Design of Cross Coupled Meander Folded Hairpin Resonator Filters

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***Abstract*—The paper presents design, simulation and measurement of microstrip cross-coupled filter based on meander folded hairpin resonators. For the design purposes, the coupling coefficient of the coupled resonators of different topologies is investigated. Each topology is simulated in full-wave electromagnetic simulator and the coupling coefficient is extracted. Design graphs for the coupling coefficients are presented. Based on the simulation results, a design examples is carried out. The design example is 4 pole cross coupled Chebyshev filter presenting two symmetrical attenuation poles. There is very good agreement between the simulation and measurement results.**

***Keywords*—microstrip cross coupled filter, folded resonator, coupling coefficient**

I. Introduction

Mobile communication systems have constant demand on miniaturized microstrip filters with stringent frequency requirements. This demand results in continuous research for miniaturized microstrip filters satisfying the frequency requirements for passband and stopband. Generally microwave filters are designed according to Chebyshev and generalized Chebyshev approximations. They result respectively in cascaded coupled resonator filters or cross coupled filters.

The first paper concerning halfwave resonator (Fig.1a) size reduction is introducing the hairpin resonator (Fig.1b) [1]. The miniaturization of filters passes through making halfwave resonators in different geometrical forms-triangular, square, penthagonal, [2-4]. In order to achieve more flexible filter design, the authors of [5] proposed hexagonal structure of the resonator. It inherits the features of the half wavelength resonator-position of the spur frequency, input impedance and slope parameter.

Another approach for compact resonator size is to load the main transmission line with distributed capacitors [6] forming slow-wave resonator (Fig.1c). The fundamental paper from Sagawa et al.[7], proposes miniaturized hairpin filter realizing the capacitive loading of the main resonator line with coupled microstrip lines (Fig.1d).

Slow-wave resonators and miniaturized hairpin resonators design process is rather difficult than the halfwave resonator's. One further step in miniaturization of the halfwave resonator is made in [8] (Fig.1e). Folding of the two of the arms of the resonators is proposed, but is not researched in depth.

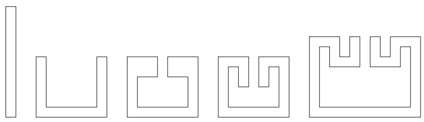

Figure 1. Types of resonators-(a) halfwave, (b) hairpin, (c) slow-wave, (d) miniaturized hairpin, (e) folded halfwave miniature

In this paper is researched in depth the resonance performance of microstrip folded halfwave resonator depending on the space and length of the folded lines for FR-4 substrate. The main coupling topologies using the folded microstrip resonators are described and the coupling coefficient is computed using full-wave electromagnetic (EM) simulator. Design graphs for the coupling coefficients are presented. Two design examples of microstrip fitlers are synthesized, simulated and measured. There is very good agreement between the theoretical and measured results.

II. Folded Miniaturized Halfwave Resonator

The detailed topology with the dimensions of the folded miniaturized resonator is shown on Fig.2.

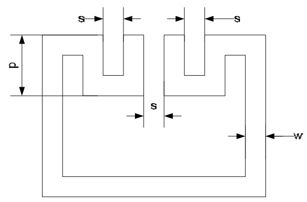

Figure 2. Topology of folded miniaturized resonator

All the simulations, design procedures and measurements in the paper are performed for dielectric substrate FR-4 with height $h = 1.5mm$, relative dielectric constant $\varepsilon_r = 4.4$ and loss tangent $tg\delta = 0.02$.

The folded miniaturized resonator width of the main line is corresponding to characteristic impedance 50Ω. For this dielectric substrate w=2.78mm. Fig.3 shows the dependence of the resonant frequency for constant length of the folded arms of the resonator and varying the space between them.

