Beta Multiplier Referenced Self-Biasing

The **beta multiplier current reference** circuit provides an alternate scheme for (possibly) obtaining a PTAT-like current without using bipolar transistors. The basic form of the beta multiplier current reference is shown in the schematic below (excluding startup circuit). Note that the overall width of M2 is K times larger than M1 and $L_1 = L_2$ to provide $\beta_2 = K \cdot \beta_1$.

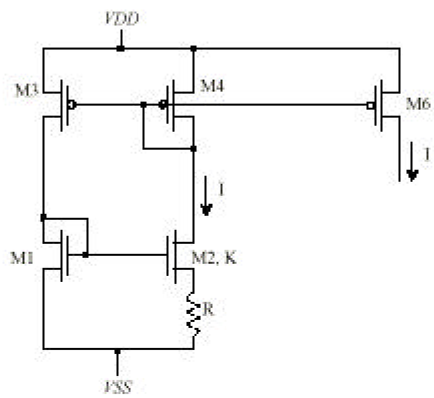

Figure 21.16 β multiplier referenced self-biasing ($\beta_2 = K \cdot \beta_1$).

Neglecting mismatch and λ effects, the current mirror M3, M4 provides $I_{D1} = I_{D2}$. Applying KVL, we obtain

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

After substituting for V_{GS} and neglecting body effect, channel-length modulation and mobility modulation,

$$\left(\sqrt{\frac{2I}{\beta_1}} + V_{THN}\right) = \left(\sqrt{\frac{2I}{\beta_2}} + V_{THN}\right) + IR$$

$$\left(\sqrt{\frac{2I}{\beta_1}} + V_{THN}\right) = \left(\sqrt{\frac{2I}{K \cdot \beta_1}} + V_{THN}\right) + IR$$

Solving for I provides the design equation for the beta multiplier current reference:

$$I = \frac{2}{R^2 \beta_1} \cdot \left(1 - \sqrt{\frac{1}{K}}\right)^2 \qquad \text{[SI sat.]}$$

Note the design variables that determine $I \Rightarrow$ ratioed gate widths (K) and the value of a single passive component (R).

The temperature coefficient for this reference's output current is described by

$$TC_I = \frac{1}{I} \cdot \frac{\partial I}{\partial T} = -2 \cdot \frac{1}{R}\frac{\partial R}{\partial T} - \frac{1}{KP(T)}\frac{\partial KP(T)}{\partial T} = -2 \cdot TCR + \frac{1.5}{T}$$

Obviously the resistor temperature coefficient (TCR) has significant impact on TC_I. In fact, TCR will determine if this reference achieves PTAT performance.

A practical example of a beta multiplier current reference is provided in example 21.4. The schematic for this example is included below. This circuit achieves a TC_I of 1,000ppm/°C at room temperature using a resistor with TCR = 2000ppm/°C. Note also the use of cascode MOSFETs to minimize the influence of finite MOSFET r_o.

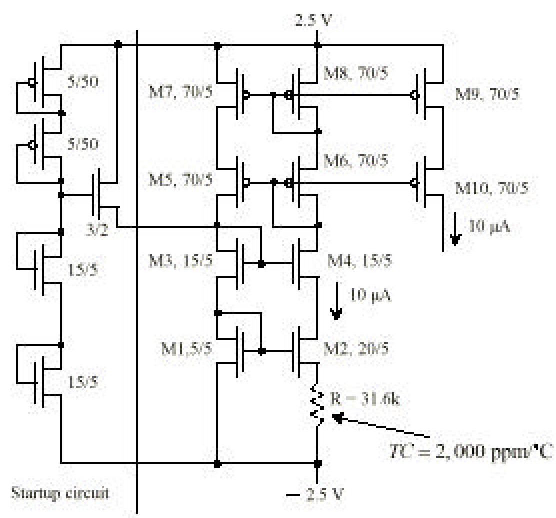

Figure 21.17 A 10 μA reference using a transconductance multiplier self-biased reference.

Recognize that the absolute accuracy of R dramatically effects the absolute accuracy of I. Additional sources of error for this circuit include device mismatch and body effect. Again, however, this technique does not require bipolar transistors that would need thorough characterization.

