

## Low voltage front-end circuits: SA601, SA620

AN1777

Author: M. B. Judson

## CONTENTS

<b>I. Introduction</b>	<b>75</b>
<b>II. Key Attributes of the SA601 and SA620</b>	<b>75</b>
<b>III. Power Consumption</b>	<b>75</b>
<b>IV. Low Noise Amplifiers</b>	<b>76</b>
<b>V. SA620 Mixer</b>	<b>76</b>
Open-Collector Output Basics	72
Why $R_C$ Acts Like A Source Resistor	72
Open-collector with $R_{LOAD}$	72
Open-collector with inductor ( $L_C$ )	74
Open-collector with Inductor ( $L_C$ ) and $R_{LOAD}$	77
<b>VI. Flexible Matching Circuit</b>	<b>82</b>
<b>VII. SA601 MIXER</b>	<b>83</b>
<b>VIII. Matching the Open-Collector Differential Output</b>	<b>84</b>
Inductor $L_2$	80
Capacitors $C_5$ and $C_7$	80
Inductor $L_3$	80
Capacitor $C_6$	80
Resistor $R_2$	80
Method of Achieving High Impedance Matching with a Network Analyzer	80
Non-Ideal Current-Combiner Ckt Considerations	81
<b>IX. Matching Examples</b>	<b>87</b>
83MHz, 1 k $\Omega$ Match	82
45MHz, 1 k $\Omega$ Match	84
110.592MHz, 50 $\Omega$ Match	85
SA601 IP <sub>3IN</sub> Considerations	87
Summary of Mixer Open-Collector Output Concepts	87
<b>X. SA601 Mixer Characterization</b>	<b>93</b>
SA601 System SINAD Performance	88
<b>XI. SA620 VCO</b>	<b>93</b>
<b>XII. <math>\mu</math>-Strip Inductor Oscillator Resonant Circuit</b>	<b>94</b>
General Theoretical Background	89
Increasing Loaded-Q	90
High-Q Short Microstrip Inductor	90
UHF VCO Using the SA620 at 900MHz	91
<b>XIII. Application Boards</b>	<b>99</b>
SA601 Applications Board	94
SA601 Application Board Modification For Increasing Mixer Gain	94
SA620 Applications Board	95
<b>XIV. Test and Measurement Tips</b>	<b>102</b>
Noise Figure and Gain	97
1dB Compression Point	98
Input Third-Order Intercept Point	99
Phase Noise	99
<b>XV. Common Questions and Answers</b>	<b>104</b>
<b>XVI. References</b>	<b>104</b>

## I. INTRODUCTION

The objectives of this application note are to highlight key features and distinguish key differences between the SA601 and SA620. The power, gain, noise figure, and third-order intercept point of the LNA and mixer will be characterized. A resonant circuit

implementing a high loaded-Q  $\mu$ -strip inductor and an extensive discussion of open-collector mixer outputs and how to match them will also be presented.

The SA601 and SA620 are products designed for high performance low power RF communication applications from 800 to 1200MHz. These chips offer the system designer an alternative to discrete front-end designs which characteristically introduce a great deal of end-product variation, require external biasing components, and require a substantial amount of LO drive. The SA601 and SA620 contain a low noise amplifier and mixer, offering an increase in manufacturability and a minimum of external biasing components due to integration. The LO drive requirements for active mixers are less stringent than passive mixers, thus minimizing LO isolation problems associated with high LO drive-levels. These chips also contain power down circuitry for turning off all or portions of the chip while not in use. This minimizes the average power consumed by the front-end circuitry. The SA620 features an internal VCO that eliminates additional cost and space needed for an external VCO. The SA601 and SA620 fit within a 20-pin surface mount plastic shrink small outline package (SSOP), thus saving a considerable amount of space.

## II. KEY ATTRIBUTES OF THE SA601 AND SA620

The primary differences between the SA601 and SA620 are the LNA power down capability, the implementation of the mixer output circuitry, and the incorporation of an integrated VCO.

Table 1 below summarizes the attributes of both parts.

Table 1. Showing SA601 and SA602 Attributes

Product	Diff. Mixer Output	LNA Thru Mode	Mixer Power Down	Int. VCO and VCO Pwr Down
SA601	Yes	No	Yes	No
SA620	No	Yes	Yes	Yes

## III. POWER CONSUMPTION

As mentioned above, the average power consumed by the front-end circuitry can be decreased by selectively turning off circuitry that is not in use. The supply current at a given voltage will decrease more than 3mA for each LNA, mixer, or VCO disabled. When the LNA is disabled on the SA620 it is replaced by a 9dB attenuator. This is useful for extending the dynamic range of the receiver when an overload condition exists. Tables 2 and 3 below contain averaged data taken on the SA601 and SA620 while in an application board environment.

Table 2. Showing SA601 Supply Current

V <sub>CC</sub>	I <sub>CC</sub> (mA)	I <sub>CC</sub> (mA) Mixer Disabled
3.0	8.4	4.9
4.0	8.4	4.7
5.0	8.3	4.5

Table 3. Showing SA620 Supply Current

V <sub>CC</sub>	I <sub>CC</sub> (mA)	I <sub>CC</sub> (mA) LNA Disabled	I <sub>CC</sub> (mA) Mixer Disabled	I <sub>CC</sub> (mA) VCO Disabled	I <sub>CC</sub> (mA) Chip Fully Powered Down
3.0	11.4	8.0	8.1	8.2	1.4
4.0	11.6	8.5	8.0	8.2	1.6
5.0	11.7	8.9	8.0	8.2	1.7

# Low voltage front-end circuits: SA601, SA620

AN1777

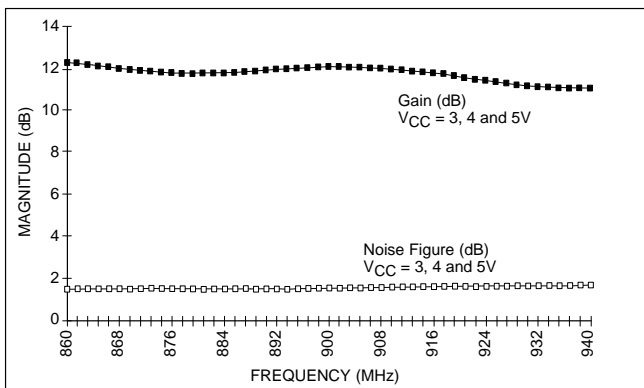


Figure 1. Noise Figure and Gain vs Frequency SA601 LNA

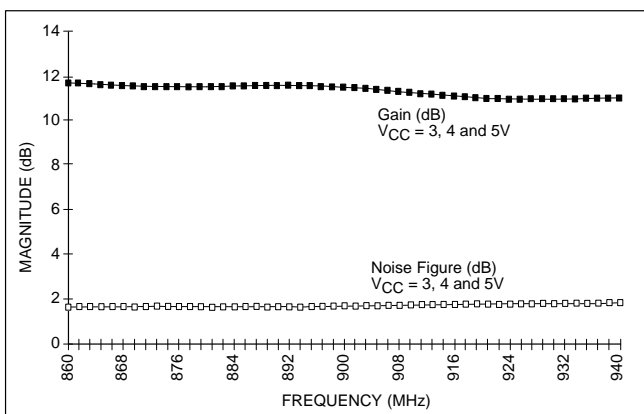


Figure 2. Noise Figure and Gain vs Frequency SA620 LNA

## IV. LOW NOISE AMPLIFIERS

The performance of the SA601 and SA620 low noise amplifiers are virtually identical. You can expect an average gain of approximately  $11.5 \pm 1.5$  dB and a noise figure of approximately  $1.6 \pm 0.3$  dB. The LNA input and output networks are matched for optimum return loss, gain and best noise figure over the 869 - 894 MHz band. They also perform well when utilized in the 902 - 928 MHz band without any additional modification to the LNA matching networks. Figures 1 and 2 show gain and noise figure of a typical SA601 and SA620 LNA in the application board environment. Both the gain and the noise figure remain almost constant as  $V_{CC}$  is adjusted to 3, 4 and 5V. Therefore, only one curve is shown for clarity.

## V. SA620 MIXER

The SA620 mixer is intended for operation with the integrated VCO and employs a single open-collector output structure. The open-collector output structure allows the designer to easily match any high impedance load for maximum power transfer with a minimum of external components. This eliminates the need for elaborate matching networks. The external mixer output circuitry also incorporates a network which distributes the power from the mixer output to two unequal loads. This enables the mixer output to be matched to a high impedance load such as a SAW bandpass filter (typically 1k $\Omega$ ) while simultaneously providing a 50 $\Omega$  test point that can be used for production diagnostics. The mixer output circuitry generates the majority of questions for those utilizing this part in their current applications, so some basic concepts regarding open-collector outputs are presented below, as well as a discussion of the network used to provide the 50 $\Omega$  diagnostic point.

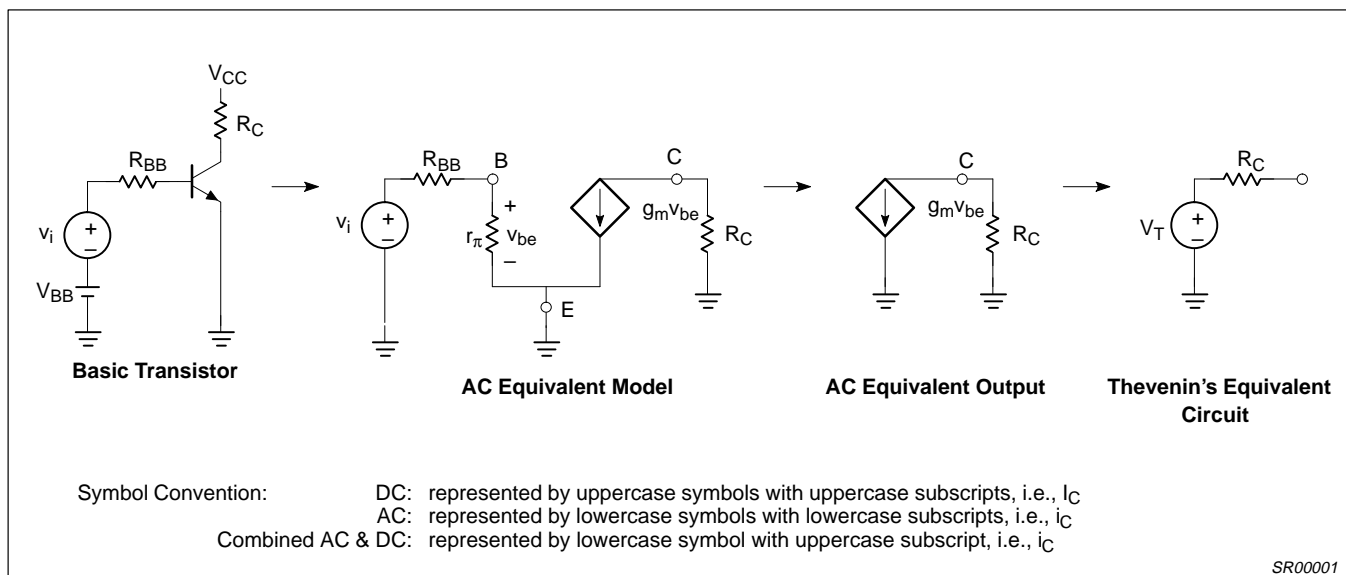


Figure 3.  $R_C$  as Source Resistor

SR00001

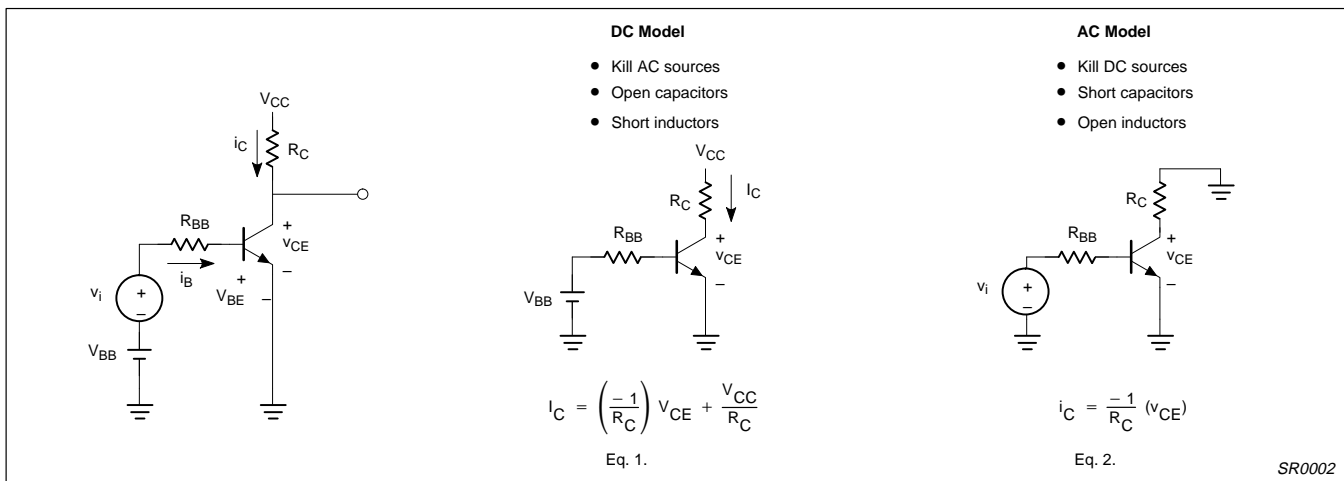


Figure 4. Basic Transistor Analysis

Open-Collector Output Basics [1, 2, 3, 6]

Why R<sub>C</sub> Acts Like A Source Resistor

An open-collector output allows a designer the flexibility to choose the value of the R<sub>C</sub> resistor. Choosing this resistor value not only sets the DC bias point of the device but also defines the source impedance value. Figure 3 shows the AC model of the transistor. Converting the output structure by applying a Norton and Thevenin transformation, one can conclude that R<sub>C</sub> becomes the source resistance. Thus, by choosing R<sub>C</sub> to be equal to the load, maximum power transfer will then occur.

Figure 4 shows an active transistor with a collector and base resistor. From basic transistor theory Equation 1 is generated and has the same form as the general equation for a straight line

$$y = mx + b$$

The slope of the DC load line is generated by the value of the collector resistor (m = -1/R<sub>C</sub>) and is shown in Figure 5. For a given small signal base current, the collector current is shown by the dotted curve.

The intersection of the dotted curve and the DC load line is called the Quiescent point (Q-point) or DC bias determinant. The location of the Q-point is important because it determines where the transistor is operated; in the cutoff, active, or saturation regions. In most cases, the Q-point should be in the active region because this is where the transistor acts like an amplifier.

Figure 6 shows the ac collector-emitter voltage (V<sub>CE</sub>) output swing with respect to an AC collector current (i<sub>c</sub>). Collector current is determined by the AC voltage presented to the input transistor's base (v<sub>i</sub>) because it effects the base current (i<sub>b</sub>) which then effects i<sub>c</sub>. This is how the v<sub>i</sub> is amplified and seen at the output. Recall that this is with no external load (R<sub>LOAD</sub>) present at the collector. Since no load is present, the AC load line has an identical slope as the DC load line as seen in equation 1 and 2 (m = -1/R<sub>C</sub>).

Open-Collector With R<sub>LOAD</sub>

A filter with some known input impedance is a typical load for the output of the transistor. For simplicity, we will assume a resistive

load (R<sub>LOAD</sub>) and neglect any reactance. Since a resistive load is used (see Figure 7), the AC output swing is measured at V<sub>OUT</sub> or V<sub>CE</sub>.

A DC blocking capacitor is used between the R<sub>LOAD</sub> and the V<sub>CE</sub> output to assure that the Q-point is not influenced by R<sub>LOAD</sub>. It is also necessary to avoid passing DC to the load in applications where the load is a SAW filter. However, R<sub>LOAD</sub> will affect the AC load line which is seen in Equation 4 in Figure 7. Notice that the V<sub>CE</sub> voltage swing is reduced and thus, the V<sub>OUT</sub> signal is reduced (see Figure 8).

Since the value of R<sub>C</sub> and R<sub>LOAD</sub> affects the AC load line slope, the value chosen is important. The higher the impedance of R<sub>LOAD</sub> and R<sub>C</sub>, the greater the AC output swing will be at the output, which means more conversion gain in a mixer. This is due to the slope getting flatter, thus allowing for more output swing.

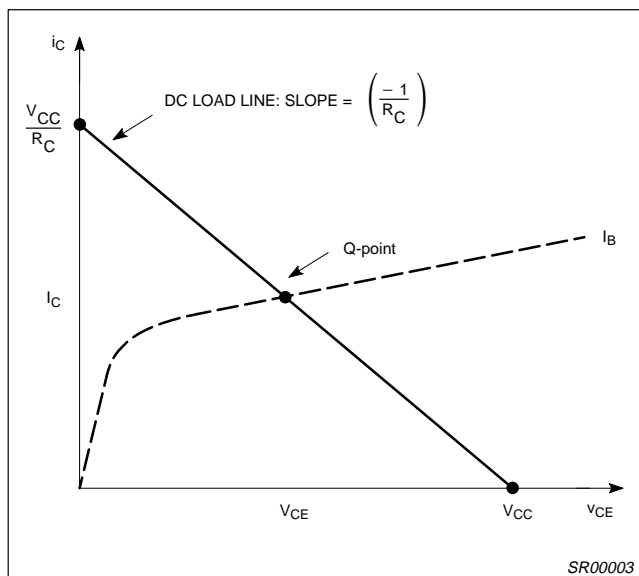
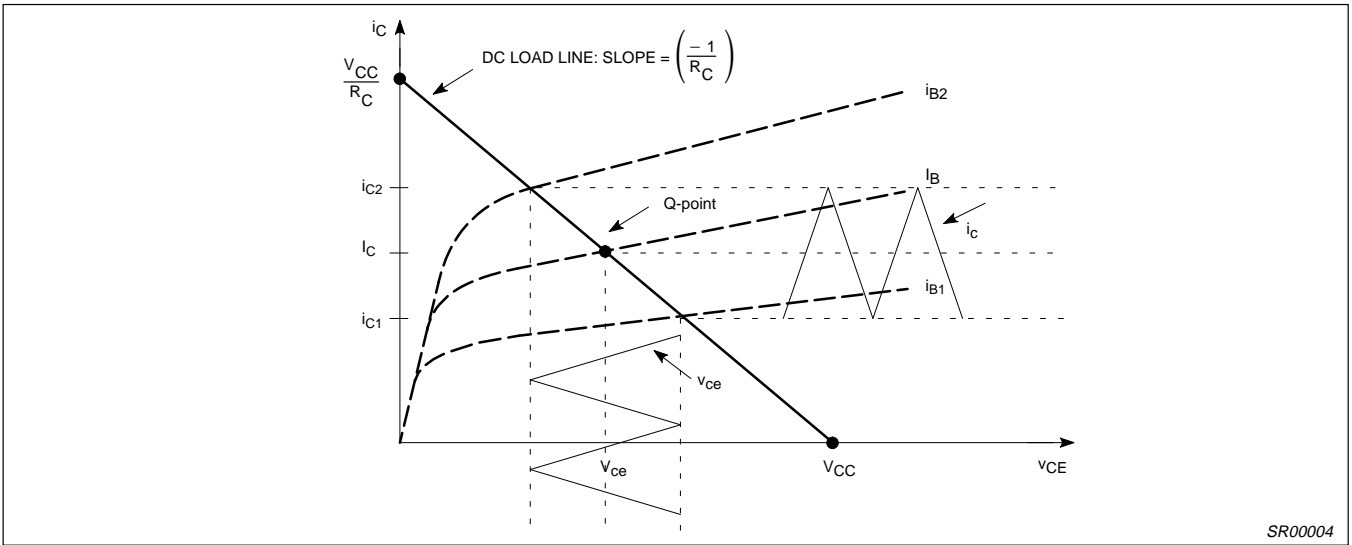
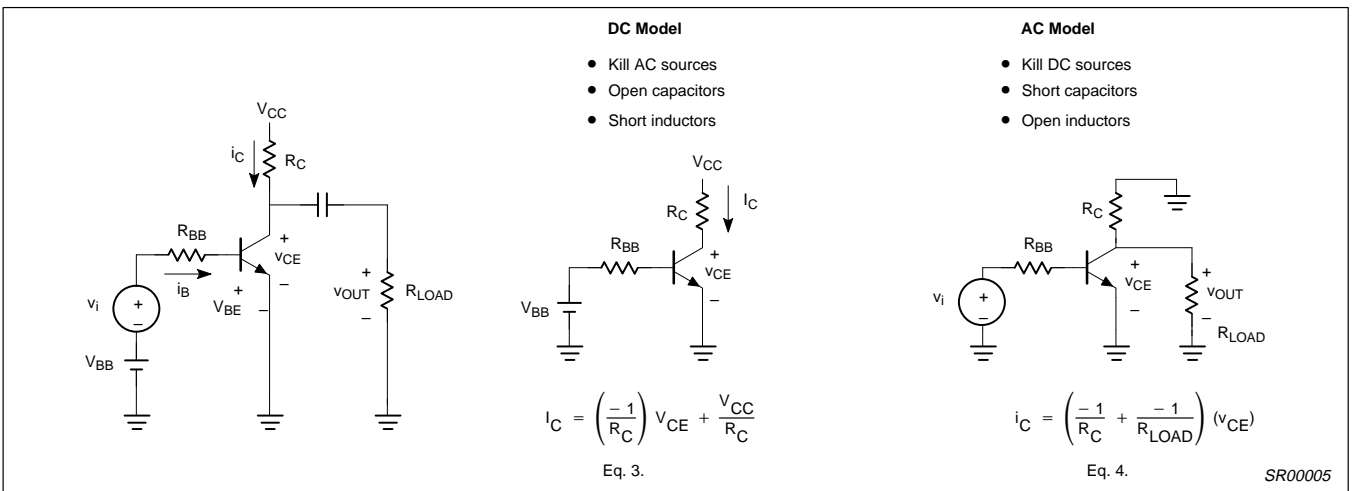


