the conventional APAAs (see Appendix A.1). However, the proposed antenna has the disadvantage of a slower beam scanning speed than the conventional APAAs. In contrast to the sub-microsecond switching speeds

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of typical digital phase shifters, micromotors (stepper motors) exhibit beam switching speeds that will reach at least the order of milliseconds (see Appendix A.2). Therefore, the proposed antenna is not suitable for systems that require high-speed beam scanning, such as radars.

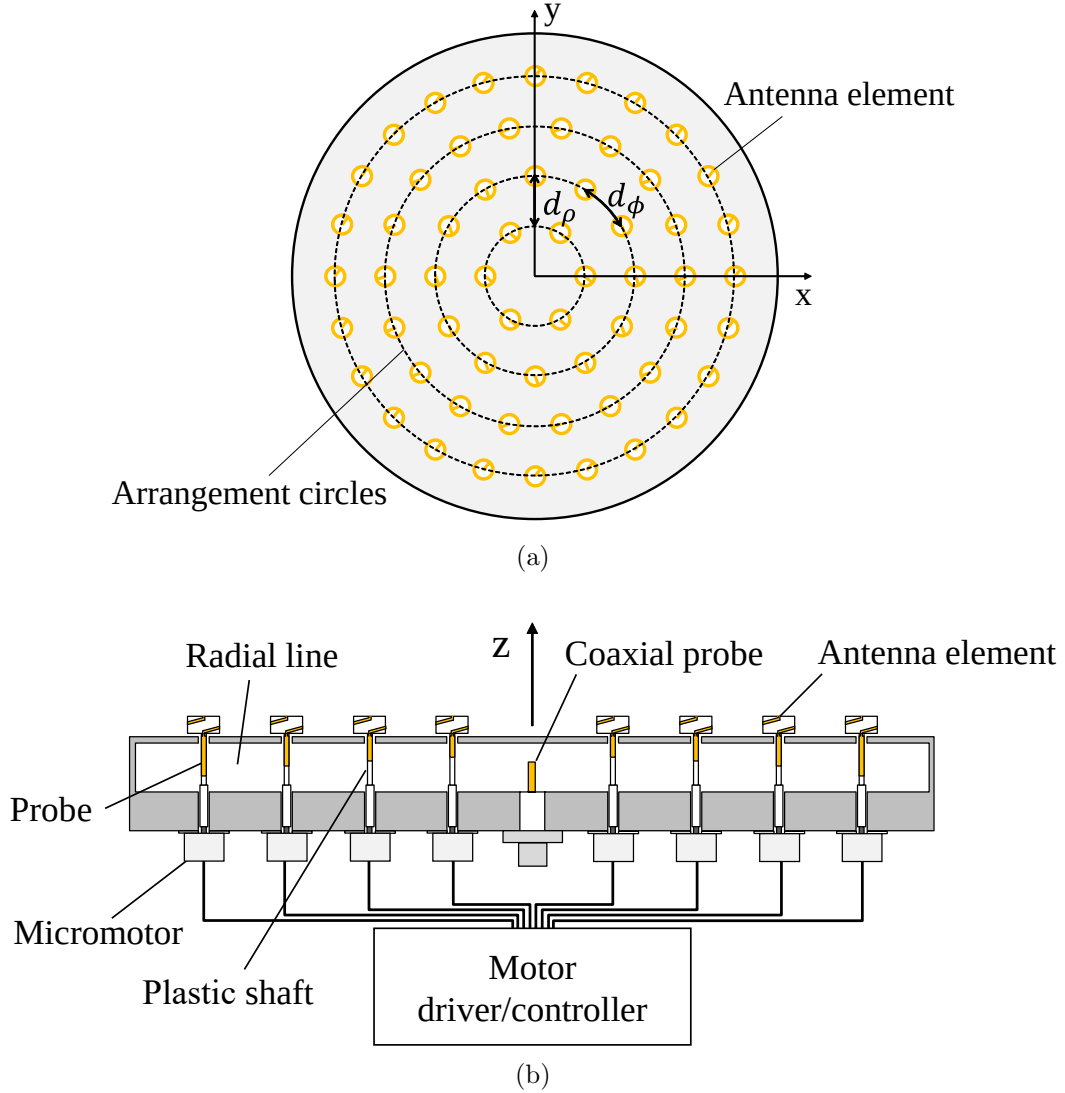


Figure 2.1: Configuration of the proposed antenna. (a) Radiating aperture and (b) cross-sectional view of the antenna.

2.3 Antenna Design

Table 2.1 lists a set of design requirements for a prototype of the proposed antenna. A 7-circle array with 168 antenna elements is designed to realize the required gain. To avoid the sidelobe level increase in the required beam-scanning coverage [2.22], the element spacings d_ρ and d_ϕ are $0.78\lambda_c = 19.2$ mm and $0.82\lambda_c = 20.1$ mm, respectively, where λ_c is the wavelength at the center frequency $f_c = 12.2$ GHz.

Table 2.1: Design Requirements for a Prototype Antenna.

Parameter	Value	Notes
Frequency band	11.7–12.7 GHz	
Coverage	$\pm 10^\circ$	Coverage without any quasi-grating lobes.
Gain	> 30 dBi	Gain in the broadside direction at f_c .
Polarization	RHCP	

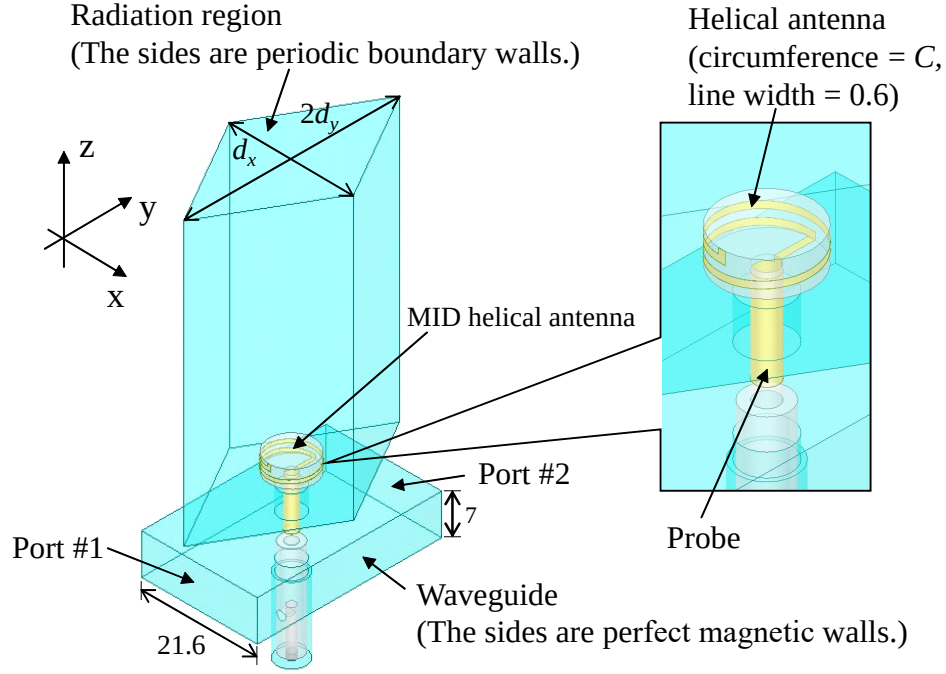
The critical parts and design parameters of an RLPA with probe-fed helical antennas to realize the desired aperture distribution and reflection characteristics are the helical antenna, probe lengths, and feeding coaxial probe. Herein, a helical antenna is first designed using an electromagnetic simulator Ansys HFSS. Next, the radiation conductances for the desired aperture distribution and impedance matching with the feeding coaxial probe are determined using an equivalent circuit model of the array; then the probe lengths are designed using HFSS to achieve the determined radiation conductances. Subsequently, the rotation angles of the antenna elements for the beam-scanning are calculated using the equivalent circuit model with the simulated load admittances. Finally, all the designed parts of the array are integrated to build a prototype array.

2.3.1 Helical Antenna Design

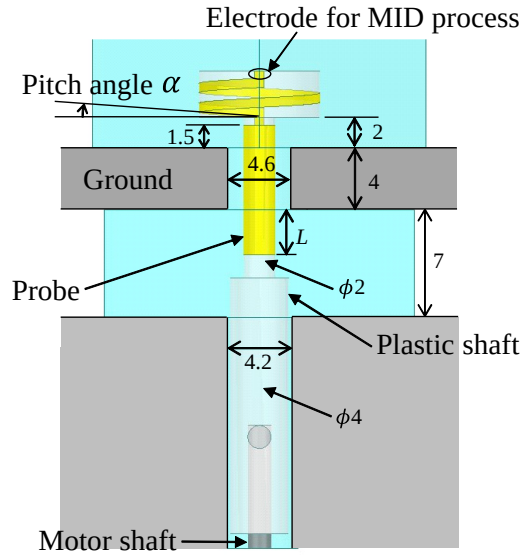
A short axial-mode helical antenna with a few turns and a small pitch angle [2.3], which is suitable for a low-profile circularly polarized array, is designed. Figure 2.2 shows a simulation model of the antenna element mounted on the radial line. The characteristics of an antenna element in an RLPA are affected by the electromagnetic couplings between neighboring antenna elements outside the radial line. A unit-cell model with periodic boundary walls is usually employed in the antenna element design for a uniformly spaced large array to consider the couplings effects between neighboring antenna elements. To apply the unit-cell model to the element design for an RLPA, which is a nonuniformly spaced array, an array of uniformly spaced elements arrayed in a triangular grid that approximates the original element positions is considered [2.20]. This approximate grid

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can be obtained by an optimization to minimize the differences from the original element positions, and the optimized triangular grid with $d_x = 21.6$ mm and $d_y = 18.7$ mm is obtained for the prototype antenna design.



(a)



(b)

Figure 2.2: Simulation model of the antenna element (dimensions in millimeter). (a) Perspective view and (b) side view of the antenna element.

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In the present model, the height of the radial line is less than $\lambda_c/2$, and the probes are assumed to be evenly distributed in the circles and only slightly perturb the field excited by the coaxial probe. In this case, the fields in the radial line are the rotationally symmetric modes TM to z , and the TEM mode is the only propagating mode [2.23]. For the prototype antenna, the height of the radial line is $0.28\lambda_c = 7.0$ mm to satisfy the above assumption. To simulate the fields in the radial line, a rectangular waveguide with perfect magnetic walls is used.

The plastic shaft is made of cyclic olefin copolymer (COC) with a dielectric constant of 2.18 and a loss tangent of 0.0006, and it has shank and head parts. The antenna is a helical conducting line with a width of 0.6 mm, which is wrapped around the plastic shaft head. One end of the conducting line is bent toward the $+z$ -direction to form an electrode for plating in the MID process. The other end of the conducting line is bent at a right angle along the bottom surface of the plastic shaft head and bent again to the $-z$ -direction to connect to the upper side of the probe at 1.5 mm above the ground plane. The circumference C , pitch angle α , and number of turns n_t of the helix are the design parameters. The probe is a conducting pattern wrapped around the plastic shaft shank and has a rotationally symmetric shape to ensure constant coupling with the radial line while being rotated. Though the interior of the probe is not filled with conductors, the electromagnetic wave cannot propagate through it because of the cut-off characteristics. Hence, the probe and the insertion hole in the top plate approximately constitute a coaxial line, whose characteristic impedance is approximately 50Ω . The probe length L is a design parameter for the radiation admittance design and is tentatively assumed to be 3 mm in the helical antenna design. The diameter of the shank enlarges to 4 mm at the bottom side to realize a hole for the press-fitting of the motor shaft whose diameter is 2 mm. The insertion hole in the bottom plate has a diameter of 4.2 mm and can be considered a dielectric-filled circular waveguide. The lowest cut-off frequency of this circular waveguide is approximately 28 GHz, and hence the electromagnetic wave in the radial line scarcely leaks from this hole.

Nakano *et al.* [2.24] have revealed that helical antennas with a pitch angle α of approximately 5° can provide a low axial ratio (AR). Therefore, the pitch angle α is taken as 5° in this design. Figure 2.3 shows the simulated AR on the z -axis with different circumferences and number of turns. The optimized helical antenna with circumference $C = 0.97\lambda_c = 23.9$ mm and the number of turns $n_t = 1.6$ has the lowest AR of 0.4 dB.

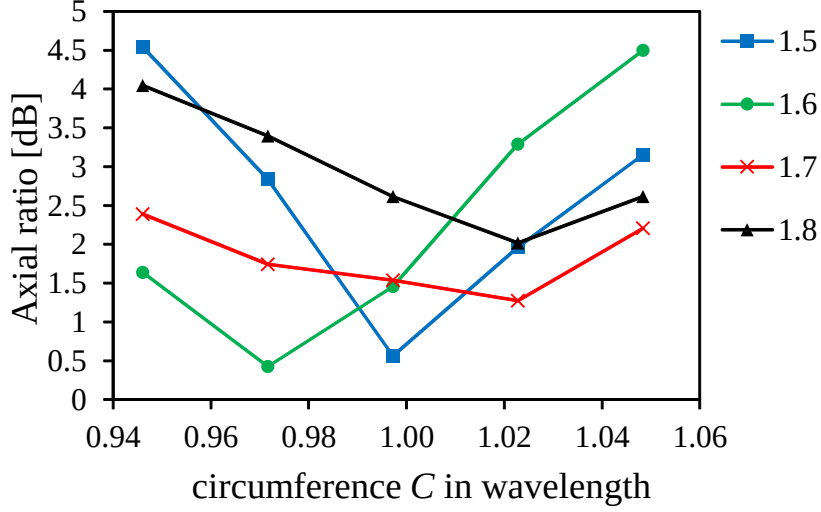


Figure 2.3: Simulated AR at $f = f_c$ with different C and n_t .

2.3.2 Radiation Conductance Design with the Equivalent Circuit Model

The probe lengths need to be appropriately adjusted to realize a desired aperture distribution. Therefore, an equivalent circuit model of the RLPA with probe-fed antenna elements is introduced to obtain a guideline for the radiation conductances for the desired aperture distribution and impedance matching.

The electromagnetic wave in the radial line is first discussed. In addition to the abovementioned assumptions, the element spacing d_ρ is assumed to be greater than $\lambda_c/2$, which is usually satisfied in practical phased arrays, so that the fields of the evanescent higher-order modes are sufficiently attenuated between the adjacent circles. The only propagating mode in the radial line is the TEM mode, whose fields are approximated to have the form of $e^{\pm jk\rho}/\sqrt{\rho}$ for $\rho > \lambda_c/2$ where k is the wavenumber and ρ is the distance from the center of the radial line to the observation point [2.23]. The field in the radial line can be analyzed using the classical transmission line theory, except it has a tapered amplitude of the form of $1/\sqrt{\rho}$.

The equivalent circuit model of the array can be represented as shown in Fig. 2.4, where Y_m^a is the active admittance of the helical antennas including the probes in the m th circle, which is normalized by $1/Z_0$ where Z_0 is the wave impedance of the radial line. In this model, the author assumes that the helical antennas are evenly distributed in the circles and have an identical active admittance in each circle but have different ones for different circles. The transmission line is short-circuited at a distance d_s from the M th circle. In this study, $d_s = 3\lambda_c/4 = 18.4$ mm to superpose the reflected wave from the termination on the forward wave in-phase at the M th circle [2.7]. Y_m^t is the

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total admittance observed from the input of the antenna element in the m th circle, which includes $Y_m^a, Y_{m+1}^a, \dots, Y_M^a$, and the short circuit. A set of Y_m^a for the desired aperture distribution is initially determined.

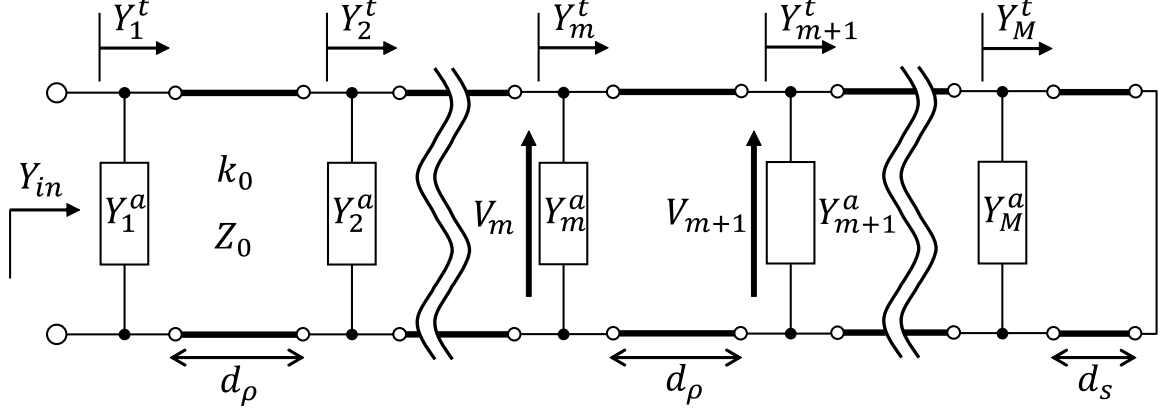


Figure 2.4: Equivalent circuit of an RLPA.

From transmission line theory, the recursive formula for Y_m^t is as follows:

$$Y_m^t = Y_m^a + \begin{cases} \frac{Y_{m+1}^t + j \tan kd_\rho}{1 + jY_{m+1}^t \tan kd_\rho} & (m < M) \\ -j \cot kd_s & (m = M). \end{cases} \quad (2.3)$$

Let V_f and V_r be the forward and reflected voltages on the antenna element in the m th circle, respectively. Considering the amplitude taper of the fields in the radial line, the voltage at a distance ρ from the input port is given by

$$V(\rho) = \frac{\sqrt{md_\rho}}{\sqrt{\rho}} [V_f e^{-jk(\rho-md_\rho)} + V_r e^{jk(\rho-md_\rho)}] \quad (2.4)$$

and the ratio between the voltages $V_m = V(md_\rho)$ and $V_{m+1} = V((m+1)d_\rho)$ is given by

$$\frac{V_{m+1}}{V_m} = \sqrt{\frac{m}{m+1}} \frac{V_f e^{-jkd_\rho} + V_r e^{jkd_\rho}}{V_f + V_r}. \quad (2.5)$$

The expression for the reflection coefficient at the antenna element in the $(m+1)$ th circle is given by

$$\Gamma_{m+1} = \frac{1 - Y_{m+1}^t}{1 + Y_{m+1}^t} = \frac{V_r e^{jkd_\rho}}{V_f e^{-jkd_\rho}} = \frac{V_r}{V_f} e^{j2kd_\rho}. \quad (2.6)$$

Using (2.6), V_f and V_r in (2.5) can be eliminated, and (2.5) can be written as

$$\frac{V_{m+1}}{V_m} = \sqrt{\frac{m}{m+1}} \frac{1}{\cos(kd_\rho) + jY_{m+1}^t \sin(kd_\rho)}. \quad (2.7)$$

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To realize the desired excitation amplitude a_m of the antenna element in the m th circle, the radiated power P_m from the corresponding antenna element should satisfy the following relationship:

$$\frac{P_m}{P_{m+1}} = \frac{Y_m^{a*} |V_m|^2}{Y_{m+1}^{a*} |V_{m+1}|^2} = \frac{a_m^2}{a_{m+1}^2}. \quad (2.8)$$

Substituting (2.7) into (2.8), a recursive expression of the radiation admittance or conductance for the desired excitation coefficients can be obtained as follows:

$$Y_m^a = \frac{a_m^2}{a_{m+1}^2} \frac{m}{m+1} \frac{Y_{m+1}^a}{\left| \cos(kd_\rho) + jY_{m+1}^t \sin(kd_\rho) \right|^2}. \quad (2.9)$$

The reflection coefficient at the input port of the array can be easily calculated from Y_1^t , and it should be zero for impedance matching, i.e.,

$$\Gamma_{in} = \frac{1 - Y_1^t}{1 + Y_1^t} = 0. \quad (2.10)$$

The radiation conductances satisfying the desired aperture distribution and impedance matching condition can be obtained by solving (2.3), (2.9), and (2.10). However, it is challenging to develop an analytical solution for Y_m^a . Consequently, a problem of finding an approximate solution that provides Γ_{in} less than a specified value is considered instead. This problem can be solved using an iterative numerical method.

Figure 2.5 shows radiation conductances calculated by the above procedure for the uniform aperture distribution and $20 \log |\Gamma_{in}| < -35$ dB. The calculated conductances range from 0.03 to 1.6 and those for the outer circles are larger.

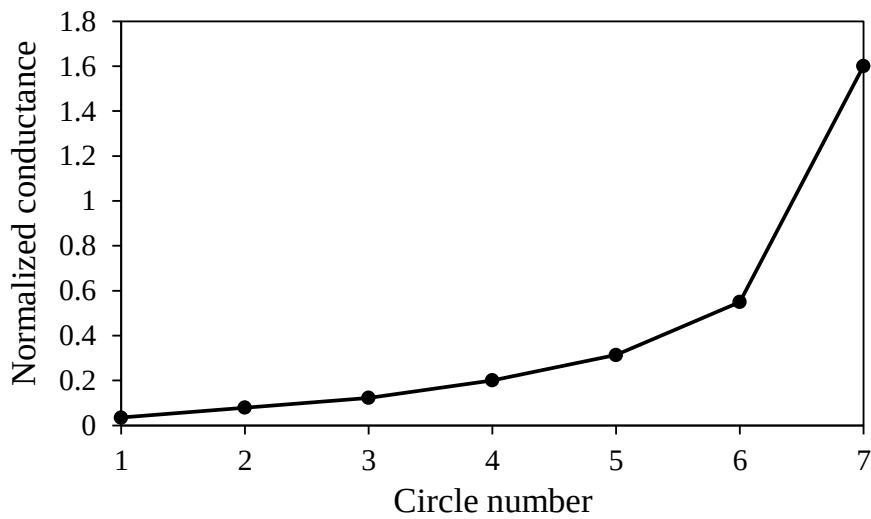


Figure 2.5: Calculated radiation conductance for the uniform aperture distribution.

2.3.3 Probe Design

Probe lengths are designed using Ansys HFSS to realize the abovementioned radiation conductances. The simulation model for the design is the same as that in Fig. 2.2, and the radiation admittance is calculated from the de-embedded S -parameters (the reference plane is set just under the probe). The antenna element can be practically represented by only a shunt admittance, which can be calculated by [2.25]

$$Y^a = \frac{-2S_{11}}{1 + S_{11}}. \quad (2.11)$$

The probe length for each circle is designed so that the calculated radiation conductance G_m^a has the predetermined values shown in Fig. 2.5. Figure 2.6 shows the designed probe lengths and the simulated admittance at the center frequency. Shorter probes have smaller radiation conductances, and the designed probe lengths range from 1.5 to 4.1 mm. In this design, the probe length in the seventh circle is chosen so that Y_7^a becomes almost required pure conductance of 1.6. Since a probe in a radial line exhibits an admittance characteristic of series resonant circuit [2.26], probes for the first to sixth circles, which are shorter than the resonant length, have capacitive susceptances B_m^a . Although these residual susceptances will disturb the aperture distribution of the array, they are permitted in this design because their effect on the aperture distribution is expected to be minimal, as will be shown later. The cancellation of these residual susceptances is one of the future works for further improvement of aperture efficiency.

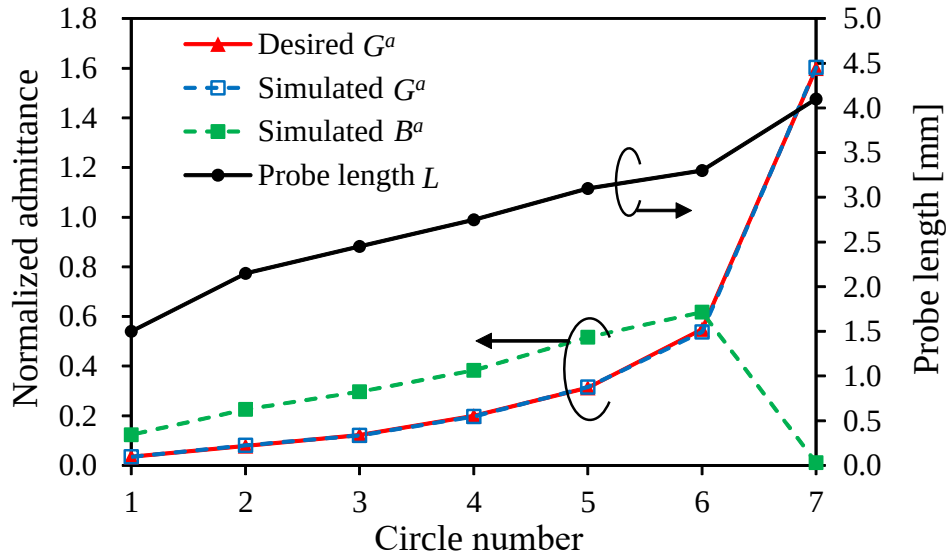


Figure 2.6: Designed probe lengths and simulated admittances for each circle.

2.3.4 Rotation Angles for Beam-scanning

When all elements in the array are installed at the same angle, the excitation amplitude a_m and phase ψ_m of the antenna elements in the m th circle can be calculated by

$$a_m^2 = \operatorname{Re} \left(\frac{1}{2} Y_m^a |V_m|^2 \right) \quad (2.12)$$

$$\psi_m = \arg(Y_m^a V_m) \quad (2.13)$$

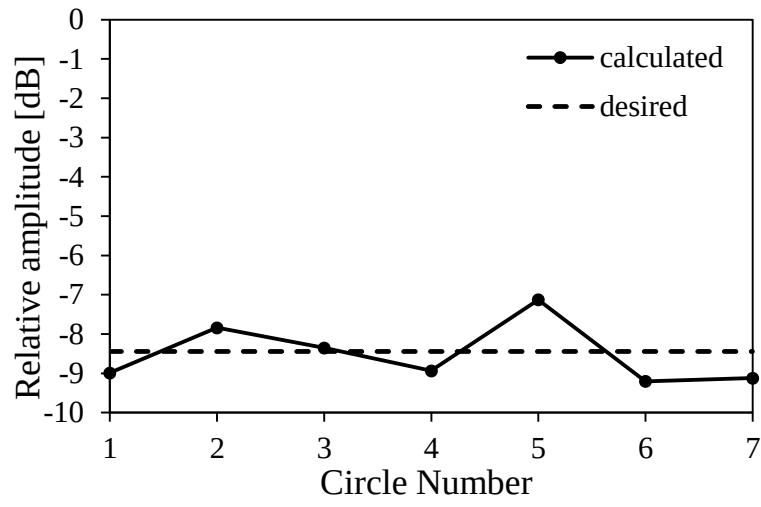
where Y_m^a denotes the calculated radiation admittance shown in Fig. 2.6 and V_m can be calculated from (2.3) and (2.7) using the calculated Y_m^a . Figure 2.7(a) and (b) show the calculated a_m and ψ_m at the center frequency, respectively. The calculated a_m values deviate from the uniform distribution, and the calculated ψ_m values include the phase shifts due to propagation in the radial line and those caused by the residual radiation susceptances.

For beam-scanning, the excitation phase shown in Fig. 2.7(b) should be compensated and the additional phase shift is set to each antenna element so that the main beam points in the direction of (θ_0, ϕ_0) , and thus the total phase shift $\Phi_{m,n}$ for the n th antenna element in the m th circle is given by

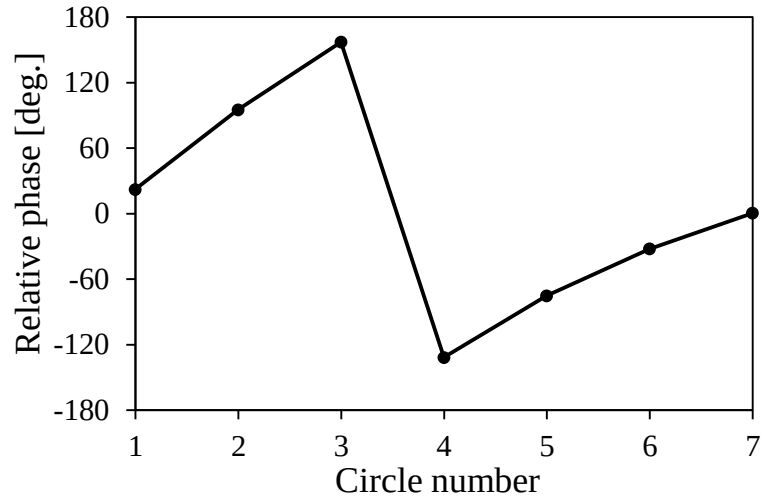
$$\Phi_{m,n} = -\psi_m - kmd_\rho \cos(\phi_0 - \phi_{m,n}) \sin \theta_0 \quad (2.14)$$

where $n = 1, 2, \dots, N_m$ and $\phi_{m,n} = \pi(n-1)/(3m)$. To realize the abovementioned phase shift, the antenna element with right-hand circular polarization (RHCP) should be rotated counterclockwise by $\Phi_{m,n}$, and that with left-hand circular polarization (LHCP) should be rotated clockwise by $\Phi_{m,n}$.

The effect of the amplitude deviation on the directivity of the array is now evaluated. Elements in each circle are assumed to have the same amplitude shown in Fig. 2.7(a) and have an identical element pattern in the form of $\cos^{1.7} \theta$ [2.3]. The calculated directivity with the uniform aperture distribution is 31.2 dBi, whereas that with the deviated amplitudes is 31.1 dBi. Therefore, the effect of the amplitude deviation on the antenna performance is expected to be negligible.



(a)



(b)

Figure 2.7: Aperture distribution calculated from the equivalent circuit model. (a) Amplitude and (b) phase.

2.4 Experiments

2.4.1 Fabrication

The prototype antenna was fabricated and evaluated to experimentally verify the feasibility of the proposed phased array and the performance of the proposed array design. All designed components of the array were integrated together into the prototype antenna. Figure 2.8 shows photographs of the fabricated prototype antenna. It has a radial line with a diameter of 305.6 mm and a circular ground plane with a diameter of 410 mm. The radial line is centrally fed by a straight coaxial probe with an insertion length of 5.5 mm. The helical antenna and the probe are metalized on an injection-molded COC plastic shaft via the MID process. The micromotor employed in the prototype antenna was a two-phase bipolar stepper motor (Minebea SMC10-20). It can provide 20 steps per rotation with an 18° angular resolution, and it also provides a higher angular resolution of 2.25° by the microstepping mode. In measurements, the micromotors were controlled in the microstepping mode to prevent gain decrease and sidelobe level increase due to phase quantization errors. The dimension of the stepper motor is an important factor in the proposed antenna because it restricts the element spacing or applicable frequency band of the proposed antenna. The stepper motors used in the prototype antenna have a diameter of 10 mm, and they can be arranged with required spacings in the prototype antenna. They are fixed on the back of the bottom plate of the radial line, and are connected with their respective motor drivers (L&F LFM9103). The control unit comprises a central processing unit board (L&F LF64) and a PC104 mainboard (Nippon Pulse Motor NPMCMNET-I/O104). The control unit receives the rotation angle of each micromotor from a computer, and dispatches the control commands to the motor drivers. Each motor driver converts these signals into pulses, which are then sent to the corresponding stepper motor.

