**Stripline Combline Filter**

**Tunes to 1300 MHz**

With the aid of 2.5D and 3D computer simulations, a novel compact stripline combline bandpass filter was designed and fabricated with a tunable passband across a wide frequency range.

The bandwidth and response of a combline filter is governed by the coupling of each resonator to its immediate neighbor. This is also a function of the resonator size, resonator spacing, and ground-plane separation. Different components and circuit elements are added to a combline BPF structure to tune the passband frequency, including PIN diodes, ferroelectric diodes, RF microelectromechanical systems (RF MEMS) devices, barium strontium titanate (BST) varactor diodes, and silicon/GaAs varactor diodes.

Microstrip has been the basis for many tunable combline filter designs. In Ref. 14, a tunable planar microstrip combline filter with multi-resonator coupling is realized to achieve high selectivity. In Ref. 15, a combline bandpass filter using step-impedance microstrip lines is investigated. Ref. 16 detailed a 1.4- to 2.0-GHz miniature filter with high linearity. Reference 17 offers results on a tunable RF MEMS microstrip filter with constant bandwidth by means of corrugated coupled resonators, while Ref. 18 features a tunable combline bandpass filter loaded with series resonator.

Reference 19 details a compact dual-mode second-order filter with constant bandwidth, while Ref. 20 presents a high-quality-factor (high-Q) fully reconfigurable tunable microstrip bandpass filter. Reference 21 details a tunable microstrip filter design with independent frequency and bandwidth control.

All of the above examples were based on microstrip. Few researchers have investigated stripline comb-line filters, although such filters can achieve much better bandwidth than their microstrip counterparts, with excellent isolation between adjacent traces.

To explore the performance capabilities of stripline combline filters, a five-pole Chebyshev stripline combline filter with tunable center frequency was designed and simulated. Tuning is performed by means of varactor diodes and capacitors that vary as a function of applied voltage. The filter provides tuning capability over a wide range of frequencies while maintaining a nearly constant fractional bandwidth and high stopband rejection.

Figure 1 depicts a simple schematic diagram of the prototype Chebyshev lowpass filter. For the required passband ripple, $L_{min}$ in dB, and the minimum stopband attenuation, $L_{min}$ in dB, at a stopband characteristic impedance of $Z_0$, in $\Omega$, the degree of the prototype lowpass Cheb filter that will meet the required specifications can be found by Eq. 1:

$$L_{min} = 20 \log_{10} \frac{Z_0}{L_{max}}$$

Where $L_{max}$ is the maximum allowed passband ripple and $Z_0$ is the characteristic impedance of the transmission line.
n ≥ \cosh^{-1}\left(\left[(10^{0.1L_{as}} - 1) /(10^{0.1L_{ar}} - 1)\right]^{0.5}\right) / \cosh^{-1}(\Omega_s) \tag{1}

By assuming that the minimum stopband attenuation at \Omega_s = 2\Omega_{cutoff} = 2 and the passband ripple have been chosen as 50 dB and 3 dB, respectively, the degree of the Chebyshev lowpass prototype is calculated to be n = 5. The lowpass prototype parameters that were calculated from Eq. 1 can be found in the table.

Figure 2 shows the general structure of a tapped combline filter. To design a combline BPF, the following parameters are also required: center passband frequency (F_0) and passband bandwidth (BW). The filter is designed for a BW of 100 MHz and a F_0 of 1,200 MHz.

To implement the proposed filter, quarter-wave transformers were employed as impedance inverters. Using parameters g_i calculated previously, it is possible to obtain the coupling coefficients of adjacent resonators M_{i, i+1} and the external quality factors (Qs) of the resonators at the inputs and outputs as^1:

\[ M_{i, i+1} = \frac{\text{BW}}{F_0(g_{i}g_{i+1})^{0.5}}, \quad i = 1, 2, 3, 4 \tag{2} \]

\[ Q_{e1} = \frac{F_0g_{0}g_{1}}{\text{BW}}, \quad Q_{e5} = \frac{F_0g_{5}g_{6}}{\text{BW}} \tag{3} \]

From the above equations, M_{1, 2} = M_{4, 5} = 0.0511; M_{2, 3} = M_{3, 4} = 0.0448; and Q_{e1} = Q_{e5} = 41.7804. The prototype stripline combline filter was designed on a circuit laminate with relative dielectric constant of 3.38 and thickness of 0.98 mm (Rogers Corp.’s RO4003C material). A line width was established at W = 0.55 mm with a line length of L = 20 mm and trace thickness t = \mu m for all of the filter line elements, except for the terminating lines. The terminating lines measured 0.93 mm wide, which matches a terminating impedance of 50 \Omega.

The coupling coefficient of the two adjacent resonators can be calculated by Eq. 4:

\[ M = \frac{F_H^2 - F_L^2}{F_H^2 + F_L^2} \tag{4} \]

where \(F_H\) and \(F_L\) are representing the frequencies of the even- and odd-mode oscillations, respectively, in a system that incorporates two coupled resonators. These correspond to the clearly pronounced peaks in the attenuation characteristic for the dual resonator circuit.

