

The Design and Integration of a Quasi-Distributed Tunable UHF Filter

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Abstract

A novel quasi-distributed filter design is used to vertically integrate a tunable UHF filter in a polymer based substrate. The filter is a sub band of a widely tunable narrowband preselect filter array and makes use of unused substrate space within the array to increase the overall tuning bandwidth of the array while keeping the substrate volume constant. This paper will discuss the use of mixed three dimensional electromagnetic and SPICE simulation in the design of the tunable filter. The fabrication of the filter using standard PCB fabrication techniques will be described and measured performance of the filter will be compared to the simulated response.

High inductor and capacitor Q's were required to meet insertion loss specifications over the entire tuning range of the filter. The design and simulation of vertically integrated coupled parallel plate inductors and the careful selection of the combination of static capacitors and tuning varactors to achieve the desired tuning and filter response will be highlighted.

1.0 Introduction

Next generation cognitive radios will require high quality, tunable filters capable of tuning across several octaves in frequency. Current Si technology is capable of providing several RF transceivers integrated on a single chip allowing the user to transmit and receive simultaneously on multiple frequencies. To take full advantage of such capabilities, low loss preselect filters must be employed to cover multiple frequency bands. Recently there has been considerable research in widely tunable filter arrays to meet these needs. Filters have been constructed based upon MEMS technology [1-3] ferroelectric thin films [4], and piezoelectric actuators [5-7]. Although the filters presented in [5-7] represent the state of the art in narrowband widely tunable preselect filters based on piezoelectric actuation in the 800 MHz to 6 GHz band, it is desirable to extend the tuning range of these filters into the UHF region without increasing the overall size of the array substrate. This can be accomplished by using unused portions of the filter substrate as shown in Figure 1. This paper will discuss the vertical integration of high Q inductors within this unused space as well as the careful selection of static and variable capacitors that preserve high Q performance. The goal of the filter is to provide a 25 MHz bandwidth filter that tunes from 650 MHz to 800 MHz with a maximum insertion loss of 3 dB. Section 2.0 covers the filter topology chosen for the filter design. Section 3.0 and Section 4.0 discuss the design of the coupled inductor section and

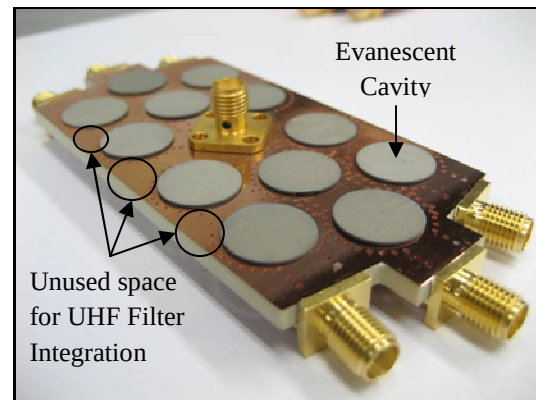


Figure 1. Widely Tunable Evanescent Mode Filter Bank

tuning capacitors respectively. Simulated and measured performance of a fabricated filter is presented in Sections 5.0 and 6.0.

2.0 Filter Design

The filter topology chosen for this design was a two pole, coupled resonator design shown in Figure 2. Capacitors C_2 and C_3 are high Q variable capacitors and are used to change the center frequency of the filter and capacitors C_1 and C_4 provide variable matching capability. Inductors L_1 and L_2 are vertically integrated loop inductors that provide a high Q inductance for the filter resonators while the coupling between the inductors, k , determines the bandwidth of the filter.

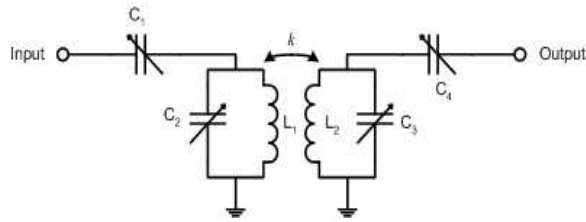


Figure 2. Lumped Element Coupled Inductor Filter Model

3.0 Coupled Inductor Design

To maintain a high resonator Q for low loss filter performance, inductor Q 's in the order of 300 or more were required to meet the insertion loss specifications for the given filter bandwidth. Traditional SMT components are not capable of providing such a high Q thus a novel distributed inductance was used for the filter design. The coupled inductors were vertically integrated in the unused space between the evanescent resonator elements of the high frequency filter array.

