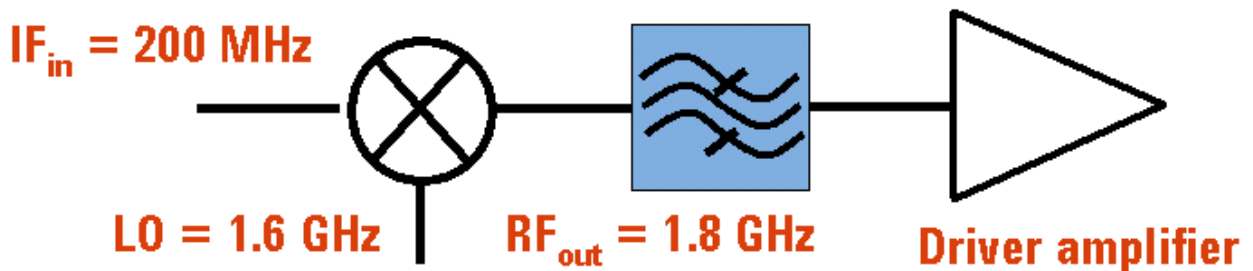


# Evaluating and Optimizing Tradeoffs in CMOS RFIC Upconversion Mixer Design

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It is not easy to design an RFIC mixer. Different, sometimes conflicting, performance specifications must be prioritized, depending on the application. Receive applications require a mixer to handle a wide range of input signal levels, and maximum linearity under large signal conditions is often more critical than the noise figure. Transmit applications have controlled signal levels, so the design strategy shifts to tradeoffs between noise and intermodulation distortion (IMD) behavior to achieve the largest useable dynamic range.

In this article, a transmit mixer is used to illustrate many of the design tasks that go into the development of a quality RFIC. The design is an upconversion mixer for a basestation transmitter (see Fig. 1) using a Gilbert-cell MOSFET double-balanced differential mixer with an input IF signal centered at 200 MHz and an output of 1.8 GHz. This design example uses 0.35  $\mu\text{m}$  MOSFETs with a default device model parameter set. Other users would substitute their own verified device models.



**Fig. 1: An Upconversion Mixer For A Basestation Transmitter Application**

The basic mixer performance of this example was evaluated, then the design was modified to improve conversion gain and image rejection by using a tuned mixer output. Finally, a differential-to-single-ended converter was added to provide the proper interface to an off-chip bandpass filter.

### The Mixer Circuit

The schematic of a MOSFET version of the Gilbert cell active double-balanced mixer is shown in Fig. 2. The lower FET differential pair serves as a transconductance amplifier, while the upper FETs operate as a fully balanced, phase-reversing current switch. Although not shown, a dc bias generator keeps the MOSFETs in their active region.

