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Metamaterial based agile frequency selective surfaces

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مُلَخَّص

مِنْ بَيْنَ أَهَمِّ تَطْبِيقَاتِ أَشْبَاهِ الْمَوَادِّ فِي مَجَالِ الْهَوَائِيِّ هِيَ تَحْسِينُ أَدَاءِ الْهَوَائِيِّ وَتَقْلِيلُ حَجْمِهِ، لَطَالَمَا التَّحَكُّمُ فِي خَصَائِصِ الْهَوَائِيِّ كَانَ هَدَفًا رَئِيسِيًّا وَذَلِكَ لِاسْتِخْدَامِ أَشْبَاهِ الْمَوَادِّ فِي مَجَالِ الْهَوَائِيِّ. وَالْهَدَفُ مِنْ ذَلِكَ هُوَ تَصْنِيمُ نِطاقِ رَادَارٍ مِنْ صَفَائِحِ إِنْتِقَائِيَةِ التَّرْدُدِ بِاسْتِعْمَالِ أَشْبَاهِ الْمَوَادِّ، إِذْ يُمَكِّنُنَا تَغْيِيرُ خَصَائِصِهَا بِاسْتِخْدَامِ نِظَامِ تَحَكُّمٍ خَارِجِيٍّ. وَهَذَا مَا يَجْعَلُ أَنْظِمَةَ الْإِتِّصَالَاتِ أَكْثَرَ مُرَوْنَةً، إِنْتِمَامًا فِي هَذِهِ الْأَطْرُوحَةِ بِالصَّفَائِحِ إِنْتِقَائِيَةِ التَّرْدُدِ بِاسْتِعْمَالِ أَشْبَاهِ الْمَوَادِّ. إِذْ يَتِمُّ تَقْدِيمُ خُفَّةِ التَّرْدُودِيَّةِ عَنْ طَرِيقِ اسْتِخْدَامِ أَشْبَاهِ مَوَادِّ تُتَبَّحُ لِلْمَرْءِ ضَبْطُ الْإِسْتِجَابَةِ الْكَهْرُومَغْنَاطِيَّيَّةِ فِي وَقْتٍ أَنْيٍّ. تُسَاعِدُ قُدْرَةُ التَّغْيِيرِ هَذِهِ عَلَى جَعْلِ الْأَجْهَزَةِ الْمُصَنَّعَةِ بِأَشْبَاهِ الْمَوَادِّ أَكْثَرَ تَنْوَعًا وَتَكْيُفًا أَوْ لِتَغْيِيرِ إِعْدَادَاتِهَا وَفَقًّا لِإِحْتِيَاجَاتِنَا.

Abstract

Among the most important metamaterial application in the antenna field we can cite the antenna performances enhancement and size reduction. However, antenna characteristics control remains a major objective of metamaterial use in the antenna field. The goal is to design an agile FSS radome based on metamaterial layers, of which it is possible to change the characteristics using an external active control system. This makes the telecommunications systems more flexible, we are interested in this thesis, with the FSS agility. The frequency agility is introduced by using agile metamaterials which allow one to adjust the electromagnetic response in real time. This turning ability helps make metamaterial devices more versatile, adapting to shifting input or changing target parameters according to our needs.

Keywords: Metamaterial, Agile frequency selective surface, radome.

Résumé

Parmi les applications métamatérielles les plus importantes dans le domaine des antennes, nous pouvons citer l'amélioration des performances de l'antenne et la réduction de sa taille. Cependant, le contrôle des caractéristiques de l'antenne reste un objectif majeur de l'utilisation de métamatériaux dans le domaine d'antenne. L'objectif est de concevoir un radôme agile du SFS basé sur des couches de métamatériaux, dont il est possible de modifier les caractéristiques à l'aide d'un système de contrôle actif externe. Cela rend les systèmes de télécommunications plus flexibles, nous nous intéressons à cette thèse, avec l'agilité FSS. L'agilité en fréquence est introduite en utilisant des métamatériaux agiles qui permettent d'ajuster la réponse électromagnétique en temps réel. Cette capacité de tournage contribue à rendre les dispositifs à métamatériaux plus polyvalents, en s'adaptant au changement d'entrée ou en modifiant les paramètres de la cible en fonction de nos besoins.

Mots clés: Métamatériaux, Surface sélective en fréquence, Radôme.

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Symbols and Abbreviations

α	The damping factor
BSE	Boresight Error
β_0	The propagation constant of the line without inclusion of the magnetic material
β_i	The propagation constant of the line with inclusion of the magnetic material
$\vec{B}$	Magnetic induction
$\langle B \rangle_v$	Average of magnetic induction in volume V
CRLH	Composite Right Left Handed
$\langle D \rangle_p$	Average electrical displacement D in the surface P
DSL	Double Square Loop
DNG	Double Negative
dB	Decibel (radiation gain unit)
$\vec{E}$	The electric field
$\langle E \rangle_v$	Average electric field E in volume V
EMI	Electromagnetic Interference
ENG	Epsilon Negative
ENZ	Epsilon Near Zero
FSS	Frequency Selective Surface
FEM	Finite Element Method
FDTD	Finite Difference Time Domain
FSM	Field Summation Method
GHZ	Gigahertz (frequency unit)
$\vec{H}$	The magnetic field
$\vec{H}_i$	The high frequency internal magnetic field

$\overrightarrow{H}_a$	The continuous internal magnetic field applied
HF	Higher Frequency
$\langle H \rangle_p$	Average of the magnetic field H in the surface P
HFSS	High Frequency Structure Simulator
ISL	Incurved Square Loop
ISM	The Industrial Scientific and Medical
LF	Lower Frequency
LHM	Left Handed Materials
MPA	Microstrip Patch Antenna
MoM	Method of Moments
MEMS	Micro Electro Mechanical Systems
MNG	Mu Negative
MNZ	Mu Near Zero
MIMO	Multiple Input Multiple Output
NRIM	Negative Refractive Index Materials
NMR	Nuclear Magnetic Resonance
NRI	Negative-Refractive Index
n	The refractive index
P_t	Transmitted powers
$(p_a)_{FD}$	Absorbed powers in the forward
$(p_a)_{BW}$	Absorbed powers in the backward
$(p_r)_{FD}$	Reflected powers in the forward
$(p_r)_{BW}$	Reflected powers in the backward
PEC	Perfect Electric Conductor
RCS	Radar Cross Section
RF	Radio Frequency

RFID	Radio Frequency Identification
SRR	Split Ring Resonator
SNG	Single Negative
T^+	The input coefficient of transmission
T^-	The transmission coefficient at the output
T	The absolute temperature
UHF	Ultra High Frequency
VSWR	Voltage Standing Wave Ratio
V_R	Volume of dielectric resin
V_F	Volume of reinforcement fibers
VHF	Very High Frequency
ω_0	The pulsation of Larmor
χ	Magnetic susceptibility
Γ^+	The reflection coefficient at the input
Γ^-	The reflection coefficient at the output
μ_0	The magnetic permeability of the vacuum
μ_r	Relative magnetic permeability
ϵ_r	The relative dielectric permittivity
λ	The wavelength in free space
ϵ_{eff}	Effective permittivity
μ_{eff}	Effective permeability
Z^+	Impedance in the direct direction
Z^-	The impedance in the opposite direction
z	impedance
Z_{FD}	Impedance for the forward propagation

Z_{BW}	Impedance for the backward propagation
σ_e	Electrical conductivity
δ	Loss tangent

Introduction

In recent years new concepts in synthesis and novel fabrication techniques have allowed the construction of structures and composite materials that mimic known material responses or that qualitatively have new physically realizable response functions that do not occur or may not be readily available in nature, it called metamaterials.

Metamaterials are artificial composite materials that are usually constructed from sub-wavelength size periodic structures and attain their properties from the unit cell shape size rather than chemical composition. They can exhibit effective material properties i.e. effective permittivity (ϵ_{eff}) and effective permeability (μ_{eff}), which are not found in naturally existing materials. These unusual properties have been achieved by optimizing the dimensions of unit cell.

Electromagnetic metamaterials are broadly defined as artificial effectively homogeneous electromagnetic structures. An effectively homogeneous structure is a structure whose structural average cell size are much smaller than the guided wavelength (all geometric dimensions of unit cell should be at least smaller than a quarter of wavelength λ).

These sub-wavelength dimensions of the unit cell lead to good absorption characteristics. Moreover, the sub-wavelength size makes the reflection and refraction phenomenon more significant than the diffraction and scattering. By manipulating the loss components of effective permittivity and magnetic permeability numerous potential applications can be addressed. The unusual properties of the metamaterials such as negative permittivity and permeability and their range of applications which covers microwave to optical frequency regime have aroused the extensive interest into these artificially engineered periodic structures over the past decades.

Designing spatial filters (frequency-selective surface FSS) is an important application of metamaterial. Once exposed to the electromagnetic radiation, a frequency-selective surface FSS acts like a spatial filter, some frequency bands are transmitted and some are reflected. Noted that, dielectric materials instead of metallic patterns are capable of achieving more functional performance in FSS design, moreover it can be used in different extreme environments, depending on their electrical, thermal or mechanical properties.

The first concept which should be understood is "Periodic Surface" since it constitutes the nature of FSS structures, a periodic surface consists of identical elements of infinite arrays in one, two or three dimensions, although, the name FSS was not used until much later on. In a way, an FSS can be a cover for hiding communication facilities. This is probably the top ten

potential applications of FSS structures, as they have actually been used as covers named radomes.

Radomes are bandpass FSS filters that are used to reduce the radar cross-section (RCS) of an antenna system outside its frequency band of operation. FSS characteristics control is a major objective of metamaterial use. Our objective is to design a radome invisible for the electromagnetic waves as much as possible, on the other hand, improve the performances of a patch antenna such as the realized gain and the radiation pattern in addition to introduce frequency agility, hence, a range of different operating band frequencies controlled using an external command system.

Metamaterial plays a significant role in improving the electromagnetic properties of radomes, so metamaterial, employed for radome material, is called metaradome.

The design of metaradome should be developed and adjusted to different application areas such as maritime, telecom, radar, aerospace, and automotive applications. Owing to their high gain behavior the proposed metamaterial radome is applicable in several radar systems. Metaradomes may help in shaping the directivity of antennas. Directional antennas are used in wireless networks especially for the 5G technology such as beam forming ..etc.

We are interested in this thesis, with the FSS agility. The agility is introduced by using agile metamaterials which allow one to adjust the electromagnetic response in real time. This turning ability helps make metamaterial devices more versatile, adapting to shifting input or changing target parameters and is the first step to go forward smart devices. Such smart devices can perform self-matching to the surrounding environment.

The complete thesis is divided into four chapters;

Chapter one, covers an overview of metamaterials, a brief history and their applications and characteristics has been provided.

In chapter two, due to the use of metamaterial on the frequency selective surface design, a general review for this last has been covered, therefore, the design approach and methodology for conformal frequency selective surfaces has also been investigated.

In chapter three, at the first, the necessity of the effective parameters extraction methods has been discussed. Thereafter the characterization of the elementary unit cell of the agile metamaterial layer by varying its physical dimensions, such as the period, the dielectric thickness, the width ...etc.

In the last chapter, a short introduction about the radomes and their characteristics has been discussed, In order to achieve an agile frequency antenna/radome, we design a metamaterial layer which has two different behaviors that can be controlled through external command system. This agile metamaterial layer is obtained by a periodic structuration based on the agile metamaterial cross type unit cell.

The agility in frequency is achieved by changing the behavior of a metamaterial FSS radome layer to be in the connected or disconnected state by modifying the wires of the cross unit cell. This agile metamaterial layer, based on connected or disconnected agile cross unit cells, can behaves as an effective medium with a refraction index close to zero (n_1) or greater than unity (n_2), respectively, depending on the switches states. With this agile unit cell it is possible to design a metamaterial multilayer FSS radome in which the behavior of each layer can be programmed individually independently from the others.

Moreover, increasing the number of layers constituting the metamaterial FSS radome and changing their behavior to be in the connected or disconnected state can improve the functioning of the patch antenna according to our needs, such as, the realized gain, the directivity and the radiation beam width.

1.1. Introduction

All natural materials are composed of atoms and can basically be classified into two types, according to these particles different arrangements: crystals (with periodic arrangement of atoms) and noncrystals (with random arrangement of atoms). In 1985, Daniel Shechtman, Nobel Prize winner in chemistry, discovered an intermediate state between crystals and noncrystals-quasicrystal, which is composed of nonperiodic atoms with certain orders [1], showing excellent performance beyond that of natural materials.

Unlike natural materials, metamaterials are macroscopic composites of periodic or nonperiodic subwavelength-scale artificial particles with certain shapes, materials, and geometries, which are usually called as meta-atoms. The main advantages of metamaterials are as follows. (i) Meta-atoms can be engineered and tailored as required, and hence have countless choices. However, the natural atoms are very limited. (ii) Meta-atoms can be arbitrarily arranged due to their flexibilities. But it is difficult to arrange natural atoms to specific orders. Therefore, we can realize homogeneous metamaterials (super crystals), inhomogeneous metamaterials (super quasicrystals), and random metamaterials.

Usually, meta-atoms are electric and/or magnetic resonators. Depending on the resonant status of meta-atoms, metamaterials are classified into resonant metamaterials and nonresonant metamaterials. For the resonant metamaterials, the meta-atoms are in the resonant state with significant changes of electromagnetic responses, which will produce extreme values of effective permittivity and/or permeability (e.g., negative permittivity, negative permeability, and zero index of refraction). Around the resonant frequency, large dynamic range of effective medium parameters can be achieved, from negative to largely positive. Such parameters will result in unusual physics, such as negative refraction [2], perfect imaging [3], and tunneling effect [4].

Due to the resonance nature. However, resonant metamaterials are always accompanied by large loss and narrow bandwidth. For nonresonant metamaterials, the meta-atoms are away from the resonant state, yielding gradient changes of the electromagnetic responses. This feature produces small loss and wideband medium parameters, which makes metamaterials practical in engineering applications. The drawback of nonresonant metamaterials is the small dynamic range of medium parameters. However, the free arrangement of meta-atoms makes it possible to realize arbitrarily inhomogeneous media.

Only when the meta-atoms have strictly circular (or spherical) symmetries, are the resulting two-dimensional (2D) or three-dimensional (3D) metamaterials isotropic. However, most practical meta-atoms do not have such symmetries, and hence most metamaterials are anisotropic. The extreme values of effective medium parameters, high inhomogeneity, and strong anisotropy enable metamaterials to have powerful abilities to control the medium parameters, and further manipulate electromagnetic waves.

1.2. A brief history

Metamaterials were first known as left-handed materials (LHM) or negative-refractive index materials (NIM). Although the concept of LHM was proposed by Veselago in 1968 [5], the negative refraction and backward-wave media had been discovered theoretically much earlier [6][7][8] [9]. The earliest publication on negative refraction was in lecture notes of Prof. Mandelshtam from Moscow University[7]. Then the Sommerfeld radiation condition in backwardwave media was studied by Malyuzhinets in 1951 [8], who showed that the phase velocity of waves pointed from infinity to the source. Furthermore in 1957, Sivukhin investigated materials with negative parameters and noticed the backward-wave property [9]. Except the theoretical study, backward-wave structures had been developed for the design of microwave tubes in 1960s [10][1][11], and the negative refraction was discovered even in periodical media [12]. A significant progress was made by Veselago in 1968, who proposed a systematic analysis of electromagnetic behaviors in materials with negative permittivity and permeability, and named the term of LHM [5]. Besides the negative refraction and backward-wave propagation, Veselago also showed some new features like reversed Doppler shift and backward Cerenkov radiation in LHM.

However, due to the non-existence in nature and lack of experimental verification, LHM had not attracted attention over 30 years in the scientific community.

The first revolution on LHM or metamaterials occurred in 1996, when Pendry et al. realized the artificially electric plasma using the wire medium whose permittivity is negative [13]. Then Pendry and his coworkers discovered the artificially magnetic plasma whose permeability is negative in 1999[14]. In this work, the well-known split-ring resonators (SRR) are used to achieve the magnetic response. The first artificial LHM was made by Smith et al. in 2001 using the combination of wires and SRRs [15]. In this famous experiment, the negative refraction phenomenon was verified. Inspired by the experimental realization, LHM has attracted growing attention in both theoretical exploration and experimental study

[15][16], including the exciting discovery of perfect lens and super lens [16]. However, LHM has unavoidable disadvantages of big loss and narrow bandwidth, and such disadvantages restrict the applications of LHM.

In the meantime, an alternative representation of LHM was presented by three groups (Eleftheriades, Oliner, and Caloz-Itoh), almost simultaneously in June 2002, using the transmission-line (TL) approach [17][18][19]. It is well known that a conventional TL is composed of distributed series inductance (L) and shunt capacitance (C), which can be shown equivalent to a one-dimensional (1D) RHM. As a dual model of the conventional TL, a new TL consisting of series capacitance and shunt inductance will support backward waves and hence can represent LHM. In the realization of left-handed (LH) TL, the distributed components have been used. Since the series capacitance is always accompanied by an inductance and the shunt inductance is accompanied by a capacitance, a general composite right-left handed (CRLH) TL model has been proposed to represent RHM and LHM [18][19]. Based on the CRLH structures, a lot of microwave components and antennas have been proposed [20].

Despite the good properties, LHM suffers from the large loss and narrow bandwidth. Hence scientists seek other features of metamaterial beyond the negative refraction. Then the second revolution on metamaterial came in 2005 when the gradient refractive index medium was realized to bend electromagnetic waves [21], and in 2006 when the optical transformation was proposed to make invisible cloaks and to control the propagation of electromagnetic waves using metamaterials [21]. From then on, metamaterials have a much broader meaning than LHM, which do not require the negative permittivity and/or negative permeability, and hence open a completely new area. After the experimental realization of invisible cloak in the microwave regime [22][23], a large explosion of interest has been paid to metamaterials and the optical transformation, and many scientific papers have been published in journals.

1.3. Metamaterial characteristics

The structural units of metamaterials can be tailored in shape, size, and the composition. The inclusions composition and morphology can be designed and artificially tuned, and placed in a predetermined manner to achieve prescribed functionalities.

The dependence of metamaterial properties on the unit cell architecture provides great flexibility to control metamaterials electromagnetic responses which are described by the electric permittivity and the magnetic permeability. So, by varying the different unit cell

geometry parameters, we can create an agile metamaterial which electromagnetic response is unavailable in nature and can be tailored in practice to a reach desired profile. Metamaterial tailoring is one of the important advantages of metamaterials.

1.4. Metamaterial classes

Permittivity ε and permeability μ are two parameters used to characterize the electric and magnetic properties of materials interacting with electromagnetic fields.

The permittivity is a measure of how much a medium changes to absorb electrical energy when it is subjected to an electric field [24][25]. It is defined as the ratio of the electric displacement $\vec{D}$ and the electric field $\vec{E}$. The common term dielectric constant is the ratio of permittivity of the material to that of free space ε_0 , which is approximately 8.854×10^{-12} F/m. It is also termed as the relative permittivity. The permeability is also defined as the ratio of the magnetic induction $\vec{H}$ and magnetic field $\vec{B}$. The free-space permeability μ_0 is approximately 1.257×10^{-6} H/m. R. Ziolkowski [26][27] has categorized metamaterials by their constitutive parameters as shown in figure 1-1. Most of the materials in nature have positive permittivity and permeability, and hence, they are referred to as ‘double-positive’ (DPS) medias or right-handed materials (RHM). In contrast, if both of these quantities are negative, these medias are called ‘double-negative’ (DNG) and are also referred to as left-handed materials (LHM). Finally, materials with one negative parameter are named ‘single-negative’ (SNG) and are further classified into two subcategories, namely, ‘epsilon-negative’ (ENG) an ‘mu-negative’ (MNG). Interestingly, natural materials such as cold plasma and silver exhibit negative permittivity at microwave and optical frequencies, respectively, and ferromagnetic materials exhibit a negative permeability behavior in the VHF and UHF regimes. However, to date, no materials that exhibit simultaneous negative permittivity and permeability have been found in nature, and hence they must be produced artificially.

The first comprehensive review of the history of negative refraction and metamaterials was given by Moroz [28]. He indicated that some of metamaterial research started long before Veselago’s work and went back to as far as 1905, when Lamb [29] suggested the existence of backward waves, which are associated with waves in which the phase propagates in a direction opposite to that of the energy flow. Moroz’s on-line article first appeared in 2003, tracing the roots of research on backward waves, negative refraction, and LHM. Only

recently, Veselago and Narimanov [30], Shivola [31], and Shamonina [32] have linked LHMs, and more generally metamaterials, to the earlier works.

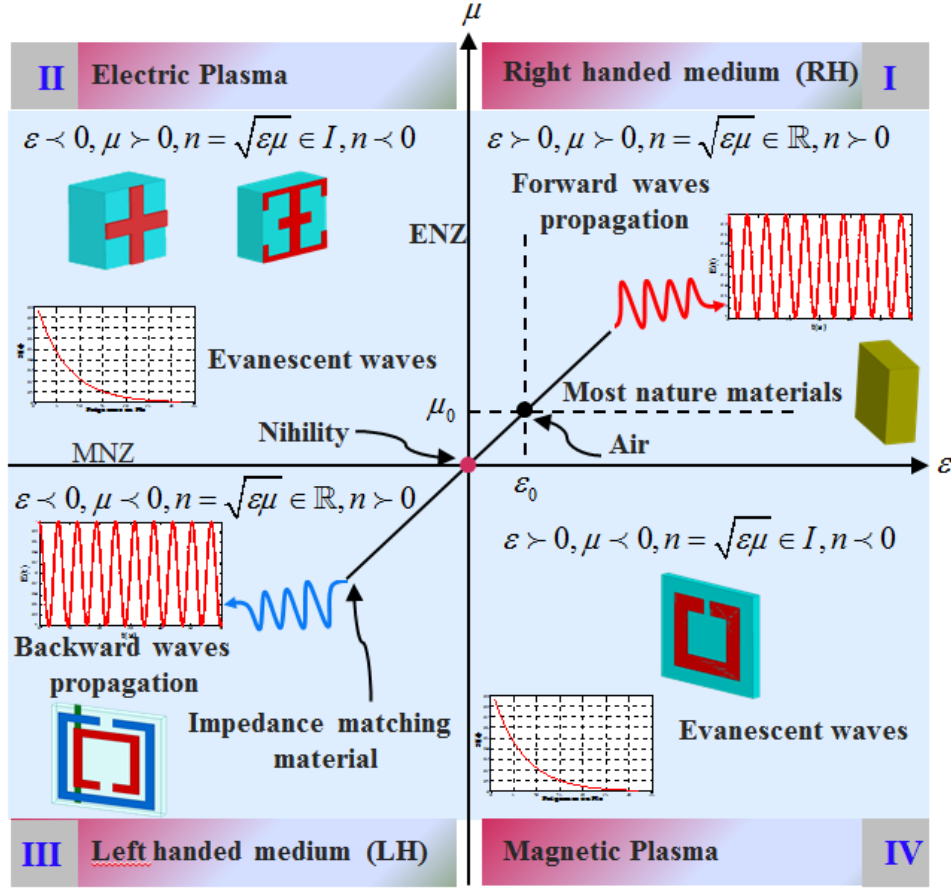


Figure 1-1: Metamaterial classes

From figure 1-1 four possible quadrants can be taken:

- Quadrant 1: double-positive media or right-handed materials $\varepsilon > 0$ and $\mu > 0$.

This quadrant represents (RHM), which support the forward propagating waves. It is well know that the constitutive parameters ε and μ are related to the refractive index n by:

$$n = \sqrt{\varepsilon_r \mu_r} \quad (1-1)$$

We have $\varepsilon > 0 \Rightarrow \varepsilon = |\varepsilon|$ and $\mu > 0 \Rightarrow \mu = |\mu|$.

From (1-1) we obtain:

$$n = \sqrt{|\varepsilon_r| |\mu_r|} = +\sqrt{|\varepsilon_r| |\mu_r|} \quad (1-2)$$

$$k = \omega \sqrt{|\varepsilon_r| |\mu_r|} = k_0 n \quad (1-3)$$

$$\gamma = jk = j\omega\sqrt{|\varepsilon_r| |\mu_r|} = \alpha + j\beta \quad (1-4)$$

$$\gamma = j\omega\sqrt{|\varepsilon_r| |\mu_r|} = \alpha + j\beta \Rightarrow \alpha = 0 \text{ and } \beta = \omega\sqrt{|\varepsilon_r| |\mu_r|} = \gamma \quad (1-5)$$

We have clearly propagating waves (propagating medium $\alpha = 0$). The expression of a propagating plane wave in such media is written as:

$$\vec{E} = \vec{E}_0 \exp(-\gamma z) = \vec{E}_0 \exp(-jkz) = \vec{E}_0 \exp(-jk_0 n z) \quad (1-6)$$

- Quadrant 2: epsilon-negative medium or electric plasma $\varepsilon < 0, \mu > 0$

In certain frequency regimes many plasmas exhibit this characteristic. For example, noble metals (e.g., silver, gold) behave in this manner in the infrared (IR) and visible frequency domains. This quadrant supports evanescent waves.

We have $\varepsilon < 0 \Rightarrow \varepsilon = -|\varepsilon|$ and $\mu > 0 \Rightarrow \mu = |\mu|$

From (1-1) we obtain:

$$n = \sqrt{-|\varepsilon_r| |\mu_r|} = j\sqrt{|\varepsilon_r| |\mu_r|} \quad (1-7)$$

$$k = j\omega\sqrt{|\varepsilon_r| |\mu_r|} = k_0 n \quad (1-8)$$

$$\gamma = jk = j \times j\omega\sqrt{|\varepsilon_r| |\mu_r|} = \alpha + j\beta \quad (1-9)$$

$$\gamma = j \times j\omega\sqrt{|\varepsilon_r| |\mu_r|} = \alpha + j\beta \Rightarrow \beta = 0 \text{ and } \alpha = -\omega\sqrt{|\varepsilon_r| |\mu_r|} < 0 \quad (1-10)$$

And then from (1-7) the expression of a propagating plane wave in such media is written as:

$$\vec{E} = \vec{E}_0 \exp(-\alpha z) \quad (1-11)$$

Since α is negative the term $\exp(-\alpha z)$ diverges when $z \rightarrow +\infty$. The evanescent waves are amplified in an ENG media.

- Quadrant 3: double-negative media or left-handed materials $\varepsilon < 0, \mu < 0$

This third case was proposed by Veselago in 1968, supporting the backward propagating waves. We have:

$$\varepsilon < 0 \Rightarrow \varepsilon = -|\varepsilon| \text{ and } \mu < 0 \Rightarrow \mu = -|\mu|$$

From equation (1-1) we obtain:

$$n = \sqrt{-|\varepsilon_r| \times -|\mu_r|} = \sqrt{|\varepsilon_r| |\mu_r|} \quad (1-12)$$

$$k = \omega \sqrt{\varepsilon_r \mu_r} = \omega \sqrt{|\varepsilon_r| |\mu_r|} \quad (1-13)$$

Comparing equations (1-9) and (1-13) then we obtain

$$\alpha = 0 \text{ and } \beta = \omega \sqrt{|\varepsilon_r| |\mu_r|} = \gamma \quad (1-14)$$

Then, the expression of the electric field corresponds to the propagating waves and can be written as:

$$\vec{E} = \vec{E}_0 \exp(-\gamma z) = \vec{E}_0 \exp(-jkz) = \vec{E}_0 \exp(-jk_0 n z) \quad (1-15)$$

- Quadrant 4: mu-negative medium or magnetic plasma $\varepsilon > 0, \mu < 0$.

In certain frequency regimes some gyrotropic materials exhibit this characteristic. This quadrant supports also evanescent waves.

$$\varepsilon > 0 \Rightarrow \varepsilon = |\varepsilon| \text{ and } \mu < 0 \Rightarrow \mu = -|\mu|$$

From equation (1-1)

$$n = \sqrt{-|\varepsilon_r| |\mu_r|} = j \sqrt{|\varepsilon_r| |\mu_r|} \quad (1-16)$$

$$k = j \omega \sqrt{|\varepsilon_r| |\mu_r|} \quad (1-17)$$

From equation (1-9) and (1-17), we find

$$\beta = 0 \text{ and } \alpha = -\omega \sqrt{|\varepsilon_r| |\mu_r|} = \gamma \quad (1-18)$$

From equation (1-6), the expression of the propagating plane wave in such medium is :

$$\vec{E} = \vec{E}_0 \exp(-\alpha z) \quad (1-19)$$

Metamaterials have much broader scope than LHM, as shown in figure 1-1. In the $\varepsilon - \mu$ domain, there are several special lines and points indicating special material properties. For example, the point $\mu = -\mu_0$ and $\varepsilon = -\varepsilon_0$ represents an anti-air in the LHM region, which will produce a perfect lens; the point $\mu = 0$ and $\varepsilon = 0$ represents a nihility, which can yield a perfect tunneling effect; the line $\mu = -\varepsilon$ in both RHM and LHM regions represents impedance-matching materials, which have perfect impedance matching with air, resulting no

reflections. Also, the vicinity of $\mu = 0$ is called as μ near zero (MNZ) material, and the vicinity of $\varepsilon = 0$ is called as ε -near zero (ENZ) material, which has special properties applied in in perfect lens [56].

1.5. Artificial dielectrics

Artificial dielectrics, the first known metamaterials, usually consist of artificially created molecules with dielectric or metallic inclusions of a certain shape [33]. These molecules can be distributed and oriented in space, as seen in figure 1-2, either in a regular lattice or in a random manner.

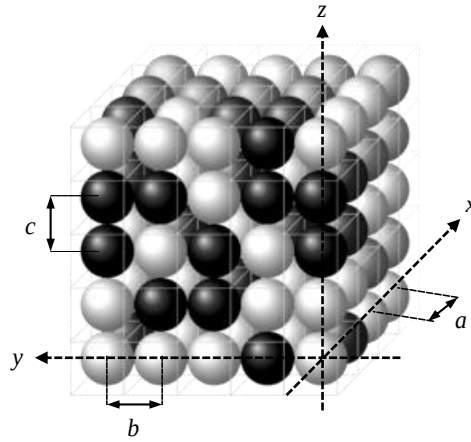


Figure 1-2 : The geometry of a generic artificial dielectric [33]

The dimensions of the molecules and characteristic distances between neighboring ones are assumed to be very small, as compared to the wavelength.

The concept of artificial dielectrics was perhaps first introduced in [34], who used it to design low-weight dielectric lenses at microwave frequencies. Artificial dielectrics also find applications where we need to use a high-permittivity material, such with the titanate which has $\varepsilon_r = 90$. Since these materials are usually expensive, an alternative is to use metallic inclusions of various shapes with sizes that are small compared to the wavelength to artificially produce a high-permittivity material, which is both low-cost and lightweight [35]. A very interesting example of artificial dielectrics is the wire medium [36] [37], which has been known since 1950s. It is formed by a regular lattice of conducting wires.

1.6. The microscopic view of metal thin wire

Based on the effective medium theory, the metamaterials can be characterized by effective electric permittivity and magnetic permeability. Pendry realized the artificially

electric plasma using a metallic wire whose permittivity is negative [38] And the artificially magnetic plasma whose permeability is negative [39]. Metamaterials open a door to realize all possible material properties (tailoring desired electromagnetic properties) by designing different unit cells and using different substrate materials [40]. Among the various, usual and unusual material parameters provided by metamaterials ($\epsilon_{eff}>0$, $\mu_{eff}>0$, $\epsilon_{eff}<0$, $\mu_{eff}<0$, $\epsilon_{eff}\approx 0$, $\mu_{eff}\approx 0$, $\eta\approx 0$...) they can lead to many interesting phenomena.

The metal thin wire medium, whose unit cell is presented in figure 1-3-(b) exhibits a plasmonic type permittivity frequency function of the form [41]:

$$\mathcal{E}_{ref}(\omega) = 1 - \frac{\omega_{pe}^2}{\omega^2 - j\Gamma_e \omega} \quad (1-27)$$

Equation (1-27) in real and imaginary parts becomes:

$$\epsilon_{ref}(\omega) = 1 - \frac{\omega_{pe}^2}{\omega^2 + \Gamma_e^2} + j \frac{\Gamma_e \omega_{pe}^2}{\omega(\omega^2 + \Gamma_e^2)} \quad (1-28)$$

The real and imaginary parts are plotted in figure 1-3 (c).

Analytic procedure is detailed in [42] from which the Drude parameters ω_{pe} and Γ_e are expressed in function of the geometric dimensions (the period a and the radius of the wire r) as:

$$\omega_{pe} = \sqrt{\frac{2\pi c^2}{a^2 \ln(a/r)}} \quad (1-29)$$

$$\Gamma_e = \epsilon_0 \left(\frac{a\omega_{pe}}{\pi\sigma r^2} \right) \quad (1-30)$$

Where c is the speed of light and σ is the conductivity of metals.

