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Novel Resonators with Applying Binomial Coefficients Admittance Values to Design Multi-Frequency Bands of Microstrip Filters

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Abstract—**This paper presents the superiority of designing microstrip line Multi-frequency band filters with a single input and a single output. Utilizing the S-parameter of a two-port network and the ABCD matrix parameters for the analysis of short-circuited and open-circuited resonators with various combinations of series and shunt sequences, Multi-frequency band resonators are proposed based on the stepped-impedance resonator (SIR) structure. With the idea of the impedance ratio of the resonators, the center frequency of the two passband is determined. Using transmission line connections of suitable characteristic impedances and the resonators by the binomial coefficient admittance values, the dual-band effect is obtained. Binomial coefficient admittances of resonators are chosen for higher order dual-band filter design and simulation. From the experimental results, the bandwidth is 500MHz to the least extent, the insertion loss is less than 1dB within the passband, and more than 50dB attenuation is permitted within 1.4GHz from the center frequencies of the two passband. The method can be applied to any other multi-frequency bands, then 5.2 GHz and 2.4 GHz filter was designed as example.**

I. INTRODUCTION

In many communication applications, especially including satellite and mobile communication, keeping RF filter structures to a minimum size, weight, and low insertion loss, high selectivity together with linear phase or flat group delay in the passband is very important. Planar filters are preferred since they can be fabricated using a printed circuit with low cost, small size and light weight. The size of multi-frequency band planar filters is too large to be used in modern wireless and mobile systems. Thus research has mainly focused on the design to expand the other channel with a single input and a single output to meet the size reduction with a compact topology in the ISM band needs. There are currently several approaches to meet this demand in [1] and [2]. One solution consists of both different filters that are made in parallel[1], but the global size may become a problem. A better approach to multi-band filter design will be one that is low cost and has a simple configuration. Examples of dual-band design can be found in [2], [3] and [4]. But all of the sizes are somewhat large and topologies are too complicated. We developed novel resonators and designed the topologies of multi-frequency filter with the idea of two symmetric bands and rejection bands.

The Stepped Impedance Resonator (SIR) filters [4] enable size reduction in connection with series transmission line structures and have an advantage over spurious response suppression. Transform functions of filters are able to place transmission zero in the resonator frequency with the average of central multi-frequency pass-band and near the cutoff frequencies of pass-band so that higher selectivity with less resonators can be obtained.

II. MULTI-FREQUENCY BAND RESONATORS

Fig. 1 shows a compact SIR which evolves from combining the both quarter-wavelength open stub and short stub.

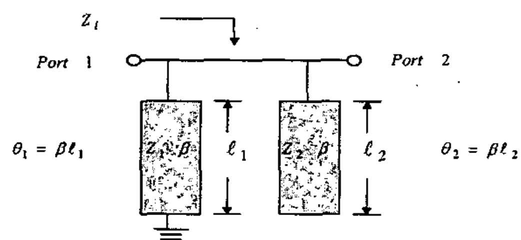

Fig.1. $\lambda / 4$ open stub and short stub resonators

Consider the input impedance of the network.

$$Z_i = Z_{i1} // Z_{i2} = \frac{Z_{i1} \cdot Z_{i2}}{Z_{i1} + Z_{i2}} = -j\frac{Z_1 Z_2 \tan\theta_1}{Z_1 \tan\theta_1 \tan\theta_2 - Z_2} \quad (1)$$

From the transform of ABCD matrix to S21

$$S_{21} = \frac{2}{A + B/Z_0 + C \cdot Z_0 + D} = \frac{2}{2 + Y_i \cdot Z_0}$$

$$= \frac{2}{2 + j\frac{Z_1 \tan\theta_1 \tan\theta_2 - Z_2}{Z_1 Z_2 \tan\theta_1} \cdot Z_0} \quad (2)$$

Analyze the two conditions for either $S_{21} = 0$ or $S_{21} = 1$.

If $S_{21} = 0 \Rightarrow Y_i \to \infty \Rightarrow \tan\theta_1 \text{or} \tan\theta_2 \to \infty$.

