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INPUT AND OUTPUT MATCHING NETWORKS DESIGN FOR RF CIRCUITS

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ABSTRACT
A design and study of input and out impedance matching for RF (radio frequency) circuits is presented in this paper. Passive elements such as inductors (L) and capacitors (C) are crucial for impedance matching, and are specifically designed such that they would satisfy the gain requirements at a specific frequency of operation. Impedance matching is necessary in RF circuit design to provide maximum possible power transfer between the source and the load. Complex tradeoffs among technology specifications and design parameters exist and should be carefully observed when designing the impedance matching networks, to optimize the performance of RF circuits.

INTRODUCTION
Impedance matching plays vital role in optimizing the performance of the RFIC (radio frequency integrated circuit) design. Matching provides maximum power transfer between the input or source and the output or the load, thus allowing the RF circuit to achieve the desired performance esp. the gain requirements. Inductors and capacitors are key passive elements that are crucial for impedance matching, and are specifically designed such that they would satisfy the gain requirements at a specific frequency or range of operation [1], [2]. Design tradeoffs between matching network parameters are inevitable, so it is crucial that inductors and capacitors be designed carefully for the specific requirements of the RF application.

METHODOLOGY
For this paper, actual scattering parameters (S-parameters) of a 300μm/0.25μm (W/L, width over length) transistor were provided, for RF circuit application of common-source amplifier topology. Required values of S-parameters for a specific frequency of operation could then be determined using linear interpolation. Table 1 shows the S-parameters of the transistor at frequency initially set to 2.6GHz, and with a 10-dB gain requirement.

Table 1. S-parameters of transistor at frequency of 2.6GHz

S-Parameters	Real	Imaginary
S_{11}	0.599858625	-0.53991373
S_{21}	-0.219423779	1.14183461
S_{12}	0.067523223	0.03730980
S_{22}	0.116580879	-0.40044436

Stability conditions of the two-port network in terms of S-parameters play an essential role in amplifier designs. Although stability is frequency dependent, we want to ensure that the amplifiers design exhibits unconditional stability esp. at higher frequencies.

There are several ways to check for the stability of the two-port network. Expressions of stability constants in Eq. (1) to (6) could be used to check for the stability of the design. Computed values are shown in Table 2. These can also be used to compute for the source and load reflection coefficients which will be shown later.

$$\Delta = det(S) = S_{11}S_{22} - S_{12}S_{21} \quad (1)$$

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} \quad (2)$$

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 \quad (3)$$

$$B_2 = 1 + |S_{22}|^2 - |S_{11}|^2 - |\Delta|^2 \quad (4)$$

Table 2. Stability constants

Stability Constants	Values
Δ	0.3825303 ∠-103.4315597°
K	1.7895714
B_1	1.3310610
B_2	0.3762802
C_1	0.6520807 ∠-44.9832792°
C_2	0.1329476 ∠-103.4852039°

To have unconditional stability, the Rollett stability factor K must be greater than unity, that is, K > 1, as well as one other condition [1], [2]. Hence, any of the following criteria is sufficient and necessary for unconditional stability:

$$K > 1 \text{ and } |\Delta| < 1 \quad (7)$$

$$K > 1 \text{ and } B_1 > 0 \quad (8)$$

$$K > 1 \text{ and } B_2 > 0 \quad (9)$$

$$K > 1 \text{ and } |S_{12}S_{21}| < 1 - |S_{11}|^2 \quad (10)$$

$$K > 1 \text{ and } |S_{12}S_{21}| < 1 - |S_{22}|^2 \quad (11)$$

Table 3 shows the condition values of all the unconditional stability criteria.

