A very low noise, high accuracy, programmable voltage source for low frequency noise measurements

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In this paper an approach for designing a programmable, very low noise, high accuracy voltage source for biasing devices under test in low frequency noise measurements is proposed. The core of the system is a supercapacitor based two pole low pass filter used for filtering out the noise produced by a standard DA converter down to 100 mHz with an attenuation in excess of 40 dB. The high leakage current of the supercapacitors, however, introduces large DC errors that need to be compensated in order to obtain high accuracy as well as very low output noise. To this end, a proper circuit topology has been developed that allows to considerably reduce the effect of the supercapacitor leakage current on the DC response of the system while maintaining a very low level of output noise. With a proper design an output noise as low as the equivalent input voltage noise of the OP27 operational amplifier, used as the output buffer of the system, can be obtained with DC accuracies better that 0.05% up to the maximum output of 8 V. The expected performances of the proposed voltage source have been confirmed both by means of SPICE simulations and by means of measurements on actual prototypes. Turn on and stabilization times for the system are of the order of a few hundred seconds. These times are fully compatible with noise measurements down to 100 mHz, since measurement times of the order of several tens of minutes are required in any case in order to reduce the statistical error in the measured spectra down to an acceptable level. © *2014 AIP Publishing LLC*. [http://dx.doi.org/10.1063/1.4870248]

I. INTRODUCTION

Low Frequency Noise Measurements (LFNMs) have proven to be one of the most sensitive tools that are available for the characterization of the quality and reliability of electron materials and devices.[1–5] The ability to reveal the voltage and/or current fluctuations that are induced at the ends of a device because of the smallest perturbations in an otherwise stationary conduction regime has also suggested the possibility of employing LFNM for the realization of selective sensor in what has become to be known as the Fluctuation Enhanced Sensing (FES) approach.[6,7] Most of the microscopic phenomena we may be interested in are characterized by power spectra decreasing with frequency, and therefore their effect is masked, at higher frequencies, by the thermal noise that does not contain any useful information. This is the main reason why we are interested in the investigation of the noise at low frequencies, where $1/f$ noise or Lorentian components produce spectral contributions that are well above the thermal noise level. In principle, there is no lower limit at the frequencies we might be interested in and indeed, because of the argument above, the lower the minimum frequency we can observe, the better. However, since the averaging time required to reduce the variance of the estimate of the spectrum amplitude at a given frequency is inversely proportional to the square root of the frequency resolution[8] and since in the case of flicker like spectra the frequency resolution should be significantly smaller than the minimum frequency of interest in order to reduce systematic errors,[9] it can be easily calculated that it is unlikely that we may obtain sensible information at frequencies below a few hundreds mHz if we need to remain within the practical limit of signal recording time of the order of a few hours at most. Excess noise is only present if the device is not at thermal equilibrium and, therefore, we are usually interested in the analysis of the fluctuations superimposed to the DC components resulting from the device biasing. Since a large gain is required in order to rise the signal of interest up to the level compatible with a spectrum analyzer, any low noise preamplifier that is employed to this purpose needs to be AC coupled in order to remove the DC component that would otherwise cause saturation and the cut-off frequency of the AC coupling filter clearly sets the minimum frequency that can be analyzed. In view of what we have discussed before, the cut-off frequency filter of the AC filter is usually selected in such a way as the useful frequency range may extend down to a few hundred mHz both in the case of commercial instrumentation and dedicated designs. We can safely assume, therefore, that 100 mHz is the de-facto practical minimum frequency that is going to be explored in a large number of experimental set-ups.

It is important to recognize, however, that noise measurements can provide sensible results only in the cases in which the instrumentation does not introduce a noise level that is much larger than the one to be detected. Even assuming that the external interferences can be properly shielded from the measurement environment (including thermal fluctuations and mechanical vibrations that may introduce significant contributions at low frequencies) there still remains the contribution of the noise introduced by the preamplifiers and by the device biasing systems. Such noise spectra have themselves flicker like spectra and it is therefore their noise performances that set the ultimate limit to the sensitivity of a LFNM system. Low noise voltage and current preamplifiers with very low background noise at low frequencies are available on the market and several designs capable of reaching even lower levels of background noise have been proposed in the literature.[10–12] Moreover, in many cases, cross correlation techniques among two or more amplification channels can be employed that allow to reach a sensitivity that is well below the background noise of each single amplifier.[13–17]

The situation is however quite different when we turn to the issue of the noise introduced by the biasing systems. Standard DA converter that could be used for the implementation of a programmable voltage source simply produces too much noise at low frequencies to be directly used as bias source in LFNM systems.[18] Indeed, the noise produced by a DA converter at frequencies as low as 100 mHz is usually a few orders of magnitudes above the equivalent input voltage noise of the best low frequency noise preamplifiers reported in the literature (40 dB or more). Moreover, conventional cross correlation techniques that are beneficial in reducing the equivalent background noise of the amplifying chain are useless in reducing the noise introduced by the biasing system as it almost always appears as a fully correlated component for all employed amplifiers.

As a matter of fact, resorting to large capacity batteries in combination with excess noise free resistors is the most common approach for the realization of low noise voltage and current bias system when very high sensitivity is required.[19,20] Resorting to batteries, however, while providing excellent performances in terms of low frequency noise, has severe limitations, the most important of which is the difficulty of realizing a completely automated measurement system for the characterization of the noise produced by passive or active devices in a number of bias conditions. When we recall that obtaining a single spectrum down to a few hundreds of mHz may require measurement times of the order of several hundreds (of even thousands) of seconds, the disadvantage of requiring the intervention of a skilled operator for manually changing the battery and resistance network configuration in order to obtain a different bias configuration becomes apparent, not to mention the fact that because of the discharge curve of the batteries and of their ageing effects, it is seldom possible to obtain repeatability and absence of significant drift in setting the desired bias point.

