A CAD Model of Triple Bandpass Filter Implemented with Mushroom Structure

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ABSTRACT: This article demonstrates the design of planar triple-band filter using mushroom resonator for significant size reduction of the order 80%. Here, mushroom type resonator is used to realize a triple passband filter using the coupling matrix approach. The electric and magnetic couplings between the mushroom resonators are achieved by proper positioning of corner via which have been analyzed and presented. Finally, the frequency transformation has been applied to design a highly selective 12-pole triple-passband cascaded-quartets filter. This symmetric triple-band filter is operative at center frequencies 3.41/3.56/3.72 GHz with 100 MHz bandwidth of each band. It is observed that the measured results are in good agreement with the simulated results. A significant size reduction has been achieved while realizing triple-passband filter, which is as small as 50 mm × 25 mm × 1.27 mm. Int J RF and Microwave CAE 24:421–428, 2014.

Keywords: triple bandpass filters; mushroom structure; electric and magnetic coupling; planar filters; coupling matrix synthesis

I. INTRODUCTION

The present modern communication system requires multi-band operation. This initiates the need for microwave devices capable of working on multiple frequency bands. In recent era, technology is more directed toward miniaturization and energy efficient systems. To minimize the size and cost of the circuit, there is a trend to design a single circuit that is capable of producing the different designated bands. Beside the multiband operation, high performance and compactness are desired features for multiband filters. Different techniques have been reported to provide passive component size reduction, among which application of the metamaterials is used widely [1]. There has been growing interest in research and development of the devices (antennas, filters, diplexer, power divider, etc) using the metamaterial concept and it has generated a great attention from the scientific and engineering community working in RF and microwave field [1]. In previous research of metamaterial filters, Martin and coworkers have demonstrated that split ring resonators (SRRs), complementary split ring resonators (CSRRs), and spirals can be applied to the design of compact narrow bandpass filters and diplexers [2]. However, they have moderate selectivity performance attributed to the smooth transmission band edges and low out-of-band rejection levels above the pass band. The effective area reduction between metamaterial filters and microstrip open loop filters resonators has been compared by Martin and coworkers [3]. It has been shown that significant reduction in area takes place. Recently, a Sievenpiper mushroom structure is used to design a bandpass filter [4, 5]. The mushroom structure consists of a patch and via. These resonators are much smaller in size compared to the conventional Right-Handed (RH) resonators and have significant advantages in building filters with small size [4]. Single bandpass filter based on mushroom structure with corner via has been presented in [6]. Similar structure is used in this article to implement triple-band filter.

To generate a multiband filter, it is still a real challenge to synthesize the coupling coefficients and the external quality factors accurately for multiple bands. The frequency mapping is a systematic procedure for designing symmetric and asymmetric multiple-band filters [7]. Frequency transformation for dual and triple passband filter is presented in [8, 9]. Similar transformation for

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triband filter combined with matrix rotations has been used [9–13]. The main purpose of this article is to apply the triple-band filter theory based on coupling matrix approach to design compact triband filter using mushroom resonators with corner vias. The paper is organized as follows: Section II presents the design and synthesis of triple passband filter. In Section III, circuit simulations are performed to incorporate the quality factor-Q of resonators in the filter synthesis. In Section IV, filter implementation is discussed along with Eigen mode analysis of mushroom resonator. The identification of the electric or magnetic coupling of the coupled mushroom structures are presented and introduced for the first time in this article. In Section V, filter simulations and measurements results are discussed and compared. The reduction in the area of proposed filter is also compared with open loop resonators based triband filter [12]. Conclusions are drawn in Section VI.