The same circuit can also used as a **beta multiplier voltage reference**. In this case, $V_{ref} = V_{GS1}$. In terms of current, we have

$$V_{ref} = V_{GS1} = \frac{2}{R\beta_1}\left(1 - \frac{1}{\sqrt{K}}\right) + V_{THN} \qquad \text{[SI sat.]}$$

Using this result, the change in V_{ref} with temperature can be predicted.

$$\frac{dV_{ref}}{dT} = \frac{dV_{THN}}{dT} - \frac{2}{R\beta_1}\left(1 - \frac{1}{\sqrt{K}}\right)\left[\frac{1}{R}\cdot\frac{\partial R}{\partial T} + \frac{1}{KP(T)}\cdot\frac{\partial KP(T)}{\partial T}\right]$$

And at room temperature, assuming TCR is 2,000ppm/°C, we obtain

$$\frac{dV_{ref}}{dT} = -2.4\frac{mV}{^\circ C} + \frac{2}{R\beta_1}\left(1 - \frac{1}{\sqrt{K}}\right)\left[-2{,}000\frac{ppm}{^\circ C} + \frac{1.5}{T}\right]$$

Selecting $K = 4$, a value for R can be determined that provides a temperature coefficient of zero for V_{ref}:

$$R = \frac{1}{0.8 \cdot \beta_1}$$

The value of V_{ref} associated with this R is

$$V_{ref} = 0.8 + V_{THN} = 1.63\text{V}$$

in CN20 technology. Sources of error include variations in V_{THN}, R, are β and the change in T. Note that V_{THN} also depends on body effect. Often times, however, body effect is ignored since V_{THN} can sometimes vary up to 20%.

Its susceptibility to process variations make the beta multiplier voltage reference difficult to use as a precision voltage reference. The bandgap voltage reference circuit is better suited for use as a precision voltage reference.

Low-power CMOS reference design can utilize MOSFETs operating in the **subthreshold region**. Circuits using subthreshold (weak inversion) MOSFETs can consume less than 100nA. At such low current levels, obviously the basic current mirror and resistor biasing scheme should not be used due to the excessively large value of resistance required.

Recall that the saturated weak inversion MOSFET's drain current is given by

$$I_{Dn} = I_{DO}\frac{W}{L}\left(e^{(V_{GS}-V_{THN})/n\cdot V_T}\right)$$

Operating the beta multiplier reference circuit in weak inversion provides

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

$$nV_T \cdot \ln\left[\frac{I\cdot L}{I_{DO}\cdot W}\right] + V_{THN} = nV_T \cdot \ln\left[\frac{I\cdot L}{I_{DO}\cdot K\cdot W}\right] + V_{THN} + IR$$

Then, solving for I, we obtain

$$I = \frac{n\cdot V_T}{R}\cdot \ln K$$

This result is identical to that of the thermal voltage referenced self-biasing technique! But without using BJTs and, most likely, less power.

Example 21.6 demonstrates using the beta multiplier in weak inversion. See the output current characteristic in Figure 21.19 of the text. The factor of two error in output current is due to body effect. Unless *p*MOS devices in the their own wells are used, the weak inversion beta multiplier will be very sensitive to body effect. Variations in R will also significantly effect the current.

Weak inversion operation of the beta multplier current reference also minimized its output voltage requirements.

Note also that in subthreshold operation the beta multiplier circuit can be used as a bandgap-like voltage reference.

Constant-G_m Biasing

Recall that using a PTAT current reference (see top of p. 66 in the notes) to bias a bipolar transistor provides constant transconductance over temperature (and also independent of supply voltage and process). How might we achieve similar results in CMOS analog circuits? Using the beta multiplier current reference.

As you may recall, the beta multiplier current reference (Figure 21.16-17 in your text – p. 69-70 in your notes), when operating in strong inversion saturation, generates an output current described by

$$I_{ref} = \frac{2}{R^2 b_1} \cdot \left(1 - \sqrt{\frac{1}{K}}\right)^2$$

Then a MOSFET (let's call it device M_A) biased by this current and operating in strong inversion saturation would have a transconductance described by

$$g_{mA} = \sqrt{2 b_A I_{ref}}$$

Substituting in for I_{ref},

$$g_{mA} = \sqrt{2 b_A \cdot \left(\frac{2}{R^2 b_1} \cdot \left(1 - \sqrt{\frac{1}{K}}\right)^2\right)} = \frac{2}{R} \cdot \left(1 - \sqrt{\frac{1}{K}}\right) \cdot \sqrt{\frac{b_A}{b_1}}$$

This result is independent of MOSFET parameters (such as m and V_{TH}) and supply voltage. The variation in g_{mA} over temperature, however, is directly affected by the temperature coefficient of R. And, of course, the absolute accuracy of R directly impacts the absolute accuracy of g_{mA}.