It is for these reasons that we have devoted some attention to this problem in the past and that we have proposed a few possible designs demonstrating that a very low noise programmable voltage source compatible with the strict requirements of a LFNM system can be indeed realized. In first design we essentially used a battery followed by a low noise buffer as the reference source for an $R/\beta R$ ($\beta > 2$) discrete resistor ladder network behaving as an high accuracy, infinite hold low noise sampling circuit.[21,22] When a reference voltage produced by a high accuracy standard DA converter is fed to the system during the setting phase, its voltage is reproduced with high accuracy at the output of the discrete components ladder network. This approach allows to obtain very low noise and high initial accuracy, but it does not prevent output voltage drift as due to the reference battery self discharge. Most importantly, the large component count due to the need of implementing relays and discrete resistor based ladder converter for ensuring very low noise resulted in a quite expensive and somewhat cumbersome system that does not make such an approach feasible for easy reproduction and routine applications to LFNM. A second approach consisted in filtering out the noise produced by a conventional DA by means of a low pass filter followed by a very low noise JFET (Junction Field Effect Transistor) acting as a buffer.[23] While this approach may appear obvious, reaching an attenuation of 40 dB at 100 mHz means that a single RC stage with capacitance values limited to a few tens of μF (conventional electrolytic capacitors are likely to introduce excess noise at low frequencies because of micro-discharge phenomena) is anything but simple. Moreover, a JFET buffer stage must be used that introduced an offset of the order of few hundred mV or more and a quite sophisticated digital control loop had to be used in order to ensure at the same time very low noise and reasonable accuracy. Once again, while effective in terms of noise level, such a system still results quite complex and with a size that makes its integration impossible in compact measurements systems such as the one described in Ref. 24 where it would be most needed for a systematic analysis of the noise produced by devices available at wafer level. Notably, most of the problems that had to be overcome for the realization of such a source can be traced back to the limited value of the non-electrolytic capacitors that were available at the time in which the design was being developed. Nowadays, supercapacitors and/or ultracapacitors have become quite standard components. Contrary to the case of standard electrolytic capacitors, experiments on supercapacitors have recently demonstrated that they can be indeed employed in very low noise systems.[25] With the availability of supercapacitors with rated voltages as high as 12 V and capacitances in the tens of mF range, several of the problems that had to be faced in Ref. 23 are greatly reduced and a simpler design of a high accuracy, very low noise voltage source characterized by a very low component count becomes possible as it will be discussed in the reminder of this paper.

II. DESIGN OF THE LOW NOISE, HIGH ACCURACY PROGRAMMABLE VOLTAGE SOURCE

In order to better understand the problems that have to be faced in designing a programmable low noise voltage source following the approach of filtering out the noise produced by a standard DA converter, let us take into consideration the simplified diagram in Fig. 1, where e_{nDA} represents the noise produced by the solid state DA converter superimposed to the DC voltage V_{DA}, R and C form a first order low pass filter, e_R represents the thermal noise of the resistance R, e_{nb}, and i_{nb} represent the equivalent input voltage and current noise sources of the buffer with an offset voltage V_{OFF}. In our discussion and experiments we will assume that we are using the 12 bit AD667 high accuracy DA converter characterized by an accuracy better than $\frac{1}{4}$ LSB.[26] The AD667 is configured in such

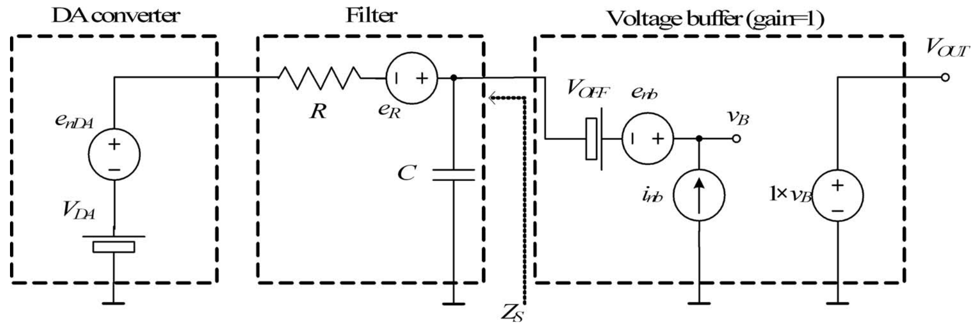

FIG. 1. Schematic of a programmable low noise voltage source following the approach of filtering out the noise produced by a standard DA converter. e_{nDA} represents the noise produced by the solid state DA converter superimposed to the DC voltage V_{DA}, R and C form a first order low pass filter, e_R represents the thermal noise of the resistance R, e_{nb} and i_{nb} represent the equivalent input voltage and current noise sources of the buffer with an offset voltage V_{OFF}.

a way as to provide a unipolar output from 0 to 10 V using its own internal 10 V reference with a resolution of about 2.5 mV (1 LSB $= 10/2^{12} = 2.44$ mV) and an accuracy, after offset and gain correction, better than 1 mV (1 LSB/4 = 0.61 mV). The voltage noise at the output of the DA converter when $V_{DA} = 8$ V is shown in Fig. 2 together with the equivalent input noise of the ultra low noise preamplifier used for performing the noise measurements.[12] With reference to the circuit in Fig. 1, it is clear that the minimum noise we can obtain at the output, provided that we can filter out the noise produced by the DA, is the noise introduced by the buffer stage. Therefore, before proceeding any further, we must discuss the options that are available for the design of the low noise output buffer. It is reasonable to assume, as a target for the performances of the programmable voltage source to be designed, a low frequency noise ($f < 1$ Hz) comparable to the equivalent input voltage noise of the voltage amplifier used for performing noise measurements in the final LFNM system. As the amplifier used for performing the measurements in Fig. 1 is among the best general purpose low noise amplifiers ever reported in the literature, we can regard its equivalent input noise at low frequencies ($f < 1$ Hz) as the target performance to be achieved. The noise introduced by the buffer is the combination of the effects of its equivalent input voltage noise and equivalent input current noise. Assuming no correlation between e_{nb} and i_{nb}, the output noise due to the buffer can be calculated as

