PROGRAMMABLE LOW-VOLTAGE CONTINUOUS-TIME FILTER FOR AUDIO APPLICATIONS

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ABSTRACT

The implementation of a Continuous-Time Filter (*CTF*) useful for audio frequency applications is presented in this paper. The filter functions can be programmed and tuned with two independent control variables. The filter here proposed has been designed to work at *1.5V* of power supply and at a maximum of *0.5μA/OTA* for the worst case current consumption. Electrical simulations of a Tow-Thomas Biquad (*TTB*) show the possibility of obtaining Low-Pass and Band-Pass filter functions over the *10Hz-40KHz* frequency range by changing a control current over four decades.

1. INTRODUCTION

Continuous-Time Filters (*CTF*) are an alternative to sampled-data filters, *Switched-Capacitor* (*SC*) [1] or *Switched-Current* (*SI*) [2] which require neither *pre* nor *post* aliasing filters. Furthermore, they do not need high component spread (capacitors, current gains, etc.) to implement the large time constants required in audio frequency applications. One of the main characteristics of *CTFs* is that tuning processes are necessary to control technological process variations but, on the contrary, clock-noise is absent. Traditionally, two approximations have been reported to implement *CTFs*: the *MOSFET-C* and the g_m-C [3]. The first one requires, in general, the linearization of *MOS* transistors used as resistors and it is limited to low and medium frequencies due to the finite bandwidths of the opamps. They use a reduced number of components (resistors, capacitors and opamps). The second one is limited by the small linear input range of the transconductors and requires a careful design of the *OTAs*. A novel technique uses the translinear principle to make *log-filter CTFs* [4]. Compressing and expanding current and voltage signals allow us to increase the dynamic range and to reduce the voltage node swing, so they seem to be a good candidate for low-voltage high-frequency operation [5,6].

In this paper, the realization of a g_m-C filter is presented. The filter, intended to audio frequency applications, has been based on the well-known transconductor presented in [7], which has excellent programming and tuning characteristics. The key points or *trade-offs* for the transconductor design are: 1) A power supply of *1.5V*; this means that the input voltage ranges needed for low voltage applications can be achieved by using low threshold voltage processes and/or with transistors operating in weak inversion region. Specific circuit techniques can be also used [7,8]. 2) Filter specifications for the whole audio frequency range. Large time-constants for low frequencies (some tens of *Hz*) need low g_m/C ratios, which force to low transconductance values and large integrating capacitors. 3) Fully differential input-output operation are required in order to reduce harmonic distortion. This requires common-mode voltage feedback to ensure that the input and output voltages are at the same quiescent voltage level, thus reducing the effects of process parameter variations.

Next, the design of an operational transconductance amplifier (OTA) with transistors working in weak inversion will be presented. It has been thought to decrease the coefficients of the g_m/C integrators. The circuit has been designed to work at *1.5V* in a low threshold voltage process (V_{tn} *~0.56V*). It has two control variables: a short range tuning variable given by a V_g voltage, and a large range programming variable, given by an I_B biasing current. Thus, the full audio-range can be selected with the same topology. A Tow-Thomas Biquad (*TTB*) section is built using this integrator, obtaining the frequency control expected, and the *Q*-tuning derived from the *TTB* topology.

2. OTA DESIGN

The transconductor used in this work is shown in Fig. 1. The operation of the circuit is based on the fact that the *M1A-M1B* transistors in the input pair work as source followers, applying to the source resistance R_s (the

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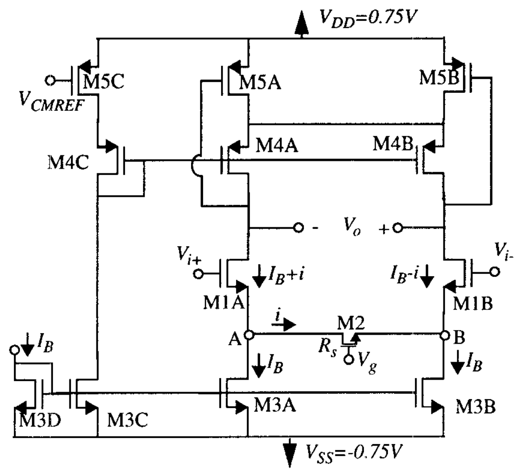

