Optimal and Suboptimal Design of SAW Bandpass Filters Using the Remez Exchange Algorithm

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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. The optimal synthesis provides uniquely the best fit to a design target, but its major drawback is an excessive amount of computations. The suboptimal synthesis technique proposed allows a considerable reduction of the amount of computations without significant sacrificing of the approximation accuracy. Thus computer runtime and storage are greatly saved if compared to the 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.**

I. Introduction

SEVERAL surface acoustic wave (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].

However, 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 design simplifications, sometimes without sufficient foundations. For example, it is usually supposed that within a filter passband a contribution $F_1(\omega)$ of an unapodized interdigital transducer (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 for the techniques above to be applied 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 the 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 this paper how an original SAW filter FR approximation problem may be converted to an auxiliary one by the proper modifying of both a desired magnitude function $F_0(\omega)$ as well as a weight function $W_0(\omega)$. The 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 the frequency response $F_1(\omega)$, both an element factor [24]–[26] and/or a multistrip frequency response [23] might also be accounted for if necessary.

The optimal synthesis uniquely provides the best fit to a design target, but its major drawback is a considerable amount of computations due to a large number of optimized variables (OV's) even if one uses the 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.

Given a band-limited frequency response, the number of variables to be optimized may be considerably reduced by applying the sampling theorem in the time or frequency domain [7], [8], [16], [27], [28], the linear or nonlinear programming techniques being used for optimizing the sample values.

Another approach to reduce the OV number is taken in [29] where only the passband Z-transform roots are to be found by the parametric optimization technique, while the stopband roots are determined in the closed form using the frequency transformations.

However, despite the relatively small OV number, these design procedures remain rather time consuming due to the intrinsic computation slowness of the optimization techniques applied.

Contrary to this, the proposed suboptimal synthesis technique ingeniously exploits the same efficient McClellan's computer program [18] for the considerably reduced OV number. This is accomplished by factorizing *a priori* the optimized function, with the majority of the stopband zeros prescribed and expressed in the closed form. Thus the storage and the computation time are greatly gained if compared to the optimal synthesis while maintaining its generality and flexibility. The detailed suboptimal synthesis theory and some practical design aspects will be discussed.

II. SAW Filter Optimal Design

A. 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

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optimized providing a Chebyshev (mini-max) approximation of the desired magnitude shape function $F_0(\omega)$. Another case, where the filter consists of two apodized IDT to be optimized, is beyond the scope of this paper and is treated separately in [12], [30]–[32].

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

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

$$\Delta F(\omega) = W_0(\omega)[F_0(\omega) - F(\omega)] \tag{1}$$

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

$$F(\omega) = \xi(\omega)F_1(\omega)F_2(\omega) \tag{2}$$

describes a linear phase SAW filter FIR. Here the functions $F_i(\omega), i = 1, 2$, are attributed to the IDT array factors [22], with the skewing factor $\xi(\omega)$ introduced to account for the IDT element factors [23]–[26] and/or the multistrip coupler FR [23], etc. For a linear phase IDT, each array factor $F_i(\omega)$ is the *cosine* or *sine* trygonometric polynomial [1] of the order $n_i = [N_i/2]$ where N_i is a total number of the IDT acoustical sources (gaps or electrodes).

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

$$\delta = \|\Delta F(\omega)\| = \max_{\omega \in \Omega_\pi} |\Delta F(\omega)| \tag{3}$$

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

A feature of the previous approximation problem is the multiplicative nature of the approximating function $F(\omega)$ given by (2), with the function $F_1(\omega)$ sign-alternated in the general case. Unfortunately, McClellan's computer program [18] cannot 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)–(3) let us consider an auxiliary one with an error function

$$\Delta \hat{F}(\omega) = \text{sign}\{F_1(\omega)\}\Delta F(\omega) = \hat{W}_0(\omega)[\hat{F}_0(\omega) - F_2(\omega)] \tag{4}$$

where

$$\hat{W}_0(\omega) = W_0(\omega)|\xi(\omega)F_1(\omega)|, \tag{5}$$

$$\hat{F}_0(\omega) = \frac{F_0(\omega)}{|\xi(\omega)F_1(\omega)|}. \tag{6}$$

One must be careful to omit in (4)–(6) 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 the 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 $F_0(\omega_i) \neq 0$. 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

