COMPACT PARALLEL COUPLED-LINE BANDPASS FILTER WITH WIDE BANDWIDTH AND SUPPRESSION OF SPURIOUS

Kuo-Sheng Chin, Yi-Ping Chen, Ken-Min Lin, and Yi-Chyun Chiang
Department of Electronic Engineering, Chang Gung University, Tao-Yuan, Taiwan, Republic of China; Corresponding author: kschin@mail.cgu.edu.tw

Received 28 October 2008

ABSTRACT: The re-entrant coupling structure is applied in a parallel-coupled microstrip filter design to ensure a wide bandwidth, compactness, and the suppression of spurious. Fan design graphs are provided to determine the dimensions of each coupled stage. The features and advantages of the proposed circuit are analyzed in detail. Two experimental filters with design bandwidths of 70 and 90% are fabricated to verify the effectiveness of the modified structure. © 2009 Wiley Periodicals, Inc. Microwave Opt Technol Lett 51: 1795–1800, 2009; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24513

Key words: parallel coupled-line filter; re-entrant; wideband; spurious suppression

1. INTRODUCTION

Parallel-coupled microstrip bandpass filters have been extensively adopted in the RF front end of microwave and wireless communication systems for decades. Two shortcomings limit the range of practical applications of coupled-line filters. Traditionally, parallel-coupled microstrip filters suffer from spurious responses at twice the fundamental frequency, degrading the frequency selectivity. This spurious response (at 2fo) is caused by unequal even- and odd-mode phase velocities. Another major limitation on parallel-coupled filters arises from the weak lateral coupling between lines in a conventional structure. When a wide bandwidth is desired (>40%), small strip widths and spacings are required to ensure tight coupling, which is difficult to obtain accurately using a microstrip fabrication process.

In the literature [1–6], various approaches have been reported to overcome the aforementioned problems. The defected ground structure (DGS) [1] provides attenuation poles for a wide stopband. Some modified coupled-line structures, such as wiggly lines [2] and meandering lines [3], have been used to attenuate the spurious response at 2fo based on a phase velocity equalization procedure. Additionally, capacitors have been placed between the coupled lines [4] to compensate for the difference between even- and odd-mode phase velocities. To solve the problem of wide bandwidth, three-line structures [5] and ground-plane aperture techniques [6] can enhance the coupling between microstrips.

The re-entrant coupling structure [7], [8], which uses a floating ground plane conductor over parallel-coupled lines, is applied in the coupled design to increase coupling and improve directivity. The design method has been developed systematically in [9]. In this work, the re-entrant coupling configuration is adopted to design parallel-coupled microstrip bandpass filters, and the associated improvement in filter performance is described originally.

2. FAN DESIGN GRAPHS

Figure 1 schematically depicts in three dimensions the re-entrant coupled-line filter. Each coupled stage of the filter is covered by nontouching floating plate overlays, which are allowed to vary according to the filter design. Figure 2 presents the equivalent capacitance network of the ith re-entrant coupled stage of Figure 1. In Figure 2, Cmi is the capacitance between two microstrips, and Cm represents the capacitance of a single microstrip line to ground. C1 and C2 are the decreased capacitance due to the finite ground and the increased fringing capacitance of the floating plate to real ground under even mode excitation, respectively. The odd- and even-mode capacitances are simply given by

$$C_o = 2(C_{mi} + C_{mu}) + C_{ni} + C_{nu} - C_1$$

$$C_e = C_{pl} + \left(\frac{1}{C_{pu} - C_1} + \frac{1}{C_2}\right)^{-1}$$

The accuracy of the dimensions of the coupled stage which correspond to the required mode impedances are critical and the most important in implementing the filter. Based on the design method proposed in [9], Figure 3(a) plots the even- and odd-mode impedance design data concerning the proposed coupled lines on a structure of h1 = 0.762 mm, \(\varepsilon_r = 3.38\), tan \(\delta_1 = 0.0025\), h2 = 0.13 mm, \(\varepsilon_{ru} = 4.3\), and tan \(\delta_1 = 0.02\). This fan design graph plots the characteristic impedances as a function of strip width and line spacing ranging from 0.1 to 1.6 mm. The corresponding circuit dimensions can be easily determined from Figure 3(a) for the required mode impedances. Figure 3(b) analyzes the effect of a thicker overlay with the same parameters as in Figure 3(a), except with \(h_2 = 0.22\) mm.