II. DESIGN AND SYNTHESIS FOR TRIPLE-BAND FILTER

The transfer and reflection functions for any two port lossless filter network consisting of a series of N-coupled resonators can be expressed as [10]:

$$S_{21}(s)=\frac{P_N(s)}{\varepsilon E_N(s)},\quad S_{11}(s)=\frac{F_N(s)}{E_N(s)} \tag{1}$$

where $s=j\omega$ and ω is the real frequency variable. For a chebyshev filtering function, ε is a constant normalizing S_{21} to the equiripple level at $s=\pm j$ as follows:

$$\varepsilon=\frac{1}{\sqrt{10^{\frac{RL}{10}}-1}}\frac{P_N(s)}{E_N(s)}\bigg|_{n=j} \tag{2}$$

where RL is the prescribed return loss level in dB and it is assumed that all the polynomials have been normalized such that their highest degree coefficients are unity. Once $P_N(s)$ and $F_N(s)$ are defined, the unitary condition of S parameters require that

$$E_N(s)E_N^*(-s)=\frac{P_N(s)P_N^*(-s)}{\varepsilon^2}+F_N(s)F_N^*(-s) \tag{3}$$

The poles can be obtained by evaluating the roots of the above expression and by selecting those with negative real part. Again, here $E_N(s)$is generated from the poles by imposing equal to 1 the highest degree coefficient. The poles and zeros of the final multiple bands filtering function can be generated using the following frequency transformations [9, 12]. Figure 1 shows the frequency response of the filter in three different frequency domains as discussed in [9, 12].

For the outer bands,

$$s=\left(\frac{s'}{c_1}+\frac{c_1'}{s'}\right)\quad\text{for}\quad \Omega'>\Omega_0',\; s=-\left(\frac{s'}{c_1}+\frac{c_1'}{s'}\right)\quad\text{for}\quad \Omega'<-\Omega_0' \tag{4}$$

For the inner band,

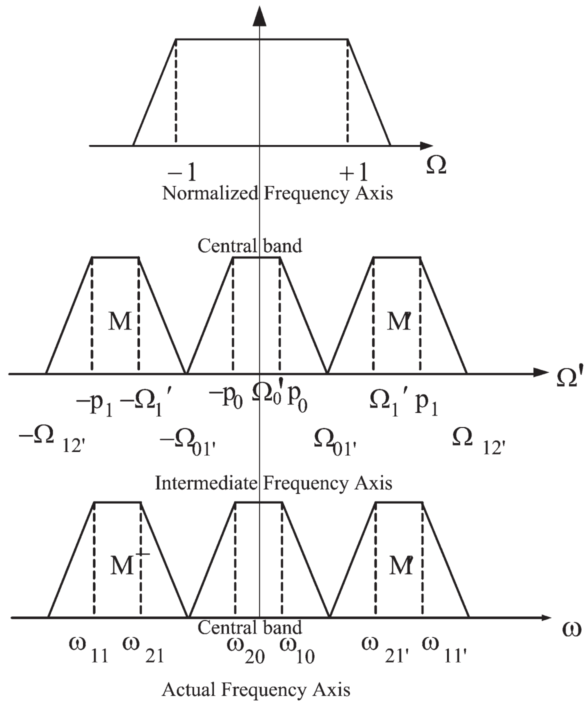

Figure 1 Frequency response of the filter in three different frequency domains, Ω is the normalized frequency for a single-passband low-pass prototype and Ω' is the normalized frequency for a triple-passband filter. The ω domain is the actual frequency domain where the filter operates [9, 12].

$$s=\left(\frac{s'}{c_0}\right)\text{for}\quad -P_0<\Omega'<P_0\quad \text{where}\, s=j\Omega\quad\text{and}\quad s'=j\Omega'. \tag{5}$$

For outer bands, 1 and -1 in the Ω-domain are transformed to p_1 and Ω_1' in the Ω'-domain for $\Omega'>\Omega_{12}'$, respectively, and 1 and -1 in the Ω-domain are transformed to $-p_1$ and $-\Omega_1'$ in the Ω'-domain for $\Omega'<-\Omega_{12}'$. Using the transformations, c_1, c_2, c_0 edge frequencies of the filter in the Ω'-domain are given as:

For outer bands: $c_1=p_1-\Omega_1'$, $c_1'=\frac{p_1\Omega_1'}{p_1-\Omega_1'}$ and for central band

$$c_0=p_0. \tag{6}$$

The transformation used for Ω'-domain to ω-domain for the outer bands are given as:

$$s=\left(\frac{s'}{d_1}+\frac{d_1'}{s'}\right),\quad d_1=\frac{\omega_{11'}-\omega_{11}}{p_1},\quad d_1'=\frac{\omega_{11'}\omega_{11}}{\omega_{11'}-\omega_{11}}p_1$$

and for central passband

$$s=\frac{s'}{d_0}. \tag{7}$$

Once the transformations are completed in all the bands, one evaluates the poles and zeros of the filtering function after each transformation. The unified filtering function is created by combining all the transformed poles and zeros [12].

