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Overview on CMOS RFIC Upconversion Mixer Design

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CMOS RFIC Upconversion Mixer Design

Optimized tradeoffs are required when implementing a mixer in an RFIC

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Designing a mixer to work within an RFIC is not an easy task. Many different performance specifications must be prioritized, based on the application. Some are intended for receive applications with a wide range of input signal levels, where maximum linearity under large signal conditions is often more critical than the noise figure. On the other hand, for transmit applications where the signal levels can be controlled, 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 tasks required to design a quality RFIC. The application is an upconversion mixer intended for a base station transmitter power amplifier section (Figure 1). The design is based on a Gilbert cell MOSFET double-balanced differential mixer with an input IF signal centered at 200 MHz and an output of 1.8 GHz. The example uses 0.35 µm MOSFETs with a default device model parameter set. For your application, you would substitute your own verified model.

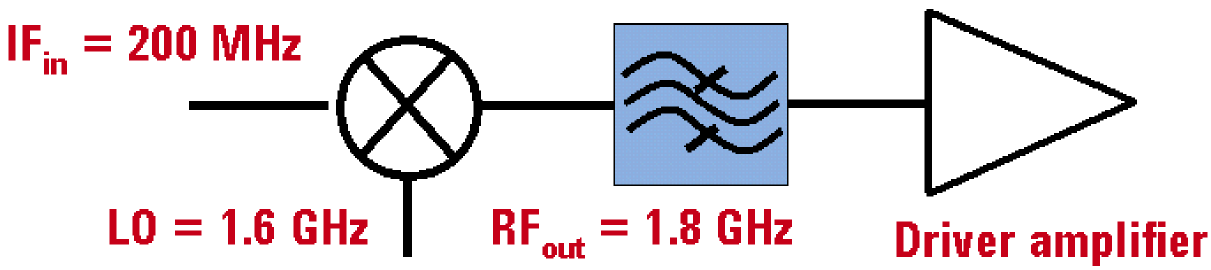


Figure 1. The upconversion mixer for the base-station transmitter power amplifier application.

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

The Gilbert cell mixer

Figure 2 shows the schematic of a MOSFET version of the Gilbert cell active double-balanced mixer. The lower FET differential pair serves as a transconductance amplifier, while the upper FETs provide a fully balanced, phase-reversing current switch. A DC bias generator (not shown) keeps the MOSFETs in their active region.

