SURFACE ACOUSTIC WAVE RESONATOR FILTERS

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ABSTRACT

Very narrow filters in the VHF and UHF frequency ranges can be designed and built with SAW resonators. A detailed design example of electrically coupled filters demonstrates the technique. Measured and predicted performance closely agree.

Introduction

One and two port Surface Acoustic Wave (SAW) Resonators using both quartz and lithium niobate have been constructed for use in the VHF and UHF frequency ranges.[1,2,3] The two port resonator (Figure 1) has the decided advantage that the interelectrode capacitance C_o does not parallel the resonant arm (Figure 2). The filters described are designed using two-port resonators with C_o incorporated into the coupling capacitance.

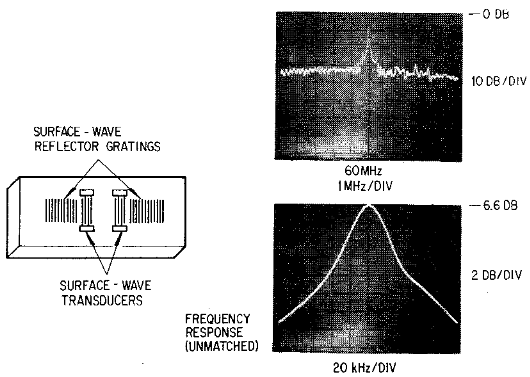

Figure 1. Two-Port SAW Resonator

Coupled Resonator Filters

The concept of coupled resonator filters is familiar to most microwave engineers. Parallel resonant circuits interspersed with admittance inverters or series resonant circuits connected by impedance inverters are used to form multipole bandpass filters.[5] These concepts are useful in the design of multipole SAW resonator filters, whether coupled electrically or acoustically. The bandpass design which lends itself to cascaded electrically coupled two-port resonators has been detailed in circuit element terms by Dishal.[6] The filter configuration is shown in Figure 3.

Performance Prediction

A knowledge of the characteristics of the available resonators is sufficient to allow prediction of performance and bandwidth limitations before detailed design begins, precluding wasted effort in the detailed design phase.

Essentially any bandshape design utilizing only poles of transmission can be realized with the configuration of Figure 3, subject to the insertion loss and bandwidth constraints discussed below. Curves describing the passband shape can be found in standard references.[6,7,8]

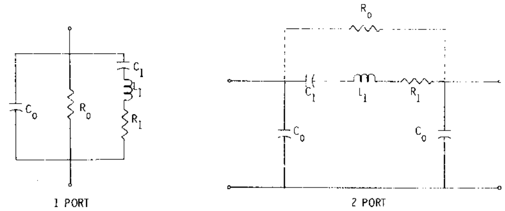

Figure 2. SAW Resonator Equivalent Circuits

For bandwidths realizeable with the configuration of Figure 3, it is necessary to compensate for the losses in the resonator if the filter shape in the passband is important.[9] Design constants for some of the more popular shapes are available.[6] Once the unloaded Q (Qu) of the resonators has been determined, the design constants and insertion loss of the completed filter can be determined. For nominally flat passband amplitude (Butterworth) filters, for example, the insertion loss has been calculated and is presented in Figure 4. Lower loss filters could be constructed using the same resonators, but not with the prescribed Butterworth shape. The unloaded Qu of the resonator can also be determined from Figure 4 using the curve n = 1. For example, the two-port resonators used in the filter of Figure 3 had bandwidth to center frequency ratio of 1/1500 ("loaded Q" = Qe of 1500) and insertion loss = 6.6 dB. From Figure 4, n = 1, normalized Q = 1.8 so Qu = 2700. If we desire 80 kHz bandwidth in a four-pole filter of Butterworth shape at 60 MHz, normalized Q_n = 2700 x 80 kHz/60 MHz = 3.6. From Figure 4, n = 4, the expected insertion loss is 12 dB.

The lower limit of bandwidth is determined by the acceptable insertion loss. The upper limit of bandwidth with the configuration of Figure 3 is set by the parasitic capacitance C_o. As can be seen from Equation 2, the maximum bandwidth occurs when $C_{\ell m} = 0$, or $\mathrm{Max}(f_{3dB}) = C_1 f_o / 2\ \mathrm{Max}\ (k_{1m})\ C_o$. (Since the k_{1m} are different, the largest one governs).

The bandwidth can be made larger by substituting inductors for the inter-resonator capacitors of Figure 3, thus "tuning out" some of the capacitive reactance of the equivalent inter-resonator capacitance $2C_o$. This configuration introduces losses (in the inductors) not included in curves such as Figure 4, which will therefore yield optimistic predictions. The inductors also introduce extra poles of transmission below the filter center frequency (Fig. 5). As the bandwidth is further increased, the separation of these poles from center frequency decrease, until in the upper limit of bandwidth they cause unacceptable distortion of passband shape. The poles cannot readily be incorporated in the design (to make a 5-pole filter from a 3-pole filter, for example) because of the low Q_u of the coils.

