

Integral Study of GaN Amplifiers and Antenna Technique for High-Power Microwave Transmission

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Integral Study of GaN Amplifiers and Antenna Technique for High Power Microwave Transmission

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Abstract

The microwave transmission recently requires a high power RF circuit and antenna system based on continuous wave (CW) for wireless power supply and space satellite applications. The GaN amplifiers and antenna technologies are necessary for the high power microwave transmission. In the modern works, these technological fields have been considered as the different field. In this work, the integration of the design methods of GaN amplifier and RF antenna technologies is studied. In addition, the GaN amplifier and RF antenna are comprehensively designed for the high efficiency and effective microwave transmission. In particular, the following programs are discussed in this work: the high power GaN amplifier design based on a combiner technique for CW operation; the high efficiency GaN amplifier design and integration with an antenna for beam combining in a free space based on active array antenna techniques; the sidelobe reduction based on the active antenna techniques.

In the study of high power GaN amplifier design based on a combiner technique for CW operation, a GaN amplifier is firstly designed and fabricated at C-band (7.1 GHz) frequency on CW condition. The fabricated GaN amplifiers are combined by a 20-way radial combiner based on a waveguide circuit. The assembled 20-way high power GaN amplifier is assumed to use for a grand base station for satellite application. In the measurement of properties of the 20-way high power GaN amplifier, 179.8 W RF output power is observed in 7.1 GHz frequency and CW condition, which stand comparison with conventional GaN amplifiers operated in pulse condition. This research achievement contributes to high power microwave transmission based on large-scale aperture antenna system with a high power amplifier.

For the effective microwave transmission, the beam combining in free space

based on active array antenna techniques is required. It is necessary for the active array antenna to design a high efficiency GaN amplifier and integrate with a antenna element. In this work, the high efficiency GaN amplifier with more than 60 % power-added efficiency (PAE) is design and fabricated by a harmonic tuning technique at C-band frequency (5.8 GHz). In this techniques, the second harmonic is treated by a harmonic tuning circuit, which makes an improvement of more than 10 point PAE at comparing the amplifier without the harmonic circuit in this work. In addition, an antenna design method for loss reduction in integration with amplifier is proposed. In this method, the impedance ratio of antenna to GaN HEMT is reduced, which contribute to high efficiency active antenna design. In fabrication of the active antenna, the integration loss is reduced at comparing conventional studies.

The low sidelobe techniques are necessary for compatibility of the high power microwave transmission system with other wireless system and ambient environment. In this work, the new beam controlling approach based on the active array antenna technique is proposed. It is possible to control the saturated output power levels of amplifier by operated DC drain bias voltage. In this approach, this amplifier property is used for a sidelobe reduction based on active array antenna, which makes possible tunably to control the sidelobe level. In an experiment, the sidelobe reduction with one dimension active array antenna based on this technique is demonstrated.

This comprehensive studies of high power and high efficiency GaN amplifiers and antenna is expected for development of the high power microwave transmission. This research is hope to contribute the progressive social system with the microwave transmission technology.

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Chapter 1

General Introduction

1.1 Microwave Transmission

Microwave Transmission is an absolutely imperative technology for saving our social living. The technologies of microwave transmission are widely used in many fields such as communication, space satellite, radar, and military. Recently, microwave Power transmission (MPT) is particularly expected for next-generation method of electric power distribution. Microwave is an electromagnetic wave with frequencies between 300 MHz and 300 GHz, with corresponding electrical wavelength between 1 m and 1 mm, respectively. The microwave is included in radio wave frequency range within the 30 kHz to 300 GHz. The location of microwave frequency band in the electromagnetic wave is shown in Fig. 1.1 [1]. The approximate band designations of microwave is described by IEEE as shown in Table 1.1. In the high-power and long-range microwave transmission, the frequency bands between L to X band are often used. The microwave transmission of these frequency bands is based on a vacuum tube and GaN semiconductor technologies which is imperative for high-power microwave generation. The typical uses of L to X band are shown as follows: military telemetry, GPS, mobile phones, and amateur radio used in L band; weather radar, surface ship radar, microwave oven, and satellite communication used in S band; long-distance radio telecommunications used in C band; satellite communications, radar, terrestrial broadband, amateur radio, molecular rotational spectroscopy used in X band. In particular, the frequencies of 915 MHz, 2.45 GHz, and 5.8 GHz between L and X band are internationally reserved for the industry science medical (ISM) bands used for radio frequency

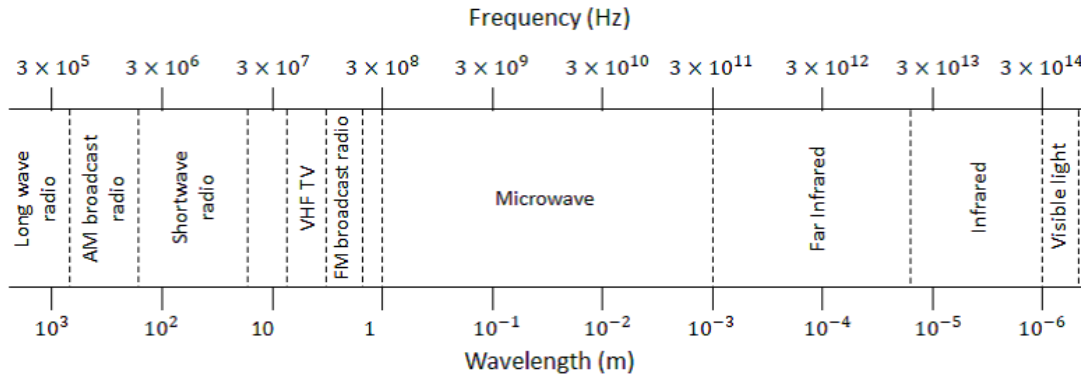


Figure 1.1: Electromagnetic wave spectrum [1].

Table 1.1: Approximate band designations of microwave described by IEEE.

Band designation	Frequency range
HF (High frequency)	3 MHz to 30 MHz
VHF (Very high frequency)	30 MHz to 300 MHz
UHF (Ultra high frequency)	300 MHz to 3 GHz
L (Long-wave) band	1 GHz to 2 GHz
S (Short-wave) band	2 GHz to 4 GHz
C (Compromise between S and X) band	4 GHz to 8 GHz
X (Exotic/Extreme-short) band	8 GHz to 12 GHz
K _u (Kurz-under) band	12 GHz to 18 GHz
K (Kurz) band	18 GHz to 27 GHz
K _a (Kurz-above) band	27 GHz to 40 GHz
V band	40 GHz to 75 GHz
W band	75 GHz to 110 GHz
mm (Millimeter) band	110 GHz to 300 GHz

(RF) energy of industrial, scientific, and medical purposes other than telecommunications. The MPT studies mainly uses the ISM band.

The microwave transmission studies are based on antenna and amplifier techniques. In particular, the long-range microwave transmission based on continuous wave (CW) operation such as MPT, satellite communication, and radar requires the high-power and high-efficiency technologies of the device and propagation. In

conventional studies of microwave transmission, the techniques of antenna and amplifier were cognized for discrete study topics. In addition, the modern microwave-transmission systems basically require the beam forming function based on the phased array technology. Therefore, it is necessary for next stage of the high-power microwave transmission to integral study of these topics. The goal of this study is to propose an integrally design techniques of GaN amplifier, antenna, and radiated beam form for high-power and high-efficiency microwave transmission.

1.2 Applications of High Power Microwave Transmission

A lot of applications with the CW-based high power microwave transmission are operated and proposed. For example, the some satellites such as PROCYON launched in 2014 by JAXA communicate to ground base station with CW beacon. The transmission distance between the ground base station and satellites is at least hundreds of kilometer, which requires the high power microwave radiation. This satellite communication system consists of the parabolic antenna system as shown in Fig. 1.2(a) for high gain transfer and receiver. On the other hand, the long-range MPT systems such as space-solar power satellite (SSPS) as shown in Fig. 1.2(b) requires the high power microwave transmission. SSPS as shown in Fig. 1.2(b) is a space-based solar power system projected by JAXA and kyoto university for future space utilization. The photograph as shown in Fig. 1.2(b) is an artist conception of the SSPS proposed by Institute for Unmanned Space Experiment Free Flyer (USEF). In addition, the MPT technologies are required to use for power supplying to electronics such as EV, mobile phone, and other devices. These MPT systems consist of phased array antenna system which is necessary for future microwave transmissions. In particular, the wireless power suppling for EV is one of the most featured application of MPT as shown in Fig. 1.2(c), which is researched and developed, recently. It is necessary for these applications with the high power microwave transmission to develop the technologies of antenna and high power excitation of microwave. In conventional system, the vacuum tube technologies such as magnetron and klystron have been used for the microwave excitation. However, the vacuum tubes are breakable and unreliable. Hence, the irrefrangible and stable high-power amplifier technologies are recently used. In

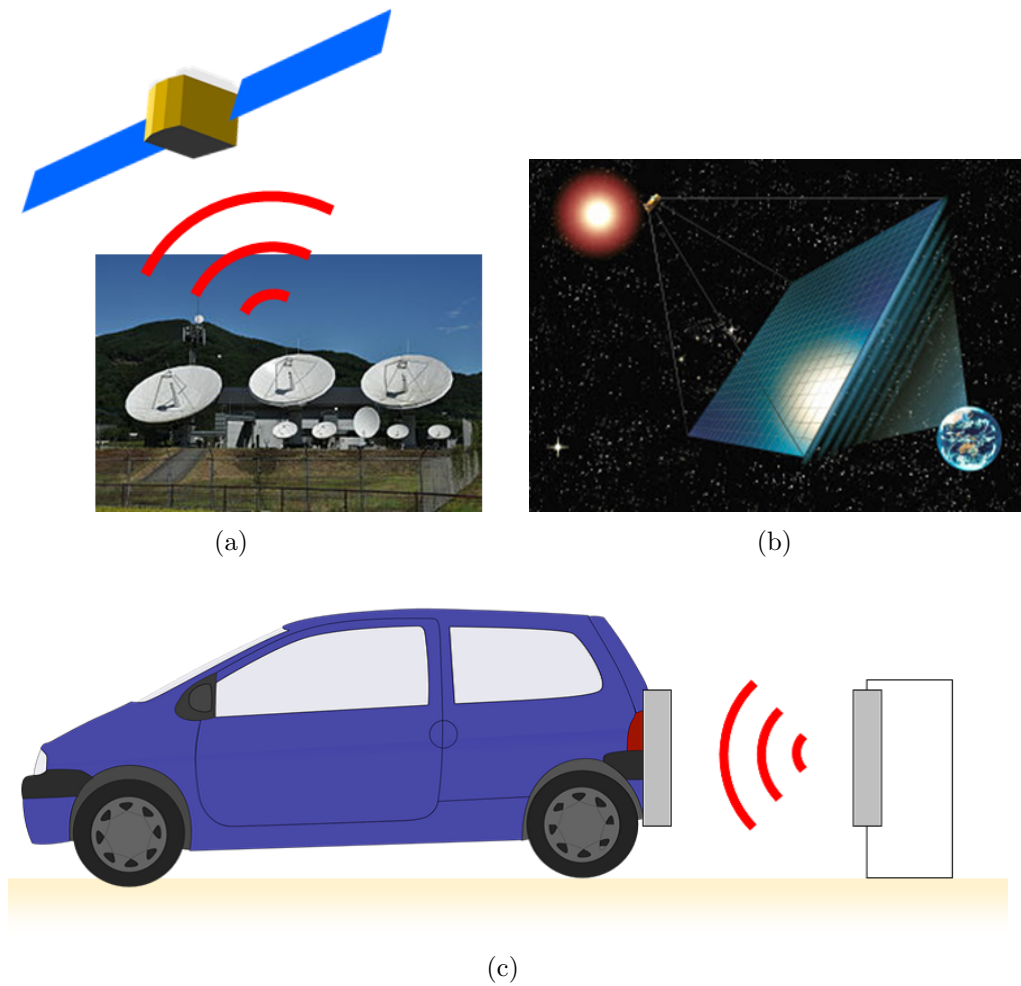


Figure 1.2: Applications of high power microwave transmissions: (a) satellite communications, (b) space solar-power satellite, and (c) wireless power supplying to EV.

this work, the high power and high efficiency amplifier and antenna technologies are comprehensively studied for development of these high power microwave transmission systems.

1.3 History of Microwave Engineering

The fundamental concepts of electromagnetics have developed since early times. Even though the field of microwave engineering has a history over 100 years, the significant developments in high-frequency solid-state devices, microwave integrated circuits, and microwave transmission systems have kept the field active and vibrant.

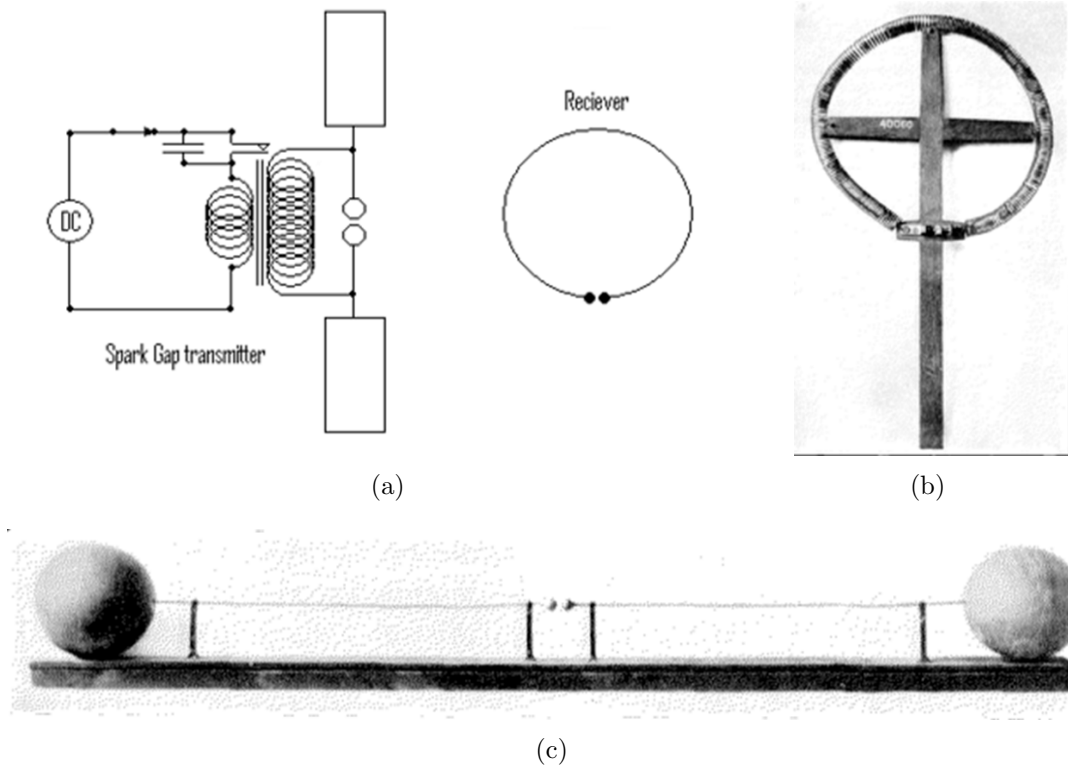


Figure 1.3: Original apparatus used by Hertz for his experiments of electromagnetic wave propagation in 1887: (a) schematic of demonstration system, (b) loop antenna for receiving electromagnetic wave, and (c) dipole resonator consisting of one meter copper wires ending in 30 cm zinc spheres for radiation.

The fundamental theory of modern electromagnetics was formulated in 1873 by Maxwell [2]. His hypothesis with mathematical considerations formulated the electromagnetics and the wave speed as fast as light speed. The original form of Maxwell's equations were formulated by 20 equations included the three orthogonal directions components.

The Maxwell's equations were reformulated to the vector form by Heaviside, during the period from 1885 to 1887. In the Heaviside's reformulation, 20 forms of Maxwell's equations were reduced to 4 forms known as Maxwell-Heaviside equations by using a vector calculus theory and removing the equations of vector potential and electrostatic potential. Heaviside also developed the transmission line theory, also known as the telegrapher's equations, which became a basis for modern microwave theory. The following terms of art in transmission line theory was coined by Heaviside: admittance in December 1887; conductance in September 1885; impedance in July 1886; inductance in September 1886; permittance which was later susceptance in June 1887; reluctance in May 1888, which contributed



Figure 1.4: Photograph of Wardencliff tower built by Tesla in 1905.

briefly to understand the theory.

The availability of the Maxwell's theory for electromagnetic wave traveling in free space was demonstrated by Hertz, during the period from 1887 to 1891. The original apparatus used by Hertz for his experiments of electromagnetic wave traveling in 1887 is shown in Fig. 1.3. In his experiments, Hertz invented a first radio transmitter of half-wave dipole resonator consisting of one meter copper wires ending in 30 cm zinc spheres and receiver of loop antenna as shown in Fig. 1.3(b) and Fig. 1.3(c). Hertz's experiment completely validated Maxwell's theory, which contributed significantly to develop the modern microwave engineering. In addition, the antenna techniques studied in this century brought innovation to the microwave transmission.

From the 20th century onward, the experimental studies of radio wave propagation officially started. A large-scale experiment of a long-range radio wave propagation was demonstrated by Tesla in the 1900s. Tesla designed and built Wardencliff tower as shown in Fig. 1.4, also known as Tesla tower, sponsored by John Pierpont Morgan in 1905. A concept of the tower was wirelessly to

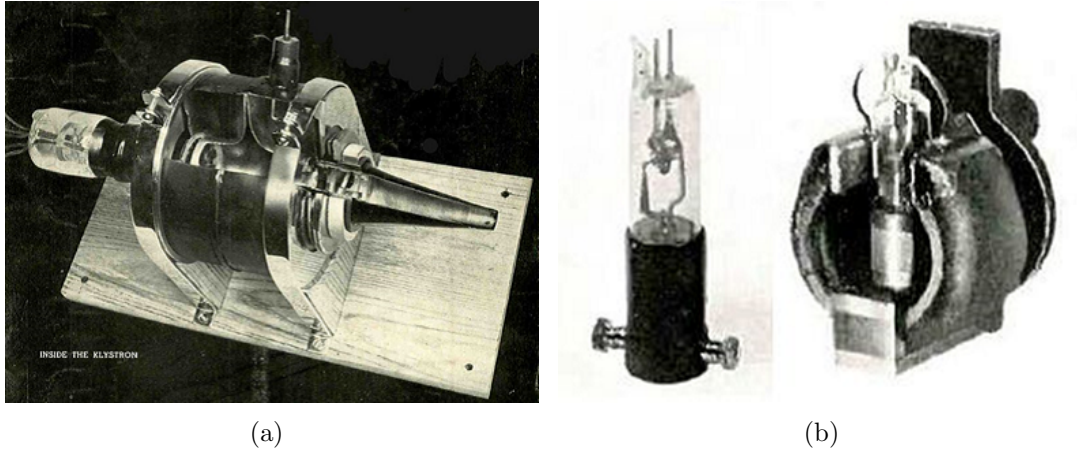


Figure 1.5: Photographs of first prototype tubes: (a) klystron manufactured by Westinghouse Electric Corporation in 1940 and (b) early forms of magnetron in 1935.

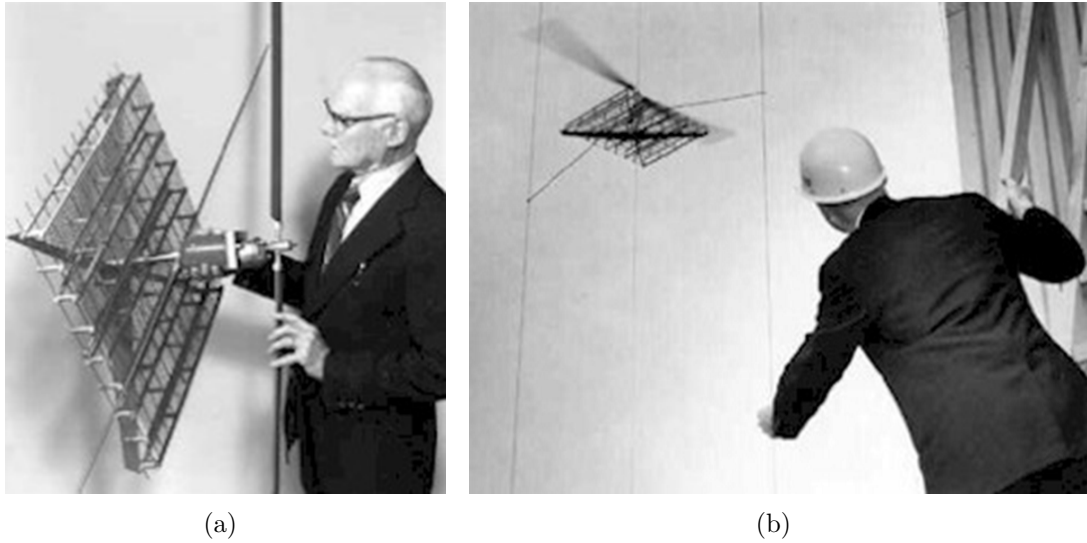


Figure 1.6: Photographs of MPT demonstration to helicopter by Brown in 1964: (a) used helicopter and (b) condition of the demonstration [5].

transfer and telegraph power, information, and image for early radio wave station. This tower used 150 kHz radio wave and 300 kW output power. In the WPT experiment, this Tesla's attempt was unsuccessful because of the low transmission efficiency. It was necessary for improving the efficiency to increase the frequency. However, the Tesla's ambitious attempt contributed the modern radio wave transmission technologies such as mobile communication and wireless power transmission (WPT).

Just before World War II, the operated frequency of microwave transmission were drastically developed by vacuum tube such as klystron and magnetron signal



Figure 1.7: Long-range MPT demonstration over a distance of 1.6 km by Brown in 1975 [6].

generators which were the first significantly powerful microwave source and antenna technologies. Klystron and magnetron as shown in Fig. 1.5 were invented for radar and radiotelephone by the Varian brothers in 1937 and Hans Gerdien in 1910, respectively. These vacuum tube technologies were used by the first commercial microwave radiotelephone system in 1934 operated between England and France at 1.8 GHz. In addition, the radar theory and engineering were developed by the microwave technologies based on the vacuum tubes during World War II.

The microwave transmission technologies were inherited to MPT. The MPT study using a rectifying antenna known as rectenna was started in 1958 by Raytheon, US Air Force, and NASA. In the first demonstration of MPT with 400 W class magnetron, the 13 % DC-to-DC efficiency and 100 W DC power received were confirmed [3], [4]. In 1964, Brown in Raytheon demonstrated a microwave-powered helicopter as shown in Fig. 1.6 [5]. In this demonstration, the helicopter is powered by 2.45 GHz microwave. Although the vertical flight distance was constrained to 170 cm, 54 % DC-to-DC efficiency was achieved. Brown also tried MPT to 1.6 km distance in 1975 as shown in Fig. 1.7, which made a success of 30 kW DC power received [6], [7]. In this long-range MPT, a 26 m-diameter parabolic antenna for

Table 1.2: Comparison of semiconductor electric properties at 300 K [9]-[11].

Properties	Si	3C-SiC	4H-SiC	6H-SiC	GaAs	GaN
E_g (eV)	1.12	2.36	3.26	3.02	1.42	3.42
E_c (MV/cm)	0.3	1.5	2.8	3.0	0.4	3.0
μ_e (cm ² /V·s)	1400	1000	1000	450	8500	1200
v_s (10 ⁷ cm/s)	1.0	2.2	2.2	2.2	1.2	2.5
σ_r (W/cm·K)	130	450	450	450	55	110
ε_r	11.9	9.7	9.7	9.7	13.1	9.5

transmission and $3.4 \text{ m} \times 7.2 \text{ m}$ rectenna were used. 450 kW microwave was radiated from transmission antenna in the demonstration.

The vacuum tube techniques have been used for decades, which contribute to develop the high-power microwave transmission technologies. In the many fields, the vacuum tube such as traveling wave tube amplifier (TWTa) typically used in C- to X-band frequency have been used for radio frequency (RF) high-power amplifier. However, the high-power tubes in microwave range is recently being replaced to solid-state devices based on composite semiconductor technology. In 1966, a GaAs metal-semiconductor field-effect transistor (MESFET) was invented by Mead, which is still used for low-noise amplifiers in up to 30 dBm power level. In 1979, a high electron mobility transistor (HEMT) was invented by Mimura, which was possible to operate at frequencies well above 30 GHz. The operated power level of the GaAs FETs were not comparable to TWTAs. However, the output power level of these solid-state devices is already significantly improved above 50 dBm by GaN semiconductor technologies, recently. The GaN extrinsic semiconductor was invented by Akasaki and Amano in the 1980s for blue light-emitting diode (LED). The first GaN MESFET was demonstrated by Khan et al. in 1993 [8]. The GaN HEMT devices which is widely used in fields such as satellite communication, base station for mobile phone, and MPT are necessary for the modern development of high-power microwave transmission technologies.

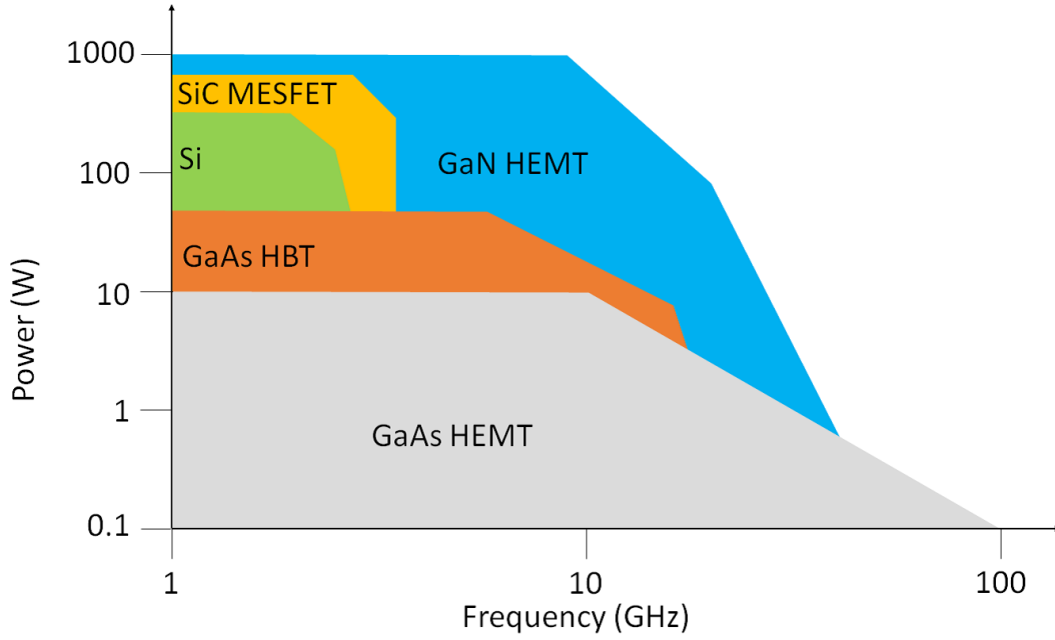


Figure 1.8: Frequency and power regions of each semiconductor devices for RF amplifier [12].

1.4 Semiconductor Technologies for Microwave Transmission

The modern microwave transmission have been developed by semiconductor technologies. Recently, the highly used devices in microwave frequency are the HEMT based on composite semiconductor such as GaN and GaAs. The electric properties of GaAs and GaN semiconductor compared to Si and SiC are shown in Table 1.2 [9]-[11]. In the Table 1.2, E_g is bandgap, E_c is breakdown field, μ_e is electron mobility, v_s is saturated electron velocity, σ_r is thermal conductivity, and ϵ_r relative dielectric constant, respectively. The crystal systems of SiC are explained for 3C (cubic), 4H (hexagonal), and 6H. According to Table 1.2, GaN has wider bandgap and higher breakdown field than Si, and GaAs, which improves the high power property of device. On the other hand, the electron mobility and saturated electron velocity of GaN are faster than SiC, which improves the high frequency property.

In the RF amplifier design, GaAs HEMT or heterojunction bipolar transistor (HBT) have conventionally used before GaN HEMT developed. The GaAs RF amplifier still be studied in fields of wide frequency band, and higher frequency than 100 GHz [13]-[15]. GaAs has a higher electron mobility than Si, SiC, and

GaN, which is suitable for improving the noise property of RF amplifier. Hence, the GaAs devices are also used for low-noise RF amplifier in the low-power region. However, the GaAs devices are not able to operate high voltage because of lower bandgap property than SiC and GaN. The operated DC voltage of GaAs devices is not more than 10 V. On the other hand, GaN absorbs higher DC voltage than GaAs because of wider bandgap, which is possible to operate higher RF output power per one cell. In general, the FET cells parallel connected in a chip for increasing output power are used for high power amplifier device. Comparing GaN HEMT to GaAs, the GaN HEMT device operated in high power is able to reduce the cells, which prevent to decrease the impedance of device. This characteristics of wide band gap device contribute to design the high power RF circuits. The frequency and power regions of each semiconductor devices for RF amplifier is shown in Fig. 1.8 [12]. According to Fig. 1.8, GaN HEMT is used in higher power than GaAs devices and higher frequency region than Si and SiC devices.

1.5 Requirement for High Power Antenna System

Recently, high power microwave transmission is used for space communication, radar, and microwave power transmission based on CW operation. In the CW operation, it is possible to use the saturated regions of active devices such as amplifier, where active devices is expected to operate under maximum efficiency. However, loss of the high power devices is converted to temperature increasing in the CW condition. Therefore, it is necessary for the CW high power device to improve the efficiency.

The integrally design contributes for the compact and lightweight microwave transmission systems. Up to kW-class microwave transmission system based on the phased array antenna consists of a lot of antenna elements. In conventional discrete approach, these antenna elements are connected to circuit by connectors such as subminiature version A (SMA). For example, MPT system of SSPS is assumed to consist of billions of antenna elements. If the every connector between antenna and circuits are removed, the weight of SSPS will be massively reduced. In particular, assuming 2 g weight of the connector for connecting a billion antenna to circuits, the total weigh of connectors above 200 Mg are calculated. The active

antenna techniques printed on one substrate and multilayer substrate are recently reported [16], [17], which are useful for large-scale microwave transmission systems.

The active antenna consisting of planar structure are also useful for integration with a cooling system such as Peltier device. The cooling system prevents the temperature increasing of amplifier which causes the adverse effects such as efficiency and output power reduction. In addition, the cooling system prevent the effect of environmental temperature and heat generation of device, which contribute a high-integrity experience. The integration design of active antenna with colling system contributes an effective cooling and a compact design of active antenna module.

In the long-range and high-power microwave transmission system, the beam forming technique are absolutely imperative for high-efficiency microwave propagation to a target and electromagnetic interference (EMI) programs. In the long-range system, the a few angle error of transmission antenna cases a critical program such as reduction of transmission efficiency. Recently, many studies for angle controlling of microwave beam based on a phased array antenna are reported [18]-[20]. The beam angle is easily controlled by phase shifters assembled in the each antenna element of phased array antenna. On the other hand, the high-power microwave has a possibility to affect the environment as an EMI. The array antenna techniques are a lot of advantage for high-power microwave transmission. However, the incidence of sidelobe is inevitable in the array antenna theory. In general, the array antenna has the strongest gain in main beam for front direction. The sidelobe is unnecessary radiation generated in sides of the main beam. The the theoretical sidelobe reduction techniques have been studied [21]-[23]. The circuit approach using a non-uniform power divider, filter, or resonator and non-uniformly spaced array antenna approach are mostly used for sidelobe reduction. However, the disadvantages of these approaches are not able tunably to control. Hence, the tunable approach for sidelobe reduction is required.

The requirement of high power microwave transmission system are concluded for the following four programs:

1. High power radiation;
2. High efficiency microwave amplification;

Table 1.3: Conduction angle of each classical mode.

Class	Normalized bias voltage (V)	Conduction angle (rad)
A	0.5	2π
AB	0 to 0.5	π to 2π
B	0	π
C	< 0	0 to π

3. Effective integration of devices including amplifier, antenna, and cooling system;
4. Beam forming.

In this work, the integrally consideration based on GaN amplifier and antenna techniques are tried. In addition, the tunable approach for sidelobe reduction by using the active array antenna integrated with high-power and high-efficiency GaN amplifier are proposed.

1.6 High Power Amplifier

GaN amplifier and antenna are important technologies for microwave transmission. The high power and high efficiency antenna systems have been studied, which are realized by effective GaN amplifier and active antenna design. The small-signal amplifier are often designed by using S parameters of transistor. However, the small-signal S parameters are not suitable for high power amplifier design. In the large-signal condition, transistor has different characteristics to small-signal condition due to the non-linear properties of transistor. Therefore, it is important for obtaining the excellent characteristics of amplifier to design with the transistor informations in high-power condition.

In the classical high power amplifier design, operated class is categorized as class A, AB, B, and C. The operated class of amplifier are defined with the aid of a bias of DC drain voltage for FET. The class A bias point is operated in half-value amplitude of RF drain voltage for providing a 2π conduction angle of drain current. In the class A operation, output signal contains the harmonics,

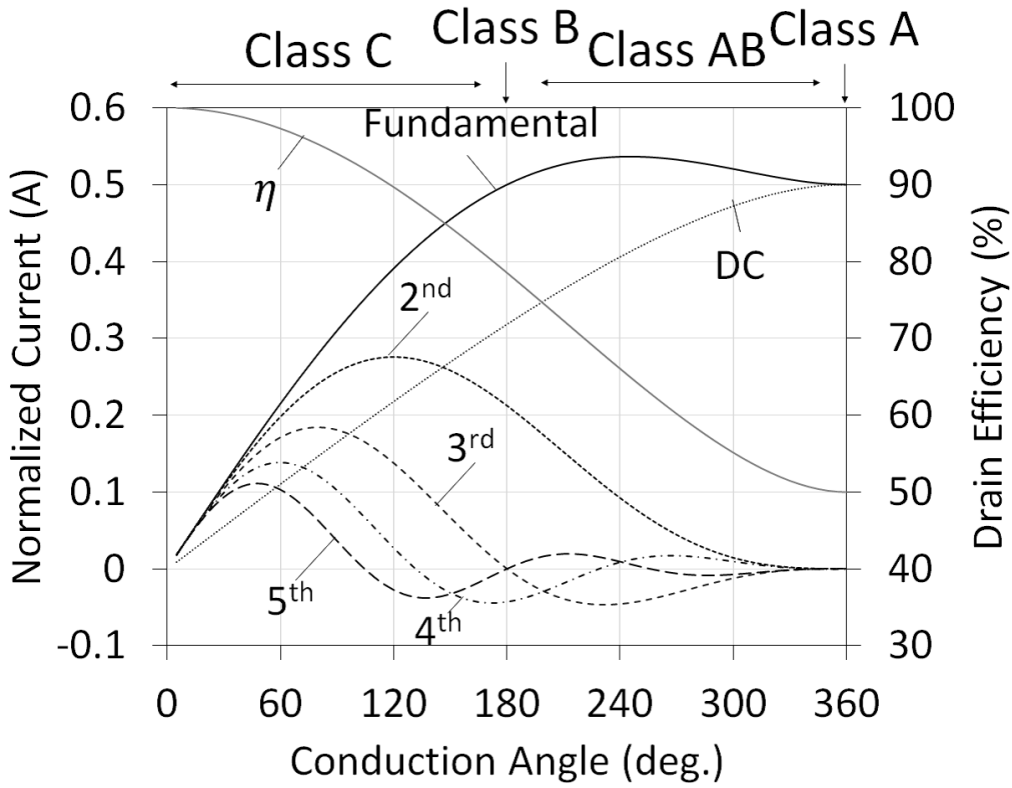


Figure 1.9: Calculated normalized current by peak drain current and drain efficiency versus conduction angle [24].

which provide a well linearity to amplifier property. However, the drain efficiency, defined by the equation $\eta = P_{\text{RFout}}/P_{\text{DCin}}$ where P_{RFout} is an output RF power and P_{DCin} is an output DC power, are theoretically up to 50 %. The practical value of drain efficiency in class A operation is about 25 % to 30 %. In the class B and C condition, the DC drain voltage is operated below 0 V, which provide up to π of the conduction angle of output current. The class B is defined only in 0 V of DC drain voltage. The theoretical drain efficiency of class B operation is 78.5 % which is increased in class C condition. The AB class amplifier is operated in the conduction angle between A and B class. The conduction angle and normalized bias voltage by peak drain voltage are concluded in Table 1.3. The normalized currents by peak drain current and drain efficiency versus conduction angle are illustrated in Fig. 1.9 [24]. In Fig. 1.9, the drain efficiency increases at decreasing the conduction angle. However, the normalized fundamental current decreases. The peak of current are confirmed at AB class condition.

The harmonics characteristics are also illustrated in Fig. 1.9. According to Fig.1.9, the harmonics are excited in the conduction angle range without class

A condition. In the AB class condition, the second harmonic current are larger than other harmonics. Hence, it is important program for high power amplifier design particularly to treat the second harmonic. In general, the harmonics are case of amplifier loss. The harmonic treatment techniques are used for modern high-efficiency amplifier design such as class F operation.

This work studies the microwave transmission system based on high-power GaN amplifier. The amplifier and active antenna are designed with the AB class operation and harmonic treatment techniques in this work.

1.7 Outline of This Thesis

The targets of this work are to propose the high-power and high-efficiency GaN amplifier design, the effective integration of the amplifier with antenna, and newly beam forming technique based on the active antenna approach. These topics are necessary techniques for the high power and high efficiency microwave transmission. The high-power GaN amplifier is required for the system based on an aperture antenna such as space applications. The high efficiency GaN amplifier, integrated active array antenna, and beam forming techniques are also necessary for the system based on an array antenna such as MPT. In this thesis, firstly, the high-power amplifier with the output power above 170 W is demonstrated, where the combiner techniques of amplifier are used for increasing output power and C-band 20-way GaN amplifier are demonstrated. Secondly, the high-efficiency amplifier with the harmonic treatment circuit is demonstrated, where the amplifier with power added efficiency (PAE) above 60 % at C-band are demonstrated. Thirdly, the effective integration technique for high-power and high-efficiency active antenna are proposed, where the low-loss active antenna with single patch antenna and GaN amplifier designed on Peltier cooling unit is demonstrated. Fourthly, the beam forming technique based on fabricated 4-way GaN active antenna for sidelobe reduction is proposed. In this technique, the sidelobe level are tunably controlled by suitable amplifier output power. Finally, the integral design techniques of GaN amplifier, antenna, and radiated beam form for high-power and high-efficiency microwave transmission are concluded.

Chapter 2

RF Antenna and Power Amplifier Technologies

2.1 RF Antenna Technologies

The RF antenna technologies have been developed and discussed since the mid 1800s. In 1842, the first radiation experiment was demonstrated by Joseph Henry. In this experiment, he threw a spark on a circuit and observed that needles were magnetized by the received current. In 1900s, a resonant length antenna such as a half-wavelength dipole was developed, which became a base of modern RF antenna technology. Recently, the RF antenna technologies used in fields of microwave transmission such as wireless communication, radar, and MPT become a necessary for wireless technologies.

The RF antenna is a device for radiating a guided wave to free space from a wire circuit and receiving the free-space wave. The often-used antenna performance parameters for long-range and high power microwave transmission are radiation pattern and gain. The radiation pattern is angular variation of radiation around the antenna. The gain is ratio of radiation intensity to isotropic or dipole antenna. The high-gain and low-sidelobe antenna is required for the high power microwave transmission. The antenna is categorized into the four basic types of electrically small antennas, resonant antennas, aperture antennas, and array antennas as shown in following list.

Electrically small antennas: The electrically antennas such as short dipole and small loop antenna are much less than a wavelength in extent. This antenna is often used for ideal antenna discussion.

Properties:

- Low gain
- Low radiation efficiency

Resonant antennas: The resonant antennas such as half-wave dipole, Yagi-uda, and microstrip patch antenna are used in wireless system with narrow frequency bands.

Properties:

- Low to moderate gain
- Narrow bandwidth

Aperture antennas: The aperture antennas such as horn and parabolic antenna has a physical aperture through microwave flow.

Properties:

- High gain
- Gain increases with frequency
- Moderate bandwidth

Array antennas: In this antenna, the low gain antenna elements such as microstrip patch antennas are arrayed at regular intervals.

