What does precision mean for an op amp?

May 2023

User guide

The following presentation is the **textual explanation of the video: "What does precision mean for an op amp."** Also available in the **audiovisual version on our YouTube channel** accessible through the following link.

[Video on Youtube]

Recommendations

In this document you will find concepts explained through informative tables, graphs, sequences of graphs, which will be followed by a detailed text explanation.

- So, when you see this icon: [Explanation] it means that in the next pages you will find an explanation text about this topic.
- Each page contains an explanation text that indicates the page or group of pages on the explanation being made: (page # to #)
- If you se this icon: [Continue] it means that the explanation text on this topic continues on the following pages

Op amps: Vio – input offset voltage

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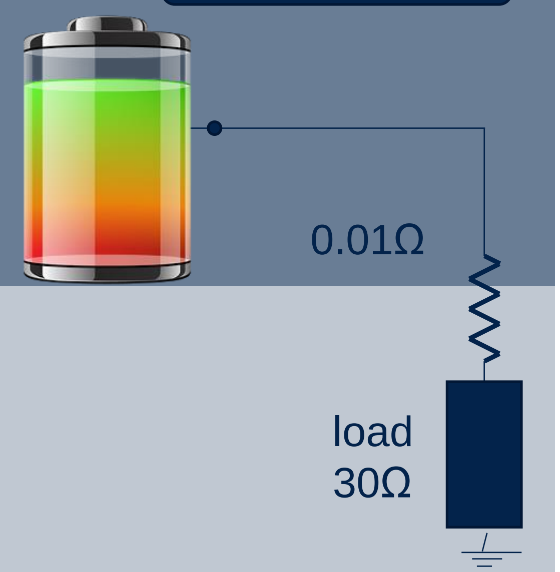
Why op amps and why precision?

Battery fuel gauging

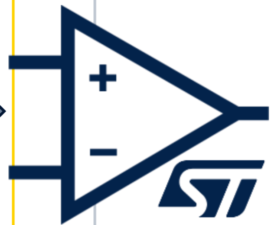
4.2V to 2.7V



0.9 mV ≤ **V1max** ≤ 1.4 mV

Very small signal

Need to amplify



STM32 power supply:
1.65V to 3.6V

ADC 12bits

LSB of the ADC
$= 3.6V / 2^{12}$
$= 0.88$ mV!

Explanation

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Why op amps and why precision?

EXPLANATION (page 4)

Today sensors are everywhere, in our modern world we need to constantly measure everything, weight, UV, temperature, speed, current, even though we live in an increasingly digital world, many are still purely analog and generally most of them provide an extremely small signal. The question raised is about how to deal with this slow signal in order to transfer it to the digital domain without adding any error that could compromise the information. The solution is to use an op amp to amplify the signal, but this op amp needs to be ideal not to introduce any offset; unfortunately, this kind of device does not exist.

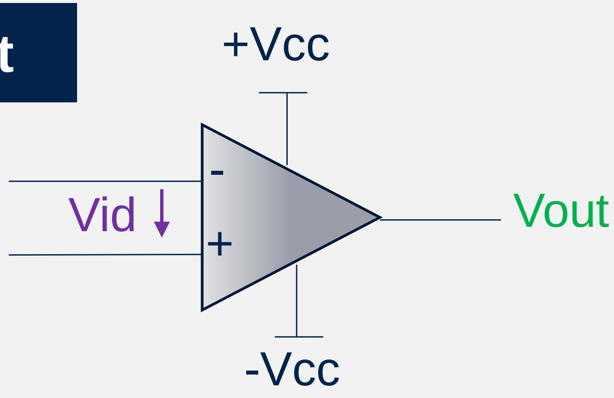
Op amps will always impact measurements, however by choosing the right op amp we can limit the error on the output as much as possible.

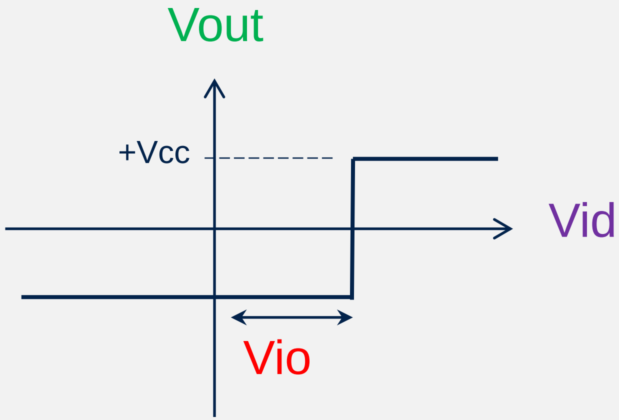
In the example shown in this slide, let’s consider a battery delivering a maximum current of 140mA through a shunt of 10mA. The resulting voltage will be 1.4mV maximum, as the LSB of the ADC is 880µV, we clearly need to amplify the signal to increase the accuracy.

The main goal of this video is to identify which op amp parameters are the most relevant with regard to precision.

Input offset voltage
What is this?

Vio offset





LM324

Symbol	Parameter	Min.	Typ.	Max.	Unit
V_{io}	Input offset voltage (1) $T_{amb} = +25°$ C LM124-LM224 LM324		2	5 7	mV
	$T_{min} \leq T_{amb} \leq T_{max}$ LM124-LM224 LM324			7 9	

TS507

V_{io}	Input offset voltage(2)	V_{icm} = 0 to 3.8 V, T=25 °C TS507C full temperature range TS507I full temperature range		25	100 250 400	µV
		V_{icm} = 0 V to 5 V, T=25 °C TS507C full temperature range TS507I full temperature range			450 550 750	

TSZ121 (Very high accuracy)

Symbol	Parameter	Conditions	Min.	Typ.	Max.	Unit
DC performance						
V_{io}	Input offset voltage	T = 25 °C		1	5	µV
		-40 °C < T< 125 °C			8	
$\Delta V_{io}/\Delta T$	Input offset voltage drift(1)	-40 °C < T< 125 °C		10	30	nV/°C

Explanation

Input offset voltage
What is this?

EXPLANATION (page 6)

The main limitations to precision in analog integrated circuits are noise and mismatch.

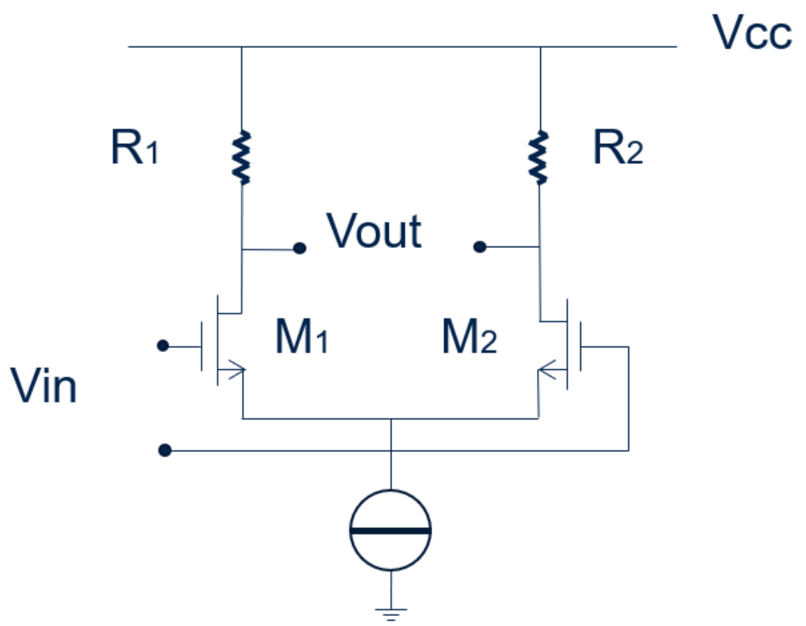
Before we start let's define a few terms:

What is offset? When a 0 differential input voltage is applied the op amp output voltage should be zero in practice this is not the case, the offset voltage by definition is the differential input voltage that is required to make the output voltage 0. The input offset voltage parameter generally called Vio is defined as a DC voltage between the noninverting and the inverting input. It is always specified in the electrical characteristics of a datasheet and can be either positive or negative.

The Vio value will be different for each op amp and to ensure good precision it is important to choose the op amp with the lowest Vio, because the Vio will be amplified by the gain and added to the total output as an error. Some op amps with a chopper architecture like ST **TSZ121** exhibit a Vio of 5µV, they are especially good for precision DC measurements.

Input offset voltage
What is this?

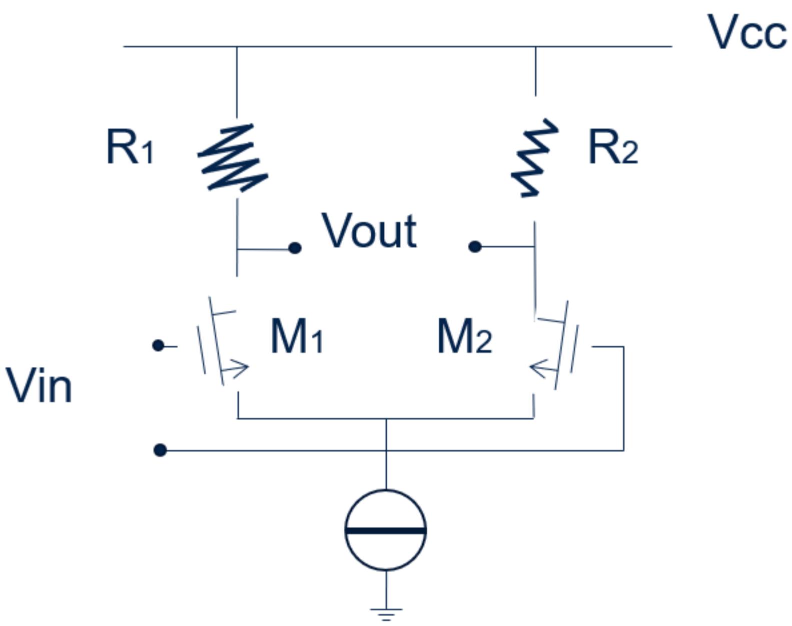
Differential input



Explanation

Input offset voltage
What is this?

Differential input



Component mismatch
$R1 \neq R2$, $M1 \neq M2 \Rightarrow$ **offset**

For CMOS technology

$$Vos = \Delta Vth + \frac{VGS - VT}{2}\left(\frac{\Delta R}{R} + \frac{\Delta k'}{k'} + \frac{\Delta W/L}{W/L}\right)$$

ΔVth linked to the substrate doping
Second term linked to the size of MOS

Mismatch is mainly due to:
- Doping variations
- Lithographic errors
- Packaging & local stress

Explanation

Input offset voltage
Where does it come from?

EXPLANATION (page 8 to 9)

The cause of the input offset voltage is well known,

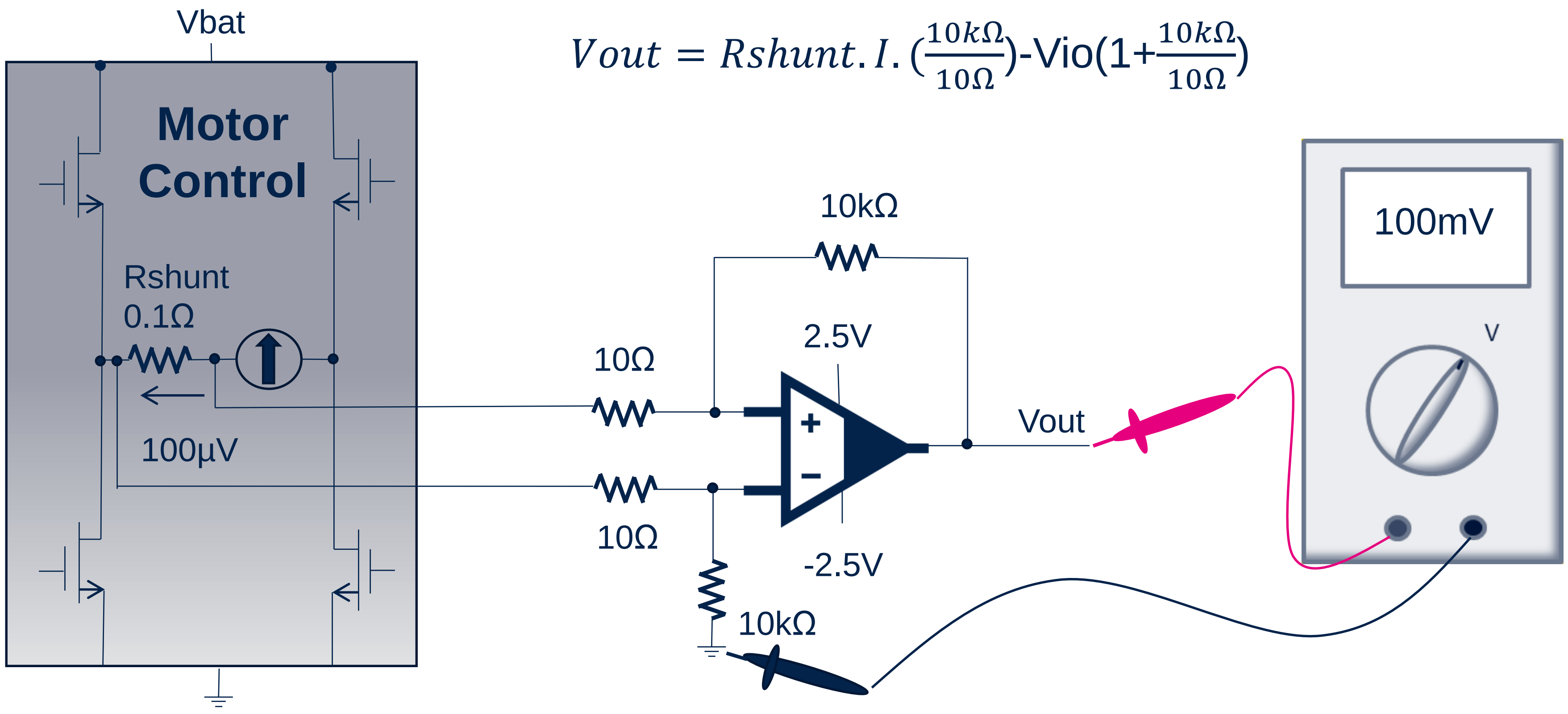
(page 9) it is due to the inherent mismatch of the input transistors and components during fabrication of the silicon die, and the stress placed on the died during the packaging process a minor contribution. These effects collectively produce a mismatch of the bias of the input circuit resulting in a differential voltage at the input terminals of the op amp.

For CMOS technologies this equation shows the root cause of the input offset voltage. We can see that the size of the MOS width and length and the doping substrate play a key role in the Vio error. The resulting expression contains four terms since all of them can be either positive or negative, they never all add up while this never occurs in practice, they never cancel each other out neither. In any case regardless of the design or process effort there always remains a small input offset voltage. This also explains why for a dual amplifier where two op amps are in the same package the input offset voltage of both channels is different.

Let’s now look at how the input offset voltage can impact the theorical measurement of an application.

Impact of V_{IO} on a real application
current sensing for Motor control

$$Vout = Rshunt.I.\left(\frac{10k\Omega}{10\Omega}\right) - Vio\left(1+\frac{10k\Omega}{10\Omega}\right)$$

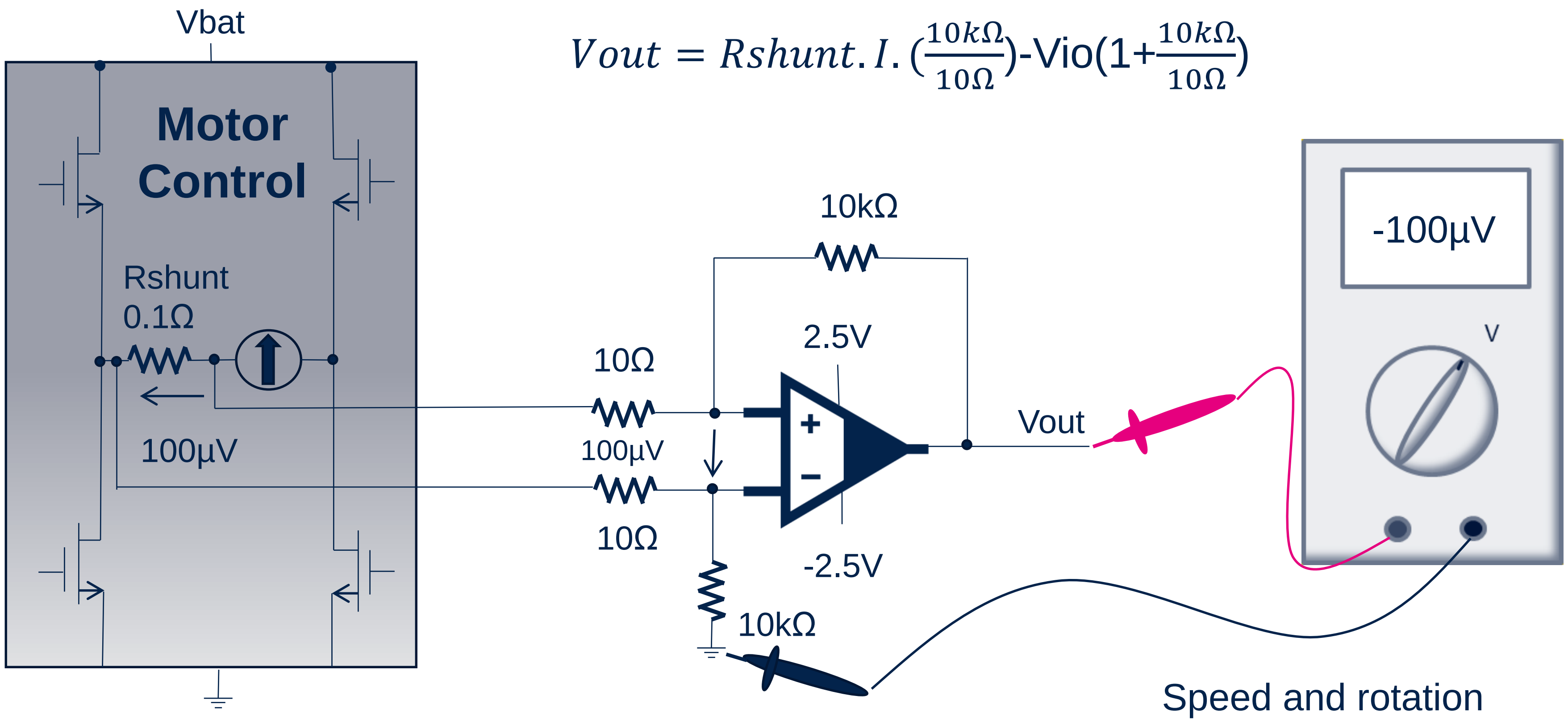


I=1mA

Impact of V_{IO} on a real application

current sensing for motor control

$$Vout = Rshunt.I.\left(\frac{10k\Omega}{10\Omega}\right)-Vio\left(1+\frac{10k\Omega}{10\Omega}\right)$$



I=1mA

TS507
Vio=100µV

Speed and rotation information is incorrect

Explanation

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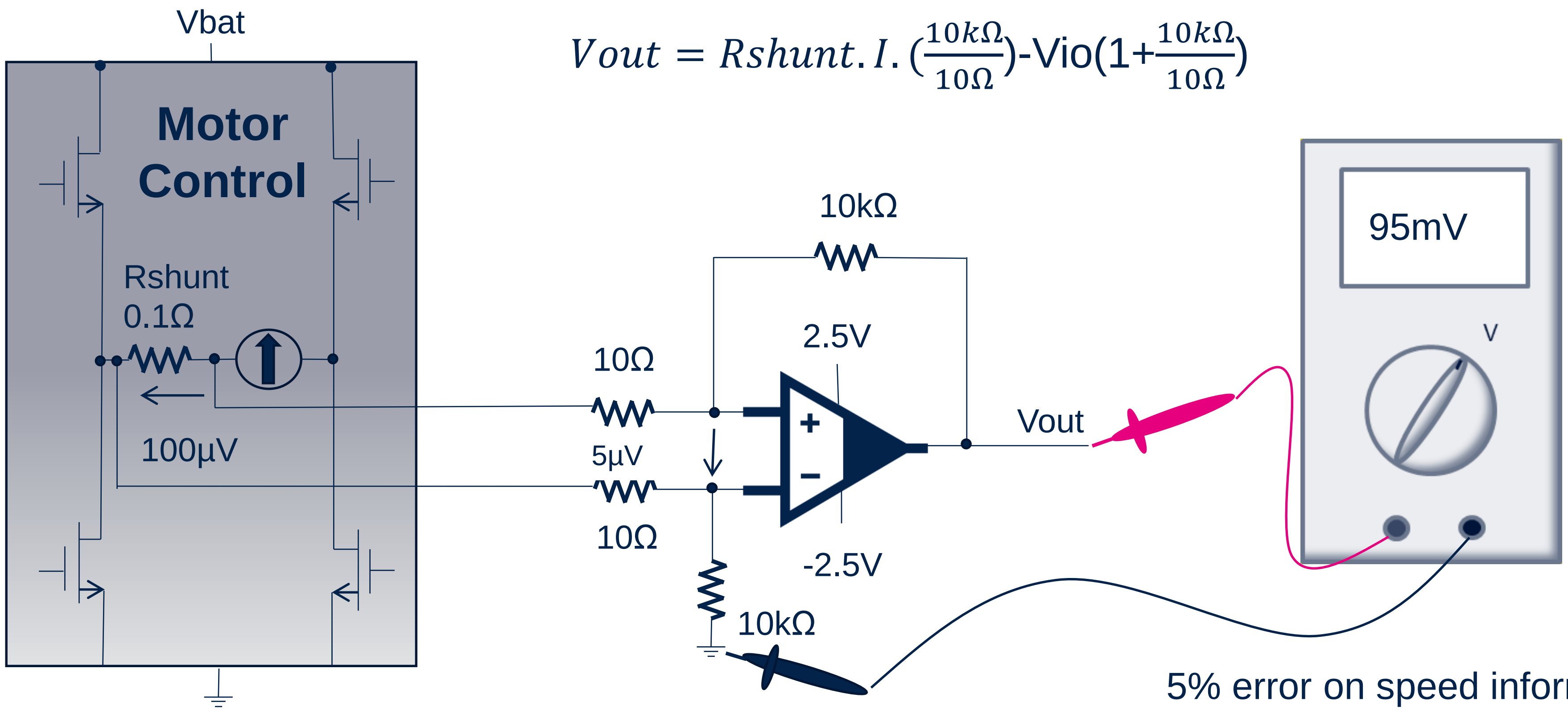
Impact of V_{IO} on a real application
current sensing for Motor control

$$Vout = Rshunt.I.(\frac{10k\Omega}{10\Omega}) - Vio(1+\frac{10k\Omega}{10\Omega})$$



