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NOVEL MICROSTRIP BANDPASS FILTERS WITH TRANSMISSION ZEROS

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Abstract—Some design methods for band-pass filters based on half-wavelength resonators have been proposed. The main feature of these methods is that n or $n+1$ transmission zeros can be generated for a structure composed of n resonators. To demonstrate the usefulness of the proposed filter structures, three kinds of two-pole compact microstrip hairpin filters are designed and fabricated. Good agreement between measured and simulated data has been demonstrated.

1. INTRODUCTION

The out-of-band characteristics of a bandpass filter can be effectively improved by generating attenuation poles in the rejection band. Some common techniques to obtain such effects are: using cross-coupling between nonadjacent resonators [1–6], applying a kind of dud-mode filter composed of ring resonators [7], adopting a technique to create multiple attenuation poles by a tap-coupling structure [8]. Recently, microstrip bandpass filters were proposed that used hairpin resonators with asymmetric input and output feed lines tapping on the first and last resonators to obtain two transmission zeros lying on either side of the passband [9, 10]. However, in [9, 10], there is only one type of electric coupling filter using skew-symmetric feed structure which is researched. The type of magnetic coupling filter is not analysed. The serial and parallel $\lambda/2$ microstrip line filters with transmission zeros are analysed by using EM simulation in [11, 12] respectively. But [11, 12] did not discuss the amount and variation of transmission zeros due to the placement of the tapping positions of the different feed lines. [13, 14] designed well-performing bandpass filters with transmission zeroes by only utilizing the open-circuited stubs and didn't consider the affect of the types of coupling.

In this paper, we propose some new design methods for bandpass filters with transmission zeros. The relations between the types of coupled microstrip transmission lines and transmission zeroes are thoroughly analysed. The transmission zeros are due to the series resonance of the quarter-wavelength open stub, and parallel coupling structure with loads at skew-symmetric ports and parallel coupled-lines with symmetric feed structure by increasing the velocity ratio. Finally, these structures are applied to the design of two-pole microstrip filters to further improve selectivity. The numerical results are verified by the simulation, and good agreement between the measured and simulated have been obtained.

2. ANALYSIS OF TRANSMISSION-ZERO CONDITIONS

Here the relations between the types of coupled microstrip transmission line and transmission zeros of transfer response are thoroughly analysed.

2.1. The Serial $\lambda/2$ Microstrip Line Filters with Symmetric and Skew-Symmetric Feed Structure

Fig. 1(a) shows a $\lambda/2$ microstrip line filter using symmetric feed structure. The input and output feed lines divide the resonators into two sections of l_1 and l_2. The total length is $l = l_1 + l_2 = \lambda_{g0}/2$, where λ_{g0} is the guided wavelength at fundamental resonance. The coupling between the two open ends of the resonators is simply expressed by the gap capacitance c_s [5]. Inspecting Fig. 1(a), the $ABCD$ matrix for the sections of the lossless circuit is

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = M_1 M_2 M_3 \tag{1}$$

with

$$M_1 = \begin{bmatrix} 1 & 0 \\ jY_0 \tan\theta_1 & 1 \end{bmatrix} = M_3$$

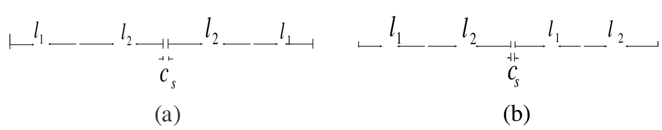


Figure 1. Equivalent circuit of the serial $\lambda/2$ microstrip line filter with (a) symmetric feed structure and (b) skew-symmetric feed structure.

$$
\begin{aligned}
M_2 &= \begin{bmatrix} A_2 & B_2 \\ C_2 & D_2 \end{bmatrix} \\
&= \begin{bmatrix} \cos(2\theta_2) + \dfrac{Y_0}{\omega c_s}\cos\theta_2 \sin\theta_2 & jZ_0\sin(2\theta_2) - j\dfrac{\cos^2\theta_2}{\omega c_s} \\ jY_0\sin(2\theta_2) + j\dfrac{Y_0^2}{\omega c_s}\sin^2\theta_2 & \cos(2\theta_2) + \dfrac{Y_0}{\omega c_s}\sin\theta_2\cos\theta_2 \end{bmatrix}
\end{aligned}
$$

