A Low-Power Low-Voltage Second-Order High-Pass Butterworth Leapfrog Filter

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Abstract

A low-power bipolar continuous-time low-frequency high-pass second-order Butterworth filter that works in the current domain and operates from a single 1.3-V battery is presented. 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 a standard full-custom IC process. A semi-custom realization is shown. The filter demonstrates operation with battery voltages down to 1 V with less than 16 μW power consumption and a dynamic range of 50 dB. Its cutoff frequency can be varied exponentially with a control current from 100 Hz - 1 kHz.

1 Introduction

As the size and power consumption of electronic circuits is becoming more and more important the demand for circuits that use one single battery and consume little current is increasing. Examples are hand-carried radiotelephones, pagers and hearing instruments. To improve the speech intelligibility in these systems, especially of consonants, often a high-pass filter is used. 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 - 1 kHz. In sections 2, 3 and 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 so called 'leapfrog filter' [1, 2].

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 [3] and has a current as the information carrying quantity, to reduce the influence of parasitic admittances [4, chapter 2]. The input-output relation $H(f)$ is given by:

$$H(f) = \frac{i_{out}}{i_{in}} = \frac{1}{j^2 + j\sqrt{2}f_c/f + f_c^2/f^2} \qquad (1)$$

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).

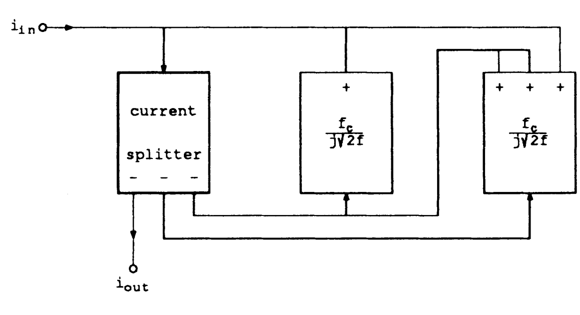

Figure 1: A second-order high-pass leapfrog filter operating in the current domain

3 The integrator blocks

Because integrating elements are only available in the form of one-ports, an integrator operating in the current domain will always contain at least one reactive and one dissipative element. Choosing a capacitance as the reactive element, Figure 2 shows an appropriate configuration: a capacitance followed by a transconductance amplifier. The amplifier has an indirect output. Hence the possible output swing is maximized. Moreover, the filter is easily tuned by changing the scaling factor n, as will be shown subsequently.

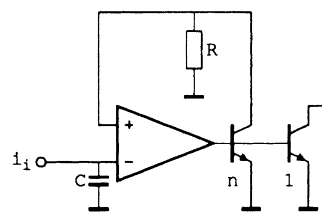

Figure 2: A capacitance-transconductance amplifier with an indirect output.

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 $v_i = \frac{i_i}{j2\pi fC}$, that in turn is transformed by Q_1, Q_2 and resistance R into a current $i'_o = v_i/R = \frac{i_i}{j2\pi fRC}$. Q_3 provides the indirect output. When the bias current through Q_3 is n times as small as the current through Q_2 (and when the Early effect is negligible) we find:

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

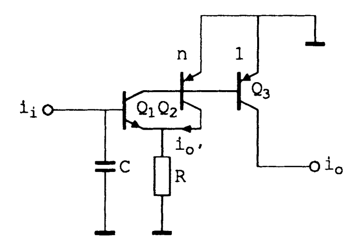

Figure 3: Circuit diagram of the capacitance-transconductance amplifier

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, $i_{n,eq}$:

$$i_{n,eq} = \sqrt{\frac{2kT(f_2 - f_1)}{n^2}\left(\frac{1}{R^2 g_m} + \frac{g_m}{4\pi^2 B_F f_1 f_2 R^2 C^2}\right)} \tag{3}$$

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, $I_{C,Q1}$, it is possible to minimize expression 3 by varying I_C. For this optimum value, $I_{C,Q1,opt}$, we find:

$$I_{C,Q1,opt} = \frac{kT}{q\sqrt{\frac{R^2}{B_F} + \frac{1}{4\pi^2 B_F C^2 f_1 f_2}}} \tag{4}$$

4.2 Collector currents of Q_2 and Q_3

The tuning of the filter is done by varying the factor n, which equals the collector current of Q_2, $I_{C,Q2}$, divided by the collector current of Q_3, $I_{C,Q3}$. As the maximum output current of the integrator does not depend on the cutoff frequency f_c of the filter and therefore does not depend on n, it is convenient to choose $I_{C,Q3}$ to be constant and vary the scaling factor n by varying $I_{C,Q2}$. The value of $I_{C,Q3}$ depends on the maximum voltage swing $v_{i,max}$ over the capacitance C. Some calculation yields:

$$I_{C,Q3} = v_{i,max} 2\pi f_{c,min} C\sqrt{2} \tag{5}$$

Because the filter has to work at voltages down to 1 V $v_{i,max}$ is chosen equal to 100 mV.

