An Improvement of Defected Ground Structure Lowpass/Bandpass Filters Using H-Slot Resonators and Coupling Matrix Method

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***Abstract*— A novel compact wideband high-rejection lowpass filter (LPF) using H-DGS is presented. The proposed filter has neither open stub nor cascaded high-low impedance elements. It consists of two coupled H-slots in the ground plane along with a compensated line. The effect of the new slot on the filter performance is examined. The comparison with the conventional filters shows that the proposed one guarantees a large rejected-band of 20dB from 2.5 to 16 GHz. Experimental measurements by means of HP8719D network analyzer agree well with simulated results which are carried out by Microwave Office. Based on H-DGS LPF Structure, a novel bandpass filter (BPF) will be is designed and tested verified by using both *J*-inverter and coupling matrix methods.**

***Index Terms*—** H-defected ground structure (H-DGS), microstrip lowpass filter (LPF), microstrip bandpass filter (BPF).

I. INTRODUCTION

Recently, both defected ground structures (DGS) and electromagnetic band gap (EBG) have received much attention because of their use in radar, satellite, microwave areas and mobile communication systems. Such systems often require circuits to be as small as possible. The DGS components are the dominant technology which can provide size reduction and has capability of harmonics and spurious suppression. The DGS can be applied to various kinds of components such as lowpass filters (LPFs) and bandpass filters (BPFs) as well as RF phase shifters. The DGSs [1-5], which were evolved from EBG and are realized by etching a certain pattern in the backside metallic ground plane which perturbs the current distribution in the ground, and hence increases the effective inductance and capacitance of the microstrip line. Therefore, a DGS cell is equivalent to an LC circuit. The classic microwave lowpass filter (LPF) is implemented either by all stunt stubs or by series connected high-low (Hi-Lo) stepped-impedance microstrip line sections. However, these designs suffer from some disadvantages such as the fabrication difficulties associated with the high impedance lines and the appearance of spurious bands.

In order to overcome these disadvantages, a method has been proposed in [6, 7], which uses both DGS resonators and a compensated microstrip line to design the desired LPF.

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In this paper, a new compact H-slot is presented to serve as a DGS cell element for the microstrip line as shown in Fig. 1. The filter has been fabricated and measured. The measured and simulated results show a good agreement.

II. CHARACTERISTICS OF THE DGS

The proposed DGS is symmetric and consists of four capacitive arms, which are connected to a rectangular slot. All are etched on the bottom of the substrate as shown in Fig. 1. The rectangular slot corresponds to an inductance.

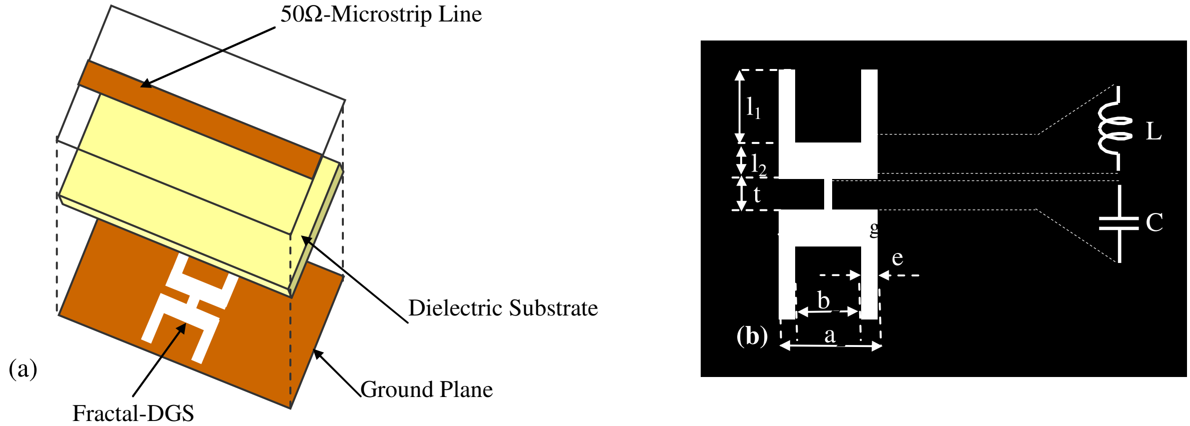

Fig. 1. (a) three and (b) one dimensional view of the DGS cell, (b) the shape of the DGS unit etched on the ground plane of a microstrip line.

The two H-heads are connected together via a slot of 0.6 mm width. This slot channel corresponds to a capacitance, as shown in Fig. 2. In many works, the conventional equivalent circuit parameters are extracted from an electromagnetic simulation by matching to a one pole Butterworth band stop filter response, as discussed in [8]-[9]. In this work, the circuit does not have an easy structure; therefore, the extraction will be realized through the optimization method. The conductor strip of the microstrip line (50Ω) on the top plane has a calculated width W of 1.9 mm. The substrate dielectric constant and thickness (h) are respectively 3.38 and 0.813 mm. The dimensions shown in Fig. 3(a) and their corresponding values are shown in Table I. The DGS cell is simulated using Microwave Office. The simulations of the DGS cell are carried out by Microwave Office and the obtained results are depicted in Fig. 3(b). It shows the characteristic of a one-pole LPF. Accordingly, we can get the attenuation pole frequency (f_0) at 5.2 GHz and the 3-dB cutoff frequency (f_c) of the filter at 2.4 GHz. The H-slot in the ground plane acts as a parallel resonant circuit; it can be modeled by an LC circuit as shown in Fig. 2(a). The values of *L*, *C* and *R* can be computed using:

$$Y = \frac{1}{R} + j2\pi\left(fC - \frac{1}{4\pi fL}\right) \tag{1}$$

$$Z = \left[\frac{1}{R} + j2\pi\left(fC - \frac{1}{4\pi fL}\right)\right]^{-1} \tag{2}$$