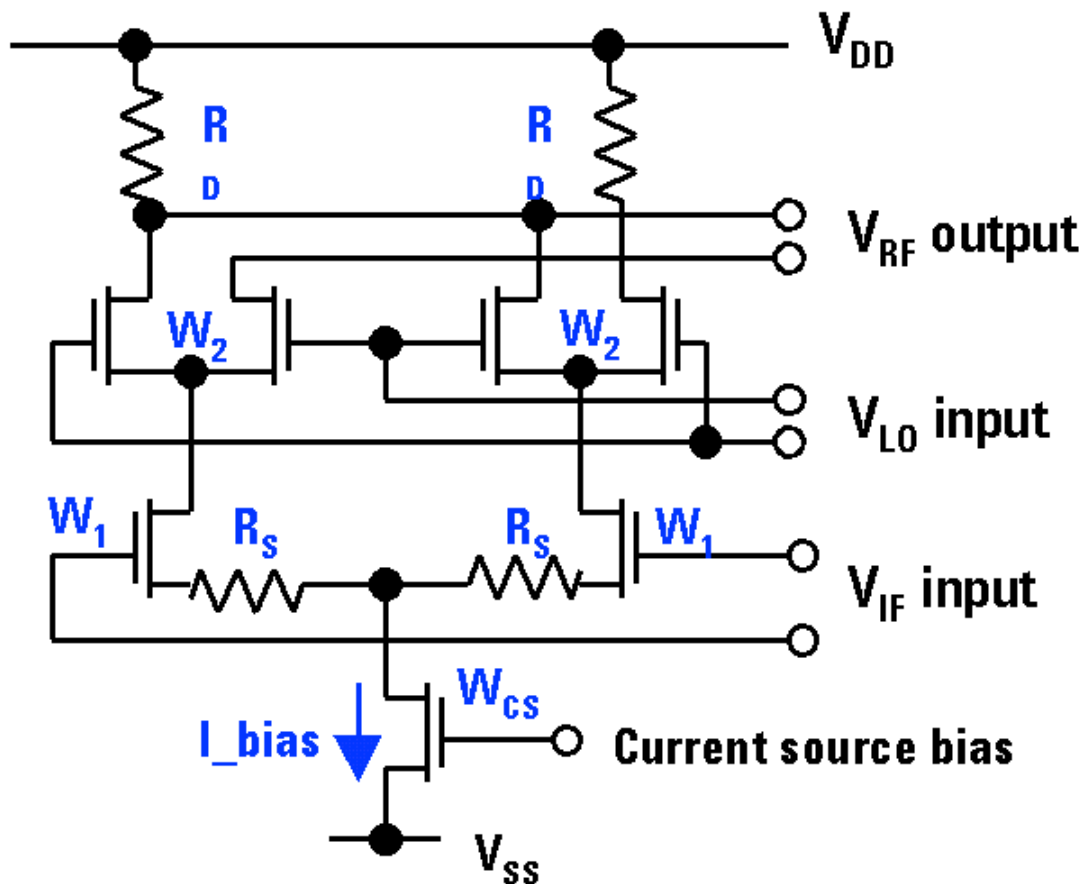


Fig. 2: The MOSFET Gilbert Cell Active Double-Balanced Mixer

The large signal performance of the mixer depends mainly upon the linearity of the transconductance amplifier, which is measured by determining the maximum input voltage,  $V_{1dB}$ , that causes a 1 dB compression in the conversion gain. In some cases it is appropriate to use power,  $P_{1dB}$ , instead of voltage. If the maximum linear input voltage range must be increased, this can be accomplished by increasing the values of the source-degeneration resistors ( $R_s$ ). Source inductance can also provide beneficial degeneration, but at the low frequency of 200 MHz, the required inductance values would be too large for RFIC implementation. Source resistors must be used, even though they add noise. The load resistors could also cause gain compression if the voltage swing at the drains is large enough to cause the output to clip under large signal drive conditions.

The double-balanced mixer design rejects IF and LO feedthrough to the output (as long as the output is taken differentially) because the LO component at the output is a common-mode signal and the RF output is differential.

## **The Design Procedure**

Basestation transmit applications require a mixer with high linearity and low noise to minimize the amount of spurious power that is spread into adjacent channels. The performance of this example mixer was optimized in the following sequence:

1. Determine the LO amplitude. The mixer commutating switch must be fully activated since a weakly conducting, or slowly activated, switch can create excess distortion. The conversion transducer gain and 1-dB gain-compression input level were used to determine when the LO voltage is sufficient.
2. Evaluate the influence of source and drain resistance on the 1-dB compression level. This information gives insight into the principal mechanisms that limit linearity.
3. Determine how the added noise of the mixer affects the minimum signal level, thus limiting dynamic range. This is necessary to evaluate the tradeoffs between noise, gain, and gain compression.
4. Evaluate how the two-tone third-order IMD power and the noise figure affect the mixer dynamic range relative to the input voltage. Since the designer has control over the input voltage in transmit applications, the optimum dynamic range, or performance “sweet spot,” must be determined. If a fixed signal level is specified, the mixer must be designed to provide the best dynamic range at that signal level.
5. Finally, test the mixer using a realistic signal input, such as a CDMA source, to emulate a multi-carrier environment. This is a more severe test than two-tone IMD, and is much more time consuming to simulate because a large number of symbols must be used for accurate results.

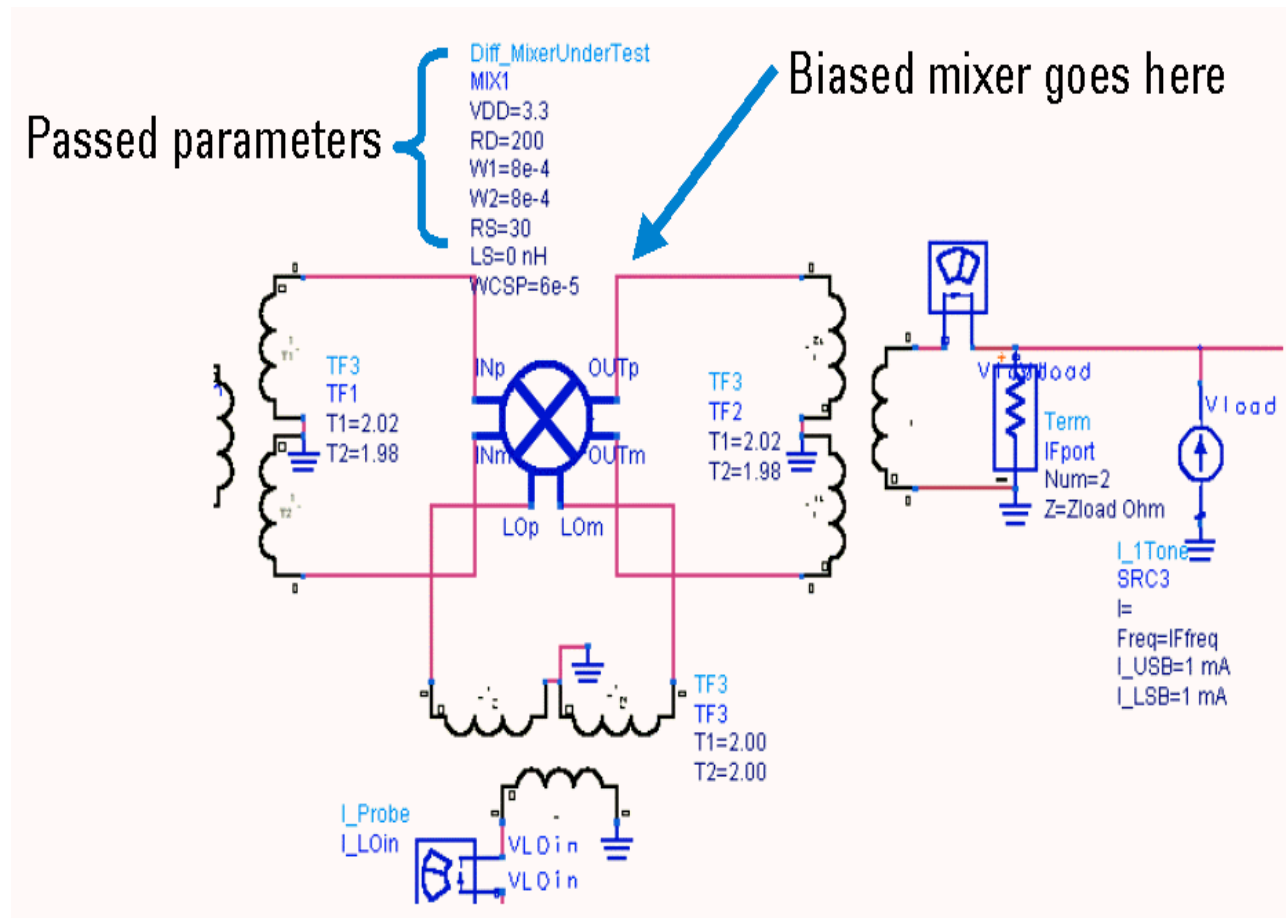
After the basic, resistively-loaded, Gilbert cell mixer was characterized, two modifications were employed to improve performance. First, the mixer drain nodes were tuned with inductors and a capacitor for resonance at the output frequency. This improves conversion gain if the IC process can fabricate inductors with reasonable Q. It also decreases the amplitude of the undesired output image signal because of its bandpass transfer function. The image must be removed anyway, and its presence can degrade the distortion of the output stage by increasing the peak voltage at its input.

