

CAD of Cross-Coupled Miniaturized Hairpin Bandpass Filters

Ilia G. Iliev¹ and Marin V. Nedelchev²

Abstract: This paper describes cross-coupled microstrip miniaturized hairpin bandpass filters. Formulas for coupling coefficients are derived. A design technique using an approximation polynomial and a filter-prototype is proposed. Synthesis results, obtained by the design technique, are shown.
Keywords: Microstrip bandpass filter, Cross-coupled resonators, Miniaturized hairpin resonators.

I. INTRODUCTION

For their compact structures, $\lambda/2$ microstrip hairpin filters are widely used in the X-band (Fig.1). However the size of these resonators is still too large to design a filter in the L-band. To make these resonators smaller in the L-band can be used microstrip line split-ring resonators, which are composed of a transmission line and a lumped element capacitor (Fig1b).

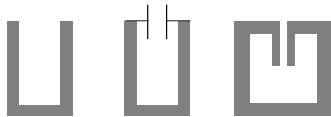


Fig. 1a Fig. 1b Fig. 1c

The authors of [1] proposed a miniaturized hairpin resonator with parallel-coupled lines instead of the lumped capacitor (Fig1c). The proposed filters are adjacent resonators coupled as it is shown in fig2. Slow-wave open-loop bandpass filters (BPF) are depicted in [2,3].



Fig.2

Mobile communication systems require a narrow-band BPF having low passband loss, high selectivity, small size and flat group delay. Such a filter response can be realized using filter with cross coupling between nonadjacent resonators. Normally this kind of filters is realized using waveguide cavities or dielectric resonators.

This paper describes a method for CAD of cross-coupled BPF using miniaturized hairpin resonator. In contrast to the design techniques in [2,3] no full-wave electromagnetic (EM) simulator is needed here. Microstrip slow-wave open-loop filter is proposed in [2,3].

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Synthesis algorithm of BPF composed of adjacent coupled resonators is described in [1,5]. Cross-coupled structures are shown in fig.3. This type of coupling reduces the filter size.



Fig.3

The design technique uses an approximation polynomial and a lowpass filter (LPF) prototype. The loaded Q factor and the coupling coefficient between the different resonators could be calculated. The main problem in the BPF synthesis is determining the coupling coefficients for different coupling structures. It is quite easy to identify in the full wave EM simulation the two split resonant frequencies, which are related to the coupling coefficient. The paper solves the problem analytically using the filter electrical parameters.

II. ANALYSIS OF COUPLING COEFFICIENTS FOR THE BASIC COUPLING STRUCTURES

Basically the coupling structures are:

A. Magnetic coupling

Magnetic coupling of two resonators is shown in fig.4.

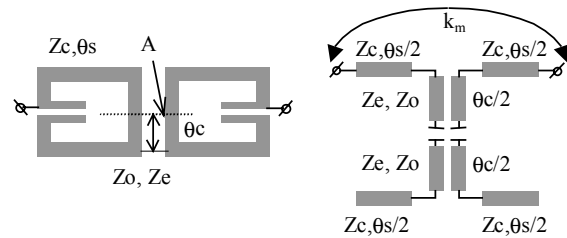


Fig.4

For the resonator symmetry in the point A electric component of the electromagnetic wave is equal to zero. So that point A can be grounded. Using the equivalent scheme shown in fig4b, coupling coefficient is given by:

$$k_m = \frac{1}{bZ_c} \left(\frac{Z_c - Z_o \operatorname{tg} \frac{\theta_c}{2} \operatorname{tg} \frac{\theta_s - \theta_c}{2}}{Z_o \operatorname{tg} \frac{\theta_c}{2} - Z_c \operatorname{tg} \frac{\theta_s - \theta_c}{2}} - \frac{Z_c - Z_e \operatorname{tg} \frac{\theta_c}{2} \operatorname{tg} \frac{\theta_s - \theta_c}{2}}{Z_e \operatorname{tg} \frac{\theta_c}{2} - Z_c \operatorname{tg} \frac{\theta_s - \theta_c}{2}} \right) \quad (1)$$

Where Z_c is the transmission line characteristic impedance, Z_e, Z_o are even and odd mode impedance of the coupled lines,

θ_s is electrical length of the resonator, θ_c is electrical length of the coupled lines defining the magnetic coupling, b is the resonator's admittance slope.

B. Electric coupling

Two resonators are electrically coupled in the case shown in fig.5a.

