The effects of ground planes in modern electromagnetic applications

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The Effects of Ground Planes in Modern Electromagnetic Applications

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in fulfillment of the requirements for the degree of

Doctor of Philosophy

2010

Statement of Originality

I hereby certify that the content of this thesis is the result of work done by me and has not been submitted for a higher degree to any other University or Institution.

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Date

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Huang Shao Ying

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Now is the future!

To my family and Yu Hai.

Summary

This thesis presents the study of the effects of ground plane geometries on the propagation, the coupling and radiation, and the scattering of electromagnetic waves. The effects are studied and demonstrated using various structures in different frequency bands in the electromagnetic spectrum.

The study of the effects on wave propagations is performed through four structures. The E-shaped patterned ground structure, the straight and the U-shaped dual planar electromagnetic band-gap structures, and the optimized conventional stepped impedance low-pass filter applying a microstrip line with a slot in the ground plane. A circuit model is proposed to ease the study of the behavior of waves within the passband and the stopband in a periodic microstrip structure.

The proposed E-shaped patterned ground structure with a single etched unit provides a sharp notched stopband with high attenuation. It provides sharper transition knees before and after the resonant frequency and a higher Q-factor (lower loss) within a smaller circuit area as compared to other patterned ground microstrip structures in the literature. This new pattern makes it possible to realize a tunable bandstop filter within a small circuit area.

A straight and a U-shaped dual planar EBG microstrip structures are then proposed to provide wide stopband with high attenuation. These two structures are periodic structures and therefore have high passband ripple levels. However, the ripple levels in the passband are effectively lowered by applying tapering tech-

niques. In order to reduce the circuit area of the straight EBG structures, the straight geometry is bent into a U-shape. This unique bent structure demonstrates a large bandwidth with high attenuation within the stopband. Its passband ripple level is low.

Other than the three new designs, the effects of ground planes on the propagations of waves are also examined through the optimization of a conventional filter structure using a patterned ground structure. Due to the wave propagation characteristics of the patterned ground structure, the optimized filter shows a sharper cutoff, a better suppression of spurious passbands, and a much smaller circuit area than the conventional filter. The optimization applying patterned ground structures not only requires no additional circuit areas, but also reduces the physical size of the original device due to the slow-wave effect.

The study of patterned ground structures naturally leads to the concern of the coupling and radiation effects of such structures due to the defected ground plane. Therefore, the study of the electromagnetic compatibility and the electromagnetic susceptibility of a dual planar EBG microstrip structure is examined. The effects of the radiation and the coupling caused by a circuit component near to the EBG structure are studied and analyzed. The radiation generated by patterned ground structures in the far-field region is examined in order to analyze the effects of the etched patterns on the radiation in detail. This study helps the understanding of the radiation and coupling properties of patterned ground structures when the structures are integrated into a microwave circuit when no additional shielding mechanisms are used.

The final part of the thesis examines the effects of ground plane geometries on the scattering. The surface-enhanced Raman scattering (SERS) active substrate at optical frequencies is detailed. A metallic ground plane is proposed to be incorporated to the bottom of a SERS substrate. It helps to reflect waves and to increase

the electric field intensity at the nanoparticle surface. An analytic method is proposed to quantify the effect of the metallic-layer-bottomed substrate in terms of the dielectric constant and the thickness of each layer on the E-fields that are available at the surface of the particles. This method is useful for the optimization of the construction of a SERS substrate so as to avoid poorly constructed substrates. This global effect of a substrate in SERS can be significant and therefore, overshadow the strong local electromagnetic effects. The optimized construction of a SERS substrate can contribute to a significant increase in the Raman enhancement factor.

The thesis contributes to the understanding of waves in a transmission line with etched pattern(s) in the ground plane. Novel patterned ground structures with high performance and small physical size are proposed. The circuit modeling of dual planar EBG structures is also proposed to ease the optimizations and the analysis of such structures. The electromagnetic compatibility and susceptibility of these patterned ground structures are presented to understand the effects of integrating such structures directly into microwave circuits. The scattering of waves in a SERS substrate with bulk metal at the bottom is detailed. The effect of a substrate in SERS experiments is presented. This thesis contributes to the understanding of the behaviors of waves when new patterns are introduced to a ground plane and when the properties of metal used as a ground plane are complicated at optical frequencies.

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List of Acronyms

CPW	Coplanar waveguide
DDA	Discrete dipole approximation
DGS	Defected ground structure
DUT	Device under test
EBG	Electromagnetic band-gap
EMC	Electromagnetic compatibility
EMS	Electromagnetic susceptibility
FDTD	Finite-difference time-domain
MLIN	Microstrip line
MMIC	Monolithic microwave integrated circuits
MoM	Method of moments
NSOM	Near-field scanning optical microscope
PBG	Photonic band-gap
PUT	Plane under test
SEM	Scanning electron microscope
SERS	Surface-enhanced raman scattering
SMA	SubMiniature version A
SOC	Short-open calibration
SPP	Surface plasmon polariton
TE	Transverse electric

TEM	Transverse electromagnetic
TM	Transverse magnetic
UC-PBG	Uniplanar compact photonic band-gap

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Chapter 1

Introduction & Literature Review

In the studies of electromagnetic waves, metallic structures always play an important role in affecting the behavior of waves in terms of propagation, coupling, radiation, and scattering. In the past decades, there has been significant progress in the research field of telecommunications. High-speed data services for voice and multimedia applications are in high demand. Devices working at microwave frequencies, ranging from 0.3 GHz to 300 GHz, are required to work at the higher end of this frequency spectrum over a broad bandwidth. Unfortunately, when the operating frequency is high, traditional microwave circuits exhibit limitations due to the increase in loss and the constraint in physical size. Planar microwave circuits (printed microwave circuits) such as patch antennas, microstrip-line-based devices, and coplanar-waveguides (CPW)-base devices show attractive properties such as low profile, light weight, compactness, ease for array arrangement, compatibility with monolithic microwave integrated circuits, and ease of fabrication. These features make them applicable to portable wireless communication devices and to phased-array devices for military applications. With the potential wide applications in communications, the research and development in planar microwave devices is one of the main areas in the microwave research community. In this

research area, modifications of the metallic layers that are acting as the ground planes of structures provide effective solutions for performance optimizations and size reductions. It is one of the research topics that have captured a lot of attention and are becoming increasingly important.

Higher up in the electromagnetic spectrum, at optical frequencies, fundamental research in the field of optical structures has made significant progress. Research on optical structures for the application of spectroscopy has made its significant advancement through the detection of single molecules using surface-enhanced Raman scattering (SERS). This progress benefits many fields, such as chemistry, biology, medicine, pharmacology, and environmental science. Not only is the research on optical structures driven by potential applications, but also it is pushed by the rapid advances in nanofabrications and by the theoretical research. Theoretical modeling of optical structures, analytically or numerically, is useful to identify new physics in this area and to guide the designs and optimizations of the structures.

Studies of the effect of new metallic components/layers in structures on the propagation, coupling, radiation, and scattering of electromagnetic waves are important to lead to novel designs meeting new requirements and to obtain a basic understanding of working principles of the structures for their optimizations. This thesis covers the propagation, coupling, radiation, and scattering of electromagnetic waves from microwave frequencies to optical frequencies. The main focus of this study is on the effects of ground planes with etched patterns in devices working at microwave frequencies and the effect of a ground plane with bulk metal in structures working at the optical frequencies. For microwave devices, the ground planes mainly affect the wave propagation, coupling, and radiation. The filtering functionality of the structures with patterned ground plane is studied in this thesis. The EMC issues due to the etched patterns is addressed and investigated. For optical structures, the effects of ground planes on the scattering of waves are

examined.

In the following section, a detailed literature review is given on the studies of patterned ground structures in terms of their propagation, coupling, and radiation. The review also includes the research progress on the effects of layers in a substrate in surface-enhanced Raman scattering. Through the review, research gaps are identified which motivate the work in this thesis. The literature review is followed by the motivations and objectives of the thesis. At the end of this chapter, the major contributions in this thesis are summarized and the organization of the thesis are presented.

1.1 Literature Review

1.1.1 Patterned Ground Structures

Patterned ground structures are transmission lines with etched patterns, such as slots or/and apertures, in the ground plane. They have been a topic of interest in the society of microwave research and engineering in recent years. These structures are often referred to as “defected ground structure”. They are called EBG structures when the etched patterns are arranged periodically. The etched patterns introduce additional equivalent inductance and capacitance seen by a transmission line. Therefore, they can cause resonance, a change in characteristic impedance, and a change in phase velocity of the waves propagating along the transmission line. Patterned ground structures show a prominent stopband with a designable center frequency. They exhibit slow-wave effects at the edge of the stopband. Because of the designable stopband, patterned ground structures can be applied to filter designs to enhance the filters’ functionalities from conventional designs [3, 31, 32, 33]. They can also be applied to microwave circuit components, such as amplifiers [34] and antennas [35], to filter out unwanted signals so as to significantly improve the

performance of the components. Due to the slow-wave effect, patterned ground structures applied to microwave circuits introduce reduction in physical size and improvement in performance. Moreover, microstrip patterned ground structures are easy to fabricate and are compatible with monolithic microwave circuits. The structures under study for this thesis (to study the effects of patterned ground planes on the propagation, coupling, and radiation of electromagnetic waves) are planar patterned ground structures, especially microstrip patterned ground structures. In the content that follows, patterned ground structures refer to microstrip patterned ground structures.

Patterned ground structures with a single etched unit in the ground plane are bandstop resonators [3, 4, 6, 7, 9, 28, 36, 37] which show narrow stopbands. The resonant frequency of the structure is determined by the geometry of the etched pattern. The geometry is also related to the quality factor (Q-factor) and physical size of the structure. When more than one etched unit are arranged periodically along the microstrip line, an increase in the bandwidth and attenuation of the stopband can be obtained [12, 32, 38, 39]. These structures with periodically arranged etched patterns are EBG structures. The center frequency of the stopband is determined by the Bragg reflection condition [40]. In this section, a review on patterned ground structures is detailed.

1.1.1.1 Patterned Ground Structures with Single Etched Units

a. Early Research on Patterned Ground Structures

Back in 1986, a microstrip line with a slot etched in the ground plane was studied for its application for antennas and for the feeding of a patch antenna [1]. Fig. 1.1 shows a three dimensional (3-D) view of the microstrip structure which is now known as a patterned ground structure. Fig. 1.2 shows an aperture coupled microstrip patch antenna where a microstrip line with a slot in the ground plane is

Figure 1.1: Geometry of a printed slot in the ground plane of an infinite microstrip line, the inset of the figure shows the equivalent circuit model of the slot in the ground plane of the microstrip line [1].

used to feed a patch antenna. In [1], the electromagnetic fields of the patterned ground structure were analyzed using the reciprocity theorem. As a result of the analysis, the slot in the ground plane can be modeled using a series impedance Z to the microstrip line as shown in the inset in Fig. 1.1. A full moment method solution was applied to the aperture distribution to calculate the series impedance Z . The calculated and measured normalized Z 's, $\bar{Z}$'s ($\bar{Z} = \bar{R} + i\bar{X}$), are shown in Fig. 1.3 (Fig. 1.3 (a) and (b) show $\bar{R}$ and $\bar{X}$, respectively). In Fig. 1.3, it can be seen that both the resistance and the reactance are dispersive. As shown in Fig. 1.3 (a), the equivalent resistance has a peak value at around 3.0 GHz whereas in Fig. 1.3 (b), it is shown that the slot is inductive at low frequencies and capacitive at high frequencies. In Fig. 1.3 (b), $\bar{X}$ is zero at a frequency slightly higher than 3.0 GHz where the resonance occurs. At the resonant frequency, $\bar{R}$ is close to

Figure 1.2: Geometry of an aperture coupled microstrip patch antenna [1].

its peak value. This indicates a severe impedance mismatch at the resonance.

Fig. 1.4 shows the simulated S-parameters of the patterned ground structure shown in Fig. 1.1 using Microwave Studio [2]. As shown in Fig. 1.4, the transmission coefficients have a minimum at 3.2 GHz whereas the reflection coefficients have a maximum at 3.0 GHz. The maximum S_{11} -parameter is due to the maximum impedance mismatch as shown in Fig. 1.3(a). The minimum S_{21} -parameter is due to two mechanisms, one is the reflections due to impedance mismatches and the other is the resonance when the stored electrical energy equals the stored mechanical energy or when the reactance of the series impedance, $\overline{X}$, equals to

zero. Therefore, the frequency corresponding to the minimum S_{21} -parameter is shifted.

A patterned ground structure with a slot as shown in Fig. 1.1 was used as a printed slot antenna at its resonance with a tuning stub on the slot to eliminate the impedance mismatch [41]. Similarly, this structure was also applied to feed a patch antenna [1, 42]. Such antennas are called aperture coupled microstrip patched antennas. In an aperture coupled microstrip patch antenna, the electrical size of the slot is smaller than its resonant size and the radiation is from the resonance of the patch.

The trends of the resistance and the reactance in Fig. 1.3 (a) and (b), respectively, suggest a parallel RLC circuit for the series Z in the equivalent circuit. For a parallel RLC circuit as shown in Fig. 1.5 (a), its input impedance can be expressed as (1.1),

$$Z = \frac{1}{\frac{1}{R} + \frac{1}{j\omega L} + j\omega C} = \frac{R^2(1 - \omega^2 LC)\omega Lj + R(\omega L)^2}{(R - R\omega^2 LC)^2 + \omega^2 L^2} = R' + jX' \quad (1.1)$$

where ω is the angular frequency. Fig. 1.5 (b) and (c) show this RLC circuit using its equivalent impedances Z and $R' + jX'$, respectively. Primes are used to distinguish the values in Z from those in the RLC circuit. Therefore,

$$R' = \frac{R(\omega L)^2}{(R - R\omega^2 LC)^2 + \omega^2 L^2} \quad (1.2)$$

$$X' = \frac{R^2(1 - \omega^2 LC)\omega L}{(R - R\omega^2 LC)^2 + \omega^2 L^2} \quad (1.3)$$

As can be seen in (1.3), X' goes from a positive value to a negative one when the frequency increases for some values of L and C . The resonant frequency ω_r can be obtained by letting $X' = 0$. Therefore, let $1 - \omega^2 LC = 0$, we obtain $\omega_r = 1/\sqrt{LC}$.

(a)

(b)

Figure 1.3: Measured (xx) and calculated (-) $\bar{Z}$ of the patterned ground structure in Fig. 1.1 with $\epsilon_r = 2.20$, $d = 0.16$ cm, $L = 4.02$ cm, $W = 0.07$ cm, and $W_f = 0.50$ cm [1]

Figure 1.4: Simulated S-parameters of the patterned ground structure in Fig. 1.1 ($\epsilon_r = 2.20$, $d = 0.16$ cm, $L = 4.02$ cm, $W = 0.07$ cm, and $W_f = 0.50$ cm) [2].

If R' is differentiated with respect to ω as shown in (1.4),

$$\frac{\delta R'}{\delta \omega} = \frac{-RL^2(-2R^2\omega^{-3} + 2\omega R^2L^2C^2)}{(L^2 - 2LCR^2 + R^2/\omega^2 + \omega^2 R^2L^2C^2)^2} \quad (1.4)$$

the maximum of R' is at the frequency ω_{mR} . It is obtained when $-2R^2\omega_{mR}^{-3} + 2\omega_{mR}R^2L^2C^2 = 0$. Therefore, $\omega_{mR} = 1/\sqrt{LC}$ when the resistance is a maximum. The resonant frequency, ω_r , can be shifted by introducing additional reactance, jX_a , as shown in Fig. 1.5 (d). Therefore, ω_{mR} and ω_r can be different and a match impedance can be obtained at resonance. It can be used to explain the difference between the resonant frequency and the frequency at which there is a maximum reflection, shown in Fig. 1.4. In Fig. 1.4, the 0.2 GHz-shift in the resonant frequency is due to the parasitic inductance of value jX_a . This concept has been applied to the design of slot antennas [41, 43].

In the 80's, significant amount of research work was done on the radiation of a patterned ground structure and its equivalent circuit for impedance matching. Al-

Figure 1.5: A RLC circuit in different forms

though a microstrip line with a slot in the ground plane shows a resonance, the frequency response is not good enough to be applied to the design of a notch bandstop filter due to the lack of sharp transitions between a passband and the attenuation at the resonant frequency. In the past decade, the research focus of patterned ground structures lies in the propagation and the circuit modeling. Since 2001, different geometries of apertures which shows different frequency responses have been reported for patterned ground structures [3, 4, 6, 7, 9, 28, 36, 37].

b. Dumbbell-shaped Patterned Ground Structures

Fig. 1.6 shows a 3-D view of patterned ground structure with a dumbbell-shaped etched pattern in the ground plane [3]. With the dumbbell-shaped etched pattern in the ground plane, there is high attenuation at the resonant frequency and sharp transitions between the passband and the attenuation. This is the first patterned ground structure with a single etched unit reported that provides a frequency response that is good enough to function as a bandstop resonator. As shown in

Figure 1.6: A 3-D view of the dumbbell-shaped patterned ground structure proposed in [3].

Fig. 1.6, the slot directly below the microstrip line gives rise to an increase in the equivalent capacitance seen by the microstrip line whereas the two rectangular apertures on the two sides of the microstrip line increases the equivalent inductance of the structure. Fig. 1.7 (a) shows typical frequency responses of the dumbbell-shaped structures when the dimensions of the rectangular apertures change. As shown in Fig. 1.7 (a), the size of the aperture affects both the cutoff and the resonant frequency. It indicates that the rectangular apertures affect the inductance of the structure. The dumbbell structure was modeled using an inductor and a capacitor in parallel as shown in Fig. 1.7 (b) [3]. This equivalent circuit agrees well with the one for a microstrip line with a slot in the ground plane when R is set to infinite.

Based on this structure, many other patterned ground structures with modified geometries of the etched patterns were proposed [4, 6, 37]. The development of patterned ground structures is oriented towards the reduction in the overall physical size of the structure. Due to the resonance nature of patterned ground structures,

(a) Frequency responses

(b) An equivalent circuit

Figure 1.7: The dumbbell-shaped defected ground structure (DGS) [3]

the Q-factor is another important parameter to be considered for the evaluation of the performance of such structures.

c. Q-Factor

Q-factor, Q , is a characteristic of resonant circuits/transmission line resonators. For a resonant circuit with RLC in series or in parallel, shown in Fig. 1.8 (a) and (b), respectively, its Q-factor is defined in the literature as (1.5) [44, 45],

$$Q = \frac{\omega W_T}{P_L} \quad (1.5)$$

where ω is the angular frequency, W_T is the total electromagnetic energy expressed

Figure 1.8: RLC resonators

Table 1.1: Q-Factors of Resonators

Resonator	Q-Factor
Parallel	$Q = \omega_0 RC = R/(\omega_0 L)$
Series	$Q = (\omega_0 L)/R = 1/(\omega_0 RC)$

in (1.6), and P_L is the power loss.

$$W_T = W_e + W_m \quad (1.6)$$

W_e and W_m are the time-averaged electric and magnetic energy, respectively.

In both the series and the parallel resonant circuits, the capacitor stores electric energy, the inductor stores magnetic energy, and the resistor dissipates power. The time-averaged electric and magnetic energy are expressed using (1.7) and (1.8), respectively.

$$W_e = \frac{1}{4} |V_0|^2 C \quad (1.7)$$

where C is the capacitance of the capacitor and V_0 is the voltage across the capacitor.

$$W_m = \frac{1}{4} |I_0|^2 L \quad (1.8)$$

where L is the inductance of the inductor and I_0 is the current in the inductor. The power loss, P_L , can be expressed as

$$P_L = \frac{1}{2} I_R^2 R = \frac{1}{2} \frac{V_R^2}{R} \quad (1.9)$$

Figure 1.9: A loaded parallel RLC resonator

where R is the resistance of the resistor and I_R and V_R are the current through and the voltage across the resistor, respectively. Therefore, at the resonance, due to the fact that $W_e = W_m$ at resonance,

$$Q = \frac{\omega_0 W_T}{P_L} = \frac{\omega_0 (W_e + W_m)}{P_L} = \frac{2\omega_0 W_e}{P_L} \quad (1.10)$$

or

$$Q = \frac{2\omega_0 W_m}{P_L} \quad (1.11)$$

ω_0 is the resonant frequency. For a parallel RLC resonator as shown in Fig.1.8 (b), V_0 equals V_R . Equation (1.10), (1.7), and (1.9) can be used and the Q-factor can be expressed as

$$Q = \frac{2\omega_0 W_e}{P_L} = \frac{2\omega_0 \frac{1}{4} V_0^2 C}{\frac{1}{2} \frac{V_0^2}{R}} = \omega_0 RC \quad (1.12)$$

If the magnetic energy is expressed in terms of V_0 as

$$W_m = \frac{1}{4} \left| \frac{V_0}{i\omega L} \right|^2 L = \frac{1}{4} \frac{V_0^2}{\omega^2 L} \quad (1.13)$$

Thus the Q-factor can be obtained using (1.14) below,

$$Q = \frac{2\omega_0 W_m}{P_L} = \frac{2\omega_0 \cdot \frac{1}{4} \frac{V_0^2}{\omega_0^2 L}}{\frac{1}{2} \frac{V_0^2}{R}} = \frac{R}{\omega_0 L} \quad (1.14)$$

For a series RLC resonator as shown in Fig.1.8 (a), I_0 equals I_R . Thus (1.11),

Figure 1.10: A ground view of the spiral-shaped patterned ground structure [4]

(1.8), and (1.9) can be used and the Q-factor of the circuit is expressed as

$$Q = \frac{2\omega_0 W_m}{P_L} = \frac{2\omega_0 \cdot \frac{1}{4} I_0^2 L}{\frac{1}{2} I_0^2 R} = \frac{\omega_0 L}{R} \quad (1.15)$$

If the electric energy is expressed in terms of I_0 as in (1.16)

$$W_e = \frac{1}{4} |I_0 \frac{1}{i\omega C}|^2 C = \frac{1}{4} \frac{I_0^2}{\omega^2 C} \quad (1.16)$$

The Q-factor can be obtained as,

$$Q = \frac{2\omega_0 W_e}{P_L} = \frac{2\omega_0 \cdot \frac{1}{4} \frac{I_0^2}{\omega_0^2 C}}{\frac{1}{2} I_0^2 R} = \frac{1}{\omega_0 RC} \quad (1.17)$$

The equations for obtaining Q-factors for a parallel and a series resonator are summarized in Table 1.1.

The Q-factor for a resonant circuit discussed above is a Q-factor when the circuit has no external circuitries. It is therefore called the unloaded Q-factor, Q_u .

Unloaded Q-factor is an important parameter that can be used to evaluate the in-

trinsic loss of a resonant circuit [44]. The Q-factor of a resonant circuit is lowered if it is connected to an external load, R_L , because R_L introduces additional loss to the circuit. Fig. 1.9 shows a loaded parallel RLC resonator, as an example. The Q-factor of a resonant circuit that has an external load is called loaded Q-factor, Q_ℓ . In order to calculate Q_ℓ , R in (1.12), (1.14), (1.15) and (1.17) is replaced by an effective resistance, R_e . The effective resistance can be obtained using (1.18)

$$R_e = \frac{RR_L}{R + R_L} \quad (1.18)$$

for a parallel resonator and (1.19)

$$R_e = R + R_L \quad (1.19)$$

for a series resonator.

For a transmission line resonator that is terminated with a matched load, the Q-factor at the resonance is defined as

$$Q = \omega_0 \frac{\text{Energy per unit length}}{\text{Power loss per unit length}} \quad (1.20)$$

where ω_0 is the angular resonant frequency. Power transmitted in the transmission line equals energy per unit length multiplied by the energy transport velocity which is the phase velocity, v_g . Therefore, Q can be expressed as

$$Q = \frac{\omega_0 P / v_g}{P'_L} \quad (1.21)$$

where P is the power transmitted in the transmission line and P'_L is the power loss per meter. If α is the attenuation constant in a lossy transmission line with the direction of wave propagation in the $\hat{z}$ -direction and P_0 is the power at $z=0$, the

(a) Frequency responses

(b) An equivalent circuit

Figure 1.11: A spiral-shaped patterned ground structure [4]

power flow, P , can be expressed as below,

$$P = P_0 e^{-2\alpha z} \quad (1.22)$$

Based on (1.22), the power decreasing rate can be obtained by taking the derivative with respect to z .

$$P'_L = -\delta P / \delta z = 2\alpha P_0 e^{-2\alpha z} \quad (1.23)$$

Figure 1.12: Slow-wave factors of CPW's [5]

Substitute (1.22) into (1.23), we have

$$P'_L = 2\alpha P \quad (1.24)$$

Based on (1.21) and (1.24), the Q-factor of a transmission line can be obtained using (1.25) below,

$$Q = \frac{\omega_0}{2\alpha v_g} \quad (1.25)$$

d. Research Progress on Patterned Ground Structures with Single Etched Units

Significant progress has been achieved in research on patterned ground structures in recent years. Most of these research focus on the improvement in filtering functionality, an increase in unloaded Q-factor, and a reduction in physical size. In [37], the shape of the aperture in a dumbbell-shaped patterned ground structure was changed from rectangular to arrow-shaped or circular. The change in the shape of the aperture has minor effect on the performance and the physical size as compared to the dumbbell-shaped patterned ground structure.

Figure 1.13: A ground view of the double-H-shaped patterned ground structure [6]

The spiral-shaped patterned ground structure proposed in [4] exhibits high performance. Fig. 1.10 shows the patterned ground plane view. With the introduction of the spiral-shaped etched pattern in the ground plane, the equivalent inductance of the microstrip line as shown in Fig. 1.7 (b) increases. As a result, in Fig. 1.11 (a), the performance of the structure is improved significantly in terms of the sharpness of the transition knee. In Fig. 1.11 (a), it is also noted that the second resonant frequency is close to the first one. In order to model the frequency response, a more sophisticated circuit as shown in Fig. 1.11 (b) was used [4]. A shorted stub with step impedances is used to model the spurious second resonance, as shown in Fig. 1.11 (b). The small difference in frequency between the first and the second resonant frequencies limits the applications of the structure. The spiral-shaped etched pattern was applied to a coplanar waveguide (CPW) [5]. For the spiral-shaped patterned ground CPW, a resonance was obtained with a sharp transition knee. The spurious second resonance is farther away from the first resonance compared to the patterned microstrip line structure. It is applied to reduce phase noise of a V-band CPW monolithic microwave integrated circuit (MMIC) oscillator [46]. Moreover, the spiral-shaped CPW shows slow-wave effect near the edge of the stopband, as shown in Fig. 1.12.

(a) 3-D view

(b) S_{21} 's when l is varied, $c = 0.2$ mm, $g = 0.2$ mm, $d = 1.0$ mm

Figure 1.14: The U-slot patterned ground structure [7]

The double-H-shaped resonant cell was proposed in [6] which shows good filtering functionality in a small circuit area. Fig. 1.13 shows the ground view of this resonant cell. Compared to the dumbbell-shaped cell with circular apertures at two sides of the microstrip line [37], the double-H-shaped structure shows a large reduction in physical size and a significant enhancement in the sharpness of the transition knee.

In terms of miniaturization and the sharpness of the transition knees, the proposed U-slot and V-slot structures show significant progress [7]. Fig. 1.14 (a) and Fig. 1.15 (a) show the 3-D views of the U-slot and V-slot structures, respectively. As shown in Fig. 1.14 (a), a U-slot structure has two slot lines in the direction of the transmission line and the rectangular etched patches in a traditional dumbbell-shaped structure are not required. Therefore, the structure occupies less circuit area and is more compact. Fig. 1.14 (b) shows the transmission coefficients when the length ℓ is varied while other dimensions are kept constant. The dielectric material of thickness 1.27 mm and of dielectric constant 10.2 was used as the substrate. As shown in Fig. 1.14 (b), the U-slot structures show sharp transition knees. The lengths of the two arms affect the resonance frequency but not the sharpness of the knees. The V-slot structure is a variant of the U-slot structure. Fig. 1.15 (b) shows the simulated S_{21} -parameters of the V-slot structure when the angle of the two arms, θ , is varied and other dimensions are kept constant. As shown in Fig. 1.15 (b), θ affects the sharpness of the knees but not the resonant frequency.

The single unit microstrip complementary split-ring resonator (CSRR) [47] also shows a notch stopband. It is the idea of split-ring resonators [48] extended to patterned ground structures. Fig. 1.16 (a) shows a square CSRR presented in [8] and Fig. 1.16 (b) shows the simulated transmission coefficients when the length of the edge of the CSRR, a , is varied. As can be seen in Fig. 1.16 (b), the structures

(a) 3-D view

(b) S_{21} 's when θ is varied, $\ell = 7.0$ mm, $g = 0.2$ mm

Figure 1.15: The V-slot patterned ground structure [7]

(a) 3-D view

(b) S_{21} -parameters

Figure 1.16: A single unit microstrip CSRR [8]

with a single CSRR show notch stopbands. It is noticed that the insertion loss in the lower passband is small while the insertion loss in the upper passbands is higher than 5 dB. Due to the high insertion loss in the upper passband, it is not suitable to be used as bandstop filters. Nevertheless, the microstrip line of this structure can be modified to further eliminate the transmission in the upper passband obtaining a low-pass filtering functionality, as presented in [8]. Moreover, the high insertion loss in the upper passband can be lowered by cascading the CSRR units in a periodic manner [47].

In [28], the conventional dumbbell-shaped patterned ground structure is further modified by inserting metal patches at the center of the rectangular aperture, as shown in Fig. 1.17 (a). In Fig. 1.17 (a), the slot connecting the two rectangular apertures directly below the microstrip line is changed to a circular aperture in order to ease fabrications. With this modification, an increase in capacitance and a reduction in circuit area are both obtained. Moreover, the structure becomes tunable simply by inserting varactors between the inserted patch and the ground plane. Fig. 1.17 (b) shows a tunable coplanar waveguide structure with a modified dumbbell-shaped etched pattern in the ground plane [36]. As shown in Fig. 1.17 (b), varactors are inserted across the gap between the inserted patch and the ground plane. This tunable patterned ground structure shows a frequency tuning range of 19%.

Recently, a T-shaped patterned ground structure was proposed to achieve a significant improvement in performance together with a large reduction in size [9]. Fig. 1.18 (a) shows the ground view of the T-shaped patterned ground structure. As shown in Fig. 1.18 (a), the slot of the dumbbell-shaped patterned ground structure in Fig. 1.6 (a) is moved to the edges of the etched patch with arms and it is lengthened, forming a T-shape. The dumbbell-shaped, double-H-shaped, and the modified dumbbell-shaped patterned ground structures are compared in terms of

(a) Microstrip structure [28]

(b) Coplanar waveguide with varactors [36]

Figure 1.17: Modified dumbbell-shaped patterned ground structures

(a) Without varactors

(b) With varactors

Figure 1.18: Ground views of T-shaped patterned ground structures [9]

performance and physical size at a fixed resonant frequency in [9]. Among the patterned ground structures under comparison, the T-shaped structure has the smallest physical size and the sharpest transition knees. Moreover, with the lengthened slot, it becomes possible to embed a varactor into the arms, allowing the tuning of the resonant frequency. Fig. 1.18 (b) shows the schematic of the tunable T-shaped structure. As shown in Fig. 1.18 (b), one varactor is added in each arm of the T-shaped structure. This tunable T-shaped structure shows a maximum frequency tuning range of 13.2%.

Besides the patterned ground structures with a single unit etched in the ground plane, there are structures with periodic etched units. They provide a wide stopband rather than a narrow stopband at the resonance. They are known as EBG

Figure 1.19: The reflection of waves at a stacked crystal lattice plane

Figure 1.20: Generic schematics of photonic band-gap structures

structures. They are widely used in the designs of microwave circuits. Next, a review will be done on EBG microstrip structures.

1.1.1.2 Electromagnetic Band-Gap (EBG) Microstrip Structures

Patterned ground structures with periodic etched patterns in the ground plane of microstrip lines are referred to as planar EBG microstrip structures in the literature. “Planar” in this content refers to the dimension of unit cells. The etched patterns can be arranged in two dimension (2-D) (multiple columns along the microstrip line) or in one dimension (1-D) (single column). It is worth mentioning that “two dimension” here refers to the arrangement of the periodic unit cells while “planar” describes the single plane on which the unit cell resides. The waves in a planar EBG microstrip structure are guided along the microstrip line in a single

Figure 1.21: Right-hand side, the electromagnetic dispersion with a forbidden gap; left-hand side, the typical electron wave dispersion of a direct-gap semiconductor, the dots represent electrons and holes [10]. h , $\hbar$, ν are the Plank's constant, the reduced Plank's constant, and the frequency of the emitted photon, respectively.

direction.

a. The Bragg Reflection Condition & Photonic Band-Gap (PBG) Structures

The Bragg reflection condition [40] proposed in 1913 describes the working principle of reflections of X-rays incident at a crystal. Fig. 1.19 shows the reflection of incident X-rays towards stacked crystal lattice planes at an angle of θ with respect to the lattice planes. The distance between successive lattice planes is d . Each lattice plane reflects waves obeying Snell's law. In order to obtain an enhancement of reflections at the output, it requires the reflected waves from each plane are in phase, which is equivalent to $X'Y'Z - XYZ = n\lambda_g$ in Fig. 1.19, where n is an integer

and λ_g is the guided wavelength in the crystal. As can be seen in Fig. 1.19, this is equivalent to $Y'Y - Y'V = n\lambda_g$. Due to Snell's law, $Y'Y = Y'W$. Thus $VW = n\lambda_g$, leading to

$$2d \sin \theta = n\lambda_g \quad (1.26)$$

Equation (1.26) is called Bragg equation and θ is called Bragg angle. For an incident wave that is perpendicular to the lattice planes, $\theta = 90^\circ$, and $n = 1$, the relation between the wavelength that corresponds to the center frequency of the stopband and the spatial periodicity d can be expressed as (1.27),

$$d = \frac{\lambda_g}{2} \quad (1.27)$$

In the analysis above, the crystal lattice plane is assumed to be infinitesimally thin and the Bragg reflections dominate. In a multi-layer structure where each layer has a finite thickness, for example, the structure as shown in Fig. 1.20 (a), the wavelength at which the reflection is enhanced may be shifted by the reflections at the boundaries of the layers.

