

Design of Low Noise Amplifiers

We have already studied amplifier design for

stability

gain

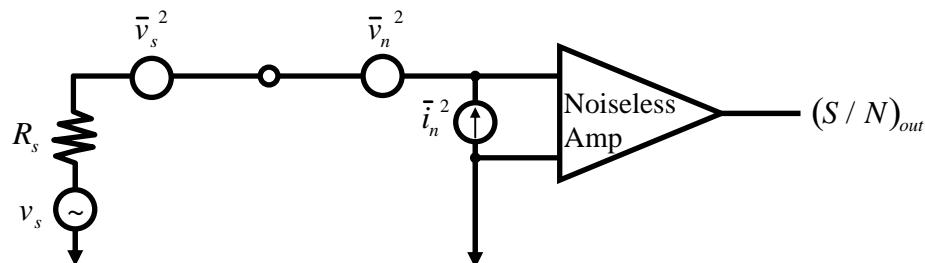
Now we will consider how to design for lowest noise.

Recall
$$F_{total} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \dots$$

- The noise factor of the first stage, F_1 , dominates the overall noise performance if G_1 is sufficiently high. Recall that these are ratios, not dB. Gains are available gains.
- We will see that maximum gain and minimum noise cannot be obtained at the same time. Noise performance is controlled by Γ_S . So, the design winds up being a compromise between gain and noise.

There are two techniques widely used.

1. Input-referred noise voltage and currents



all noise sources in the amplifier (devices, resistors) are combined to form input-equivalent voltage and current sources at the input.

We choose components and biasing in amp to minimize \bar{v}_n^2 and \bar{i}_n^2 . Then select R_S (or Z_S) to provide the highest $(S/N)_{OUT}$.

This technique is good for encouraging intuition. Similar to the use of

device model equivalent circuits. Widely used in low frequency circuits where \bar{v}_n^2 and \bar{i}_n^2 can be measured by input short and open circuits.

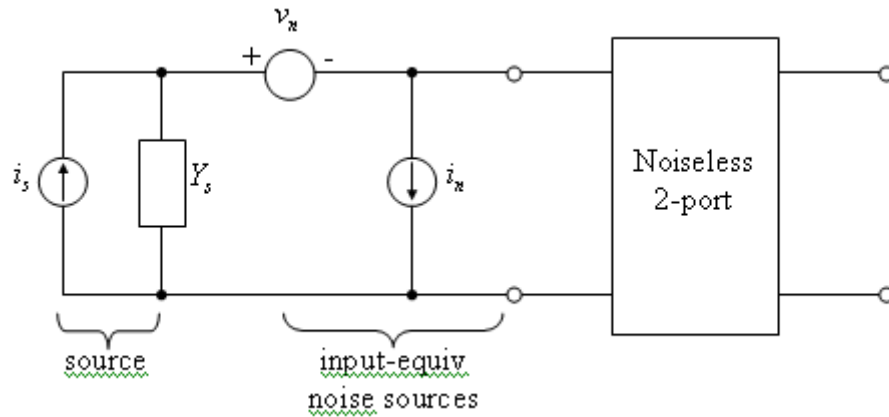
But, for the same reasons that we chose to use s-parameters for designing high-frequency circuits

what were they?

there is a more accurate technique for low noise design.

2. Two-port Noise Parameter representation

We will derive the optimum source admittance Y_s which when presented to the input of the transistor leads to the minimum degradation of S/N.



The total output noise power is proportional to \bar{i}_{sc}^2 , the mean-square input port short circuit current.

and:
$$F = \frac{\bar{i}_{sc}^2}{\bar{i}_s^2} = \frac{\text{noise power due to source and amp}}{\text{noise power of source alone}}$$

$$i_{sc} = -i_s + i_n + v_n Y_s$$

$$\bar{i}_{sc}^2 = \bar{i}_s^2 + \overline{(i_n + v_n Y_s)^2} - \underbrace{2i_s (i_{nt} v_n Y_s)}_{=0 \text{ since } i_s \text{ not correlated with } i_n \text{ or } v_n}$$

so:
$$F = 1 + \frac{\overline{(i_n + v_n Y_s)^2}}{\bar{i}_s^2}$$

here, $\bar{i}_s^2 = 4kT_o G_s B$ where G_s is the source conductance

and $\bar{v}_n^2 = 4kT_o R_N B$ where R_N is a fictitious equivalent noise resistance that represents \bar{v}_n^2 .

i_n can be represented by a correlated (with v_n) and uncorrelated part.

$$i_n = i_{nt} + i_{nc}$$

$$i_{nc} = Y_c v_n$$

(Y_c is a fictitious correlation admittance)

$$i_{nu} = 4kT_o G_u B$$

(G_u is a fictitious equivalent noise conductance)

To continue, we must minimize F .

(see Appendix L in Gonzalez for more detail)

This leads to the following:

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

where $Y_s = G_s + jB_s$ and $Y_{opt} = G_{opt} + jB_{opt}$

F_{\min} = lowest possible noise factor

Y_{opt} = optimum source admittance for minimum noise

Normalizing everything to Z_o leads to:

$$F = F_{\min} + \frac{r_n}{g_s} |y_s - y_{opt}|^2$$

Using reflection coefficients:

$$y_s = \frac{1 - \Gamma_s}{1 + \Gamma_s} \quad y_o = \frac{1 - \Gamma_{opt}}{1 + \Gamma_{opt}}$$

$$\text{thus, } F = F_{\min} + \frac{4r_n |\Gamma_s - \Gamma_{opt}|^2}{(1 - |\Gamma_s|^2)(1 + \Gamma_{opt})^2}$$

$F_{\min}, r_n, \Gamma_{opt}$ are noise parameters given by manufacturer or measured.

So what? Well, we see from the above that there is a minimum noise factor possible for a device, F_{\min} , that is achieved only when a particular reflection coefficient, Γ_{opt} is presented to the input. So, $\Gamma_s = \Gamma_{opt}$ leads to the minimum noise figure for the amplifier built with this transistor.

To measure the 2 port noise parameters:

1. Fiddle with Γ_s until min. F is observed.

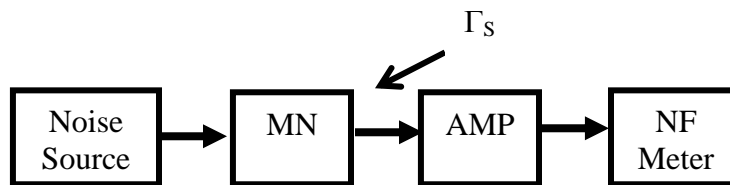
read F_{\min} from NF meter

use the network analyzer to measure $\Gamma_{opt} = \Gamma_s$

2. Set $\Gamma_s = 0$. Measure F .

(term. in Z_o)

$$r_n = (F_{\Gamma_s=0} - F_{\min}) \frac{|1 + \Gamma_{opt}|^2}{4|\Gamma_{opt}|^2}$$



Ok. But, this is equivalent to using equivalent input-referred noise voltages and currents (method 1) isn't it?

Yes, but the noise parameters in method 2 are measurable whereas in method 1 they are not at high frequencies.

Noise Figure Circles

We can also use the equation to predict noise figure vs. Γ_s . The equation can be reconfigured:

$$\frac{|\Gamma_s - \Gamma_{opt}|^2}{1 - |\Gamma_s|^2} = \underbrace{\frac{F_i - F_{\min}}{4r_n}}_{\text{constant for each } F_i} |1 + \Gamma_{opt}|^2 \equiv N_i$$

Then circles can be drawn on the source Γ_s Smith chart that correspond to a particular noise figure. When these noise circles are plotted with available gain circles, you can show the tradeoff between min. noise and gain.

