

# LNA Design

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An LNA (Low Noise Amplifier) combines a low noise figure, reasonable gain, and stability without oscillation over entire useful frequency range.

The Low Noise Amplifier (LNA) always operates in Class A, typically at 15-20% of its maximum useful current. Class A is characterized by a bias point more or less at the center of maximum current and voltage capability of the device used, and by RF current and voltages that are sufficiently small relative to the bias point that the bias point does not shift.

The smallest signal that can be received by a receiver defines the receiver sensitivity. The largest signal can be received by a receiver establishes the upper power level limit of what can be handled by the system while preserving voice or data quality. The dynamic range of the receiver, the difference between the largest possible received signal and the smallest possible received signal, defines the quality of the receiver chain. The LNA function, play an important role in the receiver designs. Its main function is to amplify extremely low signals without adding noise, thus preserving the required Signal-to-Noise Ratio (SNR) of the system at extremely low power levels. Additionally, for large signal levels, the LNA amplifies the received signal without introducing any distortions, which eliminates channel interference.

- An LNA design presents a considerable challenge because of its simultaneous requirement for high gain, low noise figure, good input and output matching and unconditional stability at the lowest possible current draw from the amplifier.
- Although Gain, Noise Figure, Stability, Linearity and input and output match are all equally important, they are interdependent and do not always work in each other's favor.
- Carefully selecting a transistor and understanding parameter trade-offs can meet most of these conditions.
- Low noise figure and good input match is really simultaneously obtained without using feedback arrangements.
- Unconditional stability will always require a certain gain reduction because of either shunt or series resistive loading of the collector. High IP3 requires higher current draw, although the lowest possible noise figure is usually achieved at lower current levels.
- Envelope termination technique can be used to improve IP3 performance while operating LNA at low current levels.
- Additional improvement of IP3 can also be achieved by proper power output matching (1dB compression point match or P1dB match). The P1dB match, being different from conjugate match, reduces the gain although improving IP3 performance.
- Transistor selection is the first and most important step in an LNA design. The designer should carefully review the transistor selection, keeping the most important LNA design trade-offs in mind.

The transistor should exhibit high gain, have a low noise figure, and offer high IP3 performance at the lowest possible current consumption, while preserving relatively easy matching at frequency of operation.

Examination of a data sheet is a good starting point in a transistor evaluation for LNA design.

The transistor's S-parameters should be published at different collector/emitter (or drain/source) voltages and different current levels for frequencies ranging from low to high values. The data sheet should also contain noise parameters, which are essential for low noise design. Spice models for the transistor are also useful for IP3 and P1dB simulations. The designer should first look at the main design parameters as: Noise, Gain, and IP3, and decide what Vce and Ic levels will produce optimal performance.

The forward transducer power gain represents the gain from transistor itself with its input and output presented with  $50\ \Omega$  impedance.

The manufacturer of the transistor at multiple frequencies and different Vce and current levels provides the S21 values.

Additional gain can be obtained from source and load matching circuits.

Maximum Stable Gain and Maximum Power Gain ( $G_{max}$ ) are good indicators of additional obtainable gain from the LNA circuit.

LNA linearity is another important parameter. A figure of merit for linearity is IP3. A two-tone test is used for derivation of IP3.

As a rule of thumb for bipolar junction transistors (BJT), the Output-IP3 can be estimated from the following formula:

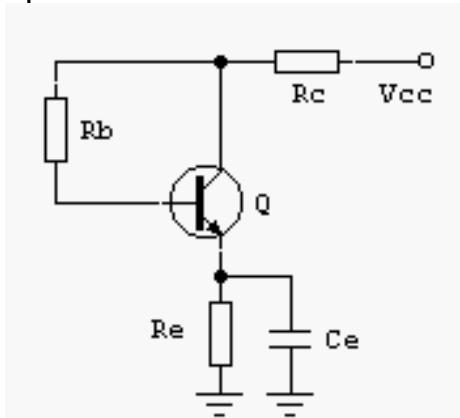
$$OIP3_{[dBm]} = 10 \log (V_{ce}^{[volt]} * I_c^{[mA]} * 5)$$

- RF performance of the LNA depends by many variables as:
  - Frequency
  - DC Biasing and Power Dissipation
  - Stability
  - Input and Output Matching
  - Layout and Grounding
  - EM Shielding
  - Supply decoupling
  - Temperature

## 1. DC Biasing (BJT)

represent the first step in LNA design.

- The chosen DC bias circuit should exhibit stable thermal performance and reduce the influence of  $h_{FE}$  spread.
- The resistive feedback arrangement is the simplest form of DC biasing that fulfills all the major requirements.
- Two bias feedback arrangements are possible: one with a combination of  $R_c$  and  $R_b$  and a second one with simple  $R_e$  and  $C_e$  combination.



The operation of the  $R_c$  and  $R_b$  is simple:  $R_c$  and  $R_b$  will establish a biasing point. If the device current increases, the voltage drop across  $R_c$  increases, reducing the voltage seen by the base, thereby providing feedback. Because the operation class of the LNA is going to be Class-A (constant current draw for dynamic range of power levels), a stable biasing point over different temperatures is required. For different lot of transistors small variation in  $h_{FE}$  can be expected. For  $R_b$  to have little influence on source matching, which is crucial for noise performance, the feedback network should be decoupled with an inductor (making biasing invisible at RF band of operation).

Another possible bias feedback can be realized with emitter resistor and capacitor.

$C_e$  should be selected carefully, because  $R_e$  will also have a direct effect on RF gain of LNA.  $C_e$  should present a short at frequency of operation to limit its influence on gain and noise performance of the circuit.

