Advances in Low-Voltage Ultra-Low-Power Analog Circuit Design

Wouter A. Serdijn, Jan Mulder, Daniel Rocha and Luís Cléber C. Marques

Abstract— *In this paper it is argued that there are good reasons to choose current as the information-carrying quantity in the case of low-voltage ultra-low-power design constraints. This paper focuses on the influence of transfer quality on that choice. To obtain power-efficient transfer quality, indirect feedback is shown to be a good alternative to traditional feedback techniques. Two recently developed analog circuit techniques that both operate in the current domain and employ indirect feedback are described, being the continuous-time dynamic-translinear technique and the discrete-time switched-MOSFET technique.*

I. Introduction

Low-voltage circuit techniques are applied in the area of battery-operated systems. For portability reasons, the size of the equipment must be small, which necessitates the maximum integration of the signal processing circuitry. However, as the size of batteries is now becoming the limiting factor, the reduction of the power dissipation has become an extra design constraint. As a consequence, the key point is to develop, simultaneously, both low-voltage (i.e. 1 – 1.5 V) and low-current (i.e. < 1 mA) operating integrated circuits in order to reduce the battery size.

Another design criterion that must be fulfilled is transfer quality. This quality is influenced by two different kinds of errors: stochastic ones and systematic ones. By stochastic errors we mean inaccuracies in the input-output relation caused by noise or interference. Though impossible to eliminate, their influence can be minimized by a proper design strategy.

Systematic errors arise from network imperfections, such as offset, non-linearity, inaccuracy of the device parameters, drift and temperature dependence. Probably the most effective method to reduce their influence, and thus to obtain an accurate transfer function, is by means of applying negative feedback, which allows us to exchange the large gain provided by the (highly non-linear) active devices for quality provided by (usually linear) passive devices.

Design strategies for the reduction of stochastic errors and systematic errors are normally not consistent with design strategies which take into account power dissipation, voltage range and current range. Therefore, it is the combination of transfer quality, low voltage and low power that must be considered during the whole design process.

II. Indirect feedback

As mentioned above, systematic errors can be reduced by means of negative feedback. Fig. 1 shows the four basic ways of applying (single-loop) direct feedback by means of two two-ports. If all the transfer parameters of two-port H approach infinity, i.e., H is a nullor, the output signal (v_L or i_L) is related to the input signal (v_S or i_S) as the inverse transfer function of the feedback network T_f.

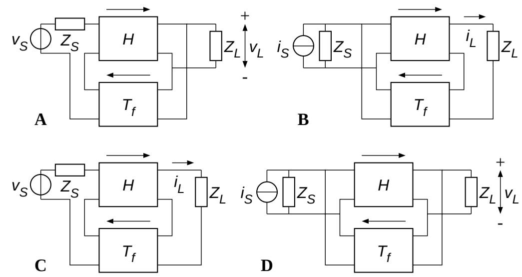

Fig. 1. Four basic direct negative-feedback amplifiers: a voltage amplifier (A), a current amplifier (B), a transconductance amplifier (C) and a transimpedance amplifier (D).

In low-voltage circuits, however, due to the restricted voltage swing, it is often not possible, or at least not preferable, to connect two ports of these two-port networks in series, thus to sense the output current or to compare the input voltage of a circuit directly. This occurs in configurations A (at the input), B (at the output) and C (at both input and output). Hence, all direct-feedback configurations, except the transimpedance amplifier (configuration D), are less suited for low-voltage applications.

To realize voltage, current and transconductance amplifiers, a useful alternative to direct negative feedback may be a technique called *indirect negative feedback* [1]. In an indirect-negative-feedback circuit, the output and/or the input stage is copied, so that it has an equivalent input-output relation, and the feedback signal is taken from and/or fed back to that copy. Thus, it is possible to obtain a circuit response which is determined by the feedback network only, assuming that the copying does not introduce errors. A voltage amplifier, a current amplifier and a transconductance amplifier, all using the indirect negative-feedback principle, are depicted in Figures 2, 3 and 4. It can be seen that series-connected ports are now avoided in all configurations.

