Unique inductive feedback LNA design

In first-stage, low-noise amplifiers, optimum noise and optimum return loss performance are often compromised by impedance matching. This load impedance mismatching and inductive feedback design offers some relief.

By David VanStone

First-stage, low-noise amplifier (LNA) designs often require both low-noise and low-input voltage-standing-wave ratios (VSWR). Unfortunately, the source reflection coefficient required for an input conjugate match (Γ_M) and the source reflection coefficient required for minimum noise (Γ_{OPT}) are rarely equal. As a result, the designer is faced with having to find a compromise input match that sacrifices both optimum noise and optimum input return-loss performance.

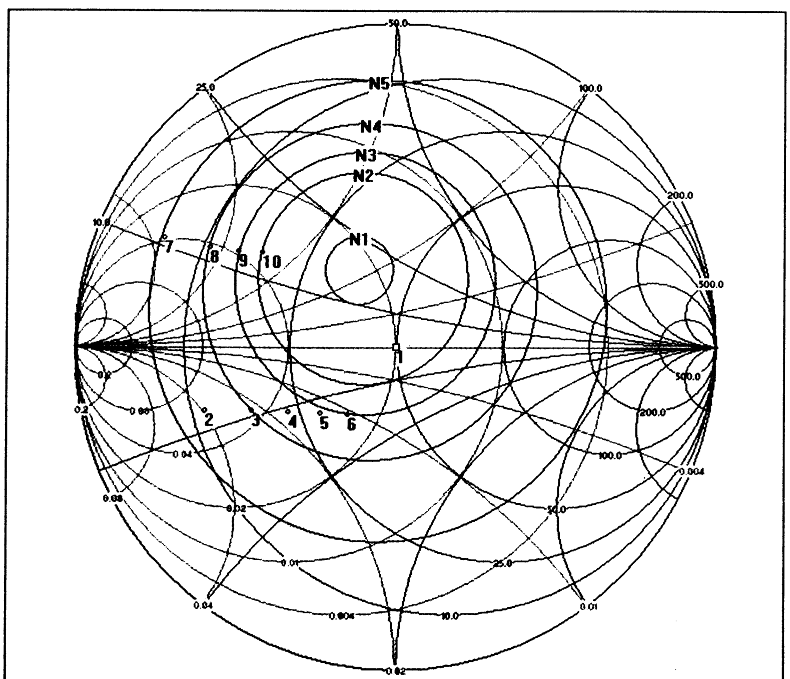

Figure 1. A plot of the input impedance of the simulated amplifier and the source-reflection coefficient required for an input conjugate match.

However, for most field-effect transistors (FETs) and bipolar junction transistors (BJTs), there is a combination of source inductance and load reflection coefficient that will produce $\Gamma_{OPT} - \Gamma_M$ coincidence. Under this condition, an amplifier can be designed that exhibits minimum noise and minimum reflected power at the input.

But this desirable condition extracts a price by presenting a poor match at the output and achieving lower power gain. However, the designer can transform the load into reflection coefficients somewhere between those producing optimum noise figure and those producing a simultaneous conjugate match and, thereby, achieve an acceptable compromise between noise figure, gain and output return loss. This article will discuss how combining a series inductive feedback with a unique, load-reflection coefficient can produce an amplifier with improved noise figure and input-matching performance compared to that produced by traditional design techniques.

The use of either emitter or source inductance feedback to increase the input resistance and increase the *k*-factor of a bipolar or field-effect transistor is well documented. Refer to references 1 to 3 for thorough theoretical and practical coverage of this technique. However, for convenience, a cursory review of theory and practice follows.

