Dynamic Translinear Circuits - An Overview

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Abstract - A promising new approach to meet the challenges that the area of analogue integrated circuits is facing, in the light of ever lower supply voltages, low power consumption and high-frequency operation, is formed by the class of dynamic translinear circuits. This paper aims to give an overview of this young, yet rapidly developing, circuit paradigm. Emphasis is placed on methods for analysis and synthesis and on state-of-the-art results obtained for both linear and non-linear applications.

I Introduction

Nowadays, the area of analogue integrated filters is facing some serious challenges due to the ongoing trends to low supply voltages, low power consumption and high-frequency operation [1–3]. The situation is even complicated by the fact that the transfer function of many circuits has to be tuneable.

A promising approach to face these challenges is given by the class of translinear (TL), log-domain or exponential state-space (ESS) filters, which are consequently receiving increasing interest [4–12]. These circuits are based on the dynamic TL principle [13], which is a generalisation of the conventional, i.e. static, TL principle, proposed by Gilbert [14–16]. Application areas where dynamic TL circuits can be successfully employed include audio filters [17], fibre optic front-ends, high-frequency filters [18], RF on-chip transmitters, high-frequency oscillators [9, 19] and low-voltage ultra-low-power applications [20, 21].

Translinear filters are inherently companding [5] (the voltage swings are logarithmically related to the currents), which is beneficial with respect to the dynamic range in low-voltage low-power environments [22–24]. In addition, TL filters are easily implemented in class AB [5, 25–27], which entails a larger dynamic range and a reduced average current consumption. Further, owing to the small voltage swings, TL filters facilitate relatively wide bandwidth operation. At high frequencies though, considerable care has to be taken regarding the influence of parasitic capacitances and resistances, which affect the exponential behaviour of the transistor.

Translinear filters are excellently tuneable across a wide range of several parameters, such as cut-off frequency, quality factor and gain, which increases their designability and makes them attractive to be used as standard cells or programmable building blocks.

The application of the dynamic TL principle is not limited to filters, i.e., linear differential equations (DE). The principle can also be applied to the structured design of non-linear DEs, e.g., oscillators [9, 19], RMS-DC converters [13], PLLs [28] and even chaos.

This paper aims to give an overview of the complete field of dynamic TL circuits. The emphasis is on structured design methods and principles, rather than on specific circuit implementations. The static and dynamic TL principles are reviewed in Sec. II. Section III gives an overview of analysis methods. The general class of TL filters contains several different types. In Sec. IV, the correspondences and differences between log-domain, tanh, sinh and $\sqrt{}$-domain filters are treated. Section V presents several synthesis methods. Finally, an overview of state-of-the-art results that have been obtained thus far is presented in Sec. VI.

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II Design principles

Translinear circuits can be divided into two major groups: static and dynamic TL circuits. The first group can be applied to realise a wide variety of linear and non-linear static transfer functions. All kinds of frequency-dependent functions can be implemented by circuits of the second group. The underlying principles of static and dynamic TL circuits are reviewed in this section.

Static translinear principle Translinear circuits are based on the exponential relation between voltage and current, characteristic for the bipolar transistor and the MOS transistor in the weak inversion region . The collector current I_C of a bipolar transistor in the active region is given by:

$$I_C = I_s e^{V_{BE}/U_T}, \tag{1}$$

where all symbols have their usual meaning.

The TL principle applies to loops of semiconductor junctions. A TL loop is characterised by an even number of junctions [15, 29]. The number of devices with a clockwise orientation equals the number of counter-clockwise oriented devices. An example of a four-transistor TL loop is shown in Fig. 1. It is assumed that the transistors are somehow biased at the collector currents I_1 through I_4. When all devices operate at the same temperature, this yields the familiar representation of TL loops in terms of products of currents:

$$I_1 I_3 = I_2 I_4. \tag{2}$$

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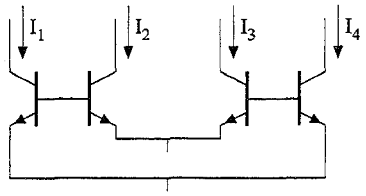

Figure 1: A four-transistor translinear loop.

