NONVOLATILE CORRECTION OF Q-OFFSETS AND INSTABILITIES IN COCHLEAR FILTERS

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ABSTRACT

We present a feedback circuit that performs nonvolatile correction of instabilities and resonant-gain offsets (Q-offsets) in individual cochlear filters. The subthreshold CMOS circuit adapts using analog floating-gate technology. We present experimental data from a working chip that illustrates the performance of the circuit. We discuss how to extend our work to do very long-term gain control in the silicon cochlea. Positive-feedback circuits, such as our cochlear filters, are very sensitive to parameter variations. This potential problem becomes an advantage in our corrective feedback loop where the hypersensitivity behaves merely like high loop gain.

1. INTRODUCTION

Second-order active filters are the basic elements of several filtering circuits. In this paper, our interest in them stems from their importance as the building blocks of electronic cochleas. The electronic cochlea is a cascade of nonlinear second-order filters with exponentially increasing time constants and a constant Q, set by a global bias voltage [1]. The filters use transconductance amplifiers with positive feedback to model the gain provided by outer hair cells in the biological cochlea. As is well known, amplifiers with positive-feedback are very sensitive to parameter variations, and are apt to go unstable. In electronic cochleas with cascades of 100–200 filters it is often hard to operate all the second-order filters with even modest Q's because parameter variations across a chip can induce instability in some errant filters; the spurious activity of these filters then propagates through the cascade and masks other legitimate signals. In order to keep the errant filters stable, it is necessary to operate all the other filters with conservatively low Q's.

In the type of positive-feedback second-order filter described above, the Q of the filter is a function of a parameter α, the ratio of two currents, and is given by

$$Q = \frac{1}{(2-\alpha)} \quad (1)$$

In eq. (1), $Q \to \infty$ as $\alpha \to 2$. For $\alpha > 2$, the filter is unstable. Thus if the α parameter is 1.0 at a particular filter, and 2 at another, due to offsets, the first filter will have a Q of 1 and the second filter will be unstable.

Active filters that use negative feedback and no positive feedback can be constructed, but they typically have higher harmonic distortion than ones built with positive feedback: In negative-feedback filters, $Q = \sqrt{\tau_2/\tau_1}$, and the corner frequency $\omega_c = 1/\sqrt{\tau_2\tau_1}$, where τ_1 and τ_2 refer to two different time constants in the filter. The differing time constants in the filter cause intermediate nodes in it to have large voltage swings even if its output appears to be well behaved, and the distortion in the filter is higher. The higher the Q, the larger is the separation between the two time constants, and the higher is the distortion. Negative-feedback filters suffer from instability problems as well, because of parasitic capacitances. Further, it is hard to change the Q's of these filters without undesirably changing their corner frequencies as well. In positive-feedback filters, the corner frequency and Q parameters are orthogonal and it is easy to change Q without affecting the corner frequency of the filter. It is also easier to get an enormous range of gain control for a small change in a system parameter in positive-feedback filters.

Instead of avoiding positive feedback in constructing its filters, Nature has evolved a solution that uses nonlinearity (in outer hair cells) and negative feedback (the olivocochlear efferent system) to control positive feedback. By and large, the auditory system does a superb job of keeping the cochlea stable even at the very high gains necessary to detect sub-Å motions at our eardrums. Nevertheless, in some humans, the instability in the cochlea is uncontrolled and gives rise to a persistent ringing in the ear called tinnitus.

In [2] we describe a cochlea with second-order active filters that performs nonlinear gain control by adapting its filters' Q's with input amplitude. In this paper, we describe a nonvolatile adaptation technique that controls instability, and compensates for Q-offsets in an array of these filters. In the present work the filters are isolated and non-interacting. In section 5. we discuss how to extend this work to do long-term gain control in the silicon cochlea where these filters interact in a cascade. Our nonvolatile adaptation uses the analog floating-gate technology reported in [3], involving tunneling and hot-electron injection, and is the first system-level application of this technology.

