6.976
High Speed Communication Circuits and Systems
Lecture 6
Enhancement Techniques for Broadband Amplifiers, Narrowband Amplifiers

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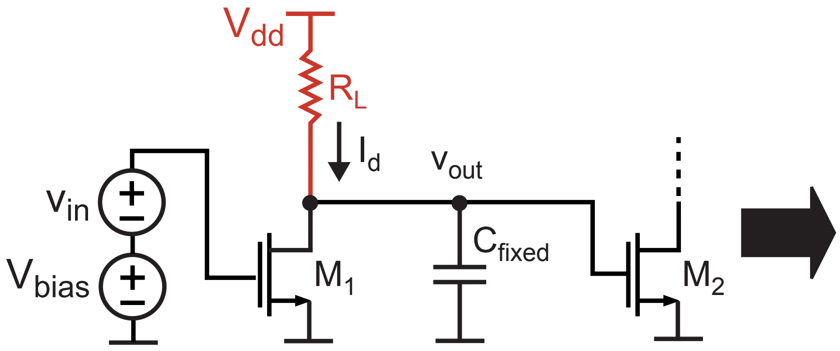
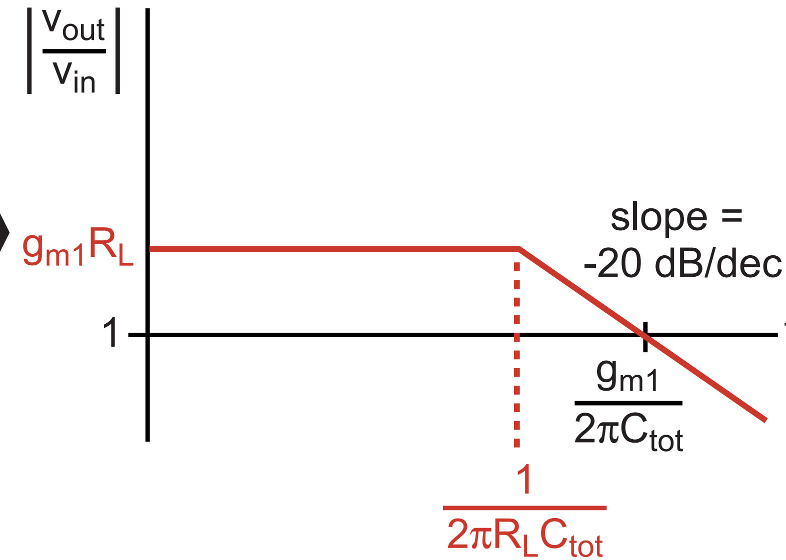
Resistor Loaded Amplifier (Unsilicided Poly)



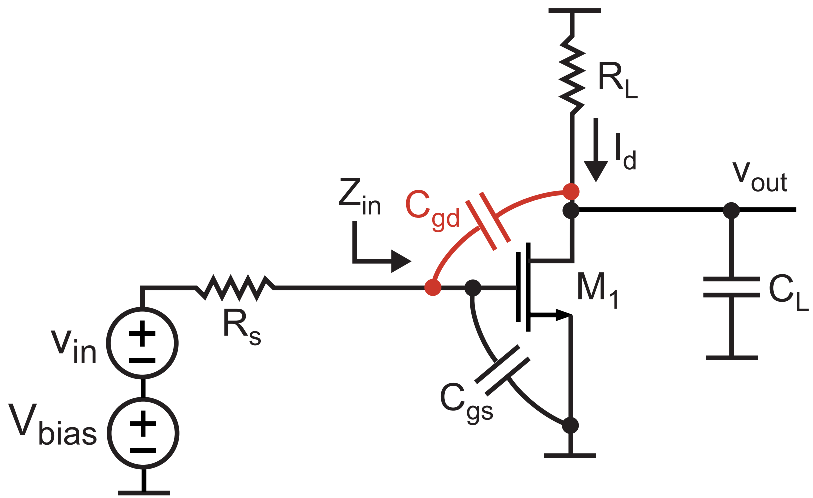
$$C_{tot} = C_{db1} + C_{RL}/2 + C_{gs2} + KC_{ov2} + C_{fixed}$$

$(+C_{ov1})$ Miller multiplication factor



- **We decided this was the fastest non-enhanced amplifier**
  - **Can we go faster? (i.e., can we enhance its bandwidth?)**
- **We will look at the following**
  - **Reduction of Miller effect on C_{gd}**
  - **Shunt, series, and zero peaking**
  - **Distributed amplification**

Miller Effect on C_{gd} Is Significant



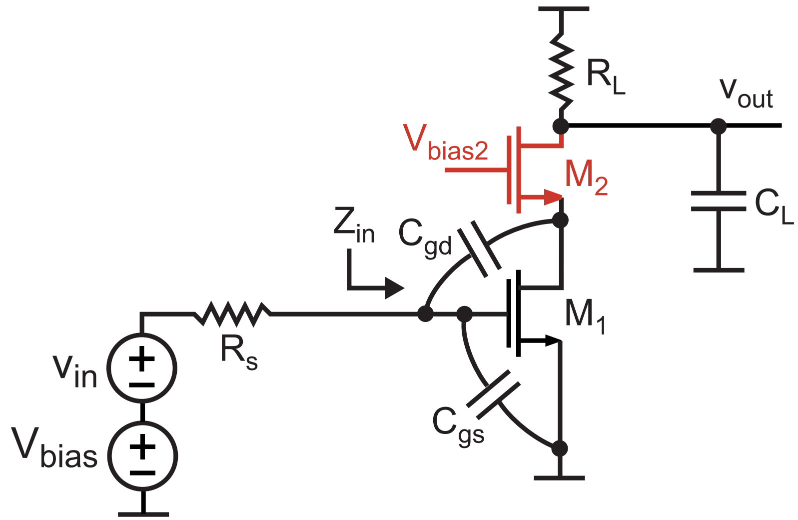
- C_{gd} is quite significant compared to C_{gs}
  - In 0.18μ CMOS, C_{gd} is about 45% the value of C_{gs}
- Input capacitance calculation

$$Z_{in} \approx \frac{1}{s(C_{gs} + C_{gd}(1 - A_v))} = \frac{1}{sC_{gs}(1 + C_{gd}/C_{gs}(1 + g_m R_L))}$$

  - For 0.18μ CMOS, gain of 3, input cap is almost tripled over C_{gs}!

$$Z_{in} \approx \frac{1}{sC_{gs}(1 + 0.45(4))} = \frac{1}{sC_{gs}2.8}$$

Reduction of C_{gd} Impact Using a Cascode Device



- **The cascode device lowers the gain seen by C_{gd} of M_1**

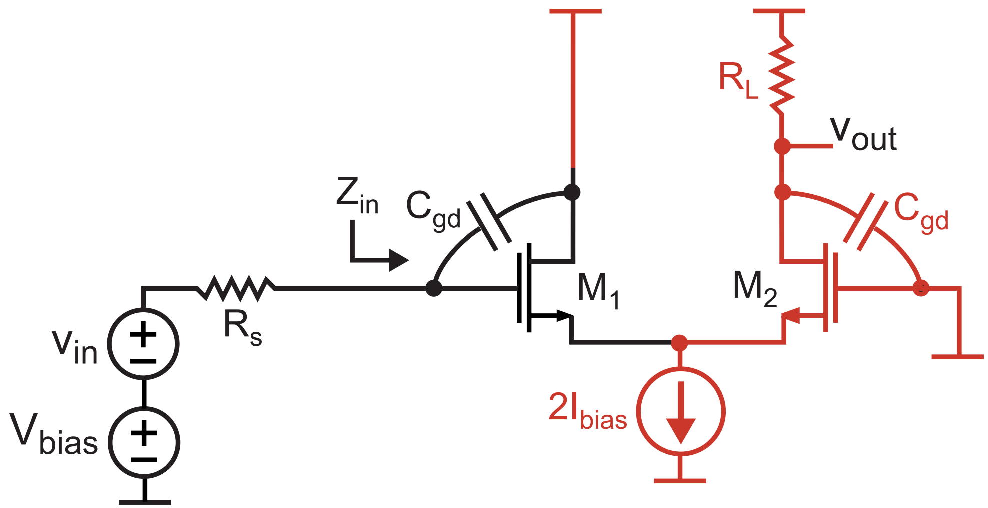
$$A_v \rightarrow g_{m1}\frac{1}{g_{m2}} \approx 1 \quad \Rightarrow \quad Z_{in} \approx \frac{1}{sC_{gs}(1 + C_{gd}/C_{gs}(2))}$$

  - **For 0.18m CMOS and gain of 3, impact of C_{gd} is reduced by 30%:**

$$Z_{in} \approx \frac{1}{sC_{gs}1.9}$$

- **Issue: cascoding lowers achievable voltage swing**

Source-Coupled Amplifier

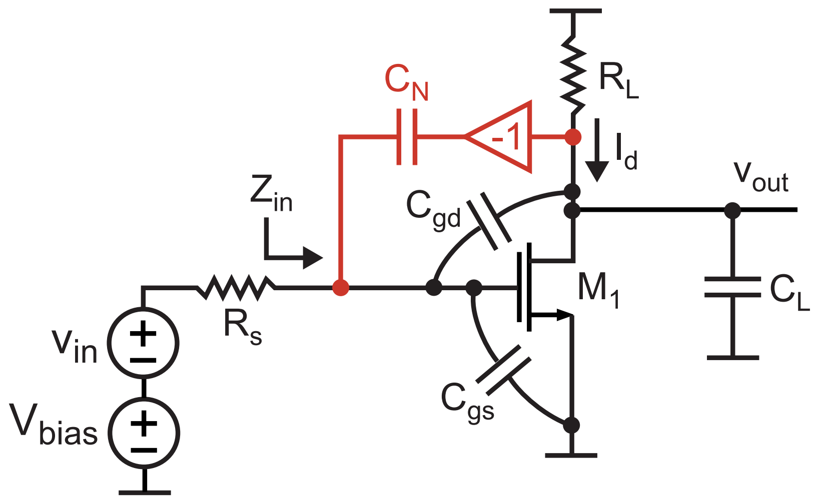


- **Remove impact of Miller effect by sending signal through source node rather than drain node**
  - **C_{gd} not Miller multiplied AND impact of C_{gs} cut in half!**

$$Z_{in} \approx \frac{1}{s(C_{gs}/2 + C_{gd})} \quad \Rightarrow \quad Z_{in} \approx \frac{1}{sC_{gs}0.95} \quad (0.18\mu \text{ CMOS})$$

