Presentation
Of
RF INTEGRATED CIRCUITS
(UNIT-II : REVIEW MOS DEVICE PHYSICS)

BY

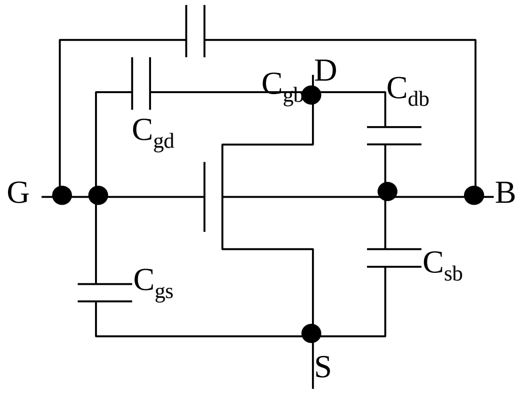
M.JYOTHIRMAI
ASSISTANT PROFESSOR
Department of ECE

RAVINDRA COLLEGE OF ENGINEERING FOR WOMEN
Kurnool-518452,Andhra Pradesh,INDIA

OUTLINE

- **1.MOS Capacitances**
- **2. Lossless Line**
- **3.Lossy Line**
- **4.HPA-Bandwidth using open circuit time constant**
- **5. HPA-Bandwidth using short circuit time constant**
- **6.Rise time and delay time**
- **7.Shunt Series Amplifiers**
- **8.Tuned Amplifiers**
- **9.Cascaded Amplifiers**

MOS Capacitances



- Masks result in some regions having overlaps, for example the gate electrode overlaps both the source and drain regions at the edges.
- Two overlap capacitances arise as a result.

- These are C_{gs} and C_{gd} respectively.
- If both the source and drain regions have the same width (W), the overlap capacitance becomes: $C_{gs}=C_{ox}WL_D$ and $C_{gd}=C_{ox}WL_D$.
- These overlap capacitances are voltage dependent.
- C_{gs}, C_{gd} and C_{gb} are voltage dependent and distributed
- They result from the interaction between the gate voltage and the channel charge.

Why this lecture is important.

We will use MOSFETs to design our circuits.

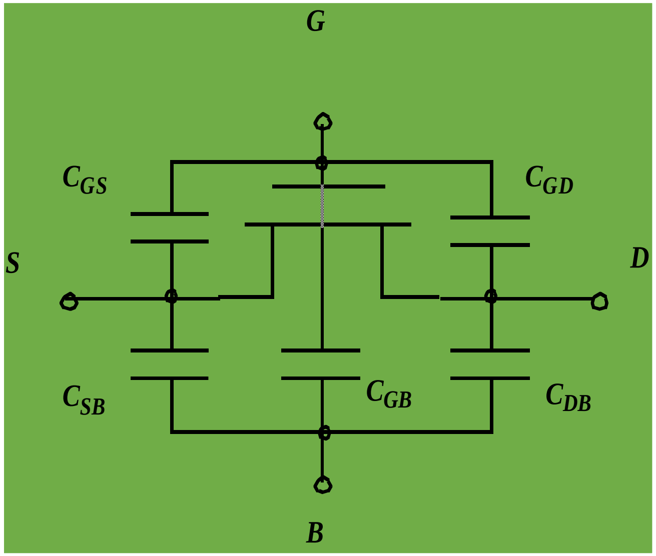
MOSFET capacitances tend to limit the frequency response of circuits.

In order to predict the circuit frequency response, we need to estimate the circuit capacitance.

We may use the MOSFET capacitance to our advantage, by intentionally implementing capacitors using MOSFETs.

Dynamic Behavior of MOS Transistor

- MOSFET is a majority carrier device (unlike *pn* junction diode)
- Delays depend on the time to (dis)charge the capacitances between MOS terminals
- Capacitances originate from three sources:
  - basic MOS *structure* (layout)
  - charge present in the *channel*
  - *depletion regions* of the reverse-biased *pn*-junctions of drain and source
- Capacitances are *non-linear* and vary with the applied voltage



MOS Structure Capacitances

Gate Capacitance

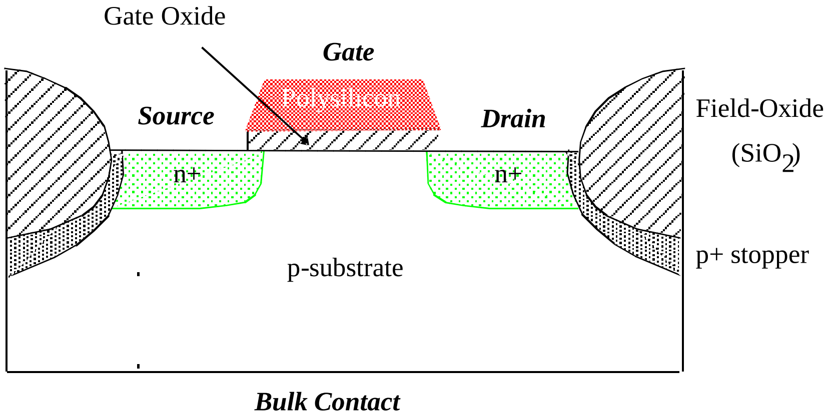
- Gate isolated from channel by gate oxide