Other biasing methods are suitable for Class-A networks. These are usually closed feedback arrangements with dynamic bias control provided by active components. Although suitable for LNA application, these active feedback bias networks increase complexity of the LNA network, introduce additional components and increase the real-estate area of the solution.

## 2. Stability Design

should be the next step in LNA design.

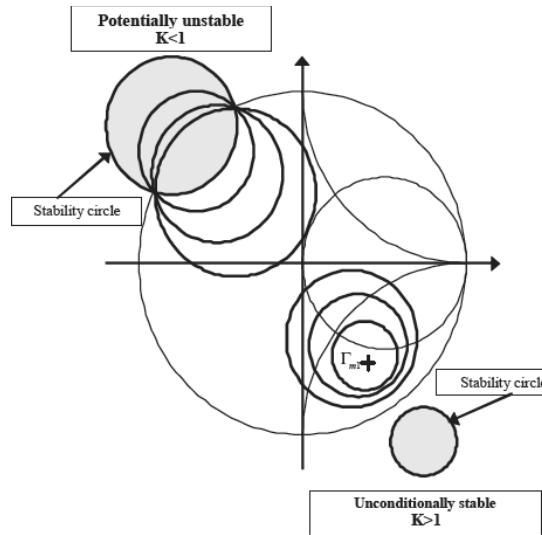
- Unconditional stability of the circuit is the goal of the LNA designer.
- Unconditional stability means that with any load present to the input or output of the device, the circuit will not become unstable – will not oscillate. Instabilities are primarily caused by three phenomena: internal feedback of the transistor, external feedback around the transistor caused by external circuit, or excess gain at frequencies outside of the band of operation.
- S-parameters provided by manufacturer of the transistor will aid in stability analysis: numerical ( $K$  and  $\mu$  factors) and graphical (stability circles on Smith Chart).
- An option of numerical analysis consists of calculating a term called Rollett Stability Factor (or K-factor). K-factor is given by:

$$K = \{1 - |S_{11}|^2 - |S_{22}|^2 + |S_{11}S_{22} - S_{12}S_{21}|^2\} / \{2|S_{12}S_{21}|\}$$

- When  $K > 1$ , the circuit will be **unconditionally stable** for any combinations of source and load impedance.
- When  $K < 1$ , the circuit is **potentially unstable** and oscillation may occur for certain combination of source and/or load impedance present to the transistor.
- However, the two necessary and sufficient conditions for unconditional stability are that the **stability factor K** is greater than unity ( $K > 1$ ) and the **stability measure b** is positive ( $b > 0$ ). Where  $b$  is given by:

$$b = 1 + |S_{11}|^2 - |S_{22}|^2 - |S_{11}S_{22} - S_{12}S_{21}|^2$$

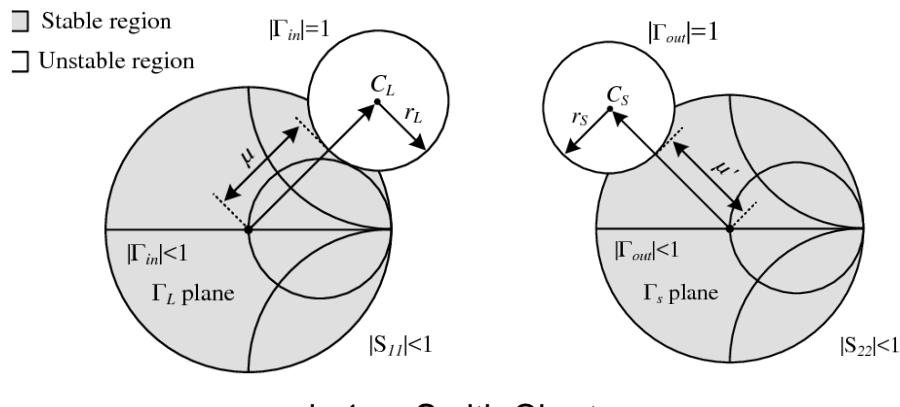
- The K-factor represents a quick check for stability at given biasing condition. A sweep of the K-factor over frequency for a given biasing point should be performed to ensure unconditional stability outside of the band of operation.
- When working with K-factor, it is important that it is greater than 1. However, the actual value is not telling anything. You cannot say that an amplifier with  $K=3$  is "more stable" than an amplifier with  $K=1.1$ .
- As noted, if  $K > 1$ , says the amplifier is stable but give no hint at how far outside the Smith Chart the stability circles are. Similarly, if you get  $K < 1$  it doesn't give you much insight as to what to do, or where to do it.



Smith Chart Stability circles and K-factor

- To determine if a two-port is unconditionally stable, also could be used the  **$\mu$  (Mu) test**.
 
$$\mu = \{1 - |\mathbf{S}_{11}|^2\} / \{|\mathbf{S}_{22} - \text{conj}(\mathbf{S}_{11}) * \Delta| + |\mathbf{S}_{12} * \mathbf{S}_{21}|\}$$

$$\mu' = \{1 - |\mathbf{S}_{22}|^2\} / \{|\mathbf{S}_{11} - \text{conj}(\mathbf{S}_{22}) * \Delta| + |\mathbf{S}_{21} * \mathbf{S}_{12}|\}$$
 where Delta is the determinant of the S-parameter matrix.
- Having both  $\mu$  and  $\mu' > 1$ , is the single necessary and sufficient condition for **unconditional stability** of the two-port network.
- The  $\mu$ -test not only is a test for unconditional stability, but the magnitudes of  $\mu$  and  $\mu'$  is a quantified measure of the stability.  
In other words, if one of two-ports has a larger  $\mu$ , it is more stable as the other.
- $\mu$  (**Mu**) is the distance from the center of the Smith Chart to nearest point of instability on the **input** stability circles.  $\mu'$  is the same, but for the **output** stability circles.
- If  $\mu$  or  $\mu' = 1$  means that a stability circle just brushes the edge of the Smith Chart.



