## *Designing with the HMC454ST89 Amplifier Utilizing Small Signal S-parameters*

### General Description

The HMC454ST89 is a high dynamic range GaAs InGaP Heterojunction Bipolar Transistor (HBT) ½ watt MMIC amplifier operating between 0.4 and 2.5 GHz. Packaged in a low cost industry standard SOT89, the amplifier gain is typically17.8 dB from 0.8 to 1.0 GHz and 12.5 dB from1.8 to 2.2 GHz. Utilizing a minimum number of external components and a single +5Vsupply, the amplifier output IP3 can be optimized to +40dBm at 0.9 GHz or +42 dBm at 2.0GHz. The high output IP3 and PAE makes the HMC454ST89 an ideal driver amplifier for Cellular/PCS/3G, WLL, ISM and Fixed Wireless applications.

### Application Problem

The expanding wireless market is placing increasing demands on semiconductor manufacturers to design smaller, lower cost and versatile parts. Many of these parts are designed for a specific frequency band which limits its' adaptability. In many cases it is desired to purchase a part which, with minor adjustment, can be used in multiple applications.  For example, there are hundreds of matched amplifiers from various manufacturers which cover discrete bands. These amplifiers are convenient since the only design required is the land pattern on the PC board. However, many times a system may require an amplifier in a band  which does not fall into one of the popular wireless bands. When this situation arises, a designer is forced to design an amplifier using a discrete device.  These packaged transistors require both a matching network and a bias network. Most experienced RF designers can easily design a matching network which will give the required response over the band of interest. However, many times the bias network is overlooked in the amplifier design. This oversight normally results in a poorly designed bias network that causes an amplifier to oscillate or to vary in gain over temperature. One solution to this problem is to integrate a bias network with the transistor therefore eliminating the need to design one.  Hittite has recently introduced the HMC454ST89 amplifier which has a specified frequency range from 400MHz to 2.5GHz but exhibits gain above and below this band. The amplifier also contains an integrated current mirror in the bias network which provides stable biasing and temperature performance for the HMC454ST89.

The matching networks shown in the HMC454ST89 data sheet are optimized for maximum OIP3. This matching network differs from its small signal counterpart in that it is normally designed from large signal s-parameters obtained from load pull data. However, acquiring load pull data is a long and tedious endeavor that many times yields inaccurate results. In cases where optimum performance in OIP3 or P1dB performance is not required, a small signal matching network will suffice.  Typically the differences in OIP3 performance will be on the order of 3dB. In this application note, three matching networks will be designed using de-embedded small signal s-parameter data obtained from a standard HMC454ST89 evaluation board. The matched amplifiers will be tested for PSAT, OP1dB, OIP3 and ACPR and the results compared to the large signal matching network.

### Obtaining de-embedded S-parameters from evaluation board S-parameters

The first step in any design is to obtain either an accurate model or set of s-parameters with the reference plane located at the package pins.  S-parameter data is easily obtained from any network analyzer.  Obtaining S-parameters where the reference plane is at the package pins presents more of a challenge.  There are basically three ways that this can be accomplished: 1) Direct measurement using coplanar probes and a Thru, Reflect, Line (TRL) calibration 2) Using the port extension of the VNA,  3) Using the standard Short, Open, Load, Transmission (SOLT) calibration and an accurate fixture model.

The first method, making direct measurements with coplanar probes, requires special equipment such as probes and custom calibration standards for the TRL calibration. If the cost of the probes and the time to develop the custom standards is not a problem, then this method should be used since it will produce the most accurate results. The second method takes advantage of a feature that many of the VNA's have as a function.  The reference ports of the VNA can be mathematically extended from the calibrated plane to the reference plane of the pins.  This method assumes an ideal transmission line between the calibrated reference plane and device.

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The drawback is that it will not account for physical losses and discontinuities in the fixture therefore not making it a practical option for most applications.  Finally, the third method makes use of simulation software, such as Eagleware, to de-embed the fixture.  The VNA is calibrated with the standard SOLT calibration technique which sets the reference plane to the ends of the cable.  Using the standard models in the simulation software, S-parameter files for the input and output of the evaluation board or test fixture are created.  Using these S-parameters and a "negation" block in Genesys[1] the effects of the input and output are removed leaving an S-parameter file with the reference plane at the package pins.  This technique was used to de-embed s-parameters of the HMC454ST89 from the Hittite evaluation board 107753-2

Figure 1 shows an evaluation board with the HMC454ST89 mounted on it.  The biasing is removed and zero ohm resistors are placed to provide a 50 ohm transmission line to the connector.  A wideband bias tee is used on the output to provide +5V bias to the HMC454ST89.
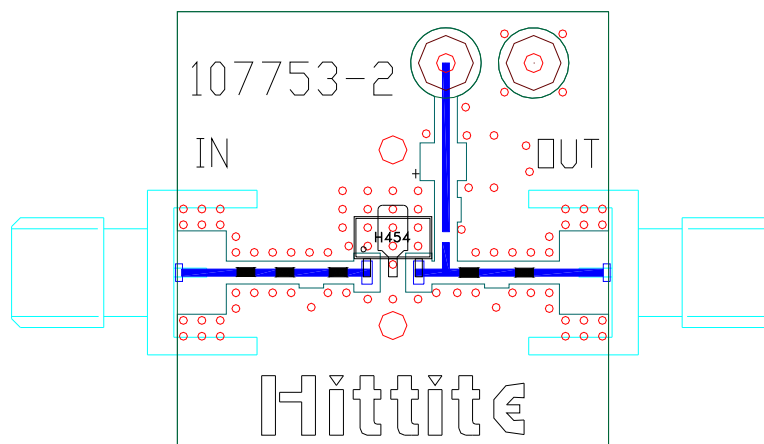


**Figure-1 HMC454ST89 Evaluation Board**

The board consists of three layers, the first layer is used for the transmission media while the remaining two layers are used as a stiffener for the board.  The top layer is 10mil thick and consists of Rogers 4350 with ½ Oz. copper cladding. The second and third layers are comprised of 16 mil thick Rogers 4403 and 30 mil thick Rogers 4350 for a total thickness of 62 mil including metallization. The transmission lines are grounded coplanar with a line width of 16mil and a gap of 13mil, which calculates to a line impedance of ~58Ω. The two SMA connectors are edge mounted and are rated for operation up to 18GHz.

In order to de-embed the S-parameters to the HMC454ST89 pins, the effects of the connector and the board must be removed. This can easily be performed with =GENESYS= if the model for the connector and line are known. The coplanar line is easily modeled using =EMPOWER=, a planar 3D simulator. The more difficult task is to model the connector. Once the line and connector are modeled then the S parameters can be easily extracted.

Figure 2 shows a cross section of a popular "edge mounted" SMA connector. The connector is rated for operation up to 18GHz. The construction of the connector consists of an outer metallic casing with prongs that fit over the board, center pin, dielectric core, and outer pin.

