Noise Considerations for Translinear Filters

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(Invited Paper)

***Abstract*—Translinear filters exhibit unconventional noise characteristics, due to the internal nonlinear behavior. In this paper, the maximal signal-to-noise-ratio and the dynamic range properties of translinear filters are discussed, and the relation between these two specifications and the inherent instantaneous companding is described.**

***Index Terms*— Companding, continuous-time filters, noise, translinear.**

I. Introduction

THE dynamic translinear (DTL) principle, or "log-domain" principle, constitutes a promising new implementation technique for linear filters [1]–[6]. Using the DTL principle, which states that a linear time derivative of a collector current can be realized through a product of a capacitance current and one or more collector currents [6], a wide variety of linear and nonlinear differential equations (DE's) can be implemented.

The dynamic range (DR) and the maximal signal-to-noise ratio (SNR), defined in Section II, are important specifications for linear filters. The noise behavior of TL filters is, however, different from conventional filter implementation techniques, due to the explicit dependence on the exponential voltage-to-current $(V–I)$ transfer function of the bipolar transistor.

This paper discusses the noise properties of TL filters. In Section III, the DR specifications of opamp-MOSFET-C, g_mC, and TL filters are compared. The companding nature of TL filters, which is related to both the DR and the maximal SNR is discussed in Section IV. Finally, the maximal SNR of TL filters is treated in Section V.

II. Definitions of DR and (Maximal) SNR

In the literature, several definitions for the DR and the maximal SNR are commonly used. Therefore, to clarify the discussion, this section explicitly defines these quantities.

By definition:

- the (maximal) SNR equals the (maximal) ratio of the signal power to the noise power *at the same time*;
- the DR equals the ratio of the maximal signal power to the minimum acceptable signal power; the latter is usually taken as equal to the noise power in the absence of any signals, and this convention will be adopted in this paper.

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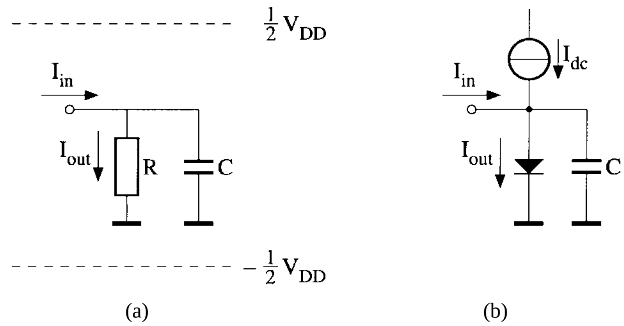

Fig. 1. Comparison of the dynamic range properties of (a) RC and (b) diode-C filter sections.

In conventional filters, the DR and the maximal SNR are equal, since the noise floor is constant. Hence, the maximal SNR is obtained for the maximum value of the signal power, which is determined by an application-specific specification of the distortion level, e.g., <1% total harmonic distortion (THD).

As pointed out in [7], a single noise figure cannot adequately describe the noise behavior of TL filters, or companding filters in general. Due to signal × noise intermodulation, the maximal SNR can be much smaller than the DR.

III. Dynamic Range

Static translinear (STL) circuits do not have a good reputation with respect to noise. Since DTL circuits are in a way very similar to STL circuits [6], it is likely that they also inherit the noise characteristics. Therefore, it is interesting to compare the DR properties of g_mC and opamp-MOSFET-C filters, the most popular implementation techniques to date [8]–[10], with the DR specifications of TL filters. Note that a comparison is made with respect to the DR and not to the maximal SNR, thus, due to the definition of the DR, excluding the signal × noise intermodulation. This intermodulation is the topic of Section V.

The comparisons are made under the practically relevant restriction of a low supply voltage. Though DTL circuits find wider application than only linear DE's, in this section, the comparisons are restricted to linear filters.