Properties:

- High gain
- Flexible radiation pattern characteristics

In the high power microwave transmission, the aperture antenna and array antenna techniques are used. For example, a large scale parabolic antenna have been used in a ground-based station for space communication system. A parabolic antenna with 64 m-diameter reflector used in S- to X-band frequency was established in Usuda Deep Space Center, Japan, in 1984 [25]. The parabolic antenna has a property of very high gain, which makes it possible to transmit to a satellite at long-range distance. In 1975, a large-scale parabolic antenna was also used in

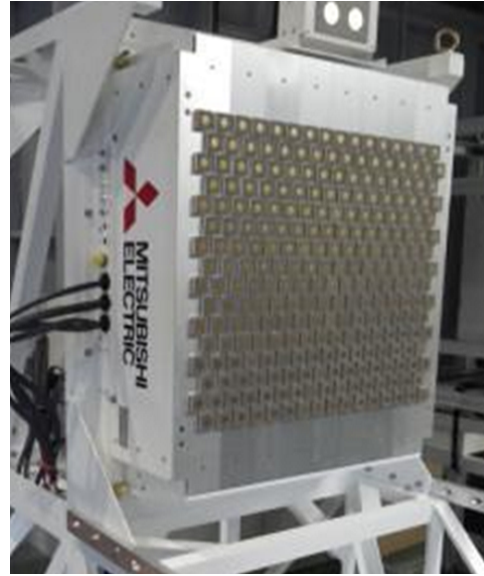
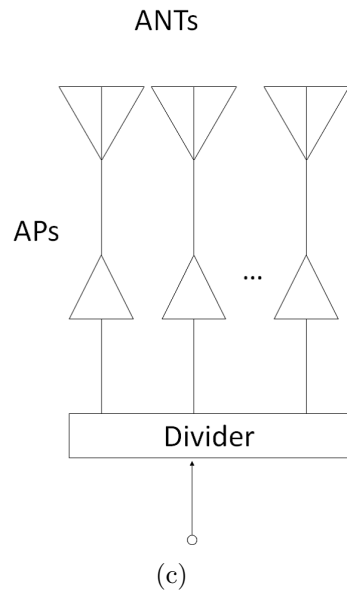
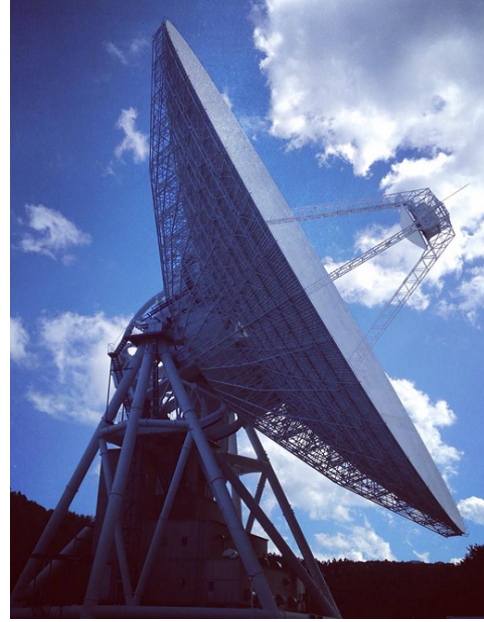
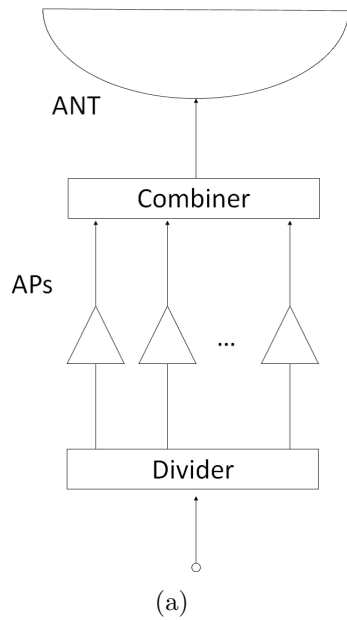


Figure 2.1: (a) Simple RF circuit configuration based on an aperture antenna, (b) photograph of a large-scale parabolic antenna, (c) simple RF circuit configuration based on an array antenna, and (d) photograph of an array antenna.

long-range transmission for MPT demonstration [6]. A lot of long-range and high power microwave transmission system have used the aperture antenna techniques. The simple RF transmission system based on an aperture antenna and a large-scale parabolic antenna are shown in Fig. 2.1(a) and Fig. 2.1(b). The aperture

antenna has one feed port. Hence, a high power amplifier is required for this antenna system. In general, in S- to X-band frequency, the operated output power of an amplifier with one chip GaN HEMT based on CW condition is up to 100 W. Therefore, in the transmission system required more than 100 W radiation based on the aperture antenna, the combiner technologies for high power amplifier is necessary. The output from amplifiers are combined on RF circuit in this system. In the aperture antenna system, beam angle and shape hardly control by electrical tuning. Therefore, the antenna system require a mechanical control system for beam forming.

On the other hand, in an array antenna technologies, the output from amplifier are combined in free space, which makes a high gain radiation pattern. Array antenna technologies are recently used for high power microwave transmission system such as MPT. One of the advantage of the array antenna is possible to add functions such as beam forming based on phase and amplitude tuning for each antenna element. Figure 2.1(c) and Fig. 2.1(d) show a simple circuit configuration of array antenna and an example of array antenna for MPT system based on phased array [26]. The phased array is an array antenna with phase shifters attached for each element, which make it possible to control the beam angle and shape by electrical tuning. In addition, amplifiers are often attached to each antenna element, directly, in the array antenna system, also known as active array antenna, which contribute to reduce the size of antenna system and to increase the strength of RF system with circuits, devices, and components for voltage.

The amplifier technologies are necessary for these high power antenna system. The aperture antenna system requires a high power amplifier based on combiner techniques in a RF circuit. The array antenna system also requires a high efficiency amplifier and integration techniques for an active array for high power microwave transmission. In this work, the high power and high efficiency amplifier for the aperture and array antenna are discussed and demonstrated. First of all, the GaN amplifier technologies are introduced and basic amplifiers are designed and fabricated in this section.

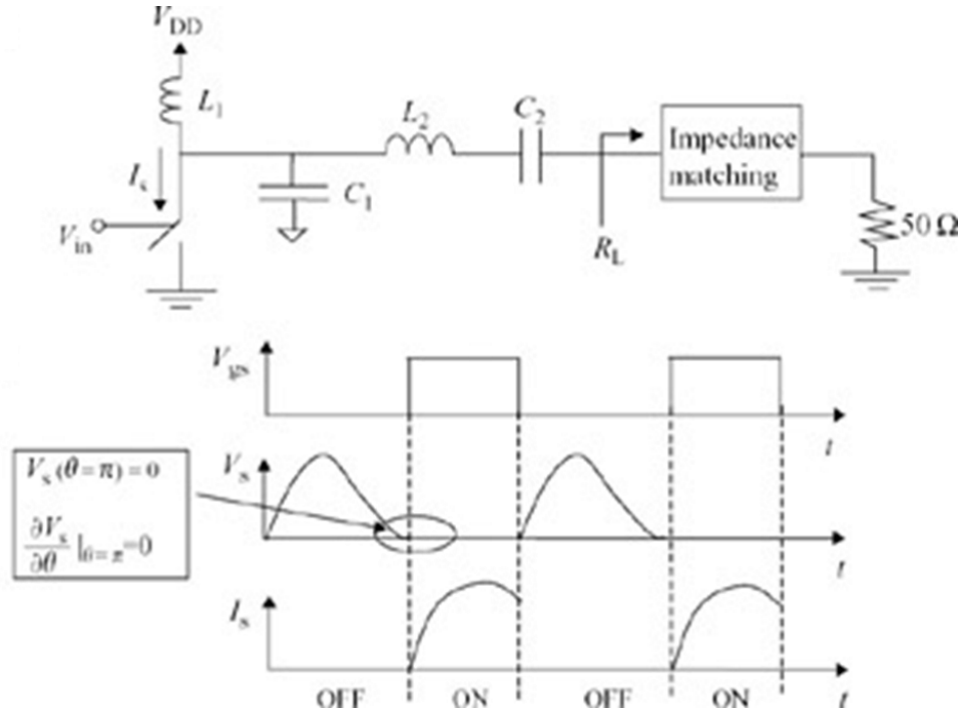


Figure 2.2: Typical schematic and waveform of a class-E power amplifier [33].

2.2 GaN Amplifier Technologies

GaN power amplifiers have been studied for RF applications. In the power amplifier design such as class A, AB, B, and C, the operated power and efficiency are defined by conduction angle and bias voltage. These amplifier design techniques are based on the recent every power amplifier. A lot of power amplifier based on these class are still developed and reported for wireless application such as LTE communication [27]-[29]. Vetury et al. reported a high-power and high-efficiency AB-class GaN amplifier for wireless base station applications with 100 W output power and 55.3 % PAE at 2.14 GHz [27]. Also, Iqbal et al. reported an AB-class GaN amplifier for LTE communication band with 5 W output power and 68 % PAE at 2.3 GHz [29], which is close to capacity of the efficiency for AB-class GaN amplifier.

The high-efficiency techniques for the power amplifier are drastically developed by switching and harmonic tuning techniques. The possibility of improving efficiency with vacuum tubes have discussed and firstly demonstrated since 1910s. Generally, the overlap of the voltage and current waveform at the output terminals produces a loss. If the overlap was completely reduced, the loss of the tube must be zero. In the high efficiency amplifier design, it is also necessary to reduce the

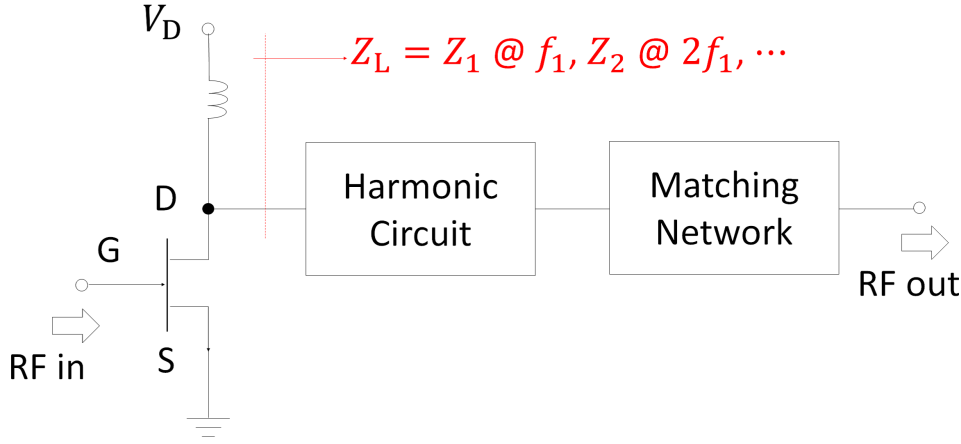


Figure 2.3: Typical schematic of a power amplifier with harmonic tuning circuit.

overlap of the output current and voltage waveform. In a switching technique, the FET with a shunt capacitor and a series filter are operated by input signal of rectangular pulse for turning on and off of the FET, which makes the current and voltage waveform switch in time domain [32]. The switching technique is also known as class-E mode. The typical schematic and waveform of a class-E is shown in Fig. 2.2 [33]. The early history of switching techniques for power amplifier was summarized in [31]. Ui et al. reported a 100 W class-E GaN amplifier with 75 % drain efficiency at 2 GHz in 2006 [32].

The harmonic tuning techniques are also developed for high efficiency amplifier design. A typical load side schematic of power amplifier with harmonic tuning circuit is shown in Fig. 2.3. In the harmonic tuning techniques, the harmonic impedances of FET are treated in frequency domain. Therefore, the load impedance Z_L is operated at different values for each harmonic frequency, which is provided by a harmonic circuit. If the suitable values was provided for the load impedance at every harmonic frequency, the loss of the amplifier at harmonic frequency ideally become zero, which is contribute the high efficiency amplifier design. The suitable impedances are generally measured by load-pull system or calculated with non-linear FET model. Kim et al. reported a high efficiency GaN amplifier based on harmonic tuning techniques with 75.1 % PAE at 2.655 GHz in 2011 [34]. The amplifier provided with zero or infinite value for load impedance at each harmonic frequency is particularly defined as class-F or inverse class-F mode. A lot of class-F and inverse class-F amplifiers are recently reported as well as the class-E and harmonic tuning amplifier. Schmelzer et al. reported a high efficiency GaN amplifier based on class F technique with 85 % PAE at 2 GHz in 2007 [35].

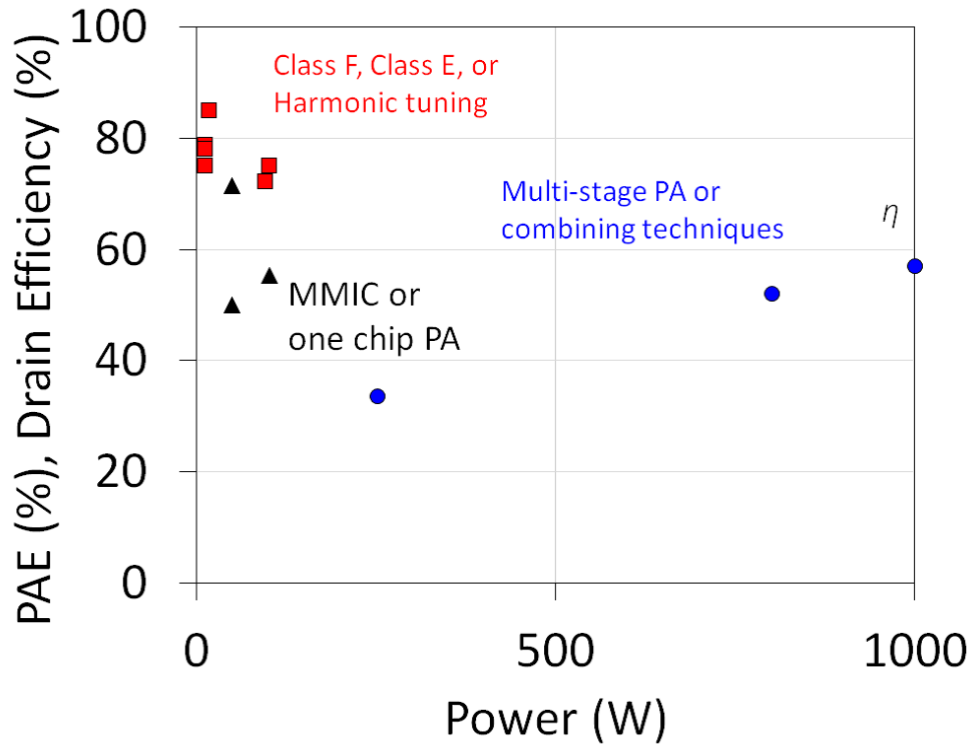


Figure 2.4: Comparison of S-band amplifiers with the state-of-the-art [38]-[47].

The combiner technique is useful for high power amplifier design. The one chip FET device has a limit of operation power. In this techniques, the output signal from single-stage amplifiers with one chip FET is combined. Shigematsu et al. reported a C-band GaN amplifier with over 300 W output power for radar application in 2008 [36]. In him report, GaN HEMTs are combined by a combiner designed on high dielectric substrate with isolation resistors. The high power amplifier with over kW-class output for space application is reported by Noji et al. in 2011 [37]. In him report, the high power GaN amplifiers are assembled by circular waveguide power combiner, which contributes the high power microwave combining.

In this section, comparisons of GaN amplifiers with the state-of-the-art are firstly discussed. Secondly, a design techniques of typical microwave amplifier is explained. Next, a basic AB-class GaN power amplifier is fabricated. In this fabrication, the output characteristics and thermal effect of GaN amplifier by fixture material mounted on GaN HEMT are discussed.

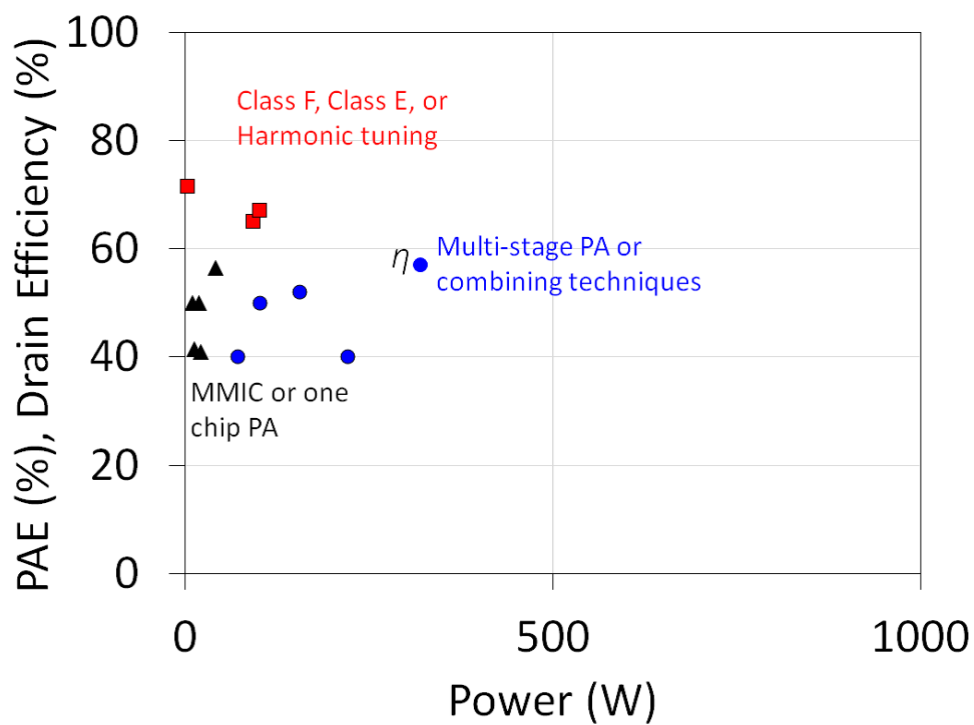


Figure 2.5: Comparison of C-band amplifiers with the state-of-the-art [48]-[59].

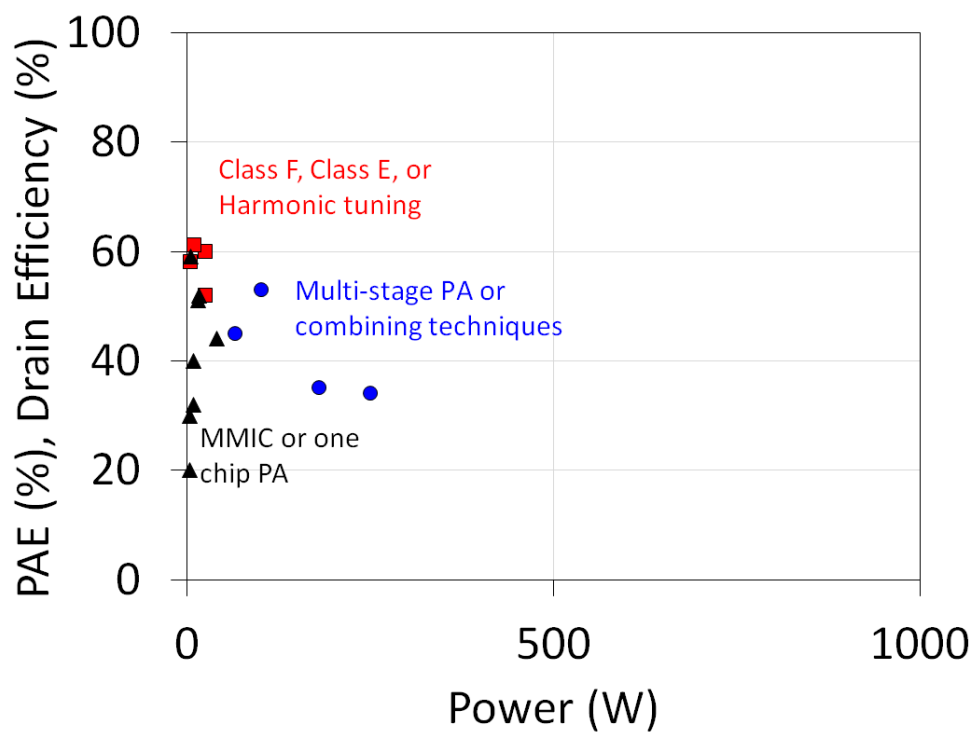


Figure 2.6: Comparison of X-band amplifiers with the state-of-the-art [60]-[76].

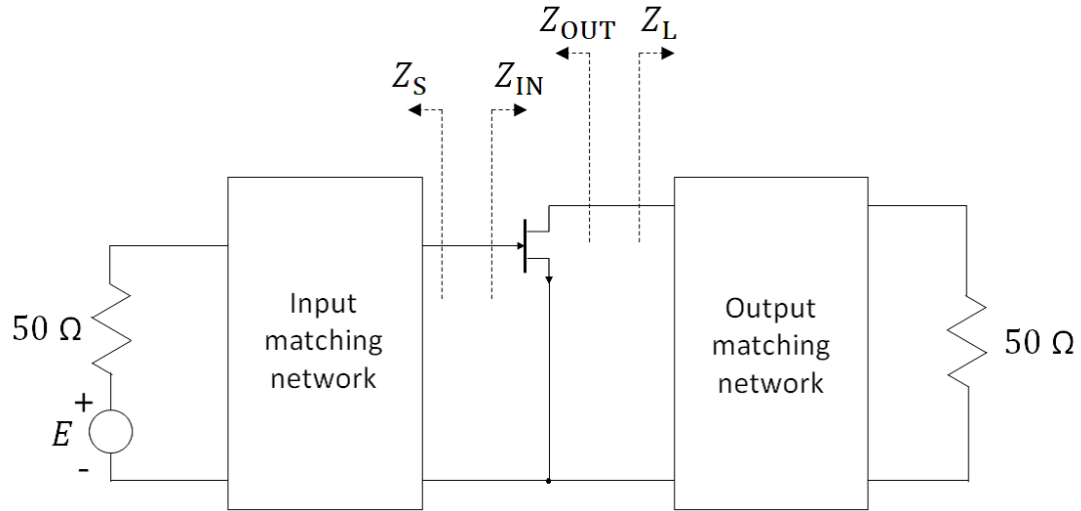


Figure 2.7: Microwave amplifier diagram for RF system.

Recently, a lot of high-power and high-efficiency GaN amplifiers are reported. The comparison of GaN amplifiers with the state-of-the-art at S-band to X-band frequency are concluded in Fig. 2.4, Fig. 2.5, and Fig. 2.6, respectively [38]-[76]. In Fig. 2.4, the recently reported efficiency properties of output power at S-band GaN amplifiers are summarized. In the S-band frequency, the high efficiency and high power GaN amplifiers with more than 80 % and 1 kW were respectively reported. The improvement of amplifier efficiency is realized by a harmonic-tuning, Class-F, and Class-E design techniques which provide the more than 10 points efficiency increase. In the C-band and X-band frequency, the efficiency is reduced to more than 70 % and 60 %, respectively. In particular, the superiority of the each high-efficiency technique disappears in X-band frequency because it is difficult to treat the harmonics in high frequency condition. On the other hand, it is confirmed the output power of the one chip GaN amplifiers decreases at comparing the amplifier specs in S- to X-band frequency condition, which affects the output power of GaN amplifier based on combiner techniques.

2.3 Microwave Amplifier Design

The basic microwave amplifier diagram is shown in Fig. 2.7. In the power amplifier design, the small-signal S parameters are not used because of the non-linear properties of FETs for input RF power. Therefore, the amplifier circuits of input and output matching networks are designed for input and output impedances,

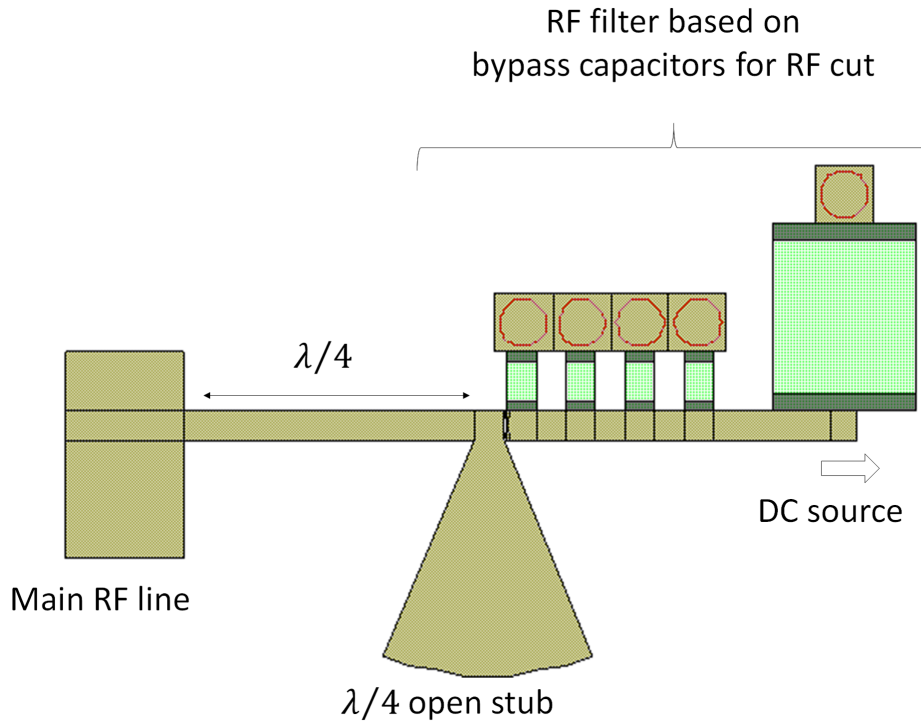


Figure 2.8: Circuit configuration of bias line.

Z_{IN} , and Z_{OUT} , of FET at large-signal condition. In general, the RF circuits are designed for $50\ \Omega$ impedances because the most RF applications use $50\ \Omega$ system. Therefore, the RF amplifier circuits such as input and output circuit as shown in Fig. 2.7 have to design for $50\ \Omega$ load.

These impedances are generally calculated by using non-linear FET model and load-pull simulation. In the circuit design used semiconductor devices sometime, SPICE simulators have been used. However, the SPICE simulators are not suitable to simulate the power saturation properties of FETs in the large-signal condition because the simulators are assumed to calculate in the linear condition. The non-linear model are design by using a harmonic balance algorithm, which make it possible to repeat the saturation properties of FET in a simulation. Recently, the most of non-linear models are provided by semiconductor-device manufacturers, which contribute to save the power amplifier design based on large-signal condition.

The input and output impedances of a FET non-linear model are calculated by load-pull simulation. In the load-pull simulation, the load and source impedances, Z_L , and Z_S , are calculated by using impedance tuners for obtaining a maximum output power or efficiency. The amplifier circuit are designed for the obtained

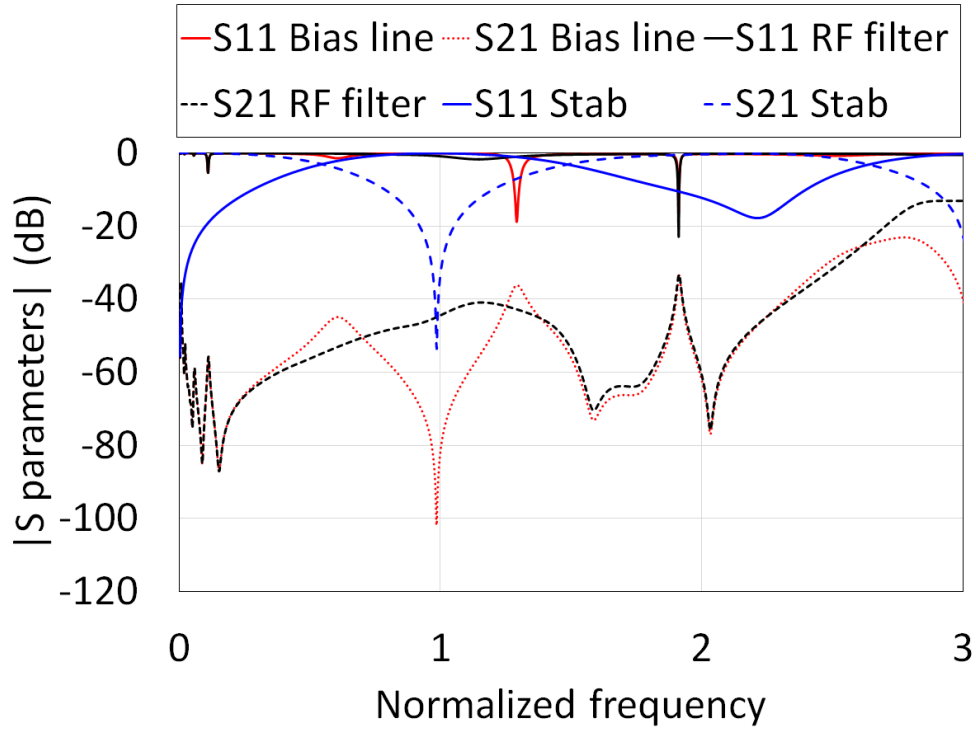


Figure 2.9: S-parameters of bias line.

impedances by the load-pull simulation. Therefore, the passable RF circuit design theory are available for the amplifier circuits design based on calculated impedances by load-pull simulation.

Noise from DC source is a cause of the undesired excitation. In addition, the leaky wave from a RF line of amplifier circuit to DC circuit system possibly prevents the correct measurement in DC system. In the amplifier circuits design, the bias line is necessary for stable amplifier operation. In general, the bias line is designed for isolating RF circuit system to DC circuit system. Therefore, the most bias lines are designed by using filter techniques. The circuit configuration of bias line employed in this work is shown in Fig. 2.8. The bias line consists of $\lambda/4$ line for a fundamental frequency with a $\lambda/4$ open stub and RF filter based on capacitors. In the $\lambda/4$ line with the stub, the RF signal at fundamental frequency is separated to DC source. In addition, the RF filter based on capacitors blocks the noise for wide band frequency. Figure 2.9 show the bias line S parameters characteristics of normalized frequency by a fundamental frequency. According to Fig. 2.9, the S_{11} of $\lambda/4$ line with the stub is adequately reduced at fundamental frequency. The S_{21} of RF filter is widely reduced. The S_{21} of bias line consisted

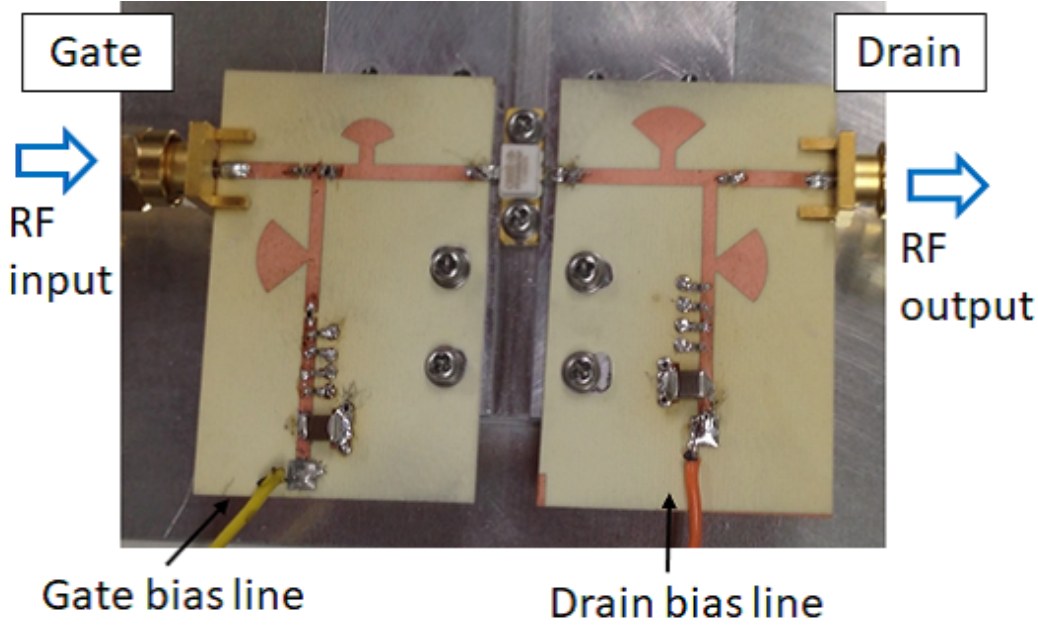


Figure 2.10: Fabricated GaN amplifier.

of $\lambda/4$ line with the stub and RF filter is also widely reduced. In particular, the S_{21} of bias line is adequately reduced at fundamental frequency.

2.4 Basic AB-Class Power Amplifier Fabrication

A 20 W-class and C-band amplifier was designed with RF circuit simulator (ADS, Agilent) and non-linear model of GaN HEMT (CGH40025F, Cree). In general, solid-state elements including GaN HEMT have a non-linear characteristics with input signal power. Therefore, ample studies of the small and large signal characteristics is required for the precise design of amplifier. In the load-pull simulation, $Z_L = 12.2 + j18.0$ and $Z_S = 7.4 - j62.9$ for max PAE of 58.31 % and 45.3 dBm P_{out} are obtained. These impedances are targets for amplifier circuits design. First of all, Gate and Drain bias lines and matching network based on radial stub circuits was designed by RF circuit simulator (Keysigt ADS). The substrate of ROGERS RO4350 (Copper thickness $T=0.018$ mm, substrate thickness $H=0.762$ mm, dielectric constant $\epsilon_r = 3.66$, dielectric tangent $\tan \delta = 0.0034$) was used. In this simulation, the DC gate voltage $V_g = -2.8$ V and the DC drain voltage $V_d = 28$ V. These bias voltages were fed through the bias lines. In this bias line, the bypass capacitors were implemented for noise filtering from DC power sources. In the

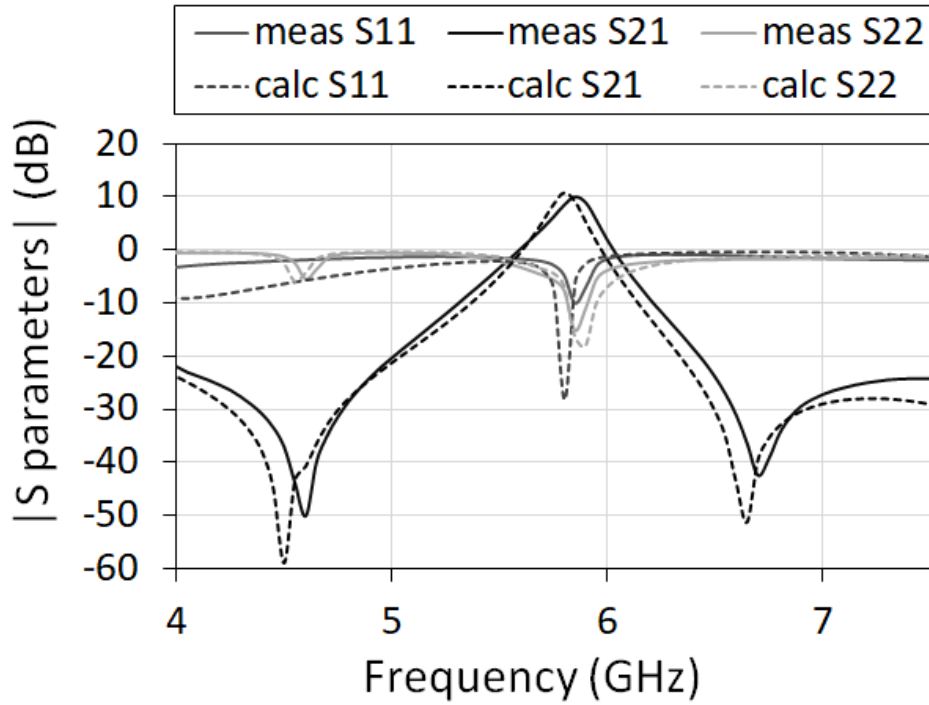


Figure 2.11: Measured amplifier S parameters.

next, the matching network was tuned in the large-signal condition and optimized for input impedance of the GaN HEMT element.

The fabricated 20-W C-band GaN amplifier is shown in Fig. 2.10. The small-signal S parameter and large-signal input-output characteristics of fabricated amplifier was measured at the same DC bias condition as simulation. The measured S parameter of amplifier as shown in Fig. 2.11 is compared to ADS simulation result $|S_{21}|$. According to Fig. 2.11, the measured $|S_{21}|$ agreed with ADS simulation result. The large-signal input-output characteristics of measurement and simulation results are shown in Fig. 2.12. The setting frequency of input signal was 5.95 GHz. According to Fig. 2.12, the measured maximum output power and power added efficiency (PAE) were 43.1 dBm and 44.5 % at input RF power of 31.5 dBm and 43.1 dBm, respectively. The maximum gain was 10.3 dB at input RF power of 1.37 dBm. However, the gain was compressed above the input RF power of 25 dBm. Comparing the measured results with simulation, the maximum PAE decreased from the simulated maximum PAE of 54.5 % because of fabrication and measurement error.

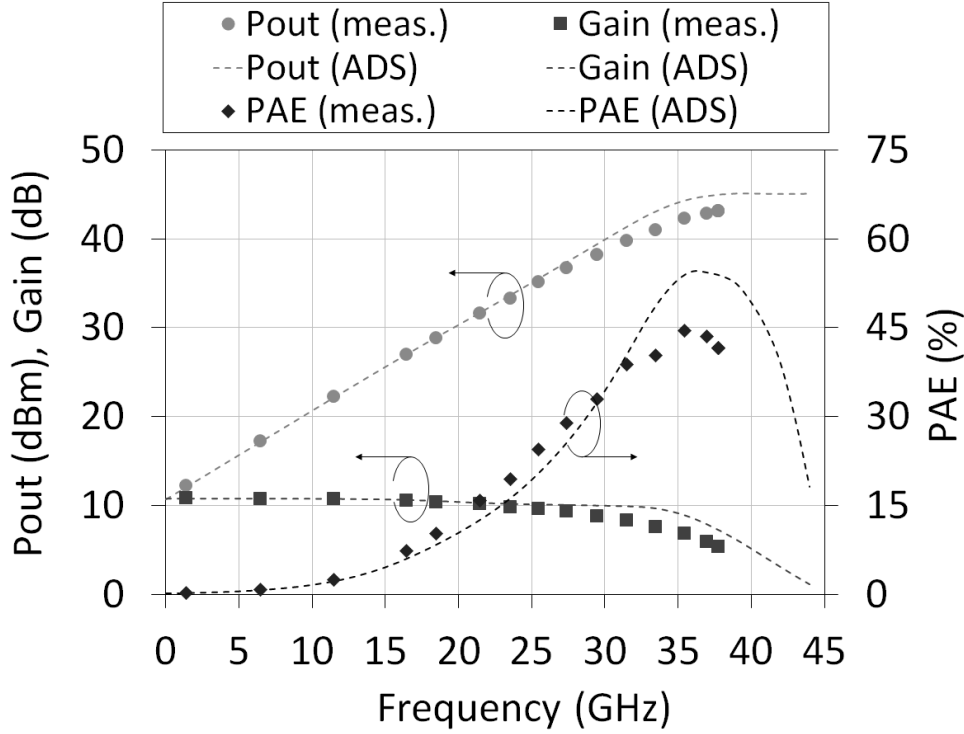


Figure 2.12: Measured input output characteristics.

2.5 Lightweight Carbon Graphite Fixture for GaN Amplifier

The weight reduction of microwave transmission systems are required. The high-power active-circuit units such as amplifier uses a copper fixture. Generally, copper has property of high thermal conductivity, but the specific weight is large. On the other hand, the carbon graphite has the properties of good thermal conductivity and lightweight. It is necessary for use of amplifier with large power consumption to install a cooling system. In our study, the peltier device was used as the cooling system. In addition, the fixture with high thermal conductivity which fix the amplifier on peltier device is necessary for sufficient cooling. The conventional fixture for amplifier have used aluminum or copper. The thermal conductivity and specific weight of aluminum, copper, and carbon graphite are shown in TABLE 2.1. According to TABLE 2.1, copper has the 1.7 times thermal conductivity as large as aluminum, which mean the copper is more suitable for the fixture than aluminum. However, the specific weight was 3.3 times as large as aluminum. The carbon graphite has 7.2 times and 4.2 times thermal conductivity as large as aluminum

and copper. In addition, the specific weight of carbon graphite is 0.8 times and 0.25 times as low as aluminum and copper. Therefore, the use of carbon graphite fixture provides benefits for weight and thermal property of amplifier.

The 60 mm (length) $\times$ 40 mm (width) $\times$ 10 mm (thickness) size fixture consisted of pure copper or carbon graphite combined with copper shown in Fig. 2.13(a) and Fig. 2.13(b) was fabricated. Fig. 2.13(c) show the image of amplifier with the carbon graphite fixture on the peltier device. The GaN HEMT was mounted on the fixture and subjected to cooling with peltier device. The weight of the amplifier including circuit and fixture is shown in the TABLE 2.1. According to the TABLE 2.1, the weight of amplifier with carbon graphite fixture was reduced by 64.2 g and the reduction rate at comparing with copper fixture was 71 %.

The C band 30 W class amplifier was designed with RF circuit simulator (ADS, Agilent) and non-linear model of GaN HEMT (CGH40025F, Cree). In general, solid-state elements including GaN HEMT have a non-linear characteristics with input signal power. Therefore, the high accuracy design under small and large signal conditions is required.

First of all, Gate and Drain bias lines and matching network was designed in the small-signal condition with ADS. The substrate of ROGERS RT6035HTC (Copper thickness: 18 μ m, substrate thickness: 0.8 mm, permittivity: 3.66, dielectric tangent: 0.0014) was used. In this simulation, the DC gate voltage $V_g = -2.8$ V and the DC drain voltage $V_d = 28$ V. These bias voltages were fed through the bias lines. In general, bias line should be isolated from RF circuit. Therefore, the designed bias line should have infinite input impedance.

The fabricated GaN amplifier was shown in the Fig. 2.14. The radial stubs on the main line and bias line realized impedance matching to GaN HEMT and isolation of the bias line, respectively. In this bias line, the bypass capacitors were implemented for noise filtering from DC power sources.

Table 2.1: Comparison of the thermal conductivity and weight among the aluminum, copper, and carbon graphite.

	Aluminum (Al)	Copper (Cu)	Carbon graphite ^a
Thermal conductivity	236 W/m/K	403 W/m/K	1500 W/m/K ^b
Specific weight	2.72	8.96	2.20 ^c
Weight of Amplifier ^d	-	221.5 g	64.2 g

^a combined material with copper shown in Fig. 2.13

^b value of thickness direction (value of pure carbon graphite is 1700 W/m/K.)

^c value of pure carbon graphite.

