AN427

# Si443x RX LNA Matching

## 1. Introduction

The purpose of this application note is to provide a description of the impedance matching of the RX differential low noise amplifier (LNA) on the Si443x family of RFICs.

We desire to simultaneously achieve two goals with the matching network:

- Match the LNA input to a 50 Ω source impedance (i.e., the antenna)
- Provide a single-ended to differential conversion function (i.e., a balun)

The matching procedure outlined in this document will allow for achieving the goals listed above.

For those users who are not interested in the theoretical derivation of the match network, but are just concerned with quickly obtaining matching component values, refer to the Summary Tables shown in "4.1.8. Summary Table of 3-Element Match Network Component Values vs. Frequency" on page 15 and "4.3.7. Summary Table of 4-Element Match Network Component Values vs. Frequency" on page 21.

Measurements were performed on the Si4432-V2 but are applicable to other revisions of the device.

## 2. Match Network Topology

The LNA on the Si443x family of chips is designed as a differential amplifier and thus has two input pins (RXp and RXn) on the RFIC. It is necessary to design a network that not only provides a conjugate match to the input impedance of the LNA but also provides a balanced-to-unbalanced conversion function (i.e., a balun).

We will consider use of two basic topologies of matching networks, with two variations on one of the topologies.

### 2.1. Three-Element Match Network

The simplest match network that may be fabricated from discrete components is comprised of three discrete elements. We will consider two different forms of this 3-element match network: one with a highpass filter (HPF) response, and one with a lowpass filter (LPF) response.

The RXp and RXn inputs of the Si443x RX LNA are internally capacitively-coupled; there is no need to provide external coupling capacitors to act as dc blocks.

#### 2.1.1. Three-Element HPF Match Network

A 3-element ($C_1$-L-$C_2$) HPF matching network is shown in Figure 1. This matching network has the virtue of requiring a minimum number of components but results in slightly sub-optimal performance. It is not theoretically possible to achieve a perfectly balanced single-ended-to-differential conversion function with this network. As will be demonstrated, the waveforms obtained at the RXp and RXn inputs to the RFIC will not be exactly 180° out of phase; the result is a very slight loss in conversion gain in the LNA and a small drop in overall sensitivity of the RFIC. The reduction in performance is typically less than 0.5 dB; many customers may view this as an acceptable trade-off for the reduction in the bill of materials (BOM).

  

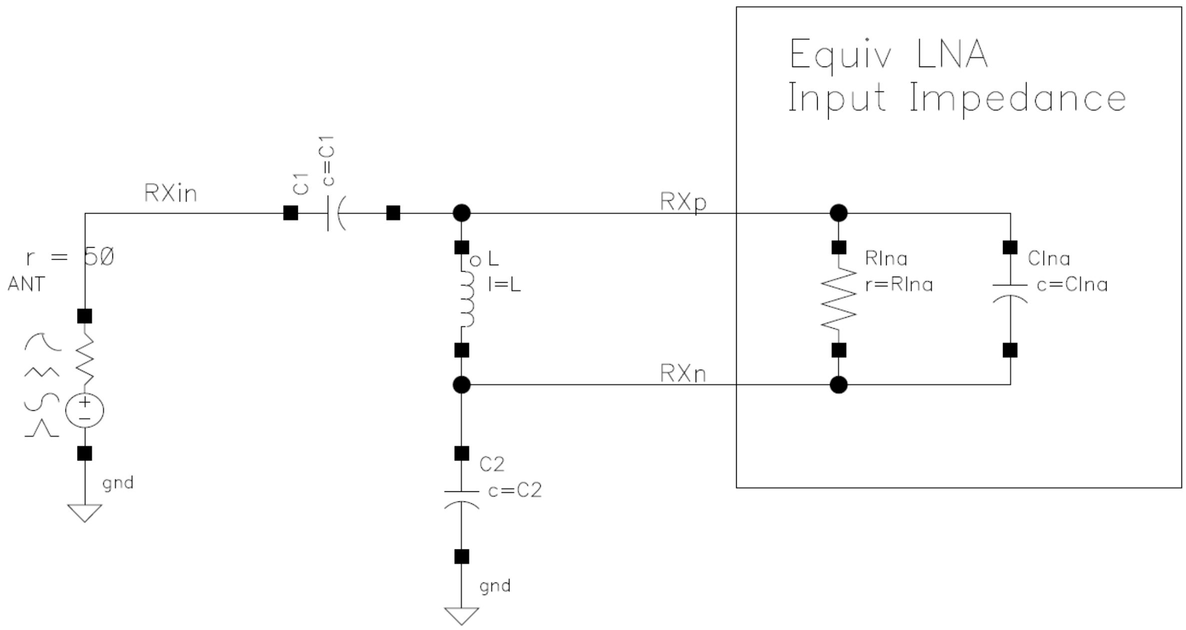


**Figure 1. HPF Three-Element Match Network**

#### 2.1.2. Three-Element LPF Match Network

A 3-element ($L_1$-C-$L_2$) LPF matching network is shown in Figure 2. This matching network also requires a minimum number of components but again results in slightly sub-optimal performance for the same reason as before: it is not theoretically possible to achieve a perfectly balanced single-ended-to-differential conversion function with this network. As will be demonstrated, this matching network may not be realizable at all operating frequencies.

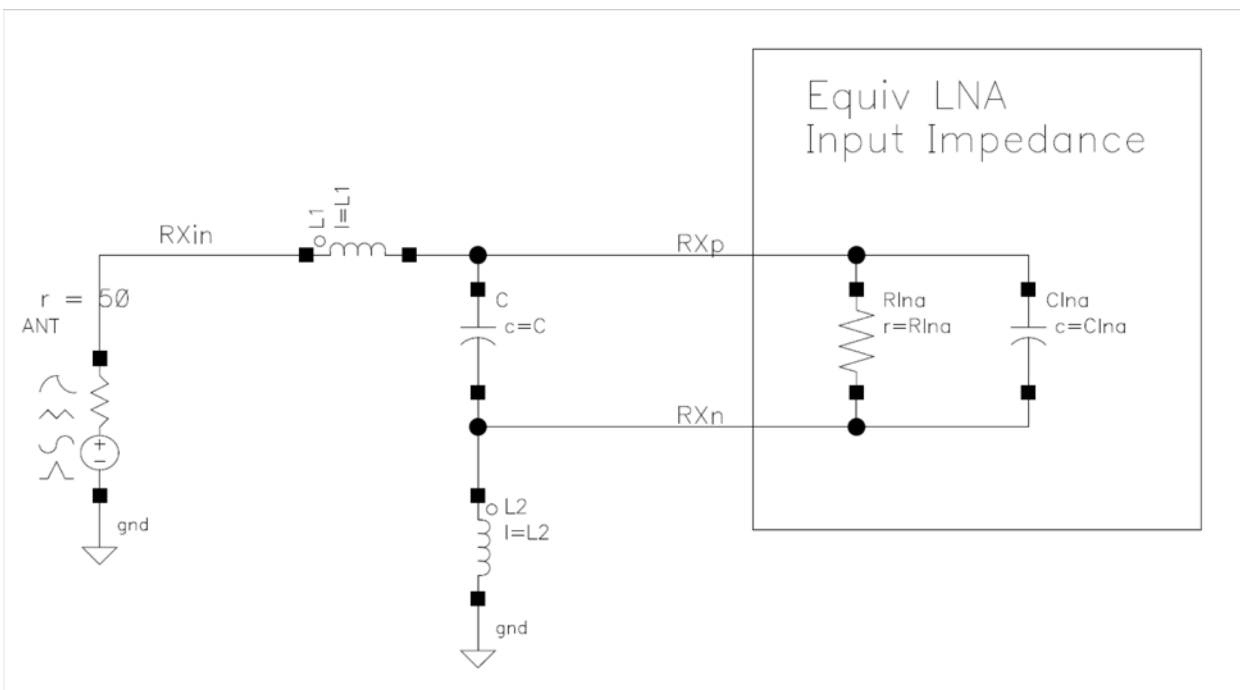


**Figure 2. LPF Three-Element Match Network**

## 2.2. Four-Element Match Network

If the customer is concerned with obtaining absolutely optimal performance, the 4-element match network of Figure 3 is recommended. This match network can provide theoretically perfect phase balance between the RXp and RXn inputs (exactly 180° out-of-phase), thus optimizing LNA conversion gain and receiver sensitivity. The only drawback is the addition of one more component (an inductor) to the BOM.

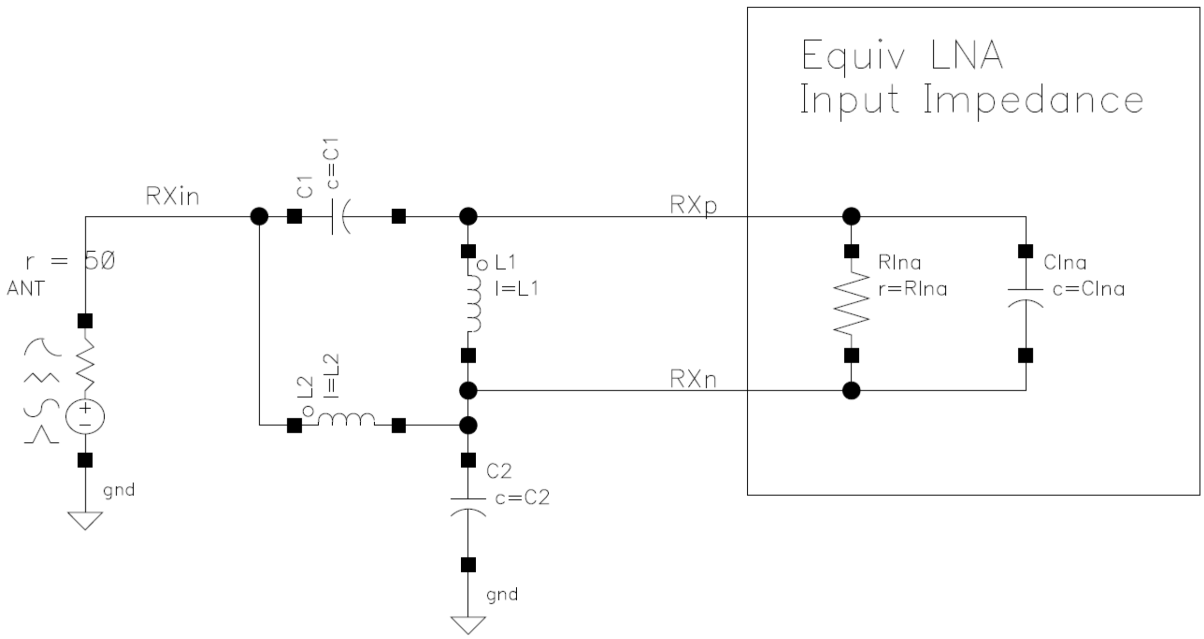


**Figure 3. Four-Element Match Network**

# 3. Si4432 Differential LNA Input Impedance

Silicon Labs has measured the differential input impedance of the Si4432 RX LNA with no matching network, directly at the RXp/RXn input pins of the RFIC.