We observe from the equation (1-28) that $\text{Re}(\epsilon_{ref}) < 0$ for $\omega < \sqrt{\omega_{pe}^2 - \Gamma_e^2}$. It means that if there are no collisions, the free electrons movements are in phase of 180° with respect to the applied electric field. In this frequency band, the propagation is not possible and the wave is evanescent and it's completely reflected at the medium interface.

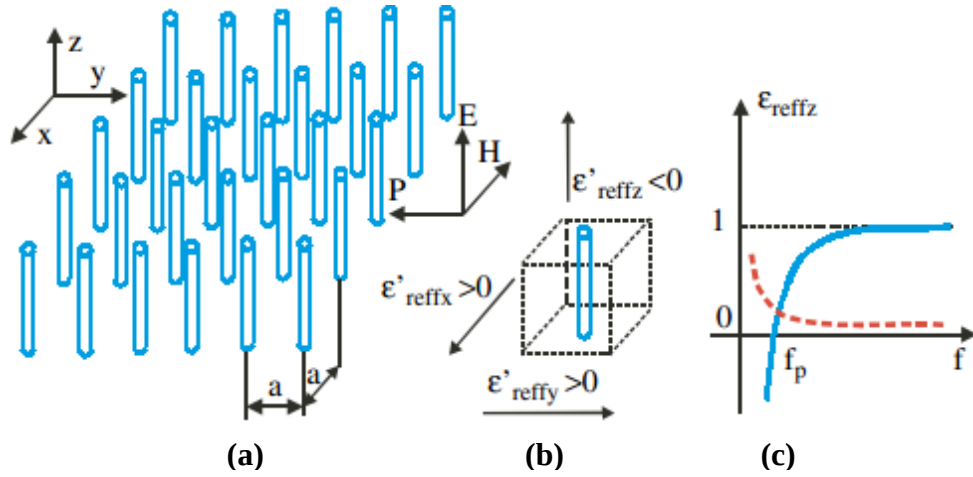


Figure 1-3 : 2D thin wires arrangement, (b) Thin wire unit cell, and. (c) Real (solid blue) and imaginary (dashed red) parts of effective permittivity of metamaterial (a) [43]

1.7. Artificial magnetic materials

Artificial magnetic materials are typically synthesized by using resonant elements, such as split-ring resonators or Swiss rolls as shown in figure 1-4. The split-ring resonators are more widely used than the Swiss rolls since they can be manufactured using printed circuit technology. Artificial magnetic materials were known as far back as the early 1950s, and a book by Schelkunoff and Friis [44] provides expressions that can be used to calculate the magnetic flux of a split metallic wire loaded with a small capacitor, formed by a pair of parallel metallic plates. Schneider et al in [45], used a split metallic tube to produce a nuclear magnetic resonance NMR probe, and, in [46] Hardy proposed a Swiss roll resonator to achieve magnetic resonance at 200-2000 MHz. Kostin et al in [47], has synthesized an artificial magnetic material with frequency dependent positive permeability by using double circular ring resonators.

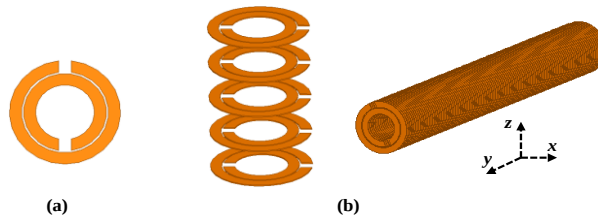


Figure 1-4 : (a) - Split-ring resonator SRR (b) - Swiss roll resonator

An artificial magnetic material, which is formed by split-ring resonators, possesses negative permeability within a frequency band (bandwidth is typically narrow) near the resonant frequency of the single split-ring resonator; it is widely used to create LHMs. Metallic waveguides filled with such artificial magnetic materials support guided waves at

frequencies below the cutoff frequencies of corresponding hollow waveguides hence such guides are also referred to as the sub-wavelength waveguides [48]. This effect can be used for miniaturization of guided wave structures [49], though they do introduce additional losses owing to the presence of high circulating currents in the inclusions when they are resonant.

1.8. Microscopic view of split ring resonator (SRR)

The SRR in figure 1-5-(b) exhibits a permeability frequency response of the form [41]:

$$\mu_{\text{reff}}(\omega) = 1 - \frac{\omega_{pm}^2}{\omega^2 - \omega_{0m}^2 + j\Gamma_m \omega} \quad (1-31)$$

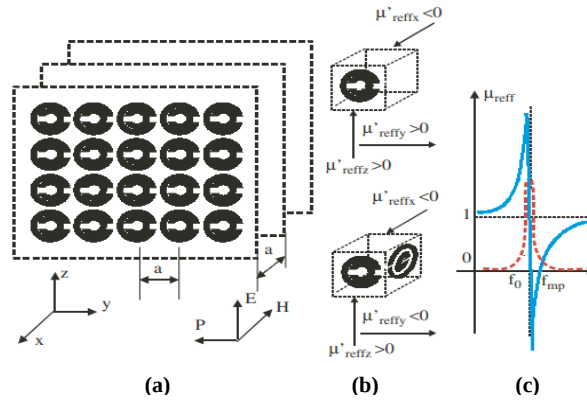


Figure 1-5 : (a).Array of SRR.(b).SRR of 1D and 2D unit cells (c). $\text{Re}(\mu_{\text{reff}})$ (solid blue), $\text{Im}(\mu_{\text{reff}})$ (dashed red) [31][43]

Equation (1-31) is divided in real and imaginary parts and presented in figure 1-5-(c):

$$\mu_{\text{reff}}(\omega) = 1 - \frac{F(\omega^2 - \omega_{0m}^2)}{(\omega^2 - \omega_{0m}^2) + (\omega\Gamma_m)^2} + j \frac{\Gamma_m \omega_{pm}^2}{(\omega^2 - \omega_{0m}^2) + (\omega\Gamma_m)^2} \quad (1-32)$$

Lorentz parameters are calculated in detail manner in [38] :

$$F = \sqrt{\pi(r/a)^2} \quad (1-33)$$

$$\omega_{0m} = c \sqrt{\frac{3a}{\pi \ln 2(2wr^3/\delta)}} \quad (1-34)$$

Where r , w and δ are the inner radius for the smaller ring, width of the ring and radial spacing between the rings.

We have also:

$$\Gamma_m = \frac{2aR'}{r\mu_0} \quad (1-35)$$

R' is the metal resistance per unit length.

We observe from (1-32) that $\mu_{\text{reff}} < 0$ for $\omega_{0m} < \omega < \frac{\omega_{0m}}{\sqrt{1-F}} = \omega_{pm}$

1.9. Metamaterial applications

Due to the exciting and unusual features, metamaterials have found and are finding a lot of applications. Metamaterials can impact on telecommunications, sensing and susceptibility, across all wave types, but especially electromagnetic and acoustic as shown in figure 1-6.

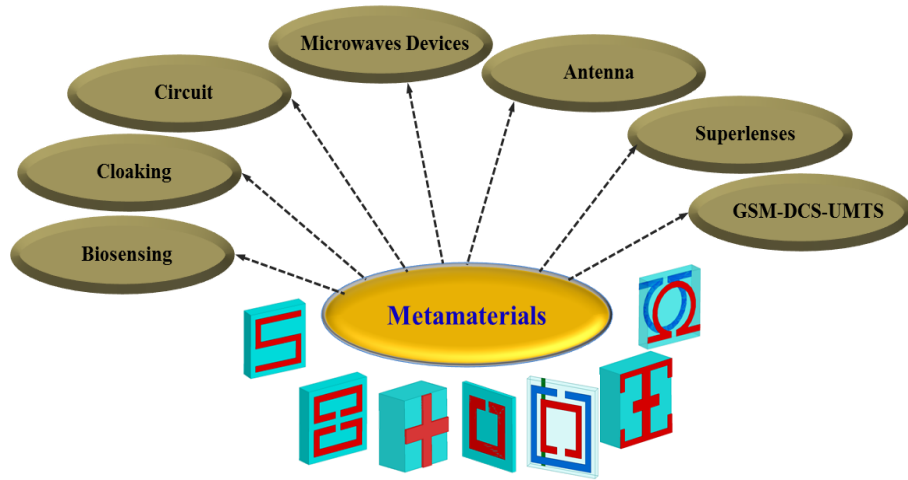


Figure 1-6 : Potential applications

1.9.1. Superlens

For left handed metamaterial, the most attractive feature is the superlens which can be widely used in super resolution medical imaging, optical imaging, and nondestructive detections [50]. A super lens or perfect lens is a lens which uses metamaterials to go beyond the diffraction limit [51]. This is an inherent limitation in conventional optical devices or lenses. In 2000, a type of lens was proposed, consisting of a metamaterial that compensates for wave decay and reconstructs images in the near field and most importantly the both propagating and evanescent waves contribute to the resolution of the image figure 1-7. In 2004, the first superlens with a negative refractive index provided resolution three times better than the diffraction limits and was demonstrated at microwave frequencies. In 2005, the first near field superlens which exceeded the diffraction limit was demonstrated [52]. The higher focusing resolution will be provided by flat LHM lens if compared to convex dielectric lens

and elliptical reflector focusing system. The LHM lens has the potential to acquire higher imaging resolution and easy in-depth scanning, which will simplify the detection system design [52].

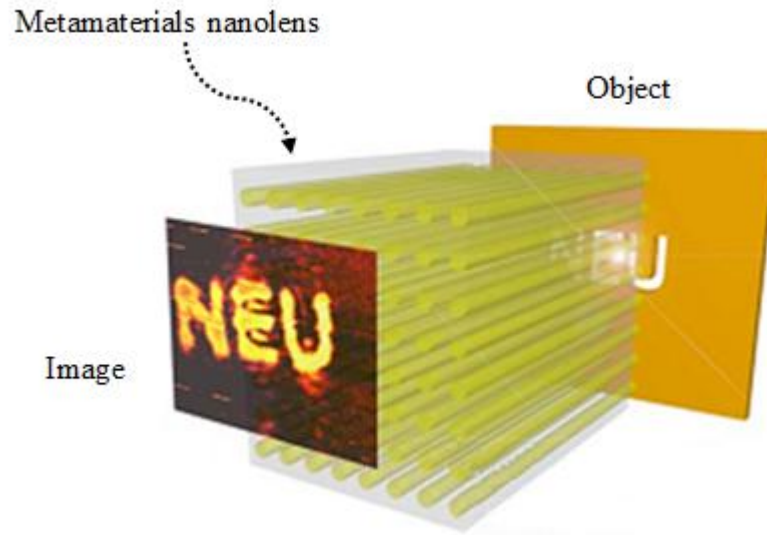


Figure 1-7 : Super-resolution imaging using a three-dimensional metamaterials nanolens

For example in biological domain, microwaves are used to destroy or ablate diseased soft-tissue by heating it to a temperature that causes cell death. This procedure is called hyperthermia. The generator produces microwave energy which is transmitted through the antennas into the patient. Elevating the temperature of tumor cells causes cell membrane damage, which leads to the destruction of the cancer cells. Hyperthermia treatment of cancer requires directing a carefully controlled dose of heat to the cancerous tumor and surrounding body tissue. The most prominent property of metamaterial lens is the ability of negative-refractive index (NRI) to focus the electromagnetic field of a source. Hence it can generate appropriate focusing spot in biological tissue as required in microwave hyperthermia treatment.

1.9.2. Antennas

Metamaterial antennas are a class of antennas which use metamaterials to enhance or increase the antenna performances [53], as shown in figure 1-8. The metamaterial could enhance the radiated power of an antenna. Materials with negative magnetic permeability could possibly allow for properties such as an electrically small antenna size, high directivity, and tunable operational frequency [54]. Furthermore, metamaterial based antennas can demonstrate improved efficiency and high bandwidth performance [55].

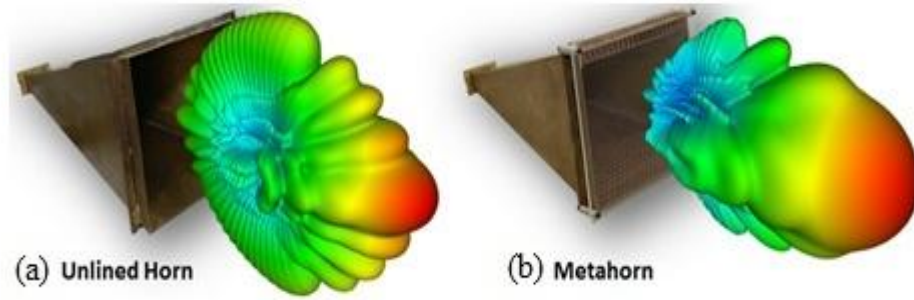


Figure 1-8 : Rectangular horn antenna (a)-alone (b)-with metamaterial

Metamaterials employed in the ground plane surrounding antennas offers improved isolation between radio frequency or microwave channels of (multiple-input multiple-output) (MIMO) antenna arrays. Metamaterial, high-impedance ground planes can also be used to improve the radiation efficiency, and axial radio performance of low-profile antennas located close to the ground plane surface.

Metamaterials have also been used to increase the beam scanning range by using both the forward and backward waves in leaky wave antennas. Various metamaterial antenna systems can be employed to support surveillance sensors, communication links, navigation systems, command and control systems [56].

The gradient refractive index metamaterials are also utilized to produce beam-bending lens and beam-focusing lenses. Based on such properties, high-gain and broadband gradient planar lens antennas and Luneberg-like lens antennas have been proposed and realized [56]. The Industrial, Scientific and Medical (ISM) radio bands are portions of the radio spectrum reserved internationally for the use of radio frequency (RF) energy for industrial, scientific and medical purposes other than communications.

1.9.3. Microwave devices

The waveguided or planar metamaterials which are composed of complimentary structures like CSRR have been used to design microwave devices like filters, power dividers, and phase shifters. Narrowband and broadband polarizers have been realized using three-dimensional (3D) anisotropic metamaterials.

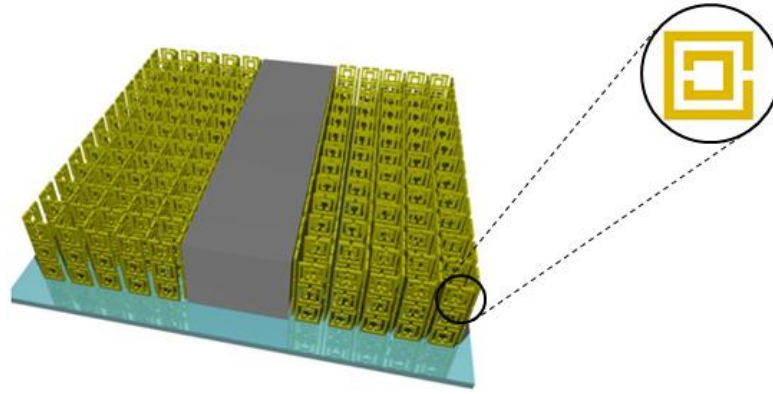


Figure 1-9 : Planar waveguides with double SRR metamaterials

Superconducting metamaterials have been achieved in addition to superconducting split rings and wires. They have been utilized to realize a number of unique applications. The combination of left-handed and right-handed propagation media creates opportunities for new types of resonant structures. By laminating two materials with opposite sense of phase it is possible to create a new class of resonant structures [57]. For example, two flat conducting plates separated by a sandwich of left-handed and right handed metamaterials constitutes such a new resonator. A wave propagating in the direction normal to the plates will suffer a combination of forward and reverse phase windings before reaching the other reflecting boundary. Under these conditions the wave could undergo a net phase shift of 0 radians and still create a resonance condition. The net phase shift could also be a positive or negative multiple of 2π radians as well, each creating a resonant condition. The result is an ultra-compact resonator whose overall dimension is no longer constrained by the wavelength of the resonant wave.

Superconductors are particularly attractive for use in ultra-compact resonators and showed resonances with indices between +2 and -6, including zero, over the broad frequency range from about 5 to 24 GHz [58]. Quality factors of the negative order resonances were on the order of 3000 at 30 K, while those for positive order resonances were below 400.

1.9.4. Circuit applications

SRR and CSRR-based transmission lines are useful for device miniaturization, and find their applications in microwave passive components such as impedance inverters, power dividers [59], couplers, and filters. SRRs are also useful particles in many other applications such as magneto inductive and electro inductive wave components [60] frequency selective surfaces [61].

1.9.5. Cloaking

Cloaking devices are advanced stealth technologies that can make objects partially or wholly invisible to some portions of the electromagnetic spectrum. Generally speaking, there are several major approaches to render objects invisible. For example, Alu and Engheta [62] proposed to use plasmonic coatings to cancel the dipolar scattering. But this technique is limited to the sub-wavelength scale of the object, and the coating depends on the geometry and material parameters of the object. The cloak deflects microwave rays, so they circumvent the object inside with little distortion figure 1-10 making it appears almost as if nothing were there at all. Milton and Nicorovici [63] discovered that using a metamaterial coating would cloak polarizable line dipoles. But the coating is affected by the objects placed inside. Leonhardt [64] and independently Pendry, Schurig and Smith [65] discovered a coordinate transformation mechanism for electromagnetic cloaking.

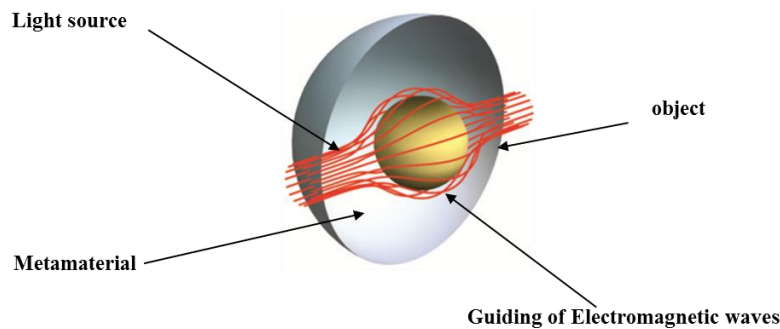


Figure 1-10 : Invisibility cloaks

Their mechanism was quite similar to that of Greenleaf et al. [66][67] introduced for conductivity. Their main idea is to guide electromagnetic wave around the cloaked region, and much later work has adopted this technique. In May 2006, the first full wave numerical simulations on cylindrical cloaking was carried out by Cummer et al. [69]. A few months later, the first experiment of such a cloak at microwave frequencies was successfully demonstrated by Schurig et al. [70], where the cloak surrounding a 25-mm-radius Cu cylinder was measured. After 2006, numerous studies have been devoted to cloaking, mainly inspired by [69]. For example, in 2008, Liang et al [71] performed a time-dependent simulation for the cylindrical cloak using finite-difference time-domain method.

1.9.6. Biosensing

Another potential application field of metamaterials is on biosensing. Conventional biosensors (such as those based on electro-mechanical transduction, fluorescence,

nanomaterials, and surface plasmon resonance) often involve labor-intensive sample preparation and very sophisticated equipment [68].

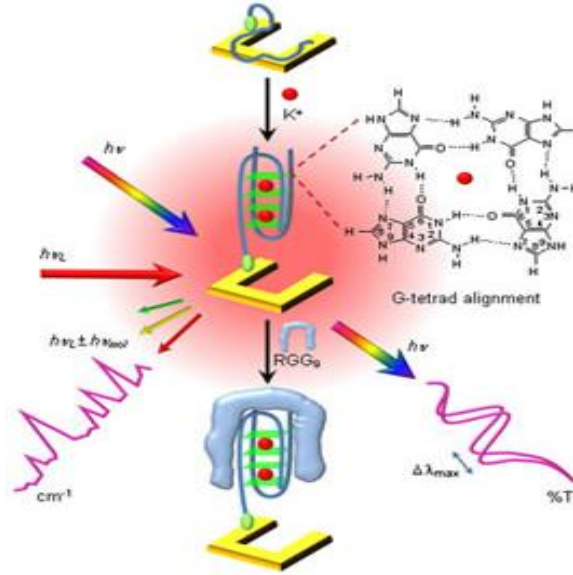


Figure 1-11 : Schematic illustration of Vis-NIR SRR metamaterials as dual transducing mode nanosensor for the detection and identification of biomolecules

In recent years, researchers have proposed to use metamaterials as candidates for detection of highly sensitive chemical, biochemical and biological analytes. For example, Lee et al. [72] studied the possibility of using split-ring resonators (SRRs) for biosensors. The basic principle is based on the fact that SRR can be considered to be a simple LC circuit with a response frequency of $f = 1/2\pi\sqrt{LC}$, which shows that the resonant frequency varies in terms of the changes in the inductance L and/or capacitance C . Hence the resonant frequency of the SRR shall be shifted before and after the introduction of biomaterials. Planar metamaterials were proposed to serve as thin-film sensors recently by O'Hara et al [73]. They found that a resonant frequency response can be tuned through metamaterial designs. Though their metamaterial design can only detect thin films having a thickness less than 100 nm, their work presents a promising outlook for THz sensing technology.

1.10. Frequency selective surface (FSS)

Frequency Selective Surfaces are planar periodic structures of identical patches or apertures of conducting elements duplicate periodically in either a one or two-dimensional array on a dielectric substrate as shown in figure 1-12. Because of their frequency selective properties, FSS are incorporated in a wide variety of applications such as the realization of reflector antennas and antenna/radome design as shown in chapter 4, making polarizers and

beam splitters, and also as radar absorbing structure [74]. The frequency behavior of the FSS depends on the shape of the elements (apertures/patches), their size, spacing, and thickness of the metal screen. Generally FSS are employed in front of a grounded dielectric slab (substrate) to synthesize high-impedance absorbing surfaces.

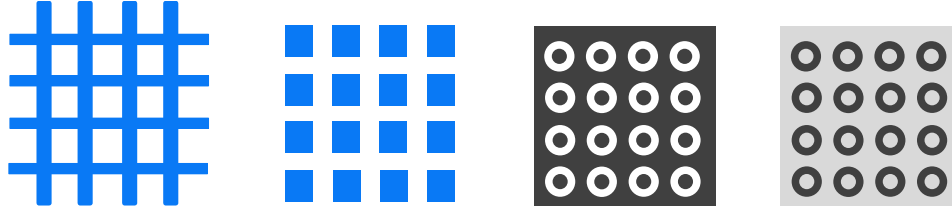


Figure 1-12 : Plane FSS with different unit cells [75][76]

1.10.1. FSS element comparison

There are four different types of possible element-type FSS arrays: center connected, loop, solid interior and combination. See figure 1-13 for some examples of each element type [74].

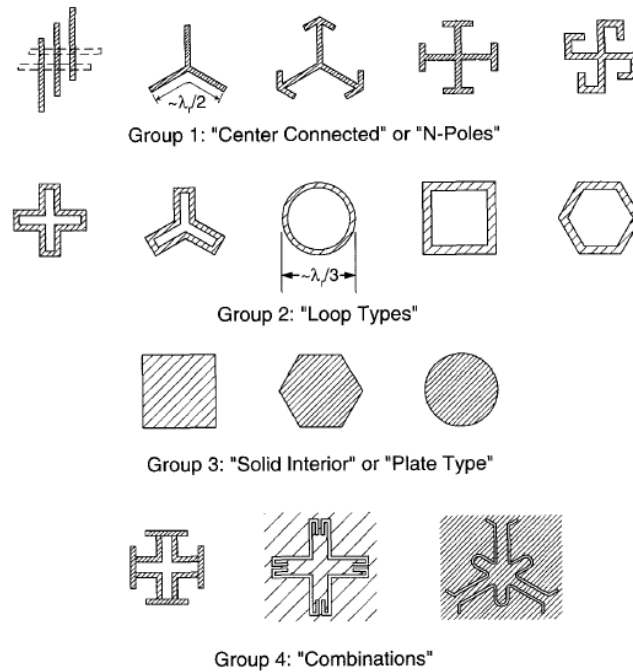


Figure 1-13 : The four major groups of FSS elements [74]

The first group, the center connected elements, are a viable candidate for both radiating and non-radiating arrays. Perhaps the most recognizable from the center connected group is the dipole [74]. The loop elements are in general an excellent choice for non-radiating element FSS arrays. These elements are typically smaller in the x and y directions than the center connected elements with respect to a wavelength and thus can be spaced close to one

another. Plate elements in general do not have very desirable characteristics. First the elements have x and y dimensions around a half of a wavelength, which limits how closely these elements can be placed next to one another. In addition plates are highly inductive elements with small capacitances between them, leading to issues with regard to achieving resonance. If the FSS never resonates, then it is impossible to achieve perfect reflection. This is due to the fact that at resonance, the FSS becomes a short circuit (assuming the materials are lossless) and thus acts as a perfect electric conductor (PEC) ground plane [74]. Finally combination elements are too diverse to generalize in any meaningful way. The possibilities in this group are as boundless as the imagination of the designer. Some possible examples of elements in this group have already been shown in figure 1-13 [74].

The FSS arrays effectively create low pass, high pass, band stop, and band pass filters respectively (see figure 1-14) to electromagnetic energy at their design frequency range. These types of filters are used in a wide variety of applications, including radomes, dichroic surfaces, and circuit analog absorbers [77].

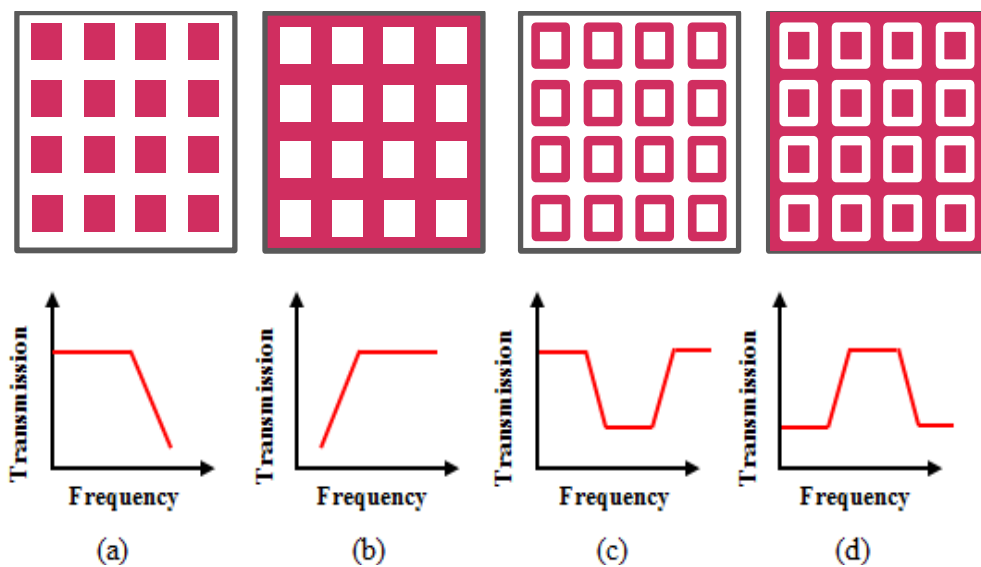


Figure 1-14 : Filter types based on FSS (a)-low pass filter (b)-high pass filter(c)-band stop filter (d)-band pass filter [78]

The FSS found their application in:

1- Selective shielding of the electromagnetic interference from high power microwave heating machines adjacent to wireless communication base stations,

2- Selective shielding of frequencies of communication in sensitive areas (military installations, airport, police.),

3- Protection from harmful electromagnetic radiation especially in the [2 , 3] GHz band in domestic environment, schools, hospitals etc. arising externally (wireless communication base stations) or internally (microwave ovens),

4- Control of radiation at unlicensed frequency bands (Bluetooth applications, 2.45 GHz),

5- Pico-cellular wireless communications in office environments such as the Personal Handy phone System in offices where to improve efficiency each room needs to prevent leakage of radio waves into and of the room. This implies that windows, floor and ceiling need to be shielded. Kajima Research Institute is developing such a film to be applied to buildings,

6- Isolation of unwanted radiation. FSS windows can be incorporate in trains prevent mobile phone frequencies; around base station antennas to filter out unwanted radiation.

1.11. Metaradomes

Metamaterial plays a significant role in improving the electromagnetic properties of the radome, so metamaterial is employed for radome material, it called metaradome.

Metaradome designs should be developed and adjusted to different application areas such as maritime, telecom, radar, aerospace, and automotive applications [79]. Metaradomes may help in shaping the directivity of antennas. Directional antennas are used in wireless networks [80], and the use of metamaterials in the antenna design is described in [80][81]. Another application is cognitive radio. Such a radio automatically detects available channels in a frequency spectrum and accordingly changes its transmission or reception parameters to have an optimum communication bandwidth within a defined spectral region. Such a software-defined radio platform should convert into a completely reconfigurable wireless system that automatically alters its communication variables according to network and demand of user [82]. For this propose, there are studies to design suitable reconfigurable antennas [83][84].

In[85] the gain of the metamaterial radome is showing improvement of 124% compared to the patch antenna alone. Owing to high gain behaviour the proposed metamaterial radome is applicable in several radar systems. The potential of metamaterials for improving the performance of microwave radomes and several examples of the use of metamaterials in the design of radomes have been given in [86].

The application of metamaterials to radomes is a new research field which is open for further analysis and development. Potential applications are diverse and promising.

1.12. Conclusion

The metamaterials are artificial materials made of a periodic arrangement of unit cells with some unique properties. Such properties include a simultaneous negative or positive permittivity and/or permeability. Its properties can be tailored in a broader manner than with classical materials. These characteristics allow metamaterials to be useful.

Metamaterials can be used for several interesting devices in microwave and terahertz engineering such as antennas, microwave devices, superlenses, absorbers, cloaks, filters and acoustics, sensors, superlens, and biomedical applications.

2.1. Introduction

Frequency Selective Surfaces (FSSs) are periodic structures in two dimensions that can provide frequency filtering to incoming electromagnetic waves [87], also they have been the subject of intensive investigation for their widespread applications as spatial microwave and optical filters for more than four decades [88][89]. Frequency selective surfaces are usually constructed from periodically arranged metallic patches of arbitrary geometries or their complimentary geometry having aperture similar to patches within a metallic screen. These surfaces exhibit total reflection or transmission, for the patches and apertures, respectively, in the neighborhood of the resonances element. The most important step in the design process of a desired FSS is the proper choice of constituting elements for the array. The element type and geometry, the substrate parameters, the presence or absence of superstrates, and inter-element spacing generally determine the overall frequency response of the structure, such as its bandwidth, transfer function, and its dependence on the incidence angle and polarization.

Although conceptually similar to a classical microwave filter, it should be emphasized that a space filter is inherently much more complicated.

First of all, a classical filter merely has a pair of input and output terminals and only the frequency is varied at the input while the response at the output is recorded. A space filter, on the other hand, has the incident field arriving at various angles of incidence as well as polarizations. This fact has a profound effect on the transmission properties; in short, the transmission curve will change dramatically unless carefully designed. The shapes and configurations that can be chosen for the FSS elements are limited only by the imagination of the designer.

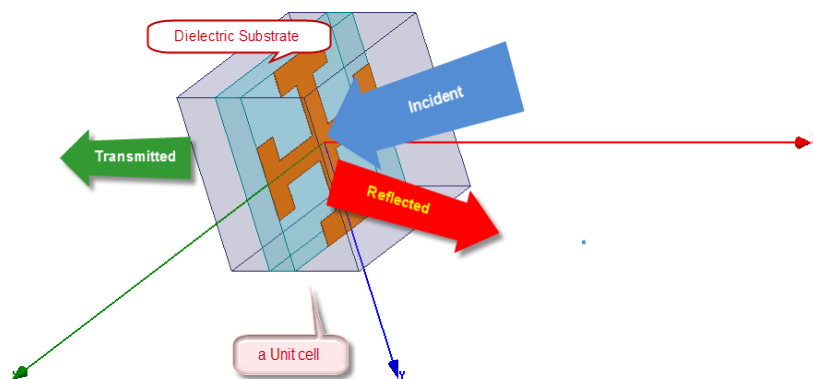


Figure 2-1: Perspective of a unit cell (Jerusalem)

The unique properties and practical uses of frequency selective surfaces realized over many years have produced an extensive body of work within both academic and industrial

sectors. Historically, the essential behavior of these surfaces stems from meshes and strip gratings concepts exploited in the optical regime. At microwave frequencies, the applications of FSSs are predominantly for antenna systems in fixed as well as mobile services. Published reports of basic properties of simple structures in the cm-wave region go back as far as 1946[90][91], although the name FSS was not used until much later on. Planar and curved FSSs are used for a variety of applications including design of antenna radomes, as mentioned in the final chapter, dichroic surfaces for reflectors and subreflectors of large aperture antennas, or even absorbers [75]. More sophisticated FSSs with multiple resonances [92] have been designed and reported in the literature with the advent of computational electromagnetics codes that began in the 1980's [93]. More recently, in [94] advanced methods based on method of moments (MoM), finite element method (FEM), and finite-difference time-domain (FDTD) with periodic boundary conditions have made the design process of FSSs substantially easier. In [95], self-similarity characteristics of fractal geometries are exploited for the design of a dual-band FSS. In [96], design of a reconfigurable frequency selective surface with traditional elements integrated with RF MEMS switches operating at Ka-band is reported. Also the application of ferroelectric substrates for tuning the surface frequency behavior is considered in [97]. The use of superconductors in the design of very low loss FSSs is comprehensively studied in [98], also FSS are used like circuit analog absorbers, filters, polarizers and some recent applications such as radio frequency identification RFID tags [75] [99], robotic guided paths, photonic bandgap structures [100], collision avoidance, electromagnetic interference EMI protection [101] and secure building applications [102] . An interesting use of FSS on fabrics for wearable applications is mentioned in [103]. The first conceptual work about "periodic surfaces" was patented by Marconi and Franklin in 1919 [104]. FSS studies starting with Marconi and Franklin's invention have been continued by many researchers up to now. There has been a great interest in these structures especially since 1960's. However, the name "Frequency Selective Surface" was patented for the first time by Yee in 1993 [105]. The definitions and discussions about FSS phenomena are clearly declared by Munk, guru of FSS, in [75]. The first concept which should be understood is "Periodic Surface" since it constitutes the nature of FSS structures. A periodic surface consists of identical elements of infinite arrays in one or two dimensions.