2. Radial Line Planar Phased Array Using Electromechanically Rotated Helical Antennas

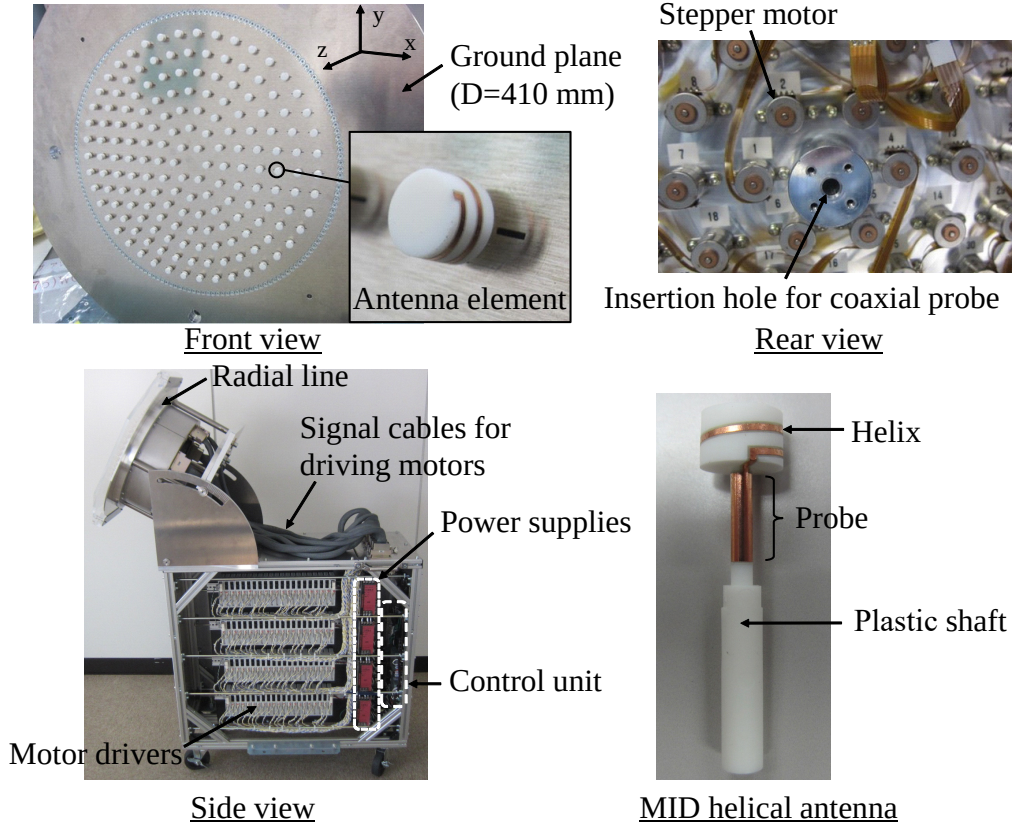


Figure 2.8: Fabricated prototype antenna.

2.4.2 Results and Discussion

Figure 2.9 shows the measured and simulated radiation patterns without any array calibration, that is, each antenna element is rotated by the angles given by (2.14) with a scan angle of $(\theta_0, \phi_0) = (0^\circ, 0^\circ)$ at f_c . A calculated radiation pattern with the excitation amplitudes shown in Fig. 2.7(a) is also plotted in Fig. 2.9. These radiation patterns are normalized by the respective RHCP gains in the broadside direction. A good agreement between the measurements and simulation is obtained in the RHCP radiation pattern, whereas there are some discrepancies in the LHCP one. Although the measurement has a larger LHCP gain than the simulation in the main beam region, it has a low cross-polarization level of less than -24 dB, that is, the AR is less than 1.1 dB. There are some differences between the measured and calculated patterns in the sidelobe region; these differences are apparently caused by a discrepancy in the aperture distribution, as will be shown later.

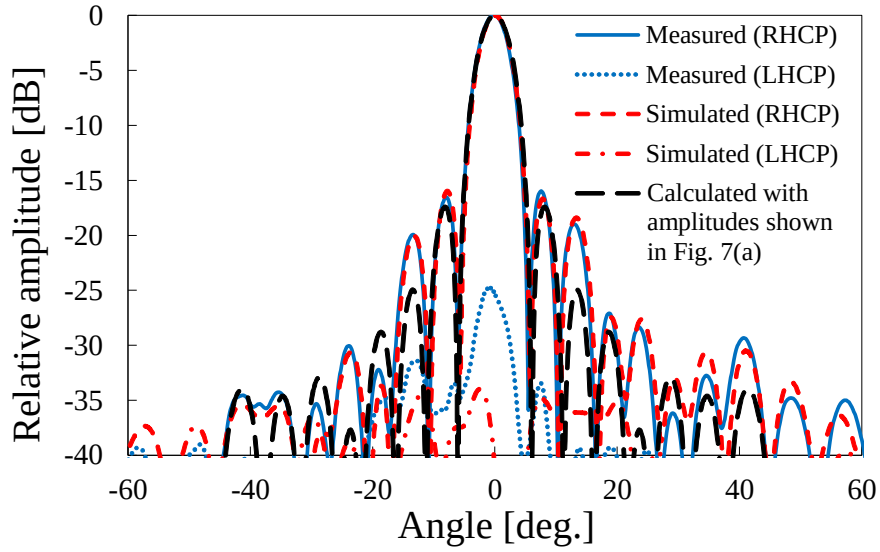


Figure 2.9: Measured, simulated, and calculated radiation patterns for 0° beam-scanning in the x - z plane.

Figure 2.10(a) shows the measured RHCP radiation patterns with the scan angles from -50° to 0° in the x - z plane at f_c . The amplitudes are normalized by the measured gain in the broadside direction for 0° beam-scanning. Figure 2.10(b) shows the measured ARs, scan losses (gain reductions), and beam pointing errors of the scanned beams in the x - z plane. Since beam-scanning performance in the y - z plane is almost the same as that in the x - z plane, the detailed results of it are omitted here. Although the prototype antenna has large scan losses, it achieves dynamic and accurate beam-scanning performance. Large scan losses are caused by the sidelobe level increases in skew directions, and they will be improved by narrowing the element spacings in the array. Measured axial ratio is less than 1.8 dB from -40° to 0° but increases to above 3 dB at -50° . This deterioration of the AR is caused by antenna element's narrow AR beamwidth. The increase in the beam-pointing errors with large scan angles is caused by the beam shift owing to the element pattern [2.27].

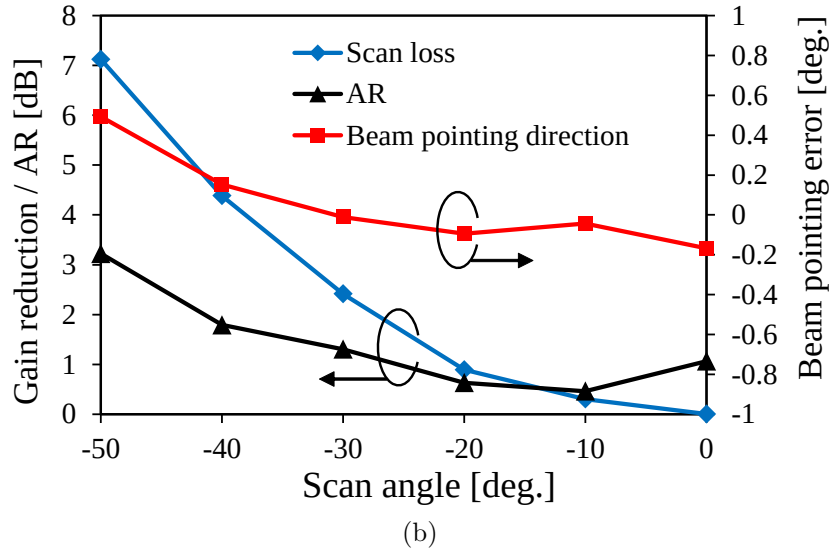
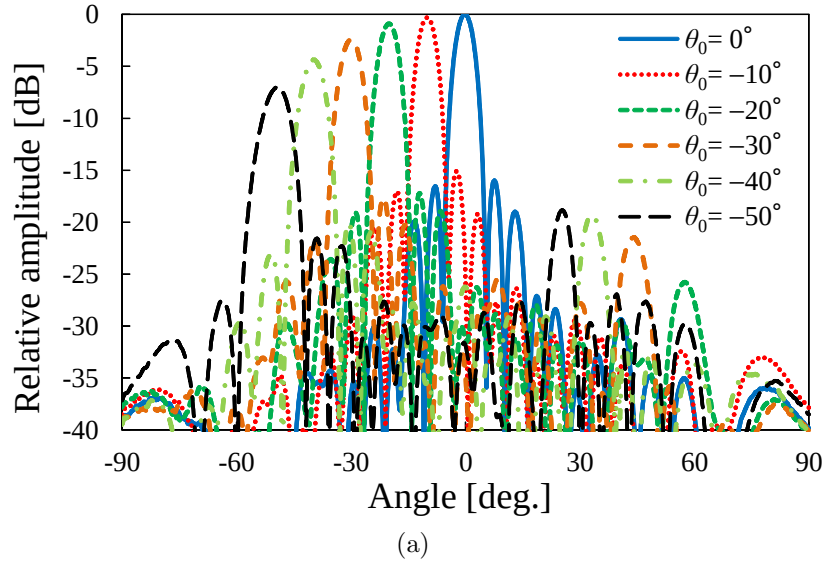


Figure 2.10: (a) Measured radiation patterns for different scan angles at 12.2 GHz; (b) measured scan losses, ARs, and beam-pointing errors.

Figure 2.11 shows examples of the measured and simulated reflection coefficients for different scan angles. The prototype antenna exhibits good impedance matching, and its measured reflection coefficient is less than -17 dB in the required frequency band regardless of the scan angles. The prototype antenna has slightly large element spacings of approximately $0.8\lambda_c$, and the effects of the mutual coupling between adjacent elements on the active admittance is relatively small. Therefore, the input impedance of the prototype antenna is less affected by the element rotations during the beam-scannings. However, if an array is designed for wide-angle beam-scanning, the element spacings must be small enough and the effects of mutual coupling on the array performance will become severe. Antenna designs for wide-angle beam-scanning is therefore a challenge for the future.

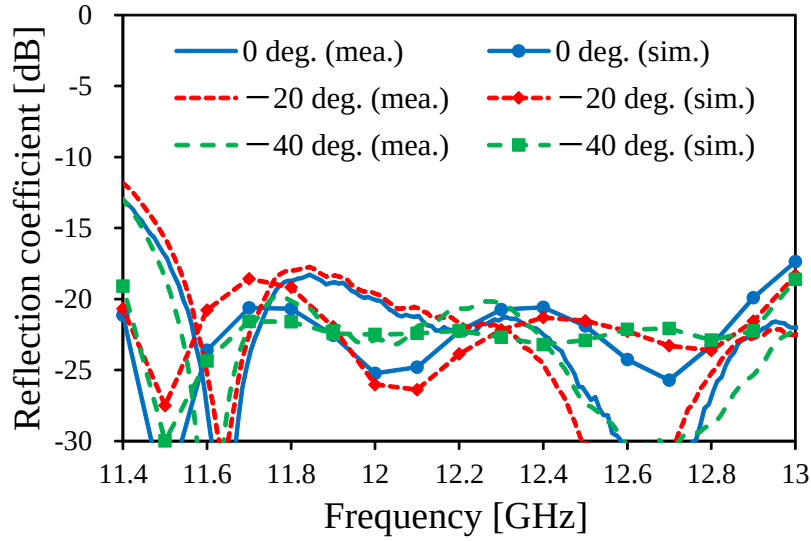


Figure 2.11: Measured and simulated reflection coefficients for different scan angles.

Figure 2.12 shows the frequency variations of the measured and simulated gain and AR in the broadside direction and the antenna efficiencies of the array for 0° beam-scanning. The antenna efficiency η is given by $\eta = G_a / (\pi^2 D_a^2 / \lambda_c^2)$ where G_a is the measured gain and D_a is the diameter of the prototype antenna, which is $2(7d_\rho + d_\rho/2) = 288$ mm. Although a frequency shift by approximately 90 MHz is observed between the measured and simulated gains, relatively high antenna efficiencies (above 60%) are obtained in the required frequency band. The measured AR, which is worse than the simulated one, is less than 1.2 dB in the required frequency band. The causes of these discrepancies between measurement and simulation appear to be fluctuations of the probe length and those of the gaps between helices and the ground plane due to the axial shaft displacement of the stepper motors. For fabrication reasons, the stepper motors have the $+z$ -directed axial shaft displacements, which make the probes shorter than the nominal lengths. This leads to a shift to higher frequency in the frequency response of each antenna's radiation admittance and thus in that of the amplitude distribution of the array. Consequently, the measured gain curve seems to shift to higher frequency from the simulated one. The measured gain at 12.2 GHz is 30.2 dBi with an antenna efficiency of 77%. The causes of the degradation in the antenna efficiency are the conduction and dielectric losses of the radial line, antenna elements, and coaxial probe; the loss due to these factors is estimated to be 0.4 dB in total by some simulations. The fluctuation of the aperture distribution is also a factor for the reduction of antenna efficiency or directivity.

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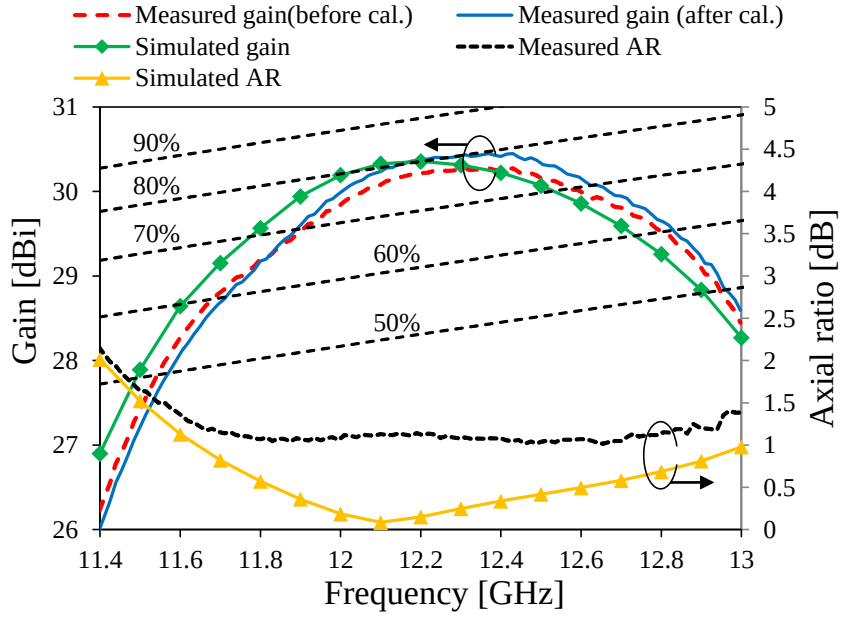


Figure 2.12: Measured and simulated gain and AR in the broadside direction for 0° beam-scanning.

To reveal the causes of the abovementioned discrepancies in the radiation pattern and degradation in directivity and enhance the array performance, array calibration using the so-called rotating element electric field vector (REV) method [2.28], [2.29] was performed. In the REV method, only the array power variation is measured while a single antenna element is successively rotated from 0° to 360° by the corresponding micromotor, and the amplitude and phase of the electric field of the corresponding element can be determined by calculation using the measured power variation. Figure 2.13 shows the measured amplitude and phase of each antenna element in the broadside direction at 12.2 GHz for 0° beam-scanning. The measured amplitudes have periodic variations in the circumference direction. Because 6m antenna elements are arranged in the m th circle in the prototype antenna, the array can be divided into six identical 60° sectors. The periodic variation of the amplitudes appears to be caused by this periodicity of the element arrangement in the array. Moreover, the measured amplitudes have larger values around the outer three circles, unlike those in Fig. 2.7(a). The radiation admittances of the antenna elements in the outer circles may differ from those shown in Fig. 2.6, which are obtained using the unit-cell simulation model, because of the edge effect of the array (variation in the mutual coupling effect in the edge elements of the array). These differences cause the variations in the excitation amplitude and appears to lead to the amplitude fluctuations, as shown in Fig. 2.13. The measured phases have a nonuniform distribution. This is apparently due to the edge effect of the array, axial shaft displacement of the stepper motors, and misalignment in the measurement setup. The calculated directivity with the measured

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amplitudes and phases is 30.6 dBi, and the gain reduction due to these fluctuations of the aperture distribution is estimated to be 0.6 dB.

Correcting the amplitude fluctuations is practically difficult because it requires modifications in the probe lengths, but the phase calibration can be easily performed by rotating the antenna elements to compensate for the measured phase errors. The frequency variation of the measured gain after the phase calibration is also plotted in Fig. 2.12. The measured gain curve after the calibration still shifts to higher frequency from the simulated one for the same reason as above. Motors with smaller axial shaft displacement should be used to prevent this frequency shift. The measured gain at 12.2 GHz improves by 0.2 dB and is 30.4 dBi with an antenna efficiency of 81% after the calibration.

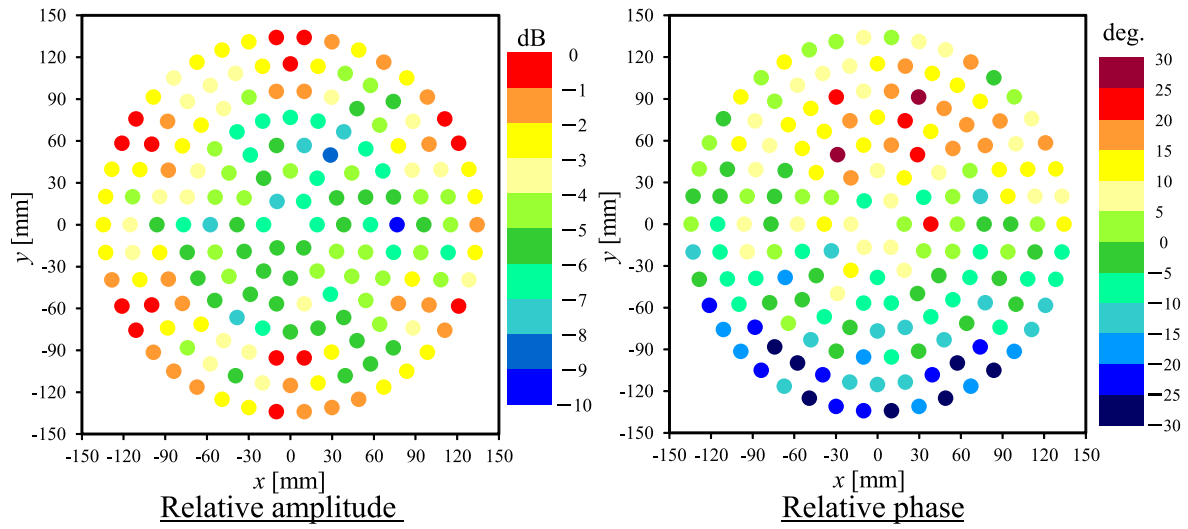


Figure 2.13: Measured amplitude and phase of each antenna element for 0° beam-scanning.

2.5 Conclusion

In this chapter, a novel radial line planar phased array using helical antenna elements individually and electromechanically rotated by micromotors was presented to achieve a PPAA with a flat structure, high efficiency, and flexible beam forming capability. A new design methodology for the RLPA was also presented, which combines the equivalent circuit theory and electromagnetic simulations. The proposed antenna was designed, fabricated, and tested. The fabricated prototype antenna exhibits dynamic and accurate beam-scanning performance; moreover, it has a low reflection coefficient of less than -17 dB and a relatively high antenna efficiency of above 77% at the center frequency, which demonstrates the effectiveness of the proposed design methodology.

Future work in this area includes the development of an array for wide-angle beam-scanning and an array design including the edge effects for further improving the antenna performance. Although the proposed antenna has the disadvantage of a slower beam scanning speed than the conventional APAAs, it can be a lower-cost alternative to the conventional APAAs. Therefore, the proposed phased array is potentially suitable for applications that do not require high-speed beam-scanning, such as satellite communication and microwave power transmission.

Reference

- [2.1] K. R. Carver, “A cavity-fed concentric ring phased array of helices for use in radio astronomy,” Ph.D. dissertation, The Ohio State Univ., 1967.
- [2.2] M. Haneishi and S. Saito, “Radiation properties of microstrip array antenna fed by radial line,” *Antennas Propag. Soc. Symp. Dig.*, 1991, vol.2, pp. 588–591.
- [2.3] H. Nakano, H. Takeda, Y. Kitamura, H. Mimaki, and J. Yamauchi, “Low-profile helical array antenna fed from a radial waveguide,” *IEEE Trans. Antennas Propag.*, vol. 40, no. 3, pp. 279–284, Mar. 1992.
- [2.4] H. Nakano, S. Okuzawa, K. Ohishi, H. Mimaki, and J. Yamauchi, “A curl antenna,” *IEEE Trans. Antennas Propag.*, vol. 41, no. 11, pp. 1570–1575, Nov. 1993.
- [2.5] T. Watanabe, M. Ogawa, K. Nishikawa, T. Harada, E. Teramoto, and M. Morita, “Mobile antenna system for direct broadcasting satellite,” *IEEE Antennas Propag. Soc. Int. Symp. Dig.*, 1996, vol. 1, pp. 70–73.
- [2.6] R. Shavit, L. Pazin, Y. Israeli, M. Sigalov, and Y. Leviatan, “Dual frequency and dual circular polarization microstrip nonresonant array pin-fed from a radial line,” *IEEE Trans. Antennas Propag.*, vol. 53, no. 12, pp. 3897–3905, Dec. 2005.
- [2.7] H. Miyashita and T. Katagi, “Radial line planar monopulse antenna,” *IEEE Trans. Antennas Propag.*, vol. 44, pp. 1158–1165, Aug. 1996.
- [2.8] Y. Liang, J. Zhang, Q. Liu, and X. Li, “High-power radial-line helical subarray for high-frequency applications,” *IEEE Trans. Antennas Propag.*, vol. 66, no. 8, pp. 4034–4041, Aug. 2018.
- [2.9] O. Shibata, S. Saito, and M. Haneishi, “Radiation properties of microstrip array antennas fed by radial line,” *Electron. Commun. Jpn.*, vol. 76, no. 12, pp.93–102, 1993.
- [2.10] R. H. Phillion and M. Okoniewski, “Improving the phase resolution of a micromotor-actuated phased reflectarray,” *Proc. Microsyst. Nanoelectron. Res. Conf.*, 2008, pp. 169–172.
- [2.11] L. Martinez-Lopez, J. Rodriguez-Cuevas, A. E. Martynyuk, and J. I. Martinez-Lopez, “Wideband-reconfigurable reflectarrays based on rotating loaded split rings,” *J. Electromag. Wave Appl.*, vol. 29, no. 2, pp. 218–232, Jan. 2015.
- [2.12] X. Yang, S. Xu, F. Yang, M. Li, Y. Hou, S. Jiang, and L. Liu, “A broadband high-efficiency reconfigurable reflectarray antenna using mechanically rotational elements,” *IEEE Trans. Antennas Propag.*, vol. 65, no. 8, pp. 3959–3966, Aug. 2017.

- [2.13] K. Yokokawa, K. Suzuki, N. Nakamoto, S. Yamaguchi, Y. Suzuki, T. Fukasawa, M. Otsuka, N. Yoneda, and H. Miyashita, "Study on rotational element electronically scanned array," *Proc. IEEE Int. Workshop Electromagn., Appl. Student Innov. Competition (iWEM)*, Aug. 2018.
- [2.14] L. Yu, C. Yuan, J. He, and Q. Zhang, "Beam steerable array antenna based on rectangular waveguide for high-power microwave applications," *IEEE Trans. Plasma Sci.*, vol. 47, no. 1, pp. 535–541, Jan. 2019.
- [2.15] A. Sharaiha and C. Terret, "Analysis of quadrifilar resonant printed helical antenna for mobile communications," *IEE Proc. H - Microw. Antennas Propag.*, vol. 140, no. 4, pp. 269–273, Aug. 1993.
- [2.16] A. Islam, H. N. Hansen, P. T. Tang, and J. Sun, "Process chains for the manufacturing of molded interconnect devices," *Int. J. Adv. Manuf. Technol.*, vol. 42, no. 9, pp. 831–841, Jun. 2009.
- [2.17] L. Pazin and Y. Leviatan, "Uniform amplitude excitation of radiating elements in array antenna pin-fed from a radial waveguide," *IEE Proc. Microw. Antennas Propag.*, vol. 148, no. 6, pp. 413–417, Dec. 2001.
- [2.18] L. Pazin and Y. Leviatan, "Design of a radial waveguide feed network for a pin-fed array antenna," *IEE Proc. Microw. Antennas Propag.*, vol. 153, no. 1, pp. 38–42, Feb. 2006.
- [2.19] N. Takemura, H. Miyashita, and S. Makino, "Wave analysis of the aperture field distribution in probe-fed radial line planar antennas," *IEICE Trans. Commun.*, vol. E89-B, no. 9, pp. 2580–2586, Sep. 2006.
- [2.20] O. Becker and R. Shavit, "Efficient full-wave method of moments analysis and design methodology for radial line planar antennas," *IEEE Trans. Antennas Propag.*, vol. 59, no. 6, pp. 2294–2302, Jun. 2011.
- [2.21] M. E. Bialkowski and P. Kabacik, "An electromagnetic-field method modeling of a radial line planar antenna with coupling probes," *IEEE Trans. Antennas Propag.*, vol. 51, no. 5, pp. 1114–1120, May 2003.
- [2.22] G. Holtrup, A. Marguinaud, and J. Citerne, "Synthesis of electronically steerable antenna arrays with elements on concentric rings with reduced sidelobes," *Proc. IEEE Antennas Propag. Soc. Int. Symp.*, vol. 3, pp. 800–803, July 2001.
- [2.23] R. F. Harrington, *Time-Harmonic Electromagnetic Fields*, New York, NY, USA: Wiley-IEEE Press, 2001.
- [2.24] H. Nakano, H. Takeda, T. Honma, H. Mimaki, and J. Yamauchi, "Extremely low-profile helix radiating a circularly polarized wave," *IEEE Trans. Antennas Propag.*, vol. 39, no. 6, pp. 754–757, Jun. 1991.

- [2.25] H. Y. Yee, "The design of large waveguide arrays of shunt slots,," IEEE Trans. Antennas Propag., vol. 40, no. 7, pp. 775–781, July 1992.
- [2.26] B. Tomasic and A. Hessel, "Linear phased array of coaxially-fed monopole elements in a parallel plate waveguide," Rome Lab In-House Report RL-TR-91-124, Rome Laboratory, Griffiss AFB, New York, April 1991.
- [2.27] S. Mano, M. Ono, and Y. Takeichi, "Effects of the element pattern on the directive properties of a scanning planar array," IEEE Trans. Antennas Propag., vol. 22, no. 2, pp. 169–172, Mar. 1974.
- [2.28] S. Mano and T. Katagi, "A method for measuring amplitude and phase of each radiating element of a phased array antenna," Trans. IECE '82/5, vol. J65-B, no. 5, pp. 555–560, May 1982. (in Japanese).
- [2.29] S. Mano and T. Katagi, "A method for measuring amplitude and phase of each radiating element of a phased array antenna," Electron. Commun. Jpn. B, vol. 65, no. 5, pp. 58–64, 1982.

Chapter 3

Waveguide Slot Array with Code-Division Multiplexing Function for Single RF Chain Digital Beamforming

3.1 Introduction

Digital beamforming (DBF) array antennas have been actively developed in recent years to meet the demands of modern wireless communication and radar systems [3.1]–[3.3]. In the conventional element-level DBF array antenna, each antenna element is connected to a separate radio frequency (RF) chain (amplifier, down-converter, and filter) and analog-to-digital converter (ADC) to obtain an individual complex received signal from each antenna element. Therefore, expensive hardware costs and increased power consumption make it difficult to realize a large-scale element-level DBF array antenna.

Many works have been reported to realize an element-level DBF array antenna that reduces the number of RF chains and ADCs [3.4]–[3.16]. To share a single RF chain among multiple antennas, received signals from multiple antennas are multiplexed into a single RF chain and ADC. Subsequently, each signal is recovered in the digital domain. In analogy with multiple access methods in wireless communication systems, three major multiplexing techniques, such as time-division multiplexing (TDM), frequency-division multiplexing (FDM), and code-division multiplexing (CDM), have been applied to realize a DBF array antenna with a single RF chain. In [3.4]–[3.7], received signals from each antenna element are transferred to a single RF chain sequentially using switches in a TDM fashion. In [3.8] and [3.9], received signals from each antenna element are converted to signals with a different intermediate frequency (IF) from each other and then multiplexed in an FDM fashion. However, TDM and FDM schemes have a major drawback: when an adjacent channel interferer (ACI) is present, recovered signals in TDM and FDM schemes

3. Waveguide Slot Array with Code-Division Multiplexing Function for Single RF Chain Digital Beamforming

suffer from corruption by the ACI [3.10], [3.11].

CDM is the most promising method because, unlike TDM and FDM, the original signals can be recovered even in the presence of ACI [3.11]. In the CDM scheme, received signals from each antenna element are multiplied by spreading codes, and the spread signals are transferred to a single ADC. The original signals are recovered by a despreading process in the digital domain. In [3.11], received signals are code-modulated in the RF domain with a code-modulating low noise amplifier (LNA). In [3.12] and [3.13], code-modulation is performed in the IF domain with multipliers after amplification and frequency conversion of received signals. In these systems, each antenna element still requires a separate amplifier, which increases cost and power consumption in a large-scale array. In [3.14], received signals are code-modulated with reflection-type $0^\circ/180^\circ$ phase shifters, and then all spread signals are combined by a microstrip power combiner and fed into a single RF chain. In [3.15] and [3.16], a metasurface with a CDM function is presented. The responses of each unit cell are code-modulated by switching their integrated diodes, and then spread signals are spatially combined and fed into a feeding antenna with a single RF chain. In these antennas, signal amplification is implemented after combining the spread signals, and thus they can reduce the number of amplifiers.

In this study, a novel waveguide slot array antenna with a CDM function is presented. The proposed antenna uses slot pairs with RF switches (e.g., PIN diodes) on the broad wall of a rectangular waveguide. Waveguide slot arrays with integrated PIN diodes are presented in several studies [3.7], [3.17]–[3.19]. In [3.17]–[3.19], slots with appropriate radiation phases are activated, and the others are deactivated by switching the diodes' bias to achieve desired radiation patterns in the RF domain. In [3.7], only one slot is activated at a time, and an activated slot is sequentially switched by switching the diodes' bias. The original signal of each slot is then recovered in the digital domain in a TDM fashion. However, in the proposed antenna, the received signal of each slot-pair can be code-modulated by switching the RF switches' bias. Subsequently, the spread signals are combined in the hollow rectangular waveguide, and fed into a single RF chain. Following digitization, the despreading process recovers the original complex signals received by each slot pair, and the DBF is performed to achieve scanned beams, multiple beams, adaptive beams, or direction of arrival (DOA) estimation. To the best of the author's knowledge, this is the first waveguide slot array with a CDM function. The proposed antenna reduces the number of RF chains, amplifiers, and ADCs and has the potential to offer higher efficiency than previous works using microstrip lines or transmittarray/reflectarray architecture.

The rest of this chapter is organized as follows: Section 3.2 explains the detailed configuration and operating principle of the proposed antenna; Section 3.3 describes the antenna design; Section 3.4 presents the experimental results of the prototype antenna; Section 3.5 presents the discussions; and Section 3.6 presents the conclusion of this chapter.

3.2 Antenna Configuration and Operating Principle

Figure 3.1 shows the configuration of the proposed antenna with four slot pairs. The antenna comprises the bottom part of a metallic waveguide, a multilayer printed circuit board (PCB), RF switches, a single RF chain, an ADC, and a digital signal processing (DSP) unit. Both surfaces of the PCB are covered with copper foil, and slot apertures are formed by etching. Moreover, the back surface of the PCB is the top wall of the rectangular waveguide. A rectangular waveguide is formed by covering the bottom part of the metallic waveguide with the PCB. The waveguide is connected to the single RF chain through its feeding port. A slot pair comprises two longitudinal slots, which are offset symmetrically about the centerline of the waveguide, with the same dimensions. An RF switch is mounted across each slot aperture on the top surface of the PCB, and a DC bias line is printed on the inner layer of the PCB. The detailed configuration of the slot pair will be presented in Section 3.3.