$$S_B = S_{enb} + S_{inb} \left|Z_S\right|^2 , \tag{1}$$

where S_B, S_{enb}, and S_{inb} are the Power Spectral Density (PSD) of the voltage fluctuations at the output of the buffer and of the noise sources e_{nb} and i_{nb}, respectively, and Z_S is the impedance seen by the input of the buffer toward the DA source. Regardless of the actual values of R and C in the filter, it is clear that, if a significant noise reduction at the minimum frequency of interest f_{min} (f_{min} = 100 mHz) has to be obtained, the filter corner frequency $f_c = 1/(2\pi RC)$ has to be much lower than f_{min}, and therefore, at all frequencies of interest, we have $|Z_S| \approx 1/(2\pi fC)$ and

$$S_B = S_{enb} + \frac{S_{inb}}{4\pi^2 f^2 C^2}. \tag{2}$$

It must be noted that conventional electrolytic capacitors are prone to instabilities that result in large spectra components at very low frequencies, while tantalum capacitors, that are easily available with capacitances up to 10 mF (10 V maximum) are characterized by high values of leakage currents (in excess of 100 μA) that limit the accuracy because of the voltage drop across the resistance R of the low pass filter. Prior to the introduction of supercapacitors, therefore, only polyester or polypropylene capacitors could be used for C, thus limiting its maximum sensible value to a few tens of μF at most. Even with $C = 100$ μF, in order to obtain a contribution due to S_{inb} below the target value at 100 mHz (about 4×10^{-16} V^2/Hz or 20 nV/$\sqrt{}$Hz from Fig. 2), S_{inb} should be below 10^{-24} A^2/Hz (or 1 pA/$\sqrt{}$Hz) which clearly rules out any bipolar input low noise operational amplifier as a possible candidate for the buffer. This can be readily verified with reference to two of the most popular low noise operational amplifiers such as the OP27 (by Analog Devices) and LT1128 (by Linear Technology): the OP27 at $f = 100$ mHz is characterized by $S_{en} = 2.4 \times 10^{-16}$ V^2/Hz and $S_{in} = 224 \times 10^{-24}$ A^2/Hz while the LTC1128, at the same frequency, is characterized by $S_{en} = 1.40 \times 10^{-16}$ V^2/Hz and $S_{in} = 1000 \times 10^{-24}$ A^2/Hz (maximum values). On the other hand, if we turn to very low input current noise operational amplifiers (that is JFET or MOSFET input operational amplifiers), their equivalent input voltage noise is much larger than the target value at f_{min}. The only way to obtain a very low equivalent input current noise to-

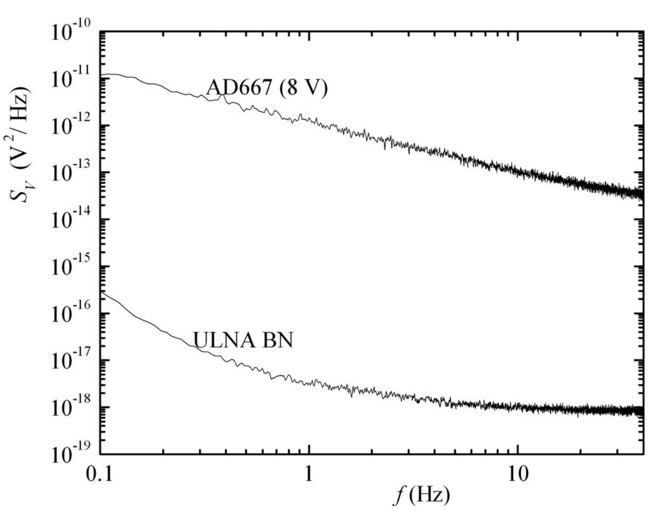

FIG. 2. Voltage noise at the output of the AD667 DA converter when V_{DA} = 8 V together with the equivalent input noise of the ultra low noise preamplifier used for performing the noise measurements.

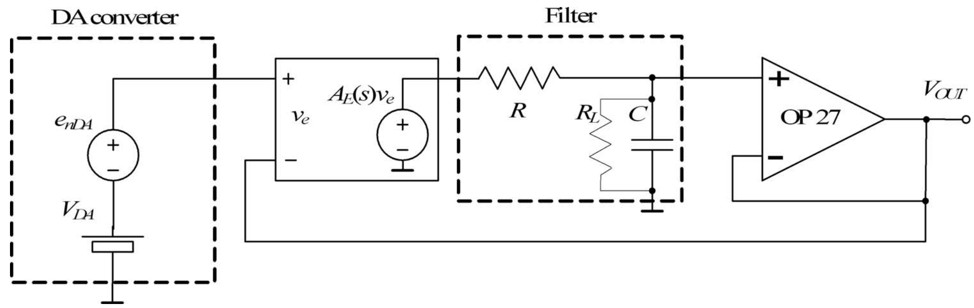

FIG. 3. Feedback configuration that could be employed to reduce errors in the value of the supplied voltage due to the leakage current in the supercapacitor (modeled by $R_L = 1$ MΩ in the schematic).