To calculate noise circles:

Define:

$$N_i = \frac{F_i - F_{\min}}{4r_n} |1 + \Gamma_{opt}|^2 = \text{noise figure parameter}$$

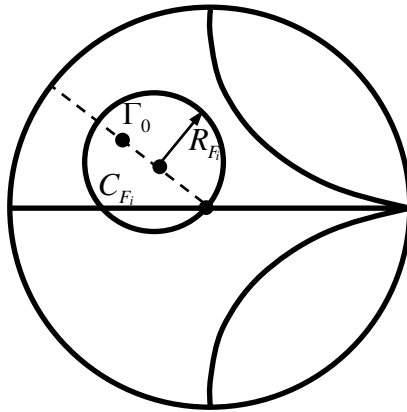
Rearrange the equation so that

$$|\Gamma_S - C_{Fi}| = r_{Fi}$$

where:

$$C_{Fi} = \frac{\Gamma_{opt}}{1 + N_i} \quad (\text{center})$$

$$r_{Fi} = \frac{1}{N_i + 1} \sqrt{N_i^2 (1 - |\Gamma_{opt}|^2)} \quad (\text{radius})$$



Note that $r_{Fi} = 0$ when $\Gamma_S = \Gamma_{opt}$

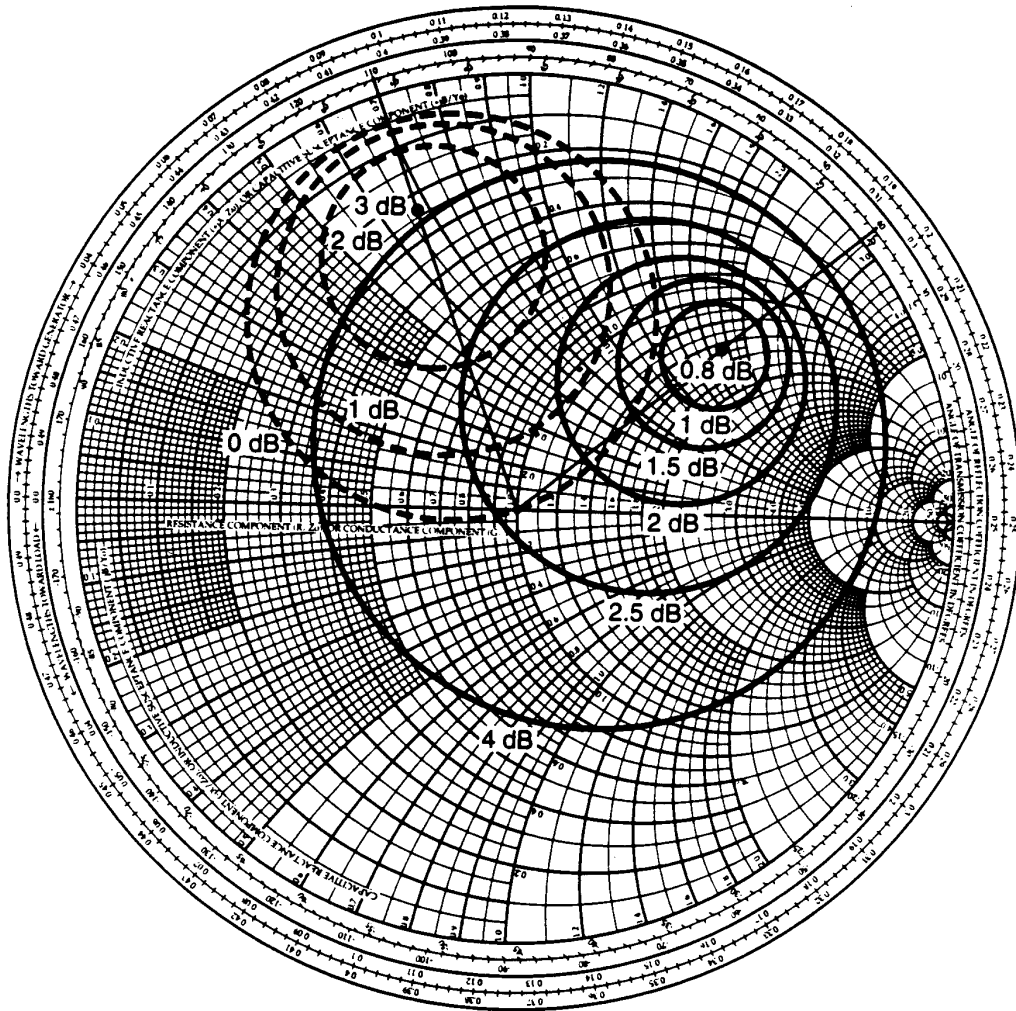


Figure 4.3.3 Noise figure circles (solid curves) and G_s constant-gain circles (dashed curves). The transistor is a GaAs FET with $V_{DS} = 4$ V, $I_{DS} = 12$ mA, and $f = 6$ GHz.

From: G. Gonzalez, *Microwave Amplifiers: Analysis and Design*, Second Ed., J. Wiley, 1997

Available gain circles


Device data sheets often plot G_T on the Γ_S plane under the assumption that the output is matched ($\Gamma_L = \Gamma_{out}^*$). This is the available power gain = G_A .

$$G_A = \frac{P_{AVN}}{P_{AVS}}$$

Since output is always matched, and Γ_{out} depends on Γ_S , G_A is independent of Γ_L .

$$G_A = \frac{1 - |\Gamma_S|^2}{|1 - S_{11}\Gamma_S|^2} |S_{21}|^2 \frac{1}{1 - |\Gamma_{OUT}|^2}$$

(since $\Gamma_L = \Gamma_{OUT}^*$)



Depends upon input match because actual power absorbed in the input is not necessarily the same as Γ_{AVS} (unless conjugately matched at input).

If input is also conjugately matched, then we get the maximum available gain:

$$\begin{aligned} G_A &= G_{A,\max} = MAG = G_{T,\max} \\ &= \frac{1}{(1 - |S_{11}|^2)} |S_{21}|^2 \frac{1}{1 - |\Gamma_{OUT}|^2} \\ &\qquad\qquad\qquad \frac{1}{1 - |S_{22}|^2} \text{ (if unilateral)} \end{aligned}$$

MRF901

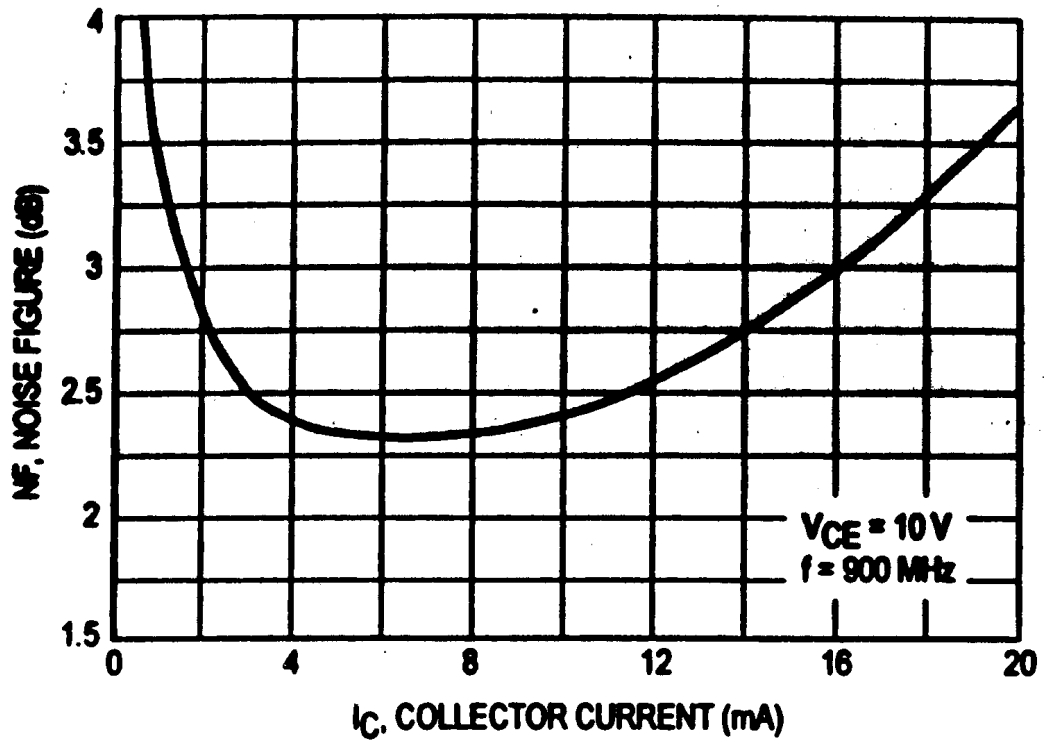


Figure 12. Noise Figure versus Collector Current

You can see that noise figure is often dependent on the bias point selection for the transistor. The manufacturer should provide noise parameters for several bias conditions, as well as plots such as this one.

Using ADS to simulate Noise Figure

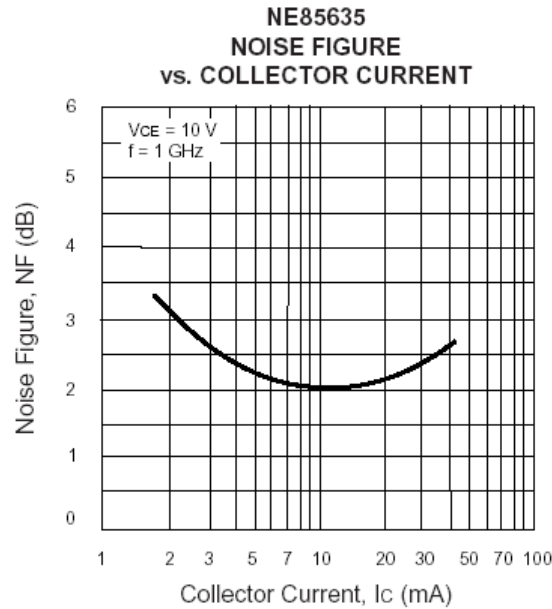
ADS can be used to design low noise amplifiers much in the same way you have already used it for MAG or MSG designs. Noise circles and available gain circles are the tools that give the most guidance on design tradeoffs. Refer to Chap. 4 and Appendices K and L of Gonzalez for the theory behind these analyses. The ADS files shown below are available from the course web page as a .zap file.