Next, a stage was added to convert the differential signal to single-ended. Because the output of the mixer must be filtered off-chip with a SAW filter – before further amplification – a single-ended output is more efficient. This circuit must have good common-mode rejection to suppress LO feedthrough and good linearity so that it doesn’t degrade dynamic range.

### **Determining LO voltage**

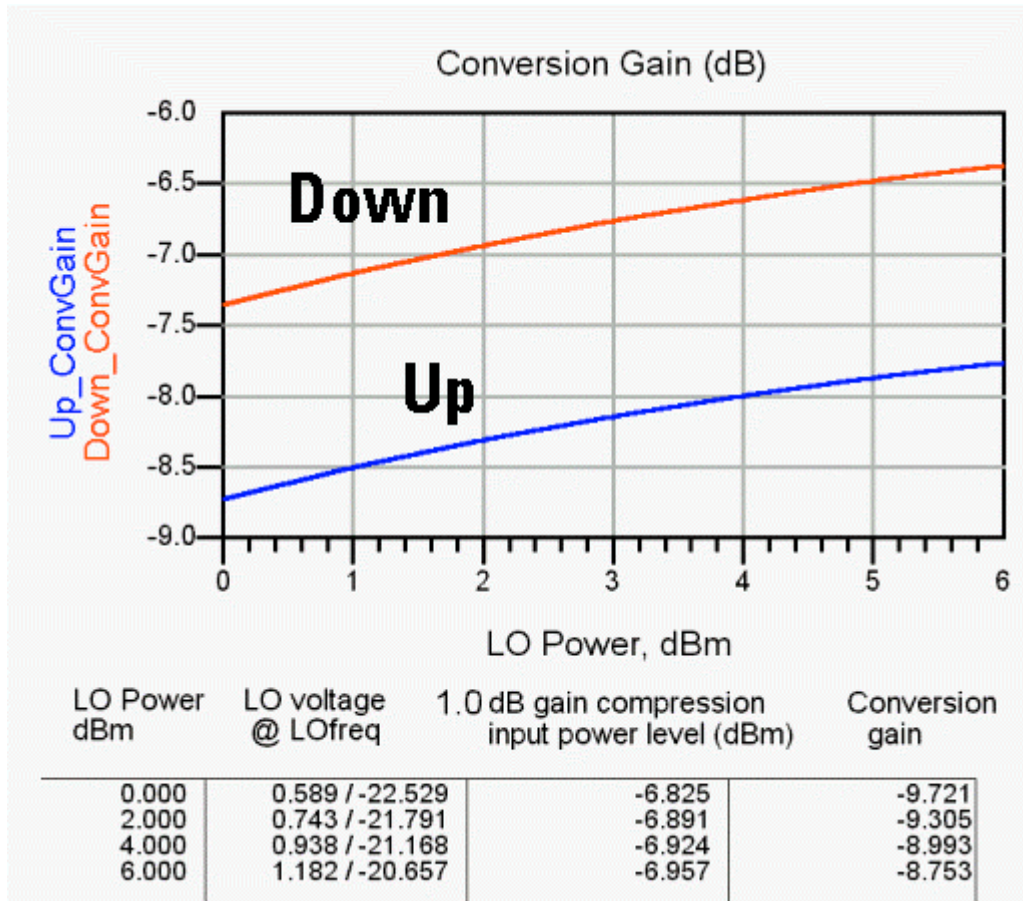
With the topology established, the first step in designing this mixer was to determine a suitable LO voltage to provide a reasonable compromise between conversion gain and LO power, but must be sufficient to avoid limiting the 1-dB gain-compression input voltage. The MOSFETs forming the commutating switch must be driven hard enough to present a low series resistance to the load. An LO power sweep and an N-dB gain compression analysis can be used to evaluate the dependence of gain compression on LO drive.

