

modifying the VSWR, gain, and radiation pattern. The error obtained between mathematical model and measurements is close to 3.3%.

## ACKNOWLEDGMENT

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# STUDY OF PARALLEL COUPLED-LINE MICROSTRIP FILTER IN BROADBAND

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**ABSTRACT:** A parallel coupled-line microstrip filter is designed and implemented on PCB with its maximum bandwidth. In this paper, this filter has presented almost very good characteristics including the simulated bandwidth of 67%, the measured bandwidth of 64%, lower insertion loss of 1.27 dB, and flat response in the pass-band. The maximum bandwidth fabricated on PCB shows good agreement with the theoretically calculated bandwidth of 70% due to the narrower gap size between the coupled lines. © 2005 Wiley Periodicals, Inc. *Microwave Opt Technol Lett* 48: 373–375, 2006; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21353

**Key words:** parallel coupled-line microstrip filter; PCB; narrower gap

## 1. INTRODUCTION

Microstrip band-pass filters [1–4] have been widely used in the RF front end of microwave transceivers because of their planar structure, ease of synthesis method, and low cost [5]. Parallel coupled-line microstrip filters (PCMFs) are usual components in microwave integrated circuits due to their simple design and versatility [6, 7]. Due to the need of newly proposed ultra-wideband (UWB) specifications, band-pass filters with GHz bandwidth will be required. The parallel coupled-line microstrip filter fabricated on a printed ceramic board (PCB) usually has a relatively small bandwidth. This is mainly due to the fact that the reductions of  $w/d$  and  $s/d$  using conventional fabrication methods were limited. However, this limitation has been surmounted, since the gap between

**TABLE 1** Characteristic Impedances of the Parallel Coupled-Line Microstrip Filter for Odd and Even Modes and Values of  $w$  and  $s$

Bandwidth	60%	70%	80%
$Z_{0o1}$ $\Omega$	136	146	157
$Z_{0e1}$ $\Omega$	45	49	52
$Z_{0o2}$ $\Omega$	115	132	150
$Z_{0e2}$ $\Omega$	40	44	50
$Z_{0o3}$ $\Omega$	125	134	143
$Z_{0e3}$ $\Omega$	43	36	47

Bandwidth [mm]	60%	70%	
$W_1$ mm	0.18	0.12	—
$S_1$ mm	0.12	0.1	—
$W_2$ mm	0.34	0.21	—
$S_2$ mm	0.15	0.13	—
$W_3$ mm	0.25	0.19	—
$S_3$ mm	0.14	0.11	—

two coupled lines can be scaled down to 4  $\mu\text{m}$  [8], which allows PCMFs to be used successfully in broader bandwidth.

Previously, we have shown coplanar planar waveguide (CPW) and microstrip ring resonators, broadband and narrowband filters, and antenna and transmission lines on ion-implanted silicon substrates with excellent RF performance up to 100 GHz [9, 11]. This achievement is partially due to a sufficiently small-sized gap fabricated on a low RF loss silicon substrate, which is generated by ion-implantation and lithography technique.

In this paper, the bandwidth calculated using the theory previously proposed in [5] is presented. The lithography technique is applied in order to experimentally realize the broadband parallel coupled-line microstrip filter on PCB. Both the theoretical and experimental bandwidths are found to be in agreement.

## 2. DESIGN AND FABRICATION

The formulas of the theoretical calculation for the parallel band-pass filter are given by [5]:

$$J_1 Z_0 = \left[ \frac{\pi \Delta}{2g_1} \right]^{1/2}, \quad (1)$$

$$J_n Z_0 = \frac{\pi \Delta}{2\sqrt{g_{n-1}g_n}}, \quad (2)$$

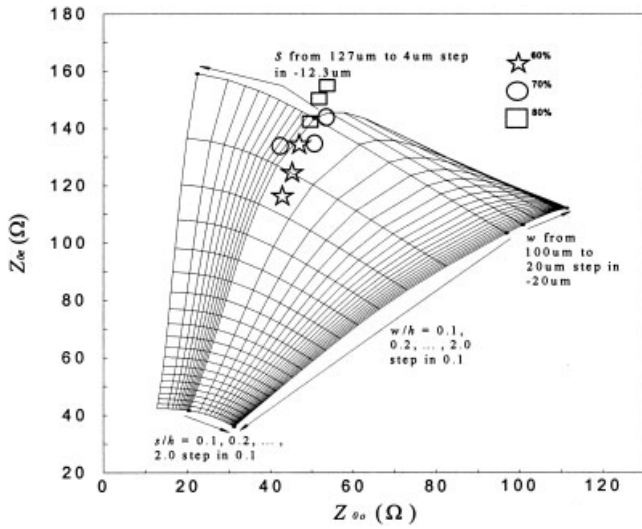
$$J_{n+1} Z_0 = \left[ \frac{\pi \Delta}{2g_n g_{n+1}} \right]^{1/2}, \quad \text{for } n = 2, 3, 4, \dots, \quad (3)$$

where  $\Delta$  is the fractional bandwidth. Here,  $g_n$  and  $Z_0 = 50\Omega$  are the element value and the characteristic impedance of the parallel band-pass filter, respectively. Based on Eqs. (1)–(3), the characteristic impedances for odd and even modes,  $Z_{0o}$  and  $Z_{0e}$ , can be determined by

