Micro-Power Analog-Filter Design

Gert Groenewold

Philips Semiconductors, Sunnyvale, CA, USA

Bert Monna

Delft University of Technology, Electronics Research Lab

Delft, The Netherlands

Bram Nauta

Philips Research Lab, Eindhoven, The Netherlands

Abstract

There are fundamental minima for the power consumption of filters. We will see how these minima can be found approximately, and also how filters that approach these minima can be designed for supply voltages down to about 1V.

1 Introduction

Most applications for low-power or low-voltage filters are battery operated. Other applications are high-frequency filters, which by their nature need much power to operate, so that the power consumption needs to be controlled to reduce chip temperature.

Especially for low-voltage filters, it is advisable to try to use rail-to-rail circuits. We will prove that this is also the case for low-power filters, even if the supply voltage does not impose a direct limit. In fact, we will see that a fundamental minimum for the power dissipation can be directly linked to the dynamic range (which is the maximal signal-to-noise ratio), and that there is no first-order link between the supply voltage and either the dynamic range or the power dissipation. Beyond the fundamental limits, there are always practical limits to what can be achieved, and here is the place where clever circuit design comes in.

Figure 1: A second-order filter network.

Figure 2: A Voltage-mode and a current-mode integrator. Both consist of a transconductor and a capacitor.

After having seen an illustrative example, we will discuss the fundamentals of power consumption in filters. This gives us the basics of how to curb it. Then we will see how with this knowledge practical circuits can be designed, and where practical limits are found.

2 An Example

A simple example may illustrate the basics of the filter design.

Figure 1 shows a network for a second-order lowpass filter. The network consists of two integrators with transfer $1/s$, and a number of interconnections. Two suitable integrator configurations are shown in Figure 2. The integrator of Figure 2(a) is a voltage-in voltage-out integrator, the other one is current-in current-out. Both consist of a capacitor and a voltage-to-current converter, also known as a transconductance stage or transconductor. The number of inputs of the voltage-mode integrator can be extended by adding transconductors and connecting their outputs to the output of the integrator. Similarly, the number of outputs of the current-mode integrator can be extended by adding transconductors and connecting their inputs to the input of the integrator. If we use either of these two integrator types, one of the two filters of Figure 3 will result.

Figure 3: A voltage-mode and a current-mode filter, based on the signal-flow graph of Figure 1 and the integrators of Figure 2.

A well known dynamic-range optimization step for filter networks like these is a process called *scaling*. The underlying idea is that if the dynamic range is maximal, all the critical internal signal levels must be equal. Intuitively, it is not difficult to see why. The filters consist of a network of active integrators. The integrators have a maximal signal amplitude at which they work correctly without generating too much distortion by saturation effects. The maximal input or output signal level of the filter is determined by the level at which at least one of the integrators starts to saturate. At this level, all the internal signal levels should be made as large as possible so that a maximal distance to the noise level is maintained. This means that all integrators should begin to saturate at the same (maximal) input signal level, although this does not necessarily need to happen for each integrator at the same frequency. If the output signal of the filter is a voltage, usually the same limit exists for the output signal as for the internal signals. Therefore, the internal signal levels also need to be made equal to the output signal level in this case.

The internal signal levels can be equalized without changing either the overall transfer function or the network topology by modifying the transconductance values. This process is called *scaling* and has for analog active filters probably been introduced by Martin and Sedra [1]. For digital filters earlier references can be found [2].

A mathematical proof of the fact that optimality requires equal signal levels exists [3]. It is more complicated than the line of thought sketched above, because in the process of scaling the noise sources are modified.

Because the internal signal levels need to be made as large as possible, the signal voltage levels in the two filters of Figure 3 will be equal if

the same transconductor type is used. The filter networks are almost identical, and therefore the noise properties of both versions will be almost identical, if the capacitance values are the same. This means that the dynamic-range properties of the voltage-mode filter and the current-mode filter are the same.

This is an important conclusion. It means that in the design of low-power low-voltage filters there is no direct advantage in using current mode. In a sense, there is no such thing as a current-mode filter, because internally there is a continuous interplay between voltage and current; the capacitances convert current to voltage, and the transconductors vice versa. The only difference between a voltage-mode and a current-mode filter is the nature of the input and output signal, which in a properly scaled filter never limit the dynamic range. We will show in the next section that there is no direct link between the dynamic range and the supply voltage of any active filter. The dynamic range can rather be linked to the power dissipation and the chip area.