The simulation setup for the initial mixer design is shown in Fig. 3. For simplicity, the mixer is implemented as a sub-network. As a sub-network, the mixer itself can be replaced or modified as necessary through out the design process while maintaining the basic simulation setup. Mixer parameters are accessible outside the sub-network and are passed to the mixer design for analysis. In this example,  $V_{DD}$  (drain voltage),  $R_D$  (drain resistance),  $W_{csp}$  (current source control width),  $W_1$  and  $W_2$  (transconductance and switch MOSFET widths),  $R_S$  (source-degeneration resistance), and  $L_S$  (source-degeneration inductance) are all available for parameter sweeps.



**Fig. 3: Simulation Set-Up For The Differential Mixer**

Simulation results showed that the input power  $P_{1dB}$  does not have a strong dependence on LO voltage, but conversion gain does depend somewhat on LO voltage (see Fig. 4.) As higher gate voltage is applied to the upper pair of MOSFETs, their series resistance becomes lower relative to the drain resistance and, thus, the conversion gain is higher. Simulation showed a conversion loss that became worse at the higher output RF frequency of 1.8 GHz, but this could be improved by tuning the RF output of the mixer.



**Fig.4: Simulations Of LO Voltage Effects On Input Power 1-dB Gain Compression And Conversion Gain**

### Gain Compression Evaluation

The 1-dB gain compression input power and voltage were found for a range of swept parameters. For this example we wanted to know the influence of  $R_S$  and  $R_D$  on  $V_{1dB}$ . The  $R_S$  sweep used an  $R_D$  of 100  $\Omega$ , and an  $R_S$  of 30  $\Omega$ . Conversion gain was measured at the 1 dB compression level.

$V_{1dB}$ , rather than  $P_{1dB}$ , is used as the input signal level parameter. In an RFIC mixer, where the input might not be matched to a source impedance, the input voltage is a more important metric of gain compression than the input power, because available power assumes a conjugate match between source and load. Also, in a multi-signal environment, the peak input voltage can be quite large at the instant in time when all signals add in phase. It is this peak voltage that determines the distortion limits of the mixer. For example, two-tone IMD simulations predict a 1-dB compression power that is 6 dB lower than a single tone simulation, because the peak voltage is twice as high for the same power per tone.

Also note that the conversion power gain varies inversely with the value of  $R_D$ . In the simulation, the external load resistance was set to  $2R_D$  so that the output power (power absorbed by the load) was also the available output power,  $P_{out} = V_{out}^2 / R_D$ . The voltage gain would be expected to follow  $R_D/R_S$ , but increased less rapidly than anticipated, probably due to the output bandwidth limitations because of the RC time-constant.

### **Determining Noise And Gain Tradeoffs**

The next step was to evaluate how dc bias current ( $I_{\text{bias}}$ ) and source resistance affect the mixer noise figure. The mixer single-sideband noise figure (SSB NF) was simulated as a function of dc bias current through the Gilbert cell (mixer core.) The dc current was varied by sweeping the width of the pMOS current source ( $W_{\text{csp}}$ ) and the mixer current mirror width ( $W_{\text{cs}}$ ) using a parameter sweep.

SSB NF was appropriate because only one input frequency was applied to the mixer, but wideband noise at the image frequency and from LO harmonics was included in the signal-to-noise calculation. The simulation showed that the NF was reduced as  $I_{\text{bias}}$  increased, but reached a point of diminishing returns. Thus, a width of 50  $\mu\text{m}$  for the current source was selected as a compromise between power and noise.

The SSB NF was also found to be strongly dependent on the source resistance. This was expected because the thermal noise contributed by the resistor is directly in the input voltage loop of the differential pair. Thus, there needs to be a tradeoff between  $V_{\text{1dB}}$  and NF to obtain the largest dynamic range.

The dynamic range at low input signal power levels is limited by the carrier-to-noise ratio. The noise power for a minimum detectable signal ( $S/N = 1$ ) depends on both NF and the noise bandwidth. This bandwidth is normally set by an external SAW filter between the mixer and the driver amplifier. The filter is also required to reject the output difference ( $F_{\text{LO}} - F_{\text{in}}$ ) image frequency at 1.4 GHz.

The drain resistor thermal noise is input-referred through the gain, thus, the conversion gain (or loss in this case) may also increase the NF. If better performance is needed to reach design goals, a tuned output should be investigated to eliminate some of this noise.

At higher input signal levels the dynamic range of the mixer is limited by distortion. The third-order IMD products are the most damaging because they show up in-band and cannot be rejected by the filter. A two-tone third-order IMD simulation with an RF power sweep was used to display the carrier-to-IMD power ratio. The IMD power present at the output increases at three times the rate of increase of input power. Thus, the difference between the output power and IMD power becomes smaller with increasing input.