The Bragg reflection condition can be applied in a 1-D, 2-D, and 3-D manners in order to modulate the refractive index of dielectric materials periodically to open up stopbands (also called forbidden gaps) in the electromagnetic spectrum in one, two, and three directions. Fig. 1.20 (a), (b), and (c) shows generic schematics of structures that have 1-D, 2-D, and 3-D periodicities. This type of structures is given a name, photonic crystals or photonic band-gap structures. A distributed-feedback laser [49] is an example of a 1-D photonic band-gap structure. In the photonic band gap of a photonic crystal, optical modes and spontaneous emission are absent. Fig. 1.21 from [10] is used to illustrate the prohibition of spontaneous emission by a photonic band gap for direct-gap semiconductors. As shown in Fig. 1.21, the right-hand side shows the photonic dispersion (frequency versus k -

(a) Top view

(b) Transmission attenuation

Figure 1.22: A fabricated photonic band-gap structure that shows a measured forbidden gap from 13 GHz to 16 GHz [10]

Figure 1.23: A rectangular waveguide loaded with capacitive diaphragms at a regular interval of l

vector). A photonic band gap can be seen at the frequencies where there are no k -vectors and it is shadowed. The left-hand side of Fig. 1.21 shows a plot of the electronic dispersion of a direct-gap semiconductor, including conduction and valence bands. The dots represent the electrons and holes. The electron k -vectors are scaled to fit in the same graph. If an electron was to recombine with a hole, they would produce a photon at the electronic-band-edge energy. In Fig. 1.21, if the photonic band gap straddles the electronic band edge, then the photon produced by electron-hole recombination would have no place to go. In such a way, the spontaneous radiative recombination of electrons and holes would be prohibited.

Fig. 1.22 (a) shows the top view photograph of a fabricated 3-D photonic band-gap structure. A forbidden gap was successfully shown in experiments. Fig. 1.22 (b) shows the measured S_{21} -parameters of this structure. As can be seen in Fig. 1.22 (b), the forbidden gap is from 13 GHz to 16 GHz at microwave frequencies [10].

When the Bragg reflection condition is applied to a waveguide or a transmission line, the 1-D periodicity is along the direction of the wave propagation generating stopbands and passbands. The studies on the propagation of waves in an infinite 1-D periodic structure is documented in Chapter 9 in R. E. Collin's *Field Theory of Guided Waves* [45]. Fig. 1.23 shows the side view of a rectangular waveguide

Figure 1.24: The unit cells from terminal plane #1 to #2 with a lossless quadrupole

loaded with capacitive diaphragms at a period of l . In Fig. 1.23, a unit cell in the structure is from terminal plane #1 to #2. In the analysis, a lossless quadrupole is used to modeled the unit cell of the periodic structure. Fig. 1.24 shows the unit cell taken out from the two terminal planes with the lossless quadrupole in the middle and $\theta/2$ electrical length of transmission line or waveguide sections at the two ends. The sections are assumed to be sufficiently long so that all higher order modes excited by the discontinuity decay to a negligible value at the terminal planes. They are also assumed to be uniform along the section so that they can be described by a fixed wave propagation constant, β_0 , and a fixed wave impedance, Z_w . The amplitudes of the forward and backward-propagating waves at the input side of the quadrupole are c'_1 and b'_1 , respectively. They are related to the amplitudes of the forward and backward-propagating waves at the output side, c'_2 and b'_2 , using the wave matrix as shown below,

$$\begin{bmatrix} c'_1 \\ b'_1 \end{bmatrix} = \begin{bmatrix} A'_{11} & A'_{12} \\ A'_{21} & A'_{22} \end{bmatrix} \begin{bmatrix} c'_2 \\ b'_2 \end{bmatrix} \quad (1.28)$$

The amplitudes of the forward and backward-propagating waves at the input

Figure 1.25: The equivalent network of a unit cell

side of the unit cell, c_1 and b_1 are related to the amplitudes of the forward and backward-propagating waves at the output side, c_2 and b_2 using the wave matrix $[A]$,

$$\begin{bmatrix} c_1 \\ b_1 \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} \begin{bmatrix} c_2 \\ b_2 \end{bmatrix} \quad (1.29)$$

The wave matrix between the the two terminals, $[A]$, can then be expressed as,

$$A = \begin{bmatrix} e^{j\theta/2} & 0 \\ 0 & e^{-j\theta/2} \end{bmatrix} \begin{bmatrix} A'_{11} & A'_{12} \\ A'_{21} & A'_{22} \end{bmatrix} \begin{bmatrix} e^{j\theta/2} & 0 \\ 0 & e^{-j\theta/2} \end{bmatrix} = \begin{bmatrix} A_{11} & A_{12} \\ A_{21} & A_{22} \end{bmatrix} \quad (1.30)$$

The unit cell in Fig. 1.24 can also be described by an equivalent T-network in Fig. 1.25. If the unit cells as shown in Fig. 1.24 are cascaded, they forms an infinitely long periodic structure as shown in Fig. 1.26. An expression for the propagation constant of the periodic structure, γ can be obtained as [45],

$$\cosh \gamma \ell = \frac{Z_{11} + Z_{22}}{2Z_{12}} \quad (1.31)$$

where $\gamma = \alpha + j\beta$, and

$$\cosh \gamma \ell = \cosh \alpha \ell \cos \beta \ell + j \sinh \alpha \ell \sin \beta \ell \quad (1.32)$$

Based on (1.31), the dispersion relation of a periodic structure can be obtained

Figure 1.26: Cascaded unit cells

according to the geometry of the unit cells. Fig. 1.27 (a) shows a periodic capacitively loaded transmission line. The characteristic impedance of the transmission line, Z_0 , is 50Ω , the period of the structure, ℓ , is 1.0 cm , and the capacitance, C_0 , is 2.67 pF . Therefore, (1.31) becomes,

$$\cosh \gamma \ell = \cos \theta - \frac{\omega C_0 Z_0}{2} \sin \theta \quad (1.33)$$

where $\theta = k\ell$ and k is the wave number in the transmission line before loading. It is assumed to be k_0 here. As shown in (1.33), $\cosh \gamma \ell$ is a real number in this case, thus $\gamma = j\beta$ and

$$\cosh \beta \ell = \cos k_0 \ell - \frac{\omega C_0 Z_0}{2} \sin k_0 \ell \quad (1.34)$$

Fig. 1.27 (b) shows the plot of β versus k_0 . As shown in Fig. 1.27 (b), there are alternative passbands and stopbands as the frequency increases. Due to the stopbands, periodic waveguides and transmission lines have been widely applied in filter designs. Fig. 1.28 (a) shows the 3-D schematic of an implementation of periodic transmission lines proposed in [11]. It is a sinusoidally width-varied microstrip periodic structure. Fig. 1.28 (b) shows the first-order approximation of its unit cell. As can be seen in Fig. 1.28 (b), the first-order approximation is a patch inserted into the microstrip line. In [11], it is reported that this periodic structure shows alternative passbands and stopbands.

Beside the stopbands, periodic waveguides and transmission lines show slow-wave effect at the edge of the stopband. The slow-wave effect of periodic patterned ground structure will be discussed in detail in Section 1.1.1.3. Besides the dispersion relation, the characteristic impedance of the 1-D periodic structure (transmission line or waveguide) can be expressed as below,

$$Z_c = \frac{Z_{11} - Z_{22}}{2} + Z_{12} \sinh \gamma \ell \quad (1.35)$$

If the infinite structure is terminated at the n th terminal plane by a load Z_L , the reflection coefficient at this plane, Γ_n is expressed as,

$$\Gamma_n = \frac{Z_L - Z_c}{Z_L + Z_c} \quad (1.36)$$

The reflection coefficient at the s th terminal plane can be obtained from (1.36) as,

$$\Gamma_s = \Gamma_n e^{2j\beta_c \ell(s-n)} \quad (1.37)$$

where β_c is the imaginary part of γ , ℓ is the period of the structure. An analysis on periodic transmission lines using $ABCD$ matrix is documented in Chapter 8.1 in D. M. Pozar's *Microwave Engineering* [50]. The dispersion relation can be expressed as,

$$\cosh \gamma \ell = \frac{A + D}{2} \quad (1.38)$$

where A and D are the parameters of the $ABCD$ matrix of the unit cell. Equation (1.38) agrees with (1.31) using $[Z]$. The wave impedances for the forward- and backward-waves are expressed as,

$$Z_c^\pm = \frac{\pm B Z_0}{\sqrt{A^2 - 1}} \quad (1.39)$$

Figure 1.27: A periodic capacitively loaded transmission line and its dispersion relation

where Z_0 is the characteristic impedance of the transmission line without any periodic loading. The plus and minus signs correspond to the forward- and the backward-waves. When the periodic structure is terminated with a load, Z_L , the reflection coefficient at the load can be expressed as,

$$\Gamma = \frac{Z_L - Z_c}{Z_L + Z_c} \quad (1.40)$$

which agrees with (1.36).

A planar EBG microstrip structure is one type of periodic transmission lines. It

(a) 3-D schematic

(b) First-order approximated unit cell

Figure 1.28: A sinusoidally width-modulated microstrip periodic structure [11]

is an application of the Bragg reflection condition to a microstrip line with unit cells etched in the ground plane. As the unit cells are simply patches etched in the ground plane, it is compatible with printed circuit board fabrication techniques and it does not require additional circuit area or space. Moreover, due to the periodicity, they show slow-wave effect with which their circuit areas can be reduced. Therefore, this is a worthwhile area to perform systematic studies and development. In the sections that follow, 2-D and 1-D EBG microstrip structures are reviewed in detail.

b. Two Dimensional EBG microstrip Structures

Fig. 1.29 shows the schematic of the first proposed 2-D EBG microstrip structure [12].

Figure 1.29: Schematic of a 2-D planar EBG structure [12]. The rectangular lattice of circles are etched in the ground plane of a microstrip line

As seen in Fig. 1.29, three columns of periodic circles are etched in the ground plane of a microstrip line where the circles form a square lattice. The radius of the etched circle is r . The distance between two successive etched circles along the microstrip line, a , is the period of the structure. With the periodically etched circles, the structure shows a wide stopband, prohibiting the propagation of electromagnetic waves over a wide frequency range. The center of the stopband is determined by the period of the structure satisfying the Bragg reflection condition [40]. In [12], this two dimensional EBG structure is bent 90°. A two dimensional EBG microstrip circuit with a compensated right-angle microstrip bend (Fig. 1.30) was found to have similar bandstop performance to that of a straight EBG microstrip structure. The coupled mode theory was formulated to analyze this two dimensional EBG microstrip structure [51]. The structures with different lattices (square, triangle, and hexagonal lattices) and different geometry of etched patches (square and triangle) are studied in [52]. The study concludes that the effect of the lattice is significant

Figure 1.30: Photograph of a fabricated 2-D planar EBG microstrip circuit with a compensated right-angle microstrip bend [12]

while that of the geometry of the etched patch is minor. For a structure with a fixed lattice, the area of the etched patch instead of the geometry affects the bandwidth of the stopband. In the same paper, the effect of the relative location of the columns to the microstrip line on the wave propagation is also reported. When the column is directly below the microstrip line, the wave propagation is severely perturbed. This is due to the high confinement of electromagnetic fields around the microstrip line.

As a good bandstop filter, the two dimensional EBG microstrip structure with circular etched patches was used to suppress coupling of two close microstrip lines or two intersecting microstrip lines working at different frequencies [53]. It was used as reflectors at two ends of a microstrip line to construct a Bragg resonator [54]. Other than bandstop filters, two dimensional planar EBG structures are applied to the designs of other types of filters. For example, in [13], a tapered two dimensional EBG structure (Fig. 1.31) is used to construct a microstrip low-pass filter with good performance.

Figure 1.31: Schematic of a microstrip low-pass filter employing 2-D planar EBG structure [13]

Besides a wide stopband, a two dimensional EBG microstrip structure in Fig. 1.29 shows slow-wave effect. It can be used as a phase shifter for phased array antenna. In [14], it was found that the electrical delay is linearly related to the number of rows at a fixed frequency in a two dimensional EBG microstrip structure. This means that the structure is suitable to use as a phase shifter [55]. Fig. 1.32 shows the schematic of a four-element patch antenna array proposed in [14] where there is a two dimensional EBG structure incorporated to the input of each antenna as a phase shifter.

An anisotropic 2-D Planar EBG structure [15] is a variant of a two dimensional EBG microstrip structure. In the anisotropic EBG structure, as Fig. 1.33 shows, the etched patches are connected by slits in one direction. Thus, the transmission characteristics of electromagnetic waves are different when the waves are guided in the y - and the z -directions. In Fig. 1.33, in the y -direction, the etched patches are not connected whereas in the z -direction, the patches are connected. The anisotropic characteristics of 2-D planar EBG structures are applied to the design of microstrip diplexer antennas to enhance the isolation between transmitting (T_x) and receiving (R_x) ports [56, 57, 58] and to the design of diplexer filters [59].

A uniplanar compact photonic band-gap (UC-PBG) structure is a special

Figure 1.32: The schematic of a 4-element patch antenna array incorporated with 2-D planar EBG structures [14]

type of two dimensional EBG microstrip structure where the etched pattern in the ground plane is uniquely designed and unit cells are arranged in two dimension [16]. Its design does not obey the Bragg reflection condition. It is named photonic band-gap structure due to its periodicity. Fig. 1.34 (a) and (b) show the top view of the UC-PBG structure proposed in [16] and the details of its unit cell, respectively. As shown in Fig. 1.34 (b), the UC-PBG unit cell consists of a square metal pad with four connecting branches. A UC-PBG structure shows a wide stopband and slow wave effect, which are shown in Fig. 1.35 and Fig. 1.36, respectively. This is similar to a two dimensional EBG microstrip structure with etched circles. Compared with the EBG structures with etched circles, a UC-PBG structure shows a wider stopband, more severe slow wave effect, and is more compact.

The UC-PBG structure has a wide range of applications. It is applied to traditional antenna designs [60] and filter designs [61, 62] to improve their performances. It

Figure 1.33: Ground plane of an anisotropic 2-D planar EBG structure [15]

is applied to both microstrip line structures and coplanar waveguides (CPW) for improving their performance [16]. It is used to build a novel TEM-waveguide which generates a relatively uniform field distribution in the desired frequency band [63] and shows potential applications in optical devices [64] and EMC testing. It is used to be a mode suppressor for waveguides such as non-radiative dielectric (NRD) waveguide in the design of hybrid planar/NRD circuits [65]. It is also used in the design of a nonlinear transmission line (NLTL) frequency multiplier in [66] to improve the conversion efficiency.

Moreover, studies on the dispersion characteristics of the UC-PBG structure are reported in [67, 68] and a proposed circuit model is presented in [69]. Variants of a UC-PBG are reported. A UC-PBG structure is modified based on the original design [16] being proposed to lower the center frequency of the stopband [70] or to achieve multiband structures [71]. Two metal planes with UC-PBG etched patterns are stacked below a microstrip line to construct a multi-layer EBG structure [72,

(a) In the ground plane of a microstrip line

(b) Details of one unit

Figure 1.34: Schematic of the uniplanar compact photonic band-gap (UC-PBG) structure [16]

73]. Compared with a single-layer UC-PBG structure reviewed previously, a multi-layer structure shows a wider stopband (no matter whether the ratio of the periods of different layers are an integer [72, 73]) and a sharper cutoff (when the ratio of the periods are not an integer [73]).

c. One Dimensional EBG microstrip Structures

A one dimensional EBG microstrip structure is a patterned ground structure where there is a single column of etched patterns directly below the microstrip line. Fig. 1.37 shows the photograph of a one dimensional EBG microstrip structure [17] and that of a two dimensional EBG structure reviewed in the previous section [12]

Figure 1.35: FDTD simulation results of S-parameters of the microstrip on the UC-PBG ground plane [16]

for comparison purposes. As shown in Fig. 1.37, the one dimensional structure has a large reduction in physical size in the transversal direction as compared to the two dimensional structure. The one dimensional EBG structure satisfies the Bragg reflection condition [17]. As reported in [17], it shows a wide stopband with high attenuation. The wave propagation of the one dimensional structure is similar to that of the two dimensional structure. It is due to the high confinement of electromagnetic fields around the microstrip line. The one dimensional EBG microstrip structure is able to control the wave propagation effectively while occupying a small circuit area. Its transmission characteristics are analyzed by applying different theories and models such as the fiber grating model [74], the coupled mode theory [75], and the short-open calibration (SOC) de-embedding technique [76].

In order to increase the bandwidth and attenuation of the stopband of a one dimensional EBG microstrip structure, conventionally, the number of unit cells has to be increased. However, this is realized at the expense of a significant increase in the physical size of the structure in the longitudinal direction. Without an increase in the physical size, a couple of novel one dimensional EBG structures based on the

Figure 1.36: Slow wave factor of a microstrip line on the UC-PBG ground plane [16] in comparison with a conventional microstrip line

structure in Fig. 1.37 were proposed with enhanced stopband performance [18, 77]. In [18], a compact EBG microstrip reflector was proposed using a meander microstrip conductor line with periodic etched circles directly below it. Fig. 1.38 shows the schematic of the proposed structure. As can be seen in Fig. 1.38, a large number of unit cells are included in a small circuit area. In this structure, high attenuation and sharp cutoff are both obtained in the stopband due to the large number of EBG unit cells and the circuit area remains small because of

Figure 1.37: Photograph of a 2-D planar EBG microstrip structure (upper) and a 1-D planar EBG microstrip structure (lower) [17]

Figure 1.38: Schematic of a compact 1-D planar EBG microstrip structure with a meander microstrip line [18]

the meander microstrip line. In [77], the high ripples caused by the periodicity of the structure and the turnings of the microstrip line were alleviated by applying tapering.

Instead of increasing the number of unit cells, a one dimensional chirped EBG microstrip structure was proposed in [19] to obtain an increase in the bandwidth of the stopband. In this structure, the chirping technique that was first used in optical fiber Bragg grating and in microwave corrugated waveguides was adopted. Fig. 1.39 shows the schematic of this proposed structure. The period of the structure a_i is given by the following equation as a function of its position in the device,

$$a_i = a_0 \cdot (1 + i\delta) \quad (1.41)$$

where a_0 is the period corresponding to the etched circle at the center of the structure, δ is the dimensional chirp parameter that controls the period variation, and $i = 0, \pm 1, \pm 2, \dots$. The one dimensional chirped EBG microstrip structure exhibits

Figure 1.39: Schematic of a chirped 1-D planar EBG microstrip structure with circles etched in the ground plane [19]

an increased width of the stopband but it also shows an increase in the ripple level within the passband. Tapering techniques were applied to flatten the ripples of the proposed chirped structure.

The one dimensional periodic etched patches are applied to the design of microwave devices in order to optimize their performances. Fig. 1.40 shows the schematic of a microstrip bandpass filter optimized using one dimensional periodic etched patches in the ground plane [20]. Four patches form a group with a period. The two groups in this structure are designed to have different periods for suppressing different higher harmonics. In Fig. 1.41, a tapered and meandered one dimensional EBG structure is employed in the design of diplexer to switch transmit/recieve paths in a multi-band transceiver antenna in [21]. Other than the examples given above, one

Figure 1.40: Layout of microstrip bandpass filter with cascaded 1-D planar EBG structures [20]

Figure 1.41: Schematic of a diplexer with two meandered, tapered 1-D planar EBG microstrip structures [21]

dimensional periodic etched patches are applied to optimize dielectric resonator filters [78], active antennas [79], and coupled ring bandpass filters [80].

For the wide stopband shown by a one dimensional EBG microstrip structure, it is worth mentioning that it can be controlled when optical sensitive materials and lights are working together [81, 82]. The optically tunable and switchable stopband of a one dimensional EBG microstrip structure could be useful in future applications.

1.1.1.3 Slow-wave Effect

Previously in Section 1.1.1.1 and 1.1.1.2, both the patterned ground structure with a single etched pattern in the ground plane of a transmission line and those with periodic etched patterns show slow-wave effect at the edges of the stopband, as shown in Fig. 1.12 in Section 1.1.1.1 and Fig. 1.36 in Section 1.1.1.2, respectively. In the studies of patterned ground structures, a parameter named *slow-wave factor* and defined as in (1.42) has been used to quantify slow-wave effect.

$$\text{Slow wave factor} = \frac{\beta}{k_0} \quad (1.42)$$

where β is the phase constant and $k_0 = \omega\sqrt{\mu_0\epsilon_0}$. An increase in slow-wave factor implies slow-wave effect [5, 83]. For quantifying how fast waves propagation, phase velocity is one of the parameters. In general, for phase velocity of waves at a single frequency, it is expressed as in (1.43),

$$v_p = \frac{\omega}{\beta} \quad (1.43)$$

In free space, as $k_0 = \omega\sqrt{\mu_0\epsilon_0} = \omega/c$ where c is the speed of light in the air, (1.43) can be expressed as,

$$v_p = c \frac{k_0}{\beta} \quad (1.44)$$

Based on (1.44), the slow-wave factor defined in (1.42) can be linked to the phase velocity as,

$$\text{Slow wave factor} = \frac{c}{v_p} \quad (1.45)$$

Thus, a small v_p corresponds to a large slow-wave factor defined in (1.42), indicating slow-wave effect.

It is worth mentioning here that the slow-wave factor used in the literature for patterned ground studies is at a single frequency. For light waves where a frequency

Figure 1.42: The equivalent circuits of (a) a microstrip line, (b) a patterned ground structure in Fig. 1.1 using R , L , and C , and (c) the patterned ground structure using R' and X'

band is usually of interest in the studies of photonic crystals, the group velocity of waves is an appropriate parameter to quantify the slow-wave effect [84].

For a patterned ground structure with a single etched patterned, the slow-wave effect can be explained from a circuit point of view. The simplest patterned ground microstrip structure as shown in Fig. 1.1 is used as an example to explain this effect. Fig. 1.42 (a) shows the equivalent circuit of a microstrip line. The characteristic impedance, Z_c , and phase velocity, V_p , can be expressed using (1.46) and (1.47) as follows [85], respectively,

$$Z_c = \sqrt{\frac{L_m}{C_m}} \quad (1.46)$$

$$v_p = \frac{1}{\sqrt{L_m C_m}} \quad (1.47)$$

where L_m and C_m are the equivalent inductance and capacitance of the microstrip line as shown in Fig. 1.42 (a). Thus, based on (1.45) and (1.47), the slow-wave factor of a microstrip line can be estimated as,

$$\text{Slow wave factor} = c\sqrt{L_m C_m} \quad (1.48)$$

When an etched pattern is introduced to the microstrip line, for example, a slot as shown in Fig. 1.1, a parallel RLC circuit is introduced in series to the equivalent inductor as shown in Fig. 1.42 (b). As presented in Section 1.1.1.1a, the input impedance of the RLC circuit is Z where $Z = R' + jX'$. The resistance R' and reac-

tance X' are related to R , L , and C using (1.2) and (1.3), respectively. Fig. 1.42 (c) shows the equivalent circuit of the patterned ground structure using $Z = R' + iX'$. Therefore, for the patterned ground structure, the characteristic impedance, Z'_c , phase velocity, V'_p , and the slow-wave factor can be expressed as,

$$Z'_c = \sqrt{\frac{L_m + X'}{C_m}} \quad (1.49)$$

$$V'_p = \frac{1}{\sqrt{(L_m + X')C_m}} \quad (1.50)$$

$$\text{Slow wave factor} = c\sqrt{(L_m + X')C_m} \quad (1.51)$$

As X' is dispersive as shown in Fig. 1.3 (b), the characteristic impedance, the phase velocity, and the slow-wave effect of the patterned ground structure are dispersive.

As shown in Fig. 1.3 (b), X' varies from positive to negative as the frequency increases and it equals to zero at the resonant frequency. Therefore, before the resonance, the slow-wave factor is larger than that of a microstrip line based on (1.51) and the structure exhibits slow-wave effect. The slow-wave factor of the pattern ground structure reaches a maximum when X' reaches a maximum before the resonance, which agrees with the results presented in Fig. 1.12.

For patterned ground structures with periodic etched patterns, the slow-wave effect can be explained using a plot of the dispersion relation. The periodic capacitively loaded transmission line in Fig. 1.27 (a) is used as an example for explanations. Fig. 1.43 shows the first quarter of the plot of the dispersion relation of this structure.

The phase velocity of waves at a specific frequency is the slope of a line connecting the origin and the corresponding point on the curve. As shown in Fig. 1.43, the slopes of the blue, green, and red lines, v_{p1} , v_{p2} , v_{p3} are the phase velocities at frequencies at point 1, 2, and 3, respectively. As shown in Fig. 1.43, as the frequency

Figure 1.43: The dispersion relation of the periodic capacitively loaded transmission line in Fig. 1.27 (a).

approaches the edge of the stopband, v_p decreases and the slow-wave factor increases based on (1.45). Therefore, at the edge of the stopband, there is slow-wave effect. This conclusion agrees well with the results shown in Fig. 1.36.

A 1-D periodic transmission line shows slow-wave effect at the edge of the stopband that are quantified using slow-wave factors. Periodic structures working at optical frequencies, such as photonic crystals, show similar effects at the edge of the stopband. They are called slow lights and are explained using zero group velocity at the edge of the stopband. In [86], slow light was demonstrated near the band edge in the passband using simple 1-D periodic gratings.

As reviewed, microstrip patterned ground structures show advantages of a designable stopband, slow-wave effect that results in small circuit area, easy fabrica-

tions, and compatibility with monolithic microwave circuits. In patterned ground structures, the etched patterns in the ground plane introduce additional useful features to a microstrip line. However, they cause EMC problems when a structure is integrated into a microwave circuit. Due to the apertures, a microstrip structure with an imperfect ground plane is more susceptible to interference compared to one with a perfect ground plane. On the other hand, they generate additional radiation through the apertures that can interfere with nearby circuit components.

1.1.2 EMC & EMI of Patterned Ground Structures

In the literature, in order to implement a patterned ground structure without affecting other circuit components in a microwave circuit or to ensure that other circuit components do not affect the implemented patterned ground structure, studies have been conducted on the effects of shielding on a patterned ground structure. In [87], it is reported that the scattering parameters of a 2-D EBG microstrip filter are influenced by a uniform shielding plate at close proximity to the structure.

Fig. 1.44 shows a 3-D view of a shielded patterned ground structure with periodic dumbbell-shaped etched patterns in the ground plane [22]. In order to shield the patterned ground structure, metal plates are placed above and below the structure and the size of the ground plane is limited. In [22], the effect of metal plates in close proximity above and below the structure as well as that of a finite ground plane on the performance of the structure were studied in detail. It was reported that the metal plates close to the structure and a finite ground plane can severely affect the frequency responses of the structure. To eliminate the effect of a finite ground, W should be larger than $5a$ as shown in Fig. 1.44 whereas to have a negligible effect of the upper and the lower metal plates, h_1 and h_2 should be at least three times and six times of the thickness of the substrate, respectively.

Figure 1.44: A 3-D view of a shielded patterned ground structure with periodic dumbbell-shaped etched patterns in the ground plane [22]

In the literature, not many papers have reported on the electromagnetic compatibility and electromagnetic interference of patterned ground structures. In [22], the use of a metallic enclosure with finite ground plane was suggested for the shielding of a patterned ground structure from interference, which makes the circuit bulky and complex. Integrations of patterned ground structures into microwave circuits without metallic enclosures are practical approaches. The knowledge on the EMC and EMI of patterned ground structures, such as the EMS of the structures due to interferences nearby and the interferences generated by the structures, is important and yet remains unexplored.

In this thesis, the EMC and EMI of patterned ground structures are to be studied in detail by examining the effect of a nearby circuit component on the performance of the structures and by examining the radiation of the structures. The study is useful to understand the effects of radiation and coupling of circuit components close by on these patterned ground structures as well as how the structures radiate. The knowledge obtained can be used to guide the integrations of patterned ground

Figure 1.45: Single Ag nanoparticles imaged with evanescent-wave excitation [23]

structures into microwave integrated circuits.

Besides the effects on the propagation, coupling, and radiation of electromagnetic waves, the effects of metallic layers on the scattering of waves are examined by studying SERS substrates. The progress in the research on SERS is reviewed in the next section.

1.1.3 Surface-enhanced Raman Scattering

SERS was first observed by *Fleischman* et al. in 1974 [88]. The unusual intensity of the Raman signals was attributed to the increased surface area of the rough substrate. It was discovered by *Jeanmaire* and *Van Duyne* [89] and *Albrecht* and *Creighton* [90] in 1977. Van Duyne and Creighton pointed out that the intensification of the effective Raman cross section was far in excess of the increase in the number of molecules interrogated as a results of surface roughness.

Twenty years after the discovery of SERS, strong scientific interest in SERS has been renewed since 1997 when it was reported that single molecules absorbed on

Figure 1.46: E-field enhancement contours of a “tip-to-side” prism dimer at 852 nm [24]

single nanoparticles can be detected using a methodology based on SERS [23, 91]. In [23], it was reported that the intrinsic Raman enhancement factor was of the order 10^{14} to 10^{15} . In the experiment, each Ag particle carries an average of one rhodamine 6G (R6G) molecule at a R6G concentration of 2×10^{-10} mol/L. Fig. 1.45 (a) and (b) show an unfiltered and a filtered photographs of single Ag nanoparticles imaged with evanescent-wave excitation, respectively, when the concentration of R6G is 2×10^{-10} mol/L. Fig. 1.45 (a) shows background scattered laser light from all particles immobilized on a polylysine-coated surface. When the laser scattering background was filtered using a high-performance bandpass filter, as shown in Fig. 1.45 (b), the bright dots (also called hot particles) are the Raman scattering particles. The number of hot particles increases as the concentration of R6G increase, but the relation is not linear. The difference in color between the laser scattering and the Raman scattering shows the Stokes-shift of SERS (toward longer wavelengths).

The ability to detect single molecules makes SERS important in many fields, such as chemistry, biology, medicine, pharmacology, and environmental science. Nu-

Figure 1.47: A nanocrescent [25]

merous studies are dedicated to model and analyze the giant SERS enhancement factor obtained in the experiment (10^{14} – 10^{15} [23]). New geometries of nanoparticles have been proposed to obtain a high SERS enhancement factor. For example, two “tip-to-side” prisms with a gap size of 2 nm as shown in Fig. 1.46 shows a simulated electric field (E-field) enhancement factor of 239 at 852 nm [24] while a crescent-shaped nano-cylinder (called nanocrescent) in Fig. 1.47 (a) shows a simulated E-field enhancement factor of 211 at 785 nm for a horizontal polarization [25]. Fig. 1.47 (b) shows the E-field contours of a nanocrescent at 785 nm for a parallel polarization. On the other hand, progress in nanofabrication techniques [92, 93] enables the practical realizations of SERS-active substrates.

In SERS, the Raman enhancement is attributed to two main mechanisms, namely

Figure 1.48: SEM images of fabricated Au nanowells observed at a tilt angle of 60° [26] where D_o and D_i are the inner and outer diameter of the nanowells and P is the pitch of the array.

electromagnetic field enhancement and chemical enhancement, with the electromagnetic mechanism providing the dominant contribution to the SERS effect. The E-fields at the surface of nanoparticles are enhanced by a large factor mainly due to the excitation of surface plasmons at the incident frequency, w_i . The enhanced E-fields is localized near the nanoparticles. They induce dipoles in the molecules adsorbed at the surface of the particle. The induced dipoles are oscillating and excite plasmons in the particles at the Stokes-shifted frequency (also called Raman scattered frequency in some articles), w_s [23, 94]. With the mechanisms above, the SERS enhancement factor is proportional to $|\bar{E}(w_i)/\bar{E}_i(w_i)|^2 |\bar{E}(w_s)/\bar{E}_i(w_i)|^2$ where $\bar{E}(w_i)$ and $\bar{E}(w_s)$ are the average E-field at the particle surface at w_i and w_s , respectively [95]. When the enhancements at the two frequencies are close to each other, the enhancement factor is approximately equal to $|\bar{E}(w_i)/\bar{E}_i(w_i)|^4$.

The enhancement of E-fields, $\bar{E}(w_i)/\bar{E}_i(w_i)$, is directly related to the SERS enhancement factor, $|\bar{E}(w_i)/\bar{E}_i(w_i)|^2 |\bar{E}(w_s)/\bar{E}_i(w_i)|^2$, in SERS experiments. Theoretical analysis of the E-fields around the nanoparticles are useful in providing basic guidance to obtain a high Raman enhancement factor. For an isolated spherical nanoparticle, Mie theory provides an analytical solution. Numerically, finite-difference time-domain (FDTD) methods and other computational electrodynamic

(a) $D_o = 180\text{nm}$, $D_i = 50\text{nm}$, $P = 600\text{nm}$ (b) $D_o = 375\text{nm}$, $D_i = 239\text{nm}$, $P = 500\text{nm}$ (c) $D_o = 492\text{nm}$, $D_i = 320\text{nm}$, $P = 620\text{nm}$

Figure 1.49: Simulated scalar cut planes (the left column) and isocurves (the right column) of E-field enhancement at 785 nm of nanowells of different dimensions, the arrow at the left top corner indicates the polarization of E-fields [26].

Figure 1.50: Calculated local fields (absolute value relative to that of the excitation field) in the equatorial plane of symmetry for the linear self-similar chain of three silver nanospheres, R_i is the radius of the i^{th} sphere and $d_{i,i+1}$ is the distance between the surfaces of the i^{th} and $(i+1)^{\text{th}}$ spheres. Inset: the geometry of the system in the cross section through the equatorial plane of symmetry [27].

methods such as discrete dipole approximation (DDA) [96] are available to determine the local E-fields by solving Maxwell's equations.

Numerous studies on the enhancement of E-fields for Raman-active substrates have been reported [24, 26, 94, 27]. The enhancement of E-fields is dependent on a number of factors. One is the shape/geometry of the particles and their physical size with respect to the incident wavelength. Besides nanocrescents as shown in Fig. 1.47, nanowells were proposed showing field enhancement at both visible and near-infrared regime [26]. Fig. 1.48 shows scanning electron microscope (SEM) images of the fabricated nanowells. They are confirmed to be SERS active by SERS spectra of both R6G and phthalazine molecules adsorbed on the nanowells. Fig. 1.49 shows the simulated scalar cut planes and isocurves of E-field enhancement of nanowells with different dimensions at 785 nm. As shown in Fig. 1.49, the maximum E-field changes when the physical size of the nanowells changes. When the outer diameter, D_0 , the inner diameter, D_i , and the pitch of the array, P , are 180 nm, 50 nm, and 600 nm, respectively, the nanostructure shows a maximum E-field enhancement factor of 269 which is comparable to that of a nanocrescent in [25].

The enhancement of E-fields is also affected by the arrangement of particles, such as the number of particles, the periodicity, and the space between the particles. A dimer as shown in Fig. 1.46 which has two particles in close proximity shows a much higher enhancement factor than an isolated particle [24]. An array of particles obtain an extra factor of 10^2 due to long-range photonic interactions caused by the periodicity [94]. The space between two particles plays an important role in the enhancement [27]. Fig. 1.50 (a) and (b) shows the calculated E-field enhancement of a self-similar chain of three silver nanospheres at different distances between successive spheres. As shown in Fig. 1.50, the enhancement factor jumps to as high as 1300 at 3.25 eV when the distance decreases.