This generic TL equation is the basis for a wide variety of static electronic functions, which are theoretically temperature and process independent.

Dynamic translinear principle The static TL principle is limited to frequency-independent transfer functions. By admitting capacitors in the TL loops, the TL principle can be generalised to include frequency dependent transfer functions. The term 'Dynamic Translinear' was coined in [13] to describe the resulting class of circuits. In contrast to other names proposed in literature, such as 'log-domain' [4], 'companding current-mode' [5], 'exponential state-space' [25], this term emphasises the TL nature of these circuits, which is a distinct advantage with respect to structured analysis and synthesis.

The dynamic TL principle can be explained with reference to the sub-circuit shown in Fig. 2. Using the current-mode approach, this circuit is described in terms of the collector current I_C and the capacitance I_{cap} flowing through the capacitance C. Note that the dc voltage source V_{const} does not affect I_{cap}. An expression for I_{cap} can be derived from the time derivative of (1) [12,29]:

$$I_{cap} = CU_T \frac{\dot{I}_C}{I_C}, \tag{3}$$

where the dot represents differentiation with respect to time.

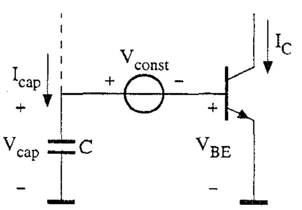

Figure 2: Principle of dynamic translinear circuits.

Equation (3) shows that I_{cap} is a non-linear function of I_C and its time derivative $\dot{I}_C$. More insight in (3) is obtained by slightly rewriting it:

$$CU_T \dot{I}_C = I_{cap} I_C. \tag{4}$$

This equation directly states the dynamic translinear principle: *A time derivative of a current can be mapped onto a product of currents.* At this point, the conventional TL principle comes into play, for, the product of currents on the right-hand side (RHS) of (4) can be realised very elegantly by means of this principle. Thus, the implementation of (part of) a DE becomes equivalent to the implementation of a product of currents.

The dynamic TL principle can be used to implement a wide variety of DEs, describing signal processing functions. For example, filters are described by linear DEs. Examples of non-linear DEs are harmonic and chaotic oscillators, PLLs and RMS-DC converters.

III Analysis

In almost all publications on dynamic translinear circuits, the emphasis has been on synthesis. Both structured design methods and new circuit realisations have been presented. Although synthesis is more powerful than analysis, it must go together with a generally applicable analysis method in the same domain. Only when this condition is met, the full potentials of a synthesis method can be exploited.

In this section, an overview is given of the analysis methods proposed in literature. It is shown that a current-mode TL approach yields a general analysis method, which is next elaborated to facilitate state-space analysis of TL filters.

Inverse transformation In [4], Adams not only presented a synthesis method, but also proposed an analysis method. The first step is to write down the node equations from the large-signal ac model of the filter. These equations are multiplied by exponential functions to eliminate the isolated derivatives. Next, the intermediate voltages have to be eliminated, such that a single equation results, expressing the relation between the compressed input and output voltage. Unfortunately, according to Adams, no systematic method might exist for this step [4]. In the last analysis step, a voltage-mode linear DE is obtained from this single equation by applying a logarithmic transformation; the inverse of the exponential transformation used during synthesis.

Implicitly, Adams' method has been applied in a number of publications to verify parts of transistor level implementations, however, never to analyse a complete higher-order TL filter.

Small-signal analysis A very simple way to calculate the transfer function of a complete filter is to analyse the small-signal equivalent circuit, see e.g. [7]. Since, by definition, a small-signal analysis results in a linear transfer function, this method yields the correct expression only when the dynamic TL circuit under consideration is globally linear and properly designed. The large-signal linearity cannot be proven and has to be verified in another way. Numerical simulations can provide some insight.

Global translinear analysis Recently, a large-signal analysis method was presented by the authors [12]. This current-mode method is based on a TL approach and is believed to be completely general. It has been tested with success on all published log-domain, tanh and sinh filters.

