LECTURE 27 – HIGH SPEED OP AMPS

LECTURE ORGANIZATION

Outline

- Extending the GB of conventional op amps
- Cascade Amplifiers
  - Voltage amplifiers
  - Voltage amplifiers using current feedback
- Summary

***CMOS Analog Circuit Design*, 3rd Edition Reference**

Pages 370-386

INCREASING THE GB OF OP AMPS

What is the Influence of *GB* on the Frequency Response?

The unity-gainbandwidth represents a limit in the trade-off between closed loop voltage gain and the closed-loop -3dB frequency.

Example of a gain of -10 voltage amplifier:

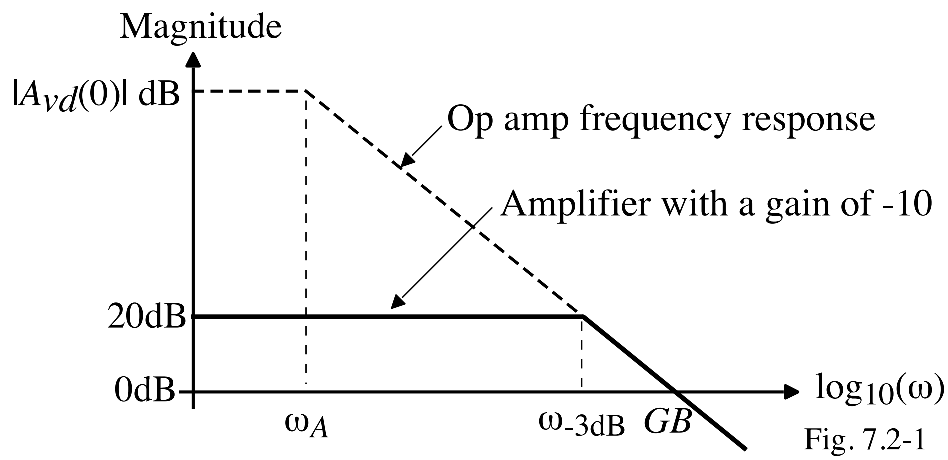


Fig. 7.2-1

What defines the *GB*?

We know that

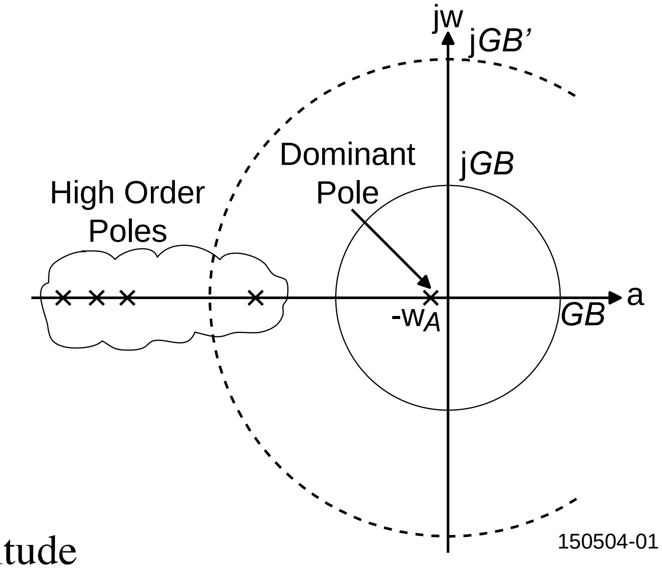
$$GB = \frac{g_m}{C}$$

where g_m is the transconductance that converts the input voltage to current and C is the capacitor that causes the dominant pole.

This relationship assumes that all higher-order poles are greater than *GB*.

What is the Limit of *GB*?

The following illustrates what happens when the next higher pole is not greater than *GB*:



For a two-stage op amp, the poles and zeros are:

1.) Dominant pole $p_1 = \dfrac{-g_{m1}}{A_v(0)C_c}$

2.) Output pole $p_2 = \dfrac{-g_{m6}}{C_L}$

3.) Mirror pole $p_3 = \dfrac{-g_{m3}}{C_{gs3}+C_{gs4}}$

and $z_3 = 2p_3$

4.) Nulling pole $p_4 = \dfrac{-1}{R_z C_I}$

5.) Nulling zero $z_1 = \dfrac{-1}{R_z C_c - (C_c/g_{m6})}$

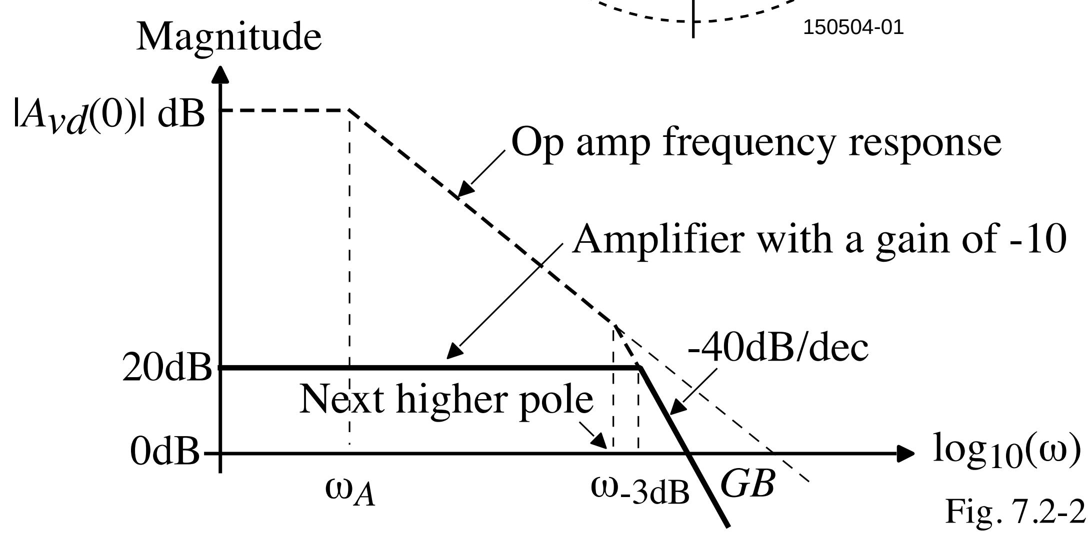
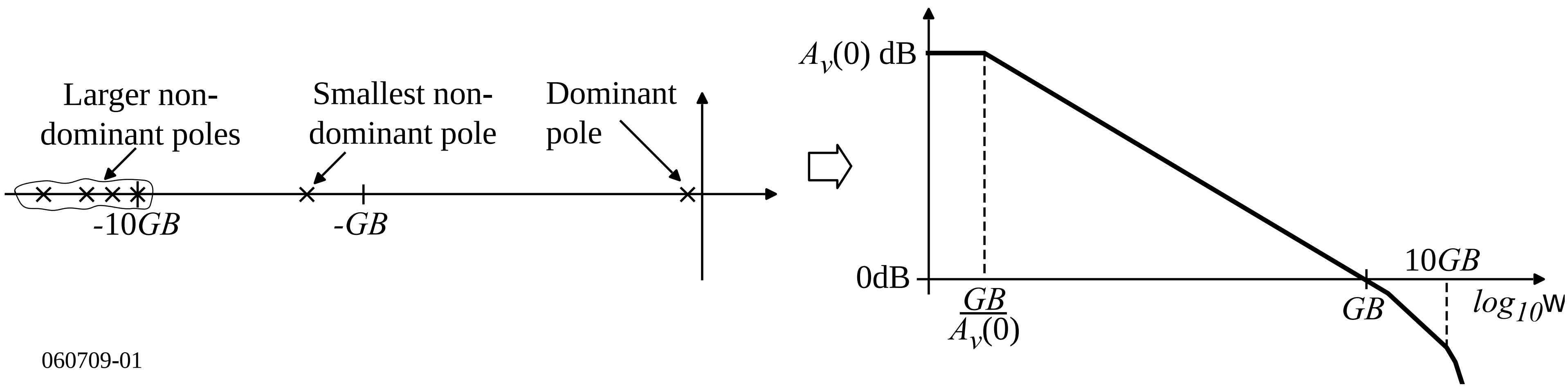


Fig. 7.2-2

Higher-Order Poles

For reasonable phase margin, the smallest higher-order pole should be 2-3 times larger than *GB* if all other higher-order poles are larger than 10*GB*.



If the higher-order poles are not greater than 10*GB*, then the distance from *GB* to the smallest non-dominant pole should be increased for reasonable phase margin.

Increasing the *GB* of a Two-Stage Op Amp

1.) Use the nulling zero to cancel the closest pole beyond the dominant pole.

2.) The maximum *GB* would be equal to the magnitude of the second closest pole beyond the dominant pole.

