Analog Integrated Circuits and Signal Processing 3, 105–112 (1993)

A Low-Voltage Low-Power Current-Mode High-Pass Leapfrog Filter

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Abstract. A low-power bipolar continuous-time low-frequency high-pass second-order Butterworth filter is presented that works in the current domain and operates from a single 1.3-V battery. The filter contains two adjustable integrators. These integrators are realized by means of a capacitance and an adjustable transconductance amplifier with an indirect output. The complete filter, including all capacitances needed, can be integrated in an ordinary full-custom IC process. A semicustom realization is shown. The filter demonstrates operation down to 1 V with less than 16 μW power consumption and a dynamic range of 50 dB. Its cutoff frequency can be exponentially tuned with a control current over a range from 100 Hz to 1 kHz.

Received November 11, 1991; Revised April 8, 1992.

1. Introduction

As the size and power consumption of electronic circuits is becoming more and more important, the demand for circuits that use a single battery and consume little current is increasing. Examples are hand-carried radiotelephones, pagers, and hearing instruments (see, e.g., [1–4]). To improve the speech intelligibility in these systems, especially of consonants, often a high-pass filter is used [5–7]. Apart from operating at "low-voltage level" (i.e., 1–1.3 V) and consuming as little current as possible to ensure long battery life, the filter bandwidth must be programmable to ensure a wide application area. Moreover, external components need to be avoided as much as possible.

This paper deals with the design and measurement of a fully integrated second-order high-pass Butterworth filter that meets all former specifications and whose cutoff frequency can be varied from 100 Hz to 1 kHz. In Sections 2–4, successively, the filter design is followed from a suitable filter architecture, via the elementary building blocks, up to their signal path. Together with a proper biasing circuit these blocks form the complete filter, as described in Section 5. Section 6 deals with a semicustom realization of which, in Section 7, measurement results are given.

2. A First Approach

To reduce the complexity and power consumption of the circuit to a minimum, we have chosen for an analog, continuous-time filter. A passive, lossless *LC* filter would offer the best solution, but as the adjustability of such filters is weak and inductors cannot be integrated, we will have to simulate the network equations of the filter by means of "analog computer techniques." This results in a "leapfrog filter" [8, 9]. The inductances and capacitances in a passive *LC* filter can be seen as elements that perform mathematical operations, like integrating and differentiating. These operations on the voltages or currents in the circuit are now performed by active circuits. Because a one-to-one relationship exists between the time constant of each active circuit and the value of a reactive element in the *LC* filter, also the sensitivity of the transfer function of a leapfrog filter to its component values is low.

Figure 1 shows a possible implementation of a second-order high-pass Butterworth leapfrog filter. The filter consists of integrators (because of their frequency stability [10]) and current mirrors with multiple outputs and has a current as the information carrying quantity, to reduce the influence of parasitic admittances [11]. The right current mirror provides three output signals: one equals the (inverted) input signal, the other two equal half the input signal. The input-output relation $H(f)$ is given by

$$H(f) = \frac{i_o}{i_i} = \frac{1}{j^2 + j\sqrt{2}f_c/f + f_c^2/f^2} \qquad (1)$$

in which f_c equals the cutoff frequency of the filter. Note that for every f_c the filter response is maximally flat (i.e., a Butterworth characteristic).

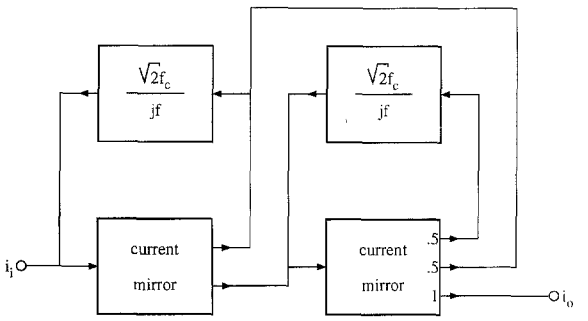

Fig. 1. Implementation of a second-order high-pass Butterworth leap-frog filter operating in the current domain.

