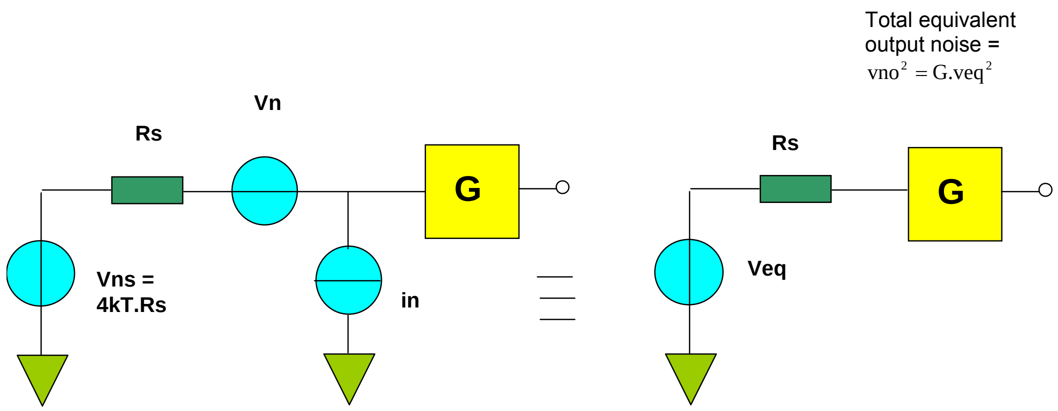
Noise Matching

From the noise tutorial the equivalent noise at the input of the device was given by:



Total equivalent input noise voltage =

$$\boxed{veq^2 = vns^2 + vn^2 + in^2Rs^2} \quad \text{where } vns^2 = 4kTRs$$

and noise factor F

$$F = \frac{\text{Total equivalent input noise power}}{\text{Input noise power due to the source only}} \approx \frac{veq^2}{vns^2} = \frac{1 + vn^2 + in^2Rs^2}{vns^2}$$

Ideally $F = 1$

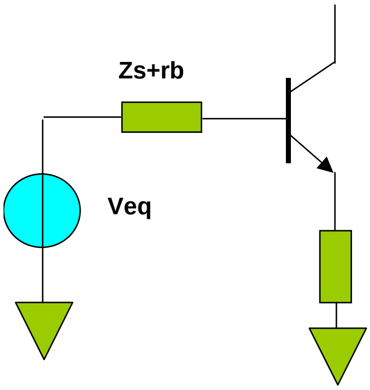
To minimise the noise factor, set dF/dRs = 0

$$F = \frac{1 + vn^2 + in^2Rs^2}{vns^2} \quad \text{differentiate w.r.t to Rs} = \frac{in^2Rs}{vns^2} = 0$$

$$\text{Rearrange to get Rs} = \frac{vns^2}{in^2}$$

Usually Rs is given and vn^2 and in^2 are optimised for best noise. Sometimes we must find a suitable value for Rs.

CE stage + series source impedance (Broadband)



The equivalent noise voltage (Veq) is given by:-

$$Veq^2 = 4kT\Delta f\left[Re(Zs) + rb + \frac{re}{2} + \frac{|rb+Zs|^2}{2\beta.re}\left(1+\frac{f_L}{f}\right)\right]$$

$\boxed{\text{Where } Re(Zs) \text{ is the real part of } Zs = Rs}$

$$\text{Where}\quad Vn^2 = 4kT\left(rb + \frac{re}{2}\right) \quad \text{and} \quad In^2 = \frac{2kT}{\beta.re}\left(1+\frac{f_L}{f}\right)$$