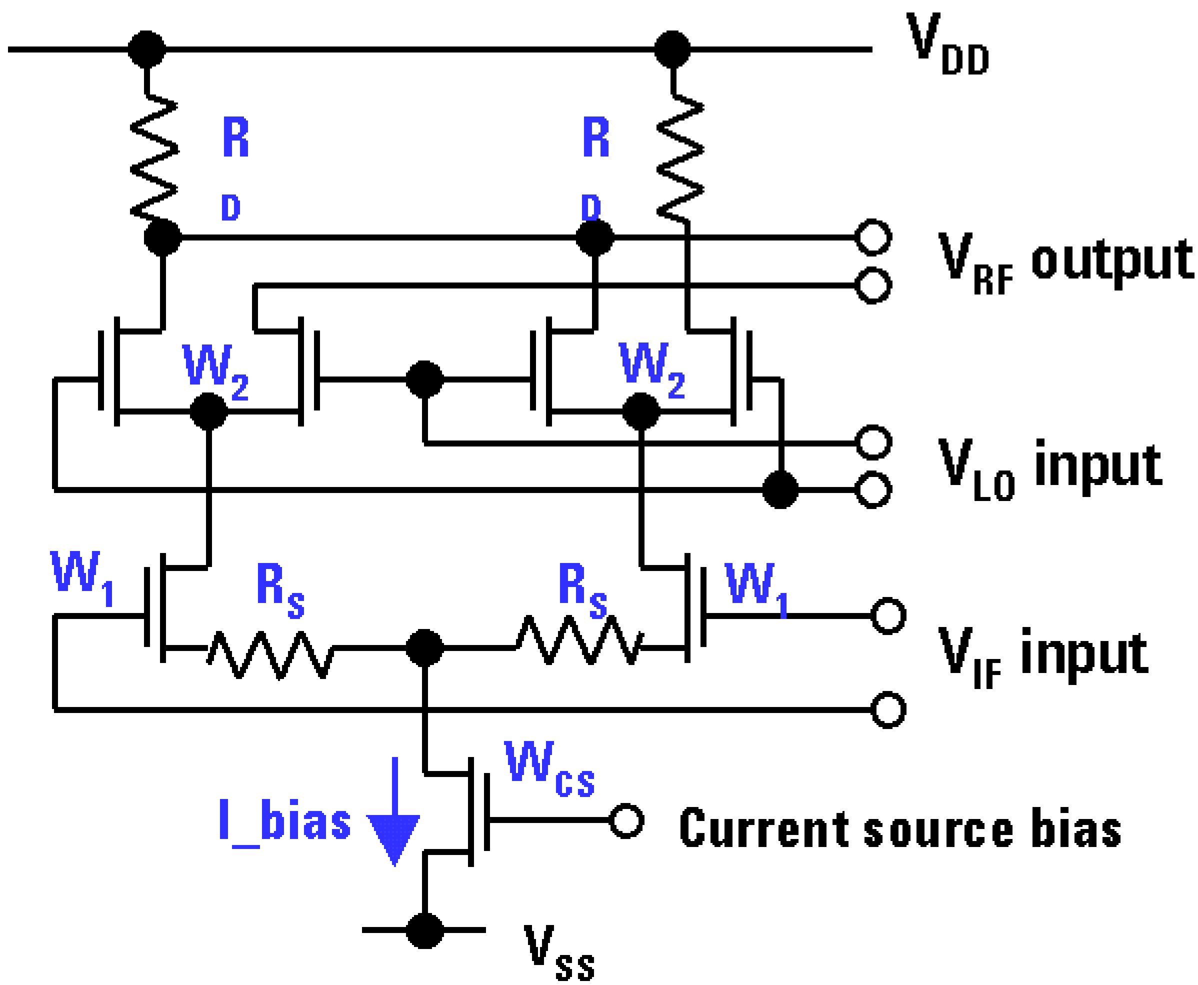


Figure 2. The MOSFET Gilbert cell active double-balanced mixer

The large signal-handling capability 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, you may use power, P_{1dB}). The maximum linear input voltage range can be increased by increasing the values of the source-degeneration resistors (R_S). Source inductance can also provide beneficial degeneration, but with a low input IF frequency of 200 MHz, the required inductance values would be too large for RFIC implementation. 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 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.

Design sequence

A mixer that is used for base station transmit applications requires 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) Determination of LO amplitude. The mixer commutating switch must be fully activated, as excess distortion can be produced with a weakly conducting or slowly activated switch. The conversion transducer gain and 1 dB gain compression input level were used to determine when the LO voltage is sufficient.

2) Evaluation of the influence of source and drain resistance on the 1 dB compression level, giving insight into the principal mechanisms that limit linearity.

3) Determination of 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) Evaluation of 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—the mixer's "sweet spot" for best performance—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, testing the mixer under a more realistic signal input, such as a CDMA source, to emulate a multi-carrier environment. This is a more severe test than the two-tone IMD one, and is much more time consuming to simulate because a large number of symbols must be used for accurate results.

Once 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 inductors with reasonable Q can be fabricated. 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 only degrade the distortion of the output stage by increasing the peak voltage present at its input.

The other change was to convert the differential signal to a single-ended one. 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. The 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

Once the topology was established, the first step in designing this mixer was to determine a suitable LO voltage. The LO level should provide a reasonable compromise between conversion gain and LO power, but should not limit the 1 dB gain-compression input voltage. The MOSFETs forming the commutating switch (upper level) 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 Figure 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.

