Gilbert Cell Mixer Design Tutorial

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1 ABSTRACT

Frequency translation in a system, is performed by a non-linear device known as a mixer. There are various topographies from simple single ended, single balanced mixers to more complicated double & triple balanced mixers that provide better isolation from the Local Oscillator (LO) and spurious.

The most popular double-balanced mixer used in RFIC designs is the Gilbert Cell mixer.

The design of this mixer is the subject of this paper.

2 INTRODUCTION

Mixers are non-linear devices used in systems to translate (multiply) one frequency to another. All mixer types work on the principle that a large Local Oscillator (LO) RF drive will cause switching/modulating the incoming Radio Frequency (RF) to the Intermediate Frequency (IF).

The multiplication process begins by inputting two signals:

$$a = A\sin(\omega_1 t + \phi_1) \text{ and signal } b = B\sin(\omega_2 t + \phi_2)$$

The resulting multiplied signal will be:

$$a.b = AB\sin(\omega_1.t + \phi_1).\sin(\omega_2.t + \phi_2)$$

This can be multiplied out thus:

$$\text{Using this trig identity .. } \sin A\sin B = -\frac{1}{2}\left[\cos(A+B) - \cos(A-B)\right]$$

$$\text{Where } A = (\omega_1.t + \phi_1) \text{ and } B = (\omega_2.t + \phi_2)$$

$$= -\frac{AB}{2}\left[\cos((\omega_1.t + \phi_1) + (\omega_2.t + \phi_2)) - \cos((\omega_1.t + \phi_1) - (\omega_2.t + \phi_2))\right]$$

$$= -\frac{AB}{2}\left[\underbrace{\cos((\omega_1 + \omega_2)t + (\phi_1 + \phi_2))}_{\text{Sum frequency (removed by filtering)}} - \underbrace{\cos((\omega_1 - \phi_1)t - (\phi_1 - \phi_2))}_{\text{Difference frequency ie I.F}}\right]$$

3 MIXER DEFINITIONS

(1) Conversion Gain: This is the ratio (in dB) between the IF signal (usually the difference frequency between the RF and LO signals) and the RF signal.

(2) Noise Figure: Noise figure is defined as the ratio of SNR at the IF port to the SNR of the RF port.

(i) Single sideband (SSB): This assumes the only noise from the signal **ω_1** and not the image frequency **ω_1-1**, this would be the case if a band-pass filter was added in front of the mixer eg.

RF = 1694 MHz, LO = 1557MHz to give an IF of 137MHz.

Also an image IF will add to 137MHz from an RF of 1420MHz ie 1557MHz-1420MHz = 137MHz

(ii) Double sideband (DSB): In DSB both sidebands are available thus it has twice as much power available at the IF port compared to the SSB signal. As a result, it's conversion loss is 3dB less than that of an SSB signal, as shown:

$$P_{(IF)DSB} = 2\,P_{(IF)SSB} \text{ and conversion loss is given by}$$

$$(LC)_{DSB} = (LC)_{SSB} - 3(dB)$$

$$\text{or in terms of loss ratios } (LC)_{DSB} = \frac{(LC)_{SSB}}{2}$$

(iii) DSB to SSB Noise Figure conversion

$$Fm = 1 + (LC - 1)\frac{T}{To}$$

Where T = temperature of mixer,
To = room temperature (273°K)

For DSB, $(LC)_{DSB} = \frac{(LC)_{SSB}}{2}$

Therefore, $Fm_{(DSB)} = 1 + \left(\frac{(LC)_{SSB}}{2} - 1\right)\frac{T}{To}$

At room temperature, ie $T = To$

$Fm_{(DSB)} = \frac{(LC)_{SSB}}{2}$ in other words

$Fm_{(DSB)}$ is half or 3dB less than $Fm_{(SSB)}$

$$\boxed{Fm_{(DSB)} = Fm_{(SSB)} - 3(dB)}$$

(3) Isolation: These parameters define how much signal leakage will occur between pairs of ports. ie RF to LO,LO to IF and RF to IF. So if for example RF to IF isolation was specified at 35dB this means that the RF at the IF port will be 35dB lower than the RF applied to RF port.

(4) Linearity

(i) 1dB Compression point: Like other non-resistive networks, a mixer is amplitude-nonlinear above a certain input level resulting in a gain compression characteristic as shown in Figure 1.

