

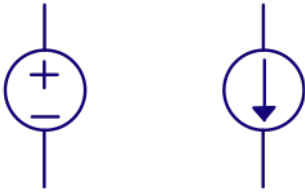
References and bias

Carsten Wulff, 2022-05-07

In our testbenches, and trial schematics, it's common to include voltage sources and current sources. However, the ideal voltage source, or ideal current source does not exist in the real world. There is no such thing.

We can come close to creating a voltage source, so a known voltage, with a low source impedance, but it won't be zero. And it won't be infinitely fast either. If we suddenly decide to pull 1 kA from a lab supply I promise you the voltage will drop.

So how do we create something that is a *good enough* voltage and current source on an IC?



Routing

Before we take a look at the voltage and current source, I want you to think about how you would route a current, or a voltage on an IC.

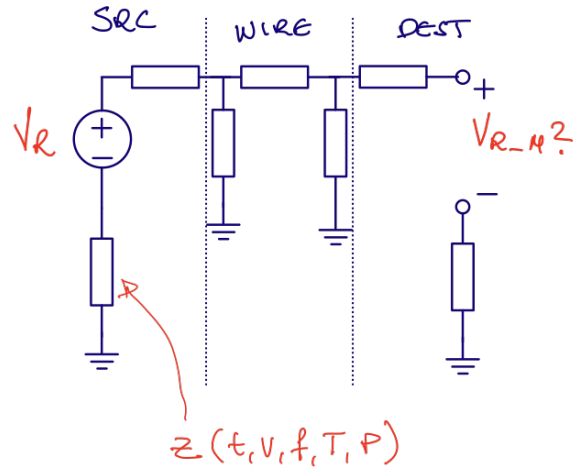
Assume we do have a known voltage on our IC. How can we make sure we can share that voltage across an IC?

A voltage is only defined between two points. There is no such thing as the *voltage at a point on a wire*, nor *voltage in a node*. Yes, I know we say that, but it's not right. What we forget is that by *voltage in a node* we always, always mean *voltage in a node referred to ground*.

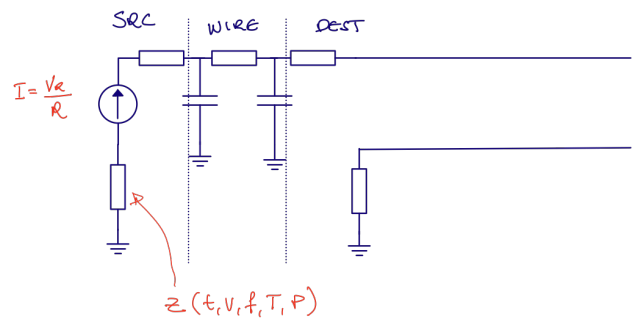
We've invented this magical place called *ground*, the final resting place of all electrons, and we have agreed that all voltages refer to that point.

So when we say "Voltage in node A is 1V", what we actually mean is "Voltage in node A is 1 V referred to ground".

Now you understand why we can't just route a known voltage across the IC, the other side might not have the same ground. The *other* side might have a different impedance to ground, and the impedance might be a function of time, voltage, frequency temperature, pressure and presence of gremlins.



Most of the time, in order not to think about the ground impedance, we choose to route a known quantity as a current instead of a voltage. That means, however, we must convert from a voltage to a current, but we can do that with a resistor (you'll see later), and as long as the resistor is the same on the other side of the IC, then we'll know what the voltage is.



BANDGAP VOLTAGE REFERENCE

A bipolar transistor, or indeed a PN diode, assuming a fixed current, will have a voltage across that is temperature dependent

$$I_D = I_S(e^{V_D/V_T} - 1)$$

As I_S is much smaller than I_D we can ignore the -1, which gives us

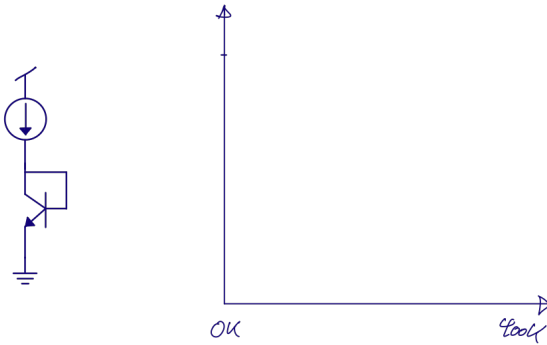
$$V_D = V_T \ln \frac{I_D}{I_S}$$

From this equation, it looks like the voltage V_D is proportional to temperature, as

$$V_T = \frac{kT}{q}$$

However, it turns out that the V_D usually decreases with temperature due to the temperature dependence of I_S . If you want to understand why, take a look at Diodes

The V_D is linear with temperature with a property that if you extrapolate the V_D line to zero Kelvin, then they all seem to meet at the bandgap voltage of silicon (approx 1.12 V).