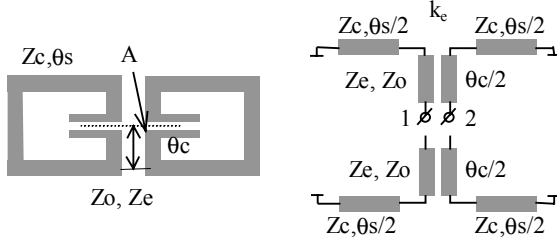


Fig.5a

Fig.5b

In the point A, magnetic component of the electromagnetic wave is equal to zero. Using the equivalent scheme shown in fig.5b, derived formula for the coupling coefficient is :

$$k_e = \frac{1}{b} \left(\frac{1}{Z_o} \frac{Z_o - Z_c \operatorname{tg} \theta_c \operatorname{tg} \left(\frac{\theta_s}{2} - \theta_c \right)}{Z_c \operatorname{tg} \left(\frac{\theta_s}{2} - \theta_c \right) - Z_o \operatorname{tg} \theta_c} - \frac{1}{Z_e} \frac{Z_e - Z_c \operatorname{tg} \theta_c \operatorname{tg} \left(\frac{\theta_s}{2} - \theta_c \right)}{Z_c \operatorname{tg} \left(\frac{\theta_s}{2} - \theta_c \right) - Z_e \operatorname{tg} \theta_c} \right) \quad (2)$$

C. Mixed coupling:

Mixed coupling is in the following two resonator structures shown in fig.6 and fig.7.

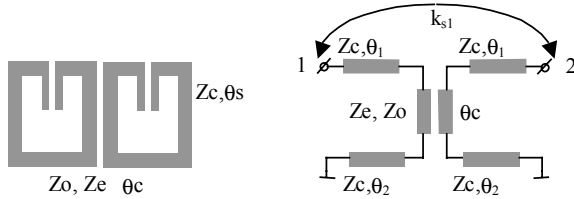


Fig.6a

Fig.6b

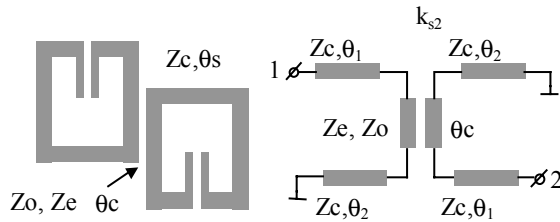


Fig.6a

Fig.6b

For the resonator configuration shown in fig.6a and using the equivalent scheme shown in fig.6b, the coupling coefficient is:

$$k_{s1} = \frac{y_{21}}{2b} \quad (3)$$

$$y_{21} = \frac{1}{|Z_{2o}|} - \frac{1}{|Z_{2e}|}, \quad Z_{2e} = Z_c \frac{Z_{1e} + jZ_c \operatorname{tg} \theta_1}{Z_c + jZ_{1e} \operatorname{tg} \theta_1},$$

$$Z_{2o} = Z_c \frac{Z_{1o} + jZ_c \operatorname{tg} \theta_1}{Z_c + jZ_{1o} \operatorname{tg} \theta_1}, \quad Z_{1e} = Z_c \frac{Z_1 + jZ_e \operatorname{tg} \theta_2}{Z_e + jZ_1 \operatorname{tg} \theta_2},$$

$$Z_{1o} = Z_c \frac{Z_1 + jZ_o \operatorname{tg} \theta_2}{Z_o + jZ_1 \operatorname{tg} \theta_2}, \quad Z_1 = jZ_c \operatorname{tg} \theta_2.$$

The coupling coefficient for the resonator configuration shown in fig.7a is derived using the equivalent scheme shown in fig.7b:

$$k_{s2} = \frac{y_{21}}{2b} \quad (4)$$

$$y_{21} = \frac{1}{|Z_{2o}|} - \frac{1}{|Z_{2e}|}, \quad Z_{io} = Z_c \frac{Z_{i1} + jZ_c \operatorname{tg} \theta_1}{Z_c + jZ_{i1} \operatorname{tg} \theta_1}, \quad Z_{i2} = \frac{1}{2y_{12a}}$$

$$Z_{ie} = Z_c \frac{Z_{ii} + jZ_c \operatorname{tg} \theta_1}{Z_c + jZ_{ii} \operatorname{tg} \theta_1}, \quad Z_{ii} = \frac{Z_{i1} Z_{i2}}{Z_{i1} + Z_{i2}}, \quad Z_{i1} = \frac{1}{y_{11a} - y_{12a}},$$

$$y_{11a} = -j \left(\frac{\cot g \theta_c}{w} + \frac{2 \cos \theta_c}{dv^2 w \sin^3 \theta_c} \right) +$$