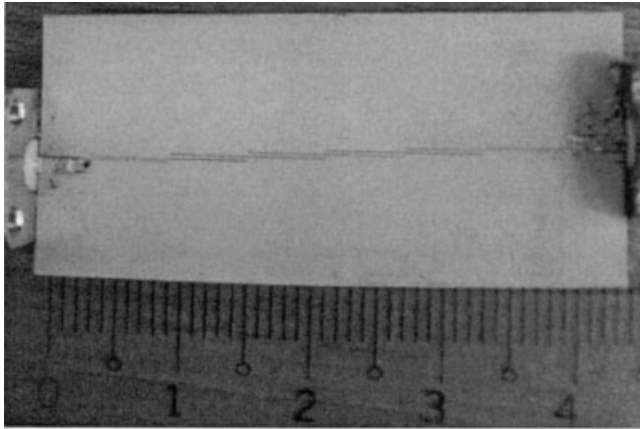
$$Z_{0en} = Z_0 [1 + J_n Z_0 + (J_n Z_0)^2], \quad (4)$$

$$Z_{0on} = Z_0 [1 - J_n Z_0 + (J_n Z_0)^2]. \quad (5)$$

Using these two expressions, the characteristic impedances of the multisection parallel coupled-line filter for odd and even modes can be carried out. In our work, we focused on the five-section parallel coupled-line filter. This filter was designed as a Cheby-



**Figure 1** The general solution of  $s/d$  and  $w/d$  under the range of 1.27 mm and this special case of the  $s/d$  and  $w/d$  from 1.27 mm into 4  $\mu\text{m}$

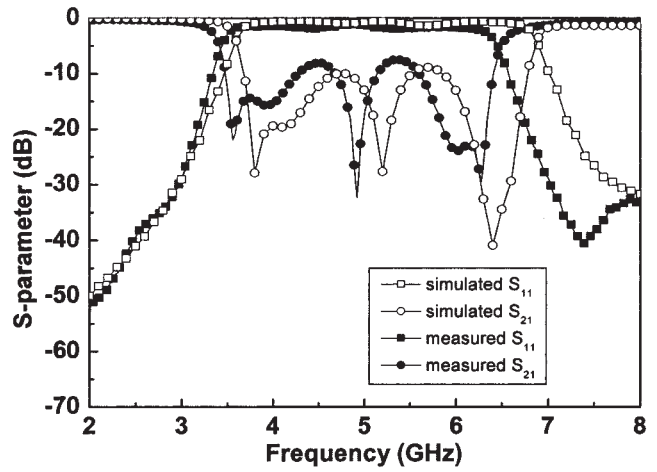


(a)



(b)

**Figure 2** The pictures of (a) fabricated and (b) enlarged structure of the five-section parallel coupled-line filter



**Figure 3** The parallel coupled-line microstrip filter with a central frequency  $f_c = 5$  GHz: (a) simulated response frequency equal as 67%; (b) measured responses frequency equal as 64%

shev-type filter, with ripple of 0.1 dB. Then the element values for our designed filter can be obtained by  $g_0 = 1$ ,  $g_1 = 1.146$ ,  $g_2 = 1.371$ , and  $g_3 = 1.975$  [3]. Substituting these four element values into Eqs. (1)–(5), the characteristic impedances of our filter for odd and even modes were calculated, and the results are listed in Table 1. These calculated characteristic impedances can help us to find the corresponding values of  $w/d$  and  $s/d$  from the calibration plot shown in Figure 1, where  $w$  is the metal width,  $s$  is the gap size, and  $d$  is the substrate thickness. The obtained values of  $w$  and  $s$  are also listed in Table 1. We can see that the maximum bandwidth is 70%. By meeting the theoretically determined values of  $w$  and  $s$  for the maximum bandwidth of 70%, we fabricated a five-section parallel coupled-line microstrip filter, the photograph and structure of which are displayed in Figures 2(a) and 2(b), respectively.