It is too difficult and complex to analyze further. For the simplification design, set

$\theta_1 = \theta_2 = \frac{\pi}{2} \Rightarrow \ell_1 = \ell_2 = \ell_0 = \frac{\lambda_0}{4}$, λ_0 is the wavelength of $f = 3.8GHz$ then $S_{21} = 1 \Rightarrow Y_i = 0 \Rightarrow Z_1 \tan\theta_1 \tan\theta_2 = Z_2$

$\Rightarrow \tan\theta_1 \tan\theta_2 = \frac{Z_2}{Z_1}$

Based on the previous concept, they are shunt by the $\lambda / 2$ and $\lambda / 4$ short stubs in Fig.2.

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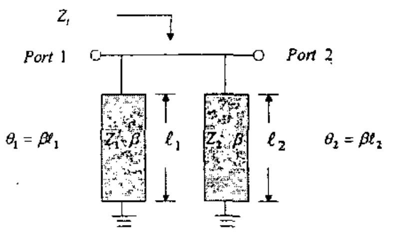


Fig.2. $\lambda/2$ and $\lambda/4$ short stub resonators

Obtain the following equation from the Fig.2 network.

$$S_{21}=\frac{2}{2+\frac{Z_1\tan\theta_1+Z_2\tan\theta_2}{jZ_1Z_2\tan\theta_1\tan\theta_2}\cdot Z_0} \quad (3)$$

If $S_{21}=0\Rightarrow Y_i\to\infty, Z_1\tan\theta_1+Z_2\tan\theta_2\to\infty$

or $Z_1Z_2\tan\theta_1\tan\theta_2=0$

$\tan\theta_1=0\Rightarrow\theta_1=\beta\ell_1=\pi\ \Rightarrow\ell_1=\frac{\lambda_0}{2}$, λ_0 is the wavelength

of $f=3.8GHz$, $\tan\theta_2\to\infty\Rightarrow\theta_2=\beta\ell_2=\frac{\pi}{2}\Rightarrow\ell_2=\frac{\lambda_0}{4}$

Then $S_{21}=1\Rightarrow Y_i=0\Rightarrow Z_1\tan\theta_1+Z_2\tan\theta_2=0$

$$\Rightarrow\frac{\tan\theta_1}{\tan\theta_2}=-\frac{Z_2}{Z_1} \quad (4)$$

After analyzing S-parameter of the SIR, the impedance ratio is obtained to control the secondary harmonic band to be the second pass-band. It is of interest to get the other point of view in the frequency domain. From the network configuration of both Fig.1 and Fig.2, it is combining a wide-BP (Bandpass) and narrow-BS (Bandstop) resonators to form the multi-frequency band resonator.

III. DESIGN HIGH ORDER MULTI-FREQUENCY BAND FILTERS

We synthesize resonators with $\lambda/4$ transmission line to make typical high-Q filters and demonstrate the development of the proposed models. However, the methodology is general and can be extended easily to a variety of other related structures. For matching the discontinuous transmission line, we apply the binomial coefficients (1, 1-1, 1-2-1) then $Y_b=2Y_a$ as shown in Fig.3, Later, we apply the binomial coefficient admittances (1, 1-1 and 1-2-1) and simulations were conducted using IE3D with the frequency range of 0-8 GHz. The both center frequencies are 2.4 GHz and 5.2 GHz. This methodology can be considered for integrating filter structures with binomial coefficient resonators. However, the both center frequencies at same time are shifted upper and inner in the frequency domain. To avoid this, the resonator shape can be optimized