Figure 5. Load Line and Q-Point Graph



SR00004

Figure 6. Graphical Analysis for the Circuitry in Figure 4



SR00005

Figure 7. Basic Transistor Analysis with R<sub>LOAD</sub>

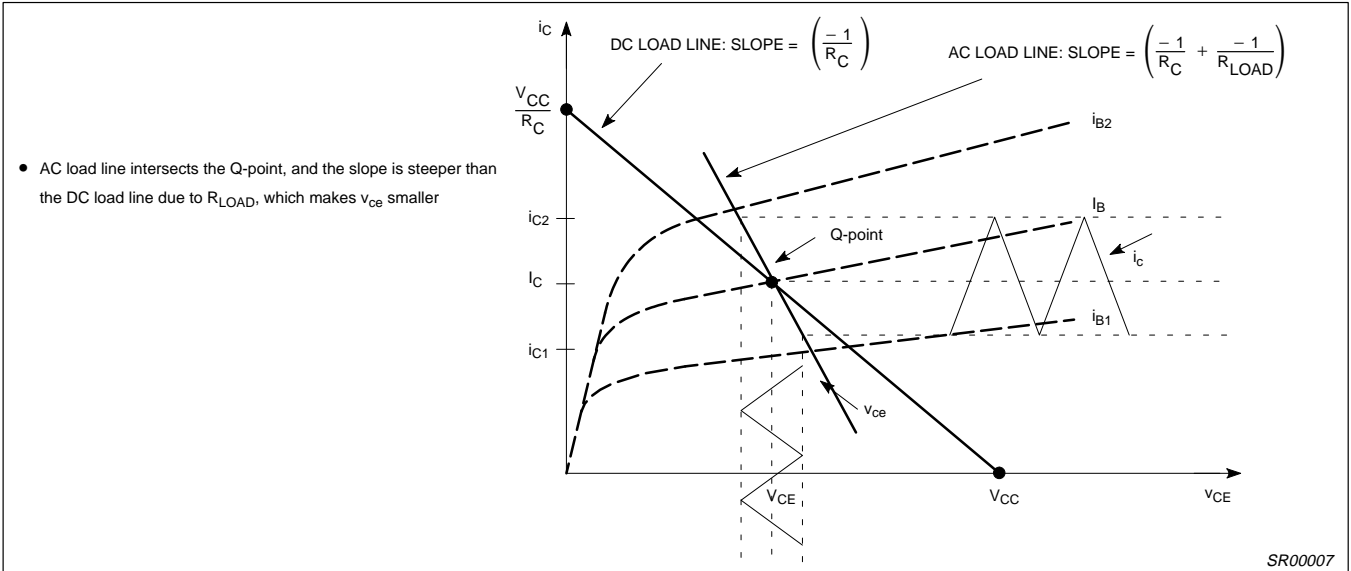


Figure 8. Graphical Analysis for the Circuitry in Figure 7

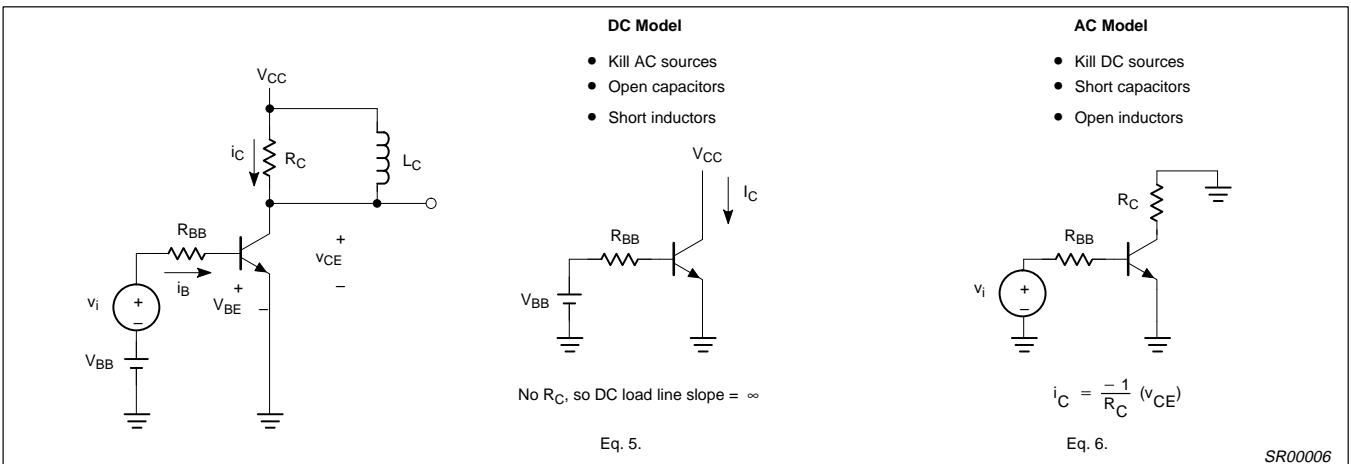


Figure 9. Basic Transistor Analysis with Inductor Added to Collector

**Open-Collector With Inductor ( $L_C$ )**

Adding an inductor in parallel with  $R_C$  can increase the AC output signal  $V_{CE}$ . Figure 9 shows the DC and AC analysis of this circuit configuration. In Equation 5, there is no  $R_C$  influence because the inductor acts like a short in the DC condition. This means the slope of the DC load line is infinite and causes the Q-point to be centered around  $V_{CC}$ , thus moving it to the right of the curve. The AC load line slope is set only by  $R_C$  because no load is present. Notice that it has the same AC load line slope as the first condition in Figure 4, Equation 2.

Referring to Figure 10, one might notice that the base current (AC and DC) curves spread open as  $V_{CE}$  increases. This is caused by a

non-infinite early voltage (see Figure 11), which causes the collector current to be dependent on  $V_{CE}$ . Taking advantage of this non-ideal condition, the peak-to-peak AC output swing  $V_{CE}$ , can thereby be increased by moving the DC Q-point to the right due to the wider spreading between the curves corresponding to different base currents. Figure 12 combines Figures 6 and 10 to show the different AC output signals with different Q-points.

Looking at the AC output level, one might ask how the  $V_{CE}$  peak voltage can exceed the supply voltage  $V_{CC}$ . Recall that the inductor is an energy storing device ( $v=Ldi/dt$ ). Therefore, total instantaneous voltage is  $V_{CC}$  plus the voltage contribution of the inductor.

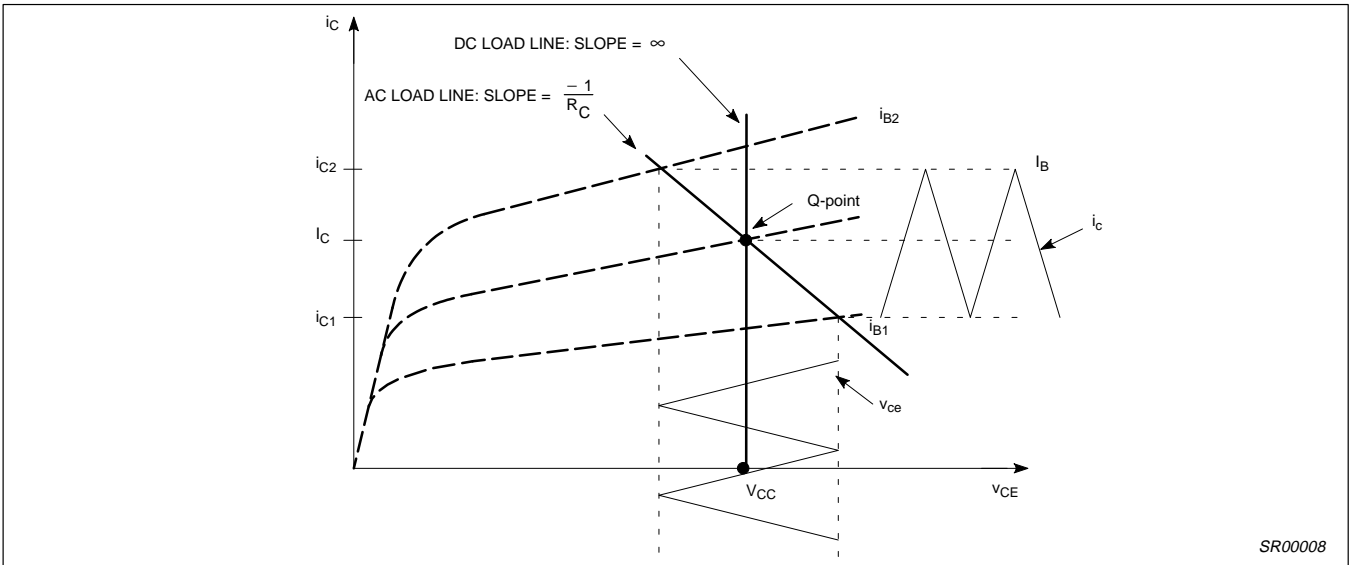


Figure 10. Graphical Analysis for the Circuitry in Figure 9

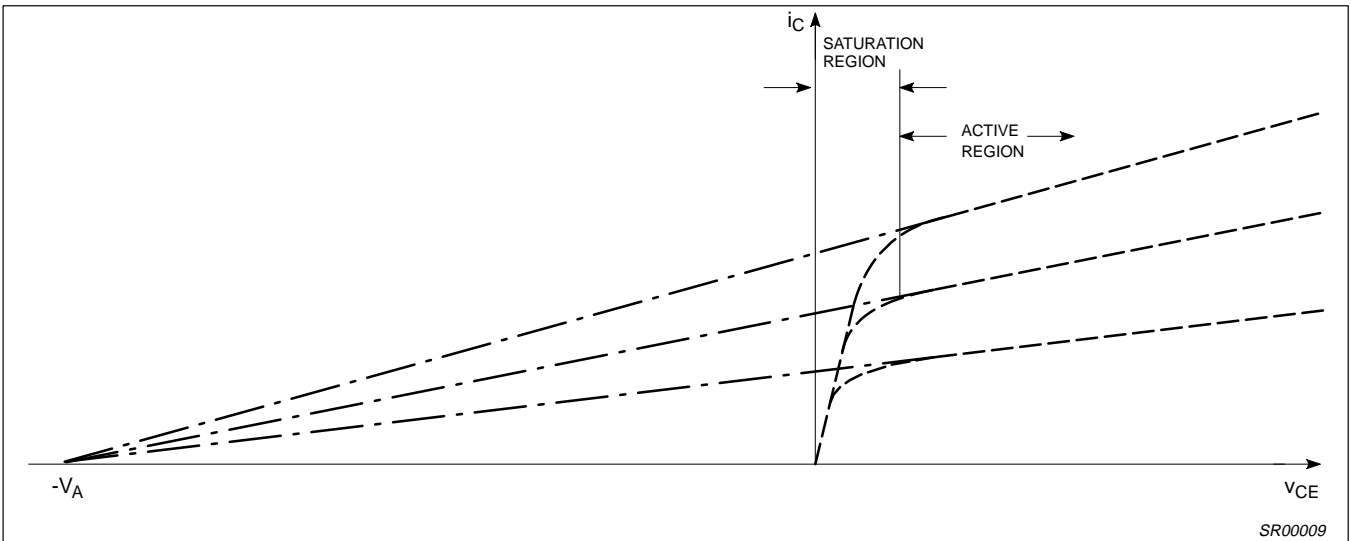
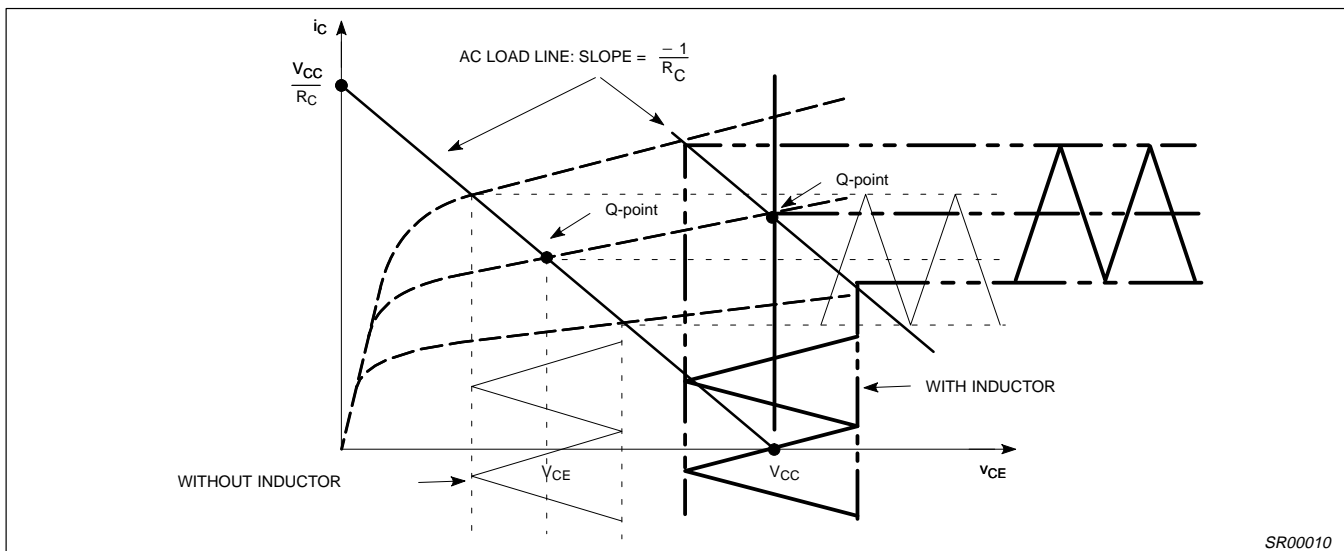
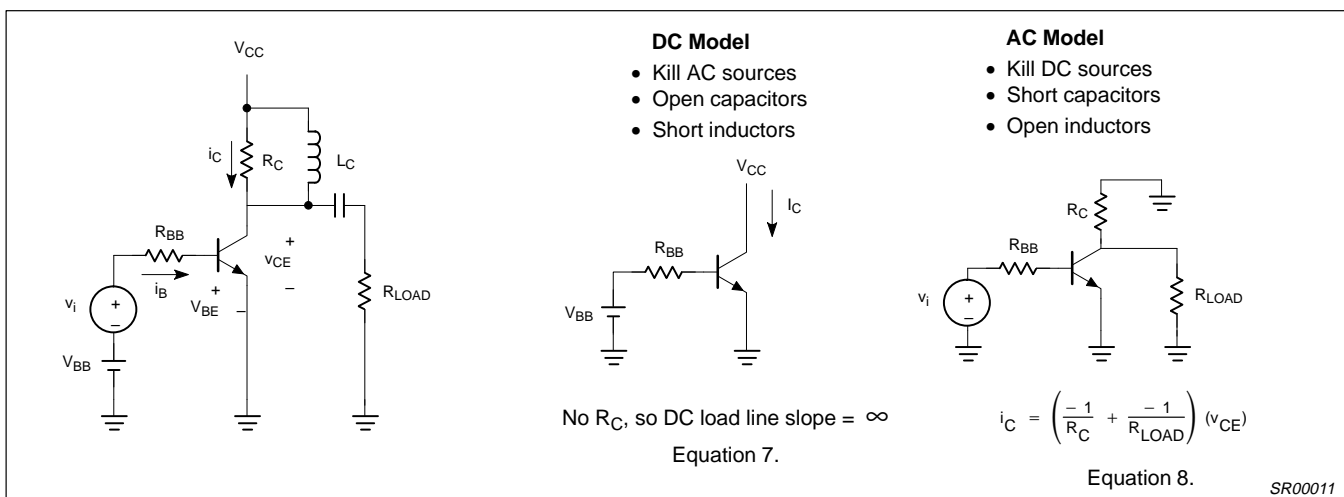


Figure 11. Graphical Representation of Early Voltage Effect



SR00010

Figure 12. Comparison of Open-Collector Circuit With Inductor vs Without Inductor



SR00011

Figure 13. Basic Transistor Analysis with Inductor and RLOAD

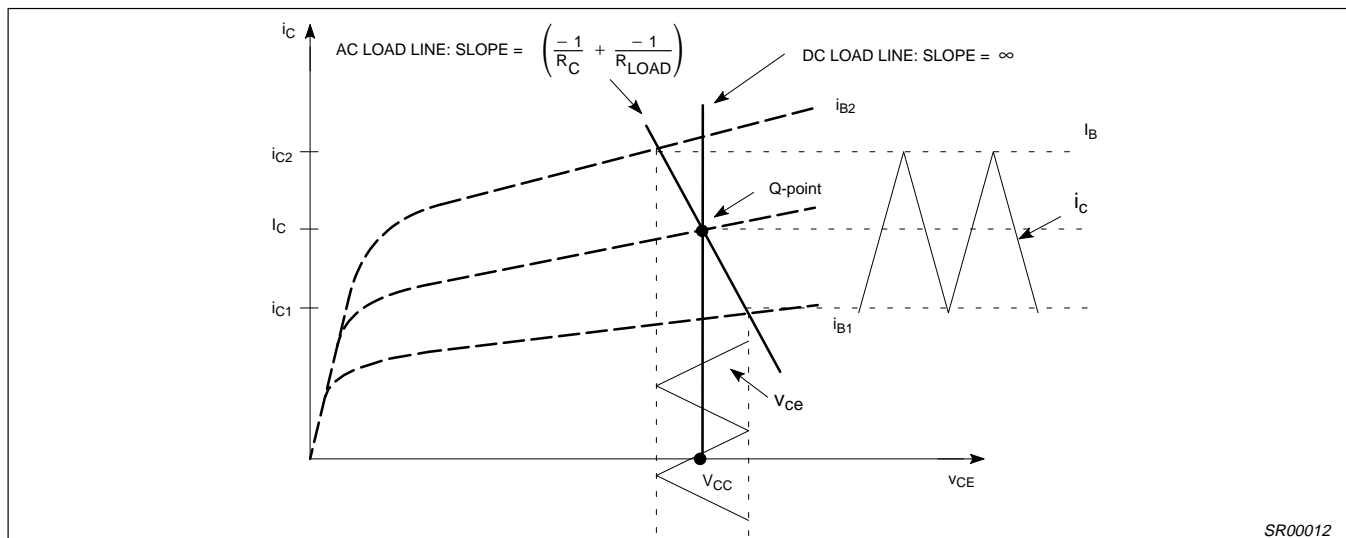


Figure 14. Graphical Analysis for the Circuitry in Figure 13

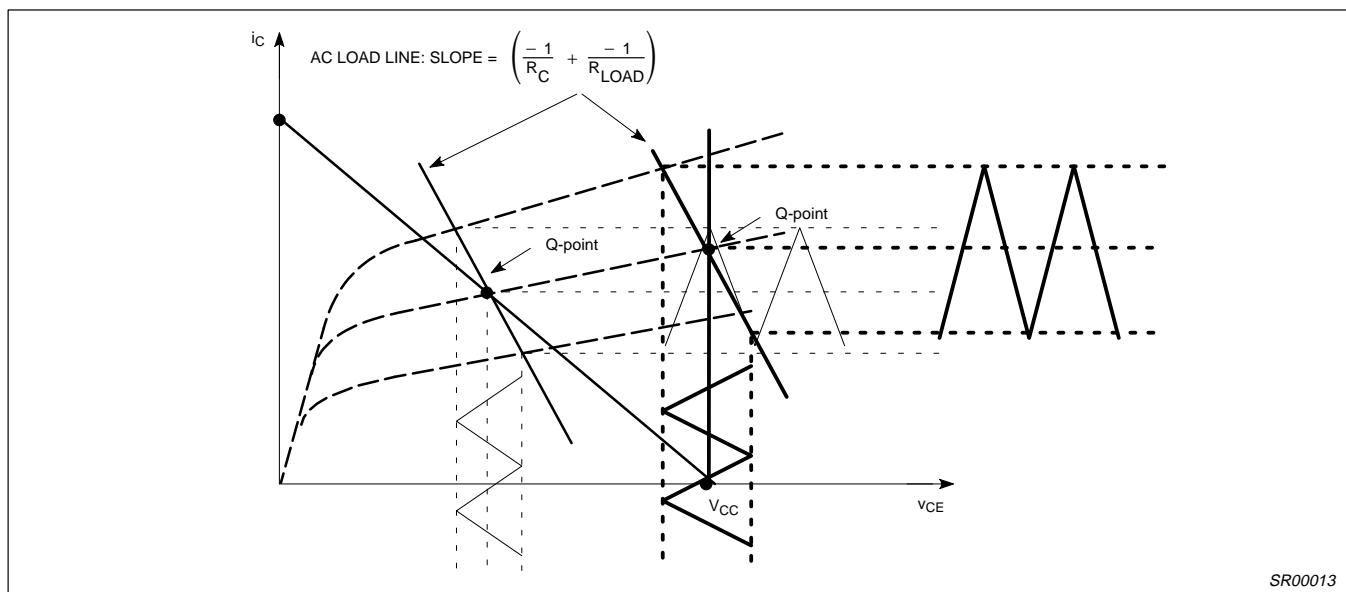


Figure 15. Comparison of Open-Collector  $R_{LOAD}$  Circuit With Inductor vs Without Inductor

**Open-Collector With Inductor ( $L_C$ ) and  $R_{LOAD}$**

Figure 13 shows the DC and AC analysis with the inductor and load resistor. Again, from the DC analysis, the inductor causes  $R_C$  to be non-existent so the DC load line is vertical. In the AC analysis, the AC load line slope is influenced by both  $R_C$  and  $R_{LOAD}$  resistors (see Equation 8). The AC load line slope is the same as the example of the open-collector without the inductor. Figure 14 shows the response for the open-collector with  $L_C$  and  $R_{LOAD}$ . Figure 15 combines Figures 8 and 14 showing the increase in AC output swing.

In conclusion, for the load line,  $R_C$  plays a role in setting up the bias determined Q-point as well as the AC source impedance. However, when an inductor is placed in parallel with  $R_C$ , a different Q-point is set and the AC source impedance is altered. Moving the Q-point takes advantage of the transistor's non-ideal  $i_c$  dependence on  $V_{CE}$  to get more signal output without having to change the base current.

Since  $R_C$  is in parallel with  $R_{LOAD}$  in the AC condition, it influences the AC load line slope.