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with help of the circuit theory the transmission coefficients are calculated as follows:

$$S_{12} = \frac{1}{1+\frac{1}{2}ZZ_0^{-1}} = \left[1+\frac{1}{2}ZZ_0^{-1}\right]^{-1} \tag{3}$$

$$for\ \ R >> Z_0 \Rightarrow Z = \left[j2\pi\left(fC - \frac{1}{4\pi fL}\right)\right]^{-1} \tag{4}$$

$$|S_{12}| = \left[1+(2Z_0)^{-2}\left(4\pi^2\left(\frac{1}{4\pi^2 fL} - fC\right)^2\right)^{-1}\right]^{\frac{1}{2}} = \left[1+(2Z_0)^{-2}\left(\frac{1}{\omega L} - \omega C\right)^{-2}\right]^{\frac{-1}{2}} \tag{5}$$

At resonance frequency, the relationship between LC and ω_0 is defined as follows:

$$L = \left[\frac{1}{\omega_0^2 C}\right] \tag{6}$$

Substituting the result of eq. (6) in the eq. (5), gives:

$$|S_{12}| = \left[1+(2Z_0)^{-2}\left(\frac{L(\omega\omega_0)^2}{\omega\left(\omega_0^2-\omega^2\right)}\right)^2\right]^{\frac{1}{2}} = \left[\frac{1}{\sqrt{2}}\right]_{\omega=\omega_c} \tag{7}$$

$$L = \frac{2Z_0}{\omega_0^2}\left[\frac{\left(\omega_0^2-\omega_c^2\right)}{\omega_0}\right] \tag{8}$$

Substituting the eq. (8) in the eq. (6), gives:

$$C = \frac{1}{2Z_0}\left[\frac{\omega_c}{\left(\omega_0^2-\omega_c^2\right)}\right] \tag{9}$$

The resistance R in the equivalent circuit model is best fitted around the resonant frequency ω_0. In this case, the equivalent impedance is $Z = R$ (eq. (2)) and the transmission loss S_{12} is:

$$S_{12}\big|_{\omega=\omega_0} = \left[1+\frac{1}{2}RZ_0^{-1}\right]^{-1} \Rightarrow R = 2Z_0\left[\left(S_{21}\big|_{\omega=\omega_0}\right)^{-1} - 1\right] \tag{10}$$

The values of f_0, f_c and $S_{12}\big|_{\omega=\omega_0}$ are easily extracted from the EM-simulation curve of the DGS element.

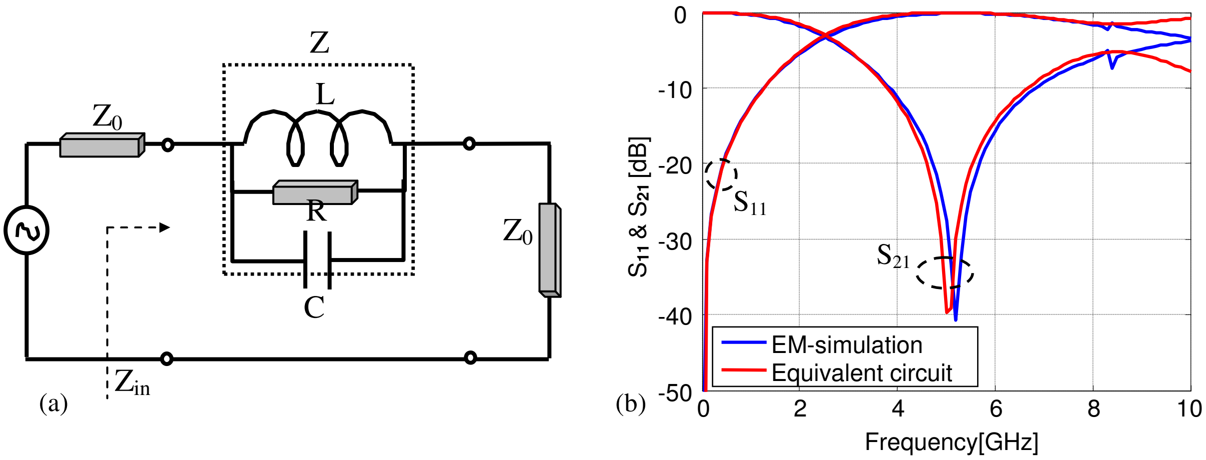

Fig. 2. (a) Equivalent circuit of the DGS cell (b) EM Simulation

III. COMPARISON OF S-PAREMETERS OF DIFFERENT SLOTHEADS

Three fabricated different DGS slot heads with their dimensions are shown in Fig. 3(a). Fig. 3(b) shows the simulation results of the different DGS cells. It is observed that all DGS resonators have the same resonance frequency of 5.2 GHz. From Fig. 3(b), it is clear that the proposed DGS is superior over conventional ones [3]. It gives a deeper rejection in the stopband and smaller losses in the passband, while having the same size as the other conventional DGS. These three DGS resonators can be used to construct a compact filter with intrinsic spurious rejection. First of all, the DGS section is an appropriate for use since it has a low insertion loss. Second, the wide and deep stopband of the DGS section can be employed to suppress the spurious pass bands at harmonic frequencies. All these advantages are due to the slow-wave effect, which appears due to the capacitive coupling and can further reduce the size of the structure. It is therefore possible to design and fabricate a compact LPF with a wide upper stopband.

