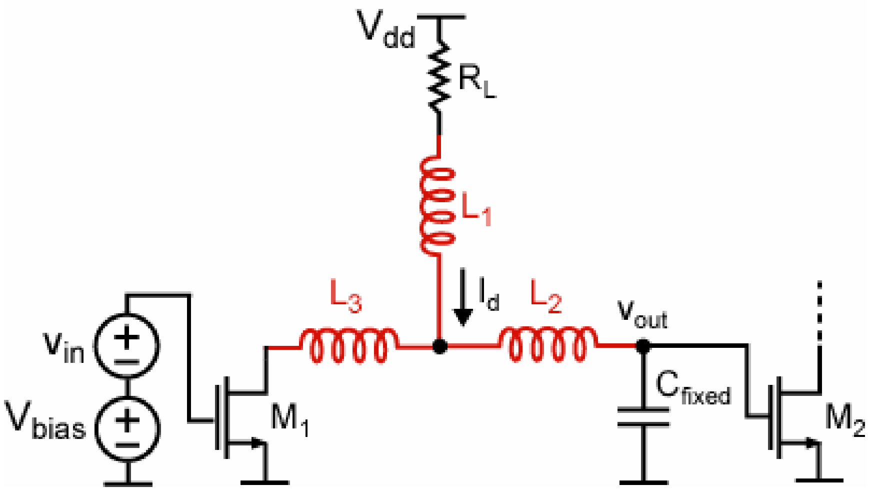
6.776
High Speed Communication Circuits and Systems
Lecture 9
Enhancement Techniques for Broadband Amplifiers, Narrowband Amplifiers

Massachusetts Institute of Technology

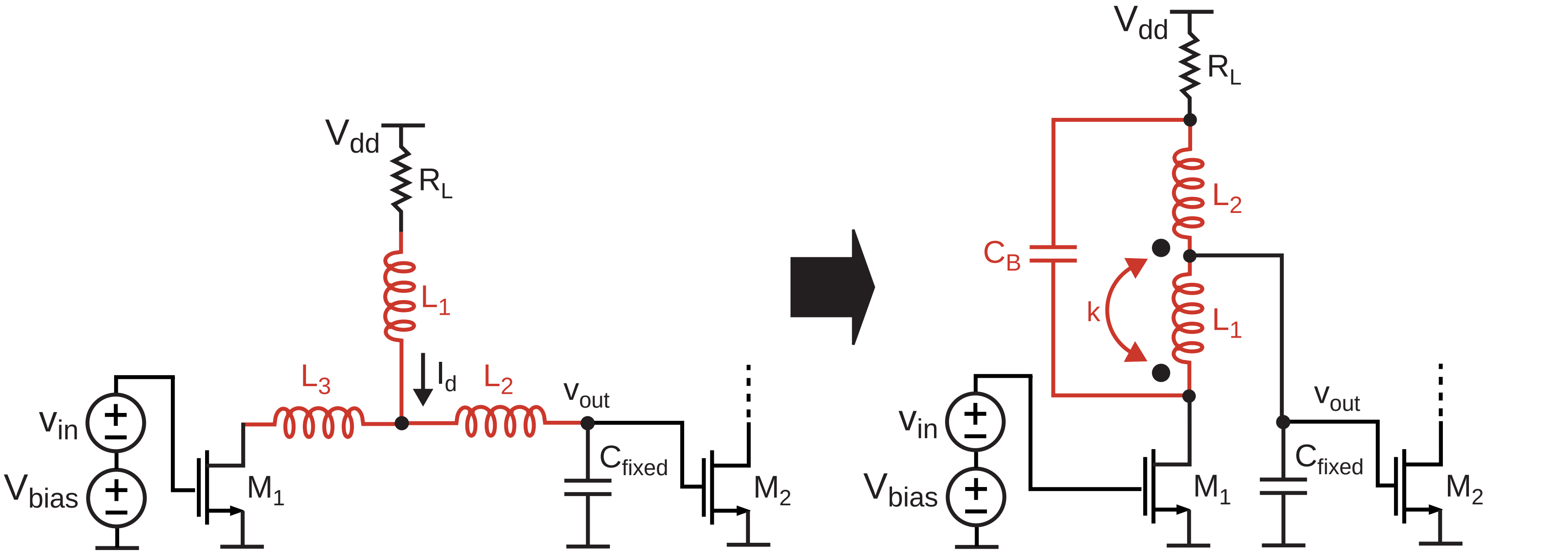
March 3, 2005

Shunt-Series Peaking



- **Series inductors isolate load capacitance from M_1: delays charging of load capacitance**
- **Trades delay for bandwidth**
- **L_1, L_2, L_3 can be implemented by 2 coupled inductors with coupling coefficient of k**

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**
- **C_B provides parallel resonance to improve bandwidth further See Chap. 9 (Ch. 8, 1st ed.) of Tom Lee's book pp 279-282 (187-191)**

T-Coil Continued

- **The self inductance L (with the other winding open-circuited) must be**

$$L = \frac{R_L^2 C_L}{2(1+k)}$$

- **The bridging capacitance**

$$C_B = \frac{C_L(1-k)}{4(1+k)}$$

- **Coupling coefficient**

k=1/3 for Butterworth response

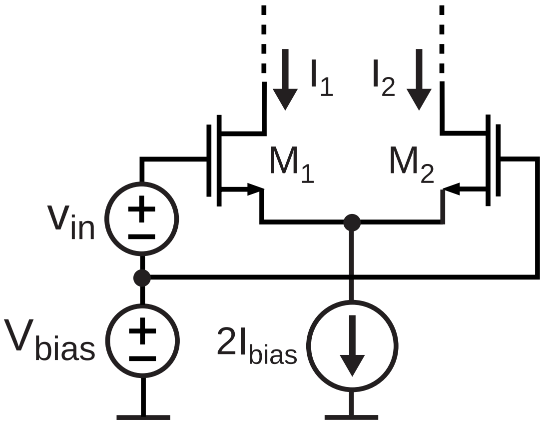
k=1/2 for maximally flat delay (linear phase)

Bandwidth extension: approximately 2.8 (Butterworth)

- 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

- Also see "Circuit Techniques for a 40 Gb/s Transmitter in 0.13um CMOS", J. Kim, et. al. ISSCC 2005, Paper 8.1

Bandwidth Enhancement With f_t Doublers



- **A MOS transistor has f_t calculated as**

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

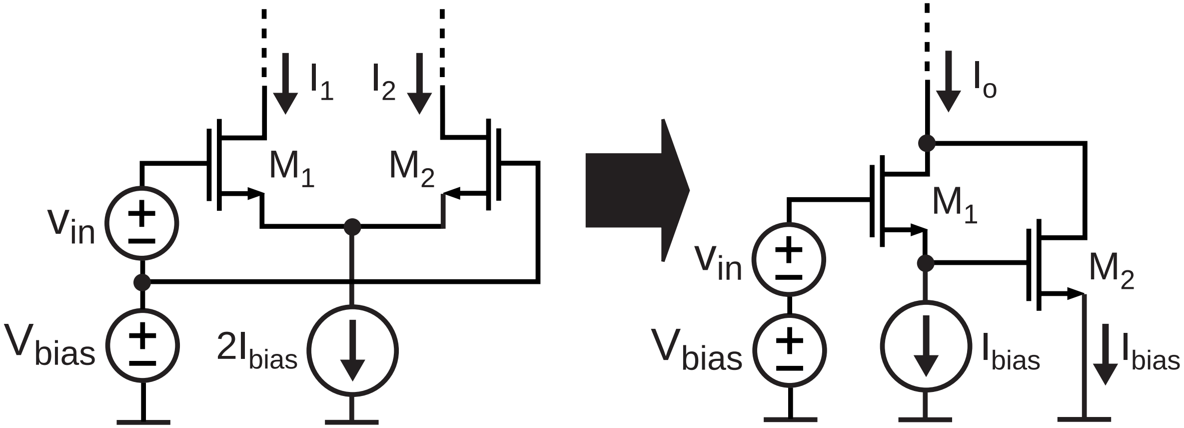
- **f_t doubler amplifiers attempt to increase the ratio of transconductance to capacitance**

- **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:**

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

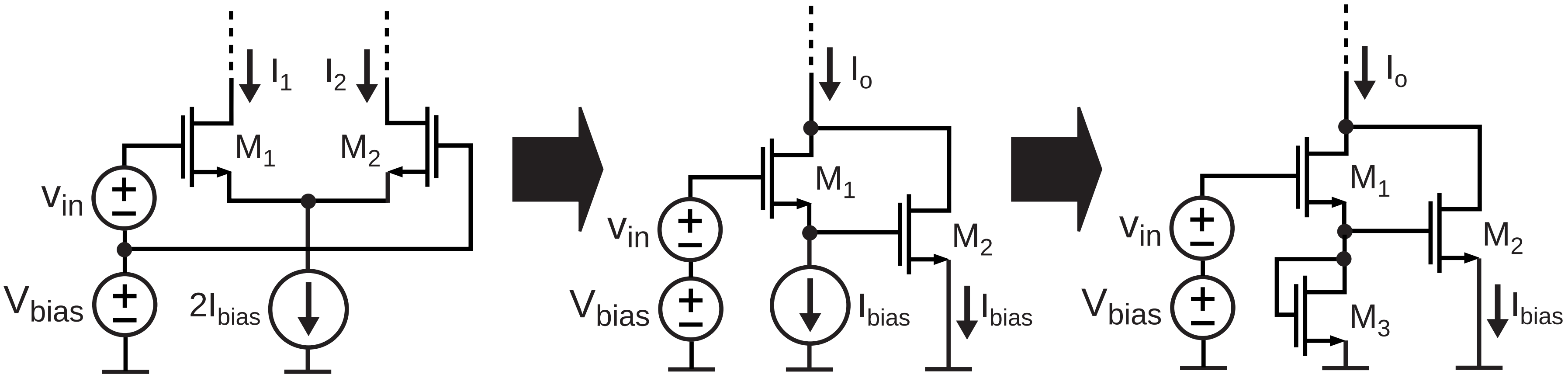
- **Transconductance to Cap ratio is doubled:** $\dfrac{2g_m}{C_{gs}}$

