\[ V_+ = 250 \text{ mV} \]
\[ A_{oc} = 2 \]
\[ V_o = 500 \text{ mV} \]
\[ \text{error} \leq 1 \text{ mV} \]

Then for an opamp in unity gain feedback:

\[ V_o = \frac{A_{oc} V_+}{1 + A_{oc} B} \]
\[ \frac{V_o - \text{error}}{V_+} = \frac{A_{oc}}{1 + A_{oc} B} \]

\[ \frac{499 \text{ mV}}{250 \text{ mV}} = \frac{A_{oc}}{1 + A_{oc} B} \]

\[ \frac{1.996}{1 + A_{oc} B} = A_{oc} \]

\[ 1.996 = A_{oc} B = \frac{1.996}{0.002} \]

\[ \Rightarrow A_{oc} = 998 \text{ for error} < 1 \text{ mV} \]

\[ \Rightarrow A_{oc} > 998 \text{ for error} < 1 \text{ mV} \]
Fig 24.59 is simulated and OP analysis are obtained to confirm the lack of control on $I_{bias}$ of the output stage.

By not controlling the current in the output stage we introduce a systematic offset at the input of the op-amp. Because, as seen in Fig 1, if $V_g$ source

$$V_g \rightarrow M_2 \rightarrow V_i$$
$$V_{bias} \rightarrow M_1$$

more current than what current source $M_2$ can sink, $M_2$ goes into triode which causes an offset in the previous stage which when referred is called input referred systematic offset.

Lack of control of output current in the simulated results.
Q#2: Output stage I_{bias} changed a lot when \( V_{DD} = 1.2 \) from

no control on I_{bias} output...

\[ I = 231.3 \mu A \]

Q#2: Output stage I_{bias} for \( V_{DD} = 1V \)

\[ I = 66.73 \mu A \]
Q #3 10 AC response from 1k to 1G

Following from Q#2, current control over output is implemented. Here I provide the AC response of the op-amp.

This figure is followed by a transient waveform, and then the schematic figures which show control over output stage bias current under Vgs variation which also results in minimizing input-referred systematic offset explained in Q#2.

Q #3 Transient Response
Q#3  Output $I_{\text{bias}}$ for $V_{dd} = 1\text{V}$

The current in the output stage is now under control.

$I = 17.539\text{uA}$
Q# 4

Input common mode range

Since
\[ V_{50\%} \rightarrow -V_{50\%} \]

then

\[ V_{\text{in(max)}} = V_{DD} - V_{SG3} + V_{SGM2} - V_{SDM1,\text{sat}} + V_{SGM2} \]
\[ = V_{DD} - V_{SG3} - V_{SGM2} + V_{PM} + V_{SGM2} \]
\[ = V_{DD} - V_{SG3} + V_{PM} \]

Using values from Table 9.2 gives

\[ V_{\text{in(max)}} = 1.1350 \text{mV} + 2.80 \text{mV} \]
\[ = 0.93 \text{V} \]

\[ V_{\text{in(min)}} = V_{SS} + V_{SG3} + V_{DG4,\text{sat}} + V_{SGM2} \]
\[ = 0 - 0.05 + 0.05 + 0.35 \]
\[ = 0.45 \text{V} \]

Output swing range

\[ V_{\text{out(max)}} = V_{DD} - V_{SD7,\text{sat}} \]
\[ = 0.95 \text{V} \]

\[ V_{\text{out(min)}} = V_{SS} + V_{DSM8} + V_{DSM8,\text{sat}} \]
\[ = 0 + 0.05 + 0.05 \]
\[ = 0.1 \text{V} \]
To have $V_{out}$ balanced at 500 mV we need $I_{M7}$ to match $I_{M8}$ if $I_{V8}$.

Any mismatch or systematic offset would take $V_{out}$ to either $V_{DD}$ rail or $V_{SS}$ (ground rail).

* If $I_{M7} > I_{M8}$
  
  then $V_{out} = V_{DD} \Rightarrow M_7$ is larger than $M_8$

* If $I_{M8} > I_{M7}$
  
  $V_{out} \approx 0 \Rightarrow M_8$ is larger than $M_7$

This is verified in the attached simulation results!
Q#4: DC sweep which shows \( V_{out} \) reaches \( \sim 0V \) when \( V_{in} \sim 350mV \)
Q#4: \( W(M_4) \uparrow \) which takes \( M_4 \) into triode

\( W(M_{13}) \)

and \( V_{out} \approx V_{DD} \)

\[ V_{out} = 898.7 \text{ mV} \]

Q#4: \( V_{out} \approx GND \) as expected

\[ V_{in} \]

\[ V_{out} = 57.84 \text{ mV} \]
$$A_{OLC} = g_{m12} \times \left( \frac{V_{dd} + 11 \times V_{dd2}}{g_{m7}} \right) \times g_{m4} \left( \frac{V_{dd}}{G_{ds1}} \right)$$

$$\approx g_{m2} \times \left( \frac{V_{dd}}{G_{ds2}} \right) \times g_{m4} \times \left( \frac{V_{dd}}{G_{ds7}} \right)$$

$$\approx 150 \mu \text{A} \times \left( \frac{111.1 \text{k}}{150 \mu \text{A}} \right) \times 383 \mu \text{A} \times \left( \frac{383 \text{k}}{24.04 \text{fF}} \right)$$

$$\approx 832.5 \approx 58.4 \text{ dB}$$

$$f_{3dB} = \frac{f_{PDominant}}{R_1 R_2 C_C g_m}$$

$$\approx \frac{1}{111.1 \text{k} \times 383 \mu \text{A} \times 24.04 \text{fF} \times 150 \mu \text{A}}$$

$$\approx 750.8 \text{ kHz}$$

$$f_{min} = A_{OLC} \times f_{3dB} \text{ I think } f_{min} \text{ can be computed directly}$$

$$\approx 625 \text{ MHz}$$

From simulations, we get \( f_{min} \approx 350 \text{ MHz} \)