The coupled inductor section used in the tunable UHF filter design is shown in Figure 3, and consist of two magnetically coupled loops each created by two vertical vias that form the sides of the loop, and strip of copper on the bottom of a printed circuit board that connects the two vertical vias creating the bottom of the inductive loop. Ansoft's HFSS simulation software was used for the design and characterization of the coupled inductor section with individual inductances optimized for inductance and Q . The spacing between the loops ϕ determined the coupling coefficient and hence the bandwidth of the filter as given in Equation (1). Given that the ratio of plate separation to width is large, parallel plate approximations for inductance and Q become inaccurate and inductance and Q are best determined by full wave simulation and Equations (2) and (3).

$$k = \frac{BW}{f_0} \quad (1)$$

Where BW is the bandwidth of the filter and f_0 is the filters center frequency.

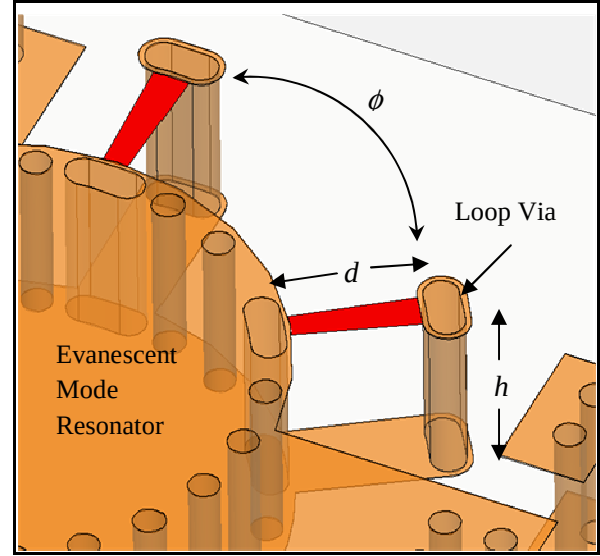


Figure 3. HFSS Model of the Coupled Inductor Section

A useful design strategy was to maximize the inductance of the filter resonator thereby decreasing the required change in capacitance needed to cover the tuning range of the filter. This can be done by increasing the distance d between the vertical vias and/or increasing the height h of the inductive loop. However, for this design, the distance d was limited to the spacing between other evanescent resonator elements in the substrate, and the height of inductor loop was determined by the thickness of the substrate. For the filter design discussed in this paper, d was limited to 3.5mm and the height of the loop h was 3.125mm. Using the full wave simulation results and equations (2) and (3) an inductance of 4.9 nH with a Q of 320 at 650MHz was obtained yielding a sufficiently high Q for the resonator inductor.

$$L_{eff} = \frac{1}{\omega(im(Y_{11}))} \quad (2)$$

$$Q \triangleq \frac{|X_l|}{R_l} = \frac{\omega L_{eff}}{R_l} = \frac{im(Zin)}{re(Zin)} \quad (3)$$

The inter-resonator coupling, k , is provided via magnetic field coupling as shown in Figure 4. The required k of 0.031 was determined using Equation (1) at a center frequency of 800 MHz and a 25 MHz bandwidth. The simulated coupling coefficient was then determined via full wave simulations using Equations (5) and (6). A parametric sweep was run using HFSS by changing the distance, ϕ , between the loops from 50 degrees to 65 degrees.

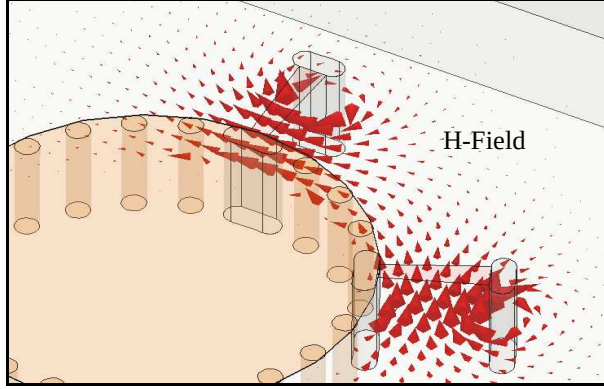


Figure 4. Magnetic Field Plot of the Coupled Inductor Loops

$$K = \frac{M}{L_{eff}} \quad (4)$$

Where

$$M = \frac{im(Z_{12})}{\omega} \quad (5)$$

The results of the simulation are shown in Figure 5, and indicate that at a separation of approximately 64 degrees, the proper coupling coefficient can be obtained. A 3D full wave simulation was completed with this spacing and the results were exported to be used in the lumped element filter simulation discussed in Section 5.0.