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 (Battjes f_t doubler)
  - Inspired by bipolar circuits (see Tom Lee's book, pp288-290 (197-199))
- 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

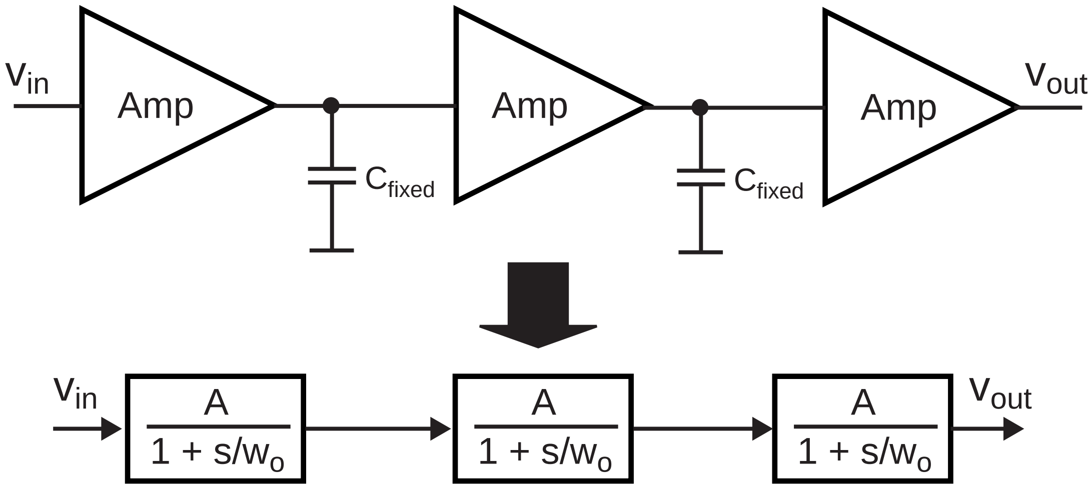
Problems of f_t Doubler in Modern CMOS RF Circuits

- **Problems:**
  - **Works if Cgs dominates capacitance , but in modern CMOS, this is not the case (for example, C_{gd}=0.45C_{gs} in 0.18 μ CMOS)**
  - **achievable bias voltage across M_1 (and M_2) is severely reducedby 2x!) (thereby reducing effective f_t of device)**
  - **Input capacitance degrades due to C_{gs}, C_{db} of M_3: at most 1.5x improvement in transconductance/capacitance ratio**

Assuming zero Cdb:

$$C_{in} = C_{gs} || 2C_{gs} = \frac{1}{1.5} C_{gs}$$

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**

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

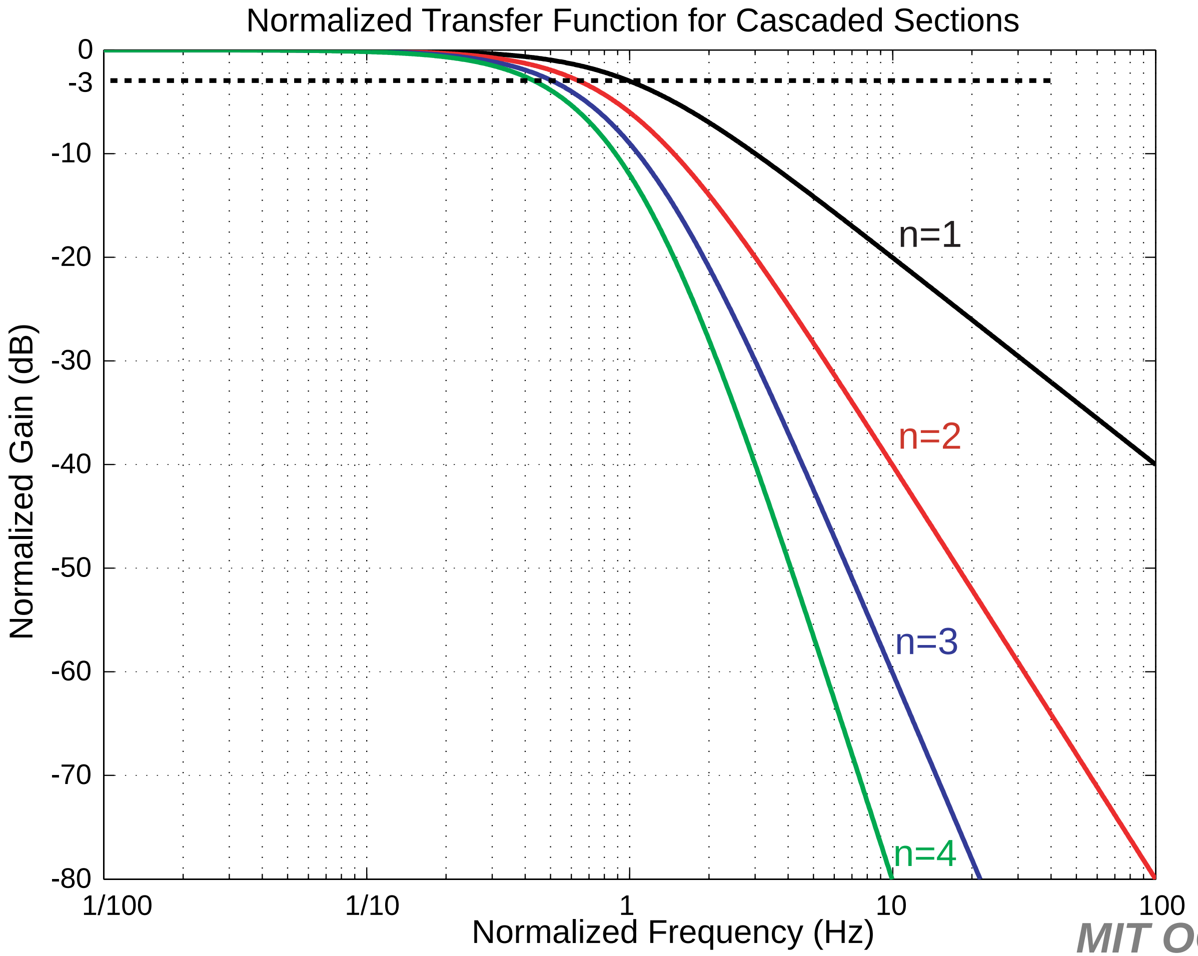
$$\Rightarrow \quad \left(1 + (w_1/w_o)^2\right)^n = 2$$

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

  - **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**

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

  - **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**

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

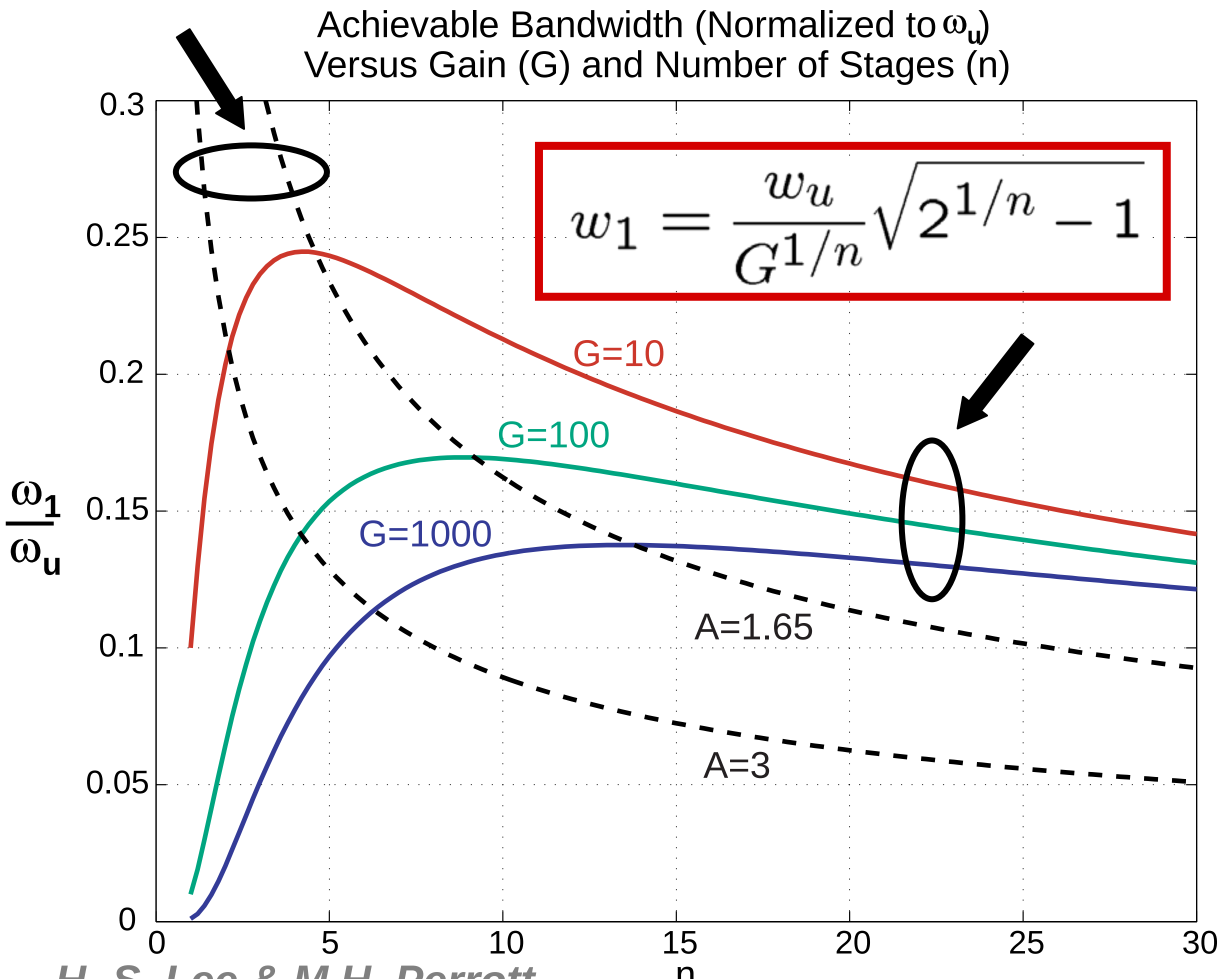
- **Assume that we want to achieve gain G with n stages**

