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Design Of Hairpin Bandpass Filters Using Coupled Resonator Theory

A. Christina Josephine Malathi
School of Electronics and Engineering, VIT University, Vellore - 632014, TamilNadu, India.
Email-id achristina@vit.ac.in

Abstract- In this paper we design a hairpin band pass filter using coupled resonators. It is realized by microstrip open-loop resonators by coupling a pair of non-adjacent resonator of standard chebyshev filter based on low pass prototype filter. A 4 pole microstrip filter is designed and simulated using HFSS software. The performance of the filter is observed from its coupling co-efficient, midband insertion Loss.

Keywords - HFSS (High Frequency structure simulator), Microstrip filter, LLR (linear lossless reciprocal).

INTRODUCTION

Miniaturized microwave bandpass filters are always in demand for systems requiring small size and light weight. This paper represents the design of a cross-coupled microstrip band-pass filter comprised of coupled hairpin resonators. The hairpin resonator filter is one of the most popular microstrip filter configurations used in the lower microwave frequencies. It is easy to manufacture because it has open-circuited ends that requires no grounding. Its form is derived from the edge-coupled resonator filter by folding back the ends of the resonators into a “U”shape. This reduces the length and improves the aspect ratio of the microstrip significantly.

The capability of placing attenuation poles near the cutoff frequencies of the passband improves the selectivity using fewer resonators. This type of filter is usually realized using waveguide cavities or dielectric-resonator-loaded cavities. However, with the advent of high-temperature superconducting (HTS) and micromachined circuit technologies, there is an increasing interest in microstrip filter structures. Two technical approaches are normally used to realize this type of filter. The first is to extract poles from both ends of a filter prototype by using shunt resonators. The size of the microstrip filter resulting from this approach may, however, be large. The second approach is to introduce a cross coupling between a pair of nonadjacent resonators .The filter employing the cross coupling generally results in a compact topology. This is obviously more attractive for those systems where size is important.

The cross-coupled filters are so attractive because they exhibit ripples in both passband and stopband, which can improve both frequency selectivity and bandpass loss and they are able to place transmission zeros near cutoff frequencies of a passband so that higher selectivity with less resonators can be obtained.

LOW PASS PROTOTYPE DESIGN

For a Linear lossless reciprocal device, the scattering parameter S_{11} is given in s=σ+jΩ domain as,

$$S_{11}(s) = \frac{1}{\varepsilon_R}\frac{F(s)}{E(s)}, \quad |S_{21}(s)|^2 = \frac{1}{1+\left|\frac{\varepsilon}{\varepsilon_R}\right|^2 \left|\frac{F(s)}{E(s)}\right|^2}$$

where F(s)/P(s) is referred to as characteristic function.
Some important properties of the polynomial are given,

1. E(s) is a Hurwitz polynomial of degree N .All its root lies in the left half-plane of s.
2. The roots of F(s) lie on the imaginary axis where the degree is N. These roots are known as reflection zeros.
3. Zeros of the polynomial P(s) are known as transmission zeros (TX zeros). All TX zeros lie on the imaginary axis or appear as pairs of zeros located symmetrically with respect to the imaginary axis.

It is customary in the filter design to synthesize the lowpass prototype (LPP) first. From the designed LPP, components of the actual filter can be obtained by using frequency transformations. Two general LPP configurations are shown in (i) LPP series (ii) LPP shunt

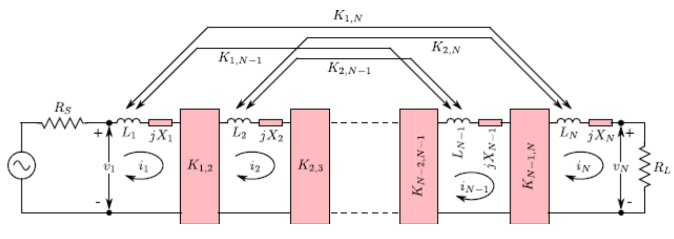

Fig.1. – LPP Series

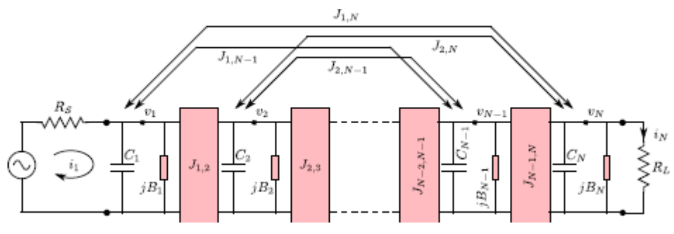

Fig.2. – LPP Shunt

In these figures, the colored components are assumed to be frequency invariant (i.e., do not change with frequency transformations). All the other components change according to the actual filter response required (such as bandpass, bandstop, etc) and their transformed values are shown in fig 3

