Iterative WLS Design of SAW Bandpass Filters

Alexander S. Rukhlenko, *Senior Member, IEEE*

***Abstract*—It has been demonstrated by several authors that the well-known weighted least squares (WLS) approximation can be equiripple if a suitable weighting function is applied. In the present paper, the WLS algorithm is generalized to SAW filter synthesis with prescribed magnitude and phase specifications. Several weighting techniques producing quasi-equiripple designs are presented. The frequency sampling technique is used for SAW filter frequency response approximation to reduce the number of optimized variables. The WLS algorithm rapidly converges both for linear and nonlinear phase SAW filters. Typically, no more than 5–10 iterations are required to obtain the WLS solution to accuracy better than 0.5–1 dB in the stopband when compared with the optimum Chebyshev approximation. Moreover, it is shown that the WLS technique can be effectively applied for second-order effect compensation.**

I. Introduction

The state-of-the-art design of surface acoustic wave (SAW) filters which are optimum in the Chebyshev (minimax) sense requires sophisticated optimization tools such as the Remez exchange algorithm or linear programming [1], [2]. However, some of the recent finite-impulse response (FIR) optimization methods [3]–[6] have not been configured for SAW filters despite their superiority over traditional design approaches. The present paper describes an application of the innovative weighted least squares (WLS) technique [3]–[6] to designing optimum SAW filters.

It is well known that for a given least squares weighting function, the optimum WLS solution can be derived analytically. The WLS software can be implemented in compact computer codes and it is included in many standard software packages. Unfortunately, the standard WLS solution suffers from the drawback that the approximation accuracy deteriorates near the band edges.

It has been reported by several authors [3]–[6] that the WLS technique produces a quasi-equiripple solution if a suitable nonuniform least squares weighting function is applied. Though there is no proof of the Chebyshev optimality for equiripple WLS designs, it has been observed that WLS solutions closely correspond to Chebyshev optimum ones.

The major problem is that the least squares weighting function producing an equiripple design cannot be specified analytically. Therefore, iterative reweighting schemes need to be applied, with the weighting function updated at each iteration using the results of the previous iteration.

In the present paper, the WLS method is generalized to SAW filters. Several reweighting schemes are discussed. The real-valued WLS approach is extended to complex-valued (nonlinear phase) SAW filter design that allows compensation for second-order effects (e.g., SAW filter frequency response distortion due to electrical source/load effects).

II. Chebyshev (Minimax) Approximation

Consider a problem of the best approximation of the prescribed complex-valued function (target function) $D(\omega)$ by the approximating function (frequency response)

$$F(\omega) = C(\omega) + jS(\omega) = \sum_{k=0}^{N-1} a_k C_k(\omega) + j \sum_{k=0}^{N-1} b_k S_k(\omega), \quad (1)$$

where $C_k(\omega)$ and $S_k(\omega)$ are the real and imaginary part basis functions, and a_k and b_k are the approximation coefficients for the real and imaginary parts, respectively. At any frequency ω, the approximation accuracy is characterized by the weighted Chebyshev error function

$$E(\omega) = W(\omega)[F(\omega) - D(\omega)], \quad (2)$$

where $W(\omega) > 0$ is the Chebyshev weighting function. The Chebyshev (minimax) approximation is the best fit to the complex-valued function $D(\omega)$ to minimize an absolute error:

$$\delta = \min_{\mathbf{a},\mathbf{b}} \|E(\omega)\| = \min_{\mathbf{a},\mathbf{b}} \{\max_{\omega\in\Omega} |E(\omega)|\} \quad (3)$$

over a set of the coefficients $\mathbf{a} = [a_k]$ and $\mathbf{b} = [b_k]$ within the approximation interval Ω. In the particular case of the real-valued functions (linear phase design), the key property of the optimum solution is the equiripple sign-alternated behavior of the Chebyshev error function given by the Chebyshev alternation theorem if the basis functions $C_k(\omega)$ or $S_k(\omega)$ meet Haar's condition [1].

III. Complex WLS (CWLS) Fit

Given the complex-valued approximating function $F(\omega)$ and the desired (target) function $D(\omega)$, the WLS error on the discrete frequency grid $\omega_i \in \Omega$, $i = 0, M - 1$ is

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The author is with Infineon Technologies, Munich, Germany (e-mail: Alexander.Rukhlenko@Infineon.com).

