Compact Quasi-Lumped Element HTS Microstrip Filters

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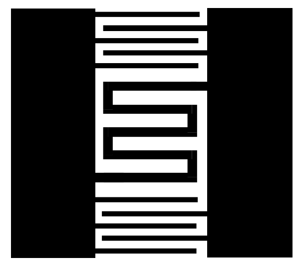
***Abstract* - This paper demonstrates the design of two quasi-elliptic and one linear phase filter. Different methods of cross coupling, which result in different advanced filtering characteristics, are investigated and verified experimentally. Each of these filters is designed to have a fractional bandwidth of 15MHz and a centre frequency of 1.7775GHz. All the filters are in basic quadruplex form and can be used for higher order filter design. These filters are good candidates for wireless and mobile communication systems.**

I. INTRODUCTION

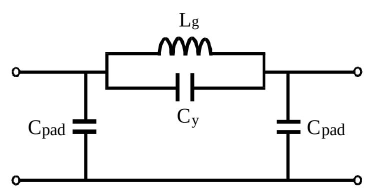
Mobile communication is placing more and more challenging requirements for high performance microwave components. High temperature superconductor (HTS) enables the design of very small microwave filters with low loss and good performance. Various high temperature superconducting microwave distributed filters with excellent performance, have been reported. Microwave filters with quasi-elliptic filtering characteristics [1]-[6] are the most popular. Compared with Chebyshev filters, the number of resonators used to achieve the same selectivity is smaller. Consequently, the filter size and weight as well as the resistive losses are reduced. This paper reports the design and implementation of a few very compact quasi-lumped element filters with quasi-elliptic and linear phase responses.

The new resonator structure consists of an interdigital capacitor in parallel with a meander-line inductor, with a shunt capacitor to earth at both ends. This is shown in Fig. 1. This structure is more compact than the resonator structure reported in [7] and the design theory is simpler. The design of the resonator structure will not be discussed here as this can be found in [8].

For the non-adjacent resonators cross couplings, at least two different approaches, to realize transmission zeros at finite frequencies, have been reported [1,4,5,6]. One uses transmission line cross coupling as demonstrated in [4]-[5] and the other uses magnetic cross coupling [1,6]. In this paper, both methods will be implemented. When transmission line cross coupling is used, it is also easy to implement a linear phase filter. The structure of this filter is very similar to that of quasi-elliptic filter but with the coupling configuration being modified in order to realize the transmission zeros at imaginary frequencies.



(a)



(b)

Fig. 1. Resonator structure and equivalent circuit.

II. FILTER SYNTHESIS AND CONFIGURATIONS

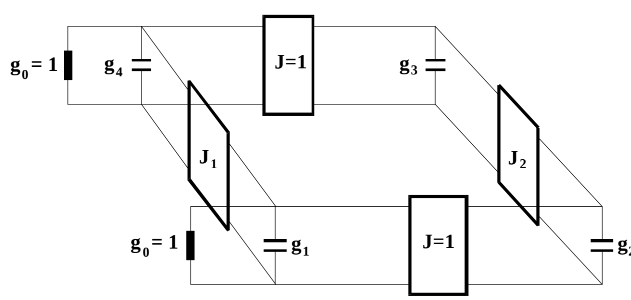


Fig. 2. Low pass prototype network.

The synthesis of quasi-elliptic and linear phase filters starts from a low pass prototype (Fig. 2). Four shunt capacitors are coupled by admittance inverters as shown. These admittance inverters and the low pass prototype capacitor values can be synthesized using [9]. For the quasi-elliptic filter, J_1 and J_2 have opposite signs whereas for linear phase filter, the signs are the same.

The next stage is to solve the bandpass prototype network which is depicted in Fig. 3.

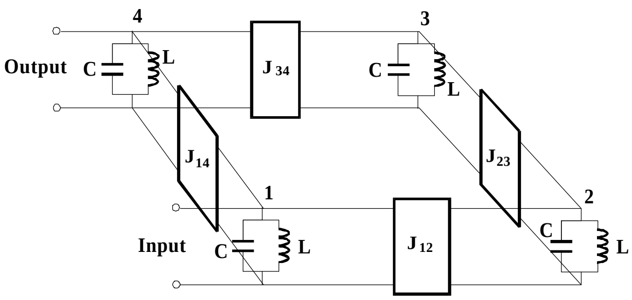


Fig. 3. Bandpass prototype network.