If we assume that Zs is resistive and ignore the flicker noise contribution the above equation simplifies to:-

$$Veq^2 = 4kT\,\Delta f\left[Rs + rb + \frac{re}{2} + \frac{|rb+Rs|^2}{2\beta.re}\right] \quad \text{differentiate w.r.t}\quad re = 4kT\,\Delta f\left[1 - \frac{|rb+Rs|^2}{2\beta.re^2}\right] = 0$$

$$\text{Therefore,}\quad re(opt) = \frac{rb+Rs}{\sqrt{\beta}}$$

$$veq^2{}_{(min)} = 4kT\,\Delta f\left[Rs + rb + re(opt)\right] = 4kT\,\Delta f(Rs+rb)\left(1+\frac{1}{\sqrt{\beta}}\right)$$

$$vns^2 = 4kT.Rs$$

$$\text{Therefore } F_{min} = \frac{veq^2}{vns^2} = \frac{4kT\Delta f\left[Rs + rb + re(opt)\right]}{4kT.Rs} = \frac{1\left[Rs + rb + re(opt)\right]}{Rs} = 1 + \frac{rb + re(opt)}{Rs}$$

$$F = 1+\frac{rb+re(opt)}{Rs} = 1+\frac{rb+\frac{rb+Rs}{\sqrt{\beta}}}{Rs} = 1+\frac{rb}{Rs}+\frac{1}{Rs}\left(\frac{rb}{\sqrt{\beta}}+\frac{Rs}{\sqrt{\beta}}\right) = 1+\frac{rb}{Rs}+\frac{rb}{Rs\sqrt{\beta}}+\frac{1}{\sqrt{\beta}}$$

Therefore, $F\min = \left(1+\frac{1}{\sqrt{\beta}}\right)\left(1+\frac{rb}{Rs}\right)$

If rb << Rs then

$$F\min = \left(1+\frac{1}{\sqrt{\beta}}\right) \text{ and } NF = 10\log_{10}\left(1+\frac{1}{\sqrt{\beta}}\right)dB$$

Eg

β	NF(dB)
10	1.19
50	0.57
100	0.41

If rb is not << Rs then assuming β=100

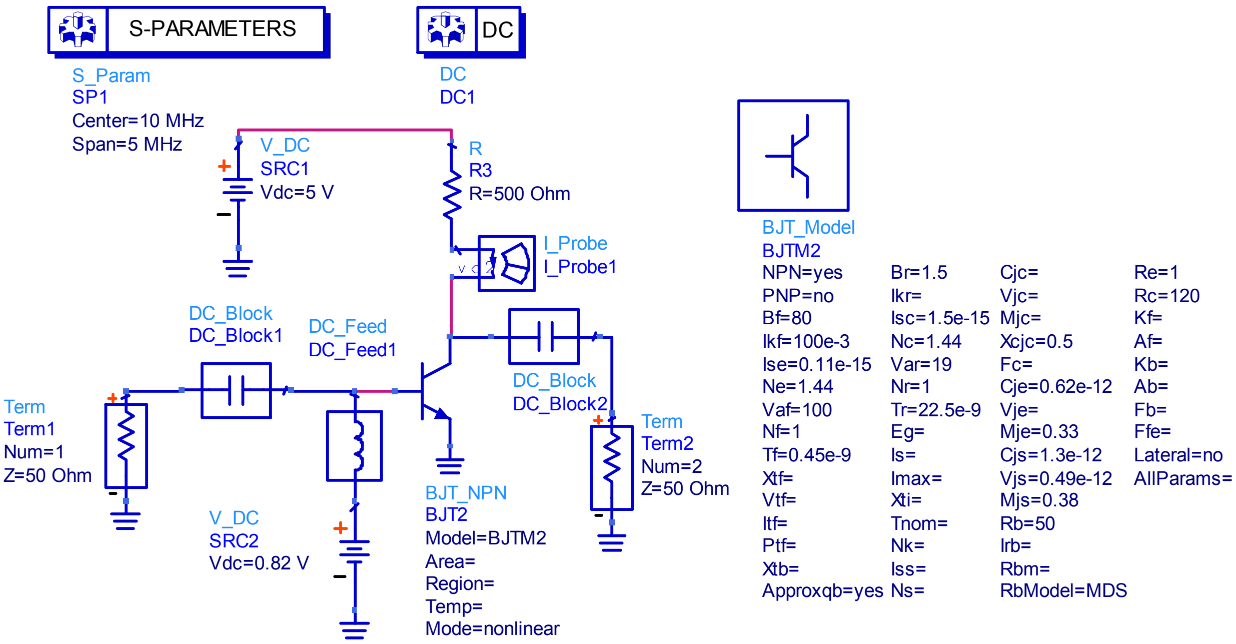
$$NF\min = 10\log_{10}\left[\left(1+\frac{1}{\sqrt{\beta}}\right)\left(1+\frac{rb}{Rs}\right)\right]dB$$

rb/Rs	NF(dB)
2	5.18
1	3.42
0.5	2.17
0.1	0.82

To minimise the noise factor we need to make rs << Rs

We could decrease the value of rb by adding transistors in parallel, however the value of cbe will double lowering the frequency reponse of the amplifier.