To obtain an indication of the DR properties of TL, g_mC and opamp-MOSFET-C filters, we compare the RC and the diode-C subcircuits shown in Fig. 1. The DR of a complete filter is strongly related to the DR of these elementary building blocks. For both filter sections, the low-pass transfer function $I_{\text{in}} \rightarrow I_{\text{out}}$ is considered. Under certain presumptions, to be discussed later, the circuit shown in Fig. 1(a) is representative for opamp-MOSFET-C filters. The circuit shown in Fig. 1(b) represents both TL filters and "bipolar g_mC" filters, for which the transconductances comprise bipolar transistors only. The DR properties of the two filter sections are first compared based on a simplified approach. The influence of low-voltage implementation issues, tunability, low-power operation, high-frequency performance, and class AB operation is discussed next.

For the RC filter section, the current signal swing is limited, due to the supply voltages, to $V_{DD}/(2R)$. Assuming class A operation, the signal swing in the diode-C filter is limited by the dc bias current $I_{\rm dc}$.

The only noise source in the RC filter is due to the resistor R. The double-sided noise current power spectral density is given by $2kT/R$. The noise bandwidth of the filter equals $1/(2RC)$. Hence, the equivalent noise power is found to be $kT/(CR^2)$.

In the diode-C circuit, the power spectral density of the shot noise in the bias point equals $qI_{\rm dc}$. In comparing the two filters, both the capacitance value and the bandwidth of the filter are assumed to be equal. As a result, the relation between R and $I_{\rm dc}$ is given by $RI_{\rm dc} = U_T$, where U_T is the thermal voltage. Hence, the noise bandwidth of the TL filter equals $I_{\rm dc}/(2CU_T)$, and the equivalent noise power is found to be $qI_{\rm dc}^2/(2CU_T)$.

Dividing the maximal signal swing by the total amount of noise, we find the dynamic ranges ${\rm DR}_{RC}$ and ${\rm DR}_{g_mC}$ for the RC and the diode-C filter sections, respectively. This yields

$$ {\rm DR}_{RC} = \frac{CV_{DD}^2}{4kT} \tag{1} $$

$$ {\rm DR}_{g_mC} = \frac{2CU_T}{q}. \tag{2} $$

Equations (1) and (2) represent upper limits of the DR. In practice, these values have to be divided by the square of the crest factor of the specific signal being processed.

To compare the dynamic range properties, we divide ${\rm DR}_{RC}$ by ${\rm DR}_{g_mC}$:

$$ \frac{{\rm DR}_{RC}}{{\rm DR}_{g_mC}} = \frac{V_{DD}^2}{8U_T^2}. \tag{3} $$

Obviously, in principle, application of filters based on linear resistors yields a much better DR. For example, even for a low supply voltage of 1 V, and $U_T = 26$ mV, ${\rm DR}_{RC}$ and ${\rm DR}_{g_mC}$ differ by a factor of 185, or equivalently, 22.7 dB. Since the minimal power consumption of a filter is fundamentally related to the desired DR [9], the voltage swings should be preferably rail-to-rail [9], [11]. This is realized in the RC section, but not in the diode-C subcircuit, where the voltage swing is only U_T, corresponding to a current swing of $I_{\rm dc}$. This explains the large difference between ${\rm DR}_{\rm RC}$ and ${\rm DR}_{g_mC}$.

However, the conclusion drawn from (3) is not absolute. Many adventitious factors that affect the DR are not incorporated in (1) and (2). Their influence will be discussed next.

A. Opamp-MOSFET-C Filters

The opamp-MOSFET-C technique is the only method to realize filters with rail-to-rail signal swings and low noise levels [9]. In opamp-MOSFET-C filters, large voltage swings are possible, due to the fact that the quadratic behavior of the MOS transistor in strong inversion is not very nonlinear. Thus, the subcircuit shown in Fig. 1(a) can be used to represent this class of filters. Based on the simple MOS square law equation, it is even possible, in theory, to obtain a perfectly linear transconductance, which extends the voltage swings [12]. Consequently, these filters can be made to approach the fundamental limit regarding the minimal power consumption for a certain specified DR [9].