^d including the circuit and fixture weight

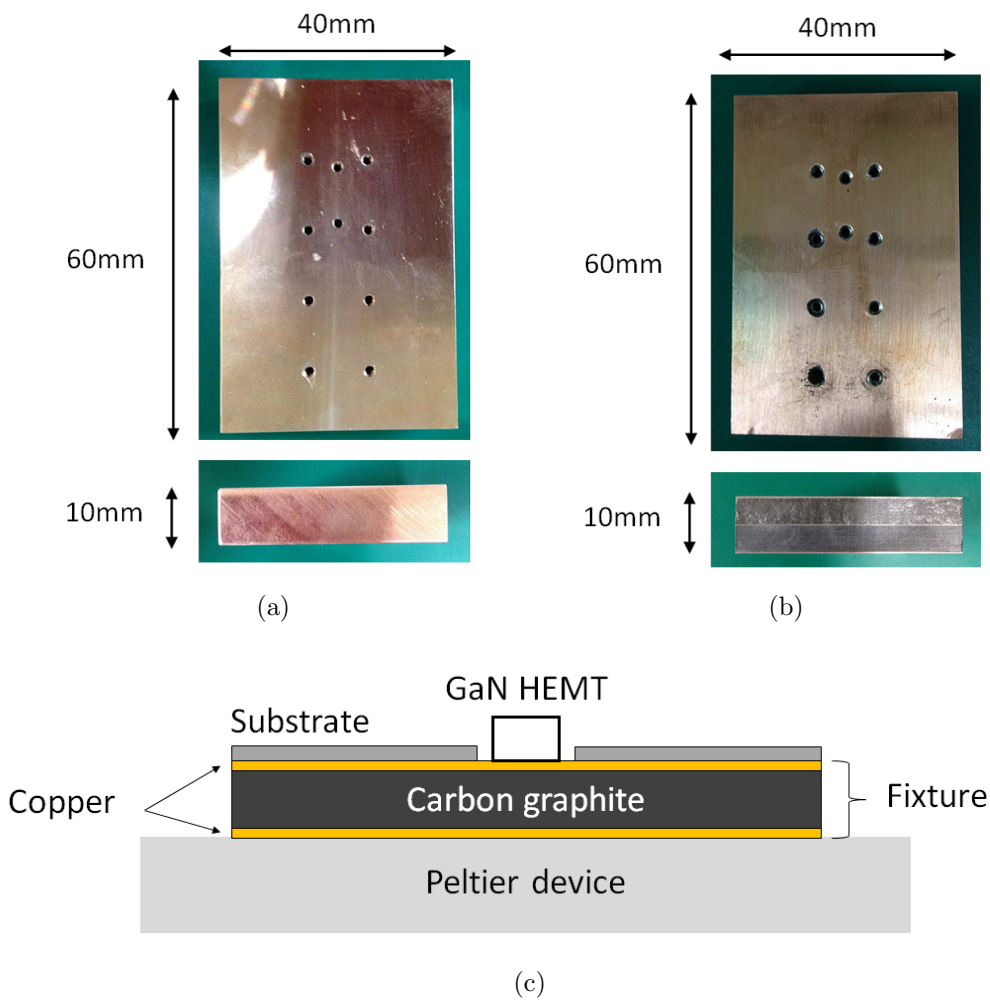


Figure 2.13: (a) pure copper, (b) carbon graphite combined with copper, (c) image of GaN amplifier with the carbon graphite fixture on the Peltier device.

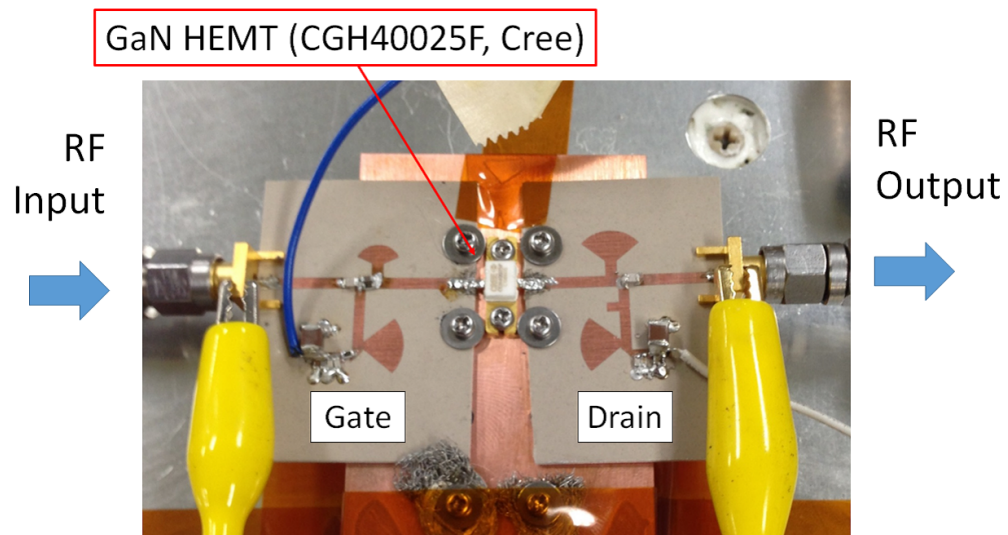


Figure 2.14: Fabricated GaN amplifier mounted on carbon graphite fixture.

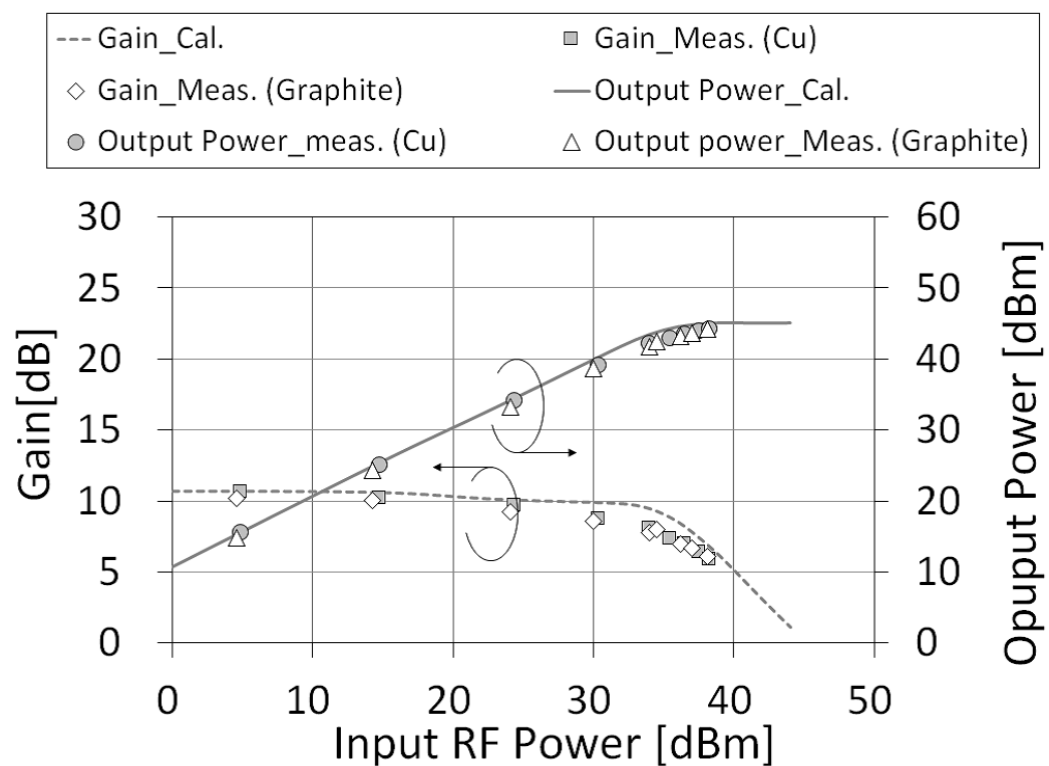


Figure 2.15: Calculated and measured large signal output power and gain.

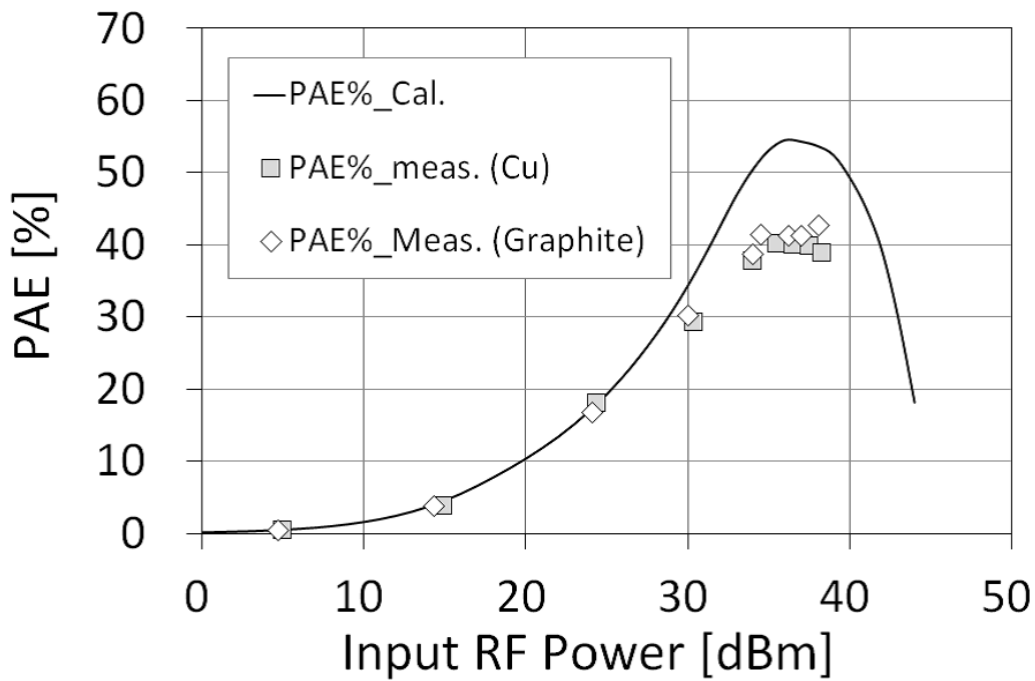


Figure 2.16: Calculated and measured PAE.

2.5.1 Measurement of RF properties and thermal conditions

The large-signal input-output characteristics of measurement and simulation results are shown in Fig. 2.15. The setting frequency of input signal was 5.8 GHz. The fabricated copier and carbon graphite fixture was used. Comparing to the results of use of copper and carbon graphite fixture, measured gain and output power corresponded with each other and ADS simulation results. According to Fig. 2.15, the measured maximum output power was 44.2 dBm (26.3 W) at use of copier and carbon graphite fixtures at 38.7 dBm RF input. The gain at 38.7 dBm RF input also was 6.1 dB and gain compression from small signal condition were 4.3 dB and 4.5 dB at use of copper and carbon graphite fixture, respectively. The measurement and simulated PAE are shown in Fig 2.16. According to Fig 2.16, the measured maximum PAEs were respectively 40.1 % and 41.4 % at used copier and carbon graphite fixture, but the maximum PAE of simulation was 54.5 %. The cause of PAE difference between measurement and simulation was considered as fabrication error or thermal property.

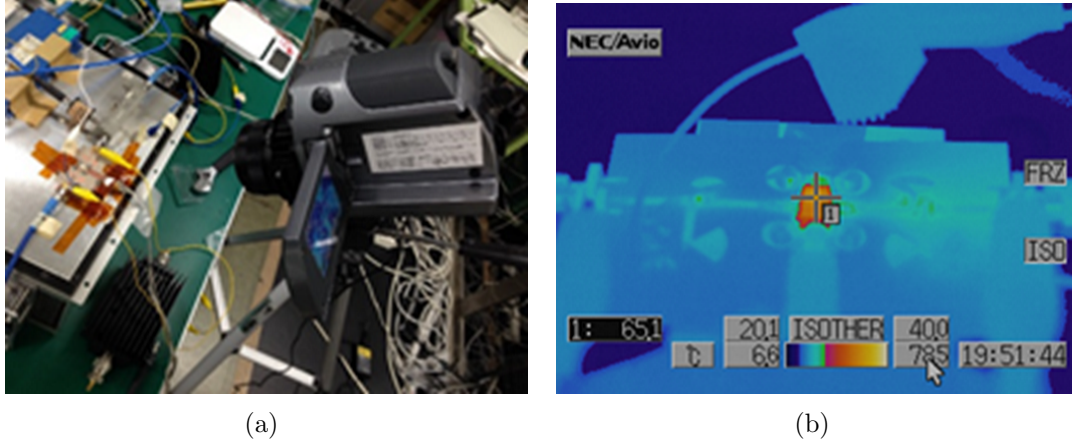


Figure 2.17: Infrared thermal camera for temperature measurement: (a) photograph of the camera system and (b) thermal image of surface temperature on GaN amplifier.

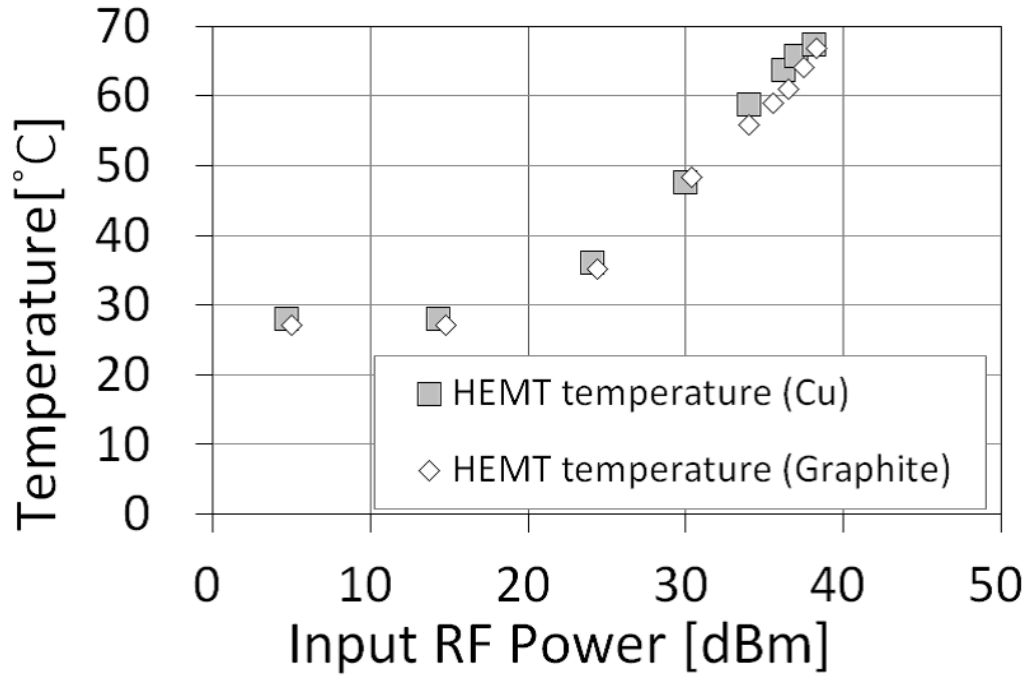


Figure 2.18: Measured surface temperature of HEMT chip.

The surface temperature of HEMT chip was measured with infrared thermal camera as shown in Fig. 2.17. The HEMT chip was subjected to cooling by 15°C peltier device through the fixture. The measured temperature as a function of input RF power was shown in Fig.2.18. According to Fig.2.18, the chip temperature increased to 67.3°C and 66.8°C at 38 dB input RF power and use of copper and

carbon graphite fixture, respectively. The simulation result of Fig 2.16 was calculated under thermal condition of 25°C HEMT temperature. In the simulation, it confirmed the maximum PAE decreased to 50.1 % under thermal condition of 66.8°C HEMT temperature. The calculated difference of maximum PAE between 25°C and 66.8°C was 4.4% .

2.6 Summary

The recent RF antenna and GaN amplifier technologies are firstly introduced. In particular, the configurations of RF circuit for aperture antenna and array antenna system are illustrated. In addition, the basic configurations of GaN amplifier and the comparison of S- to X-band GaN amplifiers with the state-of-the-art are illustrated. Next, an AB-class power amplifier is designed and fabricated for illustrating a fundamental design method of GaN amplifier in this work. In this amplifier fabrication, the 43.1 dBm output power, 44.5 % PAE, and 10.3 dB small-signal gain are obtained. Finally, the carbon graphite fixture for light weight amplifier is fabricated, which contributes to reduce 71 point weight of fixture at comparing copper one.

Chapter 3

High Power GaN Amplifier Based on Combiner Techniques

3.1 Introduction

High power microwave transmission based on aperture antenna system requires the one port high power GaN amplifier. The solid-state power amplifiers (SSPAs) have been recently used for the microwave transmission such as MPT, space applications, and radar. In general, SSPA has advantages such as longer operating life and greater reliability compared with conventionally used vacuum-tube devices such as TWTA, magnetron, and klystron. In recent research, high-power SSPA that uses GaN technology for applications such as satellite up-/down-link communication and wireless power transmission based on CW operation have been reported. Kobayashi et al. demonstrated CW driven 1-kW class S-band SSPA with 8-way amplifiers with GaN HEMT chips for replacing conventional vacuum-tube devices; the demonstrated SSPA has specifications comparable to the vacuum-tube devices [77]. In recent deep space exploration missions, S- to X-band frequencies were used for satellite communication [78], [79]. In particular, a 7.1 GHz band up-link communication was used for a recent mission –proximate object close flyby with optical navigation (PROCYON)–launched in 2014 [78]. The 7.1 GHz oscillation in the space ground base station is based on vacuum-tube devices that need to be replaced with GaN-based SSPA in the future.

Deep space communication often requires CW-drive operation [80]. In addition, SSPA is expected to be used for wireless power transmission based on CW

microwave [81], [82]. In general, the temperature rise in HEMT chips under the CW-drive condition affects the performance of SSPA. However, the design of SSPA is often assumed to only drive on the pulse condition. Therefore, it is necessary to verify the SSPA performance under the CW-drive condition for high power microwave transmission.

The GaN-based high-power amplifier is promising for SSPA for high power microwave transmission. GaN HEMT has the characteristics of wide band gap and high saturation electron velocity, which make it possible to drive it with high power and high efficiency. Several high-power amplifiers with GaN HEMT have been recently proposed in the S- to X-band frequencies [83]-[91]. Shigematsu et al. reported the over 300 W and 100 W GaN amplifier in the C-band (4.8 GHz) and the X-band (9.8 GHz) for replacing TWTA in 2009 [92]. In their research, the amplifiers were driven under the pulsed condition. This is due to the temperature rise caused by low efficiency of PAs. Therefore, higher efficiency PAs are required for CW operation.

In conventional studies, GaAs-based amplifiers are combined by N-way spatial combiners for improving output power [93]-[95]. In the GaN-based amplifiers, the same output power level as GaAs N-way amplifiers can be obtained for using a few way planer combiners on a substrate [96]-[98]. However, the output power per GaN chip decreases at higher frequencies than the C-band. Hence, low loss Combiner technique is needed for obtaining high-power SSPAs.

In this work, the 7.1 GHz high power amplifier for high power microwave transmission such as the up-link satellite communication and the MPT are demonstrated. The amplifiers with GaN HEMT chips are designed and fabricated for showing the performance of amplifier unit in SSPA. The SSPA are assembled with a 20-way radial combiner and divider. The output of the SSPA assumes a rectangular waveguide. This work targets to realize a prototype SSPA with more than 100 W output under the CW-drive condition at 7.1 GHz for high power microwave transmission by using low loss Combiner technique and higher efficiency GaN power amplifiers.

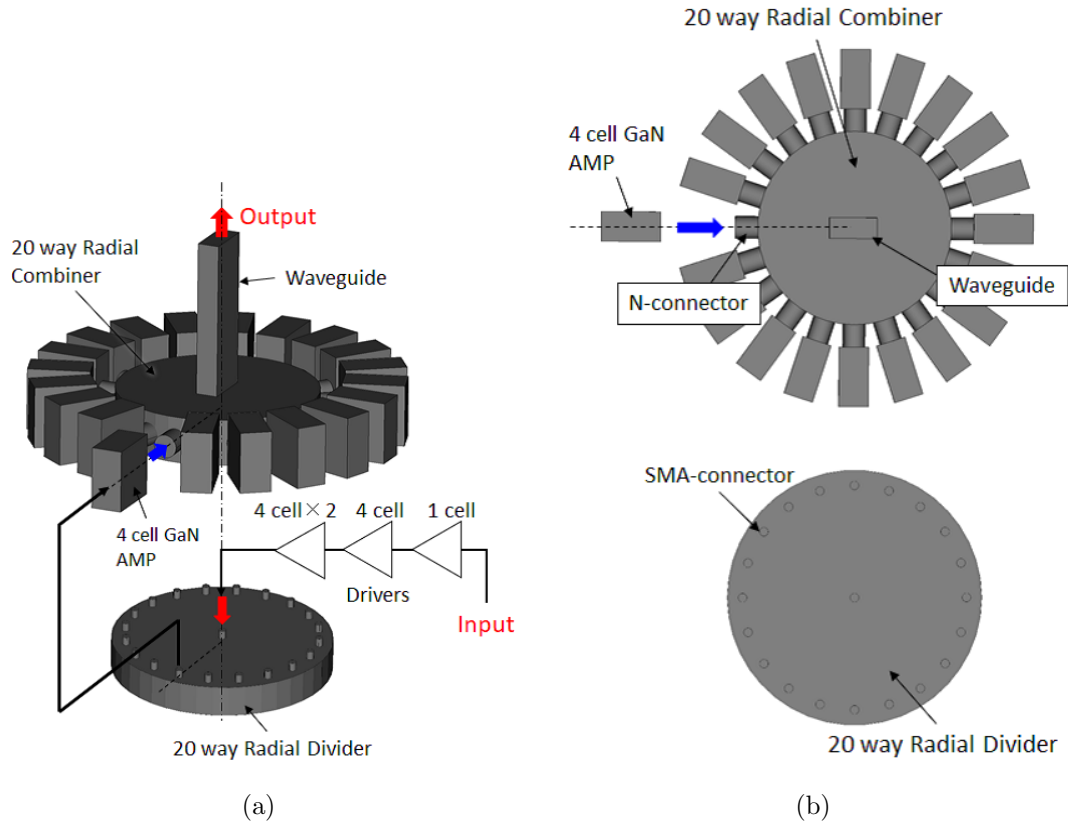


Figure 3.1: Overview of high power amplifier: (a) 20 way GaN amplifier and (b) top view of combiner and divider.

3.2 Overview of High Power Amplifier

The 7.1 GHz SSPA for high power microwave transmission are demonstrated by 20 way combiner approach in this work. Power combiner and divider are one of the key technology for high power SSPS. In the combiner and divider design, the all ports are sufficiently matched. Wilkinson and Gysel power divider/combiner are generally used in planar circuit. These dividers/combiners are used in 2 way dividing and combining. It is unsuited to assemble for large-scale and high-power SSPA with a lot of amplifiers. Therefore, in this work, the radial combiner and divider based on wave guide are employed.

The SSPA is constructed by 20 way radial combiner/divider as shown in Fig. 3.1. The designed GaN amplifiers are assembled by the combiner and divider as shown in Fig. 3.1(a), which are driven by 3 steps GaN driver amplifiers. Figure 3.1(b) shows the top view of combiner and divider. The 4 cell GaN amplifiers are attached by N- or SMA-connectors to combiner or divider. In the output of the SSPA, a rectangular waveguide is used.

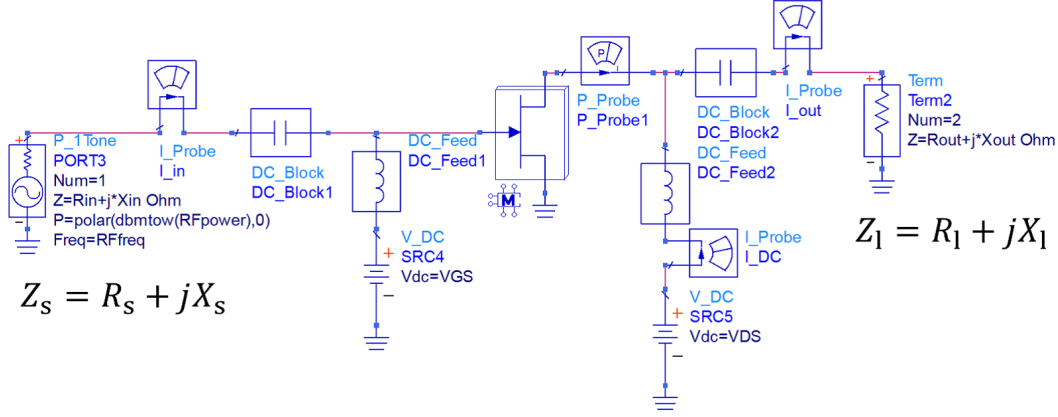


Figure 3.2: Source and load impedances calculation on ADS simulator

Table 3.1: Chip size and calculated impedance of GaN HEMTs.

Chip Size	Z_s	Z_l^*	P_{out}	PAE
1 cell	$25.8 + j5.4$	$18.5 - j33.3$	37.3 dBm	61.1 %
4 cells	$6.5 + j1.4$	$4.6 - j8.3$	43.5 dBm	61.4 %

3.3 GaN Amplifier Design

3.3.1 GaN HEMT Chips

The TriQuint TGF2023 series are selected for GaN HEMT chips in this work. The TGF2023 series are constructed by combining basic unit cells TGF2023-1. According to data sheets for the TGF2023 series, the drain voltage of 12–32 V and drain quiescent current of 62.5 mA are recommended for operating DC bias conditions. Further, the frequency range is between DC to 18 GHz.

Jeong and Yeom presented a X-band 40 W pulse-driven power amplifier with the TGF2023-10 chip constructed by combining eight cells in 2013 [99]. They researched the source and load impedances by a measurement with load-pull tuners. In this study, the impedances are calculated by optimization with non-linear models on the harmonic balance setup of circuit simulation (ADS, Keysight). The circuit for calculating the source and load impedances is shown in Fig. 3.2. The targets in the optimization are impedances of the one cell (TGF2023-1) and four cell (TGF2023-5) chips. The goal is to determine values exceeding 60 % PAE. The calculated source and conjugated load impedances, Z_s and Z_l^* , are shown in

Table 3.2: Impedance ratio to $|Z_s|$ or $|Z_1^*|$.

Chip Size	50 / $ Z_s $	50 / $ Z_1^* $	30 / $ Z_s $	30 / $ Z_1^* $
1 cell	1.9	1.3	1.1	0.8
4 cells	7.6	5.3	4.5	3.2

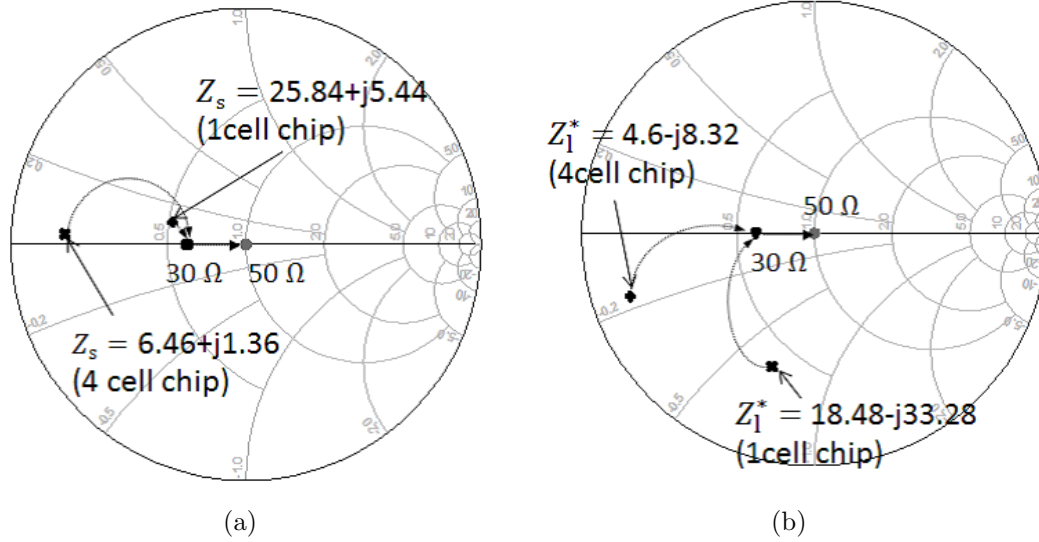


Figure 3.3: Impedance matching flow for 1 cell and 4 cell chips: (a) source impedance matching flow and (b) load impedance matching flow.

Table 3.3: Microstrip line width on ARLON AD1000 substrate

Thickness	Line width of			
	50 Ω	40 Ω	30 Ω	20 Ω
$H = 0.25$ mm	0.22 mm	0.35 mm	0.57 mm	1.03 mm
$H = 0.8$ mm	0.75 mm	1.16 mm	1.89 mm	3.39 mm
$H = 1.6$ mm	1.65 mm	2.52 mm	4.04 mm	7.20 mm

Table 3.1 on 28 V drain and AB-class bias condition. In this condition, the output power P_{out} of the one cell and four cells chips are 37.3 dBm and 43.5 dBm, respectively.

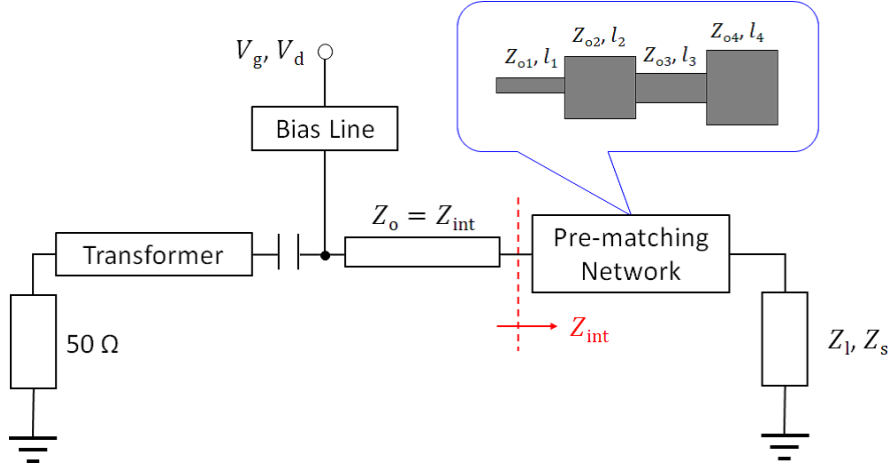


Figure 3.4: Circuit configuration for the amplifiers.

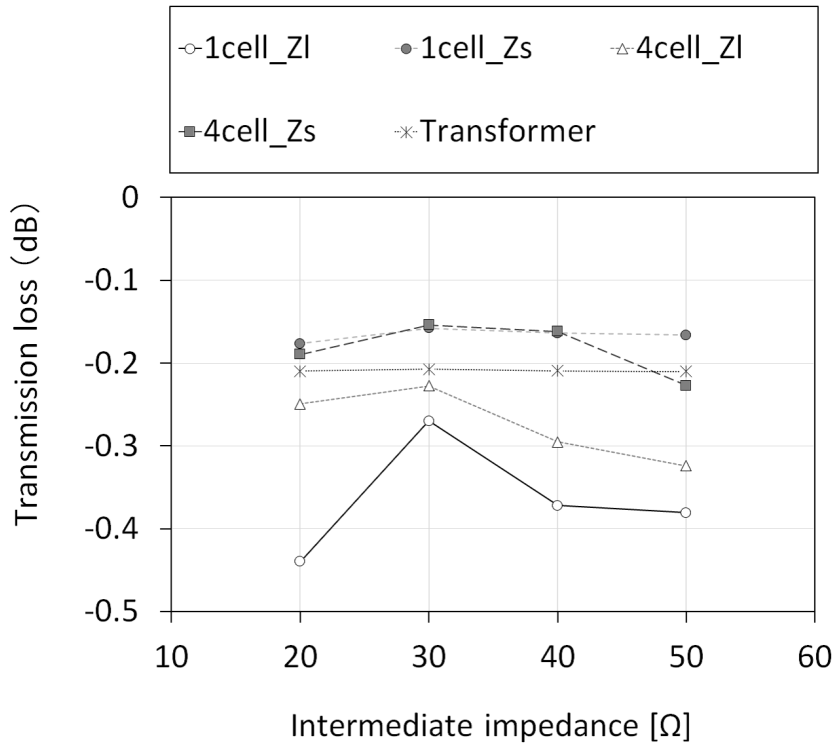


Figure 3.5: Transmission loss of pre-matching network and transformer characteristics for each intermediate impedance

3.3.2 Matching Network Configuration

The matching networks of amplifiers are designed for calculated source and load impedances. In general, RF circuits have to be transformed to 50Ω . However, according to Table 3.1, the calculated impedances have values smaller than 50Ω . A large impedance ratio leads to a narrow band in general, which makes designing

matching networks difficult. The impedance ratio of $|Z_s|$ or $|Z_1^*|$ to $50\ \Omega$ or $30\ \Omega$ are shown in Table 3.2. According to Table 3.2, the ratio decreases if the impedance is transformed to $30\ \Omega$. Therefore, the source and load impedances are transformed to $30\ \Omega$ at the 1st stage matching network in this design, which reduces the burden on the circuit design. In the 2nd stage, the impedance is transformed to $50\ \Omega$ as shown in Fig. 3.3. The circuits of the amplifier are designed with a two-stage matching network. This circuits consist of pre-matching network and transformer as shown in Fig. 3.4 which are connected through a transmission line with a $Z_o = Z_{int}$ of characteristics impedance. The pre-matching network and transformer are designed by a 4 blocks transmission line and a tapered line. It is necessary for circuit design to decide the intermediate impedance, appropriately.

In this design, the substrate of ARLON AD1000 (dielectric constant $\epsilon = 10.2$, dissipation factor $\tan \delta = 0.0023$, and substrate thickness $H = 0.25\ \text{mm}$ at $10\ \text{GHz}$) is used. In this substrate, the width of $50\ \Omega$ and $30\ \Omega$ lines are $0.22\ \text{mm}$ and $0.56\ \text{mm}$, respectively. The value of intermediate impedance is less than $50\ \Omega$ is expected. Assuming the substrate of ARLON AD100, the microstrip line width of $50\ \Omega$, $40\ \Omega$, $30\ \Omega$, and $20\ \Omega$ characteristic impedance are calculated as shown Table 3.3. In general, the line width should keep less than $\lambda/8$ for reducing a radiation loss. Calculating the $\lambda/8$ length for each thickness, the impedances of less than $30\ \Omega$ at $H = 0.8\ \text{mm}$ and less than $40\ \Omega$ at $H = 1.6\ \text{mm}$ have a line width beyond $\lambda/8$ length, respectively, which is undesirable for this amplifier design. Therefore, the $H = 0.25\ \text{mm}$ substrate is chosen for preventing the line width expanding.

Figure 3.5 show the transmission loss of the pre-matching network and the transformer for each intermediate impedance. The transmission losses are calculated by ADS simulator. According to Fig. 3.5, the losses have peak values at $30\ \Omega$. On the other hand, the tapered transformer has few difference for each impedance. Therefore, the $30\ \Omega$ intermediate impedance is appropriate for reducing the circuit loss. According to Fig. 3.5, in addition, the circuit of four cell chip has lower loss than one cell chip in addition, which expect to increase the efficiency of amplifier with four cell chip at comparing to one cell chip.

The amplifier with $30\ \Omega$ intermediate impedance is designed as shown in Fig. 3.6. This circuits consist of pre-matching networks which transform the source and load impedance of GaN HEMT chips to $30\ \Omega$ intermediate impedance and transformers which transform to $50\ \Omega$.

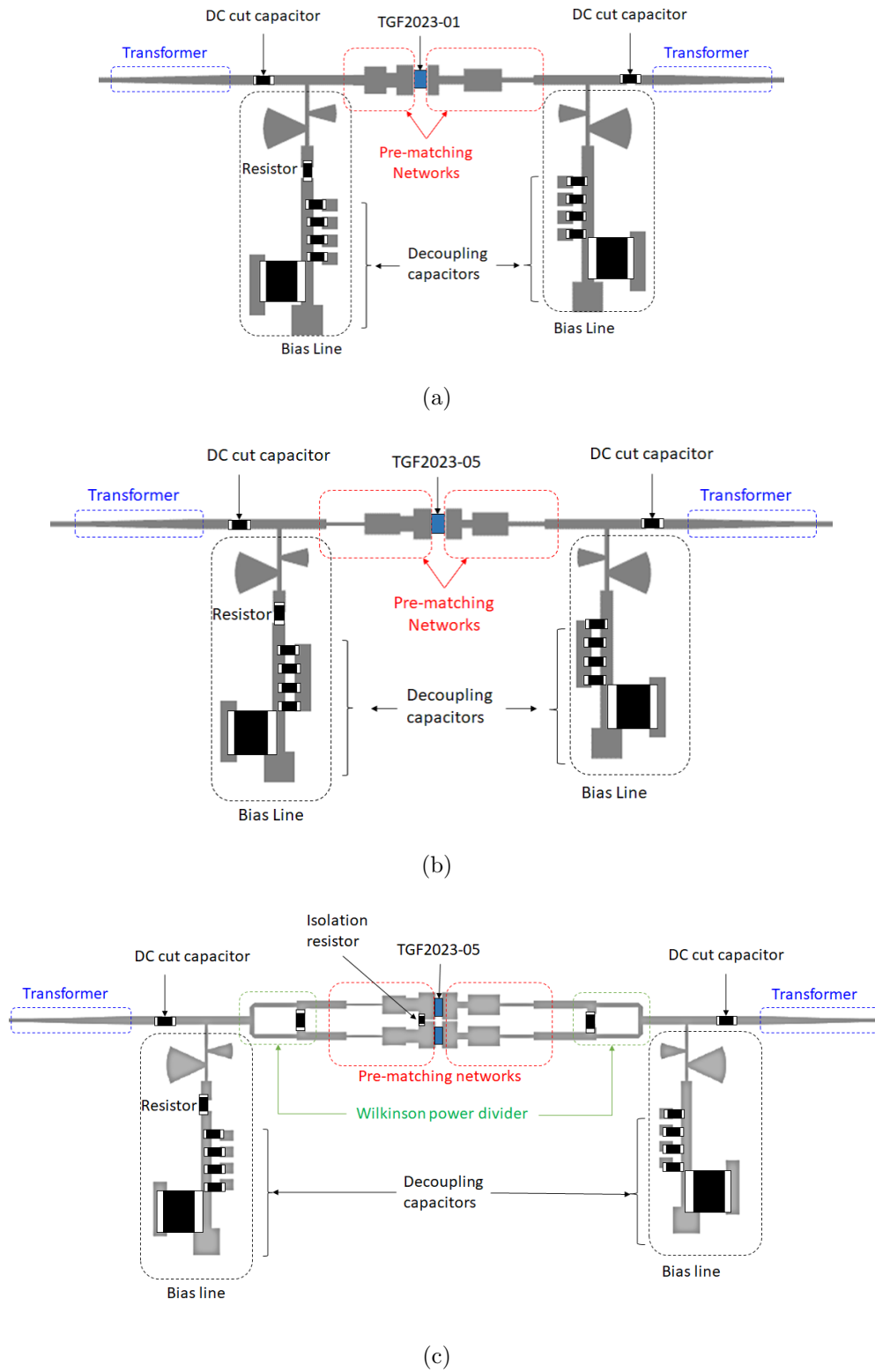


Figure 3.6: Designed amplifier layouts: (a) layout of amplifier with one cell GaN HEMT chip, (b) layout of amplifier with four cell GaN HEMT chip, and (c) layout of 2 way amplifier with four cell GaN HEMT chips and Wilkinson power dividers.

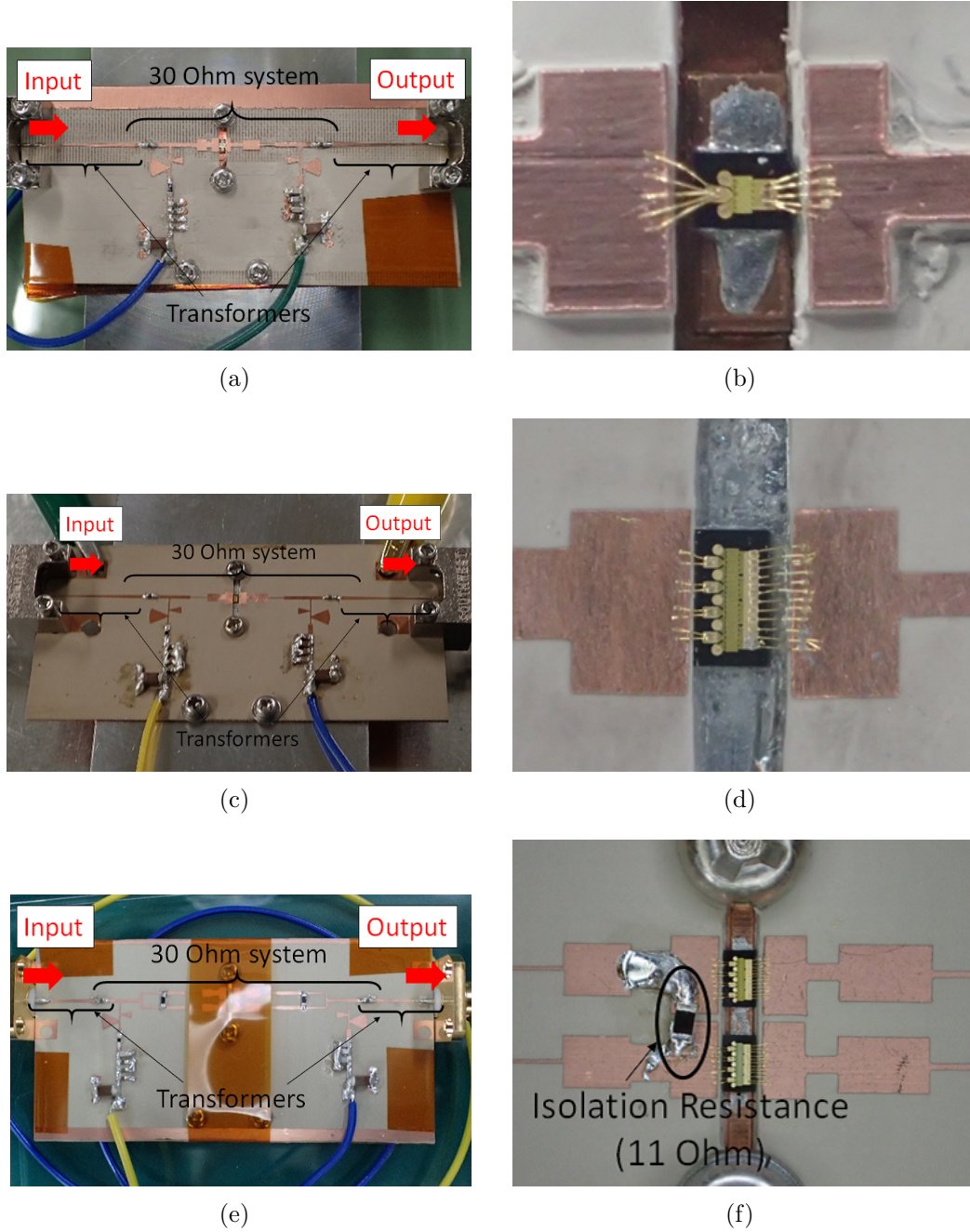


Figure 3.7: Fabricated amplifiers: (a) top view of fabricated amplifier with 1 cell TGF2023-1 chip, (b) mounted TGF2023-1 chip, (c) top view of fabricated amplifier with 4 cell TGF2023-5 chip, (d) mounted TGF2023-5 chip, (e) top view of fabricated amplifier with 2 way 4 cell TGF2023-5 chips, and (b) mounted TGF2023-5 chips.