The key to the analysis of dynamic TL circuits is formed by the capacitance currents. Basically, the only difference between static and dynamic TL circuits is the presence of some capacitances.

Static TL circuits can be analysed through the method described in [29]. The first step is to express all collector currents in terms of the current sources, which are connected to the nodes of the TL core. The collector currents are linear combinations of the input, dc bias and output currents, and of some intermediate currents in case of multiple-loop circuits.

Once the collector currents are found, the TL loop equations are derived. These are given by equations like (2).

The last analysis step is to solve the system of TL loop equations for the output current(s) by eliminating the intermediate currents.

In dynamic TL circuits, some capacitors are connected to the nodes of the TL core. Consequently, the node currents are also determined by the currents flowing through these capacitors and the capacitance currents will appear in the TL loop equations. From this point of view, the capacitors can be regarded as being a special kind of current source.

To solve the system of loop equations, the capacitance currents will have to be eliminated [12]. To this end, expressions for the capacitance currents have to be found. This is quite simple. A capacitance connected to a node of the TL core will always form a loop with one or more base-emitter junction in series. This is illustrated in Fig. 3. The capacitance voltage V_{cap} can be expressed in

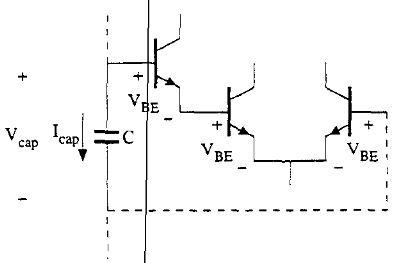

Figure 3: A capacitance in (a part of) a translinear loop.

terms of the base-emitter voltages, which in turn are expressed in terms of the collector currents flowing through these transistors. The capacitance current I_{cap} can now be calculated from the constitutive law by taking the derivative of V_{cap} with respect to time. Thus, a very simple equation is obtained [12]:

$$I_{cap} = CU_T \sum_i \pm \frac{\dot{I}_{C,i}}{I_{C,i}}. \tag{5}$$

The $\pm$ sign of each term depends on the orientation of the corresponding transistor. This equation can be applied to each capacitance in the circuit to find an expression for the current flowing through it.

An example of the application of the proposed analysis method can be found in [12], where a second-order TL low-pass filter, designed by Frey [6], is analysed.

State-space translinear analysis The state-space method can be used to break down a high-order DE into a system of first-order DEs. The method can also be used beneficially in TL filters to limit intermediate expression swell, a disadvantage of the global analysis method described above, thus obtaining linear equations at an earlier stage of the analysis.

In order to find a state-space description of a filter, it is necessary to choose state variables. For TL filters, using the capacitance voltages as the state variables is very inconvenient. Since TL filters are instantaneously companding, a better choice is to use the currents obtained from an exponential voltage-to-current expansion of the capacitance voltages, applying the exponential law describing the bipolar transistor.

As an example, regard the second-order low-pass Butterworth filter shown in Fig. 4 [30]. In this filter, ex-

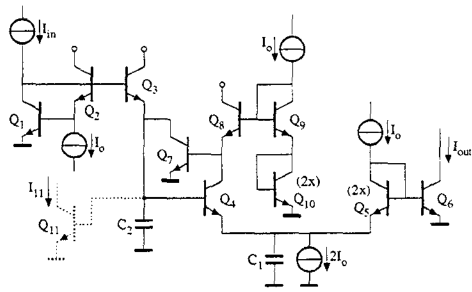

Figure 4: Sensing the states of a translinear filter.

pansion of the voltage V_{C_1} across capacitance C_1 is already implemented by means of transistor Q_6. Therefore, the collector current I_{out} of Q_6 is chosen as the first state variable. Note that Q_5 merely acts as a dc voltage source.

The voltage V_{C_2} across the second capacitance C_2 is not expanded within the filter, but this can be accomplished by adding a *fictitious* sense transistor Q_{11}, as shown in Fig. 4. The collector current I_{11} of this transistor is the second state variable to be used.