## 3.1. LNA Input Impedance in RX Mode

The plot shown in Figure 4 shows the measured input impedance in the RX mode of operation over the 240 to 960 MHz frequency band, with markers placed at various points throughout the frequency range.

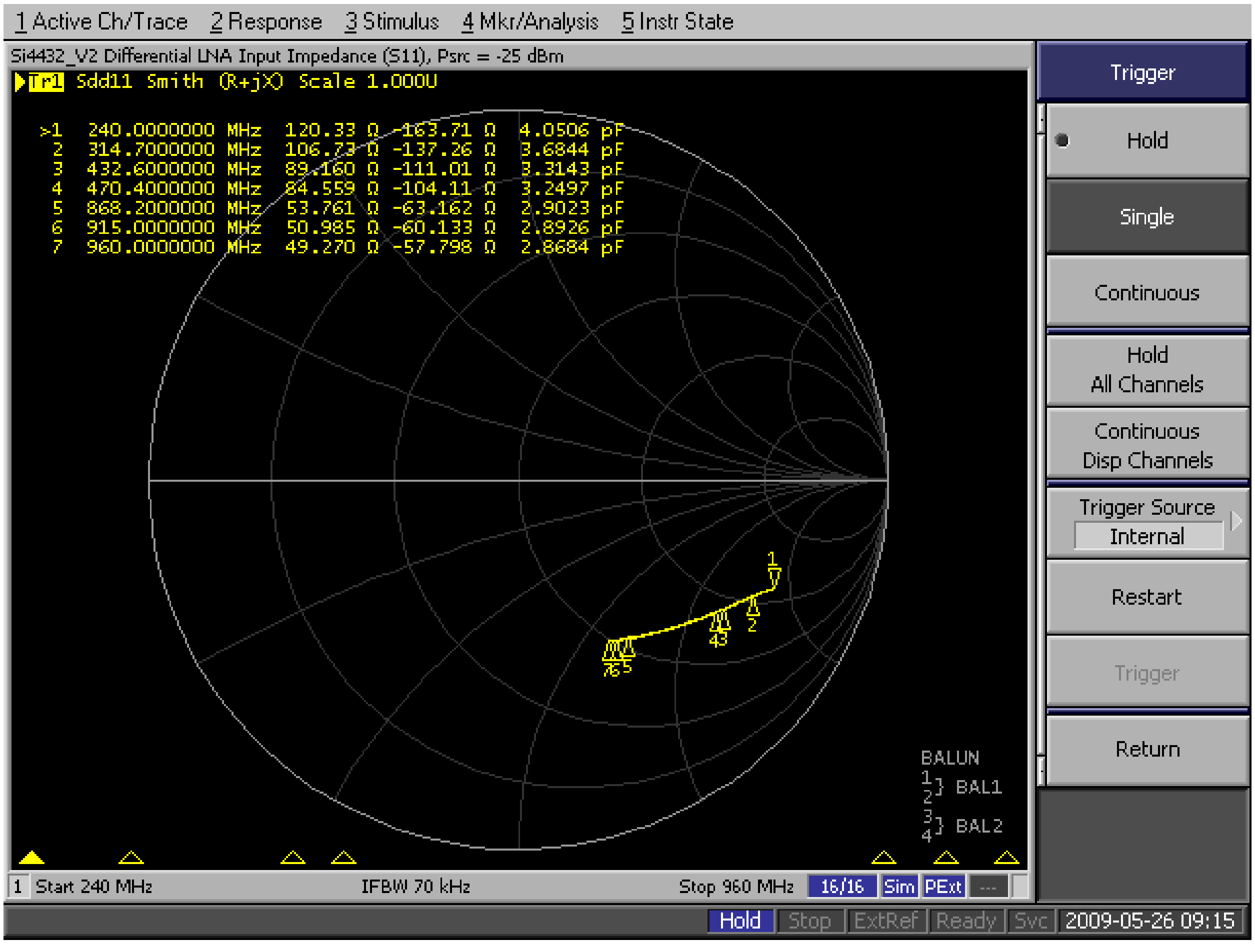


**Figure 4. Si4432 Differential RX LNA Input Impedance 240-960 MHz (RX Mode)**

As can be seen from this curve, at any given single frequency the input impedance of the LNA may be considered as some resistance, in parallel with a small amount of capacitance. That is to say, the input impedance of the LNA falls in the capacitive half of the Smith Chart across its entire operating frequency range. This measured data was obtained with a source power level from the Network Analyzer of –25 dBm; care must be taken to ensure that the input drive level to the LNA does not exceed its linear operating range, else the measured impedance data may be distorted.

The impedance curve shown in Figure 4 cannot be described by a single fixed value of resistance, placed in parallel with a single fixed value of capacitance. The equivalent values of parallel resistance and capacitance ($R_{LNA}$ and $C_{LNA}$ in Figure 1 through Figure 3) vary as a function of frequency. However, the variation with frequency is not rapid; it is possible to construct a moderately wideband (~100 MHz) matching network by simply designing for the value of $R_{LNA}$ and $C_{LNA}$ in the center of the desired frequency range.

From the differential input impedance values (Z = R + jX) shown in Figure 4, we first need to calculate the equivalent input admittance, where Y = 1/Z = G + jB. It is then a simple matter to calculate the values of the equivalent input resistor and capacitor (i.e., $R_{LNA}$ and $C_{LNA}$ in Table 1) as $R_{LNA} = 1/G$ and $C_{LNA} = B/(2\pi F_{RF})$.

Silicon Labs has performed these computational steps on the measured $S_{11}$ data of Figure 4, and the resulting equivalent values of $R_{LNA}$ and $C_{LNA}$ are shown in Table 1 as a function of frequency.

**Table 1. Equivalent $R_{LNA}$–$C_{LNA}$ from 240–960 MHz**

| Freq | $R_{LNA}$ | $C_{LNA}$ |
|---|---|---|
| 240 MHz | 343 Ω | 2.63 pF |
| 300 MHz | 292 Ω | 2.34 pF |
| 315 MHz | 283 Ω | 2.30 pF |
| 350 MHz | 263 Ω | 2.19 pF |
| 390 MHz | 244 Ω | 2.09 pF |
| 400 MHz | 240 Ω | 2.07 pF |
| 433 MHz | 227 Ω | 2.01 pF |
| 450 MHz | 220 Ω | 1.98 pF |
| 470 MHz | 213 Ω | 1.96 pF |
| 500 MHz | 204 Ω | 1.92 pF |
| 550 MHz | 189 Ω | 1.87 pF |
| 600 MHz | 176 Ω | 1.83 pF |
| 650 MHz | 165 Ω | 1.80 pF |
| 700 MHz | 155 Ω | 1.77 pF |
| 750 MHz | 146 Ω | 1.74 pF |
| 800 MHz | 138 Ω | 1.72 pF |
| 850 MHz | 131 Ω | 1.70 pF |
| 868 MHz | 128 Ω | 1.68 pF |
| 900 MHz | 125 Ω | 1.69 pF |
| 915 MHz | 122 Ω | 1.68 pF |
| 950 MHz | 118 Ω | 1.66 pF |
| 960 MHz | 117 Ω | 1.66 pF |

### 3.2. LNA Input Impedance in STANDBY Mode

The plot shown in Figure 5 below shows the measured input impedance in the STANDBY mode of operation over the 240 to 960 MHz frequency band, with markers placed at various points throughout the frequency range.

Although not covered in this Application Note, we will mention that it is possible (with certain restrictions) to construct a matching network that ties the RXp/RXn LNA input pins directly to the TX output pin, without the need for a TX/RX switch. Knowledge of the LNA impedance in the OFF state is necessary to construct this “direct tie” match.

We will not further consider the impedance of the LNA in its OFF state in this Application Note.