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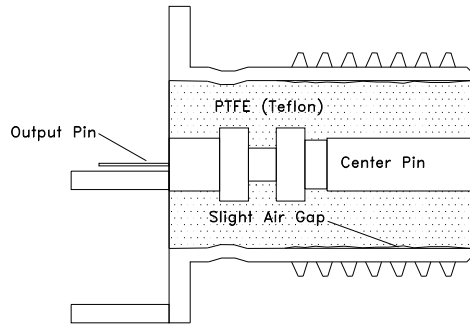


**Figure – 2 Cross Section of Edge Mounted Connector**

The connector is similar to a coaxial cable and can be modeled as such with a few exceptions. Unlike coax, the center pin of the connector is not uniform and consists of several discontinuities due to the changing diameter of the pin. This change in pin diameter represents a change in impedance and at the point of the discontinuity there is a small parasitic capacitance from the pin to ground[2]. The dielectric material surrounding the pin is Teflon.  The dielectric constant of Teflon is typically 2.2 but because of small air gaps that are present at the dielectric boundaries the dielectric constant will be slightly lower.  Also, the transition from the connector to the board introduces set of parasitics. Using Genesys, a model was created and validated.   Figure 3  shows the model created in Genesys.
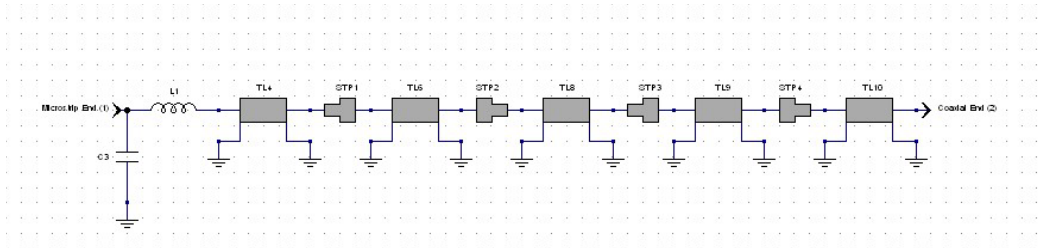


**Figure – 3 Modeled Connector Using Genesys**

The model consists of a series of coaxial line models separated by step discontinuities.  The dimensions of the lines were obtained from measurements taken from the connector.  The series inductor represents the inductance of the pin from the connector transitioning to the microstrip. The shunt capacitance accounts for the fringing capacitance between the connector and ground.

Figure 3 shows a board which has two connectors mounted back to back and separated by a 720 mil coplanar line. The board material consists of the same material and dimensions as the evaluation board 107753-2.

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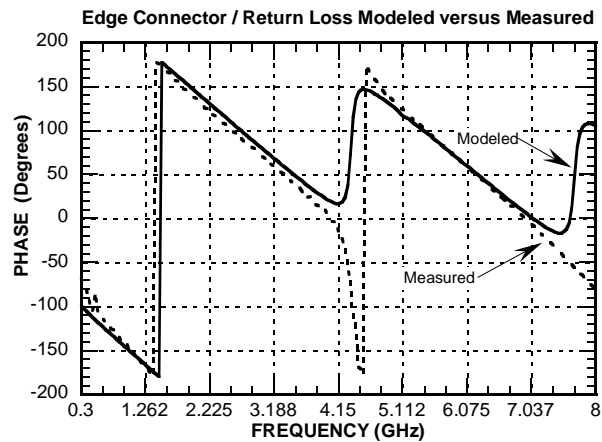## *Utilizing Small Signal S-parameters*
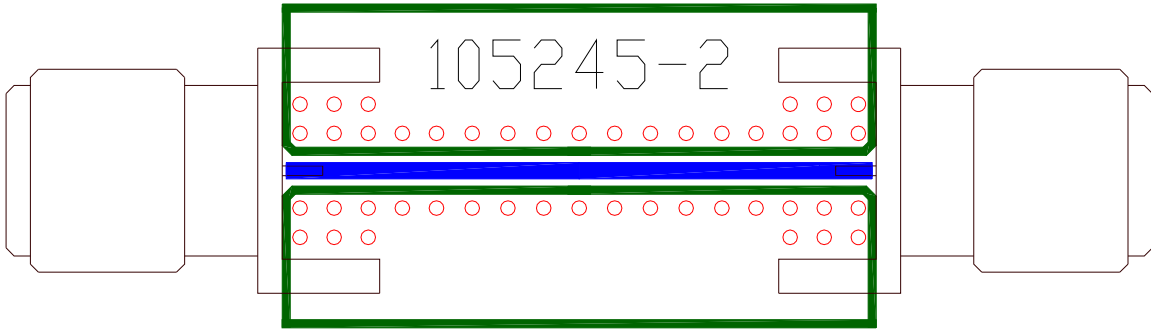


**Figure – 4 Connector Test Board**

S-parameters for the board were measured using a network analyzer and then compared to a model consisting of the connector model separated by a 720 mil coplanar line.  The shunt capacitance and series inductance were optimized to get a "best fit" to the measured data.   Since the measured pin dimensions have an associated error it was also varied slightly (+/- 2mil) to reduce the error between the modeled and measured data.  Figure 5a shows measured S11 versus modeled where figure 5b shows measured versus modeled S21.
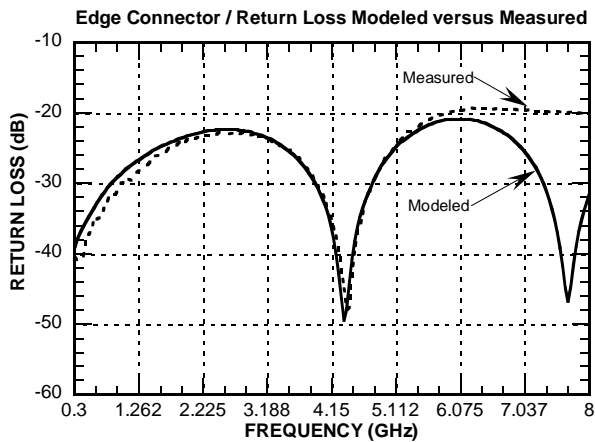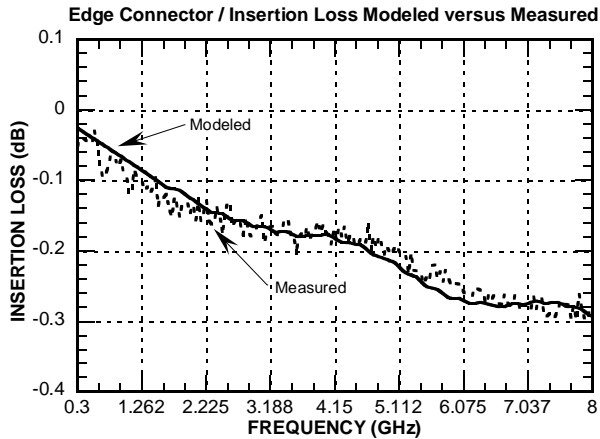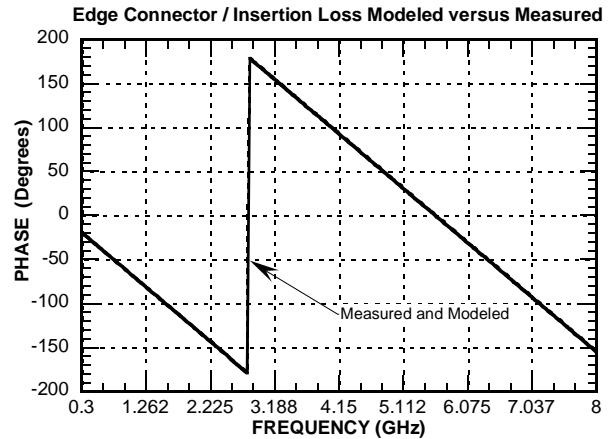


**Figure 5a S11 Magnitude**



**Figure 5b S11 Phase**

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**Edge Connector / Insertion Loss Modeled versus Measured**



**Edge Connector / Insertion Loss Modeled versus Measured**



**Figure 5c S21 Magnitude**                     **Figure 5d S21 Phase**

The data shows excellent agreement between the measured but begins to diverge at the higher frequencies particularly with the return loss.  This divergence is primarily due to higher order effects at the connector and board interface.  Since the HMC454ST89 is only specified for operation up to 2.5GHz the connector model can be used with a high degree of confidence.

