ECAD lecture: Constant Current Sources

Topics Covered: Need of constant current sources, BJT current sources: Two transistor and improved three transistor current source, Cascode current source, Wilson and Widlar current source, MOSFET current sources

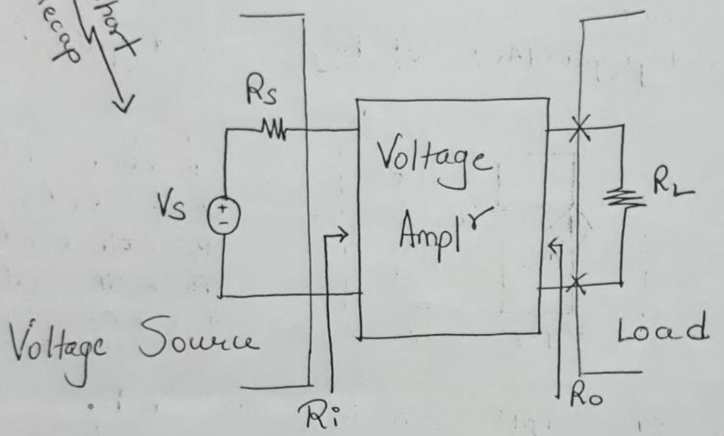
Module or sub-module covered: 6.3

Constant Current Sources

Intro : To understand current sources, let's first understand current amplifier's & its requirements.

So, far we have discussed in majority on only voltage sources and voltage amplifiers.

Short Recap



where, Rs → internal source resistance

We know following w.r.t voltage ampl^r that,

1) It's I/p impedance should be high (Ri)
2) It's o/p impedance should be low (Ro)

Requirements for voltage source, is

1) It's source resistance (internal) should be v. low ie $\boxed{R_s << R_i}$ → so that entire source voltage (Vs) reach amplifier I/P and to prevent loading effect

Load requirements:-

1) O/P impedance (Ro) of amplifier should be Low ie $\boxed{R_o << R_L}$ → so that entire o/p reaches Load (RL).

Concept:

- Current Amplifiers:-

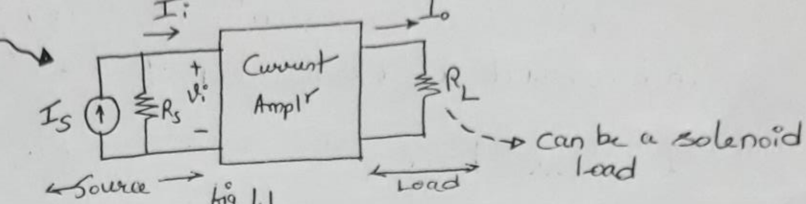


fig 1.1

→ It's o/p current (I_o) is proportional to its I/P current (I_i)
→ It's I/P is a "Current Source".
→ Provides modest voltage gain (A_v) but high current gain (A_i)

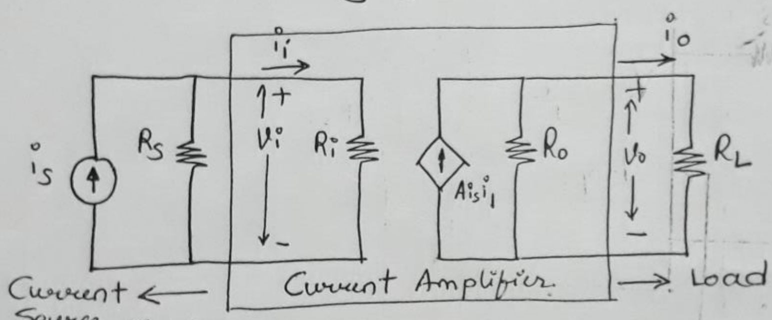


fig 1.2

where,
R_L → Load resistance
A_{is} → current gain
R_i → I/P resistance
R_o → o/p resistance
$A_{is} i_1$ → current-controlled current source

- o/p current (i_o) of the ampl^r can be obtained by using current-divider rule (CDR),

$$i_o = A_{is}\, i_i \left(\frac{R_o}{R_o + R_L}\right) \quad —\ (1)$$

From (1), we see that $R_o >> R_L$ → so that max current (i_o) flows th^r the Load

Also, i/p current i_i of the ampl^r is related to s/g source current i_s by CDR,

$$i_i = \left(\frac{R_s}{R_s + R_i}\right) i_s \quad —\ (2)$$

From (2), if $R_i << R_s$, then $\boxed{i_i = i_s}$, i.e. entire source current reaches I/P of ampl^r.

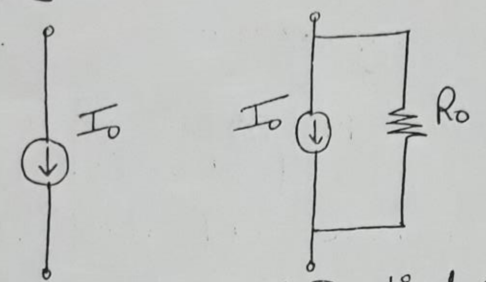
From (1) & (2), the requirements of current Amplifier are

1) $R_i << R_s$
2) $R_o >> R_L$

→ If these condition's are met, there will be no Loading effect.

· In summary, an ideal current amplifier should have $R_o \rightarrow \infty$ & $R_i \rightarrow 0$, so that there is no reduction in the current gain.

. From fig 1.2, it is evident that a current source should very $\boxed{\text{high o/p resistance}}$



where,
R_o – internal current source resistance (or o/p resistance of current source).

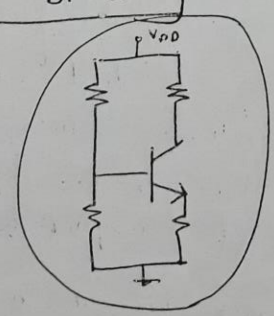
fig 1.3 a) Ideal current source ($R_o \rightarrow \infty$) b) Practical current source. ($R_o \rightarrow$ v.high)

→ An ideal current source should have <u>constant current</u> and a <u>v. large o/p resistance</u> ($R_o \rightarrow \infty$).

I. Why use Current Source's ?

1) Current-source biasing technique eliminates the need for <u>resistor-intensive biasing</u> used up to this point.

Remember this!



2) Current-source are used for biasing transistor's in Integrated circuits.

· Let's justify points (1) and (2),

- Biasing techniques for FET and BJT amplifiers employing voltage divider resistor networks can be used for <u>discrete circuits</u>, it is $\boxed{\text{NOT}}$ suitable for <u>integrated circuits.</u>

Because of the following reasons:-

1) Resistor's occupy large areas on an integrated circuit (IC) compared to transistor.
   ∴ A resistor-intensive ckt would necessitate a large chip area.

2) Transistor's can easily be fabricated in IC's with matched or identical parameters (which may not be possible in discrete circuits).

Let's consider another justification for using current source's.

- Why we do DC biasing in BJTs (let's say): Imagine a voltage-divider DC biasing ckt.
  a) to turn the device on
  b) to place Q-point in region of characteristics where device operates most Linearly so that any changes in the I/P source causes a proportional change in the o/p signal.
  c) ckt elements are selected so as to bias C-B & E-B J^n of BJT in appropriate magnitude & polarity.
  d) Since B-E J^n behaves Like a diode, transistor needs a B-E voltage V_{BE} of 0.7V to conduct.
  e) If we apply more than 0.7V, then transistor will be damaged due to excessive current.
  f) Hence, resistor (R_C, R_B, R_E) are used to limit the transistor currents (I_B, I_C, I_E).

- Recollect that voltage gain (A_v) of a CE ampl^r was

$\boxed{A_v = -g_m R_C}$ → small-signal mid-band voltage gain (with C_E, w/o R_L)

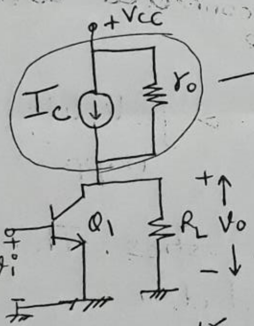
- In order to ↑se value of A_v, we need to ↑se R_C.

But ↑se in R_C will lead to more power dissipation or loss

So replace R_C by a current source

A current source has a high o/p resistance → Thereby, producing a high A_v.

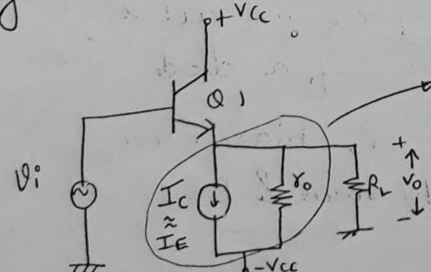
- Example:



current source → I_C flows $\boxed{\text{out}}$ of current source into Q_1.

This type of current source is referred to as "Sourcing Current source".

fig.1.3 a) CE ampl^r



current source → Source current flows from Q_1 $\boxed{\text{into}}$ the current source.

This type of current source is referred to as "Sinking Current Source".

b) Emitter follower

fig: 1.3 Amplifier with a biasing current source.

