Electrical Noise and Low Noise Amplifier presentation

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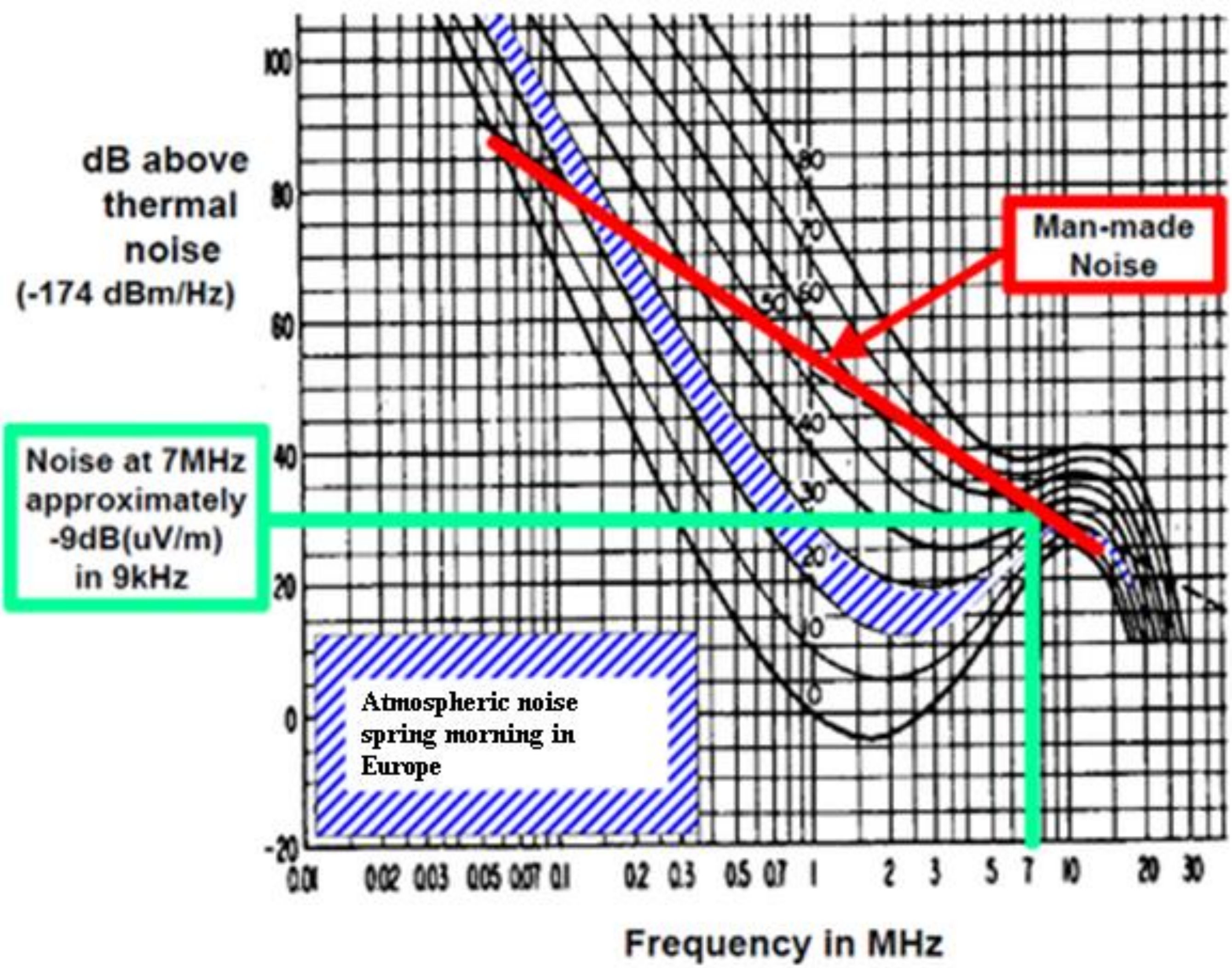
Outline

- Types of noise
- Noise and gain circles
- SNR, bandwidth, noise factor, noise figure
- Noise correlation
- Noise in 2-port devices
- Noise in semiconductors
- Quantization noise in A/D convertors
- Design cases for LNA's
- Novel design approach
- Large signal parameters

Electrical Noise

- Atmospheric noise
- Man-made noise
- Johnson noise/ thermal
- Schottky noise/shot
- Flicker noise
- Transit time noise- really only for Tubes
- Generation-recombination noise
- Avalanche noise

RF noise spectrum levels



Johnson noise

- The Johnson noise (thermal noise) is due to the movement of molecules in solid devices called Brown’s molecular movements.
- It is expressed as $v_n^2 = 4kT_0RB \quad (emf) \quad (volt^2)$
- The power can thus be written as

$$Noise\ Power = \frac{v_n^2}{4R} = kT_0B \quad (W/Hz)$$

$$for\ B = 1Hz, \quad Noise\ Power = kT_0$$

$$T = 290K\ and\ k - Boltzmann's\ const. = 1.38 \times 10^{-23}$$

$$by\ Thevenin,\ Noise\ Power = 1.38 \times 10^{-23} \times 290 = 4 \times 10^{-21}\ W$$

$$L(\omega) = 10.\log\left(\frac{v_n^2/R}{1mW}\right) = -173.97dBm\ or\ about\ -174dBm$$

- In order to reduce this noise, the only option is to lower the temperature, since noise power is directly proportional to temperature.
- The Johnson noise sets the theoretical noise floor.

Planck's radiation law

- The available noise power does not depend on the value of resistor but it is a function of temperature T. The noise temperature can thus be used as a quantity to describe the noise behavior of a general lossy one-port network.
- For high frequencies and/or low temperature a quantum mechanical correction factor has to be incorporated for the validation of equation. This correction term results from Planck's radiation law, which applies to blackbody radiation.

$$P_{av} = kT \cdot \Delta f$$

$$P_{av} = kT\Delta f \bullet p(f,T); \quad with\ p(f,T) = \left[\frac{hf}{kT} \Big/ \left(e^{\left(\frac{hf}{kT}\right)} - 1 \right) \right]$$

$$where\ h = 6.626 \cdot 10^{-34}\ J-s,\ Planck's\ \text{constant}$$

Schottky /shot noise

- The Schottky noise occurs in conducting PN junctions (semiconductor devices) where electrons are freely moving. The root mean square (RMS) noise current is given by

$$\overline{i_n^2} = 2 \times e \times I_{dc}; \quad \boldsymbol{P} = \boldsymbol{i_n^2} \times \boldsymbol{Z}$$

Where

e is the charge of the electron, P is the noise power, and $\boldsymbol{I_{dc}}$ is the dc bias current. Z is the termination load (can be complex).

- Since the origin of this noise generated is totally different, hence they are independent of each other.

Flicker noise

- The electrical properties of surfaces or boundary layers are influenced energetically by states, which are subject to statistical fluctuations and therefore, lead to the flicker noise or $1/f$ noise for the current flow.
- $1/f$ noise is observable at low frequencies and generally decreases with increasing frequency f according to the $1/f$ - law until it will be covered by frequency independent mechanism, like thermal noise or shot noise.

Flicker noise

- Example: The noise for a conducting diode is bias dependent and is expressed in terms of AF and KF.

$$\left\langle i_{Dn}^{2} \right\rangle_{AC} = 2eI_{DC}B + KF\frac{I_{DC}^{AF}}{f}B$$

- The AF is generally in range of 1 to 3 (dimensionless quantity) and is a bias dependent curve fitting term, typically 2.
- The KF value is ranging from $1E^{-12}$ to $1E^{-6}$, and defines the flicker corner frequency.

Transit time and recombination noise

- When the transit time of the carriers crossing the potential barrier is comparable to the periodic signal, some carriers diffuse back and this causes noise. This is really seen in the collector area of NPN transistor.
- The electron and hole movements are responsible for this noise. The physics for this noise has not been fully established.

Avalanche Noise

- When a reverse bias is applied to semiconductor junction, the normally small depletion region expands rapidly.
- The free holes and electrons then collide with the atoms in depletion region, thus ionizing them and produce spiked current called the avalanche current.
- The spectral density of avalanche noise is mostly flat. At higher frequencies the junction capacitor with lead inductance acts as a low-pass filter.
- Zener diodes are used as voltage reference sources and the avalanche noise needs to be reduced by big bypass capacitors!

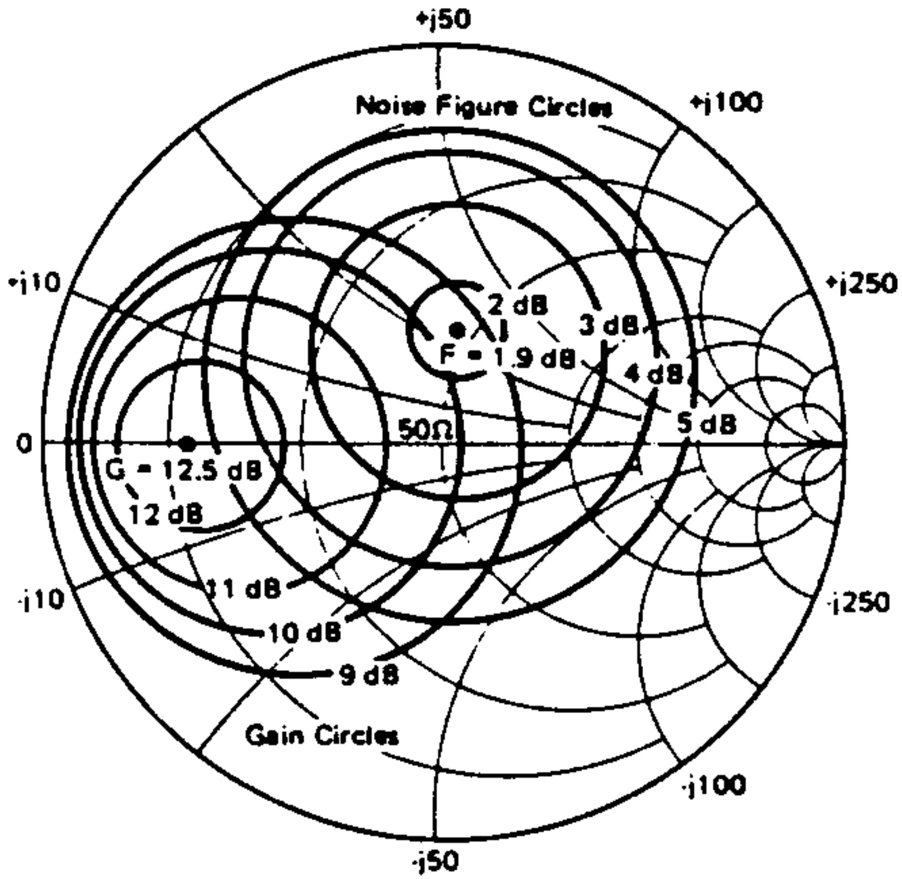
Noise and Gain circles

- These plots are used to:
- Determine noise and gain for unilateral; (S_{12}=0) or a bilateral, (S_{12}>0) condition.
- Find the termination for which the transistor remains unconditionally unstable and obtain the minimum noise figure.
- For power matching it is required to have

$$S_{11}^{*} = complex\ conjugate\ of\ S_{11}$$

- For low noise designs the source supplied at the input termination must be Γ_{opt}

Noise and Gain Circles



$f = 2GHz$
$\Gamma_L = S_{22}$
$(S_{12} \neq 0)$

Feedback

$$F = F_{min} + \frac{R_n}{G_g}\left[\left(G_{opt} - G_g\right)^2 + \left(B_{opt} - B_G\right)^2\right]$$

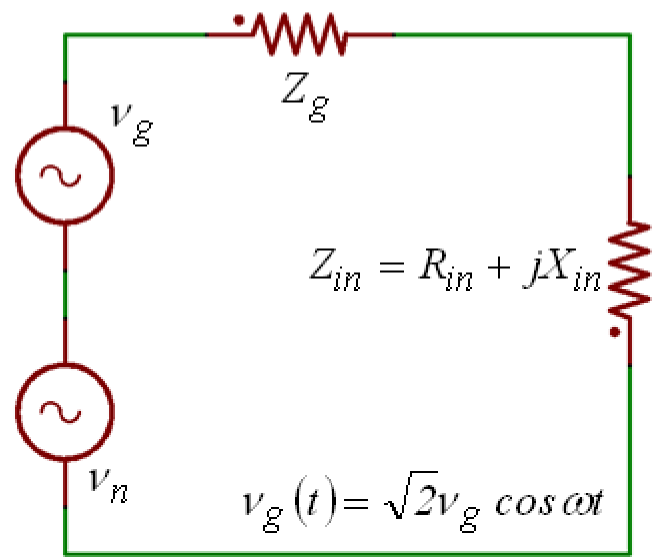
There is a range of values of input reflection coefficients over which the noise figure is constant. In plotting these points of constant noise figure, we obtain noise circles, which can be drawn on the Smith chart Γ_G plane.