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$$\begin{aligned}\varepsilon &= \sum_{i=0}^{M-1} w_i \left|F_i - D_i\right|^2 \\ &= \sum_{i=0}^{M-1} w_i \left|\sum_{k=0}^{N-1} a_k C_{ik} + j \sum_{k=0}^{N-1} b_k S_{ik} - D_i\right|^2,\end{aligned} \tag{4}$$

where $F_i = F(\omega_i)$, $D_i = D(\omega_i)$, $C_{ik} = C_k(\omega_i)$, $S_{ik} = S_k(\omega_i)$, and w_i are the WLS weights (not to be confused with the Chebyshev weights which are capitalized), or, in matrix form,

$$\varepsilon = (\mathbf{FA} - \mathbf{D})^* \mathbf{W} (\mathbf{FA} - \mathbf{D}), \tag{5}$$

where

$$\mathbf{F} = \left[\mathbf{C}\ j\mathbf{S}\right], \quad \mathbf{A} = \left[\mathbf{a}\ \mathbf{b}\right]^T,$$

$$\mathbf{C} = \begin{bmatrix} C_{0,0} & C_{0,1} & \cdots & C_{0,N-1} \\ C_{1,0} & C_{1,1} & \cdots & C_{1,N-1} \\ \vdots & & \ddots & \vdots \\ C_{M-1,0} & C_{M-1,1} & & C_{M-1,N-1} \end{bmatrix},$$

$$\mathbf{S} = \begin{bmatrix} S_{0,0} & S_{0,1} & \cdots & S_{0,N-1} \\ S_{1,0} & S_{1,1} & \cdots & S_{1,N-1} \\ \vdots & & \ddots & \vdots \\ S_{M-1,0} & S_{M-1,1} & & S_{M-1,N-1} \end{bmatrix},$$

$$\mathbf{a} = \left[a_0\ a_1\ \cdots\ a_{N-1}\right], \quad \mathbf{b} = \left[b_0\ b_1\ \cdots\ b_{N-1}\right],$$

$$\mathbf{D} = \left[D_0\ D_1\ \cdots\ D_{M-1}\right]^T, \quad \mathbf{W} = diag\left[w_0\ w_1\ \cdots\ w_{M-1}\right];$$

the superscript ^{T} stands for the matrix transpose and $*$ denotes the Hermitian conjugation (complex conjugation with transpose).

For the functions $D(\omega)$ and $F(\omega)$ specified on the discrete frequency grid $\omega_i \in \Omega$, $i = 0,\ M-1$, the complex WLS (CWLS) problem is reduced to minimizing the absolute CWLS error (5).

The closed-form CWLS solution can be found by differentiating (5) with respect to the vector of the coefficients $\mathbf{A}$ and equating the derivative to zero:

$$\frac{d\varepsilon}{d\mathbf{A}} = \mathrm{Re}\left\{\mathbf{F}^* \mathbf{W}(\mathbf{FA} - \mathbf{D})\right\} = 0. \tag{6}$$

The solution of (6) is given by

$$\mathbf{A} = \mathrm{Re}\left\{\mathbf{F}^*\mathbf{WF}\right\}^{-1} \mathrm{Re}\left\{\mathbf{F}^*\mathbf{WD}\right\}. \tag{7}$$

By splitting (7) into the real and imaginary parts and substituting $\mathbf{A} = [\mathbf{a}\,\mathbf{b}]^T$, we obtain

$$\begin{aligned}\mathbf{a}^T &= \left(\mathbf{C^T W C}\right)^{-1} \mathbf{C^T W}\mathrm{Re}\{\mathbf{D}\}, \\ \mathbf{b}^T &= \left(\mathbf{S^T W S}\right)^{-1} \mathbf{S^T W}\mathrm{Im}\{\mathbf{D}\}.\end{aligned} \tag{8}$$

Therefore, the CWLS problem is reduced to two separate real-valued WLS problems for the real and imaginary parts of $D(\omega)$ weighted by the same WLS weighting function $w(\omega)$. In other words, one needs to solve the real WLS problem twice (for the real and imaginary parts) to find the CWLS solution (8). In the particular case of the linear phase design, the WLS problem is reduced to either the real or imaginary part approximation (8).

