POLYLITHIC SAW FILTERS USING COUPLED-RESONATOR FILTER ELEMENTS

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Abstract

Discrete SAW filters consisting of electrically coupled SAW resonators in ladder structure have strong finite attenuation poles, which are a severe limitation in the achievable bandwidth and stopband attenuation, and in the realizability of arbitrary filter characteristics. Lattice structures are difficult to realize at higher frequencies.

Transversely and Longitudinally Coupled SAW filters (TCF, LCF) are the SAW analogue of monolithic crystal filter „duals" (MCF). Their simplified electrical diagram is equivalent to that of a two-pole lattice crystal filter section. Polylithic SAW Filters (PSF) can be generated by electrical coupling of several TCFs / LCFs in a tandem configuration with interstage coupling reactances. This structure allows the realization of filters of a large variety of responses with „all-pole"-characteristics, including linear-phase and selective-linear-phase filters.

The paper descibes the synthesis of PSFs based on the SAW element values. Alternatively an approximative method based on the transmission characteristic of the SAW elements is given. Experimental results achieved with TCF elements at 620 MHz are reported.

Introduction

Filters using discrete SAW resonators electrically connected in ladder configuration are widely used for narrowband applications (see Ref. [11] to [16]). Their disadvantage is the limitation in the maximum bandwidth and the low stop-band attenuation in lower-order filters due to the strong finite attenuation poles. For wider bandwidth bridge configurations in full lattice or in half-lattice form [17],[18] are reported, but suffer from their circuit complexity, because either two identical resonators per „pole" or hybrid transformers are required.

Tiersten and Smythe [1] described first acoustic waveguide-coupled SAW resonator filters as an analogue to monolithic bulk-wave quartz crystal filters. Due to the acoustic coupling of the two waveguide resonators, the fundamental mode frequency splits up into a symmetric and an antisymmetric mode frequency, whose frequency difference (mode spacing) is determined by the coupling through the spacing between the two waveguides. Staples and Smythe [19] reported first experimental results on these devices, indicating a realizable bandwidth range of 0.01 % to 0.1 % of f_0 and suggesting the realization of higher order filters by tandem connection of several coupled resonator „duals". Since then numerous papers were published on „inline" (= longitudinally) and transversely coupled SAW resonators (Ref. [20] to [24]). Because LCF's show a lower stopband attenuation than TCFs due to the „feed-through" of unwanted modes, TCF's are generally preferred for most applications.

The equivalent electrical circuit for the two first-oder modes of TCF's / LCF's was derived by Hartmann et al. [2] with the „coupling-of-modes" (COM) theory. It is given in Fig.1.