Where $D_2 = A_2,\ \theta_1 = \beta l_1,\ \theta_2 = \beta l_2,\ \beta$ is the propagation constant, ω is the angular frequency, and $Z_0 = 1/Y_0$ is the characteristic impedance of the resonator. S_{21} of the circuit can then be calculated from the $ABCD$ matrices and is expressed as

$$S_{21} = \frac{2Z_L}{S} \tag{2}$$

$$
\begin{aligned}
S &= (2A_2 + 2jB_2Y_0\tan\theta_1)Z_L + B_2 \\
&\quad + \left(j2A_2Y_0\tan\theta_1 + C_2 - B_2Y_0^2\tan^2\theta_1\right)Z_L^2
\end{aligned} \tag{2a}
$$

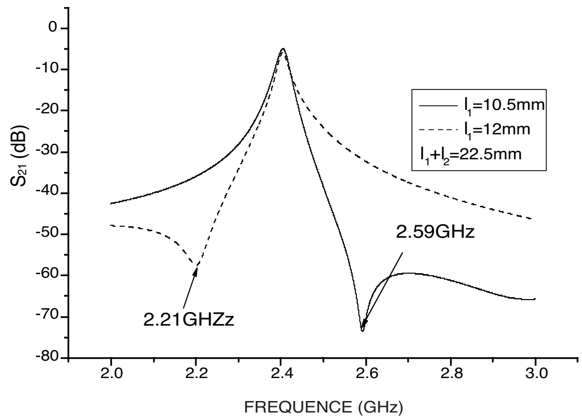
Where Z_L is the load impedance. Since the numerator of S_{21} is not equal to zero, the necessary and sufficient condition for the existence of the transmission zero is $|S| \to \infty$, namely, $\tan\theta_1 \approx \infty$, so the transmission zeros positions are $f_1 = nc/(4l_1\sqrt{\varepsilon_{eff}})$. Where ε_{eff} is the effective dielectric constant, n is the mode number, c is the speed of light in free space. The symmetrically fed bandpass filters exhibit one transmission zero only in the upper or lower stopband. In the passband, when $\theta_1 + \theta_2 \approx \pi$, the transmission coefficient can then be found as

$$S_{21} = \frac{1}{2 + j\left(Z_0\sin(2\theta_2) - \dfrac{\cos^2\theta_2}{Z_L\omega c_s}\right)} \tag{3}$$

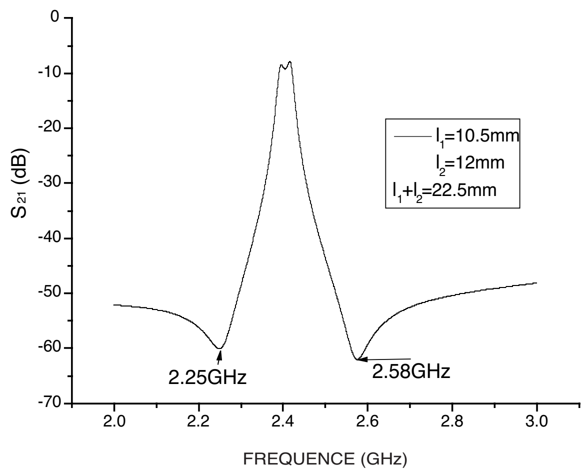
Fig. 2(a) shows the simulated results for different tapping positions on the transmission line resonators in Fig. 1(a). The filters were designed at the fundamental frequency of 2.4 GHz and fabricated on an substrate with a thickness 25 mil and a relative dielectric constant $\varepsilon_r = 9.6$. The location of transmission zeros can be calculated, $f_1 = 2.26$ GHz and 2.59 GHz. Inspecting the results, the simulation agree well with the calculations.