As an example consider the following Bipolar amplifier circuit. The device is biased at 5mA (The 500 ohm resistor thus dropping 2.5V). Initially a 50-ohm source resistor is connected to the amplifier input. (Note the DC blocks & feeds are ideal). The plot shows the noise figure (dB) at 10MHz resulting from a rb/Rs ratio of 50 ohms (rb = 50 ohms).



I_Probe1.i	nf(2)
4.860mA	3.844
	3.857
	3.873

If we decrease the source resistor Rs to 25ohms (ie rb/Rs ratio of 2) then the noise figure degrades to :-

I_Probe1.i	nf(2)
4.860mA	5.545
	5.555
	5.567

These results seem to agree with the theory. Theory suggests that increasing Rs will decrease the noise figure. However in reality this is not the case.

Repeating the simulation with rb/Rs ratio of 0.5 ie Rs = 100 we obtain the following data:-

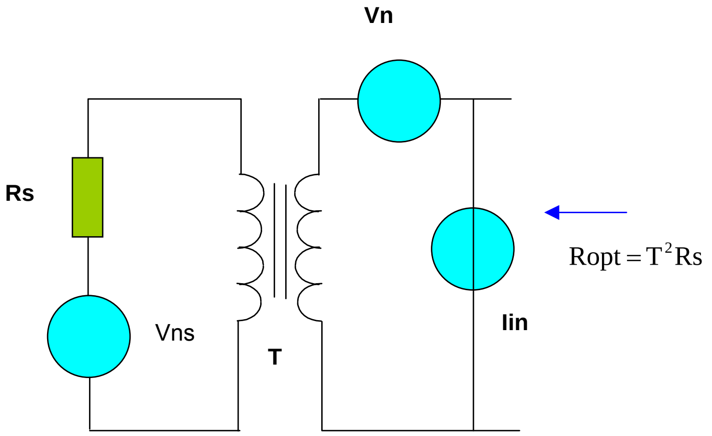
I_Probe1.i	nf(2)
4.860mA	2.775
	2.793
	2.816

An improvement on Rs = 50 ohms but not the predicted 2.2dB noise figure. The optimum noise figure of the device is limited by the device process.

The one thing this example shows is that if you want to use a source with a low source impedance then some form of low-loss matching is required. For example assume we want to connect a 10ohm source to our amplifier. The resulting simulation results in:-