Table 3. Unconditional stability criteria

Criteria	Values	Check Condition
$K > 1$	1.7895714 > 1	✓
$\|\Delta\| < 1$	0.3825303 < 1	✓
$B_1 > 0$	1.3310610 > 0	✓
$B_2 > 0$	0.3762802 > 0	✓
$\|S_{12}S_{21}\| < 1 - \|S_{11}\|^2$	0.0896990 < 0.3486628	✓
$\|S_{12}S_{21}\| < 1 - \|S_{22}\|^2$	0.0896990 < 0.8260532	✓

It can be observed that all of the conditions are met. Therefore, the two-port network in terms of S-parameters is unconditionally stable. Maximum power transfer is achieved when both the generator and load are conjugately matched to the two-port network, that is,

$$\Gamma_{in} = \Gamma_G^* \text{ and } \Gamma_{out} = \Gamma_L^* \quad (12)$$

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$$Z_{in} = Z_G^* \text{ and } Z_{out} = Z_L^* \quad (13)$$

Where
Γ_{in} = input reflection coefficient of the two-port network
Γ_{out} = output reflection coefficient of the two-port network
Γ_G = source or generator reflection coefficient
Γ_L = load reflection coefficient
Z_{in} = input impedance of the two-port network
Z_{out} = output impedance of the two-port network
Z_G = source or generator impedance
Z_L = load impedance

Figure 1 shows the block/schematic diagram of a two port network impedance matching networks.

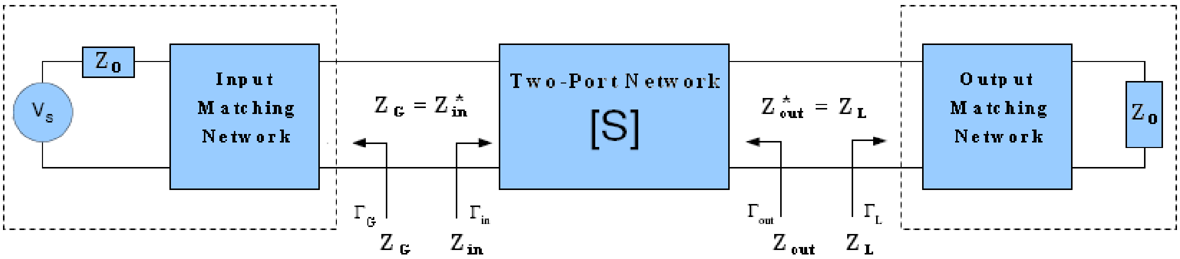

Figure 1. Two-port network with input and output matching networks

Through simultaneous conjugate matching, the following reflection coefficients can be obtained:

$$\Gamma_{in} = \Gamma_G^* = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1 - S_{22}\Gamma_L} = \frac{S_{11} - \Delta\Gamma_L}{1 - S_{22}\Gamma_L} \quad (14)$$

$$\Gamma_{out} = \Gamma_L^* = S_{22} + \frac{S_{12}S_{21}\Gamma_G}{1 - S_{11}\Gamma_G} = \frac{S_{22} - \Delta\Gamma_G}{1 - S_{11}\Gamma_G} \quad (15)$$

Alternatively, the source and load reflection coefficients in Eq. (16) and (17) could be derived using the expressions in Eq. (3) to (6).

$$\Gamma_G = \frac{B_1 - \sqrt{B_1^2 - 4|C_1|^2}}{2C_1} \quad (16)$$

$$\Gamma_L = \frac{B_2 - \sqrt{B_2^2 - 4|C_2|^2}}{2C_2} \quad (17)$$

Using the expressions in Eq. (16) and (17), source/generator and load impedances could now be obtained.

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$$Z_G = \left(\frac{1+\Gamma_G}{1-\Gamma_G}\right) Z_0 \tag{18}$$

$$Z_L = \left(\frac{1+\Gamma_L}{1-\Gamma_L}\right) Z_0 \tag{19}$$

Table 4 shows the values of all the reflection coefficients as well as the impedances, assuming normalization impedance of $Z_0 = 50\ \Omega$.

Table 4. Reflection coefficients and impedances

Γ and Z	Values
Γ_{in}	0.5775038 - j0.5771668
Γ_{out}	-0.0965024 - j0.4024191
Γ_G	0.5775038 + j0.5771668
Γ_L	-0.0965024 + j0.4024191
Z_{in}	32.5793388 - j112.81061 Ω
Z_{out}	30.3735008 - j29.49727 Ω
Z_G	32.5793388 + j112.81061 Ω
Z_L	30.3735008 + j29.49727 Ω

For the input and output matching networks, L-network is used because it is the simplest and most widely used matching network for lumped elements, as shown in Figures 2 to 3.