The waveguide is here assumed to operate in only a single dominant mode (TE_{10} mode). In that case, the transverse currents along the top inner waveguide wall are symmetrical about the centerline of the waveguide [3.20]. Because the longitudinal slot radiates by interrupting the transverse current, the field radiated by each slot of the slot pair has the same amplitude in anti-phase to each other. In the proposed antenna, one RF switch of each slot pair is turned off (OFF state), and the other is turned on (ON state), allowing one slot to receive an incident wave, whereas the other cannot. Therefore, the phase of the received signal of each slot pair can be reversed by switching the RF switches' state. The states of each switch are sequentially controlled according to high-speed spreading codes, and the received signal of each slot pair is code-modulated and spectrum-spread. Subsequently, spread signals are combined in the rectangular waveguide and fed into a single RF chain and an ADC. Following digitization, the original complex signals received by each slot pair are demultiplexed by the despreading process in the DSP unit and then used for further beamforming or DOA estimation with DSP.

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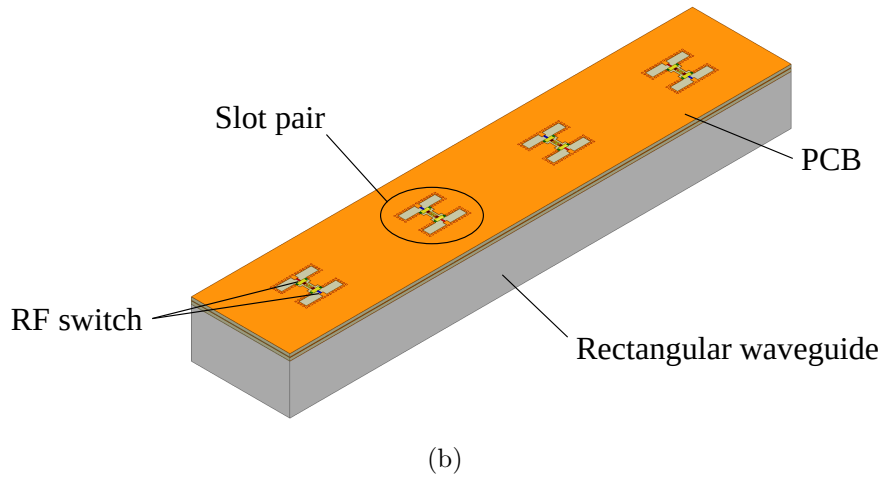
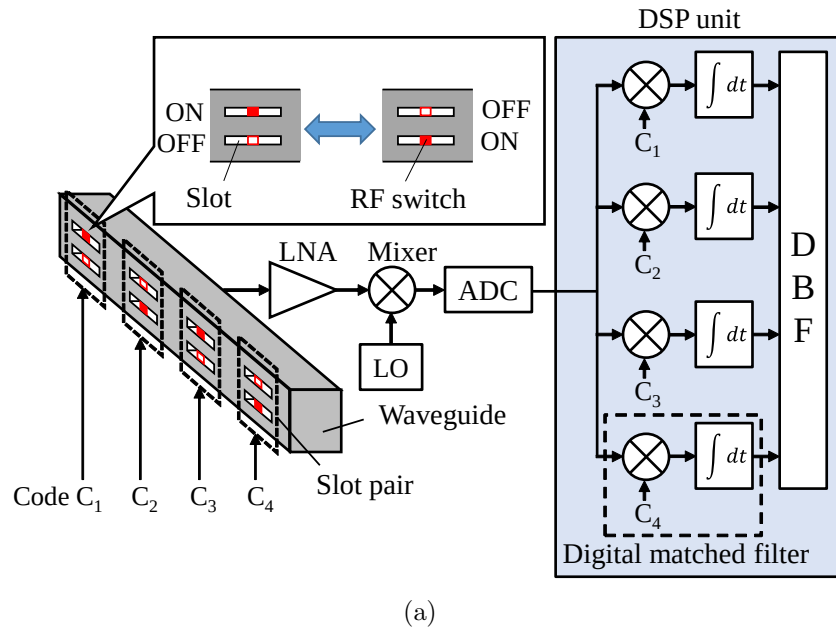


Figure 3.1: Configuration of the proposed antenna. (a) Block diagram and (b) perspective view of the waveguide slot array.

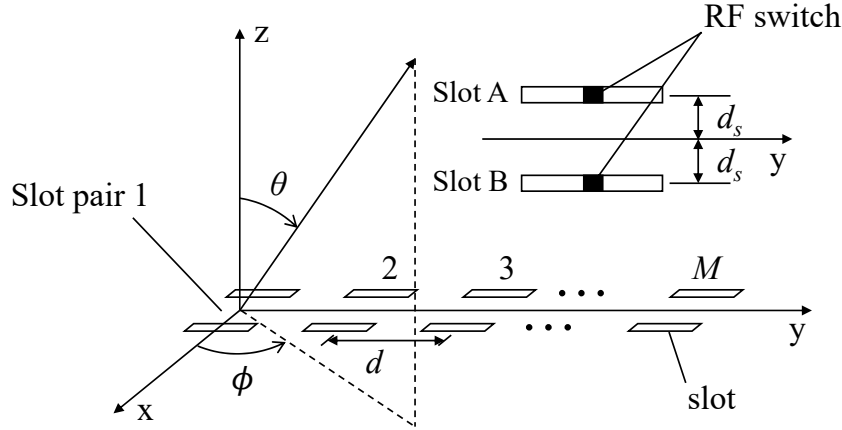


Figure 3.2: Radiating aperture of the proposed antenna with M slot pairs.

To understand the operating principle and signal-to-noise ratio (SNR) performance of the proposed antenna, a mathematical model of the proposed antenna is now described. Figure 3.2 illustrates a coordinate system of the radiating aperture of the proposed antenna. M slot pairs are arranged parallel to the y -axis, and each slot is offset by $\pm d_s$ from it. Each slot is now assumed to have a radiation pattern given by

$$\mathbf{f}^{(\pm)}(\theta, \phi) = \mathbf{f}_m(\theta, \phi) e^{\pm j k_0 d_s \sin \theta \cos \phi} \quad (3.1)$$

where the superscripts $(+)$ and $(-)$ indicate the radiation pattern of the slot A and B, respectively, $\mathbf{f}_m(\theta, \phi)$ is the far-field vector of the m th slots, and k_0 is the free space wavenumber. We consider that a plane wave arrives at the antenna with an incident angle of θ in the y - z plane ($\phi = \pi/2$). In this case, the received signal of the m th slot pair is given by

$$r_m(t) = c_m(t) f_{m\phi} s_m(t) e^{j 2\pi f_c t} \quad (3.2)$$

where $f_{m\phi}$ is the ϕ -component of $\mathbf{f}_m(\theta, \phi)$ in the y - z plane (the cross-polarization component is ignored), f_c is the carrier frequency, $s_m(t)$ is an incident signal to the slot pair, and $c_m(t)$ is a designated spreading code and indicates an operating state of the slot pair: $c_m(t) = 1$ when the slot A is activated and $c_m(t) = -1$ when the slot B is activated. Here the slot with an RF switch in the ON state is assumed to be perfectly deactivated and does not receive any signals.

All the received signals $r_m(t)$ (for $1 \leq m \leq M$) are combined by the waveguide, and the output signal is then amplified by an LNA. The resultant signal is expressed as

$$R(t) = g \left[\sum_{m=1}^M \alpha_m r_m(t) + n(t) \right] \quad (3.3)$$

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where α_m is the transmission characteristics of each slot pair through the waveguide, g is the gain of the LNA, and $n(t)$ is the white Gaussian noise referred to the input of the LNA. $R(t)$ is then downconverted to the baseband, and the baseband signal is given by

$$R_b(t) = g \left[\sum_{m=1}^M \alpha_m f_{m\phi} c_m(t) s_m(t) + n(t) \right]. \quad (3.4)$$

Following digitization, the signal is separated into I and Q components by a digital IQ demodulator, but the IQ demodulator is ignored and only a single path is shown for simplicity in Fig. 3.1(a). The signal is then input to M digital matched filters for the despreading process, where the m th matched filter is matched to the m th spreading code. When the incident signal $s_m(t)$ is assumed to vary more slowly than the spreading code $c_m(t)$, the output signal $S_m(t)$ of the m th matched filter is expressed as

$$S_m(t) = \frac{1}{T_b} \int_0^{T_b} R_b(t) c_m(t) dt = S_{m,d}(t) + S_{m,i}(t) \quad (3.5)$$

$$S_{m,d}(t) = g \alpha_m f_{m\phi} C_{m,m} s_m(t) \quad (3.6)$$

$$S_{m,i}(t) = g \sum' \alpha_{m'} f_{m'\phi} C_{m,m'} s_{m'}(t) + \frac{g}{T_b} \int_0^{T_b} n(t) c_m(t) dt \quad (3.7)$$

where T_b is the symbol duration, and $\sum'$ indicates a summation from $m' = 1$ to M except for $m' = m$. $S_{m,d}(t)$ represents the desired signal for the m th slot-pair, whereas $S_{m,i}(t)$ is the interference from other signals and noise. $C_{m,m'}$ is the cross-correlation of the spreading codes and is defined as

$$C_{m,m'} = \frac{1}{T_b} \int_0^{T_b} c_m(t) c_{m'}(t) dt. \quad (3.8)$$

In (3.7), the first term of the right side is the interference from other signals, and a spreading code with $C_{m,m'} = 0$ must be used to minimize $S_{m,i}(t)$. Therefore, although there are several options for the spreading codes [3.21], [3.22], the perfectly orthogonal codes, namely the Walsh-Hadamard (WH) codes (see Appendix B.1), with a length of $L_c \geq M$ should be used as the spreading codes. When the WH codes are used, it is important to ensure synchronization between spreading by PIN diode switching and despreading by signal processing for each slot pair, because the cross-correlation between different shifts of WH codes is not zero. In contrast to the case of CDMA for wireless communication systems, it would be relatively easy to synchronize spreading and despreading. This is because the antenna only needs to be designed in such a way that they are synchronized within the same device. When the perfect synchronization between spreading and despreading is achieved, the orthogonality is satisfied for the WH codes as follows:

$$C_{m,m'} = \begin{cases} 1 & \text{for } m = m' \\ 0 & \text{for } m \neq m', \end{cases} \quad (3.9)$$

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and $S_{m,d}(t)$ and $S_{m,i}(t)$ are reduced to

$$S_{m,d}(t) = g\alpha_m f_{m\phi} s_m(t) \quad (3.10)$$

$$S_{m,i}(t) = \frac{g}{T_b} \int_0^{T_b} n(t) c_m(t) dt. \quad (3.11)$$

As expected, the original signal can be perfectly recovered for the WH codes. The SNR at each element output becomes

$$SNR_m = \frac{|\alpha_m|^2 G_m P_s}{\sigma_n^2} \quad (3.12)$$

where $P_s = E[|s_m(t)|^2]$, $\sigma_n^2 = E[|n(t)|^2]$ ($E[\cdot]$ is an expectation operator), and $G_m = |f_{m\phi}|^2$ is the gain of each slot pair (see Appendix B.2). Similarly, SNR at the array output is given as

$$SNR_a = \frac{M\alpha_0^2 G_e P_s}{\sigma_n^2} \quad (3.13)$$

where $|\alpha_m| \simeq \alpha_0$ and $G_m = G_e$ for all slot pairs. For the hollow waveguide, $|\alpha_m|^2 = \alpha_0^2 \simeq 1/M$ in (3.12) and (3.13), and thus the SNRs are degraded to $1/M$ of those for the element-level DBF array antenna or the previous works in [3.11] and [3.12]. This is because the signal amplification is performed after signal combination, and the signal power of each slot pair is degraded due to the insertion loss of the waveguide combiner. This SNR degradation is a drawback of the proposed antenna, and there is another restriction. When the spreading code with length L_c is used, the chip duration becomes T_b/L_c . Therefore, the ADC sampling rate and PIN diode's switching rate must be faster than M/T_b . In particular, the available PIN diode's switching speed dictates the maximum number of multiplexing signals and maximum symbol rate in the proposed antenna (see Appendix B.3).

Although the receiving antenna has been considered here, the same idea is applicable to a transmitting antenna. In that case, the transmitting signals from each slot-pair are code-modulated in the digital domain and recovered by the despreading process by switching the RF switches' bias. However, there are some difficult problems for the transmitting antenna such as spurious emissions and power handling capability of the RF switch, which will hinder practical application. Therefore, this study focuses only on the receiving antenna. In addition, the incident wave is here assumed to lie in the y - z plane. For other incident angles out of this plane, the factor $e^{\pm jk_0 d_s \sin \theta \cos \phi}$ remains in (3.2), which introduces errors in the spreading process. The effect of these errors will be discussed in Section 3.4.

3.3 Antenna Design

A prototype waveguide slot array with eight slot pairs is designed as a proof of concept of the proposed antenna. Figure 3.3 demonstrates a detailed configuration of a slot pair for the prototype antenna. Each slot pair and its DC bias line are printed on a 4-layers PCB. The conductor layers are denoted as L1–L4, and two substrates with a thickness of 0.75 mm are bound with a prepreg. The total thickness of the PCB is 1.67 mm. Megtron6 ($\epsilon_r = 3.6$ and $\tan \delta = 0.003$) by Panasonic Industry is selected as a substrate material. Slot apertures with a length of L_s and a width of w_s are formed by etching all the layers. Additionally, multiple through-hole vias with a diameter of 0.4 mm are placed around the slot apertures at 0.7 mm intervals to prevent RF leakage into the parallel plate mode between layers.

PIN diode (SMP1345-079LF by Skyworks) is used as an RF switch. A single PIN diode is placed across each slot at its center on the L1 layer. DC bias lines of each PIN diode are printed on the L2 layer and connected to the bias circuit (not shown in Fig. 3.3). An RF choke inductor of 15 nH (LQW15AN15NG00 by Murata Manufacturing) is placed between the PIN diode and the bias line to prevent the RF leakage into the bias line. Two DC block capacitors of 1.2 pF (GJM1555C1H1R2WB01 by Murata Manufacturing) are placed between the component pad and the ground to isolate the bias line from the ground. In the proposed antenna, one diode of each slot pair is reverse biased (OFF state), and the other is forward biased (ON state). When the PIN diode is in the OFF state, it approximates an open circuit and has minimal effect on the radiation characteristics of the slot. In contrast, when the PIN diode is in the ON state, it approximately short-circuits the slot at its center, effectively deactivating it.

The inter-element spacing d , distance between the centers of adjacent slot pairs, is set to 33.1 mm for the prototype antenna to have a field of view of $\pm 15^\circ$ without any grating lobes at the operating frequency of 7 GHz. A standing-wave waveguide slot array is adopted in the prototype antenna, and the dimensions of the bottom part of metallic waveguide is designed such that a guided wavelength becomes twice the inter-element spacing. A single-ridged waveguide is selected as the waveguide, and the designed dimensions are listed in Table 3.1.

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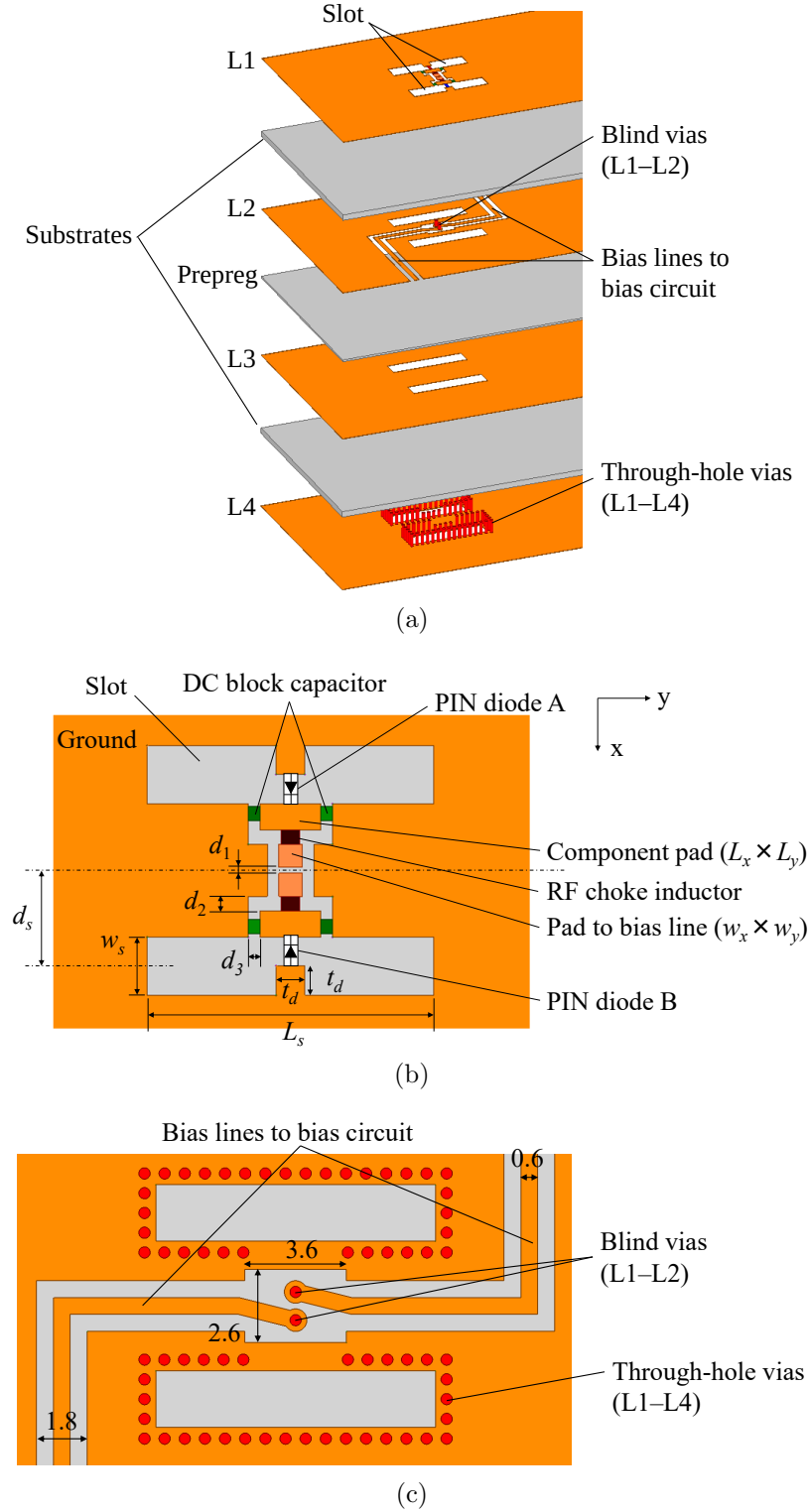


Figure 3.3: Configuration of a slot pair for the prototype antenna. (a) Expanded view of the PCB, (b) detailed configuration of slot pair with PIN diodes, RF choke inductors, and DC block capacitors, and (c) layout of the L2 layer.

A slot pair with the integrated components is designed using the electromagnetic

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simulator Ansys HFSS. In the prototype antenna, slot pairs with identical dimensions are used to simplify the antenna design, and they are determined such that the slots resonate at 7 GHz with an appropriate radiation conductance for impedance matching. The aperture distribution would degrade due to this simplified design. However, it is acceptable because the signal variations among the slot pairs caused by it can be calibrated in the digital domain after demultiplexing. In the simulations, the PIN diode is modeled as an equivalent circuit of lumped elements to consider its effect on the slot performance. Specifically, the PIN diode in the ON state is modeled as a series circuit of a $2.0\ \Omega$ resistor and a 0.7 nH inductor, whereas one in the OFF state is modeled as a 0.13 pF capacitor [3.23], [3.24]. The designed dimensions of a slot pair are shown in Table 3.1.

Table 3.1: Dimensions of the prototype antenna.

Parameter	Value [mm]	Parameter	Value [mm]
Waveguide		Pad and clearance	
inner width	23.0	L_x	0.9
inner height	11.5	L_y	2.1
ridge height	5.0	w_x	0.8
ridge width	1.0	w_y	0.8
Slot-pair		d_1	0.2
L_s	9.9	d_2	0.5
w_s	2.0	d_3	0.4
d_s	3.3	t_d	1.0

Figure 3.4 depicts an exploded view of the prototype antenna. Eight designed slot pairs are arranged linearly, and both ends of the ridge waveguide are short-circuited at 16.55 mm away from the center of the end slot pairs. The ridge waveguide is fed by an SMA coaxial probe inserted at its center from its bottom side. Figure 3.5 shows the detailed structure of the transition from the coaxial line to the ridge waveguide. The coaxial inner conductor and the dielectric are extended into the waveguide, and the ridges are linearly tapered at the transition to achieve impedance matching.

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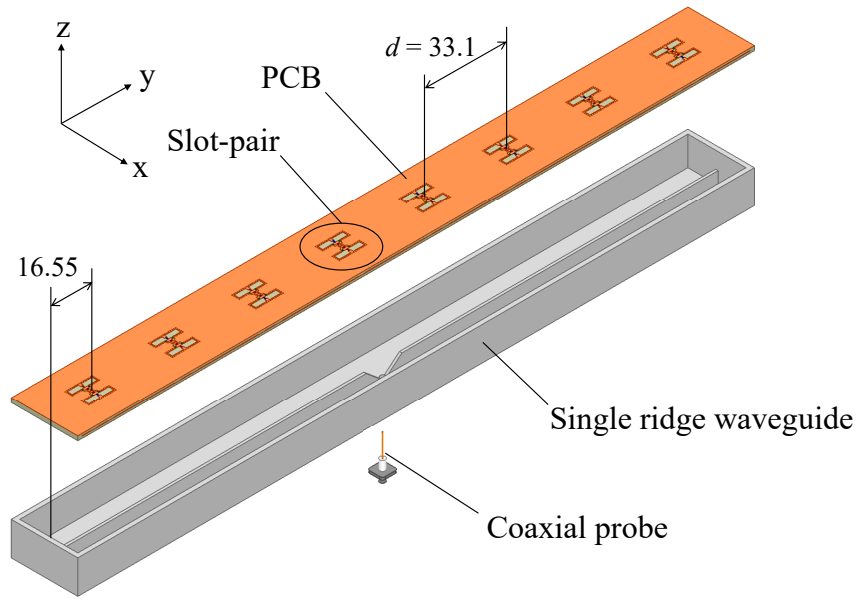


Figure 3.4: Exploded view of the prototype waveguide slot array (dimensions in millimeter).

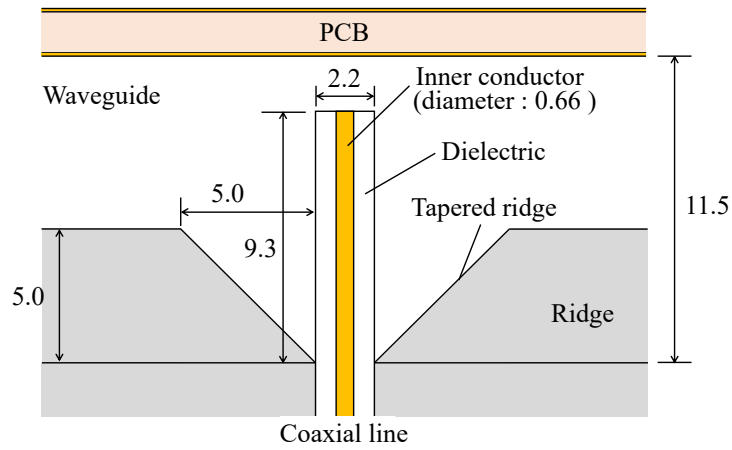


Figure 3.5: Configuration of the designed transition from the coaxial line to the single-ridged waveguide (dimensions in millimeter).

3.4 Experiments

3.4.1 Experimental Setup

The prototype antenna was fabricated and tested to verify the feasibility of the proposed antenna. Figure 3.6 shows photos of the fabricated antenna. The prototype antenna was installed in a ground plane of dimension 400 mm $\times$ 400 mm. Bias lines for PIN diodes were connected to bias circuits (DC power suppliers) through biasing wires. The PIN diodes were forward biased with the forward current of 10 mA and were reverse biased with the reverse voltage of 22 V.

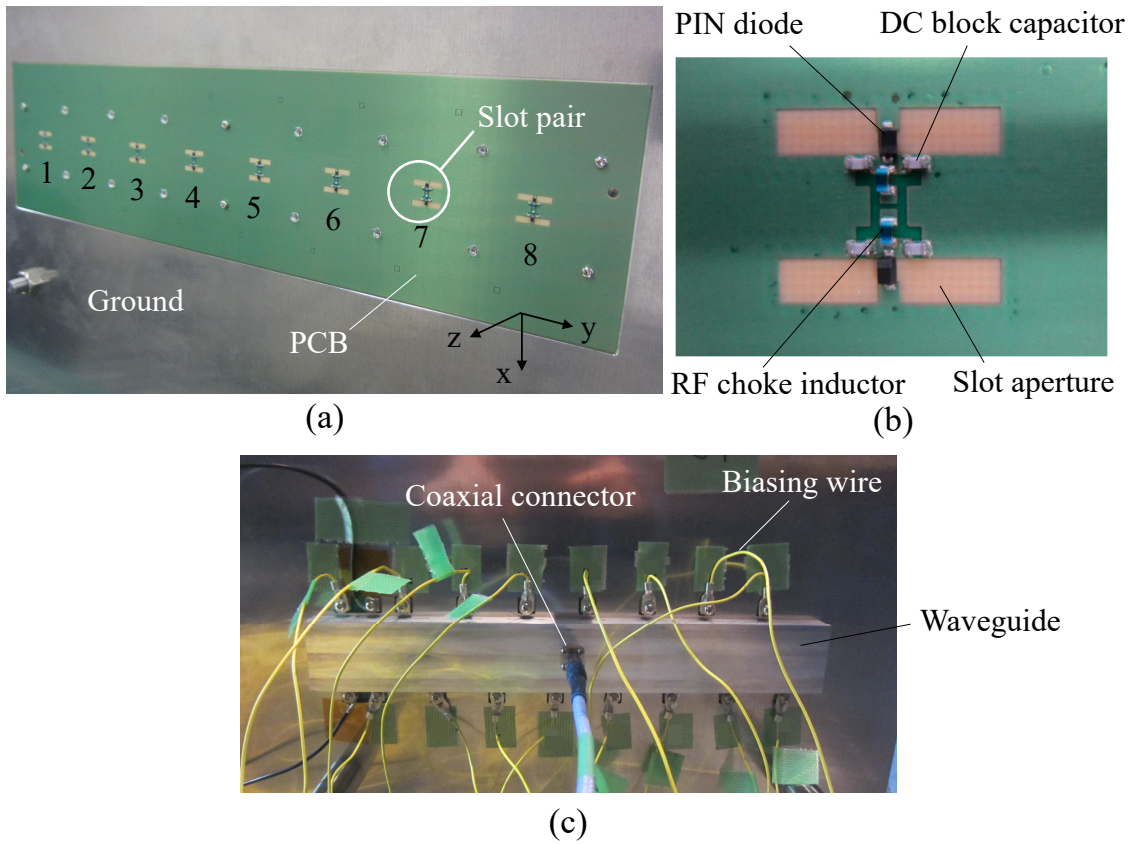


Figure 3.6: Fabricated prototype antenna.

Here we consider that a continuous wave (CW) with an amplitude s_0 impinges on the proposed antenna with the incident angle θ in the y - z plane. The incident signal to the m th slot pair is given by $s_m(t) = s_0 e^{jk_0 m d \sin \theta}$, and thus the recovered desired signal (3.10) becomes $S_{m,d}(t) = g\alpha_m s_0 f_{m\phi} e^{jk_0 m d \sin \theta}$ and is proportional to the element pattern of the m th slot pair. In the experiment, therefore, CWs at 7 GHz were transmitted and received while changing θ , and the recovered signals by the prototype antenna were then compared with simulated element patterns. The measurement was performed using

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a compact antenna test range setup in an anechoic chamber. CW transmission and reception, transmission coefficient (S_{21}) measurement, between a transmitting antenna and the prototype one was conducted with a vector network analyzer. To change the incident angle θ , the prototype antenna was rotated around the x -axis, which passed through the waveguide center, with a rotary positioner. For a proof of concept, the biasing state of each PIN diode was manually switched to the specified states described later, and the S_{21} was measured at various θ (from -90° to 90° in 0.1° increments) in each state. Subsequently, the received signal e_m of each slot-pair was recovered from the measured signals at θ by the following expression:

$$e_m(\theta) = \frac{1}{L_c} \sum_{l=1}^{L_c} E_l(\theta) c_m(l) \quad (3.14)$$

where $E_l(\theta)$ and $c_m(l)$ are the measured S_{21} at θ and the designated spreading code for the m th slot pair at the l th state, respectively. This process was performed in software. The synchronization between spreading and despreading was assumed to be perfectly achieved, and the transient characteristics of the PIN diodes was ignored. The WH codes with $L_c = 8$ or 16 were used as the spreading codes in the experiment. Figure 3.7 demonstrates the relationship between the WH codes with $L_c = 8$ and the positions of activated slots (with the PIN diode in the OFF state). In the first state, the activated slots were located to be excited in phase at 7 GHz. This arrangement was assigned to +1 of each code, and the positions of activated slots for the other states were determined as shown in Fig. 3.7. For the WH codes with $L_c = 16$, the WH code whose elements were all +1 was not used for a reason discussed later, and the codes corresponding to the second to tenth rows of the Hadamard matrix were used (see Fig. B.1). The positions of the activated slots for the WH codes with $L_c = 16$ were determined in a similar manner to $L_c = 8$, and the detailed arrangement is omitted here.

■ : activated slot
□ : deactivated slot

Slot-pair number	WH code	State number							
		1	2	3	4	5	6	7	8
1	$c_1 = (+, +, +, +, +, +, +, +)$	■	■	■	■	■	■	■	■
2	$c_2 = (+, -, +, -, +, -, +, -)$	■	□	■	□	■	□	■	□
3	$c_3 = (+, +, -, -, +, +, -, -)$	■	■	□	□	■	■	□	□
4	$c_4 = (+, -, -, +, +, -, -, +)$	■	□	□	■	■	□	□	■
5	$c_5 = (+, +, +, +, -, -, -, -)$	■	■	■	■	□	□	□	□
6	$c_6 = (+, -, +, -, -, +, -, +)$	■	□	■	□	□	■	□	■
7	$c_7 = (+, +, -, -, -, -, +, +)$	■	■	□	□	□	□	■	■
8	$c_8 = (+, -, -, +, -, +, +, -)$	■	□	□	■	□	■	■	□

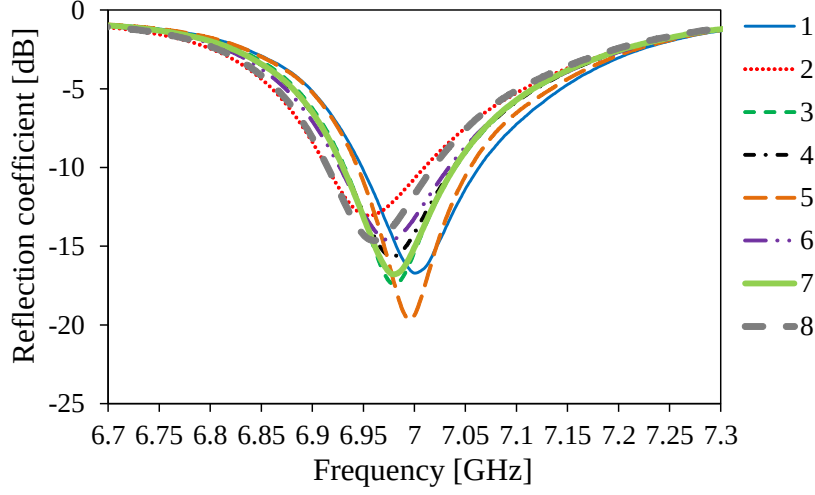
Figure 3.7: Positional relationship between the WH codes with $L_c = 8$ and the activated slots (+ and - indicate +1 and -1, respectively).