3. FILTER FABRICATION

Two third-order Chebyshev filters were synthesized and fabricated at \(f_o = 5.8\) and 5.0 GHz with \(\Delta = 70\) and 90%, respectively, for demonstration. The 25 N substrate was selected as the lower substrate in both filters with \(\varepsilon_{ru} = 3.38\), tan \(\delta_1 = 0.0025\), and h1 = 0.762 mm. Both filters had the same designed ripple level of 0.5 dB. Full-wave EM software HFSS was used in the simulation.

3.1. Filter A (\(f_o = 5.8\) GHz and \(\Delta = 70\%\))

Based on the conventional synthesis formulae [10], the even- and odd-mode impedances for the given specifications were calculated as \(Z_{ru} = 126\) Ω and \(Z_{ru} = 43\) Ω for all coupled stages. The FR4 substrate with \(\varepsilon_{ru} = 4.3\), tan \(\delta_2 = 0.02\), and \(h_1 = 0.22\) mm was used as the overlay in the design, which has metal only on the top. In Figure 3(b), the corresponding line width and spacing are 0.37 and 0.3 mm, respectively. Figure 4(a) presents the photograph of filter A, in which the overlays adhered to the lower substrate with an adhesive. Figure 4(b) shows the simulated and measured responses of filter A, and both results are in good agreement. The measured bandwidth and insertion loss were 60% and 0.7 dB, respectively. The bandwidth decreased was mainly because of the inherent narrow band approximation of the conventional synthesis formulae [11]. Notably, the wide upper stopband was achieved and

- Figure 1 Three-dimensional schematic view of proposed coupled-line filter with floating plate overlays
the rejection level in the upper stopband exceeded 30 dB at $2f_o$.

Figure 4(c) indicates that the filter A has a small group delay in the passband, ranging from 0.3 to 0.6 ns. The length of this filter was about 27 mm, which is shorter than the conventional parallel-coupled filter (34 mm), owing to its enhanced total dielectric constant.

3.2. Filter B ($f_o = 5.0$ GHz and $\Delta = 90\%$)

The even- and odd-mode impedances for filter B were $Z_{0e1} = Z_{0e4} = 141.32 \, \Omega$, $Z_{0o1} = Z_{0o4} = 47.22 \, \Omega$, $Z_{0e2} = Z_{0o2} = 160.51 \, \Omega$, and $Z_{0o3} = 53.66 \, \Omega$. The FR4 substrate with a thickness of 0.13 mm was used as the overlay. In Figure 3(a), the corresponding dimensions were $w_1 = 0.22$, $s_1 = 0.44$, $w_2 = 0.2$, and $s_3 = 0.35$ (all in mm). Figures 5(a) and 5(b) show a photograph of filter B and plot its responses. The simulated and measured bandwidths were 72% and 70%, respectively, which are very difficult to achieve using the conventional coupled-line structure. The insertion loss was 0.9 dB and the group delay variation was less than 0.5 ns over the passband.

4. CHARACTERISTIC ANALYSIS

This section describes the features of the proposed circuit configuration and its advantages over the conventional coupled line configuration.

4.1. Compact Size

Figure 6(a) plots the total dielectric constant of the proposed structure normalized with $\varepsilon_{r1}$ versus the ratio $\varepsilon_{r2}/\varepsilon_{r1}$, where $\varepsilon_{ru}$ and $\varepsilon_{rl}$ are the dielectric constants of the upper and lower substrates, respectively. Both $\varepsilon_{ru}$ and $\varepsilon_{rl}$ affect the total dielectric constant of the proposed structure. The range of the dielectric constant ratio, $1 \leq \varepsilon_{r2}/\varepsilon_{r1} \leq 3$, is set greater than one purposely to enhance coupling, and the thicknesses of the upper and lower substrates are chosen as 0.13 and 0.762 mm, respectively. It is found that the thickness of floating plates only weakly affects the total dielectric constant when $h_{un}/h_{ul} \leq 0.5$. Two curves with different dielectric constants of $\varepsilon_{r1} = 3.38$ and 6.15, respectively, are utilized to analyze the effect of using different lower substrates. From Figure 6(a), the total dielectric constant increases with $\varepsilon_{ru}/\varepsilon_{rl}$, such that the denser floating plates with a higher $\varepsilon_{ru}$ induce a larger total dielectric constant. Notably, the two curves reveal different proportional increases in the total dielectric constant, indicating that $\varepsilon_{ru}$ more strongly affects the lower substrate with small $\varepsilon_{rl}$.

