Bandpass Filters Designed by Transmission Zero Resonator Pairs With Proximity Coupling

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Abstract—The concept of transmission zero resonator pair (TZRP) is utilized in this paper. Based on this TZRP, a new method to induce a passband is demonstrated, by which different types of bandpass filters can be designed. The proposed TZRP is structured by a pair of resonators with different resonant frequencies, which lead not only to two transmission zeroes, but also to a transmission pole between them. Passband filters can then be built by designing the proximity coupling between the TZRPs, as the TZRP works as a basic resonant element. The passband of these filters can be flexibly controlled by the resonators of TZRPs, which determine the locations of the transmission zeroes and poles. By carefully allocating the transmission zeroes, high selectivity and large out-of-band rejection can be realized. This design method is applied to design filters in two different transmission media, namely, microstrip line and rectangular waveguide. Simulated and measured results demonstrate the effectiveness of this new approach of bandpass filter design. The designed filters have the properties of small size, easy fabrication, low cost, and low loss.

Index Terms—Bandpass filter, microwave filter, transmission zero resonator pairs (TZRPs), transmission zeroes, waveguide filter.

I. INTRODUCTION

Bandpass filters are generally designed by various kinds of resonators [1]–[3], which induce transmission poles at their resonances in the desired passbands. The generating mechanism of transmission pole is mostly based on the resonances of single-mode, dual-mode, or multimode resonators [4], [5]. On the other hand, bandstop filters [6]–[8] can be realized by designing resonators which produce transmission zeroes at their resonant frequencies. Besides being used for designing bandstop filters, transmission zeroes are widely used to improve the out-of-band rejection performance of bandpass filters [9]–[12]. They are realized by cross-coupling technique [13], mixed electric/magnetic coupling [14], multipath effect, and source/load couplings. The method to design bandpass filters by resonators which directly bring transmission zeroes at their resonance was proposed in [15] and [16] using dual-behavior resonators (DBRs).

However, three sections of quarter-wavelength inverters were inevitable for the second-order filters, resulting in large sizes. In addition, straight and parallel arrangement of the resonators could not achieve size reduction. The DBRs were further designed in [17] and [18]. Similar to those in [15] and [16], the resonators were separated by quarter-wavelength inverters for high-order designs. Efforts have been made to reduce the size of this kind of filter by absorbing two equivalent capacitances of the inverter into neighboring resonators [19]. However, because wavelength-related inverters were used, the above-designed bandpass filters had the penalty of narrow bandwidth (less than 10%), which would limit its applications.

In this paper, proximity coupling instead of wavelength-related inverters will be utilized for a higher-order bandpass filter design, which greatly reduces filter sizes. Moreover, the bandwidth can be flexibly designed by controlling the locations of the two lower or upper transmission zeroes. The concept of transmission zero resonator pair (TZRP) is first used instead of the DBR. Because the transmission pole and passband performance are primarily determined by the two transmission zeroes, the TZRP can emphasize the important role of the two transmission zeroes induced by a pair of resonators with different resonant frequencies. Then, on the one hand, two transmission zeroes can be generated as expected. On the other hand, a transmission pole can also be produced by this TZRP. This is a very special feature and can be used as a basic resonant element for designing a passband. With this TZRP concept, a category of bandpass filter design approach comes out. When coupling two TZRPs, rather than using quarter-wavelength inverters, compact bandpass filters with two transmission poles and four transmission zeroes can be easily obtained. Two kinds of bandpass filters are introduced to demonstrate this new design method in two most popular transmission media, namely, microstrip line and rectangular waveguide. The measured results show very good performance for both microstrip and waveguide bandpass filters. The two-pair design can be easily extended to the design of multiple pairs for dual-band bandpass filters [20].

II. CIRCUIT THEORY

For a systematic statement of the proposed method, we start from the basic element circuit. First, an LC resonator is considered as shown in Fig. 1. A transmission zero is induced by this series LC resonator. This transmission zero is located at the LC resonant frequency, which is 2.05 GHz (with \( L = 2 \) nH and \( C = 3 \) pF).