Moreover, the substrate supporting the particles shows effects on the enhancement of E-fields [97]. The effect of the substrate in SERS supporting the nanoparticles can be approximated numerically using DDA [96]. However, many of the numerical results calculate only localized fields near the surface of isolated particles, where the particles are assumed to be surrounded by homogeneous dielectric medium, without considering the effect of the substrate [24, 26]. There is an exception [97]. In [97], the effect of the substrate is studied experimentally and numerically. This study was conducted by changing substrates of different refractive indices. The effect of the refractive index of the substrate is reported in detail.

In most of the recent SERS experiments, metal particles are immobilized on a dielectric substrate for the detection of scattered SERS signals [23, 98]. The effect of a substrate on the enhancement of E-fields is important to guide the constructions of substrates in the experiments. However, it is not yet quantified rigorously. In this thesis, an analytic method is presented to study the effects of a substrate on the enhancement of E-fields in SERS. Based on this method, a metal ground plane is proposed to be incorporated to the back of a substrate, which leads to a significant increase in the Raman enhancement factor.

1.2 Motivation

As reviewed, a patterned ground structure with an etched patch or etched patches in the ground plane shows a bandstop phenomenon. They therefore provide filtering functionality without requiring additional circuit area. Patterned ground structures with a single etched unit in the ground plane are bandstop resonators [3, 4, 6, 7, 9, 28, 36, 37] which show narrow stopbands. Patterned ground structures with more than one etched unit arranged periodically along the microstrip line (EBG structures) show a wide stopband with high attenuation. They can work

as alternatives to conventional microwave planar filters to meet the stringent new requirements in modern communication systems. They can also be applied to optimizations of microwave devices, such as amplifiers [34]. Besides stopband phenomena, patterned ground structures exhibit slow wave effect [5, 16] with which they occupy small circuit area. Moreover, microstrip patterned ground structures exhibit advantages of ease of fabrication and compatibility with monolithic microwave integrated circuits.

In the research and development of patterned ground structures in the literature, the potential to increase the quality factor and to reduce the physical size of the structure can be further explored to meet new requirements in different communication systems. An etched pattern or etched patterns can be applied to optimize other existing structures with performance limitations besides the reported ones. In this part of research, the effects of a patterned ground plane on the propagation of microwaves are studied.

As the study of patterned ground structures unfolds their potential in the applications to microwave circuits, the EMC and EMI are important issues that need to be addressed. When a patterned ground structure is integrated into a microwave circuit, due to the apertures, the structure with an imperfect ground plane is more susceptible to interference from nearby circuit components compared to a structure with a perfect ground plane. Similarly, patterned ground structures generate radiation through the apertures that can interfere with nearby circuit components. The studies of the EMC and electromagnetic susceptibility (EMS) of patterned ground structures are important to the integrations of these structures into microwave circuits. The EMC and EMS issues are investigated by studying the effects of a patterned ground plane on the coupling and radiation of electromagnetic waves.

For the effect of a metallic layer (a ground plane) on scattering of electromagnetic waves, studies on SERS are reviewed. Based on the review on SERS, a large variety

of structures are observed or predicted to have a large Raman enhancement factor. The enhancement of the electric field is both affected locally by the geometry of the nanoparticles and affected globally by the particles in the neighborhood or by the substrate supporting the particles. In most of the SERS experiments, there are effects of supporting substrates on the nanoparticles since metal particles are immobilized on a dielectric substrate for the detection of scattered SERS signals [23, 98]. Reported numerical results calculate only localized fields near the surface of isolated particles, where the particles are assumed to be surrounded by homogeneous dielectric medium, without considering the effect of the substrate [24, 26], although the effect of the substrate supporting the nanoparticles can be approximated numerically using discrete dipole approximation (DDA) [96]. A substrate supporting metallic nanoparticles could consist of different dielectric and metallic layers. The effect of a substrate is yet to be studied analytically. The effect of a metallic layer becomes complicated at optical frequencies. Studies of this effect will help us to obtain fundamental understanding on the behaviors of waves in the substrate and on the surface of nanoparticles, and in turn, provide guidance on the construction of a substrate in order to obtain a high Raman enhancement factor.

1.3 Objectives

This thesis aims to study the effects of ground planes on propagation, coupling, radiation, and scattering of electromagnetic waves. The study on the effects on the former three behaviors of electromagnetic waves is to be done by investigating patterned ground structures at microwave frequencies. In view of the recent development of patterned ground structures exhibiting bandstop phenomena and slow wave effect, the aims of this part of the thesis are two folds. One is to explore possibilities for new designs with a high Q-factor and a small physical size, to apply

ground plane with etched patterns to existing structures for possible optimizations, and to do circuit modeling of the aforementioned to facilitate computer-aided designs. The other aim is to study the EMC and EMS of patterned ground structures. The EMC and EMS studies are to quantify the electromagnetic effects of nearby components to patterned ground structures and the electromagnetic effects of patterned ground structures to the components in close proximity.

The study on the effects of ground planes on the scattering of electromagnetic waves is completed by working on optical structures. With the renewed interest in SERS, in optical frequency range, the author aims to conduct an analytic study to examine the effect of a ground plane in a SERS substrate on the E-fields available to the nanoparticle surface. Through this study, the understanding on the scattering of electromagnetic waves in a layered substrate by a metallic layer at optical frequencies is to be enhanced. It will be useful to guide the construction of a substrate in SERS experiments in terms of material selection and thickness of each layer, so as to obtain a high enhancement factor.

1.4 Major Contributions of the Thesis

In this thesis, the proposed objectives have been achieved. Both patterned ground structures and SERS are studied intensively. Several important contributions have been made:

1. A high performance E-shaped patterned ground structure has been proposed [A 5]. Its circuit model has been proposed to facilitate its compatibility with computer-aided designs. It has been applied to design a compact tunable bandstop filter.
2. A dual planar EBG configuration has been proposed where periodic patches are inserted into a microstrip line and periodic patches are etched in the

ground plane [A 8]. The stopband is widened and the attenuation is increased without increasing the circuit area. The circuit area of the structure is further reduced by adopting a U-shape configuration [A 7]. The significant reduction in physical size is obtained when the performance in both the passband and the stopband are maintained.

3. A fast and accurate calculation of the transmission coefficients of a periodically capacitively modulated microstrip EBG structure has been proposed [A 2]. The calculation is an alternative to full-wave calculations that consume a considerable amount of time. This calculation helps to shorten the computational time of the computer-aided designs of this type of periodic structures.
4. A microstrip line with an aperture in the ground plane has been applied to a conventional stepped-impedance low-pass filter design to optimize its performance [A 6]. Compared with the conventional filter before optimization, the optimized stepped-impedance low-pass filter shows a sharper cutoff, higher attenuation in the stopband, a better suppression of spurious passbands, and a smaller circuit area.
5. The EMC and EMS of patterned ground structures have been studied by examining the effect of a nearby component on the performances of the structures. A radiating component [A 4] and a microstrip line [A 1] are both considered as the possible circuit components that interfere with a patterned ground structure.
6. The EMC of patterned ground structures has been studied by examining their radiation [A 10].
7. An analytic method has been proposed to study the effect of a substrate bottomed with a metallic ground layer on the E-fields available on the nanopar-

title surface in SERS [A 3].

1.5 Organization of the Thesis

Chapter 1 provides an overview of the thesis and a detailed literature review on the progress in the research and development of both patterned ground structures and SERS. It gives the motivations of the research work for the author's thesis based on the review, introduces the objectives, and lists the main contributions in her studies.

In **Chapters 2–5**, the effects of ground planes on electromagnetic waves, in terms of propagation, coupling, radiation, and scattering, are presented. **Chapters 2 and 3** report the effects of ground plane geometries on the propagation of waves. New designs of patterned ground structures are presented in Chapter 2. This includes structures with single etched patterns and periodic structures. In this chapter, the design and circuit modeling of an E-shaped patterned ground structure and its application to a tunable bandstop filter are discussed. It is followed by the design and study of an EBG dual planar configuration where periodic units are introduced to both the microstrip line and the ground plane along the microstrip line. Based on the design of a straight dual planar EBG microstrip structure, a U-shaped EBG microstrip structure are designed and studied. Chapter 3 consists of two parts. The first part reports a fast and accurate calculation of transmission coefficients for a periodically capacitively modulated microstrip EBG structure. The second part presents the application of a patterned ground plane to optimize the performance of a conventional stepped-impedance low-pass filter and the associated analysis.

Chapter 4 reports the effects of ground plane geometries on the coupling and radiation of electromagnetic waves. The studies on the EMC and EMS of patterned ground structures are presented. **Chapter 4** reports the effects of a radiating

component and that of a microstrip line on the performance of a patterned ground structure. It also details the studies of the radiation of patterned ground structures.

Chapter 5 presents an analytical method for studying the effect of a substrate with a metallic ground plane in SERS at optical frequencies.

The thesis is summarized in **Chapter 6**. Recommendations for future work are suggested in the same chapter.

Chapter 2

Effects of Ground Plane Geometries on Wave Propagation I

In this chapter and the next chapter, the study of the effects of etched patterns in the ground plane on the propagation of waves through a microstrip line is presented. In this chapter, the effects are illustrated using three new structures, a patterned ground microstrip structure with a single E-shaped etched pattern, a dual planar EBG microstrip filters, and a slot-back miniaturized stepped-impedance low-pass filter. The designs and the modeling of the structures and the analysis of their performance are presented. A fast and accurate calculation of the transmission coefficients for an EBG microstrip structure is included.

2.1 E-shaped Pattern & Tunable Bandstop Filter

In this section, an E-shaped patterned ground microstrip bandstop resonant cell is proposed. It demonstrates the significant effect of a single etched pattern on the propagation of waves through a microstrip line. In comparison to other bandstop resonators, it provides the advantages of a high Q-factor (or in other words, a sharp transition knee) within a small circuit area. This resonant cell is applied to the design of a tunable bandstop resonator. Due to the unique E-shaped pattern etched in the ground plane, the proposed resonant cell provides tuning function within a small circuit area.

Fig. 2.1 (a) to (e) show the schematics of the patterned ground structures in the literature that are reviewed in Chapter 1. The dumbbell-shaped cell [3] in Fig. 2.1 (a) is one of the early designs of patterned ground structures. Based on this structure, many other modified and new patterned ground structures with improved performances were proposed [4, 6, 37]. The development of patterned ground structure is focused on the reduction in the overall physical size of the structure and the improvement in the sharpness of the transition knee (an increase in the Q-factor). As introduced in Section 2.1.1, an unloaded Q-factor is a Q-factor when the circuit has no external circuits. It is an important parameter that can be used to evaluate the intrinsic loss of a resonant circuit [44]. The shape of the etched patch in a dumbbell-shaped patterned ground structure was changed from square to arrow-shape or circle [37]. The change in the shape of the etched patch has minor effects on the performance or on the physical size of dumbbell-shaped patterned ground structure. The performance was improved significantly in terms of the sharpness of the transition knee by the introduction of a spiral-shaped patterned ground structure [4] shown in Fig. 2.1 (b). In [6], a double-H-shaped resonant cell

Figure 2.1: Patterned ground microstrip bandstop resonator (a) dumbbell-shaped (Jan. 2001) [3], (b) spiral-shaped (Jun. 2002) [4], (c) double-H-shaped (Feb. 2006) [6], (d) modified dumbbell-shaped (Sept. 2006) [28], (e) T-shaped (Sept. 2007) [9], and (f) E-shaped.

Figure 2.2: The U-slot patterned ground microstrip bandstop resonator [7].

[Fig. 2.1 (c)] was proposed with both a significant reduction in the physical size and enhancement of the sharpness of the transition knee as compared to a dumbbell-shaped cell [37]. In [28], the conventional dumbbell-shaped patterned ground structure was further modified by inserting a patch in the middle of each etched patch [Fig. 2.1 (d)]. The modified structure shows an increase in capacitance, resulting in a reduction in circuit area. Moreover, due to the modification, the tunability of the structure becomes possible by inserting varactors between the inserted patch and the ground plane. A frequency tuning range of 19 % was demonstrated in [36]. In [9], a T-shaped patterned ground structure [Fig. 2.1 (e)] was proposed. A significant improvement in performance and a large reduction in size were both achieved. In this structure, the slot of the dumbbell-shaped patterned ground structure in Fig. 2.1 (a) was moved to the edges of the etched patch with arms and it was lengthened, forming a T-shape. Among the patterned ground structures with a fixed resonant frequency that were compared in [9], this structure has the smallest physical size and the sharpest transition knees. With the lengthened slot, it becomes possible to embed a varactor into the arms so as to allow for the tuning of the resonant frequency. A maximum frequency tuning range of 13.2% was realized. Based on the T-shaped patterned ground structure in [9], we proposed an E-shaped patterned ground structure [Fig. 2.1 (f)] to further improve the sharpness of the

transition knee and to reduce the circuit area. In order to miniaturize the proposed E-shaped structure further, the design concept of a U-slot patterned ground structure [7] was adopted. Fig. 2.2 shows a U-slot patterned ground structure proposed in [7]. As can be seen in Fig. 2.2, a U-slot cell consists of three connected slot lines in the ground plane below the microstrip line, forming a U-shaped slot. It is compact due to the unique U-shape configuration. For all six structures shown in Fig. 2.1, only in the T-shaped and the proposed E-shaped structures can a geometry similar to the U-slot structure be adopted.

In this section, the design, working principles, and circuit modeling of the E-shaped patterned ground structure are presented. In the E-shaped structure, the geometry of the U-slot structure was adopted. This structure outperforms all the patterned ground structure reviewed in the literature in terms of physical size, and unloaded Q-factor. Furthermore, the proposed structure was applied to realize a compact tunable bandstop resonator.

2.1.1 Design of the E-shaped Resonator

2.1.1.1 Design

In a dumbbell-shaped bandstop resonator, the size of the two etched rectangular patches has to be increased in order to obtain a decrease in the resonant frequency. However, this results in an overall increase in the circuit area of the cell. It has always been a challenge for a patterned ground structure to obtain a low resonant frequency and a small circuit area simultaneously. In order to obtain a compact resonator at a low frequency, a spiral-shaped etched pattern [see Fig. 2.1 (b)] is introduced in the etched patch of the dumbbell [4]. With the spiral-shaped etched pattern, the size of the cell can effectively be reduced when the resonant frequency is fixed. Similarly, a double-H-shaped resonator [6] [see Fig. 2.1 (c)] or a modified dumbbell-shaped structure (with an inserted patch in the middle of each etched

(a) The T-shaped structure [9]

(b) The proposed E-shaped structure

Figure 2.3: Slotted ground microstrip bandstop resonators adopted U-slot geometries

patch) [28] [see Fig. 2.1 (d)] can provide a reduction in circuit area of the cell at a fixed resonant frequency. Keeping the resonant frequency constant, the cell size can further be decreased by using a T-shaped etched pattern with a lengthened slot [9] [see Fig. 2.1 (e)]. The report on a U-slot structure [7] presents a big progress in the development of patterned ground structures in terms of circuit miniaturization. As shown in Fig. 2.2, a U-slot structure has two slot lines in the direction of the transmission line, keeping the structure compact. In a U-slot structure, rectangular etched patches are not required and thus the structure is more compact than the other dumbbell-shaped structures. Of the five structures discussed, only the T-shaped structure can be tailored to be of similar dimension to the U-slot structure, as shown in Fig. 2.3 (a). Moreover, a T-shaped structure is more compact than a

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Table 2.1: Dimensions of Patterned Ground Microstrip Bandstop Resonators in Fig. 2.1 and Fig. 2.2 Resonating at 5 GHz (unit: mm)

Figure	Dimensions
Fig. 2.1 (a)	$w = 1.14, a = b = 10.00, g = 0.25$
Fig. 2.1 (b)	$w = 1.14, a = b = 2.90, g = 0.25, t = 0.30$
Fig. 2.1 (c)	$w = 1.14, D_1 = 0.50, D_2 = 2.65, L_1 = 4.08, L_2 = 2.72$
Fig. 2.1 (d)	$w = 1.14, a = b = 8.20, g = 0.25, c = d = 7.70$
Fig. 2.2	$w = x = 1.40, l = 11.30, g = 0.25$
Fig. 2.3 (a)	$w = x = 1.40, a = 11.00, b = 0.80,$ $g = f = 0.25, t = t' = 0.30$
Fig. 2.3 (b)	$w = x = 1.40, a = 5.20, b = 0.80$ $g_h = g_v = 0.25, t = t' = 0.30, l_a = 4.40$

U-slot structure when they are designed to resonate at the same frequency [9].

The E-shaped patterned ground microstrip resonator proposed in this thesis is shown in Fig. 2.1 (f). As shown, the lengthened slot in a T-shaped structure is further extended to the two edges of the etched patches that are in parallel to the microstrip line, forming the E-shaped structure. The dimensions of this E-shaped structure are as labeled in Fig. 2.1 (f). The width of the the metal strip below the microstrip line (central ground line), t' , is reduced to enhance the inductive effect. Similar to the T-shaped structure, the proposed E-shaped structure can be further miniaturized by adopting a U-slot geometry, as shown in Fig. 2.3 (b). In Fig. 2.3 (b), x of the E-shaped structure is set to be as large as the width of the microstrip line, w , and t' is tailored.

The seven resonant cells from Fig. 2.1 (a) to (d), Fig. 2.2, and Fig. 2.3 (a) to (b) were designed to resonate at 5 GHz. Taconic ($\epsilon_r = 2.45$, $h = 0.4$ mm) was used as the substrate. The dimensions of the structures are tabulated in Table 2.1.

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For structures in Fig. 2.1 (a)-(d), $w = 1.14$ mm corresponding to a characteristic impedance of $50\ \Omega$ at 5 GHz; For those in Fig. 2.2, Fig. 2.3 (a), and (b), the width of the microstrip line was set to be 1.40 mm in order to accommodate the arms of the proposed E-shaped structure. It is tapered to 1.14 mm at two ends for a load matching. The width x in the U-slot, T-shaped, and E-shaped structure was set to be 1.40 mm, $x = w$. All these structures were simulated using a full-wave electromagnetic simulator, Agilent's Advanced Design System [30]. They are compared in terms of physical size, frequency response, and Q-factor.

2.1.1.2 Physical Size

Fig. 2.4 (a) and (b) show the S_{21} -parameters of these resonant cells designed to resonate at 5 GHz. As can be seen in Fig. 2.4, all the structures resonate at 5 GHz according to the design specifications. Table 2.2 tabulates the length, the width, and the area of the etched patches of these patterned ground structures in column two to four, respectively. The fifth column shows the area of the proposed E-shaped cell over that of the other resonant cells in percentage. The last column of Table 2.2 lists the calculated unloaded Q-factors of the structures which will be discussed shortly. As shown in Table 2.2, with the same resonant frequency, the proposed E-shaped cell has an etched area of 9.9 mm^2 . It's only 4.68 % of the dumbbell-shaped structure and less than half of the size of the most recently proposed T-shaped structure (47.37 %). The data imply that the proposed E-shaped structure has the smallest physical etch area in all the resonant structures under this comparison when they are designed to resonate at 5 GHz.

2.1.1.3 Frequency Response

The frequency responses of the patterned ground structures resonating at 5 GHz are compared in terms of sharpness of the transition knee, attenuation at the resonant

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(a)

(b)

Figure 2.4: S_{21} -parameters of the patterned ground structures when they are designed to resonate at $f_0 = 5$ GHz.

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Table 2.2: Physical Size of the Etched Patches of the Patterned Ground Microstrip Bandstop Resonators at $f_0 = 5$ GHz

Resonator Cell	Width (mm)	Length (mm)	Area (mm ²)	E-shaped /other cell (%)	Q-factor
Dumbbell	21.1	10.0	211.4	4.68	4.3
Spiral-shaped	6.9	2.9	20.1	49.7	144.1
Double-H	14.7	6.8	100.2	9.88	28.0
M. Dumbbell	17.5	8.2	143.9	6.88	8.3
U-slot	1.9	11.3	21.5	46.05	557.6
T-shaped	1.9	11	20.9	47.37	639.4
E-shaped	1.9	5.2	9.9	100	714.6

frequency, and insertion loss within the passband. In general, a bandstop resonator is used for harmonic suppression. When it comes to harmonic suppression, a sharp transition knee, high resonant attenuation, and small insertion loss in both lower and higher passbands are desirable. When a bandstop resonator is applied to suppress harmonics, the sharper the transition knee the wider the passbands are and the less the insertion loss it introduces to the signals at the frequencies in the neighborhood in the passbands. High attenuation is to provide effective suppression of unwanted frequencies and small insertion loss within the passbands is to minimize the loss of the wanted frequencies.

In Fig. 2.4 (a) and (b), it can be seen that the resonant cells with the same resonant frequency exhibit different sharpness of the transition knees. Comparing Fig. 2.4 (a) and (b), among all the structures, the conventional dumbbell-shaped cell has the flattest transition knee. The proposed E-shaped cell has the sharpest lower transition knee and a similar upper transition knee to the spiral-shaped cell. The E-shaped cell has the best performance in terms of transition knee. In terms of

Figure 2.5: The equivalent circuit model for the E-shaped bandstop resonator.

attenuation at the resonant frequency, the U-slot cell has the highest attenuation of -34 dB followed by the double-H-shaped cell with an attenuation of -31 dB and the T-shaped cell with an attenuation of -30 dB. The proposed E-shaped structure exhibits an attenuation of -26 dB which is the fourth highest attenuation in all the structures. Insertion loss in the lower passband is dependent on the lower transition knees whereas that in the upper passband is affected by the upper transition knee and the second resonant frequency (how close it is to the first resonance) of the structure. As shown in Fig. 2.4, the dumbbell-shaped cell shows an insertion loss higher than -10 dB in the upper passband. The spiral-shaped and the modified dumbbell-shaped cell exhibit the second resonance that is close to 11 GHz. The U-slot, the double-H-shaped, and the T-shaped cell have a decreasing insertion loss in the upper passband while the proposed E-shaped structure shows a constant insertion loss at around -1.3 dB. Compared to other patterned ground structures, the E-shaped structure shows the best transition knee performance within the smallest circuit area.

2.1.1.4 Q-Factor and Explanations

The unloaded Q-factors of the patterned ground structures resonating at 5 GHz were calculated based on simulation results and compared. The circuit model of the patterned ground structures in Fig. 2.5 was used for the estimations of the Q-factors [7]. The capacitive and inductive components are used to model the

resonance of the cell whereas the resistive component is used to model the conduction, dielectric, and radiation loss [99, p. 820-821]. The model works accurately at frequencies from the cutoff to the resonant frequencies. The parasitic effects of the structures could be modeled using additional circuit components but it is not the focus of the study here. This model does not provide phase information of the structures, but it does not affect the estimation of the Q-factors. The values of the circuit components, R , L , and C , can be extracted using the equations below,

$$C = \frac{\omega_c}{Z_0 g_1 (\omega_0^2 - \omega_c^2)} \quad (2.1)$$

$$L = \frac{1}{\omega_0^2 C} \quad (2.2)$$

$$R = \frac{2Z_0}{\sqrt{\frac{1}{|S_{11}(\omega_0)|^2} - (2Z_0(\omega_0 C - \frac{1}{\omega_0 L}))^2 - 1}} \quad (2.3)$$

where ω_c and ω_0 are the 3 dB cutoff and the center angular frequencies, respectively, Z_0 is the characteristic impedance of the transmission line, g_1 is the normalized element value in the Butterworth-type one-pole prototype low-pass filter circuit, $g_1 = 2$ [3], $S_{11}(\omega_0)$ is the reflection coefficient at the resonant frequency. Based on the simulation results at $f_0 = 5$ GHz, the values of f_c and $S_{11}(\omega_0)$ of the E-shaped structure were extracted to be 4.55 GHz and 0.99, respectively. The Q-factor without external circuit for a parallel RLC resonant circuit can be estimated using equation in (1.12) where $Q = \omega_0 RC$. By using (2.1)–(2.3) and (1.12), Q-factor for the E-shaped structure was calculated to be 714.6. Note that R in (2.3) and (1.12) accounts for the loss of the resonator including loss through heat and radiation.

Q-factors of other patterned structures in Fig. 2.1 were also estimated based on their simulated S-parameters [3, 4, 6, 7, 9, 28]. Their Q-factors are shown in the last column of Table 2.2. In Table 2.2, the conventional dumbbell-shaped structure has the lowest Q-factor of 4.3, indicating the highest loss. The Q-factor of the modified

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(a)

(b)

(c)

Figure 2.6: The simulated surface current of the ground planes of the patterned ground structures (a) the U-slot patterned structure, (b) the T-shaped structure, and (c) the E-shaped structure.

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Table 2.3: R , L , C of the Circuit Models in Fig. 2.5 for the Patterned Ground Microstrip Resonators at $f_0 = 5$ GHz

Resonator Cell	R ($k\Omega$)	L (nH)	C (pF)
U-slot	27.97	1.60	0.64
T-shaped	22.81	1.14	0.89
E-shaped	13.52	0.60	1.68

dumbbell-shaped structure is 8.3 which is close to that of the dumbbell-shaped structure. The Q-factor of the double-H-shaped structure is 28.0. It is almost seven times as much as that of the dumbbell cell. The spiral-shaped structure has a Q-factor as high as 144.1. The three cells with a U-slot geometry (including the U-slot cell itself) shows a significant increase in Q-factor. Among these three cells with high Q-factors, the proposed E-shaped cell has the highest Q-factor of 714.6. It is the highest Q-factor among all the structures under comparison.

The high Q-factor of the E-shaped structure is due to the high equivalent capacitance created by the geometries of the cell. As mentioned previously, the resonance of the cell can be modeled using a capacitor and an inductor in parallel as shown in Fig. 2.5. Thus, large equivalent capacitance and/or small equivalent inductance lead to a large Q-factor based on (1.12) and (1.14) in Chapter 1. In order to explain the high equivalent capacitance that contribute to a high Q-factor of the E-shaped structure, the working mechanisms of the structures are studied. This is done by analyzing the surface currents of the U-slot, the T-shaped, and the E-shaped structures as follows.

The three cells with U-slot geometries were simulated using CST Microwave Studio [2]. Fig. 2.6 shows the surface currents of their ground planes at the resonant frequency. The size and color of the arrows are proportional to the magnitude of the surface current. As can be seen in Fig. 2.6, currents are flowing around the etched patterns

resulting in a periodic charging and discharging processes. These charging and discharging processes generate capacitive and inductive effect. Fig. 2.6 (a) shows a sketch of the associated inductance and capacitance. With the above charging and discharging mechanisms, the capacitance is determined by the current density and the width of the slot whereas the inductance is determined by the length of the current path. Table 2.3 shows the extracted R , L , and C values from the equivalent circuit model in Fig. 2.5. The U-slot cell has the highest inductance due to the long current path (the largest longitudinal dimension, l in Fig. 2.6 (a)). For the T-shaped structure, the inductance is still high due to the long current path. Its capacitance is increased from the U-slot cell since the surface current density in the central line of the T-shape structure is increased. However, due to the large width of gaps in the longitudinal direction (b in Fig. 2.6 (b)), the capacitance is increased only by a small amount. The E-shaped structure has the highest capacitance because it has two parts of equivalent capacitance: one between the central line and the extended arms (g_e in Fig. 2.6 (c) where $g_e = b - g_h - t$) and the other is between the arms and the edge of the etched patch (g_h in Fig. 2.6 (c)). Since there is high surface current density along the arm of the E-shaped structure, as shown in Fig. 2.6 (c), the resultant capacitance is high. The inductance, on the other hand, is reduced due to the reduction in longitudinal path length ($a_e < a_t$ or $a_e < l$ in Fig. 2.6). The high capacitance contributes to the significant increase in the Q-factor of the proposed E-shaped structure.

The relative power absorbed of the structures were estimated using (2.4) as follows. It mainly includes the dissipation and radiation power of a two-port network.

$$\text{Relative power loss, \%} = (1 - |S_{11}|^2 - |S_{21}|^2) \times 100 \quad (2.4)$$

Fig. 2.7 shows the estimated relative power absorbed of the dumbbell-shaped, the U-slot, the T-shaped, and the E-shaped structure. As shown in Fig. 2.7 (a), the power absorption of the dumbbell-shaped structure is much larger than the

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other three structures. This agrees with the Q-factors of the structures shown in Table 2.2. Fig. 2.7 (a) is zoomed in and shown in Fig. 2.7 (b) where the loss of the later three structures can be seen clearly. As can be seen in Fig. 2.7 (b), the loss of the E-shaped structure is lower than the other two structures at frequencies lower than 6.5 GHz or so while the loss of the T-shaped structure is lower than the U-slot structure in the same frequency range. The observations of Fig. 2.7 (b) agree well with the comparison of Q-factors in Table 2.2. In Fig. 2.7 (b), it is noticed that the loss of the E-shaped structure is high in the upper passband. It corresponds to the high insertion loss in the upper passband as shown in Fig. 2.4 (b). It is a drawback of the E-shaped design that may need to be further addressed in the future.

2.1.2 Effects of the Dimensions

For the proposed E-shaped bandstop resonant cell, the effects of the dimensions, namely the horizontal arm length (l_a), the width of the central ground line (t'), and the gap width (g_h and g_v), on the performance of the resonator were studied based on simulation results. The structure in Fig. 2.3 (b) with $a = b = 5$ mm was studied. Fig. 2.8 shows the transmission coefficients of the E-shaped structures when l_a was varied while the rest of the dimensions of the structure was kept constant. As can be seen in Fig. 2.8, when l_a increases, the resonant frequency f_0 decreases while the cutoff frequency f_c remains the same. Moreover, the insertion loss in the upper passband increases as l_a increases. A negligible change in f_c indicates that l_a has insignificant contribution to the inductance L in the equivalent circuit. This is because that the surface current density at the end of the E-shaped arms is almost negligible, and thus the change in the length of the current path of the arms of the patterned ground structure is negligible. Since the equivalent L is a constant, the shifting of f_0 to a lower frequency as l_a increases is because of a change in capacitance C . In Fig. 2.3 (b), an increase in l_a causes an increase in the length of

(a) The view-all version

(b) The zoomed in version

Figure 2.7: The relative power absorbed of the patterned ground structures

Figure 2.8: S_{21} -parameters of the E-shaped bandstop resonators with different l_a ($w = 1.14$ mm, $a = b = 5$ mm, $g_h = g_v = 0.25$ mm, $t = t' = 0.3$ mm).

Figure 2.9: S_{21} -parameters of the E-shaped bandstop resonators with different t' ($l_a = 2.20$ mm, $w = 1.14$ mm, $a = b = 5.00$ mm, $g_h = g_v = 0.25$ mm, $t = 0.30$ mm).

Figure 2.10: S_{21} -parameters of the E-shaped bandstop resonators with different g_v or g_h ($l_a = 2.20$ mm, $w = 1.14$ mm, $a = b = 5.00$ mm, $t = t' = 0.3$ mm).

the gaps between the arms and the edge of the etched patch whose gap widths are denoted using g_h . The capacitive effect of the gaps between the central line and the extended arms whose gap widths are denoted using g_c is negligible since g_c is large in this part of study ($g_c = 4.45$ mm, $g_h = 0.25$ mm, $g_c \gg g_h$). Although the current density is not high at the end of the arms, the increase in l_a results in a notable increase in C due to the small width of the gaps. For the higher loss in the upper passband with a longer l_a , it is because the increase in l_a results in a shifting of the second resonance to lower frequencies.

Fig. 2.9 shows the simulated S_{21} -parameters of the proposed structures with a varied width of the central ground line, t' . When $t' = 1.14$ mm, the width of the central ground line equals to the width of the microstrip line above it. As seen in Fig. 2.9, an increase in t' leads to an increase in both f_0 and f_c . This is because an increase in t' shortens the current path thus reducing the equivalent inductance. The change in capacitance at g_c due to the increase in t' is negligible because g_c is large, as stated previously.

Fig. 2.10 shows the S_{21} -parameters of the E-shaped structures when g_v or g_h is varied. In Fig. 2.10, it is observed that f_c does not change with the width of the gap while f_0 increases when g_v or g_h increases. This is because the width of the gap (vertical or horizontal) affects the equivalent C but not L . The equivalent inductance L is not changed since the size of the etched patch remains the same (the length of the current path remains the same). It is also noticed that an increase in g_v introduces a larger increase in f_0 than the same amount of increase in g_h . This implies that g_v has more significant effect than g_h , which is due to the high surface current density around the microstrip line.

2.1.3 E-shaped Tunable Bandstop Resonator

2.1.3.1 Design and Circuit Model

The E-shaped tunable bandstop resonator is designed and implemented. In [9], a tunable bandstop resonator is implemented applying T-shaped patterned ground structure [Fig. 2.1 (e)] with varactors embedded in the arms. However, in a T-shaped structure, the length of the arms x has to be sufficiently large in order to accommodate the varactors, which results in a large circuit area of the structure. If the U-slot geometry is adopted to the T-shaped structure, as shown in Fig. 2.3 (a), it becomes difficult to accommodate varactors.

In this proposed E-shaped bandstop resonator shown in Fig. 2.3 (b), the extension arms make it possible for lumped circuit components to be embedded easily into the structure. Fig. 2.11 shows an E-shaped structure with two identical gaps of length d in the extension arms (the gaps in the arms). The gaps can be moved along the arms. The arms of length l'_a that are floating (named floating arms) are coupled to the central ground line through these gaps and laterally. The lateral coupling is negligible here due to the large g_c . The strength of coupling is determined by the length of the gap in the arm d .

Figure 2.11: The schematic of an E-shaped structure for realizing the tuning functionality.

Fig. 2.11 shows the E-shaped resonator with gaps in the arms. It is used to demonstrate the tuning functionality of the proposed structure. Taconic ($\epsilon_r=2.45$, $h=0.4$ mm) was used as the substrate. The dimensions are labeled as shown in Fig. 2.11. $w=1.14$ mm corresponding to a characteristic impedance of 50Ω . In order to facilitate the fabrication and testing, $g_h=g_v=0.25$ mm, $t=1.00$ mm, and $t'=0.40$ mm. The length of the vertical arm x was set to 8.01 mm in order to minimize interference caused by the embedded circuit components. This E-shaped structures with gaps and the one of the same dimensions without gaps were simulated. Fig. 2.12 shows their simulated S_{21} -parameters. As shown in Fig. 2.12, the structure without gaps shows a resonant frequency of 1.80 GHz whereas the structure with gaps shows a resonant frequency of 2.66 GHz.