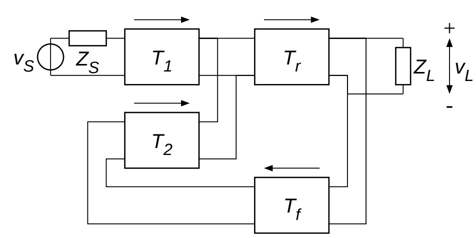

Fig. 2. A voltage amplifier with negative feedback and indirect voltage comparison.

Again, if all the transfer parameters of two-port T_r approach infinity, $T_2 = T_1$ and $T_4 = T_3$, the output signal (v_L or i_L) is related to the input signal (v_S or i_S) as the inverse transfer function of the feedback network T_f.

III. Processing in the current domain

We now investigate how applying indirect negative feedback relates to the choice of the electrical quantities inside

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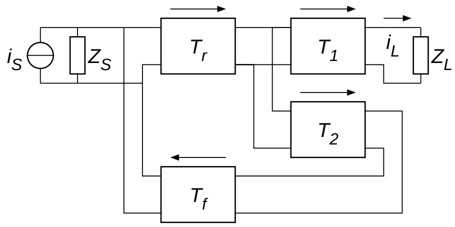

Fig. 3. A current amplifier with negative feedback and indirect current sensing.

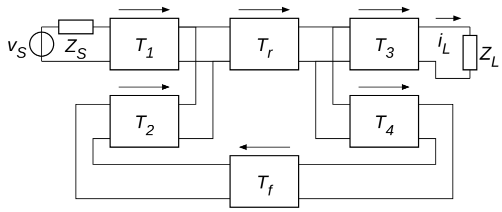

Fig. 4. A transconductance amplifier with negative feedback and indirect current sensing and indirect voltage comparison.

the system. In electronic circuits, indirect voltage comparison results in a doubled power density spectrum of the equivalent noise voltage at the input, because the outputs of the direct and indirect input stages are connected in parallel. Indirect current sensing results in a doubled power density spectrum of the noise current at the output, because the direct and indirect output are placed in parallel. In practice, often the noise is most critical at the input, so on that ground there may be a preference for current sensing and thus for current as the information-carrying quantity.

Another disadvantage of the use of voltage as the information-carrying quantity is that, when the circuits are "voltage-driven," i.e., from a low-impedance source, the equivalent input noise voltage is predominantly the result of the input noise voltage of both input stages. For bipolar transistors and CMOS transistors in weak inversion, this input noise voltage is inversely proportional to the bias (collector or drain) current, and thus, in order to obtain a low input noise voltage, these bias currents must be rather large. This, of course, is in sharp contrast with our low-power requirement.

When, however, the circuits are "current-driven," thus with a high impedance, the equivalent input noise current is mainly determined by the input noise current of the input stage. Since the equivalent input noise current of bipolar transistors is proportional to the bias current, this calls for small bias currents, which is in line with the low-power requirement. This favors the choice of current as the information-carrying quantity.

A third disadvantage of indirect voltage comparison is that, in order to compensate each other, the non-linearities of the two input stages must be symmetrical or opposite, because the sum of their output currents must be nullified by the nullor. In practice, this requires either two balanced input stages or two complementary stages in a complementary IC process. The use of two balanced input stages, since their input noise voltages are placed in series, again doubles the power density spectrum of the equivalent input noise voltage. A complementary IC process is often not available and, moreover, exact complementarity can never be accomplished.

Indirect feedback at the output, however, calls for two identical output stages, to compensate for the non-linearities. These can easily be made in any ordinary IC process. For this reason there again may be a preference for current sensing and thus for current as the information-carrying quantity.

Let us now address the influence of parasitic immitances. The influence of parasitic admittances in parallel with the signal path can be reduced by terminating the signal path with a low impedance. The parasitic admittances then have no voltages across their terminals and thus no current flows in them. The influence of parasitic impedances in series with the signal path can be reduced by terminating the signal path with a high impedance. Then no current flows in the parasitic impedances and thus there is no voltage across their terminals.

