QUASI-LUMPED ELEMENT BAND-PASS FILTERS USING DC ISOLATED SHUNT INDUCTORS

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ABSTRACT

The design of quasi-lumped element band-pass filters using DC isolated shunt inductors with finite frequency attenuation poles below the pass-band is described. The design uses coupled series transmission lines for improved tuning. Although generally applicable, these filters are particularly useful for planar thin film High Temperature Superconductors.

INTRODUCTION

Quasi-lumped band-pass filters for operation in the low microwave region and above, tend to have small circuit structures which together with process tolerances, uncertainties in dielectric constant, inaccuracies in the modeling process and spreads in other physical parameters usually require the filter to be tuned. For series resonators (and hence series connected inductors), if tuners are used to change the values of the shunt capacitors at either ends of the inductors, then as these capacitors also form part of the coupling network, it is very difficult to change the resonant frequencies without significantly affecting the coupling between resonators. A practical quasi-lumped circuit structure with better tunability is the band-pass filter with shunt inductors shown in Figure 1. The need for grounding one end of the inductor is avoided by capacitively coupling the ground end of the structure. This has the effect of introducing a pole in the susceptance of the network at a frequency below the zero of this function. A plot of the susceptance of this type of resonator showing the pole and zero is given in Figure 2. The zero corresponds to the parallel resonance of the network and traditional filter design techniques [1] using the admittance slope parameter at this resonance can be utilized to synthesize a filter. However, the introduction of the pole can cause the susceptance at the band edges to be significantly different from the simple shunt parallel LC resonator. It has been observed that with complex resonators, of the type described here, these traditional filter synthesis techniques produce inaccuracies in the computed filter performance. The design technique given below, based on the band edge susceptances, overcomes these deficiencies.

SHUNT INDUCTOR RESONATORS

The design of high-frequency, lumped-element, shunt resonator band-pass filters requires the resonator to have zero susceptance at resonance. Normally, the design is based upon the susceptance slope at the resonant frequency and the fractional bandwidth. The design approach given below modifies this approach to design the band-pass filter based upon the band edge frequencies and the susceptance at those frequencies only. This is necessary for filters using these resonators as the susceptance deviates from that of a simple parallel LC resonator, particularly as the bandwidth increases. The actual resonant frequency may occur anywhere between the band edge frequencies and is not necessarily at the geometric band center frequency. This forces the design to be accurate at the band edges of the band-pass filter and allows for the performance to deviate from that achieved from the normal parallel LC arrangement within the pass-band. For equal ripple Chebyshev designs, this causes changes in the filter attenuation zero frequencies with no degradation in the maximum pass-band ripple level. In general, the design procedure, depicted in Figure 3, matches the susceptance at the pass-band edges of the simple parallel LC arrangement to that of the new resonator structure.

As can be seen by the resonator in Figure 3(b), the structure is relatively complex but by modifying only the transmission lines shown above the inductor, one need not generate a model for the complete structure. The susceptance of this resonator at the band edges can be determined through accurate computer modeling or experimentally. With the inductor connected to the mid-point of the series transmission line, the susceptance of the two series transmission lines connected at this point can be equated to an equivalent capacitance (C_r) at ω_0 as shown in Figure 3(c).

Figure 4(a) shows a design where the susceptance slope parameters at the resonant frequency, ω_0, are set equal for both the simple LC resonator and the complex resonator used to build the filter. In general, as shown in Figure 4(a), the susceptance of these two resonators will not be equal at the edges of the pass-band. The new design

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approach is shown in Figure 4(b) where the susceptance of the two resonators are made equal at the band edge frequencies, ω_1 and ω_2. In general, this results in the resonant frequency of the two resonators occurring at different frequencies (ω_0 for the LC resonator and ω'_0 for the actual resonator).