Based on the equivalent circuit of an E-shaped structure in Fig. 2.5, a circuit model was proposed for the E-shaped structure with gaps in the arms [Fig. 2.11] and shown in Fig. 2.13 (a). As shown in Fig. 2.13 (a), C_1 corresponds to the coupling between the central ground line and the edge of the etched patches while C_2 corresponds to that part of coupling between the floating arm and the edge of the etched patch. Therefore,

$$C_1 + 2C_2 \approx C \quad (2.5)$$

Figure 2.12: The S_{21} -parameters of the E-shaped structures with or without gaps in Fig. 2.3 (b) and Fig.2.11.

where C is the capacitance in the equivalent circuit of the E-shaped structure without gaps in Fig. 2.5. C_g is the new coupling between the floating arm and the central ground line introduced by the gap. Nodes of the circuit are labeled as shown in Fig. 2.13 (a) for further analysis.

With C_g , the total capacitance between node a_1a_2 , $C_{a_{12}}$, can be estimated using (2.6) below,

$$C_{a_{12}} = (C_2 C_g) / (C_2 + C_g) \quad (2.6)$$

Based on (2.6), $C_{a_{12}} = C_2$ when there is no gap and C_g goes to infinity, resulting in $C_1 + 2C_{a_{12}} \approx C$ according to (2.5); $C_{a_{12}} < C_2$ when gaps are introduced and C_g is a finite number, leading to $C_1 + 2C_{a_{12}} < C_1 + 2C_2$; $C_{a_{12}} = 0$ and node a_1 and a_2 are disconnected when C_g is approaching 0. Based on the simulation results and (2.1)-(2.3), R , L , and C in the equivalent circuit for the E-shaped structure without gaps were obtained to be 289479.70Ω , 3.25 nH , and 2.37 pF , respectively. Using Schematic of Agilent's Advanced Design System [30], C_g can be obtained

Figure 2.13: The proposed circuit model for E-shaped structure (a) with gaps, (b) with mounted capacitors across the gaps.

by following the guidelines as followed. $C_1 + 2C_2 < C$ since a small portion of coupling between the arms to the edge of the etched patch is eliminated due to the gap; $C_1 > 2C_2$ due to the high confinement of electromagnetic fields around the microstrip line. With $C_1 = 1.1$ pF and $C_2 = 0.59$ pF, C_g was extracted to be 10^{-5} pF for the structures with gaps ($C_g \approx 0$ and node a_1 and a_2 are equivalent to open circuit). The equivalent circuits for the E-shaped structures without gaps and with gaps were simulated [30]. Fig. 2.12 shows their simulated S_{21} -parameters. As can be seen in Fig. 2.12, the simulation results of the equivalent circuits are in good agreement with those using a full-wave simulation.

In the design of an E-shaped patterned ground structure, when C_g is approaching zero as that in the case above, C_1 can be estimated using (2.7) below,

$$C_1 = 1/(\omega_{0g}^2 L) \quad (2.7)$$

Figure 2.14: The equivalent circuit of the tunable E-shaped bandstop resonator with varactors.

where ω_{0g} is the angular resonant frequency of the E-shaped structure with gaps. The gaps in the arms allow accommodation of circuit components. If a capacitor is mounted across the gap, the circuit in Fig. 2.13 (a) is modified to the circuit in Fig. 2.13 (b) where the mounted capacitor introduces a C_c that is in parallel with C_g . With the mounted capacitors across the gaps, the total capacitance between node a_1a_2 changes and denoted as $C'_{a_{12}}$. $C'_{a_{12}}$ can be expressed below when the inductance of the mounted capacitor L_c is neglected in the derivation,

$$C'_{a_{12}} = (C_2(C_g + C_c)/(C_2 + C_g + C_c) \quad (2.8)$$

When $C_g \approx 0$, $C'_{a_{12}} \approx C_2C_c/(C_2 + C_c)$. According to (2.8), $C_2 > C'_{a_{12}} > C_{a_{12}}$, thus $C > C_1 + 2C'_{a_{12}} > C_1 + 2C_{a_{12}}$ based on (2.5). Therefore, mounting capacitors across gaps tunes the resonant frequency within a range of frequencies between the frequency for the structure without gaps and the frequency for the structure with gap. The structure becomes a tunable bandstop resonator when tunable capacitors

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Figure 2.15: The measured and simulated S_{21} -parameters of the tunable E-shaped bandstop resonator when varactors are mounted across the gaps.

or varactors are used. When varactors are embedded into the structure, the tuning range of the E-shaped resonator can be obtained based on the equivalent circuit in Fig. 2.13 (b) and the specifications of the varactor.

The MV31020-46 varactors from the MDT Corporation, Westford, MA, were used in order to realize a tunable bandstop resonator. The equivalent circuit of the E-shaped structure with varactors is shown on the right of Fig. 2.14. The capacitor C_c in the equivalent circuit for the E-shaped structure with gaps in Fig. 2.13 (b) is replaced by the equivalent circuit of the varactor as shown on the left of Fig. 2.14. C_J is the junction capacitance that can be estimated using (2.9) [100],

$$C_J = \frac{C_{0j}}{(1 + \frac{V}{\Phi})^\gamma} \text{ pF} \quad (2.9)$$

where C_{0j} is the junction capacitance at 0 V in pF, Φ is the built-in potential, V is the applied reverse voltage, and γ is the tuning slope. L_s is the parasitic inductance, R_s is the parasitic resistance, and C_p is the package capacitance. Fig. 2.15 shows

Figure 2.16: The C-V characteristics of the varactor used in the proposed tunable bandstop resonator.

the simulation results of this equivalent circuit when the DC bias is set to 0 V and 15 V with $L_s = 0.170$ nH, $R_s = 0.497 \Omega$, $C_p = 0.100$ pF, $C_{0j} = 20.007$ pF, $\Phi = 1.2$ V for GaAs, and $\gamma = 1.250$ [100]. As shown in Fig. 2.15, the equivalent circuit has a resonant frequency at 1.81 GHz and 2.07 GHz when V is set to 0 V and 15 V, respectively. The predicted tuning range is 12.4 %.

2.1.3.2 Experiments

The E-shaped structure was fabricated. The measured S_{21} -parameters of the structure without gaps and with gaps are shown in Fig. 2.12 with their simulated results. As can be seen in Fig. 2.12, the E-shaped structure shows a resonant frequency 1.80 GHz when there are no gaps in the structure whereas a resonant frequency of 2.66 GHz when there are gaps. The measurement results agree well with the simulation results.

The MV31020-46 varactors were mounted across the gaps. Fig. 2.16 shows the

capacitance-voltage characteristics of this varactor. The tuning range for the capacitance is from 20 pF to 0.5 pF when the applied bias DC voltage is varied from 0 V to 15 V.

Fig. 2.17 (a) and (b) shows the photographs and the schematic of the microstrip line of the implemented tunable bandstop resonator, respectively. Fig. 2.18 shows the photograph of its ground plane. DC voltage is supplied from the plane with a microstrip line, through a via, to the varactors on the other side of the board (the ground plane). On the plane with a microstrip line, an RF choke (47 nH with a self-resonant frequency at 2.6 GHz) is placed between the DC supply and the varactor in order to minimize RF loss of the circuit in the frequency range of interest. Moreover, the DC supply and RF chokes are located outside the circuit area of the resonator so as to eliminate their possible effect on the resonant cell. In Fig. 2.18, the E-shaped pattern is etched in the ground plane. Vias are introduced to the arms and varactors are mounted across the gaps.

Fig. 2.15 shows the measured S_{21} -parameters when varactors across the gaps have DC bias. As shown in Fig. 2.15, the E-shaped structure has a resonant at 1.81 GHz, 1.90 GHz, 1.97 GHz, and 2.07 GHz when DC bias of 0 V, 5 V, 10 V, and 15 V are supplied, respectively. This proposed E-shaped tunable bandstop resonator realizes a tuning range of 12.4 %. As can be seen in Fig. 2.15, the resonant frequency increases with an increase in the DC voltage (a decrease in total capacitance of the varactors) within the expected tuning range (1.80 GHz to 2.66 GHz). The measured results when the DC bias is 0 V and 15 V are in good agreement with the simulated results of the equivalent circuit in Fig. 2.14. This indicates the accuracy of the proposed equivalent circuit in Fig. 2.14 for the E-shaped structure with mounted lumped components across the gap. Based on the measured S-parameters, the Q-factors of the implemented tunable resonator were calculated for different DC supply voltages. The Q-factor varies from 18.65 to 31.67 when the DC supply is

(a) The photograph of the microstrip line side

(b) The schematic of the microstrip line side

Figure 2.17: The fabricated tunable E-shaped bandstop resonator with varactors and DC supply

Figure 2.18: The photograph of the ground plane side of the fabricated tunable E-shaped bandstop resonator with varactors and DC supply.

changed from 0 V to 15 V. Comparing to the E-shaped structure in Fig. 2.3 (b), the tunable resonator has a lower Q-factor. This is probably due to the loss of the varactors and the DC supply circuits that are used for tuning.

In comparison to a T-shaped structure, their tuning ranges are comparable while this proposed E-shaped structure is much smaller in physical size when the center frequency of tuning is set to be the same. Furthermore, only the E-shaped structure is able to accommodate lumped components even when the length of the vertical arm x is very small.

2.2 Compact Dual Planar EBG Microstrip Structures

Besides a single etched unit in the ground plane, periodic etched units alter the transmission of waves in a microstrip line. In this section, a straight dual planar EBG structure is proposed based on two single planar EBG structures achieving

(a) 3-D view

(b) Top view

Figure 2.19: Schematic of the proposed dual planar EBG microstrip structure

a wide stopband with high attenuation without increasing the physical size of the structure. With the wide stopband with high attenuation, this type of EBG structures can filter spurious signals in a wide bandwidth in structures such as the wideband or ultra-wideband [101] microwave devices, by using a single structure rather than cascading filters with different stopbands. The size of the proposed structure is further reduced by applying a U-shaped geometry. The dual planar EBG structures with high performance within small circuit areas are promising alternatives to conventional planar filters.

2.2.1 A Straight Dual Planar EBG Microstrip Structure

2.2.1.1 Design & Numerical Results

Fig. 2.19 (a) and (b) shows a 3-D and top views of the proposed straight dual planar EBG microstrip structure, respectively. As shown in Fig. 2.19 (a), there are circular patches with a radius of r etched directly below the microstrip line at a period of a_1 and rectangular patches with a length of l_a and a width of w_a inserted into the microstrip line at a period of a_2 . The two periods were determined using the Bragg reflection condition [40] where the period equals to half of the guided wavelength of the structure at the center frequency of the stopband f_0 . They were set to be the same in this design. As shown in Fig. 2.19 (b), the width of the microstrip line w was set to correspond to a characteristic impedance of $50\ \Omega$ at f_0 . The distance between successive rectangular and circular patches is d and it was set to be half of a period. The distance between the central axis of the circular etched patches and the longitudinal center line of the microstrip line is d' and it was set to be 0 to align the two arrays. With this configuration, an increase in the variation of reactance in the longitudinal direction can be obtained due to the additional equivalent inductance and capacitance introduced by the etched circles and the inserted rectangular patches, respectively. Therefore, the stopband performance of the structure can be enhanced. In Fig. 2.19 (b), $l_a = w_a$ and they were set to be half of the period. The radius of the circular patch was set to be a quarter of the period. Thus, overlaps between successive etched patches and inserted ones are avoided and the contrast between the reactance of the successive microstrip line sections (the inductances for the microstrip line section with an etched circle and the capacitance for the microstrip line section with an inserted patch) is enhanced. The proposed dual planar EBG microstrip structure with five inserted square patches in the microstrip line and six etched circles in the ground plane was simu-

Figure 2.20: Simulated S_{21} parameters of the dual planar EBG structure and two single-plane EBG structures

lated using Agilent's Advanced Design System [30]. The two corresponding single planar EBG structures, the structure with only six etched circular patches (named SP-six-circle) and the one with only five inserted squares (named SP-five-patch), were simulated for comparison purposes. The SP-six-circle structure is the same as the 1-D EBG microstrip structure proposed in [17] whereas the SP-five-patch structure is a width-modulated microstrip periodic structure similar to the first-order approximation of the sinusoidally width-modulated microstrip structure reported in [11].

Taconic with a dielectric constant (ϵ_r) of 2.43 and a thickness (h) of 30 mils was used as the substrate. The center frequency of the stopband was set to be 10 GHz and a_1 and a_2 were determined to be 10.35 mm according to the Bragg reflection condition. The width of the microstrip line w was set to 2.29 mm, corresponding to a characteristics impedance of 50Ω at 10 GHz. The radius of the circular etched patch was set to be 2.59 mm which is a quarter of the period of the structure. The

2.2 Compact Dual Planar EBG Microstrip Structures

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Table 2.4: Simulated Performances of EBG Microstrip Structures

EBG Structure	10 dB Bandwidth	Maximum Attenuation (GHz)	f_0 (dB)	Ripple Level (dB)	
				Lower	Higher
Dual planar (uniform)	7.91 GHz	50.72	10	3.97	8.80
Single planar six-circle (uniform)	4.98 GHz	34.24	10	2.77	6.55
Single planar five-patch (uniform)	3.27 GHz	20.17	9.5	1.60	2.55
Dual planar (tapered, ground plane)	6.60 GHz	67.71	10	0.92	3.42
Dual planar (tapered, ground plane and microstrip line)	6.30 GHz	41.54	10	0.16	3.20

(a) The dimensions of the etched circular patches are tapered

(b) The dimensions of both the inserted squares and the etched circular patches are tapered

Figure 2.21: Schematics of the tapered dual planar EBG microstrip structures adopting the Chebyshev distribution

(a)

(b)

Figure 2.22: Simulated S-parameters of the tapered and uniform dual-plane C-EBG microstrip structure (a) S_{21} (b) S_{11}

length and the width of the inserted patch in the microstrip line, l_a and w_a , were both fixed to 5 mm that are half of the period.

Fig. 2.20 shows the simulated S_{21} -parameters of the dual planar EBG structure and those of the corresponding single planar EBG structures. As shown in Fig. 2.20, the dual planar structure shows a 10 dB stopband of 7.91 GHz centered at 10 GHz, a maximum attenuation of 50.72 dB in the stopband, and ripple levels of 3.97 dB and 8.80 dB in the lower and the upper passbands, respectively. The performances of the three EBG structures are tabulated in row one, two and three in Table 2.4. As shown in Table 2.4, comparing to the two single planar EBG structures, the dual planar EBG structures shows a much wider stopband bandwidth and a significant increase in the maximum attenuation of the stopband. However, it is noticed that its ripple levels in both the lower and upper passbands are much higher (at least 1.2 dB higher in the lower passband and 2.2 dB higher in the upper passband), which is due to the periodicity in both the microstrip line and the ground plane. In Table 2.4, it is also noticed that the center frequency of the five-patch single planar EBG structure is shifted by 0.5 GHz compared to the designed frequency based on the Bragg reflection condition. The shifting may be due to the strong reflections at the abrupt changes of the width of the microstrip line in this five-patch structure which affects the Bragg reflections.

2.2.1.2 Tapering

Tapering techniques were applied to the dual planar EBG structure to eliminate the ripple levels in the passbands. Fig. 2.21 (a) and (b) show the schematics of the dual planar EBG structures when the etched circular patches in the ground plane are tapered (called single tapered structure) and when both the etched patches and the inserted squares in the microstrip line are tapered (called double tapered structure), respectively. In these tapered structures, the areas of the etched or/and

(a)

(b)

Figure 2.23: Photographs of the fabricated ground tapered dual planar EBG microstrip structure (upper) and the fabricated double tapered dual planar EBG microstrip structure (lower) (a) microstrip line, (b) ground plane

(a)

(b)

Figure 2.24: Simulated and measured S-parameters of the tapered dual planar EBG microstrip structures (a) the structure with tapered etched circular patches in the ground plane, (b) the structure with inserted squares in the microstrip line.

inserted patches are tapered applying Chebyshev distribution. For the details of tapering, please refer to Appendix B in this thesis.

The two tapered structures were simulated and Fig. 2.22 (a) and (b) show their simulated S_{21} - and S_{11} -parameters and those of the dual planar structure. Their performances are tabulated in the last two rows in Table 2.4. As shown in Table 2.4, the EBG structure that has tapered etched patches in the ground plane shows a significant reduction of 76.83% and 61.14% in the ripple levels in the lower and the upper passbands, respectively, compared to its counterpart with uniform unit cells. The center frequency of its stopband is at 10 GHz, not affected by the tapering of the cells. However, it is noticed that its bandwidth of the stopband decreases from 7.91 GHz to 6.60 GHz. For the tapered structure with EBG unit cells in both the microstrip line and the ground plane tapered, the ripple levels in the lower and the upper passbands are further reduced to 0.16 dB and 3.20 dB, respectively. Its stopband bandwidth is reduced to 6.30 GHz while the center frequency of the stopband is remained to be 10 GHz.

As revealed by the results above, the ripple level of the proposed dual planar EBG structure can be tailored by applying a certain tapering technique to taper the dimension of EBG unit cells on different planes. This comes with degradations of the stopband performance. The ripple level of the passband in a double tapered EBG structure is lower than that in a single tapered EBG structure such as the ground tapered EBG structure. With the unique dual-plane configuration, the filtering functionality of the proposed structure can be adjusted to meet different requirements in various applications.

2.2.1.3 Measurement Results

Both the proposed ground tapered and double tapered five-patch-six-circle dual planar EBG microstrip structure were fabricated and tested. Fig. 2.23 (a) and (b)

shows the modulated microstrip line and the ground plane with etched circles of the fabricated structures, respectively. These structures are highly compact with a dimension of 70 mm by 26 mm as shown in Fig. 2.23. Fig. 2.24 (a) and (b) show the simulated and measured S-parameters of the fabricated ground tapered and the double tapered structures, respectively. As shown in Fig. 2.24, with only six EBG cells, the ground tapered EBG structure shows a 6.70 GHz 10-dB bandwidth of the stopband centered at 10 GHz, with a maximum attenuation of 39.00 dB, a ripple level of 1.56 dB in the lower passband, and a ripple level of 4.68 dB in the upper passband. The fabricated double tapered structure exhibits a 25-dB bandwidth of 4.87 GHz, a maximum attenuation of 52.9 dB, a ripple level of 0.51 dB in the lower passband, and a ripple level of 5.47 dB in the upper passband. A good agreement between the measurement and the simulation results has been obtained in both proposed tapered dual planar EBG structures. The slight difference is probably due to the effect of the SMA (SubMiniature version A) connectors, the lack of material uniformity and cell uniformity, the over etching, and the non-ideal soldering. The infinite ground plane assumed in the simulation is another possible cause of the difference.

Both proposed tapered dual planar EBG microstrip structures demonstrate the advantages of a large stopband bandwidth, high attenuation in the stopband, smooth passband, compact size, and easy fabrication process. The ground tapered EBG structure is able to obtain an ultra-wide stopband with very high attenuation and relatively small ripple levels in both passbands whereas the double tapered EBG structure shows a small ripple level in the passband but a relatively smaller stopband bandwidth and attenuation. Due to the common characteristics as well as the different unique features, these two proposed structures can easily be applied to various circuit applications to meet different requirements on the passband and the stopband performance.

(a) 3-D view

(b) Top view 1

(c) Top view 2

Figure 2.25: Schematic of the proposed high performance compact U-shaped dual planar EBG microstrip structure with low-pass filtering functionality.

2.2.2 A U-shaped Dual Planar EBG Microstrip Structure

2.2.2.1 Design, Numerical Results, & Measured Results

Bending or meandering a straight microstrip line is a straightforward and effective way to reduce the circuit area of the structure [18, 21]. The meander structures in [18] and [21] are shown in Fig. 1.38 and Fig. 1.41, respectively. The meander geometry of a microstrip line may affect the transmission of the structure. In order to reduce the physical size of the proposed dual planar EBG microstrip structure in Fig. 2.19, it was bent into a U-shape. Its performance is improved due to the bending. The details of the design are presented and the improvement of performance due to the bending is analyzed in this section.

Fig. 2.25 (a) shows the 3-D schematic of the U-shaped dual planar EBG microstrip structure with four inserted patches in the microstrip line and three circular etched patches in the ground plane. The microstrip line with inserted square patches were bent into a U-shape and two 90° turns were made at the center of the squares. As indicated in Fig. 2.25 (a), the substrate material has a dielectric constant of ϵ_r and a thickness of h .

Fig. 2.25 (b) shows the relative location of the patches and the circles and Fig. 2.25 (c) shows their dimensions. The width of the microstrip line is w , corresponding to a characteristic impedance of 50Ω at the center frequency of the stopband f_0 . The square patches were inserted at a period of a according to the Bragg reflection condition. Circles were etched exactly below the microstrip line with their centers allocated at the center point of two consecutive square patches ($d = a/2$). The etched circles construct another single planar EBG structure [A 8] with a period of a , forming a dual planar EBG configuration. In order to further increase the equivalent inductance of the microstrip line section with an etched circle to enhance the contrast between an inductive and a capacitive microstrip line section, the width

Table 2.5: Normalized Coefficients of Chebyshev Arrays

Chebyshev Array	Normalized Coefficients
<i>Three-Element</i>	0.73, 1, 0.73
<i>Four-Element</i>	0.48, 1, 1, 0.48

of the microstrip line above the etched circle was reduced, $w_{bi} < w$ where $i = 1, 2$. As reported in [A 9], the reduction leads to a further increase in the bandwidth and attenuation of the stopband.

The Chebyshev distribution was applied to taper the area of the inserted patches and the circular etched patches in order to eliminate ripple levels in the passband. As shown in Fig. 2.25 (c), the corresponding radii are r_1 and r_2 and the corresponding lengths of the edge are e_1 and e_2 . The radius of the center circular etched patch r_2 was set to be a quarter of the period of the structure. The length of the edge of the two central squares e_2 was set to avoid any overlap between the circle in the ground plane and the square patch in the microstrip line in order to obtain a sharp reactance contrast in an EBG cell which ensures good filtering performance of the structure. The effect of the reduction in the width of the microstrip line ($w - w_{bi}$) has been studied and it was found that good performance can be obtained when ($w - w_{bi}$) is also tapered [A 9]. A three-element and a four-element Chebyshev array with a major-to-minor ratio of 25 dB were used for the proposed design and the normalized coefficients are tabulated in Table 2.5.

The proposed U-shaped dual planar EBG microstrip structure was simulated using Agilent's ADS [30]. Taconic ($\epsilon_r = 2.43$, $h = 0.76$ mm) was used as the substrate and the center frequency of the stopband f_0 was set to 7.5 GHz. Table 2.6 shows the the parameters set. Fig 2.27 (a) and (b) show the simulated S_{21} - and S_{11} -parameters of the U-shaped structure where the effect of the periodicity and that of the bending on the transmission of the structure can be seen. As shown in

2.2 Compact Dual Planar EBG Microstrip Structures

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Table 2.6: Parameters of the Proposed Structure with Low-pass Filtering Functionality

Parameter	Dimension (mm)	Parameter	Dimension (mm)
w	2.32	r_1	2.94
a	13.78	r_2	3.45
e_1	4.79	w_{b1}	0.85
e_2	6.89	w_{b2}	0.32

Fig 2.27 (a), the U-shaped structure shows a 10 dB bandwidth of more than 8 GHz, a maximum attenuation of 40 dB in the stopband, and a ripple level of 0.20 dB in the lower passband. The cutoff of the structure is sharp, which is quantified using a parameter called selectivity. Selectivity evaluates how fast waves are attenuated as the frequency increases. It has a unit of dB/GHz. For the U-shaped structure, it has a selectivity of 22.50 dB/GHz. The U-shaped structure was fabricated and tested. Fig. 2.26 shows the microstrip line (left) and the ground plane (right) of the fabricated structure. The measured S-parameters were plotted in Fig. 2.27. As shown in Fig. 2.26, EBG cells are located in both the microstrip line (the inserted square patches) and the ground plane (the etched circles). The input port and the output port are on the same side of the circuit, which may offer flexibility in the layout designs of microwave circuits. The physical size of the proposed structure is as small as $33 \times 30 \text{ mm}^2$. The circuit area is reduced by 45.60% as compared to that of the corresponding straight structure.

The good low-pass filtering performance of the U-shaped dual planar EBG microstrip structure as shown in Fig 2.27 is the results of the periodicity of the EBG cells in both the microstrip line and the ground plane, as well as the result of the bending of the microstrip line. As the bending of the microstrip line plays an important role in improving the performance of the structure, the effect of bending is singled out and analyzed as follows.

Figure 2.26: Fabricated compact U-shaped dual planar EBG microstrip structure: microstrip line (left), ground plane (right)

2.2.2.2 Analysis on the Effect of a U-shaped Geometry on the Performance

In order to understand the good low-pass filtering performance of the U-shaped dual planar EBG microstrip structure, the effect of a U-shaped geometry on the filtering performance of an EBG structure with periodically inserted patches in the microstrip line is studied. The effect on the radiation of the structure is also presented. Fig. 2.28 (a) shows the schematic of a straight single planar EBG microstrip structure with four square patches inserted in the microstrip line at a period of a . It is sometimes called capacitively modulated EBG structure in the literature. Fig. 2.28 (b) shows the schematic of a U-shaped microstrip line where the distance between the two parallel sections is a . Fig. 2.28 (c) shows the schematic of the EBG microstrip structure in Fig. 2.28 (a) when it is bent into a U-shape. The 90° bending takes place at the center point of the second and the third patch from left to right.

The straight EBG microstrip structure in Fig. 2.28 (a) satisfies the Bragg reflection condition where a equals to half of the guided wavelength at the center frequency of the stopband f_0 . Thus,

$$f_0 = \frac{c}{2a\sqrt{\epsilon_{eff}}} \quad (2.10)$$

(a)

(b)

Figure 2.27: Simulated and measured S-parameters of the proposed U-shaped dual planar EBG microstrip structure (a) S_{21} (b) S_{11}

(a)

(b)

(c)

Figure 2.28: Schematic of (a) a straight EBG microstrip structure, (b) a U-shaped microstrip line, and (c) a U-shaped EBG microstrip structure

c is the speed of light in free space and ϵ_{eff} is the effective dielectric constant. It is not a conventional stepped-impedance microstrip low-pass filter. The length of the inserted square patch e equals to half of the period ($e = a/2$).

Compared to the straight EBG microstrip structure, the U-shaped EBG microstrip structure in Fig. 2.28(c) is more compact in physical size and may provide additional flexibility in circuit layout designs. This motivates the change of the geometry. However, the performance of the U-shaped structure is affected by the resonator created by the bends as shown by the shadowed area in Fig. 2.28(b). The resonator in a microstrip line with multiple bends (the number of bends ≥ 2)

(a)

(b)

Figure 2.29: Simulated S -parameters of structures in Fig. 2.28 (a) S_{21} , (b) S_{11}

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is the result of the reflections between two consecutive or nearby bends (with one bend in between). The resonant frequency is determined by the physical length of the resonator. Base on a study on the meander microstrip line [102], the relation between the length of the resonator and the resonant frequency can be expressed using (2.11)

$$L = n \cdot \frac{\lambda'_g}{2} \quad (2.11)$$

where L is the physical length of the resonator, n is an integer, and λ'_g is the guided wavelength corresponding to the resonant frequency f_r . With $\lambda'_g = c/(f_r \sqrt{\epsilon_{eff}})$, (2.11) is expressed as

$$L = n \cdot \frac{c}{2f_r \sqrt{\epsilon_{eff}}} \quad (2.12)$$

As can be seen in Fig. 2.28 (b), the U-shaped microstrip line contains two bends. Therefore, it has only one resonator with a length of a as shown by the shadowed area. According to (2.12), with n set to 1, its resonant frequency $f_r = c/(2a\sqrt{\epsilon_{eff}})$ is the same as the center frequency of the stopband f_0 in the straight EBG structure (Fig. 2.28 (a)) as approximated by (2.10).

Compared to the meander line in [102], the U-shaped microstrip line has less bends. As a result, U-shaped structures may have smaller power loss caused by the radiation and scattering that take place at the bend. In the U-shaped microstrip line, there is only one resonator and its length is as short as one EBG period. Due to the absence of a resonator with a length longer than a , no resonance is introduced at frequencies lower than f_0 based on (2.12).

The transmission characteristics of an EBG microstrip structure with multiple bends are determined by the responses of the two individual structures, the corresponding straight EBG structure, and the microstrip line with a corresponding bent geometry. In order to study the U-shaped EBG microstrip structure in Fig. 2.28(c), the three structures in Fig. 2.28 were simulated [30]. The center frequency of the stopband was set to be 7.5 GHz and the same dielectric material was

Figure 2.30: Simulated phase delay in the output wave with respect to the input wave in a bent unitary period and in a straight unitary period

used ($\epsilon_r = 2.43$, $h = 0.76\text{mm}$). The values for a and w are the same as those listed in Table 2.6. The dimensions of the inserted patches were set to be the same. The edge of the square patch e was set to 6.89 mm which equals to half of the period. Fig. 2.29(a) and (b) show the simulated S_{21} - and S_{11} -parameters of the structures, respectively.

As can be seen in Fig. 2.29(a), the U-shaped microstrip line shows attenuation in the high frequency range. Based on the S_{11} -parameters shown in Fig. 2.29(b), the U-shaped microstrip line shows a weak resonance centered at around 8 GHz . This resonance is due to the resonator where reflections take place between the bends. With a physical length of 13.78 mm , the predicted resonant frequency is about 7.5 GHz using (2.12) which is close to that predicted using the commercial software. In Fig. 2.29 (a), the straight EBG microstrip structure shows a prominent stopband centered at about 7 GHz , which indicates that it satisfies the Bragg reflection condition. The U-shaped EBG microstrip structure shows a wider stopband with

Figure 2.31: Simulated radiation and scattering levels for the U-shaped EBG microstrip structure and the straight EBG structure

higher attenuation and an increase in selectivity yet without compromising on the ripple level in the lower passband. The low passband ripple level of the U-shaped EBG structure is indicated by the low side lobe level of its S_{11} -parameter in Fig. 2.29 (b).

A wide stopband with high attenuation always come with an increase in ripple level in the passband in microstrip structures. However, the results above show that the U-shaped microstrip line geometry enhances the filtering functionality of the capacitively modulated EBG structure without producing any additional ripple or increasing the ripple level. The coexistence of the enhanced filtering functionality and the good performance in the passband is analyzed as follows.

One of the reasons for the coexistence of the good performances in both the passband and the stopband is the absence of the resonator with a length that is longer than one period of the structure in the U-shaped EBG structure which avoids resonances at frequencies lower than the center frequency of the stopband. On the

other hand, the weak resonance at the center of the stopband may help to increase the attenuation of the stopband. For a small ripple level in the passband of the U-shaped structure, another reason is the minimized difference of phase delay caused by the bends. In [102], it is reported that, in a meander EBG microstrip structure with circles etched in the ground plane below a bend, the bending of a unit cell introduces a reduction in the phase delay, resulting in an increase in the passband ripple level due to the phase mismatch. Fig. 2.30 shows the phase delays of a bent (shadowed area) and a straight unit cells (meshed area) of the U-shaped EBG structure in Fig. 2.28 (c). As indicated in Fig. 2.30, in the U-shaped EBG structure with a square patch inserted into the microstrip line at a bend (Fig. 2.28 (c)), the phase delay of the bent unit cell is the same as that of the straight one over a wide frequency range from DC to 6.76 GHz. Since the bending of the unit cell does not introduce any difference in phase delay, the U-shaped EBG microstrip structure in Fig. 2.28 (c) maintains a low ripple level in the passband when the stopband performance is enhanced by the bends.

Additional power loss due to the radiation that takes place at the bends is another important issue for an EBG microstrip structure with bends. In the meander EBG microstrip structure in [102, 103], the radiation at the bend with an etched circle below introduce a significant amount of additional power loss, thus degrading the performance of the structure. In this study, the radiation loss in the meander microstrip structure is approximated using (2.4) since lossless materials were used in the simulations and there is no power dissipated due to dielectric and conduction loss.

The power loss of the U-shaped EBG microstrip structure in Fig. 2.28 (c) and that of the straight structure in Fig. 2.28 (a) were estimated using (2.4) and compared. Fig. 2.31 shows the estimated radiation levels of these two EBG structures. As shown in Fig. 2.31, the power loss of the U-shaped EBG structure is similar to

that of the straight one over a frequency range DC - 9.42 GHz. This shows that the bends in Fig. 2.28 (c) do not introduce additional radiation to the structure. It is due to the square patch inserted at the bend of the microstrip line that alters the current distributions at the bend in a way that radiation is not enhanced. This is also attributed to the small number of bends in the U-shaped structure.

The U-shaped geometry increases the bandwidth and attenuation of the stopband in a capacitively modulated EBG microstrip structure while the ripple level in the lower passband is not increased and the radiation of the structure is low. The good filtering performances are attributed to the U-shape geometry and the inserted patches at the bends.

2.3 Summary

In this chapter, the effects of ground planes with etched patterns on the propagation of electromagnetic waves at microwave frequencies have been examined by presenting new designs, including both patterned ground structures with single etched units and those with periodic etched units.

A single etched unit in the ground plane of the microstrip line provides a resonance based on the geometry of the etched pattern. The geometry also affects the Q-factor and the physical size of the structure. An E-shaped etched pattern has been proposed. Compared to the patterned ground structures with single etched units in the literature, the E-shaped structure shows the sharpest transition knees before and after the resonance, thus the highest Q-factor over the smallest circuit area. The working mechanism of the structure has been studied by conducting a circuit modeling and by examining the current distributions in the ground plane. The effects of the dimensions of the structure on its performances have also been examined. The proposed etched pattern enables the design of a compact tunable

bandstop filter. Over a small circuit area, the tunable bandstop filter achieves a wide tuning range of 12.4 %. The tunable structure has been modeled to ease the study of its tuning behavior. The E-shaped and the tunable structure have both been implemented and tested. Simulation and measurement results are in good agreements for both structures, verifying the accuracy of the proposed circuit models and the superior performances of the proposed patterned ground structures. An array of etched unit cells arranged along a microstrip line (an EBG structure) exhibits a prominent stopband at microwave frequencies whose center frequency is determined by the Bragg reflection condition. A dual planar configuration has been proposed for EBG microstrip structures. When an EBG structure has unit cells in both the microstrip line and the ground plane arranged in a way that the contrast of the reactance of the successive microstrip line sections is high, a wide stopband with high attenuation can be obtained within a small physical size. Although the ripple levels in the passbands increase due to the periodicities in both the microstrip line and the ground plane, they can be effectively eliminated by applying tapering techniques, such as Chebyshev distribution. A capacitively modulated EBG microstrip structure is a microstrip line with periodic inserted patches into the microstrip line. When bending takes place around the center of the patch, it shows small phase difference and power loss as compared to its straight counterpart. A U-shaped dual planar EBG microstrip structure is constructed by bending a capacitively modulated EBG structure 90° at two successive inserted patches. It shows an increase in the attenuation and in the bandwidth of the stopband and no increase in the ripple level in the passband, as compared to the corresponding straight structure. The proposed straight and the U-shaped dual planar EBG structures with high performance have been fabricated and tested, verifying the calculated results.

