A Novel Design Method for SAW Bandpass Filter

Yang Chaobin Xu Jilin Huang Xianfu
(Dept. of Elect. Eng., UEST of China Chengdu 610054)

Abstract A novel design method of surface acoustic wave (SAW) bandpass filter is discussed. After resampling at the interval of M, the original desired weighting function is decomposed into two weighting functions with less order by the zero-separating of Z-transform, and then represented by interpolation derived from the signal reconstruction theory and can be furtherly suboptimized using the standard Remez change method. The simulation results of a Blackman spectrum-shaping filter with the center frequency of 70 MHz, the roll-off factor of 0.19, 0.35 dB ripple, 67 dB rejection, confirm that this method is very simple and effective.

Key words surface acoustic wave; resample and interpolation; zero-separating; optimization; Remez change method

The design of surface acoustic wave (SAW) filter[1] is based on the finite impulse response (FIR) digital filter theory, and the optimal synthesis method employs window functions, Remez change method or linear programming[2~5], as well as other optimal and sub-optimal algorithms.

The SAW filter can be constructed as in-line filter, consisting of an apodized interdigital transducer (IDT) and an unapodized IDT, or multistrip coupler (MSC) filter consisting of two apodized IDTs in separated acoustic tracks, coupled by a MSC. If high stopband rejections are required, the MSC configuration must be used, showed in Fig. 1.

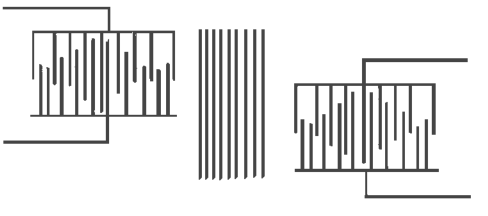

Fig. 1 The structure of MSC SAW filter

Given a band-pass frequency response, thousands of tap with small weights are required to achieve narrow transition bandwidth, wide passband and low ripple. The numbers of tap to be optimized in practical SAW filter design may be considerably reduced by sampling theory[6] in time or frequency domain. Another approach to reduce the tap is taken in Ref. [7], where only the pass-band Z-transform roots are found by the parametric optimization technology, while the stopband roots are determined in the closed-form using Z-transform. For MSC SAW filter, the desired frequency response must be separated into two parts which represents the response of the two IDT respectively. In practice, the above method can not be used directively.

In this paper, the author proposes a new decomposition method using resampling and interpolation method derived from signal reconstruction theory which can reduce the numbers of tap and combine zero-separating and optimum and suboptimum to design practical SAW filter using the standard Remez change algorithm. Section 2 presents the basic design procedure and optimal algorithm, concerning frequency response. Section 3 gives the simulation results of a SAW Blackman

(20) Received November 12, 1996

spectrum-shaping filter for a QAM digital communication. At last the conclusions are given

1 Decomposition and Suboptimization

The design principle of SAW filter is based on the delta impulse response model. Frequency response of SAW filter with two interdigital transducers is given by

$$H(f)=\sum_{m=0}^{M}\sum_{n=0}^{N}h(m,n)\exp(-jw\tau_{mn}) \tag{1}$$

where $h(m,n)$ and $\tau(m,n)$ are interactional weighting coefficient and delay between two interdigitals. For the MSC SAW filter, the above total response is equal to

$$H(f)=\sum_{n=0}^{N}h(n)\exp(-j2\pi fnT)=H_1(f)H_2(f)=$$

$$\sum_{n=0}^{N_1}h_1(n)\exp(-j2\pi fnT)\sum_{m=0}^{N_2}h_2(m)\exp(-j2\pi fmT) \tag{2}$$

where T is IDT sample interval; N_1 and N_2 are the numbers of taps in the two transducers respectively. $N=N_1+N_2$, $h(n)$, $h_1(n)$ and $h_2(n)$ are weight coefficient or overlap length of the two apodized interdigitals. In the theory, with the Z-plane representation of $H(f)$, by separating the zeros into two groups, $H(Z)$ can be written as the product of two polynomials which expresses the two transfer responses of the MSC filter