$$+ j \left[\frac{1}{d} \left(\frac{\cot g \theta_c}{w} + y_c \cot g \theta_2 \right) \left(\frac{\cot g \theta_c}{v^2} + \frac{1}{w^2 \sin^2 \theta_c} \right) \right],$$

$$y_{21a} = j \frac{1}{v \sin \theta_c} \left\{ 1 + \frac{1}{d} \left[\frac{1}{w^2 \sin^2 \theta_c} + \frac{\cot g^2 \theta_c}{v^2} - 2 \left(\frac{\cot g \theta_c}{w} + y_c \cot g \theta_2 \right) \frac{\cot g \theta_c}{w} \right] \right\},$$

$$d = -\frac{1}{v^2 \sin^2 \theta_c} + \left(\frac{\cot g \theta_c}{w} + y_c \cot g \theta_2 \right)^2,$$

$$\frac{1}{v} = \frac{1}{2} \left(\frac{1}{Z_o} - \frac{1}{Z_e} \right), \quad \frac{1}{w} = \frac{1}{2} \left(\frac{1}{Z_e} + \frac{1}{Z_o} \right).$$

D. Tapped input electrical length

Tapped input electrical length is determined in [5] (fig.7):

$$\theta_t \approx \arctg \sqrt{\frac{G_a}{bQ_{ei}}} \quad (5)$$

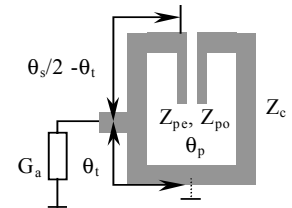


Fig.7

E. Admittance slope parameter

A closed formula for admittance slope parameter is derived in [5]:

$$b = -\frac{1}{2} \frac{A+B}{C} \quad (6),$$

$$A = \frac{(Z_{pe} - Z_{po})\theta_p}{\sin^2 \theta_p} + \frac{(Z_{pe} + Z_{po})\theta_p \cos \theta_s}{\sin^2 \theta_p} + (Z_{pe} + Z_{po})\theta_s \sin \theta_s \cot g \theta_p,$$

$$B = \theta_s \cos \theta_s \left(\frac{Z_{pe} Z_{po}}{Z_c} \cot g^2 \theta_p - Z_c \right) + \frac{2Z_{pe} Z_{po} \theta_p \cot g \theta_p \sin \theta_s}{Z_c \sin^2 \theta_p},$$

$$C = 2Z_{pe} Z_{po} \cot g^2 \theta_p \cos \theta_s - Z_c (Z_{pe} + Z_{po}) \sin \theta_s \cot g \theta_p.$$

III. DESIGN TECHNIQUE OF CROSS-COUPLED BPF USING MINIATURIZED HAIRPIN RESONATORS

1. Finding the element values of LPF prototype, using the approximate synthesis method described in [4].

The relations between the bandpass design parameters and the lowpass elements are:

$$Q_{ei} = Q_{eo} = C_i / \Delta\omega$$

$$k_{n,n-1} = k_{N-n,N-n+1} = \frac{\Delta\omega}{\sqrt{C_n C_{n+1}}}, \text{ for } n=1 \text{ to } N/2$$

$$k_{m,m+1} = \frac{\Delta\omega J_m}{C_m}, \text{ for } m=N/2,$$

$$k_{m-1,m+2} = \frac{\Delta\omega J_{m-1}}{C_{m-1}}, \text{ for } m=N/2,$$

where $\Delta\omega$ denotes the fractional bandwidth of the bandpass filter,

C is the capacitance of the lumped capacitor, J is the characteristic admittance of the inverter, N is the degree of the filter.

LPF prototype is shown in Fig.8.

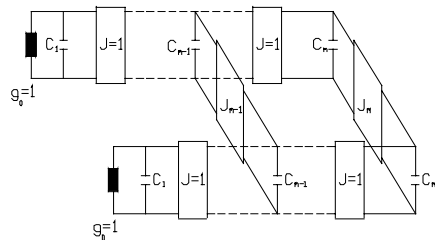


Fig.8

2. Calculating the resonator parameters.

Let choose resonator electric length θ_s , characteristic impedance Z_c , even and odd mode impedance Z_{pe} , Z_{po} of the coupled lines. The length of the coupled lines can be calculated by:

$$\cot g \theta_p = \frac{-R + \sqrt{R^2 + 4Z_c^2 \sin^2 \theta_s}}{2Z_c \sin \theta_s} \quad (7),$$

$$\text{and } R = (Z_{pe} + Z_{po}) \cos \theta_s - (Z_{pe} - Z_{po}).$$

The admittance slope parameter can be calculated by Eq.(6).

