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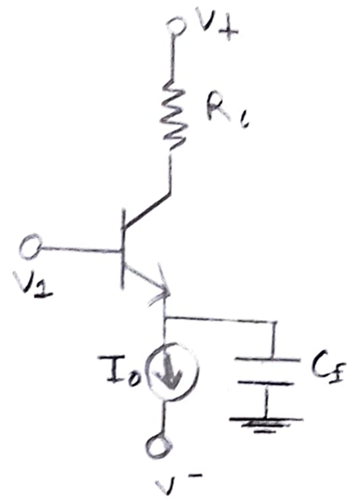
Q.14. Two transistor current source.

other name → current mirror.

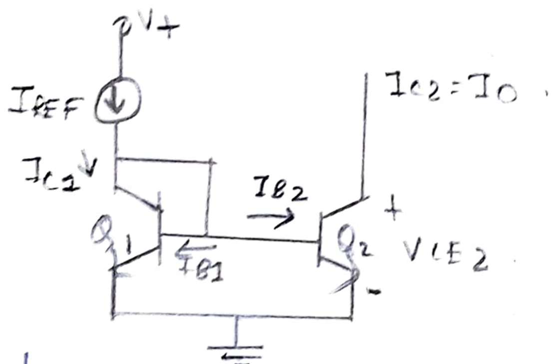
A current mirror is nothing but a current source created by a transistor circuit in order to maintain uniformity in current supply.

now, considering a BJT :-



when a BJT is connected with a Reference current IREF in short with the Base terminal, we can connect it to another transistor to get a copy of the current source IREF which is why it is called current "Mirror".

current mirror :-



↓

condition

$I_{C1} \approx I_{C2}$ for current mirror

current Relationship:-

$I_{REF} = I_{C1} + I_{B1} + I_{B2}$

$= I_{C1} + 2I_{B2}$.

⇒ Since $I_{C1} \approx I_{B2}$, $I_{B2} = I_{C2}/\beta$

⇒

$I_{REF} = I_{C2} + \frac{2I_{C2}}{\beta}$

$= I_{C2}\left(1 + \frac{2}{\beta}\right)$

$$\Rightarrow \left[I_{C2} = \frac{I_{REF}}{1+\frac{2}{\beta}} = I_o \right] \rightarrow \text{current relationship.}$$

Output resistance :-

Taking ratio of load current to reference current

$$\frac{I_o}{I_{REF}} = \frac{1}{\left(1+\frac{2}{\beta}\right)} \times \frac{\left(1+\frac{V_{CE2}}{V_A}\right)}{\left(1+\frac{V_{CE1}}{V_A}\right)}$$

V_A = Early effect voltage.

Reason for considering V_{CE}. In the previous current source circuit, we have a infinite resistance. but in practical, transistors do have a finite resistance which is why I_c becomes a function of collector emitter voltage.

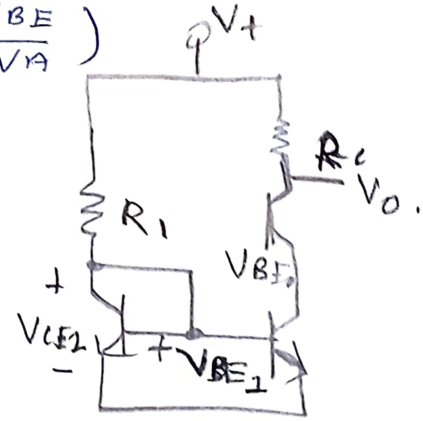
From circuit, we see that $V_{CE1} = V_{BE}$.

∴ differential change :-

$$\frac{dI_o}{dV_{CE2}} = \frac{I_{REF}}{\left(1+\frac{2}{\beta}\right)} \times \frac{1}{V_A} \times \frac{1}{\left(1+\frac{V_{BE}}{V_A}\right)}$$

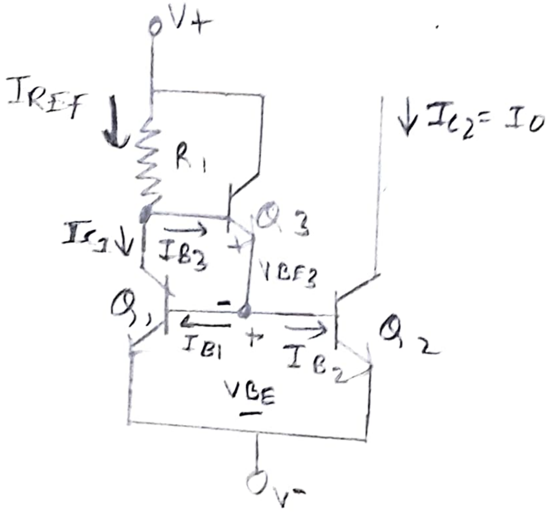
$$\Rightarrow \frac{dI_o}{dV_{CE2}} = \frac{I_o}{V_A} = \frac{1}{r_o}$$

where r_o = Output resistance.