The actual filter shown in Fig. 4 consists of two disjunct TL loops: $Q_1 - Q_6$ and $Q_7 - Q_{10}$. By adding the sense transistor Q_{11}, the first loop, $Q_1 - Q_6$, is broken into two coupled second-order loops, i.e. $Q_1 - Q_3, Q_{11}$ and $Q_{11}, Q_4 - Q_6$. Now, the filter can be described by a set of three loop equations and two expressions for the capacitance currents I_{C_1} and I_{C_2}, given by:

$$I_{in} I_o - (I_7 + I_{C_2}) I_{11} = 0, \tag{6}$$

$$I_{11} I_o - 2 (I_o + I_{C_1}) I_{out} = 0, \tag{7}$$

$$2 I_7 (I_o + I_{C_1}) - I_o^2 = 0, \tag{8}$$

$$I_{C_1} = CU_T \frac{\dot{I}_{out}}{I_{out}}, \tag{9}$$

$$I_{C_2} = CU_T \frac{\dot{I}_{11}}{I_{11}}, \tag{10}$$

where I_7 is the collector current of Q_7; an intermediate current. The factors of 2 in (7) and (8) are due to the doubled emitter areas of Q_5 and Q_{10}. All second-order effects have been neglected in this set of equations.

Next, we have to find expressions for $\dot{I}_{out}$ and $\dot{I}_{11}$ in terms of I_{in}, I_{out} and I_{11}. An equation for $\dot{I}_{out}$ is found by eliminating I_{C_1} from (7) by applying (9). This yields:

$$CU_T\dot{I}_{out} = I_o\left(\tfrac{1}{2}I_{11} - I_{out}\right). \tag{11}$$

To find an expression for $\dot{I}_{11}$, we first eliminate I_7 from (6) and (8). From the resulting equation, the capacitance currents I_{C_1} and I_{C_2} can be eliminated using (9) and (10), after the derivative $\dot{I}_{out}$ has been eliminated from (9) by using (11). This yields the second equation of the state-space description:

$$CU_T\dot{I}_{11} = I_o\left(I_{in} - I_{out}\right). \tag{12}$$

Thus, a complete current-mode state-space description of the TL filter shown in Fig. 4 is given by (11) and (12).

IV log, tanh, sinh and $\surd$-domain circuits

Within the general class of dynamic translinear circuits several different types of ESS filters have been proposed. Next to the most prevalent class of log-domain filters, the two classes of tanh and sinh filters have been proposed by Frey [25]. The extension of the underlying principles of TL filters to MOS transistors operating in the strong inversion region was proposed independently by Mulder *et al.* [31] and Eskiyerli *et al.* [32, 33]. The resulting circuits are called $\surd$-domain filters.

In this section, we describe the correspondences and differences between log-domain, tanh, sinh and $\surd$-domain filters. The characteristics of these classes can be derived from their generic output structures, which are depicted in Fig. 5.

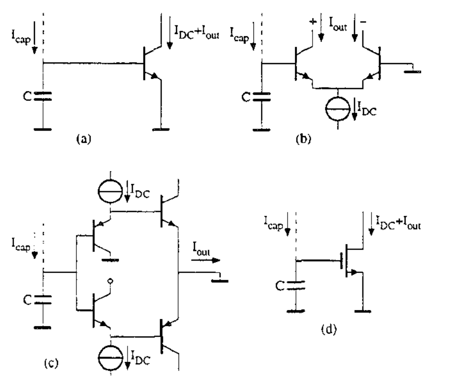

Figure 5: Generic output structures of a) log-domain, b) tanh, c) sinh, and d) $\surd$-domain filters.

log-domain filters Most published dynamic TL circuits are based on the single transistor output structure shown in Fig. 5(a), characteristic for the class of log-domain filters. The transfer function from the capacitance voltage V_{cap} to the output current I_{out} is given by the well-known exponential law (1).