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Chapter 3

Effects of Ground Plane

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Propagation II

In this chapter, the effects of ground planes with etched patterns on the propagation of electromagnetic waves at microwave frequencies are further studied. The effects are demonstrated by using etched patterns in the ground plane to optimize a conventional filter. A circuit modeling of a patterned ground structure (a capacitively modulated EBG microstrip structure) is conducted. Based on the circuit model, a time-saving method to calculate the transmission coefficients of the structure is proposed and presented.

3.1 Slot-Back Stepped-Impedance Low-Pass Filter

Etched patterns in the ground plane can be applied to optimize the performance of existing planar filter designs. In this section, a microstrip line with a longitudinal

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slot in the ground plane (slot-back microstrip line) is applied to optimize the performance of a conventional stepped-impedance low-pass filter. The slot-back microstrip line is used to implement the inductive elements in the filter. The design process is simple and this approach is effective in improving the performance of the filter structure.

In modern wireless communications, a filter is required to have high selectivity, good stopband and passband performances, and compactness. The ladder LC lumped-circuit low-pass filter prototype [50, 104] is well-known and widely used in filter designs.

The circuit components (inductors and capacitors) in the low-pass filter prototype can be implemented using various structures [3, 105, 106, 107, 108]. For example, the inductive elements of the filter prototype can be implemented using a defected ground structure (DGS) [3, 105, 106, 107]. In [105], a DGS-base low-pass filter shows an effective suppression of the harmonic response, an improvement on the cutoff, and a reduction in circuit area. However, the design demands a complicated process to obtain the dimensions of the DGS from the circuit model. In [106, 107], DGS-base low-pass filters exhibit improvement on the cutoff and the bandwidth of the stopband, while their insertion loss in the passband is considerably high. Other than a DGS, new transmission lines such as a micromachined overlay CPW [108] were applied to implement the capacitive elements of the filter prototype due to their enhanced capacitive effect. This proposed filter shows reduction in size and improvement in terms of loss, cutoff, and stopband. However, its structure and fabrication process are intrinsically complicated.

By taking the approaches above [105, 108], high performance low-pass filters can be designed. However, the design procedure and structure are highly complex and tedious. The conventional implementation for a ladder filter prototype is simple. It consists of the cascade of high- (narrow) and low-impedance (wide) microstrip line

(a) Top view

(b) Cross-sectional view

Figure 3.1: The schematic of a slot-back microstrip line.

sections. It is named stepped-impedance low-pass filter structure [104]. Stepped-impedance low-pass filters are well-known and have been widely used by engineers due to their simple design and low fabrication cost. Unfortunately, it suffers from a flat cutoff, low attenuation of the stopband, and spurious passband that is close to the cutoff. Its disadvantages prevent the filter from being used in advanced communication systems.

The problem at the cutoff and in the stopband of a conventional stepped-impedance low-pass filter can be solved by increasing the ratio of the impedances of the narrow microstrip line section to that of the wide microstrip line section. However, for a

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conventional substrate with a perfect ground plane and lossless dielectric material, this high ratio of the impedance results in a extremely narrow high impedance microstrip line section and/or an extremely wide low impedance microstrip line section. This makes it hard to be realized practically due to the fabrication limitation. Although this problem was addressed in [109], solutions were not provided in this paper. One solution was proposed by milling holes in the substrate to modify the substrate locally and increase the impedance of the narrow microstrip line sections [110]. In this approach, the ratio of impedance is increased and a sharper cutoff, a wider stopband, and a smaller physical size are obtained at the same time. However, it shows much higher insertion loss in the passband due to the milled holes in the substrate.

In terms of increasing the impedance of a microstrip line, the introduction of a longitudinal slot in the ground plane of a narrow microstrip line section (a slot-back microstrip line) has been shown to be an effective means to increase its characteristic impedance and wave propagation constant [111]. In [111], the slot-back microstrip line is used to implement the inductive elements of an elliptic low-pass filter and the modified filters show a sharper cutoff, higher attenuation than the corresponding conventional filters. The characteristic impedance of the slot-back microstrip line is presented in [111] with little detail. Fig. 2 of [111] shows that the characteristic impedance of the slot-back microstrip line increases as the width of the slot increases when the width of a microstrip line is fixed.

For this part of work, in order to improve the performance of a stepped-impedance low-pass filter, the slot-back microstrip line sections with high characteristic impedances were used to implement the inductive sections of the filter. An approach was proposed to obtain the parameters of the slot-back microstrip line with assistances of full wave calculations for the filter design. The design process is simple and the design is easy to implement. The significant increase in the characteristic

impedance of the slot-back microstrip line sections leads to an increase in the ratio of impedances between the narrow microstrip line section and the wide one of a stepped-impedance low-pass filter. With a high ratio, this proposed stepped-impedance low-pass filter shows a sharper cutoff and higher attenuation in the stopband than its conventional counterpart. Due to the etched patch in the ground plane, the suppression of spurious passbands is obtained compared to the conventional filter. This optimization is achieved with a reduction in circuit area and a small insertion loss in the passband.

3.1.1 Design

The design process of the proposed stepped-impedance low-pass filter consists of two steps, obtaining circuit parameters of a slot-back microstrip line and determining the lengths of the microstrip line sections.

3.1.1.1 Parameters of the slot-back microstrip line

Fig. 3.1 (a) shows the top view of a slot-back microstrip line. As can be seen in Fig. 3.1 (a), a longitudinal slot is etched in the ground plane below a conventional microstrip line. The length of the microstrip line and that of the slot in the ground plane are the same and denoted by ℓ . The width of the microstrip line is w_{st} and that of the slot is w_{sl} , $w_{sl} > w_{st}$. The wide slot in the ground plane eliminates the coupling between the microstrip line and the ground plane. It results in increases in both the equivalent inductance and the characteristic impedance of the microstrip line.

Fig. 3.1 (b) shows the cross-sectional view of the slot-back microstrip line. As can be seen in Fig. 3.1 (b), the slot in the ground plane are symmetric with respect to the axis of the microstrip line. In Fig. 3.1 (b), it can be seen that the structure is similar to the elevated coplanar waveguide using air-bridge technique as shown

Figure 3.2: The cross-sectional view of the elevated coplanar waveguide proposed in [29], $w = 12 \mu\text{m}$, $s = 18 \mu\text{m}$, $m = 100 \mu\text{m}$, $h_b = 2.6 \mu\text{m}$, $t = 2.9 \mu\text{m}$, $h_s = 500 \mu\text{m}$.

in Fig. 3.2 [29]. The difference between the two structures is only the sequence of layers. From top to bottom, the slot-back microstrip line in Fig. 3.1 (b) consists of a signal line, a substrate ($\epsilon_r > 1$), a ground plane with a slot filled with air, and an air half space whereas the elevated coplanar waveguide in Fig. 3.2 has a signal line, an air substrate with a thickness of $(h_b - t)$, a ground plane with a slot filled with air (the thickness is t), and a substrate layer of a finite thickness (the thickness is h_s). Thus, the electromagnetic wave behaviors are similar in these two structures and they both show high equivalent characteristic impedances [29, 111]. The analysis methods used in [29] for the elevated coplanar waveguide can be applied to analyze the slot-back microstrip line. The difference of the two structures lies in the fabrications. Due to the elevated center conductor, processes that are more complicated than common printed circuit board fabrications are required for elevated coplanar waveguides. A slot-back microstrip line only requires standard printed circuit board fabrications. The fabrication cost is low and is compatible with the fabrications of other common circuit components. In this study, slot-back microstrip line sections are applied to implement a low-pass filter circuit. An approach is proposed as follows to analyze the slot-back microstrip line with the assistance of full wave calculations.

For a slot-back microstrip line as shown in Fig. 3.1, the characteristic impedance,

3.1 Slot-Back Stepped-Impedance Low-Pass Filter

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Z_{0-sl} , can be obtained from (3.1) and (3.2) below,

$$\tau_o = \frac{\tau_i}{e^{2j\theta}} \quad (3.1)$$

$$Z_{0-sl} = Z_L \frac{1 - \tau_o}{1 + \tau_o} \quad (3.2)$$

where τ_o and τ_i are the reflection coefficients at the output and the input port of the structure, respectively. In (3.1), $\theta = -\beta_{sl}\ell$ is the phase delay at the output port with respect to the input port. In (3.2), Z_L is the load impedance and is normally $50\ \Omega$. In order to obtain Z_{0-sl} based on (3.1) and (3.2), τ_i and θ are needed and they can be obtained from the S-parameters based on full wave calculations of the structure. The other parameter needed, β_{sl} , can be extracted based on θ and the physical length of the slot-back microstrip line under simulation.

3.1.1.2 Lengths of the microstrip line sections

In the design of a conventional stepped-impedance low-pass filter, equations (3.3) or (3.4) below are used to determine the lengths of the low- and the high-impedance microstrip line sections [50], respectively.

$$\beta_\ell \ell_i = g_i \frac{Z_\ell}{R_0} \quad (3.3)$$

$$\beta_h \ell_i = g_i \frac{Z_h}{R_0} \quad (3.4)$$

where the coefficient, g_i , is the i th low-pass prototype value and R_0 is the filter impedance. Based on (3.3) or (3.4), the length of the i th line section, ℓ_i , can be determined by the propagation constant, β , and the characteristic impedance of the transmission line, Z_ℓ or Z_h , at the cutoff frequency of the filter.

In the proposed stepped-impedance low-pass filter design, the length of this slot-back microstrip line was tailored and used to implement the inductive elements in

(a) The proposed filter

(b) The conventional filter

Figure 3.3: The schematic of stepped-impedance low-pass filters.

a ladder filter prototype. The capacitive elements were implemented using wide microstrip line sections, the same as those in a stepped-impedance low-pass filter. Therefore, the proposed filter has stepped impedances and a patterned ground plane. The lengths of the slot-back microstrip line sections are determined based on (3.4) with Z_{0-sl} and β_{sl} at the cutoff frequency substituted in. The lengths of the wide microstrip sections are decided using (3.3).

The slot in the ground plane enhances the inductance and at the same time provides high characteristic impedance for the inductive elements. The slot-back microstrip line sections allow a higher impedance ratio between the narrow and the wide microstrip line section as compared to a conventional filter with a perfect ground plane. Therefore, in the proposed filter design, high performance can be expected.

Table 3.1: Dimension for Stepped-impedance Low-pass Filters (mm)

Stepped-impedance Low-pass Filter	ℓ_1	ℓ_2	ℓ_3	ℓ_4
Proposed filter	3.8	6.3	8.3	7.4
Conventional filter	3.8	11.8	8.3	13.8

3.1.2 Simulated & Measured Results

A seven-pole conventional Chebyshev low-pass filter prototype with 0.01 dB ripple was implemented using the slot-back microstrip line sections. For comparison purposes, both the conventional stepped-impedance low-pass and the proposed slot-back filter were simulated, fabricated, and measured. The cutoff frequency was set at 2 GHz. Taconic with a dielectric constant of 2.46 and a thickness of 30.5 mils was used as the substrate. According to the design process detailed above, a slot-back microstrip line with $w_{st} = 0.6$ mm, $w_{sl} = 10$ mm, and $l = 15$ mm was simulated using Advanced Design System [30]. Based on the simulation results, (3.1), and (3.2), Z_{0-sl} and β_{sl} were found to be 167.0Ω and 65.8 radian/m at 2 GHz. For comparison purposes, the characteristic impedance and wave propagation constant of the same microstrip line over a perfect ground plane Z_0 and β were obtained to be 101Ω and 41.8 radian/m at the same cutoff frequency [112]. Comparing the two characteristic impedances, the characteristic impedance of a microstrip line is increased significantly by introducing a longitudinal slot in the ground plane. It is also noticed that, with the etched patch in the ground plane, the propagation constant of the microstrip line is increased considerably, which indicates the significant slow-wave effect at the cutoff. Based on (3.4), the high propagation constant implies a reduction in physical size when the filter is implemented using slot-back microstrip line sections.

Fig. 3.3 (a) shows the schematic of the proposed stepped-impedance low-pass filter

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and Fig. 3.3 (b) shows the corresponding conventional stepped-impedance low-pass filter for comparison purposes. The width of the microstrip line at the input and the output $w_{50\Omega}$ was set to be 2.2 mm corresponding to a characteristic impedance of 50Ω at 2 GHz for both structures.

In the proposed structure in Fig. 3.3 (a), the width of the wide and the narrow microstrip line sections, w_l and w_h , were set to be 10 mm and 0.6 mm, respectively. The slot-back microstrip line section was applied to implement the inductive elements in a standard stepped-impedance low-pass filter prototype. The width of the slot in the ground plane, w_{sl} , was set to be the same as that of the wide microstrip line section, w_l , whereas the width of the microstrip line, w_{st} , was set to be the same as that of the narrow microstrip line section, w_h . With $w_{sl} = 10$ mm and $w_{st} = 0.6$ mm, $Z_{0-sl} = 167.0\Omega$ and $\beta_{sl} = 65.8$ radian/m based on the previous extraction. The length of the wide microstrip line sections and the slot-back microstrip line sections are determined using (3.3) and (3.4), respectively. They were tabulated in Table 3.1.

In the conventional structure in Fig. 3.3 (b), the width of the wide and the narrow microstrip line sections are w'_l and w'_h , respectively. For comparison purposes, $w_l = w'_l$ and $w_h = w'_h$. For the conventional stepped-impedance low-pass filter, the lengths of the wide and the narrow microstrip line sections were determined applying (3.3) and (3.4) again with the parameters of the microstrip line sections [112]. The lengths, ℓ_1 to ℓ_4 , are also tabulated in Table 3.1.

In Table 3.1, the lengths of the slot-back microstrip line sections in the proposed filter are reduced significantly from those of the narrow microstrip line sections in the conventional filter. The total length of the new filter is reduced by 24.1 % from the conventional filter. The reduction in length is due to the significant slow-wave effect of the slot-back microstrip line.

The proposed stepped-impedance low-pass filter and the conventional filter were

Figure 3.4: The simulated and measured S_{21} -parameters of the proposed and the conventional stepped-impedance low-pass filter and the simulated S_{21} -parameter of their corresponding lumped-circuit elements.

both simulated. Fig. 3.4 shows the simulated S_{21} -parameters of the two filters and the simulated S_{21} -parameters of the filter prototype lumped-circuit. As seen in Fig. 3.4, these two structures are working at the same cutoff frequency at 1.9 GHz. No shifting of frequency is introduced by replacing the narrow microstrip line sections with the slot-back microstrip line sections. As compared to the conventional filter, the proposed filter shows a frequency response that is more similar to that of the lumped-circuit. This is because the slot-back microstrip line provides a high characteristics impedance, therefore, a better approximation to the inductive lumped-circuit elements. As shown in Fig. 3.4, the proposed filter has a high selectivity of 33.8 dB/GHz whereas the conventional filter has a selectivity of 27.7 dB/GHz. The proposed filter shows a sharper cutoff performance. It is also observed that the spurious passband of the conventional filter is effectively suppressed in the proposed filter. When comparing their ripple levels in the passbands, the proposed filter has a ripple level of 0.19 dB, slightly higher than the ripple level

(a) The microstrip line

(b) The ground plane

Figure 3.5: The photographs of the fabricated proposed and conventional stepped-impedance low-pass filter.

Figure 3.6: The cross-sectional views of microstrip line sections

of the conventional filter at 0.16 dB. This shows that the etched patches in the ground plane do not generate much additional insertion loss in the passband.

The proposed stepped-impedance low-pass filter and the conventional filter were fabricated and tested. Fig. 3.5 (a) shows the photographs of their microstrip line planes. The upper structure in Fig. 3.5 (a) is the conventional filter and the lower one is the proposed filter. Fig. 3.5 (b) shows the photographs of their ground planes. As can be seen in Fig. 3.5 (b), the conventional filter has a perfect ground plane whereas the proposed filter has slots in the ground plane (a patterned ground plane). As seen in Fig. 3.5, the physical size of the proposed low-pass filter is much smaller as compared to that of the conventional filter. The measured S_{21} -parameters for the two low-pass filters are shown in Fig. 3.4. The measured results are in good agreement with the simulated results for both structures. The fabricated slot-back stepped-impedance low-pass filter shows a ripple level of 0.27 dB in the passband and a selectivity of 33.5 dB/GHz whereas the fabricated conventional low-pass filter shows a ripple level of 0.19 dB in the passband and a selectivity of 27.4 dB/GHz. Experimentally, the proposed filter shows superior filtering performance to the conventional one.

3.1.3 Analysis & Discussions

In Fig. 3.4, the conventional stepped-impedance low-pass filter shows spurious modes at frequencies higher than 7 GHz, which has been well known as one of the disadvantages of this type of filters. It is also observed that the application of the slot-back microstrip line sections to the stepped-impedance low-pass filter helps to eliminate the spurious modes. In the implementations of a low-pass filter prototype, the equivalent lumped-element values of the transmission line sections that are used to implement the inductors and capacitors depart significantly from those in the filter prototype at frequencies far away from the cutoff frequency. Transmissions at these frequencies are normally possible and these transmission modes are called spurious modes. One of the reasons for the eliminations of spurious modes by using the slot-back microstrip line sections is that the filter with slot-back microstrip line sections has its second passband at higher frequencies compared to the conventional filter. This is due to the introduction of the slots which may be modeled using a circuit with a higher order. Another reason is the high attenuation of the slot-back microstrip line at high frequencies, which is to be detailed as follows.

Fig. 3.6 (a) shows the E-fields of the transmission of waves through the narrow microstrip line section by using its cross-sectional view. As shown in Fig. 3.6 (a), the coupling between the signal line and the ground plane is always possible because the ground plane is perfect. However, if there is a slot in the ground plane, such as the slot-back microstrip line, the coupling between the signal line and the ground plane depends on the width of the slot and the working frequency. Fig. 3.6 (b) shows the cross-sectional view of a slot-back microstrip line. In Fig. 3.6 (b), the horizontal distance between the microstrip line and the edge of the ground plane is denoted as d . The electrical length of d increases as frequency increases. This prevents the waves from coupling between the microstrip line and the ground plane.

Figure 3.7: The simulated S_{21} -parameters of the slot-back microstrip line in Fig. 3.1 with $w_{st} = 0.6$ mm and w_{sl} varies.

Thus, the transmissions of waves become difficult when the frequencies are high. Due to the different coupling mechanisms, the slot-back microstrip line may have higher attenuation at high frequencies than a normal microstrip line.

The transmissions of the slot-back microstrip lines with the width of the microstrip line fixed and the width of the slot varied are examined using simulated results [30]. Fig. 3.7 shows the simulated S_{21} -parameters of the slot-back microstrip line with $w_{st} = 0.6$ mm and $w_{sl} = 10$ mm, 15 mm, and 20 mm. The input and output characteristic impedances are set to be 50Ω while the characteristic impedances of the transmission lines varies due to the different slot widths. Thus, there are impedance mismatches at the ports which are similar to the inputs and outputs of the slot-back microstrip line sections in the proposed filter structure as shown in Fig. 3.3 (a). The results of a microstrip line with a characteristic impedance of 50Ω are included as a reference.

As can be seen in Fig. 3.7, the transmission of the microstrip line is close to 0 dB in the frequency range of interest, which is mainly due to the matched impedance at

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the ports. In Fig. 3.7, the slot-back microstrip lines show attenuation and ripples in the transmissions. As compared to a normal microstrip line, the attenuation of the slot-back microstrip lines are considerable. The attenuation of each slot-back microstrip line increases as frequency increases. This is because the electrical length of d increases as frequency increases and as a result, the coupling between the microstrip line and the ground plane is eliminated and the transmissions are attenuated. Comparing the structures of different slot widths at the same frequency, it is observed that the attenuation increases as the slot width increases. The increase in attenuation is also due to the large electrical length of d . Moreover, as a combined effect of the physical distance and the frequency, it is observed that, as frequency increases, the attenuation of the transmission of the structure with a bigger slot width increases faster than that of the structure with a smaller slot width.

The results in Fig. 3.7 reveal that the transmission of a slot-back microstrip line is attenuated significantly when the slot width is large at high frequencies. The results agree with the previous analysis on the structures. Therefore, when the slot-back microstrip line sections are applied to the stepped-impedance low-pass filter, the spurious modes at high frequencies are attenuated.

Patterned ground structures of different types are useful to eliminate spurious passbands of microwave components. Besides the slot-back microstrip structure introduced above that eliminates the spurious modes of the stepped-impedance low-pass filter, patterned ground structures with single unit cells or periodic cells (EBG structures) are useful to suppress spurious transmission at high frequencies. Examples are given in the literature [34, 35]. In [35], a patterned ground structure with a single cell was applied to suppress the second harmonic of an active integrated microstrip antenna at the feeding whereas in [34], an EBG structure was

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applied to class AB GaAs field-effect transistor (FET) power amplifiers to realize harmonic suppressions. Moreover, EBG structures are applied to the designs of filters [113, 114] to suppress spurious passbands in the stopband. When patterned ground structures are used for harmonic suppressions, the stopband is designed overlapping the spurious passbands according to different design processes presented in Chapter 2.

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In an EBG microstrip structure, the periodic cells along the microstrip line are used to modulate the microstrip line, thus locally altering its effective permittivity. The modulation creates alternate passbands and stopbands in the frequency response of the structure. These periodic cells can be in the form of etched patches in the ground plane of a microstrip line with a 2-D [12] or a 1-D pattern [38]. They can also be in the form of patches inserted in the microstrip line [32, 114, 113] which is known as periodic structure with shunt-capacitive loading [114] or capacitively modulated transmission line [113]. Grating structures have been applied in the field of optics for decades [115]. Lately, they have been applied in millimeter waves technology [116] and CPW technology [117]. An EBG microstrip structure can be seen as the applications of gratings in microstrip line technology. It satisfies the Bragg reflection condition [40]. EBG structures are applied to the designs of bandpass filters [113] and ultra-wideband (UWB) bandpass filters [114] to suppress spurious passbands in the stopband. EBG structures are widely used in filter designs. However, full-wave simulations, including numerical methods [118] and commercial softwares [2, 30, 119], are required to predict the stopband performance of the structure, such as the bandwidth, center frequency, and attenuation with

Figure 3.8: The schematic of a capacitively modulated EBG microstrip structure.

high accuracy. The simulations consume a large amount of time and computational resources.

In this part, an approach is proposed to provide an accurate and time-saving calculation of the transmission coefficients of an EBG microstrip structure where square patches are periodically inserted into the microstrip line (capacitively modulated EBG microstrip structure) as shown in Fig. 3.8. The stopband performance, such as the center frequency, bandwidth, and attenuation, of these EBG structures can be predicted with high accuracy through the calculation of the transmission coefficient. The periodic waves of the structure are analyzed and the dispersion relation of electromagnetic waves in the structure is derived, plotted, and analyzed. The grating nature of the structure is demonstrated. This approach can be applied to periodic microstrip structures of a similar nature to simplify the analysis and design procedures when the unit cell is properly modeled using lump circuit elements. It can easily be generalized and applied to facilitate and accelerate the computer-aided designs of periodic microstrip structures of this type.

Figure 3.9: (a), (b) microstrip step discontinuities and (c), (d) its equivalent circuits.

3.2.1 The EBG Structure under Study

An EBG structure is an application of gratings into a microstrip line. It satisfies the Bragg reflection condition,

$$\lambda = 2a \cdot n \quad (3.5)$$

where λ is the wavelength of the reflected wave, a is the distance between two successive cells (defined as the period of the structure), and n is the refractive index $n = \sqrt{\epsilon_r \mu_r}$. Therefore, with a fixed period of the structure, the center frequency of its stopband can be estimated using (3.5). However, equation (3.5) provides no other information about the stopband, such as the attenuation and the bandwidth.

Fig. 3.8 shows the schematic of the capacitively modulated EBG microstrip structure under study. It is the same as the EBG structure shown in Fig. 2.28 (a) with additional flexibility to change the width of the microstrip line at two ends of a unit cell (w'_h). The calculation of the transmission coefficients for the structure is done by obtaining the $ABCD$ matrix and the characteristic impedance of the periodic structure. In Section 3.2.2, an equivalent circuit is proposed to obtain an $ABCD$ matrix of a unit cell of the structure, according to its geometry. An overall $ABCD$ matrix of the periodic EBG structure can then be obtained by cascading

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several unit cells. In Section 3.2.3, periodic waves in the structure are analyzed in order to obtain the characteristic impedance. With an $ABCD$ matrix and the characteristic impedance of the periodic structure, its transmission coefficient at each frequency can be calculated. The dispersion relation of the structure can be expressed and plotted.

3.2.2 The $ABCD$ Matrix of a Unit Cell

Fig. 3.8 shows a capacitively modulated EBG microstrip structure with four cells ($m = 4$). An EBG unit cell was taken out with terminal planes 1 and 2 midway between two successive inserted patches. In order to obtain the $ABCD$ matrix for the unit cell shown in Fig. 3.8, the unit cell was further divided into five sections of length d_i ($i = 1 - 5$) that are cascaded; the first section is a narrow microstrip line section with a length of $d_1 = (a - l'_i)/2$; the second section takes into account the narrow to wide discontinuity where $d_2 \approx 0$; the third section is a wide microstrip line section with $d_3 = l'_i$; the fourth section is another discontinuity from wide to narrow with $d_4 \approx 0$; and the last section is a narrow microstrip line section with $d_5 = d_1 = (a - l'_i)/2$.

The $ABCD$ matrix of a microstrip line section (in this content, the first, third, or fifth sections) can be expressed as [50],

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{mi} = \begin{bmatrix} \cos\beta l & jZ_0 \sin\beta l \\ jY_0 \sin\beta l & \cos\beta l \end{bmatrix} \quad (3.6)$$

where the subscript mi denotes the i^{th} microstrip line section. The $ABCD$ matrix of the two microstrip step discontinuity (narrow to wide and wide to narrow) is obtained as follows. Fig. 3.9 (a) and (b) shows the step-up discontinuity (narrow to wide) and the step-down discontinuity (wide to narrow), respectively. In Fig. 3.9 (a)

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and (b), W_{i-1} and W_i are the widths of the microstrip line sections from left to right. The subscripts, $(i+1)$ and i , are the indices for the microstrip line sections. Each of them can be modeled using the T-network as shown in Fig. 3.9 (c) [112]. As can be seen in Fig. 3.9 (c), L_{i-1} and L_i respectively correspond to the inductance of the microstrip line section with a width of W_{i-1} and of W_i of the step discontinuities that are shown in Fig. 3.9 (a) and (b). The excess capacitance due to the step discontinuity is C_s . The inductances, L_{i-1} and L_i can be estimated using the equations (3.7) to (3.10) below.

$$L_s = 0.000987h \left(1 - \frac{Z_{0W}}{Z_{0N}} \sqrt{\frac{\epsilon_{reW}}{\epsilon_{reN}}} \right)^2 (nH) \quad (3.7)$$

$$L_{i-1} = \frac{L_{W_{i-1}}}{L_{W_{i-1}} + L_{W_i}} L_s \quad (3.8)$$

$$L_i = \frac{L_{W_i}}{L_{W_{i-1}} + L_{W_i}} L_s \quad (3.9)$$

$$L_{W_i} = \frac{Z_{0mi} \sqrt{\epsilon_{rei}}}{c} (H/m) \quad (3.10)$$

where Z_{0W} and ϵ_{reW} are the characteristic impedance and the effective dielectric constant of the wide microstrip line section, respectively, whereas Z_{0N} and ϵ_{reN} are those of the narrow microstrip line section. The thickness of the substrate h is in micrometer. Z_{0m} and ϵ_{re} are the characteristic impedance and the effective dielectric constant of the microstrip line, respectively. In (3.10), L_{W_i} is the inductance per unit length of the microstrip line of width W_i . Z_{0mi} and ϵ_{rei} are the characteristic impedance and the effective dielectric constant of the microstrip line of width W_i , respectively. c is the speed of light in free space. The excess capacitance C_s can be estimated by using (3.12) below,

$$C_s = 0.00137 \frac{\sqrt{\epsilon_{reW}}}{Z_{0W}} \left(1 - \frac{W_N}{W_W} \right) h \left[\frac{\epsilon_{reW} + 0.3}{\epsilon_{reW} - 0.258} \right] \left[\frac{W_W/h + 0.264}{W_W/h + 0.8} \right] (pF) \quad (3.12)$$

Figure 3.10: The equivalent T-network for the unit cell of a capacitively modulated EBG microstrip structure in Fig. 3.8.

where h is in micrometer, W_N and W_W are the widths of the narrow and the wide microstrip line sections, respectively.

The T-network in Fig. 3.9 (c) is presented in an alternative way using reactance Z_{i-1} and Z_i and a susceptance Y and is shown in Fig. 3.9 (d). Therefore, the $ABCD$ matrix of this network for the step discontinuity can be obtained as shown below,

$$\begin{aligned} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{si} &= \begin{bmatrix} 1 & Z_{i-1} \\ 0 & 1 \end{bmatrix} \begin{bmatrix} 1 & 0 \\ Y & 1 \end{bmatrix} \begin{bmatrix} 1 & Z_i \\ 0 & 1 \end{bmatrix} \\ &= \begin{bmatrix} 1 + Z_{i-1}Y & Z_i + Z_i Z_{i-1}Y + Z_{i-1} \\ Y & Z_i Y + 1 \end{bmatrix} \end{aligned} \quad (3.13)$$

where the subscript, si denotes the i^{th} step discontinuity. Since the five sections are cascaded to form a unit cell, the $ABCD$ matrix for the unit cell can be obtained by multiplying the $ABCD$ matrix of each section as follows,

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix}_c = \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{m1} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{s1} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{m2} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{s2} \begin{bmatrix} A & B \\ C & D \end{bmatrix}_{m3} \quad (3.14)$$

where the subscript c denotes cell. $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{m1}$, $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{m2}$, and $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{m3}$

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can be obtained by using (3.6) whereas $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{s1}$ and $\begin{bmatrix} A & B \\ C & D \end{bmatrix}_{s2}$ can be obtained by using (3.13). Therefore, the overall $ABCD$ matrix of the EBG structure as shown in Fig. 3.8 can be obtained by cascading those of the four unit cells by using (3.14). Based on the $ABCD$ matrix, two-port network parameters, such as the $[Z]$ matrix and the $[Y]$ matrix of the structure can be obtained readily. For the prediction of the stopband performance, or in other words, for obtaining the S-parameters at each frequency, the characteristic impedance of the periodic structure Z_c is required. In the following section, Z_c is obtained for the capacitively modulated EBG microstrip structure by analyzing the periodic waves in the structure. The S-parameters are calculated and plotted. The dispersion relation of the structure is obtained and plotted. It is used to show the grating nature of the EBG microstrip structure.

3.2.3 Characteristic Impedance & Dispersion Relation of an EBG Microstrip Structure

The characteristic impedance of the EBG structure in Fig. 3.8 can be obtained by analyzing periodic waves in the structure [120]. The unit cell shown in Fig. 3.8 has a reflection coefficient Γ_1 and Γ_2 at the two terminal planes 1 (input) and 2 (output), respectively. In Fig. 3.8, c_1 and b_1 are the amplitude of the forward- and the backward-propagating wave at the input, respectively, whereas c_2 and b_2 are the amplitudes of the waves at the output. The transmission coefficients between the two terminal planes are τ_{12} and τ_{21} . Due to the reciprocity of the unit cell, $\tau_{12} = \tau_{21}$. The wave amplitudes are related using a wave matrix [120]. For a capacitively modulated EBG microstrip line structure, it consists of cascaded periodic unit cells. With a wave propagating down the structure, c_1 and c_2 , b_1 and

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b_2 are related by (3.15) and (3.16), respectively, as shown below,

$$c_2 = e^{-\gamma_c a} c_1 \quad (3.15)$$

$$b_2 = e^{-\gamma_c a} b_1 \quad (3.16)$$

where $\gamma_c = \alpha_c + j\beta_c$ is the propagation constant and the periodic structure is assumed to be infinite.

Besides the wave matrix, the amplitudes, c_1 , c_2 , b_1 , and b_2 can also be related through voltages and currents of the network. Fig. 3.10 shows the equivalent T-network for the unit cell with input and output voltages and currents. The relation between c_1 and b_1 and that between c_2 and b_2 are expressed as follows,

$$V_1 = c_1 + b_1 \quad (3.17)$$

$$V_2 = c_2 + b_2 \quad (3.18)$$

$$I_1 = (c_1 - b_1)Y_w \quad (3.19)$$

$$I_2 = (c_2 - b_2)Y_w \quad (3.20)$$

where Y_w is the wave admittance. Based on (3.17)–(3.20), wave matrix can be translated into impedance matrix $[Z]$. With algebraic manipulations, the characteristic impedance of the structure can be expressed as shown in (3.21) below [120],

$$Z_c = \frac{Z_{11} - Z_{22}}{2} + Z_{12} \sinh \gamma_c a \quad (3.21)$$

where $\sinh \gamma_c a = (\cosh^2 \gamma_c a - 1)^{\frac{1}{2}}$. The relation between γ_c and $[Z]$ can be expressed using (3.22) [120].

$$\cosh \gamma_c a = \frac{Z_{11} + Z_{22}}{2Z_{12}} \quad (3.22)$$

Based on (3.14) and (3.21), the impedance of the EBG microstrip structure is

Figure 3.11: The simulated and calculated S_{21} -parameters of the capacitively modulated EBG microstrip structure ($f_c = 3.1$ GHz).

related to the geometry of the structure and can be obtained. Therefore, the S-parameters of the EBG structure can be calculated based on the dimensions of the structure. Equation (3.22) expresses the dispersion relation of the EBG structure.

3.2.4 Prediction on S-Parameters

A four-cell capacitively modulated EBG microstrip structure as shown in Fig. 3.8 with a center frequency of the stopband at 3 GHz was designed. Taconic ($\epsilon_r = 2.46$, $h = 30.5$ mils) was used as the substrate. The width of the microstrip line at the input and the output port was set to correspond to a characteristic impedance of 50Ω at 3 GHz. The period of the structure a was determined by applying (3.5). Both l_l' and w_l' were set to $0.5a$ corresponding to an optimal filling factor of 0.5 [12]. The width of the microstrip line section between two successive inserted patches w_h' was set to 0.6 mm. The EBG microstrip structure was simulated using the Method of Moment based software, Advanced Design System (ADS) [30]. The

solid line in Fig. 3.11 shows its simulated S_{21} -parameters. The center frequency and the 15-dB bandwidth of the stopband were found to be 3.1 GHz and 3.7 GHz, respectively. The attenuation at the center of the stopband is 67.6 dB. The lower and the upper 15-dB-cutoff frequencies of this EBG structure are found to be 1.2 GHz and 4.9 GHz, respectively.