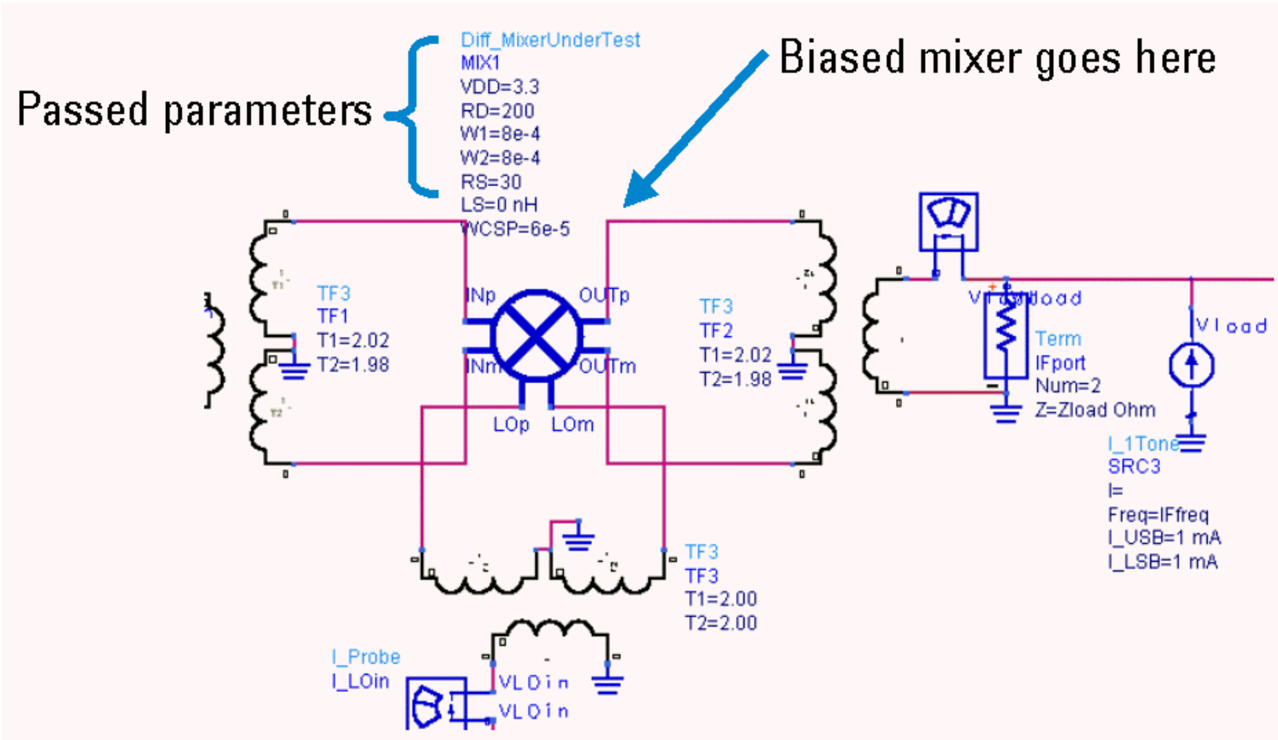
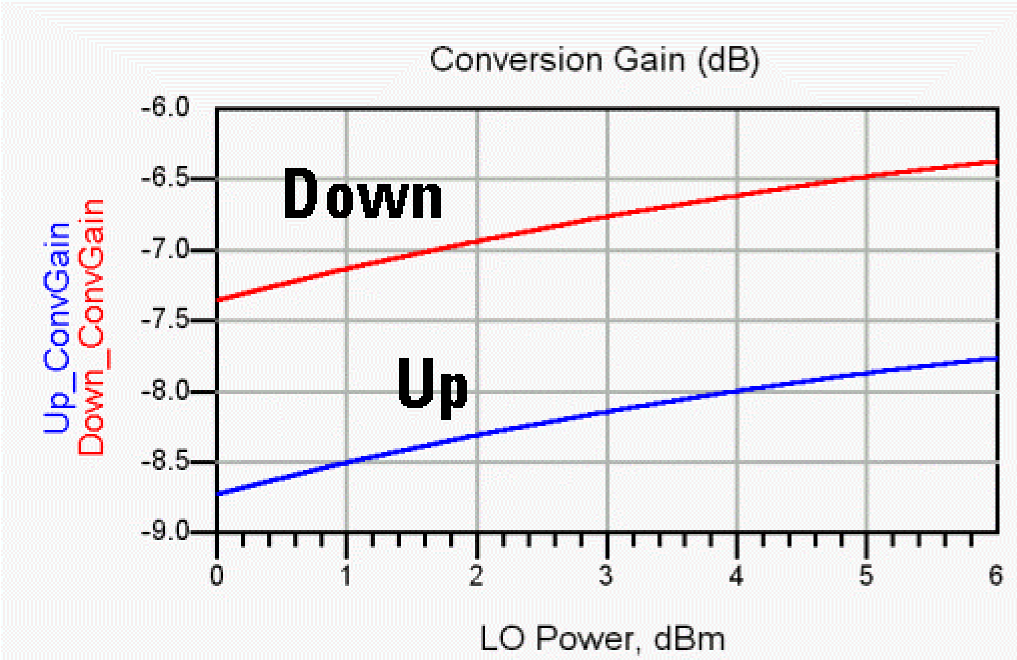