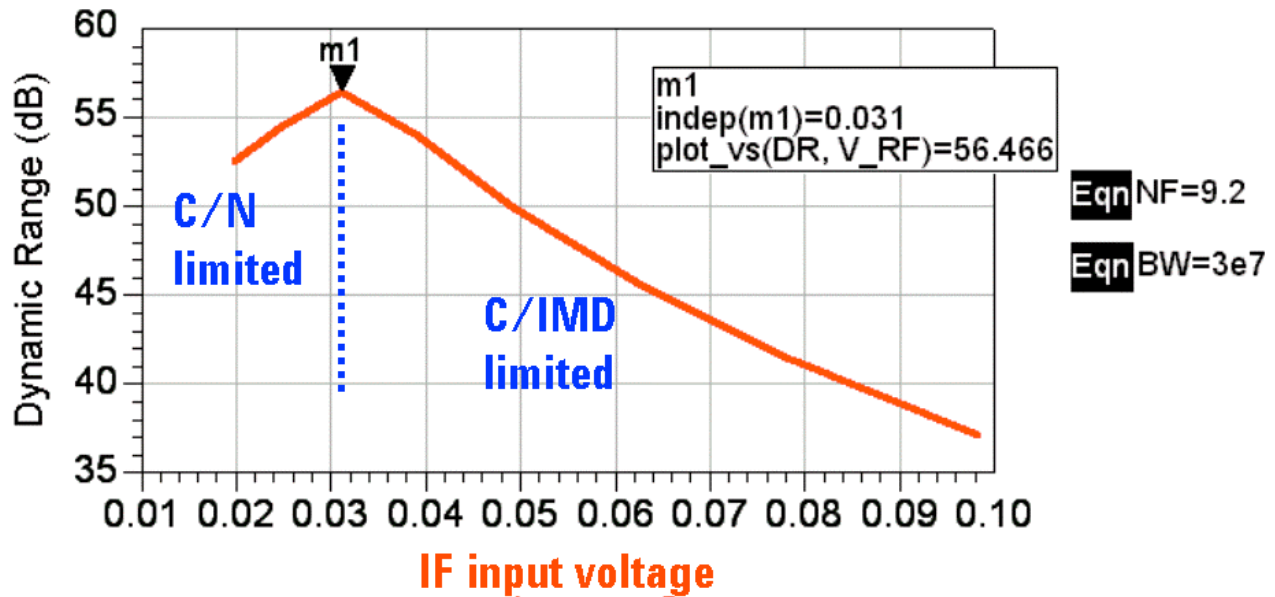
### **Dynamic Range Vs. Input Voltage**

To determine the effect of input voltage on dynamic range, two simulations were required: IMD RF power sweep and the SSB NF (see Fig. 5.) The dynamic range is determined by the least of these two conditions:

$DR = P_{\text{out}} \text{ (dBm)} - \text{MDS (dBm)}$ : Noise-limited for low input levels

$DR = P_{\text{out}} \text{ (dBm)} - P_{\text{IMD}} \text{ (dBm)}$ : Distortion-limited for higher input levels

$R_s$	DR (dB)	$V_{in}$ (V) (differential)	NF (dB)
10	57.7	0.017	6.5
20	57.3	0.025	8
30	56.4	0.031	9.2
40	56.0	0.039	10.3



**Fig. 5: The Effects Of Input Voltage On Dynamic Range**

The dynamic range peak depends on the noise bandwidth. For narrower bandwidths, the noise floor is lower and the peak DR increases but shifts to lower differential input voltage. A 30-MHz noise bandwidth was chosen because the transmitter should be capable of covering an entire frequency band in a base station application.

### Tuning The Mixer Drain Nodes

The low conversion gain of the resistively-loaded mixer resulted in higher noise due to the drain resistors. By resonating the output at 1.8 GHz, the conversion gain was increased and the gain at the image frequency (1.4 GHz) was reduced. A comparison between the resistively loaded case and the tuned case showed an increase in conversion gain of about 3.5 dB.

The resonant frequency of the tuned output is affected by the capacitance contributed by the drain-to-substrate junction. To find the resonant frequency of your specific design, perform an RF frequency sweep. From that, you can calculate the value of this capacitance and absorb it into the resonator.