Figure 1 shows the evaluation board which was used to measure the S parameters of the HMC454ST89.  These S-parameters contain the effects of the connectors and input and output transmission lines.  If the S parameters are converted to Transmission Parameters (T-parameter) then the S parameters for the HMC454ST89 can easily be de-embedded from the evaluation board ( Equation 1.1). Equation 1.2 shows the relationship between the embedded S parameters and the effects of the connector and transmission line.  The transmission matrices of the entire evaluation board are the product of the transmission matrices of the input network, HMC454ST89, and output network.

$$(0.1) \left[ S_{entire} \right] \Rightarrow \left[ T_{entire} \right]$$

$$(0.2) \left[ T_{entire} \right] = \left[ T_{input} \right] \cdot \left[ T_{HMC454ST89} \right] \cdot \left[ T_{output} \right]$$

$$(0.3) \begin{array}{l} \left[ S_{input} \right] \Rightarrow \left[ T_{input} \right] \\ \left[ S_{output} \right] \Rightarrow \left[ T_{output} \right] \end{array}$$

$$(0.4) \left[ T_{HMC454ST89} \right] = \left[ T_{input} \right]^{-1} \cdot \left[ T_{entire} \right] \cdot \left[ T_{output} \right]^{-1}$$

$$(0.5) \left[ T_{HMC454ST89} \right] \Rightarrow \left[ S_{HMC454ST89} \right]$$

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From equations 1.2 to 1.5, it can be seen that what is required to de-embed the HMC454ST89 from the evaluation board are the S-parameters for the connectors and transmission line. Using a linear simulator, and models for the transmission line and connector, this can be easily achieved. More importantly, Genesys has a function block called "negation", which will automatically perform calculations similar to those in equations 1.1 to 1.5.

### Design of Matching Networks

With a set of de-embedded small signal s parameters it is now possible to design a matching network for the HMC454ST89.  Three matching networks are designed to cover the bands, 450MHz to 500MHz, 824MHz to 960MHz, and 1950MHz to 2150MHz.  Each matching network provides the minimum gain and typical return loss as specified in the Hittite data sheet.

Figure-6a shows the de-embedded S parameters for the HMC454ST89 from 300MHz to 3GHz.  The "matched transducer power gain"[3] of the HMC454ST89 is greater than 20dB at 300MHz and drops to 5dB at 3GHz. The return loss is approximately 4dB across the entire band and the isolation 30dB to 20dB.
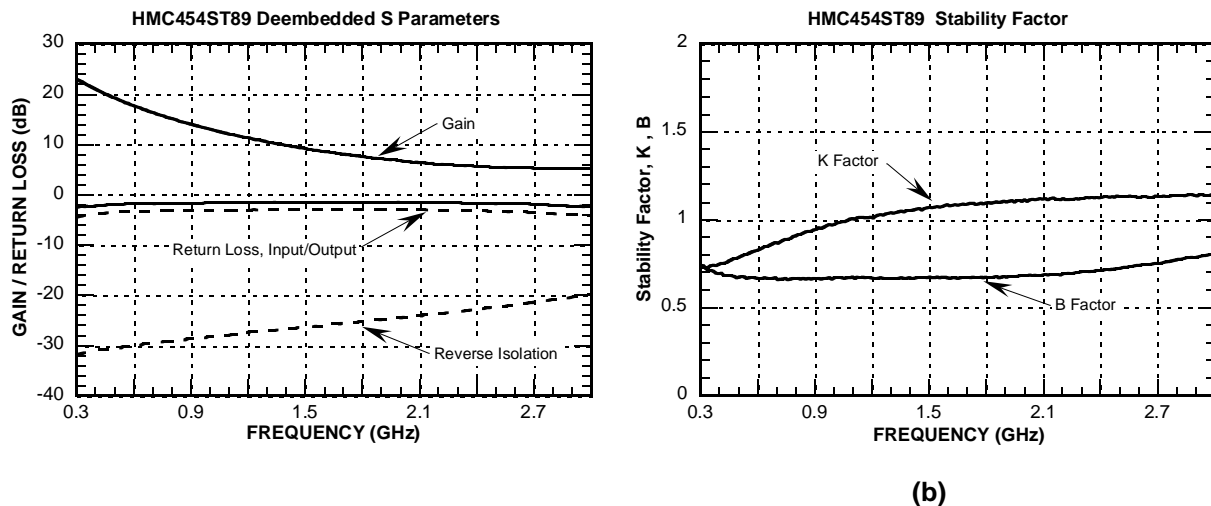


**(b)**

**Figure 6 (a) De-embedded HMC454ST89 amplifier S-parameters (b) Stability factor for the     HMC454ST89**

### Stability

The increasing gain in the lower band is significant since it shows a potential for instability. This is important since the gain is approaching the isolation of the HMC454ST89 and therefore closer to unity loop gain.  As a result, the design process of a matching network must include an analysis to determine the stability of the HMC454ST89. Stability analysis is based on a simple principal: If the magnitudes of the input impedance ( $\Gamma_{in}$ ) and output impedance ($\Gamma_{out}$ ) are greater than 1, then oscillation is possible. The analysis is detailed and beyond the scope of this paper and therefore only the results will be referenced.

In a bilateral device ( $\left| S_{12} \right|$<50dB), $\Gamma_{in}$ and $\Gamma_{out}$ are determined by the following equation:

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$$\Gamma_{in} = S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22}\Gamma_L} < 1$$

(0.6)

$$\Gamma_{out} = S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_S}{1 - S_{22}\Gamma_S} < 1$$

$\Gamma_{IN}$ = Input impedance into the device while terminated with $\Gamma_L$

$\Gamma_{OUT}$ = Output impedance into the device while terminated with $\Gamma_S$

The equation shows the dependency of the input and output impedance on the values of the isolation ($S_{12}$) and gain ($S_{21}$). The lower the isolation and the higher the gain, the more dependent the stability is on the values of $S_{11}$ and $S_{22}$ of the device and $\Gamma_L$ and $\Gamma_S$ of the matching networks.

There are four conditions that have to be met in order to insure the stability of the amplifier. The conditions are: K>1, B>0, S11<1, and S22<1.[4] S11 and S22 are the input and output return loss of the device. K is called the Rollets stability factor and is calculated using equation 1.7.

$$(0.7) \quad K = \frac{1 - |S_{11}|^2 - |S22^2| + |\Delta|^2}{2 \cdot |S_{12}| \cdot |S_{21}|} > 1$$

The equation for the Rollets stability factor is derived by manipulating equation 1.6 under the condition that it must be less than one for stability. The resulting equation shows that K must be greater than one for unconditional stability.