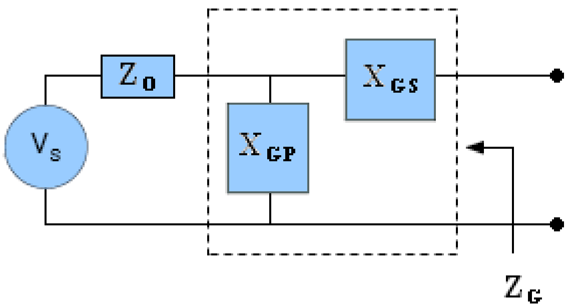

Figure 2. L-network of the input matching network

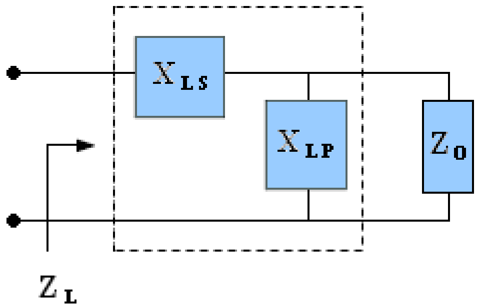

Figure 3. L-network of the output matching network

Where

X_{GS} = series reactance of the L-network of the input matching network

X_{GP} = parallel reactance of the L-network of the input matching network

X_{LS} = series reactance of the L-network of the output matching network

X_{LP} = parallel reactance of the L-network of the output matching network

The elements of the L-network for both the input and output matching network as shown in Figures 2 to 3 are arranged in such orientation given that the real components of Z_G and Z_L (or R_G and R_L) are smaller than the real component of the normalization impedance which is $Z_0 = 50\ \Omega$ (or $R_0 = 50\ \Omega$) [1], [2]. To check,

$$R_G = 32.57933879\ \Omega \quad < \quad R_0 = 50\ \Omega \tag{20}$$

$$R_L = 30.37350084\ \Omega \quad < \quad R_0 = 50\ \Omega \tag{21}$$

For the L-network of the input matching network, the elements can be solved using the following equations given that $Z_0 = 50\ \Omega$ ($R_0 = 50\ \Omega$, $X_0 = 0$):

$$Q_G = \sqrt{\frac{R_0}{R_G} - 1} \tag{22}$$

$$X_{GP} = \pm\frac{R_0}{Q_G} \tag{23}$$

or $$X_{GP1} = +\frac{R_0}{Q_G} \tag{24}$$

$$X_{GP2} = -\frac{R_0}{Q_G} \tag{25}$$

$$X_{GS} = -\left(-X_G \pm R_G Q_G\right) \tag{26}$$

or $$X_{GS1} = -\left(-X_G + R_G Q_G\right) \tag{27}$$

$$X_{GS2} = -\left(-X_G - R_G Q_G\right) \tag{28}$$

Likewise, for the L-network of the output matching network, the elements can be solved using the following equations given that $Z_0 = 50\ \Omega$ ($R_0 = 50\ \Omega$, $X_0 = 0$):

$$Q_L = \sqrt{\frac{R_0}{R_L} - 1} \tag{29}$$

$$X_{LP} = \pm\frac{R_0}{Q_L} \tag{30}$$

or $$X_{LP1} = +\frac{R_0}{Q_L} \tag{31}$$

$$X_{LP2} = -\frac{R_0}{Q_L} \tag{32}$$

$$X_{LS} = -(-X_L \pm R_L Q_L) \quad (33)$$

or $$X_{LS1} = -(-X_L + R_L Q_L) \quad (34)$$

$$X_{GS2} = -(-X_L - R_L Q_L) \quad (35)$$

Table 5 summarizes the values obtained from the expressions Eq. (29) to (35).

