Ultra-low noise, single JFET voltage pre-amplifier for low frequency noise measurements

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***Abstract*—We propose an approach to the design of voltage amplifiers for low frequency noise measurements applications that is based on a single transistor (JFET) amplifier as the first stage. With respect to differential stages, a single transistor input stage has the advantage, among others, of allowing for a lower background noise. The problem of the dispersion of the gain, in the absence of a passive feedback loop, is addressed by introducing auxiliary operational amplifier based stages in parallel to the very low noise amplifier chain and using cross correlation among the outputs at a conveniently high frequency for obtaining the accurate estimate of the gain of the main amplifier in a short time and while measurement is in progress. With the approach we propose no adjustment of any kind is required on the circuit parameters of the main amplifier prior to any new measurement.**

Keywords—Low noise voltage amplifier, cross correlation, noise measurements

I. Introduction

Performing meaningful noise measurements on electron devices and systems requires that the Background Noise (*BN*) of the measurement chain be much smaller than the noise to be detected [1]. In the case of voltage noise measurements, the *BN* of the system can be reduced to the Equivalent Input Voltage Noise (*EIVN*) of the voltage amplifier directly connected to the Device Under Test (*DUT*). A possible approach to the design of very low noise voltage amplifiers is to add a differential stage based on discrete, very low noise, devices in front of a solid state high gain stage. In this way, the entire system can be regarded as a very low noise operational amplifier whose gain can be set and stabilized by means of passive feedback [2-6]. This approach has the added advantage that, because of the input differential stage, coupling to the *DUT* down to *DC* is possible, which can be relevant in the field of Low Frequency Noise Measurements (*LFNM*) where the frequencies of interest may extend well below 1 Hz [1]. While the lowest level of *EIVN* are obtained by employing BJT based input stages [1,3], *JFET* input stages are preferred in all those cases in which noise is to be measured on nodes where a significant DC component is also present. This is often the case in the field of noise measurements on electron devices, since measurements need to be performed on biased *DUT*s. Using *BJT*s, *AC* coupling down to the tens of mHz range would be made quite difficult because of the large bias currents and equivalent input noise current of these devices.

A feedback amplifier design approach allows to compensate for the spread in the discrete device characteristics so that a stable and known gain can be obtained regardless of the individual device being used. This simplification, however, comes at a price. Compared to a Common Source (*CS*) single transistor stage, the *EIVN* is at least doubled because both transistors in the input stage equally contribute to the *BN* [5]. Moreover, obtaining stability when designing with an operational amplifier based on a discrete component input stage is never easy and frequency compensation often results in reduced bandwidth and BN increase at higher frequencies [5, 6]. It is for these reasons that we devoted our attention to the possibility of designing a low noise voltage preamplifier based on a single transistor input stage that could be effectively employed in the field of *LFNM*s.

II. Proposed Approach

The simplest possible implementation of a low noise voltage amplifier based on a single transistor stage for application in the field of *LFNM*s is reported in Fig. 1. The active device is a low noise JFET that insures that we can obtain *AC* input coupling down to very low frequencies. In our prototype we employ a IF3601 large area JFET by InterFet, capable of providing tranconductance gains (g_m) in the order of a few tens of mA/V with a drain current in the order of a few mA. This device is characterized by an equivalent input noise that can be as low as 0.3 nV/√Hz at 100 Hz. As it is characteristic of *JFET* devices, *DC* and *AC* parameters are quite spread. The datasheet for IF3601 lists a pinch off voltage in the interval form -2 up to -0.35 V, while only a minimum value for the saturation current at V_{GS}=0 is given. In the circuit configuration in Fig. 1 we have (assuming negligible gate current and the *JFET* in saturation):

$$I_D = \frac{-V_{GS} + V_{SS}}{R_S} \qquad (1)$$

While the actual value of I_D depends on the characteristic of the particular *JFET* used in the circuit, the fact that $-V_{GS}$ is certainly smaller than 2 V means that for V_{SS} equal to 6 V or more, the bias current (I_D) can be set with a reasonable error (typically less than 10%) by the ratio V_{SS}/R_S. In other words, with the simple bias configuration in Fig. 1, the effect of parameter dispersion is modest, as far as the bias point is concerned. Note, however, that this does not insure in any way