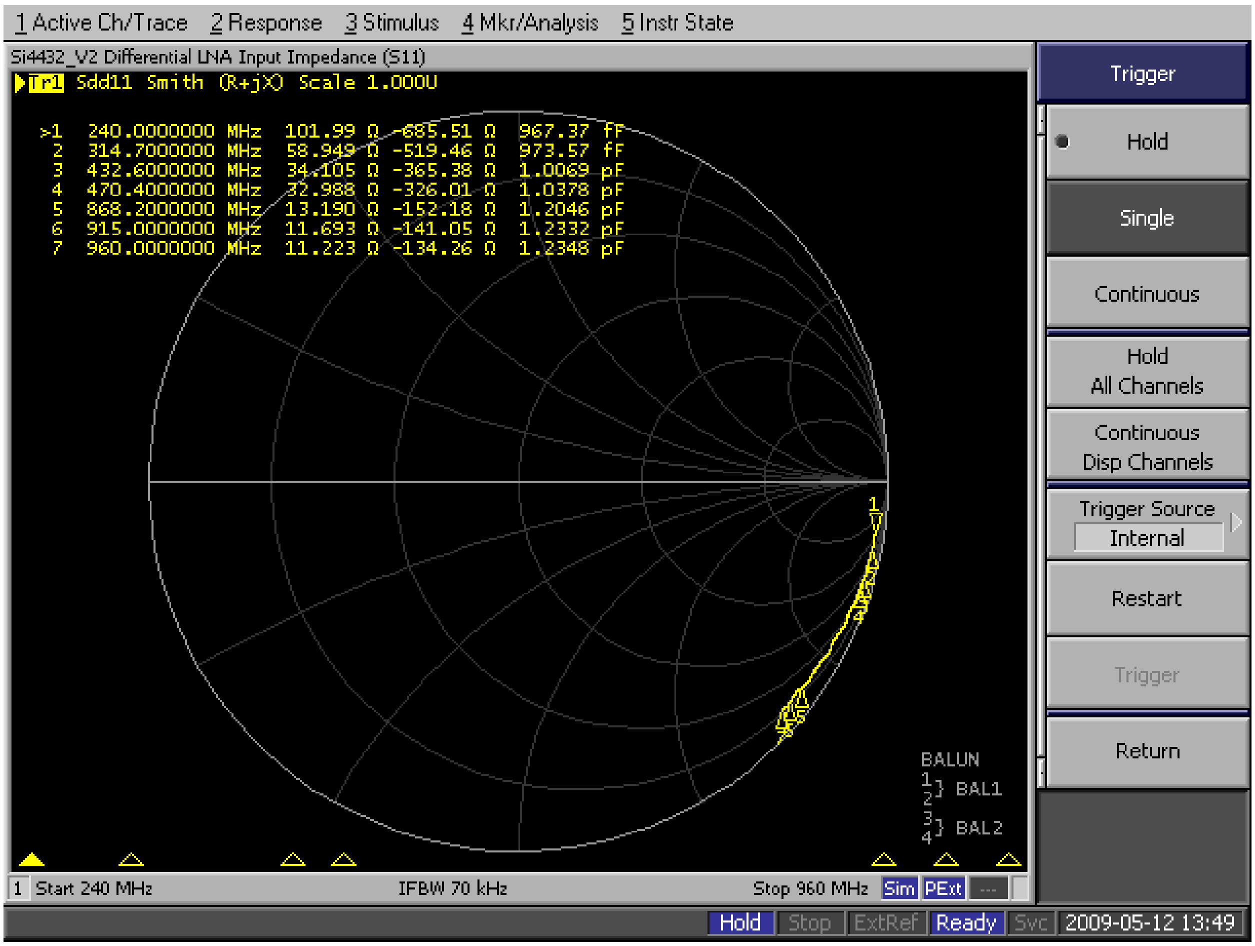


**Figure 5. Si4432 Differential RX LNA Input Impedance 240–960 MHz (STDBY Mode)**

# 4. LNA Matching Procedure for the Si443x

Armed with the measured values of unmatched differential input impedance of the Si4432 LNA, we can now proceed with constructing a matching network. For demonstration purposes, we chose a frequency of 315 MHz to illustrate our examples.

We have mentioned two different types of 3-element matching networks: one with a highpass filter (HPF) response (refer to Figure 1) and one with a lowpass filter (LPF) response (refer to Figure 2). The associated response of each type of network is somewhat intuitive, based upon the observation of the placement of series and shunt inductors and/or capacitors. For the HPF match network, greater attenuation will be provided to far out-of-band signals located below the desired frequency of operation, as compared to far out-of-band signals located above the desired frequency of operation. The opposite of course holds true for the LPF match network.

Such a choice of alternate matching networks may be of interest in those applications that operate in the presence of strong out-of-band interfering signals. In such a scenario, it may be desirable to gain some small amount of attenuation of these interfering signals, simply due to the choice of matching architecture.

## 4.1. Three-Element Matching Procedure (HPF Architecture)

The matching procedure for the 3-element ($C_1$-L-$C_2$) HPF match is outlined below.

### 4.1.1. Step #1: Plot the LNA Input Impedance

Start with the equivalent parallel $R_{LNA}$-$C_{LNA}$ circuit values shown in Figure 6. At 315 MHz, we find $R_{LNA}$ = 283 Ω and $C_{LNA}$ = 2.30 pF. It is useful to plot this value on a Smith Chart, as shown in Figure 6.

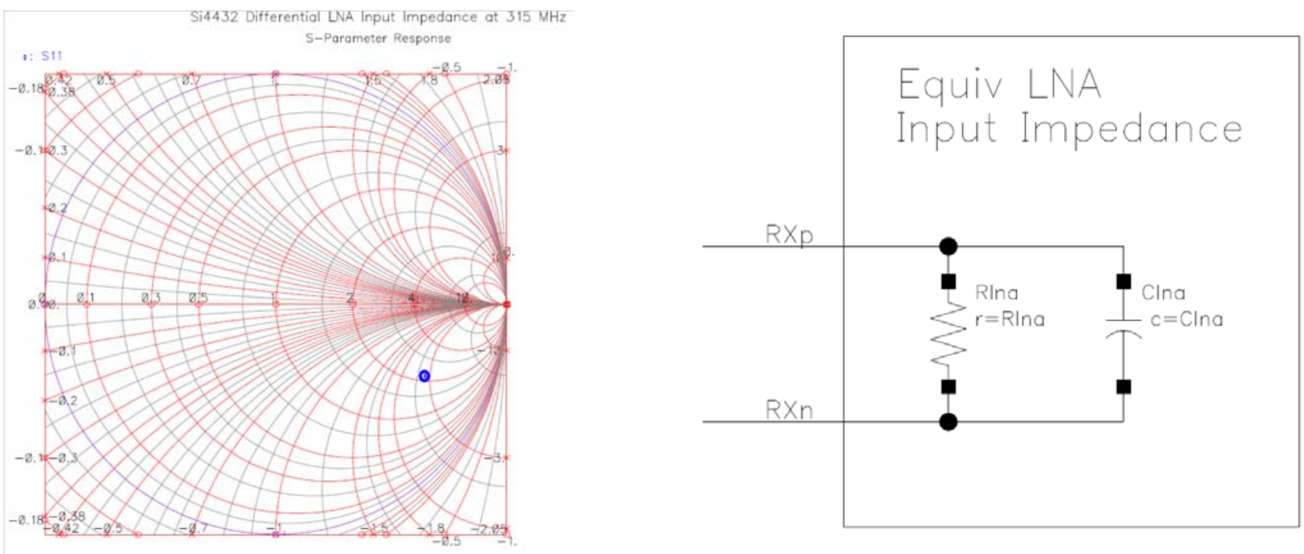


**Figure 6. Step #1: Plot LNA Input Impedance**

### 4.1.2. Step #2: Add Parallel Inductance $L_{LNA}$ to Resonate with LNA Capacitance

Although Step #2 may technically be combined with the subsequent step, the design equations are somewhat easier to manipulate if the equivalent LNA input capacitance "$C_{LNA}$" is first effectively cancelled (at the frequency of interest) by resonating it with a parallel inductance "$L_{LNA}$".

Equation 1: $$L_{LNA} = \frac{1}{\omega_{RF}{}^{2} C_{LNA}}$$

In our design example, we find this value of inductance to be equal to $L_{LNA}$ = 110.99 nH. After this amount of parallel inductance is added across the LNA inputs, the input impedance can be considered to be purely real and of a value equivalent to"$R_{LNA}$". This is shown in Figure 7.

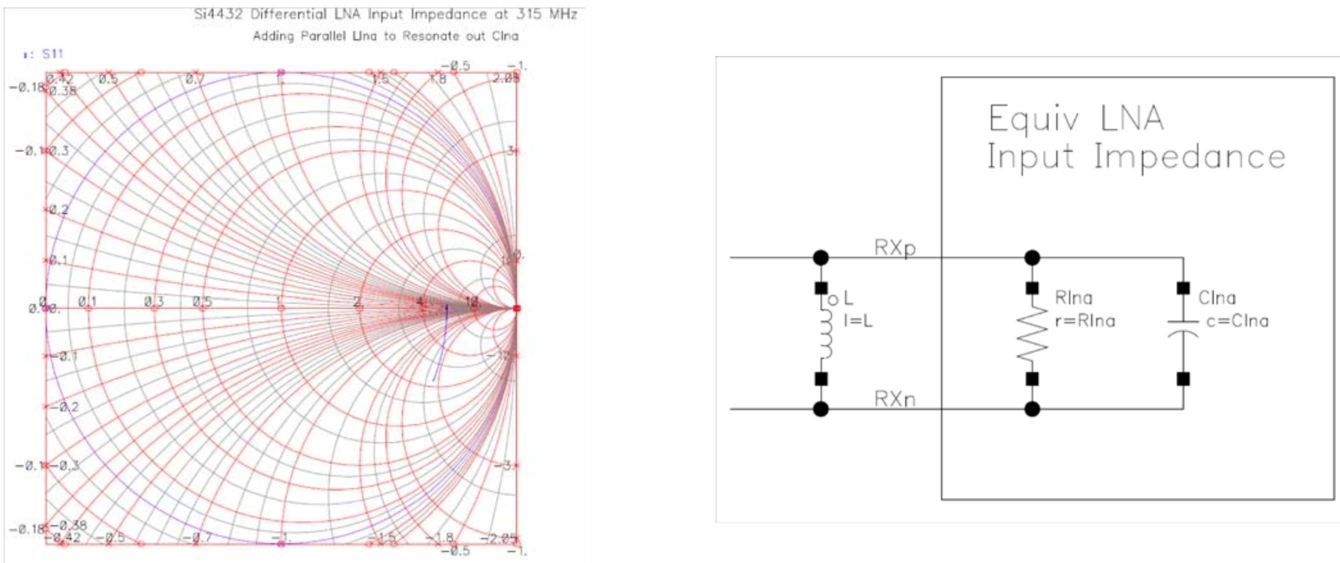


**Figure 7. Step #2: Add Parallel Inductance to Resonate $C_{LNA}$**

### 4.1.3. Step #3: Add Matching Inductance in Parallel with LNA Input

Next place an additional matching inductor "$L_M$" in parallel with the LNA input network. The value of the inductance should be chosen to further rotate the susceptance on the Smith Chart along a line of constant conductance (in the -j$B_P$ direction) until the 50 Ω circle is reached. The required value of matching inductance "$L_M$" is given by the following:

Equation 2: $$L_M = \frac{1}{\omega_{RF}\sqrt{\left(\frac{1}{50\Omega * R_{LNA}}\right) - \left(\frac{1}{R_{LNA}}\right)^2}}$$

(We will not provide the derivation for this equation in this Application Note; the full derivation may be found in a Mathcad worksheet, also available from Silicon Labs.)