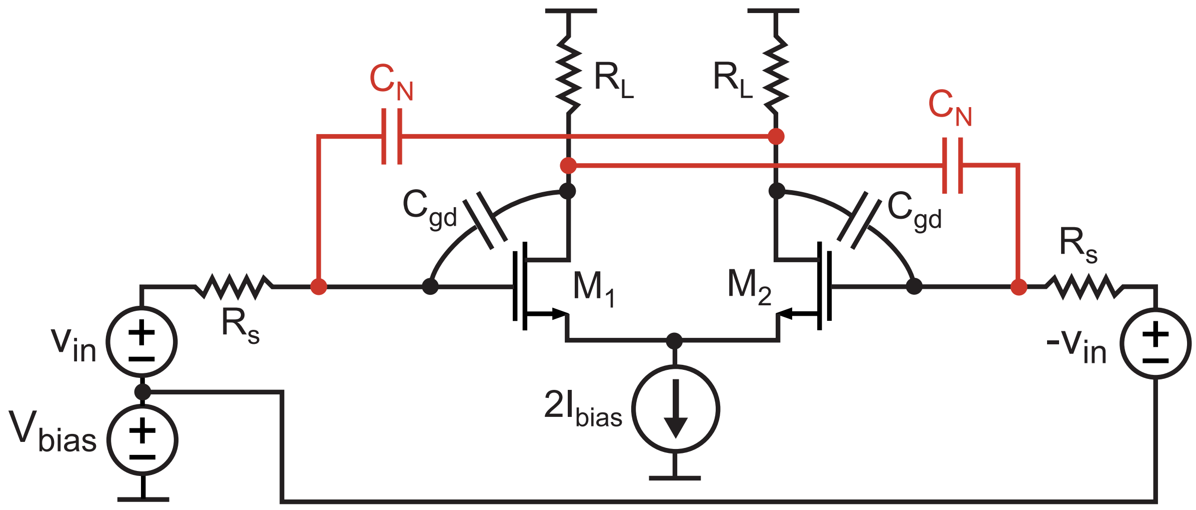
- **The bad news**
  - **Signal has to go through source node (C_{sb} significant)**
  - **Power consumption doubled**

Neutralization



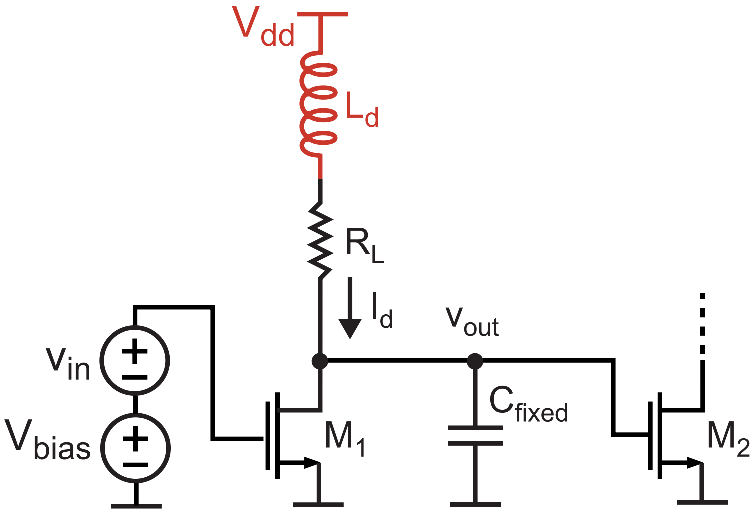
- **Consider canceling the effect of C_{gd}**
  - **Choose $C_N = C_{gd}$**
  - **Charging of C_{gd} now provided by C_N**
- **Benefit: Impact of C_{gd} removed** $\Rightarrow \quad Z_{in} \approx \frac{1}{sC_{gs}}$
- **Issues:**
  - **How do we create the inverting amplifier?**
  - **What happens if C_N is not precisely matched to C_{gd}?**

Practical Implementation of Neutralization



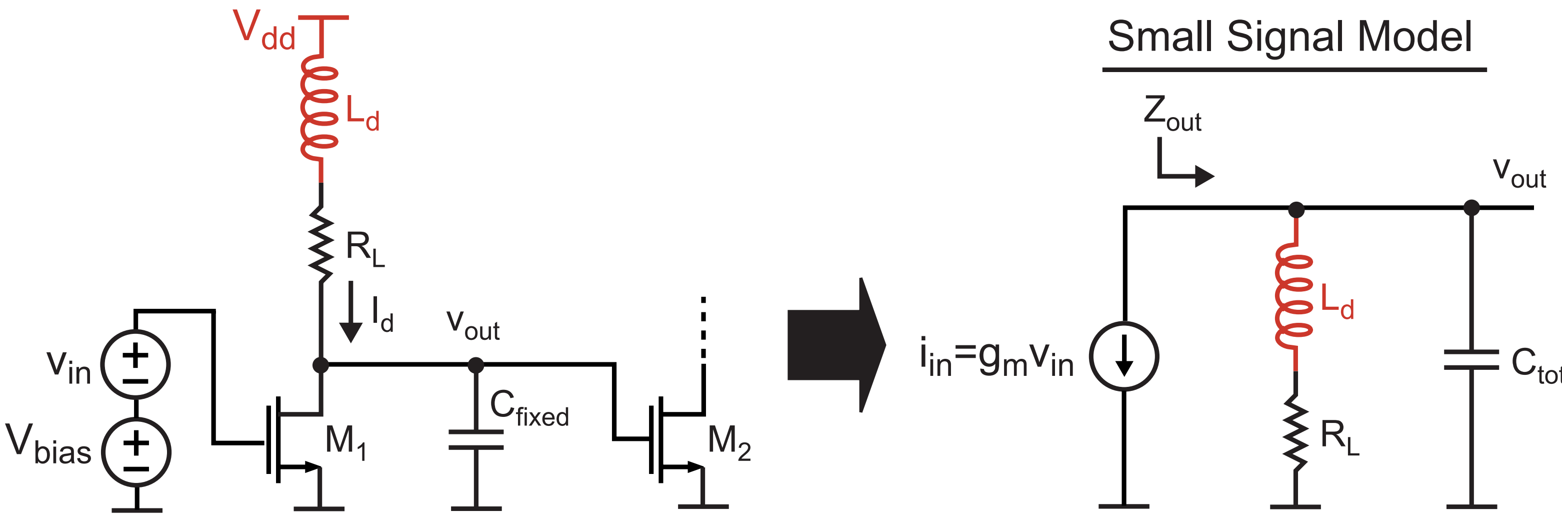
- **Leverage differential signaling to create an inverted signal**
- **Only issue left is matching C_N to C_{gd}**
  - **Often use lateral metal caps for C_N (or CMOS transistor)**
  - **If C_N too low, residual influence of C_{gd}**
  - **If C_N too high, input impedance has inductive component**
    - Causes peaking in frequency response
    - Often evaluate acceptable level of peaking using eye diagrams

Shunt-peaked Amplifier



- Use inductor in load to extend bandwidth
  - Often implemented as a spiral inductor
- We can view impact of inductor in both time and frequency
  - In frequency: peaking of frequency response
  - In time: delay of changing current in R_L
- Issue – can we extend bandwidth without significant peaking?

Shunt-peaked Amplifier - Analysis



- Expression for gain

$$A_v = g_m Z_{out} = g_m[(sL_d + R_L)||1/(sC_{tot})]$$

$$= g_m R_L \frac{s(L_d/R_L)+1}{s^2 L_d C_{tot} + sR_L C_{tot} + 1}$$

- Parameterize with

$$m = \frac{R_L C_{tot}}{\tau}, \quad \text{where } \tau = \frac{L_d}{R_L}$$

  - Corresponds to ratio of RC to LR time constants

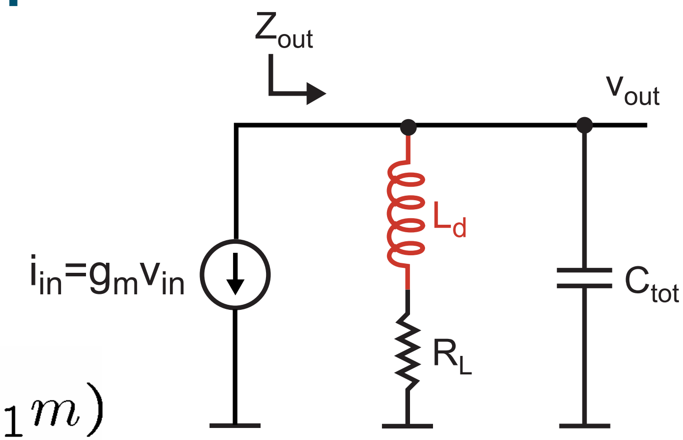
The Impact of Choosing Different Values of m – Part 1

- **Parameterized gain expression**

$$A_v = g_m R_L \frac{\tau s + 1}{s^2\tau^2 m + s\tau m + 1}$$

- **Comparison of new and old 3 dB frequencies**

Small Signal Model



set: $s = jw, w_1 = \frac{1}{RC}, \tau = 1/(w_1 m)$

$$|A_v| = g_m R_L \left| \frac{jw/(w_1 m) + 1}{-(w/(w_1 m))^2 m + jw/(w_1 m)m + 1} \right|$$

define w_2 as new 3 dB frequency, note that w_1 is old one

$$\Rightarrow \left| \frac{jw_2/(w_1 m) + 1}{-(w_2/(w_1 m))^2 m + jw_2/w_1 + 1} \right| = \frac{1}{\sqrt{2}}$$

- **Want to solve for w_2/w_1**

The Impact of Choosing Different Values of m – Part 2

- **From previous slide, we have**

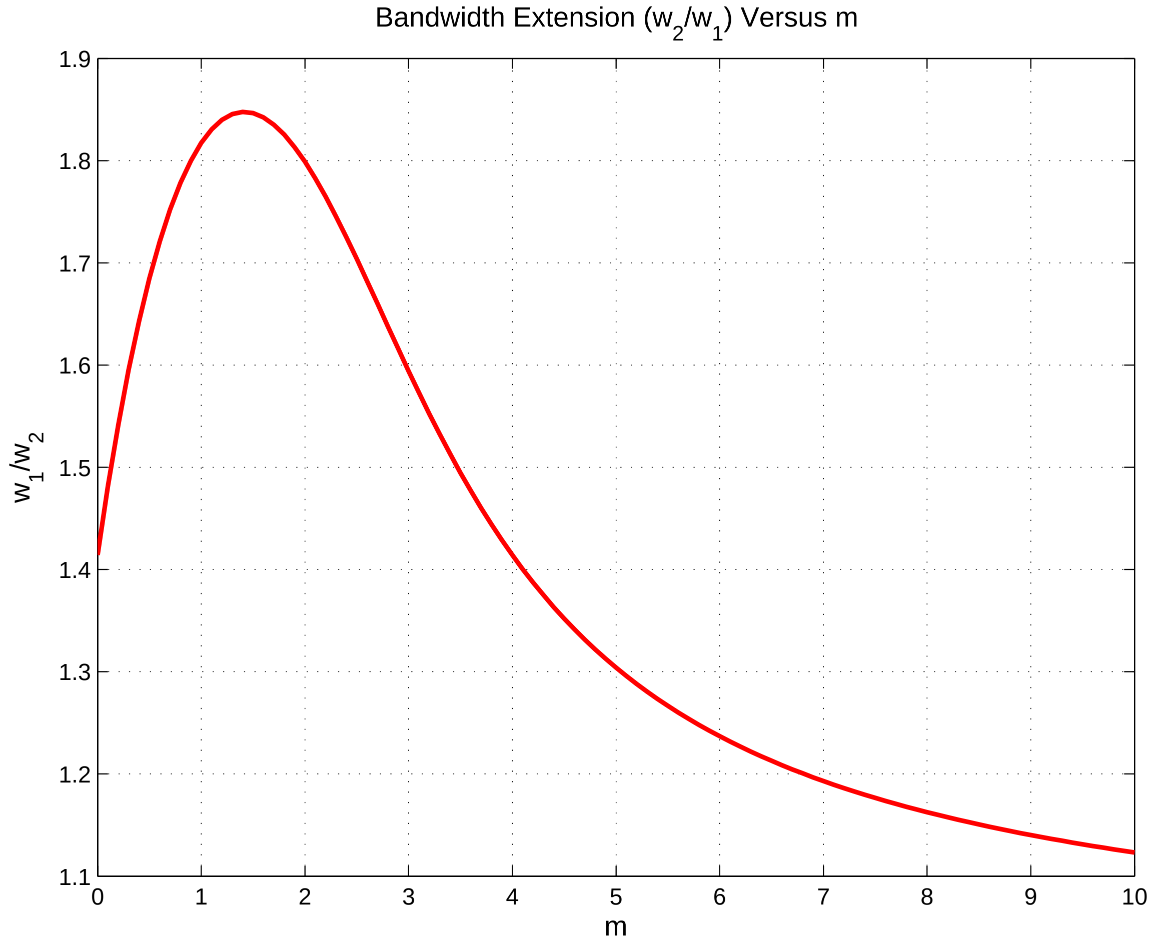
$$\left|\frac{jw_2/(w_1 m)+1}{-(w_2/(w_1 m))^2 m + jw_2/w_1 + 1}\right| = \frac{1}{\sqrt{2}}$$