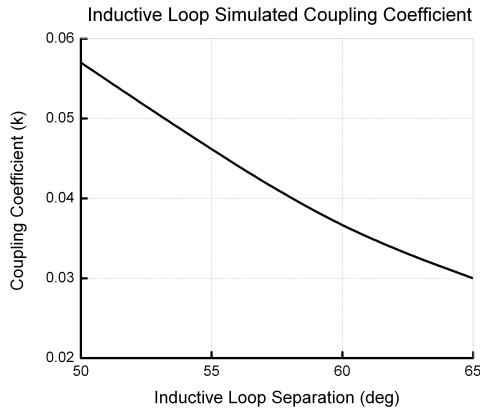


Figure 5. Coupling Coefficient Simulation Results

4.0 Capacitor Selection

Using Equation (6), the capacitances for the required tuning range were then determined to be a maximum of 12pF for 650 MHz center frequency and a minimum of 7.5pF for 800 MHz center frequency.

$$f_{Res} = \frac{1}{2\pi\sqrt{L_{Res}C_{Res}}} \quad (6)$$

Although the change in capacitance required to cover the entire tuning range could be provided by a single high-Q GaAs varactor, the overall resonator Q determined by the parallel combination of a GaAs varactor and the high Q inductive loops described in Section 3.0 would be too small to meet the filter specifications. Therefore, the capacitor configuration shown in Figure 6 was used for the variable capacitor section of the resonator. Given this configuration, it is advantageous to employ a varactor with a wide tuning range and small series resistance in order to maximize the improvement in overall Q provided by the C_2 , D_1 series branch.

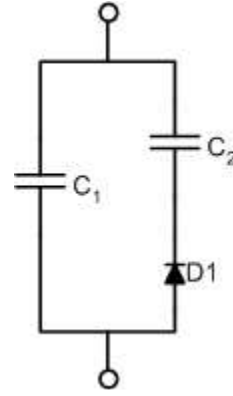


Figure 6. Filter Resonator Capacitor Tuning Section

The bulk of the resonator capacitance is provided by ultra high Q static capacitors C_1 and C_2 . C_2 also provides a DC block for the bias voltage applied to varactor D_1 . Since the Q of a component is defined as the ratio of energy stored to energy loss in the component, it is desirable to have most of the field storage in the ultra high Q static capacitors and only a fraction of the field storage in the lossy varactor. The relationship between energy stored between components in a series capacitance branch is shown in Equation (7). Thus the varactor should be made as large as possible and the combination of C_1 , C_2 , and D_1 should be chosen to provide the necessary change in capacitance to tune the filter over the desired band. For this filter Microsemi model number MV31024 GaAs varactor was chosen and C_1 and C_2 were 5.6 pF and 7.5 pF static capacitors respectively.

$$U_{C2} : U_{D1} = \frac{C_{D1}}{C_{C2}} \quad (7)$$

Where U_x is the energy stored in the component.

5.0 Simulated Filter Results

The final filter model is shown in Figure 7. Input and output varactors. Microsemi model number MV31022, and DC blocking capacitors were added to provide a variable matching circuit for the filter design. 85 k Ω bias resistors were included to prevent RF energy from leaking into the DC bias supplies. The box labeled “Coupling Section” was the linear S-parameter output of the HFSS simulation of the coupled inductor loops. SPICE models for all varactors were created using manufacture’s datasheets and the filter was simulated using various biasing voltages to adjust the center frequency from 650 MHz to 800 MHz.