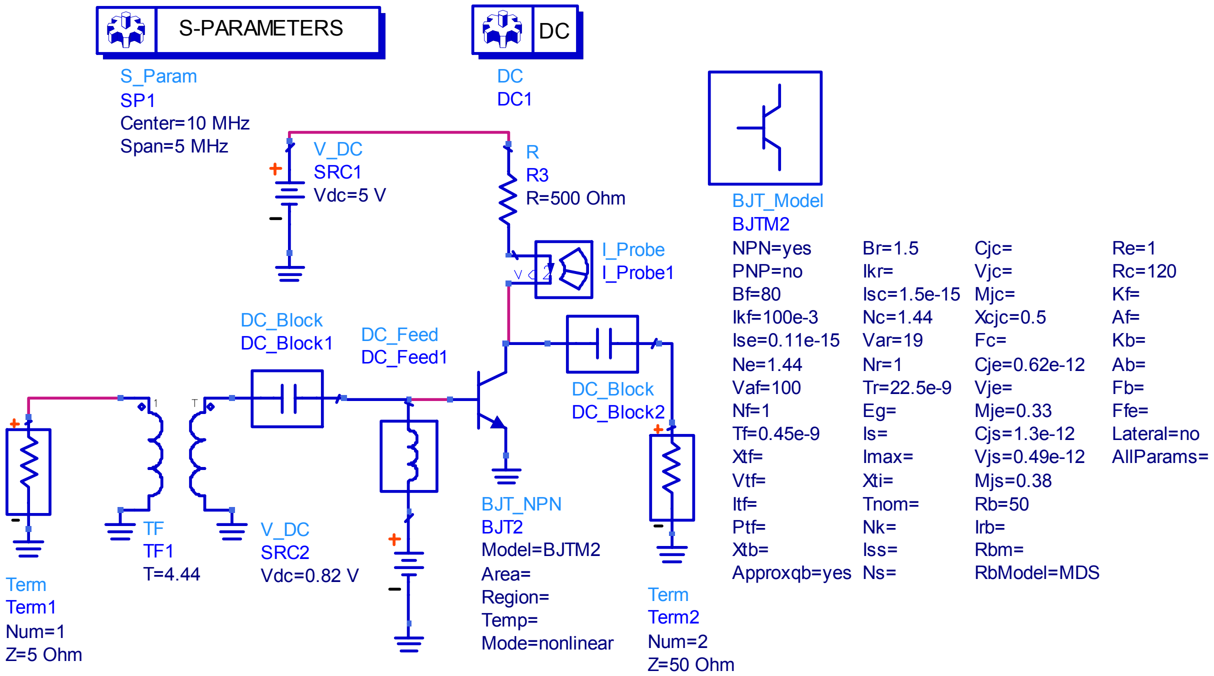
I_Probe1.i	nf(2)
4.860mA	11.182 11.190 11.199

However, we could use a transformer to match 10ohms to say 100 ohms to improve our noise figure ie



$$\text{Ropt} = \text{T}^2.\text{Rs} \quad \text{so} \quad \text{Ropt} = 100 \text{ and } \text{Rs} = 5 \text{ then } \sqrt{\text{T}} = \sqrt{\frac{100}{5}} \;;\; \text{T} = 4.4$$

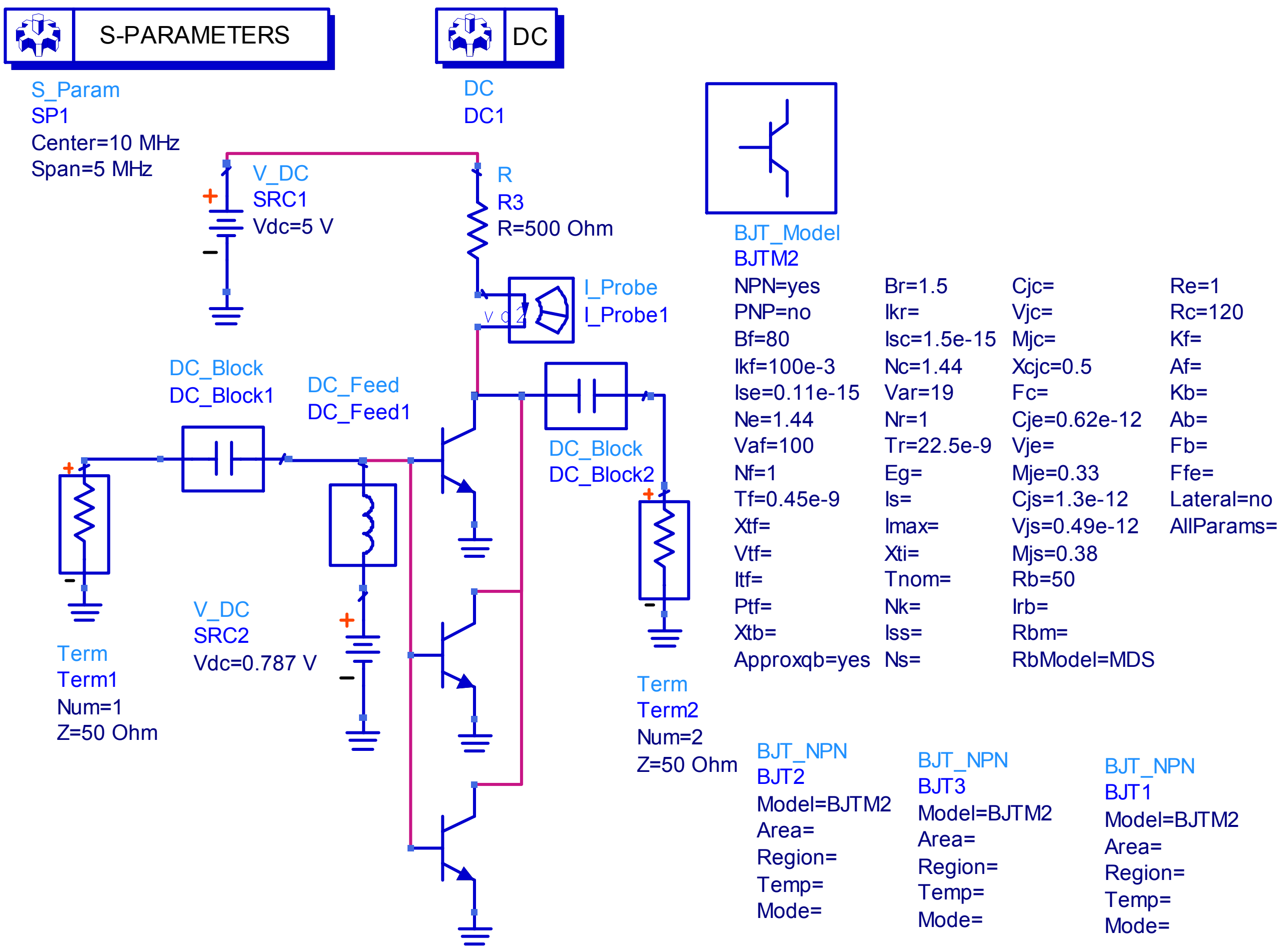
So the circuit is modified with an input transformer with a turns ration of ~4.4