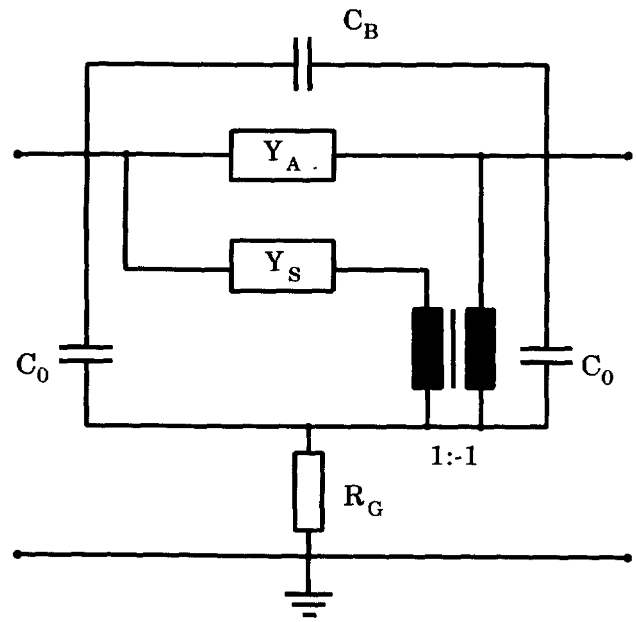

Fig.1: Equivalent electrical circuit of a coupled SAW resonator filter

The circuit is basically that of a so-called „half-lattice" crystal filter with the series resonant circuits Y_S, Y_A in the two branches of a differential transformer and C_{01}, C_{02} = C_0 for the static (transducer) capacitances. The series resonance frequencies of Y_S and Y_A are the symmetric and antisymmetric mode frequencies f_{sym}, f_{asym}. The additional elements stand for parasitic effects: the bridge capacitance C_B from input to output and the common ground bus bar resistance R_G. In the following the parasitic elements C_B and R_G will be neglected for the filter synthesis and structural symmetry ($C_S = C_A$, $R_S = R_A$, $C_{01} = C_{02}$) is assumed. In the filter analysis and optimization C_B and R_G must be considered as well as the higher order transverse modes, which can be represented by additional series resonant circuits in parallel to Y_S and Y_A.

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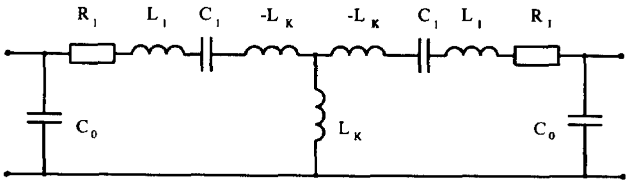
Equivalence to monolithic twopole filter

The equivalence of the circuit of Fig.1 to the electrical diagram of a bulkwave monolithic twopole filter („dual") can be shown by simple circuit transformation (without C_B and R_G) using Bartlett's bi-section theorem. Fig.2 shows the equivalence.



$R_A = R_S = R_1 / 2$ $\qquad L_A = (L_1 - L_k) / 2$

$C_A = C_S = 2 \cdot C_1$ $\qquad L_S = L_1 / 2$

Fig.2: Equivalent ladder structure of a monolithic twopole filter

The frequency f_1 of the two series resonant circuits C_1, L_1 is equal f_{sym}

(1) $\quad f_{sym} = f_1$

(2) $\quad f_{asym} = f_1 / \sqrt{1-k}$

where k is the coupling factor between the waveguides, which is described by the coupling-T network with $-L_k$, L_k, $-L_k$ [3]

(3) $\quad L_k = k \cdot L_1$

Determination of the equivalent circuit elements

Two-mode method

The circuit elements of the equivalent electrical diagram of Fig.1 can be determined by operating the device as a resonator in the symmetric and the antisymmetric mode according to Figs.3 a and 3b.

In the „symmetric mode configuration" the input and output ports are connected in parallel, in the „antisymmetric mode configuration" the device is measured between input and output pin with the ground-pin floating. [4].

The suitable measurement techniques are described in IEC standard 444-5 using a network analyser and measuring either s-parameters or amplitude and phase values in a transmission set-up in connection with error-correction [5]. Cavin and Malocha [6] describe a suitable high frequency test fixture for s-parameter measurement. In this method we were using a modified π-network according to [5].

The first measurement is done in the „symmetric configuration". From the admittance values determined at several points in the 3dB-bandwidth of the series resonance the frequency f_{sym} and the element values C_1*, R_1* can be computed (e.g.with the „two-point iterative method" [5],[7]). These values are related to the element values in the lattice circuit (Fig.1) through (see reference [4]):