One exception might be brought up. If there is internal dynamic-range compression in the voltage domain, a current-mode filter can have an advantage [4]. These filters can be used if they don't have to process large and small signal levels at the same time. They may therefore be useful in audio applications, but not in for instance intermediate-frequency filtering. The advantage lies in the fact that the input and output signals of the filter are currents, so that these do not need to be compressed.

3 Power and Noise

As early as in 1971, it was pointed out by Blom and Voorman [5, 6], that there is a fundamental link between the power dissipation and the dynamic range of an active filter. In simple terms, the signal power needs to be a larger than the noise power by a certain amount. Because the filters are active, this signal power is drawn from the power supplies. If the dynamic range is specified, we can determine the minimum power level at which the circuit can operate if we know the noise level.

Along those lines, Blom and Voorman came up with an expression for the power consumption of gyrator-capacitor filters. A simplified

derivation is given below. In this derivation, a filter is viewed as a network of integrators, which is more general than the gyrator approach.

It has been proven before [3, 7] that the noise level of a filter is dependent on the total capacitance of the filter, the transfer function, the network configuration, and the noise factor of the active circuits of the filter. The network configuration can be optimized, so that this factor can be removed from the dependency. This can only be done in an analytic way for narrow-band bandpass filters, and even in this case, it is a complex process. It is, however easily possible to make a rough estimate.

Suppose the total integration capacitance in the filter is C . If the order of the filter is n , there generally are either n or $2n$ capacitances. Let us assume that there are n capacitances C_i ($i = 1, 2, \dots, n$), each with value C/n . Each capacitance can be associated with a minimal mean-squared noise voltage over the capacitor of

$$\overline{V_{\text{noise}}^2} = \frac{kT}{C_i} = \frac{nkT}{C} \quad (1)$$

where C is the value of the capacitor, k is Boltzmann's constant, and T is the absolute temperature. The excess noise of the active circuits has not yet been taken into consideration. This can be done by multiplying this expression by the noise factor ξ of these circuits. This noise factor is defined as the ratio of the total noise power associated with the integrator and a fundamental minimum determined by the total of transconductors or resistors that set the time constants of the integrator. A noise-free active circuit would give rise to a noise factor of 1. If it is assumed that the noise associated with each capacitance appears without amplification or attenuation at the output (this generally appears to be approximately the case), the total mean-squared output noise voltage is

$$\overline{V_{\text{noise,out}}^2} = \xi \frac{n^2 kT}{C}. \quad (2)$$

We have assumed a voltage output. If the output signal is carried by a current, an effective transconductance factor needs to be taken into consideration and the derivation works the same way.

Suppose the dynamic range is required to be DR , the mean-squared

output signal voltage then needs to be

$$\overline{V_{\text{signal,out}}^2} = DR\xi \frac{n^2 kT}{C}. \quad (3)$$

If the filter is scaled to optimize dynamic range, all the internal capacitances carry the same signal voltage, which is the same as the output signal voltage, and the total mean-squared signal current in the capacitances is

$$\overline{I_{\text{signal}}^2} = \overline{V_{\text{signal,out}}^2} (\omega C)^2 = DR\xi n^2 kT C \omega^2. \quad (4)$$

This current is drawn from the supply voltages. If the supply voltage is V_{sup} , the power drawn is

$$P = V_{\text{sup}} \sqrt{\overline{I_{\text{signal}}^2}} = V_{\text{sup}} \sqrt{DR\xi n^2 kT C} \omega. \quad (5)$$

This is a very crude estimate, but it can give valuable insight into the fundamentals of power consumption in filters.

According to (5), the power dissipation is minimal if C is minimal. How small can C be made? Equation (3) links the mean-squared signal voltage at the output of the filter (and at the internal nodes) to the capacitance. If the capacitance is made smaller, the signal voltages have to be enlarged. The capacitances can be made smaller up to the point where the maximal signal voltage has been reached. If we assume that the signal is sinusoidal, its amplitude has to be equal to

$$V_{\text{ampl}} = \sqrt{2\overline{V_{\text{signal,out}}^2}} = \sqrt{2DR\xi \frac{n^2 kT}{C}}. \quad (6)$$

The maximally allowable amplitude is largest if the integrators can handle rail-to-rail signals. In that case, the maximal amplitude is half the supply voltage. This gives us with (6) a minimal capacitance value, which, when substituted into (5) yields a power estimate.

$$P = 2\sqrt{2}DR\xi n^2 kT \omega. \quad (7)$$

This estimate applies to class-B output stages. If class-A output stages are used, the supply current increases minimally by a factor π [3]. As the dynamic range of a bandpass filter is inversely proportional to its *quality factor* Q [6, 7, 8], which is defined as the ratio of its central

frequency to its bandwidth, (7) needs to be multiplied by Q in this case.

