Novel Miniaturized Cross-coupled Trisection Bandpass Filters

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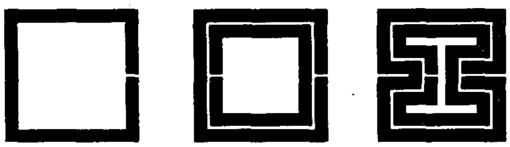
CONTENT

1. INTRODUCTION

In modern microwave communication systems, especially in satellite and mobile communications, the circuit size might be one of critical issues. These requirements are imposed to conserve the valuable frequency characteristics and enhance the performance of the systems. Obviously, the circuit size of the microstrip filter with parallel-coupled, half-wavelength resonators is too large to be used for wireless communicaion systems. Thus, size reduction as well as selectivity and passband insertion loss are important factors in the recent developments of RF filters in microwave integrated circuits (MIC's) for wireless systems. It is known that when frequency selectivity and passband loss are considered to be the important filtering properties, the optimum filters are those exhibiting ripple in both passband and stopband. Such a filter response can be realized using filters with cross couplings to give a number of alternative paths which a signal may take between nonadjacent resonators. Depending on the phases of the signals, the multipath effect may cause attenuation poles at finite frequencies or group delay flattening, or even both simultaneously. In this paper, novel cross-coupled trisection microstrip EC-SRR bandpass filters which lead to new microstrip split ring resonator bandpass filters [1,2] including the four-pole elliptic or quasi-elliptic function response filter [3] with an asymmetric frequency response filter with higher selectivity on high side of passband are introduced. As another applications of EC-SRRs, free space transmission and the first reflection measurements of a composite double negative (DNG) metamateral, also known as a left-handed material (LHM) are reported [4-6].

2.SPLITRINGRESONATOR

Square and circular loops are widely used as elements for frequency selective surfaces and reflecting arrays. These loops are symmetrical and therefore at normal incidence the scattering parameters are insensitive to polarization. However, the resonant frequency becomes a strong function of polarization when the conductor is broken at a suitable location. Two types of the EC-SRRs used in the design of the first backward mata-material [3] is sketched in Fig.1 [1]. The EC-SRR also consists of two loops while the resonators are identical and parallel to each other. This EC-SRR comprises two conductive rings printed on a thin dielectric substrate and separated by a narrow gap. Each ring has a slit, and rings are oriented in such a way that slits are on the opposite sides of line of symmetry.



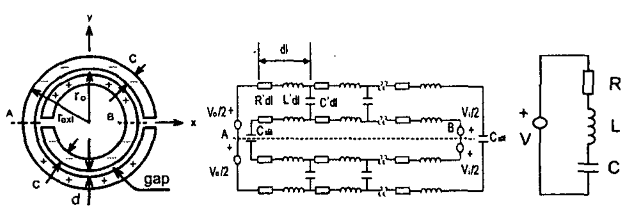
(a) (b) (c)

Fig. 1. (a) Capacitively loaded open loop resonator (OLR) (b) EC-SRR (c) novel miniaturized EC-SRR

The electrical dimensions of the SRR are much smaller than a wavelength of the impinging plane wave. The voltage induced in each SRR as two mutually coupled small loop resonators can be simply calculated from Faraday's law :

$$V = -j\omega\mu_0 AH_i \qquad (1)$$

Here, V stands for induced voltage, ω stands for radial frequency, and H_i stands for incident magnetic field, which is assumed being perpendicular to the loop, A and U_0 stand for the loop area and free-space permeability, respectively. The induced voltages are modelled as simple voltages' sources located at point A (the source V_o– outer ring) and point B (the source V_i– inner ring). The complete equivalent circuit is sketched in Fig. 2(b). The symbols R' and L' stand for distributed resistance and inductance respectively, C_{slit} stands for capacitance of the slits and C' stands for distributed capacitance of the gap between rings. From the Fig. 2(b), two main loops (inner and outer ring) which are connected across the gap via distributed capacitance C' can be identified.