3. Calculation the coupling parameters between the resonators. One of the possibilities is to choose Z_e and θ_c . Z_o can be calculated from the coupling coefficients and Eqs. (1÷4);

4. Calculation of the input tapped electrical length. Defining the tapped input electrical length is made by (5);

5. Calculation of the geometric parameters of the filter for an exact substrate;

6. Optimization filter parameters by varying the geometric dimensions.

IV. DESIGN EXAMPLE

Cross-coupled miniaturized hairpin filter design software is developed on the base of the algorithm described in the previous paragraph. Two four resonator filters with different bandwidths are synthesized. The bandwidths are respectively 1% - Filter I and 5% - Filter II. The central frequency is $f_0 = 1.44 \text{ GHz}$. The coupling coefficients are:

Filter I	Filter II
$k_{12} = k_{34} = 0.00856$	$k_{12} = k_{34} = 0.04337$
$k_{14} = -0.002195$	$k_{14} = -0.002905$
$k_{23} = 0.00786$	$k_{23} = 0.03857$

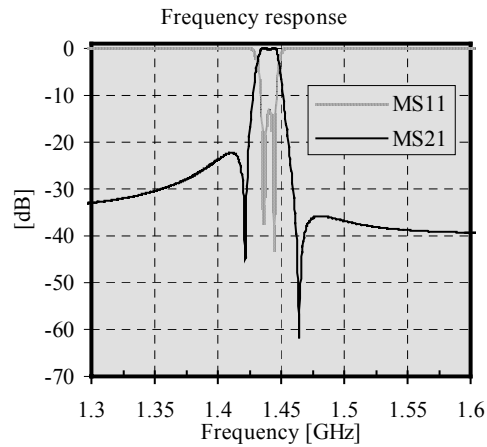


Fig.9

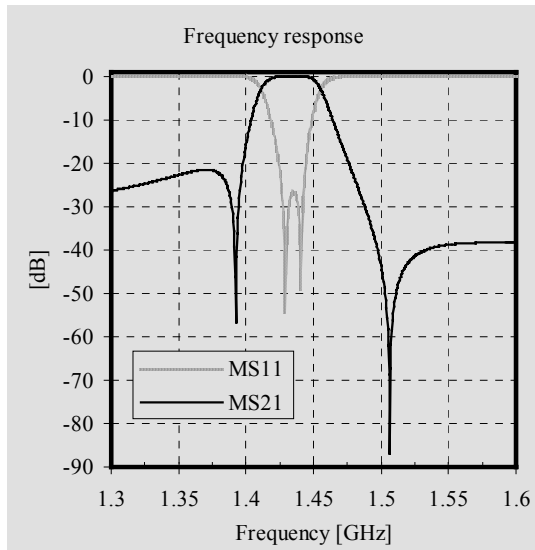


Fig.10

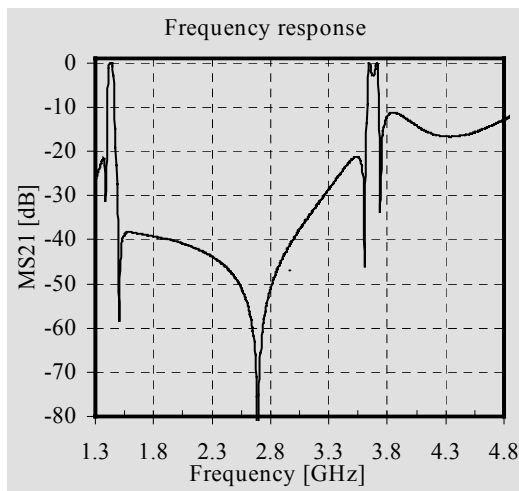


Fig.11

Figures 9, 10 and 11 show frequency responses of the synthesized filters. Filters analysis is done with the electrical parameters of the lines. The results confirm that the proposed design technique leads to filters with the desired characteristics. The cross-coupled filters have greater response skirts than the ordinary coupled filters. Figure 11 shows that the spurious bandwidth of the filter is not on a harmonic of the central frequency $f_0=1.44GHz$. This type of filters is suitable for filtering signals including high-order harmonics.

V. CONCLUSION

This paper describes a cross-coupled BPF CAD design technique. The calculation of the filter can be done without using full wave EM simulator and constraints for the line or substrate parameters. The derived formulas allow analytical calculation of the filter. The constraints of the design technique are related with the increasing inaccuracy of the method for strong coupling between the resonators. This disadvantage can be decreased by optimization process after the initial filter calculation.

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