LECTURE 170 – TEMPERATURE STABLE REFERENCES

LECTURE ORGANIZATION

Outline

- Principles of temperature stable references
- Examples of temperature stable references
- Design of bias voltages for a chip
- Summary

***CMOS Analog Circuit Design*, 2nd Edition Reference**

Pages 153-159

PRINCIPLES OF TEMPERATURE STABLE REFERENCES

Temperature Stable References

- The previous reference circuits failed to provide small values of temperature coefficient although sufficient power supply independence was achieved.
- This section introduces a temperature stable reference that cancels a positive temperature coefficient with a negative temperature coefficient. The technique is sometimes called the *bandgap reference* although it has nothing to do with the bandgap voltage.

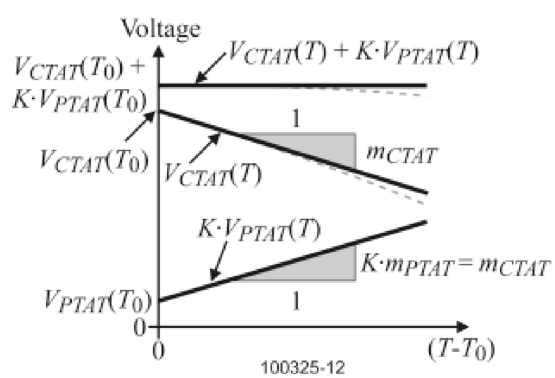
Principle

$$V_{REF}(T) = V_{PTAT}(T) + K\cdot V_{CTAT}(T)$$

where

$V_{PTAT}(T)$ is a voltage that is *proportional to absolute temperature* (PTAT)

$V_{CTAT}(T)$ is a voltage that is *complimentary to absolute temperature* (CTAT)



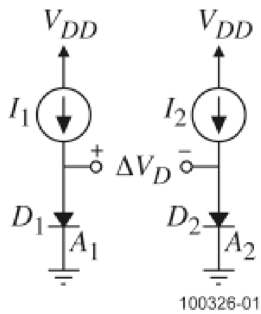
and

K is a temperature independent constant that makes $V_{REF}(T)$ independent of temperature

PTAT Voltage

The principle illustrated on the last slide requires perfectly linear positive and negative temperature coefficients to work properly. We will now show a technique of generating PTAT voltages that are linear with respect to temperature.

Implementation of a PTAT voltage:



$$V_{PTAT} = \Delta V_D = V_{D1} - V_{D2} = V_t\, ln\left(\frac{I_1}{I_{s1}}\right) - V_t\, ln\left(\frac{I_2}{I_{s2}}\right)$$

$$= V_t\, ln\left(\frac{I_1}{I_2}\frac{I_{s2}}{I_{s1}}\right) = V_t\, ln\left(\frac{I_{s2}}{I_{s1}}\right) = V_t\, ln\left(\frac{A_2}{A_1}\right) = \frac{kT}{q}\; ln\left(\frac{A_2}{A_1}\right)$$

if $I_1 = I_2$.

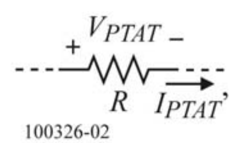
Therefore, if $A_2 = 10A_1$, ΔV_D at room temperature becomes,

$$\Delta V_D = \left[\frac{k}{q}\, ln\left(\frac{A_2}{A_1}\right)\right] T = \left[\frac{1.381\text{x}10^{-23}\text{J/°K}}{1.6\text{x}10^{-19}\text{ Coul}}\, ln(10)\right] T = (+\,0.086\text{mV/°C})T$$

$$\therefore \quad V_{PTAT} = V_t\, ln\left(\frac{A_2}{A_1}\right)$$

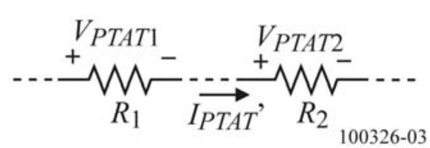
Psuedo-PTAT Currents

In developing temperature independent voltages, it is useful to show how to generate PTAT currents. A straight-forward method is to superimpose V_{PTAT} across a resistor as shown:



Because R is always dependent on temperature, this current is called a *pseudo-PTAT current* and is designated by I_{PTAT}'.

When a pseudo-PTAT current flows through a second resistor with the same temperature characteristics as the first, it creates a new V_{PTAT} voltage.



The new V_{PTAT} voltage, V_{PTAT2} is equal to,

$$V_{PTAT2} = \frac{R_2}{R_1} V_{PTAT1}$$

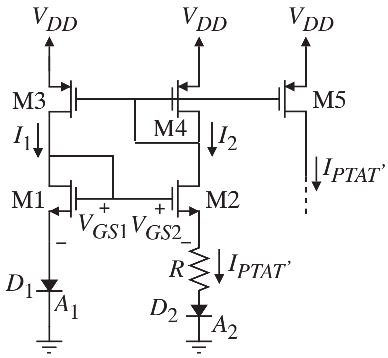
Differentiating with respect to temperature gives

$$\frac{dV_{PTAT2}}{dT} = \frac{R_2}{R_1}\left(\frac{dR_2}{R_2 dT} - \frac{dR_1}{R_1 dT}\right) + \frac{dV_{PTAT1}}{dT}$$

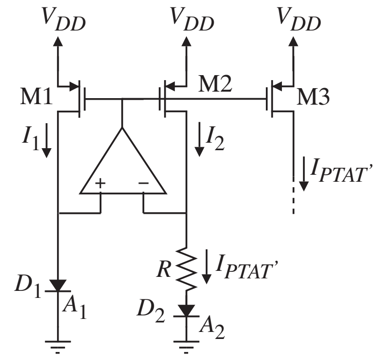
Therefore, if the temperature coefficient of R_1 and R_2 are equal, then the temperature dependence of V_{PTAT2} is the same as V_{PTAT1}.

