Voltage and Current References
References
References

- The reference standard for a kilogram is the one kilogram mass of a platinum-iridium kilogram maintained by the Bureau International des Poids et Mesures in Sèvres, France.
Historical Evolution

- 1964 – Hilbiber from Fairchild Semiconductor developed a voltage reference. Two strings with different transistor types. Obtained a constant voltage difference. However, he was unable to explain how the circuit worked.
- 1970 – Widlar was able to explain the phenomenon and came up with the “bandgap” reference.
- 1974 – Brokaw improved the bandgap reference design (Brokaw cell).
Widlar Current Source

• Typically used to create small currents without resorting to large resistors.
Widlar Current Source

\[ V_{BE1} - V_{BE2} - \frac{\beta + 1}{\beta} I_{OUT} R_2 = 0 \]

\[ V_T \ln \left( \frac{I_{C1}}{I_{S1}} \right) - V_T \ln \left( \frac{I_{OUT}}{I_{S2}} \right) - \frac{\beta + 1}{\beta} I_{OUT} R_2 = 0 \]

\[ V_T \ln \left( \frac{I_{IN}}{I_{S1}} \right) - V_T \ln \left( \frac{I_{OUT}}{I_{S2}} \right) - I_{OUT} R_2 \approx 0 \]

\[ V_T \ln \left( \frac{I_{IN}}{I_{OUT}} \right) = I_{OUT} R_2 \]
Widlar Current Source

- Take $I_{IN} = 1\text{mA}$ and $I_{OUT} = 5\mu\text{A}$.
- $I_{OUT}R_2 = \ln(200)V_T = 5.3V_T = 136.7 \text{ mV}$
- $R_2 = 27.3 \text{ K}$
- $R_1 = (5\text{V} - 0.7\text{V})/1\text{mA} = 4.3\text{K}$
- The total resistance in the circuit is only 31.7K
- If ordinary current mirror were used, $R = 4.3\text{V}/5\mu\text{A} = 1\text{M}$.
Widlar Current Source

• A MOS version is also possible

\[ V_{GS1} - V_{GS2} - I_{OUT} R_2 = 0 \]
\[ I_{OUT} R_2 + V_{ov2} - V_{ov1} = 0 \]

• The rest of the derivation can be found in your book.
Bipolar Peaking Current Source

• The Widlar source yields currents on the order of $\mu$A. When nA are needed, use the following circuit.
Bipolar Peaking Current Source

• The analysis is as follows

\[
V_{BE1} - I_{IN} R = V_{BE2} \\
V_{A} \rightarrow \infty \\
V_{T} \ln \left( \frac{I_{IN}}{I_{S1}} \right) - V_{T} \ln \left( \frac{I_{OUT}}{I_{S2}} \right) = I_{IN} R \\
I_{S1} \approx I_{S2} \\
I_{OUT} = I_{IN} \exp \left( \frac{-I_{IN} R}{V_{T}} \right) \\
R = \frac{V_{T}}{I_{IN}} \ln \left( \frac{I_{IN}}{I_{OUT}} \right)
\]
Bipolar Peaking Current Source

![Graph showing current versus sweep]
Bipolar Peaking Current Source

• When the input current is small, the voltage drop on resistor is small. Thus $V_{BE2}$ is approx. equal to $V_{BE1}$ and the currents are approx. equal.

• $V_{BE1}$ increases logarithmically with current whereas resistor drop increases linearly.

• As a result, the base voltage of Q2 initially increases, then decreases.