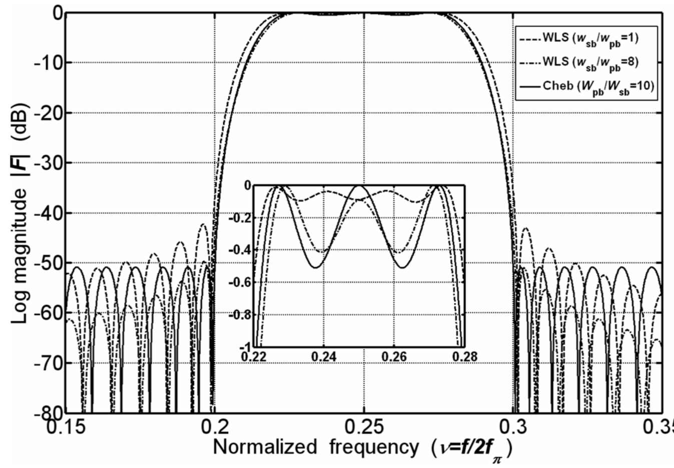

Fig. 1. Comparison of the Chebyshev ($W_{PB}/W_{SB} = 10$) and WLS ($w_{SB}/w_{PB} = 1$ and $w_{SB}/w_{PB} = 8$) approximations.

The complexity of (8) is defined by the matrix inversion and matrix multiplications. Linear algebra techniques and commercial software can be applied to find each of the real WLS solutions (8).

IV. Chebyshev and WLS Solutions

For comparison, the optimum Chebyshev approximation for the weights ratio in the passband and stopband $W_{PB}/W_{SB} = 10$ and the WLS fit for two different WLS ratios in the stopband and passband $w_{SB}/w_{PB} = 1$ and $w_{SB}/w_{PB} = 8$ are shown in Fig. 1, where $f_\pi = v/2p$ is the synchronous frequency, v is the SAW velocity, and p is the transducer period. Please note that the uppercase weights W are related to the Chebyshev approximation, whereas the lowercase weights w are attributed to the WLS fit.

As can be observed, the WLS weight ratio $w_{SB}/w_{PB} = 1$ (uniform weighting) provides very small ripple in the passband (less than 0.1 dB) and a wider passband width at the expense of poor sidelobe rejection (about -42 dB) near the stopband edges. Redistribution of the WLS weights in favor of the stopband weights ($w_{SB}/w_{PB} = 8$) improves the stopband rejection to about -50 dB, whereas the passband ripple increases to 0.4 dB. The Chebyshev approximation ($W_{PB}/W_{SB} = 10$) is superior to the WLS $w_{SB}/w_{PB} = 8$ near the band edges, whereas the WLS fit provides better accuracy elsewhere.

This example demonstrates that, by the appropriate guess for the WLS passband/stopband weights ratio, the difference between WLS and Chebyshev solutions can be reduced. Moreover, by adjusting individually the WLS weights w_i at each frequency point, in particular, by increasing the weights near the band edges, the overall approximation error can be minimized at the expense of the "over-approximated" regions.

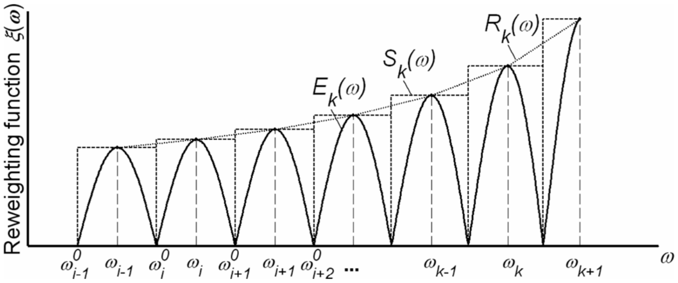

Fig. 2. Different reweighting functions: $E_k(\omega)$ is the Chebyshev error function, $S_k(\omega)$ is the step-wise error approximation, and $R_k(\omega)$ is the error function envelope.