3.4 Fabrication of Power Amplifiers

The amplifiers with one cell TGF2023-1, four cell TGF2023-5, and two-way TGF2023-5 chips are fabricated as shown in Fig. 3.7(a) to Fig. 3.7(f). The GaN HEMT chips

are mounted on the copper plate using a solder paste. The copper plates have a convex shape where the chips are mounted, which adjust the top face of the chips to the surface of the circuit on the substrate. The circuits are connected to the gate and drain pads on the chips by $18\text{ }\mu\text{m}$ radius gold wires. In the amplifier design with two-way chips as shown in Fig. 3.7(e), the circuit is divided and combined using a Wilkinson power divider and a combiner with $60\text{ }\Omega$ isolation resistor in the $30\text{ }\Omega$ system. The N-way chip configuration on a substrate is considered as the case of excitation because the loops form a coupling between the unbalanced lines explained by chip-to-chip dispersion. Therefore, the $11\text{ }\Omega$ isolation resistor is mounted for preventing the excitation as shown in Fig 3.7(f).

3.4.1 Small-Signal Characteristics

The small-signal S-parameters of fabricated amplifiers as shown in Fig. 3.8(a) to Fig. 3.8(c) are measured by a network analyzer (Keysight, E8361A) under a 28 V drain bias and AB-class condition. The drain quiescent currents are respectively 59 mA, 176 mA, and 489 mA for amplifiers with one cell, four cells, and two-way four cells under this bias condition. In Fig. 3.8(a) to Fig. 3.8(c), the dashed and solid lines show the calculated results by ADS and the measured results, respectively. According to the results, gains of 12.0 dB, 11.7 dB, and 12.7 dB are obtained at 7.1 GHz from the measurement of the amplifier with one cell, four cells, and two-way four cells, respectively. In addition, the bandwidths Δf when the gain compress to -3 dB from maximum $|S_{21}|$ are $\Delta f = 1.98\text{ GHz}$, $\Delta f = 1.24\text{ GHz}$, and $\Delta f = 1.14\text{ GHz}$ for the amplifiers with one cell, four cell, and two-way four cell chips, respectively.

3.4.2 Large Signal Performance

Large-signal performance of the amplifiers is measured. This system consists of a signal generator (MG3690B, Anritsu), a directional coupler (PE2210-20, PASTER-NACK), an isolator (0700EN, NOVA), an attenuator (50HF-030-50/18SMA, JFW), and spectrum analyzers (E4407B and N9020A, Keysight). In addition, the power level of the spectrum analyzers are calibrated using a power meter (E4416A, Keysight). In this configuration, the input loss from the spectrum analyzer to

the reference plane of the device under test (DUT) including the directional coupler, isolator, and coaxial cable, and the output loss including the attenuator and cable from the DUT are 31.5 dB and 31.1 dB, respectively. The DUTs are set on a peltier unit (FC-2414, SCNT) for cooling. PW18-1.8AQ (TEXIO) and PD36-20AD (KENWOOD) are used as the DC supply for the gate and drain bias. The drain current is monitored using a digital multimeter (34461A, Keysight). The signal generator MG3690B is able to drive a 28 dBm maximum output. However, it is necessary to input over 30 dBm to amplifier with four cell and two-way four cell chips for measuring output saturation. When measuring the amplifier with one cell chip, the system does not include a driver amplifier. On the other hand, the system includes the driver amplifier when measuring amplifiers with four cells or two-way four cell chips.

The measured large-signal performances are shown in Fig 3.9. The measured results are compared to simulation results. Figure 3.9(a) shows the graph of the output power versus input power. According to Fig. 3.9(a), the output power of the amplifiers linearly increases in the small signal region, which agrees with simulation results. In the region of the large signal condition, the output power is saturated. The measured maximum output power of the amplifier with one cell, four cell, and two-way four cell chips are 35.4 dBm (3.5 W), 42.4 dBm (16.5 W), and 44.6 dBm (28.6 W), respectively. Figure 3.9(b) shows the graph of the gain versus input power. According to Fig. 3.9(b), the -3 dB gain compression is observed at 27.1 dBm, 33.8 dBm, and 36.2 dBm input for amplifiers with one cell, four cells, and two-way four cell chips, respectively. In the small-signal condition, more than 10 dBm gain is obtained for each amplifier, which agrees with the simulation result. Figure 3.9(c) shows the graph of PAE versus input power. According to Fig. 3.9(c), 32.2 %, 43.0 %, and 44.8 % of maximum PAE are obtained for the amplifiers with one cell, four cells, and two-way four cell chips, respectively. Compared with the simulation results, there is a decrease in the PAE by 16.0, 9.2, and 9.4 points. The reason for this difference between the simulation and the measurement results is assumed to be due to manufacturing and/or measurement errors. In addition, a temperature rise in the GaN HEMT chips is also considered. The parameters of the chip models under the room temperature condition are used in this simulation.

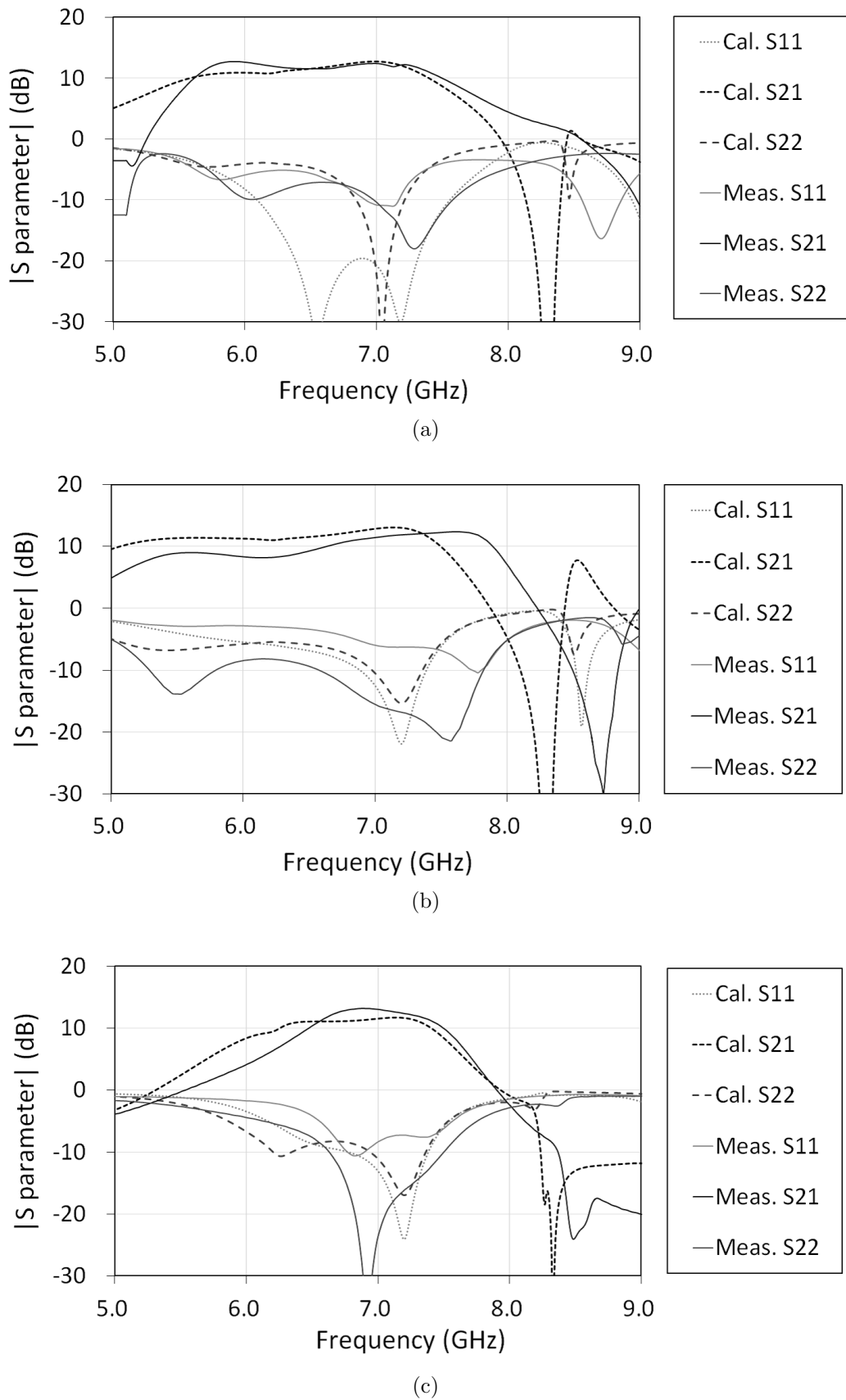
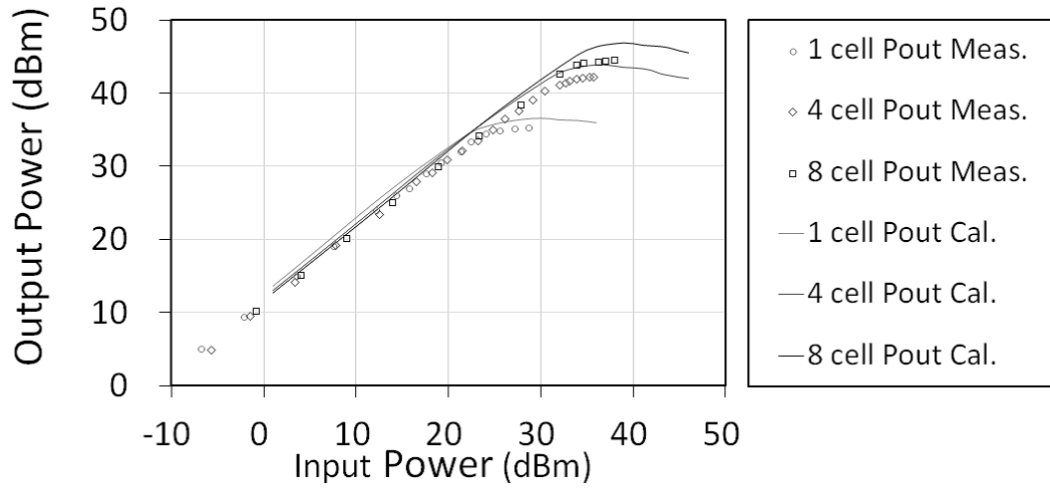
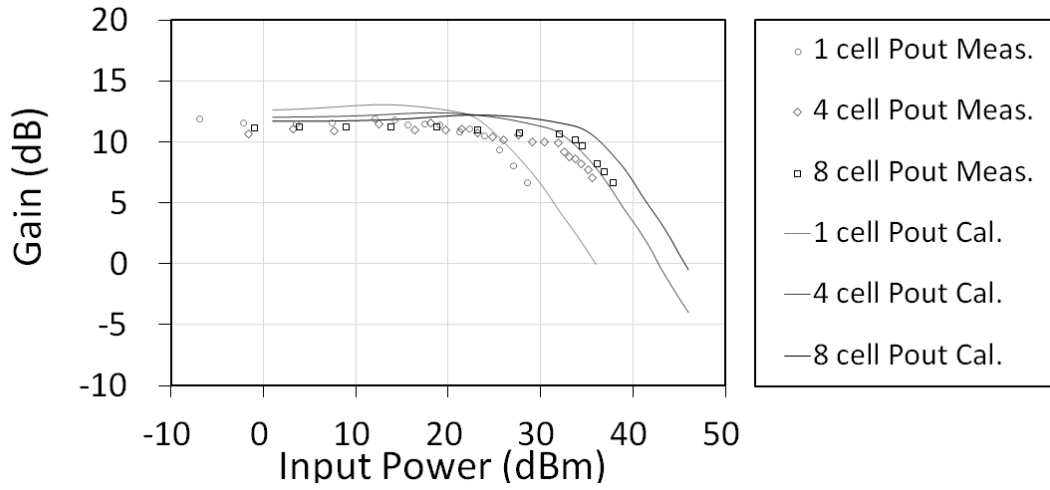


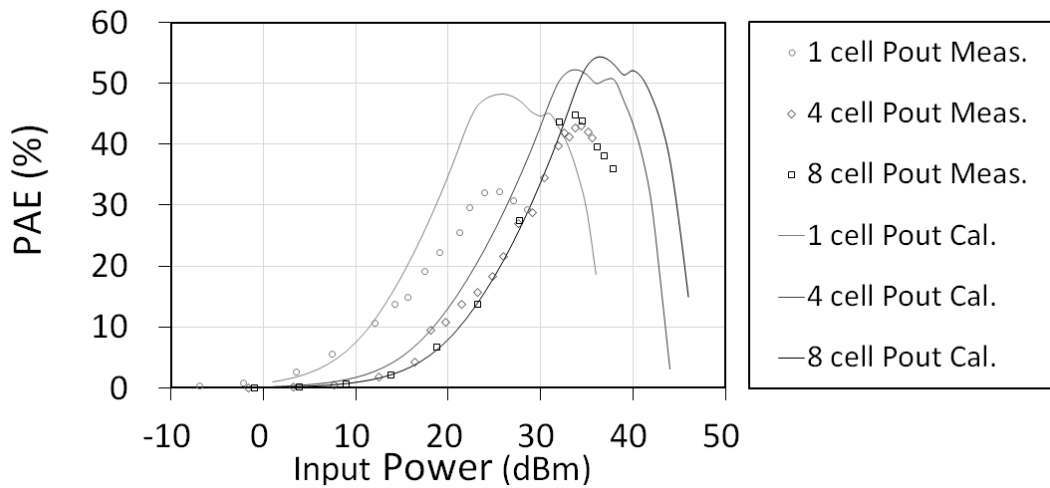
Figure 3.8: Small-signal S-parameters: (a) results of amplifier with 1 cell TGF2023-1 chip, (b) results of amplifier with 4 cell TGF2023-5 chip, and (c) results of amplifier with 2 way 4 cell TGF2023-5 chips



(a)



(b)



(c)

Figure 3.9: Input output characteristics: (a) output power, (b) gain, and (c) PAE.

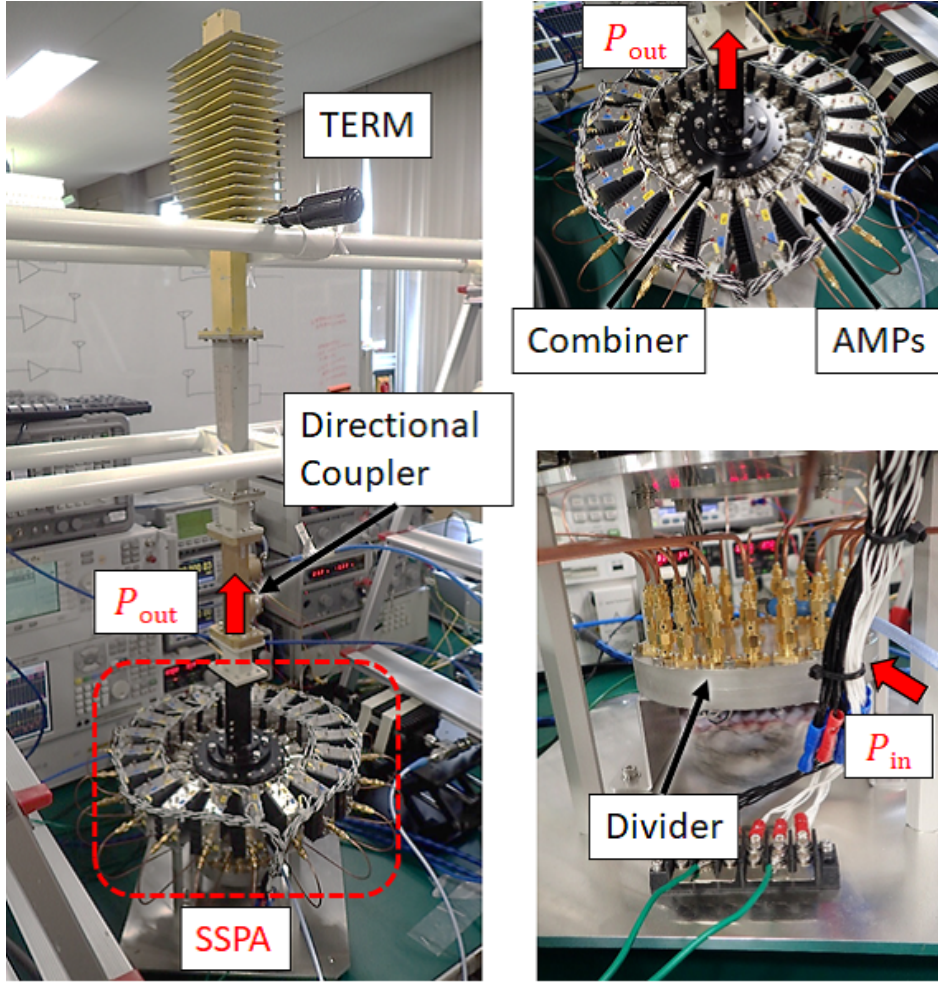


Figure 3.10: Photograph of SSPA with 20 way amplifiers.

3.5 20-way GaN Amplifier Assemble

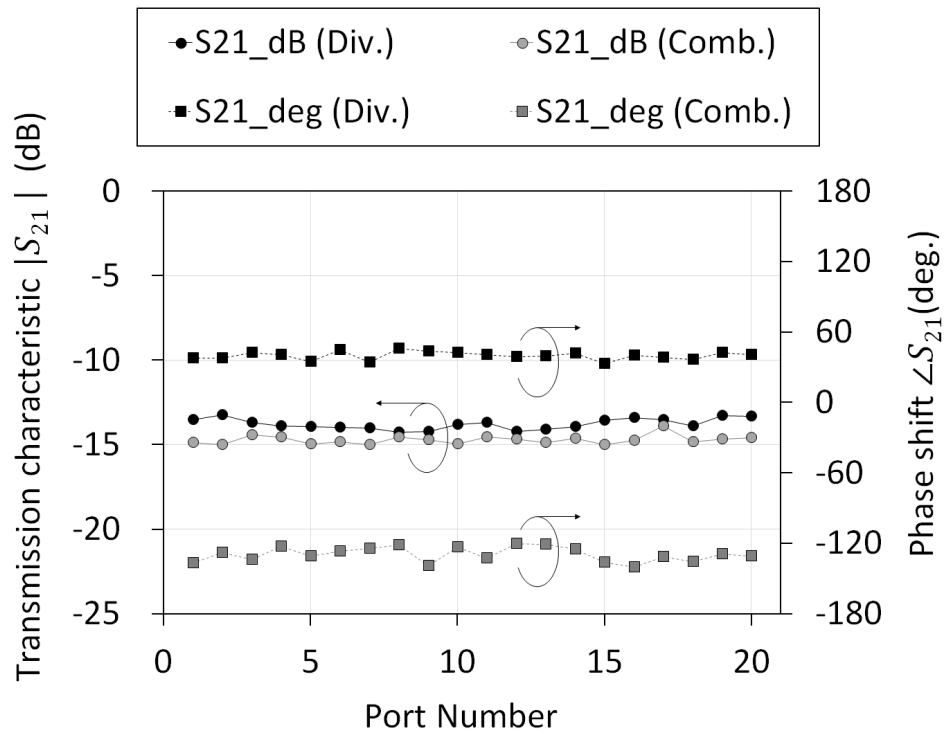
The photograph of the assembled SSPA is shown in Fig. 3.10. This SSPA consists of a 20-way radial divider, combiner, and amplifiers with the four cell TGF2023-05 chips. The amplifiers are attached to the radial divider and combiner by N-connectors and through the SMA semi-rigid cables, respectively. The commercial 20-way radial combiner and divider produced by NEC-NETS are used in this experiment. Figure 3.11(a) and Fig. 3.11(b) show the measured the reflection characteristics S_{11} and transmission characteristics S_{21} of 20-way combiner and divider. According to Fig. 3.11(a), the average $|S_{11}|$ s of divider and combiner for each port are -11.2 dB and -12.3 dB. The standard deviations are 0.4 dB and 0.2 dB, respectively. In addition, the average phase shift $\angle S_{11}$ of divider and combiner are -106.6 degree and -129.5 degree. The standard deviations are 4.7 degree and

6.2 degree, respectively. According to Fig. 3.11(b), on the other hand, the average $|S_{21}|$ s of divider and combiner for each port are -13.8 dB and -14.7 dB. The standard deviations are 0.3 dB and 0.2 dB, respectively. In addition, the average phase shift $\angle S_{21}$ of divider and combiner are 39.8 degree and -129.4 degree. The standard deviations are 3.3 degree and 6.3 degree, respectively. According to this results, the divider and combiner have the characteristics of low loss and low phase difference for each port, which contribute a high power and high efficiency SSPA assemble.

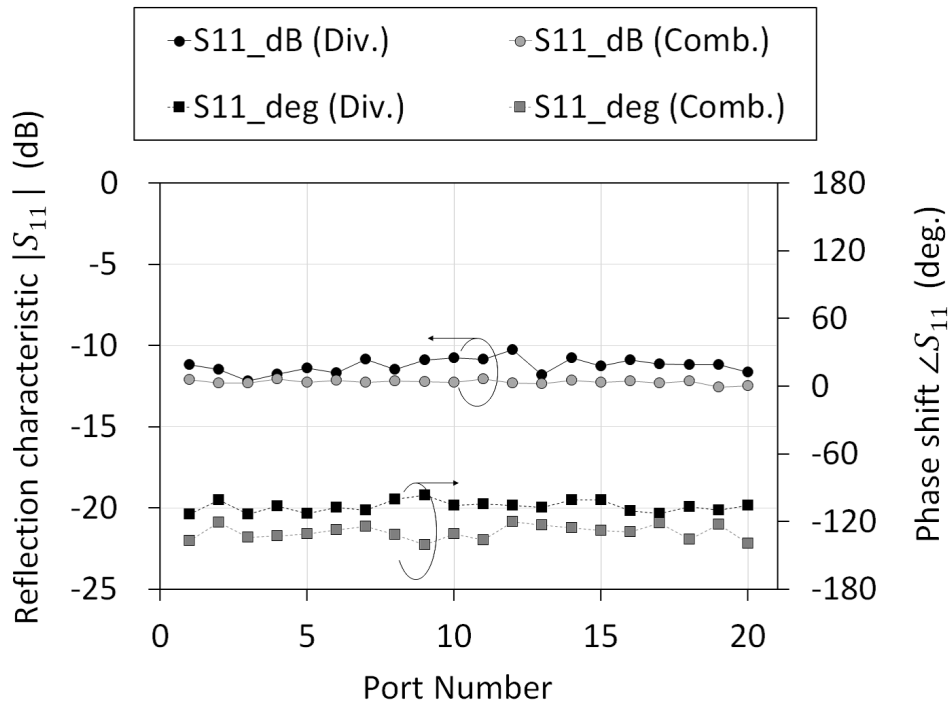
The large-signal performance of SSPA is measured. The input loss including the directional coupler, isolator, and coaxial cables is 31.5 dB. On the other hand, the output loss including the waveguide directional coupler, attenuator, and cables is -68.0 dB. The SSPA is driven with an air spot cooler (SS-25DG, Suiden). PW18-1.8AQ (TEXIO) and PS60-18 (TEXIO) $\times 2$ are used as the DC supplies for the gate and drain bias. The drain current is monitored by the digital multimeter (34461A, Keysight). The 28 V drain bias is supplied under the AB class condition. For the driver, the fabricated amplifiers with one cell, four cells, and two-way four cell chips are used. In this system, the radial combiner and divider are included in the DUT.

The measured result for the large-signal performance of the SSPA is shown in Fig. 3.12. The measured output power, gain, and PAE are shown in Fig. 3.12. According to this result, the measured maximum gain is 10.7 dB at 30.2 dBm input. The output power increases linearly in the region of the small-signal condition. In the region of the large-signal condition, the gain is compressed, which decreases the ratio of output power. In this experiment, the -3 dB gain compression at 44.5 dBm input is observed and the measurement is stopped at this time. At the -3 dB gain compression, the 52.6 dBm (179.8 W) output and 17.4% PAE are observed.

The large-signal results of SSPA and GaN amplifier are compared in Table 3.4. According to Fig. 3.11(a) and Fig. 3.11(b), the average losses of combiner are 2.0 dB is assumed. According to Fig. 3.12, therefore, the 41.6 dBm of GaN amplifier output power is calculated from the results as shown in Fig. 3.12, which agree with the result as shown in Fig. 3.9(a). In addition, calculating to the PAE from power amplifier result as shown in Fig. 3.9(c) and combiner loss, the 20.4% PAE of the SSPA is expected, which explain the PAE reduction.



(a)



(b)

Figure 3.11: S-parameters for each port of 20-way combiner and divider: (a) reflection characteristic $|S_{11}|$ and $\angle S_{11}$, and (b) transmission characteristic $|S_{21}|$ and $\angle S_{21}$.

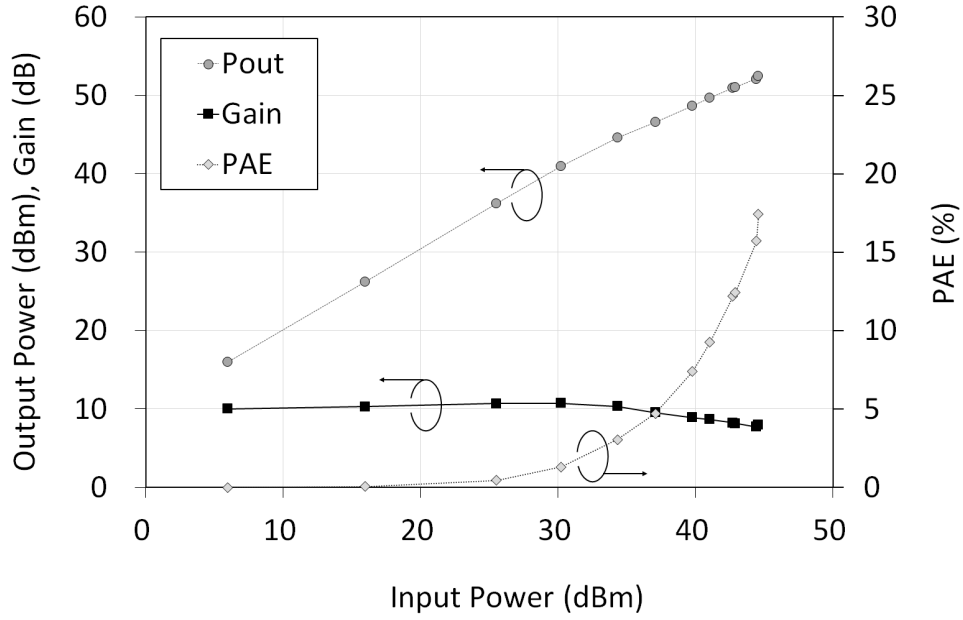


Figure 3.12: Measured large-signal performance of the SSPA.

Table 3.4: Specs of fabricated amplifiers at 7.1 GHz at -3 dB gain compression.

	Pout	PAE
AMP with one cell	35.3 dBm	30.8 %
AMP with four cells	41.7 dBm	42.6 %
AMP with 2 way four cells	44.6 dBm	44.8 %
SSPA with 20 way AMPs (80 cells)	52.6 dBm	17.4 %

The assembled SSPA realize high power and high efficiency characteristics on CW operation, which experimentally shows the availability of low loss design techniques as shown in the last section.

3.6 Summary

Amplifiers with one cell, four cell, and two-way four cells TriQuint TGF2023 series GaN HEMT chips are designed and fabricated. The circuits of amplifiers are matched to the source and load impedances of GaN chips calculated using non-linear models. The matching network consists of two stages. In the first stage, the impedance of chips is matched to 30Ω . The 30Ω impedance is transformed to 50Ω in the second stage. A Wilkinson power divider is used in the design of

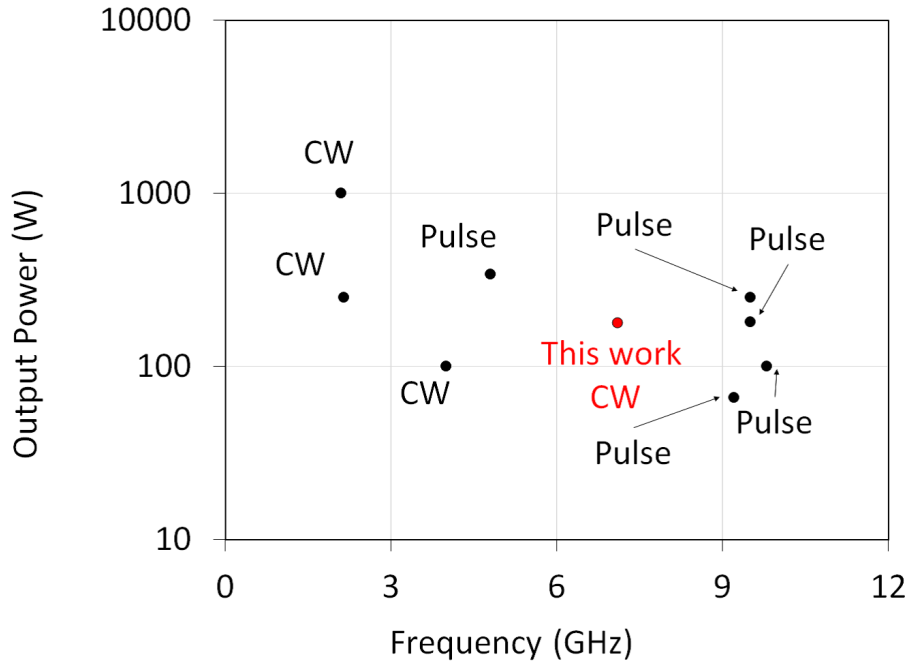


Figure 3.13: State of the arts.

the amplifier with two-way chips. According to the measurement results of the amplifiers with one cell, four cell, and two-way four cell chips, gains of 13.0 dB, 11.7 dB, and 13.7 dB are obtained under the small-signal condition. In addition, 35.3 dBm, 41.7 dBm, and 44.6 dBm output power and 30.8 %, 42.6 %, and 44.8 % PAE are observed for the amplifiers with each chips, respectively.

The 7.1 GHz SSPA with 20-way amplifiers is demonstrated. The SSPA is assembled with the 20-way amplifiers with four cell chips attached to the radial combiner and divider. In the large-signal measurement, the fabricated amplifiers are used for drivers. According to the measurement result under the large-signal condition, a 52.6 dBm output power and 17.4 % PAE are confirmed at 7.1 GHz. In addition, the measured maximum gain is 10.7 dB.

The comparison of this work with the recent researches of high-power amplifiers in the S- to X-band. The output power of SSPA is similar to the other researches on the C- to X-band, which is realized in the CW-drive condition in this work. In comparison with the researches on the S-band, the output power of the SSPA is lower because the GaN HEMT chip performance is limited on frequencies higher than the C-band. It is necessary for increasing the output power to improve the chip performance or increase the combining number in the future.

The comparison of this work with the recent researches of high-power amplifiers in the S- to X-band is shown in Fig. 3.13.

In this work, the 179.8 W output from assembled SSPA is realized, which compare favorably with recent research at the C- and X-bands. In particular, the SSPA can drive in the CW condition. Therefore, it is expected that SSPA will be employed for future high power microwave transmission such as space applications in deep space missions and MPT. In the future, the production type of the SSPA will be installed in the space ground base station for up-link satellite communication system based on large scale aperture antenna technologies. In addition, the SSPA will be used for microwave power transmission experiments.

Chapter 4

High Efficiency GaN Amplifier and Active Antenna Design

4.1 Introduction

A long-range microwave transmission system assumes the use of an active array antenna for high efficiency transfer. The active array antenna often consists of amplifiers and planar antennas. Recently, high-power microwave transmission systems have been reported with GaN HEMT technologies for use in amplifiers [100], [101].

GaN semiconductors are expected to be useful for high-power, high-frequency technologies for wireless transmission systems required for high-power RF-based microwave transfer. High-power GaN RF amplifiers have recently been reported [102], [103]. Active-antenna techniques are also expected for space applications such as satellite communication, future SSPSs, and radar systems, which require a lightweight high-power antenna system. Recently, high-power GaN amplifier units for active array antennas have been reported [104], [105]. In addition, Chung et al. reported designs that integrated a GaN amplifier into a microstrip active antenna for RF front-end applications in 2003, thereby realizing a 42 % PAE and a 30.3 dBm saturated output power with a linear gain of 8 dB in the C-band (7.25 GHz) [106].

High-efficiency GaN amplifiers with harmonic-tuning circuits have been reported [107]-[112]. Kuroda et al. reported an immensely high-efficiency GaN amplifier with a 71.4 % PAE and a 33.4 dBm output power in the C-band (5.86 GHz) [113], which is extremely useful for GaN active-antenna design.

Using a conventional active-antenna design, Qin et al. reported a high-efficiency active antenna with an E-class power amplifier [114]. A high-efficiency power amplifier with an 82.1 % PAE and a 22.08 dBm output power was fabricated in [114]. However, the PAE was reduced to about 60 % by integrating it with an antenna. Therefore, it is necessary for high-efficiency active antenna to losslessly integrate a power amplifier.

Lightweight high-power active-antenna systems also require appropriate integration with a cooling system. For instance, the SSPS unit is expected to use approximately ten billion active-antenna elements to obtain a 1 GW output power transferred by microwave to the ground [115]. Hence, it is necessary for a compact and lightweight design to effectively integrate active-antenna circuits with cooling systems such as Peltier unit.

Large-signal characteristics such as the output power, PAE, and gain of the GaN HEMT are reduced by electrothermal effects. Wang et al. reported a PAE reduction of up to 17 % between temperature of -55°C , 25°C , and 175°C in a large-signal modeling of a GaN HEMT [116]. Hence, temperature management of a power amplifier is an important factor for preventing PAE reduction.

It is necessary to prevent the reduction in efficiency seen in conventional studies when an amplifiers integrated into an antenna. The impedance matching with a large transform ratio is generally hard to losslessly design, which is one of the causes of efficiency reduction. This impedance matching procedure is applied to the high power design of the GaN-active antenna. The conventional reports scarcely refer to antenna impedance. However, the well-suited antenna impedance is able to simplify the matching network design and to realize a lossless integration with the amplifier. Therefore, an antenna design and integration with a well-suited impedance is proposed for lossless integration in a high-efficiency GaN amplifier in this work.

Here, a lossless-design method for a high-efficiency GaN active antenna is proposed to prevent PAE reduction after integration. Moreover, a high-efficiency, high-power active antenna that integrates GaN amplifiers with harmonic-tuning

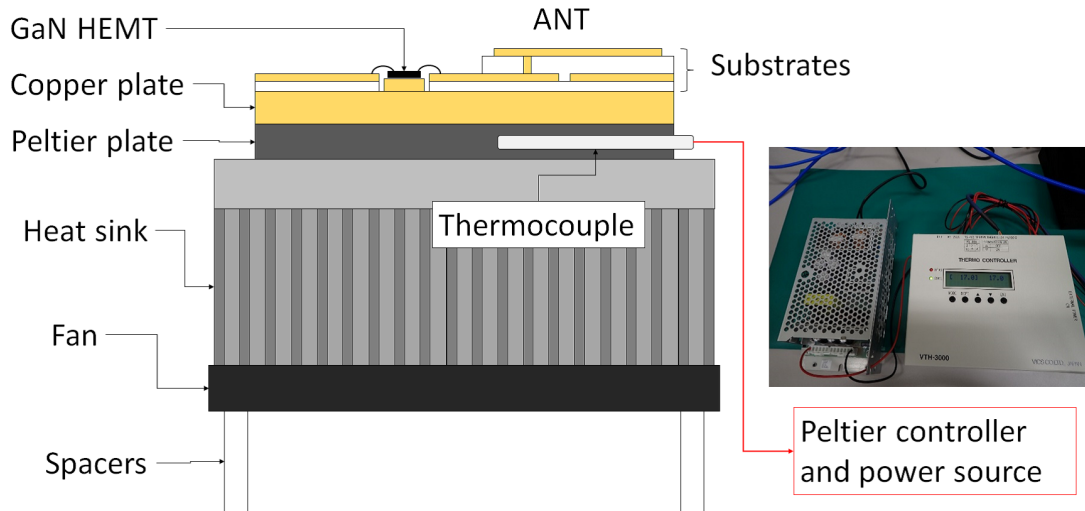


Figure 4.1: Active-antenna configuration on a Peltier cooling unit.

circuits and Peltier cooling units is designed for a center frequency of 5.8 GHz. The antenna impedance is suitably designed for the output impedance of the harmonic-tuning circuit, allowing a lossless integration. Figure 4.1 shows the active-antenna configuration. In a conventional GaN-active-antenna design, the antenna and amplifier circuits are integrated on the same dielectric constant substrate of 10.2 [106], causing degradation of the antenna gain. In general, the microstrip patch antenna gain has a negative-slope characteristic to increase dielectric constant and substrate thickness. However, the low dielectric constant expands the line width, which is the case of increasing radiation loss. In addition, the difference of dimension between GaN HEMT chip and line width makes an equivalence capacitance which reduces the calculation accuracy of circuit simulators. Therefore, a multilayer substrate consisting of low- and high-dielectric constants for reducing the line width of amplifier circuits and increasing the gain of antenna, respectively, is adopted in our design. The substrates are piled up on the Peltier cooling unit (VICS LVPU-70), as shown in Fig. 4.1. The LVPU-70 Peltier unit has a cooling capacity of up to 71 W, which is sufficient for use as the amplifier's cooling system. Moreover, the LVPU-70 is able to precisely control the surface temperature, which allows our discussion to be simplified regardless of the electro-thermal effect.

The antenna includes a harmonic-tuning circuit designed for GaN-HEMT impedance, which is calculated by load-pull simulation with a nonlinear GaN model. The antenna is also designed to reduce the transform ratio between the antenna and amplifier impedances, thereby realizing an appropriate integration of the devices.

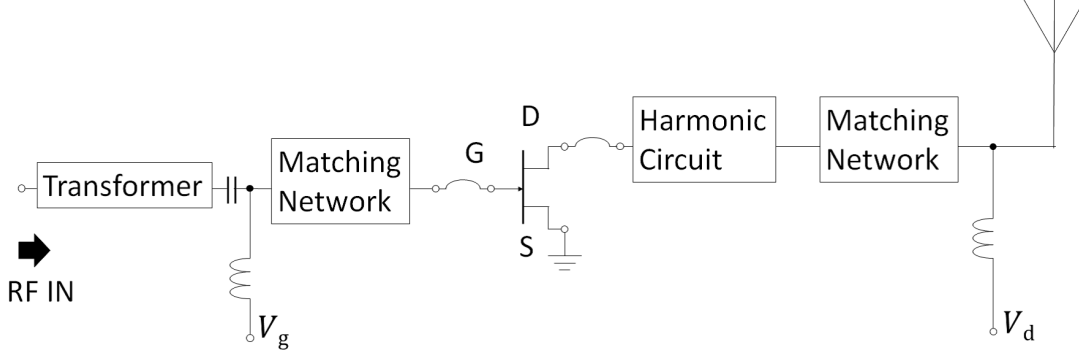


Figure 4.2: GaN-active-antenna-circuit configuration.

In this paper, the GaN amplifier is first fabricated in order to confirm the improvement of efficiency using a harmonic treatment circuit. Then, the amplifier and antenna are losslessly integrated to demonstrate the viability of our integration method.

4.2 Theory of High Efficiency Amplifier Design

The PAE improvement is explained by reduction of the harmonic-dissipation power. The instantaneous drain voltage ($V_d(t)$) and current ($I_d(t)$) can be expressed by the following equations

$$V_d(t) = V_{DC} + \sum_{n=1} \sqrt{2}|V_n| \cos(\omega_n t + \angle V_n), \quad (4.1)$$

$$I_d(t) = I_{DC} + \sum_{n=1} \sqrt{2}|I_n| \cos(\omega_n t + \angle I_n), \quad (4.2)$$

where V_{DC} and I_{DC} are DC components. The second terms of the equations show the fundamental- and harmonic-frequency components. The instantaneous average power $P(t)$, as calculated by Eqs. (1) and (2) can be expressed by the following equation:

$$P(t) = V_{DC}I_{DC} + \sum_{n=1} |V_n||I_n| \cos(\angle V_n - \angle I_n). \quad (4.3)$$

Here, for the phase difference $\angle V_n - \angle I_n = 90^\circ + 180^\circ n$ at each harmonic frequency, the dissipation power of the harmonic becomes zero, which improves the harmonic loss.

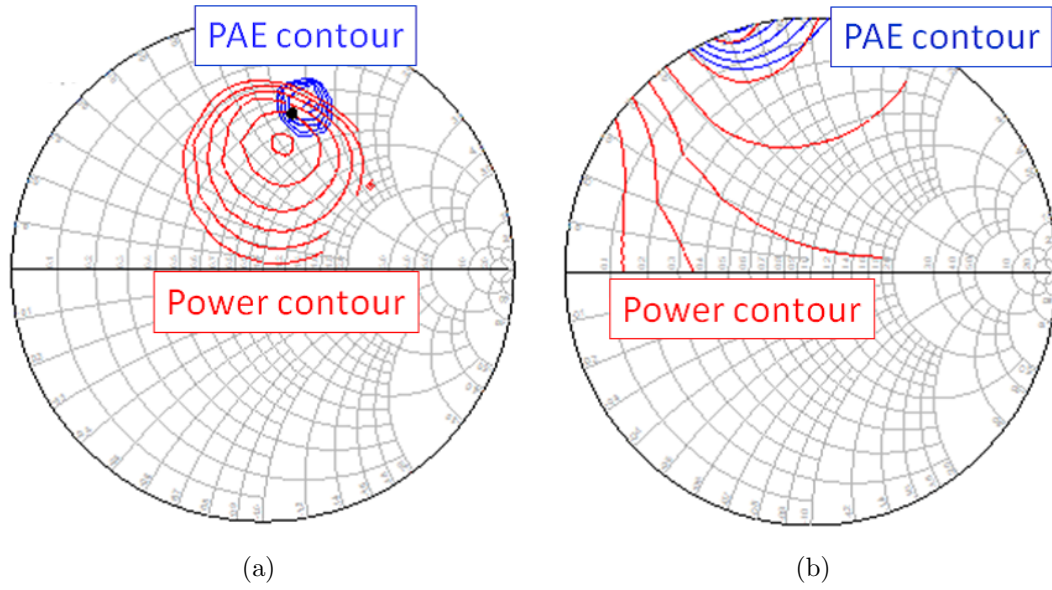


Figure 4.3: Results of load-pull simulation for a nonlinear model of a GaN HEMT. (a) Fundamental frequency of 5.8 GHz and (b) second-harmonic frequency of 11.6 GHz.