3.4.2 Experimental Results

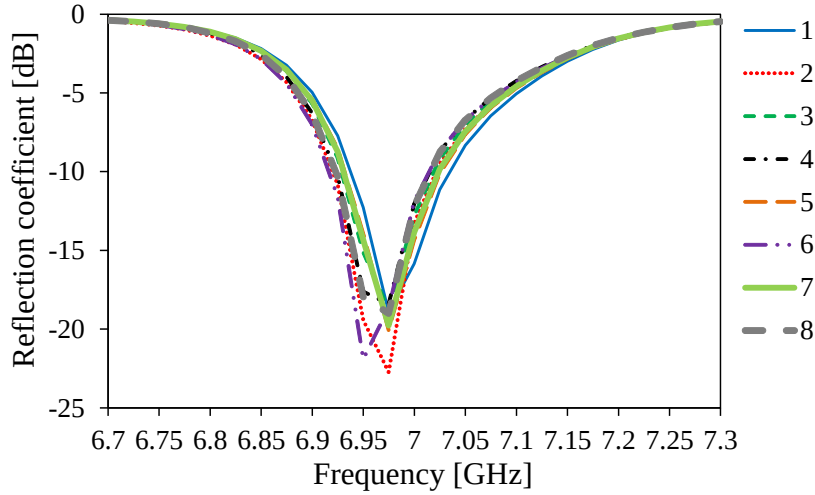
Figures 3.8(a) and (b) show the measured and simulated reflection coefficients at each state for the WH codes with $L_c = 8$, respectively. Although there are some discrepancies between the measured and simulated reflection coefficients, the prototype antenna has a reflection coefficient of less than -10 dB at the operating frequency in all states. The causes for the discrepancies seem to be fabrication errors and errors in electrical properties of the integrated components. The reflection coefficients vary depending on the states due to the variation of the active admittance of each slot pair. The prototype antenna has a bandwidth of only 1.5% with a reflection coefficient of less than -10 dB, and thus a bandwidth enhancement is one of the future works. The bandwidth of the slot can be improved by replacing the PCB material with one with a low dielectric constant, and the slot array's bandwidth can be improved by replacing its series-feed structure with a corporate-feed type [3.25]. Since the measured reflection coefficients for the WH codes with $L_c = 16$ were almost the same as those with $L_c = 8$, the detailed results of them were omitted here. The measured gain in the broadside direction at 7 GHz for the first state, corresponding to an ordinary probe-fed waveguide slot array, is 11.1 dBi, and the estimated radiation efficiency is approximately 47%. The loss due to the reverse parallel resistance R_p of the PIN diode is a significant factor in the reduction in radiation efficiency. R_p of the PIN diode for the prototype antenna is estimated to be approximately 10 k Ω by

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simulations. The radiation efficiency can be improved by replacing the PIN diode with one with a larger R_p , which is also a future challenge.



(a)



(b)

Figure 3.8: Reflection coefficients in each state for the WH codes with $L_c = 8$: (a) Measurement and (b) simulation.

Figure 3.9 shows the recovered signals of the first to fourth slot pairs at each sample angle from the measured S_{21} at 7 GHz for the WH codes with $L_c = 8$. In addition, recovered signals from the simulated S_{21} and simulated individual element patterns of each slot pair are shown in Fig. 3.9. Due to the symmetry of the antenna structure, the recovered signals of the fifth to eighth slot pairs have similar patterns, which are omitted here. The amplitudes of each signal are normalized by their respective peak amplitudes, and the phases are calibrated to be 0 deg. in the broadside direction. The

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recovered signals of each slot pair agree well with their respective element pattern in simulation, whereas there are some discrepancies in measurement. Particularly, there are larger discrepancies between measured and simulated recovered signals for the first slot pair, and demultiplexing is not working well. The cause of this deterioration in recovered signals will be discussed later. Similarly, Fig. 3.10 shows the recovered signals with the WH codes with $L_c = 16$. In this case, the recovered signals from the measured S_{21} agree with simulated element patterns for all slot pairs. Although there are some discrepancies between measured and simulated recovered signals, the prototype antenna is successful in demultiplexing.

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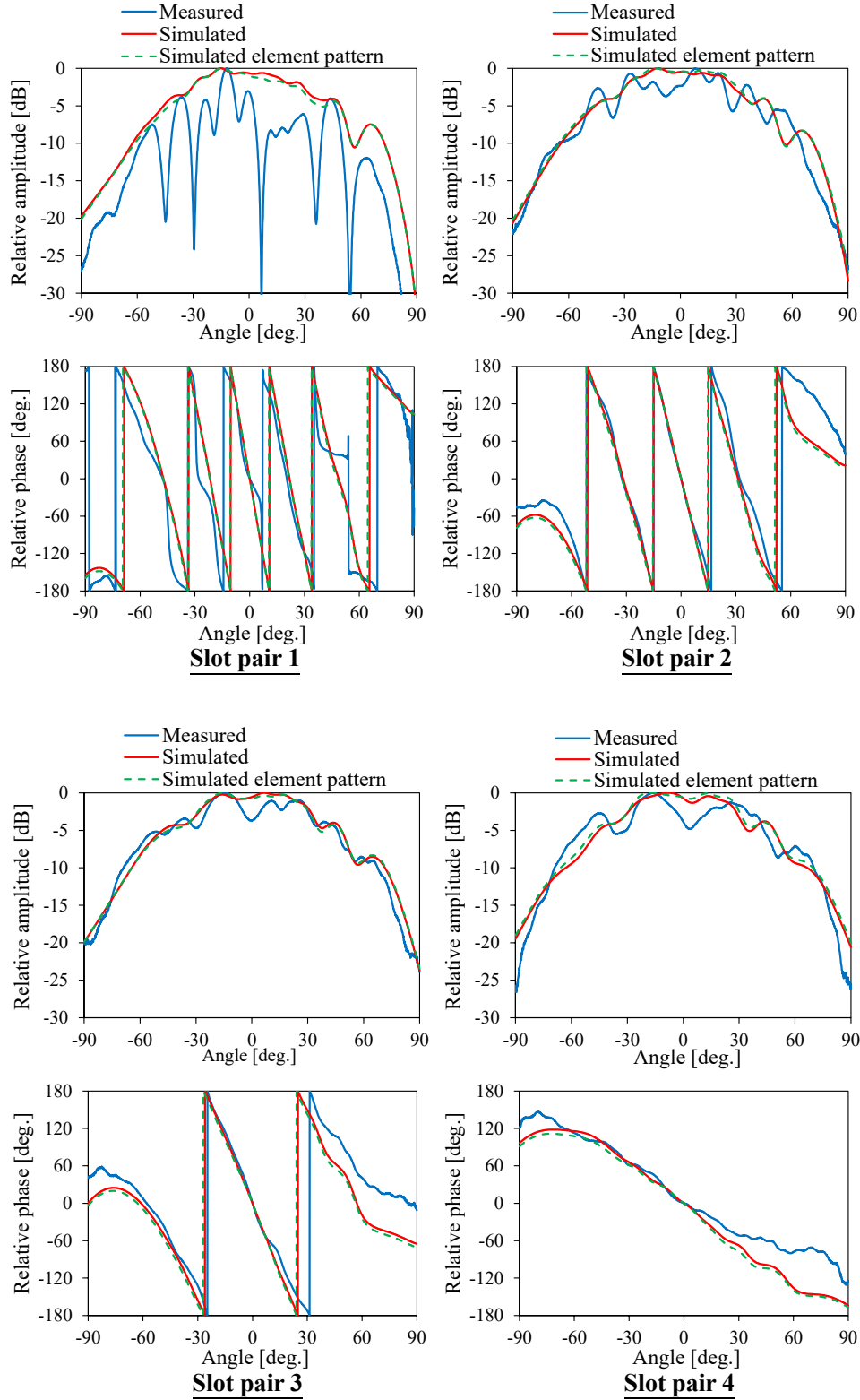


Figure 3.9: Recovered signals of the first to fourth slot-pairs with the WH codes with $L_c = 8$ at 7 GHz. Blue dashed and red solid lines indicate the recovered signals from the measured and simulated S_{21} , respectively, and long-dashed green lines indicate the simulated element patterns of each slot pair. The amplitude and phase patterns are shown in the upper and lower rows, respectively.

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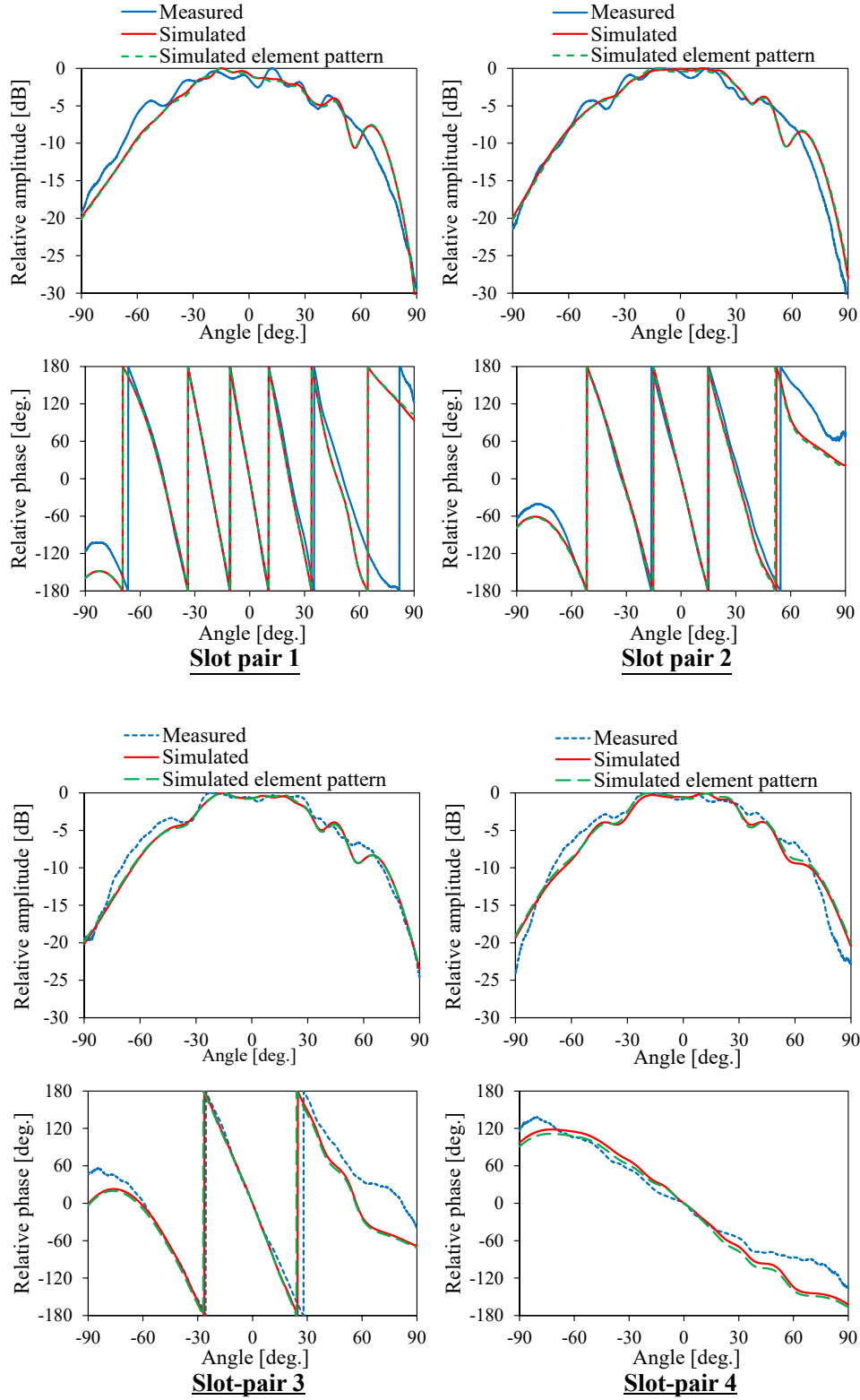


Figure 3.10: Recovered signals of the first to fourth slot-pairs with the WH codes with $L_c = 16$ at 7 GHz. Blue dashed and red solid lines indicate the recovered signals from the measured and simulated S_{21} , respectively, and long-dashed green lines indicate the simulated element patterns of each slot pair. The amplitude and phase patterns are shown in the upper and lower rows, respectively.

3.5 Discussions

3.5.1 Effect of PIN Diode Characteristic Errors

The effect of the differences in properties of each PIN diode on the recovered signals is first discussed to reveal the cause of the deterioration in the recovered signals of the first slot-pair with the WH codes with $L_c = 8$.

We now consider the same model as that in Section 3.2. When the PIN diodes have errors in their electrical properties, the received signal of each slot-pair deviates from its ideal. Now consider δ be the fractional error of the received signal due to the errors in the PIN diode's properties, and the received signal $r'_m(t)$ of the m th slot pair under these errors is then expressed by

$$r'_m(t) = \{1 + \delta_m(t)\} r_m(t). \quad (3.15)$$

The errors depend on the operating states of the slot pair and are expressed as $\delta_m(t) = \delta_m^{(\pm)}$ when the slot corresponding to ± 1 is activated. Replacing $r_m(t)$ in (3.3) with $r'_m(t)$ and using (3.4) and (3.5), we obtain the following recovered signals:

$$S'_{m,d}(t) = \frac{g\alpha_m f_{m\phi} s_m(t)}{T_b} \int_0^{T_b} c_m^2(t) \{1 + \delta_m(t)\} dt \quad (3.16)$$

$$S'_{m,i}(t) = \frac{g}{T_b} \sum' \alpha_{m'} f_{m'\phi} s_{m'}(t) \int_0^{T_b} c_{m'}(t) \{1 + \delta_{m'}(t)\} c_m(t) dt \quad (3.17)$$

where the noise is ignored for simplicity. When the WH code, whose elements are all +1, is used for the m th slot-pair, (3.16) and (3.17) are reduced to

$$S'_{m,d}(t) = g\alpha_m f_{m\phi} s_m(t) (1 + \delta_m^{(+)}) \quad (3.18)$$

$$S'_{m,i}(t) = \frac{g}{2} \sum' \alpha_{m'} f_{m'\phi} s_{m'}(t) (\delta_{m'}^{(+)} - \delta_{m'}^{(-)}). \quad (3.19)$$

According to (3.19), $S'_{m,i}(t) \neq 0$ when $\delta_{m'}^{(+)} \neq \delta_{m'}^{(-)}$, and thus the recovered signal suffers from the remaining interference from other slot-pairs' signals. However, when the WH codes, including both +1 and -1, are used for the m th slot pair, (3.16) and (3.17) are given by

$$S'_{m,d}(t) = g\alpha_m f_{m\phi} s_m(t) \left(1 + \frac{\delta_m^{(+)} + \delta_m^{(-)}}{2} \right) \quad (3.20)$$

$$S'_{m,i}(t) = 0. \quad (3.21)$$

Because $c_m(t)$ and $c_{m'}(t) \{1 + \delta_{m'}(t)\}$ are also orthogonal when $m \neq 1$ (see Appendix B.1), the interference from other slot-pairs' signals disappear in (3.21). Therefore, in an actual situation where the PIN diodes have errors in their electrical properties, the WH codes including both +1 and -1 must be used, with a length of $L_c > M$, to prevent deterioration in recovered signals.

3.5.2 Behavior at Oblique Incidence

Next, we consider that a plane wave arrives at the antenna with an incident angle (θ, ϕ) out of the y - z plane in Fig. 3.2. In this case, the received signal $r_m''(t)$ is given by

$$r_m''(t) = c_m(t) f_m e^{\pm j k_0 d_s \sin \theta \cos \phi} s_m(t) e^{j 2 \pi f_c t} \quad (3.22)$$

where $f_m = |\mathbf{f}_m(\theta, \phi)|$, and the polarization of the antenna is assumed to be matched to that of the incident plane wave. In (3.22), the factor $e^{\pm j k_0 d_s \sin \theta \cos \phi}$ causes an error depending on the operating state of the slot pair for the received signals. The effect of this error can be evaluated with (3.20) by replacing $\delta_m^{(\pm)}$ and $f_{m\phi}$ by $(e^{\pm j k_0 d_s \sin \theta \cos \phi} - 1)$ and f_m , respectively, and the recovered signal becomes

$$S_{m,d}''(t) = g \alpha_m f_m s_m(t) \cos(k_0 d_s \sin \theta \cos \phi). \quad (3.23)$$

The factor $\cos(k_0 d_s \sin \theta \cos \phi)$ introduces an additional loss into the recovered signal depending on the incident angle. Figure 3.11 shows calculated losses with various incident angles for the prototype antenna. The prototype antenna suffers from larger losses when ϕ is around 0° (for example, the loss of 0.43 dB for $(\theta, \phi) = (45^\circ, 0^\circ)$). Slot pairs with a smaller slot offset should be used to mitigate this loss, and it requires miniaturizing slot pairs. Slot offset is closely related to slot admittance and mounting space for the inductors and capacitors. Therefore, miniaturization of the slot pair is also one of the future challenges.

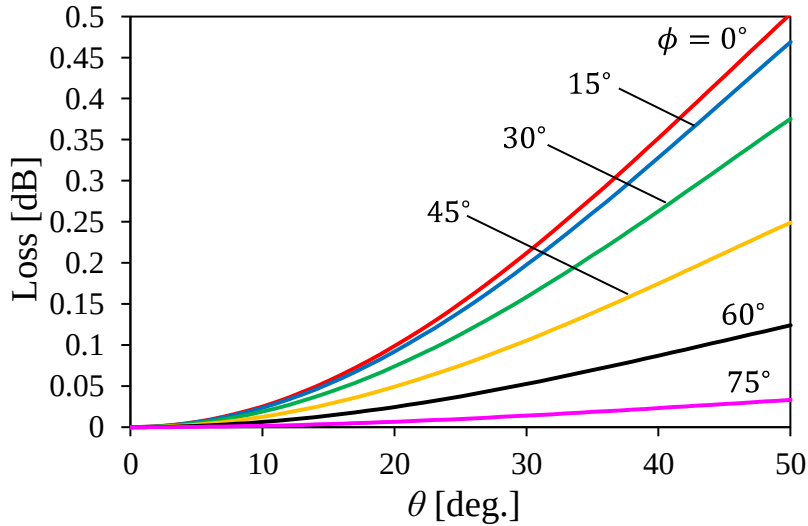


Figure 3.11: Calculated losses due to the factor $\cos(k_0 d_s \sin \theta \cos \phi)$.

3.5.3 Application Example: DOA Estimation

Simulation tests of DOA estimation are performed to demonstrate the feasibility of DBF function by the prototype antenna. In the simulation, the one-dimensional MUSIC al-

3. Waveguide Slot Array with Code-Division Multiplexing Function for Single RF Chain Digital Beamforming

gorithm [3.26] is employed, and the received signal of the antenna is calculated using the measured S_{21} . Thus, received signals at the l th state of the prototype antenna are calculated by

$$R_b(t) = \sum_{k=1}^K s^{(k)}(t, \theta_k) E_l(\theta_k) + n(t) \quad (3.24)$$

where $s^{(k)}(t, \theta_k)$ represents the k th incident signal to the antenna, θ_k is its incident angle, and K is the number of the incident signals. Recovered signals $S_m(t)$ are calculated by (3.5), and the MUSIC spectrum is then calculated by [3.25]

$$P_{MU}(\theta) = \frac{1}{\mathbf{a}^H(\theta) \mathbf{E}_n \mathbf{E}_n^H \mathbf{a}(\theta)} \quad (3.25)$$

where the superscript H represents the Hermitian transpose operation, and

$$\mathbf{a}(\theta) = [C_1 e^{-jk_0 d \sin \theta}, \dots, C_M e^{-jk_0 M d \sin \theta}].$$

C_m are calibration weights and are determined using a reference signal from $\theta = 0^\circ$, ignoring the mutual coupling effect for simplicity. $\mathbf{E}_n$ is a matrix whose columns are the noise eigenvectors of the covariance matrix $\mathbf{R} = E[\mathbf{S}(t) \mathbf{S}^H(t)]$ where $\mathbf{S}(t) = [S_1(t), \dots, S_M(t)]^T$, and the superscript T represents the transpose operation.

In this simulation, the case in which two uncorrelated QPSK signals are incident with the same power is considered. The incident angles are $(\theta_1, \theta_2) = (-10.0^\circ, 2.0^\circ)$ is considered. The operating state of the antenna is switched 16 times during each symbol according to the WH codes with $L_c = 16$. The synchronization between spreading and despreading is assumed to be perfectly achieved. The covariance matrix $\mathbf{R}$ is calculated using recovered signals of 100 symbols. Figure 3.12 shows the calculated MUSIC spatial spectrum for SNR = 20 dB. Two obvious peaks are observed in the spectrum, and the estimated incident angles are -10.0° and 1.8° , which agree well with the actual ones. The estimation error would be due to an effect of mutual couplings between slot pairs and imperfection in demultiplexing.

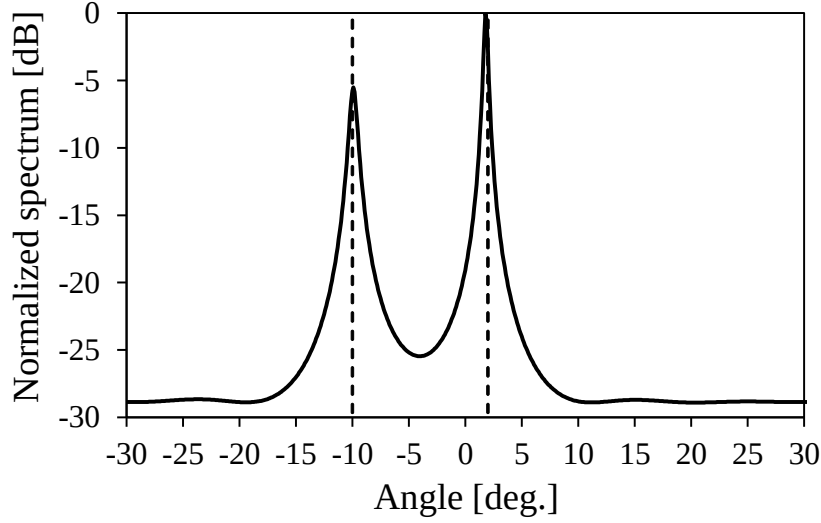


Figure 3.12: Simulated MUSIC spatial spectrum for two uncorrelated incident signals with $(\theta_1, \theta_2) = (-10.0^\circ, 2.0^\circ)$ and SNR = 20 dB.

3.5.4 Evaluation of Reduction in Cost and Power Consumption

The proposed antenna is evaluated in terms of its estimated cost and power consumption, and the potential reduction from the conventional element-level DBF array is quantified. Figure 3.13(a) and (b) show the estimated costs and power consumption, respectively, of ADCs and PIN diodes for a 128-element array antenna with a signal bandwidth of 2.5 MHz. In this estimation, each ADC has 14-bit resolution, and the costs and power consumption of ADCs and PIN diodes are based on those of commercially available ones [3.24], [3.27]. For the sake of simplicity, the component costs and power consumption presented here include only those of the ADCs and the PIN diodes. The overall component cost and power consumption can be reduced by increasing the multiplexing number and reducing the number of ADCs. However, the cost and power consumption of ADCs typically increase as the sampling rate increases. Therefore, an excessive increase in the multiplexing numbers will result in an increase in cost and power consumption. Consequently, an optimal multiplexing number exists in terms of cost and power consumption, which is 16 as shown in Fig. 3.13. In practice, however, the multiplexing number is limited by the switching speed of the PIN diodes (see Appendix B.3), and the feasible optimal multiplexing number would become eight. In this case, the component cost and power consumption are approximately 35% and 62% of that of the conventional element-level DBF array, respectively. This demonstrates the effectiveness of the proposed antenna in reducing hardware costs and power consumption.

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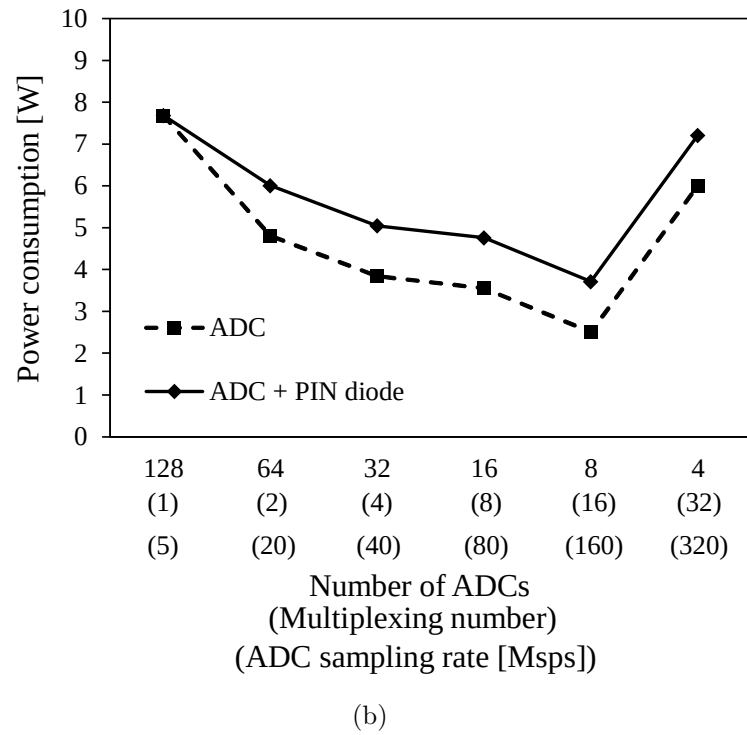
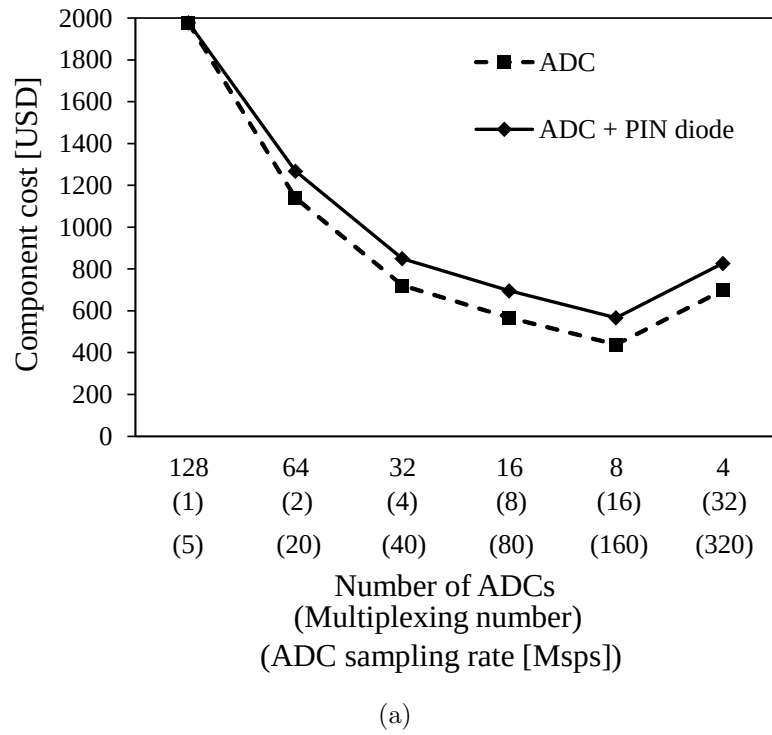


Figure 3.13: Estimated (a) hardware costs and (b) power consumption.

3.6 Conclusion

In this chapter, a novel waveguide slot array with slot pairs for the CDM function was presented. The received signals of each slot pair are multiplexed in the CDM fashion by switching the diodes' bias according to the WH codes, and they are recovered through the despreading process in the digital domain for further DSP. The operating principle of the proposed antenna was also explained, and the prototype antenna with eight slot pairs were then designed, fabricated, and tested as a proof of concept. Although the PIN diodes' bias were manually switched, the transient characteristics of the PIN diodes were ignored, and the perfect synchronization between spreading and despreading was assumed in the experiment, the experimental results demonstrated the feasibility of the proposed antenna.

Future works of the proposed antenna include further tests considering the transient characteristics of the PIN diodes and actual synchronization issues, bandwidth enhancement, radiation efficiency improvement, and miniaturization of the slot pair for mitigating the losses depending on the incident angle of signals. Although the proposed antenna has the disadvantage of lower SNR than the typical element-level DBF array antennas, it has the potential to offer a DBF antenna with the CDM-based path-sharing architecture at a lower cost and higher radiation efficiency.

Reference

- [3.1] C. Fulton, M. Yeary, D. Thompson, J. Lake, and A. Mitchell, “Digital phased arrays: challenges and opportunities,” *Proc. IEEE*, vol. 104, no. 3, pp. 487–503, March 2016.
- [3.2] P. K. Bailleul, “A new era in elemental digital beamforming for spaceborne communications phased arrays,” *Proc. IEEE*, vol. 104, no. 3, pp. 623–632, March 2016.
- [3.3] B. Yang, Z. Yu, J. Lan, R. Zhang, J. Zhou, and W. Hong, “Digital beamforming-based massive MIMO transceiver for 5G millimeter-wave communications,” *IEEE Trans. Microw. Theory Tech.*, vol. 66, no. 7, pp. 3403–3418, July 2018.
- [3.4] J. D. Fredrick, Y. Wang, and T. Itoh, “Smart antennas based on spatial multiplexing of local elements (SMILE) for mutual coupling reduction,” *IEEE Trans. Antennas Propag.*, vol. 52, no. 1, pp. 106–114, Jan. 2004.
- [3.5] K. Koga, N. Kikuma, H. Hirayama, K. Sakakibara, T. Koike, H. Iwasaki, and Y. Mizuno, “A study of switching methods for an adaptive array with a single receiver using time-division multiplexing,” *Trans. IEICE B*, vol. J96-B, no. 2, pp. 149–162, Feb. 2013 (in Japanese).
- [3.6] S. Henault, B. R. Jackson, and Y. M. M. Antar, “Compensation of time-division multiplexing distortion in switched antenna arrays with a single RF front-end and digitizer,” *IEEE Trans. Antennas Propag.*, vol. 61, no. 8, pp. 4383–4388, Aug. 2013.
- [3.7] Y. Yashchyshyn, K. Derzakowski, P. R. Bajurko, J. Marczewski, and S. Kozłowski, “Time-modulated reconfigurable antenna based on integrated S-PIN diodes for mm-wave communication systems,” *IEEE Trans. Antennas Propag.*, vol. 63, no. 9, pp. 4121–4131, Sept. 2015.
- [3.8] M. A. Johnson, “Phased-array beam steering by multiplex sampling,” *Proc. IEEE*, vol. 56, no. 11, pp. 1801–1811, Nov. 1968.
- [3.9] N. Akram et al., “Frequency-multiplexed array digitization for MIMO receivers: 4-antennas/ADC at 28 GHz on Xilinx ZCU-1285 RF SoC,” *IEEE Access*, vol. 9, pp. 142743–142753, 2021.
- [3.10] A. Jahanian, F. Tzeng, and P. Heydari, “Code-modulated path-sharing multi-antenna receivers: theory and analysis,” *IEEE Trans. Wireless Commun.*, vol. 8, no. 5, pp. 2193–2201, May 2009.
- [3.11] F. Tzeng, A. Jahanian, D. Pi, and P. Heydari, “A CMOS code-modulated path-sharing multi-antenna receiver front-end,” *IEEE J. Solid-State Circuits*, vol. 44, no.