Signal-To-Noise Ratio

- The signal power delivered to the input is given by P_{in}.
- where V_g is the rms voltage of the input signal supplied to the system, and the noise power supplied to the input is $P_{N_{in}}$, can be less than available power.
- The noise power at the input is provided by the noise energy of the real part of Z_g. The input impedance Z of the system in the form $Z = R_{in} + jX_{in}$ is assumed to be complex.

$$P_{in} = \frac{v_g^2 \, Re(Z_{in})}{\left|Z_g + Z_{in}\right|^2}$$

$$P_{N\,in} = \frac{\overline{v_n^2} \, Re(Z_{in})}{\left|Z_g + Z_{in}\right|^2}$$



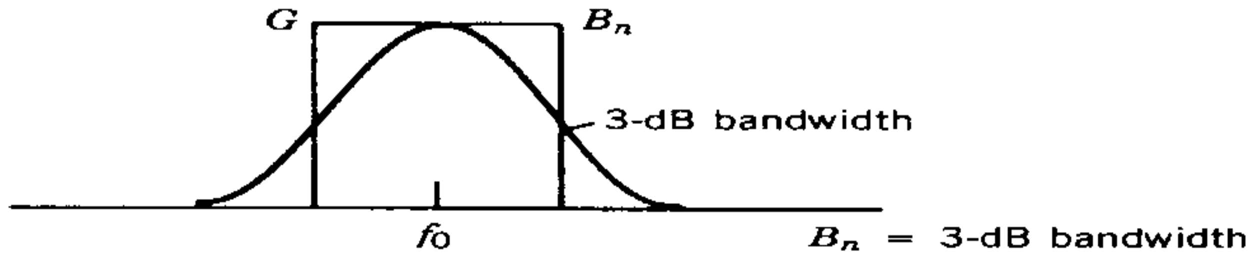
SNR (Signal-to-noise ratio)

- In all environments noise exist, so absolute value of the signal makes little difference. It is important to know how much higher the signal is in comparison to the noise. This ratio is termed as noise factor.
- In any system SNR, rather than the absolute value of noise, is more important.
- SNR is defined as power ratio, typically 3.16 or 10dB is needed for understanding voice signals correctly in a 2.4kHz bandwidth.

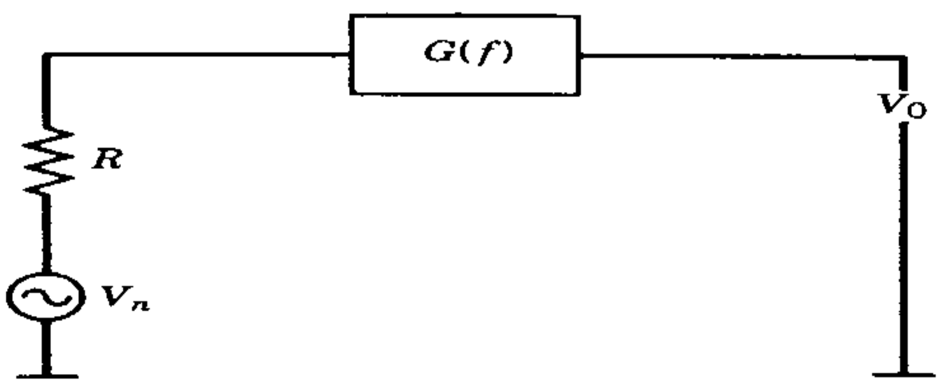
Noise Bandwidth

- Noise bandwidth, B_n, is defined as the equivalent integrated bandwidth, as shown in Figure below. For reasons of group delay correction, most practical filters have round Gaussian response rather than sharp comers (Chebyshev- type).
- Graphical and mathematical explanation of the noise bandwidth from a comparison of the Gaussian-shaped bandwidth to the rectangular filter response.

Noise Bandwidth



$$G(f) = \left| \frac{V_0(f)}{V_n(f)} \right|^2$$



$$V_o^2 = \int_0^\infty 4kT_0RG(f)\,dF$$
$$= 4kT_0R \int_0^\infty G(f)\,dF$$
$$B_n = \frac{1}{G} \int_0^\infty G(f)\,dF$$

B_n = noise bandwidth

Noise factor-noise figure

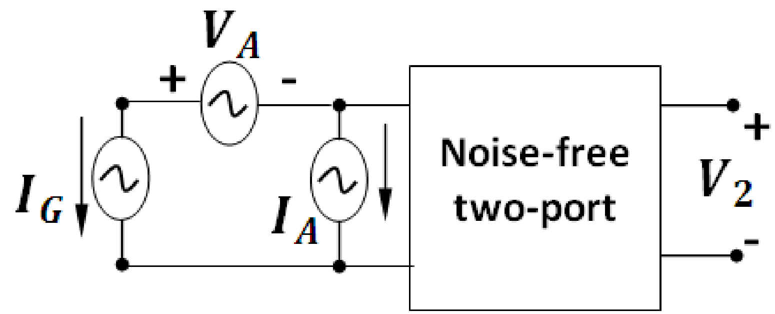
- The noise factor is a measured parameter defined as,

$$F = \frac{available\ S/N\ \ Power\ \ ratio\ at\ the\ input}{available\ S/N\ \ Power\ \ ratio\ at\ the\ output}$$

- When referred in decibels it is termed as noise figure (NF).

$$NF = (F)_{dB} = 10.log(F)$$

Noise factor-Derivation



$$F = \left|\frac{I_G}{I_G}\right|^2 + \left|\frac{I_A + Y_G V_A}{I_G}\right|^2$$

$I_G - Current\ through\ generator$

$I_A = Y_{cor} V_A + I_u$

$|V_A|^2 = 4kTBR_n \quad |I_u|^2 = 4kTBG_u \quad |I_g|^2 = 4kTBG_g$

$$\left|\frac{I_A + Y_G V_A}{I_G}\right|^2 = \left|\frac{Y_{cor} V_A + I_u + Y_G V_A}{I_G}\right|^2$$

$$= \left|\frac{(Y_{cor} + Y_G)V_A + I_u}{I_G}\right|^2; \quad Y_{cor} = G_{cor} + jB_{cor}$$

$$|Y_{cor} + Y_G|^2 = (G_{cor} + G_u)^2 + (B_{cor} + B_u)^2$$

$Where\ Y_g \rightarrow Generator\ impedance \quad V_A \rightarrow Noise\ Voltage$

$G_u \rightarrow uncorrelated\ noise\ component \quad I_A \rightarrow Noise\ Current$

$Y_{cor} \rightarrow correlated\ noise\ component \quad R_n \rightarrow Equivalent\ Noise\ Resistance$

Noise factor

- Ideally F=1 or 0dB, is the minimum possible noise figure for a system.

$$F = \frac{R_g}{R_g} + \frac{R_u}{R_g} + \frac{G_n}{R_g}\left[\left(R_g + R_{cor}\right)^2 + \left(X_g + X_{cor}\right)^2\right]$$

$$F = 1 + \frac{R_u}{R_g} + \frac{G_n}{R_g}\left[\left(R_g + R_{cor}\right)^2 + \left(X_g + X_{cor}\right)^2\right]$$

$$Where\ R_g \rightarrow Generator\ impedance$$

$$R_u \rightarrow uncorrelated\ noise\ component$$

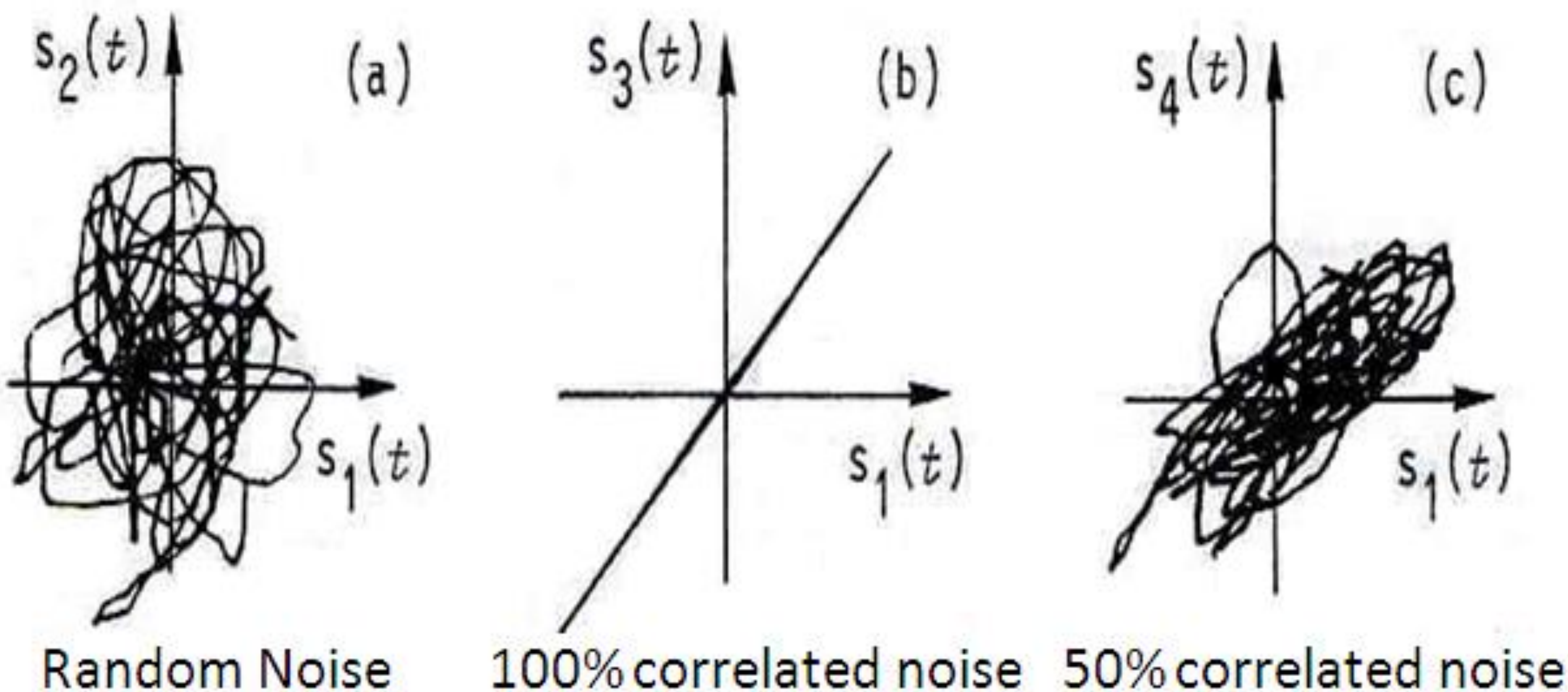
$$G_n \rightarrow equivalent\ noise\ conductance$$

- Noise correlation needs to be explained.

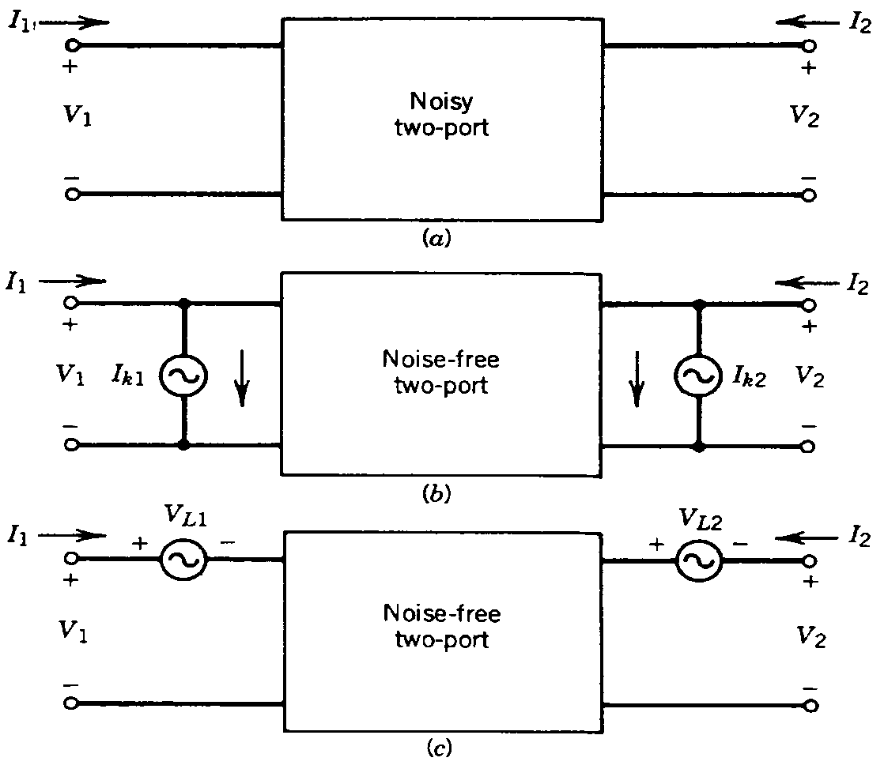
Concept of Noise Correlation

- Signals resulting from the same origin are always 100 % correlated. the origin of shot noise generated is totally different than the thermal noise, therefore, there is no correlation between the two.
- The correlation of noise sources is described by the noise correlation coefficient. The correlation coefficient "C_r" is zero for 2-totally different noise sources (Thermal noise and Schottky noise) as discussed above.
- It can be 0.5 for partially correlated sources or is 1 for 100% correlated noise sources.