In order to calculate the dispersion relation of this EBG structure, Z_{0m} and ϵ_{re} of the narrow and wide microstrip line sections were estimated using Linecalc in ADS at different frequencies based on their physical dimension [30]. The propagation constant of the microstrip lines, β_n and β_w , were determined accordingly. By applying (3.6)–(3.14), the $ABCD$ matrix for the unit cell of the capacitively modulated EBG microstrip structure was obtained. The S-parameters of the structure were calculated based on the $ABCD$ matrix and (3.21) for the characteristic impedance Z_c . Fig. 3.11 shows the calculated S_{21} -parameters. The calculated S_{21} -parameters without taking into consideration of the step discontinuities in (3.14) are included for comparison. The center frequencies and 15 dB-cutoff frequencies of the three curves in Fig. 3.11 are compared to evaluate the accuracy of the calculation. As can be seen in Fig. 3.11, for the curve that takes into account the step discontinuities, its lower and the upper 15 dB-cutoff frequencies are at 1.2 GHz and 4.8 GHz, respectively. The center frequency of the stopband is at 3.1 GHz. These calculated results are in a good agreement with the simulated results (15 dB-cutoff frequencies at 1.2 GHz and 4.9 GHz, the center frequency at 3.1 GHz). For the curve without taking into account the step discontinuities, its 15 dB-cutoff frequencies are at 1.3 GHz (lower) and 5.1 GHz (upper). It has the stopband centered at 3.2 GHz. Compared with the simulation results, it shows a shift in the center frequency of the stopband and an increase in the bandwidth. The comparison shows the importance of the step discontinuity for the calculation of the stopband performance.

Figure 3.12: Dispersion relation of the capacitively modulated EBG microstrip structure in Fig. 3.8.

The results reveal that the proposed approach for deriving the stopband performance of a capacitively modulated EBG microstrip structure is of high accuracy. It provides information on the center frequency, the bandwidth, and the attenuation of the stopband. The accuracy of the prediction is increased by taking into account the effect of the step discontinuities in the structure. This approach can be further extended to other periodic microstrip structure by modifying the circuit model in Section 3.2.2.

3.2.5 Plot of Dispersion Relation

The dispersion relation for the capacitively modulated EBG microstrip structure is calculated using (3.22) and plotted in Fig. 3.12. As can be seen in Fig. 3.12, the EBG structure shows maximum β_c at both edges of the stopband, which is similar to a typical dispersion relation of a periodically loaded transmission line [121, Chapter 2, p. 137 - 260]. The parameters of the structure will be analyzed in detail to examine its grating nature.

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Due to the lossless network that is assumed in every section of the unit cell, Z-parameters are all purely imaginary. Thus, $\cosh\gamma_c a$ in (3.22) is a real number. Since $\cosh\gamma_c a = \cosh[(\alpha_c + j\beta_c)a] = \cosh\alpha_c a \cos\beta_c a + j\sinh\alpha_c a \sin\beta_c a$, there are two possible cases:

- Case 1: $\alpha_c \neq 0$, $\beta_c = 0$ ($\cosh\gamma_c a > 0$) or $\beta_c a = \pi$ ($\cosh\gamma_c a < 0$). The wave is evanescent. In the short distance it propagates, there is no change in phase when $\beta = 0$ and there is a change in phase when $\beta = \pi$. It defines the stopbands of the structure.
- Case 2: $\alpha_c = 0$, $\beta_c \neq 0$. The wave is propagating without attenuation. It defines the passbands of the structure.

Fig. 3.12 shows a plot of the calculated $\beta_c a$ versus $k_0 a$. As can be seen in Fig. 3.12, there is a prominent stopband at a center frequency of 3.1 GHz.

Table 3.2 summarizes the passband and stopband characteristics of the capacitively modulated EBG microstrip structure based on calculated results. A and B are the parameters in the $ABCD$ matrix. They are related to γ_c , β_c , and Z_c by the expressions,

$$A = \frac{e^{\alpha_c l} + e^{-\alpha_c l}}{2} \cos(\beta_c l) + j \frac{e^{\alpha_c l} - e^{-\alpha_c l}}{2} \sin(\beta_c l) \quad (3.23)$$

$$B = jZ_c \left[\frac{e^{\alpha_c l} + e^{-\alpha_c l}}{2} \sin(\beta_c l) - j \frac{e^{\alpha_c l} - e^{-\alpha_c l}}{2} \cos(\beta_c l) \right] \quad (3.24)$$

where l is the length of the transmission line. Z_c in (3.21) is related to A-parameter as below for a reciprocal network ($Z_{11} = Z_{22}$),

$$Z_c = Z_{12}(A^2 - 1)^{\frac{1}{2}} \quad (3.25)$$

Equation (3.25) is obtained based on $Z_{11} = A/C$, $Z_{12} = (AD - BC)/C$, and $AD - BC = 1$.

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Table 3.2: Summaries on Characteristics of a Capacitively Modulated EBG

	α_c	β_c	A	B	Z_c
Passband	$= 0$	$\neq 0$	$ A < 1$	Purely imaginary	Real
Stopband	$\neq 0$	$\beta_c = 0$ $/\beta_c a = \pi$	$ A > 1$	Purely imaginary	Purely imaginary

As shown in Table 3.2, besides the differences in α_c and β_c between the passband and the stopband, it is observed that A-parameters are in different ranges of values for different frequency ranges, B-parameters are purely imaginary at all frequencies, and Z_c s are real in the passband and purely imaginary in the stopband. Based on (3.23) – (3.25), it is the differences in γ_c and β_c that lead to the different ranges of values for A-parameters. The different A-parameters result in the real Z_c in the passband and the purely imaginary Z_c in the stopband. The characteristics of γ_c , β_c , and Z_c result in the purely imaginary B-parameters in both the passband and the stopband.

In the capacitively modulated EBG microstrip structure under study, the characteristics of the parameters in the passband and the stopband, such as the propagation constant, the characteristic impedance, the A- and the B-parameters, are the same as the periodic coplanar waveguides with inductive loading as reported in [117]. The dispersion relation and the analysis in this section show the grating nature of the EBG structure. The propagation constants of the EBG microstrip structure in the passband and those of the periodically loaded CPW [117] have different trends versus frequency. The EBG microstrip structure has the maxima of β_c at both edges of the stopband while the CPW structure has the maximum value only at the lower edge of the stopband (Fig. 3 (a) and Fig. 4 (a) in [117]).

3.3 Summary

In this chapter, the effects of ground planes with etched patterns on the propagations of electromagnetic waves at microwave frequencies have been further studied. An optimization has been performed on a conventional stepped-impedance low-pass filter by using a microstrip line with a longitudinal slot in the ground plane (a slot-back microstrip line). A slot-back microstrip line has a high equivalent inductance and a high characteristic impedance. Due to the high characteristic impedance of the slot-back microstrip line, when it is applied to implement the inductive components in a ladder low-pass filter prototype, the filter has a high impedance ratio between the narrow and the wide microstrip line sections, which leads to a sharper cutoff, higher attenuation in the stopband, and a better suppression of the spurious passband, as compared to a conventional stepped-impedance low-pass filter. Due to the slow-wave effect of the slot-back microstrip line, the optimized filter has a small physical size. Moreover, its insertion loss in the passband is low. The formulae and the procedure for the optimization have been presented. This optimization is simple and the design is easy to implement.

An approach that provides an accurate and time-saving calculation of the transmission coefficients of a capacitively modulated EBG microstrip structure has been presented. It has been shown that the model of the step discontinuity in the structure plays an important role in the calculation. The stopband of the structure can be predicted with high accuracy. Furthermore, the dispersion relation of the electromagnetic waves in the structure has been derived, plotted, and analyzed. The grating nature of the structure has been shown. This proposed approach helps to simplify the analysis and design procedures of EBG microstrip structures. With proper modifications to the circuit model, the approach can be extended and applied to the calculations of other periodic microstrip structures and other patterned ground structures.

Chapter 4

Effects of Ground Plane

Geometries on Coupling and Radiation

In a patterned ground structure, a single patch or periodic patches are etched in the ground plane to prohibit the propagation of electromagnetic waves in certain frequency bands so as to provide filtering functionality. However, the existence of the etched patch(es) with the dimension less than half of the guided wavelength in the ground plane becomes a concern in terms of electromagnetic compatibility (EMC) and electromagnetic susceptibility (EMS) when a patterned ground structure is integrated into a microwave circuit. The etched patch(es) in the ground plane make the structures prone to electromagnetic interference from nearby circuit components as compared to a structure with a perfect ground plane. Moreover, a patterned ground structure such as an EBG structure can possibly generate radiation that interferes other circuit components, which is not a concern for a structure with a perfect ground plane.

There have been studies on the EMC of a patterned ground structure. It was re-

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ported that the scattering parameters (S-parameters) of a patterned ground structure with periodic etched patterns (an EBG structure) are affected by a uniform shielding plate in close proximity [87]. In [22], the susceptibility of an EBG microstrip structure with periodic dumbbell-shaped etched patterns in the ground plane was studied. The effect of a metallic enclosure on the S-parameters of a shielded EBG structure and that of a finite ground plane on the performance of an unshielded EBG structure were both studied. The effects were reported to be significant. The paper suggests the use of a metallic enclosure with a finite ground plane to shield patterned ground structures from interference. However, this makes the circuit bulky and complex. For the applications of a patterned ground structure to microwave circuits without introducing additional shielding mechanisms, the knowledge on the EMC and EMS of the structure to the circuit components nearby is important. The knowledge on the radiation of the structures is equivalently important for their integrations to a microwave circuit.

In this chapter, the electromagnetic susceptibilities of a dual planar EBG microstrip structure to radiation or coupling nearby are studied. The studies of the EMS of an EBG structure to a radiating source and a microstrip line are presented in Section 4.1 and 4.2, respectively. In Section 4.3, a study on the radiation generated by patterned ground structures is presented. The effects of the etched patterns in the ground plane on the radiation generated by the structures in the far-field region are examined.

4.1 **EMS to a Nearby Radiating Component**

In this section, the EMS of a dual planar EBG microstrip structure as shown in Fig. 2.19 to a nearby radiating circuit component, a broadband horn antenna, was studied based on experimental results.

(a)

(b)

Figure 4.1: (a) The 3-D view of a dual planar EBG microstrip structure, (b) the simulated (using ADS [30]) and measured results of the EBG structure.

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4.1.1 Device under Test

The device under test (DUT) in this part of the study is the dual planar EBG microstrip structure shown in Fig. 2.19. Fig. 2.19 is copied and shown in Fig 4.1 (a) for the convenience of reading. The DUT was fabricated using Taconic ($\epsilon_r = 2.43$, $h = 30$ mils) as the substrate of the structure. The center frequency of the stop-band f_0 was set to be 10 GHz. Thus, the periods of the structure, a_1 and a_2 , were determined to be 10.35 mm according to the Bragg reflection condition. The width of the microstrip line (MLIN) w was set to be 2.29 mm corresponding to a characteristic impedance of 50Ω at 10 GHz. The width and the length of the inserted patch in the microstrip line, w_a and l_a , were both set to be 5 mm and the radius of the etched circular patches was 2.59 mm. The structure was simulated [30], fabricated, and tested. Fig 4.1 (b) shows its simulated and measured S-parameters. As shown in Fig 4.1 (b), the measured results are in good agreement with the simulated results. The ripples in the passband can be suppressed by applying tapering techniques [A 8]. An insertion loss (ripple level) of less than 1 dB in the passband is achievable in a tapered dual planar EBG structure [A 8]. In order to simplify this study on EMS, the DUT is a non-tapered EBG structure.

4.1.2 Experimental Setup

In this part of the study, the DUT was subjected to a radiating interference source to examine its EMS. A signal was transmitted through the DUT while a broadband horn antenna (1–18 GHz, gain = 11 dBi, linear) was used to generate the radiating interference. Fig. 4.2 (a) shows the cross-sectional view of the DUT with the horn antenna. As can be seen in Fig. 4.2 (a), an interference source is placed d away from the plane under test (PUT) with an angle of incidence at θ (with respect to the PUT). The DUT was tested within both the near-field (Fresnel region) and the far-field (Fraunhofer region) region of the interference source. The inner boundary

R_1 and outer boundary R_2 of the radiating near-field region can be approximated using (4.1) and (4.2), respectively [99].

$$R_1 = 0.62\sqrt{D^3/\lambda} \quad (4.1)$$

$$R_2 = 2D^2/\lambda \quad (4.2)$$

where D is the maximum aperture dimension of the antenna and it must be greater than the wavelength λ . For testing in the near-field and the far-field region, d was set such that $R_2 > d > R_1$ and $d > R_2$, respectively. The angle θ defines the angle of incidence of the radiation with respect to the PUT. It is positive when the microstrip line plane is exposed to the interference and is negative when the ground plane is exposed to the interference. In Fig. 4.2 (a), θ equals to 90° .

Fig. 4.2 (b) shows the top view of the DUT. The polarization of the antenna is indicated by a double-headed arrow. As shown in Fig. 4.2 (b), the angle between the line of symmetry of the DUT a and the arrow is denoted by ϕ . It defines the relative polarization between the radiating interference source and the signal transmission in the DUT. In Fig. 4.2 (b), $\phi = 0^\circ$.

Fig. 4.3 shows the block diagram of the EMS test on the DUT. A frequency modulated (FM) sinusoidal signal was generated using signal generator 1 and transmitted through the DUT at a carrier frequency f_c . This FM signal is observed and demodulated at the output of structure using a spectrum analyzer. Another signal was generated using signal generator 2 and radiated using the horn antenna at a frequency of f_i , generating a radiating interference. When the DUT was subjected to this radiating interference, the noise peak-to-peak voltage of the demodulated signal at the output of the DUT was recorded.

A relative noise peak-to-peak voltage of the demodulated signal is used as a parameter to evaluate the EMS characteristics of the DUT. It is denoted using r_n

Figure 4.2: The experimental setup for the EMS test

and it is defined as a ratio between the measured noise peak-to-peak voltage of the demodulated signal when the DUT is exposed to an interference and that of the signal when there is no interference (taken as reference):

$$r_n = \frac{v_{max} - v_{min}}{v'_{max} - v'_{min}} \quad (4.3)$$

where $(v_{max} - v_{min})$ is the measured noise peak-to-peak voltage when the DUT is exposed to an interference and $(v'_{max} - v'_{min})$ is the measured noise peak-to-peak voltage when the DUT is not exposed to an interference. A high r_n implies high noise level due to the interference. The additional noise is considered to be insignif-

Figure 4.3: Block diagram of the EMC test setup.

Figure 4.4: The normalized simulated $|S_{21}|$ and the normalized noise peak-to-peak (P-P) voltage of the signal transmitted through the EBG filter structure.

icant when $r_n = 1$. This parameter r_n indicates the level of interference suffered by the DUT in an EM environment rich in radiation. It is used to evaluate the susceptibility of the DUT when it is subjected to an interference caused by nearby radiation.

Before subjecting the structure to interference, the FM signal with a carrier frequency ranging from 1 GHz to 4 GHz at a step of 0.5 GHz was transmitted through the DUT. The noise P-P voltages without any interference and only due to the structure itself were recorded and the normalized values were plotted in Fig. 4.4.

For comparison purposes, the normalized measured $|S_{21}|$ of the DUT was also

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plotted in Fig. 4.4. Interestingly, as shown in Fig. 4.4, the normalized $|S_{21}|$ has the same trend as that of the normalized $(v'_{max} - v'_{min})$ over the entire passband. When the ripple level is high, the noise peak-to-peak voltage is correspondingly high and vice versa. In order to take into account the effect of the structure for the EMS experiment, this measured noise peak-to-peak voltage, $(v'_{max} - v'_{min})$, is used as reference.

With the experimental setup introduced above, by using the relative noise peak-to-peak voltage as a parameter for the evaluation of EMS of the structure, the effect of the interference source and those of the relative position and the polarization between the DUT and the radiating interference source are studied and presented in the following sections.

4.1.3 **Frequency and Power Level of the Radiating Interference Source**

In order to study the effect of the interference source at various working frequencies and power levels, parameters that determine the relative position (or polarization) between the radiating interference source and the DUT such as θ and d (or ϕ) were fixed. The angle θ is $+90^\circ$ and -90° when the interference source was facing the microstrip line and the ground plane, respectively. The angle ϕ was set to be 0° and d was set in such a way that the DUT is in the near-field or far-field region of the interference source.

4.1.3.1 **Frequency, f_i**

An FM signal with a power level $P_c = -30$ dBm and carrier frequency $f_c = 1.5$ GHz, 2 GHz, and 3.5 GHz were transmitted through the DUT respectively. The frequency of the interference source f_i was varied from 1 GHz to 7 GHz and the power level P_i was fixed at -15 dBm. Other parameters were fixed ($d = R_1$, $\theta = \pm 90^\circ$

(a) Towards the microstrip line plane

(b) Towards the ground plane

Figure 4.5: The relative noise peak-to-peak voltage of the demodulated FM signal ($f_c = 1.5$ GHz, 2 GHz, 3.5 GHz) at the output of the EBG filter when the frequency of the interference source f_i was varied from 1 GHz to 7 GHz.

Figure 4.6: The relative noise peak-to-peak voltage of a demodulated FM signal at the output of the EBG filter when $f_i = f_c$.

(towards the MLIN plane or GND plane), $\phi = 0^\circ$). The noise peak-to-peak voltages of the demodulated FM signal were recorded and the relative noise peak-to-peak voltages were calculated using (4.3) and plotted in Fig. 4.5.

Fig. 4.5 (a) and (b) show r_n of the FM signal with a fixed f_c ($f_c = 1.5$ GHz, 2 GHz, 3.5 GHz) when $\theta = 90^\circ$ and -90° , respectively. In Fig 4.5 (a), it can be seen that, at $\theta = 90^\circ$ where the MLIN plane is exposed to the interference source, the signal in the DUT at 1.5 GHz, 2 GHz, and 3.5 GHz has a maximum relative noise P-P value (r_{n-max}) of 88.87, 65.00, and 38.00 when the working frequency of the interference source is the same as that of the signal transmitted through the DUT. Similarly, when the ground plane is exposed to the interference source, as shown in Fig 4.5 (b), the interference level is the highest when both working frequencies are the same. This is expected since when the frequency of the interferer is the same as the working carrier frequency, it will cause the most interference and thus the interference level will be the highest.

The peak values when $f_i = f_c$ are compared. A frequency sweeping over the passband (from 1 GHz to 4 GHz) at a step of 0.5 GHz was performed. The angle θ was set to be 90° and -90° . The relative noise peak-to-peak voltages of the signal r_n 's were obtained and plotted in Fig. 4.6. As can be seen in Fig. 4.6, r_n varies a lot when the frequency is varied. In both cases, the noise level of the signal with $f_i = f_c$ is frequency dependent. When the microstrip line plane is exposed to the interference source, r_n has the maximum value of 88.87 at $f_i = f_c = 1.5$ GHz and a minimum value of 2.48 at $f_i = f_c = 1$ GHz among all the frequencies. When the ground plane is exposed to the interference source, the curve also has a maximum value of 45 at 1.5 GHz and a minimum value of 3 at 1 GHz.

The results obtained reveal that the susceptibility of the EBG filter structure to the interference is frequency dependent. This particular structure has good immunity at 1 GHz even with an interference source placed in close proximity and working at the same frequency as the frequency of the signal in the structure. However, the susceptibility of the structure is considerably high at 1.5 GHz when it is working at the same frequency as that of the interference source.

In Fig. 4.6, it also shows that r_n is dependent on the relative position of the interference source with respect to the DUT, θ . It is observed that r_n is consistently higher when the microstrip line plane is exposed to the interference source than that when the ground plane is exposed to the interference source. This indicates better immunity of the defected ground plane of the DUT as compared to that of the modulated microstrip line. As a result of the observation, for the ease and accuracy of observation, f_i and f_c are both set to be 1.5 GHz in the following experiments due to the high susceptibility of the DUT at this frequency.

In addition, a comparison is made between Fig. 4.4 and Fig. 4.6. It is observed that although the reference noise peak-to-peak value is the highest at 4 GHz (Fig. 4.4), this does not necessarily suggest a poor susceptibility of the structure at the same

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frequency (Fig. 4.6). In fact, from Fig. 4.6, the susceptibility of the EBG filter is relatively good at 4 GHz as compared to that at 1.5 GHz.

Guideline 1: The working frequency or the harmonics of a nearby radiating circuit component should be different from the working frequency of a EBG microstrip filter structure.

4.1.3.2 Power, P_i

In the experiment, $\theta = 90^\circ$, $\phi = 0^\circ$, and the interference source was placed $d = [R_1 + 0.5(R_1 - R_2)]$ and $d = 1.2R_2$ away from the DUT for the near-field and far-field test, respectively. An FM signal ($f_c = 1.5$ GHz, $P_c = -30$ dBm) was transmitted through the DUT. In order to obtain a strong interference for easy detection and measurement, $f_i = f_c = 1.5$ GHz. The noise peak-to-peak values of the FM signal transmitted through the DUT were recorded at the output of the device when the power level of the interference source P_i was varied from -15 dBm to -45 dBm. The relative noise peak-to-peak values were calculated and plotted in Fig. 4.7.

Fig. 4.7 (a) and (b) show the logarithm and the real value of r_n , respectively. As can be seen in Fig. 4.7 (a), in both the near- and far-field region, the logarithm of r_n decreases linearly as the power level of the interference source decreases. This is expected since the power level of the interference source has a linear relationship with logarithm of the noise level. In Fig. 4.7 (b), a knee is observed when $P_i = -30$ dBm where r_n drops below 15.00 in both the near- and far-field regions. This indicates that the DUT is relatively unsusceptible to the interference when the power of the interference source is equal to or lower than that of the signal transmitted through the DUT.

Guideline 2: If a radiating interference source is working at the same frequency as that of the DUT, its power level should be reduced to be equal to or less than that of the DUT.

(a) Log.-Log. scale

(b) Linear-Log. scale

Figure 4.7: The relative noise P-P voltage of a demodulated FM signal at the output of the EBG filter when the power of the interference source P_i was varied from -15 dBm to -45 dBm

4.1.4 Relative Position and Polarization of the Radiating Interference Source

From the results obtained above, the effects of either the frequency or the power level of a radiating interference source on the signal transmission through the EBG structure under study are significant. However, in some applications, the working frequency and the power level of a nearby radiating circuit component are not allowed to be changed to eliminate the interference to the EBG structure. In such circumstances, the knowledge on the effect of the relative position and polarization of the radiating interference source with respect to the DUT are useful for minimizing interference. In this part of the study, the angle of incidence of the radiating interference source to the DUT θ and the polarization ϕ are examined. The frequencies of the signals in the DUT and in the interference source, f_c and f_i , were set to be 1.5 GHz to obtain a high interference level. The power of the transmitted signal through the DUT was set to be -30 dBm and the power level of the interference source was set to be -15 dBm.

4.1.4.1 Angle of Incidence, θ

As has been shown in Fig. 4.6, the FM signal transmitted through the DUT has higher r_n when the microstrip line plane is exposed to the interference source ($\theta = +90^\circ$) as compared to that when the ground plane is exposed to the interference source ($\theta = -90^\circ$). The difference between the interference levels when $\theta = +90^\circ$ and when $\theta = -90^\circ$ is dependent on the working frequency when $f_c = f_i$ as can be seen in Fig. 4.6. In this section, the effect of θ on the interference levels is examined. With $f_i = f_c = 1.5$ GHz, the interference source was placed 50 cm and 100 cm away from the DUT for the near-field and far-field test, respectively. The angle ϕ was set to be 0° and θ was varied from -90° to 90° .

The noise levels of the signals under test were measured and plotted in Fig. 4.8.

Figure 4.8: The relative noise peak-to-peak voltage of a demodulated FM signal at the output of the EBG filter when the angle of incidence of the radiating interference source to the DUT θ is varied from -90° to $+90^\circ$.

As shown in Fig. 4.8, in both the near-field and the far-field region, r_n increases as θ increases from -90° to $+90^\circ$. The interference level that the signal in the DUT suffers is higher when $\theta > 0^\circ$ than that when $\theta < 0^\circ$. This implies that the MLIN of the DUT is more susceptible to the interference than the GND. The reason is that when the MLIN plane is exposed to the interference source, the E-field of the antenna directly interferes the signal transmitted through the DUT while when the ground plane was exposed to the interference source, the interference can only get through to the signal line through the etched circles in the ground plane. As can be seen in Fig. 4.8, the DUT has a r_n larger than 20 when the radiating interference source is directed to the ground plane. It is additional interference suffered by the DUT as compared to the one with a perfect ground plane. It is introduced by the etched circular patches in the ground plane.

Guideline 3: In a multi-layer circuit, a radiating component is suggested to be placed under the ground plane of an EBG structure in order to eliminate interfer-

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ence.

4.1.4.2 Polarization, ϕ

The study on the effect of the polarization of the interference source ϕ was conducted in both the near-field ($d = 50$ cm) and the far-field region ($d = 100$ cm). The angle ϕ was varied from 0° to 90° and θ was set to be 90° and -90° .

Fig. 4.9 shows the calculated relative noise peak-to-peak values of the measured demodulated FM signals. As shown in Fig. 4.9, r_n of the signal decreases dramatically as ϕ increases from 0° to 90° in both near-field and far-field region. The relative noise peak-to-peak value decreases at an increasing speed when $\phi > 30^\circ$. The signal has an r_n of 5.00 and 4.03 at $\phi = 90^\circ$ in the near-field and far-field region, respectively, which are close to 1. It indicates that the interference is negligible at $\phi = 90^\circ$ where the polarization of the E-field of the interference source is perpendicular to the direction of the signal transmission and a minimum cross polarization interference is obtained.

Guideline 4: A relative cross polarization is suggested for a low interference level introduced by a radiating interference source to an EBG structure.

4.2 EMS to a Nearby Microstrip Line

In this section, the EM susceptibility of the dual planar EBG filter structure to a microstrip line in close proximity is studied. The studies examine the coupling effects of a nearby microstrip line on the performance of the EBG structure when the microstrip line is and is not transmitting a signal. When the microstrip line is transmitting a signal, the effect of the working frequency and that of the direction of the signal on the coupling to the EBG structure are studied. Experimental and numerical work are presented, analyzed, and verified. In this study, knowledge is

Figure 4.9: The relative noise peak-to-peak voltage of a demodulated FM signal at the output of the EBG filter when the angle of polarization, ϕ , is varied from 0° to 90° .

obtained to eliminate interference to an EBG structure from a nearby microstrip line. The results obtained are useful to the applications and integrations of EBG microstrip structures to microwave circuits. They are useful for EMC and EMS studies of other patterned ground structures/defected ground structures.

4.2.1 Device Under Test & Interference Source

The dual planar EBG structure in Fig. 4.1 (a) is the DUT for this study. The substrate material and the dimensions of the structure are the same as described in Section 4.1.1. The EMS study involves the coupling and crosstalk effects of a microstrip line which is placed close to the DUT and on the same plane as the structure. Fig. 4.10 (a) and (b) show the top view and the cross sectional view of the DUT with a nearby microstrip line, respectively. Fig. 4.10 (a) also includes the block diagram of the experimental setup for the test in case two. As shown in Fig. 4.10 (a), a microstrip line with a width of w' is placed x away from the

Figure 4.10: The structure under electromagnetic susceptibility test

DUT in the transversal direction on the plane where the microstrip line of the dual planar EBG structure is located. Both microstrip lines are bent at an angle of 150° at the input and output ports to ease the test when x is small. The bends are chamfered to compensate for additional loss introduced by the bend. The four ports are numbered in the sequence shown in Fig. 4.10 (a). The test in case one will have Port 3 and Port 4 of the microstrip line either terminated with matched loads or left as open circuits. As shown in the cross sectional view in Fig. 4.10 (b), the interference source shares the same substrate and ground plane as those of the

DUT.

4.2.2 Case One: an Inactive Microstrip Line

When the DUT is transmitting signals while there is no signal transmitting through the microstrip line, coupling takes place. The waves that are propagating in the DUT are quasi-TEM waves [85]. They induce electromagnetic waves in the nearby microstrip line via the shared substrate. The waves induced in the microstrip line couple with the waves in the DUT, thus the microstrip line interferes with the waves propagating in the DUT.

In the following subsections, coupling is studied when the microstrip line has open-circuit or matched-load terminations. The effect of the etched circles in the ground plane on the susceptibility of the DUT is examined and the physics behind are explored. The DUT was simulated using Agilent's Advanced Design System [30]. The distance x was set to 4.5 mm, 6.5 mm, 10.38 mm, 15 mm, and 20.76 mm where 10.38 mm is exactly one period of the DUT.

4.2.2.1 Microstrip Line with Open-circuit Terminations

Fig. 4.11 shows the simulated S_{21} -parameters of the DUT with an open-ended microstrip line x away from it. This figure includes the simulated data of the DUT itself for comparison purposes. As shown in Fig. 4.11, each DUT shows a stopband over the same frequency range. Thus, the stopband is not affected by the nearby microstrip line. When $x < a$, large ripples are observed in the stopband. The ripple level in the stopband is as high as -9.64 dB when $x = 4.5$ mm. The filtering performance of the DUT is degraded by the ripples. As x increases, the ripple level decreases significantly. It is found to be -40 dB when x is one period of the DUT. When $x > a$, it decreases further. For the performance in the passband, ripples are observed only when the microstrip line is 4.5 mm away from the DUT. The ripples

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Figure 4.11: The simulated S_{21} -parameter of the dual planar EBG (DP-EBG) filter structure with an open-ended microstrip line x away (The distances, 4.50 mm, 6.50 mm, 10.38 mm, 15.00 mm, 20.76 mm, are $0.22 \lambda_g$, $0.31 \lambda_g$, $0.50 \lambda_g$, $0.72 \lambda_g$, $1.00 \lambda_g$, respectively).

are caused by the induced electromagnetic waves in the microstrip line as analyzed previously. When the microstrip line has open-circuit terminations, induced waves are reflected back and forth along the microstrip line, behaving like a resonator. Thus, the coupling effect is enhanced. The strength of coupling weakens as the distance between the two structures increases. This is due to the high confinement of fields around the microstrip line of the DUT.

With open-circuit terminations, the S-parameters of the DUT is affected when x is small. In this case, high ripples are observed in the stopband when x is less than one period of the structure. The ripple level becomes acceptable when $x \geq a$. The distance a is the safe distance where the interference level becomes acceptable. The safe distance depends on the field distribution of the DUT. It varies with the geometry of the DUT and the parameters of the substrate, such as the dielectric constant.

The effect of the etched circles on the susceptibility of the DUT is examined. The DUT without etched circles in the ground plane is simulated [30] when the microstrip line is x away ($x = 4.50$ mm, 6.50 mm, 10.38 mm, 15.00 mm, and 20.76 mm). Fig. 4.12 shows the simulated data of the single planar DUT with or without a microstrip line nearby. As shown in Fig. 4.12, the single planar structures have stopbands over the same frequency range. Thus, the stopband is not affected by the microstrip line, which is similar to the dual planar DUT. High ripples in the stopband are observed when $x = 4.50$ mm. When $x \geq 6.50$ mm, the ripple levels decrease significantly and then eventually vanish. As compared to the dual planar DUT, the structure without an imperfect ground plane has a lower ripple level at a smaller x . This indicates that the etched circles cause additional susceptibility for the structure. From a wave propagation point of view, compared with the single planar DUT, the E-fields of the dual planar DUT are disturbed by the etched patches and become less confined below the microstrip line. This makes the structure more vulnerable to the interference caused by the induced electromagnetic waves in the nearby microstrip line.

4.2.2.2 Microstrip Line with Matched-Load Terminations

Fig. 4.13 shows the simulated transmission coefficients of the DUT when the microstrip line nearby is terminated with matched loads. In Fig. 4.13, it is seen that each DUT shows a stopband over the same frequency band, which is the same as that when the microstrip line is open-ended. At $x = 4.5$ mm, the attenuation in the stopband is 10 dB less than that of the DUT without a microstrip line. No ripples are observed in the stopband. When $x \geq 6.5$ mm, the frequency responses of the structures approach that of the DUT without a microstrip line. Comparing the frequency responses of the DUT when the microstrip line is open-ended with those when the microstrip line is terminated with matched loads, it is noticed that

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Figure 4.12: The simulated S_{21} -parameter of the single planar EBG (SP-EBG) filter structure (no etched patches in the ground plane) with an open-ended microstrip line x away.

Figure 4.13: The simulated S_{21} -parameter of the dual planar EBG (DP-EBG) filter structure with a microstrip line with matched load terminations x away.

Table 4.1: The Measured r_n 's of the Demodulated FM Signals

Carrier Frequency	1 GHz	2 GHz	3 GHz
r_n	40	75	96

the effect of a nearby microstrip line is much less significant when the microstrip line is terminated with matched loads. This is because, for the microstrip line with matched loads, the induced electromagnetic waves in the microstrip line are absorbed by loads at either port and the coupling is reduced.

The coupling of an inactive microstrip line on the frequency response of the DUT is much stronger when the microstrip line has open-circuit terminations than when it is terminated with matched loads. Therefore, if there is no signal being transmitted through the microstrip line, terminations with matched loads are preferred for eliminating interference. In the situation where matched-load terminations are not obtainable for a microstrip line near the DUT, a microstrip line should be placed far away from the DUT. It should be placed beyond the safe distance for the DUT where the interference experienced by the DUT is acceptable. This distance is determined by the geometry of the DUT and the parameters of the substrate. Moreover, the etched circles in the ground plane of the DUT show effect on the susceptibility of the DUT to an inactive nearby microstrip line. The single planar DUT is found to be less susceptible to the microstrip line at the same x than the dual planar one. A microstrip line can be placed closer to a single planar EBG structure due to the higher confinement of its fields compared with its dual planar counterpart.

4.2.3 Case Two: Microstrip Line with a Signal

When the DUT and the microstrip line are both transmitting signals, crosstalk is introduced by the microstrip line through the shared substrate to the DUT. The

Figure 4.14: The simulated S_{24} versus x at different carrier frequencies.

signal integrity is affected by the amount of crosstalk introduced. The experimental setup is illustrated by the block diagram in Fig. 4.10 (a). An FM signal with a carrier frequency, f_c , was transmitted through the DUT while a continuous wave (CW) at f_{MLIN} was transmitted through the microstrip line. The FM signal was demodulated at the output of the DUT and the noise peak-to-peak voltages were measured. The relative noise peak-to-peak voltages were calculated based on (4.3) introduced in Section 4.1 and used to evaluate the integrity of the received signal transmitted through the DUT. Based on the study in case one, the microstrip line was placed sufficiently far away from the DUT ($x > a$) to avoid the coupling effect. The DUT was tested when a microstrip line was 10.38 mm, 15 mm, 20.76 mm, or 25 mm away from it.