Background: series inductive feedback

A small amount of inductance, in series with the emitter or source, has three predominant effects:

- Increased input resistance
- Increased in-band k-factor (increased in-band stability)
- Decreased gain

Secondary effects include changes to input reactance and small shifts to Γ_{OPT}. Typically, the inductance is inserted by grounding the transistor through a short length of transmission line. The inductive reactance of the stubs is usually no greater than 10Ω and line lengths are typically 0.1" or less with characteristic impedances of 50Ω or greater. This kind of lossless feedback (assuming an ideal inductor) has no effect on the minimum noise figure of the device. Because it increases input resistance, source inductance usually moves the reflection coefficient required for an input conjugate match closer to Γ_{OPT}.

To illustrate the effects of source inductance, a pseudomorphic, high-electron-mobility transistor (PHEMT) amplifier (using an advanced Curtis quadratic model) was simulated with different amounts of source inductance. Figure 1 plots the input impedance (points 2 through 6) and the source reflection coefficient required for an input conjugate match (Γ_M, points 7 through 10) as a function of source inductance. Constant-noise figure circles are also plotted; the 0.43 dB noise figure circle is labeled N_1 and the minimum noise figure (F_{MIN}) is 0.42 dB. As the source trace length increases, corresponding to increased source inductance, Γ_M moves closer to Γ_{OPT} (located at the center of circle N_1).

As the source inductance is increased, the *k*-factor

Inductor length (inches) ($Z_0 = 50\Omega$)	Inductance (nH)	Z_{in}	Γ_S for simultaneous conjugate match	Noise figure circle
0.0	0	point 2	no match point, $k < 1$	n/a
0.025	0.22	point 3	point 7	N5, 1.2 dB
0.05	0.44	point 4	point 8	N4, 0.72 dB
0.075	0.66	point 5	point 9	N3, 0.58 dB
0.100	0.88	point 6	point 10	N2, 0.52 dB

Data table for Figure 1.

at higher frequencies eventually falls below 1. This effect limits the amount of source inductance that can safely be used. In Figure 2, the k-factor of the simulated amplifier is plotted for source inductance values of 0 nH to 0.44 nH.

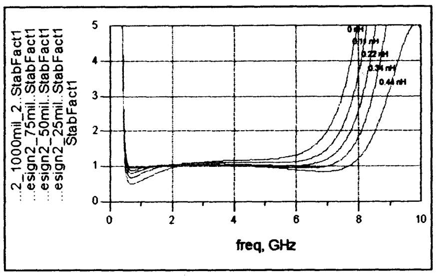

Figure 2. *K*-factor plots for source inductance values of 0 to 0.44 nH. Inductor lengths vary from 0 to 100 mils in 25 mil steps.

Design Example 1

An LNA was designed for the 1710 MHz to 1785 MHz band with both ports matched (simultaneous conjugate match). In order to produce a k-factor of 1 or greater at all frequencies, a source inductance value of 0.25 nH was used together with a 51Ω resistor at the output, grounded through a quarter-wave shorted-stub at 1745 MHz (see Figure 3). The simulation yielded the following results at midband:

Gain: 18.3 dB
Input return loss: 27 dB
Output return loss: 23.1 dB
Noise figure: 0.91 dB
Stability: unconditional at all frequencies

The amplifier has respectable performance, but the noise figure is nearly 0.5 dB greater than F_{MIN}, which is 0.42 dB. A technique for achieving both minimum noise figure and high-input return loss is developed in the next section.

Background: load impedance tuning

The basis of the technique is the interaction between the load reflection coefficient and the input reflection coefficient of an amplifier. This relationship is expressed in the following equation:

$$\Gamma_{IN} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \qquad (1)$$

If a value of Γ_L can be found that will make $\Gamma_{IN} = \Gamma^*_{OPT}$, or equivalently $\Gamma_M = \Gamma_{OPT}$, then a minimum noise-figure match and a conjugate match can be obtained simultaneously. (Recall that $\Gamma_M = \Gamma_{IN}^*$.) For some sets of S parameters, no value of $\Gamma_L \leq 1$ exists that will produce this condition. In these cases, the S parameters can be modified by using an appropriate amount of series-inductive feedback. A value of Γ_L that is ≤ 1 can usually be found using S parameters of a transistor that has been stabilized using source inductance.