Above this point the If fails to track the RF input power level – normally a 1dB rise in RF power will result in a 1dB rise in the IF power level. The 1dB compression point is measured by plotting incident RF power against IF power as shown in the figure above.

Most mixers have the 1dB compression point specified at the input ie the single-tone input signal level at which the output of the mixer has fallen 1dB below the expected output level.

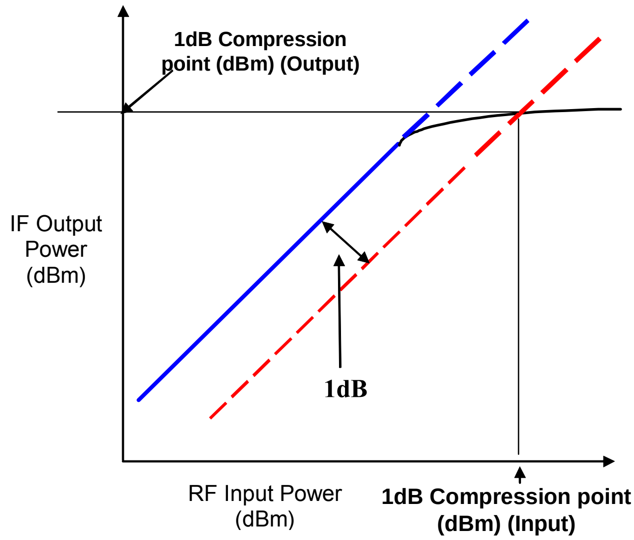


Figure 1 Typical gain compression characteristic for a non-linear Amplifier/Mixer, showing the measurement of the 1dB compression point.

For typical double balanced mixers this figure is ~ 6dB below the LO power level. (So performance can be improved by overdriving the LO port).

The 1dB compression point gives rise to the dynamic range of the mixer, which is the difference between the 1dB compression point and the minimum discernable signal (MDS – this is dependant on the noise floor of the device).

(ii) Intermodulation (IM3) performance: This parameter is the same as specified for amplifiers and measured in a similar way. It is measured by applying two closely spaced input tones at frequencies F_1 and F_2.

Third order products from the mixing of these tones with the LO (at frequency F_{LO}) occur at frequencies given by: $(2F_1 \pm F_2) \pm F_{LO}$ and $(2F_2 \pm F_1) \pm F_{LO}$. In the case of the mixer, the third order products of most interest are $(2F_1 - F_2) - F_{LO}$ and $(2F_2 - F_1) - F_{LO}$ as they fall in, or close to the IF band.

The IM3 performance is often summarised by giving the 3rd Order Intercept point (IM3 Intercept) as shown in the compression characteristic of Figure 2, where the IM3, IM5 plots intersect with the extrapolated gain plot (blue dotted line). As a rule of thumb the IM3 intercept point is approximately 10dB above the 1dB compression point.

This figure of merit gives an indication of the mixer's signal handling capability. In particular it provides an indication of the levels of third order products a mixer is likely to produce under multi-tone excitation.

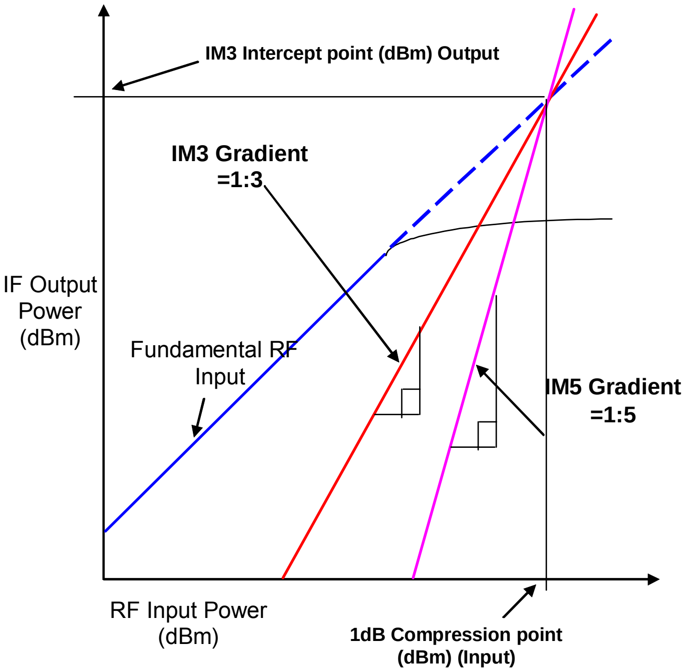


Figure 2 IM3 gain compression characteristic, as a rule of thumb the IM3 intercept point is approximately 10dB above the 1dB compression point.