$\mu$  and  $\mu'$  on Smith Chart

- The advantage of the  $\mu$ -test is that only a single parameter needs to be evaluated. There are no auxiliary conditions as the K-factor needs.

The designer's goal is to design an LNA circuit that is unconditionally stable for the complete range of frequencies where the device has a substantial gain.

An LNA designer can use at least five methods for circuit stabilization.

- The first one consists of resistive loading of the input. This method, although capable of improving the stability of the circuit, also degrades the noise of the LNA and is almost never used.

- Output resistive loading is preferred method of circuit stabilization. This method should be carefully used because its effects are lower gain and lower P1dB point (thus IP3 point).
- The third method uses collector to base resistor-inductor-capacitor (RLC) feedback to lower the gain at the lower frequencies and hence improve the stability of the circuit.
- The fourth method consists of filter matching, usually used at the output of the transistor, to decrease the gain at a specific narrow bandwidth frequency. This method is frequently used for eliminating gain at high frequencies, much above the band of operation. Short circuit quarter wave lines designed for problematic frequencies, or simple capacitors with the same resonant frequency as the frequency of oscillation (or excessive gain) can be used to stabilize the circuit.
- The final stabilization method can be realized with a simple emitter feedback inductor. A small inductor can make the circuit more stable at higher frequencies. But if the source inductance is increased, the K-factor at higher frequencies eventually falls below 1. This effect limits the amount of source inductance that can safely be used.
- To get the best LNA stability performances have to accommodate the full range of expected variations in operating parameters as:
  - Component package parasitics
  - Component values
  - Temperature
  - Supply voltage
- Most common causes for LNA instability are:
  - Insufficient RF decoupling between supply lines of the amplifier bias.
  - Parasitic inductance in GND connections.
  - Excess in-band and/or out-of-band Gain.
  - Electro-Magnetic coupling and Feedback.
- Always check stability of your LNA well beyond band-of-interest checking for both, small-signal stability and for large-signal stability.
- Use stability circles on Smith Chart (for both, source and load) to verify legitimacy of chosen  $Z_{in}$  and  $Z_{out}$  impedances.

### 3. Noise Matching and Input Return Loss (IRL)

The next step in LNA design consists of Noise Match and Input Return Loss (IRL).

- IRL defines how well the circuit is matched to  $50\ \Omega$  matching of the source.
- A typical approach in LNA design is to develop an input matching circuit that terminates the transistor with conjugate of Gamma optimum ( $\Gamma_{opt}$ ), which represents the terminating impedance of the transistor for the best noise match.

In many cases, this means that the input return loss of the LNA will be sacrificed.

The optimal IRL can be achieved only when the input-matching network terminates the device with a conjugate of  $S_{11}$ , which in many cases is different from the conjugate of  $\Gamma_{opt}$ .

- In 1928 [H. Nyquist](#) showed that the noise from any impedance is determined by its resistive component. Consequently, if an ideal lossless element is used to provide feedback, then the minimum noise measure is unaffected.
- An emitter (or source) inductor feedback can rotate  $S_{11}$  closer to  $\Gamma_{opt}$ , which can help obtaining close to minimum Noise Figure and respectable IRL simultaneously.

The additional series inductance provides lossless negative series feedback and also reduce the overall available gain of the network and can be used in balancing trade-offs between the gain, IIP3 and stability in LNA design. Have to mention that this inductive degeneration does not seriously impact Noise Figure performance, as resistive degeneration does.

At high frequencies this inductance will be achieved with small strip lines (stubs) connected directly to the emitters of the transistor.

The inductive reactance of the stubs is usually no greater than  $10\ \Omega$  and the line lengths are typically  $\sim 2\text{mm}$  or less with characteristic impedances  $50\ \Omega$  or greater.

- To design an LNA for minimum Noise Figure, determine (experimentally or from the data sheet) the source resistance and bias point that produce the minimum Noise Figure for that device.
- Then force the actual source impedance to “look like” that optimum value with all stability considerations still applying. If the Rollet stability factor ( $K$ ) is calculated to be less than 1 ( $K$  is defined as a figure of merit for LNA stability), then you must be careful in choosing the source and load-reflection coefficients.
- A typical method used in designing input matching network is to display noise circles and gain/loss circles of the input network on the same Smith chart. This provides a visual tool in establishing an input matching network for the best Input Return Loss and noise trade off.

### **Using Noise Figure from Datasheets**

Generally, for microwave transistors following a datasheet the minimum Noise Figure ( $F_{\min}$ ) at higher frequencies is based on measurements, while the  $F_{\min}$ s at lower frequencies are extrapolated.

