Design of Compact Wideband Bandpass Filters Based on Multiconductor Transmission Lines With Interconnected Alternate Lines

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***Abstract*—A new compact wideband bandpass filter consists of one series and one shunt multiconductor transmission-line (MTL) structure with short-circuited alternate lines is proposed. Based on this topology, a three-pole filter with two transmission zeros at the edges of the desired passband, generated by the shunt MTL, and with a wide stopband can be easily designed. Therefore, a comprehensive study of the filter is carried out and useful closed-form design equations are obtained and assessed by means of measurements. A filter is designed, fabricated and measured for validation and excellent agreement between analytical and measured results is obtained.**

***Index Terms*—Bandpass filter (BPF), coupled lines, filters, multiconductor transmission lines (MTLs).**

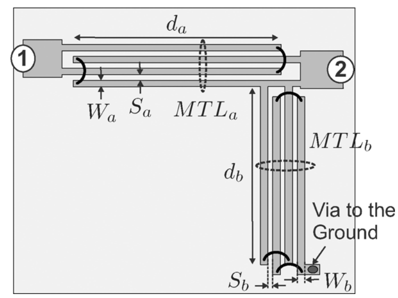

Fig. 1. Layout of the proposed wideband BPF consists of one series wire-bonded MTL and one shunt short-circuited wire-bonded MTL.

I. Introduction

THE design of filters is a long-standing issue in communication systems, where it is necessary to confine the transmitted power spectral density at the transmitter as well as reject undesired out-of-band signals and noise at the receiver. In recent years there has been a growing interest in designing bandpass filters (BPFs) with large fractional bandwidths and sharp selectivity. A design procedure to synthesize wideband BPFs by employing composite series and shunt stubs was presented in [1], but in order to enhance the selectivity it is necessary to cascade several stages. Other topologies, based on multiple-mode resonators with stepped-impedance or stub-loaded configuration have also been developed [1]–[7]. This variety of filters is designed to allocate the first resonant frequencies into the desired wide passband.

In this work, the use of shunt wire-bonded multiconductor transmission lines (MTL) is proposed to produce two transmission zeros at the cut-off frequencies and thus improving the skirt steepness. A wideband BPF consisting of one series MTL and a shunt short-circuited MTL (Fig. 1) is analyzed and a comprehensive study of the filter is presented. The use of two- and three-shunt coupled lines to design a BPF was addressed in [8], but only equations for the two-strip case were provided with no design guidelines. The proposed topology was also used in [9] to design balanced composite right/left handed transmission lines. However, the focus of that work is completely different and the objective is the design of both coupled-line sections to achieve a seamless transition between the left-handed and right-handed bands. Now, the proposed topology is analyzed by computing its transfer function and a closed-form design equation is derived for accurate and fast filter synthesis. In addition, the new proposed filter exhibits a quasi-elliptic function response with both good in-band flatness and high skirt selectivity.

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II. Analysis and Design of the Proposed Filter

The scheme of the proposed filter is depicted in Fig. 1. This filter consists of one series wire-bonded MTL and a shunt short-circuited wire-bonded MTL, where the shunt section allows the enhancement of the selectivity and aims to form a single wide pass band. The wire-bonded MTL is a specific case of multiconductor transmission lines in which bonding wires interconnect the ends of alternate conductors [10]. Assuming a lossless medium, the admittance matrix of the series wire-bonded MTL can be computed as [11]

$$[Y]_a = \begin{bmatrix} Y_{11_a} & Y_{12_a} \\ Y_{21_a} & Y_{22_a} \end{bmatrix} = \frac{-jc_a \sin\theta_a \cos\theta_a}{Z_{0_a}\left(c_a^2 - \cos^2\theta_a\right)} \begin{bmatrix} 1 & \frac{-c_a}{\cos\theta_a} \\ \frac{-c_a}{\cos\theta_a} & 1 \end{bmatrix} \quad (1)$$

where the characteristic impedance Z_{0_a} and the coupling factor c_a, related to the even- and odd-mode impedances, can be calculated from [12]

$$Z_{0_i} = \frac{(k_i-1)\left(Z_{oe_i} - Z_{oo_i}\right)\left(Z_{oe_i} + Z_{oo_i}\right)^2}{2\left[(k_i-1)Z_{oe_i} + Z_{oo_i}\right]\left[(k_i-1)Z_{oo_i} + Z_{oe_i}\right]} \quad (2)$$

$$c_i = \frac{(k_i-1)(Z_{oe_i}^2 - Z_{oo_i}^2)}{2Z_{oe_i}Z_{oo_i} + (k_i-1)(Z_{oe_i}^2 + Z_{oo_i}^2)}. \quad (3)$$