The second condition, B>0, is again derived from equation 1.6. In this case, $\Gamma_{in}$ and $\Gamma_{out}$ are set to give maximum gain, or $\Gamma_{in} = \Gamma_S *$ and $\Gamma_{out} = \Gamma_L *$. Substituting into equation 1.6 and solving the simultaneous equation will result in B as a key component of the equation from which a condition for stability can be determined. Equation 1.8 shows the equation for B.

$$(0.8) \quad B = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 > 0$$

One interesting parameter which is contained in equation 1.8 is the parameter $|\Delta|^2$. This parameter is often used in determining stability where the condition of $|\Delta| < 1$ must be satisfied for unconditional stability. An evaluation of equation 1.8 shows that $|\Delta| < 1$ must be satisfied in order for the condition B>0.

An examination of figure 6b shows that B>0 across the entire band and K<1 below 1.2GHz. Because K is not greater than 1 across the entire band, the part is only conditionally stable. A conditionally stable part can be designed for stable operation if the proper impedances to the matching networks are chosen. The part can also be made unconditionally stable by placing resistance at either the input or output of the amplifier. It will be shown that at the lower frequencies, an input series resistance will be required in order to stabilize the HMC454ST89, while at the higher frequencies (>1.5GHz), a matching network with acceptable impedance can be used.

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**Matching Network Design**

Figure 6a shows the isolation of the HMC454ST89 going from -30dB to -20dB across the frequency band. Because the isolation is not greater than 50dB the amplifier is considered bilateral. The design of matching networks for a bilateral device is complicated due to the input and output loads affecting each other. In order to accommodate the design, available power gain and operating power gain circles will be used. These circles are derived from the available power gain ( $G_A$ ) and the operating power gain ( $G_P$ ) equations which are shown in equations 1.9 and 1.10.

Equation 1.9 calculates the gain contribution from the input when the output is matched, while equation 1.10 calculates the gain contribution of the output when the input is matched. Equation 1.9 is a function of the device S parameters and source impedance. Since the available power gain is independent of the output load it can be used to design the input matching network when the device is bilateral. Equation 1.10 is similar to 1.9 with the exception that the operating power gain is a function of the device S parameters and the load impedance and therefore can be used to design the output matching network.

$$(0.9) \quad G_A(S, \Gamma_S) = \frac{1 - |\Gamma_S|^2}{|1 - S_{11} \cdot \Gamma_S|^2} \cdot |S_{11}|^2 \cdot \frac{1}{1 - |\Gamma_{OUT}|^2} \qquad \Gamma_{OUT} = S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_S}{1 - S_{11} \cdot \Gamma_S}$$

$$(0.10) \quad G_P(S, \Gamma_L) = \frac{1}{1 - |\Gamma_{IN}|^2} \cdot |S_{21}|^2 \cdot \frac{1 - |\Gamma_L|^2}{|1 - S_{22} \cdot \Gamma_L|^2} \qquad \Gamma_{IN} = S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22} \cdot \Gamma_L}$$

Since maximum power is achieved when $\Gamma_{IN} = \Gamma_S^*$ and $\Gamma_{out} = \Gamma_L^*$ it is possible to determine the values $\Gamma_S$ and $\Gamma_L$ which will give a simultaneous conjugate match. Substituting the conditions for maximum power into equation 1.6 and taking its conjugate allows equation 1.6 to be rewritten as:

$$\Gamma_L = \left[ S_{22} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_S}{1 - S_{11} \cdot \Gamma_S} \right]^*$$

$$(0.11)$$

$$\Gamma_S = \left[ S_{11} + \frac{S_{12} \cdot S_{21} \cdot \Gamma_L}{1 - S_{22} \cdot \Gamma_L} \right]^*$$

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Equation 1.11 is two equations with two unknowns and can therefore be solved simultaneously for $\Gamma_S$ and $\Gamma_L$ giving the values for a simultaneous conjugate match.

Figure 7 (a) shows the available gain circle that was generated using =GENSYS=. The frequency for this particular set of circles is ~880MHz. The line running through the circles represent the center of the maximum gain circle which also corresponds to the impedance required for a simultaneous conjugate match ( S11*) to the input.  The corresponding surrounding circles represent the gain that is 1dB lower from the maximum gain.



**( a )**                                                    **(b)**
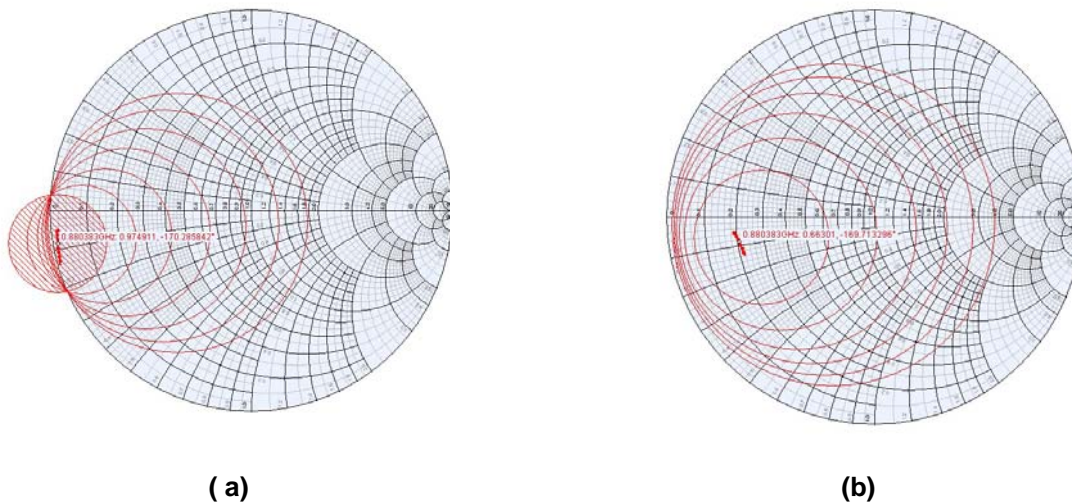
**Figure-7  Available Gain Circles  for 880MHz. (a) No input resistor. (b) with 4.7 ohm input resistor**

The first circle is shaded which indicates that the HMC454ST89 is potentially unstable in that region of the smith chart. This is expected since the K factor is less than 1 at 880MHz.  Even if the HMC454ST89 was stable, the impedance required to achieve a conjugate match would be difficult due to its low impedance.  An alternative would be to match to an impedance associated with one of the gain circles but this would sacrifice both gain and return loss and therefore should be avoided if possible. One technique used to improve the stability and to shift the gain circles further onto the smith chart, is to add a series resistor at the input of the HMC454ST89.  Figure 7 (b) shows the available gain circles when a $4.7\,\Omega$  series resistor is placed at the input of the amplifier. The gain circles are now further into the smith chart and are not shaded which indicates that the region is stable.

Figure 8 (a)  shows the  operating power gain circles for the output with no input stabilization resistor.  Again the circles are located at the edge and off the smith chart with a region of instability.  When the 4.7 ohm resistor is placed at the input of the HMC454ST89 the gain circles move toward the center of the smith chart and the HMC454ST89 becomes  stable as shown in figure 8 ( b ).