### Gain Reduction Due To Low Inductor Q

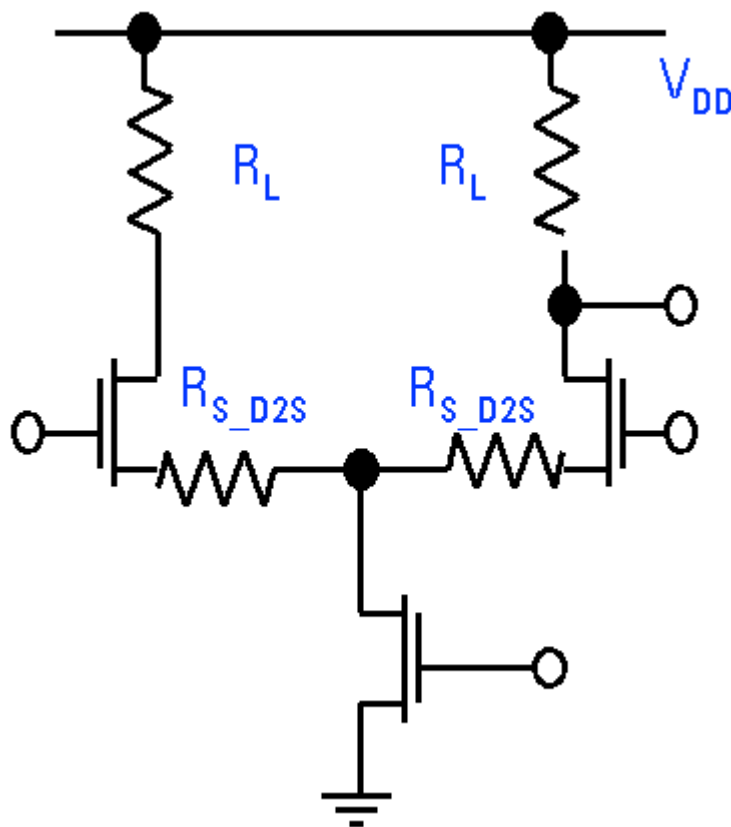
In bulk silicon processes, on-chip inductor Q is limited by metal losses and substrate conduction. An ordinary digital IC process produces low Q in spiral inductors. CMOS or BiCMOS RFIC

processes can achieve higher Q inductors by using thicker dielectrics and thicker metal. Q values in the range of 5 to 15 are typical.

Unfortunately, for realistic unloaded inductor Q values on the order of 5, the benefits of tuned output are diminished. The conversion gain is improved by about 4 dB, but the NF is improved by just 0.5 dB. A tuned output would be of greater benefit on a CMOS RF analog, SOI, or GaAs process, where higher Q values can be obtained.

### Differential-To-Single-Ended Conversion

An active balun was used to convert the RF output from differential to single-ended. Instead of taking one output from the mixer, this conversion maintains a differential output, which is necessary for rejection of LO feedthrough. The single-ended output drives the SAW filter between the mixer output and the driver stage. Although passive baluns can be made for 1.8 GHz, an active on-chip balun provides cost and size benefits. The differential to single-ended amplifier stage is shown in Fig. 6. The gate capacitances of the D2SE stage can be absorbed into the resonator at the mixer drain nodes. Also, the D2SE stage must be designed so that it does not dominate the IMD generation of the mixer.  $R_{S\_D2SE}$  can be adjusted to set the  $V_{1dB}$  level.



**Fig. 6: The Differential-To-Single-Ended Conversion Stage**

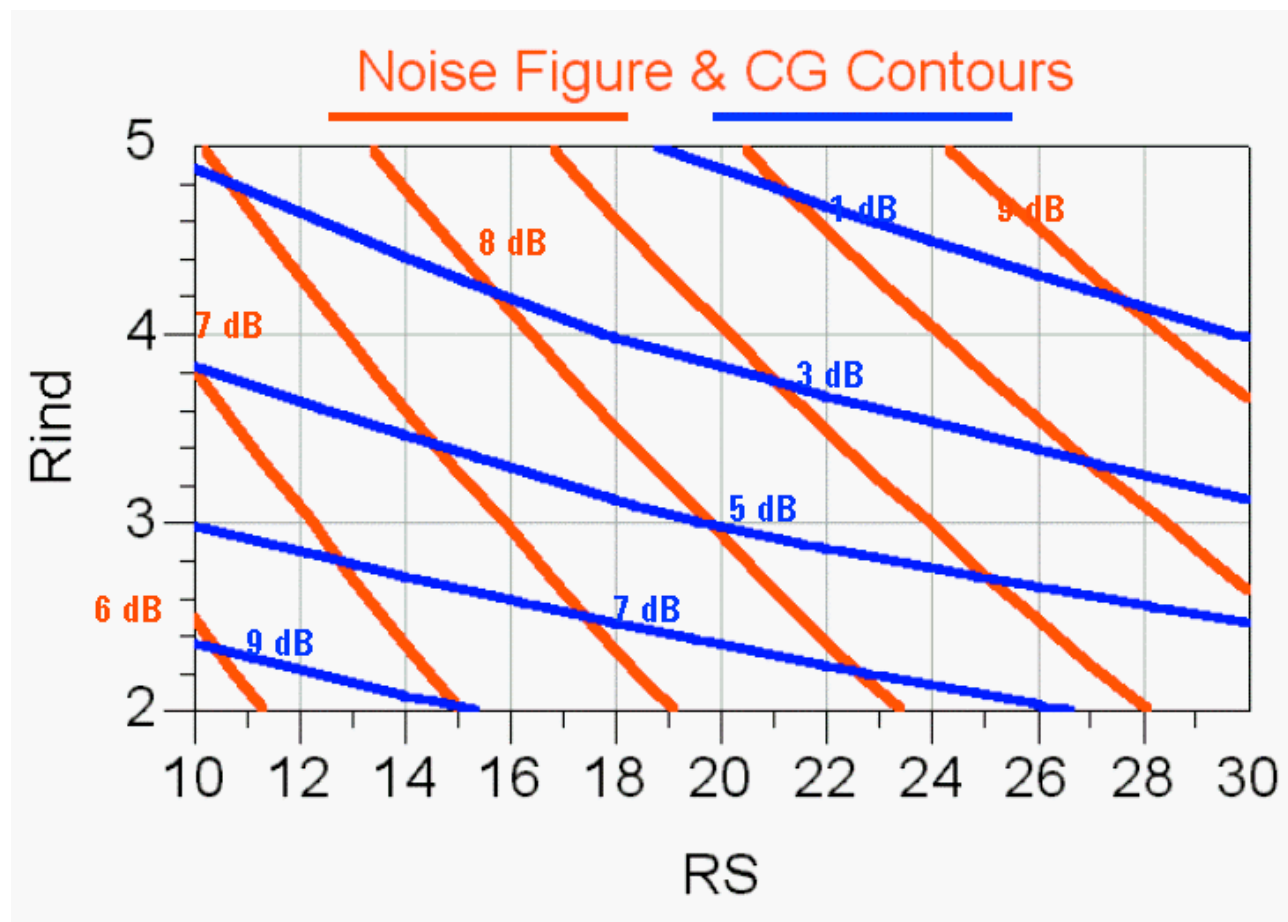
The output driver could use an off-chip load resistance with an open drain output connection, as suggested by Fig. 6. The load resistance would then be determined either by the filter impedance or by a transmission line impedance, which would then dictate the bias current for the D2SE converter