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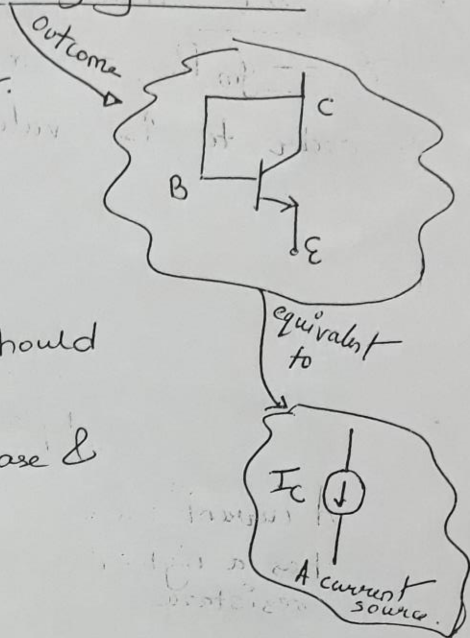
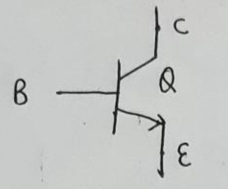
· How to make current sources utilizing BJT ?



→ Let us consider a npn transistor.



- For amplifier application's, BJT should be in forward-active region.
- So, what if we connect (short), Base & Collector terminal together, we get,

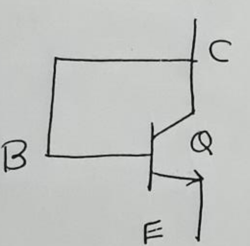


fig 1.4

- B–C J^n of a transistor if shorted → following happen's
  ie $V_{CB} = 0$

1) B–E Junction exhibits a diode characteristics. Then this transistor is said to be diode connected.

2) Still transistor operates internally in active region, but it exhibits the characteristics of a diode.

3) Even if V_{CB} is 0, still transistor is biased in forward-active mode. (coz BEJ^n is F/w biased & B-CJ^n is zero-biased or R/w biased)

* So above configuration of fig 1.4 behaves as a current source. Thus, transistor's can generate the characteristics of a constant-current source. An ideal current source has a very high o/p resistance, and its o/p current is $\boxed{\text{not sensitive}}$ to the transistor parameter β.

I. Need of Constant current sources:

. For increasing CMRR in case of diffamp's, either emitter or source resistance (R_E or R_S) has to increase which has the following problems.

1) It is very difficult to fabricated very large resistance values in Integrated ckts (ICs) since $R = \rho\left(\frac{L}{A}\right)$, where L ↑ses if R is required of high value.

2) Any ↑se in R_E or R_S will affect or change Q point (I_{CQ} or I_{DQ}).

3) In order to keep Q point constant for large value of R_E or R_S, very large negative power supplies ($-V_{EE}$ or $-V_{SS}$) are required.

4) Such a large value of R_E or R_S leads to very high power dissipation ⇒ means loss of dc power.

Hence for all practical diffamp's applications, the source or emitter resistance (R_E or R_S) are replaced by a (CCS) constant current source, which only ↑se value of R_E or R_S electrically, not physically at the same time providing constant bias current (I_{CQ} or I_{DQ}).

. This is possible becoz, o/p resistance (R_o) of current source is very high.

→ This explain's the need of CCS in diffamp.

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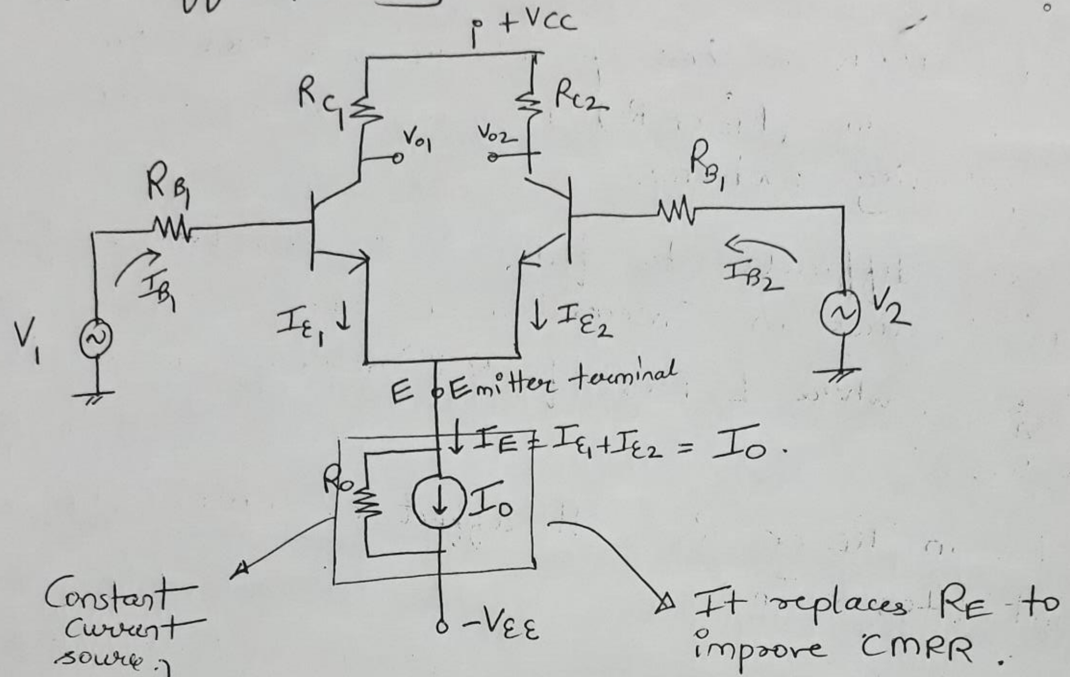
→ Diffamp using Constant current source



↳ It's o/p resistance (R_o) → is extremely large compared to $\underline{\underline{R_E}}$.

→ Now, this constant current source ckt's we are going to study and analyze in details.

ie We will consider types of ckts that can be designed to establish the bias current I_o.

BJT Current Sources:-

Commonly used current sources:-

a) Two transistor current source
b) Improved three transistor current source
c) Cascode current source
d) Wilson current source
e) Widlar current source.

A] Two transistor (BJT) Current Source:

- It is also called "current mirror."
- It is the basic building block in the design of integrated ckt current source.

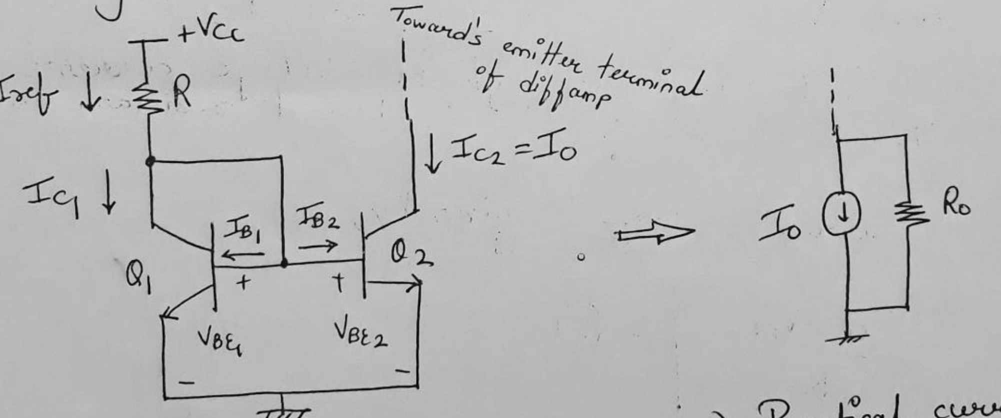


fig 2.1 a) Two transistor current source b) Practical current source.

- Transistor Q_1 is diode connected & its V_{CB} is forced to 0. Still Q_1 will operate in the active region. Q_2 can be in active as well as in the saturation region.

→ The current source in fig 2.1(a) behaves as "current sink".

- When supply vtg ($+V_{cc}$) is applied, B-E J^n of Q_1 is F/w-biased & a reference current (I_{ref}) is established. ie V_{BE_1} is established.
- Once V_{BE_1} is established, it is applied to the B-E junction of Q_2. The applied V_{BE_2} turns Q_2 ON and generates the load current $I_o (= I_{C2})$, which is used to bias a transistor ckt.

Note: I_{ref} is established by connecting a resistor R to the voltage source ($+V_{cc}$).

- In fig (2.1 a), the two transistor Q_1 & Q_2 are considered to be matched or identical transistor's, operating at the same temperature, with their base terminals and emitter terminals connected together.

  $\therefore V_{BE_1} = V_{BE_2} = V_{BE}$, $\therefore I_{B_1} = I_{B_2}$ & $I_{C_1} = I_{C_2}$

- Also, these two tr's Q_1 & Q_2 have (negligible leakage currents & whose o/p resistance are ∞) which is another assumption.

- The reference current I_{ref} is given by,

$$\boxed{I_{ref} = \frac{V_{cc} - V_{BE_1}}{R}} \quad \text{--- (1)}$$

Note: In fig (2.1a), if instead of gnd, $+V_{EE}$ (-ve supply) is connected then eqn (1) becomes, $\boxed{I_{ref} = \dfrac{V_{cc} - V_{BE} - V_{EE}}{R}}$

Point to be noted → To allow for a wide o/p voltage swing, an ampl^r is often connected to 2 DC supplies.

Current Relationships:- (2 transistor current source)

. Fig 2.1(a), shows the currents in the two-transistor current source. Since V_{BE} in both Q_1 & Q_2 are same, & the transistors are identical, then $I_{B_1} = I_{B_2}$ and $I_{C_1} = I_{C_2}$. Transistor Q_2 is assumed to be biased in the forward-active region.