Concept of Noise Correlation



Noise in 2-Port



- Noise in two-ports:

(a) general form;

(b) admittance form;

(c) impedance form

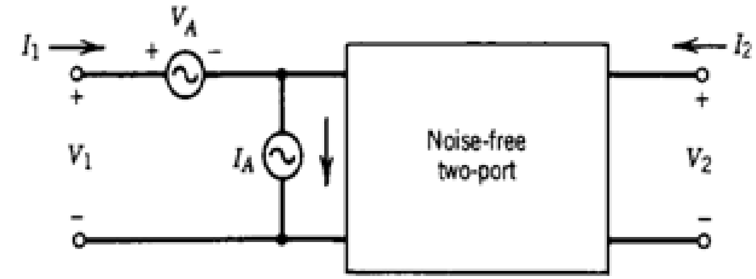
$$I_1 = y_{11}V_1 + y_{12}V_2 + I_{k1}$$
$$I_2 = y_{21}V_1 + y_{22}V_2 + I_{k2}$$
$$V_1 = z_{11}I_1 + z_{12}I_2 + V_{L1}$$
$$V_2 = z_{21}I_1 + z_{22}I_2 + V_{L2}$$

- where the external noise sources are I_{K1}, I_{K2}, V_{L1}, and V_{L2}.

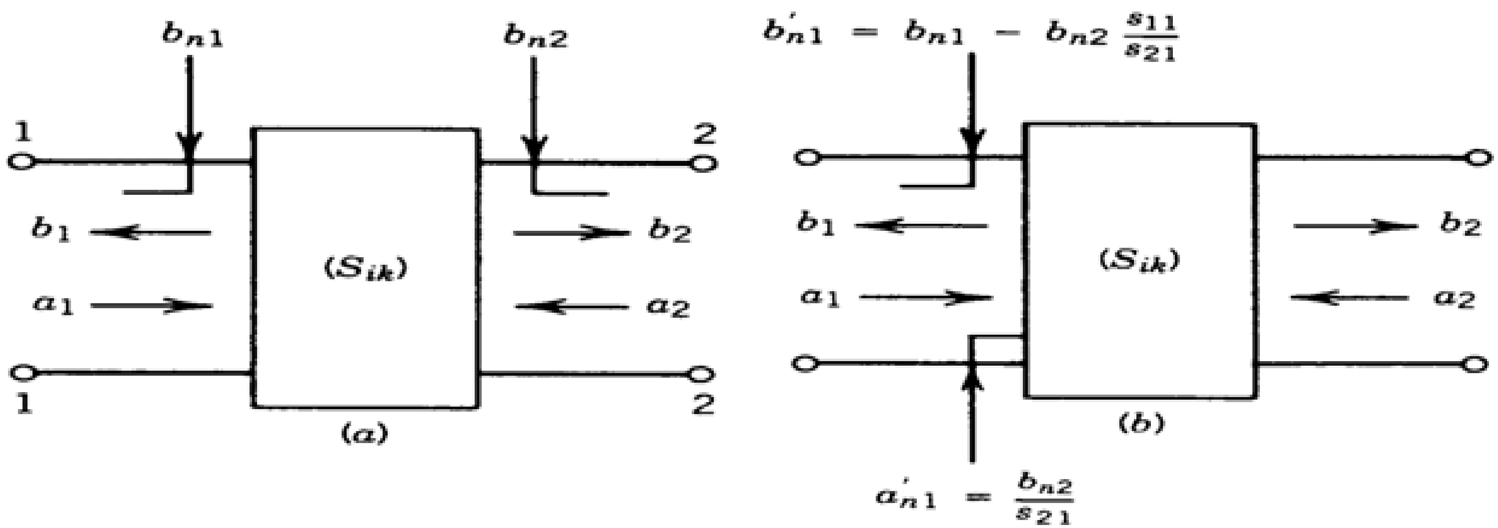
Noise in 2-Port

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} + \begin{bmatrix} b_{n1} \\ b_{n2} \end{bmatrix}$$

S-parameters of 2-port circuit



Chain Matrix form of 2-port circuit



S-parameters representation of noisy 2-port circuit

Noise in 2-Port

- Many sources of noise are represented by only two noise sources at the device input, the two equivalent input noise sources are often a complicated combination of the circuit internal noise sources. Often, some fraction of V_A and I_A is related to the same noise source. This means that V_A and I_A are not independent in general.

- So the correlation between the V_A and I_A needs to be computed. The noise source V_A represents all of the device noise referred to the input when the generator impedance is zero; that is, the input is short-circuited. The noise source I_A represents all of the device noise referred to the input when the generator admittance is zero; that is, the input is open circuited.

Noise in 2-Port

- By defining correlation admittance, we can simplify the mathematics and get some physical intuition for the relationship between noise figure and generator admittance. Since some fraction of I_A will be correlated with V_A, I_A can be split into correlated and uncorrelated parts.
- I_u is the part of I_A uncorrelated with V_A. Since I_n is correlated with V_A, I_n is proportional to V_A and the constant of proportionality is the correlation Admittance Y_{cor} .
- Y_{cor} is not a physical component located somewhere in the circuit. Y_{cor} is a complex number derived by correlating the random variables I_A and V_A.

$$\boldsymbol{I_A = I_n + I_u}$$

$$\boldsymbol{I_A = Y_{cor} V_A + I_u}$$

$$\boldsymbol{I_n = Y_{cor} V_A}$$

Noise in 2-Port

- Multiply each side of by V_A^* and average the result
- where the I_u term averaged to zero since it is uncorrelated with V_A.

$$I_A = Y_{cor} V_A + I_u$$

$$V_A^* I_A = Y_{cor} V_A^* V_A + V_A^* I_u \Rightarrow V_A^* I_A = Y_{cor} \overline{V_A^2}$$

where

$$V_A^* = \text{complex conjugate of } V_A$$

- “Correlation coefficient" is a normalized dimensionless quantity.

Noise in 2-Port

- Note that the dual of this admittance description is the impedance description.

$$Y_{cor} = \frac{\overline{V_A^* I_A}}{\overline{V_A^2}}$$

- Thus the impedance representation has the same equations as above with *Y* replaced by *Z*, *I* replaced by *V*, and *V* replaced by *I*.

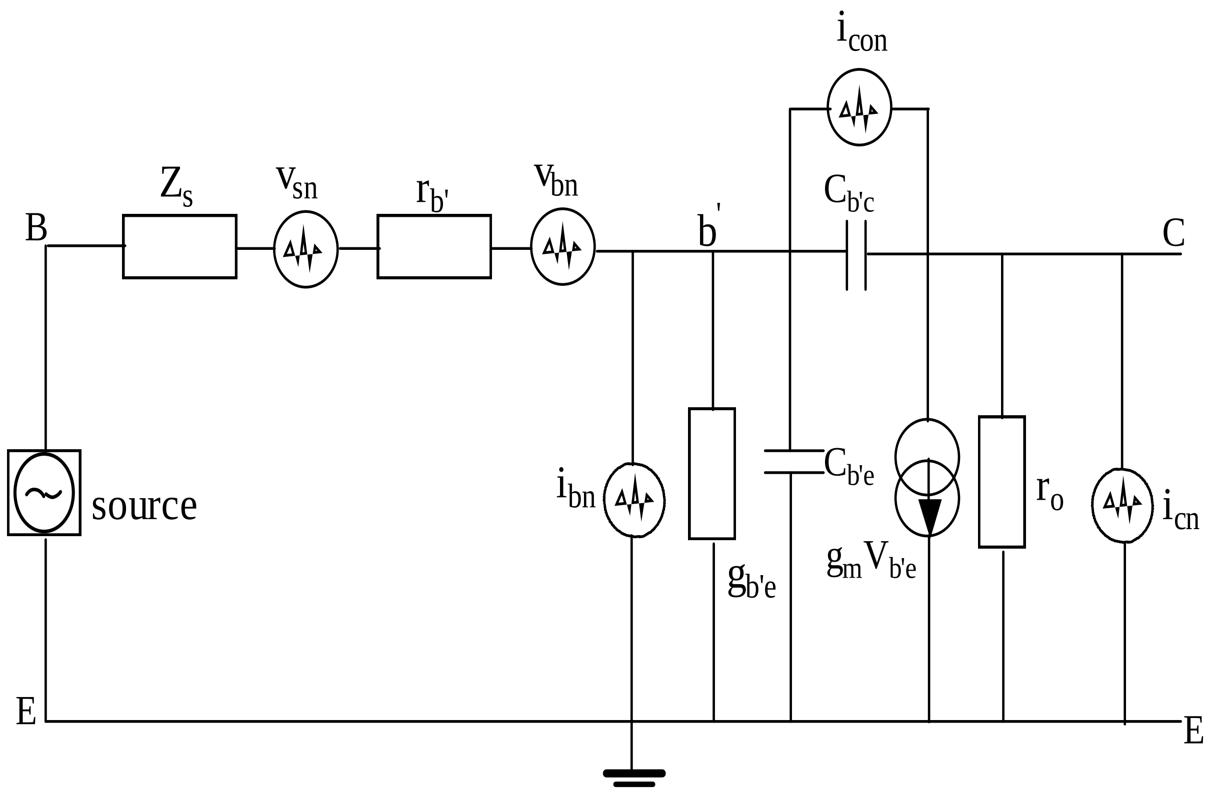
$$\frac{\overline{V_A^* I_A}}{\sqrt{\overline{V_A^2}\,\overline{I_A^2}}} \Rightarrow C_r = Y_{cor}\sqrt{\frac{\overline{V_A^2}}{\overline{I_A^2}}}; \qquad C_r = Z_{cor}\sqrt{\frac{\overline{I_A^2}}{\overline{V_A^2}}};$$

Noise in semiconductors

- Diodes
- BJT / HBT
- FET / JFET
- MOSFET
- GaAs FET

Noisy Bipolar transistor

The noise equations for BJT and HBT have little difference. The GaAs based transistor have large flicker noise and different behavior at microwave frequencies while the silicon HBT's are more predictable.



- Hybrid-π Configuration (Grounded Emitter)

Bipolar transistor noise

$$G_{opt} = Rg_{opt} + jXg_{opt}$$

$$G_{opt} = \sqrt{\frac{C_{ii^\bullet}}{C_{uu^\bullet}} - \left[Im\left(\frac{C_{ui^\bullet}}{C_{uu^\bullet}} \right) \right]^2} + j\,Im\left(\frac{C_{ui^\bullet}}{C_{uu^\bullet}} \right);$$

$$R_n = \frac{C_{uu^\bullet}}{2kT} \qquad F_{min} = 1 + \frac{C_{ui^\bullet} + C_{uu^\bullet} G^\bullet_{opt}}{KT}$$

$$C = \begin{bmatrix} C_{uu^\bullet} & C_{ui^\bullet} \\ C_{u^\bullet i} & C_{ii^\bullet} \end{bmatrix} = \begin{bmatrix} R_n & \frac{F_{min} - 1}{2} - R_n G^\bullet_{opt} \\ \frac{F_{min} - 1}{2} - R_n G_{opt} & R_n \left| G_{opt} \right|^2 \end{bmatrix};$$

- The noise correlation matrix C contains all necessary information about the four extrinsic noise parameters F_{min}, Rg_{opt}, Xg_{opt} and R_n of the bipolar

Bipolar transistor noise

$$R_n = \frac{C_{uu^\bullet}}{2kT} = r_b\left(\frac{1+\left(f/f_b\right)^2}{\alpha_0^2} - \frac{1}{\beta_0}\right) + \frac{r_e}{2}\left[\frac{1+\left(f/f_b\right)^2}{\alpha_0^2} + \left(g_e r_b\right)^2\left\{1-\alpha_0+\left(\frac{f}{f_b}\right)^2+\left(\frac{f}{f_e}\right)^2+\left(\frac{1}{\beta_0}-\left(\frac{f}{f_b}\right)\left(\frac{f}{f_e}\right)\right)^2\right\}\right]$$