(4) $\quad f_S = f_{sym}$

(5) $\quad C_S = C_A = C_1{}^*/4$

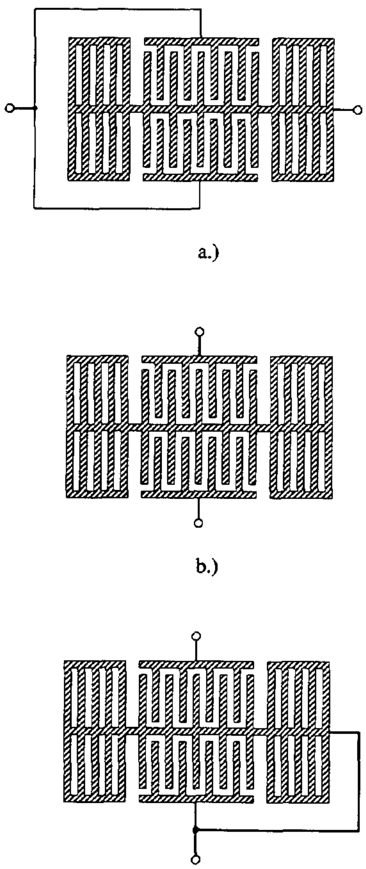
(6) $\quad L_S = \dfrac{1}{\omega_{sym}{}^2 \cdot C_S}$

(7) $\quad R_S = R_1{}^* \cdot 4$

The second measurement in the „antisymmetric configuration" is only used to determine the frequency f_{asym}, which is in Fig.1

(8) $\quad f_A = f_{asym}$

(9) $\quad L_A = \dfrac{1}{\omega_{asym}{}^2 \cdot C_A}$



a.)

b.)

c.)

Fig.3: Excitation of symmetric and antisymmetric mode

(10a) $\quad R_A \approx R_S$

The static capacitance C_0* is derived from measurement points far off resonance in the „symmetric configuration"

(10b) $\quad C_{01} = C_{02} = C_0{}^*/2$

Note:

The factors „4" in eq.(4) and (7) are correct, while in Ref.[4] these factors are erraneously set to „2".

One-port method

The circuit elements of the two motional branches can be easier determined by measuring the admittance at the input port Y_{11} with the output port short-circuited, e.g. by s_{11}-measurement according to Fig. 3c.Y_{11} shows the resonances at the symmetric and the antisymmetric frequency, which are sufficiently isolated. The element values can be determined directly from admittance values within the 3dB-bandwidth of each resonance as in IEC 444-5 without using the transformation factors of eqn. (5),(7), (10b). Furthermore the assumption of structural symmetry of eq.(5) and (10a) must not be maintained.

Note:

An alternative approach published in 1985 by Peach et. al. [8] using a parameter fitting technique is found to be more sensitive to measurement errors and has a tendency to yield unrealistic parameter sets.

Filter design equations

The Polylithic SAW filter structure for a filter of N'th order („N-pole") consists of N/2 SAW twopole resonator units connected in tandem (see Fig.4). At the node between each two units a coupling network is inserted, which in the simplest case is a capacitance C_c to ground, in the other cases it is a π-configuration of three capacitances C_{ca}, C_{cb}, C_{cc}. This is due to a Norton transformation which is necessary to achieve equal motional capacitances for all stages. The filter is terminated at the input with a resistance R_{in} in parallel with a capacitance C_{in} and at the output with R_{out} // C_{out}. In many cases the synthesis delivers negative capacitance values, which must be realized by inductors of the same reactance, i.e. $L_{eq} = 1/(\omega_c^2 |C|)$

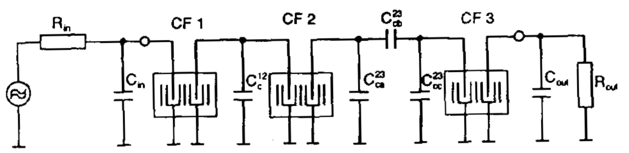


Fig.4:Polylithic SAW filter structure

The filter design is based on the normalized lowpass, whose transformation to a bandpass leads to a ladder circuit as in Fig.2. The design fomulas below refer to the element values as in Fig.1. Any arbitrary „all-pole" filter characteristic - i.e. with attenuation poles at infinity - can be chosen, from which the normalized k_{ij}- and q-factors (computed from the normalized lowpass elements) must be known [9]. To enable the usage of SAW resonator filters with identical motional capacitance for all stages, a so-called „equimotional" design was chosen.
In the following the equations up to 6^{th} order are summarized, for higher orders please refer to the filter literature [9],[10].