$$C_{ox} = \varepsilon_{ox} / t_{ox}$$

- t_{ox} is very small <10nm

- Results in *gate capacitance C_g*

$$C_g = C_{ox} W L$$



CROSS-SECTION of NMOS Transistor

The Gate Capacitance

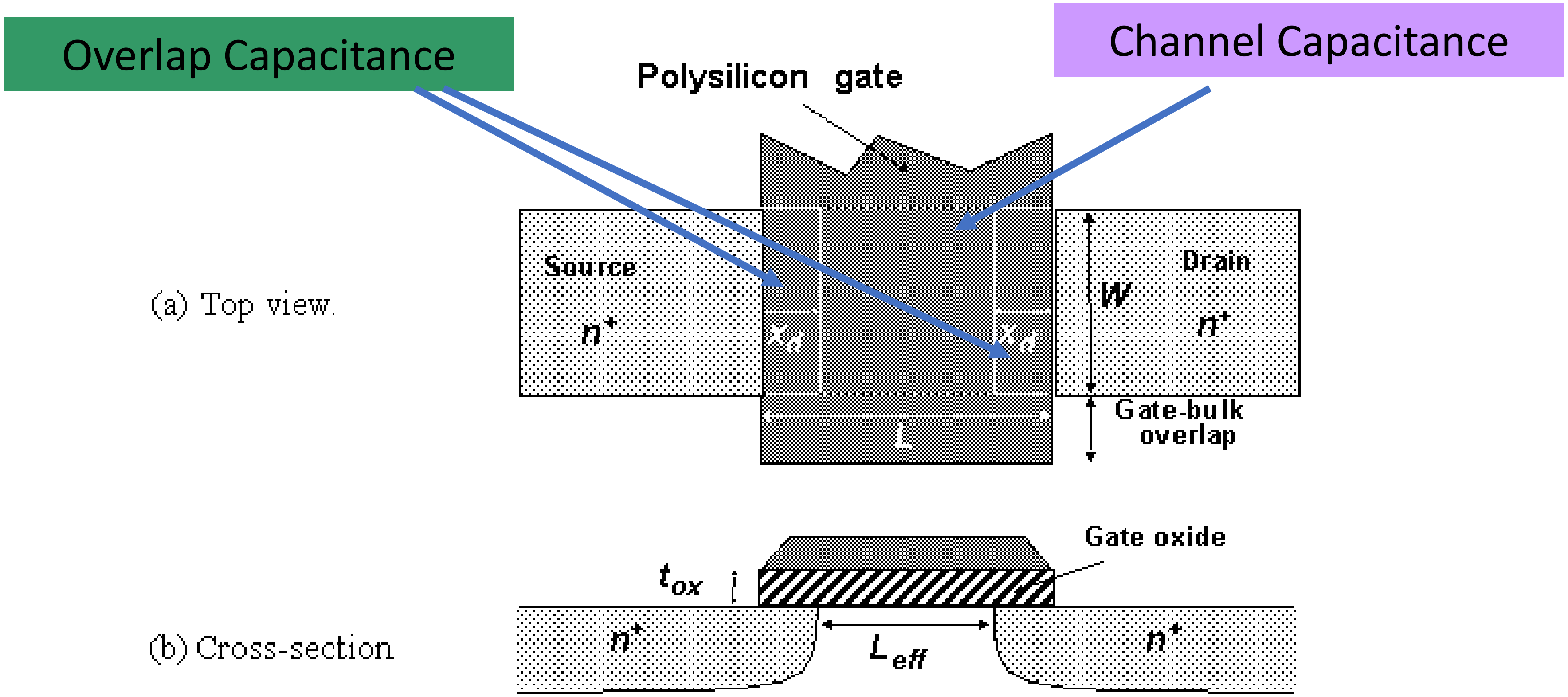
Gate Capacitance depends on

- channel charge (non-linear)
- topology

Capacitance due to topology

- Source and drain extend below the gate oxide by x_d (*lateral diffusion*)
- Effective length of the channel L_{eff} is shorter than the drawn length by factor of $2x_d$
- Cause of parasitic overlap capacitance, C_{gsO}, between gate and source (drain)

The Gate Capacitance



$$C_{gate} = \frac{\varepsilon_{ox}}{t_{ox}} WL$$

The Channel Capacitance

Channel Capacitance has three components

- capacitance between gate and source, C_{gs}
- capacitance between gate and drain, C_{gd}
- capacitance between gate and bulk region, C_{gb}

Channel Capacitance values

- non-linear, depends on operating region
- averaged to simplify analysis

Capacitive Device Model

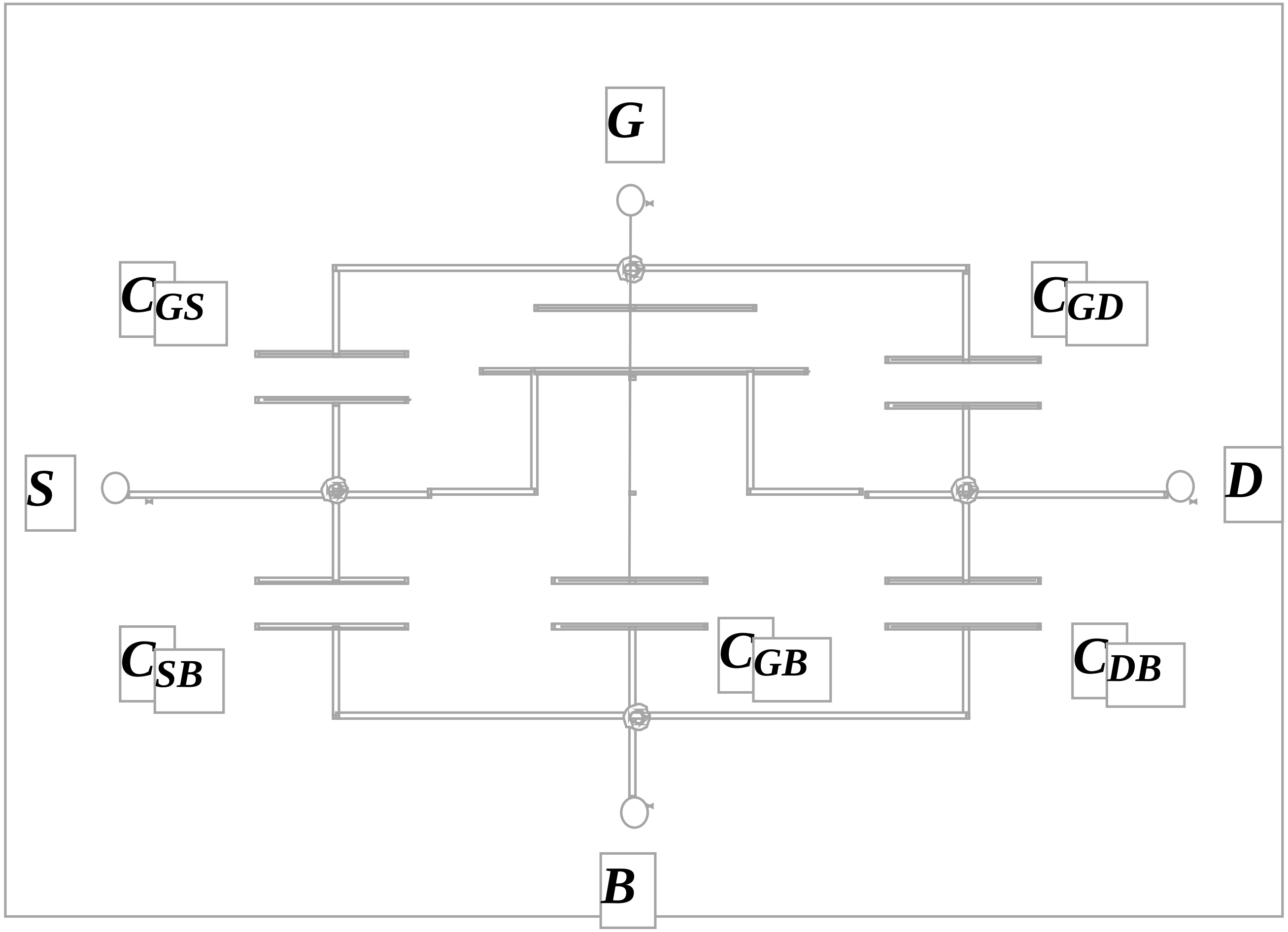
$C_{GS} = C_{gs} + C_{gsO}$

$C_{GD} = C_{gd} + C_{gdO}$

$C_{GB} = C_{gb}$

$C_{SB} = C_{Sdiff}$

$C_{DB} = C_{Ddiff}$



Transmission lines

1.Lossless line

$$\gamma = \alpha + j\beta = j\omega\sqrt{LC} \qquad \beta = \omega\sqrt{LC} \qquad \alpha = 0$$

$$Z_0 = \sqrt{\frac{L}{C}} \qquad V(z) = V_0^+ e^{-i\beta z} + V_0^- e^{i\beta z}$$

$$I(z) = \frac{V_0^+}{Z_0} e^{-j\beta z} - \frac{V_0^-}{Z_0} e^{j\beta z} \qquad \lambda = \frac{2\pi}{\beta} = \frac{2\pi}{\omega\sqrt{LC}}$$

$$v_p = \frac{\omega}{\beta} = \frac{\omega}{\omega\sqrt{LC}} = \frac{1}{\sqrt{LC}}$$

2.Lossy line

- One type of metal loss is I^2R *loss*
- In transmission lines,
  - the resistance of the conductors is never equal to zero
    - except for superconductors
- Whenever current flows through one of these conductors,
  - some energy is dissipated in the form of heat

Lossy line ideal characteristics

$$\gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} = \sqrt{(\mathrm{j}\omega\mathrm{L})(\mathrm{j}\omega\mathrm{C})\left((\frac{\mathrm{R}}{\mathrm{j}\omega\mathrm{L}} + 1)(\frac{\mathrm{G}}{\mathrm{j}\omega\mathrm{C}} + 1) \right)}$$

$$= \mathrm{j}\omega\sqrt{\mathrm{LC}}\sqrt{1 - \left[\left(\frac{\mathrm{R}}{\omega\mathrm{L}} + \frac{\mathrm{G}}{\omega\mathrm{C}} \right)\mathrm{j} - \frac{\mathrm{RG}}{\omega^2\mathrm{LC}} \right]}$$