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(a)                                    (b)
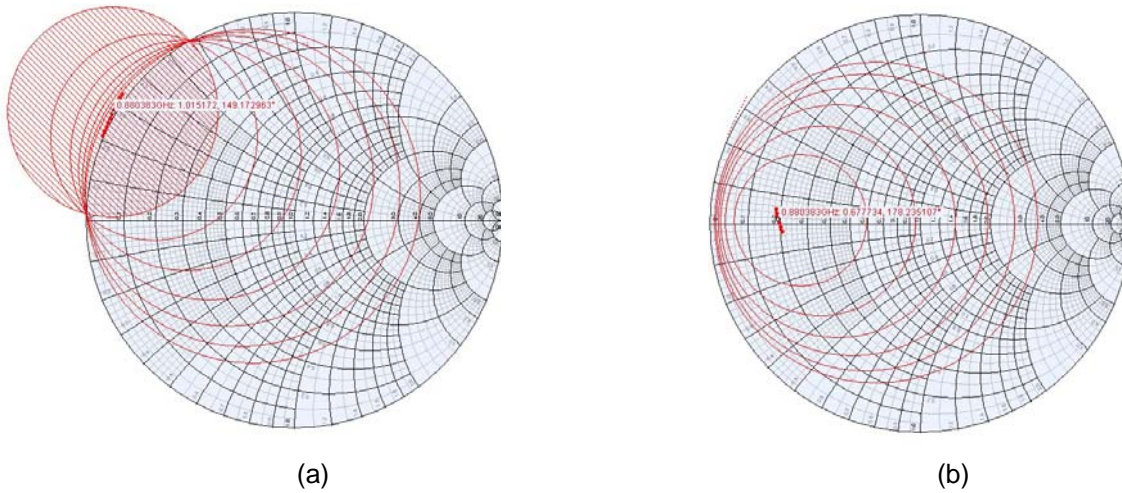
**Figure-8  (a) Operating power gain circles before 4.7ohm input resistor. (b) Operating power gain circle after 4.7ohm input resistor**

To determine stability across the entire band, the K-Factor and B-Factor are recalculated with a 4.7 ohm series resistance in place.  Figure 9 shows the recalculated values.
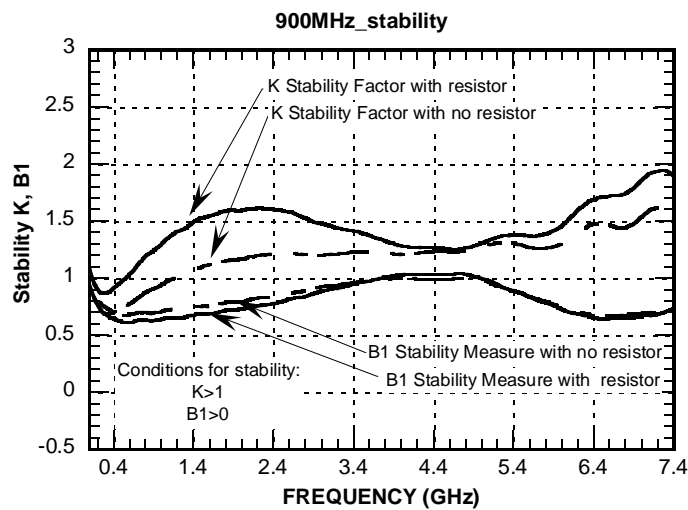


**Figure-9  Stability factor when 4.7 Ohm series resistor is used at input**

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The K factor is now greater than one and the B factor greater than zero at 880MHz making the HMC454ST89 unconditionally stable from ~560MHz and above.  The HMC454ST89 is only conditionally stable from 130MHz to 560MHz therefore care must be taken in the design of the matching network in order to avoid these regions of instability.
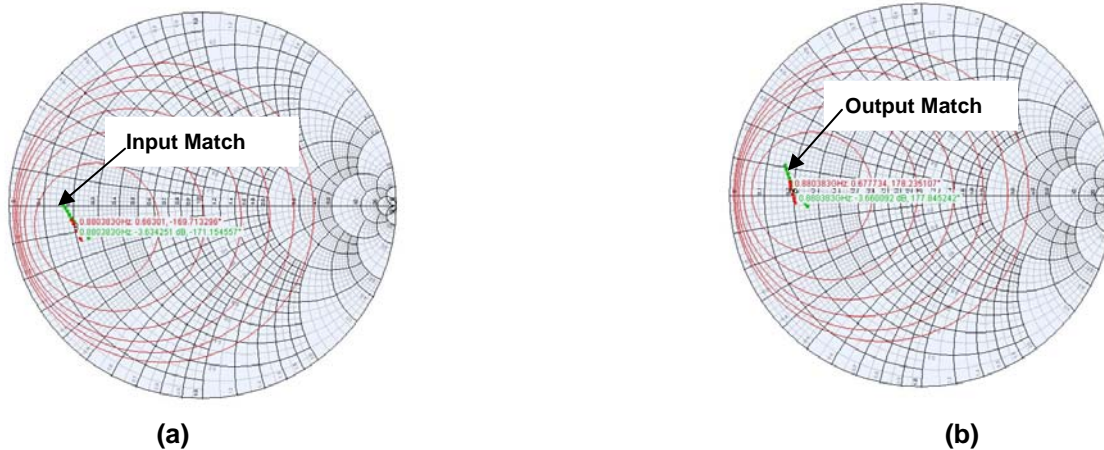
Since the HMC454ST89 is unconditionally stable at 880MHz, the input matching networks can be designed to achieve maximum gain.  However, since there are regions of instability below 480MHz, the matching networks must be evaluated at these lower frequencies to insure that they do not present unstable impedance.
Matching networks for both the input and out were designed using =Genesys= with Modelithic[5] surface mount component models.  The input matching network consists of a series 4.7 ohm resistor ( stability) a 2.2nH series inductor ( Coilcraft) followed by a 6.8pF shunt capacitor ( ATC).  The output matching network has a similar topology with the exception that there is no series resistor and the value of the series inductor is 2.4nH.  Figure 10 shows the configuration of the matching networks as implemented on the evaluation board.
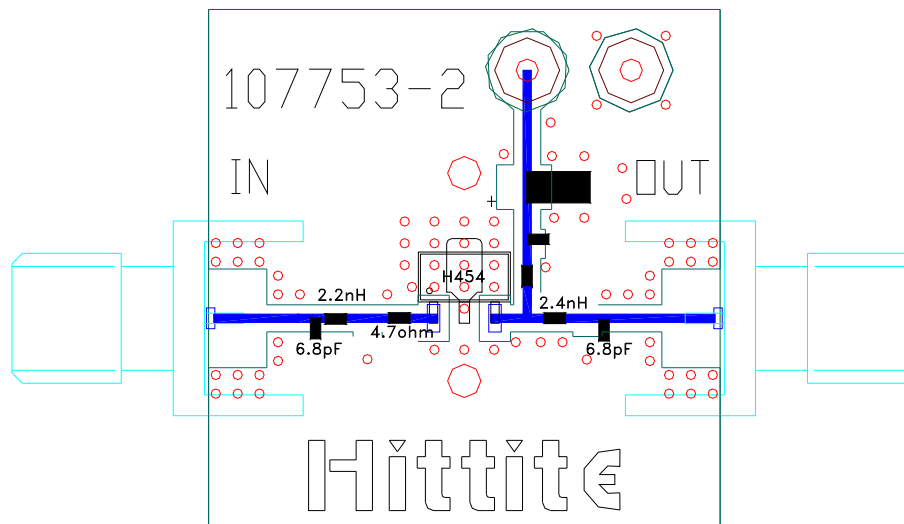


**Figure-10  HMC454ST89 evaluation board with 880MHz matching networks**

 Figure 11a shows the Available Power Gain Circle at 880MHz with the associated impedance to the input matching network. As it turns out, the impedance that the input matching network presents to HMC454ST89 is a conjugate match. This is not always possible since there are limitations in component values offered by the manufacturer amongst other factors . Figure 11b shows the operating power gain circles with the input impedance presented by the output matching network.  Again, the output matching network presents a conjugate impedance to the output of the HMC454ST89.