**VI. FLEXIBLE MATCHING CIRCUIT [6]**

A useful variation of the open collector matching concepts previously outlined provides the capability of delivering equal power to two unequal resistive loads. This allows the power delivered to the load to be measured indirectly at another test point in the circuit where the impedance can be arbitrarily defined. If this impedance is defined to be  $50\Omega$ , a spectrum analyzer can be easily placed directly into the circuit. This is an excellent troubleshooting technique and a valuable option to have available in high production environments.

Figure 16 shows the schematic for this flexible matching circuit. In this circuit,  $C_B$  functions only as a DC blocking capacitor and presents a negligible impedance at the frequency of interest. Recall from the previous open collector matching discussions that, when  $R_C$



# Low voltage front-end circuits: SA601, SA620

AN1777

is placed in parallel with an inductor, it has no effect on the Q-point, but does influence the slope of the AC load line as

$$\text{Slope} = \frac{-1}{R_C} + \frac{-1}{R_{LOAD}}$$

The capacitor  $C_S$  functions not only as a DC blocking capacitor, but is also chosen such that the impedance presented by the combination of  $L$ ,  $R_C$  and  $C_S$  is equal to  $R_{LOAD}$  for optimum power transfer. The analysis is done in the following manner. First, note that inductor  $L$  is connected to  $V_{CC}$  which is an effective AC ground. So,  $L$  can be redrawn to ground. Next,  $R_C$  and  $C_S$  are converted to their parallel equivalent values as shown in Figure 17.

The resulting parallel LCR circuit is shown in Figure 18. At resonance, the parallel  $L$ ,  $C_P$  combination will be an effective open

circuit leaving only  $R_P$ .  $R_P$  is then simply chosen to be equal to  $R_{LOAD}$ .

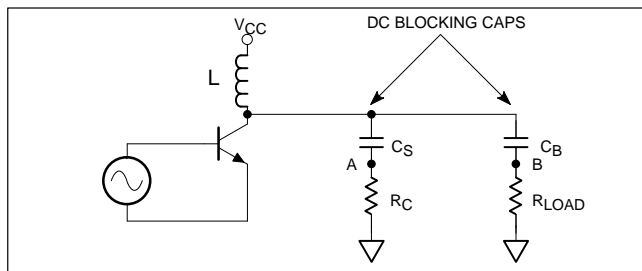


Figure 16. Flexible Matching Circuit

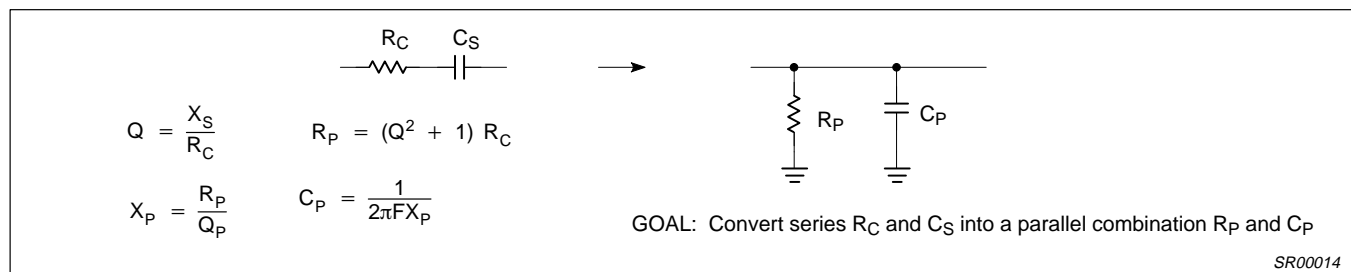


Figure 17. Converting from Series to Parallel Configuration

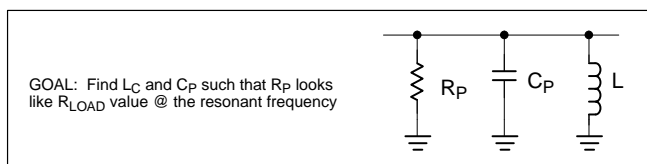


Figure 18. Converting from Series to Parallel Configuration

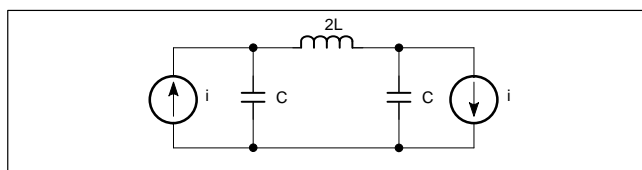


Figure 21. Ideal AC Equivalent Circuit

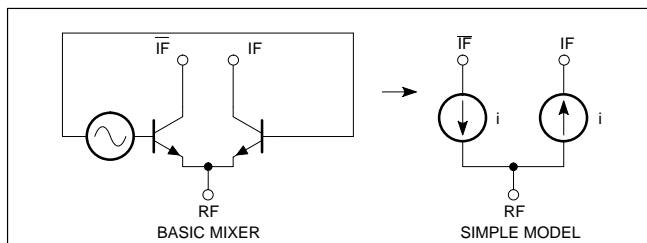


Figure 19. Circuit Model of Differential Open-Collector Output

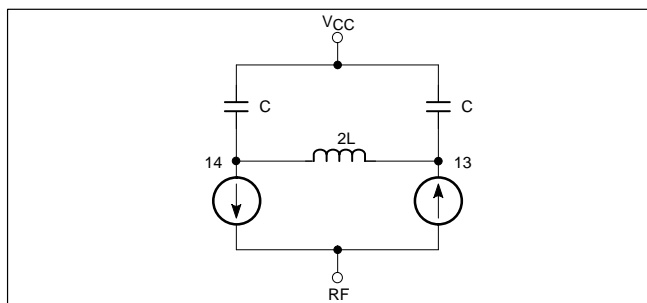


Figure 20. External current-combiner Circuit

## VII. SA601 MIXER [6]

The SA601 mixer is intended for operation with an external VCO and employs a differential output structure. The current wireless markets demand front-end solutions with low power, and high gain. The differential output offers higher gain than the conventional single-ended open-collector output without an increase in supply current. An equivalent model can be seen in Figure 19. A characteristic of the differential output is that the two output currents are 180° out of phase. This is why the mixer output is labeled IF and IF.

The current-combiner circuit shown in Figure 20 consists of two capacitors and one inductor. The purpose of the current-combiner is to combine the currents such that they are in phase with one another. By aligning the currents in phase with one another, the output will have a larger AC output swing due to the increased signal current.

Figure 21 shows the ideal AC equivalent model. In the ideal case, it is assumed that all component Q's are high enough to be neglected and the output impedances of the current sources are also high enough to be neglected. By source transformation, the parallel capacitor and current source can be converted to a voltage source and a source capacitor. The inductor, 2L, can be split into two inductors where L becomes the new value (See Step 2 in Figure 22). Since two inductors of equal value in series will be twice that

# Low voltage front-end circuits: SA601, SA620

AN1777

value, we can split the one inductor into two series inductors each equal to L.

The series capacitor and inductor (L) at resonance will act like a short circuit. Therefore, for this analysis we can redraw the circuit, as seen in Figure 22, step 3.

In Step 4, the voltage source and series inductor (L) is converted back into a current source and parallel inductor. Source transformation, in this simple form, the values of L and C do not change. It is the value of the current and voltage source that changes.

Using Ohm's Law,  $i=v/z$  and  $v = i(1/j\omega c)$ , while  $Z =j\omega L$ , the imaginary j causes the current to be negative at the resonant frequency

specified in Equation 9 of Figure 22. Therefore, by switching the current's direction, the negative sign disappears and the current source is aligned in the same direction as the other one.

## VIII. MATCHING THE OPEN-COLLECTOR DIFFERENTIAL OUTPUT

Figure 23 shows the current-combiner and the open-collector matching circuit. The collector current is increased and passes through the load resistor which allows for more AC output swing.

Since the SA601 has differential open-collector outputs, it is possible to implement both a current-combiner and a flexible matching circuit (see Figure 24).

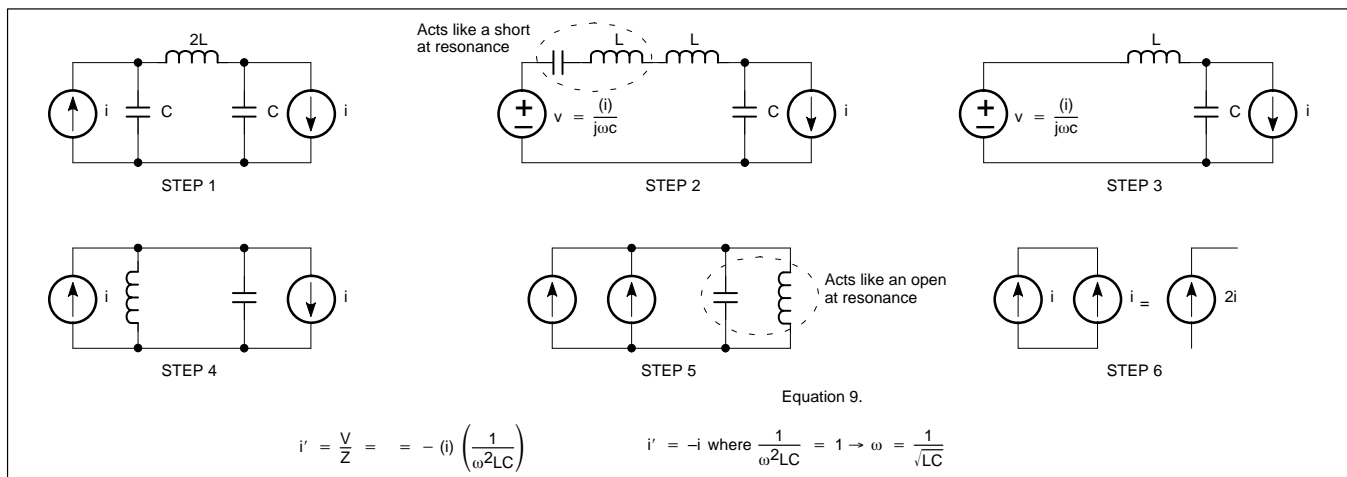


Figure 22. Equivalent Circuit Transformations at Resonance

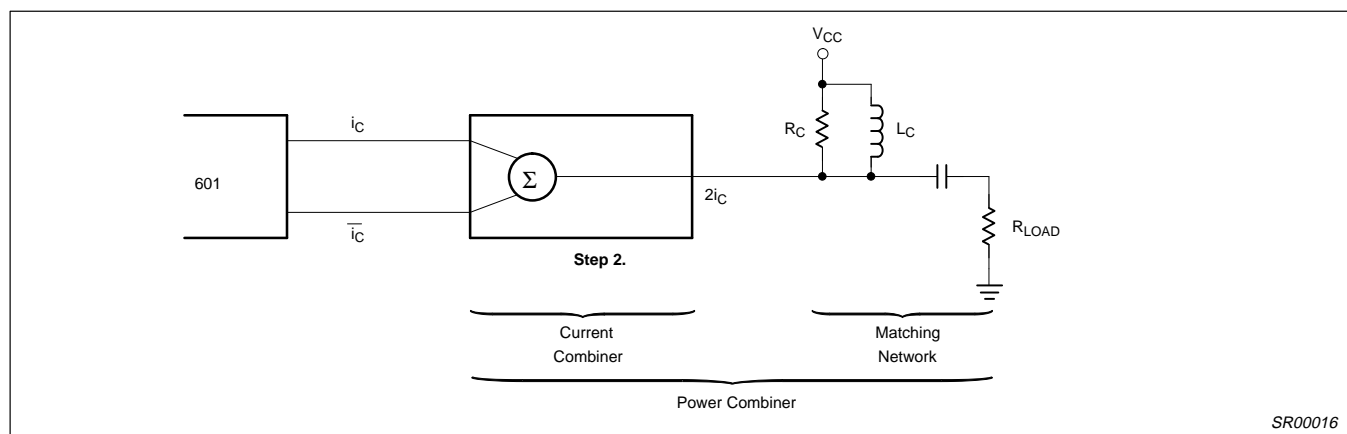


Figure 23. Power Combiner

## Low voltage front-end circuits: SA601, SA620

AN1777

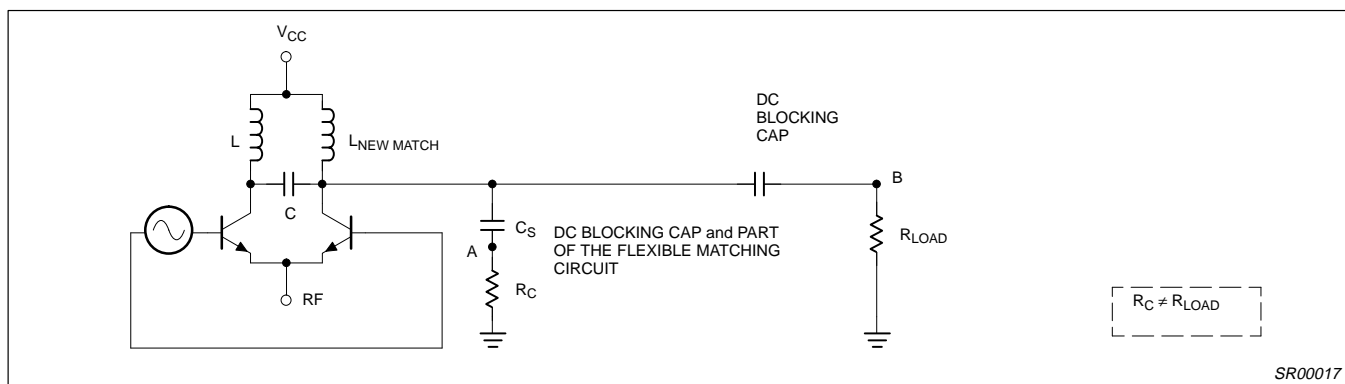


Figure 24. current-combiner with Flexible Matching Circuit

Understanding and implementing the mixer output match is most likely the biggest challenge for those using the SA601. As in most cases, there are many solutions to obtaining the required match for a given circuit. The method selected for the following examples was chosen in order to simply demonstrate some key principles involved in obtaining both high impedance and 50Ω matching while also maintaining some continuity to the previous discussions pertaining to the ideal current-combiner circuit. Each component in the current-combiner/matching circuit will be identified as well as the role each plays in obtaining the required matched condition. In addition, a general procedure for acquiring a good matching circuit along with Smith chart documentation will be provided.

The schematic of the mixer output circuitry utilized on the SA601 demo-board is reproduced here in Figure 25 for convenience.

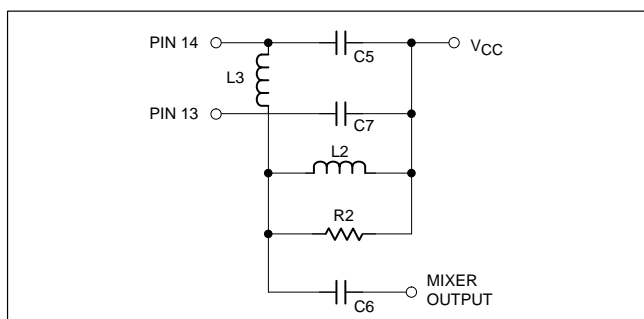


Figure 25. External Mixer Output Circuit

The following is a descriptive summary of the components comprising the external mixer output circuitry.

**Inductor L<sub>2</sub>**

In order to minimize the deviation from the ideal current-combiner circuit, this inductor is chosen to be as large as possible. The inductor will then function only as a choke at the IF frequency and, therefore, should not interact with the current-combiner circuit or effect the matching conditions. This inductor is also necessary to provide the needed DC path from V<sub>CC</sub> to Pins 13 and 14. In all the examples to follow, a 6.8μH inductor was used for L<sub>2</sub>.

**Capacitors C<sub>5</sub> and C<sub>7</sub>**

Also, in order to adhere to the analysis set forth in the ideal current-combiner discussions, these capacitors will be set equal to each other in the examples to follow. The main function of these capacitors is to define the resonant frequency of the current-combiner. They also play a secondary role in defining the output impedance.

**Inductor L<sub>3</sub>**

This inductor also defines the resonant frequency of the current-combiner as well as the output impedance of the current-combiner at resonance.

**Capacitor C<sub>6</sub>**

This capacitor is used to determine the output impedance for 50Ω matches only. For high impedance matches greater than or equal to 1kΩ this capacitor is not very effective. Thus, in this case it is usually replaced with a large capacitance value, adding almost no contribution to the match. A 1000pF is used in the high impedance examples that follow.

**Resistor R<sub>2</sub>**

This resistor is used to simplify the high impedance matching process. In a 50Ω match, the series impedance presented by capacitor C<sub>6</sub> greatly simplifies the process of moving to the 50Ω point on the Smith chart. At high impedance, C<sub>6</sub> is no longer effective and it is often extremely difficult to obtain the proper match by varying just the components associated with the current-combiner circuit. A simple solution to this problem is to obtain the proper resonance condition at a higher impedance than the targeted impedance and then reduce it by placing a resistor of the correct value in parallel with it.

The best way to configure the mixer output circuitry for optimum gain is to optimize the return loss (S<sub>11</sub>) for this port at the required IF. This, by itself, does not guarantee that optimum gain will occur at this frequency due to the phase relationships of the signals inside the current-combiner, but it is usually very close. The best tool available for this is a network analyzer.

**Method of Achieving High Impedance Matching with a Network Analyzer**

The network analyzer lends itself very nicely to obtaining 50Ω matches. However, for high impedance matches the Smith chart data will be far to the right of the chart and will be inaccurate, hard to read, and hard to interpret. If the network analyzer were normalized to the impedance value to which you want to match, the data would be in the center of the Smith chart and easy to interpret.

One method of doing this is to disconnect the 'A' port from an HP network analyzer and attach a high impedance probe to this port. Next take an SMA connector (or whatever connector type your analyzer uses) and solder two resistors each equal to the target impedance in the following manner: Solder one end of one of the resistors to the center lead of the connector and leave the other end open. This resistor will define the normalization on the network analyzer during calibration. Next, solder one end of the other resistor to the ground of the connector and leave the other end

# Low voltage front-end circuits: SA601, SA620

AN1777

open. This resistor will function as a dummy load during calibration. Connect this SMA connector to port '1' on the network analyzer. You are now ready to calibrate. Prior to initiating the open portion of the one-port calibration procedure, contact the open end of the resistor soldered to the center lead of the connector with the high impedance probe (see Figure 26). After this has been completed, contact ground on the connector with the high impedance probe prior to initiating the short portion of the one-port calibration procedure (see Figure 26). Then connect the two open ends of the resistors with solder. Prior to initiating the load portion of the one-port calibration procedure, contact the connection between the resistors with the high impedance probe (see Figure 27). The network analyzer should now be normalized to your target impedance. Thus, the optimum match will once again be in the center of the chart. When making connection to a circuit it is necessary to include a resistor of the same value used during the calibration between the center lead of the connector and its connection to the circuit. Impedance measurements are taken by contacting the high impedance probe at the end of this resistor nearest the circuit (see Figure 28).

Another tip concerning the calibration of the network analyzer should be mentioned. It is often useful to be able to look at a Smith chart over more than one frequency range. A wide frequency range is useful initially when your results are far off the target. Then the narrower range is useful for fine tuning your results. So, it is recommended that you calibrate in both frequency ranges and save the settings in the internal registers of the network analyzer if possible.

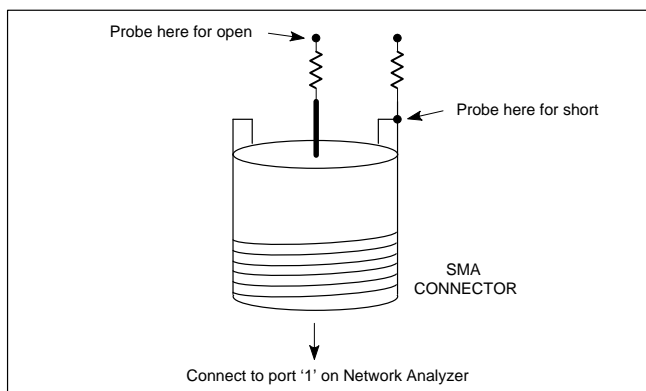


Figure 26. Open- and Short-circuit Calibration Locations

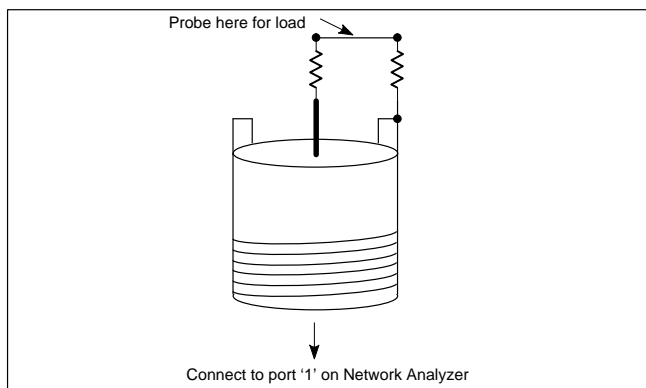


Figure 27. Load Impedance Calibration Location

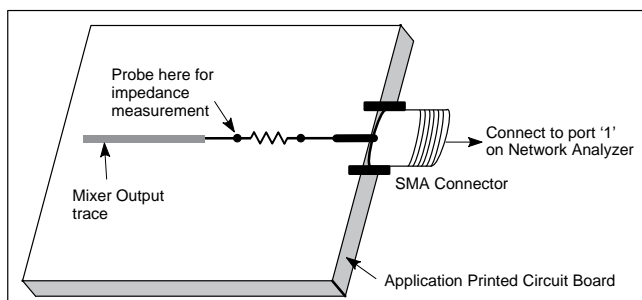


Figure 28. PCB Impedance Measurement Configuration

## Non-Ideal Current-Combiner Ckt Considerations

Before we go through some actual impedance matching examples, some key differences between the ideal current-combiner circuit presented earlier and a non-ideal circuit which takes into account the finite Q of inductor  $L_3$  as well as the output impedances of the current sources should be discussed. Through a series of Thevenin/Norton conversions and Series/Parallel equivalent impedance conversions similar to the analysis of the ideal circuit, the mixer output circuit can be modeled by the circuit shown in Figure 29.