4.3 High Efficiency GaN Amplifier Design

The GaN active-antenna-circuit configuration is shown in Fig. 4.2. This configuration consists of a harmonic tuning circuit, matching networks, bias lines, a transformer, and an antenna. First of all, a single-stage GaN amplifier is designed and fabricated.

4.3.1 Load-pull Simulation

The load- and source-side circuits are designed for GaN-HEMT impedances, which are calculated by load-pull simulation with a high-power GaN HEMT SG0601C (Sumitomo Electric) nonlinear model in a Keysight ADS-RF circuit simulator. The SG0601C is operated with a 50 V drain voltage and a -2.5 V gate voltage, which provide an AB-class dc-bias condition. The results of load-pull simulation under this dc-bias condition are illustrated in Fig 4.3. Figs. 4.3(a) and (b) show the results at a 5.8-GHz fundamental frequency and an 11.6 GHz second-harmonic frequency, respectively. The input power is 25 dBm in this simulation. The red and blue lines show the output power and PAE contours, respectively. These contours show the 0.25 dB and 1.0 point step lines. According to the results, the maximum output power and PAE of 41.18 dBm and 67.11 %, respectively, are confirmed.

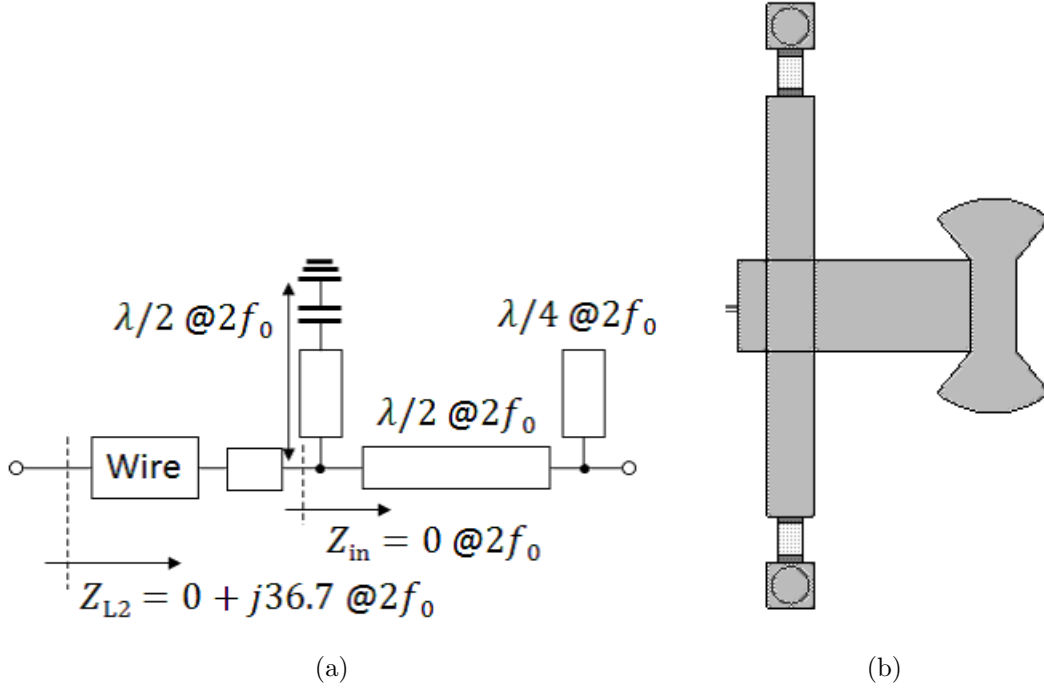


Figure 4.4: (a) Harmonic-treatment-circuit schematic and (b) layout.

In this work, the impedance for maximum PAE is targeted. According to Fig. 4.3, the fundamental and second-harmonic GaN HEMT impedances of $Z_{L1} = 25.6 + j53.3 \, \Omega$ and $Z_{L2} = 0 + j36.7 \, \Omega$ are calculated to obtain a maximum PAE. In this condition, the source-side input impedance is $Z_S = 2.9 + j11.6 \, \Omega$ at the fundamental frequency.

4.3.2 Amplifier Circuit Design

A design technique for harmonic-tuned power amplifiers was reported in [119], whose approach was based on a methodology developed for the design of multi-frequency passive matching networks. In this work, a power amplifier is designed assuming the second-harmonic tuning.

The harmonic-tuning circuit is designed for the calculated impedance of the second-harmonic frequency. Figs. 4.4(a) and (b) show the harmonic-tuning-circuit schematic and layout, respectively. In this circuit, the appropriate input impedance at the second-harmonic frequency is provided by open and short stubs, as shown in Fig. 4.4(a). The microstrip circuits are connected by wire bonding to a GaN-HEMT chip. The 2- and 6-segment wires are used for gate- and drain-side bonding.

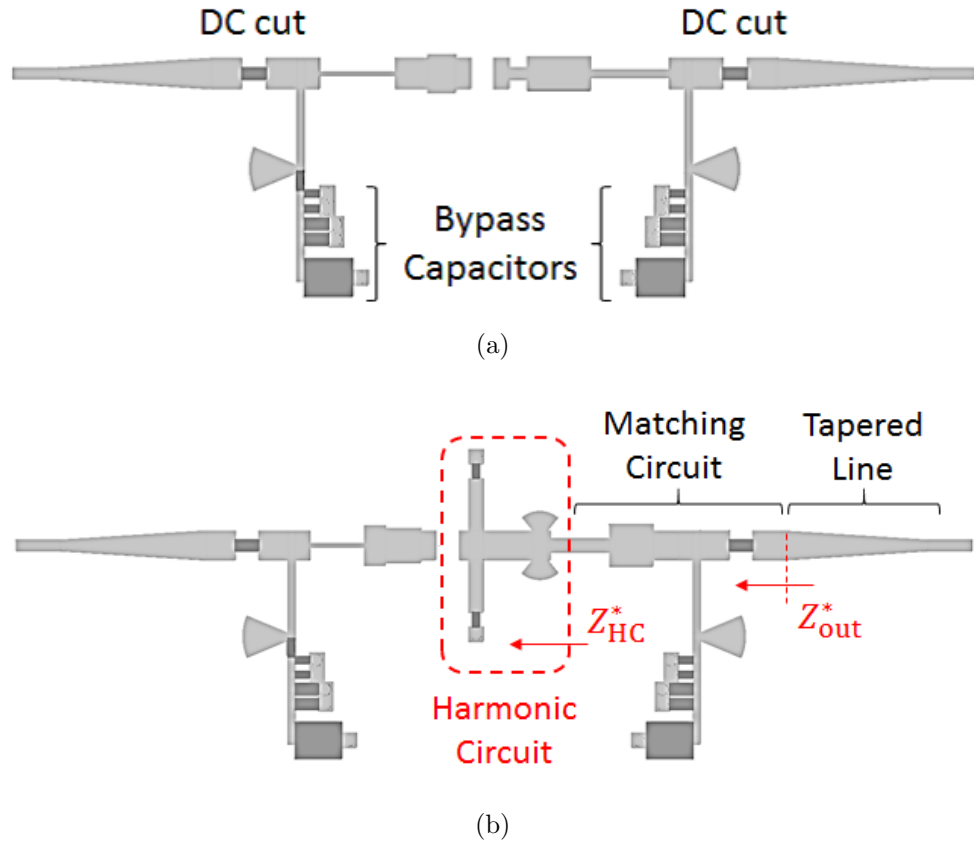


Figure 4.5: Amplifier layouts (a) without and (b) with a harmonic tuning circuit.

A wire length of 0.5 mm is assumed, which is sufficiently shorter than the wavelength. The 2-segment-wire impedances for gate-side bonding at 5.8 GHz and 11.6 GHz are $0.11 + j8.79 \Omega$ and $0.15 + j17.52 \Omega$, respectively. In addition, the 6 segment wire impedance for drain side bonding at 5.8 GHz and 11.6 GHz are $0.04 + j4.07 \Omega$ and $0.05 + j8.13 \Omega$, respectively.

The layouts of the designed amplifier circuits are shown in Fig. 4.5. These circuits are designed on an Arlon AD1000 substrate. The substrate's dielectric constant, dielectric-dissipation factor, and thickness are $\epsilon_r = 10.2$, $\tan \delta = 0.0023$, and $H = 0.8$ mm, respectively. Figs. 4.5(a) and (b) show the amplifier layout without and with the harmonic-tuning circuit, respectively.

The output impedance of the harmonic circuit Z_{HC}^* which has $32.6 - j24.8 \Omega$ for the maximum PAE at the fundamental frequency is transformed to 50Ω by the designed circuit as shown in Fig. 4.5(b). In the matching circuit of Fig. 4.5(b), the imaginary part of Z_{HC}^* is reduced. Hence, Z_{HC}^* is transformed to $Z_{out}^* = 27.57 - 1.71 \Omega$ by the matching circuit. Next, Z_{out}^* is transformed to 50Ω by the

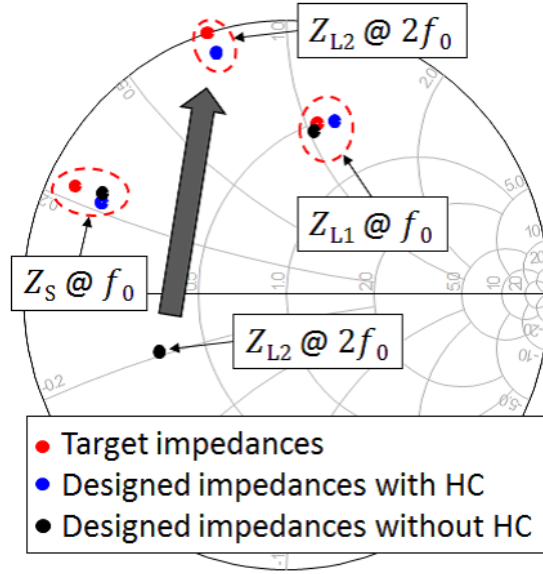


Figure 4.6: Target impedance calculated by load-pull simulation for maximum efficiency and the designed impedance of the load- and source-amplifier circuits with and without a harmonic tuning circuit.

tapered line. In the design of the gate-side and amplifier circuits without harmonic treatment, the same method is used.

The load- and source-side input impedances, Z_{L1} , Z_{L2} , and Z_S , of the designed circuits are shown in Fig. 4.6. The target impedances show the results of the load-pull simulation. According to Fig. 4.6, the impedances Z_{L1} and Z_S of each circuit layout agree with the target impedance. However, the impedance Z_{L2} of the circuit without a harmonic-tuning circuit deviates from the target. On the other hand, this is not the case when a harmonic tuning circuit is included.

Tables 4.1 and Table 4.2 show the calculated parameters of the amplifiers without and with harmonic-treatment circuits, respectively, at fundamental f_1 to the fourth-harmonic frequency f_4 by ADS. According to Tables 4.1 and Table 4.2, the harmonic power of each frequency is reduced comparing the amplifier with the harmonic-treatment circuit to that without this circuit. In particular, the phase difference of the amplifier with a harmonic-treatment circuit at the second-harmonic frequency becomes a value close to $270^\circ (= 90^\circ + 180^\circ)$. The value of the phase difference at the second harmonic close to $90^\circ + 180^\circ n$ is considered to be the confrontive cause of efficiency improvement. The ratios of total harmonic power to fundamental power are 13.47% and 3.34% in the Tables 1 and 2, respectively. Therefore, a 10.13-point increase of the ratio is confirmed by these results, demonstrating the efficiency improvement by the harmonic treatment circuit.

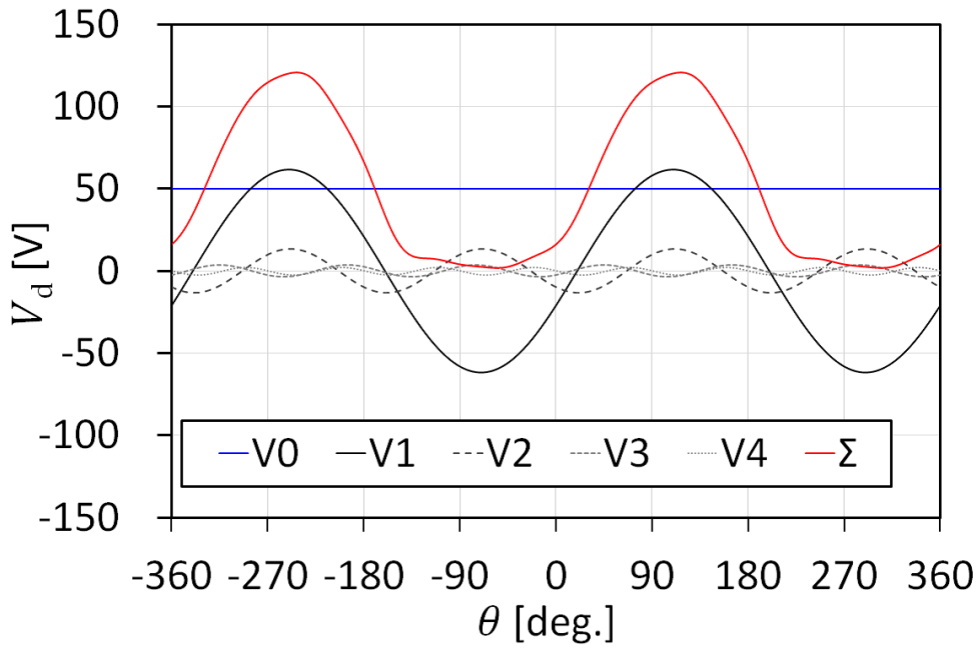
Table 4.1: Harmonic parameters of the amplifier without a harmonic-treatment circuit

	$ V [\text{V}]$	$\angle V [^\circ]$	$ I [\text{A}]$	$\angle I [^\circ]$	$ \angle V_n - \angle I_n $	$ P [\text{W}]$
f_1	53.11	106.38	0.46	11.78	118.20	5.71
f_2	5.89	107.34	0.16	-42.62	149.96	0.41
f_3	4.10	-119.09	0.26	-3.88	115.21	0.23
f_4	3.05	98.51	0.11	-114.49	213.00	0.77

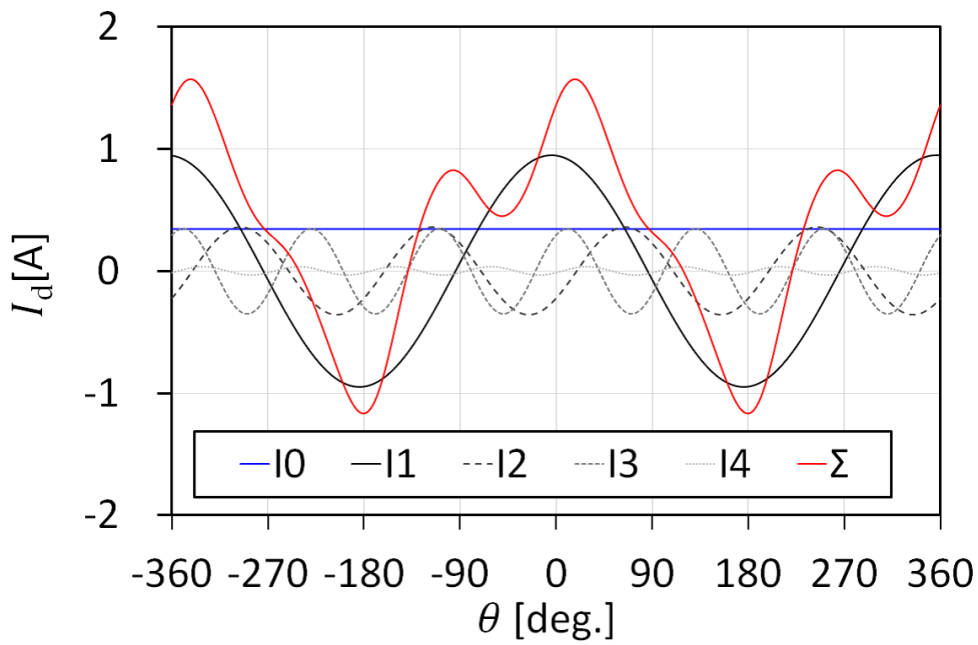
Table 4.2: Harmonic parameters of the amplifier with a harmonic-treatment circuit

	$ V [\text{V}]$	$\angle V [^\circ]$	$ I [\text{A}]$	$\angle I [^\circ]$	$ \angle V_n - \angle I_n $	$ P [\text{W}]$
f_1	61.74	-109.82	0.475	4.32	114.15	6.00
f_2	13.37	136.75	0.18	-128.21	264.96	0.11
f_3	3.51	-135.75	0.18	-30.91	104.69	0.08
f_4	2.31	80.76	0.02	-115.13	195.89	0.02

The drain voltage and current waves based on Tables 4.1 are calculated as shown in Fig. 4.7. These wave form are calculated by using Eq. 4.1 and Eq. 4.2. In these graph, V_0 is a drain voltage at DC condition, V_1 is a drain voltage at fundamental frequency, V_2 to V_4 are drain voltage at harmonic frequencies, and Σ is the calculated result by using Eq. 4.1. In addition, I_0 is a drain current at DC condition, I_1 is a drain current at fundamental frequency, I_2 to I_4 are drain current at harmonic frequencies, and Σ is the calculated result by using Eq. 4.2. According to Fig. 4.7(a), the half-wave rectified wave for instantaneous drain voltage is confirmed. The current wave form don't confirm a characteristic property. In the much lossless design, the overlap between the drain voltage and current wave form should be reduced, which is theoretically realized by treating higher-order harmonics.



(a)



(b)

Figure 4.7: Calculated wave forms of drain voltage and current at each harmonic frequencies.

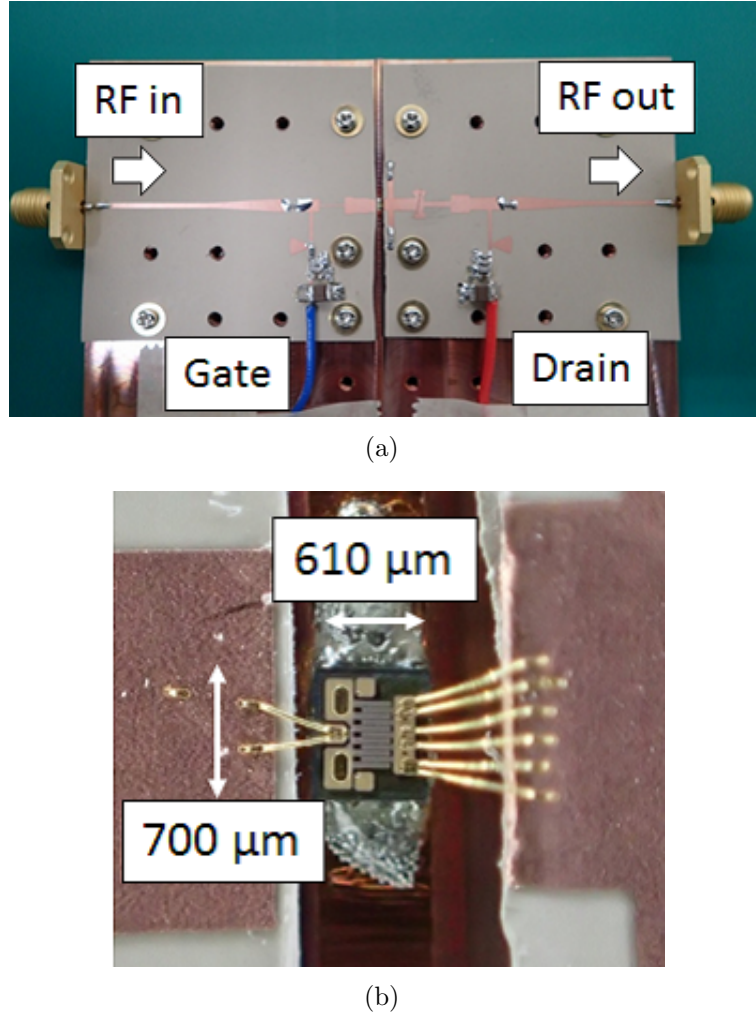


Figure 4.8: (a) Top view of the fabricated amplifier and (b) a mounted GaN HEMT chip.

4.3.3 Amplifier Fabrication and Measurement

The photograph shown in Fig. 4.8 illustrates the fabricated single-stage GaN amplifier with a harmonic-tuning circuit. Figures. 4.8(a) and (b) show the top view of the amplifier and the mounted GaN HEMT. The GaN HEMT chip is mounted on a copper plate and connected to lines by wire bonding. The amplifier is set on the Peltier unit for cooling. The GaN amplifier including the copper fixer has a surface area of 81 cm², a thickness of 0.8 cm, and a weight of 528 g. On the other hand, the Peltier-unit size including the heatsink, fan, and spacer has a surface area of 100 cm², a height of 11 cm, and a weight of 738 g.

At 5.8 GHz, the output power (P_{out}), amplifier gain (G_{ap}), and PAE are measured as large-signal characteristics and reported in Fig. 4.9. According to this

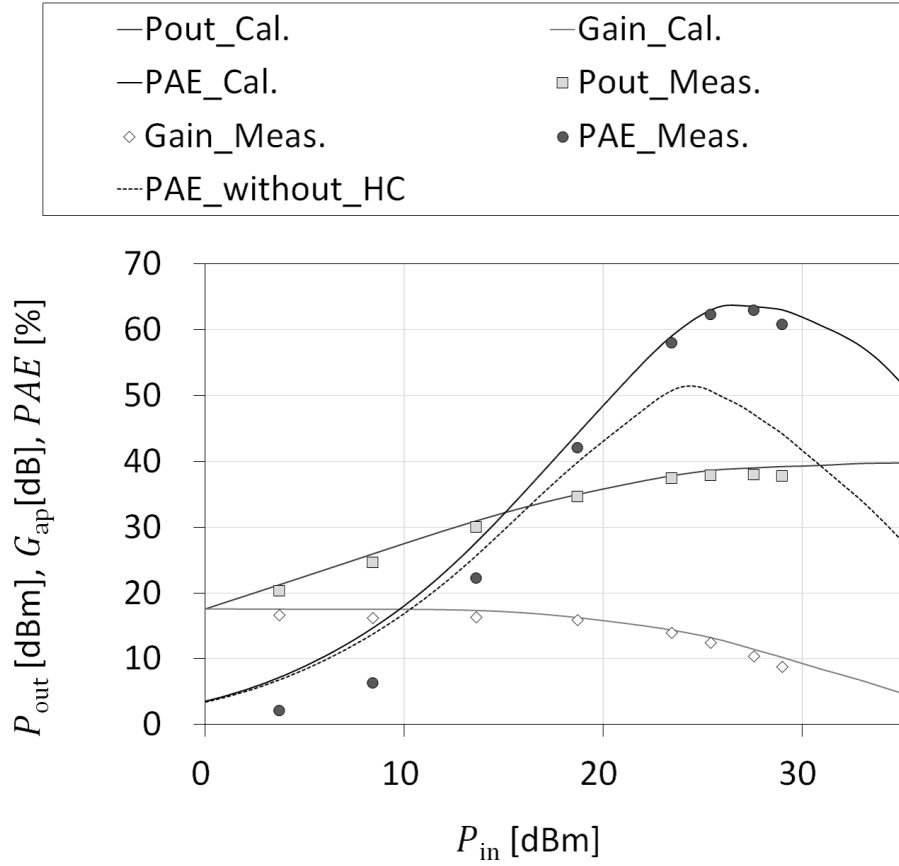


Figure 4.9: Large-signal characteristics of the amplifier.

measurement, the amplifier is operated with a 50 V drain voltage and a -2.5 V gate voltage. The temperature of the Peltier unit is 20°C . According to Fig. 4.9, the measured results agree with the simulated values. The measured maximum output power and PAE are 37.95 dBm (6.24 W) and 62.94 % at an input of 27.58 dBm, respectively. Under these conditions, an amplifier gain of 10.37 dB is confirmed, providing 6.23 dB of gain compression. The dashed line in Fig. 4.9 shows the simulated PAE of the amplifier without a harmonic-tuning circuit, with the maximum PAE of 51.37%. Comparing the maximum PAE of the amplifier with the harmonic tuning circuit to that of the amplifier without the circuit, an 11.57 % improvement is confirmed.

According to the measurement results, the amplifier loss at a 27.58 dBm input is $P = P_{dc} + P_{RF_{in}} - P_{RF_{out}} = 3.85$ W, where P_{dc} , $P_{RF_{in}}$, and $P_{RF_{out}}$ are the dc-input power, RF-input power, and RF-output power, respectively. The amplifier loss is mainly wasted by the Peltier unit with a 20°C surface temperature, which is sufficient to reduce the GaN chip temperature. Assuming that a heat sink is used

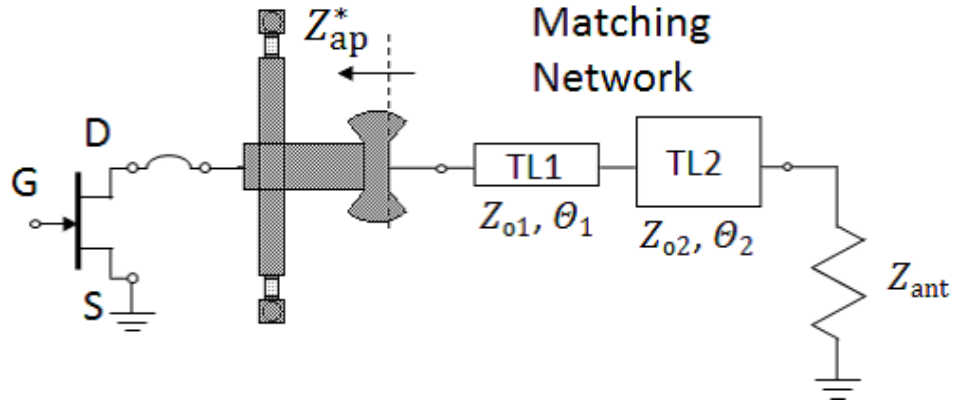


Figure 4.10: Load-side circuit for antenna impedance.

Table 4.3: Microstrip line parameters for impedance matching condition

	Z_{o1} [Ω]	θ_1 [$^\circ$]	Z_{o2} [Ω]	θ_2 [$^\circ$]
	(W [mm])	(L [mm])	(W [mm])	(L [mm])
to 50 Ω	43.28	23.09	27.28	102.87
	(1.00)	(1.25)	(2.17)	(5.30)
to 40 Ω	62.12	16.16	28.33	106.35
	(0.45)	(0.91)	(2.05)	(5.50)
to 30 Ω	68.86	19.31	29.66	25.21
	(0.34)	(1.10)	(1.91)	(1.31)
to 20 Ω	39.78	11.19	44.69	15.65
	(1.17)	(0.60)	(0.94)	(0.85)

under natural convection conditions, a surface size of about 0.14 m² is designed to reduce the chip temperature to 20 °C.

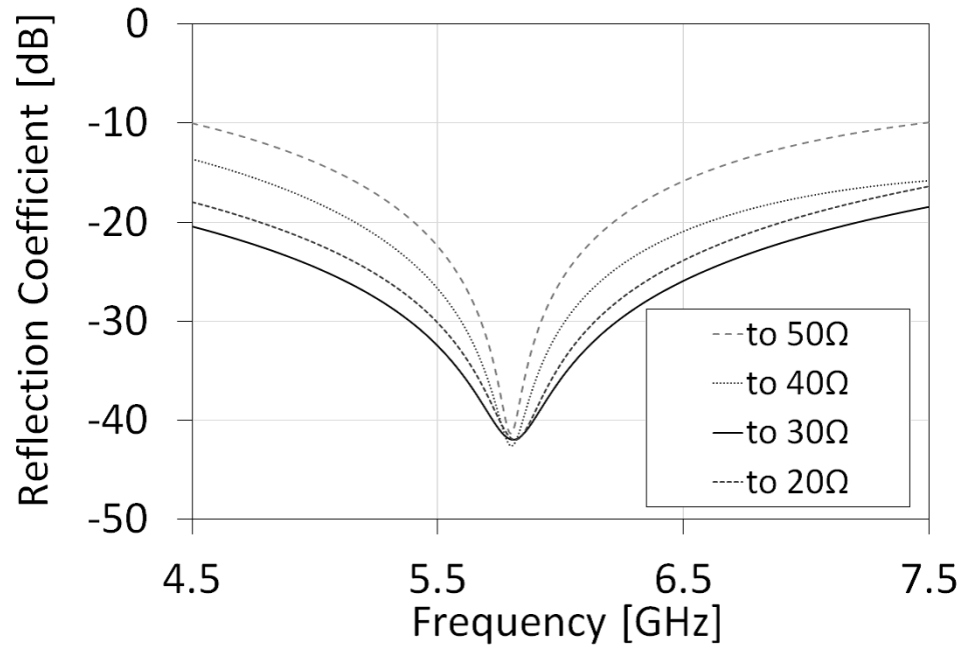
4.4 High Efficiency GaN Active Antenna

The designed GaN amplifier is integrated with an antenna. For this active-antenna design, we discuss an high efficiency integration method. For the high efficiency circuit design, it is important to reduce the loss of matching circuits. In the previous single amplifier design, the tapered line for low-loss matching technique was used. However, the tapered line circuit requires a long line length, which is

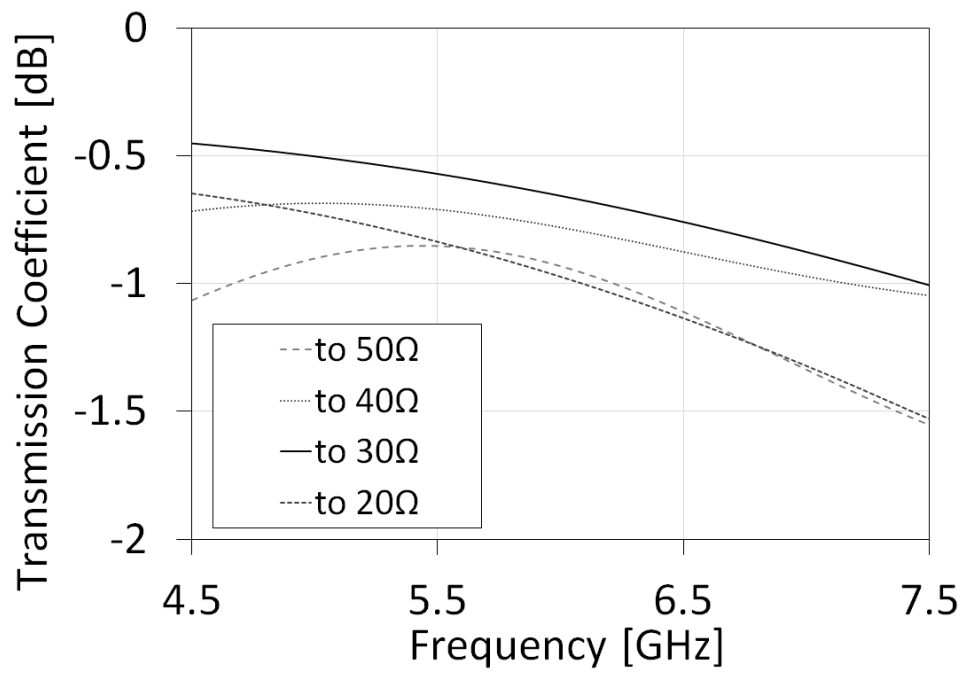
inadequacy for active integrated array antenna design. In this work, the antenna-input impedance is designed to decrease the impedance-transform ratio between the GaN HEMT with the harmonic tuning circuit and the antenna, with the expectation of preventing degradation of circuit characteristics such as bandwidth and transmission efficiency. In addition, the antenna is designed on a multilayer substrate, with the expectation of improving antenna gain. Figure 4.10 shows the load-side circuit configuration, which consists of a harmonic tuning circuit and a matching network for antenna impedance, Z_{ant} . This matching network is designed with two block-transmission lines, TL1 and TL2, with characteristic impedances and phases of Z_{o1} and θ_1 , and Z_{o2} and θ_2 , respectively. The microstrip-line parameters used to provide an impedance-matching condition for loads of $50\ \Omega$ to $20\ \Omega$ are reported in Table 4.3. The values in brackets show the microstrip-line width (W) and length (L).

In the $30\ \Omega$ load condition, the impedance, Z_{o1} , is designed to be $68.86\ \Omega$ and the phase, θ_1 , is designed to be 19.31° , providing an output impedance of $28.25 - j4.75\ \Omega$ for TL1. The imaginary part of Z_{ap}^* is almost entirely eliminated by the TL1 transmission-line block. Therefore, the TL2 transmission line hardly affects matching under the $30\ \Omega$ load condition.

The reflection and transmission coefficients of the matching network for each load designed on a finite-element-method (FEM) simulator (AWR Analyst) are reported in Figs. 4.11(a) and (b). According to this result, the reflection and transmission coefficients have superior wide-band and effective-transmission characteristics under a load of $30\ \Omega$ than at other loads. Figure 4.12 show the calculated PAE of GaN amplifier for each antenna impedance of $50\ \Omega$ to $20\ \Omega$. According to Fig. 4.12, it is confirmed that more than 5 % PAE is reduced at comparing $30\ \Omega$ to $50\ \Omega$ and $20\ \Omega$ load impedance conditions. The impedance of Z_{ap}^* , as shown in Fig. 4.10, is $32.6 - j24.8\ \Omega$ for the maximum PAE, which is calculated by load-pull simulation for the GaN-HEMT model with a harmonic-tuning circuit. Therefore, when a $30\ \Omega$ load is provided as antenna impedance, the impedance-transform ratio is reduced, which we expect will enable the high efficiency integration of the GaN amplifier and an antenna.



(a)



(b)

Figure 4.11: (a) Reflection and (b) transmission coefficients of matching networks for each antenna impedance of $50\ \Omega$ to $20\ \Omega$.

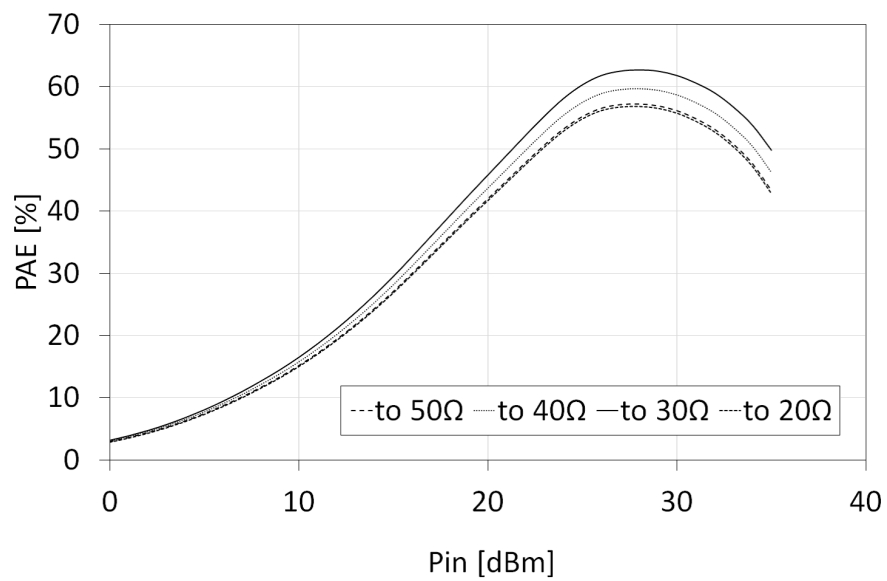


Figure 4.12: Calculated PAE for each antenna impedance of 50Ω to 20Ω .

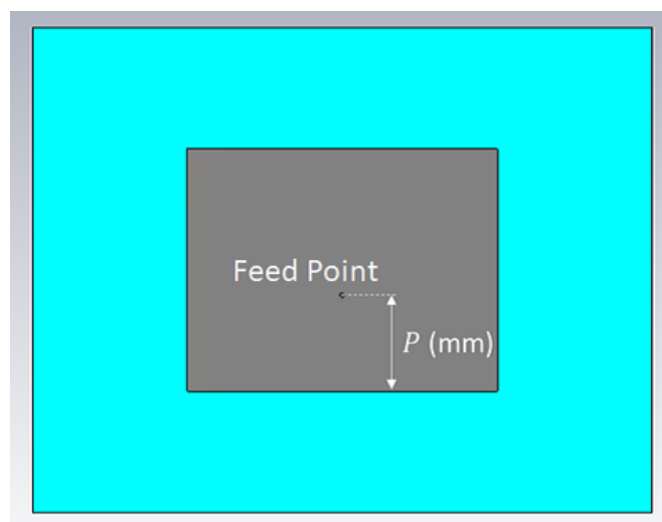
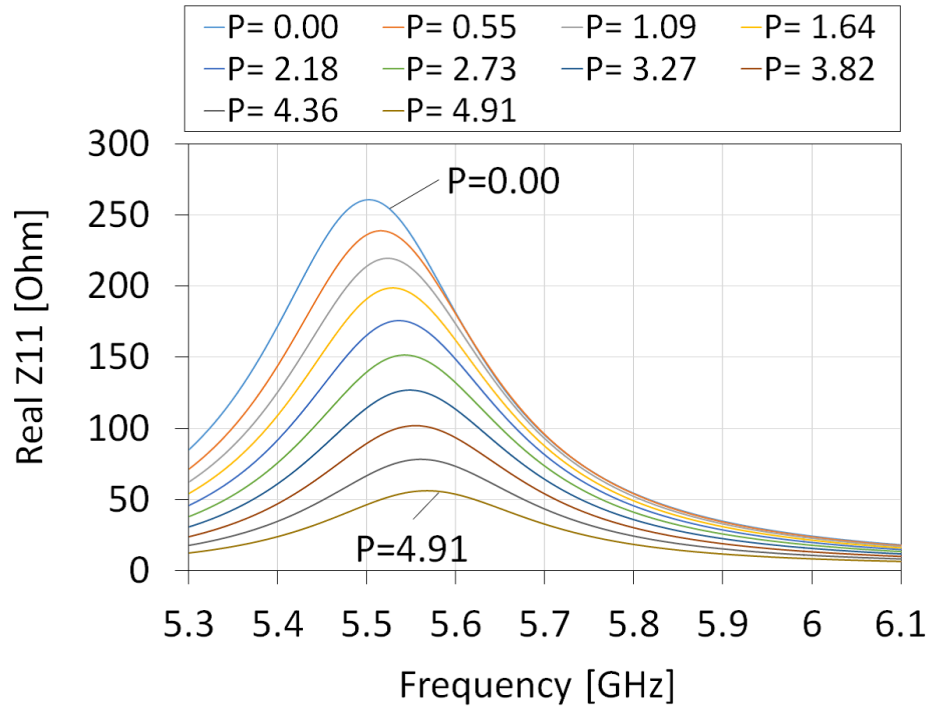


Figure 4.13: Rectangular antenna model in electromagnetic simulator.

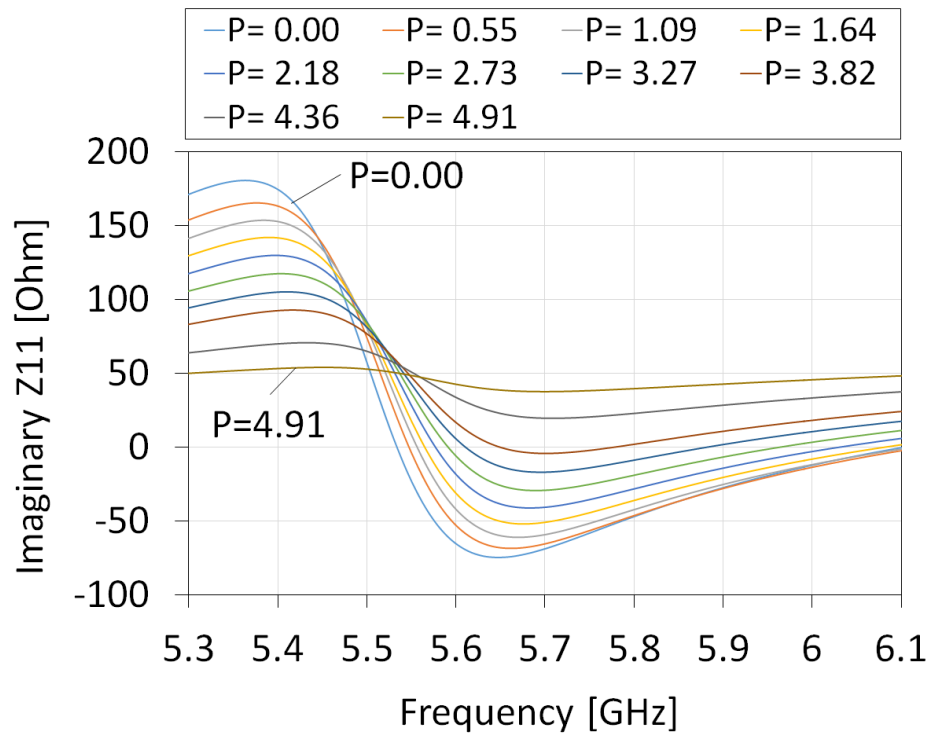
4.4.1 Single Antenna Design and Fabrication

The impedance of patch antenna is defined by the position of feed point, which is calculated by the ratio of E-field and H-field on the antenna. Assuming a rectangular patch antenna as shown in Fig. 4.13, the microwave signal is given through a pin from back side of antenna. The real part and imaginary part impedance of the antenna for each position P of the feeding pin are shown in Fig. 4.14. According to Fig. 4.14, when $P = 3.82$ mm, $30.05 + j1.83 \Omega$ of antenna impedance is obtained, which is a close value to 30Ω of appropriate load impedance for GaN HEMT with the harmonic tuning circuit. Therefore, the parameter of $P = 3.82$ mm is employed for antenna design.