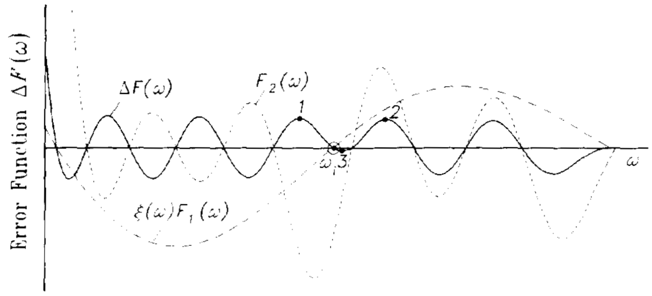


Fig. 1. Optimal solution error function $\Delta F(\omega)$ (—) and the arrangement of zeros and extremuma of the functions $\xi(\omega)F_1(\omega)$ (— — —) and $F_2(\omega)$ (- - -).

an error at these points and their omitting does not influence an approximation accuracy.

Now the auxiliary approximation problem with the error function $\Delta \hat{F}(\omega)$ may be solved on the subset $\hat{\Omega}_\pi = \{\omega \epsilon [0, \omega_\pi], \omega \neq \omega_i\}$ 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 (5) and (6).

The optimal solution obtained has some interesting properties that will be discussed in the following.

B. Optimal Solution Properties

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

A feature of the optimal solution is the behavior of the error function $\Delta F(\omega)$ within a filter stopband resulting from (4) and from the Chebyshev alternation theorem [1]. According to this theorem, the error function $\Delta \hat{F}(\omega)$, and hence, the error function $\Delta F(\omega)$ must exhibit on the subset $\hat{\Omega}_\pi$ at least $n_2 + 1$ equi-ripple extremuma, with n_2 being the 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 the 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 neighboring equi-ripple alternation ones to appear.

Indeed, unlike the usual alternation law, when two neighbor equi-ripple extremuma are always of the opposite signs, it follows from (4) 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 neighborhood 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 law.

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 neighborhood of zeros and vice versa.

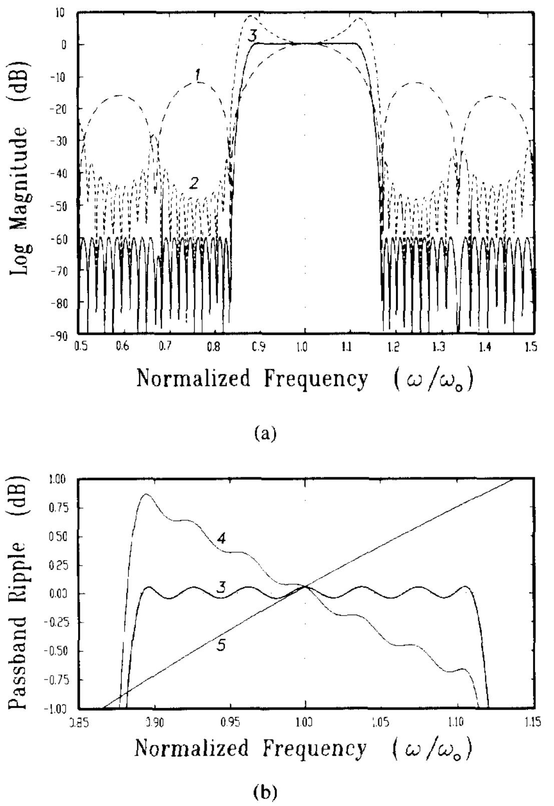


Fig. 2. Optimal synthesis design example. (a) Magnitude response. (b) Passband ripple. (1) Unapodized IDT response $\sqrt{\xi(\omega)}F_1(\omega)$ (— — —). (2) Apodized IDT response $\sqrt{\xi(\omega)}F_2(\omega)$ (- - -). (3) Overall optimal magnitude response $\xi(\omega)F_1(\omega)F_2(\omega)$(—). (4) Polynomial product $F_1(\omega)F_2(\omega)$ (—). (5) The skewing factor $\xi(\omega)$ (—).