Fig. 1(b) shows one type of $\lambda/2$ microstrip line filter using skew-symmetric feed structure. S_{21} of the circuit can be calculated from the $ABCD$ matrices and is expressed as

$$S_{21} = \frac{2Z_L}{S} \tag{4}$$



(a)



(b)

Figure 2. Simulated frequency responses for (a) symmetric feed structure and (b) skew-symmetric feed structure.

$$S = (A_2 + jB_2Y_0 \tan\theta_1 + jB_2Y_0 \tan\theta_2 + D_2)Z_L + B_2 + \left(jA_2Y_0 \tan\theta_1 + C_2 - B_2Y_0^2 \tan\theta_1 \tan\theta_2 + jD_2Y_0 \tan\theta_2\right) Z_L^2 \quad (4a)$$

The necessary and sufficient condition for the existence of the transmission zero is $\tan\theta_1 \approx \infty$ or $\tan\theta_2 \approx \infty$. The transmission zeros positions are $f_1 = nc/(4l_1\sqrt{\varepsilon_{eff}})$ or $f_2 = nc/(4l_2\sqrt{\varepsilon_{eff}})$. The zero frequencies correspond to the quarter-wavelength resonances in the open-circuited stubs with the lengths l_1 and l_2, respectively. However, in [9, 10], the type of electric coupling hairpin filter using skew-symmetric feed structure has two extra transmission zeros because the electrical delays of the upper path and the lower path are the same.

In the passband, when $\theta_1 + \theta_2 \approx \pi$, the transmission coefficient can then be found as

$$S_{21} = \frac{-1}{2 - j\dfrac{\cos^2\theta_1}{Z_L \omega c_s}} \tag{5}$$

Compared with (3), the passband transmission response of this skew-symmetric feed structure has only a 180-degree phase difference (Fig. 3). The transmission coefficient of the electric coupling hairpin filter using skew-symmetric feed structure is

$$S'_{21} = \frac{-1}{2 - j\dfrac{\cos^2\theta_1}{2Z_L \omega c_s}} \tag{6}$$

Inspecting (5) and (6), when θ_1 and c_s keep constant, $|S'_{21}| > |S_{21}|$. In other words, the electric coupling hairpin filter has lower insertion loss in the passband.

Fig. 2(b) shows the simulated results for the transmission line resonators in Fig. 1(b). The filter was designed at the fundamental frequency of 2.4 GHz and fabricated on an substrate with a thickness 25 mil and a relative dielectric constant $\varepsilon = 9.6$. The location of

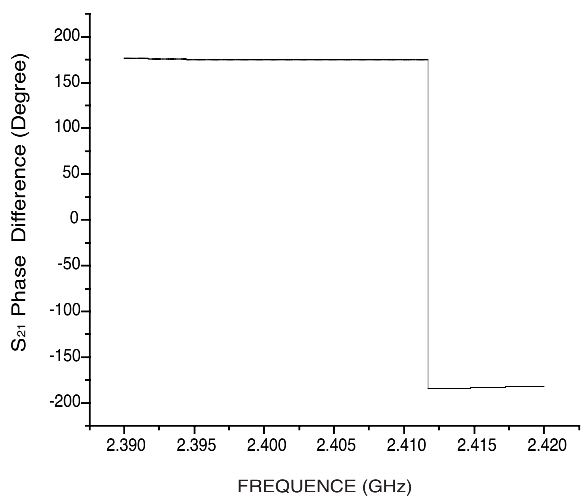


Figure 3. Difference of simulated phase responses.

transmission zeros can be calculated, $f_1 = 2.26\,\mathrm{GHz}$, $f_2 = 2.59\,\mathrm{GHz}$. Inspecting the result, the simulation agree well with the calculations.

2.2. The Parallel Microstrip Line Filters with Skew-symmetric and Symmetric Feed Structure

It is well known that the stripline parallel coupling structure with open circuits at the skew-symmetric ports has a transmission zero at the frequency when the electrical lengths of coupled lines are equal to π. This transmission zero is rarely used because its frequence is usually higher than the first spurious modes of open-line planar filters. However, it has been found that the position of the transmission zero of this type of coupling structures could vary when the coupling structure is inhomogeneous or the loads at the skew-symmetric ports are changed. Under these circumstances, the created transmission zeros may occurs at lower frequencies, and these the coupling structures can be applied to design the low-loss and high stopband rejection filters.