The factors affecting the performance of FSS structures can be listed as:

- Type of elements (patches vs. slots),

- Shape of elements,
- Dimensions of elements,
- Inter element spacing,
- Array configuration (i.e. rectangular, triangular, hexagonal etc.),
- Alignment of elements in multi-layer structures.

Moreover, the electrical properties of the substrate material also determine the performance of FSS. Therefore, there are numerous combinations of various parameters in order to reach an optimum solution for a specific problem.

In modern military platforms such as ships, aircrafts and missiles, frequency selective surfaces are widely used for antennas and radar cross section (RCS) reduction. The RCSs of complicated objects such as antennas are difficult or impossible to control over a wide frequency range [106]. The most efficient and cost-effective approach in these situations is to shield the scattering object from the threat radar. This can be accomplished by retracting the object and covering the cavity in which it is housed with a shutter. If the object is an antenna, then, obviously, the system served by this antenna cannot operate when it is stowed. An alternate approach is to cover the antenna with an FSS that is transparent at the antenna operating frequency, yet opaque at the threat radar frequency. Considering the shape and structural constraints of military platforms, the antenna and the FSS radome should be carefully designed as an integrated module such that the module should have the radiation performance of the antenna, and the filtering performance for anti-interference and low RCS (see chapter 4). Otherwise, a bad design such as setting up the antenna behind an FSS radome blindly may lead to an increase in the RCS, a degradation of the radiation performance of the antenna as well as its impedance matching properties, strong undesired electromagnetic interference (that may destroy the whole electronics of the system), etc.

In literature very few papers can be found about an integrated design of antenna and FSS radome (see chapter 4). Structural integration of antenna and an FSS structure was firstly reported in [107]. But in [107] FSS was used only to enhance the antenna radiation efficiency due to its high impedance characteristic. Multilayer FSS combined with open-end waveguide radiator arrays have been presented in [103][104], in which only filtering performance of the integrated module has been investigated using multi mode equivalent network method.

In this section, a detailed investigation of periodic surfaces including their resonance mechanisms, analysis method and some unit element types are provided (with HFSS).

2.2. Type of elements: patches and slots (apertures)

FSS structures are constructed with any of two types of elements: metallic patches on dielectric slabs or slots in metals. The metallic patches exhibit a capacitive behavior and low pass filtering characteristics (see figure 2-2.a). If the patches and apertures are replaced as shown in figure 2-2.b, the structure can be made inductive, which results in high-pass filtering properties [108]. In order to govern desired filtering characteristics, capacitive and inductive models can be arranged together.

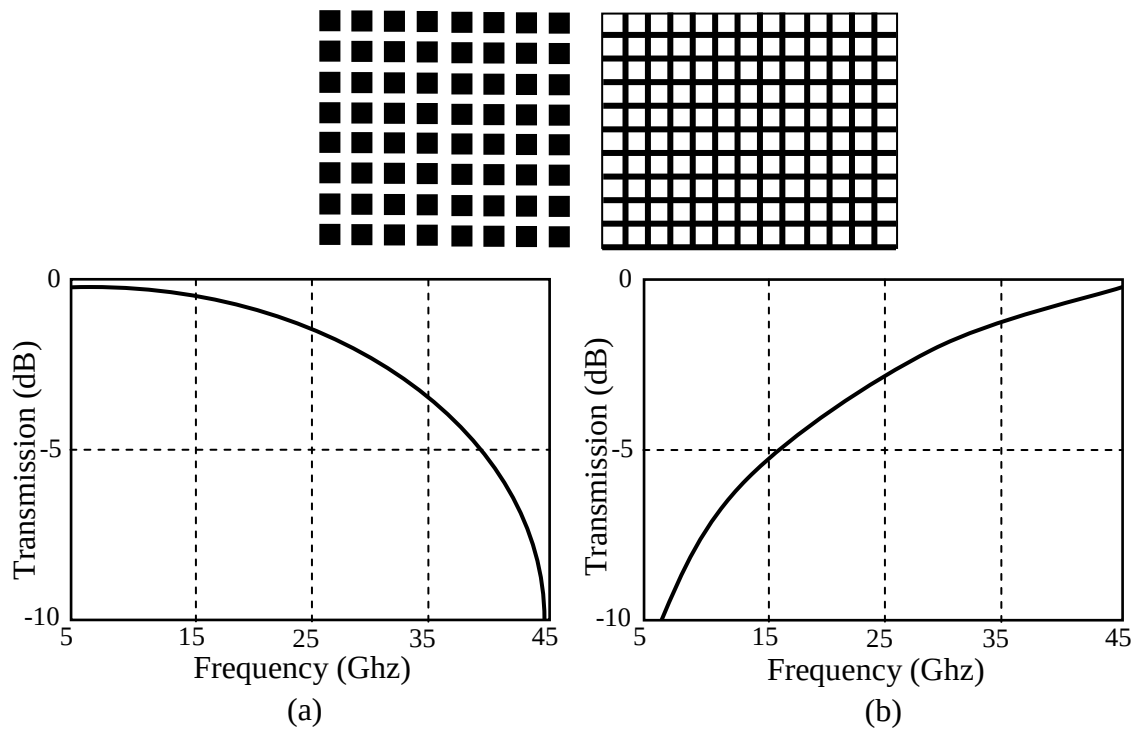


Figure 2-2: Transmission characteristics and models of FSS with respect to electrical properties:

(a) Capacitive (b) Inductive [108]

By modification and utilization of different element types together, different filtering characteristics like low-pass, high-pass, band-pass and band-stop, can be obtained. As mentioned above, if the elements are pure capacitive, corresponding is a low-pass filter. Reversely, if the elements are pure inductive, a high-pass characteristics comes up. The size of patches in figure 2-2.a is greater than the size of the apertures, which results a high capacitance and a negligible inductance with respect to the capacitance. In figure 2-2.b, the same situation is present for the apertures against small patches which creates an almost pure inductive medium. These characteristics are modified by making the inter-element apertures between the patches comparable and vice versa. In this way, FSS elements can be arranged in order to include both capacitive and inductive properties together. If patches are modified in

this manner a band-stop filtering effect is observed. Similarly, due to the duality principle, apertures can be utilized in order to achieve band-pass performance.

2.3. Shape of elements

Munk groups FSS elements according to their shapes in four main categories [101] (see figure 2-3), which are namely;

1. The center connected or N-poles
2. The loop types
3. Solid interior or plate types
4. Combinations

The center connected or N-poles is the basic form for many FSS structures such as single dipole, three-legged, anchor, the Jerusalem cross and square spiral elements. Simple dipoles are arrayed in two dimensionally periodic sense which demonstrate a band-stop characteristics when illuminated.

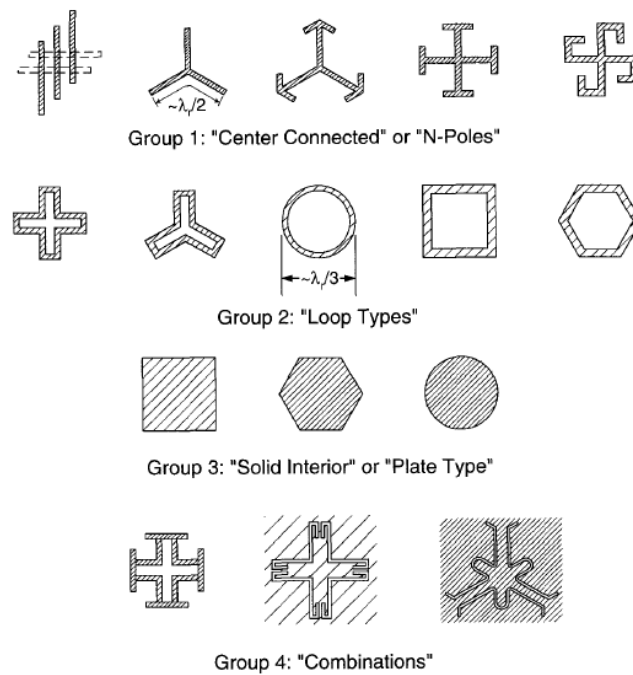


Figure 2-3 : Classification of FSS elements [75]

A patent for FSS was granted in 1974 to Munk for his work about the stabilization of FSS performance against oblique incidence [109]. Munk determined that the spacings between the unit cell elements must be smaller than 0.4λ so as to preserve the FSS characteristics for different scan angles. Furthermore, in order to provide small inter-element spacing, he

claimed that the elements should be miniaturized. The most commonly used method to miniaturize a resonance structure is to load it either capacitively or inductively.

As shown in figure 2-4, the dipole elements (Figure 2-4.a) are loaded by two shorted wires segment (figure 2-4.b) which supplies the required inductance to obtain the resonance frequency of half wavelength dipole elements in smaller sizes. The characteristics of the unloaded dipole elements are very sensitive to the angle of impinging waves as seen in figure 2-5. When the shorted loading segment is inserted between smaller dipole elements, the stabilization is observed to increase significantly.

According to Munk [101], the reason is the impedance of the shorted dipole elements which is much higher compared to the mutual impedance. Thus, the effect of the mutual impedance, which normally changes with the scan angle and modifies the behavior, is diminished. The loaded model has narrow-band characteristics since the impedance of the shorted dipoles varies dramatically with frequency. Therefore, Munk suggests that kind of FSS elements for applications where wide scanning and minimum bore-sight error are desired [101].

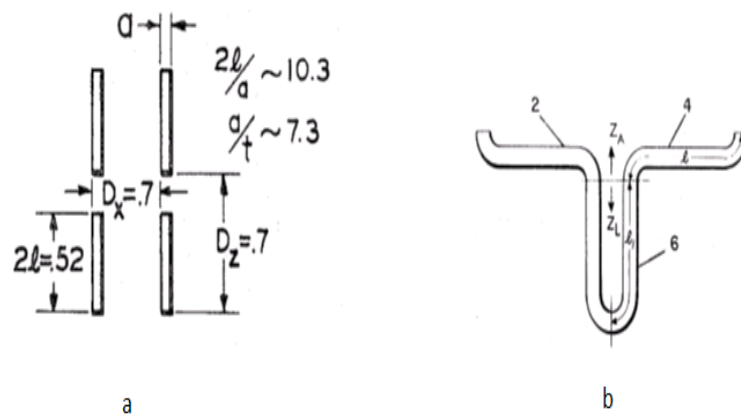


Figure 2-4 : Dipole elements a) Regular half wavelength dipoles b) Loaded dipoles[109]

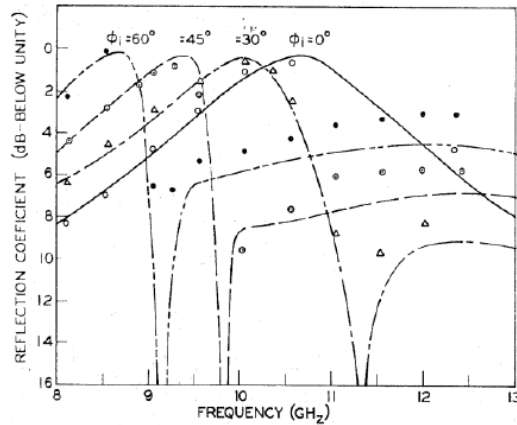


Figure 2-5 : Reflection curves of standard dipoles for different angles of incidence [109]

The band reject performance of the single dipoles is poor since secondary resonance occurs in oblique incidence cases for parallel polarization. This situation may cause transmission of undesired signals of stop-frequency band for parallel polarization cases. The unloaded tripole element improves the bandwidth performance with respect to single dipoles. The smaller inter-element spacing is another advantage of the unloaded tripoles. However, the second harmonic excited by parallel polarization still occurs for these elements. The anchor elements are constructed by tuning the tripole element with an end capacitance. The modification results in a broader bandwidth and smaller spacing. The Jerusalem cross consists of crossed dipoles loaded at the ends similar to anchor elements. The capacitive loading decreases the fundamental resonance frequency. The second harmonic null shifts upward in frequency. For Jerusalem cross, another null also takes place in upper band, which is defined as "free space grating lobe" by Munk [101]. The last member of "center connected" elements is the square spiral. The interesting feature of the square spiral element is its dual filtering characteristics. In other words, the element can be used as either band-pass or band-stop filter due to the higher null position in frequency.

The second group of elements called "loop types" consists of three or four-legged, circular, square and hexagonal loops. The invention of the four legged loaded element by Munk in [109], aims to decrease the inter-element spacing in order to strengthen the stability against oblique incidence and to increase the bandwidth of operation. Since the model is symmetric in two orthogonal axes, both vertical and horizontal polarizations are filtered equally. As a natural consequence, the same equivalent circuit model parameters apply for both polarizations.

The three-legged loaded element is modeled as a consequence of the four-legged element. The bandwidth and stability performance is more powerful with respect to unloaded model. The three-legged loaded element has a wider broadband than the four-legged elements since the inter-element spacing is significantly smaller which can be observed from figure 2-6.

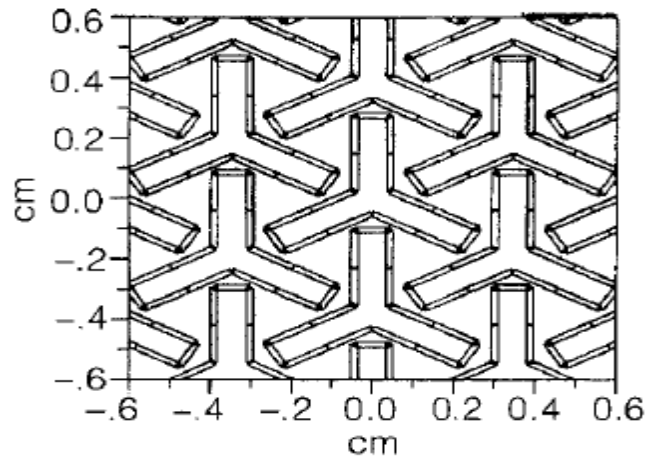


Figure 2-6 : Array of three-legged loaded FSS elements [101]

The hexagonal loop element is better than the other members of loop family in terms of the location of the first null which occurs at a higher frequency like the square spiral of the first group. Solid interior FSS elements contain the circular, triangular, square, hexagonal plates or the apertures of the corresponding shapes for the reflecting or transmitting characteristics, respectively. According to Munk [101], the main disadvantage of this group is the size of elements, which should be around $\lambda/2$ and results in a higher inter-element spacing. Thus, these elements are more sensitive to the angle of incidence.

The last group of elements is constructed combining the elements of the three groups above. The number of possible combinations is countless. As a consequence of the capacitive effect of loading, the element sizes and inter-element spacing decrease. Thus, the oblique incidence performance of the four-legged loaded element enhances. Any combination of the elements can be applied in order to benefit the major advantages of different types.

The performance parameters for FSS structures can be listed as follows:

- Bandwidth
- Sensitivity to angle of incidence
- Pass-band flatness
- Selectivity

Bandwidth definition for FSS model can be made with respect to either transmission or reflection performances. In terms of transmission properties, the frequency points where the transmitted signal is 3dB below of maximum transmission value can be defined as a bandwidth criterion. It is also possible to define bandwidth of operation using reflection characteristics of FSS model. In this case, a pre-determined value for reflected signals, i.e. -10 dB, can be regarded as the limits for bandwidth determination. These definitions are valid for band-pass cases. However, for band-stop FSS, duals of the definitions are applicable. For example, 3dB below of resonance reflection of a band-reject FSS can be chosen as bandwidth measure.

In aspects of bandwidth characteristics, FSS can be constructed for narrow or wide band operations. Single element dipoles create a narrow-band filtering behavior. For the applications where a wide spectrum of frequency is required, wideband FSS structures can be employed. Numerous articles about wideband FSS can be found. For example, Zhou et. al. designed an ultra-wideband FSS consisting of three metallic layers and two very thin dielectric layers [110]. The upper and bottom layers of the model are rectangular patches while the middle layer is a square grid as shown in figure 2-7.a. The prototype fabrication of the FSS seen in figure 2-7.b consists of an array of 67x67 unit cells.

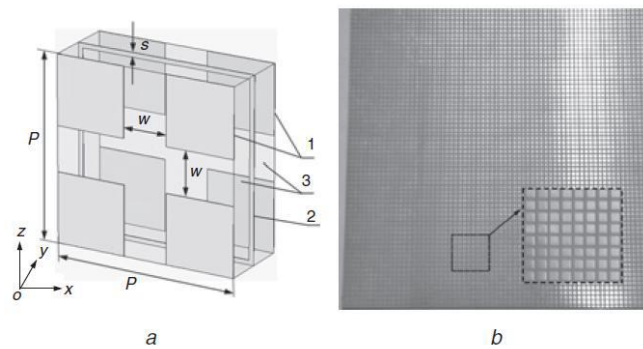


Figure 2-7 : Wideband FSS structure a) Three dimensional unit cell model b) Prototype of fabricated FSS [110]

Rather than single band of operation, it is also possible to achieve dual-band filtering responses with FSS structures. figure 2-8 show a dual-band reject FSS which consists of two nested rectangular loops placed on the same surface. The model can be further modified such that a band-pass filtering occurs between two nulls of rejection frequencies.

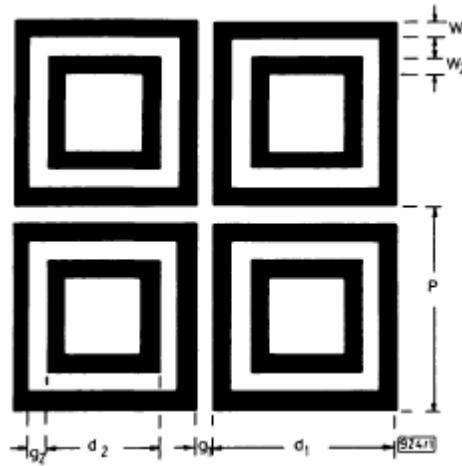


Figure 2-8 : Double square loop FSS model [111]

Another dual-band FSS model described by Yee in [105] consists of three metal layers and two dielectric layers between them. The outer metallic sheets have arrays of apertures while the inner sheet has patches. Yee also suggests the implementation of both low and high frequency FSS arrays on the same metallic sheet as shown in figure 2-9. With this configuration, a dual-band frequency selective surface is obtained. When the crossed dipole elements are used, FSS rejects the signals of two predetermined frequencies. If the elements are crossed apertures, FSS transmits the impinging waves of two different wavelengths. The sheets are kept apart from each other using the dielectric substrates.

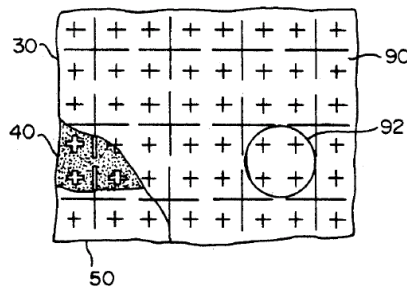


Figure 2-9 : Dual-band FSS model [105]

An extension to dual-band FSS could be the concept of "multi-band filtering". The multi-band pass or reject FSS can be designed by applying different sized elements together or fractal elements. As an example of the recent work about this concept, a tri-band stop FSS developed by Kim and Choi can be found in [112]. In this model, three distinct resonance frequencies are created using the ternary tree shaped loop loaded with a tripole element (see figure 2-10). Two of the resonance frequencies originate from the loop and the third one from the tripole placed inside the loop.

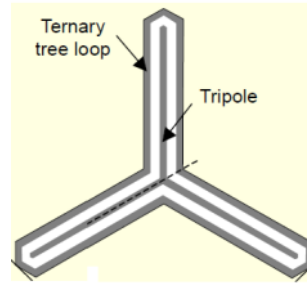


Figure 2-10 : Tri-band FSS model [112]

As the design and fabrication technologies developed, new types and applications of FSS structures have arisen enormously. Fractal elements have unique properties like small element size due to the confined. Her interesting implementation is the introduction of the fractal elements for multiband geometry which creates a long curve in a low space [106]. Thus, the spacings between the fractal elements become closer than the non-fractal ones.

The major point about the fractal elements is the self-similarity property which provides multiband behavior [107]. Self-similarity is the replication of the major element at a different scale with iterations (see figure 2-11). The multiband frequencies change with respect to the replication scale.

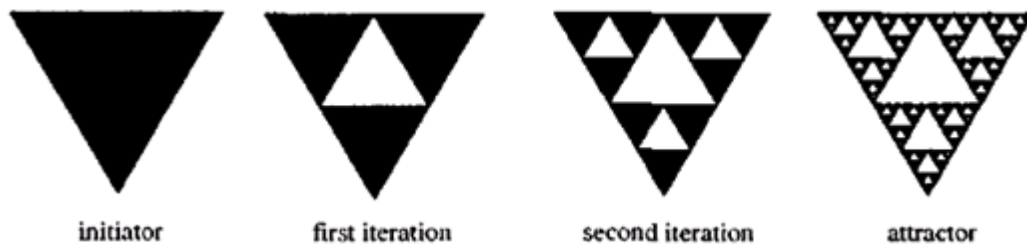


Figure 2-11 : Iterations of the Sierpinski FSS [107]

Sensitivity to angle of incidence, top flatness and selectivity properties of FSS structures are directly related with the constructional aspects. Instead of single metal sheet, FSS structures can consist of double metallic layers with dielectric substrates between the FSS layers or superstrates on them. The main targets for such modifications to FSS structures are to create a flat-top pass-band and to increase the selectivity with a fast roll-off. First of all, using two cascaded FSS layers, as shown in figure 2-12, enhances the before-mentioned properties (see figure 2-13). However, the structure is still weak for the illuminations at different angles of incidences.

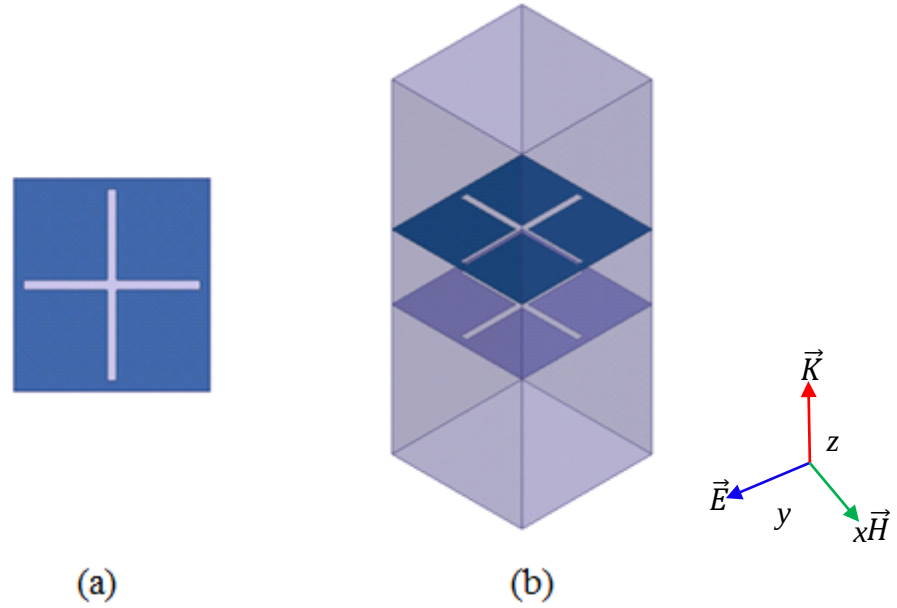


Figure 2-12 : Crossed dipole aperture FSS model: a) Single layer b) Double layer

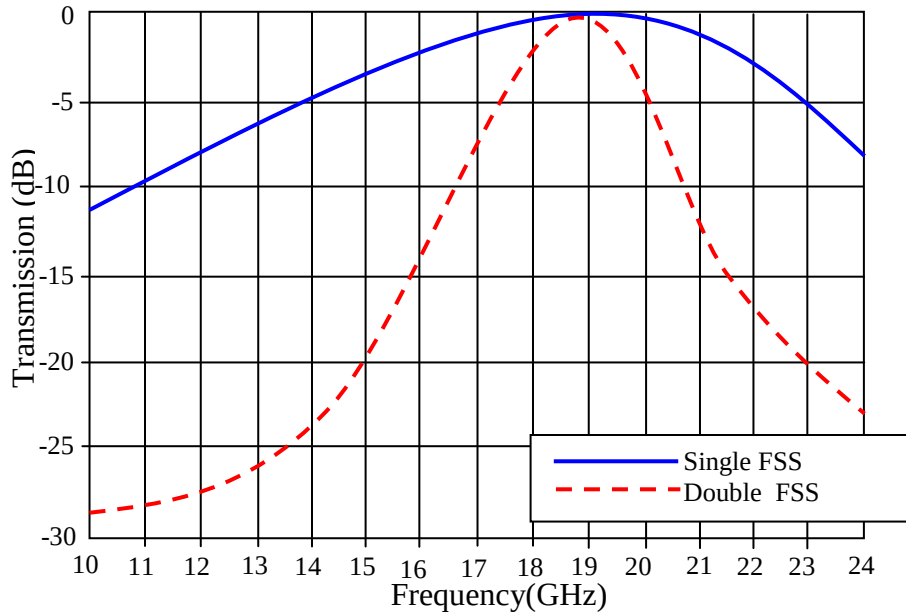


Figure 2-13 : Comparison of transmission performances of single and double layer crossed dipole aperture FSS

Another adaptation is to place dielectric layers of equal thickness and dielectric coefficient on both sides of metallic FSS layer as shown in figure 2-14. The single FSS with dielectric layers favors in broadening the bandwidth with respect to the case without dielectric layers. Nevertheless, the model suffers from the rough in-band characteristics. In other words, the flatness cannot be provided with this modification. As observed from figure 2-15, the resonance frequency of single FSS layer moves down from f_0 to $f_0/\sqrt{\epsilon_r}$.

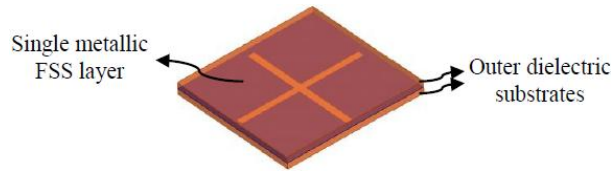


Figure 2-14 : Single crossed dipole aperture FSS with dielectric slabs

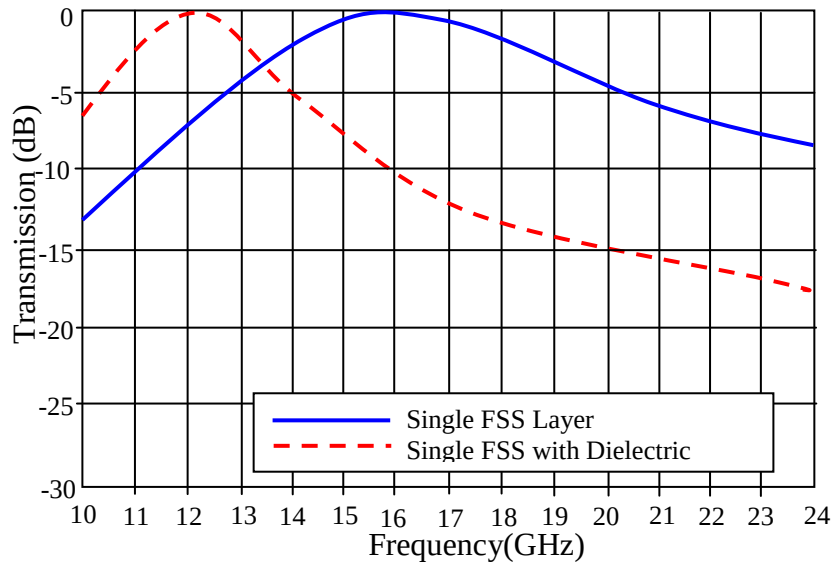


Figure 2-15 : Transmission responses of single FSS layer with and without dielectric slabs

In order to overcome both flatness and selectivity problems with a stable oblique incidence performance, "hybrid" FSS structures can be utilized. According to Munk, a real hybrid FSS consists of two outer dielectric layers, two cascaded metallic FSS sheets and a dielectric slab between them [101]. The outer dielectric layers have the same thickness and dielectric constant and serve for constant bandwidth performance against oblique incidence and polarizations. The middle dielectric layer determines the flatness of transmission curve in the pass band. FSS sheets are responsible for the bandwidth and resonance frequency. The invention patented by Munk called "Space Filter", a sandwich type FSS consisting of slot metal sheets and dielectric layers, was patented in 1978 [113]. With this invention, Munk claimed that the bandwidth of the FSS remains constant over a wide range of angle of incidence. The new structure includes the cascaded dielectric and metal layers with different thicknesses and models respectively. It is possible to construct a band-pass or band-stop space filter using the corresponding elements, i.e., slots or patches. The drawing in figure 2-16 exhibits the multilayer hybrid FSS model. The outermost dielectric layers are used to provide the necessary conductance yielding to constant bandwidth for varying angles of incidence.

The intermediate dielectric layer ensures the coupling between the periodic arrays. The total thickness of the FSS is around an odd multiple of one quarter wavelength.

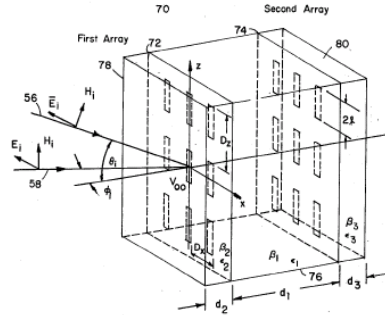


Figure 2-16 : The schematic of a multilayer “Space Filter” [113]

Hybrid configuration of crossed dipole aperture FSS is constructed as described above (figure 2-17). The transmission response of the hybrid FSS model is observed to be enhanced in terms of band-pass flatness and bandwidth with respect to single metallic layer of crossed dipole aperture FSS (Figure 2-18). Hybrid model is also more powerful against oblique incidence cases when compared to single layer FSS.

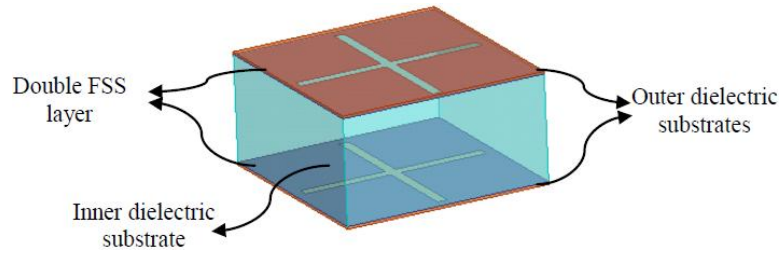


Figure 2-17 : Hybrid FSS configuration of crossed dipole aperture element

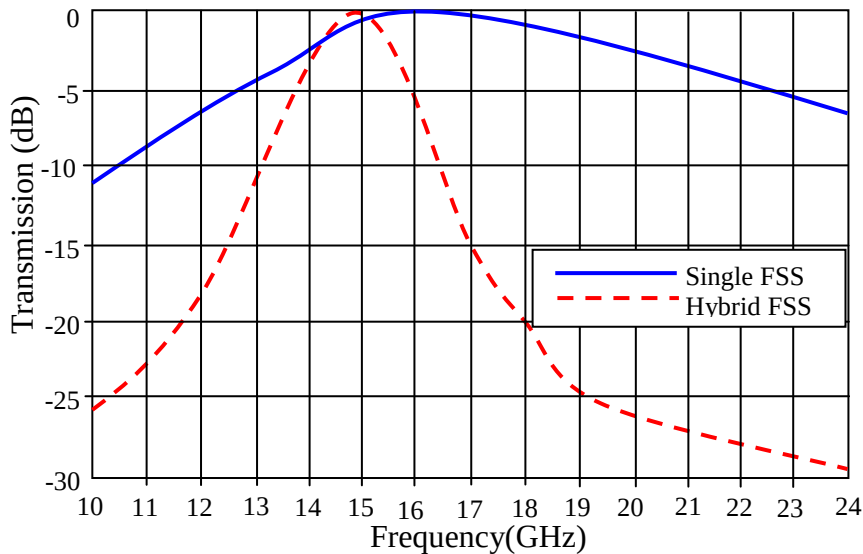


Figure 2-18 : Transmission responses of single layer and hybrid crossed dipole aperture FSS

So far, the characteristics of planar FSS are discussed, however, for some applications like nose radomes on the aircrafts, the angle of incidence of the incoming wave can reach to very high values due to the shape of the radome. Therefore, conformal FSS structures are required in order to decrease the RCS of curved structures successively, while creating no significant effect on the performance of the antenna. Implementation of FSS on curved radomes decreases the out-of-band target RCS of the antenna effectively since the incoming signal is scattered to other directions, not reflected back. Other major application of conformal FSS is the dual band reflector antenna systems. In this case a conformal FSS can be designed as a sub-reflector, which reflecting the signals of one of the operation frequencies to main reflector and transmitting the other (see Figure 2-19).

