SECTION 4.5 - CURRENT AND VOLTAGE REFERENCES

Characteristics of a Voltage or Current Reference

What is a Voltage or Current Reference?

A voltage or current reference is an independent voltage or current source that has a high degree of precision and stability.

Requirements of a Reference Circuit:

- Should be independent of power supply
- Should be independent of temperature
- Should be independent of processing variations
- Should be independent of noise and other interference

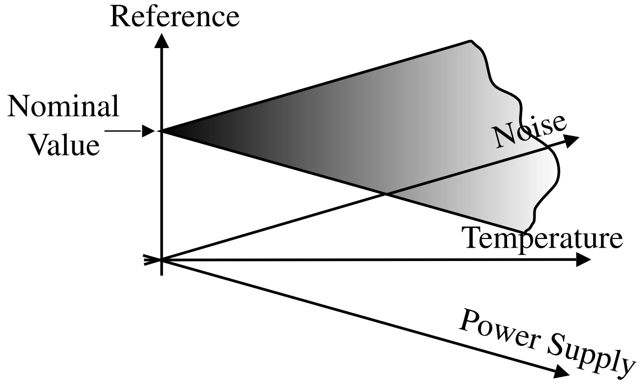


Fig. 4.5-1

REFERENCES WITH POWER SUPPLY INDEPENDENCE

Power Supply Independence

How do you characterize power supply independence?

Use the concept of:

$$S_{V_{DD}}^{I_{REF}} = \frac{\partial I_{REF}/I_{REF}}{\partial V_{DD}/V_{DD}} = \frac{V_{DD}}{I_{REF}}\left(\frac{\partial I_{REF}}{\partial V_{DD}}\right)$$

Application of sensitivity to determining power supply dependence:

$$\frac{\partial I_{REF}}{I_{REF}} = \left(S_{V_{DD}}^{I_{REF}}\right)\frac{\partial V_{DD}}{V_{DD}}$$

Thus, the fractional change in the reference voltage is equal to the sensitivity times the fractional change in the power supply voltage.

For example, if the sensitivity is 1, then a 10% change in V_{DD} will cause a 10% change in I_{REF}.

Ideally, we want $S_{V_{DD}}^{I_{REF}}$ to be zero for power supply independence.

Simple Current Reference

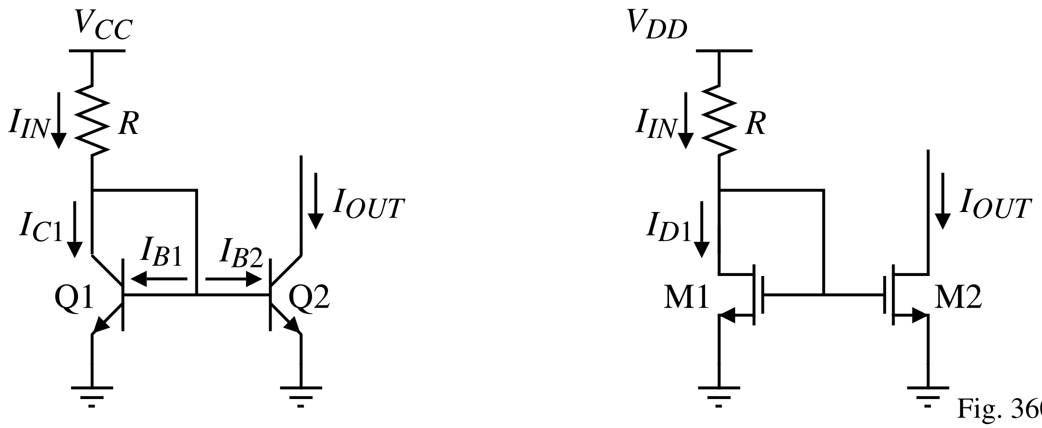


Fig. 360-02

$$I_{OUT} \approx \frac{V_{CC}-V_{BE}}{R}\left(\frac{1}{1+\frac{2}{\beta_F}}\right) \qquad I_{OUT} \approx \frac{V_{DD}-V_{GS}}{R} = \frac{V_{DD} - \sqrt{\frac{2I_{IN}}{\beta_1}} - V_T}{R}$$

$$S_{V_{CC}}^{I_{REF}} = 1 \qquad S_{V_{DD}}^{I_{REF}} = 1$$

Temperature and process dependence?

MOS Widlar Current Reference

Operation:

$$V_{GS1} - V_{GS2} - I_{OUT}R_2 = 0$$

$$I_{OUT}R_2 + V_{ON2} - V_{ON1} = 0$$

Assuming strong inversion and $\lambda \rightarrow 0$,

$$I_{OUT}R_2 + \sqrt{\frac{2I_{OUT}}{K'(W_2/L_2)}} - V_{ON1} = 0$$

Solving for $\sqrt{I_{OUT}}$ gives,

$$\sqrt{I_{OUT}} = \frac{-\sqrt{\frac{2}{K'(W_2/L_2)}} + \sqrt{\frac{2}{K'(W_2/L_2)} + 4R_2V_{ON1}}}{2R_2}$$

where $V_{ON1} = \sqrt{\frac{2I_{IN}}{K'(W_1/L_1)}}$

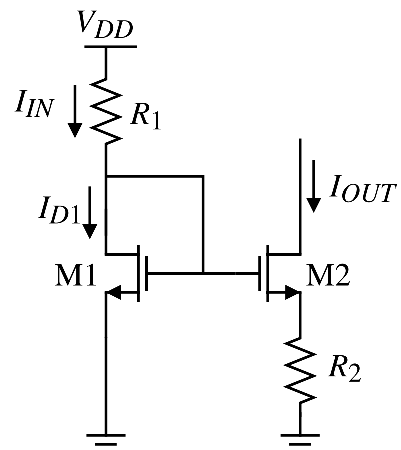


Fig. 360-04

Differentiating I_{OUT} with respect to V_{DD} gives,

$$\frac{1}{2\sqrt{I_{OUT}}}\frac{dI_{OUT}}{dV_{DD}} = \frac{1}{\sqrt{2/(K'\,W_2/L_2) + 4R_2V_{ON1}}}\frac{dV_{ON1}}{dV_{DD}}, \qquad \frac{dV_{ON1}}{dV_{DD}} = \frac{V_{ON1}}{2I_{IN}}\frac{dI_{IN}}{dV_{DD}}$$

$$\therefore \quad S_{V_{DD}}^{I_{REF}} = S_{V_{DD}}^{I_{OUT}} = \frac{V_{ON1}}{\sqrt{V_{ON2}{}^2 + 4\,I_{OUT}R_2V_{ON1}}}S_{V_{DD}}^{I_{IN}} \approx \frac{V_{ON1}}{\sqrt{4V_{ON1}{}^2}}S_{V_{DD}}^{I_{IN}} = 0.5S_{V_{DD}}^{I_{IN}}$$

Example 4.5-1

For the MOS Widlar current reference, find I_{OUT} if I_{IN} = 100µA, R_2 = 4kΩ, K' = 200µA/V^2, and $W_2/L_2 = W_1/L_1 = 25$. Assume the temperature is 27°C and that n = 1.5. Find the sensitivity of I_{OUT} with respect to V_{DD}.

Solution

$$V_{ON1} = \sqrt{\frac{2I_{IN}}{K'(W_1/L_1)}} = \sqrt{\frac{2\cdot 100}{200\cdot 25}} = 0.2\text{V}$$

$$\sqrt{I_{OUT}} = \frac{-\sqrt{\frac{2}{200\cdot 25}} + \sqrt{\frac{2}{200\cdot 25} + 4(0.004)0.2}}{20.004}\sqrt{\mu A} = 5\sqrt{\mu A} \quad\Rightarrow\quad I_{OUT} = 25\mu A$$

Note that $V_{ON2} = V_{ON1} - I_{OUT}R_2$ = 0.2-(25)(0.004) = 0.1V > $2nV_t$ = 78mV so both transistors are in strong inversion.

For the sensitivity calculations, assume that $V_{DD} >> V_{GS1}$. Therefore $I_{IN} \approx V_{DD}/R_1$.

$$S_{V_{DD}}^{I_{REF}} = \frac{V_{ON1}}{\sqrt{4V_{ON2}{}^2}} S_{V_{DD}}^{I_{IN}} \approx \frac{V_{ON1}}{\sqrt{4V_{ON2}{}^2}} = 0.5$$

Therefore, a 10% variation in V_{DD} causes a 5% variation in I_{OUT}.

MOS Peaking Current Reference

Circuit:

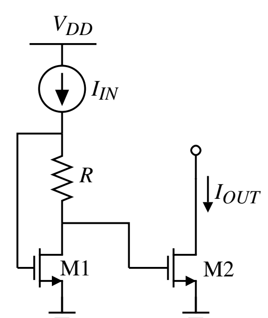


Fig. 360-7

Strong Inversion Operation:

$$V_{GS1} - I_{IN}R - V_{GS2} = 0$$

$$V_{ON2} = V_{ON1} - I_{IN}R$$

$$I_{OUT} = \frac{K'(W_2/L_2)}{2} V_{ON2}{}^2 = \frac{K'(W_2/L_2)}{2}(V_{ON1} - I_{IN}R)^2$$

where

$$V_{ON1} = \sqrt{\frac{2I_{IN}}{K'(W_1/L_1)}}$$

Weak Inversion Operation:

$$V_{GS2} - V_T \approx nV_t \ln\left(\frac{I_{IN}}{(W_1/L_1)I_T}\right) - I_{IN}R$$

If the transistors are identical and $V_{DS2} > 3V_T$,

$$I_{OUT} = \frac{W_1}{L_1} I_T \exp\left(\frac{V_{GS2} - V_T}{nV_t}\right) \approx I_{IN} \exp\left(\frac{-I_{IN}R}{nV_t}\right)$$

Transfer Characteristics:

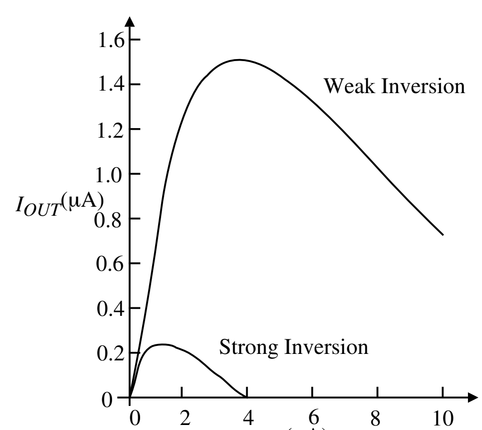


Fig. 360-8

Threshold Referenced Current Reference

Circuit:

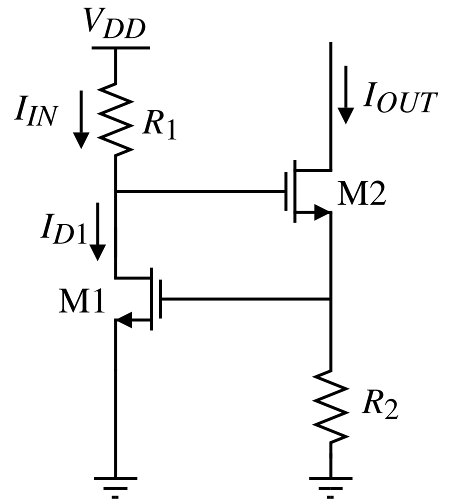


Fig. 360-10

Operation:

$$I_{OUT} = \frac{V_{GS1}}{R_2} = \frac{V_T + \sqrt{\frac{2I_{IN}}{K'(W_1/L_1)}}}{R_2}$$

$$\approx \frac{V_T}{R_2} \text{ if } V_T > V_{ON1}$$

The sensitivity of I_{OUT} with respect to V_{DD} is

$$S_{V_{DD}}^{I_{OUT}} = \left(\frac{V_{ON1}}{I_{OUT}R_2}\right) S_{V_{DD}}^{I_{IN}} = \left(\frac{V_{ON1}}{2V_{GS1}}\right) S_{V_{DD}}^{I_{IN}}$$

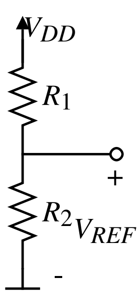
For example, if $V_T = 1\text{V}$, $V_{ON1} = 0.1\text{V}$ and $S_{V_{DD}}^{I_{IN}} \approx 1$, then

$$S_{V_{DD}}^{I_{OUT}} = \left(\frac{0.1}{2\cdot 1.1}\right) = 0.045$$

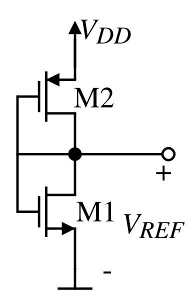
Therefore, if V_{DD} changes by 10%, I_{REF} or I_{OUT} changes by 0.45%.

SIMPLE BIAS/REFERENCE CIRCUITS

Voltage References using Voltage Division



Resistor voltage divider.



Active device voltage divider. Fig. 370-01

$$V_{REF} = \frac{R_2}{R_1+R_2} V_{DD}$$

$$V_{REF} = \frac{V_{TN} + \sqrt{(\beta_P/\beta_N)}\,(V_{DD}-|V_{TP}|)}{1 + \sqrt{(\beta_P/\beta_N)}}$$

$$S_{V_{DD}}^{V_{REF}} = 1$$

$$S_{V_{DD}}^{V_{REF}} = \frac{V_{DD}}{V_{REF}}\left(\frac{\sqrt{(\beta_P/\beta_N)}}{1+\sqrt{(\beta_P/\beta_N)}}\right) = \frac{V_{DD}\sqrt{(\beta_P/\beta_N)}}{V_{TN} + \sqrt{(\beta_P/\beta_N)}\,(V_{DD}-|V_{TP}|)}$$

$$= \frac{V_{DD}\sqrt{(\beta_P/\beta_N)}}{V_{TN} + \sqrt{(\beta_P/\beta_N)}\,(V_{DD}-|V_{TP}|)}$$

Assume $\beta_N = \beta_P$ and $V_{TN} = |V_{TP}|$ $\Rightarrow$ $S_{V_{DD}}^{V_{REF}} = 1$

References with Sensitivity Less than One

In order to get sensitivities less than one, the upper and lower circuits must be different with the lower circuit less dependent on V_{DD}.

In otherwords, the upper circuit should act like a current source and the lower circuit like a voltage source.

Principle:

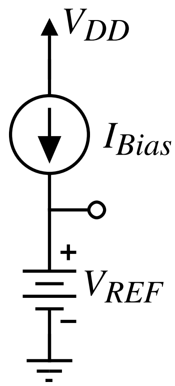


Fig. 370-02

MOSFET-Resistance Voltage References

Fig. 370-03

$$V_{REF} = V_{GS} = V_T + \sqrt{\frac{2(V_{DD}-V_{REF})}{\beta R}}$$

or $V_{REF} = V_T - \frac{1}{\beta R} + \sqrt{\frac{2(V_{DD}-V_T)}{\beta R} + \frac{1}{(\beta R)^2}}$

$$S_{V_{DD}}^{V_{REF}} = \left(\frac{1}{1+\beta(V_{REF}-V_T)R}\right)\left(\frac{V_{DD}}{V_{REF}}\right)$$

Assume that V_{DD} = 5V, W/L = 2 and R = 100kΩ,

Thus, $V_{REF} \approx 1.281\text{V}$ and $S_{V_{REF}}^{V_{DD}} = 0.283$

This circuit allows V_{REF} to be larger.

If the current in R_1 (and R_2) is small compared to the current flowing through the transistor, then

$$V_{REF} \approx \left(\frac{R_1 + R_2}{R_2}\right) V_{GS}$$

Bipolar-Resistance Voltage References

Fig. 370-04

$$V_{REF} = V_{EB} = \frac{kT}{q} \ln\left(\frac{I}{I_s}\right)$$

$$I = \frac{V_{CC} - V_{EB}}{R} \cong \frac{V_{CC}}{R}$$

$$V_{REF} \cong \frac{kT}{q} \ln\left(\frac{V_{CC}}{RI_s}\right)$$

$$S_{V_{CC}}^{V_{REF}} = \frac{1}{\ln[V_{CC}/(RI_s)]} = \frac{1}{\ln(I/I_s)}$$

If V_{CC}=5V, R = 4.3kΩ and I_s = 1fA, then V_{REF} = 0.719V.

Also, $S_{V_{CC}}^{V_{REF}} = 0.0362$

If the current in R_1 (and R_2) is small compared to the current flowing through the transistor, then

$$V_{REF} \approx \left(\frac{R_1 + R_2}{R_2}\right) V_{EB}$$

Example 1 - Design of a Higher-Voltage Bipolar Voltage Reference

Use the circuit on the previous slide to design a voltage reference having V_{REF} = 2.5V when V_{CC} = 5V. Assume I_s = 1fA and β_F = 100. Evaluate the sensitivity of V_{REF} with respect to V_{CC}.

Solution

Choose I (the current flowing through R) to be 100µA.

Therefore $R = \frac{V_{CC}\text{-}V_{REF}}{100\mu A} = \frac{2.5V}{100\mu A} = 25k\Omega$.

Choose I_1(the current flowing through R_1) to be 50µA. Therefore the current flowing in the emitter is 50µA. The value of $V_{EB} = V_t \ln\left(\frac{50\mu A}{1fA}\right) = 0.638V$.

$\therefore\ R_1 = \frac{0.638V}{50\mu A} = 12.76k\Omega$

With 50µA in the emitter, the base current is approximately 5µA.

Therefore, the current through R_2 is 55µA.

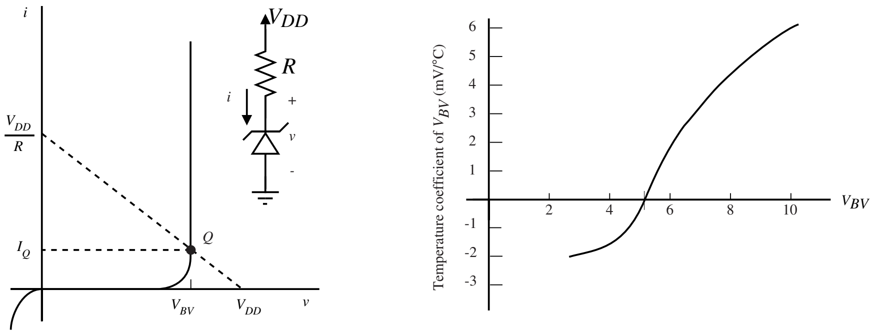
Since $V_{REF} = I_{R2}R_2 + 0.638V = 2.5V$, we get $R_2 = \left(\frac{2.5V\text{-}0.638V}{55\mu A}\right) = 33.85k\Omega$.

The sensitivity of V_{REF} with respect to V_{CC} is

$$S_{V_{CC}}^{V_{REF}} = \left(\frac{R_1 + R_2}{R_1}\right) S_{V_{CC}}^{V_{EB}} = \left(\frac{12.76k\Omega + 33.85k\Omega}{12.76k\Omega}\right)\left(\frac{1}{\ln(I_Q/I_s)}\right) = 3.652(0.0406) = 0.148$$

Breakdown Diode Voltage References

If the power supply voltage is high enough, i.e. $V_{DD} \approx 10\text{V}$, the breakdown diode can be used as a voltage reference.



V-I characteristics of a breakdown diode.

Variation of the temperature coefficient of the breakdown diode as a function of the breakdown voltage, BV.

Fig. 370-05

$$V_{REF} = V_{BV}$$

$$S_{V_{DD}}^{V_{REF}} = \left(\frac{\partial V_{\text{REF}}}{\partial V_{DD}}\right)\left(\frac{V_{DD}}{V_{\text{REF}}}\right) \cong \left(\frac{v_{ref}}{v_{dd}}\right)\left(\frac{V_{DD}}{V_{BV}}\right) = \left(\frac{r_Z}{r_Z + R}\right)\left(\frac{V_{DD}}{V_{BV}}\right)$$

where r_z is the small-signal impedance of the breakdown diode at I_Q (30 to 100Ω).

Typical sensitivities are 0.02 to 0.05.

Note that the temperature dependence could be zero if V_B was a variable.

BOOTSTRAPPED BIAS/REFERENCE CIRCUITS

Bootstrapped Current Source

So far, none of the previous references except the base-emitter and threshold-referenced sources have shown very good independence from power supply. Let us now examine a technique which does achieve the desired independence.

