Non-Iterative Finite Impulse Response Design Techniques

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NON-ITERATIVE FINITE IMPULSE RESPONSE DESIGN TECHNIQUES

BY

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ABSTRACT

A general, non-iterative design technique for low shapefactor, transversal filters is presented. This design approach uses two cosine series to specify appropriate eigenfunctions. An infinite set of such eigenfunctions are defined and the method for choosing the coefficients is discussed.

The total filter response is specified as the product of two individual frequency responses. The impulse response of each is then determined by applying the superposition of appropriate eigenfunctions. The criteria for choosing the appropriate eigenfunctions is discussed.

A synthesis procedure for designing surface acoustic wave filters is presented. The effects of truncating the impulse response are also explored. A design example is shown for a filter with 10 percent fractional bandwidth and a shapefactor of 1.15.
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CHAPTER I

INTRODUCTION

Many approaches have been presented for designing finite impulse response filters. Parks and McClennan describe a commonly used technique which utilizes the Remez exchange algorithm to obtain a near optimum finite impulse response [1]. This is an iterative approach which requires convergence to the optimum response. Other approaches have also been investigated which enhance the optimization process or use alternative linear programming techniques [2,3].

An alternate approach is to apply superposition techniques to finite impulse response (FIR) filter design. DeVries described a building block approach which utilized the superposition principal [4]. Vasile presented a filter synthesis technique which described an eigenfunction for approximating a triangle in the frequency domain [5].

In some technologies, such as Surface Acoustic Wave Devices (SAW), the frequency response of a filter is implemented using two transducers each with a separate frequency response [6]. Previous approaches have normally specified only a single frequency response which then must be split between the two transducers. One approach for dividing the frequency response between two transducers utilizes a zero extraction technique [7]. This technique splits the zeros of the transfer response into two sets such that the zeros are exclusive.
Then appropriate polynomial expressions are determined which describe the desired frequency responses. The technique is very useful. However, time consuming complex calculations and proper choice of zero separation are required.

The superposition approach to designing FIR filters requires the specification of appropriate basis functions. C.L. Dolph introduced two sets of functions for improving the beam pattern of a linear array [8]. The coefficients for the terms of the functions are determined using complex power series.

The objective of this work is to develop a new, alternate approach to designing FIR filters which is non-iterative and will reduce the required design time for SAW filters. The resulting designs are required to meet all passband specifications within 5% and meet or exceed all transition and stop band specifications.
CHAPTER II
DEFINITION OF EIGENFUNCTIONS

Characteristics of Eigenfunctions

In designing FIR filters, there are certain advantages to defining a desired frequency response as the product of two separate responses. Two functions with the following properties are being proposed to specify the responses:

1. Each function is finite in time.
2. The convolution of the time responses, or conversely the product of their frequency transfer functions, produce the desired overall transfer function of the filter.
3. In the passband region, the two transfer functions have maximum overlap.
4. Half the zeros in the stop band of the overall frequency transfer function exist in each of the eigenfunctions.
5. The zeros of each function are equidistant from each other, and the peaks of one function’s sidelobes fall at the nulls of the other.
6. The functions are well behaved and can be described analytically.
7. The zeros of the components of the functions coincide.
8. The IR of the transducers are implementable.
9. As the number of terms, N, approaches infinity, the time functions approach an impulse and the frequency responses approach unity.

Definition of Basis Functions

The following general set of functions is presented for consideration as eigen pairs

\[
\begin{align*}
   h_{N1}(t) &= \sum_{n=0}^{N-1} a_n \cos \left( \frac{2n\pi t}{T} \right) \\
   h_{N2}(t) &= \sum_{n=0}^{N-1} a_{n+1} \cos \left( \frac{(2n-1)\pi t}{T} \right)
\end{align*}
\]

The first subscript denotes the number of terms used in the Nth cosine series of the eigenfunction. The second subscript denotes the even or odd eigenfunction determined by the terms in the arguments of the cosine given by 2n and 2n-1 respectively.

As N approaches infinity, there are an infinite number of eigenfunctions possible. In addition, a criteria must be imposed on the coefficients to obtain a function having the required properties.