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(a)                                                                 (b)

**Figure 11 (a) Available power gain circles with input matching impedance**
**(b) Operating power gain circles with output matching impedance**

The HMC454ST89 is only conditionally stable from 130MHz to 560MHz and therefore an examination of the stability circles versus the input/output impedance is required. Figure 12a shows a plot of the input stability circles versus the impedance that is presented to the input of the HMC454ST89. There are three circles plotted at 130MHz, 280MHz and 560MHz. Circles above and below this range are off the smith chart and therefore do not need to be considered. Figure 12a clearly shows that the impedance presented by the input matching network is well away from any region of instability. Figure 12b shows the same circles but for the output of the HMC454ST89. Again, the impedance from the output matching network is located away from any unstable region.



(a)                                                                 (b)
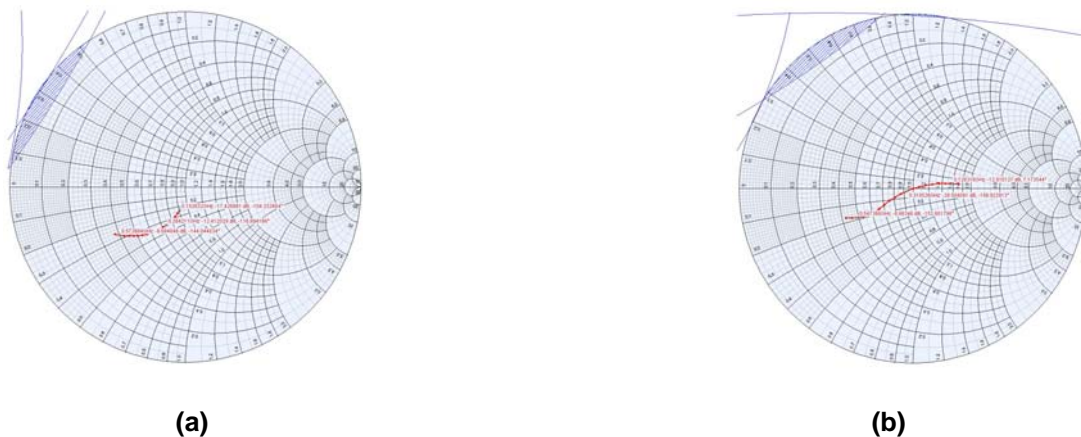
**Figure 12 (a) Input stability circles versus input matching network.**
**(b) Output stability circles versus output matching network.**

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**Measured Results**

The input and output matching networks were implemented onto Hittite evaluation board 107753-2 as shown in figure 10.  Figure 13a shows the measured S21 and S11 versus the simulated and figure 13b measured S22 versus simulated.  Despite the difficulty of modeling the abrupt interface at the input and output pins, the agreement between the measured data and simulated is quite good, suggesting that the de-embedded s-parameters are representative of the HMC454ST89.
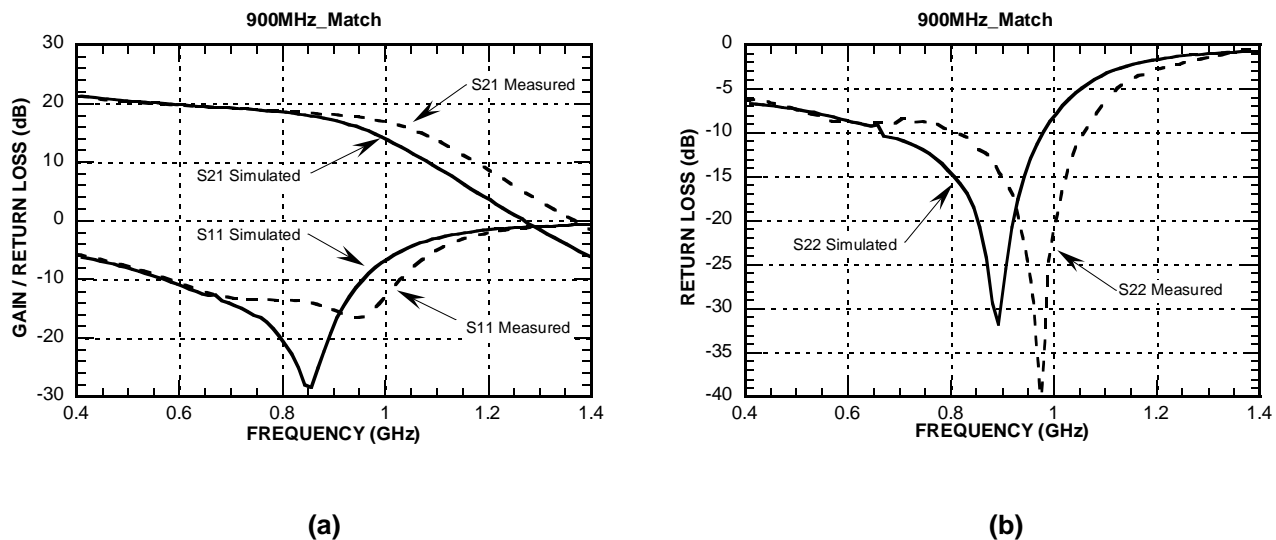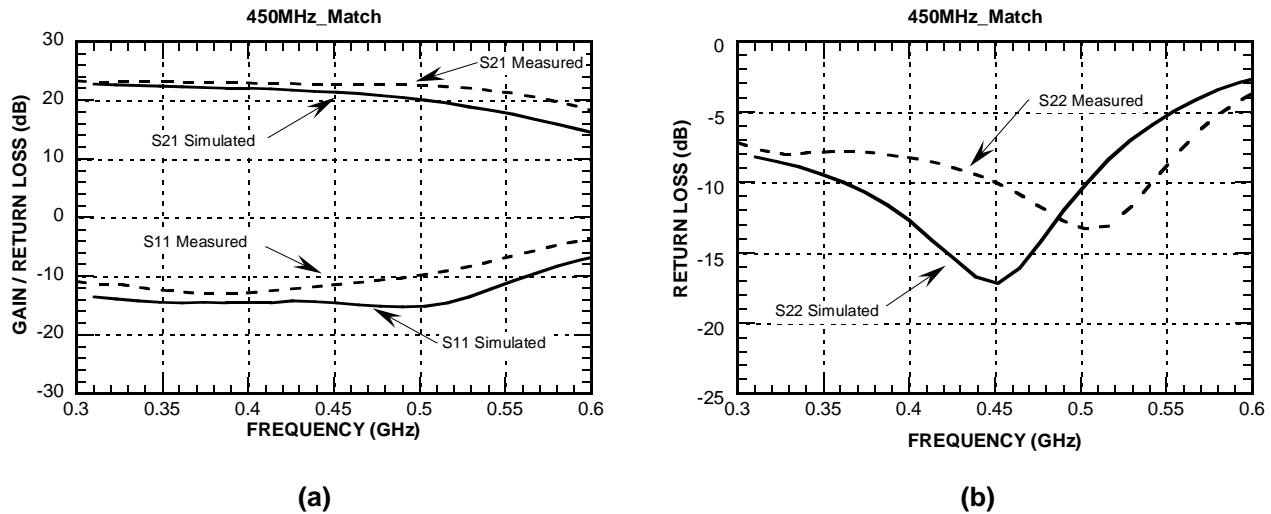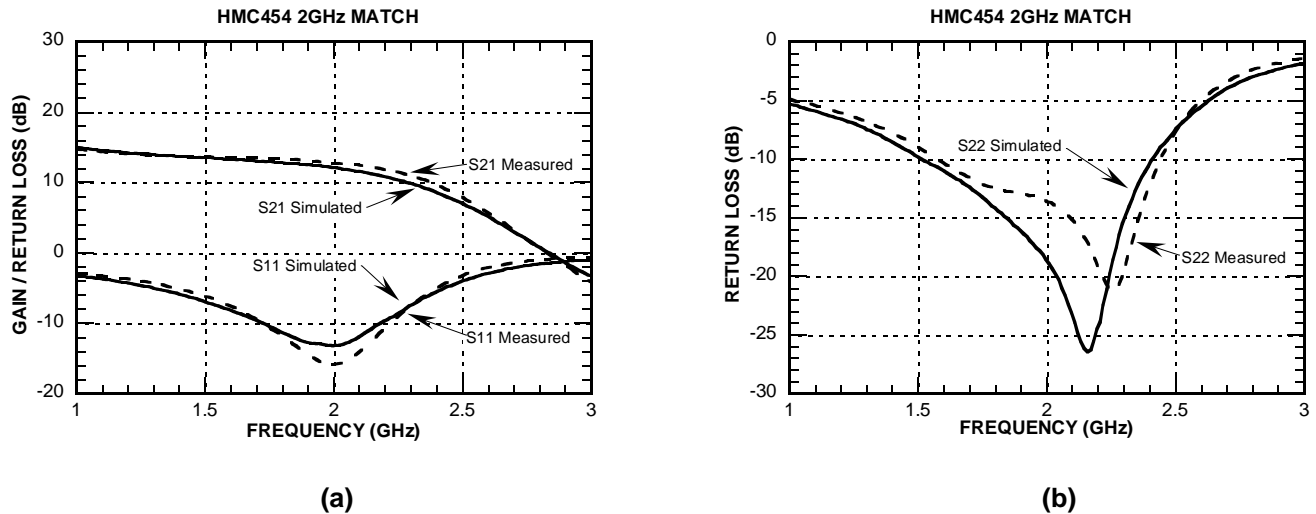
**(a)**                                                            **(b)**