I=1mA

TSZ121
Vio=5µV

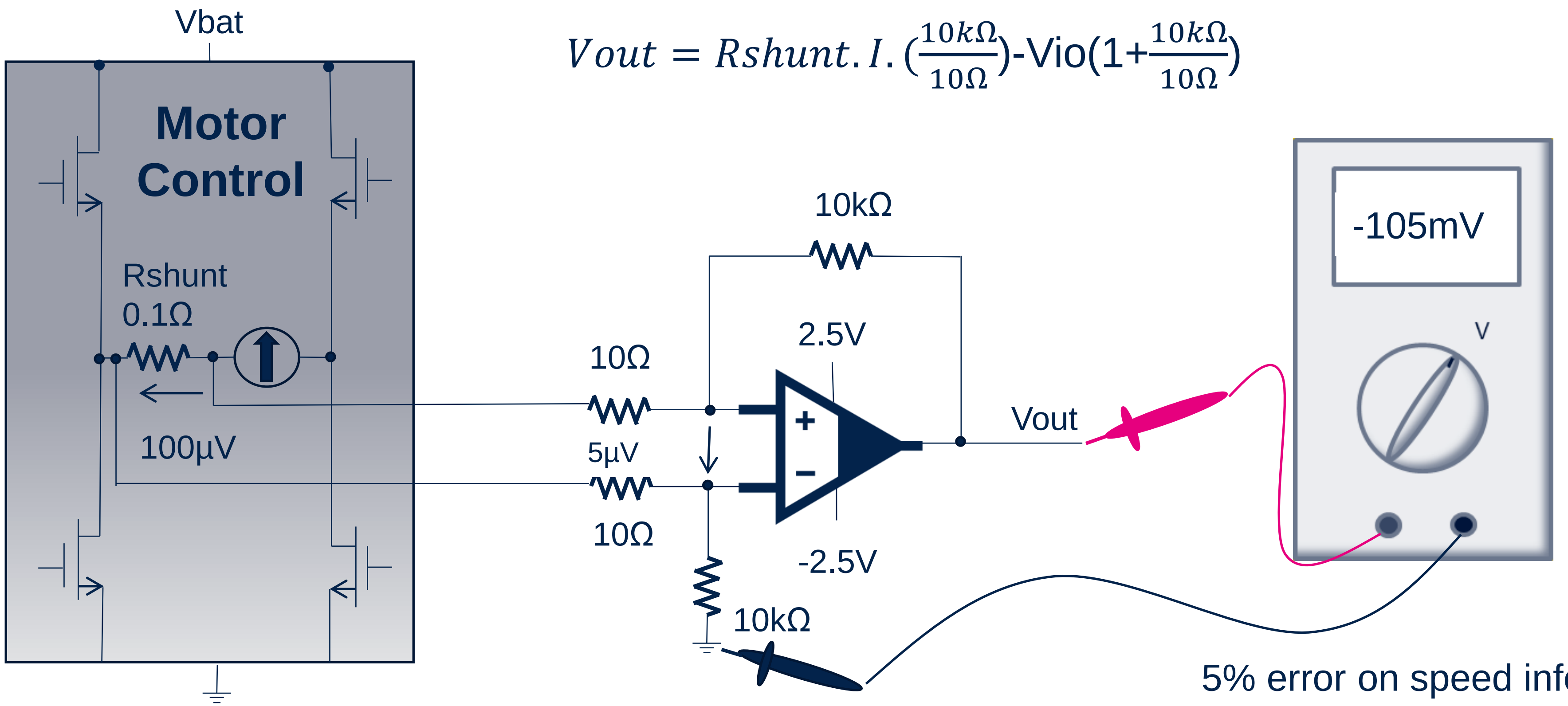
5% error on speed information
rotation information is correct

Explanation

www.st.com/OpAmps

Impact of V_{IO} on a real application
current sensing for Motor control

$$Vout = Rshunt.I.(\frac{10k\Omega}{10\Omega}) - Vio(1+\frac{10k\Omega}{10\Omega})$$



I=1mA

TSZ121
Vio=5µV

5% error on speed information
rotation information is correct

Explanation

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Impact of V_{IO} on a real application
current sensing for Motor control

EXPLANATION (page 11 to 14)

Shunt current sensors are used in precision current sources for feedback control systems. They are also used in a variety of other applications, including battery fuel gauging, and torque feedback controls in electric power steering and precision power metering.

In this slide we will look at a motor control application.
Thanks to the H bridge, the motor can be driven in both directions. The current measurement helps to know the motor's speed and its rotation direction. In such applications, it is desirable to use a shunt with a very low resistance to minimize the series voltage drop; this wastes less power and allows the measurement of high currents without a significant voltage drop. A typical shunt might be 0.1 Ω (ohm).

The lower the current, the lower the resulting voltage through the shunt resistor. So, at high current the Vio of the op amp has little impact. For low current it may become critical.

When the current is only few amps, the shunt's output signal is only a few hundred of millivolts. For example, at 1mA, the voltage through the shunt is 100µV to which the Vio must be added. So, the op amp demands a very low offset voltage and drift to maintain absolute accuracy.

Continue

Impact of V_{IO} on a real application
current sensing for motor control

EXPLANATION (page 11 to 14)

(Page 11) the op amp is used as a differential op amp in order to amplify the voltage drop appearing through the shunt resistor. In this example the signal is amplified by 1000. With an ideal op amp Vio = 0 mV. So only the first term of the equation is valid. Normally we read an output voltage of 100mV, but as mentioned previously, the world is not perfect, and the ideal op amp doesn't exist.

(Page 12) If we use a **TS507** op amp with a maximum positive Vio of 100µV at ambient temperature, we can see that the output voltage is -100µV. This is far from the expected value 100mV meaning that the speed information is completely incorrect. Moreover, this negative output will indicate to the MCU that the motor is rotating in the reverse way, which is totally incorrect. Here, we can see how the Vio of the op amp can impact of the entire application if the op amp is not chosen carefully.

(Page 13) if we use a precision amplifier, with a chopper architecture, like the ST's **TSZ121** with a Vio of 5µV at ambient temperature, we can see that we will make an error of 5% on the motor speed, which is acceptable when we use a gain of 1000.

(Page 14) if the motor rotation direction is changed, the theoretical output value considering an ideal op amp will be -100mV. We can see by using the **TSZ121** that even in a reverse rotation, the measurement remains valid. 5% error on the output.

ST Op Amps

Summary of V_{IO} impact on motor control applications

Op amp	Offset @ 25°C	V_{OUT} for a current I = 1 mA	Comment
Ideal	0 µV	100 mV	Theoretical measurement in a perfect world!
TS507	+100 µV	-100 µV	Speed of the motor is incorrect. Information about the motor rotation is incorrect
	-100 µV	200 mV	100% error on motor speed Information about the motor rotation is correct
TSZ121	+5 µV	95 mV	5% error on the motor speed Information about the motor rotation is correct
	-5 µV	-105 mV	5% error on the motor speed Information about motor rotation is correct

Explanation

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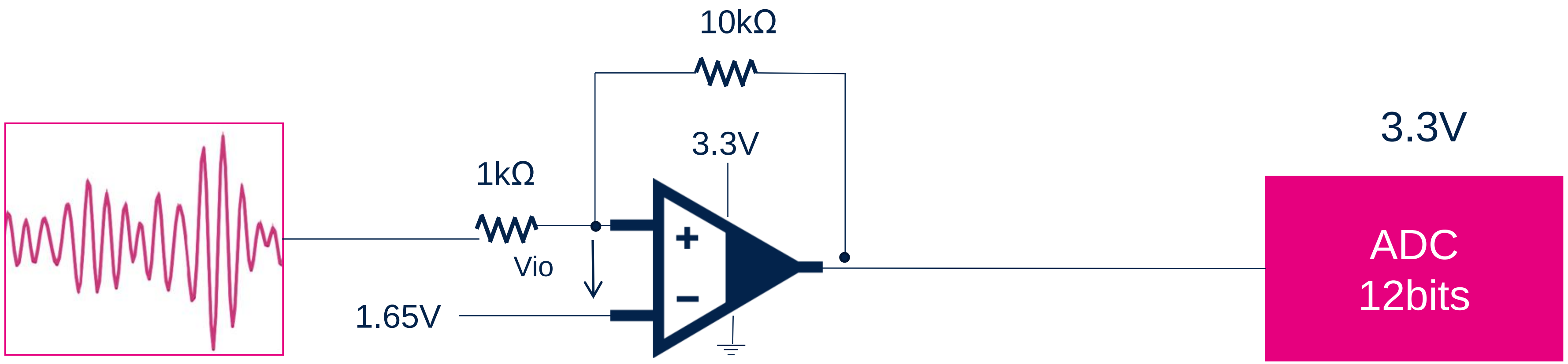
Summary of V_{IO} impact on motor control applications

EXPLANATION (page 17)

An op amp Vio can be either positive or negative, this table summarizes the real impact of the Vio with two different amplifiers **TS507** and **TSZ121**.

With the **TS507** and a positive Vio of +100µV, the result on the output is completely incorrect even in regard to the motors speed and rotation direction. If the Vio is negative the rotation direction of the motor is okay, but we make an error of 100% on the motor speed. Unfortunately, we cannot predict the polarity of the Vio from one part to another. Therefore, when designing, an important fact to take into consideration is that the Vio might be positive or negative, however by using the **TSZ121,** a precision amplifier op amp with an extremely low Vio the impact on the output result will be limited. The fact that the Vio can be either positive or negative will not impact the expected results.

The real cost of V_{IO}!



The LSB of the ADC is 3.3 V/2^{12} = 805 µV
The input signal is amplified by -10, and the V_{IO} by 11

	Maximum V_{IO}	Maximum offset at ADC	Equivalent effective ADC
TSZ121	5 µV	55 µV	~12 bits
TS507	100 µV	1.1 mV	~11 bits
TS512A	500 µV	5.5 mV	~9 bits
TS512	2.5 mV	27.5 mV	~7 bits

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Explanation

The real cost of V_{IO}!

EXPLANATION (page 19)

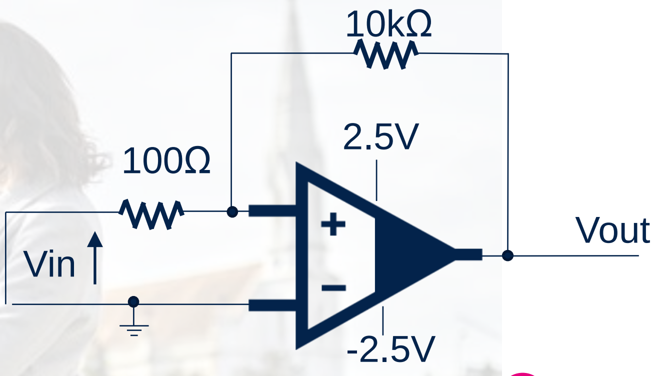
Another point we need to understand is the real impact of the Vio on an application in terms of effective cost.

The Vio also reduces the dynamic range of an ADC. The loss of dynamic range affects the resolution of ADC circuits because the maximum dynamic range is required for maximum resolution.

The exposed table, shows the equivalent resolution of the ADC for various input offset-voltage ranges. Usually, an op amp can be chosen with a Vio low enough to meet the desired accuracy. It is easy to find an op amp that meets the Vio specification for an 8- or 10-bit converter, but it becomes increasingly difficult as the resolution increases.

We can see that if the op amp is not chosen carefully all the money spent on a 12-bit ADC is lost.
Indeed, in this example, if the **TS512** op amp is chosen to amplify the signal, only 7 bits of the 12-bit ADC are effective. Whereas, by choosing the **TSZ121** which has an extremely low Vio of 5µV no LSB will be lost when using a 12-bit ADC and the entire resolution of the ADC can be used. Therefore, it is not necessary to spend money on a precision ADC if the op amp is not correctly chosen.

ΔVio/ΔT and calibration

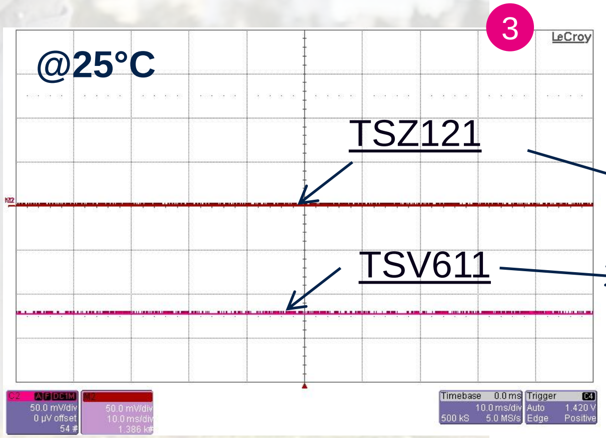


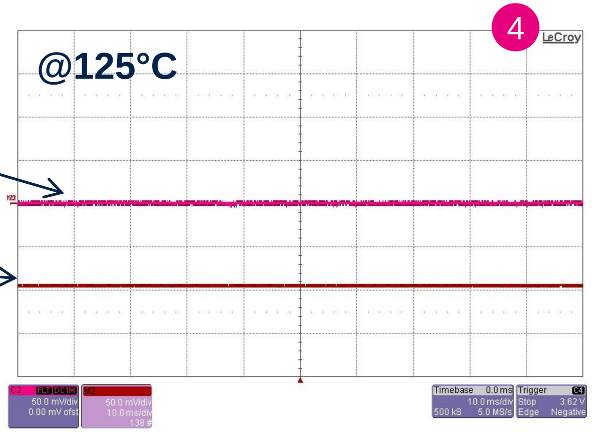
1

$$Vout = Vin(\frac{-10k\Omega}{100\Omega}) \pm Vio(1+\frac{10k\Omega}{100\Omega})$$

2

$$Vout = Vin(\frac{-10k\Omega}{100\Omega}) \pm (Vio \pm dT(\frac{\Delta Vio}{\Delta T}))(1+\frac{10k\Omega}{100\Omega})$$





3	Vio @25°C max	ΔVio/Δt max
TSZ121	5µV	30nV/°C
TSV611	4mV	10µV/°C

ΔVio/ΔT and calibration

EXPLANATION (page 21)

As the Vio is principally due to the mismatching of the input components, this parameter will also vary depending on the ambient temperature. It is though important to take into consideration in a precision environment the offset drift in temperature which is generally called ΔVio/ΔT .

1 The Vio is always multiplied by the noninverting gain of the op amp and added to the signal amplified by the circuit, which is the -100 in this example. The transfer function is determined by equation 1, where Vio is the maximum value written in the datasheet.

2 Adding the effects of the temperature to equation 1 gives equation 2.

This allows a quite accurate calculation of the worst-case change in output due to Vio, neglecting the effect of the resistors. However, the resistor values also change with temperature and will also affect the gain.

Typical drift values for general-purpose precision op amps lie in the rage of 1 to 10µV/°C.

In this schematic then equals 0 as the inputs are grounded, so the output of the op amp reveals Vio multiply -101.

3 This slide compares two amplifiers the **TSV611** a standard CMOS op amp with a maximum Vio of 4mV and the **TSZ121** a chopper op amp with a maximum Vio of 5µV.

Continue

ΔVio/ΔT and calibration

EXPLANATION (page 21)

3 At 25°C we can see that the **TSV611** shows an output offset of -125mV, whereas **TSZ121** is close to 0 micro volts. This shows that if the **TSV611** is used in an application requiring accuracy, a calibration must be made at manufacturing level to eliminate the offset introduced by the op amp.

4 However, when the temperature changes, we can clearly see on the oscilloscope that the output offset voltage of the **TSV611** changes from -125mV to - 90mV.

This means when that when using the **TSV611** a calibration also must be made on the temperature at the manufacturing level, which is very costly. While using the **TSZ121, which** is a zero-drift amplifier, we do not necessarily need this calibration phase as the Vio stays very low even with a large temperature variation, because it exhibits a delta Vio/delta T of 30nV/°C, so when the temperature increases by 100°C, the Vio will vary within a range of 3µV.

Op amps: CMRR – common-mode rejection ratio

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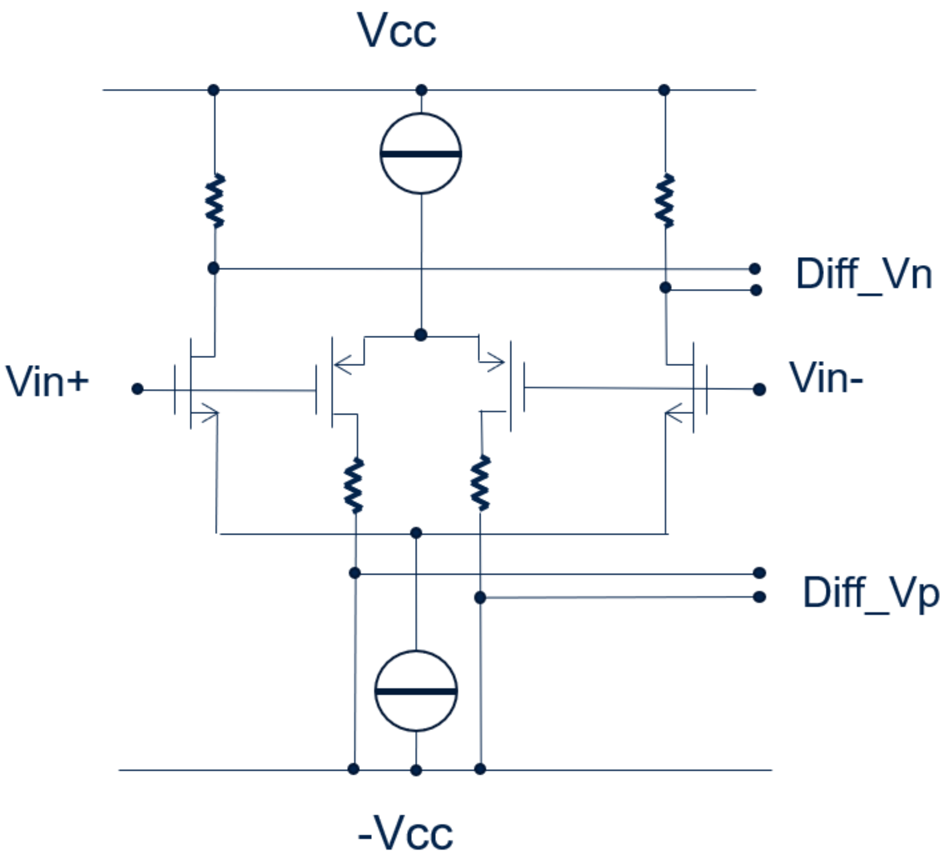
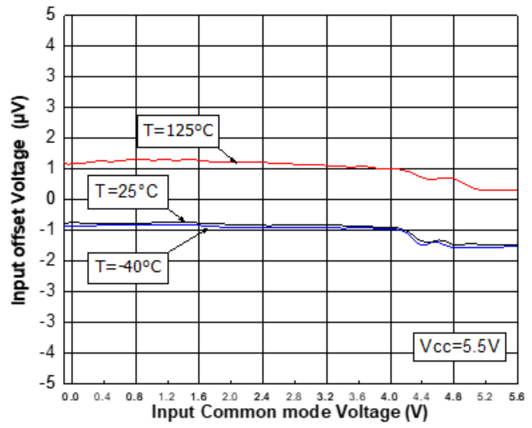
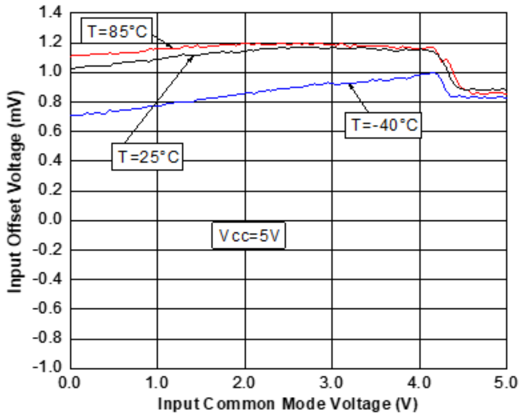
Common-mode rejection ratio