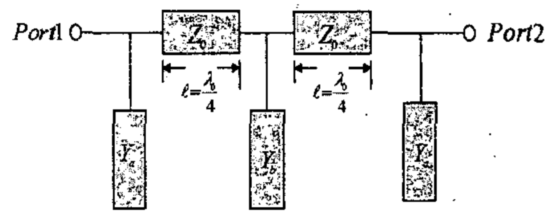
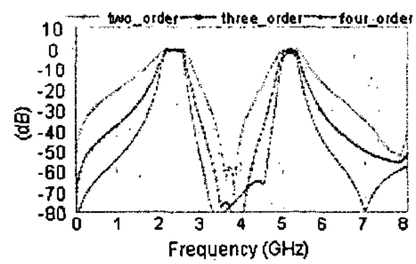
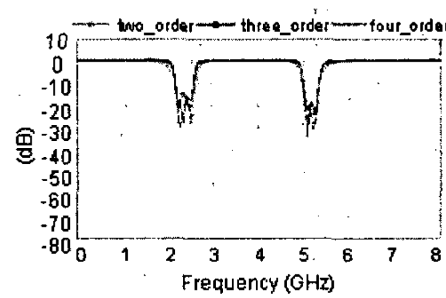


Fig.3. 3rd-order multi-frequency filter with binomial coefficient admittance values 1-2-1 in connecting the $\lambda/4$ microstrip line

for impedance values and electrical length. As a result, the optimization task is carried out, by annealing algorithm to synthesize the filter. The main technology of this algorithm is to derive a general system transfer function. Based on the idea of this algorithm, our approach first transforms the chain ABCD of short stubs and transmission-line sections to equivalent S_{21} and S_{11}, then adjusts a number of microstrip line characteristic impedances and electrical lengths which carefully chooses each microstrip line length and width. After performing the computer aided design, we get the optimal impedance value and electrical length of each stub. Then we transfer them into physical dimension to run EM simulation (IE3D) on S_{21} and S_{11} of $\lambda/4$ open stub and short stub configurations for 2, 3 and 4 orders respectively are shown in Fig.4. From the simulation data, the transmission zero point is found to be 3.8 GHz and it is sharper in the inner band which. It is the average of (2.4+5.2) GHz, so the methodology can be applied to any multi-frequency bands, as example in the 2.4 and 5.2 GHz.



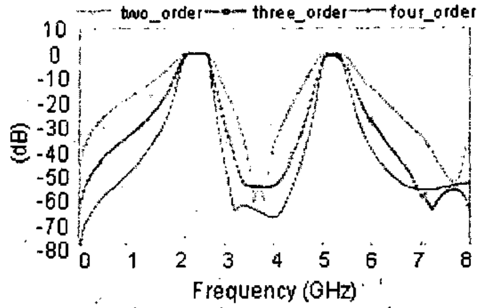
(a) S_{21}



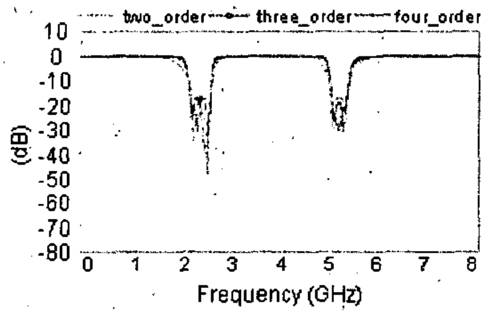
(b) S_{11}

Fig.4. Comparing S_{21} and S_{11} of the 2, 3 and 4 order multi-frequency filter with binomial coefficient admittance values in the $\lambda/4$ open stub and short stub resonators

Similarly, the EM simulation of $\lambda/2$ and $\lambda/4$ short stub is given in Fig. 5.



(a) S_{21}

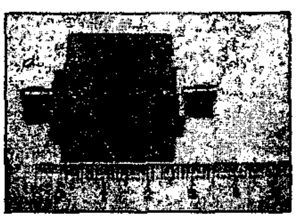


(b) S_{11}

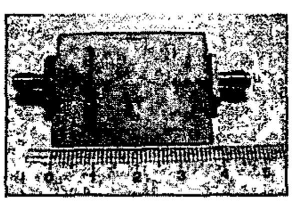
Fig.5. Comparison of S21 and S11 of the 2nd, 3rd and 4th order multi-frequency filters with binomial coefficients admittance value in the $\lambda/2$ and $\lambda/4$ short stub resonators

IV. FABRICATION AND MEASRUEMENT

The substrate used in this paper has thickness of 0.38mm. The metallic strip is 0.018mm, loss tangent is 0.0025, relative permittivity (ε_r) is 3.25. Measuring instrument is Advantest R3765/67G Series Network Analyzer.