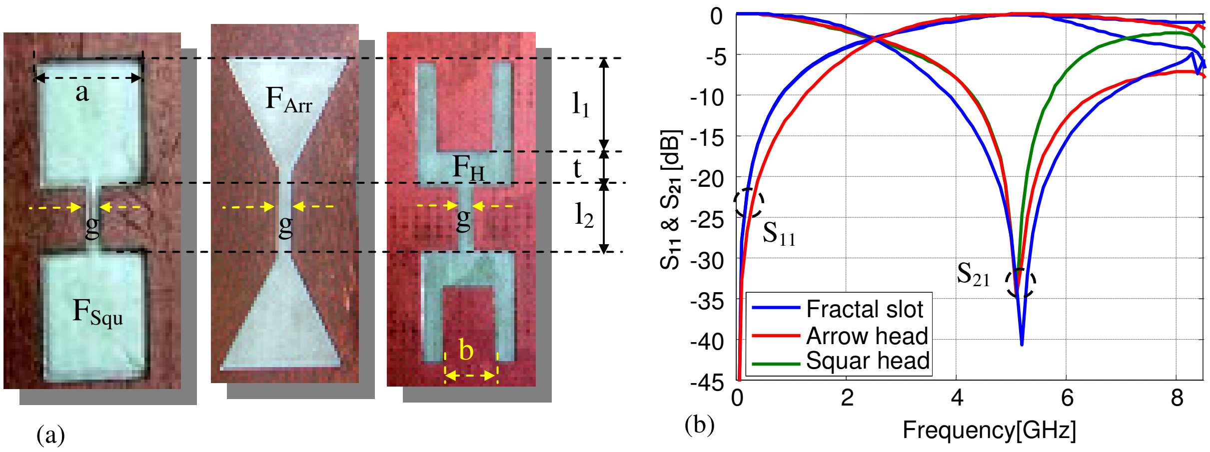

Fig. 3. (a) Three fabricated different DGS slots, (b) S-parameters of the different DGS cells.

The proposed H-filters make use of this fact. In order to prove that, the suggested H-DGS is more favorable than the other conventional DGSs, the slow wave effect method will be used, by placing the three slots on the ground plane of a 50Ω-microstrip line. Their heads will be changed, while keeping the slots' widths (gap g) constant. This loaded microstrip line shows a phase delay of $\pi/2$ at 3.48 GHz. Table I shows the comparison of the three shaped DGS dimensions. It can be observed that the H-DGS has the lowest occupying etched area (area without copper). (see Table I).

TABLE I. DIMENSIONS (IN MM) OF THE DIFFERENT DGS RESONATORS

Square head	**Arrowhead**	**H-head**
$l_1 = 5$	$l_1 = 5$	$l_1 = 5$
$l_2 = 3$	$l_2 = 3$	$l_2 = 3$, $b = 3.2$
$g = 0.6$, a = 5	$g = 0.6$	$g = 0.9$, t = 2
$F_{squa} = 40\text{mm}^2$	$F_{arrow} = 37.5\text{mm}^2$	$F_H = 28.2\text{mm}^2$

IV. INFLUENCE OF H-DGS DIMENSIONS ON THE ATTENNUATION POLE FREQUENCY

The proposed slot is shown in Fig.1. It can provide a cutoff frequency and an attenuation pole at some frequency without any periodic array of DGS [4]. In order to investigate the frequency characteristics of the etched slot, we simulated the DGS unit section using Microwave Office. The placement of the DGS under the microstrip line involves the appearance of a resonance frequency. This effect is due to the increase of both the effective permittivity and the effective inductance of the microstrip. The variation of the dimensions of the etched square area (FH) and the etched gap shifts the cutoff frequency and the attenuation pole location in the frequency domain. As is well known, a resonance frequency can be generated by a combination of inductive and capacitive elements. Therefore, in order to explain the simulated frequency response of the proposed DGS section, we introduced a capacitance in the equivalent circuit. The etched gap area, which is placed under the microstrip line, corresponds to the capacitance and the head area is equivalent to a series inductance. Consequently, the DGS unit is equivalent to a resonant circuit as shown in Fig. 3. The parameters of this DGS equivalent circuit have been found using curve-fitting. They are: C = 1.1pF and L = 2.9nH. In order to investigate the effect of the slot-arm dimensions, the etched gap [5]-[6], which is related to the gap capacitance, was kept constant at 0.9mm and the etched rectangular area was varied. The simulation results are illustrated in Fig. 4(a). As the arm-lengths are increased, both the characteristic impedance and the series inductance of the microstrip line are increased while the cutoff and resonance frequency are decreased. Next, we investigated the effect of the etched gap distance. The slot head dimensions were kept constant and the etched gap distance was varied [7]. The simulation results are shown in Fig. 4(b). We can note that there is no change in the cutoff frequency. In addition, the attenuation pole location moves up to a higher frequency. This means that the gap corresponds only to a capacitance. In the last case, all parameters were kept constant and t was varied, the

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simulation results show that there is no modification. Fig. 4(c) shows strong suppression of the S_{21} response at higher frequencies [3, 6]. This means that the parameter t affects the reject band. By adjusting the value of t we could obtain a wide stopband. These results prove the advantages of the suggested H-DGS compared to the conventional ones.