V. Iterative Reweighting Schemes

A. Statement of the Problem

There is no analytical method for deriving the WLS weighting function $w(\omega)$, which would produce a minimax equiripple design. Therefore, an iterative approach needs to be applied, with the function $w(\omega)$ redefined at each k-th iteration in the following multiplicative form [4]:

$$w_{k+1}(\omega) = \xi_k^\theta(\omega) w_k(\omega), \tag{9}$$

where $\xi_k(\omega) > 0$ is the weight correction (update) function and θ is the empirical convergence factor. The normalized function $\xi_k(\omega)$ must be updated at each iteration so that $\xi_k(\omega) > 1$ at frequencies where the error $|E_k(\omega)|$ needs to be reduced. At the next iteration, the error $|E_{k+1}(\omega)|$ will decrease in this region at the expense of increasing in other regions where the WLS weights are changed in the opposite direction or remain unchanged. The effect of reweighting can be emphasized or damped from iteration to iteration by selection of the exponential convergence factor θ. There are several reweighting schemes leading to the quasi-equiripple Chebyshev solution [4]–[6].

B. Lawson's Algorithm

The weight correction function is taken to be proportional to the weighted absolute Chebyshev error function at the k-th iteration [3] (Fig. 2):

$$\xi_k(\omega) = |E_k(\omega)| = W(\omega)\,|F_k(\omega) - D(\omega)|\,. \tag{10}$$

Therefore, Lawson's algorithm requires calculation of just the error function value $E_k(\omega)$ at each frequency. Due to its simplicity, Lawson's algorithm is characterized by relatively slow convergence [4].

It is believed that Lawson's algorithm might fail at points where $E_k(\omega) = 0$, since these points give zero WLS weights $w_{k+1}(\omega) = 0$ at all further iterations that may cause divergence [4], [5]. However, the author's design experience has not confirmed this concern, and Lawson's algorithm demonstrates stable convergence in all practical designs of SAW bandpass filters. This fact can be justified by observing that WLS weights degenerate to small values in the frequency regions where the Chebyshev error function does not tend to increase, whereas at the critical extremal points the weights are well controlled by keeping their values adjusted during WLS iterations.

C. Step-Wise Error Approximation

More sophisticated reweighting schemes based on the search of error function extremal frequencies have been suggested in [4], [5]. They provide faster convergence if compared with Lawson's algorithm. In particular, the error function can be step-wise approximated as [5]

$$\xi_k(\omega) = S_k(\omega) = \max_{\omega_i^0 \le \omega \le \omega_{i+1}^0} \{|E_k(\omega)|\} = |E_k(\omega_i)|\,, \tag{11}$$

where ω_i is the position of the local maximum, and ω_i^0 and ω_{i+1}^0 are the positions of the local minima (valley frequencies) of the absolute Chebyshev error function $|E_k(\omega)|$ (Fig. 2). Within one sidelobe of the error function $|E_k(\omega)|$ at the frequencies $\omega_i^0 \le \omega \le \omega_{i+1}^0$, the update function $\xi_k(\omega)$ is approximated by a step of the magnitude $|E_k(\omega_i)|$, where ω_i is the frequency of the local maximum at this interval. At each iteration, the update function $\xi_k(\omega)$ is given by a set of steps having different widths and heights. In addition to calculation of the Chebyshev error function $|E_k(\omega)|$, this algorithm requires a search for extremal frequencies (both local minima and maxima).

D. Error Envelope Approximation

The best convergence is obtained using a reweighting scheme based on the following approximation [4]:

$$\xi_k(\omega) = R_k(\omega), \tag{12}$$

where $R_k(\omega)$ is the envelope of the absolute Chebyshev error function $|E_k(\omega)|$ reconstructed as a set of line segments connecting the extremal frequencies in a particular frequency subband (Fig. 2). Contrary to the aforementioned reweighting techniques, this method results in smooth and monotone WLS weighting function segments demonstrating the best convergence rate.

It is worth noting that the two last weighting schemes involve searching for extremal frequencies, with special care taken in the extrema interpretation at the band edges [4]. However, the simplest algorithms for extrema search can be applied, since the iterative WLS method does not require high accuracy in the location and evaluation of extrema.

