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OPTIMAL AND SUBOPTIMAL SYNTHESIS OF SURFACE ACOUSTIC WAVE FILTERS

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Abstract

Optimal and suboptimal design techniques of surface acoustic wave (SAW) linear phase filters both based on the Remez exchange algorithm and the McClellan's computer program are considered.

An optimal synthesis provides the best fit to a design target and leads to an apodized interdigital transducer (IDT) of minimal length. The optimal solution properties are studied.

A suboptimal synthesis technique proposed allows to reduce considerably an amount of computations without significant sacrificing the approximation accuracy. Thus computer run time and storage are greatly saved if compared to an optimal synthesis. The detailed suboptimal theory and some practical design aspects are also discussed.

The design examples are presented which confirm the efficiency and the flexibility of the synthesis techniques proposed.

Introduction

Several SAW filter synthesis techniques based on the finite impulse response (FIR) digital filter theory [1] have been proposed in the past [2-15]. The most wide spread are 1) the windowing techniques [2-6], 2) the linear programming techniques [7-10], and 3) the Remez exchange algorithm techniques [11-15]. A comprehensive review of these may be found elsewhere [16].

But up to now there have been some problems in applying these techniques to a SAW filter design. That is why this is a common practice to use some simplifications, sometimes without sufficient foundations. For example, it is usually supposed that within a filter passband a contribution $F_1(\omega)$ of an unapodized IDT to the overall filter transfer function $F(\omega)=F_1(\omega)F_2(\omega)$ is negligible, i.e. $F_1(\omega)\approx 1$. Then the filter frequency response (FR) $F(\omega)\approx F_2(\omega)$ depends on the function $F_2(\omega)$ only, and consequently within the limits of the δ-function model [22,23] a SAW filter synthesis becomes equivalent to a FIR digital filter synthesis [1]. Unfortunately, this is the only case to apply the techniques above without any adaptations.

Further, while in the passband the roll-off of the overall function $F(\omega)$ due to the frequency response $F_1(\omega)$ might easily be compensated by a proper predistortion of the desired magnitude function $F_0(\omega)$, this is not the case in the filter stopband where the function $F_1(\omega)$ is usually sign-alternated.

It is yet shown in the paper how an original SAW filter FR approximation problem may be converted to an auxiliary one by a proper modifying of both: a desired magnitude function $F_0(\omega)$ as well as a weight function $W_0(\omega)$. An auxiliary approximation problem is solvable by means of standard linear Chebyshev approximation techniques using the Remez exchange algorithm by virtue of the McClellan's computer program [18] for example. In addition to a frequency response $F_1(\omega)$, both an element factor [24-26] and a multistrip frequency response [23] might also be accounted for if necessary.

An optimal synthesis provides uniquely the best fit to a design target and leads to an apodized IDT of minimal length. But its drawback is a considerable amount of computations due to a large number of optimized variables (OV) even if one uses an efficient McClellan's program. There have been some efforts to accelerate the algorithm convergence [20,21], but the results obtained are not sufficient for a real-time design.

To overcome this problem a suboptimal SAW filter synthesis technique also based on the McClellan's computer program [18] is proposed in the paper. While maintaining optimal synthesis generality and flexibility, the latter considerably reduces an OV number and hence the storage and the computation time. The detailed suboptimal synthesis theory and some practical design aspects will be discussed.

1. SAW Filter Optimal Synthesis

1.1. Optimal Approximation Problem Formulation and Solution

The SAW filter to be designed consists of two linear phase IDT. Frequency response $F_1(\omega)$ of one of them is supposed to be given a priori while the other's $F_2(\omega)$ is optimized providing a Chebyshev (minimax) approximation of the desired magnitude shape function $F_0(\omega)$. Another case, where the fil-

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ter consists of two apodized IDT to be optimized, is beyond the scope of this paper and is treated separately [12,27-29].

There are no constraints on a magnitude shape function $F_0(\omega)$ imposed. It may be symmetrical, non-symmetrical, multipassband etc. [15,17].