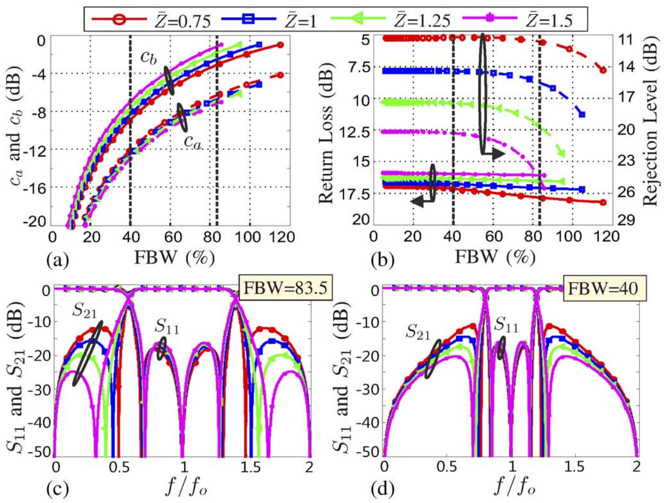

Fig. 2. (a) Coupling values c_a and c_b, (b) return loss and stopband minimum rejection level as a function of the 3 dB fractional bandwidth (FBW) for different impedance ratios $\bar{Z}$. (c)-(d) Magnitude of S-parameters for two particular FBW, 83.5% and 40%.

k_i denotes the number of conductors and the subindex i can be a or b to refer to the series or shunt MTL, respectively. The electrical length of the conductors is θ_a and is calculated as the arithmetic mean value of the even- and odd-mode electrical lengths θ_e and θ_o, respectively [11], [12].

The shunt coupled-line section results from short-circuiting to ground the output port of a two-port wire-bonded MTL. This element, recently analyzed in [11], is equivalent to a pair of series short-circuited and open-circuited shunt stubs, and its input admittance is given by

$$Y_b = j\frac{c_b}{Z_{0_b}\left[(1-c_b^2)\cot\theta_b - c_b^2\tan\theta_b\right]} \tag{4}$$

where θ_b is the electrical length of the MTL and the characteristic impedance Z_{0_b} and the coupling factor c_b are calculated by means of (2) and (3), respectively. Therefore, from (1) and (4) the S-parameters of the filter can be obtained, but these equations are quite involved. Nevertheless, as both MTLs will be designed to be a quarter-wavelength long at the design center frequency f_o to obtain the maximum operating bandwidth, it can be considered that θ_a and θ_b are similar and equal to $\theta = (\pi/2)(f/f_o)$. Hence, after some algebraic manipulations, the following relations can be deduced to calculate the magnitude of the S-parameters as

$$|S_{21}|^2 = \frac{1}{1+F^2(\theta)}, \quad |S_{11}|^2 = \frac{F^2(\theta)}{1+F^2(\theta)}, \tag{5}$$

$$F^2(\theta) = \frac{(g_6\cos^6\theta + g_4\cos^4\theta + g_2\cos^2\theta + g_0)\cos^2\theta}{4c_a^4\left(c_b^2 - \cos^2\theta\right)^2\sin^2\theta} \tag{6a}$$

$$g_6 = \left(c_a^2 - 1\right)\left[(c_a + \bar{Z}c_b)^2 - 1\right] \tag{6b}$$

$$g_4 = \bar{Z}^2c_b^2 - 2c_b\left(c_a^2-1\right)\left[\bar{Z}c_a\left(1+c_b^2\right) + c_b(c_a^2-1)\right] \tag{6c}$$

$$g_2 = -\bar{Z}^2c_a^2c_b^2\left(1+c_a^2\right) + c_b^2\left(c_a^2-1\right)\left[2\bar{Z}c_ac_b + c_b^2\left(c_a^2-1\right)\right] \tag{6d}$$