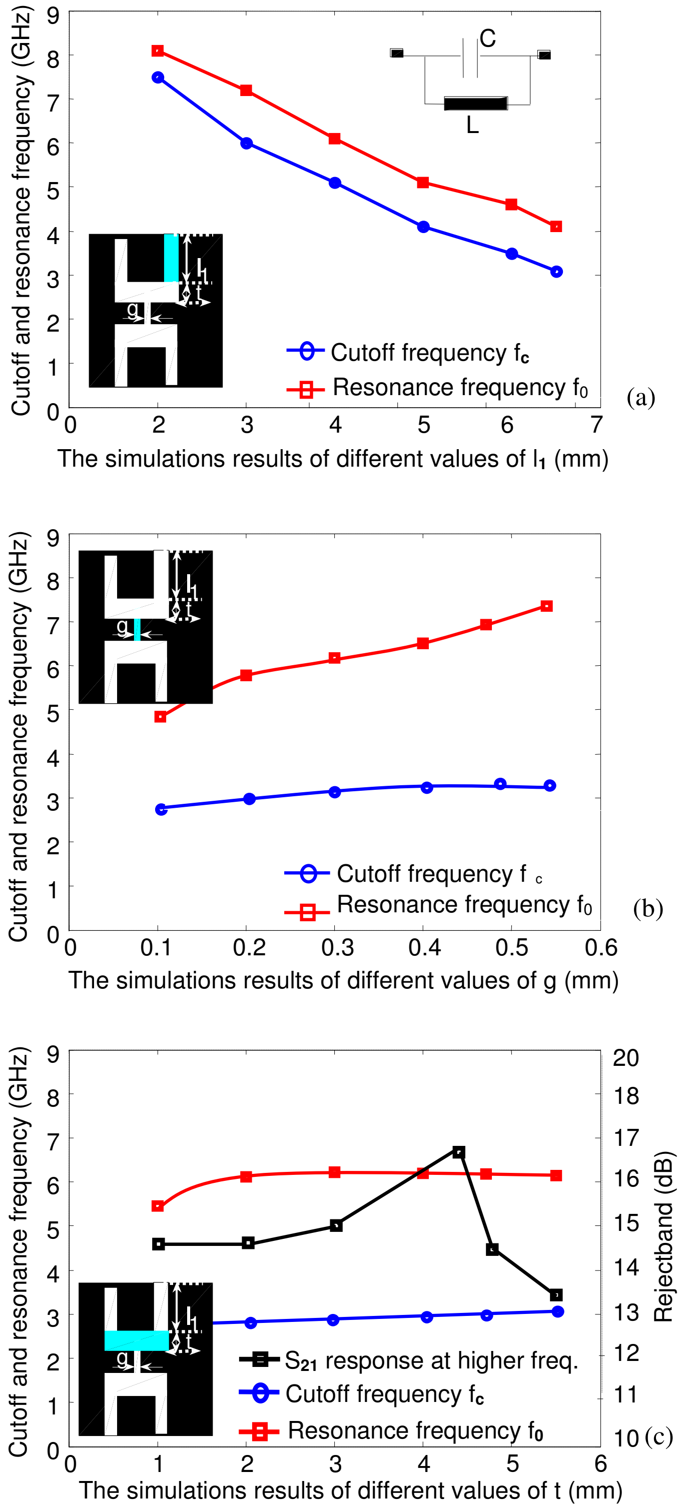

Fig. 4. Cutoff and resonance frequencies of the H-DGS. (a) Effect of variation of l_1, (b) Effect of variation of g and (c) Effect of variation of t.

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V. DESIGN AND MEASUREMENT OF THE IMPROVED LOWPASS FILTER

The optimized DGS has been used to design a LPF, which was fabricated on a ($30\times20mm^2$) substrate with a relative dielectric constant (εr) of 3.38 and a thickness (h) of 0.813 mm. Figs. 5(a) and 5(b) show respectively the layout and the equivalent circuit of the improved LPF. The extracted values C and L are respectively equal to 1.1pF and 2.9nH. The Photography of the fabricated filter is shown in Fig. 6(a). The measurements were carried out on an HP8719D network analyzer and are shown in Fig. 6(b). It can be seen that the measured results show a good consistency with both the EM and circuit simulations. The fabricated LPF has a 3dB cutoff frequency at 2.4GHz and a suppression level of 25dB from 3.85 to 14GHz; the insertion loss in the passband is about 0.65dB. Therefore, we have demonstrated that the proposed coupled DGS LPF is very favourable than the designed LPFs in [1]-[2].