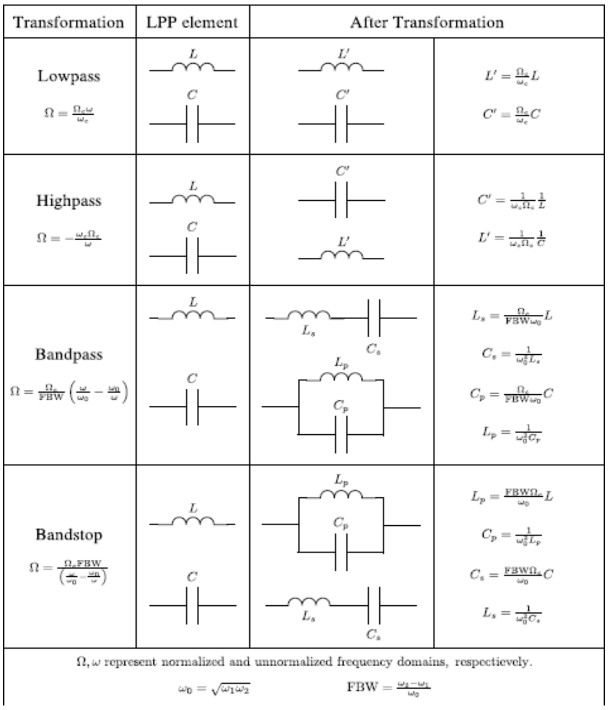

Fig. 3

Also, characteristics of the immittance inverters (both K & J) used in the LPPs are shown in Fig.4a, 4b

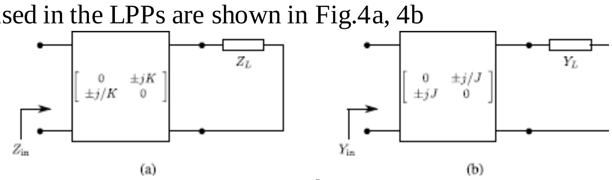

Fig. 4a, 4b

EXTRACTION OF COUPLING CO-EFFICIENTS

It is clear that any coupling encountered in the cross-coupled microstrip hairpin-resonator filters is that of the proximity coupling, which is basically through fringe fields. The nature and the extent of the fringe fields determine the nature and the strength of the coupling. However, the semiopen configuration and inhomogeneous dielectric medium of coupled hairpin resonators make the characterization of couplings even more complicated. On the other hand, it is well known that two resonant peaks in association with the coupling can be observed if the coupled resonators are over-coupled, which occurs when the corresponding coupling coefficient is larger than a critical value amounting to $1/Q_0$, where Q_0 is the quality factor of the resonator circuit. Due to the two split resonant frequencies being quite easily identified in the full-wave EM simulation, the coupling coefficient can be determined from the relationships between the coupling coefficient and the two split resonant frequencies

The electric and magnetic coupling methods are very important in filter designing. These two coupling phenomena's are described in fig.5a, 5b, 5c and fig.6a, 6b, 6c

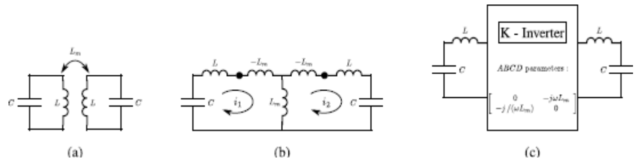

Fig. 5

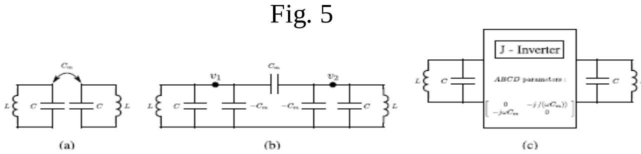

Fig. 6

KVL and KCL equations related to both electric as well as magnetic coupling are as given below:

$$Magnetic\ Coupling: \left(j\omega L + \frac{1}{j\omega C}\right) i_1 + jKi_2 = 0, where\ K = -\omega L_m$$

$$Electric\ Coupling: \left(j\omega C + \frac{1}{j\omega L}\right) v_1 + jJv_2 = 0, where\ J = -\omega C_m$$

If each resonator is isolated from the other, then their resonant frequencies are equal. However, when these two resonators are brought closer to each other, coupling between them yields two distinct resonant frequencies, usually known as f_{even} and f_{odd}.