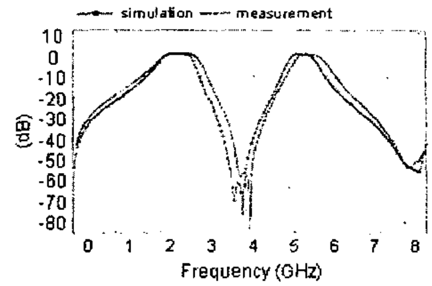
(a) 2nd(1-1) order circuit



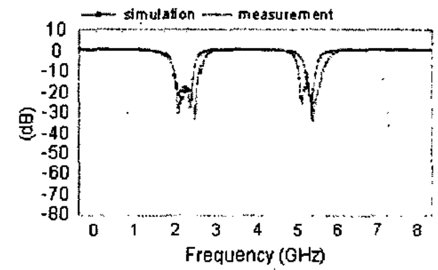
(b) 3rd(1-2-1) order circuit

Fig.6. Fabrication of layout in the $\lambda/4$ open stub and short stub resonators

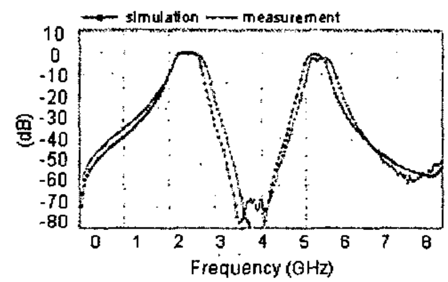
Based on the computational data, a prototype filter was fabricated and measured. We compare the simulation and measurement data between the first, second and the third resonators.



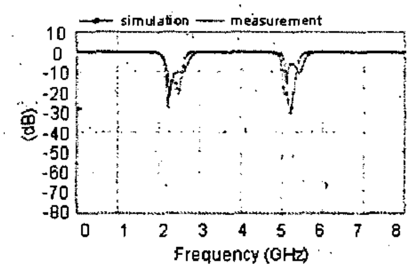
(a) Comparison of simulation and measurement of 2nd S_{21}



(b) Comparison of simulation and measurement of 2nd S_{11}



(c) Comparison of simulation and measurement of 3rd S_{21}



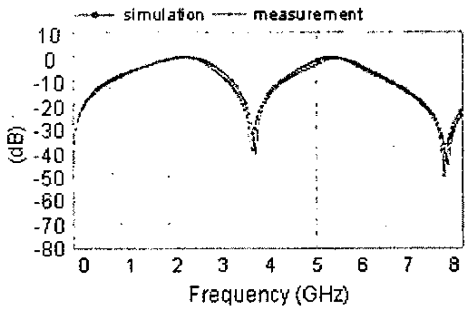
(d) Comparison of simulation and measurement of 3rd S_{11}

Fig.7. Comparison of simulation and measurement of S21 and S11 of the 1, 2, 3 and 4 order multi-frequency filter with binomial coefficients admittance value in the $\lambda/4$ open stub and short stub resonators

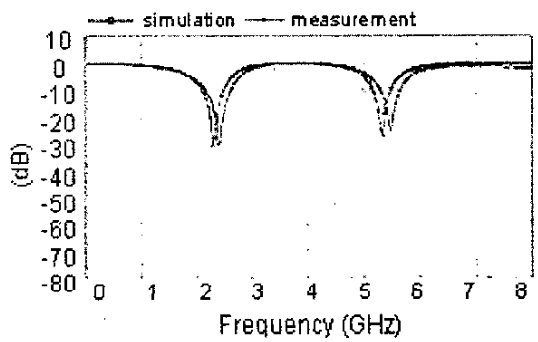
Similarly, we fabricate the layout in the Fig. 8. and compare the measurement and simulation in the Fig.9.