By changing the spacing between the resonators and calculating the coupling coefficient using Eq. 4, the design curve of the coupling coefficient versus spacing, s, can be obtained. The dimensions of the spacing required to achieve the design requirements for the resonant circuit for the design parameters can then be found from a curve of coupling coefficient versus s (Fig. 3) as: s_{1, 2} = s_{4, 5} = 0.45 mm and s_{2, 3} = s_{3, 4} = 0.53 mm.

7. These simulated responses for the five-pole filter at 913 MHz show (a) S_{21} and (b) S_{11}.
The external Q of the filter can also be extracted from Eq. 5:

\[ Q_e = \frac{\omega_0}{\Delta \omega_{\text{eq}}} \]  

(5)

where \( \omega_0 \) is the resonant frequency of the filter and \( \Delta \omega_{\text{eq}} \) is the absolute bandwidth between ±90-deg. points in the phase of \( S_11 \). By changing the taping position \( y \) and calculating the external Q of the filter using Eq. 5, it is possible to generate a design curve of external filter Q versus \( y \). With this curve (Fig. 4), the required taping position is obtained as \( y = 0.30 \) mm.

For tuning capability, varactor diodes were added to the proposed filter as variable capacitors, in place of the short-circuited end of the combline filter. Due to the small size of the proposed filter and fabrication issues involving small circuit dimensions, there is insufficient space to add varactor diodes in place of the short-circuited end of the coupling lines without interference. Therefore, the ends of the coupling lines are bent so that they do not interfere with each other after adding the varactor diodes. Figure 5 shows the modified structure of the combline filter after the coupling lines have been bending the end of the couple lines. A bending angle of 45 deg. was chosen for minimum impact on the design.

**SIZING UP SIMULATIONS**

The filter circuit’s performance possibilities were simulated with the Advanced Design System (ADS 2009) computer-aided-engineering (CAE) simulation software from Agilent Technologies (now Keysight Technologies; www.keysight.com). With the aid of the modeling software, the dimensions of the filter and the capacitance values were optimized with the following goals: The filter passband loss should be less than 2 dB and the filter stopband rejection should be more than 30 dB.

Figure 6 shows the simulated frequency response of the bandpass filter. As displayed in Fig. 6, the filter has a bandwidth of 100 MHz with center frequency (\( f_0 \)) of 1,200 MHz. The stopband rejection level and passband ripple of the filter are about 50 dB and 3 dB, respectively, which are in keeping with the design goals.

To demonstrate filtering tuning, a constant capacitance value (C) was added to the capacitance of each varactor diode. Fig. 7 shows the simulated filter frequency response for different values of C. As the plots show, by changing the capacitance of the varactor diodes, the center frequency of the filter can be tuned across a frequency range of 900 to 1,300 MHz with good impedance matching.

**Figure 8** shows the filter’s fractional bandwidth as a function of capacitance, C. The fractional bandwidth of the filter (about 9%) is very stable when tuning the center frequency, due to the constant coupling coefficient of adjacent resonators during tuning.

The simulation results were achieved using two- and one-half-dimension (2.5D) ADS simulation tools. Although more time-consuming, a three-dimensional (3D) software solver was needed for greater simulation accuracy. But even a 3D solver may not properly predict the effects of certain circuit structures, such as vias holes.

For improved accuracy, the High-Frequency Structure Simulator (HFSS) full-wave electromagnetic (EM) simulator from Ansys (www.ansys.com) was used in the filter simulations. Figure 9 shows a 3D HFSS schematic diagram of the stripline combline BPF, with each via-hole modeled as a cylindrical structure.

As Fig. 10 shows, the presence of via-holes will change the frequency response of the combline BPF. As an example, the 3D-modeled filter has a narrower transition band than simulations from the 2.5D ADS analysis. Still, the effects of...
the vias on filter frequency response are negligible, meaning that the 2.5D simulations are still valid and accurate.

A hardware version of the filter was fabricated on RO4003C laminate using surface-mount-technology (SMT) packaged varactor diodes to provide the variable capacitances needed for filter passband tuning (Fig. 11). Unfortunately, these diodes contribute to the losses of the filter and, for some applications, an additional amplifier may be needed to compensate for undesirable passband losses.

The filter was characterized (Fig. 12) with a model FSL6 spectrum analyzer from Rohde & Schwarz (www.rohde-schwarz.com). The filter tuning range from 0.9 to 1.3 GHz had a nearly constant fractional bandwidth (about 9%). The rejection level at both upper and lower stopbands was about 50 dB, and measured passband ripple was about 3 dB. The frequency range and response measured with the spectrum analyzer was found to be quite similar to the performance predicted by the computer simulations.

REFERENCES