We now obtain the following noise figure:-

I_Probe1.i	nf(2)
4.860mA	2.790 2.808 2.831

Also the noise figure can be reduced, by adding more transistors in parallel. The basic amplifier with Rs = 50 ohms yielded a noise figure of 3.8. Adding two more transistors (and readjusting the bias to maintain 5mA in the load) improves the noise figure to:-



freq	I_Probe1.i	freq	nf(2)
0.0000 ...	5.153mA	7.500M...	1.902
		10.00M...	1.912
		12.50M...	1.925

Single-frequency noise matching

The previous example was a broad-band LNA as the matching was purely resistive, however for frequency matching we need to add frequency dependant components in the matching circuit ie inductance and capacitance.

$$Veq^2 = 4kT\Delta f\left[Rs + rb + \frac{re}{2} + \frac{\left|rb + (Rs + j\omega.Ls)\right|^2}{2\beta.re}\right] \text{ neglecting flicker noise contribution}$$

$$= 4kT\Delta f\left[Rs + rb + \frac{re}{2} + \frac{rb^2 + Rs^2 + \omega.^2 Ls^2}{2\beta.re}\right] = 4kT\Delta f\left[Rs + rb + \frac{re}{2} + \frac{(rb + Rs)^2 + (2\pi.f.Ls)^2}{2\beta.re}\right] V^2/Hz$$

To minimise the input noise at a single frequency , we set $d(veq^2)/dre = 0$.

$$4kT\Delta f\left[Rs + rb + \frac{re}{2} + \frac{(rb + Rs)^2 + (2\pi.f.Ls)^2}{2\beta.re}\right] \quad \Delta f = 1 \text{ (Single frequency)}$$

$$4kT\left[\frac{1}{2} - \frac{(rb + Rs)^2 + (2\pi.f.Ls)^2}{2\beta.re^2}\right] = 0 \text{ Eliminate constants}$$