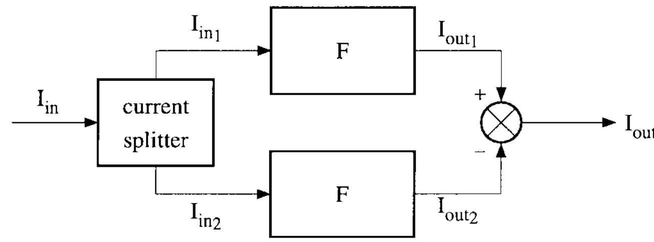

Fig. 2. Setup for class AB operation.

Unfortunately, at low supply voltages, opamp-MOSFET-C filters become difficult to implement [8]–[10], resulting in a lower DR than indicated by (1). Due to the requirement for strong inversion operation, very low voltage operation becomes only possible by using an on-chip charge pump to drive the gate voltages high. In addition, the tuning range of these filters is quite limited; it is only just enough to cope with process tolerances [8].

B. MOS g_m-C Filters

The class of g_mC filters can be divided into the categories of "MOS g_mC" and "bipolar g_mC," based on MOS and bipolar transconductors, respectively. With respect to opamp-MOSFET-C filters, the excess noise of the transconductors in MOS g_mC filters results in a factor 2 to 3 lower DR. Since most of the other characteristics of MOS g_mC filters are very similar to opamp-MOSFET-C filters, MOS g_mC filters will not be discussed here.

C. Bipolar g_m-C Filters

Since the bipolar transistor is an exponential device, the circuit shown in Fig. 1(b) can be used to represent bipolar g_mC filters. Equation (3) shows that the DR of bipolar g_mC filters is generally worse in comparison with opamp-MOSFET-C filters, since the voltage swings are limited to U_T. In practice, however, the voltage swings are even smaller due to the strongly nonlinear nature of the bipolar transistor. Therefore, most often the differential pair is used instead of a single transistor to eliminate even order distortion. Nevertheless, the voltage swings remain limited to only $0.5U_T$ for a THD of 1% [10]. The application of emitter degeneration resistors is often not allowed as this severely reduces the tuning range.

Transconductance linearization techniques using linear combinations of collector currents are not as effective for bipolar as for MOS transconductors and cannot increase the maximal voltage swings above 100 $\rm mV_{pp}$ [13]. Whereas for MOS transconductors exact linearization is possible, as the square

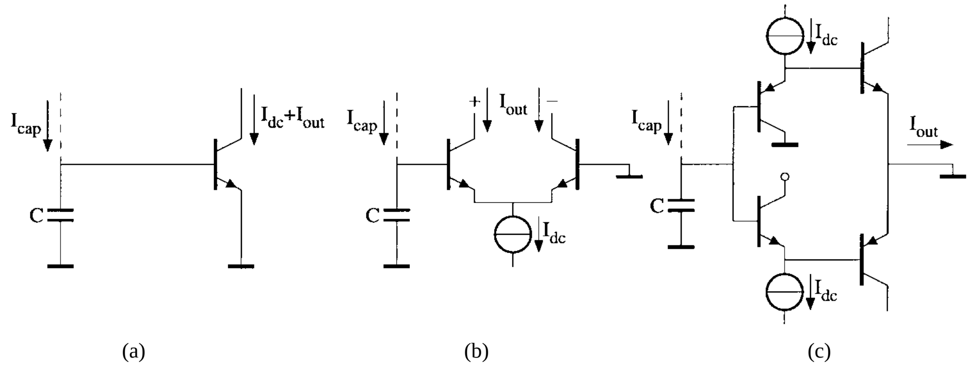

Fig. 3. Generic output structures of (a) log-domain filters, (b) $\tanh$ filters, and (c) $\sinh$ filters.

law is a polynomial, exact linearization of a bipolar transconductance is fundamentally impossible, since the exponential function is transcendental.