3. The Integrator Blocks

As the integrator is to be realized in an integrated circuit, the integrating element must be a capacitor. The input signal of this capacitor is a current, whereas the output signal is a voltage. Therefore, an integrator operating in the current domain will always consist of a capacitor followed by a transconductance amplifier. This configuration is shown in figure 2. The amplifier has an indirect output [12]. Hence the possible output swing is maximized. Moreover, the filter is easily tuned by changing the scaling factor n, as will be shown subsequently.

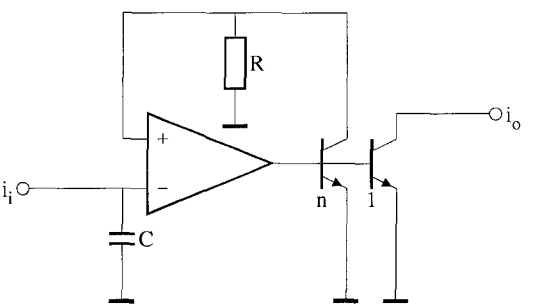

Fig. 2. Implementation of the integrator: a capacitance-transconductance amplifier with an indirect output. n is a scaling factor.

4. The Capacitance-Transconductance Amplifier

A possible implementation of the capacitance-transconductance amplifier is shown in figure 3. The capacitance C transforms the input current i_i into a voltage $v_i = i_i/j2\pi fC$, that in turn is transformed by Q_1, Q_2, and resistance R into a current $i_o' = v_i/R = i_i/j2\pi fRC$. Q_3 provides the indirect output. When the current through Q_3 is n times as small as the current

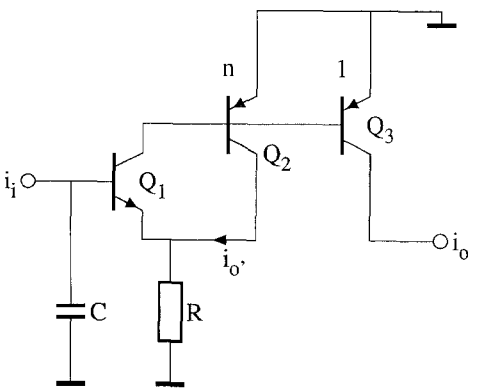

Fig. 3. Circuit diagram of the capacitance-transconductance amplifier.

through Q_2 (and when the Early effect is negligible) we find

$$i_o = \frac{i_i}{j2\pi fnRC} \tag{2}$$

Combining (1) and (2) we find that the transfer functions H_t must satisfy

$$H_t = \frac{1}{nR} = 2\pi\sqrt{2}\, f_c C \tag{3}$$

4.1. Noise Optimization

To calculate the amount of noise the capacitance-transconductance amplifier contributes to its output current, we shift the dominating noise sources (i.e., the noise sources of Q_1) to the output, integrate the noise-power density spectrum over the total audio frequency range, and we find an expression for the equivalent noise current at the output, $i_{n,\mathrm{eq}}$ (see also appendix):

$$i_{n,\mathrm{eq}} = \sqrt{\frac{2kT(f_2 - f_1)}{n^2}\left(\frac{1}{R^2 g_m} + \frac{g_m}{4\pi^2 B_F f_1 f_2 R^2 C^2}\right)} \tag{4}$$

with f_1 the lowest frequency, f_2 the highest frequency, g_m the transconductance factor of Q_1, and B_F the (low-frequency) current gain factor of Q_1. Obviously, C has to be chosen as large as possible. In addition, this expression can be minimized by varying g_m. Because g_m is proportional to the collector current of Q_1, I_{C,Q_1}, it is possible to minimize (4) by varying I_C. For this optimum value, $I_{C,Q_1,\mathrm{opt}}$, we find

$$I_{C,Q_1,\mathrm{opt}} = \frac{kT}{q\sqrt{R^2/B_F + 1/4\pi^2 B_F C^2 f_1 f_2}} \tag{5}$$