Pseudo-PTAT Currents - Continued

This can be done through the circuits below which use only MOSFETs and *pn* junctions or MOSFETs, an op amp and *pn* junctions.



Psuedo-PTAT current generator using only MOSFETs and pn junctions.



Psuedo-PTAT current generator using MOSFETs, an op amp and pn junctions.

100326-04

In these circuits, $I_1 = I_2$ and the voltage across D_1 is made equal to the voltage across the series combination of R and D_2 to create the pseudo-PTAT current,

$$I_{PTAT}' = \frac{V_{D1} - V_{D2}}{R} = \frac{kT}{Rq}\ ln\left(\frac{A_2}{A_1}\right)$$

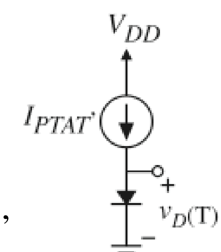
where $V_{GS1} = V_{GS2}$ for the MOSFET only version.

CTAT Voltage

This becomes more challenging because a true CTAT voltage does not exist. The best approach is to examine the *pn* junction (can be a diode or BJT).

The current through a *pn* junction shown can be written as,

$$J_D = \left[\frac{qD_n n_i^2}{L_n N_A} + \frac{qD_p p_{no}}{L_p}\right]\left(\frac{(v_D - V_{G0})}{V_t}\right) = AT^{\gamma}\exp\left(\frac{v_D - V_{G0}}{V_t}\right)$$



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Consider the circuit shown. It can be shown, that $v_D(T)$ can be given as,

$$v_D(T) = V_{GO}\left(1 - \frac{T}{T_0}\right) + v_{D0}\left(\frac{T}{T_0}\right) + \frac{\gamma kT}{q}\ \ln\left(\frac{T_0}{T}\right) + \frac{kT}{q}\ln\left(\frac{J_D}{J_{D0}}\right)$$

where,

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

T_0 = a reference temperature about which T varies

γ = a temperature coefficient for the *pn* junction saturation current ($\gamma \approx 3$)

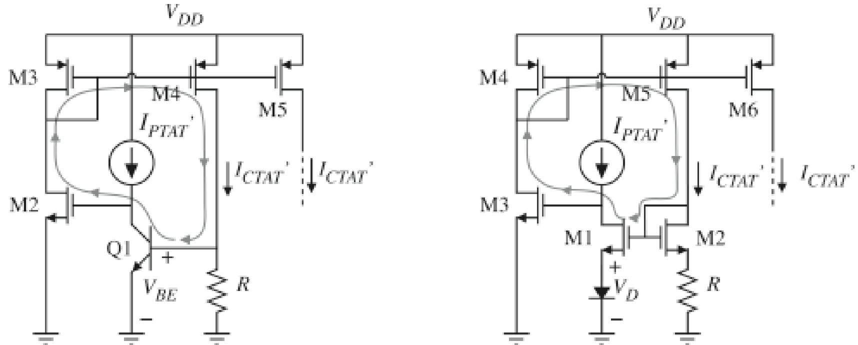
J_D = pn junction current density

In the above expression for $v_D(T)$ the term $\frac{\gamma kT}{q}\ln\left(\frac{T_0}{T}\right)$ is not linear with T!!

This term will create a problem called “bandgap curvature problem” because a perfectly linear PTAT function cannot be cancelled by a term that is not truly CTAT.

Pseudo CTAT Currents

The circuits below show two ways of creating a pseudo CTAT current using negative feedback:†



Generation of a pseudo CTAT current using a bipolar transistor.
Generation of a pseudo CTAT current using a diode.

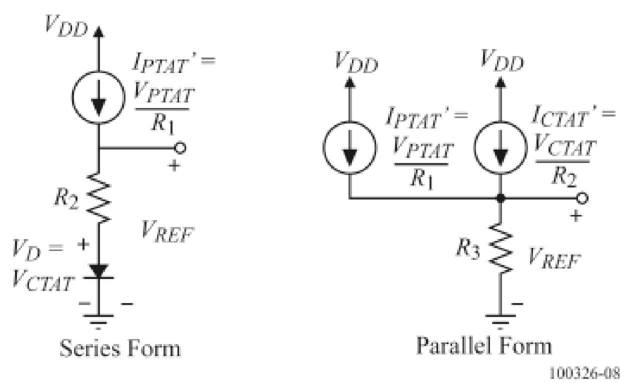
The negative feedback loop shown causes the current designated as I_{CTAT}' to be,

$$I_{CTAT}' = \frac{V_{BE}}{R} = \frac{V_D}{R}$$

† I.M. Gunawan, G.C.M. Jeijer, J. Fonderie, and J.H. Huijsing, "A Curvature-Corrected Low-Voltage Bandgap Reference, *IEEE J. Solid-state Circuits*, vol. SC-28, No. 6, June 1993, pp. 677-670.