In low-power integrated circuits, often the parasitic admittances, i.e., the node capacitances, e.g., the transistors junction capacitances, due to their (non-linear) voltage dependency, have more influence on the signal behavior than the parasitic impedances, i.e., the branch inductances and resistances, e.g., the transistors' bulk resistances. Therefore it is convenient to terminate the signal paths with low impedances as much as possible. In this situation it is best to choose current as the information-carrying quantity.

This argument is also at the base of the popularity of various 'current-mode' techniques, of which it is rightly stated that they have an inherent ability to exhibit good high-frequency properties.

Finally, we have to consider the power supply. In practice, this power supply is a voltage source (battery), giving a limitation in voltage. The limitation in current is only indirectly given by a limitation in the power of the battery and might be less restricting than that of the voltage. This favors the choice of current as the information-carrying quantity. However, not using the total range of this supply voltage for signal swing gives rise to waste of power.

From the above discussion, it will be clear that for low-voltage ultra-low-power analog IC's the total design process must be considered, in which transfer quality plays a dominant role. Designs, based on the design principles shown above, confirm that current becomes more favorable than voltage as the information-carrying quantity and that indirect feedback is to be preferred in a low-voltage low-power environment.

In the following sections, two recently developed analog circuit techniques that both apply indirect negative feedback and process the information in the current domain are presented: the continuous-time *dynamic translinear* technique and the discrete-time *switched MOSFET* technique.

IV. Dynamic translinear circuits

Dynamic translinear (DTL) circuits, of which recently an all-encompassing current-mode analysis and synthesis theory has been developed in Delft [2], are based on the DTL principle, which can be regarded as a generalization of the well-known 'static' translinear principle, formulated by Gilbert in 1975 [3].

The first DTL circuit was originally introduced by Adams in 1979 [4], being a first-order lowpass filter. Although not recognized then, this was actually the first time a first-order linear differential equation was implemented using translinear (TL) circuit techniques. In 1990, Seevinck introduced a 'companding current-mode integrator' [5] and since then the principle of TL filtering has been extensively studied by Frey, Punzenberger and Enz, Toumazou et al., Roberts et al., Fox et al., Tsividis et al., Mulder and Serdijn, and others. See [2] for a detailed list of references.

However, the DTL principle is not limited to filters, i.e. linear differential equations. By using the DTL principle, it is possible to implement linear *and* nonlinear differential equations.

DTL circuits are based on two circuit principles, being the *static translinear principle* and the *dynamic translinear principle*. The first principle can be applied to realize a wide variety of linear and non-linear static transfer functions. All kinds of frequency-dependent functions can be implemented by additionally applying the second principle.

A. Static translinear principle

TL circuits are based on the exponential relation between voltage and current, characteristic for diodes, bipolar transistors and MOS transistors in the weak inversion region. In the following discussion, bipolar transistors are assumed. The collector current I_C of a bipolar transistor in the active region is given by:

$$I_C = I_S \mathrm{e}^{\mathrm{V_{BE}}/\mathrm{V_T}}, \tag{1}$$

where all symbols have their usual meaning.

The TL principle applies to loops of semiconductor junctions. A TL loop is characterized by an even number of junctions [3]. 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. 5. It is assumed that the transistors are somehow biased at the collector currents I_1 through I_4. When all devices are equivalent and operate at the same temperature, this yields the familiar representation of TL loops in terms of products of currents:

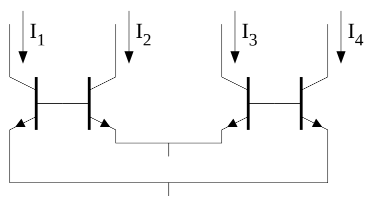

Fig. 5. A four-transistor translinear loop.

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

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

B. 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 generalized to include frequency-dependent transfer functions. The term 'Dynamic Translinear' was coined in [6] 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' [7], this term emphasizes the TL nature of these circuits, which is a distinct advantage with respect to structured analysis and synthesis.

The DTL principle can be explained with reference to the sub-circuit shown in Fig. 6. Using a current-mode approach, this circuit is described in terms of the collector current I_C and the current 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) [5, 6]:

$$I_{\mathrm{cap}} = C V_T \frac{\dot{I}_C}{I_C}, \tag{3}$$

where the dot represents differentiation with respect to time.