4.2. Collector Currents of Q_2 and Q_3

The maximum output signal, $i_{t,\max}$ of the transconductance amplifier can be determined from its maximum input signal, $v_{t,\max}$, which is, in fact, the maximum voltage across the capacitor. Therefore

$$v_{t,\max} = \frac{i_{t,\max}}{H_{t,\min}} \tag{6}$$

With (3) it follows that

$$i_{t,\max} = v_{t,\max} H_{t,\min} = v_{t,\max}\, 2\pi\sqrt{2}\, f_{c,\min} C \tag{7}$$

As $i_{t,\max}$ is delivered by Q_3, the collector current of Q_3, I_{C,Q_3} is chosen best at least to be $i_{t,\max}$. A practical value of 1.2 times $i_{t,\max}$ will do. As the filter has to operate at voltages down to 1 V, $v_{t,\max}$ has been chosen equal to 100 mV.

The tuning of the filter is done by varying the factor n, which equals the collector current of Q_2, I_{C,Q_2} divided by the collector current of Q_3, I_{C,Q_3}. With (3) it follows that

$$I_{C,Q_2} = nI_{C,Q_3} = \frac{1}{2\pi\sqrt{2}\, f_c RC} I_{c,Q_3} \tag{8}$$

4.3. Efficiency of the Capacitance-Transconductance Amplifier

Only a proper choice of the resistance R remains. Therefore we look at the efficiency η of the integrator, defined as follows:

$$\eta = \frac{\text{maximum signal current at the output}}{\text{total current of the integrator}} \tag{9}$$

or

$$\eta = \frac{I_{C,Q_3}}{I_{C,Q_1} + (n+1) I_{C,Q_3}} \tag{10}$$

In order to obtain a good efficiency we see that it is convenient to choose n smaller than one for all cutoff frequencies possible. With (3) it follows that

$$n = \frac{1}{2\pi\sqrt{2}\, f_c RC} > 1 \quad \forall f_c \tag{11}$$

and thus

$$R > \frac{1}{2\pi\sqrt{2}\, f_{c,\min} C} \tag{12}$$

In practice this easily leads to values of R that cannot be integrated. For example, if $f_{c,\min} = 100$ Hz and $C = 400$ pF this results in $R > 2.8$ MΩ. In most standard bipolar IC processes the value of a diffused resistor is limited to several hundred kilohms. So in that case the value of R can best be chosen maximally.

5. The Complete Filter

When we combine two capacitance-transconductance amplifiers with two current mirrors having multiple outputs according to the block diagram of figure 1, we have completed the signal path of the complete filter (figure 4). The (adjustable) scale factor n, which determines the cutoff frequency of the filter, is obtained by means of an adjustable voltage source V_f:

$$n = e^{V_f/V_T} \tag{13}$$

With (3) it follows that

$$f_c = \frac{1}{2\pi\sqrt{2}\, nRC} = \frac{e^{-V_f/V_T}}{2\pi\sqrt{2}\, RC} \tag{14}$$

This exponential relation between V_f and the cutoff frequency enables us to adjust the cutoff requency over a wide range. If V_f is made proportional to the absolute temperature (PTAT) the cutoff frequency is independent of the temperature.

