

EFFICIENCY IMPROVEMENT FOR ELECTRONICALLY STEERED DUAL-BEAM ANTENNA BY USING METAMATERIAL BASED RADOME

Auychai Yatongchai, Peerasan Khamsalee and Rangsan Wongsan*

Received: March 25, 2023; Revised: July 09, 2023; Accepted: July 18, 2023

Abstract

The efficiency improvement for an electronically steered dual-beam antenna by using a metamaterial-based radome is a PCB-based circular patch antenna design for dual-beam steering. This design employs the original circular patch antenna, which consists of eight small square patches placed symmetrically on eight angles and separated by slots. The beam-steering property of this antenna can be implemented by controlling the on/off state of the PIN diodes, which are connected between each small square patch and the main patch at each angle of the structure. When the diode is on, the small square patch becomes a part of the radiating patch, whereas is off, and the small square patches become the parasitic element. The antenna is designed for future high-altitude platforms such as drones, aircraft, rockets, etc. This can be tracked or commanded from the control station within a given range without disconnection. This antenna is investigated at 5.8 GHz of frequency and provided a bandwidth of 113 MHz, which is high enough for unmanned aerial vehicle UAVs applications. It provides a gain of 7.13 dB without a meta radome and is switchable to the dual beam in eight different space quadrants for the azimuth plane by exciting forward bias with binary sequence to the PIN diodes. While it provides a constant tilted beam of 35° and angular width (3 dB) of 52.4° in the elevation plane, and efficiency improvement for the antenna provides a gain of 10.18 dB with a meta-radome

Keywords: Beam steering, Dual beam, PIN diode, Patch, Metamaterial, MetaRadome, Electromagnetic Band Gap (EBG), UAV

Introduction

In recent years, the use of unmanned aerial vehicles (UAV) has attracted significant interest in the military, communications, and commercial markets. Unmanned aerial vehicles, commonly known as drones, are aircraft without a pilot, crew, or passengers on board, and have become one of the

most interesting areas of the Defence Technology Institute (DTI), Ministry of Defence, the Kingdom of Thailand. In addition to the ongoing projects implemented by DTI, a modified anti-aircraft rocket project with a prototype developed into a modified rocket propulsion system and testing to verify the

School of telecommunication engineering, Institute of engineering, Suranaree university of technology, Nakhon Ratchasima, 30000, Thailand. E-mail: auychai_ya@hotmail.com; mpeerasan.k@gmail.com; rangsan@sut.ac.th

** Corresponding author*

DOI: <https://doi.org/10.55766/sujst-2023-04-e0862>

design results is one of the projects of which DTI is proud. This project was integrated into the Department of Royal Rainmaking and Agricultural Aviation. This Project was used to support the Royal Rainmaking Project. It is an extension of the knowledge of rocket production to support projects through royal initiatives and for the defense of the country (Defence Technology Institute, 2016). In Thailand, high-altitude platform stations are to be operated in the frequency spectrum of the C-band and X-band to support operations in fixed and mobile services. The high-altitude platform station is 17 km above the Earth's surface and serves a ground area of 60-200 km in diameter, with an elevation angle from the ground of up to 30° "General Use" is meant in accordance with the use purpose of UAV which is externally controlled for aviation as stipulated in the Notification of the Ministry of Transport on Rules for permission and conditions for control and launch of unmanned aircraft in the category of remotely piloted aircraft and subsequent amendment. The use of radio frequency and equivalent isotropically radiated power (e. i. r. p.) are determined at 5.725-5.850 GHz and 1000 mW, respectively (Government Gazette, 2018; Muhammad *et al.*, 2011). The antennas commonly used with UAVs are dipoles, patches, arrays, skew planars, and helical antennas, which are omnidirectional and single-beam directional antennas, which generally have a gain of about 4 dB-7 dB (Zhang and Yang, 2017, 2018; Xinhuan *et al.*, 2021). An interesting antenna that Muhammad *et al.* (2011) have designed is a dual beam antenna using metamaterial based on a complementary split ring resonator, mounted on both wings of the aircraft. which the antenna provides a gain of ~7 dB. Therefore, researchers are interested in the design of beam-steering antennas to compensate for blind angles from aerobatic flight, increased gain, and radiated waves for longer distances and to cover more space.

Recently, beam-steering antennas have become one of the most interesting areas of wireless communication, especially antennas for future high-altitude platforms such as drones, missiles, and aircraft. The beam-steering antenna on such platforms can be continuously tracked and commanded from their control stations within a specific distance without disconnection. At present, there are various techniques for steering antenna beams, such as the use of switchable parasitic elements, leaky-wave antennas on PCB, comb line techniques on microstrip arrays, and steerable beams with metamaterial structures (Jais *et al.*, 2013; Iyemeh and James, 2016; Hisamatsu *et al.*, 2017; Asaduzzaman *et al.*, 2018). In this study, we focus on a PCB-based microstrip patch

antenna owing to its simple structure, ease of fabrication, lightweight, and low cost. To steer the beam of this antenna, beam steering can be implemented by controlling the on/off state of the PIN diodes, which are connected to a small square patch of the structure. When the diode is on, the small square patch becomes part of the radiating patch, whereas is off and the small square patch becomes a parasitic element. If we can design an electronically steered dual-beam antenna, the switching beam can be decreased by half compared to that of a conventional single-beam antenna to reduce the signal loss during switchover. An electronically steered dual-beam antenna with a metamaterial-based radome was presented theoretically.

Subsequently, the influence of the dimensions and positions of the small square patch was described. When these parameters were found, eight elements of the small square patch addition and microstrip patches were investigated to meet the requirements of the proposed dual beam. The simulation results obtained using CST software show the curves of the reflection characteristics, S_{11} of the patch antenna, and dual beam, which can be improved by varying the number and dimension of the small square patches.

In this study, a method to increase the efficiency of antennas using a new technology called metamaterial structure is proposed because it can increase the radiation efficiency of microwave antennas (Yang and Rahmatsamii, 2009; Tie *et al.*, 2013). In addition, the design of the metamaterial-based radome covering the electronically steered dual-beam antenna protected the antenna effect from the environment, and the antenna gain could be increased to at least 3 dB, radiating the waves for longer distances and covering more space. Finally, the experimental results of the dual-beam steering were measured and validated with the simulation results.

Theory and Configuration of The Proposed Antenna

Initially to design the dual-beam antenna, the rectangular patch theory was used to transform to a circular patch, which includes the PIN diode and bias T are used to control the dual-beam steering of the antenna. Therefore, it is necessary to understand the principles and theories to design dual-beam antennas at the desired frequency, which are detailed in the following topics.

Circular Patch Design Method

Suitable and similar substrates must be selected to provide a common platform for antenna simulations. This greatly reduces any discrepancies,

instead of simulating the fabrication of different microwave laminates. The chosen substrate has a dielectric constant (ϵ_r), dielectric loss tangent ($\tan\delta$), and substrate height for a single layer. The substrate thickness and cutoff frequency can be characterized using the following formula. The patch has an effective dielectric constant.

$$\epsilon_{\text{reff}} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left[1 + 12 \frac{h}{W} \right]^{-\frac{1}{2}} \quad (1)$$

Substrate thickness ;

$$h \leq \frac{1}{4f\sqrt{\epsilon_{\text{reff}} - 1}} \quad (2)$$

where c is the free-space velocity, ϵ_{reff} the effective dielectric constant, and f the operating frequency. The cutoff frequency was calculated using the following formula:

$$f_c = \frac{c}{\sqrt{\epsilon_{\text{reff}} [2W0.8h]}} \quad (3)$$

where f_c is the cutoff frequency, W is the width of the patch, and h is the substrate height.

The patch width should be selected to obtain good radiation efficiency if the real estate requirements or grating lobe are not overriding factors. It has been suggested that $1 < W/L < 2$ for an operating frequency 5.8 GHz. The width of the TM₁₀ Mode was given by (Balanis, 2005).

$$W = \frac{1}{2f_r\sqrt{\mu_0\epsilon_0}} \sqrt{\frac{2}{\epsilon_r + 1}} = \frac{c}{2f_r} \sqrt{\frac{2}{\epsilon_r + 1}} \quad (4)$$

where ϵ_r is the relative dielectric constant, ϵ_0 is the free-space permittivity, μ_0 is the permeability of air, f_r is the resonant frequency, and c is the free-space velocity.

