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Microstrip Spiral-Coupled Scheme Bandpass Filters with Mutual Electric and Magnetic Coupling

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Abstract—This paper presents miniaturized microstrip bandpass filters based on the quarter-wavelength ($\lambda/4$) spiral-coupled scheme (SCS). By introducing the mutual electric and magnetic coupling between two SCS, two resonant modes can be well adjusted to form a passband and four transmission zeros can be generated to obtain a good out-of-band rejection. Without modifying the initial design parameters, another transmission zero with strong attenuation can be further introduced by loading a parallel-coupled line section along the symmetrical plane, thus largely widening the upper stopband. To verify our proposal, two filters are finally implemented and fabricated, which exhibit the following attractive features: low insertion loss, good passband selectivity, wide stopband with a high rejection level, and compact size.

Index Terms—Bandpass filter, dual-mode, mutual electric and magnetic coupling, parallel-coupled line, spiral-coupled scheme (SCS).

I. INTRODUCTION

For a design of microwave bandpass filter, compact size, good selectivity, and wide stopband are generally important features in the practical applications of the modern communication systems. To meet these requirements, multiple transmission zeros are always preferred to not only enhance the passband selectivity, but also extend the upper stopband bandwidth. It is very known that the cross-coupling between resonators can introduce multiple transmission zeros at two sides of passband to enhance the passband selectivity [1]-[2]. Meanwhile, the periodically nonuniform-coupled line or over coupled scheme can be utilized to compensate two different phase velocities in the parallel-coupled line section, thus suppressing a harmonic passband at the upper stopband. Moreover, to further improve the upper stopband performance, a parallel-coupled line is installed along the symmetrical plane, which can allocate another transmission zero at upper stopband. As a result, the totally five transmission zeros are generated outside the operating passband, which largely improve the out-of-band rejection and widen the stopband bandwidth.

II. FILTER SCHEMATIC AND OPERATIONAL PRINCIPLE

Fig. 1(a) and (b) illustrate the layout and topology of the proposed $\lambda/4$ SCSs with involved mutual electric and magnetic coupling. (a) Layout. (b) Topology. (c) Equivalent transmission line model. (MEMC: mutual electric and magnetic coupling.)

magnetic coupling can not only provide the finely adjusted two resonances for the passband operation, but also introduce four transmission zeros for the stopband enhancement. Meanwhile, to further improve the upper stopband performance, a parallel-coupled line is installed along the symmetrical plane, which can allocate another transmission zero at upper stopband. As a result, the totally five transmission zeros are generated outside the operating passband, which largely improve the out-of-band rejection and widen the stopband bandwidth.
the $\lambda/4$ SCSs. The even- and odd-mode equivalent circuits with the involved mutual electric and magnetic coupling are shown in Fig. 2. Then, the even- and odd-mode characteristic impedances (i.e., $Z_{\text{ine}}$ and $Z_{\text{ino}}$) of the structure depicted in Fig. 1 can be derived as follows

$$Z_{\text{ine,o}} = \sqrt{Z_{\text{oe}}Z_{\text{oo}}Z_{\text{Le,o}} + Z_{\text{oe}}Z_{\text{oo}}\tan^2 \beta l}$$

(1)

where

$$Z_{\text{Le}} = -j\omega \left(2\omega^2 A L_{\text{via}} C_p - A - 2l_{\text{via}}\right)$$

$$Z_{\text{Lo}} = \frac{j\omega B}{1 - \omega^2 (C_p + 2C_m) B}$$

$$A = L_p + L_m$$

$$B = L_p - L_m$$

(2)-(5)

$Z_{\text{oe}}$ and $Z_{\text{oo}}$ are the even- and odd-mode impedances of the SCS, and $\beta l$ indicates its electric length. It is important to notice that the electric length (i.e., $\beta l$) of the involved mutual electric and magnetic coupling is much smaller than a quarter wavelength ($\lambda/4$). Thus, the inductances (i.e., $L_p$, $L_m$, and $L_{\text{via}}$) and capacitances (i.e., $C_p$ and $C_m$) can be expressed as follows [11]-[12], respectively

$$L_p = \frac{\left(Z_{\text{oe}} + Z_{\text{oo}}'\right) \sin \beta l}{4\pi f}$$

$$C_p = \frac{\tan \beta l}{2\pi f Z_{\text{oe}}}$$

$$L_m = \frac{Z_{\text{oe}} - Z_{\text{oo}}}{Z_{\text{oe}}' + Z_{\text{oo}}} L_p$$

$$C_m = \frac{1}{Z_{\text{oo}}' - Z_{\text{oe}}} - \frac{1}{Z_{\text{oo}}} \tan \beta l - \frac{1}{4\pi f}$$

$$L_{\text{via}} = \frac{\mu_0}{2\pi} \left[ h \ln \left( \frac{h + \sqrt{r^2 + h^2}}{r} \right) + \frac{3}{2} \left( r - \sqrt{r^2 + h^2} \right) \right]$$

(6)-(10)

where $r$ (i.e., $2r = d$) is the radius of the metal-via from the strip conductor to the ground, and $h$ is the thickness of the dielectric substrate, and $Z_{\text{oe}}$ and $Z_{\text{oo}}$ are the characteristic impedances of the even- and odd-mode for the involved mutual electric and magnetic coupling. From (1), the $S_{21}$ and $S_{11}$ of the proposed structure shown in Fig. 1 can be derived as [1]

$$S_{21} = \frac{Z_0 (Z_{\text{ine}} - Z_{\text{ino}})}{(Z_{\text{ine}} + Z_{\text{ino}})(Z_{\text{ine}} - Z_{\text{ino}})}$$

$$S_{11} = \frac{Z_{\text{ine}}Z_{\text{ino}} - Z_0^2}{(Z_{\text{ine}} + Z_{\text{ino}})(Z_{\text{ine}} - Z_{\text{ino}})}$$