stage. The device widths must also be chosen so that they can handle the necessary drain current and provide adequate voltage gain. The addition of a source follower at the output is another option for driving a low load impedance.

### Evaluation Of The Design

Earlier in the design process, it was easier to measure the differential output so that tradeoffs and comparisons could be made between the differential tuned mixer and the mixer with an output buffer. Once the design was complete, the mixer could then be evaluated in a single-ended configuration.

The SSB NF simulation was performed again with parameter sweeps for  $R_S$  and  $R_{ind}$ . Fig. 7 shows that there is little noise sensitivity to  $R_{ind}$ ; however, it strongly affects the conversion gain.  $R_S$  affects both NF and conversion gain and also the carrier-to-IMD ratio versus IF input voltage. The mixer TOI/IMD simulation was performed again for an  $R_S$  of 10, 20, and 30  $\Omega$ . The dynamic range slowly improves for smaller  $R_S$ , but is strongly dependent on the noise bandwidth.



**Fig. 7: Noise Figure And Conversion Gain Contours**

To speed up the evaluation process, an ADS 2001 Mixer DesignGuide schematic intended for evaluation of single-ended mixers was copied from the menu and modified as shown in Fig. 8. The tuned mixer with the D2SE output stage was then inserted from the component library. Unused inputs were terminated, the input was grounded and the output terminated in a large resistance. To

obtain a differential LO, a transformer and source were copied from a differential test schematic and pasted into this schematic.

