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DESIGN NOTE

The accurate measurement of load currents provided by high-voltage DC power supplies

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Abstract. The simplest way to measure a current is to measure the voltage drop across a resistor in the same current loop. A resistor in the power return is the preferred solution, even with sophisticated loads such as particle detectors. Unfortunately, most detectors of this type cannot operate with a floating low-voltage terminal and other solutions must be adopted. Various circuits to solve the problem of detector current measurement are proposed.

Keywords: high voltage, low current, ground loop, DC power supply, linear regulator, low current monitoring, standard components, design and error budget

Introduction

Voltage regulators with linear stabilization and designed with a bipolar junction (BJT) or field effect transistor (FET) in the output stage normally cover a voltage range up to 1200 V. Since they have excellent low-noise performance they are often used in particle detectors which require a precision high-voltage (HV) supply. Such detectors may draw a current as low as a few nanoamperes, which in many experiments must be accurately measured. Monitoring of a low current provided by a HV linear regulator can cause some non-trivial design problems especially if a high resolution is desired and at the same time the HV across the detector has to be kept stable.

1. Sensing resistor at the output

If the HV is allowed to vary with changes of the detector's current within a measurable value, but, for any reason, a resistor in the power return cannot be used, then the usual way of current monitoring is to measure the voltage drop across a sensing resistor in the output (figure 1). However, this is not normally feasible because the mean potential of the sensing resistor is far beyond the maximum input voltage rating of every opamp connected to ground.

To convert the voltage drop across a sensing resistor in a HV output to a ground related potential and, if necessary, to perform an amplification, a well-known circuit [1] can be used (figure 2).

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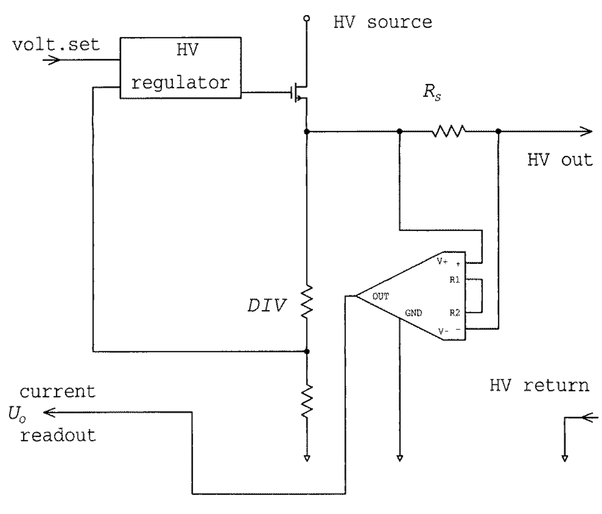

Figure 1. Sensing resistor at the output of a voltage regulator.

The output voltage U_O corresponds to the load current I_L:

$$U_O = I_L \frac{R_S R_D}{R_{SC}} + U_{Off} \frac{R_D}{R_{SC}} + I_B(R_S - R_{SC}) \qquad (1)$$

where U_{Off} is the input offset voltage and I_B the input bias current of the opamp.

The error budget of the circuit contains a scale error caused by the resistors' tolerances and an offset error caused by the input offset voltage U_{Off} and the input bias current I_B of the opamp. The accuracy of the resistor must be selected to match the required accuracy of the current measurement.

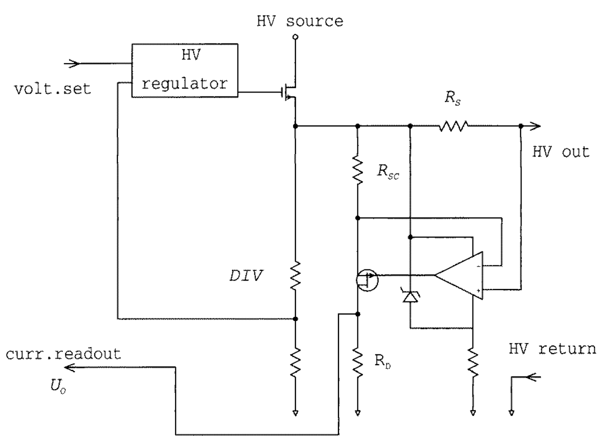

Figure 2. Sensing resistor at the output of a HV regulator.

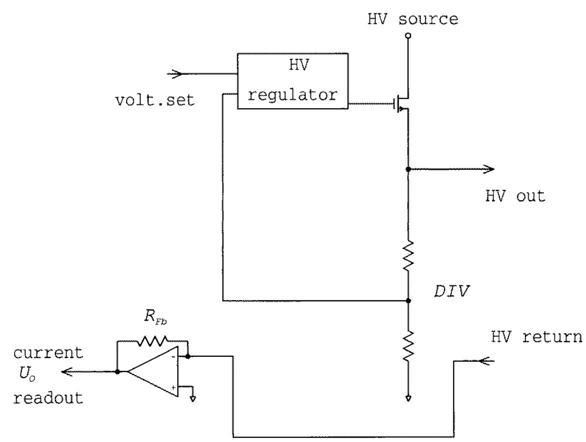

Figure 3. I-to-V converter in the HV power return.