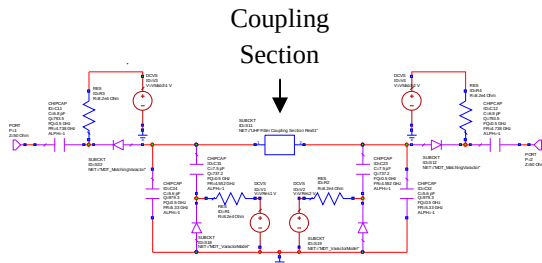


Figure 7. Final Coupled Inductor Filter Design

The results of the simulations are shown in Figure 8, and summarized in Table 1. Table 1 shows that the desired 25 MHz bandwidth is achieved at approximately 800 MHz. center frequency. However, as the filter is tuned to lower frequencies, the bandwidth of the filter decreases to approximately 23 MHz at a center frequency of 650 MHz. This decrease in bandwidth can be attributed to the near constant coupling coefficient across the frequency band of interest. As described in Section 3.0, the coupled inductors were designed to have the required coupling coefficient at 800 MHz. However, since k does not increase with a decrease in frequency, a coupling coefficient of 0.032 at 650 MHz would yield a lower filter bandwidth as indicated in the simulation results.

Center Frequency (MHz)	Insertion Loss (dB)	Bandwidth (MHz)
650	2.5	22.5
725	1.9	23.8
800	1.7	26.1

Table 1. Simulated Filter Performance Summary

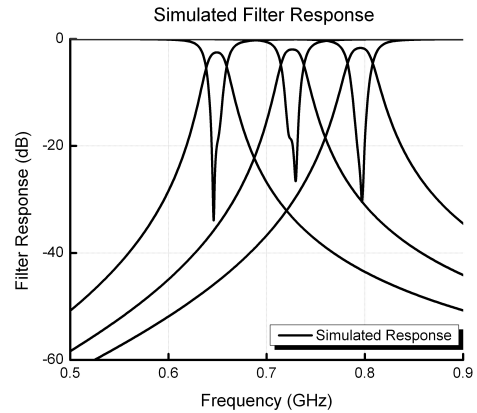


Figure 8. Simulated Filter Performance

6.0 Measured Performance

Once the filter performance was verified using lumped element simulations, the filter was fabricated using standard PCB processes on Rogers TMM3 substrate. A picture of a fabricated filter is shown in Figure 9. SMA connectors were soldered to the input and output of the filter to facilitate testing with an Agilent PNA. The measured results across the band of interest are shown in Figure 10, and show close agreement to the simulated response. There is a slight increase in filter bandwidth for all frequencies that can be attributed to excess coupling between resonators not accounted for in the simplified full wave simulation discussed in Section 3.0.

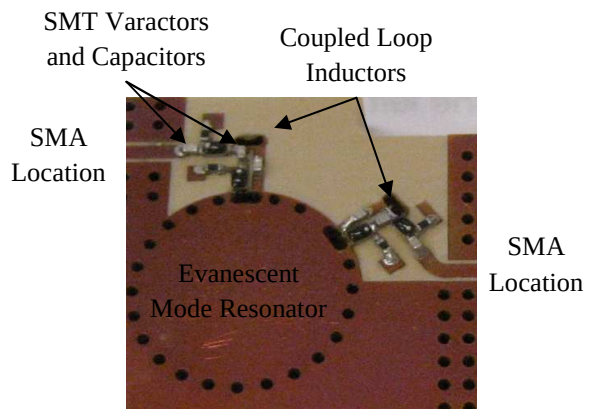


Figure 9. Fabricated Filter

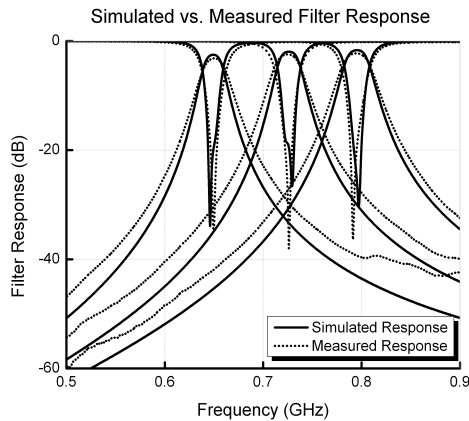


Figure 10. Measured Filter Response

7.0 Conclusions

It was shown that a 2-pole, quasi-distributed filter could be integrated in the unused space of a large tunable filter array. By utilizing this unused substrate, the tuning range of the filter bank was increased without increasing the overall size of the substrate. By integrating high Q inductive elements and carefully selecting capacitive tuning elements, a low loss tunable UHF filter suitable for use as a preselect filter has been created.

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