A MATLAB program was written that will find the load-reflection coefficient (if one exists) that will make $\Gamma_M = \Gamma_{OPT}$ for a given set of S-parameters. The program is listed in Appendix A (see page 82).

Before running the MATLAB program, it may be best to determine how much source inductance, if any, is required to produce $\Gamma_M = \Gamma_{OPT}$. This can be determined by mapping the Γ_L plane onto the Γ_{IN} plane. Mapping capability is included in most simulation programs, but a simple mapping program can be readily written. It should map the unit circle of the Γ_L plane (Γ from $1 \angle 0°$ to $1 \angle 360°$) to the Γ_{IN} plane using equation 1 and the S- parameters for the amplifier configuration under study. The image of the unit circle will usually be a smaller circle on the Γ_{IN} plane. This circle, which may extend beyond the perimeters of the conventional Smith chart, must enclose Γ_{OPT}^* for the condition $\Gamma_M = \Gamma_{OPT}$ to occur with a passive load impedance, $\Gamma_L \leq 1$. Figure 4 shows plots of four Γ_L-to-Γ_{IN} mappings using values of source inductance from 0 to 0.44 nH. For the PHEMT device under study, a source inductance of at least

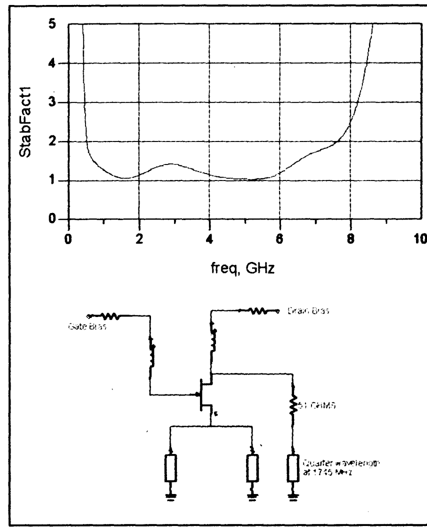

Figure 3. (a) *K*-factor plot of stabilized PHEMT amplifier and (b) basic stabilized amplifier topology.

0.1 nH is required to produce $\Gamma_M = \Gamma_{OPT}$ with a $\Gamma_L \leq 1$.

Design example 2

The second design example will use the MATLAB program to help design an LNA with a minimum noise figure match and minimum reflected power at the input (conjugate match).

The same PHEMT is used as in the first example, with the same amount of source inductance and the same stabilizing network at the output. The S parameters of this network are found using the simulation program. These parameters, along with the value for Γ_{OPT}, are input to the MATLAB program. The program then finds a value for Γ_L of $0.34 \angle -143°$. When this load impedance is presented to the output of the amplifier, Γ_M will be equal to $0.27 \angle 116°$, which is Γ_{OPT} for this device at this frequency.

Impedance-transforming networks were designed and the amplifier was simulated. The simulation yielded the following results at midband:

Gain: 16.5 dB

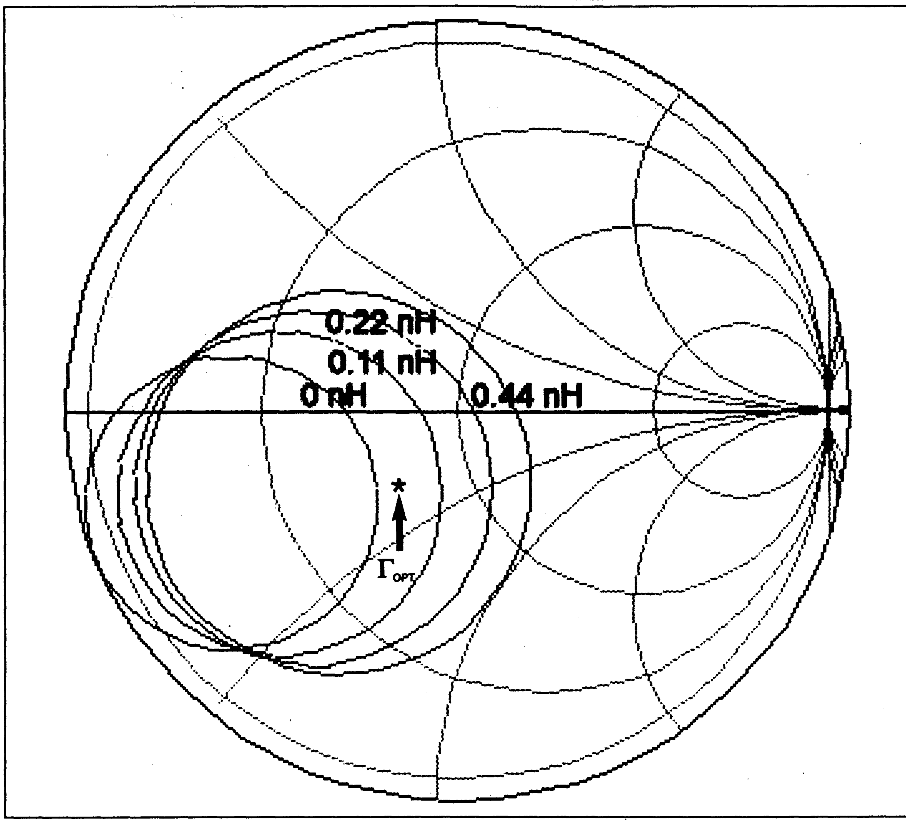

Figure 4. Unit circles mapped from the Γ_L plane to the Γ_{IN} plane.

a conjugate input match (Γ_M).

$$\Gamma_M = \left[S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} \right]^* \qquad (2)$$

As the load reflection coefficient moves from point 1 to point 2, gain and output return loss will improve as noise figure degrades. By trial-and-error, a Γ_L will eventually be found that produces an acceptable compromise between gain, noise figure and output return loss.

Design example 3

An amplifier was designed using this technique, again using the same PHEMT with the same amount of source inductance and the stabilizing network as in the first two examples. The load-reflection coefficient that produces $\Gamma_{OPT} - \Gamma_M$ coincidence and the load-reflection coefficient for a simultaneous conjugate match are plotted as shown in Figure 4. A load-reflection coefficient of 0.21 ∠ 102° (point 3 on Fig. 5) was eventually tried that, by using equation 2, yielded a Γ_M of 0.46 ∠

Input return loss: 30.3 dB
Output return loss: 6.6 dB
Noise figure: 0.51 dB
Stability: unconditional at all frequencies

When we compare these results to those of the first example, we note the following changes to the amplifier's performance:

1.8 dB reduction in gain
0.4 dB improvement in noise figure
Output return loss degraded by > 16 dB

The design trades off gain and output return loss for improved noise-figure performance. If the insertion-loss of the input noise-matching network is kept low by using low-loss components, this technique allows the designer to attain an amplifier noise figure close to F_{min}.

If the sacrifice of gain and output match using this technique is unacceptable, a "compromise" between the minimum noise technique and the simultaneous-match technique can be found. This involves plotting both the load-reflection coefficient required for a simultaneous conjugate match and the load-reflection coefficient required for $\Gamma_{OPT} - \Gamma_M$ coincidence on the same Smith chart. A straight line (see Figure 4) connects the two points.