Then, for a pair of series LC resonators, as shown in Fig. 2(a), which are denoted by \( R1 \) consisting of \( L1 \) and \( C1 \)
and \( R_2 \) consisting of \( L_2 \) and \( C_2 \), two transmission zeroes are definitely induced at their resonant frequencies, and denoted by \( \omega_{z1} \) and \( \omega_{z2} \), respectively. So, this pair of resonators can be called TZRP, which is constructed by two series \( LC \) resonators with different resonant frequencies. The two transmission zeroes are

\[
\omega_{z1}^2 = \frac{1}{L_1C_1} \quad (1) \\
\omega_{z2}^2 = \frac{1}{L_2C_2}. \quad (2)
\]

\( R_1 \) behaves as an inductor when \( \omega > \omega_{z1} \), while it equals a capacitor when \( \omega < \omega_{z1} \). Similarly, for \( R_2 \), it behaves as an inductor when \( \omega > \omega_{z2} \), while it equals a capacitor when \( \omega < \omega_{z2} \). Assuming \( \omega_{z2} > \omega_{z1} \), for a frequency \( \omega (\omega_{z1} < \omega < \omega_{z2}) \), \( R_1 \) is an equivalent inductor, which is denoted by \( L_e \), as shown in Fig. 2(b). \( L_e \) is calculated by

\[
L_e = L_1 - \frac{1}{\omega^2C_1}. \quad (3)
\]

For \( R_2 \), it behaves as a capacitor denoted by \( C_e \), which is

\[
C_e = \frac{C_2}{1 - \omega^2L_2C_2}. \quad (4)
\]

Then, \( L_e \) from \( R_1 \) and \( C_e \) from \( R_2 \) characterize a parallel \( LC \) resonator denoted by \( R_e \), which would lead to a transmission pole at its resonant frequency. The transmission pole is determined by

\[
\omega_p^2 = \frac{1}{L_eC_e} = \frac{1}{\left(\frac{1}{L_1} - \frac{1}{\omega^2C_1}\right)\left(\frac{1}{L_2} - \frac{1}{\omega^2L_2C_2}\right)}. \quad (5)
\]

It can be seen from (5) that a transmission pole \( \omega_p \) can then be produced as

\[
\omega_p^2 = \frac{1 + \frac{C_2}{L_2C_2}}{L_1C_2}. \quad (6)
\]

For simplicity of description, a cross-resonance frequency is defined as

\[
\omega_{\text{z1}}^2 = \frac{1}{L_2C_1}. \quad (7)
\]

By substituting (1), (2), and (7) into (6), the following can be obtained:

\[
\omega_p^2 = \frac{\omega_{z1}^2\omega_{z2}^2}{\omega_{z1}^2 + \omega_{z2}^2}. \quad (8)
\]

Equation (8) shows that the newly induced transmission pole between the transmission zeroes \( \omega_{z1} \) and \( \omega_{z2} \) of the TZRP is determined by \( \omega_{z1} \), \( \omega_{z2} \), and \( \omega_{z1} \). So the bandpass performance of a TZPR is defined by its two resonators with transmission zeroes. An example is given choosing \( L_1 = 2 \text{ nH}, C_1 = 3 \text{ pF}, L_2 = 1 \text{ nH}, \) and \( C_2 = 3 \text{ pF} \). It can be seen that a TZRP can not only give two transmission zeroes but also a transmission pole between the two transmission zeroes, as shown in Fig. 3.

It is concluded from the above analysis that a TZRP consisting of two resonators with different resonating frequencies obtains three resonating points, which lead to a transmission pole \( \omega_p \) and two transmission zeroes \( \omega_{z1} \) and \( \omega_{z2} \). The TZRP itself can be viewed as a resonator as well, so that this TZRP, like other conventional resonators, can be employed to build passband filters. Unlike conventional filters, no extra measure is needed in this TZRP bandpass filter to generate transmission zeroes because the TZRP has inherent transmission zeroes. The bandpass filters are designed by coupling two identical TZRs and thus are equipped with two transmission poles and four transmission zeroes. Mixed couplings between the TZRs are modeled as presented in Fig. 4(a), which include magnetic coupling referred to as \( L_p \) and electric couplings \( C_{p1} \) and \( C_{p2} \) for \( R_1 (R_1') \) and \( R_2 (R_2') \), respectively. Adjusting the equivalent coupling parameters \( L_p, C_{p1}, \) and \( C_{p2} \), the performances of the bandpass filter are determined. For example, based on the TZRP in Fig. 3,
Fig. 4. Designed passband filter by TZRPs and the coupling between them. (a) Schematic. (b) Simulation results of an example versus $L_p$ with $C_{p1} = 0.1 \text{ pF}$ and $C_{p2} = 0.1 \text{ pF}$. (c) Simulation results versus $C_{p1}$ with $L_p = 1.2 \text{ nH}$ and $C_{p2} = 0.1 \text{ pF}$. (d) Simulation results versus $C_{p2}$ with $L_p = 1.2 \text{ nH}$ and $C_{p1} = 0.1 \text{ pF}$.