A. Synthesis of Triple-Bandpass Filter

To design a 12-pole symmetric triple bandpass filter of center frequencies 3.41/3.56/3.72 GHz having an overall bandwidth of 400 MHz, a 4-pole low pass prototype for single band having transmission zeros at $s = \pm j\ 2.0$ with return loss (RL) 20 dB has been chosen. In Figure 1 for $p_1 = 1$, $p_0 = 0.20$ and $\Omega_1' = 0.60$, the unknown coefficients are calculated using Eq. (6), $c_1 = 0.40$, $c_1' = 1.5$ and $c_0 = 0.2$. Using the frequency transformation given in [12], we can find the poles and zeros of the triple-passband filter in the Ω' domain as shown in Eq. (8). In the case that some of the transmission zeros are redundant for satisfying the frequency selectivity, then those overlapped transmission zeros are removed and other remaining poles and zeros are rearranged using trial and error method as discussed in [12]. The new obtained poles and zeros in the Ω' domains are given in Eq. (9). Figure 2 shows the theoretical response of S_{11} and S_{21} in Ω' domain of the triple-passband filter using MATLAB. From these poles and zeros, the coefficients of transfer and reflection polynomials are generated as discussed in [10] and shown below in Eq. (10).

$$\begin{aligned} \text{Poles} &= [\pm\ j1.5543,\ \ \pm j\ 0.4943,\ \ \pm j0.3890,\ \ \pm j0.3890] \\ \text{Zeros} &= [\pm j\ 0.9845,\ \ \pm j0.8683,\ \ \pm j0.7140,\ \ \pm j0.6298, \\ &\qquad \pm j0.1867,\ \ \pm j0.0812] \end{aligned} \tag{8}$$

$$\begin{aligned} \text{Poles}_{\text{new}} &= [\pm\ j1.2540,\ \ \pm j\ 0.4943,\ \ \pm j0\ \ \pm j0.40] \\ \text{Zeros}_{\text{new}} &= [\pm j\ 0.9845, \pm j0.8683, \pm j0.7140, \pm j0.6298, \\ &\qquad \pm j0.1867, \pm j0.0812] \end{aligned} \tag{9}$$

$$P_N = [0.0135\ 0\ \ 0.2396\ 0\ 1.4183\ 0\ 2.9628\ 0\ 1], \tag{10}$$

$$\begin{aligned} F_N = [&3.4e{-5}\ 0\ 0.0064\ 0\ 0.1902\ 0\ 1.1149\ 0\ 2.6042\ 0 \\ &2.6777\ 0\ 1], \end{aligned} \tag{11}$$

$$\begin{aligned} E_N = [&0.0004\ 0.0038\ 0.0261\ 0.1135\ 0.4017\ 0.8593 \\ &1.9697\ 2.5020\ 3.9888\ 3.4044\ 1.2111\ 1]. \end{aligned} \tag{12}$$

From the above polynomials, the coupling matrix can be determined in the folded canonical form with the procedure described in [10].To achieve a higher selectivity and simpler tuning of triple band filter, a canonical prototype is then transformed into a cascaded-quartets (CQ) topology using the matrix rotation technique [11] and the obtained final coupling matrix is shown in Table I. The significant advantage of CQ filters is simpler tunability because the effect of each cross-coupling is independent. The coupling and routing diagram of triple bandpass CQ filter is shown in Figure 3.