3.) Adjust the dominant pole so that $2.2GB \approx$ (second closest pole beyond the dominant pole)

Illustration which assumes that p_2 is the next closest pole beyond the dominant pole:

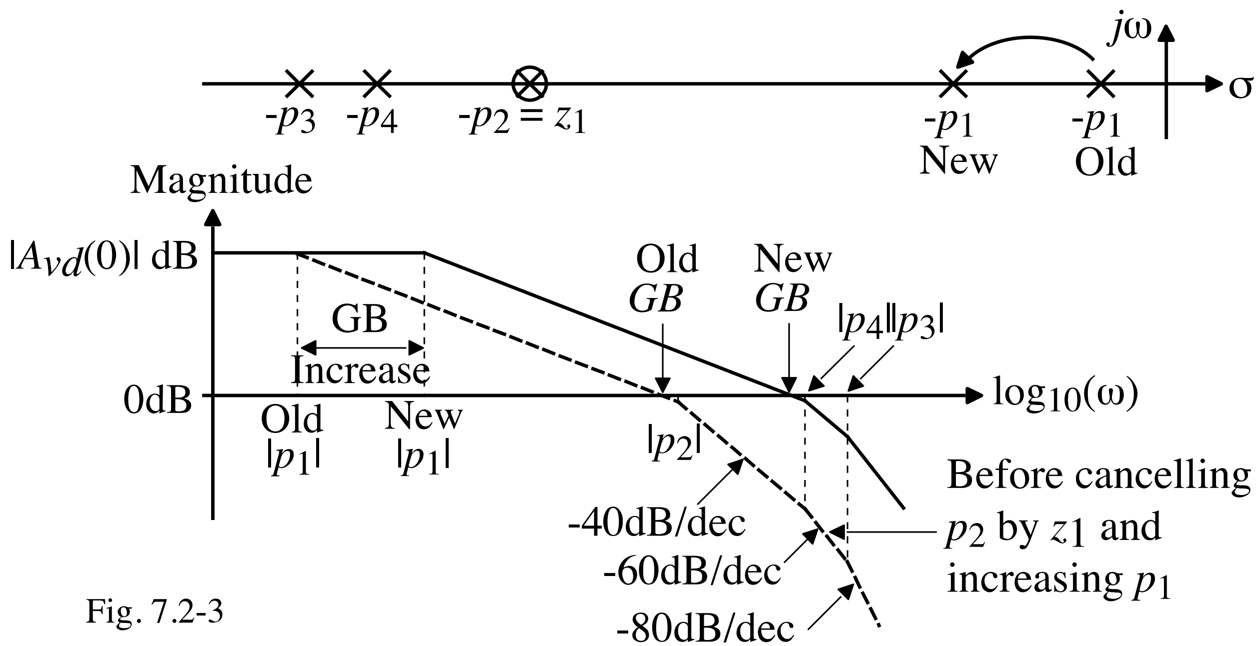
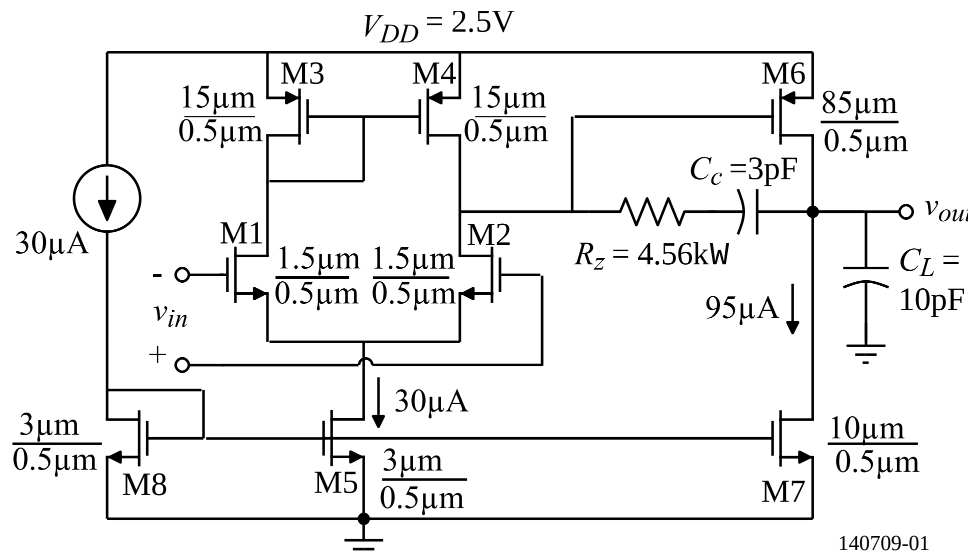


Fig. 7.2-3

Example 27-1 - Increasing the *GB* of the Op Amp Designed in Ex. 23-1

Use the two-stage op amp designed in Examples 23-1 and 23-2 and apply the above approach to increase the gainbandwidth as much as possible. Use the capacitor values in the table shown along with C_{ox} = 6fF/μm^2.



Type	P-Channel	N-Channel	Units
CGSO	220×10^{-12}	220×10^{-12}	F/m
CGDO	220×10^{-12}	220×10^{-12}	F/m
CGBO	700×10^{-12}	700×10^{-12}	F/m
CJ	560×10^{-6}	770×10^{-6}	F/m^2
CJSW	350×10^{-12}	380×10^{-12}	F/m
MJ	0.5	0.5	
MJSW	0.35	0.38	

Solution

1.) First find the values of p_2, p_3, and p_4.

a.) From Ex. 23-2, we see that

p_2 = -95x10^6 rads/sec.

b.) p_3 was found in Ex. 23-1 as

p_3 = -1.25x10^9 rads/sec. (also there is a zero at -2.5x10^9 rads/sec.)

Example 27-1 - Continued

(c.) To find p_4, we must find C_I which is the output capacitance of the first stage of the op amp. C_I consists of the following capacitors,

$$C_I = C_{bd2} + C_{bd4} + C_{gs6} + C_{gd2} + C_{gd4}$$

For C_{bd2} the width is 1.5µm $\Rightarrow$ L1+L2+L3=3µm $\Rightarrow$ AS/AD=4.5µm^2 and PS/PD=9µm.
For C_{bd4} the width is 15µm $\Rightarrow$ L1+L2+L3=3µm $\Rightarrow$ AS/AD=45µm^2 and PS/PD=36µm.
From Table 3.2-1:

$$C_{bd2} = (4.5\mu m^2)(770\text{x}10^{-6}\text{F/m}^2) + (9\mu m)(380\text{x}10^{-12}\text{F/m}) = 3.47\text{fF}+3.42\text{fF} \approx 6.89\text{fF}$$

$$C_{bd4} = (45\mu m^2)(560\text{x}10^{-6}\text{F/m}^2) + (36\mu m)(350\text{x}10^{-12}\text{F/m}) = 25.2\text{fF}+12.6\text{F} \approx 37.8\text{fF}$$

C_{gs6} in saturation is,

$$C_{gs6}=CGDO\cdot W_6+0.67(C_{ox}W_6L_6)=(220\text{x}10^{-12})(85\text{x}10^{-6})+(0.67)(6\text{x}10^{-15})(42.5)$$

$$= 18.7\text{fF} + 255\text{fF} = 273.7\text{fF}$$

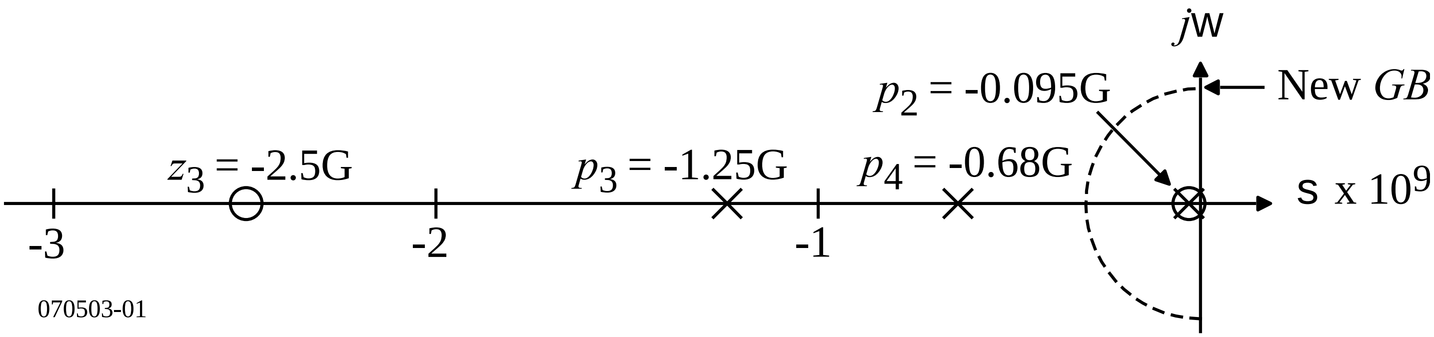
$$C_{gd2} = 220\text{x}10^{-12}\text{x}1.5\mu m = 0.33\text{fF and } C_{gd4} = 220\text{x}10^{-12}\text{x}15\mu m = 3.3\text{fF}$$

Therefore, C_I = 6.9fF + 37.8fF + 273.7fF + 0.33fF + 3.3fF = 322fF. Although C_{bd2} and C_{bd4} will be reduced with a reverse bias, let us use these values to provide a margin. Thus let C_I be 322fF.

