$$e_j^2 = \frac{1}{l}\sum_{i=1}^{l}\left[\log_{10}[|H_{casc}(i)|/|H_{casc}(max)|] - \log_{10}[|H_{test}(i)|/|H_{test}(max)|]\right] \qquad (9)$$

for $j$ = 1, 2, ..., $t_{order}$, where $t_{order}$ = number of filters in the set, $H_{casc}(i)$ = the sampled cascade form frequency response, $H_{test}(i)$ = the sampled frequency response of the filter under test, and $H_x(max)$ = the maximum value of $|H_x(i)|$.

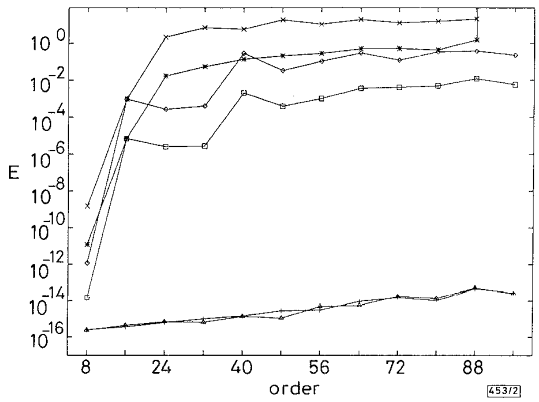


**Fig. 2** *Maximum, minimum and average errors of synthesis algorithms*

◇ new algorithm: maximum error
| new algorithm: minimum error
□ new algorithm: average error
× standard method: maximum error
△ standard method: minimum error
✳ standard method: average error

Since no transformations were required to obtain the cascade form realisations, it was reasonable to assume that they were accurate representations of the filters. An overall average error $E_{order}$ dB, for each set of filters was computed by averaging the average spectral deviations $e_j$, according to eqn. 10.

$$E_{order} = 10\left[\frac{\sum_{j=1}^{t_{order}} e_j^2}{t_{order}}\right]^{\frac{1}{2}} \qquad (10)$$

The average errors $E_{order}$ are plotted together with the maximum and minimum errors for each order, in Fig. 2. This shows that the new algorithm yielded smaller average errors than the DDPP method. Moreover, we found that the new algorithm yielded smaller errors than the DDPP method for almost every filter tested.

*Conclusions:* A new parallel form filter synthesis algorithm has been defined. The new algorithm has a clear performance advantage, in terms of accuracy, over the standard parallel form synthesis method, even for low filter orders.

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**References**

1 HENRICI, P.: 'Partial fractions' *in* 'Applied and computational complex analysis' (Wiley, 1974), Vol. 1, Chap. 7

2 SANDLER, M.B.: 'High order LPC analysis/synthesis of atonal percussion', *Proc. Inst. Acoustics*, 1988, **10**, (7), pp. 113–120

# Dynamic translinear RMS-DC converter

J. Mulder, W.A. Serdijn, A.C. van der Woerd and A.H.M. van Roermund

*Indexing terms: Convertors, Nonlinear filters*

Translinear, log-domain or exponential state-space filters constitute only a subclass of dynamic translinear circuits. The dynamic translinear principle can also be used to implement nonlinear dynamics. This is illustrated by the design of an RMS-DC converter.

*Introduction:* One of the possible techniques to cope with the restrictions imposed by low supply voltages is instantaneous companding [1], which is an inherent characteristic of translinear (TL) filters [2], also called exponential state-space filters [3]. TL filters exploit the exponential large-signal behaviour of the bipolar, or weak inversion MOS, transistor both for an expanding capacitance voltage to output-current conversion and to implement multiplications of currents, based on the translinear principle [4]. Since TL loops are the basic elements of these filters, we propose the term: dynamic translinear principle, thus emphasising that translinear dynamic circuits are an extension of the conventional, i.e. static, translinear principle proposed by Gilbert [4].

TL filters inherit the advantages of conventional TL circuits: insensitivity to variations in temperature and processing, current controllability, large bandwidth and high functional density [5].

Another important characteristic of TL circuits is that they can implement nonlinear transfer functions. Extending this to dynamic TL circuits implies that it will be possible to implement not only linear dynamic transfer functions, i.e. filters, but also nonlinear dynamics. Consequently, the dynamic translinear principle can be used to implement circuits ranging from filters to oscillators and even PLLs and chaotic circuits.

*Circuit description:* A well-known example of a nonlinear dynamic operation is RMS-DC conversion, which can be performed by solving the equation

$$I_{out} = \left\langle \frac{I_{in}^2}{I_{out}} \right\rangle \qquad (1)$$

where <···> represents an averaging operation.