Input stage of a CMOS op amp



TSZ121





TSV611

Explanation

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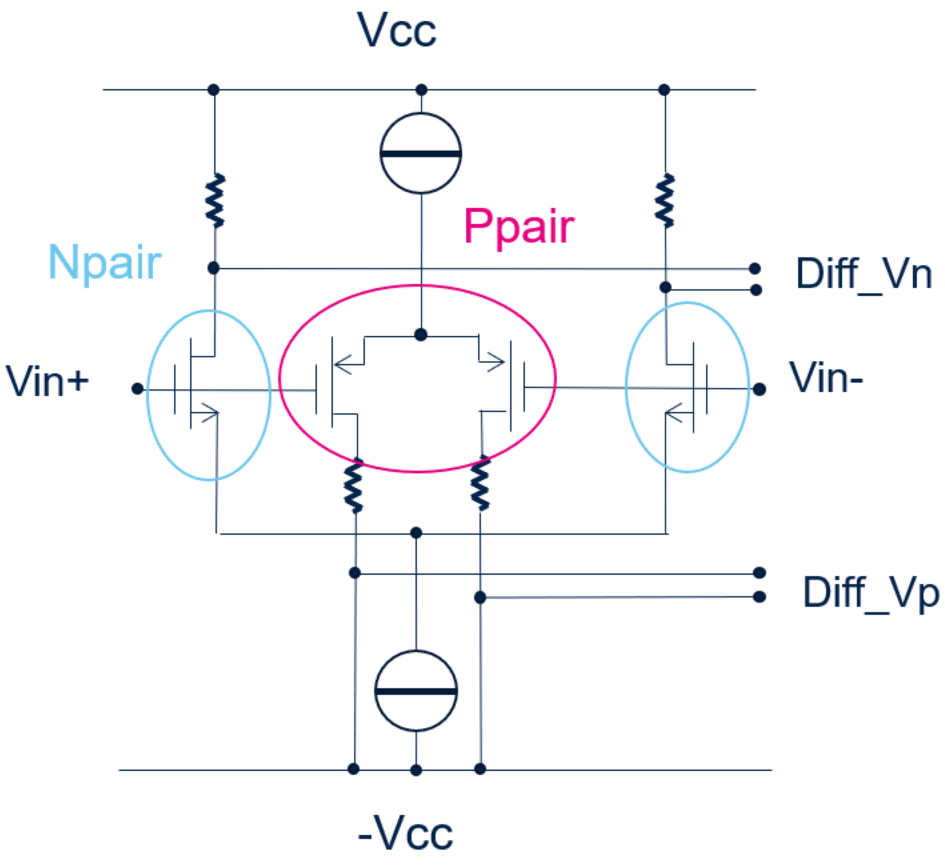
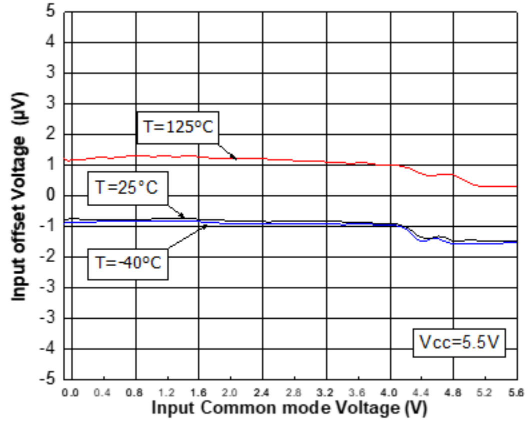
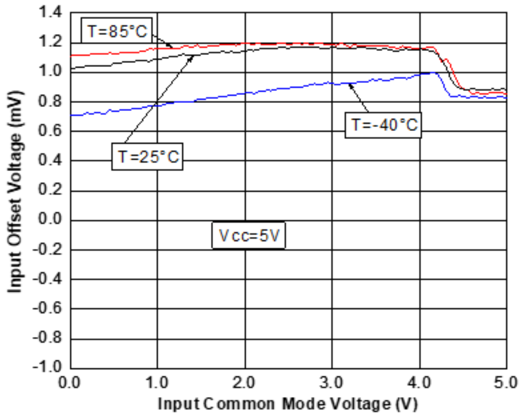
Common-mode rejection ratio

Input stage of a CMOS op amp



TSZ121





TSV611

Explanation

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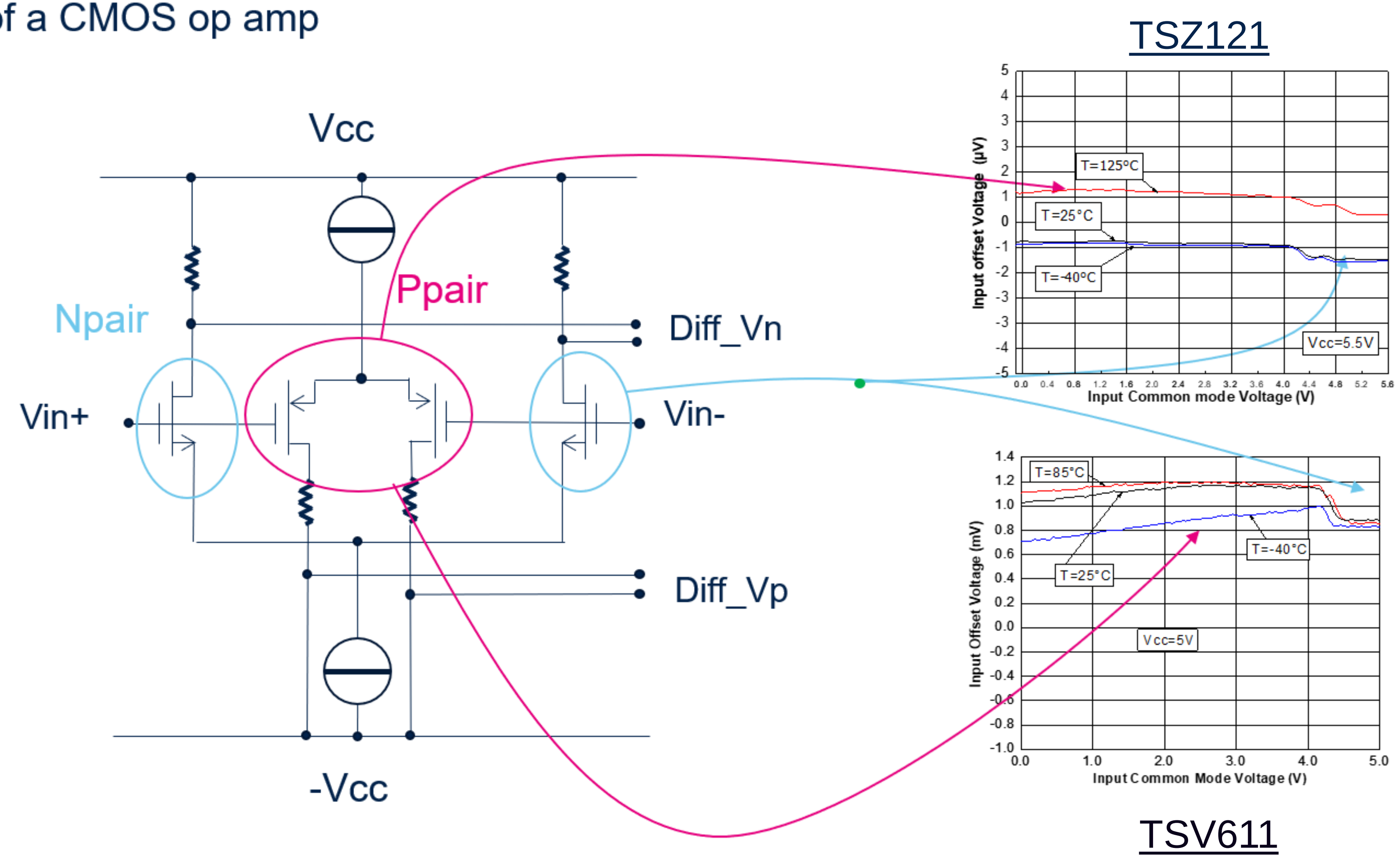
Common-mode rejection ratio

Input stage of a CMOS op amp



Explanation

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Common-mode-rejection ratio

EXPLANATION (page 25 to 27)

When we speak about precision, another important point must be taken into consideration: the CMRR

The common-mode rejection ratio is defined as the ratio of the differential voltage amplification to the common-mode voltage amplification. This is measured by determining the ratio of a change in input common-mode voltage to the resulting change in the input offset voltage. The common-mode input voltage affects the bias point of the input differential pair. Because of the inherent mismatches in the circuitry, changing the bias point changes the offset voltage, which in turn changes the output voltage.

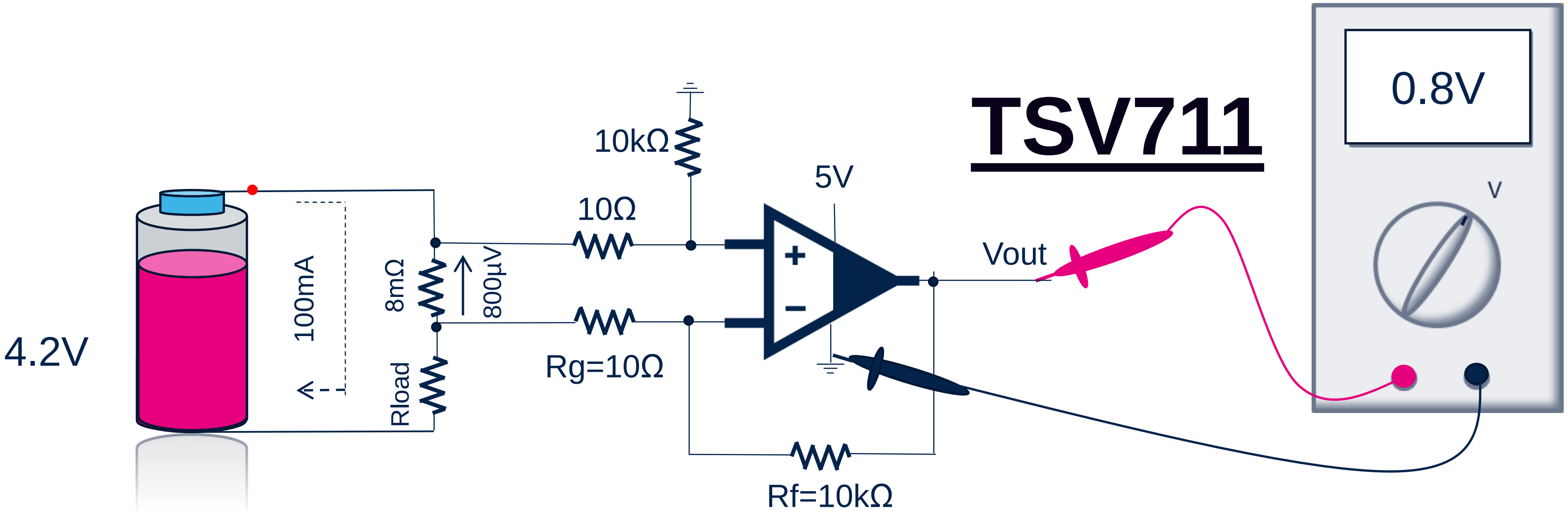
In general, a rail-to-rail op amp has parallel input stages made of a Ppair, which work on the low input common-mode voltage and a Npair, which work on the high input common-mode voltage.

(Page 25) As seen previously the mismatch between two NMOS or two PMOS is responsible for the Vio,
(Page 26) but there is no link between the mismatch of the NMOS and the mismatch of the PMOS.
(Page 27) This means that each pair will generate its own Vio. So, depending on the common-mode voltage used in the application the Vio might be different.

In a precision environment the main goal is to achieve the lowest Vio jump, when the signal switches from one pair to another.

Impact of CMRR on a battery monitoring
High-side current sensing

Theoritical



Explanation

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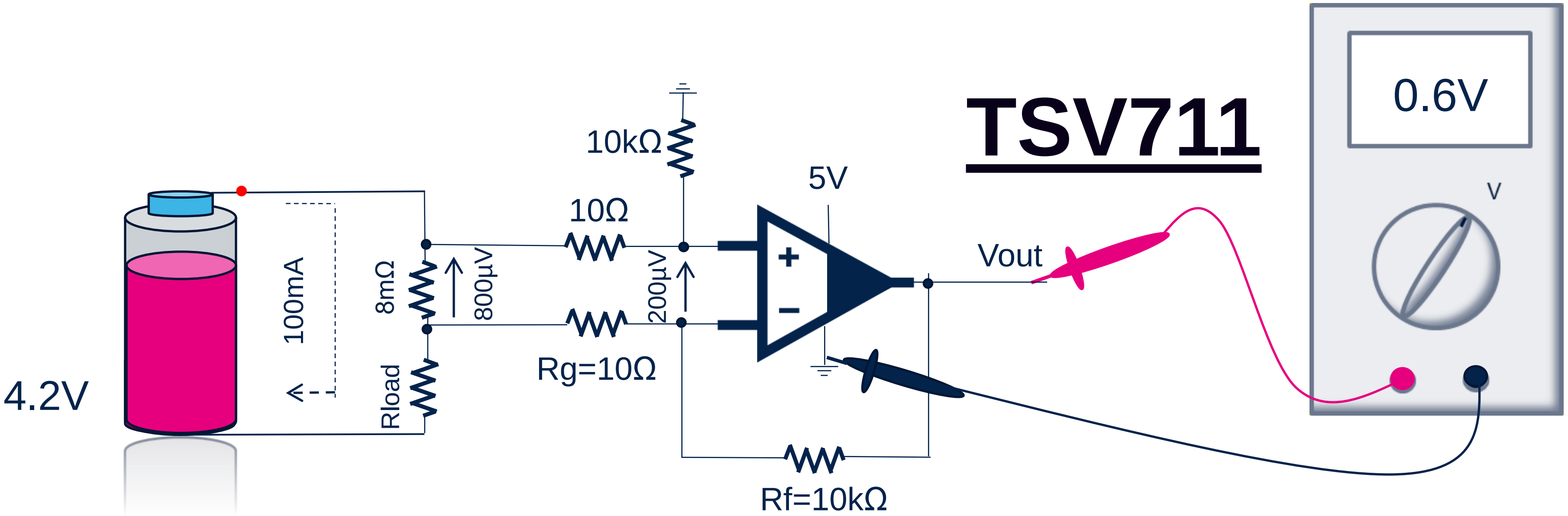
Impact of CMRR on a battery monitoring High-side current sensing

$$Vout = 0.8 - \left(1+\frac{Rf}{Rg}\right).\ Vio$$



TSV711	Impact on Vout	Error %
Vio	0.2V	25%

Explanation

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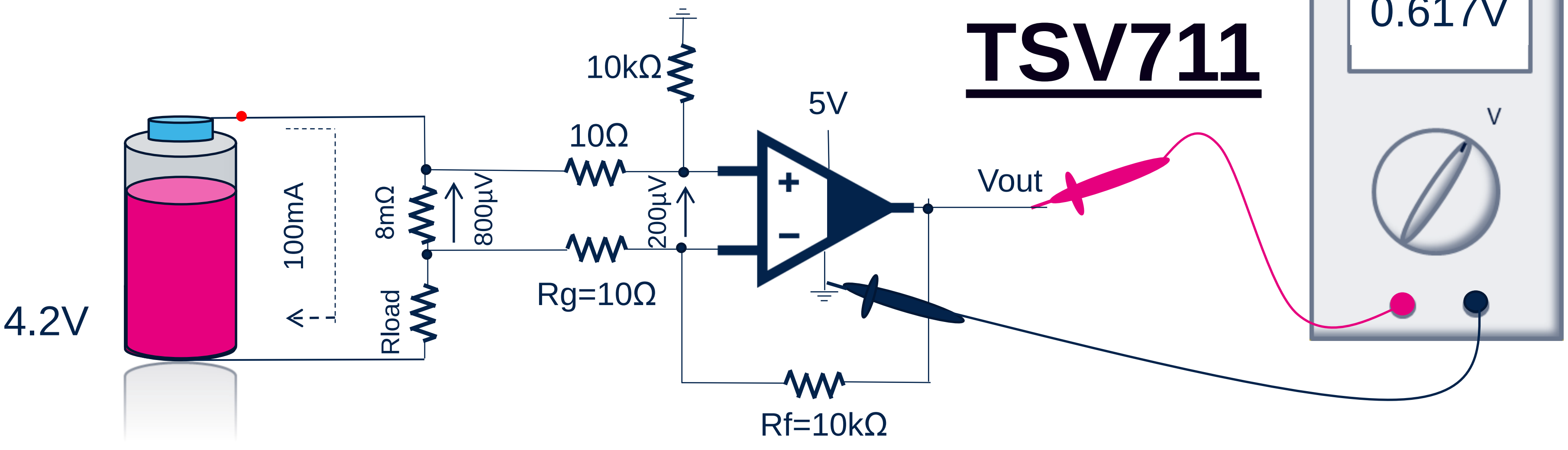
Impact of CMRR on a battery monitoring High-side current sensing

$$CMRRres = \frac{1+\frac{Rf}{Rg}}{4\varepsilon}$$

With ε=0.1% precision resistance
and a gain of 1000
The CMRR of the whole schematics is 250250 (108 dB).

$$Vout = 0.8 - \left(1+\frac{Rf}{Rg}\right).Vio \pm \frac{vbat}{CMRRres}\left(\frac{Rf}{Rg}\right)$$



TSV711	Impact on Vout	Error %
Vio	0.2V	25%
CMRRres @4.2V (108dB)	16.8mV	2.1%

Explanation

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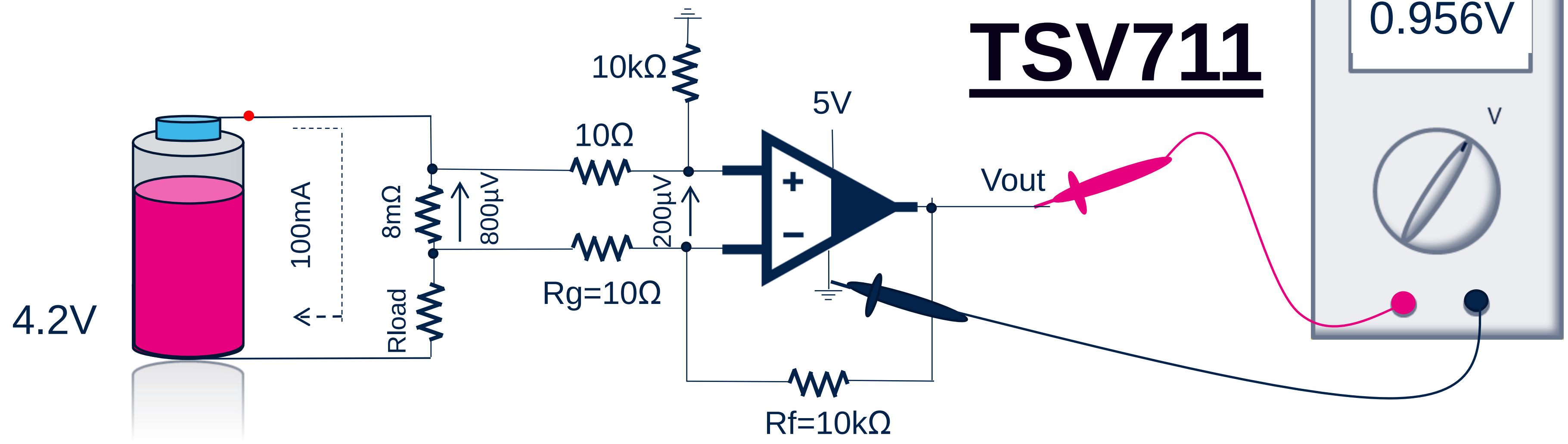
Impact of CMRR on a battery monitoring High-side current sensing

$$CMRRres = \frac{1+\frac{Rf}{Rg}}{4\varepsilon}$$

$$Vout = 0.8 - \left(1+\frac{Rf}{Rg}\right).Vio \pm \frac{vbat}{CMRRres}\left(\frac{Rf}{Rg}\right) \pm \frac{vicm - vcc/2}{CMRRop}\left(1+\frac{Rf}{Rg}\right)$$

With ε=0.1% precision resistance
and a gain of 1000
The CMRR of the whole schematics is 250250 (108 dB).



Explanation

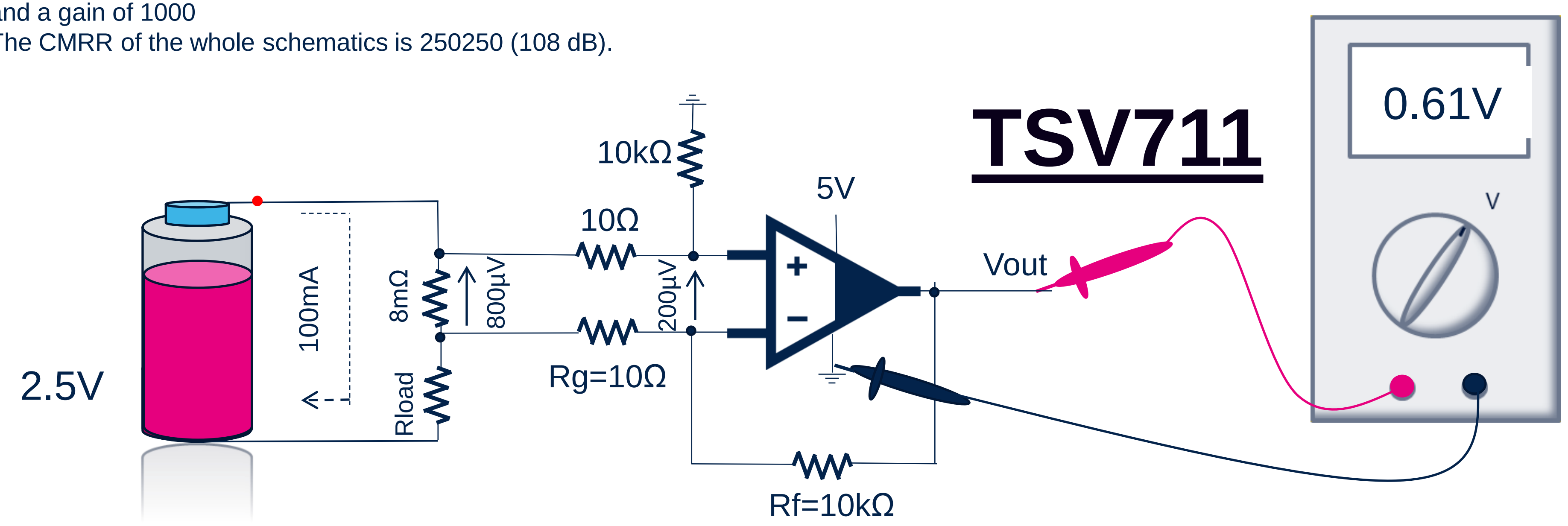
TSV711	Impact on Vout	Error %
Vio	0.2V	25%
CMRRres @4.2V (108dB)	16.8mV	2.1%
CMRRop @4.2V (74dB)	340mV	42.5%

Impact of CMRR on a battery monitoring High-side current sensing

$$CMRRres = \frac{1+\frac{Rf}{Rg}}{4\varepsilon}$$

$$Vout = 0.8 - \left(1+\frac{Rf}{Rg}\right).Vio \pm \frac{vbat}{CMRRres}\left(\frac{Rf}{Rg}\right) \pm \frac{vicm - vcc/2}{CMRRop}\left(1+\frac{Rf}{Rg}\right)$$

With ε=0.1% precision resistance
and a gain of 1000
The CMRR of the whole schematics is 250250 (108 dB).