Although the small voltage swings in bipolar g_m-C filters have a negative influence on the DR properties, on the other hand, it makes them very suitable for operation at low supply voltages [8], [10], [13]. The DR is, to first order, independent of the supply voltage. Further, bipolar g_m-C filters exhibit very wide tuning ranges, and potential for high-frequency and low-power operation.

D. Translinear Filters

Since the diode-C circuit shown in Fig. 1(b) represents both bipolar g_mC and TL filters, these two types of filters have many characteristics in common, e.g., excellent tunability and potential for low-voltage and low-power operation.

A major difference is formed by the possible signal swings. Owing to the application of the DTL principle, in theory, TL filters offer a perfectly linear current-mode transfer function. Hence, the maximal signal swings in TL filters are larger than in bipolar g_mC filters. A DR comparison between a bipolar g_mC, a log-domain, and a $\tanh$ filter, reported in [3], shows that the latter two outperform the g_mC filter by 13 and 10 dB, respectively.

A very important aspect of log-domain filters is the possibility of class AB operation, which can be used to increase the DR [3], [4]. For example, in [4], a DR of 65 dB is reported, in connection with a maximal SNR of 52.5 dB. A generic class AB setup is shown in Fig. 2. A bipolar valued signal is split up into two unipolar signals, which are subsequently processed by different signal paths. Since the two different signal paths only have to process unipolar signals, no dc bias current is needed. Hence, the noise floor is decoupled from the maximal current signal swings, in contrast with the class A setup shown in Fig. 1(b). Class AB operation is possible due to the fact that the linearization mechanism of TL filters is theoretically exact.

E. Summary

To conclude, in principle, opamp-MOSFET-C filters are the best choice for a large DR, in the area of tuneable continuous-time filters. However, in low voltage environments, or for applications where a large tuning range is required, TL filters and bipolar g_m-C are more suitable.

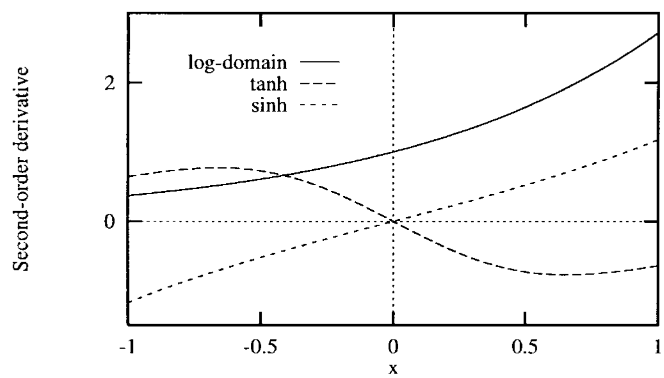

Fig. 4. The second-order derivatives of the V–I transfer functions of the circuits shown in Fig. 3.

Owing to the theoretically exact linearization mechanism, TL filters form an interesting and competitive alternative to g_m-C filters. Especially when class AB operation is applied, the DR of TL filters exceeds the DR obtainable with bipolar g_m-C filters.

IV. COMPANDING

In TL filters, the voltages are logarithmically related to the currents. Therefore, these circuits are in some way instantaneously companding. The exact nature of the companding affects the relation between the DR and the maximal SNR of a circuit. In this section, we apply the strict definition of companding, where the input compression and output expansion factors are larger than one. Only for filters that are companding according to this definition can the DR be significantly larger than the maximal SNR.

The companding properties of log-domain, $\tanh$, and $\sinh$ filters [3] can be characterized by the V–I transfer functions of the generic output structures of these circuits, depicted in Fig. 3. The V–I "expansion" functions are given by $\exp x, \tanh x$, and $\sinh x$, respectively, where x represents the normalized capacitance voltage swing. Fig. 4 depicts the second-order derivatives of the V–I transfer functions. For an output stage complying to the strict definition of (instantaneous) expansion, this second-order derivative should be strictly positive for $x > 0$ and strictly negative for $x < 0$.