Using this equation, or by employing graphical methods on the Smith Chart, we find that an additional parallel inductance of $L_M$ = 66.24 nH is required to reach the 50 Ω circle.

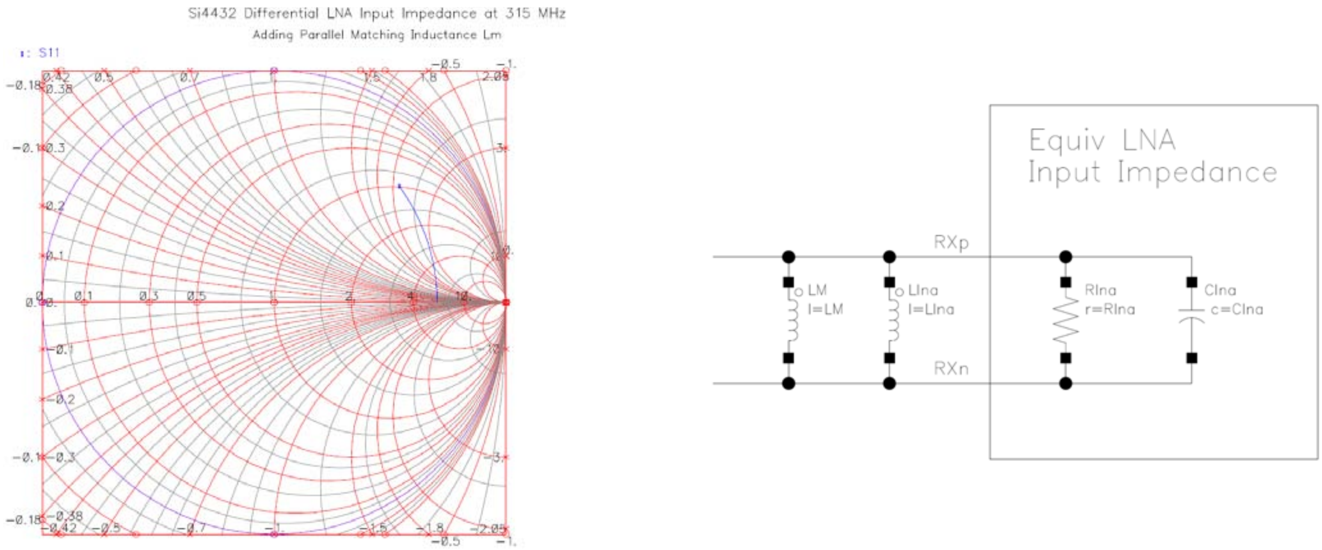


**Figure 8. Step #3: Add Parallel Matching Inductance $L_M$**

Note that as $L_{LNA}$ and $L_M$ are in parallel with each other, they may be combined into one equivalent inductance "L".

Equation 3: $L = \dfrac{L_{LNA} L_M}{L_{LNA} + L_M}$

Using this equation, we quickly find that a single inductor of value L = 41.4 nH may be used in place of $L_{LNA}$ and $L_M$.

### 4.1.4. Step #4: Determine Total Amount of Series Capacitive Reactance

We next determine the total amount of series capacitive reactance ($-jX_{CTOTAL}$), required to match this point to 50 Ω. That is to say, we desire to rotate the reactance along a line of constant resistance until we arrive at the center of the Smith Chart. The required value of total capacitance is given by the following:

Equation 4: $C_{TOTAL} = \dfrac{1}{\omega_{RF}\, 50\Omega \sqrt{\left(\dfrac{R_{LNA}}{50\Omega}\right) - 1}}$

(This equation is again offered without proof here; refer to the Mathcad sheet for the full derivation.)

Using this equation, or by employing graphical methods on the Smith Chart, we find that a total series capacitance of $C_{TOTAL}$ = 4.68 pF is required to reach the 50 Ω origin of the Smith Chart.

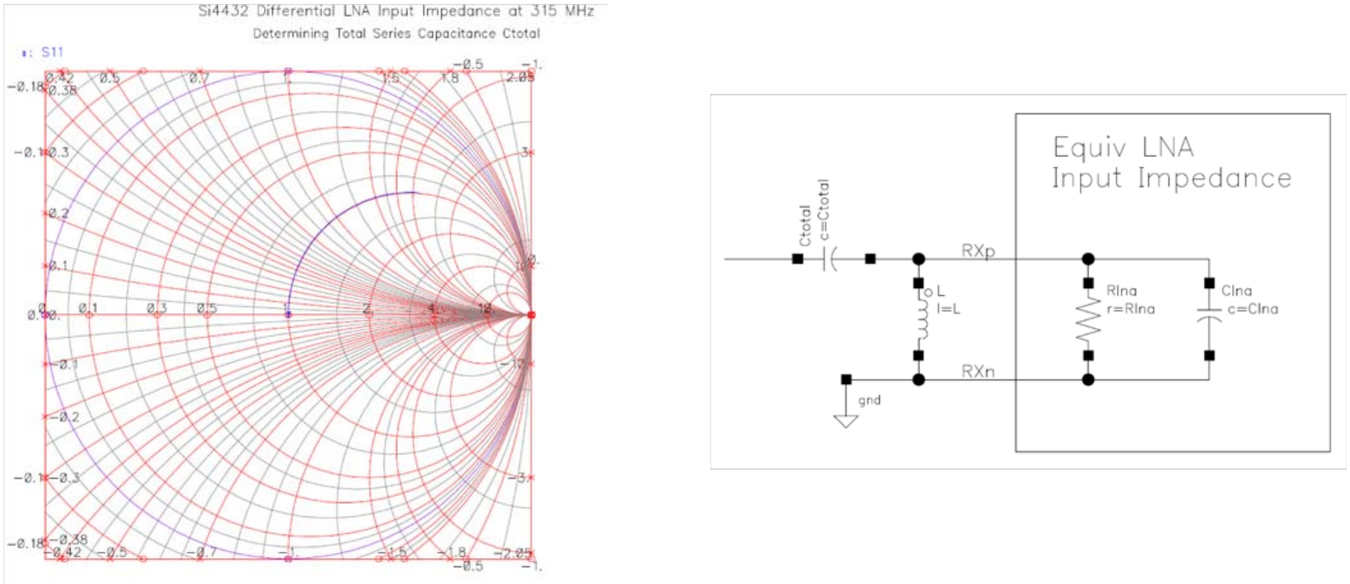


**Figure 9. Step #4: Determine TOTAL Series Capacitive Reactance**

#### 4.1.5. Step #5: Allocate Total Series Capacitance Between $C_1$ and $C_2$

The final step is to properly allocate this total required series capacitive reactance between $C_1$ and $C_2$.

There are an infinite number of possible matching networks which achieve a perfect match to 50 Ω. However, only one of these solutions also achieves the best possible equal-amplitude-with-180°-phase relationship between the waveforms at the RXp / RXn inputs.

For example, we could set the value of $C_2$ so large that it provides essentially 0 Ω of capacitive reactance and essentially ac-shorts the RXn pin to GND. Under this condition, it would be possible to set the value of $C_1$ to provide all of the required series capacitive reactance (determined in Step #4 above) and still achieve a perfect match to 50 Ω; however, the waveforms at the RXp and RXn nodes would not be balanced. The voltage at the RXn pin in this scenario would be zero (ac-shorted to GND by $C_2$). From an ac standpoint, this is equivalent to the schematic shown in Figure 9.

To properly allocate the total series capacitive reactance between $C_1$ and $C_2$, we must first recognize the required relationship between L and $C_2$. We want the voltages at the RXp and RXn pins to be equal in amplitude but opposite in phase, and the voltage developed "across" the parallel network of L-$R_{LNA}$-$C_{LNA}$ must be twice the amplitude (and of opposite polarity) as the voltage that exists at the RXn node.

We recall that a portion of the parallel inductance "L" is simply used to resonate out the capacitance $C_{LNA}$. As shown in Steps #2 and #3, it was useful to consider the inductance "L" as consisting of two inductors in parallel: $L_{LNA}$ and $L_M$, as re-drawn in Figure 10.