- **After much algebra**

$$\frac{w_2}{w_1} = \sqrt{\left(-\frac{m^2}{2}+m+1\right)+\sqrt{(-\left(\frac{m^2}{2}+m+1\right)^2+m^2}}$$

- **We see that m directly sets the amount of bandwidth extension!**
  - **Once m is chosen, inductor value is** $\boxed{L_d = \frac{R_L^2 C_{tot}}{m}}$

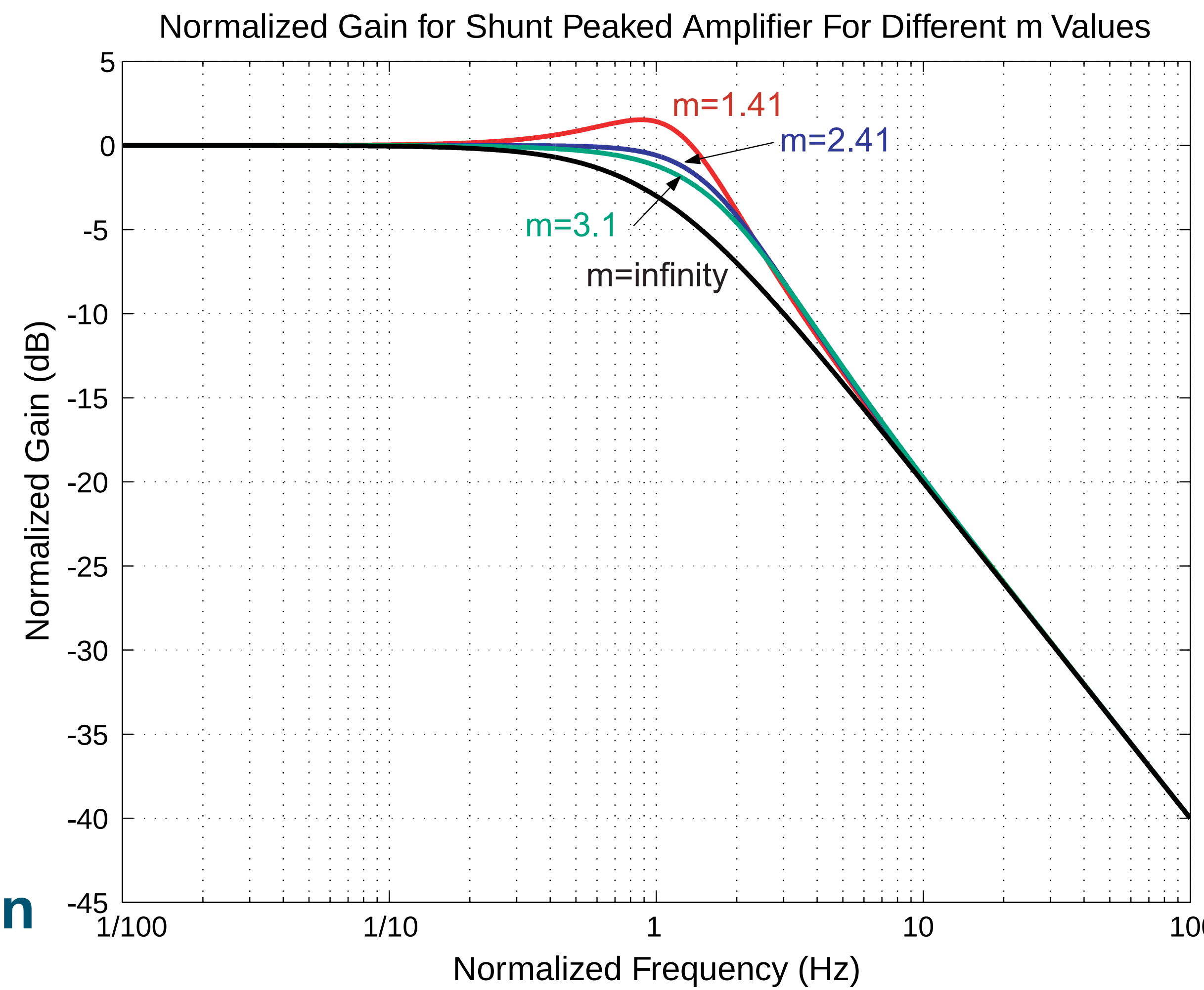
Plot of Bandwidth Extension Versus m



- **Highest extension: w_2/w_1 = 1.85 at m $\approx$ 1.41**
  - **However, peaking occurs!**

Plot of Transfer Function Versus m

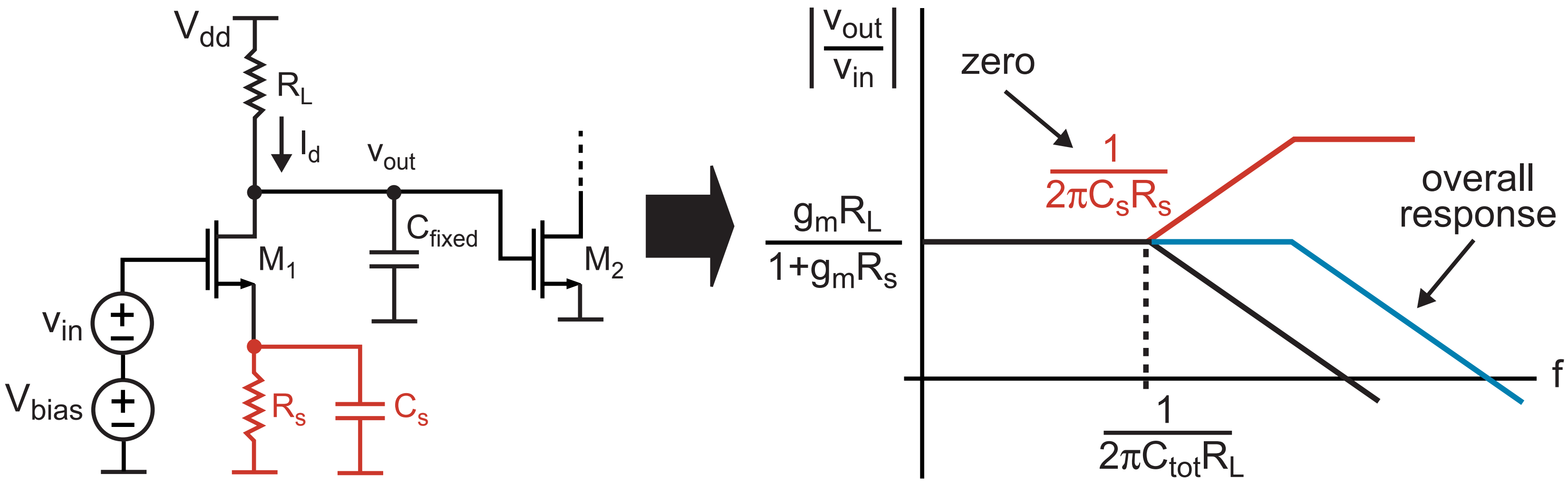
- **Maximum bandwidth: m = 1.41 (extension = 1.85)**
- **Maximally flat response: m = 2.41 (extension = 1.72)**
- **Best phase response: m = 3.1 (extension = 1.6)**
- **No peaking: m = infinity**
- **Eye diagrams often used to evaluate best m**