If we take two diodes, or bipolars, biased at different current densities, for example

$$V_{D1} = V_T \ln \frac{I_{D1}}{I_S}$$

$$V_{D2} = V_T \ln \frac{I_{D2}}{I_S}$$

where $I_{D1} = 8I_{D2}$, then

$$V_{D1} - V_{D2} = V_T \ln \frac{I_{D1}}{I_S} - V_T \ln \frac{I_{D2}}{I_S} = V_T \ln \frac{I_{D1}}{I_{D2}} = V_T \ln 8$$

This is a remarkable result. The difference between two voltages is only defined by boltzmann's constant, temperature, charge, and a known current relationship.

This differential voltage can be used to read out directly the temperature on an IC, provided we have a known voltage to compare with.

We often call this voltage ΔV_D or ΔV_{BE} , and we can clearly see it's proportional to absolute temperature.

We know that the V_D decreases linearly with temperature, so if we combined a multiple of the ΔV_{BE} with a V_D voltage, then we should get a constant voltage.

The Brokaw reference below does exactly that. The opamp ensures the two bipolars have the same current. Q_1 is larger than Q_2 . The ΔV_{BE} is across the R_2 , so we

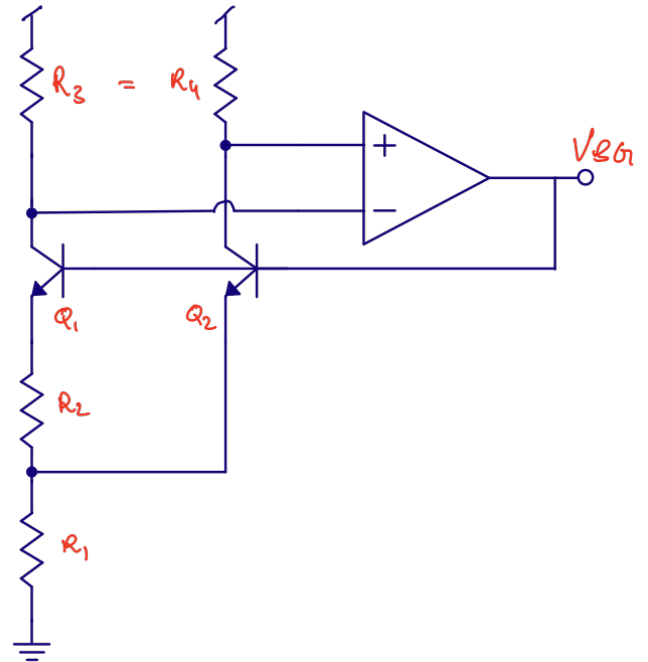
know the current I . We know that R_1 must then have $2I$. The voltage at the output will then be.

$$V_{BG} = V_{G0} + (m-1) \frac{kT}{q} \ln \frac{T_0}{T} + T \left[\frac{k}{q} \ln \frac{J_2}{J_1} \frac{2R_2}{R_1} - \frac{V_{G0} - V_{be0}}{T_0} \right]$$

where V_{G0} is the bandgap, V_{be0} is the base emitter measured at a temperature T_0 and the J 's are the current densities.

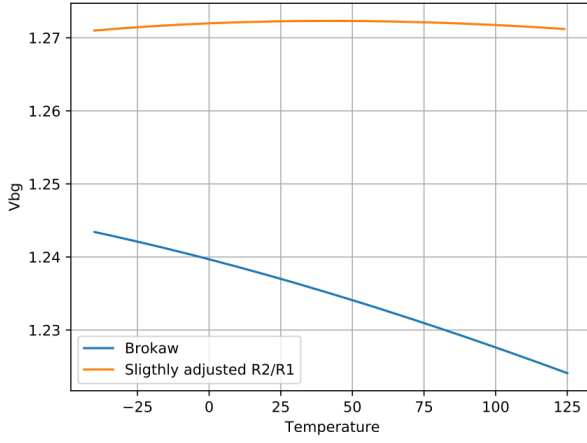
To get a constant output voltage, the relationship between the resistors should be approximately

$$\frac{R_2}{R_1} = \frac{V_{G0} - V_{be0}}{2T_0 \frac{k}{q} \ln(\frac{J_2}{J_1})}$$