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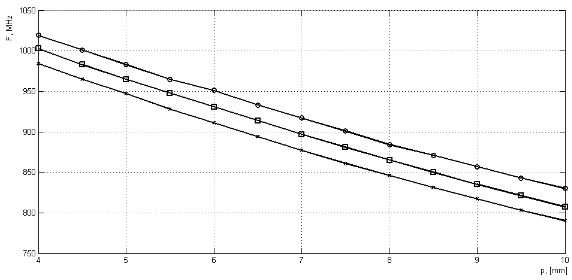


Figure 3. Resonance frequency of the folded miniaturized resonator for constant length p and s=0.5mm (circle), 1.5mm (square), 2.5mm (cross)

From Fig.3 is seen that the resonance frequency is almost linear with the increase of the arms' length *p*. When the arms are longer, the resonance frequency decreases as it is a function of the wavelength.

Fig.4 shows the dependence of the resonance frequency of the space between the folded arms for constant length *p*.

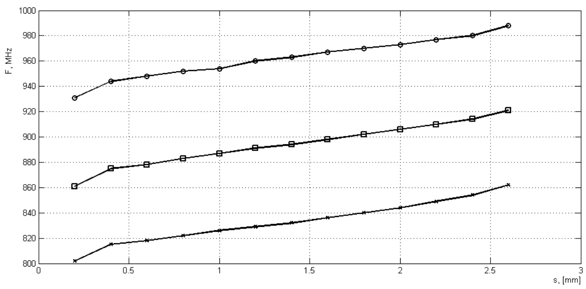


Figure 4. Resonance frequency of the folded miniaturized resonator for constant spaces s and p=5mm (circle), 7mm (square), 9mm (cross)

It is seen from Fig.4 that if the space between the folded arms is relatively small, the resonance frequency is lower due to the increased capacity between the arms. This leads to the equivalent increasing of the length of the main resonator line and lowering the resonance frequency.

III. FOLDED MINIATURIZED RESONATOR COUPLING TOPOLOGIES

The coupling mechanism is based on the fringe fields of closely placed resonators. The nature of coupling depends on the resonator configuration. It is clear that half wavelength folded resonator is symmetrical along the middle of the main transmission line. The coupling coefficient for synchronously tuned resonators can be calculated easily by the resonance frequencies of even and odd mode [3], when the coupled resonators are overcoupled:

$$k = \frac{f_e^2 - f_o^2}{f_e^2 + f_o^2} \qquad (1)$$

A full wave EM simulator based on the Method of the Moments (MoM) is used to identify the resonance frequencies. Most of the coupling structures are simulated using the symmetry in their topology for electrical and magnetic wall introducing in-between. When symmetry does not present, the whole structure is simulated. This does not constrain in any way the obtained results.

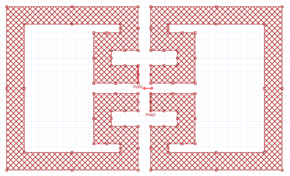


Figure 5. Topology of electrical coupled folded resonators

Even and odd mode frequencies are derived simulating the topology shown on Fig.5. The electrical field is stronger around the open ends of the resonator arms. This determines the electrical nature of the coupling. The electrical coupling coefficient is computed using full wave EM simulations. It is with negative sign and the graphical results are shown on Fig.6.

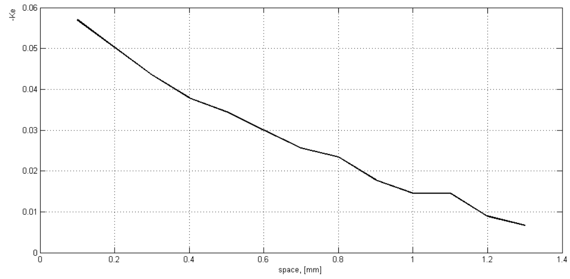


Figure 6. Coupling coefficient for electrical coupling

The coupling topology performing magnetic nature of the coupling mechanism is shown on Fig.7. The magnetic field is predominant over the electrical field around the point of symmetry of the resonator.

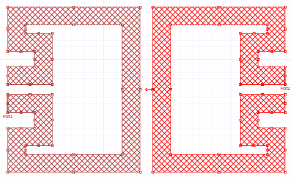


Figure 7. Topology of magnetic coupled folded resonators

The results for the magnetic coupling coefficient are shown on Fig.8.