In Ex. 23-2, R_z was 4.564kΩ which gives p_4 = - 0.680x10^9 rads/sec.

Example 27-1 - Continued

Therefore, the roots are:



When p_2 is cancelled, the next smaller pole is p_4 which will define the new *GB*. 2.) Using the nulling zero, z_1, to cancel p_2, gives p_4 as the next smallest pole. For 60° phase margin $GB = |p_4|/2.2$ if the next smallest pole is more than 10*GB*.

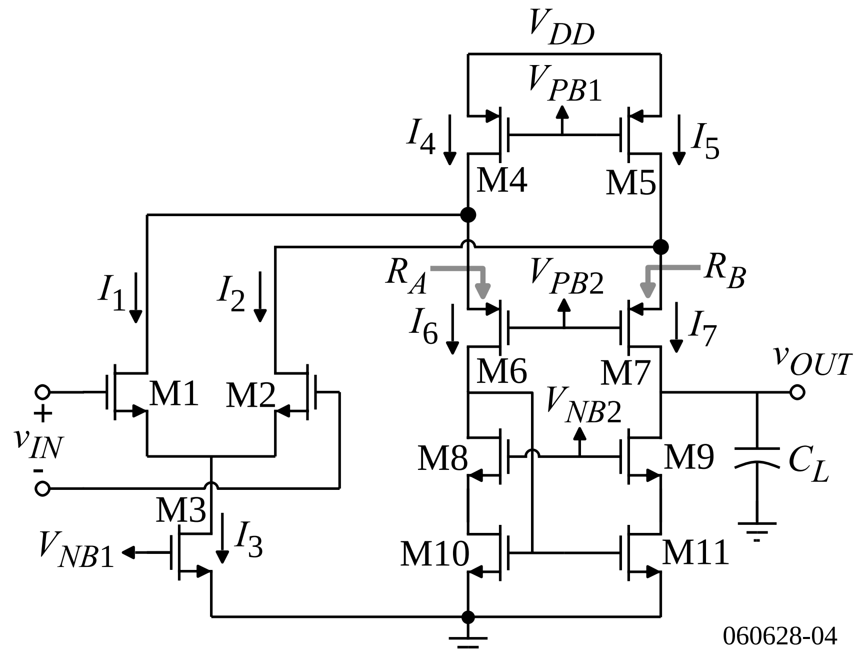
$$\therefore \quad GB = 0.680\text{x}10^9/2.2 = 0.309\text{x}10^9 \text{ rads/sec. or 49.2MHz.}$$

This value of *GB* is designed from the relationship that $GB = g_{m1}/C_c$. Assuming g_{m1} is constant, then $C_c = g_{m1}/GB = (94.25\text{x}10^{-6})/(0.309\text{x}10^9) = 307\text{fF}$. It might be useful to increase g_{m1} in order to keep C_c above the surrounding parasitic capacitors (C_{gd6} = 18.7fF). The success of this method assumes that there are no other roots with a magnitude smaller than 10*GB*.

The result of this example is to increase the *GB* from 5MHz to 49MHz.

Example 27-2 - Increasing the *GB* of the Folded Cascode

Use the folded-cascode op amp designed in Example 24-4 and apply the above approach to increase the gainbandwidth as much as possible. Assume that the drain/source areas are equal to 2μm times the width of the transistor and that all voltage dependent capacitors are at zero voltage.



Solution

The poles of the folded cascode op amp are:

$$p_A \approx \frac{-1}{R_A C_A} \quad \text{(the pole at the source of M6)}$$

$$p_B \approx \frac{-1}{R_B C_B} \quad \text{(the pole at the source of M7)}$$

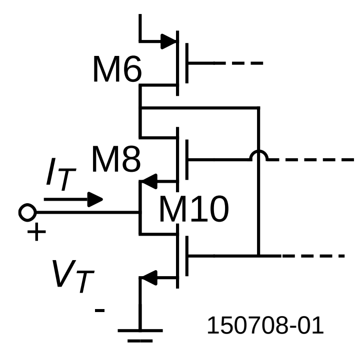
$$p_6 \approx \frac{-g_{m10}}{C_6} \quad \text{(the pole at the drain of M6)}$$

$$p_8 \approx \frac{-g_{m8} r_{ds8} g_{m10}}{C_8} \quad \text{(the pole at the source of M8)}$$

$$p_9 \approx \frac{-g_{m9}}{C_9} \quad \text{(the pole at the source of M9)}$$

$$I_T = g_{m8} V_T r_{ds8} g_{m10}$$

$$R_8 = \frac{V_T}{I_T} = \frac{1}{g_{m8} r_{ds8} g_{m10}}$$



Example 27-2 - Continued

Let us evaluate each of these poles.

1.) For p_A, the resistance R_A is approximately equal to g_{m6} and C_A is given as

$$C_A = C_{gs6} + C_{bd1} + C_{gd1} + C_{bd4} + C_{bs6} + C_{gd4}$$

From Ex. 24-4, g_{m6} = 774.6µS and capacitors giving C_A are found as,

$$C_{gs6} = (220\text{x}10^{-12}\cdot 80\text{x}10^{-6}) + (0.67)(80\mu\text{m}\cdot 0.5\mu\text{m}\cdot 6\text{fF}/\mu\text{m}^2) = 177.6\text{fF}$$

$$C_{bd1} = (770\text{x}10^{-6})(16.5\text{x}10^{-6}\cdot 2\text{x}10^{-6}) + (380\text{x}10^{-12})(37\text{x}10^{-6}) = 39.5\text{fF}$$

$$C_{gd1} = (220\text{x}10^{-12}\cdot 16.5\text{x}10^{-6}) = 3.6\text{fF}$$

$$C_{bd4} = C_{bs6} = (560\text{x}10^{-6})(80\text{x}10^{-6}\cdot 2\text{x}10^{-6}) + (350\text{x}10^{-12})(2\cdot 82\text{x}10^{-6}) = 147\text{fF}$$

and

$$C_{gd4} = (220\text{x}10^{-12})(80\text{x}10^{-6}) = 17.6\text{fF}$$

Therefore,

$$C_A = 177.6\text{fF} + 39.5\text{fF} + 3.6\text{fF} + 147\text{fF} + 17.6\text{fF} + 147\text{fF} = 0.532\text{pF}$$

Thus,

$$p_A = \frac{-774.6\text{x}10^{-6}}{0.532\text{x}10^{-12}} = -1.456\text{x}10^{9} \text{ rads/sec.}$$

2.) For the pole, p_B, the capacitance connected to this node is

$$C_B = C_{gd2} + C_{bd2} + C_{gs7} + C_{gd5} + C_{bd5} + C_{bs7}$$

Example 27-2 - Continued

The various capacitors above are found as

$$C_{gd2} = (220\text{x}10^{-12}\cdot 16.5\text{x}10^{-6}) = 3.6\text{fF}$$

$$C_{bd2} = (770\text{x}10^{-6})(16.5\text{x}10^{-6}\cdot 2\text{x}10^{-6}) + (380\text{x}10^{-12})(37\text{x}10^{-6}) = 39.5\text{fF}$$

$$C_{gs7} = (220\text{x}10^{-12}\cdot 80\text{x}10^{-6}) + (0.67)(80\mu\text{m}\cdot 0.5\mu\text{m}\cdot 6\text{fF}/\mu\text{m}^2) = 177.6\text{fF}$$

$$C_{gd5} = (220\text{x}10^{-12})(80\text{x}10^{-6}) = 17.6\text{fF}$$

and

$$C_{bd5} = C_{bs7} = (560\text{x}10^{-6})(80\text{x}10^{-6}\cdot 2\text{x}10^{-6}) + (350\text{x}10^{-12})(2\cdot 82\text{x}10^{-6}) = 147\text{fF}$$

The value of C_B is the same as C_A and g_{m6} is assumed to be the same as g_{m7} giving $p_B = p_A = -1.456\text{x}10^9$ rads/sec.