The length of the patch determines the resonant frequency and is an important design parameter owing to its inherently narrow bandwidth. The extended incremental length of patch ΔL was:

$$\frac{\Delta L}{h} = 0.412 \frac{(\epsilon_{\text{reff}} + 1) \left(\frac{W}{h} + 0.264 \right)}{(\epsilon_{\text{reff}} - 0.258) \left(\frac{W}{h} + 0.8 \right)} \quad (5)$$

Thus, the actual length of the patch is;

$$L = \frac{\lambda}{2} - 2\Delta L \quad (6)$$

This gives out the effective length;

$$L_e = L + 2\Delta L \quad (7)$$

The conversion from a common rectangular shape to a circular shape is given by the following equation: $A_{\text{rectangular}} (\text{Width} \times \text{Length}) = A_{\text{circular}} (\pi r^2)$; The diagram is shown in Figure 1.

Calculation of parameters for a basic rectangular patch shape transforms to a circular patch shape, which considers frequency, width, length, radius, and area, as calculated values are shown in Table 1, to be used to configure the design of the antenna.

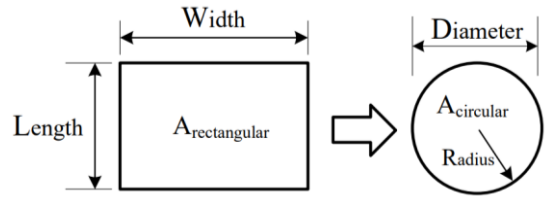


Figure 1. Diagram of changing rectangular shape to circular shape

There are several types of beam-steering antennas, such as patches, circular plates, SRRs, and cross-shaped structures (Jais *et al.*, 2013; Hisamatsu *et al.*, 2017). In the current design, we chose square and circular patches because of their simplicity and symmetrical geometry for supporting the dual beam. Figure 2 shows the slit-top uniform square patch and its dimensions. The parameters of the microstrip patch antenna are given as follows: D is the antenna diameter, w_p is the small patch width, w_g is the air-gap width, W is the patch diameter, w is the circular patch diameter, and h is the substrate thickness.

Table 1. Calculation of the basic patch shape of width, length, and radius

f_r (GHz)	f_c (GHz)	λ (mm)	ϵ_r (F/m)	ϵ_{reff} (F/m)	W (mm)	L (mm)	ΔL (mm)	L_e (mm)	$L_e \times W$ (mm ²)	r (mm)	πr^2 (mm ²)
5.8	15.33	51.72	4.4	3.49	42.49	24.09	0.89	25.86	1098.79	18.70	1098.79

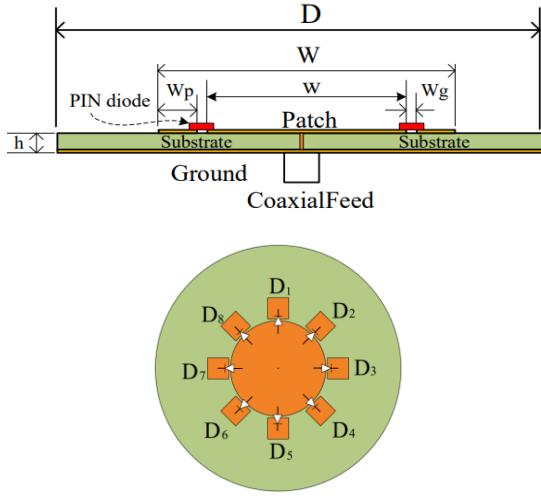


Figure 2. The geometry of proposed slit-top uniform square patch and circular patch

Plane-wave analysis was performed to study the transmission and reaction characteristics of the microstrip patches with different incidence angles. The boundary condition is denoted as a perfect electric conductor (PEC) along both ends of the unit cell. In addition, the resonant frequency of the small square patch structure can be tuned by adjusting the air-gap widths and small square patch sizes, which affect the bandwidth. The bandwidth of the uniform circular patch was also improved, partly because of the increase in the substrate thickness (h) and partially because of the decrease in the equivalent dielectric constant (ϵ_r). The configuration of the circular patch structure with an additional air gap is illustrated in Figure 2. It is composed of a one-layer substrate of thickness (h) with a dielectric constant (ϵ_r). The method for calculating the patch width (W) applied to a frequency of 5.8 GHz is calculated by $2r$ equal to 37.4 mm. Those parameters are shown in Table 1, where w_p , w_g , and D are described in the section on The simulation results of the antenna (see Figure 7).

PIN Diode Circuits for Controlled Dual-Beam Antennas and Bias Tee

PIN diodes have many applications in electronic circuits and RF devices, owing to their ability to act as fast switches. Description of the SMP1345 chip package. This PIN diode is designed for high-throughput devices such as wireless local area networks (WLAN), low-noise blocks (LNB), and switch applications from 10 MHz to 6 GHz. The short carrier lifetime is typically 100 ns (typical), combined with an I-region width as thin as 10 μm (nominal), resulting in a group of high-speed RF-

switching PIN diodes. The SMP1345 series exhibited RF performance because of its very low capacitance (0.15 pF) and impedance (1.5 Ω at 10 mA).

Bias T is a special 3-port device that is mainly used when trying to use DC power (DC Bias) to protect RF components from being damaged by DC bias. L_1 and L_2 prevent the reverse current of the RF signal while the PIN diode is biased (see Figure 3). A typical Bias T circuit looks like this: see the circuit is 'T' shaped, which is why it is called bias T. The performance characteristics are as follows:

$$X_c = \frac{1}{j2\pi fC} \quad (8)$$

$$X_L = j2\pi fL \quad (9)$$

$$L_{fr} = \frac{c}{4f_r\sqrt{\epsilon_{\text{reff}}}} \quad (10)$$

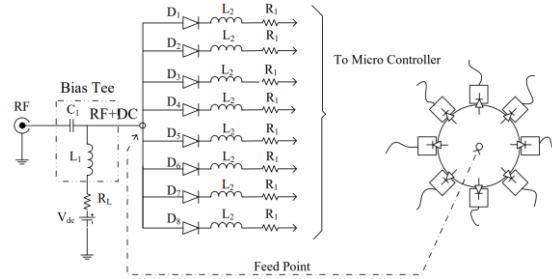


Figure 3. Bias tee and PIN diode circuits

Theory and Configuration of the Meta Radome

In order to design the meta-radome, metamaterial theory was used to design the unit cell, which includes a Jerusalem cross-frequency selective surface (JC-FSS) and a square loop-frequency selective surface (SL-FSS). Therefore, it is necessary to understand the equivalent circuits of unit cells. In order to be able to design and simulate the meta-radome, it will be explained in the section on the equivalent circuit of the unit cells.

Equivalent Circuit of Square Loop Frequency Selective Surface (SL-FSS)

This section explains the understanding of equivalent circuits and their parameters of the unit cells, which are described by combining the two types of unit cells. In general, the basic structures of the frequency selective surfaces (FSS) consists of the square grids (SG), the square patches (SP), and

the square loops (SL). They are shown in Figure 4. In the general case, these metallic arrays may be in a homogeneous medium or at the interface of two different dielectric substrates. In reference Yuan and Mang (2019), the surface impedance of the SG and SP were studied detail in Figure 4(a-b), the characteristics of the lumped equivalent circuit of the SL-FSS structures is the band-stop filter shown in Figure 4(c) were derived. The equivalent circuits of the conventional SL-FSSs were defined as the values of the terms seen in Figure 4 that were derived in (Lee and Langley 1985; Parker, 1991). The admittance and impedance of the SL-FSS can be calculated using the equivalent circuit shown in Figure 4(c) given by Parker (1991).

$$Z_{01} = j \left[\omega L_1 - \frac{1}{\omega C_1} \right] = \frac{-j}{\omega C_1} (1 - \omega^2 L_1 C_1) \quad (11)$$

$$Y_1 = \frac{1}{Z_{01}} = \frac{1}{\frac{-j}{\omega C_1} (1 - \omega^2 L_1 C_1)}$$

$$Y_1 = \frac{\omega C_1}{-j(1 - \omega^2 L_1 C_1)} = \frac{j\omega C_1}{-j^2(1 - \omega^2 L_1 C_1)} ; j^2 = -1$$

$$Y_1 = j \left(\frac{\omega C_1}{(1 - \omega^2 L_1 C_1)} \right) \quad (12)$$

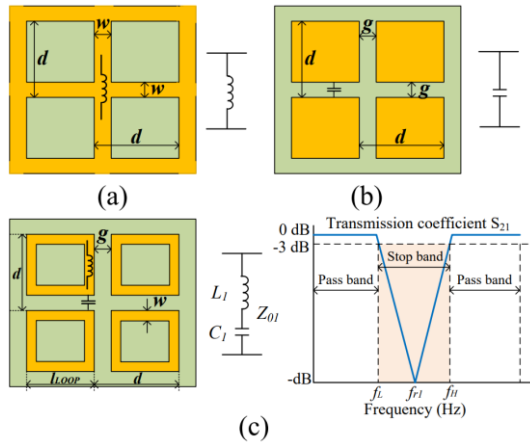


Figure 4. (a) Square grids, (b) Square patches, and (c) The band stop equivalent circuit of Square loops