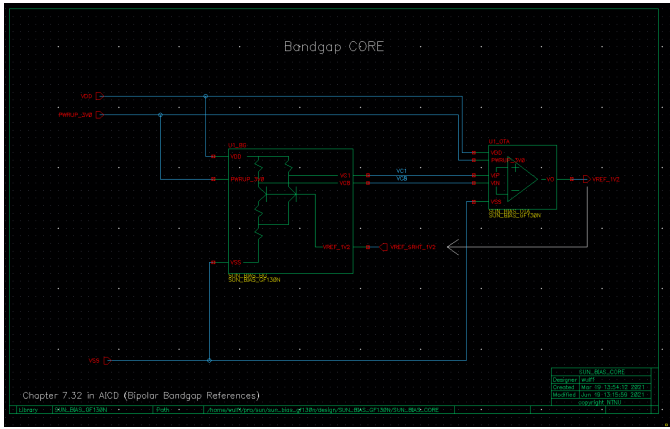


If you have access to SUN_BIAS_GF130N, then you can check the go program that was used to calculate the resistor relationship. The value was slightly tweaked in the design to get the desired output voltage.

sun/sun_bias_gf130n/go/



sun/sun_bias_gf130n/design/SUN_BIAS_GF130N/SUN_BIA:

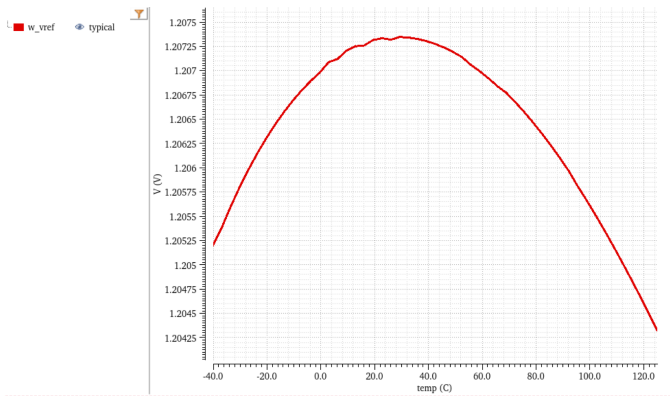


In typical simulations, the variation is extremely low over the temperature range. The second order error is the remaining error from

$$V_{BG} = V_{G0} + (m-1) \frac{kT}{q} \ln \frac{T_0}{T} + T \left[\frac{k}{q} \ln \frac{J_2}{J_1} \frac{2R_2}{R_1} - \frac{V_{G0} - V_{be0}}{T_0} \right]$$

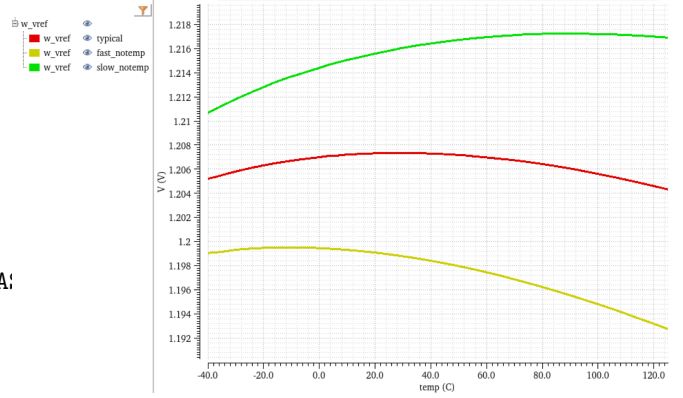
Where the last term is zero, so

$$V_{BG} = V_{G0} + (m-1) \frac{kT}{q} \ln \frac{T_0}{T}$$



Over corners, I do expect that there is variation. It may be that the V_D modeling is not perfect, which means the cancellation of the last term is incomplete.

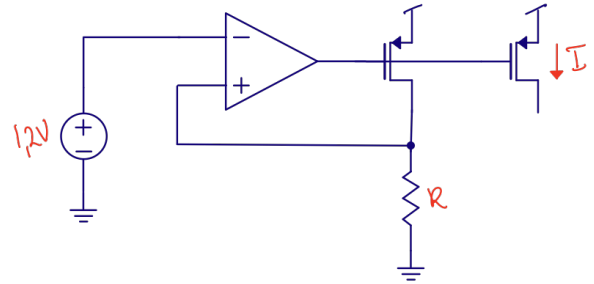
We could include trimming of R_1 to calibrate for the remaining error, however, if we wanted to remove the linear gradient, we would need a two point temperature test of every IC, which too expensive for low-cost devices.