Choosing Coefficients

Figures 1a and 1b are plots of the components of the three term eigenfunction series. Notice that in each series the zeros of the components coincide and that the peak of each component occurs at a null of its adjacent neighbor. To obtain the properties desired, the
The following criteria is applied to the series to select the coefficients:

1. The sum of the coefficients is unity.
2. The first N-1 natural sidelobe peaks in the frequency response are set to zero.

These conditions describe a set of equations that can be solved for the coefficients, \( a_{nj} \). These equations can be written in closed form as

\[
\begin{bmatrix}
1 & \cdots & 1 \\
\lambda_{10} & \cdots & \lambda_{1}(N-1) \\
\vdots & \ddots & \vdots \\
\lambda_{(N-1)0} & \cdots & \lambda_{(N-1)(N-1)}
\end{bmatrix}
\begin{bmatrix}
a_{0j} \\
a_{1j} \\
\vdots \\
a_{(N-1)j}
\end{bmatrix}
= 
\begin{bmatrix}
1 \\
0 \\
\vdots \\
0
\end{bmatrix}
\]

where \( \lambda_{mn} \) is the value of the \( m^{th} \) sidelobe peak of the \( n^{th} \) component of the appropriate series.

To determine the values of \( \lambda_{mn} \) for a particular series, the frequency response of the series must be examined. \( H_{N1}(f,T) \) can be written in general form as

\[
H_{N1}(f,T) = \sum_{n=0}^{N-1} a_{n1}T[\text{sinc}(f-nT) + \text{sinc}(f+nT)]
\]

Let \( f_m(N,T) \) designate the frequency of the \( m^{th} \) sidelobe peak of a \( N \) term series. The first sidelobe peak of any term of \( H_{N1}(f,T) \) occurs at \( \frac{3\pi}{2} \) from the center frequency of the component. This frequency for the \( N^{th} \) term can be written in the form
Figure 1. Normalized terms (a) $h_{31}$, (b) $h_{32}$
\[ f_m(N, T) = \frac{2m + 2N - 3}{2T} \]

At this frequency, the value of the sine function will be \( \pm 1 \). Each term has two frequency components. These components for \( H_{N1} \) are of the same sign. Keeping all these items in mind and with some manipulation, \( \lambda_{mn} \) can be written in the form

\[ \lambda_{mn} = (-1)^{m+N+n-1} \left[ \frac{1}{k-2n-1} + \frac{1}{k+2n-3} \right] \]

\[ n=1, 2, 3, \ldots, N; \ m=2, 3, 4, \ldots, N; \ k=2m+2N \]

\( \lambda_{mn} \) for \( H_{N2} \) can be derived in a similar manner except that the components of the frequency response for each term are of opposite sign.

\( \lambda_{mn} \) for \( H_{N2} \) can be written as

\[ \lambda_{mn} = (-1)^{m+N+n-1} \left[ \frac{1}{k-2n-1} - \frac{1}{k+2n-3} \right] \]

With \( \lambda_{mn} \) determined, the coefficients, \( a_{nj} \), can be solved for using Gaussian elimination.

**Choosing the number of terms**

The choice of the number of terms, \( N \), in each function is dictated by the out-of-band rejection. The larger the number of terms, the greater the sidelobe rejection. As the number of terms in the eigenfunctions increase, the impulse response becomes more smooth, i.e., the function approaches zero at a slower rate at \( |t|=\frac{\pi}{2} \). This is demonstrated in Figure 2. Notice that the increased smoothness concentrates the energy in the IR toward the center which increases the
time-bandwidth product. Due to implementation constraints, the number of terms, \( N \), will be restricted to a maximum of three.

![Figure 2. First three time eigenfunction pairs](image)

**Eigenfunctions**

The eigen time functions under consideration are

**One term functions**

\[
h_{11}(t) = 1 \quad |t| < \frac{T}{2}
\]

\[
h_{12}(t) = \cos \frac{\pi t}{T}
\]

**Two term functions**

\[
h_{21}(t) = 0.537 + 0.453 \cos \frac{2\pi t}{T} \quad |t| < \frac{T}{2}
\]

\[
h_{22}(t) = 0.796 \cos \frac{\pi t}{T} + 0.204 \cos \frac{3\pi t}{T}
\]
Three term functions

\[ h_{31}(t) = 0.427 + 0.497\cos\frac{2\pi t}{T} + 0.0768\cos\frac{4\pi t}{T} \quad \text{(|t| < \frac{T}{2})} \]  

\[ h_{32}(t) = 0.690\cos\frac{\pi t}{T} + 0.277\cos\frac{3\pi t}{T} + 0.0324\cos\frac{5\pi t}{T} \]  

Note that the IR length, T, is the same for both transducers. This result would be expected for a near optimum design.

