INTEGRATED CIRCUITS

APPLICATION NOTE

AN2000

Techniques for optimizing UHF front-end integrated circuits

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ABSTRACT

This application note discusses S-parameter theory and applications and a method for obtaining device noise parameter data for the SA621 and SA611 UHF front-end integrated circuits. These devices are high performance, low-power communications systems, optimized over the 800-1000MHz frequency range. The SA621 contains a low noise amplifier (LNA), mixer and voltage controlled oscillator (VCO), while the SA611 is a simplified version containing only the LNA and mixer. This application note will focus primarily on matching the LNA input and output to achieve stability and optimal gain using S-parameter data and related calculations. Discussed is a procedure developed for obtaining device noise parameters without the need for an automated data-acquisition setup. In addition, performance trade-offs for the LNA, mixer, and VCO circuits are addressed.

INTRODUCTION

The high frequency communication industry is growing so rapidly that the design phase of product development is becoming shorter and shorter. Time-to-market is one of the critical factors in the success or failure of a product. Given that, long periods of experimentation are no longer feasible, designers need quick and reliable ways to evaluate integrated circuits to decide which parts are best suited for their products.

One way to reduce trial-and-error experimentation is by judiciously using S-parameters. In order to optimize a design, accurate S-parameters may need to be taken when not available from published data sheets. Through use of these measurements, stability can be determined and optimal gain achieved. Another significant way which design uncertainties can be reduced is by an experimental noise figure procedure capable of producing accurate results when automated data-acquisition equipment is unavailable. This application note presents and discusses S-parameter techniques that can be readily applied using a hand-held scientific calculator. This is particularly useful for obtaining desired, first order design approximations. Later in the application note trade-offs for optimizing particular aspects of front-end performance are presented in regards to the LNA, mixer, and VCO circuits of the SA611 and SA621.

SA621/SA611 LNA AND S-PARAMETERS

To prepare the reader for the examples to follow, a short description of the SA621/SA611 is presented. The LNA performance of the SA621 and SA611 are virtually identical. They have an average gain of 15dB with an achievable noise figure of approximately 1.7dB, and an input IP3 of –7dBm over the Advanced Mobile Phones System (AMPS) receive frequency range (869–894MHz). The LNA also performs well in the Industrial, Scientific and Medical (ISM) band (902–928 MHz.).

S-parameters play a critical role in designing for optimum gain and noise figure performance. Understanding how they are obtained and what to do with them can greatly decrease the experimental work necessary to make the LNA function properly. Knowing exactly where data sheet S-parameters are measured is the first step in applying them. The SA621/SA611 data sheet records typical measurements taken at *the pins of the IC.* They were obtained using a calibration board designed specifically for this purpose (Figure 1). Thus, any matching networks will refer to an actual pin connection on the IC. Measurements were made using a Hewlett Packard network analyzer (HP8753D) to automatically log data over a swept range from 100 to 1200 MHz. This instrument measures reflection coefficients at the ends of its calibrated coaxial connection points. These points are connected to the IC pins through SMA connectors and short PCB microstrip transmission lines. Time delays associated with these connectors and lines at the input and output (Figure 2 a) were rotated out using the “Port Extensions” function located in the “Calibration” menu of the HP8753D. Open and short termination results were observed with no chip on the board. If the microstrip transmission line and its associated connector (input or output) have negligible effect on each measurement over this frequency range, then the pad is open circuited and the rotated display should ideally be a single point on the outermost circle of the Smith Chart at R = ∞ (Figure 2 b). When the pad is shorted a similar point would be observed at R = 0. Following this port extension calibration, a chip was then soldered onto the calibration board and actual S-parameter data automatically gathered. Note that these parameters cannot be obtained from the demonstration(demo) board available from Philips Semiconductors because the external matching networks and associated transmission lines are in the way.

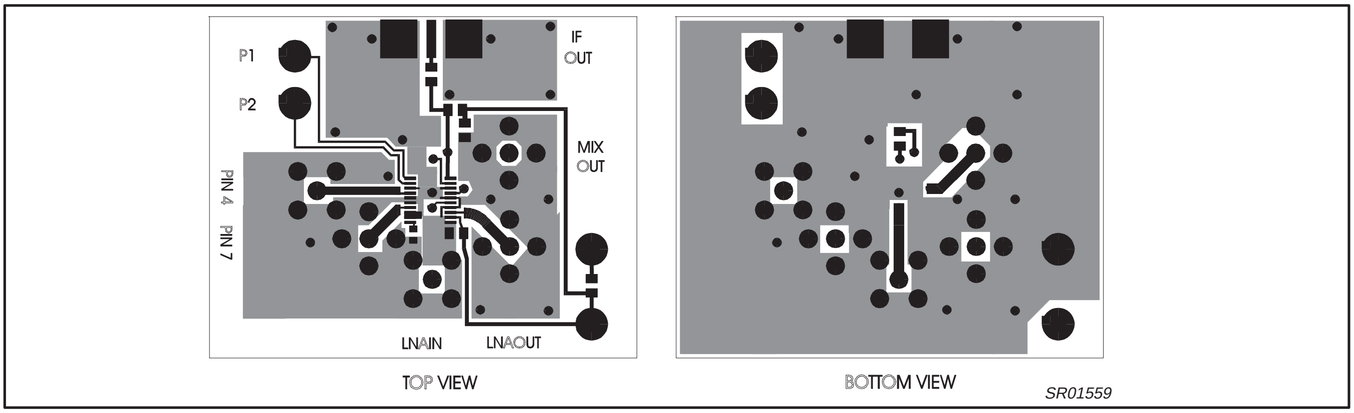


Figure 1 . Calibration Board Layout

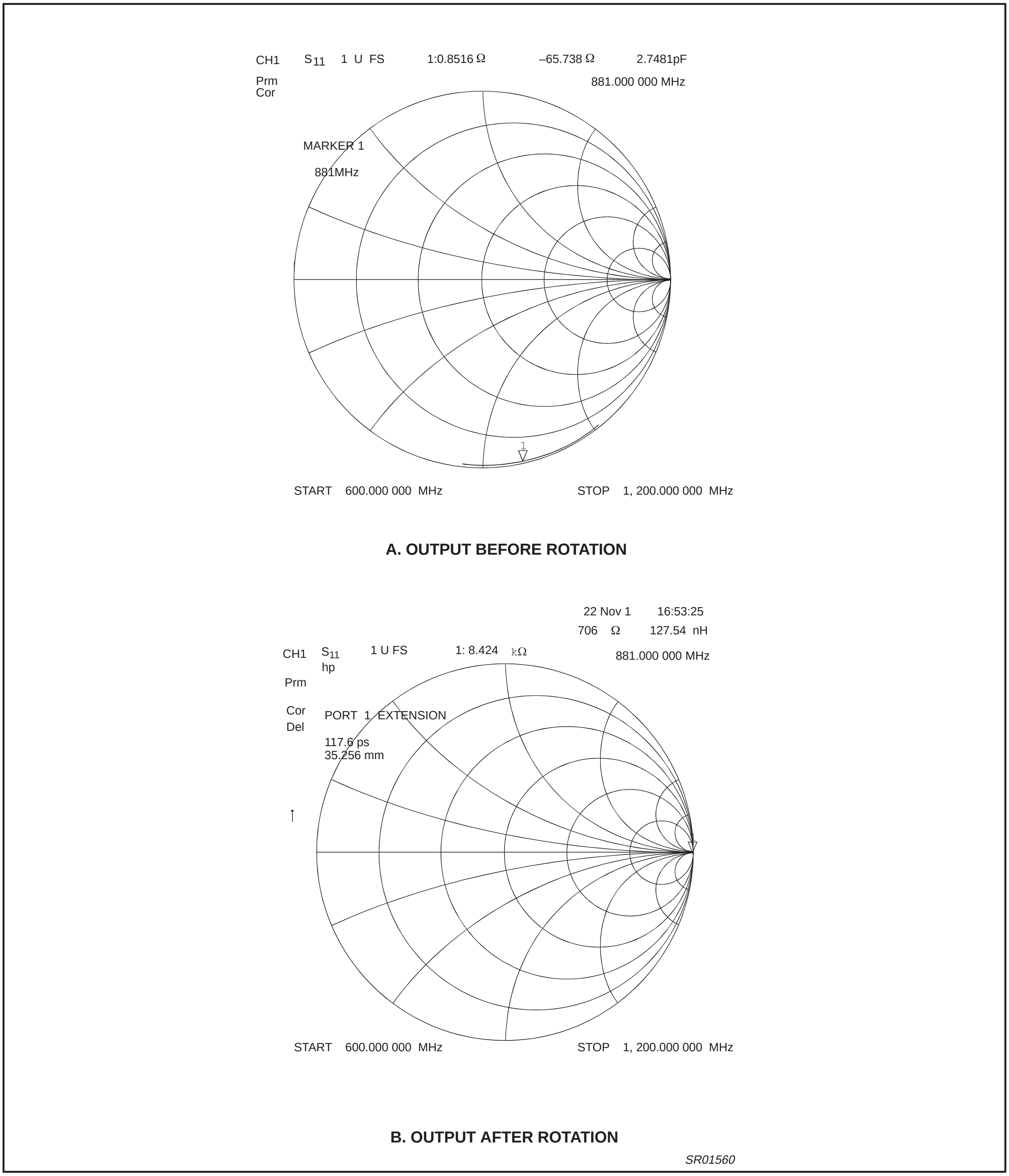
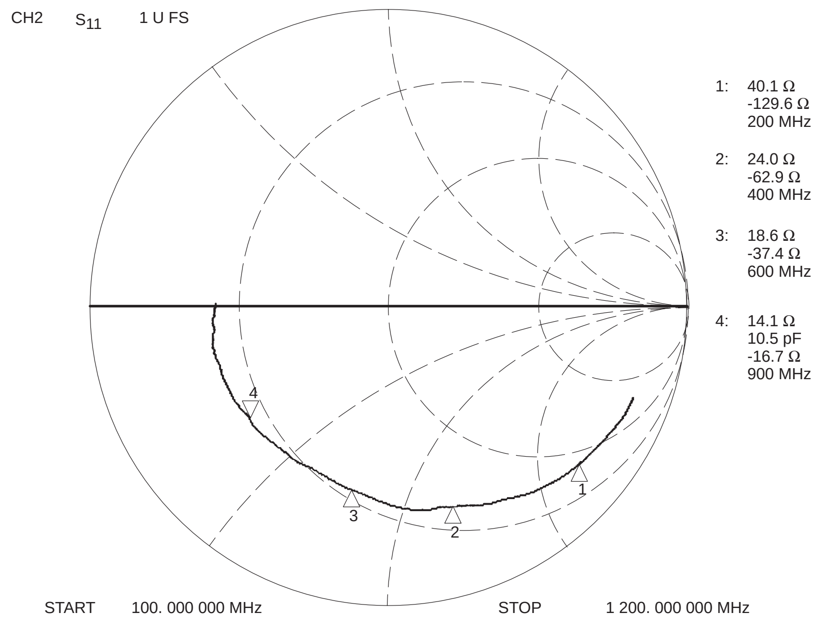
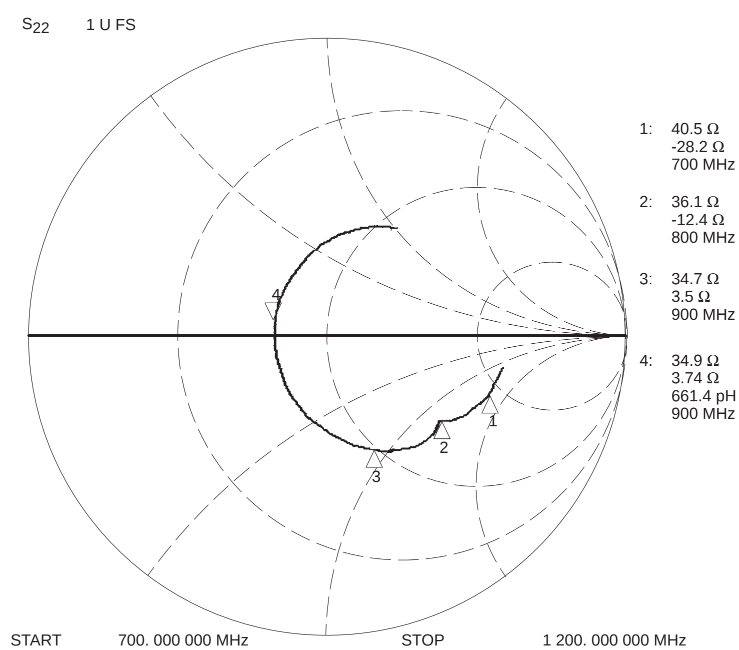