**Figure 13 (a) 880MHz match S21 and S11, measured versus simulated.**
**(b) 880MHz match S22, measured versus simulated.**

Using the same design method, two additional matching networks were designed, built, and tested for the 450MHz CDMA band and the 2.14GHz W-CDMA band.  The measured results versus simulated are shown in figures 14 and 15, respectively.

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**Figure 14 (a) 450MHz match S21 and S11, measured versus simulated.**
**(b) 450MHz match S22, measured versus simulated.**



**Figure 15 (a) 2140MHz match S21 and S11, measured versus simulated.**
**(b) 2140MHz match S22, measured versus simulated.**

The S-parameter data in figures 13,14, and 15 meet or exceed the specified data as stated in the Hittite data sheet for the HMC454ST89 amplifier. These values are compared to the small signal data in Table-1. The primary difference between the matching networks presented on the data sheet and the newly designed matching

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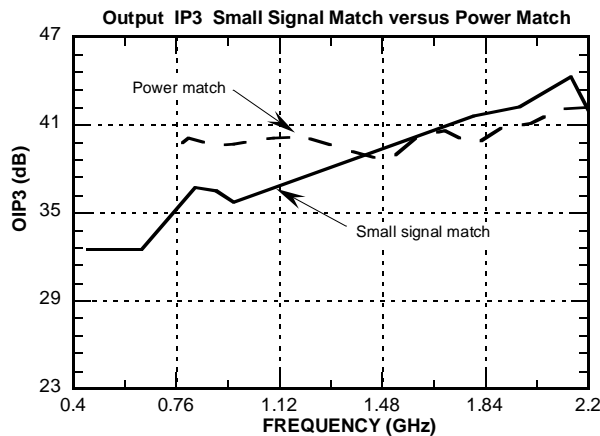## *Utilizing Small Signal S-parameters*

networks is that the data sheet matching networks are designed for large signal operation.   In order to determine the operational differences between the matching networks the amplifier was measured for OP1dB, OIP3, Psat, and ACPR.  The comparison between the small signal match and the large signal match are shown in figures 16 a,b , 17, and18a,b.

Figure 16a compares the large signal match Output 1dB Compression Point (OPIdB), to  the small signal match. The small signal match has the same OP1dB in the 2140MHz band but has a significantly lower OP1dB in the 880MHz band and below.
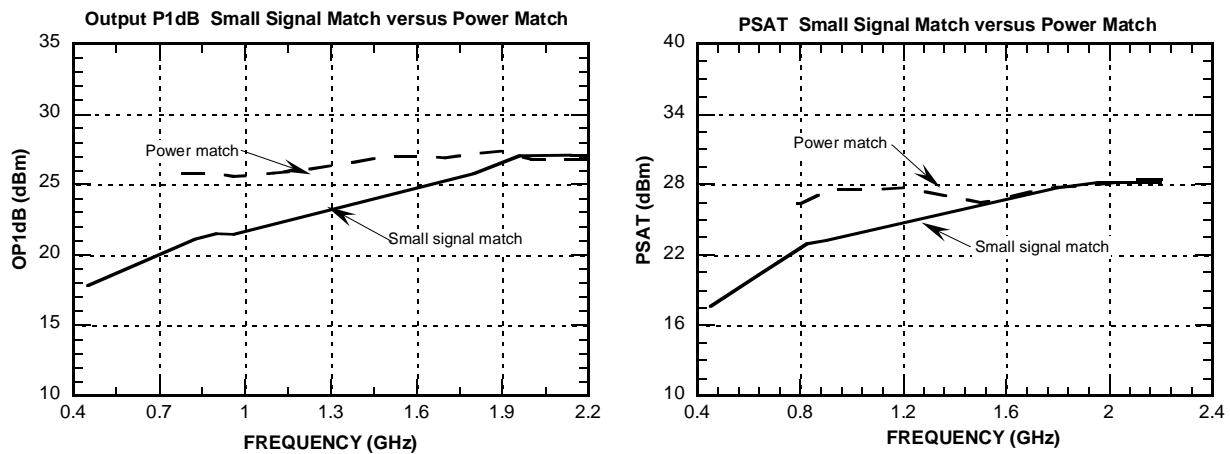
**Figure-16  (a) Large signalOP1dB versus small signal match
(b) Large signal Psat versus small signal match**

Figure 17 is a comparison between the Output 3$^{rd}$ Order Intercept Point (OIP3).  In this case the OIP3 in the 2140MHz band is slightly higher but lower in the 880MHz band.