C. Optimal Synthesis Design Example

As an example, the optimal synthesis procedure was applied to design a SAW filter with the -3-dB fractional bandwidth of 25 % and the $-3/-40$-dB shape factor of 1.25. The quasi-static approximation [23, eq. (4.97)] was used for the SAW filter frequency response simulation. The design results obtained are shown in Fig. 2 where the curves 1 and 2 correspond to the unapodized IDT FR $\sqrt{\xi(\omega)}F_1(\omega)$ and the apodized IDT FR $\sqrt{\xi(\omega)}F_2(\omega)$ respectively, with the curve 3 being the optimal overall magnitude response $F(\omega) = \xi(\omega)F_1(\omega)F_2(\omega)$. For clarity, the polynomial product $F_1(\omega)F_2(\omega)$ (curve 4) and the skewing factor $\xi(\omega)$ (curve 5) are also shown together with the filter passband ripple (curve 3) in Fig. 2 (b). The function $\xi(\omega) = \omega\zeta^2(\omega)$ comprising the skewing frequency factor ω and the gap-weighted IDT element factor $\zeta(\omega)$ [23, eq. (4.96)] for the metallization ratio of 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 the filter central frequency. The optimal solution was obtained using the McClellan's computer program [18] after the variables substitution (5), (6).

The IDT electrode numbers are $N_1 = 24$ and $N_2 = 200$, respectively. The IDT synchronism frequency is $\omega_\pi = 2\omega_0$ which corresponds to both IDT structures with split electrodes [33].

The optimization was performed on the discrete frequency grid containing $N_g = 943$ points with a discretization step $\delta\omega = 0.1\Delta\omega, \Delta\omega = 2\omega_\pi/N_2$ being the frequency sampling interval.

As we can see from Fig. 2, the out-of-band attenuation is better than -60 dB and the passband peak-to-peak ripple is less than 0.1 dB. It took 110 iterations to obtain an optimal solution, the computation time being 32 min on a personal computer IBM PC/AT 286 with a math co-processor.

III. Saw Filter Suboptimal Synthesis Technique

A. Suboptimal Approximation Problem Formulation and Solution

The optimal solution described previously uniquely provides the best fit to a design target within a total approximation interval Ω_π. However, the most serious drawback of the optimal synthesis is an excessive amount of computations. It is very desirable to find some way to reduce the OV's number n_2, and hence, the computation time and the memory size needed. With this aim the suboptimal synthesis technique was elaborated, which considerably reduces an OV number without significant sacrificing of the approximation accuracy.

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

$$F_2(\omega) = \overline{F}_2(\omega)\tilde{F}_2(\omega), \tag{7}$$

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

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

$$F(\omega) = \xi(\omega)\overline{F}_1(\omega)\tilde{F}_2(\omega) \tag{8}$$

where $\overline{F}_1(\omega) = F_1(\omega)\overline{F}_2(\omega)$, we note the function $F(\omega)$ to be of the same structure (2), but with the function $F_1(\omega)$ replaced by the function $\overline{F}_1(\omega)$ and the optimized function $F_2(\omega)$ of the order n_2 replaced by the function $\tilde{F}_2(\omega)$ of the reduced order $\tilde{n}_2 < n_2$. Therefore, a suboptimal approximation problem with the approximating function (8) can be converted to an auxiliary one and solved like the optimal one considered previously, with the order $\tilde{n}_2$ of the optimized function $\tilde{F}_2(\omega)$ decreased.

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

$$n_2 = \overline{n}_2 + \tilde{n}_2 - 1. \tag{9}$$

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

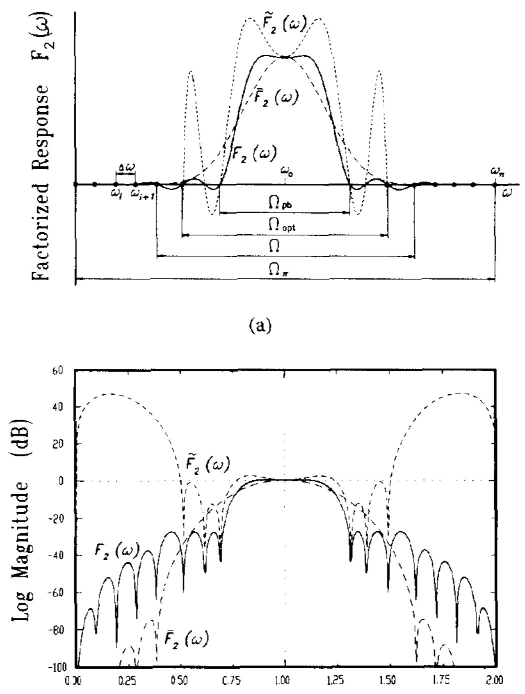


Fig. 3. Suboptimal factorization of the function $F_2(\omega)$(—) comprising the window-type function $\overline{F}_2(\omega)$ (— — —) and the optimized function $\tilde{F}_2(\omega)$ (- - -). (a) Linear scale magnitude response. (b) Log magnitude response.