Temperature Independent Voltage References

Basic structures:



Series Form
Parallel Form

Series form:

$$V_{REF} = I_{PTAT}'R_2 + V_D = \left(\frac{R_2}{R_1}\right)V_{PTAT} + V_{CTAT}$$

Parallel form:

$$V_{REF} = (I_{PTAT}' + I_{CTAT}')R_3 + V_D = \left(\frac{R_3}{R_1}\right)V_{PTAT} + \left(\frac{R_3}{R_2}\right)V_{CTAT} = \left(\frac{R_3}{R_2}\right)\left[\left(\frac{R_2}{R_1}\right)V_{PTAT} + V_{CTAT}\right]$$

To achieve temperature independence, V_{REF} must be differentiated with respect to temperature and set equal to zero. The resistor ratios and other parameters can be used to achieve temperature independence.

Conditions for Temperature Independence

Differentiating either the series or parallel form with respect to temperature and equating to zero gives,

$$K = \frac{R_2}{R_1} = -\frac{dV_{CTAT}/dT}{dV_{PTAT}/dT}$$

The slopes of V_{CTAT} and V_{PTAT} at a given temperature, T_0, are:

$$m_{CTAT} = \frac{dV_{CTAT}}{dT}\Big|_{T=T_0} = \frac{V_D - V_{GO}}{T_0} + (\alpha-\gamma)\left(\frac{k}{q}\right) = \frac{V_{CTAT} - V_{GO}}{T_0} + (\alpha-\gamma)\left(\frac{V_{t0}}{T_0}\right)$$

where α = temperature dependence of J_D [$J_D(T) \propto T^{\alpha}$, where $\alpha = 1$ for PTAT current flowing through the pn junction]

and

$$m_{PTAT} = \frac{dV_{PTAT}}{dT}\Big|_{T=T_0} = \frac{k}{q}\,ln\left(\frac{J_{D2}}{J_{D1}}\right) = \frac{k}{q}\,ln\left(\frac{A_2}{A_1}\right) = \frac{V_{t0}}{T_0}\,ln\left(\frac{A_2}{A_1}\right) = \frac{V_{PTAT}}{T_0}$$

Therefore, the temperature independent constant multiplying V_{PTAT} is

$$\text{Temp. independent constant} = K = \frac{R_2}{R_1} = \frac{V_{GO} - V_{CTAT} + (\gamma-\alpha)V_{t0}}{V_{PTAT}}$$

Therefore,

$$V_{REF} = V_{GO} - V_{CTAT} + (\gamma-\alpha)V_{t0} + V_{CTAT} = V_{GO} + (\gamma-\alpha)V_{t0} \approx 1.205\text{V}+0.057 = 1.262\text{V}$$

Example 170-1 – Temp. Independent Constant for Series and Parallel References

(a.) Design the ratio of R_2/R_1 for the series configuration if $V_{CTAT} = 0.6$V and $A_2/A_1 = 10$ for room temperature ($V_t = 0.026$V). Assume $\gamma = 3.2$ and $\alpha = 1$. Find the value of V_{REF}.

$$\frac{R_2}{R_1} = \frac{V_{GO} - V_{CTAT} + (\gamma-\alpha)V_{t0}}{V_{PTAT}} = \frac{1.205 - 0.6 + 2.2(0.026)}{0.026(2.3026)} = 11.06$$

$$V_{REF} = 1.205 + 2.2(0.026) = 1.262\text{V}$$

(b.) For the parallel configuration find the values of R_2/R_1 and R_3/R_2 if $V_{REF} = 0.5$V.

From (a.) we know that $R_2/R_1 = 11.05$. We also know that,

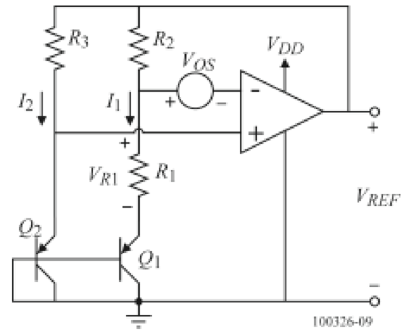
$$V_{REF} = \left(\frac{R_3}{R_1}\right)V_{PTAT} + \left(\frac{R_3}{R_2}\right)V_{CTAT} = \left(\frac{R_3}{R_2}\right)\left[\left(\frac{R_2}{R_1}\right)V_{PTAT} + V_{CTAT}\right]$$

$$= (R_3/R_2)[11.05\,ln(10)(0.026) + 0.6] = (R_3/R_2)1.262 = 0.5$$

$\therefore \quad (R_3/R_2) = 0.3963$

If $R_1 = 1\text{k}\Omega$, then $R_2 = 11.05\text{k}\Omega$ and $R_3 = 4.378\text{k}\Omega$

A Series Temperature Independent Voltage Reference



An early realization of the series form is shown below[†]:

Assuming $V_{OS} = 0$, then V_{R1} is

$$V_{R1} = V_{EB2} - V_{EB1} = V_t\, ln\left(\frac{J_2}{J_{s2}}\right) - V_t\, ln\left(\frac{J_1}{J_{s1}}\right)$$

$$= V_t\, ln\left(\frac{I_2 A_{E1}}{I_1 A_{E2}}\right) = V_t\, ln\left(\frac{R_2 A_{E1}}{R_1 A_{E2}}\right)$$

The op amp forces the relationship $I_1 R_2 = I_2 R_3$

$$\therefore V_{REF} = V_{EB2} + I_2 R_3 = V_{EB2} + V_{R1}\left(\frac{R_2}{R_1}\right) = V_{EB2} + \left(\frac{R_2}{R_1}\right) V_t ln\left(\frac{R_2 A_{E1}}{R_1 A_{E2}}\right) = V_{CTAT} + \left(\frac{R_2}{R_1}\right) ln\left(\frac{R_2 A_{E1}}{R_1 A_{E2}}\right) V_t$$