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

- **From this, optimum gain/stage ≈ sqrt(e) = 1.65**
  - **See Tom Lee's book, pp 299-302 (207-211,1st ed.)**

Achievable Bandwidth Versus G and n

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

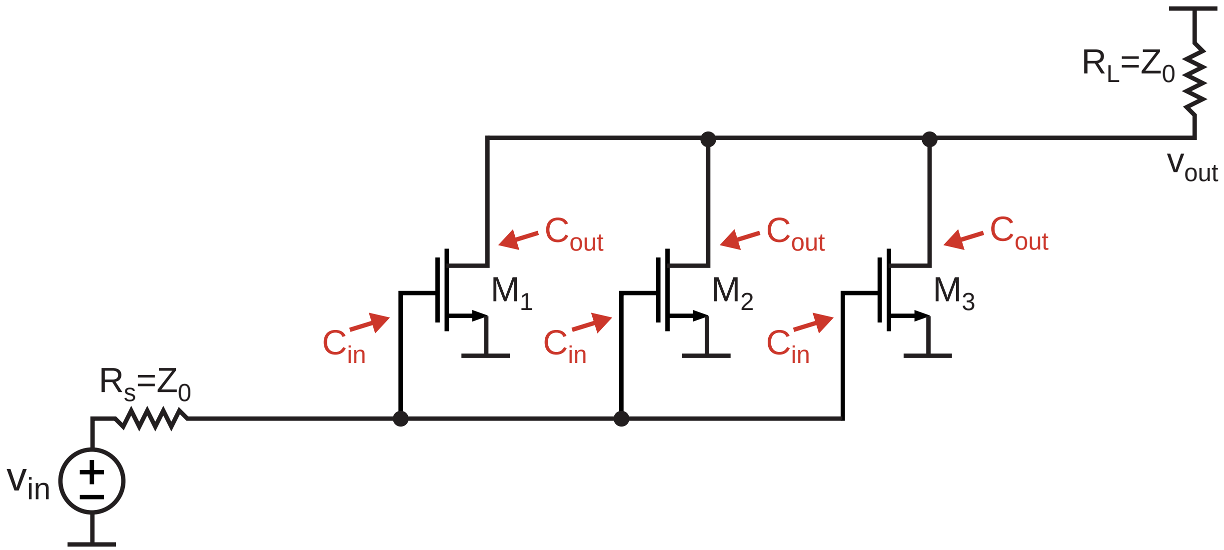


- **Optimum gain per stage is about 1.65**
  - **Note than gain per stage derived from plot as**

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

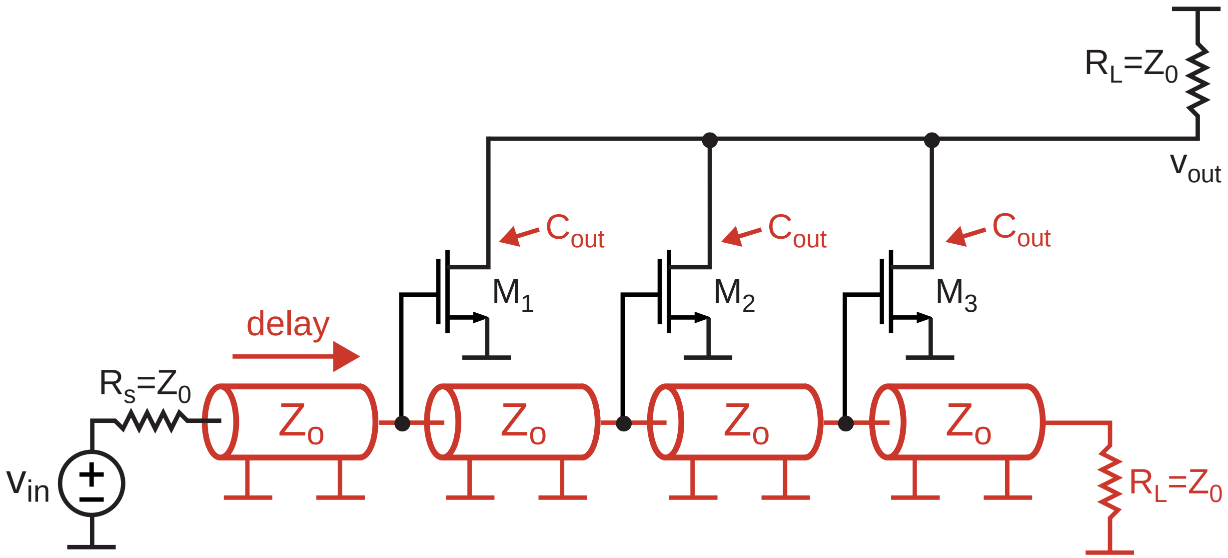
  - **Maximum is fairly soft, though**
- **Can dramatically lower power (and improve noise) by using larger gain per stage**

Motivation for Distributed Amplifiers



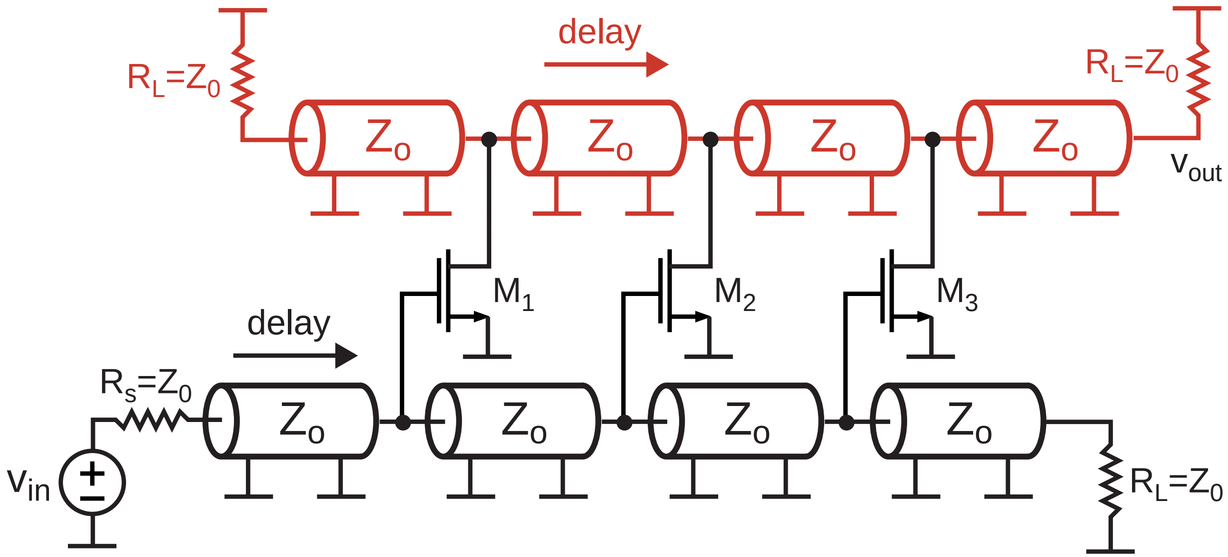
- **We achieve higher gain for a given load resistance by increasing the device size (i.e., increase g_m)**
  - **Increased capacitance lowers bandwidth**
    - We therefore get a relatively constant gain-bandwidth product
- **We know that transmission lines have (ideally) infinite bandwidth, but can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

Distributing the Input Capacitance



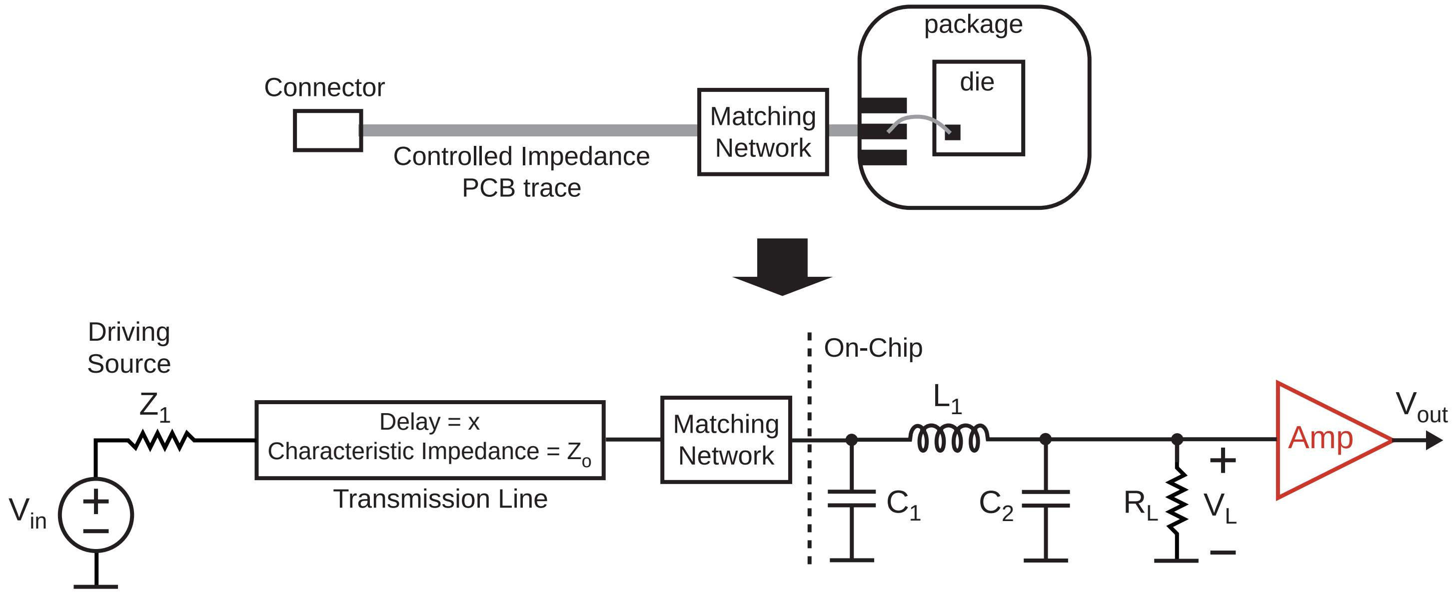
- **Lump input capacitance into LC network corresponding to a transmission line**
  - **Signal ideally sees $Z_o=R_L$ rather than an RC lowpass**
  - **Often implemented as lumped networks such as T-coils**
  - **We can now trade delay (rather than bandwidth) for gain**
- **Issue: outputs are delayed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

