More Economical State-Space Digital-Filter Structures
Which are Free of Constant-Input Limit Cycles

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Abstract—A theorem that establishes conditions for the elimination of constant-input limit cycles is used as the basis for a procedure for the synthesis of state-space structures. Three different second-order structures are obtained, and in each, granularity and overflow limit cycles can be eliminated under zero- and constant-input conditions. Detailed procedures are then presented for the application of one of the proposed structures to the design of parallel and cascade filters. The paper concludes with extensive comparisons which show that more economical low-noise and low-sensitivity state-space filter implementations are possible through the use of one of the proposed structures.

I. INTRODUCTION

SEVERAL approaches for the realization of recursive digital filters have evolved in recent years. A particular approach that received significant attention in the past is the state-space approach [1]–[6]. The interest in this approach is due to the fact that it leads to optimal digital-filter structures with respect to roundoff noise [1]–[4]. State-space structures are relatively uneconomical in practice since they entail a large number of multiplications. Nevertheless, their use should seriously be considered in applications where low-noise narrow-bandwidth filters are required.

In this paper, an alternative procedure to the synthesis of second-order state-space structures is proposed. The procedure leads to more economical low-noise state-space structures in which granularity and overflow limit cycles [7]–[10] can be eliminated under zero- and constant-input conditions.

II. SECTION-OPTIMAL STRUCTURE

Given a second-order transfer function

\[ H(z) = \frac{\beta_1 z + \beta_2}{z^2 + \alpha_1 z + \alpha_2} \]  

(1c)

a digital-filter structure can be obtained which is characterized by

\[ x(k + 1) = Ax(k) + Bu(k) \]  

(2a)

\[ y(k) = Cx(k) + du(k) \]  

(2b)

where

\[ A = \begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix}, \quad B = \begin{bmatrix} b_1 \\ b_2 \end{bmatrix}, \quad C = [c_1, c_2]. \]

If

\[ a_{11} = a_{22} \]  

(3a)

\[ \frac{b_1}{b_2} = \frac{c_2}{c_1} \]  

(3b)

the output roundoff noise is minimized, as was demonstrated in [4]. Such a structure is said to be section optimal and its transition matrix assumes the form [4], [5]

\[ A = \begin{bmatrix} a & -
\frac{\xi}{\sigma} \\ \frac{\xi}{\sigma} & a \end{bmatrix} \]  

(4)

where \( a, \xi, \) and \( \sigma \) are constants.

III. ELIMINATION OF LIMIT CYCLES

A. Zero-Input and Overflow Limit Cycles

Zero-input limit cycles can be eliminated in a recursive structure if there exists a positive-definite diagonal matrix \( G \) such that \( G - A^TG \) is positive definite [7]. This condition is satisfied if [8]

\[ a_{12} a_{21} \geq 0 \]  

(5a)

or if

\[ a_{12} a_{21} < 0 \quad \text{and} \quad |a_{11} - a_{22}| + \det(A) < 1. \]  

(5b)

In the section-optimal structure, the parameters of matrix \( A \) satisfy (5b) since \( a_{11} = a_{22} \), and \( \det(A) < 1 \) for a stable filter. In effect, zero-input limit cycles can be eliminated in this structure by quantizing the state variables such that

\[ |x_i(k)|_{\text{Q}} \leq |x_i(k)| \quad \forall \ k \]
where \( [\cdot]_Q \) is the quantized value of \([\cdot]\). Overflow limit cycles, on the other hand, can be eliminated as in wave digital filters by using the approach described in [9].

**B. Constant-Input Limit Cycles**

In this section, a theorem is proved which establishes sufficient conditions for the elimination of constant-input limit cycles in a general digital-filter structure [11]. This will be used in Section IV for the derivation of new state-space structures in which constant-input limit cycles can be eliminated.