(a) (b) (c)

Fig. 2. (a) The EC-SRR (b) The completely equivalent circuit (c) The simplified equivalent circuit

It is important to notice that voltage V_o is always higher than voltage V_i by the ratio of areas formed by outer and inner ring, respectively (square of the ratio of radius of outer and inner rings). Thus current essentially flows from the outer ring into inner ring across the gap. It flows through many branches formed by distributed capacitance C'. Due to this branching, current in outer ring changes with the location at the ring. It is maximal at point A, then decreases along the ring and reaches minimal value at the slit. All the currents, which flow from the outer ring into the inner ring, contribute to the net current in inner ring. Therefore, the current in an inner ring exhibits the maximum at the location of voltage source V_i (point B), then decreases along the ring and reaches the minimum at the slit. If the current which flows across the slit to the net ring current should be negligible, the whole circuit with a much simpler form is possible as shown in Fig. 2(c). It comprises a single voltage source and a serial tank circuit. The magnetic resonant frequency of parallel SRR or broadside coupled SRR (BC-SRR) proposed in [2], is lower than that of the EC-SRR introduced in this paper. This is because the mutual capacitance between the parallel broken rings is now capacitance of a conventional parallel-plate capacitor and is significantly higher than the mutual capacitance of two coplanar split rings. However if the upper half of the ring 1 is charged negatively, the positively induced charges appear in the upper half of the ring 2. The same situation happens for the EC-SRR, but not so effectively since the strips are coplanar and weakly interacted. These split rings are coupled by means of a strong distributed capacitance in the region between the rings. When a time-harmonic external (or local) magnetic field of angular frequency, ω is applied along the z axis of these structures, an electromotive force will appear around the EC-SRR. Provided that the electrical size of the EC-SRR can be considered small, a quasi static behavior is expected. With this quasi static model in mind, it is not difficult to see how the induced current lines will pass from one ring to the other through the capacitive gaps between them in the form of field displacement current lines. Therefore, the total current intensity flowing on both rings remains the same for any cross section of the structure (i.e., it is independent of the angular polar coordinate). The whole device then behaves as an LC circuit driven by an external electromotive force. The total capacitance of this LC circuit will be the series capacitance of the upper and the lower halves of the SRR and the resonance frequency ωo is given by

$$\omega_0 = \sqrt{\frac{2}{\pi\, r_0 L C_{pul}}} \quad (2)$$

where C_{pul} is the capacitance per unit length between the rings and L is the total inductance of the EC-SRR. By comparing BC-SRR, the EC-SRR particle is presenting the cross-polarization effects. The resulting equations can be summarized as follows:

$$m_z = \alpha_{zz}^{mm} B_z^{ext} - j\alpha_{yz}^{em} E_y^{ext} \quad (3)$$

$$p_y = (\alpha_{yy}^{ee} + \alpha_{yy}^{\prime ee}) E_y^{ext} + j\alpha_{yz}^{em} B_z^{ext} \quad (4)$$

$$p_x = \alpha_{xx}^{ee} E_x^{ext} \quad (5)$$

where m and p are the magnetic and electric induced dipoles, B^{ext} and E^{ext} the external fields and α the polarizabilities. The above results for the polarizabilities are confirmed by a more detailed electro- magnetic analysis [2].

3. BASIC COUPLING STRUCTURES

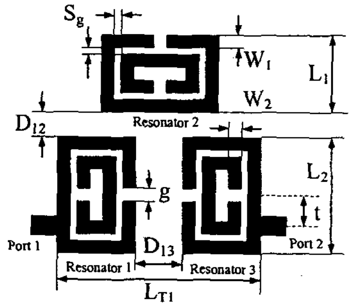
A three-pole trisection filter is not only the simplest CT filter by itself, but also the basic unit for construction of higher-degree CT filters. Cross-coupled trisection microstrip EC-SRR bandpass filter are designed and fabricated for the demonstration [7-8]. In the case of cross-coupled trisection microstip EC-SRR bandpass filters, with the aid of the lowpass prototype filter, the bandpass filters with an attenuation pole could be synthesized. The simulated attenuation pole of both EC-SRR and novel miniaturized EC-SRR should be located on Ω_a =2.2516. The lumped circuit element values of the lowpass prototype filter are found to be g_0=1, g_1=g_3=0.695, g_2=1.245, J_{12}=J_{23}=1, J_{13}=-0.457, B_1=B_3=0.185, B_2=-0.615. The cross couplings can be synthesized using the method described in [9]. The tapped-line feed is used for the loaded external Q_e though the other means of feed such as the coupled-line feed is feasible [10].

$$\omega_{0i} = \omega_0 \bullet \sqrt{1 - \frac{B_i}{\omega_0 / FBW + B_i/2}} \quad (6)$$