Table 5. L-network elements

Q and Z	Values
Q_G	0.73124
X_{GP1}	68.37681 Ω
X_{GP2}	-68.37681 Ω
X_{GS1}	88.98723 Ω
X_{GS2}	136.63400 Ω
Q_L	0.80385
X_{LP1}	62.20081 Ω
X_{LP2}	-62.20081 Ω
X_{LS1}	5.08159 Ω
X_{LS2}	53.91296 Ω

Actual capacitor and inductor values at frequency of 2.6GHz can be computed from the L-network reactances. Positive reactance denotes an inductive component while a negative reactance implies a capacitive component.

$$jX_{GP1} = j\omega L_{GP1} \quad (36)$$

$$\boxed{L_{GP1} = \frac{X_{GP1}}{\omega} = \frac{68.37680755}{2\pi(2.6GHz)} = 4.185579582nH} \quad (37)$$

$$-jX_{GP2} = \frac{1}{j\omega C_{GP2}} \quad (38)$$

$$\boxed{C_{GP2} = \frac{1}{\omega X_{GP2}} = \frac{1}{2\pi(2.6GHz)(68.37680755)} = 0.895236877\,pF} \quad (39)$$

$$jX_{GS1} = j\omega L_{GS1} \quad (40)$$

$$\boxed{L_{GS1} = \frac{X_{GS1}}{\omega} = \frac{88.98722805}{2\pi(2.6GHz)} = 5.447214314nH} \quad (41)$$

$$jX_{GS2} = j\omega L_{GS2} \quad (42)$$

$$\boxed{L_{GS2} = \frac{X_{GS2}}{\omega} = \frac{136.633996}{2\pi(2.6GHz)} = 8.363836868nH} \quad (43)$$

$$jX_{LP1} = j\omega L_{LP1} \quad (44)$$

$$\boxed{L_{LP1} = \frac{X_{LP1}}{\omega} = \frac{62.20080573}{2\pi(2.6GHz)} = 3.807525268nH} \quad (45)$$

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$$-jX_{LP2} = \frac{1}{j\omega C_{LP2}} \tag{46}$$

$$C_{LP2} = \frac{1}{\omega X_{LP2}} = \frac{1}{2\pi(2.6GHz)(62.20080573)} = 0.9841261529\,pF \tag{47}$$

$$jX_{LS1} = j\omega L_{LS1} \tag{48}$$

$$L_{LS1} = \frac{X_{LS1}}{\omega} = \frac{5.081592616}{2\pi(2.6GHz)} = 0.3110617629nH \tag{49}$$

$$jX_{LS2} = j\omega L_{LS2} \tag{50}$$

$$L_{LS2} = \frac{X_{LS2}}{\omega} = \frac{53.91295498}{2\pi(2.6GHz)} = 3.300197416nH \tag{51}$$

Two sets of values will be used in the simulation to check if the whole circuit is really matched at the frequency of operation which is 2.6GHz. Design1 is comprised of L_{GS1} and L_{GP1} for the input matching network and L_{LS1} and L_{LP1} for the output matching network. On the other hand, Design2 is composed of L_{GS2} and C_{GP2} for the input matching network and L_{LS2} and C_{LP2} for the output matching network. The actual values of inductors and capacitors are listed in Table 6. Note that the gain requirement for the amplifier design is set at 10dB, and hopefully the computed L and C for impedance matching networks could help achieve the target.

Table 6. Actual L-network elements

L and *C*	Values
L_{GP1}	4.186 nH
C_{GP2}	0.895 pF
L_{GS1}	5.447 nH
L_{GS2}	8.364 nH
L_{LP1}	3.808 nH
C_{LP2}	0.984 pF
L_{LS1}	0.311 nH
L_{LS2}	3.300 nH

SIMULATION RESULTS AND ANALYSIS

Two designs were simulated using the two sets of values of the input and output matching networks, with values previously summarized in Table 6. The actual values are shown in Table 6. Figures 4 and 5 shows the complete schematic circuit designs of Design1 and Design2.

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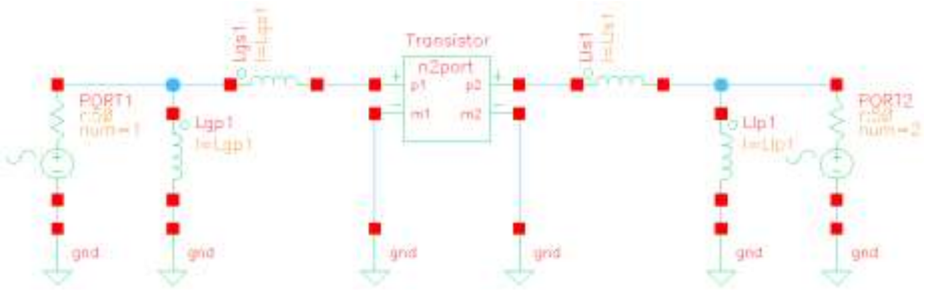