Distributing the Output Capacitance



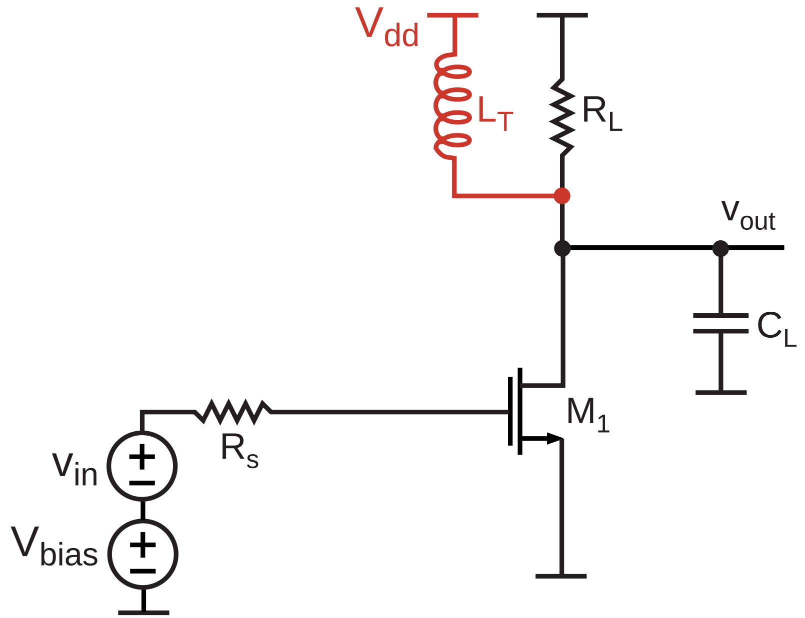
- **Delay the outputs same amount as the inputs**
  - **Now the signals match up**
  - **We have also distributed the output capacitance**
- **Benefit – high bandwidth**
- **Negatives – high power, poorer noise performance, expensive in terms of chip area**
  - **Each transistor gain is adding rather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

Narrowband Amplifiers



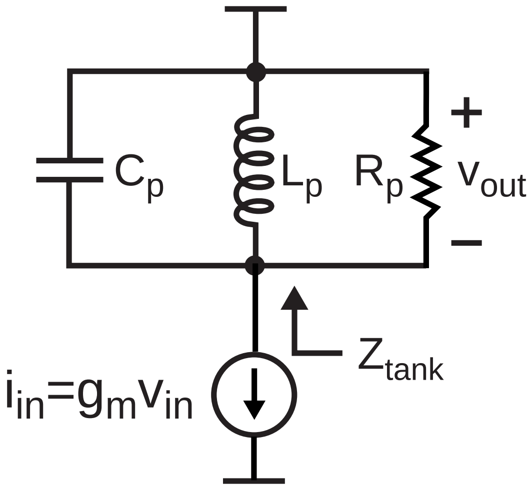
- For wireless systems, we are interested in conditioning and amplifying the signal over a narrow frequency range centered at a high frequency
  - Allows us to apply narrowband transformers to create matching networks
- Can we take advantage of this fact when designing the amplifier?

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**
- **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 = jω**

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

- **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.)

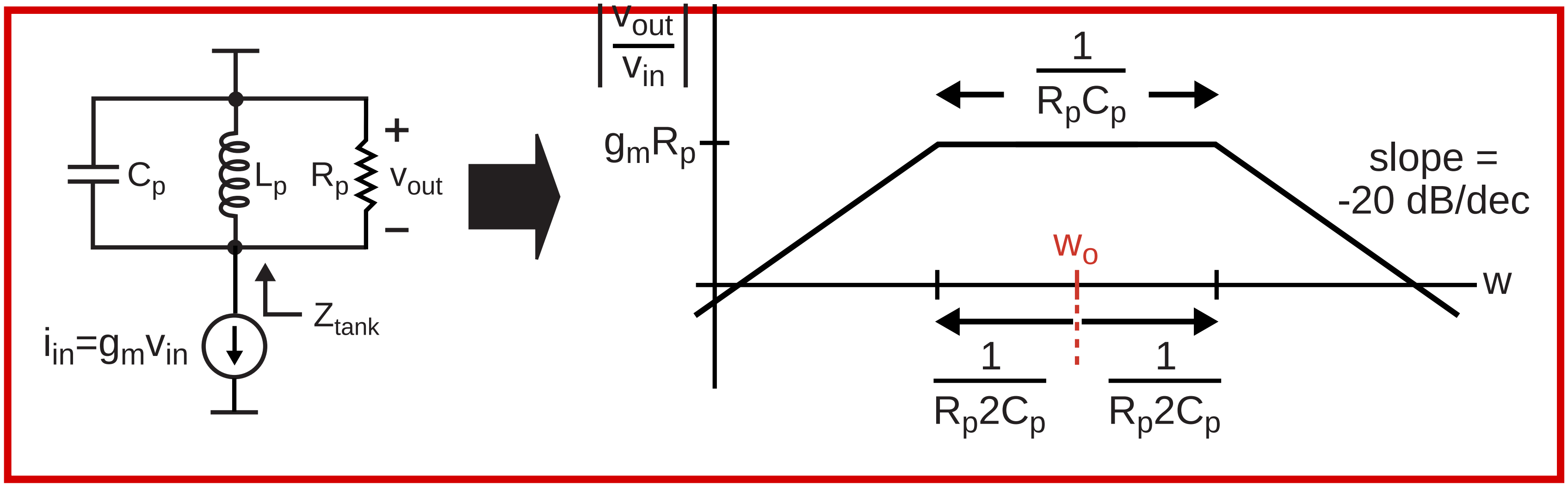
- **From previous slide**

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

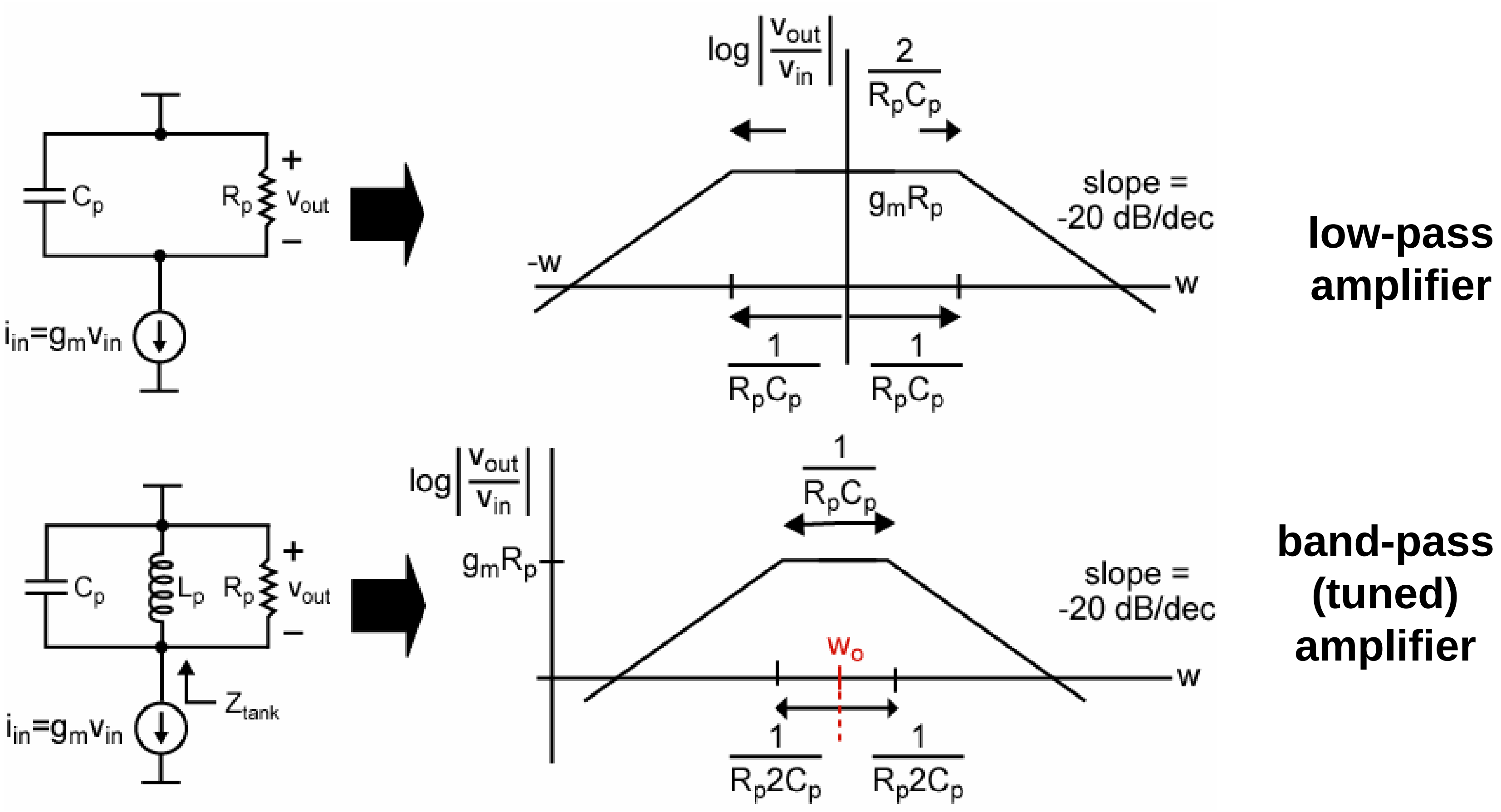
$$\approx \frac{1}{R_p} + \frac{j}{w_o L_p}(2 w_o \Delta w L_p C_p) = \frac{1}{R_p} + j \Delta w 2 C_p$$