Applying KCL at the collector of Q_1, will give

$$I_{ref} = I_{C_1} + I_{B_1} + I_{B_2}$$

$$I_{ref} = I_{C_1} + 2I_{B_2} \quad -(2) \quad (\because I_{B_1} = I_{B_2})$$

$$\text{ie } I_{ref} = I_{C_2} + \frac{2I_{C_2}}{\beta} \quad (\because I_{C_2} = \beta I_{B_2} \text{ \& } I_{C_1} = I_{C_2})$$

$$\text{ie } I_{ref} = I_{C_2}\left(1 + \frac{2}{\beta}\right) \quad -(3)$$

From fig 2.1 (a), we have

$$I_0 = I_{C_2} \quad -(4)$$

$$\therefore \boxed{I_0 = I_{C_2} = \left(\frac{I_{ref}}{1 + 2/\beta}\right)} \quad -(5) \quad (\text{From } (3))$$

Here, since $\beta >> 1$, $\therefore \frac{2}{\beta} << 1$

$$\text{ie } \boxed{I_0 = I_{ref}} \quad -(6) \quad \text{ie } I_{C_1} = I_{C_2} = I_{ref}$$

ie For 2 identical transistor's, the reference & o/p current's are almost equal.

ie I_{C_2} which is the mirror image of I_{C_1}, is known as the "<u>mirror current</u>" of I_{C_1}

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KVL in C-E loop of Q_1 gives,

$$+V_{CC} - I_{ref} \cdot R - V_{CE_1} = 0$$

$$I_{ref} = \frac{V_{CC} - V_{CE_1}}{R_1}$$

However, Q_1's base & collector are connected together giving $V_{CE_1} = V_{BE}$, hence we can write

$$\boxed{I_{ref} = \frac{V_{CC} - V_{BE_1}}{R_1}} \quad — (7)$$

→ constant current

Assuming $V_{CC} >> V_{BE_1}$, where $V_{BE_1} \approx 0.7V$ (Si), hence eq^n (7) will act as a constant reference current. which is mirrored at the o/p.

$$\therefore \quad \boxed{I_0 = \frac{I_{ref}}{1 + 2/\beta}}$$

→ gives constant current. & becomes a constant current source as desired.

→ gives the ideal o/p current of the two-transistor current source.

→ For a transistor with finite o/p resistance, the effect of Early voltage V_A should be considered,

$$I_C = I_S \left[\exp\left(\frac{V_{BE}}{V_T}\right) - 1 \right] \left(1 + \frac{V_{CE}}{V_A} \right) \quad — (a)$$

O/P Resistance :

- In actual transistor's, the Early voltage is finite, which means that the collector current is a function of V_{CE}.

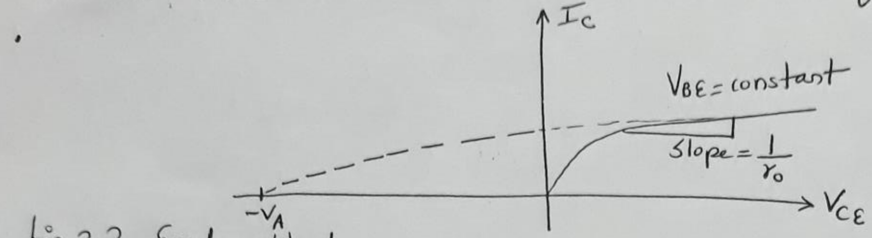


fig 2.2 Early effect.

- The stability of a Load current (I_o) generated in a constant-current source is a function of the $\boxed{\text{O/P resistance Looking back into the O/P transistor}}$

- From eq^n(a), ratio of 2 collector's are,

$$\frac{I_{C2}}{I_{C1}} = \frac{(1+V_{CE2}/V_A)}{(1+V_{CE1}/V_A)} \quad \text{— (b)}$$

From eq^n (5), we get

$$\frac{I_o}{I_{ref}} = \frac{1}{\left(\frac{2}{\beta}+1\right)} \frac{(1+V_{CE2}/V_A)}{(1+V_{CE1}/V_A)} \quad \text{— (c)}$$

where, V_A is the Early voltage.

→ From the ckt configuration, we see that $V_{CE1} = V_{BE1}$, which is essentially a constant,
Differentiating eq^n (c), we get.

$$\therefore \frac{dI_o}{dV_{CE2}} = \frac{I_{ref}}{(1+2/\beta)} \times \frac{1}{V_A} \times \frac{1}{(1+V_{BE}/V_A)}$$

change in Load current produced by change in V_{CE} of Q_2.

If we assume $V_{BE} << V_A$, then it becomes

$$\boxed{\frac{dI_o}{dV_{CE_2}} \simeq \frac{I_o}{V_A} \approx \frac{1}{r_o}} \quad \text{— (d)}$$

where, r_o is the small-s/g o/p resistance Looking into the collector of o/p transistor (Q_2).

where $\boxed{r_o = \frac{|V_A|}{I_o}}$

→ O/P resistance 'R_o' of 2 transistor Current Source:-

→ To find o/p resistance of constant-current ckt, place a test voltage 'V_x' at o/p node & analyze small-signal equivalent ckt.

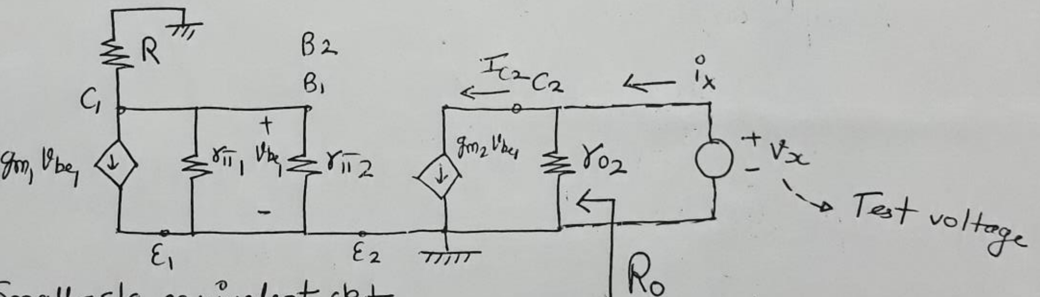


fig: 2.3: Small-s/g equivalent ckt of 2.1(a).

Note: Since I_{ref} is constant, it will make some of base voltages constant or at ac gnd.

In fig 2.3, B_1 & B_2 are at ac gnd, $g_{m_2} v_{be_1} \approx 0$

∴ O/P resistance R_o is the same as r_{o2}.

i.e $\boxed{R_o = \frac{v_x}{i_x} = r_{o2} = \frac{V_A}{I_{C_2}}}$ ⟶ (For 2 transistor current source)

—x—

· Design the basic two-transistor current source to give an o/p current of $I_o = 5\mu A$.
The transistor parameters are $\beta = 100$, $V_{CC} = 30V$, $V_{BE_1} = V_{BE_2} = V_{CE_1} = 0.7V$ and $V_A = 150V$.

Sol^n:- $I_o = I_{C2} = I_{C1} = \dfrac{I_{ref}}{1 + 2/\beta}$

ie $I_{ref} = I_o(1 + 2/\beta) = 5.1\mu A$.

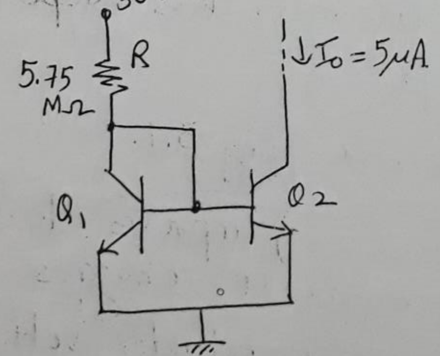
Also, $I_{ref} = \dfrac{V_{CC} - V_{BE1}}{R}$.

$\therefore \quad R = \dfrac{30V - 0.7V}{I_{ref}} = 5.75 M\Omega$

Also, now o/p resistance of Q_2 is.

$$R_o = \frac{V_A}{I_{C_2}} = \frac{150V}{5\mu A} = 30M\Omega$$

Designed ckt is



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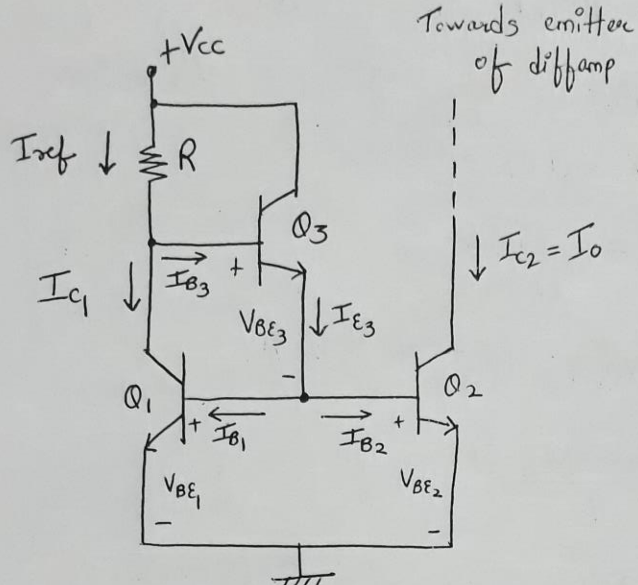
B] Improved three transistor current source:



Assumption's :-

Both Q_1 & Q_2 are identical. $\therefore V_{BE_1} \approx V_{BE_2}$
$\beta_1 \approx \beta_2 = \beta$ & $I_{B_1} = I_{B_2}$
thus $I_{C_1} = I_{C_2}$.

fig 3.1 Basic 3 transistor current source.

- In a two-transistor current source, from eq^n 5, the collector current $I_{C_2} (= I_{C_1})$ differs from the reference current I_{ref} by a factor of $(1 + 2/\beta)$.
- For Low-gain transistors, I_{C_2} can differ significantly from I_{ref}.
- The error can be reduced by adding another transistor so that I_{C_2} becomes less dependent on the transistor parameter β.
- This type of ckt is shown in fig 3.1.