This is a lumped element used to analyze the surface impedance formula. Base on (Yuan and Mang, 2019), described the method for estimating the per-unit-length inductance (L_1) of the strip in the SL-FSS structure and the per-unit-length capacitance (C_1) of the slot between two adjacent patches in the SL-FSS structure. It can be estimated using Equations (13) and (14) as follows:

$$L_1 = \mu_0 \frac{l_{Loop}}{2\pi} \ln \left(\frac{1}{\sin \left(\frac{\pi w}{2d} \right)} \right) \quad (13)$$

$$C_1 = \epsilon_0 \epsilon_{reff} \frac{2l_{Loop}}{\pi} \ln \left(\frac{1}{\sin \left(\frac{\pi g}{2d} \right)} \right) \quad (14)$$

Where the resonant frequency (f_r) can be expressed as the following equation:

$$f_{r1} = \frac{1}{2\pi \sqrt{C_1 L_1}} \quad (15)$$

where $g = (d - l_{Loop})$ is the slot width and w is the strip width, as shown in Figure 4(b). L_1 and C_1 represent the inductance and capacitance of the equivalent circuit model of the SL-FSS array, respectively.

Equivalent Circuit Model of Jerusalem Cross-Frequency Selective Surface (JC-FSS).

Consider Figure 5(a), the geometry of a Jerusalem cross-frequency selective surface (JC-FSS) with a cross-shaped metal plate on a dielectric substrate. The JC-FSS element and ground plane are considered perfectly conductive when the period is small compared to the wavelength of interest. The equivalent circuit and characteristics of the artificial magnetic conductor (AMC) can be described using the effective surface impedance model shown in Figure 5(b). The JC-FSS equivalent circuit in Figure 5(c) is the configured band-stop filter, where C_2 is the capacitance C_g and L_2 is the sum of the inductances L_s and L_d .

When the AMC structure has an important parameter, the surface impedance (Z_s) plays an important role in determining the resonance frequency and the phase of the reflection coefficient. The surface impedance (Z_s) can be taken in a transmission line approach, explaining that the JC-FSS impedance (Z_g) and the surface impedance of the grounded dielectric slab (Z_d) are connected in parallel (Hosseini and Hakkak, 2008). Therefore, in Figure 5(b), the surface impedance can be expressed as.

$$Z_s(\omega) = Z_g \parallel Z_d = j\omega L_d \frac{1 - \omega^2 L_g C_g}{1 - \omega^2 C_g (L_d + L_g)} \quad (16)$$

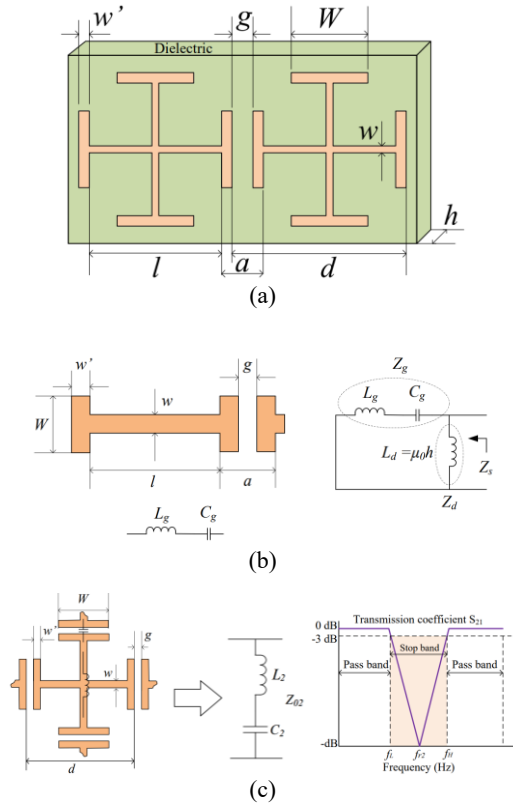


Figure 5. Shows (a) The geometry of JC-FSS, (b) The surface impedance, and (c) the band stop equivalent circuit model of JC-FSS array

In Equation (16), ω is the angular frequency, L_d is the inductance of the grounded dielectric slab, and L_g is the inductance of the JC-FSS. Where C_g is the capacitance formed by the adjacent JC-FSS junction, which has very little influence on Z_d . When the denominator of Equation (16) is equal to zero, the resonant frequency (f_r) can be expressed as the following equation:

$$f_{r2} = \frac{1}{2\pi\sqrt{C_2 L_2}} \quad (17)$$

$$f_{r2} = \frac{1}{2\pi\sqrt{C_g (L_g + L_d)}} \quad (18)$$

Therefore, Z_s is very high and the reflection phase coefficient is zero at the resonant frequency (f_r).

In (Sievenpiper, 1999), a simple estimation method for determining capacitance C_g , Sievenpiper described an approach to extract the capacitance of two parallel patches placed on a dielectric slab.

$$C_g = \frac{2W}{\pi} \epsilon_0 \epsilon_{\text{reff}} \text{Cosh}^{-1} \left(\frac{a}{g} \right) \quad (19)$$

The inductive component (L_g) is the length of l , the width of w using the microstrip transmission line theory, and c is the velocity of light in free space, which can be estimated using the formula (Bahl, 2003):

$$L_g = \frac{Z_0 \sqrt{\epsilon_{\text{reff}}}}{c} l \quad (20)$$

When, the effective dielectric constant can be calculated using equation (21) (Gupta *et al.*, 1996)

$$\epsilon_{\text{reff}} = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2} \left(1 + \frac{12h}{w} \right)^{-\frac{1}{2}} + 0.041 \left(1 - \frac{w}{h} \right)^2; \frac{w}{h} \leq 1 \quad (21)$$

where h is the thickness of the dielectric slab, w is the width of the microstrip line, and η is the impedance of the free space. L_2 and C_2 represent the inductance and capacitance of the equivalent circuit model of the JC-FSS array theory, respectively, as shown in Figure 5(c), which was described by (Pristin, 2013), Z_{02} and Y_2 are described by Equation (22)-(23), respectively.

$$Z_{02} = j \left[\omega L_2 - \frac{1}{\omega C_2} \right] \quad (22)$$

$$Y_2 = j \left(\frac{\omega C_2}{(1 - \omega^2 L_2 C_2)} \right) \quad (23)$$

The inductance of a grounded dielectric slab is an incident wave normal to the surface of the JC AMC, and the impedance of a metal-backed dielectric slab with thickness h can be expressed as (Simovski *et al.*, 2005).

$$Z_d = \left(\frac{j\eta}{\sqrt{\epsilon_r}} \right) \tanh kh \quad (24)$$

In Figure 5(a), Z_d is modeled by inductance (L_d). We have $Z_d = j\omega L_d$; therefore, we can write it immediately as follows:

$$L_d = \left(\frac{\eta}{\omega \sqrt{\epsilon_r}} \right) \tanh kh \quad (25)$$

where $k = \omega\sqrt{\epsilon_r\epsilon_0\mu_0}$ is the wave number and $\eta = \sqrt{\mu_0/\epsilon_0}$ is the free-space impedance

From Equation (24), when the substrate is smaller than the wavelength ($k, h \ll 1$), inductance can be calculated using the formula presented by (Sievenpiper, 1999).

$$L_d = \mu_0 h \quad (26)$$

Equivalent Circuit Model of Combined JC-FSS with Conventional SL-FSS

Finally, Figure 6(a) shows the geometry and dimensions of the combined JC-FSS with the conventional SL-FSS for the design unit cell, and the admittance of the combined JC-FSS with the conventional SL-FSS is calculated using the L_1C_1/L_2C_2 equivalent circuit as a dual-band stop, where a pass band appears between the two band stops, as shown in Figure 6(b).