Inputs:

filter characteristic given by the k_{ij}- and q_1,q_n parameters e.g. from filter tables [9], which relate to the normalized lowpass elements l_i, c_i, r_{in}, r_{out} by (for N even):

$k_{ij} = 1/\sqrt{lc_i\ lc_j}$

lc_k = k-th lowpass element (l_k or c_k)

$q_1 = l_1/r_{in}$

$q_n = c_n\ r_{out}$

filter order	N = 2,4,6,8...
centre frequency	f_c
bandwidth (3dB)	BW
motional capacitance	C_S
static capacitance	C_0
resonator Q (11)	$Q = 1/(\omega_S*C_S*R_S)$

Definitions:

(12) $q_0 = Q*BW/f_c$
(13) $q_S = 1/(1/q_1-1/q_0)$
(14) $q_L = 1/(1/q_n-1/q_0)$
(15) $C_{ref} = 2*C_S\ f_c/BW$
(16) $R_{ref} = 1/(\omega_C\ C_{ref})$

Design equations:

2-pole

(17) $f_{sym} = f_c - BW/2*(1/q_S+k_{12})$
(18) $f_{asym} = f_c - BW/2*(1/q_S-k_{12})$
(19) $R_{in} = 2/q_S*R_{ref}$
(20) $C_{in} = q_S/2*C_{ref} - C_0$
(21) $R_{out} = q_L*(1/q_S^2+1/q_L^2)*R_{ref}$
(22) $C_{out} = 1/(q_S*(1/q_S^2+1/q_L^2))*C_{ref} - C_0$

4-pole

(23) $f_{sym}(1) = f_c - BW/2*(k_{23}+k_{12})$
(24) $f_{asym}(1) = f_c - BW/2*(k_{23}-k_{12})$
(25) $f_{sym}(2) = f_c - BW/2*(k_{23}+k_{34})$
(26) $f_{asym}(2) = f_c - BW/2*(k_{23}-k_{34})$
(27) $R_{in} = (q_S*k_{23}^2+1/q_S)*R_{ref}$
(28) $C_{in} = k_{23}/(k_{23}^2+1/q_S^2)*C_{ref} - C_0$
(29) $R_{out} = (q_L*k_{23}^2+1/q_L)*R_{ref}$
(30) $C_{out} = k_{23}/(k_{23}^2+1/q_L^2)*C_{ref} - C_0$
(31) $C_c = C_{ref}/k_{23} - 2*C_0$

6-pole

Set k_{max}= MAX(k_{23},k_{45})

(32) $f_{sym}(1) = f_c - BW/2*(k_{max}+k_{12})$
(33) $f_{asym}(1) = f_c - BW/2*(k_{max}-k_{12})$
(34) $f_{sym}(2) = f_c - BW/2*(k_{max}+k_{34})$
(35) $f_{asym}(2) = f_c - BW/2*(k_{max}-k_{34})$
(36) $f_{sym}(3) = f_c - BW/2*(k_{max}+k_{56})$
(37) $f_{asym}(3) = f_c - BW/2*(k_{max}-k_{56})$
(38) $R_{in} = (1/q_S+q_S*k_{max}^2)*R_{ref}$
(39) $C_{in} = (k_{max}/(k_{max}^2+1/q_S^2))*C_{ref} - C_0$
(40) $R_{out} = (1/q_L+q_L*k_{max}^2)*R_{ref}$
(41) $C_{out} = (k_{max}/(k_{max}^2+1/q_L^2))*C_{ref} - C_0$

If $k_{max} = k_{23}$, then C_{c1}= single capacitor, C_{c2}= π-coupler.

(42) $C_{c1} = 2*C_{ref}/(k_{max}+k_{23}) - 2\ C_0$
(43) $C_{c2a} = C_{ref}/(k_{max}+k_{45}) - C_0$
(44) $C_{c2b} = C_{ref}*k_{45}/(k_{max}^2-k_{45}^2)$
(45) $C_{c2c} = C_{ref}/(k_{max}+k_{45}) - C_0$

If $k_{max} = k_{45}$, then C_{c1}= π-coupler, C_{c2}= single capacitor.