• Output current is max when $V_{BE2}$ is max.
Bipolar Peaking Current Source

![Graph of Bipolar Peaking Current Source]

- Voltage (mV) vs. Sweep (uA)
- Curve labeled v(3)
MOS Peaking Current Source

- Operation is similar.

\[ I_{OUT} = \frac{k'(W/L)^2}{2} (V_{ov2})^2 = \frac{k'(W/L)^2}{2} (V_{ov1} - I_{IN}R)^2 \]

- However, transistors are probably in weak inversion (subthreshold)

\[ I_{OUT} \approx I_{IN} \exp \left( -\frac{I_{IN}R}{nV_T} \right) \]
Supply Insensitive Biasing

• Simple bipolar current mirror with resistor,

\[ I_{OUT} \approx I_{IN} = \frac{V_{CC} - V_{BE}}{R} \approx \frac{V_{CC}}{R} \]

• Define Sensitivity as

\[ S^y_x = \lim_{\Delta x \to 0} \frac{\Delta y/y}{\Delta x/x} = \frac{x}{y} \frac{\partial y}{\partial x} \]

• Applying this definition, \( S^I_{V_{SUP}} \approx 1 \)
Supply Insensitive Biasing

• Apply the definition to a bipolar Widlar source

\[
V_T \frac{\partial}{\partial V_{CC}} \left[ \ln \left( \frac{I_{IN}}{I_{OUT}} \right) \right] = R_2 \frac{\partial I_{OUT}}{\partial V_{CC}}
\]

\[
\frac{\partial I_{OUT}}{\partial V_{CC}} = \left( 1 + \frac{I_{OUT} R_2}{V_T} \right) \frac{I_{OUT}}{I_{IN}} \frac{\partial I_{IN}}{\partial V_{CC}}
\]

\[
S'_{V_{CC}} = \left( 1 + \frac{I_{OUT} R_2}{V_T} \right) S'_{V_{CC}} \approx \frac{1}{1 + \frac{I_{OUT} R_2}{V_T}}
\]
Supply Independent Biasing

• This sensitivity is still not good enough for most applications.
• One idea is to use base-emitter, threshold. Or zener voltages.
• Both base-emitter and threshold voltages are temperature dependent.
• Zener voltages are too large and require large currents. They also produce noise.
$V_{BE}$ Referenced Biasing

• Try a modified version of the Wilson Current Mirror
V_{BE} Referenced Biasing

• The output current is determined by the ratio of V_{BE1} and R_2.

\[ V_{BE1} = V_T \ln \left( \frac{I_{IN}}{I_{S1}} \right) \]

\[ I_{OUT} = \frac{V_{BE1}}{R_2} = \frac{V_T}{R_2} \ln \left( \frac{I_{IN}}{I_{S1}} \right) \]

\[ S_{VCC}^I_{OUT} = \frac{V_T}{I_{OUT} R_2} S_{VCC}^I_{IN} \approx \frac{V_T}{V_{BE1}} \approx \frac{26mV}{700mV} = 0.037 \]

• Similar analysis for the MOSFET yields

\[ S_{V_{DD}}^I_{OUT} \approx \frac{V_{ov}}{2V_{GS1}} \approx \frac{0.2V}{1.4V} = 0.14 \]
$V_{BE}$ Referenced Biasing

- The BJT version of the circuit designed for 100µA.
Self Biasing

Current Mirror

Current Source
Self Biased $V_{BE}$ Reference
Self Biased $V_{BE}$ Reference

- Two output currents $I_{BIAS1}$ and $I_{BIAS2}$ are taken over the additional transistors Q3 and Q6.
- Q1 and Q2 form the current reference circuit along with the resistor R1.
- Q4 and Q5 form the current mirror which stabilizes the current.
- The CMOS version has exactly the same circuit diagram.
Self Biased $V_T$ Reference Simulation

- Current through the self biased section:
Self Biased $V_T$ Reference Simulation

- Output Currents
Self Biased Micro-current Generator

• To obtain very small currents, large resistances are required.
• This can be solved by making the current dependent not on the $V_{GS}$ or $V_{BE}$, but on a difference of two $V_{GS}$ or $V_{BE}$ values.
• Lower voltage yields lower current.
Self Biased Micro-current Generator
Self Biased Circuits

• Self biased circuits need a start-up circuit because I=0 is also a stable point for the circuit.