Using the band edge frequencies and susceptance, the resulting filter will have identical performance to the LC resonator filter at the pass-band edges. This will result in some deviation from the ideal LC resonator design, but the width of the pass-band and band edge frequencies will be maintained. From a design standpoint, one can use the LC resonator model to design the filter once the susceptances of the two resonators are equalized at the band edges. This then allows the design to proceed in a similar manner to Ref. [1], Fig. 8.02-4, with the expression given in equation (1) used for the susceptance slope parameter, b.

$$b = \frac{B}{w} \qquad \text{.................................} \ (1)$$

Where, $B(\omega_2) = B = -B(\omega_1)$ and $w = \dfrac{\omega_2 - \omega_1}{\sqrt{\omega_1 \omega_2}}$

SHUNT INDUCTOR RESONATOR BAND-PASS FILTER DESIGN

The design is based upon a generalized band-pass filter circuit using admittance inverters, as shown in Figure 5. Given the band edge frequencies, ω_1 and ω_2, and the susceptance, B, of the resonator at these frequencies, the LC resonator equivalent susceptance slope parameter, b, can be derived as shown in equation (1) and the equations (2) below can be used to design the filter.

$$\text{If} \quad \omega_0 = \sqrt{\omega_1 \omega_2} \qquad \text{................. ....} (2)$$

$$J_{01} = \sqrt{\frac{G_A w b}{g_0 g_1 \omega'_1}}$$

$$J_{j,j+1} = \frac{wb}{\omega'_1 \sqrt{g_j g_{j+1}}}$$

$$J_{n,n+1} = \sqrt{\frac{G_B w b}{g_n g_{n+1} \omega'_1}}$$

$$C_{01} = \frac{C_0 + G_A \sqrt{\dfrac{C_0^2}{J_{01}^2} + \dfrac{1}{\omega_o^2}\left(\dfrac{G_A^2}{J_{01}^2} - 1\right)}}{\left(\dfrac{G_A^2}{J_{01}^2} - 1\right)}$$

$$C_{j,j+1} = \frac{J_{j,j+1}}{\omega_0}$$

$$C_{n,n+1} = \frac{C_{n+1} + G_B \sqrt{\dfrac{C_{n+1}^2}{J_{n,n+1}^2} + \dfrac{1}{\omega_o^2}\left(\dfrac{G_B^2}{J_{n,n+1}^2} - 1\right)}}{\left(\dfrac{G_B^2}{J_{n\,n+1}^2} - 1\right)}$$

$$C_1 = -\left(C_{01}^e + C_{12}\right) \qquad \text{..... ....(2) cont}$$

$$C_j = -\left(C_{j-1,j} + C_{j,j+1}\right)$$

$$C_n = -\left(C_{n,n+1}^e + C_{n-1,n}\right)$$

$$C_{01}^e = \frac{C_{01}\left[G_A^2 + \omega_o^2 C_0 (C_0 + C_{01})\right]}{\left[G_A^2 + \omega_o^2 (C_0 + C_{01})^2\right]}$$

$$C_{n,n+1}^e = \frac{C_{n,n+1}\left[G_B^2 + \omega_o^2 C_{n+1} (C_{n+1} + C_{n,n+1})\right]}{\left[G_B^2 + \omega_o^2 (C_{n+1} + C_{n,n+1})^2\right]}$$

In the equations (2), ω'_1 is the low-pass prototype cut-off frequency and can be set equal to 1. The resulting filter is of the form shown in Figure 6. The matching transformers at the ends of the filter have shunt capacitors, C_0 and C_{n+1}, the value of which can be selected at will (including equal to zero) in order to make the end elements realizable.

The negative capacitors from the admittance inverters, C_1 through C_n, can be absorbed into the first element (i.e. the open circuit series transmission line of shunt resonator or it's equivalent capacitance, C_r), at each node of the filter. For lumped element resonators that are connected as 1-port networks, this design procedure is adequate. However, coupling to the center of the transmission lines to adjacent resonators is not practical. The coupled 1-port resonator must be modified to connect the resonators as 2-port networks.