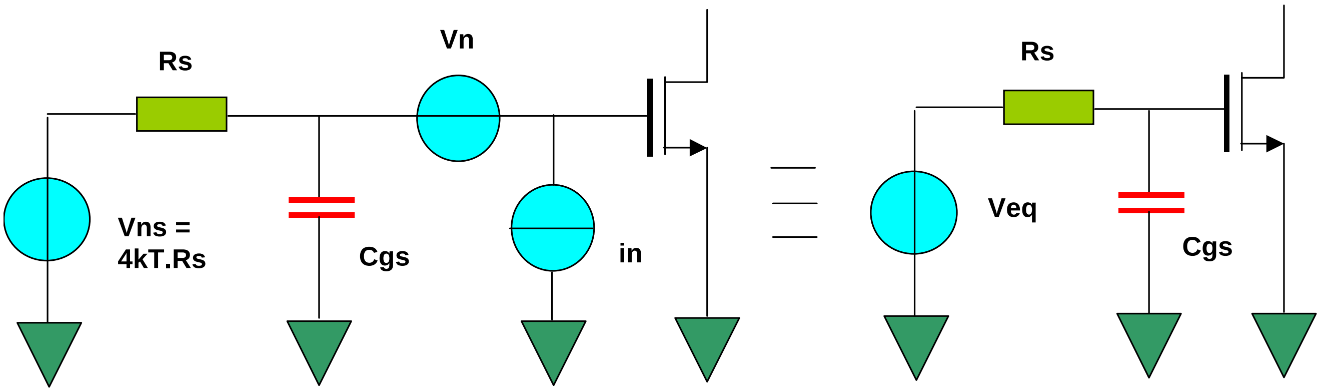
$$\frac{(rb + Rs)^2 + (2\pi.f.Ls)^2}{2\beta.re^2} = 0$$

$$\boxed{\text{Rearrange to get } re(opt) = \sqrt{\frac{(rb + Rs)^2 + (2\pi.f.Ls)^2}{2\beta}}}$$

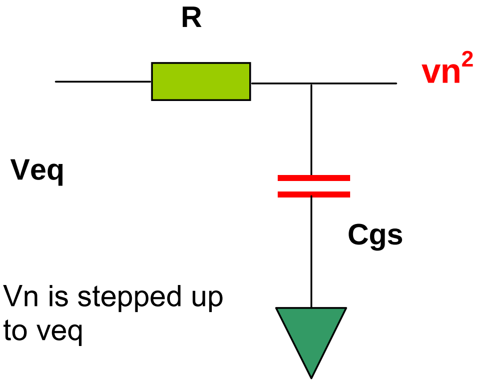
To minimise the noise over a given bandwidth (f_2-f_1) we perform a similar process to obtain:-

$$re(opt) = \frac{1}{\sqrt{\beta}}\left\{(rb + Rs)^2 + \frac{(2\pi.Ls)^2}{3}\left(\frac{f_2^3 - f_1^3}{f_2 - f_1}\right)\right\}^{1/2}$$

CS stage with Series Source Impedance



For this case we refer back to the input ie



Ignoring effects of in^2

$$veq^2 = vns^2 + vn^2 \left| \frac{1+Zs}{Xcgs} \right|^2$$

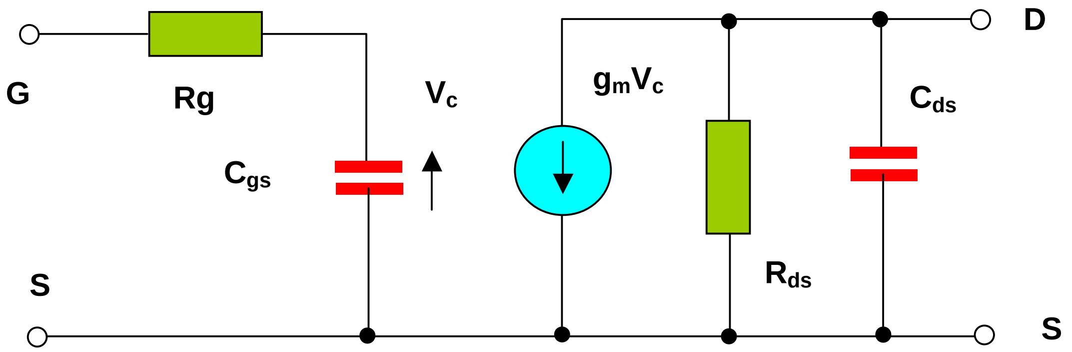
Assuming that the source is resistive

$$veq^2 = \left(4kTRs + \frac{4kT.\gamma}{gm}\left(\frac{1+f_1}{f}\right)\left(1+\omega^2.Cgs^2.Rs^2\right) \right) \quad V^2/Hz$$