For the LC parallel resonator, an appropriate inductor, L value is chosen. Based on [10] and [11], the design formulas for the central inverter, J_{23} and the cross-coupled inverter, J_{14} are given as :-

$$J_{23} = \frac{\mathrm{FBW} \cdot J_2 \cdot \sqrt{b_2 \cdot b_3}}{\sqrt{g_2 \cdot g_3}} \qquad (1)$$

$$J_{14} = \frac{\mathrm{FBW} \cdot J_1 \cdot \sqrt{b_1 \cdot b_4}}{\sqrt{g_1 \cdot g_4}} \qquad (2)$$

where FBW is the fractional bandwidth of filter, and *b* is the susceptance slope parameter of the resonator. The J values of the other admittance inverters can be calculated using formulas found in [10]. Because of the symmetry, calculation of one side of the network is sufficient. J is represented by a π capacitive structure (positive series capacitor with a negative capacitor to earth at each end) with the capacitor values equal to J/ω_0. The negative capacitances are absorbed into the adjacent resonator capacitors. The input and output matching circuit design can be found in [8]. Because of lack of proven processes to make via-holes on substrates to realize the inductor to ground, we use π-resonator as shown in Fig. 1 instead of LC parallel resonator. By equating the resonant frequency of the π-resonator and the LC parallel resonator, and their susceptance slope parameters, *b*, the relationship between L_g, C_y, C_{pad} and L, C are given as

$$L_g = 4 \cdot L \qquad (3)$$

$$C_{pad} = \frac{C}{2} - 2 \cdot C_y \qquad (4)$$

The positive and negative values of the cross coupling, J_{14} are achieved by using different arrangement of the coupling line.

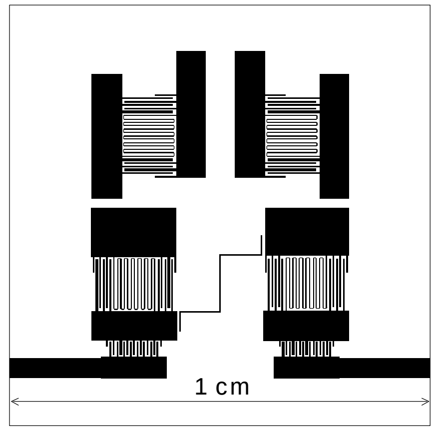


Fig. 4. Quasi-elliptic filter layout.

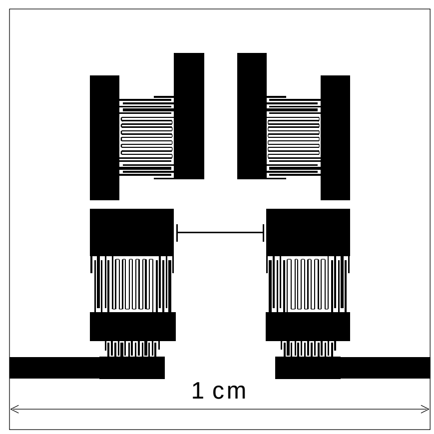


Fig. 5. Linear phase filter layout.

Fig. 4 shows the layout of the quasi-elliptic microstrip filter. The cross coupling is obtained by a short transmission line as shown, and provides 180 degree phase difference (negative J_{14}) as compared with coupling between resonators 2 and 3 (positive J_{23}). This gives a pair of transmission zeros located at filter passband edges in order to improve the filter selectivity. The small coupling gaps between the transmission line and the resonators control the location of the transmission zeros. The smaller the coupling gaps, the closer the transmission zeros are to the passband edges. The overall effect of the capacitances of the two narrow gaps should be a capacitance with magnitude J_{14}/ω_0. The arrangement of the resonators 1 and 4 not only reduces the filter size, but also keeps the coupling line as short as possible.