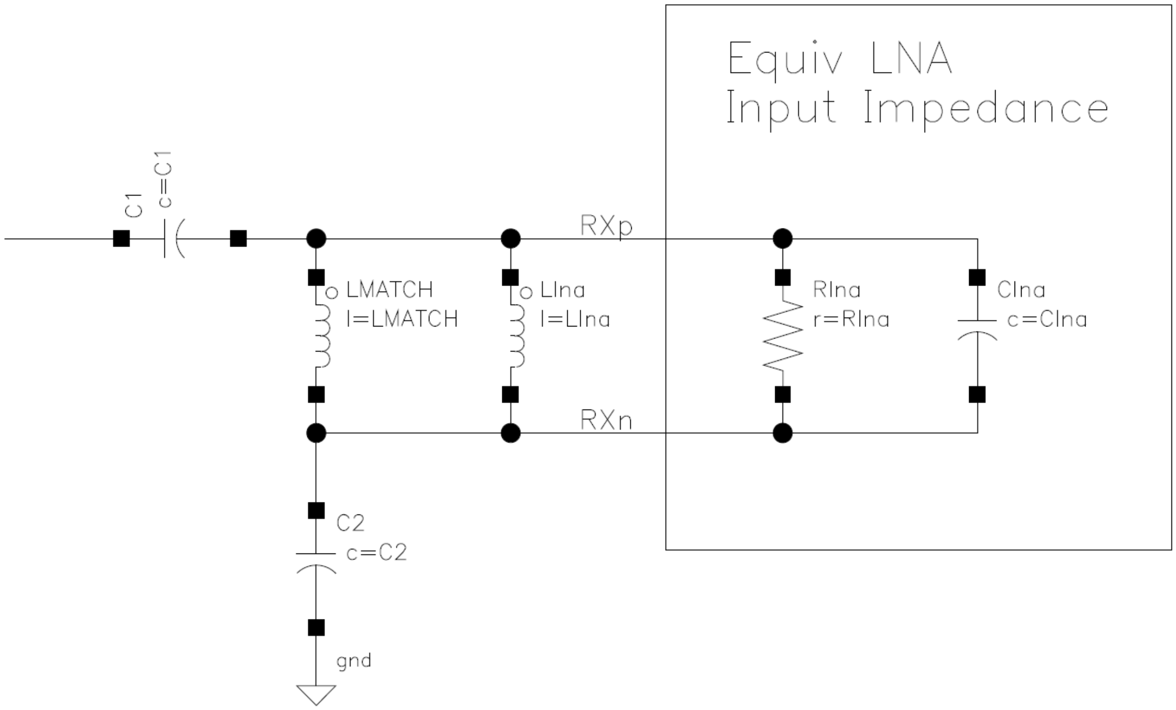


**Figure 10. Resolving "L" into Two Parts**

We have already determined the value of these two inductances as $L_{LNA}$ = 110.99 nH and $L_M$ = 66.24 nH. As the inductance "$L_{LNA}$" is simply used to resonate with "$C_{LNA}$", the match network may thus be re-drawn (at the operating frequency) as shown in Figure 11.

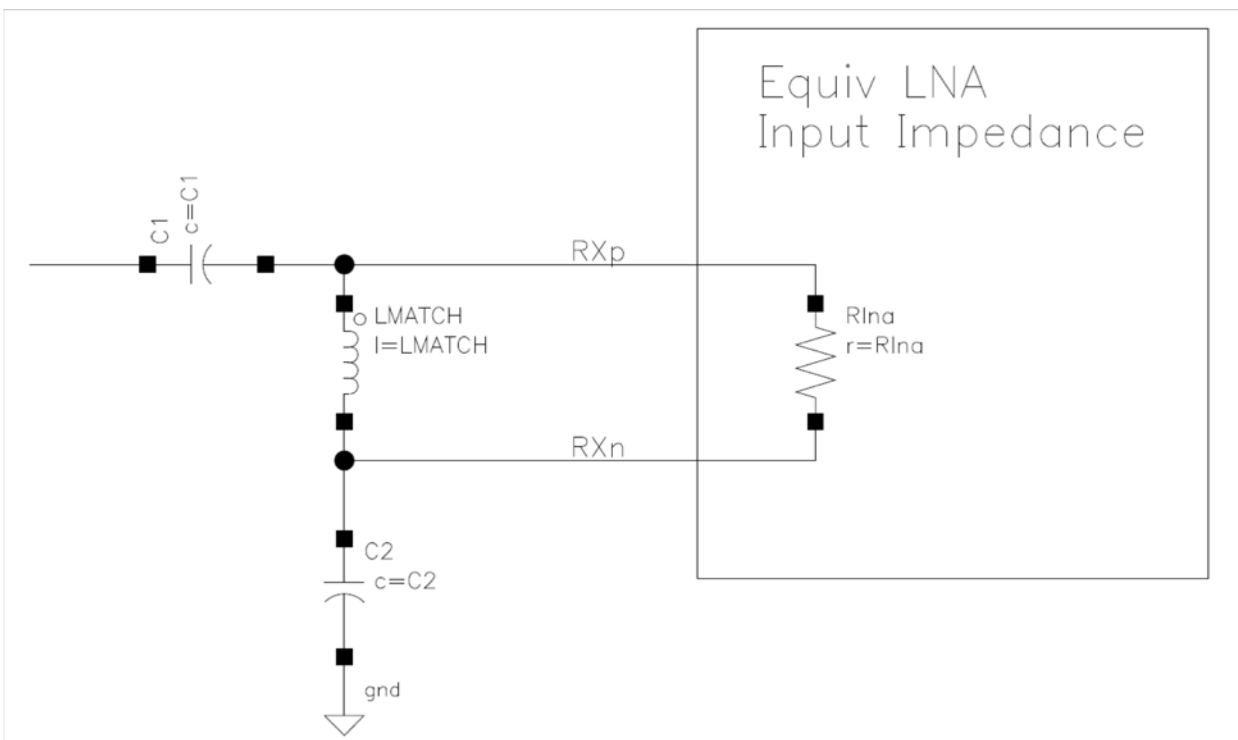


**Figure 11. Equivalent Match Network at Operating Frequency**

We next recall that we desire the voltage across $L_M$ to be twice the amplitude (and opposite in phase) to the voltage across $C_2$. Temporarily ignoring the effects of $R_{LNA}$, we arrive at the following requirement:

Equation 5: $X_{LM} = 2 * X_{C2}$

As we have already determined the required value of inductance $L_M$, the required value for $C_2$ follow immediately from the previously-derived equation for $L_M$.

Equation 6: $$C_2 = \frac{2\sqrt{\left(\frac{R_{LNA}}{50\Omega}\right) - 1}}{\omega_{RF} R_{LNA}}$$

Using this equation, we arrive at the value for $C_2$ = 7.71 pF. It is then a simple matter to allocate the remaining portion of total required series capacitive reactance to $C_1$.

Equation 7: $$C_1 = \frac{1}{\left(\frac{1}{C_{TOTAL}}\right) - \left(\frac{1}{C_2}\right)}$$

From this equation, we find that $C_1$ = 11.92 pF. Thus we have now determined all of the components in our 3-element match network:

- $C_1$ = 11.92 pF
- L = 41.48 nH
- $C_2$ = 7.71 pF

### 4.1.6. Phase Imbalance of RXp/RXn Signals

If the input impedance of the LNA were infinite ($R_{LNA} = \infty$), this procedure would result in equal-amplitude perfectly-balanced (180° out-of-phase) waveforms at the RXp and RXn nodes. However, a finite value for $R_{LNA}$ has the effect of shifting the phase of the signal developed across the parallel combination of L-$R_{LNA}$-$C_{LNA}$; thus the voltage developed at the RXp node can never be exactly 180° out-of-phase with respect to the voltage at the RXn node. This effect may be clearly seen in the simulated results of Figure 12; the differential voltages are equal in amplitude but not quite opposite in phase.

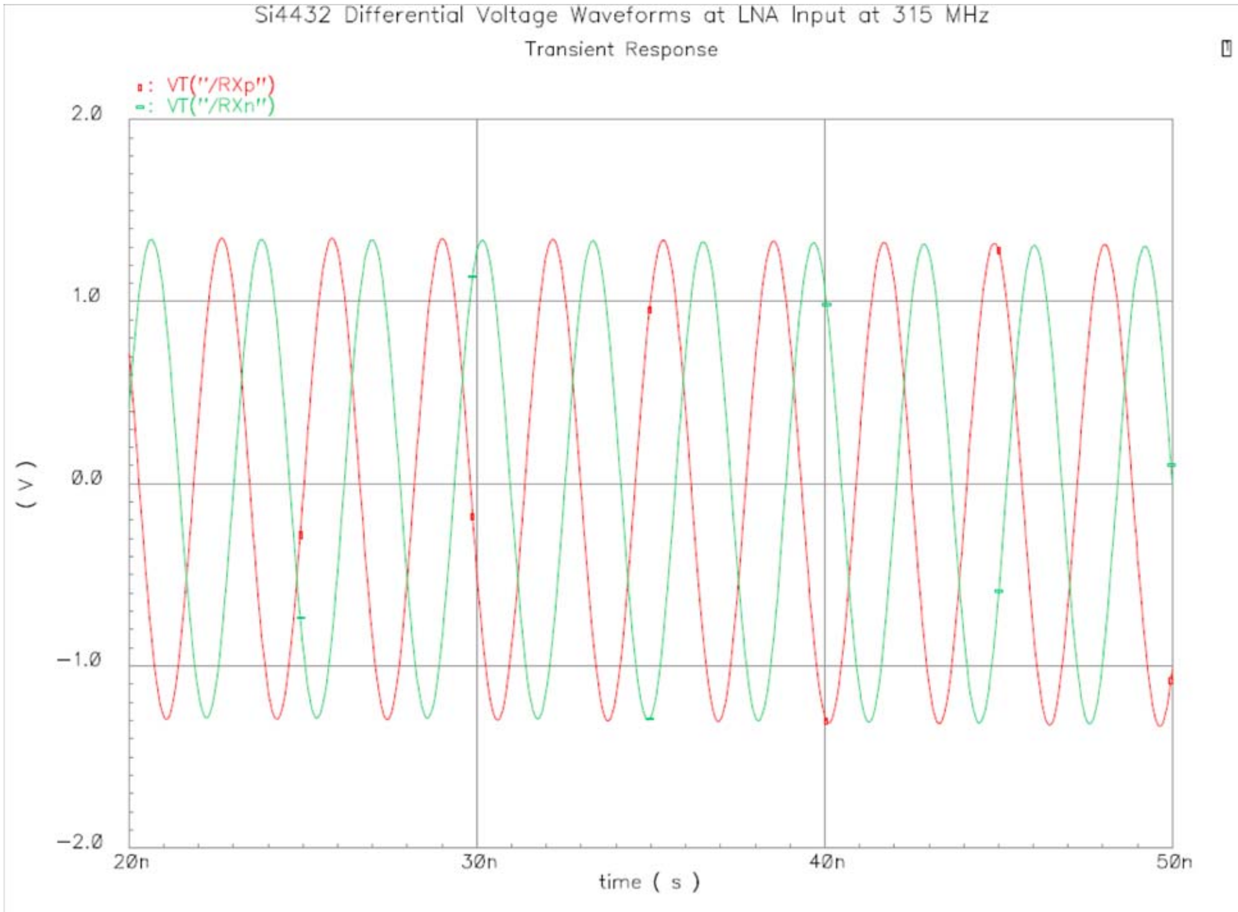


**Figure 12. Differential Voltage Waveforms at LNA Input (3-Element HPF Match)**

This is why we earlier stated that the 3-element match network provides slightly less-than-optimal performance when compared to a perfect balun.