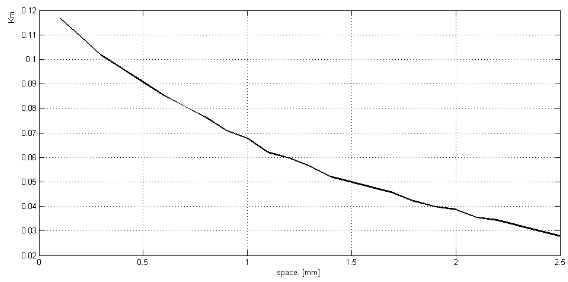


Figure 8. Coupling coefficient for magnetic coupling

The coupling scheme shown on Fig.9 is for mixed coupling. The coupling resonators are opposite situated in the plane. The currents in the coupled lines are equal in amplitude and in-phase. The value of the coupling coefficient is with positive sign.

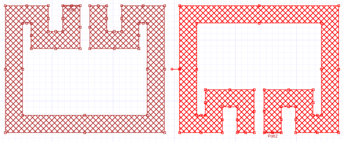


Figure 9. Topology of magnetic coupled folded resonators

The coupling coefficient for mixed coupled resonators, obtained by full-wave EM simulations, is shown on Fig.10.

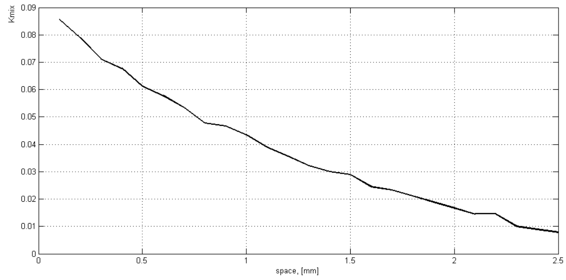


Figure 10. Coupling coefficient for mixed coupling

As it is seen from Fig.6, 8 and 10, the coupling coefficients decrease with the increase of the space between the coupled lines.

In order to determine the exact tapping position of the input/output lines, it is necessary to derive the external quality factor Q_e according to the topology shown on Fig.11.

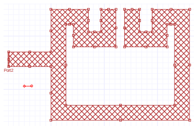


Fig.11 Tap position for derivation of the external Q factor

The mathematical expression for it is [3]:

$$Q_e = \frac{f_0}{\Delta f_{\pm\frac{\pi}{2}}} \quad (2),$$

Where f_0 is the resonance frequency of the resonator, and $\Delta f_{\pm\frac{\pi}{2}}$ is the bandwidth at which the phase of the s_{11} shifts to $\pm\pi/2, rad$ with respect to the phase at the resonance frequency. It is extracted the external quality factor and the results are shown on Fig.12.

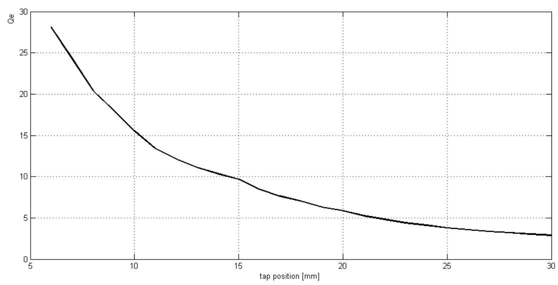


Figure 12 External quality factor versus the tap position.

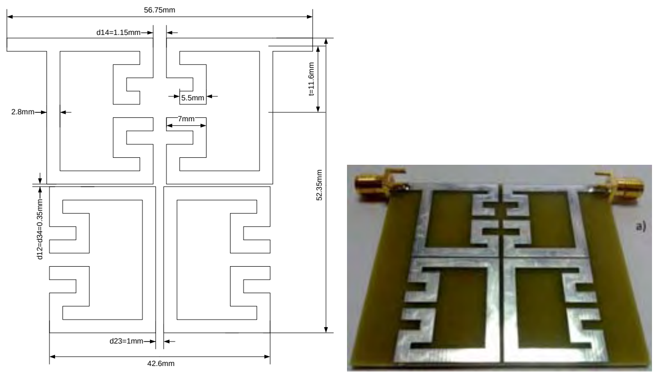
IV. DESIGN EXAMPLE

A cross-coupled microstrip filters is synthesized, simulated and measured in order to prove the design procedure. The design procedure starts with computation of the elements of the coupling matrix corresponding to the filter topology. The next step is to define the coupling structures for realizing the coupling coefficients. From the presented figures (3-10), the resonator dimensions and the distance between the resonators are found. The tapping position can be found by Fig.12.