The ultimate optimum WLS solution does not depend on the reweighting scheme applied, whereas the convergence speed and hence the number of iterations are sensitive to reweighting.

E. Convergence

The number of WLS iterations needed to closely approximate the optimum Chebyshev solution depends on

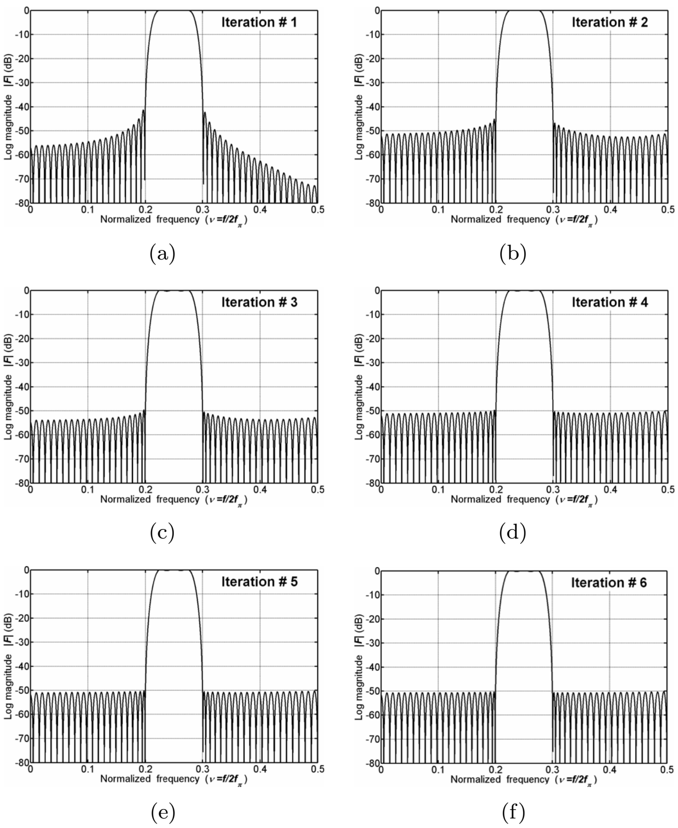

Fig. 3. Iterative WLS convergence in the stopband ($\theta = 0.6$).

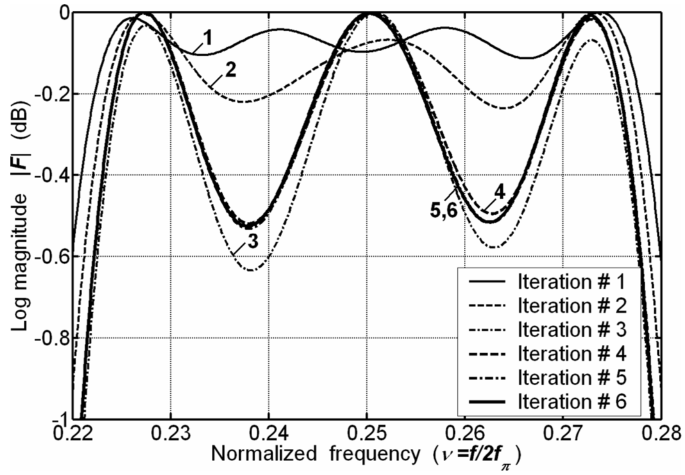

Fig. 4. Iterative WLS convergence in the passband ($\theta = 0.6$).

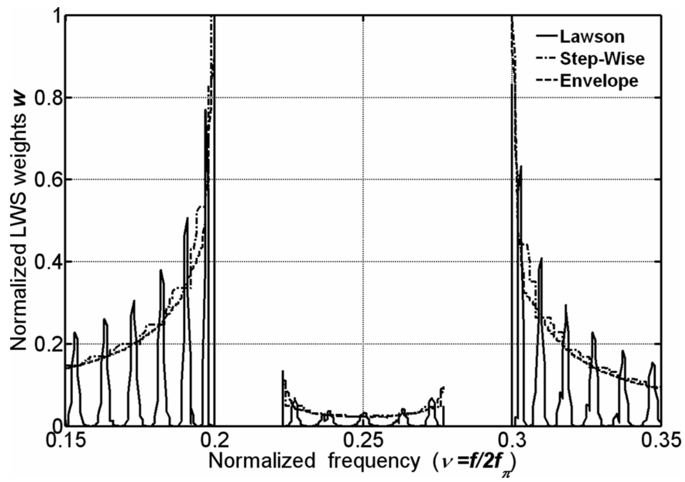

Fig. 5. Normalized WLS weights for different reweighting schemes.