Q.15. Basic three transistor current source.

circuit diagram :-



Assuming all transistors are identical
now, since B-E voltage is same for Q_1 & Q_2,

$\Rightarrow I_{B1} = I_{B2}$,

$I_{C1} = I_{C2}$

currents are suplied to Q_2 & Q_2 transistors by Q_3

now

Sum of currents at collector node of Q_1 :-

we get :-

$$I_{REF} = I_{C1} + I_{B3} \quad \text{— (1)}$$

$$\Rightarrow I_{B1} = I_{B2} = 2I_{B2} = I_{E3} \quad \text{—— (2)}$$

$$\Rightarrow I_{E3} = (1+\beta_3) I_{B3} \quad \text{—— (3)}$$

Solving (1), (2), (3), we get :-

~~$I_{E3} = (1+\beta_3) I_{B3}$.)~~

$$I_{REF} = I_{C1} + \frac{I_{E3}}{(1+\beta_3)} = I_{C1} + \frac{2I_{B2}}{(1+\beta_3)}$$

$\Rightarrow$

$I_{C1} \Rightarrow I_{C2}$ & $I_{B2} = I_{C2}/\beta$; we get :-

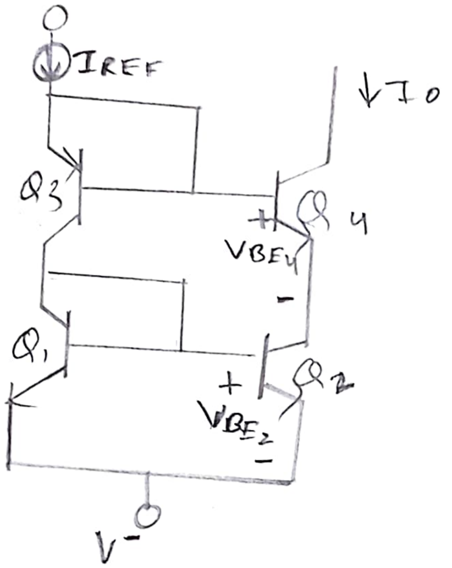
$$I_{REF} = I_{C2} + \frac{2I_{C2}}{\beta(1+\beta_3)} = I_{C2}\left[1 + \frac{2}{\beta(1+\beta_3)}\right]$$

∴ The output current I_O :- (I_{C2})

$$\Rightarrow I_O = I_{C2} = \frac{I_{REF}}{\left[1 + \frac{2}{\beta(1+\beta_3)}\right]}$$

Q.16.

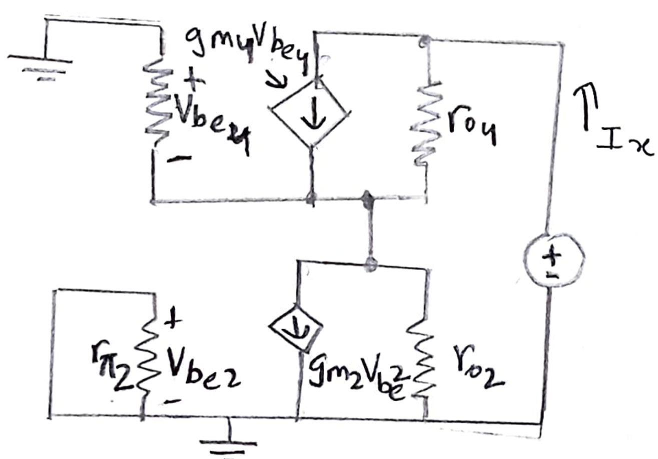
Cascode Current Mirror



If the transistors are matched, i.e. if their device parameters are identical,

then the load and reference currents are equal.

Small signal equivalent of above circuit.



In above circuit since $g_{m2} V_{be2} = 0$,

$V_{be4} = -I_x (r_{o2} \| r_{\pi 4})$.

$\Rightarrow \quad I_x = g_{m4} V_{be4} + \left(\dfrac{V_x - I_x (r_o \parallel r_{\pi 4})}{r_{o4}} \right)$

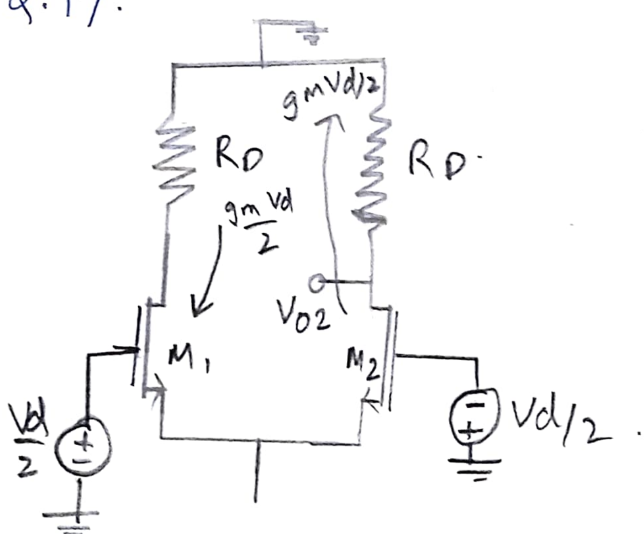
$= -g_{m4} I_x (r_{o2} \parallel r_{\pi 4}) + \left(\dfrac{V_x - I_x (r_{o2} \parallel r_{\pi 4})}{r_{o4}} \right)$

assuming $r_{\pi 4} << r_{o2}$,

$$\left[R_o = \frac{V_x}{I_x} = r_{o4}(1+\beta) + r_{\pi 4} \cong \beta r_{o4}. \right]$$

$\Rightarrow$ The output resistance has increased by a factor of β compared to the two transistor current source which increases the stability of current source with changes in output voltage.