(11)-(12)

From (11), the transmission zeros can be estimated when

$$Z_{\text{ine}} = Z_{\text{ino}}$$

(13)
Therefore, based on the above analysis, the frequency characteristics of the proposed scheme can be well predicted. To further investigate and demonstrate the characteristics of the proposed scheme, the dielectric substrate RT5880 (i.e., $\epsilon_r = 2.2$ and a thickness of 0.508 mm) and EM-simulator IE3D are used. Fig. 3 and Fig. 4 depict the involved two resonances and four transmission zero frequencies versus the coupling dimensions (i.e., $l_p$ and $w_p$) and metal-via radius $r$ (i.e., $d/2$), respectively. From Fig. 3, as $w_p$ increases, the $f_1$, $f_{z1}$, and $f_{z4}$ increase, whereas the $f_2$, $f_{z2}$, and $f_{z3}$ decrease. Thus, the relatively narrow passband and wide stopband can be achieved when the larger $w_p$ is chosen. On the other hand, it can be found in Fig. 4 that as the metal-via radius $r$ increases, the $f_{z3}$ decreases, whereas the $f_1$, $f_2$, $f_{z1}$, $f_{z2}$, and $f_{z4}$ remain almost the same. Therefore, by carefully adjusting $r$, the desired location of $f_{z3}$ can be easily achieved, which could improve the rejection level of the upper stopband. Based on the investigation above, two functions of the involved mutual electric and magnetic coupling for the filter design can be concluded: 1) the capacitive coupling employed by the mutual electric and magnetic coupling in the SCSs (i.e., $l_p$ and $w_p$) is the primary element to determine the required resonances and transmission zeros. 2) the good stopband rejection level can be achieved by the inductive coupling introduced by the metal-via (i.e., $r$). To further extend the upper stopband bandwidth for the filter design with enhanced performance, a parallel-coupled lines are tapped at the symmetrical plane as shown in Fig. 5. This structure could affect the mutual capacitive coupling and introduce an additional transmission zero $f_{z5}$ with deep attenuation, which can be utilized to suppress the spurious frequency response at upper stopband. Fig. 6 depicts the transmission zero frequency $f_{z5}$ versus the dimensions (i.e., $w_{p1}$ and $l_{p1}$) of the loaded parallel coupled-line, which can be adjusted in a relatively large range.

III. FILTER DESIGN AND EXPERIMENTAL RESULTS

The design procedures of the proposed structures for filter design are shown as follows. The first step is to achieve the desired dual-resonances for the specified passband and allocate transmission zeros in the stopband, by properly tuning the mutual electric and magnetic coupling (i.e., dimensions i.e., $l_p$, $w_p$, and $r$) of the $\lambda/4$ SCSs. The second step is to determine the dimensions of tapping and coupling structures to satisfy the requirement of external quality factor ($Q_e$) and coupling coefficient ($k$), which could be obtained using the synthesis methods [1]. Fig. 7 illustrates the simulated passband current-density distribution of the proposed SCSs. By making the width $w$ wider and gap $s$ narrower, the peak of the edge-current can be effectively reduced, then a smoother current distribution away from the edges and strong current distribution on the involved mutual electric and magnetic coupling can be obtained. Therefore, the Q-factor of the resonator could be improved with less conductor loss. For this reason, the ratio of the width $w$ to gap $s$ in the $\lambda/4$ SCS is optimally chosen at 4:1. We can find that the gap $s$ of the tapped feed-lines and SCSs are the critical elements to affect $k$. According to the filter specifications, the initial dimensions of the filter can be determined.

Next, to verify the operation and design procedure above, two bandpass filters (i.e., Filter I and Filter II) are implemented and fabricated, as illustrated in Fig. 8. The measured results shown in Fig. 9 and Fig. 10 are obtained from the Agilent 5230A network analyzer. The filters are designed to operate
IV. CONCLUSION

In this paper, two compact bandpass filters using the \( \lambda/4 \) SCS have been designed at 3.37 GHz and 3.33 GHz with a 3 dB fractional bandwidth (FBW) of 19% and 18.5%, respectively. The typical in-band insertion loss for Filter I and Filter II is 0.53 dB and 0.55 dB, respectively. Specifically, the upper stopband bandwidth of Filter II is about \( 5f_0 \) (i.e., \( f_0 \) is 3.33 GHz) of the center resonant frequency with a typical rejection level of 40 dB. In addition, both filters have the merits of miniaturized size. The size of the Filter I is 7.4 mm \( \times \) 6.8 mm (i.e., 0.11 \( \lambda_g \) \( \times \) 0.1 \( \lambda_g \), where \( \lambda_g \) is the microstrip guided wavelength on the substrate at the center frequency of 3.37 GHz), including the 50\( \Omega \) I/O ports. Similarly, the size of the Filter II is only 8.9 mm \( \times \) 6.8 mm (i.e., 0.135 \( \lambda_g \) \( \times \) 0.1 \( \lambda_g \), where \( \lambda_g \) is the microstrip guided wavelength on the substrate at the center frequency of 3.33 GHz), including the 50\( \Omega \) I/O ports.

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