- $F_{\min}$  represents the true minimum Noise Figure of the device when the device is presented with an impedance matching network that transforms the source impedance, typically  $50\Omega$ , to an impedance represented by the reflection coefficient  $\Gamma_{\text{opt}}$ .
- The designer must develop a matching network that will present  $\Gamma_{\text{opt}}$  to the device with minimal associated circuit losses. To accomplish this have to minimize the number of components needed on the LNA input.
- The Noise Figure of the completed amplifier is equal to the Noise Figure of the device plus the losses of the matching network preceding the device.
- The Noise Figure of the device is equal to  $F_{\min}$  only when the device is presented with  $\Gamma_{\text{opt}}$ .
- If the reflection coefficient of the matching network is other than  $\Gamma_{\text{opt}}$ , then the Noise Figure of the device will be greater than  $F_{\min}$ .
- The losses of the matching networks are non-zero and they will also add to the noise figure of the device creating a higher amplifier noise figure. The losses of the matching networks are related to the Q of the components and associated printed circuit board loss.
- $\Gamma_{\text{opt}}$  is typically fairly low at higher frequencies and increases as frequency is lowered.
- For FET devices larger gate width devices will typically have a lower  $\Gamma_{\text{opt}}$  as compared to narrower gate width devices.
- Typically for FETs, the higher  $\Gamma_{\text{opt}}$  usually infers that an impedance much higher than  $50\Omega$  is required for the device to produce  $F_{\min}$ . At VHF frequencies and even lower L Band frequencies, the required impedance can be in the vicinity of several thousand ohms. Matching to such high impedance requires very hi-Q components in order to minimize circuit losses.

#### 4. Output Matching

The last step in LNA design involves output matching of the transistor. An additional resistor, either in series or parallel, has been placed on the collector of the transistor for circuit stabilization.

Conjugate matching has been exclusively used for narrow band LNA design to maximize the gain out of the circuit.

With additional IP3 requirement forced on the LNA, the trade-off between IP3 and gain must be considered.

Linearity matching is widely known by high-power amplifier designers. The so-called load pulling is used to establish IP3 and gain impedance contours. The load pulling can be realized by using the non-linear Spice model of the transistor with simulation software.

Harmonic balance can be used for establishing two-tone environment.

The load pulling method sweeps impedance of the whole Smith chart and plots contours of the constant gain and IP3 numbers. The optimal gain impedance does not match the optimal IP3 point, which means that the design will have to be realized by means of a trade off.

Typically, the designer should design the LNA circuit at the point where the gain does not degrade as much, and the IP3 is still respectable. If one were to draw a line between the optimal gain and IP3 impedance points, every point on that straight line will represent a good area of trade-off, with the ends representing the two optimal points.

The rule of thumb for 1dB gain compression point ( $P_{1dB}$ ) and IP3 is:

$$IP3 = P_{1dB} + 10 \text{ [dBm]}$$

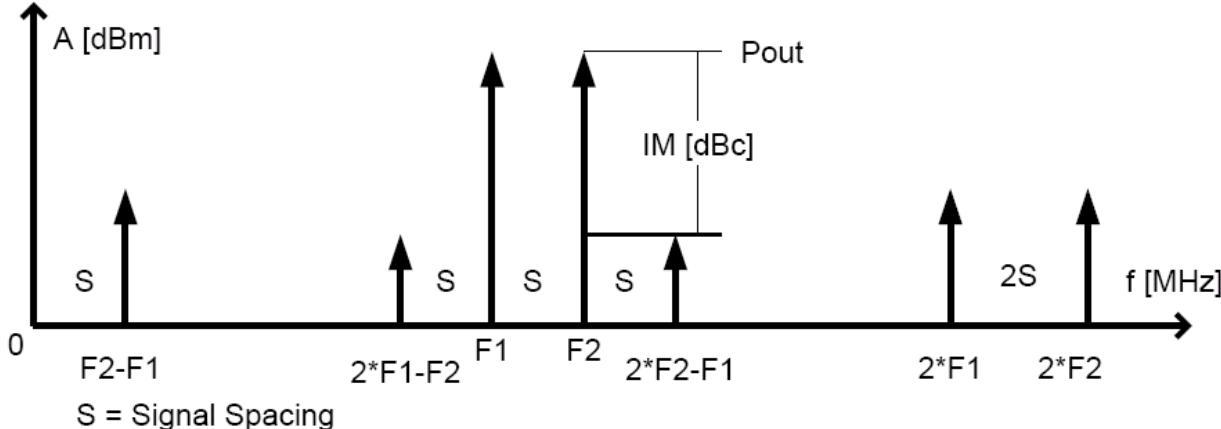
That means that by knowing the gain compression point ( $P_{1dB}$ ), can estimate the IP3 levels.

- The 10dB rule can further be improved with appropriate bypassing of the base and the collector.

As previously indicated, the IIP3 is established by injecting two equal-in-magnitude signals with small frequency offset ( $S$ ) into an active circuit.

As the active circuit approaches non-linear region, close to  $P_{1dB}$ , the two carriers will generate distortion products, both in and out of band.

In example below we have two signals, with output levels  $P_{out}$  and frequencies  $F_1$  and  $F_2$ .