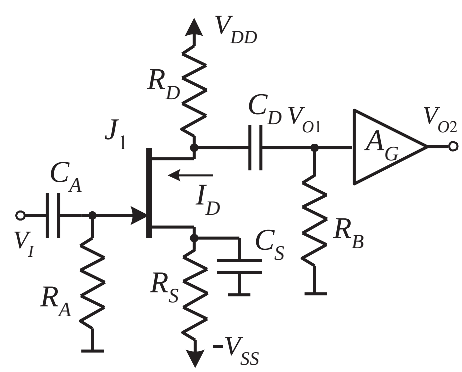

Fig. 1. Schematic of a low noise voltage amplifier using a Common Source (CS) JFET amplifier as a first gain stage.

that we obtain a predictable and known gain for the amplifier, independent of the particular device being used. Dealing with the problem of a largely unpredictable overall gain without resorting to time consuming calibration procedures that require the modification of circuit parameters, is the main subject of the paper. However, if a circuit like the one reported in Fig. 1 is to be used for *LFNM*, we must insure a passband extending well below 1 Hz, a result that is not easily obtained. The small signal equivalent circuit for gain and noise calculation in the circuit in Fig. 1 is reported in Fig. 2. In the circuit in Fig. 2 e_{nRX} is the source representing the thermal noise of the resistance R_X (we assume that all resistors in the circuit are excess noise free resistors); the parameter g_m is the transconductance of the *JFET* corresponding to the bias point in Fig. 1; e_{nj} is the noise source representing the noise generated by the JFET and e_{nAG} is the equivalent input noise voltage of the solid stage amplifier cascaded to the first stage (we assume, for the sake of simplicity, negligible equivalent input current source). For the time being, let us assume that all noise sources are turned off (replaced by short circuits) so that we can proceed with the evaluation of the transfer function. Applying standard circuit solving techniques we obtain:

$$A_V = -A_G \frac{s\tau_A}{1+s\tau_A} \cdot \frac{s\tau_D}{1+s\tau_D} \cdot \frac{1+s\tau_S}{1+s\tau'_S} \cdot \frac{g_m(R_D||R_B)}{1+g_m R_S} \quad (2)$$

where

$$\tau_A = R_A C_A \,; \tau_D = (R_D + R_B)C_D \,; \tau_S = R_S C_S$$
$$\tau'_S = \frac{\tau_S}{(1+g_m R_S)} = R'_S C_S \,;\ R'_S = \frac{R_S}{(1+g_m R_S)} \quad (3)$$

From (2) it is apparent that, since ($\tau'_s<\tau_s$), a constant frequency response is obtained for frequencies larger than the one corresponding to the pole with the largest magnitude. With resistances R_A and R_B in the order of a few MΩ and C_A and C_D in the order of a few tens of μF, the pole frequencies corresponding to τ_A and τ_D can be easily set to be below 10 mHz, but in the case of the pole frequency corresponding to τ'_S we have the problem that R_S must be in the order of 1 kΩ to obtain bias currents in the order of a few mA according to (1). Moreover, since g_m is in the order of a few tens of mA/V, R'_S in (3) may be as low as a few tens of Ω. Therefore, the only way to obtain a value of τ'_S in the order of a few seconds or tens of seconds (to obtain a pole frequency well below 1 Hz) is to resort to a capacitor value in the order of 1 F or more for C_S. While up to a few years ago this would have been largely impractical, nowadays supercapacitors in very compact size are easily available with capacitances ranging from a few mF to a few F. Note that, because of the circuit configuration in Fig. 1, the DC voltage drop across the capacitor C_S is $-V_{GS}$ that, in the case of the *JFET* we employ in our design, is typically below 1 V. Unlike other high specific capacitance devices (electrolytic capacitors) supercapacitor have been shown to be compatible with instrumentation intended for *LFNM* applications [7-9]. From

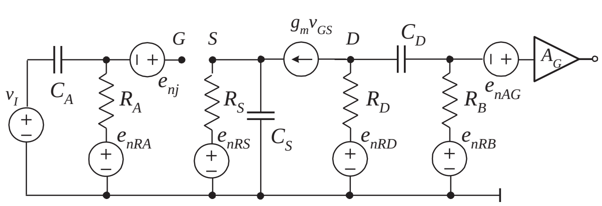

Fig. 2. Small signal equivalent for the circuit in Fig. 1. All relevant noise sources are explicitly shown.