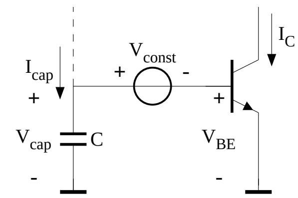

Fig. 6. Principle of dynamic translinear circuits.

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:

$$C V_T \dot{I}_C = I_{\mathrm{cap}} I_C. \tag{4}$$

This equation directly states the DTL 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, since the product of currents on the right-hand side (RHS) of (4) can be realized very elegantly by means of this principle. Thus, the implementation of (part of) a differential equation (DE) becomes equivalent to the implementation of a product of currents.

The DTL principle, in combination with the static 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.

V. SWITCHED MOSFET CIRCUITS

The switched MOSFET (SM) technique has recently been developed at UFSC, Brazil [8–10] and is a discrete-time technique suitable for low supply voltage operation as the MOSFET switches, in contrast to those required in switched capacitor or switched current circuits, do not suffer from the so-called "conduction gap."

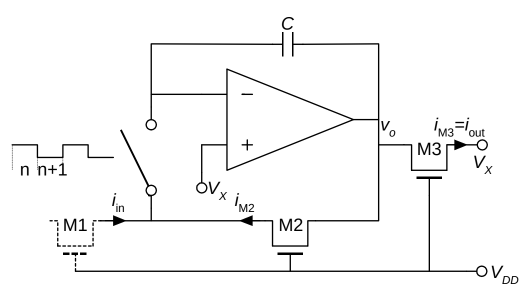

Fig. 7. Switched-MOSFET delay cell.

The basic delay cell of the SM technique is depicted in Fig. 7 [8]. Just like the basic switched-current cell, its operation can be analyzed by considering it to be a switched current mirror [11], albeit that in a SM delay cell the storage is performed by a transcapacitance instead of a capacitance. Its operation is as follows. When the switch is closed, assuming similar transistors, the transcapacitance, composed of capacitance C and the operational amplifier, integrates the sum of $i_{\rm in}$ and $i_{\rm M2}$, resulting in a voltage v_o. This voltage is transformed into similar currents by transistors M2 and M3. Due to the feedback operation the sum of $i_{\rm in}$ and $i_{\rm M2}$, in steady state, equals zero and since M2 and M3 are biased with the same set of voltages v_o, V_X, V_{DD}, it follows:

$$\frac{i_{\rm out}}{i_{\rm in}} = -\frac{(W/L)_{\rm M3}}{(W/L)_{\rm M2}} \tag{5}$$

Thus, the output current $i_{\rm out}$ becomes an inverted replica of the input current $i_{\rm in}$. When the switch opens, voltage v_o is stored at the output of the transcapacitance and both currents, $i_{\rm out}$ and $i_{\rm M2}$ are sustained.

The switch can be implemented by a single n-MOS transistor, since it operates at a constant voltage V_X which is close to V_{SS} and can be generated by the series connection of two identical transistors as shown in Fig. 8 [8]. This choice for V_X ensures equal output current swings in both directions and also guarantees that the switch operates outside the conduction gap.

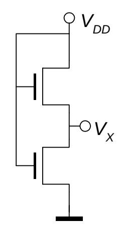

Fig. 8. Generation of voltage source V_X. Both transistors are identical.

Programmability of the above SM delay cell can be readily achieved by replacing either M2 or M3 by a MOSFET-only current divider (MOCD) as introduced by Bult and Geelen [12]. See Fig. 9. Applying a MOCD at the output of a SM delay cell yields a delay cell of which the output current $i_{\rm out}$ is a digitally controlled fraction of the delayed input current. Using this combination four times in a particular feedback configuration results in a universal SM biquad, as presented in [9], which, for the sake of brevity, is not discussed here. Employing the universal SM biquad, in turn, facilitates the implementation of all types of filter transfer funcions.