$$H(Z)=\sum_{n=0}^{N}h(n)Z^{-n}=h(0)Z^{-N}\prod_{n=1}^{N}(Z-Z_{0n})= \tag{3}$$

$$h_1(0)Z^{-N_1}\prod_{n=1}^{N_1}(Z-Z_{1n})\times h_2(0)\prod_{n=1}^{N_2}(Z-Z_{2n})=H_1(Z)H_2(Z) \tag{4}$$

where Z_{0n}, Z_{1n}, Z_{2n} are zeros assigned to $H(Z)$, $H_1(Z)$ and $H_2(z)$ respectively. $h(0)=h_1(0)^*h_2(0)$.

Since all weights $h(n)$ are real, the zeros are either real or occur in conjugate complex pair. To achieve narrow and precise transition bandwidth, thousands of tap with small weight are required, which makes the separating of zeros very difficult. In this case, a novel decomposition is used.

Firstly, the original tap weights $h(n)$ are resampled at the intervals of M (this resample is different from Nyquist sample), then the decomposition is applied to the resampled weighted function $h_s(n)$ with more less order which may still be written as following with zeros separating

$$H_s(Z)=\sum_{m=0}^{N/M}h(Mm)Z^{-m}=h_s(0)\prod_{m=1}^{N/M}(Z-Z_{osm})= \tag{5}$$

$$h_{1s}(0)\prod_{i=1}^{N_{1s}}(Z-Z_{10si})h_{2s}(0)\prod_{j=1}^{N_{2s}}(Z-Z_{20sj})=\sum_{i=0}^{N_{1s}}h_{1s}(i)Z^{-i}\sum_{j=0}^{N_{2s}}h_{2s}(j)Z^{-j} \tag{6}$$

where $N_{1s}+N_{2s}=N/M$, $h_s(0)=h_{1s}(0)^*h_{2s}(0)$.

In this case, the zeros must be distributed appropriately between two IDTs to minimize the difference between the response of the two sub-filters and the insertion loss, and to achieve precise response. Using Eq. (6), two decomposed resampled weighting sequences $h_{1s}(n)$ and $h_{2s}(n)$ are obtained, and can be considered as two staggered sampling sequences which are zeros at the non-resampled points as well as they may be represented by interpolation method according to the signal reconstruction theory

$$h_1(n)=\sum_{k=0}^{N_{1s}}h_{1s}(k)\operatorname{sinc}[\pi(n/m-k)] \qquad n=0\text{ to }N_1 \tag{7}$$

$$h_2(n) = \sum_{k=0}^{N_{2s}} h_{2s}(k)\,\mathrm{sinc}[c(n/M - k)] \qquad n = 0 \text{ to } n_2 \tag{8}$$

and

$$H(k) = H_1(k) \times H_2(k) \tag{9}$$

Therefore, with these two original weighting value in Eqs. (7) and (8), we suppose that the function $H_1(k)$ is fixed, and choose a priori and the function $H_2(k)$ of the reduced order $N_2 < N_1 < N$. In order to solve the original optimization problem: given a desired response $H_D(k)$ and error weight function $W_e(k)$, we suppose to minimize the error function

$$\Delta H(k) = |W_e(k)[H_D(k) - H_1(k)H_2(k)| \tag{10}$$

within the interval $K = \{k \in (0, k_x)\}$.

Instead of Eq. (9), we consider a suboptimal approximation function which can be converted to an auxiliary

$$\Delta H(k) = |\overline{W}_e(k)[H_D(k)/H_1(k) - H_2(k)]| \tag{11}$$

where $\overline{W}_e(k) = W_e(k)H_1(k)$ and solved like an optimal one obtained by standard Remex algorithm previously within the same internal, but with less order N_2. For $H_1(k)$, the above procedure may be done again. Thus the function $H_1(k)$ has a two-fold role; to dectease the number of variable to be optimized and to secure at the same time a sufficient approximation accuracy.