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

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



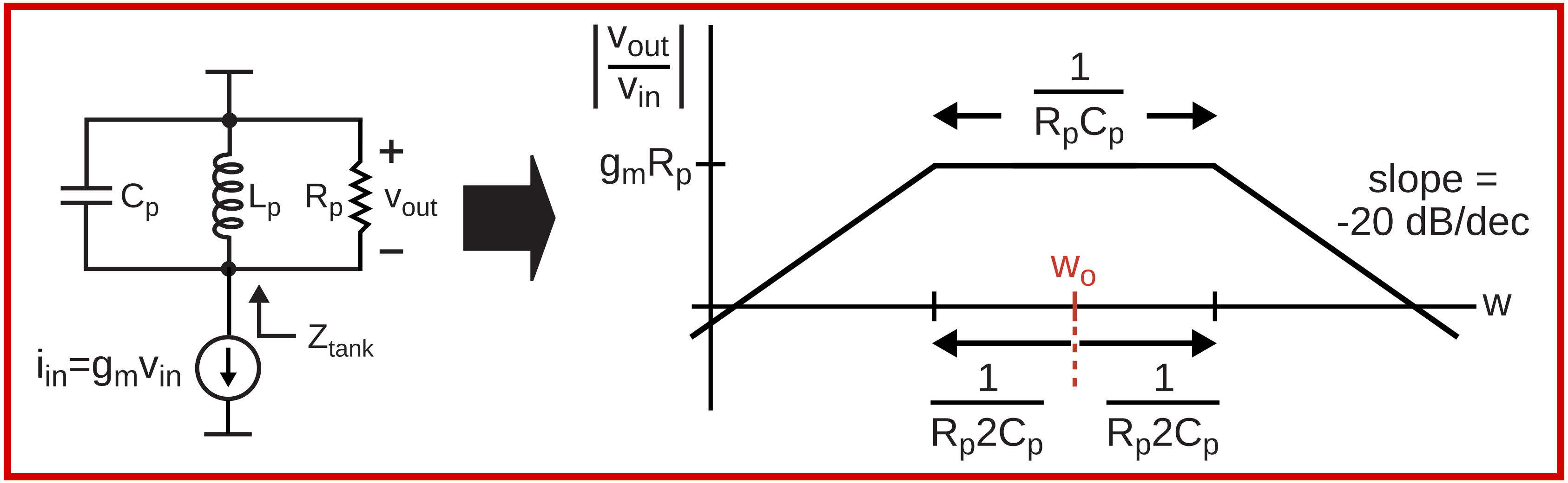
Comparison between Low-Pass and Band-Pass



Comparison between Low-Pass and Band-Pass

- **Tuned amplifier characteristic is a frequency translated version of the low-pass amplifier**
- **The band-pass bandwidth is equal to the low-pass bandwidth (the band-pass *shape* is 2x narrower but upper and lower sidebands give the same bandwidth as LP)**
- **We are *tuning out* the effect of capacitor (parallel LC looks like an open circuit at resonance, so C doesn't load the amplifier)**
- **This is often called low-pass to band-pass transform in filter design:**
  - **Replace C with parallel LC tank**
  - **Replace L with series LC tank**

Gain-Bandwidth Product for Tuned Amplifiers

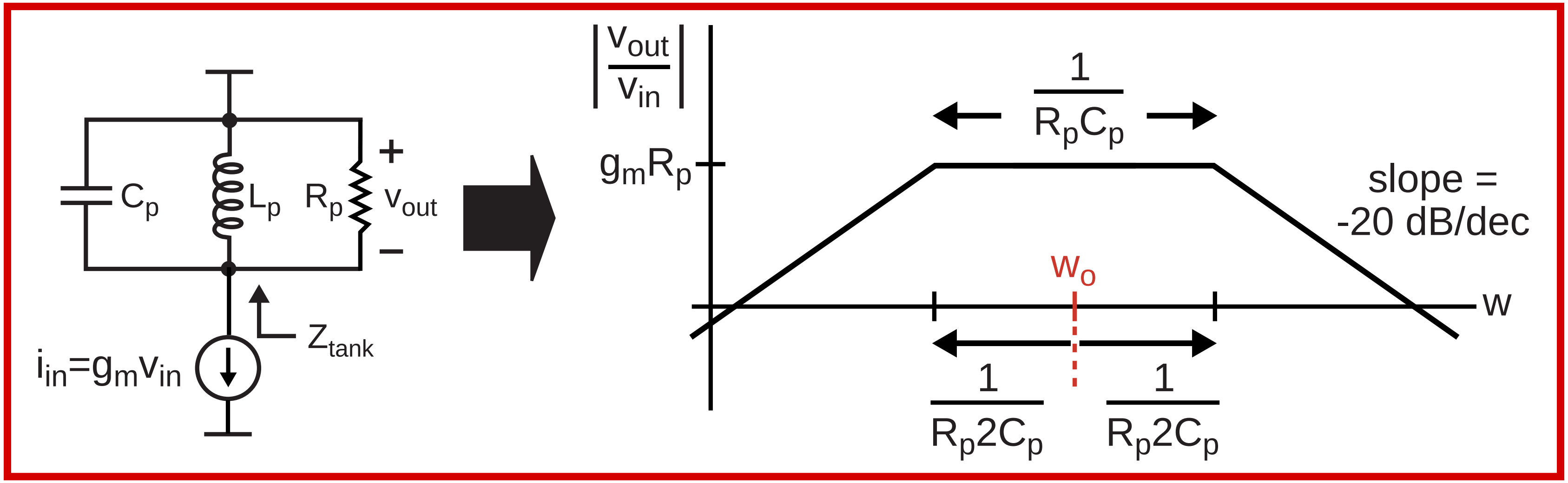


- **The gain-bandwidth product:**

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

- **The above expression is just like the low-pass and 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

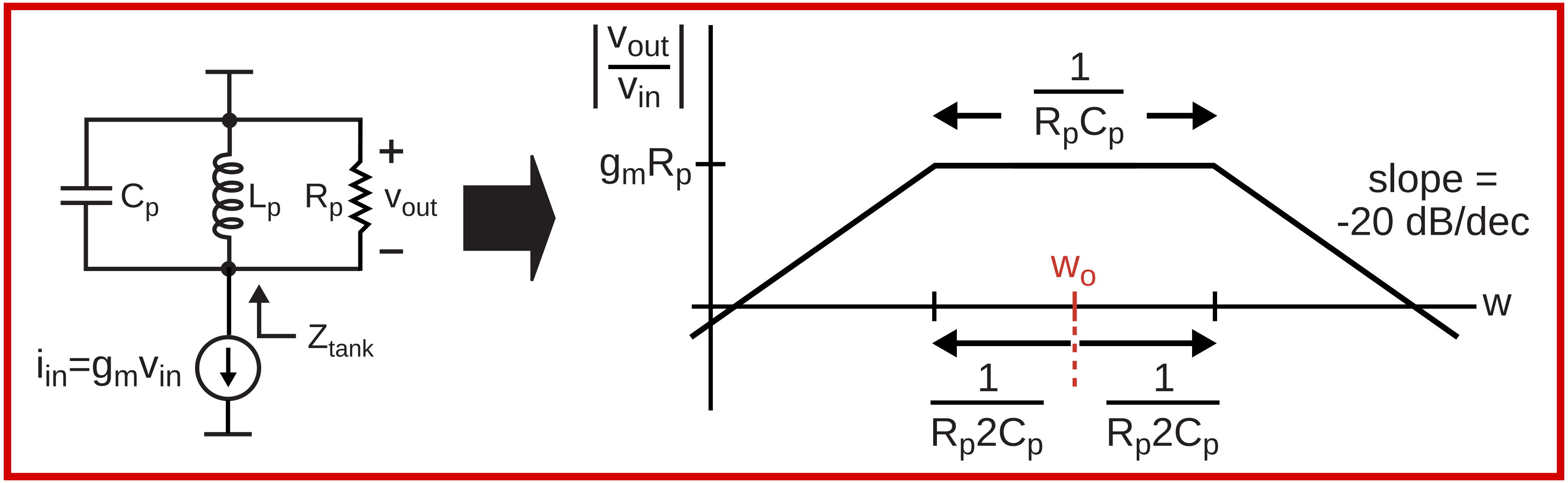


- **By definition** $$Q = w\frac{\text{energy stored}}{\text{average power dissipated}}$$
- **For parallel tank**

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

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

Design of Tuned Amplifiers

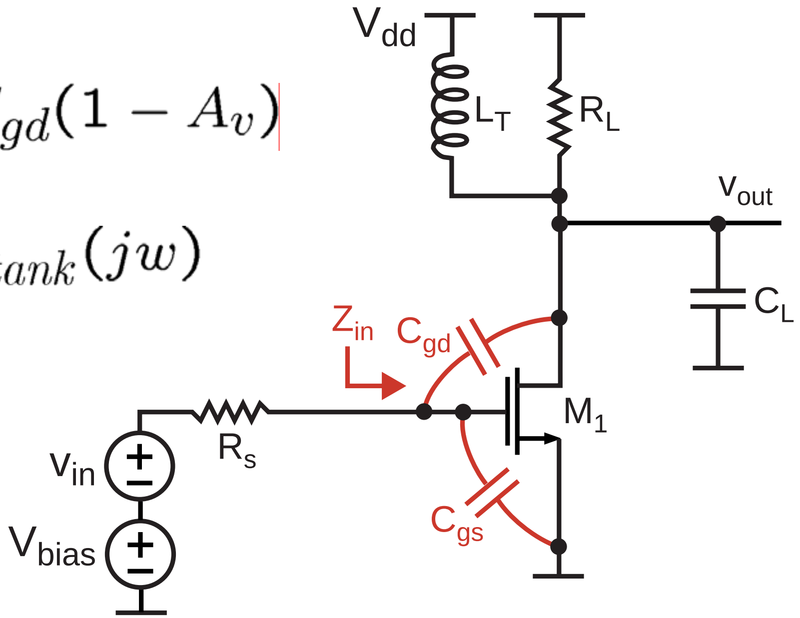


- **Three key parameters**
  - **Gain = $g_m R_p$**
  - **Center frequency = ω_o**
  - **Q = ω_o/BW**
- **Impact of high Q**
  - **Benefit: allows achievement of high gain with low power**
  - **Problem: makes circuit sensitive to process/temp variations**