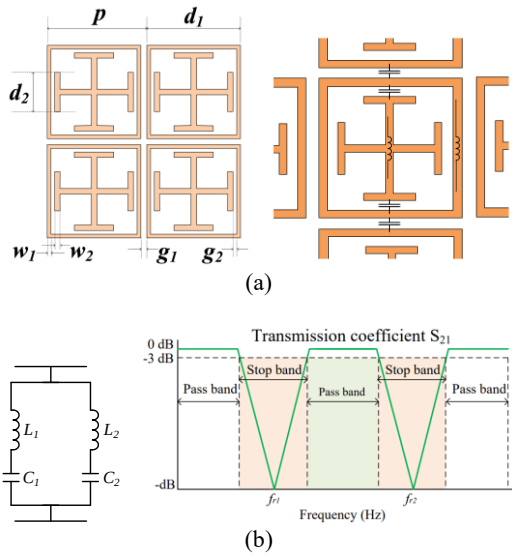


Figure 6. Shows (a) To combined JC- FSS with SL-FSS, and (b) The dual-band stop filter equivalent circuits model of JC-FSS with SL-FSS

The admittance can be calculated using the equivalent circuit shown in Figure 6

$$Y = Y_1 + Y_2 \quad (27)$$

$$Y = \left(j \frac{\omega C_1}{(1 - \omega^2 C_1 L_1)} + j \frac{\omega C_2}{(1 - \omega^2 C_2 L_2)} \right)$$

$$Y = j \left(\frac{(\omega C_1 - \omega^3 C_1 C_2 L_2) + (\omega C_2 - \omega^3 C_1 C_2 L_1)}{(1 - \omega^2 C_1 L_1)(1 - \omega^2 C_2 L_2)} \right)$$

$$Y = j \omega \left(\frac{C_1 + C_2 - \omega^2 C_1 C_2 (L_1 + L_2)}{(1 - \omega^2 C_1 L_1)(1 - \omega^2 C_2 L_2)} \right) \quad (28)$$

We define C_1 and L_1 as the capacitance and inductance of the SL-FSS equivalent circuit, respectively, and C_2 and L_2 are the capacitance and inductance of the JC-FSS equivalent circuit, respectively. Therefore, we can estimate Z_0 from the LC parallel equivalent circuit from the $Z_{LC||LC} = 1/Y$ using Equation (30).

$$Z_0 = Z_{01} || Z_{02} = \frac{1}{Y} \quad (29)$$

$$Z_{LC||LC} = \frac{1}{j \omega} \left(\frac{(1 - \omega^2 C_1 L_1)(1 - \omega^2 C_2 L_2)}{C_1 + C_2 - \omega^2 C_1 C_2 (L_1 + L_2)} \right) \quad (30)$$

Simulation Results of Antenna

This topic shows the simulated results for the reflection performance of the square patch shown in Figure 7 when its patch width (w_p) is kept constant and the gap width (w_g) is varied. The capacitance ratio is varied, as well as the values of L and C of the gap. As the patch width (w_g) is reduced, the values of L and C decrease, which is assigned to characterize the reflection curve that occurs in the required frequency band. First, in Figure 7(a), the value of w_p is fixed, whereas the value of w_g is varied to determine the proper slit width for the patch. We found that the width of w_g was related to the position of the resonant frequency. In Figure 7(b), to consider the effect of patch width, w_p is varied from 4.5 mm to 6.5 mm while its length (w_g) is fixed at 0.4 mm. It was found that the bandwidth of the passband increased when the width of the patch was enlarged. These properties can be exploited in the design of slit-patch structures to increase frequency response. In Figure 7(c), the effect of the antenna diameter (D) varied from 65 mm to 105 mm. It was observed that the value of S_{11} decreased. Then, considering the reflection coefficient curves of w_g , w_p , and D , will be used to determine the width of the gap and the size of the small patch, in order to design and configure the antenna as follows in Figure 2: $D = 75$ mm, $w_p = 5.5$ mm, $w_g = 0.4$ mm, $w = 26.2$ mm, $W = \sim 37.4$ mm (for customization choose to configure at 38 mm), $h = 1.6$ mm, $\epsilon_r = 4.4$, and loss tangent ($\tan\delta$) = 0.025.

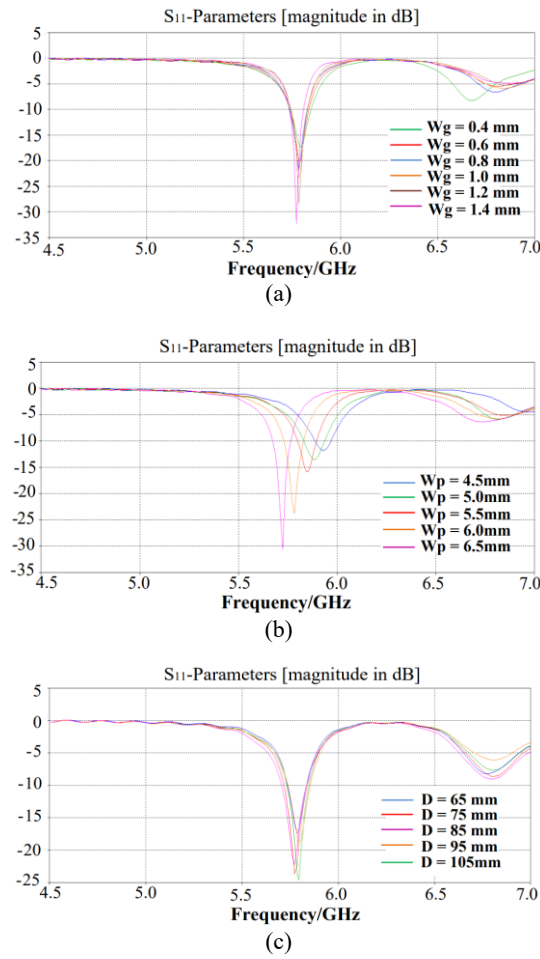


Figure 7. Reflection curves for tuning the gap of slit top, (a) w_g is varied from 0.4 to 1.4 mm, (b) w_p is varied from 4.5 to 6.5 mm ($w_g=0.4$ mm. is Fixed), and (c) D is varied from 65 to 105 mm

In Table 2, the main beam direction and half-power beamwidth depend on the antenna diameter, resulting in a change in the angle by varying the antenna diameter (D) from 45 to 95 mm. When reducing the width D by 10 mm steps, it can be

observed that the gain is reduced, but the degrees of the main beam direction and half-power beamwidth are increased, which is very useful for extending the flight range of UAVs. The 75 mm of substrate and ground are suitable diameters for the design of the beam-steering antennas at a constant tilted beam of 35° and angular width (3 dB) of 52.4° (17.2° - 69.7°) in the elevation plane. The simulation results for the reflection coefficients of the dual beam-steering antennas using a metasurface formed by a small square patch are shown in Figure 8(a). We found that the proposed antenna is able to retain the frequency bandwidth minimum 113 MHz [5.737 GHz to 5.850 GHz] around 5.8 GHz at $S_{11} = -23.17$ dB in Figure 8(b). The simulation results of the 3D and 2D patterns for the elevation and azimuth planes show that the antenna provides a gain of 7.13 dB in Figures 8(c-f).