SERIES TRANSMISSION LINE COUPLING STRUCTURE

The problem of connecting the shunt resonator as a 2-port network rather than a 1-port network can be solved by modifying the admittance inverters. The transformation of the lumped element resonator with capacitive admittance inverter, Figure 7a, to a series of coupled transmission lines, Figure 7b, can be accomplished by equating the two networks. In the equations (3) below, C_r is half of the equivalent shunt capacitor from the resonator, C_f is the fringing capacitance from the ends of the transmission lines and C_c is the series coupling capacitance between the transmission lines (all capacitance determined at ω_0).

$$C_c = \frac{Y_0\left(Y_0 - \omega_0 C_f \tan\theta\right)\left(Y_0 \tan\theta \left[1 + C_r C_f \left(\dfrac{\omega_0}{Y_0}\right)^2\right] - \omega_0\left[C_r - C_f\right]\right)}{\omega_0\left\{\tan^2\left(Y_0^2 + 2\omega_0^2 C_r C_f\right) - 2Y_0 \omega_0 \left(C_r - C_f\right)\tan\theta - Y_0^2\right\}}$$

$$\tan\theta = \frac{\omega_0 \left(C_r - C_f - C\right)}{Y_0\left[1 + C_f (C_r - C)\left(\dfrac{\omega_0}{Y_0}\right)^2\right]} \qquad \text{..... .... (3)}$$

where, $\omega_0 = \sqrt{\omega_1 \omega_2}$

Y_0 = Admittance of transmission line

$\theta = \beta l$ = Phase length of transmission line

This transformation preserves the 90^0 phase shift from the 1-port node of one resonator to the node of the adjacent resonators. The effect of this transformation is to re-form the transmission line that C_r was derived from, but with a different length than used with the original resonator.

End coupling to the transmission lines at the input and output ends of the filter can be derived in a similar manner. However, the coupling capacitance that results tends to be unrealizably large, resulting in an unacceptably small gap, and therefore alternative coupling techniques usually need to be employed.

DESIGN EXAMPLE

To show this design procedure, a trial filter is shown in Figure 8. The design uses the same basic resonator throughout the filter. TBCCO superconductor on a 0.020" thick MgO substrate was used to construct the circuit. The susceptance at the band edges for the resonators was determined to be $|B| = 0.0002632$ mhos using an in-house generated software program. The input and output coupling was achieved using asymmetric coupled microstrip transmission lines, similar to those shown in Figure 1. The input/output coupling capacitance and associated shunt capacitors are those that would give the same susceptance at ω_0 as the asymmetric coupled lines, with the length of the end series transmission lines adjusted to maintain resonance. Using this coupling technique avoids unrealizably small coupling gaps. The 0.24569pF input/output coupling capacitor was realized by coupling the 50 Ω input/output transmission lines to the 0.040" width resonator lines with a length of 0.1213" and a gap of 0.0024". The element values given in Figure 8 are readily realizable in printed form on a dielectric substrate. The predicted performance of the trial filter is shown in Figure 9, using the ideal capacitor values for the input/output coupling network. Center frequency tuning can be achieved by the addition of shunt capacitance to either of the transmission lines of the resonator or by perturbing the capacitance across the inductor. Also, the coupling gap is readily accessible for tuning bandwidth.

Figure 10 shows the measured performance of the above filter at 77K. Although the performance of this filter is reasonably good, it was about 80 MHz higher than the design center frequency and the return loss was not as good as predicted. The frequency error was greater than could be adjusted using tuning and was attributed to an error in the derivation of the resonator susceptance. The return loss was due to inaccuracies in the derivation of input/output coupling structure (as closed form approximations were used together with approximate fringing capacitances).

CONCLUSION

The band-pass filter design given above allows for practical narrow band-pass filters to be generated, without the need for DC ground connections to the inductor. By the use of numerical techniques, electromagnetic simulation on a computer can be used to predict an individual resonator's susceptance. Using the above design procedure, a complete band-pass filter can then be systematically constructed. The resulting filter is very realizable as a planar circuit pattern created on a dielectric substrate and is inherently tunable as compared to more common series inductor realizations.