Issue: C_{gd} Can Cause Undesired Oscillation

$$Y_{in}(w) = jwC_{gs} + jwC_{gd}(1 - A_v)$$

$$\text{where } A_v = -g_m Z_{tank}(jw)$$



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

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

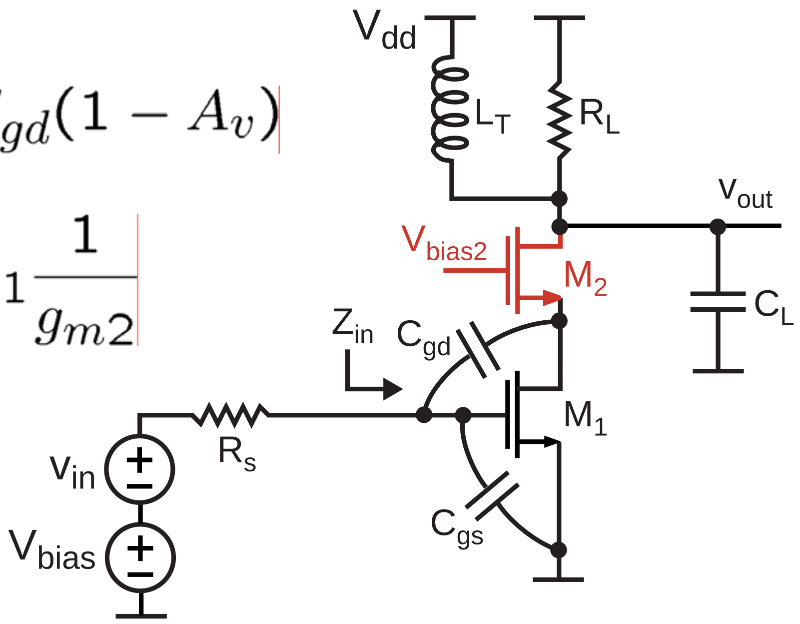
$$\Rightarrow \quad Y_{in}(w) \approx jwC_{gs} + jwC_{gd} - w^2 g_m C_{gd} L$$

Negative Resistance!

Use Cascode Device to Remove Impact of C_{gd}

$$Y_{in}(w) = jwC_{gs} + jwC_{gd}(1 - A_v)$$

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



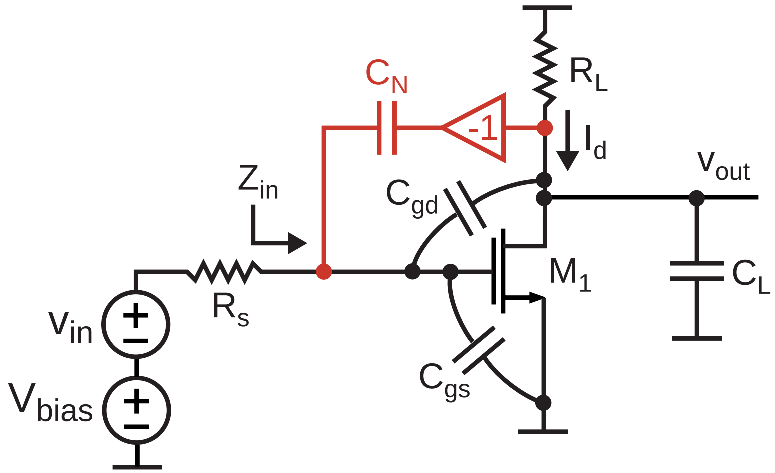
- **At frequencies above and below resonance**

$$Y_{in}(w) = jwC_{gs} + jwC_{gd}(1 + g_{m1}/g_{m2})$$

Purely Capacitive!

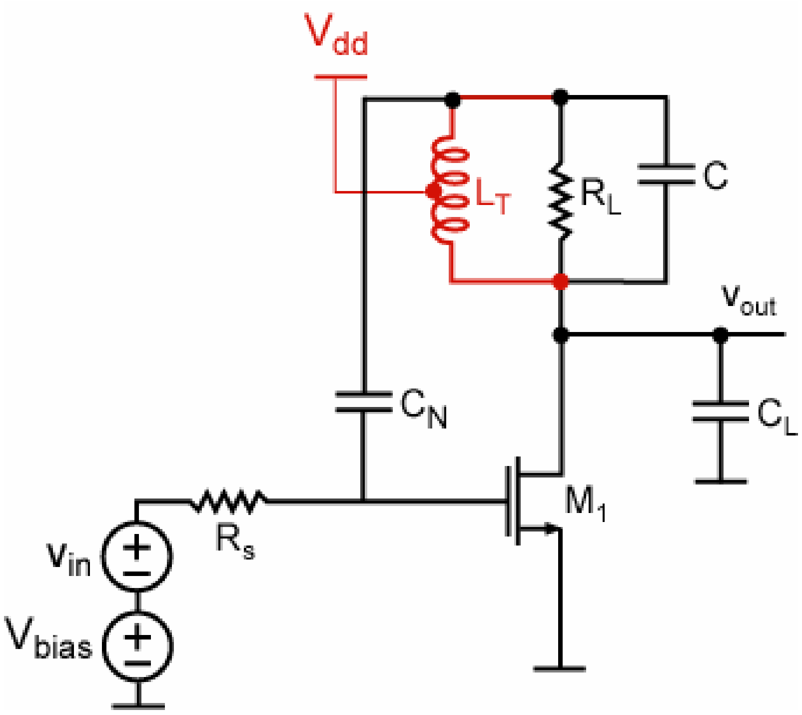
Neutralization in Tuned Amplifier

Recall the neutralization for broadband amplifier



For narrowband amplifier, the inverting signal can be generated by a tapped transformer

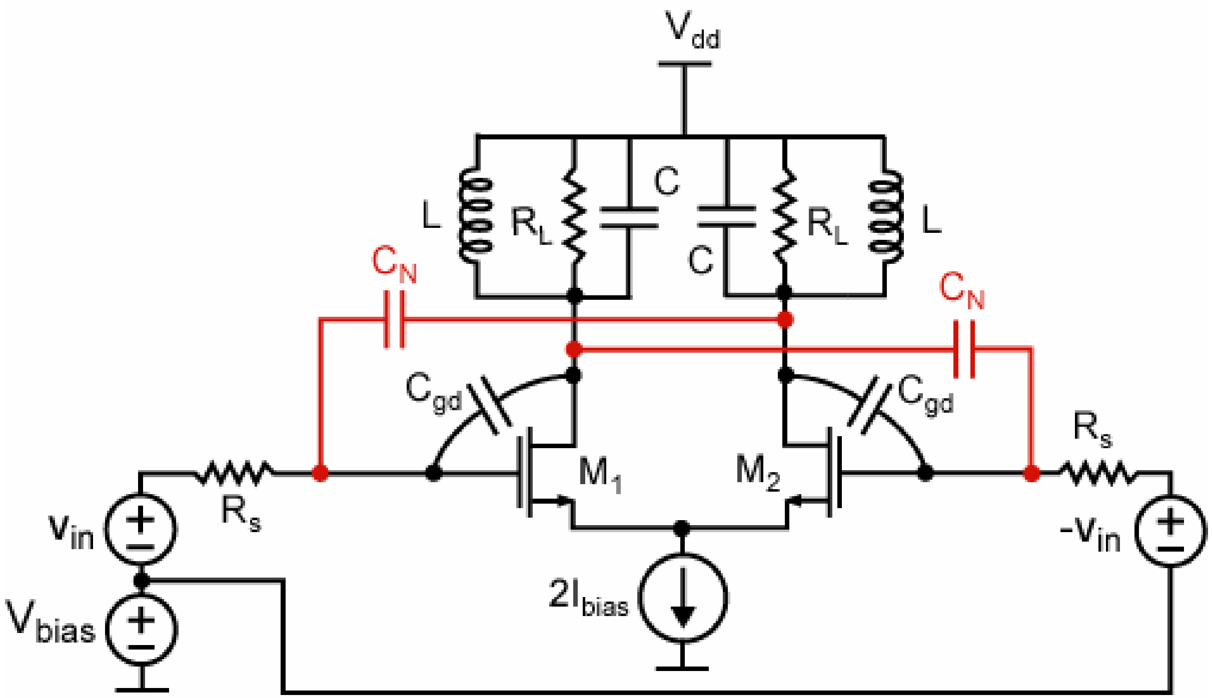
Neutralization with Tapped Transformer



Problems: Area and quality of on chip transformer

The neutralization cap C_N must be matched to C_{gd}

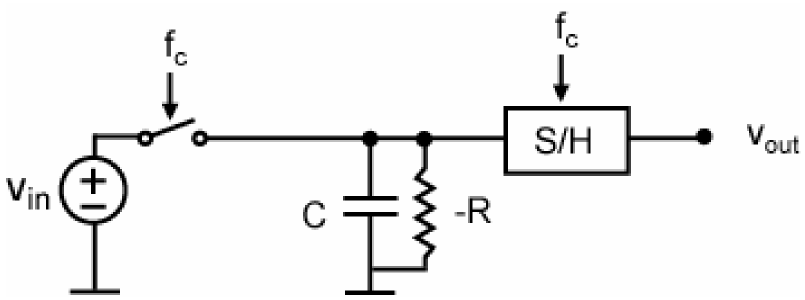
Differential Neutralization for Narrow Band Amplifier



- **Same principle as differential neutralization in broadband amp**
- **Only issue left is matching C_N to C_{gd}**
  - **Often use lateral metal caps for C_N (or CMOS transistor)**

Superregenerative Amplifier

quench



$$v_{out}(t) = v_{in}(0)e^{\frac{t}{RC}}$$

V_{in} is sampled at the rate f_c (in actual implementation f_c may be input level dependent)

The sampled output voltage

$$v_{out}(T) = v_{in}(0)e^{\frac{T}{RC}} \qquad T = \frac{1}{f_c}$$

Can be used for both broadband and narrowband – we'll do a simple analysis for broadband amp.