The baseband eigen frequency functions are given by

\[ H_{11}(f) = T\text{sinc}(fT) \]
\[ H_{12}(f) = \frac{T}{2}[\text{sinc}(f-\frac{1}{2T})T + \text{sinc}(f+\frac{1}{2T})T] \]
\[ H_{21}(f) = 0.537T\text{sinc}(fT) + 0.226T[\text{sinc}(f-\frac{1}{T})T + \text{sinc}(f+\frac{1}{T})T] \]
\[ H_{22}(f) = 0.398T[\text{sinc}(f-\frac{1}{2T})T + \text{sinc}(f+\frac{1}{2T})T] \\
+ 0.102T[\text{sinc}(f-\frac{3}{2T})T + \text{sinc}(f+\frac{3}{2T})T] \]
\[ H_{31}(f) = 0.427T\text{sinc}(fT) + 0.249T[\text{sinc}(f-\frac{1}{T})T + \text{sinc}(f+\frac{1}{T})T] \\
+ 0.0384T[\text{sinc}(f-\frac{2}{T})T + \text{sinc}(f+\frac{2}{T})T] \]
\[ H_{32}(f) = 0.345T[\text{sinc}(f-\frac{1}{2T})T + \text{sinc}(f+\frac{1}{2T})T] \\
+ 0.139T[\text{sinc}(f-\frac{3}{2T})T + \text{sinc}(f+\frac{3}{2T})T] \\
+ 0.0162T[\text{sinc}(f-\frac{5}{2T})T + \text{sinc}(f+\frac{5}{2T})T] \]

In Figure 3a, the frequency responses for the one term eigenfunctions are plotted. Observe that there is maximum overlap of the two functions in the passband and that in the stopband the peaks of one function's sidelobes fall at the nulls of the other's sidelobes. Since the overall frequency response of the filter is the product of the odd and even eigenfunctions, the orthogonality of the sidelobes about doubles the overall out-of-band rejection, as measured in dB,
Figure 3. (a) Frequency Responses of One-term even and odd eigenfunctions $H_{11}$ and $H_{12}$. (b) Frequency responses of the two term even and odd functions $H_{21}$ and $H_{22}$. (c) Frequency responses of even and odd functions $H_{31}$ and $H_{32}$. 
over that achieved by either function alone. This characteristic is
demonstrated in Figure 4. Figures 3b and 4b show the frequency re-
sponses for the two term eigenfunctions. Notice that the characteris-
tics pointed out for one term functions also hold for the two term
functions, i.e., overlap in the passband and orthogonal sidelobes.
Also, the sidelobes of the three term functions, Figure 3c, are lower
than those of either $H_{1n}(f)$ or $H_{2n}(f)$. The sidelobes are not equirip-
ple, but this is not very important considering the other properties
these functions have.

The Dolph-Chebyshev window function has been shown to be nearly
optimum in terms of minimizing the time-bandwidth product for a speci-
ficied sidelobe level. The solid line in Figure 5 is an approximation
of the expected sidelobe level versus the time-bandwidth product for a
Dolph-Chebyshev window function [6]. The three eigenfunction pairs
are also plotted on the graph. Notice that the sidelobe level for a
given time-bandwidth product for each of the eigenfunctions is nearly
as good as the ideal Dolph-Chebyshev window function. The product of
eigenfunction pairs appears less than optimum, however, indicating
that truncation may be used to shorten the IR length. The advantage
of the eigenfunction pairs over the Dolph-Chebyshev window function is
that the zeros of the IR can be evenly divided between two trans-
ducers.

In general for a given N, the transition bandwidth (TBW) is
greater for the even function than for the odd function. While the
sidelobe level is lower for the odd function than for the even func-
Figure 4. (a) Product of $H_{11}$ and $H_{12}$. (b) Product of $H_{21}$ and $H_{22}$. (c) Product of $H_{31}$ and $H_{32}$. 
tion. In all cases of the even-odd function pair, the nulls of one function fall at the peak of the other, except in the vicinity of the null bandwidth where the peaks and nulls do not necessarily occur at the same frequency.