The two main things to note in this circuit are the presence of the shunt resistors,  $R_O$  and  $R_Q$ , and that the current-combiner circuit looks like a simple parallel LRC circuit to capacitor  $C_6$ .

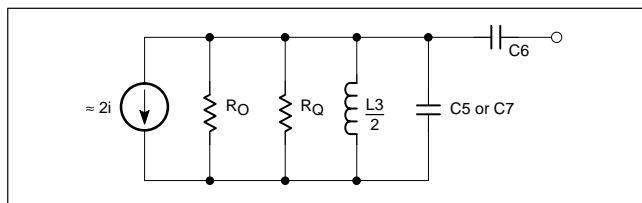


Figure 29. Non-ideal Mixer Output Equivalent Circuit

The significance of the resistors is their role in determining the output impedance of the overall circuit.  $R_Q$  is smaller in value than  $R_O$  which is the relatively high impedance of the open-collector outputs. Thus, because they are in parallel,  $R_Q$  defines the output impedance of the current-combiner circuit at resonance. The value of  $R_Q$  is a function of frequency, the component Q of inductor  $L_3$ , and the inductance value of inductor  $L_3$ . We do not have direct control over component Q or frequency, so the value of  $R_Q$  is adjusted by simply changing the value of inductor  $L_3$ . A more detailed explanation of this will be shown in the examples to follow.

It is not intuitively obvious why the matching portion of the mixer output circuitry used to obtain the  $50\Omega$  matches is composed of a single series capacitance. Many customers using the SA601 have asked why this works because it does not seem to adhere to basic two element matching concepts. To answer this, let's first take a quick look at the general parallel LC circuit shown in Figure 30. This circuit will resonate according to the simple resonance calculation

$$\omega = \frac{1}{\sqrt{LC}}$$

If the capacitor of this circuit is cut in half and the inductor is equivalently represented by two parallel inductors, the circuit in Figure 31 results. At the original resonant frequency  $\omega = \frac{1}{\sqrt{LC}}$ , the capacitor and one of the inductors will resonate  $\omega = [(0.5C)(2L)]^{-\frac{1}{2}} = (LC)^{-\frac{1}{2}}$ . What is left over is a shunt inductance of  $2L$  presented to the output. Thus, the general

## Low voltage front-end circuits: SA601, SA620

AN1777

principle is that by decreasing the capacitance of a parallel LC circuit, you will present a shunt inductance to the output of that circuit at the same frequency. This concept when applied to the circuit in Figure 29 offers an explanation of where the missing shunt inductance element is coming from. If capacitors  $C_5$  and  $C_7$  are decreased, this will present a shunt inductance to capacitor  $C_6$ . Thus, two element matching concepts would still apply to this circuit. How much inductance is referred to  $C_6$ , and what value of  $C_6$  it takes to obtain the matching conditions, is difficult to predict. So, we will rely on the network analyzer to point us in the right direction.

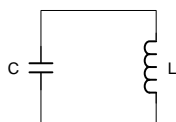


Figure 30. Parallel LC Network

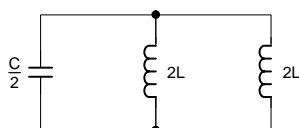


Figure 31. Parallel LC Network with Decreased Capacitance

One of the most frustrating parts of matching on the network analyzer is it often does not give you very much information until you are at least "in the ball park" of the component values required. The first example will purposely begin with component values to create this condition to give an idea of what might initially be encountered. Additionally, Table 4 at the end of this section contains values for the components required to obtain optimum gain from the mixer at IF frequencies of 45, 83 and 110MHz at both a high impedance of 1k $\Omega$  and a low impedance of 50 $\Omega$ . Using this table you should be able to obtain a good guess for the initial values for your particular application.

## IX. MATCHING EXAMPLES

A procedure for acquiring a good high impedance matching circuit for providing optimum gain from the mixer output can be summarized as follows:

1. Normalize the network analyzer to the impedance of interest.
2. Set inductor  $L_2$  to be a large inductance value.
3. Choose "ball park" values for the initial component values based on the simple resonance calculation and /or the tabulated data provided in Table 4.
4. Adjust  $C_{5,7}$  to obtain the required match.
5. If the output impedance at resonance is above its target and resonance occurs at the targeted IF, the required match can be obtained with the appropriate value for  $R_2$ .
6. If the resonant frequency is above the target IF and the output impedance at resonance is below its target, change inductor  $L_3$  to a higher value and return to step 4.
7. If the resonant frequency is below the target IF and the output impedance at resonance is above its target, change inductor  $L_3$  to a lower value and return to step 4.

8. Design a matching circuit to bring the impedance back down to 50 $\Omega$ .
9. Make final fine tuning adjustments to  $C_{5,7}$  based on the frequency response as observed on a spectrum analyzer.

A procedure for acquiring a good 50 $\Omega$  impedance matching circuit for providing optimum gain from the mixer output can be summarized as follows:

1. Set inductor  $L_2$  to be a large inductance value.
2. Choose "ball park" values for the initial component values based on the simple resonance calculation and/or the tabulated data provided.
3. If the output impedance at resonance is greater than its target, increase capacitor  $C_6$  until the curve on the Smith chart passes through the center of the chart.
4. If the output impedance at resonance is less than its target, decrease capacitor  $C_6$  until the curve on the Smith chart passes through the center of the chart.
5. If the resonant frequency is greater than the target IF, increase the values of  $C_5$  and  $C_6$  until resonance occurs at the target IF.
6. If the resonant frequency is less than the target IF, decrease the values of  $C_5$  and  $C_6$  until resonance occurs at the target IF.
7. Make final fine tuning adjustments to  $C_{5,7}$  based on the frequency response as observed on a spectrum analyzer.

### 83MHz, 1 k $\Omega$ Match

The objective of the first matching circuit we will look at is intended to provide a 1k $\Omega$  match at 83MHz. It has inductance values  $L_2 = 6.8\mu\text{H}$  and  $L_3 = 270\text{nH}$ . (These will be the values for these inductors in all the examples unless stated otherwise.) This match is a high impedance 1k $\Omega$  match so capacitance  $C_6$  was set to 1000 pF. The simple resonant calculation

$$\omega = [(0.5CL)]^{-\frac{1}{2}} = [(0.5)(270\text{nH})(C)]^{-\frac{1}{2}} = 2\pi(83\text{MHz})$$

suggests that  $C_{5,7}$  should be approximately 27pF. The actual value needed to optimize the gain is always less than the value predicted by this equation. Figure 32 shows what an unfavorable Smith Chart might look like when you go in the wrong direction. In this example  $C_{5,7}$  was set to 33pF. The chart shows resonance does not occur anywhere between 75 and 95MHz. If you were to continue this curve, it would eventually hit the real axis at a much lower frequency. According to the simple resonance calculation this suggests that our capacitance value is too large. As  $C_{5,7}$  is decreased, the plot on the Smith chart starts to resemble a constant admittance circle. Figure 33 shows that a  $C_{5,7}$  value of 23pF yields a more favorable curve where resonance occurs at 83MHz as desired. The only problem is the impedance at resonance is not 1k $\Omega$  as desired. The Smith chart in Figure 33 has been normalized to 1k $\Omega$ . The real part coordinate of 61.984 $\Omega$ , shown on the Smith chart must be converted in the following manner:

$$Z_O = (61.984/50)(1\text{k}\Omega) = 1.24\text{k}\Omega$$

We could at this point choose another value for inductor  $L_3$  and then again find the right value for  $C_{5,7}$  to acquire resonance. However, we will take this opportunity to demonstrate how resistor  $R_2$  might be used. The needed shunt resistance  $R_2$  can be calculated from the simple formula for combining parallel resistances.

$$R_2 = \frac{(Z_O)(Z_{\text{TARGET}})}{(Z_O - Z_{\text{TARGET}})} = \frac{(1.24)(1)}{(1.24 - 1)} = 5.17\text{k}\Omega$$

Figures 34 and 35 show the resulting output match when a 5k $\Omega$  resistor is used for  $R_2$ .

# Low voltage front-end circuits: SA601, SA620

AN1777

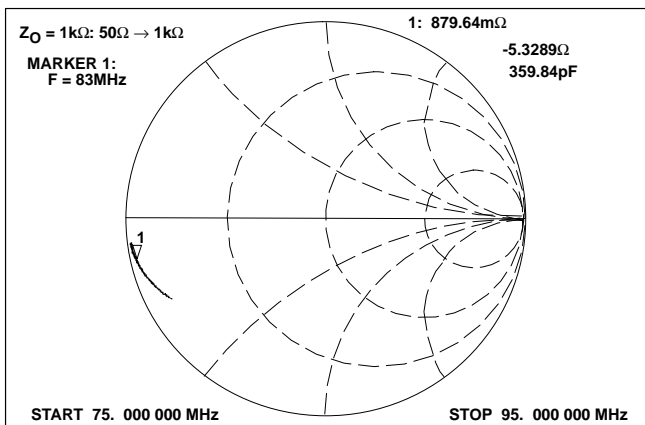


Figure 32. Smith Chart  $S_{11}$ : Showing Poor Initial Conditions

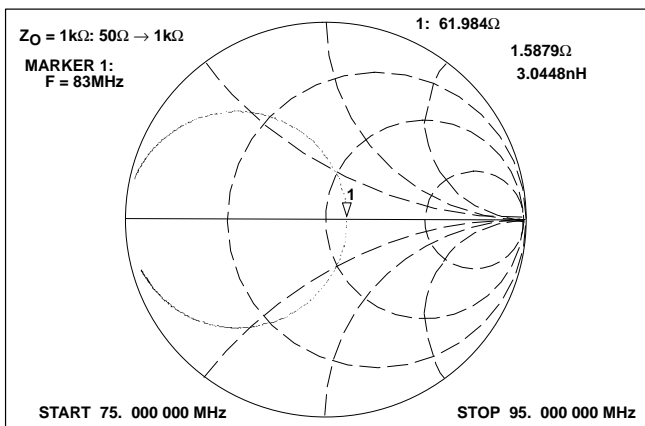


Figure 33. Smith Chart  $S_{11}$ : Showing Targeted Resonant Frequency, But Output Impedance Is Too High

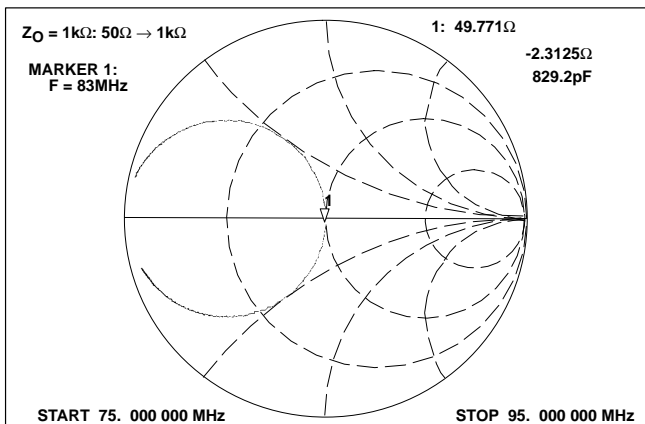


Figure 34. Smith Chart  $S_{11}$ : Showing Targeted Resonant Frequency And Output Impedance

it can then be terminated into a standard 50Ω test equipment port. Figures 36, 37 and 38 show the design of such a circuit. A detailed discussion of the design of this type of circuit can be found in [1]. Figure 39 shows the frequency response of the mixer output as the LO was varied to change the IF output. The envelope of this response has been added for clarity. The RF input signal to the mixer was -30dBm. So, the plot shows a very favorable gain of approximately 10.6dB at 81.75MHz. At 83MHz, which is the intended IF, the gain is only about 1dB lower.

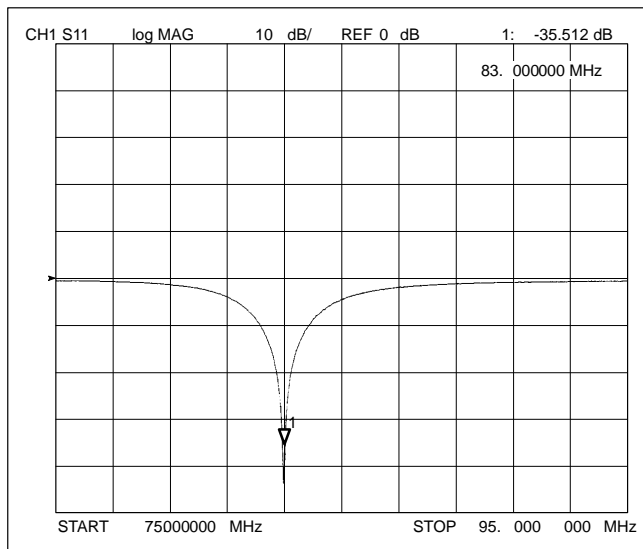


Figure 35.  $S_{11}$  Reflection at Mixer Output

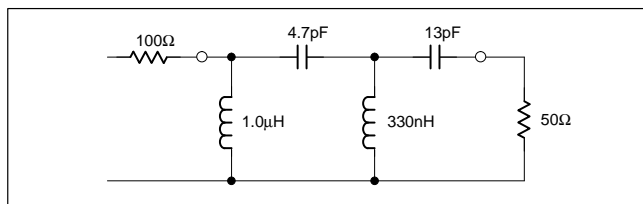


Figure 36. Wideband 1kΩ to 50Ω Matching Network

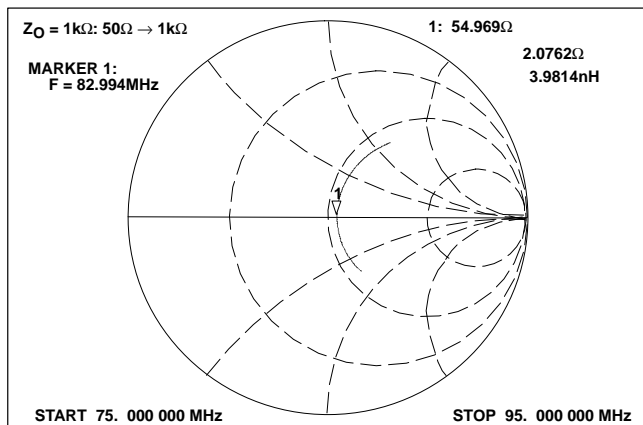
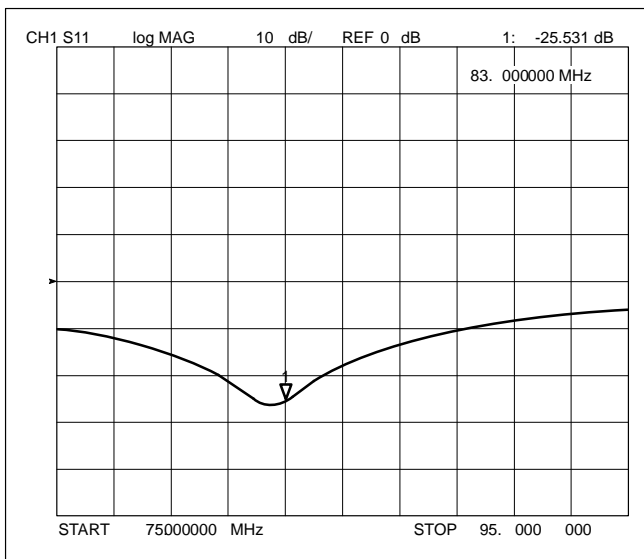


Figure 37. Smith Chart  $S_{11}$ : Showing Wideband Matching of Circuit in Figure 36

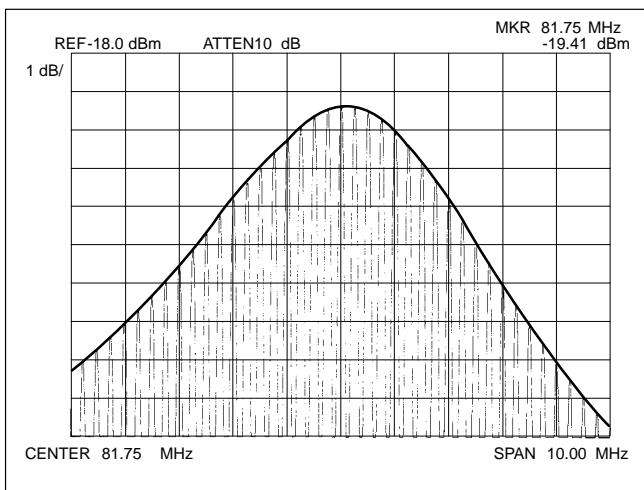
In order to verify the high impedance match in the absence of a system in which to test it, a wideband matching circuit was designed to convert the high impedance 1kΩ match back down to 50Ω where

# Low voltage front-end circuits: SA601, SA620

AN1777



**Figure 38. Smith Chart S<sub>11</sub>: Showing Wideband Matching of Circuit in Figure 36**



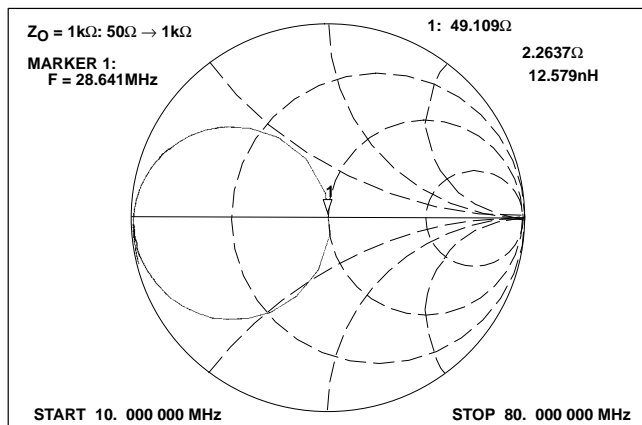
**Figure 39. Showing the Frequency Response of the Mixer Output**

## 45MHz, 1 kΩ Match

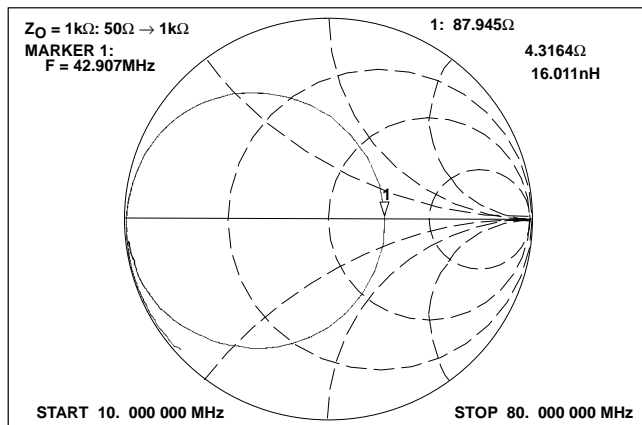
The next example is designed to show how to determine when inductor L<sub>3</sub> is too low or too high to enable you to acquire the proper matching conditions. The objective is to provide a 1kΩ match at the much lower IF of 45MHz. Before we can determine if we have the wrong inductance value for L<sub>3</sub>, we must first understand what effect adjusting C<sub>5</sub> and C<sub>7</sub> has on the circuit. The simple resonance calculation suggests that a decrease in C<sub>5,7</sub> should increase the frequency at which resonance occurs. This also causes a change in the output impedance at resonance. When C<sub>5,7</sub> is decreased the output impedance at the new higher resonant frequency will also be higher. As C<sub>5,7</sub> is adjusted the resonant frequency and the output impedance at resonance will change in the same direction. In other words, both will increase as C<sub>5,7</sub> is decreased or both will decrease as C<sub>5,7</sub> is increased. This means that if a condition exists where the resonant frequency is greater than the target IF and the impedance at resonance is below its target, the inductance value used for

inductor L<sub>3</sub> is too small. Conversely, if the resonant frequency is less than the target IF and the impedance at resonance is greater than its target, the inductance value used for L<sub>3</sub> is too large.

To demonstrate these concepts, inductor L<sub>3</sub> was chosen to be larger than usual (L<sub>3</sub> = 750nH) and C<sub>5,7</sub> was also chosen to be much larger than usual (C<sub>5,7</sub> = 82pF). Capacitor C<sub>6</sub> was again set to 1000pF making it negligible during high impedance matching. Again, it should be noted that the following Smith charts have been normalized to 1kΩ. Figure 40 shows that the component values listed above yield a 1kΩ output impedance at a resonant frequency of just 28MHz instead of the 45MHz target IF. In an effort to increase the resonant frequency C<sub>5,7</sub> was decreased to 33pF. Figure 41 shows that the resonant frequency is close but below the target IF and the impedance is well above 1kΩ. According to the discussions above, this suggests that inductor L<sub>3</sub> is too large. Figure 42 shows the results when inductor L<sub>3</sub> is decreased to 620nH. This chart shows that both the resonant frequency and the impedance at resonance are greater than their targeted values. This indicates that there is still a chance that the match can be made with some further adjustment of C<sub>5,7</sub>. C<sub>5,7</sub> was increased to 39pF to lower the resonant frequency. Figure 43 shows that when this was done the condition that suggests that inductor L<sub>3</sub> is too large still exists.