**Figure-17 Large signal (data sheet) OIP3 versus small signal match**

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Figure 18a is a measurement of the ACPR of the HMC454ST89 at 880MHz. The modulation is CDMA, IS95. The ACPR for the small signal match is significantly higher than the large signal match. This would be expected since both the OP1dB and OIP3 are low at 880MHz. Figure 18b shows the ACPR at 2140MHz. The modulation is W-CDMA. In this case, the small signal match has slightly lower ACPR than the large signal match as the OIP3 data suggests.
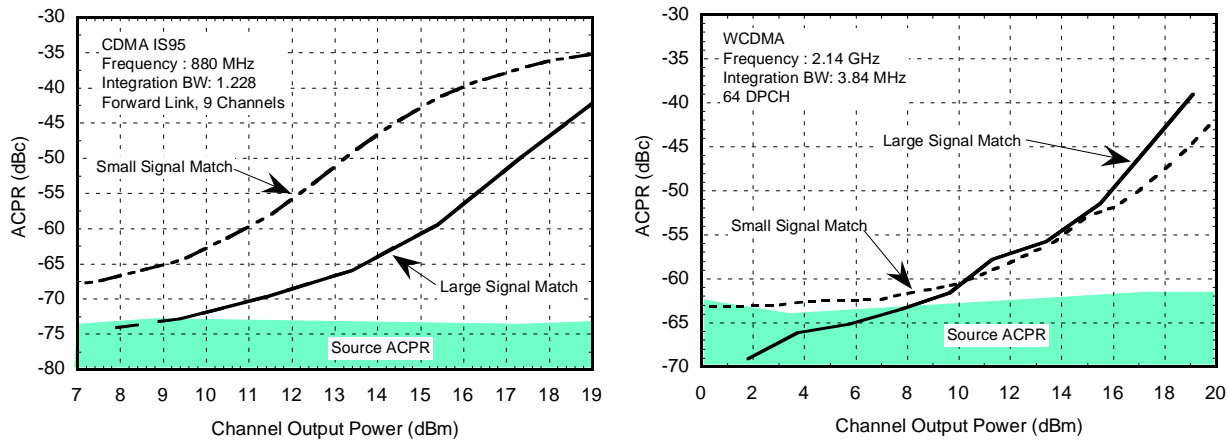


**Figure-18 (a) Large signal ACPR versus small signal match at 880MHz**
**(b) Large signal ACPR versus small signal match at 2140MHz**

Typically, the small signal match has a 2-4 dB lower linearity than the same amplifier matched to large signal[6]. However, this may not always be the case as demonstrated. It is likely that the small signal matching network impedance for the 2140MHz matching network resides on the same power contour as the large signal matching network. To see if this is the case, the large signal load matches were probed and plotted on the smith chart along with the small signal load matches and power contours. The power contours were calculated using a technique that was developed by Steve C. Cripps[7]. This technique uses the operating conditions of the amplifier to roughly determine the location of the power contours. Figure 19 shows that the 2140MHz small signal match resides near the probed large signal 2140MHz match and within the 2140MHz power contour. The 880MHz and 450MHz small signal matches reside away from the respective probed large signal matches and outside the associated power contour. The positions of the small signal loads explain why the optimum large signal performance was obtained for the 2140MHz match but not for the 450MHz and 880MHz matches.

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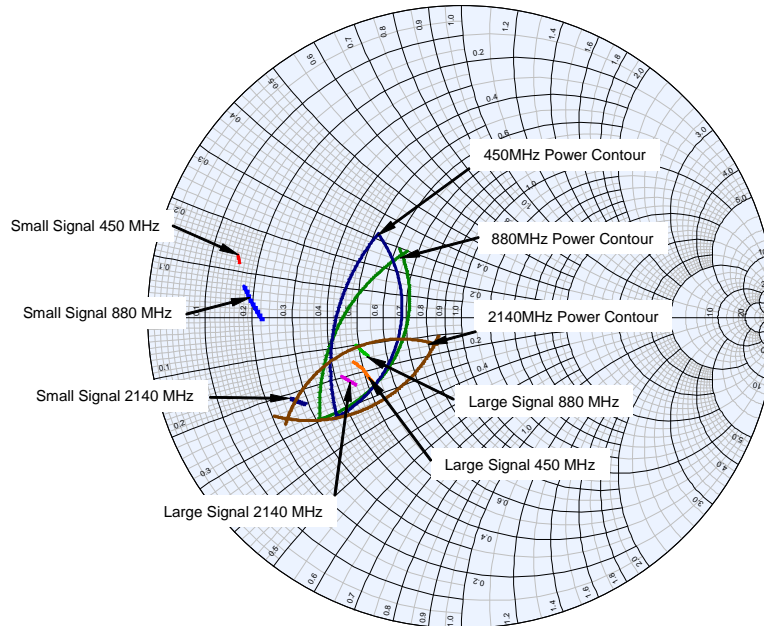


**Figure-19 Large signal match versus small signal match**

## Conclusion

S-parameters for the HMC454ST89 on a standard Hittite evaluation board are measured.  Using =GENESYS= and models for the edge mounted connector and transmission line, the S-parameters are de-embedded to the pins on the HMC454ST89.  These small signal, de-embedded S-parameters are then used to design matching networks at 450MHz, 880MHz, and 2140MHz.  The performances of the small signal matching networks are compared to the datasheet large signal matching networks as shown in Table-1 and the probed large signal matches.

| Parameters | Large Signal/Data Sheet | | Small Signal | | | Units | Comments |
|---|---|---|---|---|---|---|---|
| Frequency | 880 | 2140 | 450 | 880 | 2140 | MHz | |
| Gain | 17.8 | 12.5 | 22.8 | 18.2 | 12.2 | dB | |
| Input Return Loss | 9 | 12 | 11.5 | 14.1 | 12.3 | dB | |
| Output Return Loss | 13 | 19 | 9.9 | 11.1 | 16.8 | dB | |
| Output 1dB Compression (OP1dB) | 24.5 | 27.5 | 17.8 | 21.5 | 27.1 | dBm | |
| Saturated Output Power ( Psat) | 25.5 | 28.5 | 18.6 | 23.2 | 28.2 | dBm | Note 1 |
| Output Third Order Intercept (OIP3) | 40 | 42 | 32 | 36.5 | 44.4 | dBm | Note 2 |
| Adjacent Channel Power ( ACPR) | 47 | 43 | - | 36 | 47 | dBc | Note 3 |

Notes:
1) Measured at 5dB into compression
2) Two tones separated by 1MHz @ P=-10dBm
3) Measured at 18dBm output power, W-CDMA, 64DPCH @ 2140MHz : CDMA IS95, Forward  Link 9 Channels @ 880MHz

**Table-1 Comparison of performance between large signal versus small signal matching**

## *Designing with the HMC454ST89 Amplifier Utilizing Small Signal S-parameters*

The large signal performance for the 2140MHz small signal matching network is comparable to the data sheet ( large signal matching network) performance.  However, the large signal performance for the 880MHz matching network is lower compared to the data sheet as would be expected since the match is not optimally located.

Matching the HMC454ST89 ( or any amplifier) with small signal s parameters will typically yield lower than optimum large signal performance.

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[1] Linear Simulator, Eagleware Corporation, Norwalk, Georgia

[2] Mattei,Young and Jone ##########################

[3] Matched transducer power gain defined as the gain when both input and output network equal the source and load impedance.  In this case 50 ohms.

[4] Guillemo Gonzalez, "Microwave Transistor Amplifiers, Analysis and Design", Prentice Hall, Englewood Cliffs, New Jersey, 1984.  pg 99-113.

[5] Substrate dependent model library for surface mount components, Modelithics Corporation, Tampa, Florida.

[6] Steve C. Cripps, "RF Power Amplifiers for Wireless Communications", Artech House, Bostano, Mass.,1999, pg 19-20.

[7] Cripps, S.," A Method for the Prediction of Load-Pull Power Contours in GaAs MESFETs",Proc. IEEE Intl. Microw. Symp., MTT-S, 1983, pp 221-223.