Explanation

TSV711	Impact on Vout	Error %
Vio	0.2V	25%
CMRRres @4.2V (108dB)	16.8mV	2.1%
CMRRop @4.2V (74dB)	340mV	42.5%
CMRRres @2.5V (108dB)	10mV	1.2%
CMRRop @2.5V (74dB)	0mV	0%

Impact of CMRR on a battery monitoring High-side current sensing

EXPLANATION (page 29 to 33)

We can have an idea of the impact of CMRR on an op amp used in differential mode to sense a current through a shunt.

High-side current sensing is typically selected in applications where ground disturbance is not tolerated, and short circuit detection is required, such as battery current monitoring.

The application shows a single op amp used in a differential amplifier made of the **TSV711** op amp and four external resistors. It amplifies the small voltage drop across the sensing resistor of 8mohm *(milli ohms)* by the gain Rf/Rg, or 1000, while rejecting the common-mode input voltage.

(Page 29) First the battery is fully charged, and its voltage is at 4.2Volts. The Rload sources 100mA, so through the 8mohms shunt, a differential voltage of 800µV appears on the input of the amplification stage. If we consider a perfect world, the input should be amplified by the gain so on the output we should have 800mV, but despite a lot of effort we are still not in a perfect world and some error must be taken into account.

Continue

Impact of CMRR on a battery monitoring High-side current sensing

EXPLANATION (page 29 to 33)

(Page 30) As seen previously it is important to take the Vio into consideration. The **TSV711**, which is already a precision amplifier, shows a maximum Vio of 200µV (microvolts) at 25°C. So, on the output we will not see 800mV but 600mV. So, there is a 25% error due to the Vio.

(Page 31) The CMRR due to the mismatch of the resistance must be taken into account, and depending on the precision of the resistance the CMRRres might be predominant in the total output error.

The CMRR of the differential amplifier is given by this equation, and if we consider that the precision of the 4 resistors is 0.1% with a gain of 1000, we can obtain a CMRR of 108dB. So, when the battery is fully charged at 4.2V the CMRR due to the mismatch of the resistance will add an error on the output of roughly 17mV.

(Page 32) The **TSV711** has its own CMRR specified in the datasheet, which is 74 dB. In this case, the **TSV711**'s CMRR will be the predominant one and it will cause an error of 340mV on the output so an error of more than 42% compared to the theoretical value!

Continue

Impact of CMRR on a battery monitoring High-side current sensing

EXPLANATION (page 29 to 33)

(Page 33) When the battery discharges, the input common mode voltage will change, as the schematic shows high side current sensing.

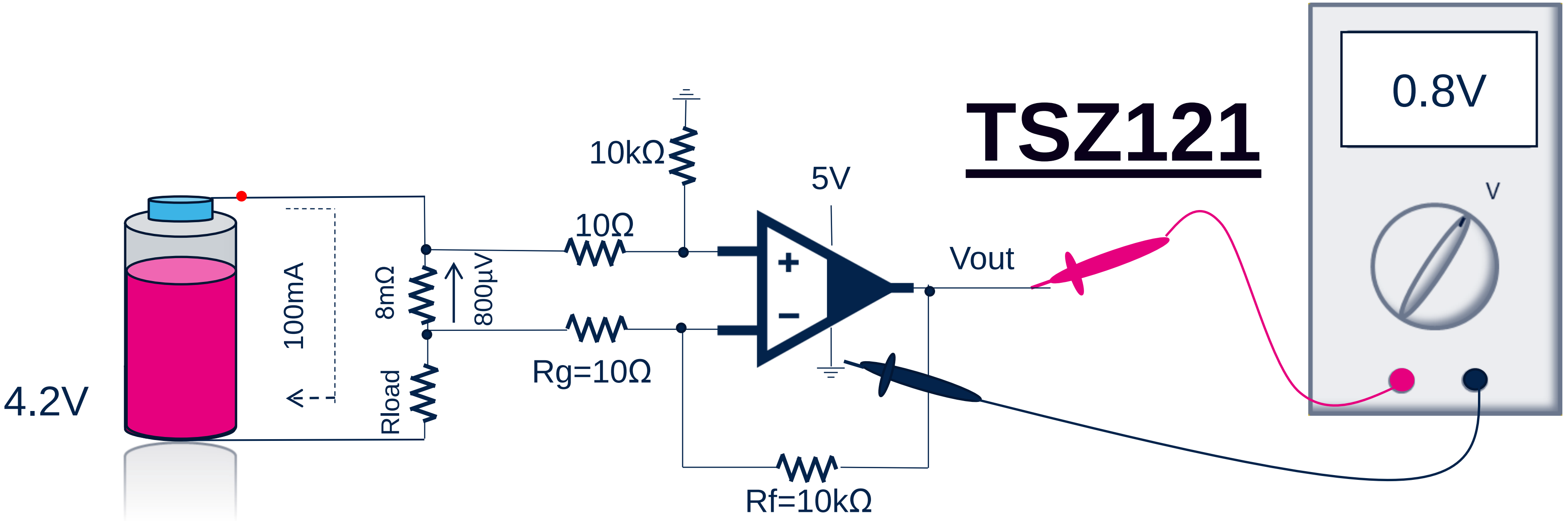
In the datasheet, the Vio is defined at Vcc/2, so in this case when the battery discharges, Vcc is equal to 2.5. this means that the **TSV711**'s CMRR will not have any impact on the output. Only the CMRR due to the mismatch of the resistance will play a role by adding an error of more than 1%.

The table summarizes the impact of the CMRR at different battery voltages when using the **TSV711**.

We can see that when the battery is fully charged the op amp's CMRR is the main contributor to the accuracy of the current measurement. We can also clearly understand that the precision of the measurements will change depending on the voltage level of the battery, due to the CMRR of the **TSV711** amplifier.

Impact of CMRR on a battery monitoring High-side current sensing

Theoritical



Explanation

TSV711	Impact on Vout	Error %
Vio	0.2V	25%
CMRRres @4.2V (108dB)	16.8mV	2.1%
CMRRop @4.2V (74dB)	340mV	42.5%
CMRRres @2.5V (108dB)	10mV	1.2%
CMRRop @2.5V (74dB)	0mV	0%

TSZ121	Impact on Vout	Error %

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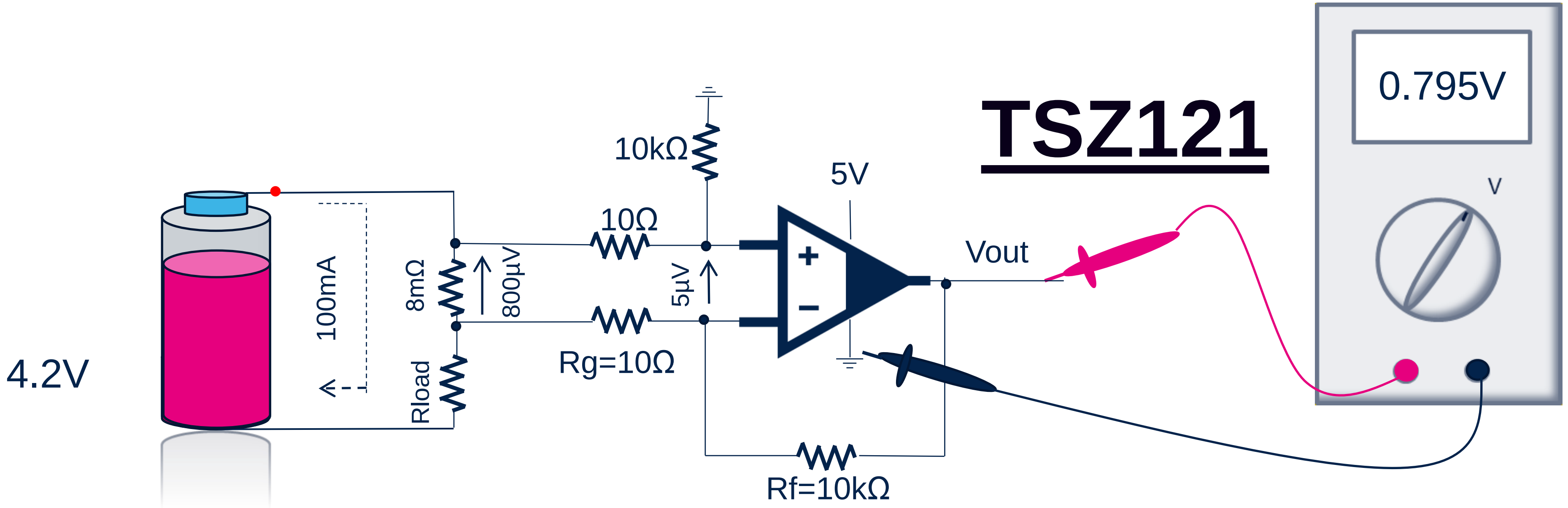
Impact of CMRR on a battery monitoring High-side current sensing

$$Vout=0.8-\left(1+\frac{Rf}{Rg}\right).\ Vio$$



Explanation

TSV711	Impact on Vout	Error %
Vio	0.2V	25%
CMRRres @4.2V (108dB)	16.8mV	2.1%
CMRRop @4.2V (74dB)	340mV	42.5%
CMRRres @2.5V (108dB)	10mV	1.2%
CMRRop @2.5V (74dB)	0mV	0%

TSZ121	Impact on Vout	Error %
Vio	0.005V	0.5%

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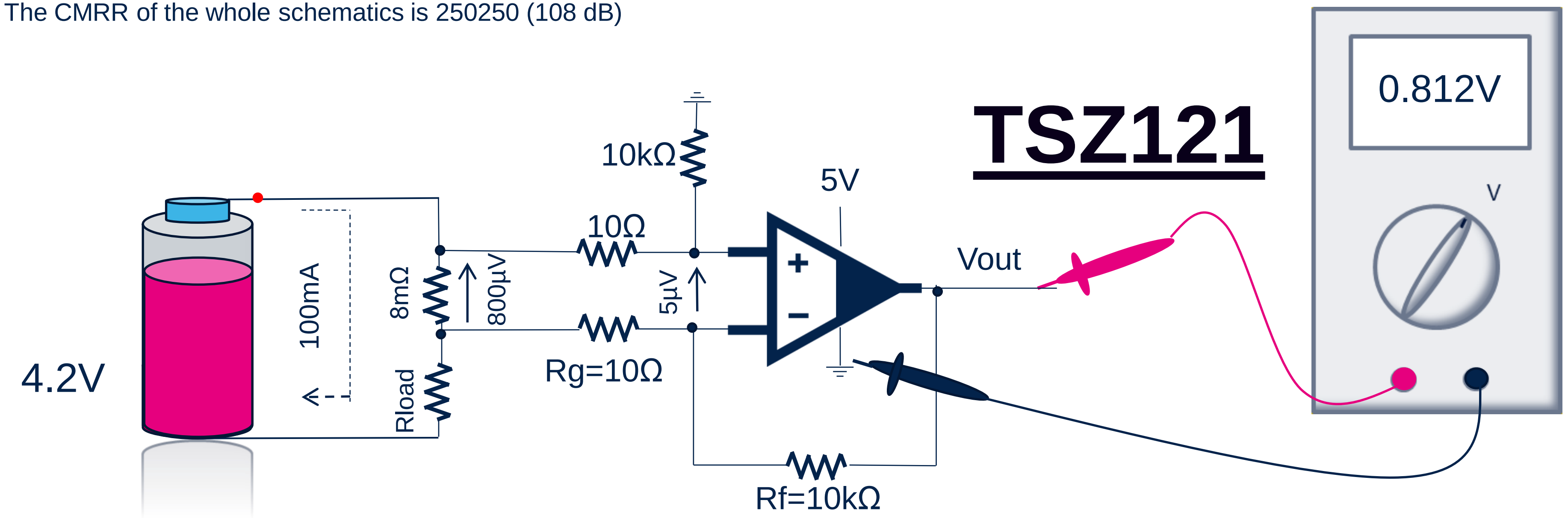
Impact of CMRR on a battery monitoring
High-side current sensing

$$CMRR_{res} = \frac{1+\frac{Rf}{Rg}}{4\varepsilon}$$

With ε=0.1% precision resistance
and a gain of 1000
The CMRR of the whole schematics is 250250 (108 dB)

$$Vout = 0.8 - \left(1+\frac{Rf}{Rg}\right).Vio \pm \frac{vbat}{CMRRres}\left(\frac{Rf}{Rg}\right)$$



Explanation

TSV711	Impact on Vout	Error %
Vio	0.2V	25%
CMRRres @4.2V (108dB)	16.8mV	2.1%
CMRRop @4.2V (74dB)	340mV	42.5%
CMRRres @2.5V (108dB)	10mV	1.2%
CMRRop @2.5V (74dB)	0mV	0%

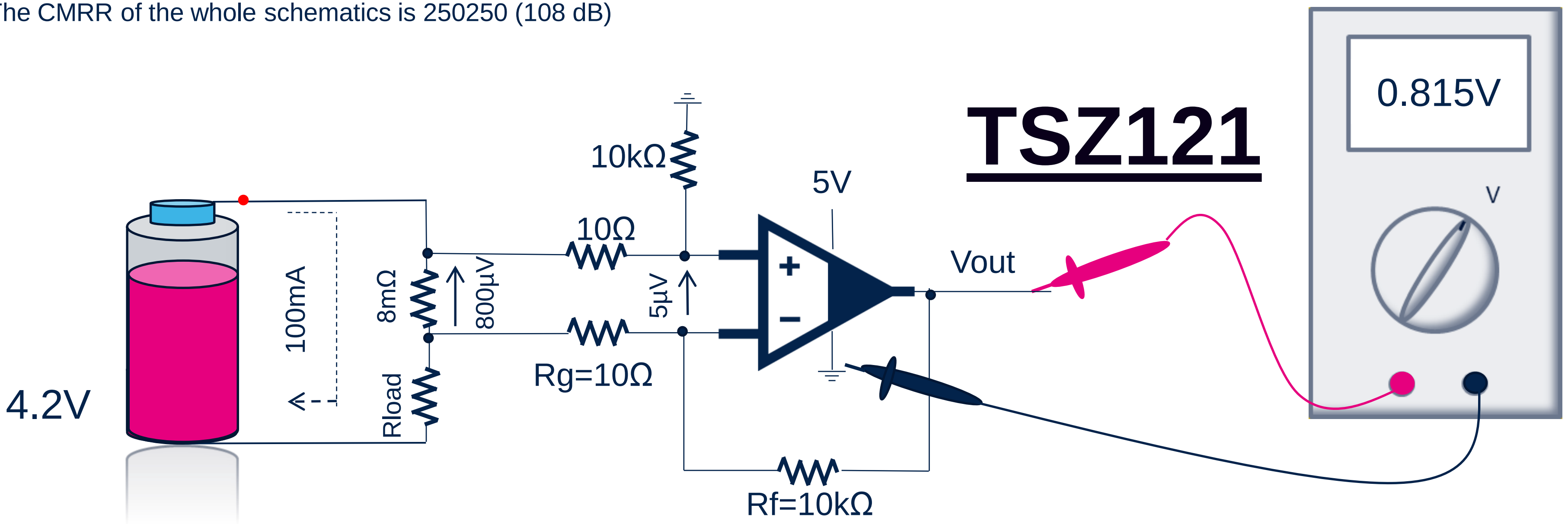
TSZ121	Impact on Vout	Error %
Vio	0.005V	0.5%
CMRRres @4.2V (108dB)	16.8mV	2.1%

Impact of CMRR on a battery monitoring
High-side current sensing

$$CMRR_{res} = \frac{1+\frac{Rf}{Rg}}{4\varepsilon}$$

$$Vout = 0.8 - (1+\frac{Rf}{Rg}).Vio \pm \frac{vbat}{CMRRres}(\frac{Rf}{Rg}) \pm \frac{vicm - vcc/2}{CMRRop}(1+\frac{Rf}{Rg})$$

With ε=0.1% precision resistance
and a gain of 1000
The CMRR of the whole schematics is 250250 (108 dB)



Explanation

TSV711	Impact on Vout	Error %
Vio	0.2V	25%
CMRRres @4.2V (108dB)	16.8mV	2.1%
CMRRop @4.2V (74dB)	340mV	42.5%
CMRRres @2.5V (108dB)	10mV	1.2%
CMRRop @2.5V (74dB)	0mV	0%

TSZ121	Impact on Vout	Error %
Vio	0.005V	0.5%
CMRRres @4.2V (108dB)	16.8mV	2.1%
CMRRop @4.2V (115dB)	3mV	0.4%

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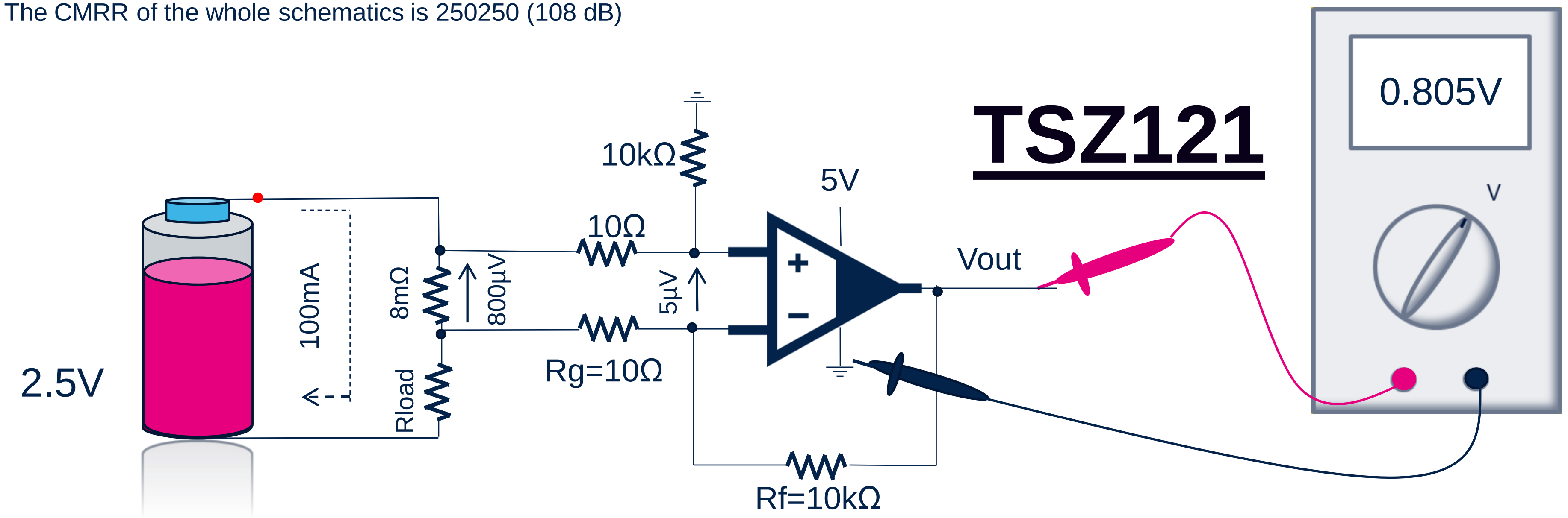
Impact of CMRR on a battery monitoring
High-side current sensing

$$CMRR_{res} = \frac{1+\frac{Rf}{Rg}}{4\varepsilon}$$

$$Vout = 0.8 - \left(1+\frac{Rf}{Rg}\right).Vio \pm \frac{vbat}{CMRRres}\left(\frac{Rf}{Rg}\right) \pm \frac{vicm - vcc/2}{CMRRop}\left(1+\frac{Rf}{Rg}\right)$$

With ε=0.1% precision resistance
and a gain of 1000
The CMRR of the whole schematics is 250250 (108 dB)



Explanation

TSV711	Impact on Vout	Error %
Vio	0.2V	25%
CMRRres @4.2V (108dB)	16.8mV	2.1%
CMRRop @4.2V (74dB)	340mV	42.5%
CMRRres @2.5V (108dB)	10mV	1.2%
CMRRop @2.5V (74dB)	0mV	0%

TSZ121	Impact on Vout	Error %
Vio	0.005V	0.5%
CMRRres @4.2V (108dB)	16.8mV	2.1%
CMRRop @4.2V (115dB)	3mV	0.4%
CMRRres @2.5V (108dB)	10mV	1.2%
CMRRop @2.5V (115dB)	0mV	0%

Impact of CMRR on a battery monitoring High-side current sensing

EXPLANATION (page 37 to 41)

Let's keep this application in mind but replace the **TSV711** by the **TSZ121** op amp.

(Page 37) The theoretical value of the output is still 800mV without any error introduced by the op amp itself.