Differentiating the above with respect to temperature and setting the result to zero, gives

$$\left(\frac{R_2}{R_1}\right) ln\left(\frac{R_2 A_{E1}}{R_1 A_{E2}}\right) = \frac{V_{GO} - V_{CTAT} + (\gamma\text{-}\alpha)V_{t0}}{V_t}$$

If $V_{OS} \neq 0$, then V_{REF} becomes,

$$V_{REF} = V_{EB2} - \left(1 + \frac{R_2}{R_1}\right) V_{OS} + \frac{R_2}{R_1} V_t\, ln\left[\frac{R_2 A_{E1}}{R_1 A_{E2}}\left(1 - \frac{V_{OS}}{I_1 R_2}\right)\right]$$

† K.E. Kujik, "A Precision Reference Voltage Source," *IEEE Journal of Solid-State Circuits*, Vol. SC-8, No. 3 (June 1973) pp. 222-226.

Example 170-2 – Design of the Previous Temperature Independent Reference

Assume that A_{E1} = 10 A_{E2}, V_{EB2} = 0.7 V, $R_2 = R_3$, and V_t = 0.026 V at room temperature for temperature independent reference on the previous slide. Find R_2/R_1 to give a zero temperature coefficient at room temperature. If V_{OS} = 10 mV, find the change in V_{REF}. Note that $I_1 R_2 = V_{REF} - V_{EB2} - V_{OS}$.

Evaluating the temperature independent constant gives

$$\left(\frac{R_2}{R_1}\right) \ln\left(\frac{R_2\, A_{E1}}{R_3 A_{E2}}\right) = \frac{V_{GO} - V_{CTAT} + (\gamma\text{-}\alpha)V_{t0}}{V_{PTAT}} = \frac{1.205 - 0.7 + (2.2)(0.026)}{0.026} = 21.62$$

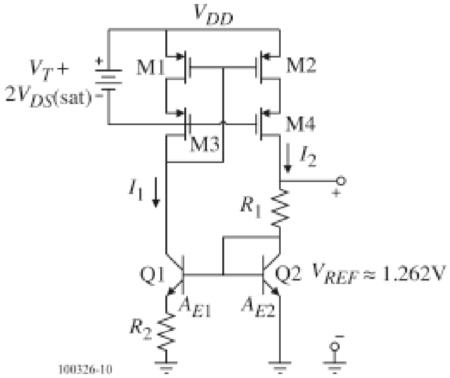
Therefore, R_2/R_1 = 9.39. In order to use the equation for V_{REF} with $V_{OS} \neq 0$, we must know the approximate value of V_{REF} and iterate if necessary because I_1 is a function of V_{REF}. Assuming V_{REF} to be 1.262, we obtain from

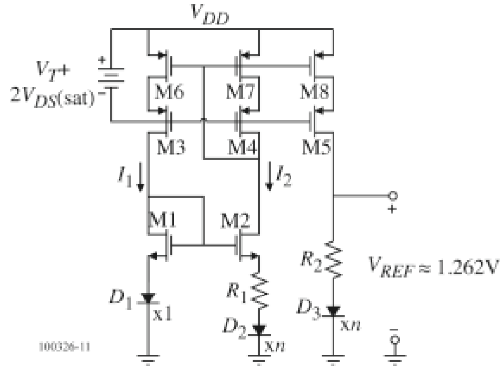
$$V_{REF} = V_{EB2} - \left(1 + \frac{R_2}{R_1}\right) V_{OS} + \frac{R_2}{R_1} V_t\, ln\left[\frac{R_2 A_{E1}}{R_1 A_{E2}}\left(1 - \frac{V_{OS}}{V_{REF} - V_{EB2} - V_{OS}}\right)\right]$$

a new value V_{REF} = 1.153 V. The second iteration makes little difference on the result because V_{REF} is in the argument of the logarithm

Series Temperature Independent Voltage References

The references shown do not use an op amp and avoid the issues with offset voltage and PSRR.





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

$$= \frac{V_t}{R_2}\ln\left(\frac{I_{s1}}{I_{s2}}\right) = \frac{V_t}{R_2}\ln\left(\frac{A_{E1}}{A_{E2}}\right)$$

Since $I_1 = I_2$, $V_{REF} = V_{BE2} + I_1R_1 = V_{BE1} + \left(\frac{R_1}{R_2}\ln\left(\frac{A_{E1}}{A_{E2}}\right)\right)V_t$

$$= V_{CTAT} + \left(\frac{R_1}{R_2}\ln\left(\frac{A_{E1}}{A_{E2}}\right)\right)V_{PTAT}$$

$$V_{D1} = I_2R_1 + V_{D2}$$

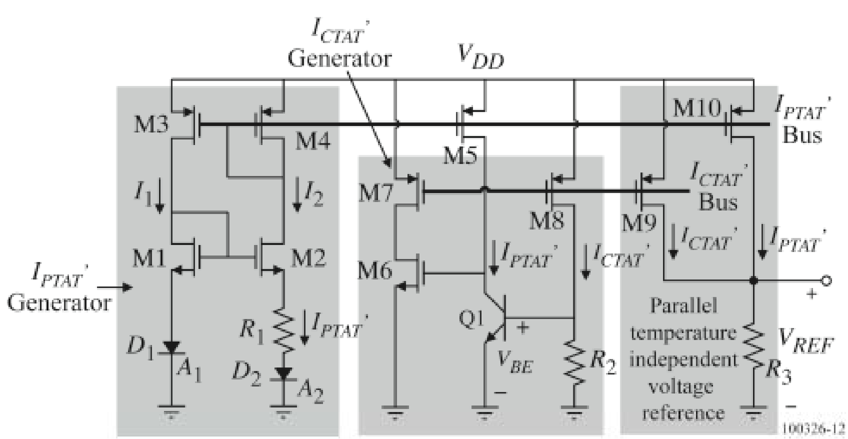
$$I_3 = I_2 = I_{PTAT}' = \frac{V_t}{R}\,ln(n)$$

$$V_{REF} = V_{D3} + I_3(kR) = V_{D3} + kV_t\ln(n)$$

$$= V_{CTAT} + k\ln(n)V_{PTAT}$$

Parallel Temperature Independent Voltage Reference

A parallel form of the temperature independent voltage reference is shown below:



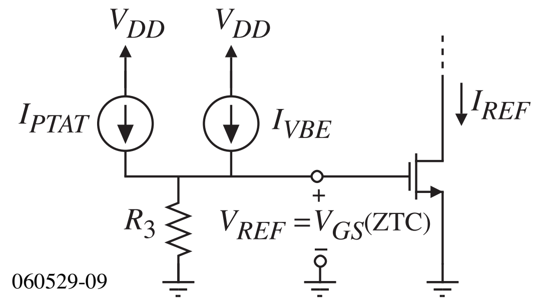
$$V_{REF} = \left(\frac{R_3}{R_1}\right)V_{PTAT} + \left(\frac{R_3}{R_2}\right)V_{CTAT}$$

Comments:

- The BJT of the I_{CTAT}' generator can be replaced with an MOSFET-diode equivalent
- Any value of V_{REF} can be achieved
- Part (b.) of Example 170-1 showed how to design the resistors of this implementation

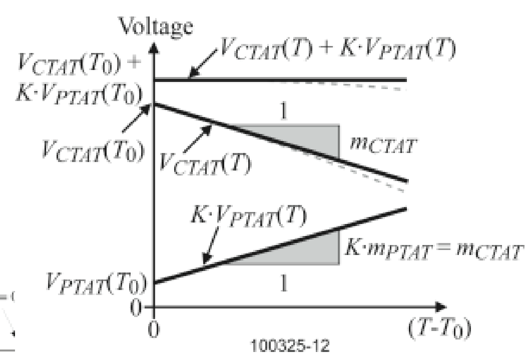
How Can a Bandgap "Current" Reference be Obtained?

Use a MOSFET under ZTC operation and design the parallel form of the bandgap voltage reference to give a value of V_{ZTC}.

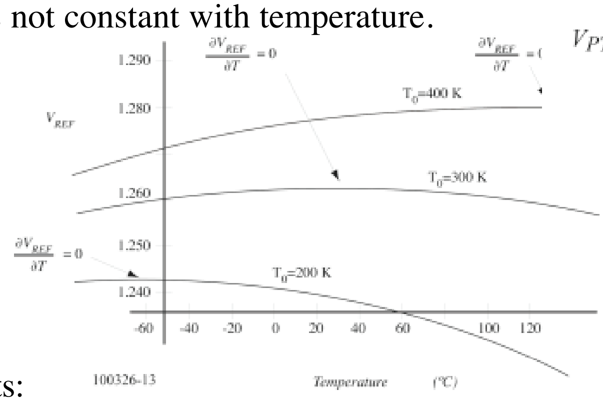


Bandgap Curvature Problem

Unfortunately, the $\frac{\gamma kT}{q}\ln\left(\frac{T_0}{T}\right)$ term of the *pn* junction contributed a nonlinearity to the CTAT realization. This is illustrated by the dashed lines in the plot shown.



The result is shown below where the reference voltage is not constant with temperature.



Comments:

- True temperature independence is only achieved over a small range of temperatures
- References that do not correct this problem have a temperature dependence of 10 ppm°/C to 50 ppm/°C over 0°C to 70°C.

Some Curvature Correction Techniques

- Squared PTAT Correction:

  Temperature coefficient ≈ 1-20 ppm/°C

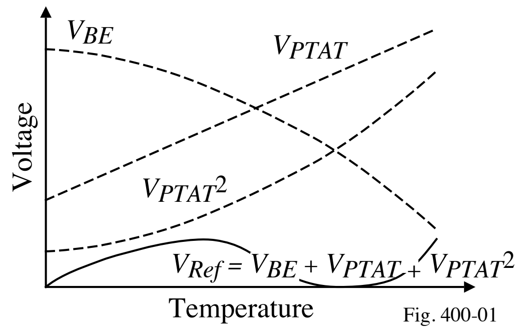


Fig. 400-01

- 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:

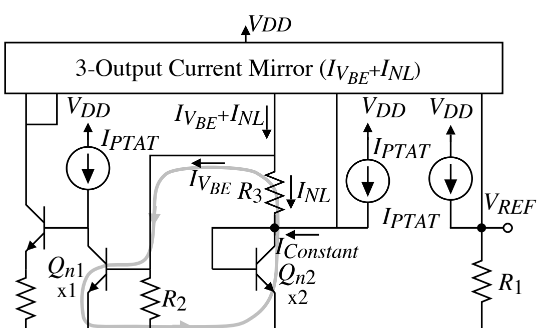


Fig. 400-02

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:

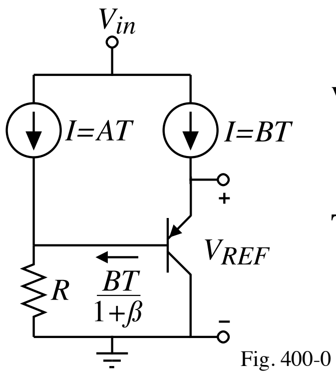


Fig. 400-0

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}.