In conventional RMS-DC converters the squaring and division operations are performed by a second-order TL loop [6]. A lowpass filter implements the averaging operation. Often, this is a first-order RC section. However, this filter can also be realised in the translinear domain. Thus, all functions can be merged into one nonlinear dynamic TL circuit, as will be shown.

A TL first-order lowpass filter is described by [7]

$$CU_T\dot{I}_y + I_0 I_y = I_0 I_x \qquad (2)$$

where $I_x$ and $I_y$ represent the input and output current of the filter, respectively, $U_T$ is the thermal voltage and $I_0$ DC current. The dot represents a time derivative. The cutoff frequency of this filter is given by $\omega_c = I_0/(CU_T)$.

When we apply eqn. 2 to implement the averaging operation in eqn. 1, the input current $I_x$ of the filter is not the input current $I_{in}$ of the RMS-DC converter, but the fraction $I_{in}^2/I_{out}$. Therefore, a dynamic TL first-order RMS-DC converter is described by the differential equation

$$CU_T\dot{I}_{out}I_{out} + I_0 I_{out}^2 = I_0 I_{in}^2 \qquad (3)$$

which is nonlinear.

To find a corresponding TL circuit we eliminate the derivative $I_{out}$ from eqn. 3. The capacitance current $I_{cap}$ in the substructure, characteristic for TL filters, depicted in Fig. 1 can be described by

$$I_{cap} = CU_T\frac{\dot{I}_{out}}{I_{out}} \qquad (4)$$

Note that a DC voltage source in series with the base does not influence $I_{cap}$.

Applying eqn. 4, we can eliminate $\dot{I}_{out}$ from eqn. 3. A polynomial equation is obtained, in which all variables are currents:

$$(I_{cap} + I_0)I_{out}^2 = I_0 I_{in}^2 \qquad (5)$$

If the input current $I_{in}$ is full-wave rectified, which is common practice [6], all factors in this equation are positive. In that case, these factors can directly represent collector currents in a TL loop and the polynomial can be implemented through the circuit shown in Fig. 2.

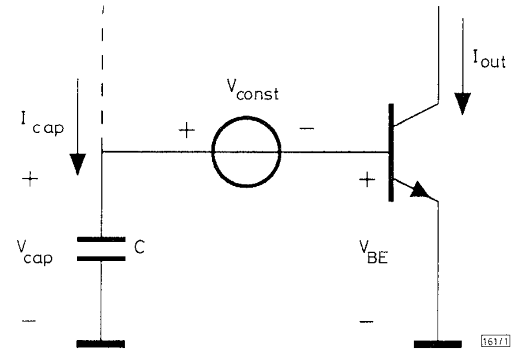


**Fig. 1** *Principle of dynamic translinear circuits*

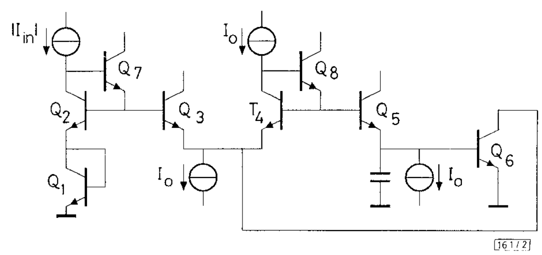


**Fig. 2** *Dynamic translinear RMSDC converter*

Transistors $Q_1$–$Q_6$ constitute a third-order TL loop, implementing eqn. 5. The output current $I_{out}$ is the collector current of $Q_6$. It is fed back to $Q_4$. The substructure $C$–$Q_6$ generates the capacitance current $I_{cap}$ flowing through $Q_5$. $Q_7$ and $Q_8$ constitute two simple amplifier implementations, reducing the effects of finite $\beta_F$. In particular, the base current of $Q_5$, conducting $I_{cap}+I_0$, becomes relatively large for input frequencies around $\omega_c$.

SPICE simulations, using realistic minimum-sized transistor models, were performed to verify the correct operation of the circuit.

*Conclusions:* TL filters can be regarded as an extension of the TL principle. Therefore, the term dynamic translinear principle was proposed. This principle can be used to implement both linear and nonlinear differential equations. As an example of a nonlinear dynamic TL circuit, an RMSDC converter was designed. All necessary functions are merged into one TL loop, thus demonstrating the high functional density obtainable with TL circuits. Correct operation of the circuit was verified by simulations.