REFERENCES

[1] G. L. Matthaei, L. Young, and E. M. T. Jones, "Microwave Filters, Impedance-Matching Networks, and Coupling Structures", Norwood, MA: Artech House, 1980.

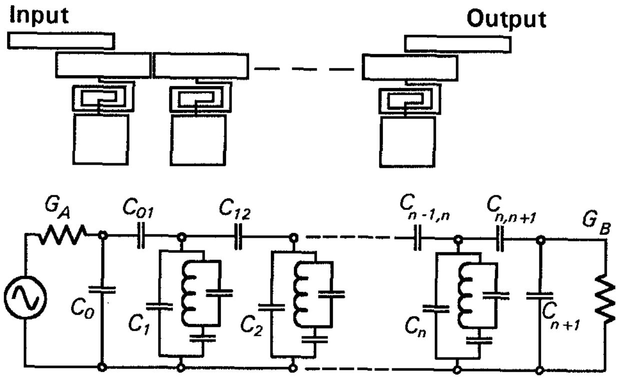

Figure 1. DC isolated shunt inductor BPF.

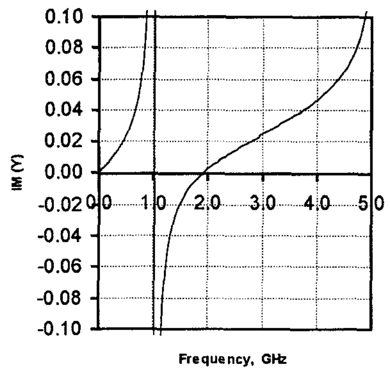

Figure 2. Typical susceptance of the complex resonator.

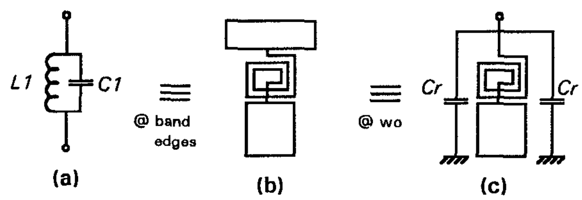

Figure 3. The susceptance of a simple LC resonator (a) is matched to the susceptance of a complex resonator (b) at ω_1 and ω_2, (c) shows the equivalent shunt capacitances, C_r, of the transmission line used to transform to a 2-port device.

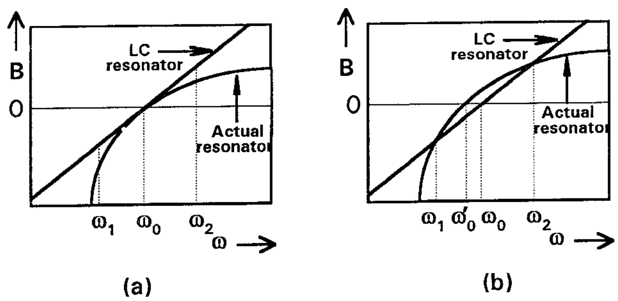

Figure 4. In (a) the resonant frequency of both the LC and complex resonators have the same susceptance slope at resonance. In (b) the LC and complex resonators have the same susceptance at the band edges of the band-pass filter.

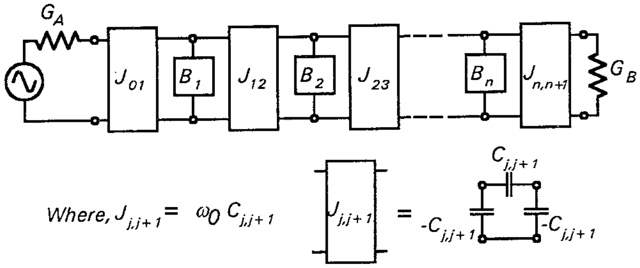

Figure 5. Band-pass filter using shunt resonators and admittance inverters.