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

Series Temperature Independent Voltage Reference with Curvature Correction

Objective: Eliminate nonlinear term from V_{CTAT}.

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

Operation:

$$V_{REF} = V_{PTAT} + 3V_{CTAT} - 2V_{Constant}$$

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

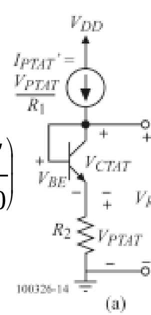
and $I_{Constant} \Rightarrow I_c \propto T^0 \Rightarrow \alpha = 0$,

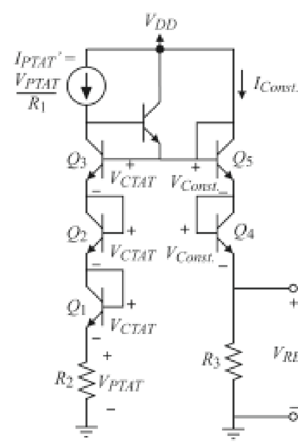
Previously we found,

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

so that

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





and

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

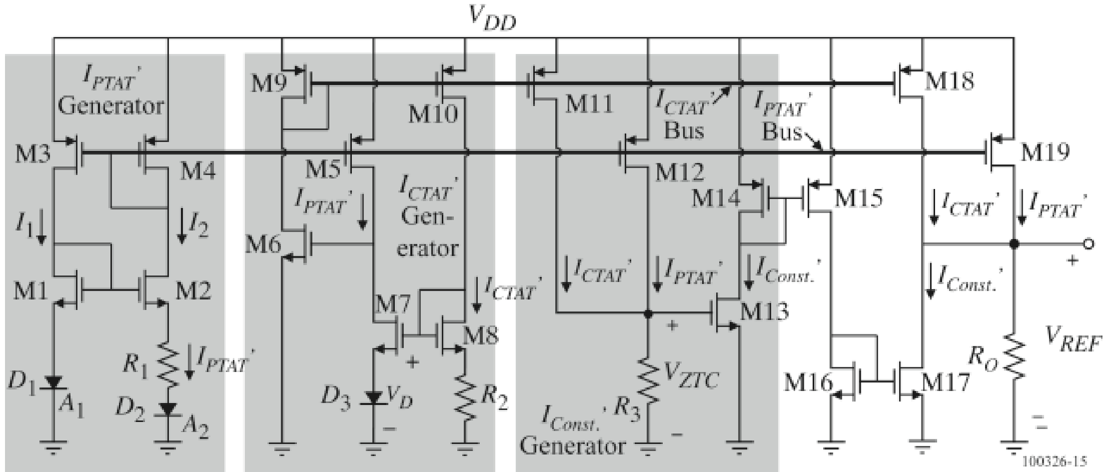
Combining the above relationships gives,

$$V_{REF}(T) = V_{PTAT} + V_{GO} - (T/T_0)[V_{GO} - V_{CTAT}(T_0)] - [\gamma - 3] \, V_t \, ln((T/T_0))$$

If $\gamma \approx 3$, then $V_{REF}(T) \approx V_{PTAT} + V_{GO}(1 - (T/T_0)) + V_{CTAT}(T_0)(T/T_0)$

A Parallel Version of the Nonlinear Curvature Correction Technique

The last idea was good in concept but not appropriate for CMOS implementation. The following is a better implementation.



$$V_{REF} = R_0[I_{PTAT}' + I_{CTAT}' - I_{const.}] = \frac{R_0}{R_1} V_{PTAT} + \frac{R_0}{R_2} V_{CTAT} - \frac{R_0}{R_3} V_{const.}$$

Use the resistor ratios to eliminate the nonlinear term given γ and α.

Parallel Curvature Correction Reference - Continued

Substitute for V_{CTAT} and $V_{const.}$ in the expression for V_{REF}.

$$V_{REF} = \frac{R_0}{R_1} V_{PTAT} + \frac{R_0}{R_2}\left[V_{GO} - \frac{T}{T_0}[V_{GO} - V_{CTAT}(T_0)] - (\gamma-1)V_t \ln\left(\frac{T}{T_0}\right)\right]$$

$$- \frac{R_0}{R_3}\left[V_{GO} - \frac{T}{T_0}[V_{GO} - V_{CTAT}(T_0)] - \gamma V_t \ln\left(\frac{T}{T_0}\right)\right]$$

To cancel the nonlinear CTAT term, we want the following relationship to hold:

$$\frac{R_0}{R_2}(\gamma-1) = \frac{R_0}{R_3}\gamma \quad \Rightarrow \quad \frac{R_2}{R_3} = \frac{(\gamma-1)}{\gamma}$$ (Fortunately γ is always greater than 1)

With these constraints, we find the voltage reference to be,

$$V_{REF} = \frac{R_0}{R_1} V_{PTAT} + \left(\frac{R_0}{R_2} - \frac{R_0}{R_3}\right)\left[V_{GO} - \frac{T}{T_0}[V_{GO} - V_{CTAT}(T_0)]\right]$$

$$= \frac{R_0}{R_1} V_{PTAT} + \frac{1}{\gamma}\frac{R_0}{R_2}\left[V_{GO} - \frac{T}{T_0}[V_{GO} - V_{CTAT}(T_0)]\right]$$

$$= \frac{R_0}{\gamma R_2}\left\{\frac{\gamma R_2}{R_1} V_{PTAT} + \left[V_{GO} - \frac{T}{T_0}[V_{GO} - V_{CTAT}(T_0)]\right]\right\} = \frac{R_0}{\gamma R_2}\left[\frac{\gamma R_2}{R_1} V_{PTAT} + V_{CTAT}(T_0)\right], (T = T_0)$$

Design $\gamma R_2/R_1$ to achieve temperature independence and R_0/R_2 to get V_{REF}.