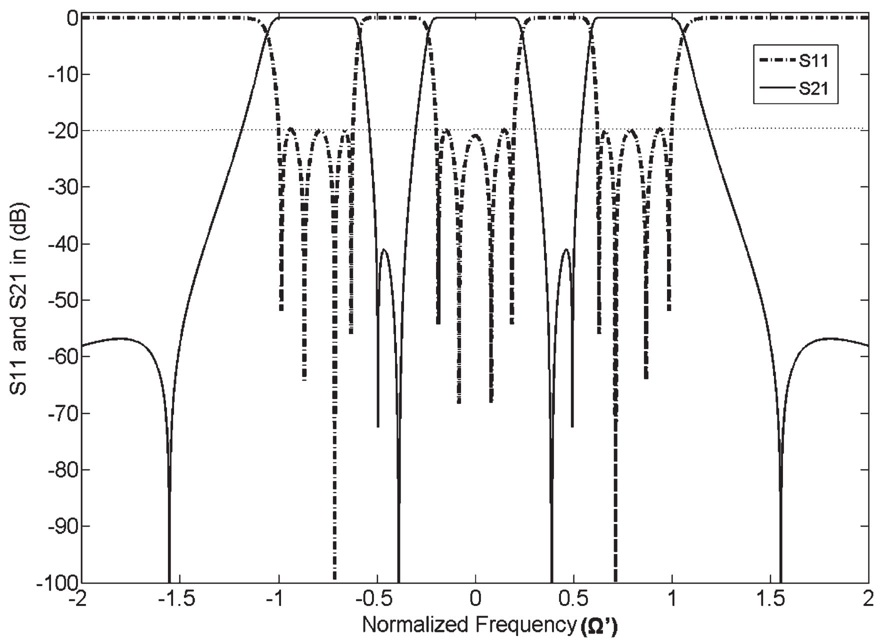

Figure 2 Theoretical response of 12-pole triple bandpass filters from the coupling matrix [transmission zero at $s = \pm j2.0$, $c_1 = 0.40$, $c_1' = 1.5$ and $c0 = 0.2$, RL 20 dB].

TABLE I Coupling Matrix for a 12-Pole Symmetric Triple-Band Filter

	S	1	2	3	4	5	6	7	8	9	10	11	12	L
S	0	0.8576	0	0	0	0	0	0	0	0	0	0	0	0
1	0.8576	0	0.7801	0	0	0	0	0	0	0	0	0	0	0
2	0	0.7801	0	0.5738	0	0	0	0	0	0	0	0	0	0
3	0	0	0.5738	0	0.4987	0	0	0	0	0	0	0	0	0
4	0	0	0	0.4987	0	0.4695	0	−0.4016	0	0	0	0	0	0
5	0	0	0	0	0.4695	0	0.3449	0	0	0	0	0	0	0
6	0	0	0	0	0	0.3449	0	0.1516	0	0	0	0	0	0
7	0	0	0	0	−0.4016	0	0.1516	0	0.6727	0	0	0	0	0
8	0	0	0	0	0	0	0	0.6727	0	0.2166	0	−0.4253	0	0
9	0	0	0	0	0	0	0	0	0.2166	0	0.2548	0	0	0
10	0	0	0	0	0	0	0	0	0	0.2548	0	0.3851	0	0
11	0	0	0	0	0	0	0	0	−0.4253	0	0.3851	0	0.7801	0
12	0	0	0	0	0	0	0	0	0	0	0	0.7801	0	0.8576
L	0	0	0	0	0	0	0	0	0	0	0	0	0.8576	0

III. CIRCUIT SIMULATIONS OF TRIPLE BANDPASS FILTER

It was observed that frequency transformation thus does not account for the losses due to the finite Q factor of the resonators in filter synthesis. To incorporate the Q factor of resonator in filter synthesis, circuit simulations have been performed in advanced design system (ADS) to synthesize the final coupling matrix as shown in Table I. The finite $Q = 320$ is calculated by simulating resonator patch under eigen mode analysis in high frequency structure simulator (HFSS). The design specifications for the 12-pole triple-band pass filter are as follows: center frequencies 3.41/3.56/3.72GHz with 400 MHz overall bandwidth and each band is having 100 MHz bandwidth. The results S_{11}, S_{21}, and Group delay are shown in Figure 4. The obtained insertion loss is 2.61/2.33/2.62 dB and return loss is 25 dB of all the three bands.

IV. TRIPLE-BAND FILTER IMPLEMENTATION

The coupling matrix given in Table I can be realized for many types of filter structures. In this article, we use a compact mushroom resonator for the realization of the filter which is proposed by Sievenpiper et al. [5].