Figure 2 . Network Analyzer Calibration

Measured LNA input and output S-parameters (S11 and S22) are shown below (Figures NO TAGa and NO TAGb). A full set of S-parameter data is available in the SA621 and SA611 data sheets.



A. S11 DATA



B. S22 DATA

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Figure 3 . Measured LNA S11 and S22 Data

MAXIMUM GAIN

Data sheet S-parameters only inform us of the LNA's performance when the source and load terminations are 50 ohms. Since the input and output impedances of this device are not 50 ohms, some return loss will be associated with the mismatch to 50 ohms at each port. These losses can be eliminated simply by conjugately matching each of them to 50 ohms. The standard first-order equation describing the resulting maximum unilateral gain (G_{umax}) is given by (1.1a) below (many good references exist treating the theory and practice of S-parameters [1] and [2]).

$$G_{umax} = \frac{1}{1 - |S_{11}|^2} \, |S_{21}|^2 \, \frac{1}{1 - |S_{22}|^2} \, , \quad (S_{12} = 0) \qquad (1.1a)$$

This equation assumes the LNA is unconditionally stable when conjugately matched. The first and last terms describe the result of conjugately matching both the input and output ports respectively, thereby gaining back the losses due to mismatches present when S_{11} and S_{22} were determined at whatever impedance they were measured at, 50 ohms with the network analyzer in this case. It should be clear that if both LNA ports miraculously exhibited 50 ohm impedances, S_{11} and S_{22} would be 0 (no reflections!) and these terms would reduce to unity leaving only $|S_{21}|^2$, the unilateral transducer gain (remember, since scattering parameters are voltage ratios, they must be squared to obtain power). This term is an invariant property of the LNA itself that depends externally only on frequency but not source and load terminations. Finally, this is a *practical* equation since it assumes that the reverse transmission gain is zero (S_{12}= 0), making the device *unilateral*. Specifically, it does not account for the fact that when the load termination changes S_{11} also changes slightly and vice-versa.

The complete expressions for the non-unilateral or *bilateral* maximum gain G_{max} when $S_{12} \neq 0$ are:

$$G_{max} = \left|\frac{S_{21}}{S_{12}}\right| \left(K + \sqrt{K^2 - 1}\right) \text{ when } B < 0 \qquad (1.1b)$$

$$G_{max} = \left|\frac{S_{21}}{S_{12}}\right| \left(K - \sqrt{K^2 - 1}\right) \text{ when } B > 0 \qquad (1.1c)$$

where the parameter $B = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2$, and K and Δ are from (1.2) and (1.3) which are described next. G_{umax} is much easier to calculate, and usually is close enough to the actual value Gmax to warrant using it as a good approximation for design purposes.

STABILITY

As noted above, (1.1a,b,c) can only be used if the amplifier in question is stable when conjugately matched. Some amplifiers will oscillate with certain source and load terminations; these devices are *conditionally stable.* Otherwise, they are *unconditionally stable* since no passive source and load terminations exist that will cause oscillation. Stability is easily determined from measured S-parameters. This involves finding a stability factor *K*(scalar number) and Δ (a complex number), where

$$K = \frac{1 + |S_{11}S_{22} - S_{12}S_{21}|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{21}S_{12}|} = \frac{1 + |\Delta|^2 - |S_{11}|^2 - |S_{22}|^2}{2|S_{21}S_{12}|} \qquad (1.2)$$

and

$$\Delta = S_{11}S_{22} - S_{12}S_{21} \qquad (1.3)$$

If $K > 1$, the device will be *at least* stable when simultaneously conjugately matched to 50 ohms, and Gumax or Gmax can then be found. If $K>1$ *and* $|\Delta|<1$, then the device is also unconditionally stable with any termination. Note that *K* only tells us whether the device is stable when conjugately matched to the same characteristic impedance used to obtain the measured S-parameters, usually 50 ohms. The SA621/611 LNA is unconditionally stable at all frequencies between 100 and 1200 MHz.

Conditionally stable devices require further analysis to determine regions on the Smith Chart where they are stable. This is done by determining classic *source* and *load* stability circles. Not only must their location and radii be determined, but also whether the regions inside or outside the circles constitute stable terminations. This treatment is beyond the scope of this application note and will not be discussed here. Those interested in an excellent and readable treatment should refer to [2] as a good example.

EXAMPLE 1

Example showing the LNA to be unconditionally stable at 870 MHz:

Substitute S-parameters from Table 1 below into stability equations (1.2) and (1.3). Note: S-parameter data are vector quantities having both magnitude and phase; therefore, care must be taken to use vector arithmetic where necessary in the equations given in this application note.

Table 1. S-Parameter Data at 870 MHz

$\|S_{11}\|$	∠°	$\|S_{12}\|$	∠°	$\|S_{21}\|$	∠°	$\|S_{22}\|$	∠°
0.596	–135.8	0.027	111.9	4.81	26.9	0.179	–175.1

Find $|\Delta|$ first, then stability factor *K*:

$$|\Delta| = |S_{11}S_{22} - S_{12}S_{21}|$$

$$|\Delta| = |0.596 \cdot 0.179 \angle(-135.8 - 175.1) - 0.027 \cdot 4.81 \angle(111.9+26.9)| = 0.168$$

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

$$K = \frac{1 + 0.168^2 - 0.596^2 - 0.179^2}{2 \cdot 0.027 \cdot 4.81} = 2.25$$

Since $K>1$ and $|\Delta|<1$, the LNA is unconditionally stable at 870 MHz.

Thus, the maximum unilateral gain can be found from (1.1a),

$$G_{umax} = \frac{1}{1 - |S_{11}|^2} |S_{21}|^2 \frac{1}{1 - |S_{22}|^2} = (1.551)\ (4.81)^2\ (1.033) = 37.1 = 15.69\ \text{dB}$$

and the bilateral gain from (1.1b,c), found by computer,

G_{max} = 15.77 dB.

These two results differ by less than 0.08 dB and show the value of the simpler unilateral form. Note also that gain measured by the network analyzer would be just $10\log(|S_{21}|^2)$ = 13.6 dB. Thus, we conclude that without any matching (50 ohm source and load terminations) we can expect a gain of 13.6 dB, while the most obtainable under any matching conditions is about 15.8 dB.

GAIN AND NOISE CIRCLES

Gain Circles

Input gain and noise circles plotted on the Smith Chart greatly facilitate the design of input matching networks by graphically showing the designer the tradeoff between any particular gain and wanted noise figure. The use of input gain and noise circles to achieve this end is discussed next. With the calibration board, S-parameter data is accurately referred right to the pin of the IC. From the measured S-parameters, LNA gain can be calculated for *any* matching network attached to either the input or the output pins.

Consider the case when the input matching network has a source reflection coefficient Γ_S (scattering parameter looking *into* the *output* of the network connected to the LNA input pin) equal to the conjugate of the input reflection coefficient S_{11} (scattering parameter seen looking *into* the LNA input IC pin) (Figure 4).

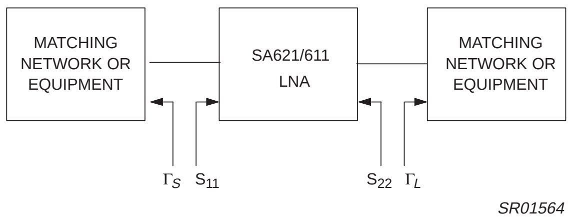

Figure 4 . Amplifier Setup

Thus,

$$\Gamma_S = S_{11}^*, \text{ unilateral device } (S_{12} = 0). \tag{1.4a}$$

$$\Gamma_S = \left(S_{11} + \frac{S_{12}S_{21}\Gamma_S}{1 - \Gamma_L S_{22}}\right)^*, \text{ bilateral device } (S_{12} \neq 0). \tag{1.4b}$$

This condition will plot Γ_S as a single point on the Smith Chart that can be used to design a suitable matching network to whatever source is driving the network's input, typically 50 ohms. Note that (1.4a) is quite simple and often differs little from (1.4b), especially for quick estimations. Comparison of (1.4a) and (1.4b) clearly shows the effect the load has on the LNA input when $S_{12} \neq 0$.

If a similar procedure is used to match the output, we have the unique situation described by equation (1.1a), where the LNA exhibits its maximum possible gain, G_{umax} If we stipulate that the output always remain conjugately matched, but introduce a deliberate input mismatch, the gain G will obviously be less than the maximum possible. For any particular $G<G_{max}$, many different Γ_S exist that will result in this same gain. It can be shown that the locus of these various Γ_S all lie on a single circle on the Smith Chart [1] [2]. Therefore, choosing any point on this circle defines a particular matching network all of which yield the same return loss or mismatch and thus the same gain G. In the time domain, this simply means the amplitude ratio of the forward and reverse traveling waves remains unchanged, but the phase between them will be different for each point on the circle.