Advantage → 3 transistor current source improves load current (I_0) stability against changes in β and changes in o/p collector voltages.

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- Current relationship (3 transistor current source)

Here (fig 3.1), transistor Q_3 supplies the base currents to Q_1 and Q_2, so these base currents should be less dependent on the reference current (I_{ref}).

KVL in the CE loop of Q_1 gives,

$$V_{CC} - I_{ref} \cdot R - V_{CE_1} = 0$$

where $V_{CE_1} = V_{BE_3} + V_{BE_1} \approx 1.4V.$

$$\text{ie } \boxed{I_{ref} = \frac{V_{CC} - V_{BE_3} - V_{BE_1}}{R}} \simeq \frac{V_{CC} - 1.4V}{R} \quad - (1)$$

Hence, Reference current is almost constant.

KCL at emitter of transistor Q_3, gives

$$I_{E_3} = I_{B_1} + I_{B_2} = \frac{I_{C_2}}{\beta} + \frac{I_{C_2}}{\beta} \quad \left(\because \begin{array}{l} I_C = \beta I_B \\ I_{C_1} = I_{C_2} \end{array}\right)$$

$$\therefore I_{E_3} = \frac{2 I_{C_2}}{\beta} \quad - (2)$$

$$\text{Now, } I_{B_3} = \frac{I_{E_3}}{1+\beta} = \frac{2}{\beta(1+\beta)} I_{C_2} \quad - (3) \quad (\because I_E = (1+\beta) I_B)$$

Now, applying KCL at the collector of Q_1 gives,

$$I_{ref} = I_{C_1} + I_{B_3} = I_{C_1} + \frac{2}{\beta(1+\beta)} I_{C_2}$$

$$\text{ie } I_{ref} = I_{C_2} + \frac{2}{\beta(1+\beta)} I_{C_2} \quad (\because I_{C_1} = I_{C_2})$$

(Q_1 & Q_2 are identical transistor's)

$$I_{ref} = I_{C_2}\left(1 + \frac{2}{\beta(1+\beta)}\right)$$

$\therefore$ The o/p current $I_0 (= I_{C_2})$ is

$$I_0 = I_{C_2} = \frac{I_{ref}}{1 + \frac{2}{\beta(1+\beta)}}$$

$$\boxed{I_0 = \left[\frac{I_{ref}}{1 + \left(\frac{2}{\beta^2 + \beta}\right)}\right]} \quad \text{--- (4)}$$

A comparison of eq^n of I_0 for a two-transistor current source and a three-transistor current source shows that the approximation of $I_0 \approx I_{ref}$ is better for the 3-transistor ckt.

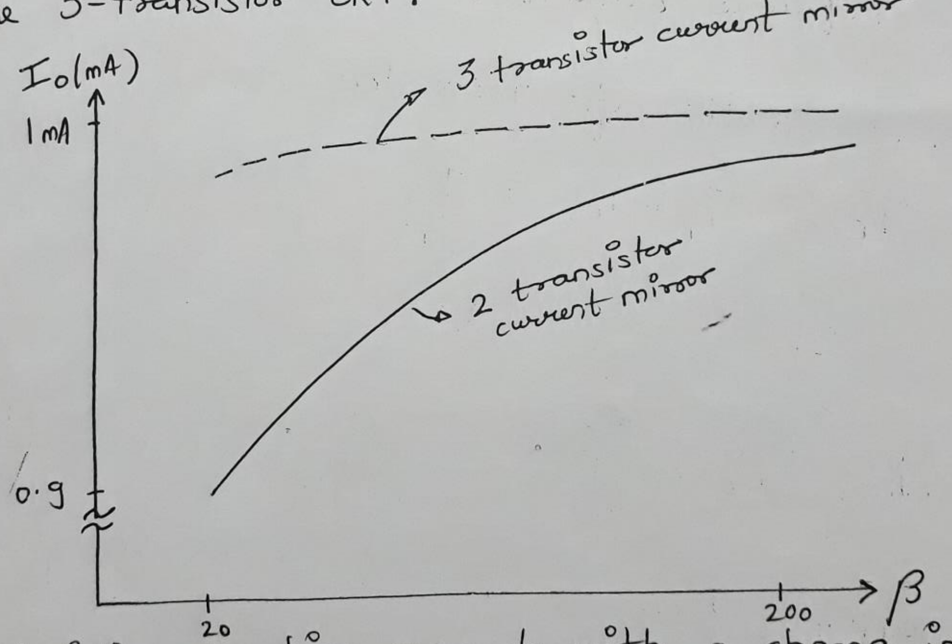

fig 3.2 Variation in bias currents with a change in β.

- Above fig 3.2, shows that change in Load current (I_0) with a change in β is much smaller in the three transistor current source.

O/P resistance (R_o) → (3 transistor current source)

The small-signal equivalent ckt for determining the o/p resistance is shown below (fig 3.3)

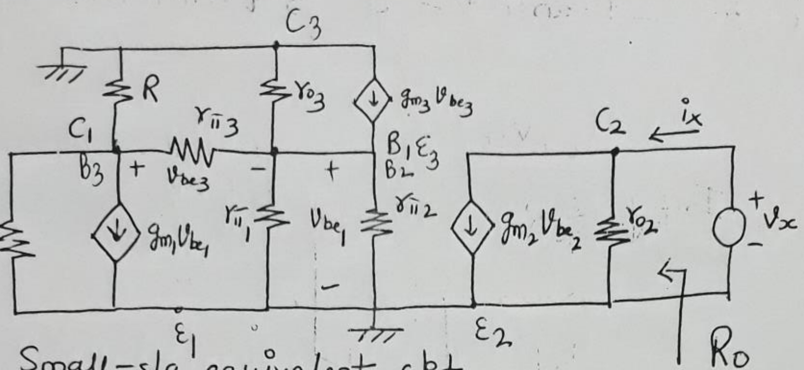

fig 3.3 Small-s/g equivalent ckt

The o/p resistance looking into o/p transistor Q_2 is

$$\boxed{R_o = \frac{v_x}{i_x} = r_{o2} = \frac{V_A}{I_{C2}}} \quad \text{---} \left(\begin{array}{c}\text{For 3 transistor} \\ \text{current source}\end{array}\right)$$

This o/p resistance is same as that of the two-transistor s current source.

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C] Cascode Current Source -

→ This configuration gives a larger O/P resistance compared to 2 transistor & 3 transistor current sources.

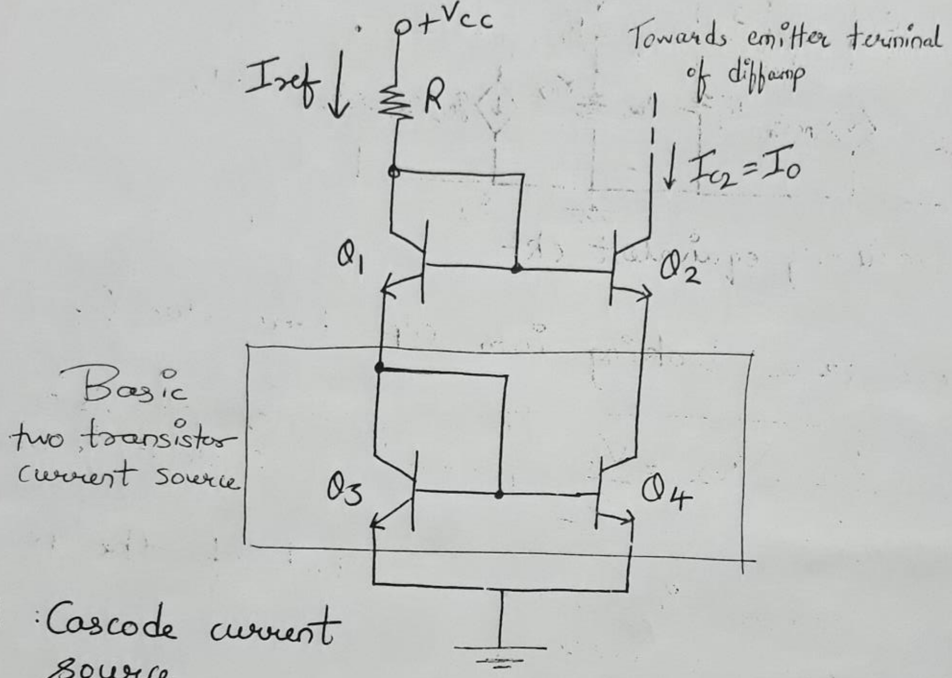


fig: 4.1 : Cascode current source

- In a cascode like connection, two or more transistors are connected in series so that their collector biasing currents are almost identical (eg Q_1 & Q_3), whereas in a cascode like connection, the transistor's operate in parallel fashion so that one transistor drives the other (eg Q_3 and Q_4).