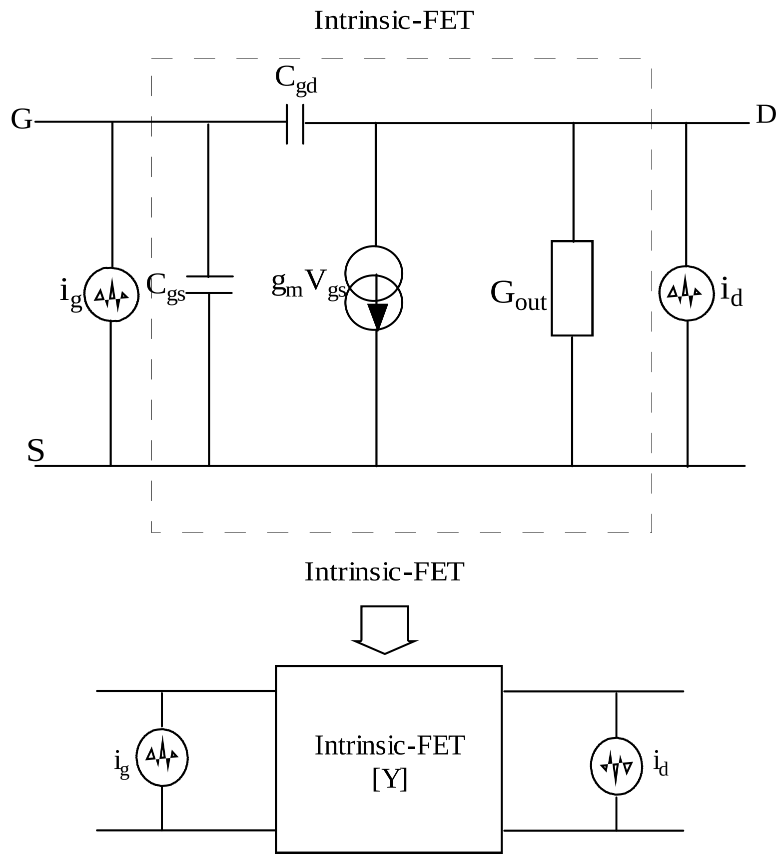
$f_b \rightarrow$ *cutoff frequency of base alone*

$f_e = 1/\left(2\pi C_{Te} r_e\right) \rightarrow$ *emitter cutoff frequency*

- R_n (Noise resistance) gives the sensitivity of the noise figure to the source admittance. The noise factor F is given as

$$F = F_{min} + \frac{R_n}{G_g}\left[\left(G_{opt} - G_g\right)^2 + \left(B_{opt} - B_G\right)^2\right]$$

FET Noise



$$[Y]_{FET-Intrinsic} = \begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix}$$

$$[C_Y] = [N]_{noise-matrix} = \begin{bmatrix} \overline{i_g i_g^{\bullet}} & \overline{i_g i_d^{\bullet}} \\ \overline{i_d i_g^{\bullet}} & \overline{i_d i_d^{\bullet}} \end{bmatrix}$$

$$[C_Y]_{FET} = 4kT \begin{bmatrix} \dfrac{w^2 c_{gs}^2 R}{g_m} & -jwc_{gs} C\sqrt{PR} \\ jwc_{gs} C\sqrt{PR} & g_m P \end{bmatrix}$$

- Intrinsic FET with noise sources at input and output

- Intrinsic FET has no noise contribution from a base spreading resistor from a transistor.
- The Key contributors are C,R,P C_{gs} and g_m

FET Noise

Where P,R and C are experimentally found FET noise coefficients:

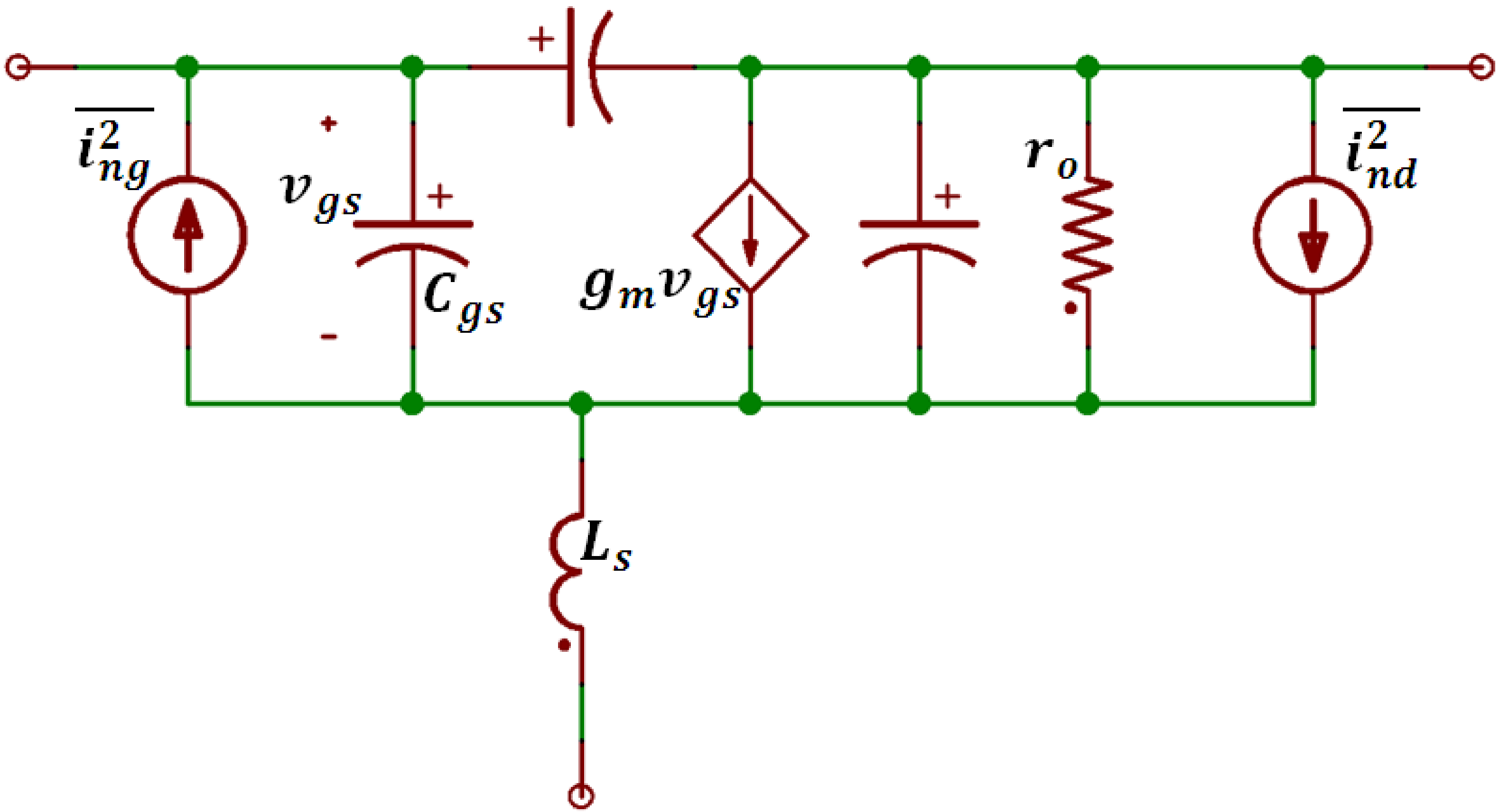
$$P = \left[\frac{1}{4kT_{g_m}}\right]\overline{i_d^2}\Big/Hz\ ; \qquad p = 0.67 \ \ for\ JFETs\ and\ 1.2 \ \ for\ MESFETs$$

$$R = \left[\frac{g_m}{4kT\omega^2 C_{gs}^2}\right]\overline{i_g^2}\Big/Hz\ ; \qquad R = 0.2 \ \ for\ JFETs\ and\ 0.4 \ \ for\ MESFETs$$

$$C = -j\left[\frac{\overline{i_g i_d^{\bullet}}}{\sqrt{\overline{i_d^2}\,\overline{i_g^2}}}\right]; \qquad C = 0.4 \ \ for\ JFETs\ and\ 0.6 - 0.9 \ \ for\ MESFETs$$

Note: C has no real part. It is only imaginary!

MOSFET noise



MOSFET noise

$$R_N = \frac{4kT\Delta f\left(\gamma g_{d0}\right)}{4kT\Delta f g_m^2} = \frac{\gamma g_{d0}}{g_m^2}$$

$$G_u = \frac{\delta\omega^2 C_{gs}^2\left(1-|c|^2\right)}{5g_{d0}}$$

$$\Gamma_{opt} = \omega C_{gs}\frac{g_m}{g_{d0}}\sqrt{\frac{\delta}{5\gamma}\left(1-|c|^2\right)} - j\omega C_{gs}\left(1+\frac{g_m}{g_{d0}}|c|\sqrt{\frac{\delta}{5\gamma}}\right)$$

$$F_{min} = 1 + \frac{2\omega}{\sqrt{5}\omega_T}\sqrt{\delta\gamma\left(1-|c|^2\right)}$$

MOSFET noise-Define c,γ and δ

$$\textit{Typically}$$

$$|c| = 0.395;\ \gamma = 2;\ \text{and}\ \delta = 4$$

$$\text{Correlation coefficient } c \equiv \frac{\overline{i_{ng} \cdot i_{nd}^{*}}}{\sqrt{\overline{i_{ng}^{2}} \cdot \overline{i_{nd}^{2}}}};$$

$$i_{ng} \rightarrow \text{gate current noise},\ i_{nd} \rightarrow \text{drain current noise}$$

$$\text{Gamma},\ \gamma = \frac{\Delta V_t}{\sqrt{2\phi_F - V_{BS}} - \sqrt{2\phi_F}};\quad \text{level} - 1 \text{ spice model}$$

$$V_t \rightarrow \text{Threshold voltage}\quad \phi_F \rightarrow \text{Fermi level depth}$$

$$V_{BS} \rightarrow \text{bulk} - \text{to} - \text{source voltage}$$

$$\text{delta},\ \delta = \sqrt{\frac{2\varepsilon_{Si}}{qNsub}\left(V_{ds} - V_{dsat}\right)};\quad \text{depletion layer's extent}$$

Why Do I need Amplifier in System Design

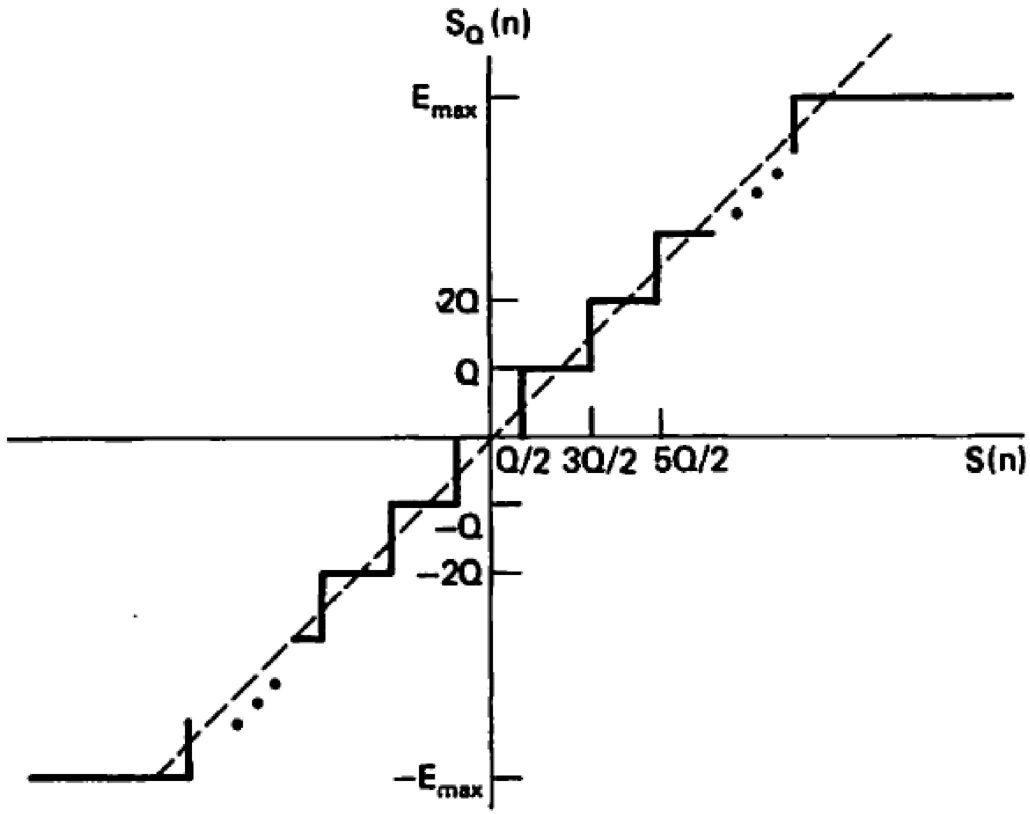
- To overcome quantization noise, caused by digitization of analog signal.
- We need a pre-amplifier stage for proper detection of small amplitude signals to overcome the mixer or A-to-D noise.
- The quantizing noise sounds like crackling of glass pieces under your feet on a concrete road.
- For an multiple stage amplifier the overall noise of the system is computed by Friis's formula as

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \cdots$$

Quantization noise in A/D convertors

- For example in a linear quantizer the input signal gets divided into series of steps each q volt, which is then coded to provide an digital output.
- An input voltage that falls between (2k+1)Q/2 and (2k+3)Q/2 is represented by an output kQ.
- The output voltage waveform is equivalent to input waveform plus random error e(t), varying between ±Q/2.
- The RMS voltage of e(t),Q/Sqrt(12) is quantizing noise and its spectrum is uniform.