A weighted error function $\Delta F(\omega)$ can be written in the form

$$\Delta F(\omega)=W_0(\omega)[F_0(\omega)-F(\omega)] \quad (1)$$

where $W_0(\omega)>0$ is a positive-defined weight function and the function

$$F(\omega)=\xi(\omega)F_1(\omega)F_2(\omega) \quad (2)$$

describes a linear phase SAW filter FR, with the functions $F_i(\omega)$, i=1,2, being attributed to the IDT array factors and the function

$$\xi(\omega)=\omega\xi_1(\omega)\xi_2(\omega)\xi_3(\omega) \quad (3)$$

accounting for IDT element factors $\xi_1(\omega)$ and $\xi_2(\omega)$ [24-26] and multistrip coupler FR $\xi_3(\omega)$ [23] if necessary. Hereafter we suppose $\xi_3(\omega)=1$ for simplicity.

It follows from the symmetry relations for a linear phase IDT that the array factors $F_i(\omega)$, i=1,2, can take one of the four following forms:

$$F_i(\varphi)=\begin{cases}\sum_{k=0}^{n_i} a_k\cos k\varphi \\ \sum_{k=1}^{n_i} b_k\sin k\varphi\end{cases} \quad \text{for an odd number } N_i=2n_i+1 \quad (4)$$

or

$$F_i(\varphi)=\begin{cases}\sum_{k=1}^{n_i} a_k\cos(k-1/2)\varphi \\ \sum_{k=1}^{n_i} b_k\sin(k-1/2)\varphi\end{cases} \quad \text{for an even number } N_i=2n_i, \quad (5)$$

where $\varphi=\pi\omega/\omega_\pi$, $\omega_\pi=v/2p$ is a synchronism frequency, with v being a SAW velocity and p being an IDT period, and $N_{1,2}$ is a total number of the IDT acoustical sources (gaps or electrodes), with amplitudes $A_0=a_0$ and $A_{\pm k}=(a_k\pm b_k)/2$, $k\neq0$, if numbered relatively to an IDT center.

In the quasistatic approximation [23] the element factors $\xi_i(\omega)$, i=1,2, are described by the expression [24-26]

$$\xi_i(\nu)=\frac{P_n(\cos\Delta)}{P_{-\nu}(-\cos\Delta)}\times\begin{cases}1 & \text{gap-weighted IDT}\\ \sin\pi\nu & \text{electrode-weighted IDT}\end{cases} \quad (6)$$

where

$\nu=\frac{\varphi}{2\pi}=\frac{\omega}{2\omega_\pi}$ dimensionless frequency variable;
$\Delta=\pi\eta$, $\eta=w/p$ being metallization ratio;
w, p electrode width and period (pitch);
$P_n(\cos\Delta)$ Legendre polynom of an order n;
$P_{-\nu}(-\cos\Delta)$ Legendre function;
$n=[\nu]$ spatial harmonic number $(n\leq\nu<n+1)$.

The approximation problem can be stated as follows: given a desired magnitude function $F_0(\omega)$ and a weight function $W_0(\omega)$, one wishes to minimize the absolute weighted error function

$$\delta=\|\Delta F(\omega)\|=\max_{\omega\in\Omega_\pi}|\Delta F(w)| \quad (7)$$

within an approximation interval $\Omega_\pi=\{\omega\in[0,\omega_\pi]\}$ over the set of coefficients of the optimized function $F_2(\omega)$.

A feature of the approximation problem above is a multiplicative nature of the approximating function $F(\omega)$ given by the expression (2), with the function $F_1(\omega)$ sign-alternated in general case. Unfortunately, the McClellan's computer program [18] can not be directly applied to solve this problem, with the only exception of a special case $\xi(\omega)F_1(\omega)=1$.

Instead of the initial problem (1)-(7) let us consider an auxiliary one with an error function

$$\hat{\Delta F}(\omega)=sign\{F_1(\omega)\}\Delta F(\omega)=\hat{W}_0(\omega)[\hat{F}_0(\omega)-F_2(\omega)] \quad (8)$$

where

$$\hat{W}_0(\omega)=W_0(\omega)\ |\xi(\omega)F_1(\omega)| \quad (9)$$

$$\hat{F}_0(\omega)=F_0(\omega)/|\xi(\omega)F_1(\omega)| \quad (10)$$

One must be careful to omit in (8)-(10) those frequencies ω_i at which $\xi(\omega)F_1(\omega)=0$. The function $\xi(\omega)$ is usually monotone within Ω_π, the points ω_i being zeros of the sign-alternated function $F_1(\omega)$ only. At these frequencies an error function $\Delta F(\omega)$ takes the fixed values $\Delta F(\omega_i)=W_0(\omega_i)F_0(\omega_i)$, and the approximation might fail if the desired function $F_0(\omega_i)\neq0$. Fortunately, if all the points ω_i are located in a filter stopband, where $F_0(\omega)=0$, then the error function $\Delta F(\omega_i)$ is also equal to zero. Therefore, there is no need to minimize an error at these points and their omitting does not influence an approximation accuracy.