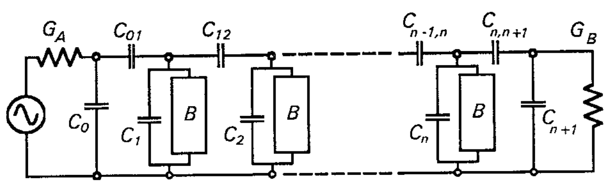

Figure 6. Realization of band-pass filter using shunt resonators, all of the same form with band-edge susceptance B, and capacitive admittance inverters.

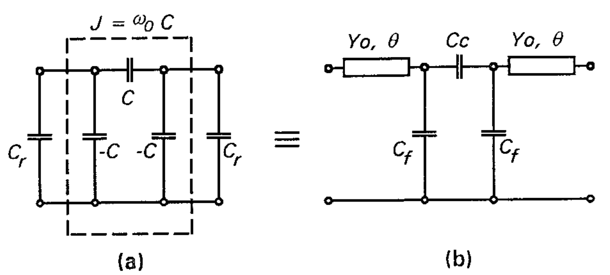

Figure 7. The lumped capacitor admittance inverter shown in (a) together with C_r is equivalent to the capacitively coupled series transmission lines shown in (b).

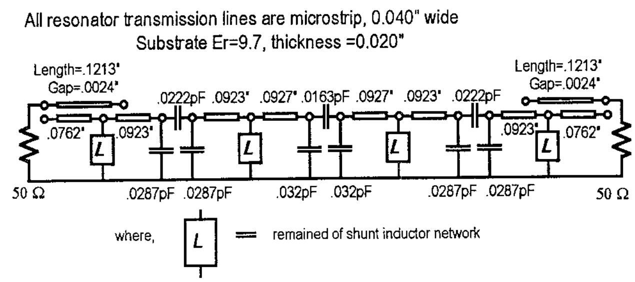

Figure 8. Component values for a trial shunt resonator band-pass filter example (n=4, 0.01 dB equal ripple Chebyshev, fo=1902.5 MHz, BW=17 MHz).

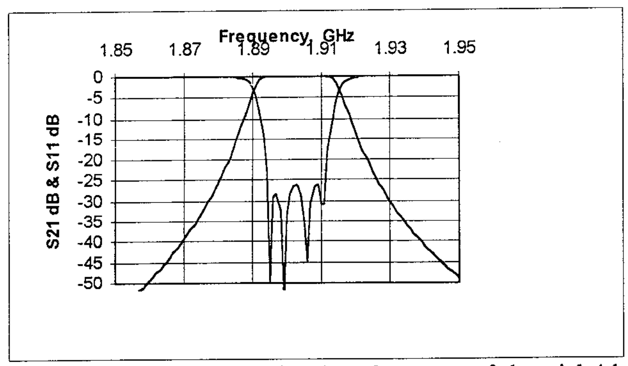

Figure 9. Computer predicted performance of the trial 4th order, 0.01 dB equal ripple Chebyshev band-pass filter.

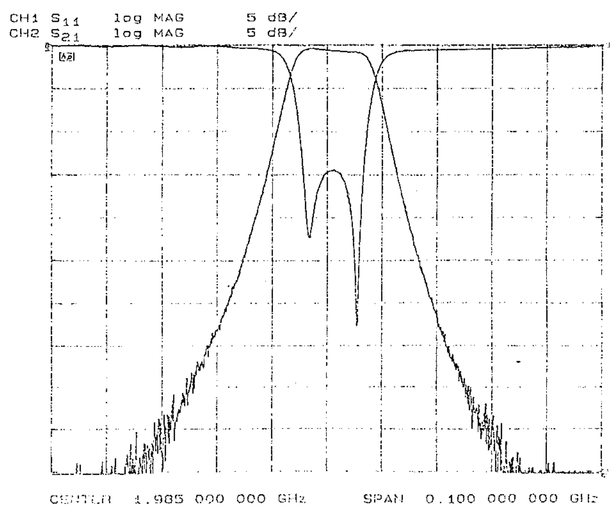

Figure 10. Measured performance of the trial 4th order, 0.01 dB equal ripple Chebyshev band-pass filter.