Calculating input gain circles can be accomplished using either simplified unilateral equations or the more complicated bilateral forms. The former are useful for practical designs and can easily be done by any good engineering calculator. For an in–depth discussion into this topic refer to *Hewlett Packard Application Note 154* [1]. Reference [2] contains a well-developed complete treatment for an actual bilateral device like the SA621/SA611, with calculations that are most efficiently handled by a computer or calculator program.

Since the device will no longer be conjugately matched, (1.5) expresses the unilateral gain form of (1.1a) rewritten for *any* source or load terminations:

$$G_u = \frac{1 - |\Gamma_S|^2}{|1 - S_{11}\Gamma_S|^2} |S_{12}|^2 \frac{1 - |\Gamma_L|^2}{|1 - S_{22}\Gamma_L|^2} \text{ where } G_u \text{ is always } \leq G_{umax}. \tag{1.5}$$

As stated earlier, the output will remain conjugately matched; only the input mismatch will be varied through Γ_S. For this special case, (1.5) becomes:

$$G_{su} = \frac{1 - |\Gamma_S|^2}{|1 - S_{11}\Gamma_S|^2} |S_{21}|^2 \frac{1}{1 - |S_{22}|^2} \tag{1.6}$$

Since any arbitrary value of G_{su} is between 0 and G_{umax}, solutions for Γ_S lie on a circle. We can choose any Γ_S and calculate its gain (G_{su}). So if we choose Γ_S along a 15.2dB circle, we would have a G_{su} of 15.2dB. Equations (1.7 – 1.9) from [1] determine where a gain circle can be found on the Smith Chart.

$$d_{ui} = \frac{g_{ui}|S11|}{1 - |S_{11}|^2(1 - g_{ui})} \angle S_{11}^*, \text{ gain circle center} \quad (1.7)$$

$$R_{ui} = \frac{\sqrt{1 - g_{ui}}\ (1 - |S_{11}|^2)}{1 - |S_{11}|^2\ (1 - g_{ui})}, \text{ gain circle radius} \quad (1.8)$$

where g_{ui} is a parameter expressing the gain circle's wanted gain G_{ui} normalized to G_{max}.

$$g_{ui} = G_{ui}(1 - |S_{11}|^2) = \frac{G_{ui}}{G_{umax}} \quad (1.9)$$

Note that d_{ui} is the distance from the center of the Smith Chart to the center of the gain circle along the conjugate S_{11}vector (S_{11}*) and R_{ui} is the radius of the circle at that point.

EXAMPLE 2

Calculate and plot on the Smith Chart an input gain circle 0.5dB below the maximum gain for the SA621/SA611 at 870 MHz using the simplified unilateral assumption and compare it to the general bilateral case.

Begin by finding g_{ui}:

$$g_{ui} = G_i(1 - |S_{11}|^2) = \frac{G_i}{G_{umax}}$$

Choose $G_i = G_{umax}$ – 0.5 dB = 15.7 dB – 0.5 dB = 15.2 dB.

Linearize the gain and find g_{ui}:

$$g_{ui} = 10^{\frac{15.2 - 15.7}{10}} = 0.891$$

Find the center, C_{ui}:

$$C_{ui} = \frac{g_i|S_{11}|}{1 - |S_{11}|^2\ (1 - g_{ui})} \angle S_{11}^*$$

$$C_{ui} = \frac{(0.891)\ (0.596)}{1 - (0.596)^2\ (1 - 0.891)} \angle - (-135.8)^o = \frac{0.530}{0.9614} \angle 135.8^o = 0.55\angle 135.8^o$$

Find the radius, R_{ui}:

$$R_{ui} = \frac{\sqrt{1 - g_{ui}}\ \left(1 - |S_{11}|^2\right)}{1 - |S_{11}|^2(1 - g_{ui})} = \frac{\sqrt{1 - 0.891}\ \left(1 - (0.596)^2\right)}{0.9614} = 0.22$$

Thus, the center will be located at $0.55\angle 135.8°$ with radius of 0.22.

Compare this result with a computer derived bilateral center and radius for the same G_{max} – 0.5 dB gain circle [2]:

Gain Circle for G_{max}– 0.5 *dB* = 15.3 dB

$d_{ci} = 0.56\ \angle 133.4°$

$R_{ci} = 0.22$

The radii are equal with the centers differing only slightly and the gains by 0.1 dB. Now that all the information is found, a designer can draw these gain circles on the Smith Chart for visual comparison. Figure 5 shows these two gain circles plotted on the Smith Chart.

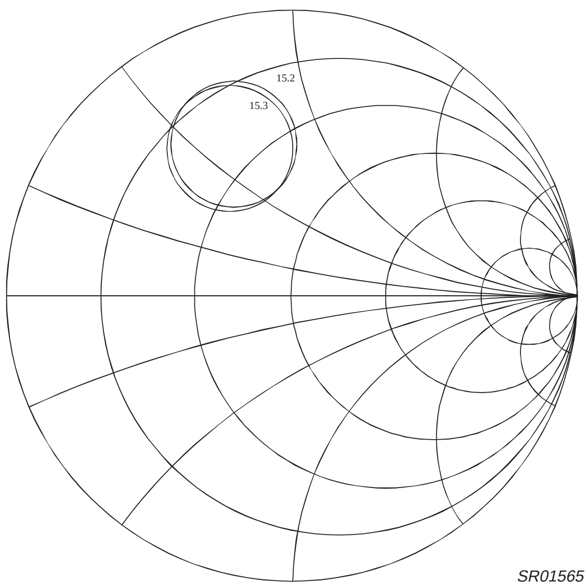

Figure 5 . Plot of Example–1: Unilateral (15.2dB) and Bilateral (15.3dB) Gain Circle

Noise Figure Circles

Noise factor *NF* is defined as the ratio of the signal-to-noise ratio at the input to the signal-to-noise ratio at the output,

$$NF = \frac{S_i/N_i}{S_o/N_o}$$

It is always greater than 1 for practical devices. Noise figure *F* is simply *NF* expressed in dB, *F = 10log(NF)*. Note that a perfect device having no internal noise generation has noise factor of 1 which yields a noise figure of 0 dB.

The LNA noise figure depends on the source impedance connected to its input port and frequency of operation. Further, the minimum noise figure, F_{min}, increases monotonically with frequency. Like most devices, F_{min} does not occur when the source is 50 Ωs. Thus, if a designer conjugately matches the input to realize maximum gain, observed noise figure *F* will be greater than F_{min}. How much greater? Similar to gain circles, noise circles can also be determined and plotted as an indispensable design aid to answering this question. Each circle describes a locus of source reflection coefficients Γ_S all having a constant $F>F_{min}$. Essentially, this is the same thing we did for gain circles except here the parameter is noise figure rather than gain.

Remember that gain circles are found by calculation employing the four basic S-parameters known at a particular frequency. Noise figure circles are based on three additional parameters: F_{min} at Γ_{min} and R_n. If these are not known and a designer does not have automated testing equipment, accurately determining them can be an exacting and time-consuming chore. An experimental method for obtaining them is presented in *Hewlett Packard Application Note 154* [1]. This method requires two basic measurements: One to determine what input reflection coefficient $\Gamma_S = \Gamma_{min}$ is required for best noise figure F_{min} and one additional point used to find the equivalent noise resistance R_n, described below in (1.10). The value and limitations of this method when applied to the LNA will be discussed later. After obtaining these quantities, any wanted noise circle may be found and plotted on the Smith Chart.

The following equation [1][2] expresses the fundamental relation between any noise figure *F* and the minimum noise figure F_{min} (at Γ_{min}) as a function of Γ_S.

$$F = F_{min} + 4r_n \frac{|\Gamma_s - \Gamma_{min}|^2}{|1 + \Gamma_{min}|^2 \, (1 - |\Gamma_{min}|^2)} \tag{1.10}$$

where

$$r_n = \frac{R_n}{Z_o} \tag{1.11}$$

F	noise figure $\geq F_{min}$
S	source reflection coefficient at F
F_{min}	minimum noise figure possible for the device
Γ_{min}	source reflection coefficient at minimum noise figure
R_n	equivalent noise resistance in Ohms
Z_o	characteristic impedance of system, usually 50 Ohms
r_n	normalized equivalent noise resistance

Note that R_n is a parameter having dimensions of Ωs. It is *not* the actual Thevenin source resistance required for minimum noise figure; that can be found from Γ_{min}. Rather, it is a scalar factor gauging the dependence of *F* on Γ_S. This should be evident by inspection of (1.10). Also, Smith Charts usually have the center normalized to be 1 Ω, so only normalized quantities can be plotted. Thus, the normalized equivalent noise resistance r_n must be found.

Once r_n is found, it is used with min and F_{min} to determine a family of circles representing different noise figures found by applying (1.13) and (1.14) for each one as follows.

$$C_{Fi} = \frac{\Gamma_{min}}{1 + N_i} \qquad \text{noise circle center} \qquad (1.12)$$

$$R_{Fi} = \frac{1}{1 + N_i} \sqrt{N_i^2 + N_i \left(1 - |\Gamma_{min}|^2\right)} \qquad \text{noise circle radius} \qquad (1.13)$$

where

$$N_i = \frac{F_i - F_{min}}{4r_n} \left[1 + \Gamma_{min}\right] \qquad F_i > F_{min} \qquad (1.14)$$

C_{Fi} determines where the noise circle is located, and is found by measuring C_{Fi} from the center of the Smith Chart along the vector $\angle C_{Fi.}$ R_{Fi} is the radius of the circle then constructed at that point. Note that S_{12} does not appear in these equations (1.10 to 1.14), so they are applicable to both unilateral and bilateral devices.