Here are 3 cases that you might encounter with device models when analyzing a low noise amplifier.

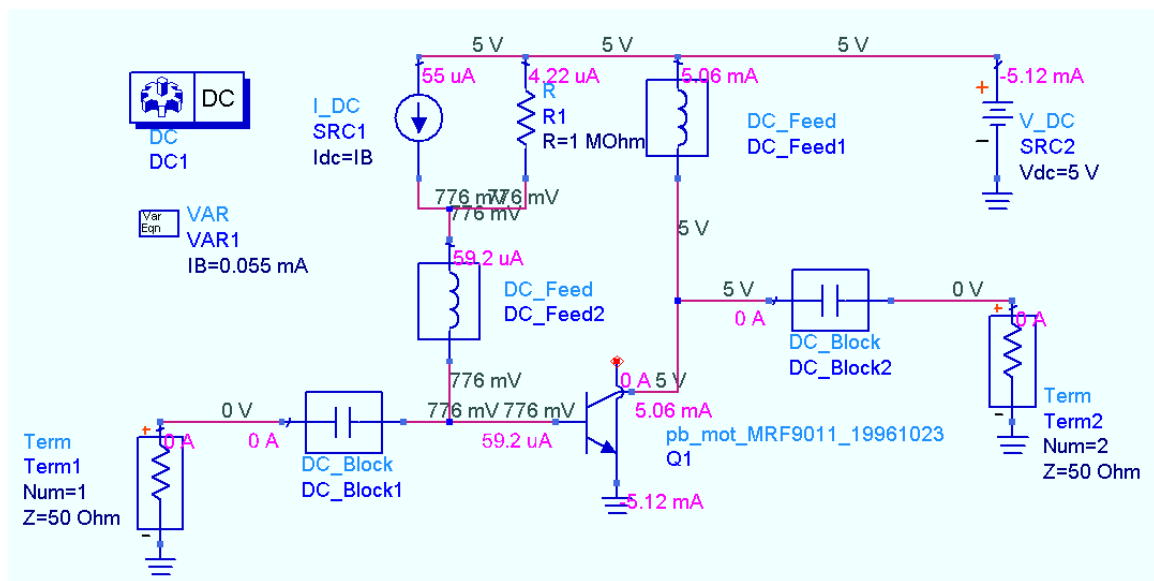
1. The ADS large signal transistor model is used to represent the device. This is the ideal case, but unfortunately, the large signal models sometimes do not produce accurate S parameters. If you use this model, you should check the simulated S parameters with the manufacturer's data sheet to verify that it provides reasonably accurate results. It should be useful for DC simulations however.
2. The ADS S-parameter transistor model is used to represent the device. This is the most accurate case. Of course, no DC simulations will be possible with this model, but it will represent the S parameters and noise parameters accurately. Be sure to select the model that represents the actual bias condition to be used in your analysis.
3. No S parameter model is available for the device in the ADS library. In this case, check the manufacturer's web site and download an S2P file for the device. Place an S2P block from the Data Items menu in your schematic and identify the file name. ADS will look for the file in the ADS project's data directory. See the appendix at the end of this tutorial for more information on S2P files. In this case, you may have to enter the noise parameters as well using equations on the data display panel.

Case 1. Using ADS large signal model library.

First you must bias the transistor. Let's say $V_{CE} = 5V$ and $I_C = 5\text{ mA}$. This bias condition might be selected from the plot of noise figure vs. bias condition provided by the manufacturer.

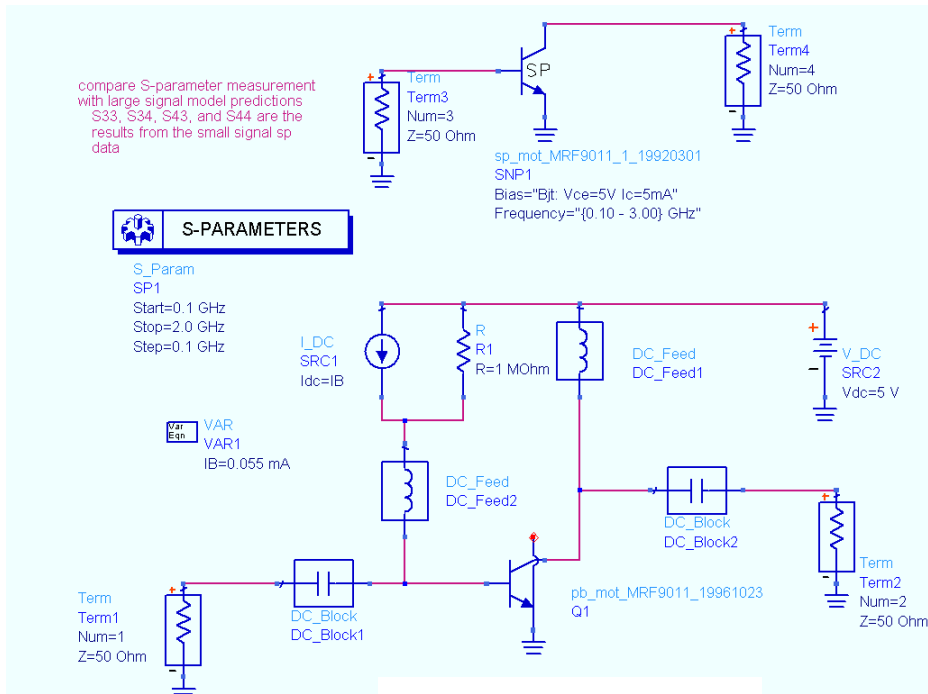


Set up a biasing circuit such as the one below. Select a large signal device model from the Analog/RF – RF Transistor/Packaged BJT library. Then perform a DC simulation. To see the results of the DC simulation, you go to the Simulate Menu > Annotate DC solution. Sometimes it is helpful to move component text aside so that the annotation is easier to read. Use F5 to move text.

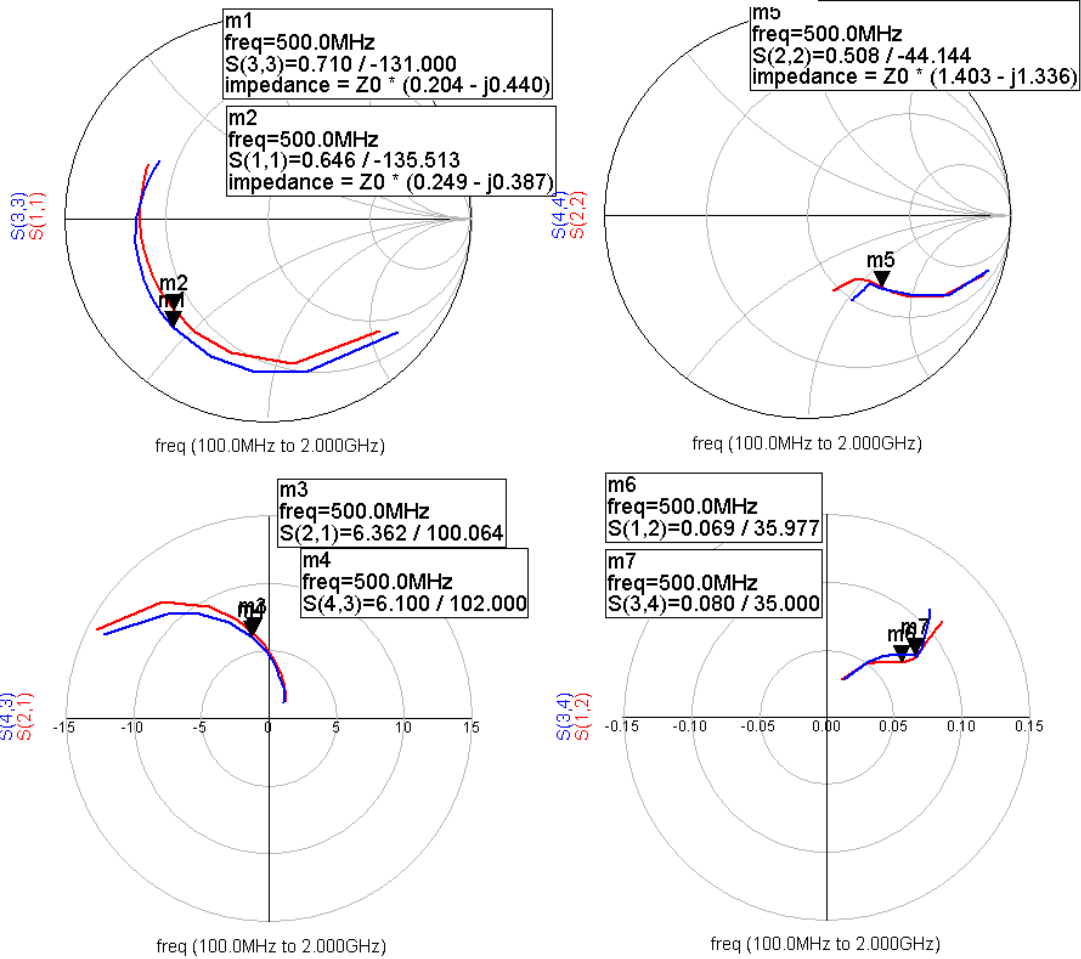


This is the result. We have $V_{CE} = 5V$ and $I_C = 5.06\text{ mA}$. (MRF901_DC schematic file)

Next, compare the S parameters from the large signal model with those from the small signal measured S parameter model. This model can be found in the Analog/RF S Parameter component library. Simulate over the specified frequency range.