Figure 4: Design1 schematic diagram

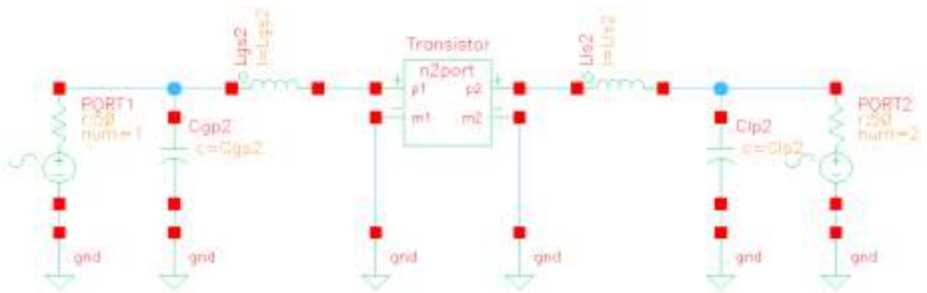

Figure 5: Design2 schematic diagram

The n2port from the analogLib library is used for the two-port network. Although spectre-format file is preferred for the S-parameter file input of the n2port component, touchstone-format can still be used. In this study, the touchstone-format S-parameter file is used since the actual S-parameters are given in touchstone format. Still, touchstone-formatted file can be converted to spectre-format using the command sptr. Figures 6 to 9 shows the comparison of the results of the S-parameter plots of the two designs.

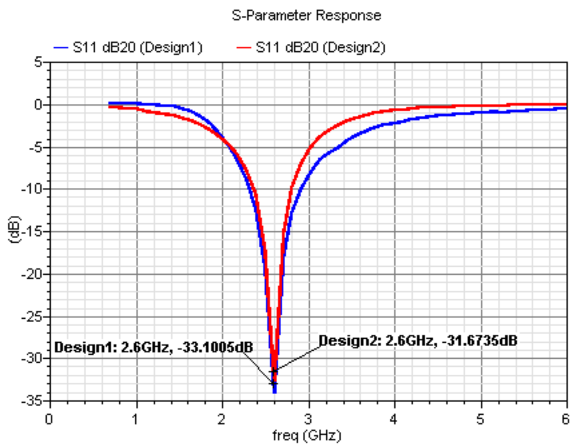

Figure 6. S_{11} plot (in dB) versus frequency

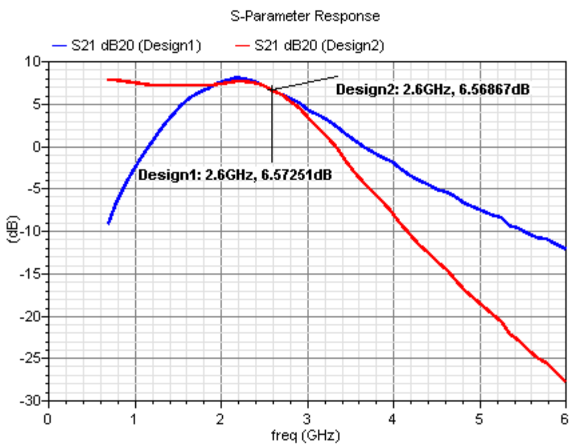

Figure 7. S_{21} plot (in dB) versus frequency

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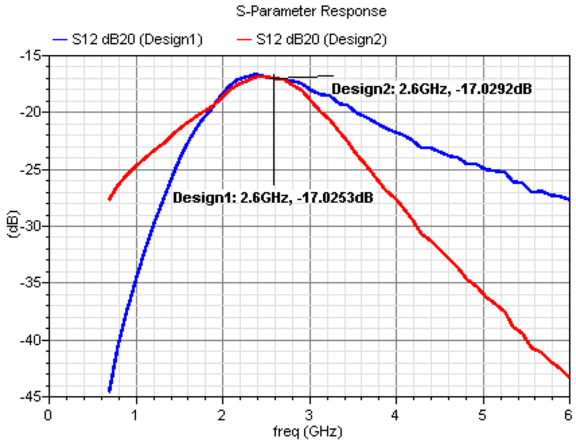