Low loss case

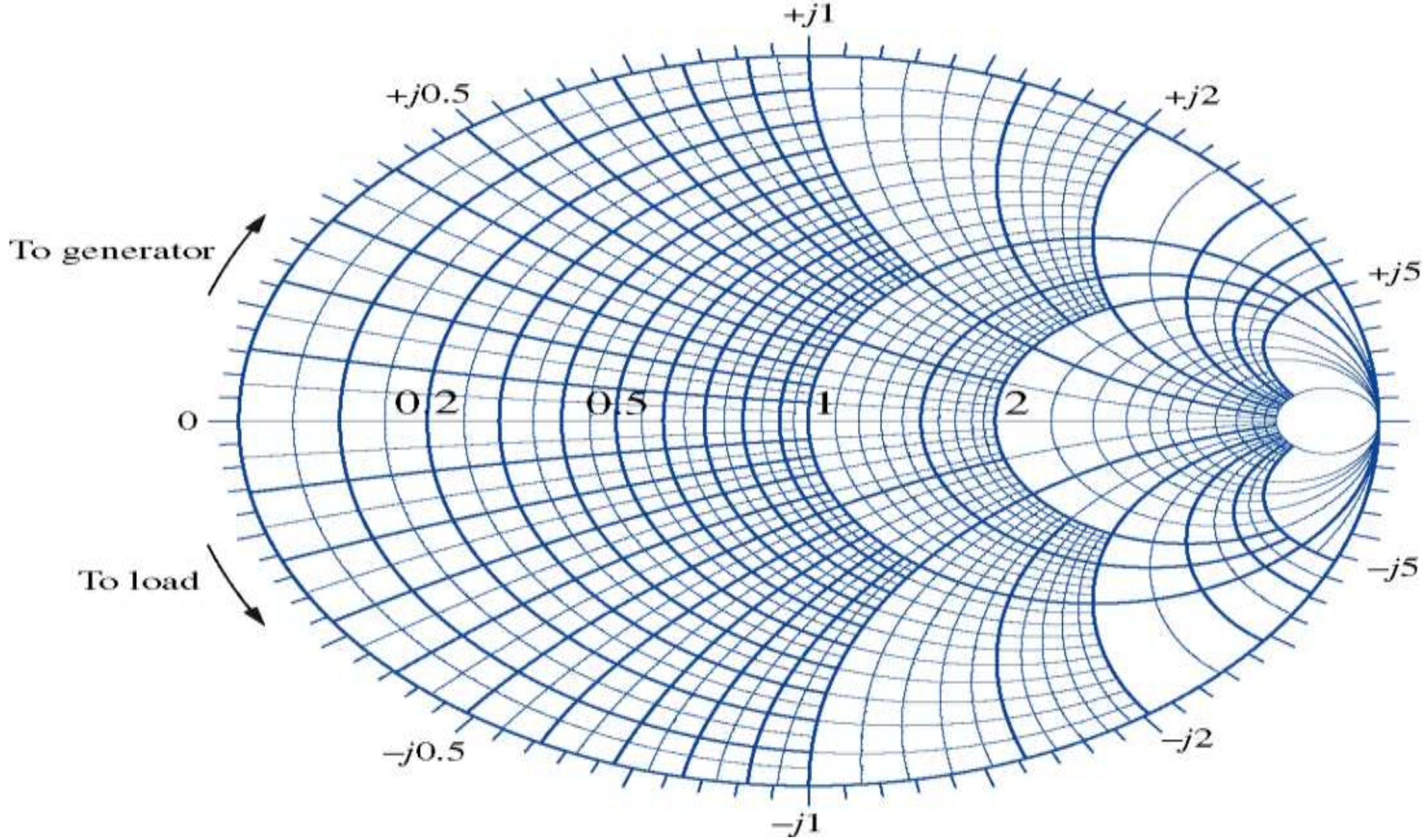
$$\mathrm{R} << \omega\mathrm{L}, \mathrm{G} << \omega\mathrm{C} \quad \mathrm{RG} << \omega^2\mathrm{LC} \quad \gamma \cong j\omega\sqrt{LC}\left[1 - \frac{1}{2} j \left(\frac{R}{\omega L} + \frac{G}{\omega C} \right) \right]$$

$$\therefore \alpha \cong \frac{1}{2}\left(R\sqrt{\frac{C}{L}} + G\sqrt{\frac{L}{C}} \right) = \frac{1}{2}\left(\frac{R}{Z_0} + GZ_0 \right) \quad \beta \cong \omega\sqrt{\mathrm{LC}} \quad Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \cong \sqrt{\frac{L}{C}}$$

- Q is a measure of loss of a resonant circuit,
  - lower loss implies higher Q and
  - high Q implies narrower bandwidth
- As R increases,
  - power loss increases and
  - quality factor decreases
- Let us see what the approximate Z_{in} near resonance
- The input impedance can be rewritten in the following form:

$$Z_{in} = R + j\omega L\left(1 - \frac{1}{\omega^2 LC}\right) = R + j\omega L\left(1 - \frac{{\omega_0}^2}{\omega^2}\right) = R + j\omega L\left(\frac{\omega^2 - {\omega_0}^2}{\omega^2}\right)$$

Smith chart



Smith chart (Example)

Example 9.9: On the simplified Smith chart, locate the following normalized impedances:

$a)\ z = 1 + j0;$

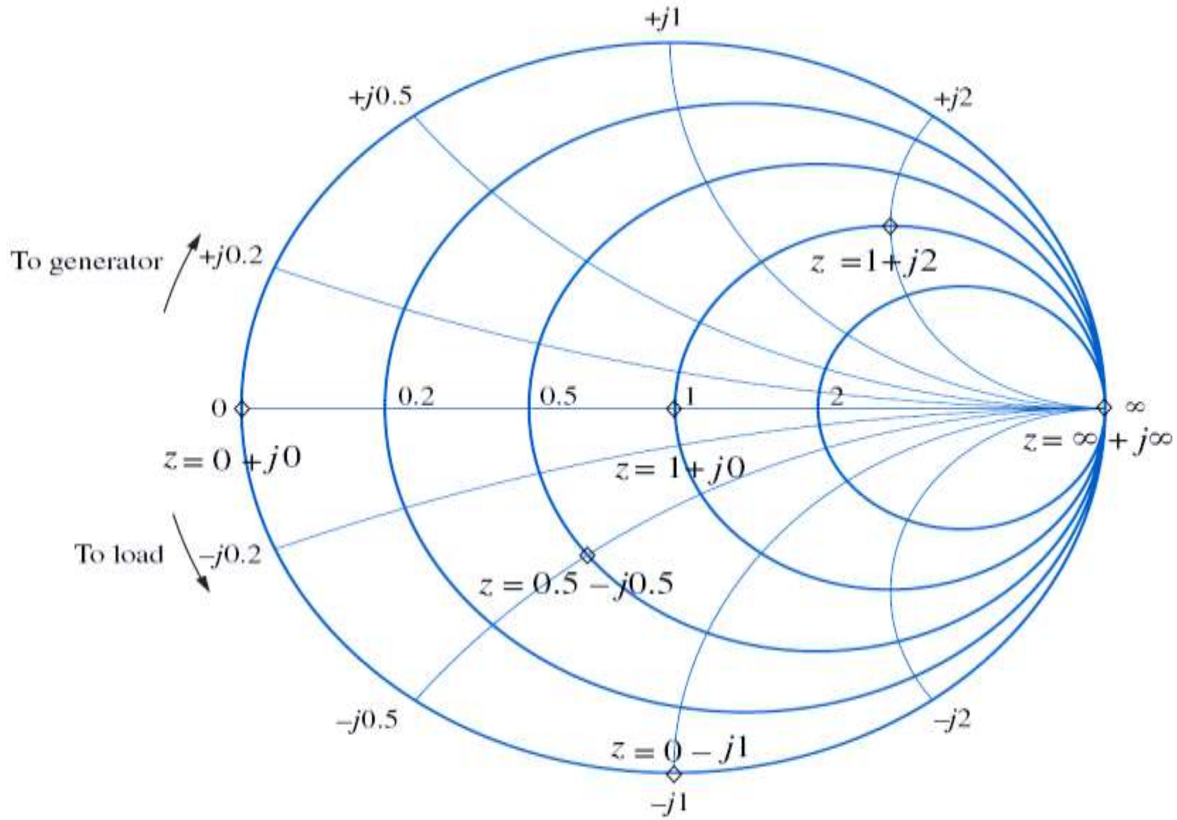
$b)\ z = 0.5 - j0.5$

$c)\ z = 0 + j0;$

$d)\ z = 0 - j1;$

$e)\ z = 1 + j2;$

$f)\ z = \infty$



To find reflection coefficient

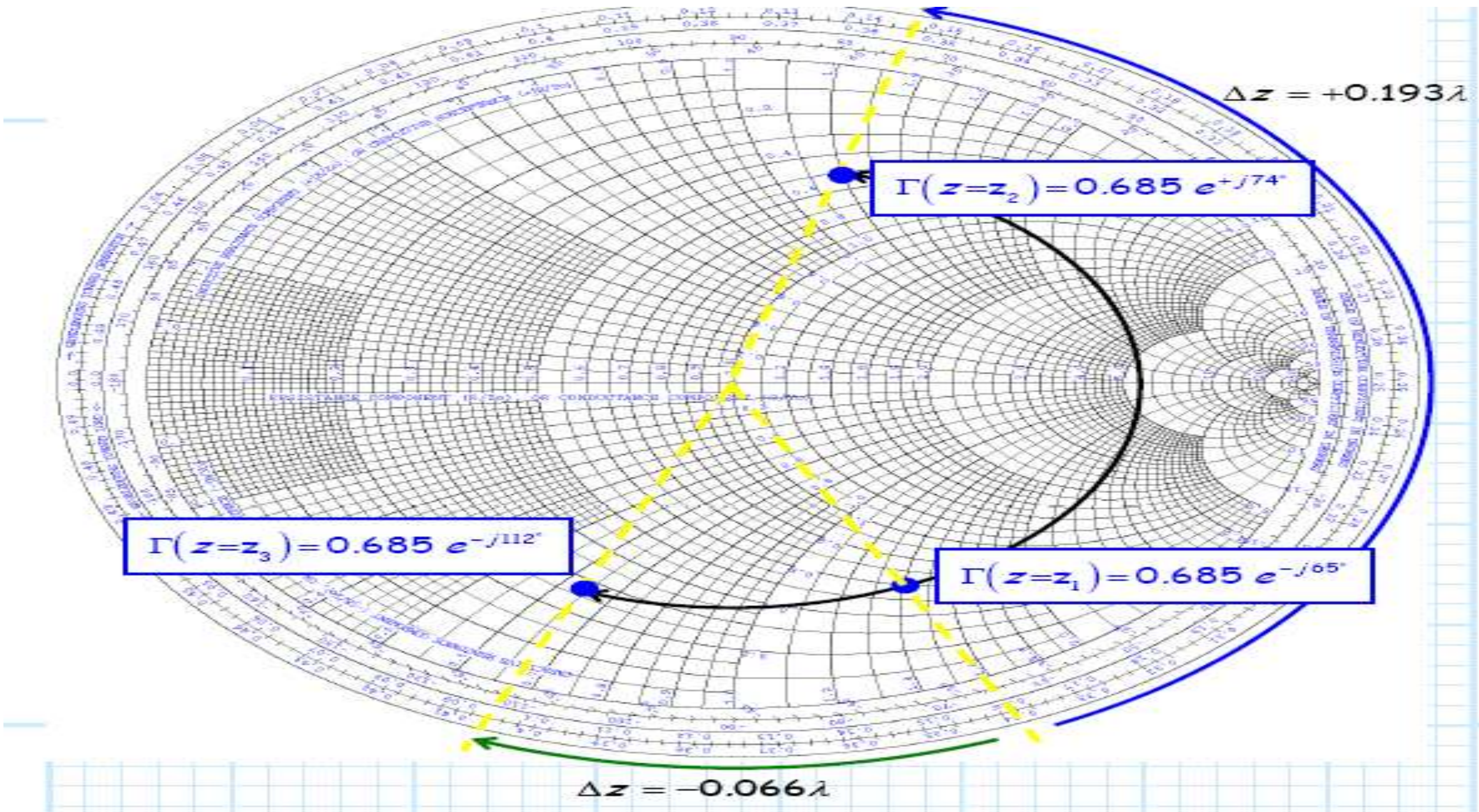
1.Get the normalized impedance

2. Plot normalized impedance on sm chart

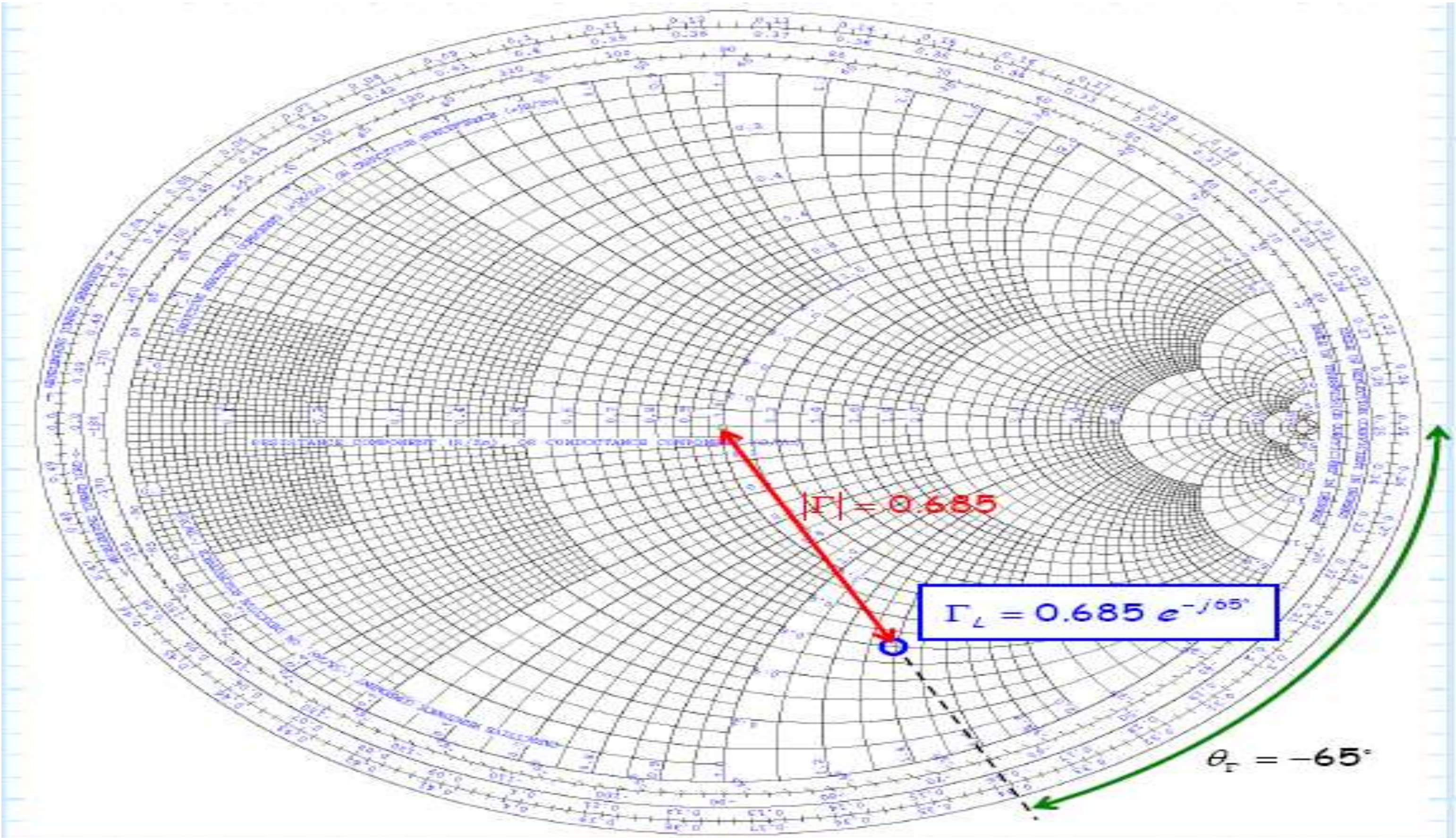
2.Draw the VSWR circle

3.Measure the distance from the center of the smith chart to the circle on the reflection coefficient scale