Circuit:

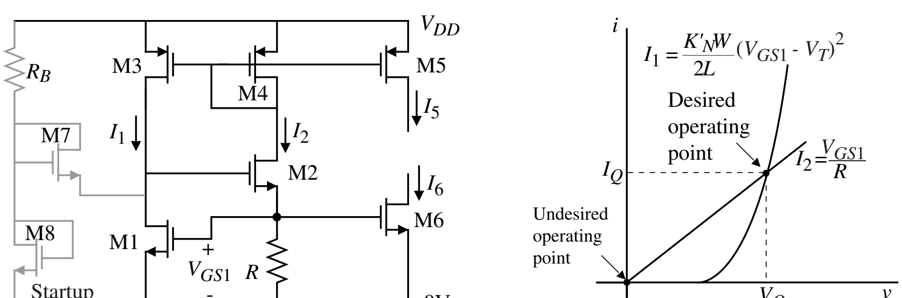


Fig. 370-06

Principle:

If M3 = M4, then $I_1 \approx I_2$. However, the M1-R loop gives $V_{GS1} = V_{T1} + \sqrt{\dfrac{2I_1}{K_N'(W_1/L_1)}}$

Solving these two equations gives $I_2 = \dfrac{V_{GS1}}{R} = \dfrac{V_{T1}}{R} + \left(\dfrac{1}{R}\right)\sqrt{\dfrac{2I_1}{K_N'(W_1/L_1)}}$

The output current, $I_{out} = I_1 = I_2$ can be solved as $I_{out} = \dfrac{V_{T1}}{R} + \dfrac{1}{\beta_1 R^2} + \dfrac{1}{R}\sqrt{\dfrac{2V_{T1}}{\beta_1 R} + \dfrac{1}{(\beta_1 R)^2}}$

Simulation Results for the Bootstrapped Current Source

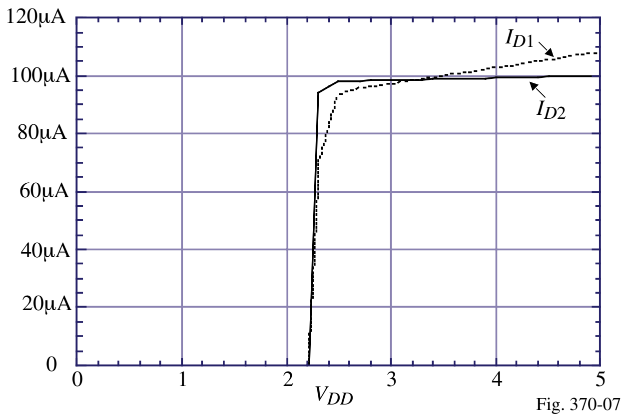


Fig. 370-07

The current I_{D2} appears to be okay, why is I_{D1} increasing?

Apparently, the channel modulation on the current mirror M3-M4 is large.

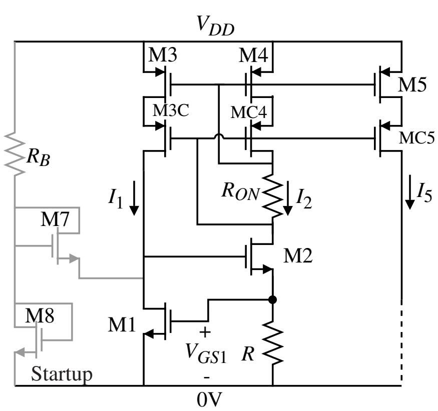
At $V_{DD} = 5V$, $V_{SD3} = 2.83V$ and $V_{SD4} = 1.09V$ which gives $I_{D3} = 1.067 I_{D4} \approx 107\mu A$

Need to cascode the upper current mirror.

SPICE Input File:

```
Simple, Bootstrap Current Reference
VDD 1 0 DC 5.0
VSS 9 0 DC 0.0
M1 5 7 9 9 N W=20U L=1U
M2 3 5 7 9 N W=20U L=1U
M3 5 3 1 1 P W=25U L=1U
M4 3 3 1 1 P W=25U L=1U
M5 9 3 1 1 P W=25U L=1U
R 7 9 10KILOHM
M8 6 6 9 9 N W=1U L=1U
M7 6 6 5 9 N W=20U L=1U
RB 1 6 100KILOHM
.OP
.DC VDD 0 5 0.1
.MODEL N NMOS VTO=0.7 KP=110U
GAMMA=0.4 +PHI=0.7 LAMBDA=0.04
.MODEL P PMOS VTO=-0.7 KP=50U
GAMMA=0.57 +PHI=0.8 LAMBDA=0.05
.PRINT DC ID(M1) ID(M2) ID(M5)
.PROBE
.END
```


Cascoded Bootstrapped Current Source



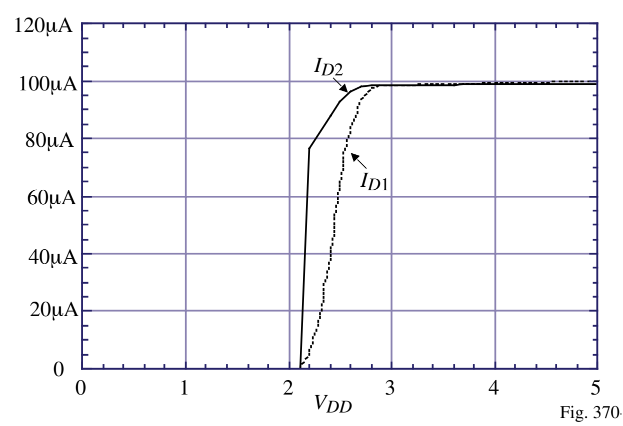


Fig. 370-

SPICE Input File:

```
Cascode, Bootstrap Current Reference
VDD 1 0 DC 5.0
VSS 9 0 DC 0.0
M1 5 7 9 9 N W=20U L=1U
M2 4 5 7 9 N W=20U L=1U
M3 2 3 1 1 P W=25U L=1U
M4 8 3 1 1 P W=25U L=1U
M3C 5 4 2 1 P W=25U L=1U
MC4 3 4 8 1 P W=25U L=1U
RON 3 4 4KILOHM
M5 9 3 1 1 P W=25U L=1U
R 7 9 10KILOHM
M8 6 6 9 9 N W=1U L=1U
M7 6 6 5 9 N W=20U L=1U
RB 1 6 100KILOHM
.OP
.DC VDD 0 5 0.1
.MODEL N NMOS VTO=0.7
KP=110U GAMMA=0.4 PHI=0.7
LAMBDA=0.04
.MODEL P PMOS VTO=-0.7
KP=50U GAMMA=0.57 PHI=0.8
LAMBDA=0.05
.PRINT DC ID(M1) ID(M2) ID(M5)
.PROBE
.END
```

Base-Emitter Referenced Circuit

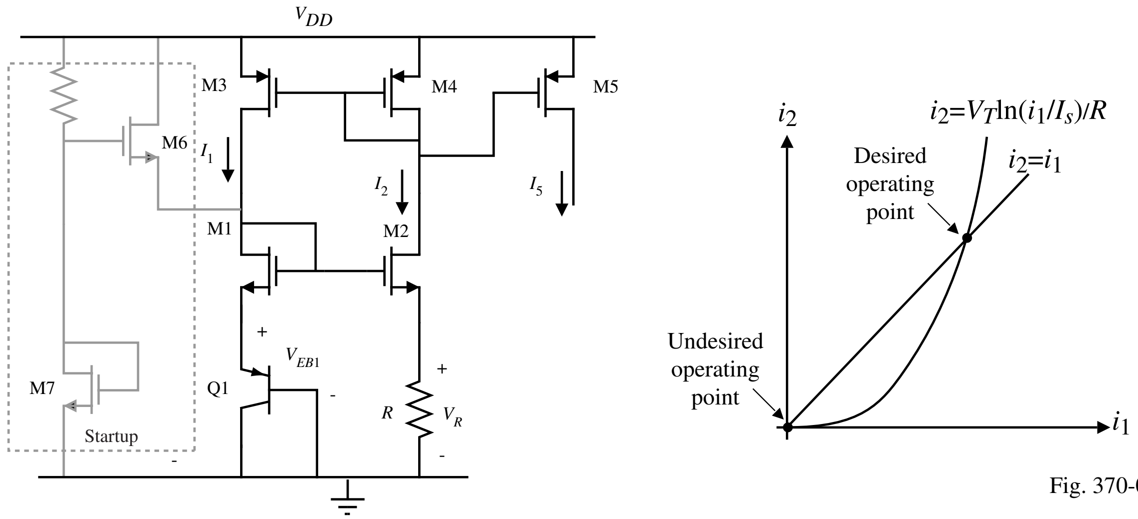


Fig. 370-09

$$I_{out} = I_2 = \frac{V_{EB1}}{R}$$

BJT can be a MOSFET in weak inversion.

Low Voltage Bootstrap MOS Circuit

The previous bootstrap circuits required at least 2 volts across the power supply before operating.

A low-voltage bootstrap circuit:

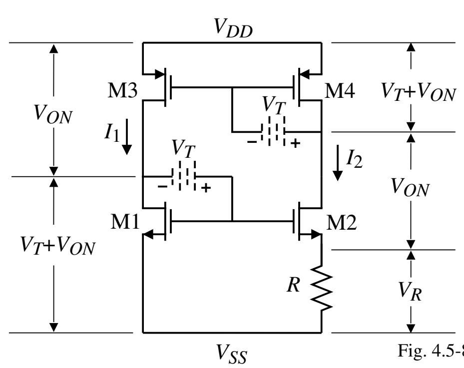


Fig. 4.5-8A

Without the batteries, V_T, the minimum power supply is $V_T+2V_{ON}+V_R$.