$$g_0 = \bar{Z}^2c_a^4c_b^2 \tag{6e}$$

and $\bar{Z} = Z_{0_a}/Z_{0_b}$. In (6) it was considered that $Z_{0_a} = Z_0$, being Z_0 the reference impedance, to achieve a wideband impedance-matched filter. As seen, the S-parameters depend on the values of c_a and c_b (3) but also on the impedance ratio $\bar{Z}$. Taking into account the function $F^2(\theta)$, it is clear that the filter can be designed to have a quasi-elliptic frequency response with three transmission poles. Besides, the transmission zeros of the circuit can be obtained from (5) as

$$\theta_{z_1} = 0, \quad \theta_{z_2} = \pi, \quad \theta_{z_3} = \arccos(c_b), \quad \theta_{z_4} = \pi - \theta_{z3}. \tag{7}$$

The symmetrical transmission zeros z_3 and z_4, that exclusively depend on the coupling factor c_b of the shunt MTL, can be placed at the edges of the passband to increase the roll-off rejection. However, the design condition to achieve a equiripple response is not straightforward and it requires determining the relation among c_a, c_b and $\bar{Z}$. From (6), and after some rigorous transformations, this simplified design equation is found

$$c_a = -\frac{c_b}{2}\left(\bar{Z} - \sqrt{\bar{Z}^2+4}\right) \tag{8}$$

that specifies the relationships between the series and shunt wire-bonded MTLs to obtain a quasi-elliptic function response. Fig. 2 depicts the required coupling values c_a and c_b as well as the expected return loss and stopband minimum rejection level as a function of the 3 dB fractional bandwidth (FBW). These curves are obtained from (5) by imposing the design condition (8). As seen, the rejection level can be improved by increasing the impedance ratio $\bar{Z}$ and, as known, the greater the coupling factors, the broader the passband. To visualize the filtering characteristics of the proposed filter, the magnitude of S-parameters are depicted for two particular FBW values; 83.5% and 40%, respectively. It is clear that the filter presents three poles and two transmission zeros at the lower and upper cut-off frequencies, and that the higher the impedance ratio, the greater the rejection level. However, it is important to remark that the possible values of c_a, c_b and $\bar{Z}$ are directly related to the manufacturing capabilities.

Finally, given (8) ($Z_{0_a} = Z_0$, $Z_{0_b} = Z_{0_a}/\bar{Z}$) and using (2) and (3), the equations to calculate the required even- and odd-mode impedances of both wire-bonded MTLs are given by

$$\frac{Z_{oe_i}}{Z_{0_i}} = \frac{c_i+1}{2c_i^2}\Big[(k_i-1)(c_i-1)+c_i + \sqrt{(k_i-1)^2(1-c_i^2)+c_i^2}\Big] \tag{9a}$$

$$Z_{oo_i} = Z_{oe_i}\frac{(k_i-1)(1-c_i)}{c_i+\sqrt{(k_i-1)^2(1-c_i^2)+c_i^2}}. \tag{9b}$$

Equations (9a) and (9b), when $Z_{0_i} = Z_0$ and $k = 2$, are equivalent to that given in [13] to synthesize dc blocks with a Butterworth frequency response.

III. Experimental Validation

Based on the above analysis, a filter prototype has been designed at a center frequency $f_o = 3.5$ GHz, with a 3 dB fractional bandwidth of about 84%. The value of c_b is first calculated to allocate the two transmission zeros at the edges of the passband (7), and then, the coupling factor c_a is obtained by means of (8). The design parameters are $c_a = -6.6$ dB, $c_b = -2.53$ dB and $\bar{Z} = 1$ and the prototype is implemented on a Rogers 4350B substrate with relative permittivity of 3.66 and thickness of 30 mil. Once the values of c_a, c_b and $\bar{Z}$ are obtained

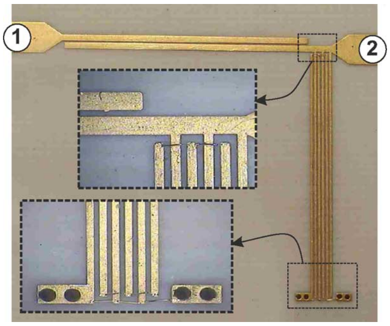

Fig. 3. Photograph of the fabricated filter.