	Magnetic Coupling	Electric Coupling
f_{even}	$\frac{1}{2\pi\sqrt{C(L+L_m)}}$	$\frac{1}{2\pi\sqrt{C(C-C_m)}}$
f_{odd}	$\frac{1}{2\pi\sqrt{C(L-L_m)}}$	$\frac{1}{2\pi\sqrt{C(C+C_m)}}$
$\frac{L_m}{L}$ or $\frac{C_m}{C}$	$\frac{f_{odd}^2-f_{even}^2}{f_{odd}^2+f_{even}^2}$	$\frac{f_{even}^2-f_{odd}^2}{f_{even}^2+f_{odd}^2}$

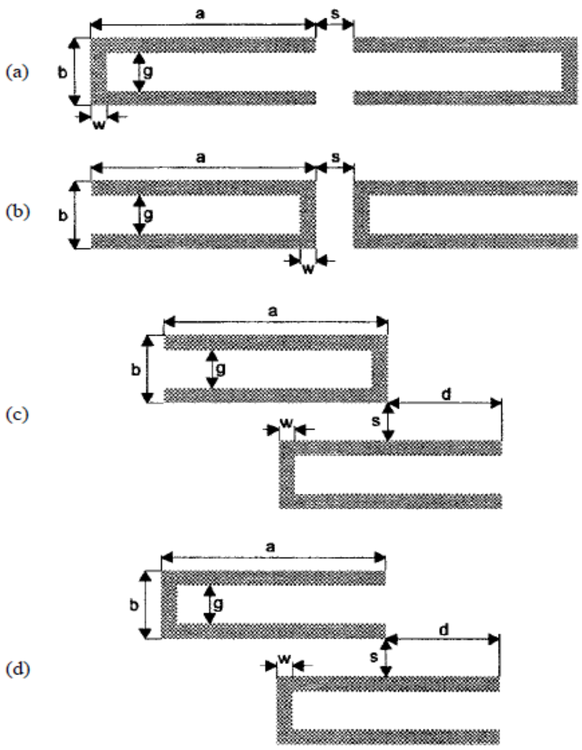

Fig.7.Basic coupling structures of coupled microstrip hairpin resonators with a relative dielectric constant and a thickness h. (a) Electric coupling. (b) Magnetic coupling. (c)

The first type of mixed coupling. (d) The second type of mixed coupling.

Figure 7 shows the four basic coupled structures for realization of cross-coupled filters. It is obvious that at resonance the electric fringe field is much stronger near the open ends, while the magnetic fringe field is much stronger near the middle of each resonator. This leads to the electric coupling of Fig. 7(a) and the magnetic coupling of Fig. 7(b). In the coupling structure of Fig. 7(c), the two hairpin resonators in the opposite orientation are coupled with each other with a separation s and offset d. In this case, both the electric and magnetic couplings occur, and may be referred to as the first type of mixed coupling. It can be shown that the first type of mixed coupling has resulted from the superposition of the magnetic and electric couplings which are in-phase. The coupling structure of Fig. 7(d) consists of the two coupled hairpin resonators in the same orientation with a separation s and offset d. Although both the electric and magnetic couplings occur as the first mixed coupling structure discussed above, this coupling structure exhibits a rather different coupling characteristic. Hence, in this case the coupling may be referred to as the second type of mixed coupling. It can also be shown that the magnetic and electric couplings are out-phase in the second type of mixed coupling.
Using HFSS we simulated the frequency responses of the four basic coupling structures. The coupling coefficients would then be extracted from the simulated frequency responses by using

$$k = \frac{f_{p2}^2 - f_{p1}^2}{f_{p2}^2 + f_{p1}^2}$$

where f_{p1} and f_{p2} are the two spilt resonant frequencies.

The coupling between any pair of coupled resonators is obtained using HFSS simulation.