It is apparent that:

\[ h_{12}(t) = \text{Cosine Function} \]
\[ h_{21}(t) = \text{Hamming Function} \]
\[ h_{31}(t) = \text{Vasile’s Eigen Function} \]

These are functions which are also applied to window function techniques [9]. In addition, there are two other functions which have been presented which could be used as window functions. As \( N \) is
increased, there are an infinite number of possible window functions with ever decreasing sidelobes.

The number of terms in the even and odd functions need not be the same. $H_{N1}(f)$ can be used with $H_{(N-1)2}(f)$. The sidelobe level will be higher than for the $N^{th}$ term set, but will be lower than the $(N-1)^{th}$ term set.

The synthesis procedure presented in Chapter 4 is developed using the two term eigenfunctions because the functions offer a good trade-off between impulse response length, out of band rejection and realizability.
DeVries describes an approach to designing FIR filters using a rectangular window function as a basic building block [4]. This method utilizes the superposition of rectangular time functions with appropriate center frequencies, impulse response lengths, and amplitudes to realize a specified frequency response. An arbitrary filter response can be constructed. However, some of the basic blocks must be used to cancel the large sidelobes of the rectangular window function, and small sidelobes are difficult to obtain.

Vasile describes another superposition approach which uses $h_{31}$ as the basis function [5]. The frequency response of $h_{31}$ is assumed to be triangular in shape. The parameters of the eigenfunctions, center frequency, IR length, and amplitude, are chosen to implement a particular frequency response using the triangular approximation. The eigen frequency function is not triangular in shape and, as Vasile states a triangular frequency function is undesirable since its IR is not finite. The deviations from the ideal triangular shape introduce errors in the frequency response of the resulting filter. In many applications, implementation constraints require that the IR be truncated introducing more errors in the resulting frequency response.

Both of these approaches provide a good basis for a synthesis procedure. However the accuracy of the methods are inadequate. By
introducing an infinite set of functions from which to choose basis functions, an additional degree of freedom is obtained for designing filters.

The total filter response will be given by the product of the individual transducer filter responses given by

\[ G_T(f) = G_1(f) \cdot G_2(f) \]

The individual transfer functions, \( G_1(f) \), \( G_2(f) \) are defined as

\[ G_i(f) = \sum_{m=1}^{M} c_m H_i(f-f_{m',T}), \text{ f>0} \]

where \( c_m \) is a real amplitude coefficient and \( H_i(f,T) \) is the frequency eigenfunction which is normalized to unity at \( f=0 \). The individual frequency responses are a summation of properly weighted and shifted eigenfunctions. This is similar to Vasile's approach. However no approximation to a triangle is assumed and the proper frequency translation criteria is yet to be defined [5].
CHAPTER IV
SYNTHESIS PROCEDURE

There are three regions of interest in the frequency response of a filter: the passband, the stopbands, and the transition bands.

In the stopband regions, the main concern is the level of the highest sidelobe. Because a SAW filter is a linear system, the property of superposition applies. Therefore, the orthogonality of the sidelobes between \( H_{21}(f-f_n) \) and \( H_{22}(f-f_n) \) is valid for all \( n \) and for the sum of multiple functions. Hence, the sidelobe level is set by the characteristics of the functions as previously discussed.

Four degrees of freedom exist for the design of the filter, the IR length, the number of frequency translated functions, the amplitudes of the functions, and their center frequencies. For a flat passband, the amplitudes of the functions are all set to unity. For a shaped passband, the amplitudes can be scaled accordingly.

Vasile describes a technique for constructing a filter using the sum of frequency translated triangles [5]. In this method, the center frequencies of the triangles are chosen such that adjacent triangles cross at the 6 dB points. The functions to be used in this synthesis procedure are given by

\[
G_1(f,T) = \sum_{m=1}^{M} .537T \text{sinc}(f-f_m)T + .266T[ \text{sinc}(f+f_m)T + \text{sinc}(f-f_m)T]
\]

17
\[ G_2(f,T) = \sum_{m=1}^{M} \left[ .398T[sinc(f-\frac{1}{2T}f_m)T + sinc(f+\frac{1}{2T}f_m)T] + .102T[sinc(f-\frac{3}{2T}f_m)T + sinc(f+\frac{3}{2T}f_m)T] \right] \]