• Start-up circuits operate either by always having a small current through the circuit or by sensing the zero current state and forcing the reference circuit out of it.
Self Biased $V_{BE}$ Reference with Startup
Self Biased $V_T$ Reference with Start-up
The previous two circuits were static start-up circuits.

Dynamic start-up also possible by using a resistor and a capacitor.

Dynamic circuits do not monitor the currents in the circuit, but simply provide a glitch which should start-up the circuit.
Start-up Circuits
Temperature Dependence

• Define a fractional temperature coefficient as

\[ TC_F = \frac{1}{I_{OUT}} \frac{\partial I_{OUT}}{\partial T} \]

• For the \( V_{BE} \) referenced circuit,

\[
I_{OUT} = \frac{V_{BE1}}{R}
\]

\[
\frac{\partial I_{OUT}}{\partial T} = \frac{1}{R} \frac{\partial V_{BE1}}{\partial T} - \frac{V_{BE1}}{R^2} \frac{\partial R}{\partial T}
\]

\[
= I_{OUT} \left( \frac{1}{V_{BE1}} \frac{\partial V_{BE1}}{\partial T} - \frac{1}{R} \frac{\partial R}{\partial T} \right)
\]

\[
TC_F = \frac{1}{V_{BE1}} \frac{\partial V_{BE1}}{\partial T} - \frac{1}{R} \frac{\partial R}{\partial T}
\]
Temperature Dependence

• Note that $V_{BE}$ has a negative $TC_F$ (-2mV/°C) and $R$ has a positive $TC_F$ (around 1500 ppm/°C).

• The two effects add up to worsen the temperature dependence.

• The threshold voltage has a similar temperature dependence as the $V_{BE}$ (-2mV/°C)

• Thus, the MOSFET source has similar temperature performance.
$V_{BE}$ Referenced Self Biased Circuits

- $V_{BE}$ referenced self biased circuits can also be implemented in CMOS by using a parasitic pnp device inherent in p-substrate CMOS.
- The same is true for a parasitic npn in an n-substrate CMOS.
$V_{BE}$ Based Reference
References Using Thermal Voltage
References Using Thermal Voltage

• Using our experience from the Widlar current source,

\[ I_{OUT}R_1 = V_T \ln \left( \frac{I_{IN}}{I_{OUT}} - \frac{I_{S2}}{I_{S1}} \right) \]

• Choose Q3 and Q4 to be identical

\[ I_{OUT}R_1 = V_T \ln(2) \]
References Using Thermal Voltage

• The temperature dependence can be calculated as,

\[
\frac{\partial I_{OUT}}{\partial T} = \ln(2) \frac{R_1 \frac{\partial V_T}{\partial T} - V_T \frac{\partial R_2}{\partial T}}{R_1^2} = \frac{V_T}{R_1} \ln(2) \left( \frac{1}{V_T} \frac{\partial V_T}{\partial T} - \frac{1}{R_1} \frac{\partial R_1}{\partial T} \right)
\]

\[
TC_F = \frac{1}{I_{OUT}} \frac{\partial I_{OUT}}{\partial T} = \frac{1}{V_T} \frac{\partial V_T}{\partial T} - \frac{1}{R_1} \frac{\partial R_1}{\partial T}
\]

• Note that the two terms compensate for each other. Also, the ratio does not have to be 2, but is chosen for optimum sizing.
References Using Thermal Voltage

• The same idea can be utilized in CMOS with parasitic BJT’s.
• Note from previous discussion that cascoding may also be a good option to reduce power supply dependence.
References Using Thermal Voltage
Voltage Biasing

• The previous circuits have been for current bias only.
• However, we would also like to use reference voltages in addition to currents in many analog circuits.
• The quick and dirty solution is to use a voltage divider from the supply.
Voltage Dividers
Voltage Dividers