- In this current source, if the transistor's are matched, then the load and reference currents are essentially equal.

ie $\boxed{I_0 \approx I_{ref}}$ --- (For a Cascode current Source)

. O/P resistance Ro (Cascode current mirror): 21

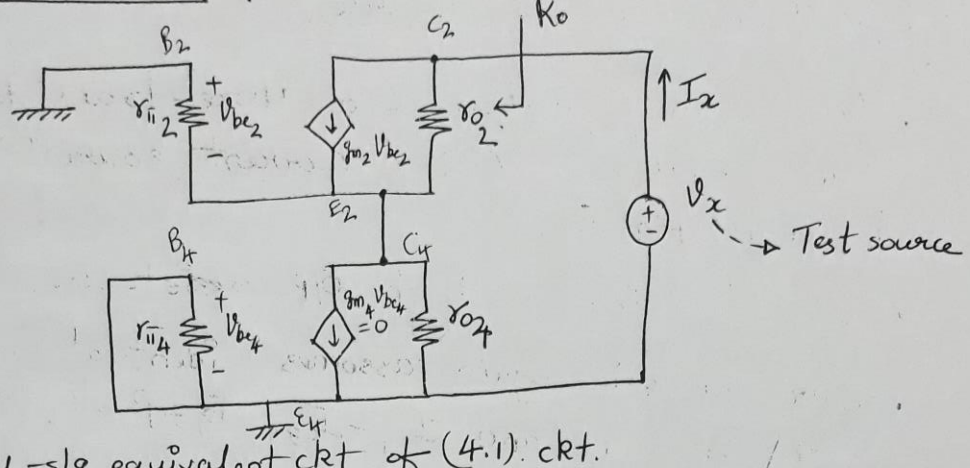


fig 4.2: Small-s/g equivalent ckt of (4.1) ckt.

- We may calculate the O/P resistance Ro by considering fig 4.2.
- For a constant reference current (Iref), the base voltages of Q_2 and Q_4 are constant, which implies these terminals are at s/g ground.
- ie $g_{m_2} v_{be_2} \approx 0$, then Ro is found out as,

$$R_o = \frac{v_x}{I_x} \approx r_{o2}\left[1 + g_{m2}(r_{o4} \| r_{\pi 2})\right]$$

$$\approx r_{o2}\left(1 + g_{m2} r_{\pi 2}\right) \qquad (\because r_{o4} >> r_{\pi 2} \approx r_{\pi 2})$$

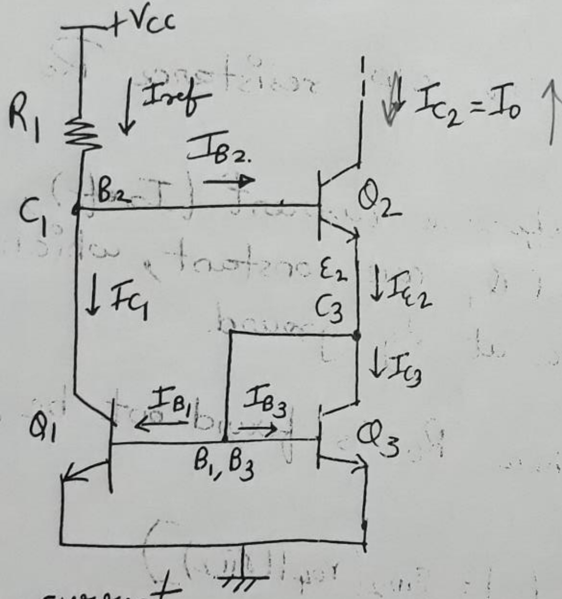
$$\boxed{R_o \approx r_{o2}(1+\beta)} \qquad (\because g_m r_\pi = \beta)$$

For a (Cascode current source)

Thus, Ro has ↑sed by a factor of $(1+\beta)$ compared to two-transistor current source, which increases the stability of current source with changes in O/P voltage.

D] Wilson Current Source:

- Another configuration of a three-transistor current source called a "Wilson current source" is shown below in fig 5.1
- It also gives a high o/p resistance.
- Our analysis again assumes identical transistors, with $I_{B_1} = I_{B_3}$ and $I_{C_1} = I_{C_3}$, $\beta_1 = \beta_2 = \beta_3 = \beta$.



5.1 : Wilson current source

Current Relationship:-

From fig 5.1, we can write

$$I_{E_2} = I_{C_3} + I_{B_3} + I_{B_1} = I_{C_3} + \frac{I_{C_3}}{\beta} + \frac{I_{C_3}}{\beta} \quad \left(\because I_C = \beta I_B,\ I_{C_1} = I_{C_3} \right)$$

$$I_{E_2} = I_{C_3}\left(1 + \frac{2}{\beta}\right) \quad \text{—} \quad (1)$$

Now, $I_{C_2} = \alpha I_{E_2} = I_{E_2}\left(\frac{\beta}{1+\beta}\right)$ $\quad (\because I_C = \alpha I_E)$

$$I_{C_2} = I_{C_3}\left(1+\frac{2}{\beta}\right)\left(\frac{\beta}{1+\beta}\right) \quad -(\text{From } 1)$$

ie $I_{C_2} = I_{C_3}\left(\frac{2+\beta}{1+\beta}\right)$

ie $I_{C_3} = I_{C_2}\left(\frac{1+\beta}{2+\beta}\right)$ — 2

Also, $I_{C_1} = I_{Ref} - I_{B_2} = I_{Ref} - \frac{I_{C_2}}{\beta} = I_{C_3}$ — 3

Equating 2 & 3, we get

$$I_{C_2}\left(\frac{1+\beta}{2+\beta}\right) = I_{Ref} - \frac{I_{C_2}}{\beta}$$

ie $I_{ref}\left(\frac{2+\beta}{1+\beta}\right) = I_{C_2} + I_{C_2}\left(\frac{2+\beta}{\beta(1+\beta)}\right)$

ie $I_{ref} = \left(\frac{1+\beta}{2+\beta}\right)\left[1+\frac{2+\beta}{\beta(1+\beta)}\right] I_{C_2}$

ie $I_{ref} = \left(\frac{\cancel{1+\beta}}{2+\beta}\right)\left(\frac{\beta+\beta^2+2+\beta}{\beta(\cancel{1+\beta})}\right) I_{C_2}$

$$I_{ref} = \left(\frac{\beta^2+2\beta+2}{(2+\beta)\beta}\right) I_{C_2}$$

ie $I_0 = I_{C_2} = I_{ref}\left(\frac{(2+\beta)\beta}{\beta^2+2\beta+2}\right)$

ie $\boxed{I_0 = I_{ref}\left[1 - \frac{2}{\beta^2+2\beta+2}\right]}$ - 4

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Since, $\beta >> 1$, eq^n (4) can be approximated by,

$$\boxed{I_0 \approx I_{ref}} \quad -(5)$$

Thus O/P current (I_0) almost equals the reference current (I_{ref}) and is less sensitive to the current gain β, which varies in response to temperature changes.

O/P resistance (Ro) :

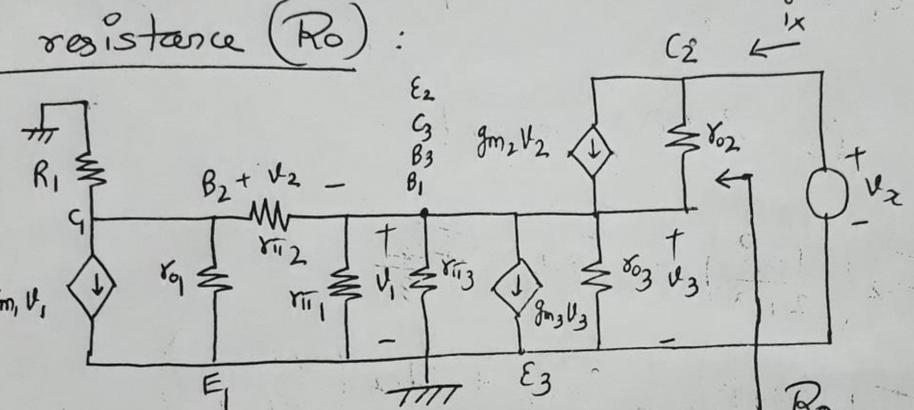

fig 5.2: Small-signal AC equivalent ckt

O/P resistance of a Wilson current source looking into collector of Q_2 is

$$R_o \approx r_{o2}\left(1 + \frac{g_{m2} r_{\pi 2}}{2}\right)$$

$$\boxed{R_o \approx r_{o2}\left(1 + \frac{\beta}{2}\right)} \longrightarrow \text{(For a Wilson current source)}$$

. O/P resistance is $\left(1+\frac{\beta}{2}\right)$ factor times larger than that of either the two-transistor or two-transistor current source.

E] Widlar Current Source

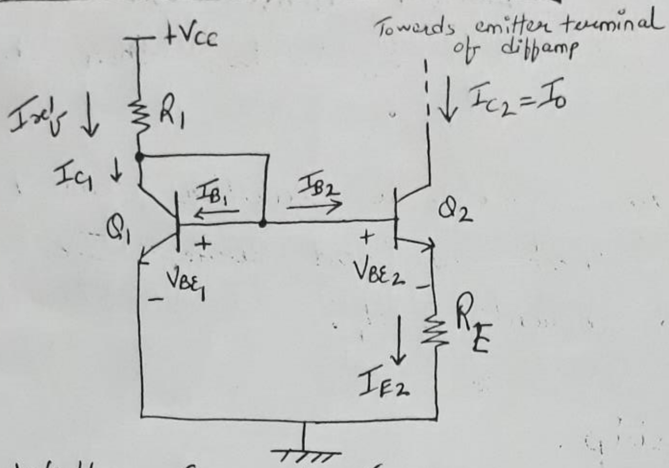

fig: 6.1 Widlar Current Source.

- It helps design biasing currents of low magnitudes (of order of 5μA).
- In the current-sources ckts considered so far, the Load (I_o) and reference (I_{ref}) currents have been nearly equal.
- For a two-transistor current source, if we require a load current of say $I_o = 10\mu A$, then for $V_{CC} = 5V$ and $V_{EE} = -5V$, the required resistance value is

$$R_1 = \frac{V_{CC} - V_{BE} - V_{EE}}{I_{ref}} \approx 930k\Omega$$

In integrated ckts, resistors on the order of 1MΩ require large areas and are difficult to fabricate. We therefore need to limit, resistor values to low kΩ range.

