Open-Loop Dual-Mode Microstrip Filters

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***Abstract* — This paper presents the design of compact second-order bandpass filters based on dual-mode open-loop resonator. A filter design procedure is provided to facilitate the design process. The paper also describes the nature of the inherent transmission zero associated with the structure and presents a method of generating two additional zeros for improving stop-band performance. Finally, a filter design example is presented to validate the argument.**

***Index Terms* — Dual-mode filters, microstrip filters, microstrip resonators.**

I. Introduction

MICROSTRIP filters are popular due to size, cost, weight, and fabrication factors and find extensive applications in low to medium power RF transceivers. High performance bandpass filters having a low insertion loss, compact size, wide stop-band and high selectivity are important for modern communication systems [1].

Filters based on the single mode open-loop resonator (OLR) such as in [2] and [3] focus only on the odd mode resonance. Although an even mode resonance is present, this is approximately at twice the fundamental resonant frequency and therefore is of little use in single band filter synthesis. Consequently, the even mode will appear as the first spurious harmonic, which degrades the filter response. Dual-mode filters also make use of the even-mode and therefore behave as a doubly tuned circuit. These filters are not only more compact but also offer significantly less insertion loss.

This paper presents a novel design procedure for the second order dual-mode open-loop filter. Design equations are provided to facilitate filter development for second order filters and finally stop-band improvements are suggested and demonstrated through an example.

II. Dual-Mode Open-Loop Filters

An open-loop filter may be modified to behave as a doubly-tuned filter. The structure proposed in [4] illustrates that the even mode resonance can be lowered to operate closer to the odd mode. The availability of two poles generates a second order response. The layout of the dual-mode filter is illustrated in Fig. 1. (a).

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The additional open-circuited stub placed in the centre of the filter lowers the even mode resonant frequency. The extension shown has no effect on the odd mode [4] so the two modes may be independently adjusted.

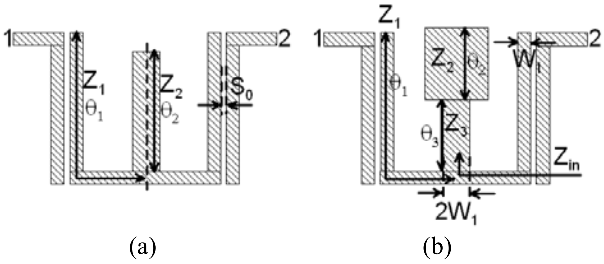

Fig. 1. (a) Dual-mode resonator (b) Stepped-impedance resonator.

The even mode resonator is an open-circuited half wavelength type resonator while the odd mode is a short circuited quarter wavelength resonator. While the even mode resonator is elongated by θ_2, the added equivalent stub impedance is not equal to Z_2 since the effective width of the added element is halved due to the virtual open-circuit in the symmetry plane as indicated by the dashed line in Fig. 1. (a).

One design approach is to first select a suitable open loop structure. Then the initial open loop dimensions are determined to obtain the desired odd mode resonance. Lastly, the dimensions of the open-circuited extension are determined for positioning the second filter transmission pole. The simplest case of dual-mode resonance can be illustrated by using the uniform impedance case where $Z_1 = \alpha Z_2$. This equality can be approximated by setting the line width of the open stub to be twice the resonator line width.

As a design example, for the substrate parameters of $h = 1.575\,\text{mm}$ and $\varepsilon_r = 2.2$ where $\varepsilon_{\text{eff}} = 1.85$, the specification is to have a 2$^{\text{nd}}$ order filter with a bandwidth of $200\,\text{MHz}$ and centre frequency of $2\,\text{GHz}$. It is desired to place poles at $1.95\,\text{GHz}$ and $2.05\,\text{GHz}$ to achieve desired pass-band ripple.

Dimensions were calculated using equations

$$l_1 \cong \frac{c}{4 f_{\text{odd}} \sqrt{\varepsilon_{\text{eff}}}} \tag{1}$$

$$l_2 \cong \frac{c}{2 f_{\text{even}} \sqrt{\varepsilon_{\text{eff}}}} - l_1 \tag{2}$$

where l_x ($x = 1,2$) corresponds to length of section θ_x in Fig. 1. (a) and c is the speed of light in vacuum. Then a suitable coupling gap spacing S_0 was chosen to be $0.4\,\text{mm}$ in order to obtain desired Q-factor. The full-wave EM simulation results are illustrated in Fig. 2., while Fig. 3. details the filter layout. The response may be fine tuned to match the exact design specification.