The choice of the opamp is more critical because the following features must be taken into account.

- *Input offset voltage and input bias current.* The best way to avoid these error sources is to use a FET opamp with external offset adjustment by a potentiometer.
- *Input/output performance.* The input common mode voltage range and the output voltage swing must include the positive supply rail.
- *Supply current.* Because the negative supply has to float parallel to the positive supply, a Zener diode connected between the supply pins and to a high value resistor connected to ground represents a possible solution. In this case the opamp has to be a low-power type.

One advantage of this circuit is the possibility to connect several detectors with their own current measurements to one HV regulator. The circuit is relatively simple, but its application is limited because p-channel FETs with a breakdown voltage over 500 V are not yet commercially available.

2. *I*-to-*V* converter in the power return

If a resistor's voltage drop reduces the stability of the output voltage to an unacceptable level, then it cannot be used in the output. A simple way to provide HV on a floating load and to measure the load current is to place an I-to-V converter in the power return of the regulator (figure 3). The inverting input of the opamp is a virtual ground, so there is no decrease of the output voltage across the detector as there would be with a resistor in the output.

The output voltage U_O as a function of the load current I_L is given by:

$$U_O = I_L R_{Fb} + U_{Off} + I_B R_{Fb} \tag{2}$$

where I_B is the input bias current of the opamp.

The error budget looks much better than in the circuit of figure 2 (equation (1)). Resolving low currents down to a few nanoamperes is possible without using exotic components as long as the input bias current I_B of the opamp is negligible compared with the measured current. The accuracy of the measured voltage as a function of the current flowing to ground depends on the precision of the feedback resistor and on the input offset of the opamp—both values are easy to control by selecting the proper type of components. However, this simple and reliable design cannot be used if the detector is *a priori* grounded (to physical ground) or if, in the case of a multichannel power supply, the detector modules have a common power return. Both conditions often occur in the use of particle detectors where proper grounding of the detector is an important condition for the reliable performance of its wideband readout electronics.

3. *I*-to-*V* converter at the HV output

Nevertheless the I-to-V converter with an opamp is not a useless design: the opamp simply has to be moved from the power return to the HV output of the regulator (figure 4). In this case the HV output is connected to the inverting input of the opamp and the feedback resistor R_{Fb} is the source of the output current. As with the circuit in figure 3 the voltage across the feedback resistor corresponds to the current drawn by the detector. Relative to ground, the opamp's output voltage is the sum of its own output voltage and the pre-set output HV. To subtract the last term and to get the difference value relative to ground, this sum voltage is applied to a second voltage divider with a division ratio RAT_2 nominally equal to the main voltage divider ratio RAT_1. An instrumentation amplifier connected to the lower ends of both dividers performs the subtraction and compensates for the attenuation of the dividers, amplifying the result by a factor A_{IA} which is also nominally equal to the dividers' ratio. Thus the output voltage of the instrumentation amplifier should be equal to:

$$U_O = \left(\frac{I_L R_{Fb} + I_B R_{Fb} + U_{Off} + U_{HV}}{RAT_2} - \frac{U_{HV}}{RAT_1} \right) A_{IA}. \tag{3}$$

As the error budget shows, a serious drawback of this circuit is its sensitivity to the difference between the effective values of RAT_1 and RAT_2, since the instrumentation amplifier which compensates the attenuation of both dividers amplifies this difference as well. A manual correction made with a balancing potentiometer between the ground terminals of both dividers is an appropriate cure.

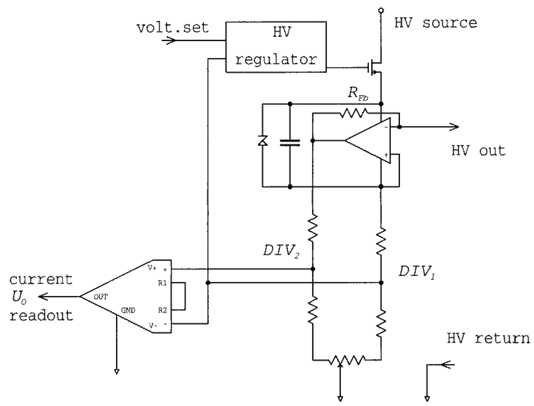

Figure 4. I-to-V converter in the HV output.

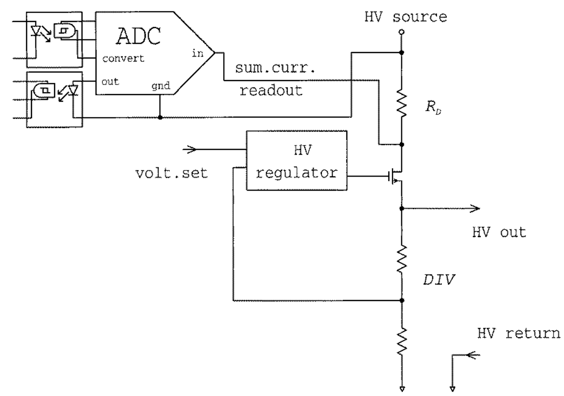

Figure 5. Current readout via a drain resistor.