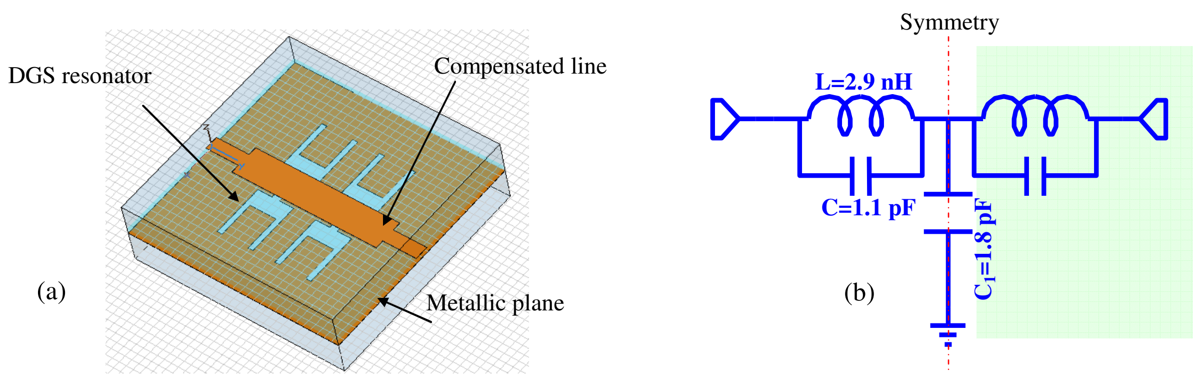

Fig. 5. (a) Three dimensional view of the LPF layout. (b) Its equivalent circuit.

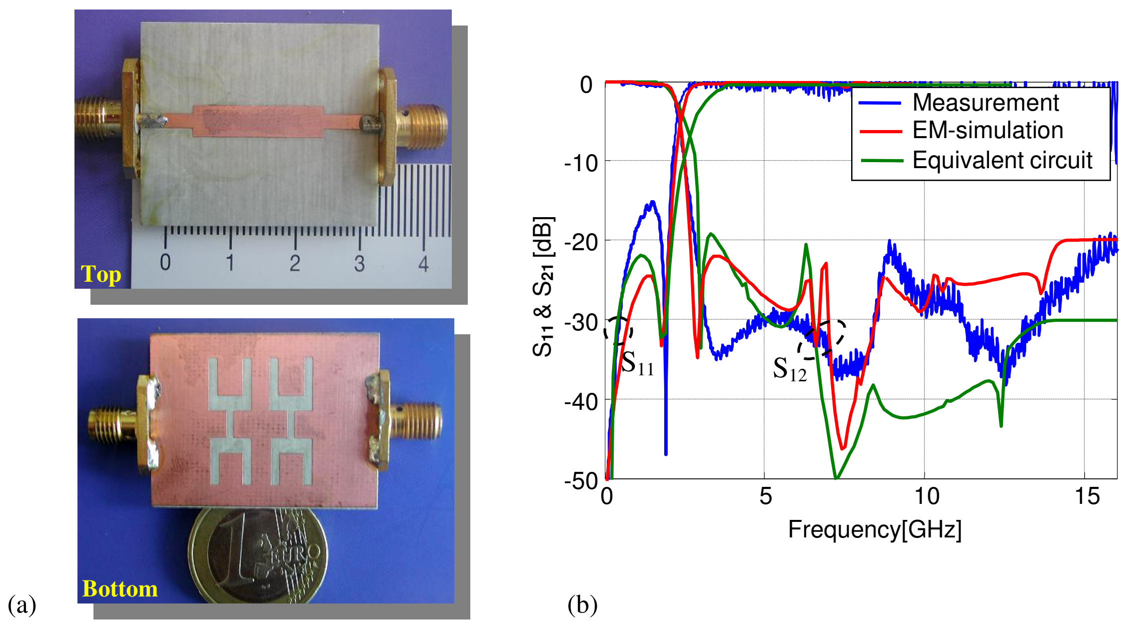

Fig. 6. (a) The photography of the proposed H-DGS LPF, (b) Comparison between the measurement, equivalent circuit simulation and EM simulation results.

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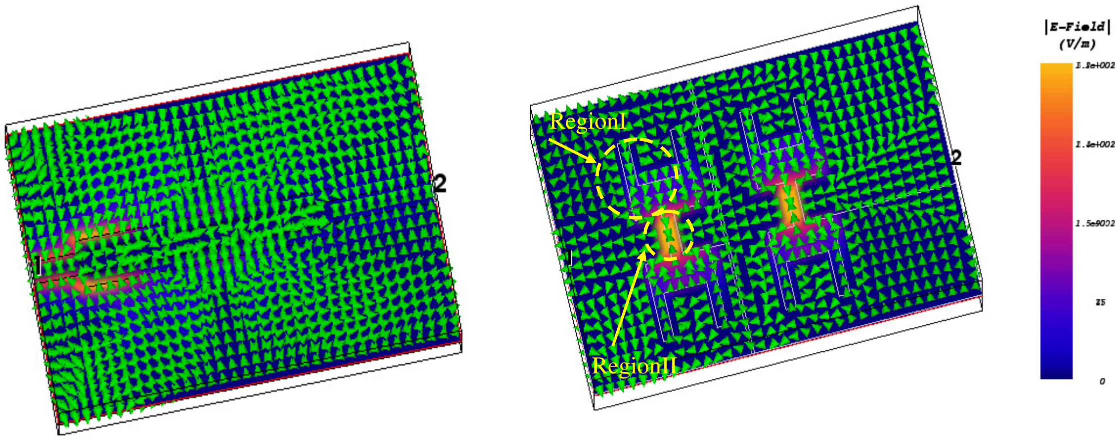
VI. DISTRIBUTION OF THE ELECTRIC FIELD AT THE PASSBAND AND THE STOPBAND