The necessary parameters were determined using the following setup:

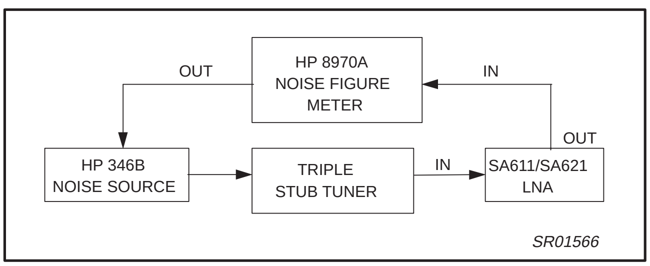

Figure 6 . Noise Circle Setup for Measuring Noise Figure

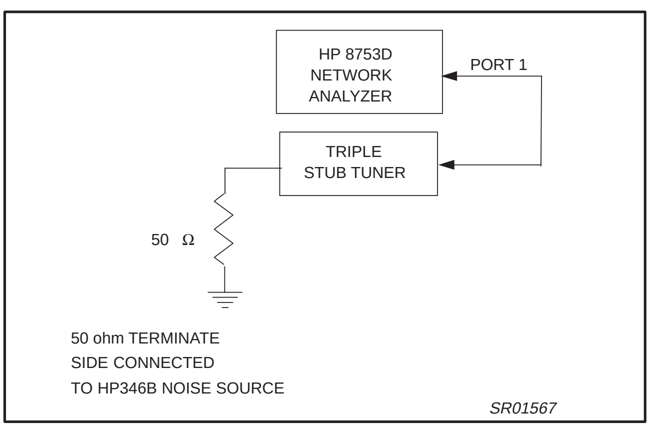

Figure 7 . Noise Circle Setup To Determine Location

The setup consists of two major pieces of equipment, a noise figure meter and a network analyzer and follows the general method discussed in *Hewlett Packard Application Note 154* [1]. The following is for those familiar with test procedures when measuring source reflection coefficients, Γ_S. The point to remember, here, is we are finding reflection coefficients looking *into* the *output* of the network that was connected to the *input* of the LNA. When we measure Γ looking into the LNA we label it S_{11}. When we measure Γ looking into the output of a network attached to the LNA's input, we label it Γ_S.

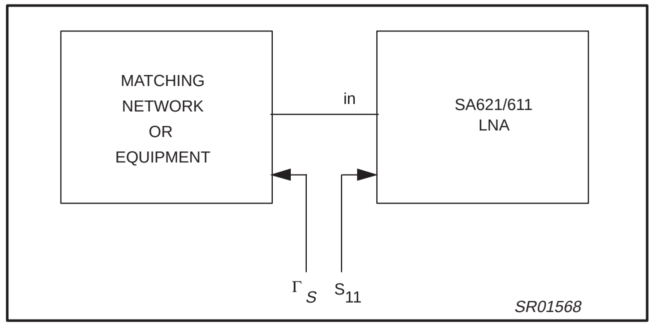

Figure 8 . Input Matching Network Showing Source Reflection Coefficient Γ_S and Input Reflection Coefficient S_{11}

1. Using the noise figure meter (Figure 6), find the minimum noise figure F_{min} of the device by carefully tuning the triple-stub tuner. Most noise figure meters express to 3 significant digits (e.g. 1.73 dB), so some physical interpolation of the stub elements may be necessary to set the triple-stub to a good estimate of where the minimum actually occurs. The reflection coefficient looking into the stub will be equal to Γ_{min}.
2. To measure Γ_{min}, connect a 50 Ω termination to the triple-stub tuner on the *same* side where the noise source was connected (Figure 7). Measure the reflection coefficient Γ of the tuner on a network analyzer. This will be the same as $F_{s\,=\,}F_{min}$. Note that on the HP8753D network analyzer is displayed when the format is "Polar". If the format is "Smith", then the display will give impedance Z even though the reflection coefficient plane is actually being displayed. The formulas require reflection coefficients not impedances. However, one can easily convert Z to by using the familiar relation:

$$\Gamma = \frac{Z_S - Z_L}{Z_S + Z_L} \qquad (1.15)$$

where, Z_S and Z_L are the source and load impedance, respectively. Because the HP network analyzer normalizes to 50 Ωs, the equation becomes:

$$\Gamma = \frac{Z_S - 1}{Z_S + 1} \qquad (1.16)$$

3. This procedure can be repeated for measurements of different noise figures, not just F_{min} .
4. Please keep in mind that *any* added lengths and attenuations need to be *accurately* accounted for. Depending on hardware, this can be tedious and complicated. For a simple example, if an SMA-to-SMA connector adaptor, often referred to as a barrel or through, is attached to the triple-stub tuner (that was not used when the stub was connected to the LNA) to obtain Γ_S, it will exhibit some time delay causing rotation on the Smith Chart as well as attenuation. Particular attention must be given to attenuations not accounted for when the noise figure apparatus was calibrated. Thus, attenuations of *all additional* RF hardware needs to be accounted for: including the stub-tuner, barrels, etc. Errors of 0.05 dB will typically be observable as changes in the diameter of the plotted noise circles.

EXAMPLE 3

Using the described noise circle setup, determine and plot the 2.0 dB noise circle on the Smith Chart for the SA621/SA611 at 881 MHz.

This is accomplished by using the triple-stub tuner to find all the necessary variables mentioned above. Following the HP procedure as a guide, we first find F_{min} and Γ_{min} using the noise figure meter and triple-stub tuner. Finding r_n requires another determination of any F and its associated Γ_S. Eqn. (1.10) is then solved for r_n, the named quantities substituted and r_n found. The HP procedure uses an especially simple and mathematically convenient way for finding this by letting $\Gamma_{S=0}$; this occurs when the source termination is 50 Ω or just the noise figure reading *without* the triple-stub tuner. This makes sense, but unfortunately does not work well with the SA621/SA611 LNA. It turns out the two noise figures are quite close, F_{min} and F at 50 Ω, and each of these is expressed to only three significant figures; after linearizing their decibel forms, the resulting numerical error can be quite large. Further, the uncertainty in finding Γ_{min} with the stub tuner contributes to a larger overall uncertainty. Using this method, r_n has been observed to vary between 3 and 20 Ω, an unacceptably large variance. Philips has developed a better method to experimentally determine and verify these parameters.

The two re-written forms of (1.10), general and special cases respectively, solved for r_n, are given as:

$$r_n = \frac{\left(F - F_{min}\right)\left(\left|1 + \Gamma_{min}\right|^2 (1 - \left|\Gamma_S\right|^2)\right)}{4\left(\left|\Gamma_S - \Gamma_{min}\right|^2\right)} \tag{1.17a}$$

$$r_n = \frac{\left(F - F_{min}\right)\left(\left|1 + \Gamma_{min}\right|^2\right)}{4\left(\left|\Gamma_{min}\right|^2\right)} \tag{1.17b}$$

Eqn. (1.17b) is used with the HP procedure, while (1.17a) the Philips procedure. Additional measurements are taken at random noise figures, ideally scattered around the Smith Chart, each yielding an associated r_n found by applying (1.12a) and multiplying by Z_0. Simple mean and variance or sample standard deviation are then calculated for the ensemble of r_n. The mean is taken to be the best estimate of r_n. Further, inspection of the r_n series can usually pin-point experimental errors when especially bad points turn up.

Assuming rotations and attentions have been properly accounted for (see point 4 above), the weak point in this procedure is the estimate of Γ_{min} since it is based on only one measurement subject to considerable error. This fact can only be appreciated by actually doing the laboratory measurement and getting a feel for the latitude one has in setting the triple-stub tuner at minimum noise figure. A further refinement is then possible using the measurement series of ordered pairs F and Γ_S for each r_n. A gradient search algorithm can be developed that finds a better estimate of Γ_{min} based on minimizing the ratio of the mean to variance of the ensemble r_n for each candidate Γ_{min}. Clearly, this is not suitable for hand calculator computations, but is easily done by a computer. First order approximations suitable for experimental design may be composed of several (at least 3) measurements for F and Γ_S to ascertain “whether we are in the ballpark” in the value settled upon for r_n. This is easily done on a hand calculator or with computer assistance using such common programs as Matlab.

Let’s continue with this example using the Philips’ method. A particular session in the screen room yielded the following actual set of experimental measurements using the calibration board for the LNA at 881 MHz:

Table 2. Noise Figure Measurements for Nine Samples

n	F(dB)	GAMMA	
SAMPLE	NOISE FIGURE	MAGNITUDE	ANGLE
1	1.60	0.170	126.1
2	4.53	0.737	15.0
3	3.15	0.551	–96.8
4	2.04	0.320	–131.1
5	2.45	0.428	–105.6
6	3.89	0.684	–102.2
7	3.38	0.706	–165.0
8	1.73	0.058	–91.0

Sample calculation using (1.17a for F = 2.03 dB at $\Gamma_S = 0.32 \angle -131.1°$,

$$r_4 = \frac{\left(\log^{-1}\left(\frac{2.03}{10}\right) - \log^{-1}\left(\frac{1.60}{10}\right)\right)\left(|1\angle 0^o + 0.170\angle 126.1^o|^2\left(1 - |0.32|^2\right)\right)}{4\left(|0.32\angle -131.1^o - 0.170\angle 126.1^o|^2\right)} = \frac{(0.150)(0.791)(0.898)}{4(0.131)}$$

$r_4 = 0.203\ \Omega$

$R_4 = r_4 Z_o = (0.203)\ (50) = 10.2\ \Omega$

Using a computer, better estimates of Γ_{min} and r_n were found to be:

$$\hat{\Gamma}_{min} = 0.174\angle 131.6^o, \text{ so } \hat{R}_n = 9.0\Omega$$

To find the 2.0 dB noise figure circle we apply eqns.(1.12–1.14) using the non-optimized values for Γ_{min} and r_n for comparison purposes.

Find parameter N_I:

$$N_i = \frac{F_i - F_{min}}{4r_n}\left[1 + \Gamma_{min}\right]$$

$$N_i = \frac{\log^{-1}\left(\frac{2.0}{10}\right) - \log^{-1}\left(\frac{1.60}{10}\right)}{4\left(\frac{10.2}{50}\right)}|1\angle 0^o + 0.170\angle 126.1^o|^2 = \frac{(0.140)(0.791)}{0.816} = 0.135$$

Find where the noise circle is centered, C_{FI}:

$$C_{Fi} = \frac{\Gamma_{min}}{1 + N_i} = \frac{0.170\angle 126.1^o}{1.135} = 0.150\angle 126.1^o$$

Find the noise circle's radius, R_{FI}:

$$R_{Fi} = \frac{1}{1 + N_i}\sqrt{N_i^2 + N_i\left(1 - |\Gamma_{min}|^2\right)} = 0.881\sqrt{0.149} = 0.340$$