The parallel coupling structure with open circuits at the symmetric ports is shown in Fig. 4. This kind of structure is widely used to provide the necessary coupling between resonators of transmission-line filters. However, its transmission-zero condition is not yet fully studied, especially for inhomogeneous coupled lines. The circuit is a symmetric two-port network and can be analyzed by the classical method of even- and odd-mode excitations [15]. The even-

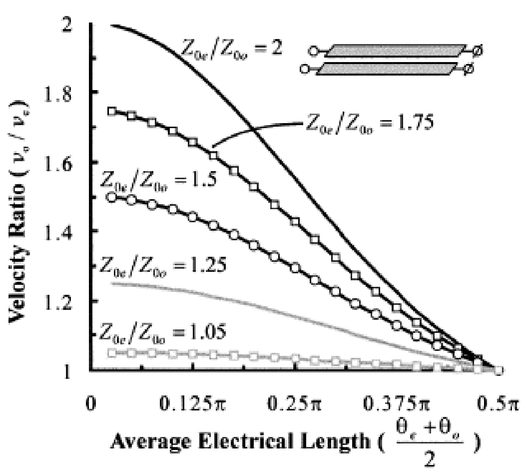


Figure 4. Transmission-zero conditions for coupled lines with open circuits at one end.

and odd-mode input impedances are given as

$$Z_{ino} = jZ_{0o} \cot\theta_o \tag{7}$$

$$Z_{ine} = jZ_{0e} \cot\theta_e \tag{8}$$

Where Z_{0o} and Z_{0e} are the characteristic impedances of the coupled lines, θ_o and θ_e are the electrical lengths for even- and odd-mode excitations, respectively. S_{21} could be found by the superposition of the even and odd modes and is given by

$$S_{21} = \frac{Z_L(Z_{ine} - Z_{ino})}{(Z_{ine} + Z_L)(Z_{ino} + Z_L)} \tag{9}$$

The equation for the transmission zero is reduced to

$$Z_{0e} \cot\theta_e = Z_{0e} \cot\theta_o \tag{10}$$

For coupled microstrip lines used in filter design, however, the created transmission zero would be at a frequency when $(\theta_o + \theta_e)/2 < \pi/2$, because Z_{0e} and θ_e are larger than Z_{0o} and θ_o, respectively. The transmission zero occurs no longer at the frequency when the average electrical length is equal to $\pi/2$. Fig. 4 is drawn based on solving (10) numerically and shows the relations between the impedance ratio Z_{0e}/Z_{0o}, the odd and even-mode velocity ratio v_o/v_e, and the average electrical length $\theta = (\theta_o + \theta_e)/2$ of coupled microstrip lines when a transmission zero is created. It is clear that, for a given impedance ratio, the velocity ratio must be less than a certain value if a transmission zero is desired. Moreover, this maximum value of the velocity ratio will be reduced if the impedance ratio is decreased. In the circumstance that the impedance ratio is fixed, the transmission-zero position can be tuned to be at a lower frequency by increasing the velocity ratio. Furthermore, when the velocity ratio is fixed, the lower the impedance ratio, the shorter the average electrical length for a transmission zero. In other words, the created transmission zero will move to a lower frequency [15].

3. FILTERS DESIGN AND MEASUREMENT

To demonstrate the usefulness of the proposed filter structures, some two-pole filters are designed.

3.1. A Two-pole Hexagonal Magnetic Coupling Filter

Assuming that the filter has a fractional bandwidth of 3% at 2.45 GHz, an in-band return loss of $R > 20\,\text{dB}$ and a minimum out-band loss of

$L > 25$ dB in the upper stopband, the photograph of the fabricated filter is shown in Fig. 5. The width of the microstrip of the resonator is 1.5 mm, and the feed line 0.56 mm. The substrate used in the simulation has a relative dielectric constant of 9.6 and a thickness of 20 mil. The transmission zeros are obtained near the passband at $f_1 = 2.63$ GHz and $f_2 = 2.97$ GHz. There is a coincidence between the simulated and measured results (Fig. 6). It is obvious that the asymmetrical fed magnetic coupling filters are preferable designs if high rejection is needed only in the upper stopband. Fig. 6 shows the two transmission zeros can be generated in the upper stopband. The principle is that a tapped half-wavelength resonator forms one open stub and parallel coupling with loads at skew-symmetric ports.