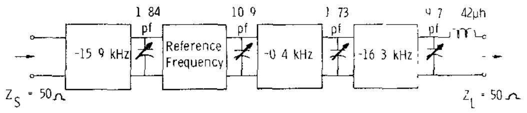

Figure 3. Four-Pole Butterworth Filter with Two-Port Resonators

As can be seen from the various filter response figures, these filters do not follow the classical curves to infinite attenuation out of the passband. There are "shoulders" of spurious responses around the passband. These are due to the transducers coupling directly from one to the other off resonance. At frequencies where the spacing of reflectors in the reflective arrays is different from 1/2 wavelength, the reflective arrays are essentially transparent, and the transducer coupling follows the classical sin x/x response, the peak of which is represented by R_o in Figure 1.[4] The model of Figure 1 does not predict the performance exactly, therefore, since it does not account for the frequency dependence of the transducer coupling. The rejection at the "shoulder" can be estimated from the detailed design by presuming the impedance of the resonant arms to be high (off resonance condition) and calculating the attenuation of the resultant R_o C_{mn} ladder network. A generalized procedure for predicting the attenuation at the "shoulder" without resorting to a detailed design has yet to be developed.

Design Procedure

For ladder network narrowband filters, the configuration of Figure 3 is suggested by the form of the equivalent circuit of the two-port resonator (Fig. 1). The design equations for this configuration are presented in Reference 2 in terms of the normalized coupling and loading coefficients ($k_{\ell m}$ and q_j). Because the resonators are nearly identical, these equations reduce to:

$$R_{Lj} = \frac{2\pi f_{3dB} L_1}{q_j} - R_1 \ , \ j = 1 \text{ or } n \qquad (1)$$

$$C_{\ell m} = \frac{C_1}{k_{\ell m}} \cdot \frac{f_o}{f_{3dB}} - 2 \cdot C_o \qquad (2)$$

Where R_{Lj} are the source and load resistances, and C_o, C_1, L_1, and R_1 are defined in Figure 1, $C_{\ell m}$ is the additional coupling capacitance required between resonators ℓ and m, f_o is the center frequency, and f_{3dB} is the desired 3 dB bandwidth of the n-pole filter.

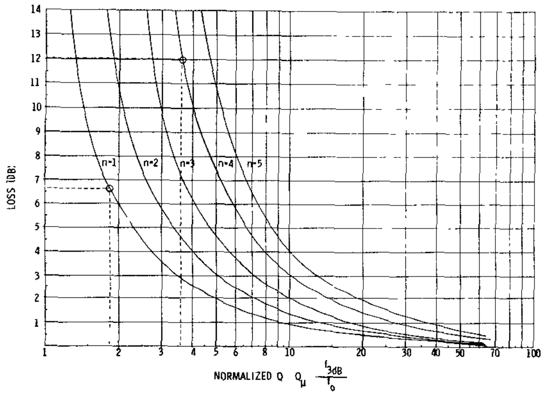

Figure 4. Insertion Loss of Multipole Butterworth Filters

The design procedure is best described with an illustrative example. It is desired to construct a 4-pole Butterworth filter centered near 60 MHz using two port resonators with the following characteristics:

$$f_o = 59.855 \text{ MHz}$$

$$IL = 6.6 \text{ dB}$$

$$f_{3dB} = 40 \text{ kHz}$$

$$C_o = 3.5 \text{ pf}$$

The unloaded Q_u is determined from Figure 5 as discussed above in the section on performance prediction to be 2700. R_1 is determined from the insertion loss at resonance to be 105 ohms (which was incidentally confirmed with a measurement of S_{11} at resonance). C_o is determined from S_{11} far from resonance. Since $Q_u = WL_1/R_1$, $L_1 = 0.74$ mh. From the resonance equation, $C_1 = 9.38 \times 10^{-3}$ pf. As pointed out in the prediction example above, for an 80 kHz bandwidth, $Q_n = 3.6$. The values of $k_{\ell m}$ and q_j are read from design curves (for example Figure 45, page 8-37, Reference 7). Equations 1 & 2 are then used to calculate the $C_{\ell m}$ and R_{Lj} listed in Table 1 alongside the corresponding q_j and k_{lm} values.

q_1 = 0.61	R_{L1} = 502 ohm
k_{12} = 0.82	C_{12} = 1.73 pf
k_{23} = 0.40	C_{23} = 10.9 pf
k_{34} = 0.81	C_{34} = 1.84 pf
q_4 = 2.42	R_{L4} = 48 ohm

Table 1. BUTTERWORTH FILTER PARAMETERS

A schematic of the completed filter with L-section matching network to transform R_{L1} (502 ohm) to 50 ohm test equipment, is shown in Figure 3.