Filter example: The center frequency of the 4-th order cross coupled Chebyshev filter is $f_0 = 900MHz$, the bandwidth is $\Delta f = 100MHz$ with maximum return loss in the passband $RL = -20dB$. The two prescribed symmetrical transmission zeros are placed at 820MHz and 970MHz.

The coupling coefficients and the external quality factor are: $M_{S1} = 0.138$, $M_{12} = M_{34} = 0.11$, $M_{23} = 0.1088$, $M_{41} = -0.04024$.

The topology with the dimensions and the picture of the designed filter are shown on Fig.16.



(a) (b)

Figure 16. Topology with dimensions (a) and photograph of 4 order cross coupled Chebyshev filter (b)

The simulation results of the designed filter are shown on Fig.17.

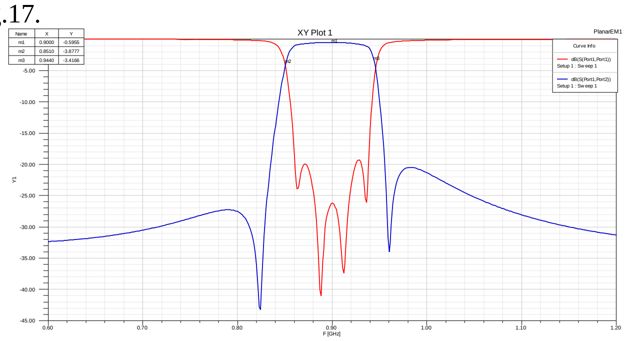


Figure 17. Simulation results of the filter. Transmission coefficient (solid), reflection coefficient (dashline)

From simulation results on Fig.17 is seen that the bandwidth of the filter is slightly less than the required 90MHz. The transmission zeros are situated exactly on prescribed frequencies. The minimum return loss value is -19.8dB.

After the fabrication process the filter frequency responses are measured with SignalHound Spectrum Analyzer SA44B with tracking generator TG44A.

The measurement results are shown on Fig.18.

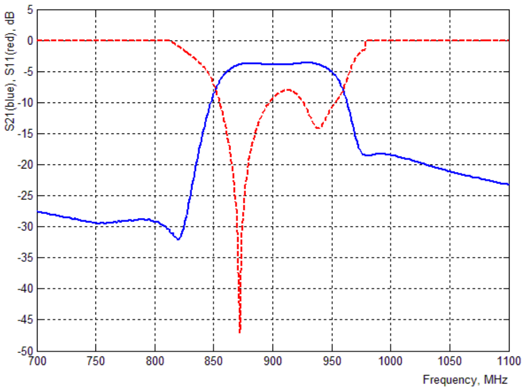


Figure 18. Measurement results of Filter 2. Transmission coefficient (solid), reflection coefficient (dashline)

From the measurement results, shown on Fig.18, is clearly seen the symmetrical transmission zeroes on 820MHz and 970MHz. They are exactly on the prescribed frequencies, but with different value of the transmission coefficient, because of the frequency dependence of the electrical coupling. The passband width is 90MHz and the insertion loss of the filter is 4dB. The losses are mainly due to the high dielectric loss in the substrate. The maximum return loss in the passband is -7dB. The size of the designed filter is $(0.3225\lambda)x(0.3138\lambda)$ and is 30% of the corresponding hairpin filter.

V. Conclusion

In this paper is researched in depth the resonance performance of microstrip folded halfwave resonator depending on the space and length of the folded lines for FR-4 substrate. The main coupling topologies using the folded microstrip resonators are described and the coupling coefficient is computed using full-wave electromagnetic simulator. Design graphs for the coupling coefficients are presented. A design example of microstrip fitler is synthesized, simulated and measured. There is very good agreement between the theory and measured results.

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