Now an auxiliary approximation problem with an error function $\hat{\Delta F}_0(\omega)$ may be solved on a subset $\Omega_\pi=\{\omega\in[0,\omega_\pi],\ \omega\neq\omega_i\}$ by any linear Chebyshev approximation technique [7-19]. The McClellan's computer program [18] can easily be applied, with the initial data changed according to the formulae (9) and (10).

An optimal solution obtained has some interesting properties that will be discussed below.

1.2. Optimal Solution Properties

Due to optimality and uniqueness properties an optimal solution does not depend on the optimization technique applied, would it be the Remez exchange algorithm or the linear programming.

A feature of an optimal solution is a behaviour of an error function $\Delta F(\omega)$ within a filter stopband resulting from the relation (8) and from the Chebyshev alternation theorem [1]. According to this theorem, the error function $\Delta\hat{F}(\omega)$ and hence the error function $\Delta F(w)$ must exhibit on the subset $\hat{\Omega}_{\pi}$ at least n_2+1 equiripple extremuma, with n_2 being an order of the trigonometric polynomial $F_2(\omega)$. On the other hand the maximum number of the overall function $F(\omega)$ extremuma is defined by an order $n=n_1+n_2-1>n_2$ of the polynomial product $F_1(\omega)F_2(\omega)$. It is the difference between the extremuma number n and the alternation extremuma number n_2 that makes it possible for some extra extremuma between two neighbouring equiripple alternation ones to appear.

Indeed, unlike the usual alternation law, when two neighbour equiripple extremuma are always of the opposite signs, it follows from (8) that two extremuma 1 and 2 (Fig. 1) must have the same sign if an odd order real zero ω_i of the function $F_1(\omega)$ is placed between them. This results in an extra extremum 3 of the opposite sign and of the lower amplitude to appear between extremuma 1 and 2 in the neighbourhood of the frequency ω_i. It is worth noting that even order real zeros or complex-valued roots of the function $F_1(\omega)$ do not violate the habitual alternation low.

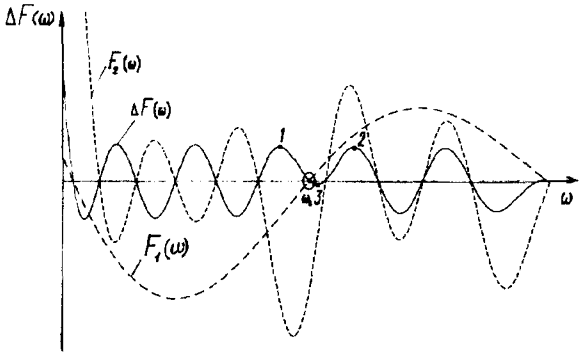
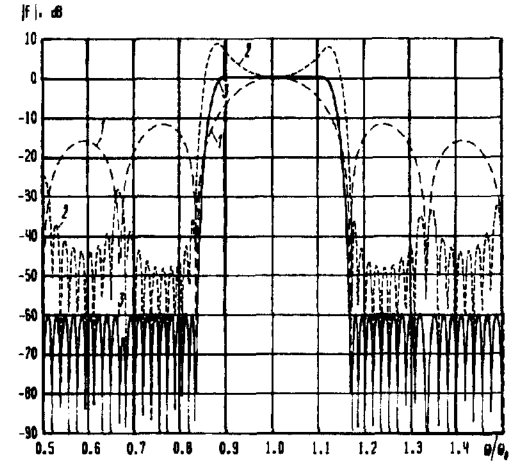


Fig. 1. Optimal Solution Error Function $\Delta F(\omega)$ and an Arrangement of Zeros and Extremuma of the Functions F1(w) and F2(w)