The objective of this short investigation is to verify the dependence of the equivalent circuit elements (capacitance and inductance) on the surface as the distribution electromagnetic field. The simulation results are shown in Fig. 7(a) and Fig. 7(b). The microstrip structure is divided into two regions. In the region I, the electric field is highly concentrated in the gap, hence any change in dimensions of the gap affects the effective capacitance of the structure. In the region II, the electric field is nearly vanishes. It means that the length of the arms (l_1) does not affect the effective capacitance of the filter structure. On the other hand, the current is distributed throughout the whole structure. Therefore any change in the length of the H-arm strongly affects the magnetic field distribution and hence the surface current, which in turn leads to a change in the effective inductance of the structure. Therefore, those regions I and II correspond respectively to a capacitance and an inductance. The full structure corresponds then to an *LC*-resonator.



(a) (b)

Fig. 7. Electric field distribution at (a) f_0 = 3GHz, (b) at f = 0.5GHz.

VII. CHARACTERISTICS OF THE DGS WITH GAP (S)

In order to design and to realize the band pass filters, the complicated method are often used. The idea in this work is that instead to use these conventional methods, a new transformation based on a small opening (or gap) will be used (see Fig. 7). The gap is placed on the top layer (discontinuity on microstrip feed) and plays a role of π-inverter; therefore, the signal will be suppressed in both lower and higher frequency domain, which leads to a property of the BPF. The H-DGS slot is modeled by a complex LC resonant circuit [6]. The effect of the gap is equivalent to a π-network consisting of shunt (C_p = 0.0047 pF) and series (C_g = 0.013 pF) capacitances [10]-[11]. They can be calculated with the following expressions:

$$C_p = 0.5C_e \tag{11}$$

$$C_g = 0.5C_0 - 0.25C_e \tag{12}$$

$$\frac{C_0}{w} = \left[\frac{\varepsilon_r}{9.6}\right]^{0.8} \left[\frac{s}{w}\right]^{m_0} \exp(k_0) \quad \text{(pF/m)} \tag{13}$$

$$\frac{C_e}{w} = 12\left[\frac{\varepsilon_r}{9.6}\right]^{0.9} \left[\frac{s}{w}\right]^{m_e} \exp(k_e) \ \text{(pF/m)} \tag{14}$$

$$\left.\begin{array}{l} m_0 = t\ [0.62\log(t) - 0.38] \\ k_0 = 4.26 - 1.45\log(t) \end{array}\right\} \ for \quad 0.1 \le \mathrm{t} = \frac{\mathrm{s}}{\mathrm{w}} \le 1.0 \tag{15}$$

$$\left.\begin{array}{l} m_e = \left[\dfrac{1.56}{t^{0.16}}\right] - 1 \\ k_e = \left[1.97 - \dfrac{0.03}{t}\right] \end{array}\right\} \ for \quad 0.3 \le \mathrm{t} \le 1.0 \tag{16}$$

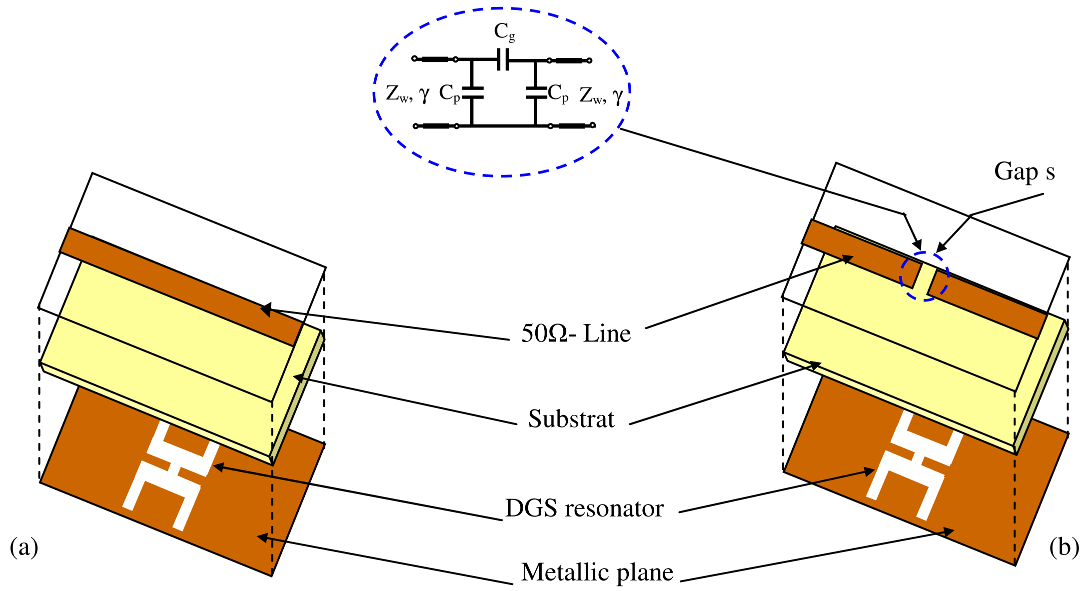


Fig. 8. Three-dimensional view of the DGS cell. (a) The conventional DGS. (b) The new H-DGS with gap.