(46) $C_{c1a} = C_{ref}/(k_{max}+k_{23}) - C_0$
(47) $C_{c1b} = C_{ref}*k_{23}/(k_{max}^2-k_{23}^2)$
(48) $C_{c1c} = C_{ref}/(k_{max}+k_{23}) - C_0$
(49) $C_{c2} = 2*C_{ref}/(k_{max}+k_{45}) - 2\ C_0$

transducer attenuation

A fairly good approximation for all filter orders and characteristics can be computed [11] from the normalized Q-value q_0 and the sum of the normalized lowpass element values Σlc_i:

(50) $a_t\ [dB] = 4.34 \cdot \Sigma lc_i\ /\ q_0$

Specification of the SAW two-pole resonators

The SAW devices industry is used to specify and measure their devices in filter terms, i.e. centre frequency, bandwidth, termination impedance etc. Therefore it is recommended to translate the above used terms of f_{sym}, f_{asym}, and C_A of each SAW twopole into filter terms. Within certain limits this is possible in the following way:

A filter with a maximally flat top is a Butterworth filter. For the ideal case of infinite resonator Q ($q_0=\infty$) for a two-pole Butterworth filter it is

(51) $q_S = q_L = q_1 = q_n = \sqrt{2}$

(52) $k_{12} = 1/\sqrt{2}$

From eqn. (17),(18) the mode frequencies for this are

(53) $f_{sym} = f_c - BW/\sqrt{2}$

(54) $f_{asym} = f_c$

From eqn.(15),(16),(19) the motional capacitance can be translated into the termination impedance:

(55) $$R_{in} = R_{out} = \frac{BW}{f_c}\frac{1}{\sqrt{2}\,\omega_c C_S}$$

Thus a SAW-twopole defined by its mode frequencies f_{sym}, f_{asym} and the motional capacitance C_S can be specified as a filter unit with Butterworth (flat-top) response with

(56) centre frequency $f_{c2p} = f_{asym}$

(57) bandwidth $BW_{2p} = \sqrt{2}\,(f_{asym}-f_{sym})$

(58) termination $$R_{in2p} = \frac{f_{asym}-f_{sym}}{\sqrt{2}\cdot\pi\cdot f_{asym}^2\cdot C_S}$$

Limits of realizability

The maximum bandwith for a SAW two-pole is limited by the coupling factor k according to eqn.(1), (2) and (57):

(59) $BW \approx k\,f_c/\sqrt{2}$

and by the filter impedance R_{in}, R_{out}, which is proportional to the relative bandwidth BW/f_c. Spurious modes falling into the filter passband also may degrade the filter performance.

For polylithic SAW filters ($N\geq4$) a further limitation is given by the limited Q of the interstage coupling inductors L_{ci} used in place of the negative coupling capacitors C_{ci}, which causes a distortion of the filter response. For large bandwidth the inductance approaches

(60) $L_{ci} \rightarrow 1/(\omega_c^2\,2\,C_0)$

For low distortion its loss resistance

(61) $R_{Lci} = Q_{Lci}\,\omega_c\,L_{ci}$

must be large compared to the filter termination.impedances. Therefore

(62) $BW_{max}/f_c \ll Q_{Lci}\,C_S/C_0$

The minimum bandwidth is limited by the Q-factor of the SAW resonators. Insufficient Q increases the transducer attenuation and leads to distortion of the passband response.

From eq. (50) and from the approximation $\Sigma lc_i \approx N \ldots 1.6\,N$ the minimum relative bandwidth is

(63) $$BW_{min}/f_c \approx \frac{(4\ldots7)\cdot N}{Q\;a_i[dB]}$$

Further limitations are the frequency adjustment accuracy of the SAW device and the stability over the required operating temperature range and over life time (aging).

As a rule of thumb, the realizable relative bandwidth range of Coupled Resonator Filters is 0.03 % to 0.2 % in the frequency range of 150 MHz to 1000 MHz.