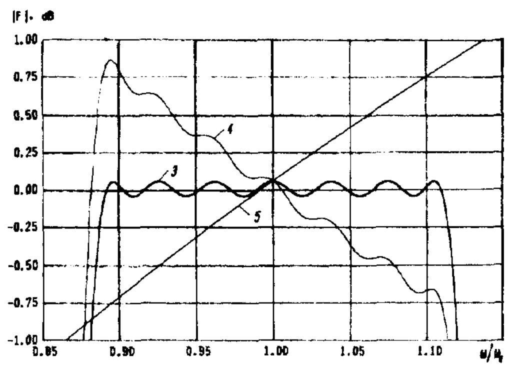
Besides, it is such a special arrangement of zeros and extremuma of the functions $F_1(\omega)$ and $F_2(\omega)$ in a filter stopband that ensures to a large degree a solution optimality, extremuma located in the neighbourhood of zeros and vice versa.

1.3. An Optimal Synthesis Design Example

The optimal synthesis technique was successfully applied to a SAW filter design. Fig. 2 shows an example of the optimal FR, with the curve 1 the FR $\sqrt{\omega}\,\xi_1(\omega)F_1(\omega)$ of the unapodized IDT, the curve 2 the FR $\sqrt{\omega}\,\xi_2(\omega)F_2(\omega)$ of the apodized IDT after optimization, and the curve 3 the overall optimal magnitude response $F(\omega)=\xi(\omega)F_1(\omega)F_2(\omega)$. For clarity, the polynomial product $F_1(\omega)F_2(\omega)$ (the curve 4) and the gap-weighted factor $\xi(\omega)=\omega\xi_1(\omega)\xi_2(\omega)$ (the curve 5) are also shown together with the filter passband ripple (the curve 3) in Fig. 2b. The function $\xi(\omega)$ for the metallization ratio $\eta=0,5$ was accounted for within a filter passband only. For convenience, the frequency characteristics are plotted versus a normalized frequency ω/ω_0 where ω_0 is a filter central frequency. An optimal solution was obtained using the McClellan's computer program [18] after a variable substitution (9), (10).



a) magnitude response



b) passband ripple

Fig. 2. SAW Filter Optimal Frequency Response

The design specifications are the following: the -3 dB passband width $\Delta\omega_0/\omega_0=25\%$, and the -3/-40 dB shape factor K=1,25. The IDT synchronism frequency is $\omega_{\pi}=2\omega_0$ that corresponds to IDT structures with splitted electrodes [30]. The IDT electrode numbers are $N_1=24$ and $N_2=200$ respectively.

As we can see from Fig. 2, the outofband attenuation is better than -60 dB and the passband peak-to-peak ripple is less than 0,1 dB.

The optimization was performed on the discrete frequency grid containing N_g=943 points with a discretization step $\Delta\omega_g=0,1\Delta\omega$, $\Delta\omega=2\omega_\pi/N_2$ being frequency sampling interval.

It took 110 iterations to obtain an optimal solution, the computation time being 32 minutes on a personal computer IBM PC/AT 286 with a math coprocessor.

2. SAW Filter Suboptimal Synthesis Technique

2.1. Suboptimal Approximation Problem Formulation and Solution

An optimal solution provides the best fit to a design target within a total approximation interval Ω_π and leads uniquely to an apodized IDT of minimal length. But the most serious drawback of an optimal synthesis is an excessive amount of computations. It is very desirable to find some way to reduce an optimized variables (OV) number n_2 and hence the computation time and the memory size needed. With this aim a suboptimal synthesis technique was elaborated which considerably reduces an OV number without significant sacrificing the approximation accuracy.

The key point of the suboptimal synthesis technique proposed is a splitting of the function $F_2(\omega)$ into two factors

$$F_2(\omega)=\bar{F}_2(\omega)\tilde{F}_2(\omega), \qquad (11)$$

with the function $\bar{F}_2(\omega)$ fixed and chosen a priori and the function $\tilde{F}_2(\omega)$ of the order $\tilde{n}_2< n_2$ optimized within an approximation subinterval $\Omega\in\Omega_\pi$. Outside the subinterval Ω approximation accuracy depends mainly on the function $\bar{F}_2(\omega)$ which must secure a sufficient outofband attenuation. A synthesis technique of such a wideband window-type function $\bar{F}(\omega)$ will be discussed some later.