This power estimate only includes the power that is necessary to charge and discharge the filter capacitances with an amplitude large enough to satisfy the dynamic-range requirements. In practice, some extra power is necessary to bias stages, as we will see in Section 4.

Some important conclusions can now be drawn. To minimize the power dissipation of active filters, we need to:

- minimize the noise factor of the active circuits,
- use rail-to-rail class-B output stages.

Equation (7) shows that the power consumption is independent of the supply voltage. In fact, it suggests that any dynamic range can be realized at with any supply voltage. In a theoretical sense, this is true. Practically, the capacitance that is needed to maintain the DR increases if the supply voltage is reduced. This imposes a practical limit. Another limit stems from the fact that at low supply voltages it is difficult to realize the active circuits.

4 Circuits

Traditionally, there are two methods available to realize integrators, which are the filter building blocks. One is known as the transconductor-C method, the other as the opamp-R-C method. An example of a transconductor integrator is shown in Figure 4. It consists of a differential pair loaded with a capacitor. This circuit is very non-linear, but several linearization methods can be applied [9, 10, 11]. The main advantages of this stage are its enormous tuning range and its potential for high-frequency operation. It is, however, not extremely suitable for low-power applications, because it can not process rail-to-rail signals. The input and output swing can be maximized by linearizing and by reducing the saturation voltages of the bias current sources. The latter can be done by reducing the emitter resistors of the current-source transistors. Even if these resistors are zero valued, the stage can still not process rail-to-rail signals. Also, if the emitter resistors are reduced

Figure 4: A bipolar transconductance integrator.

Figure 5: An opamp- MOSFET-C filter.

in value, the noise generation by the bias circuits increases, which forces us to dissipate even more power to maintain a dynamic range.

By the knowledge of the authors (based on an extensive investigation [12]), there is only one way to realize rail-to-rail integrators with a low noise factor, viz. the opamp-MOSFET-C (or opamp-R-C) method. This method was initiated by Banu and Tsividis [13, 14] and immediately yielded dynamic ranges of 94 and 100dB, figures that have since then almost never been equaled. The 94dB figure had been obtained with a 20mW dissipation. With (7), a lower limit for the power consumption for class-A output stages is estimated at $44\mu\text{W}$.

Another example of such a filter, designed by Jaap van der Plas [15], is given in Figure 5. It is a fifth-order filter, consisting of a network of five

Figure 6: An opamp-MOSFET-C integrator.

Figure 7: The opamp used in the filter of Figure 5.

integrators. Each integrator has been realized with a balanced opamp, two capacitors, and a number of MOSFET's that serve as tunable resistors. One such integrator is shown in Figure 6. For this integrator to be able to process rail-to-rail signals at its output, the opamp needs to have a rail-to-rail output stage. Figure 7 shows the opamp. As NMOS transistors generate much more $1/f$ -noise than PMOS transistors, no NMOS transistors have been used. For that reason, the input differential pair (P_1 and P_2) has been loaded with resistors (R_d) instead of NMOS transistors, even if this reduces the opamp gain. For common-mode stability as well as for even-order nonlinearity cancellation, an accurate common-mode feedback circuit (P_3 - P_6 and the circuitry around it) has been added. This particular common-mode feedback configuration has been originally proposed by Banu *et al.* [16]. The output stage gives the opamp and the integrators output rail-to-rail capabilities.

The integrators can have a rail-to-rail input swing if the gate voltages

of the MOSFET's are higher than the supply voltage plus the threshold voltage of the MOSFET's. If the MOSFET's are of the enhancement type, this would mean that the gate voltages need to be higher than the supply. Because the gates do not draw current, this is not difficult to accomplish. Several charge pumps have been developed for this purpose (e.g., [17]).

If depletion MOSFET's are available, their gate voltages could be kept below the supply voltage, eliminating the need for a charge pump. It is unfortunate that very few processes allow for the usage of these devices. Many companies develop and use BiCMOS processes for the development of low-power low-voltage circuits. The step from CMOS to BiCMOS that is made, is a lot more involved than a step that would add depletion MOSFET's, while this step simplifies the design of low-voltage filters greatly.

In an initial implementation, this filter had a 98dB dynamic range with a power consumption of 10mW. A minimum power consumption by (7) is 0.13mW for a class-A output stage. In a later study [12] it appeared that with the same power consumption, the dynamic range could be increased to 104.5dB, which would be estimated by (7) to cost minimally 0.6mW. Even with this improvement, the actual filter still consumes more than necessary. To get closer to the fundamental minima, bipolar class-AB opamps should be used.