The most important characteristic of a dynamic TL output structure is the current-mode expression for the capacitance current I_{cap}. For log-domain filters, I_{cap} is given by (3), where $I_C = I_{DC} + I_{out}$. As was shown in Sec. II, a linear derivative $\dot{I}_{out}$ is obtained by multiplying I_{cap} by $I_{DC} + I_{out}$.

Typically, log-domain filters operate in class A. The actual ac signal I_{out} is superposed on a dc bias current I_{DC}. As a consequence, the output signal swing is limited to $I_{out} > -I_{DC}$. Note that this limitation is single sided, which is advantageous if a-symmetrical input wave forms have to be processed. This characteristic can be exploited to enable class AB operation [5, 26]. Using a class AB set-up, see Fig. 6, the dynamic range can be enlarged without increasing the quiescent power consumption. Using a current splitter, the input current I_{in} is divided into two currents I_{in1} and I_{in2}, which are both strictly positive, and related to I_{in} by: $I_{in} = I_{in1} - I_{in2}$. The current splitter impresses a constant geometric or harmonic mean on I_{in1} and I_{in2}. Next, I_{in1} and I_{in2} can be processed by two class A log-domain filters.

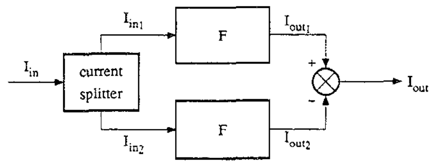

Figure 6: Set-up for class AB operation.

tanh filters Instead of a single transistor in common-emitter configuration, the class of tanh filters is characterised by a differential pair output structure, shown in Fig. 5(b). The name of this class of filters is derived from the well-known hyperbolic tangent V-I transfer function.

The tail current of the differential pair is a dc current I_{DC}, and therefore, tanh filters also operate in class A. The output current I_{out} is the difference of the two collector currents. The output swing is limited to $-I_{DC} < I_{out} < I_{DC}$. Since this interval is symmetrical, the class AB set-up shown in Fig. 6 cannot be applied to tanh filters. Alternatively, the dynamic range of a tanh filter can be enlarged using syllabic companding [34].

Using (5), the capacitance current I_{cap} is found to be:

$$I_{cap} = CU_T\left(\frac{\dot{I}_{out}}{I_{DC} + I_{out}} - \frac{-\dot{I}_{out}}{I_{DC} - I_{out}}\right). \tag{13}$$

A linear derivative $\dot{I}_{out}$ is obtained by multiplying this equation by $(I_{DC} + I_{out})(I_{DC} - I_{out})$:

$$2CU_T I_{DC}\dot{I}_{out} = I_{cap}(I_{DC} + I_{out})(I_{DC} - I_{out}). \tag{14}$$

Comparing (4) and (14), we can see that the RHS of (14) is third order, whereas the RHS of (4) is only second order. Consequently, in general, TL loops of higher order are necessary to implement a tanh filter, resulting in a more complex circuit.

sinh filters The third type of ESS filters is the class of sinh filters [25, 27, 35]. The V-I transfer function of the output structure shown in Fig. 5(c) is a hyperbolic sine function. The output structure is a complete second-order TL loop. It implements the geometric mean function $I_{DC}^2 = I_{out1}I_{out2}$, where I_{out1} and I_{out2} are the collector currents of the two right-most transistors shown in Fig. 5(c). The output current I_{out} is the difference of I_{out1} and I_{out2}. Since both I_{out1} and I_{out2} are always positive, the sinh output structure operates in class AB, which is beneficial with respect to the dynamic range [25, 27, 35].

The current-mode expression for the capacitance current I_{cap} is given by:

$$I_{cap} = CU_T \frac{\dot{I}_{out}}{I_{out1} + I_{out2}}. \tag{15}$$

A linear derivative $\dot{I}_{out}$ is obtained by multiplying I_{cap} by the sum $I_{out1} + I_{out2}$. It is interesting to note that the voltage V_{cap} and the current $I_{out1} + I_{out2}$ are related through a hyperbolic cosine function.