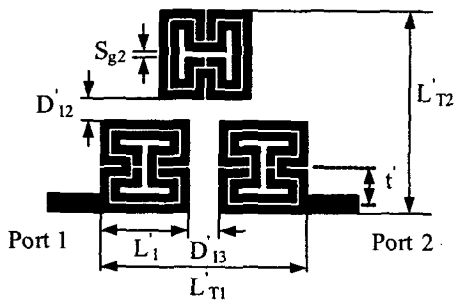
$$Q_{en} = \frac{\omega_{0n}}{\omega_0 g_{n+1}} \bullet \left(\frac{g_n}{FBW} + \frac{B_n}{2}\right) \quad (7)$$

$$M_{ij}\big|_{i\neq j} = \frac{\omega_0}{\sqrt{\omega_{0i}\omega_{0j}}} \bullet \frac{FBW \bullet J_{ij}}{\sqrt{(g_i + FBW \bullet B_i/2)\bullet(g_j + FBW \bullet B_j/2)}} \quad (8)$$

where n is the degree of the filter or the number of the resonators. The couplings between adjacent resonators are indicated by the coupling coefficients M_{12} and M_{23} and the cross coupling is denoted by M_{13}. Q_{e1} and Q_{e3} are the external quality factors denoting the input and output couplings, respectively.



(a)

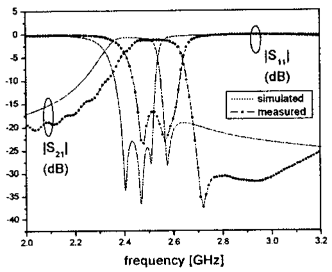


(b)

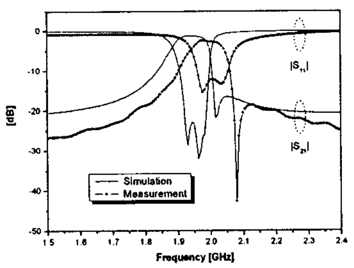
Fig. 3. Layout of cross-coupled trisection microstrip filters. (a) EC-SRR (b) novel miniaturized EC-SRR. Geometric parameters for each isolated resonator : D_{12}=0.5, D_{13}=1.21, Sg=0.2, t=2.50, g=0.4, L_1=4, L_2=8, L_{T1}=0.197λ_{g0} ; D'_{12}=0.55, D'_{13}=0.91, S_{g2}=0.5, t'=2.36, L'_1=6.11, L'_{T1}=0.222λ_{g0}, L'_{T2}=0.216λ_{g0}. All are in mm.

4. RESULTS AND MEASUREMENTS

The filters are fabricated on a dielectric substrate with a relative dielectric constant of 10.2 and a thickness of 1.524 mm. The layout and dimensions of these filters are shown in Fig. 3. The cross-coupled trisection EC-SRR bandpass filter specifications are the center frequency of 2.46 GHz, the fractional bandwidth FBW=4% and the corss-coupled trisection novel miniaturized EC-SRR bandpass filter specifications are the center frequency of 1.95 GHz, the fractional bandwidth FBW=3.077%. In layout of both Fig. 3(a) and (b), it can be tuned to the gap widths with a very little different value each other. The fabricated filters are measured on an HP 8510C network analyzer. The measured and theoretical performance for the circuits are shown in Fig. 4 and the photos of these circuits are shown in Fig. 5.

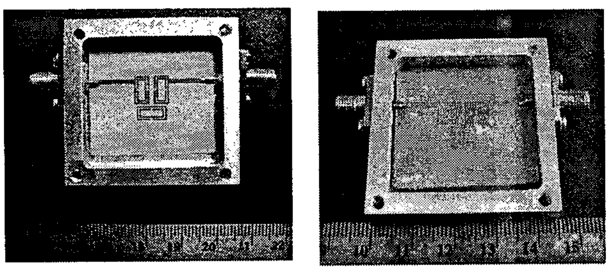


(a)



(b)

Fig. 4. Theoretical and measured performance for two filter types of Fig. 3.



(a) (b)

Fig. 5. Photographs of cross-coupled trisection microstrip filters. (a) EC-SRR (b) novel miniaturized EC-SRR.

Although the measurement results are a little shifted in the frequency responses, they are well agreed with simulation data. The passband insertion loss of both the cross-coupled trisection microstrip EC-SRR and novel miniaturized EC-SRR bandpass filters is about -1.30 dB and -2.77 dB, which is mainly resulted from the conductor loss, at the measured mid-band frequency in Fig. 4, respectively.

5. CONCLUSION

New planar cross-coupled bandpass filters using EC-SRRs have been introduced and the feasibility of realizing the cross-coupled filters such as a CT filter has been discussed. Since EC-SRR dimensions are much smaller than signal wavelength, the proposed filters are extremely compact and can be used to reject frequency parasitics in microstrip structures by simply patterning properly tuned EC-SRRs in the upper side metal. It is shown that the capability of implementing an attenuation pole, the compactness and flexibility in size, and the simplicity in both the design and fabrication make the cross-coupled trisection microstrip EC-SRR bandpass filters attractive for further development toward new applications.

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