VIII. DESIGN OF COMPACT BAND PASS FILTER USING J-INVERTER AND TWO COUPLED H-DGS

Figs. 8(a) and 8(b) show respectively the 3-D schematic view of the proposed compact BPF and its corresponding equivalent circuit. The proposed BPF has a discontinuity in the feed microstrip line as compared to a continuous feed line of the LPF. In this case, the transmission characteristics are inverted, causing the structure to act as a BPF with passband between fc and 2fc. The EM simulation results of this bandpass structure are shown in Fig. 9.

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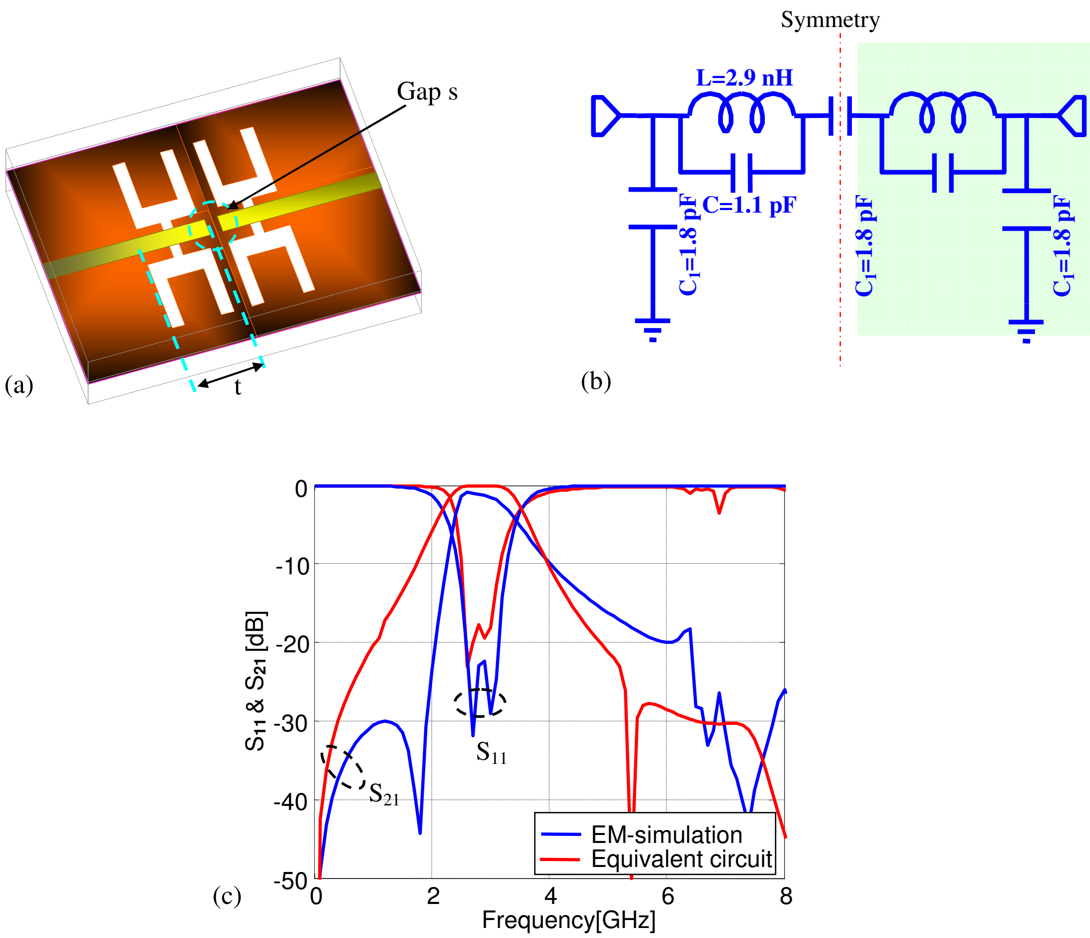

Fig. 9. (a) Schematic view of **H**-DGS BPF, (b) Its equivalent circuit, (c) Comparison between EM and circuit simulations results of the H- DGS BPF

IX. THE THEORY OF COUPLING MATRIX METHOD

In order to realize a coupling matrix which conforms to a chosen topology, it is necessary to give first the specifications of the filter. The desired parameters will be then extracted by using an optimization-based scheme [10]. The coupling coefficient and quality factor curves [10] are then used to realize the obtained coupling coefficients. In our case, the second order filter is with a bandwidth BW = 700MHz, return loss RL = 20dB, and the centre frequency f_0 = 3GHz. The obtained coupling matrix from the optimization scheme is:

$$m = \begin{bmatrix} 0 & 1.021 \\ 1.021 & 0 \end{bmatrix}, \tag{17}$$

The external quality factors are $q_1 = q_2 = 0.772$. To realize the normalized coupling matrix and quality factors, we use the required fractional bandwidth $FBW = BW/f_0$, the actual (denormalized) coupling matrix becomes

$$M = \begin{bmatrix} 0 & 0.238 \\ 0.238 & 0 \end{bmatrix}, \tag{18}$$

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and $Q_1 = Q_2 = Q = 3.24$ where $M = FBW \times m$, and $Q = q/FBW$.