Rewriting the approximating function $F(\omega)$ in the form

$$F(\omega)=\xi(\omega)\bar{F}_1(\omega)\tilde{F}_2(\omega), \qquad (12)$$

where $\bar{F}_1(\omega)=F_1(\omega)\bar{F}_2(\omega)$, we note the function $F(w)$ to be of the same structure (2), but with the function $F_1(\omega)$ replaced by the function $\bar{F}_1(\omega)$ and the optimized function $F_2(\omega)$ of the order n_2 replaced by the function $\tilde{F}_2(\omega)$ of the order $\tilde{n}_2< n_2$. Therefore, a suboptimal approximation problem with the approximating function (12) can be converted to an auxiliary one and solved like an optimal one considered above, with an order $\tilde{n}_2$ of the optimized function $\tilde{F}_2(\omega)$ decreased.

The Fourier coefficients of the functions $F_2(\omega)$, $\bar{F}_2(\omega)$, and $\tilde{F}_2(\omega)$ are related via the convolution [1], that yields the relation

$$n_2=\bar{n}_2+\tilde{n}_2-1. \qquad (13)$$

The higher is the order $\bar{n}_2$ of the fixed function $\bar{F}_2(\omega)$, the lower is the order $\tilde{n}_2$ of the optimized function $\tilde{F}_2(\omega)$. Hence it appears desirable to increase the order $\bar{n}_2$ of the function $\bar{F}_2(\omega)$ until an approximation accuracy deteriorating. As a matter of fact a suboptimal solution is inevitably inferior if compared to an optimal one, but by a proper choice of the function $\bar{F}_2(\omega)$ the difference might be made negligible, with the low order function $\tilde{F}_2(\omega)$ optimized.

Thus the function $\bar{F}_2(\omega)$ performs here a twofold role: to decrease an OV number and to secure at the same time a sufficient approximation accuracy.

2.2. Window Function Construction

Let us suppose that the function $\bar{F}_2(\omega)$ to be constructed is completely defined by the Z-transform roots [1] $z_i=e^{j\varphi_i}$ at the points $\omega_i\notin\Omega_{op}$ allocated outside the optimization subinterval $\Omega_{opt}\in\Omega$ of the width $\Delta\omega_{opt}$, where $\omega_i=i\Delta\omega$, i=0,1,2, ..., with $\Delta\omega=2\omega_\pi/N_2$ being the frequency sampling interval (Fig.3). Such a window-type function $\bar{F}_2(\omega)$ is to force the frequency samples

$$F_2(\omega_i)=\bar{F}_2(\omega_i)\tilde{F}_2(\omega_i) \qquad (14)$$

to be zero outside the optimization subinterval Ω_{opt}, while the other frequency samples at the points $\omega_i\in\Omega_{opt}$ optimized. It has a bell-like magnitude response, its sidelobes rapidly decreasing outside Ω_{opt}.

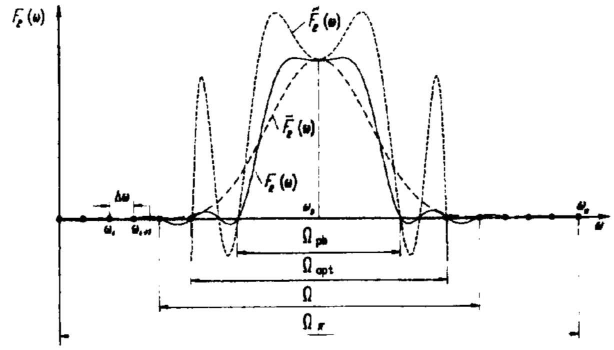
Using Z-transform properties we can write $\bar{F}_2(z)$ in the following form

$$\bar{F}_2(z)=\sum_{k=0}^{\bar{N}_2-1}\bar{A}_k z^k=\prod_{\omega_i\notin\Omega_{opt}}D_i(z)=\frac{1-z^{N_2}}{\prod_{\omega_i\in\Omega_{opt}}D_i(z)}, \qquad (15)$$

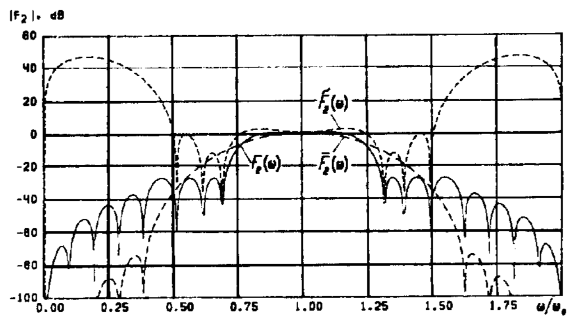
where

$$D_i(z)=\begin{cases}(z-z_i)(z-z_i^*), & z_i\neq\pm1;\\ (z-z_i), & z_i=+1 \text{ or } z_i=-1.\end{cases} \qquad (16)$$