Now, pick a value for Γ_L along this line and, using equation 2, find the source-reflection coefficient required for

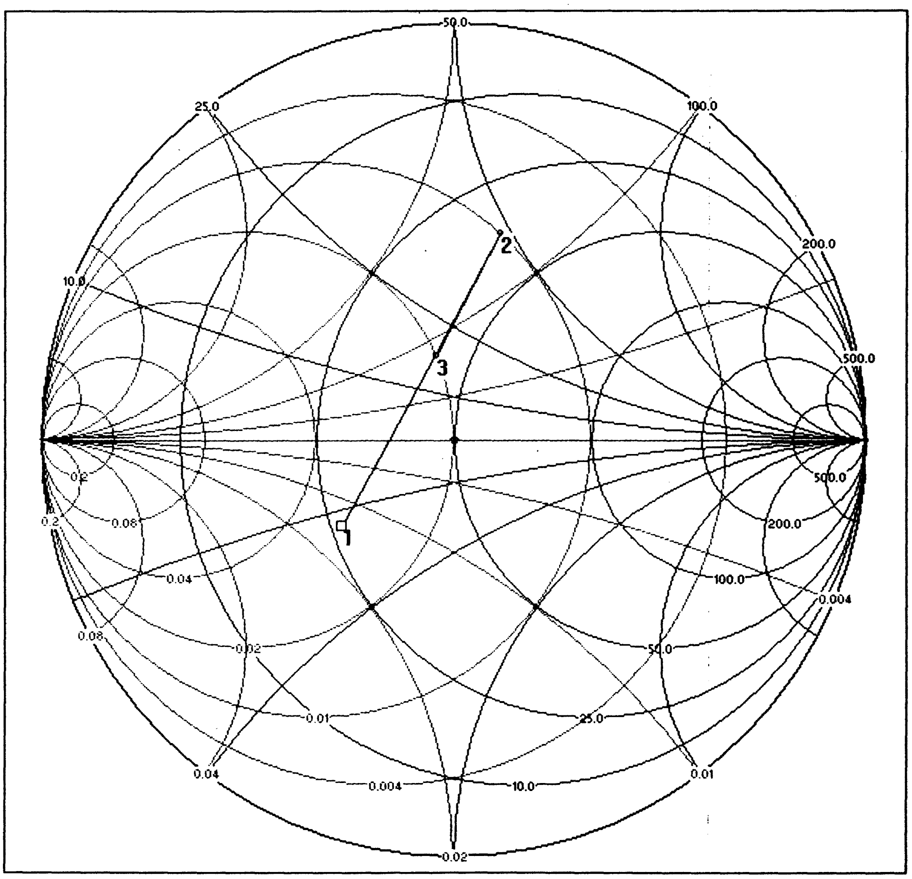

Figure 5. Point 1 is the load-reflection coefficient that produces $\Gamma_{OPT} - \Gamma_M$ coincidence. Point 2 is the load-reflection coefficient for a simultaneous conjugate match. Γ_L (point 3) is a compromise load-reflection coefficient, 0.21 ∠ 102°.

143°. The simulation yielded the following results at midband:

Gain: 17.9 dB
Input return loss: 23 dB
Output return loss: 11.6 dB
Noise figure: 0,66 dB
Stability: unconditional at all frequencies

Compared to Example 1, there is only 0.3 dB lower gain and a 3 dB difference in input return loss. The output return loss of 11.6 dB was deemed by the author as an acceptable compromise in obtaining a noise figure of 0.66 dB, an improvement of 0.25 dB over Example 1.

Conclusion

This article has presented a method for obtaining the optimum noise figure from an amplifying device while also achieving an excellent input match. The method requires a good computer simulation program, a Smith chart and the simple MATLAB program written by the author. This technique will find an application anytime a designer needs to squeeze out the last bit of noise performance from a transistor amplifier that is preceded by a device requiring a good termination.

RF

References

1] M. Murphy, "Applying the Series Feedback Technique to LNA Design," *Microwave Journal*, Nov. 1989.

2] D. Henkes, "LNA Design Uses Series Feedback to Achieve Simultaneous Low Input VSWR and Low Noise," *Applied Microwave & Wireless*, Oct. 1998

3] Agilent Technologies "High Intercept Low Noise Amplifier for the 1850 MHz through 1910 MHz PCS Band using the ATF-54143 Enhancement Mode PHEMT."