gether with a very low equivalent voltage noise is to resort to discrete JFET devices, which is in fact what we did in the past for obtaining a programmable low noise voltage reference according to the basic approach in Fig. 1.[23] With the availability of supercapacitors, that have been proven to be compatible with the design of low noise measurement instrumentation,[25] the situation dramatically changes since with $C = 15$ mF, the maximum current noise that can be tolerated at 100 mHz is of the order of 200×10^{-24} A^2/Hz that it is close to the typical values observed in low noise operational amplifiers such as the OP27, thus paving the way for the much simpler and compact design that we will be discussing in the following. Unfortunately, also supercapacitors suffer from a relatively large leakage current that can reach the order of 10 μA for a 12 V 15 mF device. Although much lower than in the case of tantalum capacitors, the leakage current adversely affects the accuracy that can be obtained. This can be clearly understood with a quite simple calculation. Since, as it is apparent from Fig. 2, the minimum attenuation of the DA noise that is required at 100 mHz is of the order of 40 dB, this means that the corner frequency of the filter in Fig. 1 should be less than 1 mHz, thus resulting, with $C = 15$ mF, in a resistance R above 60 kΩ. Therefore, a leakage current of 10 μA would produce a voltage drop across R of the order of 600 mV, thus resulting in large and, in most cases, unacceptable errors in the value of the supplied voltage. In order to correct this error, we could resort to a feedback configuration such as the one represented in Fig. 3 where an error amplifier with a proper gain $A_E(s)$ could allow to compensate for the error introduced by the leakage resistance R_L (R_L is used to provide for a simple model of the leakage current in the supercapacitor). However, such a system is extremely difficult to design if we are willing to obtain, at the same time, low noise at very low frequencies and high accuracy. Indeed, in order to obtain a DC accuracy of the order of 0.1% (10 mV for a supplied voltage of 10 V), the DC gain of the error amplifier $A_E(0)$ should be larger than 1000 at DC and should assume a value of 1 or less at 100 mHz since otherwise the noise produced by the DA would also be amplified in the frequency range of interest, thus essentially neutralizing the filtering effect of the RC stage. This would mean designing a single pole error amplifier with a DC gain of 1000 and a pole frequency of 100 μHz or below, that is anything but an easy task, not to mention the problem of the additional noise introduced by the error amplifier itself.

In order to overcome all these limitations, we devised the approach illustrated in Fig. 4. In this approach, the filter is fed with the output of the DA to which the error amplifier adds a voltage proportional to the error between the input (the DA output voltage) and the output (the actual output of the system). With this approach, as it will be presently demonstrated, a much smaller gain for the error amplifier is required, with respect to the approach in Fig. 3, for obtaining a given accuracy, thus significantly reducing the problem of the increase of the noise at the input of the filter stage due to the presence of the error amplifier itself. The actual circuit implementing the approach in Fig. 4 is reported in Fig. 5. The control strategy in Fig. 4 is obtained in a quite straightforward way by resorting to the high precision low noise instrumentation amplifier INA101. The reference input of the INA101 is indeed connected to the output of the DA so that its output voltage is

$$V_{OI} = V_{DA} + A_I(V_{DA} - V_{OUT}), \tag{3}$$

where A_I is the differential gain set by means of the gain setting resistance R_G. For gains up to 100, the frequency re-

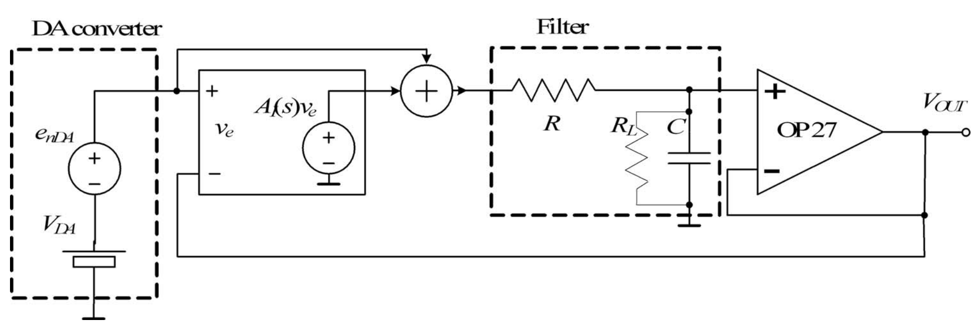

FIG. 4. Proposed approach that, with respect to the approach in Fig. 3, allows to significantly reduce the problem of the increase of the noise at the input of the filter stage due to the presence of the error amplifier.

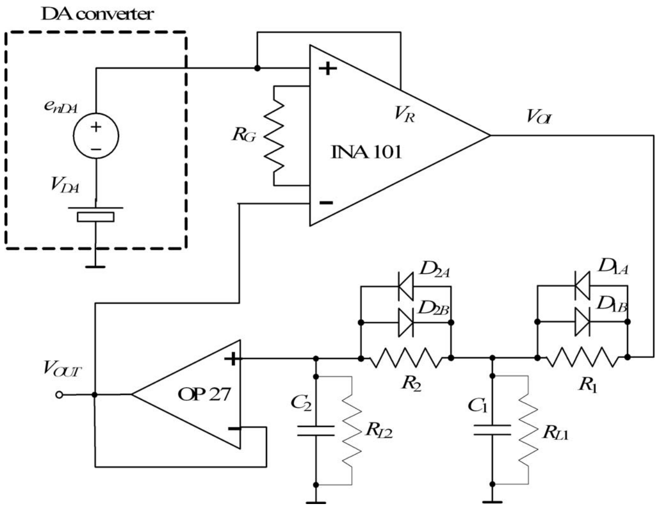

FIG. 5. Actual circuit implementing the approach in Fig. 4.