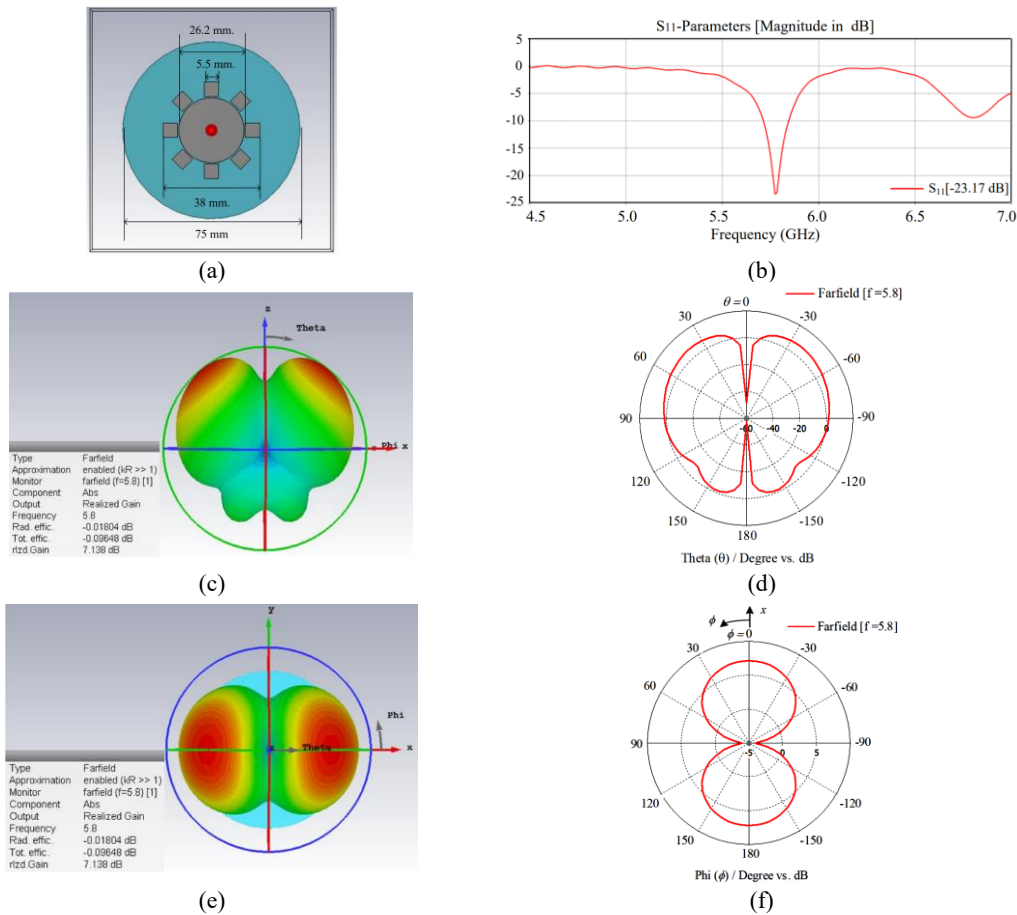
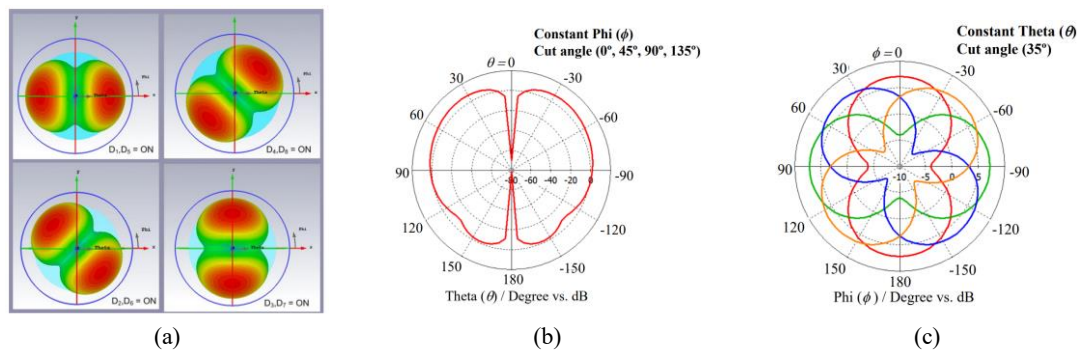
The antenna operates on the 5.8 GHz test frequency band and covers the C-band. The eight small square patches were arranged symmetrically with respect to the center of the antenna, and the center circular patch was fed coaxially. When two small square patches opposite each other are activated, and the feed point of the remaining small square patch is an open circuit by the PIN diode, which is explained by the binary code of the pin diode position shown in Table 3, is a suitable position for the bias pin diode and shows binary code control steering beams. The antenna generates a tilted beam of 35° in all elevation planes and a rotated dual beam at each 45° azimuth plane, as shown in Figure 9(a-b). The antenna can switch the tilted beam in eight directions on the space quadrants in the azimuth plane by exciting forward bias pin diodes in the space quadrants, which are binary sequences. At a test frequency of 5.8 GHz, the antenna provided a gain of 7.13 dB. This antenna is capable of steering its dual beam (1st and 2nd beams) to 0° with 180° , 45° with -135° , 90° with -90° , and 135° with -45° angles and an angular width of 96° at 3 dB in the azimuth plane, which is sufficient for pattern overlapping, as shown in Figure 9(c).

Table 2. Shows the Performance of the antenna diameters (D) are varied from 45 to 95 mm

Performance	The 5.8 GHz dual-beam antenna diameters (D), substrate, and ground					
	95 mm	85 mm	75 mm	65 mm	55 mm	45mm
Gian	7.53 dB	6.63 dB	7.13 dB	6.75 dB	5.80 dB	5.34 dB
Main Beam direction	50°	40°	35°	40°	45°	45°
Half power beamwidth	46°	61.1°	52.4°	52.8°	57.2°	59.2°
	(31.0° - 77.0°)	(14.5° - 76.6°)	(17.2° - 69.7°)	(18.7° - 71.5°)	(20.3° - 77.5°)	(21.5° - 80.7°)
Side lobe level	-7.0 dB	-8.2 dB	-9.1 dB	-8.7 dB	-7.5 dB	-6.7 dB
Total efficiency	-0.095 dB	-0.149 dB	-0.075 dB	-0.033 dB	-0.168 dB	-0.143 dB
Null beamwidth of dual-beam	32°	17°	16°	18°	22°	10°

Table 3. Shows binary code of the PIN diode is a suitable position

Number	D ₁	D ₂	D ₃	D ₄	D ₅	D ₆	D ₇	D ₈	Direction	Main lobe direction	
										Azimuth	Elevation
1	L	H	H	H	L	H	H	H	90°	90°	-35°
2	H	L	H	H	H	L	H	H	45°	45°	35°
3	H	H	L	H	H	H	L	H	0°	0°	-35°
4	H	H	H	L	H	H	H	L	135°	135°	35°

**Figure 8. Simulated characteristics of (a) Geometry of proposed antenna, (b) Reflection coefficients (S_{11}) of the dual beam-steering antennas, (c) 3D elevation plane, (d) Pattern of elevation plane, (e) 3D azimuth plane, (f) Pattern of Azimuth plane****Figure 9. Simulated results of dual beam steering antennas, (a) 3D azimuth plane, (b) Elevation plane and (c) Azimuth plane**

Simulation Results of Antenna with Meta-Radomes

This section presents the simulation results for an antenna with a meta-radome. Starting by simulating each unit cell type at f_c of 5.8 GHz, to analyze the characteristics of the metamaterial structure, the S-parameters were extracted to calculate the effective permittivity (ϵ_{eff}) and effective permeability (μ_{eff}), which are derived as follows (Kamonsin *et al.*, 2020):

$$\epsilon_{eff} = \frac{2}{jkh} \times \left(\frac{1 - (S_{21} + S_{11})}{1 + (S_{21} + S_{11})} \right) \quad (31)$$

$$\mu_{eff} = \frac{2}{jkh} \times \left(\frac{1 - (S_{21} - S_{11})}{1 + (S_{21} - S_{11})} \right) \quad (32)$$

where S_{11} is the reflection coefficient, S_{21} is the transmission coefficient, h is the substrate thickness, and k is the wave-number.

Using CST simulation software, which is based on the finite element method, the geometric parameters of Our proposed unit cells in Figure 6(a), printed on a FR4 substrate ($\epsilon_r = 4.4$, $\tan\delta = 0.025$) was defined as the JC-FSS combined SL-FSS with dimensions of $D = 75$ mm, $d_1 = 10$ mm, $d_2 = 4.2$ mm, $p = 11.8$ mm, $w_1 = 0.55$ mm, $w_2 = 0.6$ mm, $g_1 = 1.8$ mm, $g_2 = 0.8$ mm and $h = 1.6$ mm, which is more compact by almost 1/3 compared to the conventional structure as shown in Figure 10(a), see Figure 10(b) shows the simulated results of S_{11} and S_{21} for SL-FSS and JC-FSS, which both characterize the bandpass filter at f_c 5.8 GHz as observed in the LC equivalent circuit which provides a transmission coefficient S_{21} of band stops at f_r 2.76 GHz and 3.96 GHz frequencies, respectively. They both have a bandpass bandwidth, which is wider than the combination of JC-FSS with SL-FSS, but they both have a large size, as shown in Figure 10(a). Consider Figure 10(c) shows the simulated results of S_{11} for different types of unit cells: the conventional SL-FSS and Combined JC-FSS with square loop, in which the combination of JC-FSS with SL-FSS can reduce the size of the meta radome is decreased, see Figure 10(d) shows S_{11} of g_2 is varied from 0.4 to 1.2 mm, the values of L and C decrease, which is assigned to characterize the reflection curve that occurs in the required frequency band and Figure 10(e) shows the simulated reflection coefficient S_{11} and transmission coefficient S_{21} of JC-FSS with SL-FSS, which characterize the bandpass filter at f_c 5.8 GHz, 425 MHz of bandwidth as observed in the L_1C_1/L_2C_2 equivalent circuit, which provides a transmission coefficient S_{21} of two band stops at 4.4

GHz and 9.0 GHz frequencies, the bandwidth of the pass band can be increased by increasing the distance of the gap (g).