3.) For the pole, p_6, the capacitance connected to this node is

$$C_6 = C_{bd6} + C_{gd6} + C_{gs10} + C_{gs11} + C_{bd8} + C_{gd8}$$

The various capacitors above are found as

$$C_{bd6} = (560\text{x}10^{-6})(80\text{x}10^{-6}\cdot 2\text{x}10^{-6}) + (350\text{x}10^{-12})(2\cdot 82\text{x}10^{-6}) = 147\text{fF}$$

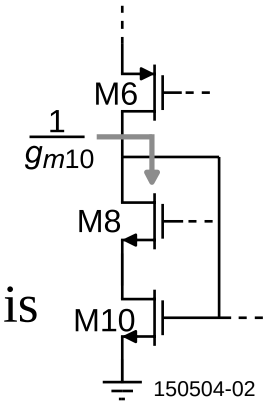
$$C_{gs10} = C_{gs11} = (220\text{x}10^{-12}\cdot 10\text{x}10^{-6}) + (0.67)(10\mu\text{m}\cdot 0.5\mu\text{m}\cdot 6\text{fF}/\mu\text{m}^2) = 22.2\text{fF}$$

$$C_{bd8} = (770\text{x}10^{-6})(10\text{x}10^{-6}\cdot 2\text{x}10^{-6}) + (380\text{x}10^{-12})(2\cdot 12\text{x}10^{-6}) = 24.5\text{fF}$$

$C_{gd8} = (220\text{x}10^{-12})(10\text{x}10^{-6}) = 2.2\text{fF}$ and $C_{gd6} = C_{gd5} = 17.6\text{fF}$

Therefore, $C_6 = 147\text{fF} + 17.6\text{fF} + 22.2\text{fF} + 22.2\text{fF} + 2.2\text{fF} + 17.6\text{fF} = 0.229\text{pF}$

Example 27-2 - Continued



From Ex. 24-4, g_{m10} = 600x10^{-6}. Therefore, p_6, can be expressed as

$$-p_6 = \frac{600\text{x}10^{-6}}{0.229\text{x}10^{-12}} = 2.62\text{x}10^9 \text{ rads/sec.}$$

4.) Next, we consider the pole, p_8. The capacitance connected to this node is

$$C_8 = C_{bd10} + C_{gd10} + C_{gs8} + C_{bs8}$$

These capacitors are given as,

$$C_{bs8} = C_{bd10} = (770\text{x}10^{-6})(10\text{x}10^{-6}\cdot 2\text{x}10^{-6}) + (380\text{x}10^{-12})(2\cdot 12\text{x}10^{-6}) = 24.5\text{fF}$$

$$C_{gs8} = (220\text{x}10^{-12}\cdot 10\text{x}10^{-6}) + (0.67)(10\mu\text{m}\cdot 0.5\mu\text{m}\cdot 6\text{fF//}\mu\text{m}^2) = 22.2\text{fF}$$

and

$$C_{gd10} = (220\text{x}10^{-12})(10\text{x}10^{-6}) = 2.2\text{fF}$$

The capacitance C_8 is equal to

$$C_8 = 24.5\text{fF} + 2.2\text{fF} + 22.2\text{fF} + 24.5\text{fF} = 73.4\text{FF}$$

Using the values of Ex. 24-4 of 600µS, the pole p_8 is found as,

$$-p_8 = g_{m8} r_{ds8}\, g_{m10}/C_8 = -600\mu\text{S}\cdot 600\mu\text{S}\cdot/4.5\mu\text{S}\cdot 73.4\text{fF} = -1090\text{x}10^9 \text{ rads/sec.}$$

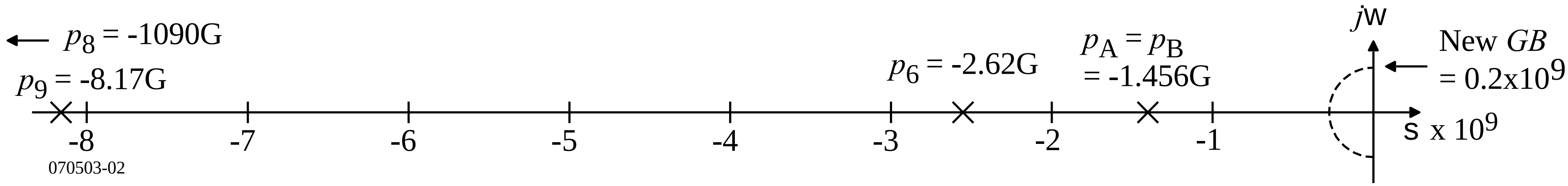
5.) The capacitance for the pole at p_9 is identical with C_8. Therefore, since g_{m9} is 600µS, the pole p_9 is $-p_9 = 8.17\text{x}10^9$ rads/sec.

Example 27-2 - Continued

The poles are summarized below:

p_A = -1.456x10^9 rads/sec p_B = -1.456x10^9 rads/sec p_6 = -2.62x10^9 rads/sec

p_8 = -1090x10^9 rads/sec p_9 = -8.17x10^9 rads/sec



The smallest of these poles is p_A or p_B. Since p_6 is not much larger than p_A or p_B, we will find the new *GB* by dividing p_A or p_B by 4 (which is a guess rather than 2.2) to get 364x10^6 rads/sec. Thus the new *GB* will be 364/2π or 58MHz.

Checking our guess gives a phase margin of,

$$PM = 90° - 2\tan^{-1}(0.364/1.456) - \tan^{-1}(0.364/2.62) = 54° \text{ which is okay}$$

The magnitude of the dominant pole is given as

$$p_{dominant} = GB/A_{vd}(0) = 364\text{x}10^6/3{,}678 = 99{,}000 \text{ rads/sec.}$$

The value of load capacitor that will give this pole is

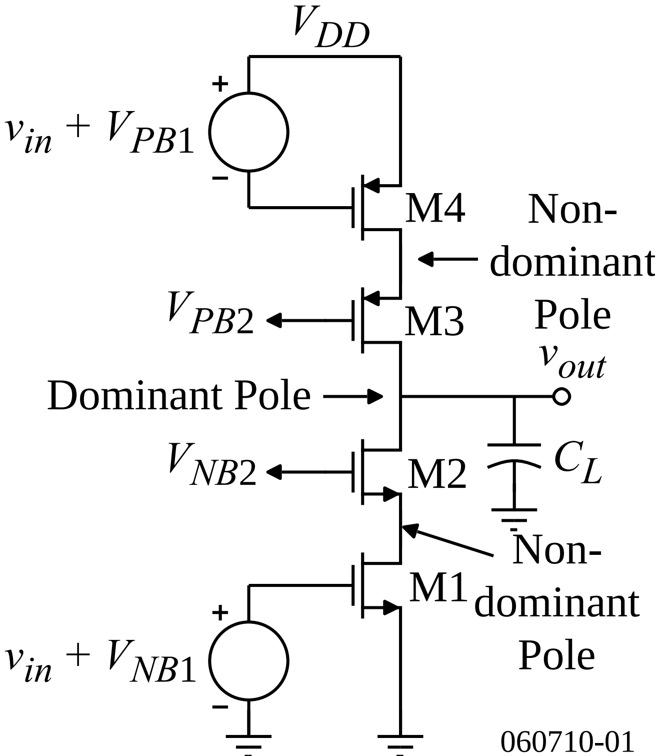
$$C_L = (p_{\text{dominant}}\cdot R_{out})^{-1} = (99\text{x}10^3\cdot 7.44\text{M}\Omega)^{-1} = 1.36\text{pF}$$

Therefore, the new *GB* = 58MHz compared with the old *GB* = 10MHz.

Elimination of Higher-Order Poles

Principle - minimize the number of nodes in the amplifier.

The minimum circuitry for a cascode op amp is shown below:



If the source-drain area between M1 and M2 and M3 and M4 can be minimized, the non-dominant poles will be quite large.