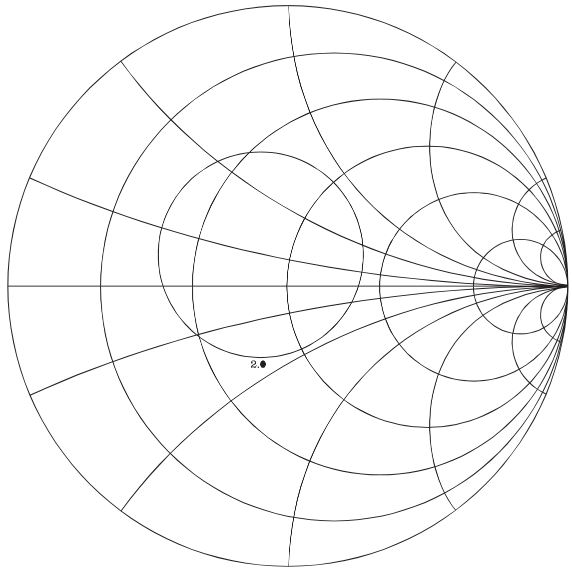

Figure 9 . 2.0 dB Noise Circle at 881 MHz

Piecing everything together

When gain circles, noise circles, and stability are all determined, the designer is ready to design a match to realize the optimal noise figure and gain combination for a specified application. Figure 10 shows the combined gain and noise circles on a single Smith Chart computed for 881 MHz. The noise circles were found from the data set in example 3 above with Γ_{min} and R_n optimized by a gradient search algorithm with the experimental data from Table 2. Compare especially the 2.0 dB noise circle with that found using the non-optimized Γ_{min} and r_n from example 3 shown in Figure 9 . The optimized location for this circle is $\hat{C}_{Fi} = 0.151\angle 131.6^{o}$ with a radius $\hat{R}_{Fi} = 0.361$

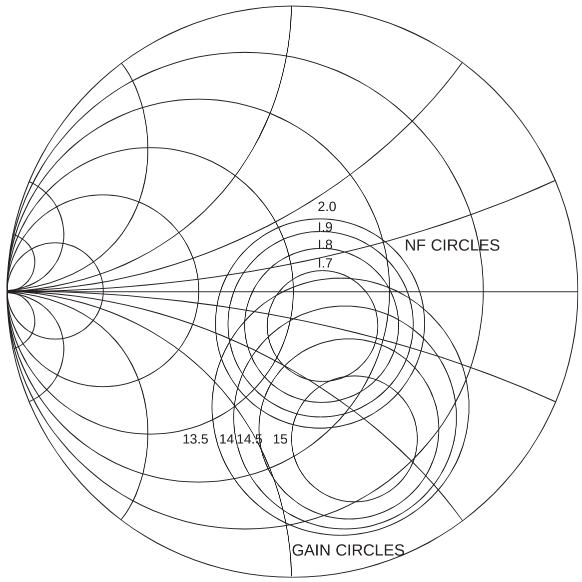

Figure 10 . Noise and Gain Circles at 881 MHz

Since the LNA is unconditionally stable at 881 MHz , there are no matching conditions that will cause it to oscillate. Therefore, interpreting the data in Figure 10 is easy: any input reflection coefficient Γ_S can be located as a point on the Smith Chart. The gain and noise circles tell us immediately what the expected gain and noise performance will be at this point, and we can proceed to design a suitable input matching network to realize this particular Γ_S.

Say a designer wants a noise figure of 1.8dB and a gain of 15dB when the source driving the LNA input is 50 Ω. On the Smith Chart, this noise figure and gain point are at approximately $Z_S = 31 + j20\ \Omega$. Since the LNA "sees" Γ_S, we begin from the center of the Chart, corresponding to the source of impedance of 50 Ω and move using suitable series or shunt elements to the Γ_S point. Thus, we choose components to match to the ending point (connected to the input of the LNA). Usually, more than one solution exists. In this case shunt C and series L are used. (Figure 11).

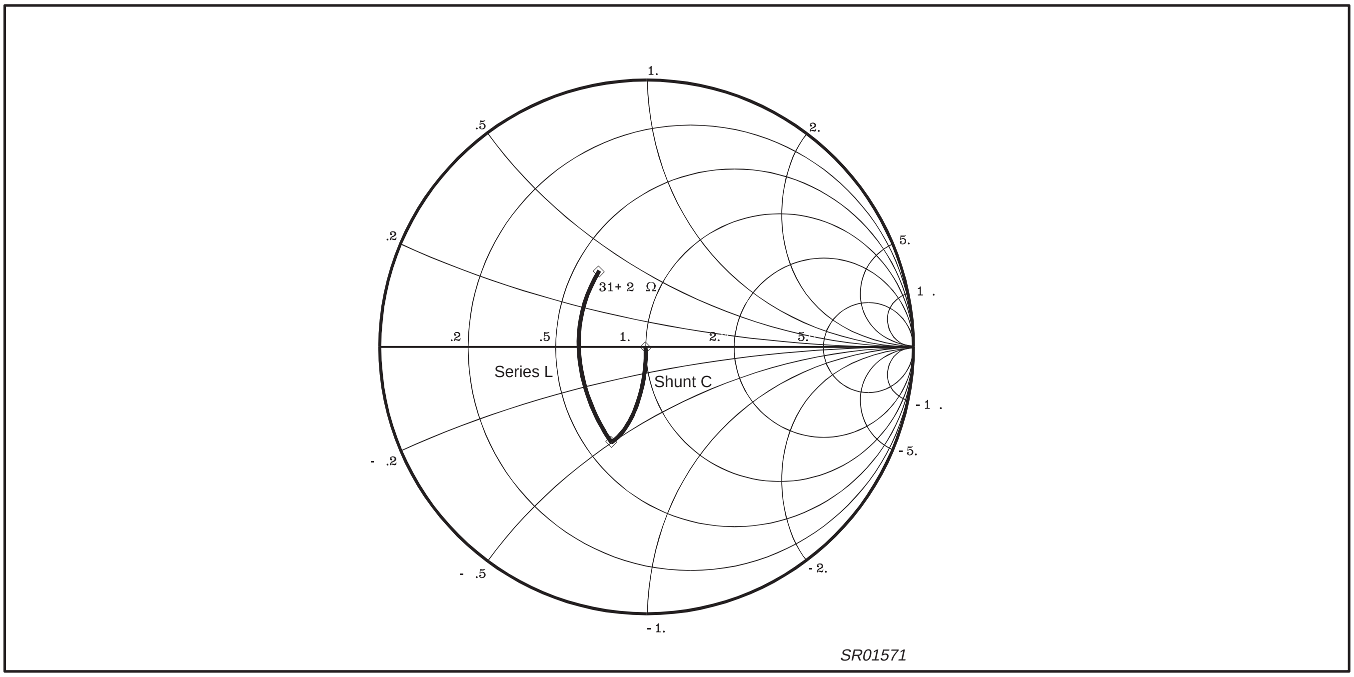

Figure 11 . Smith Chart Display of Matching

The circuit to match the input of the LNA is seen in Figure 11.

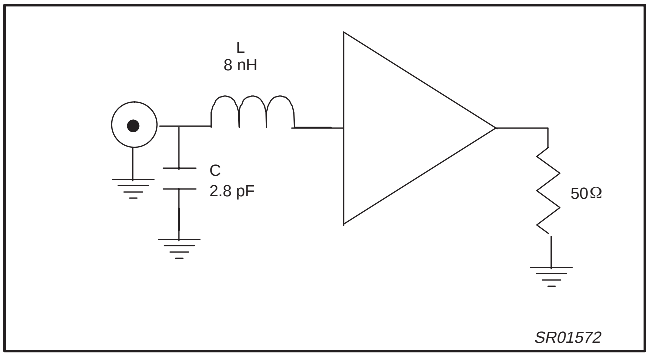

Figure 12 . LNA Input Matching

APPLICATION DEMONSTRATION BOARD (DEMO BOARD)

SA621/611 LNA

By conjugately matching the output port to 50 Ωs, the designer can easily improve the gain. 50 Ωs is desirable because most RF test equipment is designed for a 50 Ω characteristic impedance, and most image reject filters are specified in a 50 Ω environment.

On the application board however, no output matching was used because the output return loss without matching is about 10dB, so only about 0.5dB of gain is lost due to impedance mismatch. However, a designer could improve the output match, and thus the LNA gain, by 0.4dB by using the low pass circuit below (Figure 13), which also has the advantage of reducing high frequency noise. One is not constrained to use only this circuit, but may choose to use a high pass circuit which uses one less external component for those situations where parts count is more important than noise filtering.

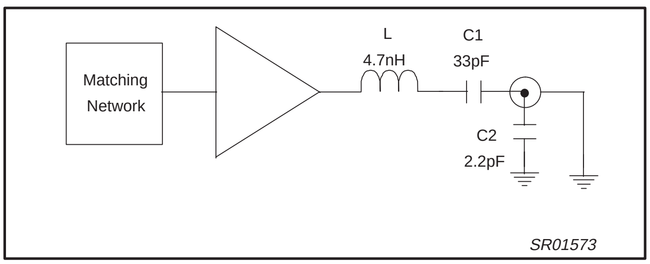

Figure 13 . LNA Output Circuit

The input matching circuit (Figure 14) is basically a simple shunt L match. The network may seem trivial but its realization is not. For example, placement of the 0805 size SMD inductor is critical because the necessary interconnecting transmission lines and the spatial size of the inductor itself will rotate the theoretical point to another position on the Smith chart. Therefore, the actual impedance being matched will vary depending on where the designer places the 6.8nH inductor with respect to the LNA input pin. This is always a consequence of using lumped parameter solutions at microwave frequencies rather than transmission line segments; the advantage is reduced board space at the expense of greater difficulty in getting the match to work properly. These devices should be as small as possible consistent with wanted efficiency or element Q. Inductors are typically size 0805 or larger while capacitors are 0603 or larger.

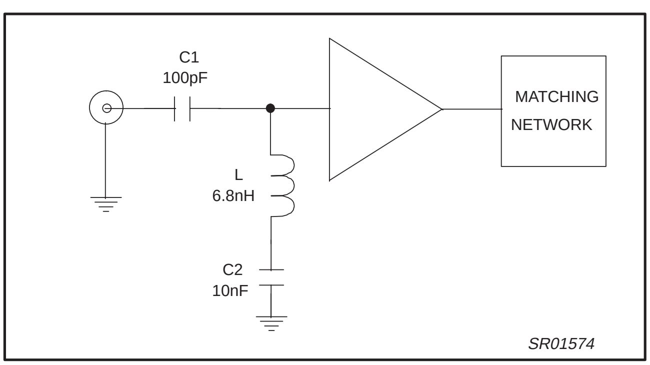

Figure 14 . LNA Input Match

Note that the input matching network also includes a 100pF capacitor for DC blocking as well as a 10nF capacitor. The 10nF capacitor plays two roles in this circuit. First, DC-wise, the capacitor prevents shorting the DC bias potential present on the input pin through the 6.8nH inductor to GND. Second, the 10nF capacitor introduces destructive phase cancellation of the third order intermodulation products present at the input and output thus improving the overall LNA IP3 performance. As this capacitor is increased, IP3 performance improves at the expense of LNA power-up turn-on time (turning “on” the LNA after powering it down). Table 3 shows the compromise between turn-on time and input IP3.