### 4.1.7. Highpass Filter AC Frequency Response of Match Network

We previously stated that this type of matching architecture resulted in a HPF response. This is demonstrated by performing a small-signal ac simulation of the matching network's frequency response, shown in Figure 13. It can clearly be seen that this matching network provides slightly greater attenuation to out-of-band signals located below the desired operating frequency, as compared to out-of-band signals located above the desired operating frequency.

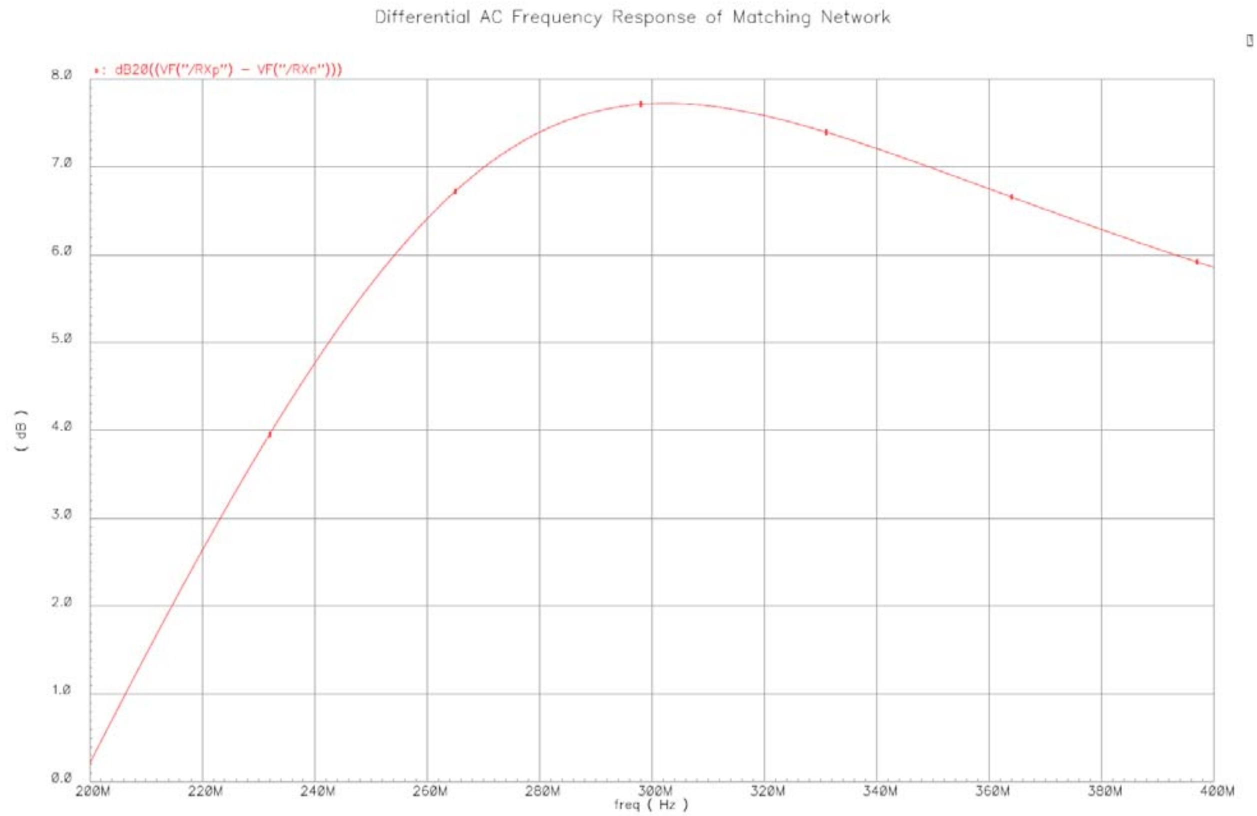


**Figure 13. Highpass Filter Response of 3-Element Match Network**

#### 4.1.8. Summary Table of 3-Element Match Network Component Values vs. Frequency

Some users may not be greatly interested in the theoretical development of the matching network, but are concerned only with quickly obtaining a set of component values for a given desired frequency of operation. For those users, we summarize the resulting component values for the 3-element HPF match network for multiple frequencies across the operating range of the Si4432 RFIC.

**Table 2. 3-Element HPF Match Network Component Values (Calculated)**

| Freq | $R_{LNA}$ | $C_{LNA}$ | $C_1$ | L | $C_2$ |
|---|---|---|---|---|---|
| 240 MHz | 343 Ω | 2.63 pF | 13.21 pF | 60.16 nH | 9.36 pF |
| 300 MHz | 292 Ω | 2.34 pF | 12.16 pF | 44.41 nH | 7.99 pF |
| 315 MHz | 283 Ω | 2.30 pF | 11.92 pF | 41.48 nH | 7.71 pF |
| 350 MHz | 263 Ω | 2.19 pF | 11.52 pF | 35.23 nH | 7.14 pF |
| 390 MHz | 244 Ω | 2.09 pF | 11.16 pF | 30.37 nH | 6.59 pF |
| 400 MHz | 240 Ω | 2.07 pF | 11.08 pF | 29.75 nH | 6.46 pF |
| 433 MHz | 227 Ω | 2.01 pF | 10.89 pF | 26.41 nH | 6.09 pF |
| 450 MHz | 220 Ω | 1.98 pF | 10.87 pF | 25.40 nH | 5.93 pF |
| 470 MHz | 213 Ω | 1.96 pF | 10.82 pF | 23.50 nH | 5.74 pF |
| 500 MHz | 204 Ω | 1.92 pF | 10.74 pF | 21.75 nH | 5.48 pF |
| 550 MHz | 189 Ω | 1.87 pF | 10.84 pF | 18.72 nH | 5.11 pF |
| 600 MHz | 176 Ω | 1.83 pF | 11.08 pF | 16.66 nH | 4.79 pF |
| 650 MHz | 165 Ω | 1.80 pF | 11.43 pF | 14.80 nH | 4.50 pF |
| 700 MHz | 155 Ω | 1.77 pF | 11.98 pF | 13.27 nH | 4.25 pF |
| 750 MHz | 146 Ω | 1.74 pF | 12.78 pF | 12.00 nH | 4.03 pF |
| 800 MHz | 138 Ω | 1.72 pF | 13.89 pF | 10.90 nH | 3.83 pF |
| 850 MHz | 131 Ω | 1.70 pF | 15.38 pF | 9.96 nH | 3.64 pF |
| 868 MHz | 128 Ω | 1.68 pF | 16.36 pF | 9.69 nH | 3.58 pF |
| 900 MHz | 125 Ω | 1.69 pF | 17.33 pF | 9.14 nH | 3.47 pF |
| 915 MHz | 122 Ω | 1.68 pF | 18.98 pF | 8.92 nH | 3.42 pF |
| 950 MHz | 118 Ω | 1.66 pF | 21.71 pF | 8.46 nH | 3.31 pF |
| 960 MHz | 117 Ω | 1.66 pF | 22.58 pF | 8.33 nH | 3.28 pF |

#### 4.1.9. Final Tweaking of Component Values

All of the above analysis assumes use of ideal discrete components in the matching network. We recognize the fact that surface-mount 0603- or 0402-size components themselves contain parasitic elements that modify their effective values at the frequency of interest. Additionally, the analysis presented above does not take make allowance for any PCB parasitics, such as trace inductance, component pad capacitance, etc. Furthermore, it is convenient to use the nearest-available 5% or 10% component value; the component values shown above represent results of exact mathematical calculations.

This means that it will almost certainly be necessary to “tweak” the final matching values for a specific application and board layout. The above component values should be used as starting points, and the values modified slightly to zero-in on the best match to 50 Ω, and the best RX sensitivity.

Silicon Labs has empirically determined the optimum matching network values at a variety of frequencies, using RF Test Boards designed by (and available from) Silicon Labs. By comparing the empirical values of Table 3 with the calculated values of Table 2, the reader may observe that the component values are quite close in agreement at frequencies below 500 MHz. However, somewhat larger deviations in value occur at higher frequencies, primarily due to the unmodelled parasitic effects of the PCB traces and discrete components. As was mentioned previously, the calculated matching component values of Table 2 should be used as a starting point and adjusted for best performance.

**Table 3. 3-Element HPF Match Network Component Values (Empirical)**

| Freq | $R_{LNA}$ | $C_{LNA}$ | $C_1$ | L | $C_2$ |
|---|---|---|---|---|---|
| 240 MHz | 343 Ω | 2.63 pF | 12.0 pF | 56 nH | 9.1 pF |
| 315 MHz | 283 Ω | 2.30 pF | 15.0 pF | 47 nH | 5.6 pF |
| 433 MHz | 227 Ω | 2.01 pF | 10.0 pF | 33 nH | 4.7 pF |
| 470 MHz | 213 Ω | 1.96 pF | 12.0 pF | 27 nH | 4.7 pF |
| 868 MHz | 128 Ω | 1.68 pF | 6.8 pF | 11 nH | 3.9 pF |
| 915 MHz | 122 Ω | 1.68 pF | 6.8 pF | 11 nH | 3.3 pF |
| 950 MHz | 118 Ω | 1.66 pF | 8.2 pF | 10 nH | 3.9 pF |

## 4.2. Three-Element Matching Procedure (LPF Architecture)

We have demonstrated that it is possible to construct a 3-element match network that has a slight HPF frequency response. It was previously stated (without proof) that it was also possible to construct a 3-element match network that has a slight LPF response. We now turn to investigating the conditions under which such a LPF match network may be constructed and how to do so.

The HPF 3-element match network could be described as a C-L-C network; that is to say, the matching network consisted of a series capacitor ($C_1$), a shunt inductor (L), and another series capacitor ($C_2$).