To Do

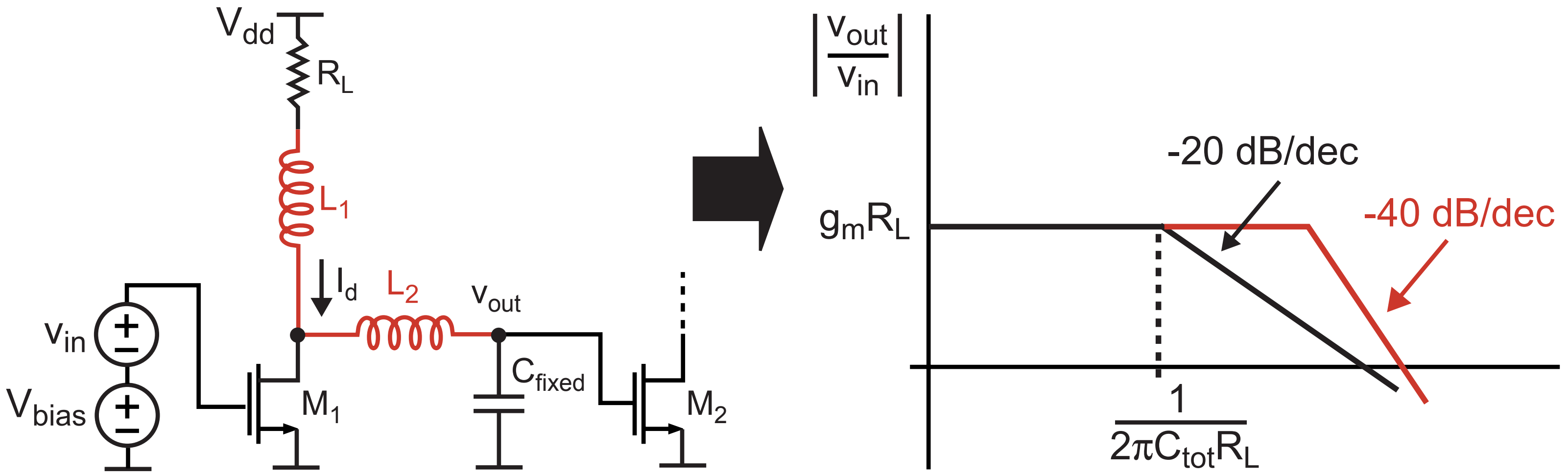
- **Add eye diagrams**

Zero-peaked Common Source Amplifier



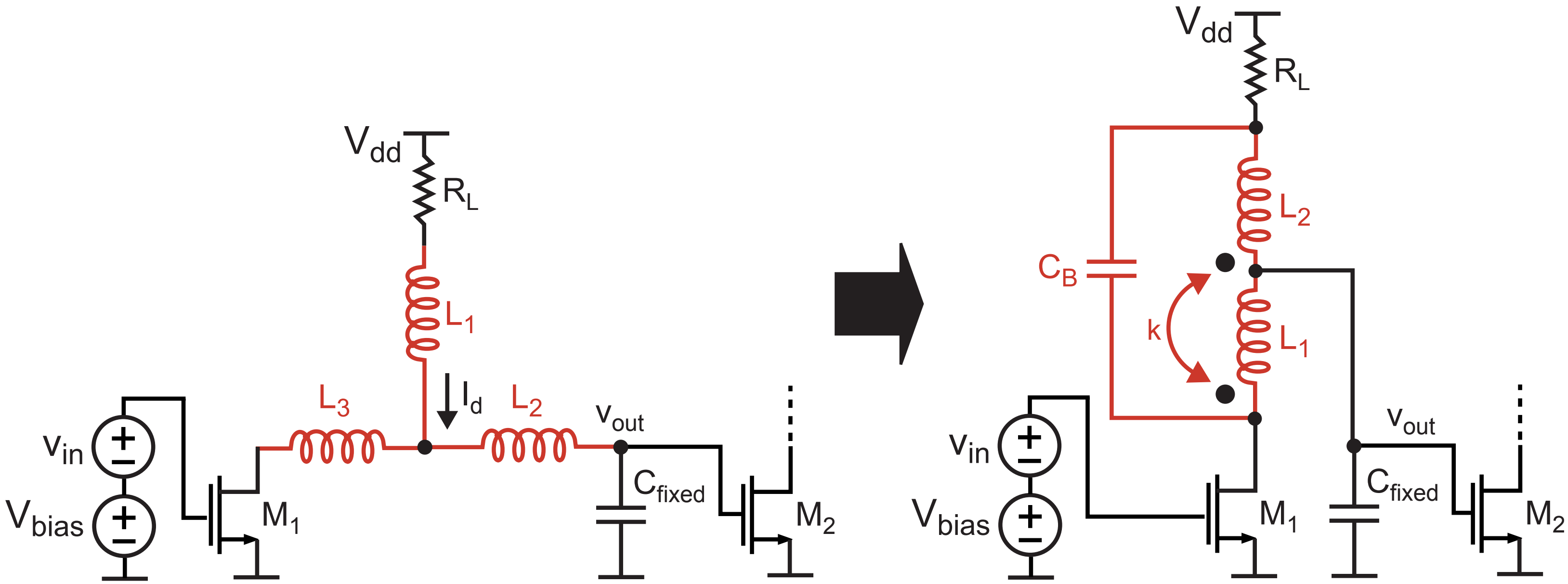
- **Inductors are expensive with respect to die area**
- **We can instead achieve bandwidth extension with capacitor**
  - **Idea: degenerate gain at low frequencies, remove degeneration at higher frequencies (i.e., create a zero)**
- **Issues:**
  - **Must increase R_L to keep same gain (lowers pole)**
  - **Lowers achievable gate voltage bias (lowers device f_t)**

Back to Inductors – Shunt and Series Peaking



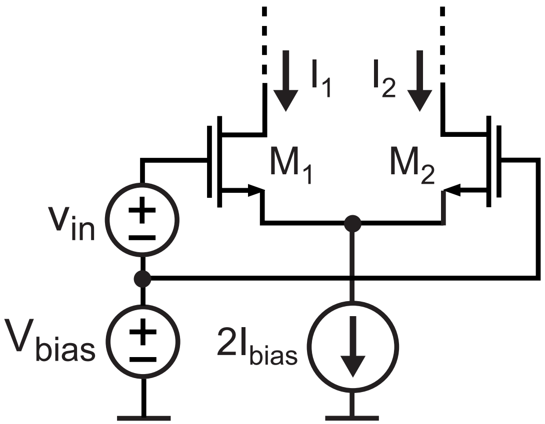
- **Combine shunt peaking with a series inductor**
  - **Bandwidth extension by converting to a second order filter response**
    - Can be designed for proper peaking
- **Increases delay of amplifier**

T-Coil Bandwidth Enhancement



- **Uses coupled inductors to realize T inductor network**
  - **Works best if capacitance at drain of M_1 is much less than the capacitance being driven at the output load**
- **See Chap. 8 of Tom Lee's book (pp 187-191) for analysis**
- **See S. Galal, B. Ravazi, "10 Gb/s Limiting Amplifier and Laser/Modulator Driver in 0.18u CMOS", ISSCC 2003, pp 188-189 and "Broadband ESD Protection …", pp. 182-183**

Bandwidth Enhancement With f_t Doublers



- A MOS transistor has f_t calculated as

$$2\pi f_t = \frac{g_m}{C_{gs} + C_{gd}} \approx \frac{g_m}{C_{gs}}$$

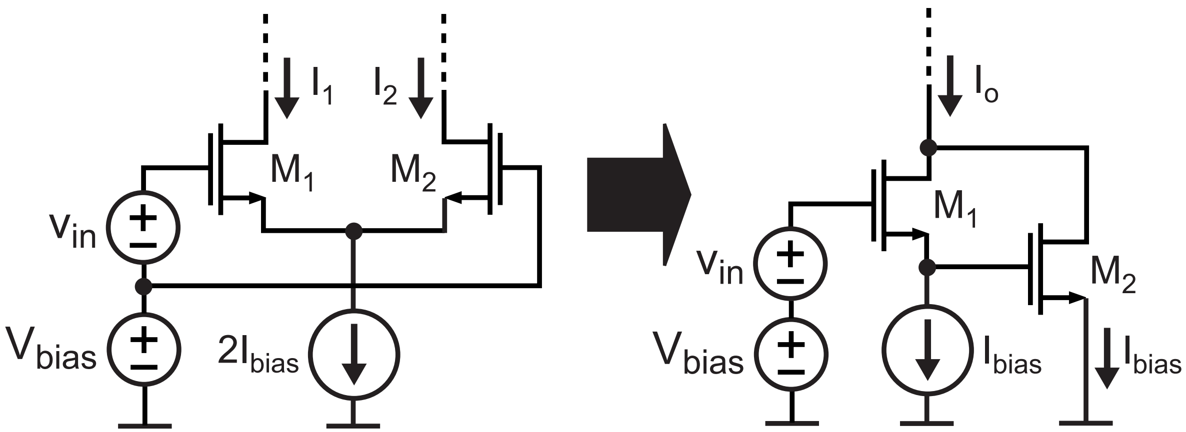
- f_t doubler amplifiers attempt to increase the ratio of transconductance to capacitance

- We can make the argument that differential amplifiers are f_t doublers
  - Capacitance seen by V_{in} for single-ended input: $C_{gs}/2$
  - Difference in current:

$$i_2 - i_1 = \frac{v_{in}}{2} g_m - \left(-\frac{v_{in}}{2}\right) g_m = v_{in} g_m$$

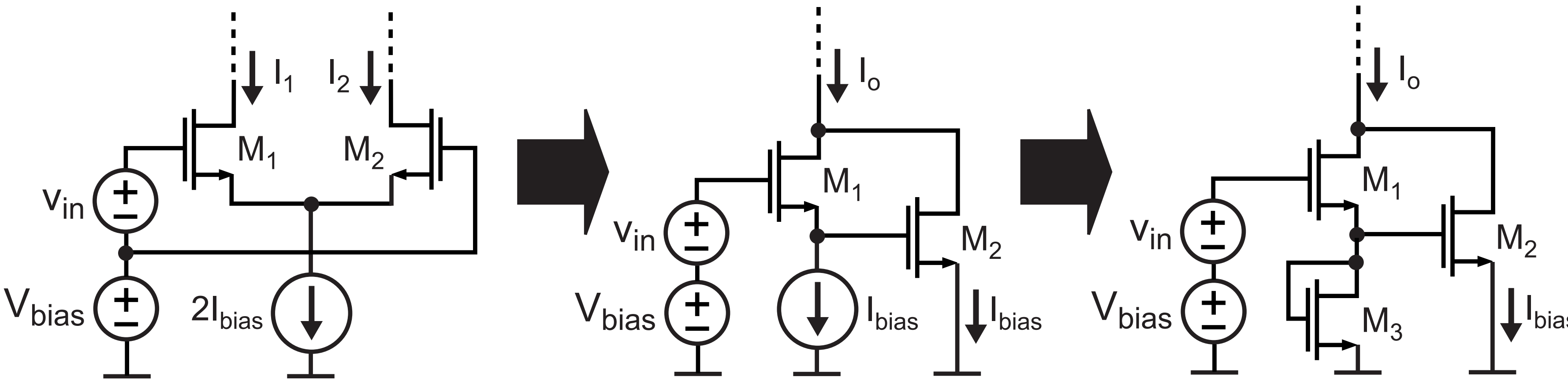
- Transconductance to Cap ratio is doubled: $\frac{2g_m}{C_{gs}}$

Creating a Single-Ended Output



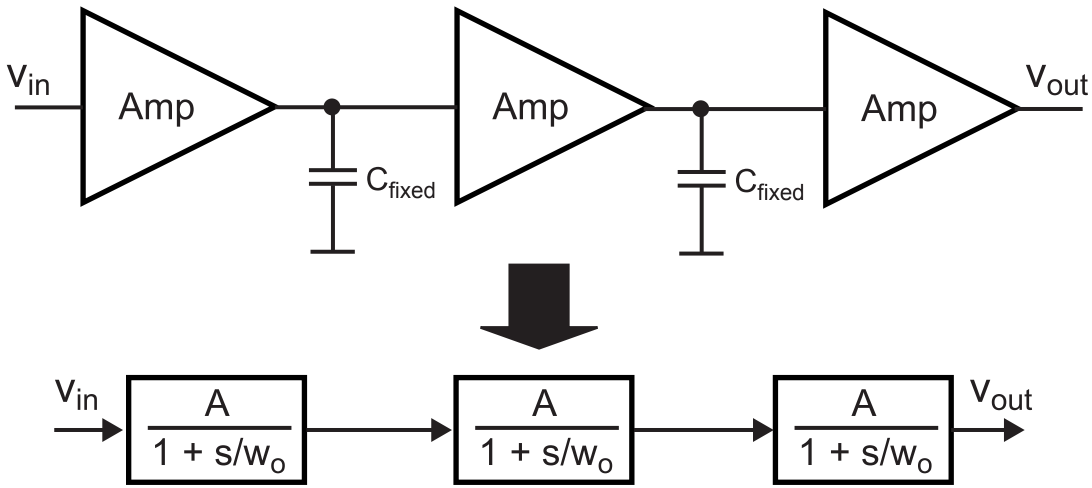
- **Input voltage is again dropped across two transistors**
  - **Ratio given by voltage divider in capacitance**
    - Ideally is ½ of input voltage on C_{gs} of each device
- **Input voltage source sees the series combination of the capacitances of each device**
  - **Ideally sees ½ of the C_{gs} of M_1**
- **Currents of each device add to ideally yield ratio:** $\frac{2g_m}{C_{gs}}$

Creating the Bias for M_2



- **Use current mirror for bias**
  - **Inspired by bipolar circuits (see Tom Lee's book, page 198)**
- **Need to set V_{bias} such that current through M_1 has the desired current of I_{bias}**
  - **The current through M_2 will ideally match that of M_1**
- **Problem: achievable bias voltage across M_1 (and M_2) is severely reduced (thereby reducing effective f_t of device)**
  - **Do f_t doublers have an advantage in CMOS?**

Increasing Gain-Bandwidth Product Through Cascading



- **We can significantly increase the gain of an amplifier by cascading n stages**

$$\Rightarrow \quad \frac{v_{out}}{v_{in}} = \left(\frac{A}{1 + s/w_o}\right)^n = A^n \frac{1}{(1 + s/w_o)^n}$$

  - **Issue – bandwidth degrades, but by how much?**

Analytical Derivation of Overall Bandwidth

- **The overall 3-db bandwidth of the amplifier is where**

$$\left|\frac{v_{out}}{v_{in}}\right| = \left|\frac{A}{1+jw_1/w_o}\right|^n = \frac{A^n}{\sqrt{2}}$$