**Theorem 1**: Assume that the digital-filter structure of Fig. 1 is free of zero-input limit cycles and that

\[
\begin{align*}
  x(k+1) &= [Ax(k) + Bu(k)]_Q \\ \\
  y(k+1) &= Cx(k) + du(k)
\end{align*}
\]  

(6a)  

(6b)

where \([\cdot]_Q\) is the quantized value of \([\cdot]\).

Constant-input limit cycles can be eliminated by modifying the structure of Fig. 1 as depicted in Fig. 2 where \( P \) is given by

\[
P = [p_1 \ p_2 \ \cdots \ p_n]^T = (I - A)^{-1}B
\]

(7)

provided that \( Pu_0 \) is machine representable where \( u_0 \) is the constant input.

**Proof**: Since the structure of Fig. 1 is assumed to be free from zero-input limit cycles, the equation

\[
x(k+1) = [Ax(k)]_Q
\]

(8)

describes a stable system such that

\[
\lim_{k \to \infty} x(k) = [0 \ 0 \ \cdots \ 0]^T.
\]

If the input is constant, i.e., \( u(k) = u_0 \), the modified structure of Fig. 2 is characterized by

\[
x(k+1) = [Ax(k) - Pu_0 + Bu_0]_Q + Pu_0
\]

and if (7) holds, we have

\[
x(k+1) = [Ax(k) - I(I - A)^{-1} Bu_0 \\
+ (I - A)(I - A)^{-1} Bu_0]_Q + Pu_0
\]

\[
= [A\{x(k) - Pu_0\}]_Q + Pu_0
\]

Hence,

\[
x'(k+1) = [Ax'(k)]_Q
\]

(9)

where

\[
x'(k) = x(k) - Pu_0.
\]

Evidently, (9) is of the same form as (8), except for the translation in the state variables and, therefore, it represents a stable system. Stability can be guaranteed if (7) is satisfied exactly and, therefore, \( Pu_0 \) is required to be machine representable.

Q.E.D.

The application of the above theorem is straightforward and does not affect the output equations of the structure. It should be mentioned that if the quantization in Fig. 2 is carried out by means of magnitude truncation, an implementation of the controlled-rounding arithmetic proposed by Butterweck [10] is achieved.

**IV. New State-Space Synthesis**

For a second-order state-space structure, column vector \( P \) may assume the machine-representable forms

- **case I**: \( P = [\pm 1 \ 0]^T \)
- **case II**: \( P = [0 \ \pm 1]^T \)
- **case III**: \( P = [\pm 1 \ \pm 1]^T \)

and for each case, \( B \) can be chosen so as to assure the elimination of constant-input limit cycles. From (7), \( B \) can be chosen as

- **case I**: \( b_1 = \pm(1 - a_{11}), \ b_2 = \mp a_{21} \)
- **case II**: \( b_1 = \mp a_{12}, \ b_2 = \pm(1 - a_{22}) \)
- **case III**: \( b_1 = \pm(1 - a_{11}) \neq a_{12}, \ b_2 = \mp a_{21} \pm (1 - a_{22}) \)

(10a)  

(10b)  

(10c)

On the basis of the above assignments, structures I–III illustrated in Fig. 3(a)–(c) can be derived. Adders and quantizers needed for the elimination of constant-input limit cycles are omitted for the sake of simplicity. As can be seen, coefficients \( b_1 \) and \( b_2 \) are formed without the need of additional multipliers and, consequently, these structures require fewer multipliers than the section-optimal structure. Structures I and II entail a similar computational complexity. Structure III, however, requires five extra additions and it will not be considered further.

The multiplier constants for structure I can be deduced from (1), (2), and (4) as

\[
a_{11} = a_{22} = a
\]

(11a)
In order to eliminate the possibility of overflow, the input signal can be scaled through the use of a scaling multiplier $\lambda$ as depicted in Fig. 3(d).

V. DESIGN CONSIDERATIONS

A. Dynamic Range

The signal-to-noise ratio in a fixed-point digital-filter implementation can be increased by increasing the dynamic range, and the latter can be increased by equalizing the maximum signal levels at the inputs of multipliers. For structures I and II, the maximum signal levels to be equalized are state variables $x_i(k)$.