Table 3. LNA Input IP3 vs Switching Time

RF Input Level = –30dBm		Gain	Switching Time
Capacitance	IP3in		
10nF	–7.5	15	84uS
22nF	–6.3	14.8	220uS
47nF	–5	15	490uS
100nF	–5	15	720uS

With any UHF RF application, layout is critically important to circuit performance. For matching and RF signal routing, board traces must be viewed and treated as transmission lines or antenna segments. As transmission lines, they affect nodal impedances (impedance at any point on the segment or at its ends) and cause rotations on the Smith Chart. The most common model treats them using classic microstrip theory where they are implemented as traces over a single ground plane. Without a ground plane, a microstrip becomes a trace or wire behaving as an antenna segment susceptible to unwanted coupling and radiation. Thus, designer’s must account for length and width of traces as well as their proximity and relation to associated ground planes. Generally, narrow traces should be avoided because of their high characteristic impedance, greater phase delay and radiation losses. The demo board has microstrip lines connected to both ports of the LNA with associated ground planes on opposite sides of the board. Their lengths were kept as short and as wide as practical. Since truly 50 Ω lines are impractical, being over 100 mils wide on 62 mil thick FR4/5 PCB, the demo board uses traces having characteristic impedances between approximately 75 and 90 Ωs. This makes introducing lumped matching elements simple, because they have predictable electrical length (phase delay) and low radiation thereby improving isolation and minimizing unwanted coupling. Discrete capacitors or inductors should generally be mounted over a ground plane for best isolation. Radiation, especially from inductors, may also need to be taken into account. A FET probe on a spectrum analyzer can be useful in determining whether such elements need shielding or re-location. Elements carrying comparatively high RF currents, in oscillator tank circuits for example, will nearly always require shielding for good isolation and spurious radiation attenuation.

Good port-to-port isolation is necessary to minimize unwanted output-to-input feedback that might result in parasitic oscillations or poor stability. Poor isolation also increases the affect of output termination on the actual input impedance of the device and vice-versa, modifying the actual $S_{11}$and S_{22} seen at input and outputs respectively with particular port terminations. Essentially, this modifies S_{12}. This can make realizing a particular gain and noise figure by design difficult or impossible.

Board layout will always degrade isolation. For example, on the SA621/SA611 demo board, the LNA input is pin 15 and the output pin 13. Since these pins are so close together, special attention must be given to obtaining good isolation between them. This is best accomplished by having the input and output ports on opposite sides of the board to take advantage of ground plane isolation. This layout achieved more than 25dB of isolation.

Another aspect to the layout is providing proper RF ground returns to the LNA pins 14 and 16. Input RF currents should return to pin 14 while output RF currents should return to pin 16. One must keep the ground lengths associated with these port's external components as short as physically possible. Failure to do so can result in poor isolation, instability, and parasitic oscillations.

SA621/611 Mixer

The SA621 mixer utilizes the local oscillator (LO), internal to the IC, as one of the inputs to the mixer. The mixer employs an open-collector output structure which allows the designer to match any high impedance load for maximum power transfer with a minimum number of external components. The SA621/611mixer outputs include internal 10pF capacitors that have an effective value of about 12pF because of parasitic capacitance from the internal bond wires and lead lengths from packaging. The mixer output enables the designer to match for high impedance devices (SAW or Crystal filters) or to 50 Ωs for testing. Reference Application note 1777 to review the details of the mixer output circuit commonly referred to as the current combiner. This portion of the application note gives the designer an intuitive feel of the performance of the mixer and the design trade-offs associated with it.

For the SA611, the mixer circuitry is basically the same except that the LO is driven externally. This allows designers to use a VCO component for the LO.

SA621/611 mixer performance

Table 4 is a comparison of the SA611 and SA621 mixer typical performance parameters.

The SA611 and SA621 has the same performance when comparing gain, 8.7dB of typical. As for input IP3 and noise figure, these parts exhibit some differences. For the SA611 the input IP3 is measured to be about +6dBm, while the SA621 measures typically +4.5dBm. The typical noise figure for the SA611 is 12.5dB while for the SA621 it is 12dB. Not too much difference when looking at noise figure. Lastly, the current consumption of both devices are different because the SA611 does not incorporate the SA621's oscillator section.

Improving SA621/611 mixer Noise Figure

Noise figure can be improved in several ways, however there are some trade-offs. The discussion below will go over these trade-offs.

As always, noise figure can be improved by matching, as seen from the LNA discussion earlier. However, with the mixer, noise circles cannot be easily obtained by the same method used for the LNA because the assumption was made that the LNA output was terminated in 50 Ωs. The mixer, on the other hand, has an open collector output structure where by the output is not 50 Ω terminated. To achieve this the output must be matched to 50 Ω, and because of this match, external component issues become a factor which could yield inaccurate results. All the user can do is match for the best noise figure with the triple stub tuner to achieve minimum noise figure. With the triple stub tuner adjusted for a desired noise figure, the designer can then match from any desired impedance to where the triple stub tuner's impedance lands on the smith chart. In effect the stub tuner is being replaced by a matching network that presents the same load impedance to the mixer as the stub tuner; hence, the same noise figure performance should be obtained.

Table 4. Typical performance comparison of the SA611 and SA621

Part	Gain	Input IP3	Noise Figure	Current
SA611	8.7 dB	+6 dBm	12.5 dB	8mA
SA621	8.7 dB	+4.5 dBm	12 dB	13mA

Method of noise figure improving

As mentioned before, there are trade-offs when improving noise figure. The trade-off here is that component count will increase for improved noise figure, while gain and IP3 remain about the same.

Figure 15 is the schematic of the existing mixer input match on the application demo board. From the S11 data from this application note, the match was determined. This is a low pass matching network to reject high frequencies at the IF output, namely the LO.

An alternative is to use a high pass match (Figure 16), however, experience has shown that the noise figure will increase by 0.5dB to 1dB. The benefit of the high pass match is that the component count is reduced by one because only a shunt inductor and a series cap are needed. Also, because there is one less component, the gain is improved by about 0.25dB while the input IP3 is maintained.

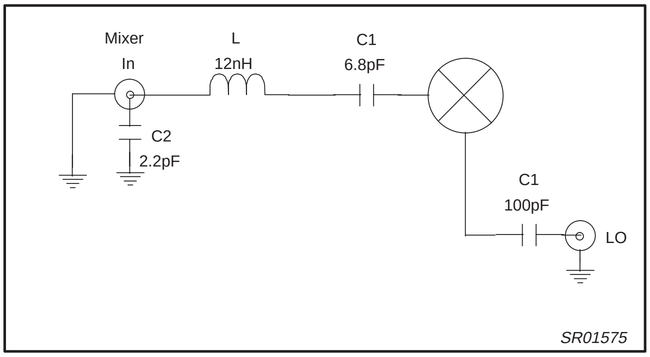

Figure 15 . Existing Mixer Configuration (Low Pass)

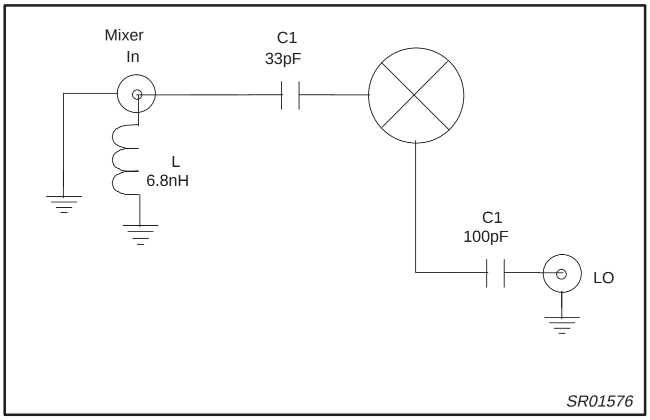

Figure 16 . High Pass Mixer Input Match. (NF Degradation of 0.5db to 1db, Gain Improvement .25db)

If the designer is not pleased with the existing match, he or she could continue to optimize by improving the LO input match. The improvement should be about 0.3dB in noise figure, while gain and input IP3 performance remain the same. Figure 17 shows an additional 3.3nH and 470pF bypass cap. This matching is used to shunt out some of the low frequency noise and improve matching to the LO port.

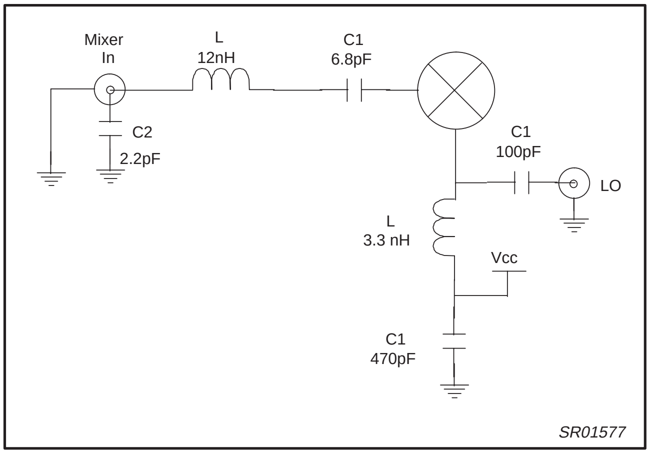

Figure 17 . Mixer with LO matching. (NF Improvement of 0.3dB)

If the designer wishes to improve the mixer noise figure even more at the expense of gain and input IP3, a 6.8KΩ resistor could be used to affect the biasing of the mixer (Figure 18). This improvement does two things, (1) the noise figure is improved nearly 1dB and (2) the mixer current is reduced about 0.5mA. The drawbacks are mixer gain degrades 0.3dB and the input IP3 drops to approximately 1 to 4dBm depending on the current consumption of the part. Further improvements to noise figure can perhaps be achieved by decreasing the resistor even more, however, no attempt was made to do this for this application note.

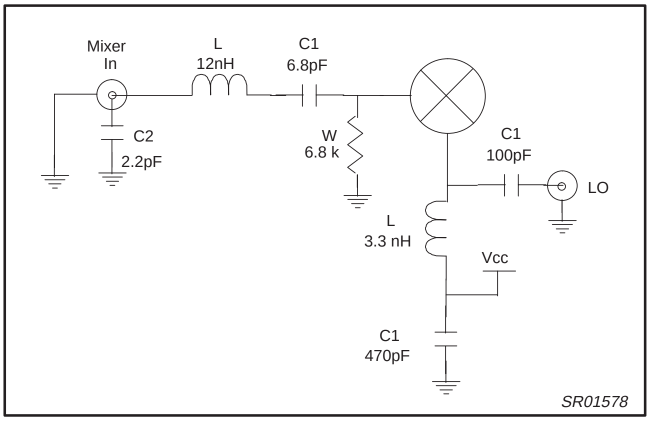

Figure 18 . Mixer Match With Resistor. (Nf Improves 1db, Current Drops 0.5ma, and Gain and IP3in Degrades)

Step by step method to matching the mixer out to 50 Ωs

Many designers have shown interest in using the Philips front-ends but were confused as to how to properly match the mixer output section. The following steps will simplify the task of matching the mixer output at any particular IF frequency or output impedance.