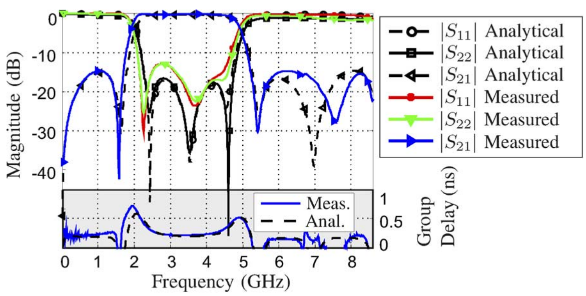

Fig. 4. Measured and analytical (5) S-parameters and group delay of the fabricated filter.

TABLE I
COMPARISON WITH OTHER TOPOLOGIES

Ref.	f_o (GHz)	FBW (%)	IL (dB)	RL (dB)	Roll-off slope lower & upper (dB/GHz)	Size ($\lambda_o \times \lambda_o$)
[2]	5.95	80	0.9	14	52 & 40	0.17×0.17
	6.85	105	0.8	17	50 & 35	0.18×0.17
[3]	5.3	45	1.2	15	13 & 18	0.5 ×0.18
[4]	6.85	110	1.5	11	56 & 35	0.51×0.31
[5]	3	63	1	11	36 & 41	0.24×0.07
[6]	6.85	110	1	11	14 & 11	0.07×0.12
[7]	6.85	120	1.5	10	35 & 24	0.51×0.33
⋆	3.5	83.7	0.6	13	80 & 52	0.17×0.16

f_o: design center frequency; IL & RL: insertion and return loss;
roll-off slope: defined between 3 and 30 dB rejection frequency points;
λ_o: free-space wavelength at design center frequency; ⋆: This work.

(8), the required even- and odd-mode impedances are easily calculated as a function of the number of conductors (9) and translated into physical dimensions. The computed dimensions in this experiment are $k_a = 2$, $W_a = 320\ \mu\text{m}$, $S_a = 115\ \mu\text{m}$ for the series wire-bonded MTL and $k_b = 6$, $W_b = 110\ \mu\text{m}$ and $S_b = 100\ \mu\text{m}$ for the shunt MTL. The length of both MTLs is $d_a = d_b = 13.5\ \text{mm}$. It is important to mention that these theoretical dimensions have not been optimized with any EM simulator and that the number of conductors and the value of $\bar{Z} = 1$ have been chosen to meet our fabrication capability ($> 100\ \mu\text{m}$). A photograph of the filter is shown in Fig. 3.

Fig. 4 shows the measured and analytical (5) S-parameters and the group delay of the designed and fabricated BPF. It can be observed that a very good agreement is seen between measurements and theory, with insertion and return losses less than 0.6 dB and greater than 13 dB in the passband, respectively. The group delay is less than 0.7 ns and the two symmetrical transmission zeros at $f_{z_3} = 1.6$ GHz and $f_{z_4} = 5.4$ GHz (7) provide sharp cut-off slopes and a wide upper stopband is achieved with a rejection level greater than 15 dB. Table I shows a comparison between our proposed filter and others reported topologies. It can be seen that there is a good performance according to the insertion loss, return losses and attenuation slopes. Notwithstanding, it is necessary to mention that the slight deviation in measurements are due to fabrication tolerances (variations of up to 10% have been measured) and because of the unequal even- and odd-mode phase velocities.

IV. CONCLUSION

A new compact wideband BPF consisting of just two quarter-wavelength wire-bonded MTLs has been presented and thoroughly analyzed. It has been proved that by connecting a shunt wire-bonded MTL it is possible to generate two transmission zeros at the lower and upper cut-off frequencies and, thus, to achieve sharpened roll-off skirts. As an important result, closed-form equations have been obtained to design three-pole BPFs with a quasi-elliptic response. A BPF with a 3 dB fractional bandwidth of about 84% and with a good performance and a wide stopband has been designed, fabricated and measured. The very good agreement between the analytical and measured results allows the use of the presented analytical equations to design filters by means of a quick and reliable procedure.

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