(Page 38) the **TSZ121** has a maximum Vio of 5µV at 25°C so by considering this error and a gain of 1000 the output will be 795mV and therefore result in an error of 0.5%

(Page 39) As the resistors are still precise at 0.1%, the error on the output is still the same and add 2.1% of error.

Continue

Impact of CMRR on a battery monitoring High-side current sensing

EXPLANATION (page 37 to 41)

(Page 40) In this case, the **TSZ121** has a CMRR of 115 dB so its own impact on the total output error is not predominant. When Vbat equals 4.2V it represents only 3mV on the output, so just 0.4% of the error.

As already seen with the **TSV711** the main error was introduced by the CMRR of the op amp. In this case by choosing a more precise op amp, such as the **TSZ121**, which exhibits a very high CMRR of 115 dB, the main error is now introduced by the external resistors. With a VCC = 4.2V, the total error has been divided by 10 by using the **TSZ121**. Moreover, we can also see that with a very precise op amp the error on the output will be roughly the same even with a different Vbat, which was not the case when using the **TSV711**.

(Page 41) When the battery discharges, the input common-mode voltage will change, and in this case, the resistors are impacted the most, so the error is the same as previously with the **TSV711**.

V_{IO}, CMRR, PSRR and A_{VD}

$$Vid = Vio + \frac{\partial Vid}{\partial Vout}\Delta Vout + \frac{\partial Vid}{\partial Vicm}\Delta Vicm + \frac{\partial Vid}{\partial Vcc}\Delta Vcc + \frac{\partial Vid}{\partial T}\Delta T \quad (1)$$

1. Input Offset
2. AVD
3. CMRR
4. PSRR
5. Input Offset drift

$$We\ define: Avd = -20\log\left(\left|\frac{\partial Vid}{\partial Vout}\right|\right),\ CMRR = -20\log\left(\left|\frac{\partial Vid}{\partial Vicm}\right|\right),\ SVR = -20\log\left(\left|\frac{\partial Vid}{\partial Vcc}\right|\right)\ and\ DV_{io} = \left|\frac{\partial Vid}{\partial T}\right|$$

Explanation

V_{IO}, CMRR, PSRR, and A_{VD}

EXPLANATION (page 44)

The power supply rejection ratio PSRR and the differential voltage amplification AVD are also important parameters when making precision measurements, but the op amps PSRR will have a low impact if the power supplies are well decoupled. It is the same thing for the AVD, if the gain of the amplifier is not so high less than 1.000 this parameter will not create any issues.

In order to take into account, all the parameters likely to have an impact on precision we can use this equation. Where the first parameter expresses the input offset voltage, the second parameter the AVD, the third parameter expresses the CMRR, the fourth expresses the PSRR, and the last parameter expresses the input voltage drift int temperature, where:

1. the first parameter expresses the input offset voltage
2. the second parameter the **AVD**
3. the third parameter expresses the CMRR
4. the fourth expresses the PSRR
5. and the last parameter expresses the input voltage drift in temperature.

Continue

Op amps: Iib – Input bias current

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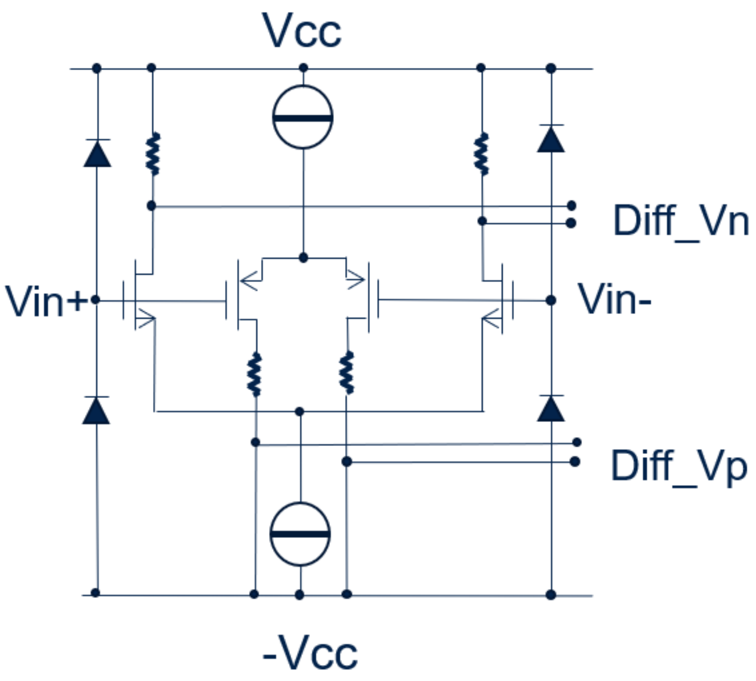
Input bias current

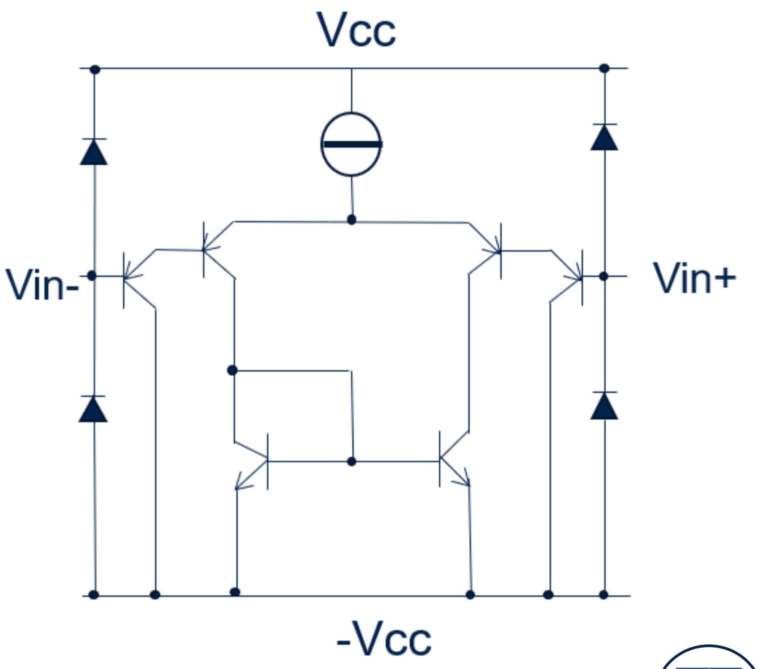
CMOS

No gate current only diode leakage



BIPOLAR

Current in/out (NPN/PNP) in the base



$$I_{ib} = \left| \frac{I_{ibn} + I_{ibp}}{2} \right|$$

Explanation

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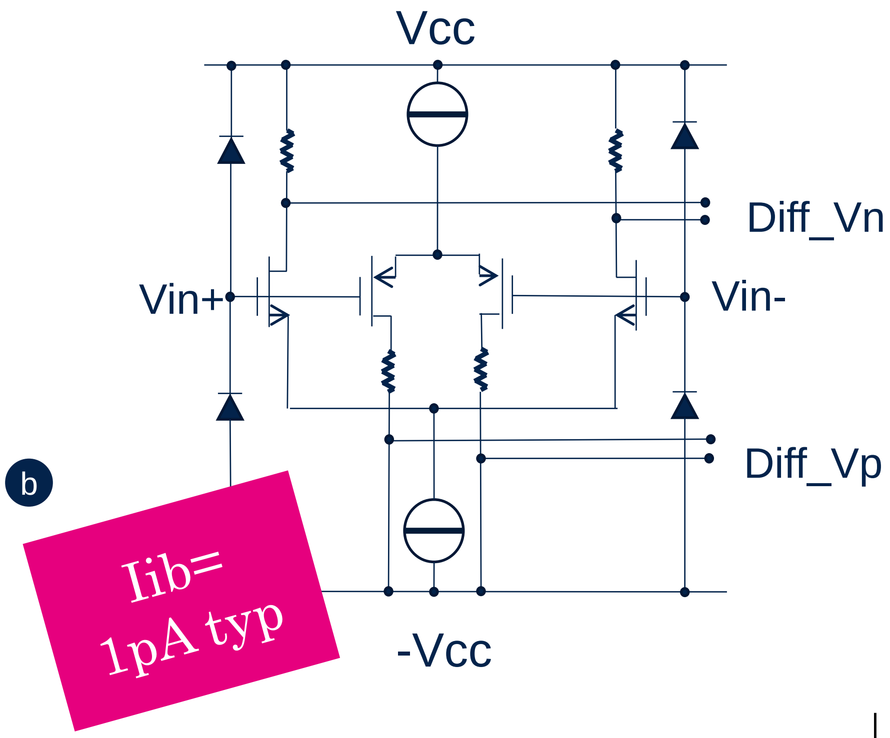
Input bias current

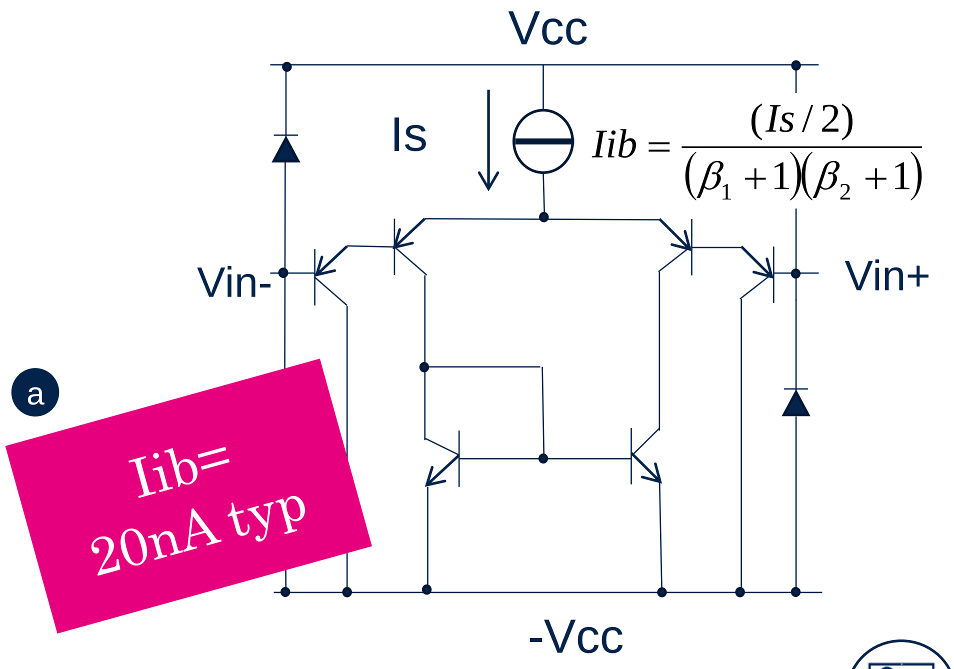
CMOS

No gate current only diode leakage



BIPOLAR

Current in/out (NPN/PNP) in the base



$$I_{ib} = \left|\frac{I_{ibn} + I_{ibp}}{2}\right|$$

Explanation

Input bias current

EXPLANATION (page 47 to 48)

(Page 47) In addition to errors in the voltage domain, that is, voltage offset (Vio) and input voltage noise density (en); current domain errors, such as the input current (Iin) are also important sources of error, especially for high source impedances (>100kΩ (*Kiloohm*)).The technology used for the op amp can significantly impact the whole precision of a system

The input bias current parameter, Iib, is defined as the average of the current into the two input terminals with the output at a specified level.

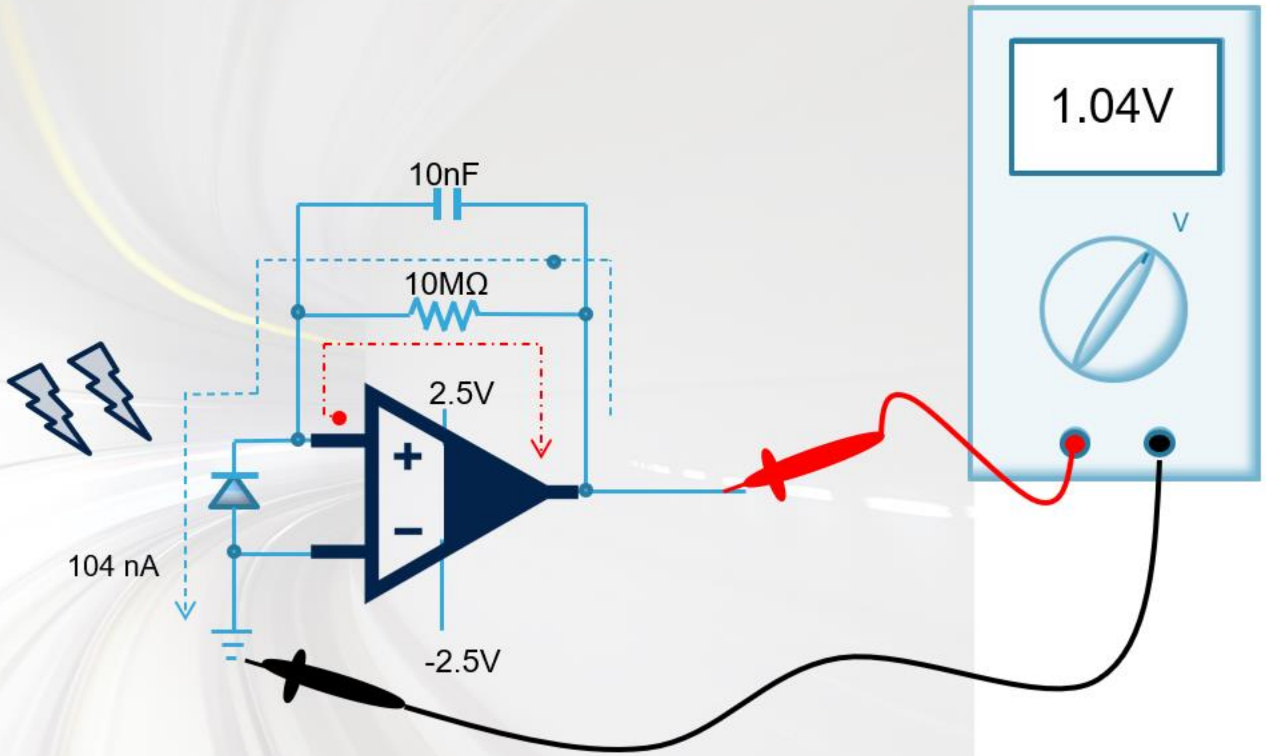
The input circuitry of all op amps requires a certain amount of bias current to operate properly.
The input bias is defined by the shown formula (at the bottom of the slide).
CMOS and JFET have a much lower input current than a standard bipolar.

(Page 48) Indeed, for a bipolar architecture, part of the current coming from the current source will flow in the input with a ratio of (a) $1/\beta^2$. (b) The CMOS transistor is driven by a gate, and there is an insignificant current inside. The small input current that can appear in a CMOS technology is mainly due to leakage of the ESD diode.

For sensors with small source impedances, voltage domain errors dominate while for higher source impedances, current domain errors dominate especially for bipolar op amp.

UV sensor application

UV source Index 4

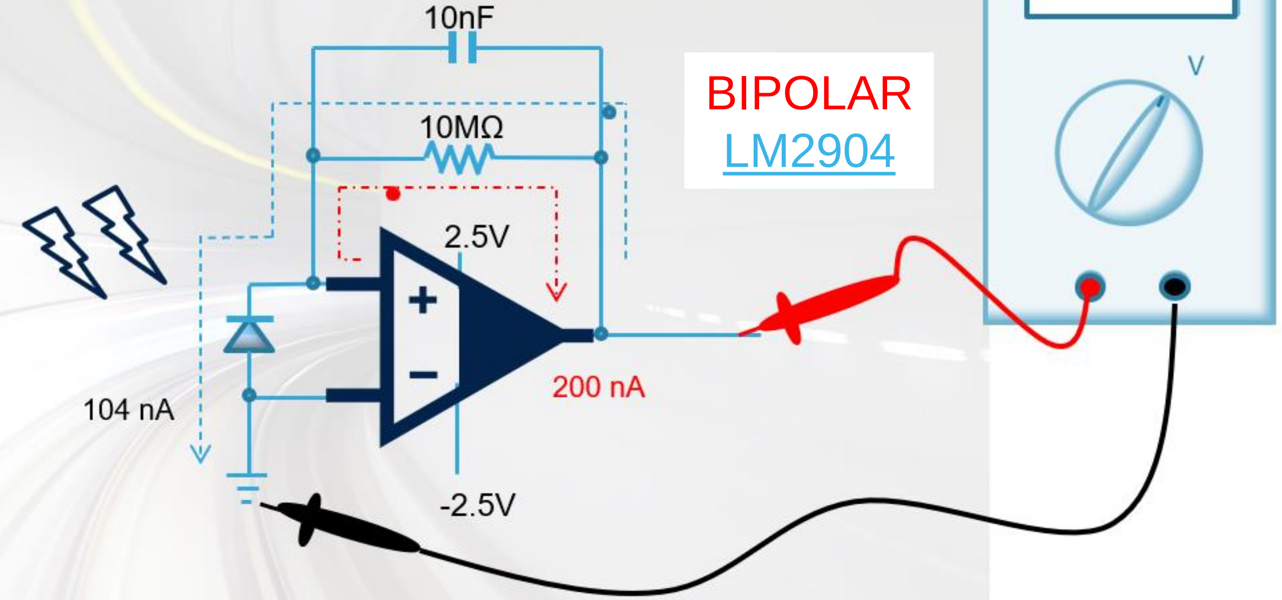


UV table translation for the UV sensor and Gain of 10M						
UV1	UV2	UV3	UV4	UV5	UV6	UV7
0.26 V	0.52 V	0.78 V	1.04 V	1.3 V	1.56 V	1.82 V

Explanation

UV sensor application

UV source Index 4



UV table translation for the UV sensor and Gain of 10M						
UV1	UV2	UV3	UV4	UV5	UV6	UV7
0.26 V	0.52 V	0.78 V	1.04 V	1.3 V	1.56 V	1.82 V

Explanation

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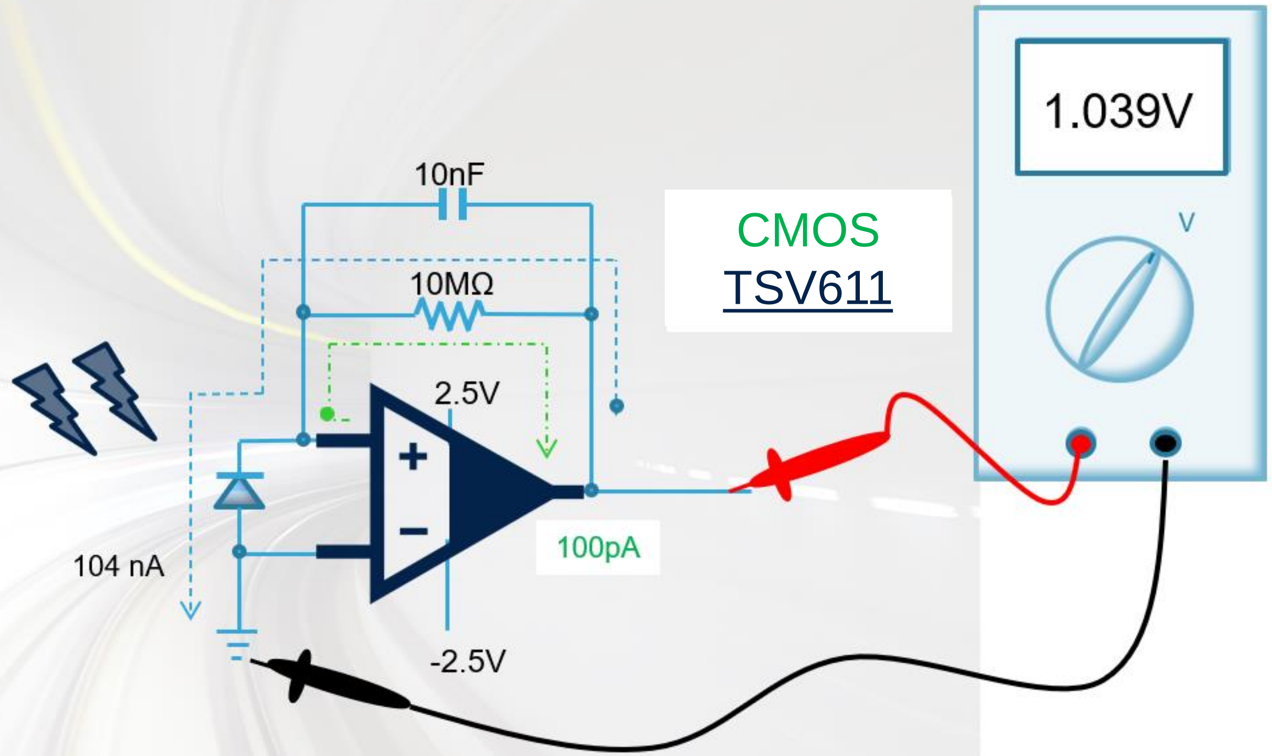
UV sensor application

UV source Index 4



UV table translation for the UV sensor and Gain of 10M						
UV1	UV2	UV3	UV4	UV5	UV6	UV7
0.26 V	0.52 V	0.78 V	1.04 V	1.3 V	1.56 V	1.82 V

Explanation

UV sensor application

EXPLANATION (page 50 to 52)

The input bias current, even if it represents a very small current, might affect the precision of a measurement, especially when we need to measure a low current using an op amp.

In this application we want to measure UV radiation using a UV sensor.
The UV sensor delivers a small current depending on the intensity of the UV-Source. A transimpedance circuit is used to convert the current delivered by the UV sensor thanks to the feedback resistance (10MΩ ($megaohms$)).The capacitance in the feedback helps to stabilize the system.

