AC amplifiers with ultra-low corner frequency by using bootstrapping

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A novel architecture for an AC (i.e. high-pass) amplifier is proposed allowing a drastic reduction of the cutoff frequency to the sub-Hertz range. It builds upon the classic AC configuration with a high gain amplifier and a parallel RC circuit in the feedback loop, by increasing the feedback resistance through bootstrapping. Resistance multiplying factors higher than four orders of magnitude are easily achievable. The basic principle can be applied to several practical implementations, though in this letter it is demonstrate with measurement results of an op-amp based discrete implementation.

Introduction: Electronic circuits and particularly amplifiers, which provide DC or very low frequency blocking, have received considerable attention in the last decades. This is not strange because they are needed in a host of applications, amongst them, seismic amplifiers and remarkably, bioamplifiers for ECG, EEG, or EMG signals [1, 2]. Implementations are highly application dependent, which has resulted in a large number of practical circuits that are difficult to categorize and classify. The type of architecture, the continuous time, mixed and chopped nature, the differential or single ended character, the implementation in either integrated or discrete form, or the technology used, are criteria which hardly help to clarify the picture. References [3, 4] are examples of recent attempts to review and classify ‘low frequency and offset rejection circuits’, mainly focused on EEG signals.

Architectures for ultra-low cutoff frequency: Having in mind this scenario, there are two facts that are almost constant and true for most designs. First, the use of the well-known architecture shown in Figure 1 based on a high-gain amplifier (e.g. op-amp), with gain settled by the ratio of two capacitors, $C_1/C_2$ and cutoff frequency defined by the feedback capacitor $C_2$ and resistor $R$; and second, resorting to mechanisms, either active or passive, to increase the feedback resistance in order to obtain a very low cutoff frequency.

Focusing on this second issue there are many known ways to implement a high-valued feedback resistance ranging from the always effective resistive T-network to voltage controlled resistors with MOS transistors in different configurations and operation regimes [5, 6]. It is important to note, however, that a symmetrical floating resistor is not needed, and what is really relevant is the resistance ‘seen’ from the inverting terminal of the amplifier, which determines the cutoff frequency. This fact suggests a modification of the basic architecture in Figure 1, as shown in Figure 2a. Assuming a finite, but high, gain $A$ for the amplifier, the equivalent grounded resistance value is boosted by the reduced voltage value between its terminals by a factor $A$. Whereas amplifiers in Figures 1 and 2a share the same bandpass gain, given by:

$$V_{out}/V_{in} = \frac{C_1}{C_2} \frac{1}{1 + \frac{1}{A} \left( 1 + \frac{C_1}{C_2} \right)}.$$

For the sake of demonstration of the capabilities of the proposed architecture in Figure 3, we will show here the implementation of a discrete circuit making use of commercial op-amps, shown in Figure 4. Assuming constant gains $A_1$ and $A_2$ for the op-amps, it is routine to calculate its transfer function as:

$$V_{out}/V_{in} = \frac{C_1}{1 + \frac{1}{A_1} \left( 1 + \frac{C_1}{C_2} \right)} \frac{1}{1 + \frac{1}{ARC_2(1+A_1)(1+A_2)} \left( 1 + \frac{1}{ARC_2} \right)}.$$

Passband gain is the same as given in Equation (1) and thus close to the ideal value $-C_1/C_2$ when $A_1$ is high. Cutoff frequency depends now,
under the same condition by the following expression:

\[ \omega_c \approx \frac{1}{A_1 (1 + A_2) RC}, \]  

which clearly shows how cutoff frequency can be dramatically reduced from the value settled by the passive components \( R \) and \( C_2 \). This result can be easily explained along the lines drawn in the previous section: the buffer made with op-amp 2 having gain \( K = A_2/(1 + A_2) \approx 1 \) boosts resistance \( R \) rendering an input impedance \( (1 + A_2) R \) for the block marked in dashed lines. Both Miller and bootstrapping effects can be invoked to interpret this result [7]. This grounded impedance is further boosted by the effect of the reduced voltage swing across op-amp 1 input, with an additional factor \( A_1 \).

**Measurement results:** The circuit in Figure 4 has been extensively tested to measure its frequency response and demonstrate its potential to achieve such anticipated low corner frequencies. The results shown here correspond to cutoff frequencies in a range from approximately 20 mHz to 2 Hz, with gain values between 20 and 40 dBs. Capacitor \( C_2 \) has been varied from 1 to 10 nF whereas \( R \) is in the range of few kΩ. The particular values for the curves shown in Figure 5 are provided in the upper part of Table 1.

The measured responses have been obtained with a dynamic signal analyzer, HP 89440A, though using an external signal source HP 33522A to generate sine sweeps, instead of using the internal source of the analyzer. In this way, we had more freedom to improve the measurements accuracy. Anyway, quality of measurements worsens as the cutoff frequency gets lower as can be appreciated in the spikes around corner frequencies. This is mainly due to the limited acquisition time of the instrument, and the constant-frequency resolution inherent to dynamic analyzers. Note also that measuring frequencies in the mHz range requires acquisition times as long as several minutes, which multiply if averaging or other kind of estimator is used.

The experimental frequency response curves have been fitted by splines and from them the 3 dB cutoff frequency is estimated. Besides, cutoff frequency has been estimated by means of transients (using a Tektronix TDS 5104 digital oscilloscope), measuring time constants assuming a first order exponential response under a step excitation. In this latter case, we note that the results slightly differ for positive and negative steps, bringing out the asymmetrical behaviour of the op-amps. In Table 1 we also show the cutoff frequency estimations for both frequency and time domain measurements, compared against the reference value 1/(2πRC). The reduction factor in the cutoff frequency is very high, far beyond four orders of magnitude, though lower than the theoretical factor given by the product, \( A_1 (1 + A_2) \), of the DC gain of the op-amps. In these measurements we have used the TL081 op-amp with typical DC gain around 150000. Extensive simulations with both Spice and symbolic analysis have shown that most of the discrepancy can be explained by the effects of the op-amp dominant pole, which reduces its effective gain and, remarkably, op-amp output impedances.

**Conclusion:** Based on the well-known basic topology of the AC (high-pass) amplifier, we have proposed a novel architecture with the potential to dramatically reduce the cutoff frequency by means of bootstrapping. We have theoretically and practically demonstrated the principle of operation with a discrete op-amp implementation.

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### Table 1. Configuration and measurement values of the proposed ultra-low cutoff frequency AC amplifier in Figure 4

| \( C_2 \) (nF) | \( C_1 \) (nF) | \( R \) (kΩ) | \( r \) (Ω) |
|--------------|--------------|------------|-----------|
| 1            | 0.98         | 98.2       | 5.57      | 465       |
| 2            | 0.98         | 98.2       | 38.4      | 465       |
| 3            | 9.55         | 97.2       | 5.57      | 465       |
| 4            | 9.55         | 97.2       | 38.4      | 465       |

\( 1/(2\pi RC) \) Cutoff frequency estimation by transient measurements

\( Av \) Cutoff frequency estimation by frequency sweep measurements

![FIG 4](image4.png) Schematic of the proposed ultra-low corner frequency AC amplifier

![FIG 5](image5.png) Measured frequency response of the proposed AC amplifier (Figure 4) in four different configurations (Table 1)
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