(2) we can calculate that the passband gain A_{PB} (assuming R_B>>R_D) is:

$$A_{PB} = -A_G g_m R_D \quad (4)$$

We can now evaluate the Power Spectral Density (PSD) of the equivalent input voltage noise of the amplifier using Fig. 2. We will limit our analysis to frequencies well into the passband (all capacitors replaced by short circuits) in the reasonable hypothesis of uncorrelated noise sources. As before, we will assume R_B>>R_D. We have:

$$S_{ENI} = S_{ENJ} + \frac{4kTR_D}{(g_m R_D)^2} + \frac{S_{ENAG}}{(g_m R_D)^2} \quad (5)$$

where S_{ENI} is the *PSD* of the overall equivalent input noise of the amplifier, S_{ENAG} is the equivalent input noise of the solid state amplifier with gain A_G, k is the Boltzmann constant and T is the absolute temperature. With reference to (5) it is important to note that even assuming conservative values for g_m and R_D (say g_m=10 mA/V, R_D=1 kΩ), the noise coming from the resistance R_D and the equivalent input noise of the second stage would be divided by a factor 100, so that neither of them, with proper design of the solid state amplifier, contribute significantly to the overall equivalent input noise. The *EIVN* of the amplifier can be therefore reduced to the equivalent input noise of the JFET alone. While the extremely low noise represents an advantage with respect to feedback configurations [5], we are left with the problem of a gain that is directly proportional to g_m that, as we have noted before, does change considerably from one device to another. Our approach to address this issue is illustrated in Fig. 3.

With reference to Fig. 3, the uppermost amplifier (A_1) in the figure represents a very low noise amplifier with an unknown gain G_1, while the other two (A_2 and A_3) are Operational Amplifier (OA) based amplifiers with stable and well known gains G_2=G_3=G. Let us assume that we can perform cross correlation among all the three output channels. For the sake of simplicity we will assume all gains to be real. Ideally, the equivalent input noise of the amplifiers are cancelled out in the cross correlation process so that:

$$S_{O13} = S_I G_1 G$$
$$S_{O23} = S_I G^2 \quad (6)$$

where S_{Oxy} is the cross spectrum between outputs V_{Ox} and V_{Oy} and S_I is the *PSD* of the input V_I. Using (6) we have:

$$G_1 = G\frac{S_{O13}}{S_{O23}} \quad (7)$$

that means that we can calibrate the unknown gain G_1 from the known gain G and the measurement of the cross spectra between channels 1 and 2 and 2 and 3 independently of the shape of S_I. At a first sight, the following obvious question may arise: why bother with the design of a low noise amplifier

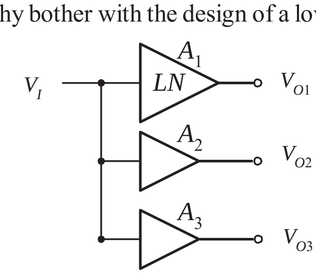

Fig. 3. Simplified diagram illustrating the approach we propose for the calibration of the gain of the main amplifier (A_1) using two auxiliary amplifiers connected to the input (A_2 and A_3).

if one can obtain the correct estimate of the input noise by cross correlation between channels 2 and 3 regardless of the level of equivalent input noise, as suggested in (6)? The fact is, as it is very well known, that reducing the uncorrelated noise contribution in cross correlation measurements requires averaging the results over several independent time records [10]. The level of uncorrelated noise decreases with the square root of the number of averages and therefore, the time required to uncover the correlated component (the one we want to measure) from the uncorrelated ones depends on the level of uncorrelated noise with respect to the correlated one. Suppose the uncorrelated noise is 10 dB above the noise to be measured; in order to reduce it to 10 dB below the noise to be measured the averaging of about 10^4 records is required. In *FFT* based spectrum analyzes, the duration of the record used for elaboration is the inverse of the resolution bandwidth Δf used for analysis. Especially in the case of the analysis of flicker spectra, the resolution bandwidth required to insure the absence of systematic errors must be significantly smaller than the minimum frequency of interest [11]. If the minimum frequency of interest is 100 mHz, as it can be the case in measurements on electron devices [1], the resolution bandwidth Δf must be in the order of 10 mHz and therefore the record length is in the order of 100 seconds ($1/\Delta f$). This means that performing 10^4 averages, as in the example above, would require more than ten days of uninterrupted measurement. If we go back to the approach we propose, assuming that the gains of the amplifiers in Fig. 3 be constant over a sufficient frequency interval, we can use a conveniently larger resolution bandwidth since we are only interested in calculating the gain G_1 of the amplifier A_1 according to (7). This allows to considerably reduce the time required to reach the correct estimate of the cross spectra: with Δf =100 Hz, for instance, 10^4 averages can be performed in less than two minutes. With these considerations in mind, our approach can be summarized as follows: we design a low noise amplifier (main amplifier) whose first stage is based on a single active device for obtaining the lowest possible *BN* at very low frequencies; we add two auxiliary OA based amplifiers to obtain the configuration in Fig. 3; we employ cross correlation with a large resolution bandwidth among the outputs in Fig. 3 in order to obtain an accurate estimate of the gain of the main amplifier; the estimate gain is used to calculate the *PSD* of the noise at the input of the main amplifier form the measurement of the PSD at the output V_{O1} in Fig. 3.