The UV source is set with an index of 4 and for this level of radiation, the UV sensor will generate a current of 104nA. This very small current will be amplified by the 10MΩ resistor resulting theoretically in an output voltage of 1.04V.

First, let's use a bipolar op amp such as the **LM2904**. This kind of op amp might have an input bias current up to 200nA, or twice the current we want to measure, so it will completely affect the output results.

The voltmeter will display -0.96V (volts) which corresponds to Index 1 in the UV sensor conversion table.

Continue

Input bias current

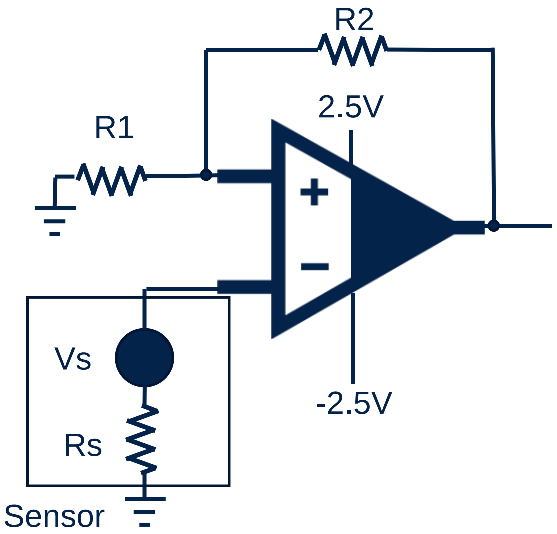
EXPLANATION (page 50 to 52)

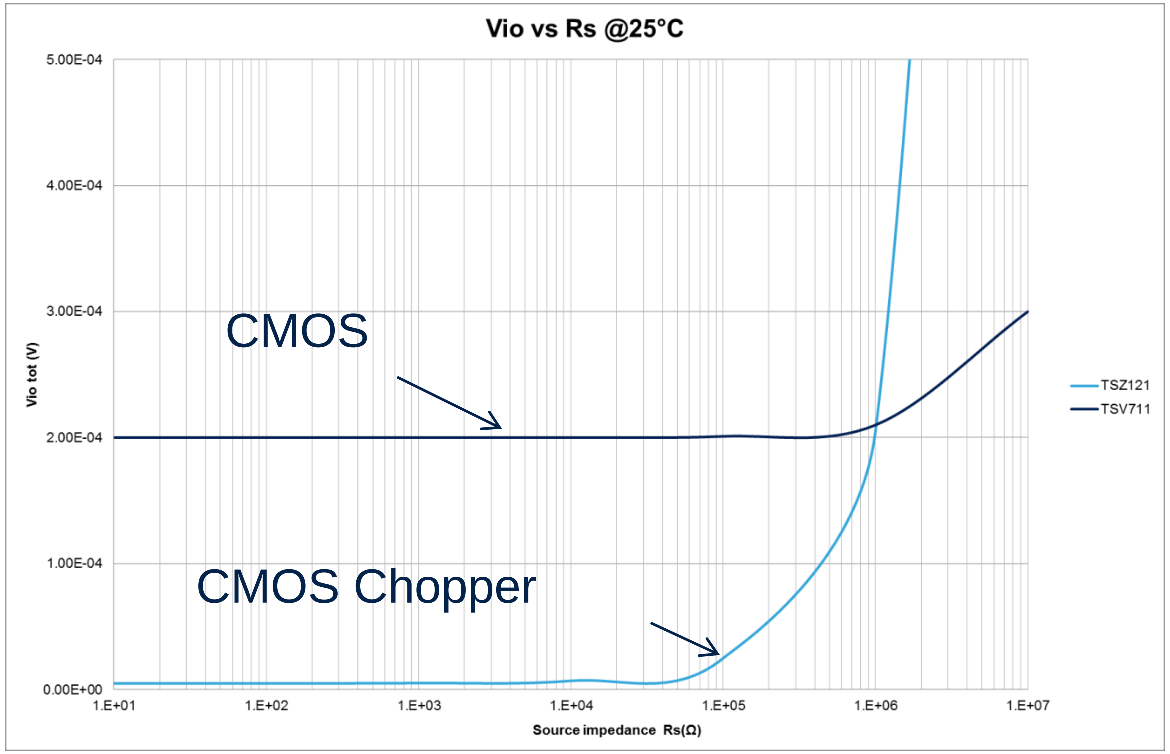
Now, we can keep the configuration as it is, and just replace the **LM2904** with a CMOS op amp such as the **TSV611**.

In this case, we can see that the output is close to the theoretical value and the level of output voltage corresponds to UV Index 4.

In case of an application where the current to be measured is extremely low or where the input impedance is very high, it is mandatory to use a CMOS op amp, so as not to affect the measurement.

Is the TSZ121 chopper always a good choice?





Vio tot= $Vio + Rs.In$ 1

Rs $> \frac{Vio}{In+}$ 2

Explanation

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Is the TSZ121 chopper always a good choice?

EXPLANATION (page 55)

In the precision domain, the **TSZ121** is generally the best candidate as it offers extremely good parameters due to its chopper architecture, but the input stage of chopper stabilized amplifiers does not behave like conventional amplifier input stages.

The **TSZ121** uses switches on the inputs that continually "chops" the input signal at 100 kHz to reduce input offset voltage down to 5µV. The dynamic behavior of these switches induces a charge injection current on the input terminals of the amplifier. The charge injection current has a DC path to ground through the resistors seen at the input terminals of the amplifier. Higher input impedance causes an apparent shift in the input bias current of the amplifier resulting in a higher input bias current than conventional CMOS op amps.

1 It is hard to find op amps that can be used across a wide range of source impedances, e.g. (10Ω to 10MΩ), and still achieve DC precision. A comparison of the state-of-the-art DC specifications of CMOS and precision chopper op amps is shown in the following graph. If these op amps are used to interface a sensor with certain source impedance, Rs, the resulting offset is given by equation 1

Continue

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Is the TSZ121 chopper always a good choice?

EXPLANATION (page 55)

2 Although the offset performance of a chopper op amp is better than the rest of the competition across a wide range of impedances, we can see that the offset performance starts to degrade rapidly when the source impedance, Rs, exceeds a threshold given by, equation 2.

We can clearly see that if the sensor used has an impedance higher than 1MΩ it is better to choose the **TSV711** rather than the **TSZ121**.

Op amps: Noise

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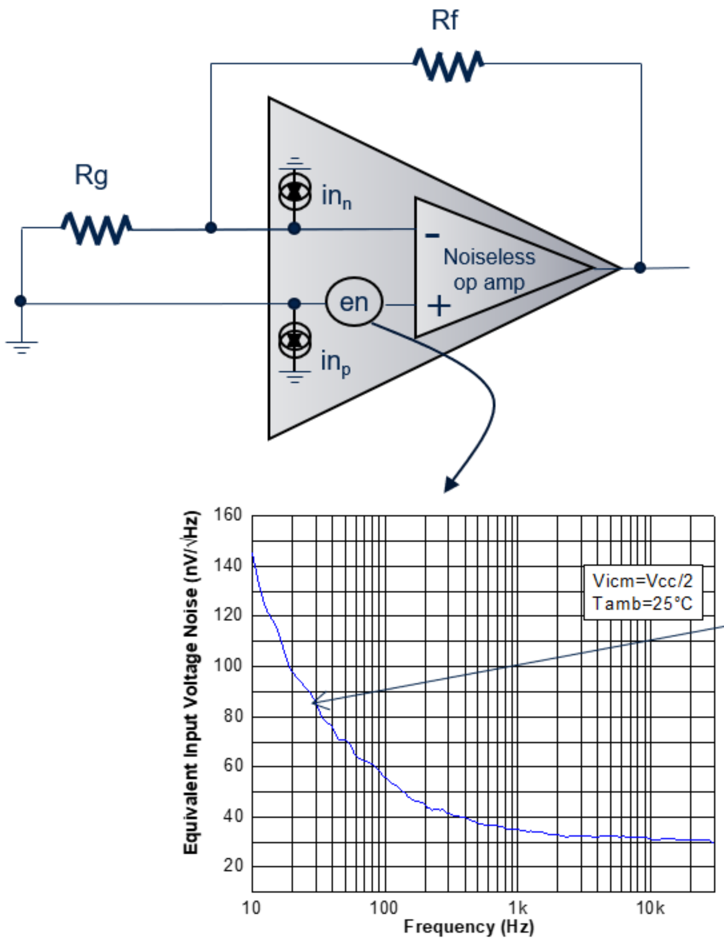
Noise sources of an op amp



$\frac{1}{f}$ noise (flicker noise)

$$e_{nf}(f) = \sqrt{\frac{enf\ (1Hz)}{f}}\ V/\sqrt{hz}$$

Explanation

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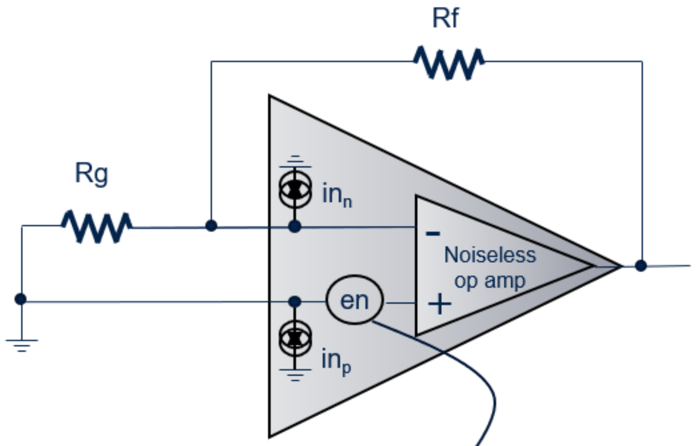
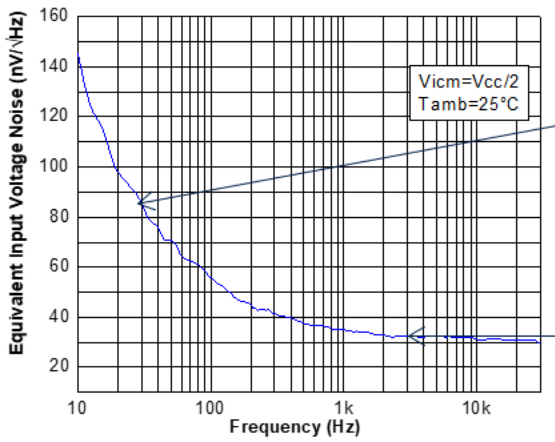
Noise sources of an op amp





$\frac{1}{f}$ noise (flicker noise)

$$e_{nf}(f) = \sqrt{\frac{enf\ (1Hz)}{f}}\ \text{V}/\sqrt{\text{hz}}$$

White noise

en V/√hz

Explanation

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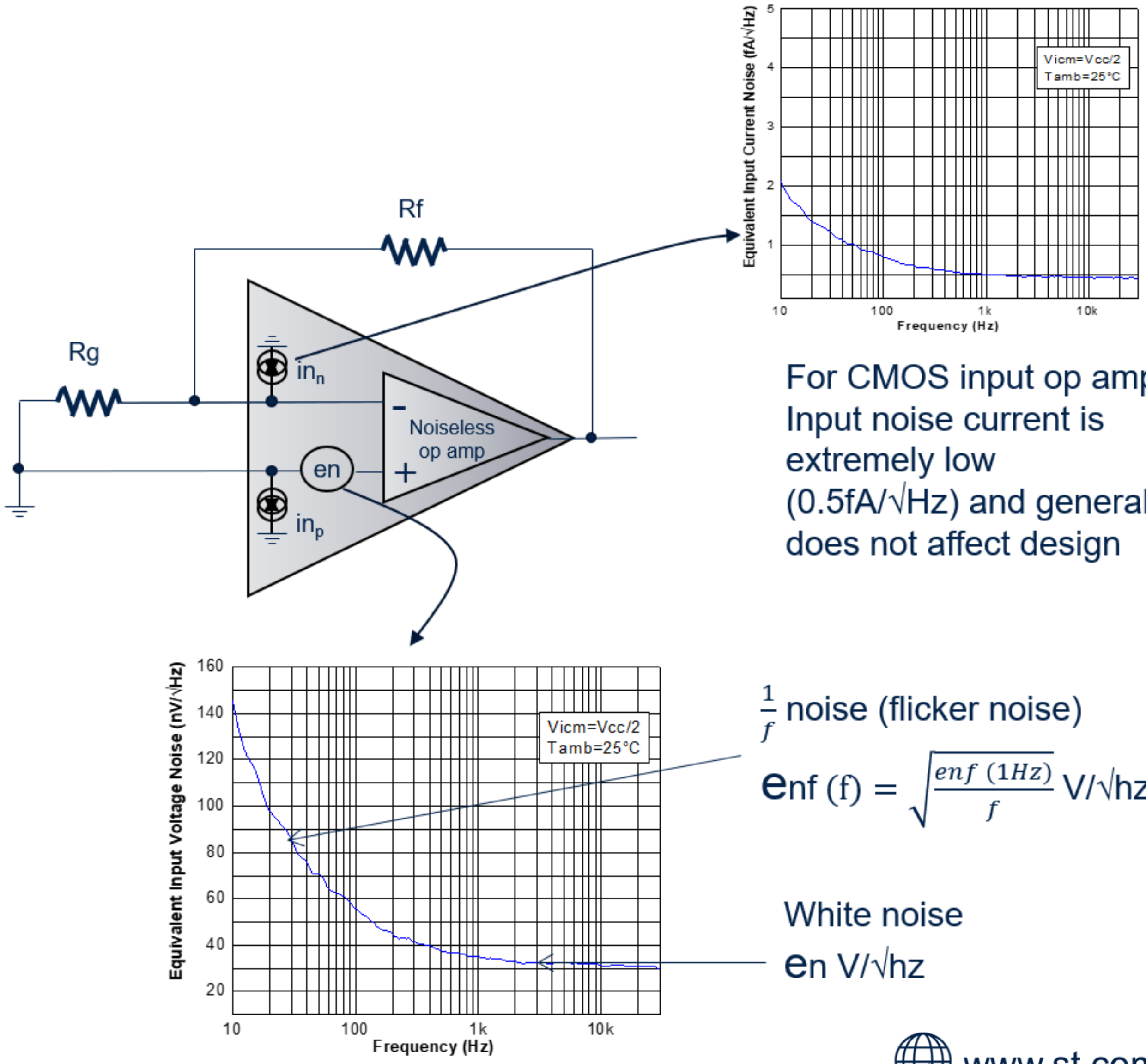
Noise sources of an op amp



For CMOS input op amps Input noise current is extremely low (0.5fA/√Hz) and generally does not affect design

$\frac{1}{f}$ noise (flicker noise)

$$en_f(f) = \sqrt{\frac{enf\ (1Hz)}{f}}\ \text{V}/\sqrt{\text{hz}}$$

White noise

en V/√hz

Explanation

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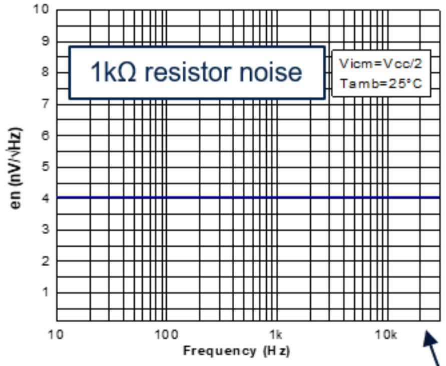
Noise sources of an op amp

There are 5 sources of noise



Resistors generate a white noise with a spectral density of:

$$e_n = \sqrt{4kTR} \quad VHz^{-\frac{1}{2}}$$

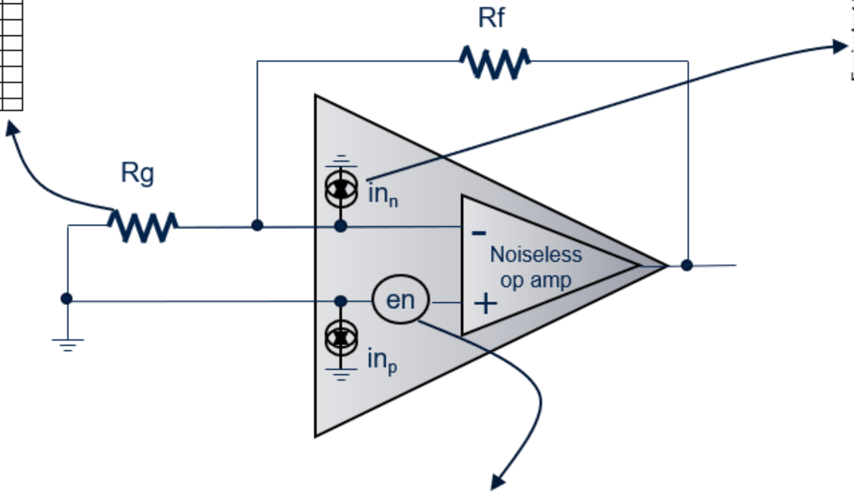
Where

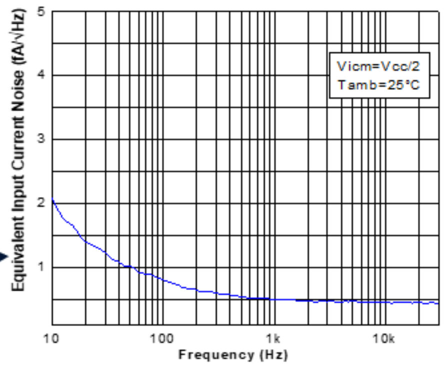
$$k = 1.38\,10^{-23} \quad JK^{-1}$$

(Boltzmann's constant)

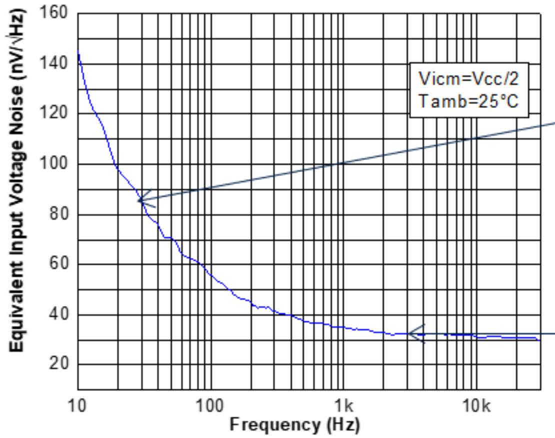
$$T = T(^\circ C) + 273.15$$

(Temperature in Kelvin)





For CMOS input op amps Input noise current is extremely low (0.5fA/√Hz) and generally does not affect design



$\frac{1}{f}$ noise (flicker noise)

$$enf\,(f) = \sqrt{\frac{enf\ (1Hz)}{f}}\ V/\sqrt{hz}$$

White noise

en V/√hz

Explanation

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Noise sources of an op amp

EXPLANATION (page 59 to 62)

Noise is also a key parameter in a precision environment.

It is part of life, and we have to deal with it. When an electronics component, even passive, is added to a system it will add noise that will impact the signal to noise ratio.

Noise is not easy to understand as it is nonperiodic, and it must be considered using statistics.
The easiest approach is to think of it being in the frequency domain even if engineers generally prefer the time domain.

All internal sources of noise contribute to the overall noise generated by the operational amplifier. The op amp noise is modeled with 3 noise sources. One source for the input noise voltage (en) and 2 sources for the input noise current (in). A current issued from a current noise source and flowing into a resistor generates voltage noise according to ohm law. All sources are physically independent; and therefore uncorrelated.

2 other sources of noise can be added due to the gain resistances, Rg and Rf.

Continue

Noise sources of an op amp

EXPLANATION (page 59 to 62)

These noise sources can be expressed:

- As a spectral density in nV/√Hz for voltage sources or pA /√Hz for current sources, which can be seen as the noise energy at a given frequency
- As an RMS value for a given bandwidth,

Let's have a look at each noise source.

(Page 59) the noise voltage source in the classical op amp architecture showing a combination of two different noise types. At low frequencies, generally lower than 500 Hz, flicker noise also called 1/f noise or pink noise appears.