The difference is because of the fact that when a single MOS is divided into two as done here then the lower MOS operates in the trade region, hence \( R_1 \) decreases which results in a lower \( f_{min} \).
Q#5 a) AC Response

Q # 5(b) 

Gain appears to drop
Q#5 (b)  

- Gain drop delayed when C is increased.
Slight drop in voltage due to gate leakage current.
Input CMR of Fig 24.29

\[ V_{\text{in(max)}} = V_{DD} - V_{SGB3} - V_{DS1T(sat)} + V_{DS2(sat)} + V_{SMB2} \]
\[ = 1 - 0.35 - 0.25 - 0.05 + 0.35 \]
\[ = 0.9 \text{ V} \]

\[ V_{\text{in(min)}} = V_{SS} + V_{DS6(sat)} + V_{DS6L(sat)} + V_{SGM1} \]
\[ = 0 + 0.05 + 0.05 + 0.35 \]
\[ = 0.45 \text{ V} \]

Input CMR of given op-amp

\[ V_{\text{in(max)}} = V_{DD} - V_{SGB3} - V_{SGM1T} + V_{DS2(sat)} + V_{SMB1} \]
\[ = 1 - 0.35 - 0.35 + 0.05 + 0.35 \]
\[ = 0.7 \text{ V} \]

\[ V_{\text{in(min)}} = V_{SS} + V_{DS6(sat)} + V_{DS6L(sat)} + V_{SGM1} \]
\[ = 0.05 + 0.05 + 0.35 \]
\[ = 0.45 \text{ V} \]

The topology in the question removes the center MOS in the diff pair which decrease the quiescent current of the op-amp. Hence power saving.
Maintains CM through two Isink branches in the diff pair!
b) \( f_1: \)

For Miller compensation due to \( C_c \)

\[
\omega_1 \cong \frac{1}{\pi [C_c + C_2]. R_2 + (C_1 + C_c(1+g_m R_2)). R_2]
\]

\( R_1 \) is the output \( R \) of first stage which is a telescopic diff. pair

\( R_2 \) is the output \( R \) of 2nd output stage

\[
A_{OL DC} = g_m (R_4) (g_{m1} g_{m1}) (V_{ol} / 2)
\]

\[
V_{dd} \quad V_{dd}
\]

\[
M_1 \quad M_2
\]

\[
R_1 \quad C_1 \quad C_2 \quad C_3
\]

\[
R_2 \quad R_3
\]

\[
V_{p} \quad V_{p}
\]

\( R_1, R_2, \) \( R_3 \) are degenerated resistances

\[
\Rightarrow \quad R_1 = g_m \frac{V_{dd}}{2}
\]

\[
R_2 = g_m \frac{V_{dd}}{2}
\]

\[
R_3 = g_m \frac{V_{dd}}{2}
\]

Using values from Table 9.2

\[
A_{OL DC} = 150 \mu (150 \mu \times (333 \text{X})^2) / 150 \mu (167 \text{K})^2
\]

\[
\times 150 \mu \times (333 \text{X}) / 167 \text{K}
\]

\[
\approx 150 \mu (333 \times 66 \text{X}) / 1118 \text{X}
\]

\[
\times 150 \mu (111.2 \mu)
\]

\[
= 62.5 \text{dB}
\]
\[ f_1 = \frac{1}{2\pi R_1 R_2 C C \text{gm}} \]
\[ = \frac{1}{\text{gm}} \left( \text{gm} - \text{\text{gm}^2} \cdot \frac{\text{nm} \cdot \text{m}^2}{\text{nm} \cdot \text{m}^2} \right) \times C C \times \left( \text{gm} \cdot \text{\text{gm}^2} \cdot \frac{\text{nm} \cdot \text{m}^2}{\text{nm} \cdot \text{m}^2} \right) \]
\[ = \frac{1}{150 \mu (1663350 \times 14183350) \times 2.4 \times F \times (111.1 \text{ k})} \]
\[ \approx \frac{1}{74 \text{ kHz}} \approx 11.9 \text{ kHz} \]

\[ f_2 \text{ would be at the output} \]
\[ \text{total cap} = C_L + C_C \text{ at the output} \]
\[ \text{from indirect compensation we have} \]
\[ f_2 \approx \frac{\text{gm} \cdot C_C \times 10}{2\pi C_C (C_L + C_C)} \]
\[ \approx \frac{150 \mu \times 2.4 \times 10^{-12}}{2 \pi \times 12.5 \times F \times (240 \times 10^{-6} + 10^{-6})} \approx 12.5 \text{ kHz} \]

\[ f_2 \approx \frac{\text{gm} \cdot C_L}{2\pi C_C} \approx 458.59 \mu \times 10 \]
\[ \approx 440.78 \text{ kHz} \]

\[ f_2 \text{ gain through common-gate} \approx \frac{1}{2 \pi 2.46 \times 10^{-12}} = \text{ very high in } 100 \text{ s of MHz} \]

\[ f_m \approx \frac{\text{gm}}{2\pi C_C} \]
\[ \approx \frac{200}{\text{MHz}} \]
C) As M4 are the current sources it is not a good idea to put feedback compensation in between two current source (FMOs). This may result in non-compensation and very low PM as seen in the figures attached!
PM is lost considerably as compensation is not suitable

$f_2 \approx 10 \text{kHz}$

$f_z = 200 \text{kHz}$ in the band pass range as expected

$f_{un} = 100 \text{kHz}$ as calculated