the exponential convergence factor θ. Again, there is no analytical technique for obtaining θ that provides the best algorithm convergence. The smaller θ values result in underweighting and the larger values cause overweighting that increases the required number of iterations in both cases. From our experience, good convergence has been observed for $\theta \approx 0.5$–0.75.

The iterative WLS algorithms can be terminated after the prescribed number of iterations or after checking the equiripple behavior to a required accuracy. Typically, the WLS algorithm converges within 5–10 iterations to accuracy better than 0.5–1 dB in the stopband when compared with the optimum Chebyshev approximation (Fig. 3). For example, it has taken 15 Lawson's iterations to obtain an equiripple Chebyshev approximation shown in Fig. 1 with accuracy better than 0.5 dB. About half the number of iterations based on the step-wise or envelope error function approximations are needed to achieve the same accuracy. However, the latter reweighting techniques are more sophisticated for programming, and Lawson's algorithm can be suggested as a good choice to get started in the WLS implementation for designing SAW bandpass filters.

The WLS algorithm convergence is illustrated in Fig. 3 (stopband) and Fig. 4 (passband), where the magnitude response at the first six iterations is shown. The most effective reweighting scheme based on the error function envelope approximation has been applied. The uniform WLS weights $w_i = 1$ have been used as the initial guess in both the passband and the stopband, and the value of the convergence factor has been chosen as $\theta = 0.6$. As can be seen, practically no further improvement in the approximation accuracy is observed after the fifth iteration, for this example.

The ultimate normalized WLS weights in the passband and stopband are shown in Fig. 5 for three different reweighting schemes that give the optimum Chebyshev solution in Fig. 1. As anticipated, Lawson's algorithm gives the degenerate weights in some frequencies regions without sacrificing the overall approximation accuracy. Despite the considerable differences in the WLS weights, all these reweighting schemes converge to the same optimum Chebyshev solution.

The WLS computational time is comparable to that of linear programming [6], whereas the latter is much more sophisticated in programming and cumbersome for practical use. However, the WLS computational time can be reduced by the appropriate choice of the number of iterations to achieve an acceptable accuracy that makes it superior over the linear programming from the computational point of view.

The computational time (number of iterations) can be further reduced if the initial guess for the WLS weights is improved. Empirical relations for the best initial stopband/passband WLS weights ratio can be deduced based on the typical passband ripple and stopband rejection specifications. However, for simplicity, we have applied uniform WLS weighting $w_i = 1$ as the initial guess in all our designs, since it requires just a few additional WLS iterations.

VI. SAW Filter Design

A. CWLS Algorithm Modifications

The iterative WLS technique can be applied to SAW filter synthesis with linear and nonlinear phase responses. In the general case, the input SAW transducer response and the element factor should be accounted for in the passband by modifying the weighting function $W(\omega)$ and target function $D(\omega)$ in the Chebyshev approximation problem [7].

To reduce the WLS problem size (and hence the computational time), the frequency sampling technique [1], [2] can be applied:

$$C(\varphi) = \frac{1}{N} \sum_i C(\varphi_i) \left\{ \operatorname{sinc}(\varphi - \varphi_i) + \operatorname{sinc}(\varphi + \varphi_i) \right\},$$

$$S(\varphi) = \frac{1}{N} \sum_i S(\varphi_i) \left\{ \operatorname{sinc}(\varphi - \varphi_i) - \operatorname{sinc}(\varphi + \varphi_i) \right\}, \quad (13)$$

$$\operatorname{sinc}(\varphi) = \frac{\sin N\varphi/2}{\sin \varphi/2}, \quad \varphi = \beta p = \pi f / f_\pi, \quad (14)$$

where $\varphi_i = i\Delta\varphi$ (i is integer) are the sampling points, $\Delta\varphi = 2\pi/N$ is the sampling (discretization) interval, $\beta = \omega/v$ is the SAW wave number, and N is the number of transducer fingers. The second term in (13) accounts for the contribution to the frequency response of the "mirror" band (with respect to the synchronous frequency f_π) that is in-phase (+ sign) with the baseband response for the real part $C(\varphi)$ and in anti-phase (− sign) for the imaginary part $S(\varphi)$. For bandpass SAW filters, most of the frequency samples at the frequency points φ_i can be set to zero values without significantly sacrificing the approximation accuracy.