4.Draw the straight line through center of the sm chart to normalized impedance we can get the angle

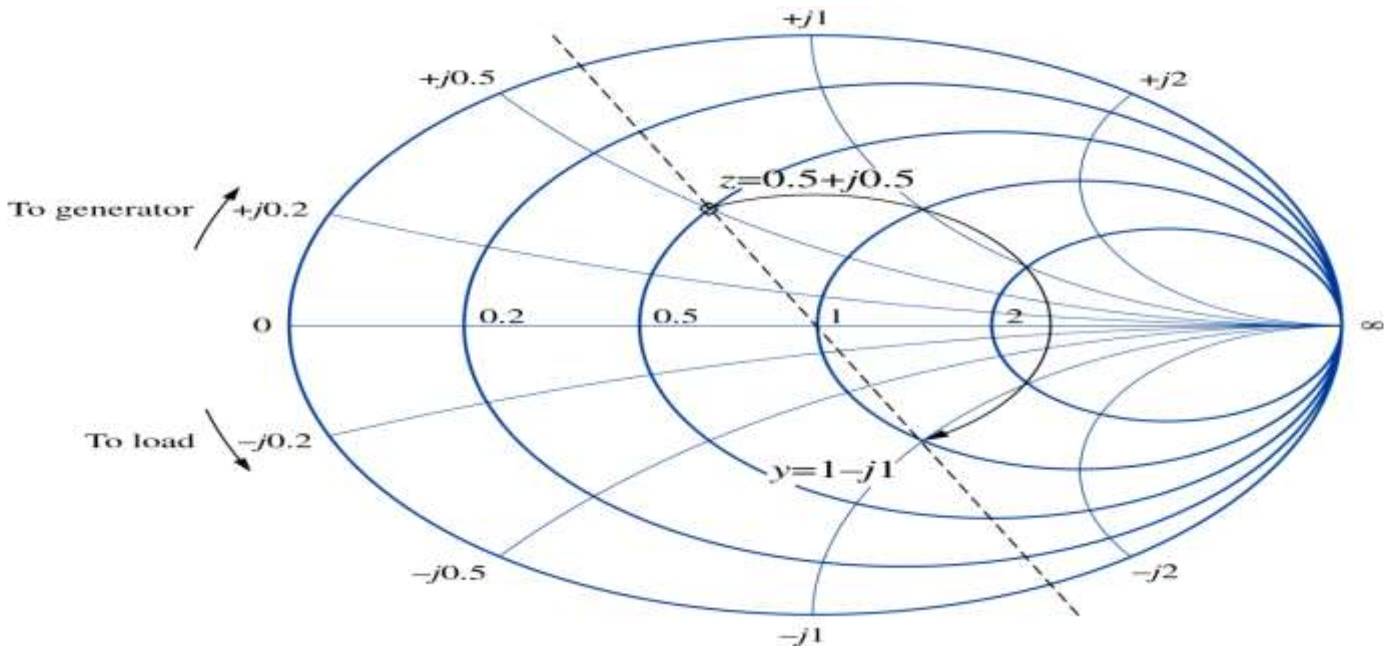


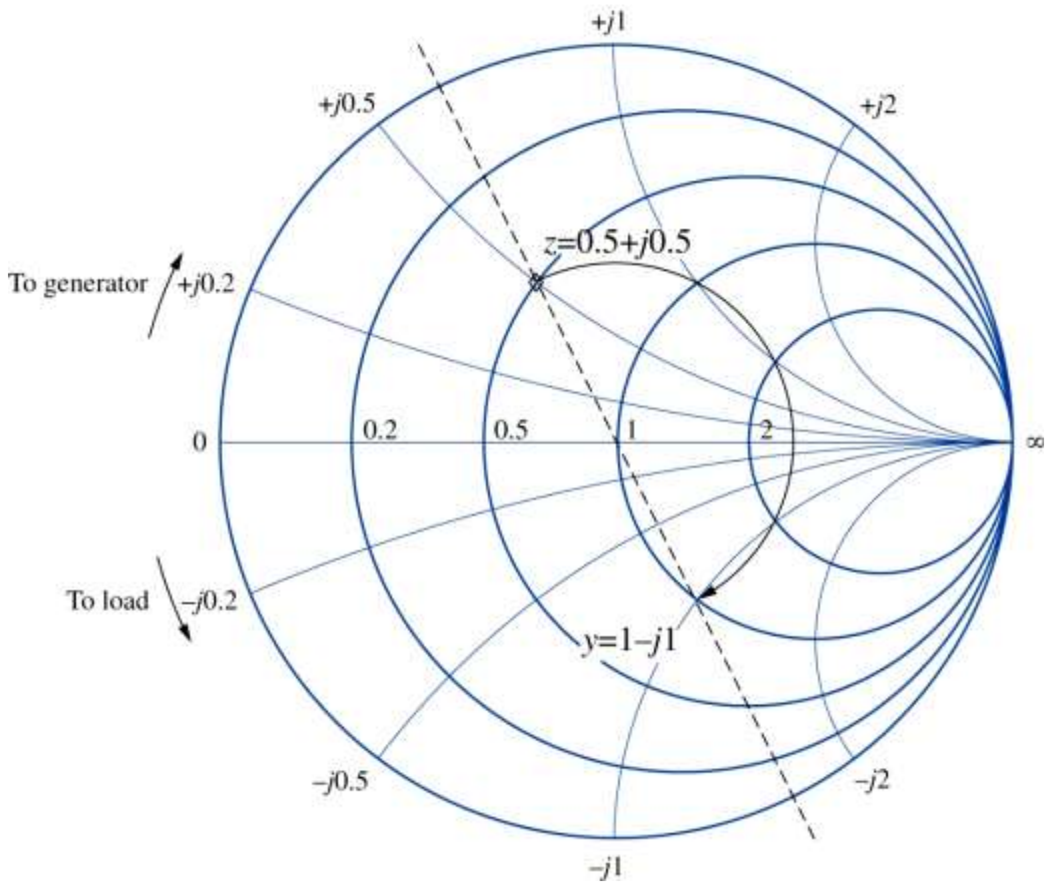
$\Delta z = +0.193\lambda$
$\Gamma(z=z_2)=0.685\ e^{+j74^\circ}$
$\Gamma(z=z_3)=0.685\ e^{-j112^\circ}$
$\Gamma(z=z_1)=0.685\ e^{-j65^\circ}$
$\Delta z = -0.066\lambda$


$|\Gamma| = 0.685$
$\Gamma_L = 0.685\, e^{-j65^\circ}$
$\theta_\Gamma = -65^\circ$