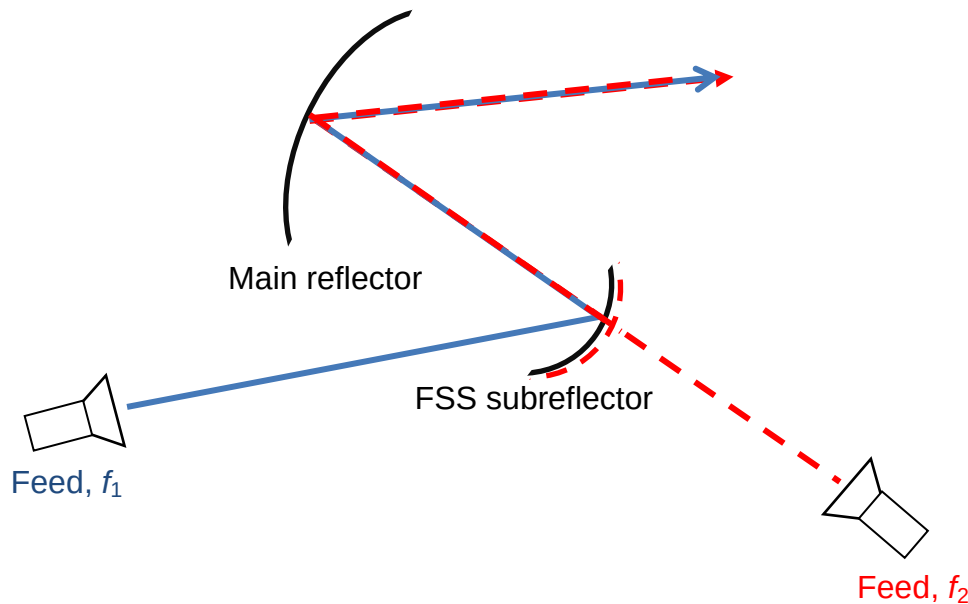


Figure 2-19 : Utilization of conformal FSS in a dual band reflector system [114]

In [115], a multi-layer array approach using open-ended waveguide antenna array combined with FSS structures conformed on a cylindrical surface is presented. A novel fabrication approach for conformal FSS structures on doubly curved bodies is explained in [116]. Design of a conformal FSS as a sub-reflector in a satellite is studied in [117].

Since the shape of the application surface is limited to one or doubly curved bodies, FSS structures are necessarily to be designed as conformal implementation surfaces. The researchers have thoroughly studied about FSS problems from 1960's up to date. With the aid of new advanced CPUs, the analysis and design durations have dramatically decreased. Different FSS structures are discovered as the fabrication and analysis techniques are improved. However, still there are challenging topics about FSS to be intensively investigated

and improved. One of these challenging issues is the application of FSS structures to conformal bodies. The conformal application brings extra difficulties different from that of classical planar problems. The effect of the elements at the end of the arrays can be neglected for large arrays of FSS. However, in curved FSS, the periodicity cannot be kept in two dimensions. On singly curved bodies like cylinders, only the periodicity in longitudinal dimension is conserved. Thus, after a planar unit cell solution, one has to prove the design with a curved structure analysis. There are some extra difficulties in designing conformal FSS. For example, the stable performances against angle of incidence are a must for conformal implementations since the incident wave can impinge on the surface of a curved FSS structure at very large angles. Thus, FSS model has to maintain its characteristics under these conditions. The orientation and registration of the elements on a curved surface also play a key role in the performance of FSS.

2.4. Analysis methods for fss structures

In order to efficiently analyze various FSS problems, different solution methods may be applicable. The analysis methods used to design and improve the performance of the FSS structures can be detailed as follows. The pioneers of the FSS design were first applied the mode matching technique, which was normally used for waveguides, to solve the FSS problems. After mode matching analysis, the equivalent circuit approach is adopted to FSS structures from a perspective of transmission lines [118]. Due to the incapacibilities of both techniques to handle the problems for the types different from the standard models like rectangular patches and apertures, more accurate numerical methods, such as method of moments, finite element and finite difference methods, are developed.

To start with the mode matching technique, it is applied by calculating the waveguide modes at the discontinuities on a basis of entire domain [118]. The procedure includes equating the expansions of fields to the boundary conditions of the discontinuity region [119]. When the scattered fields are desired to be calculated from periodic structures, the expansion into Floquet modes is required. In mode matching technique, the appropriate basis and test functions are used to find the unknowns of the problem, i.e, amplitudes of the fields. The transmission and reflection parameters are directly gathered in terms of field amplitudes into the S-matrix of the layer. The multiple layer problems are also possible to be handled with mode matching technique in a manner of cascaded networks. In [120], the numerical analysis of an arbitrarily shaped FSS with a combination of mode matching and S-matrix methods is revealed.

Another method derived for FSS analysis is the circuit equivalent method which was firstly applied by Anderson [121]. To him, the method is best fitted for the periodic, free standing, thin and perfectly conducting metal structures where the dimensions are smaller than the wavelength of the impinging wave. The impedance between the strips is described as inductive if the polarization of the wave is parallel with respect to strip alignment. Shunt capacitive elements exist if the polarization is perpendicular. The oblique incidence cases for the equivalent circuits of gridded square and double square FSS are analyzed up to 45° referenced to the normal [122]. The method is repeated with a novel attitude for the ring patch FSS in [123]. The equivalent network parameters for ring elements under oblique incidence are determined and the transmission characteristics of the analysis and measurements are compared. Some differences between the results are observed by the authors [124]. They attribute the disagreement mainly to the test environment and the weak accuracy of the circuit equivalent method than other full wave numerical solutions. If the FSS model is easy to model by lumped elements, the equivalent circuit approach can be employed to solve such kind of problems. When the FSS model gets complicated, i.e. hybridized, multilayer or conformal designs, the circuit equivalent method becomes inaccurate when compared to full wave solutions. The main drawbacks of circuit equivalent method can be listed as follows:

- The cross-polarization effects are ignored, i.e., the equivalent circuit parameters are valid for only the illumination.
- Method is incapable to handle FSS problems with large angle incidence.
- The coupling between the adjacent FSS elements is not considered.
- Analytical solutions of FSS parameters exist for a limited number of element types.

Hence, the circuit equivalent method is not adequate to solve all FSS problems. However, if one will work with a well-known FSS model like rectangular loop or the Jerusalem cross, whose parameters are defined analytically, circuit equivalent approach can be a good starting point to observe the effects of any parameter quickly.

Method of Moments (MoM) can be utilized for the solution of the operator equations of the induced currents or fields defined on an illuminated surface [125]. The characteristics of the FSS structures, i.e., transmission and reflection properties, can be extracted from these solutions [126]. The convergence of the solutions is mainly determined by the basis and test functions. MoM application for FSS brings some difficulties when the medium is multi-

layered or a thickness is defined for metal surfaces since it becomes more difficult to handle the Green's functions for 3-D structures [119].

Finite Element Method (FEM) is normally applied for bounded domain structures. The unbounded nature of FSS problems require some specific boundary conditions which eliminate the reflected fields while absorbing the incident and scattered fields appropriately. FEM is easily adapted to complex geometries and anisotropic media [127]. However, the solutions of the FEM equations require a high calculation capacity which is achieved by the high performance work stations.

Finite Difference Time Domain (FDTD) method is implemented in order to obtain broadband responses since a wide band analysis can be gathered in a single simulation. The detailed analysis of FSS with FDTD method is investigated in [128]. An important advantage of FDTD method is that there is no need for any matrix conversion to solve the problems which makes it easier when compared to other methods. However, in FDTD, the data load can be very excessive since storage at any time instant is required.

The general approach to begin an FSS design is the "unit cell" characteristics. Due to the periodic nature of FSS structures, infinite array approximations are applicable. Unit cell method employs the periodic solution of an FSS model. Using the unit cell model, the performance of an FSS structure under the infinite array conditions can be determined with the help of an electromagnetic simulation software like HFSS®, CST®, etc. [129]. In unit cell analysis, it is possible to create phase difference between the periodic boundaries, which corresponds to the oblique incidence case. The performance of a curved FSS structure has to remain stable under varying angles of incidence. Hence, this feature is required to be tested with the unit cell analysis at first.

2.5. Design approaches for curved FSS structures

The conformal FSS design requires extra points to consider when compared to planar FSS applications. The model of FSS structure is critical since the good conformability properties are needed. The circular elements are generally preferred by the designers due to good conforming properties. The singly curved structures are easier compared to the doubly ones since the periodic properties can still be maintained on a singly curved body. However, for the doubly curved cases, the periodic nature of the FSS array is very difficult to be preserved. For example, on the surface of a sphere, it is impossible to arrange the FSS elements in a two dimensionally periodic sense. Thus, different approaches should be

introduced to realize the FSS performance on a doubly curved structure. One of the approaches is to use an FSS model which is insensitive to shifts in the alignment of elements in multilayered structures. In the single element planar unit cell design period, the performance of the FSS model can be tested in aspects of the predefined properties.

Furthermore, a different periodicity approach can be implemented which preserves the period between the inter-elements in one dimension and the subsequent sequences of FSS elements in the orthogonal dimension. Another method for the doubly curved structures can be to optimize the performance of the total structure by arranging the positions of the elements using the finite model analysis. This method would give the most reliable results, however, it is inefficient and time-consuming compared to other approaches.

2.6. Design methodology for conformal frequency selective surfaces

The design procedure for the conformal FSS applications starts with the determination of the unit cell model definitions as in the planar cases. The required characteristics like band pass/band stop filtering, levels of insertion loss or reflection coefficients, oblique incidence performance etc., are specified. Then, the starting points for the unit cell parameters are determined either by using circuit equivalent approach if applicable or through simulation based fast search or by scaling the known model parameters for a specific application. The starting point determination brings the designer to a feasible solution region instead of searching the final optimum sizes in the whole space. Generally, the optimum design parameters will be within the close neighborhood of these starting values. The next step is to analyze the FSS unit cell. The design is first verified for the normal incidence case with 2D infinite planar array model. FSS performance for the oblique incidence cases can also be observed and fine tuning of unit cell parameters to satisfy the design goals for oblique incidences is pursued. Oblique incidence conditions are created by applying phase differences to the periodic boundaries of the unit cell [130].

The array topology is very important for the conformal FSS applications. The placement of the elements on a doubly curved structure is not straightforward as for the planar or singly curved applications. When the surface is planar or singly curved, i.e. cylindrical surfaces, the periodicity between the FSS elements can be preserved in two dimensions. However, if the surface is doubly curved, i.e. spherical surfaces, the simple periodicity techniques are not applicable since it is impossible to maintain the periodicity in two dimensions on these surfaces. Instead, alternative techniques for the arrangement of the FSS elements are required

as mentioned before in this chapter. Therefore, the designer is strongly suggested to determine the final array parameters like inter-element spacing and array configuration. These modifications will benefit in the realization of the eventual FSS radome structure as shown in chapter 4. FSS parameters can also be optimized for the new array topologies [130].

Verifying FSS performance using the curved models is essential since the real application surface will not be planar. The unit cell analysis of the single element is utilized for planar and infinite conditions, which does not fit the real operation conditions. The interaction between the conformal elements may be different from the planar arrays. Furthermore, the edge diffractions can be taken into account with only a finite conformal analysis. Therefore, the analysis of the FSS model with singly or doubly curved models is required in order to observe the filtering characteristics of curved FSS elements. However, the finite models usually create a heavy calculation load which cannot be solved by even high performance work stations [130].

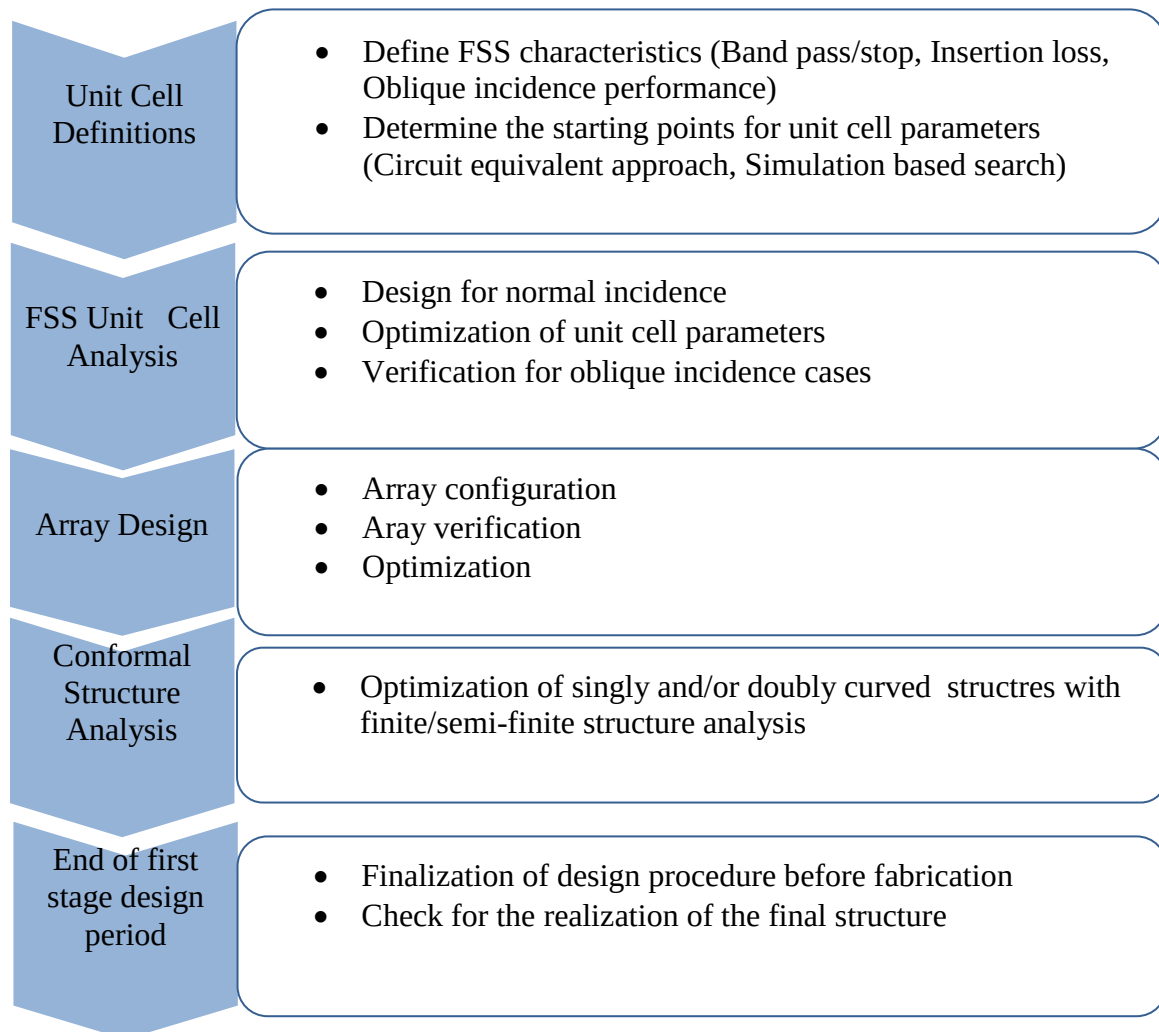


Figure 2-20 : Flowchart representation of conformal FSS design procedure

We can overcome this problem by a semi-finite solution, the edge diffractions and the performance of curved elements can be observed in the finite dimension. On the other hand, the periodicity in the longitudinal dimension makes the problem convenient to analyze by reducing the calculation load [130].

When the finite or semi-finite curved FSS analysis is completed, the design procedure before fabrication is finalized. Nonetheless, the designer needs to check the final version of the FSS structure in aspects of realizability for fabrication process. The design procedure for the conformal FSS applications is also summarized with a flow chart in figure 2-20.

2.7. Conclusion

This chapter set out to better understand the frequency selective surface and to gain a better understanding of its design methodology and the results of this investigation show that:

- There are four main categories of FSS.
- There are many factors affecting the performance of FSS structures, such as the Bandwidth ,Sensitivity to angle of incidence ,Pass-band flatness ,Selectivity .
- There are many different solution methods may be applicable to design and improve the performance of the FSS structures, such as the electrical properties.
- They are constructed with any of metallic patches on dielectric slabs or slots in metals.

3.1. Introduction

In the last few years, different applications and expected uses of metamaterials in electromagnetic engineering require a precise knowledge of their electromagnetic properties and several steps of design, in any case the metamaterial homogenization is an important step in the design process, because it is related to the environmental scale and to the effective medium nature (isotropy, chirality ...).

This chapter introduces the metamaterials homogenization. It is divided in two principal parts: the first one treats the effective medium properties and the homogenization criterion. The second one deals with the methods used to extract the effective electromagnetic parameters of the homogenized medium. This last is described by the effective dielectric permittivity ε , magnetic permeability μ and magneto-electric coupling ζ_0 . These effective parameters extraction methods use either the S parameters method or the field averaging (field summation) method in the material region of interest.

3.2. The effective media

The most elegant and concise ways to state the fundamentals of electricity and magnetism are represented by Maxwell's equations, which are a set of equations describing the interaction between fields, sources, and material properties. Impinging fields in a system can influence the organization of electric and magnetic dipoles in a medium. Fields can induce polarization and magnetization to some degree, depending on the particular material involved. The electromagnetic properties of a material are described by two material parameters: the dielectric permittivity ε and the magnetic permeability μ , which characterize the coupling between the material and the electric and magnetic field components, respectively. These two parameters, along with other values, which are the refractive index $n = \sqrt{\varepsilon\mu}$ and the impedance $z = \sqrt{\frac{\mu}{\varepsilon}}$, are essentially macroscopic effective parameters. They are used to describe the overall average response of the material as a whole. Microscopically, a piece of crystal consists of atoms arranged in a periodic or non periodic manner with a lattice constant of a few angstroms. On the atomic scale, in each atom or molecule, tiny electric dipoles can be excited by the electric component of incident wave and subsequent radiation of the energy in the dipoles occurs with a certain delay in time. The excited dipoles create a periodic local field in the crystal, referred to as Lorentz local field; therefore the field distribution inside the crystal is certainly not uniform. However, the incident wave does not really feel the

underlying inhomogeneity in the crystal, nor does it feel the processes of absorption and radiation. On the macroscopic scale, the detailed features and responses of the inhomogeneous structure are averaged, and relationships can be established between the macroscopic field vectors in Maxwell's equations, namely the electric field $\vec{E}$, the magnetic field $\vec{H}$, the electric displacement field $\vec{D}$ and the magnetic flux density $\vec{B}$. This is the origin of the permittivity and permeability parameters of materials. At different frequency range, this complicated physics is usually described using the transmittance and reflectance of waves with a certain retardation (delay). In addition, absorbed waves account for losses in the material and can be related to the imaginary parts of the complex constitutive parameters that are electric permittivity, magnetic permeability, and refractive index.

Similarly, the scale of inhomogeneities in a metamaterial is much smaller than the wavelength of interest. The inhomogeneity scale corresponds to the lattice constant of the artificial structure for the case of periodic metamaterial as shown in figure 3-1.

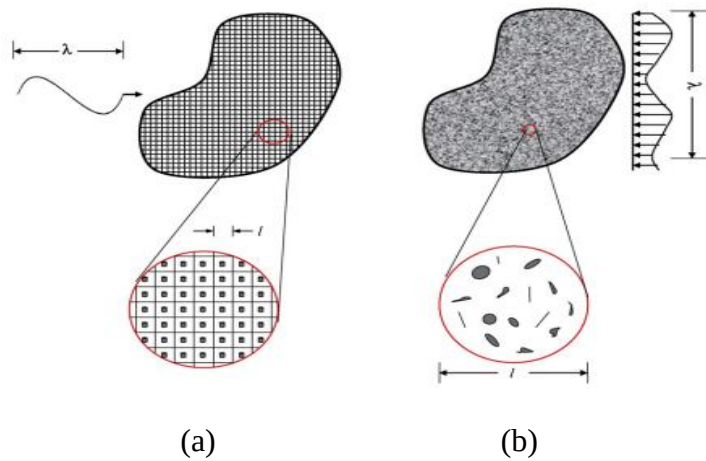


Figure 3-1: Composite material with (a) periodic (b) non-periodic structure

In figure 3-1, the largest heterogeneity length scale l should be small enough compared to the smallest wavelength scale λ , that is of physical importance. For a periodic medium (in the left of the figure) l is the unit cell size while in a random medium (in the right of the figure) it should be the smallest size of a sphere which at any position is a statistical representative of the entire body.

Then the key of homogenization is when the size of the unit cell is much smaller than the operating wavelength [131][132]. This property is called the homogenization criterion. Since the metamaterial can be seen as a homogeneous medium, the effective parameters can be determined and the effective electromagnetic response is validated as shown in figure 3-2.

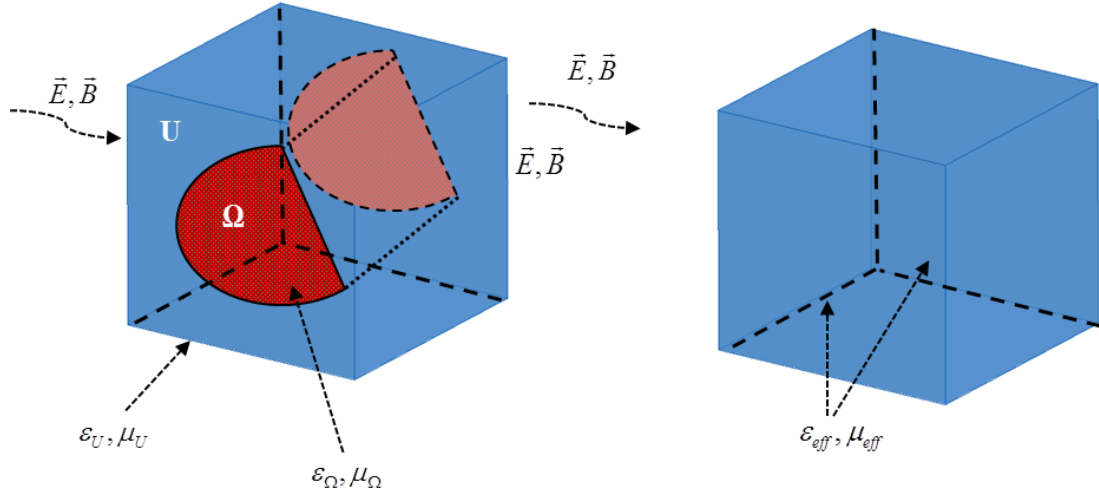


Figure 3-2 : A two-phase heterogeneous unit cell (left) and its equivalent to homogeneous unit cell [133]

This fact explains again why metamaterials made from basic constituents are identified as materials instead of devices [134]. From the point of view of Maxwell's equations, a material is a collection of sub-wavelength units with global properties described by ϵ and μ . Through the dedicated design of meta-atoms, which is usually a delicate metal-dielectric structure, metamaterial research allows us to tailor the electromagnetic response of media in an unprecedented manner. Finally, there are cases such as bianisotropic media where the electromagnetic response is not sufficiently described with ϵ and μ tensors because of the complexity of the unit scatterers. Such structures may require additional material parameters like magnetoelectric coefficients, which link the electric field vectors and the magnetic ones. Although some important metamaterial elements like split-ring resonators are bianisotropic to some extent.

3.3. Effective parameters extraction

In cases where spacing between scattering elements are much smaller than the operating wave length, the metamaterial can be considered as homogenous and it responds to electromagnetic waves in a similar way [135]. In these cases, It is possible to describe metamaterials as continuous materials [136] being characterized by effective parameters without considering local fields distributions. This part of this chapter treats classical and numerical homogenization approach, which is the S parameters extraction method (Fresnel inversion method). However, there is another method called the field summation method which is an averaging approach [137].

3.3.1. Fresnel inversion extraction methods

Smith et al [138] proposed a method to extract the constitutive parameters by using scattering parameters S , which is suitable to be applied for both numerical simulation and experimental measurements. Several reports have been published [139], most of them deal with the isotropic parameters of permittivity and permeability. However, it has already been demonstrated [140] that most metamaterials are intrinsically bianisotropic due to the asymmetry of the composite components such as split-ring resonators. For a bianisotropic metamaterial, the propagation wave vector is related not only to permittivity and permeability but also to the magnetoelectric coupling coefficient [140]. Moreover, there are some special metamaterial designs that take advantage of the property of bianisotropy. Consequently, in order to obtain a true picture for these metamaterials, it is important to present not only the effective parameters of permittivity and permeability but also some other important parameters, such as the magnetoelectric coupling coefficient ζ_0 . Only then one can provide true data for the effective parameters. Figure 3-3 shows the schematics of a homogeneous bianisotropic material slab.

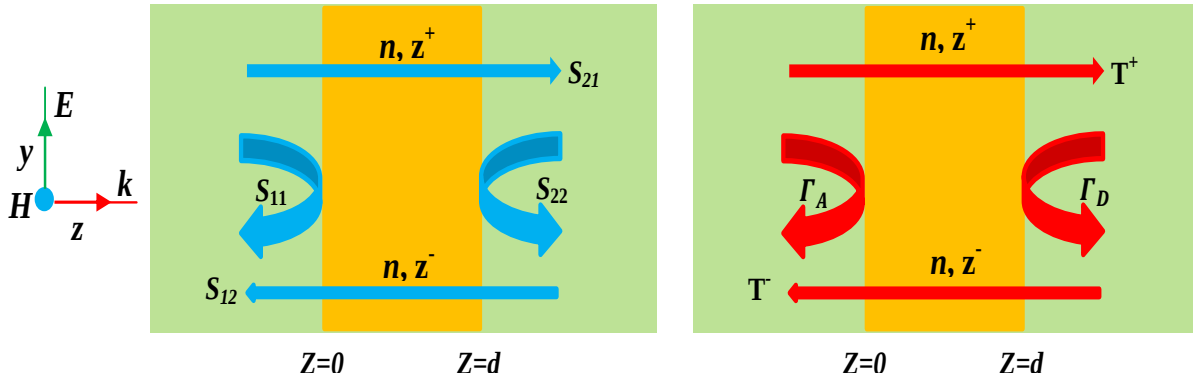


Figure 3-3: Transmission and reflection coefficients in bianisotropic metamaterial

The metamaterial effective parameters extraction requires two steps [141]:

1- Computation of the homogenized metamaterial forward and backward wave impedance (z^+ , z^-) and complex refractive index (n) which are expressed in terms of reflection and transmission coefficients (S_{ij}) for a unit cell.

2- Retrieval of the electromagnetic constitutive parameters (permittivity, permeability and magnetoelectric coupling coefficient) from the S_{ij} of the unit cell.

3.3.1.1 Metamaterial wave impedances and refractive index computation

In figure 3-3, a plane wave polarized in the y axis propagates along the z direction and is incident upon the structure. We suppose that the configuration, in figure 3-3 has a bianisotropic property, the electric field in the y direction induces a magnetic dipole in the x direction, while the magnetic field in the x direction also induces an electric dipole in the y direction [142].

By assuming that the medium is reciprocal [143] and that the harmonic time dependence is $e^{-j\omega t}$, implying that $\varepsilon'' > 0$ and $\mu'' > 0$ simultaneously at all frequencies. We use primes and double primes for denoting real and imaginary parts of a complex quantity.

From Maxwell's curl equations [143] and according to [137] we obtain the following expressions:

$$z^+ = \frac{-\Lambda_2 \pm \sqrt{\Lambda_2^2 - 4\Lambda_1\Lambda_3}}{2\Lambda_1} \quad (3-1)$$

$$z^- = \frac{\Lambda_4 z^+ + 1}{z^+ + \Lambda_4} \quad (3-2)$$

$$\text{Where} \quad \Lambda_1 = S_{21}^2 - (1 - S_{11})(1 - S_{22}) \quad (3-3)$$

$$\Lambda_2 = 2(S_{11} - S_{22}) \quad (3-4)$$

$$\Lambda_3 = (1 + S_{11})(1 + S_{22}) - S_{21}^2 \quad (3-5)$$

$$\Lambda_4 = \frac{S_{11} + S_{22}}{S_{11} - S_{22}} \quad (3-6)$$

$$\text{And} \quad n = \frac{1}{k_0 d} \left[\mp (\ln T)'' \mp 2\pi m - j(\ln T)' \right] \quad (3-7)$$

Where

$$T = e^{jk_0 nd} \quad (3-8)$$

And $(m = 1, 2, 3, \dots)$ denotes the branch index value. It can be seen from equation (3-7) that the index n have multiple solutions depending on m values, the correct choice of m for unique retrieval of n can be realized by different technique [144].

Note that the correct sign of z^+ , z^- and n corresponds to that for which the passivity conditions are satisfied: $(z^+)' > 0$, $(z^-)' > 0$, and $(n)'' \geq 0$.

3.3.1.2 Calculation of the constitutive electromagnetic parameters

According to [137]. We deduce the effective parameters expressions:

$$\zeta_0 = \frac{(2 - \Gamma_C)^2 - (4\Gamma_A^2 + \Gamma_C^2)}{4i\Gamma_A\Gamma_C / n} \quad (3-9)$$

$$\mu_x = \frac{\Gamma_C(n^2 + \zeta_0^2)}{n(2 - \Gamma_C) - 2n\Gamma_A - i\Gamma_C\zeta_0} \quad (3-10)$$

$$\varepsilon_y = \frac{n^2 + \zeta_0^2}{\mu_x} \quad (3-11)$$

3.3.2. Validation of the constitutive parameters result by inversion method FI

In order to validate and test the retrieval method, based on Fresnel inversion method, we compare the results obtained with those available in the literature [138].

As a test structure we use that of smith et al in [138]. This later, shown in figure 3-4, consists of two types of conducting elements, a split ring resonator (SRR) and wire both made of copper (thickness 17 μm). The split ring and the wire are positioned on opposite sides of the dielectric slab of thickness 0.25 mm.

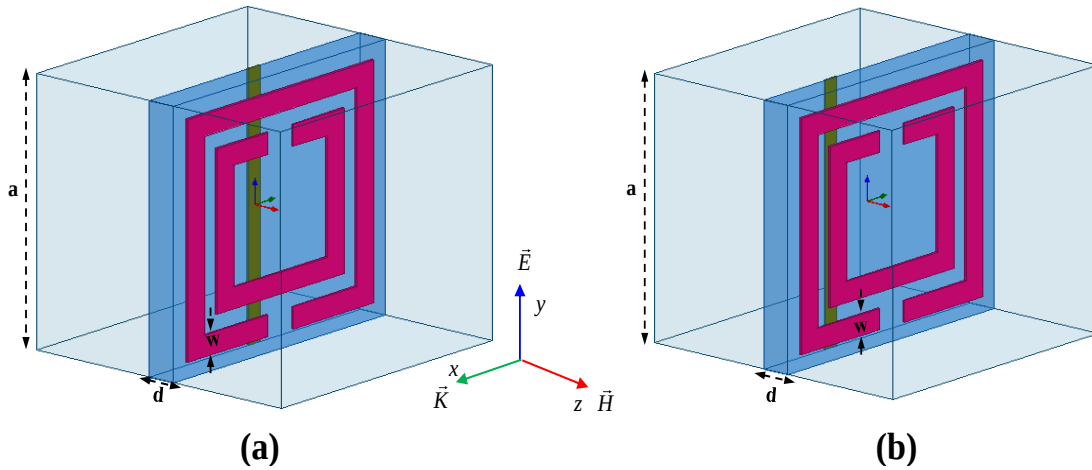


Figure 3-4 : The split ring resonator SRR (a)-symmetric (b)-Asymmetric

The extracted electromagnetic parameters of the symmetric and asymmetric unit cell are shown in figures 3-5 to 3-7 respectively.