4.2.3.1 Frequency of the Interference Signal, f_{MLIN}

The effect of the frequency at which the interference source is working is examined. FM signals with a power level of -30 dBm and carrier frequencies of 1 GHz, 2 GHz,

and 3 GHz were transmitted one at a time through the DUT from Port 1 to Port 2. It was exposed to the interference source (a microstrip line) that was one period of the DUT away from it ($x = 10.38\text{mm}$). The interference signal was pumped in from Port 4 and terminated with a matched load at Port 3. Its frequency was swept from 1 GHz to 6 GHz with a fixed power level of -20 dBm . The noise peak-to-peak voltages of the demodulated FM signals were recorded and r_n 's in (4.3) were calculated. Based on the measured results, it is found that r_n of the signal transmitted through the DUT is the highest when the frequency of the interference source is the same as that of the signal through the DUT. In other words, peak values are found when the microstrip line and the DUT are operating at the same frequency. This holds for all carrier frequencies under study. Table 4.1 shows the peak values at each carrier frequency. In Table 4.1, it is noticed that the peak interference level increases considerably as f_c increases.

Theoretical studies on coupled lines in [122] can be used to explain the observations above. In [122], the coupling effect of two coupled microstrip lines is studied numerically using coupled-mode theory [123]. Based on the studies, the coupling level of a pair of coupled microstrip lines goes higher as the working frequency increases. The observation in Table 4.1 agrees well with the conclusions in [122].

In [122], it is reported that the coupling effect of a multi-port network predicted by its S-parameters has a good agreement with that predicted by numerical results using coupled-mode theory. The four-port network in Fig. 4.10 (a) was simulated to analyze the coupling effect in this group of experiments where the two signals are transmitted from the left-hand-side to the right-hand-side of the system as shown in Fig. 4.10 (a). Fig. 4.14 shows the simulated S_{24} at working frequencies of 1 GHz, 2 GHz, and 3 GHz versus the distance between the DUT and the microstrip line. As can be seen in Fig. 4.14, at $x = 10.38\text{ mm}$ for example, the coupling becomes stronger as the frequency increases (from -39 dB at 1 GHz to -31 dB at 3 GHz). It

Figure 4.15: The real value, logarithm and its linear fitting line of the measured relative noise peak-to-peak voltage of the demodulated FM signal ($P_c = -30$ dBm, $f_c = 2$ GHz, $x = a$) when the power level of the interference source ($f_{MLIN} = 2$ GHz) is varied.

holds as the distance between the two structures varies. The simulation results are in good agreement with the observation from the measurement results in Table 4.1. They also agree with the numerical results in [122].

Based on the results, being similar to two identical coupled microstrip lines, the microstrip line introduces the highest interference level to the DUT when they are working at the same frequency. Therefore, when a dual planar EBG is incorporated into a microwave system, a nearby microstrip line should work at a different frequency and/or different harmonic frequency from that of the EBG structure. Low frequencies are preferred since the interference becomes severe as the working frequency increases.

4.2.3.2 Directions of the Interference Signal

In this section, the effect of the direction of the interference signal is studied. An FM signal ($P_c = -30$ dBm; $f_c = 2$ GHz) was transmitted through the DUT (from Port 1 to Port 2) and a CW was transmitted through the microstrip line with a

power level that was varied from -15 dBm to -50 dBm. In order to make it easy to observe the interference level, f_{MLIN} was set to be equal to f_c to maximize the interference level according to the previous study. At each power level, the signal was pumped into the microstrip line in two different directions: from Port 4 to Port 3 which is in the same direction as that of the signal through the DUT; and from Port 3 to Port 4 that is in the reversed direction to the signal through the DUT. The noise peak-to-peak voltages were recorded at each power level of the interference in each case and r_n 's were calculated and plotted.

Fig. 4.15 shows the real values and its logarithms with linear fitting lines of the relative noise peak-to-peak voltages of the DUT versus the power level when the two signals are in the same or in opposite directions ($x = a$, $f_c = 2$ GHz). As shown in Fig. 4.15, regardless of the direction of the signal through the microstrip line, the logarithm of r_n decreases linearly with the power level of the interference source, which is as expected. In Fig. 4.15, it is observed that the direction of the interference signal affects the interference level experienced by the DUT. As shown in Fig. 4.15, at 2 GHz, the interference level of the DUT is higher when the two signals are in the same direction than that when they are in opposite directions. The difference of coupling due to the directions is counterintuitive for a frequency lower than 5 GHz. It is different from a coupled system with two identical microstrip line with matched loads.

In [122], it is reported that, in a coupled system with two identical microstrip lines terminated with matched loads, if the sequence of ports is the same as that in Fig. 4.10 (a), the coupling from Port 3 to Port 2 is higher than that from Port 4 to Port 2 at frequencies below 5 GHz. In the same paper, it is also reported that the difference in coupling levels due to the directions of signals changes as the reflections in the coupled system change. An example is given showing that, by introducing reflections using unmatched loads at all the ports, the difference be-

Figure 4.16: The simulated coupling coefficients versus x at 2 GHz.

tween the coupling from Port 3 to Port 2 and that from Port 4 to Port 2 varies with frequencies. At different frequencies, the coupling level from Port 3 to Port 2 can be higher, the same, or lower than that from Port 4 to Port 2.

In this system under study where the DUT and the microstrip line are coupled, additional reflections are generated by the inserted patches in the microstrip line and the etched circles in the ground plane. They alter the coupling from the microstrip line to the DUT considerably. Fig. 4.16 shows the simulated coupling coefficients, S_{23} and S_{24} , at a working frequency of 2 GHz versus the distance between the two structures. The two parameters, S_{23} and S_{24} , correspond to the input of the interference source being pumped in from Port 3 and Port 4, respectively. As shown in Fig. 4.16, both S_{23} and S_{24} decrease as x increases. It is observed that S_{23} is always smaller than S_{24} for the same distance. This differs from the coupling between two identical microstrip line which has $S_{23} > S_{24}$ at 2 GHz [122]. The difference is due to reflections occurring in the DUT with periodic unit cells in both the microstrip line and the ground plane.

Based on the results obtained above, the interference level that the DUT expe-

riences depends on the direction of signal in the microstrip line with respect to that of the signal in the DUT. Reflections in a coupled system alter the coupling between the coupled structures from that of a standard system with two identical coupled lines. The difference in coupling due to the directions of signals in the structures depends on the geometries of each structure and the frequency. Calculations of S-parameters predict the coupling levels between the signals and the interferences in different relative directions. Based on the calculations, appropriate directions of the signals in the coupled structures can be obtained in which the interference level experienced by the DUT can be lowered.

4.3 The Radiation of Patterned Ground Structures

In this section, the radiation of patterned ground structures in the far-field region is studied theoretically and experimentally. The effects of the etched patterns on the radiation are examined. The structures include a patterned ground structure with a slot (an elongated rectangular aperture) in the ground plane and its variants. The studies are conducted by analyzing the radiation based on the fields in the aperture, the simulation results, and the measured results. It serves as a starting point for the further study of the radiation of patterned ground structures where the realistic near-field radiation shall be looked into.

4.3.1 Patterned Ground Structure with a Slot

The study starts with a patterned ground structure of the simplest form where there is a slot in the ground plane. Fig. 4.17 (a) and (b) show the 3-D and bottom view of this patterned ground structure, respectively. In this structure, the slot is perpendicular to the microstrip line in the ground plane. It has a stopband which is

(a) 3-D view

(b) Bottom view

Figure 4.17: A patterned ground structure with a slot (an elongated rectangular aperture) in the ground plane.

Figure 4.18: The sketch of fields in the patterned ground structure in Fig. 4.17.

determined by the dimension of the slot. As shown in Fig. 4.17, the length and the width of the slot are denoted as a and b , respectively. The length and the width of the microstrip line are ℓ and w , respectively. The thickness of the substrate is denoted using h . The azimuth angle is ϕ and θ is the inclination angle from the z -axis. The angles $0^\circ < \theta < 90^\circ$ correspond to the region on the side of the microstrip line whereas $90^\circ < \theta < 180^\circ$ correspond to that on the side of the ground plane. The angles $\theta = 0^\circ$ and 180° correspond to the locations directly above the microstrip line and directly below the ground plane, respectively. The plane $\phi = 0^\circ$ is perpendicular to the microstrip line cutting along the x -axis.

The patterned ground structure shown in Fig. 4.17 has been studied by applying the reciprocity theorem and a full moment method solution to obtain an equivalent circuit, back in 1986 [1]. In terms of radiation, it is reported in [1] that the dielectric

(a) E-field

(b) H-field

Figure 4.19: The snapshots of the simulated fields of the patterned ground structure with a slot in Fig. 4.17 at 3 GHz at the same moment.

substrate has no significant effect on the radiation patterns of the slot. This implies that there is radiation on both sides of the structure. Our study is focused on the radiation in the region on the side of the ground plane in the far field, in order to study the effects of the etched patterns in the ground plane on the radiation generated by the structure. The radiation at a distance of r from the origin as shown in Fig. 4.17 (a) is investigated, $r = \sqrt{x^2 + y^2 + z^2}$.

Firstly, the fields in the slot in the ground plane were analyzed. Fig. 4.18 (a) and (b) show the side view and the bottom view of the sketches of the fields in the

Figure 4.20: Equivalent models for a slot in the ground plane as shown in Fig. 4.17.

patterned ground structure, respectively. Before the slot, the quasi-transverse electromagnetic (TEM) waves are propagating along the microstrip line [124]. As can be seen in Fig. 4.18 (a), on the yz -plane ($x=0$), the E-fields (upwards and downwards), the H-fields (into and out of the paper), and the wave propagation are orthogonal to one another. $\vec{J}_s$ is the surface current of the microstrip line or the ground plane. In the slot, the quasi-TEM mode is disturbed and the aperture fields are generated [1]. Fig. 4.18 (b) shows a sketch of the H-fields and the distribution of the y -components of the E-fields in the slot. As can be seen in Fig. 4.18 (b), the H-fields are in the $\hat{x}$ - and the $\pm \hat{z}$ -direction forming a loop. The y -components of the E-fields are not evenly distributed in the slot. There is a higher confinement of fields in the area directly below the microstrip line.

The patterned ground structure was simulated using CST Microwave Studio [2]. Taconic with a permittivity (ϵ_r) of 2.2 and a thickness (h) of 0.65 mm was used as the substrate. The length of the microstrip line ℓ was set to 60 mm. The width w equals to 2 mm corresponding to a characteristic impedance of around 50Ω at

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frequencies below 10 GHz. The dimensions of the slot, a and b , were set to be 18 mm and 1 mm, respectively. Both ports of the structure have a characteristic impedance of $50\ \Omega$. In the simulated results, the structure shows a resonance at 7.5 GHz and a 3 dB-cutoff at 5.24 GHz. Respectively, Fig. 4.19 (a) and (b) show the snapshots of the distributions of the simulated E- and H-fields in the slot of the patterned ground structure at 3 GHz at the same moment. The frequency 3 GHz is a frequency in the passband. As shown in Fig. 4.19 (a), the E-fields in the slot mainly point in the positive $\hat{y}$ -direction. Both the color and the size of the arrow indicate the strength of the E-field. The size of the arrow is proportional to the logarithm of the strength of the E-field. As shown in Fig. 4.19 (a), the E-fields are high at the center of the slot below the microstrip line, and low at the two ends. This agrees with the sketches of the E-fields shown in Fig. 4.18 (b). For the H-fields shown in Fig. 4.19 (b), they are pointing from right to left, which agrees with the sketches of the H-fields in Fig. 4.18 (b).

The radiation of the slot in the far-field region (at a distance of r) is analyzed as follows based on the E-fields in the slot. Fig. 4.20 (a) shows the yz -plane of the ground plane (GND) of the structure with a slot. The region at the right hand side of the ground plane is the focus of this study. The majority of E-fields are along the $\hat{y}$ -direction. In Fig. 4.20 (a), the E-fields in the slot in the $\hat{y}$ -direction are shown. As analyzed, the E-field is a function of x . Thus, $\bar{E}(x') = E(x')\hat{y}$. Prime is used to indicate that it is the source of radiation. The normal of the plane is $\hat{n}$, $\hat{n} = -\hat{z}$. For the analysis on radiation, the field equivalence principle is applied. A flat plane S , from negative infinite y to positive infinite y on the ground plane, is chosen and shown in Fig. 4.20 (b). Over the infinite plane, the equivalent sources $\bar{J}_s$ and $\bar{M}_s$ are formed. The electric current densities are unknown. The magnetic current density $\bar{M}_s$ is non-zero only in the slot since the E-fields only exist in the slot. In Fig. 4.20 (c), an imaginary flat electric conductor is assumed to approach

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the surface S , and it shorts out all the electric current density $\bar{J}_s$ everywhere. By applying image theorem, the problem in Fig. 4.20 (c) is reduced to the equivalent problem as shown in Fig. 4.20 (d) where the radiation at the right hand side of S can be estimated by the radiation generated by the magnetic current density $\bar{M}_s = -2\hat{n} \times \bar{E}(x')$. Therefore, the radiation at the right hand side of the ground plane in Fig. 4.20 (a) can be estimated using this magnetic current.

As $\bar{E}(x') = E(x')\hat{y}$ and $\hat{n} = -\hat{z}$,

$$\bar{M}_s = -2\hat{n} \times \bar{E} = -2(-\hat{z}) \times E(x')\hat{y} = -2E(x')\hat{x} \quad (4.4)$$

The magnetic current density $\bar{M}_s$ is a function of x' and in the $\hat{x}$ -direction. With the magnetic current, the electric vector $\bar{F}(x, y, z)$ can be estimated using (4.5) below,

$$\bar{F}(x, y, z) = \frac{\epsilon}{4\pi} \int_{y'} \int_{x'} -2E(x')\hat{x} \frac{e^{-ikr}}{r} dx' dy' \quad (4.5)$$

where r is the distance from the source point to the observation point. $\bar{F}(x, y, z)$ has only x -components, $\bar{F}(x, y, z) = F_x\hat{x}$.

Therefore, the E-fields $\bar{E}_F$ due to the electric vector $\bar{F}$ in the far-field region can be estimated using (4.6) below,

$$\bar{E}_F = -\hat{r}\eta \times \bar{H}_F \quad (4.6)$$

where

$$\bar{H}_F = -j\omega\bar{F} = -j\omega F_x\hat{x} \quad (4.7)$$

$$\hat{r} = \hat{x} \cos \phi \sin \theta + \hat{y} \sin \phi \sin \theta + \hat{z} \cos \theta \quad (4.8)$$

and η is the wave impedance of the air. When (4.7) and (4.8) are substituted into

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(4.6), $\bar{E}_F$ can be expressed as

$$\begin{aligned}\bar{E}_F &= -\eta \begin{vmatrix} \hat{x} & \hat{y} & \hat{z} \\ \cos \phi \sin \theta & \sin \phi \sin \theta & \cos \theta \\ -j\omega F_x & 0 & 0 \end{vmatrix} \\ &= j\omega\eta F_x (\hat{y} \cos \theta - \hat{z} \sin \phi \sin \theta)\end{aligned}\quad (4.9)$$

As shown in (4.9), at $\phi=0$, the E-fields in the far-field region, $\bar{E}_F$, have only y -components. The analysis reveals that for the slot in the ground plane, the radiation it generates has only y -components when the E-fields in the slot are in the $\hat{y}$ -direction. Similarly, it can be proved that, when the E-fields in the slot are in the $\hat{x}$ -direction, the radiation it generates has only x -components. The radiated E-fields by a slot have the same direction as that of the E-fields in the slot. For the patterned ground structure shown in Fig. 4.17, its radiation is mainly in the $\hat{y}$ -direction since the E-fields in the slot mainly point in the $\hat{y}$ -direction. Its radiation is polarization sensitive.

Fig. 4.21 (a) and (b) show the magnitudes of the fields of the structure (simulated using CST Microwave Studio) versus θ in the far-field region at 3 GHz and 7.5 GHz, respectively. The azimuth angle, $\phi=0^\circ$. In each figure in Fig. 4.21, the total E-fields and their x - and y -components, and the total H-fields are included. The frequency, 3 GHz, is in the passband and 7.5 GHz is the resonant frequency of the patterned ground structure.

In the passband, as shown in Fig. 4.21 (a), the y -components of the E-fields and the total E-fields overlap each other. They are 25 dB or more higher than the x -components of the E-fields. It indicates that, for E-fields, the y -component dominates and therefore, the radiation of the patterned ground structure is highly polarization sensitive. Moreover, it is noticed that the total E-fields on the side of the microstrip line ($0^\circ < \theta < 90^\circ$) and those on the side of the ground plane

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(a) 3 GHz

(b) 7.5 GHz

Figure 4.21: Simulated E-fields and H-fields in the far-field region of the patterned ground structure in Fig. 4.17 at different frequencies.

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($90^\circ < \theta < 180^\circ$) are comparable, which is the same as reported in [1]. The E-fields at the back of the ground plane are slightly higher than those above the microstrip line [see the dashed line]. In the same figure, the magnitudes of the total H-fields of the structure in the far-field region are below -50 dB.

As shown in Fig. 4.21 (b), at the resonant frequency, the magnitudes of the total E-fields and their y -components are 10 dB or more above those of the corresponding x -components. Due to the reduced difference between the y - and the x -components at this frequency, the total E-field is slightly different from the corresponding y -component. For the total H-fields, it is observed that their magnitudes are below -40 dB. The results in Fig. 4.21 (b) reveal that the radiation of the structure at the resonant frequency is also polarization sensitive.

In Fig. 4.21, the results show that the radiation of the structure depends on polarizations at both frequencies. As frequency changes, the dependence on polarizations differs. By comparing the magnitudes of the fields in Fig. 4.21 (a) to those in Fig. 4.21 (b), the radiation at the resonance is much higher than that at frequencies in the passband, which is expected in a structure that resonates. Therefore, operating at the resonant frequency of a patterned ground structure should be avoided if possible.

4.3.2 Variants

Fig. 4.22 (a) and (b) show the variants of the patterned ground structure with a slot in the ground plane in Fig. 4.17. In Fig. 4.22 (a), the slot is bent 90° in opposite directions forming a Z-shape. In Fig. 4.22 (b), the slot is bent in the same direction forming a U-shape. The U-shaped structure was studied in terms of frequency response in [7]. The length of the arm is denoted by c and the distance between the center line of the microstrip line and that of the arm is d . The radiations of these two variants are analyzed by applying the same analysis presented previously

(a) Z-shaped

(b) U-shaped

Figure 4.22: Bottom views of the patterned ground structures.

for the patterned ground structure in Fig. 4.17.

For the patterned ground structure with a slot in the ground plane in Fig. 4.17, the slot exhibits waveguide-like behavior, being similar to a waveguiding mechanism. In order to study the fields and the radiation of the Z-shaped and the U-shaped patterned ground structures, they were simulated [2]. The total lengths of the bent slots are 18 mm which is the same as that of the straight slot in Fig. 4.17. The lengths were chosen for a fair comparison of the three structures in terms of frequency responses and far-field radiation. With the length of the

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Figure 4.23: The snapshots of the simulated E-fields of the patterned ground structures at 3 GHz.

slots set, $2c + w = a = 18$ mm, $c = 8$ mm, and $d = 1.5$ mm. In the simulated results, the Z-shaped and the U-shaped structures show resonant frequencies of 7.45 GHz and 7.47 GHz, respectively, which indicates that the resonant frequency of the patterned ground structure is mainly determined by the total length of the slot. Bending of the slot only has a minor effect on shifting the resonant frequency of the structure.

Fig. 4.23 (a) and (b) show the snapshots of the simulated E-fields in the Z-shaped and the U-shaped slots at 3 GHz, respectively. As can be seen in Fig. 4.23, when the slot is bent, the directions of the E-fields are changed with the bending of the slot, maintaining being perpendicular to the longer edges of the slot. For the Z-shaped structure, as shown in Fig. 4.23 (a), the E-fields in the two arms are mainly along the $\hat{x}$ -direction, pointing in the same direction, while those of the E-fields directly below the microstrip line in the slot are pointing in the $\hat{y}$ -direction. According to the analysis of radiation based on the E-fields in the slot previously, the E-fields in the Z-shaped slot may lead to high radiation along the $\hat{x}$ -direction.

Figure 4.24: The simulated E-fields of the microstrip structures in the far-field region at 3 GHz

Figure 4.25: The measured S_{21} -parameters of the testing system in Fig. 4.27 at 3 GHz

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Depending on the locations defined by θ , enhancement of radiation to different extents in this direction in the far-field region can be expected. For the U-shaped structure, as shown in Fig. 4.23 (b), the E-fields in the two arms are pointing in opposite directions along the $\hat{x}$ -direction. This results in a possible cancelation of radiation in the far-field region in the $\hat{x}$ -direction. Thus, an elimination of radiation for the U-shaped structure can be expected in this direction.

Fig 4.24 shows the simulated E-fields in the far-field region for the three patterned ground structures at 3 GHz. The angle θ is swept from 90° to 180° for the region on the side of the ground plane. Both the x - and the y -components of the E-fields are included. The E-field components of a microstrip line with the same length and width to those of the patterned ground structures are included for reference.

As shown in Fig 4.24, at 3 GHz in the passband, all structures except the Z-shaped structure show higher magnitudes of the y -components than those of the x -components everywhere in the far-field region. The microstrip line shows the lowest E-fields for both components since it has a perfect ground plane. The structure with a slot shows much higher magnitudes of the y -components than those of the x -components because its E-fields in the slot are mainly in the $\hat{y}$ -direction as explained. Its x -components are close to those of the microstrip line due to the lack of E-fields in the $\hat{x}$ -direction in the slot. The U-shaped structure shows a minimum of 10 dB difference between the magnitudes of the x -components and those of the y -components. The magnitudes of the x -components are low because the E-fields at the two arms are in opposite directions along the $\hat{x}$ -direction and they cancel each other to different extents depending on θ . It is observed that the magnitudes of the x -components of the U-shaped structure are close to those of a microstrip line. Nulls are observed near $\theta = 90^\circ$ and 180° . The Z-shaped structure shows an increasing magnitude of the x -component of the E-field as θ increases. It has the highest magnitudes of the x -components in all the structures at each θ . It

Figure 4.26: The measured S_{21} -parameters of the fabricated patterned ground structures and the microstrip line.

shows a higher magnitude of the x -component than that of the y -component when θ goes beyond 115° and it has a maximum at $\theta = 180^\circ$. For the y -components, the Z-shaped and the U-shaped structures show similar magnitudes because they have a similar portion of the slot directly below the microstrip line where the E-fields are in the $\hat{y}$ -direction.

Comparing the three patterned ground structures, the structure with a rectangular and a Z-shaped slot show high radiation in the $\hat{y}$ - and the $\hat{x}$ -direction, respectively. Thus, the radiation and the interference they generate can be eliminated by depolarizing the structures and the circuit components nearby. The U-shaped structure shows relatively low radiation in both directions. Therefore, for the U-shaped structure, a low interference level to other circuit components and low susceptibility to interference from the working environment can be expected.

4.3.3 Experiments

The three patterned ground structures and the microstrip line were fabricated. Fig. 4.26 shows the measured S_{21} -parameters of the structures. As can be seen in Fig. 4.26, all the patterned ground structures show resonances at around 7.5 GHz, which agrees with the simulated results. At the resonant frequency of the patterned ground structures, relatively high attenuation is observed. With this fact, the patterned ground structures can be used to filter unwanted signals at the resonant frequency. Moreover, it is observed that 3 GHz is a frequency within the passbands. Experiments were designed to measure the relative radiation of the patterned ground structures. Fig. 4.27 shows the setup of the experiment. The system consists of the fabricated patterned ground structures as the devices under test (DUT), a vector network analyzer (VNA), a wire dipole antenna, and a test platform to position the components. Absorbers were used to eliminate reflections from the surrounding. Port 1 of the VNA was connected to the DUT while the DUT was terminated with a $50\ \Omega$ -load. The DUT was positioned with the microstrip line perpendicular to the test platform (xz -plane). The dipole antenna was connected to Port 2 and used to collect the signal from the DUT. S_{21} -parameters of the system were recorded and used to evaluate the relative radiation from the DUT.

In order to test the y - and x -components of the radiation, the S_{21} -parameters in the $\hat{\phi}$ -, the $\hat{\theta}$ -, and the $\hat{r}$ -directions were measured and converted to those in the $\hat{y}$ - and the $\hat{x}$ -directions. For measuring the S_{21} -parameters in the $\hat{\phi}$ -, $\hat{\theta}$ -, and $\hat{r}$ -directions, the dipole antenna was placed in the $\hat{y}$ - (vertically polarized), $\hat{x}$ - (horizontally polarized), and $\hat{z}$ -direction, respectively, and the DUT was rotated in $\hat{\theta}$ -direction. The y -axis is the axis of rotation. The angle θ defines the relative location of the antenna with respect to the DUT at a fixed distance, R . The center of the dipole is aligned to the center of the DUT to obtain data on the $\phi = 0^\circ$ plane. In Fig. 4.27, θ equals 180° . Starting at $\theta = 90^\circ$, when the DUT is

Figure 4.27: The setup of the experiment.

rotating 90° clockwise, θ is varied from 90° to 180° and the far-field region on the side of the ground plane is scanned.

Fig. 4.25 shows the measured S_{21} -parameters of the testing system versus θ at 3 GHz. The S_{21} -parameters in both the $\hat{x}$ - and the $\hat{y}$ -directions are included. Fig. 4.25 is placed directly below Fig. 4.24 for the convenience of comparisons. As the experiment was designed to measure the relative radiation level of the patterned ground structures, the same levels of E-fields of the measured results as those of the simulated results are not expected. Due to the system loss, connection mismatch, and the non-unit gain of the antenna used in the experiment, the measured results

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are lower than the simulated results.

In the measured results in Fig. 4.25, the high magnitudes of the y -components of the structure with a slot and the high magnitudes of the x -components of the Z-shaped structures are observed. The U-shaped structure shows low magnitudes of both the x - and the y -components. Moreover, Fig. 4.25 shows the nulls of the x -components near $\theta = 90^\circ$ and 180° for the microstrip line, the structures with a rectangular and a U-shaped slot at 3 GHz. In Fig. 4.25, the observed trends of the relative radiation (S_{21} -parameters of the system in Fig. 4.27) agree well with those of the simulated E-fields in the far-field region as shown in Fig. 4.24. The trends are not exactly the same because matchings at the inputs of the DUT's are not ideal in the experiments and the reflections at the inputs vary.

In this part of study, the radiation of patterned ground plane structures has been studied theoretically and experimentally using a microstrip line with a slot in the ground plane and its variants. Fields in the apertures of patterned ground structures have been analyzed. In the apertures, E-fields are excited in the direction of the microstrip line. They are guided along the aperture even when it is bent. Thus, E-fields in a slot can be guided to obtain enhancement or elimination of radiation at certain polarization. Examples have been presented to demonstrate the enhancement and the elimination. The patterned ground structure with a slot and the Z-shaped structure show high radiation in one direction and eliminations of radiation in the perpendicular direction. For these structures, the radiation and the interference they generate can be eliminated by depolarizing the structures with respect to the circuit components nearby. The U-shaped structure shows low radiation levels in both directions under study. Therefore, it generates relatively less interference to the environment and reciprocally, it is more immune to a radiation source nearby. Moreover, frequencies play an important role. Operating at the resonant frequency should be avoided due to the high radiation level at the

resonance.

4.4 Summary

In this chapter, detailed studies of the electromagnetic susceptibility and electromagnetic compatibility of patterned ground structures have been presented. The susceptibility of the dual planar EBG microstrip structure proposed by the author [A 8] has been studied when the structure is exposed to radiation or coupling.

For a radiating interference source, the effects of the frequency and the power of the interference source are severe. The interference level is the highest when the frequency of the signal transmitted in the DUT equals that of the interference source. The amount of interference decreases linearly as the power level of the interference source decreases. Besides the interference source, the relative position and polarization between the DUT and the interference source play an important role. The ground plane with etched patches has a better immunity to radiation compared to the microstrip line plane. Besides a large distance between the interference and the EBG structure, a cross polarization between the two structures is an effective way to eliminate the interference suffered by the structure.

For an interference source through coupling, the susceptibility of the EBG structure has been examined when the interference source (a microstrip line) nearby is transmitting or is not transmitting a signal. When the microstrip line is not transmitting a signal, it should be terminated with matched loads to eliminate interference. If matched-load terminations are not possible, it should be kept beyond a safe distance where the interference level experienced by the victim structure is acceptable. This safe distance is determined by the geometry of the EBG structure and the parameters of the substrate. When both the interference source and the

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EBG are transmitting signals, to lower the crosstalk level, the microstrip line in close proximity should work at a frequency that is different from the frequency and/or the harmonics of the signal in the EBG structure. Moreover, it should work at a lower power level than that of the signal in the EBG structure. If the working conditions above can not be satisfied, a low working frequency is preferred. Moreover, the crosstalk level can be lowered by increasing the physical distance between the two structure. It can be significantly lowered by guiding signals in appropriate directions.

The radiation generated by patterned ground structures in the far-field region has also been studied. The radiation of a patterned ground structure with a slot (an elongated rectangular aperture) in the ground plane and its variants, a Z-shaped and a U-shaped structures, have been studied. The studies of radiation are based on the analysis of fields in the slots, simulations, and experimental results. Based on different configurations, polarized high radiation has been found in the structures with a rectangular and a Z-shaped slots while low radiation has been exhibited in the U-shaped structure.

In this chapter, the interference introduced by a source nearby to an EBG microstrip filter through either radiation or coupling has been studied and the corresponding solutions to lower the interference levels have been proposed. Study on the interference generated by patterned ground structures has been started and the possible eliminations have been suggested. Through the study in this chapter, knowledge on the effects of radiation and coupling on patterned ground structures and the radiation generated by the structures has been obtained. It provides useful guidance for the applications and integrations of EBG microstrip structures to microwave circuits without any additional shielding mechanism. An EBG microstrip structure is a typical microstrip structure with a patterned ground plane. This study is useful to EMS and EMC studies of patterned ground structures in

4.4 Summary

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general.

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Chapter 5

Effects of Ground Plane Geometries on Scattering

The effects of ground planes with etched patterns on the propagation, coupling and radiation of electromagnetic waves have been studied in microwave devices and the studies have been presented in the previous chapters. The effects of ground planes on the fourth behavior of electromagnetic waves, the scattering, are to be studied at optical frequencies by studying surface-enhanced Raman scattering (SERS).

Recently, SERS has been the subject of renewed interest, more than 30 years after the original observation of this effect. This strong interest, both in the experimental and simulation areas, is attributable to progress toward the practical realization of SERS, enabled by new nanofabrication techniques [125, 93]. Besides the technology “push”, there has been growing interest on the application “pull” side for chemical sensing and biomolecule analysis. High intensity of E-fields available at the surfaces of nanoparticles is desired in a SERS substrate.

In this chapter, the effects of a ground plane on the scattering are examined using SERS active substrates working at optical frequencies. A ground plane with bulk metal is proposed to be incorporated to the bottom of a SERS substrate to reflect

(a) 3-D view

(b) 2-D side view of the unit cell of a dimer

Figure 5.1: Periodic dimers on a layered substrate.

waves and to increase the intensity of E-fields at the nanoparticle surfaces. In this chapter, an analytical method to quantify the effect of a substrate on the E-fields at the nanoparticle surfaces in SERS is presented. This approach provides fundamental guidance to the construction of a full SERS device, including nanoparticles and the supporting substrate. It can be used to maximize the E-field available at the surfaces of the nanoparticles and to avoid a cancelation of the fields. The maximization can contribute to a considerable increase in the overall Raman enhancement factor. The method is verified using numerical simulations. With the

proposed method, the application of a ground plane in a SERS substrate is proved to be effective in increasing the intensity of E-fields at the nanoparticle surfaces.

5.1 Analytical Method

An ideal SERS substrate may consist of a periodic array of nanostructures, each consisting of a dimer of metal nano-islands, on a layered substrate. Fig. 5.1 (a) and (b) show the three-dimensional (3D) view of such an array on top of a layered substrate and the 2D side view of the unit cell of a dimer, respectively. As can be seen in Fig. 5.1 (a), a dimer has two islands with a gap of distance g between them. They are repeated with a periodicity in the $\hat{x}$ - and $\hat{y}$ -directions, p_x and p_y , respectively. They are located on a substrate comprising a dielectric layer of thickness t_1 and a metal layer of thickness t_2 . The dielectric material can be Si or SiO₂ [23][98] whereas the metal at the bottom can be gold, silver, aluminum etc. Fig. 5.1 (b) shows the side view of the unit cell of a dimer when viewed along the $\hat{y}$ -direction (either positive or negative). In what follows, however, we will analyze only arrays of monomers, since our goal is to highlight the effect of the substrate and the fractional composition of the array layer. Further optimization of the array nanostructures, for instance by designing dimers in the unit cell, is the subject of separate studies.

5.1.1 E-Fields on the Surface of a Substrate

The effect of a substrate on the E-field available on the surface of metal nanoparticles is global as compared to the effect due to the metal nanoparticle on the localized E-fields near the particles. Therefore, in the first stage of the analysis, we study the effect of a substrate on the E-fields on the surface of the substrate, while the islands are temporarily ignored. Fig. 5.2 shows this geometry. The substrate

Figure 5.2: Schematic of the layered substrate.

consists of a Si- and a Au-layer with a boundary at $z = d_1$ between them. They are both assumed to be homogeneous and infinite in the $\hat{x}$ -direction, thus fringing fields can be neglected and $\partial/\partial x = 0$. The thickness of the Si-layer on the top is d_{Si} ($d_{Si} = d_1 - d_0$) and that of the Au-layer at the bottom is d_{Au} ($d_{Au} = d_2 - d_1$). The substrate has a surface at $z = d_0$ and a bottom at $z = d_2$. It splits the space into four regions as numbered in Fig. 5.2. Region 1 and Region 4 are filled with air. The relative permittivity and permeability of the material in the n^{th} region are denoted as ϵ_n and μ_n , respectively, where $n = 1, 2, 3$, and 4. The permittivity and permeability of air are ϵ_0 and μ_0 , respectively.

When a TEM plane wave is normally incident onto the stratified substrate as shown in Fig. 5.2, the total field in region n can be expressed as [121, Chapter 4, p. 365 - 486]

$$\bar{E}_n = \hat{x}(A_n e^{ik_n z} + B_n e^{-ik_n z}) \quad (5.1)$$

$$\bar{H}_n = \frac{1}{\omega \mu_n \mu_0} \bar{k}_n \times \bar{E}_n \quad (5.2)$$

where $\bar{k}_n$ is the wave vector and $\bar{E}_n$ is the E-field strength. A_n and B_n are the E-field amplitudes of the downward and upward waves, respectively. $\bar{H}_n$ is the

magnetic field strength. Subscript n indicates the n^{th} region.