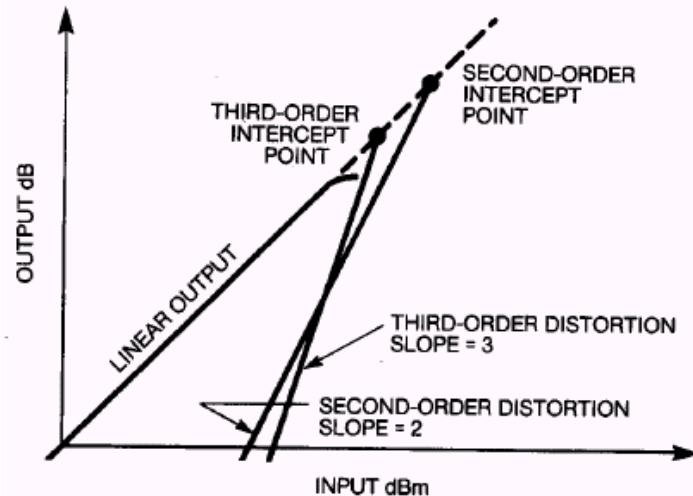


$$OIP3[\text{dBm}] = P_{out}[\text{dBm}] + IM[\text{dBc}] / 2$$

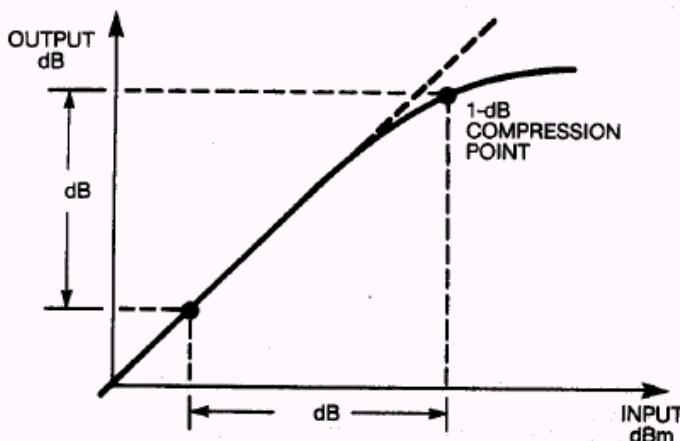
Where IM [dBc] is the difference in amplitude between one of the two equal amplitude test tones, present at the amplifier output, and the level of the highest 3rd-order distortion product.

- For every dB increase in input power, the third-order products (IM3) will increase 3dB.
- For every dB increase in input power, the second-order products (IM2) will increase 2dB.

Plotting third-order products versus input power predicts a 3:1 response which intersects the 1:1 response at the third-order intercept point.



Second and Third-Order Distortion Slopes



1dB Gain Compression Point (P1dB)

The relation between Input-IP3 (IIP3) and Output-IP3 (OIP3) is defined as:

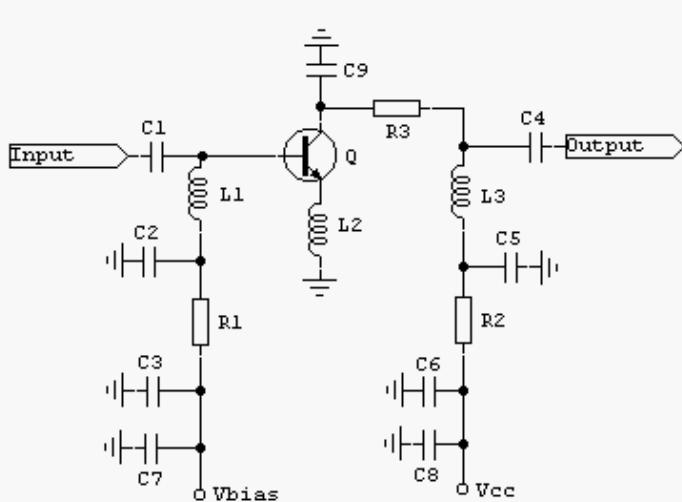
$$\text{IIP3}_{\text{[dBm]}} = \text{OIP3}_{\text{[dBm]}} - \text{Gain}_{\text{[dB]}}$$

The low frequency IM2 products (F2-F1), can modulate the base-emitter and collector-emitter LNA supply voltages. To improve the linearity the fluctuation of the base and the collector shall be stabilized with low impedance at so called video frequencies or baseband frequencies (between DC and usually up to 40MHz).

- The designer should exhibit caution during bypassing design. A poor selection of the by-pass capacitors could also degrade IP3 performance.
- As a rule of thumb, the impedance of bypassing circuit should be lower than 25% of the input impedance of the transistor at particular frequency spacing.

- Although preserving the gain performance of the LNA, the bypassing method (also known as an envelope termination technique) can improve LNA's IIP3 performance without increasing current consumption.

## 5. LNA components and the effect on IP3



- Any mismatch due to noise matching C1/L1 improves Input-IP3.
- Increasing L2 reduces gain and improves Input-IP3, but watch for microwave oscillation with excessive inductance.
- C9 can use to improve IP3- provides gain roll-off at  $2^*F_1$  or  $2^*F_2$ .
- Printed circuit board losses R3 provide Q1 stability while reducing IP3. R3 less than  $27\ \Omega$  for about a dB reduction in IP3.
- C3 and C6 provide a HF/VHF termination for Q1. Depending on spacing of signals used to test IP3, values may not be large enough – may necessitate additional low frequency bypassing in the form of C7 and C8. Typical values are 0.01 to 0.1 uF.
- The combination of C2/C5, C3/C7 and C6/C8 must provide low impedance at  $F_2 - F_1$ . May have to add resistance between caps to decrease Q.
- C7 and C8 also used to minimize power supply noise from modulating the DC.
- Capacitor C2, C3 and C7 performs the low-frequency bypass function and an improvement in IP3 of approximately 5 to 10 dB can be expected by using this method. Using extra charge storage on the drain may see the same effect, but the results are not nearly as dramatic.
- The closer together the two input test tones F1 and F2 are in frequency, the lower frequency the product or beat tone ( $F_2 - F_1$ ) is. Therefore, as input test tones F1 and F2 come closer together, more capacitance is needed to achieve best possible bypassing of the low frequency product ( $F_2 - F_1$ ).