With the batteries, V_T, the minimum power supply is $2V_{ON}+V_R \approx 0.5$V

Summary of Power-Supply Independent References

- Reasonably good, simple references are possible
- Best power supply sensitivity is approximately 0.01
  (10% change in power supply causes a 0.1% change in reference)
- Typical simple reference temperature dependence is ≈ 1000 ppm/°C
- Can obtain zero temperature coefficient over a limited range of operation

Type of Reference	$S_{V_{PP}}^{V_{REF}}$
Voltage division	1
MOSFET-R	<1
BJT-R	<<1
Threshold Referenced	<<1
Base-emitter Referenced	<<1

REFERENCES WITH TEMPERATURE INDEPENDENCE

Characterization of Temperature Dependence

The objective is to minimize the fractional temperature coefficient defined as,

$$TC_F = \frac{1}{V_{REF}}\left(\frac{\partial V_{REF}}{\partial T}\right) = \frac{1}{T}\, S_T^{V_{REF}} \text{ parts per million per °C or ppm/°C}$$

Temperature dependence of PN junctions:

$$\left.\begin{array}{l} i \approx I_s \exp\left(\frac{v}{V_t}\right) \\ I_s = KT^3 \exp\left(\frac{-V_{GO}}{V_t}\right) \end{array}\right\} \quad \frac{1}{I_s}\left(\frac{\partial I_s}{\partial T}\right) = \frac{\partial(\ln I_s)}{\partial T} = \frac{3}{T} + \frac{V_{GO}}{TV_t} \approx \frac{V_{GO}}{TV_t}$$

$$\frac{dv_{BE}}{dT} \approx \frac{V_{BE} - V_{GO}}{T} = -2\text{mV/°C at room temperature}$$

(V_{GO} = 1.205 V at room temperature and is called the bandgap voltage)

Temperature dependence of MOSFET in strong inversion:

$$\left.\begin{array}{l} \frac{dv_{GS}}{dT} = \frac{dV_T}{dT} + \sqrt{\frac{2L}{WC_{ox}}}\,\frac{d}{dT}\left(\sqrt{\frac{i_D}{\mu_o}}\right) \\ \mu_o = KT^{-1.5} \\ V_T(T) = V_T(T_o) - \alpha(T-T_o) \end{array}\right\} \quad \frac{dv_{GS}}{dT} \approx -\alpha \approx -2.3\frac{\text{mV}}{\text{°C}}$$

Resistors:

$(1/R)(dR/dT)$ ppm/°C

Bipolar-Resistance Voltage References

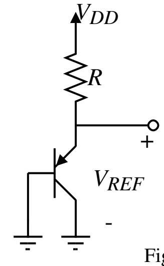


Fig. 380-1

From previous work we know that,

$$V_{REF} = \frac{kT}{q}\ln\left(\frac{V_{DD} - V_{REF}}{RI_s}\right)$$

However, not only is V_{REF} a function of T, but R and I_s are also functions of T.

$$\therefore\quad \frac{dV_{REF}}{dT} = \frac{k}{q}\ln\left(\frac{V_{DD}\text{-}V_{REF}}{RI_s}\right) + \frac{kT}{q}\left(\frac{RI_s}{V_{DD}\text{-}V_{REF}}\right)\left[\frac{-1}{RI_s}\frac{dV_{REF}}{dT} - \left(\frac{V_{DD}\text{-}V_{REF}}{RI_s}\right)\left(\frac{dR}{RdT} + \frac{dI_s}{I_s dT}\right)\right]$$

$$= \frac{V_{REF}}{T} - \frac{V_t}{V_{DD}\text{-}V_{REF}}\frac{dV_{REF}}{dT} - V_t\left(\frac{dR}{RdT} + \frac{dI_s}{I_s dT}\right) = \frac{V_{REF}\text{-}V_{GO}}{T} - \frac{V_t}{V_{DD}\text{-}V_{REF}}\frac{dV_{REF}}{dT} - \frac{3V_t}{T} - \frac{V_t}{R}\frac{dR}{dT}$$

$$\therefore\quad \frac{dV_{REF}}{dT} = \frac{\frac{V_{REF}\text{-}V_{GO}}{T} - V_t\frac{dR}{RdT} - \frac{3V_t}{T}}{1 + \frac{V_t}{V_{DD}\text{-}V_{REF}}} \approx \frac{V_{REF}\text{-}V_{GO}}{T} - V_t\frac{dR}{RdT} - \frac{3V_t}{T}$$

$$TC_F = \frac{1}{V_{REF}}\frac{dV_{REF}}{dT} = \frac{V_{REF}\text{-}V_{GO}}{V_{REF}\cdot T} - \frac{V_t}{V_{REF}}\frac{dR}{RdT} - \frac{3V_t}{V_{REF}\cdot T}$$

If $V_{REF} = 0.6$V, $V_t = 0.026$V, and the R is polysilicon, then at 27°K the TC_F is

$$TC_F = \frac{0.6\text{-}1.205}{0.6\cdot 300} - \frac{0.026\cdot 0.0015}{0.6} - \frac{3\cdot 0.026}{0.6\cdot 300} = 33110^{-6}\text{-}65\text{x}10^{-6}\text{-}433\text{x}10^{-6} = \text{-}3859\text{ppm/°C}$$

MOSFET Resistor Voltage Reference

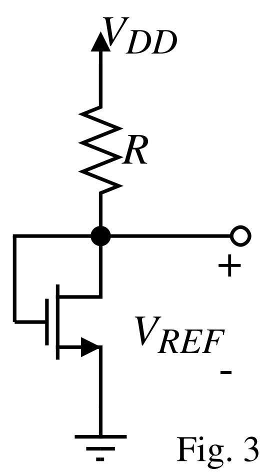


Fig. 380-02

From previous results we know that

$$V_{REF} = V_{GS} = V_T + \sqrt{\frac{2(V_{DD}\text{-}V_{REF})}{\beta R}}$$

or $$V_{REF} = V_T - \frac{1}{\beta R} + \sqrt{\frac{2(V_{DD}\text{-}V_T)}{\beta R} + \frac{1}{(\beta R)^2}}$$

Note that V_{REF}, V_T, β, and R are all functions of temperature.

It can be shown that the TC_F of this reference is

$$\frac{dV_{\text{REF}}}{dT} = \frac{-\alpha + \sqrt{\frac{V_{DD} - V_{\text{REF}}}{2\beta R}}\left(\frac{1.5}{T} - \frac{1}{R}\frac{dR}{dT}\right)}{1 + \frac{1}{\sqrt{2\beta R\,(V_{DD} - V_{\text{REF}})}}}$$

$$\therefore\quad TC_F = \frac{-\alpha + \sqrt{\frac{V_{DD} - V_{\text{REF}}}{2\beta R}}\left(\frac{1.5}{T} - \frac{1}{R}\frac{dR}{dT}\right)}{V_{REF}\left(1 + \frac{1}{\sqrt{2\beta R\,(V_{DD} - V_{\text{REF}})}}\right)}$$

Example 4.5-1 - Calculation of MOSFET-Resistor Voltage Reference TC_F

Calculate the temperature coefficient of the MOSFET-Resistor voltage reference where W/L=2, V_{DD}=5V, R=100kΩ using the parameters of Table 3.1-2. The resistor, R, is polysilicon and has a temperature coefficient of 1500 ppm/°C.

Solution

First, calculate V_{REF}. Note that $\beta R = 220 \times 10^{-6} \times 10^5 = 22$ and $\frac{dR}{RdT} = 1500\text{ppm/°C}$

$$\therefore \quad V_{REF} = 0.7 - \frac{1}{22} + \sqrt{\frac{2(5-0.7)}{22} + \left(\frac{1}{22}\right)^2} = 1.281\text{V}$$

$$\text{Now,} \quad \frac{dV_{REF}}{dT} = \frac{-2.3\times10^{-3} + \sqrt{\frac{5-1.281}{2(22)}}\left(\frac{1.5}{300} - 1500\times10^{-6}\right)}{1 + \frac{1}{\sqrt{2(22)\ (5-1.281)}}} = -1.189\text{x}10^{-3}\text{V/°C}$$

The fractional temperature coefficient is given by

$$TC_F = -1.189 \times 10^{-3}\left(\frac{1}{1.281}\right) = -928\text{ ppm/°C}$$

Bootstrapped Current Source/Sink

Gate-source referenced source:

The output current was given as, $I_{out} = \frac{V_{T1}}{R} + \frac{1}{\beta_1 R^2} + \frac{1}{R}\sqrt{\frac{2V_{T1}}{\beta_1 R} + \frac{1}{(\beta_1 R)^2}}$

Although we could grind out the derivative of I_{out} with respect to T, the temperature performance of this circuit is not that good to spend the time to do so. Therefore, let us assume that $V_{GS1} \approx V_{T1}$ which gives

$$I_{out} \approx \frac{V_{T1}}{R} \quad \Rightarrow \quad \frac{dI_{out}}{dT} = \frac{1}{R}\frac{dV_{T1}}{dT} - \frac{1}{R^2}\frac{dR}{dT}$$

In the resistor is polysilicon, then

$$TC_F = \frac{1}{I_{out}}\frac{dI_{out}}{dT} = \frac{1}{V_{T1}}\frac{dV_{T1}}{dT} - \frac{1}{R}\frac{dR}{dT} = \frac{-\alpha}{V_{T1}} - \frac{1}{R}\frac{dR}{dT} = \frac{-2.3\text{x}10^{-3}}{0.7} - 1.5\text{x}10^{-3} = -4786\text{ppm/°C}$$

Base-emitter referenced source:

The output current was given as, $I_{out} = I_2 = \frac{V_{BE1}}{R}$

The $TC_F = \frac{1}{V_{BE1}}\frac{dV_{BE1}}{dT} - \frac{1}{R}\frac{dR}{dT}$

If V_{BE1} = 0.6V and R is poly, then the $TC_F = \frac{1}{0.6}(-2\text{x}10^{-3}) - 1.5\text{x}10^{-3} = -4833\text{ppm/°C}$.

Technique to Make g_m Dependent on a Resistor

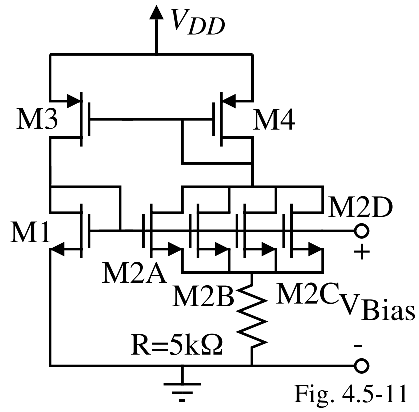


Fig. 4.5-11

Consider the following circuit with all transistors having a W/L = 10. This is a bootstrapped reference which creates a V_{bias} independent of V_{DD}. The two key equations are:

$$I_3 = I_4 \Rightarrow I_1 = I_2$$

and

$$V_{GS1} = V_{GS2} + I_2R$$

Solving for I_2 gives:

$$I_2 = \frac{V_{GS1}-V_{GS2}}{R} = \frac{1}{R}\left(\sqrt{\frac{2I_1}{\beta_1}} - \sqrt{\frac{2I_2}{\beta_2}}\right) = \frac{\sqrt{2I_1}}{R\sqrt{\beta_1}}\left(1 - \frac{1}{2}\right)$$

$$\therefore \quad \sqrt{I_2} = \frac{1}{R\sqrt{2\beta_1}} \quad \Rightarrow I_2 = I_1 = \frac{1}{2\beta_1 R^2} = \frac{1}{2\cdot 110\text{x}10^{-6}\cdot 10\cdot 25\text{x}10^6} = 18.18\mu\text{A}$$

Now, V_{bias} can be written as

$$V_{bias}=V_{GS1}=\sqrt{\frac{2I_2}{\beta_1}}+V_{TN} = \frac{1}{\beta_1 R}+V_{TN} = \frac{1}{110\text{x}10^{-6}\cdot 10\cdot 5\text{x}10^3} + 0.7 = 0.1818+0.7 = 0.8818\text{V}$$

Any transistor with $V_{GS} = V_{bias}$ will have a current flow that is given by $1/2\beta R^2$.