Figure 1: Transconductor circuit.

resistance of *M2* operating in the linear region) a proportional voltage to the input. The equivalent small signal transconductance of the circuit [7] is approximately given by,

$$G_m = \frac{i}{v_i} = \frac{1}{\frac{2}{g_{m1}} + \frac{1}{g_s}} \qquad (1)$$

where $v_i = V_{i+} - V_{i-}$, $g_s = 1/R_s$, and g_{m1} is the small-signal transconductance of both *M1A* and *M1B*. The *MOS* implementation of R_S allows us to define its value by a control voltage (the gate voltage, V_g, of the *MOS* transistor working at the linear region), resulting in the tuning of the G_m value through V_g. When both *M1A* and *M1B* operate in strong inversion mode, the g_{m1} value is given by the expression,

$$g_{m1} = \sqrt{2\mu_n C_{ox}\left(\frac{W}{L}\right)_1 I_B} \qquad (2)$$

where I_B is the biasing current, which can be also used as a control variable for tuning processes. The source follower operation of the *M1A-M1B* pair is also obtained when both transistors operate in weak inversion saturation region, resulting a transconductance value of,

$$g_{m1} = \frac{I_B}{nU_t} \qquad (3)$$

being U_t the thermal voltage (*KT/q*) and having *n* a value between *1* and *2*. For the same decreasing values of I_B current, eq. (3) provides lower proportional values of the transconductance g_{m1} than eq. (2). On the other hand, the source follower performance relies on large $(W/L)_1$ aspect-ratios, which lead to increasing g_{m1} values in strong inversion operation. For weak inversion, the trade-off can be obtained by decreasing I_B in eq.(3) and increasing the $(W/L)_1$ aspect ratio independently.

An approximated close expression for the small-signal transconductance valid in saturation for strong and weak modes has been reported in [9]. This can be resumed as,

$$g_m = \frac{I_B}{nU_t} \cdot \frac{1 - e^{-\sqrt{IC}}}{\sqrt{IC}} \qquad (4)$$

where *IC* is the *Inversion Coefficient* defined as the relation I_{Dsat}/I_S, and $I_S = 2nU_t^2\beta$, with $\beta = \mu C_{ox}W/L$. When the *IC* value is below *0.1*, it can be considered as a weak inversion operation, while for values greater than *10*, it is thought to be a strong inversion operation. In Fig. 2, the *IC* evolution is represented when the aspect-ratio of a *NMOS* transistor changes for different values of I_{Dsat}. This graphic gives us a guide to select the transistor aspect-ratio for a given biasing current and an operation mode of the transistor. For weak inversion mode, low currents and large aspect-ratios are required, while for strong inversion, large currents and very small aspect-ratios. The sizeas of the transistors must agree with the circuit specifications, that is, transconductance values must lead to the audio frequency range required. For a frequency range of $f \in [10Hz, 40KHz]$, and an integrating capacitor of *5pF*, the corresponding minimum and maximum transconductances ($g_m = 1/2\pi Cf$) are close to *0.3nA/V* and *1.2μA/V*, respectively. Fig. 3 represents the transconductance given by eq. (4) for the same values of *(W/L)* and I_{Dsat} than in Fig. 2. The boundaries for each mode of operation are marked. Using strong inversion does not fulfill the filter specifications. However, weak inversion includes the whole range of transconductance for a wide range of *(W/L)* values. For this application, weak inversion mode is necessary, and an aspect-ratio of *20* was chosen in the design.

The correction of the transconductor common-mode voltage deviations is performed by the transistors *M4* and *M5*. They define the output common-mode voltage, V_{CM}, by means of a reference voltage, V_{CMREF}, established by a bias circuit. For a $\pm 0.75V$ power supply range, the calculated input common-mode is located in the range of *0* to *250mV* approximately, for a bias current of *200nA*. The common-mode feedback gain for $R_s = 100K\Omega$ is *35.1dB*. Decreasing I_B means to make g_{m1} dominant versus R_s in the G_m equation. Thus, the decreasing values of I_B will be transformed into approximately the same decreasing quantity on the cut-off frequency, being possible to implement the lower audio frequencies, about *10Hz*, with bias currents of *20pA*.