- 5, pp. 1321–1335, May 2009.
- [3.12] E. A. Alwan, A. Akhiyat, M. LaRue, W. Khalil, and J. L. Volakis, “Low cost, power efficient, on-site coding receiver (OSCR) for ultra-wideband digital beamforming,” 2013 IEEE Int. Symp. Phased Array Systems Tech., pp. 202–206, 2013.
- [3.13] S. B. Venkatakrishnan, D. K. Papantonis, A. A. Akhiyat, E. A. Alwan, and J. L. Volakis, “Experimental validation of on-site coding digital beamformer with ultra-wideband antenna arrays,” IEEE Trans. Microw. Theory Tech., vol. 65, no. 11, pp. 4408–4417, Nov. 2017.
- [3.14] J. Zhang, W. Wu, and D. -G. Fang, “Single RF channel digital beamforming multi-beam antenna array based on time sequence phase weighting,” IEEE Antennas Wireless Propag. Lett., vol. 10, pp. 514–516, 2011.
- [3.15] X. Wang and C. Caloz, “Direction-of-Arrival (DOA) Estimation based on Spacetime-Modulated Metasurface,” 2019 IEEE Int. Symp. Antennas Propag. and USNC-URSI Radio Science Meeting, pp. 1613–1614, 2019.
- [3.16] X. Wang and C. Caloz, “Pseudorandom sequence (space–) time-modulated metasurfaces: principles, operations, and applications, IEEE Antennas Propag. Mag., vol. 64, no. 4, pp. 135–144, Aug. 2022.
- [3.17] Y. Yashchyshyn, J. Marczewski, K. Derzakowski, J. W. Modelski, and P. B. Grabciec, “Development and investigation of an antenna system with reconfigurable aperture,” IEEE Trans. Antennas Propag., vol. 57, no. 1, pp. 2–8, Jan. 2009.
- [3.18] Z. Zhou, H. Su, B. Yang, S. Wu, and X. Y. Zhang, “Digital encoding holographic antenna based on SIW waveguide,” 2021 IEEE Int. Workshop on Electromagnetics: Applications and Student Innovation Competition (iWEM), 2021.
- [3.19] H. Li, S. Li, B. Hou, X. Zhang, W. Wen, and C. Hu. “A digital SIW-slot antenna array with FPGA implementation of beamforming,” Sci. Rep. 12, no. 8927, May 2022.
- [3.20] R. E. Collin, “Waveguides and cavities,” in Field Theory of Guided Waves , pp.329–410, IEEE, 1991.
- [3.21] E. H. Dinan and B. Jabbari, “Spreading codes for direct sequence CDMA and wide-band CDMA cellular networks,” IEEE Commun. Mag., vol. 36, no. 9, pp. 48–54, Sept. 1998.
- [3.22] E. A. Alwan, S. B. Venkatakrishnan, A. A. Akhiyat, W. Khalil, and J. L. Volakis, “Code Optimization for a Code-Modulated RF Front End,” IEEE Access, vol. 3, pp. 260–273, 2015.

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- [3.23] Skyworks, “Design with PIN diodes,” Skyworks Solutions, Inc., https://www.skyworksinc.com/-/media/SkyWorks/Documents/Products/1-100/Design_With_PIN_Diodes_200312E.pdf, accessed June 15. 2024.
- [3.24] Skyworks, “SMP1345 Series: Very Low Capacitance Plastic Packaged Silicon PIN Diodes,” Skyworks Solutions, Inc., <https://www.skyworksinc.com/Products/Diodes/SMP1345-Series>, accessed June 15. 2024.
- [3.25] 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.
- [3.26] R. Schmidt, “Multiple emitter location and signal parameter estimation,” *IEEE Trans. Antennas Propag.*, vol. 34, no. 3, pp. 276–280, March 1986.
- [3.27] Analog Devices, Inc., “High Speed A/D Converters ≥ 10 MSPS,” https://www.analog.com/en/parametricsearch/10826#/, accessed June 15. 2024.

Chapter 4

Axial Ratio Improvement for Wide-angle Scanning Circularly-Polarized Phased Array of Circular Patch Antennas

4.1 Introduction

Circularly-polarized (CP) antennas have been used in various applications: global navigation satellite systems, satellite communication (SATCOM), wireless body area network systems, vital sensing systems, and microwave power transmission (MPT) systems [4.1]–[4.6]. In particular, user terminals for SATCOM and MPT systems for multiple and moving electronic devices require CP antennas with beam-scanning capability. Although conventional mechanical beam-scanning reflector antennas have advantages in high-gain with high-efficiency, low axial ratio (AR), simple structure, and low cost, they are restricted to their installation locations because of their bulky structures. Therefore, CP phased arrays with low-profile and electronically beam-scanning capability have been actively developed recently [4.7]–[4.12]. In CP phased arrays, AR is an important performance metric as well as bandwidth, gain, and side-lobe level. A poor AR causes a cross-polarization interference which is a critical concern in a dual polarization communication system. Especially for user terminals for SATCOM, a maximum AR is specified in some standards [4.13], [4.14] to prevent interference. A poor AR also causes a polarization mismatch loss and decreases the power transmission efficiency, which is a significant concern in MPT systems. The AR tends to degrade away from the broadside direction of an antenna due to the imbalance in radiated fields with the vertical polarization (V-pol.) and the horizontal one (H-pol.). Therefore, CP phased arrays suffer from the AR degradation, especially when the beam is scanned to wide-angle directions. Specifically, the AR is typically degraded to 5–10 dB at 60° scan [4.7]–[4.9].

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Several studies have been reported in antennas with low AR for wide-angle scanning CP phased arrays [4.15]–[4.26]. They can be categorized into four approaches: antenna element designs [4.15]–[4.17], sequential rotation techniques [4.18], [4.19], polarizers [4.20], [4.21], and calibration or compensation techniques [4.22]–[4.26]. Although the first three approaches are all effective and necessary for achieving low AR at wide-angle directions, the AR performance is fixed once the antenna design and fabrication are completed. Thus, redesign and refabrication will be required when the AR is deteriorated by fabrication errors. However, compensation techniques are applicable and effective even if the AR is deteriorated after antenna fabrication. In [4.22] and [4.23], excitation phases of each antenna element are adjusted so that the cross-polarized fields of each antenna element cancel each other out. Although this method can be applied to a phased array using single-polarized antenna elements, an optimization procedure is required at every scan angle to obtain appropriate excitation phases. In [4.24]–[4.26], dual-polarized antenna elements with two feeding channels are used in phased arrays, and excitation amplitudes and phases (compensation weights) of each feeding channels are adjusted to compensate differences in amplitude and phase of V-pol. and H-pol. fields. This method requires a polarization reconfigurable function for each antenna element, which currently can be relatively easily obtained by using commercially available multi-channel beamforming radio frequency integrated circuits (RFICs) [4.27]. In [4.24] and [4.25], following the setting of the phases for beam scanning, the V-pol. and H-pol. fields of the array are measured in the scanning direction, and the compensation weights of each feeding channel are determined by ignoring the cross-polarization component radiated from each feeding port. Therefore, its effectiveness will be insufficient when the cross-polarization caused by the mutual coupling between the feeding ports is significant. On the other hand, in [4.26], the compensation weights of each feeding channel are determined based on the full polarization measurement, the measurement of the V-pol. and H-pol. fields of each feeding channel, which improves AR more effectively. Although the appropriate compensation weights can be easily obtained from the measured full polarization information, they vary with direction. Consequently, the full polarization measurement should be conducted for each scan angle to implement this technique. This is a time-consuming task and a practical challenge for wide-angle scanning phased arrays.

In this study, a novel calibration technique using a closed-form design formula for compensation weights is presented to solve the aforementioned problem. This study particularly focuses on a CP phased array of dual-polarized circular patch antennas, which are a promising option for the antenna element of the phased array due to their low-profile structure, ease of fabrication, and low cost [4.10], [4.28]–[4.31]. The proposed design formula is obtained by a fitting function of an active element pattern (AEP) of dual-polarized circular patch antenna considering the effects of the mutual coupling between

antenna elements and the higher order modes of the patch. The proposed fitting function is constructed with characteristic basis function patterns (CBFPs) [4.32] expressed in cylindrical Bessel functions, which will offer an efficient approximation of the AEP of the circular patch antenna. In addition, the proposed fitting function is determined with the measured AEPs of the CP fields at a reduced number of measurement points. Consequently, the proposed technique eliminates the necessity for the individual measurement of each feeding port of the antenna at each scan angle. This will result in a reduction in the measurement effort required for its implementation.

The rest of this chapter is organized as follows. Section 4.2 explains the theory of the proposed technique; Section 4.3 presents the simulation results to evaluate calculation conditions for the design formula and the effectiveness of the proposed technique; Section 4.4 presents the experimental results using a 1024-element phased array; and Section 4.5 presents the conclusions of this chapter.

4.2 Principle of the Proposed Technique

4.2.1 Antenna Configuration

Figure 4.1(a) shows a configuration of a CP phased array system in this study. The system comprises dual-polarized circular patch antennas with two feeding ports, variable gain amplifier (VGA) and digital phase shifter connected to each feeding port, and a power dividing network. The amplitude and phase of transmitted fields from each feeding port can be adjusted individually by the VGA and digital phase shifter. A configuration of the dual-polarized circular patch antenna is shown in Figs. 4.1(b)–(d) [4.30]. The lower circular patch is fed by two feeding probes 1 and 2 located at $\phi = \phi_1$ and $\phi = \phi_2 = \phi_1 + \pi/2$, respectively. The stacked circular patch is a parasitic patch, and a spacer with a cylindrical hole for a conducting cavity supports the substrate for the stacked patch. A portion of the sidewall of the cavity is formed by via-hole array in the substrates. The patch antenna is enclosed in a cavity to suppress the mutual coupling [4.33] and has a superstrate for wide-angle impedance matching [4.34].

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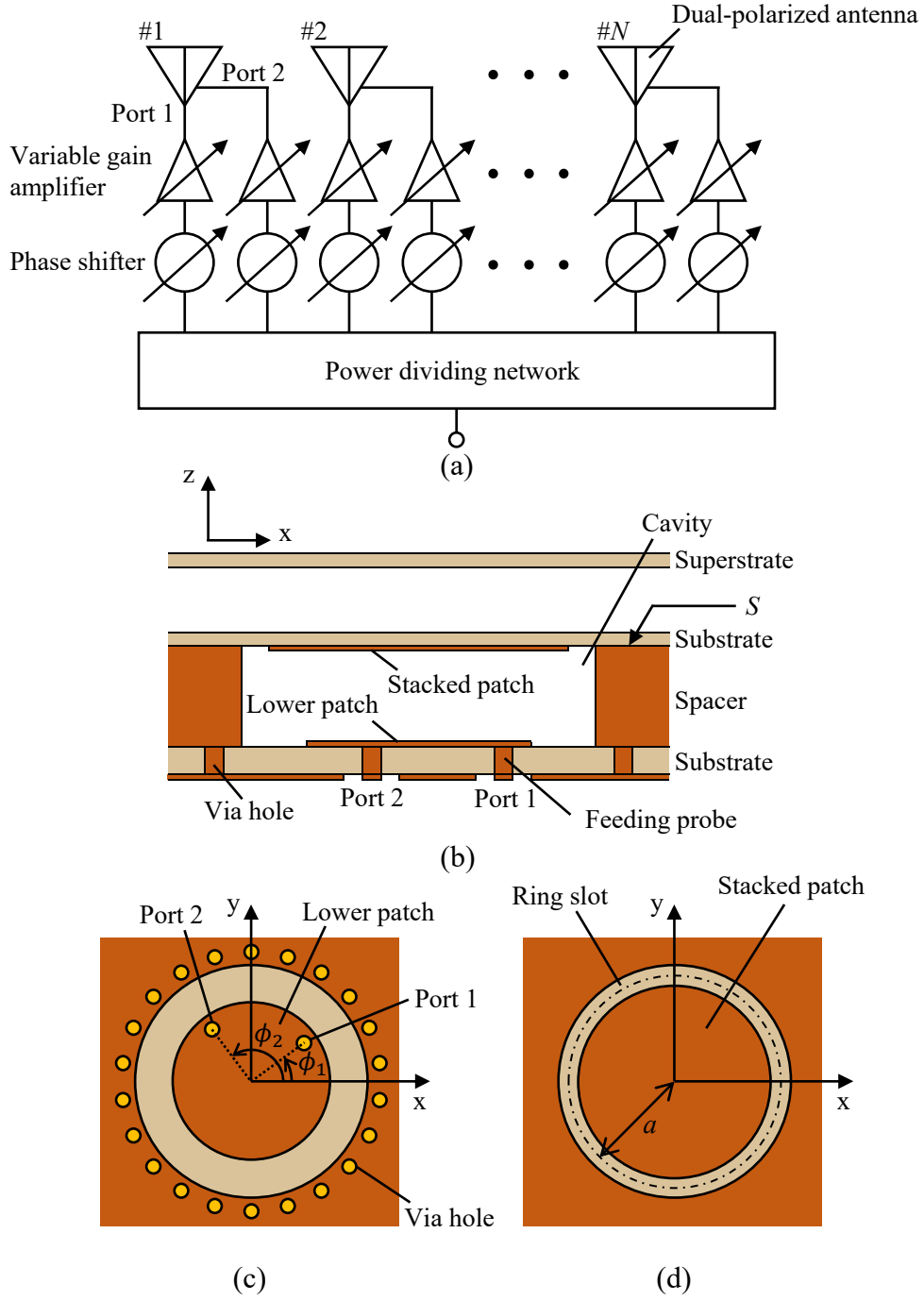


Figure 4.1: Configuration of a CP phased array system. (a) Block diagram, (b) side view of the dual-polarized circular patch antenna, (c) top view of the lower patch layer, and (d) top view of the stacked patch layer.

4.2.2 Construction of a Fitting Function of the Active Element Pattern

Now consider a phased array sufficiently large that an AEP can be approximated as the same for all antenna elements in the array; thus, an array pattern can be approximated by a product of an array factor and the AEP. In this case, the AR of the phased array at a certain scan angle is the same as that of the AEP at the same angle. In this study, therefore, the AEP is first formulated, and the expression for compensation weights is derived based on the formulated AEP.

First consider a single circular patch antenna, only with the feeding probe 1 at $\phi = \phi_1$ and without the superstrate, is recessed in an infinite ground plane, which is coincident with the surface S on the stacked patch. The substrate for the stacked patch is assumed to be considerably thin that its effect on the radiation can be ignored. Applying the Love's equivalence principle together with the image theory, the radiation from the patch antenna can be approximately calculated by the radiation in free-space from an equivalent magnetic current loop on the ring slot. When the feeding port 1 is excited with a unit amplitude current, the equivalent magnetic current loop M is expressed by the following Fourier cosine series:

$$M(\phi) = \sum_{n=0}^N M_n \cos n(\phi - \phi_1) \quad (4.1)$$

where M_n is the mode coefficient of the n th mode and N is the maximum mode number. The far-field of the antenna at (θ, ϕ) in a spherical coordinate system, whose origin is coincident with the center of the equivalent magnetic current loop, can be approximated by [4.35]

$$e_{\theta}^s(\phi_1) = -e_0 \sum_{n=0}^N j^n M_n \cos n(\phi - \phi_1) J_n^{(-)}(\theta) \quad (4.2)$$

$$e_{\phi}^s(\phi_1) = e_0 \cos \theta \sum_{n=1}^N j^n M_n \sin n(\phi - \phi_1) J_n^{(+)}(\theta) \quad (4.3)$$

$$J_n^{(+)}(\theta) = J_{n-1}(k_0 a \sin \theta) + J_{n+1}(k_0 a \sin \theta) \quad (4.4)$$

$$J_n^{(-)}(\theta) = J_{n-1}(k_0 a \sin \theta) - J_{n+1}(k_0 a \sin \theta) \quad (4.5)$$

where the subscripts θ and ϕ indicate the θ and ϕ components of the field, respectively, e_0 is an amplitude coefficient, a is the radius of the ring slot, k_0 is the free-space wavenumber, and $J_n(x)$ is the Bessel function of the first kind of the n th order. Similarly, the far-field of the antenna only with the feeding probe 2 can be obtained by replacing ϕ_1 in (4.2) and (4.3) with ϕ_2 . $e_{\theta}^s(\phi_1)$, $e_{\theta}^s(\phi_2)$, $e_{\phi}^s(\phi_1)$, and $e_{\phi}^s(\phi_2)$ are the primary CBFPs for the antenna under consideration. Considering reradiation due to the mutual coupling between the

4. Axial Ratio Improvement for Wide-angle Scanning Circularly-Polarized Phased Array of Circular Patch Antennas

feeding ports, we obtain the following isolated far-field of the dual-polarized circular patch antenna:

$$\begin{aligned} e_{\theta}^i &= (I_1 + r_0 I_2) e_{\theta}^s(\phi_1) + (r_0 I_1 + I_2) e_{\theta}^s(\phi_2) \\ &= -e_0 \sum_{n=0}^N j^n M_n \left[(I_1 + r_0 I_2) \cos n(\phi - \phi_1) \right. \\ &\quad \left. + (r_0 I_1 + I_2) \cos n(\phi - \phi_2) \right] J_n^{(-)}(\theta) \end{aligned} \quad (4.6)$$

$$\begin{aligned} e_{\phi}^i &= (I_1 + r_0 I_2) e_{\phi}^s(\phi_1) + (r_0 I_1 + I_2) e_{\phi}^s(\phi_2) \\ &= e_0 \cos \theta \sum_{n=1}^N j^n M_n \left[(I_1 + r_0 I_2) \sin n(\phi - \phi_1) \right. \\ &\quad \left. + (r_0 I_1 + I_2) \sin n(\phi - \phi_2) \right] J_n^{(+)}(\theta) \end{aligned} \quad (4.7)$$

where the feeding ports 1 and 2 are simultaneously excited with I_1 and I_2 , respectively, and r_0 is a coupling coefficient, which represents the reradiation due to the mutual coupling between the feeding ports 1 and 2.

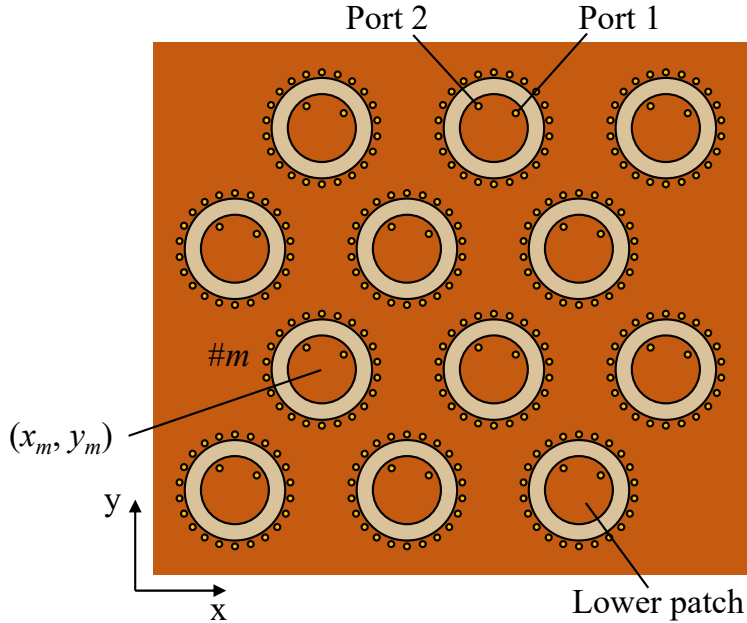


Figure 4.2: Arrangement of the array of the dual-polarized circular patch antennas (the stacked patches and the superstrate are ignored).

Subsequently, an array of the dual-polarized circular patch antennas without the superstrate is considered. The antenna elements are uniformly arranged with the same orientation as Fig. 4.1(c), as shown in Fig. 4.2. Although the antenna elements are arranged in a triangular lattice in Fig. 4.2, they can be arranged in a rectangular one. When

4. Axial Ratio Improvement for Wide-angle Scanning Circularly-Polarized Phased Array of Circular Patch Antennas

only one antenna element is excited and the rest are terminated, the AEP can be approximated by a sum of the radiation from the excited antenna element and the reradiation component from the neighboring M_e elements due to mutual couplings. Assuming that all antenna elements have the same radiation mode, i.e., M_n is the same for all antenna elements, we obtain the following fitting function:

$$\begin{aligned} e_\theta^a &= \sum_{m=1}^{M_e} e^{j\psi_m(\theta, \phi)} \left[(r_{11}^{(m)} I_1 + r_{12}^{(m)} I_2) e_\theta^s(\phi_1) + (r_{21}^{(m)} I_1 + r_{22}^{(m)} I_2) e_\theta^s(\phi_2) \right] \\ &= -e_0 \sum_{m=1}^{M_e} e^{j\psi_m(\theta, \phi)} \sum_{n=0}^N j^n M_n \left[(r_{11}^{(m)} I_1 + r_{12}^{(m)} I_2) \cos n(\phi - \phi_1) \right. \\ &\quad \left. + (r_{21}^{(m)} I_1 + r_{22}^{(m)} I_2) \cos n(\phi - \phi_2) \right] J_n^{(-)}(\theta) \end{aligned} \quad (4.8)$$

$$\begin{aligned} e_\phi^a &= \sum_{m=1}^{M_e} e^{j\psi_m(\theta, \phi)} \left[(r_{11}^{(m)} I_1 + r_{12}^{(m)} I_2) e_\phi^s(\phi_1) + (r_{21}^{(m)} I_1 + r_{22}^{(m)} I_2) e_\phi^s(\phi_2) \right] \\ &= e_0 \cos \theta \sum_{m=1}^{M_e} e^{j\psi_m(\theta, \phi)} \sum_{n=1}^N j^n M_n \left[(r_{11}^{(m)} I_1 + r_{12}^{(m)} I_2) \sin n(\phi - \phi_1) \right. \\ &\quad \left. + (r_{21}^{(m)} I_1 + r_{22}^{(m)} I_2) \sin n(\phi - \phi_2) \right] J_n^{(+)}(\theta) \end{aligned} \quad (4.9)$$

where $\psi_m(\theta, \phi) = k_0(x_m \cos \phi \sin \theta + y_m \sin \phi \sin \theta)$, (x_m, y_m) is the position of the m th antenna element, and $r_{kl}^{(m)}(k, l = 1, 2)$ are coupling coefficients, which represent reradiation components due to the mutual coupling between the feeding probe l of the excited antenna element (the first antenna element) and the feeding probe k of the m th antenna element. For $m = 1$, $r_{11}^{(1)} = r_{22}^{(1)} = 1$ and $r_{21}^{(1)} = r_{12}^{(1)} = r_0$.

Here, the superstrate is introduced to the array. The superstrate reduces the variation of the active impedance with scan angles and also affects the radiation pattern (see Appendix C.1). For the effects on the AEP, modification functions are introduced into (4.8) and (4.9) as follows [4.36]:

$$E_\theta = e_\theta^a f_\theta(\theta); \quad E_\phi = e_\phi^a f_\phi(\theta) \quad (4.10)$$

where $f_\theta(\theta) = \cos^\alpha \theta$, $f_\phi(\theta) = \cos^\beta \theta$, and α and β are parameters to be determined. Finally, the fitting functions for the CP fields can be calculated by

$$E_R = \frac{1}{\sqrt{2}} (E_\theta + jE_\phi); \quad E_L = \frac{1}{\sqrt{2}} (E_\theta - jE_\phi) \quad (4.11)$$

where the subscripts R and L indicate the right-hand circular polarization (RHCP) and left-hand circular polarization (LHCP) components of the field, respectively.

To calculate the fitting function using (4.8)–(4.11), the unknown parameters, M_n , $r_{kl}^{(m)}$, α , and β , should be determined. Because it is difficult to analytically calculate or directly measure them in an actual phased array, they are determined to ensure that the calculated

fitting function approximates the measured or simulated AEPs. Specifically, a problem to minimize the difference between the calculated fitting functions and the measured or simulated AEPs, i.e., the following objective function, is considered.

$$F(M_n, r_{kl}^{(m)}, \alpha, \beta) = \sum_{p=1}^P \sum_{q=1}^Q \left[\left| E_R(\theta_p, \phi_q) - E_R^{(t)}(\theta_p, \phi_q) \right|^2 + \left| E_L(\theta_p, \phi_q) - E_L^{(t)}(\theta_p, \phi_q) \right|^2 \right] \quad (4.12)$$

where $E_R^{(t)}$ and $E_L^{(t)}$ are the measured or simulated complex AEPs, and P and Q are numbers of the sample directions θ_p and ϕ_q , respectively. This minimization problem can be solved by some nonlinear optimization algorithms, and the GlobalSearch function provided by MATLAB [4.37] is used in this study.

4.2.3 Design Formula for Compensation Weight

Compensation weights for AR improvement are determined to ensure that each AEP has the lowest AR, namely to satisfy the following equation [4.38]:

$$\text{AR} = \left| \frac{|E_R| + |E_L|}{|E_R| - |E_L|} \right| = 1. \quad (4.13)$$

This equation is equivalent to $E_L = 0$ for the RHCP excitation, setting the RHCP field as the co-polarization, and $E_R = 0$ for the LHCP excitation, setting the LHCP field as the co-polarization. Substituting (4.8)–(4.11) into these equivalent equations yields the following equation:

$$[c_{11} + c_{21} \pm (s_{11} + s_{21})] I_1 + [c_{12} + c_{22} \pm (s_{12} + s_{22})] I_2 = 0 \quad (4.14)$$

where the plus and minus signs are for the RHCP and LHCP excitations, respectively, and

$$c_{kl} = f_\theta(\theta) \sum_{m=1}^{M_e} r_{kl}^{(m)} e^{j\psi_m(\theta, \phi)} \sum_{n=0}^N j^n M_n \cos n(\phi - \phi_k) J_n^{(-)}(\theta) \quad (4.15)$$

$$s_{kl} = \cos \theta f_\phi(\theta) \sum_{m=1}^{M_e} r_{kl}^{(m)} e^{j\psi_m(\theta, \phi)} \sum_{n=0}^N j^{n+1} M_n \sin n(\phi - \phi_k) J_n^{(+)}(\theta) \quad (4.16)$$

for $k, l = 1, 2$. Because (4.14) is an indeterminate equation with infinitely many solutions, it is necessary to impose a constraint to uniquely specify a solution. The amplitude control is usually accomplished by reducing the gain of the VGA. This results in undesirable reduction in transmitting power. Therefore, the amplitude and phase of only one feeding channel are adjusted so that the amplitude control becomes minimal. According

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to the magnitude relationship between the coefficients of I_1 and I_2 in (4.14), the desired excitation is determined as follows:

$$\begin{aligned} & \text{for } |c_{11} + c_{21} \pm (s_{11} + s_{21})| \geq |c_{12} + c_{22} \pm (s_{12} + s_{22})|, \\ & I_1 = -\frac{c_{12} + c_{22} \pm (s_{12} + s_{22})}{c_{11} + c_{21} \pm (s_{11} + s_{21})} I_2 \end{aligned} \quad (4.17)$$

where I_2 is set to a nominal excitation value, and

$$\begin{aligned} & \text{for } |c_{11} + c_{21} \pm (s_{11} + s_{21})| < |c_{12} + c_{22} \pm (s_{12} + s_{22})|, \\ & I_2 = -\frac{c_{11} + c_{21} \pm (s_{11} + s_{21})}{c_{12} + c_{22} \pm (s_{12} + s_{22})} I_1 \end{aligned} \quad (4.18)$$

where I_1 is set to the nominal excitation value. The nominal excitation values for the RHCP excitation are $I_1 = j$ and $I_2 = 1$, and the compensation weights are determined by

$$\begin{aligned} & \text{for } |c_{11} + c_{21} + s_{11} + s_{21}| \geq |c_{12} + c_{22} + s_{12} + s_{22}|, \\ & C_1 = j \frac{c_{12} + c_{22} + s_{12} + s_{22}}{c_{11} + c_{21} + s_{11} + s_{21}}, \quad C_2 = 1 \end{aligned} \quad (4.19)$$

$$\begin{aligned} & \text{for } |c_{11} + c_{21} + s_{11} + s_{21}| < |c_{12} + c_{22} + s_{12} + s_{22}|, \\ & C_1 = 1, \quad C_2 = -j \frac{c_{11} + c_{21} + s_{11} + s_{21}}{c_{12} + c_{22} + s_{12} + s_{22}}. \end{aligned} \quad (4.20)$$

Similarly, the nominal excitation values for the LHCP excitation are $I_1 = -j$ and $I_2 = 1$, and the compensation weights for the LHCP excitation are determined as follows:

$$\begin{aligned} & \text{for } |c_{11} + c_{21} - s_{11} - s_{21}| \geq |c_{12} + c_{22} - s_{12} - s_{22}|, \\ & C_1 = -j \frac{c_{12} + c_{22} - s_{12} - s_{22}}{c_{11} + c_{21} - s_{11} - s_{21}}, \quad C_2 = 1 \end{aligned} \quad (4.21)$$

$$\begin{aligned} & \text{for } |c_{11} + c_{21} - s_{11} - s_{21}| < |c_{12} + c_{22} - s_{12} - s_{22}|, \\ & C_1 = 1, \quad C_2 = j \frac{c_{11} + c_{21} - s_{11} - s_{21}}{c_{12} + c_{22} - s_{12} - s_{22}}. \end{aligned} \quad (4.22)$$

Figure 4.3 shows a flowchart of the proposed technique. The measured or simulated AEPs (Step 1) are first prepared, and then the unknown parameters M_n , $r_{kl}^{(m)}$, α , and β are determined by minimizing the objective function (4.12) using given AEPs (Step 2). Next, the scan angle (θ_s, ϕ_s) and the excitation condition are specified (Step 3), and c_{kl} and s_{kl} are calculated by (4.15) and (4.16) with $(\theta, \phi) = (\theta_s, \phi_s)$ (Step 4). Subsequently, the compensation weights C_1 and C_2 are calculated by (4.19) or (4.20) for the RHCP excitation and (4.21) or (4.22) for the LHCP excitation (Step 5). Finally, the gains of each VGA are reduced by $-20 \log_{10} |C_1|$ dB and $-20 \log_{10} |C_2|$ dB from the typical gains,

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and each phase shifter is controlled to make the phase shifts of $\arg(C_1)$ and $\arg(C_2)$ from the nominal phase shifts (Step 6). Once the unknown parameters are determined in Step 2, only Steps 3 through 6 need to be performed for a given scan angle and excitation condition.

In contrast to the conventional technique [4.26], the proposed technique determines the compensation weights using the measured AEPs of the circular-polarized fields, thereby eliminating the necessity for measurements of the V-pol. and H-pol. fields of each feeding port, as well as for measurements in each scanning direction. This will result in a reduction in the measurement effort required for its implementation by at least half.

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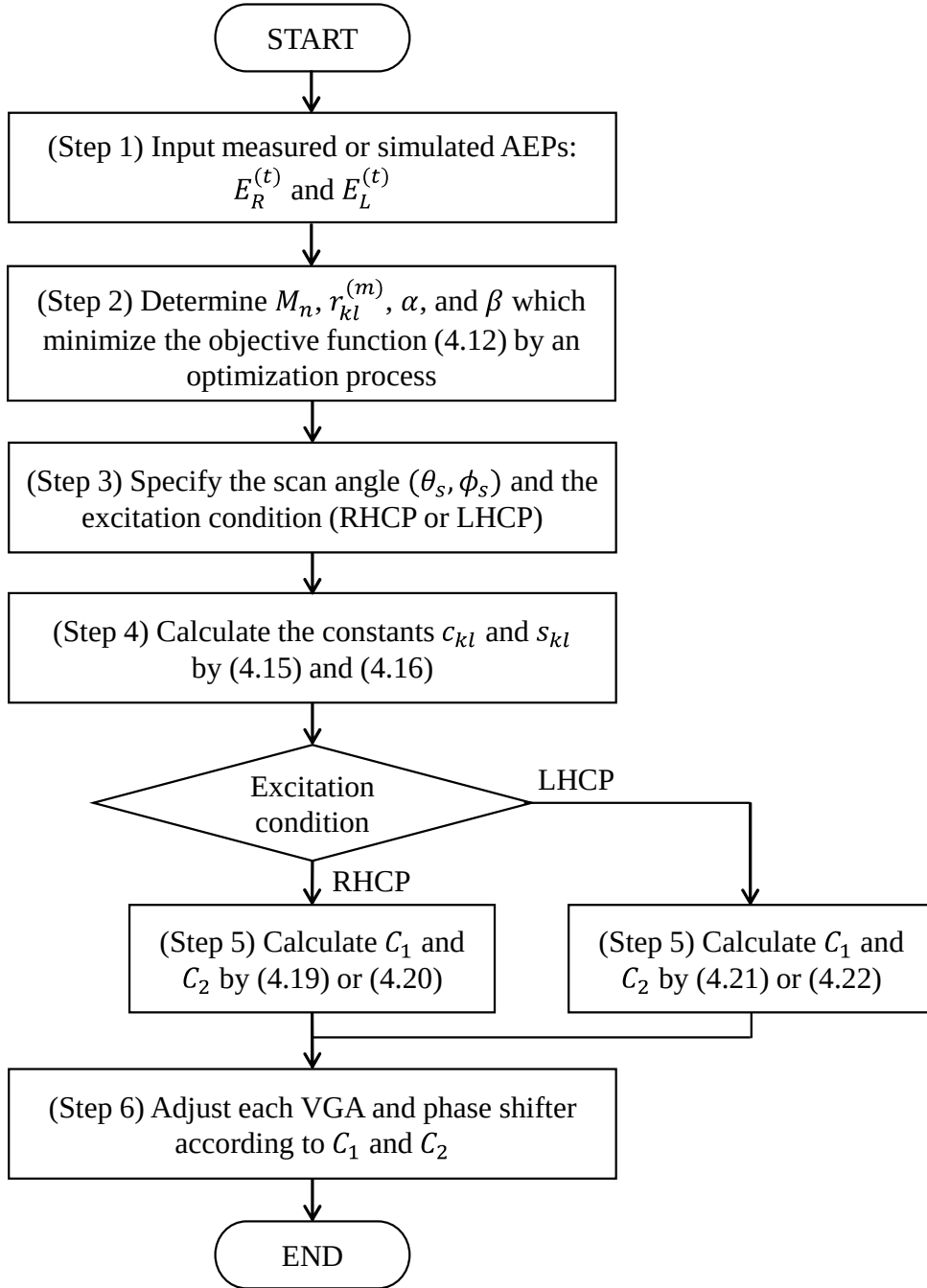


Figure 4.3: Flowchart of the proposed technique.