(MRF901_sparamtest)



From this simulation, we can see that the agreement at our design frequency of 500 MHz is fairly good. S12 is off by about 20%, which will have an effect on the predictions of MSG, but the other parameters fit pretty well at this frequency. Let's assume this is sufficient. Now we must test for stability.

S-PARAMETERS Noise, stability, and gain calculation from large signal model.

S_Param
SP1
CalcNoise=yes
Freq=500 MHz

If the device model includes noise data, you need to enable the noise simulation in the S-Parameters controller.

OPTIONS

Options
Options1
Temp=16.85
Tnom=25

The standard temperature for calculation of noise figure is 290K (16.85 C)

Term1
Num=1
Z=50 Ohm

DC_Block1
DC_Block2

Q1
pb_mot_MRF9011_19961023

Term2
Num=2
Z=50 Ohm

VAR
VAR1
IB=0.055 mA

MeasEqn
meas5
MSG=10*log(mag(S21)/mag(S12))

MeasEqn
Meas1
mag_delta=mag(S11*S22-S21*S12)

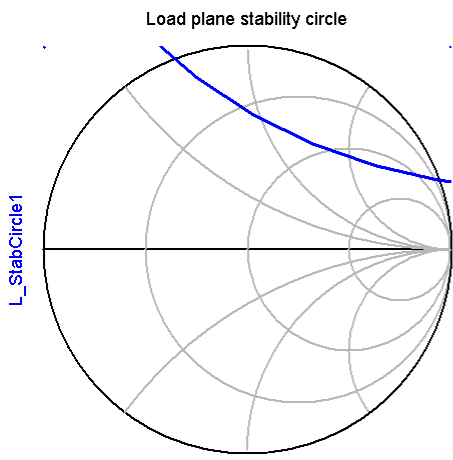
MaxGain
MaxGain1
MaxGain1=max_gain(S)

S_StabCircle
S_StabCircle1
S_StabCircle1=s_stab_circle(S,51)

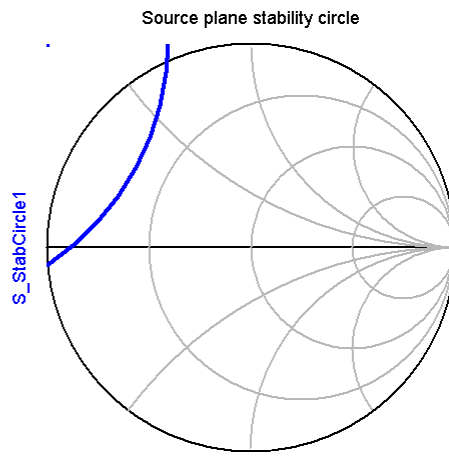
L_StabCircle
L_StabCircle1
L_StabCircle1=l_stab_circle(S,51)

StabFact
StabFact1
StabFact1=stab_fact(S)

MRF901_large_sig



indep(L_StabCircle1) (0.000 to 51.000)



indep(S_StabCircle1) (0.000 to 51.000)

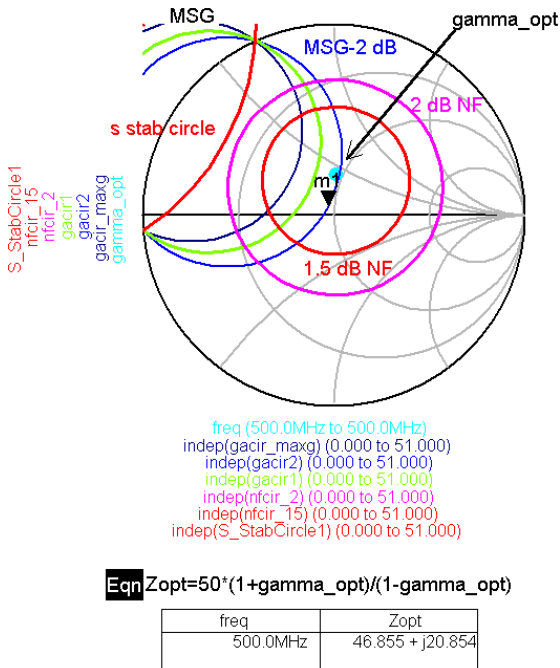
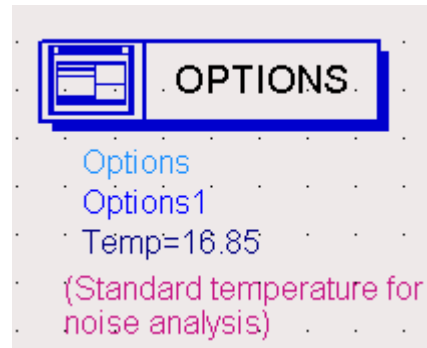
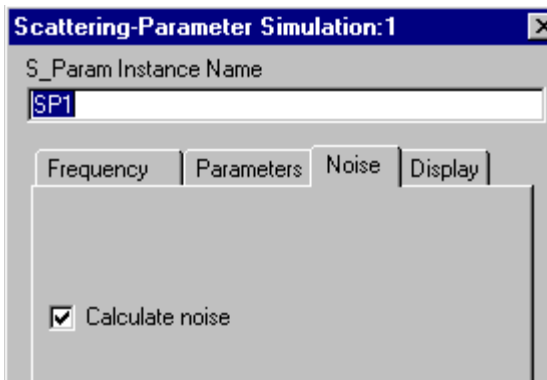
freq	mag_delta	StabFact1
500.0MHz	0.299	0.485

freq	MSG	MaxGain1
500.0MHz	19.769	19.769

(Display file: stability.dds)

We can see that the device is potentially unstable, but if we are careful and can give up some gain, we may be able to find a stable solution and still retain low noise figure. This may give us lower noise figure than if we first make the device unconditionally stable with resistive loading. To evaluate this possibility, we need to look at the available gain and noise figure circles. It is more convenient to calculate the gain and noise circles (Available gain for noise calculations) on the display panel rather than the schematic panel so you can change the noise figure without having to resimulate the circuit.

ADS can be used to calculate the noise parameters from the large signal transistor models. You must first enable the S Parameter controller icon for noise simulation. You will generally want to calculate noise parameters at one frequency, so use the Single Point Sweep option on the frequency panel. Also choose Options and set the temperature to 16.85 degrees (standard temperature for noise is 290K).



```

m1
indep(m1)=47
gacir2=0.052 / 122.559
freq=500.0000MHz
impedance = Z0 * (0.942 + j0.083)
    
```

freq	NFmin	Rn	Sopt
500.0MHz	1.167	7.560	0.213 / 86.426

freq	MaxGain1	MSG
500.0MHz	19.769	19.769

```

calculate available gain circles relative to MSG
or MAG:
Eqn gacir_maxg=ga_circle(S,MaxGain1,51)
Eqn gacir1=ga_circle(S,MaxGain1-1,51)
Eqn gacir2=ga_circle(S,MaxGain1-2,51)
Eqn gacir3=ga_circle(S,MaxGain1-3,51)
    
```

```

Eqn gamma_opt=Sopt
calculate noise circles at 1.5 and 2.0 dB
Eqn nfcir_15=ns_circle(1.5,NFmin,gamma_opt,Rn/50,51)
Eqn nfcir_2=ns_circle(2,NFmin,gamma_opt,Rn/50,51)
    
```