Quantization noise-Define Q, k



- Linear quantizer input-output curve

Q is the quantizing level

- k is the sample number

Design cases for LNA's

- There are four possible amplifier configurations. These are:
- Grounded source/emitter
- Grounded gate/base
- Source follower (not very relevant)
- Hybrid or gate-source feedback circuits

Amplifier with Grounded Source/Emitter

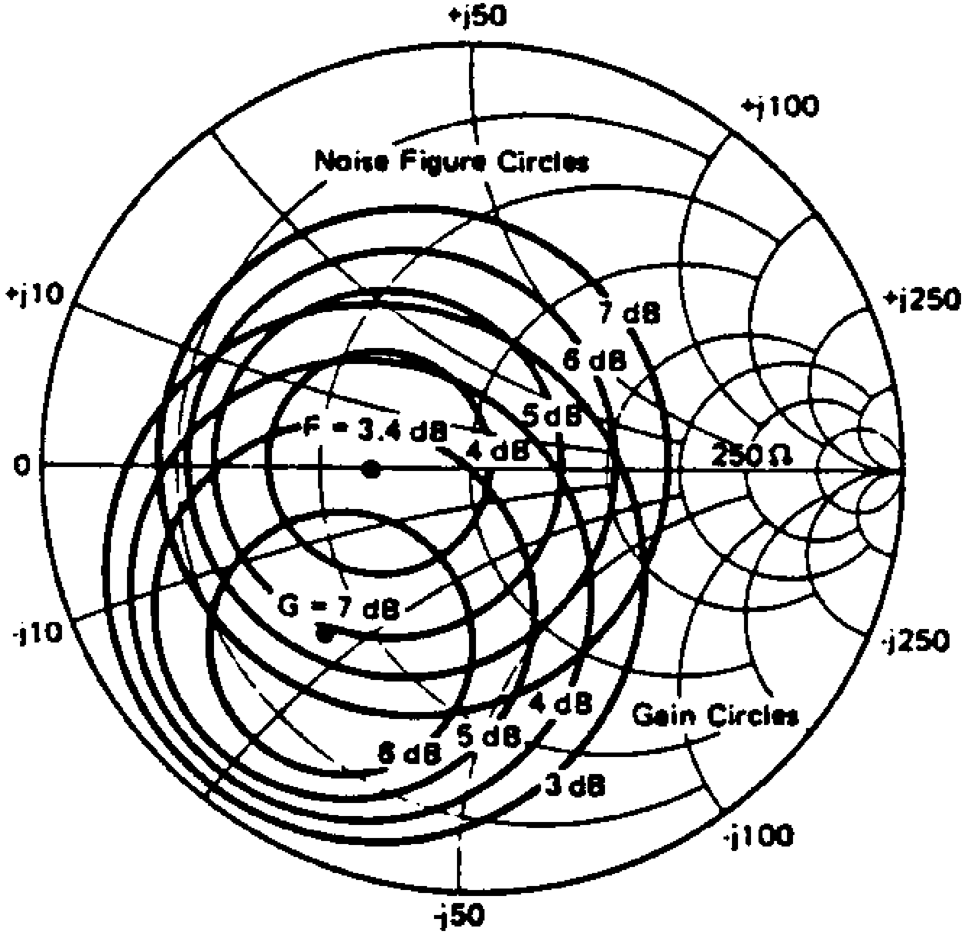
- Cannot be used for wideband impedance matching unless a negative feedback resistor is used.
- It has both voltage and current gain.
- High input impedance, bipolar transistor

$$Z_{in} = \beta . r_d = \frac{\beta * 26mA}{I_c}$$

$$Power\ Gain = gm^2 * Z_L . Z_s . N^2 * \left(\frac{\beta_0}{gm . Z_s + \beta_0} \right)^2$$

- The output impedance is in the vicinity of 400Ω (MESFET)
- Noise matching can be tricky as S11 is far apart from Γopt

Amplifier with Grounded Source/Emitter



$f = 4\text{GHz}$
$\Gamma_L = S_{22}$
$(S_{12} \neq 0)$

The center of the noise circle and the radius of the noise circle is computed using the formulae.

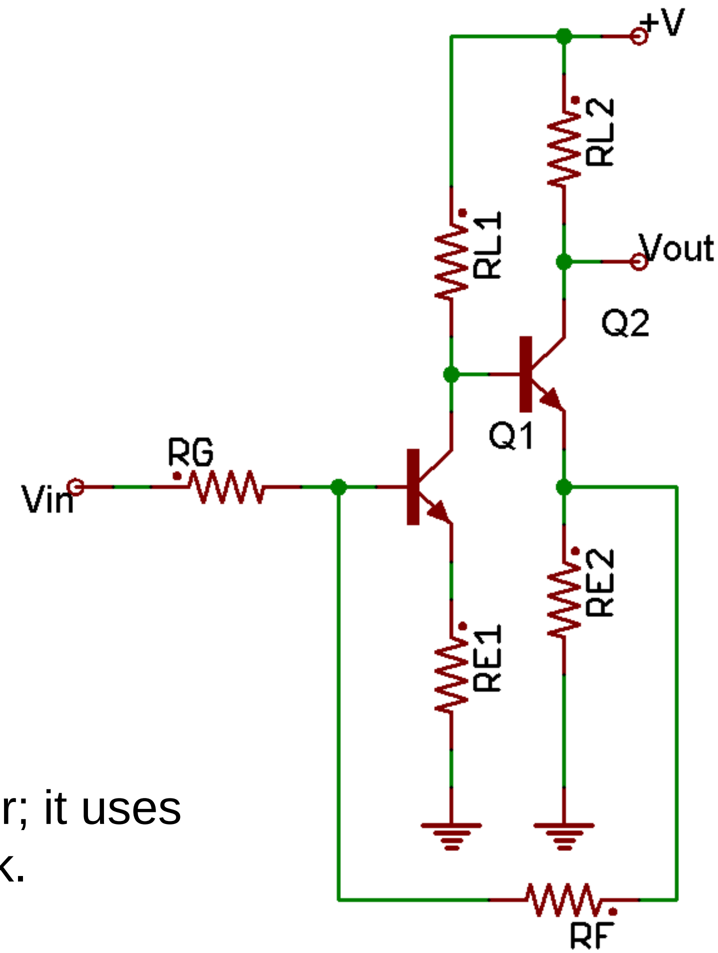
$$\Gamma_{on} = \frac{Z_{on} - Z_o}{Z_{on} + Z_o}$$

$$C_i = \frac{\Gamma_{on}}{1 + N_i}; \qquad N_i \to \text{input noise}$$

$$r_i = \frac{\sqrt{N_i^2 + N_i\left(1 - |\Gamma_{on}|^2\right)}}{1 + N_i}$$

Amplifier-Grounded Source/Emitter

- Also adding the source inductor may cause the circuit to be unstable.
- Fmin is same for all configurations.

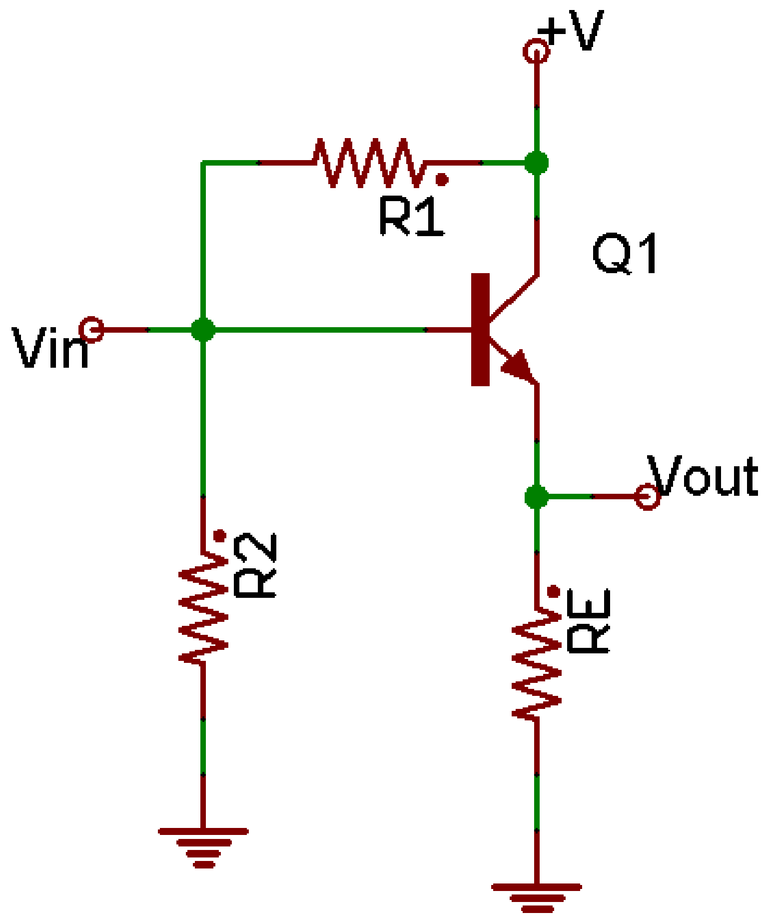


Two stage feedback amplifier; it uses voltage and current feedback.

Amplifier-Source/Emitter follower

- It has voltage gain of 0.99 ($\beta/(\beta+1)$)and a very high current gain.
- This amplifier is mainly used in Colpitts oscillator with added feedback because it maintains a constant phase shift and is more universal.
- But this configuration has fairly high impedance where Γopt is S11 conjugate.
- It is also know as Common Collector/Drain Configuration

Amplifier-Source/Emitter follower



Amplifier-Grounded Gate/Base

- It is used in simple isolation stages for systems. It offers constant wideband resistive impedance matching.
- It has power gain but no current gain.
- Input impedance is