2. OVERVIEW

In our scheme, we first shut off all input to the filters and change their α parameters, so that all of them are unstable. We detect instability in each filter with an inner-hair-cell circuit that compares the filter's level of unstable activity with a threshold level of activity. If its level of activity is

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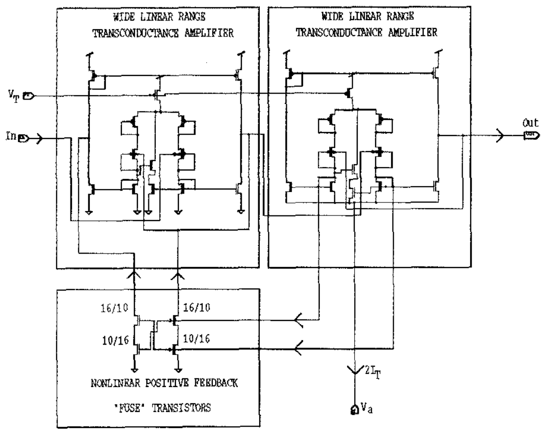

Figure 2. The figure shows the Fuse second-order filter circuit that is described in detail in a companion paper in this conference [2]. It is also briefly described in the text of this paper.

$$\Rightarrow I_{hr} \quad = \quad (\omega C_{hr} V_0 \cos(\omega t))\, u\left(-\frac{dV_{out}}{dt}\right) \qquad (5)$$

where V_0 represents the amplitude of the unstable sinusoidal oscillation of the fuse filter, ω represents its frequency, and $u(x)$ is the unit step function.

The threshold-comparison circuit compares I_{hr} with a current that is proportional to I_T, the bias current of the filter shown in Figure 2. The proportionality comes about because the gate voltage V_T that determines the bias current of the SOS filter is identical to the gate voltage of the transistor carrying the threshold-comparison current. The corner frequency of the filter ω_c is given by

$$\omega_c = \frac{I_T}{CV_L} \qquad (6)$$

where C is the capacitance in the SOS filter and V_L is the linear range of the transconductor in the filter. Thus, we have

$$I_{th} \quad = \quad I_T e^{-q(V_{DD}-V_{th})/kT}, \qquad (7)$$

$$\quad = \quad \omega_c C V_L e^{-q(V_{DD}-V_{th})/kT} \qquad (8)$$

where V_{th} is the bias voltage in Figure 1. When the filter is unstable and oscillating, its frequency is at or proportional to ω_c. Thus, when the threshold-comparison circuit compares the hair cell current of eq.(5) with the threshold-comparison current of eq.(8) the ω_c's cancel and the comparison is scale invariant. In other words, the output of the threshold-comparison circuit depends only on the amplitude of oscillation V_0 and not on the frequency of oscillation. The voltage V_{th} controls the threshold level of activity, and is typically 100-200 mV below V_{DD}. The slow floating-gate circuitry filters out the cycle-by-cycle variations in I_{hr}, so that the overall feedback correction happens gradually.

The output of the threshold-comparison circuit feeds into an inverter that controls the drain voltage of an NFET transistor in strongly-doped P-substrate. This transistor is responsible for hot-electron injection, and is marked with a circle in Figure 1. We use the Pbase layer that is normally used to construct bipolar transistors as our strongly-doped P-substrate. The magnitude of the injection currents and injection rate may be increased/decreased by lowering/raising the bias voltage V_{ig}. The maximum value of the injection drain voltage is determined by the bias voltage V_{im}; this voltage is not set at $V_{DD} = 5$ V but at about 3.5 V so that it is always lower than the final floating-gate values in our circuit (3.7 V—4.5 V) [3]. The tunneling voltage $V_{tn} \approx 40$ V couples to the floating-gate node via the capacitor C_{sm} and the floating-gate charge is stored on a capacitor called C_{BG}. We use tunneling only to initialize all floating gates to V_{DD} and then turn it off. The floating-gate voltage V_{fg} feeds into the alpha-control circuit described separately in Figure 3.