• The analysis is quite simple

\[
\frac{\mu_1 C_{ox}}{2} \left( \frac{W}{L} \right)_1 (V_{DS1} - V_{Th1})^2 = \frac{\mu_2 C_{ox}}{2} \left( \frac{W}{L} \right)_2 (V_{DS2} - V_{Th2})^2
\]

\[
V_{DS1} + V_{DS2} = V_{DD}
\]

\[
V_1 = V_{DS1} = \frac{\alpha_2}{\alpha_1 + \alpha_2} V_{DD} + \frac{\alpha_1 V_{Th1} - \alpha_2 V_{Th2}}{\alpha_1 + \alpha_2}
\]

\[
\alpha_1 = \sqrt{\mu_1 \left( \frac{W}{L} \right)_1}
\]

\[
\alpha_2 = \sqrt{\mu_2 \left( \frac{W}{L} \right)_2}
\]
Voltage References

• We have available in a good CMOS technology the following options to use as a reference:
  – $V_{BE}$ of a parasitic bipolar transistor
  – Threshold difference of MOS transistors
  – The thermal voltage $V_T$
$V_{BE}$ Multiplier
V\textsubscript{BE} Multiplier

- The resistors can be made of the same type of material. Hence, the matching between them can be quite good in a wide range of temperatures.
- However, V\textsubscript{BE} changes with about -2mV/°C.
- Not good as a reference under changing temperature.
$V_T$ Multiplier
**$V_T$ Multiplier**

- The resistors can be made of the same type of material. Hence, the matching between them can be quite good in a wide range of temperatures.
- However, $V_T$ depends directly on temperature
  \[ V_T = \frac{kT}{q} \]
- Not good as a reference under changing temperature.
Voltage Reference Based on Threshold Differences
Voltage Reference Based on Threshold Differences

• M1 is a transistor with a different threshold voltage.
• \( V_{\text{out}} = -V_{\text{th1}} + V_{\text{th2}} \)
• The threshold voltages follow each other.
• The \( V_{DS} \) and overdrive voltages are similar.
• However, different threshold voltage devices are not typically available.
Temperature Insensitive Biasing

- Imagine a circuit which adds the $V_{BE}$ of a BJT with several times $V_T$.
- $V_{REF} = V_{BE} + mV_T$
- At room temp., $V_{BE}$ changes by -2.2 mV/°C and $V_T$ by 0.086 mV/°C.
- If you choose $m$ as 25.6, they will cancel each other out perfectly at room temperature.
- In this case, $V_{REF} = 1.31V$
- Note that $V_T = kT/q$ and is Proportional To Absolute Temperature (PTAT)
Temperature Insensitive Biasing

• Let us now do a more rigorous analysis

\[ V_{BE(ON)} = V_T \ln \left( \frac{I_1}{I_S} \right) \]

\[ I_S = \frac{qAn_i D_n}{Q_B} = Bn_i^2 \bar{D}_n = B'n_i^2 T \bar{\mu}_n \]

\[ \bar{\mu}_n = CT^{-n} \]

\[ n_i^2 = DT^3 \exp \left( -\frac{V_{G0}}{V_T} \right) \]
Temperature Insensitive Biasing

• Combining all of these,

\[ V_{BE(ON)} = V_T \ln \left( I_1 T^{-\gamma} E \exp \left( -\frac{V_{G0}}{V_T} \right) \right) \]

\[ \gamma = 4 - n \]

• However, in practical cases, \( I_1 \) is also temperature dependent.

\[ I_1 = G T^\alpha \]
Temperature Insensitive Biasing

• Thus, combining these and rearranging,

\[ V_{BE(ON)} = V_{G0} - V_T \left[ (\gamma - \alpha) \ln T - \ln(EG) \right] \]

• Now, let us add the several \( V_T \),

\[ V_{OUT} = V_{BE(ON)} + M V_T \]

\[ V_{OUT} = V_{G0} - V_T (\gamma - \alpha) \ln T + V_T \left[ M + \ln(EG) \right] \]
Temperature Independent Biasing

• To obtain temperature independence, take derivative and equate to zero.