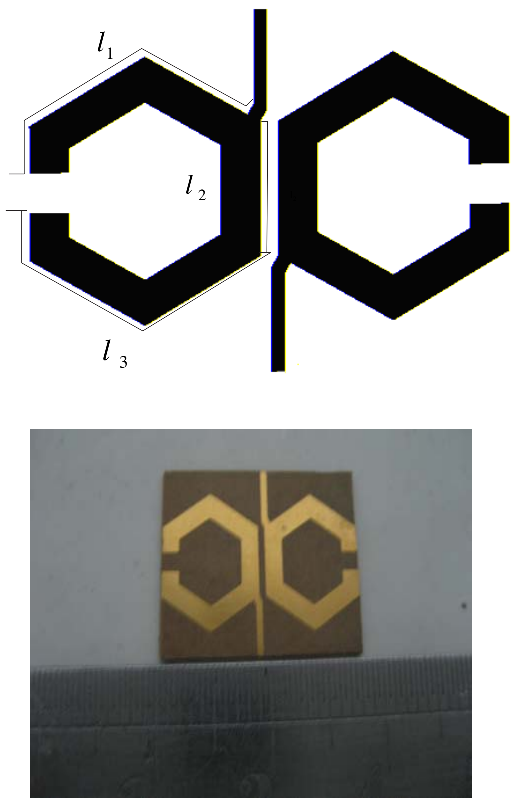


Figure 5. Layout and photograph of the hexagonal filter of magnetic coupling.

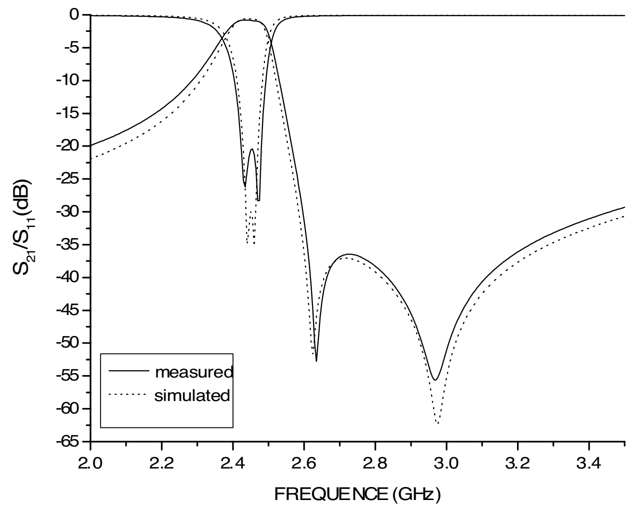


Figure 6. The simulated and measured results.

3.2. A Two-pole Asymmetric Feed Microstrip Filter

Assuming that the filter has a fractional bandwidth of 2% at 2.4 GHz, an in-band return loss of 15 dB and a minimum out-band loss of 25 dB in the lower stopband, the filter is fabricated on an substrate with a thickness 25 mil and a relative dielectric constant $\varepsilon_r = 9.6$. Fig. 7 shows the configuration of the filter. $l_1 = 10.8$ mm, $l_2 = 13.6$ mm, and $l_3 = 10$ mm. Because of asymmetric feed structure, it can obtain two transmission zeros, the location of transmission zeros can be calculated, $f_1 = 2.15$ GHz, $f_2 = 2.68$ GHz. The third transmission zero is created for coupled microstrip lines (l_3). The transmission zero is varied with the variation of the coupled gap. The photograph of the fabricated filter is shown in Fig. 7. From the measured data, it is also found that there are three transmission zeros in the stopband response of the filter. One of them is at 2.27 GHz, another is at 2.67 GHz and the third is at 1.74 GHz (Fig. 8).

3.3. A Two-pole Asymmetrical Fed Filter with Cross-coupling

Assuming that the filter has a fractional bandwidth of 2% at 0.94 GHz, the filter is fabricated on an substrate with a thickness 50 mil and a relative dielectric constant $\varepsilon_r = 10.8$ (Fig. 9). The width of the microstrip of the resonator is 1.4 mm, and the feed line 1.1 mm.

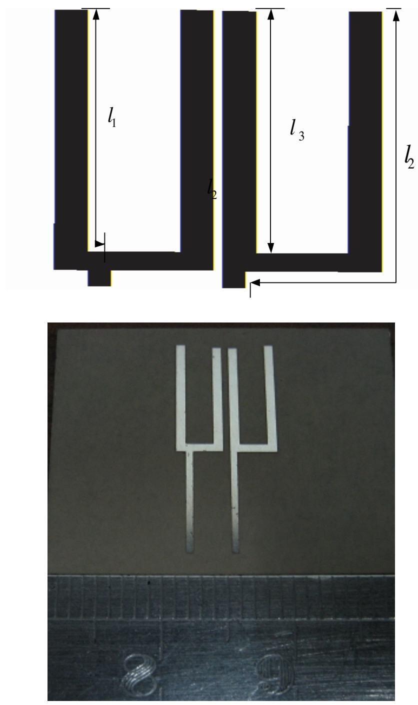


Figure 7. Layout and photograph of two-pole filter.