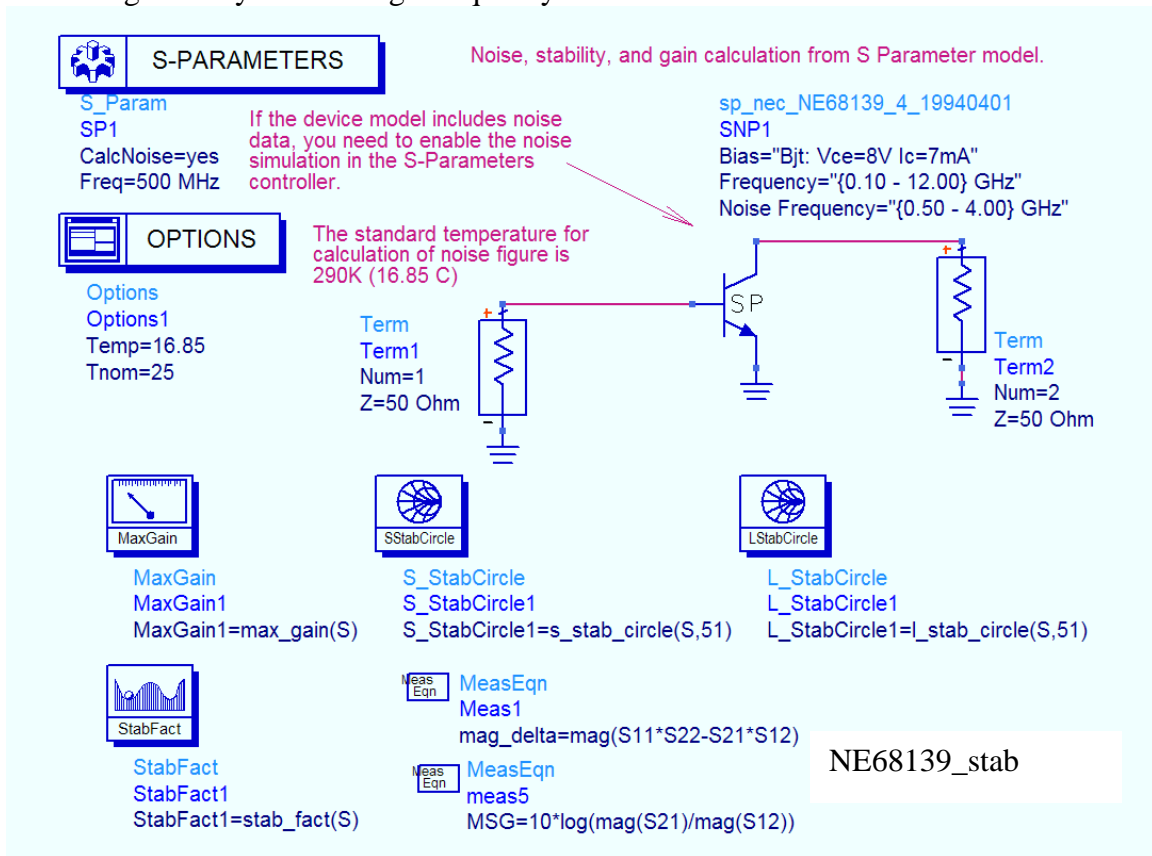
(Display file: circles.dds Use page menu to select gain and noise circles page)

The syntax used in the available gain and noise circle equations will be discussed in Section 2. However, you can see in this example, that the MSG is high (19.8 dB), and if we give up 2 dB in gain, a perfect noise source match can be achieved. The noise parameters for the transistor under the given bias condition are shown in the table. These were calculated from the large signal model.

Now, you could go back and vary the bias to determine whether you could improve on gain or noise, or you could complete the design by determining matching networks, biasing and wideband stability. Completing the design follows the same procedure described below in Section 2.

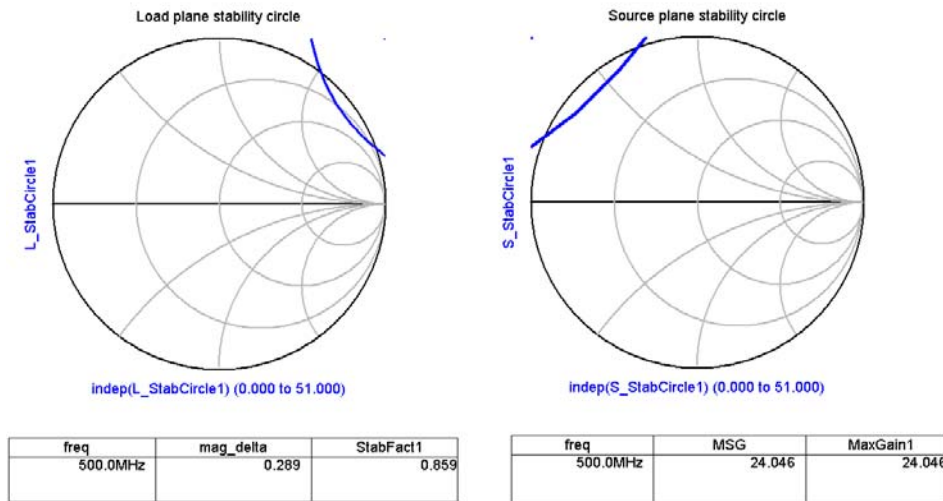
2. Simulation from S Parameter Model Library

The next example illustrates how the noise figure simulations can be carried out using the S Parameter Model Library in ADS. Choosing a different device, we start again by evaluating stability at the design frequency of 500 MHz.



If the device model includes noise data as this one does in the example above, enable the noise simulation in the S Parameter controller. This will then calculate NF_{min} , Γ_{opt} (called S_{opt}), and R_n (un-normalized). Also include the maximum gain (MaxGain1) equation (calculates either Maximum Available Gain MAG or Maximum Stable Gain, MSG) and stability circle equations on the schematic panel. You must first determine stability before proceeding with the noise analysis. Since this is a low noise design, we

will try first for a solution that does not require resistive stabilization which would degrade the noise figure.



(Display file: stability.dds)

We can see that the device is potentially unstable, but if we are careful and can give up some gain, we may be able to find a stable solution and still retain low noise figure. To evaluate this, we need to look at the available gain and noise figure circles. The noise parameters needed for noise figure calculation are not included in the S Parameter model library, so they must be added to the display panel. It is more convenient to calculate the gain and noise circles (Available gain for noise calculations) on the display panel rather than the schematic panel so you can change the noise figure without having to resimulate the circuit.

The syntax for calculating available gain circles (ga_circle equation) is the same as that for power gain circles. It is usually convenient to plot the gains relative to the MaxGain1 value. The Gav circles must be plotted on the source plane. They assume that the load is conjugately matched to the output for all Γ_S values.

The noise circle parameters (for the ns_circle equation) are defined as:

ns_circle(nf, NFmin, gamma_opt, rn, 51)

nf = noise figure of the device represented by Γ_S values that fall on the circle.

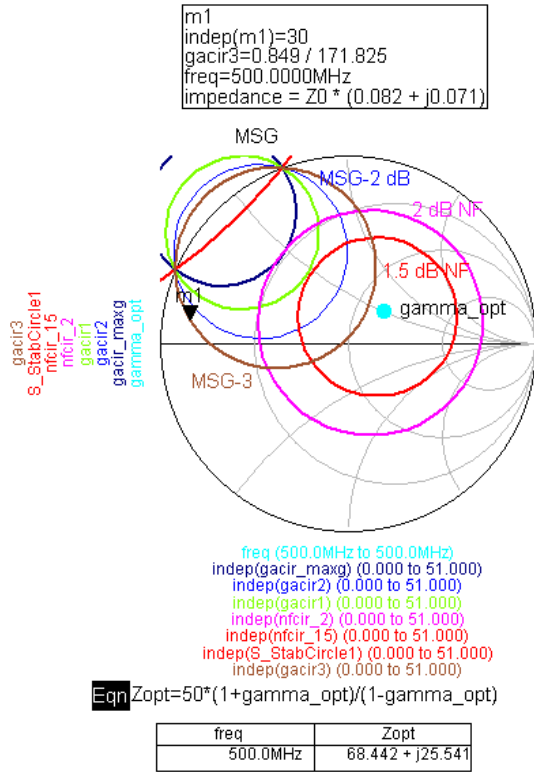
NFmin = minimum possible noise figure of device at this bias and frequency

Gamma_opt = Γ_{opt} = the optimum input match for best noise figure

rn = the noise resistance parameter (normalized to 50 ohms)

51 = number of points plotted on the circle

If the transistor model does not include noise data (not all of them do), you must enter the NFmin, gamma_opt, and rn from the transistor data sheet manually using equations (Eqn) in order to calculate noise circles.