given transmission zeroes of $R1$ and $R2$, the couplings are critical for the features of the filter as well. Changing the magnetic coupling between the TZRPs affects the in-band transmission performance as shown in Fig. 4(b). The two transmission poles get closer and the bandwidth gets narrower when the magnetic coupling $L_p$ is increased, while the four transmission zeroes remain the same. Changing the electric coupling $C_{p1}$ between $R1$ and $R1'$ in Fig. 4(c), the two lower transmission zeroes are allocated flexibly, while the two upper transmission zeroes are kept unchanged. The lower cutoff frequency is pushed toward higher frequencies when increasing $C_{p1}$. Similarly, the two upper transmission zeroes can be determined by $C_{p2}$, while the two lower transmission zeroes remain unchanged. The upper cutoff frequency is pushed toward lower frequencies when increasing $C_{p2}$. Actually, the TZRPs and proximity couplings can be designed flexibly in the real implementation circuits. In the following section, this will be demonstrated.

III. MICROSTRIP BANDPASS FILTER

As a typical transmission medium, the microstrip line has been widely used to design various components. Moreover, various filters based on microstrip line [21], [22] have been presented. As an example, microstrip line is used to implement a TZRP bandpass filter.

A. Design of Microstrip TZRP

As shown in Fig. 5, a $\lambda_{g}/4$ microstrip open-circuited stub resonator (with a width of 0.5 mm) contributes a transmission zero. A larger $l$ results in a lower frequency of transmission zero. By folding the resonator, a much smaller size is achieved. So in the following design, the microstrip resonators are folded for compact filters.

In order to design a TZRP, another resonator, with a different resonant frequency, is required. Here, two folded $\lambda_{g}/4$ microstrip open-circuited resonators are used as shown in Fig. 6 for simple structures. Two transmission zeroes and a transmission pole are obtained by this microstrip TZRP. The transmission pole is between the two transmission zeroes.

B. Design of Microstrip Filter

With this designed TZRP, a passband filter is designed easily. As illustrated in Fig. 7(a), two identical TZRP’s are placed face-to-face, i.e., the $R1'$ and $R2'$ are mirrored by $R1$ and $R2$,
respectively. The proximity coupling between the two TZRPs is implemented by a section of microstrip line and the gaps ($g_{11}$ and $g_{22}$) between the TZRPs according to the schematic in Fig. 6(a). A TZRP offers a transmission pole and two transmission zeroes, so two TZRPs, coupled by $L_p$, $C_{p1}$, and $C_{p2}$, will bring two transmission poles and four transmission zeroes. Simulation results of a sample are given in Fig. 7(b), in which two transmission poles and four transmission zeroes are observed clearly. The four transmission zeroes are located on both sides of the passband, which improve the out-of-band rejection performance and the selectivity of passband.

In this designed filter, despite the multiple transmission zeroes are obtained on both sides of the passband, another advantage is that the positions of two lower transmission zeroes and two upper transmission zeroes can be tuned individually. This property brings much flexibility to design the bandpass filter at the desired center frequency with a desired bandwidth.

The transmission zeroes are determined by the TZRPs and their proximity couplings. In this section, the $\lambda_g/4$ microstrip resonators of a TZRP are explored to adjust the performance of filters. In more detail, the frequencies of two lower transmission zeroes can be adjusted by tuning the overall length of $R_1 (R_1')$. It can be seen from Fig. 8(a) (changing $l_{11}$ as an example) that when the length of $R_1 (R_1')$ increases, the two lower transmission zeroes are decreased as expected and so does the lower cutoff frequency. In the meantime, the upper two transmission zeroes and the upper cutoff frequency keep unchanged. Therefore, the bandwidth of the designed filter becomes narrower as overall length of $R_1 (R_1')$ decreasing.

The positions of two upper transmission zeroes can be easily adjusted by tuning the overall length of $R_2 (R_2')$. By changing the length of $R_2 (R_2')$, the upper two transmission zeroes and the upper cutoff frequency are tuned while the lower ones are kept the same, as shown in Fig. 8(b). As the length of $R_2 (R_2')$ increases, the two upper transmission zeroes and the upper cutoff frequency shift down. So the lower and upper cutoff frequencies can be determined individually. With this flexible mechanism of allocating lower and upper transmission zeroes, the passband of the filter can be easily controlled. As examples, two filters operating at the same center frequency of 2 GHz but with 3-dB bandwidths of 7% and 26% are demonstrated in Fig. 8(c). Considering all the losses of metal and substrate, the simulated loss for narrowband bandpass filter is 1 dB.