For a test tone separation of 1MHz, 0.1uF was found to be more adequate for this application. For best results, the transistor should see a low impedance path at low frequencies between this additional bypass caps and its terminals. For this reason, a coil rather than a high value resistor is used to bring the gate bias voltage and isolate the RF from the DC bias network. For example, a value of 15nH for L1, has negligible impedance up to tens of MHz, but provides enough impedance at 2 GHz to nearly isolate the gate of the transistor from the bias network within LNA's normal operating frequency range.

It is important to note that bypassing the F2 – F1 product as described here does not affect the compression point of the amplifier, but only the IP3 (3rd-order intercept point). As a result, if this bypassing used, the general rule of thumb stating that are approximately a 10dB difference between IP3 and 1 dB gain compression point (P1dB) is no longer valid.

## 6. Real issues in LNA design

- An LNA is a design that minimizes the Noise Figure of the system by matching the device to its noise matching impedance, or Gamma optimum ( $\Gamma_{\text{opt}}$ ).
- Gamma optimum ( $\Gamma_{\text{opt}}$ ) occurs at impedance where the noise of the device is terminated.
- All devices exhibit noise energy. To minimize this noise as seen from the output port, one must match the input load to the conjugate noise impedance of the device. Otherwise the noise will be reflected back from the load to the device and amplified. While this gives a minimum noise figure, it often results in slightly reduced gain as well as possibility increasing the potential instabilities.

Noise match often comes close to S11 conjugate ( $S11^*$ ) under non-feedback conditions. As a result, the input impedance to the amplifier will not be matched to 50 ohms.  $\Gamma_{\text{opt}}$ , as presented in data sheets, is the actual measured load at which the minimum noise figure is found.

- Noise Figure for BJT LNA increases more rapidly than FET LNA, as Collector/Drain current is increased.
- FETs allows for better trade-off between high-linearity and low Noise Figure than BJTs.

A further complication on LNA design is that the input load of the amplifier is usually less than ideal. It is either connected to an antenna, which can change its impedance with changing the environment, or to a filter, which by very physics of a reflective network will have very bad match out of band. These mismatches could cause the device to become unstable out of band and some cases in band. As the gain of the device increases, the difficulties in yielding a stable design become increasingly more challenging.

To avoid overloading the LNA, an input filter is commonly used. Since the device is not matched to  $S11^*$ , the input of the LNA will not be 50 ohms. This can cause distortions in the pass band of the filter when connected to the input of the LNA, as filter are intended to operate in their characteristic impedance, typically 50 ohms.

Printed inductors or transmission lines are free as compared to SMT inductors, which typically cost 10 to 25 times as much as resistors or capacitors in volume. Printing an inductor is easy and results in highly repeatable results. Printed inductors usually exhibit poor Q due to the lossy dielectric, and, if a ground plane exists, they are no more than a high impedance transmission line.

As shown a transmission line can replace an inductor to some degree, but inductors and high impedance transmission lines have a different trajectory on the Smith Chart. High impedance transmission line can be made to look more like printed inductors in cases where the backside of the PCB is suspended away from a grounded chassis. This is accomplished by removing the backside ground plane of the PCB directly under the printed inductor. In this case beware of digital noise coupling into the input of the LNA from circuitry on the opposite side.

- The next concern is what load impedance to match. Remember matching to the conjugate of  $S_{22}^*$  is only valid if the input is conjugate matched. Since  $S_{12}$  is non-zero, whatever load is present to the input will cause the output load change.
- Another issue is stability, especially if a filter is going to be used at the input. The output port can potentially give difficulties since the input is very restricted by its match.

The designer must replace the ideal sources in the bias circuit and ideal values in the matching circuit with equivalent real components. This often presents the designer with a new set of problems. First, the bias network must be robust enough to function properly over a range of power-supply voltages and temperatures. This introduces additional complexity into the bias network. The real components in the bias network the resistors and large capacitors operate at DC voltages, so frequency effects are not a problem. The matching network, however, contains real capacitors and inductors that operate at RF frequencies.

Real components differ from ideal ones in several respects. First, real components have a price associated with them. There is a trade-off between price and performance of these parts. The competitiveness of today's markets often forces designers to use inexpensive components in their designs.

- Real discrete components have a finite resistance called Equivalent Series Resistance (ESR). The ESR introduces losses that result in lower gain and noise figure. Although typically only a few tenths of an ohm in value, ESR will affect the matching networks.
- Discrete components also have a Q value, measured at a particular frequency that can contribute to unwanted resonance.
- High-Q networks are sensitive to variations in process, voltage, temperature, and component value.
- A component's Series Resonant Frequency (SRF) is the frequency where it will behave erratically. For example, if an inductor is operated at or above its SRF, it might behave as a capacitor. To avoid this, select components where the SRF is much higher than the operating frequency.
- Also, lead-through-hole parts have leads that add series inductance to a design, and surface-mount parts have pads that add shunt capacitance to a circuit.
- Another issue is that of packaging a completed design. If the circuit is to be integrated and sold as an Integrated Circuit (IC), it must be packaged. The package introduces several negative effects. In an IC, the bond wires add unwanted inductance (L) and the bond pads add unwanted capacitance (C).
- Isolation between pins in the package is also important. Lack of pin-to-pin isolation in a feedback circuit can lead to major reliability problems and stability concerns. The additional inductance in the emitter of the collector-emitter section can severely degrade the noise figure of the circuit.
- Additionally, several grounds are usually needed to improve the performance of RF circuits, but the package has a limited number of pins. After using the input, output, and power-supply pins, there may not be enough ground pins to accommodate an adequate design.

All of these factors can degrade the circuit's performance from the ideal, and the designer must carefully take them into account.

