A Wideband Bandpass Filter With Wide Upper Stopband Using Stepped-Impedance Cascadable 180° Hybrid Rings

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***Abstract*—This letter proposes a wideband bandpass filter by cascading two 3-dB stepped-impedance cascadable 180° hybrid rings with a pair of stepped-impedance lines. Thanks to the stepped-impedance lines, a broad upper stopband is achieved. The stepped-impedance vertically installed planar (VIP) coupler is used to implement the ideal 180° phase inverter and crossovers. The experimental results show that this 2-GHz center frequency, fourth order filter has a 10-dB return loss bandwidth of 92.5% and upper stopband rejection levels of better than −20 dB up to 6.8 GHz. This proposed filter achieves wide passband and broad stopband performance simultaneously.**

***Index Terms*—Phase inverter, stepped-impedance, vertically installed planar (VIP) coupler, wideband bandpass filter (BPF).**

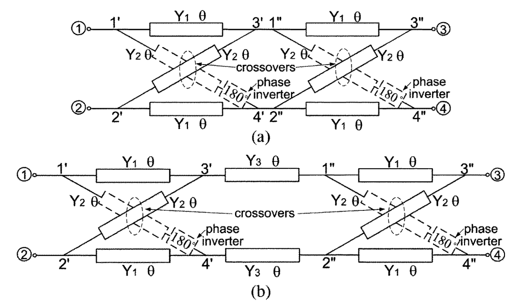

Fig. 1. Schematics of the proposed wideband filter: (a) basic structure and (b) modified structure where $Y_i = 1/Z_i, i = 1, 2\ 3$, are normalized values.

I. Introduction

In recent years, the design of a wideband bandpass filter (BPF) is an important issue for the wideband communication systems. This type of filter needs not only wide bandwidth to meet the required large data rate but also broad stopband to suppress harmonics. Based on the insertion loss method [1], a wideband BPF function of Chebyshev type can be easily synthesized. However, to realize a wideband BPF with the conventional microstrip half-wavelength parallel coupled-line structure, very small gaps between coupled strips are required to achieve a tight coupling. This may cause some difficulties in the fabrication processes. Moreover, due to the unequal even- and odd-mode phase velocity, this filter has a first spurious passband at $2f_0$, resulting in poor harmonic suppression. To improve bandwidth performance, several methods [2]–[4] have been proposed. In [2], cascading of two cascadable 180° hybrid ring couplers could form a wideband filter. In [3], a multiple-mode resonator (MMR) is used to construct a wideband filter. In [4], a microstrip ring BPF with dual stopbands below 3.1 GHz and above 10.6 GHz was built up. However, all of the above-mentioned filters show narrow upper stopband performance and some of them even show notch-like upper stopband response. Therefore, it is still a challenge to design a single filter having wide passband and broad stopband characteristics simultaneously.

In this letter, we modify the cascadable 180° hybrid ring in [5] with a stepped-impedance structure [6] to design a wideband BPF so that the above-described shortcomings can be solved. This filter with the stepped-impedance structure acquires not only size reduction, but also better upper stopband clearance. The experimental results show good agreements with the simulated results.

Manuscript received February 8, 2007; revised March 22, 2007. This work was supported by the National Science Council of Taiwan, R.O.C., under Grant NSC 95-2221-E-009-042-MY3.

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Digital Object Identifier 10.1109/LMWC.2007.901771

II. Analysis and Design of the Proposed BPF

The basic structure of the proposed BPF is shown in Fig. 1(a) where the electrical length (θ) of each transmission line is 90° and the admittances Y_1 and Y_2 are normalized values with respect to system admittance Y_0. Note that the filter configuration shown in Fig. 1(a) is different from the filter in [2] where, in [2], the second rat-race ring is flipped with respect to horizontal central line of Fig. 1(a). From [5], at the center frequency, the S-parameters of the circuit shown in Fig. 1(a) is

$$\begin{aligned} S_{11} &= S_{22} = 0 \\ S_{21} &= S_{43} = 0 \\ S_{31} &= S_{42} = -\left(Y_1^2 - Y_2^2\right)/Y_1^2 + Y_2^2 \\ S_{41} &= -S_{32} = 2Y_1Y_2/\left(Y_1^2 + Y_2^2\right). \end{aligned} \quad (1)$$