4.3 Simulation Results

The calculation condition and effectiveness of the proposed technique are evaluated using simulation results of a 41-element array of the dual-polarized circular patch antennas. Figure 4.4 shows a configuration of the array under simulation. The array is designed to scan the beam up to $\theta_s = \pm 70^\circ$. The antenna elements are arranged in an equilateral triangular lattice with an element spacing of $0.56\lambda_0$, where λ_0 is the free-space wavelength at the operating frequency f_0 . The antenna element has a stacked patch with a radius of $0.17\lambda_0$ and is enclosed in a conducting cavity with a radius of $0.23\lambda_0$; thus, the radius of the ring slot is $a = 0.20\lambda_0$. The lower patch is fed at $\phi_1 = 135^\circ$ and $\phi_2 = 225^\circ$. The substrate for the stacked patch has a dielectric constant of 3.3 and a thickness of $0.02\lambda_0$. The superstrate has a dielectric constant of 3.6 and a thickness of $0.02\lambda_0$, and is located at $0.26\lambda_0$ above the substrate for the stacked patch. The electromagnetic simulation is performed using Ansys HFSS. In order to determine M_n , $r_{kl}^{(m)}$, α , and β , the simulated AEPs when only the antenna element at the center of the array is fed for the RHCP excitation, $I_1 = j$ and $I_2 = 1$, are used.

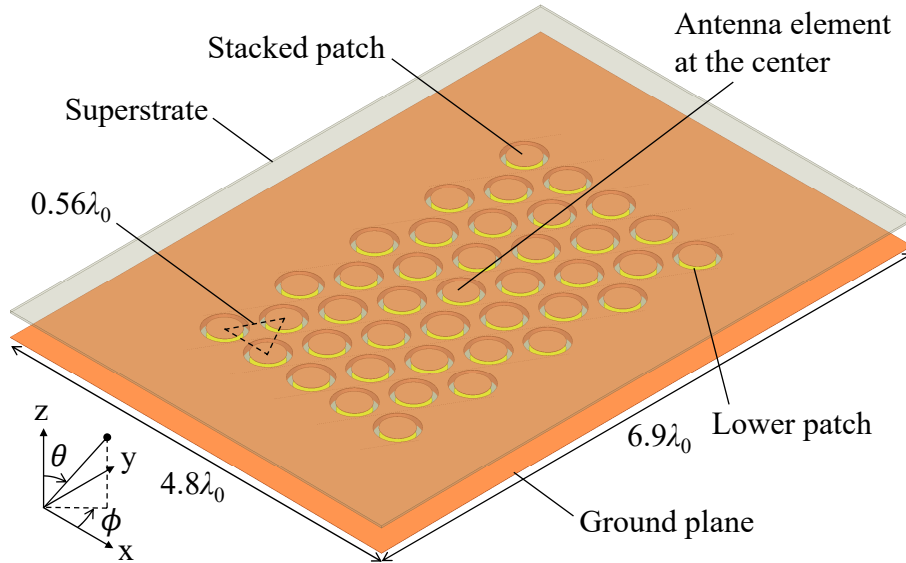


Figure 4.4: Configuration of the 41-element array in the simulation. The substrate for the stacked patches is omitted for simplicity.

4.3.1 Calculation Conditions for the Accurate Fitting Function

First, M_e used to calculate the fitting functions is considered. Figure 4.5 shows the antenna element positions and the average of the simulated S-parameters between the feeding ports of the antenna element at the center of the array and those of each antenna

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element:

$$\bar{S}_m = \begin{cases} \frac{1}{2} \left(|S_{21}^{(1)}| + |S_{12}^{(1)}| \right) & \text{for } m = 1 \\ \frac{1}{4} \sum_{k=1}^2 \sum_{l=1}^2 |S_{kl}^{(m)}| & \text{for } m \geq 2 \end{cases} \quad (4.23)$$

where $S_{kl}^{(m)}$ is the simulated transmission coefficient between the feeding probe l of the antenna element at the center of the array (the first antenna element) and the feeding probe k of the m th antenna element at f_0 . Antenna elements except for 18 neighboring antenna elements have small mutual couplings of less than -32.5 dB, and the reradiation from them seems to be negligible. Therefore, in the following calculations, $M_e = 19$ and the reradiations from only 18 neighboring elements shown in Fig. 4.5 are considered to reduce the computation time.

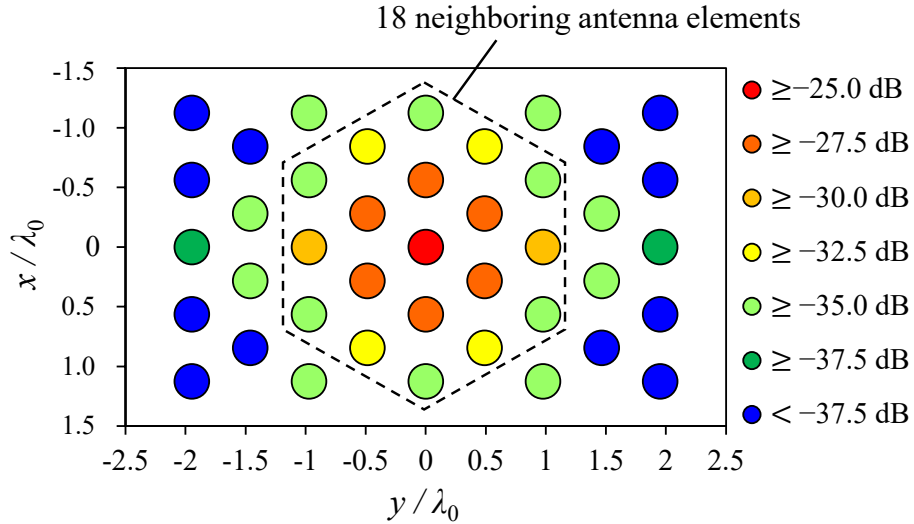


Figure 4.5: Average of the simulated S-parameters between the antenna element at the center and the others.

The maximum mode number N is decided based on the degree of contribution of each mode to the AEPs in the θ and ϕ directions. This is because the AEPs in both directions depend on the mode number n , as expressed in (4.8) and (4.9). The contribution of each mode to the AEP in the ϕ direction is first evaluated by the averaged discrete Fourier transform (DFT) of the simulated AEP in the ϕ direction, which is calculated by

$$D_\theta(n) = \frac{1}{P'} \sum_{p'=1}^{P'} \left| \sum_{q'=0}^{Q'-1} E_\theta^{(t)}(\theta_{p'}, \phi_{q'}) e^{-j \frac{2\pi}{Q'} q' n} \right| \quad (4.24)$$

where $E_\theta^{(t)}(\theta_{p'}, \phi_{q'})$ is the θ component of the simulated AEP for the RHCP excitation. Figure 4.6(a) shows the amplitudes of the averaged DFT, with $\theta_{p'} = 0^\circ$ to 90° in 10°

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increments ($P' = 10$) and $\phi_{q'} = -180^\circ$ to 180° in 1° increments ($Q' = 361$). The first mode dominates over the other modes and especially, $D_\theta(n)$ for $n \geq 8$ are less than -30 dB of that of the first mode. Therefore, $N = 7$ and $Q = 18$ ($\phi_q = 0^\circ$ to 340° in 20° increments) are tentatively selected.

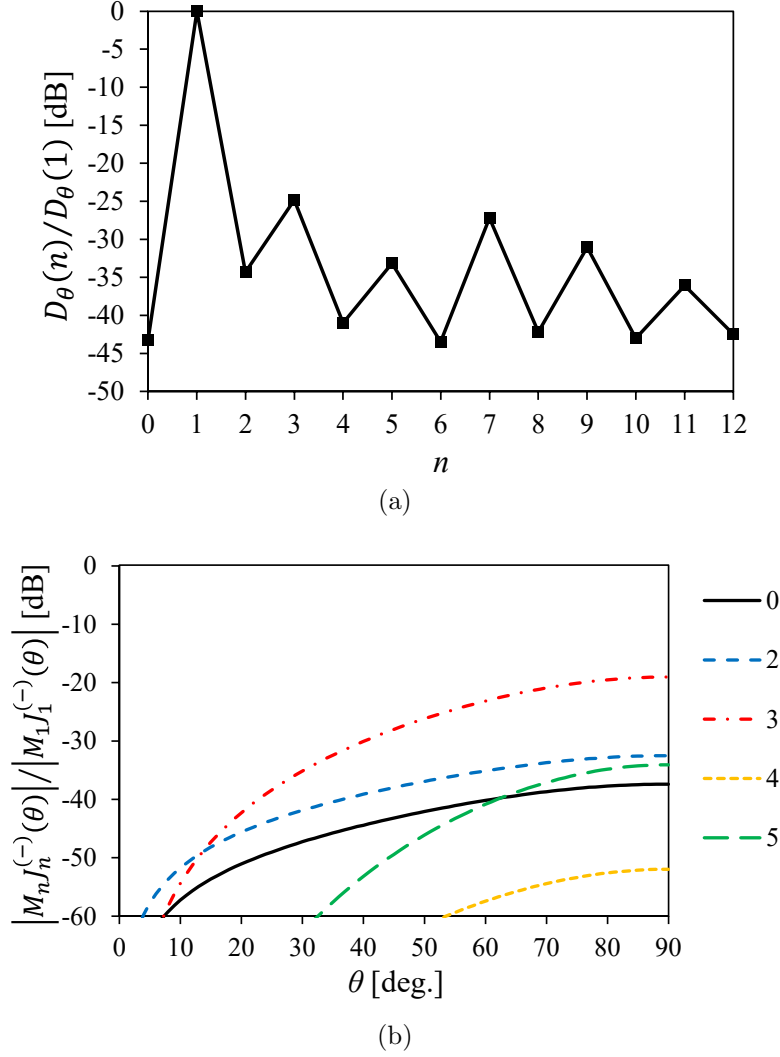


Figure 4.6: Contribution of each mode to the radiation pattern. (a) Amplitude of averaged DFT of the simulated AEPs. (b) Calculated mode coefficients M_n .

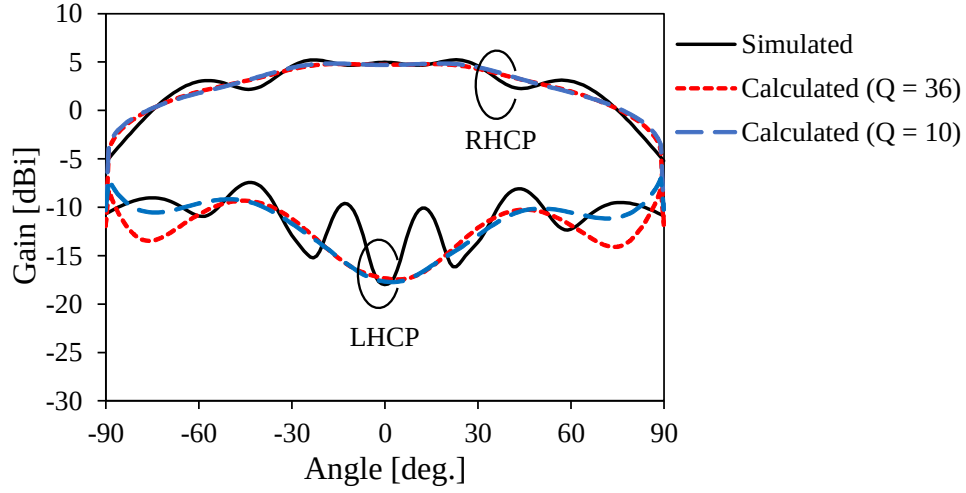
Subsequently, the contribution of each mode to the AEP in the θ direction is evaluated by the calculated $|M_n J_n^{(-)}(\theta)|$, which represents the radiation pattern of each mode in the θ direction. Figure 4.6(b) shows the calculated $|M_n J_n^{(-)}(\theta)|$, which are normalized by that for the first mode. Because the values for $n \geq 6$ are less than -60 dB, they are omitted in Fig. 4.6(b). The first mode dominates again over the other modes; thus, only the first mode seems to be sufficient to roughly approximate the simulated AEP. However, because the radiation pattern for the first mode, $J_1^{(\pm)}(\theta)$, is an even function and cannot describe

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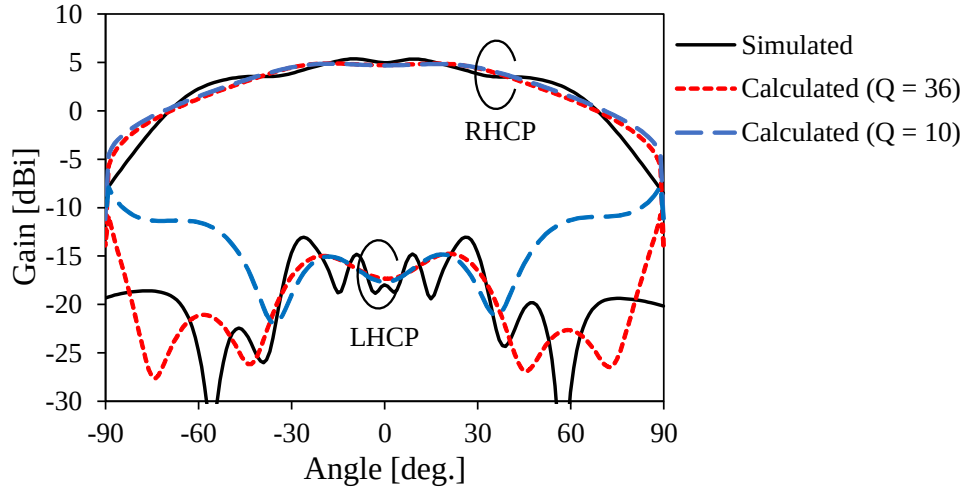
the asymmetric radiation patterns, the higher order modes, particularly the even order modes, should be considered for a more accurate approximation. The second mode has the largest contribution among the even order modes, and the third mode has a larger contribution than the second one. Consequently, $N = 3$ is selected, and the modes of $n = 0$ and $n \geq 4$ are ignored in the following calculations.

Next, P and Q used in the minimization of (4.12) are considered. When the simulated AEPs are used in (4.12), the AEPs can easily be collected in many sample directions, allowing for large P and Q . However, when using the measured AEPs, measurement of the AEPs at many sample points in both θ - and ϕ - directions becomes time-consuming task. In the experiments presented in Section 4.4, AEPs are measured while rotating the antenna in the θ -direction with a rotary positioner. Therefore, as in the experiment, the AEPs at many sample points in the θ -direction are used, and $P = 91$ ($\theta_p = 0^\circ$ to 90° in 1° increments). On the other hand, Q should be small to reduce the measurement effort but affects the accuracy of the fitting function, resulting in AR compensation performance. Figure 4.7(a) and (b) show the simulated AEP and the calculated fitting functions with $Q = 36$ and 10 at $\phi = 0^\circ$ and 90° , respectively, where $\phi_q = 0^\circ$ to 350° in 10° increments for $Q = 36$ and $\phi_q = 0^\circ$ to 324° in 36° increments for $Q = 10$. The fitting functions and simulated AEPs are in good agreement for the RHCP regardless of Q , whereas there is a discrepancy between them for the LHCP. One of the causes for the discrepancy is the ripple on the simulated AEPs due to the diffraction at the array edges. In particular, for $Q = 10$, the discrepancy is large at wide-angle directions at $\phi = 90^\circ$. This is because (4.12) is minimized without the simulated AEP at $\phi = 90^\circ$ when $Q = 10$. This large discrepancy will cause the performance degradation of AR compensation, and thus, $Q = 36$ is selected in the following calculations.

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(a)



(b)

Figure 4.7: Calculated fitting functions and simulated AEPs for the RHCP excitation at (a) $\phi = 0^\circ$ and (b) $\phi = 90^\circ$.

4.3.2 Performance of AR Compensation

Figures 4.8 (a) shows the simulated radiation patterns before and after the AR compensation when the beam is scanned to $(\theta_s, \phi_s) = (70^\circ, 90^\circ)$ for the RHCP excitation. A comparison of the AR patterns is shown in Fig. 4.8 (b). The calculated compensation weights are $C_1 = 1$ and $C_2 = 0.91 - 0.05j$. In the simulation, the compensation weights and the phase shifts for beam scanning are set to each feeding port without digitization. The AR compensation hardly changes the RHCP pattern but successfully reduces the LHCP field around the scan direction. The AR at the scan direction is 0.8 dB before the AR compensation and is improved to 0.1 dB by the AR compensation.

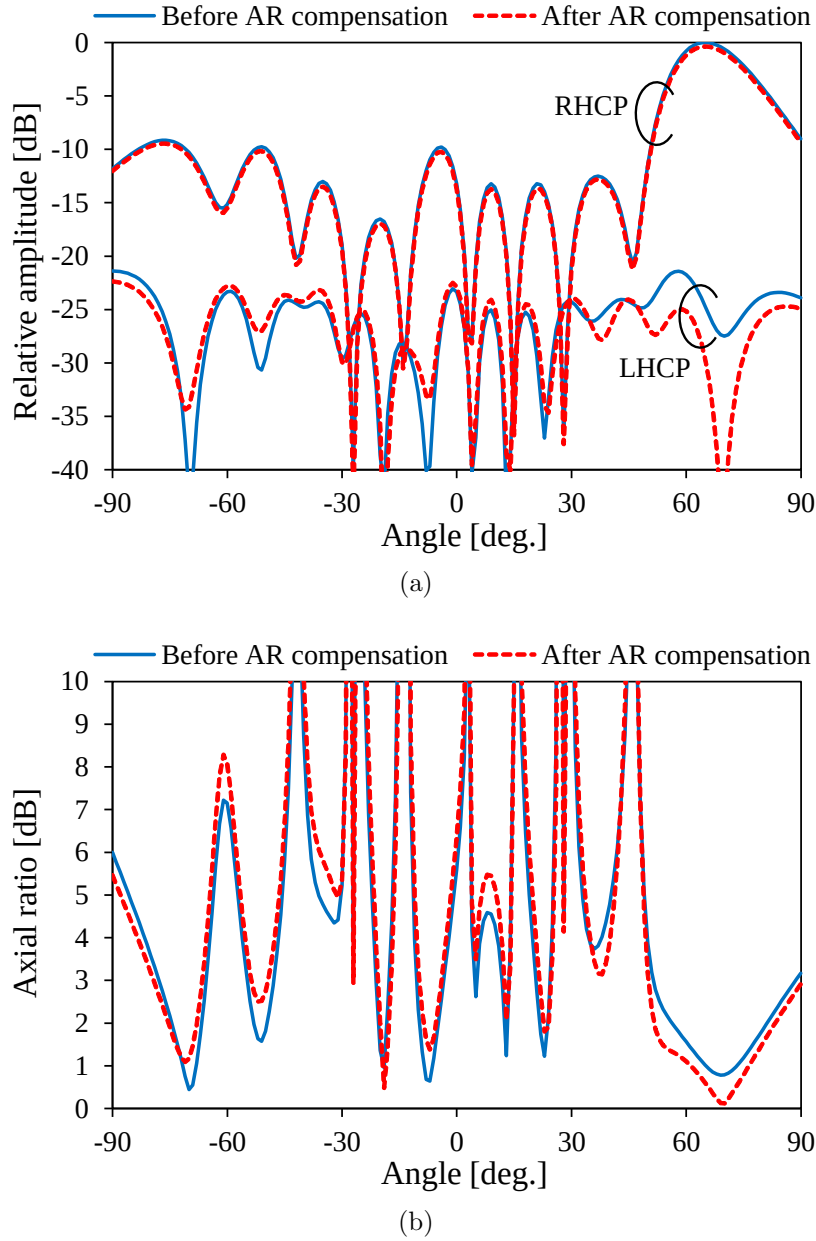


Figure 4.8: Comparison of the simulated array patterns before and after the AR compensation at $(\theta_s, \phi_s) = (70^\circ, 90^\circ)$ scan: (a) radiation patterns and (b) AR patterns at $\phi = 90^\circ$.

Figure 4.9(a) and (b) show the AR variations with the scan angle before and after the AR compensation, respectively. Figure 4.9(c) and (d) show the AR improvement and gain reduction at each scan angle by the AR compensation, respectively. In Fig. 4.9(c), positive values (warm colors) and negative values (cold colors) indicate that the AR is improved and degraded by the AR compensation, respectively. The ARs are degraded around $(\theta_s, \phi_s) = (\pm 65^\circ, 0^\circ)$ and at wide-angle directions before the AR compensation, whereas low ARs are observed within the designated coverage range $(-70^\circ \leq \theta_s \leq 70^\circ)$.

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after the AR compensation. The ARs are successfully improved in most scan angles, and significant improvements are observed especially at wide-angle scanning. The worst AR in the designated coverage range is 5.8 dB before the AR compensation and 2.0 dB after the AR compensation. Relatively large gain reduction is observed in the region where the AR is high before the AR compensation, and the largest gain reduction due to the AR compensation is approximately 1.7 dB within the coverage range. It is unavoidable that the proposed technique involves a reduction in gain. Therefore, in a system such as MPT, where power transmission efficiency is particularly important, it is necessary to evaluate both the impact of this gain reduction and the improvement in polarization mismatch loss that the proposed technique offers.

Figure 4.10 shows the calculated improvement in power transmission efficiency at each scan angle after the AR compensation. In this calculation, an opposite antenna is assumed to have an AR of 3 dB. The positive values (warm colors) and negative values (cold colors) indicate that the power transmission efficiency is improved and diminished, respectively, by the AR compensation. Although the AR compensation improves the efficiency at certain scan angles, it diminishes the efficiency at most scan angles. The maximum improvement and degradation are approximately 0.4 dB and 1.2 dB, respectively, within the designated coverage range. In order to prevent the deterioration in power transmission efficiency, it is necessary to ascertain in advance whether the proposed technique can improve the efficiency. This can be achieved by using the obtained fitting function, which is one of the future challenges for implementing the proposed technique into the systems where power transmission efficiency is particularly important.

Although the results for the RHCP excitation have been presented, the results for the LHCP excitation are almost the same, and thus the detailed data for them are excluded due to space limitations.

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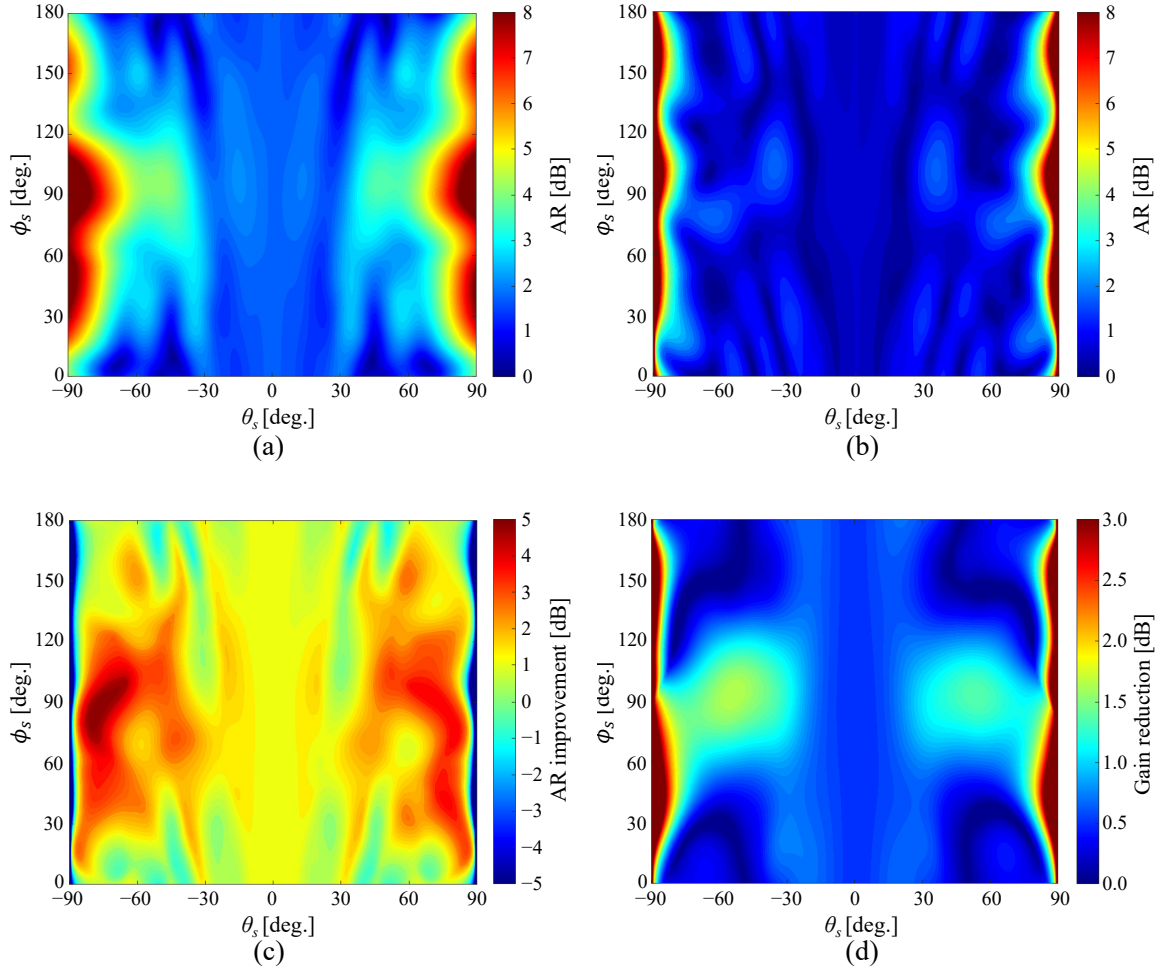


Figure 4.9: Comparison of AR and gain before and after the AR compensation for the RHCP excitation. AR variations with the scan angle: (a) before and (b) after the AR compensation. (c) AR improvement and (d) RHCP gain reduction at each scan angle by the AR compensation.

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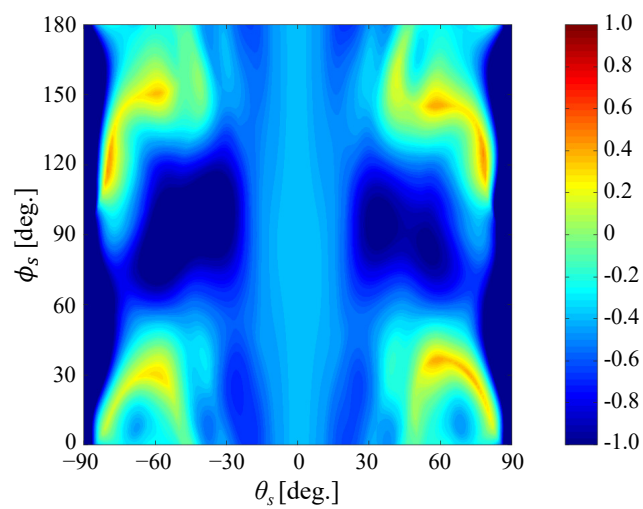


Figure 4.10: Improvement in power transmission efficiency at each scan angle after the AR compensation.

4.4 Experiment

The proposed technique is experimentally tested using a 1024-element transmitting phased array of the dual-polarized circular patch antennas. The antenna elements of the phased array antenna under test (AUT) have the same configuration as those in the simulation except for the feeding locations. The lower patches of the AUT are fed at $\phi_1 = 225^\circ$ and $\phi_2 = 315^\circ$, and antenna elements are arranged in a 32×32 equilateral triangular lattice shown in Fig. 4.11. The AUT uses commercial available 8-channel beamforming RFICs, and they are integrated in the back side of the printed circuit board (PCB) for the antenna. The RFICs have a 5-bit digital phase shifter covering 360° phase shift and a VGA with a gain control accuracy of less than 0.5 dB at each channel. The power dividing network is integrated in the PCB. Before the AR compensation, all VGAs are controlled to have their maximum gain, and each phase shifter is controlled such that the phase difference of radiated fields from two feeding ports of each antenna element becomes 90° in the broadside direction. In this experiment, the radiation patterns are measured in the θ - ϕ coordinate system using a rotary positioner and a polarization positioner (for rotation around the z -axis).

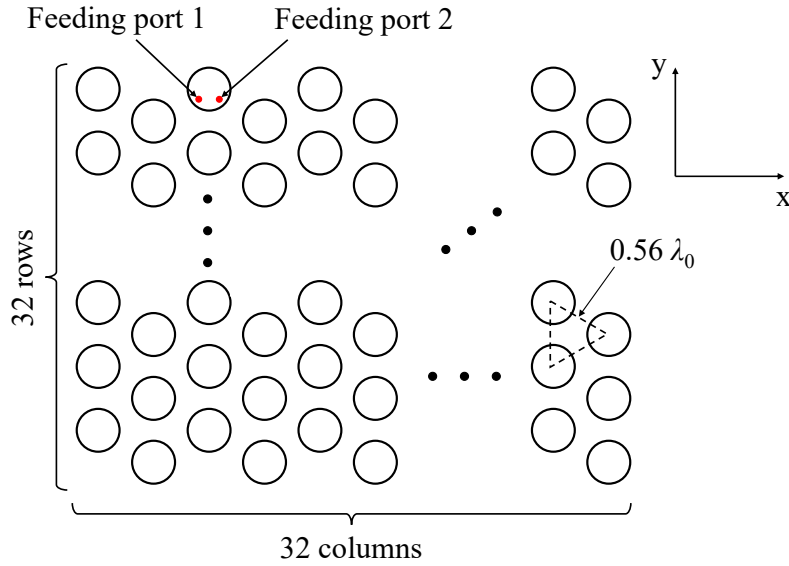


Figure 4.11: Antenna element arrangement of the AUT.