For example, the impedance
$z = 0.5 + j\,0.5$
corresponds to the admittance

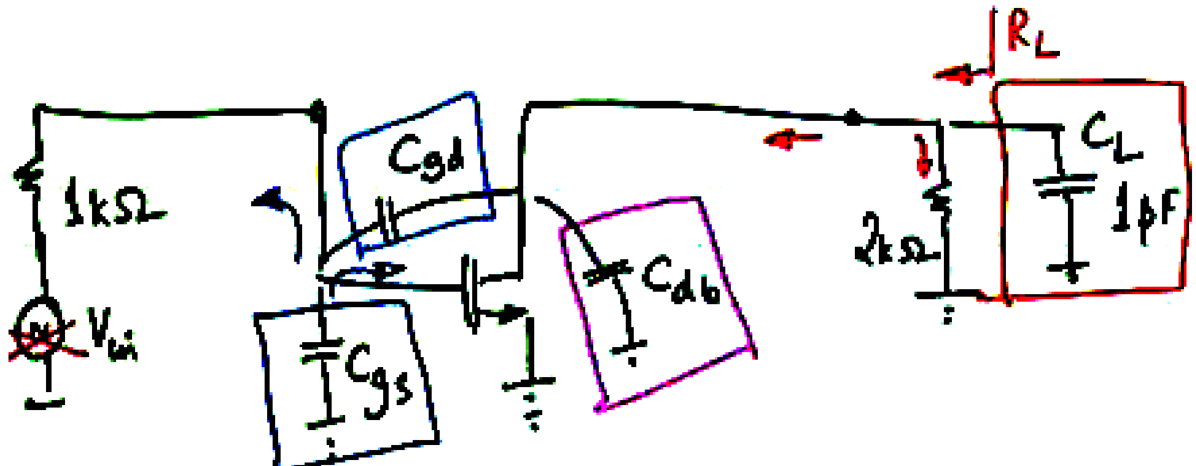
$$y = \frac{1}{0.5 + j0.5} = 1 - j1$$





+j1
+j0.5
+j2
z=0.5+j0.5
To generator
+j0.2
0
0.2
0.5
1
2
∞
To load
−j0.2
y=1−j1
−j0.5
−j2
−j1

HPA-BW USING OPEN CKT TIME CONSTANT



$R_L = 1k\Omega \quad C_L = 1pF \longrightarrow \tau_L = 1000$ psec

$R_{db} = 1k\Omega \quad C_{db} = 100fF \longrightarrow \tau_{db} = 100$ psec

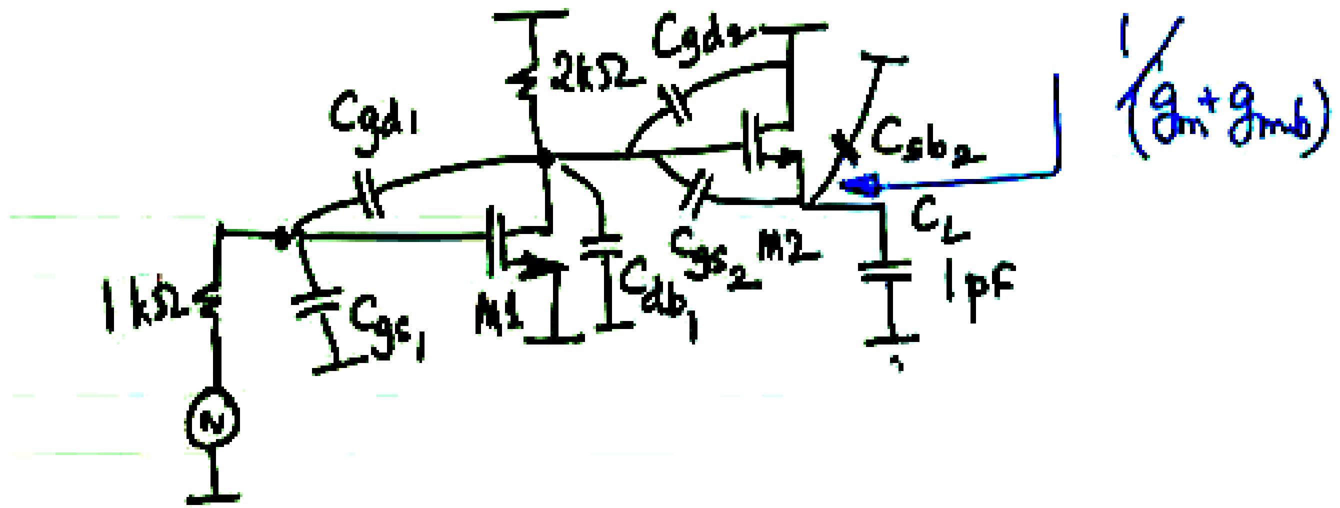
$R_{gs} = 1k\Omega \quad C_{gs} = 200fF \longrightarrow \tau_{gs} = 200$ psec

$R_{gd} : 12k\Omega \quad C_{gd} = 50fF \longrightarrow \tau_{gd} = 600$ psec

$\tau = 1900$ psec

$f_{-3dB} = 84$ MHz

HPA-BW USING OPEN CKT TIME CONSTANT



$C_{gs_1} \rightarrow \tau_{gs_1} = 200$ psec

$C_{gd_1} \rightarrow \tau_{gd_1} = 600$ psec

$C_{db_1} \rightarrow \tau_{db_1} = 100$ psec

$C_{gd_2} \rightarrow R_{gd_2} = 1k\Omega \rightarrow \tau_{gd_2} = 50$ psec

$C_{sb_2} \rightarrow R_{sb_2} = 90\Omega \rightarrow \tau_{sb_2} = 14$ psec

$C_L \rightarrow R_L = 90\Omega \rightarrow \tau_L = 90$ psec

$C_{gs_2} \rightarrow R_{gs} = 174\,\Omega \rightarrow \tau_{gs} = 35$ [illegible]

$\tau_{sb_2} = 87\Omega \times 150fF = 13\,ps$

$\frac{2k\Omega}{1 + 11mS \cdot 2k\Omega}$

$\tau_l = 87\Omega \times 1pF = 87\,ps$

$\tau_{ga} \quad 1.92k\Omega \times 50fF = 96\,ps$

$2k\Omega \parallel (11mS \cdot 2k\Omega \cdot 2k\Omega + 2k\Omega + 2k\Omega)$

$\tau_{gs_2} = 2.2k\Omega \times 200fF = 430\,ps$

$\tau_{db_3} = 1.92k\Omega \times 100fF = 192\,ps$

$\tau = 1332\,psec$

$\tau_{sb_3} = 160\Omega \times 150fF = 24\,ps$

751 Mrad/sec

$\tau_{gs_3} = 160\Omega \times 200fF = 32\,ps$

119 MHz

$\tau_{gd} = 1.92k\Omega \times 50fF = 96\,ps$

SHORT CKT TIME CONSTANT

$$\frac{V_o(s)}{V_i(s)} = \frac{ks^n}{(s+s_1)(s+s_2)\ldots(s+s_n)}$$

$$s^n + b_1 s^{n-1} + \ldots + b_{n-1} s + b_n$$

$$\frac{V_o(s)}{V_i(s)} \approx \frac{ks^n}{s^n + b_1 s^{n-1}} = \frac{ks}{s + \sum_{i=1}^{n} s_i}$$

SHORT CKT TIME CONSTANT

$$\omega_l \approx b_1 = \sum_{i=1}^{n} s_i = \omega_{l,\, est}$$

Now, at our estimated –3 dB frequency, the original denominator polynomial is:

$$-\omega_{l,\, est}^2 + j\omega_{l,\, est}(s_1 + s_2) + s_1 s_2$$

Substituting our expression for the estimated –3 dB point, we obtain:

$$-[s_1^2 + s_2^2 + 2s_1 s_2] + j[s_1^2 + s_2^2 + 2s_1 s_2] + s_1 s_2$$

SHORT CKT TIME CONSTANT

1) Compute the effective resistance R_{js} facing each jth capacitor with all of the other capacitors *short*-circuited (the subscript "*s*" refers to the short-circuit condition for each capacitor);

2) Compute the "short-circuit frequency" $1/(R_{js}C_j)$;

3) Sum all m such short-circuit frequencies.