Figure 8. S_{12} plot (in dB) versus frequency

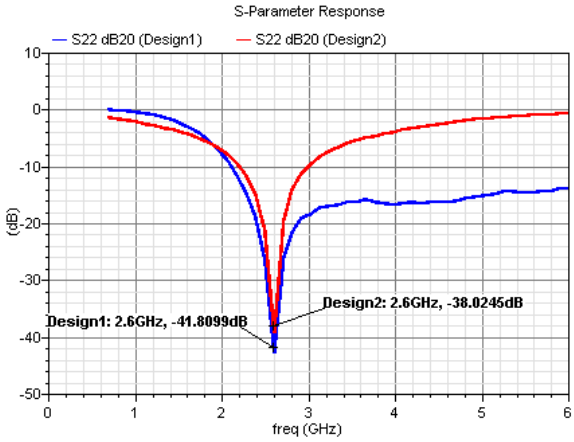

Figure 9. S_{22} plot (in dB) versus frequency

S-parameter plots were obtained using the sp (s-parameter) analysis. It can be shown in Figure 6 that the two designs are somehow matched at frequency of 2.6GHz. The values of the S-parameters for the two designs at 2.6GHz are comparable and relatively close to each other. But it can be observed that the S-parameter plots of Design2 are smoother than the plots of Design1 at frequencies greater than 2.6GHz. The difference is evident esp. in the S22 plot in Figure 9. This signifies that Design2, which is comprised of inductor-capacitor combination in the L-matching networks, exhibits a more stable behavior for higher frequencies than the Design1 which is an all-inductor design. Moreover, the S11 and S22 plots of Design2 are more symmetric in reference to the frequency of operation which is 2.6GHz compared to the Design1. A summary of S-parameters values are shown in Table 7.

Table 7. S-parameters response at 2.6GHz

S-Parameters	Design1	Design2
S_{11}	-33.101 dB	-31.674 dB
S_{21}	6.573 dB	6.569 dB
S_{12}	-17.025 dB	-17.029 dB
S_{22}	-41.810 dB	-38.025 dB

The gain of the transistor or the amplifier is shown in the S21 plot in Figure 7. At frequency of 2.6GHz, the gain is only 6.573dB for the Design1 and 6.569dB for the Design2. It is almost 3.5dB less than the 10dB gain target. This is because as the frequency increases in the higher frequencies esp. beyond the frequency of operation, the gain decreases. If the gain-bandwidth product is to be remained constant, then as the bandwidth or the frequency increases, the gain should compensate, thus decreasing the gain. Figures 10 to 15 shows the S-parameter plots in Smith charts.

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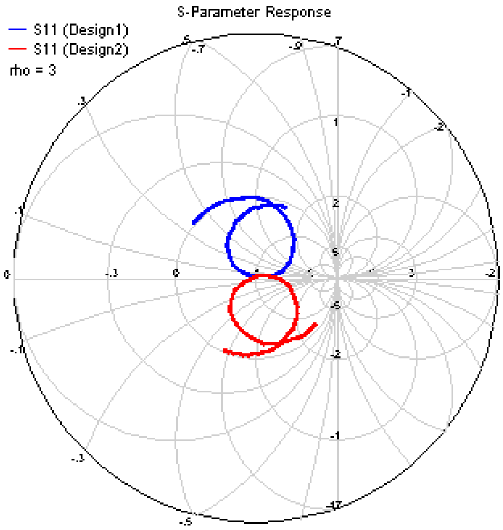