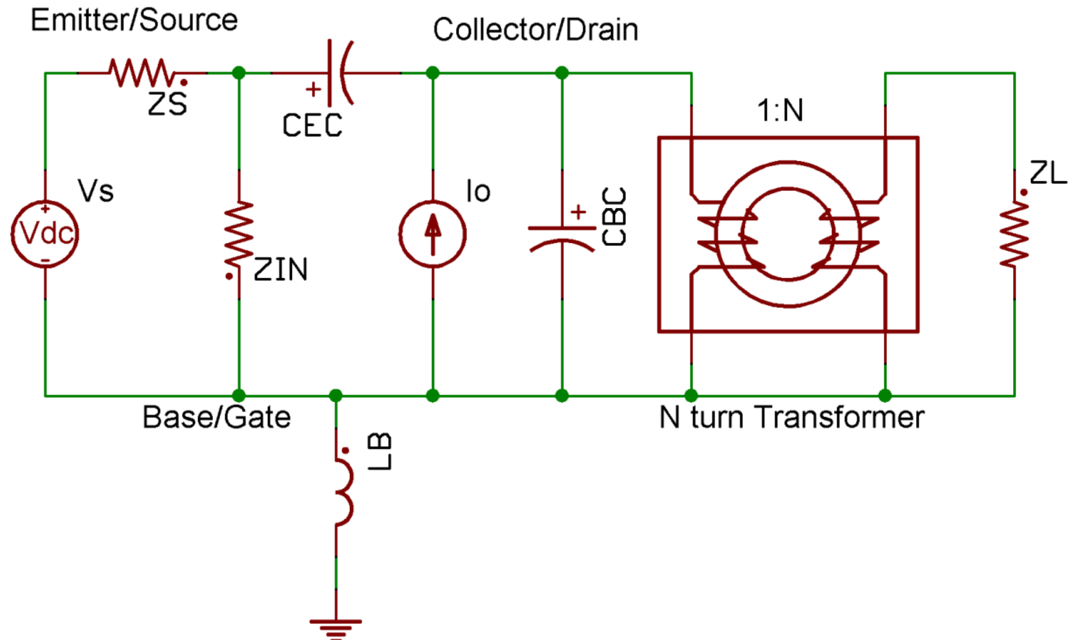
$$Z_{in} = \frac{1}{gm}$$

$$Power\ Gain = gm^2 * Z_L . Z_S . N^2 * \left(\frac{1}{gm.Z_S + 1} \right)^2$$

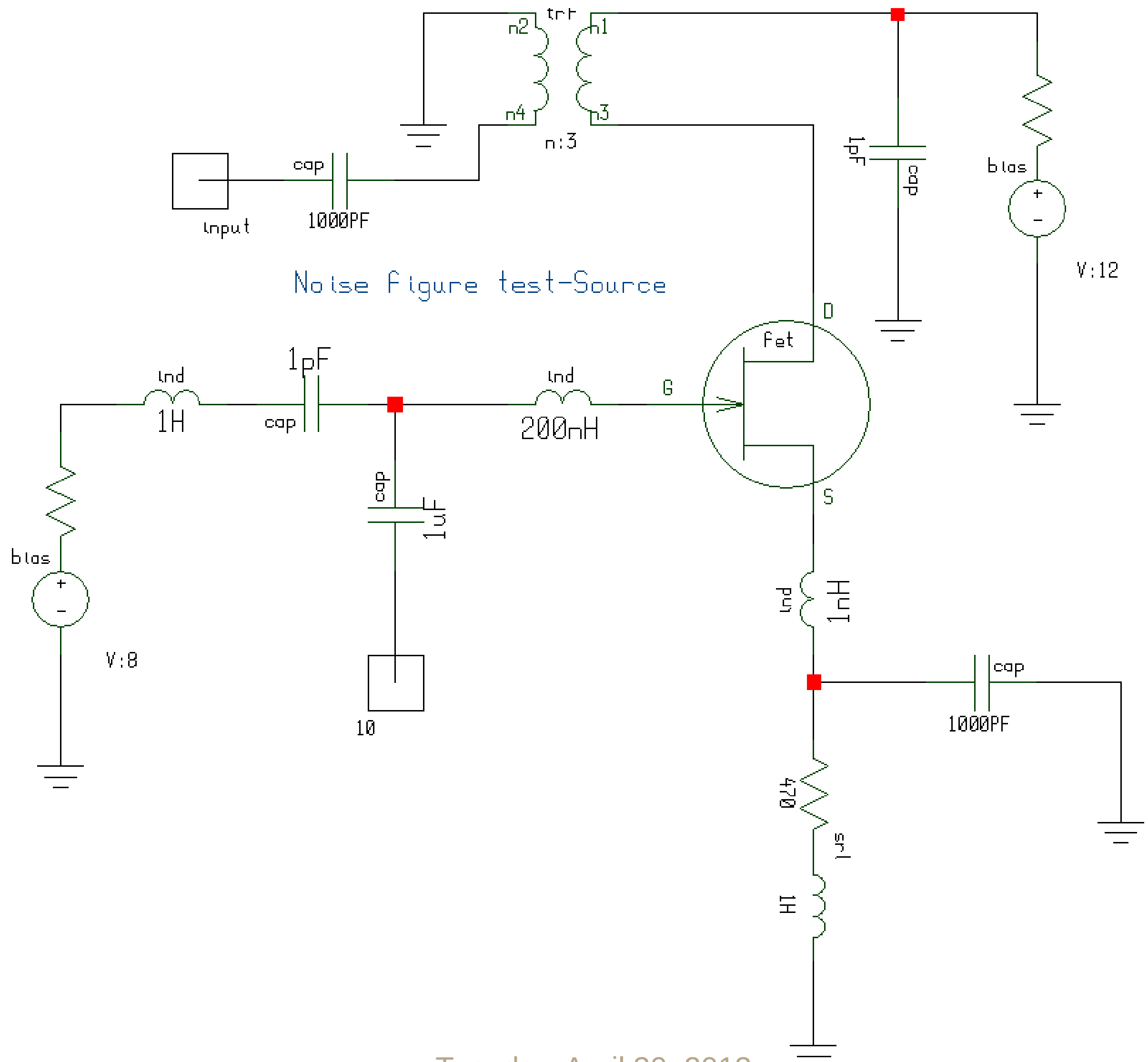
- High output impedance
- Input impedance varies by the amount of feedback caused by S12, if larger than zero.

Amplifier-Grounded Gate/Base

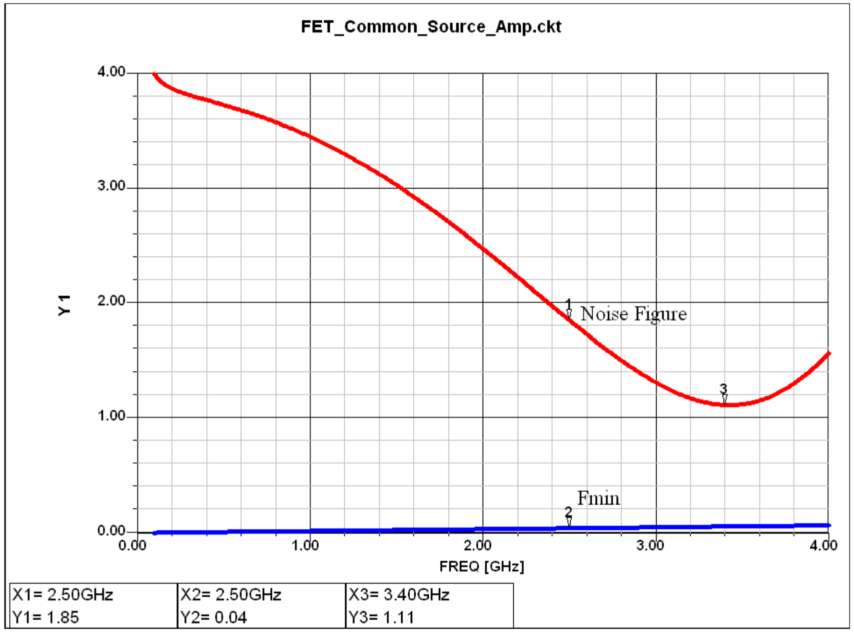
- Avoid building an amplifier with CB configuration. The lead inductance of base drives it in unstable region and makes it suitable for VCO.
- Decent gain and noise matching not practically possible.

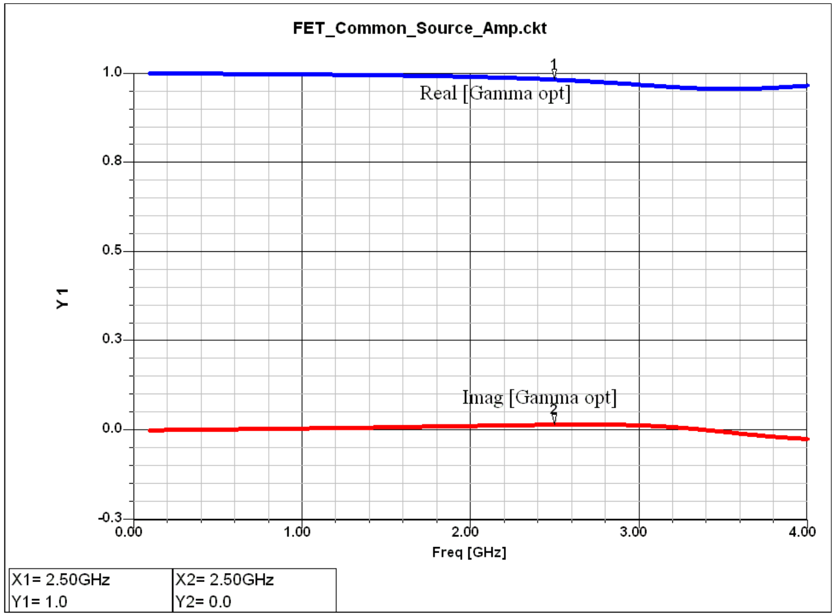


Common Source FET Amplifier

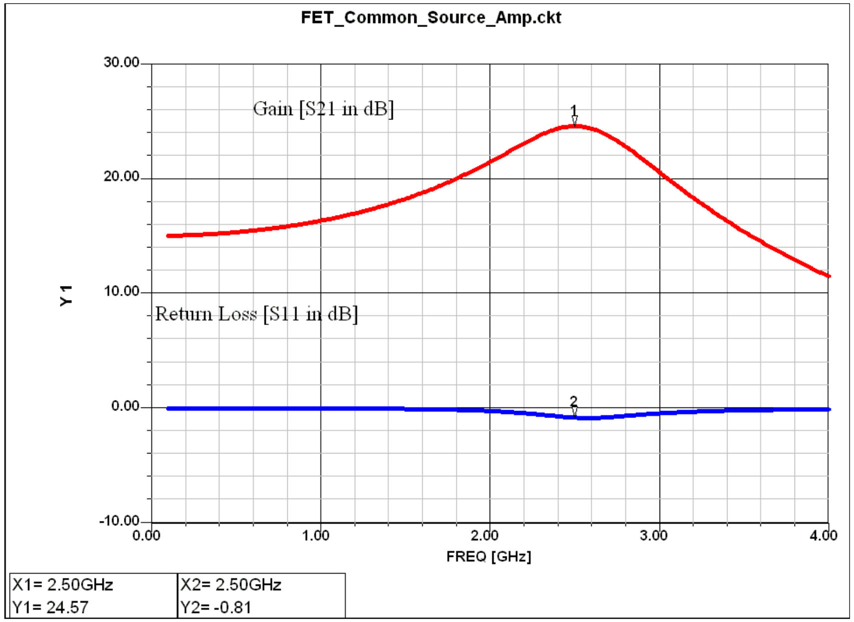


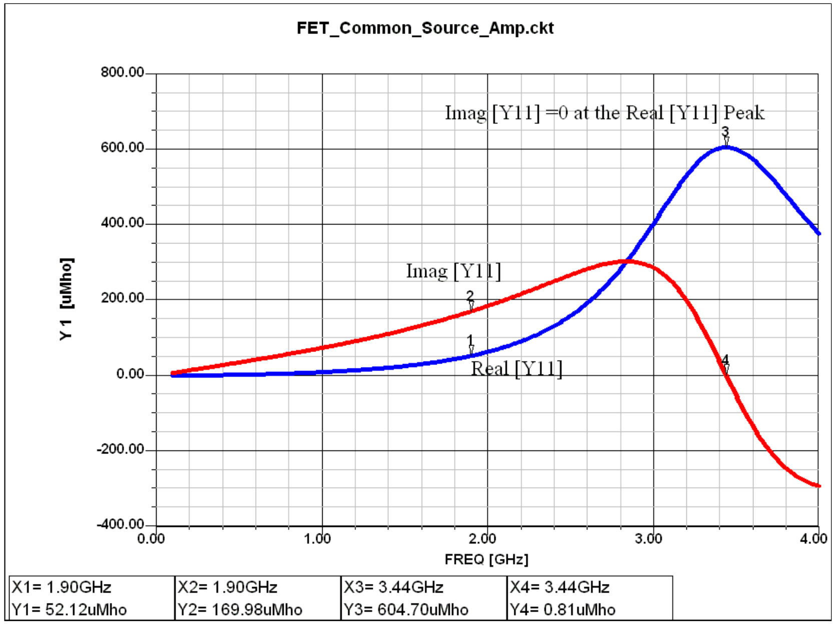
Noise figure, Minimum noise factor & Gamma opt.