## 7. CMOS LNA Design

- A few comparison characteristics between CMOS and BJT LNAs:
  - The DC currents of CMOS and BJT LNA's are close; therefore, the transconductance ( $g_m$ ) of CMOS transistor is lower than the BJT one's.
  - The  $g_m/I$  ratio of CMOS is lower than that of BJT.
  - In CMOS technologies, a high  $f_T$  is achieved through a smaller  $C_{gs}$ , while in BJT technologies the same  $f_T$  is obtained through a higher  $g_m$ .
  - Smaller  $C_{gs}$  means CMOS tuned circuits tend to have higher Q, a disadvantage in withstanding component or process variation.

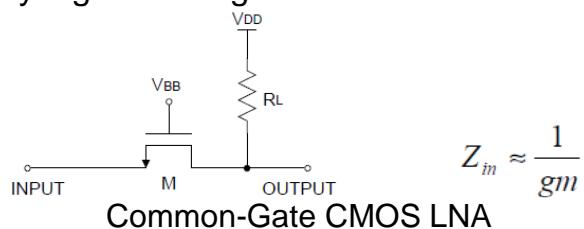
The CMOS LNA input quality factor (independent of  $L_s$ ) is defined as follows:

$$Q_{gs} = \frac{1}{\omega_o C_{gs} Z_0}$$

There are two types of methods commonly used to design an LNA in CMOS circuits:

- Common-Gate
- Cascode amplifier

While the Common-Gate stage provides a wide-band input matching and is less sensitive to parasitics, it has an inherently high noise figure.

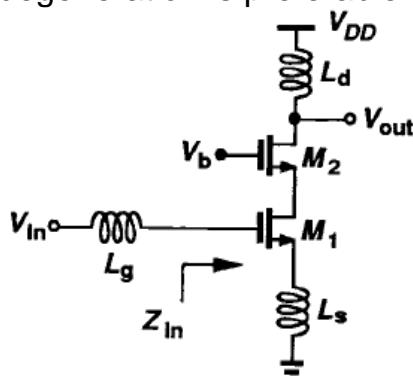


$$Z_{in} \approx \frac{1}{gm}$$

Common-Gate CMOS LNA

- With the increasing of the operating frequency, the parasitic transistor capacitance  $C_{gs}$  starts playing roles, which degrades the amplifier performance in the high frequency. In the narrow band application, a shunt inductor is added in the input to resonate with  $C_{gs}$  to have a good impedance matching in the designed frequency.
- Due to the lower quality factor of the resonant network, Common-Gate it is more robust against the process and electrical variation.
- Due to the missing of the  $C_{gd}$  path from the input to the output, the Common-Gate LNA shows better reverse isolation and stability versus Common-Source LNA.

Therefore, in most CMOS applications where the noise figure is critical issue, a cascode LNA with inductive degeneration is preferable.



$$Z_{in}(s) = L_s \frac{g_m}{C_{gs}} + \frac{1}{C_{gs}s} + L_s s$$

Cascode CMOS LNA

- For a given unit-gain frequency as we lower the bias current, the noise figure decreases.
- For a given Q, higher  $g_m$  improves the noise figure.
- For a given source resistance of 50 ohms, as we reduce  $L_s$ , unit-gain frequency increases but the minimum value of  $L_s$  is limited by parasitic and sensitivity issues.
- By proper choice of  $g_m$ ,  $L_s$ , and  $C_{gs}$ , the input resistance can be equal to 50 ohms source resistance and the input reactance (imaginary part of impedance) can be resonated out by a series inductor ( $L_s$ ).
- Inductor degeneration ( $L_s$ ) also improves the linearity by forming a negative series feedback.
- As we lower the bias current, while keeping unit-gain frequency constant,  $C_{gs}$  decreases, leading to higher Q.
- A high Q matching networks has several drawbacks:
  - Circuit becomes very sensitive to component variations and parasitics.
  - The input matching circuit which in this case contains a series inductor, inserts a large amount of loss at the input (even for a high-Q off-chip inductor)
- Another source of noise in the cascode topology is the noise introduced by the cascode device, M2, added to improve stability of the amplifier.
  - At high frequencies the capacitance at the drain of M1 reduces the impedance of this node, increasing the output noise, so to minimize the noise is very important to minimize this capacitance.
  - To improve the noise performance of the cascode design, the parasitic capacitance at the drain of M1 is resonated out by adding an inductor to the source of cascode.
  - This inductor should be sized carefully in order to resonate the unwanted capacitances at the desired frequency of operation.

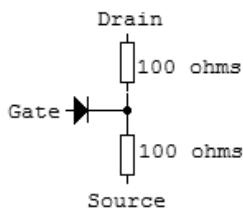
## 8. GaAs FET LNA Design

In case of GaAs FETs generally the input matching network transforms 50 ohms to the input impedance of the transistor, which typically at VHF frequencies is about 3000 ohms. A high-quality circuit usually is used to transform the impedance up and in the same time to filter out unwanted signals.

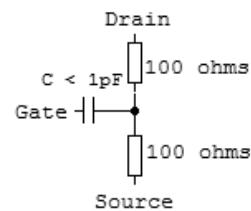
As frequency increases to microwave region (above 10GHz), the GaAs FET input impedance drops to few tens of ohms. HEMT's type transistors have even higher impedance at VHF frequencies (about 5000 ohms) dropping to 50 ohms above 10GHz.

### Biassing the GaAs FET LNA

The equivalent of the input of a GaAs FET is a small Schottky diode, when the input of a HEMT is basically a small Tunnel diode. In normal operation, at RF frequencies, these diodes are negatively biased, effectively making the gate a low value capacitor.