Alignment Procedure

Each mesh of the filter must resonate at f_o. Since the $C_{\ell m}$ are, in general, different, the self resonant frequencies of the resonators measured in a 50 ohm transmission system will differ. The offset in resonant frequency required of each resonator can be easily calculated by substituting for C_1 the series combination C_1, $C_{\ell m} + 2\ C_o$, and $C_{\ell+1} + 2\ C_o$. These offsets are shown in Figure 3, referred to the highest frequency resonator. Observe that, although the stand alone resonant frequencies of each resonator are different, they are not progressively staggered to achieve the desired shape. Since the techniques of tuning surface wave resonators are either unidirectional (adding or removing material) or involve extra components, a non-iterative adjustment technique is advisable for aligning the completed filter.[10]

Filter Examples

184 MHz Quartz Resonator Filters

A filter constructed to achieve very narrow bandwidth, good out of band rejection, and simplicity of construction is shown in Figure 6. The shape of the filter in the passband was not a consideration in this filter, so a formal design was not made. It is included here to demonstrate the consequence of bandwidth changes by contrasting with the two-pole Butterworth filter of the same center frequency shown in Figure 5. Both filters were made with two-port quartz SWD resonators with $Q_u \cong 20{,}000$. Notice that as the bandwidth increases, the in band insertion loss and out of band rejection decrease.

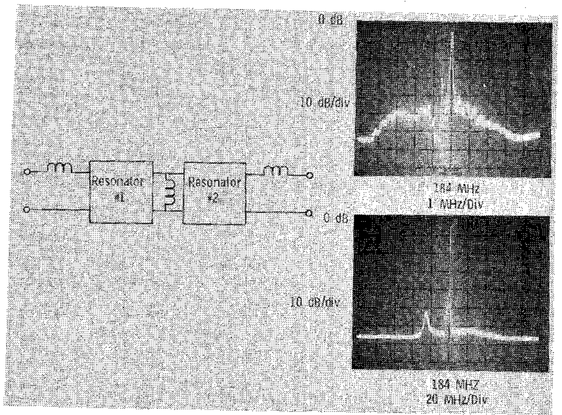

Figure 5. Two-Pole 184 MHz Butterworth Filter

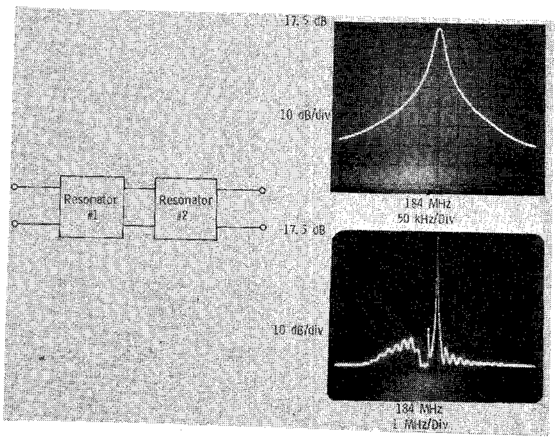

Figure 6. Two-Pole 184 MHz Narrow Band Filter

Four-Pole Filter at 60 MHz

The filter of the design example discussed above was constructed using the Lithium Niobate resonators at 59.9 MHz shown in Figure 1. The result is shown in Figure 7. Notice the flat response in the passband corresponding to Butterworth shape. The slight "bump" in the high frequency rolloff region is due to transversal modes in the resonators not totally suppressed. The discrepancy in the predicted and measured insertion loss is due to the unloaded Q_u of two of the resonators being slightly lower than the one characterized for the design.

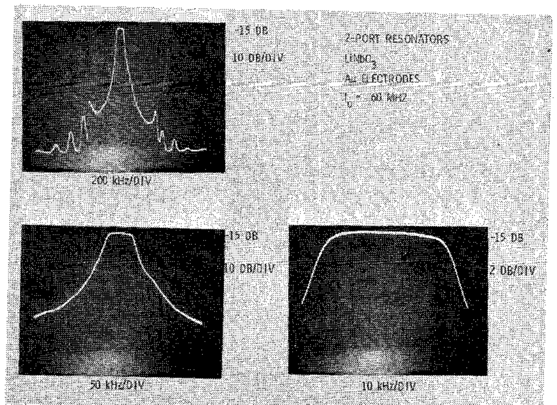

Figure 7. Four-Pole Butterworth Filters Using Two-Port Resonators at 59.9 MHz

Conclusions

Surface Acoustic Wave resonators can be used to construct useful, practical filters in the VHF frequency range. It is believed feasible to extend these devices and filters into the UHF and possibly the low microwave frequency ranges. The coupled resonator design concepts and reflectometer alignment technique will be useful in designing and aligning SAW resonator filters whether electrically or acoustically coupled. To reduce component count, complexity, and size to a minimum, and to widen the choice of design bandwidth, acoustic coupling techniques provide a fruitful area for further research.

Acknowledgements

The author has benefited from discussions with W. R. Shreve and D. T. Bell, Jr. Device design and fabrication by W. R. Shreve, R. E. Stigall, and J. Chtay was key to the finished filters. The filters were constructed, aligned, and tested by R. C. Bennett and D. R. Saunders.

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