The transfer functions from input node $u(k)$ to the state-variable nodes $x_i(k)$, designated as $F_i(z)$, can readily be obtained from (2) and (4). For structure I, we have
\[ F_1(z) = \frac{(1 - a)z + (\xi^2 - a + \alpha^2)}{z^2 - 2az + \alpha^2 + \xi^2} \]
and
\[ F_2(z) = \sigma F_1(z) \]
where
\[ F_2(z) = \frac{-\xi z + \xi}{z^2 - 2az + \alpha^2 + \xi^2}. \]

The maximum signal levels at the inputs of the quantizers can be equalized by letting
\[ \|F_1(z)\|_p = \|\sigma F_2(z)\|_p \]
where \( p = \infty \) or 2, and therefore,
\[ \sigma = \frac{\|F_1(z)\|_p}{\|F_2(z)\|_p}. \]  \hspace{1cm} (15)

It should be mentioned here that the derivation of \( A \) given in (4) is based on the use of the \( L_2 \) norm for scaling. However, as stated in [4], this choice of \( A \) is also good if the \( L_\infty \) norm is used.

A similar analysis can be applied to structure II, but the equalized structure turns out to be equivalent to equalized structure I. For this reason, structure II will not be considered further.

Overflow can be eliminated by choosing scaling multiplier constant \( \lambda \) for each section [see Fig. 3(d)] as described in Section VI.

In the above considerations, it is inherently assumed that the structural modifications needed for the elimination of constant-input limit cycles do not increase \( \|F_1(z)\|_p \) for \( i = 1, 2 \). This, indeed, is the case as can be shown through a somewhat complicated proof.

B. Optimality with Respect to Roundoff Noise

In the section-optimal structure, optimality with respect to roundoff noise is assured by the conditions in (3a) and (3b). For structure I, (4) shows that the condition in (3a) is satisfied. From (10) and (12),
\[ \frac{b_1}{b_2} = \frac{(2 + \alpha_1)}{2\sigma \xi} \]
and
\[ \frac{c_2}{c_1} = -\frac{(\alpha_1 + 2\alpha_2) \beta_1 + (2 + \alpha_1) \beta_2}{2\sigma \xi (\beta_1 + \beta_2)}. \]
Hence, (3b) will be satisfied if and only if
\[ \frac{\beta_1}{\beta_2} = \frac{\alpha_1 + 2}{\alpha_2 - 1}. \]  \hspace{1cm} (16)
This limitation is imposed on structure I by choosing \( B \) as in (10a) for the purpose of eliminating constant-input limit cycles.

If the zeros of \( H(z) \) are located at \( z = 1 \), as in the case of Butterworth, Chebyshev, and Bessel high-pass filters, (1) gives
\[ \beta_1 = -\gamma_0 (2 + \alpha_1) \]  \hspace{1cm} (17a)
\[ \beta_2 = \gamma_0 (1 - \alpha_2) \]  \hspace{1cm} (17b)
and, therefore, from (16) and (17), we note that (3b) is satisfied, i.e., optimality with respect to output roundoff noise is assured for these filters.

Although optimality cannot be assured in general, extensive experimental results presented in Section IX show that the output roundoff noise in structure I is comparable to that in the section-optimal structure of [4] in practice.

A final point of interest here is that choosing \( \sigma \) as in (15) leads to the minimum output roundoff noise with respect to \( \sigma \). The proof is omitted for the sake of brevity.

VI. DESIGN PROCEDURE

Structure I can be used for the design of high-order parallel as well as cascade filters. The following procedures can be used.