\[ 0 = \frac{dV_{OUT}}{dT} \bigg|_{T=T_0} = \frac{V_{T0}}{T_0} \left[ M + \ln(EG) \right] - \frac{V_{T0}}{T_0} (\gamma - \alpha) \ln T_0 - \frac{V_{T0}}{T_0} (\gamma - \alpha) \]

\[ M + \ln(EG) = (\gamma - \alpha) \ln T_0 + (\gamma - \alpha) \]

• Substituting this \( T_0 \),

\[ V_{OUT} = V_{G0} + V_T (\gamma - \alpha) \left( 1 + \ln \left( \frac{T_0}{T} \right) \right) \]

\[ V_{OUT} \bigg|_{T=T_0} = V_{G0} + V_{T0} (\gamma - \alpha) \]
Temperature Independent Biasing

• $V_{G0}$ is the band-gap voltage of silicon and is 1.205V.

• $\gamma$ is a constant between 3 and 3.5.

• $\alpha$ is typically 1.

• $V_{OUT}$ is 1.262V for $\gamma = 3.2$.

• $V_{OUT}$ is close to the band-gap voltage of silicon. Hence, this approach is called bandgap reference.
Temperature Independent Biasing

• Note that the temperature coefficient is zero only at a single voltage value, $T_0$. Use the M to adjust this temperature.

• One can also define a temperature range for this approach.

\[
TC_{F(\text{eff})} = \frac{1}{V_{OUT}} \left( \frac{V_{\text{MAX}} - V_{\text{MIN}}}{T_{\text{MAX}} - T_{\text{MIN}}} \right)
\]
Temperature Independent Biasing
Temperature Independent Biasing

• The circuit is called a Widlar band-gap reference.
• $V_{\text{OUT}}$ is given by $V_{\text{BE3}} + V_{\text{R2}}$.
• $V_{\text{R2}} = V_{\text{R3}} \times (R_2/R_3)$
• $V_{\text{R3}}$ is the $V_{\text{BE}}$ difference between Q1 and Q2.
• This circuit ignores the power supply dependence of I.
Temperature Independent Biasing
Temperature Independent Biasing

- The reference creates its own potential, thereby counteracting the power supply dependence.

\[
V_{R3} = \Delta V_{BE} = V_{BE1} - V_{BE2} = V_T \ln \left( \frac{I_1}{I_2} \frac{I_{S2}}{I_{S1}} \right) = V_T \ln \left( \frac{R_2}{R_1} \frac{I_{S2}}{I_{S1}} \right)
\]

\[
V_{R2} = \frac{R_2}{R_3} V_{R3} = \frac{R_2}{R_3} V_T \ln \left( \frac{R_2}{R_1} \frac{I_{S2}}{I_{S1}} \right)
\]

\[
V_{OUT} = V_{BE2} + V_{R3} + V_{R2} = V_{BE2} + \left(1 + \frac{R_2}{R_3}\right) V_T \ln \left( \frac{R_2}{R_1} \frac{I_{S2}}{I_{S1}} \right) = V_{BE2} + MV_T
\]
Temperature Independent Biasing

• Bandgap reference circuits can also be used in CMOS with parasitic BJT’s.
• Remember that the offset of CMOS OPAMP’s is much larger than BJT OPAMP’s.
• Offset voltage effects the Bias generator in an undesired manner.
Bandgap References
Bandgap References

• The circuit adds the two quantities in voltage mode.

\[ \Delta V_{BE} = V_T \ln \left( \frac{I_1 I_{S2}}{I_2 I_{S1}} \right) = R_3 I_2 \]

\[ \frac{I_1}{I_2} = \left( \frac{W}{L} \right)_1 \left( \frac{W}{L} \right)_2 \]

\[ V_{BG} = V_{BE1} + \frac{R_2}{R_3} \Delta V_{BE} = V_{BE1} + V_T \frac{R_2}{R_3} \ln \left( \frac{\left( \frac{W}{L} \right)_1 A_2}{\left( \frac{W}{L} \right)_2 A_1} \right) \]
This example is based on current processing.