A. Eigen Mode Analysis of Mushroom Structure

The resonance frequencies of the mushroom resonator are quite dependent on the substrate thickness and permittivity of the material. The center frequency of the resonator also depends on the change in the via position. In the case of mushroom structure with via at lower edge, the bandgap occurs at lower frequencies as compared to the resonator with via at center position [14]. Instead of placing the via at lower edge, we shift it to corner position, to decrease the resonance frequency. So we can conclude that a change in the via position to corner affects the resonance shift toward lower frequencies and reduces the size of the structure [6]. It is verified by performing eigen mode analysis in HFSS. The mushroom resonator size is 4.0 mm × 4.0 mm with via of diameter = 0.3 mm fabricated on RT duroid 6010 substrate($\varepsilon_r = 10.2$, thickness = 1.27 mm, tan $\delta = 0.0023$). It is observed that if we shift the via position from center to corner and keep the overall dimensions the same then the resonance frequency shift by around 0.8 GHz as mentioned in Table II.

B. Identification of Magnetic and Electric Coupling between Resonators

In the design of cross-coupled bandpass filters, it is important to identify the type of coupling between the resonators. The nature of the coupling is directly related to the sign of the corresponding coefficients of the coupling matrix given in Table I. The magnitude of the coupling coefficient can be extracted from the two resonance frequencies of the pair of coupled resonators, (Case: 1) If the

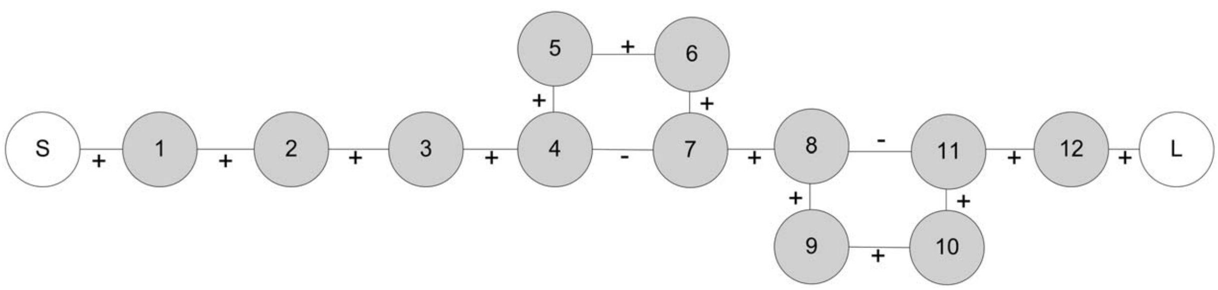

Figure 3 Coupling and routing diagram of 12-pole triple bandpass filter.

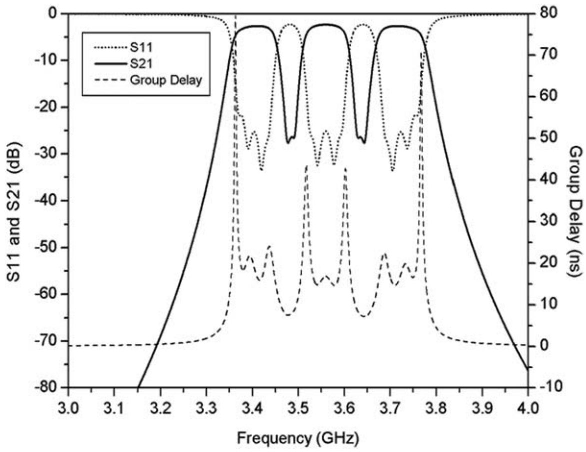

Figure 4 Frequency response of $S11/S21$ and group delay in circuit simulation.

coupling corresponding to the electric field is much stronger than that of the magnetic field, the type of coupling becomes electric and (Case: 2) if the coupling corresponding to the magnetic field is much stronger than that of the electric field, the type of coupling becomes magnetic [15]. Identification of electric and magnetic coupling can also be verified by comparing the phase responses of S_{21} of coupled structure as discussed in [13]. Figure 5 shows simulated S_{21} response of the coupled resonator structures in which ports are weakly coupled to the coupled resonator structure. The coupled mushroom resonator structure has been analyzed using Momentum. By comparing the phase responses in Figures 5a and 5b, we can observe that both are out of phase. This is the evidence that the two couplings coefficients extracted have opposite signs. So, mushroom resonators with edge via can provide both the electric and magnetic couplings between two resonators and hence it can be used for realizing both positive and negative coupling, as required by Table I.