With a known voltage, we can convert to a known current with the circuit below. On-chip we don't have accurate resistors, but for bias currents, it's usually ok with ± 20 variation (the variation of R).

Across a IC, we can expect to match the R within a few percent, so we can recreate a voltage with a accuracy of a few percent difference from the original.

If we wanted to create an accurate current, then we'd trim the R until the current is what we want.



LOW VOLTAGE BANDGAP

The Brokaw reference, and others, have a 1.2 V output voltage, which is hard if your supply is below about 1.4 V. As such, people have investigated making lower voltage references.

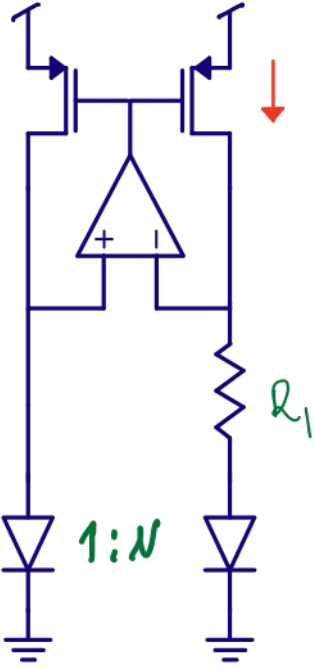
In real ICs though, you should ask yourself long and hard whether you really need these low-voltage references. Most ICs today still have a high voltage, either 1.8 V or 3.0 V.

If you do need them though, consider the circuit below. We have two diodes at different current densities. The ΔV_D will be across R_1 . The voltage at the input of the OTA will be V_D and the OTA will ensure the both are equal.

The current will then be

$$I_1 = \frac{\Delta V_D}{R_1}$$

and we know the current increases with temperature, since ΔV_D increases with temperature.

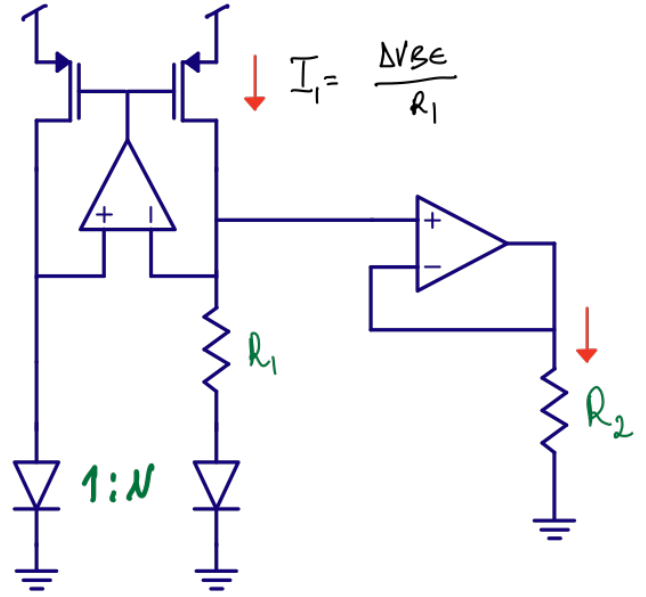


In the figure below I've used ΔV_{BE} , it's the same as ΔV_D , so ignore that error.

Assume we copy the V_D to another node, and place it across a second resistor R_2 , as shown in the figure below. The current in this second resistor is then

$$I_2 = \frac{V_D}{R_2}$$

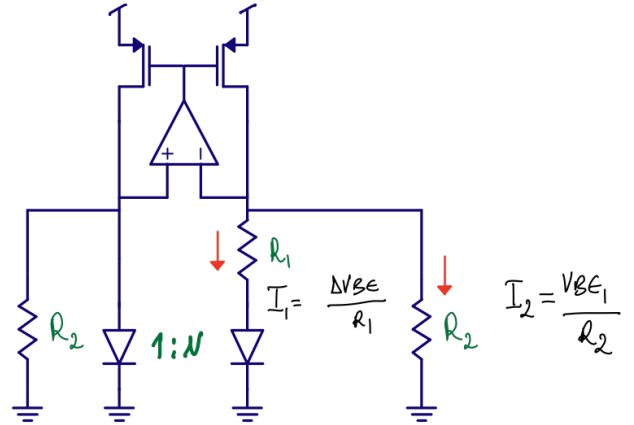
and we know the current decreases with temperature, since V_D decreases with temperature.