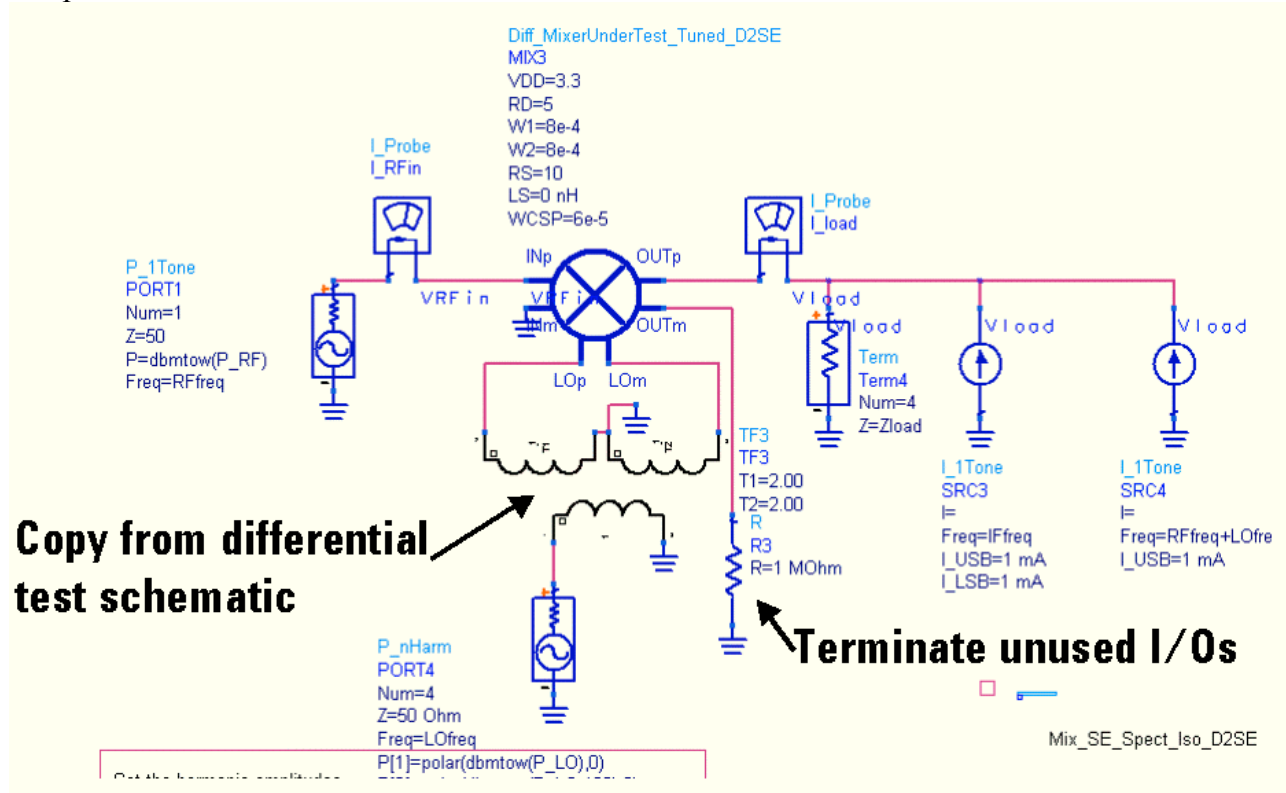


Fig. 8: A Single-Ended Mixer Modified To Evaluate The Designed Mixer

Again, NF and IMD-versus-RF power sweeps were performed for a range of  $R_S$  values from 10 to 30  $\Omega$ . This was combined to determine dynamic range, plotted in Fig. 9. An  $R_S$  of 10  $\Omega$  produced the best result: a peak dynamic range of 57.5 dB at an input voltage of 14 mV.

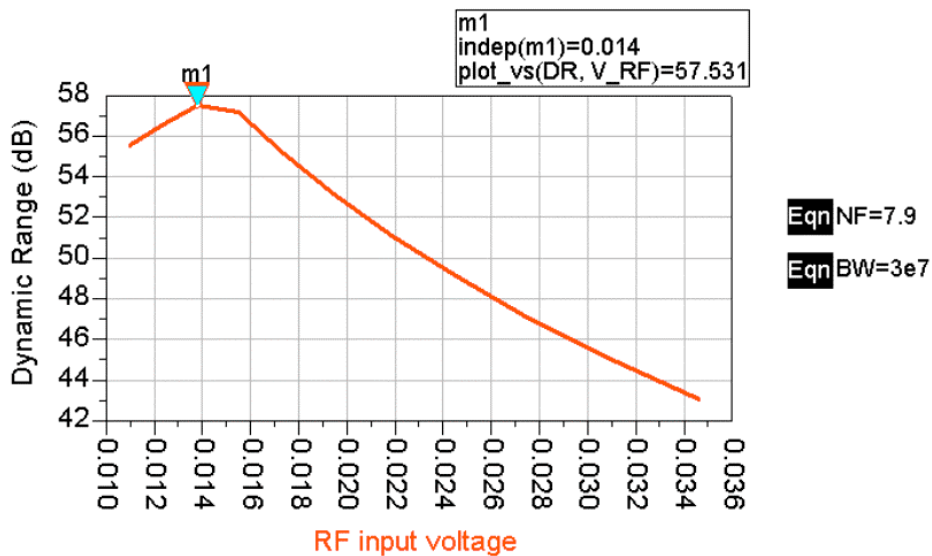


Fig. 9: The Dynamic Range Peak At 57.5 dB Is Obtained At An Input Level Of 14 mV

Simulation with a digital signal source such as CDMA is a more severe test of linearity. To evaluate performance under these conditions, an IS-95 CDMA source with very good ACPR was used to drive the mixer input. When the input RF signal level was set to the optimum value for mixer dynamic range, relatively little spectral regrowth was observed.

Readers should note that the input of the mixer is badly mismatched. This is not a big concern if the baseband and IF driver circuits are on the same chip with the upconversion mixer, where the voltage levels are of greater interest. However, if the mixer is driven from off-chip circuitry, the input impedance will be dominated by capacitive reactance and a matching network could significantly increase the conversion gain.

### **Summary**

This study of the design and optimization of an RFIC upconversion transmit mixer shows how design and analysis tools can help you evaluate the performance of your design and determine ways to improve it. The mixer design presented in this paper was achieved using the Mixer DesignGuide in Agilent EEsof EDA's Advanced Design System 2002C. The Mixer DesignGuide was used to create simulation setups, data displays and impedance matching.

To download the ADS project file used in this design or to see a design seminar on mixer design using ADS and the Mixer DesignGuide please visit <http://www.agilent.com/find/eesof>.

### **Author Bio**

*Stephen Long earned his BS degree in Engineering Physics from UC Berkeley and MS and PhD in Electrical Engineering from Cornell University. He has been a professor of electrical and computer engineering at UC Santa Barbara since 1981. The central theme of his current research projects is rather practical: using unconventional digital and analog circuits, high performance devices and fabrication technologies to address significant problems in high-speed electronics such as low-power IC interconnections, very high-speed digital ICs, and microwave analog integrated circuits for RF communications. He teaches classes on communication electronics and high-speed digital IC design.*

*Prior to UCSB, he was a Senior Engineer at Varian Associates and then he was at Rockwell International Science Center as a member of the technical staff.*

*Dr. Long received the IEEE Microwave Applications Award in 1978 for development of InP millimeter wave devices and in 1988 he was a research visitor at GEC Hirst Research Centre in the U.K. In 1994 he was a Fulbright research visitor at the Signal Processing Laboratory, Tampere, and a visiting professor at the Electromagnetics Institute, Technical University of Denmark. He is a senior member of the IEEE and a member of the American Scientific Affiliation.*