Superregenerative Amplifier

Gain calculation

$$v_{out}(T) = v_{in}(0)e^{\frac{T}{RC}}$$

$$A_v = \frac{v_{out}}{v_{in}} = e^{\frac{T}{RC}}$$

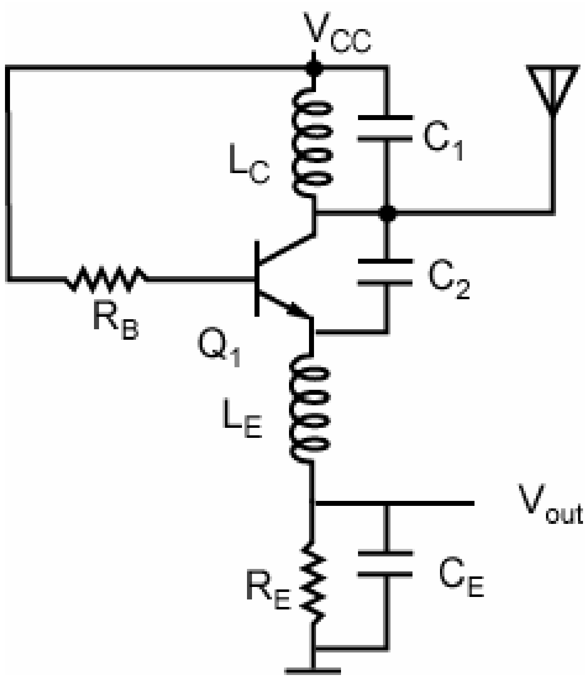
Nyquist theorem

$$BW < \frac{f_C}{2} = \frac{1}{2T}$$

$$BW|_{max} \bullet lnA_v = \frac{1}{2RC}$$

Trades bandwidth only logarithmically with gain!

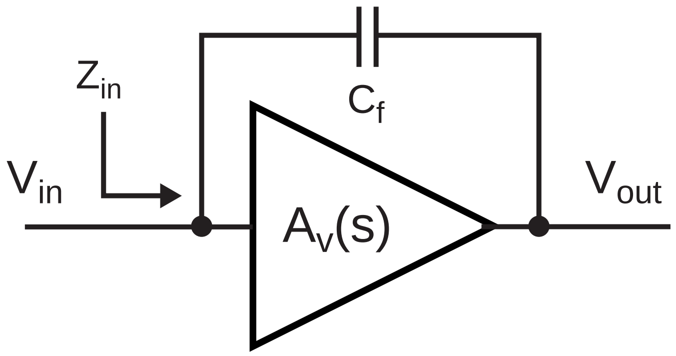
Superregenerative Amplifier Example (Narrowband)



R_B: DC bias resistor
L_C, C_1: Tuning LC
C_2 : positive feedback
L_E: RF choke (large inductance)

- **When the RF amplitude becomes large, it is rectified at the emitter of Q1**
- **This raises the DC potential at the emitter Q1 eventually turning it off**
- **The RF oscillation dies (quenched), and the DC potential at emitter of Q1 returns**
- **Amplitude of oscillation grows again due to positive feedback**

Active Real Impedance Generator



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

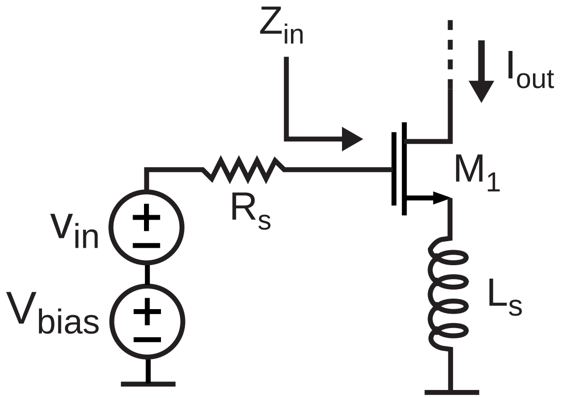
- **Input admittance:**

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

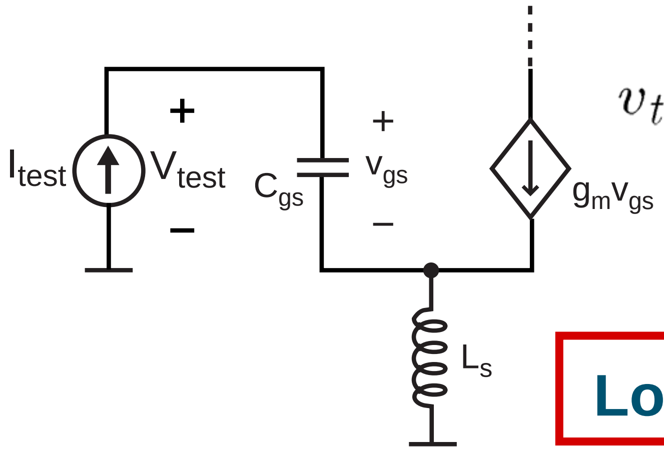
$$= jwC_f(1 + A_o \cos\Phi) + A_o w C_f \sin\Phi$$

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})**

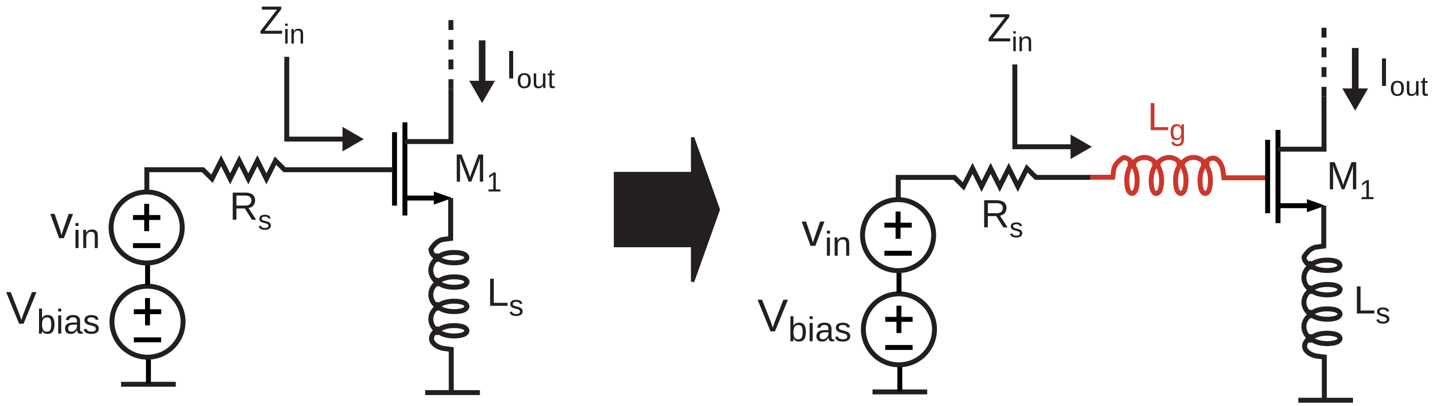


$$v_{test} = i_{test}\left(\frac{1}{sC_{gs}} + \left(1 + g_m\frac{1}{sC_{gs}}\right)sL_s\right)$$

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

Looks like series resonant circuit!

Use A Series Inductor to Tune Resonant Frequency



- **Calculate input impedance with added inductor (in order to choose resonance freq. and input resistance separately)**

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

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

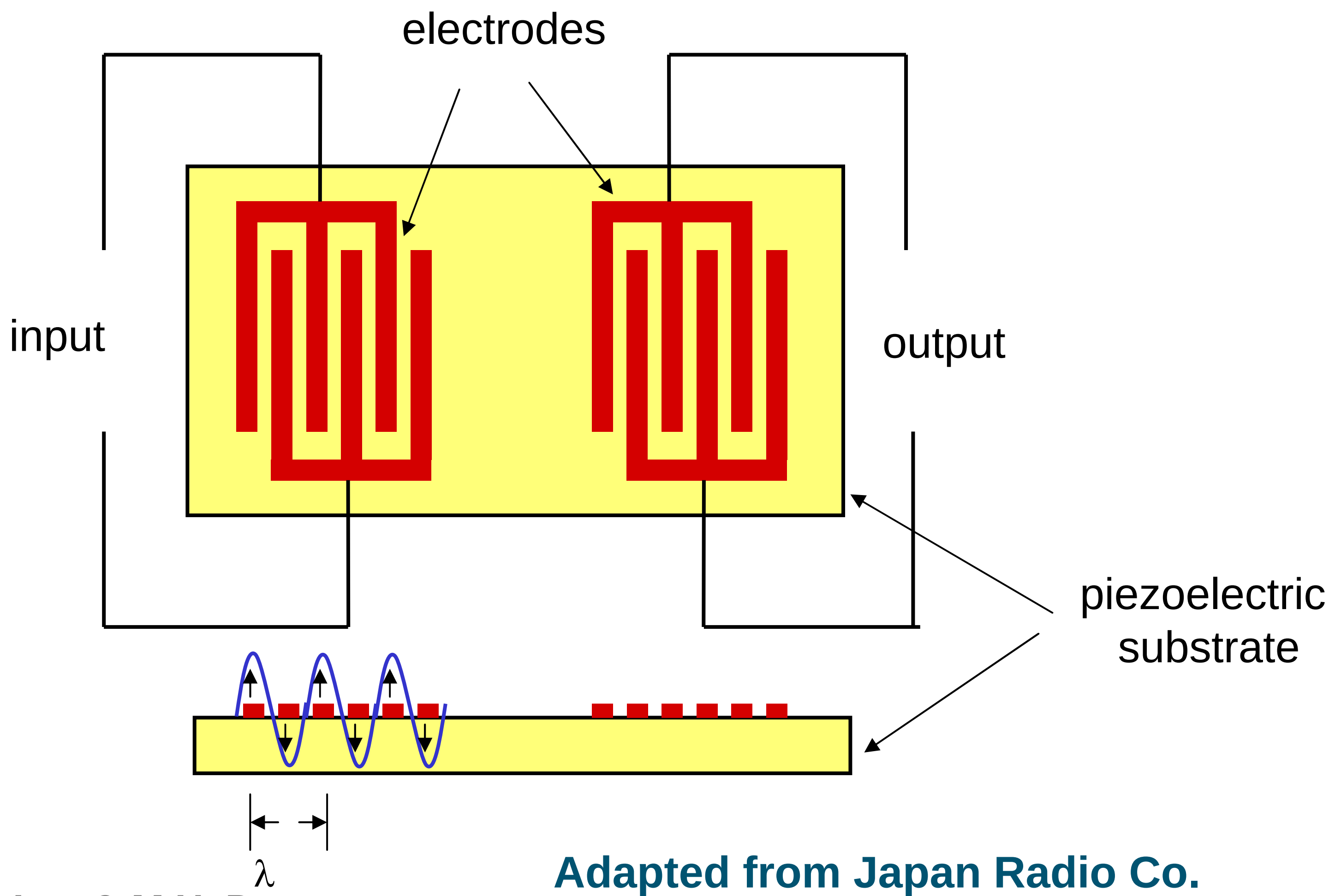
$$\text{i.e., want } \frac{1}{\sqrt{(L_s + L_g)C_{gs}}} = w_o$$