To realize the smaller gap width in this filter, the circuit was fabricated using a conventional low-cost IC process with a  $>1\text{-}\mu\text{m}$  resolution (nearly two decades older than the current VLSI technology) and standard FeCl etching. This technology has a better pattern transfer and sharper corner edge that can help extend the PCB process for making a narrower metal width and interval gap. In our experiment, for fabrication, we used a Duroid/6010 substrate with substrate thickness of 1.27 mm, metal thickness of 1/2 oz., and dielectric constant of 10. The circuit measurements were performed using an Agilent/HP 8720C vector network analyzer.

### 3. EXPERIMENTAL RESULTS AND DISCUSSION

#### 3.1. Parallel Coupled-Line Microstrip Filter for 70% Bandwidth

As shown in Figure 3, the simulated fractional bandwidth was found to be 67% close to the ideal destination of 70%. In the pass-band, the best simulated insertion return losses were 1.76 and 12.99 dB, respectively. On the other hand, the measured fractional bandwidth result was 64%, which agrees with the simulated result. Some mismatch may be due to the small power reflection of the nonideal situation. The best in-band insertion and return losses are only 1.27 and 16.7 dB, respectively. It can be observed that the smaller coupling-gap size not only achieves small insertion loss, but also exhibits the broadband characteristics of a band-pass filter. Its characteristics can modify the conventional type of this filter in order to obtain more advantages for broadband communications.

#### 4. CONCLUSION

With the rapid development of broadband communications, flatter coupling response and wider bandwidth are very important issues for UWB applications. In this paper, we have used the lithography technique to improve the gap width, and this technique can help us to achieve the wider bandwidth, low insertion loss, and high-performance characteristics of designed parallel coupled-line microstrip filters. The theoretically calculated results show that the maximum bandwidth obtained was up to 70%. The experimental data also show good agreement with the calculated results.

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## PRACTICAL IMPEDANCE MATCHING USING GENETIC PROGRAMMING

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**ABSTRACT:** A genetic programming method for impedance matching is presented. The method is applied to an impedance-matching problem where the element values available are restricted to a set of preferred

values. The experimental results show that this method can effectively generate practical and economical solutions. Since it can additionally deal with other design considerations, this method is useful for practical impedance matching. © 2005 Wiley Periodicals, Inc. Microwave Opt Technol Lett 48: 375–377, 2006; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.21354

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#### 1. INTRODUCTION

In order to maximize power transmission, an impedance-matching network is required between the given source and load impedances. In conventional design approaches [1–5], the element values are assumed to be continuous (that is, they can be any real number). In reality, however, the element values are discrete. This limitation occurs in the design of discrete-element circuit because the element values available from vendors are restricted to a set of preferred values. It also occurs in integrated circuit design for high-frequency applications, such as RF or microwave integrated circuits (MICs), where capacitors may be implemented using an interdigital structure and inductors using a spiral structure [6]. Because the number of fingers or turns is an integer, the element values that are a function of integers are discrete.

In this paper, we propose a genetic programming (GP) method, based on our previous work [7, 8], to automatically generate impedance-matching networks consisting of elements with restricted values. The experimental results show that the proposed method can effectively generate practical and economical solutions. In addition, this method has the following advantages [7, 8]. First, the tree structure of GP is dynamic rather than fixed length, which is a desirable property in the design cases where the circuit complexity and topology cannot be predicted. Second, a circuit-analysis algorithm based on the tree structure is fast and suitable for symbolic circuit analysis, which is convenient for calculating circuit characteristics such as element sensitivity, group delay, and step response. Third, the tree structure is capable of representing any series-parallel RLC circuits; therefore, a resonant structure is allowed. Finally, the GP method can deal with other design considerations; for example, the element values are frequency dependent and restricted to a set of preferred values. The abovementioned advantages make it a practical method for solving impedance-matching problems.

#### 2. GENETIC PROGRAMMING

The GP method is based on previous work [7, 8] in which we proposed a binary tree structure to represent series-parallel RLC circuits. Figure 1 shows the template of impedance-matching problems, where  $Z_S$  and  $Z_L$  denote the source and load impedances, respectively. The kernel of this method is the GP algorithm described below. At first, an initial population is created, which is composed of 500 individuals randomly chosen among the building blocks, as shown in Figure 2. The inductor  $L$  or capacitor  $C$  node is assigned a value randomly selected from a set of element values. The crossover and mutation operations are then applied to the population in order to explore the circuit topology and element values, respectively. The evolutionary process continues iteratively until a fully compliant solution is found or a maximum number of iterations is reached.

For a specified minimum transducer power gain ( $G_{T,min}$ ), the GP algorithm incorporating the following fitness function can be used to find compliant networks with size less than or equal to the maximum allowable circuit size,  $max$ :

$$fitness = \begin{cases} -error; & n \leq max \\ -M; & elsewhere \end{cases} \quad (1)$$