In general, substrates with lower dielectric constants provide a higher antenna gain. On the other hand, it is necessary for a compact and lightweight circuit design to use a substrate with a high-dielectric constant. Therefore, a single-patch antenna is designed on a multilayer substrate, as shown in Fig. 4.15(a). This is a low-dielectric substrate consisting of Arlon AD1000 with $\epsilon_r = 2.55$ (Nippon Piller Packing NPC-F260A). The thickness and dielectric dissipation of NPC-F260A are $H = 1.6$ mm and $\tan \delta = 0.0018$, respectively. The parameters of Arlon AD1000 have the same values as listed in the amplifier-circuit-design section of this paper. Each layer is connected by a through hole, as shown in Fig. 4.15(a). Figs. 4.15(b) and (c) show layers 1 and 2 from the top view of the designed single antenna with $30\text{-}\Omega$ input impedance. The antenna is designed using a FEM simulator (CST Microwave Studio). The backside of this substrate serves as a ground for the circuit. This ground is connected to the ground pad on layer 1 as shown in Figs. 4.15(a) and (b), which in turn is used for the antenna ground on layer 2.



(a)



(b)

Figure 4.14: (a) Real part and (b) imaginary part of antenna impedance for each position of feed point.

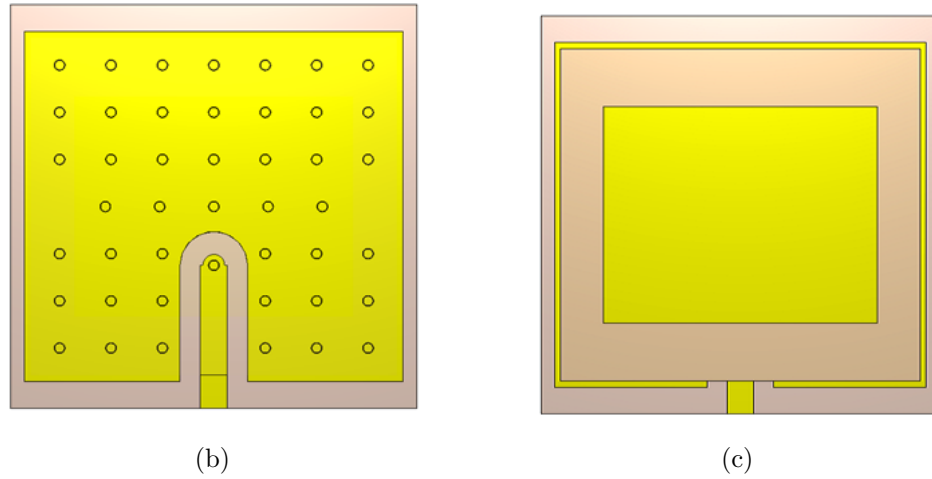
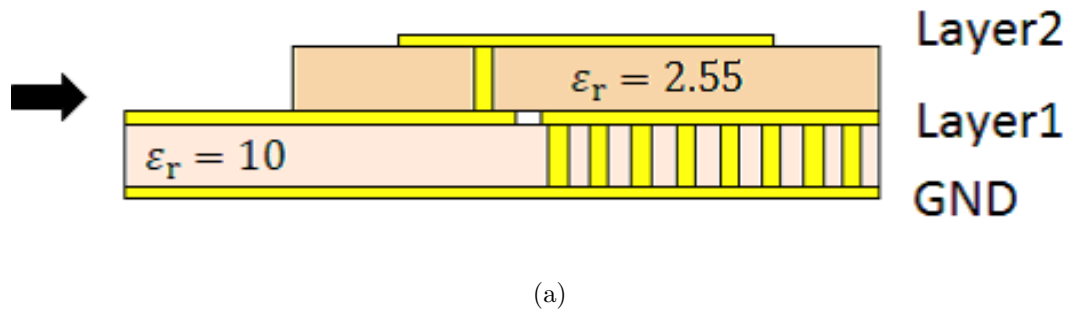


Figure 4.15: (a) antenna layer configuration, top view of (b) layer 1, and (c) layer 2 of simulation model.

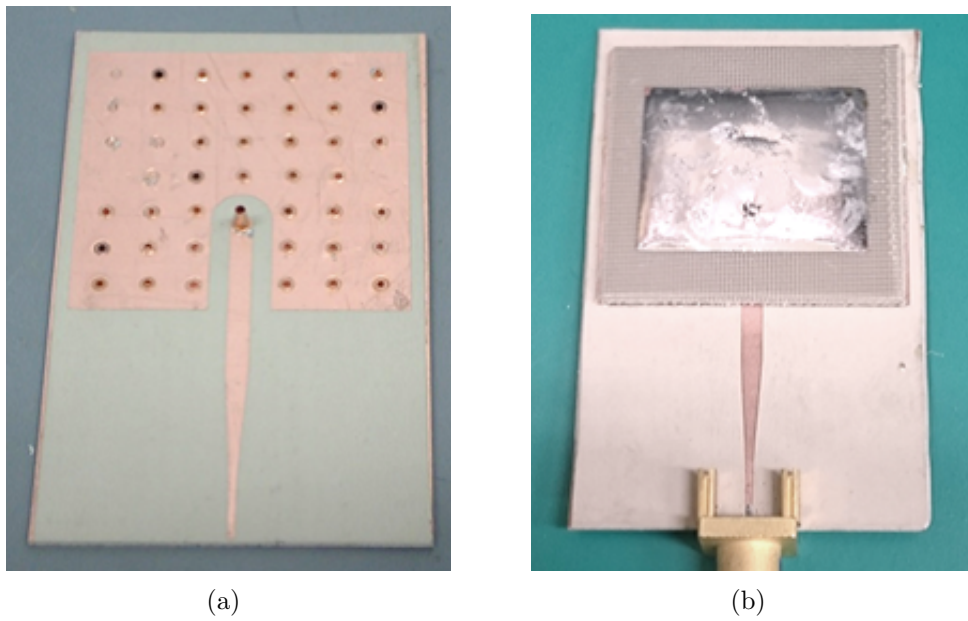


Figure 4.16: Fabricated single-antenna photographs (a) of the layer 1 circuit and (b) of each substrate combined.

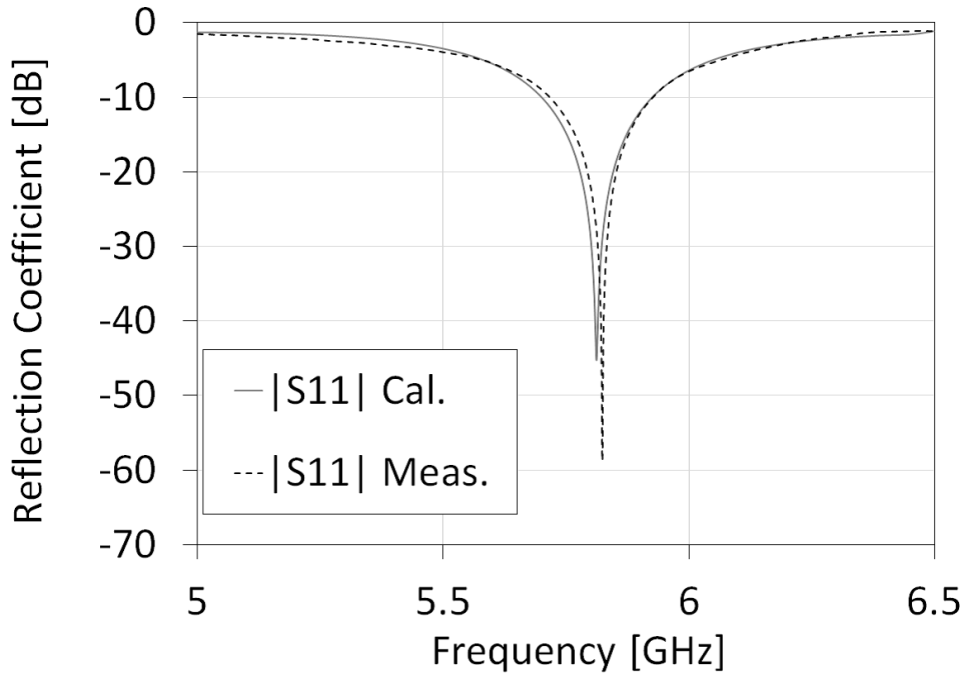


Figure 4.17: Reflection coefficient of the single antenna.

Figs. 4.16(a) and (b) show layers 1 and 2 of the fabricated single antenna. Here the antenna impedance is transformed by a tapered line for measurement in the $50\ \Omega$ system. The calculated loss of the tapered line is 0.07 dB at 5.8 GHz. In every single-antenna-measurement result, the loss of the tapered line is corrected. The reflection coefficients of the single antenna are reported in Fig. 4.17. According to Fig. 4.17, the reflection from the fabricated antenna is reduced at 5.8 GHz, agreeing with the simulation result.

4.4.2 Active Antenna Fabrication and Measurement

The GaN active antenna is fabricated on a Peltier unit, as shown in Fig. 4.18(a). Fig. 4.18(b) shows the active-antenna-circuit layout, by which the designed antenna and the GaN amplifier are integrated with the harmonic-tuning circuit. The GaN-HEMT chip is mounted on a copper plate, which is connected to the circuit through wire bonding. In this measurement, the GaN active antenna is operated with a 50 V drain voltage, a -2.5 V gate voltage, and a $20\ ^\circ\text{C}$ temperature of Peliter unit at a frequency of 5.8 GHz.

The radiation patterns of the active and single antennas under small-signal conditions are reported in Fig. 4.19, as is measured by the gain-comparison method

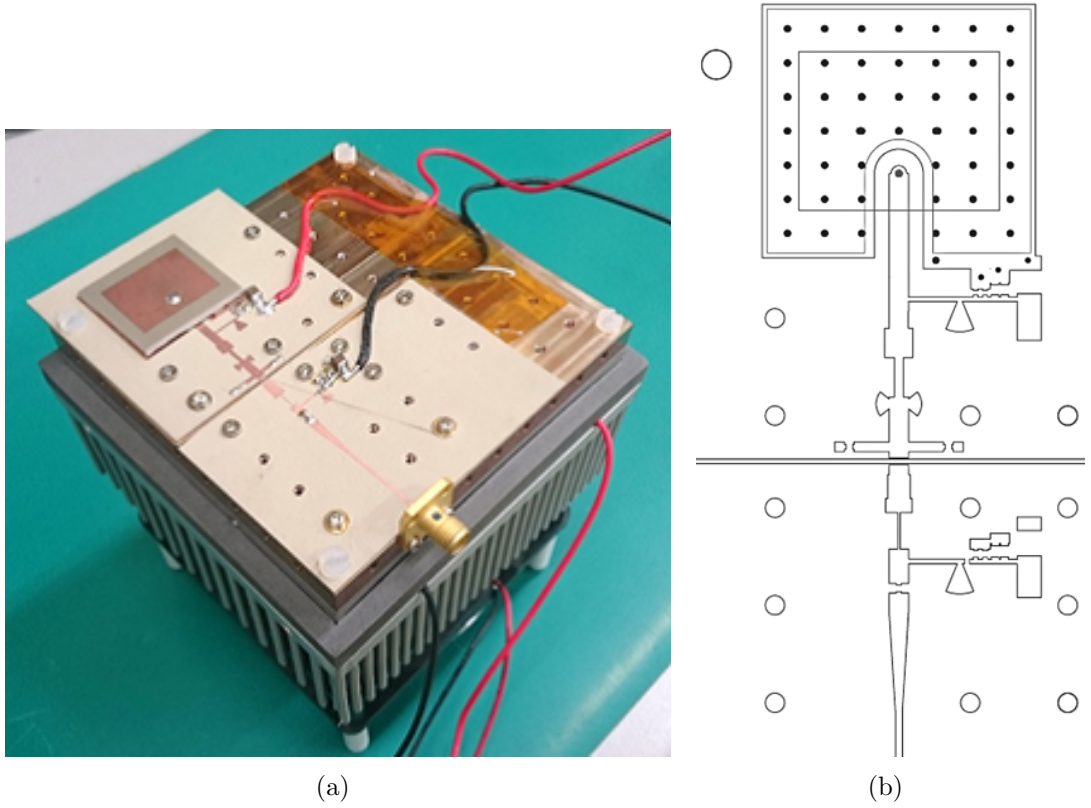


Figure 4.18: (a) Fabricated Active antenna on the Peltier unit and (b) layout of the active antenna.

with a standard dipole antenna in a far-field condition. In the far-field measurement system, the antenna gain is calculated by comparison method. In this method, the gain of AUT (antenna under test) is obtained by comparing to a standard antennas and calculated based on the following equation:

$$(G_{\text{AUT}})_{\text{dB}} = (G_{\text{ref}})_{\text{dB}} + (|S_{21}|_{\text{AUT}})_{\text{dB}} - (|S_{21}|_{\text{ref}})_{\text{dB}} + (M_{\text{AUT}})_{\text{dB}}, \quad (4.4)$$

where G_{AUT} is gain of AUT, G_{ref} is gain of a standard antenna, $|S_{21}|_{\text{AUT}}$ is transmission property with the AUT between Tx and Rx antenna, $|S_{21}|_{\text{ref}}$ is transmission property with a standard antenna. In this measurement, the dipole antenna was used for the standard antenna. In Fig. 4.19, the black line shows the single-antenna gain added to the small-signal-amplifier gain, as calculated in Fig. 4.9. According to Fig. 4.19, the radiation patterns of fabricated single and active antennas agree with the simulation results. In addition, the obtained antenna and active-antenna gains in the front direction are 6.40 dBi and 22.43 dBi, respectively. Hence, the 16.03 dB gain increase by the active antenna is confirmed by comparison to the single-antenna gain.

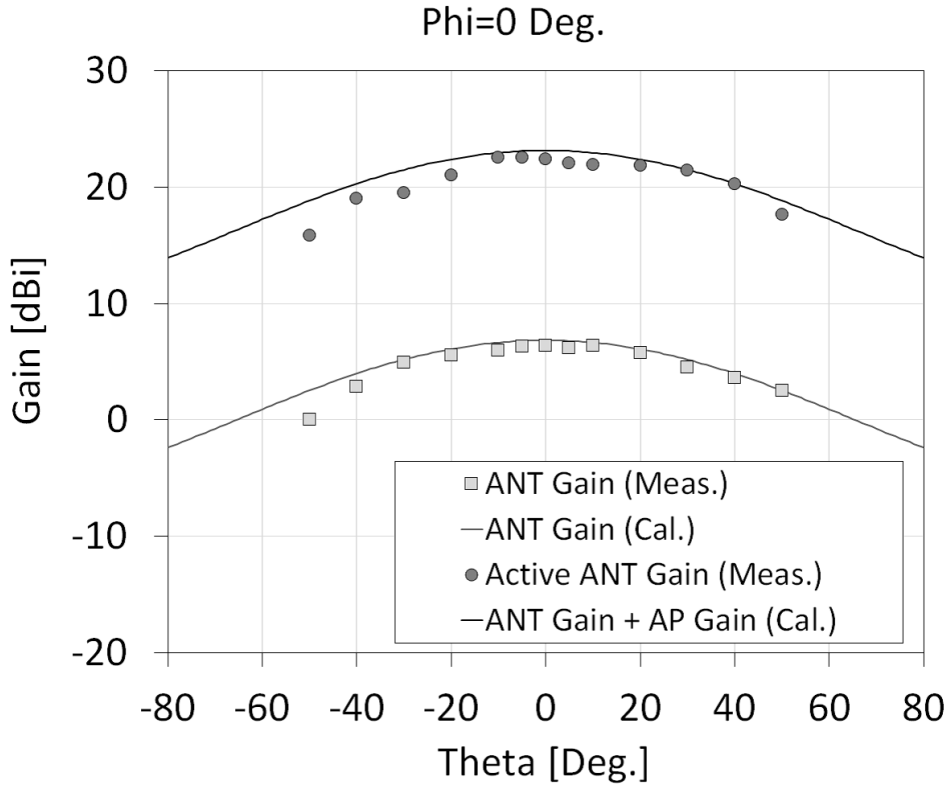


Figure 4.19: Radiation pattern of single antenna and active antenna.

The EIRP is an important characteristic for indicating the electric- and magnetic-field intensities of antenna outputs. It is defined by the equation

$$EIRP = P_{in}G_{ant}, \quad (4.5)$$

where P_{in} and G_{ant} show the input power to the antenna and the gain of the antenna, respectively. The measured large-signal characteristics of the single- and active-antenna gain and EIRP are reported in Figs. 4.20(a) and (b). The gain is measured by the gain-comparison method with a standard dipole antenna. According to Fig. 4.20(a), the active-antenna gain experiences compression in the input region above 20 dBm, which agrees with the single-stage-amplifier characteristics, as shown in Fig. 4.9. The EIRP, as shown in Fig. 4.20(b), also experiences saturation in this region. Moreover, the 44.61 dBm (28.91 W) maximum EIRP at an input of 25.70 dBm is obtained.

Finally, the validity of the active-antenna-fabrication technique is discussed. The output power P_{out} gain G_{ap} and PAE of the GaN amplifier in the active antenna are illustrated in Fig. 4.21, which is calculated based on the results in

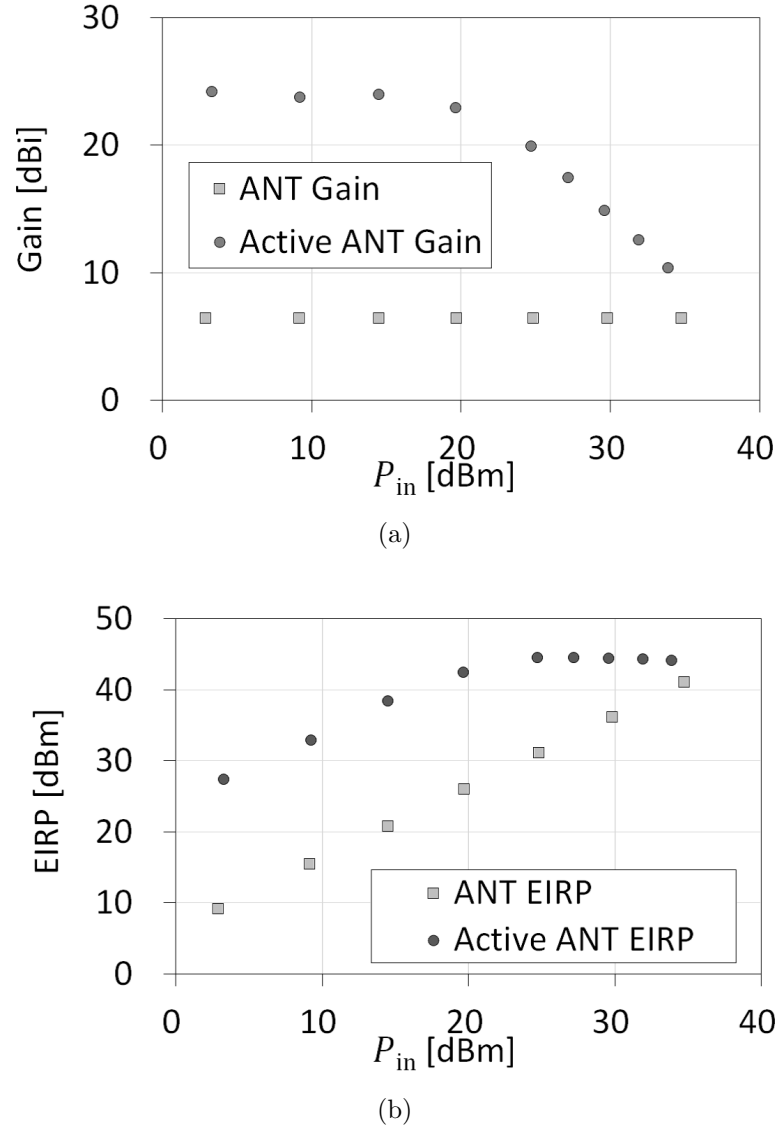


Figure 4.20: Large-signal characteristics of (a) gain and (b) EIRP of the single antenna and the active antenna in front direction.

Fig. 4.20 using the following equations:

$$(G_{ap})_{dB} = (G_{act})_{dBi} - (G_{ant})_{dBi}, \quad (4.6)$$

$$(P_{out})_{dBm} = (EIRP)_{dBm} - (G_{ant})_{dBi}, \quad (4.7)$$

where G_{act} shows the active-antenna gain and PAE is calculated based on P_{out} . In this calculation, the antenna gain, G_{ant} , is assumed to be 6.40 dBi, which is the measured result in Fig. 4.19. In Fig. 4.21, the continuous lines show the simulated result on ADS with the antenna-input impedance as the load. According to Fig. 4.21, the large-signal characteristics of the GaN amplifier in the active antenna agree with the simulated results. In addition, the maximum output powers and

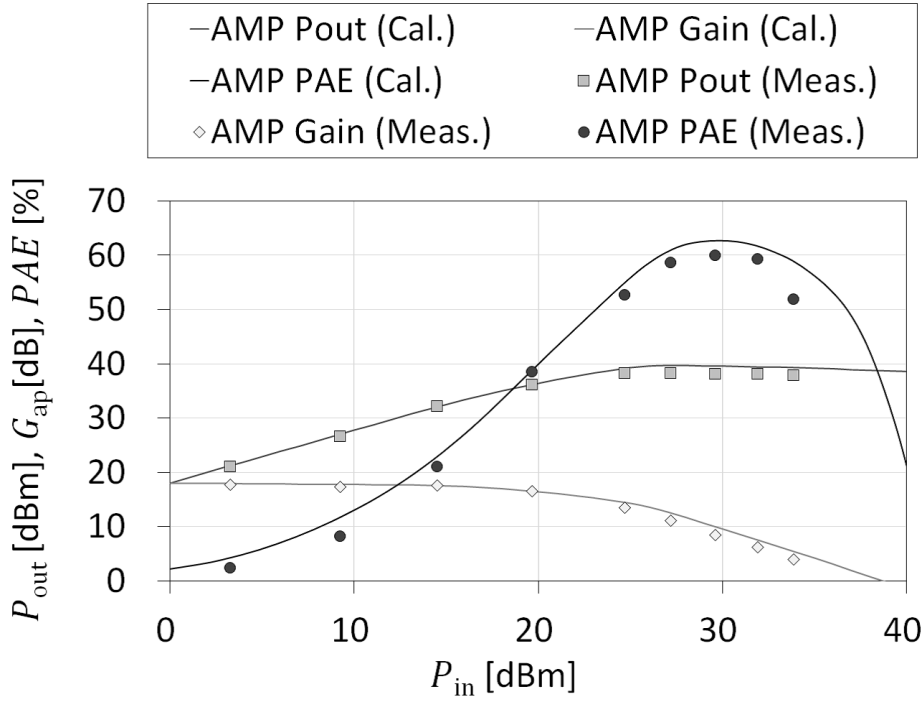


Figure 4.21: Extrapolated large-signal characteristics of the GaN amplifier in the active-antenna circuit.

PAEs of 38.21 dBm (6.62 W) at a 24.70 dBm input and 59.93 % at a 29.61 dBm input, respectively, are obtained, agreeing with the results of single-stage GaN-amplifier fabrication, as shown in Fig. 4.9. Under these input-power conditions, the amplifier gains of 13.50 dB and 8.46 dB are confirmed, yielding gain compressions of 4.27 dB and 9.31 dB, respectively. These results show that the GaN amplifier and antenna are losslessly integrated. Therefore, the design method for the GaN active antenna reported in this work is effective.

4.5 Summary

In this work, a GaN amplifier with a harmonic-tuning circuit and an antenna with a multilayer substrate were effectively integrated on a Peltier unit. First, the single-stage GaN amplifier was designed and fabricated. For this fabrication, the measured maximum output power of 37.95 dBm (6.24 W) at a 27.58 dBm input, and a PAE of 62.94 % at a 27.58 dBm input were confirmed for the large-signal condition. Comparing the PAE of the amplifier with a harmonic tuning circuit to that of the amplifier without this circuit, an 11.57 point PAE improvement

Table 4.4: Comparison of integrated GaN active antenna with the state of the art

Refs	Frequency	PAE	P_{out}	EIRP
[106]*	7.25 GHz	42 %	30.3 dBm	-
[114] *	1.99 GHz	82.1 %	22.08 dBm	-
[114]	2 GHz	>60 %	21.7 dBm	39.7 dBm **
This work *	5.8 GHz	62.94 %	37.95 dBm	-
This work	5.8 GHz	59.93 %	38.21 dBm	44.61 dBm

* specs of the single stage amplifier.

** ERP of the 2×2 active array antenna.

was confirmed. Next, an antenna on the multilayer substrate with a $30\ \Omega$ input impedance was designed and fabricated. An antenna gain of 6.40 dBi was measured in the front direction. Finally, the fabricated GaN amplifier and antenna were integrated to form an active antenna with a 22.43 dBi gain in the front direction at a small-signal condition. In addition, the 44.61 dBm (28.91-W) maximum EIRP was confirmed at a large-signal condition for the 25.70 dBm input. The measured output power and PAE characteristics of the GaN amplifier in the active antenna were 38.21 dBm at a 24.70 dBm input and 59.93 % at a 29.61 dBm input, respectively. Therefore, the GaN amplifier and the antenna were losslessly integrated, showing the viability of the active-antenna-fabrication method reported in this work. A comparison of the integrated GaN active antenna with the state of the art is provided in Table 2. According to Table 2, this work successfully achieved relatively lossless active-antenna integration compared to the conventional works.

Recently, antenna fabrication on monolithic-microwave-integrated circuits (MMICs) [117], GaN MMIC amplifiers [118]-[121], and a hybrid-semiconductor-integrated circuits (HySiCs) have all been reported [122]. In the future, a more lightweight and compact GaN active antenna will be achieved using MMIC and HySiC technologies. We expect that our work will be of interest for the design of future GaN active-antenna technologies.

Chapter 5

Beam Forming for Sidelobe Reduction with Active Antenna

5.1 Introduction

Recently, a lot of microwave transmission system are proposed such as an ultra-long-range MPT and wireless-power suppling for sensor networks, electric vehicles (EVs), and mobile phones [123], [124]. These researches mostly base a microwave beam forming by array-antenna techniques. The resent advance of high power microwave transmission is inseparable relationship with antenna and RF power semiconductor technologies.

The resent long-range microwave transmission systems assumes to use the array antenna techniques for controlling microwave beam. In the microwave transmission system, the beam control techniques are absolutely imperative for improving the transmission efficiency and the sidelobe reduction. Ishikawa et al. reported a flat-topped beam forming experiment for microwave power transfer system to a vehicle roof in 2015 [125]. They created a flat-topped beam by a phase tuning for each antenna in [125].

On the other hand, the sidelobe reduction techniques has studied based on passive antenna design in conventional. In general, the transmission power is larger than the microwave power for communication. The high power microwave transmission system must not prevent the other traditional wireless communications and affect the immediate environment. Hence, it is desirable to reduce the

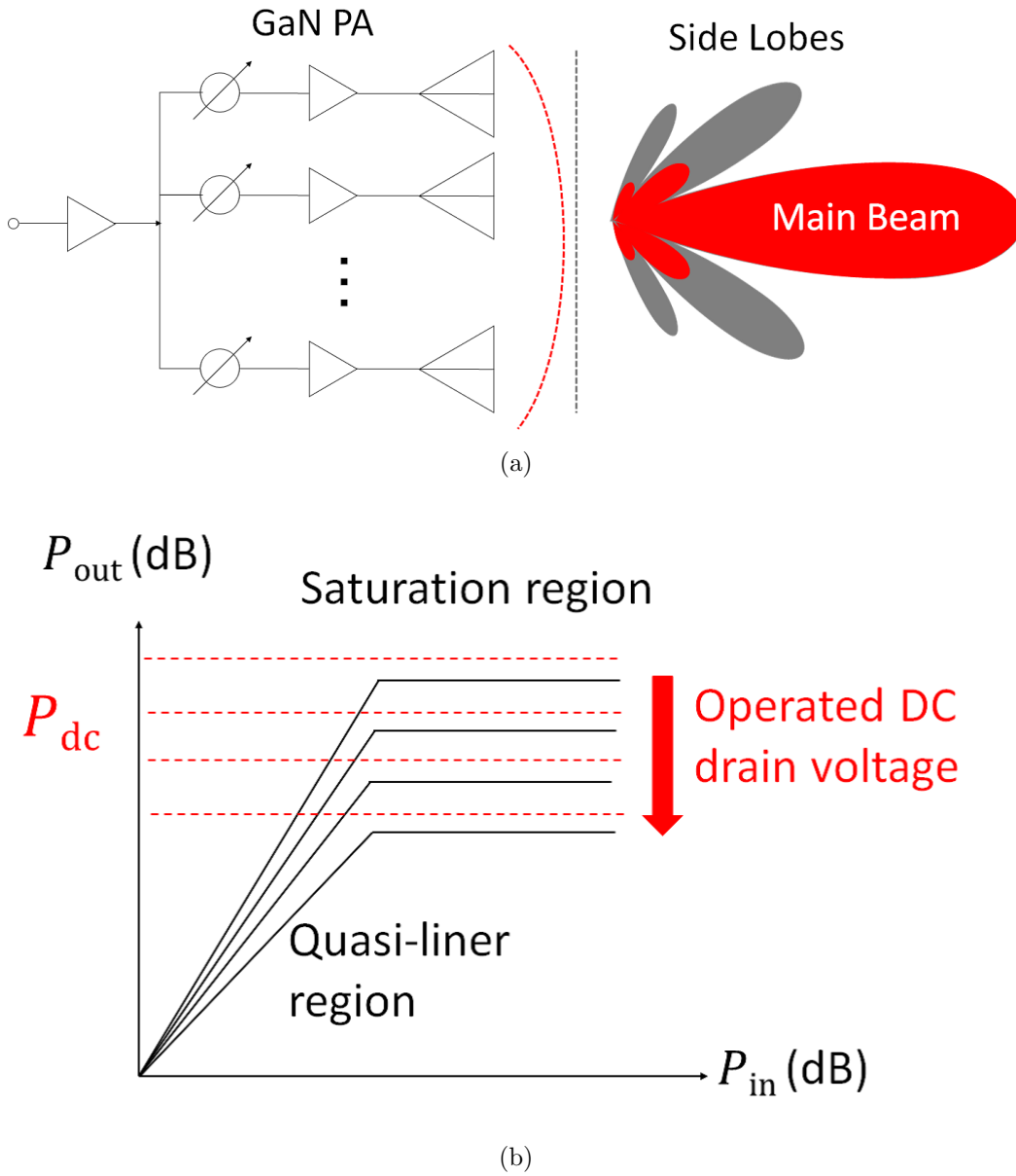


Figure 5.1: Low-sidelobe active-array antenna: (a) active array antenna configuration and (b) amplifier input-output characteristics for various operating DC drain voltages.

sidelobe derived from array antenna. The low sidelobe antenna techniques with a phase and elements space controlling are reported [126]-[131]. In addition, the sidelobe reduction by the amplitude control has been studied [132]-[133]. For instance, Chen et al. proposed a low sidelobe array antenna with a filtering microstrip antenna in 2017 [134]. In these conventional design, the sidelobe is mainly reduced by an un-uniform dividing techniques.

It is difficult to control the sidelobe level in the conventional circuit-approach

Table 5.1: Calculated load-pull impedances at the fundamental and second harmonic frequencies with 30 dBm RF input

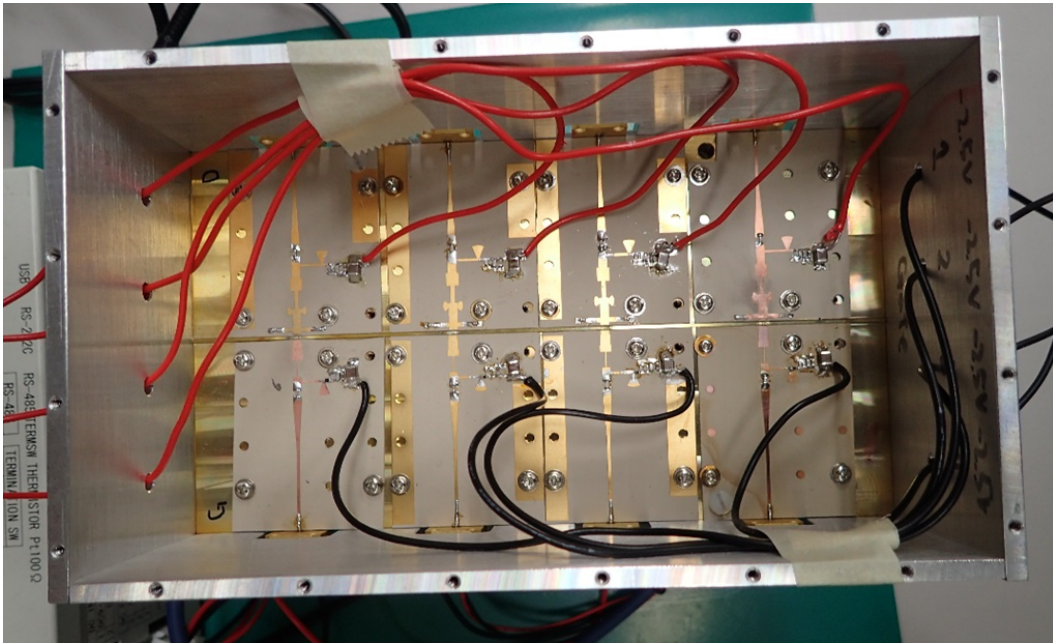
	Impedances (Ω)	PAE (%)	P_{out} (dBm)
Z_{L1}	$25.6 + j53.3$		
Z_{L2}	$0 + j36.7$	67.1	41.2
Z_S	$2.9 + j11.6$		

sidelobe reduction. However, the tunable sidelobe control according to the situation is required for the developed microwave transmission systems such as MPT toward EVs, mobile phones, and wireless sensors is expected. In the long-range and high-power microwave transmission system, the active array antenna is mainly used. In this paper, the active array antenna as shown in Fig. 5.1(a) is assumed. This active array antenna consists of the antenna elements with GaN power amplifiers. In the conventional study, the sidelobe reduction by an effectual amplitude distribution on the surface of antenna is reported by Taylor in 1955 [22]. In this work, the sidelobe reduction is executed based on a GaN active antenna techniques. The input-output characteristics of an amplifier is reported in Fig. 5.1(b). In general, it is possible to control the saturated output power of the amplifier by the operated drain DC power levels. In this work, the controlled saturated output power of GaN amplifiers is used for the sidelobe reduction. The advantage of the sidelobe-reduction method based on an active antenna approach is that it is possible to tune the sidelobe by operated DC power of the amplifier.

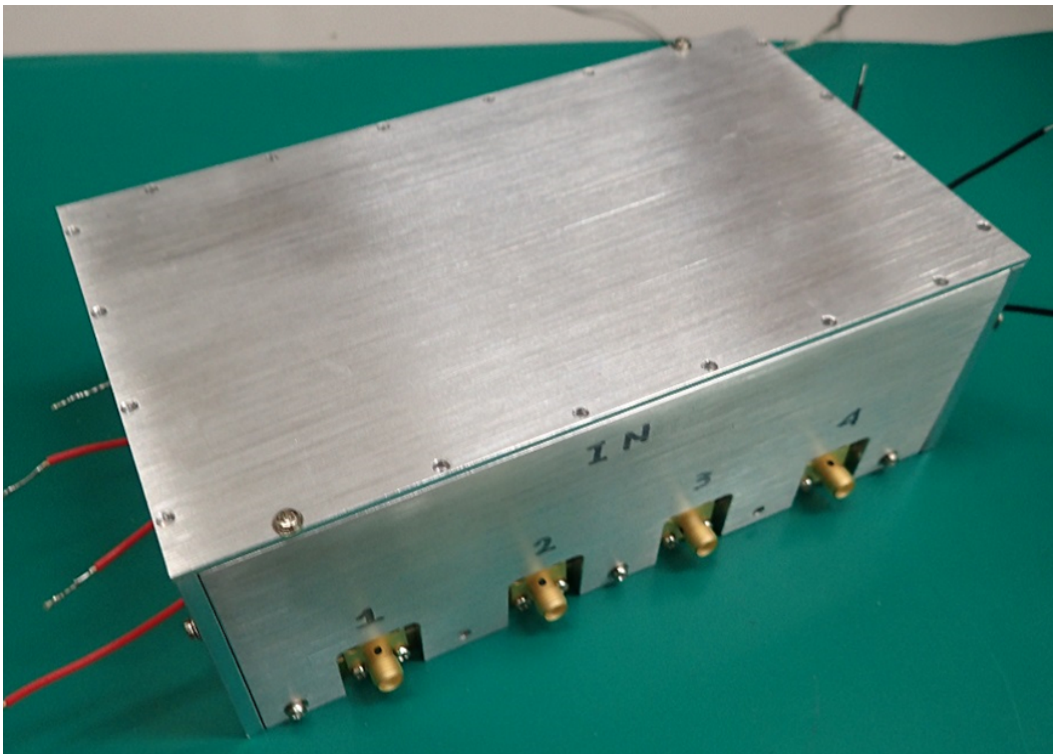
In this work, the tunable sidelobe reduction techniques with a GaN active array antenna is proposed. First of all, a single stage GaN amplifier is designed and fabricated. Secondly, 4 way GaN active array antenna is designed and fabricated. Next, experiment results of low sidelobe active antenna is reported. Finally, this paper is concluded.

5.1.1 4 Way GaN Amplifier Fabrication

In this study, the sidelobe reduction is based on a 4-way GaN active-array antenna. Hence, a 4-way amplifier is designed and fabricated, as shown in Fig. 5.2. The fabricated amplifier circuits are mounted to a copper plate and packed in an aluminum case of 90 mm width, 162 mm length, and 58 mm height.

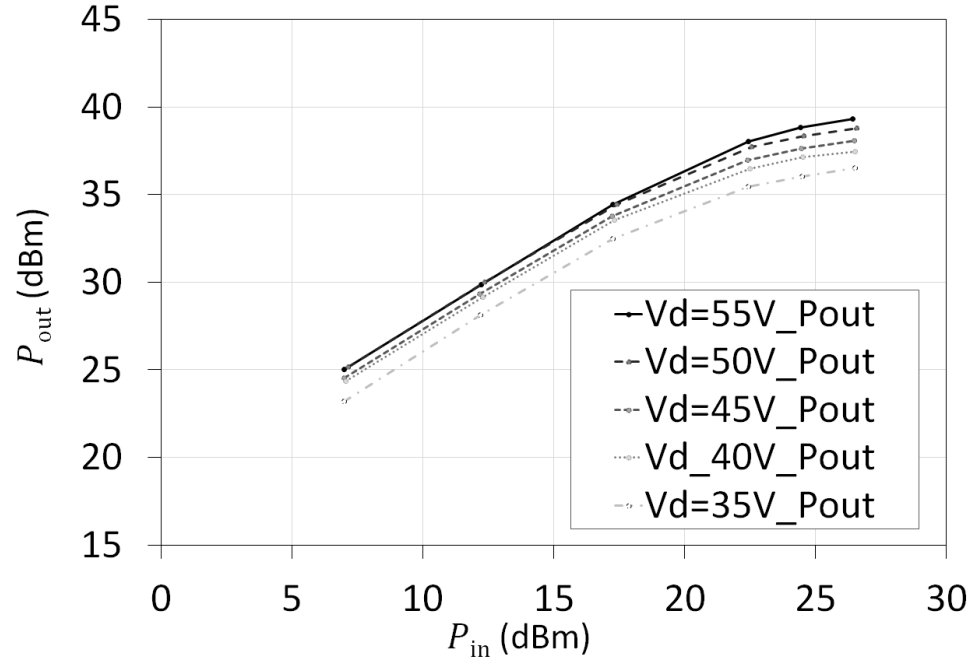


(a)

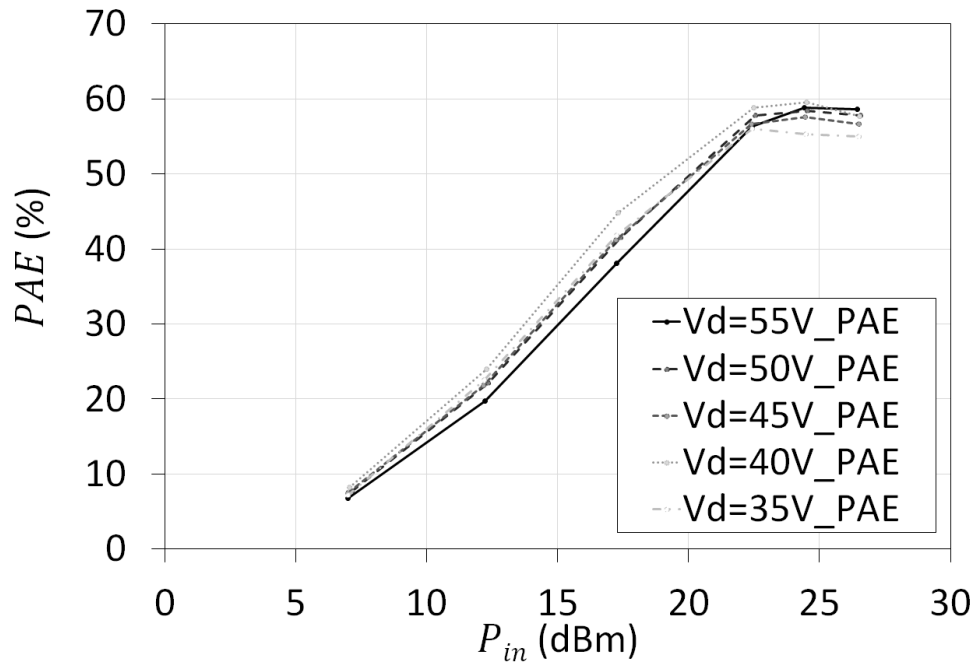


(b)

Figure 5.2: Fabricated 4 way GaN amplifier.



(a)



(b)

Figure 5.3: Measured input output characteristics at operating the DC drain voltage between 35 V and 55 V: (a) measured output power P_{out} and (b) measured PAE.