Figure 10. S11 impedance Smith chart plot

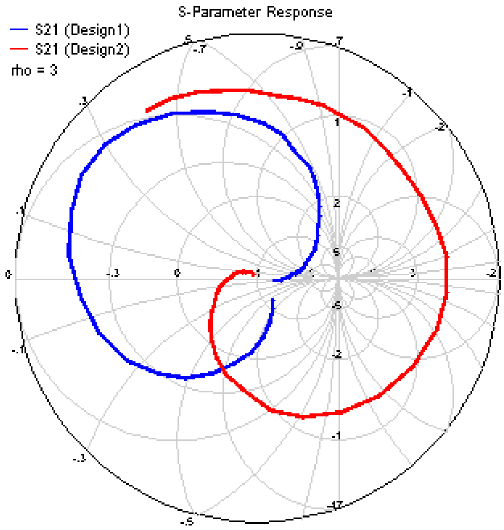

Figure 11. S21 impedance Smith chart plot

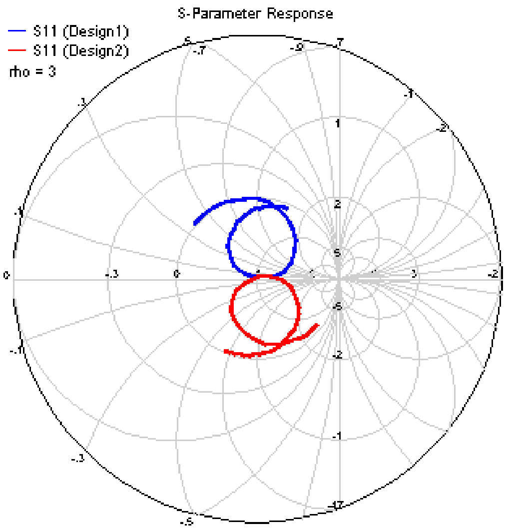

Figure 10. S11 impedance Smith chart plot

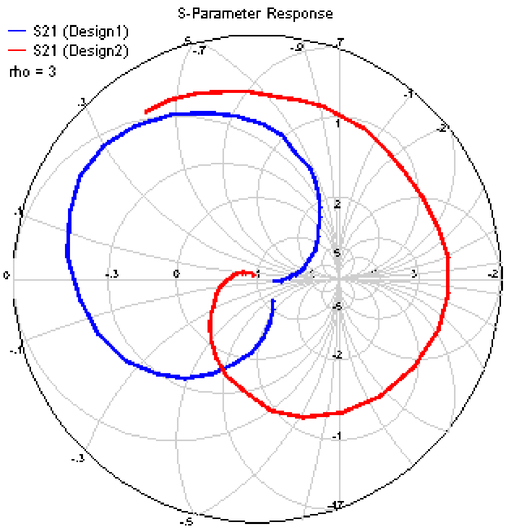

Figure 11. S21 impedance Smith chart plot

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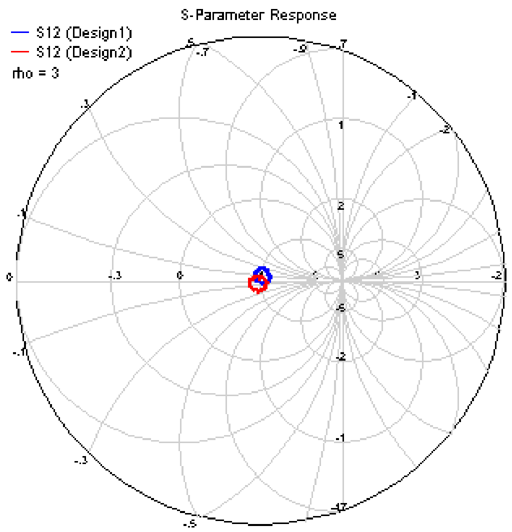