sponse of the INA101 can be considered to be constant at least up to 10 kHz. In order to compensate for the increased noise at the input of the filter stage because of the presence of the error amplifier, its structure is modified into a two stage RC ladder network so that we can obtain a much larger attenuation at all frequencies of interest for the same voltage drop due to the capacitor leakage currents. The two stages ladder network introduces two poles in the control loop, but due to the fact that the next pole (due to the INA101) is at a very high frequency, compared to the pole frequencies of the filter, the system stability is ensured for a large set of operating conditions. The diodes across R_1 and R_2 allow to shorten the time required to reach a near steady state condition, but can be neglected as far as the noise analysis of the system is concerned, since their equivalent resistance at or near equilibrium is of the order of a few MΩ (all diodes are 1N4148). The leakage current (at steady state) in the supercapacitors is modeled by means of the resistances R_L. Assuming that, at all operating frequencies we may be interested in, both the instrumentation amplifier gain and the buffer gain are constant (independent of the frequency), we can quite easily estimate the transfer function from the output of the DA to the output of the system. Assuming $R_{L1} = R_{L2} = R_L$, $R_1 = R_2 = R$, $C_1 = C_2 = C$ and with the following positions

$$\frac{R}{R_L} = \alpha \quad \omega_{RC} = \frac{1}{RC}, \tag{4}$$

we have

$$v_{OUT}(s) = v_{OI}(s)\frac{1}{(1+3\alpha+\alpha^2)+(3+2\alpha)\dfrac{s}{\omega_{RC}}+\dfrac{s^2}{\omega_{RC}^2}} \tag{5}$$

and, using Eq. (3),

$$v_{OUT}(s) = v_{DA}(s)\frac{1+A_I}{(1+A_I+3\alpha+\alpha^2)+(3+2\alpha)\dfrac{s}{\omega_{RC}}+\dfrac{s^2}{\omega_{RC}^2}} \tag{6}$$

that can also be written in the form

$$\frac{v_{OUT}(s)}{v_{DA}(s)} = H(s) = \frac{H_0}{1+\dfrac{s}{Q\omega_0}+\dfrac{s^2}{\omega_0^2}} \tag{7}$$

with

$$H_0 = \frac{1+A_I}{1+A_I+3\alpha+\alpha^2}; \quad \omega_0 = \omega_{RC}\sqrt{1+A_I+3\alpha+\alpha^2};$$
$$Q = \frac{\sqrt{1+A_I+3\alpha+\alpha^2}}{3+2\alpha}. \tag{8}$$

The relative error ε_R at DC steady state can be calculated as follows:

$$\varepsilon_R = \left|\frac{v_{OUT}(0)-v_{DA}(0)}{v_{DA}(0)}\right| = |H_0 - 1| = \frac{3\alpha+\alpha^2}{1+A_I+3\alpha+\alpha^2}. \tag{9}$$

Since we expect $\alpha \ll 1$ (that is $R_L \gg R$) the following approximations hold:

$$\varepsilon_R \approx \frac{3\alpha}{1+A_I}; \quad \omega_0 = \omega_{RC}\sqrt{1+A_I}; \quad Q = \frac{\sqrt{1+A_I}}{3}. \tag{10}$$

Since at the minimum frequency of interest f_{min} we need to have a large attenuation of the noise produced by the DA converter, it is safe to assume

$$\begin{aligned}\omega_0 \ll 2\pi f_{\min} \Rightarrow |H(j2\pi f_{\min})|^2 &\approx \frac{H_0^2\omega_0^4}{(2\pi f_{\min})^4}\\ &\approx \left(\frac{1}{2\pi RCf_{\min}}\right)^4 (1+A_I)^2 \approx \left(\frac{1}{2\pi RCf_{\min}}\right)^4 \frac{9\alpha^2}{\varepsilon_R^2}\\ &= \left(\frac{1}{2\pi R_L Cf_{\min}}\right)^4 \frac{9}{\alpha^2\varepsilon_R^2} = \left(\frac{1}{2\pi \tau_L f_{\min}}\right)^4 \frac{9}{\alpha^2\varepsilon_R^2}.\end{aligned} \tag{11}$$

Because of the frequency dependence of H, we can also assume that at all frequencies larger than f_{min}, we will obtain a larger noise attenuation (a smaller value of $|H|^2$). If we define the Noise Attenuation Factor (N_{AF}) as the inverse of $|H|^2$ at the minimum frequency of interest f_{min}, we obtain, from Eqs. (10) and (11),

$$\frac{\varepsilon_R^2}{N_{AF}} \approx \left(\frac{1}{2\pi\tau_L f_{\min}}\right)^4 \frac{9}{\alpha^2}. \tag{12}$$

Equation (12) just derived is extremely useful for appreciating the potentialities of the design we propose and to help in the selection of the circuit parameters. If we regard the characteristics of the supercapacitors as the starting point of our analysis, Eq. (12) simply states that, once a supercapacitor type is selected, the ratio between ${\varepsilon_R}^2$ and N_{AF} is set by the value chosen for R within the limits of the approximation we have made. Another interesting way to look at Eq. (12) is that, once the supercapacitor type and the value of R are chosen, changing the gain A_I of the instrumentation amplifier affects both N_{AF} and ε_R in such a way as to maintain the ratio ${\varepsilon_R}^2/N_{AF}$ constant. This means, in turn, that by increasing the gain A_I we increase the DC accuracy at the cost of the noise reduction factor and vice versa. It is also clear that increasing α

(selecting a larger value of R with respect to R_L) allows to obtain either larger N_{AF} for the same accuracy, or better accuracy for the same N_{AF}. Unfortunately, α cannot be increased indefinitely for a number of reasons, besides the fact that the results obtained so far only hold for $\alpha \ll 1$. First of all, we must recognize that since there is a limit to the output voltage swing of the INA101 operational amplifier, values of R close to R_L, that would result in a large voltage drop across the filter, would limit the output dynamic range of the source. Moreover, increasing α for obtaining the same accuracy, but a larger N_{AF}, requires increasing A_I (Eq. (10)) that results, in turn, in a larger value of Q and lower value of ω_0, thus resulting in a longer settling time (for obtaining the same ε_R, A_I must be increased as we increase α). Indeed, from the point of view of the settling time and the dynamic range of the system, we would require a value of α as small as possible.