**Figure 40. Smith Chart S<sub>11</sub>: Showing Resonant Frequency Well Below 45MHz Target**



**Figure 41. Smith Chart S<sub>11</sub>: Showing Output Impedance Well Above the 1kΩ Target**

# Low voltage front-end circuits: SA601, SA620

AN1777

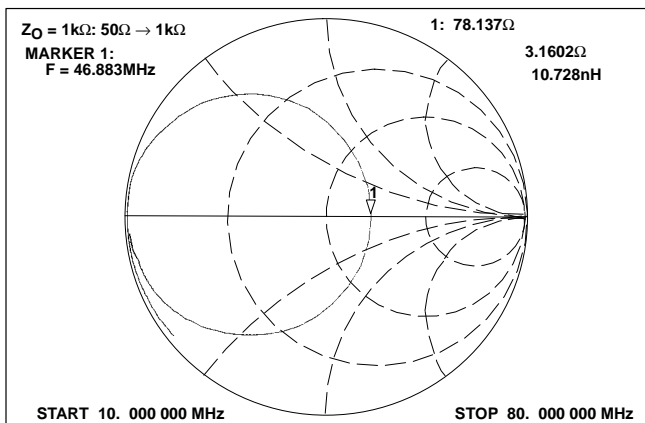


Figure 42. Smith Chart  $S_{11}$ ,  $L_3 = 620\text{nH}$ : Showing the Resonant Frequency and the Output Impedance are Both Above the 45MHz, 1k $\Omega$  Target Values

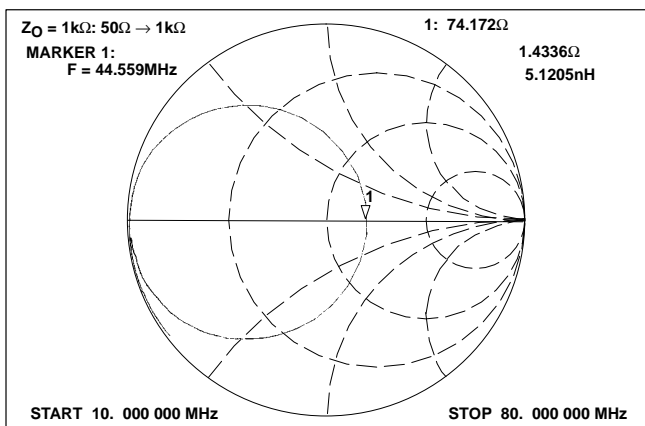


Figure 43. Smith Chart  $S_{11}$ ,  $L_3 = 620\text{nH}$ : Showing the Resonant Frequency Below 45MHz Target and Output Impedance Above 1k $\Omega$  Target Values

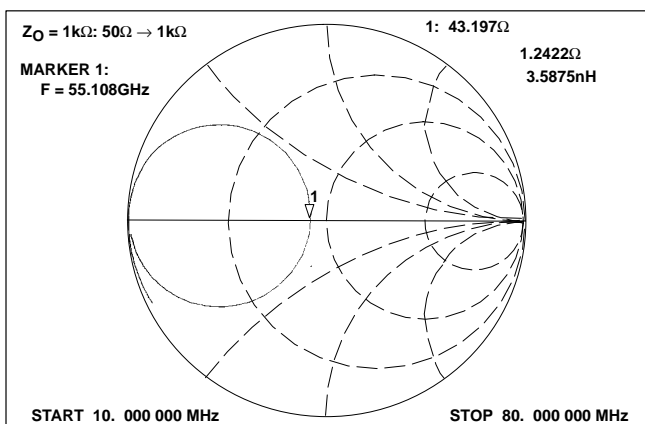


Figure 44. Smith Chart  $S_{11}$ ,  $L_3 = 270\text{nH}$ : Showing the Resonant Frequency Above 45MHz Target and the Output Impedance Less than 1k $\Omega$  Target Values

Next, in order to demonstrate the other condition, inductance  $L_3$  was changed to 270nH and  $C_{5,7}$  was changed to 56pF. Figure 44 shows that the resonant frequency is greater than the target IF and the output impedance at resonance is less than 1k $\Omega$ . According to the

discussions above, this is the condition which suggests that inductance  $L_3$  is too small. So, inductance  $L_3$  was then increased to 390nH. Figure 45 shows that the 390nH inductance yields a resonant frequency and an output impedance at resonance that are both slightly above their targets. At this point you are close enough to use another wideband matching circuit for dropping the 1k $\Omega$  impedance to 50 $\Omega$  allowing you to check the frequency response directly. It was determined that changing  $C_{5,7}$  from 56pF to 62pF yielded optimum gain at exactly the target IF. Figure 46 shows this gain to be approximately 10dB.

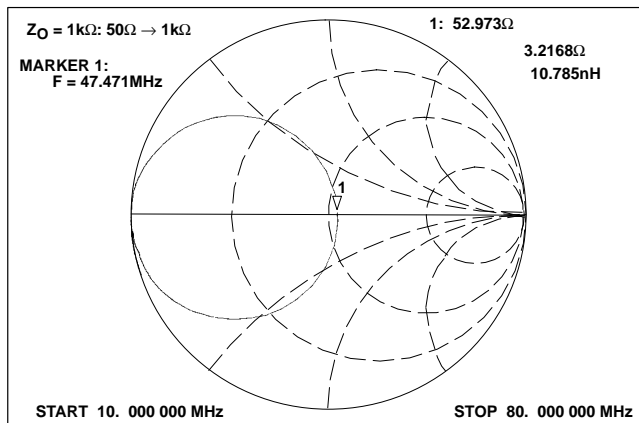


Figure 45. Smith Chart  $S_{11}$ ,  $L_3 = 390\text{nH}$

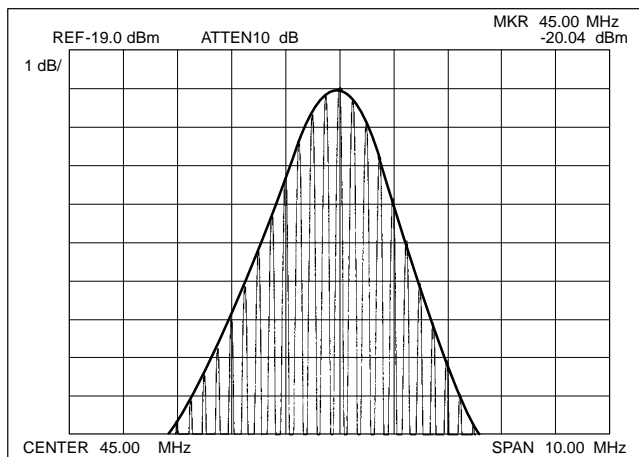


Figure 46. Frequency Response of the Mixer Output

### 110.592MHz, 50 $\Omega$ Match

The next example will demonstrate how to obtain a 50 $\Omega$  match. There are a couple of things to note concerning matching at a higher IF. The first is to expect an increase in the general sensitivity of the matching conditions with relatively small component variations. Secondly, the difference in frequency between the point of optimum gain and the point where the return loss ( $S_{11}$ ) is minimized greatly increases when matching at a higher IF. To demonstrate these things the objective of the following example will be to provide a 50 $\Omega$  match at 110.592MHz, which is a relatively high 1st IF.

The initial component values for this matching circuit (see Figure 25) are 6.8uH and 270nH for inductances  $L_2$  and  $L_3$ , respectively. Capacitors  $C_5$ ,  $C_6$  and  $C_7$  were all set to 10pF. It should be noted that  $C_6$  is no longer set at 1000pF and plays a major role in determining the required match at 50 $\Omega$  as we will see. Figure 47 shows the results of the initial component values. The Smith chart



# Low voltage front-end circuits: SA601, SA620

AN1777

indicates that the resonance frequency is much too low and the impedance at resonance is too high. In our previous high impedance matching discussions this meant that the inductor  $L_3$  was too large. In  $50\Omega$  matching this is no longer the case. Due to the lower impedance matching condition, the output impedance at resonance can now be significantly altered by varying the series capacitance element  $C_6$ . As mentioned previously, the matching circuit is much more sensitive to component variations at a higher IF. Figure 48 shows a small change in  $C_{5,7}$  from  $10\text{pF}$  to  $4.7\text{pF}$  caused the resonant frequency to shift from well above the target IF to well below the target IF with very little change in the output impedance at resonance.

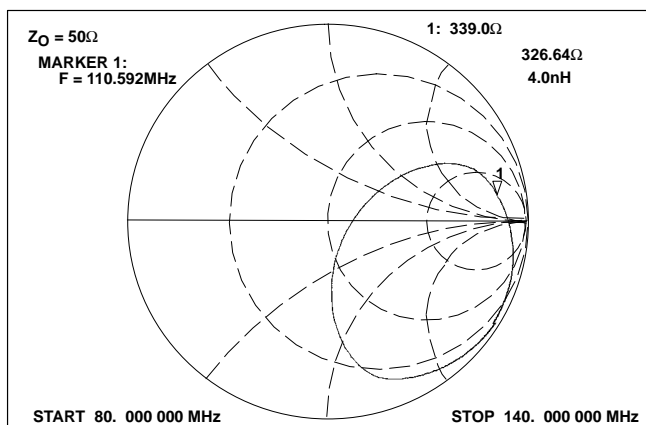


Figure 47. Smith Chart  $S_{11}$ : Showing Initial Component Results

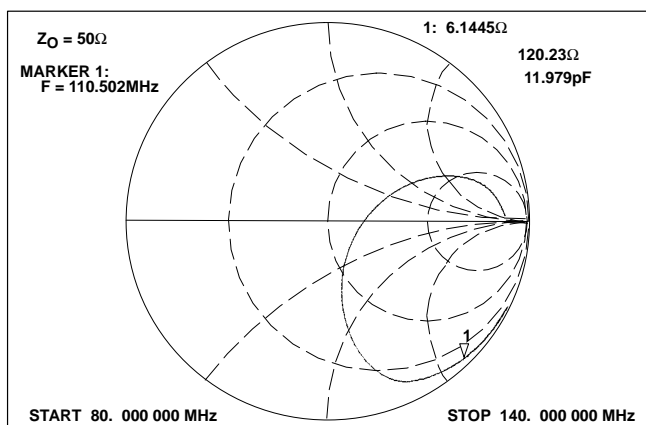


Figure 48. Smith Chart  $S_{11}$ : Showing Sensitivity to Small Changes in  $C_{5,7}$

These observations suggest a general method of attack for obtaining the output impedance at  $50\Omega$ . When the series capacitance element  $C_6$  is increased/decreased it will cause the somewhat circular curve on the Smith chart to increase/decrease in diameter. An effective method for obtaining the required match would be to adjust  $C_6$  such that this curve passes through the center of the Smith chart and then make the necessary adjustments to  $C_{5,7}$  to set the resonant frequency at the target IF. Looking at Figure 48 again, we see that the output impedance at resonance is too high. We should be able to decrease this by increasing the series capacitance element  $C_6$ . In addition, from Figures 47 and 48 we know that the correct value for  $C_{5,7}$  to obtain resonance at  $110.592\text{MHz}$  is between  $4.7\text{pF}$  and  $10\text{pF}$ . So, we will increase  $C_6$  to  $12\text{pF}$  and set  $C_{5,7}$  to  $6\text{pF}$  and see what happens. Figure 49 shows

the output impedance at resonance is very close to the target but the resonant frequency is still a bit too high. Figures 50 and 51 show that by adding  $1\text{pF}$  to  $C_{5,7}$  the required match is obtained. An additional matching network is not needed to evaluate the circuit because we are already at  $50\Omega$ .

In previous discussions, it was mentioned that the frequency at which the return loss is obtained does not guarantee optimum gain at this same frequency due to the phase relationships of the signals within the current-combiner circuitry. Figure 52 shows that approximately  $12\text{dB}$  of gain is obtained but at a frequency of  $115\text{MHz}$ , which is approximately  $5\text{MHz}$  away from the targeted IF. To correct this we simply adjusted  $C_{5,7}$  to  $8.5\text{pF}$  to obtain a similar condition  $5\text{MHz}$  lower than the previous result as shown in Figures 53, 54 and 55.

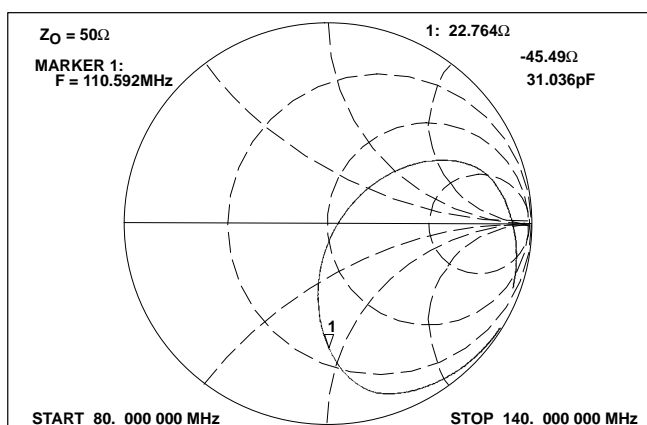


Figure 49. Smith Chart  $S_{11}$ : Showing Results of Adjustment to Capacitor  $C_{5, 6, 7}$

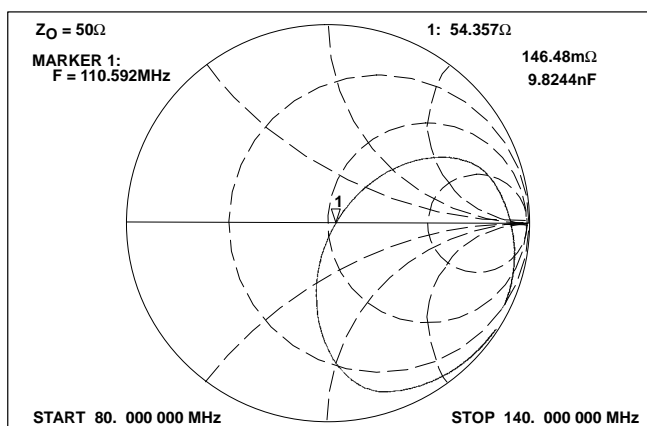


Figure 50. Smith Chart  $S_{11}$ : Showing Final  $110.592\text{MHz}$ ,  $50\Omega$  Results

Table 4. Mixer Output Component Values

	50Ω Match			1kΩ Match		
	MHz			MHz		
	45	83	110	45	83	110
$L_1$	6.8μH			6.8μH		
$L_3$	270nH			390nH	270nH	
$C_{3, 7}$	80pF	18pF	8.5pF	62pF	22pF	10pF
$C_6$	22pF	15pF	12pF	1000pF		
$R_2$	-----			—	5k	—

# Low voltage front-end circuits: SA601, SA620

AN1777

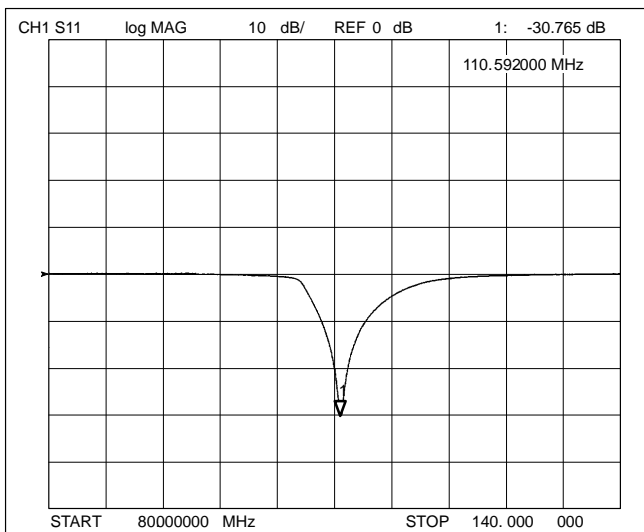


Figure 51. Final 110.592MHz, 50Ω Results

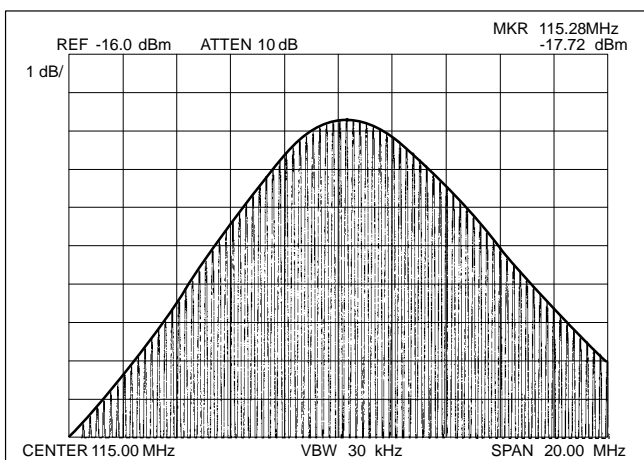


Figure 52. Mixer Output Freq. Resp. Peak is 5MHz Too High

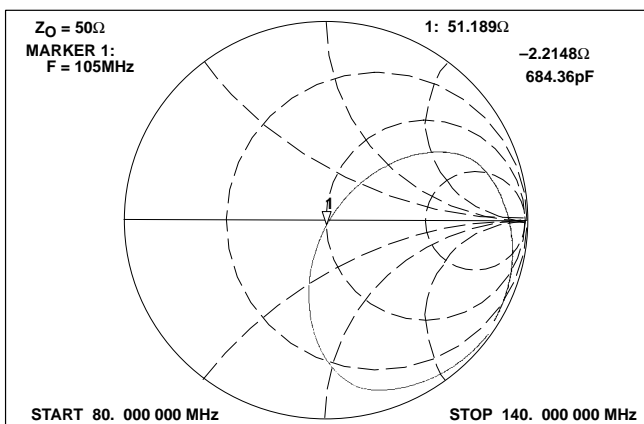


Figure 53. Smith Chart  $S_{11}$ : Showing Match With  $C_{5,7} = 8.5\text{pF}$

## SA601 IP<sub>3IN</sub> Considerations

It should be noted that it is often necessary to give up some mixer gain of the device in order to obtain acceptable IP<sub>3IN</sub> performance. The previous examples and procedures demonstrated how to optimize the gain. To meet the IP<sub>3IN</sub> specifications of your particular

application, it may be necessary to make some changes. The demoboard schematic for the SA601 in Figure 67 shows these changes. One difference between the previous discussions and this schematic is that  $C_5$  and  $C_7$  are not of equal value. It has been found that loading the differential output in an asymmetrical fashion by making  $C_5$  less than  $C_7$  is beneficial to IP<sub>3IN</sub> performance. So, your frequency adjustments would then be made by keeping  $C_7$  constant and varying  $C_5$ . Also, inductor  $L_2$  can no longer be chosen arbitrarily large. It must be chosen such that a high impedance parallel resonance condition occurs with  $C_7$  at the frequency of interest. Resistance  $R_2$  can then be used to obtain the required high impedance match.  $C_6$  is used to acquire the 50Ω match exactly as before. The IP<sub>3IN</sub> and gain performance for this configuration at 83MHz is shown in Figures 56 and 57. Notice the gain is approximately 8dB which is approximately 2dB less than that obtained in the previous examples. This was the trade-off for obtaining the better IP<sub>3IN</sub> performance (see Figure 57).

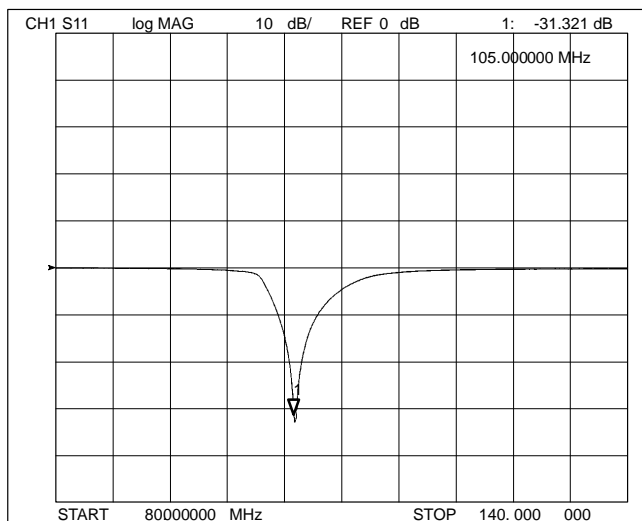


Figure 54. Mixer Output Set 5MHz Below Target to Get Final Result at 110.592MHz

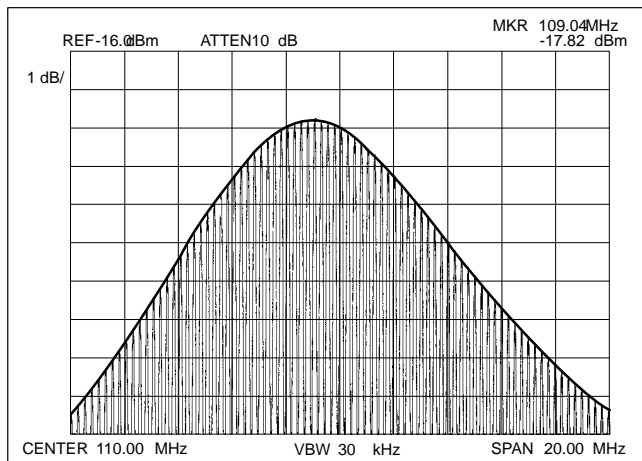


Figure 55. Mixer Output Set 5MHz Frequency Response Peak Very Near 110.592MHz Target

## Summary of Mixer Open-Collector Output Concepts

The open-collector of a transistor offers a designer using the SA601 and SA620 the flexibility to provide impedance matching with

# Low voltage front-end circuits: SA601, SA620

AN1777

minimal external components. Additionally, when an inductor is strategically located in the circuit, more AC output swing will occur due to the Early-Voltage effect of the device without adding any additional current.

When using the SA601 differential open-collector output, a designer can use the current-combiner circuit to combine the currents such that additional output signal swing is achieved. This preserves the RF signal and thus eliminates the need for interstage amplifiers. The current-combiner differential mixer output is not available on the SA620.

A flexible matching circuit can be used to deliver an equal amount of power to unequal resistive loads. The flexible matching circuit is ideal for trouble-shooting.