B. Second-Order Effect Compensation

The iterative CWLS algorithm can be applied to compensation of the second-order effects, in particular, electrical source/load, or circuit, effects. The SAW filter transfer function is given by

$$S_{12} = \frac{Y_{12} Y_L}{Y_{12} Y_{21} - (Y_{11} + Y_0)(Y_{22} + Y_L)} \approx \frac{-Y_{12} Y_L}{(Y_{11} + Y_0)(Y_{22} + Y_L)}, \quad (15)$$

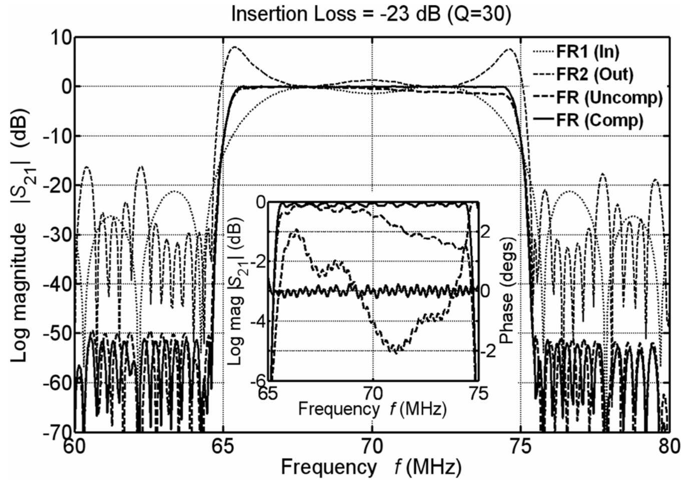

Fig. 6. SAW filter design example with the electrical circuit effects compensation.

where Y_{ik}, $i, k = 1, 2$ are Y-parameters, and $Y_{0,L}$ are the source and load admittances, respectively. Matching circuits can be also accounted for in (15).

In the quasi-static approximation, it is the SAW filter trans-admittance $Y_{12}(\omega) = Y_{21}(\omega)$ that depends on the linear function of the transducer tap weights or frequency samples. As S_{12} depends on all Y-parameters, the synthesis problem is nonlinear with respect to the optimized variables, and an iterative compensation procedure is needed, with the target function predistorted in correspondence with the simulation results from the previous iteration.

Therefore, we are approaching a "moving target" using as reference the "ideal" transfer function $S_{12}^{(0)}$ at each compensation iteration, so that

$$D^{(i)} = D^{(i-1)} \frac{S_{12}^{(0)}}{S_{12}^{(i-1)}}, \quad D^{(0)} = D, \quad S_{12}^{(0)} = Y_{12}^{(0)} / Y_0. \quad (16)$$

At each compensation iteration, the CWLS problem is constructed and solved by using (8), (9).

C. SAW Filter Design Example

The iterative CWLS method has been applied for compensation of the electrical circuit effects in 50-Ω system. The results of compensation are presented in Fig. 6, where the frequency responses FR_1 and FR_2 of the input withdrawal-weighted transducer and output apodized SAW transducers, respectively, and the SAW filter frequency response before and after compensation are shown.

Two series inductors $L_1 = 240$ nH and $L_2 = 160$ nH ($Q = 30$) have been used for matching, with their values remaining fixed during the compensation procedure. The substrate material is $LiTaO_3$, and the acoustic aperture is $W = 2.2$ mm (67 λ). The number of transducer fingers is $N_1 = 100$ (withdrawal-weighted) and $N_2 = 700$ (apodized), respectively. The number of the optimized frequency samples is 50. The discrete frequency grid comprises 500 points in the frequency range 60–80 MHz.