Let's remove the OTA, and connect R_2 directly to V_D nodes, you should convince yourself of the fact that this does not change I_1 at all. It does, however, change the current in the PMOS.

Provided we scale R_2 correctly, then the positive I_1 can be compensated by the negative I_2 , and we have a current that is independent of temperature.

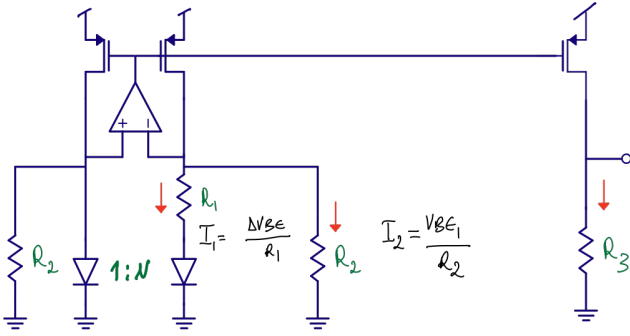
$$I_{PMOS} = \frac{V_D}{R_2} + \frac{\Delta V_D}{R_1}$$



Assuming we copy the current into another resistor R_3 , as shown below, we can get a voltage that is

$$V_{OUT} = R_3 \left[\frac{V_D}{R_2} + \frac{\Delta V_D}{R_1} \right]$$

Where the output voltage can be chosen freely, and indeed be lower than 1.2 V.



BIAS CURRENT

Sometimes we don't need a full bandgap reference. In those cases, we can use a GM cell as shown below, where the impedance could be a resistor, in which case

$$V_o = V_{GS1} - V_{GS2} = V_{eff1} + V_{tn} - V_{eff2} - V_{tn} = V_{eff1} - V_{eff2}$$

Assuming strong inversion, then

$$I_{D1} = \frac{1}{2} \mu_n C_{ox} \frac{W_1}{L_1} V_{eff1}^2$$

$$I_{D2} = \frac{1}{2} \mu_n C_{ox} 4 \frac{W_1}{L_1} V_{eff2}^2$$

$$I_{D1} = I_{D2}$$

$$\frac{1}{2} \mu_n C_{ox} \frac{W_1}{L_1} V_{eff1}^2 = \frac{1}{2} \mu_n C_{ox} 4 \frac{W_1}{L_1} V_{eff2}^2$$

$$V_{eff1} = 2V_{eff2}$$

Inserted into above

$$V_o = V_{eff1} - \frac{1}{2} V_{eff1} = \frac{1}{2} V_{eff1}$$

Still assuming strong inversion, such that

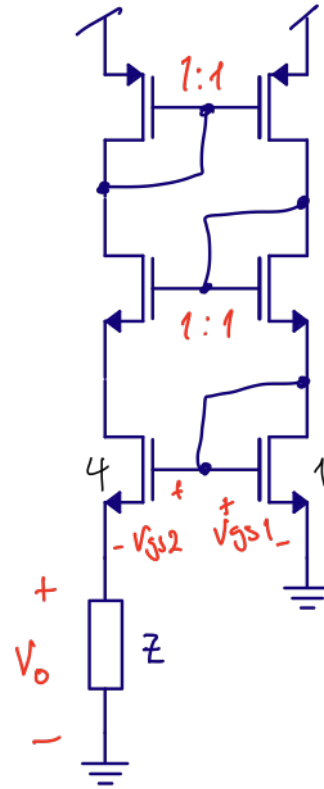
$$g_m = \frac{2I_d}{V_{eff}}$$

we find that

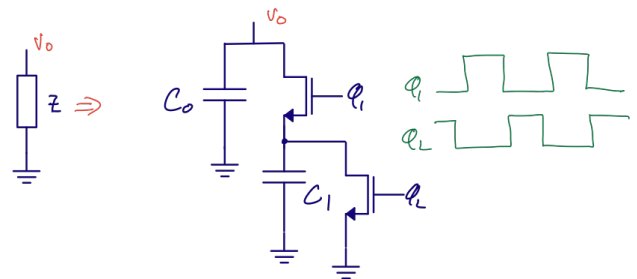
$$I = \frac{1}{2} V_{eff1} Z$$

$$Z \Rightarrow \frac{1}{g_m}$$

We can use a resistor for Z, or indeed we could use other things.



Like a switched capacitor



For bias circuits, it's common to have stable state if no current flows, as such, we usually include some form of startup circuit to get them going. A full bias circuit can be seen below.