2 Simulation Results

According to the novel optimization procedure in the prevlous section, a 70 MHz Blackman SAW spectrum-shaping filter (SSF) with roll-off factor of 0.19 is designed. The total system frequency response $H(f)$ and $h(n)$ is shown in Fig. 2 and Fig. 3. The weight coefficience gives the ripple of 0.5 dB in the passband and the attenuation of 60 dB in rejection bands. The BW is about 30% , and transition BW is about 3% of the center the frequency. This requires about 1 000 taps with many small weights.

The orignal weighting function is resampled with the perlod M of 25. Then 40 taps are used to calculate the zeros shown in Fig. 4 for the decomposition. Fig. 5 shows the separated two decomposed weighting function, and the represented functions of two IDTs of MSC filter by interpolation are shown in Fig. 6.

In practical case, the digital rectangular pulse has a sinc$X$ spectrum. So, it is necessary to compensate by dividing this shaping factor into the transmitting or receiving SSF specification.

With above data, the interpolated two function $h_1(n)$ and $h_2(n)$, and the help of standard Remez optimum, the two part-responses of the two apodized IDTs and the overall responses are obtained, shown in Fig. 6. Of the simulation results, the ripple in passband is about 0.35 dB, stopband rejection is 67 dB, 1 dB~ 60 dB transition band is 1.5 MHz. This filter can meet the designed demands of SSF. This confirms the method is correct and effective.

In the above practical simulation, the order $N_1 \approx N_2$, Hence, it appears desirable to increase the order N_1 of $H_1(k)$ until the approximation accuracy is deteriorating.

As well as the second-order effects (charge distribution, end effect, diffraction, circuit efects) in the MSC will distort the transfer response obtained by the delta impulse model, this method requires more accurate model of MSC to compensate the second-effects, which is to be researched.