Figure 3. Differential mixer simulation setup.

The simulations showed that the input power at which gain compresses by 1 dB (P_{1dB}) does not have a strong dependence on LO voltage, but conversion gain does depend somewhat on LO voltage (Figure 4). As more 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. There was 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.



LO Power dBm	LO voltage @ LOfreq	1.0 dB gain compression input power level (dBm)	Conversion gain
0.000	0.589 / -22.529	-6.825	-9.721
2.000	0.743 / -21.791	-6.891	-9.305
4.000	0.938 / -21.168	-6.924	-8.993
6.000	1.182 / -20.657	-6.957	-8.753

Figure 4. Simulations showing the effects of LO voltage on the input power at which gain compresses by 1 dB, and conversion gain.

Gain compression evaluation

Next, gain compression was evaluated. The 1 dB gain compression input power and input 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 ohms, and the R_D sweep set used an R_S of 30 ohms. Conversion gain was measured at the 1 dB compressed 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 predicted a 1 dB compression power that was 6 dB lower than predicted by single tone simulations because the peak voltage was twice as high for the same power per tone.

It is also noteworthy 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 RC time-constant bandwidth limitations.

Determining noise tradeoffs

The next step was to evaluate how DC bias current (I_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_{csp}) and the mixer current mirror width (W_{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 with increasing I_bias, but reached a point of diminishing returns. Thus, a width of 50 μm 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_{1dB} and NF to obtain the largest dynamic range of the mixer.

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 ($F_{LO} – F_{in}$) image frequency at 1.4 GHz.

The conversion gain (or loss in this case) may also increase the noise figure because the drain resistor thermal noise is input-referred through the gain. If the design goals require it, 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 in the output increases at three times the rate of increase of input power. Thus, the difference between output power and IMD power shrinks with increasing input.

Dynamic range versus input voltage

Determining the effect of input voltage on dynamic range required the output from two simulations: IMD RF power sweep and the SSB NF (Figure 5). The dynamic range is controlled by the least of these two conditions:

DR = P_{out} (dBm) – MDS (dBm)   (noise-limited for low input levels)

DR = P_{out} (dBm) – P_{IMD} (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