The floating gate adds a correction current I_{fg} that sums with the bias current I_{QT} in Figure 3 and lowers V_a. The voltage V_a is operated a few mV above Ground, so that the only NFET of this circuit is never in saturation. Recall that V_a determines α and thus Q from eq. (2). If, for the sake of simplicity, we assume that all the NFETs and PFETs have subthreshold exponential coefficients of 1.0, and that the NFETs and PFETs have identical conductance strengths, then we get a simple form for the dependence of V_a on the correction current I_{fg}:

$$V_a = \frac{kT}{q} \ln\left(1 + \frac{I_T}{(I_{fg} + I_{QT})}\right) \qquad (9)$$

In practice, these are oversimplifying assumptions, and the dependence of V_a on I_{fg} is more complicated and unsolvable in closed form. However, the qualitative form of the dependence, as revealed by experimental measurements on a separate test circuit, remains the same. In particular V_a is still a function of $I_T/(I_{fg} + I_{QT})$ and the overall shape of the function is very similar. The circuit lowers V_a when I_{fg} rises, V_a is always close to ground so that the amplifiers in the filters are in a well-defined regime of operation, the power consumption is only as high as it needs to be because I_{fg} and I_{QT} scale with the bias current I_T, and all of these nice properties are obtained with four transistors.

4. PERFORMANCE

Figure 4 shows the Q's of an array of 21 filters before and after nonvolatile adaptation. Before nonvolatile adaptation, we notice that one filter has a Q less than 2 but several others have Q's that are near 4 and 5 and the pattern of Q-offsets is erratic. After nonvolatile adaptation and a global lowering of α to 1.5, all the filters have Q's very close to 2. Any offsets in Q that remain at the edge of instability (due to mismatches in the adaptation circuitry, and the overshoot of the injector past the instability point), translate into small mismatches in α; this is mathematically illustrated by inverting eq. (1) with α near 2.0. When we lower α globally, the Q's at the lower values of α, say $\alpha = 1.5$, match well because Q is not hypersensitive to α, when α is near 1.5; *i.e.*, the small mismatches in α that remain after adaptation don't matter very much for low Q's where we operate these filters. Thus, we have exploited the extreme sensitivity of positive feedback to parameter variations, by