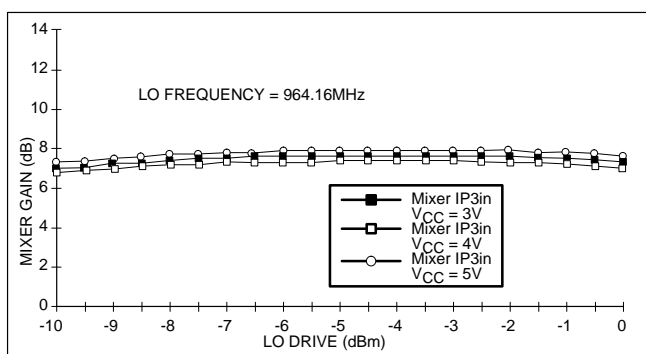


Figure 56. Mixer Gain vs LO Drive SA601

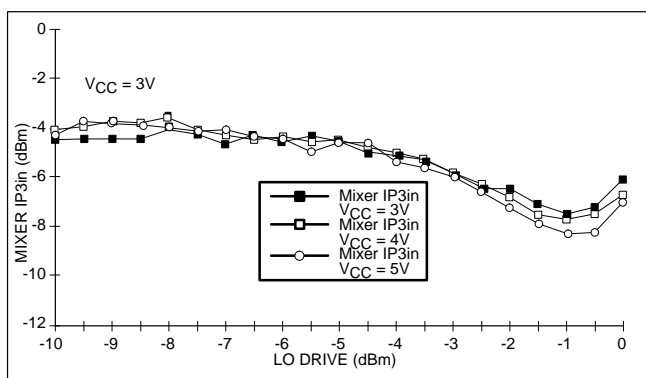


Figure 57. Mixer IP<sub>3IN</sub> vs LO Drive SA601

## X. SA601 MIXER CHARACTERIZATION

Figure 56 shows a mixer gain vs. LO drive curve for a SA601 utilizing the current-combiner circuit in an application board environment. It shows that over an LO drive-level range of -10dBm to 0dBm the mixer gain deviates from an average value of 7.5dB by less than 0.5dB. It also shows that the gain will change by less than 0.5dB when V<sub>CC</sub> is varied from 3V to 5V.

The noise figure of the SA601 generally increases with decreasing LO drive. Over the LO drive-level range of -10dBm to 0dBm the noise figure varied from approximately -12dB to approximately -8dB, respectively. The noise figure at -5dBm LO drive is approximately 10dB.

Figure 57 shows the mixer's 60kHz IP<sub>3IN</sub> vs. LO drive curve for the SA601 in an application board environment. It should be noted that the IP<sub>3IN</sub> was measured with a -35dBm RF input at 881MHz and an offset of just 60kHz. 881MHz is the center of the 869 to 894 IS-54

Rx band, -35dBm was selected to ensure P<sub>IN</sub> vs P<sub>OUT</sub> linearity, and 60kHz was chosen as a suitable representation of current application alternate-channel constraints. The figure shows IP<sub>3IN</sub> to be constant at approximately -4.3dB over an LO drive-level range from -10dBm to approximately -5dBm. For LO drive-levels greater than -5dBm, IP<sub>3IN</sub> will decrease by approximately 1dB for every 1dB increase in LO drive. It is for this reason, that even though the mixer noise figure continues to decrease with larger LO drives, the LO drive-level which optimizes gain, noise figure and IP<sub>3IN</sub> for current IS-54 applications is approximately -5dBm.

## SA601 System 12dB SINAD Performance

Figure 59 shows the 12dB SINAD vs RF input frequency of a receiver system composed of the SA601 utilizing the differential mixer output and the SA606 low-voltage FM-IF as shown in Figure 58. Data was taken over an input frequency range covering the 869 - 894MHz IS-54/AMPS Rx band as well as the 902 - 928 ISM band. The 12dB SINAD performance in both bands was approximately -121 to -122dBm without a duplexer. The system 12dB SINAD with a duplexer present will typically increase by approximately 3dB.

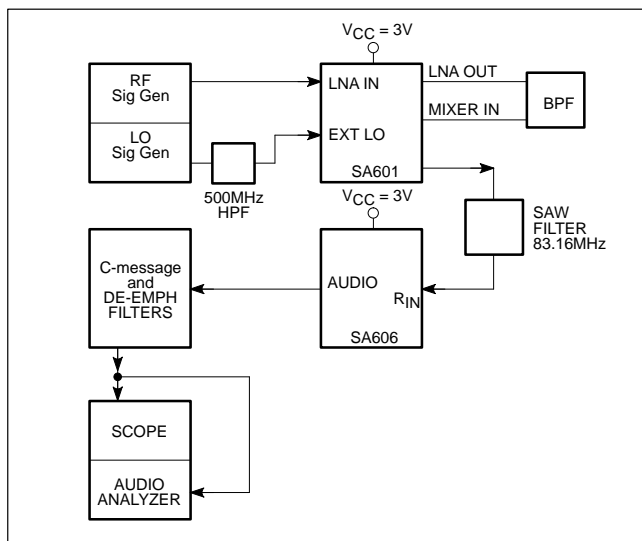


Figure 58. SA601 12dB SINAD System Test Configuration

## XI. SA620 VCO

The SA620 features a low-power integrated VCO for providing an LO to the mixer. Thus, the additional cost and space associated with the use of an external VCO are eliminated. In the Philips SA620 application board environment shown in Figure 71, the VCO can be operated from 900MHz to approximately 1200MHz with the frequency range (using a 5V control voltage) increasing from 20MHz to 70MHz, respectively. Figure 60 shows the LO frequency vs. control voltage for two LO ranges centered at 960 and 995 MHz.

Figure 61 shows the VCO output power for the same LO ranges shown in Figure 60. The average VCO output power is approximately -20dBm and varies less than 0.5dB over the 5V control voltage range.

Figure 62 shows the VCO phase noise at a 60kHz offset for the same LO ranges shown in Figures 60 and 61. The average phase noise (60kHz offset) is approximately -103dBc/Hz on non-silver-plated application boards and varies approximately 1dB across the 5V control voltage range. Phase noise performance may be improved another 1 - 2 dBc/Hz in with silver plating (see

# Low voltage front-end circuits: SA601, SA620

AN1777

discussion below). It should be noted that a significant degradation of the phase noise performance was observed in the application board environment when the VCO was operated at frequencies above 1100MHz.

## XII. μ-STRIP INDUCTOR OSCILLATOR RESONANT CIRCUIT [4, 5, 8]

The SA620 application board utilizes a high Q, μ-strip inductor in its oscillator resonant circuit to improve phase noise performance. Information concerning phase noise, μ-strips, and their implementation on the SA620 applications board is presented below

### General Theoretical Background

Phase noise  $S_{\theta}(f_m)$  is a performance parameter of an oscillator circuit's spectral purity describing the energy it produces at frequencies  $f_m$  on either side of the wanted carrier frequency  $f_c$ . To express how this works, an oscillator can be modeled as a feedback loop employing a phase modulator and noise-free amplifier to show how the resonator effects the phase characteristics of the resulting output. (A complete analysis of this model can be found in [4].)

In Figure 63, the phase spectral-density noise factor  $S_{\theta IN}$  models internal phase noise produced by the oscillator's active components (e.g. BJT, FET etc.). Essentially, this factor is filtered by an external resonator that accepts energy from the noiseless amplifier which has sufficient gain  $G_O$  to sustain closed loop oscillations. Active circuitry with low internal phase noise will produce cleaner outputs for a given degree of filtering. Similarly, increasing the filtering will produce a cleaner output for a given internal spectral phase noise. Note that because of the feedback arrangement, the output spectral phase noise can never be less than that produced internally. These relations are summarized in the following equations, assuming the external resonator to be a simple parallel resonant LC tank circuit.

$$S_{\theta OUT}(f_m) = |H(j\theta)|^2 S_{\theta IN}(f_m) \tag{10}$$

$$S_{\theta OUT}(f_m) = \left[ 1 + \left( \frac{f_o}{2Q_L f_m} \right)^2 \right] S_{\theta IN}(f_m); (f_m) > 0 \tag{11}$$

In practice, a spectrum analyzer is commonly used to obtain an indirect measure of phase noise by measuring power vs. frequency, or the power spectrum on some specified bandwidth at various carrier offset frequencies  $m$  and then normalizing each reading to the equivalent power in a bandwidth of 1Hz. This technique of measuring SSB phase noise  $L(m)$  usually takes the peak carrier power as the 0dB reference. Thus, Equation 11 can be expressed as

$$L_{OUT}(f_m) = \frac{1}{2} \left[ 1 + \left( \frac{f_o}{2Q_L f_m} \right)^2 \right] S_{\theta IN}(f_m); (f_m) > 0 \tag{12}$$

SSB Phase Noise

and

$$L(f_m) \equiv \frac{\text{noise power (in 1 Hz) with instrument corrections at } f_m \text{ offset from the carrier } f_c}{\text{carrier power}} = \frac{N}{C} \tag{13}$$

Equation 12 shows two possible ways to decrease an oscillators output phase noise  $L_{OUT}(f_m)$ . Basically, we can decrease  $S_{\theta IN}(f_m)$  or increase the loaded-Q,  $Q_L$  of the tank circuit. In the case of the SA620, only the latter is feasibly controllable since  $S_{\theta IN}(f_m)$  is a property of the internal circuitry and, as already noted, places a lower attainable bound on  $L_{OUT}(f_m)$  as  $Q_L \Rightarrow \infty$ .

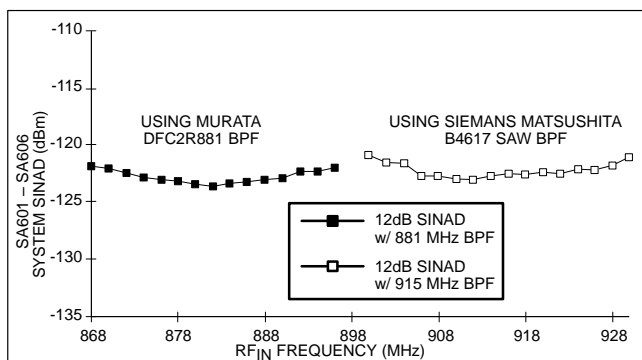


Figure 59. SA601 - SA606 System 12dB SINAD vs RF<sub>IN</sub> Frequency

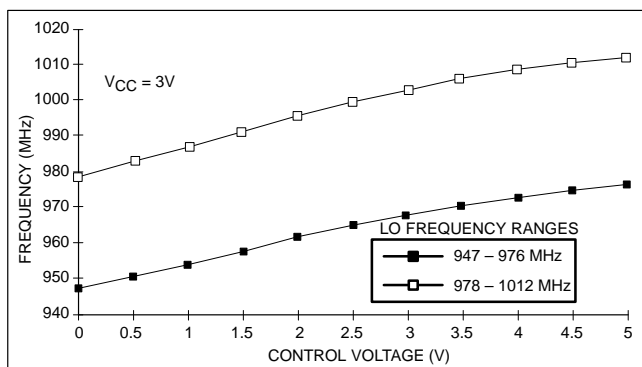


Figure 60. LO Frequency vs Control Voltage SA620 VCO

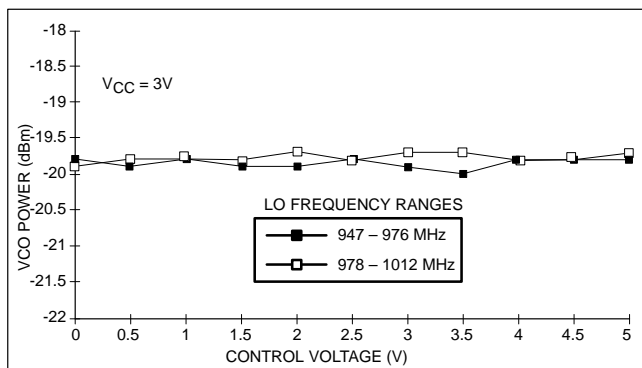


Figure 61. VCO Power vs Control Voltage SA620 VCO

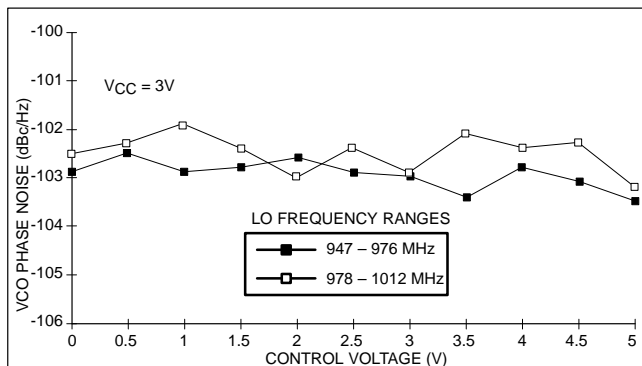


Figure 62. VCO Phase Noise vs Control Voltage SA620 VCO

# Low voltage front-end circuits: SA601, SA620

AN1777

## Increasing Loaded-Q

Figure 64 shows the oscillator section of the SA620 with a conventional 2nd order parallel tuned tank circuit configured as the external resonator.

From the basic equations for parallel resonance, the resonator's loaded-Q is given by

$$Q_L = R_P \left[ \frac{C_T}{L_T} \right]^{(1/2)} \tag{14}$$

where

$$R_P = R_T \parallel R_C = \left[ \left( \frac{1}{R_T} \right) + \left( \frac{1}{R_C} \right) \right]^{-1} \tag{15}$$

represents the net shunt tank resistance appearing across the network at resonance.  $R_T$  represents losses in the inductance  $L_T$  and capacitance  $C_T$  (almost always dominated by inductor losses), and  $R_C$  is the active circuits load impedance at resonance.

Improving the quality (i.e., their Q) of the tank components,  $L_T$  and  $C_T$  will increase  $R_T$ , but since  $R_C$  is low in the case of the SA620, the resulting increase in  $R_P$  is relatively small. We can increase  $R_P$  by decoupling  $Z_C$  from the tank circuit (note that  $R_C$  is the real part of  $Z_C$  at resonance). Decoupling this impedance by using either tapped-L or tapped-C tank configurations is possible. Inspection of the circuit shows that DC biasing is necessary for Pins 9 and 10 (osc1 and osc2, respectively), so a tapped-C approach would require shunt-feeding these pins to  $V_{CC}$ . Thus, the most practical way is to employ a tapped-L network.

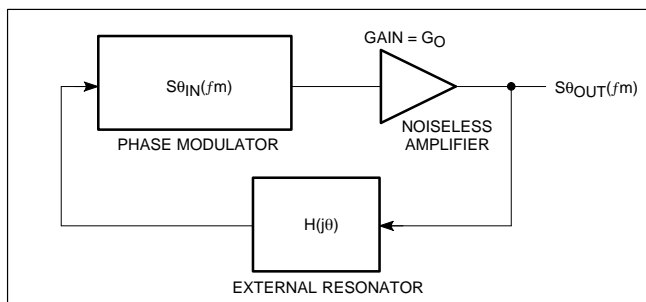


Figure 63. Oscillator Phase Noise Model

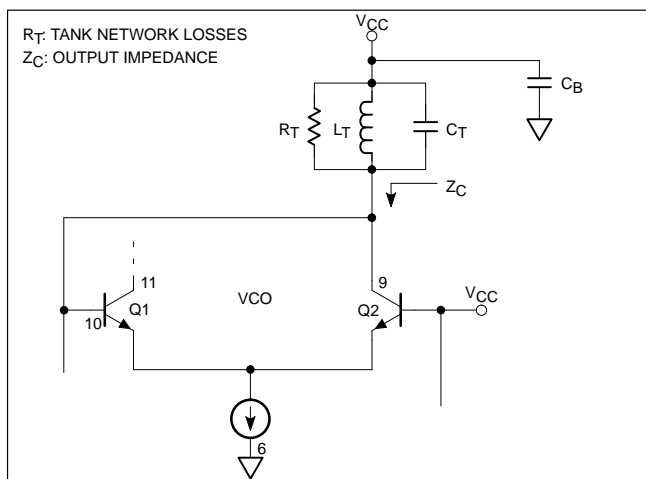


Figure 64. SA620 With External Resonator

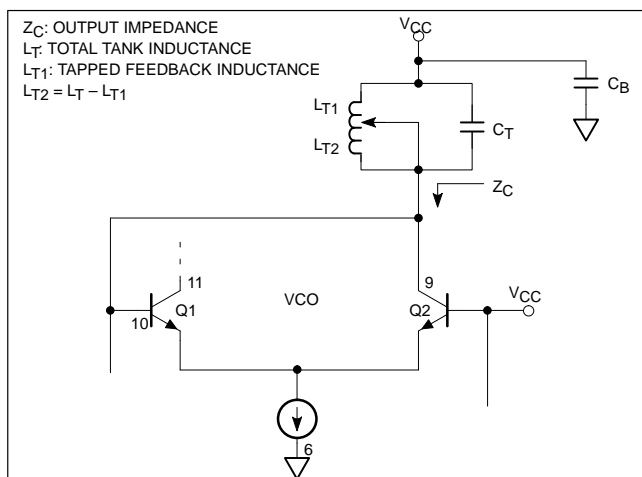


Figure 65. SA620 With Tapped-L Resonator

Figure 65 shows the basic tapped-L circuit. The effective multiplied shunt impedance at resonance across  $C_T$  is given by

$$\hat{R}_P = R_P \left[ \left( \frac{L_{T2}}{L_{T1}} \right) + 1 \right]^{1/2} \tag{16}$$

with a corresponding increase in loaded Q

$$\hat{Q}_L = Q_L \left[ \left( \frac{L_{T2}}{L_{T1}} \right) + 1 \right]^{1/2} \tag{17}$$

Feedback is caused by an RF voltage appearing across  $L_T$ , coupled through bypass capacitor  $C_B$  and its ground return to the cold end of the current source at Pin 6. The significance of this path increases with frequency. At 900MHz it becomes very critical to obtaining stable oscillations and must be kept as short as possible. Attention to layout detail cannot be overlooked here!

## High-Q Short $\mu$ -Strip Inductor

Realizing a stable tapped-L network using lumped surface-mount inductors is impractical due to their low unloaded-Q and finite physical size. Another approach utilizes a high-Q short microstrip inductor. The conventional approach to microstrip resonators treats them as a length of transmission line terminated with a short which reflects back an inductive impedance which simulates a lumped inductor. The approach presented here deviates from this technique by specifically exploiting the very high unloaded-Q attainable with short microstrip inductors where we design for a large C/L ratio by making the microstrip a specific but short length. Resonance is achieved by replacing the short with whatever capacitance is needed for proper resonance. One equation for the inductance of a strip of metal having length  $l$  (mil) and width  $w$  (mil) is

$$L = [5.08E - 3] l \left[ \ln(l/w) + 1.193 + 0.224(w/l) \right] \text{ [nH/mil]} \tag{4} \tag{18}$$

This technique results in a much shorter strip of metal for a given inductance. One intriguing property of microstrip inductors is that they are capable of very high unloaded-Q's. An equation describing the quality factor for a "wide" microstrip line is

$$Q_C = 0.63h \left[ \sigma f_{GHz} \right]^{1/2} \tag{5} \tag{19}$$

where  $h$  is the dielectric thickness in centimeters,  $\sigma$  is the conductivity in [S/m] and  $f$  is frequency in GHz. This equation predicts an unloaded-Q exceeding 700 when silver-coated copper is used. Also, wide microstrip lines are defined as those whose strip width  $w$  to height  $h$  ratio is approximately or greater than 1, i.e.  $w/h > 1$ . The parallel capacitor formed by the metal strip over the

## Low voltage front-end circuits: SA601, SA620

## AN1777

ground plane should also be included and may be calculated by the classic formula and included in the total needed to resonate with the inductance result found from equation (18). This "strip" capacitance is given by

$$C_S = K\epsilon_0\epsilon_r [w/h] \text{ [Farad]} \quad (20)$$

where  $\epsilon_0 = 8.86E-12$  [F/cm] is the permittivity of free space,  $r$  is the relative dielectric constant of the substrate. A "fringe factor"  $K$  is included to account for necessary fringing (est 2% to 15%). These equations are meant to provide insight into circuit behavior and should, therefore, be applied cautiously to specific applications.

### UHF VCO Using the SA620 at 900MHz

The basic electrical circuit involving the SA620 is shown in Figure 66. Feedback occurs when sufficient RF voltage develops across the tap inductances  $L_{T1}$  and  $L_{T2}$  (due to tap-1 and tap-2 respectively); note that inductance is reckoned from the cold end of the tank at node A. Taps 1 and 2 should be as close as possible to Pins 10 and 9, respectively. Also, nodes A (cold end of tank) and B (Pin 6) must be as physically close together and exhibit as low an impedance as possible to discourage parasitic oscillations. This "inner-loop" composed of  $L_{T2}$ , inductance of taps 1 and 2 connecting to Pins 10 and 9, and stray low Q inductance between nodes A and B create a parasitic loop that will favor oscillations that no longer depend on the full tapped-L tank circuit. This low-Q loop will exhibit very poor phase noise and typically oscillate well above 1200MHz; it has been observed as high as 1600MHz.

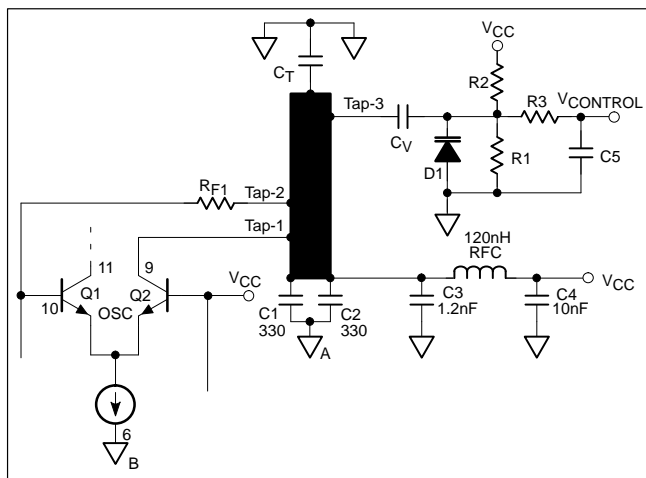


Figure 66. SA620 with Tapped-L  $\mu$ -Strip Resonator

Another factor also affects this unwanted parasitic oscillation. As the loaded-Q of the tapped tank circuit decreases, the circuit becomes conditionally stable and will eventually favor the parasitic loop exclusively. This occurs because tapped-L loaded-Q affects the magnitude of the RF voltage appearing across the entire microstrip tank. Thus, feedback at taps 1 and 2 is reduced as the Q decreases. This increases the likelihood of the inner-loop parasitic controlling the oscillator, since its feedback voltage is largely independent of the  $\mu$ -strip tank Q.

Equation 19 shows that microstrip inductors are capable of very high unloaded-Q's. This depends on a number of physical factors: frequency, dielectric thickness and its quality, and the skin depth conductivity of the metal strip itself. For example, at 900MHz, when coated with silver, unloaded-Q is somewhere about 750; when coated with lead (or sloppy soldering!) it drops to less than 230. If

the unloaded-Q is too low, the loaded-Q also drops and the circuit becomes conditionally stable, and now will either oscillate at the higher parasitic inner-loop or the wanted tapped-L lower complete tank circuit frequency. Components should be connected by narrow traces closely connected to component pads to avoid soldering on the microstrip layer.