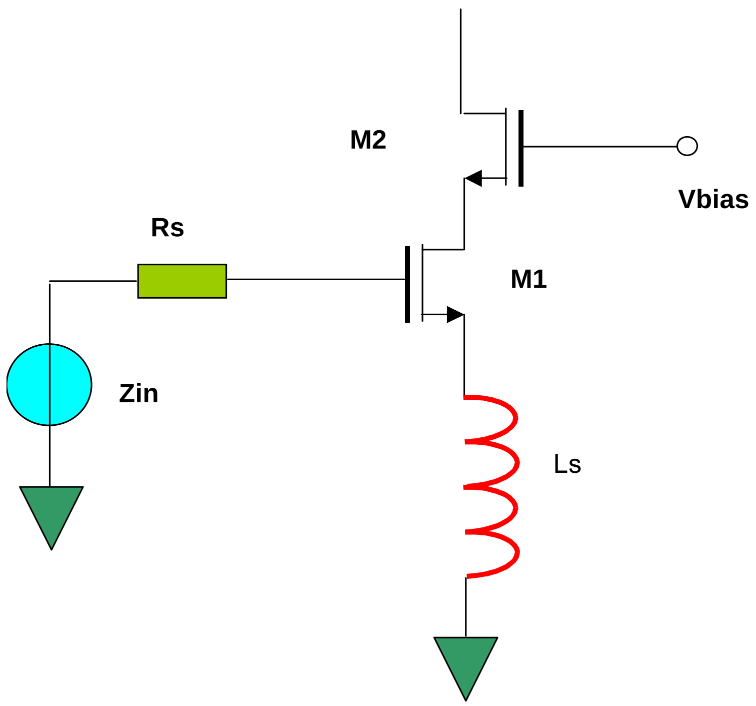
We can see from the above equation that making gm as large as possible will minimise the second noise contribution and is achieved by making the **W/L** ratio large. (There is little we can do about the first contribution ie the source noise). However making W/L large will increase Cgs and hence lower the operating bandwidth of the device.

Inductive Source Degeneration Input matching

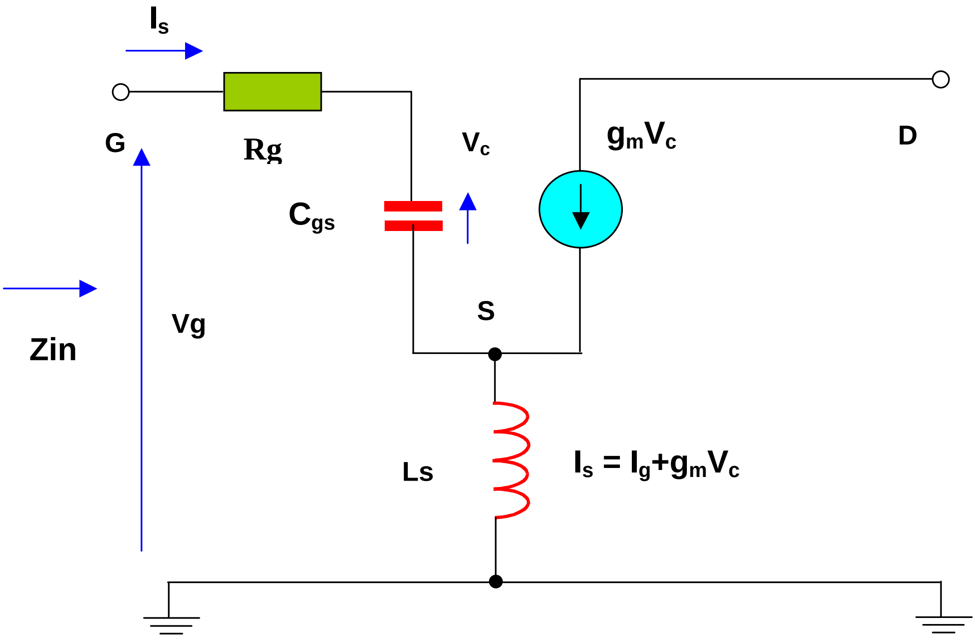
Previous design examples have assumed that the input is resistive and offered the best value of source resistance (Rs) to obtain best noise match. In reality the input of the device is reactive with a real and capacitive impedance. The circuit below shows the equivalent model of a MOSFet:-