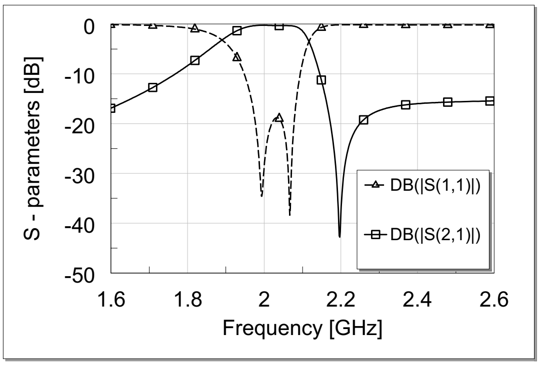

Fig. 2. Response of preliminary design.

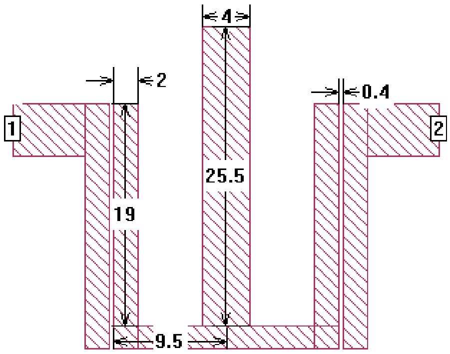

Fig. 3. Layout and dimensions of designed filter in millimeters.

III. Stepped-Impedance Filters

It may be desirable to use stepped impedances to reduce the length of the open circuited stub employed to achieve dual-mode performance [5]. The open circuited stub will consist of two sections of different impedances as depicted in Fig. 1. (b). The first section (connecting the stub to the rest of the resonator) will as before be designed such that it is twice as wide as W_1. This ensures that the characteristic impedance of the line at even mode resonance is approximately Z_1. The open-ended section however will usually be much wider and will have lower characteristic impedance Z_2.

Let αZ_2 and βZ_3 represent the even mode equivalent impedances of the sections with impedance Z_2 and Z_3. Let $R = \beta Z_3 / \alpha Z_2$ so $R > 1$ for the stepped impedance case where $\beta Z_3 > \alpha Z_2$. For every value of R there exists a particular value of θ_2

$$\theta_2 = \cos^{-1}\left(\sqrt{\frac{R(R-1)}{(R^2-1)}}\right) \tag{3}$$

for which the overall length of the open circuited stub is shortest. This particular value of θ_2 (given by equation (3)) may be used to derive the physical length (l_2) of this section and this may be computed using equation

$$l_2 \cong \frac{c}{2\pi f_{\text{even}}\sqrt{\varepsilon_{\text{eff}}}} \operatorname{acos}\left(\sqrt{\frac{R(R-1)}{(R^2-1)}}\right). \tag{4}$$

The length (l_3) of the section corresponding to θ_3 may be determined from

$$l_3 \cong \left(\frac{c}{2\pi f_{\text{even}}\sqrt{\varepsilon_{\text{eff}}}}\right)\left(\pi + \operatorname{atan}\left[-R\tan(\theta_2)\right]\right) - l_1. \tag{5}$$

These equations will provide a good starting point for the design, which may then be tuned to optimize the response.

The filter response has an inherent transmission zero (TZ) as observed in Fig. 2. The TZ causes the response to be asymmetric. It can be shown that the TZ is produced as a direct result of the open-circuited stub. This stub behaves as a virtual short to ground at the zero frequency. The condition for this to occur is simply $Z_{\text{in}} = 0$ (Fig. 1. (b)) and this condition may be generally expressed as:

$$\frac{Z_3}{Z_2} = \frac{\cot(\theta_2)}{\tan(\theta_3)}. \tag{6}$$

The zero and the even-mode pole are both dependent on the dimensions of the stub. An interesting observation is that when the even mode resonant frequency falls below that of the odd, the TZ actually appears on the lower stop-band. This property enables the designer to improve selectivity of either the upper or the lower stop band.

It is possible to introduce extra zeros in addition to the inherent TZ. One approach is to degrade the effectiveness of the input/output coupling at the desired zero frequency and this may be achieved by extending input/output coupled lines with electrical length θ_4 as shown in Fig. 4.