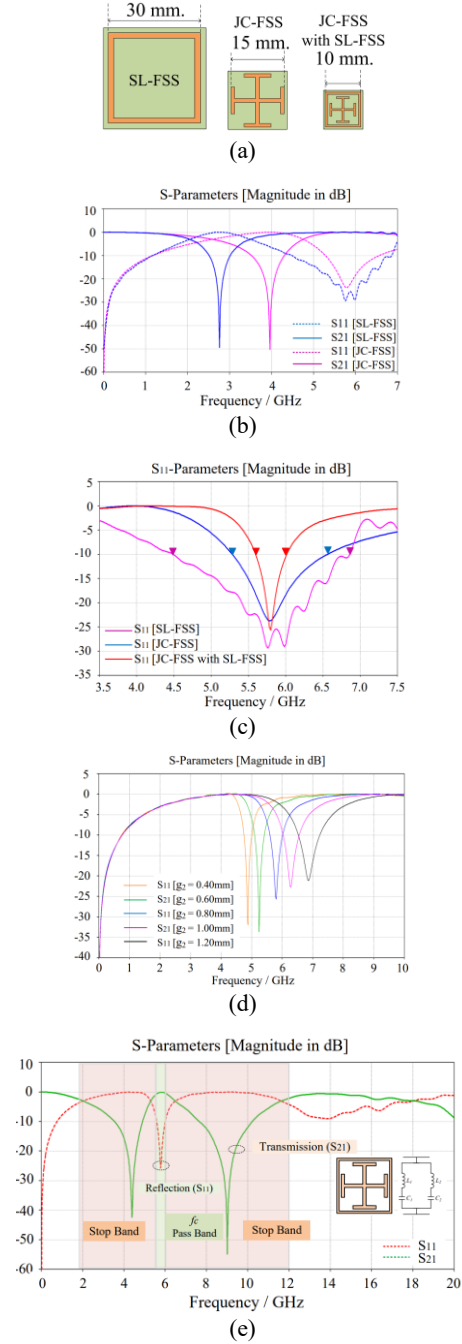


Figure 10. (a) Geometry of proposed types of unit cells, (b) Simulated results of S_{11} and S_{21} for SL-FSS and JC-FSS, (c) S_{11} of SL-FSS, JC-FSS and Combined JC-FSS with SL-FSS, (d) S_{11} of g_2 is varied from 0.4 to 1.2 mm, and (e) The simulated S_{11} and S_{21} of JC-FSS with SL-FSS

In Figure 11, shows the multilayer of the combined JC-FSS with a conventional SL-FSS. The air-gap distance (λ) between each unit cell directly influenced the bandwidth of S_{11} . Any air gap distance can increase the gain of the antenna except when the air gap distance is zero, but we chose a distance between the air gap and h of ($\lambda/4$) for a dual-beam antenna, which provides a high gain, and the propagation pattern has the least distortion.

Another consideration is the origin of the number and layout of the unit cells. Comparing the 2×2 , 3×3 , 4×4 , and 5×5 arrays, in Table 4 notice the 4×4 array arranged $\lambda/4$ from the antenna has a maximum gain of 7.65 dB, which is the main beam direction, and the half power beamwidth is suitable to be used as the meta-radome.

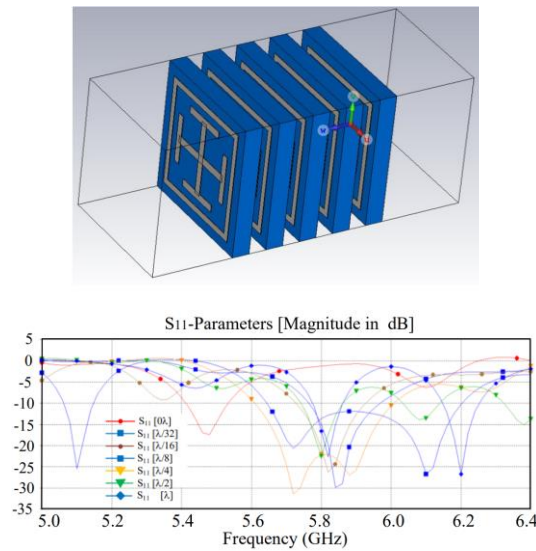


Figure 11. Shows the multilayer of combined JC-FSS with conventional SL-FSS and Simulated results of S_{11} for the unit cell distance arrangement

Table 4. Performance of the unit cells array

Performance of Meta radom at $\lambda/4$	The results for the number of unit cells			
	2x2	3x3	4x4	5x5
Gian (dB)	7.45	7.61	7.65	7.42
Main beam direction	35°	35°	35°	40°
Half power beamwidth	46.9°	48.5°	49.9°	50.0°
Side lobe level (dB)	-9.9	-10.4	-11.0	-11.1

Lastly, Another unusual property of metamaterials is that they have attracted considerable attention. It is a material with a refractive index of zero or near zero (Zero Refractive Index: ZRI or Near Zero Refractive Index: NZI) from the Refractive Index $\eta = \pm \sqrt{\mu/\epsilon}$. Another unusual property of metamaterials is that

they have attracted considerable attention. It is a material with a refractive index of zero or near zero (Zero Refractive Index: ZRI or Near Zero Refractive Index: NZI) from the Refractive Index $\eta = \pm \sqrt{\mu/\epsilon}$. Therefore, cases in which $n=0$ or $n \rightarrow 0$ can occur at all. Three cases: 1) When $\epsilon = 0$ or $\epsilon \rightarrow 0$ over μ is greater than or equal to 1 ($\mu \geq 1$), the case is called ENZ (Epsilon Near Zero), 2) when ($\mu = 0$) or $\mu \rightarrow 0$ over ϵ greater than or equal to 1 ($\epsilon \geq 1$) is called MNZ (Mu Near Zero), and 3) when $\epsilon = \mu = 0$ is called DZI (Double Zero Index) or case $\epsilon \rightarrow 0$ and $\mu \rightarrow 0$ is called MENZ (Mu-Epsilon Near Zero), although in three cases it has a value of $n = 0$ or $n \rightarrow 0$, when considering it, there are other properties that are different, for example, the wave impedance in the case of ENZ, will find $\eta \rightarrow \infty$, in the case of MNZ, $\eta \rightarrow 0$ and $\eta = 1$ in the case of DZI, the case in which $\eta = 1$ may be called an impedance matched to the space wave impedance (Matched-impedance Medium), Based on the above information, we may describe the properties of the material at the macro level with the parameters of electrical permittivity and magnetic permeability (ϵ and μ , respectively), both parameters (Spada, 2017). This can be grouped into groups of materials, as shown in Figure 12(a), in which the metamaterial-based radome is near zero ($\epsilon = 0.0142$ and $\mu = 0.0134$, respectively), as shown in Figure 12(b). It is the NZI for wave propagation in 1st quadrants of terminology and classification.

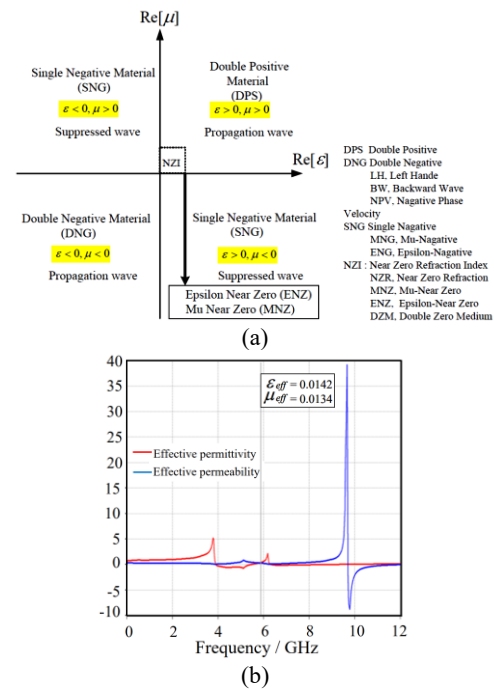
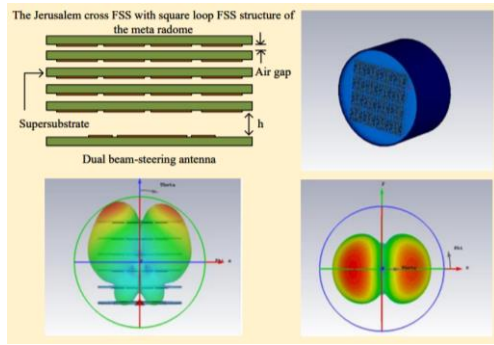


Figure 12. (a) Terminology and classification, and (b) Tables of permittivity and permeability

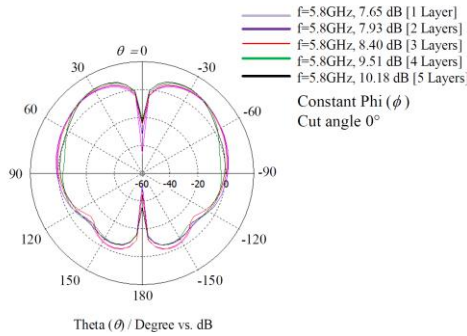
In Figure 13, shows the simulated results of the dual beam-steering antennas with a meta-radome. Figure 13(a) shows the 3D structure of the antennas with a meta-radome. The microstrip antenna was grounded at the bottom, and the meta radome was placed on top of a mirror plate with a very high reflection coefficient and two parallel plates at a reasonable distance (typically half the wavelength). The air gap can be adjusted by the wavelength ratio (Chun-Yih *et al.*, 2007; Foroozesh and Shafai, 2010; Tanan *et al.*, 2018). If the air gap

narrows at the appropriate wavelengths, the meta-radome increases the antenna gain. For the design and experiment, we selected a distance between the air gap and h of ~ 12.9 mm. ($\lambda/4$), which provides high gain, and the propagation pattern has the least distortion. In particular, the angle of the tilted beam must be at the same position or must have the least deviation. Figure 13(b) gain of the antenna to increase when increasing the number of the meta radome which gains are 7.65 dB at 1 Layer, 7.93 dB at 2 Layers, 8.40 dB at 3 Layers, 9.51 dB at 4 Layers and 10.18 dB at 5 Layers in Elevation plane and Figure 13(c) Azimuth plane.