The complete filter, including its biasing scheme, is depicted in figure 5. GM-compensated mirrors [13] with multiple outputs provide the bias currents of every stage. When the voltage across R_1 and R_2 equals the thermal voltage V_T (= kT/q), the output of the mirror is insensitive to its input (I_1 and I_2). At this point I_1 and I_2 equal e times the output currents $I_{C,Q_1,\text{opt}}$ and I_{C,Q_3}, respectively. With (5) and (7) it follows that

$$I_1 = eI_{C,Q_1,\text{opt}} = e\frac{kT}{q\sqrt{R^2/B_F + 1/4\pi^2 B_F C^2 f_1 f_2}} \tag{15}$$

$$I_2 = eI_{C,Q_3} = e1.2v_{t,\max}\, 2\pi\sqrt{2}\, f_{c,\min} C \tag{16}$$

$$R_1 = \frac{kT}{qI_1} \tag{17}$$

$$R_2 = \frac{kT}{qI_2} \tag{18}$$

The currents I_1 and I_2 come from an external circuit which controls whether the filter is on or off (standby position). Transistors Q_A through Q_E provide the collector currents of Q_{2A} and Q_{2B}. The voltage source V_f

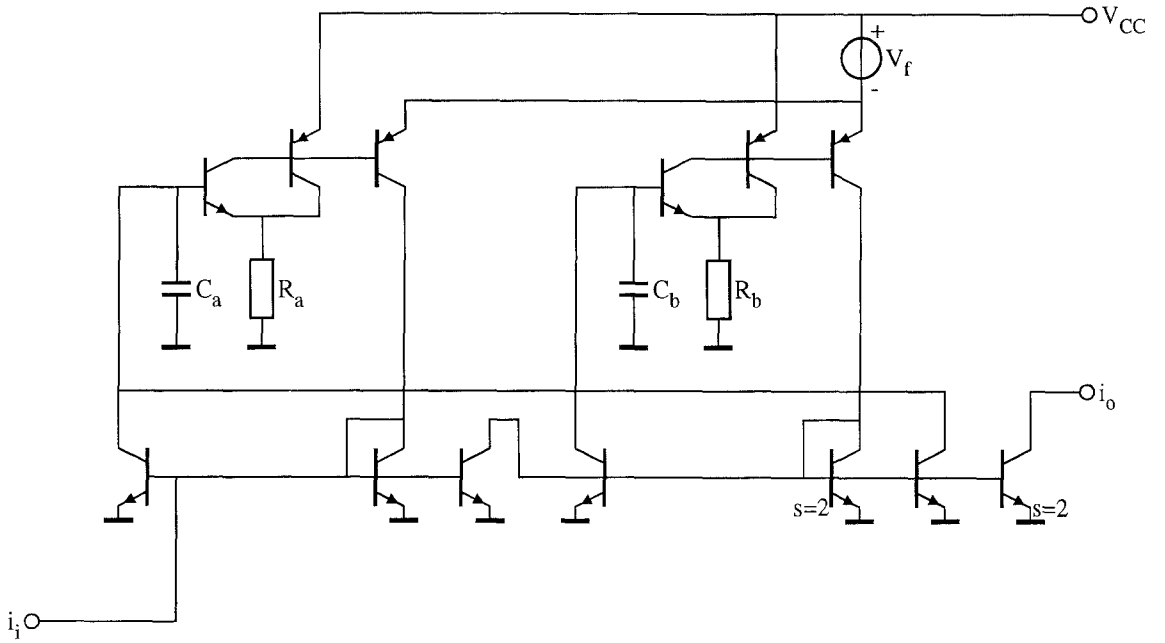

Fig. 4. Signal path of the complete filter, $s = 1$ means a doubled emitter area ration.