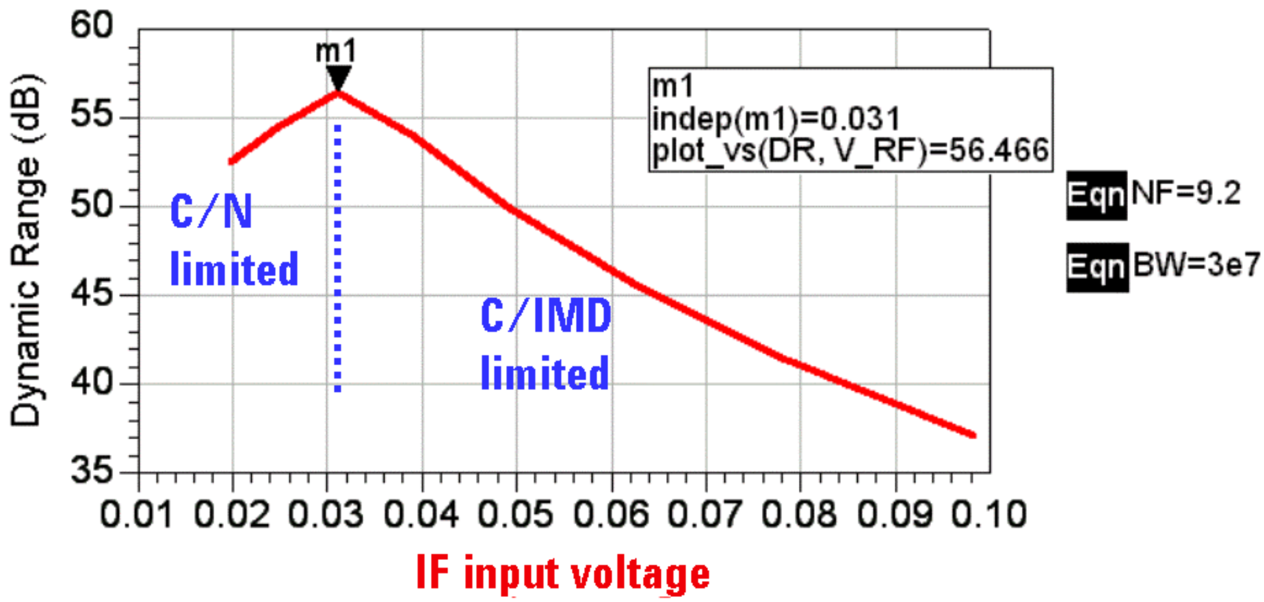


Figure 5. The effects of input voltage on dynamic range.

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

Tuning mixer drain nodes

The low conversion gain of the resistively loaded mixer caused 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 (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.

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

Gain reduction due to 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 noise figure is imporved by only 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

The next step in the design was to convert the RF output from differential to single-ended with an active balun. Rather than taking one output from the mixer, this conversion is required to maintain a differential output, which is necessary for rejection of LO feedthrough. A single-ended output is sufficient to drive the SAW filter that is needed between the mixer output and the driver stage. Although passive baluns can be made for 1.8 GHz, placing an active balun on-chip provides cost and size benefits. The differential amplifier stage shown in Figure 6 converts the differential output of the tuned mixer to a single-ended output. 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.

The output driver could use an off-chip load resistance with an open drain output connection, as suggested by Figure 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 to the output is another option.