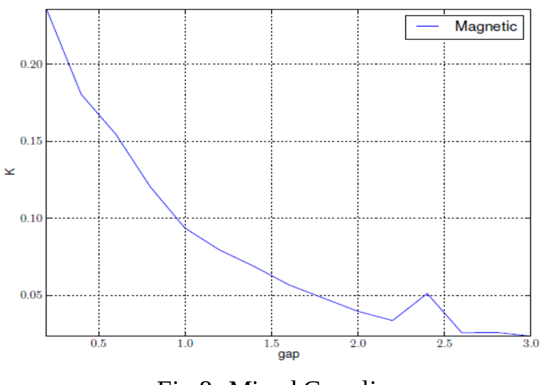

Fig 8- Mixed Coupling

COUPLING MATRIX REDUCTION

The coupling matrix M that emerge from the synthesis procedure all have non zero values. The nonzero values represent the offsets from the centre frequency of each resonance (i.e. Coupling exists between each resonator).Since it is impractical it is usual to annihilate the couplings with similarity transforms. The use of similarity transform ensures that the eigenvalues and eigenvectors of the coupling matrix M are preserved such that under analysis, the transformed matrix yields exactly the same transfer and reflection characteristics as the original coupling matrix. The Original coupling matrix obtained is

```
M:
[[-0.       -0.55561  0.55561 -0.      ]
 [-0.55561 -0.63418 -0.       -0.55561]
 [ 0.55561 -0.        0.63418 -0.55561]
 [-0.       -0.55561 -0.55561  0.      ]]
```

After similarity transform the coupling matrix is ,

```
[[-0.        -0.15715   -0.         -0.       ]
 [-0.15715   -0.         -0.126836  -0.       ]
 [-0.         -0.126836  -0.         -0.15715 ]
 [-0.         -0.         -0.15715    0.       ]]
```

CONCLUSION

As we have emphasized, the characterization of couplings of the basic coupling structures plays an important role in the design of micro strip hairpin-resonator filters. Accordingly, a four-pole hairpin band pass filter was designed and simulated. The filter specifications are the center frequency of 2 GHz, the pass band bandwidth of 20 MHz (or the fractional bandwidth FBW = 2.07%),gap between the resonator 1 and 2 =0.65mm, gap between the resonator 2 and 3 =0.75, gap between the resonator 3 and 4 =0.65mm,resonator width=2mm,resonator trace width =1mm,feed=4.2mm. The design parameters of the band pass filter (namely, the elements of coupling matrix and input/output single-loaded external Qe) could then be calculated as follows:

M_{12}= M_{21} = M_{34}= M_{43} = -0.1571
M_{23}= M_{32} = -0.1268
Q =6.2647

The negative coupling coefficients are realized by the mixed coupling. These coupling coefficients are within the range of coupling coefficients so that the design dimensions could be found. The tapped-line feed was used for the loaded external Qe, though the other means of feed such as the coupled-line feed is feasible. The return loss obtained at 2 GHz is 43.53dB.

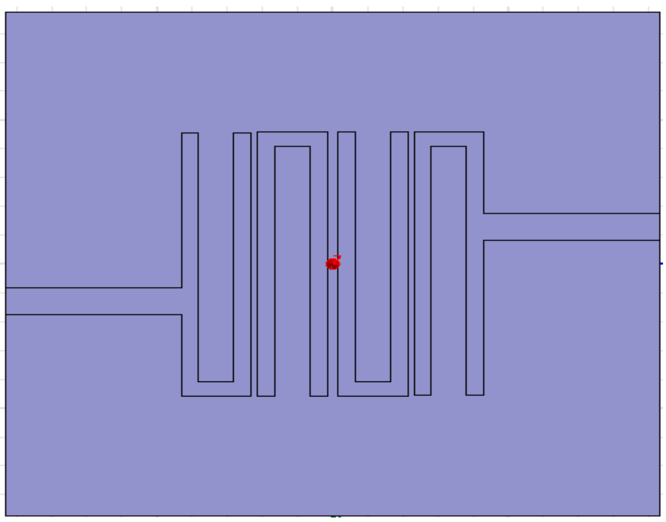

Fig.9. 4 pole Filter

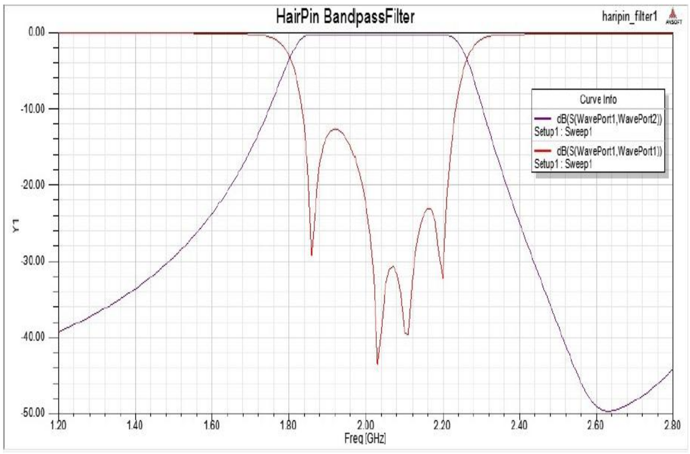

Fig.10. HairPin Bandpass Filter Response

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