Therefore,
$$g_m = \sqrt{2I\beta} = \sqrt{\frac{2\beta}{2\beta R^2}} = \frac{1}{R} \quad \Rightarrow \quad \boxed{g_m = \frac{1}{R}}$$

(This means that the temperature dependence of g_m will be that of 1/R which can be used to achieve temperature controlled performance.)

Summary of Reference Performance

Type of Reference	$S_{V_{DD}}^{V_{REF}}$	TC_F	Comments
MOSFET-R	<1	>1000ppm/°C	
BJT-R	<<1	>1000ppm/°C	
Breakdown Diode	<<1	Can be very small	BV too large
Bootstrap Gate-Source Referenced	Good if currents are matched	>1000ppm/°C	Requires start-up circuit
Bootstrap Base-emitter Referenced	Good if currents are matched	>1000ppm/°C	Requires start-up circuit

- A MOSFET can have zero temperature dependence of i_D for a certain v_{GS}
- If one is careful, very good independence of power supply can be achieved
- None of the above references have really good temperature independence

Consider the following example:

A 10 bit ADC has a reference voltage of 1V. The LSB is approximately 0.001V. Therefore, the voltage reference must be stable to within 0.1%. If a 100°C change in temperature is experienced, then the TC_F must be 0.001%/C or multiplying by 10^4 gives a TC_F = 10ppm/°C.

SECTION 4.6 - BANDGAP REFERENCES

Temperature Stable References

• The previous reference circuits failed to provide small values of temperature coefficient although sufficient power supply independence was achieved.

• This lecture introduces the bandgap voltage concept combined with power supply independence to create a very stable voltage reference in regard to both temperature and power supply variations.

Bandgap Voltage Reference Principle

The principle of the bandgap voltage reference is to balance the negative temperature coefficient of a pn junction with the positive temperature coefficient of the thermal voltage, $V_t = kT/q$.

Concept:

Result: References with TC_F's approaching 10 ppm/°C.

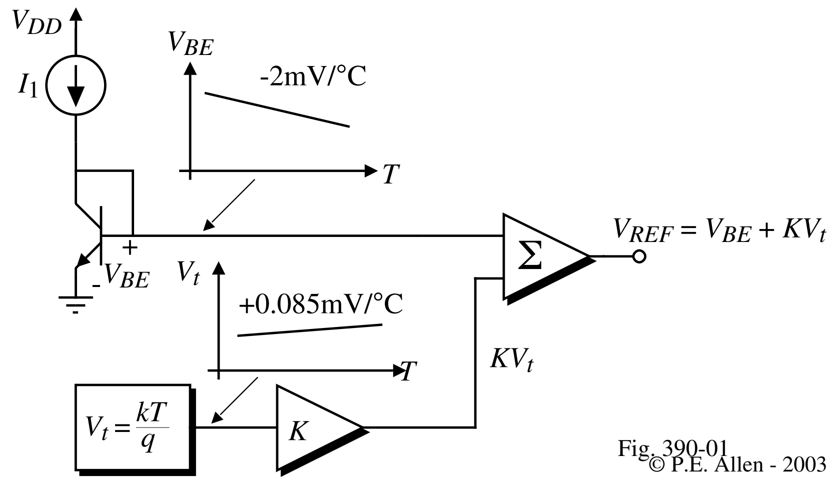


Fig. 390-01

Derivation of the Temperature Coefficient of the Base-Emitter Voltage

For small TC_F's the dependence V_{BE} must be known more precisely than ≈ -2mV/°C.

1.) Start with the collector current density, J_C:

$$J_C = \frac{q\overline{D_n}\, n_{po}}{W_B} \exp\left(\frac{V_{BE}}{V_t}\right)$$

where, $J_C = I_C/\text{Area}$ = collector current density

$\overline{D_n}$ = average diffusion constant for electrons

W_B = base width

V_{BE} = base-emitter voltage

$V_t = kT/q$

k = Boltzmann's constant ($1.38\text{x}10^{-23}$J/°K)

T = Absolute temperature

$n_{po} = n_i^2/N_A$ = equilibrium concentration of electrons in the base

$n_i^2 = DT^3 \exp\left(\frac{-V_{GO}}{V_t}\right)$ = intrinsic concentration of carriers

D = temperature independent constant

V_{GO} = bandgap voltage of silicon (1.205V)

N_A = acceptor impurity concentration

Derivation of the Temperature Coefficient of the Base-Emitter Voltage - Continued

2.) Combine the above relationships into one:

$$J_C = \frac{q\overline{D_n}}{N_A W_B} DT^3 \exp\left(\frac{V_{BE} - V_{GO}}{V_t}\right) = AT^\gamma \exp\left(\frac{V_{BE} - V_{GO}}{V_t}\right) \qquad \text{where,} \quad \gamma = 3$$

3.) The value of J_C at a reference temperature of $T = T_0$ is

$$J_{C0} = AT_0^\gamma \exp\left[\frac{q}{kT_0}(V_{BE} - V_{GO})\right]$$

while the value of J_C at a general temperature, T, is

$$J_C = AT^\gamma \exp\left[\frac{q}{kT}(V_{BE} - V_{GO})\right]$$

4.) The ratio of J_C/J_{C0} can be expressed as,

$$\frac{J_C}{J_{C0}} = \left(\frac{T}{T_0}\right)^\gamma \exp\left[\frac{q}{k}\left(\frac{V_{BE} - V_{G0}}{T} - \frac{V_{BE0} - V_{G0}}{T_0}\right)\right]$$

or

$$\ln\left(\frac{J_C}{J_{C0}}\right) = \gamma \ln\left(\frac{T}{T_0}\right) + \frac{q}{kT}\left[V_{BE} - V_{GO} - \frac{T}{T_0}(V_{BE0} - V_{GO})\right]$$

where V_{BE0} is the value of V_{BE} at $T = T_0$.

5.) Solving for V_{BE} from the above results gives,

$$\boxed{V_{BE}(T) = V_{GO}\left(1 - \frac{T}{T_0}\right) + V_{BE0}\left(\frac{T}{T_0}\right) + \frac{\gamma kT}{q} \ln\left(\frac{T_0}{T}\right) + \frac{kT}{q}\ln\left(\frac{J_C}{J_{C0}}\right)}$$

Derivation of the Temperature Coefficient of the Base-Emitter Voltage - Continued

6.) Next, assume $J_C \propto T^\alpha$ and find $\partial V_{BE}/\partial T$.

$$\frac{\partial V_{BE}}{\partial T} = \frac{\partial V_{GO}}{\partial T}\left(1 - \frac{T}{T_0}\right) - \frac{V_{GO}}{T_0} + \frac{V_{BE0}}{T_0} + \frac{\gamma kT}{q}\frac{\partial \ln(T_0/T)}{\partial T} + \ln\left(\frac{T_0}{T}\right)\frac{\partial(\gamma kT/q)}{\partial T} + \frac{kT}{q}\left(\frac{\partial \ln\left(\frac{J_C}{J_{C0}}\right)}{\partial T}\right) + \frac{k}{q}\ln\left(\frac{J_C}{J_{C0}}\right)$$

7.) Assume that $T = T_0$ which means $J_C = J_{C0}$. Since, $\partial V_{GO}/\partial T = 0$,

$$\left.\frac{\partial V_{BE}}{\partial T}\right|_{T=T_0} = -\frac{V_{GO}}{T_0} + \frac{V_{BE0}}{T_0} + \frac{\gamma kT}{q}\cdot\frac{\partial \ln(T_0/T)}{\partial T} + \frac{kT}{q}\left(\frac{\partial \ln(J_C/J_{C0})}{\partial T}\right)$$

8.) Note that,

$$\frac{\partial \ln(T_0/T)}{\partial T} = \frac{T}{T_0}\frac{\partial(T_0/T)}{\partial T} = \frac{T}{T_0}\left(\frac{-T_0}{T^2}\right) = \frac{-1}{T} \quad \text{and} \quad \frac{\partial \ln(J_C/J_{C0})}{\partial T} = \frac{J_{C0}}{J_C}\frac{\partial(J_C/J_{C0})}{\partial T} = \frac{J_{C0}}{J_C}\left(\frac{\alpha}{T}\frac{J_C}{J_{C0}}\right) = \frac{\alpha}{T}$$

Therefore,

$$\left.\frac{\partial V_{BE}}{\partial T}\right|_{T=T_0} = -\frac{V_{GO}}{T_0} + \frac{V_{BE0}}{T_0} - \frac{\gamma k}{q} + \frac{\alpha k}{q} \quad \text{or} \quad \boxed{\left.\frac{\partial V_{BE}}{\partial T}\right|_{T=T_0} = \frac{V_{BE0} - V_{GO}}{T_0} + (\alpha - \gamma)\left(\frac{k}{q}\right)}$$

Typical values of α and γ are 1 and 3.2. If $V_{BE0} = 0.6$V, then at room temperature:

$$\left.\frac{\partial V_{BE}}{\partial T}\right|_{T=T_0} = \frac{0.6-1.205}{300} + (1-3.2)\left(\frac{0.026}{300}\right) = \frac{0.6-1.205-0.1092}{300} = -1.826\text{mV/°C}$$

Derivation of the Temperature Coefficient of the Thermal Voltage (*kT/q*)

1.) Consider two identical pn junctions having different current densities,

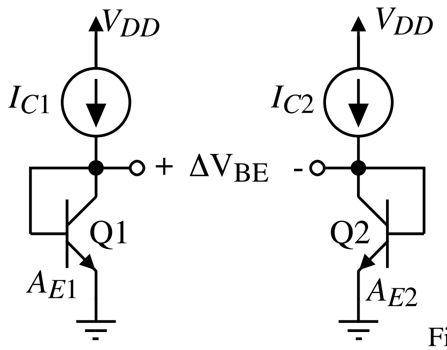


Fig. 390-02

$$\Delta V_{BE} = V_{BE1} - V_{BE2} = \frac{kT}{q}\ln\left(\frac{J_{C1}}{J_{C2}}\right)$$

- Find $\partial(\Delta V_{BE})/\partial T$,

$$\boxed{\frac{\partial(\Delta V_{BE})}{\partial T} = \frac{V_t}{T}\ln\left(\frac{J_{C1}}{J_{C2}}\right) = \frac{k}{q}\ln\left(\frac{J_{C1}}{J_{C2}}\right)}$$

Derivation of the Gain, K, for the Bandgap Voltage Reference

1.) In order to achieve a zero temperature coefficient at $T = T_0$, the following equation must be satisfied:

$$0 = \frac{\partial V_{BE}}{\partial T}\Big|_{T=T_0} + K''\frac{\partial(\Delta V_{BE})}{\partial T}$$ where K" is a constant that satisfies the equation.