Figure 6(b) plots the length of the proposed coupled stage versus $\varepsilon_{ru}/\varepsilon_{rl}$, normalized to the conventional coupled line length. As presented in Figure 6(b), the length ratio is less than 85% revealing that the proposed structure has a reduced size. A higher $\varepsilon_{ru}/\varepsilon_{rl}$ ratio is associated with a larger size reduction ratio, because of the enhancement of the total dielectric constant. The largest reduction ratio of 0.63 in Figure 6(b) is achieved when $\varepsilon_{ru}/\varepsilon_{rl} = 2.96$. 

Figure 2 Equivalent capacitance network of $i$th re-entrant coupled stage

Figure 3 (a) Even- and odd-mode characteristic impedances of the proposed structure with $h_u = 0.762$ mm, $\varepsilon_{ri} = 3.38$, $\tan \delta_{ri} = 0.0025$, $h_l = 0.13$ mm, $\varepsilon_{ro} = 4.3$, and $\tan \delta_{ro} = 0.02$. (b) Even- and odd-mode impedances ($h_u = 0.22$ mm)
Figure 4  (a) Photograph of experimental filter A with $\Delta = 70\%$. (b) Simulated and measured responses of filter A. $w = 0.37$ mm and $s = 0.3$ mm. (c) Simulated and measured group delay. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
Figure 5  (a) Photograph of filter B ($\Delta = 90\%$). (b) Simulated and measured responses. $w_1 = 0.22$, $s_1 = 0.44$, $w_2 = 0.2$, and $s_2 = 0.35$ (all in mm). (c) Group delay. [Color figure can be viewed in the online issue, which is available at www.interscience.wiley.com]
4.2. Released Line Spacing Limitation
Because the re-entrant structure can increase the coupling between microstrips, the line spacing can apparently be released in a filter with a wide bandwidth. Figure 7 plots the fan design graphs of coupled lines: the dotted curve is obtained using the conventional structure and the solid graph is calculated with the proposed structures. Notably, the substrate parameters of the proposed structure in Figure 7 are the same as in Figure 3(b), whereas the conventional structure only has the lower substrate. The solid circles represent the required line width and spacing of filter A for each coupled stage. For the very wide design bandwidth of 70%, if the conventional structure is used, the minimum line spacing must be less than 0.07 mm, as shown in Figure 7, and is difficult to realize because of the constraints of the microstrip fabrication process. Fortunately, the line spacing exceeds 0.3 mm when the proposed structure is adopted, advantageously relaxing the limitation on the line spacing. Similarly, the minimum line spacing of filter B is 0.2 mm and can still be realized by general microstrip fabrication.

4.3. Suppression of Spurious
As presented in Figures 4(b) and 5(b), the rejection levels of filters A and B exceed 30 dB and 25 dB at $2f_o$, respectively, indicating the favorable performance of spurious suppression. This fact can be explained as follows. From (1), the odd- and even-mode capacitances are increased to $(2C_{o} + C_{p} - C_i)$ and $[1/(C_{o} - C_i) + 1/C_i]^{-1}$, respectively by the floating plate overlay. Because (1a) and (1b) have the same $(C_{o} - C_i)$ term, $C_i$ clearly exceeds $C_o$ because (1b) has the term $1/C_i$ in the denominator, leading to a small summation. The larger odd-mode capacitance has a longer phase length than the even-mode capacitance. Moreover, because the phase velocity of the odd-mode exceeds that of the even-mode, the increased phase length just compensates for the difference between the odd- and even-mode phase velocities. Thus, the spurious response of the coupled line filter at $2f_o$ can be reduced by the phase velocity equalization.

5. CONCLUSIONS
Three major advantages of the re-entrant coupled-line filters—wide bandwidth, compact size, and wide upper stopband—are achieved simultaneously by using the re-entrant structure in filter design. Fan design graphs are given to determine accurately the dimensions of each re-entrant coupled stage. The validity of the proposed structure was experimentally confirmed by realizing two Chebyshev filters with wide bandwidths. The characteristics and improvements of the modified coupled-line bandpass filters are investigated in detail.