By changing the coupling position at resonator 1, as shown in Fig. 5, one can realize the linear-phase filter in which phase non-linearity is reduced. This filter has the advantage of flatter group delay at passband frequencies at the cost of less selectivity. The coupling between resonators 1 and 4 is in-phase with the coupling between resonators 2 and 3.

The input and the output port of the filters are designed to feed horizontally instead of vertically in order to minimize the unwanted coupling between ports.

Shown in Fig. 6 is another implementation of a quasi-elliptic filter. It uses magnetic cross coupling between resonators 1 and 4 which is out-of-phase compared with the capacitive coupling between resonators 2 and 3. For a suitable magnitude of magnetic coupling, resonators 1 and 4 need to be modified so that the inductor protrudes as shown.

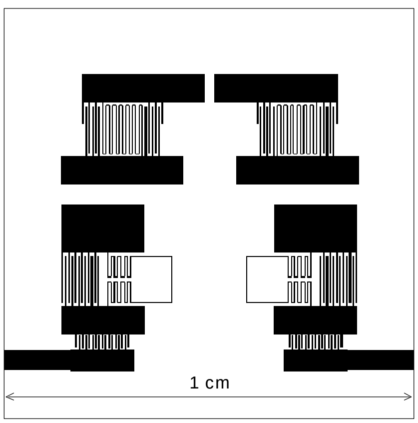


Fig. 6. Second quasi-elliptic filter layout

III. EXPERIMENTAL RESULTS

The filters were patterned on a 50mm diameter, 0.5mm thick MgO substrate with YBCO thin film deposited on both sides. Each filter occupies an area of substrate of $1cm^2$. On superconducting ground plane of the filter was deposited a thin layer of gold film in order to make electrical contact to the packaging. The filter is mounted onto a gold-plated titanium carrier inside a gold-plated titanium box using a thin layer of conductive film. The filter was cooled in a cryogenic cooler and was measured using an HP8720 network analyzer.

Fig. 7 shows the measured results at 30K for the filter in Fig. 4 compared with theoretical lumped equivalent circuit analysis. They show good agreement. This filter is tuned using sapphire tuning screws above each resonator. The minimum passband insertion loss observed is about -1.2dB which corresponds to resonator unloaded Q of about 2000 at 30K, suggesting a relatively poor quality film.

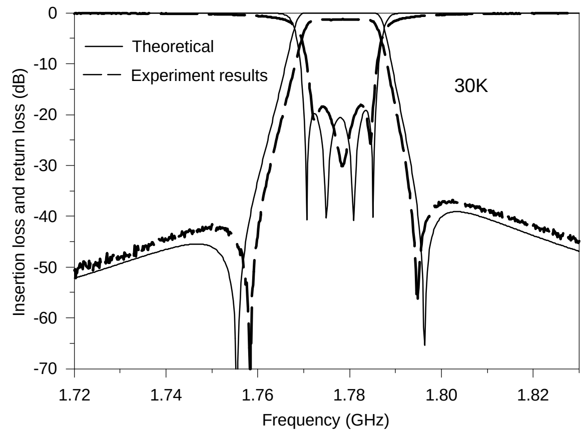


Fig. 7. Quasi-elliptic filter responses.

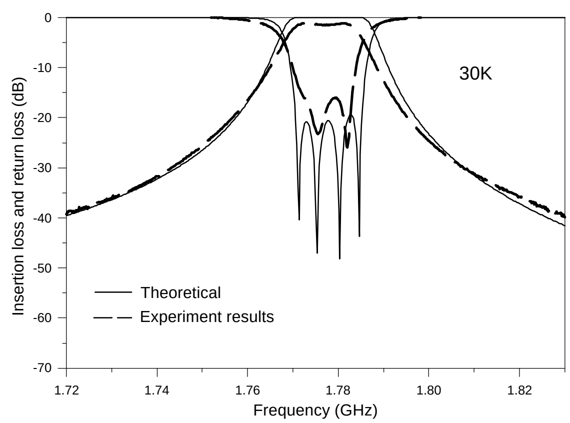


Fig. 8. Linear phase filter responses.