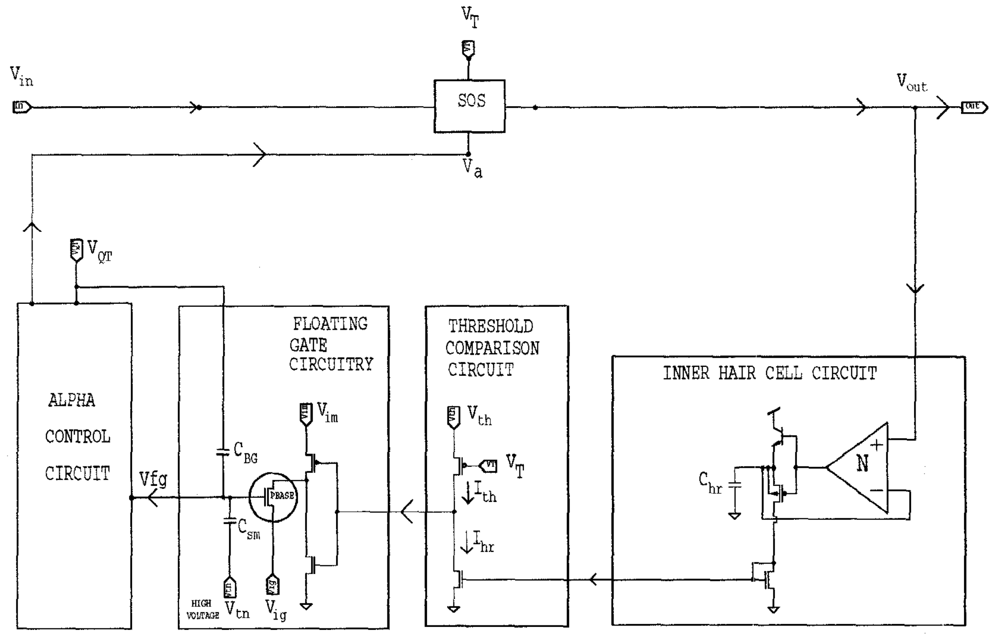

Figure 1. The figures shows the overall system-level circuit. The SOS circuit is shown separately in Figure 2 and the alpha-control circuit is shown separately in Figure 3. A description of the system may be found in the text.

greater than this threshold level, we turn on an injector, and inject electrons onto a floating gate that has been initialized to a value near V_{DD} by tunneling. The injection causes the floating gate voltage to drop, and the current in a PFET tied to this floating gate to increase. The increasing current in the PFET decreases the α parameter of the filter by adding a correction current to an alpha-control circuit. As α decreases, the amplitude of the unstable waveform decreases. When α drops just below 2, the instability vanishes, the amplitude of the unstable waveform drops abruptly below threshold, and the injector shuts off. At this point all filters are just balanced on the edge of instability with α's very near 2. We now decrease α globally for all filters, by a constant fraction. If this fraction was 0.75, all filters would have α's near 1.5, and Q's near 2. The injector is now deactivated permanently, and the filters are ready to use with real inputs.

3. CIRCUIT DETAILS

In Figure 1 there are five blocks: the second-order filter section labelled SOS, the activity sensing inner-hair-cell circuit, a threshold-comparison circuit, a floating-gate circuitry block, and an alpha-control circuit. The alpha-control circuit translates the floating-gate correction into a change in the α, and thus the Q of the filter. We shall now describe each block in a separate paragraph.

The second-order filter is described in detail in a companion paper at this conference [2]. Figure 2 is a reproduction of the circuit from that paper. The boxes labelled Transconductance Amplifier and Nonlinear Positive Feedback implement the Fuse second-order filters described in that paper. The corner frequency of the filter is proportional to the bias current I_T, which is in turn determined by the bias voltage V_T. The voltage V_a sets the α, and thus the Q of the filter; it is controlled by adaptation circuitry, shown in Figures 1 and 3. The relationship between α and V_a is given by

$$\alpha = \left(\frac{w_2 w_1}{w_2 + w_1}\right)\left(\frac{w_2 - w_1}{w_2 + w_1}\right) e^{\frac{qV_a}{kT}} \tag{2}$$

where $w_2 = 16/10$, and $w_1 = 10/16$ are the W/L ratios of the fuse transistors described in [2], and kT/q is the thermal voltage.

The negative input of the amplifier in the hair-cell circuit closely follows the positive input, except for very small input amplitudes (less than about 1–2 mV) that we shall not concern ourselves with. The top/bottom transistor thus charges/discharges the capacitor C_{hr} so that the negative input is at V_{out}. The bottom transistor's current is mirrored to form I_{hr}, the output of the haircell. Thus, the inner-hair cell senses a current proportional to the derivative of the output of the filter, on the negative half cycles of the output—it functions as a V-to-I transducer, rectifier and differentiator, all-in-one. Only a.c. activity in the filter's output is sensed and reported as a current. Or in equations,

$$I_{hr} = \left(C_{hr}\frac{dV_{out}}{dt}\right) u\left(-\frac{dV_{out}}{dt}\right) \tag{3}$$

$$V_{out} = V_0 \sin(\omega t) \tag{4}$$

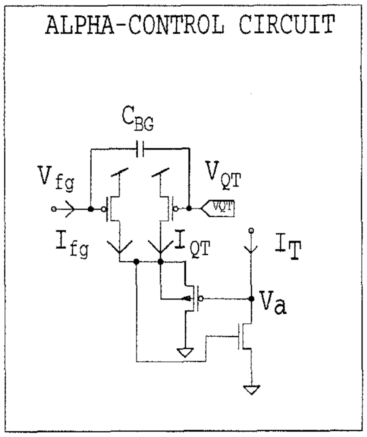

Figure 3. The alpha-control circuit is shown in the figure. A description of the circuit may be found in the text.