For a complete analysis of the current combiner theory, please refer to AN1777. The procedure to achieve a match to 50 Ωs will be discussed next.

1. Review of the current combiner circuit.

The current combiner circuit consist of two 10pF internal capacitors plus approximately 2pF parasitic pin capacitance due to internal bond wires, packaging, and the external inductor (Figure 19).

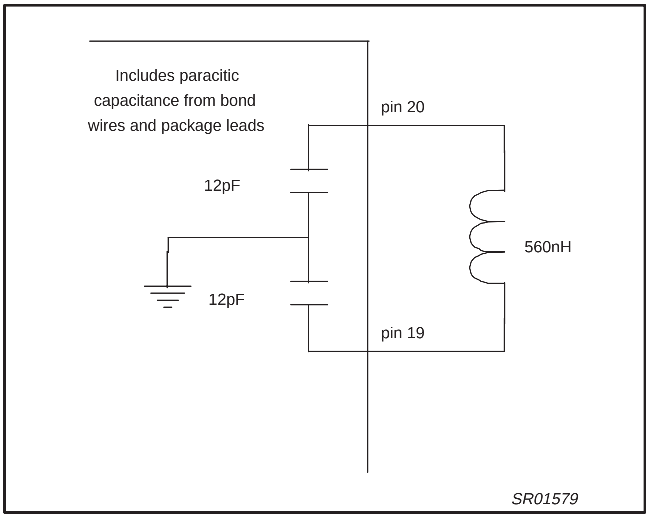

Figure 19 . Current Combiner Circuit

In this case an 83MHz intermediate frequency (IF) was used. The current combiner calculations for 83MHz IF and 12pF internal capacitors dictates that the inductor be 612nH. Since 612nH is not a standard value, 560nH was used instead.

After placing a 560nH inductor into the circuit, the current combiner's output impedance must be determined. Theoretically, the current combiner circuit at resonance will appear to be a pure "real" impedance; however, this is not an ideal case so some reactance will still be present as will be shown later.

First, a determination of a frame of reference for the network analyzer is required in order to begin the match to 50 Ωs. On the demo board, the inductor where Vcc is connected is temporally "shorted" to GND and a 0.1uF capacitor is placed in series before the output connector (Figure 20). This is done to allow the movement of the reference plane from the SMA connector on the network analyzers test cable to a location on the board where a temporary short circuit has been placed. On the network analyzer the port extensions are adjusted until a dot is achieved at the short-circuit point on the smith chart indicating that a new reference plane has been established at the location of the short on the board. This is the same concept that was used for the LNA but this time the dot will be located on the short-circuit side of the smith chart on the left versus the open-circuit side on the right. After achieving this, the temporary short is removed and the output impedance of the current combiner, at the new reference point, is measured.

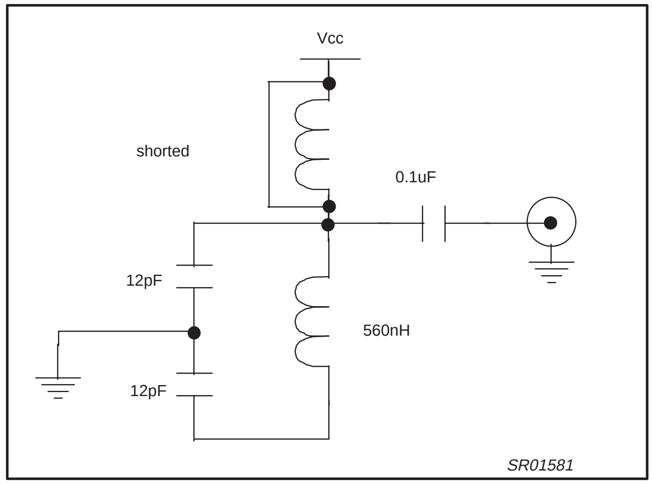

Figure 20 . Board Preparation

The network analyzer should show a reading close to the pure resistance axis of the smith chart at the desired IF frequency, indicating that resonance has been achieved at the proper frequency (Figure 21).

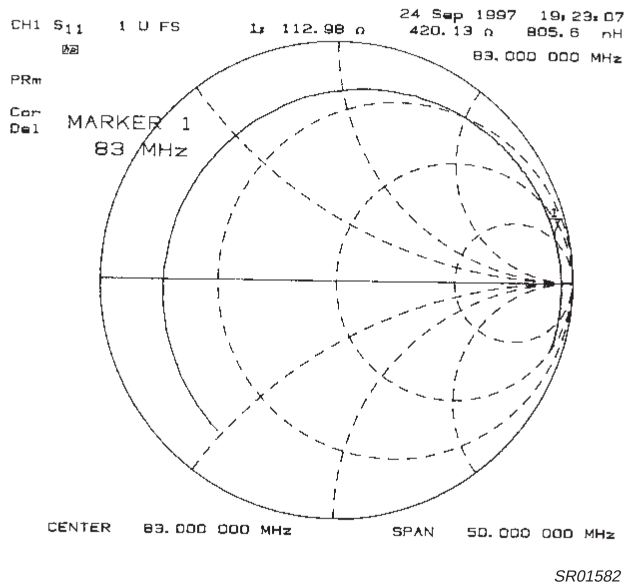

Figure 21 . Measured Output Impedance of the Current Combiner Circuit Shown in Figure 20

Note the residual reactance still present at the desired 83 MHz IF frequency.

To determine the output impedance of the current combiner circuit, the following steps were performed on an HP8753D network analyzer. Press the Marker button, then screen key “marker mode menu” , then screen key “smith marker menu”, and finally screen key “R+jX”. Once this is done, place the marker at the desired IF frequency, read the real portion of the impedance. In this case the impedance was approximately 420 Ωs in series with 800nH.

2. Match to 50Ωs

Matching to 50 Ωs is a two step process. (1) Add the pull-up inductor to rotate the current combiner impedance to the 50 Ω circle, and (2) add series capacitance to rotate the impedance down to the 50 Ω point located at the center of the smith chart

The same network analyzer calibration that was used for the current combiner output impedance measurement can again be used here. In this case make sure that the port extensions are turned “off” because the interface where the 50 Ω match is desired is now at the end of the network analyzer’s test port SMA connector. For the SA621/611 mixer output, a 560nH inductor is used to rotate the current combiner’s output impedance up near the 50Ω circle on the smith chart at 83MHz (Figure 22), followed by a series capacitor to rotate the impedance down to around the 50 Ω point on the chart (Figure 23). It was found that a 6.8pF capacitor was required to achieve the desired match.

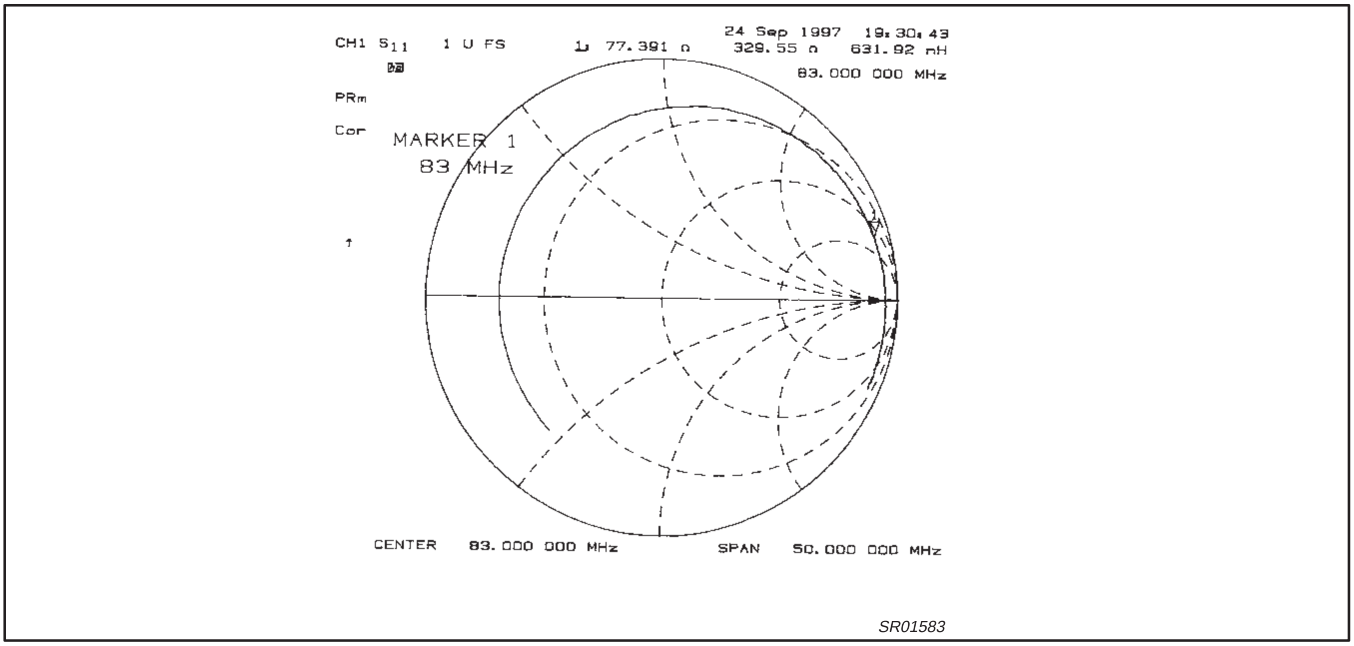

Figure 22 . Measured Output Impedance of the Current Combiner after Rotation on to the 50 Ω Circle at 83 MHz Using a 560nH Inductor to V_{CC}.