Fig. 4 demonstrates that the $\sinh$ transconductor is indeed companding. This fact is related to the inherent class AB

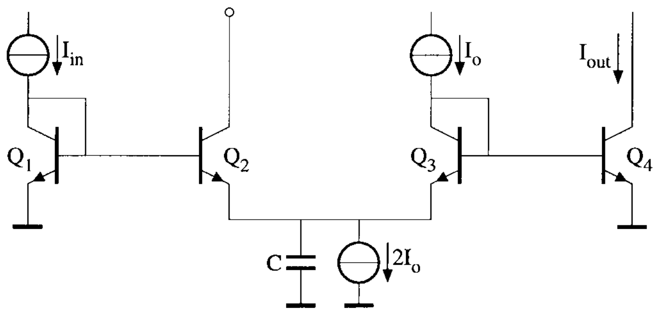

Fig. 5. A first-order translinear low-pass filter.

properties of $\sinh$ filters [3]. Large output current swings are thus facilitated without the requirement of a large dc bias current. Hence, the DR can be much larger than the maximal SNR.

The second-order derivative of the $\tanh$ function shows the opposite behavior with respect to the $\sinh$ transconductor. In other words, the differential pair *output* stage, shown in Fig. 3(b), implements a *compression* operation [5]. The signal swing is limited by the tail current. As a result, the DR and the maximal SNR will not differ much.

A similar conclusion can be reached for log-domain filters operated in class A. The second-order derivative of the function $\exp x$ is positive both for $x > 0$ and $x < 0$. Thus, these filters are not strictly companding, and hence, the maximal SNR and the DR will be approximately equal.

However, this conclusion is not valid for log-domain filters operated in class AB. Using the setup shown in Fig. 2, only positive signals have to be processed. Since for $x > 0$ the second-order derivative of the function $\exp x$ does correspond to the characteristics of an expansion stage, class AB operated log-domain filters are companding. This results in a significantly improved DR over the class A counterpart. It is interesting to note that $\tanh$ filters cannot be operated in class AB, since $\tanh x$ constitutes a compression for both positive and negative signals.

V. Maximal SNR

Due to the internal nonlinear behavior, causing signal $\times$ noise intermodulation, the noise floor in a TL filter is signal-dependent. Since class A operated TL filters, i.e., log-domain and $\tanh$ filters, are not companding, nonlinear noise effects are not very pronounced. In class AB operated filters, however, the intermodulation noise will dominate for large values of the signal power. The intermodulation noise power is proportional to the signal power, and consequently, the maximal SNR is fundamentally limited. Thus, only the DR, and not the maximal SNR, can be improved through companding [10].

The limited value of the maximal SNR can be explained intuitively from the first-order TL filter depicted in Fig. 5. For this circuit, class AB operation can be established using the setup shown in Fig. 2.

Fig. 6 shows that this filter can be redrawn as a cascade of two ports. It is obvious that the SNR of a chain of two ports is limited by the element(s) having the lowest SNR. A translation of this fact to the circuit shown in Fig. 6 indicates that the maximal SNR of the class AB TL filter is limited by the transistors with the lowest SNR value. In this context, the "SNR of a bipolar transistor" equals the ratio of the mean square value of the collector current I_C, to the power of the collector shot noise current i_C.[1] Since the mean square value P_{I_C} of the collector current is proportional to I_C^2, and the noise power P_{i_C}, over a certain noise bandwidth B_n, is proportional to I_C, the SNR of a bipolar transistor is proportional to I_C.

In Fig. 5, for increasing input signal swings and class AB operation, the average collector currents of Q_1 and Q_4 increase accordingly. Hence, the SNR of Q_1 and Q_4 increases. The average collector currents of Q_2 and Q_3, however, remain equal to I_o. Thus, for large signals, the SNR of these two transistors dictates the maximal SNR value of the complete filter. For Q_3, P_{I_C} equals I_o^2. The (double-sided) power spectral density of i_C is equal to qI_o, and the noise bandwidth B_n of the filter equals $I_o/(2CU_T)$. Hence, P_{i_C} becomes $qI_o^2/(2CU_T)$, and the SNR of Q_3 is found to be $2CU_T/q$. Since Q_2 and Q_3 have the same SNR, the overall SNR is reduced by a factor two. This yields