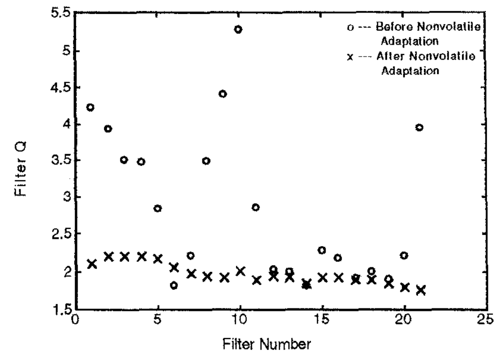

Figure 4. The figure shows that after nonvolatile adaptation and a global lowering of α to 1.5, the Q-offsets are compensated for permanently, and the Q's of all filters are near 2.0.

taking advantage of the high loop gain at the edge of instability.

Note that the floating gate capacitor in Figure 3 is referenced to the bias voltage V_{QT} rather than to ground. When we lower α globally after adaptation, we do so by decrementing the V_{QT} voltage in each filter by a constant amount. Since the floating gate is referenced to V_{QT}, it also decrements by the same amount, provided parasitic capacitance is negligible. In the subthreshold regime of operation, a constant change in gate voltage translates to a constant fractional change in current. Thus, we decrease I_{QT} and I_{fg} by the same fraction. Now, all the filters have adapted to the point where the ratio $I_T/(I_{fg}+I_{QT})$ puts V_a and thus α near 2. If we increase I_{fg} and I_{QT} in the same proportion, then the value of V_a and thus α are lowered to nearly the same value for all filters from eqs. (2) and (9).

5. EXTENSIONS OF OUR WORK TO THE COCHLEA

In the silicon cochlea, where the cochlear filters are arranged in a cascade, multiple filters can go unstable at once. It is important when adapting the Q's to distinguish between two classes of activity at the outputs of the cochlear filters:

1. Activity due to a local instability.
2. Activity due to the propagation of an unstable signal from previous filters in the cascade, or from a real legitimate input.

If these classes are not distinguished appropriately, there may be insufficient or unnecessary adaptation of Q. Sensing the output activity of a cochlear filter alone does not provide enough information to distinguish between the two possibilities. One must compare the sensed activity at a given location with that at a previous location to distinguish between the possibilities. An abrupt change in the amount of activity between a filter's input and output is a sign that the filter is unstable or has a very large Q, that must be corrected for by feedback. A mild increase in activity, especially at low signal levels must not be corrected for strongly, since it can hinder the amplification of small signals. There are many possible gain control strategies that one can pick and we are exploring some of them. All sensible schemes have to involve spatial interaction and nonlinearity. The cochlea is a collective system, and it is not important that the Q's of all the filters be regulated to exactly the same value, since the overall transfer function at any cochlear location will average out small variations in Q over the preceding cochlear taps. Cochlear adaptation must be continuous and on a slow time scale to compensate for drifts and temperature variations. This implies that the small tunneling and injection currents must both be continuously active, in order to constantly adapt floating gates in the upward or downward direction.

Our inner-hair-cell circuit, floating-gate correction circuit, and alpha-control circuit will generalize to a cochlear gain-control scheme. However, in a cochlear gain-control scheme, our threshold-comparison circuit would need to be replaced by a more complicated block that implements the desired nonlinear interaction between adjacent hair-cell currents.

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