For a propagating plane wave, the E-field amplitude of the upward wave in Region 4 is $B_4 = 0$, and A_1 in Region 1 is assumed to be 1 for the purpose of normalization. In each region, A_n and B_n are related by boundary conditions. At $z = d_n$, E_x and H_y are continuous. By matching boundary conditions, a recurrent expression for B_n/A_n can be obtained as below [121, Chapter 4, p. 365 - 486],

$$\frac{B_n}{A_n} = \frac{R_{n(n+1)}e^{i2k_{n+1}d_n} + (B_{n+1}/A_{n+1})e^{i2k_nd_n}}{e^{i2k_{n+1}d_n} + R_{n(n+1)}(B_{n+1}/A_{n+1})} \quad (5.3)$$

where

$$R_{n(n+1)} = \frac{1 - p_{n(n+1)}}{1 + p_{n(n+1)}} \quad (5.4)$$

$$p_{n(n+1)} = \frac{\mu_n k_{n+1}}{\mu_{n+1} k_n} \quad (5.5)$$

and B_n/A_n is expressed in terms of B_{n+1}/A_{n+1} . Therefore, with an incident wave propagating towards the slab, the reflected E-field can be obtained at any point along the z axis when the parameters of the material and the thickness of each layer are known.

Next, the effect of d_{Si} and d_{Au} on the E-fields at $z = d_0$ is studied. A TEM plane wave at 785 nm incident normally towards the stratified substrate in the $\hat{z}$ -direction is considered. The permeabilities, μ_2 and μ_3 both equal to 1. The loss tangent of Si is 0.004 and its relative permittivity, ϵ_2 , is $11.92 + 0.04i$ at 785 nm. The relative permittivity of Au at 785 nm is $-24.70 + 1.76i$ [126]. Based on (5.3)–(5.5), the magnitude of the sum of E-fields at $z = d_0$, including the incident and the reflected E-fields, is calculated when the thickness of the Si-layer is varied and $d_{Au} = 0$ nm, 25 nm, 50 nm, and 100 nm. As $A_1 = 1$, the sum of the E-fields is normalized and can be expressed as $|\overline{E}_i + \overline{E}_r|/|\overline{E}_i|$, where $\overline{E}_i$ and $\overline{E}_r$ are the incident and the reflected E-fields, respectively. The calculated results are plotted in Fig. 5.3.

Figure 5.3: The normalized magnitude of the E-fields on the surface of the substrate ($z = d_0$) when the thickness of the Si-layer is varied and $d_{Au} = 0$ nm, 25 nm, 50 nm, and 100 nm.

In Fig. 5.3, the solid line depicts the magnitude of the E-fields on the surface of the substrate when there is no Au-layer at the bottom ($d_{Au} = 0$ nm). When $d_{Si} = 0$ nm and there is no substrate, the magnitude of the E-field is 1.00. This magnitude varies between 0.16 to 1.00 as the thickness of the Si-layer changes from 0 nm to 200 nm. It has the first peak value of 1.00 at $d_{Si} = 0$ nm, the first minimum value of 0.16 at $d_{Si} = 57$ nm, and the second peak value of 0.99 at $d_{Si} = 114$ nm. The peak value decreases as the thickness of the Si-layer increases, which is due to absorptive losses in Si. When a Au-layer is incorporated into the substrate, the maximum magnitude of the E-field available on the surface of the substrate increases significantly, as shown by the other lines in Fig. 5.3. We note that, in all cases with a finite Au-layer, the magnitude of the E-field varies between 0 and 2 as d_{Si} varies. When the thickness of the Au-layer is 100 nm, as shown by the dashed line in Fig. 5.3, the magnitude of the E-field varies from 0.01 to 1.87 when d_{Si} varies. It has peak values of 1.87 and 1.84 when d_{Si} is 35 nm and 148 nm, respectively. As

Figure 5.4: Schematic of the layered substrate with an island-layer.

in the case when there is no Au-layer in the substrate, the peak value decreases as d_{Si} increases, owing to the losses in Si. When the thickness of the Au-layer at the bottom decreases, the maximum magnitude of the E-field at $z = d_0$ decreases. When $d_{\text{Au}} = 50$ nm, the magnitude has a maximum and a minimum value of 1.82 and 0.02 at $d_{\text{Si}} = 34$ nm and 91 nm, respectively. When $d_{\text{Au}} = 25$ nm, the maximum magnitude is found to be 1.56 at $d_{\text{Si}} = 31$ nm while the minimum value is 0.05 at $d_{\text{Si}} = 87$ nm. The thickness of the Si-layer where the maximum E-field is found decreases as d_{Au} decreases.

The results above clearly indicate that the thicknesses of both the Si-layer and the Au-layer in the substrate play an important role in determining the magnitude of the E-field on the surface of the substrate. The magnitude of the E-field depends on the properties of the materials, the thickness of each layer, and the frequency of the wave. Without a Au-layer at the bottom, there is no gain of total incident E-field on the surface of the substrate. Moreover, the thickness of the Si-layer should be set appropriately in order to avoid a large cancelation by the reflected wave

Figure 5.5: 3-D view of the unit cell of a monomer.

(to avoid the points where $|\vec{E}_i + \vec{E}_r|/|\vec{E}_i|$ is minimum). By using a Au-layer, the available E-field on the surface of the substrate can reach a factor of 1.8 or more of the incident E-field (capped by a factor of 2). Using this method, an optimal thickness of each layer can be obtained analytically so as to maximize the E-field on the surface of the substrate.

5.1.2 E-Fields on the Surface of Nanoparticles

In the next stage of our analysis, the E-field on the surface of the nanoparticles is studied by adding islands onto the substrate. The islands form an additional layer that we call island-layer. Fig. 5.4 shows the schematic of an island-layer on top of the substrate that was studied previously. The island-layer extends from $z = d_0$ to $z = d_1$. Its thickness is d_{isl} . ($d_{isl} = d_1 - d_0$). Following the same analytical method introduced in the previous section, with the known permittivity, permeability, and thickness of each layer, the magnitude of the E-field on the surface of the nanoparticles ($z = d_0$) can be obtained using (5.3)–(5.5) above.

The relative permittivity of the island-layer, ϵ_2 , is an effective permittivity which depends on the percentage of islands in a unit cell. The Maxwell Garnett equation

shown below is used to estimate the effective permittivity, ϵ_2 [127].

$$\epsilon_{eff} = \epsilon_b \left[\frac{1 + 2f(\epsilon_i - \epsilon_b)/(\epsilon_i + 2\epsilon_b)}{1 - f(\epsilon_i - \epsilon_b)/(\epsilon_i + 2\epsilon_b)} \right] \quad (5.6)$$

where f is the fraction of the metallic inclusion (the islands) in the island-layer, ϵ_i and ϵ_b are the permittivity of the inclusion and of the background material, respectively. The fraction f is determined by the volume of the nanoparticle and its periodicities, p_x and p_y [see Fig. 5.1].

As an example, periodic monomers with a square island in each unit cell are studied. Fig. 5.5 shows the 3D view of such a monomer where a , b , and c are the lengths of the monomer in the $\hat{x}$ -, $\hat{y}$ -, and $\hat{z}$ -directions, respectively ($c = d_{\text{isl}}$). The dimensions of the monomer in the $\hat{x}$ - and $\hat{y}$ -directions, a and b , respectively, are both set to be 164.4 nm. The periodicities, p_x and p_y , are both set to be 400 nm. The fraction f is determined to be 16.89% with air as the background material. $\epsilon_2 = \epsilon_{eff}$ is calculated to be $1.717 + 0.004i$. Based on (5.3)–(5.5), the magnitude of the E-field on the nanoparticle surface ($z = d_0$) is calculated when $f = 16.89\%$ and compared to the case without islands, i.e., $f = 0\%$. The thickness of the island-layer is set to be 25 nm and that of the Au-layer at the bottom is set to be 50 nm. Fig. 5.6 (a) shows the calculated magnitudes of the E-fields versus the thickness of the Si-layer. The case when $f = 16.89\%$ and there is no Au-layer at the bottom of the substrate ($d_{\text{Au}} = 0$ nm) is also included.

As shown in Fig. 5.6 (a), when $f = 16.89\%$ and no Au-layer is at the bottom, the E-field is 0.98 when $d_{\text{Si}} = 0$ nm. This means that the E-field available on the surface of the islands (nanoparticles) has a magnitude of less than one when no substrate is supporting the islands. It is not the maximum value of E-field in this case. As shown by the same curve in Fig. 5.6 (a), when d_{Si} increases and the islands are located on a Si slab, the magnitude of the E-field on the surface of the islands

(a) Islands on Si-layer

(a) Islands on Au film

Figure 5.6: The normalized magnitude of the electric field at the first boundary ($z = d_0$) of the layered substrate in Fig. 5.4.

varies. It has the first minimum value of 0.21 at $d_{\text{Si}} = 36$ nm and a maximum value of 1.05 at $d_{\text{Si}} = 106$ nm. When there is a 50 nm-Au-layer at the bottom and $f = 16.89\%$, the maximum magnitude increases significantly, which is shown by the dashed curve in Fig. 5.6 (a). In this case, the magnitude of the E-field varies from 1.83 to 0.03 when $d_{\text{Si}} = 30$ nm and 68 nm, respectively. The magnitude changes considerably for different thicknesses of the Si-layer. As revealed by these results, nanoparticles with a reflecting Au-layer have an additional enhancement factor of the E-field of 1.74 ($1.83/1.05$) over the case without a Au-layer. Comparing within the case when $d_{\text{Au}} = 50$ nm, an optimal substrate is possible to be obtained providing an additional enhancement factor of the E-field of more than 50 ($1.83/0.03$) over a poorly constructed substrate. This is equivalent to a differential factor of $(|\bar{E}|/|\bar{E}_i|)^4 > 10^7$ ($(1.83/0.03)^4$) where $\bar{E}$ is the total E-field on the surface of the nanoparticles.

At $f = 0\%$, the magnitude of the E-field has a maximum value of 1.83 at $d_{\text{Si}} = 32$ nm, which is close to that at $f = 16.89\%$. Thus, in these two cases where Si is in the middle of the island-layer and the Au-layer, the maximum E-field and d_{Si} corresponding to the maximum do not vary much as the percentage of the islands in the island-layer changes ($f \leq 16.89\%$). The difference in d_{Si} could be large if Si is replaced by other dielectric materials such as SiO_2 . Here we have analyzed Si specifically, because of its good thermal conductivity.

In the case where metal nanoparticles are deposited directly onto a metal film, this model also provides estimations of the global effect of the substrate. In this case, d_{Si} in Fig. 5.4 is set to be zero. Fig. 5.6 (b) shows the calculated magnitude of the E-field at $z = d_0$ when islands ($164.4 \text{ nm} \times 164.4 \text{ nm} \times 25 \text{ nm}$) are located on top of a Au film at a periodicity of 400 nm in both the $\hat{x}$ - and $\hat{y}$ -directions. The horizontal axis in Fig. 5.6 (b) is the thickness of the Au film. Both cases when $f = 16.89\%$ and $f = 0\%$ are included. At $f = 16.89\%$, the magnitude of the E-field has a peak

Table 5.1: E-fields on the Surface of Nanoparticles (on Si-Layer)

d_{Si} (nm)	Max. $ \overline{E} / \overline{E}_i $ (CST)	$ \overline{E}_i + \overline{E}_r / \overline{E}_i $ at $z = d_0$
15	10.25	1.25
30	18.90	1.83
50	5.18	0.61
68	3.11	0.03

Table 5.2: E-fields on the Surface of Nanoparticles (on Si-Layer, Optimized)

$a = b$ (nm)	d_{Si} (nm)	Max. $ \overline{E} / \overline{E}_i $ (CST)
222.5	30	20.66
40	68	0.54

value of 1.08 at $d_{\text{Au}} = 5$ nm and it is largely unchanged at $d_{\text{Au}} > 60$ nm. At $f = 0\%$, the peak value of the magnitude is 1.06 when the thickness of the Au film is 4 nm. It is also unchanged at $d_{\text{Au}} > 60$ nm. For Au nanoparticles on Au film, the optimal thickness of the metal film does not vary significantly as the percentage of the nanoparticles in the island-layer changes ($f \leq 16.89\%$).

5.2 Computational Verifications

The analytical results obtained in the previous section are verified using a numerical simulation tool, CST Microwave Studio [2]. Fig. 5.5 shows the 3D view of the unit cell of a monomer under simulation. The dimensions of the monomer in the $\hat{x}$ - and $\hat{y}$ -directions, a and b are both fixed at 164.4 nm and the thickness of the island, c is 25 nm ($c = d_{\text{isl.}}$). The period is 400 nm in both the $\hat{x}$ - and $\hat{y}$ -directions. The thickness of the Au-layer at the bottom, d_{Au} is 50 nm, and d_{Si} is set to be 15 nm,

Table 5.3: E-fields on the Surface of Nanoparticles (on Au-Layer)

d_{Au} (nm)	Max. $ \overline{E} / \overline{E}_i $ (CST)	$ \overline{E}_i + \overline{E}_r / \overline{E}_i $ at $z = d_0$
5	8.58	1.07
20	5.17	0.90
60	4.79	0.80
80	2.38	0.80
100	3.67	0.80

30 nm, 50 nm, and 68 nm. The simulated maximum magnitude of the E-fields on the surface of the monomer are extracted, normalized to that of the incident E-field, and tabulated in Table 5.1. The calculated magnitudes of the E-fields at the corresponding d_{Si} 's are extracted from Fig. 5.6 (a) and tabulated in Table 5.1. In the simulation results, the dimension of the island in the $\hat{x}$ -, $\hat{y}$ -, and $\hat{z}$ -directions and the periodicity in the $\hat{x}$ - and $\hat{y}$ -directions affect the E-field on the surface of the islands. Therefore, the simulated $|\overline{E}|/|\overline{E}_i|$ includes both the local effect in each direction and the global effects in both the $\hat{x}$ - and $\hat{y}$ -directions. This is a more accurate depiction of the E-field behavior than the analytical calculations in Section 5.1, which address only global effects. Nevertheless, the trends are quite similar, as shown in Table 5.1. When d_{Si} is changed, the simulation results still show an increase and then a decrease, with CST-calculated absolute values being $\sim 8-10\times$ higher than those obtained by using the analytical method.

The dimensions of the islands on top of a 30 nm- and 68 nm-Si-layer can be further optimized to obtain local resonances at 785 nm in each case, by changing the dimensions of the island. In these simulations, there is a 50 nm Au-layer at the bottom. Table 5.2 shows the simulated normalized E-fields in both cases. As can

be seen in Table 5.2, the substrate affects the local resonance in the island. A $222.5 \text{ nm} \times 222.5 \text{ nm} \times 25 \text{ nm}$ island on 30 nm-Si-layer has a resonance at 785 nm and it shows a maximum normalized magnitude of the E-field of 20.66. On the other hand, the monomer on 68 nm-Si-layer has a resonance at the same wavelength when the island has a dimensions of 40 nm, 40 nm, and 25 nm in length, width, and height, respectively. It shows $|\bar{E}|/|\bar{E}_i| = 0.54$. The results reveal that, even at the local resonance of the monomer that has a 68 nm-Si-layer, the monomer with an optimal 30 nm-Si-layer can provide a factor of 38.26 for the enhancement of the E-field over that of the 68 nm-Si-layer. The factor of 38.26 for the enhancement of the E-field corresponds to a SERS enhancement factor of $(|\bar{E}|/|\bar{E}_i|)^4 > 10^6$. This comparison indicates that the effect of a substrate in SERS can even overshadow the local effects and can become very significant. Therefore, this effect cannot be neglected and should be carefully addressed in any design of SERS devices.

Au nanoparticles on a Au film were also simulated with fixed dimensions of the monomer ($164.4 \text{ nm} \times 164.4 \text{ nm} \times 25 \text{ nm}$), while the thickness of the Au film was varied. The periodicities, p_x and p_y were both fixed at 400 nm. Table 5.3 shows the simulated normalized magnitudes of the E-fields on the surface of the nanoparticles when the periodic monomers are located directly on a Au film. It also tabulates the extracted calculated magnitudes of the E-fields at the corresponding thicknesses of the Au film using the proposed analytical method. Since the dimension of the island and the periodicity are fixed in the simulations, the local and the global effects in the $\hat{x}$ - and $\hat{y}$ -directions still remain. While the absolute values shown in Table 5.3 differ, the trend of the computationally obtained data is in good qualitative agreement with that of the results based on the proposed analytical method of Section II. The fluctuations in the computationally simulated results when $d_{\text{Au}} \geq 60 \text{ nm}$ may be caused by the local resonances in the islands, and these are of course not accounted for in the analytical model.

5.3 Summary

In this chapter, an analytical method to study the global effect of a substrate on the surface E-field of nanoparticles in SERS has been proposed. The effect of the substrate is quantified in terms of the parameters of material and the thickness of each layer. A Au-layer at the bottom of the substrate provides an additional factor of 1.7 of the magnitude of the E-field over that when there is no Au-layer incorporated. By applying the analytical method, an optimal construction of the substrate can be obtained while a poor one can be avoided. An optimized substrate results in an E-field that is a factor of more than 50 higher than that of a poorly constructed one. A substrate consisting of a Si-layer and a Au-layer, as well as the case where nanoparticles are directly located on a Au film, have both been analyzed. It is shown that the global effect of the substrate can become significant and even overshadow the strong local electromagnetic effects in SERS. An optimized substrate can contribute to a significant increase in the Raman enhancement factor.

Chapter 6

Conclusions and Recommendations

In this thesis, the effects of ground plane geometries on electromagnetic waves in terms of propagations, coupling and radiation, and scattering have been studied and presented using various structures in different frequency bands in the electromagnetic spectrum.

The effects on wave propagations have been demonstrated using an E-shaped patterned ground structure, a straight and a U-shaped dual planar electromagnetic band-gap structures, and an optimization of conventional stepped-impedance low-pass filter using a microstrip line with a slot in the ground plane. A single etched unit and an array of periodic etched patterns in the ground plane of a microstrip line provide a sharp notch and a wide stopband, respectively. They both have high attenuation in the stopbands. A circuit model has been successfully proposed to study the bandstop and bandpass behaviors of waves in a periodic microstrip structure.

The proposed E-shaped patterned ground structure has a single etched unit in the ground plane of a microstrip line. It shows a resonance with sharper transition

knees in the frequency response, a smaller circuit area, and a higher Q-factor (lower loss) than other patterned ground microstrip structures in the literature. This new etched pattern leads to the realization of a tunable bandstop filter with a small circuit area. The tunable E-shaped bandstop filter achieves a tuning range of 12.4 %.

Besides patterned ground structures with a single etched pattern, periodic etched patterns in the ground plane of an EBG structure have been proposed to work with periodically inserted patches in the microstrip line forming a dual planar configuration. With the proposed dual planar configuration, the stopband is widened and its attenuation is increased without introducing any increase in the circuit area. The ripple levels in the passbands are low due to the application of tapering techniques. They are less than 1 dB in the designs.

The proposed dual planar EBG structure was bent into a U-shape to achieve a further reduction of the circuit area. The circuit area of the U-shaped structure is only 54.4 % of the corresponding straight structure. Taking the advantages of a bent capacitively modulated EBG structure (where there are periodically inserted patches in the microstrip line), such as no additional phase shift and no high power loss due to bending, the U-shaped EBG structure shows a wide stopband with high attenuation and a small ripple level in the passband.

Other than new designs, the effects of ground planes on the propagations of waves have also been demonstrated in an optimization of a conventional filter structure using a patterned ground structure. A microstrip line with a longitudinal slot has been applied to implement the inductors in a ladder low-pass filter prototype, resulting a slot-back stepped-impedance low-pass filter. Due to the enhanced inductance and the high impedance of the slot-back microstrip line, the optimized filter shows a shaper cutoff, a better suppression of spurious passbands, and a much smaller circuit area than a conventional stepped-impedance low-pass filter.

The unique advantage of the optimizations applying patterned ground structures is that the optimizations not only require no additional circuit areas because of the location where the patterns are etched, but also reduce the size of the original device due to the slow-wave effect.

The effects of ground plane geometries on coupling and radiation have been shown through the study of the electromagnetic compatibility and electromagnetic susceptibility of a dual planar EBG microstrip structure. The effects of both radiating components and transmission lines on the performances of patterned ground structures have been examined and analyzed. The radiation generated by patterned ground structures in the far-field region has also been studied to examine the electromagnetic effects of the structures on other circuit components. The knowledge obtained are useful to guide the integrations and incorporations of a patterned ground microstrip structure to a microwave integrated circuit without using any additional shielding mechanism.

The effects of ground planar geometries on scattering have been examined using a surface-enhanced Raman scattering (SERS) active substrate at optical frequencies. A metallic ground plane has been proposed at the bottom of a SERS substrate to reflect waves and to increase the E-field intensity at the surfaces of nanoparticles. In the example that has been shown, a Au-layer at the bottom of the substrate provides an additional factor of 1.7 of the magnitude of the E-field over that when there is no Au-layer incorporated. An analytic method has been proposed to quantify the effect of the metallic-layer-bottomed substrate in terms of the dielectric constant and the thickness of each layer on the E-fields that are available at the surfaces of the particles. This method can be used to optimize the construction of a SERS substrate and to avoid poorly constructed substrates. It has been shown that this global effect of a substrate in SERS can become significant and even overshadow the strong local electromagnetic effects. An optimized SERS substrate can

contribute to a significant increase in the Raman enhancement factor.

In this thesis, the effects of ground planes on electromagnetic waves have been investigated for different mechanisms of electromagnetic waves. The thesis contributes to the understanding of waves in a transmission line with patterned ground planes. Novel structures with low loss, high performances, and small physical sizes have been proposed, analyzed, and implemented. The thesis also contributes to the knowledge on how a microstrip line with patterned ground structure is affected by radiation and coupling nearby as well as how it radiates. It is useful for understanding the EMC properties of this type of structures and for guiding the integrations of the structures into a microwave circuit directly. Moreover, this thesis elaborates the scattering of waves in a SERS substrate with bulk metal at the bottom, uncovering rigorously the effect of a substrate in SERS experiments which is crucial but has been an unknown for years.

Some of the work presented in this thesis are still at their initial stages of research. They have great potential for being further explored and there would be a lot of more work that should be done in the near future.

- The method to calculate the transmission coefficients of a capacitively modulated EBG structure is worth being extended to calculate the transmission coefficients of patterned ground structures with arbitrary etched patterns (a single etched unit or an array of periodic units).
- The radiation generated by patterned ground structures in the near-field region is a challenging but useful research area. The related research will supplement the knowledge on the EMC of patterned ground structures. It will be useful to solve realistic EMC problems when a patterned ground structure is integrated into a microwave circuit.
- The research on electromagnetic modeling of nanostructures at optical fre-

quencies is at its initial stage. Experimental results dominate in this area. More theoretical studies should be done to model and to understand the interactions of electromagnetic waves with nanostructures.

Appendix A

Author's Publications

Journal Paper

- [A 1] S. Y. Huang and Y. H. Lee, "Investigation into the susceptibility of an electromagnetic band-gap microstrip filter," to appear in *IEEE Transactions on Electromagnetic Compatibility*, May, 2010.
- [A 2] S. Y. Huang and Y. H. Lee, "Fast and Accurate Calculation of Transmission Coefficients for an EBG Microstrip Structure," *Microwave and Optical Technology Letters*, vol. 52, no. 4, pp. 793-797, Feb., 2010.
- [A 3] S. Y. Huang, Bae-Ian Wu, Baile Zhang, Y. H. Lee, Vladimir Liberman, and Mordechai Rothschild, "An Analytical Method to Study the Effects of a Substrate in Surface-Enhanced Raman Scattering," *Journal of Applied Physics*, 106, 114306, Dec., 2009.
- [A 4] Y. H. Lee, and S. Y. Huang, "Electromagnetic susceptibility of an electromagnetic band-gap filter structure," *Progress in Electromagnetics Research B*, vol. 15, pp. 31-56, 2009.
- [A 5] S. Y. Huang, and Y. H. Lee, "A compact E-shaped patterned ground structure and its applications to tunable bandstop resonator," *IEEE Transactions*

on *Microwave Theory and Techniques*, vol. 57, no. 3, pp. 657-666, March, 2009.

[A 6] S. Y. Huang, and Y. H. Lee, "Compact stepped-impedance low-pass filter with a slot-back microstrip line," *Microwave and Optical Technology Letters*, vol. 50, no. 4, pp. 1058-1061, April, 2008.

[A 7] S. Y. Huang, and Y. H. Lee, "Compact U-shaped dual planar EBG microstrip lowpass filter," *IEEE Transactions on Microwave Theory and Techniques*, vol. 53, no. 12, pp. 3799-3805, Dec., 2005.

[A 8] S. Y. Huang, and Y. H. Lee, "Tapered dual-plane compact electromagnetic band-gap microstrip filter structures," *IEEE Transactions on Microwave Theory and Techniques*, vol. 53, no. 9, pp. 2656-2664, Sep., 2005.

[A 9] S. Y. Huang and Y. H. Lee, "A Tapered Small-Size EBG Microstrip Bandstop Filter Design with Triple EBG Structures," *Microwave and Optical Technology Letters*, vol. 46, no. 2, pp. 154-158, July, 2005.

Manuscript under Preparation

[A 10] S. Y. Huang and Y. H. Lee, "Studies of the Radiation of Patterned Ground Structures," plan to submit to *IEEE Transaction on Electromagnetic Compatibility*.

Conference Papers

[A 11] S. Y. Huang, and Y. H. Lee, "Development of ultra-wideband (UWB) filters," *2008 URSI General*, Chicago, Illinois, USA, on Aug. 2008.

[A 12] Y. H. Lee, and S. Y. Huang, "Further Investigation into the Electromagnetic Susceptibility of an EBG Microstrip Filter," *Proc. Asian Pacific Microwave Conference 2007*, Bangkok, Thailand, Dec. 2007.

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- [A 13] D. Ran, S. Y. Huang, and Y. L. Lu, "Optimization of Broadband Bandpass Filter Using Tapered Electromagnetic Bandgap Structures," *Proc. Asian Pacific Microwave Conference 2007*, Bangkok, Thailand, Dec. 2007.
 - [A 14] Y. H. Lee, and S. Y. Huang, "A Study on the Electromagnetic Susceptibility of an Electromagnetic Band-gap Microstrip Filter Structure," *Proc. Eleventh URSI Commission F Triennial Open Symposium On Radio Wave Propagation and Remote Sensing*, Rio de Janeiro, Brazil, Oct.-Nov. 2007
 - [A 15] S. Y. Huang, and Y. H. Lee, "A new optimization approach for high performance meander EBG filter design," *Proc. 2007 IEEE International Symposium on Antennas and Propagation*, Honolulu, Hawaii, USA, 10 - 15 Jun. 2007.
 - [A 16] S. Y. Huang, and Y. H. Lee, "A distinction between a step impedance lowpass filter and an EBG microstrip structure," *Proc. 2007 IEEE International Symposium on Antennas and Propagation*, Honolulu, Hawaii, USA, 10 - 15 Jun. 2007.
 - [A 17] Y. H. Lee, and S. Y. Huang, "Microstrip line coupling to a dual-plane Electromagnetic Band-Gap microstrip filter structure," *Proc. EMC Europe 2006* Barcelona, Barcelona, Spain, 4 - 8 Sep. 2006.
 - [A 18] Y. H. Lee, and S. Y. Huang, "Electromagnetic compatibility of a dual-planar EBG microstrip filter structure," *Proc. 17th International Zurich Symposium On Electromagnetic Compatibility*, Singapore, 27 Feb - 3 March. 2006.
 - [A 19] S. Y. Huang, and Y. H. Lee, "Small-size dual-plane EBG microstrip lowpass filter with a U-shaped microstrip line geometry," *Proc. 1st IEEE International Workshop On Radio-Frequency Integration Technology 2005*, Singapore, Dec. 2005.

- [A 20] S. Y. Huang, and Y. H. Lee, "High performance, compact meander dual planar EBG microstrip lowpass filter design," *Proc. 2nd IASTED International Conference on Antennas, Radar, and Wave Propagation*, Banff, Canada, July. 2005.
- [A 21] S. Y. Huang, Y. H. Lee, and Y. Lu, "A novel compact electromagnetic band-gap filter design," *Proc. 2004 Asia-Pacific Microwave Conference*, New Delhi, India, 15-18 December, 2004
- [A 22] S. Y. Huang, Y. H. Lee, and Y. Lu, "Tapered electromagnetic band-gap filter structures," *Proc. 2004 Asia-Pacific Microwave Conference*, New Delhi, India, 15-18 December, 2004
- [A 23] S. Y. Huang and Y. H. Lee, "A novel approach for the control of wave propagation in electromagnetic band-gap structures," *Proc. 4th International Conference on Microwave and Millimeter Wave Technology*, Beijing, P. R. China, 19-22 August, 2004, B9.1
- [A 24] S. Y. Huang and Y. H. Lee, "A novel dual-plane compact electromagnetic band-gap (DPC-EBG) filter design," *Proc. 4th International Conference on Microwave and Millimeter Wave Technology*, Beijing, P. R. China, 19-22 August, 2004, B9.1

Appendix B

Tapering Techniques for Electromagnetic Band-gap Structures

Tapering techniques are effective for eliminating ripples in the passband caused by the periodicity in an EBG microstrip structure. In some tapered EBG microstrip structures, tapering techniques are applied by directly adopting tapering functions (such as Hamming tapering function [128] and Kaiser tapering function [102]) to modify the dimension of the periodic elements. The effects of adopting Bartlett, Hanning, Hamming, Nuttall, and Kaiser tapering function on the transmission of a 1-D planar EBG microstrip structure are compared and studied in [13]. The other approach to taper EBG structures is based on low side lobe array theory [99, p. 294 - 299] including Binomial array, Dolph-Tschebyscheff array (Chebyshev array), and Taylor array. The significant effects of Binomial and Chebyshev distribution on the passband performance of an EBG microstrip structure have been shown by Karmakar *et al* [129]. The effect of the Taylor distribution on the performance of a 1-D EBG coplanar waveguide (CPW) structure was shown in [130].

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For a 1-D planar EBG microstrip structure with a single column of m circles etched in the ground plane of a microstrip line, the distribution of the dimension of the circle follows the following expression:

$$b_i = b_c \cdot T(z_i) \quad (\text{B.1})$$

when tapering functions are applied, and

$$b_i = b_c \cdot a_i \quad (\text{B.2})$$

when low side lobe array theory is applied. i is an integer with a range from 1 to $m/2$ in the case that m is an even number and from 1 to $(m+1)/2$ in the case that m is an odd number. b_i is the radius (or the area) of the i^{th} circle, b_c is the radius (or the area) of the central circle. In (B.1), $T(z)$ is the tapering function, and z_i is the normalized distance between the center of the i^{th} circle and the center point of the structure that can be determined by the following expression:

$$z_i = \frac{2d_i}{L_t} \quad (\text{B.3})$$

where d_i is the distance between the center of the i^{th} circle and the center point of the microstrip line, and L_t is the total length of the EBG microstrip structure. In (B.2), a_i is the i^{th} coefficient in the array. In the case that m is even, the central circle does not exist but it is used to determine the dimension of the circle when employing tapering functions. Whereas when the low side lobe array theory is applied, the two circles at the center are regarded as central circles in the calculation [129].

Table B.1 shows some popular tapering functions used in EBG technology and their plots. In the Kaiser tapering function, $I_0(x)$ is the modified Bessel function of the

Table B.1: Tapering Functions

Type	Tapering Function	Plot
	$x \in [-1, 1]; y \in [0, 1]; a = 1$	
Bartlett	$1 - \frac{ x }{a}$
Connes	$(1 - \frac{x^2}{a^2})^2$
Hamming	$Hm_A(x) = 0.54 + 0.46 \cos(\frac{\pi x}{a})$
Hanning	$Hn_A(x) = \cos^2(\frac{\pi x}{2a})$ $= \frac{1}{2}(1 + \cos(\frac{\pi x}{a}))$
Cosine	$\cos(\frac{\pi x}{2a})$
Welch	$1 - \frac{x^2}{a^2}$
Kaiser	$\frac{I_0(4\sqrt{1-(2x)^2})}{I_0(4)} \quad x \in [-\frac{1}{2}, \frac{1}{2}]$
Blackman	$B_A(x) = 0.42 + 0.5 \cos(\frac{\pi x}{a})$ $+ 0.08 \cos(\frac{2\pi x}{a})$
Gaussian	$e^{-x^2/(2\sigma^2)}$

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Table B.2: Values of Tapering Functions and Normalized Coefficients

Type	$T(z_1)$	$T(z_2)$	$T(z_3)$
Bartlett	0.82	0.46	0.09
Blackman	0.87	0.27	0.01
Connes	0.94	0.49	0.03
Cosine	0.96	0.66	0.14
Gaussian	0.98	0.86	0.66
Hamming	0.93	0.48	0.10
Hanning	0.92	0.43	0.02
Welch	0.97	0.70	0.17
Kaiser	0.94	0.58	0.16
Type	a_1	a_2	a_3
Binomial	1	0.50	0.10
Chebyshev	1	0.73	0.39

first kind. The variance σ^2 in the Gaussian tapering function is generally set to 1 for a normal distribution. According to the definition, the value of the function varies in the range from 0 to 1 corresponding to $x \in [-1, 1]$ with the exception in Kaiser tapering function where $x \in [-\frac{1}{2}, \frac{1}{2}]$.

The coefficients of a Binomial array are determined by the series binomial expansion of the function $(1+x)^{m-1}$ as shown below:

$$\begin{aligned}
 (1+x)^{m-1} = & 1 + (m-1)x + \frac{(m-1)(m-2)}{2!}x^2 \\
 & + \frac{(m-1)(m-2)(m-3)}{3!}x^3 + \dots
 \end{aligned} \tag{B.4}$$

The positive coefficients of the series expansion for a value of m can be used for tapering in a m -element 1-D EBG microstrip structure. The Binomial array has wide variations between the coefficients, especially for an array with a large number of elements. Chebyshev array is a compromise between uniform and Binomial

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array. Its coefficients are related to Tschebyscheff polynomial that satisfies the recurrence relations expressed as following

$$T_m(z) = z \cdot T_{m-1}(z) - \sqrt{(1 - z^2)\{1 - [T_{m-1}(z)]^2\}} \quad (\text{B.5})$$

$$T_m(z) = 2z \cdot T_{m-1}(z) - T_{m-2}(z) \quad (\text{B.6})$$

where $T_1(z) = z$. Letting $z \equiv \cos \theta$, $T_m(z)$ is expressed as

$$T_m(z) = \cos(m \cdot \theta) = \cos[m \cdot \arccos(z)] \quad (\text{B.7})$$

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