As we have noted before, since it does not make any sense to reduce the noise produced by the DA below the noise level that is introduced by the buffer stage, we can set $N_{AF} = 10^4$ (40 dB) as the required noise reduction factor, after which Eq. (12) allows to estimate, starting from the supercapacitors characteristics, the required value of R for obtaining any given desired accuracy.

In the prototype we have built and tested, the supercapacitors are 15 mF, 12 V rated (Cellergy CLG12P015L28) with $R_L \approx 1$ MΩ (10 μA leakage current). With $f_{min} = 100$ mHz and $N_{AF} = 10^4$, if we require a relative accuracy of 0.1% (an error of 5 mV for a DC output of 5 V) from Eq. (12) we obtain $\alpha = 3.4 \times 10^{-3}$ and, hence, $R = 3.4$ kΩ. In the actual prototype we have built and tested, R was 10 kΩ, meaning that it is possible to obtain a larger N_{AF} for the same error ($N_{AF} \approx 10^5$ or 50 dB with $e_R = 0.1\%$) or a better accuracy for the same N_{AF} ($e_R < 0.04\%$ with $N_{AF} = 10^4$) depending on the instrumentation amplifier gain setting. So far we have been working in the assumption that the only relevant sources of noise are the noise introduced by the DA and the noise introduced by the output buffer. While it would be quite easy to demonstrate that this is indeed the case by means of proper calculations, we prefer to discuss this specific issue by resorting to the result of proper SPICE simulations as we believe that they can provide, within the framework of the general design guidelines that we have developed, a better picture of the general behavior of the system. SPICE simulations are needed in any case in order to investigate on the transient behavior of the system that is strongly affected by the dynamic limitations of the INA101 and by the presence of the diodes in parallel to the resistances of the filter.

III. CIRCUIT SIMULATION AND EXPERIMENTAL RESULTS

SPICE simulations have been used to investigate on the general behavior of the system in Fig. 5 as far as the frequency response, the noise, and the transient behavior are concerned. In all simulations we set $C = 15$ mF, $R = 10$ kΩ, $R_L = 1$ MΩ. The supply voltage was set to ±12 V. Two different values for the setting gain resistance were used, namely, $R_{G1} = 10$ kΩ and $R_{G2} = 500$ Ω corresponding to a differential gain of $A_{I1} = 5$ and $A_{I2} = 81$, respectively. The SPICE models for the INA101 and the OP27 were obtained from the manufacturers. The OP27 noise model as provided by the manufacturer is devised in such a way as to be compatible with the worst case scenario. Since in actual measurements we usually obtain an equivalent input voltage noise very close to the typical case reported in the data-sheet, we adjusted the noise parameter in the SPICE model in such a way as to closely reproduce such behavior.

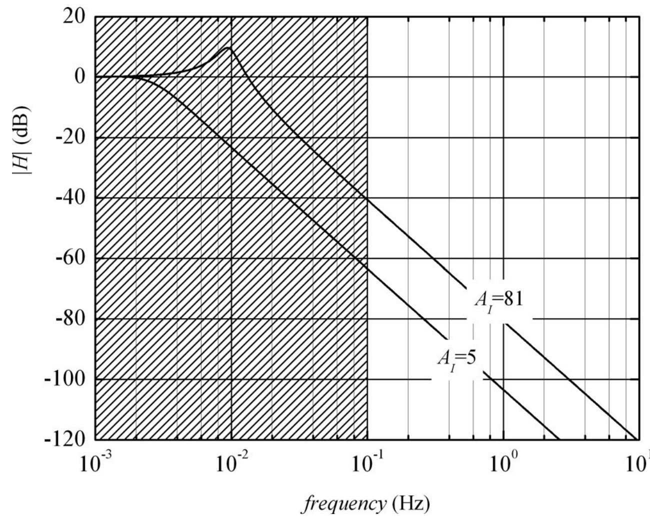

FIG. 6. Magnitude of the frequency response H as obtained from SPICE simulations for $A_I = 5$ and $A_I = 81$.

The magnitude of the frequency response H as obtained from SPICE simulations is reported in Fig. 6 for $A_I = 5$ and $A_I = 81$. As expected, increasing A_I decreases the attenuation of the DA noise that is however better than 40 dB even for $A_I = 81$ (corresponding to a DC relative error of about 0.04%). Because of the two poles introduced by the filter, the value of Q in Eq. (8) increases as A_I increases. This is clearly observable in Fig. 6 where the frequency response for $A_I = 81$ peaks at $f \approx 10$ mHz (outside the frequency range of interest) and this is expected to result in a significant overshoot in the time domain response to a change in the DA setting. If the system was not affected by nonlinearity, with the frequency response corresponding to $A_I = 81$ we should expect an overshoot of 60%. The presence of the diodes in parallel to the resistances of the filter, however, acts in such a way as to considerably limit the overshoot even for quite low voltage steps. This is clearly shown in Fig. 7 where the simulated response of the system is reported as a consequence of a 100 mV step change in the DA setting (from 4.9 V to 5 V) in the case in which the diodes are not present and in the case of the actual circuit in Fig. 5. The voltage step is small enough that the INA101 is in linearity at all times even with $A_I = 81$. As it is apparent, the diodes effectively suppress the oscillations that would be otherwise present. Fig. 7 also allows to appreciate the difference in the DC accuracy that is obtained with the two different gain settings. Finally, Fig. 7 also demonstrates that, at least in the case of a small step at the input, the system reaches DC steady state in less than 10 min, which is a time largely acceptable for the applications the system is intended for. When a larger step is forced at the input, the INA101 is likely to saturate,