2.) Therefore, we get

$$0 = K''\left(\frac{V_{t0}}{T_0}\right)\ln\left(\frac{J_{C1}}{J_{C2}}\right) + \frac{V_{BE0} - V_{GO}}{T_0} + \frac{(\alpha - \gamma)V_{t0}}{T_0}$$

3.) Define $K = K''\ln\left(\frac{J_{C1}}{J_{C2}}\right)$, therefore

$$0 = K\left(\frac{V_{t0}}{T_0}\right) + \frac{V_{BE0} - V_{GO}}{T_0} + \frac{(\alpha - \gamma)V_{t0}}{T_0}$$

4.) Solving for K gives $$\boxed{K = \frac{V_{GO} - V_{BE0} - V_{t0}(\alpha-\gamma)}{V_{t0}}}$$

Assuming that $J_{C1}/J_{C2} = A_{E1}/A_{E2} = 10$ and $V_{BE0} = 0.6V$ gives,

$$K = \frac{1.205 - 0.6 + (2.2)(0.026)}{0.026} = 25.469$$

5.) The output voltage of the bandgap voltage reference is found as,

$$V_{REF}\Big|_{T=T_0} = V_{BE0} + KV_{t0} = V_{BE0} + V_{GO} - V_{BE0} + (\gamma-\alpha)V_{t0}$$ or $$\boxed{V_{REF} = V_{GO} + (\gamma-\alpha)V_{t0}}$$

For the previous values, $V_{REF} = 1.205 + 0.026(2.2) = 1.262V$.

Variation of the Bandgap Reference Voltage with respect to Temperature

The previous derivation is only valid at a given temperature, T_0. As the temperature changes away from T_0, the value of $\partial V_{REF}/\partial T$ is no longer zero.

Illustration:

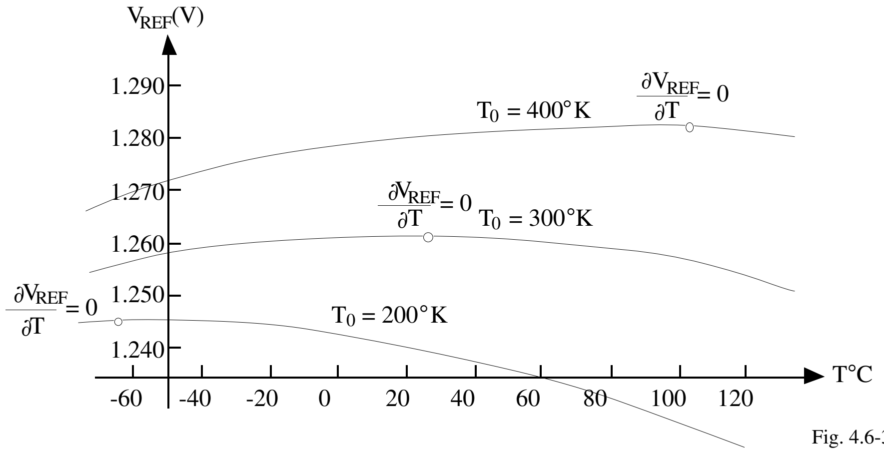


Fig. 4.6-3

Bandgap curvature correction will be necessary for low ppm/C bandgap references.

Classical Widlar Bandgap Voltage Reference†

Operation:

$$V_{BE1} = V_{BE2} + I_2R_3$$

gives

$$\Delta V_{BE} = V_{BE1} - V_{BE2} = I_2R_3$$

But,

$$\Delta V_{BE} = V_t \ln\left(\frac{I_1}{I_{s1}}\right) - V_t \ln\left(\frac{I_2}{I_{s2}}\right) = V_t \ln\left(\frac{I_1 I_{s2}}{I_2 I_{s1}}\right)$$

Assume $V_{BE1} \approx V_{BE3}$, we get $\quad I_1R_1 = I_2R_2$

Therefore,

$$I_2 = \frac{\Delta V_{BE}}{R_3} = \frac{V_t}{R_3} \ln\left(\frac{I_1 I_{s2}}{I_2 I_{s1}}\right) = \frac{V_t}{R_3} \ln\left(\frac{R_2 I_{s2}}{R_1 I_{s1}}\right)$$

Now we can express V_{REF} as

$$V_{REF} = I_2R_2 + V_{BE3} = \frac{R_2}{R_3} V_t \ln\left(\frac{R_2 I_{s2}}{R_1 I_{s1}}\right) + V_{BE3} = KV_t + V_{BE}$$

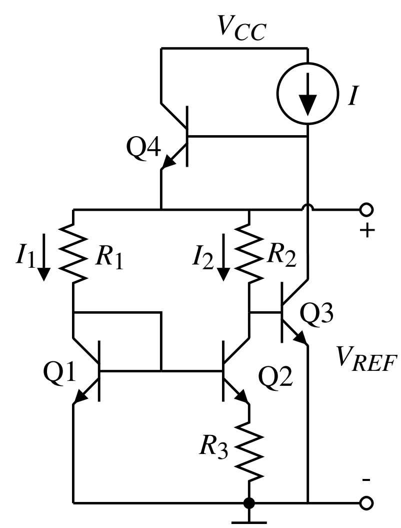


Fig. 390-04

Design R_1, R_2, I_{s1}, and I_{s2} to get the desired K.

Let K = 25 and $I_{s2} = 10I_{s1}$ and design R_1, R_2, and R_3. Choose $R_2 = 10R_1 = 10\text{k}\Omega$. Therefore, ln(100) = 4.602. Therefore $R_2/R_3 = 25/4.602$ or $R_3 = R_2/5.4287 = 1.842\text{k}\Omega$.

† R.J. Widlar, "New Developments in IC Voltage Regulators," *IEEE J. of Solid-State Circuits*, Vol. SC-6, pp. 2-7, February 1971.

A CMOS Bandgap Reference using PNP Lateral BJTs

Bootstrapped Voltage Reference using PNP Laterals-

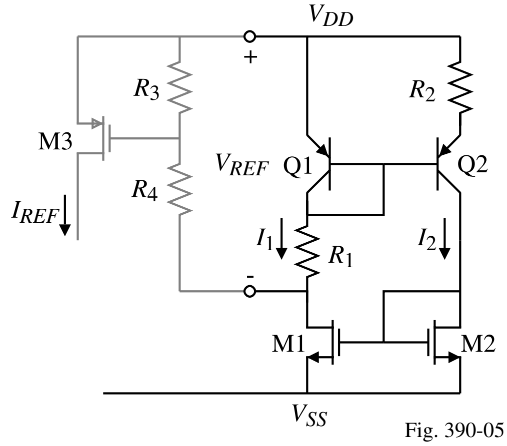


Fig. 390-05

$$I_2 = \frac{V_{BE1} - V_{BE2}}{R_2} = \frac{V_t}{R_2}\left[\ln\left(\frac{I_1}{I_{s1}}\right) - \ln\left(\frac{I_2}{I_{s2}}\right)\right] = \frac{V_t}{R_2}\ \ln\left(\frac{I_{s2}}{I_{s1}}\right) = \frac{V_t}{R_2}\ \ln\left(\frac{A_{E2}}{A_{E1}}\right)$$

if $I_1 = I_2$ which is forced by the current mirror consisting of M1 and M2.

$$\therefore \quad V_{REF} = V_{BE1} + I_1R_1 = V_{BE1} + \left(\frac{R_1}{R_2}\ln\left(\frac{A_{E2}}{A_{E1}}\right)\right) V_t = V_{BE1} + KV_t$$

While an op amp could be used to make $I_1 = I_2$ it suffers from offset and noise and leads to deterioration of the bandgap temperature performance.

V_{REF} is with respect to V_{DD} and therefore is susceptible to changes on V_{DD}.

A CMOS Bandgap Reference using Substrate PNP BJTs

Operation:

The cascode mirror (M5-M8) keeps the currents in Q1, Q2, and Q3 identical.

Thus,

$$V_{BE1} = I_2R + V_{BE2}$$

or

$$I_2 = \frac{V_t}{R}\ \ln(n)$$

Therefore,

$$V_{REF} = V_{BE3} + I_2(kR) = V_{BE3} + kV_t{\cdot}\ln(n)$$

Use k and n to design the desired value of K (n is an integer greater than 1).

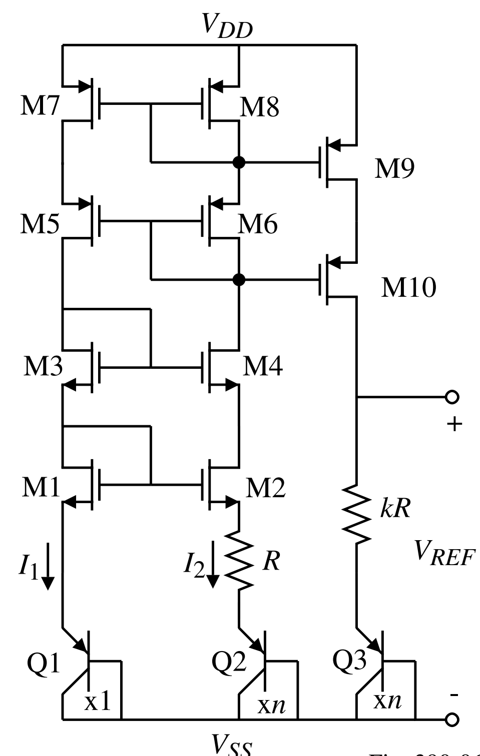


Fig. 390-06

Weak Inversion Bandgap Voltage Reference

Circuit:

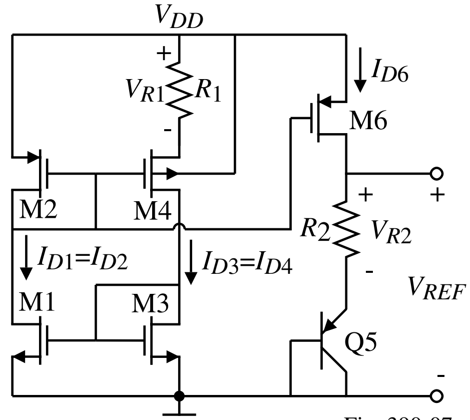


Fig. 390-07

Analysis:

For the p-channel transistors:

$$I_D = I_{DO}(W/L)\exp\left(\frac{V_{BG}}{nV_t}\right)\left[\exp\left(\frac{-V_{BS}}{V_t}\right) - \exp\left(\frac{-V_{BD}}{V_t}\right)\right]$$

where $V_t = kT/q$.