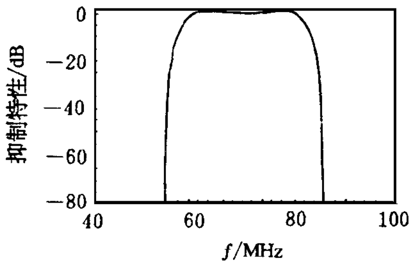

Fig. 2 Desired SSF $H(f)$

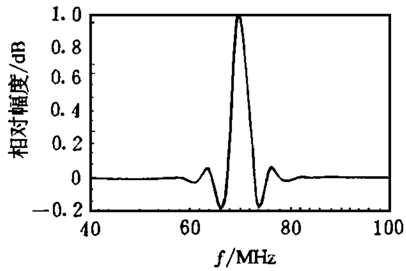

Fig. 3 Desired SSF $h(n)$

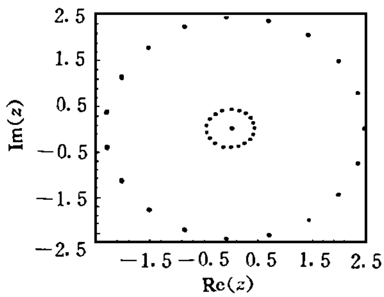

Fig. 4 Zeros' distribution of $H_s(Z)$

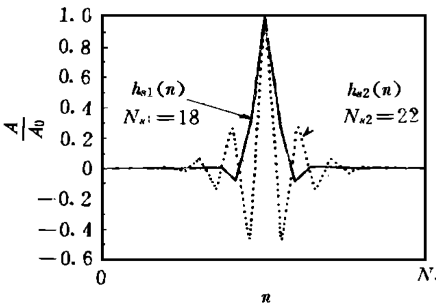

Fig. 5 The decomposed $h_{s_1}(n)$ and $h_{s_2}(n)$

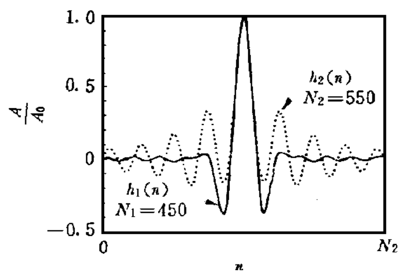

Fig. 6 Interpolated $h_1(n)$ and $h_2(n)$

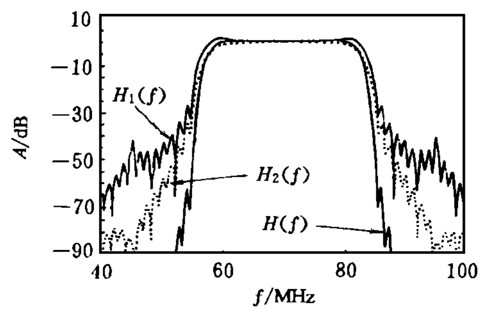

Fig. 7 The resulted $H_1(f)$ and $H_2(f)$

4 Conclusions and Discusses

In MSC SAW filter, the desired frequency response must be decomposed into two weight functions for the two IDTs. It is very difficult to apply the ordinary method referred to as zero-separating when thousands of taps are required. In this paper, original weighting function with many small weights is first resampled at an internal of M to decrease the number of variable to be optimized and to use aero separating technology. The resampled two functions are approximated by interpolation method, and one of them with less order is optimized furtherly by Remez optimal algorithm previously used, and the other is supposed to be fixed. This novel optimization has another important role: to secure at the same time a sufficient approximation accuracy. The simulation results of a SAW MSC Blackman spectrum shaping filter for high speed and bandwidth efficient digital communication system indicate that this method is very simple and effective. And the development of an accurate model for the second-order effects of the MSC is to be studied furtherly[8,9].

References

1 Morgan D P. Surface wave devices for signal processing. New York: Elsevier, 1985
2 Rupple C C W, Sachs A A. A review of optimization algorithms for the design of SAW transducers. IEEE Ultrason Symp, 1991: 1- 11
3 Reddy A R. Design of SAW bandpass filters using new window functions with applications to SAW filters. IEEE Tran 1988, UFFC-35(1): 50- 56
4 Rukhlenko A S. Optimal and suboptimal design of SAW bandpass filters using the Remez exchange algorithm. IEEE Tras, 1993, UFFC-40(5): 453- 458
5 Peach R. The use of linear programming for the design of SAW and filterbanks. IEEE Trans, 1994, UFFC-41(4): 532- 541
6 Kodama T. Sampling theorem application to SAW diffraction simulation. Electron Lett, 1980, (4): 462- 465
7 Shida S. Fushimi S. FIR low-pass filter design using parametric filter technology. IEEE Trans Circuits Sys, 1984(5): 447- 452
8 Abbott B P, Malocha D C. Closed form solutions for multistrip coupler operation including the effects of electrode resistivity. IEEE Ultrason Symp, 1985: 7- 10
9 Ruppel C, Falkenau E E, Stocker H R. Compensation algorithm for SAW second order effects in MSC filner. IEEE Ultrason Symp, 1985: 7- 10

一种新颖的 SAW带通滤波器设计方法

杨朝斌* 徐继麟 黄香馥
(电子科技大学电子工程学院 成都 610054)

【摘要】 提出了一种新的声表面波(SAW)滤波器设计方法。先对设计目标函数采样减少变量数,用 Z变换零点分开并通过内插重构得到两组加权初值。假定其中一个固定而对另一加权系数利用标准的雷米兹交换算法进行优化。对 Blackman谱成形成滤波器的模拟设计表明,该方法计算精确方便,简明实用。

关 键 词 声表面波; 采样与内插; 零点分开; 优化; 雷米兹交换算法
中图分类号 TN713.5

编辑 叶 红

⒇ 1996年 11月 12日收稿
* 男 31岁 博士生