ACKNOWLEDGMENTS
The authors would like to thank the National Science Council, Taiwan, R.O.C., (Contract No. NSC 96-2221-E-182-002) and Chang Gung University (Contract No. UERP270031) for partially supporting this research.
ABSTRACT: An optical extinction ratio monitoring technique for wave-length-division-multiplexed passive optical network systems is proposed. This technique is based on the optical spectrum analysis. It can accurately measure the optical extinction ratio of optical signals and does not change with the transmission distance. In our experiments, the transient chirp hardly affects the measured results if the optical extinction ratio is lower than 7 dB, and the measured errors are less than 0.5 dB. © 2009 Wiley Periodicals, Inc. Microwave Opt Technol Lett 51: 1800–1803, 2009; Published online in Wiley InterScience (www.interscience.wiley.com). DOI 10.1002/mop.24512

Key words: optical extinction ratio; optical spectrum analysis; optical performance monitoring; passive optical network (PON); wavelength-division-multiplexed passive optical network (WDM-PON)

1. INTRODUCTION

Passive optical networks (PONs) are now in widespread use, especially for broadband access networks. The latest development in PON systems is evolving toward higher data rates and long-reach transmission distances. Therefore, it is becoming attractive to use wavelength division multiplexing (WDM) technologies. Several wavelength-division-multiplexed passive optical networks (WDM-PONs) that utilizes the dense-WDM technology have been demonstrated to increase the transmission capacity and provide flexible services for the next generation networks [1, 2].

To make WDM-PONs practical in broadband access networks, low-cost light sources are the key challenges. Various colorless light sources like injection-locked lasers and the remodulated semiconduct optical amplifiers (SOAs) can be utilized in optical network units (ONUs). Alternative light source such as DFB-lasers and optical modulators can be used to provide much higher data rates up to 10 Gbps [3, 4].

For WDM-PON systems, the installation of an optical performance monitoring unit is very important to provide quality of service and ensure survivability. The performance monitoring units for WDM-PONs are generally built in the optical line terminal (OLT) to reduce cost. An optical spectrum analyzer (OSA) is frequently applied to monitoring channel wavelength and fiber link loss [5, 6]. In this article, we propose an optical signal extinction ratio (ER) monitoring technique using an OSA to monitor both the upstream and downstream signals. This method does not require the synchronization and clock recovery that a time-domain waveform monitoring usually needs. This method is data-rate insensitive. It is also independent of the transmission distance, so it is very suitable for monitoring optical signals remotely.

2. OPERATION PRINCIPLES

The monitoring of ER for a directly modulated DFB laser (DML) signal can be obtained from an OSA. It is based on the chirp-induced separation of the spectral peaks for different data levels. It is well known that the chirp $\Delta v(t)$ of a DML is related to the laser output optical power $P(t)$ by the expression [7] as follows:

$$\Delta v(t) = \alpha \frac{d}{dt} \left[ \ln(P(t)) \right] + kP(t),$$

(1)

where $\alpha$ is the linewidth enhancement factor and $k$ is the adiabatic chirp coefficient [7]. The first term in (1) is the transient chirp, whereas the second term is a structure-dependent adiabatic chirp [7]. When the chirp of a DML is dominated by the adiabatic chirp, the output power waveform resembles the output chirp waveform. The chirp waveforms of a DML exhibit damping oscillations and large frequency difference between steady-state “1”s and “0”s, so the two optical spectral peaks of bit “1” ($P_1$) and “0” ($P_0$) can be clearly observed by using an OSA. The two spectral peaks represent the average powers of bit “1” ($P_1$) and “0” ($P_0$), respectively. Therefore, the value of ER (dB), i.e., the power ratio of $P_1$ over $P_0$ can be directly read from an OSA. By this scheme, the values of ER for multiple channels can be easily obtained from an optical spectrum.

3. EXPERIMENTAL RESULTS

The experimental setup is shown in Figure 1. To verify the feasibility of this ER monitoring technique, a directly modulated DFB laser diode (NEL, NLK5CSE2KA) was adopted as the DML light source and was modulated by a 2$^7$:1 pseudorandom bit sequence (PRBS) with data rates of 2.5 or 10 Gbps. The driving current amplitude was 40 mA. The resolution bandwidth (RBW) of the OSA was set at 0.06 nm. The measurements are carried out for