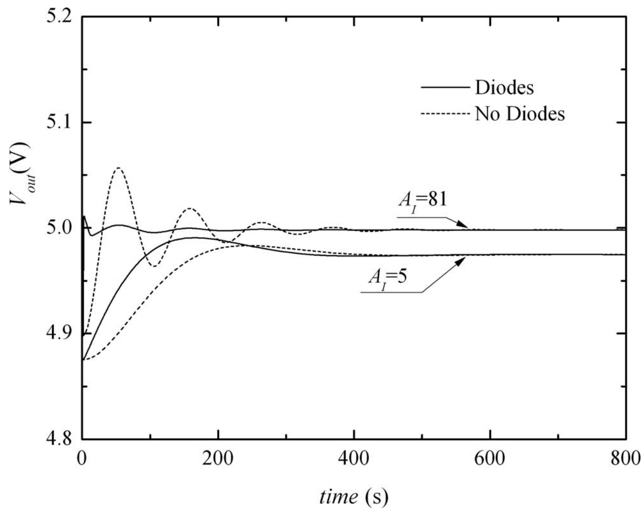

FIG. 7. Simulated response of the system as a consequence of a 100 mV step change in the DA setting (from 4.9 V to 5 V) in the case in which the diodes are not present and in the case of the actual circuit in Fig. 5.

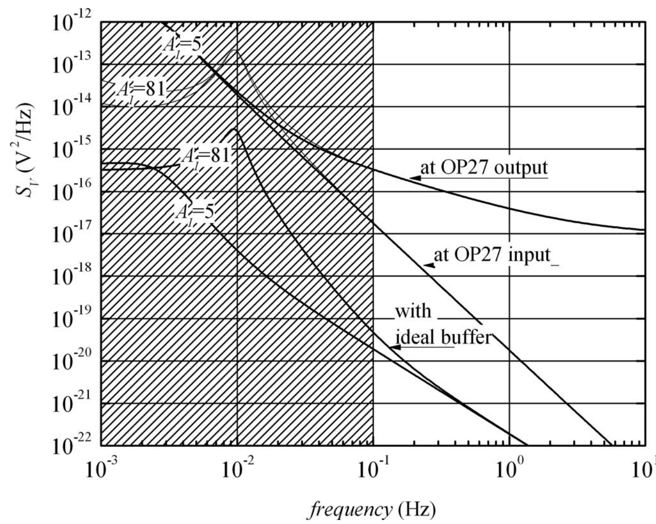

FIG. 9. Simulated output noise PSD assuming a noiseless DA at the input.

especially in the case of large values for A_I. As a result of the saturation of the output of the instrumentation amplifier, the charge transient, with no diodes across the resistances of the filter, would be much longer, as it is clearly shown in Fig. 8 where the simulated response to a 5 V input step with and without diodes across the resistances is reported in the case $A_I = 81$. Because of the presence of the diodes, no significant overshoot is observed and the output is at steady state in a matter of a few minutes. As far as the output noise is concerned, Fig. 9 shows the output noise assuming a noiseless DA at the input. We extracted the noise at the output of the actual system, the noise at the input of the OP27 and the noise that would be obtained if it were possible to employ a noiseless buffer instead of the OP27. The plots in Fig. 9 indicate that:

- the noise introduced by the system is in the frequency range of interest ($f > 100$ mHz) essentially independent of A_I;
- the noise at the output of the system is dominated by the voltage noise source of the OP27. This is clearly deduced by the fact that the noise at the input of the OP27 is significantly lower. Notably, in the investigated frequency range, the noise at the input of the buffer is dominated by the equivalent current noise at the input of the OP27 in combination with the output impedance of the filter that essentially reduces to that offered by the capacitor C_2 in Fig. 5. Indeed, if a noiseless buffer could be used, the residual noise (due to the filter resistances and to the contribution from the INA101) would be much lower.

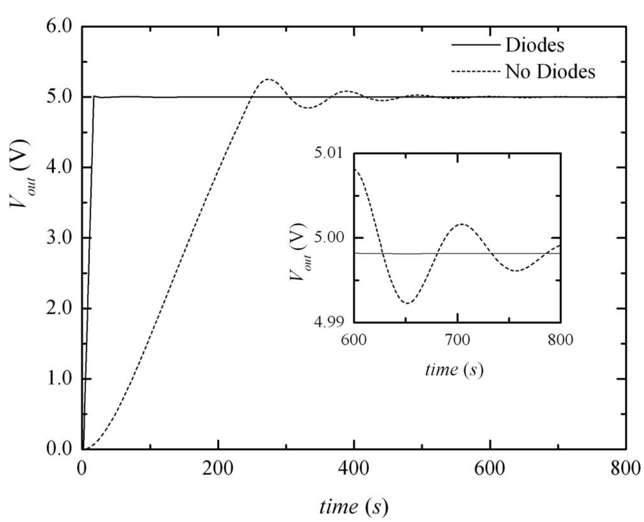

FIG. 8. Simulated response of the circuit to a 5 V input step with and without diodes across the resistances in the case $A_I = 81$.

The results of the noise simulation confirm what was indeed expected: if the noise coming from the DA can be sufficiently attenuated in the frequency range of interest the output noise of the system reduces to the equivalent input voltage noise of the output buffer.