To maintain the orthogonality of the sidelobes between \( G_1(f,T) \) and \( G_2(f,T) \), the center frequencies \( (f_m) \) of the individual eigenfunctions must be the same for the even and odd functions. Instead of using the criteria that adjacent functions cross at their 6 dB points [5], the minimization of the passband ripple of the overall filter is used as the criteria for the optimum spacing between functions. If the functions in frequency were ideal triangles, the ripple in the passband would not exist. However, the functions are rounded at the peaks and have sidelobes. Both of these items produce ripple in the passband. By choosing the spacing between the functions properly, a compromise can be reached between the two effects to produce minimum passband ripple. For the two functions presented the spacing is \( \frac{1.385}{T} \). This number was arrived at empirically and holds only for the two term eigenfunctions, \( H_{21}(f) \) and \( H_{22}(f) \).

Using this spacing, it is observed that the peak amplitude of the overall filter response is higher than that of a single eigenfunction. Therefore the passband and transition bandwidths (BW and TBW respectively) are not only a function of the IR length but also dependent on the number of eigenfunctions used. This effect is demonstrated in Figure 6.
Figure 6. (a) Frequency Response of the sum of 4 even Eigenfunctions. (b) the sum of 4 odd Eigenfunctions
It is well known from the properties of the Fourier Transform that as the IR length is decreased the transition bandwidth increases. The BW of a filter can be expressed as the TBW times an arbitrary constant ($\beta$), normally greater than 1. Remembering that the spacing between functions makes the TBW a function of the number of eigenfunctions used, the TBW is proportional to $\frac{F(N)}{L}$, where $F(N)$ describes the TBW as a function of the number of eigenfunctions. Let the shapefactor (SHP) be defined as

$$\text{SHP} = \frac{\text{BW} + 2\text{TBW}}{\text{BW}}$$

Substituting the relationship between the BW and the TBW the shapefactor can be rewritten as

$$\text{SHP} = \frac{\beta \text{TBW} + 2\text{TBW}}{\beta \text{TBW}} = \frac{\beta + 2}{\beta}$$

From this last relationship, it is apparent that the shapefactor can be defined as a function of $\beta$ alone. The BW of a filter, as shown in Figure 7, is defined at an amplitude function of the IR length but

![Figure 7. Filter Frequency Response Definitions](image-url)
also dependent on the number of the frequency response \( L_1 \). The TBW is the change in frequency between the BW amplitude, \( L_1 \), and a null level, \( L_2 \). \( L_1 \) normally ranges between 0 and -6 dB and \( L_2 \) ranges, in most applications, between -30 and -60 dB. It is obvious that the shapefactor, and therefore the minimum number of eigenfunctions, is a function of both \( L_1 \) and \( L_2 \). The minimum number of eigenfunctions required for a particular shapefactor was found empirically as

\[
N = 0.57(SHP) - 0.000282(NL)^2 + 0.0487(NL) - 308
\]

where

\( N \) — minimum number of eigenfunctions
\( SHP \) — shapefactor
\( NL \) — null level in dB

This relationship was arrived at with the passbandwidth defined at the -6dB level. For a passbandwidth defined at the -3dB level the relationship is

\[
N = 0.975(SHP) + 0.02373(NL) + 23132
\]

With the minimum number of eigenfunctions \( N \) determined, the IR length \( (T) \) can now be determined using the passbandwidth at the 6 dB levels. Taking into account the additional amplitude in the passband due to overlap between eigenfunctions, the IR length is given as

\[
T = 1.38099(N) - 75121
\]
For the passbandwidth defined at the 3 dB levels

\[ T = \frac{1.38209(N) - 1.35127}{BW} \]  

(12)

Notice that the relationship for the minimum number of eigenfunctions may not produce an integer number. The number is rounded up to the nearest integer. The IR length is then calculated using the rounded number of functions and the passbandwidth, forcing the passbandwidth to the desired width. By using more than the minimum number of functions, the TBW will be less than or equal to the desired TBW. In most filter applications, the TBW is required to be less than a specified value. With a TBW narrower than specified, the filter is not quite optimum, but will always meet specifications.
CHAPTER V
EFFECTS OF IR TRUNCATION

Original Functions

For low shapefactor filters, implementation constraints quite often require that the IR be truncated. The following effects have been noted when \( h_{21} \) and \( h_{22} \) were truncated:

1. The zeros of the components moved from their original positions.
2. The periodicity of the sidelobes is dependent on the fractional truncation point.
3. The property that the peak of one function's sidelobes fall at the nulls of the other's no longer holds.
4. As the truncation is increased the sidelobes become in phase.
5. The TBW decreases with \( T \) held constant and hence the time-transition bandwidth product decreases.
6. The energy is redistributed increasing the sidelobe level and widening the passband.