$\sqrt{}$-domain filters The static and dynamic TL principle can be implemented using MOS transistors by operating them in the weak inversion region. However, weak inversion operation is limited to low frequencies. Therefore, several authors have proposed to generalise the underlying principles of dynamic TL filters to MOS transistors operating in strong inversion [31–33, 35], which are characterised by the well-known square law model.

Replacing the bipolar transistor shown in Fig. 5(a) by a MOS transistor, we obtain the sub-circuit depicted in Fig. 5(d), which is also characterised by a non-linear relation between the capacitance current I_{cap} and the output current I_{out}:

$$I_{cap} = \frac{C}{\sqrt{2\beta}} \frac{\dot{I}_{out}}{\sqrt{I_{out}}}, \tag{16}$$

where β is the transconductance factor of the MOS transistor.

Obviously, in this case, I_{cap} has to be multiplied by $\sqrt{I_{out}}$ to obtain a linear derivative $\dot{I}_{out}$. Based on this relation, two similar $\sqrt{}$-domain integrator circuits were proposed independently in [31] and [33].

V Synthesis methods

Several synthesis methods for translinear filters have been proposed in literature. An overview of these methods is given in this section. The synthesis methods have been classified into three groups. The first is based on exponential transformations of state-space filter descriptions. The second group is based on component substitution of LC or g_mC prototype filters. The last method follows a current-mode TL approach.

Exponential transformations Synthesis of TL filters using exponential transformations was introduced by Adams in [4] and generalised to filters of arbitrary order by Frey [6]. The design of a TL filter begins with a state-space filter description:

$$\dot{\vec{x}} = A\vec{x} + BU, \qquad Y = C\vec{x} + DU, \tag{17}$$

where $\vec{x} = (x_1, \ldots, x_n)^T$ is the state vector, A, B, C and D are matrices, U is the input signal and Y the output signal.

Next, the state vector $\vec{x}$ and the input signal U are transformed to the new vector $\vec{v}$ and input u using the exponential function [6]:

$$x_i = e^{v_i/U_T}, \qquad U = I_{DC}e^{u/U_T}. \tag{18}$$

This procedure is only valid if both x_i and U are strictly positive. This restriction is satisfied by adding a dc component to U and applying linear transformations to (17), through trial-and-error, such that $x_i > 0 \; \forall i$ in the resulting state-space description [6].

Thus, a set of equations results describing the TL filter in term of voltages and exponential functions. These equations are interpreted as the nodal equations of the TL filter. The circuit implementation has to be derived directly, or possibly after some rearranging [25, 35], from these equations.

Next to the exponential function, compound exponential functions like the hyperbolic tangent and hyperbolic sine function can be used to transform the state-space description [25, 35].

Component substitution Another approach to the design of TL filters is based on component substitution of prototype LC [8, 36] or g_mC filters [10, 37]. The general idea is to replace elements from the prototype filter by parts of TL loops. In [10, 37], the transconductances are replaced by a single transistor and a level shift; the resistors are replaced by dc current sources; the capacitors remain the same.

Application of these component substitution based synthesis methods is simple. Yet, an important disadvantage seems that the designer cannot make any choices along the synthesis path. In general, for each LC or g_mC prototype filter, exactly one TL filter results. Consequently, a designer has little control over the specifications a TL filter has to meet.

Translinear synthesis In [5], two integrators were designed using a current-mode TL approach. The underlying method was generalised to filters of arbitrary order in [38]. Whereas the previously described synthesis methods seem limited to the design of linear filters, the method presented in this section can also be applied to the structured design of non-linear dynamic functions. Another major advantage is that this method fits directly onto the synthesis method for conventional TL circuits described in [29]. Consequently, all existing theory and experience on static TL circuits can be employed in the design of dynamic TL circuits.

Synthesis of a dynamic circuit, be it linear or nonlinear, starts with a DE describing its function. Often, it is more convenient to use a state-space description, which is mathematically equivalent. The structured synthesis method for dynamic TL circuits is illustrated here by the design of a second-order low-pass filter with a Q of two, described by:

$$\ddot{z} + \dot{z}/2 + z = x, \tag{19}$$

where x and z are the input and output signal, respectively, and the dot represents differentiation with respect to the dimensionless time τ.