- The transistor ckt in fig (6.1), called a "Widlar current source", meets this objective.

- A voltage difference is produced across resistor R_2, so that the B-E voltage of Q_2 is less than the B-E voltage of Q_1.
- A smaller B-E voltage produces a smaller collector current which in turn means that the load current $\boxed{I_o}$ is less than the reference current $\boxed{I_{ref}}$.

- Current Relationship:

If the two transistors Q_1 & Q_2 are identical, then their leakage current's $I_{S_1} = I_{S_2}$ are same.

$\therefore \quad I_{C_1} \approx I_S \, e^{V_{BE_1}/V_T}$ where, I_S = reverse saturatn current.

ie $\quad I_{ref} \approx I_{C_1} = I_S \, e^{(V_{BE_1}/V_T)} \quad — (1)$ $\quad V_T$ = thermal voltage

and

$$I_o = I_{C_2} = I_S \, e^{(V_{BE_2}/V_T)} \quad — (2)$$

Consider eq^n (1) and solve for B-E voltage, we get

$$\frac{I_{ref}}{I_S} = e^{(V_{BE_1}/V_T)} \rightarrow \ln\left(\frac{I_{ref}}{I_S}\right) = \frac{V_{BE_1}}{V_T}$$

$$\therefore \quad V_{BE_1} = V_T \ln\left(\frac{I_{ref}}{I_S}\right) \quad — (3)$$

Similarly,

$$V_{BE_2} = V_T \ln\left(\frac{I_o}{I_S}\right) \quad — (4)$$

Eq^n ③ − Eq^n ④, yields

$$V_{BE_1} - V_{BE_2} = V_T \ln\left[\frac{I_{ref}}{I_0}\right] \quad — ⑤$$

From the fig 6.1, we see that
KVL to B-E Loop of Q_1 & Q_2,

$$V_{BE_1} - V_{BE_2} - I_{E_2} R_E = 0$$

where $I_{E_2} \approx I_{C_2}$ & $I_{C_2} \approx I_0$

$$\therefore \quad V_{BE_1} - V_{BE_2} \approx I_0 R_E \quad — ⑥$$

Combining ⑤ and ⑥, yields

$$\boxed{I_0 R_E = V_T \ln\left(\frac{I_{ref}}{I_0}\right)} \quad — ⑦$$

Eq^n ⑦ gives the relationship between the reference and bias currents.

The reference current I_{ref} can be found from,

$$\boxed{I_{ref} = \frac{V_{CC} - V_{BE1}}{R_1}} \quad — ⑧$$

Eq^n ⑦ shows that

$I_{C_2} = I_0$ is a non-Linear function of I_{ref} and R_E.

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O/P Resistance (R_o):

The change in load current with a change in V_{C2} of Widlar current source can be written as,

$$\frac{dI_o}{dV_c} \approx \frac{1}{R_o}$$

where R_o is the o/p resistance looking into the collector of Q_2. This R_o can be determined by using the small-signal equivalent ckt in fig 6.2

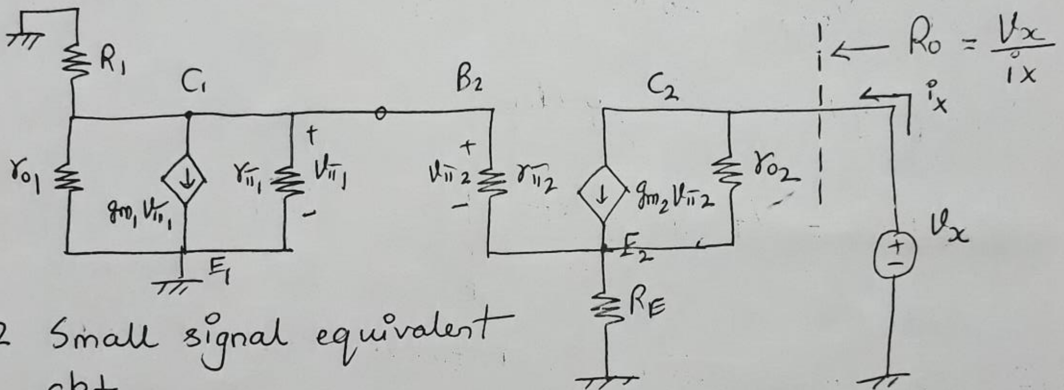

fig 6.2 Small signal equivalent ckt.

- O/P resistance is found out to be,

$$R_o = \frac{V_x}{I_x} \approx r_{o2}(1 + g_{m2}R_E)$$

R_o of widlar current source is a factor $(1+g_{m2}R_E)$ larger than that of simple 2 transistor current source

$$\boxed{R_o \approx r_{o2}\left(1 + \frac{I_{C2}R_E}{V_T}\right)} \quad \text{—(a)}$$

From eq^n(a), R_o depends on $I_{C2}R_E$, → which is the DC voltage drop across R_2. The larger this drop is made, the higher the O/P resistance becomes.

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Comparison of BJT current sources:-

Current Source Type	O/P resistance R_o	Relation between I_0 & I_{ref} $(\beta >> 1)$	Comments
Two-transistor current source	r_{o2}	$I_0 \approx \left(\frac{I_{ref}}{1+2/\beta}\right)$ ie $I_0 \approx I_{ref}$	· Not suitable for low currents $(< 0.3mA)$ · moderate R_o
Three-transistor current source	r_{o2}	$I_0 \approx \left(\frac{I_{ref}}{1+\frac{2}{\beta^2+\beta}}\right)$ ie $I_0 \approx I_{ref}$	· Not suitable for Low currents $(< 0.3mA)$ · moderate R_o
Cascode Current source	$r_{o2}(1+\beta)$	$I_0 \approx I_{ref}$	· Not suitable for Low currents · higher o/p resistance
Wilson current source	$r_{o2}\left(1+\frac{\beta}{2}\right)$	$I_0 \approx I_{ref}\left(1-\frac{2}{\beta^2+2\beta+2}\right)$ ie $I_0 = I_{ref}$	· Not suitable for Low currents · higher o/p resistance · $I_0 = I_{ref}$
Widlar current source	$r_{o2}(1+g_{m2}R_E)$	Non-Linear relation $I_0 << I_{ref}$	· Current as low as $5\mu A$ · $I_0 << I_{ref}$ · higher o/p resistance.

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• Advantages of using Constant Current Sources :-

1. Transistor current sources are widely used in analog Integrated ckts both as biasing elements and as Loads for amplifying stages.
2. Active current sources helps in $\uparrow$sing the voltage gain of an amplifier.
3. Constant current sources have a very high o/p resistance (R_o) and its O/P current is not sensitive to the transistor parameter β.
4. Current sources are less sensitive to variation's in DC power supply and temperature.
5. Especially, for a small value of bias current, the current sources are more economical than resistor's in terms of die area required for resistors.

→ A current source can be designed by using either MOSFET's or BJT's.

We have analyzed various BJT current sources'.

→ Let's analyze MOSFET current source's next.

• MOSFET current sources are analogous to BJT current sources

• MOSFET Current Sources

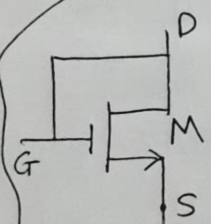
→ The choice of a BJT current source or a MOSFET current source generally depends on the type of integrated ckt involved (eg bipolar or MOS).

→ BJT current sources have some advantages over MOSFET current sources, such as a wider compliance range and a higher O/P resistance.

→ However, a higher o/p resistance can be obtained by cascode-like connections of MOSFETs.

• How MOSFET can be used as a Current Source?

We need to achieve this

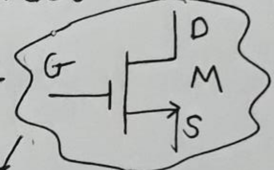


nMOS with gate & drain terminals shorted

behaves Like

Consider a nMOS (E-type)

Start here →



Current Io flowing in T^r M will be constant, if M is biased in saturation region

For MOSFET to be in saturat^n cond^n is

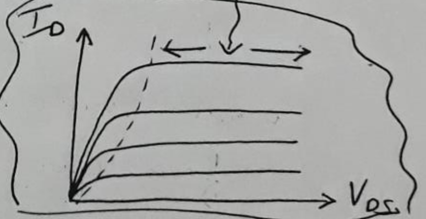
$V_{GS} \geq V_{TN}$

$V_{DS} \geq V_{GS} - V_{TN}$

Remember this

For this case, $V_{DS} \approx V_{GS}$ always so these cond^ns will be satisfied

Thus, M will be always be biased in saturat^n region



NB: Voltage compliance range :- It is the voltage range over which the ckt can maintain a constant current.

V_{TN} → Threshold voltage of NMOS-E type.

Types of MOSFET constant current sources which we will be studying are

a) Basic two-transistor current source

b) ~~Three-~~transistor current source.

c) Cascoded current source

d) Wilson current source.

A] Basic two-transistor MOSFET Current Source.