The AEPs are usually measured by feeding one antenna element, with all others terminated in matched loads. However, the RFICs of the AUT do not have the ability to switch their VGAs to terminators, and thus it is difficult to directly measure the AEPs of the AUT. In this experiment, therefore, the AEP is estimated by a simplified version of the method proposed in [4.39]. Specifically, all antenna elements in the AUT are assumed to have the same AEP, and it is estimated by solving (7) in [4.39] with the multiple mea-

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sured scanned patterns. In this method, for accurate estimation, the measured scanned patterns whose main beams fill the angular range where the AEP is to be estimated should be used. Therefore, if all antenna elements in the AUT are used, many scanned patterns are required due to a narrow beamwidth of the AUT. In this experiment, to estimate the AEP with a reduced number of scanned patterns, the scanned patterns were measured using only 6×6 antenna elements, which were located around the center of the array aperture. Figure 4.12 shows an estimated AEP for the RHCP excitation at $\phi = 0^\circ$. Seven scanned patterns with $\theta_s = 0^\circ, \pm 20^\circ, \pm 40^\circ, \pm 60^\circ$ in each cut-plane at $\phi = 0^\circ$ to 165° in 15° increments were used in the estimation. The estimated AEP for the RHCP has a hollow around the broadside direction and has large cross polarizations around $\pm 60^\circ$ – 70° .

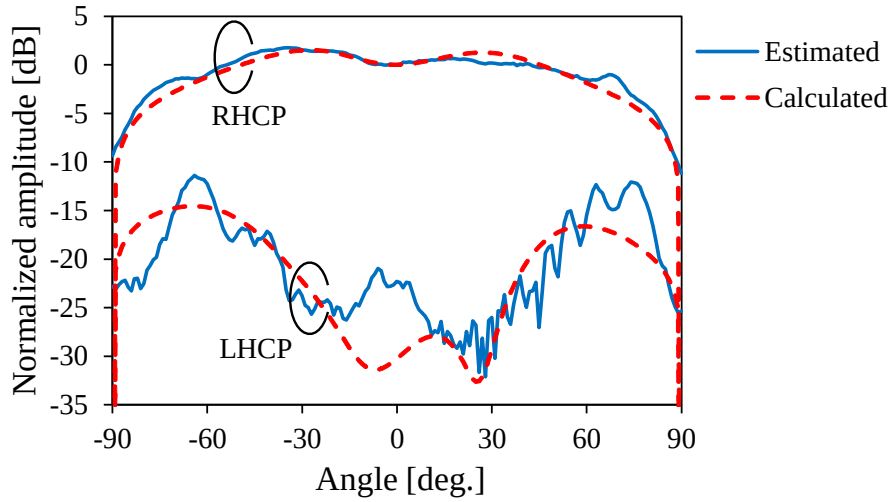


Figure 4.12: Estimated AEP and calculated fitting function of the AUT at $\phi = 0^\circ$.

The unknown parameters were then determined using the estimated AEP, and a design formula for compensation weights was constructed. In the optimization process, $M_e = 19$, $N = 3$, $P = 91$ ($\theta_p = 0^\circ$ to 90° in 1° increments), and $Q = 24$ ($\phi_q = 0^\circ$ to 345° in 15° increments). The fitting function, calculated using the determined parameters, is shown in Fig. 4.12. The calculated fitting function has a hollow around the broadside direction for the RHCP and peaks around $\pm 60^\circ$ for the LHCP, and thus generally approximates the estimated AEP.

Figures 4.13(a) and (b) show a comparison of the measured radiation patterns and AR patterns before and after the AR compensation when the beam is scanned to $(\theta_s, \phi_s) = (70^\circ, 0^\circ)$ for the RHCP excitation, respectively. Each radiation pattern is normalized by the peak gain for the RHCP before the AR compensation. The compensation weights are $C_1 = 0.57 + 0.05j$ and $C_2 = 1$. The LHCP field after the AR compensation is successfully reduced around the scan direction. The AR at the scan direction is 3.9 dB before the AR compensation and is improved to 0.9 dB by the AR compensation. The gain reduction

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by the AR compensation is 2.0 dB, which generally agrees with the expected value from the compensation weights.

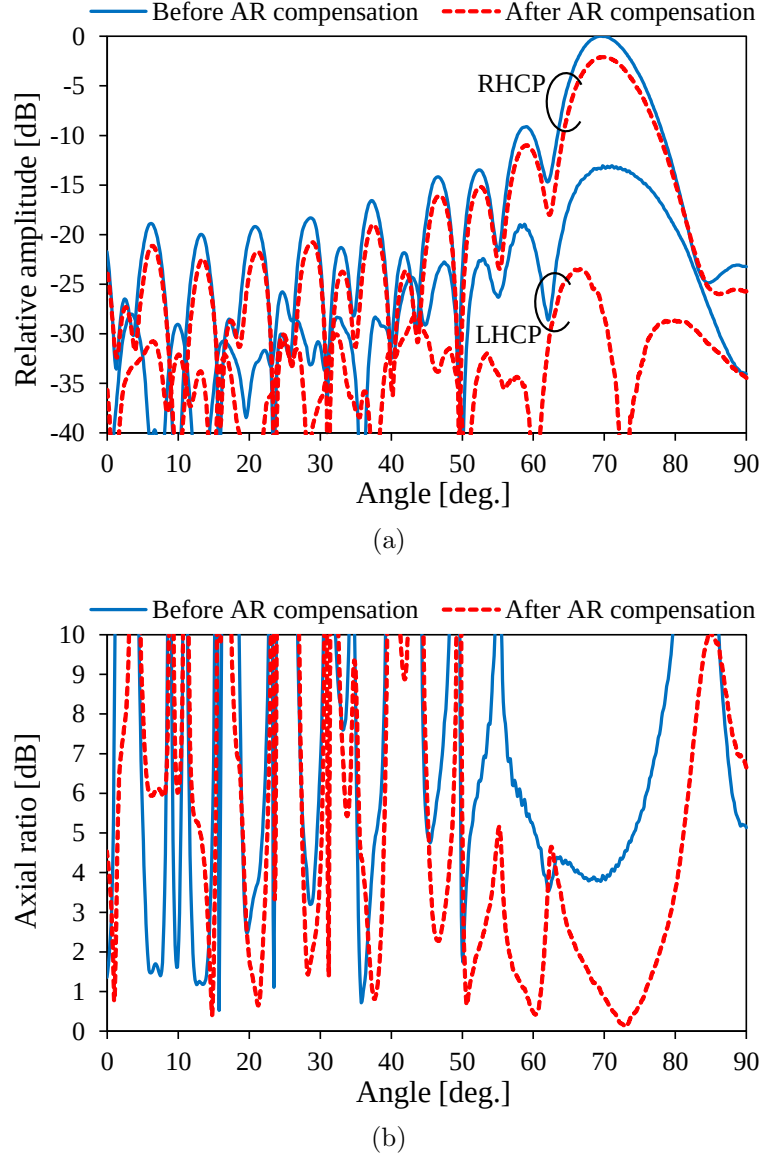


Figure 4.13: Comparison of the measured array patterns before and after the AR compensation at $(\theta_s, \phi_s) = (70^\circ, 0^\circ)$ scan: (a) radiation patterns and (b) AR patterns.

Figures 4.14 to 4.17 show the measured radiation patterns for the RHCP and AR patterns with various scan angles, $\theta_s = 0^\circ$ to 70° in 10° increments at $\phi = 0^\circ, 30^\circ, 60^\circ$, and 90° , respectively. The amplitudes of the measured radiation patterns are normalized by the peak gain before the AR compensation at $\theta_s = 0^\circ$ scan. For simplicity, the results only in the range of $\theta_s - 5^\circ \leq \theta \leq \theta_s + 5^\circ$ for each pattern are shown. The ARs are improved by the AR compensation for all scan angles except for $\theta_s = 0^\circ$. The worst AR before the AR compensation is 5.8 dB at $(\theta_s, \phi_s) = (70^\circ, 90^\circ)$ scan, whereas the one

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after the AR compensation is 1.9 dB at $(\theta_s, \phi_s) = (70^\circ, 30^\circ)$ scan. The maximum gain reduction by the AR compensation is 2.1 dB at $(\theta_s, \phi_s) = (60^\circ, 90^\circ)$ scan.

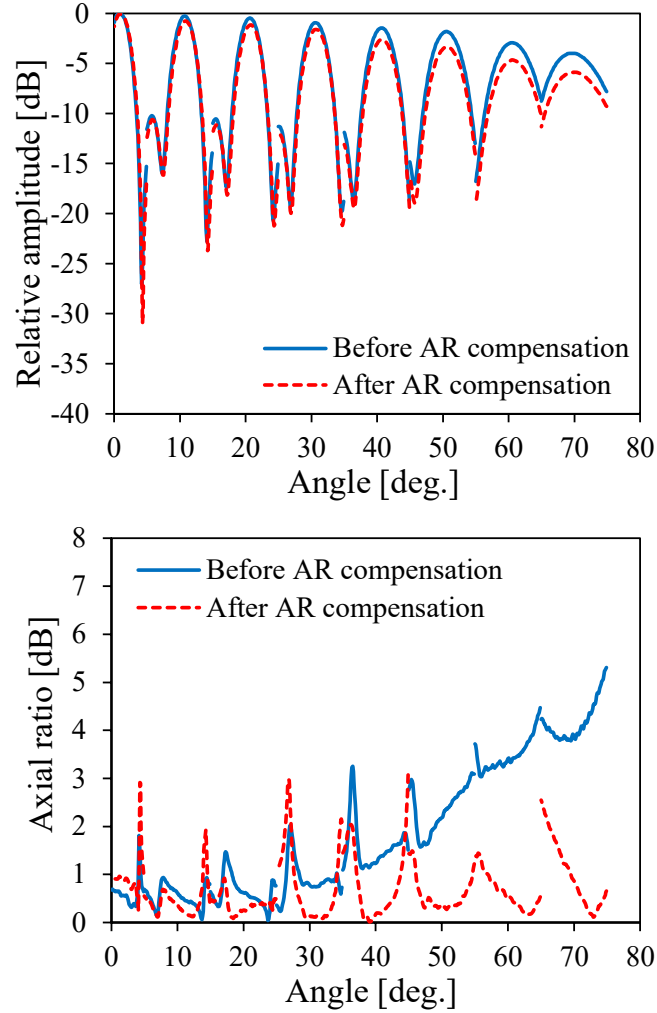


Figure 4.14: Comparison of the measured array patterns before and after the AR compensation with various scan angles: $\theta_s = 0^\circ$ to 70° in 10° increments at $\phi = 0^\circ$. The radiation patterns for the RHCP and AR patterns are shown in the upper and lower rows, respectively.

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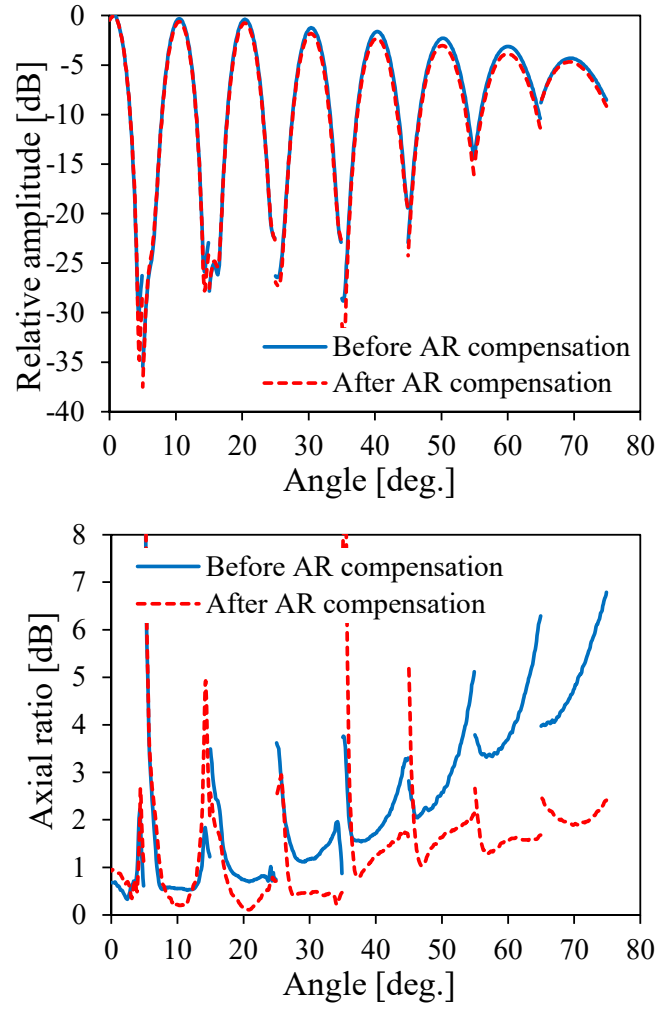


Figure 4.15: Comparison of the measured array patterns before and after the AR compensation with various scan angles: $\theta_s = 0^\circ$ to 70° in 10° increments at $\phi = 30^\circ$. The radiation patterns for the RHCP and AR patterns are shown in the upper and lower rows, respectively.

4. Axial Ratio Improvement for Wide-angle Scanning Circularly-Polarized Phased Array of Circular Patch Antennas

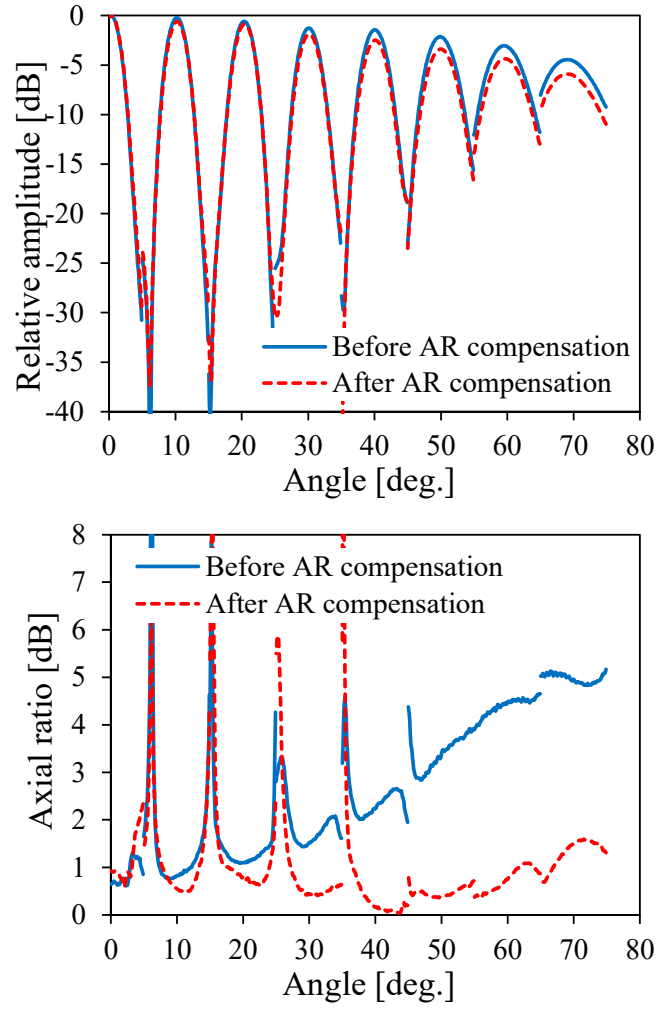


Figure 4.16: Comparison of the measured array patterns before and after the AR compensation with various scan angles: $\theta_s = 0^\circ$ to 70° in 10° increments at $\phi = 60^\circ$. The radiation patterns for the RHCP and AR patterns are shown in the upper and lower rows, respectively.

4. Axial Ratio Improvement for Wide-angle Scanning Circularly-Polarized Phased Array of Circular Patch Antennas

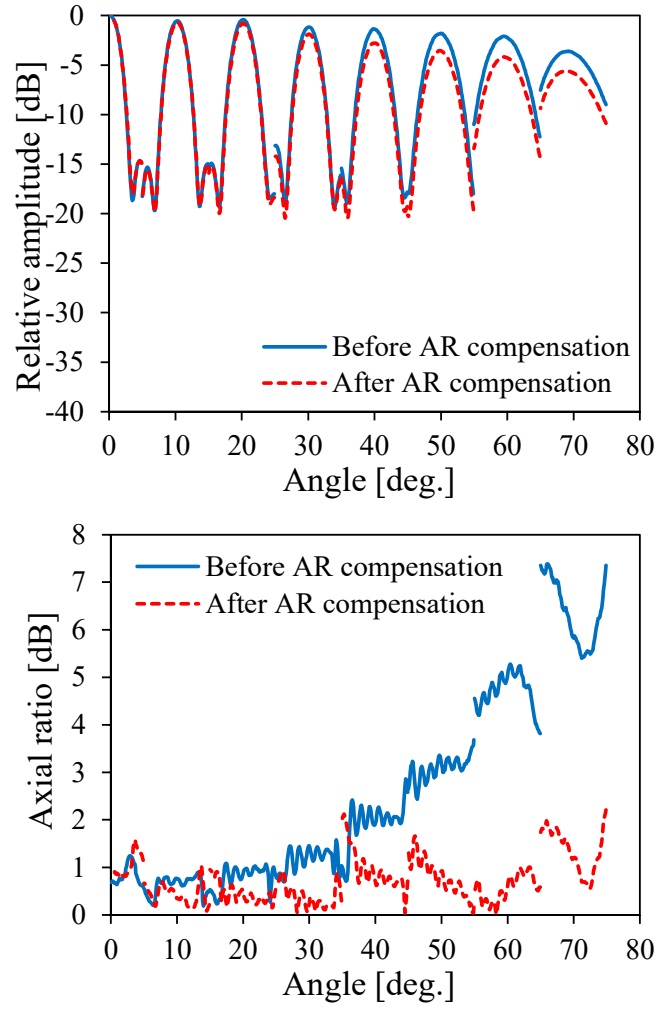


Figure 4.17: Comparison of the measured array patterns before and after the AR compensation with various scan angles: $\theta_s = 0^\circ$ to 70° in 10° increments at $\phi = 90^\circ$. The radiation patterns for the RHCP and AR patterns are shown in the upper and lower rows, respectively.

4.5 Conclusion

In this chapter, a novel calibration technique was presented to improve AR for wide-angle scanning phased arrays of reconfigurable dual-polarized circular patch antennas. The proposed technique uses a closed-form design formula for compensation weights, which is obtained by a fitting function of an AEP of the circular patch antenna array. The proposed fitting function is constructed with cylindrical Bessel functions, which will offer an efficient approximation of the AEP of the circular patch antenna. Although the unknown parameters for the fitting function are determined by an optimization process, once they are determined, the compensation weights for a given scan angle can be easily obtained using the design formula. In addition, the proposed technique eliminates the necessity for the individual measurements of each feeding port of the dual-polarized antennas at each scan angle, thereby reducing the effort required for its implementation. The calculation conditions for the fitting function and the effectiveness of the proposed technique were evaluated by the simulation of a 41-element array, and the worst AR in the designated coverage range of $\pm 70^\circ$ was improved from 5.8 dB to 2.0 dB by the proposed technique. In addition, the proposed technique was experimentally tested using a 1024-element phased array. The measured worst AR was successfully improved from 5.8 dB to 1.9 dB at 70° scan by the proposed technique.

Future works of the proposed technique include the evaluation and enhancement of frequency performance of the proposed technique and further AR improvement. In addition, it is unavoidable that the proposed technique involves a reduction in gain. When applying the proposed technique to a system where transmission efficiency is of particular importance, it is necessary to ascertain in advance whether the transmission efficiency can be improved using the obtained fitting function. This is also one of the future challenges of the proposed technique. Consequently, the proposed technique is highly effective for PAAs in systems where the AR is of particular significance.

Reference

- [4.1] J. J. H. Wang, “Antennas for global navigation satellite system (GNSS),” *Proc. IEEE*, vol. 100, no. 7, pp. 2349–2355, July 2012.
- [4.2] M. Ando, K. Sakurai, N. Goto, K. Arimura, and Y. Ito, “A radial line slot antenna for 12 GHz satellite TV reception,” *IEEE Trans. Antennas Propag.*, vol. 33, no. 12, pp. 1347–1353, Dec. 1985.
- [4.3] P. Petropoulou, E. T. Michailidis, A. D. Panagopoulos, and A. G. Kanatas, “Radio propagation channel measurements for multi-antenna satellite communication systems: a survey,” *IEEE Antennas Propag. Mag.*, vol. 56, no. 6, pp. 102–122, Dec. 2014.
- [4.4] Z. H. Jiang, Z. Cui, T. Yue, Y. Zhu, and D. H. Werner, “Compact, highly efficient, and fully flexible circularly polarized antenna enabled by silver nanowires for wireless body-area networks,” *IEEE Trans. Biomed. Circuits Syst.*, vol. 11, no. 4, pp. 920–932, Aug. 2017.
- [4.5] J. -H. Lee, J. M. Hwang, D. H. Choi, and S. -O. Park, “Noninvasive biosignal detection radar system using circular polarization,” *IEEE Trans. Inf. Technol. Biomed.*, vol. 13, no. 3, pp. 400–404, May 2009.
- [4.6] N. Shinohara, “Beam control technologies with a high-efficiency phased array for microwave power transmission in Japan,” *Proc. IEEE*, vol. 101, no. 6, pp. 1448–1463, June 2013.
- [4.7] X. Luo et al., “A scalable Ka-band 1024-element transmit dual-circularly-polarized planar phased array for SATCOM application,” *IEEE Access*, vol. 8, pp. 156084–156095, 2020.
- [4.8] K. -X. Li, Z. -J. Guo, and Z. -C. Hao, “A multipolarized planar phased array for LEO SATCOM applications,” *IEEE Antennas Wireless Propag. Lett.*, vol. 21, no. 11, pp. 2273–2277, Nov. 2022.
- [4.9] R. S. Hao, J. F. Zhang, S. C. Jin, D. G. Liu, T. J. Li, and Y. J. Cheng, “K-/Ka-band shared-aperture phased array with wide bandwidth and wide beam coverage for LEO satellite communication,” *IEEE Trans. Antennas Propag.*, vol. 71, no. 1, pp. 672–680, Jan. 2023.
- [4.10] T. Takahashi et al., “Phased array system for high efficiency and high accuracy microwave power transmission,” 2016 *IEEE Int. Symp. Phased Array Syst. Technol.*, pp. 1–7, 2016.
- [4.11] J. H. Park, N. M. Tran, S. I. Hwang, D. I. Kim, and K. W. Choi, “Design and implementation of 5.8 GHz RF wireless power transfer system,” *IEEE Access*, vol.

- 9, pp. 168520–168534, 2021.
- [4.12] J. Shin et al., “5.8 GHz 4-channel beamforming Tx IC for microwave power transfer,” *IEEE Access*, vol. 9, pp. 72316–72325, 2021.
 - [4.13] European Telecommunications Standards Institute, “Satellite earth stations and systems (SES); transmit/receive very small aperture terminals (VSATs) used for data communications operating in the fixed satellite service (FSS) 6 GHz and 4 GHz frequency bands,” *European Telecommunication Standard*, ETS 300 332, July 1997.
 - [4.14] Defense Information Systems Agency, “Interoperability of superhigh frequency (SHF) satellite communications terminals,” *Department of Defense Interface Standard*, MIL-STD-188-164C, Nov. 2018.
 - [4.15] Y. -Q. Wen, B. -Z. Wang, and X. Ding, “Wide-beam circularly polarized microstrip magnetic-electric dipole antenna for wide-angle scanning phased array,” *IEEE Antennas Wireless Propag. Lett.*, vol. 16, pp. 428–431, 2017.
 - [4.16] C. H. Lee and J. H. Lee, “Wide angle scanning circular polarized meta-structured antenna array,” *IEICE Trans. Commun.*, vol. E101-B, no. 9, pp.2017–2023, Sept. 2018.
 - [4.17] L. -K. Zhang, Y. -X. Wang, J. -Y. Li, Y. Feng, and W. Zhang, “Cavity-backed circularly polarized cross-dipole phased arrays,” *IEEE Antennas Wireless Propag. Lett.*, vol. 20, no. 9, pp. 1656–1660, Sept. 2021.
 - [4.18] L. Baggen, S. Holzwarth, W. Simon, and O. Litschke, “Phased array using the sequential rotation principle: analysis of coupling effects,” *IEEE Int. Symp. Phased Array Syst. Technol.*, pp. 571–576, 2003.
 - [4.19] J. -C. S. Chieh et al., “Development of flat panel active phased array antennas using 5G silicon RFICs at Ku- and Ka-bands,” *IEEE Access*, vol. 8, pp. 192669–192681, 2020.
 - [4.20] X. Chang, H. B. Wang, T. J. Li, S. C. Jin, D. G. Liu, and Y. J. Cheng, “Shared-aperture phased array antenna with codesigned near-field coupled circular polarizer loaded for K/Ka-band wide-angle satellite communication,” *IEEE Trans. Antennas Propag.*, vol. 70, no. 9, pp. 7478–7490, Sept. 2022.
 - [4.21] Y. Yang, Y. -L. Ban, Q. Yang, J. -W. Lian, Q. Sun, and G. Wu, “Millimeter wave wide-angle scanning circularly polarized antenna array with a novel polarizer,” *IEEE Trans. Antennas Propag.*, vol. 70, no. 2, pp. 1077–1086, Feb. 2022.
 - [4.22] I. Chiba, M. Ohtsuka, S. Matsumoto, and T. Katagi, “Phase controlling for improving the gain or providing the low cross-polarization level in circularly polarized phased array antennas,” *IEICE Trans. Commun.*, vol. J75-B-II, no.7 pp.456–463, Sept. 2018 (in Japanese).

- [4.23] I. Chiba, M. Ohtsuka, S. Matsumoto, and T. Katagi, "Phase controlling for improving the gain or providing the low cross-polarization level in circularly polarized phased array antennas," *Electron. Commun. Jap. Pt. I*, vol. 76, no. 6, pp.92–100, 1993.
- [4.24] A. H. Aljuhani, T. Kanar, S. Zahir, and G. M. Rebeiz, "A 256-element Ku-band polarization agile SATCOM receive phased array with wide-angle scanning and high polarization purity," *IEEE Trans. Microw. Theory Tech.*, vol. 69, no. 5, pp. 2609–2628, May 2021.
- [4.25] J. -W. Kim, S. -C. Chae, H. -W. Jo, T. -D. Yeo, and J. -W. Yu, "Wideband circularly polarized phased array antenna system for wide axial ratio scanning," *IEEE Trans. Antennas Propag.*, vol. 70, no. 2, pp. 1523–1528, Feb. 2022.
- [4.26] T. Jaschke and A. F. Jacob, "A dual-polarized SIW lens antenna array for Rx-/Tx-integration at K/Ka-band," *IEEE Trans. Antennas Propag.*, vol. 71, no. 3, pp. 2443–2453, March 2023.
- [4.27] T. Chaloun et al., "Electronically steerable antennas for future heterogeneous communication networks: review and perspectives," *IEEE J. Microw.*, vol. 2, no. 4, pp. 545–581, Oct. 2022.
- [4.28] E. S. Moghaddam, "Design of a compact multilayer circularly polarized phased array transmit antenna system for satellite applications", *Int. J. Electron. Commun.*, vol. 70, no. 9, pp. 1142–1155, Sep. 2016.
- [4.29] K. Y. Kapusuz, S. Lemey, and H. Rogier, "Dual-polarized 28-GHz air-filled SIW phased antenna array for next-generation cellular systems," *2019 IEEE Int. Symp. Phased Array Syst. Technol.*, pp. 1–6, 2019.
- [4.30] S. Ikeda, K. Yokokawa, N. Nakamoto, T. Fukasawa, M. Ohtsuka, and Y. Inasawa, "A circularly polarized cavity-backed stacked patch antenna for wide-angle beam scanning millimeter-wave phased array," *2020 Int. Sympo. Antennas Propag. (ISAP)*, pp. 721–722, 2021.
- [4.31] R. Maaskant, M. V. Ivashina, S. J. Wijnholds and K. F. Warnick, "Efficient Prediction of Array Element Patterns Using Physics-Based Expansions and a Single Far-Field Measurement," *IEEE Trans. Antennas Propag.*, vol. 60, no. 8, pp. 3614–3621, Aug. 2012.
- [4.32] A. T. Muriel-Barrado, J. Calatayud-Maeso, A. Rodríguez-Gallego, P. Sánchez-Olivares, J. M. Fernández-González, and M. Sierra-Pérez, "Evaluation of a planar reconfigurable phased array antenna driven by a multi-channel beamforming module at Ka band," *IEEE Access*, vol. 9, pp. 63752–63766, 2021.

- [4.33] N. C. Karmakar, “Investigations into a cavity-backed circular-patch antenna,” *IEEE Trans. Antennas Propag.*, vol. 50, no. 12, pp. 1706–1715, Dec. 2002.
- [4.34] E. Magill and H. Wheeler, “Wide-angle impedance matching of a planar array antenna by a dielectric sheet,” *IEEE Trans. Antennas Propag.*, vol. 14, no. 1, pp. 49–53, Jan. 1966.
- [4.35] D. H. Werner, “An exact integration procedure for vector potentials of thin circular loop antennas,” *IEEE Trans. Antennas Propag.*, vol. 44, no. 2, pp. 157–165, Feb. 1996.
- [4.36] H. Attia, L. Yousefi, and O. M. Ramahi, “Analytical model for calculating the radiation field of microstrip antennas with artificial magnetic superstrates: theory and experiment,” *IEEE Trans. Antennas Propag.*, vol. 59, no. 5, pp. 1438–1445, May 2011.
- [4.37] MathWorks, “GlobalSearch,” <https://www.mathworks.com/help/gads/globalsearch.html>, July 2023.
- [4.38] M. L. Kales, “Techniques for handling elliptically polarized waves with special reference to antennas: Part III–Elliptically polarized waves and antennas,” *Proc. IRE*, vol. 39, no. 5, pp. 544–549, May 1951.
- [4.39] L. Sung, S. Noh, and D. -h. Cho, “Antenna element pattern estimation based on array beam pattern,” 2018 International Conference on Information and Communication Technology Convergence. (ICTC), pp.789–791, 2018.

Chapter 5

Conclusions

The goal of this dissertation was to realize PAAs with high performance and advanced functionality at a low cost. As an introduction to this dissertation, Chapter 1 first described the trends of the PAAs in various wireless systems and presented that PAAs with lower cost and higher performance, and more advanced functionalities using DBF are required for the further spread of PAAs. In addition, the previous studies on low-cost PAAs, particularly PPAA, low-cost DBF array antennas, and calibration techniques, which are crucial for achieving the desired performance of PAAs, were briefly reviewed, and their challenges were discussed. As a result of this discussion, this dissertation has focused on the following three technologies:

- (1) Low-cost PPAA with flat structure, high efficiency, and individual phase control capability.
- (2) DBF array antenna based on the CDM-based path-sharing architecture with high antenna efficiency.
- (3) Efficient PAA calibration technique for cross-polarization reduction.

Major results obtained in this study are summarized as follows.

In Chapter 2, a novel radial line planar phased array, in which low-profile helical antenna elements are individually rotated by their respective connected micromotors, was presented to achieve a PPAA with a flat structure, high efficiency, and flexible beam forming capability. To the best of the author's knowledge, this is the first RLPA that has antenna elements electromechanically rotated by their individual micromotors. A novel design methodology for the RLPA was also described, which combines the equivalent circuit theory and electromagnetic simulations of the unit cell element. The proposed procedure is practical to design an RLPA of antenna elements with arbitrary probe shape without large-scale full-wave analysis of the whole structure of the RLPA. The proposed antenna was designed, fabricated, and tested. The fabricated prototype antenna exhibited flexible and accurate beam-scanning performance, a low reflection coefficient, and a

relatively high antenna efficiency, which demonstrates the effectiveness of the proposed design methodology. Future work in this area includes the development of an array for wide-angle beam-scanning and an array design for further improving the antenna performance. Although the proposed antenna has the disadvantage of a slower beam scanning speed than the conventional APAAs, it can be a lower-cost alternative to the conventional APAAs. Therefore, the proposed antenna is potentially suitable for applications such as satellite communication and MPT that do not require high-speed beam scanning.