By decreasing the time-transition bandwidth product of the basis functions, the sidelobe level for a particular time-transition bandwidth product is better optimized. This gain is offset, however, by the sidelobes being in phase, i.e., the zeros are not exclusive. The zeros of \( H_{21} \) and \( H_{22} \) can be made exclusive by re-optimizing the coefficients.
Re-optimizing the Coefficients

The original criteria for choosing the coefficients was applied to $h_{21}$ as a first attempt to re-optimize the coefficients. The nulls in the resulting frequency response were not located where expected. The reason for this behavior is not fully understood at this time and will require further investigation. If a null is placed near a natural null, the null moves to its new position as predicted. The sidelobe level for a particular time-transition bandwidth product is improved as would be expected.

The behavior of the nulls of $H_{22}$ are more predictable than those of $H_{21}$ when re-optimizing the coefficients. Therefore the null point for $H_{22}$ can be chosen such that the zeros between $H_{21}$ and $H_{22}$ are exclusive. The sidelobe level for the product of the two functions is also improved for a particular time-transition bandwidth product as expected. The truncated functions with the re-optimized coefficients can be implemented without further truncation as desired.

To date, only the truncation of the two term eigenfunctions has been investigated. Preliminary results indicate that the behavior of the functions and hence the re-optimization method can be extended to any number of terms. This area will require further investigation, such that the number of terms can be incorporated into the synthesis procedure to set the level of the sidelobes.
CHAPTER VI

DESIGN EXAMPLE

To demonstrate the synthesis procedure, a filter was designed. The desired and designed characteristics are presented in Table 1.

<table>
<thead>
<tr>
<th>Desired</th>
<th>Designed</th>
</tr>
</thead>
<tbody>
<tr>
<td>BW (6 dB)</td>
<td>Fo</td>
</tr>
<tr>
<td>6 MHz</td>
<td>60 MHz</td>
</tr>
<tr>
<td>6 MHz</td>
<td>60 MHz</td>
</tr>
</tbody>
</table>

The frequency response of the resulting filter is shown in Figure 8. Figure 9 shows the impulse responses for the two functions. Notice that in the odd function, the tap weights of the last two time side-lobes are smaller than desired. The IR was truncated to exclude these last two sidelobes. As shown in Figure 10, the effect on the frequency response of the filter due to the truncation was to increase the frequency sidelobe level and also to increase the passband ripple. For the truncated IR the passband ripple is .48 dB (p-p) and the out of band rejection is -78 dB. It is interesting to note that truncating the IR only significantly effects the passband ripple, which is initially small. This synthesis procedure requires both transducers to be weighted.
Figure 8. (a) Frequency response of design example
(b) Frequency response of passband
Figure 9. (a) Impulse response of design example, even function, (b) Impulse response of design example, odd function.
Figure 10. (a) Frequency response of truncated impulse responses, (b) Passband of truncated impulse responses.
CHAPTER VII
CONCLUSIONS

A non-iterative technique using the superposition principal has been presented for designing FIR filters. An infinite number of basis functions were defined and the method for determining the proper coefficients was described. It has been shown that these eigenfunctions are nearly optimum in terms of the sidelobe level for a given time-transition bandwidth product. Using the characteristics of the functions, a synthesis procedure was developed. The synthesis procedure was developed using the two term series due to the implementation constraints of SAW device technology.

It has further been shown that truncation of the impulse response can be used to improve the sidelobe level for a given time-transition bandwidth product. Techniques similar to those used for choosing the original coefficients can be used to re-optimize the truncated functions. These re-optimized functions show even more improved sidelobe level, and hence passband ripple, for a particular time-transition bandwidth product.