$$\mathrm{SNR}_{\max} = \frac{CU_T}{q}. \tag{4}$$

The exact value of the SNR of the filter for low and intermediate signal levels can only be obtained from detailed calculations, using the method outlined in [14]. In Fig. 7, the solid line shows the SNR of the class AB filter depicted in Fig. 5, using the setup shown in Fig. 2. It is assumed that the noise contribution of the dc current sources is negligible. A sinusoidal input signal is applied, i.e., $I_{\mathrm{in}} = m\ I_{\mathrm{dc}} \sin \omega t$, where I_{dc} is the quiescent current level of the (geometric mean) class AB splitter, m is the modulation index, and ω is an in-band frequency. The parameter values used in this plot are $I_{\mathrm{dc}} = I_o = 1\ \mu$A, $C = 10$ pF, and $U_T = 26$ mV. For low values of m, the SNR increases by 20 dB per decade. For high values of m, the SNR saturates to a value of 62.1 dB, as predicted by (4).

A. Seevinck's Class AB Integrator

Although exhibiting an externally linear transfer function, the class AB integrator proposed by Seevinck in [2], and shown in Fig. 8, is in a way a nonlinear DTL circuit as it implements two *nonlinear* DE's. The two TL loops of this integrator are described by

$$CU_T\dot{I}_{\mathrm{out1,2}} + I_{\mathrm{out1}} I_{\mathrm{out2}} = I_o I_{\mathrm{in1,2}}. \tag{5}$$

A linear integrator is obtained through subtraction of these equations. The overall input and output current equal $I_{\mathrm{in}} = I_{\mathrm{in1}} - I_{\mathrm{in2}}$ and $I_{\mathrm{out}} = I_{\mathrm{out1}} - I_{\mathrm{out2}}$, respectively.

Fig. 7 shows that the noise performance of both class AB filters is almost identical. The SNR of both filters converges to the same maximal value given by (4). However, for low values of the modulation index, Seevinck's circuit performs slightly better, about 0.6 dB, due to a lower noise floor in the quiescent point. This noise floor can be decreased further, up to 1 dB, by increasing the term proportional to $I_{out1} I_{out2}$ in (5).

[1] The influence of the base resistance noise is assumed to be negligible.

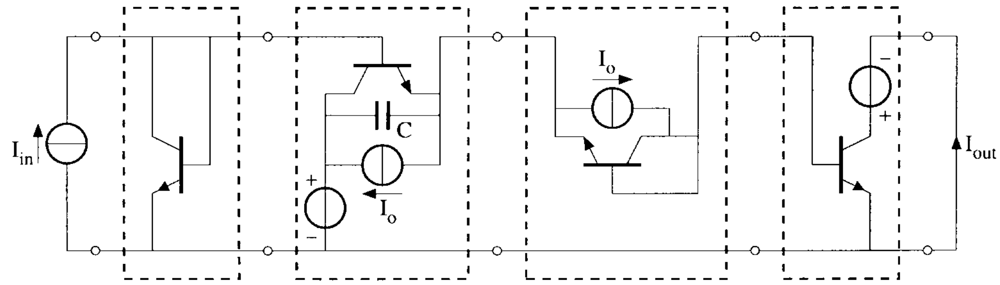

Fig. 6. A translinear filter consisting of a cascade of two ports.

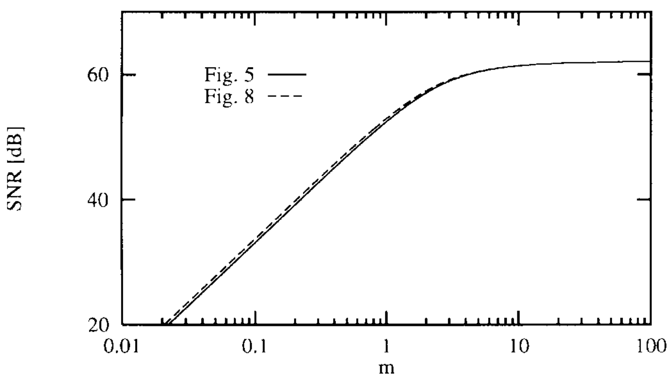

Fig. 7. SNR of a class AB translinear filter.