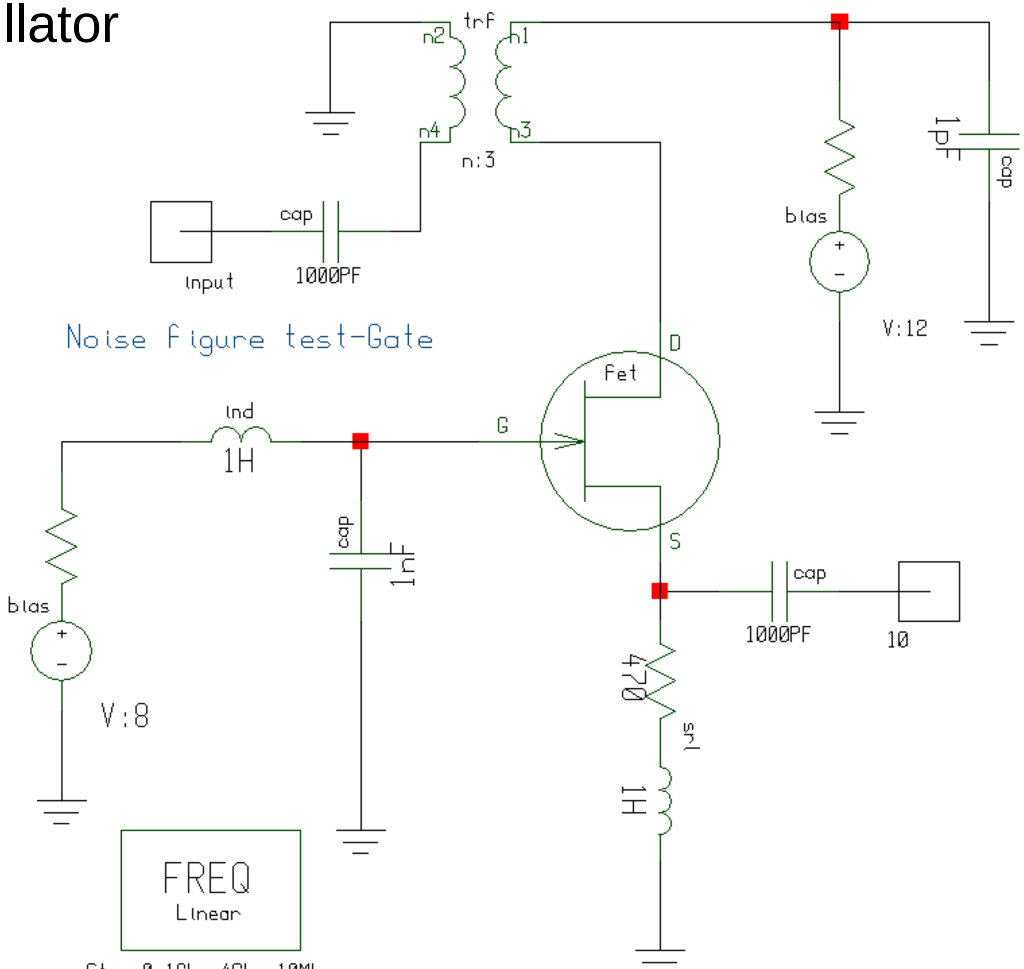
Gain, Return loss & Y11



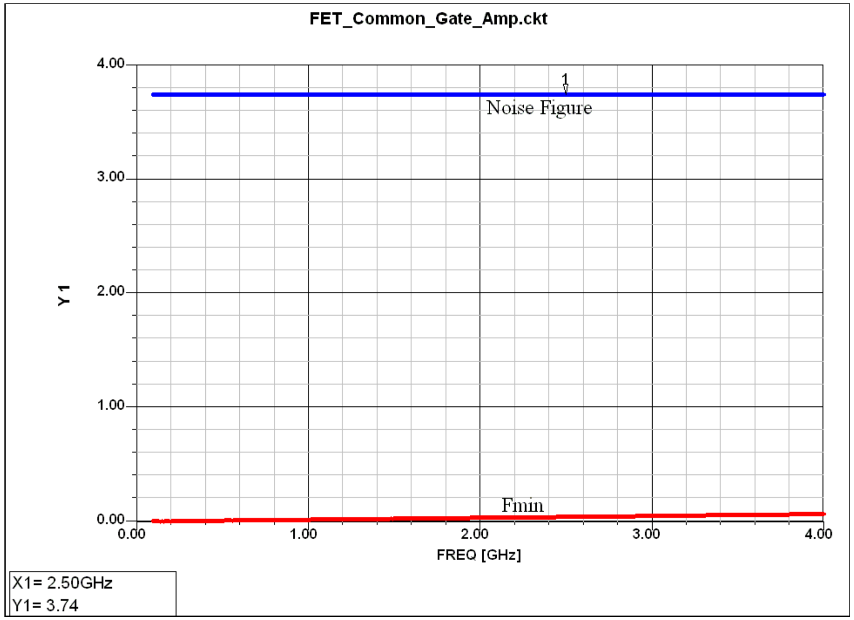


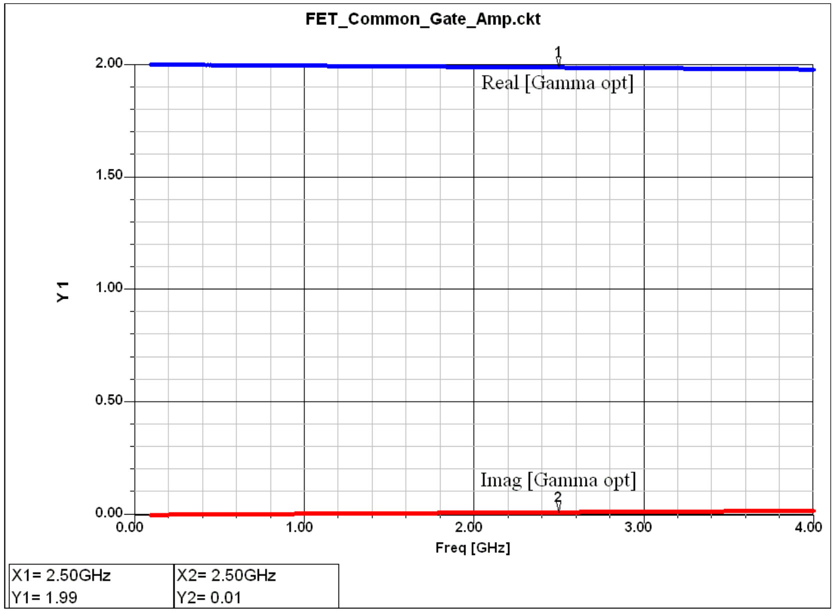
Common Gate FET Amplifier

- Surprise!! Placing the inductor at Gate and Source makes it an Oscillator

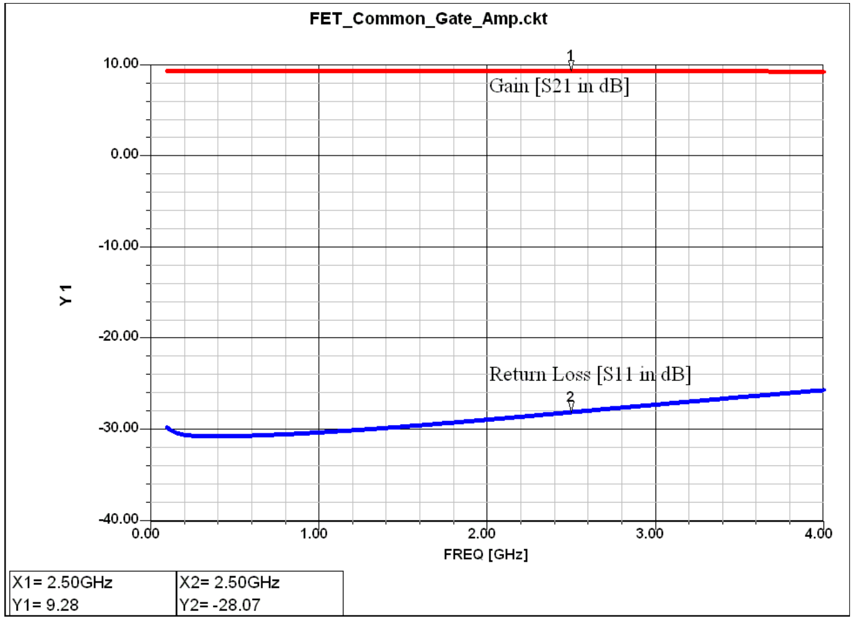


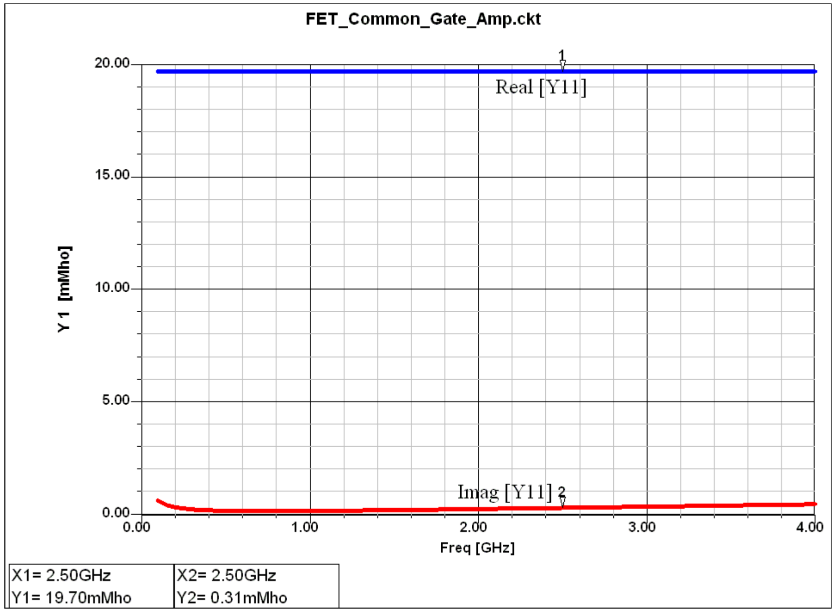
Noise figure, Minimum noise factor & Gamma opt.