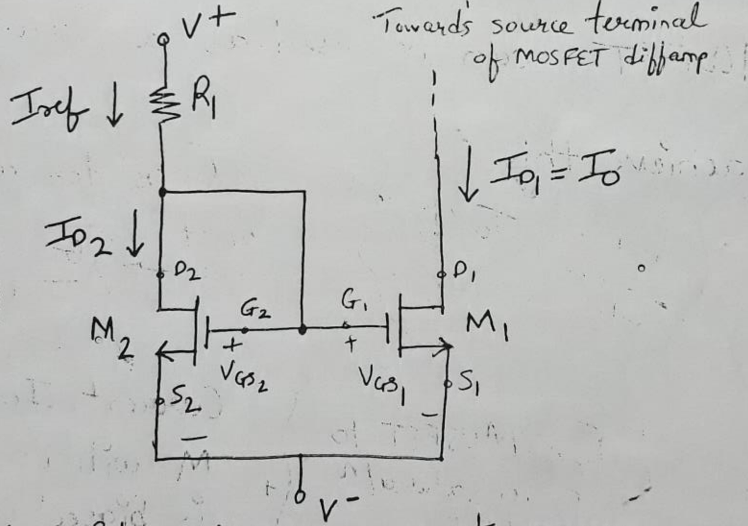


fig 7.1 Two-transistor NMOS current source.

- The direction of current flow in fig 7.1 is into the current source; this type of constant-current source is often referred to as a current sink.

- let us assume that the two transistors M_1 and M_2 are identical (ie $V_{GS_1} = V_{GS_2}$, ie $I_{D_1} = I_{D_2}$)

∴ Thus, the o/p current $I_o (= I_{D1})$ will be the mirror of I_{D2}.

- Also, since $V_{DS2} = V_{GS2}$, M_2 is always biased in the saturation region.
- Let V_{TN1} and V_{TN2} be the threshold voltages of M_1 & M_2.
- For M_1 to be in saturation, V_{DS1} which is greater than or equal to $(V_{GS2} - V_{TN2})$, must be greater than $(V_{GS1} - V_{TN1})$.
- This condition reduces the voltage compliance range of the MOSFET current source and prevents it from operating from a low power supply.

Current Relationship:

- Drain current eqn of an E-type NMOS biased in saturation region is,

$$I_{Dsat} = K_n (V_{GS} - V_T)^2 (1 + \lambda V_{DS}) \quad \text{— (a)}$$

where, $\boxed{K_n = \mu_n C_{ox} \left(\frac{W}{L}\right) \times \frac{1}{2}}$, λ - channel length modulation parameter

- The o/p current 'I_o', which is equal to drain current of M_1 is given by,

$$I_{D1} = I_o = K_{n1} (V_{GS1} - V_{TN1})^2 (1 + \lambda V_{DS1}) \quad \text{— (1)}$$

- Drain current I_{D2} which is equal to the reference current I_{ref} is

$$I_{D2} = I_{ref} = K_{n2} (V_{GS2} - V_{TN2})^2 (1 + \lambda V_{DS2}) \quad \text{— (2)}$$

- Since, all the components of the current source are processed on the same IC, all the physical parameters such as (V_{TN}, μ_n, C_{ox} & λ) are identical for both devices (M_1 and M_2)

$\therefore$ Ratio of I_o to I_{ref} is given by (using ① & ②),

$$\frac{I_o}{I_{ref}} = \frac{K_{n1}(V_{GS_1}-V_{TN_1})^2(1+\lambda V_{DS1})}{K_{n2}(V_{GS_2}-V_{TN_2})^2(1+\lambda V_{DS2})} \quad -③$$

$$= \frac{\cancel{\mu_n C_{ox}}\left(\frac{W}{L}\right)_1 \times \cancel{\frac{1}{2}}(V_{GS_1}-V_{TN_1})^2(1+\lambda V_{DS1})}{\cancel{\mu_n C_{ox}}\left(\frac{W}{L}\right)_2 \times \cancel{\frac{1}{2}}(V_{GS_2}-V_{TN_2})^2(1+\lambda V_{DS2})}$$

$$\boxed{\frac{I_o}{I_{ref}} = \frac{(W/L)_1}{(W/L)_2} \times \frac{(1+\lambda V_{DS1})}{(1+\lambda V_{DS2})}} \quad -④$$

In practice, $\lambda V_{DS} << 1$, then eq^n ④ can be approximatied by,

$$\boxed{\frac{I_o}{I_{ref}} = \frac{(W/L)_1}{(W/L)_2}} \quad -⑤$$

ie $\boxed{I_o = \frac{(W/L)_1}{(W/L)_2} \cdot I_{ref}}$ ⟶ By controlling the ratio (W/L), we can change the o/p current.

By choosing identical transistors with $W_1=W_2$ and $L_1=L_2$, a designer can ensure that the o/p current I_o is almost equal to the reference current I_{ref}.

From fig 7.1,

- Since $V_{GS_2} = V_{DD} - I_{ref} R_1$ and $V_{DS_2} = V_{GS_2}$ (From eq^n 2)

$$\therefore \boxed{I_{Ref} = I_{O_2} = K_{n_2}(V_{DD} - I_{ref} R_1 - V_{TN_2})^2} \quad -6$$

I_{ref} can be solved for known values of V_{TN_2}, K_{n_2}, V_{DD} and R_1.

- O/P resistance:

The stability of the Load current as a function of V_{DS} voltage is an important consideration in many applicat^{n}s.

- The small-s/g o/p resistance of the current source can be estimated as follows, (fig 7.2)

$$\boxed{R_O = r_{o_1}} \quad ; \quad r_{o_1} = \frac{V_x}{I_x} \quad , \quad \left\{ \text{since } g_{m_2} V_{gs_2} \approx 0,\ V_{gs_1} \text{ so } g_{m_1} V_{gs_1} \text{ becomes redundant} \right\}$$

where, r_{o_1} is o/P resistance looking into transistor M_1

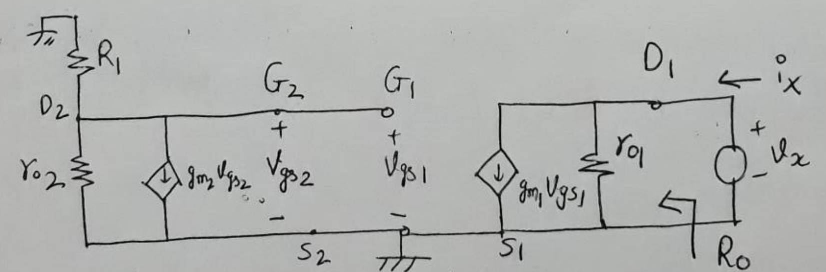


fig 7.2 : Small-s/g equivalent ckt. of fig 7.1

- The stability of the load current ($I_O = I_{D1}$) can be described in terms of o/P resistance,

From fig 7.1, $V_{DS_2} = V_{GS_2}$ = constant for a given reference current

Normally, $\lambda V_{DS_2} = \lambda V_{GS_2} << 1$, & if $(W/L)_1 = (W/L)_2$] (a)

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Then the change in bias current w.r.t to a change in V_{DS_1} is

$$\frac{1}{R_0} = \frac{\partial I_0}{\partial V_{DS_1}} = \frac{1}{r_{o_1}} = \frac{\partial I_{D_1}}{\partial V_{DS_1}} \quad --- (b)$$

from (a) & eq^n ① and ④,

$$\frac{\partial I_{D1}}{\partial V_{DS_1}} \approx \frac{K_{n_1}(V_{GS_1}-V_{TN_1})^2(1+\lambda V_{DS1})}{\lambda}$$

$$\approx I_{D_1}\lambda$$

ie $\frac{1}{r_{o_1}} \approx I_0 \lambda$

$\therefore \boxed{r_{o_1} \approx \frac{1}{\lambda I_0}}$ where, r_{o_1} is the o/p resistor of the transistor M_1.

$\therefore \boxed{R_0 = r_{o_1} = \frac{1}{\lambda I_0}}$ which is relatively small.

· This small o/p resistance is a disadvantage of having only one MOSFET M_1 at the o/p side of a current source.

——x——

8) Modified three-transistor MOSFET Current Source:

- Reference current in BJT current sources is generally established by bias voltage and a resistor. Since MOSFETs can be configured to act like a resistor, the reference current in MOSFET current sources is usually established by using additional transistor.
- The reference resistance R_1 in fig 7.1 can be replaced by another MOSFET M_3 as shown in fig 8.1.

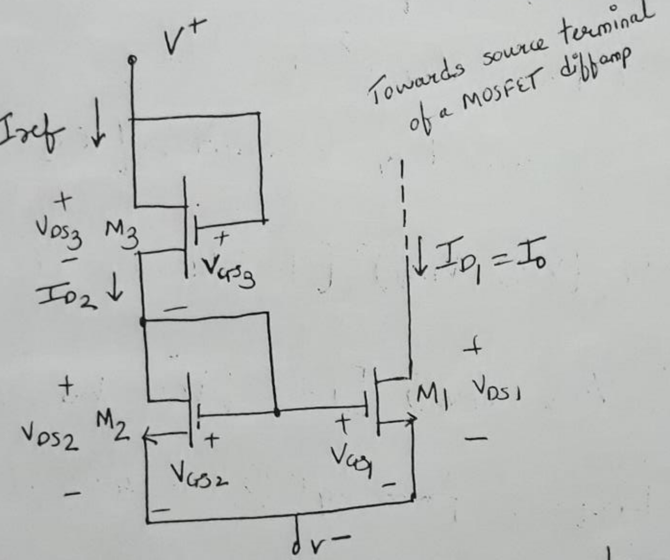

fig 7.1 → Three-transistor MOSFET current source.