What we would like to do is remove the capacitive reactance (and therefore restore the FET input to a pure resistance) and one way of achieving this is to add inductive feedback to the source. The input impedance of the MOSFet circuit below will be capacitive due to the gate-source capacitance.



If we add lossless inductive feedback to the source the above circuit is again re-drawn as shown below:



$$Z_{in} = \frac{V_g}{I_g} = \frac{(I_gR_g + Vc + j\omega.I_s.L_s)}{I_g} \qquad - \ (1)$$

$$V_c = \frac{I_g}{S.C_{gs}} \quad \text{and} \quad I_s = I_g + g_mV_c \quad \text{sub into (1)}$$

Let $S = j\omega$

$$Z_{in} = \frac{\left(I_g R_g + \frac{I_g}{S.C_{gs}} + S.\left(I_g + g_m \frac{I_g}{S.C_{gs}}\right).L_s\right)}{I_g}$$ Divide through by Ig

$$Z_{in} = \frac{I_g R_g}{I_g} + \frac{I_g}{I_g.S.C_{gs}} + S.\left(I_g + g_m \frac{I_g}{S.C_{gs}}\right).\frac{L_s}{I_g} =$$

$$= R_g + \frac{1}{S.C_{gs}} + S.L_s + S.L_s.g_m \frac{1}{S.C_{gs}}$$

$$= R_g + \frac{1}{S.C_{gs}} + S.L_s + \frac{S.L_s.g_m}{S.C_{gs}} = R_g + \frac{1}{S.C_{gs}} + S.L_s + \frac{L_s.g_m}{C_{gs}}$$

$$Rin = Rg + \frac{L_s.g_m}{C_{gs}} + S\left(\frac{1}{S^2.C_{gs}} + L_s\right)$$ Sub in $S = j\omega$

$$Rin = Rg + \frac{L_s.g_m}{C_{gs}} + j\omega\left(\frac{1}{j\omega^2.C_{gs}} + L_s\right) = Rg + \frac{L_s.g_m}{C_{gs}} + j\omega\left(L_s - \frac{1}{\omega^2.C_{gs}}\right)$$

$$Rin = Rg + \frac{L_s.g_m}{C_{gs}} + j\left(\omega L_s - \frac{\omega}{\omega^2.C_{gs}}\right)$$

$$Rin = Rg + \frac{L_s.g_m}{C_{gs}} + j\left(\omega L_s - \frac{1}{\omega C_{gs}}\right)$$ Can be re - written as

$$Rin = Rg + Ra + j[X_{LS} - X_{CGS}]$$ Where $Ra = \frac{L_s.g_m}{C_{gs}}$

Therefore, the impedance of the MOSFET without feedback is:-

$$Rin = Rg - jX_{CGS}$$

Adding series feedback adds the following term to the original input impedance:-

$$Ra + jX_{LS}$$

Additionally, another inductor is added in series with the gate **Lg** that is selected to resonate with the **Cgs** Capacitor.

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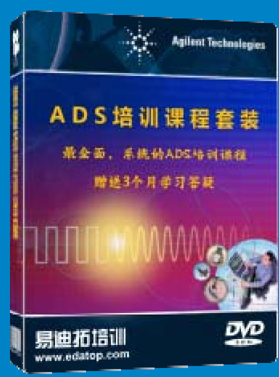

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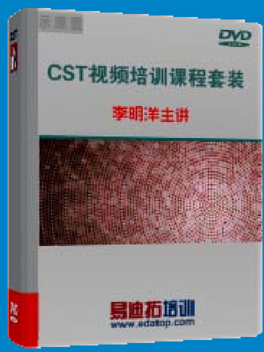

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