If $V_{BD} >> V_t$, then $I_D = I_{DO}(W/L)\exp\left(\frac{V_{BG}}{nV_t} - \frac{V_{BS}}{V_t}\right)$.

The various transistor currents can be expressed as:

$$I_{D1} = I_{D2} = I_{DO}(W_2/L_2)\exp\left(\frac{V_{BG2}}{nV_t}\right) \text{ and } I_{D3} = I_{D4} = I_{DO}(W_4/L_4)\exp\left(\frac{V_{BG4}}{nV_t} - \frac{V_{BS4}}{V_t}\right)$$

Note that $V_{BG2} = V_{BG4}$ and $V_{BS4} = V_{R1}$.

Therefore,

$$\frac{I_{D1}}{I_{D3}} = \frac{W_2/L_2}{W_4/L_4}\exp\left(\frac{V_{R1}}{V_t}\right)$$

where

$$V_{R1} = V_t \ln\left(\frac{W_1W_4L_2L_3}{L_1L_4W_2W_3}\right) \quad \text{and} \quad I_{R1} = \frac{V_{R1}}{R_1}$$

Weak Inversion Bandgap Voltage Reference - Continued

The reference voltage can be expressed as,

$$V_{REF} = R_2I_6 + V_{BE5}$$

However,

$$I_6 = \frac{W_6L_3}{L_6W_3} I_{R1} = \frac{W_6L_3}{L_6W_3}\frac{V_t}{R_1} \ln\left(\frac{W_1W_4L_2L_3}{L_1L_4W_2W_3}\right) .$$

Substituting I_6 and the previously derived expression for $V_{BE}(T)$ in V_{REF} gives,

$$V_{REF} = \frac{W_6L_3}{L_6W_3}\frac{R_2}{R_1}V_t\ln\left(\frac{W_1W_4L_2L_3}{L_1L_4W_2W_3}\right) + V_{GO}\left(1 - \frac{T}{T_0}\right) + V_{BE0}\left(\frac{T}{T_0}\right) + 3V_t\ln\left(\frac{T_0}{T}\right)$$

To achieve $\partial V_{REF}/\partial T = 0$ at $T = T_0$, we get

$$\frac{\partial V_{REF}}{\partial T} = \left(\frac{k}{q}\right)\left(\frac{R_2}{R_1}\right)\left(\frac{W_6L_3}{L_6W_3}\right)\ln\left(\frac{W_1W_4L_2L_3}{L_1L_4W_2W_3}\right) - \frac{V_{GO}}{T_0} + \frac{V_{BE0}}{T_0} + \frac{3k}{q}$$

Therefore,

$$\boxed{\frac{R_2W_6L_3}{R_1L_6W_3}\ln\left(\frac{W_1W_4L_2L_3}{L_1L_4W_2W_3}\right) = \frac{q}{kT_0}(V_{GO} - V_{BE0}) - 3}$$

Under the above constraint, V_{REF} has an ≈ zero TC_F at $T = T_0$ and has a value of

$$V_{REF} = V_{GO} + \frac{3kT}{q}\left[1 + \ln\left(\frac{T_0}{T}\right)\right] = V_{GO} + \frac{3kT}{q}$$

Practical values of $\partial V_{REF}/\partial T$ for the weak inversion bandgap are less than 100 ppm/°C.

Curvature Correction Techniques:

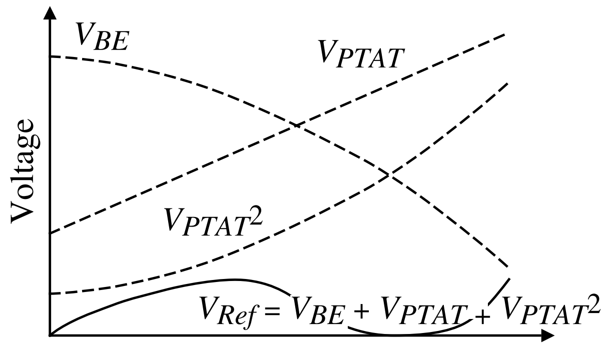


Fig. 400-01

- Squared PTAT Correction:
  Temperature coefficient ≈ 1-20 ppm/°C

- V_{BE} loop
  M. Gunaway, *et. al.*, “A Curvature-Corrected Low-Voltage Bandgap Reference,” *IEEE Journal of Solid-State Circuits*, vol. 28, no. 6, pp. 667-670, June 1993.

- β compensation
  I. Lee *et. al.*, “Exponential Curvature-Compensated BiCMOS Bandgap References,” *IEEE Journal of Solid-State Circuits*, vol. 29, no. 11, pp. 1396-1403, Nov. 1994.

- Nonlinear cancellation
  G.M. Meijer *et. al.*, “A New Curvature-Corrected Bandgap Reference,” *IEEE Journal of Solid-State Circuits*, vol. 17, no. 6, pp. 1139-1143, December 1982.

V_{BE} Loop Curvature Correction Technique

Circuit:

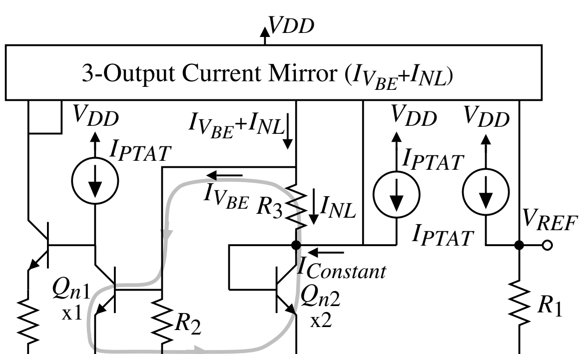


Fig. 400-02

Operation:

$$I_{NL} = \frac{V_{BE1}\text{-}V_{BE2}}{R_3} = \frac{V_t}{R_3}\,ln\left(\frac{I_{c1}A_2}{A_1 I_{c2}}\right)$$

$$= \frac{V_t}{R_3}\,ln\left(\frac{2I_{PTAT}}{I_{NL}+I_{Constant}}\right)$$

where

$$I_{constant} = I_{NL} + I_{PTAT} + I_{VBE}$$

$$\approx I_{NL} + \frac{V_t}{R_x} + \frac{V_{BE}}{R_2}$$

(a quasi-temperature independent current subject to the TC_F of the resistors)
where

$V_t = kT/q$

I_{c1} and I_{c2} are the collector currents of Q_{n1} and Q_{n2}, respectively

R_x = a resistor used to define I_{PTAT}

$$\therefore \quad V_{REF} = \left[\frac{V_{BE}}{R_2} + \frac{V_t}{R_3}\,ln\left(\frac{2I_{PTAT}}{I_{NL}+I_{constant}}\right) + I_{PTAT}\right] R_1$$

Temperature coefficient ≈ 3 ppm/°C with a total quiescent current of 95µA..

β Compensation Curvature Correction Technique

Circuit:

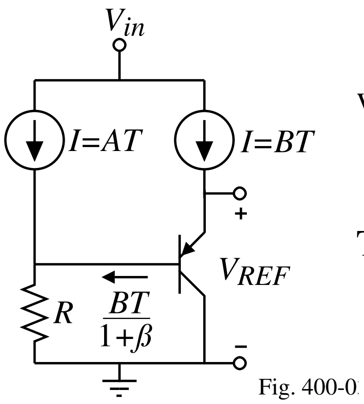


Fig. 400-0

Operation:

$$V_{REF} = V_{BE} + \left(AT + \frac{BT}{(1+\beta)}\right) R \approx V_{BE} + \left(AT + \frac{BT}{\beta}\right) R$$

where

A and B are constant

T = temperature

The temperature dependence of β is

$$\beta(T) \propto e^{-1/T} \Rightarrow \beta(T) = Ce^{-1/T}$$

$$\therefore \; V_{REF} = V_{BE}(T) + \left(AT + \frac{BTe^{1/T}}{C}\right)$$

Not good for small values of V_{in}.

$$V_{in} \geq V_{REF} + V_{sat.} = V_{GO} + V_{sat.} = 1.4\text{V}$$

Nonlinear Cancellation Curvature Correction Technique

Objective: Eliminate nonlinear term from the BE.

Result: 0.5 ppm/°C from -25°C to 85°C.

Operation: From above,

$$V_{REF} = V_{PTAT} + 4V_{BE}(I_{PTAT}) - 3V_{BE}(I_{Constant})$$

Note that, $I_{PTAT} \Rightarrow I_c \propto T^1 \quad \Rightarrow \alpha = 1$

and $I_{constant} \Rightarrow I_c \propto T^0 \quad \Rightarrow \alpha = 0,$

Previously we found,

$$V_{BE}(T) \approx V_{GO} - \frac{T}{T_0}[V_{GO}-V_{BE}(T_0)] - (\gamma - \alpha)V_t \ln\left(\frac{T}{T_0}\right)$$

so that

$$V_{BE}(I_{PTAT}) = V_{GO} - \frac{T}{T_0}[V_{GO}-V_{BE}(T_0)] - (\gamma-1)V_t \ln\left(\frac{T}{T_0}\right)$$

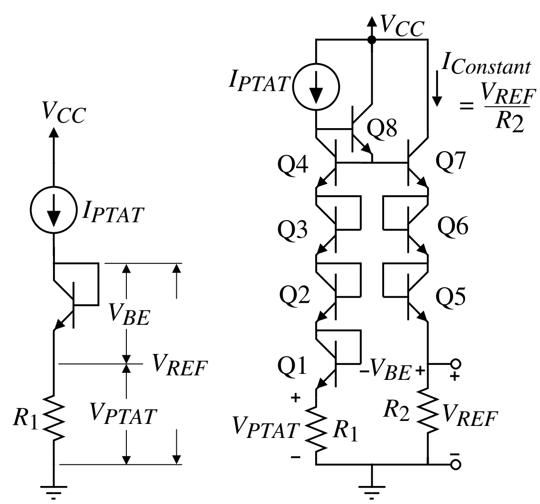
and

$$V_{BE}(I_{Constant}) = V_{GO} - \frac{T}{T_0}[V_{GO} - V_{BE}(T_0)] - \gamma V_t \ln\left(\frac{T}{T_0}\right)$$

Combining the above relationships gives,

$$V_{REF}(T) = V_{PTAT} + V_{GO} - (T/T_0)[V_{GO} - V_{BE}(T_0)] - [\gamma - 4]\, V_t \ln((T/T_0))$$

If $\gamma \approx 4$, then $\boxed{V_{REF}(T) \approx V_{PTAT} + V_{GO}(1 - (T/T_0)) + V_{BE}(T_0)(T/T_0)}$



Conventional Bandgap Reference

Curvature Corrected Bandgap Reference

Fig. 400-04

Other Characteristics of Bandgap Voltage References

Noise

Voltage references for high-resolution ADCs are particularly sensitive to noise.
Noise sources: Op amp, resistors, switches, etc.