Example 170-3 – Design of a Zero Temperature Coefficient Voltage Reference

Assume that $V_{CTAT} = 0.7$ V, R_3 = 10kΩ, $\gamma = 3.2$, $A_2 = 10A_1$, and $V_t = 0.026$ V at room temperature for the parallel curvature correction circuit. Find R_2 and R_3 to give a zero temperature coefficient at room temperature and a reference voltage of 1.0V.

To eliminate the nonlinear CTAT term,

$$\frac{R_2}{R_3} = \frac{(\gamma-1)}{\gamma} = \frac{(2.2)}{3.2} = 0.6875 \qquad \Rightarrow \quad R_2 = 6.88\text{k}\Omega$$

To cancel the temperature dependence,

$$\text{Temp. independent constant} = K = \frac{\gamma R_2}{R_1} = \frac{V_{GO} - V_{CTAT} + (\gamma-\alpha)V_{t0}}{V_{PTAT}}$$

or

$$\frac{\gamma R_2}{R_1} = \frac{V_{GO} - V_{CTAT} + (\gamma-\alpha)V_{t0}}{V_{PTAT}} = \frac{(1.205 - 0.7 + (3.2\text{-}1)(0.026)}{(0.026)(2.3026)} = 9.3907 \Rightarrow R_1 = 2.34\text{k}\Omega$$

The reference voltage can be written as,

$$V_{REF} = \frac{R_0}{\gamma R_2}\left[\frac{\gamma R_2}{R_1} V_{PTAT} + V_{CTAT}(T_0)\right] = \frac{R_0}{\gamma R_2}[9.3907(0.026)(2.3026) + 0.7]$$

$$\frac{R_0}{R_2} = \frac{(3.2)}{1.262} = 2.535 \qquad \Rightarrow \quad R_0 = 2.535R_2 = 17.44\text{k}\Omega$$

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.

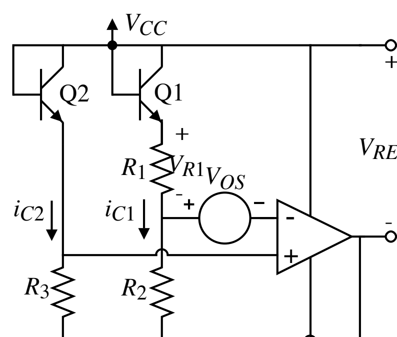


Fig. 400-05

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

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

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

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

Therefore,

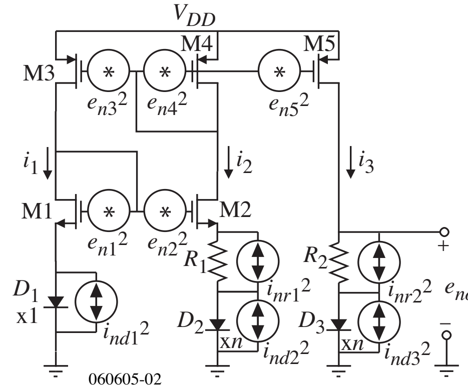
$$V_{R1} = -V_{OS} + V_t \ln\left[\frac{R_2A_{E1}}{R_3A_{E2}}\left(1 + \frac{V_{OS}}{i_{C1}R_2}\right)\right]$$

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

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

Noise Analysis of a Bandgap Reference

Consider the simple classical BG reference shown (R_2 = 10 R_1 = 10kΩ):



The open-circuit output noise voltage squared is found as,

$$e_{no}^2 = [e_{n1}^2/R_1^2 + e_{n2}^2/R_1^2 + g_{m5}^2e_{n3}^2 + g_{m5}^2e_{n4}^2 + g_{m5}^2e_{n5}^2 + i_{nd1}^2/(g_{m1}^2R_1^2) + i_{nd2}^2 + i_{nd3}^2 + i_{nr1}^2 + i_{nr2}^2]\, R_2^2$$

Assuming the MOSFETs are matched and the dc currents in D_1, D_2, and D_3 are equal gives,

$$e_{no}^2 \approx [g_{m5}^2(e_{n3}^2+e_{n4}^2+e_{n5}^2) + i_{nd2}^2 + i_{nd3}^2 + i_{nr1}^2 + i_{nr2}^2]\, R_2^2$$

Thermal noise gives (g_{m5} = 400μS),

$$e_{no}^2 = 8kTg_{m5}^2R_2^2 + 4qI_1 + \left(\frac{4kT}{R_1} + \frac{4kT}{R_2}\right)R_2^2 \approx 5.3\text{x}10^{-19} + 6.4\text{x}10^{-23} + 1.7\text{x}10^{-15}(\text{V}^2/\text{Hz})$$