A. Parallel Design

1) Express the transfer function of the filter as
\[ T(z) = d + \sum_{i=1}^{m} H_i(z) \]
where each \( H_i(z) \) is of the form given by (1c).
2) Compute \( a \) and \( \xi \) for each \( H_i(z) \) using (13) and (14).
3) Compute \( \sigma \) for each section according to (15). If the \( L_2 \) norm is used for scaling,
\[ \sigma = \sqrt{\frac{(2 + \alpha_1)^2[(1 + \alpha_2)(1 + \mu^2) - 2\alpha_1 \mu]}{8\xi^2(1 + \alpha_1 + \alpha_2)}} \]
where
\[ \mu = \frac{\alpha_1 + 2\alpha_2}{\alpha_1 + 2} \]
and if the \( L_\infty \) norm is used for scaling,
\[ \sigma = \frac{2 + \alpha_1}{2\xi} \sqrt{\frac{1 + f^2 + 2f \cos \omega_0}{2(1 - \cos \omega_0)}} \]
where \( \omega_0 \) is the frequency of the pole and
\[ f = \frac{\alpha_1}{2} + \frac{2\xi^2}{2 + \alpha_1}. \]
4) Compute \( A \) and \( C \) for each section using (11) and (12).
5) Compute the scaling constant \( \lambda \) for each section as
\[ \lambda = \frac{1}{\|F_2(z)\|_p}. \]
If \( p = 2 \),
\[ \lambda = \sqrt{\frac{(1 - \alpha_1 + \alpha_2)(1 - \alpha_2)}{2\sigma^2 \xi^2}} \]
and if $p = \infty$,

$$
\lambda = \frac{1 - r}{\sigma^2} \sqrt{\frac{1 + r^2 - 2r \cos 2\omega_0}{2(1 - \cos \omega_0)}}
$$

where $r$ is the radius of the pole.

6) In order to restore the signal level at the output of each section, replace $c_1$ and $c_2$ by $c'_1$ and $c'_2$, respectively, where

$$
c'_1 = c_1 / \lambda
$$
$$
c'_2 = c_2 / \lambda.
$$

B. Cascade Design

1) Express the transfer function as

$$
T(z) = \prod_{i=1}^{m} H_i(z)
$$

$$
= H_0 \prod_{i=1}^{m} \frac{z^2 + \gamma_{i1}z + \gamma_{i2}}{z^2 + \mu_{i1}z + \mu_{i2}}
$$

$$
= \prod_{i=1}^{m} [d_i + H'_i(z)]
$$

where $H_i(z)$ and $H'_i(z)$ are of the form given by (1a) and (1c), respectively.

2) Compute $\sigma$ and $\lambda$ for each section as

$$
\sigma_i = \left\| \left[ \prod_{j=1}^{i-1} H_j(z) \right] F_{ii}(z) \right\|_p
$$

$$
\lambda_i = \frac{1}{\left\| \left[ \prod_{j=1}^{i-1} H_j(z) \right] F_{ii}(z) \right\|_p}
$$

(18)

3) Compute $\alpha$, $\xi$, $A$, and $C$ as in the parallel design.

4) Compute multiplier coefficient $d$ for each section as

$$
d_i = \frac{1}{\prod_{j=1}^{i} H_j(z)} \text{ for } i = 1, 2, \ldots, m - 1
$$

$$
d_m = \frac{H_0}{\prod_{i=1}^{m-1} d_i}
$$

in order to satisfy the required overflow constraints at the output of each section.

5) Incorporate the scaling multipliers of sections 2, 3, $\ldots$, $m$ in the output multipliers of sections 1, 2, $\ldots$, $m - 1$ by replacing constants $c_{1i}$, $c_{2i}$, and $d_i$ by $c'_{1i}$, $c'_{2i}$, and $d'_i$, respectively, where

$$
c'_{1i} = c_{1i} \lambda_{i+1} / \lambda_i
$$
$$
c'_{2i} = c_{2i} \lambda_{i+1} / \lambda_i
$$
$$
d'_i = d_i \lambda_{i+1} / \lambda_i.
$$