Conversely, the LPF 3-element match network demonstrated here may be described as an L-C-L network; that is to say, the matching network consists of a series inductor ($L_1$), a shunt capacitor (C), and another series inductor ($L_2$). This matching topology is shown in Figure 14.

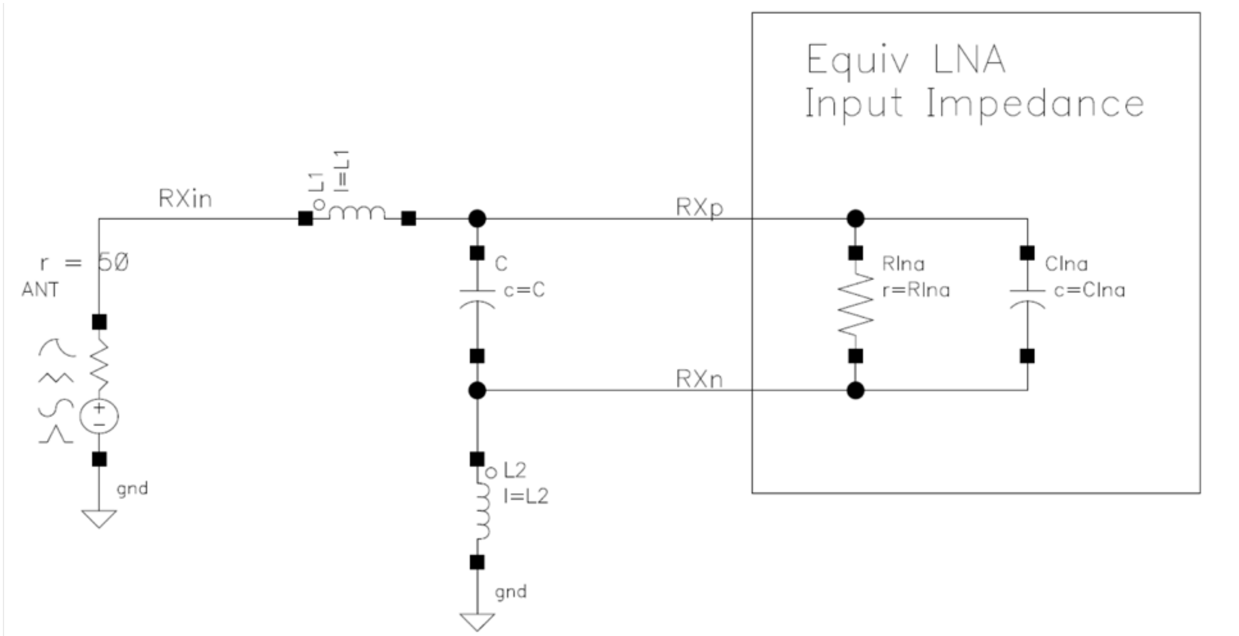


**Figure 14. LPF Three-Element Match Network**

The matching philosophy for this topology is similar in concept but proceeds around the Smith Chart in the opposite direction. In the previous example of the HPF match, we added inductance "L" in parallel with the equivalent LNA input impedance in order to rotate the susceptance on the Smith Chart along a line of constant conductance (in the $-jB_P$ direction) until the 50 Ω circle was reached. In the case of the LPF match, we instead add capacitance "C" in parallel with the LNA input impedance to rotate the susceptance along a line of constant conductance *in the opposite direction* (in the $+jB_P$ direction) until the 50 Ω circle is reached.

The total amount of series inductive reactance ($+jX_L$) required to match this point to 50 Ω is then determined. That is to say, we simply rotate this reactance along a line of constant resistance until we arrive at the center of the Smith Chart.

The final step in the LPF matching process is to properly allocate this total amount of inductance between $L_1$ and $L_2$. As before, the proper allocation is determined by the values that yield the best equal-amplitude, out-of-phase signals at the RXp and RXn inputs of the LNA. As the calculation steps for the LPF match are quite similar to those for the HPF match, we do not repeat them here.

An LPF match is not realizable at all frequencies. Referring back to Figure 4, we find that for frequencies near the upper end (e.g., above ~915 MHz) of the Si443x frequency range the real part of the LNA input impedance is already less than 50 Ω. Adding capacitance in parallel with this LNA input impedance (i.e., the first step in the LPF matching process) simply rotates the impedance on the Smith Chart to an even lower real impedance. That is to say, it is not possible to rotate the LNA input impedance in the $+jB_P$ direction to reach the 50 Ω circle, because the starting LNA input impedance is *already beyond* the 50 Ω circle. At these frequencies, the only matching option is to move in the other direction on the Smith Chart by adding parallel inductance (i.e., the HPF matching approach).

Fortunately, this restriction is usually not that burdensome. The operating frequencies at which a LPF match is realizable fall towards the lower end of the Si443x operating range (e.g., 315 MHz), but it is at these operating frequencies that a LPF match may also be found to be the most desirable. This is because the strongest signals that the user is likely to encounter are generally signals from cell phone base station towers at ~900 MHz, and the additional attenuation of these strong signals due to use of a LPF match is a welcome feature. Conversely, at frequencies near the high end of the Si443x operating range (where a LPF match is not realizable), the strongest out-of-band interfering signals are generally found to occur *below* the desired operating frequency, and thus a HPF match is preferred in any case.

## 4.3. Four-Element Matching Procedure

As discussed previously, it is possible to achieve a theoretically-perfect match with the 4-element match network shown in Figure 3. The complete mathematical derivation of the equations for the required component values is beyond the scope of this Application note; a Mathcad worksheet containing the complete derivation is available from Silicon Labs.

The matching procedure for the 3-element network was readily understood and explained by plotting each step on a Smith Chart. This graphical approach is somewhat less intuitive for the 4-element matching procedure. Instead, we will present a textual description of the main steps in the mathematical derivation, along with the important equations resulting from following these steps.

### 4.3.1. Step #1: Voltage at the RXn Node ($V_{RXn}$)

As the first step, we recognize that *if* we are successful in creating a match to a purely-real input impedance of $Z_{IN}$ = 50 Ω, then the input current $I_{IN}$ will also be purely real (arbitrarily assuming an input voltage from the source generator $V_{IN}$ of unity magnitude and zero phase). This input current passes through capacitor $C_2$ to develop the voltage at the RXn node ($V_{RXn}$). It is apparent that this voltage $V_{RXn}$ exhibits a –90° phase shift with respect to the input current $I_{IN}$, due to the capacitive reactance of $C_2$.

### 4.3.2. Step #2: Voltage at the RXp Node ($V_{RXp}$)

We next observe that the voltage at the RXp node ($V_{RXp}$) must be equal in amplitude to $V_{RXn}$ but opposite in phase. For this condition to be satisfied, the voltage *across* the LNA input pins must be *twice* the amplitude of $V_{RXN}$, as well as exactly opposite in phase. That is to say, if the phase of $V_{RXn}$ is –90°, the phase of $V_{RXp}$ must be +90°.

### 4.3.3. Step #3: Splitting the Input Current

Although the phase of the voltage across the LNA input pins must be +90°, we recognize that the input impedance of the LNA network is not purely inductive (unless $R_{LNA} = \infty$). Thus in order for the voltage across the LNA network to be purely reactive, the phase of the current through the LNA network must compensate for the phase shift introduced by $R_{LNA}$. As a result, it is necessary that the current through the LNA network be different from the current through $C_2$.

Thus the purpose of inductor $L_2$ is to split the input current $I_{IN}$ into two different components, with the current passing through the LNA network being of the appropriate phase to produce a voltage of opposite phase to $V_{RXn}$.

### 4.3.4. Equations for Component Values

Following these derivational steps, it is possible to obtain the following set of design equations for the necessary component values.

Equation 8: $$L_2 = \frac{\sqrt{50\Omega * R_{LNA}}}{\omega_{RF}}$$

Equation 9: $$C_1 = \frac{1}{{\omega_{RF}}^2 L_2}$$

Equation 10: $$C_2 = 2 * C_1$$

Equation 11: $$L_{LNA} = \frac{1}{{\omega_{RF}}^2 C_{LNA}}$$

Equation 12: $$L_M = 2 * L_2$$

Equation 13: $$L_1 = \frac{L_{LNA} L_M}{L_{LNA} + L_M}$$

Using these equations, we calculate all of the component values in our 4-element match network for our example design at $F_{RF}$ = 315 MHz:

- $C_1$ = 4.25 pF
- $L_1$ = 57.70 nH
- $C_2$ = 8.49 pF
- $L_2$ = 60.10 nH

### 4.3.5. Phase Balance of RXp/RXn Signals

We stated earlier that an advantage of the 4-element match network was the ability to achieve perfect phase balance (180 degrees) between the RXp and RXn input nodes. This effect may be clearly seen in Figure 15; the differential voltages are now both equal in amplitude *and* perfectly opposite in phase.