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

B. Window Function Construction

Assume the function $\overline{F}_2(\omega)$ to be completely defined by its Z-transform roots [1] $z_i = e^{j\varphi_i}$ at the frequency points $\omega_i \notin \Omega_{\text{opt}}$ allocated outside the optimization subinterval $\Omega_{\text{pb}} \subseteq \Omega_{\text{opt}} \subseteq \Omega$ of the width $\Delta\omega_{\text{opt}}$, where $\omega_i = i\Delta\omega, i = 0, 1, 2, \cdots$, with $\Delta\omega = 2\omega_\pi / N_2$ being the frequency sampling interval and Ω_{pb} defining the filter total passband (Fig. 3). Such a window-type function $\overline{F}_2(\omega)$ forces the frequency samples

$$F_2(\omega_i) = \overline{F}_2(\omega_i)\tilde{F}_2(\omega_i) \tag{10}$$

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

In the function of the angle variable $\varphi = \pi\omega/\omega_\pi, \omega_\pi$ being a synchronism frequency, the response $\overline{F}_2(\varphi)$ is a Z-transform $\overline{F}_2(z)$ evaluated on the unit circle $z = e^{j\varphi}$, and all the roots $z_i = e^{j\varphi_i}$ are allocated on this circle.

Using Z-transform properties [1] we can derive the following analytical expression:

$$\overline{F}_2(\varphi) = \sum_{k=0}^{\overline{N}_2 - 1} \overline{A}_k e^{jk\varphi} = \prod_{\omega_i \notin \Omega_{\text{opt}}} D_i(\varphi) = \frac{\sin \frac{N_2}{2}\varphi}{\prod_{\omega_i \in \Omega_{\text{opt}}} D_i(\varphi)} \tag{11}$$

where the elemental root factors $D_i(\varphi)$ are as follows

$$D_i(\varphi) = \begin{cases} \cos\varphi - \cos\varphi_i, & \varphi_i \neq n\pi \\ j\sin\frac{\varphi}{2}, & \varphi_i = 2n\pi, \quad n = 0, \pm 1, \pm 2, \cdots \\ \cos\frac{\varphi}{2}, & \varphi_i = (2n+1)\pi \end{cases} \tag{12}$$

The coefficients $\overline{A}_k$ of the function $\overline{F}_2(\omega)$ of the order $\overline{N}_2$ could easily be calculated using the recurrent convolution of the elemental root factors $D_i(\varphi)$. However, there is usually no need to know explicitly the coefficients $\overline{A}_k$, and one can directly multiply the root factors (12) in (11) where for the rational form representation of the function $\overline{F}_2(\varphi)$ an uncertainty at the singularity points $\varphi = \varphi_i$ can be avoided by using Lopital's rule.

The wider the bandwidth $\Delta\omega_{\text{opt}}$ of the optimization subinterval Ω_{opt}, the higher the stopband attenuation of the function $\overline{F}_2(\omega)$ and the closer the suboptimal solution to the optimal one.

The number m of the frequency samples to be optimized is found from the simple relation

$$m = \frac{\Delta\omega_{\text{opt}}}{\Delta\omega} + 1 = \frac{1}{2}N_2\frac{\Delta\omega_{\text{opt}}}{\omega_\pi} + 1. \tag{13}$$

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

$$\frac{\tilde{n}_2}{n_2} \approx \frac{m}{n_2} \approx \frac{\Delta\omega_{\text{opt}}}{\omega_\pi}. \tag{14}$$

In other words, the gain in the OV number is roughly proportional to the relative bandwidth $\Delta\omega_{\text{opt}}/\omega_\pi$ of the optimization subinterval Ω_{opt} which *cannot* be narrower than a filter total passband Ω_{pb}, giving the low limit for the OV number reduction.

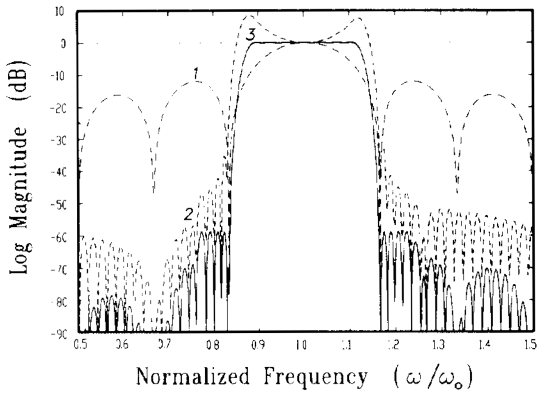
As a rule of thumb, we usually choose the optimization subinterval Ω_{opt} to be (6–10) $\Delta\omega$ wider than a filter total passband Ω_{pb} (Fig. 3). In turn, the approximation subinterval Ω might be equal or (1–2)$\Delta\omega$ wider than the optimization subinterval $\Omega_{\text{opt}} \subseteq \Omega$. It is a rapid sidelobe attenuation of the window function $\overline{F}_2(\omega)$ outside Ω_{opt} that secures the approximation accuracy over the total interval Ω_π, to be comparable with that obtained within $\Omega \subset \Omega_\pi$ where the error function is minimized.