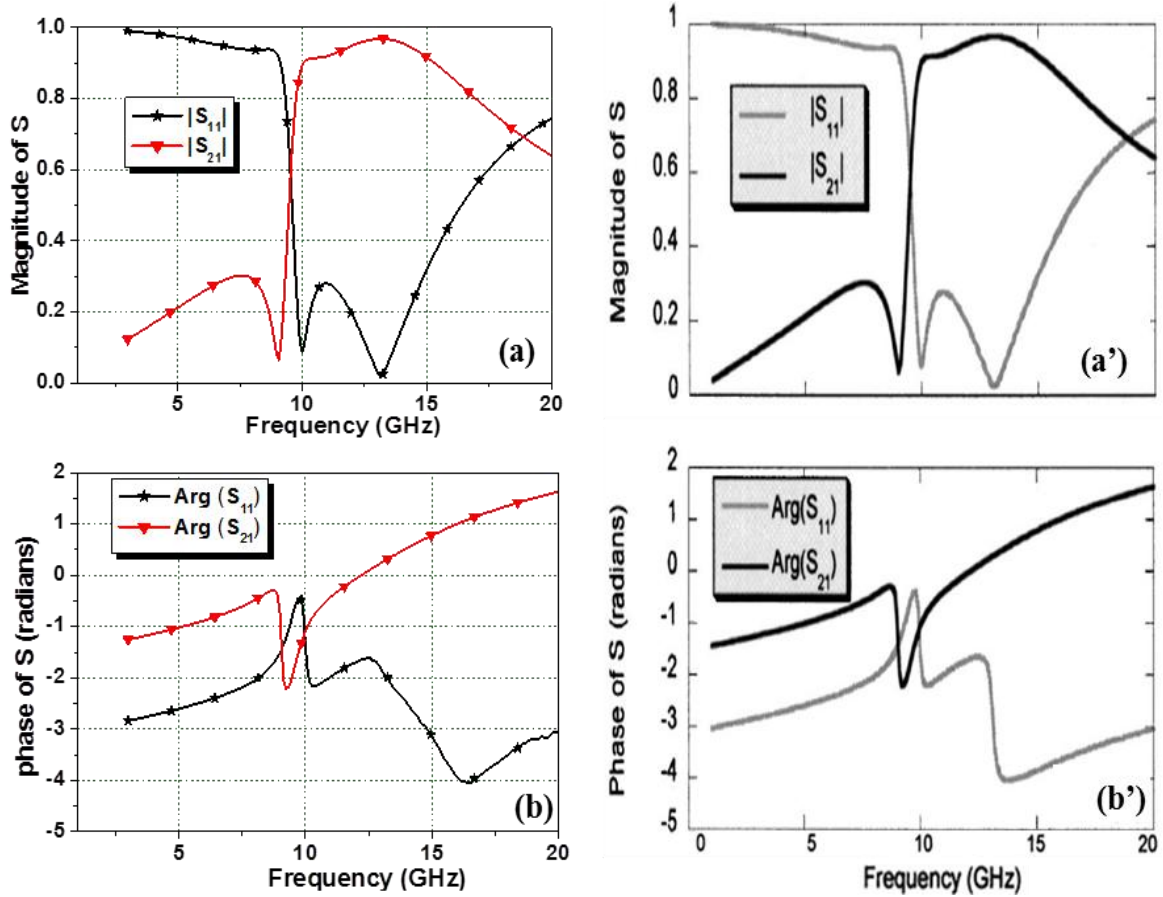


Figure 3-5 : Comparison between our extraction program results (left) and those obtained from [138] (right) for symmetric unit cell: (a)-(a') Magnitude of S, (b)-(b') phase of S

Figure 3-5 shows a good agreement between our results and those obtained by Smith and al in [138] for the symmetrical unit cells.

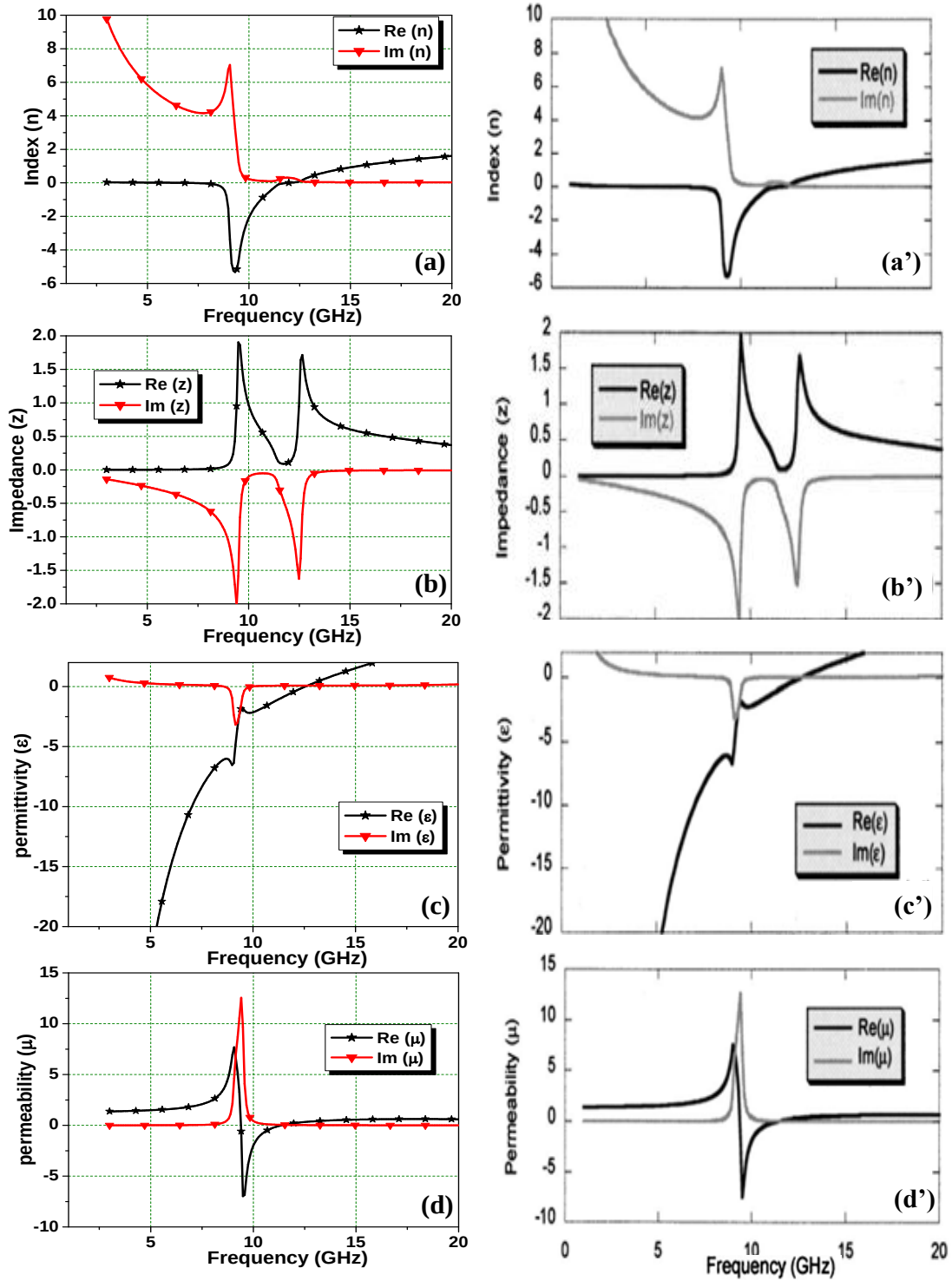


Figure 3-6 : Comparison between our extraction program results (left) and those obtained from [138] (right) for symmetric unit cell: (a)-(a') refraction Index n , (b)-(b') Impedance z , (c)-(c') relative dielectric permittivity ϵ , (d)-(d') relative magnetic permeability μ

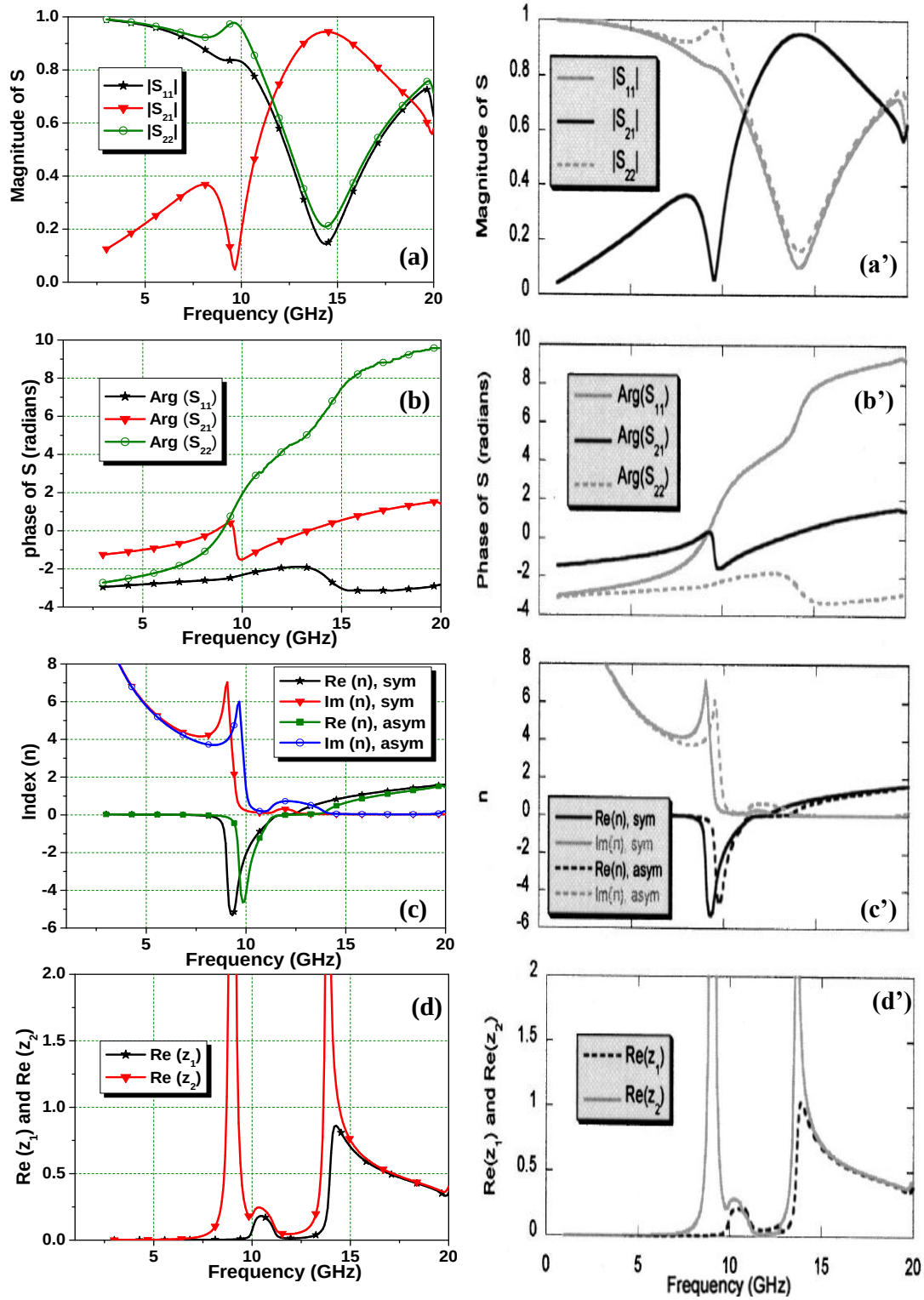


Figure 3-7 : Comparison between our extraction program results (left) and those obtained from [138] (right) for asymmetric unit cell: (a)-(a') Magnitude of S ; (b)-(b') phase of S ; (c)-(c') Refraction Index n ; (d)-(d') Impedance z

Figures 3-5 to 3-7 show a good agreement between our results and those obtained by Smith and al in [138] for the symmetrical and asymmetrical unit cells. The asymmetrical unit cell is obtained by shifting the conductor wire 0.75 mm away from its centered position corresponding to the symmetrical one.

3.4. Design of the homogenized metamaterial

Our planar metamaterial is a periodic arrangement of cross unit cells, it has two equal periods a in the x and y directions as shown in

figure 3-8-(a). The cross type unit cell consists of a metallic cross shape conductor printed on a dielectric substrate of thickness d . The metallic conductor has a length L , a width w and a thickness t , as shown in figure 3-8-a.

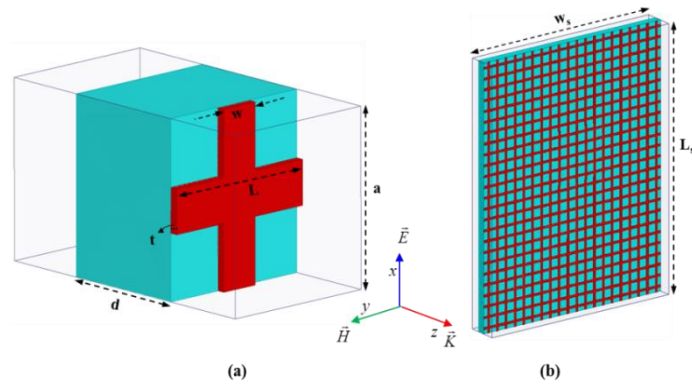


Figure 3-8 : Homogenized metamaterial: (a) unit cell, (b) metasurface

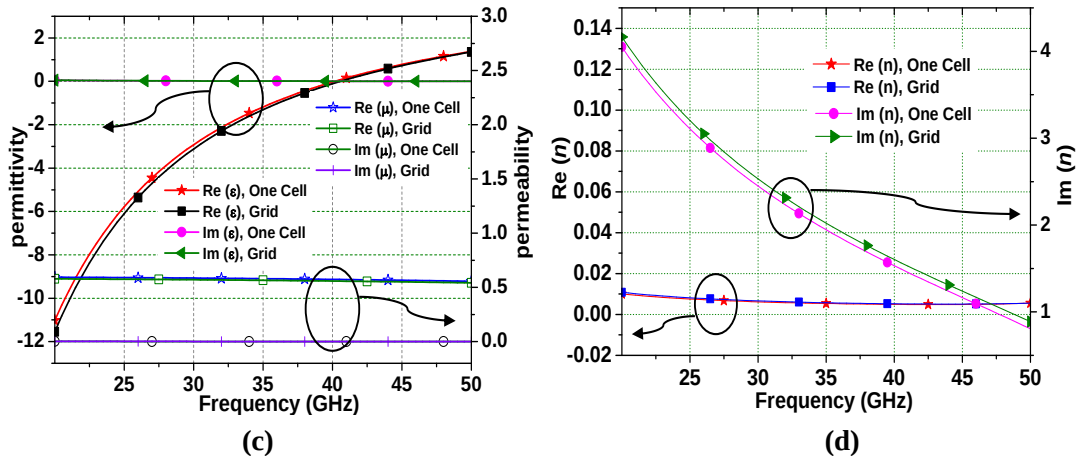


Figure 3-9 : (c) permittivity and permeability. (d) Refractive index

3.5. Symmetric and asymmetric connected cross type metamaterial

The difference between the symmetric and asymmetric connected cross unit cells can be shown in the scattering parameters S . The scattering parameters are the means of characterizing the artificially structured metamaterial.

This section gives a comparative study of parameters for symmetric and asymmetric metamaterial structures.

We have in section 3.3, methods used to calculate all the material parameters for a general bianisotropic structure based on numerical simulation of wave propagation in periodic media.

The bianisotropy arises due to coupling between the electric and magnetic fields. We can easily verify the bianisotropic character of the structure, if there reflection scattering parameters are different ($S_{11} \neq S_{22}$).

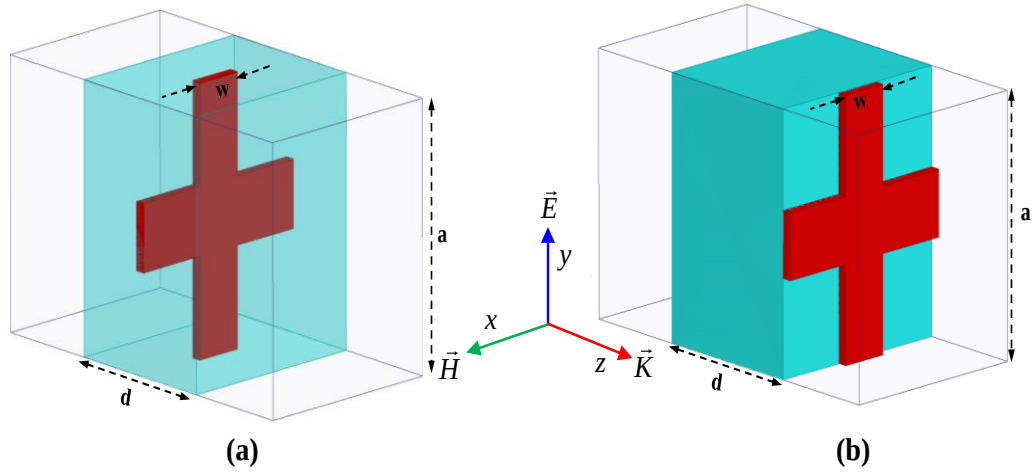


Figure 3-10 : Metamaterial connected cross unit cell: (a) symmetric; (b) asymmetric

The incident wave in the figure 3-10 is a plane wave with its wave vector $\vec{k}$ in the z -direction and the $\vec{E}$ field polarized in the y -direction, while the magnetic field vector $\vec{H}$ oriented along x -direction.

The dimensions of the symmetric cross unit cell are the same as for the asymmetric structure, where the period $a = 1$ mm, the thickness $d = 1.168$ mm, the conducting strip width $w = 0.1$ mm and the metallization thickness $t = 0.035$ mm.

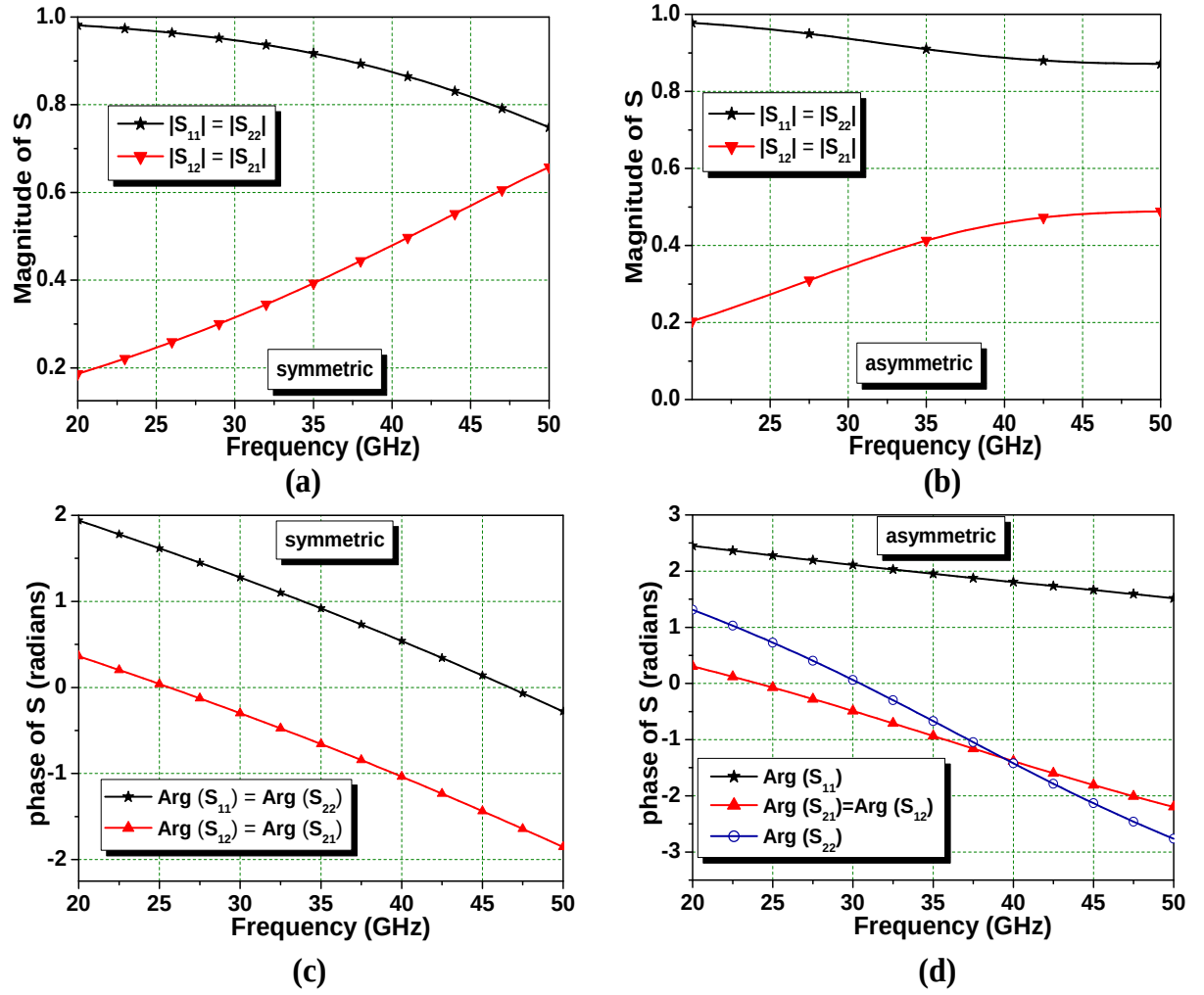


Figure 3-11 : (a) Magnitude and (b) phase of the calculated S parameters for the symmetric unit cell (c) Magnitude and (d) phase of the calculated S parameters for the asymmetric unit cell

Figure 3-11, shows the amplitude and the phase information of the calculated S parameters for the symmetric and asymmetric cross type metamaterial.

It can be seen that S_{11} is equal to S_{22} and S_{12} is equal S_{21} for the symmetric cell. However, for the asymmetric structure there is a great contrast in the phase of S_{11} and S_{22} .

The extracted electromagnetic parameters of the symmetric and asymmetric unit cell for different values of the thickness d are shown in figure 3-13.

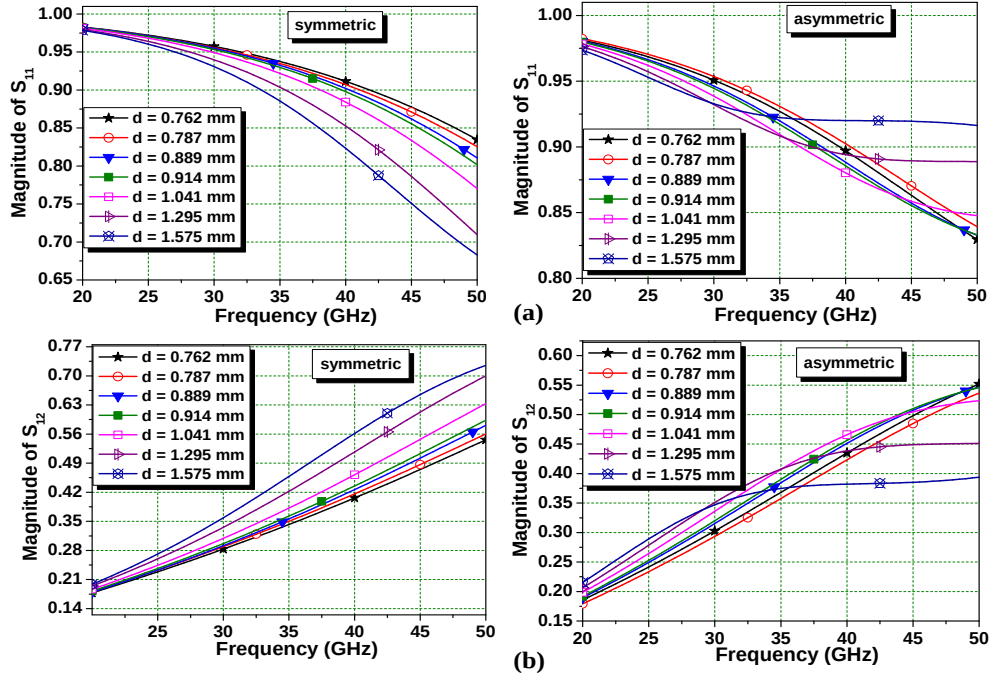


Figure 3-12 : Symmetric and asymmetric unit cell S parameters for different values of d (a) - S_{11}
(b) - S_{12}

Evidently ζ of symmetric unit cell is close to zero, while ζ of asymmetric structure shows great variations. Indeed, in general the values of ζ depend strongly on how the surface of the composite is terminated.

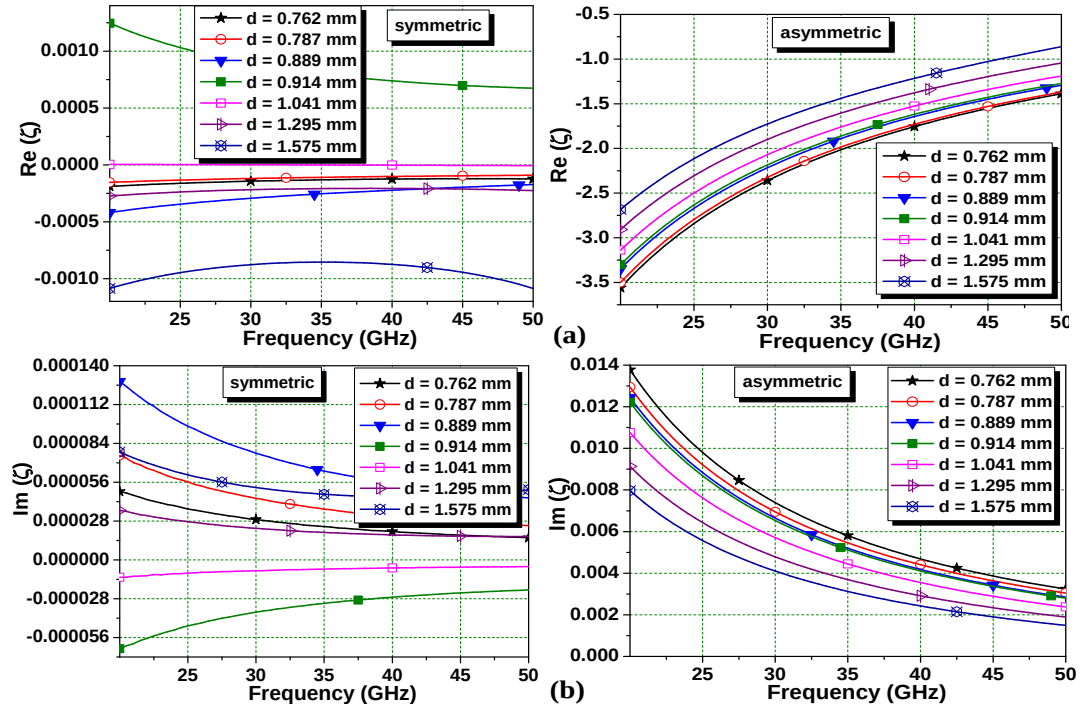


Figure 3-13 : Symmetric and asymmetric unit cell magnetolectric coupling coefficient ζ for different values of d , (a) real parts, (b) imaginary parts

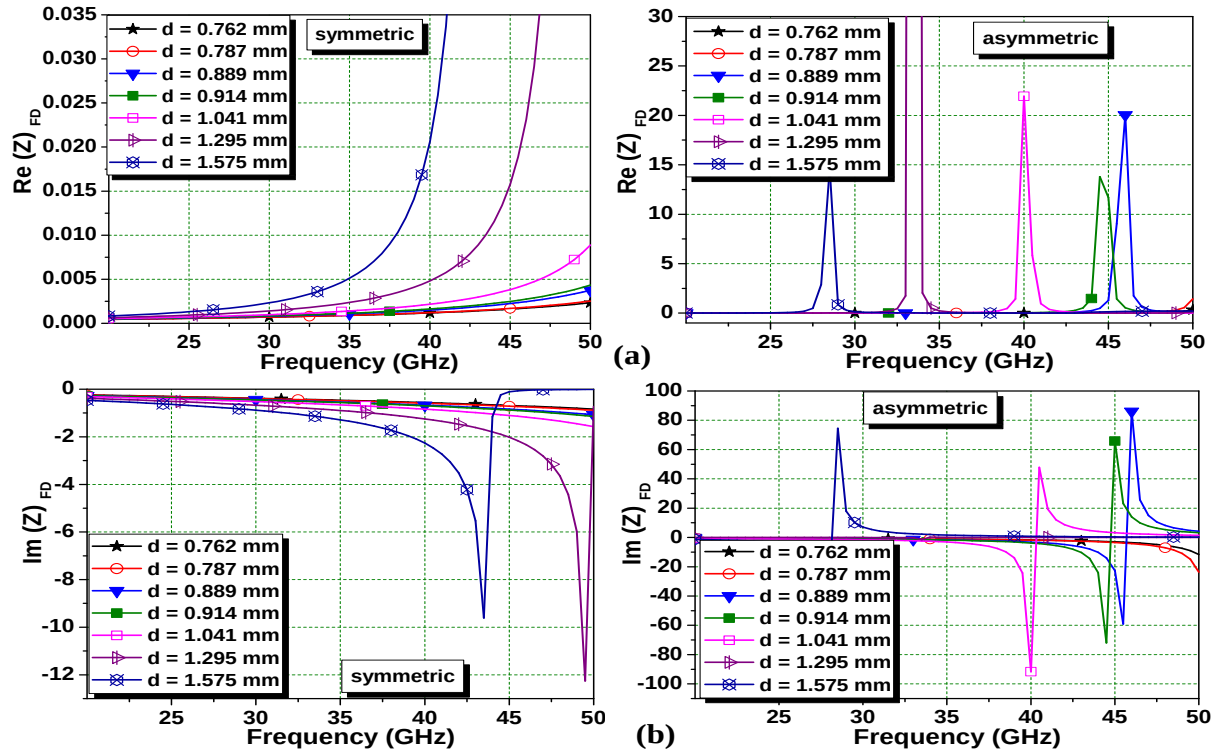


Figure 3-14 : Forward impedance $(Z)_{FD}$ (or Z^+) for the symmetric and asymmetric unit cell for different values of d , (a) real part, (b) imaginary part

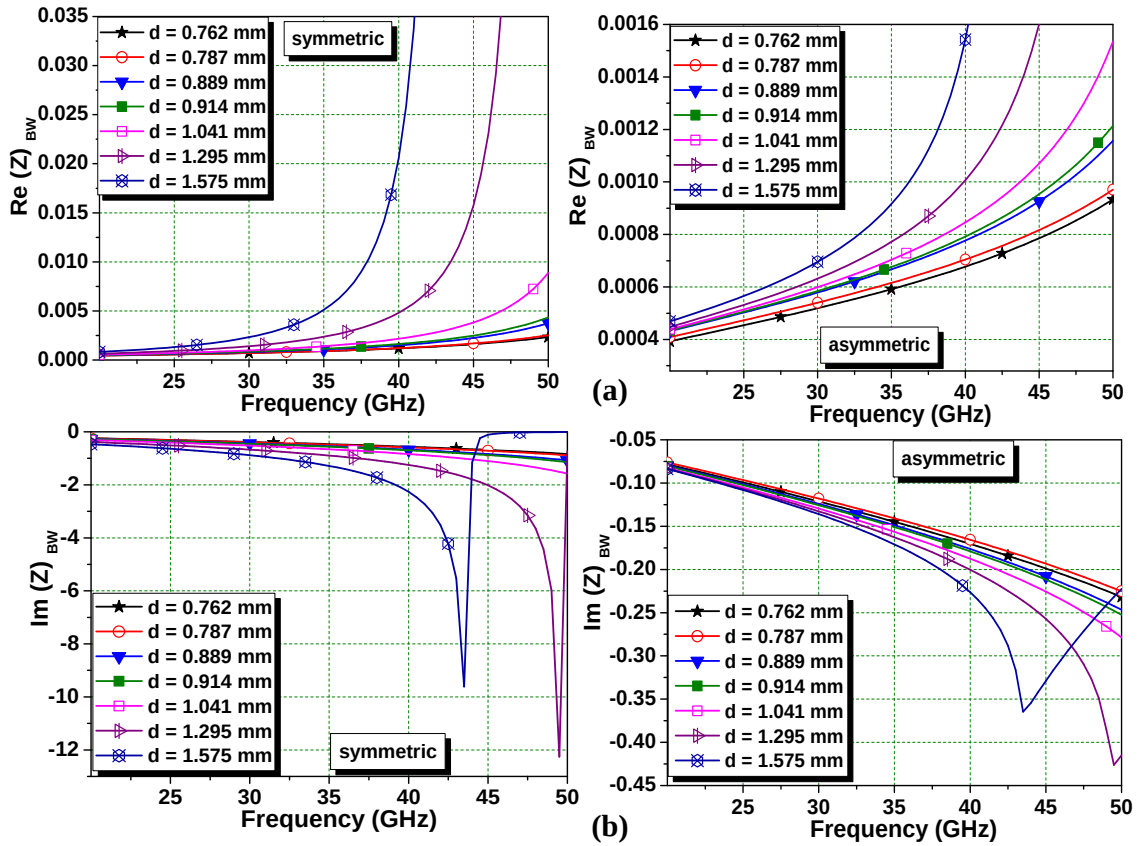


Figure 3-15 : Backward impedance $(Z)_{BW}$ (or Z) for the symmetric and asymmetric unit cell for different values of d , (a) real part (b), imaginary part

The impedance is clearly different, as shown in figure 3-14 and figure 3-15. Obviously the symmetric unit cell presents the same impedance for the forward and backward propagation ($(Z)_{FD}=(Z)_{BW}$). However, the asymmetric unit cell presents different wave impedances for the forward and backward propagation.