C. Filter Layout

The physical dimension of the filter structure can be obtained using the following well known expressions given in Eq. (14). M_{ij} is the electromagnetic coupling coefficient between the resonators i and j. From Table I, $M_{S1} = M_{12L} = 0.8576$, $M_{12} = M_{1112} = 0.7801$, $M_{23} = 0.5738$, $M_{34} = 0.4987$, $M_{45} = 0.4695$, $M_{56} = 0.3449$, $M_{67} = 0.1516$, $M_{78} = 0.6727$, $M_{89} = 0.2166$, $M_{910} = 0.2548$, $M_{1011} = 0.3851$, $M_{47} = -0.4016$, $M_{811} = -0.4253$.

TABLE II Resonance Frequency of Metamaterial Mushroom Resonator

Via position (Coordinate (x,y)) mm	Center via (0,0)	Corner via (3.65,3.65)
Resonant frequency	4.36 GHz	3.56 GHz

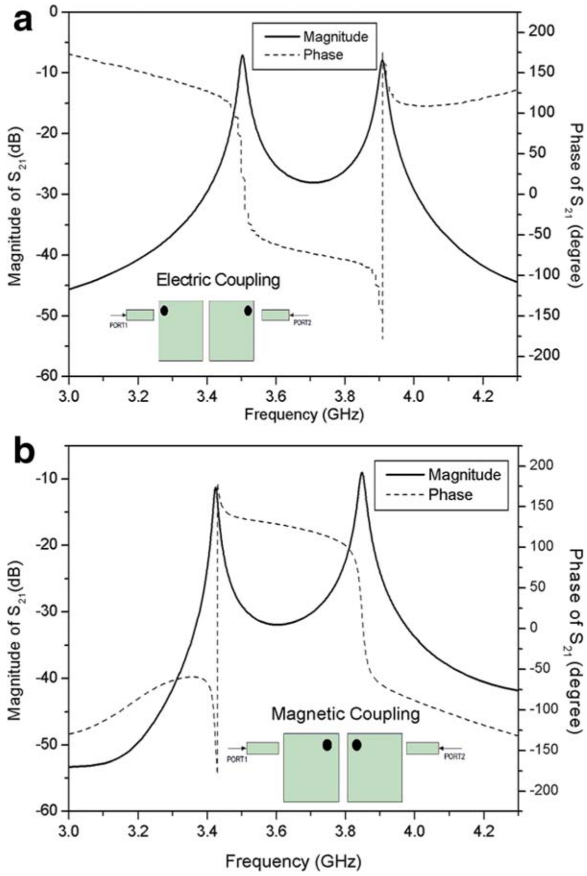

Figure 5 S_{21} response of weakly coupled mushroom structure. (a) Electric coupling. (b) Magnetic Coupling.

$$Q_{\mathrm{EXT,S}} = Q_{12,\mathrm{L}} = \frac{1}{\left(\frac{BW}{fo}\right) \times M_{\mathrm{S1}}^2} = 12.1$$

$$K_{ij} = \left(\frac{BW}{fo}\right) \times M_{ij} \tag{13}$$

The external Q's and coupling coefficient K_{ij} can be evaluated using Eq. (13),

$$K_{12} = K_{1112} = 0.087,\ K_{23} = 0.064,\ K_{34} = 0.056,\ K_{45} = 0.052,\ K_{56} = 0.038,\ K_{67} = 0.017,\ K_{78} = 0.075,\ K_{89} = 0.024,\ K_{910} = 0.028,\ K_{1011} = 0.043,\ K_{47} = -0.045, K_{811} = -0.04. \tag{14}$$

The gap g_{ij} between the resonators are calculated by using Figure 6 and after that optimized using ADS. The physical layout of the filter is shown in Figure 7. In this filter all the resonators are the same in size (4 mm × 4 mm) except resonator 1 and 12 (4 mm × 4.1 mm) and the via diameter is 0.3 mm.