C. Experiment and Comparison

For verification, a bandpass filter at 1.8 GHz is designed, fabricated, and measured. The Duriod 6010 substrate, with a thickness of 1.27 mm and relative permittivity of $\varepsilon_r = 10.2$, is used. The photograph is given in Fig. 9, and the fabricated filter is 8.2 mm $\times$ 11.8 mm except the input and output length. The input and output ports are 50 $\Omega$, which are connected to the SMA connectors for test. This filter has been measured by Agilent Network Analyzer E5071C. Comparisons of the simulation and measurement results are presented in Fig. 10. Despite of a slightly frequency shift, the measured results agree well with the simulation ones. Four measured transmission zeroes located at 1.19, 1.44, 1.94, and 2.52 GHz are obtained on both sides of the passband. Large rejection and high selectivity are realized. The measured minimum insertion loss is 0.7 dB (including the losses of two SMA connectors) in the passband. The tested filter operates at 1.72 GHz with a 3-dB bandwidth of 16.3% (0.28 GHz), while the simulated one is at 1.8 GHz with a 3-dB bandwidth of 15.5%. A 0.08-GHz frequency shift is observed in the experiment due to the fabrication and assembly errors. The comparison with the filters designed by DBRs, which induce transmission zeroes as well, is given in Table I. Due to the wavelength-related invertors, the bandwidths in [15] and [16] were limited. Wideband bandpass
A filter with compact size and low insertion loss is realized by this proposed method.

IV. WAVEGUIDE E-PLANE BANDPASS FILTER

In this section, the waveguide filter would be designed based on the concept of the TZRP.

A. Design of the Waveguide TZRP

The standard waveguide equals a kind of transmission line and the transmission zeroes can be easily induced by a metal strip at the center E-plane of the waveguide, as shown in Fig. 11. This transmission zero is obtained from the resonance of the strip resonator which is about half waveguide wavelength. The strip can be folded for generating a lower transmission zeroes and possessing more design freedom. As shown in Fig. 12, two folded strips (end-folded and center-folded strips) with different lengths are placed at the top and bottom sides of the E-plane substrate to form a waveguide TZRP. Transmission zeroes of the two folded strips are at 29.5 and 37.4 GHz, respectively. A transmission pole at 32.8 GHz is brought by this waveguide TZRP.
B. Design of the Waveguide TZRP Bandpass Filter

Once waveguide TZRP is obtained, it’s easy to design a waveguide bandpass filter by putting two identical TZRPs face-to-face for proximity coupling. A waveguide TZRP offers a transmission pole and two transmission zeroes, so two coupled waveguide TZRPs, as shown in Fig. 13, will bring two transmission poles and four transmission zeroes. The coupling strength between these two TZRPs can be adjusted flexibly by the gaps between them or even inserting metal patches in between.

The merit of individually controllable mechanism of transmission zeroes is inherited from this proposed bandpass method. The lengths of resonators of waveguide TZRP play important roles in determining the locations of transmission zeroes and bandwidth of passband as the microstrip TZRP. In the microstrip filter design, the transmission zeroes and poles, tuned by changing the lengths of resonators, are discussed in Fig. 8. In this waveguide TZRP bandpass filter, the design of proximity couplings between the two TZRPs is focused. So the gaps \(g_1\) and \(g_2\) are designed for illustrating.

The gap \(g_1\) is used to adjust coupling between two end-folded resonators of this TZRP, so the locations of the two lower transmission zeroes are determined as shown in Fig. 14. From the simulation results, it can be seen that stronger coupling obtained when the gap \(g_1\) is smaller. The selectivity is improved as \(g_1\) decreases as well. The lower out-of-band rejection range is enlarged while the rejection level becomes higher. Meanwhile, the upper two transmission zeroes and the upper cutoff frequency on the upper side of the passband are kept constant.

Similarly, the gap \(g_2\) can change the coupling between two center-folded strips, so the frequencies of the two upper transmission zeroes are determined as shown in Fig. 15. Form the simulation results, it can be seen that stronger coupling obtained when the gap \(g_1\) is smaller. The selectivity is improved as \(g_1\) decreases as well. The lower out-of-band rejection range is enlarged while the rejection level becomes higher. Meanwhile, the upper two transmission zeroes and the upper cutoff frequency on the upper side of the passband are kept constant.