The window function $\bar{F}_2(\varphi)$ is a Z-transform $\bar{F}_2(z)$ evaluated on the unit circle $z=e^{j\varphi}$.



a) linear scale magnitude response



b) Log magnitude response

Fig.3. Suboptimal Factorization of the Function F2(ω)

The coefficients $\bar{A}_k$ of the function

$$\bar{F}_2(\varphi)=\prod_{\omega_i \notin \Omega_{opt}} D_i(\varphi) \qquad (17)$$

of the order $\bar{N}_2$ could easily be calculated using the recurrent convolution of the root factors

$$D_i(\varphi)=\begin{cases} \cos\varphi-\cos\varphi_i, & \varphi_i \neq n\pi; \\ j\sin\varphi/2, & \varphi_i=2n\pi; \\ \cos\varphi/2, & \varphi_i=(2n+1)\pi, \end{cases} \qquad (18)$$

where $n=0,\pm1,\pm2,\ldots$. But usually there is no need to know the coefficients $\bar{A}_k$, and one can directly apply the formula (17) for calculations.

The wider is the bandwidth $\Delta\omega_{opt}$ of the optimization subinterval Ω_{opt}, the higher is a stopband attenuation of the function $\bar{F}_2(\omega)$ and the closer is a suboptimal solution to an optimal one. The number m of the frequency samples to be optimized is found from the simple relation

$$m = \frac{\Delta\omega_{opt}}{\Delta\omega}+1 = \frac{1}{2}N_2\frac{\Delta\omega_{opt}}{\omega_\pi}+1. \qquad (19)$$

For m > 10-15 one can use the following approximation for the OV number gain estimation

$$\frac{\tilde{n}_2}{\bar{n}_2} \approx \frac{m}{\bar{n}_2} \approx \frac{\Delta\omega_{opt}}{\omega_\pi}. \qquad (20)$$

In other words, the gain in the OV number is roughly proportional to the relative bandwidth $\Delta\omega_{opt}/\omega_\pi$ of the optimization subinterval Ω_{opt}.

The function $\bar{F}_2(\omega)$ may also be constructed by another way using a zero-extracting technique as opposed to a zero-inserting one above.

Indeed, an equivalent representation of the function $\bar{F}_2(\omega)$ follows from (15)

$$\bar{F}_2(\varphi) = \frac{\sin\frac{N_2}{2}\varphi}{\prod_{\omega_i \in \Omega_{opt}} D_i(\varphi)}, \qquad (21)$$

where an uncertainty at the singularity points $\varphi=\varphi_i$ can be avoided by using the Lopital's rule.

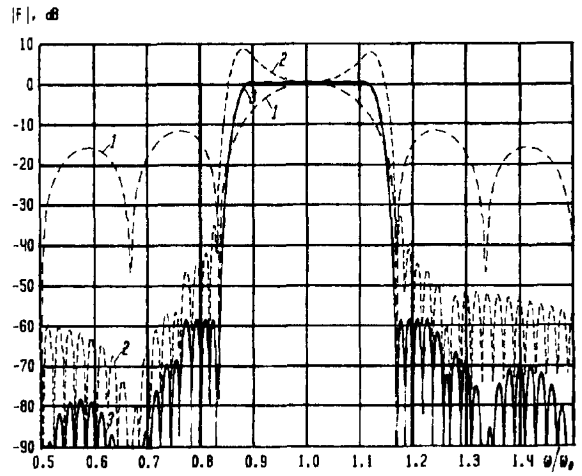
Here an amount of computations is proportional to $\tilde{n}_2$ instead of $\bar{n}_2$ in the formula (17). Hence the formula (21) is more preferable for $\tilde{n}_2 < \bar{n}_2$. Thus a rational choice of the function $\bar{F}_2(\omega)$ representation form allows somewhat to minimize an amount of computations in the function $\bar{F}_2(\omega)$ calculating.