  - **w_1 is the overall bandwidth**
  - **A and w_o are the gain and bandwidth of each section**

$$\Rightarrow \quad \left(\frac{A}{\sqrt{1+(w_1/w_o)^2}}\right)^n = \frac{A^n}{\sqrt{2}}$$

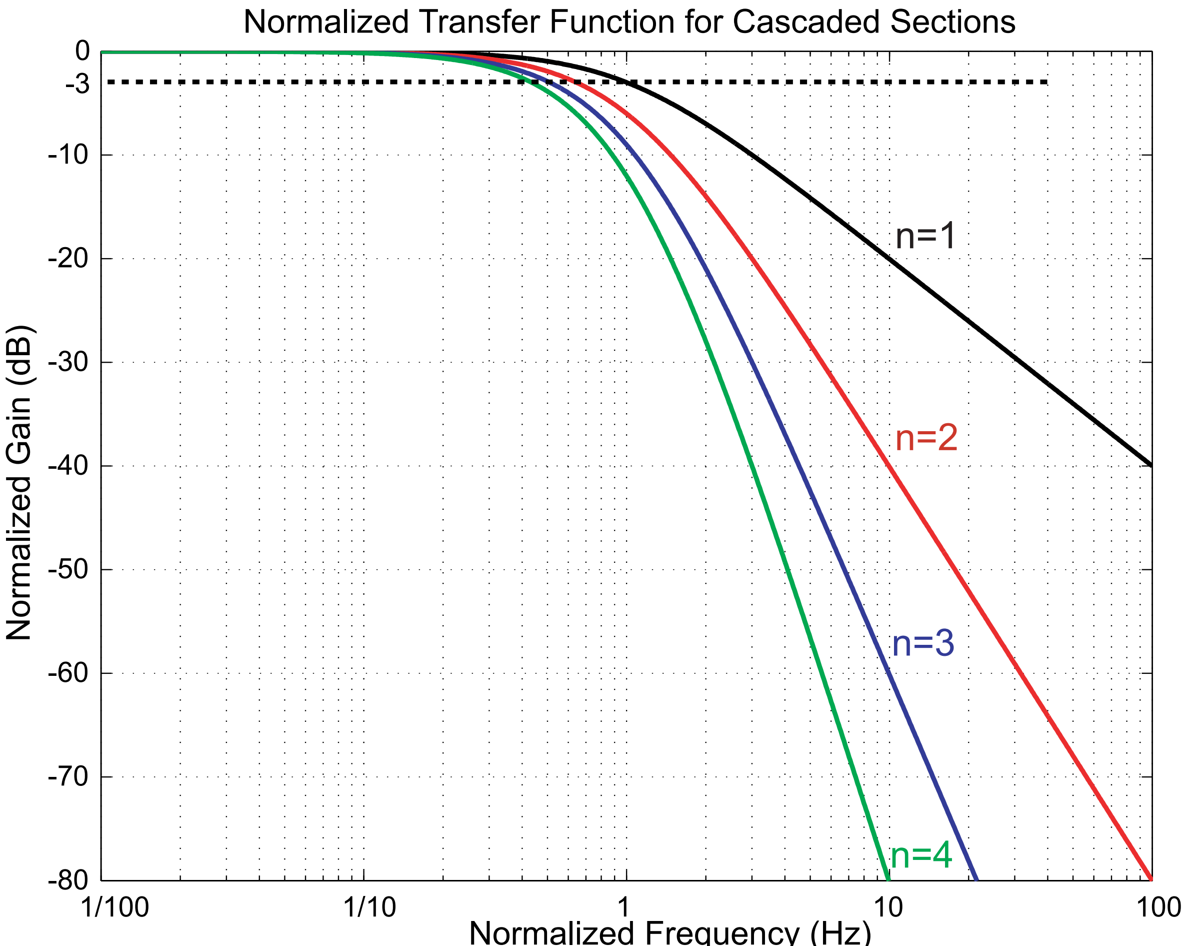
$$\Rightarrow \quad \left(1+(w_1/w_o)^2\right)^n = 2$$

$$\boxed{\Rightarrow \quad w_1 = w_o\sqrt{2^{1/n}-1}}$$

  - **Bandwidth decreases much slower than gain increases!**
    - Overall gain bandwidth product of amp can be increased!

Transfer Function for Cascaded Sections

$$H(f) = \left|\frac{1}{1 + j2\pi f}\right|^n$$



Choosing the Optimal Number of Stages

- To first order, there is a constant gain-bandwidth product for each stage

$$\Rightarrow \quad Aw_o = w_t \quad \Rightarrow \quad w_o = w_t/A$$

  - Increasing the bandwidth of each stage requires that we lower its gain
  - Can make up for lost gain by cascading more stages
- We found that the overall bandwidth is calculated as

$$w_1 = w_o\sqrt{2^{1/n} - 1} = \boxed{\frac{w_t}{A}\sqrt{2^{1/n} - 1}}$$

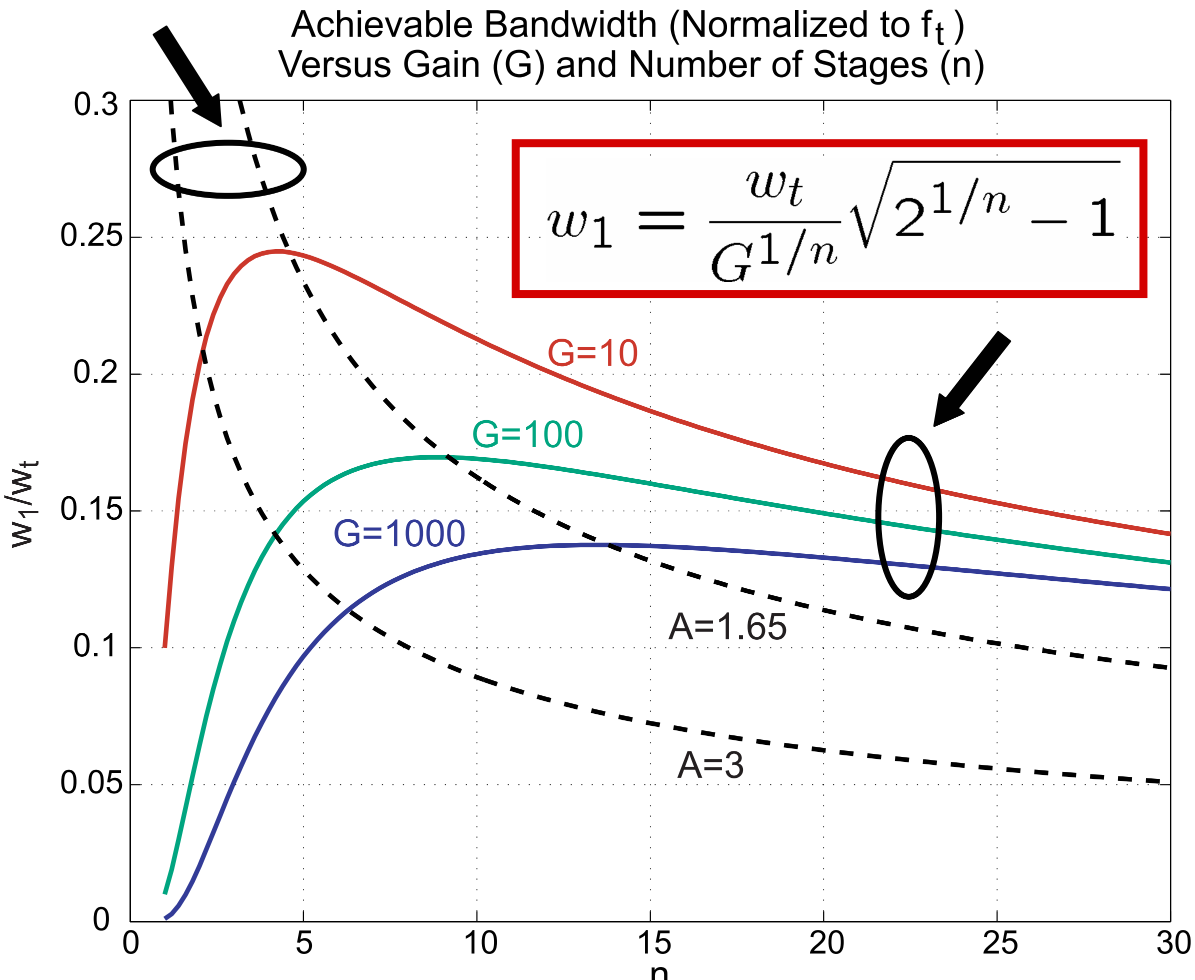
- Assume that we want to achieve gain G with n stages

$$\Rightarrow \quad A = G^{1/n} \quad \Rightarrow \quad \boxed{w_1 = \frac{w_t}{G^{1/n}}\sqrt{2^{1/n} - 1}}$$

- From this, Tom Lee finds optimum gain $\approx$ 1.65
  - See Tom Lee's book, pp 207-211

Achievable Bandwidth Versus G and n

$$\frac{w_t}{A}\sqrt{2^{1/n}-1}$$

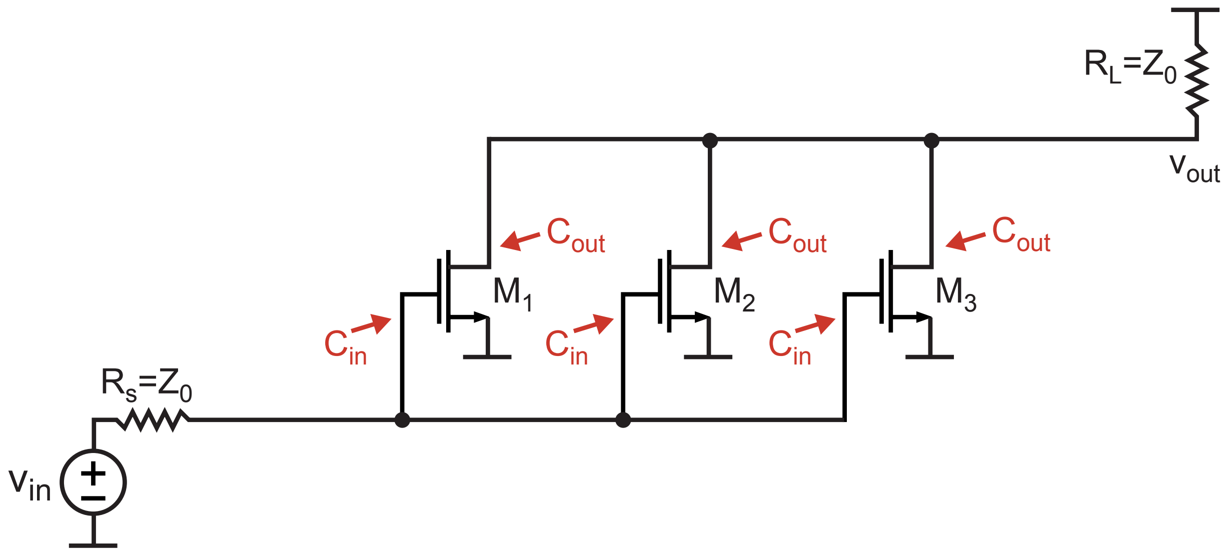


- **Optimum gain per stage is about 1.65**
  - **Note than gain per stage derived from plot as**