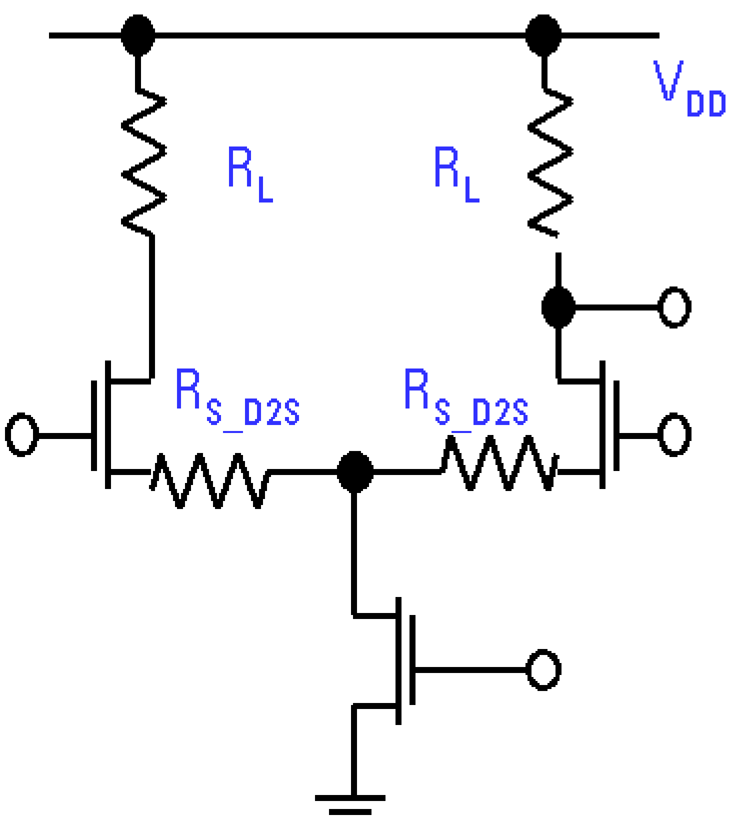


Figure 6. The differential amplifier stage used to the differential output to single-ended.

Design evaluation

For the initial design evaluation, 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}. Figure 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 ohms. The dynamic range slowly improves for smaller R_S, but is very dependent on the noise bandwidth.