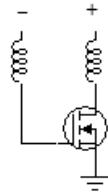
DC Equivalent Model of a GaAs FET



RF Equivalent of a biased GaAs FET

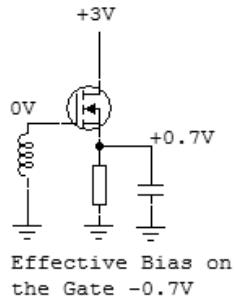
Are two common ways to supply the required negative voltage to the gate of a GaAs FET:

**Method 1:**



- This method requires power supply to be more complex than the LNA, but usually is the preferred method in most of the designs.
- Have the advantage to ground directly the source, and have excellent control of all DC parameters of the GaAs device.
- One of biggest issue of this biasing method is, the gate negative supply must turn-ON first (before drain voltage), and turn-OFF last, making the power supply to be complex.

**Method 2:**

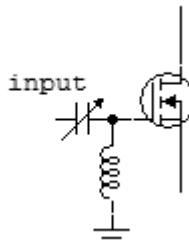


- The self-bias method makes the voltage drop across the source resistor raises the substrate of the device above DC ground, giving to the gate a reverse bias.
- The method has the advantage using only one positive supply and also the thermal run-away is almost eliminated.
- The device operates in a current limiting mode; more current gives a greater negative bias to the gate, turning OFF the device.
- The RF bypassing of the source becomes difficult at microwave frequencies, and need to use capacitors with low parasitic inductance.

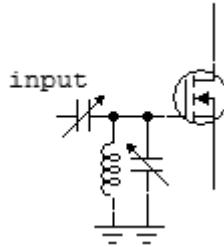
### GaAs FET LNA - Input Matching

There are few topologies for matching of GaAs FETs, each of them having pros and cons.

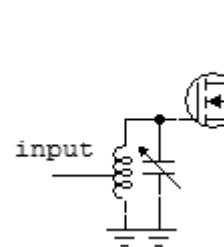
**Remember that, not always want to tune for best Noise Figure or highest gain.**



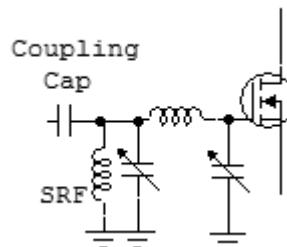
- This low insertion loss and simple input match circuit, works well up to UHF frequencies. There are not too many choices of tuning and a match from 50 ohms to Gamma Optimum (best Noise Figure) depends on the FET's internal stray capacitance from gate to the ground.



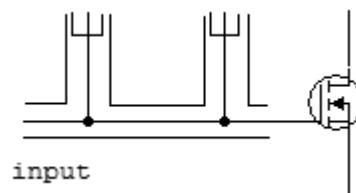
- This input match gives a high performance below 500MHz.
- Can get the best Noise Figure, Gamma Optimum that can be reached.
- Because the input impedance of the GaAs FET below 500MHz is high, the Q or Bandwidth of the input circuit can be varied with little impact on the Noise Figure.



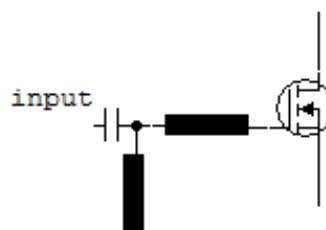
- The tapped L input is fine for VHF range high performance LNAs.
- It has very low loss, input is grounded, but the circuit has very wide bandwidth.



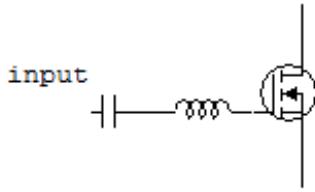
- The Pi input matching network works from low frequencies up to few GHz, and virtually can match any impedance.
- In the Pi matching network, the second capacitor is tuned at a very low value, due to the FET input stray capacitance.
- This method requires high quality capacitors.



- The stub circuit is versatile and capable of matching almost of any input impedance, but the assembly is very large below 1GHz and relative difficult to supply the bias voltage.



- The microstrip stub circuit is limited to microwave frequencies, is narrowband, but has great repeatability.



- Just a small series inductance in the gate lead can match the impedance in the 1GHz to 2GHz region. The circuit is simple and has low loss, broad bandwidth and excellent Noise Figure.

## GaAs FET LNA - Construction and Operating Tips

- Always leave the LNA powered up. When operating, the gate of the GaAs FET is negatively biased at 0.5V-0.7V. The FET is damaged when the gate has positive voltage and conducts current. You need enough RF at the input to overcome the negative bias before damage the transistor.  
Typically, the circuit needs 4 times more RF leakage power to damage a turned-ON GaAs than a turned-OFF GaAs FET device.
- Use very short leads when the circuit works at microwave frequencies.
- Use small coils and avoid mutual coupling between them placing the inductors 90° to each other.
- Use Ferrite beads which behave as both, RF absorber and a low-Q RF choke.
- If use trimmer caps, connect their rotors to the ground or to the connectors.  
When the microwave LNA is placed inside of a shielded box this become a cavity at some frequency, and the LNA might become an oscillator. Absorbers placed inside of the box help preventing this happen.

## Typical characteristics of different configurations of High-Frequency GaAs FET LNAs

Characteristic	Common-Source (CS)	Common-Gate (CG)	Cascode
<b>NFmin</b>	Best	Better	Good
<b>Gain</b>	Moderate	Low	High
<b>Bandwidth</b>	Narrow	Very Wide	Wide
<b>Stability</b>	Compensation often required	RF Decoupling of the Gate is important	Good
<b>Reverse Isolation</b>	Lower	Moderate	Best

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