    $$A = G^{1/n}$$

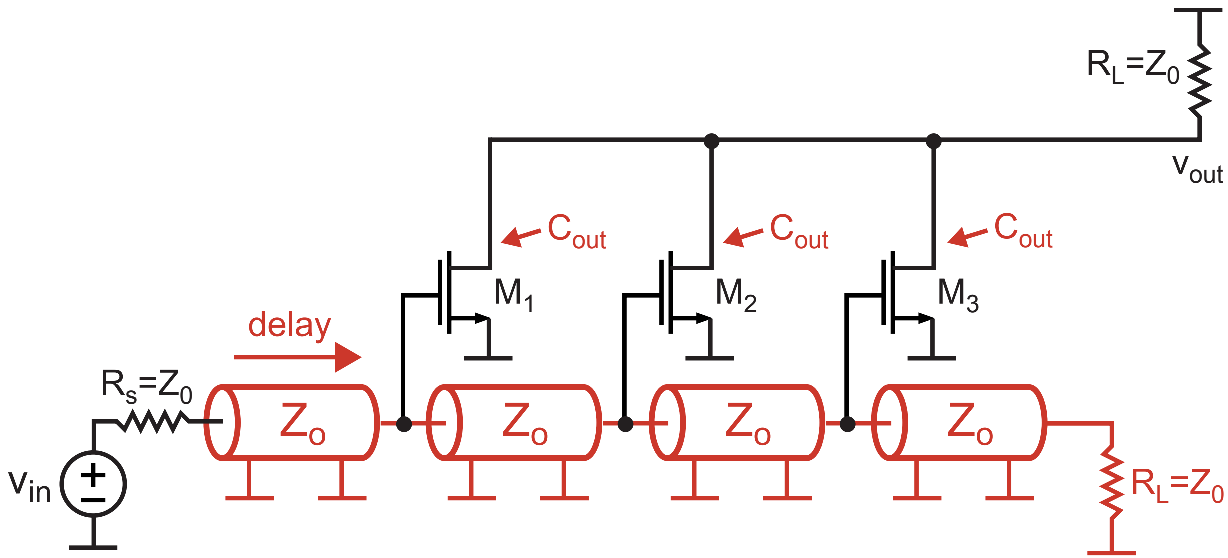
  - **Maximum is fairly soft, though**
- **Can dramatically lower power (and improve noise) by using larger gain per stage**

Motivation for Distributed Amplifiers



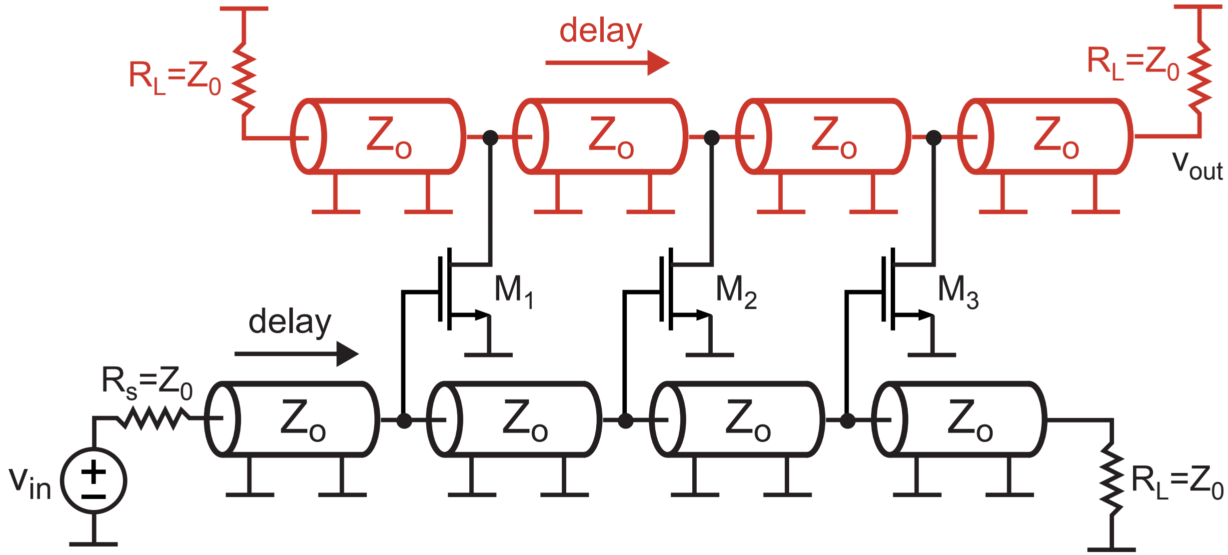
- **We achieve higher gain for a given load resistance by increasing the device size (i.e., increase g_m)**
  - **Increased capacitance lowers bandwidth**
    - We therefore get a relatively constant gain-bandwidth product
- **We know that transmission lines have (ideally) infinite bandwidth, but can be modeled as LC networks**
  - **Can we lump device capacitances into transmission line?**

Distributing the Input Capacitance



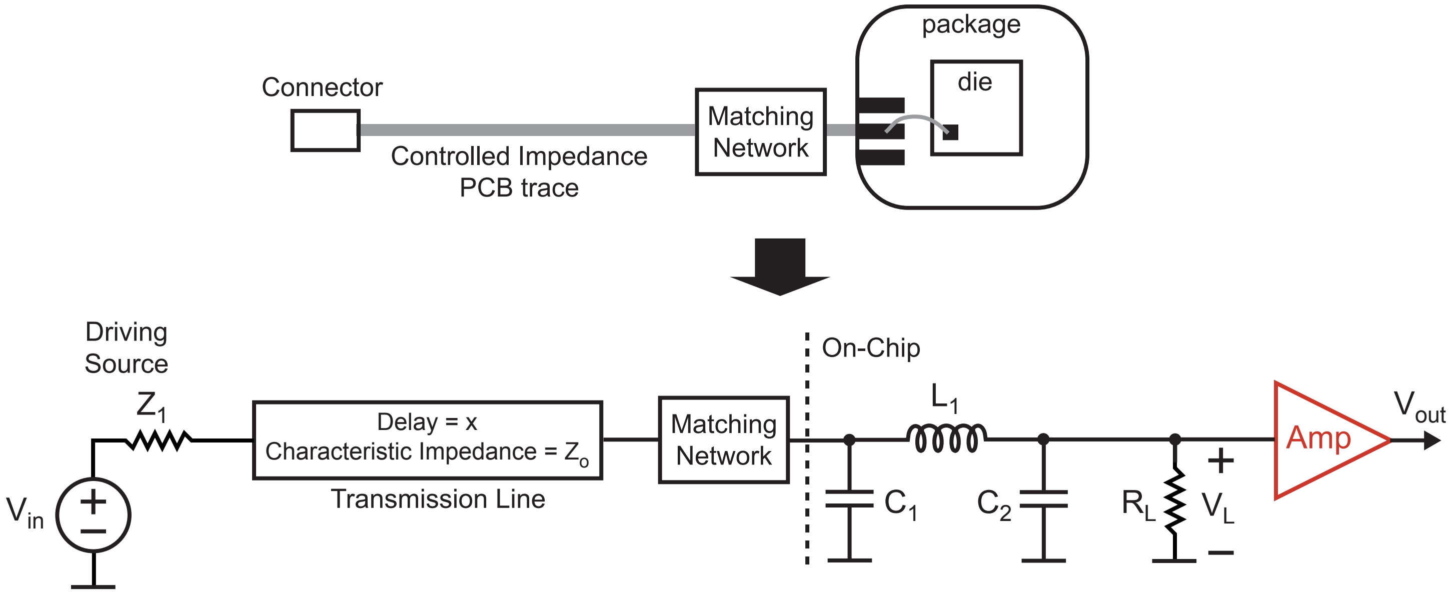
- **Lump input capacitance into LC network corresponding to a transmission line**
  - **Signal ideally sees a real impedance rather than an RC lowpass**
  - **Often implemented as lumped networks such as T-coils**
  - **We can now trade delay (rather than bandwidth) for gain**
- **Issue: outputs are delayed from each other**

Distributing the Output Capacitance



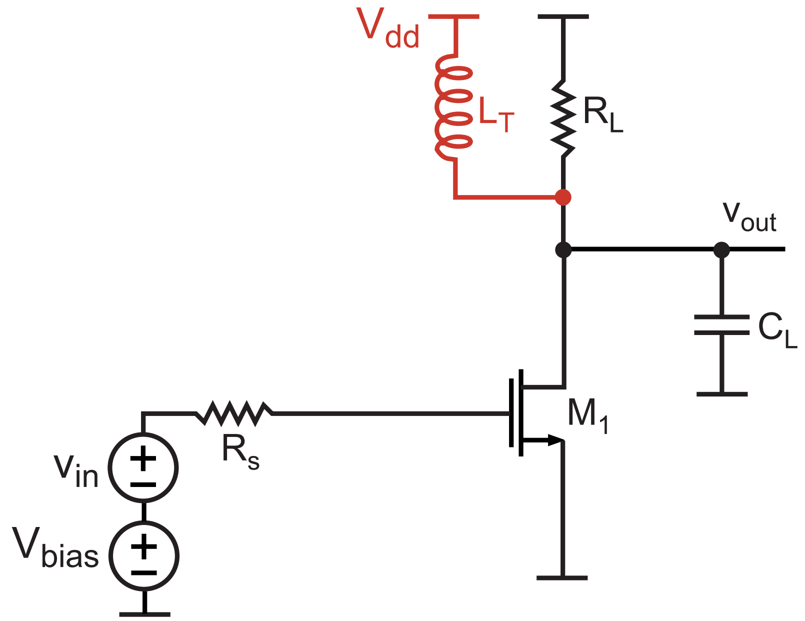
- **Delay the outputs same amount as the inputs**
  - **Now the signals match up**
  - **We have also distributed the output capacitance!**
- **Benefit – high bandwidth**
- **Negatives – high power, poorer noise performance, expensive in terms of chip area**
  - **Each transistor gain is adding rather than multiplying!**

Narrowband Amplifiers



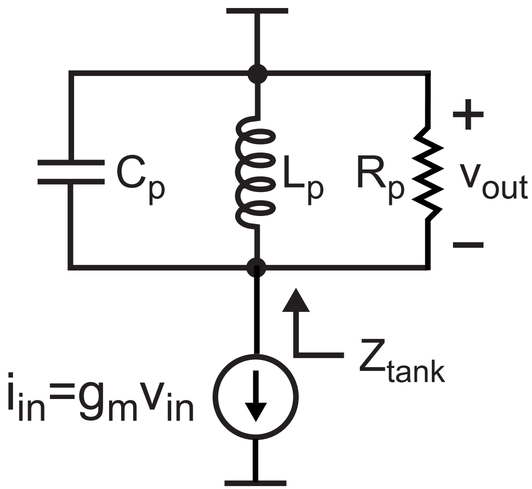
- **For wireless systems, we are interested in conditioning and amplifying the signal over a narrow frequency range centered at a high frequency**
  - **Allows us to apply narrowband transformers to create matching networks**
- **Can we take advantage of this fact when designing the amplifier?**

Tuned Amplifiers



- **Put inductor in parallel across R_L to create bandpass filter**
  - **It will turn out that the gain-bandwidth product is roughly conserved regardless of the center frequency!**
    - Assumes that center frequency (in Hz) << f_t
- **To see this and other design issues, we must look closer at the parallel resonant circuit**

Tuned Amp Transfer Function About Resonance



- **Amplifier transfer function**

$$\frac{v_{out}}{v_{in}} = g_m Z_{tank}(s) = \frac{g_m}{Y_{tank}(s)}$$

- **Note that conductances add in parallel**

$$Y_{tank}(s) = \frac{1}{R_p} + \frac{1}{sL_p} + sC_p$$

- **Evaluate at s = jw**

$$Y_{tank}(w) = \frac{1}{R_p} - \frac{j}{wL_p} + jwC_p = \frac{1}{R_p} + \frac{j}{wL_p}\left(-1 + w^2 L_p C_p\right)$$

- **Look at frequencies about resonance:** $w = w_o + \Delta w$

$$\Rightarrow Y_{tank}(\Delta w) = \frac{1}{R_p} + \frac{j}{(w_o + \Delta w)L_p}\left(-1 + (w_o + \Delta w)^2 L_p C_p\right)$$

$$\approx \frac{1}{R_p} + \frac{j}{w_o L_p}\left(-1 + w_o^2 L_p C_p + 2 w_o \Delta w L_p C_p\right)$$

Tuned Amp Transfer Function About Resonance (Cont.)