The addition of  $R_{F1}$  introduces necessary losses to control the inner-loop parasitic oscillation. Its size should be made as small as possible consistent with stability, startup and good phase noise performance. Since it also decreases feedback, LO injection falls off as it increases. Typically, experimental values from  $4\Omega$  to  $30\Omega$  have proved sufficient. Stability as a function of  $V_{CC}$  is a good way to assess conditional stability. Provided the microstrip tank has a high loaded-Q, stability should be independent of  $V_{CC}$  down to less than 2.5V, or so. When loaded-Q decreases, so as to favor the inner-loop parasitic, conditional stability will occur. This can readily be observed by increasing  $V_{CC}$  beginning at about 1.0V and noting whether the VCO jumps back and forth unpredictably. Mixer and LO port loading also affect this condition and should be considered.

Dimensions for the microstrip resonator are based on several criteria. The strip must be "wide": its width to height must be on the order of 1 or greater. Minimum strip length seems to be about 200 mils for a 62 mil thick board. L/C ratios somewhere about 500 have yielded quite good results. Note that we usually specify the "L/C" ratio because it is always greater than 1, even though in a parallel resonant circuit it appears in the Q equation as C/L. Thus, decreasing the "L/C ratio" by making the microstrip shorter helps the loaded inductor Q, and the circuit loaded- $Q_L$  increase (see Equation 14). However, the effect of the lower Q of  $C_T$ , even when using a "hi-Q" SM type, decreased the expected larger increase in the net loaded circuit Q. Thus, the expected increase in  $Q_L$  may not be fully realizable. Assuming a 62 mil thick board with FR5 dielectric, a strip 50 by 300 mils gave very good experimental results where  $C_T$  was about 4.3pF with oscillations occurring from about 950 to 1000MHz for various test boards. A shorter inductor designed to increase the C/L ratio requiring 7.5pF experimentally resulted in only about 1-2dB improvement at 950MHz.

Tap-1 may be anywhere between the cold end of the tank (where  $C_1$  and  $C_2$  are located) and tap-2. Tap-2 yields good results at about 1/3 the strip length. Making it too close to the cold end in an effort to increase  $Q_L$  will result in loss of control over the inner-loop parasitic. To keep  $Q_L$  as large as possible,  $C_T$  should be a Hi-Q SM type. Differences greater than 5dB in SSB phase noise have been observed between a generic NPO SM and Hi-Q NPO SM capacitor.

Tap-3 can be at the end of the microstrip where  $C_T$  is connected. However, the varactor inevitably will cause a decrease in overall loaded-Q, typically by as much as 5dB. Moving it back some distance from the high end of the tank will decrease this effect and yield better results. A good starting point is about 1/4 back or 3/4 of the length from the cold end.  $C_1$  and  $C_2$  are paralleled lower value capacitors to yield a better low-impedance ground return to Pin 6 (node B) since relatively large RF currents flow through them. Note that as  $Q_L$  increases the peak circulating RF current will also increase, as will the RF voltage at the high end of the microstrip. At 900MHz good results have been obtained when both are about 330pF. RFC was chosen to be approximately series-resonant around 900MHz and constitutes the series feed path for DC biasing. It may be possible to neglect this component entirely, provided the PCB connection to the cold end of the tank is very close to RF ground. Finally, note that shielding of the entire microstrip may be necessary to meet part 15 emission limitations (where applicable).

# Low voltage front-end circuits: SA601, SA620

AN1777

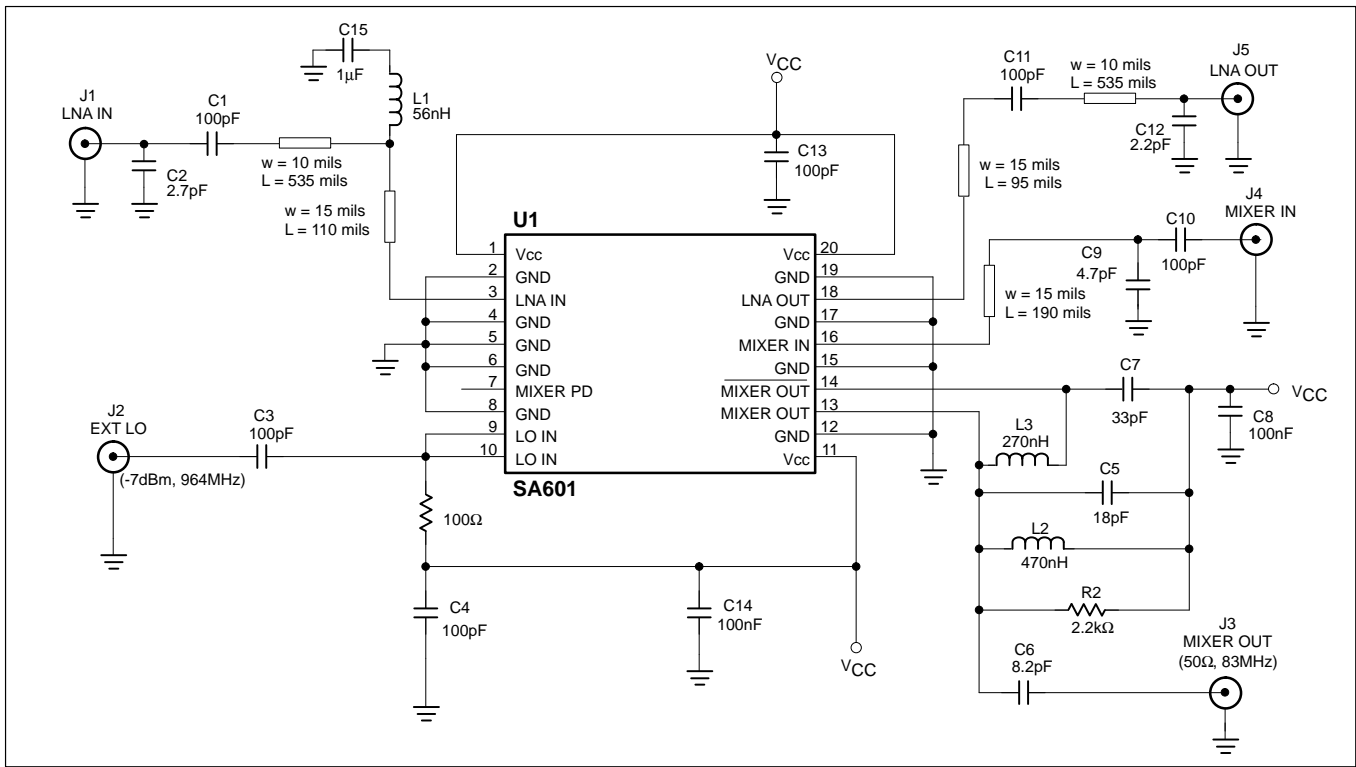
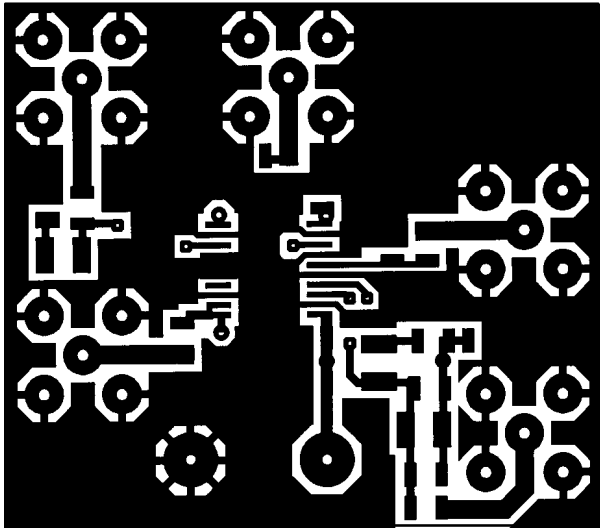


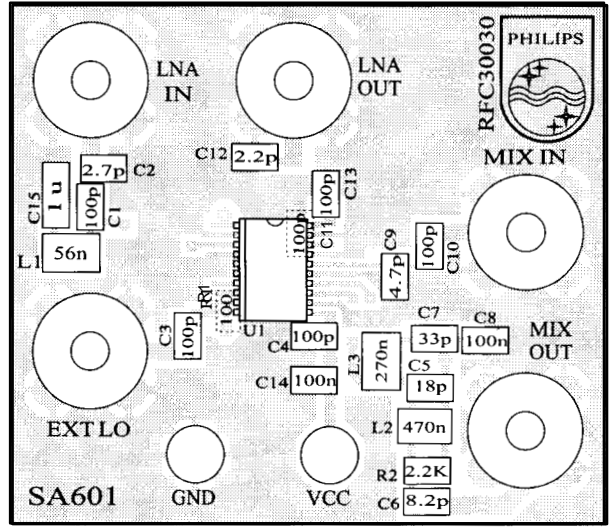
Figure 67. SA601 Application Board Schematic

Low voltage front-end circuits: SA601, SA620

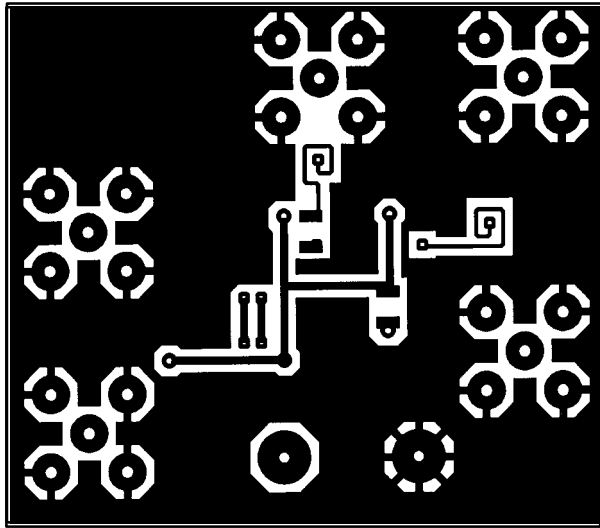
AN1777



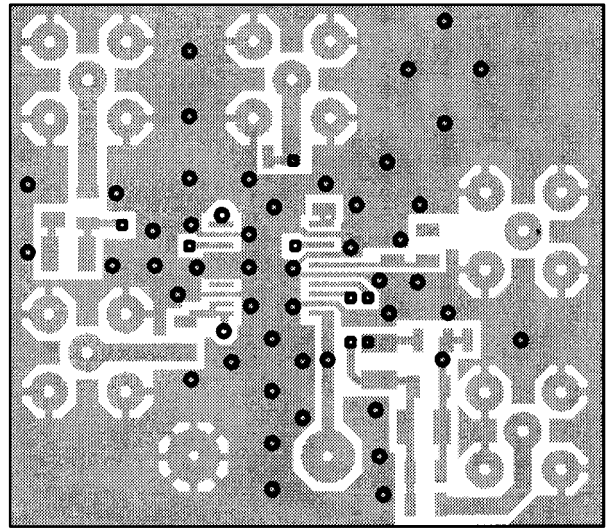
Top View



Silk Screen



Bottom View



Via Layer

Figure 68. SA601 Demoboard Layout (Not Actual Size)



## Low voltage front-end circuits: SA601, SA620

AN1777

Table 5. Customer Application Component List for SA601

Qty.	Part Value	Volt	Part Reference	Part Description	Vendor	Mfg	Part Number
<b>Surface Mount Capacitors</b>							
1	2.2pF	50V	C12	Cer Cap 0805 NPO $\pm$ .25pF	Garrett	Philips	0805CG229C9BB0
1	2.7pF	50V	C2	Cer Cap 0805 NPO $\pm$ .25pF	Garrett	Philips	0805CG279C9BB0
1	4.7pF	50V	C9	Cer Cap 0805 NPO $\pm$ .25pF	Garrett	Philips	0805CG479C9BB0
1	8.2pF	50V	C6	Cer Cap 0805 NPO $\pm$ 0.5pF	Garrett	Philips	0805CG829C9BB0
1	18pF	50V	C5	Cer Cap 0805 NPO $\pm$ 5%	Garrett	Philips	0805CG180J9BB0
1	33pF	50V	C7	Cer Cap 0805 NPO $\pm$ 5%	Garrett	Philips	0805CG330J9BB0
6	100pF	50V	C1, C3, C4, C10, C11, C13	Cer Cap 0805 NPO $\pm$ 5%	Garrett	Philips	0805CG101J9BB0
2	0.1 $\mu$ F	50V	C8, C14	Cer Cap 0805 Z5U $\pm$ 20%	Garrett	Philips	0805E104M9BB0
1	1 $\mu$ F	25V	C15	Cer Cap 1206 Y5V $\pm$ 20%	Garrett	Rohm	MCH312F105ZP
<b>Surface Mount Resistors</b>							
1	100 $\Omega$	50V	R1	Chip Res. 1/16W 0603 $\pm$ 5%	Garrett	Rohm	MCR10JW101
1	2.2k $\Omega$	50V	R2	Chip Res. 1/16W 0603 $\pm$ 5%	Garrett	Rohm	MCR10JW222
<b>Surface Mount Inductors</b>							
1	56nH		L1	1008 size chip inductor $\pm$ 10%	Coilcraft	Coilcraft	1008HS-560XKBB
1	270nH		L3	1008 size chip inductor $\pm$ 10%	Coilcraft	Coilcraft	1008HS-271XKBB
1	470nH		L2	1008 size chip inductor $\pm$ 10%	Coilcraft	Coilcraft	1008HS-471XKBB
<b>Surface Mount Integrated Circuit</b>							
1		3V	U1	RF low noise amplifier / mixer	Philips	Philips	SA601DK
<b>Miscellaneous</b>							
5				SMA gold connector	Newark	EF-Johnson	142-0701-801
2				Terminal	Newark	Cambion	160-1558-02-01
1				Printed Circuit Board	Excel	Philips	SA601 - RFC30030
<b>29 Total Parts</b>							

## XIII. APPLICATION BOARDS

## SA601 Applications Board

Figure 67 shows the schematic of the current application board for the SA601.

The functional description of each pin and its associated external circuitry is summarized below.

**Pins 1, 11 and 20** are all  $V_{CC}$  supply pins. Capacitor  $C_{13}$  is for decoupling purposes.

**Pins 2, 4, 5, 6, 8, 12, 15, 17 and 19** are all ground connections and should be tied to a common ground plane and as close to the chip as possible.

**Pin 3** is the LNA mixer input pin. The LNA input has an associated network for the purpose of optimizing the return loss while minimizing the degradation of the noise figure. This network is composed of Capacitor  $C_2$ ,  $L_1$  and the 535 mil spiral inductor. Capacitor  $C_1$  is a DC blocking cap.

Capacitor  $C_{15}$  and inductor  $L_1$  also provide the LNA with voltage compensation.

**Pin 7** is the mixer power-down pin. It will disable the mixer if set low. The application board simply leaves this pin open.

**Pin 9 and Pin 10** are both connected to the local oscillator input. It is important to have good return loss at this pin to allow small LO drive-levels to be used. Capacitors  $C_4$  and  $C_{14}$  are decoupling caps.  $C_3$  is a DC blocking cap.

**Pin 13 and Pin 14** are the differential mixer outputs. Inductor  $L_3$  and capacitors  $C_5$  and  $C_7$  comprise the differential-to-single-ended

translation circuit which combines the output currents and forces them to be in phase with each other. This circuit is extensively described in the previous discussions.  $R_2$  sets the output impedance of the mixer output and  $L_2$  is the DC bypass inductor which increases the gain of the open-collector output. Capacitor  $C_6$ , in conjunction with  $L_2$ , is used to match the mixer output port to 50 $\Omega$ .

**Pin 16** is the mixer input pin. Capacitor  $C_9$  is used to optimize return loss and noise figure. Capacitor  $C_{10}$  is a DC blocking cap.

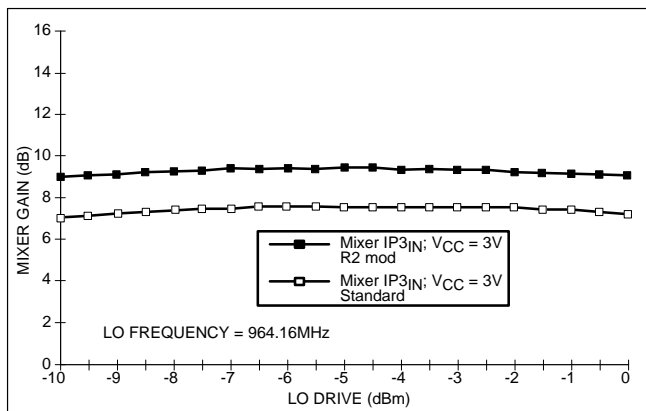
**Pin 18** is the LNA output pin. Capacitor  $C_{12}$  in conjunction with the 535 mil spiral inductor comprise a network for optimizing return loss and obtaining best noise figure.  $C_{11}$  is a DC blocking cap.

## SA601 Application Board Modification For Increasing Mixer Gain

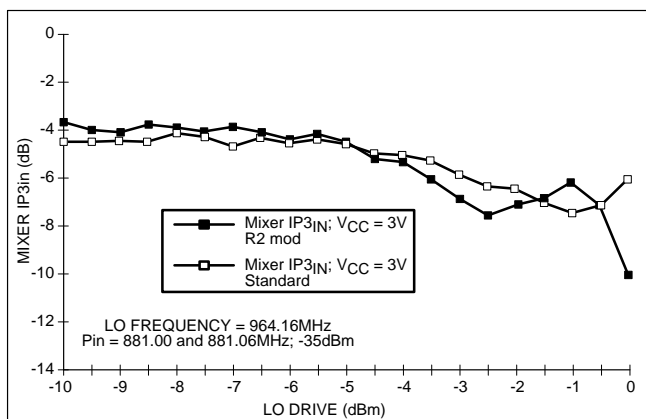
The SA601 application board can be modified by disconnecting the 2.2k $\Omega$  resistor R2 in Figure 67 and reconnecting it in parallel with C7. Figure 69 shows a comparison of the mixer gain with and without this modification. This modification yields a 2dB improvement in mixer conversion gain for LO drive-levels from -10dBm to 0dBm. Figure 70 shows a comparison of the mixer IP3<sub>IN</sub> with and without this modification over the same LO drive-level range as in Figure 69. This data shows that the IP3<sub>IN</sub> performance of the mixer is not degraded for LO drive-levels less than or equal to -5dBm. Some degradation does occur for LO drive-levels above -5dBm. Laboratory comparison data of noise figure performance also showed that over the LO drive-level range of -10dBm to 0dBm the modification of  $R_2$  yielded a decrease in noise figure of approximately 0.3dB. The  $R_2$  modification does draw about 0.3mA more supply current, however.

# Low voltage front-end circuits: SA601, SA620

# AN1777



**Figure 69. SA601 Mixer Gain vs LO Drive Standard/R2 Modification Comparison**



**Figure 70. SA601 Mixer IP3IN vs LO Drive Standard/R2 Modification Comparison**

## SA620 Applications Board [7]

Figure 71 shows the schematic of the current application board for the SA620. The functional description of each pin and its associated circuitry is summarized below.

**Pin 1** is the LNA enable pin. This pin is used for selecting the gain in the LNA path.

The logic levels for the SA620 LNA enable are specified as:

Logic "1" level (LNA on): +2.0V to  $V_{CC}$   
 Logic "0" level (LNA off): -0.3 to +0.8V

Referring to the application board schematic, resistor  $R_3$  and capacitor  $C_{25}$  are used to improve LNA enable switching time. If fast switching is not a requirement, or if the LNA does not need to be disabled, these two components can be eliminated.

**Pins 2, 4, 5 and 19** are ground connections. These pins provide the RF ground path for the LNA and should be tied to a common ground plane as close to the IC, and each other, as possible.

**Pin 3** is the LNA input. The LNA input exhibits a return loss of 7dB at 900MHz. However, the designer can implement an external matching circuit to further improve the match. For this reason,  $S_{11}$  plots of the LNA input impedance measured right at Pin 3 are provided in the SA620 data sheet.

Referring again to the application board schematic, the 260 mil long transmission line, inductors  $L_1$ , the 4.7nH spiral inductor, and capacitor  $C_2$  are the matching elements. Capacitor  $C_1$  is a DC blocking capacitor. Conjugately matching the LNA input for optimum gain performance will degrade the noise figure slightly. Typically, 1.6dB noise figure can be achieved when the return loss is improved to 10dB.

**Pin 6 and Pin 12** are the oscillator ground connections. These pins provide the RF ground path for the oscillator section of the SA620 and should be tied to a common ground plane as close to the IC, and each other, as possible.

**Pin 7 and Pin 8** are the mixer and oscillator power down pins, respectively. These pins can be used to individually power down the mixer and the oscillator sections of the SA620.

The DC levels at Pins 7 and 8 are approximately 2.3V and 1.5V, respectively, and these levels must be established by the IC and not the power down circuitry. Placing a Schottky diode ( $V_f = 0.3V$ ) between each power down pin and the LNA ENABLE pin will allow the designer to power-down the IC via a single control voltage, the LNA enable, while allowing the voltages at Pins 7 and 8 to float at the IC levels when in the power-up mode.

**Pin 9 and Pin 10** drive the base of one and the collector of another of the broadband VCO differential pair input stage transistors as shown in Figure 64. Components  $C_5, C_6, C_7, C_8, C_9, C_{10}, C_{11}, L_2, R_1, R_2, R_3, R_4, D_1$  and the 300 mil  $\mu$ -strip comprise the resonant tank circuit described in detail above and shown in Figure 66.

**Pin 11** is the open-collector VCO output and is provided to couple out VCO energy for frequency synthesis for example. Resistor  $R_5$  is chosen to be small so as not to excessively load down the VCO. Hence, only about -20dBm of VCO fundamental output power will be measured at the VCO OUT connector on the application board. Note that the second harmonic power level will actually be slightly higher than that of the fundamental. This is not a problem because this isn't the same signal that is driving the mixer, and the mixer develops its own second harmonic as a result of the mixing process anyway. The 4.7nH spiral inductor and capacitor  $C_{12}$  on the application board are the 50 $\Omega$  matching elements. Capacitor  $C_{13}$  is a decoupling capacitor.