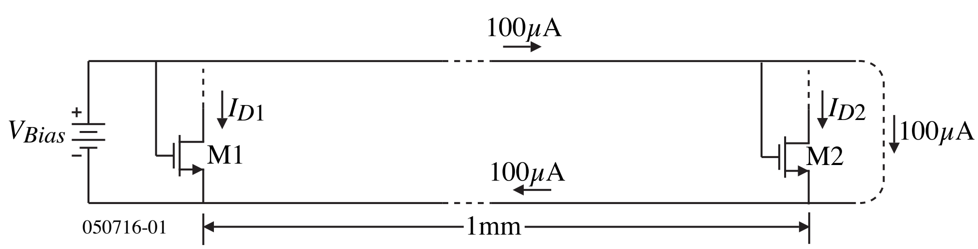
1/*f* noise gives,

$$e_{no}^2 = 3g_{m5}^2\frac{KF}{2fC_{ox}WLK'}$$

DESIGN OF BIAS VOLTAGES FOR A CHIP

Distributing Bias Voltages over a Distance

The major problem is the *IR* drops in busses. For example,



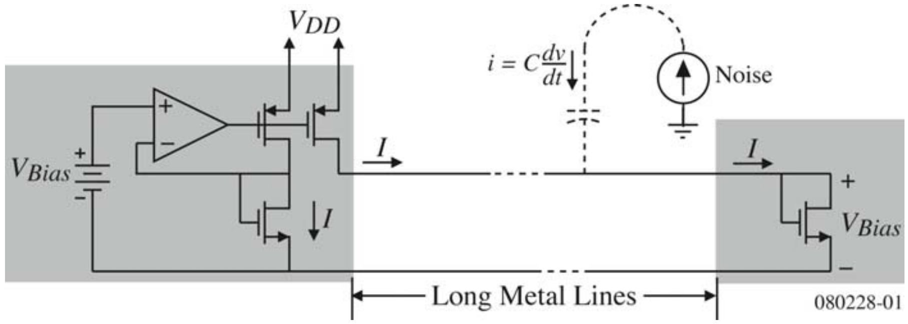
If the bus metal is 50mΩ/sq. and is 5μm wide, the resistance of the bus in one direction is (50mΩ/sq.)x(1000μm/5μm) = 10Ω. The difference in drain currents for an overdrive of 0.1V is,

$$V_{GS1} = 1\text{mV} + V_{GS2} + 1\text{mV} = V_{GS2} + 2\text{mV}$$

$$\frac{I_{D1}}{I_{D2}} = \frac{(V_{GS1}-V_{TN})^2}{(V_{GS2}-V_{TN})^2} = \frac{(V_{GS2}-V_{TN}+2\text{mV})^2}{(V_{GS2}-V_{TN})^2} = \left(\frac{0.1+0.002}{0.1}\right)^2 = 1.04$$

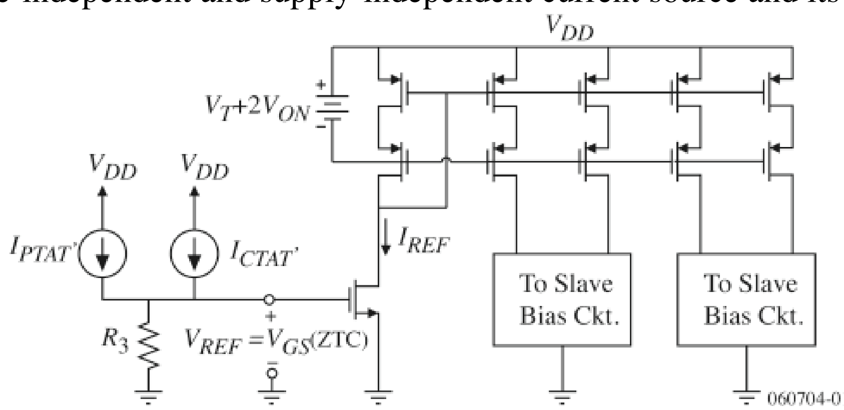
Use Current to Avoid *IR* Drops in Long Metal Lines

Example:



Practical Aspects of Temperature-Independent and Supply-Independent Biasing

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



The currents are used to distribute the bias voltages to remote sections of the chip.

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:

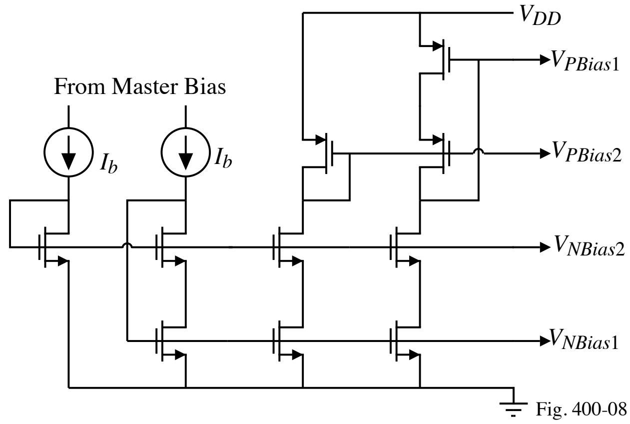


Fig. 400-08

From here on out in these notes,

$$V_{PBias1} = V_{PB1} = V_{DD}-|V_{TP}|-V_{SD}(\text{sat}) \quad V_{PBias2} = V_{PB2} = V_{DD}-|V_{TP}|-2V_{SD}(\text{sat})$$

and

$$V_{NBias1} = V_{NB1} = V_{TN} + V_{DS}(\text{sat}) \qquad V_{NBias2} = V_{NB2} = V_{TN} + 2V_{DS}(\text{sat})$$

SUMMARY OF TEMPERATURE STABLE REFERENCES

- The classical form of the temperature stable reference has a value of voltage close to the bandgap voltage and is called the “bandgap voltage reference”.
- 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.
- Distribution of bias voltages over a long distance should be done by current rather than voltage.