For the case where there is no capacitive coupling between the two extensions (i.e. when S_1 is large), the TZ simply occurs at the frequency where $\theta_4 = 90^\circ$. However, when there is coupling between the two sections, the TZ will split. The even mode zero condition is still $\theta_4 = 90^\circ$. The odd mode zero condition may be approximately summarized as

$$\tan(\theta_4) = \frac{1}{4\pi f_{\text{zero}} C_0 Z_4} \tag{7}$$

where C_0 is the capacitance associated with the gap. The additional TZ obtained as a consequence of the coupling effect is extremely useful in further improving stop-band performance. It is generally less attractive to realize these zeros on the lower stop band as it requires longer line lengths.

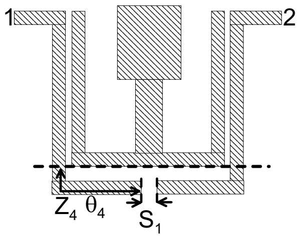

Fig. 4. Modification for producing more zeros.

IV. Filter Applications

For demonstration, a 2nd order bandpass filter was realized on a substrate of height $h = 1.575\,\text{mm}$, $\varepsilon_r = 2.2$, and $\tan\delta = 0.0012$. The dimensions of the filter are detailed in Fig. 5. The dimensions of interdigital capacitor (Fig. 5.) are: the finger width is $0.6\,\text{mm}$ and the space (between fingers) is $0.3\,\text{mm}$. The simulated *S*-parameters are shown in Fig. 6. Pass-band insertion loss is around $0.6\,\text{dB}$.

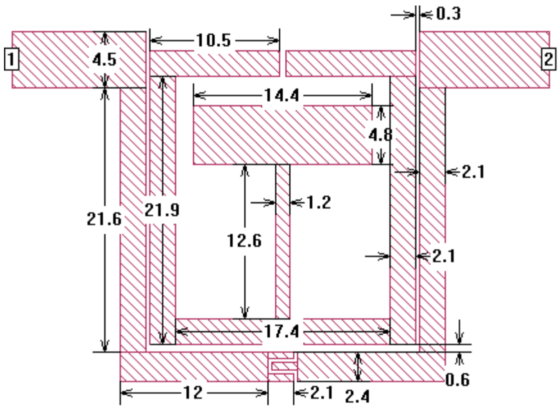

Fig. 5. Dimensions in millimeters of the realized second order filter.

V. Conclusion

The filters discussed in this paper offer excellent miniaturization. The analysis of the compact dual-mode open-loop filter presented allows for this type of filter to be designed relatively simply. The additional stop-band improvements suggested may be employed to generate sharp asymmetric filtering characteristics. It is also possible to have zeros on either side of the pass-band if desired. Moreover, these filters are an excellent candidate for applications where space is limited. Such filters may be cascaded as demonstrated in [4] to achieve higher order filtering characteristics.

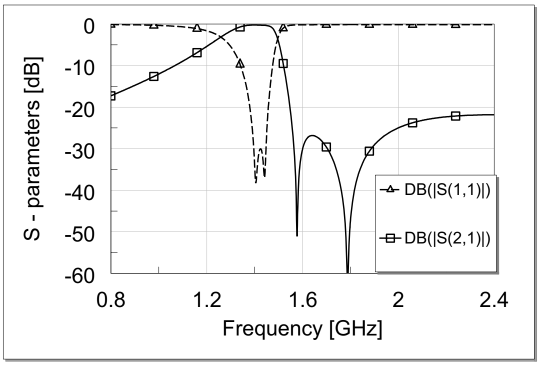

Fig. 6. Simulated *S*-Parameters of realized filter.

Acknowledgment

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References

[1] J.-S. Hong and M. J. Lancaster, *Microstrip Filters for RF/Microwave Applications*. New York: Wiley, 2001.

[2] J.-S. Hong and M. J. Lancaster, “Theory and experiment of novel microstrip slow-wave open-loop resonator filters,” *IEEE Trans. Microw. Theory Tech.*, vol. 45, no. 12, pp. 2358-2365, Dec. 1997.

[3] J.-S. Hong, M. J Lancaster, “Design of highly selective microstrip bandpass filters with a single pair of attenuation poles at finite frequencies,” *IEEE Trans. Mirow. Theory Tech.*, vol. 48, no. 7, pp. 1098-1107, Jul 2000.

[4] J.-S. Hong, H. Shaman, and Y.-H. Chun, “Dual-Mode Microstrip Open-Loop Resonators and Filters,” *IEEE Trans. Mirow. Theory Tech.*, vol. 55, no. 8, pp. 1764-1770, Aug. 2007.

[5] M. Makimoto and S. Yamashita, *Microwave Resonators and Filters for Wireless Communication: Theory, Design and Application*. New York: Springer-Verlag, 2001.