Obviously, the opamp used in this circuit cannot be of arbitrary type.

- It has to be an ultra micropower type, because the only way to supply it is to use the same current flowing through the main voltage divider (a few microamperes).
- Its common mode input voltage range and its output voltage swing have to include the negative supply rail. The choice of such opamps has been small up to now, but might increase in the near future with the rapid development of micropower devices.

The opamp MAX 406 ACP (Maxim) meets all these requirements quite well. It is an input and output rail-to-rail amplifier with a supply current of 1.2 μA [2]. This opamp was used for tests of this design. A 1 MΩ 0.1% precision resistor in the feedback provides a 1 V μA^{-1} output swing. To regain the attenuation introduced by the two dividers (Caddoc, thick film, 1 MΩ, ratio 1:100) a chopper-stabilized instrumentation amplifier, type LTC1100 (Linear Technology) with a gain of $\times$100 was used. The design provides a measurement accuracy better than 2 nA over an output current range of 0–1 μA.

4. Sensing drain resistor

For regulators using a FET (but not a BJT) in the output stage there is another possible solution to the problem described above (figure 5). In this circuit a resistor is inserted in the drain of the output FET, where it does not affect the performance of the regulator. The voltage drop across the drain resistor corresponds to the sum of the divider current and the load current. If an analogue to digital converter (ADC) is shifted up to the potential of the non-stabilized HV source, this drop can be measured directly as a negative voltage relative to the ground of the shifted ADC. Because the divider current depends only on the pre-set output HV, and because the regulator's design usually includes an ADC for the readout of this value, digital subtraction of the divider current from the sum is a task for a processor which controls the operation of both ADCs.

The problem of processing the digital readout and control signals of the ADC on a high potential is easily handled with a few optocouplers. However, the supply of the few opamps, optocouplers and the ADC on the potential of the HV source requires special attention. If all the electronics must work in a strong magnetic field of a few tesla, as is often the case with particle detectors, iron or ferrite transformers cannot be employed. The use of capacitor-coupled charge pumps, piezo transformers, or photovoltaic devices is recommended.

The main problem with the readout of the load current via the voltage drop across a drain resistor is not the requirement of a floating supply, but the fact that only part of this drop is caused by the load current. The remaining part is caused by the divider current:

$$U_{RD} = R_D(I_L + I_{DIV}). \tag{4}$$

A simple calculation shows that measurement of the load current is more accurate when the load current is several times larger than the divider current. Unfortunately, in most practical applications the opposite is true. So the drop across the drain resistor of the output FET caused by the load current is only a small part of the total voltage drop. In these conditions a significant part of the resolution of the current readout ADC is lost.

The use of this circuit is based on the assumption that there exists a common HV DC source (stabilized or not) which supplies the drains of all output FETs for all HV regulators. Thus the shifted-up ADC for the current readout becomes a device common to all regulators in the assembly. The number of components for this circuit looks quite high, but the more HV regulators it serves the more attractive it becomes.

5. A remark

All the circuits discussed above refer to HV regulators for positive output voltages only. In principle they could be used for negative voltages as well but the choice of HV semiconductor components currently on the market is limited. The main problem in the design of negative HV linear regulators, even without any current measurement, is the type of output transistor. The lack of p-channel FETs with a breakdown voltage exceeding 500 V has already been mentioned, but HV p–n–p BJTs are also difficult to source. Fortunately, for most types of detector, either HV polarity may be used.

6. Conclusion

Choosing the appropriate circuit depends upon a number of conditions.

- A sensing resistor in the power return should be employed if the detector will tolerate a limited variation of the HV with load current, and if the detector can be tied to the sensing resistor instead of ground.
- A sensing resistor at the HV output should be used if the detector will tolerate a HV variation with load current, but requires ground. Depending on the output voltage range of the HV regulator and the resistance of the HV divider, there are some limitations on the choice of active components (opamps, p-channel FETs) for the HV output stage.
- If the detector requires a very stable HV but will accept a virtual ground, the use of an I-to-V converter in the power return is recommended. However, whether the detector can actually accept virtual ground cannot be calculated and is quite difficult to estimate. It depends not only on the detector itself, but also (and mainly) on the features of the connected detector readout assembly and its grounding. If there is a suspicion that oscillations will occur there, an I-to-V converter at the HV output has to be preferred.
- If the complete system contains a large number of HV regulators tied to a common HV DC source and every regulator has a single load which requires its own current measurement, it is reasonable to use a drain resistor as current sensor.

Acknowledgments

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[2] 1998 *Maxim Full—Line Data Catalog on CD ROM* 1998 Edition Version 2.0, Document 19-4739, Rev 2, 7/93