Noise parameters calculated from device model

freq	NFmin	Rn	Sopt
500.0MHz	1.150	8.500	0.260 / 42.000

if noise parameters are all equal to 0, then the transistor model does not include noise data. You must then use equations (Eqn) to manually enter NFmin, m, and Sopt

freq	MaxGain1	MSG
500.0MHz	24.046	24.046

calculate available gain circles relative to MSG or MAG:

```

Eqn gacir_maxg=ga_circle(S,MaxGain1,51)
Eqn gacir1=ga_circle(S,MaxGain1-1,51)
Eqn gacir2=ga_circle(S,MaxGain1-2,51)
Eqn gacir3=ga_circle(S,MaxGain1-3,51)
        
```

```

Eqn gamma_opt=Sopt Eqn rn=Rn/50
        
```

calculate noise circles at 1.5 and 2.0 dB

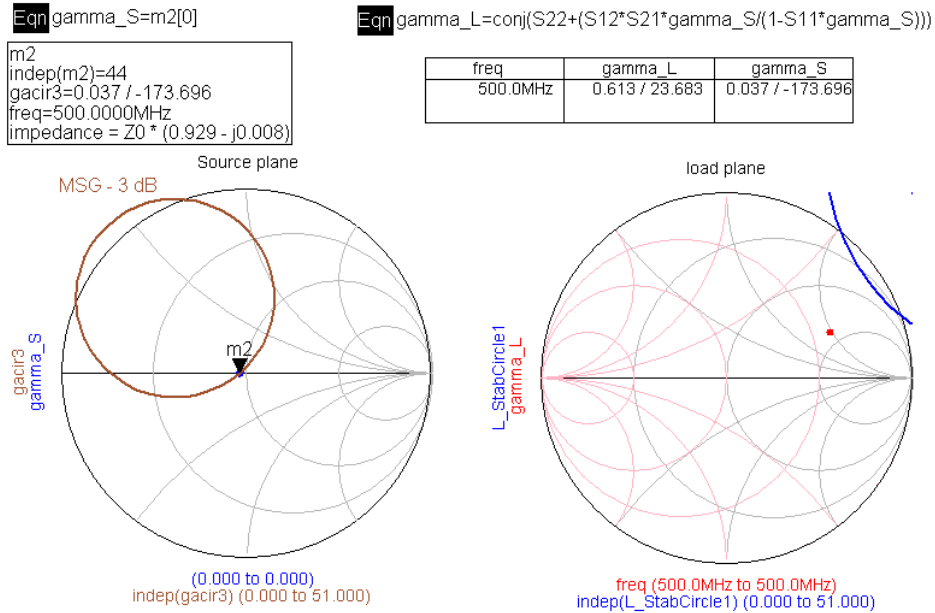
```

Eqn nfcir_15=ns_circle(1.5,NFmin,gamma_opt,rn,51)
Eqn nfcir_2=ns_circle(2,NFmin,gamma_opt,rn,51)
        
```

(Display file: circles.dds Use page menu to select gain and noise circles page)

In this example, you can see that MSG is quite high: 24 dB at 500 MHz. If we mismatch the input such that we give up 3 dB of available gain (brown gain circle), we can come quite close to the minimum noise figure of the device (1.15 dB). In fact, a $\Gamma_S = 0$ (very simple match!) could give a NF somewhere between 1.15 and 1.5 dB.

Once you have decided on the Γ_S value, place the marker there on the display. You need to design matching networks for the source and load. Calculate the corresponding Γ_L with the usual equation as shown below. Using the circles data display file, use the page menu to select the Gamma L page. Check the load stability circle to make sure Γ_L does not fall in an unstable region, and design the output matching network to provide Γ_L .



(Display file: circles.dds Use page menu to select gamma L page)

Then, design the matching networks, provide for bias insertion, and simulate the amplifier over a range of frequencies to verify stability out of band just as you did in the Stability and Gain tutorial.

SUMMARY

1. select device model; S-parameter model is preferred.
2. evaluate stability at design frequency. Don't use stabilizing resistors at input – these will greatly increase noise figure.
3. Plot available gain (GA) and noise figure circles. Tradeoff between gain and noise.
4. Design matching network.
5. Include bias insertion.
6. Simulate over wide frequency range. Must have stability at all frequencies. Modify circuit if necessary with frequency-dependent stabilizing circuits.

Appendix 1: Representing devices as an S2P file

In the case of the NE34018, the ADS component library does not contain a set of S parameters for the device. Instead, we can use an S2P file provided by NEC. The file for our device was measured at the VDS = 2V; ID = 5 mA bias point. An S2P file is simply a tab delimited ASCII text file containing S parameters measured on a 2 port device. The format is:

Frequency	S11		S21		S12		S22	
In GHz	mag	ang	mag	ang	mag	ang	mag	ang

Example file:

```
# GHZ S MA R 50
0.500 0.978 -17.0 6.806 162.8 0.018 80.1 0.723 -7.5
0.600 0.969 -20.3 6.731 159.7 0.022 78.7 0.719 -9.1
0.700 0.960 -23.6 6.691 156.5 0.025 77.3 0.712 -10.5
0.800 0.949 -26.9 6.624 153.4 0.028 75.5 0.706 -12.0
```

! NOISE PARAMETERS ADDED.....

! NOTE: The following data was just taken from the pdf file of NE34018 data sheet,

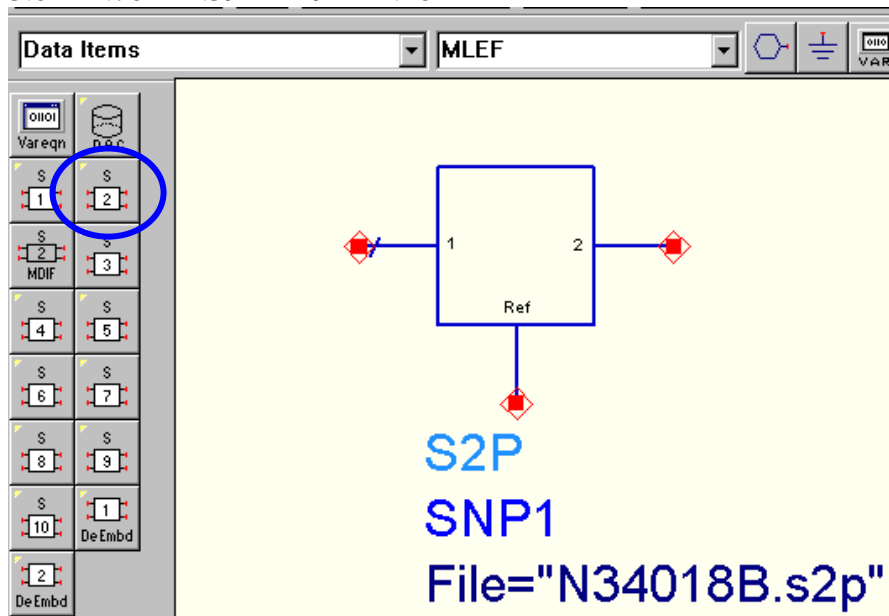
! For the Bias of VDS=2V and IDS=10mA.

! For Syntax, refer to the ADS S2P file format

! Ghz NFmin mag(gammaopt) phase(gammaopt) Rn/50

!NOISE DATA

0.9	.56	.76	30	.45
2.0	.63	.61	41	.28
2.5	.68	.49	51	.18
3.0	.70	.39	49	.16



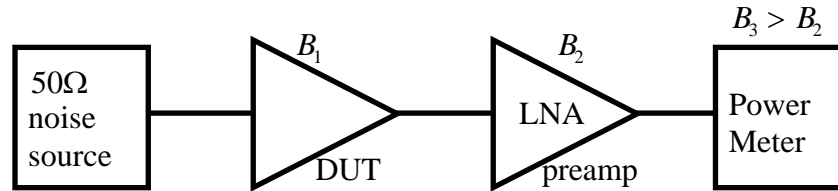
Place an S2P block from the Data Items menu and identify the file name. ADS will look for the file in the ADS project's data directory.

ADS Low Noise Design Tutorial

November 2004

Measuring Noise Figure in the Lab:**Method #3: HOT-COLD NF**

Use a calibrated noise source for measuring NF.



$$B_n \gg B_1 < B_2$$

The advantage here is that we don't need to know noise equivalent BW accurately.

Noise source has very wide BW compared with system under test.

$$P_H = \text{noise power with source on} = kT_H B$$

$$T_H = \text{effective noise temp. of source}$$

$$P_o = kT_o B = \text{noise power with source off.}$$

$$T_o = 290k$$

$$\text{Excess Noise Ratio} = ENR = \frac{P_H - P_o}{P_o} = \frac{T_H}{T_o} - 1$$

$$ENR(dB) = 10 \log_{10} \left(\frac{T_H}{T_o} - 1 \right)$$

Y factor for noise source:

$$Y_s = \frac{P_H}{P_o} = \frac{T_H}{T_o}$$

Two steps:

1. Source off. Noise power at meter:

$$P_1 = F kT_o B G_T$$

↑
↑

total noise factor
transducer gain

2. Source on.

$$P_2 = P_1 + Y_S kT_o B G_T$$

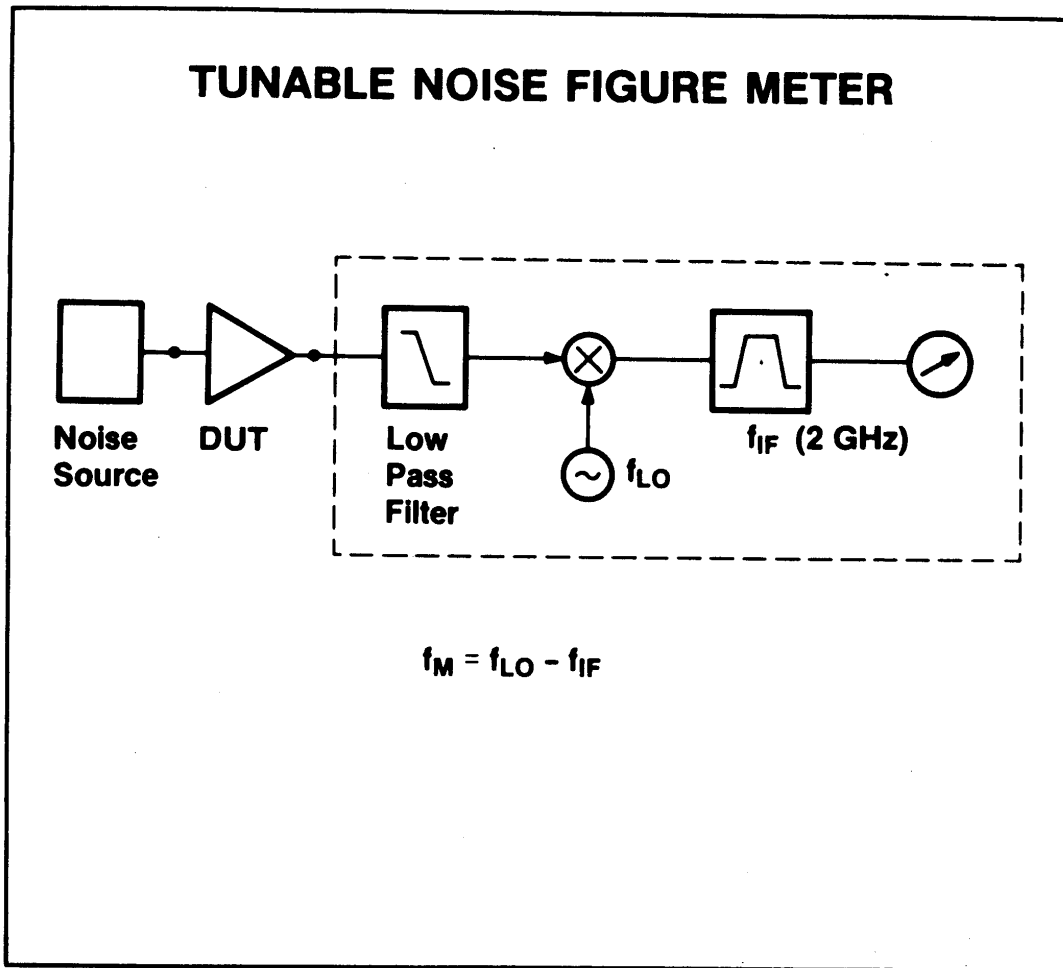
Divide: $\frac{P_2}{P_1} = Y = 1 + \frac{Y_S}{F}$

again, the transducer gain cancels, and now B cancels too. We can solve for F from the measured P_2/P_1 .

$$F = \frac{Y_S}{Y - 1} \quad \text{Noise factor – numerical ratios, not } dB.$$

and

$$\underline{NF = 10 \log F (dB)}$$



The tunable noise figure meter is a receiver. The mixer block upconverts the input noise signal to a 2 GHz power meter. $f_{in} = 2 \text{ GHz} - f_{LO}$

Thus, by choosing the local oscillator frequency f_{LO} , we measure the noise power within the bandwidth of the IF filter. The noise figure meter also applies a square wave to turn the noise source on and off, obtaining the HOT/COLD input noise condition needed to determine F. As an added bonus, the meter also measures the gain of the device under test.

Free Space Propagation Model

Used to calculate path loss between a transmitter and receiver.

Radiated power density:

$$S = \left(\frac{P_T}{4\pi r^2} \right) D$$

D = directivity = G_T = peak radiation power/average radiation power

P_T = transmitter power

A receiving antenna is described by its effective area A_e .

$$A_e = \frac{P_r}{S} = \frac{\text{power delivered to input}}{\text{incident power at antenna aperture}}$$

This relates to the directivity, D:

$$A_e = \frac{\lambda^2}{4\pi} D = \frac{\lambda^2 G_R}{4\pi}$$

(Wavelength enters into the calculation because A_e must scale with λ^2 to maintain a given directivity or gain.)

The power at the receiver is:

$$P_R = S A_e = \left(\frac{P_T G_T}{4\pi r^2} \right) \left(\frac{\lambda^2 G_R}{4\pi} \right) = \frac{P_T G_R G_T \lambda^2}{(4\pi r)^2 L}$$

Friis Equation

P_r = received power

P_t = transmit power

G_t = transmit antenna gain

G_r = receive antenna gain

λ = wavelength (meters)

$L \geq 1$ (system loss factor)

due to transmission line losses, filter attenuation

d = distance from transmitting antenna to receiving antenna

P_r falls off as square of distance

(20dB/decade rate)

$$\text{Path loss} = 10 \log \frac{P_t}{P_r} = -10 \log \left[\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2} \right]$$

(by convention, path loss is represented as a positive number)

Example. Terrestrial Microwave Link

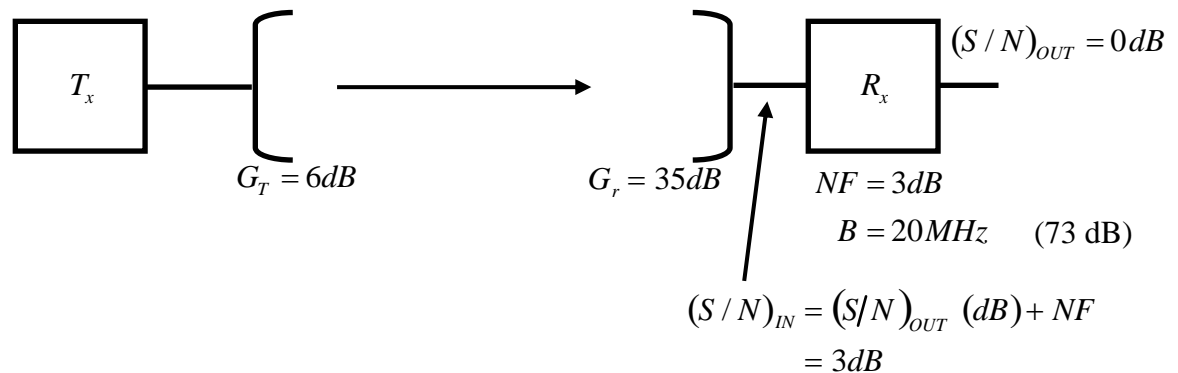
How much transmit power is needed to get $(S/N)_{OUT} = 0dB$?

Assume antenna sees standard terrestrial background temperature of 290K.

Frequency = 2.4 GHz. ($\lambda = 12.5$ cm)

$d = 50$ km.

Assume $L = 1$ (no loss – not a good assumption, but we will address this later)



Next, determine path loss:

$$10\log\left(\frac{P_r}{P_t}\right) = -10\log\left[\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2 L}\right]$$

z

How much power is needed at the input of the receiver?

$$S_{avi} = P_t - \text{path loss} = P_t \text{ (dBm)} - 134dB + 6 + 35 = P_t - 93 \text{ dB}$$

$$N_{avi} = kT_o B = -174 \text{ dBm} + 73 \text{ dB} = -101 \text{ dBm}$$

$$S_{avi} = N_{avi} + NF = N_{avi} - 3 \text{ dB} = -98 \text{ dBm}$$

$$\text{Thus,} \quad P_t \text{ (dBm)} = -98 \text{ dBm} + 93 \text{ dB} = -5 \text{ dBm}$$

More realistically, we need much higher S/N than 0 dB!

Antenna noise model. (example courtesy of D. Leeson)

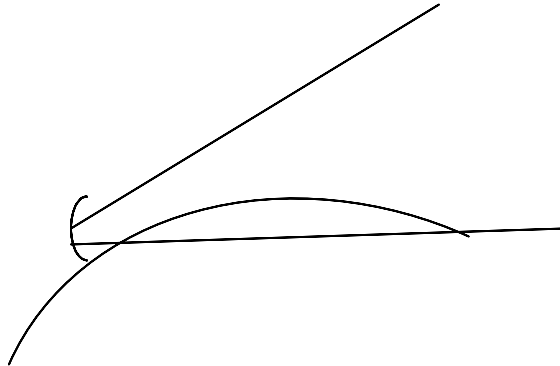
The antenna acts as a transducer for all incident radiation.

Objects in the view of the antenna radiate blackbody noise at their surface temperature.