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:

$$\eta = \frac{\text{maximum signal current at the output}}{\text{total bias current of the integrator}} \tag{6}$$

or:

$$\eta = \frac{I_{C,Q3}}{I_{C,Q1} + (n+1)I_{C,Q3}} \tag{7}$$

We see that it is convenient to choose n smaller than one for all possible cutoff frequencies. With expressions 2 and 5, it follows:

$$R > \frac{1}{2\pi f_{c,min} C\sqrt{2}} \tag{8}$$

In practice this easily leads to values of R that cannot be integrated (in most standard bipolar IC processes the value of a diffused resistor is limited to several hundreds of kilo ohms). In that case the value of R can best be chosen maximally.

5 The complete filter

When we combine two capacitance-transconductance amplifiers having two current mirrors with multiple outputs according to the block diagram of Figure 1 together with its biasing circuitry, we have completed the design of the filter (Figure 4). The (adjustable) cutoff frequency of the filter is controlled by means of an adjustable current I_f, coming from an external circuit (e.g. a potentiometer or a programmable current source):

$$I_f = -\frac{V_T}{R_f} \ln 2\pi f_c RC\sqrt{2} \qquad (9)$$

GM-compensated mirrors [4, chapter 6] with multiple outputs provide the bias currents of every stage. 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}.

6 Semi-custom realization

The circuit shown in Figure 4 was integrated in a semi-custom chip in the *LA251* process [5]. Figure 5 shows a micro-photograph of the chip. The two integrating capacitors C_A and C_B were chosen to be equal to 400 pF. The two (diffused) resistors R_A and R_B were chosen to be equal to 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 equal to 300 K, 100, 100 Hz, 10 kHz and 100 Hz, respectively, this results in 1.6 μA, 84 nA, 16 kΩ and 310 kΩ for I_1, I_2, R_1 and R_2, respectively. For R_f a 6 kΩ resistor has been chosen. I_f therefore varies between 1.5 μA and 11 μA.

7 Measurements

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 6. We observe a second-order Butterworth characteristic between 100 Hz and 10 kHz. The loss in the pass-band is less than 1 dB. When the total harmonic distortion is kept less than 5 % and the bandwidth equals 10 kHz, the dynamic range amounts to 50 dB. The maximum signal current (at both the input and output) equals 25 nA (peakvalue). The filter operates well at voltages down to 1.0 V and consumes less than 16 μA. No instability occurs.

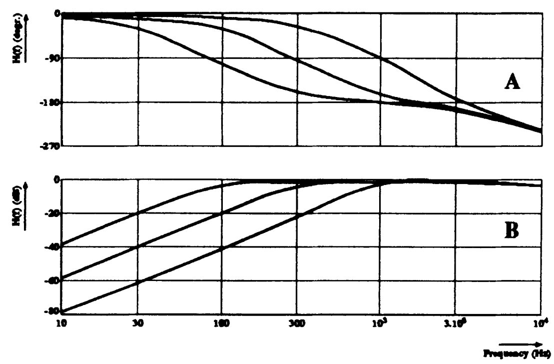

Figure 6: Phase (A) and gain (B) transfer of the filter

8 Conclusions

A low-power, low-voltage fully integrated second-order high-pass filter that works in the current domain has been presented. 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.