These amplifier circuits are fabricated on a dielectric substrate of Arlon AD1000 (with $\epsilon_r = 10.2$, $\tan \delta = 0.0023$, and thickness of 0.8 mm), which uses circuit parameters based on the amplifier designed in previous section. The load- and source-side input impedances of the designed circuit are $Z_{L1} = 28.2 + j53.6 \Omega$, $Z_{L2} = 5.8 + j35.4 \Omega$, and $Z_S = 6.5 + j10.1 \Omega$. The GaN HEMT chip is mounted on a copper plate which is connected to the circuits by wires. The circuit substrates are also mounted on the copper plate.

The measured input-output characteristics of the one amplifier are shown in Fig. 5.3. In Fig. 5.3(a) and Fig. 5.3(b), the effects of the drain voltage levels on the output power, P_{out} , and on the PAE are illustrated. According to Fig. 5.3(a), the maximum output power is 39.3 dBm when $V_d = 40$ V, which decreases to 36.5 dBm when the drain voltage level is reduced to $V_d = 35$ V. In contrast, the maximum PAE peaks at 59.6 % when $V_d = 40$ V, $P_{\text{in}} = 24.5$ dBm. In each drain voltage condition tested, maximum PAEs over 50 % are confirmed. According to the results, the amplifier output power can be controlled between 36.5 dBm and 39.3 dBm with minimal reduction in the PAE, which facilitates sidelobe reduction in the GaN active-array antenna.

The measured maximum PAE and P_{out} under difference applied drain voltages are illustrated in Fig. 5.4. In these measurements, the maximum input power is 25.36 dBm. The measured maximum output power increases as the drain voltage increase. On the other hand, the measured PAE decreases with increasing drain voltage, reaching a minimum of 50.1 % when the drain voltage is 55 V. However, the PAE remains within a range of 50 % to 60 % for all drain voltages tested.

The factors affecting the input output characteristics, such as the maximum PAE and output power, are measured. However, the output phase of the amplifier may be shifted for each drain voltage. Therefore, the GaN amplifier is measured using a system comprising of a vector network analyzer, a driver amplifier, and other RF components as shown in Fig. 5.5. Using this system, the S-parameters of the amplifier, including the isolators, the 40 dB driver amplifier, the directional coupler, and the 30 dB attenuator, are measured. The amplitude and phase compositions of the S-parameters include the independent characteristics of these components. However, the effect of the drain voltage on the fabricated GaN amplifier, which hardly depends on the component characteristics, is also considered using this measurement system.

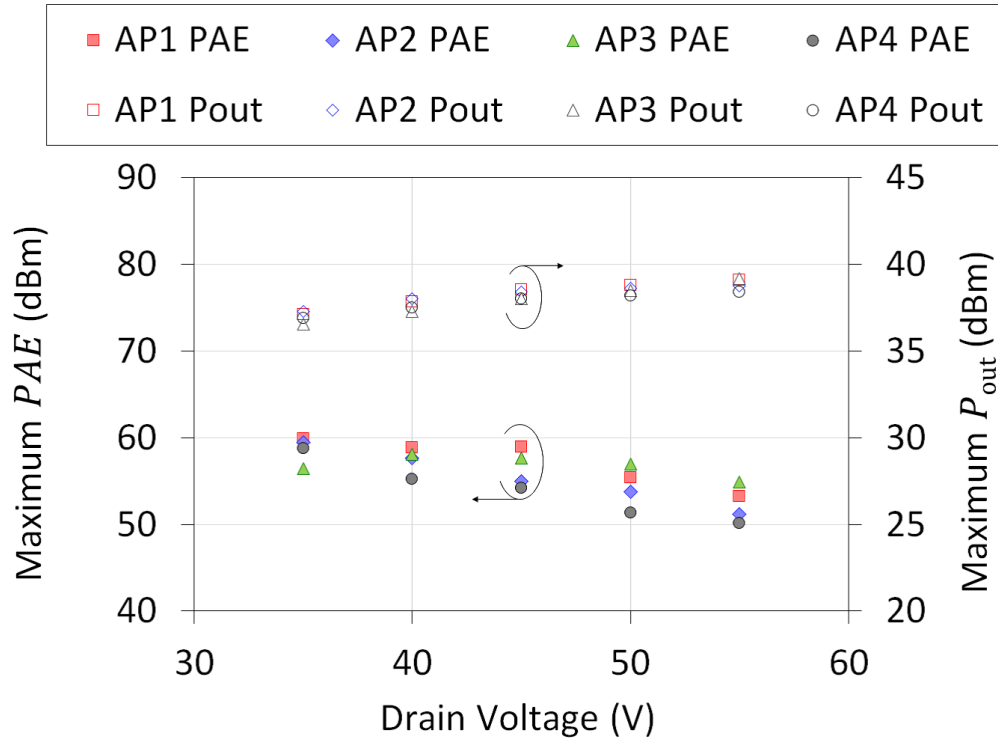
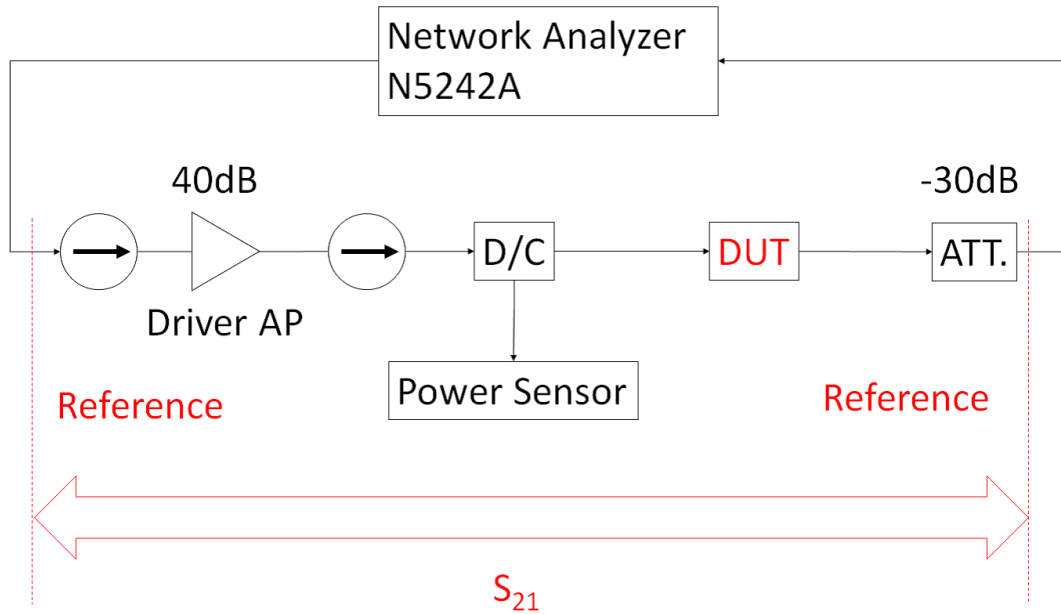
Figure 5.4: Measured maximum PAE and P_{out} as a function of the drain voltage.

Figure 5.5: Measurement system for S parameters of GaN amplifier under high-power conditions using a network analyzer.

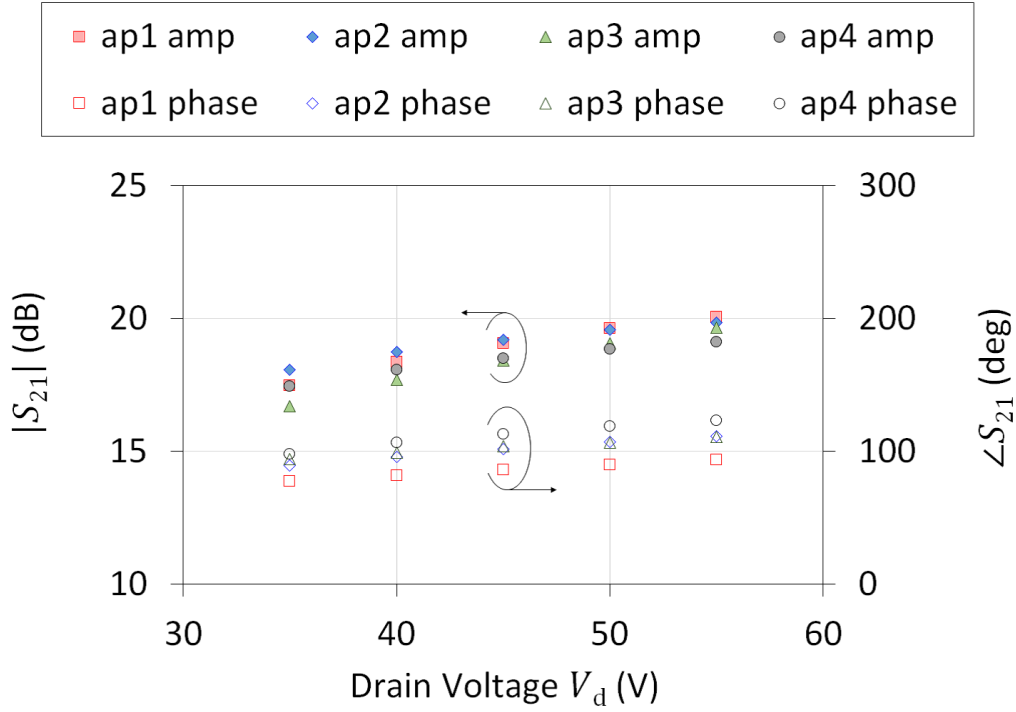


Figure 5.6: Measured GaN amplifier S-parameters associated with different drain voltages.

5.2 Low Sidelobe Antenna Design

The 1×4 array antenna with a circular patch antenna, shown in Fig. 5.7(a), is designed and fabricated. A high-frequency 3D EM field simulator (CST Microwave Studio) is used in this design. This antenna is fabricated on a dielectric Nippon Piller Packing NPC-F260A ($\epsilon_r = 2.55$, $\tan \delta = 0.0018$, and thickness of 1.6 mm) substrate. The diameter of the patch antenna is 17.4 mm and the feed pin is shifted 2.7 mm from the center. The calculated and measured values of the reflection coefficient $|S_{11}|$ s are illustrated in Fig. 5.7(b). Both theoretical and experimental results show that $|S_{11}|$ is reduced to less than -10 dB at 5.8 GHz.

The antenna radiation patterns were calculated based on the measured amplifier output power and phase; the results are illustrated in Fig. 5.8. In the calculation, the port phase and amplitude of the edge of the array antenna are fixed to the values associated with a 35 V drain voltage for the amplifier based on the measured correlation shown in Fig. 5.6. On the other hand, the center antennas are tuned between the amplitude and phase associated with the 35 V and 55 V drain voltage. According to Fig. 5.8, the second sidelobe level of 6.6 dB is reduced by tuning the port amplitude and phase conditions. The main beam

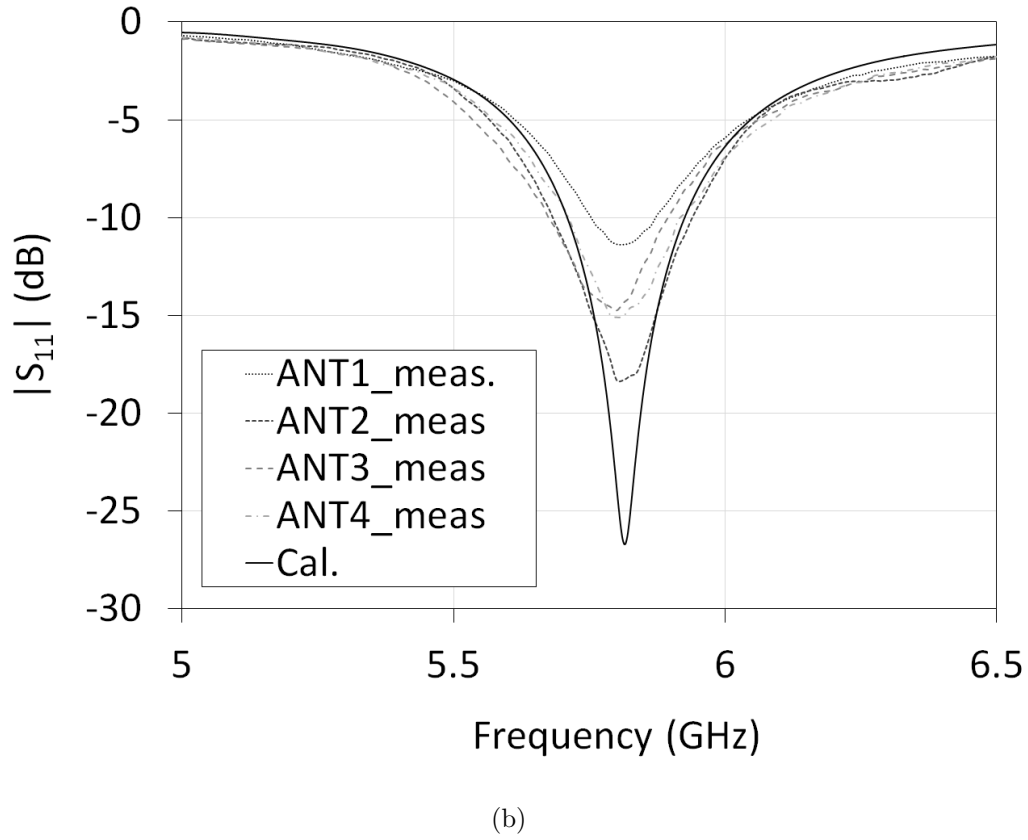
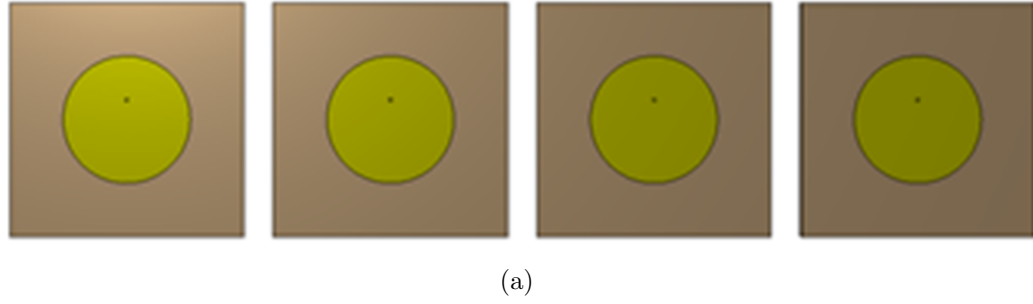


Figure 5.7: Designed 1×4 array antenna: (a) antenna layout and (b) measured and calculated values of the antenna reflection coefficient $|S_{11}|$ s .

amplitude is hardly reduced. The first null point and sidelobe vanish when the 55 V conditions are applied to the center antennas because of the phase difference of the amplifiers for each drain voltage condition. According to Fig. 5.8, removing the phase difference of the amplifiers causes the null point and sidelobe to unfold. Therefore, it is concluded that the removal of the null point and sidelobe vanish is caused by the phase difference of the amplifiers.

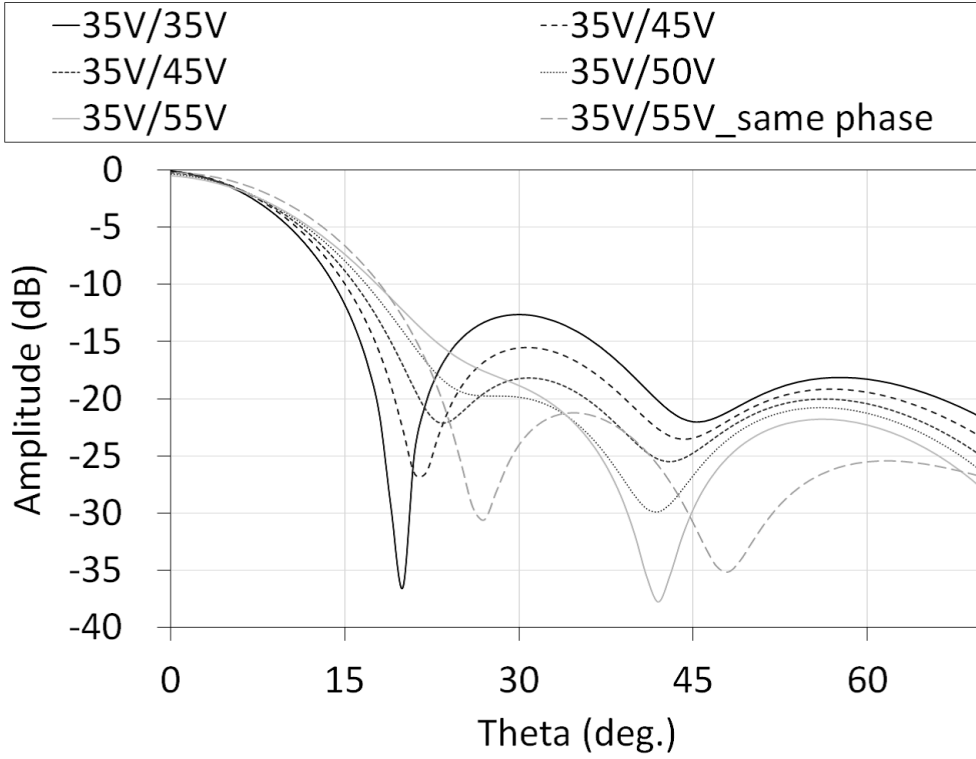


Figure 5.8: Calculated antenna radiation patterns based on measured 4-way amplifier output characteristics for each drain voltage condition.

5.3 Characterization of Low-Sidelobe Active Antenna

The fabricated 4-way amplifier and antenna are combined, as shown in Fig. 5.9(a). The active antenna component consists of the GaN amplifier, the antennas, and the power divider as shown in Fig. 5.9(b). The GaN amplifier is placed on a Peltier cooler (VICS LVPU-70), which serves to precisely control the surface temperature; the cooler can absorb up to 70 W, which is sufficient to eliminate the thermal effects on the GaN amplifier.

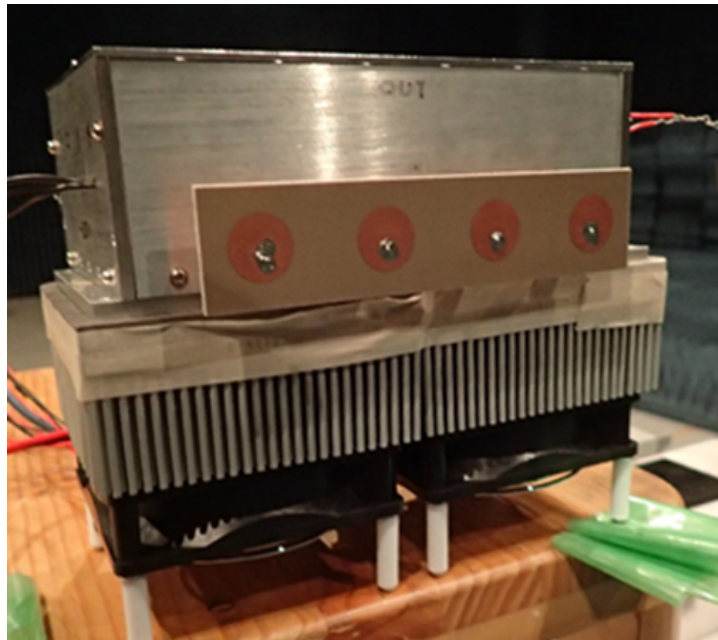
The measured main beam and sidelobe levels for various drain voltage conditions are illustrated in Fig. 5.10. In this measurement, the drain voltage of the center amplifiers is tuned between 35 V and 55 V and the edge amplifiers are fixed to 35 V. According to Fig. 5.10, the measured main beam levels agree with the calculated results and sidelobe levels decrease as the drain voltage of the center amplifiers increase. In addition, by tuning the center-amplifier drain voltage from 35 V to 55 V, the sidelobe level of 4.52 dB is reduced. In this experiment, the

ability to reduce the sidelobe level of more than 50 % by tuning the amplifier drain voltage is demonstrated. The difference between sidelobe levels of calculation and measurement is expanded to 2.63 dB at driving the center two amplifiers on 50 V drain voltage condition. A special thing about the difference is the difference of gradients, which indicates the amplifiers make mischief. Thus, we predict the cause of the difference based on output phase from the amplifiers. In Fig. 5.11, the radiation patterns with output phase error of the entire edge of amplifiers is recalculated. In this calculation, the output phase errors of 0 deg., -5 deg., and -10 deg. are assumed. According to Fig. 5.11, 2.03 dB of sidelobe level is increased by -10 deg. output phase error of amplifier. In the S-parameter measurement of 4 way antenna as shown in Fig. 5.7(b), the fabrication error is confirmed. The antenna fabrication error potentially affects the properties of amplifier. In Fig. 5.12, the recalculated output phase of amplifier schematics in RF circuit simulator at terminating with each antenna S-parameters included antenna fabrication error. In this calculation, we confirmed the output phase difference of 5.87 deg at 55 V drain voltage condition for each antenna load condition. In addition, the gradual difference for each antenna load condition is confirmed. This result is calculated on the simulation schematic without the amplifier fabrication error. The amplifier fabrication error probably affects the output phase, which expands the difference as shown in Fig. 5.12. We confirm that difference between simulation and measurement as shown in Fig. 5.10 is expected by output phase error based on fabrication error of amplifier and antenna.

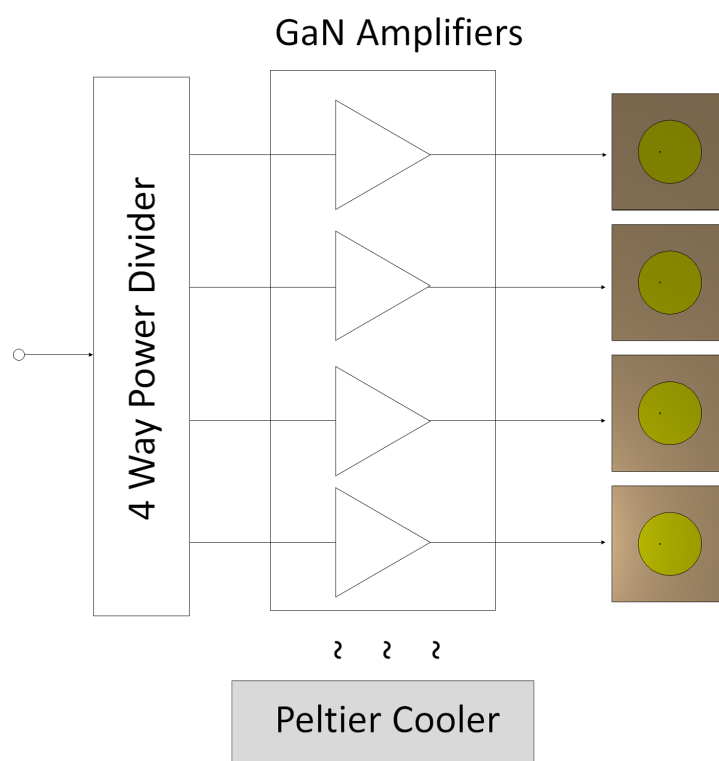
Furthermore, the equivalent isotropically radiated power (EIRP) for the front direction of the antenna is measured, which is important to characterize the performance of the active antenna. The EIRP is defined by the following equation,

$$EIRP = P_{in} \times G_{ant}, \quad (5.1)$$

where P_{in} is the input power to the antenna under test and G_{ant} is the isotropic gain of antenna. In this study, G_{ant} is measured by the gain-transfer method with a gain-standard dipole antenna. The measured EIRP of the GaN active antenna and dipole antenna are shown in Fig. 5.13. The measured maximum EIRP is 55.6 dBm at the 35 V/55 V drain voltage condition; this falls to 52.1 dBm in the 35 V/35 V condition because the total output power from the GaN amplifiers is reduced.



(a)



(b)

Figure 5.9: Fabricated 1×4 GaN active array antenna: (a) image and (b) configuration of GaN active-array antenna.

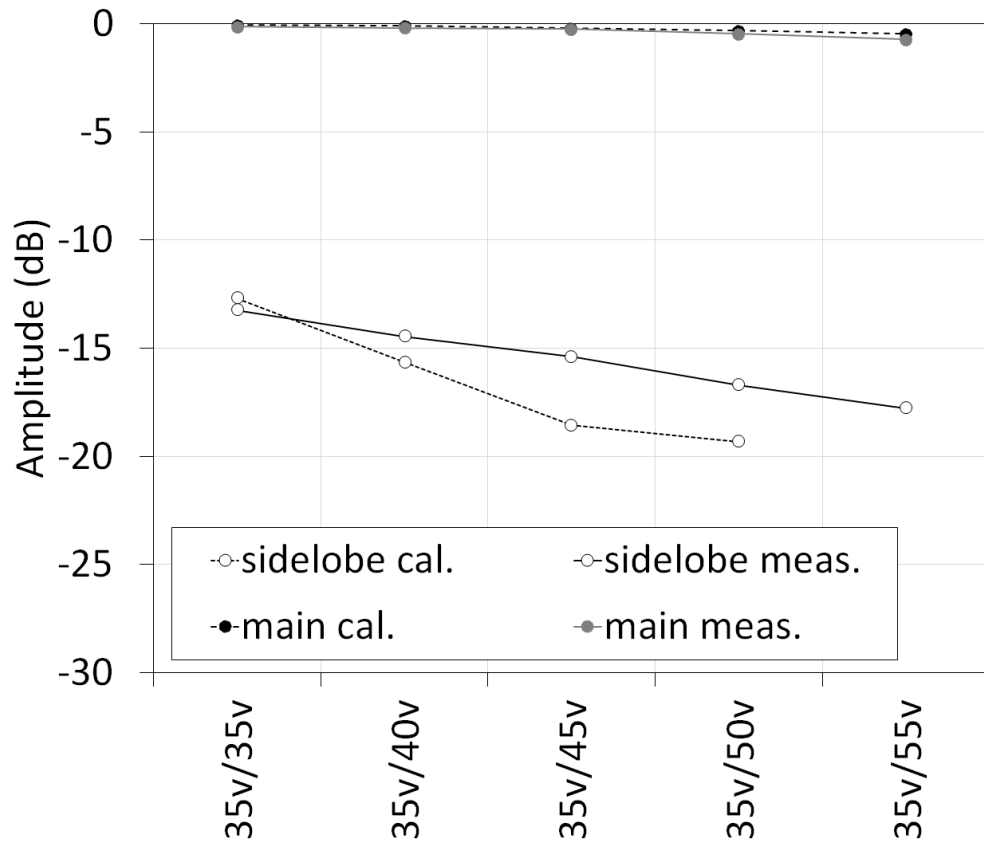


Figure 5.10: Measured and calculated main beam and sidelobe amplitude for each drain voltage condition

5.4 Sidelobe Level Calculation based on Chebyshev Distribution

In the large-scale array antenna system, the distributions such as Chebyshev, Taylor, and Gauss are used for sidelobe reduction. We investigate the relationship between sidelobe level and amplifier efficiency. In this investigation, the radiation pattern of 10 way liner array antenna based on circular patch is calculated by CST Microwave Studio as shown in Fig. 5.14(a). In this calculation, we investigate the three types of chebyshev distribution which provide -20 dB, -30 dB, and -40 dB sidelobe levels, respectively. The port amplitudes as shown in Fig. 5.14(b) for each antenna element in the three types of distribution are provided. According to Fig. 5.14(b), in the -40 dB chebyshev distribution, -8.86 dB port amplitude of the entire edge of 10 way amplifiers is confirmed at comparing to the amplitude of center antenna element, which probably reduces the amplifier efficiency. Figure 5.15 shows the drain voltage properties of the calculated normalized output

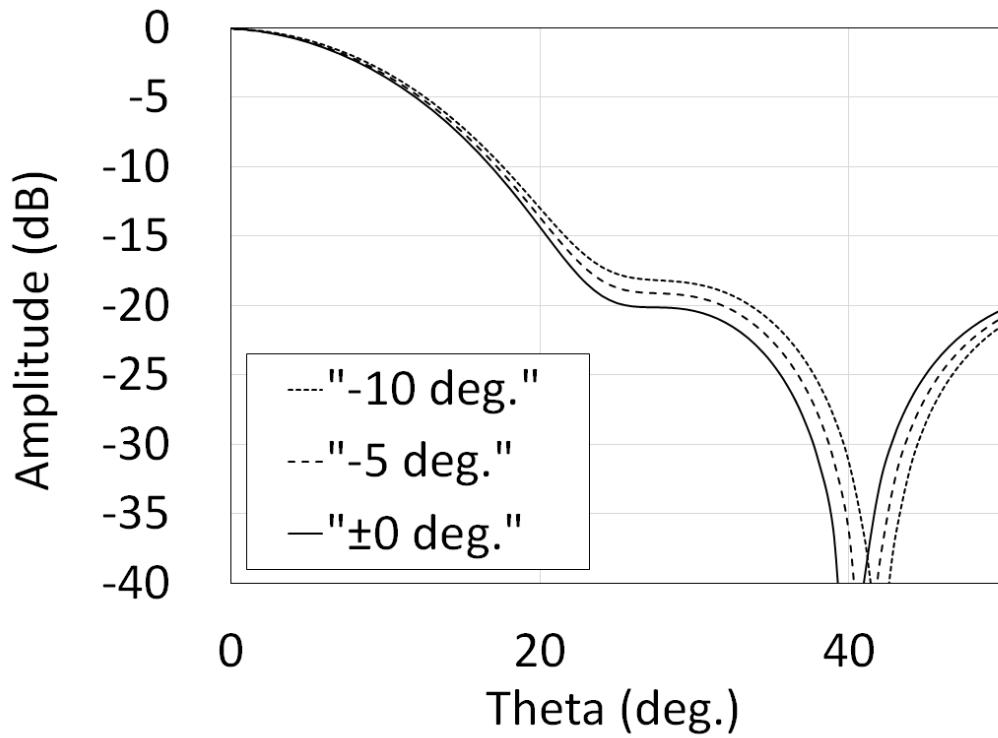


Figure 5.11: Calculated radiation patterns affected by output phase error of the entire edge of 4 way amplifiers

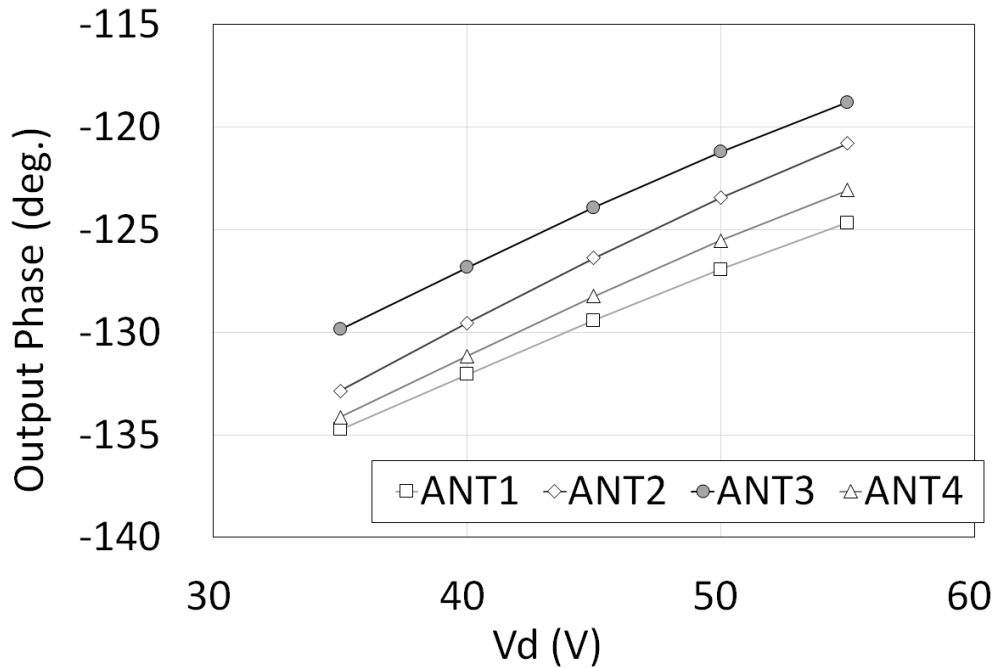


Figure 5.12: Recalculated output phase of amplifier at terminating with each antenna S-parameters included fabrication error.

power by a value at 55 V condition and PAE of GaN amplifier designed in previous section. According to Fig. 5.15, the PAE decrease of more than 18 points is

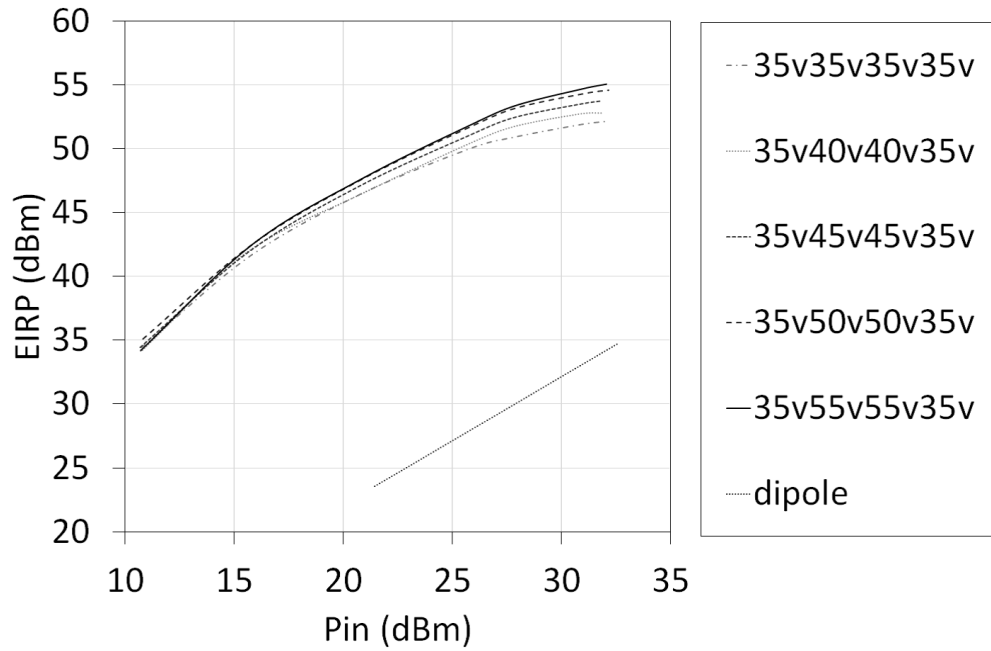
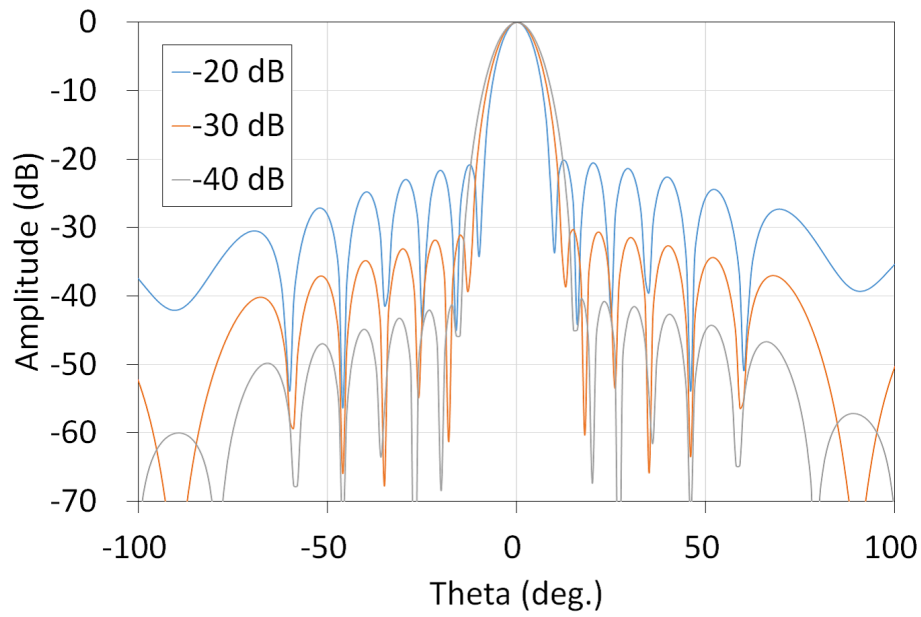


Figure 5.13: Measured front-direction EIRP of GaN active array antenna

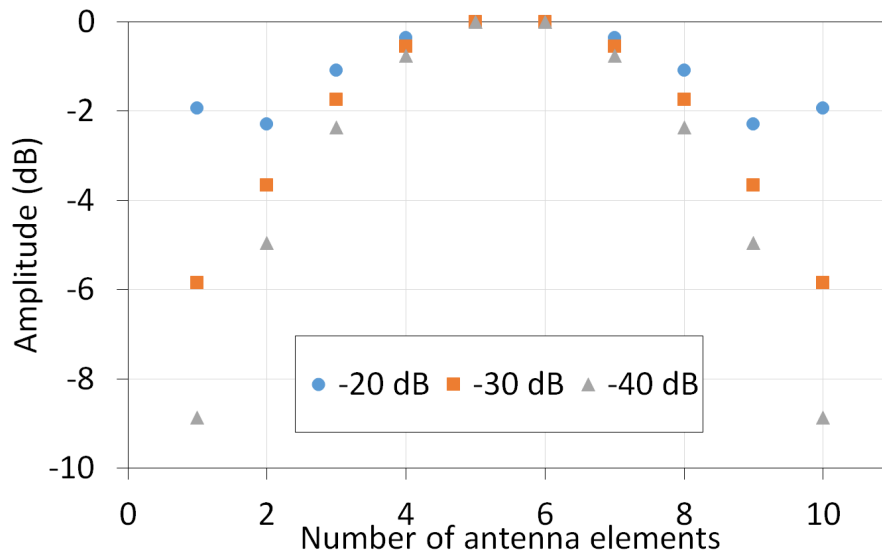
confirmed for -8.86 dB output power reduction. Assuming to operate the amplifier within 10 points range of PAE reduction, up to 7 dB of output power reduction is available, which require the drain voltage above 20 V. Therefore, the -30 dB chebyshev distribution requiring up to 6 dB output power reduction for the entire edge of 10 way amplifier is available within 10 points range of PAE reduction.

5.5 Summary

This study proposes a sidelobe reduction method based on a GaN active array antenna. In this method, the sidelobe level is tuned by adjusting the operating DC drain voltage of the amplifiers, which enables the situational switching of the radiation pattern. In this study, single-stage GaN amplifiers designed in the previous section are evaluated. The measured maximum output power is 39.3 dBm and the PAE is 59.6 % at a 55 V DC drain voltage; when the drain voltage is decreased to 35 V, the output power decreases to 36.5 dBm. The amplifier output difference is proactively leveraged to realize sidelobe reduction. To do this, a 4-way GaN amplifier is fabricated. The characteristics of the amplifiers are measured by a network analyzer and further used to calculate the antenna radiation pattern. The 4-way GaN active-array antenna was tested, with the



(a)



(b)

Figure 5.14: (a) Antenna radiation patterns based on chebyshev distribution and (b) amplitude of each antenna port.

amplifier DC drain voltage varying from 35 V to 55 V. According to the results, the 4.5 dB sidelobe level is reduced by tuning the operating center-amplifier drain voltage from 35 V to 55 V while the edge amplifiers are fixed to the 35 V condition. By this approach, it is possible to reduce the sidelobe level of more than 50 % by tuning the amplifier drain voltage. The maximum EIRP for the front direction is measured as 55.6 dBm when the 35 V conditions are set for the edge amplifiers and

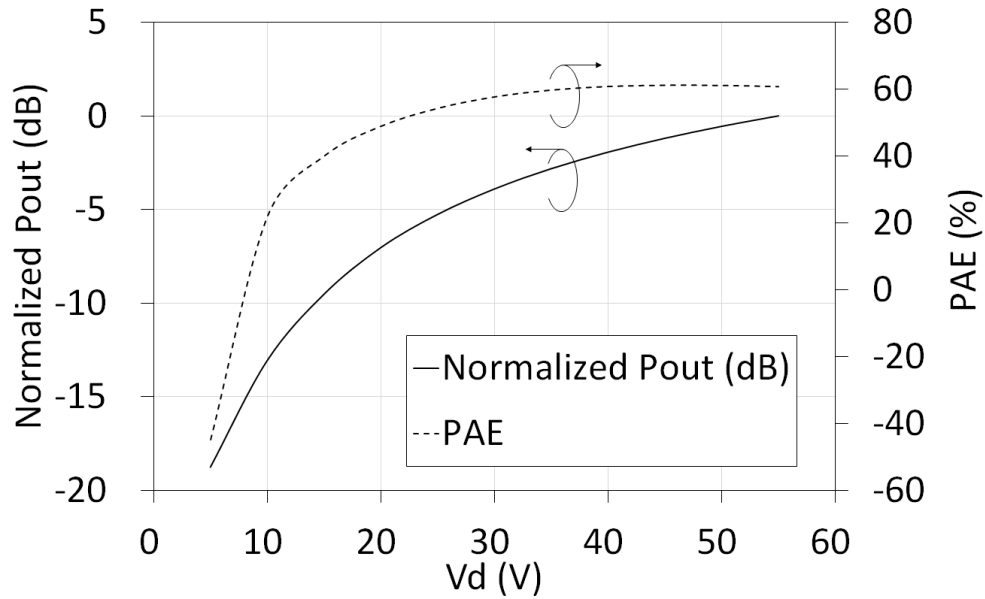


Figure 5.15: Drain voltage properties of the calculated normalized output power and PAE of GaN amplifier.

the 55 V conditions are provided for the center amplifiers; this value is reduced to 52.1 dBm upon decreasing the center amplifier DC drain voltage condition to 35 V. Finally, the antenna radiation patterns based on chebyshev distribution are calculated. In this calculation, the amplitude of each antenna port for -20 dB, -30 dB, and -40 dB chebyshev distribution and the decrease ratio of PAE for the port amplitude decrease.

In this study, tunable sidelobe reduction based on the GaN active antenna-technique has been demonstrated effectively. We expect that the proposed technique will contribute to the broad realization and implementation of MPT systems. Sidelobes can cause electromagnetic interference, affect the human body, and cause a myriad of problems. Thus, this technique is useful for the situational sidelobe reduction necessary for a variety of future long-range MPT system for EVs, mobile phones, and sensor networks.