Dynamically Biased, Push-Pull, Cascode Op Amp

Push-pull, cascode amplifier: M1-M2 and M3-M4

Bias circuitry: M5-M6-C_2 and M7-M8-C_1

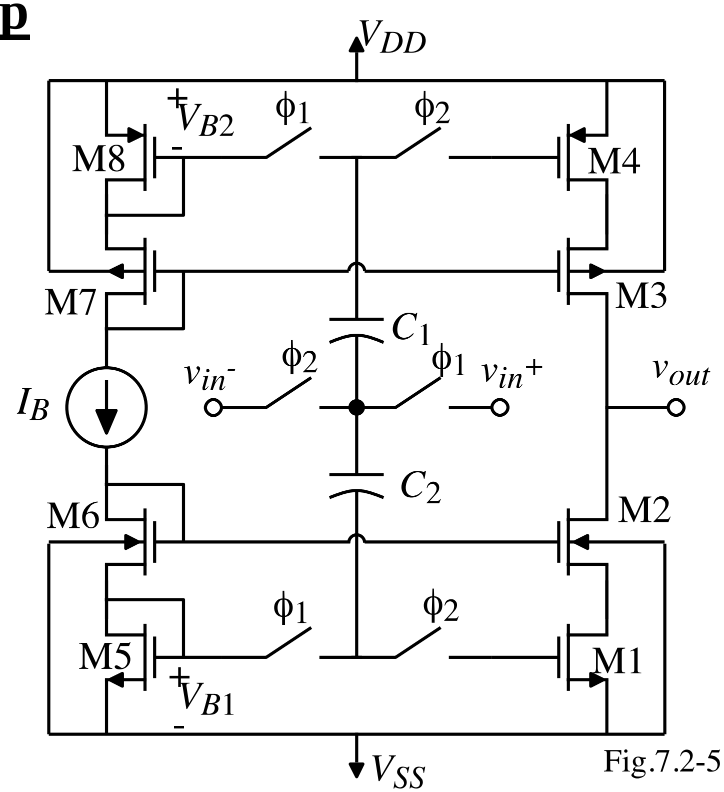
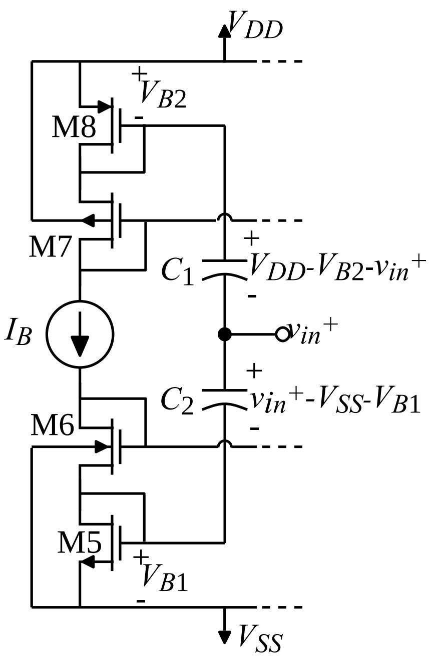


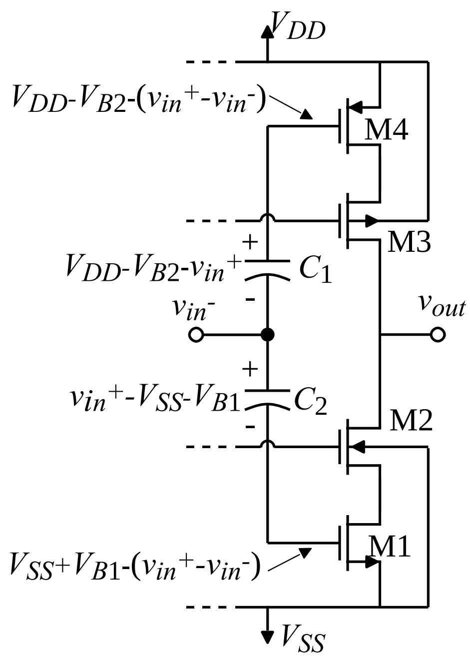
Fig.7.2-5

Operation:



Equivalent circuit during the ϕ_1 clock period

120523-07



Equivalent circuit during the ϕ_2 clock period.

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Dynamically Biased, Push-Pull, Cascode Op Amp - Continued

This circuit will operate on both clock phases† .

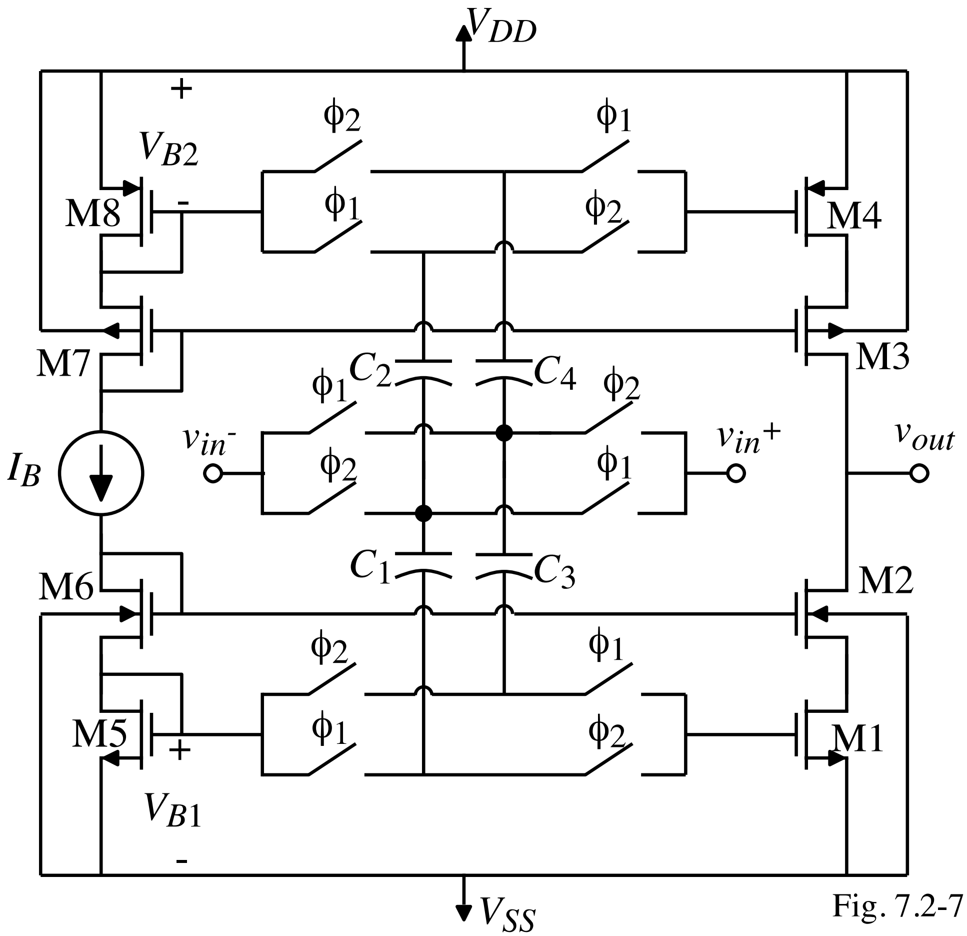


Fig. 7.2-7

Performance (1.5µm CMOS):

- 1.6mW dissipation
- $GB \approx$ 130MHz (C_L=2.2pF)
- Settling time of 10ns (C_L=10pF)

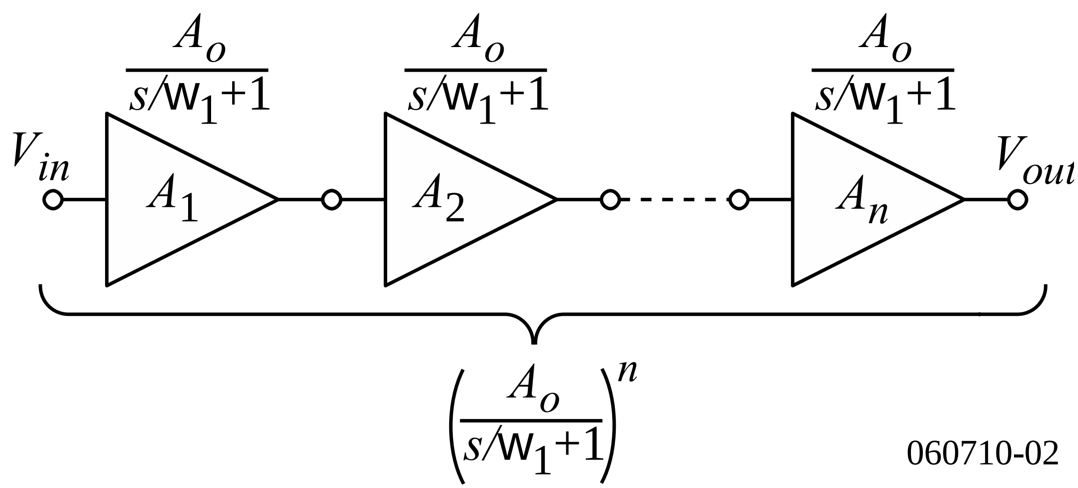
This amplifier was used with a 28.6MHz clock to realize a 5th-order switched capacitor filter having a cutoff frequency of 3.5MHz.

† S. Masuda, et. al., "CMOS Sampled Differential Push-Pull Cascode Op Amp," *Proc. of 1984 International Symposium on Circuits and Systems*, Montreal, Canada, May 1984, pp. 1211-12-14.