III. System prototype and experimental results

The prototype of the system used for the demonstration of the approach we propose is reported in Fig. 4. With respect to the circuit in Fig. 1, the output stage is modified in order to allow to employ a transeristance amplifier based on the low noise OP27 as a gain stage. The configuration we employ is similar to the one in [12]. The overall passband gain of the amplifier is $g_m R_R$, and can reach values in the order of 60 dB for g_m close to 20 mA/V. C_S is chosen in such a way that the associated pole frequency remains below 10 mHz regardless of the actual value of *gm* in the expected interval between 10 and 50 mA/V. In this way the lower frequency corner is set by C_D together with R_D (10 mHz). This means that we can expect an essentially flat response starting from 100-200 mHz, depending on the position of the pole that set by g_m. The amplifiers A_2 and A_3 are nominally identical voltage amplifier based on TLC070 operational amplifiers as in Fig. 5. They are characterized by extremely low input bias current and equivalent input current noise (so that they do not interfere

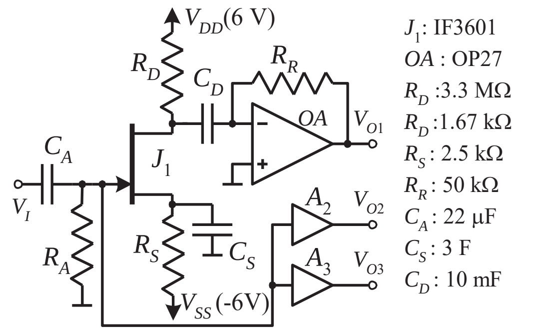

Fig. 4. Actual schematic of the circuit used for demonstrating the approach we propose. The schematic of the auxiliary amplifiers (A_2 and A_3) is shown in Fig. 5.

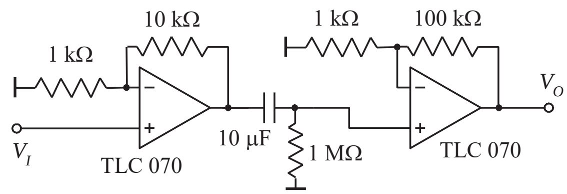

Fig. 5. Schematic of the auxiliary amplifier. The TLC070 operational amplifier is chosen because of its extremely low level of bias current and equivalent input current noise.

appreciably with the main amplifier circuit), have a passband gain of 61 dB and a low and high frequency corner of 16 mHz and 100 kHz, respectively. The equivalent input noise at 1 kHz is in the order of 10 nV/√Hz, and therefore much larger than the *EIVN* of the low noise amplifier that is expected to be below 1 nV/√Hz. Noise measurements are performed using a 4 channel PCI-4462 DSA board. The software for data acquisition and spectra and cross spectra elaboration has been developed around the QLSA library [13]. QLSA, as discussed in [13], allows to perform spectral estimation using a quasi logarithmic frequency approach. In the simplest terms, QLSA behaves as a set of conventional spectral analyzers in parallel, each covering a different frequency range with proper resolution bandwidth. This allows to estimate spectra and cross spectra at higher frequencies with large resolution bandwidth, and spectra at lower frequencies with lower and lower resolution bandwidths. Using QLSA is particularly useful in this application since cross spectra calculated at higher frequencies (with larger resolution bandwidths) can be used to calibrate the amplifier gain, while at the same time measurement in the very low frequency range are being performed with the required much smaller resolution bandwidth. In principle, using QLSA, gain calibration and corresponding spectra correction can be performed on line, during a single measurement session. As a first test we used a 50 kΩ resistance as a DUT. The thermal noise of the DUT at room temperature is 29 nV/√Hz, well above the expected equivalent input noise of main amplifier down to the hundreds of mHz range. The result of the estimation of the *PSD* of the noise at the output V_{O1} in Fig. 4 is reported in Fig. 6 (curve labeled S_{11}) and demonstrates a flat response down to the hundreds of mHz range. The decrease in the spectrum at frequencies above 2 kHz is due to the effect of the equivalent capacitance at the gate of the *JFET* that, because of the large area, is in the order of 0.5 nF, as can be verified in the datasheet. The spectrum at the output of one of the auxiliary