A design example was used to illustrate the synthesis procedure. For a filter with 10% fractional bandwidth and a shapefactor of 1.15, the resultant design meets the criteria such that the bandwidth be within 5% and the transition bandwidth be less than or equal to the designed value.
It was noted that the smaller taps of the impulse response could not be implemented using SAW device technology. These taps were excluded and the only significant change in the frequency response was an increase in passband ripple.

The synthesis procedure presented will reduce the design time required for low shapefactor SAW filters. Further work needs to be continued to use truncated eigenfunctions with more than two terms.
APPENDIX A: PROGRAM LISTING

common/file/ amp(4096),phase(4096),nfft,itype
common/dat/ fo,tflo,tfhi,num
dimension aeig(50),foeig(50),a(50),b(2,2),s(4096)
real nl,pi
c
character ans*1

c
pi=3.14159265
iccont=10

c
write(6,*),'EIGEN SYNTHESIS PROGRAM'
write(6,*)'UNIVERSITY OF CENTRAL FLORIDA'
write(6,*)'COLLEGE OF ENGINEERING'
write(6,*)'ELECTRICAL ENGINEERING AND COMMUNICATION
+ SCIENCES DEPARTMENT'
write(6,*)'ORLANDO, FLORIDA'
write(6,*)'REV. 1, JUNE 18, 1984'
write(6,*)'WRITTEN BY CARLTON BISHOP'

READ IN DESIGN PARAMETERS

c
write(6,*),'Enter passbandwidth'
read(5,*)pbw
write(6,*),'Enter level where BW defined (3 or 6 dB)'
read(6,*)bwd
write(6,*),'Enter Fsamp, Fo'
read(5,*)fs,fo
write(6,*),'Enter TBW'
read(5,*)tbw
write(6,*),'Enter null level (dB)'
read(5,*)nl
write(6,*),'Enter passband slope (dB across passband)'
read(5,*)s1db

c
Calculate Filter Parameters

c
b(1,1)=.543
b(1,2)=.457
b(2,1)=.796
b(2,2)=.204
shp=(pbw+2*tbw)/pbw
if(bwd.eq.6.0) then
    itry=int((.55*shp-.000282*nl**2+.0487*nl-.308)/(shp-1)+.99)
    tau=(1.38099*float(ityr)-.75121)/pbw
else
    itry=int((.975*shp+.02373*nl+.23132)/(shp-1)+.99)

31
\[ \tau = (1.38209 \times \text{float}(\text{ity}r) - 1.35127) / \text{pbw} \]

endif

slope = \frac{1.0 - 10^{-5} \times (\text{abs(sldb/2.0)} / 20))}{(1.385 \times \text{float(ityr))}}

if (sldb > 0.0) slope = -slope

write(6,*,'Enter fractional truncation point')

read(*,*)tf

tau2 = tau * tf

fmin = fo - float(ityr-1) * 1.385 / (2 * tau)

do 100 i = 1, itry

if (i.eq.1) then
    aeig(1) = 1.0
else
    aeig(i) = aeig(i-1) + 1.385 * slope
endif

100

foeig(i) = fmin + float(i-1) * 1.385 / tau

itype = -1

nfft = \text{int}(fs * tau2/2) * 2 + 1

num = nfft

c

c Calculate Impulse Response

c
dt = 1 / fs

tflo = -float(nfft/2) * dt

tfhi = tflo

samp = fs / fo

if (samp eq 4.0) then
    tshift = 1.0 / (2.0 * fs)
else
    tshift = 0.0
endif

do 200 m = 1, 2

t = tflo - tshift

ymax = -1.0e38

do 300 i = 1, nfft

if (m.eq.2) goto 400

s(i) = 0.0

do 500 n = 1, itry

v = \cos(2\pi \times \text{foeig(n)} \times t)

s(i) = s(i) + v

400 if (m.eq.1) then

amp(i) = s(i) * (b(1,1) + b(1,2) * \cos(2.0 * pi * t / tau))

else

amp(i) = s(i) * (b(2,1) * \cos(pi * t / tau) + b(2,2) * \cos(3. * pi * t / tau))
endif

t = t + dt

if (amp(i).gt.ymax) ymax = amp(i)

500 phase(i) = 0.0

do 600 n = 1, nfft

600 amp(i) = amp(i) / ymax

write(6,700) m
700  format( ' Writing data for transducer ',i4)  
call writeo(icont)  
200  continue
    end
REFERENCES