In Chapter 3, a novel waveguide slot array with the CDM-based path-sharing architecture was presented to provide DBF capabilities at low hardware cost. In the proposed antenna, the received signals from each slot pair are multiplexed in the CDM fashion by switching the diodes' bias according to the WH codes, and they are recovered through the despreading process in the digital domain. The operating principle and SNR performance of the proposed antenna were also explained by a mathematical model of the antenna. The prototype antenna with eight slot pairs were then designed, fabricated, and tested as a proof of concept. Although, in the experiment, the PIN diodes' bias were manually switched, the transient characteristics of the PIN diodes were ignored, and the perfect synchronization between spreading and despreading was assumed, the experimental results demonstrated the feasibility of the proposed antenna. Future works of the proposed antenna include further tests considering the transient characteristics of the PIN diodes and actual synchronization issues, bandwidth enhancement, radiation efficiency improvement, and miniaturization of the slot pair for mitigating the losses depending on the incident angle of signals. The proposed antenna has the potential to offer DBF capability at a lower cost and with a higher radiation efficiency.

In Chapter 4, a novel efficient calibration technique was presented to improve the AR of a dual-circularly polarized APAA. The proposed technique was designed to be applied to phased arrays of reconfigurable dual-polarized circular patch antennas. In the proposed technique, the compensation weights for AR improvement are controlled based on a design formula obtained by a fitting function. The fitting function for the AEP of the circular patch antenna was constructed using cylindrical Bessel functions, which are well-suited for accurately describing the radiation from the circular patch antenna. The calculation condition for constructing the fitting function and the effectiveness of the proposed technique were evaluated by the simulation. Furthermore, the proposed technique was experimentally tested using a 1024-element APAA, which demonstrated successful AR improvement performance. Although the unknown parameters for the fitting function were determined by an optimization process using the measured AEPs of the CP fields, the proposed method does not require the individual measurement of each feeding channel of the antenna. Consequently, it can reduce the measurement effort compared to the conventional techniques. Future works of the proposed technique include

the evaluation and enhancement of frequency performance of the proposed technique and further AR improvement. In addition, in principle, the gain reduction is unavoidable due to the gain control of the amplifiers. Therefore, the proposed technique will be well-suited to systems where the AR is a particularly important performance metric, such as satellite communications.

As described above, this dissertation has sought to develop PAAs with high performance and advanced functionality at a low cost, and has presented the technologies required to achieve this goal. The adoption of PAAs is expected to continue to grow to meet the demand for more sophisticated wireless systems in the future. The author believes that the results of this dissertation will contribute to ensuring this.

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Publication List

Journal papers

- [1] N. Nakamoto, H. Watanabe, T. Takahashi, Y. Inasawa, and N. Shinohara, “Axial Ratio Improvement for Wide-angle Scanning Circularly-Polarized Phased Array of Circular Patch Antennas,” *IEICE Trans. Commun.*, submitted for publication.
- [2] N. Nakamoto, K. Kihira, T. Fukasawa, Y. Inasawa, and N. Shinohara, “Waveguide slot array with code-division multiplexing function for single RF chain digital beam-forming,” *IEICE Trans. Commun.*, vol. E107-B, no. 8, pp. 541–551, Aug. 2024.
- [3] K. Konno, N. Haga, J. Chakaroathai, Q. Chen, N. Nakamoto, and T. Takahashi, “A Novel Method of Moments for Numerical Analysis of Antennas Over 2-D Infinite Periodic Arrays of Scatterers,” *IEEE Trans. Antennas Propag.*, vol. 72, no. 1, pp. 50–60, Jan. 2024.
- [4] N. Nakamoto, Y. Suzuki, S. Yamaguchi, T. Fukasawa, N. Yoneda, H. Miyashita, and N. Shinohara, “Radial Line Planar Phased Array Using Electromechanically Rotated Helical Antennas,” *IEICE Trans. Commun.*, vol. E106-B, no. 2, pp. 174–183, Feb. 2023.
- [5] N. Nakamoto, T. Takahashi, T. Fukasawa, N. Yoneda, and H. Miyashita, “A Suspended Stripline Fed Dual-Polarized Open-Ended Waveguide Subarray with Metal Posts for Phased Array Antennas,” *IEICE Trans. Commun.*, vol. E104-B, no. 3, pp. 295–303, March 2021.
- [6] N. Nakamoto, T. Takahashi, T. Fukasawa, N. Yoneda, and H. Miyashita, “RCS Synthesis of Array Antenna With Circulators and Phase Shifters and Measurement Method for Deterministic RCS Reduction,” *IEEE Trans. Antennas Propag.*, vol. 69, no. 1, pp. 135–145, Jan. 2021.
- [7] H. Watanabe, S. Yamaguchi, T. Ohba, N. Nakamoto, T. Takahashi, T. Fukasawa, M. Otsuka, and H. Miyashita, “Waveguide Edge-shunt Slot Array Antenna with Z-shaped Slot Partially Overlapping the Waveguide Wall,” *IEICE Trans. Commun.*, vol. J100-B, no. 9, pp. 807–816, Feb. 2017 (in Japanese).
- [8] N. Nakamoto, T. Takahashi, M. Otsuka, and H. Miyashita, “Phase-Only Synthesis for Phased Array Antenna with Element Field Errors at Every Phase Shift,” *IEICE Trans. Commun.*, vol. J98-B, no. 12, pp. 1255–1265, Dec. 2015 (in Japanese).
- [9] N. Nakamoto, H. Ban, T. Oka, S. Kitazawa, K. Kobayashi, N. Kikuchi, H. Hatamoto, S. Shimizu, and M. Hara, “Wireless Harness Inside ICT Equipments,” *ICACT Trans. Advanced Commun. Technol.*, vol. 2, no. 1, pp. 135–143, Jan. 2013.

- [10] N. Nakamoto, T. Oka, S. Kitazawa, H. Ban, and K. Kobayashi, “A Theoretical Analysis of a Circular Microstrip Antenna in a Parallel-Plate Waveguide,” *IEICE Trans. Commun.*, vol. E96-B, no. 1, pp. 173–181, Jan. 2013.
- [11] T. Takahashi, N. Nakamoto, M. Otsuka, T. Aoki, Y. Konishi, I. Chiba, and M. Yajima, “On-Board Calibration Methods for Mechanical Distortions of Satellite Phased Array Antennas,” *IEEE Trans. Antennas Propag.*, vol. 60, no. 3, pp. 1362–1372, March 2012.
- [12] N. Nakamoto, T. Takahashi, M. Otsuka, Y. Konishi, T. Aoki, T. Nishino, and M. Yajima, “A Theoretical Study on Accuracy of Phased Array Antenna Calibration Using the REV Method,” *IEICE Trans. Commun.*, vol. J93-B, no. 9, pp. 1312–1321, Sept. 2010 (in Japanese).
- [13] T. Takahashi, N. Nakamoto, M. Otsuka, T. Sakura, T. Aoki, Y. Konishi, and M. Yajima, “A Theoretical Study on Measurement Accuracy of the REV Method for Phased Array Measurement System Designs,” *IEICE Trans. Commun.*, vol. J92-B, no. 2, pp. 446–457, Feb. 2009 (in Japanese).

International conference proceedings

- [1] T. Uno, T. Uesaka, N. Nakamoto, T. Fukasawa, T. Takahashi, Y. Inasawa, T. Yamamoto, T. Koyanagi, I. Kakimoto, and Y. Konishi, “Design of a Dual-Polarized Low Sidelobe Slotted Waveguide Antenna for C-Band Phased Array Weather Radar,” 2022 IEEE Int. Symp. Phased Array Syst. Technol. (PAST), Oct. 2022.
- [2] N. Nakamoto, J. Goto, Y. Suzuki, T. Fukasawa, T. Takahashi, Y. Inasawa, and H. Miyashita, “Radial Line Helical Phased Array with Antenna Elements Rotated by Motors for Microwave Power Transmissions,” 2022 16th Eur. Conf. Antennas Propag. (EuCAP), March 2022.
- [3] N. Nakamoto, Y. Suzuki, S. Yamaguchi, T. Fukasawa, Y. Inasawa, and H. Miyashita, “Demonstration of Radial Line Helical Phased Array with Antenna Elements Rotated by Motors,” 2021 Int. Symp. Antennas Propag. (ISAP), Oct. 2021.
- [4] N. Nakamoto, Y. Suzuki, S. Yamaguchi, T. Fukasawa, N. Yoneda, and H. Miyashita, “Design and Experiments of Radial Line Helical Phased Array with Antenna Elements Rotated by Motors,” 2021 IEEE-APS Topical Conf. Antennas Propag. in Wireless Commun. (APWC), Aug. 2021.
- [5] T. Uno, T. Uesaka, N. Nakamoto, T. Fukasawa, Y. Inasawa, T. Yamamoto, T. Koyanagi, I. Kakimoto, and Y. Konishi, “A Low Side-Lobe Slotted Ridge Waveguide Array Manufactured by Resin Injection Molding,” 2020 Int. Symp. Antennas Propag. (ISAP), pp. 701–702, Jan. 2021.

- [6] S. Ikeda, K. Yokokawa, N. Nakamoto, T. Fukasawa, M. Ohtsuka and Y. Inasawa, "A Circularly Polarized Cavity-Backed Stacked Patch Antenna for Wide-Angle Beam Scanning Millimeter-Wave Phased Array," 2020 Int. Symp. Antennas Propag. (ISAP), pp. 721–722, Jan. 2021.
- [7] T. Uesaka, T. Uno, N. Nakamoto, T. Fukasawa, Y. Inasawa, T. Yamamoto, T. Koyanagi, I. Kakimoto, and Y. Konishi, "A Suspended Line to Waveguide Transition for Dual-polarized Slotted Waveguide Array Antenna," 2020 Int. Symp. Antennas Propag. (ISAP), pp. 759–760, Jan. 2021.
- [8] H. Watanabe, T. Maruyama, N. Nakamoto, T. Fukasawa, and N. Yoneda, "2-D Pattern Reconfigurable Array Antenna with Excitation Phase Difference Circuit," 2019 IEEE Int. Symp. Antennas Propag. USNC-URSI Radio Sci. Meeting, pp. 681-682, July 2019.
- [9] T. Uno, N. Nakamoto, T. Fukasawa, T. Yamamoto, I. Kakimoto, N. Yoneda, and Y. Konishi, "A Study of an In-line Slot Array Antenna Fed by a Zigzag Ridge Waveguide," 2018 Int. Symp. Antennas Propag. (ISAP), Oct. 2018.
- [10] K. Yokokawa, K. Suzuki, N. Nakamoto, S. Yamaguchi, Y. Suzuki, T. Fukasawa, M. Otsuka, N. Yoneda, and H. Miyashita, "Study on Rotational Element Electronically Scanned Array," 2018 IEEE Int. Workshop Electromag.:Appl. Student Innovation Competition (iWEM), Aug. 2018.
- [11] K. Yokokawa, N. Nakamoto, S. Yamaguchi, T. Fukasawa, J. Goto, M. Otsuka, K. Kihira, T. Takahashi, T. Mizuno, and H. Miyashita, "Study on reconfigurable corrugated waveguide using moveable shorting plate," 2017 IEEE Conf. Antenna Meas. Appl. (CAMA), pp. 49–50, Dec. 2017.
- [12] N. Nakamoto, T. Takahashi, T. Fukasawa and N. Yoneda, "RCS reduction of array antenna using circulator and phase shifter," 2017 IEEE Conf. Antenna Meas.. Appl. (CAMA), pp. 190–193, Dec. 2017.
- [13] N. Nakamoto, T. Takahashi, A. Ono, M. Nakashima, M. Otsuka, and H. Miyashita, "A Dual Polarized Suspended Stripline Fed Open-Ended Waveguide Antenna Array With Metal Post Structures," 2016 IEEE Int. Symp. Phased Array Syst. Technol., Oct. 2016.
- [14] N. Nakamoto, T. Takahashi, A. Ono, M. Nakashima, M. Otsuka, and H. Miyashita, "A dual polarized suspended stripline fed open-ended waveguide antenna subarray for phased arrays," 2015 Int. Symp. Antennas Propag. (ISAP), Nov. 2015.
- [15] H. Watanabe, T. Ohba, S. Yamaguchi, N. Nakamoto, T. Takahashi, M. Otsuka, and H. Miyashita, "Waveguide edge-shunt slot array antenna with Z-shaped Slot partially

- overlapping the waveguide wall,” 2014 Asia-Pacific Microw. Conf., pp. 723-725, Nov. 2014.
- [16] N. Nakamoto, T. Takahashi, T. Nomura, M. Otsuka, and H. Miyashita, “A method to measure the antenna mode and structural mode for antenna RCS reduction using circulator and phase shifter,” 2014 Int. Symp. Antennas Propag. (ISAP), pp. 21–22, Dec. 2014.
 - [17] N. Nakamoto, T. Takahashi, Y. Konishi and I. Chiba, “Phase optimization for accurate beam forming of phased array with element field errors at every phase shift,” 2013 IEEE Int. Symp. Phased Array Syst. Technol., pp. 693–697, Oct. 2013.
 - [18] N. Nakamoto, H. Ban and K. Kobayashi, “Resonant frequency of a circular microstrip antenna in a parallel-plate waveguide,” 2012 Int. Symp. Antennas Propag. (ISAP), pp. 150–153, Oct. 2012.
 - [19] N. Nakamoto, T. Oka, S. Kitazawa, H. Ban and K. Kobayashi, “A theoretical study on a circular microstrip antenna in a parallel plate waveguide,” 2012 6th Eur. Conf. Antennas Propag. (EUCAP), pp. 2483–2487, March 2012.
 - [20] A. Bekkali, T. Masuo, T. Tominaga, N. Nakamoto, and H. Ban, “Gaussian processes for learning-based indoor localization,” 2011 IEEE Int. Conf. Signal Process. Commun. Comput. (ICSPCC), Sept. 2011.
 - [21] N. Nakamoto, T. Takahashi, M. Otsuka, T. Aoki, Y. Konishi, T. Nishino, I. Chiba, and M. Yajima, “Study on Calibration Accuracy of Phased Array Antenna Calibration Using the REV Method,” 2009 Int. Symp. Antennas Propag. (ISAP), pp. 636–639, Oct. 2009.

Appendix A

Appendix for Chapter 2

A.1 Hardware Cost Estimation

In order to ascertain the extent of the cost reduction that would be achieved by the proposed radial line planar phased array, the hardware costs of the conventional APAA and the proposed antenna are estimated. For this estimation, 168-element C-band antennas with a transmitting power of 100 W are assumed as an example. Table A.1 shows the estimated hardware costs. The component costs are estimated based on the selling prices of one of the industry's leading manufacturers, information obtained through interviews with the manufacturers, and the actual prices for the prototype antenna [A.1]–[A.4]. In the conventional APAA, the costs of amplifiers and phase shifters account for nearly 95% of the total cost, indicating that they are the primary factors of the high hardware cost. In contrast, the proposed antenna employs a single high-power amplifier with a high price, but the total cost is relatively low due to the low prices of micromotors and motor driver ICs. Consequently, the estimated hardware cost of the proposed antenna is approximately 30% of that of a conventional APAA.

Table A.1: Hardware cost comparison between the conventional APAA and the proposed radial line planar phased array.

Conventional APAA

Component	Unit price [USD]	Number	Price [USD]	Note
Antenna element	0.83	168	140	PCB
Amplifier	53	168	8904	CMD169P4, a product of Qorvo Inc. [A.1]
Phase shifter	63	168	10584	TGP2108-SM, a product of Qorvo Inc. [A.2]
Feeding circuit	560	1	560	PCB
Total (A)			20188	

Proposed antenna

Component	Unit price [USD]	Number	Price [USD]	Note
Antenna element	15	168	2520	actual price of the prototype antenna (except for the initial cost)
Motor	10	168	1680	through the interviews with manufacturers
Motor driver	1	168	168	commercially available stepper motor driver IC [A.3]
Feeding circuit	500	1	700	
Amplifier	990	1	990	QPM1017, a product of Qorvo Inc. [A.4]
Total (B)			6058	

Comparison (B/A)

0.30

A.2 Beam Scanning Speed

The beam scanning speed of the proposed radial line planar phased array depends on the speed of the micromotor. The speed of a stepper motor is typically expressed in units of pulses per second (pps). When a stepper motor has a speed of P pps, the time per step is $1/P$. Consequently, the time per rotation is given by S/P when the stepper motor has S steps per rotation. The typical speed of the stepper motor is 1,000 to 2,000 pps. The calculated speeds per pulse and per rotation with $S = 160$ (the stepper motors used in the prototype antenna have 160 steps in the microstepping mode) are shown in Table A.2. The beam switching speed of the proposed antenna is approximately 80 to 160 ms when the motors are rotated by approximately one rotation in order to switch the beam pattern.

As a case study, consider the case of tracking a moving target with the prototype antenna. When the scanned direction of the prototype antenna is changed by 0.5° (the 0.1 dB beamwidth of the proposed antenna is approximately 1°), the required rotation angles for it can be calculated using (2.14), and the maximum value of them is approximately 17° (7.6 steps). Consequently, the beam scanning speed is approximately 5 ms. If the target is a drone with a speed of 10 m/s in flight at a distance of 10 meters, it passes through a distance equal to 0.5° by 8.7 ms. Therefore, the proposed antenna would be capable of successfully tracking it.

Table A.2: Rotational speeds of the typical stepper motor.

P [pps]	time / pulse [ms]	time/rotation [ms]
1000	1	160
1500	0.67	106.7
2000	0.5	80

Reference

- [A.1] Qorvo, Inc, “CMD169P4 5–7 GHz (C Band) Driver Amplifier MMIC,” <https://www.qorvo.com/products/p/CMD169P4>, accessed June 15. 2024.
- [A.2] Qorvo, Inc, “TGP2108-SM 2.5–4GHz 6-Bit Digital Phase Shifter,” <https://www.qorvo.com/products/p/TGP2108-SM>, accessed June 15. 2024.
- [A.3] Texas Instruments Incorporated, “Stepper motor drivers,” <https://www.ti.com/motor-drivers/stepper-driver/products.html#2716max=7%3B12&>, accessed June 15. 2024.
- [A.4] Qorvo, Inc, “QPM1017 5.7–7.0GHz 100 Watt GaN Power Amplifier Module,” <https://www.qorvo.com/products/p/QPM1017>, accessed June 15. 2024.

Appendix B

Appendix for Chapter 3

B.1 Walsh Hadamard Code

The Walsh Hadamard (WH) codes are based on the Hadamard matrices, which are square arrays of $+1$ and -1 and whose rows or columns are completely orthogonal to each other. The Hadamard matrices are generated as follows [B.1]:

$$\mathbf{H}_2 = \begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix}, \quad \mathbf{H}_{2N} = \begin{bmatrix} \mathbf{H}_N & \mathbf{H}_N \\ \mathbf{H}_N & -\mathbf{H}_N \end{bmatrix} \quad (\text{B.1})$$

where the subscripts indicate the orders of the Hadamard matrices. We show the Hadamard matrices of orders 8 and 16 in Fig. B.1 (the signs of ± 1 are only shown). The WH code corresponds to one of the row(column) vectors of the Hadamard matrices. The orthogonality of the WH codes can also be expressed as follows:

$$\frac{1}{N} \mathbf{h}_p \mathbf{h}_q^\top = \begin{cases} 1 & \text{for } p = q \\ 0 & \text{for } p \neq q \end{cases} \quad (\text{B.2})$$

where $\mathbf{h}_p$ and $\mathbf{h}_q$ are the p th and q th row vectors of the Hadamard matrix of order N , respectively.

To obtain (3.21), consider the orthogonality in the presence of errors. We now consider a vector Δ_q obtained by replacing $+1$ of $\mathbf{h}_q$ by $(1 + \delta_q^{(+)})$ and -1 by $-(1 + \delta_q^{(-)})$. Using a property of an orthogonal matrix [B.2], the Euclidean norm of $\mathbf{H}_N \Delta_q^\top$ can be expressed as follows:

$$\frac{1}{N} \|\mathbf{H}_N \Delta_q^\top\|^2 = \|\Delta_q^\top\|^2. \quad (\text{B.3})$$

When $q \neq 1$, the right side of (B.3) is $N\{(1 + \delta_q^{(+)})^2 + (1 + \delta_q^{(-)})^2\}/2$. The first and q th elements of $\mathbf{H}_N \Delta_q^\top$ are $N(\delta_q^{(+)} - \delta_q^{(-)})/2$ and $N(2 + \delta_q^{(+)} + \delta_q^{(-)})/2$, respectively, and a sum of squares of them becomes $N^2\{(1 + \delta_q^{(+)})^2 + (1 + \delta_q^{(-)})^2\}/2$. To satisfy (B.3), therefore, the other elements of $\mathbf{H}_N \Delta_q^\top$ must be all 0. Similarly, when $q = 1$, the right side of (B.4) is $N(1 + \delta_q^{(+)})^2$, and the first element of $\mathbf{H}_N \Delta_q^\top$ is $N(1 + \delta_q^{(+)})$. Therefore, the other elements

of $\mathbf{H}_N \mathbf{\Delta}_q^\top$ must be all 0. Since each element of $\mathbf{H}_N \mathbf{\Delta}_q^\top$ is a dot product of $\mathbf{h}_p$ and $\mathbf{\Delta}_q^\top$, the following orthogonality is then obtained:

$$\mathbf{h}_p \mathbf{\Delta}_q^\top = 0 \text{ for } p \neq q, p \neq 1. \quad (\text{B.4})$$

$$\begin{bmatrix} + & + & + & + & + & + & + & + \\ + & - & + & - & + & - & + & - \\ + & + & - & - & + & + & - & - \\ + & - & - & + & + & - & - & + \\ + & + & + & + & - & - & - & - \\ + & - & + & - & - & + & - & + \\ + & + & - & - & - & - & + & + \\ + & - & - & + & - & + & + & - \end{bmatrix}$$

$N = 8$

$$\begin{bmatrix} + & + & + & + & + & + & + & + & + & + & + & + & + & + & + \\ + & - & + & - & + & - & + & - & + & - & + & - & + & - & + \\ + & + & - & - & + & + & - & - & + & + & - & - & + & + & - \\ + & - & - & + & + & - & - & + & + & - & - & + & + & - & + \\ + & + & + & + & - & - & - & - & + & + & + & + & - & - & - \\ + & - & + & - & - & + & - & + & + & - & + & - & - & + & + \\ + & + & - & - & + & + & - & + & - & - & + & - & - & + & - \\ + & + & + & + & + & + & + & + & - & - & - & - & - & - & - \\ + & - & + & - & + & - & + & - & - & + & - & + & - & + & + \\ + & + & - & - & + & + & - & - & - & + & + & - & - & + & + \\ + & - & - & + & + & - & - & + & - & + & + & - & - & + & + \\ + & + & + & + & - & - & - & - & - & - & - & - & + & + & + \\ + & - & + & - & - & + & - & + & - & + & - & + & + & - & + \\ + & + & - & - & - & - & + & + & - & - & + & + & + & - & - \\ + & - & - & + & - & + & + & - & - & + & + & - & + & - & + \end{bmatrix}$$

$N = 16$

Figure B.1: The Hadamard matrices of orders 8 and 16. + and - indicate +1 and -1, respectively.

B.2 SNR Caluculation

The SNR of the proposed waveguide slot array with the CDM function is formulated as follows. The average power of the interference signal for the m th slot pair is given by

$$E[|S_{m,i}(t)|^2] = E \left[\left| \frac{g}{T_b} \int_0^{T_b} n(t) c_m(t) dt \right|^2 \right]. \quad (\text{B.5})$$

Using the Schwarz's inequality [B.3], the right side of (B.5) becomes

$$\begin{aligned} E \left[\left| \frac{g}{T_b} \int_0^{T_b} n(t) c_m(t) dt \right|^2 \right] &\leq \frac{|g|^2}{T_b^2} E \left[\int_0^{T_b} |n(t)|^2 dt \cdot \int_0^{T_b} |c_m(t)|^2 dt \right] \\ &= \frac{|g|^2}{T_b} \int_0^{T_b} E[|n(t)|^2] dt = |g|^2 \sigma_n^2. \end{aligned} \quad (\text{B.6})$$

Therefore, the worst-case SNR at the element output becomes (3.12).

Now consider an SNR at the array output. If all incident signals to the slot pairs have an equal power, the gain of each slot pair are identical ($|f_{m\phi}|^2 = G_e$), and the amplitude of transmission characteristics of each slot pair through the waveguide are identical ($|\alpha_m| = \alpha_0$), the maximum average power of the combined desired signal is calculated as

$$\begin{aligned} E \left[\left| \sum_{m=1}^M w_m S_{m,d}(t) \right|^2 \right] &= E \left[\left| g \sum_{m=1}^M w_m \alpha_m f_{m\phi} s_m(t) \right|^2 \right] \\ &= M^2 |g|^2 \alpha_0^2 G_e P_s. \end{aligned} \quad (\text{B.7})$$

where w_m are excitation weights and have an unity amplitude. Similarly, a combined interference signal at the array output is given by

$$\begin{aligned} \sum_{m=1}^M w_m S_{m,i}(t) &= \frac{g}{T_b} \sum_{m=1}^M w_m \int_0^{T_b} n(t) c_m(t) dt \\ &= \frac{g}{T_b} \int_0^{T_b} n(t) \sum_{m=1}^M w_m c_m(t) dt. \end{aligned} \quad (\text{B.8})$$

Using the Schwarz's inequality and the orthogonality of the WH codes, an average power of the combined interference signal is then calculated as

$$\begin{aligned} &E \left[\left| \frac{g}{T_b} \int_0^{T_b} n(t) \sum_{m=1}^M w_m c_m(t) dt \right|^2 \right] \\ &\leq \frac{|g|^2}{T_b^2} \int_0^{T_b} E[|n(t)|^2] dt \cdot E \left[\int_0^{T_b} \left| \sum_{m=1}^M w_m c_m(t) \right|^2 dt \right] \\ &= \frac{|g|^2 \sigma_n^2}{T_b} E \left[\int_0^{T_b} \sum_{m=1}^M |w_m|^2 c_m^2(t) dt \right] \\ &= \frac{|g|^2 \sigma_n^2}{T_b} \cdot MT_b = M |g|^2 \sigma_n^2. \end{aligned} \quad (\text{B.9})$$

Using (B.8) and (B.9), the SNR at the array output becomes (3.13).

B.3 Signal Bandwidth

In order to prevent deterioration in recovered signals, it is necessary to use the WH codes with a length of $L_c > M$. This requirement and the available PIN diode's switching speed determine the maximum number of multiplexing signals and maximum signal bandwidth in the proposed antenna. Table B.1 shows the requisite switching speeds of PIN diode, which are determined by the signal bandwidth and the multiplexing number. The switching time of PIN diodes, even those of the fastest type, is approximately 20 ns [B.4], [B.5].

This limits the signal bandwidth and the multiplexing numbers that can be achieved by the proposed antenna. For example, when multiplexing eight signals with the proposed antenna, the maximum signal bandwidth is approximately 2.5 MHz.

Table B.1: Switching speeds of PIN diode required for the proposed waveguide slot array.

Bandwidth [MHz]	Multiplexing number			
	2 to 3	4 to 7	8 to 15	16 to 31
0.625	400	200	100	50
1.25	200	100	50	25
2.5	100	50	25	12.5
5	50	25	12.5	6.25
10	25	12.5	6.25	3.125

Unit : ns

Reference

- [B.1] W. K. Pratt, J. Kane, and H. C. Andrews, “Hadamard transform image coding,” Proc. IEEE, vol. 57, no. 1, pp. 58–68, Jan. 1969.
- [B.2] M. P. Deisenroth, A. A. Faisal, and C. S Ong, Mathematics for Machine Learning, Cambridge University Press, Cambridge, 2020.
- [B.3] G. B. Arfken and H. J. Weber, Mathematical methods for physicists, 5th ed., Academic Press, San Diego, 2000.
- [B.4] G. A. Kumar and A. Kumar, “Design of low loss, high speed, and high isolation Ka-band SPST pin switch in suspended stripline configuration,” Microw. Opt. Technol. Lett., vol. 54, no. 8, pp. 1808–1810, Aug. 2012.
- [B.5] Millimeter Wave Products Inc., “Pin Diode MMIC Switches,” <https://www.miwv.com/pin-diode-switches/>, accessed June 15. 2024.

Appendix C

Appendix for Chapter 4

C.1 Effect of the Modification Functions in (4.10)

The modification functions $f_\theta(\theta)$ and $f_\phi(\theta)$ are introduced in (4.10) to approximate the effect of the superstrate. Here, the effectiveness of the modification functions is evaluated by comparing the calculation results with and without them. Figures C.1 shows the calculated and simulated linearly polarized AEPs at the E-plane ($\phi = 135^\circ$) and the H-plane ($\phi = 45^\circ$), respectively, when only the feeding port 1 is excited. Because the AEPs in each plane have almost symmetrical patterns, only one side of them is shown in Fig. C.1. The optimized parameters for modification functions are $\alpha = 0.02$ and $\beta = -0.5$. The AEPs calculated by the optimization without the modification functions and the simulated AEPs without the superstrate are also shown in Figs. C.1 for comparison. The superstrate has an effect of enhancing E_ϕ especially at wide-angle directions at the H-plane, whereas it has a slight effect on E_θ at the E-plane. There is a large discrepancy between the calculated E_ϕ without the modification functions and the simulated one with the superstrate especially at $|\theta| > 60^\circ$. This discrepancy is mitigated by introducing the modification functions, and the calculated E_ϕ with the modification functions well agrees with the simulated one even at wide-angle directions.

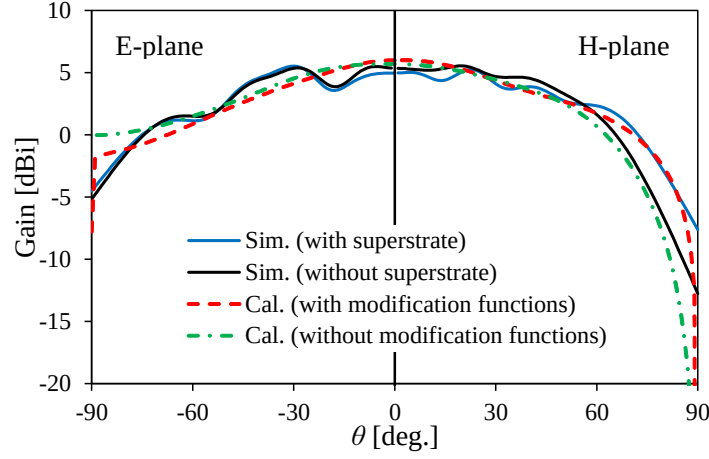


Figure C.1: Simulated and calculated the AEPs with and without the superstrate. The left and right sides show E_θ at the E-plane and E_ϕ at the H-plane, respectively.

Fig. C.2 shows the AR variations with the scan angle after the AR compensation, where the compensation weights are determined by a design formula without the modification functions. Compared with Fig. 4.9(b), the AR improvement is generally smaller at $|\theta_s| > 65^\circ$, and the worst AR in the coverage range is 2.7 dB after the AR compensation, which is 0.7 dB higher than that in Fig. 4.9(b).

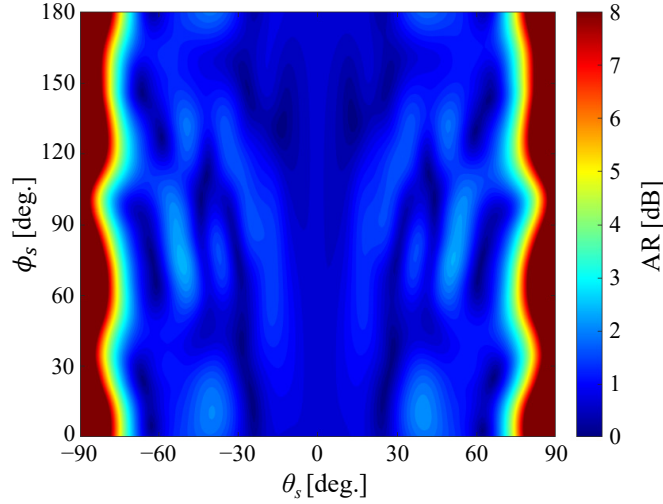


Figure C.2: AR variations with the scan angle after the AR compensation. without the superstrate: (a) E-plane and (b) H-plane.