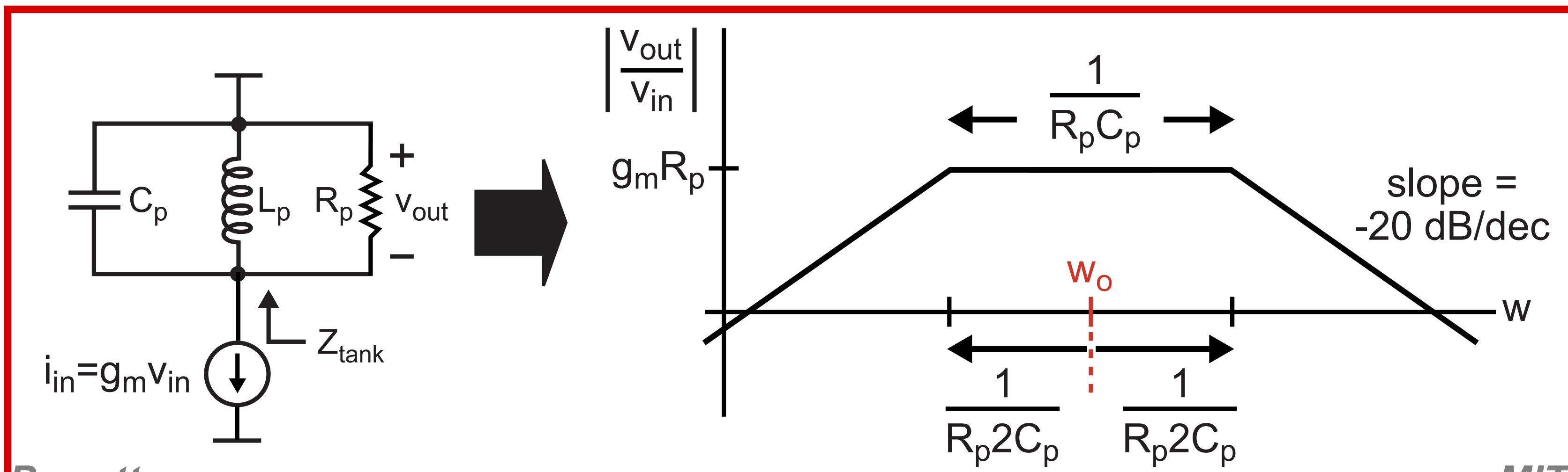
- **From previous slide**

$$Y_{tank}(\Delta w) \approx \frac{1}{R_p} + \frac{j}{w_o L_p}\Big(\underbrace{-1 + w_o^2 L_p C_p}_{=0} + 2 w_o \Delta w L_p C_p\Big)$$

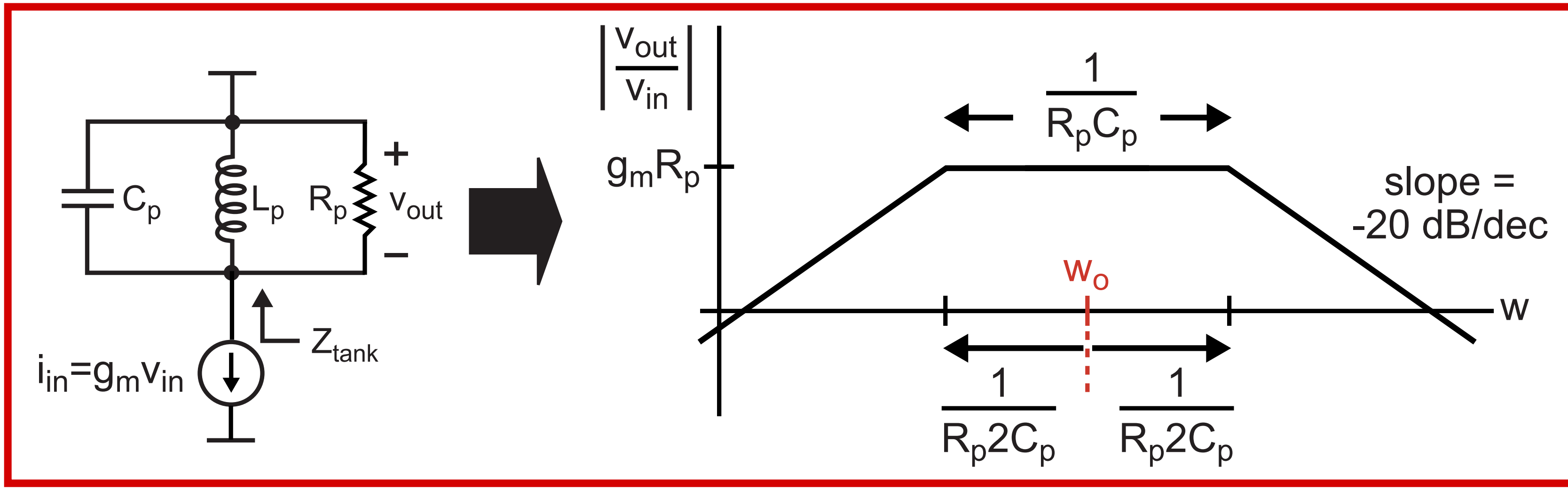
$$\approx \frac{1}{R_p} + \frac{j}{w_o L_p}(2 w_o \Delta w L_p C_p) = \frac{1}{R_p} + j \Delta w 2 C_p$$

- **Simplifies to RC circuit for bandwidth calculation!**

$$Z_{tank}(\Delta w) \approx R_p || \frac{1}{j \Delta w 2 C_p}$$



Gain-Bandwidth Product for Tuned Amplifiers

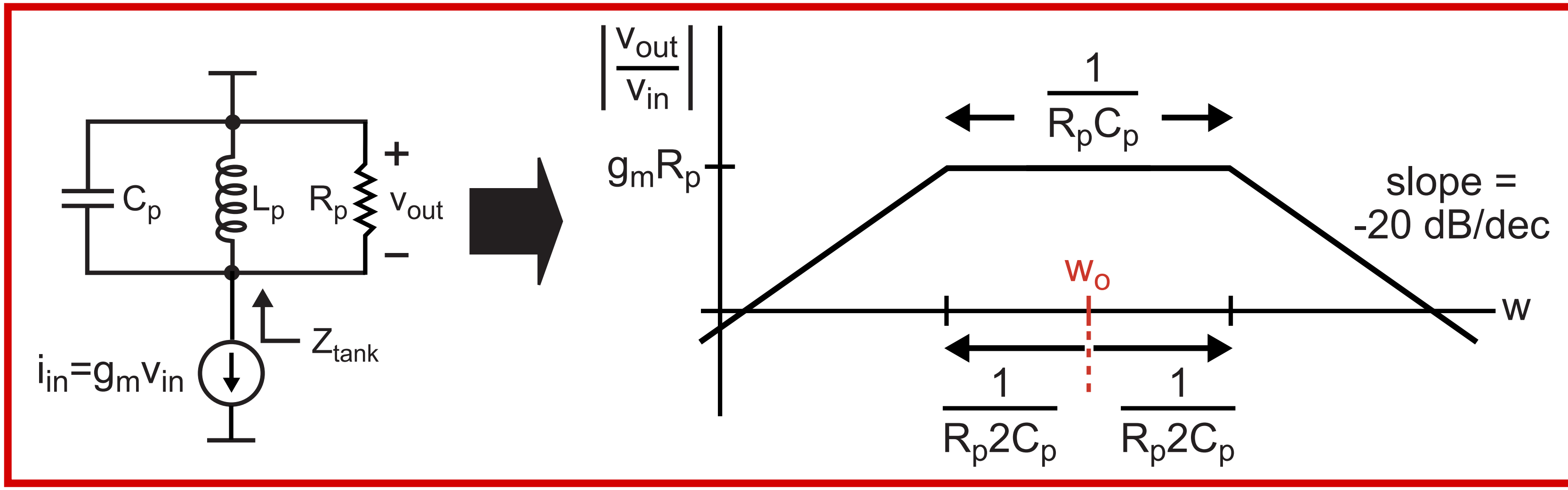


- The gain-bandwidth product:

$$G \cdot BW = g_m R_p \frac{1}{R_p C_p} = \frac{g_m}{C_p}$$

- The above expression is independent of center frequency!
  - In practice, we need to operate at a frequency less than the f_t of the device

The Issue of Q

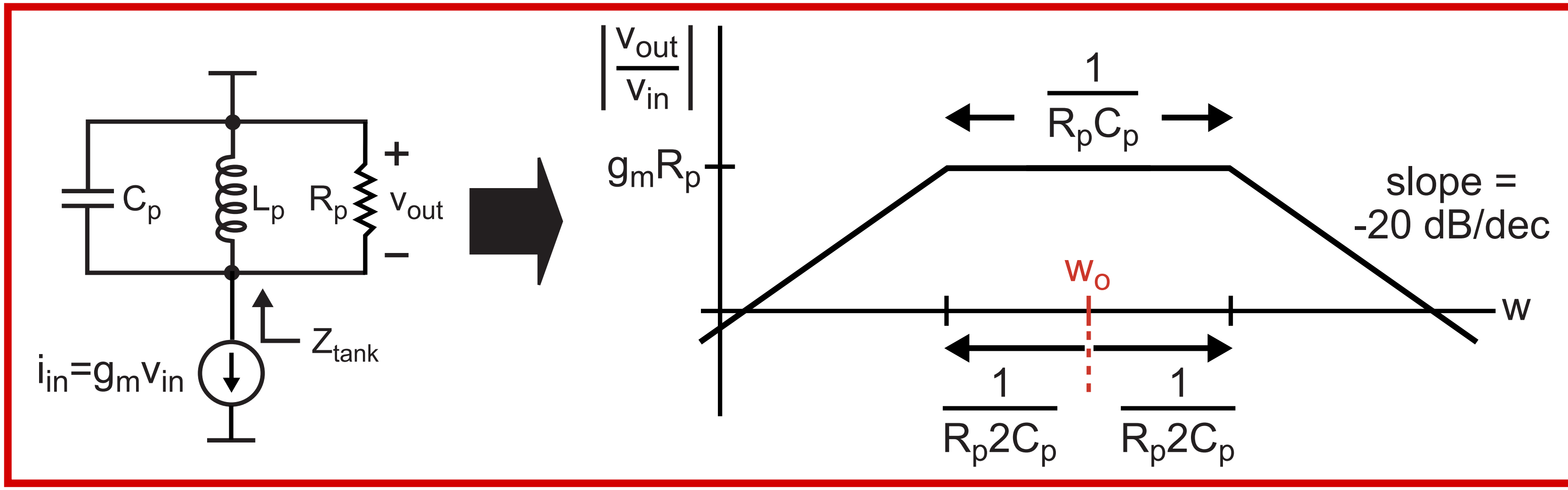


- **By definition** $$Q = w\frac{\text{energy stored}}{\text{average power dissipated}}$$
- **For parallel tank (see Tom Lee's book, pp 88-89)**

$$\text{at resonance: } Q = \frac{R_p}{w_o L_p} = w_o R_p C_p$$

- **Comparing to above:** $Q = w_o R_p C_p = \frac{w_o}{1/(R_pC_p)} = \boxed{\frac{w_o}{BW}}$