In the pure mathematical domain, equations are dimensionless. However, as soon as we enter the electronics domain to find an implementation of the equation, we are bound to quantities having dimensions. In the case of TL circuits, all time-varying signals in the DE (input signal, output signal and tunable parameters) have to be transformed into currents. For (19), x and z can be transformed into the currents I_{in} and I_{out} through the relations: $x = I_{in}/I_o$, $z = I_{out}/I_o$. The dimensionless time τ has to be transformed into the time t with its usual dimension $[s]$, using the equivalence relation given by:

$$\frac{\partial}{\partial \tau} = \frac{CU_T}{I_o}\frac{\partial}{\partial t}. \tag{20}$$

The presence of the current I_o in this expression explains the excellent linear tunability of TL filters.

Using these transformations, (19) becomes:

$$2C^2U_T^2\ddot{I}_{out} + CU_TI_o\dot{I}_{out} + 2I_o^2I_{out} = 2I_o^2I_{in}. \tag{21}$$

Conventional TL circuits are described by multivariate polynomials, in which all variables are currents. The gap between these current-mode polynomials and the DE can be bridged by the introduction of capacitance currents. For, the dynamic TL principle states that a derivative can be replaced by a product of currents.

The capacitance currents can be introduced simply by defining them. To this end, the general equation of a TL capacitance current, given by (5), can be used. This equation has two important characteristics. First, the denominators on the RHS are collector currents. This implies that these currents have to be strictly positive. Second, the numerators on the RHS are the time derivatives of the denominators.

With these characteristics in mind, we can define the capacitance currents, one by one. As the capacitance currents will be used to eliminate the derivatives from the DE, in the definitions of these currents, the derivatives present in the DE have to be used. In the definition of the first capacitance current I_{C_1}, the currents with the highest order of derivative have to be accommodated. In (21), this is the output current I_{out}. A possible definition of I_{C_1} is given by:

$$I_{C_1} = CU_T\frac{\dot{I}_{out}}{I_{out}}. \tag{22}$$

Note that this equation can be implemented using the substructure depicted in Fig. 2, when $I_C = I_{out}$. The denominator of (22) is strictly positive if the filter operates in class A.

From (22), expressions can be obtained for $\dot{I}_{out}$ and for $\ddot{I}_{out}$. If the resulting equations are used to eliminate $\ddot{I}_{out}$ and $\dot{I}_{out}$ from (21), a *first-order* DE in I_{C_1} results:

$$I_{out}\left(2CU_T\dot{I}_{C_1} + 2I_{C_1}^2 + I_{C_1}I_o + 2I_o^2\right) = 2I_o^2I_{in}. \tag{23}$$

The same process is repeated for I_{C_2}, the second capacitance current introduced. Since $\dot{I}_{C_1}$ is the only derivative in (23), I_{C_1} has to be present in the definition of I_{C_2}. A possible definition is:

$$I_{C_2} = CU_T\left(\frac{\dot{I}_{out}}{I_{out}} + \frac{\dot{I}_{C_1}}{2I_o + I_{C_1}}\right). \tag{24}$$

Using this equation to eliminate $\dot{I}_{C_1}$ from (23), a polynomial (a degenerate DE) is obtained:

$$I_{out}\left(I_{C_1}(2I_{C_2} - 3I_o) + 4I_{C_2}I_o + 2I_o^2\right) = 2I_o^2I_{in}. \tag{25}$$