The sum of the reciprocal short-circuit time constants formed in step 3) turns out to be precisely equal to the sum of the pole frequencies, b_1. Thus, at last, we have:

$$\omega_{l,\,est} = \sum_{j=1}^{m} \frac{1}{R_{js}C_j}$$

Delay time in cascaded system

$$T_D \equiv \frac{\int_{-\infty}^{\infty} t\,h(t)\,dt}{\int_{-\infty}^{\infty} h(t)\,dt}$$

Rise time

$$\left(\frac{\Delta T}{2}\right)^2 = \left[\frac{\int_{-\infty}^{\infty} t^2 h(t)\,dt}{\int_{-\infty}^{\infty} h(t)\,dt} - (T_D)^2\right]$$

Again this definition allows the use of Fourier transform identities. In particular

$$\int_{-\infty}^{\infty} t^2 h(t)\,dt = -\frac{1}{(2\pi)^2}\frac{d^2}{df^2}H(f)\Big|_{f=0}$$

so that

$$t_{rise}^2 = (\Delta T)^2 = 4\left[\frac{-\frac{1}{(2\pi)^2}\frac{d^2}{df^2}H(f)\Big|_{f=0}}{H(0)} - \left(-\frac{1}{j2\pi H(0)}\frac{d}{df}H(f)\Big|_{f=0}\right)^2\right]$$

Rise time

$$t_{rise}^2 = \frac{4}{(2\pi)^2 H(0)}\left[-\frac{d^2}{df^2}H(f)\bigg|_{f=0} - \frac{1}{H(0)}\left(\frac{d}{df}H(f)\bigg|_{f=0}\right)^2\right]$$

The Elmore risetime for a single-pole low-pass system is 2τ.

Proceeding as in Section5.2, now consider two systems, each with its own risetime. Then

$$t_{rise(tot)}^2 = \frac{4}{(2\pi)^2 H_1(0)H_2(0)}\left[-\frac{d^2}{df^2}H_1H_2\Big|_{f=0} - \frac{1}{H_1(0)H_2(0)}\left(\frac{d}{df}H_1H_2\Big|_{f=0}\right)^2\right]$$

which, after a small algebraic miracle, leads to the desired result at last:

$$t_{rise(tot)}^2 = t_{rise1}^2 + t_{rise2}^2$$

Thus we see that the *squares* of the individual risetimes add linearly to yield the square of the overall risetime. Stated alternatively, the individual risetimes add in root-sum-squared (RSS) fashion to yield the overall risetime:

$$t_{rise(tot)} = \sqrt{t_{rise1}^2 + t_{rise2}^2}$$

Bandwidth

$$BW \approx \frac{1}{\sqrt{\sum_{1}^{N} \tau^2}} = \frac{1}{\tau\sqrt{N}}$$

Compare that approximate result with the more exact (but still approximate) relationship (see the chapter on high-frequency amplifier design):

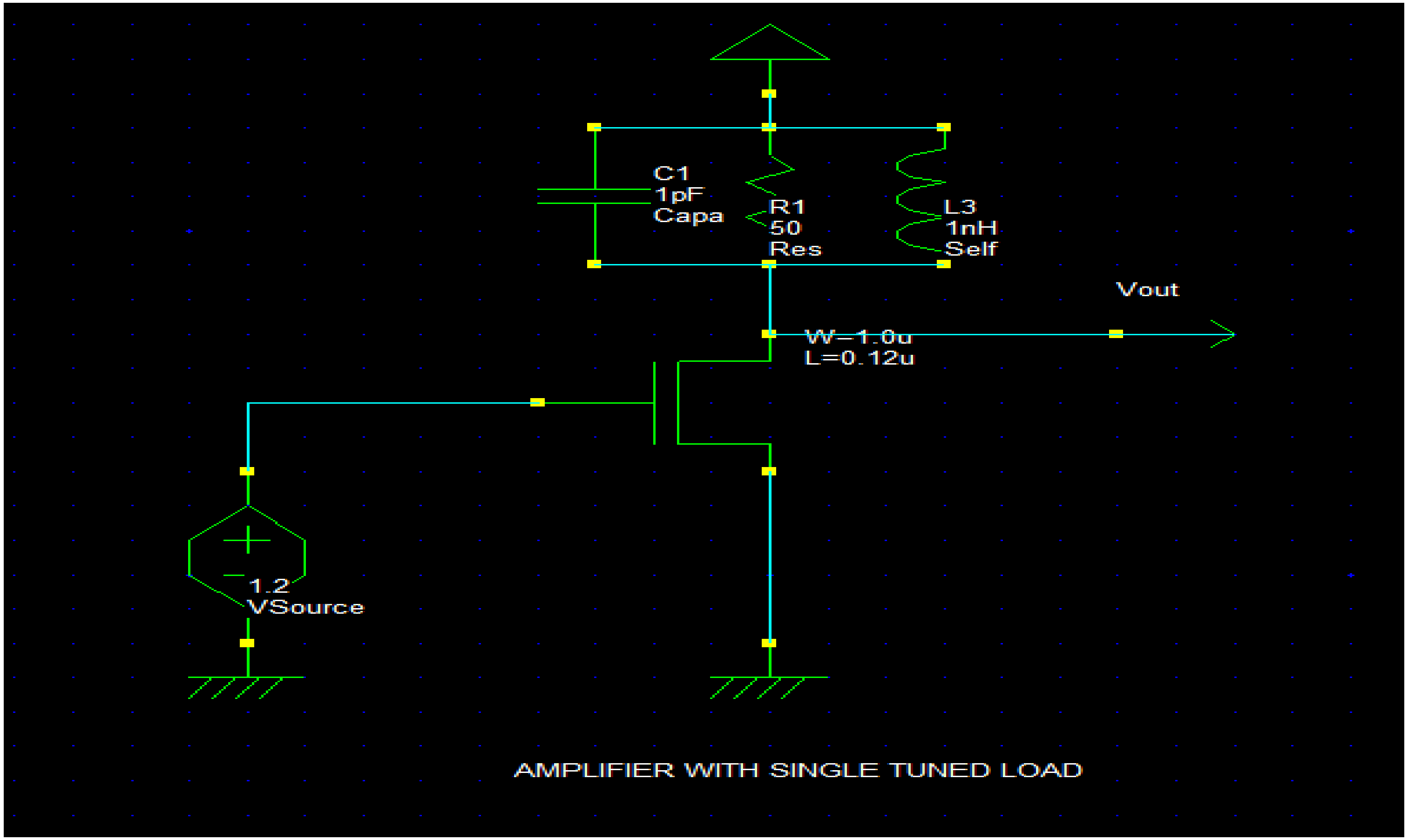
$$BW \approx \frac{\sqrt{ln\,2}}{\tau\sqrt{N}} \approx \frac{0.833}{\tau\sqrt{N}}$$