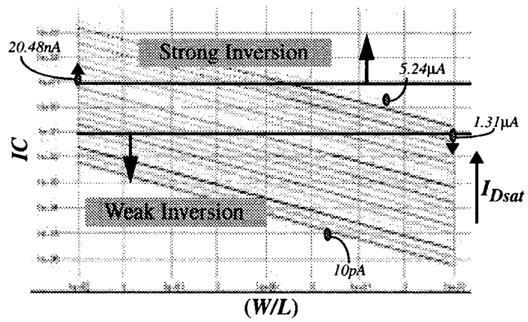

Figure 2: Inversion coefficient (*IC*) versus aspect-ratio (*W/L*) when I_{Dsat} changes from *10pA* to *10μA*.

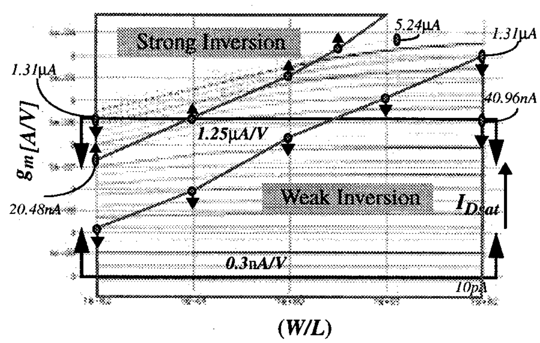

Figure 3: Transconductance versus aspect-ratio (*W/L*) when I_{Dsat} changes from *10pA* to *10μA*.

3. TOW-THOMAS BIQUAD

Here, the use of the transconductor to implement a *TTB* for audio frequencies is presented. The biquad schematic is shown in Fig. 4. It has two outputs allowing two different filters functions: a low pass function (V_{LP}/V_{in}) and a band pass function (V_{BP}/V_{in}), given by

$$\frac{V_{BP}(s)}{V_{in}(s)} = -\frac{s \cdot g_{m1}/C_1}{s^2 + s \cdot \frac{g_{m2}}{C_1} + \frac{g_{m3}g_{m4}}{C_1C_2}} \tag{5}$$

$$\frac{V_{LP}(s)}{V_{in}(s)} = \frac{(g_{m1}g_{m3})/(C_1C_2)}{s^2 + s \cdot \frac{g_{m2}}{C_1} + \frac{g_{m3}g_{m4}}{C_1C_2}} \tag{6}$$

where the typical filter parameters of a second order section are,

$$\omega_o = \sqrt{\frac{g_{m3}g_{m4}}{C_1C_2}} \tag{7}$$

$$Q = \frac{\sqrt{g_{m3}g_{m4} \cdot C_1/C_2}}{g_{m2}} \tag{8}$$

The simplest design choice is to take all $g_{mi} = g_m$ $(i=1,2,3,4)$ and $C_1=C_2=C$. This means that the cut-off frequency is $f_o=g_m/2\pi C$ and $Q=1$. For $f_o=100Hz$, and a capacitor of $C=5pF$, a $g_m=3.1nA/V$, is needed. From eq. (2), the value $I_B=122pA$ is necessary. For strong inversion mode, these values lead to excessively low $(W/L)_1$ ratios; so weak inversion mode seems to be mandatory.

Electrical simulations have been performed for a *20mV* peak-to-peak differential input voltage with an I_B current range from *20pA* to *200nA* and a power supply of *1.5V* (*HSPICE BSIM3* models have been used for *MOS* transistors). The transfer functions obtained are shown in Fig. 5. Both low-pass and band-pass functions are in the frequency range of *7Hz* to *45KHz*. For four decades of change in I_B, the cutoff frequency moves *3.8* decades approximately. This dependence can be explained by eq. (1), considering $g_s >> g_{m1}$, and weak inversion mode for *M1A* and *M1B*. This can be observed in Fig. 6, where we have represented the g_{m1} transconductace versus the biasing current I_B.