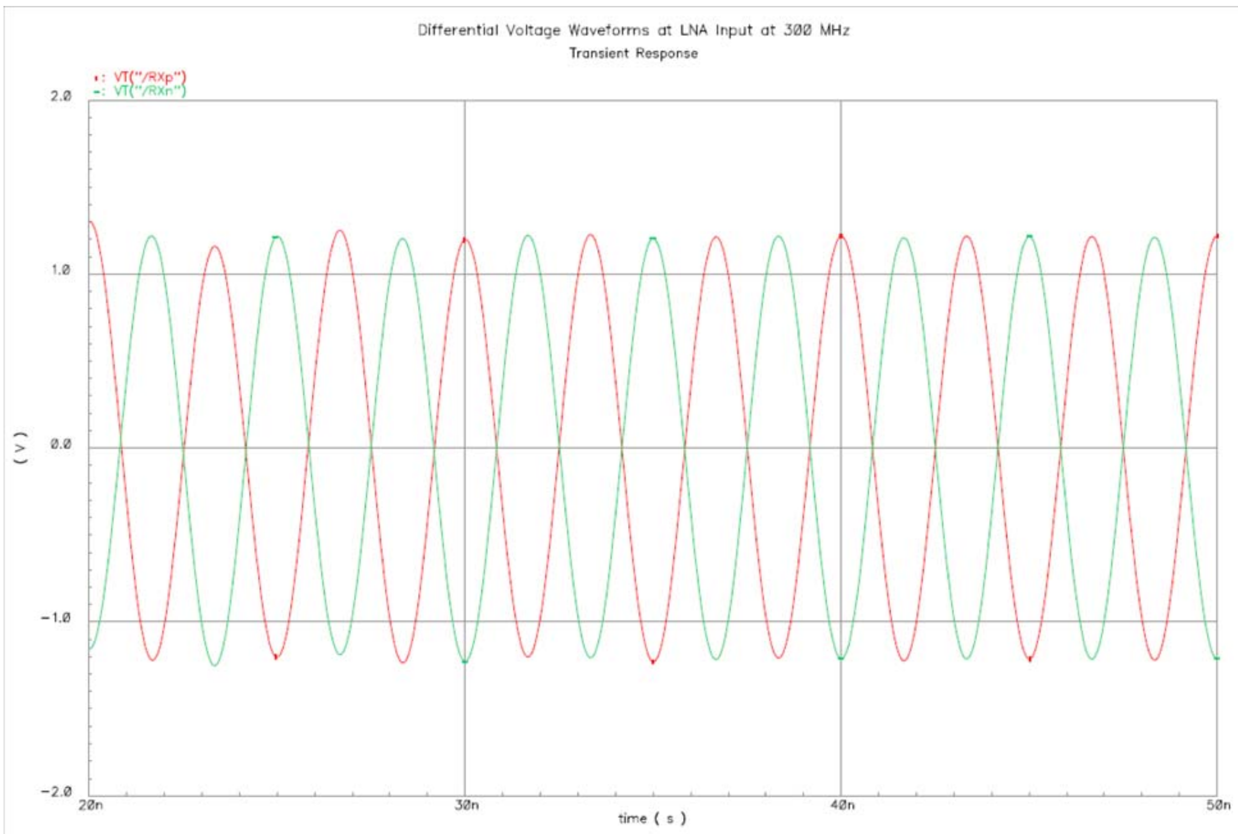


**Figure 15. Differential Voltage Waveforms at LNA Input (4-Element Match)**

### 4.3.6. AC Frequency Response of 4-Element Match Network

The 3-element match network resulted in either a HPF frequency response or a LPF frequency response, depending upon whether a C-L-C (HPF) or L-C-L (LPF) topology was chosen. The ac frequency response of the 4-element match network, however, is more symmetrical with respect to frequency. This is demonstrated by the small-signal ac simulation shown in Figure 16. It can be seen that this matching network provides a more symmetrical frequency response.

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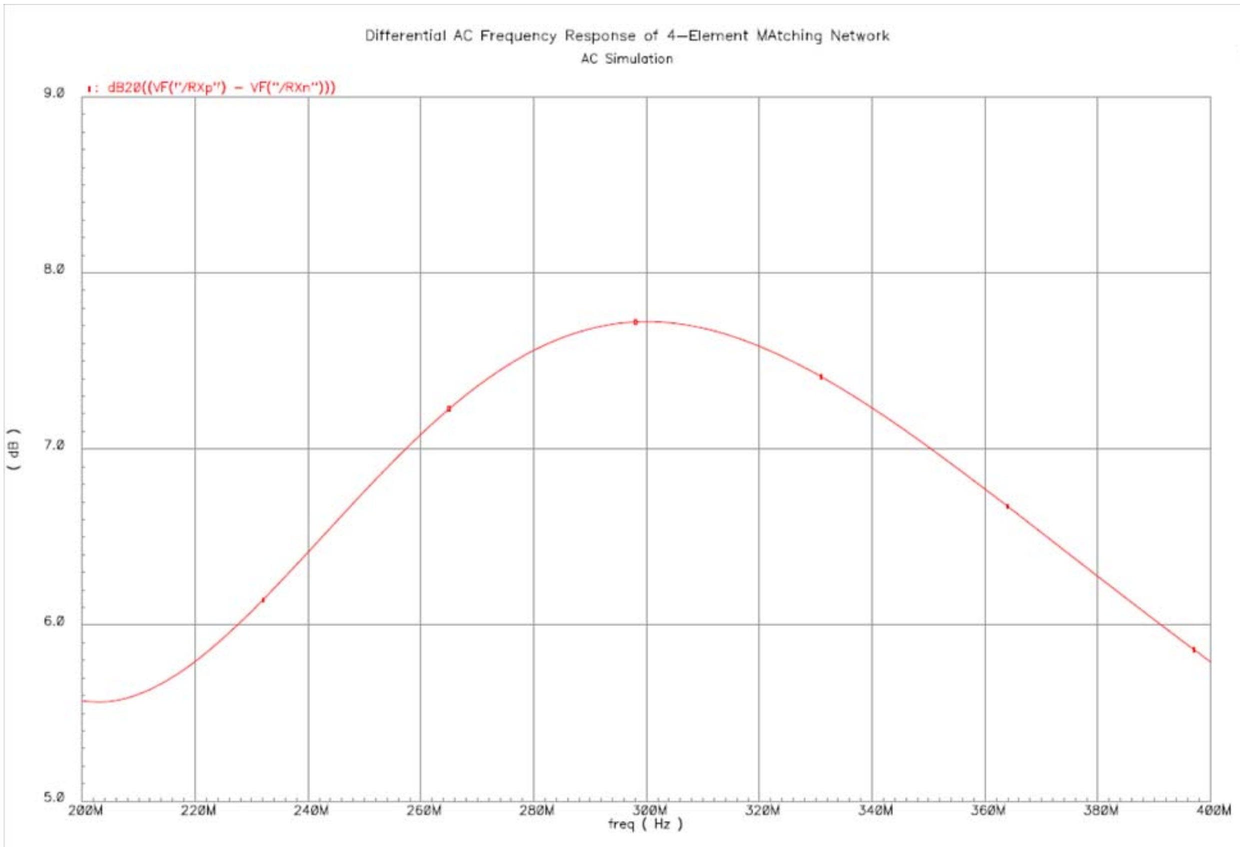


**Figure 16. AC Frequency Response of 4-Element Match Network**

#### 4.3.7. Summary Table of 4-Element Match Network Component Values vs. Frequency

Some users may not be greatly interested in the theoretical development of the matching network, but are concerned only with quickly obtaining a set of component values for a given desired frequency of operation. For those users, we summarize the calculated component values for the 4-element match network for multiple frequencies across the operating range of the Si4432 RFIC.

**Table 4. 4-Element Match Network Component Values (Calculated)**

| Freq | $R_{LNA}$ | $C_{LNA}$ | $C_1$ | $L_1$ | $C_2$ | $L_2$ |
|---|---|---|---|---|---|---|
| 240 MHz | 343 Ω | 2.63 pF | 5.06 pF | 85.19 nH | 10.13 pF | 86.84 nH |
| 300 MHz | 292 Ω | 2.34 pF | 4.39 pF | 62.06 nH | 8.78 pF | 64.10 nH |
| 315 MHz | 283 Ω | 2.30 pF | 4.25 pF | 57.71 nH | 8.49 pF | 60.10 nH |
| 350 MHz | 263 Ω | 2.19 pF | 3.97 pF | 49.55 nH | 7.93 pF | 52.15 nH |
| 390 MHz | 244 Ω | 2.09 pF | 3.95 pF | 47.15 nH | 7.90 pF | 52.34 nH |
| 400 MHz | 240 Ω | 2.07 pF | 3.63 pF | 40.74 nH | 7.26 pF | 43.59 nH |
| 433 MHz | 227 Ω | 2.01 pF | 3.45 pF | 36.17 nH | 6.90 pF | 39.16 nH |
| 450 MHz | 220 Ω | 1.98 pF | 3.37 pF | 34.12 nH | 6.74 pF | 37.09 nH |
| 470 MHz | 213 Ω | 1.96 pF | 3.28 pF | 31.85 nH | 6.56 pF | 34.95 nH |
| 500 MHz | 204 Ω | 1.92 pF | 3.15 pF | 28.98 nH | 6.30 pF | 32.15 nH |
| 550 MHz | 189 Ω | 1.87 pF | 2.98 pF | 24.93 nH | 5.95 pF | 28.13 nH |
| 600 MHz | 176 Ω | 1.83 pF | 2.83 pF | 21.69 nH | 5.66 pF | 24.88 nH |
| 650 MHz | 165 Ω | 1.80 pF | 2.70 pF | 19.05 nH | 5.39 pF | 22.24 nH |
| 700 MHz | 155 Ω | 1.77 pF | 2.58 pF | 16.89 nH | 5.17 pF | 20.02 nH |
| 750 MHz | 146 Ω | 1.74 pF | 2.48 pF | 15.10 nH | 4.97 pF | 18.13 nH |
| 800 MHz | 138 Ω | 1.72 pF | 2.40 pF | 13.57 nH | 4.79 pF | 16.53 nH |
| 850 MHz | 131 Ω | 1.70 pF | 2.32 pF | 12.27 nH | 4.63 pF | 15.15 nH |
| 868 MHz | 128 Ω | 1.68 pF | 2.29 pF | 11.90 nH | 4.58 pF | 14.67 nH |
| 900 MHz | 125 Ω | 1.69 pF | 2.24 pF | 11.14 nH | 4.47 pF | 13.98 nH |
| 915 MHz | 122 Ω | 1.68 pF | 2.23 pF | 10.83 nH | 4.45 pF | 13.59 nH |
| 950 MHz | 118 Ω | 1.66 pF | 2.18 pF | 10.20 nH | 4.36 pF | 12.87 nH |
| 960 MHz | 117 Ω | 1.66 pF | 2.17 pF | 10.02 nH | 4.34 pF | 12.68 nH |

Similar to the 3-element match network, it will almost certainly be necessary to “tweak” the final matching values for a specific application and board layout, due to parasitic effects of PCB traces and non-ideal discrete components. The above component values should be used as starting points, and the values modified slightly to zero-in on the best match to 50 Ω, and the best RX sensitivity.

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