VII. Output-Noise Calculations

For a design comprising $m$ parallel sections, the relative power spectral density (RPSD) of the output noise [12] can be obtained as

$$
\text{RPSD} = 1 + \sum_{i=1}^{m} \left[ 2|G_{1i}(e^{j\omega T})|^2 + 2|G_{2i}(e^{j\omega T})|^2 + 3 \right]
$$

where

$$
G_{1i}(z) = \frac{c'_{1i}(2z + \alpha_{1i} + 2\xi)}{2(z^2 + \alpha_{1i}z + \alpha_{2i})}
$$

$$
G_{2i}(z) = \frac{c'_{2i}(2z + \alpha_{1i} - 2\xi^2/\xi)}{2(z^2 + \alpha_{1i}z + \alpha_{2i})}
$$

and

$$
\xi = \frac{-(\alpha_{1i} + 2\alpha_{2i}) \beta_{1i} + (2 + \alpha_{1i}) \beta_{2i}}{2(\beta_{1i} + \beta_{2i})}.
$$

$G_{1i}(z)$ and $G_{2i}(z)$ are the transfer functions from the state variable nodes $x_{1i}(k+1)$ and $x_{2i}(k+1)$, respectively, to the output of the $i$th section.

Similarly, for a design comprising $m$ cascade sections, the RPSD of the output noise is given by

$$
\text{RPSD} = \sum_{i=1}^{m} \frac{1}{\lambda_{i+1}^2} \left[ 2|G_{1i}(e^{j\omega T})|^2 + 2|G_{2i}(e^{j\omega T})|^2 + 3 \right]
$$

$$
\cdot \left[ \prod_{j=i+1}^{m} H_j(e^{j\omega T}) \right]^2
$$

(19)

where $H_i(z)$ is given by (1a), $\lambda_i$ is given by (18), and $\lambda_{m+1} = 1$.

VIII. Comparison of Computational Complexity

The computational complexity inherent in structure I is compared to that of the sectional-optimal structure in Table I for the case where an $n$th-order transfer function is realized by means of parallel or cascade sections. As can be seen, for an even-order filter, the new structure reduces the number of multiplications by $n/2$ in a parallel design or by $n - 1$ in a cascade design, although the number of additions is increased by $n/2$ in each case.

IX. Roundoff Noise and Sensitivity Comparisons

For the sake of comparison, the cascade approach was used to design four sixth-order filters which included an elliptic low-pass, a Chebyshev high-pass, an elliptic bandpass, and a Butterworth bandstop filter. The specifications of the various filters are given in Table II where
TABLE I
ARITHMETIC OPERATIONS

<table>
<thead>
<tr>
<th></th>
<th>Parallel</th>
<th>Cascade</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$n$ even</td>
<td>$n$ odd</td>
</tr>
<tr>
<td>Multiplications</td>
<td>$\frac{7n}{2} + 1$</td>
<td>$\frac{7n}{2}$</td>
</tr>
<tr>
<td>Additions</td>
<td>$\frac{7n}{2} + 1$</td>
<td>$\frac{7n - 3}{2}$</td>
</tr>
<tr>
<td>New Structure</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Multiplications</td>
<td>$\frac{4n}{2} + 1$</td>
<td>$\frac{4n}{2}$</td>
</tr>
<tr>
<td>Additions</td>
<td>$\frac{3n}{2} + 1$</td>
<td>$\frac{3n - 1}{2}$</td>
</tr>
</tbody>
</table>