From this point on, the synthesis theory for static TL circuits can be used [29], since both sides of (25) are now described by a current-mode multivariant polynomial. The next synthesis step is TL decomposition. The polynomial has to be mapped on the loop equations given by (2). For example, a possible parametric decomposition of (25) is given by:

$$I_{out}(2I_o + I_{C_1})p = I_o^2I_{in}, \tag{26}$$

$$(2I_o + I_{C_1})(p - I_{C_2} + 3I_o/2) = 4I_o^2. \tag{27}$$

The last synthesis step is biasing. The TL decomposition has to be mapped on a TL circuit topology and the correct collector currents have to be forced through the transistors. Biasing methods for bipolar all-NPN TL topologies are presented in [29]. Additional biasing methods include the use of (vertical) PNPs, compound transistors or (simple) nullor implementations. If subthreshold MOSTs are used, some additional possibilities are the application of the back gate [39] and operation in the triode region [40,41]. An all-NPN biasing scheme was designed for the parametric TL decomposition given by (26) and (27). The resulting prototype circuit is depicted in Fig. 7. After biasing, the prototype circuit can be analysed for second-order effects. At this stage, an analysis method in the same domain is indispensable.

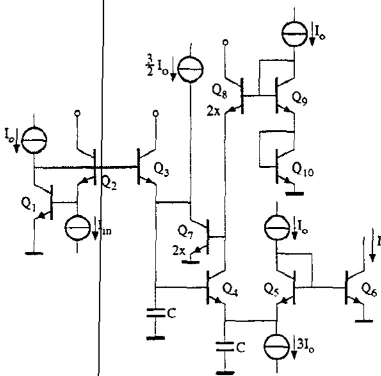

Figure 7: Prototype second-order low-pass translinear filter.

VI State-of-the-art

Many skilfully designed dynamic TL circuits have been presented in literature. Due to the limited space available, the state-of-the-art overview given in this section is limited to those designs that have been experimentally verified. To the authors' knowledge, this amounts to a total of seven different TL filter designs [8, 17, 18, 20, 37, 42–47] The specifications of these filters are summarised in table 1. Values between round brackets were calculated or estimated from the data presented in the cited publications.

Probably the most important filter specification is the dynamic range (DR). Although the area of TL filters is still quite young, the results already compare well with the specifications that are typically obtained using g_mC filters, i.e. 40-70 dB [1].

Comparing the DR specification of the different TL filters in table 1 is not easy, since the DR is influenced by many factors, e.g. power consumption and total capacitance. It is however interesting to note the difference in DR between class A and class AB operated TL filters. The first four filters [8, 17, 18, 20, 42, 43] are biased in class A, the last three [37, 44–47] operate in class AB, which explains why the latter have structurally better DR specifications.

The dynamic TL principle is not limited to the implementation of filters. Non-linear DEs can be realised just as well. As of today, the list of experimentally verified designs is still quite short. Only four designs are known to the authors: two oscillators [19, 48], an RMS-DC converter [49] and a PLL [28]. The interested readers are referred to these papers for measurement results.

With the increasing number of researchers of dynamic TL circuits, we can expect to see more and more realisations, both for linear and non-linear applications. It will be interesting to see to what extend the specifications obtained thus far can be improved in the future.

VII Conclusions

Dynamic translinear circuits constitute an exciting new approach to the integration of analogue signal processing functions. These circuits might prove to be the best approach to face the dynamic range limitations, conventional integrated circuits are facing due to low-power, low-voltage and high-frequency demands. This field is receiving increasing interest and encouraging results have been obtained thus far. The future will learn to what extend these results can be improved.

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Table 1: Comparison of experimentally obtained TL filter specifications.

	[8,42]	[20]	[18]	[17,43]	[37,44,45]	[46]	[47]
Process	Bipolar	0.8μ BiCMOS	Bipolar	Bipolar	2μ BiCMOS	1μ BiCMOS	Bipolar
Filter	LPF, 5	LPF, 4	BPF[Q:3.6-66], 5	LPF, 2	LPF, 3	LPF, 3	LPF, 1
f_c [Hz]	40k	100 - 10k	125M - 430M	1.6k - 8k	10k -100k	10k - 15M	1k - 8k
DR [dB]	52	60	(55)	57	-	65 @ 320 kHz	73
Total C [pF]	-	55	18	100	500	59	100
Power [W]	(25m)	1μ	81m @ 430 MHz	6μ	180μ	65μ @ 320 kHz	2μ
Supply [V]	10	5	2.7	1	4	0.9	0.95

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