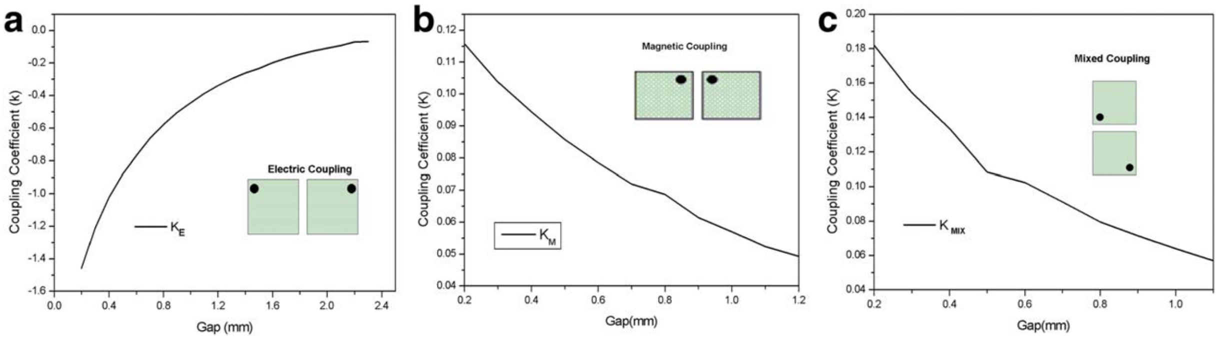

Figure 6 Mapping of the different coupling coefficients as a function of the distance (gap) between coupled mushroom resonators. (a) Electric coupling K_E. (b) Magnetic coupling K_M. (c) Mixed coupling K_{MIX}.

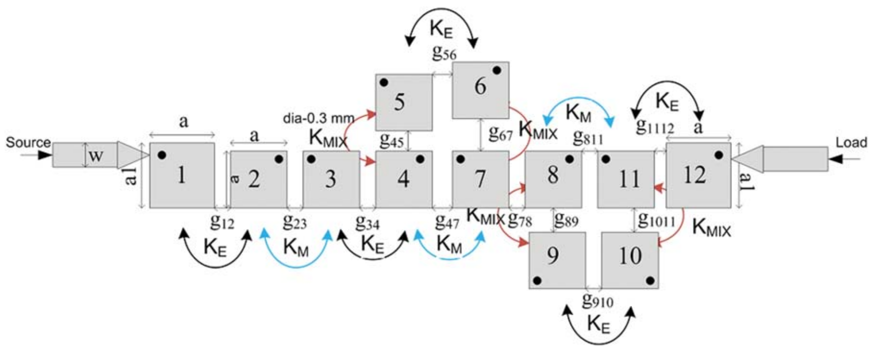

Figure 7 Triple-band filter layout [$a = 4$, $a1 = 4.1$, $W = 1.13$ $g_{12} = g_{1112} = 0.36$, $g_{23} = 0.58$, $g_{34} = 0.66$, $g_{45} = 0.83$, $g_{56} = 0.73$, $g_{67} = 1.62$, $g_{47} = 0.86$, $g_{78} = 0.5$, $g_{89} = g_{1011} = 1.43$, $g_{910} = 0.95$, $g_{811} = 0.89$, $g_{1112} = 1.58$, via diameter = 0.3, all dimensions are in mm].