About the author

David VanStone works at Motorola's GTSS Wireless Systems Group as an RF Design Engineer. He designs and develops low-noise front ends for base station receivers. His previous experience includes designing microwave synthesizers and low-noise crystal oscillators. He holds a B.S.E.E. from Illinois Institute of Technology. He can be reached at 847.632.5829 or by e-mail at *dvansto1@email.mot.com*

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APPENDIX A – The MATLAB Program

```
%Program calculates the load reflection coefficient for coincident gamma_opt and gamma source

fprintf('\nInput S parameters to calculate Load Gamma. This gamma will force GammaOpt = GammaMS or GammaIn = GammaOpt*\n');
S11mag = input('\nEnter S11 magnitude: ');
S11ang_deg = input('Enter S11 angle: ');
S21mag = input('Enter S21 magnitude: ');
S21ang_deg = input('Enter S21 angle: ');
S22mag = input('Enter S22 magnitude: ');
S22ang_deg = input('Enter S22 angle: ');
S12mag = input('Enter S12 magnitude: ');
S12ang_deg = input('Enter S12 angle: ');
GammaOptMag = input('\nEnter gamma_opt magnitude: ');
GammaOptAng_deg = input('Enter gamma_opt angle: ');
S11ang_rad = S11ang_deg * 2 * pi / 360;
S21ang_rad = S21ang_deg * 2 * pi / 360;
S22ang_rad = S22ang_deg * 2 * pi / 360;
S12ang_rad = S12ang_deg * 2 * pi / 360;
LoadGammaMag = 0.01;
LoadGammaAng_deg = 0;
% convert S22 to rectangular
S22real = S22mag*cos(S22ang_rad);
S22imag = S22mag*sin(S22ang_rad);
% convert S11 to rectangular
S11real = S11mag*cos(S11ang_rad);
S11imag = S11mag*sin(S11ang_rad);
Convergence = 0;
while (LoadGammaMag < 1) & (Convergence == 0)
   while (LoadGammaAng_deg < 360) & (Convergence == 0) % find source gamma
LoadGammaAng_rad = LoadGammaAng_deg * 2 * pi / 360;
GsNumeratorMag = S12mag * S21mag * LoadGammaMag;
GsNumeratorAng = S12ang_rad + S21ang_rad + LoadGammaAng_rad;
Dmag = LoadGammaMag * S22mag;
Dang = LoadGammaAng_rad + S22ang_rad;
DmagRect = Dmag * cos( Dang ) + i * Dmag * sin( Dang );
GsDenominatorRect = 1 - real( DmagRect ) - i*imag( DmagRect);
%Divide numerator by denominator
GammaS1Mag = GsNumeratorMag / abs(GsDenominatorRect);
GammaS1Ang = GsNumeratorAng - angle(GsDenominatorRect);
% convert to rect.
GammaS1Rect = GammaS1Mag * cos( GammaS1Ang ) + i * (GammaS1Mag * sin(GammaS1Ang));
SourceGammaRect = (S11real + real(GammaS1Rect))+ i*( S11imag + imag( GammaS1Rect ));
SourceGammaRect = conj(SourceGammaRect);
SourceGammaMag = abs( SourceGammaRect );
SourceGammaAng_rad = angle( SourceGammaRect );
SourceGammaAng_deg = SourceGammaAng_rad * 360 / (2*pi);
if abs(SourceGammaMag - GammaOptMag) <0.01 & abs(SourceGammaAng_deg - GammaOptAng_deg) < 1
Convergence = 1;
fprintf('\nLoad reflection coefficient for GammaOpt/GammaSource coincidence: %1.3f < %3.1f\n', LoadGammaMag, LoadGammaAng_deg );
else
   LoadGammaAng_deg = LoadGammaAng_deg + 1;
end
end
LoadGammaMag = LoadGammaMag + 0.01;
LoadGammaAng_deg = 0;
end
if Convergence == 0
fprintf('\nNo convergence. No value of load impedance will force GammaOpt=GammaMS.\n');
end
% END PROGRAM
```