1/f noise is caused by defects, at atomic level, in semiconductor devices. This noise is the main contributor at low frequency. It is generally expressed in nv/√(Hz)

Continue

Noise sources of an op amp

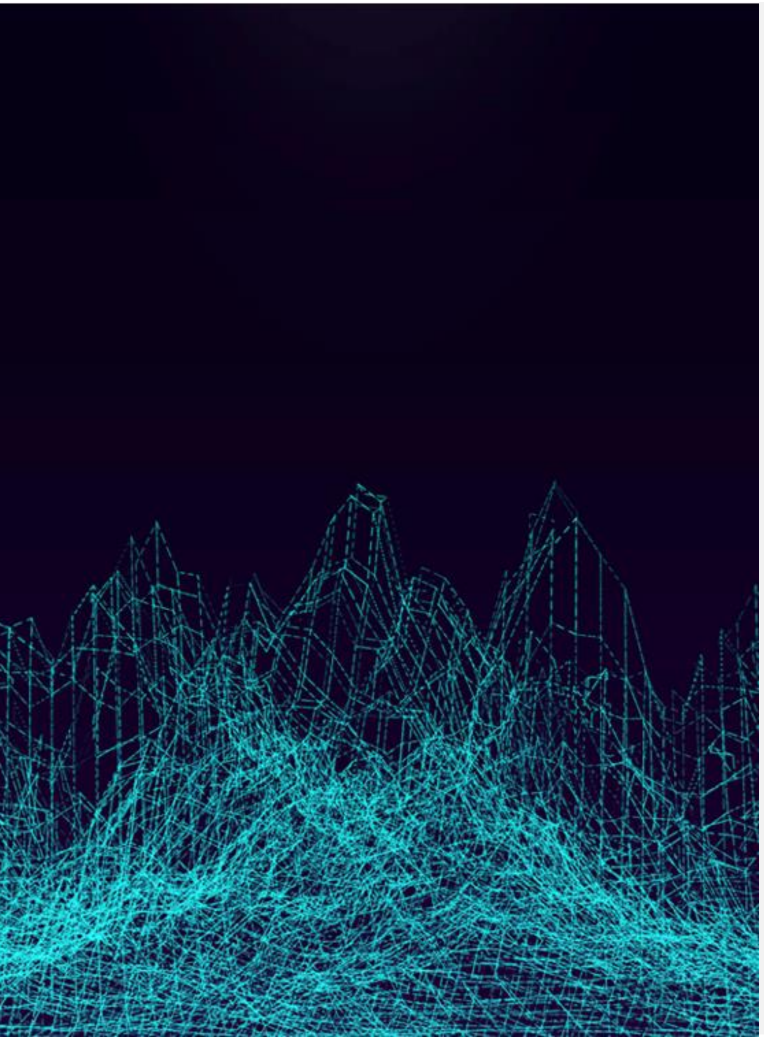
EXPLANATION (page 59 to 62)

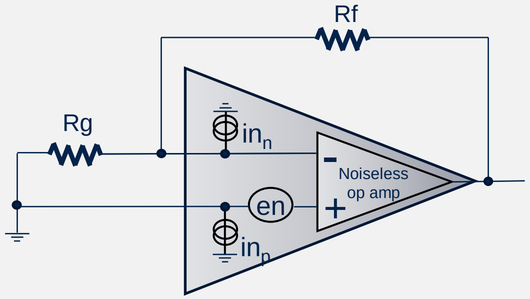
(Page 60) After the f corner frequency, the voltage noise source becomes white noise. It is a result of thermal agitation of the charges in an electric conductor, and it is also expressed in nv/√ (Hz). It is the main contributor of noise at higher frequencies. This is why, generally, in the datasheet the noise spectral density is provided at different frequencies.

(Page 61) The current noise source also adds its contribution to the overall noise, especially if the op amp is surrounded by high impedances, but the input noise current for CMOS input op amp is extremely small and generally does not affect the design as it is roughly 5fA/√Hz, but the input current noise for bipolar op amp or chopper architectures is in the range of 100pA/√ Hz.

(Page 62) Resistors will add also white noise according to the equation √ (4.k.T.R). We can see that the greater the resistance, the higher the noise.

Contribution of each source of noise





1

$$Vout_{rms} = \sqrt{\int_{fL}^{fH} \left[e_n^2 \left(1+\frac{R_f}{R_g}\right)^2 + R_f^2 I_{nn}^2 + 4kTR_g \left(\frac{R_f}{R_g}\right)^2 + 4kTR_f \right] df}$$

Noise source	Spectral density noise referred to the output	RMS noise value over a given bandwidth referred to the output
e_n	$en.(1+\frac{Rf}{Rg})$	$(1+\frac{Rf}{Rg}).\sqrt{en^2(FH-FL)}$ if white noise $(1+\frac{Rf}{Rg}).\sqrt{en^2(1Hz).ln(\frac{FH}{FL})}$ if 1/f noise
I_{nn}	$Inn.Rf$	$Rf.\sqrt{Inn^2(FH-FL)}$ if white noise
R_g	$\frac{Rf}{Rg}.\sqrt{4.k.T.Rg}$	$\frac{Rf}{Rg}.\sqrt{4.k.T.Rg.(FH-FL)}$
R_f	$\sqrt{4.k.T.Rf}$	$\sqrt{4.k.T.Rf.(FH-FL)}$

Contribution of each source of noise

EXPLANATION (page 66)

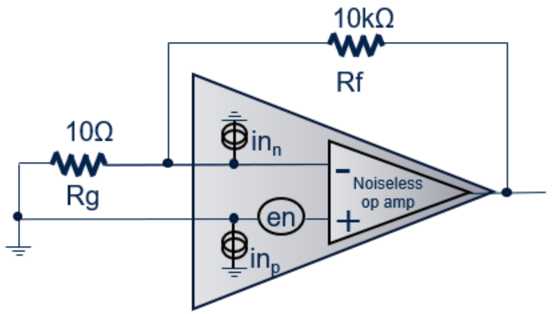
This table summarizes the different noise sources from a mathematical point of view.

The second column represents the noise of each source expressed in spectral noise density nV/√Hz. As it is a noise referred to the output, it is generally multiplied by the gain of the circuit.

The last column expresses the RMS value of the same noise. It is the integration of the spectral noise over the bandwidth of interest. 1 It is important to consider that the output noise is added in a quadratic sum. This equation expresses overall RMS noise on the output. In order to significantly reduce the noise level in an application, it is important to minimize the value of the resistors and reduce the bandwidth, as the wider the bandwidth the higher the RMS value.

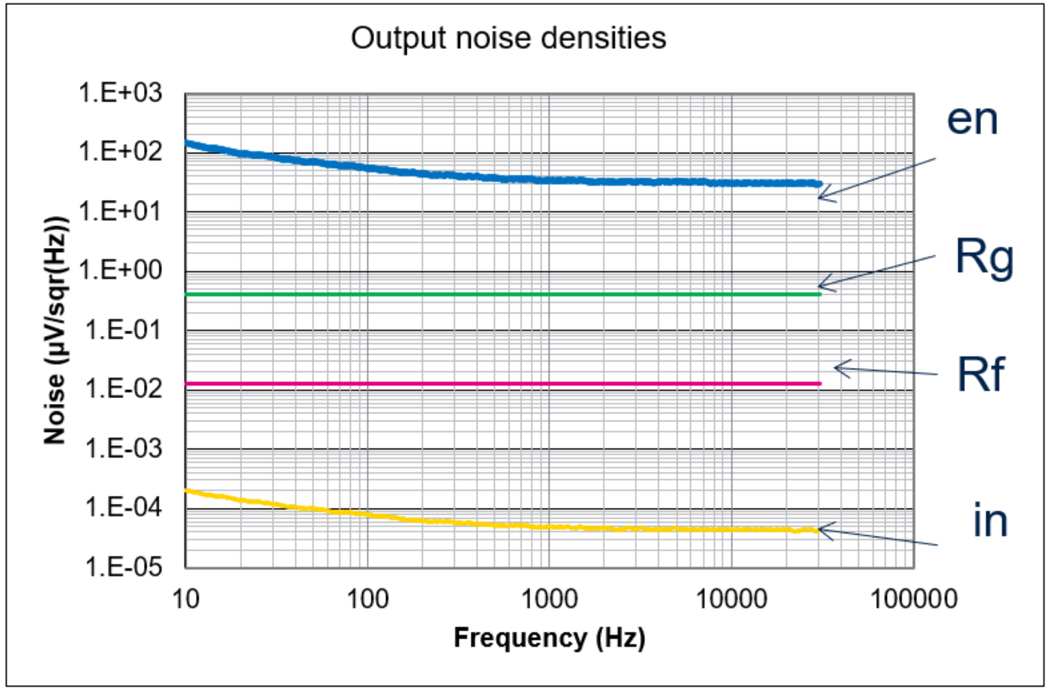
This can be easily done by inserting a capacitor in parallel with resistor RF. The cutoff frequency can be calculated to give minus -3dB at 5- or 10-times highest frequency to pass.

Impact of noise in an application



TSV731

Explanation

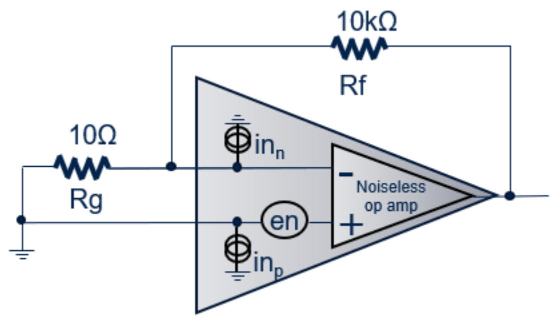


Noise voltage contribution to the output BW=30kHz

Noise Source		en Vrms
OPAMP	en	
	In	
THERMAL	Rf	
	Rg	

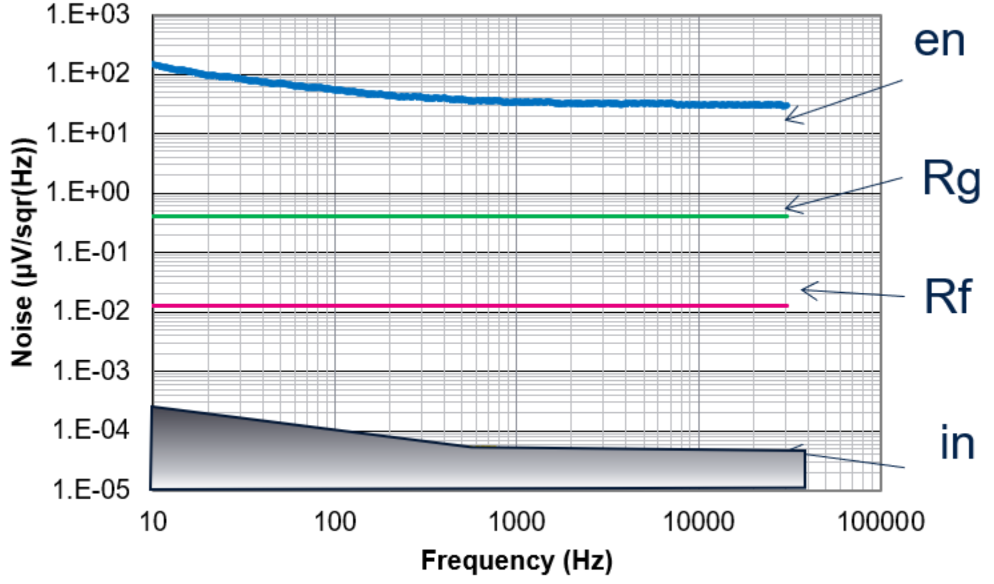
$$VoutRms = \sqrt{en^2 + In^2 + Rf^2 + Rg^2}$$

Impact of noise in an application



TSV731

Explanation

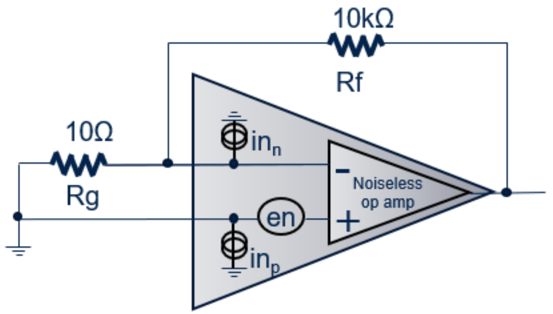


Noise voltage contribution to the output BW=30kHz

Noise Source		en Vrms
OPAMP	en	
	In	$8.66\ 10^{-9}$
THERMAL	Rf	
	Rg	

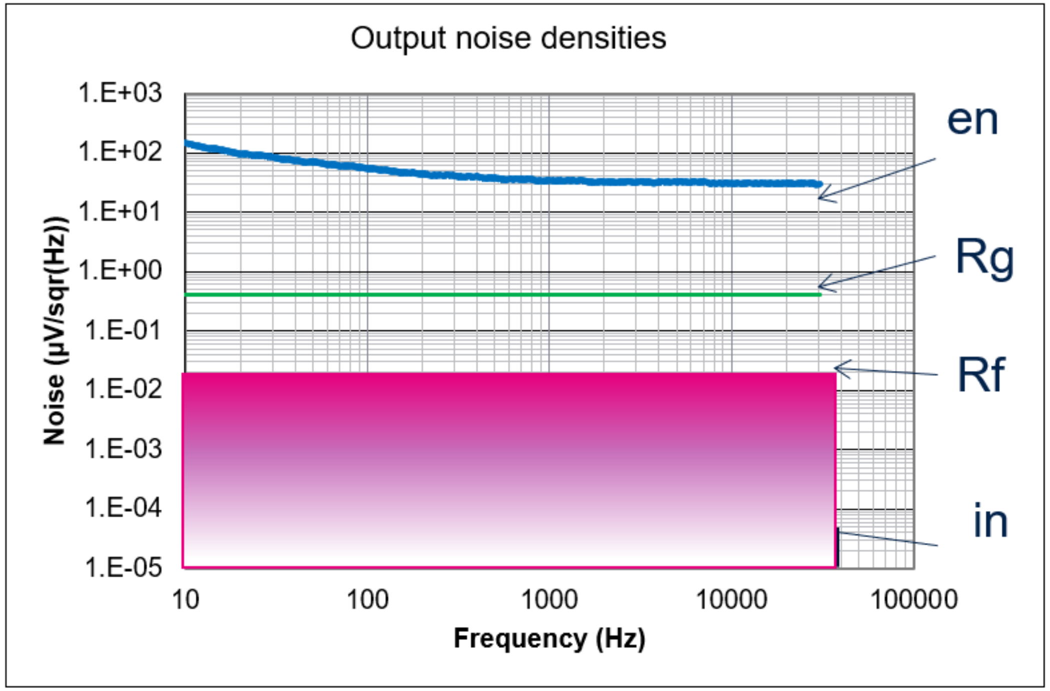
$$VoutRms = \sqrt{en^2 + In^2 + Rf^2 + Rg^2}$$

Impact of noise in an application



TSV731

Explanation



Noise voltage contribution to the output BW=30kHz

Noise Source		en Vrms
OPAMP	en	
	In	$8.66\ 10^{-9}$
THERMAL	Rf	$2.2\ \ 10^{-6}$
	Rg	

$$VoutRms=\sqrt{en^2+In^2+Rf^2+Rg^2}$$

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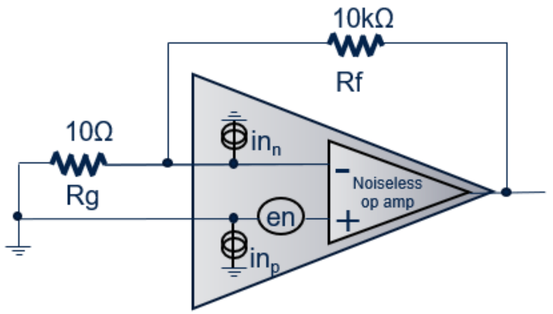
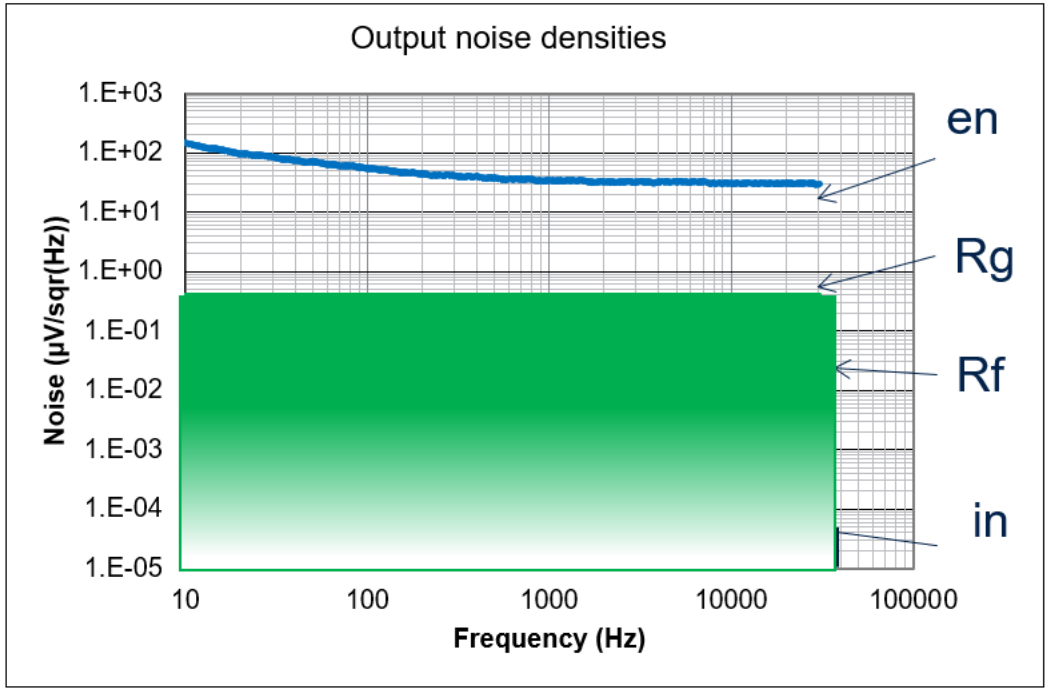
Impact of noise in an application



TSV731

Explanation



Noise voltage contribution to the output BW=30kHz

Noise Source		en Vrms
OPAMP	en	
	In	$8.66\ 10^{-9}$
THERMAL	Rf	$2.2\ 10^{-6}$
	Rg	$70.5\ 10^{-6}$

$$VoutRms = \sqrt{en^2 + In^2 + Rf^2 + Rg^2}$$

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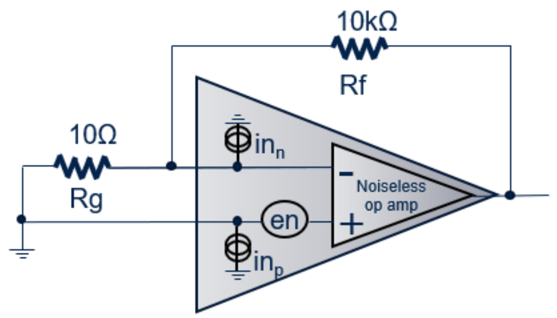
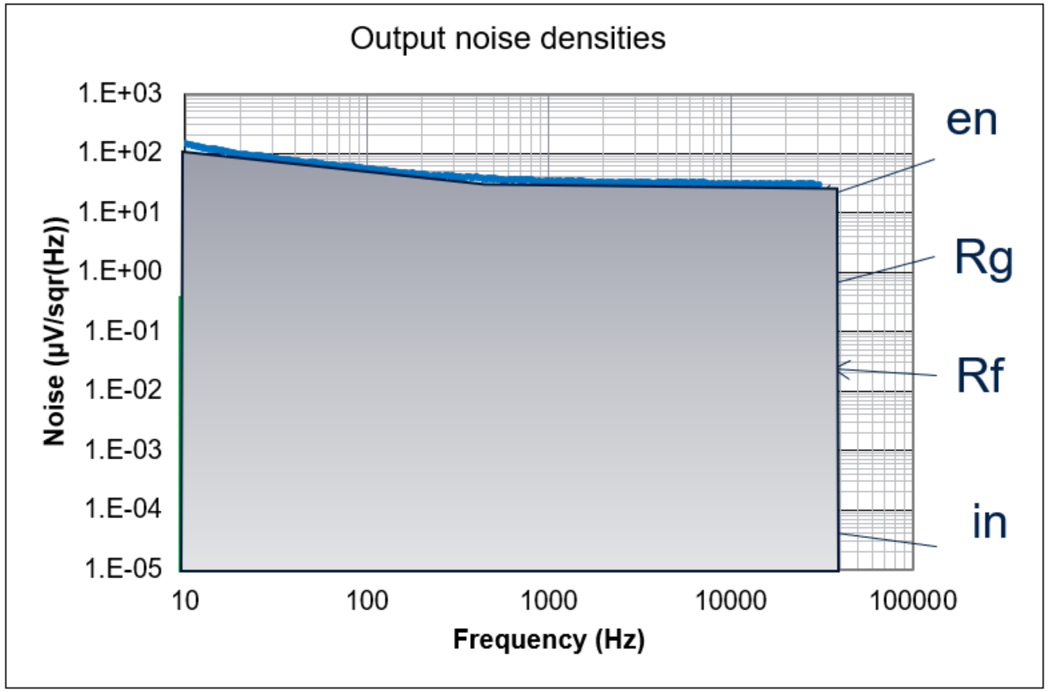
Impact of noise in an application



TSV731

Explanation



Noise voltage contribution to the output BW=30kHz

Noise Source		en Vrms
OPAMP	en	$5.37\ 10^{-3}$
	In	$8.66\ 10^{-9}$
THERMAL	Rf	$2.2\ 10^{-6}$
	Rg	$70.5\ 10^{-6}$

$$VoutRms = \sqrt{en^2 + In^2 + Rf^2 + Rg^2}$$

Impact of noise in an application

EXPLANATION (page 69 to 72)

In this example we can see the contribution of each source.

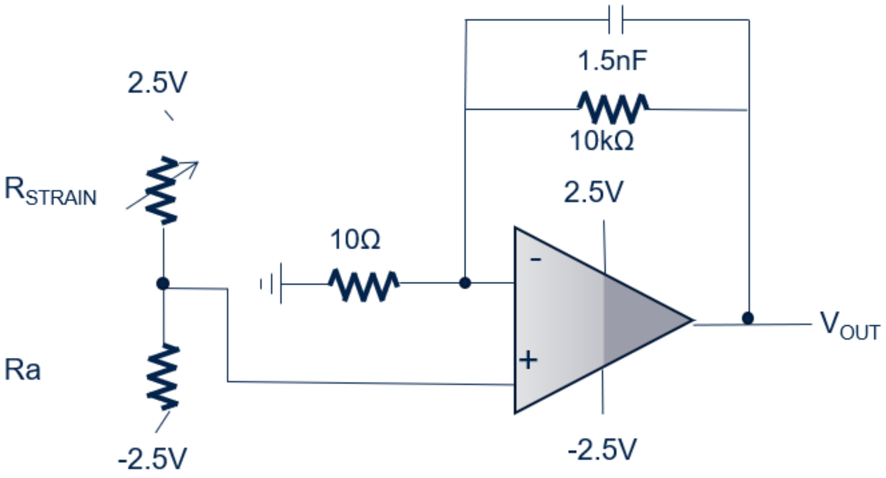
Here the noise is expressed in spectral density. To have a better understanding regarding the real impact it has on the output, let's transpose it to a Vrms value for a bandwidth of 30 kHz.