Experimental results

For verification several polylithic SAW filters (PSF) at 622 MHz were realized using Transversely Coupled SAW Filter units (TCFs) on ST-quartz and the described synthesis equations. The seperately housed TCFs were arranged on a pc-board „in-line", the L-C-networks for impedance transformation and coupling consist of chip-inductors and chip-capacitors. Figs. 5 to 10 show the results achieved for two-pole, four-pole and six-pole filters.

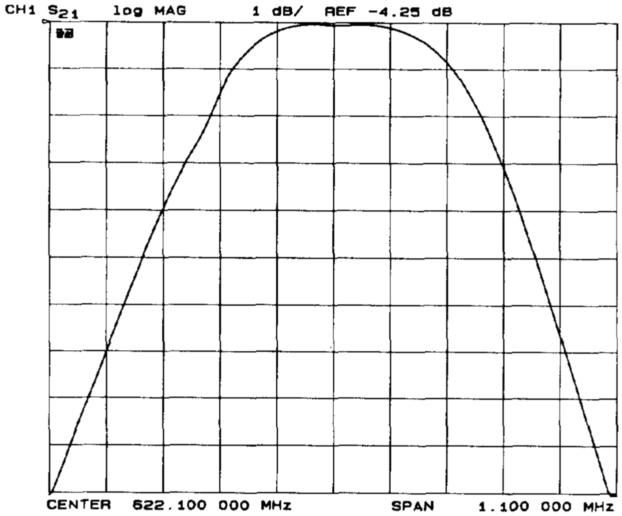


Fig. 5: two-pole TCF filter (passband)

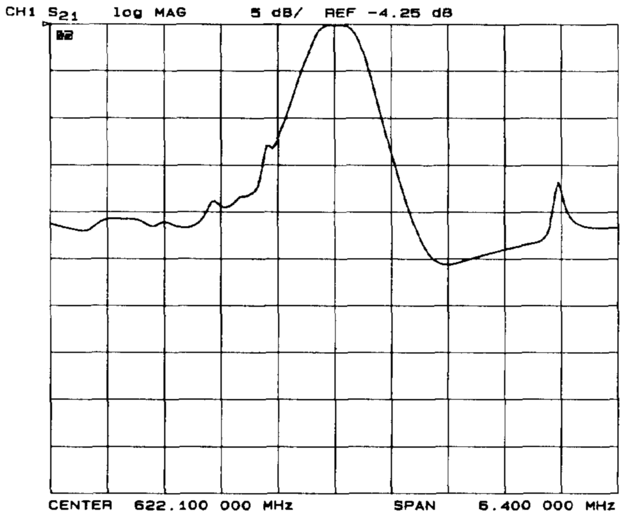


Fig. 6: two-pole TCF filter (stopband)

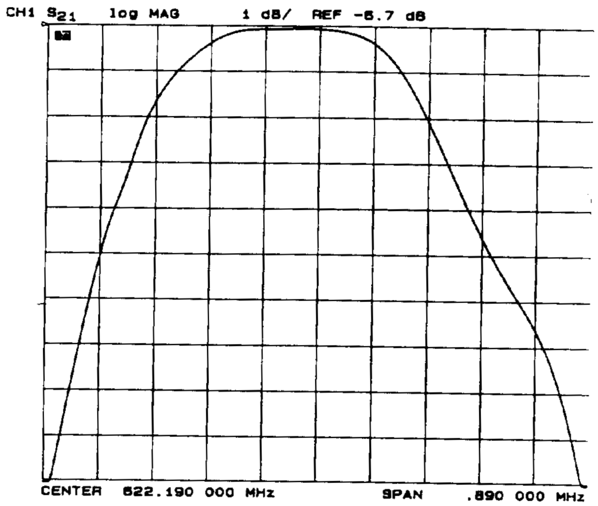


Fig. 7: four-pole polylithic SAW filter (passband)