CASCADED AMPLIFIERS USING VOLTAGE AMPLIFIERS

Bandwidth of Cascaded Amplifiers

Cascading of low-gain, wide-bandwidth amplifiers:



Overall gain is $A_o{}^n$

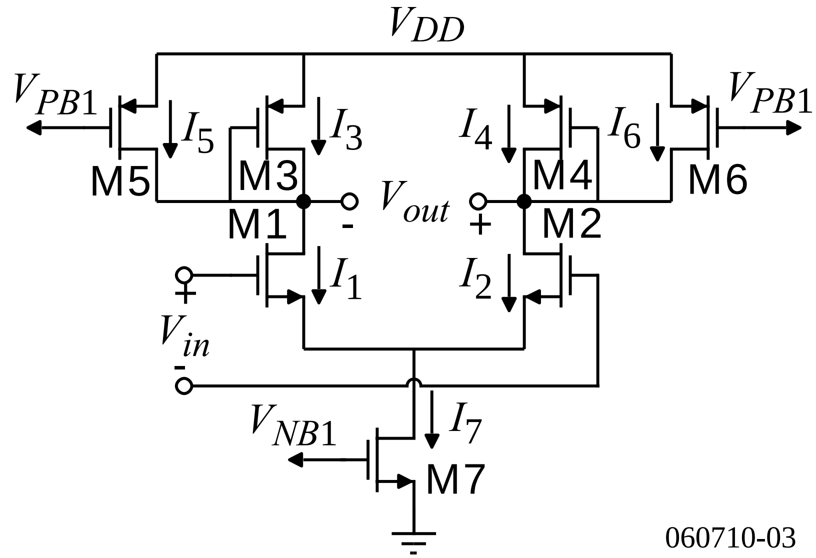
-3dB frequency is,

$$\omega_{-3dB} = \omega_1\sqrt{2^{1/n}-1}$$

If $A_o = 10$, $\omega_1 = 300\pi\text{x}10^6$ rads/sec. and $n = 3$, then

Overall gain is 60dB and $\omega_{-3dB} = 0.51\omega_1 = 480\text{x}10^6$ rads/sec. → 76.5 MHz

Voltage Amplifier Suitable for Cascading



Voltage Gain:

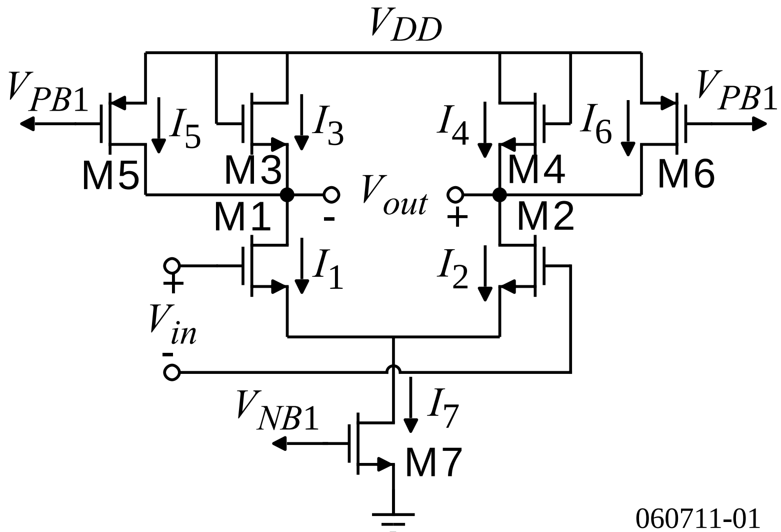
$$\frac{V_{out}}{V_{in}} = \frac{g_{m1}}{g_{m3}} = \sqrt{\frac{K_n'(W_1/L_1)(I_3+I_5)}{K_p'(W_3/L_3)I_3}}$$

$$\omega_{-3\text{dB}} \approx \frac{g_{m3}}{C_{gs1}}$$

Ex. 27-3 - Design of a Voltage Amplifier for Cascading

Design the previous voltage amplifier for a gain of A_o = 10 and a power dissipation of no more than 1mW. The design should permit A_o to be well defined. What is the ω_{-3dB} for this amplifier and what would be the ω_{-3dB} for a cascade of three identical amplifiers?

Solution



To enhance the accuracy of the gain, we replace M3 and M4 with NMOS transistors to avoid the variation of the transconductance parameter. This assumes a *p*-well technology to avoid bulk effects. The gain of 10 requires,

$$\frac{W_1}{L_1}(I_3+I_5) = 100\frac{W_3}{L_3}I_3$$

If V_{DD} = 2.5V, then $2(I_3+I_5)\cdot 2.5\text{V} = 1000\mu\text{W}$.

Therefore, I_3+I_5 = 200µA. Let I_3 = 20µA and $W_1/L_1 = 10W_3/L_3$.

Choose W_1/L_1 = 5µm/0.5µm which gives W_3/L_3 = 0.5µm/0.5µm. M5 and M6 are designed to give I_5 = 180µA and M7 is designed to give I_7 = 400µA.

The dominant pole is g_{m3}/C_{out}.

Ex. 27-3 – Continued

$$C_{out} = C_{gs3}+C_{bs3}+C_{bd1}+C_{bd5}+C_{gd1}+C_{gd5}+C_{gs1}\text{(next stage)} \approx C_{gs3} + C_{gs1}$$

Using C_{ox} = 60.6x10^{-4} F/m^2, we get,

$$C_{out} \approx (2.5+2.5)\text{x}10^{-12}\ \text{m}^2\text{x } 60.6\text{x}10^{-4}\ \text{F/m}^2 = 30.3\text{fF} \ \rightarrow C_{out} \approx 30\text{fF}$$

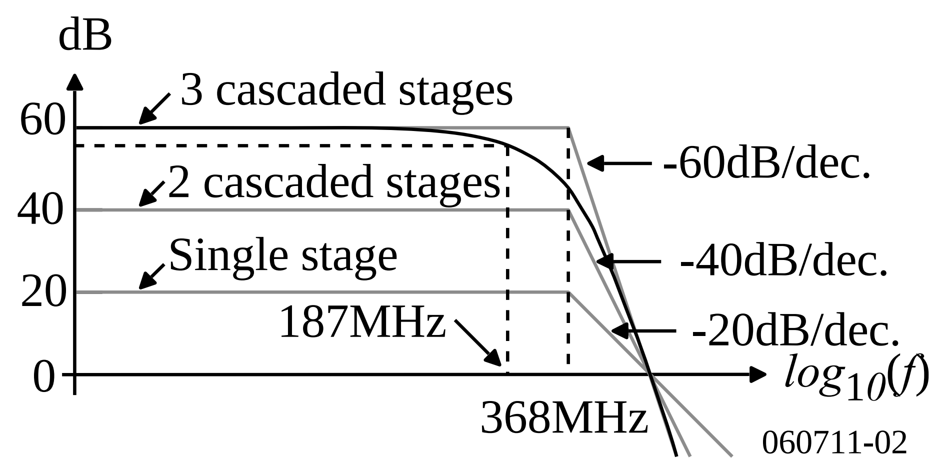
$$g_{m3} = \sqrt{2\cdot 120\cdot 1\cdot 20}\ \mu\text{S} = 69.3\mu\text{S}$$

$\therefore$ Dominant pole $\approx$ 69.3µS/30fF = 23.1x10^8 rads/sec. $\rightarrow$ f_{-3dB} = 368MHz

The bandwidth of three identical cascaded amplifiers giving a low-frequency gain of 60dB would have a f_{-3dB} of

$$f_{-3dB}\text{(Overall)} = f_{-3dB}\sqrt{2^{1/3}-1} = 368\text{ MHz } (0.5098) = 187\text{ MHz.}$$

P_{diss} = 3mW



CASCADED AMPLIFIERS USING CURRENT FEEDBACK AMPLIFIERS

Advantages of Using Current Feedback

Why current feedback?

- Higher *GB*
- Less voltage swing $\Rightarrow$ more dynamic range

What is a current amplifier?