As a rule of thumb we usually choose the optimization interval Ω_{opt} to be (6-10)$\Delta\omega$ wider than a filter passband Ω_{pb} (Fig.3). In turn an approximation subinterval Ω might be equal or (1-2)$\Delta\omega$ wider than optimization subinterval $\Omega_{opt} \in \Omega$. It is a rapid sidelobe attenuation of the window function on $\bar{F}_2(\omega)$ outside Ω_{opt} that secures an approximation accuracy to be not worse than that obtained within Ω, where the error function is minimized.

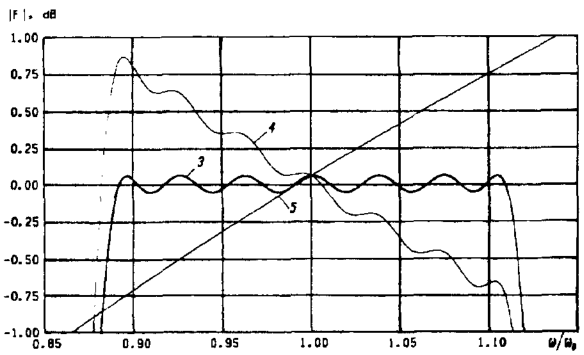
2.3. A Suboptimal Synthesis Design Example

For the comparison's sake it is convenient to use the same design example as for an optimal synthesis above. The suboptimal FR obtained is plotted in Fig.4. The initial specifications and the denominations are the same as for an optimal synthesis.

The suboptimal synthesis data are the following. The bandwidth of the optimization interval is $\Delta\omega_{opt}/\omega_{\pi}$=25%, with an approximation subinterval $\Omega\approx\Omega_{opt}$. The optimization was performed on the frequency grid containing N_g=209 points as opposed to the optimal synthesis, where the grid point number was as large as N_g=943.



a) magnitude response



b) passband ripple

Fig. 4. SAW Filter Suboptimal Frequency Response

The IDT electrode numbers are the same N_1=24 and N_2=200, but due to an a priori factorization of the function $\tilde{F}_2(\omega)$ the OV number was decreased from n_2=100 to $\tilde{n}_2$=25, with the number $m=n_2$ of the frequency samples optimized.

The stopband attenuation is -58,9 dB and the passband ripple is ≈0,11 dB. The detailed comparison of Fig. 2 and Fig. 4 shows that the solutions practically coincide within the subinterval Ω.

The difference in approximation accuracy is about 1,3 dB in a filter stopband and 0,015 dB in a filter passband that is quite negligible from the practical point of view.

It is the OV number gain $\tilde{n}_2/n_2$=25% in conjunction with a sufficient reduction of the frequency grid point number N_g that allows the computation time to be drastically reduced from 32 minutes to 15 seconds only, i.e. more than 125 times.

Conclusion

The optimal and suboptimal SAW filter design techniques have been considered above, both based on the McClellan's computer program [18].

Unfortunately, being of great theoretical importance, an optimal synthesis is rather impracticable one for a real-time design due to an excessive amount of computations.

While maintaining optimal synthesis generality and flexibility, the suboptimal synthesis technique allows to considerably reduce an OV number and hence the storage and the computation time nearly without sacrificing the approximation accuracy. Usually the difference between optimal and suboptimal approximations does not exceed 1-2 dB in a filter stopband and 0,01-0,05 dB within a passband that is more than acceptable for practical design purposes. Moreover, this slight discrepancy might easily be compensated by a small electrode number of an apodized IDT increasing if one wishes.

A feature of an suboptimal synthesis above is that an amount of computations depends mainly on the filter magnitude shape specifications but not on its central frequency. Consequently, for the most SAW filters the computation time is fairly small taking from some seconds to some minutes on a personal computer IBM PC/AT 286 with a math coprocessor. As a result a narrowband fast cut-off filter synthesis with an electrode number of several hundred and even thousand becomes possible due to dramatic OV number reducing.

It is the inherent efficiency of the Remez exchange algorithm in conjunction with a low order optimized function that makes the suboptimal synthesis technique proposed very attractive one for a SAW filter computer-aided design.

The design experience confirmed fast convergence, high computation speed, reliability, and flexibility of the suboptimal synthesis technique proposed, and good agreement between theory and experiment was obtained within the limits of the model and design constraints applied.

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