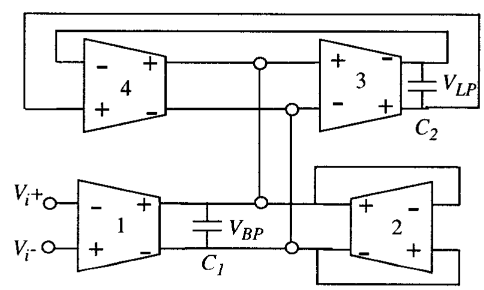

Figure 4: Tow-Thomas Biquad.

Tuning can be also performed changing the V_g value. Fig. 7 shows three set of curves for both *LP* and *BP* transfer functions. The bias currents for *OTAs* are *20pA*, *2nA* and *200nA* from low to high frequencies respectively, while V_g moves in the *[-0.1V, 0.1V]* range. Finally, according to eq. (8), the *Q-factor* can be exclusively defined by g_{m2}. Fig. 8 shows this effect for two filter transfer functions. The biasing current of transconductor 2, I_{B2}, has been changed from *20pA* to *20nA*, obtaining smaller and higher values of *Q* than the unity, which makes the *Q-tuning* of the filters feasible and easy.

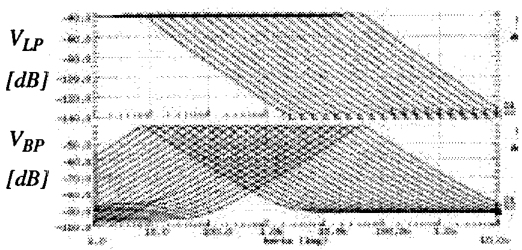

Figure 5: Transfer functions for $Q=1$.

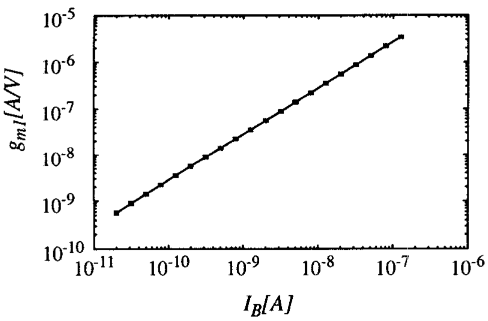

Figure 6: g_{m1} versus I_B.

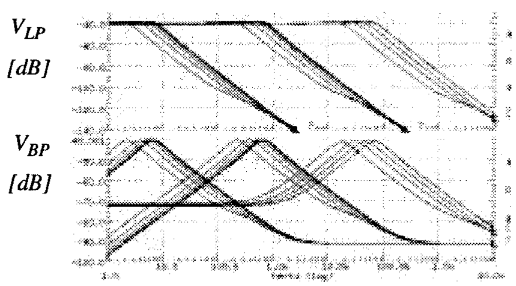

Figure 7: Frequency tuning through V_g variable. I_B=(20pA,2nA, 200nA) V_g=[-0.1V,0.1V].

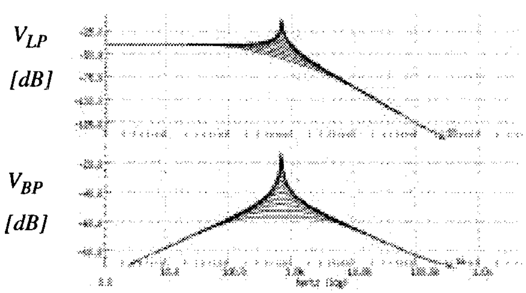

Figure 8: *Q-tuning* through g_{m2}. For $I_{B1} = I_{B3} = I_{B4}$ =2nA, and $I_{B2} \in$ [20pA,20nA].

An important filter parameter is *THD*, which gives a measure of the linear input output relationship. Simulations have been realized over a filter of the *LP* family with a cutoff frequency about *6KHz* and a sinusoidal input signal of *500Hz*. Harmonic cancellation of the even components is obtained, reducing the total distortion below *47dB* for an peak-to-peak differential input voltage of *100mV*.

4. CONCLUSIONS

The design of an Operational Transconductor Amplifiers for audio-filter applications and *1.5V* of power supply has been presented. The suitability of weak inversion operation for programming and tuning processes has been proved. Electrical simulation results show the expected transfer functions for a Tow-Thomas Biquad, with excellent ω_o and Q tuning characteristics. The maximum power consumption is below *0.75µW/OTA* for the maximum bias current.

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