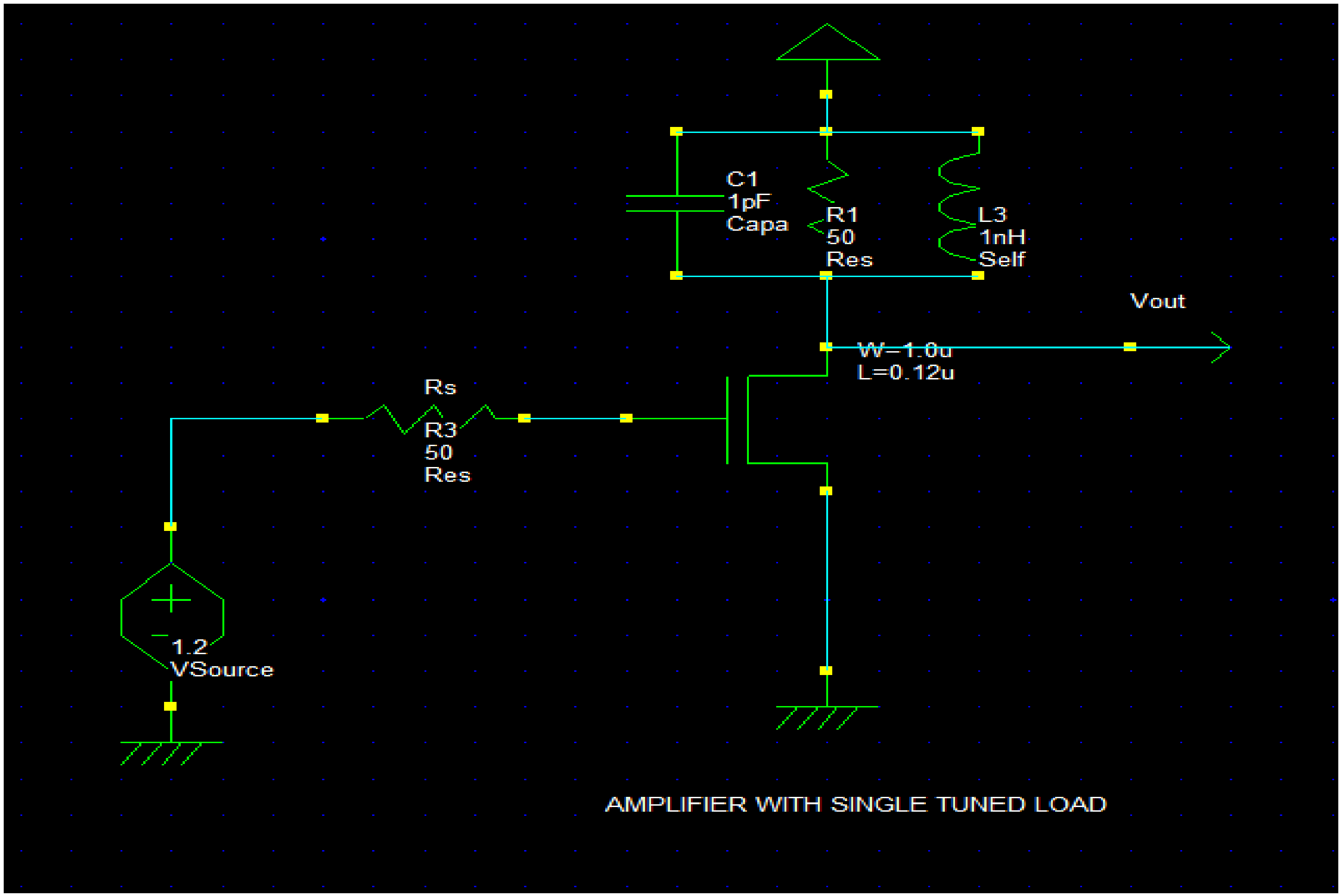
As can be seen, the functional dependence on N is the same; the equations differ only by a relatively small multiplicative factor.[5]

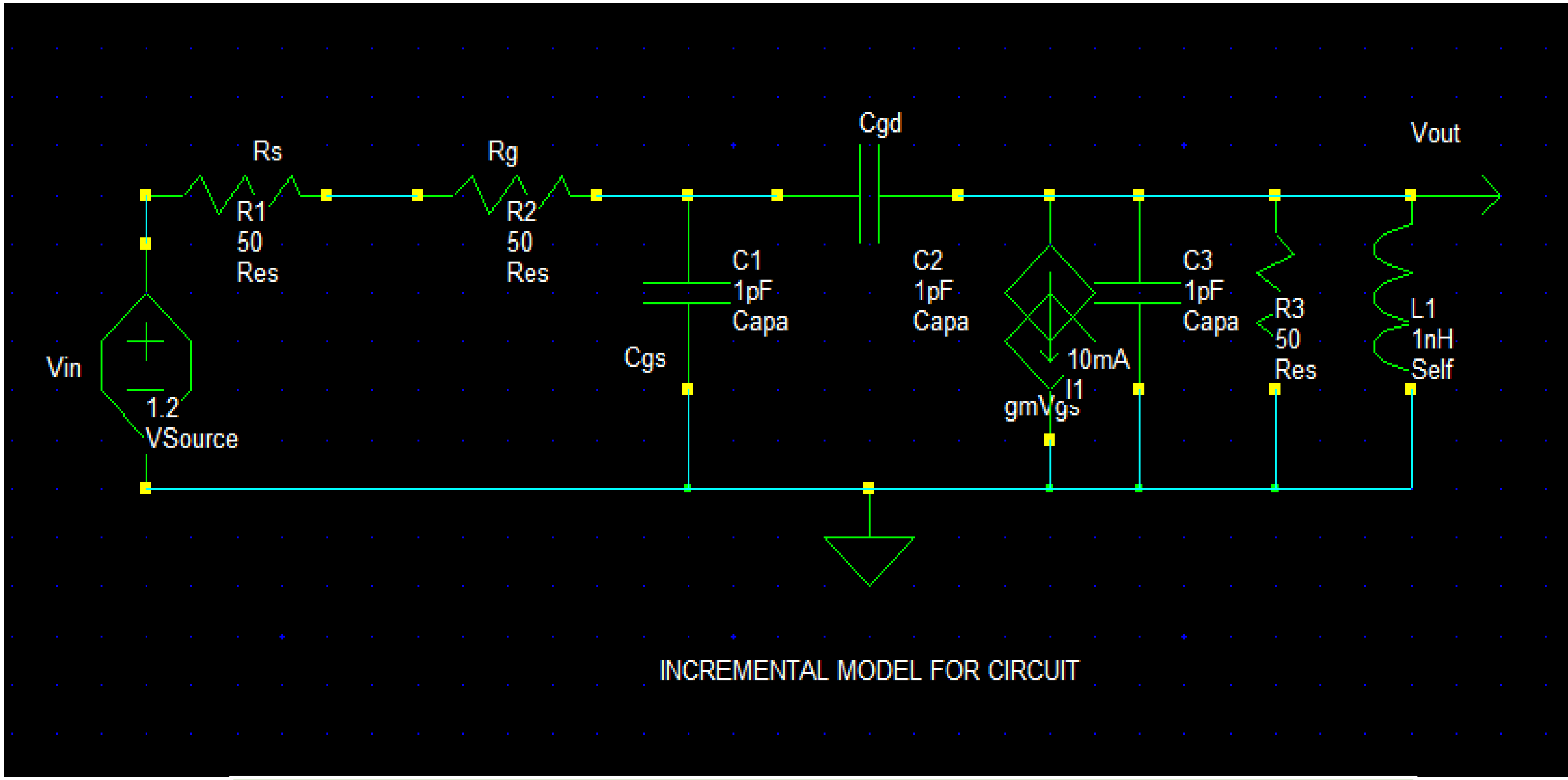
Note that the method of open-circuit time constants would predict quite a different result. Since the effective time constant is found there by summing all the individual time constants, the OCτ-estimated bandwidth would be

$$BW \approx \frac{1}{\tau N}$$

TUNED AMPLIFIERS




C1
1pF
Capa
R1
50
Res
L3
1nH
Self
Vout
W=1.0u
L=0.12u
Rs
R3
50
Res
1.2
VSource
AMPLIFIER WITH SINGLE TUNED LOAD



$$C_{eq} = C_{gd}\,[\,1 + g_m R_{eq}\,] = C_{gd}\,[\,1 + g_m\,(\,R_s + r_g\,)\,]$$

Thank you