(Pag 69) This is the noise voltage contribution to the output of the current noise source
(Pag 70) This is the contribution of the Rf resistance
(Pag 71) This is the Rg resistance
(Pag 72) This is the equivalent voltage noise source of the op amp

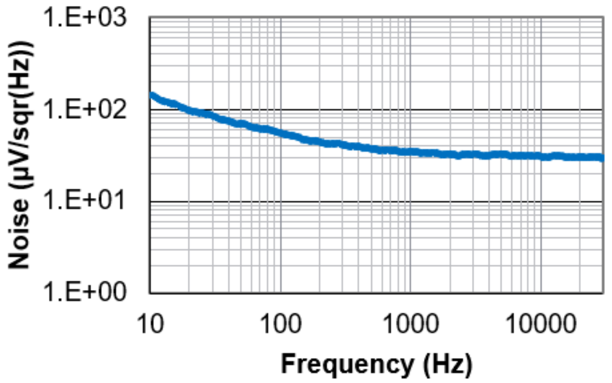
We can see that the impact of the current noise on the output is negligible compared to the other noise sources. The voltage noise source of the op amp represent the main part of the noise in this application.

We can see that the contribution of the voltage noise of the op amp is much higher than the other noise sources. However, if no care is taken regarding the value of the external resistor, their impact will become non negligible. To get the total error generated by the noise sources, you have to calculate the quadratic sum of each noise sources.
In this example, the overall output noise will be 5.4mVRms.

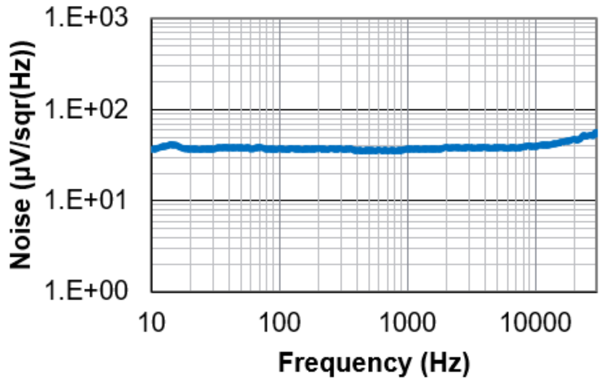
Impact of noise in an application



Output noise densities
TSV731

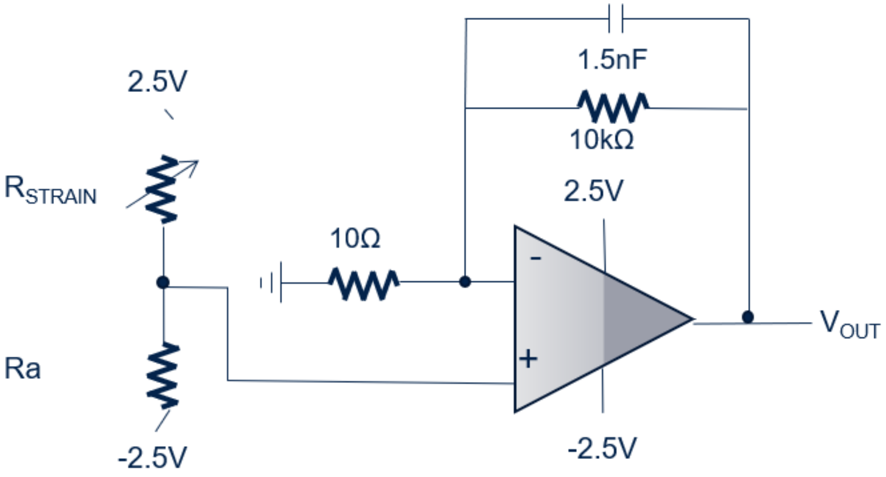


Output noise densities
TSZ121

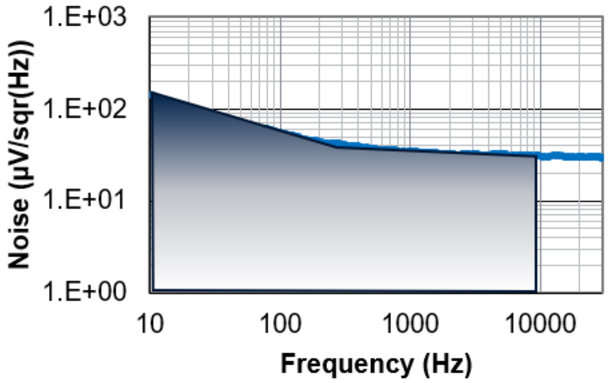


Explanation

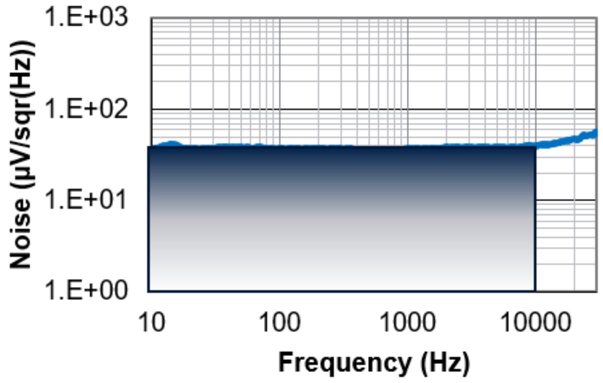
Impact of noise in an application



Output noise densities
TSV731



Output noise densities
TSZ121



Bandwidth 10kHz

en_Vout_{Rms}=3.1mVRms

en_Vout_{Rms}=3.7mVRms

Explanation

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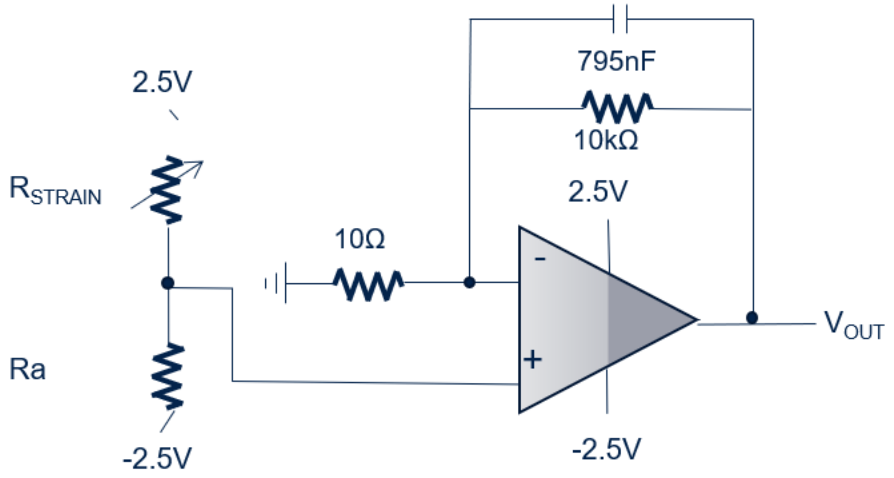
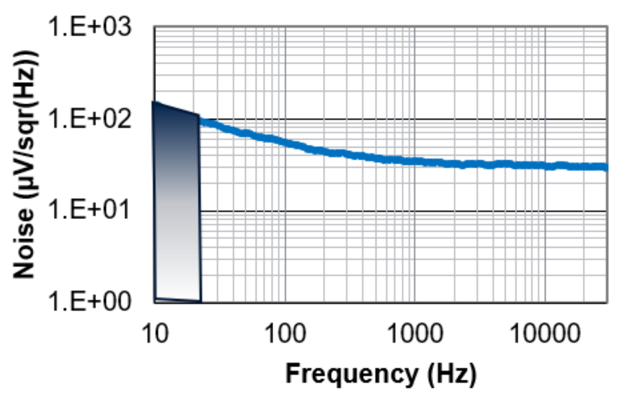
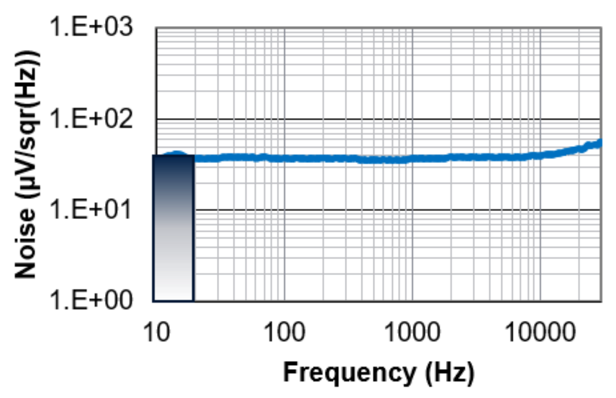
Impact of noise in an application



Output noise densities
TSV731



Output noise densities
TSZ121



Bandwidth 20Hz

en_Vout_{Rms}=0.69mVRms

en_Vout_{Rms}=0.16mVRms

Explanation

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Impact of noise in an application

EXPLANATION (page 74 to 76)

Here is an example of a half-bridge strain gauge.

The variation of the strain resistance will make a small variation of the input voltage. This small voltage variation must be detected with precision to give a correct value of the weight applied on the strain gauge.

Previously we have seen that the Vio is extremely important for this kind of measurement, but if the Vio introduces a large error, there is still a possibility to calibrate the system and completely remove the Vio, but the noise is a nonperiodic signal, and it cannot be calibrated. So, for an application requiring precision, it is also important to take noise into account.

In this case, the bandwidth is limited to 10 kHz thanks to the 1.5nF capacitor in the feedback.
Let's have a look at the impact of the noise with two different op amps: the **TSV731, which** is a high accuracy amplifier and the **TSZ121, which** is a precision chopper amplifier.

The calculation shows the noise impact on the output. For sure the Vio, and Dvio/DT must be taken into consideration regarding the whole error.

Continue

Impact of noise in an application

EXPLANATION (page 74 to 76)

(Page 75) If the bandwidth of the application is quite large there is no difference in term of noise error between the two op amps as they have the same white noise, around 35nv/√Hz.

The strain gauge is a DC application, so it is important to limit the bandwidth of the application in order to reduce the noise amplitude.

(Page 76) Here the bandwidth is limited to 20 Hz thanks to the 795nF capacitor, which was added in the feedback. In this case we can see a significant difference between the two devices. Effectively, the **TSV731** shows a 1/f noise that the **TSZ121** does not have (due to its chopper architecture). With the **TSZ121**, the noise impact on the output, is reduced by 4.

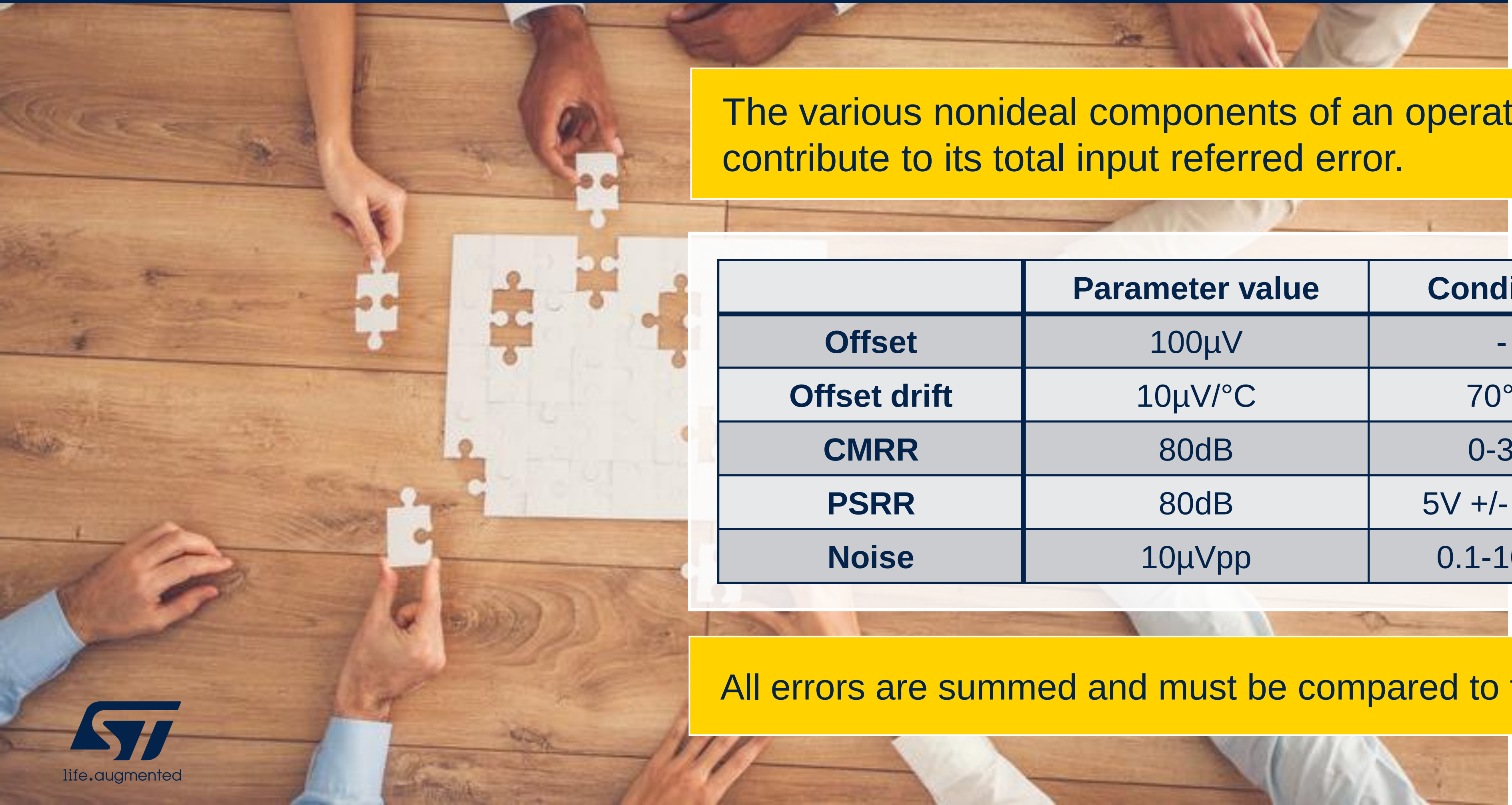
The noise can also be expressed in Vpp and in case of the **TSZ121** with a bandwidth limited to 20Hz the noise will be around 960µVpp. (Micro VPP).

It is a first order noise calculation, since we must consider a stop bandwidth with very a sharp edge. This means in the worst case roughly 600µV (micro volts) of error might be added to the output value caused by the noise.

Summary of errors impacting precision

Input referred error:

Explanation



The various nonideal components of an operational amplifier all contribute to its total input referred error.

	Parameter value	Condition	Real Value
Offset	100µV	-	100µV
Offset drift	10µV/°C	70°C	700µV
CMRR	80dB	0-3V	300µV
PSRR	80dB	5V +/- 10%	50µV
Noise	10µVpp	0.1-10Hz	10µVpp

All errors are summed and must be compared to the input signal.

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Summary of errors impacting precision

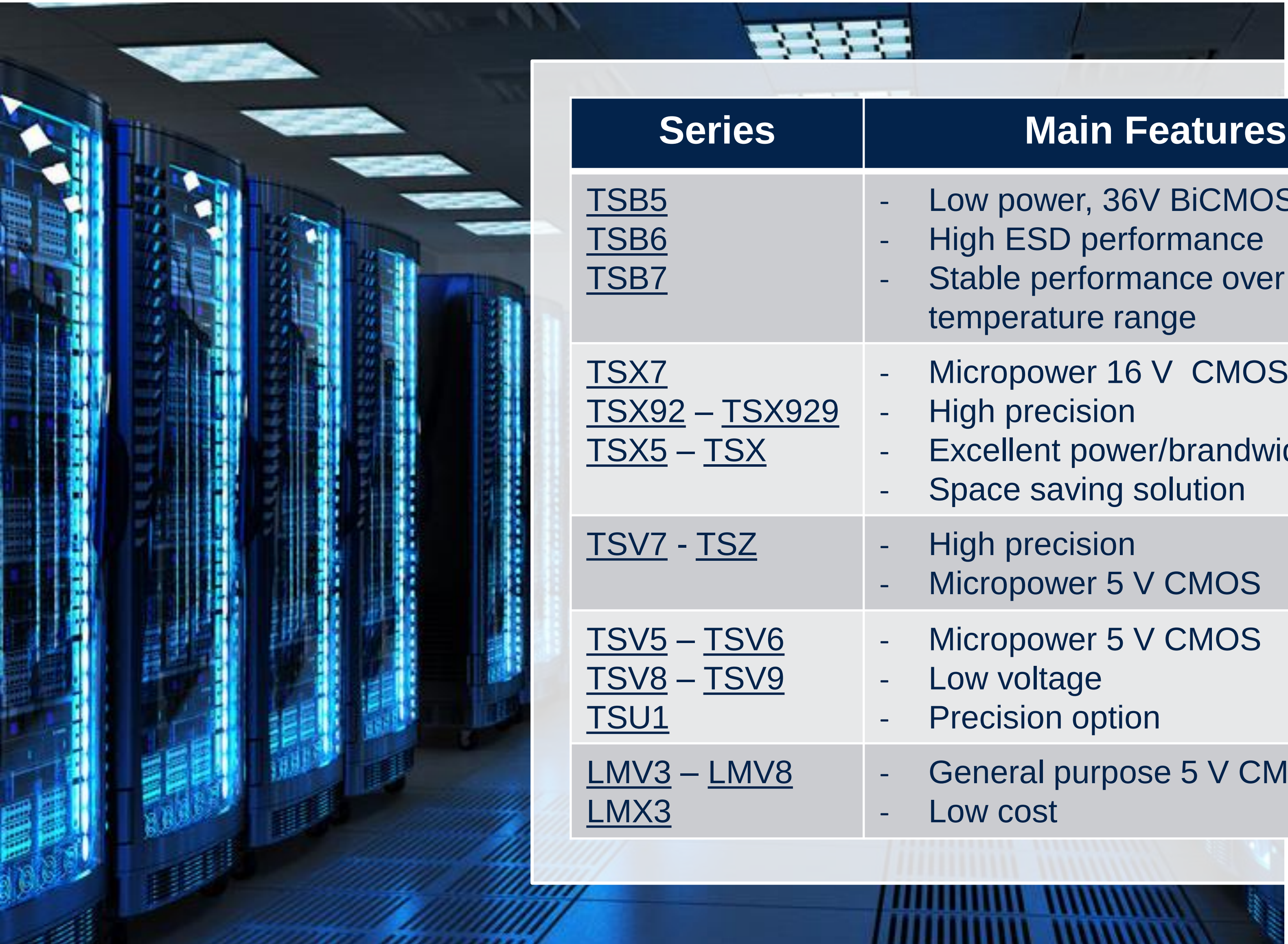
EXPLANATION (page 79)

This is a summary of the most important parameters to take into consideration when we speak about precision for an op amp. It represents the input referred error that an op amp might introduce in the whole measurement, and of course this error must be multiplied by the gain of the configuration.

If the op amp has an offset drift over temperature of 10µV/°C, it means that if the ambient temperature increases by 70 degrees Celsius, the input offset will increase by 700µV. If the common mode voltage of the application can vary from 0 to 3 volts, and if the amplifier has a CMRR of 80 dB you can expect a Vio variation of 300µV. If the power supply used to power the op amp varies by 10%, and if the amplifier shows a PSRR of 80dB you can expect a Vio variation of 50µV.

Noise is expressed on a 10 hertz bandwidth, and in the application. It is necessary to add a sharp 10 hertz filter to consider this 10 micro VPP otherwise the noise must be integrated over the entire bandwidth. The **TSZ121** which is a chopper op amp shows very good characteristics for all five parameters.

Precision op amps portfolio



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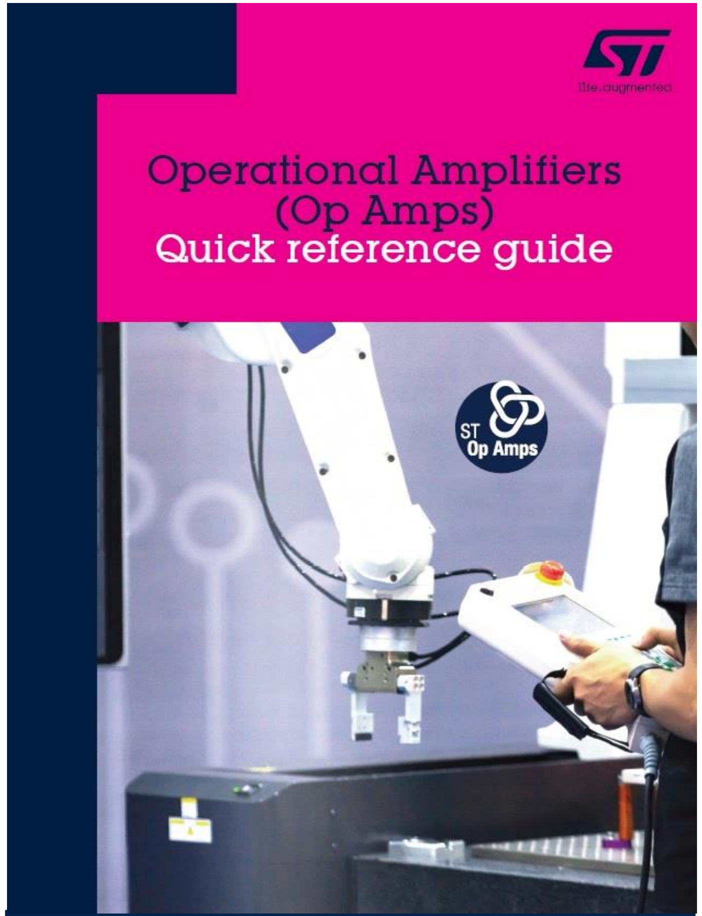

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