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**References**

1 SEEVINCK, E.: 'Companding current-mode integrator: A new circuit principle for continuous-time monolithic filters', *Electron. Lett.*, 1990, **26**, (24), pp. 2046–2047

2 MULDER, J., WOERD, A.C., SERDIJN, W.A., and ROERMUND, A.H.M.: 'General current-mode analysis method of translinear filters', *IEEE Trans.*, 1996, **CAS-I**,

3 FREY, D.R.: 'A general class of current mode filters', *Proc. ISCAS*, 1993, **2**, pp. 1435–1438

4 GILBERT, B.: 'Translinear circuits: A proposed classification', *Electron. Lett.*, 1975, **11**, (1), pp. 14–16

5 SEEVINCK, E.: 'Analysis and synthesis of translinear integrated circuits' (Elsevier, Amsterdam, 1988)

6 GILBERT, B.: 'Translinear circuits – 25 years on – Part III: developments', *Electron. Eng.*, 1993, pp. 51–56

7 ADAMS, R.W.: 'Filtering in the log domain'. 63rd Convention A.E.S., Los Angeles, preprint 1470, May 1979

## Optimum eigenfilters and matched filters

I. Lakkis and D. McLernon

*Indexing terms: Matched filters, Filters*

The authors propose three criteria for maximising the signal-to-noise ratio in detecting a signal in noise. In the stochastic case, the optimal filter maximising the average power of the signal to the average power of the noise is an eigenfilter, and is usually considered as the stochastic counterpart of the matched filter. This interpretation is inexact. However, the optimum eigenfilter can be viewed as a filter matched to the principal component of the signal's power spectral density. A fast, efficient technique for the computation of the optimum eigenfilter is proposed. Simulation results show the efficiency of the proposed algorithm.

*Introduction:* The detection of a signal embedded in additive noise is of special interest in signal processing. Applications can be found in radar, digital communications, and passive sonar. Fig. 1 depicts the detection problem under consideration. The received signal $z(n)$ consists of either a white Gaussian noise $b(n)$ of power spectral density $N_0$ or the noise $b(n)$ plus a signal $x(n)$. Using an FIR linear time-invariant filter of order $N$ characterised by an impulse response $h(n)$ of unit energy, the function of the receiver is to make a decision in favour of one of two hypotheses:

$$\begin{aligned} H_1 &: \quad z(n) = x(n) + b(n) \\ H_0 &: \quad z(n) = b(n) \end{aligned} \qquad n = 0, 1, \ldots \quad (1)$$

We assume that $x(n) = 0$ for $n \notin [0, N-1]$, and that the filter's memory is at least equal to the signal's length ($N \geq L$). Let $x_o(n)$ and $b_o(n)$ denote the signal and noise components of the filter output $y(n)$ under hypothesis $H_1$,

$$\begin{aligned} x_o(n) &= \sum_{k=0}^{N-1} h(k)x(n-k) = \mathbf{x}^T(n)\mathbf{h} \\ b_o(n) &= \sum_{k=0}^{N-1} h(k)b(n-k) = \mathbf{b}^T(\mathbf{n})\mathbf{h} \end{aligned} \quad (2)$$

where the vectors in eqn. 2 have their normal meanings. The detection of the signal reduces to a comparison of the filter output to a threshold [1] as shown in Fig. 1. In this Letter, the optimum filter maximising the output signal-to-noise ratio $(\mathrm{SNR})_o$ for three different criteria is considered, and the analogy between them is analysed.

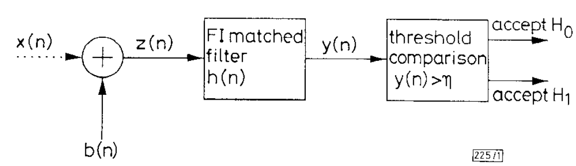


**Fig. 1** *Receiver structure*

(i) *First case:* The signal $x(n)$ is a known deterministic signal of energy $E_x$. We maximise $(SNR)_{o,1}$ which is defined as the instantaneous power in the output signal measured at an arbitrary time instant $m \geq L$ to the average power of the output noise,

$$(SNR)_{o,1} = \frac{|x_o(m)|^2}{E[|b_o(n)|^2]} = \frac{|\mathbf{x}^T(m)\mathbf{h}|^2}{N_0 \mathbf{h}^H \mathbf{h}} \quad (3)$$

The $(SNR)_{o,1}$ is proportional to the square of the scalar product of $\mathbf{h}$ and $\mathbf{x}^T(m)$. Consequently, the maximum can easily be shown to occur when $\mathbf{h}^T = x^H(m)/\sqrt{(E_x)}$ or $h(n) = x^*(m-n)\sqrt{(E_x)}$, i.e. $h(n)$ is a matched filter to the signal component, and the corresponding $(SNR)o$,1 is