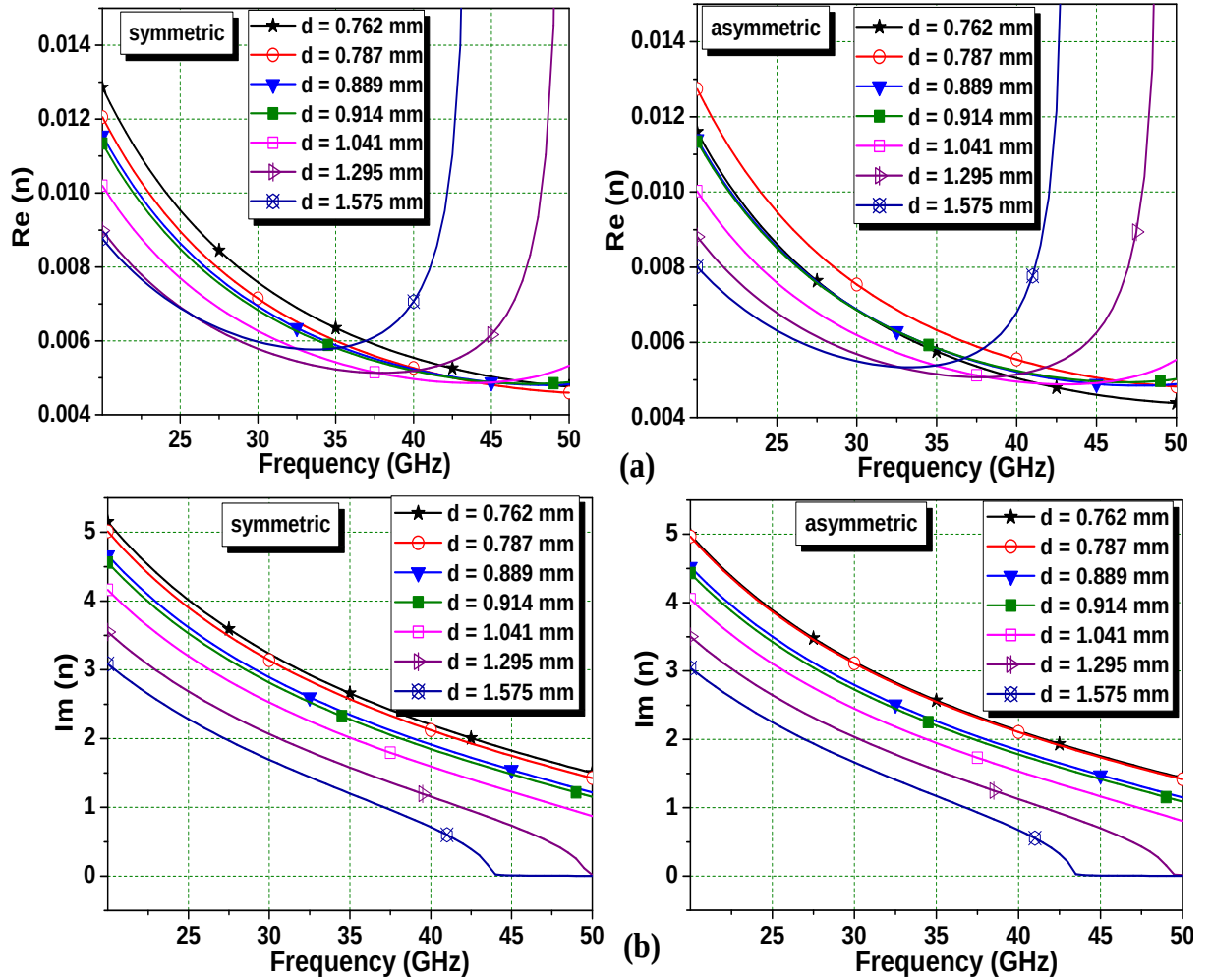


Figure 3-16 : Refractive index n , of the symmetric and asymmetric unit cell for different values of d , (a) real part, (b) imaginary part

The refraction index n in figure 3-16 is nearly the same for the symmetric and asymmetric structures.

From the point view of fabrication process, it is clear that the asymmetric structure with the conductor cross printed in the top face of a printed circuit board, is easier to fabricate. For this reason, we have chosen to use in our design the asymmetric cross unit cell.

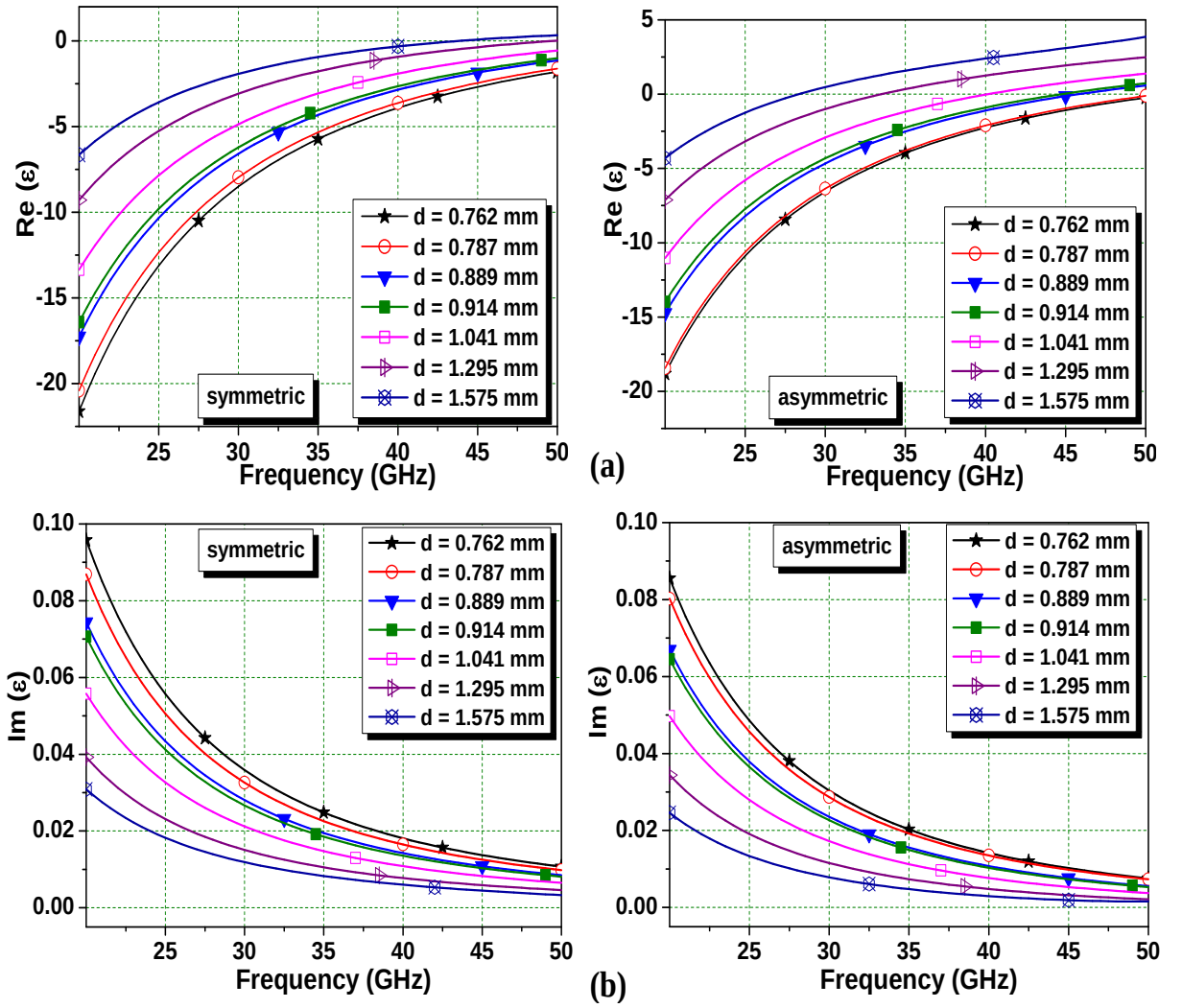


Figure 3-17 : Effective permittivity ϵ of the symmetric and asymmetric structures for different values of d , (a) real part, (b) imaginary part

From figure 3-17, it can be seen that the plasma frequency shifts to lower values when the value of d increases.

The plasma frequency shifting it's not surprising, as there is likely an interaction between the substrate and the conductor that leads to somewhat different material properties depending on their relative positions.

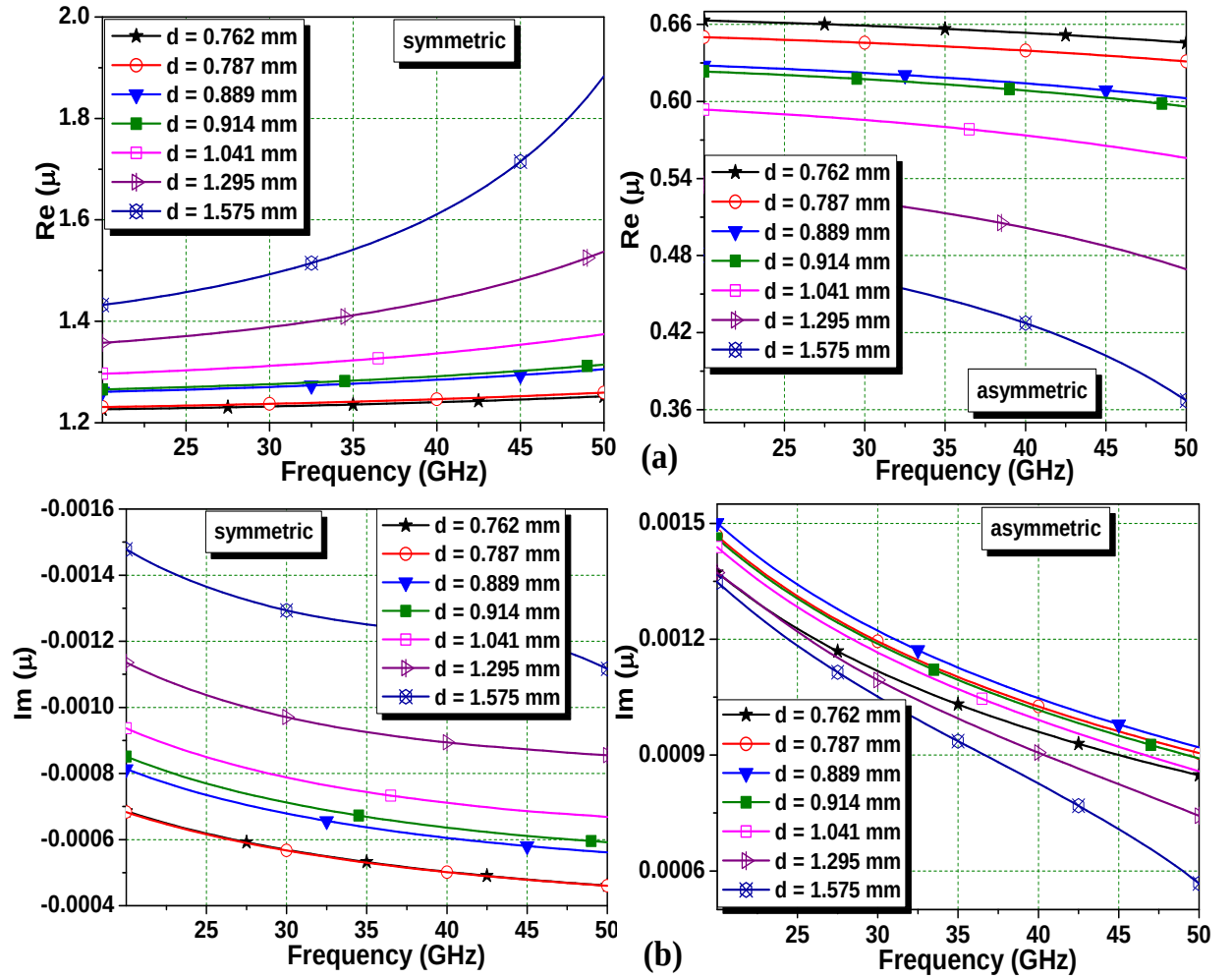


Figure 3-18 : Effective permeability μ of the symmetric and asymmetric structures for different values of d , (a) real part, (b) imaginary part

We note in figure 3-18 a great contrast for the permeability between the symmetric and asymmetric structures. As we know a straight current carrying conductor produces circular magnetic field lines around itself at all points along its length. The direction of rotation of this magnetic field depends upon the direction of current flow through the conductor, the “right hand” rule.

The real part of permeability in the symmetric unit cell is greater than in the asymmetric cell. Due to the magnetic field lines in the symmetric structure are more than number and intensity of which are in the asymmetric structure as shown in figure 3-19.

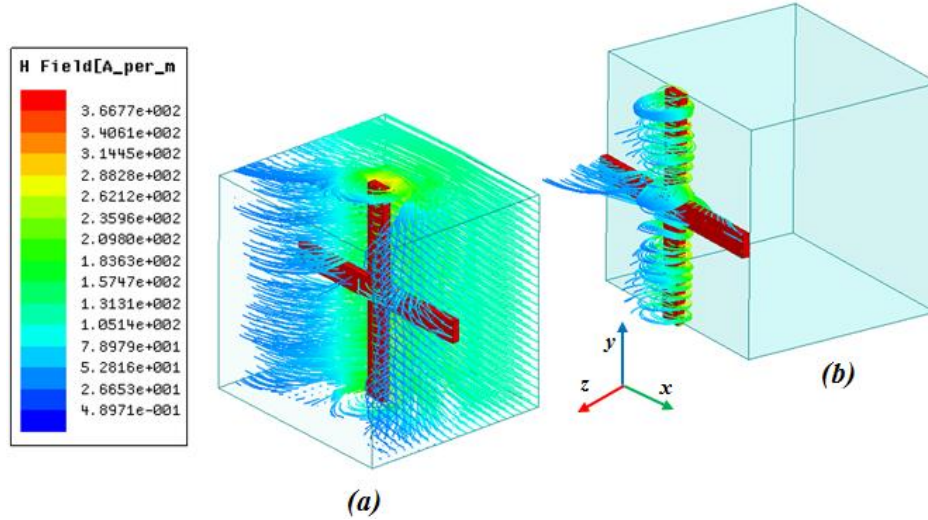


Figure 3-19 : Pattern of magnetic field lines around a conductor in (a) symmetric unit cell, (b) asymmetric unit cell

The refractive properties of the two structures are very similar. The refractive index is nearly the same for the two structures, as it can be seen in figure 3-16. From figure 3-11, it is clear that the magnitude of S_{11} and S_{22} for the two structures are very close. However, their phases are very different. The position of the cross within the dielectric slab affects the S_{ij} parameters phase, which lead to great difference between the backward and the forward impedances of the asymmetric structure as shown in figure 3-14 and figure 3-15.

The imaginary parts ε'' , μ'' and n'' account for losses in the structure caused by the electric and magnetic damping and the finite conductivity of conductors. Note that ε'' , μ'' and n'' must be positive in a passive medium due to the conservation of energy [145]. Due to their passive nature, we observe from figure 3-16, figure 3-17, and figure 3-18 that the constitutive parameters (permeability, permittivity and refractive index) of the symmetric and asymmetric unit cells have positive imaginary parts. Except for the symmetric unit cell which presents a negative permeability but the imaginary part it's very close to zero.

3.5.1. Power analysis

In this section we are concerned with the power properties of the wave propagation in the forward and backward directions on the symmetrical and asymmetrical structure.

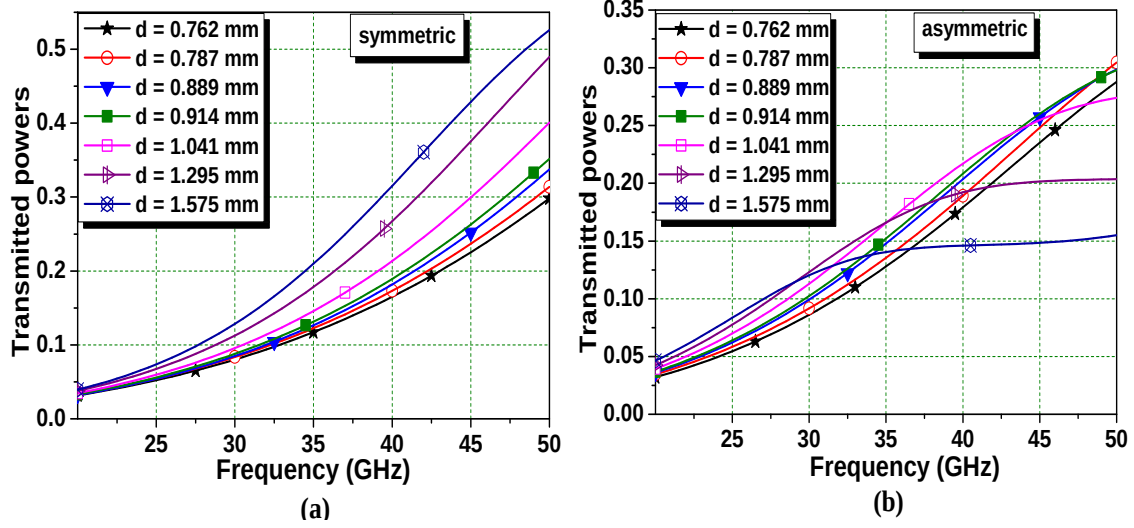


Figure 3-20 : Transmitted powers P_t (a) symmetrical, (b) asymmetrical, for different values of d

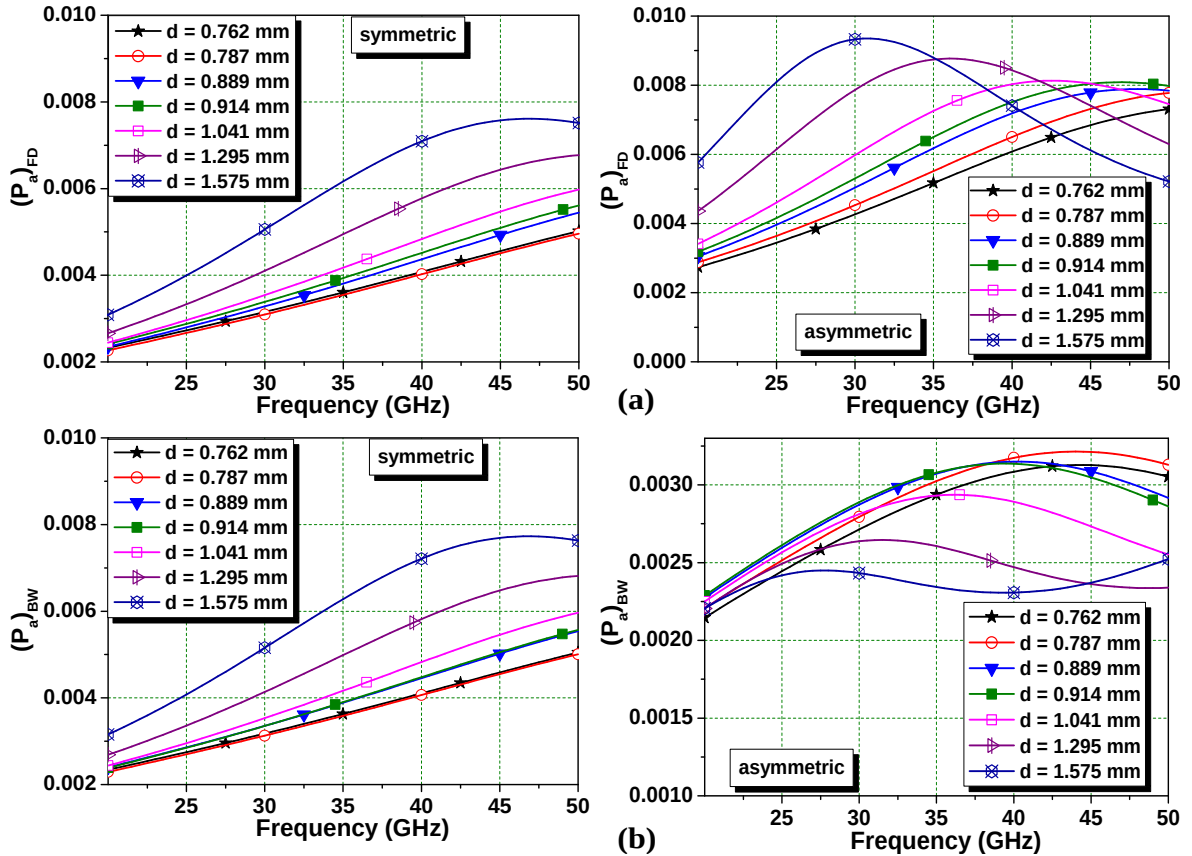


Figure 3-21 : (a) Absorbed powers in the forward $(p_a)_{FD}$, and (b) in the backward $(p_a)_{BW}$ for symmetrical and asymmetrical structure for different values of d

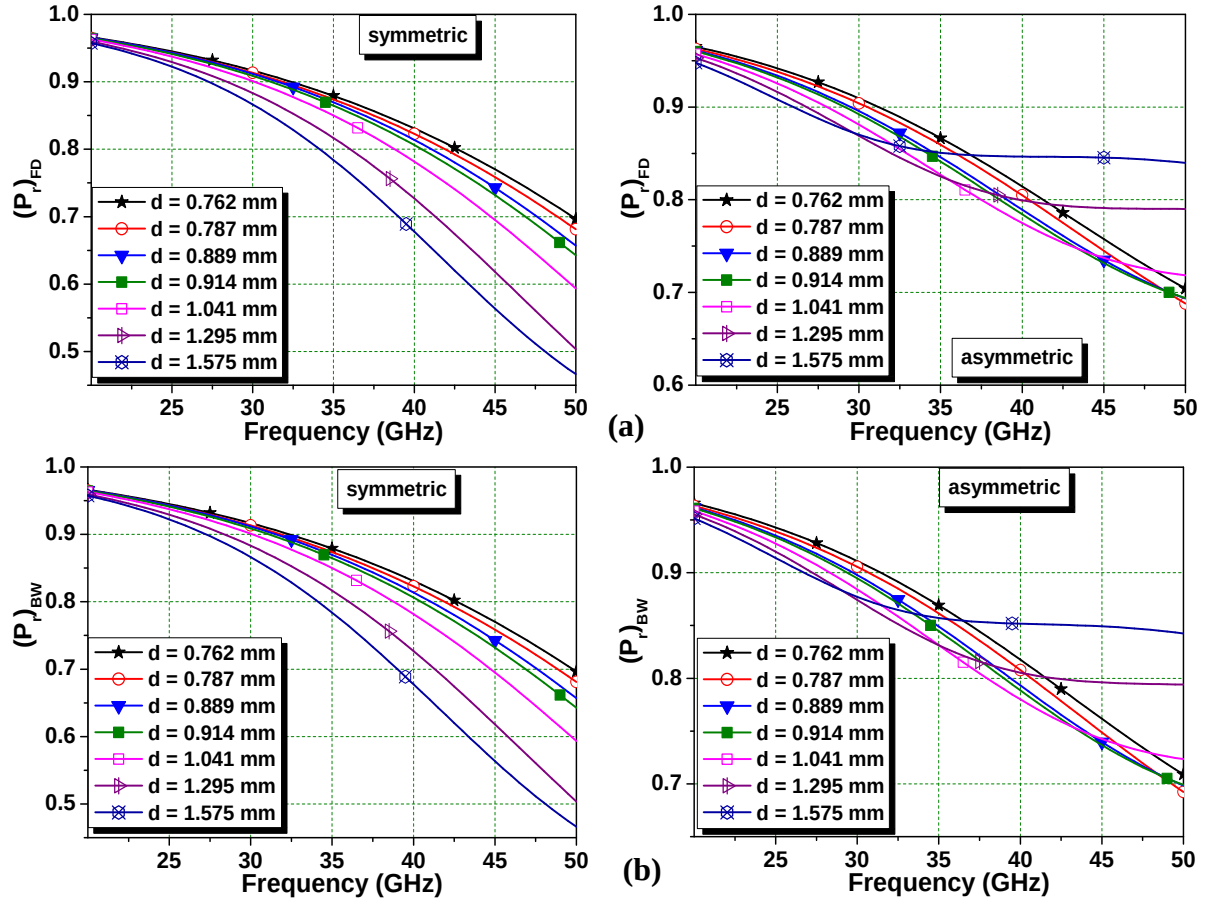


Figure 3-22 : (a) Reflected powers in the forward $(p_r)_{FD}$, and (b) in the backward $(p_r)_{BW}$ for symmetrical and asymmetrical structure for different values of d

3.5.2. Transition from the symmetric to the asymmetric unit cell

The passage from the symmetric to the asymmetric unit cell is shown in figure 3-24, which represent the different positions of the cross like conductor in the dielectric slab.

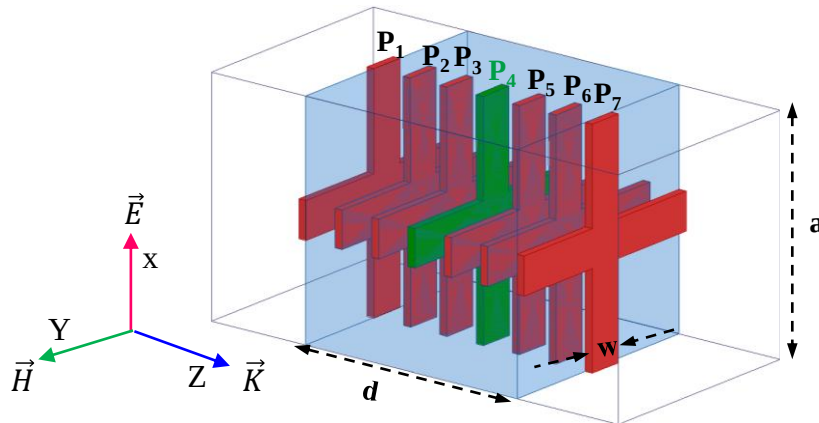


Figure 3-23 : Different positions of the conductor

Figure 3-24, shows the results for the magnetoelectric coupling parameter ζ when we shift the conductor from position P_4 to P_7 and from P_4 to P_1 .

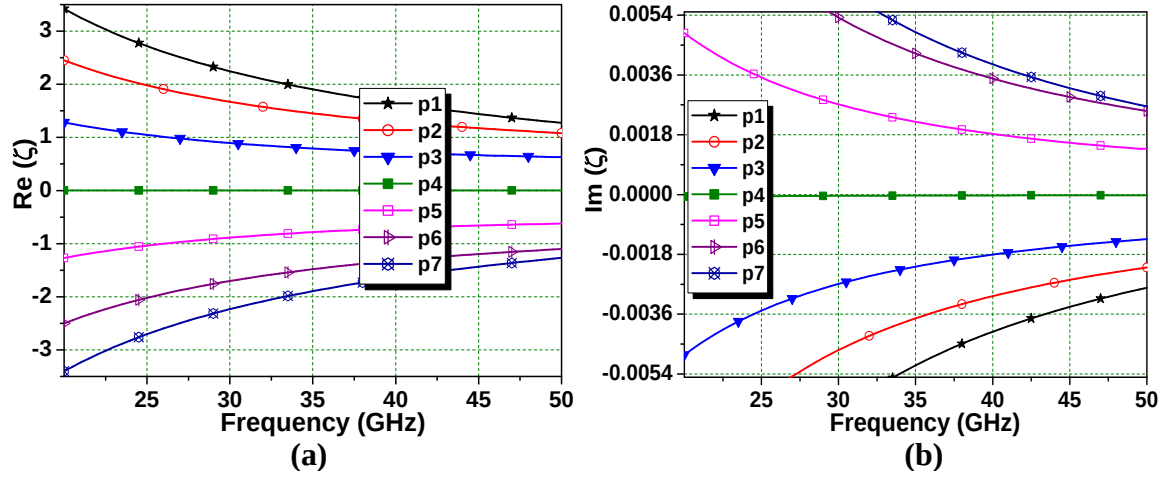


Figure 3-24 : Results for magnetoelectric coupling parameter ζ (a) real part, (b) imaginary part for different values of positions P

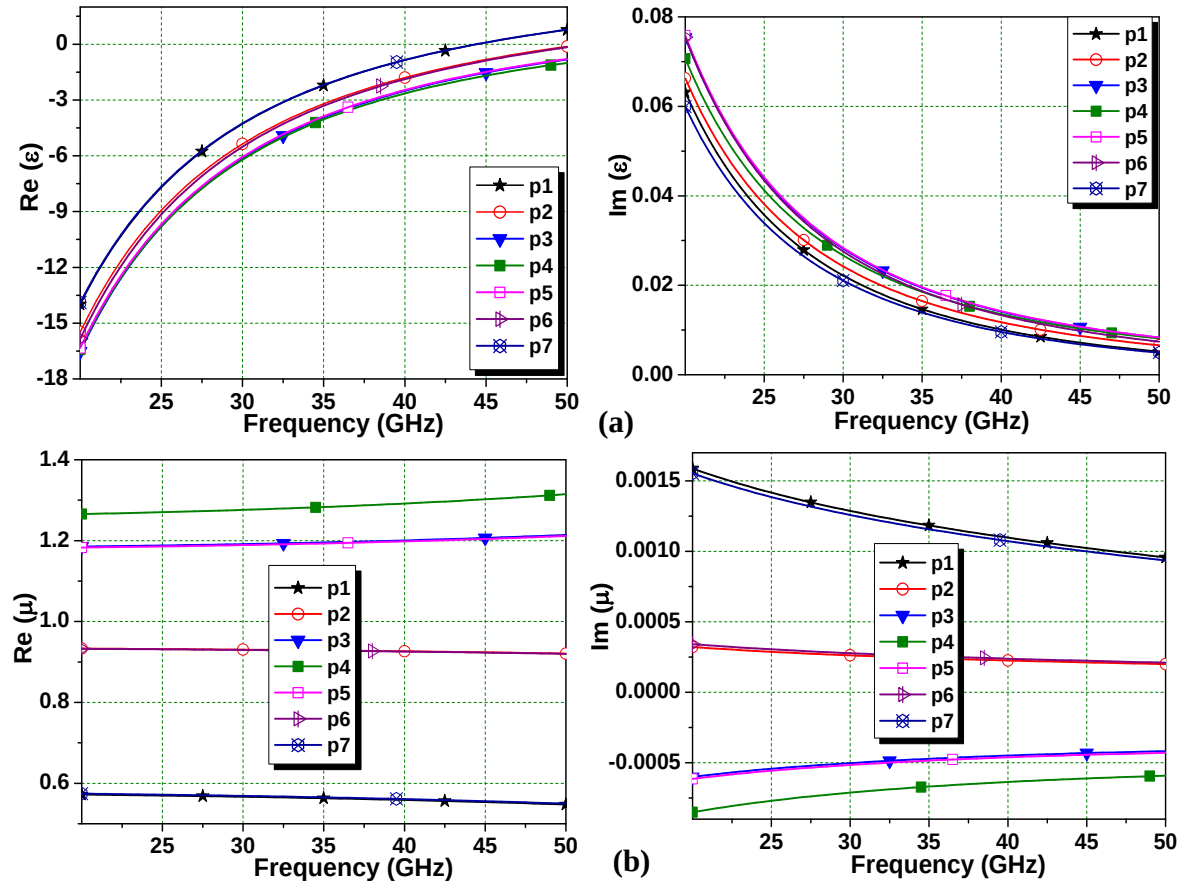


Figure 3-25 : (a) Effective permittivity ϵ and, (b) effective permeability μ for different values of positions P