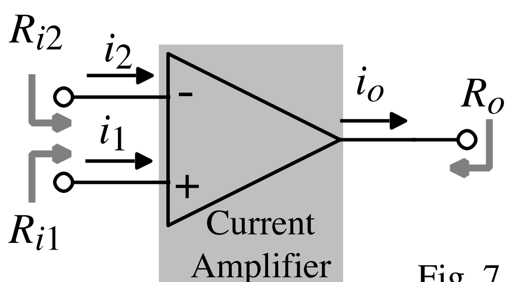


Fig. 7.2-8A

Requirements:

$i_o = A_i(i_1 - i_2)$

$R_{i1} = R_{i2} = 0\Omega$

$R_o = \infty$

Ideal source and load requirements:

$R_{source} = \infty$

$R_{Load} = 0\Omega$

Bandwidth Advantage of a Current Feedback Amplifier

Consider the inverting voltage amplifier shown using a current amplifier with negative current feedback:

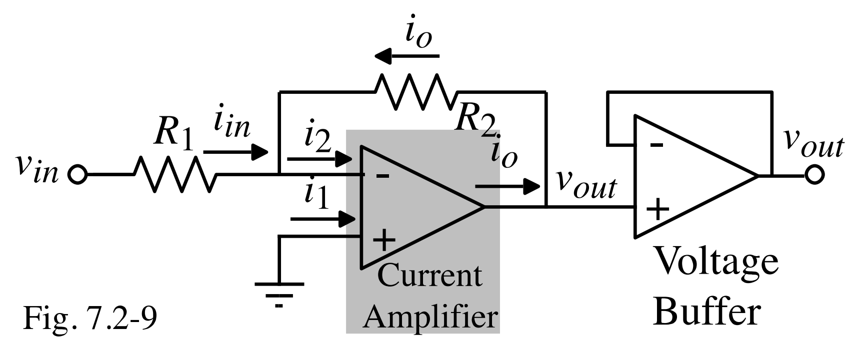


Fig. 7.2-9

The output current, i_o, of the current amplifier can be written as

$$i_o = A_i(s)(i_1 - i_2) = -A_i(s)(i_{in} + i_o)$$

The closed-loop current gain, i_o/i_{in}, can be found as

$$\frac{i_o}{i_{in}} = \frac{-A_i(s)}{1+A_i(s)}$$

However, $v_{out} = i_oR_2$ and $v_{in} = i_{in}R_1$. Solving for the voltage gain, v_{out}/v_{in} gives

$$\frac{v_{out}}{v_{in}} = \frac{i_oR_2}{i_{in}R_1} = \left(\frac{-R_2}{R_1}\right)\left(\frac{A_i(s)}{1+A_i(s)}\right)$$

If $A_i(s) = \dfrac{A_o}{(s/\omega_A) + 1}$, then

$$\frac{v_{out}}{v_{in}} = \left(\frac{-R_2}{R_1}\right)\left(\frac{A_o}{1+A_o}\right)\left(\frac{\omega_A(1+A_o)}{s + \omega_A(1+A_o)}\right) \quad \Rightarrow \quad A_v(0) = \frac{-R_2A_o}{R_1(1+A_o)} \quad \text{and} \quad \boxed{\omega_{-3dB} = \omega_A(1+A_o)}$$

Bandwidth Advantage of a Current Feedback Amplifier - Continued

The unity-gainbandwidth is,

$$GB = |A_v(0)|\,\omega_{-3dB} = \frac{R_2 A_o}{R_1(1+A_o)} \cdot \omega_A(1+A_o) = \frac{R_2}{R_1} A_o \cdot \omega_A = \frac{R_2}{R_1} GB_i$$

where GB_i is the unity-gainbandwidth of the current amplifier.

Note that if GB_i is constant, then increasing R_2/R_1 (the voltage gain) increases GB.

Illustration:

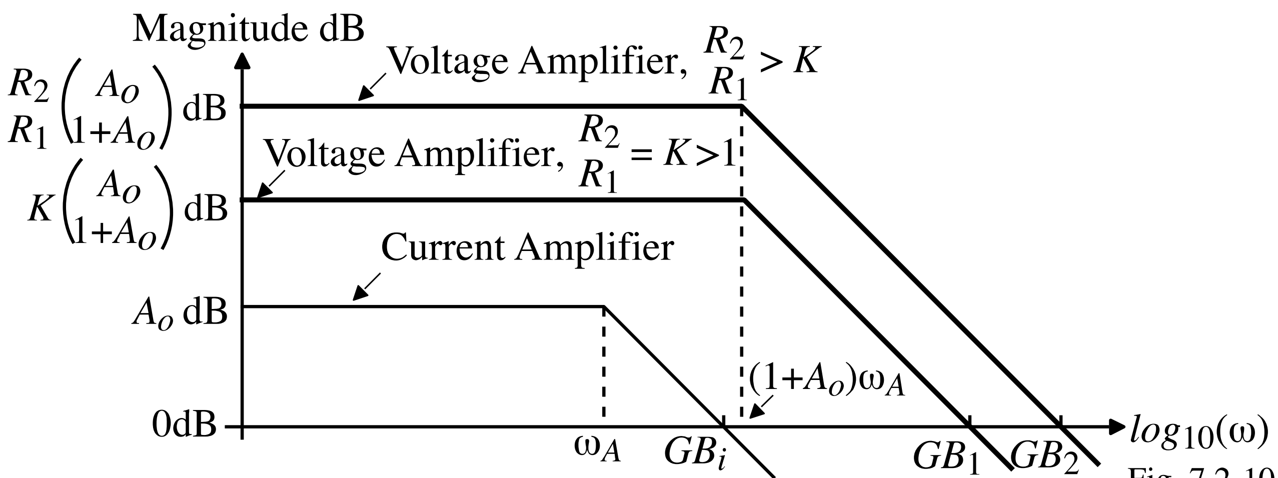


Fig. 7.2-10

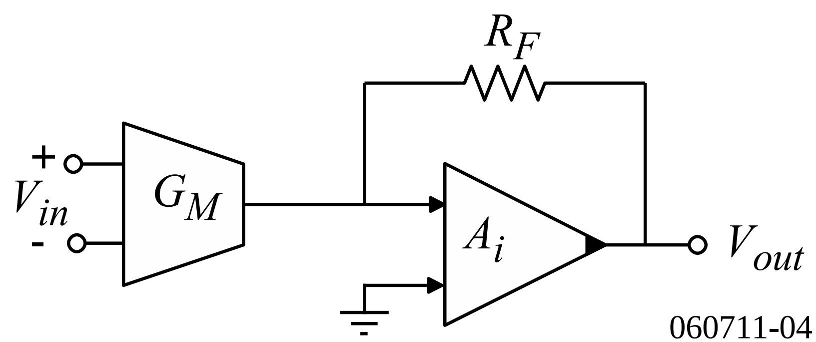
Note that $GB_2 > GB_1 > GB_i$

The above illustration assumes that the *GB* of the voltage amplifier realizing the voltage buffer is greater than the *GB* achieved from the above method.

Current Feedback Amplifier

In a current mirror implementation of the current amplifier, it is difficult to make the input resistance sufficiently small compared to R_1.

This problem can be solved using a transconductance input stage shown in the following block diagram:



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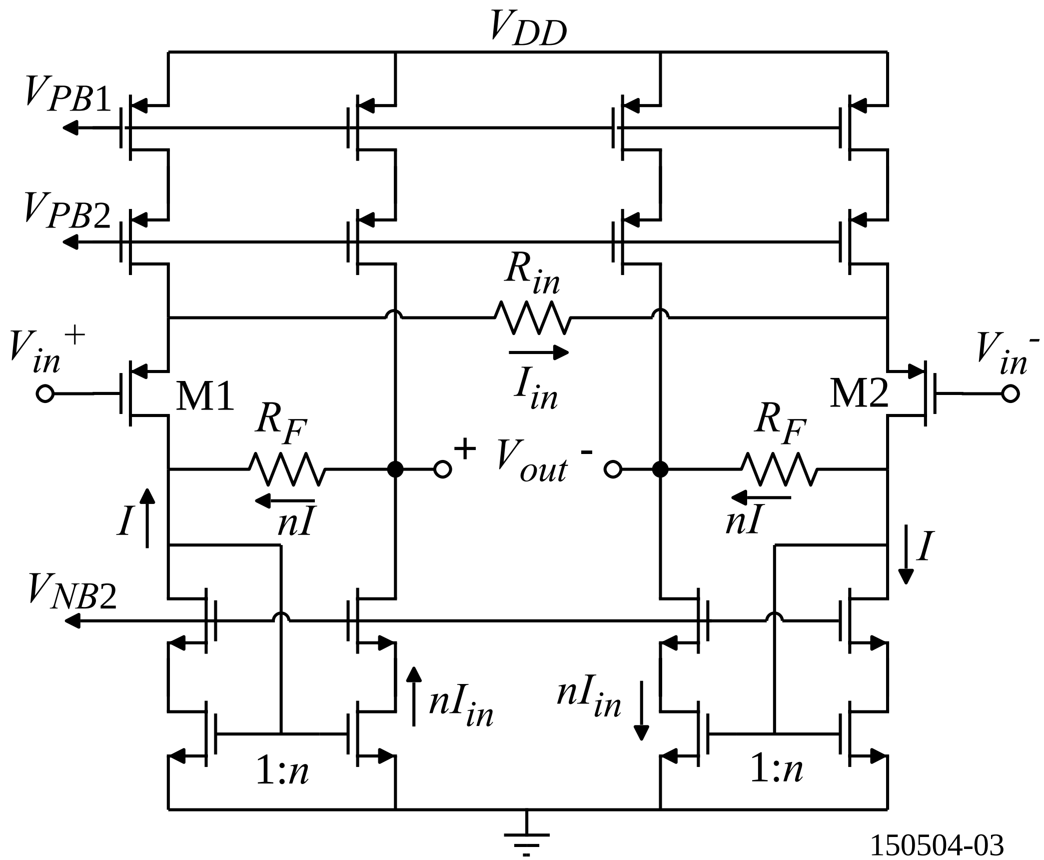
$$\frac{V_{out}}{V_{in}} = \frac{-G_M R_F A_i}{1 + A_i}$$