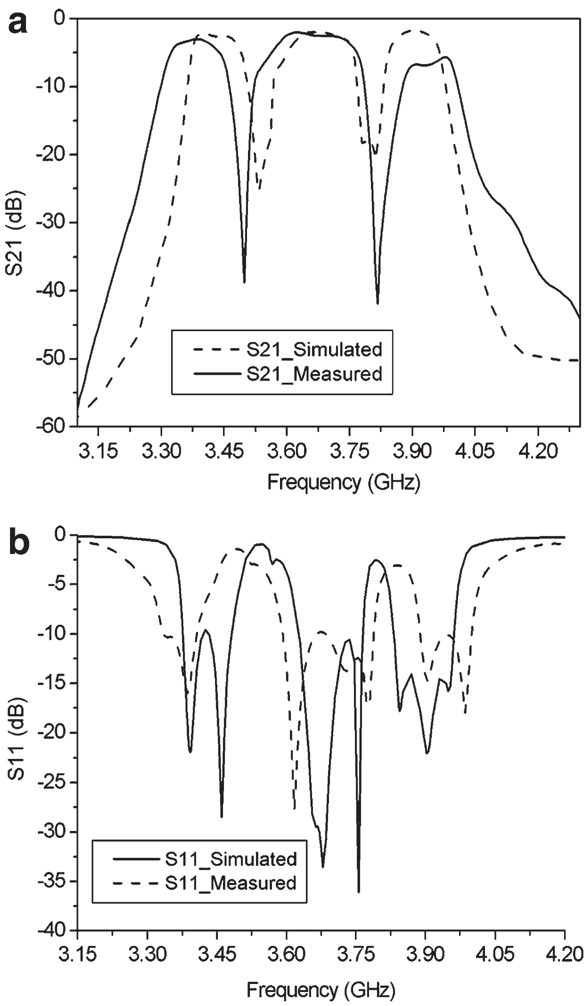

Figure 8 EM simulated/measured results of triple bandpass filter. (a) S_{21}, (b) S_{11}.

TABLE III Comparision of Circuit Simulations, EM Simulations, and Measured Results of Triple Band Pass Filter

Filter specifications	Circuit Simulation	EM Simulation	Measurements
Center frequencies (GHz)	3.41/3.56/3.72	3.40/3.66/3.90	3.37/3.66/3.96
Bandwidths (MHz)	100/100/100	130/140/160	140/170/170
Insertion loss (dB)	2.61/2.33/2.62	2.4/2.1/2.0	3.8/2.3/3.7
Return loss (dB)	−25/−25/−25	−10/−10/−14	−10/−10/−10

V. RESULTS AND DISCUSSION

A. Simulated and Measured Results

The filter has been simulated and optimized using ADS Momentum. The RT duroid 6010 substrate(ε_r = 10.2, thickness = 1.27 mm, tan $\delta = 0.0023$) has been used for filter fabrication. The simulated and measured S_{21} and S_{11} are depicted in Figures 8a and 8b. Instead of having one passband as normal filters, the filter has three passbands and two stopbands which need to be tuned simultaneously. The simulated band edge frequencies are 3.37 GHz and 3.77 GHz. The circuit simulation, EM simulation, and measured results are compared in Table III. In EM simulation, center frequencies of second and third bands are shifted by 0.1/0.2 GHz as compared to the circuit simulation. In measured results frequency selectivity and out of band performance are good. The center frequencies of measured results are shifted slightly 30 MHz/60 MHz of first and third band as compared to the EM simulated response. This can be explained by limitation of software or/and defects in various levels of the fabrication. However, apart from the frequency shift, there is good agreement between the measured and simulated results in Figure 8. This shift may be attributed due to the minute error during fabrication.

B. Comparison of Size reduction

Figure 9 shows a fabricated 12-pole triple bandpass filter. The size of the device, which is as small as 50 mm × 25 mm × 1.27 mm, is also significant. A reduction by a factor 80% has been obtained for the Mushroom resonator based filter as compared to the open-loop resonator based filter [12].

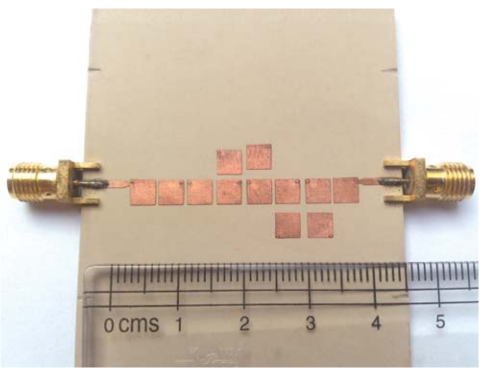

Figure 9 Fabricated picture of triple-bandpass filter using metamaterial mushroom resonators.

VI. CONCLUSION

In this article, a novel compact mushroom resonator based triple band filter with corner vias is presented by using the coupling matrix approach. The magnetic and electric character of the coupling coefficient is analyzed in detail. It has been shown that the use of resonators provides a significant reduction of the order (80%) in the active area of planar microwave filters. The resonators sizes are very small, so multiband filter can be implemented by using these resonators. Such type of compact filters can be used in satellite, military applications.

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BIOGRAPHIES

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