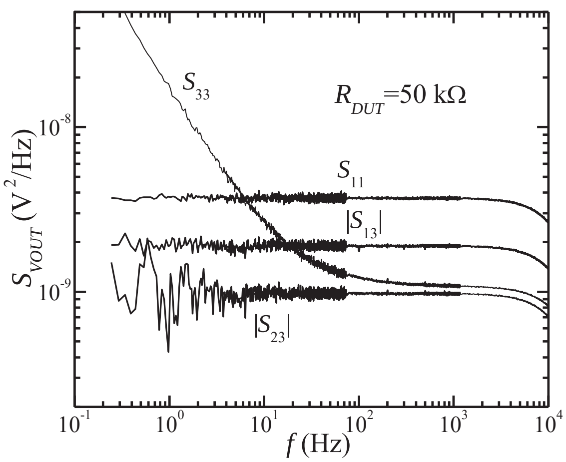

Fig. 6. Spectra and cross spectra as evaluated at the outputs of the circuit in Fig. 4 when a 50 kΩ resistor is used as a *DUT*.

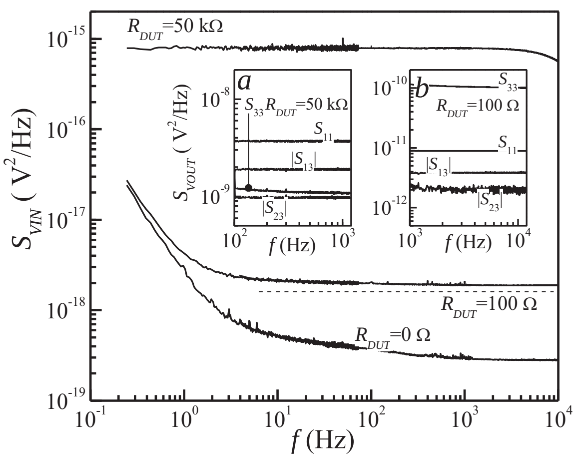

Fig. 7. Equivalent input noise as measured after gain calibration when a 50 kΩ, 100 Ω and 0 Ω are at the input of the amplifier. The insets reports the noise level (measured at the outputs) and used for gain calibration.

amplifiers is also shown in Fig. 6 (curve S_{33}) together with the cross spectra S_{13} and S_{23} required to calibrate the gain according to (7). Using the value of the cross spectra in the frequency range from 100 Hz to 1 kHz, we estimate a value of 2160 for the gain $g_m R_R$, corresponding to a transconductance value g_m=43 mA/V. With this value for the gain, we can calculate the equivalent noise at the input of the amplifier that, as shown in Fig. 7, corresponds to the expected value for the thermal noise of 50 kΩ (8.3×10^{-16} V^2/Hz). The process is repeated for a resistance of 100 Ω as a *DUT*, that is in a situation in which the noise of the *DUT* is much smaller than the EIVN of the auxiliary amplifiers (inset *b* in Fig. 7). As it is apparent from Fig. 7, the resulting equivalent input noise after gain calibration is larger than the one expected form a 100 Ω resistance (dashed line) and this is due to the contribution of the *EIVN* of the main amplifier that is no longer negligible. In order to estimate the *EIVN* of the main amplifier, we measured the *PSD* at the output V_{O1} and calculated the equivalent input *PSD* using the value of the gain measured in the case of R_{DUT}=100 Ω. As reported in Fig. 7, the *EIVN* of the amplifier that has been build is below 1 nV/√Hz above 2 Hz, and approaches 0.5 nV/√Hz for frequencies above 1 kHz.

IV. Conclusions

We proposed an approach for the realization of a very low noise voltage amplifier for applications in the field of *LFNM* by using a single *JFET CS* stage as the input stage. The approach we propose results in extremely simple hardware and in the minimum possible level of background noise for a given device, as it is reduced to the one introduced by the *JFET* alone. By employing supercapacitors for bias network decoupling, the passband can easily extend down to the hundreds of mHz range and below. No adjustment to the circuit parameters is required to calibrate the gain: the value of the gain is obtained during the actual measurement on the *DUT* by resorting to the properties of cross correlation applied to a proper configuration including, besides the main amplifier, two more *OA* based preamplifiers characterized by stable and known gain.

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