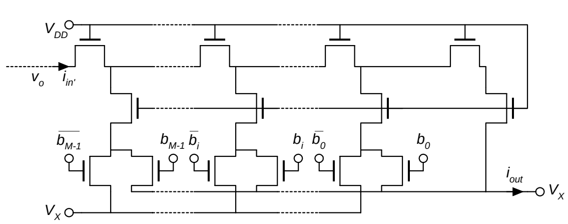

Fig. 9. Principle of a MOSFET-only current divider [12].

VI. Advantages of dynamic translinear and switched MOSFET circuits

Both DTL and SM circuits apply indirect negative feedback, process the information in the current domain and are implemented using transistors and capacitors only. Hence, a high functional density can be obtained, and the absence of large resistors makes them especially interesting for ultra-low-power applications.

DTL and SM circuits are inherently companding (the voltage swings are logarithmically related to the currents), which is beneficial with respect to the dynamic range in low-voltage environments. In addition, DTL and SM circuits are easily implemented in class AB, which entails a larger dynamic range and a reduced average current consumption. Further, owing to the processing in the current domain, DTL and SM circuits facilitate relatively wide bandwidth operation. At high frequencies though, considerable care has to be taken regarding the influence of parasitic resistances, which affect the exponential behavior of the transistors.

DTL and SM circuits are excellently tunable, by means of currents and MOCD's, respectively , 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.

Application areas where DTL and SM circuits can be successfully used include audio signal processing, radio-frequency transceivers, infra-red and fiber-optic front-ends and biomedical applications. The interested reader is referred to the open literature to find more information on these challenging and promising new circuit paradigms.

Acknowledgement

The authors would like to acknowledge the financial support of part of this work by CNPq, the Brazilian Research Council and by STW, the Dutch Technology Foundation.

References

[1] E.H. Nordholt: "Design of high-performance negative-feedback amplifiers," Elsevier, Amsterdam, 1983.

[2] J. Mulder, W.A. Serdijn, A.C. van der Woerd and A.H.M. van Roermund, "Dynamic translinear and log-domain circuits: analysis and synthesis," Kluwer Academic Publishers, Boston, 1998.

[3] B. Gilbert, "Translinear circuits: a proposed classification," Electronics Letters, Vol. 11, No. 1, January 1975, pp. 14-16.

[4] R.W. Adams, "Filtering in the log domain," Preprint No. 1470, presented at the 63rd AES Conference, New York, May 1979.

[5] E. Seevinck, "Companding current-mode integrator: a new circuit principle for continuous-time monolithic filters," Electronics Letters, 22nd November 1990, Vol. 26, No. 24, pp. 2046-2047

[6] J. Mulder, A.C. van der Woerd, W.A. Serdijn and A.H.M. van Roermund, "An RMS-DC converter based on the dynamical translinear principle," Proc. ESSCIRC'96, Neuchatel, Switzerland, 1996, pp. 312-315.

[7] Frey, D.R., "General class of current mode filters," Proceedings - IEEE International Symposium on Circuits and Systems, Vol. 2, 1993, pp. 1435-1437.

[8] R.T. Conçalves, S. Noceti F., M.C. Schneider and C. Galup-Montoro, "Digitally programmable switched current filters," Proc. ISCAS, Atlanta, Georgia, USA, 1996, pp. 258-261.

[9] L.C.C. Marques, C. Galup-Montoro, S. Noceti F. and M.C. Schneider, "Switched-MOSFET technique for programmable filters operating at low-voltage supply," Proc. SBMicro, Manaus, Brazil, 2000, pp. 18-24.

[10] F.A. Farag, C. Galup-Montoro and M.C. Schneider, "Digitally programmable switched-current FIR filter for low-voltage applications," IEEE J. Solid-State Circuits, Vol. 35, No. 4, April 2000, pp. 637-641.

[11] C. Toumazou, J.B. Hughes and N.C. Battersby (editors), "Switched currents: an analogue technique for digital technology," Peter Peregrinus, London, 1993.

[12] K. Bult and G.J.G.M. Geelen, "An inherently linear and compact MOST-only current division technique," IEEE J. Solid-State Circuits, Vol. 27, No. 12, December 1992, pp. 1730-1735.