Figure 8 shows a 3V class-AB opamp designed by Anton de Graauw [18]. The use of bipolar transistors at the input was favorable for noise performance. The bias current for the input stage has been optimized for noise minimization. The noise factor of the integrators that have been realized with this opamp is as low as 1.18, which corresponds to 0.7dB.

For the output stage a fully differential rail-to-rail class-AB configuration has been chosen [19, 20]. The main transistors in one output stage are the CE stages Q_3 and Q_4 , resp. Q_5 and Q_6 . The common-mode feedback circuit consists of R_{s1} , R_{s2} , Q_{18} , Q_{17} , Q_{15} , Q_{16} and U'_{cr} . Apart from the mentioned functions of such a circuit to ensure common-mode stability and reducing even-order distortion, it also serves to annihilate common-mode to differential-mode conversion caused by NPN-PNP asymmetry in the output stage. The circuit consisting of Q_7 to Q_{14} and the surrounding current sources biases the output stages at $1.5\mu\text{A}$ each.

Figure 8: A 3V class-AB opamp.

Figure 9: A class-A opamp.

With a 3V supply voltage, the opamp can handle a single-ended output swing of 2.8V, which brings the opamp to within 0.6dB from rail-to-rail operation.

The opamp was designed for application in a bandpass filter with a central frequency of 100kHz and a bandwidth of 7 kHz. The filter contains 16 integration capacitors of 10pF each. With the above opamp properties, the dynamic range of the filter is 75dB with a total static power consumption of 0.94mW. If the filter is processing maximum-level signals, the power consumption increases by about 0.2mW. This means that still most of the supply power is used for biasing the active circuits. Especially the current that is necessary to bias the input stage for optimal noise performance contributes to this. The actual power consumption compares reasonably to the estimate of 0.25mW that is obtained via (7).

Another 3V opamp is depicted in Figure 9. This opamp has been designed by Marcel Lugthart [21] for application in a 100kHz notch filter with a Q of 33. The input stage is a differential pair, optimized to a noise factor of 1.11 in the same way as in the previous opamp. The output stage is a rail-to-rail CE-stage, biased in class A at $60\mu\text{A}$.

The dynamic range of the second-order notch filter is about 80dB. With corrections for the quality factor and the fact that the output stage

is class A, an estimate for the lower limit of the power consumption according to (7) is 0.34mW. The actual circuit consumes 1.5mW. The difference is largely due to overdimension of the output stage.

A noteworthy detail of this opamp is its phase compensation. It can be proven [21] that the effects of the non-idealities of the opamp on the phase shift of the integrator add up to zero at the notch frequency if the dominant pole of the opamp is situated at the same frequency. This has been accomplished by choosing proper values for the common-mode feedback resistors R_{s1} and R_{s2} . This method makes it possible to construct high-frequency opamp-R-C filters without placing extreme demands on the bandwidth of the opamps.

If filters have to operate at even lower supply voltages, it becomes increasingly important to maintain a high swing and minimize the noise generation. Below a supply voltage of 1.2V the swing is mainly limited at the input of the opamp. This is because in a practical filter circuit the input bias voltage of the opamp is equal to the output bias voltage. So for an optimal swing, the DC input bias voltage must be half the supply voltage, which means that below 2V it is not very well possible anymore to use a differential pair. The input bias voltage can not be brought closer than one basis-emitter voltage to both supply lines at the same time, and this minimum is attained by using an input CE stage with the emitter grounded, as shown in Figure 10. The common-mode stabilization circuits that have been used in the above opamps can not be used here. Common-mode feedback has been implemented with a resistor bridge. For common-mode stability these resistors can not be made too large, so that the noise factor of this opamp can not be made smaller than 2. The input bias voltage has been reduced by choosing a large input device. A rail-to-rail swing is attainable at a supply voltage of 1.2V. Below this voltage, the swing rapidly decreases, and the circuit works to about 0.8V.

If this opamp were to be used in the eighth-order filter mentioned above, with 16 10pF integration capacitors, a dynamic range of 64dB can be realized with a 1.2V supply voltage. According to (7), corrected for the quality factor and class-A operation, a minimum power consumption of 0.11mW is necessary. The static power consumption of the actual filter would be 0.24mW, which compares reasonably.

Figure 10: A 1V opamp.