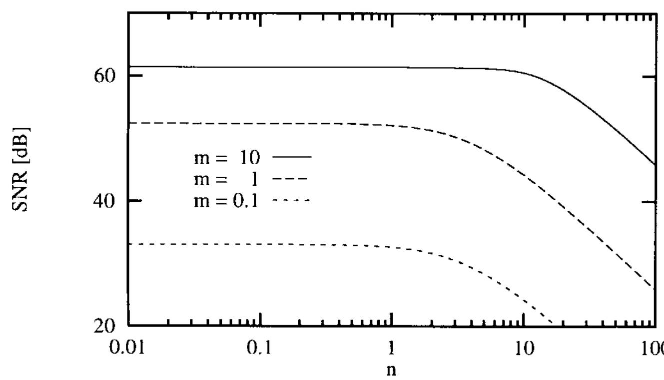

Fig. 9. SNR in the presence of an out-of-band signal.

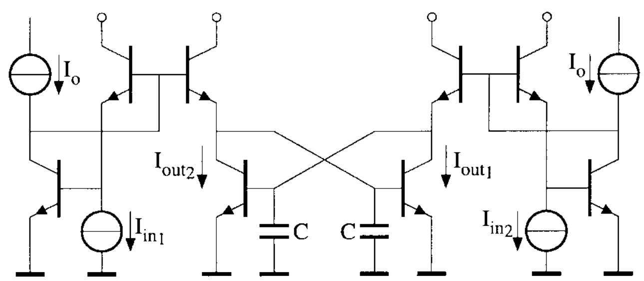

Fig. 8. Seevinck's class AB translinear integrator [2].

B. Noise Due to Out-of-Band Signals

An interesting situation is the coexistence of a large signal and a small signal in a companding TL filter. Suppose the small signal is the desired output signal and the large signal is outside the pass band. Now, in conventional filters, the large signal will limit the maximal SNR at the output, as it occupies a large part of the available DR of the filter. Naturally, the same effect applies to companding filters. However, in these filters, the SNR at the output will be further decreased, as the large out-of-band signal will increase the internal noise level [7]. This effect makes companding filters less suitable, e.g., for intermediate-frequency filtering [9], unless some form of linear prefiltering is used [7].

Fig. 9 displays the resulting SNR for an in-band signal as a function of the amplitude of an out-of-band signal, for the class AB filter shown in Fig. 5. The applied input signal equals $I_{\text{dc}}\ (m \sin \omega_1 t + n \sin \omega_2 t)$, where ω_1 and ω_2 are the in-band and out-of-band frequencies, respectively. The modulation index m of the in-band signal equals $[0.1, 1, 10]$, and n denotes the modulation index of the out-of-band signal. The figure demonstrates the expected behavior. Clearly, the effect is more pronounced for in-band signals with a small amplitude, i.e., a low value of m.

VI. Conclusions

A comparison of the DR properties of different filter implementation techniques shows that class AB operated TL filters constitute an interesting and competitive alternative for low-voltage or low-power applications. The maximal SNR of a TL filter, however, is fundamentally limited. This is due to signal $\times$ noise intermodulation, which causes the noise level to increase when large signals (even when out-of-band signals) are being processed.

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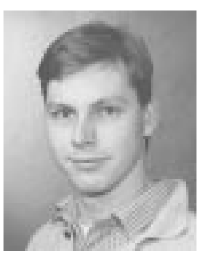

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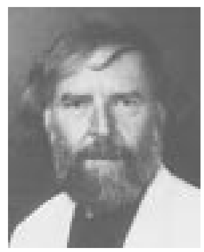

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