- Transistor M_2 and M_3 are used as voltage divider's to control the gate-source voltage of M_1.
- If M_1 and M_2 are identical, the o/p current (I_0) exactly mirrors the drain current through M_2 and M_3

$$\underline{I_0 \approx I_{D_1}} \quad \text{(From fig 7.1)}$$

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Current Relationship:

Since $V_{GS_1} = V_{GS_2}$, from fig (7.1)

The drain current $I_{D_1} (= I_O)$ is equal to the drain current I_{D_2} and is given by

$$I_{D_1} = I_{D_2} = I_{ref} = K_{n_1}(V_{GS_1} - V_{TN_1})^2 (1 + \lambda V_{GS_1}) \quad — ①$$

Since $V_{DS_2} = V_{GS_2} \rightarrow I_{DS}$ is equal to reference current I_R,

$$I_{D_2} = I_{Ref} = K_{n_2}(V_{GS_2} - V_{TN_2})^2 (1 + \lambda V_{GS_2}) \quad — ②$$

Since $V_{GS_3} = V^+ - V_{GS_2}$, the drain current of M_3 is

$$I_{D_3} = I_{ref} = K_{n_3}(V_{GS_3} - V_{TN_3})^2 [1 + \lambda V_{DS_3}] \quad — ③$$

$$I_{D_3} = K_{n_3}(V^+ - V_{GS_2} - V_{TN_3})^2 [1 + \lambda(V^+ - V_{GS_2})] \quad — ④$$

Since $I_{D_2} = I_{D_3} = I_{Ref}$, from eq^n ② & ④, we get

$$\frac{K_{n_2}(V_{GS_2} - V_{TN_2})^2 (1 + \lambda V_{GS_2})}{K_{n_3}(V^+ - V_{GS_2} - V_{TN_3})^2 [1 + \lambda(V^+ - V_{GS_2})]} = 1 \quad — ⑤$$

Thus, by controlling the constants K_{n_2} & K_{n_3}, we can obtained the desired value of $V_{GS_2} = V_{GS_1}$, which will give the desired o/p current (I_O) (from eq^n ①).

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Finally the Load current, for $\lambda = 0$, is given by

$$\boxed{I_O = \frac{k_n'}{2}\left(\frac{W}{L}\right)_1 (V_{GS_1} - V_{TN_1})^2} \quad — (6)$$

where $k_n' = \mu_n C_{ox}$

- Since the designer has control over the (W/L) ratios of transistors, there is considerable flexibility in the design of MOSFET current sources.

- O/P resistance (R_o):

→ The o/p resistance of three-transistor MOSFET constant current source can be determined by looking into O/P transistor M_1.

→ It is similar to two-transistor's o/p resistance

$$\therefore \boxed{R_o = r_{o_1} = \frac{1}{\lambda I_{D_1}}}$$

which is relatively small.

- This small o/p resistance is a disadvantage of having only one MOSFET M_1 at the o/p side of a current source.

→ The o/p resistance of a MOSFET basic current source & modified MOSFET current source can increased by connecting MOSFETs in a cascode-like connection.

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c) Cascode Current Source :-

→ The O/P resistance of a two-transistor MOSFET current source can be ↑sed by adding two more MOSFETs in a cascode-like connection, as shown in fig 8.1.

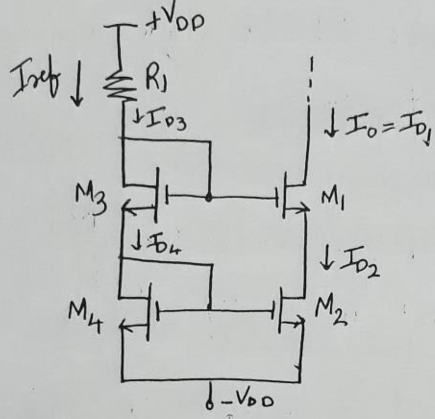

fig 8.1 Cascode Current Source

- Assuming all transistor's are identical. (ie M_3 & M_1 and M_4 & M_2), we can write $I_{ref} \simeq I_{D3} \approx I_{D4}$

Also, $I_{D4} = I_{D2} = I_{D1}$ ie $I_{D1} = I_0$

$\therefore$ $\boxed{I_0 \simeq I_{ref}}$

Thus, O/P load current I_0 is approximately equal to the reference current I_{ref}.

O/P resistance (R_o) :

To determine the O/P resistance at the drain of M_1, we use the small-signal equivalent ckt. fig 8.3.

Since, I_{ref} is a constant, which make gate voltage to M_1 & M_2 and M_3 & M_4 are constant. This is equivalent to an ac short ckt.

• The ac equivalent ckt for calculating the O/P resistance is shown in fig 8.2

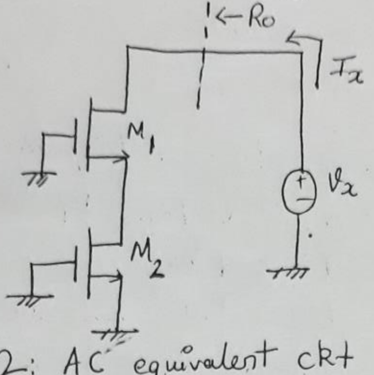

fig 8.2: AC equivalent ckt

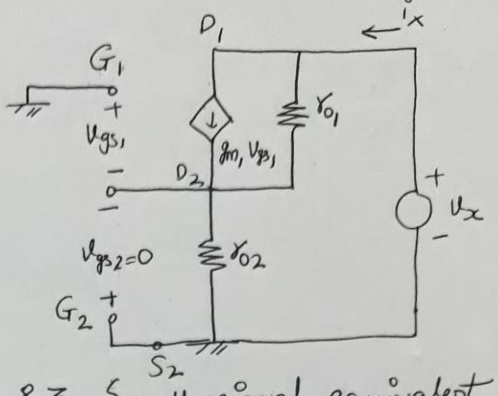

fig 8.3: Small-signal equivalent circuit.

• r_{o_2} is the O/P resistance of transistor M_2.

$$V_{gs_1} = -r_{o_2} i_x \quad —(1)$$

Also, from fig 8.1, and fig 8.3, we get

$$V_x = r_{o_1}(i_x - g_{m_1} V_{gs_1}) + r_{o_2} i_x$$

$$V_x = r_{o_1}\left[i_x + g_{m_1} r_{o2} i_x\right] + r_{o_2} i_x$$

ie $V_x = i_x\left[r_{o_1}(1 + g_{m_1} r_{o2}) + r_{o2}\right]$

ie $R_o = \dfrac{V_x}{i_x} = r_{o1}(1 + g_{m_1} r_{o2}) + r_{o2}$

For identical transistors, $r_{o_1} = r_{o_2} = r_o$ and R_o becomes

$$R_o = r_o(2 + g_{m_1} r_o)$$

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$\boxed{R_o \approx g_{m_1} r_o^2}$ ----→ (O/P resistance of Cascode Current source)

Thus, the O/P resistance can be significantly increased, to a level comparable to that of a BJT current source & indeed, then its two-transistor MOSFET current source counterpart.

D) Wilson Current Source : (MOSFET version)

- The MOSFET version of the Wilson current source is shown in fig 9.1(a).

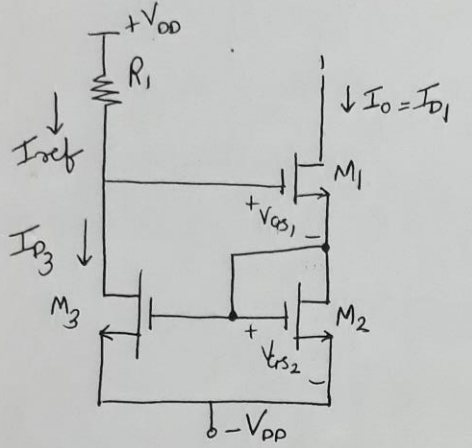

fig: 9.1(a) Wilson Current Source.

- In this configuration fig 9.1(a), drain voltages V_{D_1} and V_{D_3} of M_1 and M_3 are unequal. As a result, their drain currents I_{D_1} and I_{D_3} are also unequal.

→ This problem can be solved by adding one diode-connected MOSFET M_4 as shown in fig 9.2.

- This modification ensures that M_1 and M_3 have equal drain voltages and thus equal drain currents ie $\boxed{I_0 \simeq I_{ref}}$

- O/P resistance (R_0):
  - R_0 of Wilson current source in fig 9.1(a) can be determined by its small-s/g equivalent ckt. in fig 9.1(b) R_0 can be found out as,

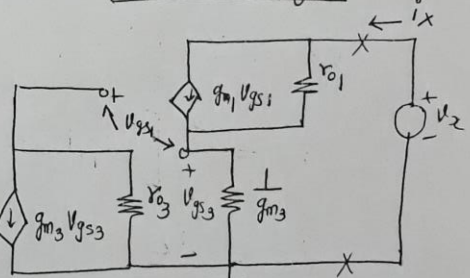

fig 9.1(b) Equivalent ckt for finding R_0.

- $R_0 = \frac{v_x}{i_x} \simeq r_{o_1} + r_{o_1}(1 + g_{m_3} r_{o_3})$

Assuming identical transistor's

$g_{m_1} = g_{m_2} = g_{m_3} = g_m$

$\therefore$ $\boxed{R_0 \simeq r_{o_1} + r_{o_1}(1 + g_m r_{o_3})}$

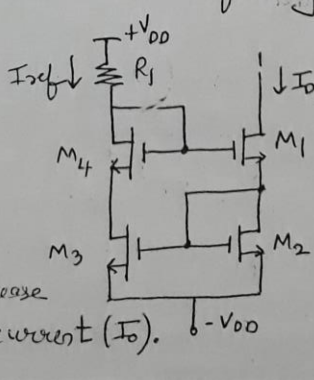

fig 9.2 Modified Wilson current source

Primary advantage of these ckts is the increase in R_0, which further stabilizes the Load current (I_0).