Similarly, the gap \(g_2\) can change the coupling between two center-folded strips, so the frequencies of the two upper transmission zeroes are determined as shown in Fig. 15. Increasing \(g_2\) reduces the coupling, so the two transmission zeroes get closer to each other. The 3-dB bandwidth can be altered, because the upper cutoff frequency decreases as the coupling is stronger. At the same time, the lower two transmission zeroes and the lower cutoff frequency keep unchanged. So the two lower transmission zeroes or the upper transmission zeroes
can be controlled individually and flexibly. The transmission zeroes contribute to high selectivity and large out-of-band rejection. Moreover, the locations of transmission zeroes can be designed at any frequencies, so the operating frequency and bandwidth of the presented filter can be adjusted flexibly as well. Finally, the length parameters can be used to design the frequencies of transmission zeroes as well. They are not discussed here.

C. Experiment and Comparison

An E-plane waveguide filter operating at 33 GHz with absolute bandwidth of 2 GHz is designed and fabricated. The Duriod 5880 substrate with a thickness of 0.254 mm and relative permittivity of $\varepsilon_r = 2.2$ is used. Two TZRPs, consisting of four folded strips with two of them on the top side and the other two and small metal patches at the bottom of the substrate, are fabricated. Then the substrate is inserted into the central E-plane of the standard waveguide WR-28. The parameters of experimented filter are as follows: $l_1 = 3.2$ mm, $l_2 = 0.45$ mm, $l_3 = 0.5$ mm, $l_4 = 2$ mm, $l_5 = 0.1$ mm, $l_6 = 0.6$ mm, $l_7 = 0.7$ mm, $g_1 = 0.1$ mm, $g_2 = 0.15$ mm, $g_3 = 1.5$ mm, and $w_1 = w_2 = 0.1$ mm. The size along waveguide transmission direction equals $l_6 + 2(g_2 + w_2 + l_5) = 1.3$ mm. Although the fabricated filter is only 1.3 mm in length of a waveguide, the standard WR-28 and the substrate with a length of 20 mm are used for easy resembling and measurement. The photograph is given in Fig. 16.

![Fig. 16. Photograph of the broadband filter.](image)

![Fig. 17. Comparisons of simulation and measurement results.](image)

This filter has been measured by Agilent Network Analyzer N5247A. Comparisons of the simulation and measurement results are presented in Fig. 17. Four transmission zeroes located at 27.7, 30.2, 35.5, and 37.6 GHz are obtained on both sides of the passband. Large rejection and high selectivity are realized. The measured minimum insertion loss is 0.95 dB in the passband. A good agreement between the simulation and measurement results is achieved. Some comparisons are given in Table II. It can be seen that our designed filter has the largest number of transmission zeroes and realizes compact size with comparable insertion losses.

V. Conclusion

In this paper, a new method to design a kind of bandpass filter is proposed. The concept of TZRP is presented, which can bring a transmission pole and two transmission zeroes. Then the proximity coupling between two TZRPs is designed for a second-order passband filter realizing two transmission poles in the passband and four transmission zeroes outside the passband. Based on this design method, two basic filter configurations are discussed in detail, i.e., microstrip bandpass filter and waveguide bandpass filter. A 1.8-GHz microstrip bandpass filter and a 33.1-GHz waveguide bandpass filter were fabricated and tested. Both the designed filters have the properties of compact size and low insertion loss. Moreover, the bandwidth of these proposed filters can be adjusted flexibly. Simulation and measurement results demonstrated the effectiveness of this new approach of bandpass filter design. This filter using TZRPs has the merits of simplicity in structure and design, compact size (only a small fraction of a wavelength), high out-of-band rejection, low and flat passband loss, and high flexibility in tuning the passband. This design method can be used to design multiband bandpass filter by properly designing the locations of transmission zeroes and transmission poles.

### TABLE II

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<td>1.3mm ($0.11\lambda_d$)</td>
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* IL, FBW, No. of TZs and $\lambda_d$ denote the Minimum Insertion Loss, 3-dB Fractional Bandwidth, Number of Transmission zeroes, and waveguide wavelength of central frequency $f_0$ with vacuum filled in corresponding standard waveguides (WR-90 or WR-28).

[23] * means the filter III of [23], with the minimum total insert length.

### REFERENCES


Jun Ye Jin (S’11) was born in Inner Mongolia, China. She received the B.S., M.S., and Ph.D. degrees in electronic engineering from the University of Electronic Science and Technology of China (UESTC), Chengdu, China, in 1988, 1991, and 1993, respectively.

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