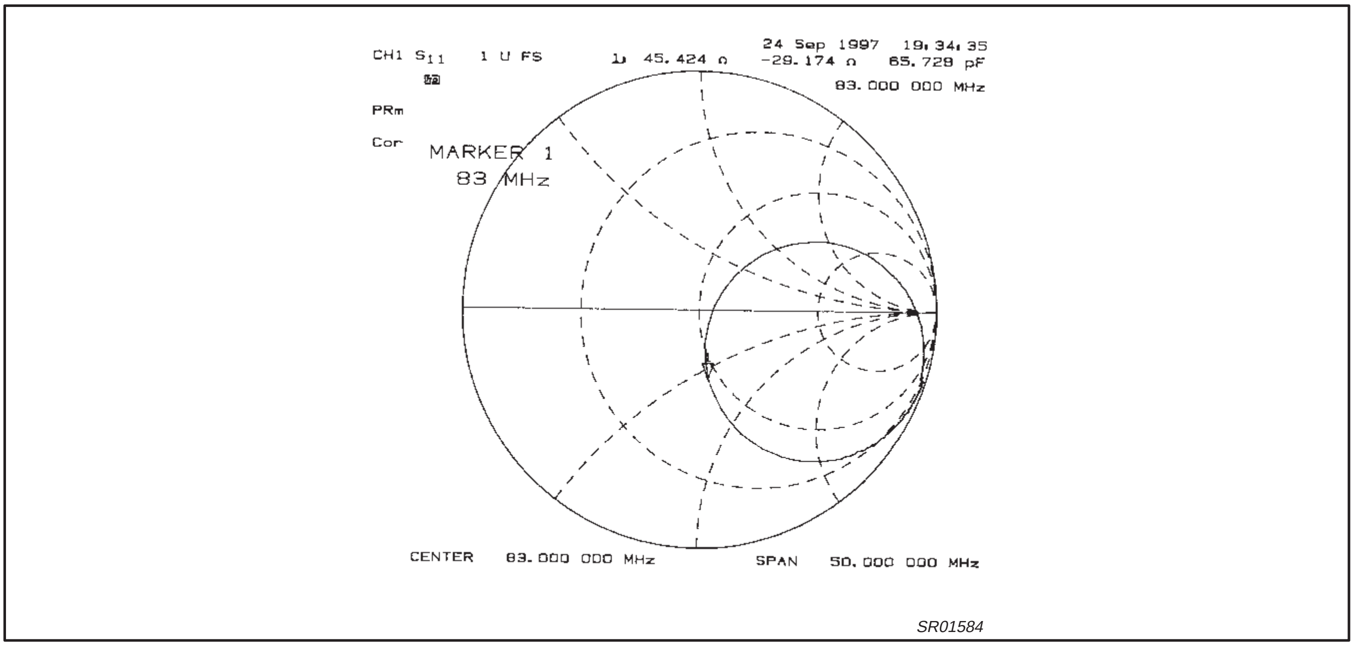

Figure 23 . Measured Output Impedance of the Current Combiner After Rotation To Approximately 50 Ωs at 83 MHz Using a Series 6.8pf Capacitor

When matching to a different frequency the designer should follow the same steps. This method is the quickest way to match to 50 Ωs or to any other impedance for that matter. Note that matching to higher impedances is more difficult but not impossible.

Layout considerations for the mixer

As with the LNA, care must also be taken in the layout of the mixer circuitry. The layout of the mixer input must be done to provide isolation between it and the LNA output in order to prevent signals from bypassing the image reject filter that is usually present between these two pins. If not done properly, degraded sensitivity performance can result from image band interference. For this reason, the LNA output and Mixer input connectors and traces are located on opposite sides of the board to achieve maximum isolation. (Figure 1 in previous LNA section).

In addition to isolation, parasitic capacitance plays a role with the mixer output circuit. A thinner board will have higher parasitic capacitance which will affect the current combiner circuit and impedance matching network. Designers should be careful when using calculated values such as those found in AN1777. These values are approximate and do not take into account parasitic capacitances, so some adjustments will be necessary to obtain the optimal match to a SAW or Crystal IF filter.

SA621 VCO

The SA621 contains a low-noise active circuit meant to be configured with an externally connected passive tank circuit to realize a voltage controlled oscillator (VCO). Feedback is fixed internally and is optimized to operate from approximately 800 to 1000MHz. This stage is capable of very low phase noise performance and is limited only by the external circuitry. A technique using dielectric resonators is presented on the Philips' demonstration (demo) board. Injection amplitude into the mixer LO port depends on oscillator activity, which is principally a function of oscillator circuit Q and impedance loading at pin-4. This pin is also used to sample the LO output for use with an external PLL.

The demonstration board VCO utilizes a physically small, 2mm, $\lambda/4$ dielectric resonator (DRO) coupled to pin-7 of the SA621 through a 1.5pF coupling capacitor, C1 (Figure 24). Decreasing this value improves phase noise, however, the trade-off is that the VCO may have start-up problems and/or the VCO may need a higher control voltage before oscillations can begin.

Choosing the dielectric resonator is critical. The designer should keep in mind that the board layout and external components have a significant effect on what frequency of oscillation is achieved. The resonator's center frequency is 1025MHz on the SA621 application board. Phase noise performance is better than –118 dBc/Hz at 60kHz carrier offset and better than –120 dBc/Hz with a 4mm DRO. As mentioned earlier layout adds parasitic capacitance and one method to decrease this capacitance is to minimize the grounding under the DRO. Doing this improved VCO phase noise and increased the oscillation frequency slightly.

Frequency control is normally obtained through use of an appropriately chosen high-Q, low loss, reverse-biased varactor diode lightly coupled through a capacitor in shunt with the DRO. Even with very low capacity varactors, sufficient coupling required to obtain an acceptable tuning range of 26MHz moves resonance to well below 900MHz. Thus, whatever varactor-controlled tuning arrangement is used, it must also introduce a negative susceptance (inductive) to raise the frequency of oscillations back to the target center frequency of 870MHz. This was accomplished on the demo board by coupling a high-Q shunt inductor, L1 and C2 of Figure 24 across the DRO. The varactor, D1, then moves this frequency down by an external DC control voltage, normally the filtered error voltage from an external PLL.

Lastly, pin 10 of the SA621 is used for bypassing. When a 10pF capacitor is used, the phase noise is improved due to internal filtering of unwanted noise; however, this can only be seen when no VCO circuitry is present (only C1 and the DRO). With the rest of the VCO circuit present, the improvement with the 10pF capacitor is not readily apparent because of the higher phase noise due to the presence of the rest of the VCO circuitry.

The Table 5 below summarizes the effects on the VCO of each component in the VCO circuit.

Table 5. VCO summary

PART	FUNCTION
C1	Coupling capacitor to lightly couple pin 7 to the DRO. Increasing its value will lower the VCO frequency and degrade phase noise. Decreasing its value will improve phase noise but oscillator start-up could be a problem.
C2	Coupling capacitor to the varactor diode (D1). This capacitor will affect the tuning range of the VCO. Increasing its value will provide a wider tuning range while sacrificing phase noise. Decreasing its value will have the opposite effect.
C3	Bypass capacitor of VCO. A value of 10pF will improve phase noise with only C1 and the DRO present; however, with more circuitry, the phase noise is degraded to a point where the improvement is insignificant. 220pF was chosen.
D1	Varactor diode used to allow tuning the VCO to different frequencies. As its voltage is increased, the capacitance of the varactor diode decreases, thus phase noise performance is better at higher control voltages than at lower control voltages.
L1	High Q inductor used to raise the frequency of oscillation. The total susceptance of the varactor diode and coupling cap (C2) is reduced thus improving phase noise while trading-off tuning range.
R1	Resistor used to reduce power supply noise from coupling into the VCO circuitry.
Dielectric Resonator (DRO)	The heart of VCO circuit. The higher Q resonators will deliver the best phase noise performance. Resonators come in 2mm, 3mm, 4mm, and 6mm sizes with the larger resonators providing the highest Qs

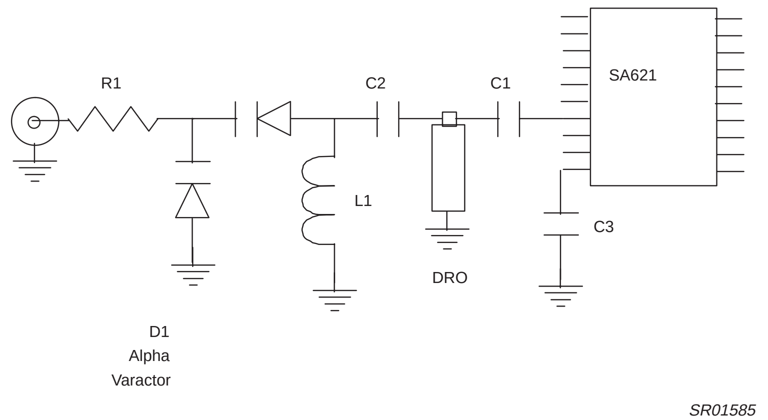

Figure 24 . SA621 VCO

Designers should be aware that this circuit was developed for demonstration purposes only. To limit spurious radiation and influence by stray magnetic fields (typically 50 or 60 Hz), the inductor would certainly require shielding. Shielding will also cause changes in VCO center frequency and tuning range, factors that would have to be accounted for in a real design.

CONCLUSION

This application note has presented a review of RF amplifier matching network design theory from a practical standpoint using S-parameters. For simplicity, unconditionally stable devices were considered, although the methods could easily be extended to conditionally stable devices as well. The Philips SA621/SA611 integrated UHF front-end chips were presented as typical devices. Stability and input gain and noise circles were defined with examples given for each. Note that output gain circles are also obtainable using techniques similar to those employed in finding the input gain circles, but this topic was not treated, since they are usually not required. The utility of input gain circles is obvious owing to their immediate value in graphical comparison to noise circles, as both types vary with changes in the input reflection coefficient Γ_S. Also, since useful noise performance data is often lacking in published device data sheets, an experimental method was presented that makes obtaining them for design purposes reasonably quick and accurate. Interested designers can easily construct calibration boards for rapid in-house evaluation of candidate devices when data sheet S-parameters or noise data are deemed insufficient for a particular application.

Layout and construction issues were also discussed from the standpoint of exploiting hybrid designs using a mix of transmission line segments and classic low-frequency lumped parameter elements in the form of surface mount inductors, resistors and capacitors. This technique can save considerable board real-estate but suffers from requiring much time to realize a suitable design. Understanding S-parameters, the Smith Chart, and layout issues can greatly decrease development and/or evaluation time without expensive simulation packages.

Also presented in this application note were trade-offs of the mixer and VCO circuits. This portion of the application note was intended to give the designer an intuitive feel of how to best implement the SA621 VCO. In summary, the SA611 and SA621 provide low-cost, high performance Front-End solutions for the growing wireless communications industry.

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NOTES

Definitions

Short-form specification — The data in a short-form specification is extracted from a full data sheet with the same type number and title. For detailed information see the relevant data sheet or data handbook.

Limiting values definition — Limiting values given are in accordance with the Absolute Maximum Rating System (IEC 134). Stress above one or more of the limiting values may cause permanent damage to the device. These are stress ratings only and operation of the device at these or at any other conditions above those given in the Characteristics sections of the specification is not implied. Exposure to limiting values for extended periods may affect device reliability.

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