5 Conclusions

We have seen that there is a fundamental lower limit to the power consumption of a continuous-time active filter. This limit can be directly linked to the dynamic range of the filter, and there is no direct connection between the dynamic range and the supply voltage. We came to a very rough approximation of this limit in (7). This limit is set by the power that is necessary to charge and discharge the capacitors in the filter with a certain frequency and amplitude. Practical filters will always dissipate more power, because some power is needed to bias circuits, especially input circuits that need to be biased for optimal noise performance.

In order to approach the fundamental limit as close as possible, rail-to-rail circuits with a low noise factor should be used. The only candidates that we found were opamp-MOSFET-C filters. These circuits can be configured in a close-to-optimum way down to supply voltages of 1.2V, although they can work less optimally to 0.8V.

References

- [1] Ken Martin and Adel S. Sedra. Designing leap-frog and SFG filters with optimum dynamic range. *Proceedings of the IEEE*, 65(8):1210–1211, August 1977.
- [2] L.B. Jackson. On the interaction of roundoff noise and dynamic range in digital filters. *The Bell System Technical Journal*, 47(2):159–184, February 1970.
- [3] Gert Groenewold. *Optimal Dynamic Range Integrated Continuous-Time Filters*. PhD thesis, Delft University of Technology, 1992.
- [4] E. Seevinck. Companding current-mode integrator: A new circuit principle for continuous-time monolithic filters. *Electronics Letters*, 26(24):2046–2047, November 1990.
- [5] D. Blom and J.O. Voorman. Noise and dissipation of electronic gyrators. *Philips Research Reports*, 26:103–113, 1971.
- [6] J.O. Voorman. *The Gyrator as a Monolithic Circuit in Electronic Systems*. PhD thesis, University of Nijmegen, Nijmegen, 1977.
- [7] Gert Groenewold. The design of high dynamic range continuous-time integratable bandpass filters. *IEEE Transactions on Circuits and Systems*, 38(8):838–852, August 1991.
- [8] Haideh Khorramabadi and Paul R. Gray. High-frequency CMOS continuous-time filters. *IEEE Journal of Solid State Circuits*, SC-19(6):939–948, December 1984.
- [9] J.O. Voorman, W.H.A. Bröls, and P.J. Barth. Bipolar integration of analog gyrator and laguerre type filters (transconductor-capacitor filters). In *Proceedings ECCTD'83*, pages 108–110, Stuttgart, September 1983.
- [10] Johannes O. Voorman. Transconductance amplifier. U.S. Patent 4,723,110, February 2, 1988.
- [11] Hiroshi Tanimoto, Mikio Koyama, and Yoshihiro Yoshida. Realization of a 1-V active filter using a linearization technique employing

- plurality of emitter-coupled pairs. *IEEE Journal of Solid State Circuits*, 26(7):937–945, July 1991.
- [12] Gert Groenewold. Optimal dynamic range integrators. *IEEE Transactions on Circuits and Systems—I: Fundamental Theory And Applications*, 39(8):614–627, August 1992.
- [13] Mihai Banu and Yannis Tsividis. Fully integrated active RC filters in MOS technology. *IEEE Journal of Solid State Circuits*, SC-18(6):644–651, December 1983.
- [14] Mihai Banu and Yannis Tsividis. An elliptic continuous-time CMOS filter with on-chip automatic tuning. *IEEE Journal of Solid State Circuits*, SC-20(6):1114–1121, December 1985.
- [15] Jaap van der Plas. MOSFET-C filter with low excess noise and accurate automatic tuning. *IEEE Journal of Solid State Circuits*, 26(7):922–929, July 1991.
- [16] Mihai Banu, John M. Khoury, and Yannis Tsividis. Fully differential operational amplifiers with accurate output balancing. *IEEE Journal of Solid State Circuits*, 23(6):1410–1414, 1988.
- [17] J.C. Sandee. Een on-chip automatisch afstemcircuit voor tijdcontinue filters met een optimaal dynamisch bereik. Master's thesis, Delft University of Technology, Delft, November 1992.
- [18] A.J.M. de Graauw. Een operationele versterker voor MOSFET-C filters. Master's thesis, Delft University of Technology, Delft, December 1992.
- [19] Johan H. Huijsing. Private Communication.
- [20] Joseph N. Babanezhad. A low-output-impedance fully differential op amp with large output swing and continuous-time common-mode feedback. *IEEE Journal of Solid State Circuits*, 26(12):1825–1833, December 1991.
- [21] M.L. Lugthart. Realisatie van een geïntegreerd notchfilter. Master's thesis, Delft University of Technology, Delft, November 1993.