PSRR

Maximize the PSRR of the op amp.

Offset Voltages

Becomes a problem when op amps are used.

$V_{BE2} = V_{BE1} + V_{R1} + V_{OS}$

$\Delta V_{BE} = V_{BE2} - V_{BE1} = V_{R1} + V_{OS} = V_t \ln\left(\frac{i_{C2}A_{E1}}{i_{C1}A_{E2}}\right)$

Since $i_{C2}R_3 = i_{C1}R_2 - V_{OS}$

then $\frac{i_{C2}}{i_{C1}} = \frac{R_2}{R_3} - \frac{V_{OS}}{i_{C1}R_3} = \frac{R_2}{R_3}\left(1 + \frac{V_{OS}}{i_{C1}R_2}\right)$

Therefore,

$$V_{R1} = -V_{OS} + V_t \ln\left[\frac{R_2A_{E1}}{R_3A_{E2}}\left(1 + \frac{V_{OS}}{i_{C1}R_2}\right)\right]$$

$$V_{REF} = V_{BE2} - V_{OS} + i_{C1}R_2 = V_{BE2} - V_{OS} + \left(\frac{V_{R1}}{R_1}\right)R_2 = V_{BE2} - V_{OS} + \left(\frac{R_2}{R_1}\right)V_{R1}$$

$$\boxed{V_{REF} = V_{BE2} - V_{OS}\left(1+\frac{R_2}{R_1}\right) + \frac{R_2}{R_1} V_t \ln\left[\frac{R_2A_{E1}}{R_3A_{E2}}\left(1 - \frac{V_{OS}}{i_{C1}R_2}\right)\right]}$$

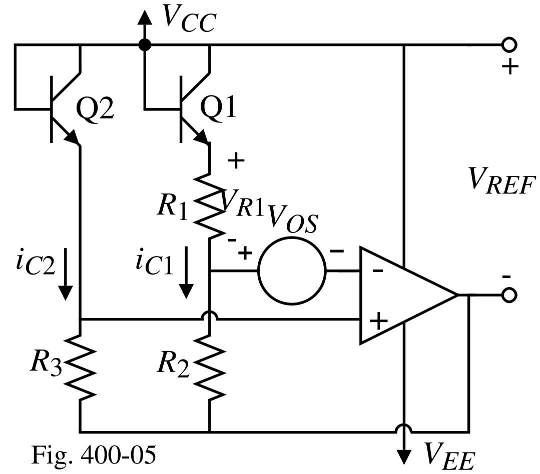


Fig. 400-05

How do you get a Stable Reference Current from the Bandgap?

Assume that a temperature stable reference voltage is available (i.e. bandgap reference) and use the zero TC NMOS current sink.

The problem is that V_{REF} may not be equal to the value of V_{GS} that gives zero TC.

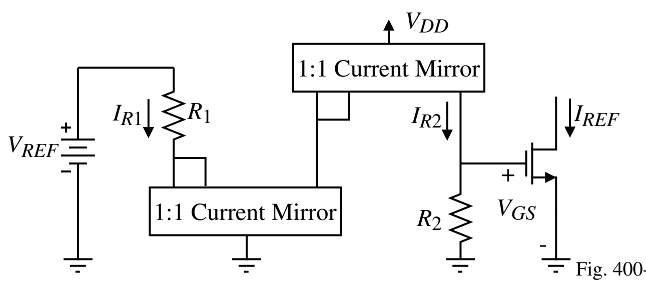


Fig. 400-06

$$V_{GS} = I_{R2}R_2 = R_2\left(\frac{V_{REF}}{R_1}\right) = \left(\frac{R_2}{R_1}\right)V_{REF}$$

$$\therefore \quad \frac{dV_{GS}}{dT} = \left(\frac{R_2}{R_1}\right)\frac{dV_{REF}}{dT} + \frac{V_{REF}}{R_1}\frac{dR_2}{dT} - \frac{R_2}{R_1^2}\frac{dR_1}{dT} = \frac{R_2}{R_1}\left[\frac{dV_{REF}}{dT} + \frac{dR_2}{dT} - \frac{dR_1}{dT}\right]$$

If the temperature coefficients of R_1 and R_2 are equal $\left(\frac{dR_1}{dT} = \frac{dR_2}{dT}\right)$, then

$\frac{dV_{GS}}{dT} = \frac{R_2}{R_1}\frac{dV_{REF}}{dT}$ and V_{GS} is proportional to the temperature dependence of V_{REF}.

If the MOSFET is biased at the zero TC point, then the current should have the same dependence on temperature as V_{REF}.

Practical Aspects of Temperature-Independent and Supply-Independent Biasing

A temperature-independent and supply-independent current source and its distribution:

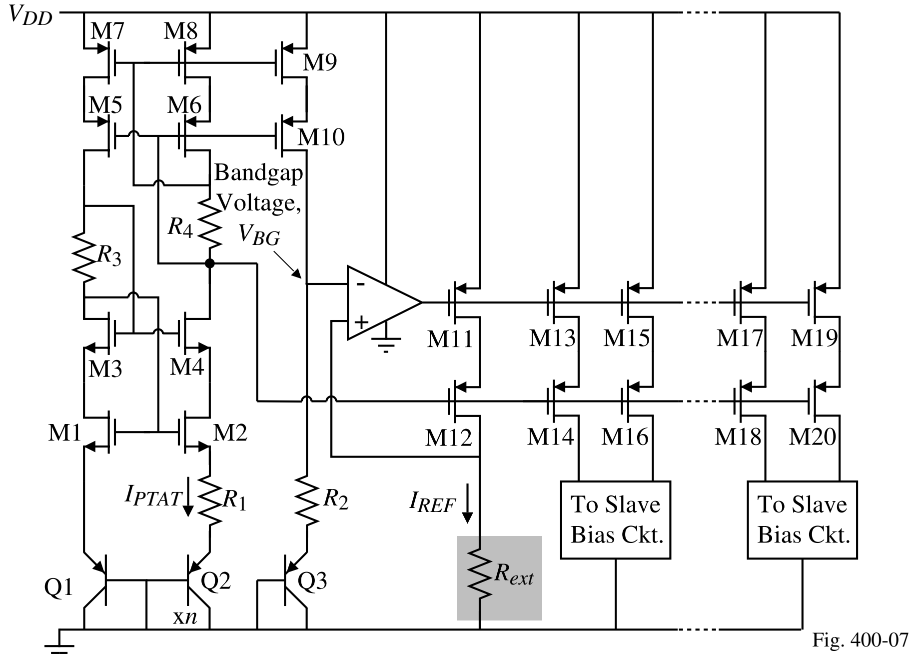


Fig. 400-07

Constant current:

$$I_{REF} = \frac{V_{BG}}{R_{ext}} \quad \text{where} \quad V_{BG} = V_{BE3} + I_{PTAT}R_2 = V_{BE3} + \frac{V_T}{R_1}\, ln(n)\cdot R_2$$

Practical Aspects of Bias Distribution Circuits - Continued

Distribution of the current avoids change in bias voltage due to *IR* drop in bias lines.

Slave bias circuit:

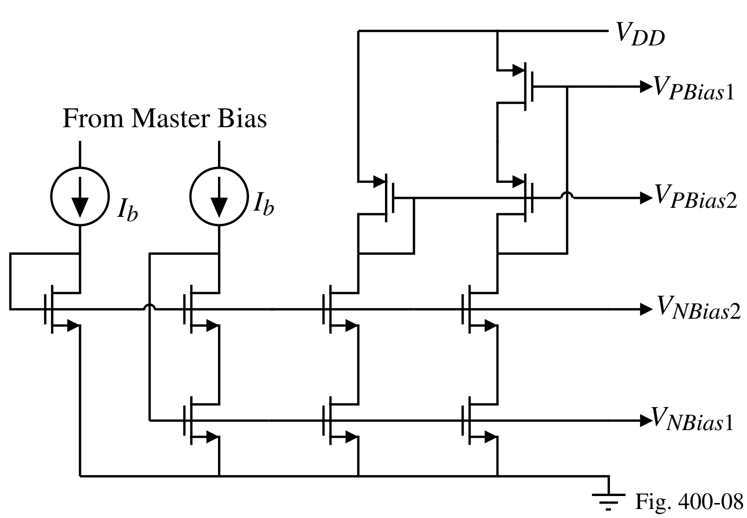


Fig. 400-08

SUMMARY OF VOLTAGE AND CURRENT REFERENCES

- Reasonably good, simple references are possible
- Best power supply sensitivity is approximately 0.01
  (10% change in power supply causes a 0.1% change in reference)
- Typical simple reference temperature dependence is ≈ 1000 ppm/°C
- Can obtain zero temperature coefficient over a limited range of operation
- Bandgap voltage references can achieve temperature dependence less than 50 ppm/°C
- Correction of second-order effects in the bandgap voltage reference can achieve very stable (1 ppm/°C) voltage references.
- Watch out for second-order effects such as noise when using the bandgap voltage reference in sensitive applications.

We will examine bandgap voltage references once again when we consider low voltage circuits in Section 6 of Chapter 7.

CHAPTER 4 - SUMMARY

- This chapter covered the analysis and design of sub-blocks or subcircuits including:
  - Switches
  - MOS diode and floating resistor realizations
  - Current sinks and sources
  - Current mirrors (amplifiers)
  - Current and voltage references
  - Bandgap reference
- Subcircuits represent primitives of circuit design and do not stand alone
- The current sink/source is an important subcircuit which is used for biases and ac loads
- A current sink/source is characterized by
  1.) The independence of the current on the voltage across it (r_{out})
  2.) The voltage range over which the current is not independent of the voltage (V_{MIN})
- A current mirror is characterized by
  1.) The independence of the output current on the voltage across it ($r_{out} \rightarrow$ large)
  2.) The output voltage range over which output current is dependent (V_{MIN}(out))
  3.) The independence of the input voltage on the input current ($r_{in} \rightarrow$ small)
  4.) The range of input voltage over which the input current is independent (V_{MIN}(in))
  5.) The accuracy of the current out as a function of the current in ratio.
- A voltage or current reference is independent of power supply and temperature
- The bandgap reference is the best realization of a voltage reference