Design of Tuned Amplifiers

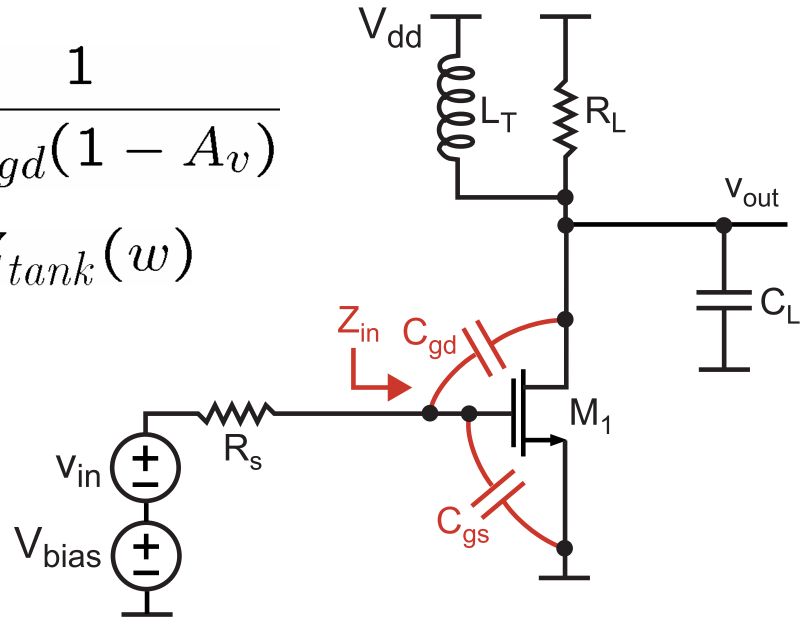


- **Three key parameters**
  - **Gain = $g_m R_p$**
  - **Center frequency = w_o**
  - **Q = w_o/BW**
- **Impact of high Q**
  - **Benefit: allows achievement of high gain with low power**
  - **Problem: makes circuit sensitive to process/temp variations**

Issue: C_{gd} Can Cause Undesired Oscillation

$$Z_{in}(w) = \frac{1}{jwC_{gs}} || \frac{1}{jwC_{gd}(1-A_v)}$$

where $A_v = -g_m Z_{tank}(w)$



- **At frequencies below resonance, tank looks inductive**

$$A_v \approx -g_m(jwL) \quad \Rightarrow \quad Z_{in}(w) \approx \frac{1}{jwC_{gs}} || \frac{1}{jwC_{gd}(1+g_m(jwL))}$$

$$\Rightarrow \quad Z_{in}(w) \approx \frac{1}{jwC_{gs}} || \frac{1}{jwC_{gd} - w^2 g_m C_{gd} L}$$

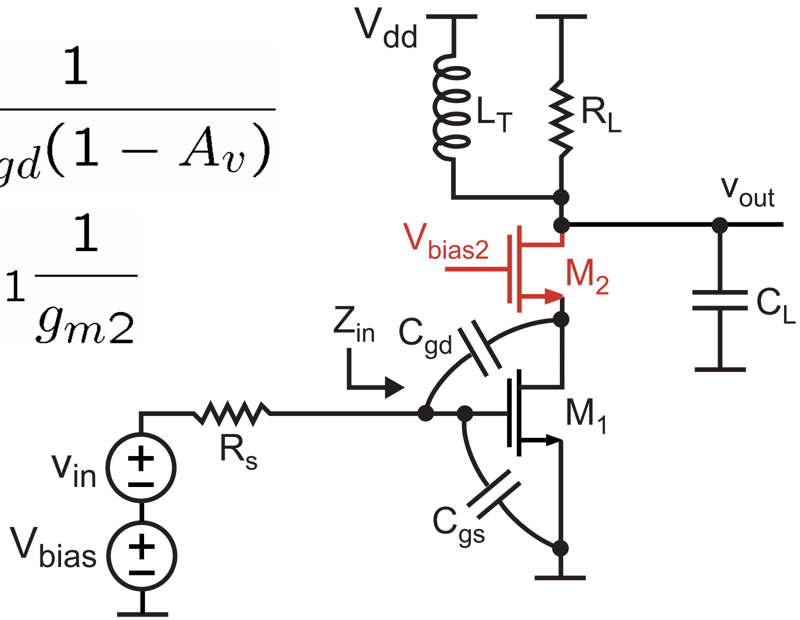
$$\Rightarrow \quad Z_{in}(w) \approx \frac{1}{jwC_{gs}} || \frac{1}{jwC_{gd}} || \frac{-1}{w^2 g_m C_{gd} L}$$

Negative Resistance!

Use Cascode Device to Remove Impact of C_{gd}

$$Z_{in}(w) = \frac{1}{jwC_{gs}} \,||\, \frac{1}{jwC_{gd}(1-A_v)}$$

$$\text{where } A_v = -g_{m1}\frac{1}{g_{m2}}$$

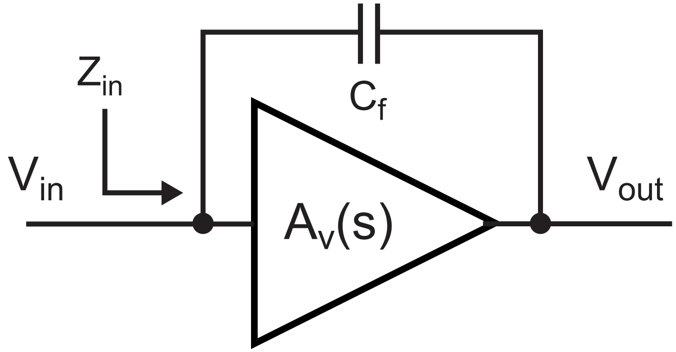


- **At frequencies above and below resonance**

$$Z_{in}(w) = \frac{1}{jwC_{gs}} \,||\, \frac{1}{jwC_{gd}(1+g_{m1}/g_{m2})}$$

Purely Capacitive!

Active Real Impedance Generator



$A_v(s) = -A_o e^{-j\Phi}$

- **Input impedance:**

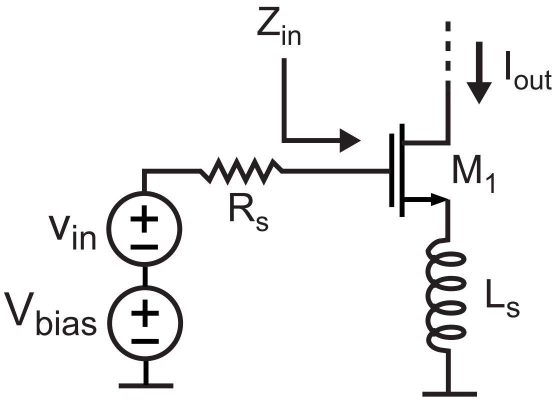
$$Z_{in}(w) = \frac{1}{jwC_f(1 - A_v)} = \frac{1}{jwC_f(1 + A_o e^{-j\Phi})}$$

$$= \frac{1}{jwC_f(1 + A_o \cos\Phi) + A_o wC \sin\Phi}$$

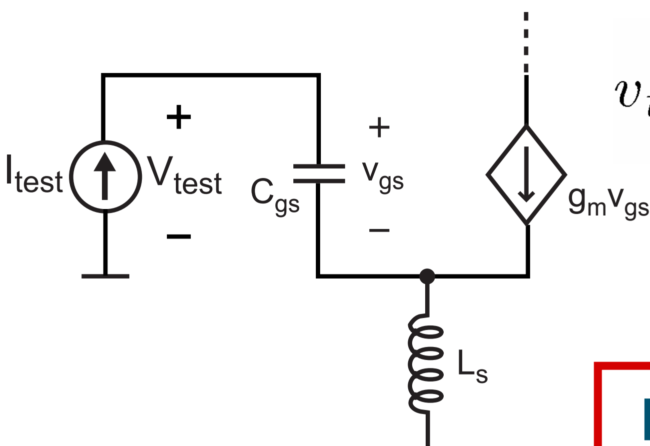
$$= \frac{1}{jwC_f(1 + A_o \cos\Phi)} \,||\, \frac{1}{A_o wC \sin\Phi}$$

Resistive component!

This Principle Can Be Applied To Impedance Matching



- **We will see that it's advantageous to make Z_{in} real without using resistors**
  - **For the above circuit (ignoring C_{gd})**

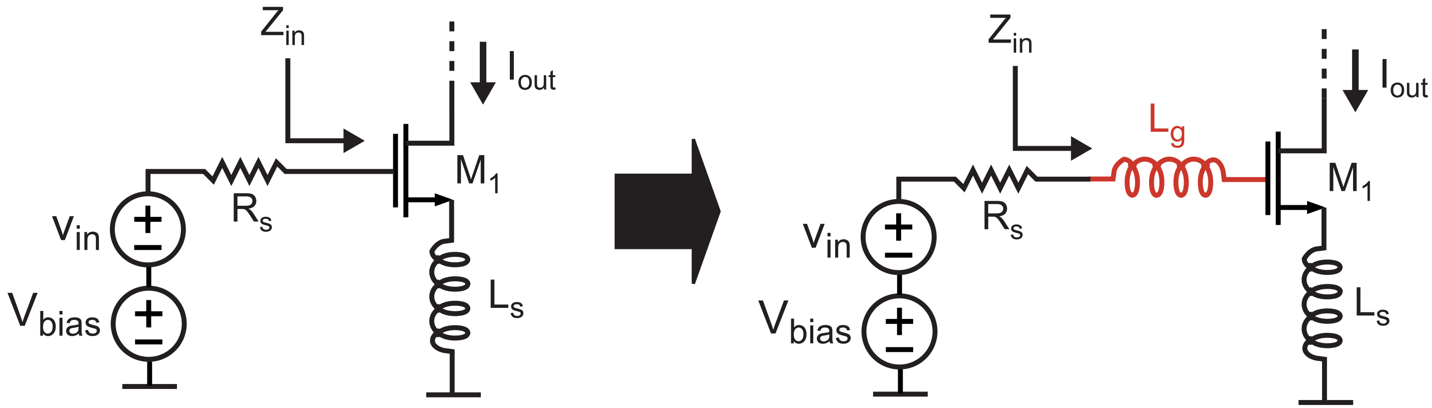


$$v_{test} = i_{test}\left(\frac{1}{sC_{gs}} + \left(1 + g_m\frac{1}{sC_{gs}}\right)sL_s\right)$$

$$\Rightarrow \quad Z_{in}(s) = \frac{1}{sC_{gs}} + sL_s + \frac{g_m}{C_{gs}}L_s$$

$$\boxed{\text{Looks like series resonant circuit!}}$$

Use A Series Inductor to Tune Resonant Frequency



- **Calculate input impedance with added inductor**

$$Z_{in}(s) = \frac{1}{sC_{gs}} + s(L_s + L_g) + \frac{g_m}{C_{gs}} L_s$$

- **Often want purely resistive component at frequency w_o**
  - **Choose L_g such that resonant frequency = w_o**

$$\text{i.e., want} \quad \frac{1}{\sqrt{(L_s + L_g)C_{gs}}} = w_o$$