Q.17.



$\rightarrow$ differential voltage gain.

$$A_d = \frac{V_o}{V_d} = \frac{g_m R_D}{2}$$

$$= \sqrt{\frac{k_n I_Q}{2}} \times R_D.$$

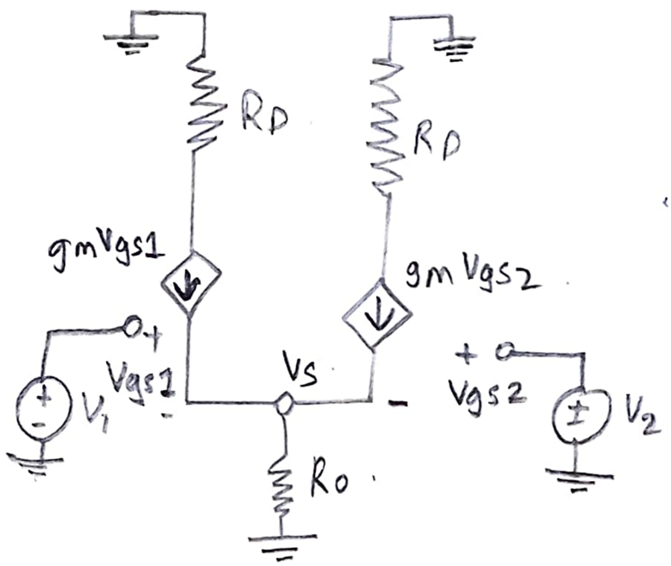
$\downarrow$

Small signal equivalent :-

All voltages are represented by their phasor components :-

Two transistors are biased at the same current

$\Rightarrow \quad g_{m1} = g_{m2} = g_m.$



KCL at Vs:-

$$g_m V_{gs1} + g_m V_{gs2} = \frac{V_S}{R_O}.$$

$\because V_{g1} = V_1 - V_S, \quad V_{gs2} = V_2 - V_S.$

$$g_m(V_1 + V_2 - 2V_S) = \frac{V_S}{R_O}.$$

Solving for Vs:-

$$V_S = \frac{V_1 + V_2}{2 + \frac{1}{g_m R_O}}, \quad V_O = V_{d2} = -(g_m V_{gs2}) R_D = -(g_m R_D)(V_2 - V_S)$$

using above two equations we get:-

$$V_O = -g_m R_D \left[\frac{V_2\left(1 + \frac{1}{g_m R_O}\right) - V_1}{2 + \frac{1}{g_m R_O}} \right] \quad \text{—— (1)}$$

now Equation (1) can be written as.

$$V_O = \frac{g_m R_D V_d}{2} - \frac{g_m R_D}{1 + 2 g_m R_O} V_{cm}. \quad \text{—— (2)}$$

Output voltage general form:-

$$V_O = A_d V_d + A_{cm} V_{cm} \quad \text{—— (3).}$$

→ transconductance:-

$$g_m = 2\sqrt{K_n I_{DQ}} = \sqrt{2 K_n I_Q}.$$

⇒ comparing (2) & (3).

$$A_d = \frac{g_m R_D}{2} = \sqrt{2 K_n I_Q}\left(\frac{R_D}{2}\right) = \sqrt{\frac{K_n I_Q}{2}}\, R_D.$$

& common mode gain:-

$$A_{cm} = \frac{-g_m R_D}{1+2g_m R_o} = \frac{-\sqrt{2k_n I_Q} \cdot R_D}{1+2\sqrt{2k_n I_Q} \cdot R_o}.$$

now, we know

CMRR (common mode rejection ratio)

$= |A_d/A_{cm}|$

$\therefore$ from above equations:-

$$CMRR = \frac{1}{2}\left[1 + 2\sqrt{2k_n I_Q} \cdot R_o\right]$$

Q18.

$$R_o = \frac{1}{\lambda I_Q} = \frac{1}{(0.02)(0.5)} = 100k\Omega.$$

$$A_d = \sqrt{\frac{k_n I_Q}{2}} = 8 \quad , \quad A_{cm} = \sqrt{2(1)(0.5)(16)}$$

where

$k_n = 1$ here.

$$CMRR = 20 \log 10\left(\frac{8}{0.0798}\right)$$

$= 40.045.$