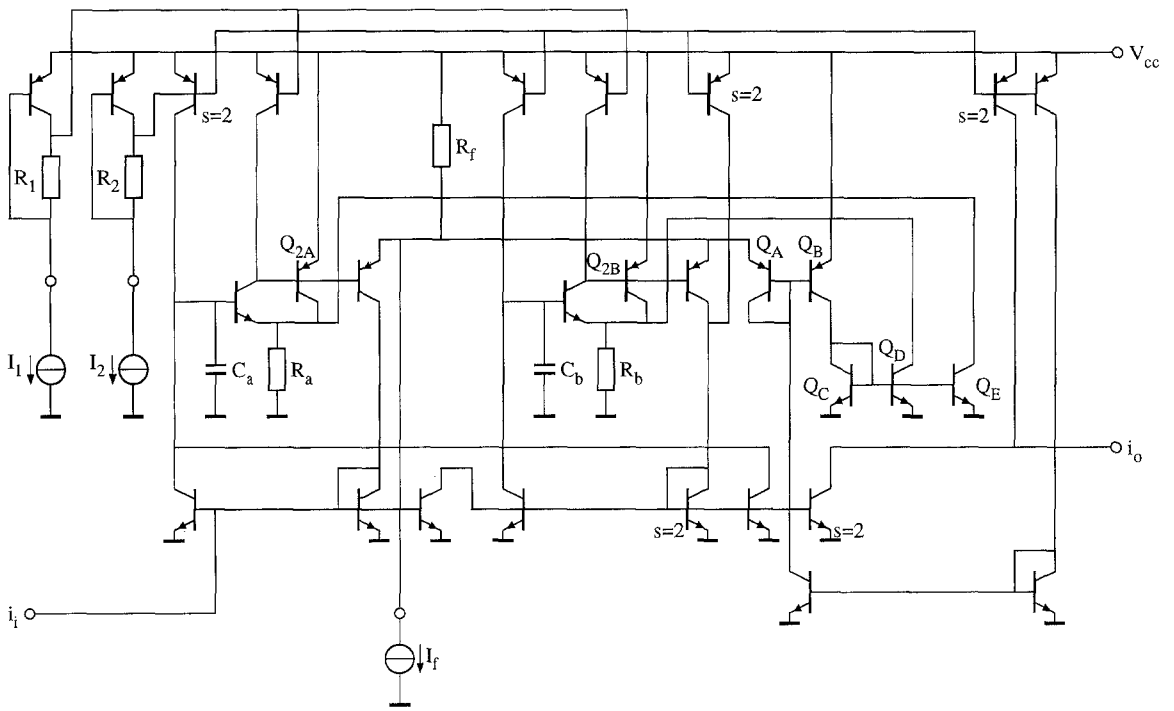

Fig. 5. The filter including its biasing scheme. $s = 2$ means a doubled emitter area ratio.

is realized by a single resistor R_f through which flows a current I_f that is derived from an external circuit (e.g., a potentiometer or a programmable current source):

$$I_f = \frac{V_f}{R_f} = -\frac{V_T}{R_f} \ln (2\pi\sqrt{2}\, f_c RC) \tag{19}$$

Besides this current I_f, two emitter currents also contribute to the voltage over R_f. Because they contain an

ac component as well, this results in distortion of the output signal at higher input signal levels. This distortion can be made sufficiently small by choosing R_f smaller and I_f larger. Unfortunately, this degrades the efficiency of the filter. An acceptable compromise is therefore to be found with the aid of computer simulations. It turns out that this is possible. In critical situations the distortion can be reduced by means of a buffered voltage source [14].

6. Semicustom Realization

The active circuitry of figure 5 was integrated in a semicustom chip in the LA251 process [15]. Figure 6 shows a microphotograph of the chip. The two integrating capacitors C_A and C_B were chosen 400 pF. The two (diffused) resistors R_A and R_B were chosen to be 200 kΩ. These values can be integrated easily in an ordinary full-custom process. With T, B_F, f_1, f_2, and $f_{c,\min}$ assumed to be 300 K, 100, 100 Hz, 10 kHz, and 100 Hz, respectively, (15)–(18) result in

$$I_1 = 1.6\ \mu\text{A}, \quad I_2 = 120\ n\text{A},$$
$$R_1 = 16\ k\Omega, \quad R_2 = 220\ k\Omega$$

For R_f a 6-kΩ resistor has been chosen. I_f therefore varies between 1.5 and 11 μA.