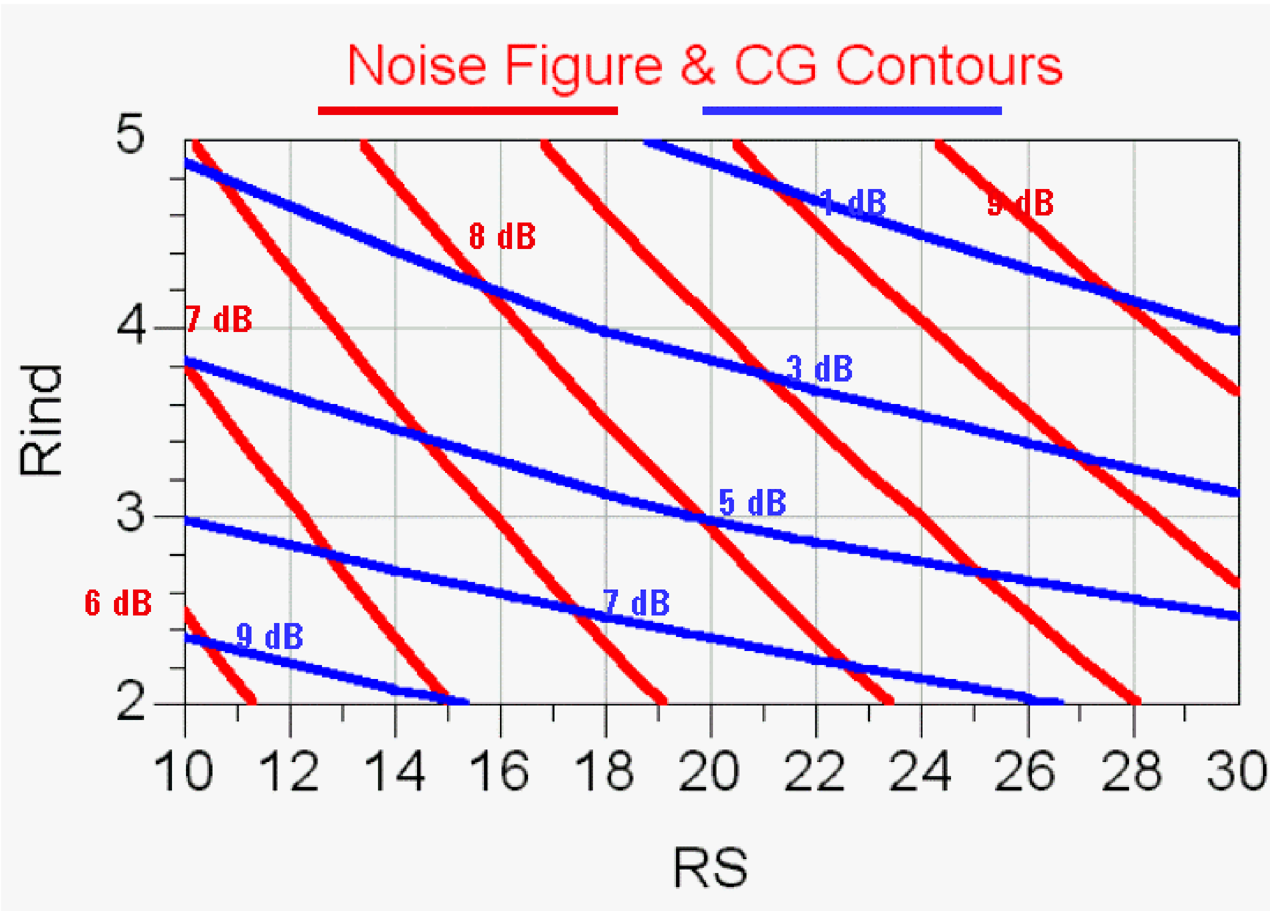


Figure 7. Noise figure and conversion gain contours.

To speed up the process, an ADS 2001 Mixer DesignGuide schematic intended for evaluation of single-ended mixers was copied from the menu and modified as shown in Figure 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. An active LO single-ended-to-differential stage could also be designed and added to the mixer.

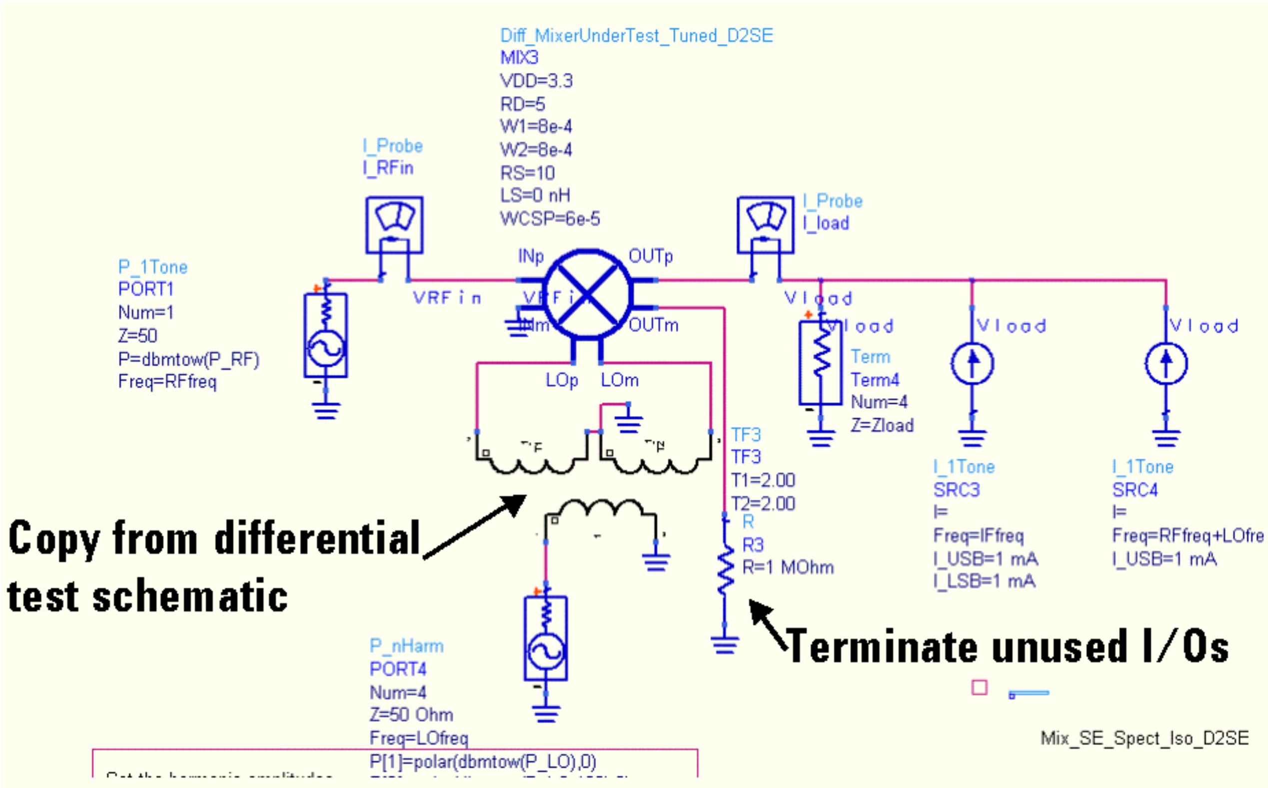


Figure 8. A single-ended mixer modified to evaluate the designed mixer.