Narrowband Alternative to LC

- **On-chip inductors take up considerable die area and have relatively poor Q. Is there any other alternatives?**
  - **Use quarter-wavelength transmission line (waveguide) resonator?**
  - **In Lecture 5 we found the $\lambda/4$ waveguide with shorted load behave much like a parallel LC circuit, while with open load it behaves like a series LC**
  - **The problem is the dimension. For 900MHz mobile phone frequency, $\lambda/4$ in free space is 3.25 inches!**
  - **With high permittivity dielectric material (ceramic), the size can be reduced to a reasonable dimensions. With ε_r=10, the length of waveguide is only about inch.**
  - **Different configurations of filters can be built by combining sections of series and parallel LC equivalents**
  - **More appropriate at frequencies over GHz**
- **SAW (Surface Acoustic Wave) filters are another popular alternative**

SAW Filters

- **SAW filters use piezoelectric substrate to generate surface acoustic wave**



Adapted from Japan Radio Co.

SAW Filters

- Piezoelectric substrate
  - $LiTaO_3$
  - $LiNbO_3$
  - Quartz
- Filter Structures
  - Longitudinal Filter
  - Transversal Filter
  - Ladder Filter

- Saw filters have high selectivity and low insertion loss (down to a fraction of % fractional bandwidth, ~2dB insertion loss

- Wide range of enter frequency (few 100 kHz-GHz)

- At 1-2 GHz, the dimensions of SAW filters are 1-2mm
- For more information on SAW filters look over www.njr.com

SAW Filters in Mobile Phones

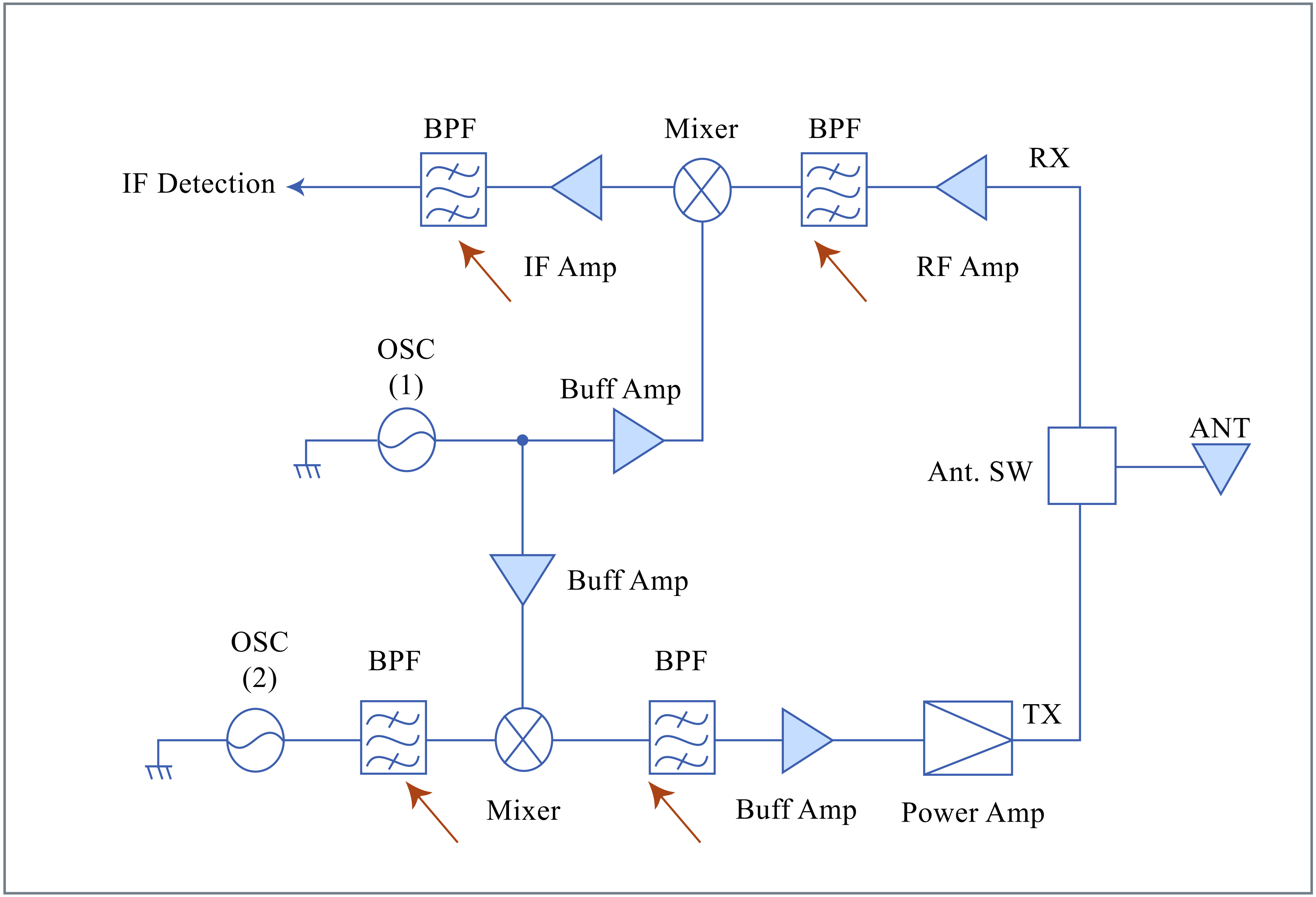


Figure by MIT OCW.

Adapted from Japan Radio Co.

SAW Filter Example

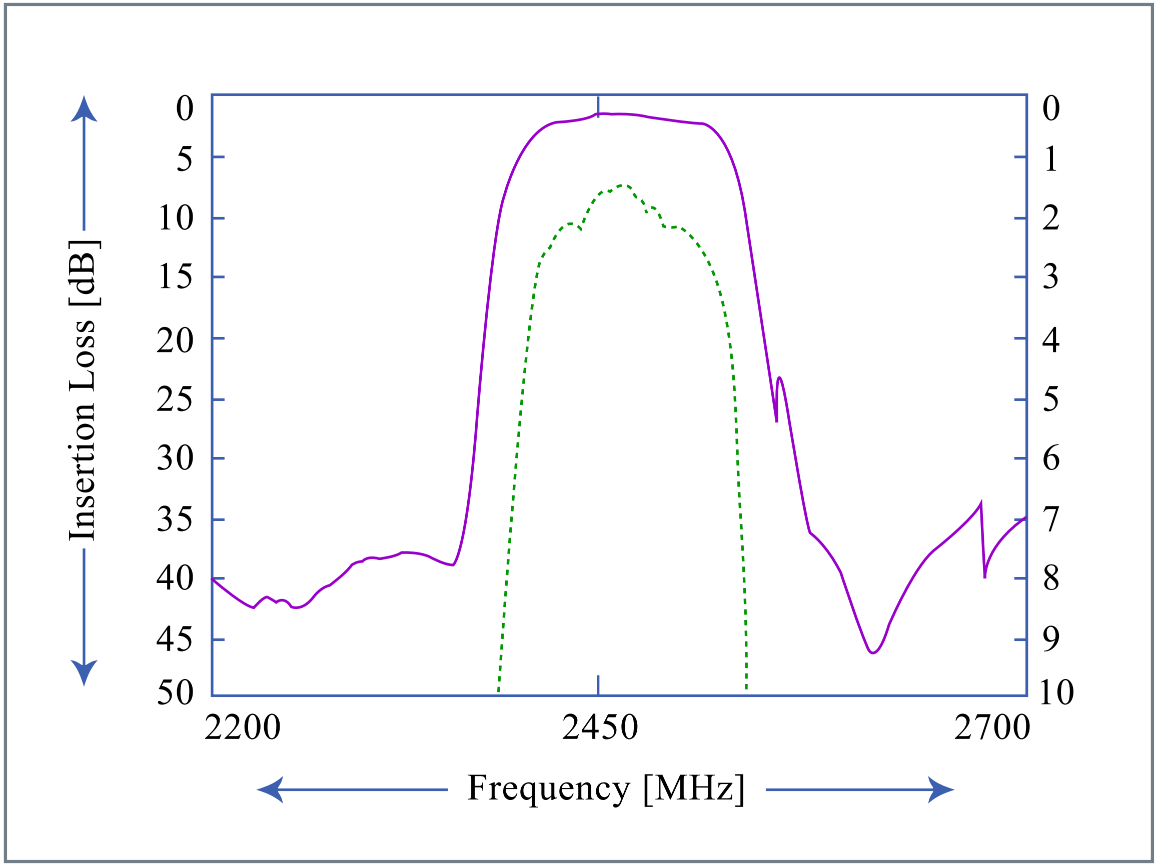


Figure by MIT OCW.

JRC NSVS754 2.4 GHz RF SAW Filter Charactersitic

Adapted from Japan Radio Co.