7. Measurements

The communication between the filter and the measuring instruments was done as follows. In order to perform the voltage-current conversion, a 100-MΩ resistor has been chosen. The output signal was made measurable by an op amp with a 1-MΩ resistor in transimpedance configuration, to perform the current-voltage conversion. For three different values of the control current I_f, corresponding with cutoff frequencies of 100, 320, and 1000 Hz, respectively, the gain and phase of the transfer were measured. The result is shown in figure 7. We observe a second-order Butterworth characteristic between 10 Hz and 10 kHz. The loss in the passband is less than 1 dB. The bandwidth equals 10 kHz. The filter linearity has been evaluated measuring the total harmonic distortion for various input signals. The maximum current (at both the input and output) corresponding to a 5% total harmonic distortion amounts to 25 nA (peakvalue). Combining this number with the measured output noise of 57 pA_{rms}

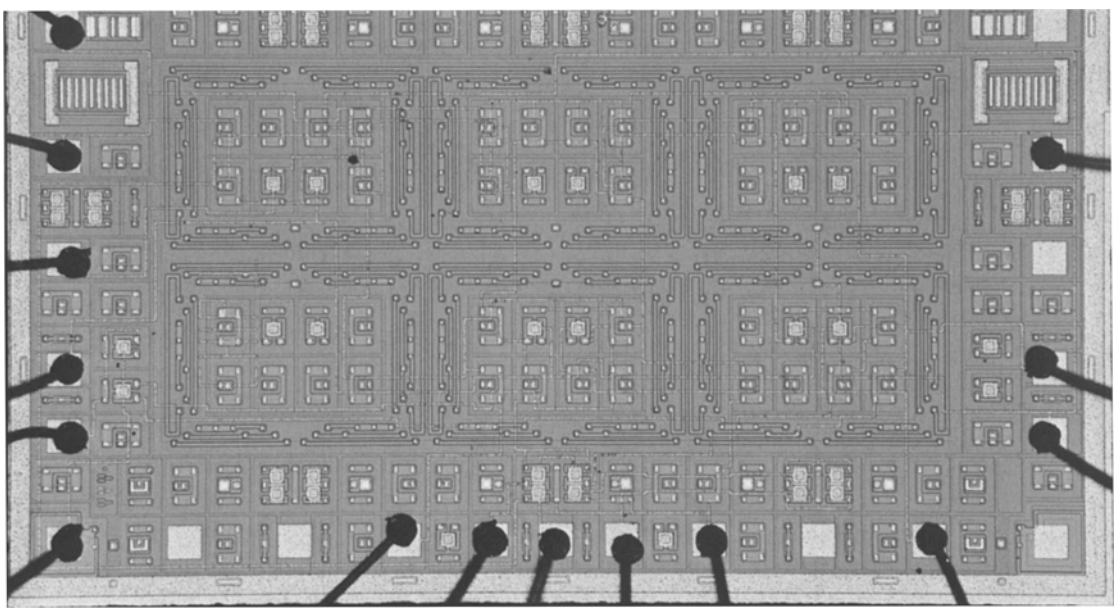

Fig. 6. Microphotograph of the semicustom chip.