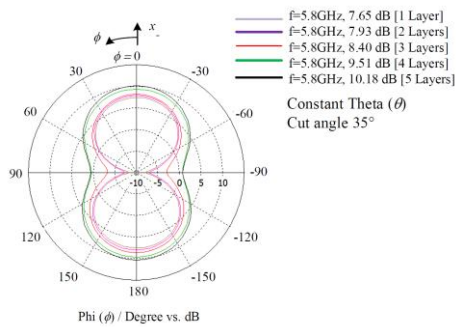
Finally, Figure 14 shows the simulation results at $\phi = 0^\circ$, $\phi = 45^\circ$, and $\phi = 90^\circ$ respectively for the far field of the antennas installed on the two wings of an aircraft. The coupling fields of the two antennas, cause the half-power beam width to have a wider elevation angle, the propagation to cover the entire area according to the elevation angle. Steering according to the azimuth angle will result in waves covering all directions, and the gain of the antennas is increased by approximately 2dB, corresponding to Muhammad *et al.* (2011).



(a)



(b)



(c)

Figure 13. Simulated results of antennas with a meta-radome, (a) 3D antennas structure, (b) Elevation plane, and (c) Azimuth plane

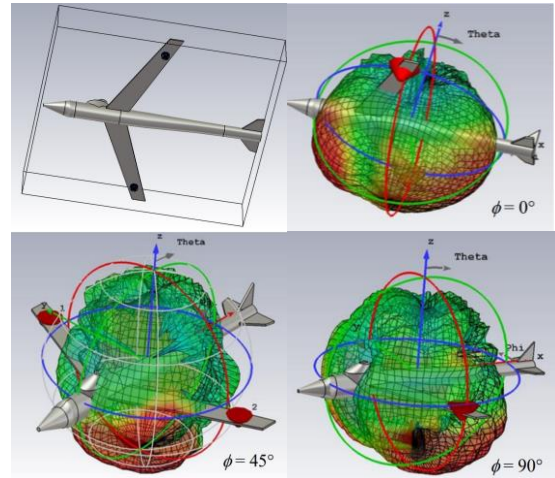


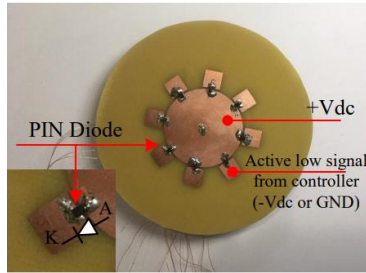
Figure 14. Simulated results of far field for antennas installed on two wings of an aircraft

Fabrications and Tests

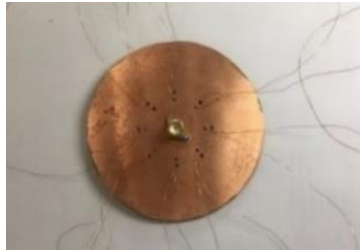
A prototype of the proposed structure was fabricated to test and validate the simulation results, as shown in Figure 15. Figure 15(a) the front of the dual beam-steering antenna with a PIN diode connected between the circular patches and square patches, Figure 15(b) the back of the dual beam-

steering antenna with holes drilled for copper wires to connect to the controller, and Figure 15(c) the prototype of 4×4 unit cells for the fabricated meta-radomes. The proposed unit cells were printed on an FR4 substrate ($\epsilon_r = 4.4$, $\tan\delta = 0.025$, $h = 1.6$ mm).

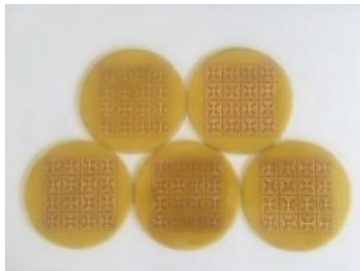
which are shown in Figures 16(c-d), were compared between the factory value and the measurement value, and it can be seen that the PIN diode was transmitted completely.



(a)



(b)

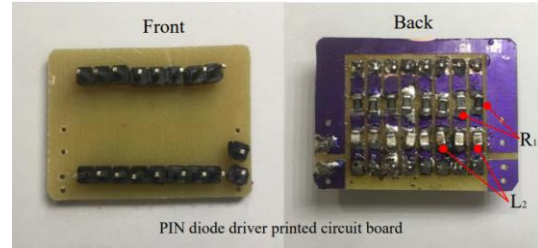


(c)

Figure 15. (a) Front of a dual-beam steering antenna, (b) The back of the dual-beam steering antenna, and (c) The prototype of meta-radomes

PIN Diode Driver Printed Circuit Board

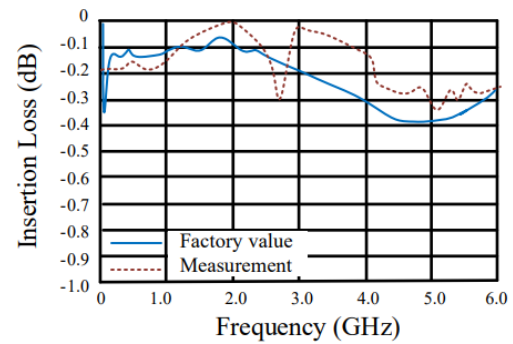
A prototype of the proposed structure for testing and controlling PIN Diode switches according to the designed circuit, as shown in Figure 3, was fabricated, as shown in Figure 16(a). The PIN diode driver printed circuit board consists of the following devices: chip inductor (L_2) is 47 nH, resistor (R_L) is 100 Ω , and chip resistor (R_1) is $\leq 430 \Omega$, Figure 16(b) shows the bias tee module. For forward biasing the diode, a 5 V, DC power supply is required. A current of 50 mA was set for the bias T module and 10 mA for the PIN diodes. The RF transmittance measurements, the results of



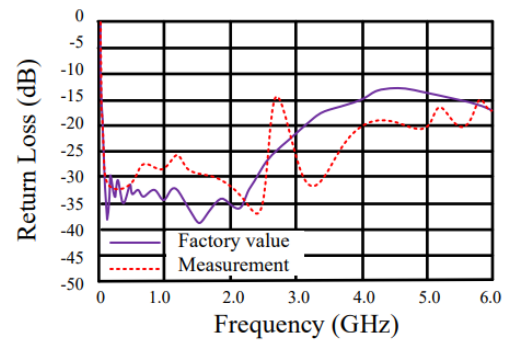
(a)



(b)



(c)



(d)

Figure 16. Shows (a) PIN Diode printed circuit board, (b) Bias tee module, (c) The insertion loss of PIN diode for forward bias, and (d) The return loss of PIN diode for forward bias

The Measurement Setup for Fabricated Dual Beam Antenna

To characterize the reflection and transmission coefficients of a dual-beam steering antenna. The measurement setup is shown in Figure 17. A prototype of the dual-beam steering antenna was assembled and tested in a semi-acoustic chamber at Telecommunications engineering laboratory, Suranaree University of Technology, Thailand. The receiving and transmitting antennas are located at a distance in the far-field region for measuring S_{11} and S_{21} of the dual-beam steering antenna, which is connected to port 1 of a vector network analyzer, whereas the receiving ultra-wideband tapered slot antenna is connected to port 2.