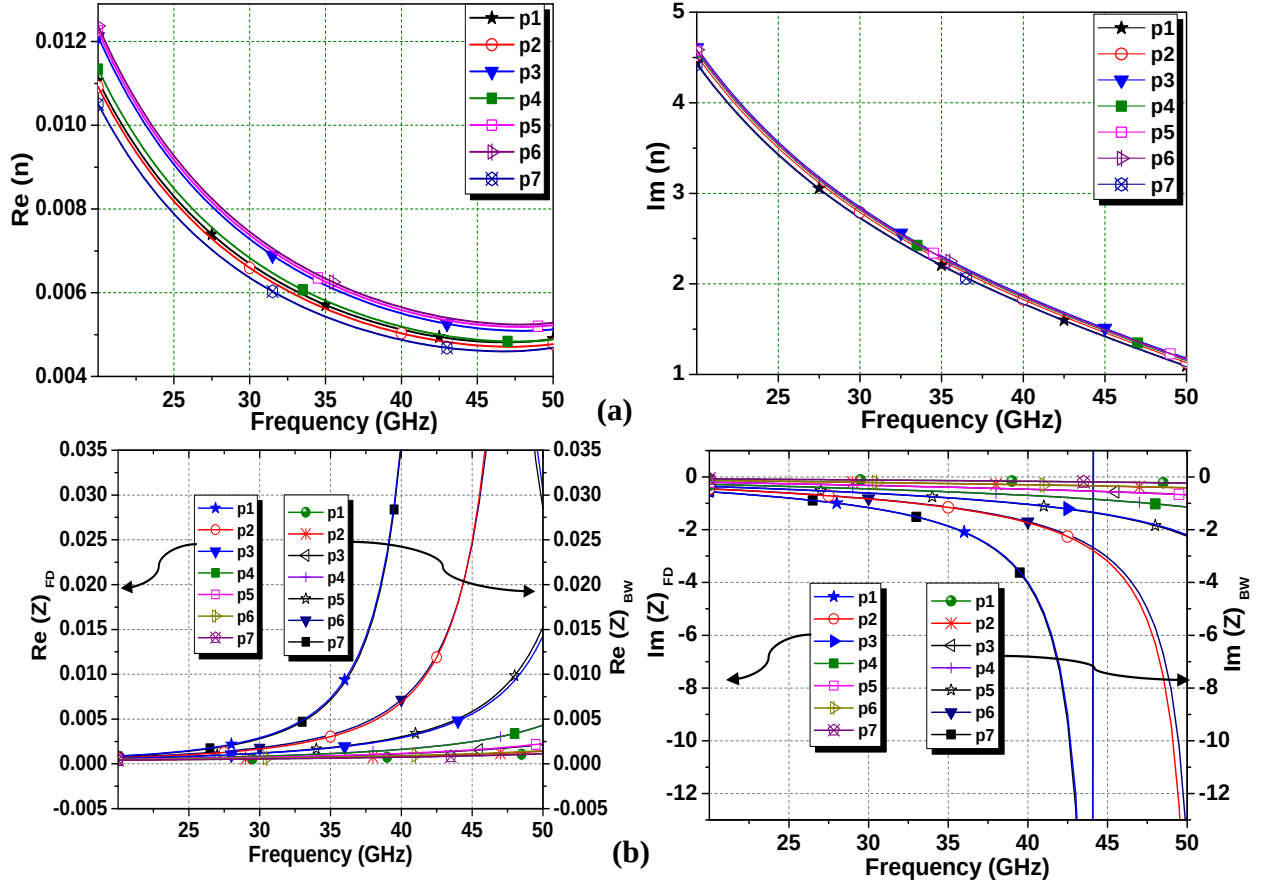


Figure 3-26 : (a) Refractive index (n) and, (b) Impedance (Z) for different values of positions P

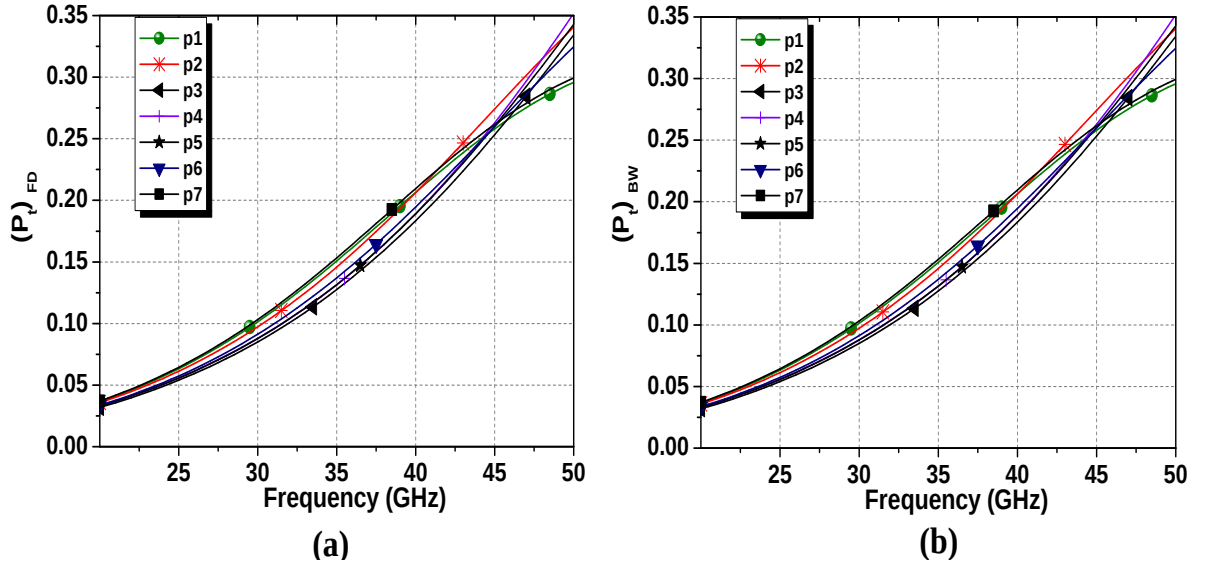


Figure 3-27 : Transmitted powers P_t (a) backward direction, (b) forward direction for different values of p

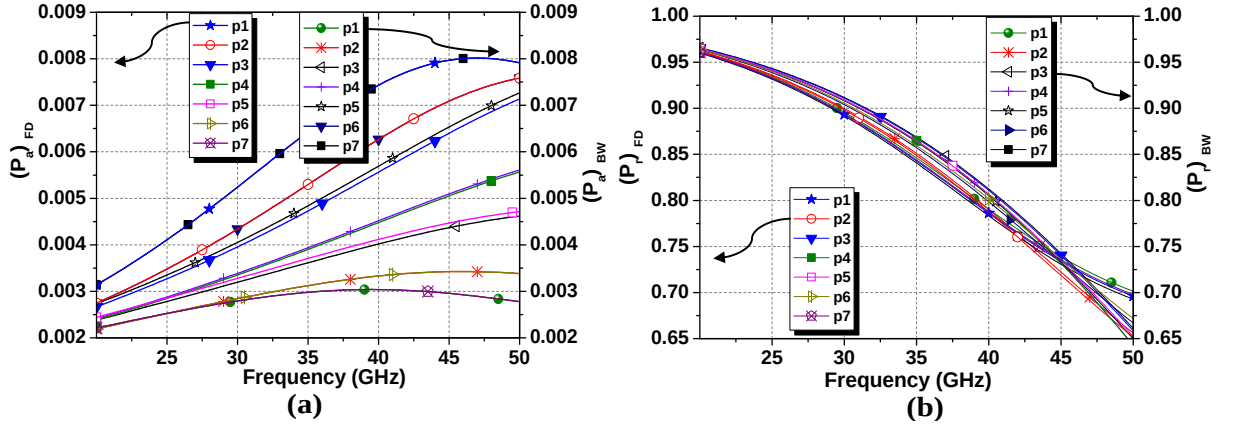


Figure 3-28 : (a) Absorbed powers forward $(P_a)_{FD}$ and backward $(P_a)_{BW}$, (b) Reflected powers forward $(P_r)_{FD}$ and backward $(P_r)_{BW}$ for different values of positions P

From figure 3-25-a in the real part of the permittivity effective, we note that the plasma frequency shifts to the lower frequencies range when the conductor cross position shifts away from the center (P_4). The shifting of the plasma frequency depends only on the distance between the conducting strip position and the center of the dielectric slab (P_4) it's independent from the shifting direction (backward or forward). Hence, for positions which are symmetric to the medium position P_4 , such as P_3 and P_5 , P_2 and P_6 , P_1 and P_7 , the unit cell has the same behavior. It presents the same plasma frequency and the forward and backward impedances are equal. For the central position P_4 , which corresponds the symmetric unit cell, the magnetoelectric coupling coefficient is null as it can be seen in figure 3-24.

As it can be seen in figure 3-24, that for $d = 1.041$ mm the metamaterial presents a plasma frequency close to 40 GHz and the refractive index in figure 3-26-a presents a flat profile close to zero corresponding to the desired ultra-refraction ($0 < n \ll 1$).

3.5.3. The parameters of the unit cell used

In this section we are concerned with the power properties of the wave propagation in the forward and backward directions on the new unit cell.

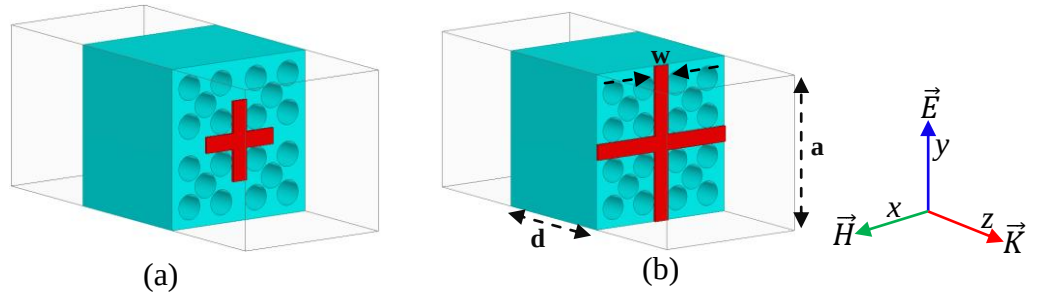


Figure 3-29 : Unit cell chosen, (a) disconnected type, (b) connected type, $d=0.914$ mm, $w=0.1$ mm, $a=1$ mm

The constitutive parameters (ϵ_{reff} and μ_{reff}) and the refraction index η of the metamaterial are obtained by the Fresnel inversion method using the S parameters of the unit cell.

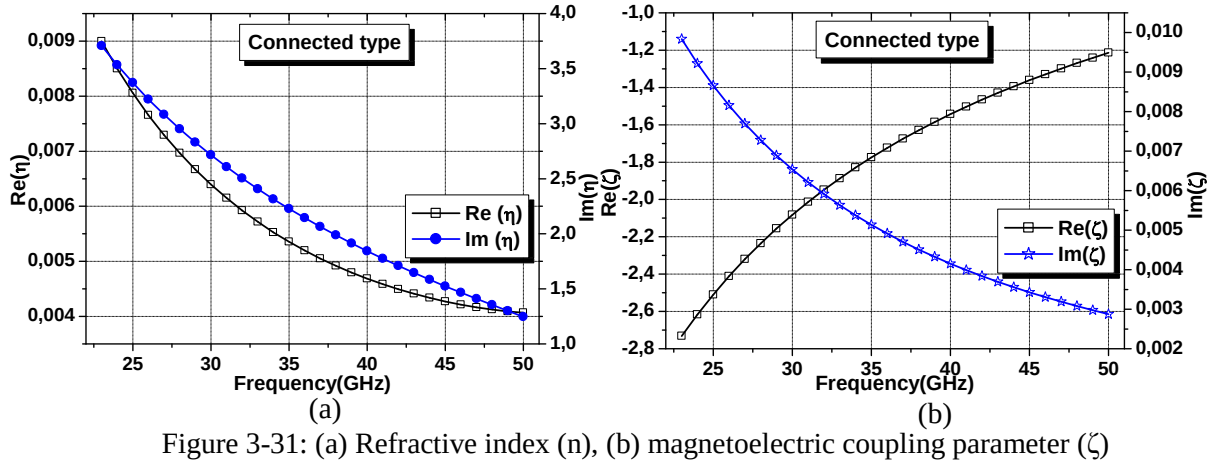
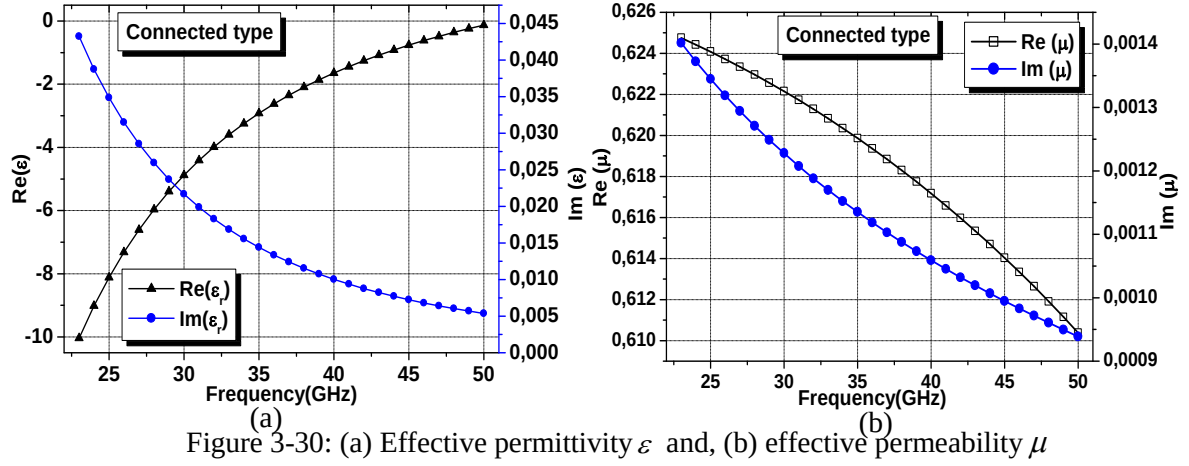
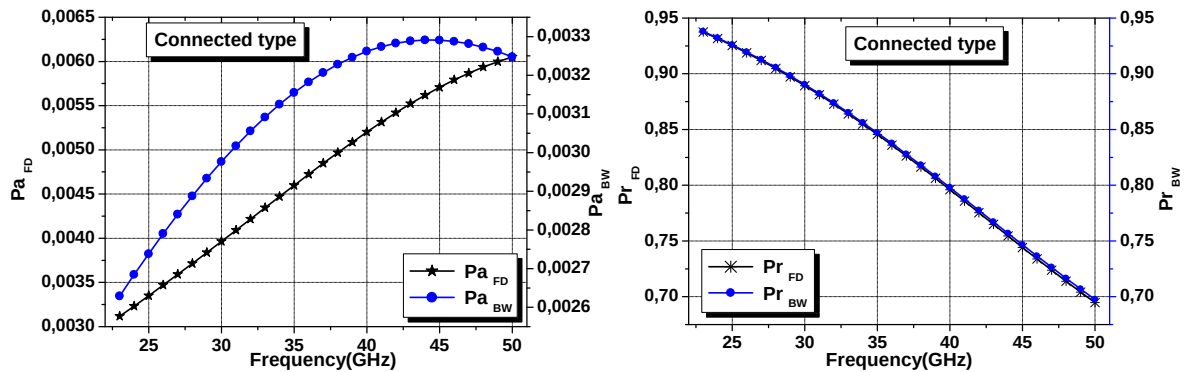


Figure 3-31 shows the real part and the imaginary part of the refractive index η and magneto-electric coupling coefficients ζ .



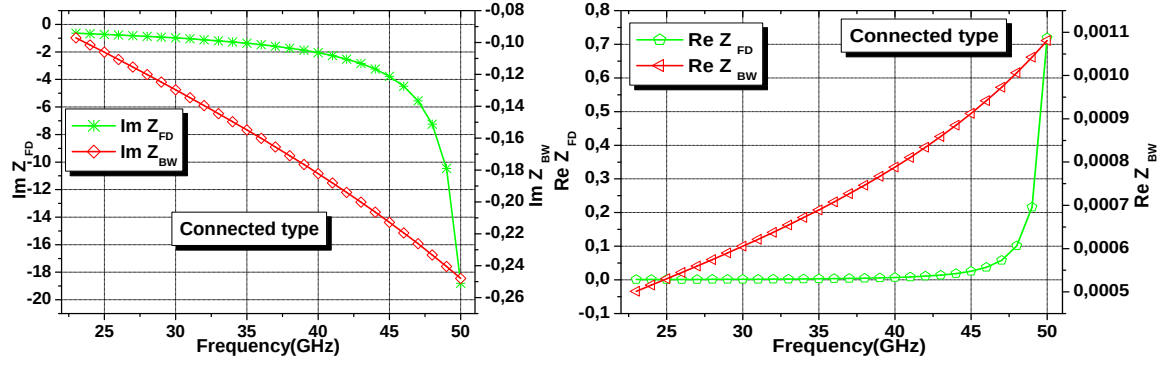


Figure 3-33: Impedance (Z) forward and backward

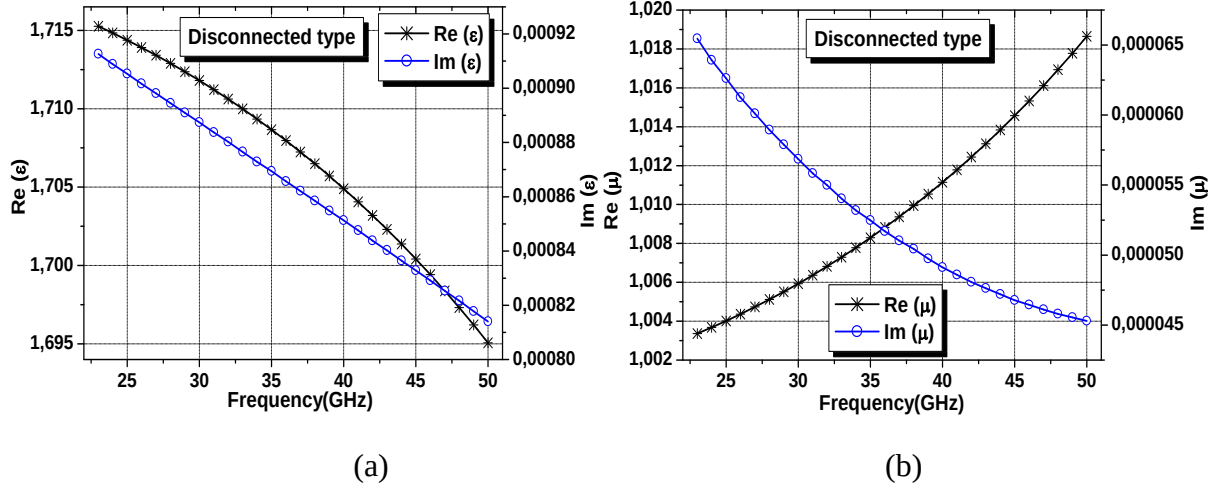


Figure 3-34:(a) Effective permittivity ε and, (b) effective permeability μ

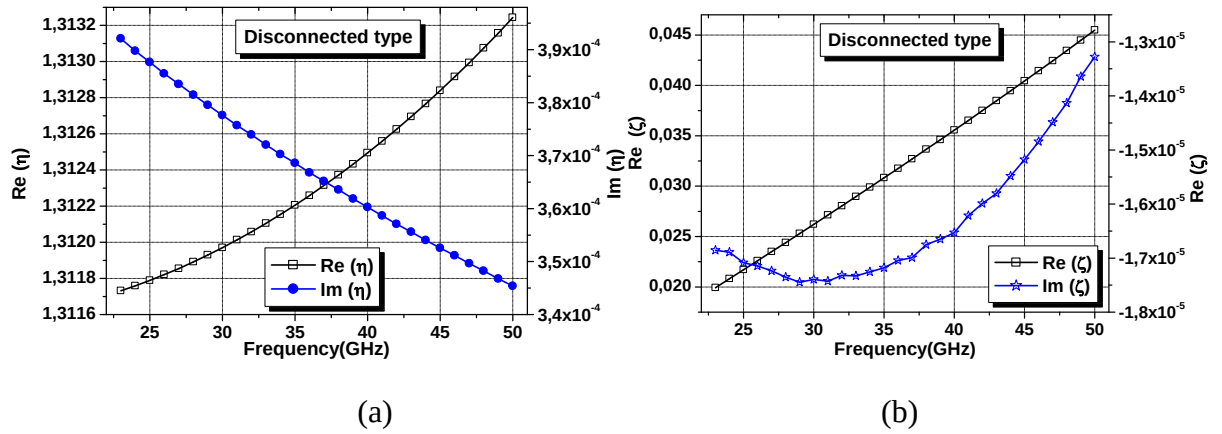


Figure 3-35:(a) Refractive index (n), (b) magnetoelectric coupling parameter (ζ)

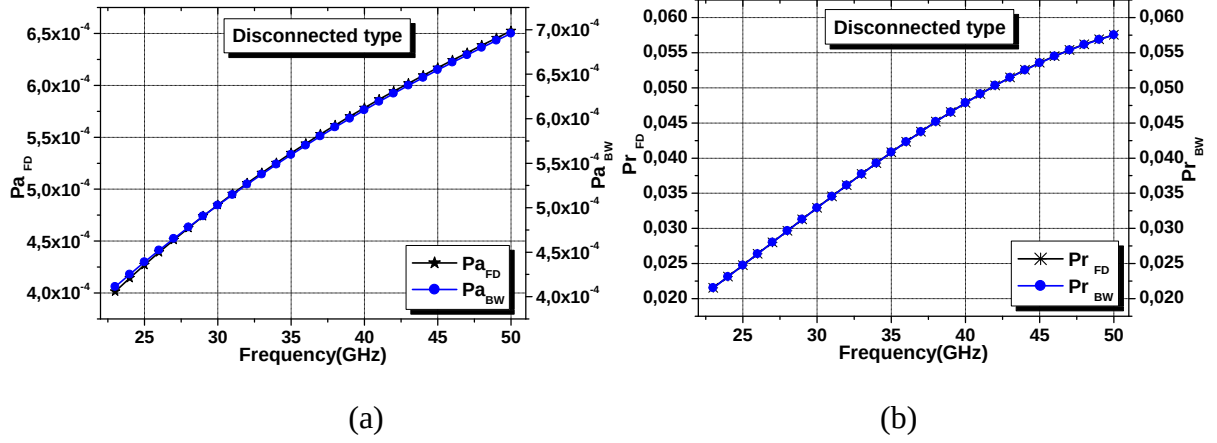


Figure 3-36: (a)-Absorbed powers forward (P_a)_{FD} and backward (P_a)_{BW}, (b)- Reflected powers forward (P_r)_{FD} and backward (P_r)_{BW}

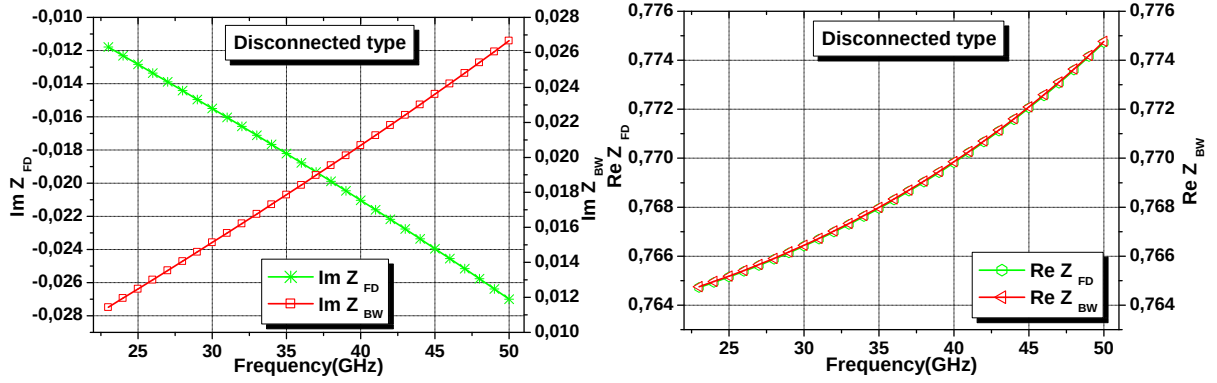


Figure 3-37: Impedance (Z) forward and backward

Figure 3-30 and figure 3-34 shows the effective permittivity ϵ and the effective permeability μ for our unit cell in the two cases obtained by a finite element method based commercial simulation code. It can be seen that the real part of the permittivity (ϵ') and the permeability (μ') and the refractive index (n) and the magneto-electric coupling (ζ) in the disconnected unit cell is greater than in the connected unit cell.

3.6. Conclusion

This chapter has argued that the homogenization theories are typically valid when the unit-cell size is much small compared to the wavelength λ .

There are two methods to extract the electromagnetic properties of metamaterials. The first one based on S -parameter which is suitable to be applied for both numerical simulation and experimental measurements. The second one is based on the electromagnetic field distribution within the unit cell (field cards) known as the field summation method (FSM). We have compared S - parameters extraction methods with different published litterateur and we found that generally there is a good agreement between each other.

4.1. Introduction

Within the huge number of new applications of metamaterials, one of the most productive ones is the implementation of multilayer FSS radome in the near-field region of a microstrip patch antenna.

A radome can protect the antenna against malfunctions or serious damages caused by debris or ice. Other than protecting radiating elements from their physical environment, another crucial important function is to reduce the Radar Cross Section (RCS) [146]. Frequency selective surface (FSS) could transmit the in-band wave and reflect the out-of-band power to other unimportant direction. The reflecting signal may also be detected by multistatic radar system, hence this approach can be classified as a fictitious RCS reduction technique [147]. Different from traditional FSS, an absorptive/transmissive FSS was proposed firstly in a patent [148]. According to the patent, the absorptive/transmissive structure could reduce the electromagnetic signature of the radome and antenna without disrupting the electromagnetic waves generated by the antenna, while absorbing the energy of electromagnetic wave at a particular frequency or range of frequencies. However, this patent only stated the working principle of the radome, and no any quantitative results were presented. Costa and Monorchio synthesized an absorptive/transmissive radome using designed resistive surface and metallic periodic surface [146]. Zhou et al. designed and fabricated another absorptive/transmissive radome using magnetic material. Its transmitting and absorbing properties were verified by simulation and experiments [149]. Liu et al. proposed and manufactured an absorptive/transmissive radome using resistors loaded double square loop (DSL) and convoluted slot [147]. Chen and Fu verified its performance and influence on antenna using a planar antenna [150]. They improved the radome by replacing the resistors loaded DSL metal layers and convoluted slots with resistors loaded incurved square loop (ISL) and miniaturized FSS, respectively [151]. They also designed an absorptive frequency selective surface with good transmission at high frequency [152][153]. All the above structures had only one passband. But in modern society, many equipments, such as mobile phone, have two or more working frequencies. In other words, two or more passbands are needed in order to guarantee their normal operation, cause of this in the following section will deal with an agile antenna/radome system consisting in of four layers of metasurfaces provide 11 operating frequencies by an outside command, in addition to that, the antenna/radome system gain is improved and the radiation pattern half-power beamwidth is reduces in both E- and H-planes.

4.2. Radome characteristics and types

The radome cover of a radar sensor is the backbone of the design as it can have influence on sensitivity, radiation pattern and immunity to vibrations. The motive of an efficient radome design is to reduce reflections at the surface of the cover and transmit waves losslessly [154]. Thus, the cover material can be thought of as a lens which either focuses or disperses waves uniformly. For this purpose a constant and uniform thickness of the material is required and it must also have a good surface smoothness. The radome surface should preferably be covered with a hydrophobic substance like a variant of silicon or Teflon.

It was from the time of World War II, 1940's that radome become such a boiling topic in the microwave community. Materials such as plywood, polyseterene fiber, plexi glass etc. were employed and it became the state of the art. Significant contributions were made during that period by Cady [155][156] who described the electrical design of normal and streamlined radomes and their installation, together with the theory of reflection and transmission of electromagnetic waves through dielectric materials. Hansen, [156] focused on large ground radomes, their environmental, structural and design problems, as well as on methods of construction, including metal spaced frame radomes covered with thin dielectric materials, etc. Nowadays Frequency Selective Surfaces, Transformation optics, Rasorbers, etc are used to improve the radome performances. As antennas operating frequency increases, the radome materials also changed and so did their construction, installation, tooling and transportation.

Classically the radome constructions can be classified as follows:

- Air pressure supported dielectric structure
- Self-supporting dielectric structure
- Widely spaced metallic or non-metallic frames supporting dielectric windows
- Metal surfaces covered with dielectric covered slotted or other shaped windows.

One can also distinguish them based on their dielectric wall construction: [157]

- a. Style A: Radomes are half wave wall solid radomes (see figure 4-1),
- b. Style B: Radomes are thin-wall monolithic or non-monolithic structures with a wall thickness equal to or less than 0.1λ wavelengths at the highest operating frequency, as shown in figure 4-1,



Figure 4-1 : Solid wall [156]

c. Style C: Radomes are also called A-sandwich multilayer walls. These consist of three layers: two high-density skins and a low-density core material. The dielectric constant of the skins is greater than the dielectric constant of the core material.

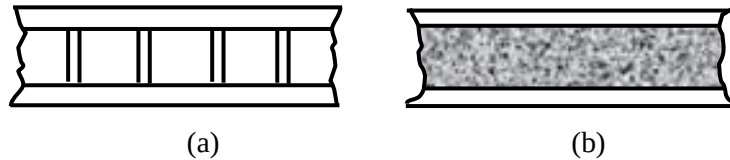


Figure 4-2 : (a) Low dielectric constant with honey comb core, (b) Low dielectric constant with foam core

d. Style D: Radomes (C-Sandwich) are multilayer walled structures having five or more dielectric layers. These have an odd number of high-density skin layers and an even number of low-density core layers. The dielectric constants of the skins are greater than the dielectric constants of the core materials. As the number of layers is increased, the broadband frequency performance is improved.

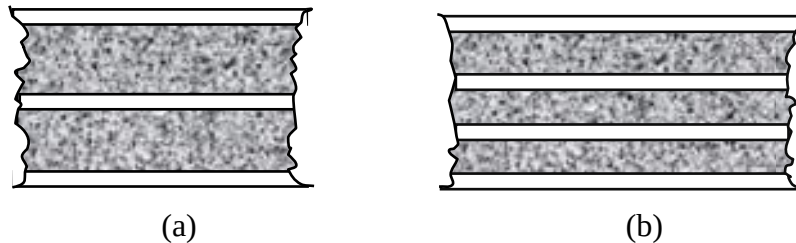


Figure 4-3: (a) Double sandwich, (b) More than five layers

e. Style E: Radomes are defined here in as all possible radome wall construction not fitting into styles A through D. These include the B-sandwich and others. The B-sandwich is similar to the A-sandwich except it consists of two low-density skins and a high-density core material. The dielectric constant of the skins is less than the dielectric constant of the core material.

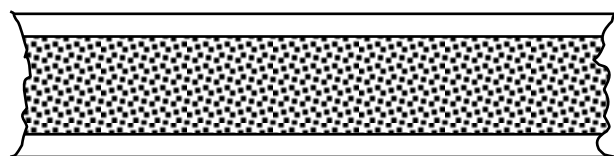


Figure 4-4 : High dielectric constant core

4.3. Radome interaction with the antenna

The antennas employed in the radar application have very strict requirements and hence it needs to be thoroughly checked as the radome can alter their performances [157] [156] in the following ways:

a. Transmission loss: This results from the reflection against the radome walls, reflection and diffraction around the corners and edges as well as polarization shift. The remainder of the loss occurs from dissipation within the dielectric layers. The material loss tangent or $\tan\delta$ is a measure of these dissipative losses.

b. Pattern and polarization distortions: This result in the change of main beam direction and shape, increase in side lobe level and cross polarization components also increases.

c. Boresight Error (BSE): BSE is a bending of the angle of arrival of a received signal relative to its actual angle of arrival or line of sight. This phenomenon occurs due to distortions of the electromagnetic wave-front as it propagates through a dielectric radome wall. BSE depends on frequency, polarization, and antenna orientation. In production, the radome manufacturing tolerance affects the BSES characteristics of the radome.

d. Registration error: Registration error is the difference in uplink frequency and downlink frequency BSE in the case of radome-enclosed SATCOM antennas. It is important that this parameter be sufficiently small to assure that both the transmit beam and receive beam peaks are coaligned on the same satellite.

e. Antenna Voltage Standing Wave Ratio (VSWR): The antenna VSWR can greatly increase from RF power reflected from the inner radome wall surface. In spite of these disadvantages, on the other hand one gets a large number of operational, structural as well as economic advantages when the radome is mounted. To mention a few, the antenna now is protected from the harsh environment and does not need any special maintenance. Although the radome, will wither with time. The over-all system is much more accurate and reliable as it is not susceptible to any changes.

4.4. Basic radome requirements

Radomes have to satisfy aerodynamic, structural and electromagnetic requirements, as well as environmental and installation problems, such as occur when they are installed on ship, submarine, ground, aircraft, missile, vehicle etc. The electromagnetic requirements could call for certain limits to transmission loss, aberration, polarization, pattern distortion,

reflection levels etc., while the structural design must satisfy the shape, size and load requirements under all conditions with an adequate safety factor.

4.4.1. Environmental requirements [157]

Typical environmental conditions that may be called for are:

- a. Temperature: Ground radomes -56°C to $+80^{\circ}\text{C}$,
- b. Relative humidity: 100%,
- c. Pressure: Ground radomes may have to stand pressures resulting from wind speeds of up to 240km/h,
- d. Loads: Ice, hail and snow loads have to be allowed for, together with possible impact loads from birds, stones etc,
- e. Rain: On ground radomes, rain adhesion should be minimized,
- f. Vibration, shock and acceleration: Ground radomes should withstand seismic loads, e.g. up to 1g, depending upon site,
- g. Fire, lightning and radiation hazards: Materials should be fire retardant, and the radome should withstand salt, dust, insects, fungi, fuels, etc.

4.5. Advances in radome material

The key to obtain an optimized radome performance is evaluating the electrical properties of the material at possible wavelengths. The influential parameters are the material dielectric constant and the dissipation loss factor at the frequency of interest.[155]

The radome wall constructions are typically of either the monolithic type or the sandwich design type by a multilayer FSS type metamaterial. Since the focus in this thesis is mostly on multilayer FSS radomes, a detailed outline and some commonly used materials shall be presented.