When $Y_1 = Y_2$, the signal excited at port 1 is equally divided between $3'$ and $4'$ and no power reaches $2'$. The signal arriving at $3'$ and $4'$ are out of phase. Then, the signal at $3'$ is equally divided between $3''$ and $4''$ and the signal arriving at $3''$ and $4''$ are out of phase. The signal at $4'$ is equally divided between $3''$ and $4''$ and the signal arriving at $3''$ and $4''$ are in phase. Therefore, all of the power excited at port 1 will be delivered to port 4 and no power reaches port 3 and port 2 (corresponding to a 0-dB coupler). This means that a bandpass response can be obtained by cascading two 3-dB 180° hybrid rings. With the design method described in [5], the bandwidth of this filter corresponding to specific return losses can be obtained. Here, to

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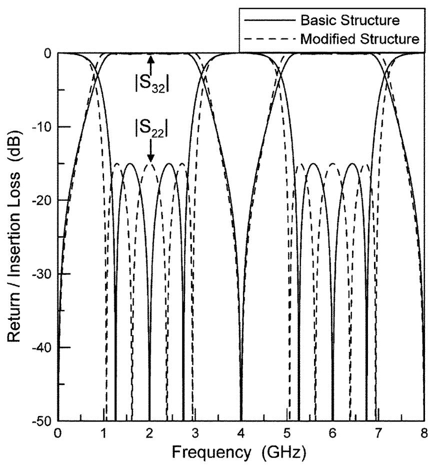

Fig. 2. Simulated results of the proposed wideband filter: (a) basic structure and (b) modified structure.

increase the order and the bandwidth of the proposed filter, a 90° transmission line of impedance Z_3 can be added between two 180° hybrids [7], [8] as shown in Fig. 1(b).

The simulated results of the filters in Fig. 1 with 15-dB return loss are depicted in Fig. 2 that they correspond to third ($Z_1 = Z_2 = 53.31\ \Omega$) and fourth ($Z_1 = Z_2 = 58.65\ \Omega$ and $Z_3 = 41.16\ \Omega$) order filter respectively. Although wide passband can be easily achieved, the first spurious passband at $3f_0$ results in poor upper stopband performance. In [9], the frequency of the first spurious response of the 180° hybrid ring can be moved using the stepped-impedance structure by adjusting its structural parameters. Thus, by applying the stepped-impedance structure to Fig. 1(b), a wideband filter with broad upper stopband characteristics can be obtained.

A quarter-wave transmission line of impedance Z is shown in Fig. 3(a) and it serves as a basic building block of the proposed BPF. The quarter-wave transmission line can be replaced by a stepped-impedance structure shown in Fig. 3(b) where Z_H and Z_L are the characteristic impedances of the cascaded sections and θ_H and θ_L are the corresponding electrical lengths. By choosing $Z_H > Z_L$, the overall electrical length of the stepped-impedance section can be shorter than 90°. Equating the ABCD matrix of the quarter-wave transmission line and the stepped-impedance structure shown in Fig. 3, we have

$$\frac{Z_H Z_L \cos\theta_H \cos 2\theta_L - (Z_H^2 + Z_L^2)\cos\theta_L \sin\theta_H \sin\theta_L}{Z_H Z_L} = 0 \quad (2)$$

$$\frac{Z_H^2 \cos^2\theta_L \sin\theta_H + Z_L(-Z_L \sin\theta_H \sin^2\theta_L + Z_H \cos\theta_H \sin 2\theta_L)}{Z_H} - Z = 0 \quad (3)$$

$$\frac{Z_L^2 \cos^2\theta_L \sin\theta_H + Z_H(-Z_H \sin\theta_H \sin^2\theta_L + Z_L \cos\theta_H \sin 2\theta_L)}{Z_H Z_L^2} - \frac{1}{Z} = 0. \quad (4)$$