Figure 12. S12 impedance Smith chart plot

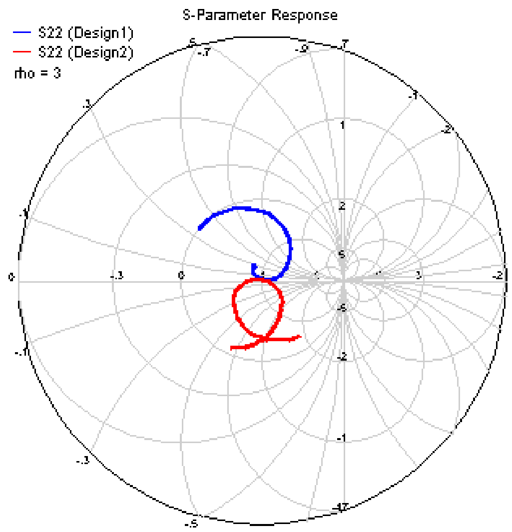

Figure 13. S22 impedance Smith chart plot

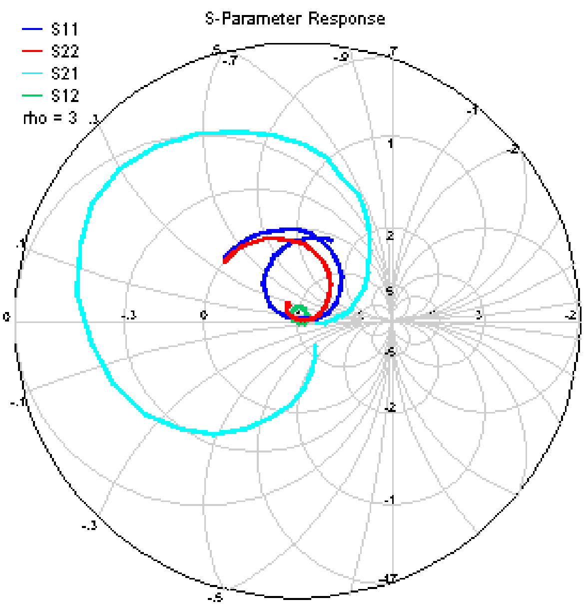

Figure 14. Impedance Smith chart plot of S-parameters of Design1

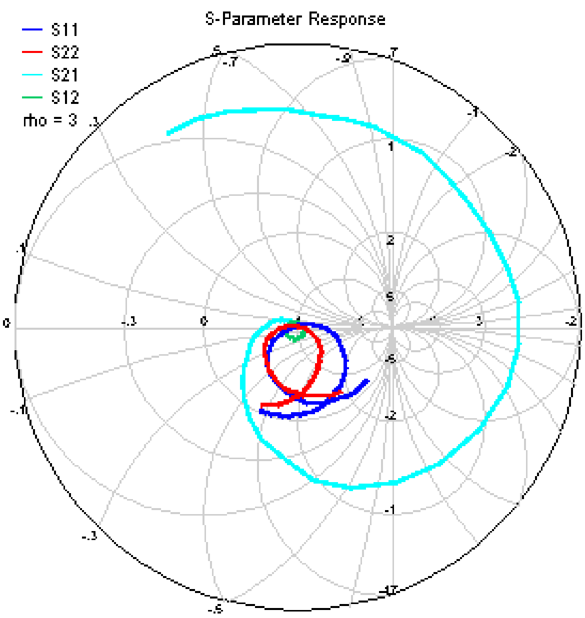

Figure 15. Impedance Smith chart plot of S-parameters of Design2

Since Design1 is an all-inductor design, the responses of S-parameters in the impedance Smith chart are more on the inductive half of the Smith chart, evident in Figures 10 to 14. On the other hand, Design2 has a capacitor on the matching networks, thus the impedance Smith chart responses of the S-parameters are more on the capacitive half of the Smith chart as evident in the charts shown in Figures 10-13, 15.

CONCLUSIONS AND RECOMMENDATIONS

Impedance matching is necessary in RF circuit design to provide maximum possible power transfer between the source or generator and the output load. In this study, two designs were modeled and investigated. The design (Design2) which comprised of an inductor-capacitor combination in the input and output matching networks resulted to a smoother response or a more stable behavior for higher frequencies than the design (Design1) with all inductors in the matching networks. All designs achieved gain of 3dB versus the target of 10dB. One factor is the limitation of the initially provided actual S-parameters of the transistor in touchstone format, which were used for the transistor model using n2port two-port network component. Regardless, complex tradeoffs among technology specifications and design parameters exist and should be carefully handled when designing the impedance matching networks, to optimize the performance of the RF circuit.

Design and study of particular passive components could be helpful in understanding and finally designing the matching networks. Software tools like ASITIC (analysis and simulation of spiral inductors and transformers for ICs) [3], [4] and SpiralCalc (integrated spiral inductor calculator) [5], [6], [7], which are available for non-commercial purposes, could be used for this particular study.

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