TABLE II
FILTER SPECIFICATIONS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Elliptic Low-Pass Filter</th>
<th>Chebyshev High-Pass Filter</th>
<th>Elliptic Bandpass Filter</th>
<th>Butterworth Bandstop Filter</th>
</tr>
</thead>
<tbody>
<tr>
<td>$A_p$, dB</td>
<td>1.0</td>
<td>0.6</td>
<td>0.5</td>
<td>2.0</td>
</tr>
<tr>
<td>$A_n$, dB</td>
<td>72.9</td>
<td>48.6</td>
<td>65.6</td>
<td>300.0</td>
</tr>
<tr>
<td>$\omega_p$, rad/s</td>
<td>250.0</td>
<td>4000.0</td>
<td>980.0</td>
<td>1000.0</td>
</tr>
<tr>
<td>$\omega_n$, rad/s</td>
<td>-</td>
<td>-</td>
<td>1020.0</td>
<td>1000.0</td>
</tr>
<tr>
<td>$\omega_{a1}$, rad/s</td>
<td>400.0</td>
<td>3300.0</td>
<td>850.0</td>
<td>650.0</td>
</tr>
<tr>
<td>$\omega_{a2}$, rad/s</td>
<td>-</td>
<td>-</td>
<td>1150.0</td>
<td>850.0</td>
</tr>
<tr>
<td>$\omega_s$, rad/s</td>
<td>10 000.0</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

$A_p$ = maximum passband ripple, dB
$A_n$ = minimum stopband attenuation, dB
$\omega_p$, $\omega_n$ = passband edges, rad/s
$\omega_{a1}$, $\omega_{a2}$ = stopband edges, rad/s
$\omega_s$ = sampling frequency, rad/s.

Designs were obtained with structure I illustrated in Fig. 3(a) and with the section-optimal structure described in [4]. Signal scaling, based on the $L_\infty$ norm, was applied in all designs, and the section ordering was chosen to minimize the output noise in each case. The implementation was assumed to be in terms of fixed-point arithmetic, and quantization of coefficients and products was assumed to be by rounding. The various designs were compared on the basis of computed output-noise spectra, and on the basis of actual amplitude responses with the coefficients quantized.

The roundoff-noise analysis involved the computation of the relative power spectral density in the various filter designs [see (19)]. The results obtained are depicted in Fig. 4(a)–(d). As can be seen, for the low-pass, bandpass, and bandstop filters, structure I leads to somewhat better results than the section-optimal structure, and for the high-pass filter, it leads to somewhat inferior results.

The results of the sensitivity analysis are depicted on Fig. 5(a)–(d). The coefficient wordlengths for the low-pass, high-pass, bandpass, and bandstop designs were assumed to be 8, 7, 9, and 7 bits, respectively. As can be seen, the performance of the new structure is very similar to that of the section-optimal structure in all examples.

X. CONCLUSIONS

A theorem that establishes sufficient conditions which assure the elimination of constant-input limit cycles in a general digital-filter structure was proved. It was then used as the basis of a procedure for the synthesis of state-space structures in which granularity and overflow limit cycles can be eliminated under zero- and constant-input conditions.

Three different second-order structures were obtained. Two of them, namely, structures I and II, become equivalent if optimum signal scaling is applied, and structure III is somewhat uneconomical with respect to the required number of additions.

Detailed procedures were presented for the application of structure I to the design of parallel and cascade filters. In addition, extensive comparisons were undertaken whereby structure I was compared to the section-optimal structure of [4] with respect to the computational complexity, output roundoff noise, and sensitivity. These results have shown that structure I, like the section-optimal structure, leads to low output roundoff noise and low sensitivity while being more economical with respect to the number of multiplications required. Furthermore, unlike the section-optimal structure, structure I permits the elimination of constant-input limit cycles, as was stated above.
Fig. 4. Output-noise spectra. (a) Elliptic low-pass filter. (b) Chebyshev high-pass filter. (c) Elliptic bandpass filter. (d) Butterworth bandstop filter. ——— Section-optimal structure. ———— Structure I.
Fig. 5. Amplitude responses. (a) Elliptic low-pass filter (wordlength = 8 bits). (b) Chebyshev high-pass filter (wordlength = 7 bits). (c) Elliptic bandpass filter (wordlength = 9 bits). (d) Butterworth bandstop filter (wordlength = 7 bits). —— Ideal. —— Section-optimal structure. ——— Structure I.
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REFERENCES


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