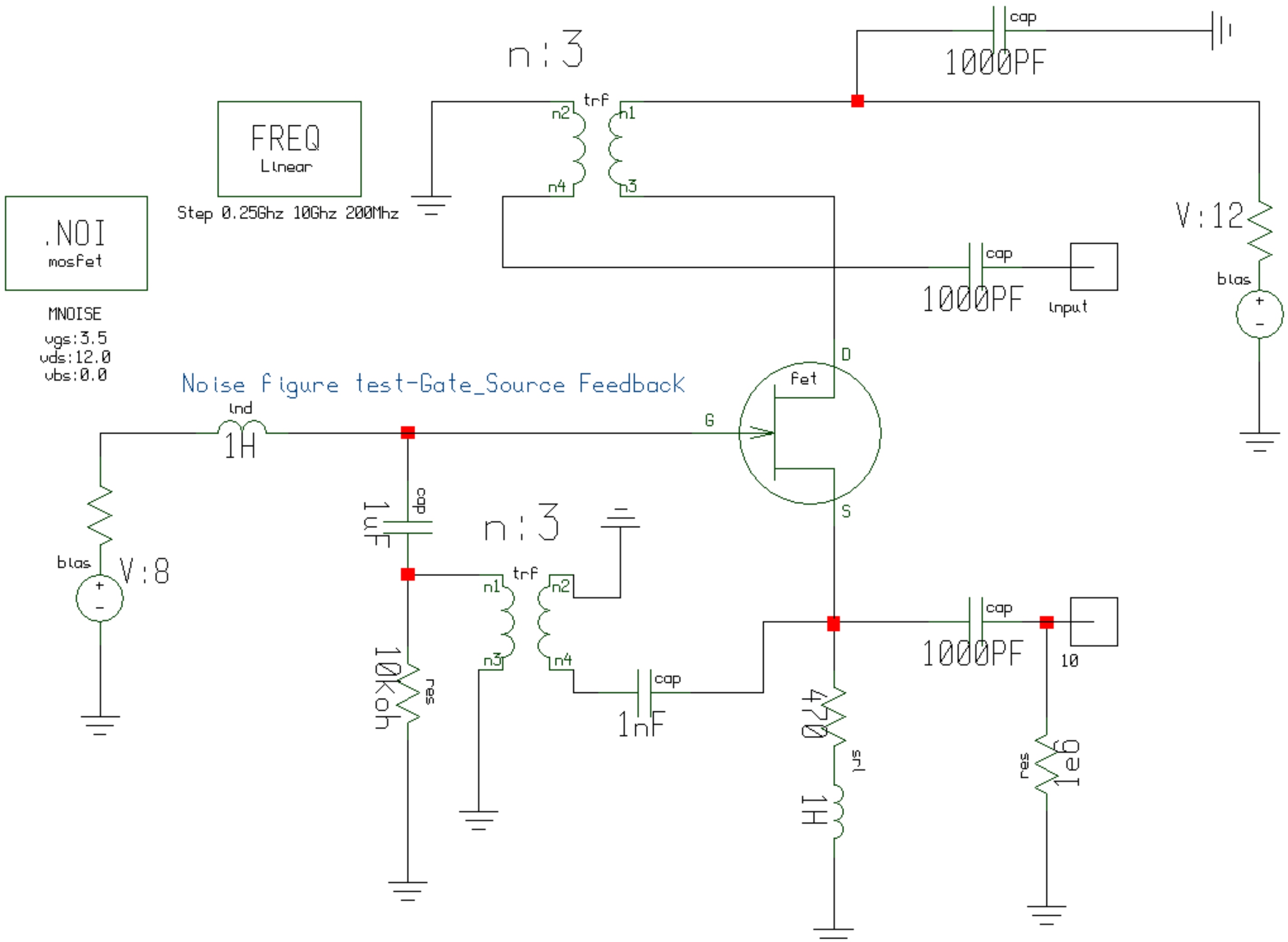


Gain, Return loss & Y11

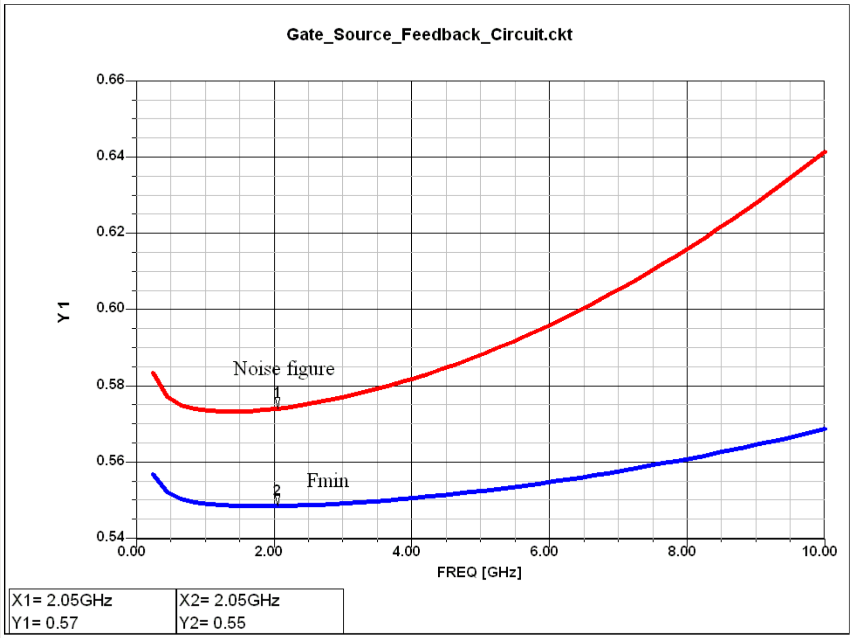


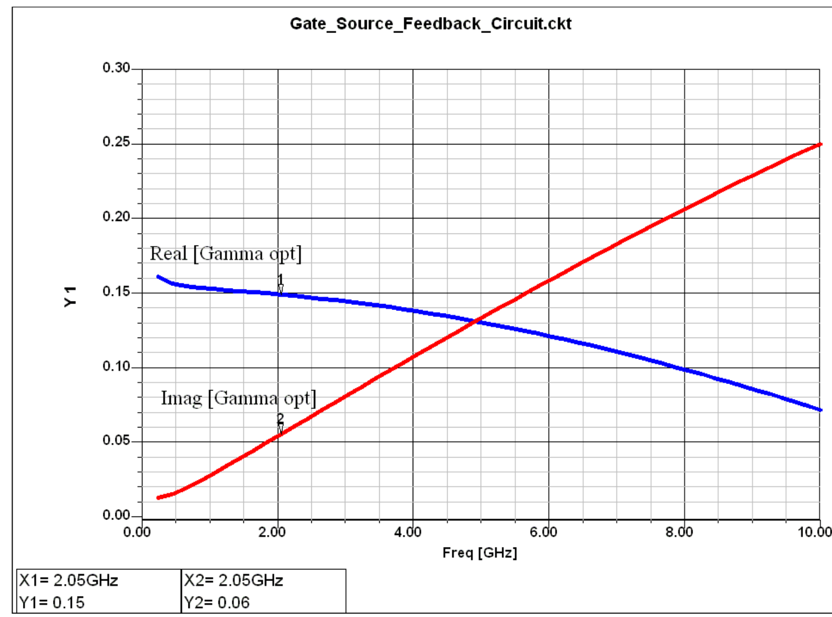


Novel Circuit: Gate-Source Feedback

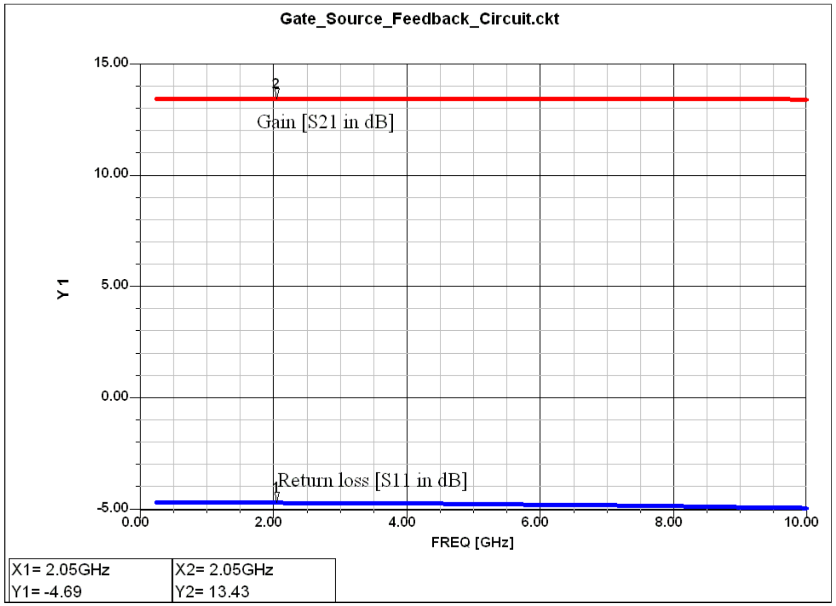


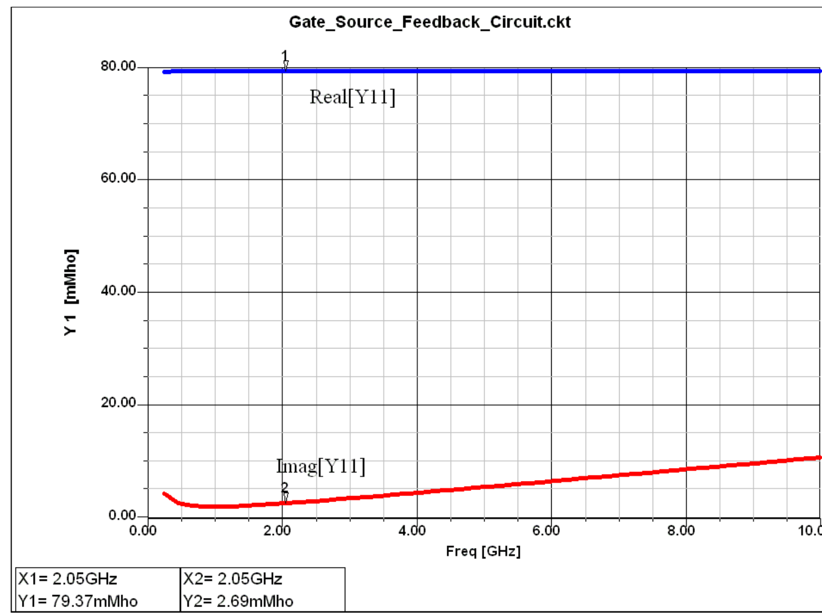
Noise figure, Minimum noise factor & Gamma opt.





Gain, Return loss & Y11





Large Signal Parameters

- Small Signal Parameters
- The RF signals are in the vicinity of micro-amps and micro-volts.

$$Y_{21}\Big|_{small-signal} = \frac{q.I_{DC}}{kT} = gm$$

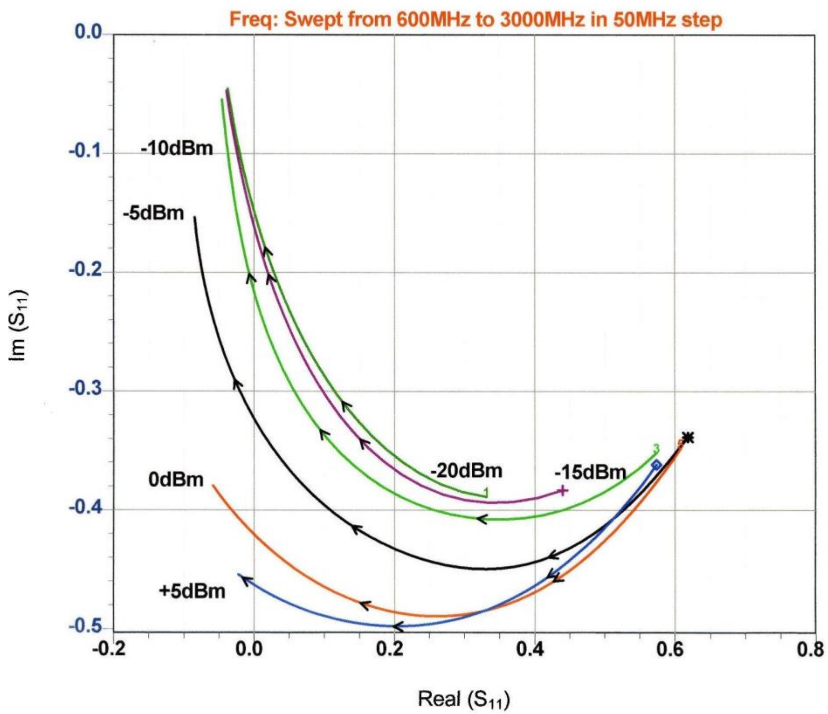
- Large Signal Parameters
- RF and DC voltage and currents are in the same range

$$Y_{21}\Big|_{l\,arg\,e-signal} = \left(\frac{q.I_{DC}}{kT}\right)\left(\frac{1}{x}\right)\left[\frac{2I_1(x)}{I_0(x)}\right]$$

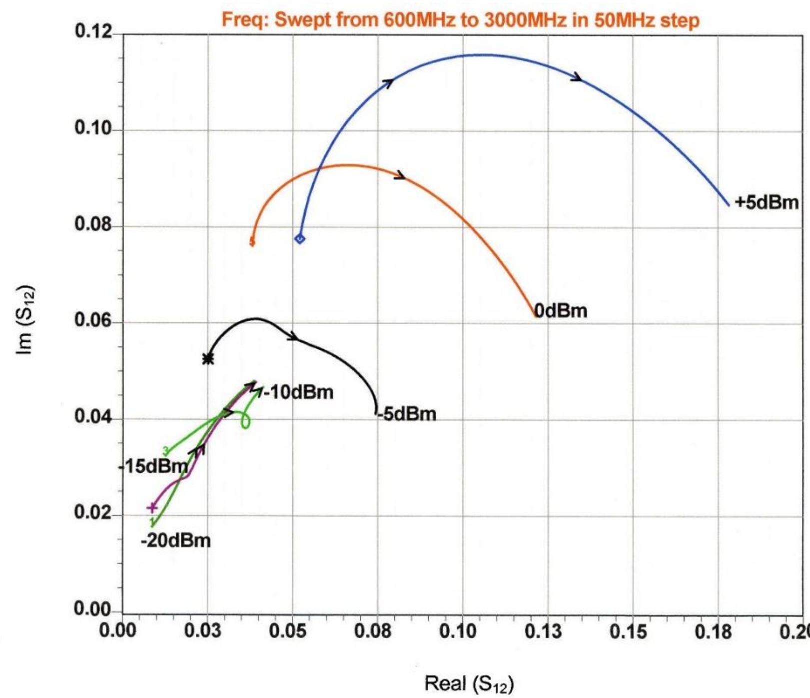
$$= \frac{gm}{x}\cdot\left[\frac{2I_1(x)}{I_0(x)}\right]$$

Do not confuse with small signal parameters

Typical example of measured large-signal parameters

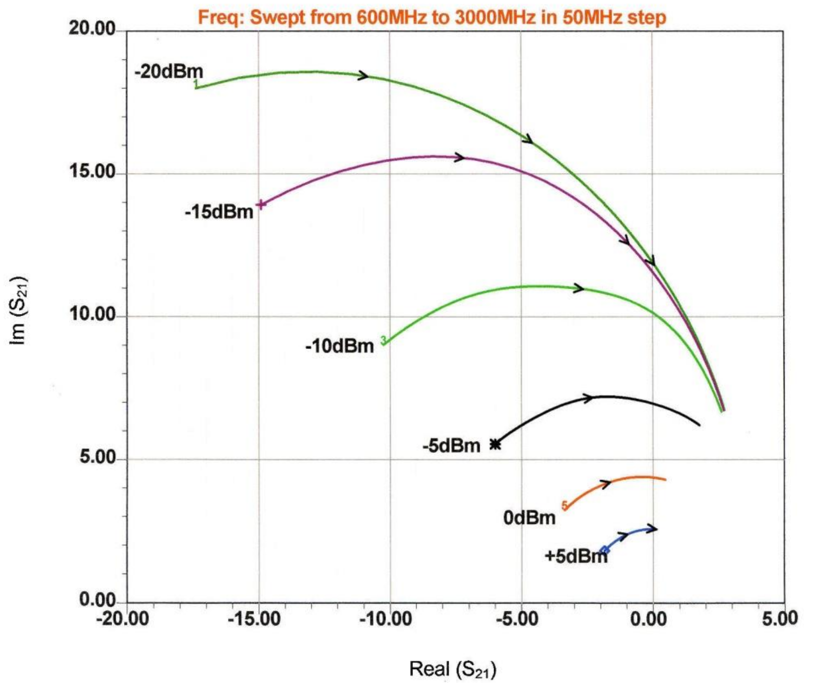


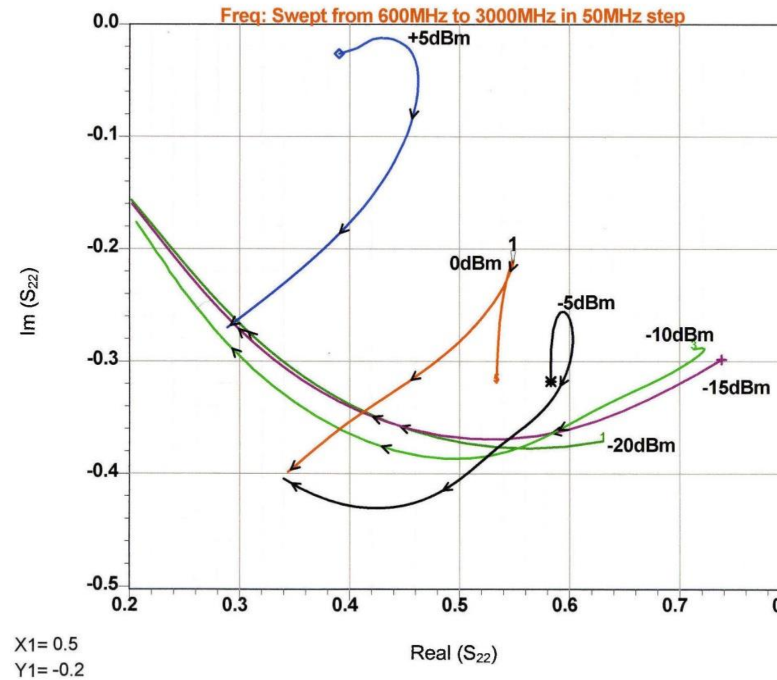
Measured large-signal S_{11} of the BFP520.



Measured large-signal S_{12} of the BFP520.

Typical example of measured large-signal parameters





Measured large-signal S_{21} of the BFP520.

Measured large-signal S_{22} of the BFP520.

Summary

- The novel approach presented using the gate-source feedback for best power matching and low noise design is so far the best ever demonstrated.

References

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