Chapter 6

Concluding Remarks

6.1 Summary and Conclusion

The aperture antenna and array antenna systems are necessary for high power microwave transmission such as space satellite communication, microwave power supplying, and other high power RF applications. The GaN amplifiers and RF antenna technologies have been studied in different technical fields. In this work, the design method of GaN amplifier and RF antenna technologies are integrated for high efficiency and effective microwave transmission. In the each chapter, the high power GaN amplifier for aperture antenna system, high efficiency GaN active antenna, and sidelobe reduction based on the designed GaN active array antenna are demonstrated. The summary of this work is hereunder illustrated.

In Chapter 2, the RF antenna and GaN amplifier technologies for high power microwave transmission are introduced. The comparison of S- to X-band GaN amplifier with the state-of-the-art is also concluded. In addition, the basic AB class GaN amplifiers are design and fabricated. The fabricated GaN amplifier with 43.1 dBm output power, 44.5 % PAE, and 10.3 dB small-signal gain is demonstrated. Next, the carbon graphite fixture for light weight amplifier is fabricated, which contribute to reduce 71 point weight of fixture at comparing copper one. These fabrication method of GaN amplifier becomes the base for high power and high efficiency amplifier design, which is used for 20-way GaN amplifier and high efficiency GaN active antenna design in this work.

In Chapter 3, the high power GaN amplifier with 20-way combiner technology is demonstrated for a long-range microwave transmission assuming a large-scale aperture antenna system based on 7.1 GHz CW operation. Firstly, an amplifier with 4 cell chip of GaN HEMT is design and fabricated, which is assembled to a radial combiner and divider. In addition, the amplifiers with one cell chip and 2-way 4 cell chip of GaN HEMT based on Wilkinson divider technique for driver amplifiers. The obtained small signal gain of the amplifiers with one cell, 4 cell, and 2-way 4 cell chips of GaN HEMT are 13.0 dB, 11.7 dB, and 13.7 dB, respectively. The obtained output power of each amplifier are 35.3 dBm, 41.7 dBm, and 44.6 dBm, respectively. In addition, the obtained PAE of each amplifier are 30.8 %, 42.6 %, and 44.8 %, respectively. The SSPA is assembled with the 20-way amplifiers with four cell chips attached to the radial combiner and divider. In the large-signal measurement, the fabricated amplifiers are used for drivers. According to the measurement result under the large-signal condition, a 52.6 dBm output power and 17.4 % PAE are confirmed at 7.1 GHz. In addition, the measured maximum gain is 10.7 dB. These properties of the 20-way amplifier is agree with the calculated value based on the properties of GaN amplifier with 4 cell chip, combiner, and divider. At comparing to the reported conventional works, the demonstrated 20-way amplifier in this work has a performance comparable to the pulse condition GaN amplifiers in C-band frequency. In the long-range microwave transmission based on an aperture antenna system such as space application, the high power GaN amplifier with one output port is significantly required. In particular, the high power GaN amplifier with more than 100 W output at CW condition contributes the space application such as ground base station for satellite communication and the future microwave power supplying systems.

In Chapter 4, a high efficiency GaN amplifier and integration technique to antenna is proposed. In particular, the amplifier and antenna are integrated based on a low impedance ratio condition, which contribute to reduce the loss of the matching circuit. In the single-stage amplifier design, a GaN amplifier with a harmonic-tuning circuit and an antenna with a multilayer substrate were effectively integrated on a Peltier unit. In the high efficiency amplifier design, the measured maximum output power of 37.95 dBm (6.24 W) at a 27.58 dBm input, and a PAE of 62.94 % at a 27.58 dBm input were confirmed for the large-signal condition. Comparing the PAE of the amplifier with a harmonic tuning circuit to that of the amplifier without this circuit, an 11.57 point PAE improvement was confirmed. The principle of the efficiency improvement is cleared in this work. In the antenna

design, the multilayer substrate is used for reducing the line width of amplifier circuits and increasing the gain of antenna. In general, the substrate with a high dielectric constant is useful for the circuit size reduction which contribute to reduce the radiation loss from RF circuit. In addition, the high dielectric constant privets to expand the line width, which reduces the value of equivalent capacitor caused as the result of size difference between RF line and chip. However, the high dielectric constant generally reduces the antenna gain. In this design with the multi-layer substrate, the antenna gain of 6.40 dBi was measured in the front direction. In the active antenna design, the fabricated GaN amplifier and antenna were integrated to form an active antenna with a 22.43 dBi gain in the front direction at a small-signal condition. In addition, the 44.61 dBm (28.91-W) maximum EIRP was confirmed at a large-signal condition for the 25.70 dBm input. The measured output power and PAE characteristics of the GaN amplifier in the active antenna were 38.21 dBm at a 24.70 dBm input and 59.93 % at a 29.61 dBm input, respectively. Therefore, the GaN amplifier and the antenna were losslessly integrated, showing the viability of the active-antenna-fabrication method. The array antenna technologies for beam combing in free space requires a compact active antenna integrated high efficiency GaN amplifier. The proposed active antenna design method contributes to realize the high efficiency active array antenna for high power microwave transmission.

In Chapter 5, the new approach of tunable sidelobe reduction based on GaN active array antenna is proposed and demonstrated. In this approach, the sidelobe level is tuned by adjusting the operating DC drain voltage of the amplifiers, which enables the situational switching of the radiation pattern. Firstly, a single-stage GaN amplifier is assemble for 4-way amplifier module. The measured maximum output power of the amplifier is 39.3 dBm and the PAE is 59.6 % at a 55 V DC drain voltage; when the drain voltage is decreased to 35 V, the output power decreases to 36.5 dBm. The amplifier output difference is proactively leveraged to realize sidelobe reduction. The characteristics of the amplifiers are measured by a network analyzer and further used to calculate the antenna radiation pattern. The 4-way GaN active-array antenna was tested, with the amplifier DC drain voltage varying from 35 V to 55 V. According to the results, the 4.5 dB sidelobe level is reduced by tuning the operating center-amplifier drain voltage from 35 V to 55 V while the edge amplifiers are fixed to the 35 V condition. By this approach, it is possible to reduce the sidelobe level of more than 50 % by tuning the amplifier drain voltage. Finally, the maximum EIRP for the front direction is measured as 55.6 dBm when the 35 V conditions are set for the edge amplifiers and the 55 V

conditions are provided for the center amplifiers; this value is reduced to 52.1 dBm upon decreasing the center amplifier DC drain voltage condition to 35 V. In the conventional, low-sidelobe array antenna is designed by using RF circuit approach with un-uniform power divider or filter techniques. In this study, tunable side-lobe reduction based on the GaN active antenna-technique is demonstrated, which make it possible for the low-sidelobe antenna to tunably control the sidelobe. The proposed technique will contribute to the broad realization and implementation of high power microwave transmission such as MPT systems. Sidelobes can cause electromagnetic interference, affect the human body, and cause a myriad of problems. Thus, this technique is useful for the situational sidelobe reduction necessary for a variety of future long-range MPT system for EVs, mobile phones, and sensor networks.

In conclusion, the series of researches through high power and high efficiency GaN amplifier to RF antenna for integral study of microwave transmission are reported. The researches of the high power GaN amplifier with 20-way combiner, the high efficiency GaN amplifier with harmonic tuning circuit, the high efficiency active antenna based on the effective integration technique, and the tunable side-lobe reduction based on the GaN active array antenna technique are newly developed in this work. These researches must contribute the microwave transmission technologies which is expected to advance our wireless society. In future, it is expected to grow availability of space application, and microwave power suppling which require the high power microwave transmission technologies. In addition, this technologies has compatibility with the EV, mobile phone, and each and every electric products expected to develop in future IoT (Internet of Things) society. Finally, it is hoped the space solar power satellite is realized by development of this study.

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2. N. Hasegawa, Hyeonjae Ju, S. Yoshida, A. Miyachi, M. Matsunoshita, K. Nishikawa, and S. Kawasaki, “Dual-band Dual-pole Antenna for compatibility of MPT with communication”, in Proc. IEICE General Conf., Fukuoka, CK-2-5, Mar. 2016. (Invited)

Awards

1. Young Researcher Best Presentation Award, “Dual-band Dual-pole Antenna for compatibility of MPT with communication”, Thailand-Japan Microwave Conf., Bangkok, Thailand, TH1-02, Aug. 2015.

Appendix A

History of Wireless Power Transmission

WPT which is a technology to deliver the electric power without wire, connector, and other metallic attachments has attracted attention since early time. The WPT technologies has developed as well as wireless communication and other wireless technologies since electromagnetic fields equations were systematically described by Maxwell in 1864. The WPT studies were historically started by Tesla, which is an essential significant for WPT history. Tesla built Wardencllyffe tower for wirelessly transfer and telegraph power and information to everyone. In second World War, a high power microwave tube known as magnetron developed, which contributed the development of the wireless technologies such as radar, communication, and power transmission. In the half century later from the Tesla's attempt, WPT technologies are brought back into the spotlight for MPT by this tube development.

After the Brown demonstrations, a large-scale applications based on MPT techniques attracted much attentions. Glaser proposed a space solar power satellite (SSPS) based on the MPT technology in 1968 as shown in Fig. A.1 [135]. The concept of SSPS was that the generated power in a geostationary satellite orbit by solar cells was transmitted to earth receiving station by MPT. In the proposed SSPS, the collected solar energy, in the range of the hundreds of kilovolts, directly transformed to microwave by klystrons which is a specialized liner-beam vacuum tube used as a radio frequency (RF) amplifier. In the earth station, the 2×10^7 kW

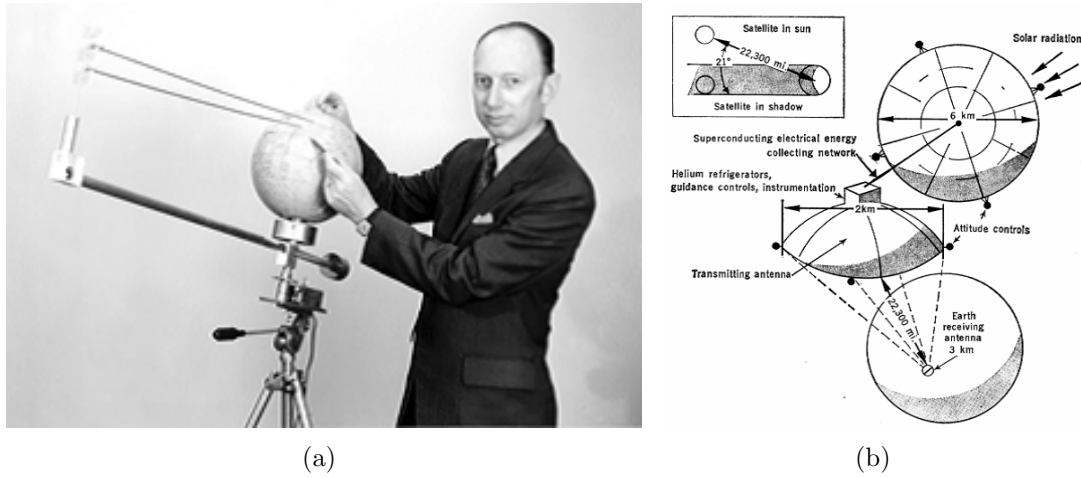


Figure A.1: Proposed SSPS in 1968: (a) photograph of Glaser portrait and (b) concept for satellite system [135].

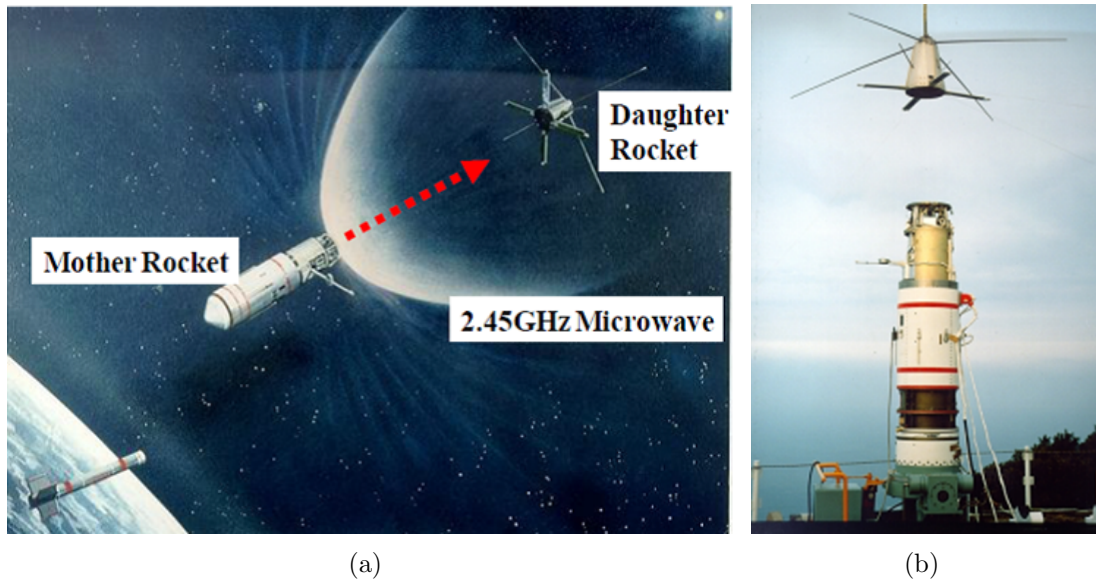


Figure A.2: MILAX rocket demonstration in 1983: (a) image of rocket experiment in ionosphere and (b) Photograph of MINIX rocket [136].

power receiving was assumed. A dish antenna of about 2-km diameter for transmission and receiving antenna of about 3-km diameter in earth were assumed in this system. This SSPS concept is ongoing high-profile for a final target of MPT studies.

From the 1980s onward, the MPT studies were officially started in Japan. Matsumoto et al. demonstrated the MINIX (microwave ionosphere nonlinear interaction experiment) project which is the world-first MPT experiment in the ionosphere in 1983 [136]-[139]. In this demonstration, the 2.45 GHz microwave which

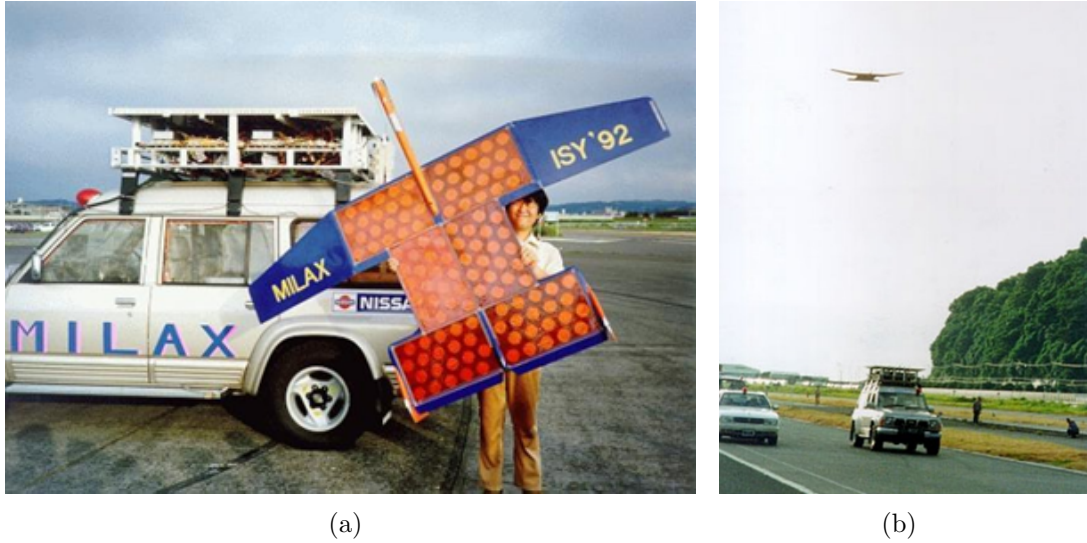


Figure A.3: MILAX airplane experiment in 1992: (a) photograph of airplane and (b) MPT experiment for flying airplane [140].

was generated by magnetron was transmitted from a launched mother vehicle to a daughter one. Matsumoto et al. also demonstrated the MILAX (microwave lifted airplane experiment) project as shown in Fig. A.3 [140]. The MILAX wirelessly powered by a 2.41 GHz microwave from a phased array antenna on a car. The phased array antenna which is able to control the phase relationship of each array antenna elements is a critical technology for modern long-range MPT systems for resolving programs of efficiency degradation due to a position error between transmitting and receiving antenna and the electromagnetic immunity due to an unwanted radiation such as grating lobes from the transmitting antenna. Recently, the output power and operated frequency of the phased array antenna were substantially improved with the performance upgrade of semiconductor technologies, which contributed the progress of MPT technologies as well as radar and wireless communication technologies.

The microwave technologies have contributed greatly to advance the long-range WPT study. On the other hand, the short- and medium-range WPT techniques with non-radiation methods based on inductive coupling and resonant coupling have developed. The inductive coupling in which the power is transferred between coils by a magnetic field for short-range WPT was already confirmed in the 1800s. In the 1990s, the study of RFID (radio frequency identification) technologies based on the inductive coupling, which is recently used in proximity cards and contact-less smartcards, was started. In 2007, MIT team demonstrated the



Figure A.4: WPT demonstration with a resonance inductive coupling techniques by MIT team in 2007 [141].

Table A.1: Comparison of each WPT-technique characteristic.

	Inductive Coupling	Resonant Coupling	Microwave
Frequency	Hz – MHz	kHz – GHz	GHz
Method	Wire coils	Resonators	Antennas
Distance	Short	Medium	Short to long
Efficiency	High	High	High to low

resonant inductive coupling for medium-range WPT as shown in Fig. A.4 [141], [142]. In this demonstration, the 60 W power with less than 40 % efficiency was transferred to a receiving coil across 2 m distance from a transmission coil.

Recently, the studies for WPT techniques of inductive coupling, resonance coupling, and microwave are aggressively reported. Vigneau et al. reported an inductive coupling technique with 98.8 % transmission efficiency at 1 MHz across 1.6 mm distance [143]. Jonah et al. reported a resonance coupling technique with 76.4 % transmission efficiency at 122 MHz across 15 cm distance [144]. The transmission efficiency of MPT system under far field condition is calculated by Friis transmission equation derived in 1945, which explains the transmission efficiency decreases as a function of distance. The comparison of each WPT-technique characteristic is summarized in Table A.1.

Appendix B

Dual-band Dual-pole Antenna for Compatibility of MPT with Communication

The studies of application such as ubiquitous power source, information communication and power trans-mission system, and full-wireless sensor system in a spacecraft, which use MPT technology, are started [145], [81]. The proposed systems are based on coexistence of MPT and information communication. The circuits of MPT system with communication tend to get complex and become a large scale. In particular, it is necessary for the system with frequency separation method to prepare the antennas for MPT and communication separately in conventional systems. Therefore, a dual-band dual-pole antenna for MPT and X-band communication designed on a multilayer substrate is proposed, which expect to reduce the size of the system, in this work. A proto-type of antenna is demonstrated for MPT with the communication in this paper. In addition, 20 W Class GaN high power amplifier for MPT system as shown in Fig. B.1 was fabricated.

B.1 Antenna Configuration

Figure B.2(b) shows the antenna configuration designed on multilayer substrate. The multilayer circuit consists of 8.4 GHz antenna layer, 7.1 GHz antenna layer, and GND layer. The polarizations of each antenna are crossed by 90 degrees,

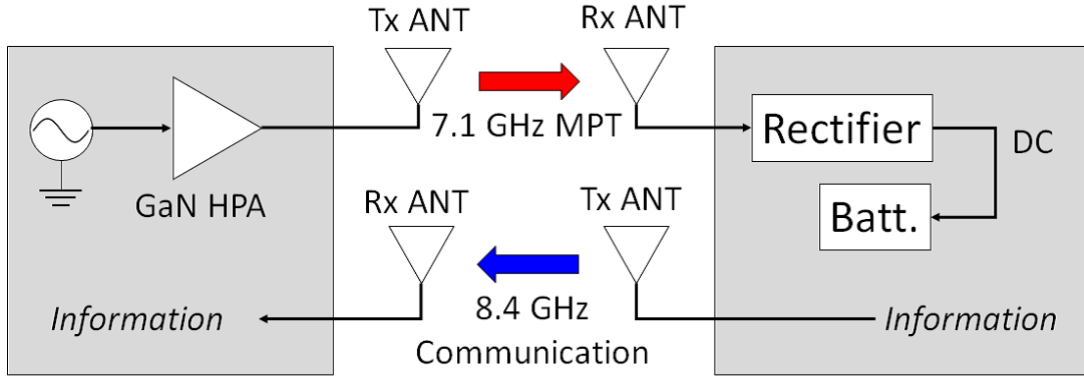


Figure B.1: MPT system with X-band communication.

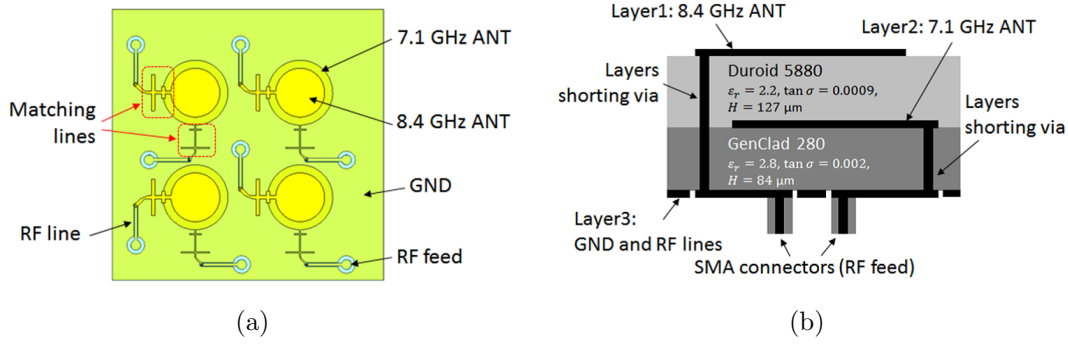


Figure B.2: Dual-band Dual-pole antenna design on a multi-layer substrate: (a) simulation model and (b) layer configuration.

which contributes high isolation between MPT and communication. Duroid 5880 ($\epsilon_r=2.2$, $\tan \delta=0.0009$, $H=0.127$ mm) and GenClad 280 ($\epsilon_r=2.8$, $\tan \delta=0.002$, $H=0.084$ mm) are used and compounded for multilayer substrate. In the back side of antenna, SMA connector is mounted and RF signal or power are fed to each antenna through layers shorting via holes. The fabricated 2×2 antenna is shown in Fig. 2. The isolation between each frequency antenna port should be prevented less than -20 dB for compatibility of MPT and communication. Resulting of S-parameters measurement by network analyzer (Agilent, E8361A), -24.1 dB and -25.3 dB of local maximal $|S_{21}|$ at 7.18 GHz and 8.29 GHz are respectively confirmed from single element.

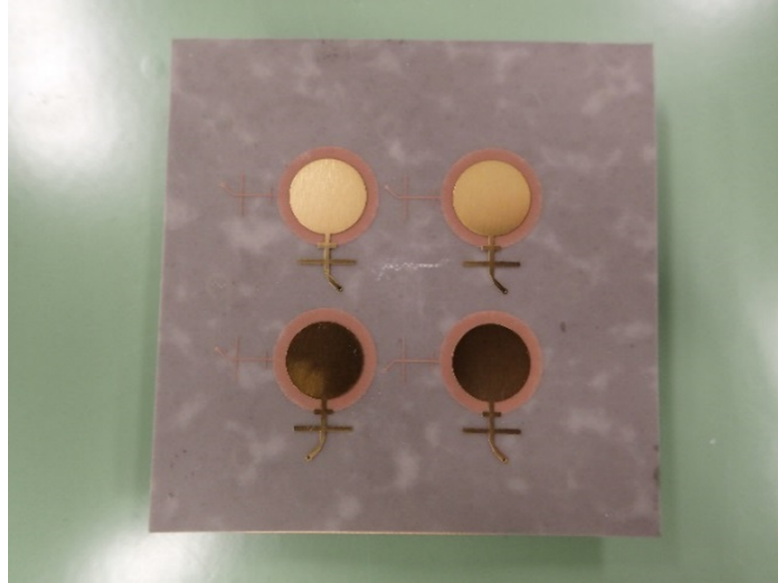


Figure B.3: Top view of fabricated dual-band dual-pole antenna.

B.2 Antenna Fabrication

The dual-band dual-pole antenna is fabricated as shown in Fig. B.3 which is a top-view photograph of the antenna. The measured and calculated S parameters of the antenna are shown in Fig. B.4. In Fig. B.4, the red line shows the $|S_{11}|$ for 7 GHz port and blue line shows the $|S_{11}|$ for 8 GHz port. According to Fig. B.4, these $|S_{11}|$ s are reduce at 7.11 GHz and 8.29 GHz. The black line shows $|S_{21}|$ through the each port, which is less than the -20 dB at each frequency. In addition, these measurement results are agree with the calculated results.

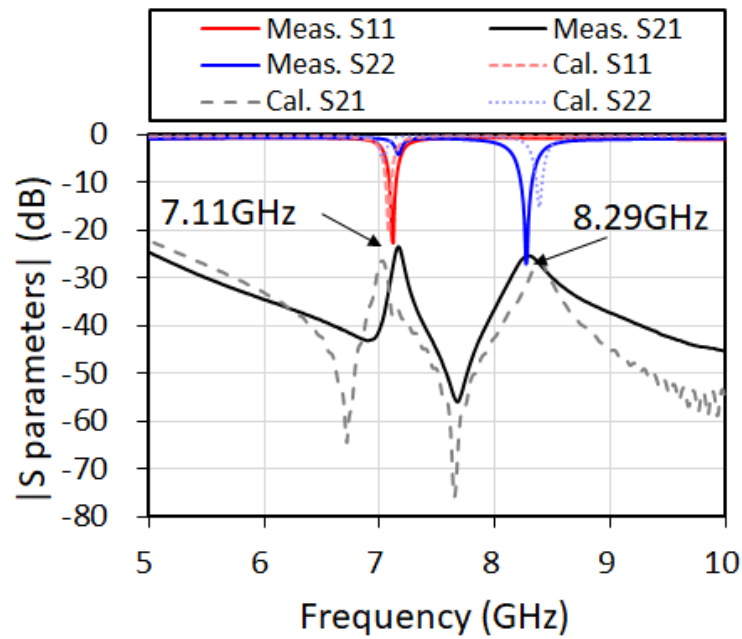


Figure B.4: Measured S parameters of dual-band dual-pole antenna.

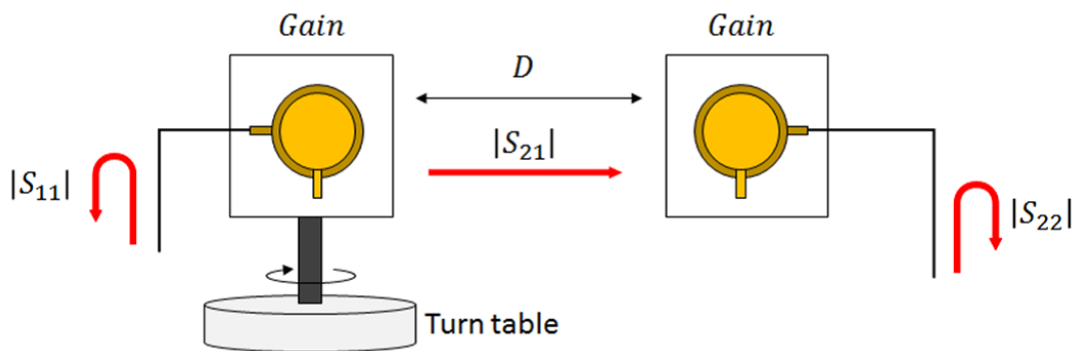


Figure B.5: Far-field measurement system based on two-antenna method.

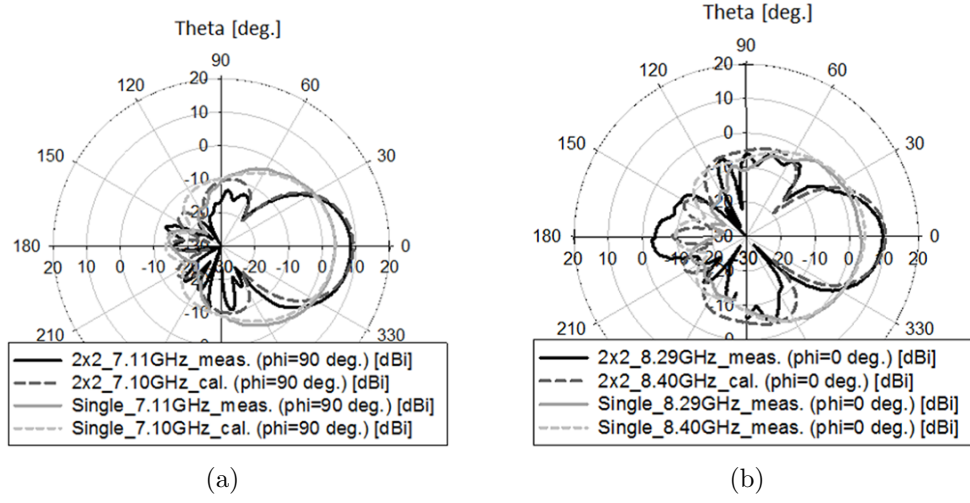


Figure B.6: Measured antenna gain. (a) 7.11 GHz at phi=90 deg. plane, (b) 8.29 GHz at phi=0 deg.

B.3 Far Field Measurement

The gain of fabricated antenna is measured by the far-field system in a microwave anechoic chamber. In this system, AUTs (Antenna Under Tests) are set on turn table and angular characteristics of gain are measured. The antenna gain is measured by two-antenna method which use same antennas for Tx and Rx. In this method, the antenna gain G (dBi) is calculated by following equation,

$$G = \frac{1}{2} \cdot \left[20 \log_{10} \left(\frac{4\pi D}{\lambda_0} \right) + (|S_{21}|)_{\text{dB}} + (|M1|)_{\text{dB}} + (|M2|)_{\text{dB}} \right], \quad (\text{B.1})$$

where D is distance between Tx and Rx antennas, $M1$ and $M2$ are return loss of each antenna. The distance between antennas is 57 cm in this measurement. Figures B.6(a) and B.6(b) show the measured antenna gain at 7.11 GHz and 8.29 GHz. According to the results, the gain of 8.49 dBi at 7.11 GHz and 9.61 dBi at 8.29 GHz are confirmed in a front direction at each polarization plane of 2x2 antenna. The front direction gain of single antenna are 3.91 dBi at 7.11 GHz and 3.57 dBi at 8.29 GHz.

Appendix C

Hardware Retro-directive system with a Dual-Band Antenna

The retro-directive is a beam controlling technique for targeting to the incoming angle of received pilot wave. Figure C.1 show the basic configuration of retro-directive system which consist of antennas, filters, and mixers. Assuming to income a RF wave from a target to received antenna with an input angle, the input RF wave to mixer is defined as following equation:

$$RF = E_{\text{in}} \cos(\omega_{\text{in}}t + \Delta\psi_{\text{in}}), \quad (\text{C.1})$$

where E_{in} is an electric intensity of input wave, ω_{in} is an angular frequency of input wave, and $\Delta\psi_{\text{in}}$ is a phase difference of input wave between antennas. In the mixer, the RF signal is mixed with a LO signal which make a filtered IF signal for output wave as following equation:

$$IF = E_{\text{out}} \cos(\omega_{\text{out}}t - \Delta\psi_{\text{out}}), \quad (\text{C.2})$$

$$= \frac{E_{\text{in}}E_{\text{lo}}}{2} \cos[(\omega_{\text{in}} - \omega_{\text{lo}})t - \Delta\psi_{\text{in}}] \quad (\text{C.3})$$

where E_{out} is an electric intensity of IF signal, ω_{out} is an angular frequency of IF wave, $\Delta\psi_{\text{out}}$ is a phase difference between IF waves from each mixer, E_{lo} is an electric intensity of LO signal, and ω_{lo} is an angular frequency of LO wave. According to this equation, radiation wave of the antennas based on IF signal of mixer has an inverse value of the phase difference of input wave between antennas, which provide the system with a property about same angle reradiation to input

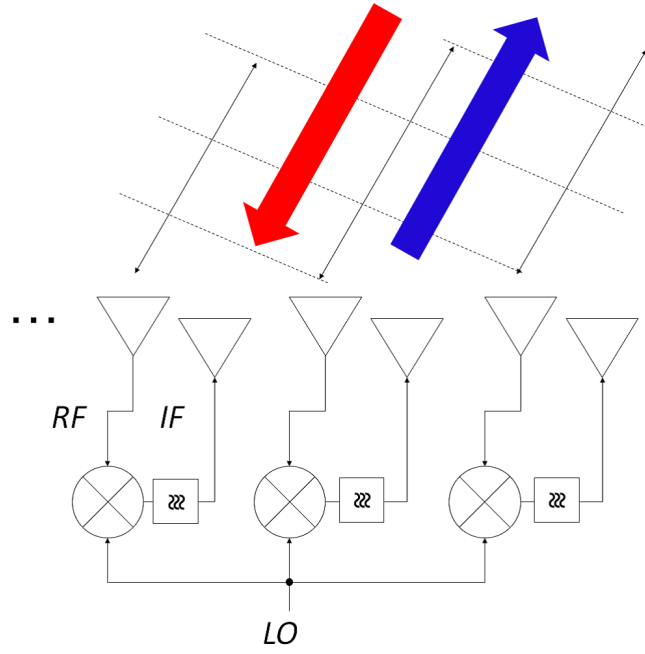


Figure C.1: Basic configuration of retro-directive system.

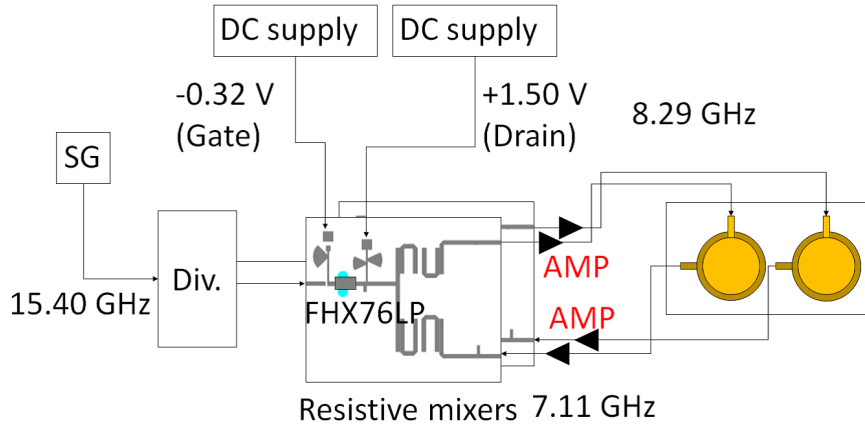


Figure C.2: Designed retro-directive system.

angle of pilot wave from target. In this work, the retro-directive system with 7.1 GHz input and 8.4 GHz output based on the dual-band antenna fabricated in previous section is designed and fabricated.

Figure C.2 show the designed retro-directive system which consists on resistive mixers and a dual-band antenna. The LO signal is input to the mixers through divider from a signal generator. The RF and IF signals are amplified by an amplifiers and separated by a diplexer. The mixer and amplifier are designed based on a GaAs FET, FHX76LP. The fabricated mixer integrated the diplexer

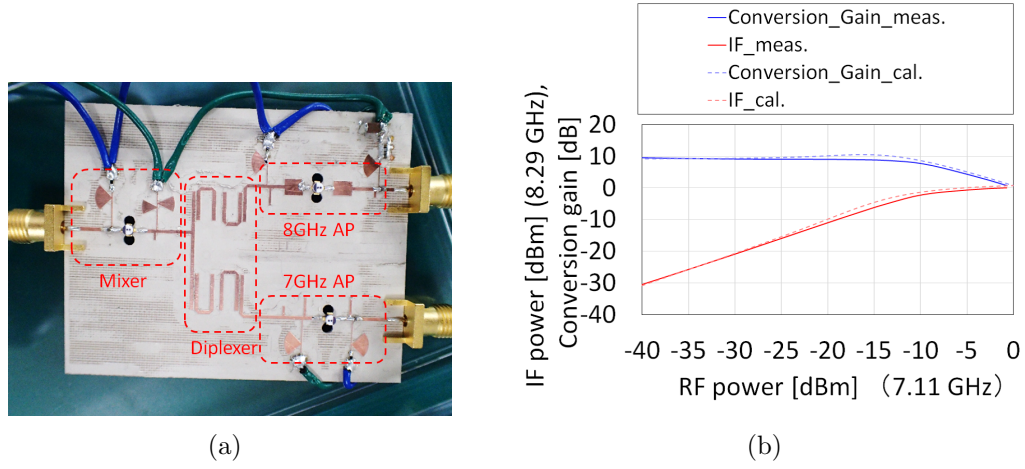


Figure C.3: fabricated mixer integrated with diplexer and amplifiers: (a) photograph of the mixer and (b) measured properties of the mixer.

and amplifier is shown in Fig. C.3(a). The measured properties of IF power and conversion gain for RF power is shown in Fig. C.3(b).

The measured beam pattern of the retro-directive system is shown in Fig. C.4. In this experiment, the radiation beam pattern are measured at 0 degrees to 30 degrees by 15 degrees. The ideal phase difference between antennas are calculated by the following equation:

$$\Delta\psi_{\text{out}} = -\Delta\psi_{\text{in}} = -\frac{2\pi d_{\text{in}} \sin(\theta_{\text{in}})}{\lambda_{\text{in}}} = \frac{2\pi d_{\text{out}} \sin(\theta_{\text{out}})}{\lambda_{\text{out}}}, \quad (\text{C.4})$$

where d is a distance between antenna elements, θ is beam angle, and λ is free space wavelength. Therefore, the reradiation angle is calculated by the following equation:

$$\sin(\theta_{\text{out}}) = -\frac{\lambda_{\text{out}}}{\lambda_{\text{in}}} \sin(\theta_{\text{in}}). \quad (\text{C.5})$$

The calculated and measured input angle of pilot wave, and reradiation angle are shown in Table C.1. According to results, it is confirmed the calculated and measured reradiation angles has an error based on equation C.5 for input angle. In the hardware retro-directive with difference frequency between input and reradiation wave, it is necessary to solve the error based on ratio of output wavelength to input wavelength.

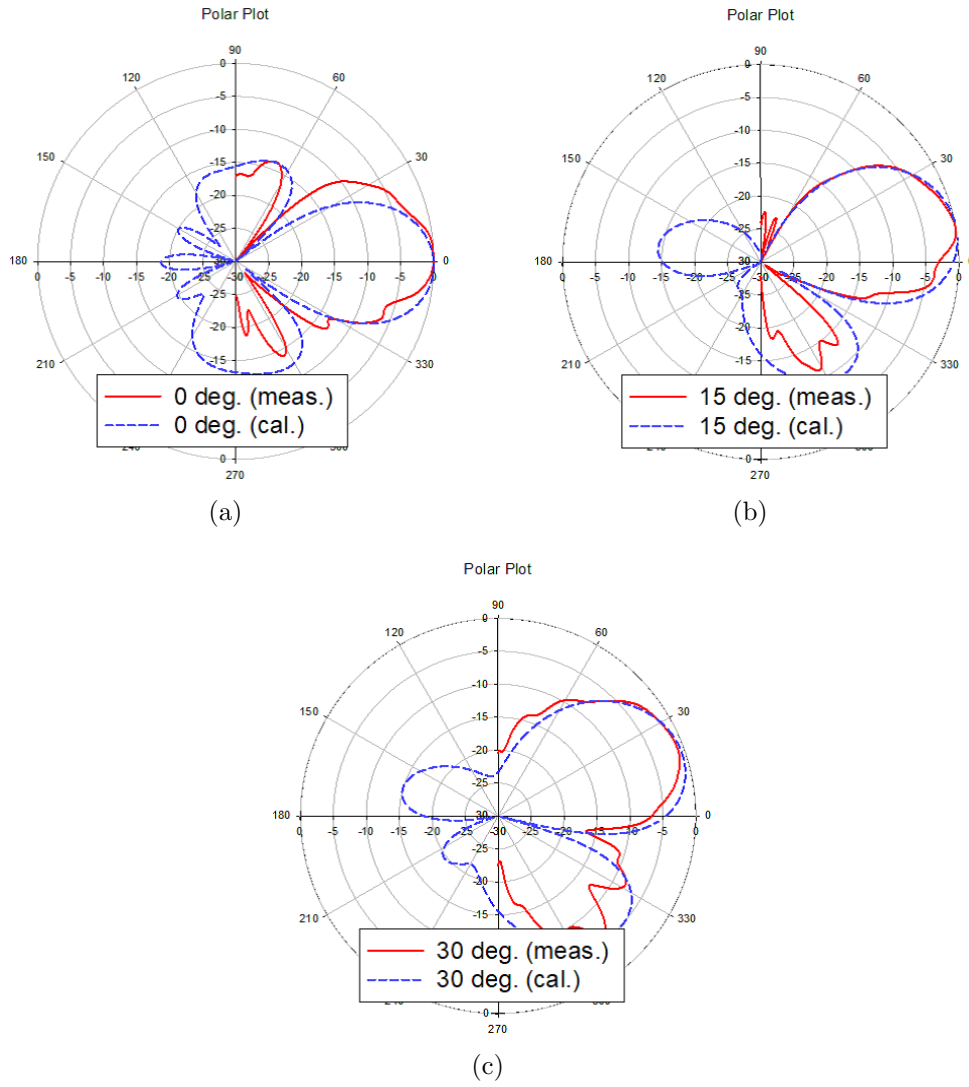


Figure C.4: Measured reradiation pattern of the retro-directive system.

Table C.1: Reradiation angles for input angle of pilot wave.

Input angle	0 deg.	15 deg.	30 deg.
Reradiation angle (Cal.)	0 deg.	12.6 deg.	25.0 deg.
Reradiation angle (Meas.)	0 deg.	12.5 deg.	21.5 deg.