The first step to implement the proposed wideband stepped-impedance BPF is to specify the required return loss and to obtain the corresponding circuit parameters in Fig. 1(b). This can be easily done using optimization process provided in most circuit simulators because only two circuit parameters ($Z_1 = Z_2 = 58.65\ \Omega$ and $Z_3 = 41.16\ \Omega$ in this example) need to be determined. The next step is to employ the stepped-impedance structure to each line section in Fig. 1(b). Followings show the procedures. First, arbitrarily choose the θ_H and θ_L and then use (2)–(4) to find Z_H and Z_L. Second, check the obtained Z_H and Z_L to be realizable or not, if not, redo the first step again. Finally, fine-tuning the obtained circuit parameters might be required to approach the equal ripple response in the passband. The optimized circuit parameters are shown in Table I. It should be emphasized that previously described procedures are based on an ideal 180° phase inverter.

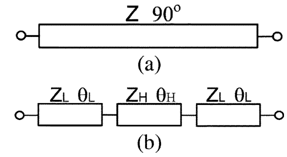

Fig. 3. (a) Quarter-wave transmission line and (b) stepped-impedance circuit equivalent to a quarter-wave transmission line.

TABLE I
PARAMETERS OF THE STEPPED-IMPEDANCE BPF

	Z_H (Ω)	Z_L (Ω)	θ_H (Degree)	θ_L (Degree)
λ/4, Z=58.65Ω	97	29	30.0	19.5
λ/4, Z=41.16Ω	84	23	22.3	20.0

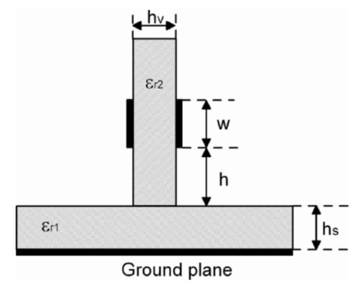

Fig. 4. Cross-sectional view of the VIP coupler.

TABLE II
COMPUTED RESULTS BY HFSS ($f_0 = 2$ GHz)

Z (Ω)	h (mil)	W (mil)	ε_{re}	ε_{ro}	Z_{0e} (Ω)	Z_{0o} (Ω)
97	104	29	1.35	2.71	320.0	42.5
29	0	133	1.84	3.06	107.0	12.7

III. REALIZATION AND MEASUREMENT OF THE PROPOSED BPF

Following the above-described design steps, the circuit parameters of the proposed BPF can be obtained. However, the realization of an ideal 180° phase inverter on microstrip circuits would be a problem. Fortunately, this can be solved by using an opposing ends short-circuited $\lambda/4$ coupled line (we simply call it "short-ended coupled line") [10] with the vertically installed planar (VIP) structure [11]. Therefore, the ideal 180° phase inverter and crossovers, shown in Fig. 1, can be implemented by a $\lambda/4$ short-ended coupled line with the VIP structure [5]. Then, we employ stepped-impedance structure to the $\lambda/4$ short-ended VIP coupler. To simplify this design, the short-ended stepped-impedance VIP coupler is approximately

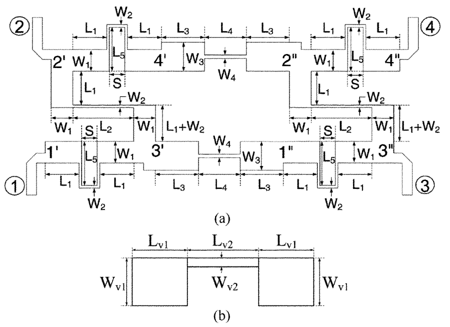

Fig. 5. Geometry of the fabricated filter (a) main circuit section $W_1 = 101$ mil, $W_2 = 12$ mil, $W_3 = 135$ mil, $W_4 = 17$ mil, $S = 82$ mil, $L_1 = 158$ mil, $L_2 = 283$ mil, $L_3 = 200$ mil, $L_4 = 195$ mil, $L_5 = 201$ mil and (b) VIP stepped-impedance coupler. $W_{v1} = 133$ mil, $W_{v2} = 29$ mil, $L_{v1} = 184$ mil, $L_{v2} = 296$ mil.