It is worthy to note that some unpredictable accuracy deterioration may arise in the proximity of the approximation subinterval Ω. However, this undesirable error function behavior can be precluded by imposing the more stringent specifications on the stopband attenuation at the end points (or at the narrow-end segments) of the subinterval Ω (say -70 dB instead of the desired -60 dB).

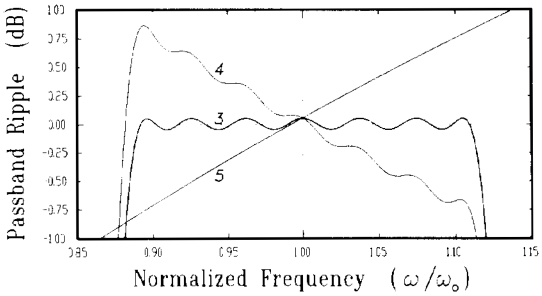
From the aforementioned, one can also conclude that this is virtually an approach to implicitly apply the McClellan's computer program [18] to the frequency sampling optimization [1], [7]–[10] instead of the time consuming linear programming technique commonly used.

C. Suboptimal Synthesis Design Example

For the comparison's sake, it is convenient to use the same design example as for the optimal synthesis described



(a)



(b)

Fig. 4. Suboptimal synthesis design example. (a) Magnitude response. (b) Passband ripple. (1) Unapodized IDT response $\sqrt{\xi(\omega)}F_1(\omega)$ (— — —). (2) Apodized IDT response $\sqrt{\xi(\omega)}F_2(\omega)$ (- - -). (3) Overall optimal magnitude response $\xi(\omega)F_1(\omega)F_2(\omega)$(____). (4) Polynomial product $F_1(\omega)F_2(\omega)$ (—). (5) The skewing factor $\xi(\omega)$ (—).

earlier. The suboptimal FR obtained is plotted in Fig. 4. The initial specifications and denominations are the same as for the optimal synthesis.

The suboptimal synthesis data are the following. The relative bandwidth of the optimization subinterval Ω_{opt} is $\Delta\omega_{\text{opt}}/\omega_\pi = 25\%$, with an approximation subinterval $\Omega \approx \Omega_{\text{opt}}$. 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 $N_g = 943$.

The IDT electrode numbers are the same, $N_1 = 24$ and $N_2 = 200$, but due to the *a priori* factorization of the function $F_2(\omega)$ the OV number was decreased from $n_2 = 100$ to $\tilde{n}_2 = 25$, with the number $m = \tilde{n}_2$ of the frequency samples optimized.

The detailed comparison of Figs. 2 and 4 shows that both the solutions practically coincide within the subinterval Ω; the stopband attenuation and the passband ripple being -58.9 dB and ≈ 0.11 dB for the suboptimal synthesis. The difference in the approximation accuracy for the optimal and suboptimal design examples of about 1.3 dB in the filter stopband and of less than 0.015 dB in the filter passband 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 min to 15 s only, i.e., more than 125 times.

IV. Conclusion

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

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

While maintaining optimal synthesis generality and flexibility, the suboptimal synthesis technique allows a considerable reduction of 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 increasing of the apodized IDT electrode number if one wishes.

A feature of the suboptimal synthesis above is that the amount of computations depends mainly on the filter magnitude shape specifications but not on its center frequency. Consequently, for 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 co-processor. As a result, a narrowband fast cut-off filter synthesis with an electrode number of several hundred and even of several thousand becomes possible due to dramatic OV number reducing.

It is the inherent efficiency of the Remez exchange algorithm in conjunction with a reduced order optimized function that makes the proposed suboptimal synthesis technique 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, and good agreement between theory and experiment was obtained within the limits of the model and design constraints applied.

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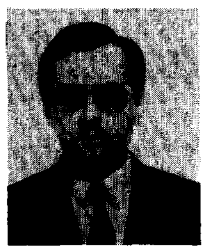

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