References

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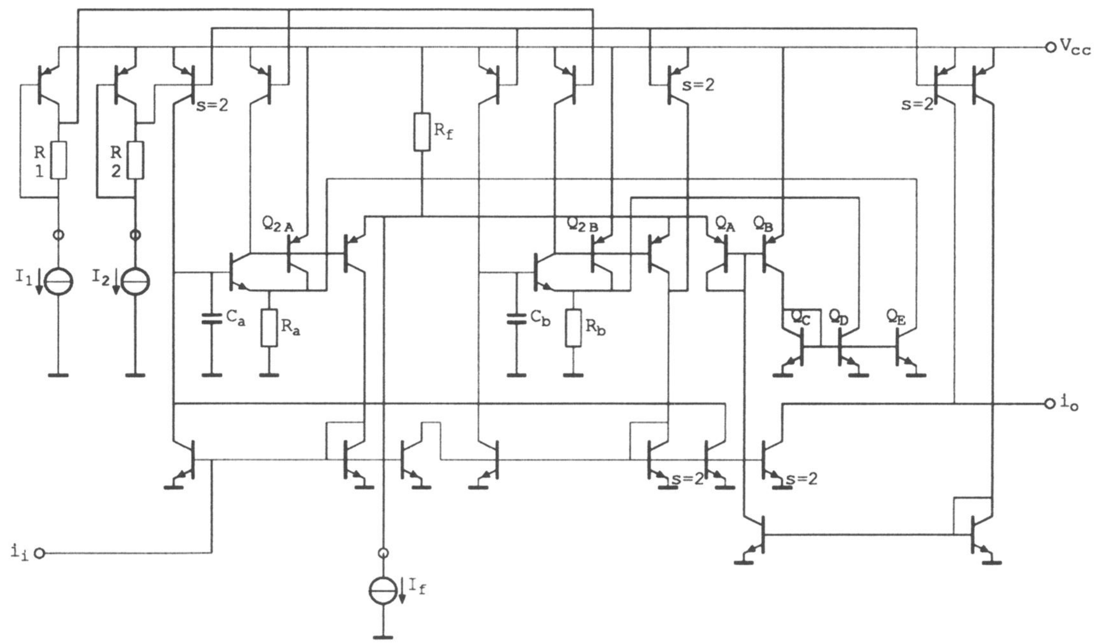

Figure 4: The filter including its biasing scheme. $s = 2$ means a doubled emitter area ratio.

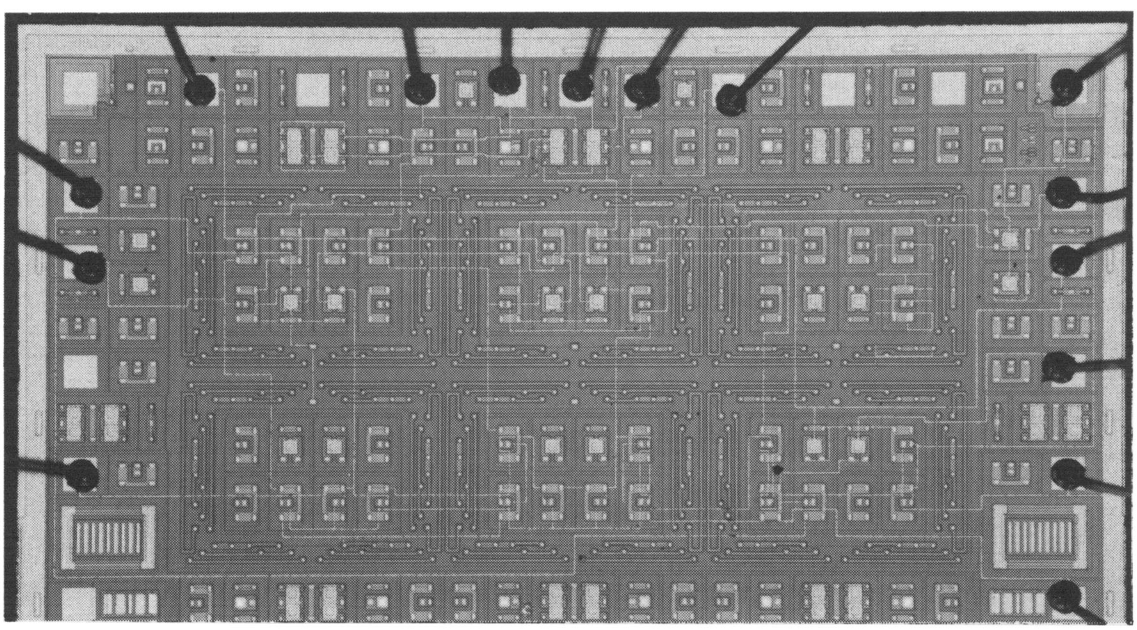

Figure 5: Micro-photograph of the semi-custom chip.