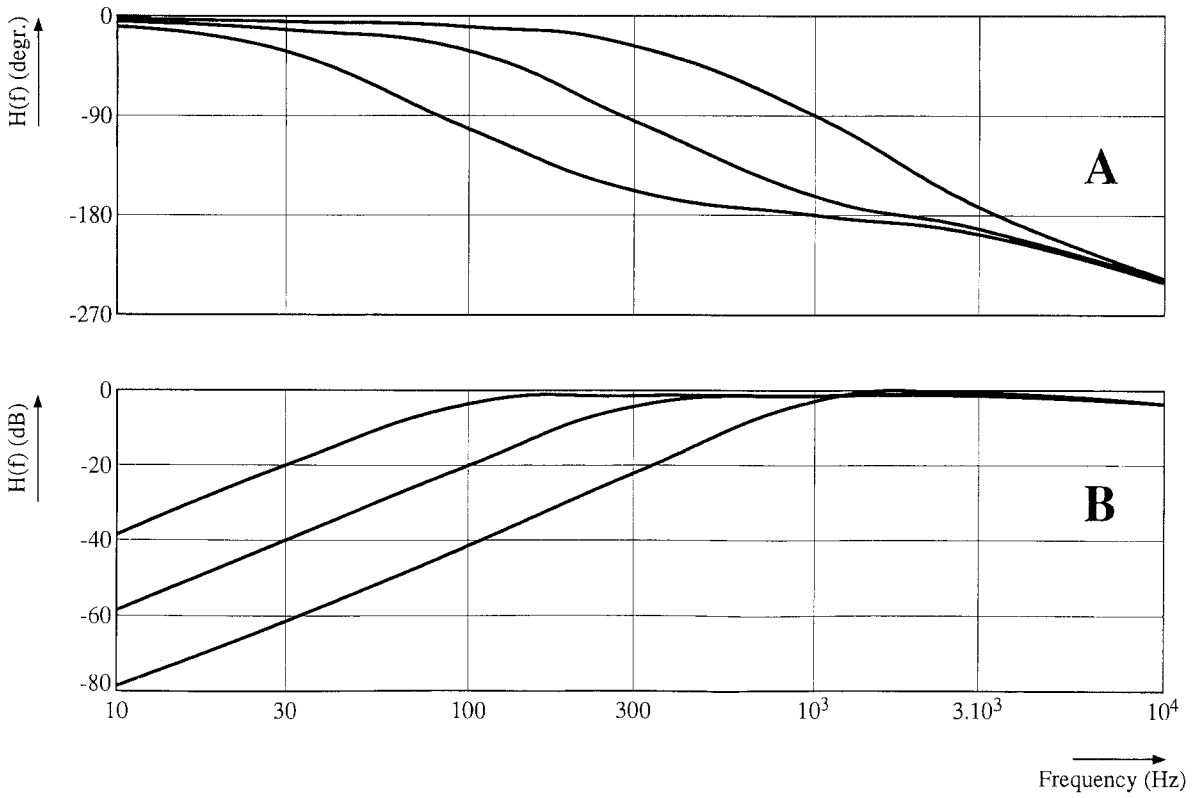

Fig. 7. Phase (A) and gain (B) transfer of the filter.

results in a dynamic range of 50 dB. The filter operates well at voltages down to 1.0 V and consumes less than 16 μA. No instability occurs.

8. Conclusions

A low-power, low-voltage fully integrated second-order high-pass filter has been presented that works in the current domain. The filter is a two-integrator type, in which an integrator is realized by a capacitance and an adjustable transconductance amplifier with an indirect output. The test chip demonstrates operation down to 1 V with less than 16 μW power consumption and a dynamic range of 50 dB.

9. Appendix: Noise Transformation

In order to find an expression for the equivalent noise current at the output of the integrator we will use the simple noise model of the bipolar transistor as presented in figure 8 [16].

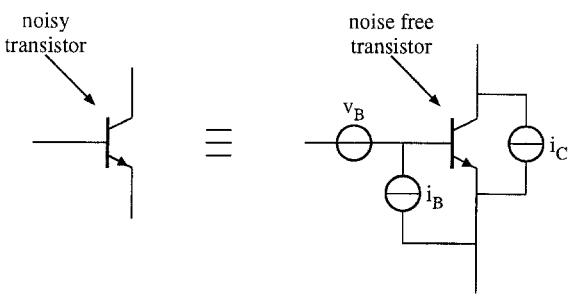

Fig. 8. Representation of a noisy transistor by a noise free transistor together with two noise current sources and one noise voltage source.