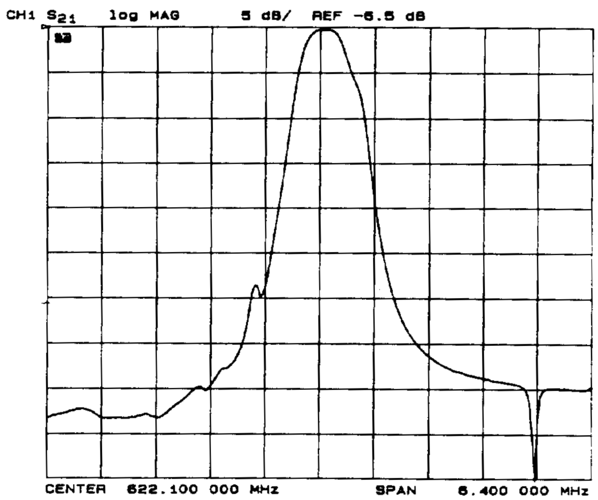


Fig. 8: four-pole polylithic SAW filter (stopband)

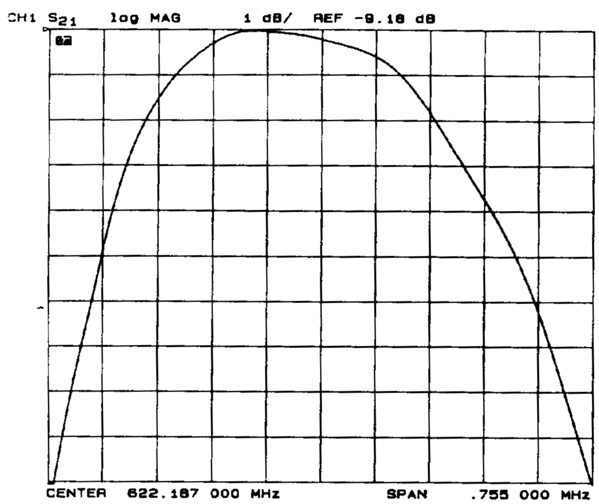


Fig. 9: six-pole polylithic SAW filter (passband)

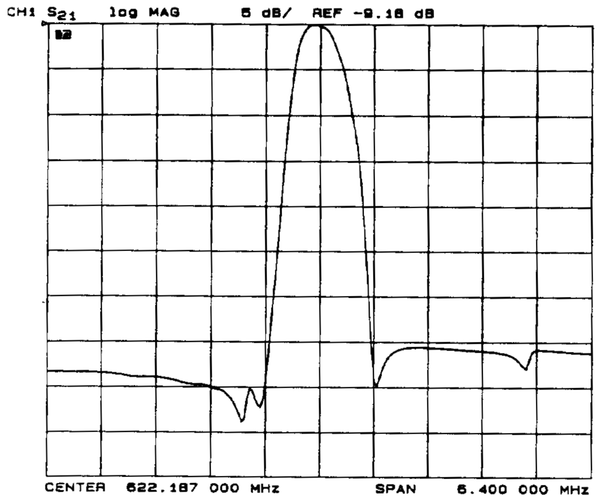


Fig. 10: six-pole polylithic SAW filter (stopband)

1.8 dB of the insertion attenuation is caused by the loss in the impedance-transformation LC-networks, the resonator-Q of approx. 5000 contributes about 2.5dB per resonator. The passband of the six-pole filter is sloping to the high-frequency side due to the loss in the interstage coupling network. The ultimate attenuation of the 6-pole degrades below 40 dB due to direct coupling on the pc-board. By suitable construction it can be easily improved to 60dB.
Wider bandwidth and smaller size can be achieved with bare SAW dice bonded on one common substrate.

Conclusions

The paper summarizes the design procedure for Polylithic SAW Filters (PSF) up to 6[th] order and describes how to set up the specification for the SAW elements. The theoretical limitations for the minimum and the maximum bandwidth are determined. Experimental results with SAW devices at 620 MHz are demonstrated which confirm the validity of the synthesis.
Future realisations will use bare SAW dice bonded on one common substrate.

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