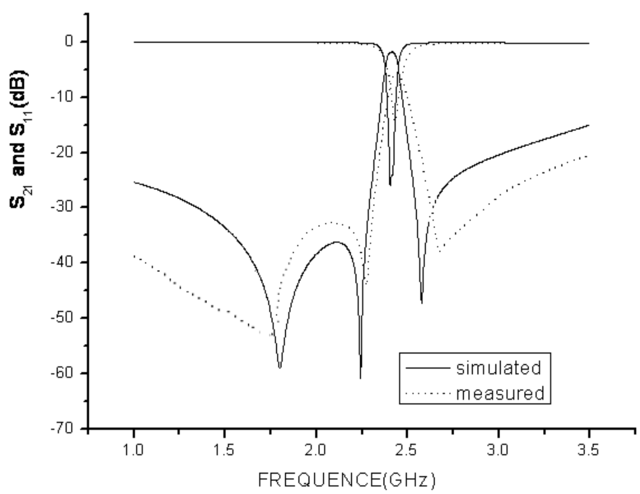


Figure 8. Response of filter.

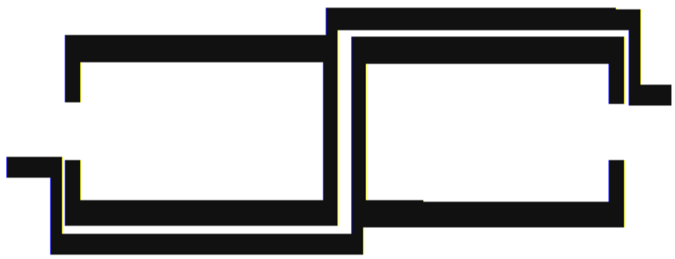


Figure 9. Layout of two-pole filter.

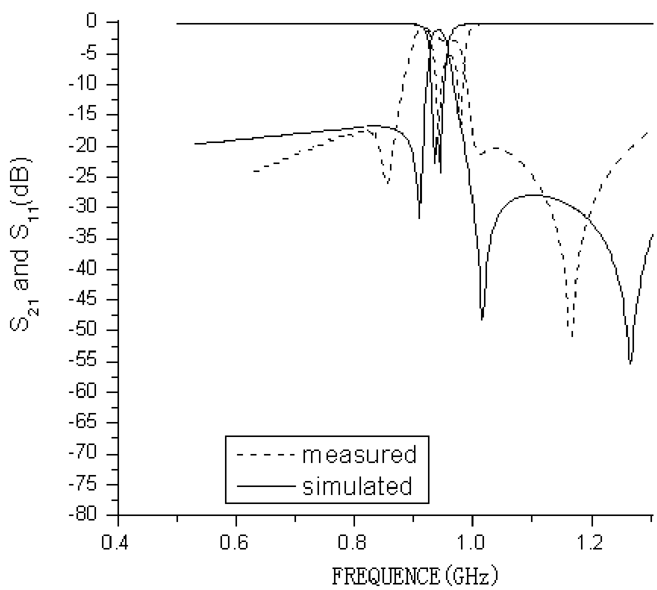


Figure 10. Response of filter.

Because of asymmetric feed structure and cross-coupling, it can obtain three transmission zeros. From the measured data, the location of transmission zeros, $f_1 = 0.86\,\text{GHz}$, $f_2 = 1.01\,\text{GHz}$. $f_3 = 1.16\,\text{GHz}$. The principle is that a tapped half-wavelength resonator forms one open stub, parallel coupling with skew-symmetric feed structure and coupling the source(or the load) to the last (or the first) two resonators. There is a coincidence between the simulated and measured results (Fig. 10).

4. CONCLUSION

Compared with conventional hairpin filter, the proposed filters in the paper can obtain more transmission zeros and the maximum of attenuation in the stopband of transmission response by utilizing open-circuited stubs, parallel coupled-lines and cross-coupling. It is good agreement between the measured and simulation results.

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