Actual measurements performed on a prototype of the system in Fig. 5 with $A_I = 81$ confirm such expectations. In the prototype the DA converter for setting the voltage is an AD667 set in unipolar mode from 0 to 10 V using its own internal 10 V reference. The system is supplied by two 12 V lead acid batteries. With this supply, because of the limitations in the output stage of the INA101, the maximum output voltage that can be obtained with the system in linearity at steady state is 8 V. The noise at the output of the system when the DA is set to 8 V is reported in Fig. 10 together with the background noise of the amplifier (ULNA BN) employed for the noise measurement. Note that the BN of the amplifier is not subtracted from the measured noise at the output. As it can be clearly appreciated, the output noise essentially coincides with the noise introduced by the equivalent input voltage noise of the OP27 operational amplifier.

We also measured the actual time response of the system to an input voltage step of 5 V. The resulting output voltage vs time is reported in Fig. 11. As it can be noted, the initial transient closely reproduces the simulated behavior. However, as it can be noted in the inset, a slow transient is still developing even after half an hour from the application of the input step.

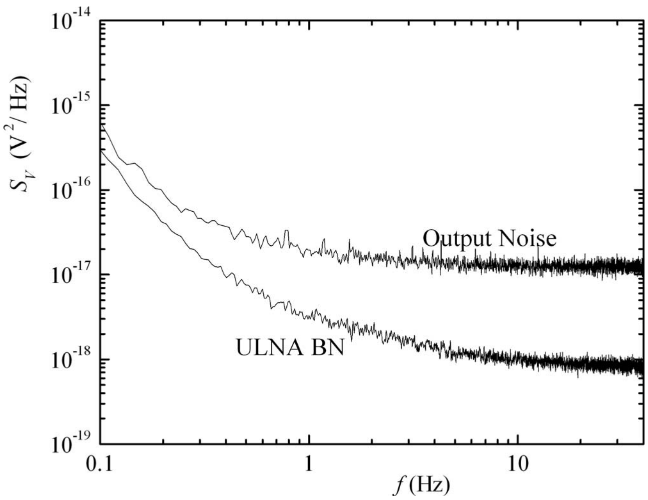

FIG. 10. Noise PSD at the output of the system when the DA is set to 8 V together with the background noise of the amplifier (ULNA BN) employed for the noise measurement and the noise at the output of the DA.

This residual transient can be traced back to the quite complex behavior of a supercapacitor during charging transients as it was observed in Ref. 25. Indeed, modeling a supercapacitor as the parallel combination of an ideal capacitor and an ideal leakage resistance is quite a crude approximation of its actual dynamic. It can be easily observed that a supercapacitor continues to draw a significant amount of current (that can be several times larger than the steady state leakage current) for a significant time after the voltage at its ends has essentially reached a constant value. This excess current is responsible for the long transient observed in Fig. 11 and as it is intrinsically due to the nature of the charge storing mechanism of supercapacitors it cannot be avoided. However, we can also observe that the output voltage is within 0.05% of the set voltage in a matter of about 10 min and that the residual transient leading to the final accuracy of less than 0.04% (that is reached within one hour) can be tolerated as it is comparable or even less pronounced than the drift that would be experienced in any case in providing the bias by means of batteries because of their discharge curves. Moreover, in the case of smaller voltage steps, as it is expected to be the case in a system used for exploring the noise behavior of a device in a range of bias conditions, the residual transient amplitude also reduces.

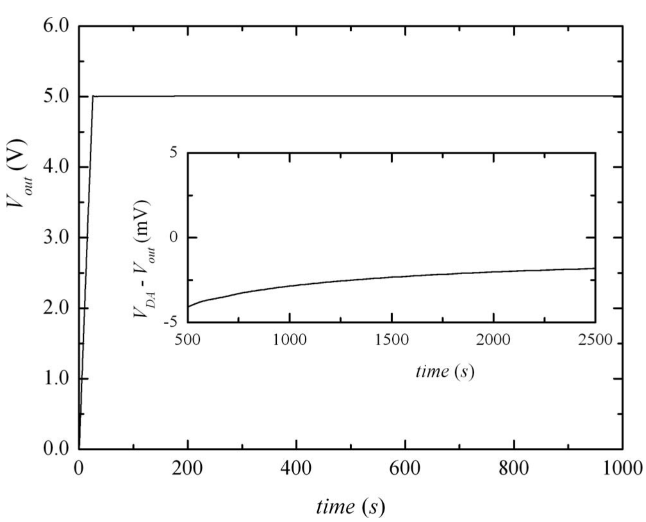

FIG. 11. Actual time response of the system to an input voltage step of 5 V.

IV. CONCLUSIONS

In this paper a new topology of a programmable, very low noise, high accuracy voltage source has been presented that takes advantage of the large capacitance of supercapacitors for realizing a low pass filter with a very low frequency corner for filtering out the noise produced by a solid state high accuracy DA converter. A solution employing an error amplifier has been designed, which requires a quite small gain for obtaining a given accuracy, thus significantly reducing the problem of the increase of the noise at the input of the filter stage due to the presence of the error amplifier itself. A prototype of the proposed voltage source has been realized and tested, demonstrating that the output noise essentially reduces to the noise introduced by the equivalent input voltage noise of the OP27 operational amplifier, used as the output buffer stage. Since a very low frequency corner must be employed in order to obtain the desired attenuation of the noise down to a few hundred mHz, a very long transient upon each voltage change may occur. Nevertheless, it has been observed that the output voltage is within 0.05% of the set voltage in a matter of about 10 min and that the residual transient leading to the final accuracy of less than 0.04% (that is reached within one hour) can be tolerated as it is comparable or even less pronounced than the drift that would be experienced in any case in providing the bias by means of batteries because of their discharge curves. Moreover, in the case of smaller voltage steps, as it is expected to be the case in a system used for exploring the noise behavior of a device in a range of bias conditions, the residual transient amplitude also reduces.

ACKNOWLEDGMENTS

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[26]See http://www.analog.com/en/digital-to-analog-converters/da-converters/ad667/products/product.html for AD667 datasheet.