**Pin 13** is the open-collector mixer output. Referring to the application board schematic,  $R_6$  establishes the output impedance to be 1k $\Omega$ . Inductor  $L_3$  and capacitors  $C_{15}$  and  $C_{16}$  are 50 $\Omega$  matching elements.

**Pin 14** is the mixer bypass pin. This is not a signal output. It is provided for designers to optimize the IP3 performance of the mixer circuit. If you look closely at the application board schematic, you will notice that the two circuit elements at Pin 14, the 2.5 - 6.0pF capacitor and the 4.7nH spiral inductor, are very small impedances at the 83MHz IF frequency. The intent is to reflect RF leakage energy back into Pin 14 to achieve phase cancellation of this energy inside the IC, thus reducing the level of third-order intermodulation products. While looking at the third-order intermodulation products on a spectrum analyzer, the designer can tune with  $C_{17}$  to minimize the level of these products at the IF output. To obtain the best IP3IN performance over the entire frequency range of the VCO the following procedure is recommended:

1. Set the VCO control voltage to its mid-range value of 2.5V.
2. Adjust capacitor  $C_9$  to obtain the desired mid-range frequency,
3. Tune  $C_{17}$  for best IP3IN .

# Low voltage front-end circuits: SA601, SA620

AN1777

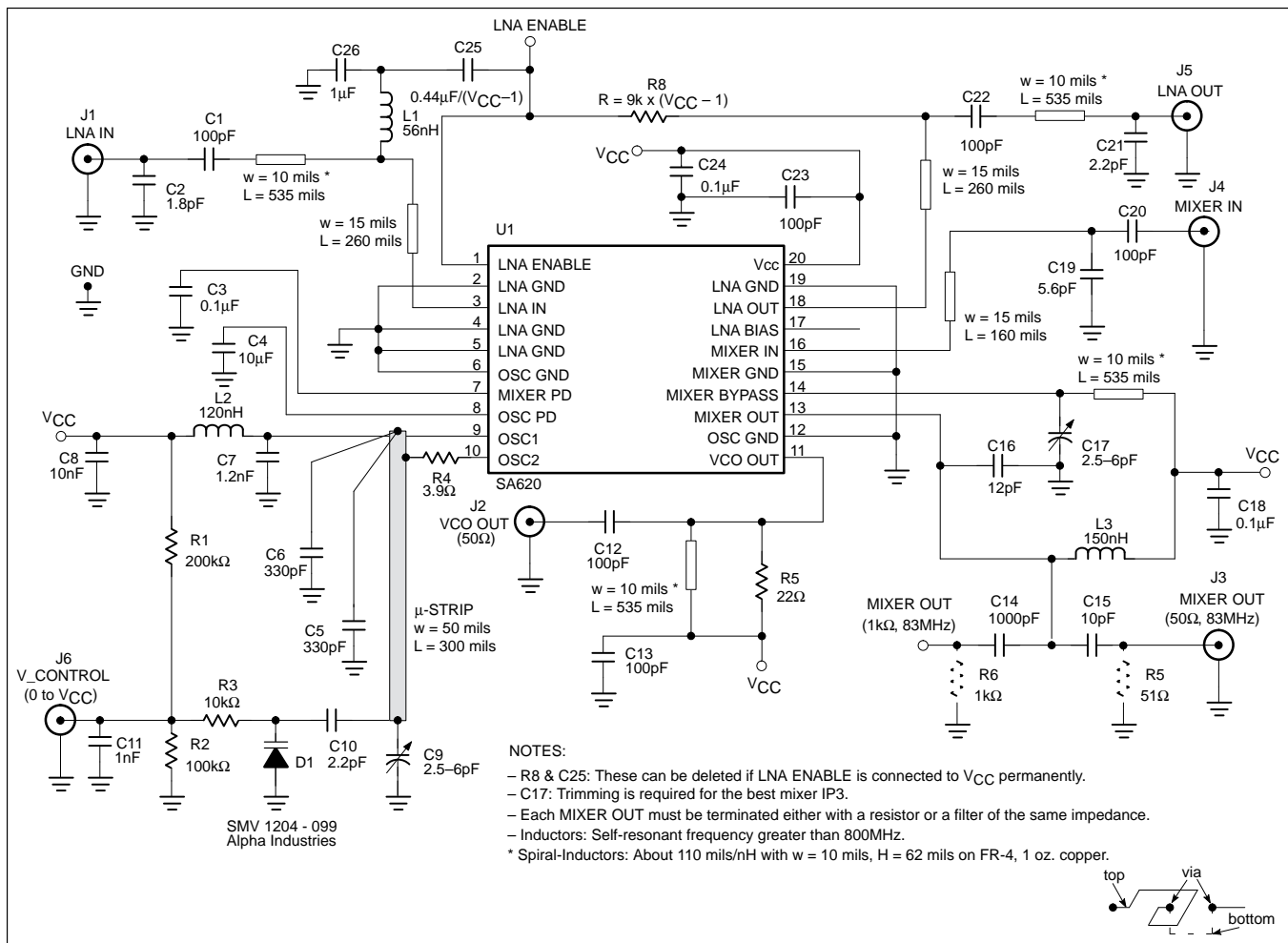


Figure 71. SA620 Application Board

**Pin 15** is the mixer ground connection. This pin provides the RF ground path for the mixer section of the SA620 and should be tied to a common ground plane as close to the IC as possible.

**Pin 16** is the mixer input pin. The mixer input is not matched to 50Ω, therefore, external matching elements are required to achieve optimum performance. The 160 mil long transmission line and capacitor C19 are the matching elements on the application board. C20 is a DC blocking capacitor. If no matching elements are used, the IP3 performance is typically maximized at +3dBm at the expense of gain, which will be reduced by 3dB.

**Pin 17** is the LNA bias pin. This is a DC check pin for use during

the manufacture of the SA620 only. Customers should float this pin in their designs.

**Pin 18** is the LNA output pin. The LNA output has an intrinsic return loss of approximately 9dB at 900MHz. However, to achieve optimum performance from the SA620, the designer can implement an external matching circuit to further improve the match. For this reason, S<sub>22</sub> plots of the LNA output impedance measured right at Pin 18 are provided in the SA620 data sheet. Referring again to the application board schematic, the 260 mil long transmission line, the 4.7nH spiral inductor, and capacitor C21 are the matching elements. Capacitor C22 is a DC blocking capacitor.

**Pin 20** is the DC power input to the SA620 where capacitors C23 and C24 are strictly for decoupling purposes.

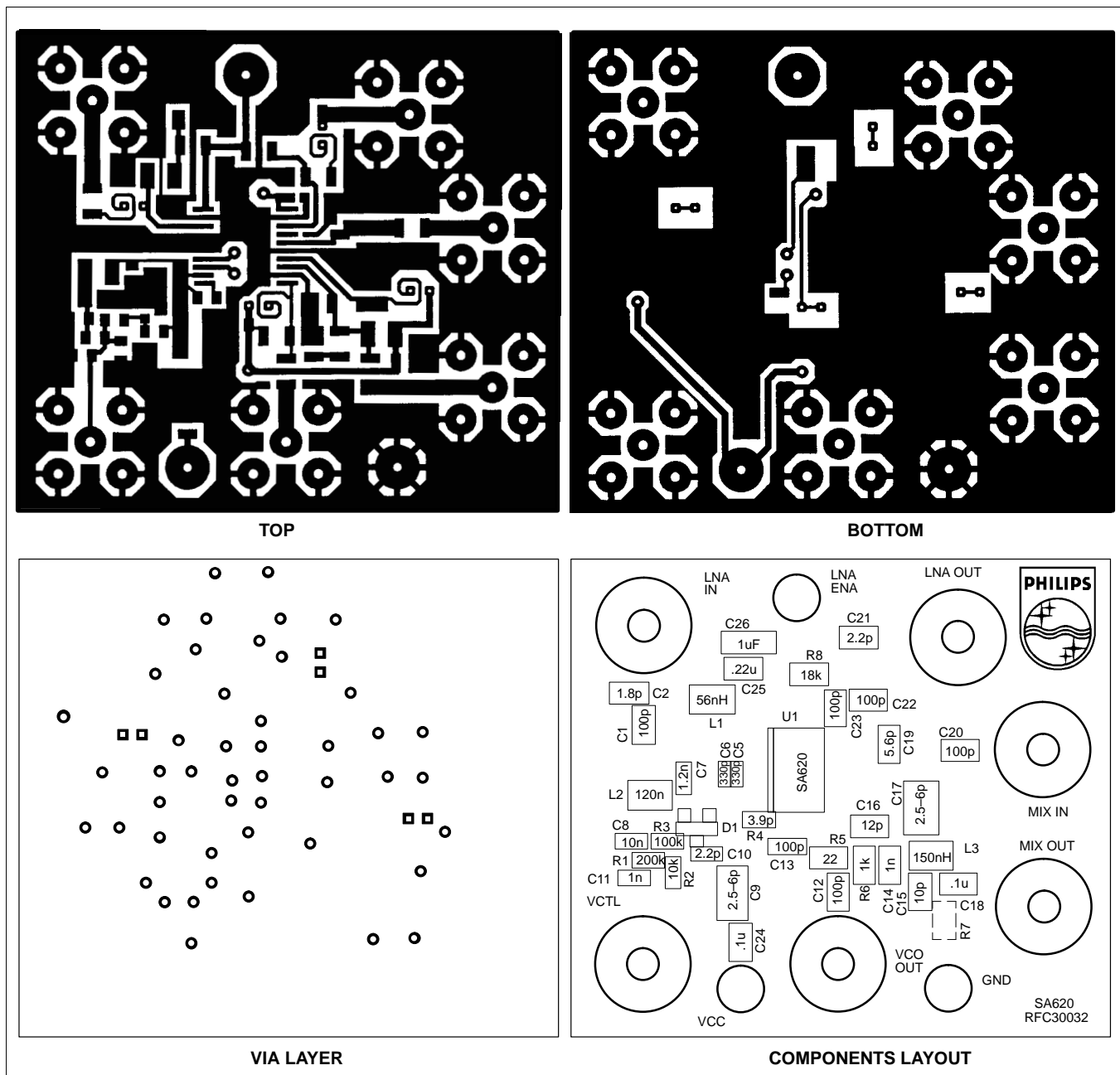


Figure 72. SA620 Demoboard Layout (Not Actual Size)

**XIV. TEST AND MEASUREMENT TIPS**

**Noise Figure and Gain**

The LNA of both the SA601 and SA620 can be tested identically. The gain and noise figure can be measured on a noise figure meter such as the HP8970. The correct measurement mode on this particular piece of equipment is obtained by selecting special function 1.0. It is important to do all noise figure measurements within a screen room to ensure the receiver of the noise figure meter is not picking up any stray signals. The LNA gain can also be measured on a spectrum analyzer. For accurate measurements, be sure to properly compensate for losses in the cables and connectors used. Also, make sure that the input power used is well below the

1dB compression point for the device.  $P_{IN} = -25\text{dBm}$  is a good value.

The SA601 and SA620 mixers require more precaution while testing. The correct measurement mode for mixer noise figure and gain measurements is obtained on the HP8970 by selecting special function 1.4. It should be stated again that all noise figure measurements should be done within a screen room. The noise source used in conjunction with the noise figure meter is generally a wideband noise source. So, for a given LO frequency there are two regions of the wideband noise power spectrum which will mix to the specified IF frequency. Thus, an image-reject bandpass filter should be placed between the noise source and the mixer input. In this way the unwanted input noise power will be rejected yielding proper

## Low voltage front-end circuits: SA601, SA620

AN1777

single sideband noise figure measurements. The loss through this image-reject filter should also be accounted for. Losses prior to the mixer input can be compensated for on the HP8970 by selecting special function 34.2. The noise figure of the mixers is highly dependent on the strength and quality of the LO signal. The LO signal should be passed through a high-pass filter. A 500MHz HPF is typically used. The noise figure is dependent on the LO drive-level. Generally, the noise figure will decrease as the LO

drive-level is increased. The noise figure has also been observed to show some dependence on the range setting of the signal generator. The high-frequency components present at the IF output of the mixer should also be filtered out prior to entering the input port of the noise figure meter. This is typically done with a 300MHz low-pass filter. If the conversion gain of the mixer is to be measured directly on a spectrum analyzer, be sure that the input power is well below the 1dB compression point. Again, -25dBm is a good value.

**Table 6. Customer Application Component List for SA620**

Qty.	Part Value	Volt	Part Reference	Part Description	Vendor	Mfg	Part Number
<b>Surface Mount Capacitors</b>							
1	2.2pF	50V	C10	Cer Cap 0603 NPO $\pm$ .25pF	Garrett	Rohm	MCH185A2R2CK
2	330pF	50V	C5, C6	Cer Cap 0603 NPO $\pm$ 5%	Garrett	Rohm	MCH185A331JK
1	1000pF	50V	C11	Cer Cap 0603 X7R $\pm$ 10%	Garrett	Rohm	MCH185C102KK
1	1200pF	50V	C7	Cer Cap 0603 X7R $\pm$ 10%	Garrett	Rohm	MCH185C122KK
1	0.01 $\mu$ F	25V	C8	Cer Cap 0603 X7R $\pm$ 10%	Garrett	Rohm	MCH185C103KK
1	1.8pF	50V	C2	Cer Cap 0805 NPO $\pm$ .25pF	Garrett	Philips	0805CG189C9BB0
1	2.2pF	50V	C21	Cer Cap 0805 NPO $\pm$ .25pF	Garrett	Philips	0805CG229C9BB0
1	5.6pF	50V	C19	Cer Cap 0805 NPO $\pm$ 0.5pF	Garrett	Philips	0805CG569C9BB0
1	10pF	50V	C15	Cer Cap 0805 NPO $\pm$ 5%	Garrett	Philips	0805CG100J9BB0
1	12pF	50V	C16	Cer Cap 0805 NPO $\pm$ 5%	Garrett	Philips	0805CG120J9BB0
6	100pF	50V	C1, C12, C13, C20, C22, C23	Cer Cap 0805 NPO $\pm$ 5%	Garrett	Philips	0805CG101J9BB0
1	100pF	50V	C14	Cer Cap 0805 NPO $\pm$ 5%	Garrett	Philips	0805CG102J9BB0
3	0.1 $\mu$ F	50V	C3, C18, C24	Cer Cap 0805 Z5U $\pm$ 20%	Garrett	Philips	0805E104M9BB0
1	0.22 $\mu$ F	50V	C25	Cer Cap 0805 Y5V $\pm$ 20%	Garrett	Rohm	MCH212F224ZK
1	10 $\mu$ F	10V	C4	Tant Chip Cap B 3528 $\pm$ 10%	Garrett	Philips	49MC106C010KOAS
1	1 $\mu$ F		C26	Cer Cap 1206 Y5V $\pm$ 20%	Garrett	Rohm	MCH312F105ZP
<b>Surface Mount Variable Capacitors</b>							
2	2–6pF		C9, C17	Trimmer capacitor	Murata	Murata	TZV02Z060A110
<b>Surface Mount Resistors</b>							
1	3.9 $\Omega$		R4	Chip res. 1/16W 0603 $\pm$ 5%	Garrett	Rohm	MCR03JW3.9
1	10k $\Omega$		R3	Chip res. 1/16W 0603 $\pm$ 5%	Garrett	Rohm	MCR03JW103
1	100k $\Omega$		R2	Chip res. 1/16W 0603 $\pm$ 5%	Garrett	Rohm	MCR03JW104
1	200k $\Omega$		R1	Chip res. 1/16W 0603 $\pm$ 5%	Garrett	Rohm	MCR03JW204
1	22 $\Omega$		R5	Chip res. 1/10W 0805 $\pm$ 5%	Garrett	Rohm	MCR10JW220
1	51 $\Omega$		R7	Chip res. 1/10W 0805 $\pm$ 5%	Garrett	Rohm	MCR10JW510
1	1k $\Omega$		R6	Chip res. 1/10W 0805 $\pm$ 5%	Garrett	Rohm	MCR10JW102
1	18k $\Omega$		R8	Chip res. 1/10W 0805 $\pm$ 5%	Garrett	Rohm	MCR10JW183
<b>Surface Mount Inductors</b>							
1	56nH		L1	Chip inductor 1008CS $\pm$ 10%	Coilcraft	Coilcraft	1008CS-560XKBB
1	120nH		L2	Chip inductor 1008CS $\pm$ 10%	Coilcraft	Coilcraft	1008CS-121XKBB
1	150nH		L3	Chip inductor 1008CS $\pm$ 10%	Coilcraft	Coilcraft	1008CS-151XKBB
<b>Surface Mount Integrated Circuit</b>							
1		3V	U1	Low voltage LNA & Mixer	Philips	Philips	SA620DK
<b>Miscellaneous</b>							
1			D1	Varactor	Wireless Components	Alpha Industries	SMV 1204-099
6				SMA gold connector	Newark	EF-Johnson	142-0701-801
3				Terminal	Newark	Cambion	160-1558-02-01
1				Printed Circuit Board	Excel	Philips	SA620 - RFC30032
<b>49 Total Parts</b>							

**1dB Compression Point**

This is determined by taking  $P_{OUT}$  vs.  $P_{IN}$  data at very low input power where the relationship between these quantities is linear. An

input power range between -35dBm and -30dBm is sufficient for both the LNA and mixer. The input power is then increased to the point where the difference between an extrapolation of the low input

# Low voltage front-end circuits: SA601, SA620

AN1777

vs output power line and the output power is 1dB. The input power at which this occurs is the 1dB compression point. A shorter and more practical method of obtaining this parameter is to adjust the input power at a low level such that the output power is some integer number. Then increase the input power in 1dB increments. The output should also increase in 1dB increments. When the output power becomes exactly 1dB less than the expected value, the input power at which this occurs is the 1dB compression point.

## Input Third-Order Intercept Point

$IP_{3IN}$  is typically measured on a spectrum analyzer by combining two input signals of equal power that are detuned by approximately 60kHz. This is then connected to the input of the device. Be sure to measure the output power of the two peaks after the power combiner because the two ports of the power combiner rarely attenuate equally.  $IP_{3IN}$  is read off the spectrum analyzer by measuring the difference between the two carrier peaks and the intermodulation peaks, then adding half this difference to the input power.

It is extremely important to take this data at very low input power because this calculation assumes linearity. Input power of -35dBm is sufficient.

## Phase Noise

The most economical way to measure phase noise is using the spectrum analyzer. As previously stated, this is done very simply by measuring the difference between the carrier frequency peak power and the power at some specified offset from the carrier.

Unfortunately, the spectrum analyzer does not have the capability of measuring this with a 1Hz resolution bandwidth. Thus, a calculation must be made as follows:

$$\text{Phase Noise (dBc/Hz)} = -(|\Delta| + 10\log(\text{RBW}))$$

where  $\Delta$  is the difference between the peak power at the carrier frequency and the power at the specified offset frequency, and RBW is the resolution bandwidth of the spectrum analyzer while taking the measurement.

## XV. COMMON QUESTIONS AND ANSWERS

- Q.** I'm not getting the Mixer Noise Figure that is stated in the databook. What could be wrong?
- A.** There are quite a few things that can affect your noise figure. The most important thing to do initially is to check your measurement setup by following the suggestions in the test and measurement section of this application note. Aside from that, maintaining a good match at the LO input is very important and using the same LO drive-level specified in the databook is also necessary.
- Q.** Does the SA620 VCO meet AMPS stringent phase noise requirements?
- A.** The SA620 VCO phase noise is typically about -102dBc/Hz @ 60kHz. This does not meet AMPS specifications of -110dBc/Hz @ 60kHz.

- Q.** Can the SA620 be used with an external VCO to get better phase noise performance?
- A.** Although possible, this is not recommended due to the higher output power of the external VCO causing LO leakage problems at the mixer output.
- Q.** Why was the current-combiner implemented with a C - L - C arrangement instead of its dual circuit approach?
- A.** The C - L - C arrangement has the advantage of being low-pass in nature. Thus, it supplies additional rejection of the higher frequency  $RF_{IN}$  and LO signals at the mixer output.
- Q.** Why am I not getting the  $IP_{3IN}$  values that are stated in the databook?
- A.** In addition to checking your test setup, you might want to check what the offset is between your two RF input signals. Some of the databook values are taken with an offset of 1MHz. Chances are that your offset is much smaller. For instance, AMPS alternate channel spec is only 60kHz.  $IP_{3IN}$  will decrease when this offset is decreased.
- Q.** What are the necessary components needed to ensure unconditional stability of the LNAs.
- A.** Capacitor  $C_{15}$  and inductor  $L_1$  on the SA601 schematic of Figure 67 are the necessary components needed to ensure stability. Without these components, the amplifier will generate spurs approximately 1.5MHz off the carrier at  $V_{CC} = 3V$ . The similar components on the SA620 schematic in Figure 71 are capacitor  $C_{26}$  and inductor  $L_1$ .

## XVI. REFERENCES

- 'RF Circuit Design', Bowick C., Indiana; Howard W. Sams & Co., 1982.
- 'Microelectronic Circuits', Sedra/Smith, Third Edition, Chicago, Saunders College Publishing, 1991.
- 'Analysis and Design of Analog Integrated Circuits', Gray P. and Meyer R., Second Edition, N.Y., John Wiley & Sons, 1984.
- 'Design of Amplifiers and Oscillators by the S-Parameter Method', Vendelin, George D.; Wiley, 1982.
- 'Microwave Devices and Circuits', 2nd Edition, Liao, Samuel Y.; Prentice Hall, 1985.
- 'A current-combiner Circuit for Better Mixer Conversion Gain', Lee, S.H., Wong, A.K., Wong, M.G., Philips Semiconductors Applications Document.
- 'Getting to know the SA620', Wong, M.G., Philips Semiconductors Applications Document.
- '900MHz  $\mu$ -Strip', S. C. Peterson P.E., Philips Semiconductors Applications Document
- Intellectual Property Document, Nov. 30, 1993, Docket # PHA 01243