Differential Implementation of the Current Feedback Amplifier



$$I_{in} = \frac{g_{m1}}{1+0.5g_{m1}R_{in}}\left(\frac{V_{in}^{+} - V_{in}^{-}}{2}\right), \quad I_{in} = (1+n)I, \text{ and} \qquad V_{out} = \frac{n\,(2R_F)}{1+n}\, I_{in}$$

$$\therefore \frac{V_{out}}{V_{in}} \approx \frac{2nR_F}{(1+n)\mathrm{R_{in}}}$$

A 20dB Voltage Amplifier using a Current Amplifier

The following circuit is a programmable voltage amplifier with up to 20dB gain:

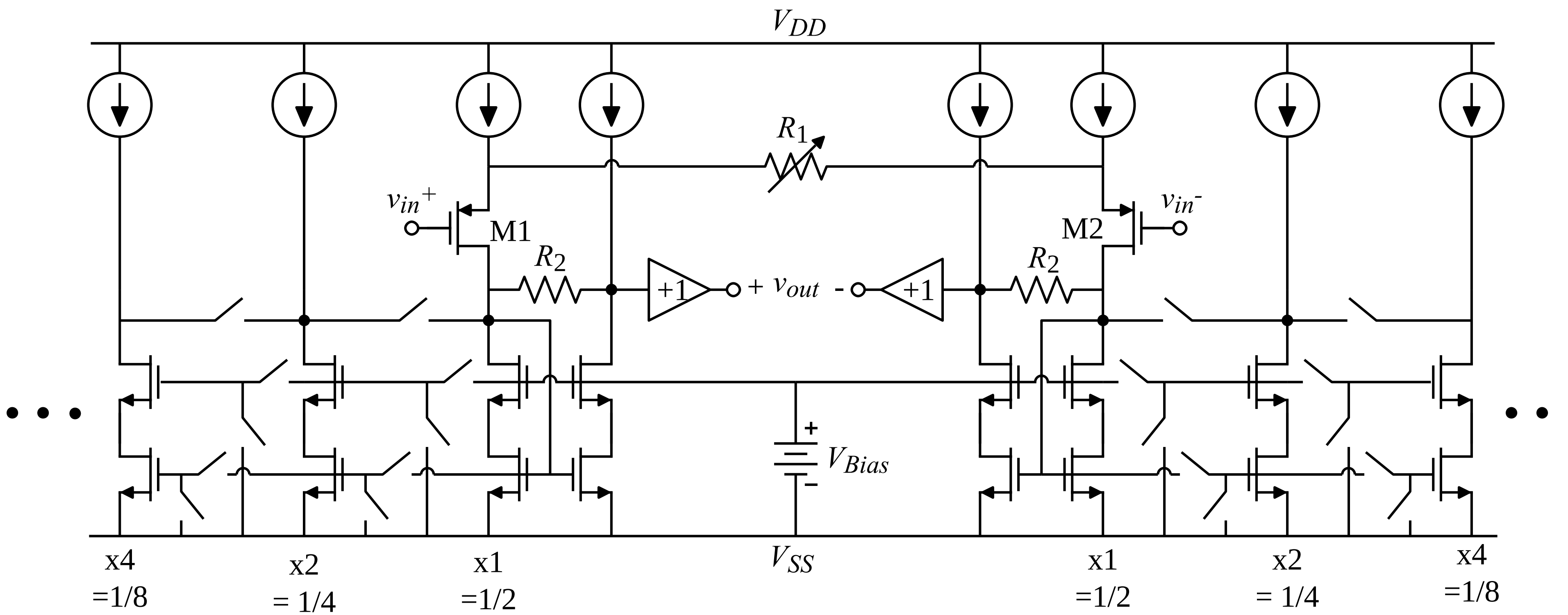
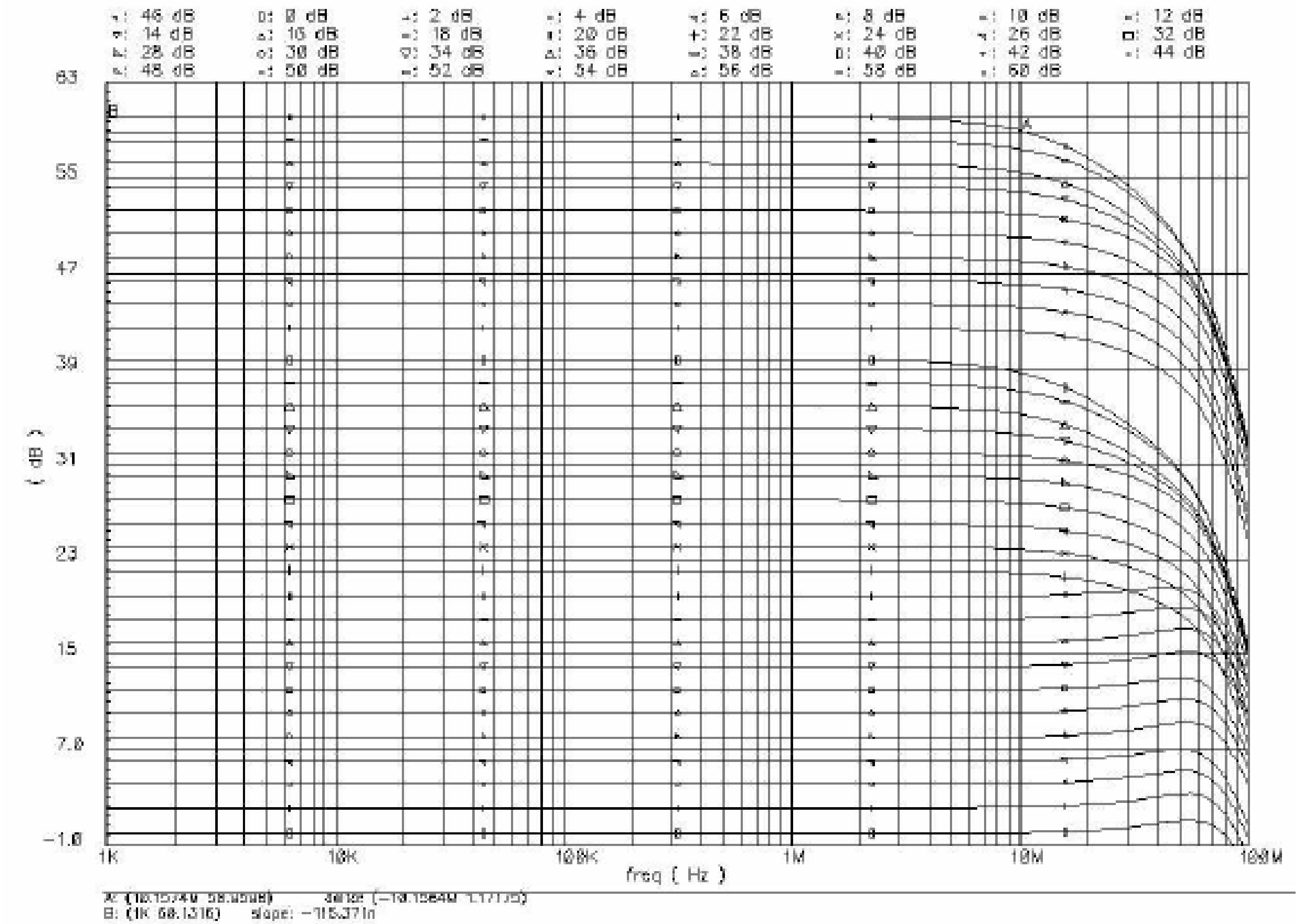


Fig. 7.2-135A

R_1 and the current mirrors are used for gain variation while R_2 is fixed.

Frequency Response of a 60dB PGA

Includes output buffer:



SUMMARY

- Increasing the *GB* of an op amp requires that the magnitude of all non-dominant poles are much greater than *GB* from the origin of the complex frequency plane
- The practical limit of *GB* for an op amp is approximately 5-10 times less than the magnitude of the smallest non-dominant pole (≈ 100MHz)
- To achieve high values of *GB* it is necessary to eliminate the non-dominant poles (which come from parasitics) or increase the magnitude of the non-dominant poles
- The best way to achieve high-bandwidth amplifiers is to cascade high-bandwidth voltage amplifiers
- If the gain of the high-bandwidth voltage amplifiers is well defined, then it is not necessary to use negative feedback around the amplifier
- Amplifiers with well-defined gains are achievable with a -3dB bandwidth of 100MHz