The radomes often incorporate fiber reinforcement in order to enhance the mechanical properties. Thus the relative dielectric constant that we obtain from the mixture [155] is as follows:

$$\epsilon_m = \frac{V_R \log \epsilon_R + V_F \log \epsilon_F}{V_R + V_F} \quad (4-1)$$

Where,

ϵ_m = Relative dielectric constant of mixture

ϵ_R = Relative dielectric constant of the resin

V_R = Volume of dielectric resin

V_F = Volume of reinforcement fibers

Equation (4-1) assumes that the mixture is uniform and isotropic.

Table 4-1 lists down few materials which are commonly used in the X band. But as we are dealing in the mm-wave band, radomes are manufactured without reinforcement. These are mostly termed as the organic materials. There is also another type of material called the inorganic or ceramic material. The advantage that inorganic materials have is that they can last at very high temperatures such as 500°C. These materials are suitable for high velocity applications such as missile systems. To name a few, aluminium oxide, Rayceram, Pyroceram, etc.

After 1999, when Rodger M. Wasler formally defined metamaterials, there was an advancement to study antenna radome metamaterials. Research was done using multilayered structures composed of two materials which together had a refractive index as close to air or below (mention reference number Liu et al). They can be used to compensate BSE, improve gain, directivity, etc. but yes they come at the price of a narrower beam-width.

Another type of inclusions on the radome walls are commonly known as Frequency Selective Surfaces (FSS) [158]. This layer can be arranged inside, outside or even embedded within the radome wall. They are usually a periodic array of etched antenna elements that can be used to either reflect or transmit electromagnetic waves with a frequency discretion. A FSS can take many forms and strips, slots, cross, square loops, omega units and Jerusalem crosses have all been used in array configuration on the radome surfaces.

Table 4-1 : Some characteristics of radome materials.

Material	Permittivity (ϵ)	Loss tangent (δ)
Polycarbonate	2.9	0.005
E-glass	6.06	0.004
Epoxy	4.4	0.016
Nylon	3.03 – 3.21	0.014 – 0.02
Teflon	2.1	0.0003
Polystyrene	2.55	0.0004
Rogers	2.2	1

4.6. Microstrip antenna characteristics

The radiating element, of the antenna/radome system, is a microstrip patch antenna, shown in Figure 4-5, operating at 43.5 GHz.

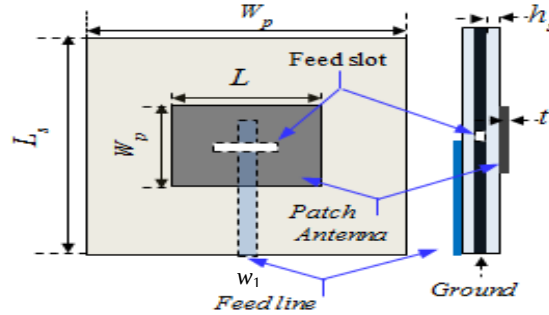


Figure 4-5 : Antenna slot feeding details ; $l = 0.5$, $L_s = 25$, $W_s = 20$, $h_s = 0.05$, $L_p = 0.257$, $W_p = 0.184$, $w_1 = 8$ [units = mm]

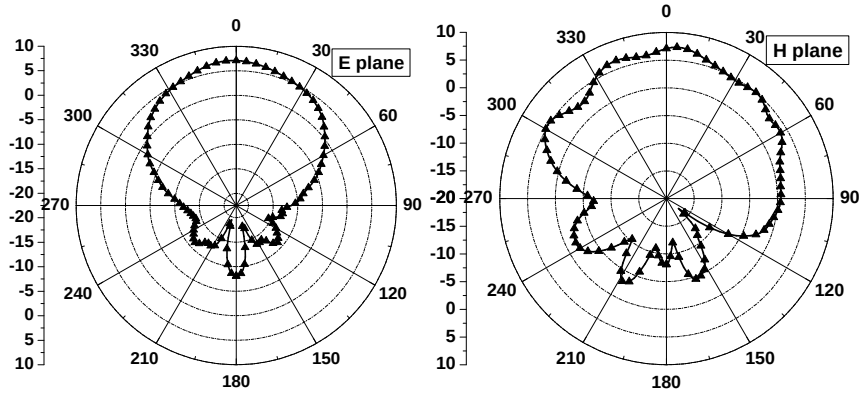


Figure 4-6 : The radiation patterns of the patch antenna, (left) E-plane, (right) H-plane

The radiation pattern, in the -E and -H plane, of the proposed antenna is shown in Figure 4-6 .It is clear from the plots that the antenna is directional and the maximum realized gain obtained is 7.3 dB. The main lobe beam-width in both E and H plane are 66 ° and 80°, respectively.

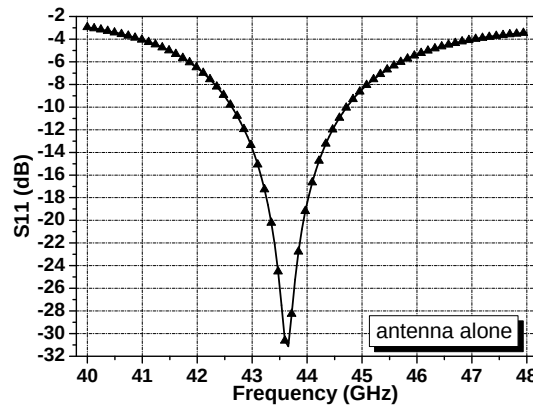


Figure 4-7 : The reflection coefficient of the patch antenna

Figure 4-7, shows the simulated return loss of the antenna alone. The S_{11} holds the importance parameter in any antenna design as it depicts how well the signal could be transmitted from the antenna.

4.7. Antenna/radome design

In this section we use a new FSS metamaterial surface as a radome with single layer superstrate suspended above a microstrip patch antenna (MPA), operating at 43.5 GHz, for the gain enhancement. The single metamaterial layer superstrate is composed of a periodic arrangement of FSS unit-cells, Figure 4-8 shows the antenna/radome system with one layer.

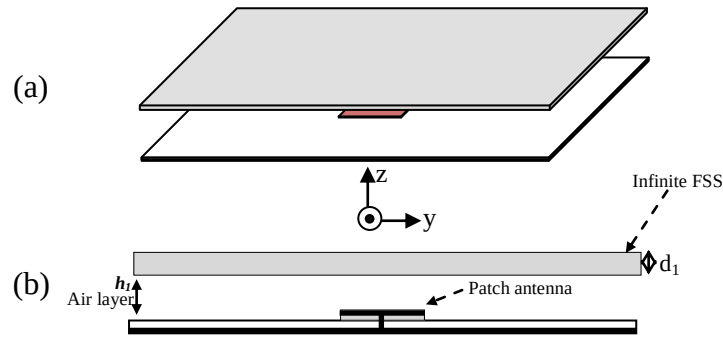


Figure 4-8 : Configuration of the proposed patch antenna covered with metamaterial FSS radome, $h_1=3.2\text{mm}$, $d_1=1.24\text{mm}$ (a) perspective view and (b) side view

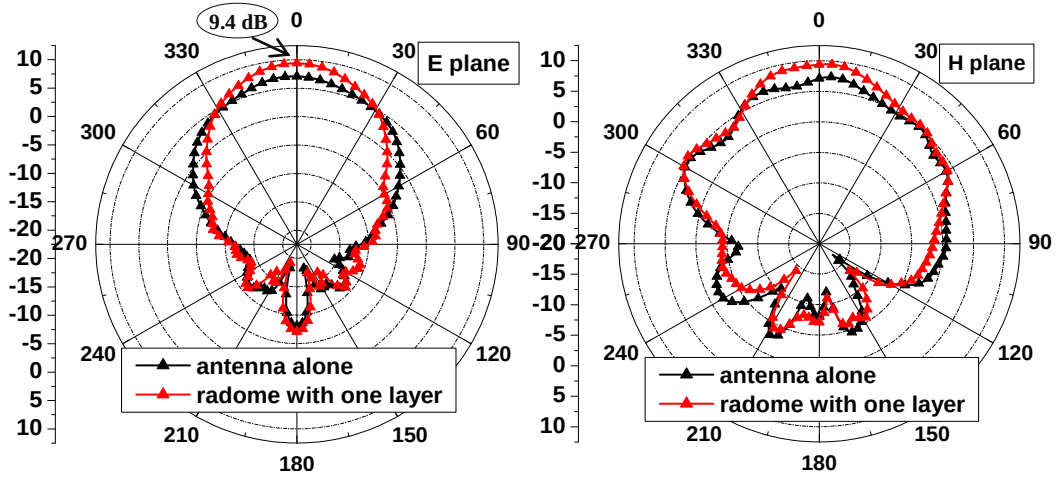


Figure 4-9 : Comparison of the radiation patterns of the proposed patch antenna and the proposed radome with one layer, (left) E-plane, (right) H-plane

It can be observed from Figure 4-9 that the FSS radome increases considerably the antenna/ radome system gain, and enhances the directivity, in both E- and H-planes. The maximum realized gain obtained in this case is 9.4 dB. The main lobe beam-width in both E and H plane are 48° and 52° , respectively.

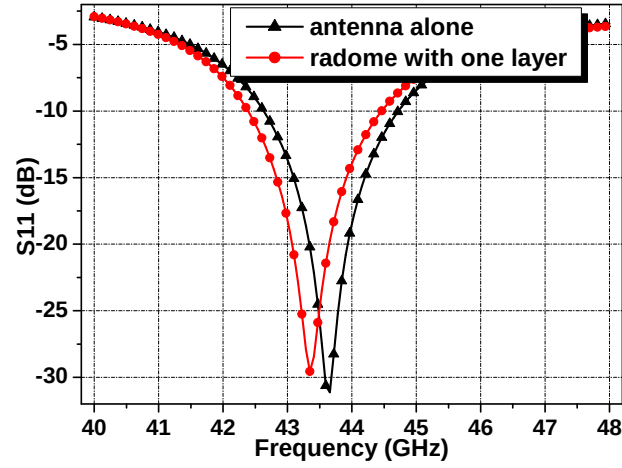


Figure 4-10 : Comparison of the reflection coefficient of the patch antenna alone and the proposed radome with one layer

It is observed from Figure 4-10 that the resonance frequency of the patch antenna is slightly shifted to lower frequency when it is loaded with the FSS radome.

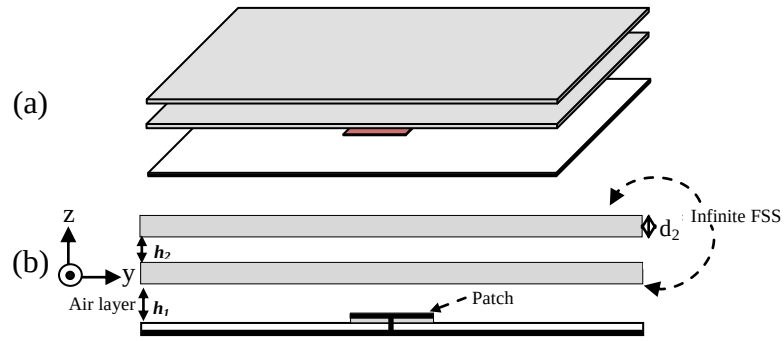


Figure 4-11 : Configuration of the proposed patch antenna covered with metamaterial FSS radome, $h_2=1.96$ mm, $d_2=1.6$ mm, (a) perspective view and (b) side view

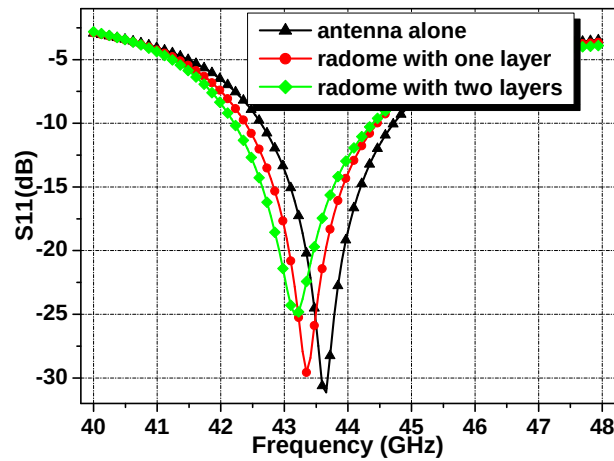


Figure 4-12 : Comparison of the reflection coefficient of the patch antenna alone and the proposed radome with two layers.

From the figure 4-12, it's can be seen that the resonance frequency of the patch antenna is slightly shifted to lower frequency region due to the FSS radome.

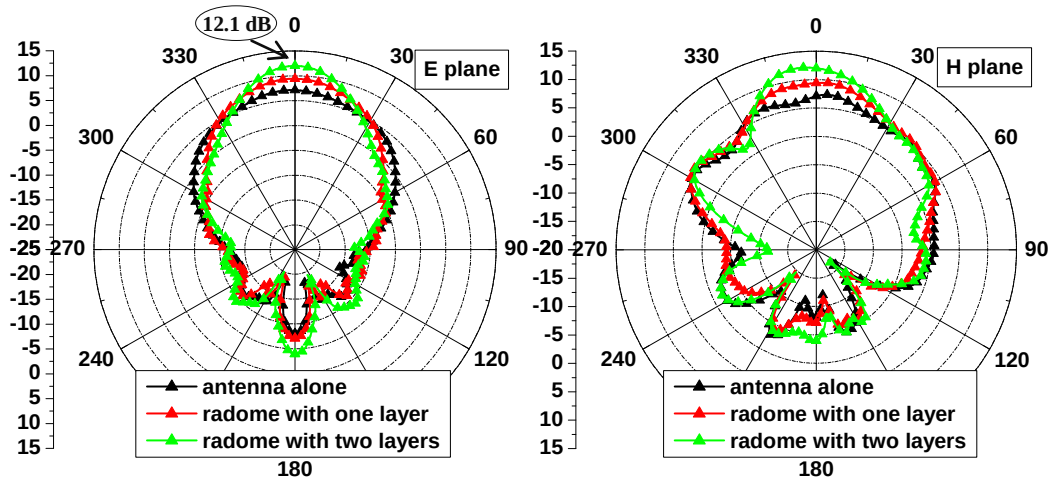


Figure 4-13 : Comparison of the radiation patterns of the patch antenna alone and the proposed radome with one and two layers, (left) E-plane, (right) H-plane

It can be seen from the figure 4-13 that loaded a patch antenna by two FSS radome layers can:

- Increases the maximum realized gain from 7.3 dB to 12.1 dB.
- Reduces the main lobe beam-width in both E-and H-plane from 66° and 80° to 32° and 38° , respectively.

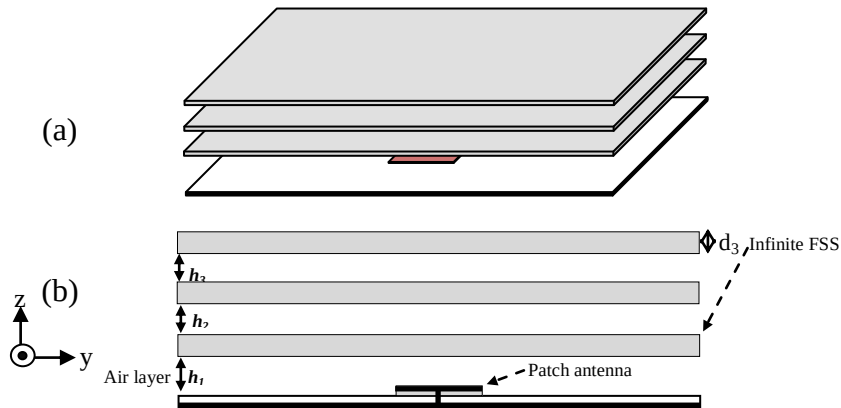


Figure 4-14 : Configuration of the proposed patch antenna covered with metamaterial FSS radome, $h_3=1.6$ mm, $d_3=1.4$ mm, (a) perspective view, (b) side view

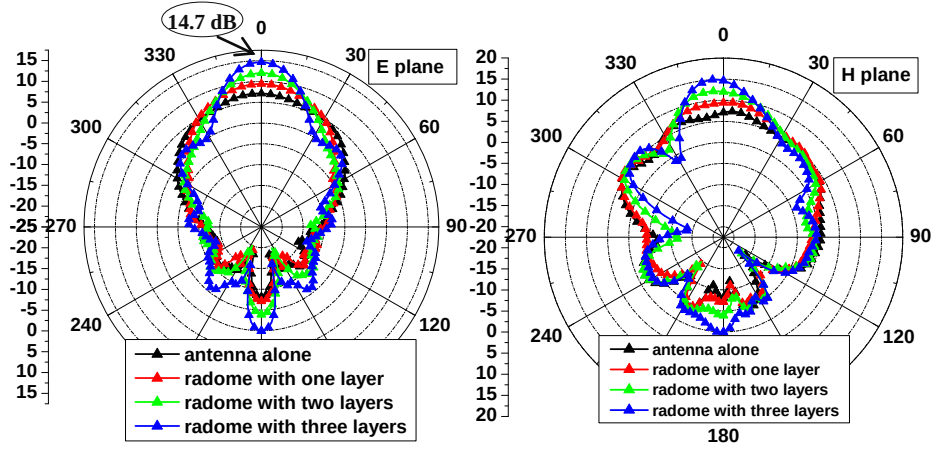


Figure 4-15 : Comparison of the radiation patterns of the patch antenna alone and the proposed radome with one and two and three layers, (left) E-plane, (right) H-plane

It's clear from the figure 4-15, that the maximum realized gain obtained in this case is 14.7 dB. The main lobe beam-width in both E and H plane are 24° and 26° , respectively.

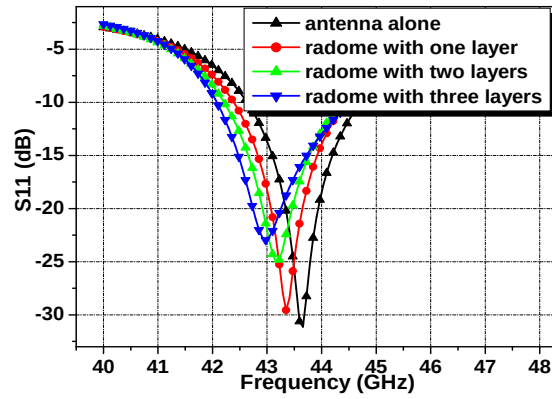


Figure 4-16 : Comparison of the reflection coefficient of the patch antenna alone and the proposed radome with three layers

As shown in Figure 4-16 the resonance frequency of the antenna is slightly shifted to lower frequency region.

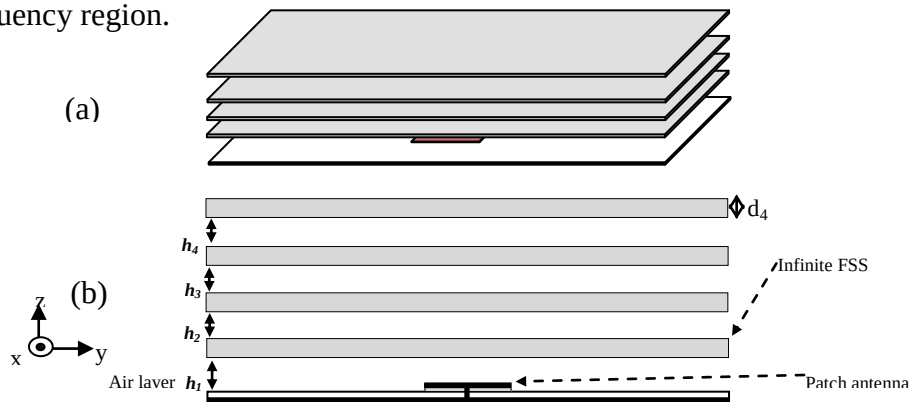


Figure 4-17 : Configuration of the proposed patch antenna covered with metamaterial FSS radome, $h_4=1.8$ mm, $d_4=1.8$ mm, (a) perspective view, (b) side view

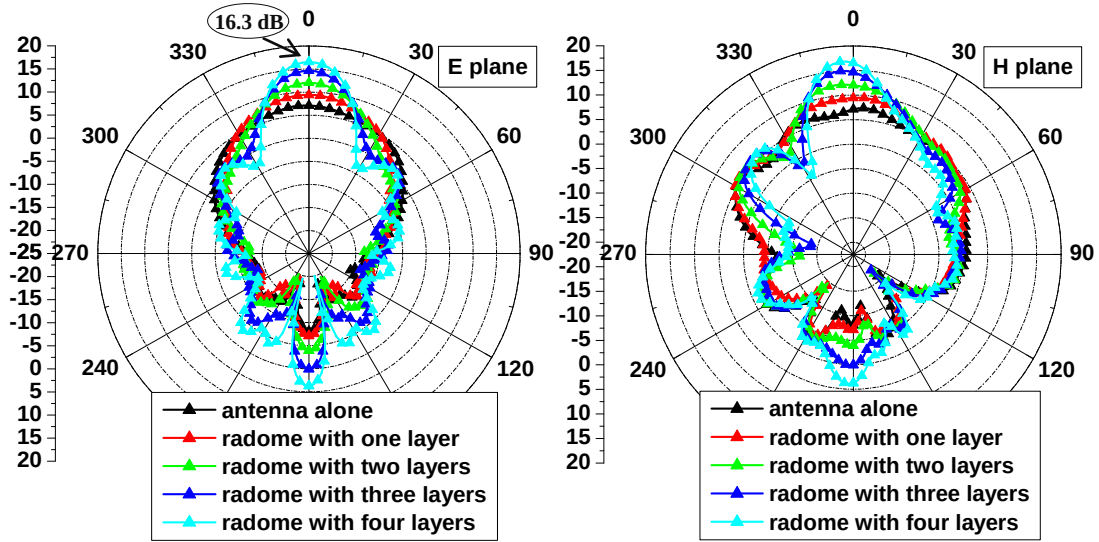


Figure 4-18 : Comparison of the radiation patterns of the patch antenna alone and the proposed radome with one, two, three and four layers, (left) E-plane, (right) H-plane

From the figure 4-18 we can conclude that the maximum realized gain obtained in this case is 16.3 dB and the main lobe beam-width in both E and H plane are 20° , it is clear from the plot that loading a microstrip antenna with a metamaterial FSS radome can focus the radiation beam width and thereby enhances the antenna directivity. The half-power beam width of the antennas, at the operating frequency in E-and H-plane respectively, are observed to be 66° and 80° for the antenna alone, 20° in both E-and H-plane for the antenna with a FSS radome composed by four metamaterial layers.

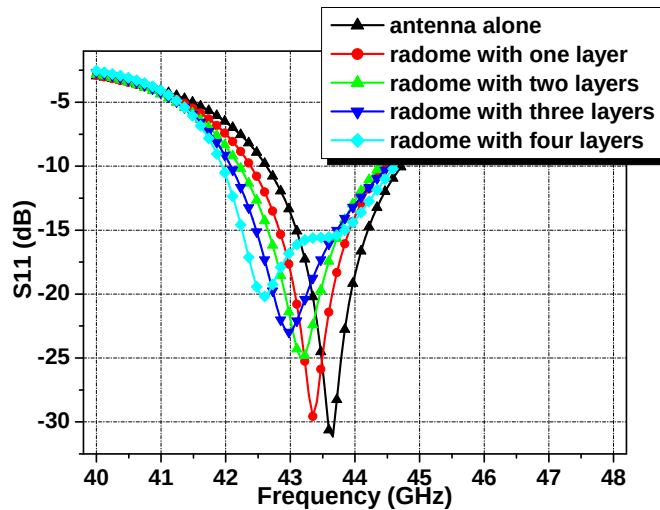


Figure 4-19 : Comparison of the reflection coefficient of the patch antenna alone and the proposed radome with four layers

From figure 4-19, it can be seen that, the resonance frequency of the patch antenna is slightly shifted to lower frequency region, by about 1 %, when loaded with the FSS radome layers.

4.8. Frequency agile antenna/radome system

The agile antenna/radome system is composed by four layers and it has two configurations. The two configurations of the agile antenna/radome system are shown in figure 4-20 and figure 4-22 respectively, the FSS radome layers are composed by a connected cross type metamaterial unit cell or disconnected cross type metamaterial unit cell, (see chapter 3, figure 3-29), for each configuration we have four combinations, In the first four cases we command each disconnected FSS layer to be a connected FSS layer using switches, we start with the first position then in each case the connected FSS layer change the position to the upper position.

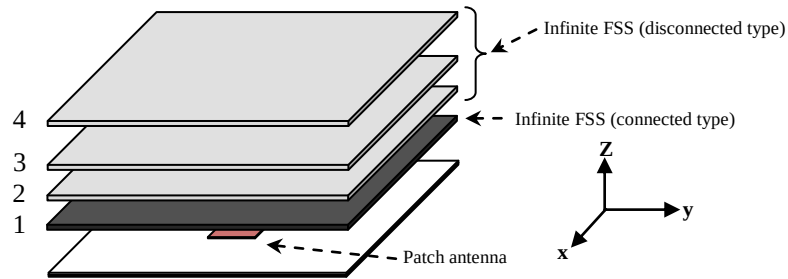


Figure 4-20 : Agile antenna/radome system (first configuration)

The simulation results of the resonance frequency of the proposed agile antenna/radome system for the first four cases are shown in figure 4-21. It can be seen that the resonance frequency shift to lower frequency band when a FSS layer made by a disconnected cross type metamaterial unit cell commanded to be a connected FSS layer in different positions.

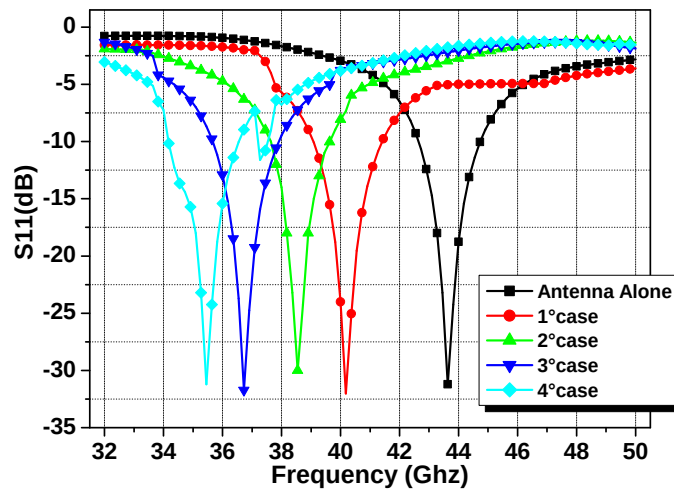


Figure 4-21 : Resonance frequency of each possible combination

Figure 4-22 shows the last four combinations of two connected FSS layers, firstly we command the two first disconnected FSS layer to be a connected FSS, then the others.

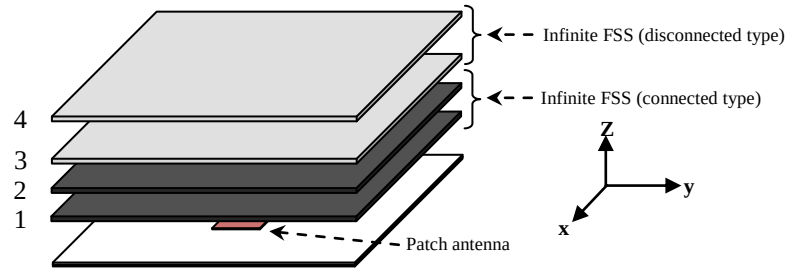


Figure 4-22 : Agile antenna/radome system (second configuration)

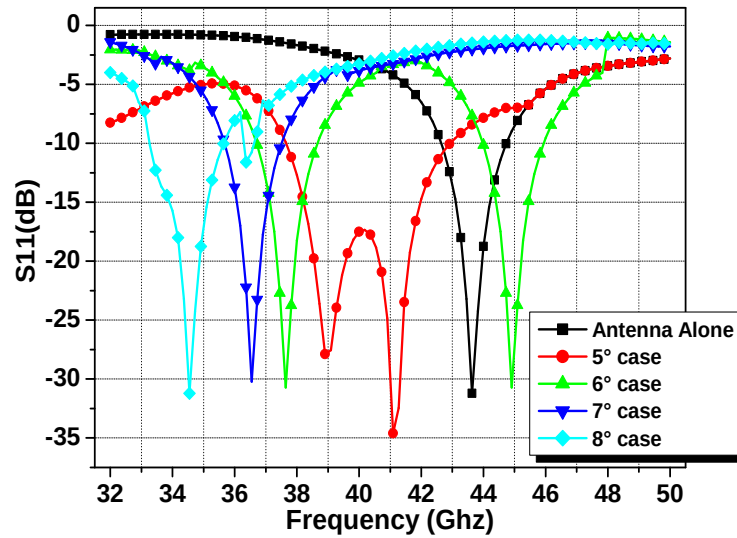


Figure 4-23 : Resonance frequency of each possible combination

We'll wrap up with the table 4-2 that its shown the resonance frequency of all the simulated possible combinations, it can be seen that our proposed agile antenna/radome system based on metamaterial FSS layers is programmable according to the wished resonance frequency, some cases has two resonance frequencies like the case number 5 and 6, notice that the lower frequency LF it's in the case number 8 and the higher frequency HF it's in the case number 6.

Table 4-2 : Resonance frequency of all cases.

Agile antenna/radome system					Resonance frequency (Ghz)		Gain (dB)
Antenna Alone					43.63		7.3
Cases	Layer 1	Layer 2	Layer 3	Layer 4	--		--
Case 1	Connected	Disconnected	Disconnected	Disconnected	40.18		12.58
Case 2	Disconnected	Connected	Disconnected	Disconnected	38.54		12.81
Case 3	Disconnected	Disconnected	Connected	Disconnected	36.72		12.97
Case 4	Disconnected	Disconnected	Disconnected	Connected	35.45		12.09
Case 5	Connected	Connected	Disconnected	Disconnected	38.9	41.09	13.01
Case 6	Connected	Disconnected	Connected	Disconnected	37.63	44.9	12.88
Case 7	Connected	Disconnected	Disconnected	Connected	36.54		12.82
Case 8	Disconnected	Disconnected	Connected	Connected	34.54		12.93

4.9. Conclusion

The work presented in this chapter examined the possibility of focusing the antenna radiation pattern using an FSS radome, the obtained results show that:

- Loading a microstrip patch antenna with a FSS radome can focus the radiation beam width to 20° in both E- and H-planes and thereby enhances the patch antenna directivity and then increase the realized gain from 7.3 dB to 16.3 dB, so the realized gain is increasing by about 9 dB compared to that of the patch antenna alone.
- Changing the behavior of a metamaterial FSS radome to be in the connected or disconnected state in different positions can shift the resonance frequency to lower or upper wished operating frequency.

Much more work is still needed in the understanding, analysis, design, and fabrication of both metamaterials and metasheets for radomes.

Conclusion

Our goal in this work is to design an agile FSS radome based on metamaterial FSS layers invisible to the electromagnetic waves as much as possible, on the other hand, improve the performances of a patch antenna such as the realized gain and the radiation pattern and agile frequency system, hence, a range of different operating frequencies controlled, using an external active control system, that's why we design an agile FSS radome layer which has two different behaviors and they can be controlled through external command system. This agile metamaterial layer is obtained by a periodic structure based on agile cross type unit cell. The agility is achieved by connecting or disconnecting the wires of the cross. This agile metamaterial, based on connected or disconnected cross unit cell, behaves as an effective medium with a refraction index close to zero (n_1) or greater than unity (n_2), respectively.

This thesis has provided a deeper insight that when a patch antenna loaded with a metamaterial FSS radome can:

- Change the resonance frequency to lower or upper wished one in a fairly important frequency range; these findings suggest that in general we can change the operating frequency of an integrated patch antenna covered by a FSS radome without changing the patch antenna itself, according to the final chapter the FSS radome proposed, provide eleven different operating frequencies, these frequencies are obtained as a simulation results but we can expect in advance some frequencies that we looking for by modifying the FSS radome characteristics, such as the type design of the cross unit cell composing the FSS layer, the thickness of the metamaterial FSS layer, the number of the FSS layers and the position or the distance between each other.

- Increase the realized gain by about 9 dB compared to that of the patch antenna alone; the patch antenna gain improvement without changing the antenna dimensions becomes one of most researchers questions, because of insufficient space between the other electrical components, the evidence reviewed in this thesis seems to suggest that increasing the number of FSS radome layers, from 1 to 4, increases the antenna/radome system gain from 9.4 to 16.3 dB, respectively.

- Focus the radiation beam width to 20° in both E- and H-planes; according to the results obtained in the last chapter, as the number of the FSS radome layers increases, we conclude that the radiating system half-power beamwidth decreases from 66° (antenna alone) to

20° (antenna/radome with 4 layers) in the E-plane and from 80° (antenna alone) to 20° (antenna/radome with 4 layers) in the H-plane.

- Enhance the patch antenna directivity; these findings provide additional evidence that as the number of the FSS radome layers increases, the energy radiated by the patch antenna becomes more concentrated, therefore more directive and presents a higher gain.

Taken together, these studies support the notion that adding a FSS radome layer or changing the behavior of a metamaterial FSS radome layer to be in the connected or disconnected state in different positions by an external active control system can improve the functioning of the patch antenna according to our needs.

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