The M-coupling coefficients and Q-quality factor will be inserted in the experimental curve [10-17] in order to get the optimal distance between the DGS resonators and the shifting distance d between the feed line and DGS resonator. The unknowns distances s and t are respectively equals to 700 µm and 7 mm (see Fig. 8(a)).

X. FABRICATION, MEASUREMENT AND DISCUSSION

The optimized DGS and the microstrip line gap width g were used to transform a LPF to a BPF. They were fabricated on a $(0.55\lambda_g \times 0.46\lambda_g)$ substrate with $\lambda_g = 0.054$m. The Photography of the filter and its simulated and mesaured results are shown respectively in Fig. 10(a) and Fig. 10(b). Measurements were carried out on an HP8719D network analyzer. It can be seen from Fig. 10(b) that the measurement results show a good agreement with both EM and equivalent circuit simulations. The combination of the H-DGS and the feed microstrip line with the gap suppresses the transmitted signal at lower frequencies and at the second harmonic reasonably well as shown in Fig. 10(b). The new structure has the same dimensions as shown in Fig. 6(a). The measured insertion and return losses are less than respectively to -0.55dB and -25dB at the passband of H-DGS BPF with a center frequency of 3 GHz. The second harmonic appeared at 6 GHz which is approximately equal to $2f_0$.

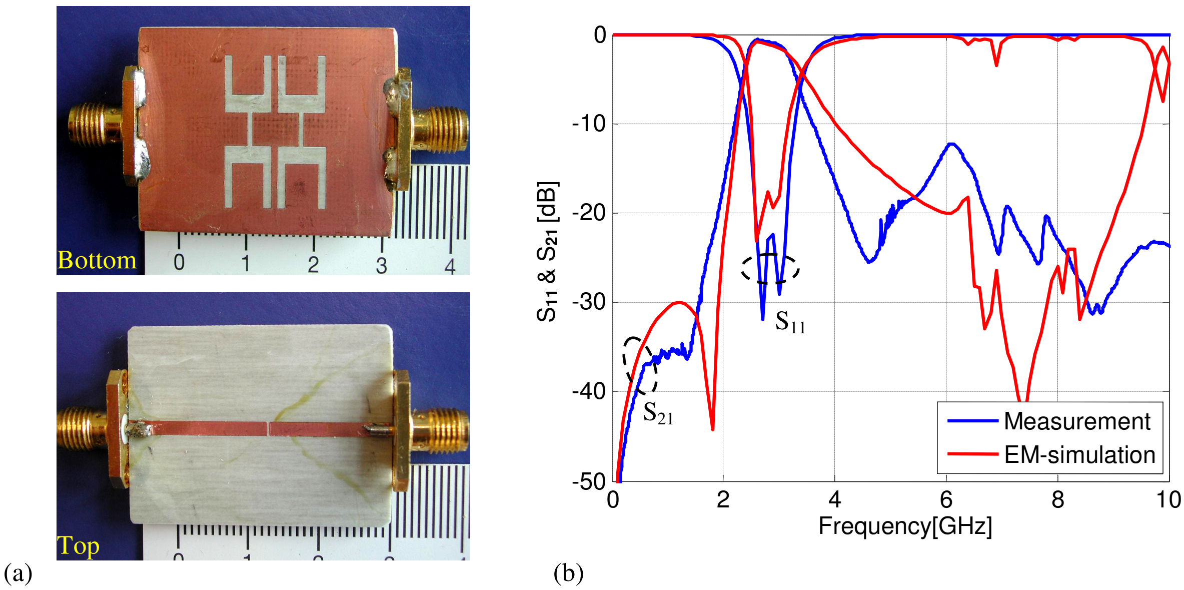

Fig. 10. (a) The photography of the proposed H-DGS BPF, (b) Comparison between the measured and simulated results.

In addition to LPF, the gap g caused a BPF characteristics, the first harmonic response has a center frequency (f_0) equal to 3GHz. The unwished second harmonics is supressed and the stop band characteristics of the fabricated H-DGS BPF is better than -15 up to 10 GHz. The experimental measurement show a very good agreement with the simulation results.

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XI. CONCLUSION

A new concept of a compact LPF using two H-DGSs has been proposed. This lowpass filter is not only of compact size, but also offers control of the cutoff frequency and transmission zero by means of the H-arm structure. To verify the performance, the filter was fabricated, simulated and measured. The measurements show a good consistency with the simulations. The LPF has a low insertion loss in the passband and a very high rejection in the stopband from 2.4 to 14GHz.

In order to change the H-DGS LPF to H-DGS BPF, a novel LPF-BPF-transformation method using DGS resonator and gap s as J-inverter has been presented. To demonstrate its potential, H- bandpass filters with suppressed spurious bands have been designed and fabricated. Two transmission zeroes can be positioned above and below the passband which give good attenuation characteristics. The design example in a microstrip DGS form provides a good agreement between the measured and simulated frequency response of the filter and verifies thereby the proposed method. Therefore, it is expected that the proposed structures with its compactness, simplicity and large stopband characteristics will be a strong candidate for applications in various integrated microwave circuits as well as other types of filters.

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