(a) 2nd(1-1) order circuit (b) 3rd(1-2-1) order circuit

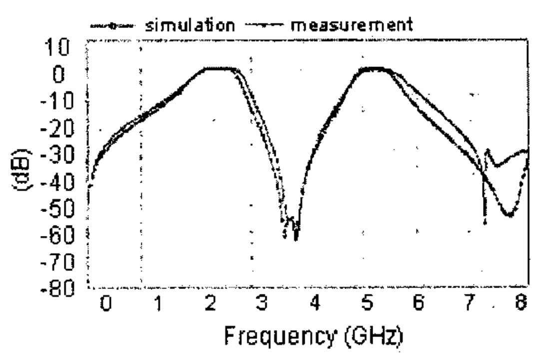
Fig.8. Fabrication of layout in the $\lambda/2$ and $\lambda/4$ short stub resonators



(a) Comparison of simulation and measurement of 1st S_{21}

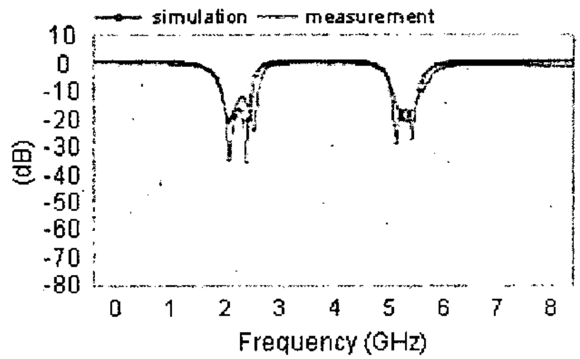


(b) S_{11}

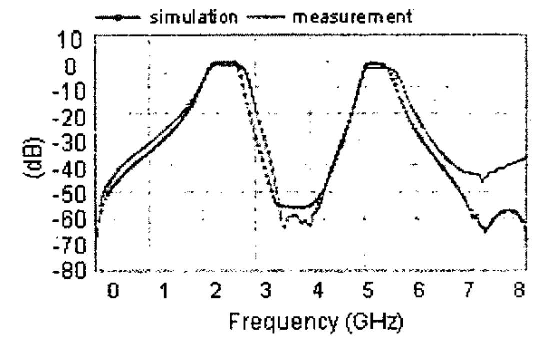


(c) 2 nd order S_{21}

Analysis of the simulation and measurement of curves in Fig7 and Fig9, they show a good agreement especially in the center frequencies and the bandwidths of the filter of S_{21}. There are very small ripple in the pass-band. The measured insertion loss is higher than simulation result because connection losses are not included in the simulation.



(d) 2nd order S_{11}



(e) 3rd order S_{21}

Fig.9. Comparison of simulation and measurement of S21 and S11 of the 1st, 2nd, 3rd and 4th order multi-frequency filters

Table1. Physical dimension of layout

	$Z_1(\Omega)$ $\theta_1 = \pi/2$		Layout Dimension	
	$Z_2(\Omega)$ $\theta_2 = \pi/2$		Width(w)	Length(l)
	Y_a	$Z_1 = 20$	3.1mm	12.17mm
		$Z_2 = 36$	1.44mm	12.57mm
	Y_b	$Z_1 = 10$	7mm	11.55mm
f=3.8(GHz)		$Z_2 = 36$	3.53mm	11.82mm
fc1=2.4(GHz)	$Z_1(\Omega)$ $\theta_1 = \pi$		Layout Dimension	
fc2=5.2(GHz)	$Z_2(\Omega)\theta_2 = \pi/2$		Width(w)	Length(l)
	Y_a	$Z_1 = 20$	3.1mm	24.55mm
		$Z_2 = 50$	0.87mm	13.05mm
	Y_b	$Z_1 = 10$	7mm	23.3mm
		$Z_2 = 25$	2.35mm	12.1mm

V. CONCLUSION

The novel wide dual-band filters (2.4 and 5.2 GHz) have been developed. The design methodology, based on the binomial coefficient admittances to synthesize the microstrip components, creates a compact and high performance structure. It also allows control of each center frequency, bandwidth and shape in the frequency domain, independently. The simulated and measured results show the validity of the proposed design method. The minimum insertion loss was less than 1 dB within 500 MHz bandwidth with a symmetric dual-passband and the return loss is great than 10 dB.

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