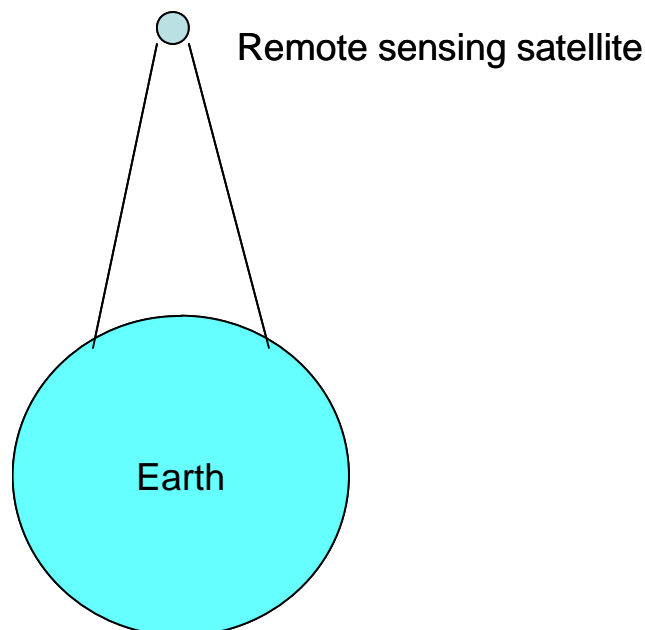
In the example above, the microwave link was point-to-point terrestrial. Thus, the antenna sees a noise temperature of 290K, and $N_{\text{avi}} = k 290 B$.

If the antenna is pointed toward space, but at a low angle to the horizon, part of the field of view will be the earth and part will be space. You can calculate an equivalent noise temperature for the antenna by the relative areas subtended by the solid angle of view of the antenna. Here, γ is the area of view of the earth.

$$T_a = \gamma T_{\text{earth}} + (1 - \gamma) T_{\text{sky}}$$



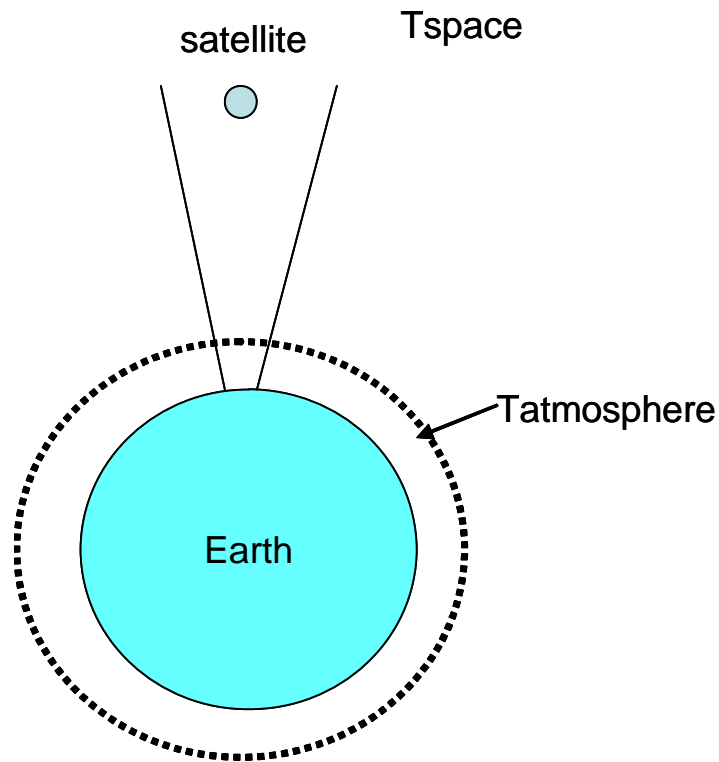
In the case above of the remote sensing satellite, its field of view is earth, so the antenna



temperature will be 290K once again.

If it were much farther away from the earth so that the antenna also was viewing partially earth and part space, the antenna temperature would be lower in proportion to the solid angles of each.

Earth satellite receiver



Now, the antenna sees only space. Atmospheric attenuation reduces the noise from space but also replaces it with noise from a hotter atmosphere.

η_{atm} = transmission coefficient of atmosphere.

$$T_{\text{ant_space}} = \eta_{\text{atm}} T_{\text{space}} + (1 - \eta_{\text{atm}}) T_{\text{atm}}$$

For a vertical pass on a clear day at 12 GHz: $\eta_{\text{atm}} = 0.98$

$$T_{\text{atm}} = 150\text{K}$$

T_{space} can vary depending on what is in view of the antenna between 3K and 100K. Of course, you want to avoid pointing the antenna at the sun!

Let's say $T_{\text{space}} = 10\text{K}$.

$$\text{Then, } T_{\text{ant_space}} = T_{\text{space}} \eta_{\text{atm}} + (1 - \eta_{\text{atm}}) T_{\text{atm}} = 12.8\text{K}$$

Now, we must consider other noise sources that add to the system noise temperature.

1. Antenna efficiency (ohmic loss)
2. Sidelobe noise
3. Feedline loss
4. LNA excess noise

1. Antenna efficiency. This consists of the ratio of radiation resistance to total resistance:

$$\eta_{ant} = \frac{R_{rad}}{R_{rad} + R_{loss}}$$

Since the antenna is at 290K, if $\eta_{ant} = 0.98$, then

$$T_{ant_loss} = T_{earth} * (1 - \eta_{ant}) = 5.8K$$

This adds to the effective antenna noise temperature.

2. Sidelobe noise. Antennas are not perfect. The radiation pattern also includes sidelobes with reduced gain, but still significant noise or interference can be added by these from undesired directions. Suppose we have a sidelobe gain of say – 14 dB, then the sidelobe gain

$$G_S = 0.04.$$

The sidelobes may see the earth, so suppose

$$T_{sidelobe} = G_S T_{earth} = 0.04 * 290K = 11.6K$$

We can now add the temperatures to obtain a total antenna temperature.

$$T_{ant} = 12.8 + 5.8 + 11.6 = 30.2K$$

Now we will include system losses between the antenna and receiver.

3. Feedline loss. This might also include filter losses. $T = 290K$. If the loss is = 0.3 dB, then the feedline “gain”, $G_F = 0.94$. This reduces antenna noise but adds much more thermal noise.

$$T_{ant_withfeed} = G_F * T_{ant} + (1 - G_F) T_{earth} = 0.94 * 30.2 + 0.06 * 290 = 45.8 K$$

This is a lot of extra noise. This is why the LNA is usually located on the antenna right at the feed point.

4. LNA and system. Now, we consider the system noise: $T_{\text{system}} = 35\text{K}$

Adding all contributors together give us a total noise temperature of:

$$T_{\text{ant_space}} = 12.8\text{K}$$

$$T_{\text{ant_loss}} = 5.8\text{K}$$

$$T_{\text{sidelobe}} = 11.6\text{K}$$

$$\text{So, } T_{\text{ant}} = 30.2\text{K}$$

Then, including feedline noise: $T_{\text{ant_withfeed}} = 45.8\text{K}$

Including system noise:

$$T_{\text{TOTAL}} = 45.8 + 35 = 80.8 \text{ K}$$

Converting to available noise power:

$$N_{\text{avi}} = k T_{\text{TOTAL}} = 1.38 \times 10^{-23} * 80.8 = 1.12 \times 10^{-21} \text{ W/Hz or } -179.5 \text{ dBm/Hz}$$

Now, let's apply this to the satellite example:

Suppose the path loss is -180 dB .

Transmitter power = 10W (40 dBm)

Transmit antenna gain = $G_T = 10 \text{ dB}$

Receive antenna gain = $G_R = 30 \text{ dB}$

Signal bandwidth = 10 MHz

Signal available power:

$$P_{\text{SIG}} = 40 \text{ dBm} - 180 + 10 + 30 = -100 \text{ dBm}$$

Available noise power:

$$P_N = -179.5 \text{ dBm} + 10 \log B = -109.5 \text{ dBm}$$

This gives us a S/N ratio of 9.5 dB .

What if the system noise temperature had been 70K rather than 35K . How much would that degrade S/N?

Our total noise temperature now increases to: $T_{\text{TOTAL}} = 45.8 + 70 = 115.8 \text{ K}$

$$N_{\text{avi}} = k T_{\text{TOTAL}} = 1.38 \times 10^{-23} * 115.8 = 1.60 \times 10^{-21} \text{ W/Hz or } -178 \text{ dBm/Hz}$$

Available noise power:

$$P_N = -178 \text{ dBm} + 10 \log B = -108 \text{ dBm}$$

This reduces the S/N ratio by 1.5 dB.

The transmitter power on the satellite would need to increase from 10W to 14.1 W. This can be very expensive on a satellite as it increases the system power and cooling requirements.

Compare noise figure of the system in these two cases. Remember that noise factor is defined in terms of noise temperature by:

$$F = 1 + T_{\text{system}}/290$$

$$F_{\text{case1}} = 1 + 35/290 = 1.12 \quad (0.5 \text{ dB NF})$$

$$F_{\text{case2}} = 1 + 70/290 = 1.24 \quad (0.93 \text{ dB NF})$$

We see that going from 70 to 35K improves the noise figure by only 0.43 dB but improves S/N by 1.5 dB. This discrepancy is due to the fact that we are looking out into a cooler background (12.8K) rather than 290K. Excess noise contribution at 290K really hurts S/N.

The S/N improvement would be even greater if some of the other noise contributors (sidelobe noise, feedline loss) were smaller.