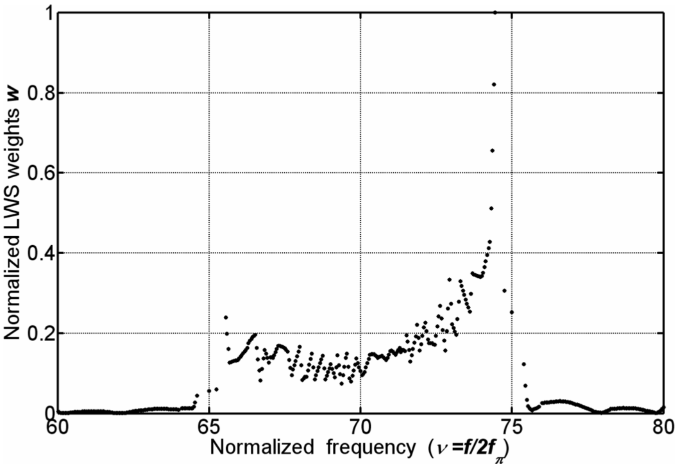

Fig. 7. WLS weights of a SAW filter after compensation for the electrical circuit effects.

The WLS weights after compensation are shown in Fig. 7. As can be seen from the comparison of Fig. 5 and Fig. 7, the WLS weights in the case of a SAW filter are distributed in a more complicated way (due to accounting for the input transducer response and compensation for the circuit effects). Therefore, more CWLS iterations are required if compared with the regular bandpass design shown in Figs. 3–5, where, for simplicity, the response of the input SAW transducer has been neglected.

The number of iterations for compensation of the electrical circuit effects was five, while the number of the WLS iterations at each compensation step was ten. The error function envelope technique was applied for reweighting. There is almost no residual magnitude and phase distortion in the passband after the compensation.

VII. Conclusions

An iterative WLS procedure has been developed for SAW filter design with arbitrary magnitude and phase specifications. Due to its simplicity and straightforward implementation, the proposed WLS approach to the Chebyshev approximation is a powerful alternative to linear programming and the Remez exchange algorithm. The WLS technique is computationally efficient and easy for programming. Three different reweighting schemes resulting in the same Chebyshev solution have been discussed, with the only difference being the number of WLS iterations required to achieve an acceptable accuracy. Typically, the Chebyshev approximation can be achieved in just 5–10 WLS iterations to accuracy better than 0.5–1 dB in the SAW filter stopband if compared with the rigorous Chebyshev problem solution. It has been demonstrated that the iterative WLS method can be effectively applied for compensation of second-order effects. The SAW filter design examples presented confirm the advantages of the iterative WLS design technique.

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Alexander S. Rukhlenko (M'92) was born in the Republic of Belarus on June 19, 1956. He received the M.Sc. degree in radiophysics and electronics (with distinction) from the Byelorussian State University, Minsk, in 1978 and the Ph.D. degree in acoustoelectronics from the Minsk Radio Engineering Institute, in 1989.

From 1980 to 1993, he was with the Laboratory of Acousto- and Optoelectronics at the Minsk Radio Engineering Institute where he was engaged in the research and development of SAW devices and their applications to signal processing. In 1993, he joined the Semiconductor Physics Department at the Belarusian State University to continue his SAW research. He worked as Principal Researcher, heading the SAW devices group where he managed multiple SAW research projects. In 1995, he was a winner of the SAW filter design contest organized by the American company SAWTEK, Inc., in the former USSR. In 1995–2001, he combined his research activities with consulting SAW companies in the USA, Germany, South Korea, China, etc., in the field of computer-aided design of SAW devices. He worked as Principal Engineer at the Swiss company TEMEX SAW, Neuchâtel, Switzerland, where he was responsible for design and manufacture of IF SAW filters in 2001–2002. He was with the Institute of Microtechnology, SAMLAB (Neuchâtel) in 2004–2005. Currently, he works as Senior Staff Engineer at Infineon (Germany) where he is engaged in the computer-aided design and simulation of FBAR filters.

Dr. Rukhlenko is an author of the comprehensive SAW software and he has more than 50 publications on acoustoelectronics. His present research interests comprise modeling, simulation, and computer-aided design of BAW and SAW devices.