Responses of the linear phase filter in Fig. 5, also tuned and measured at 30K, are shown in Fig. 8. The minimum passband insertion loss is also about -1.2dB. Fig. 9 illustrates the measured group delay in comparison with the theoretical curve. It also shows that the linear phase group delay is better than that of the quasi-elliptic filter. For the magnetic cross-coupled quasi-elliptic filter (Fig. 6), the measured response is shown in Fig. 10.

For the filter in Fig. 6, it should be mentioned that the protruding inductor lines have approximately same length as the non-protruding inductor lines. This implies that the quality factor of this filter should be close to that of the other two filters. The higher insertion loss found in this filter may be due to non-uniformity of the 50mm diameter HTS film [12]. Evaluation of surface resistances of the three $1cm^2$ HTS films based on the measured filter insertion loss using full-wave simulator suggest that the surface resistances can varies from $46\mu\Omega$ to $125\mu\Omega$.

Fig. 11 shows the wide band responses for the above filters. The spurious response does not occur until about 7GHz; at this frequency the lumped-element approximation no longer holds.

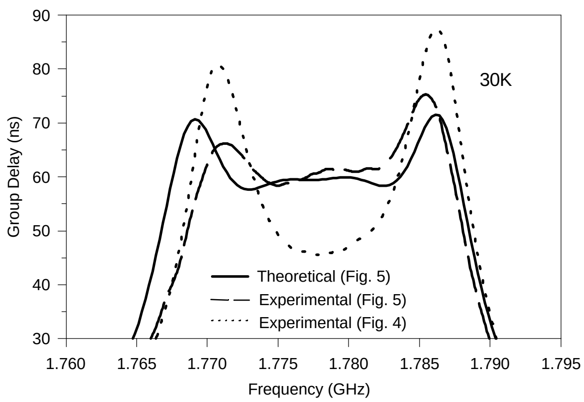


Fig. 9. Comparison of group delays.

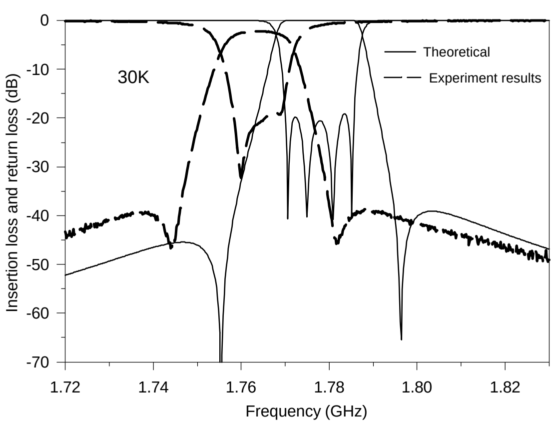


Fig. 10. Quasi-elliptic filter responses.

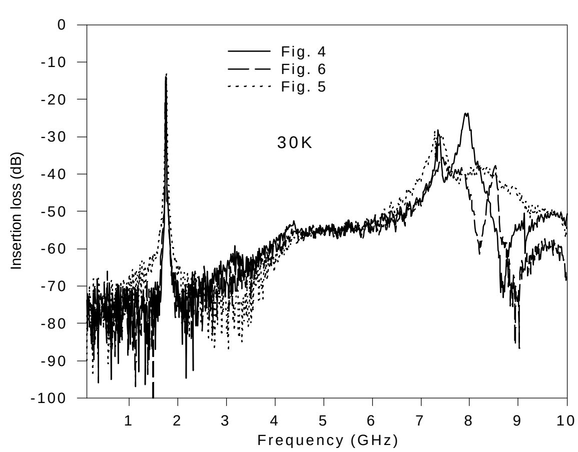


Fig. 11. Spurious responses of the filters.

IV. CONCLUSIONS

The design of two quasi-elliptic and one linear phase filter have been discussed and verified experimentally. Use of the transmission line and the magnetic coupling to achieve the required cross coupling for quasi-elliptic response have been demonstrated. For the linear phase filter, the expected flattening of group delay at passband frequencies is observed. The compactness of the filter design have enabled each of these filters to be fabricated on a $1cm^2$ substrate.

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