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

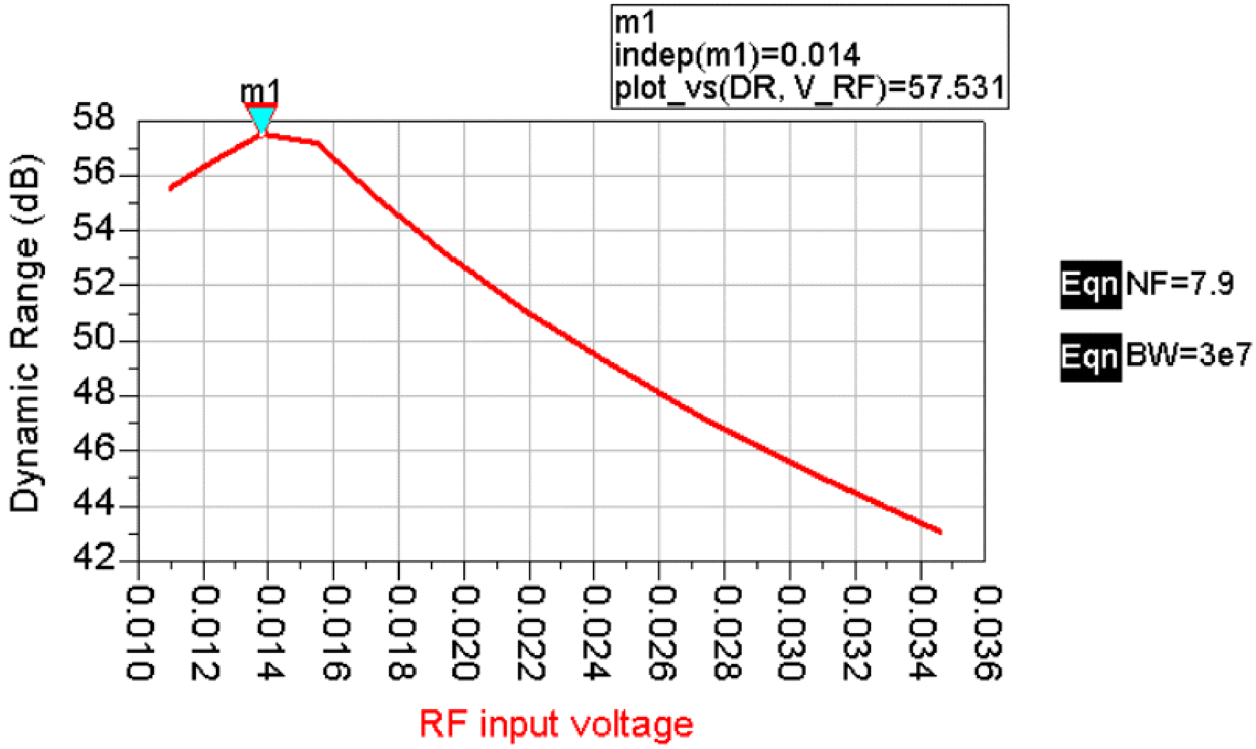


Figure 9. The dynamic range peak at 57.5 dB is obtained at a input level of 14 mV.

Simulation with a digital signal source such as CDMA is a more severe test of linearity. 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.

The input of the mixer is badly mismatched, which may not be of much concern if the baseband and IF driver circuits are on the same chip with the upconversion mixer. In that case, the voltage levels are of greater interest. 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 2001 (ADS 2001). The Mixer DesignGuide was used to create simulation setups, data displays and impedance matching.

To download the ADS 2001 project file used in this design or to see a design seminar on mixer design using ADS and the Mixer DesignGuide please visit the Agilent web site: http://www.agilent.com/eesof-eda.

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Revised: March 27, 2008

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Printed in USA, November 01, 2001
5989-8993EN

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