Figure 17. Shows the measurement setup for fabricated dual beam antenna

A simulation with the measurement results of the reflection coefficients (S_{11}) of the dual beam-steering antennas is shown in Figure 18(a). We found that the proposed antenna is able to retain the frequency bandwidth minimum 113MHz [5.737 GHz to 5.850 GHz] around 5.8 GHz at $S_{11} = -23.17$ dB and gain is 7.13 dB. Figure 18(b-c) shows measured radiation patterns in the elevation plane and azimuth plane are $S_{11} = -28.5$ dB and gain are 7.21 dB at 0° , 90° and gain are 7.18 dB at 45° , -45° , and the frequency bandwidth is ~ 134 MHz [5.722 GHz to 5.856 GHz]. It provides a constant tilted beam of 35° with an angular width of 50° (19° - 69°) in the elevation plane, and an angular width of $\sim 94^\circ$ at 3 dB in the azimuth plane.

Finally, for the design and experiment, we selected a distance between the air gap and h of ~ 12.9 mm. ($1/4\lambda$), which provides high gain, and the propagation pattern has the least distortion. Figure 19 shows a prototype of the proposed structure for testing the antenna with the super-substrate multi layer meta-radomes by stacking five meta-radomes at a spacing of 12.9 mm, which improves the efficiency of the electronically steered dual-beam antenna.

The measurement setup for the fabricated dual-beam antenna with a meta-radome of a combined JC-FSS with SL-FSS is shown in Figure 20(a). The simulation results for the

reflection coefficients of the dual beam-steering antennas using a meta-radome are shown in Figure 20(b). We found that the proposed antenna is able to retain the frequency bandwidth minimum 125 MHz [5.729 GHz to 5.854 GHz] around 5.8 GHz at $S_{11} = -33.5$ dB and gain is 10.18 dB. Figure 20(c-d) shows measured radiation patterns in elevation plane and azimuth plane measured radiation patterns in the elevation plane and azimuth plane are $S_{11} = -33.8$ dB and gain are 10.23 dB at 0° , 90° and gain is 10.19 dB at 45° , -45° and the frequency bandwidth is ~ 138 MHz [5.724 GHz to 5.862 GHz]. While It provides a constant tilted beam of 35° with an angular width of 40° (15° - 55°) at 3 dB in the elevation plane, and an angular width of $\sim 74^\circ$ at 3 dB in the azimuth plane, which is sufficient for pattern overlap.

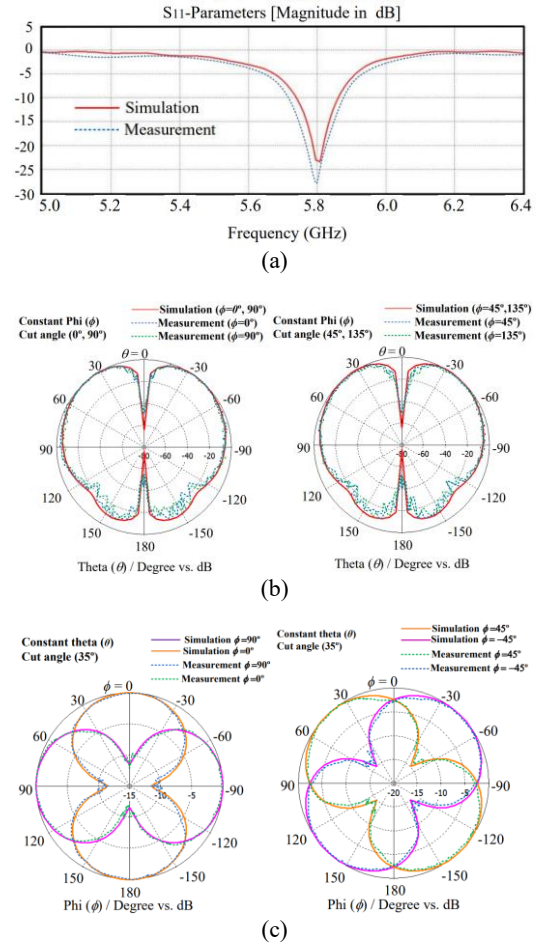


Figure 18. Shows (a) S_{11} of the simulation with measurement, (b) Radiation patterns in elevation plane at 0° , 90° , and 45° , -45° , and (c) Radiation patterns in azimuth plane at 0° , 90° , and at 45° , -45° .

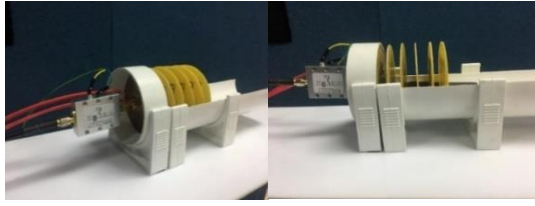
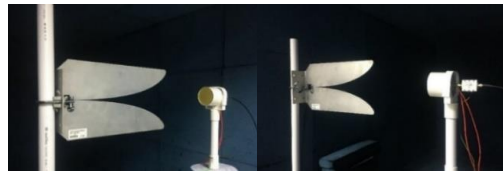
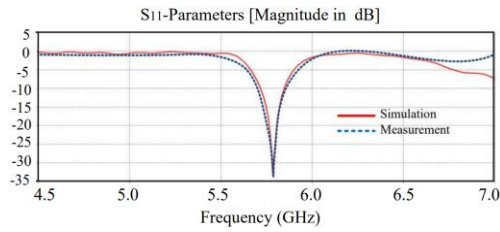


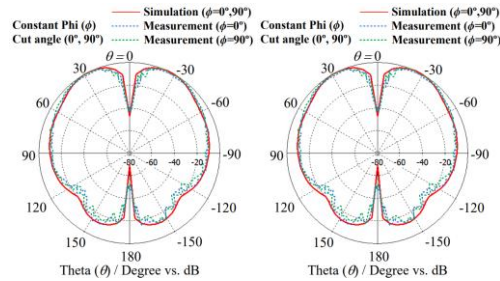
Figure 19. A prototype of proposed structure for testing antenna with the multi layer radomes



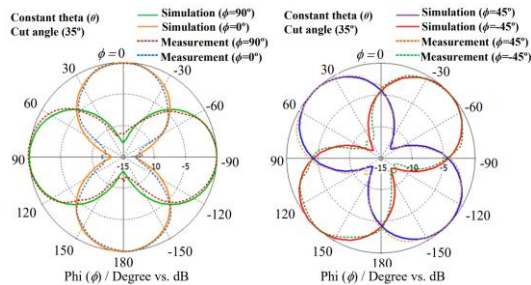
(a)



(b)



(c)



(d)

Figure 20. Shows (a) The measurement setup for fabricated dual beam antenna with meta-radome, (b) S_{11} parameter of antenna with meta radome, (c) Radiation patterns in elevation plane, at 0° , 90° and 45° , -45° , and (d) Radiation patterns in azimuth plane at 0° , 90° and at 45° , -45°

Conclusions

A technique for Efficiency improvement of an electronically steered dual-beam antenna using a metamaterial radome operates over a test frequency band of 5.8 GHz, with bandwidths of approximately 113 MHz (at -10 dB), which covers the C-band, which is high enough for the UAVs applications, such as antennas installed on two wings of an aircraft that can be tracked or commanded from its control station within a given range and without disconnection. This antenna can increase or decrease the beam width at the elevation angle by adjusting the diameter of the dielectric and ground, expandable 35° - 50° of the main beam direction and 59.2° (21.5° - 80.7°) of the maximum half-power beam width. Our antennas can provide higher gain than antennas arranged with the unit cell array in the same plane and a typical UAV antenna has a gain of only ~ 4 -7 dB. Eight square patches were arranged symmetrical with respect to the center of the antenna, and the center circular patch was fed coaxially. When activating two small square patches opposite to each other and the feed point of the remaining small square patch is an open circuit by the PIN diode or terminated at 50Ω impedance, the antenna is investigated at 5.8 GHz of frequency, which generates a tilted beam of 35° in the elevation plane. While the antenna provides a gain of ~ 7.21 dB without a meta-radome and ~ 10.23 dB with a meta-radome, the frequency bandwidth is ~ 138 MHz around 5.8 GHz. This antenna is capable of steering its dual beam (1st and 2nd beams) to 0° at 180° , 45° at -135° , 90° at -90° , and 135° at -45° angles, which the antenna has permittivity and permeability, is near zero ($\epsilon = 0.0142$ and $\mu = 0.0134$), respectively.

The direction of future research, the designed meta-radome can be applied to a variety of antennas, and provide high gain and accurate beam direction with very little error, such as a single beam antenna, the gain can be increased by more than 3 dB by using only one to three layers or fewer than five layers of meta-radomes. This combination technique is applied to combining other unit cell types to obtain the desired material characteristics. The size and weight of the antenna and radome can be reduced by using thin or flexible PCBs.

Acknowledgement

This work was supported by the Research Department Institute of Engineering, Suranaree University of Technology, Nakhon Ratchasima, Thailand.

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