In the bipolar transistor two noise sources are found whose spectra are proportional to the bias current: the shot noise (i_C) of the collector current, with a power density spectrum (in A^2/Hz) $S(i_C) = 2qI_C$, and the shot noise (i_B) of the base current, with spectrum $S(i_B) = 2qI_B$. Bipolar transistors also produce thermal noise due to their base resistance R_B. Its power density spectrum $S(v_B)$ (in V^2/Hz) equals $4kTR_B$.

By using $g_m = qI_C/kT$ and $I_B = I_C/B_F$ we are able to transform i_C to the input into two (correlated) noise sources $v_n = i_C/g_m$ and $i_n \approx i_C/B_F$ (low-frequency

approximation). See figure 9. As $S(v_B) = 4kTR_B$ and $S(v_n)$ $(=S(i_C)/g_m^2 = 2qI_C/g_m^2)$ equals $4kT/2g_m$, v_B is negligible if $R_B \ll 1/2\ g_m$. A typical value of R_B is 500 Ω. Therefore the influence of R_B can be neglected as long as I_C does not exceed several tens of microamps. As $S(i_B) = 2qI_B$ and $S(i_n)$ $(\approx S(i_C)/B_F^2 = 2qI_C/B_F^2)$ approximately equals $2qI_B/B_F$, i_n is always negligible with respect to i_B. The remaining two (uncorrelated) sources v_n and i_B can be summed in one voltage source v_{in} in series with the capacitor using the Blakesley transformation for voltage sources and an equivalent transformation for current sources (see [12], for example). We find

$$S(v_{in}) = S(v_n) + S(i_B)\left(R^2 + \frac{1}{4\pi^2 f^2 C^2}\right) = \frac{2kT}{g_m} + \frac{2kTg_m}{B_F}\left(R^2 + \frac{1}{4\pi^2 f^2 C^2}\right) \tag{20}$$

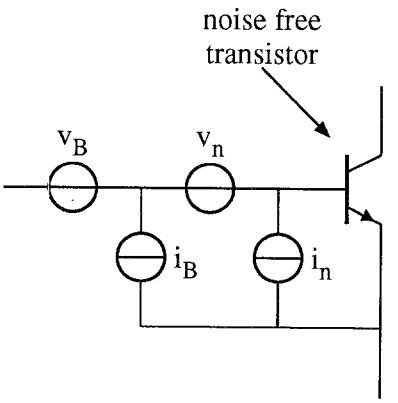

Fig. 9. Representation of the same noisy transistor by a noise-free transistor and four noise sources at the input.

In many practical situations $1/4\pi^2 f^2 C^2 \gg R^2$. Therefore $S(v_{in})$ can be approximated by

$$S(v_{in}) = S(v_n) + S(i_B)\frac{1}{4\pi^2 f^2 C^2} = \frac{2kT}{g_m} + \frac{2kTg_m}{B_F}\frac{1}{4\pi^2 f^2 C^2} \tag{21}$$

When v_{in} is transformed to the output into an equivalent current source $i_{n,eq}$ we obtain

$$S(i_{n,eq}) = S(v_{in})H_t^2 = S(v_{in})\frac{1}{n^2R^2} = \frac{2kT}{n^2}\left(\frac{1}{R^2 g_m} + \frac{g_m}{4\pi^2 f^2 B_F R^2 C^2}\right) \tag{22}$$

To obtain $i_{n,eq}$ we integrate $S(i_{n,eq})$ over the total audio frequency range (from f_1 to f_2) and square-root the result. This results in

$$i_{n,eq} = \sqrt{\frac{2kT(f_2 - f_1)}{n^2}\left(\frac{1}{R^2 g_m} + \frac{g_m}{4\pi^2 B_F f_1 f_2 R^2 C^2}\right)} \tag{23}$$

Acknowledgment

The author wishes to thank Albert van der Woerd and Rob van Beijnhem for valuable discussions and the Delft Institute for Microelectronics and Submicron Technology (DIMES) for processing the chip.

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