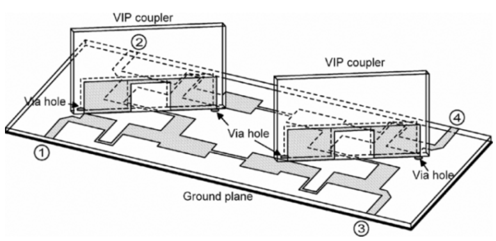

Fig. 6. 3-D structure of the proposed wideband filter with the stepped-impedance structure.

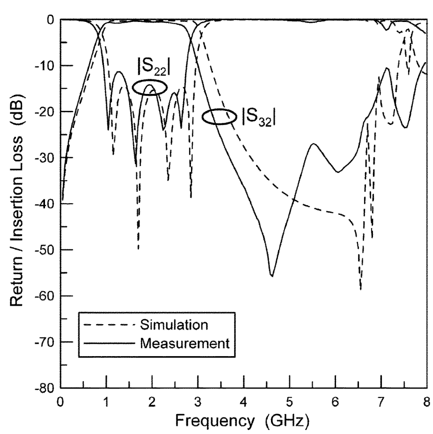

Fig. 7. Measured and simulated results of the proposed wideband filter with the stepped-impedance structure.

equivalent to a balanced stepped-impedance transmission line with 180° of twist (physically 180° of twist equivalent to an ideal 180° phase inverter). This equivalence is valid as long as the even- to odd-mode impedance ratio is large. The characteristic impedance of each balanced line section approximately equals to $2Z_{0e}Z_{0o}/(Z_{0e} - Z_{0o})$. Its physical length can be determined by the effective odd-mode dielectric constant because the signal going through it is mainly odd-mode. Fine tuning the length of each VIP line section to fit the insertion phase of other three step impedance lines at the center frequency may be required. Fig. 4 shows the cross-sectional view of the VIP coupler. Both the main circuit and the VIP couplers are implemented on a RO4003 substrate with thickness (h_v and h_s) of 20-mil and dielectric constant (ε_{r1} and ε_{r2}) of 3.38. The filter is designed to operate at the center frequency of 2 GHz. Based on the above-described method, the characteristic impedances and effective dielectric constants for even- and odd-modes of the VIP coupler can be obtained by the 3-D EM simulator (Ansoft HFSS) as given in Table II. Fig. 5 indicates the geometry of the fabricated filter. The 3-D structure of the proposed wideband filter is shown in Fig. 6 where port 2 and port 3 are the input and output port, respectively. Both port 1 and port 4 are isolated ports and are terminated with a 50 Ω load.

Fig. 7 shows the simulated and measured results of the proposed filter and the measured insertion loss is approximately 0.4 dB. Note that the simulated results of the proposed filter are simulated with the computed parameters in Table II by the circuit simulator (AWR Microwave Office). The measured 10-dB return loss bandwidth is from 0.95 to 2.8 GHz (92.5%). The stopband rejection is better than −20 dB from 3.3 to 6.8 GHz. The measured bandwidth is a little narrower than the simulated responses and the experimental stopband rejection is a little worse than the simulated results. This may be caused by the junction effect (especially junctions at microstrip and VIP coupler) and circuit fabricating imperfections.

IV. Conclusion

In this letter, by cascading two 3-dB 180° hybrid rings with a pair of 90° transmission lines, a stepped-impedance wideband BPF has been designed and fabricated to exhibit wide passband and broad stopband performance simultaneously. The short-ended step impedance VIP coupler has provided a good phase inverter as well as a crossover in this realized filter. In addition, the bandwidth can be further enhanced by cascading more sections of 90° transmission lines in this proposed structure. The experimental results are in good agreement with the simulated results.

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