The IM3 and IM5 graphs will intercept the fundamental graph at the intercept point. (Note the IM2 intercept point will be different and usually a lot higher).

Again, for mixers the measurement is referred to the input ($IP_{3,in}$) and is given by:-

$$IP_{n,in} = \frac{IMR}{(n-1)} + \text{Input power(dBm)}$$

Where IMR = Intermodulation ratio (The difference in dB between the desired output and spurious signal) and n = the IM order.

Typically, for double balanced mixers $IM_{3,in}$ is ~ 14dB greater than the single tone 1dB compression point and ~ 8dB greater than the LO power.

We can get a rough estimate of the gain compression of the LNA a non-linear expression of the input and output parameters can be expanded using Taylor's theorem. This results in the following equations for 1dB gain compression point and IM3 [1].

$$I_{DSAT} = W_{vsat} \cdot C_{ox} \cdot \frac{V_{od}^{\;2}}{V_{od} + E_{sat} \cdot L}$$

With

V_{od} = Voltage overdrive = Vgs - Vt and

E_{sat} = Velocity saturation field strength given by

$$E_{sat} = \frac{2V_{sat}}{\mu_{eff}} \quad \text{Where} \quad \mu_{eff} = \frac{\mu_o}{1 + \vartheta . V_{od}}$$

$$P_{1dB} \sim 0.29 \frac{V_{sat}.L}{\mu_1.Rs} V_{od} \left(1 + \frac{\mu_1.V_{od}}{4V_{sat}.L}\right)\left(1 + \frac{\mu_1.V_{od}}{2V_{sat}.L}\right)^2 \quad \text{- (1)}$$

to convert to dBm = 10 log (1000*P1dB)

$$P_{IIP3} \sim \frac{8}{3} \frac{V_{sat}.L}{\mu_1.Rs} V_{od} \left(1 + \frac{\mu_1.V_{od}}{4V_{sat}.L}\right)\left(1 + \frac{\mu_1.V_{od}}{2V_{sat}.L}\right)^2 \quad \text{- (2)}$$

$$\mu_1 \cong \mu_o + 2\vartheta \, \text{Vsat.L}$$

4 GILBERT CELL MIXER[1]

There are two types of mixer, passive and active. Generally passive types (although have better IM3 performance) have higher conversion losses and hence higher noise figures than active mixers.

Additionally, there are single balanced mixers and double balanced mixers. Single balanced mixers are much less complex, but have inferior performance in terms of RF to IF and LO to IF rejection, compared to double balanced mixers.

Given below are the advantages and dis-advantages of double balanced mixers.

Advantages:
(1) Both LO and RF are balanced, providing both LO and RF Rejection at the IF output.
(2) All ports of the mixer are inherently isolated from each other.
(3)Increased linearity compared to singly balanced.
(4)Improved suppression of spurious products (all even order products of the LO and/or the RF are suppressed).
(8)High intercept points.

(9) Less susceptable to supply voltage noise due to differential topography.

Disadvantages:

(1)Require a higher LO drive level.
(2)Require two baluns (although mixer will usually be connected to differential amplifiers).
(3)Ports highly sensitive to reactive terminations.

The most popular active, double balanced mixer topography in RFIC design is the **Gilbert Cell** mixer, the circuit of which is shown in **Figure 3**. This type of mixer exploits symmetry to remove the unwanted RF & LO output signals from the IF by cancellation.

4.1 MIXER OPERATION

The RF signal is applied to the transistors M2 & M3 which perform a voltage to current conversion. For correct operation these devices should not be driven into saturation and therefore, signals considerably less than the 1dB compression point should be used. Performance can be improved by adding degeneration resistors, on the source terminals of M2 & M3.

MOSFets M4 to M7 form a multiplication function, multiplying the linear RF signal current from M2 and M3 with the LO signal applied across M4 to M7 which provide the switching function.

M2 and M3 provide +/- RF current and M4 & M7 switch between them to provide the RF signal or the inverted RF signal to the left hand load. M5 & M7 switch between them for the right hand load.

The two load resistors form a current to voltage transformation giving differential output IF signals.

4.2 DESIGN GUIDELINES

Depending on the application the mixer may be designed with a low SSB noise figure, a particular gain or a high linearity. A good starting point is to use the differential LNA and add the switching FET's with the same W/L ratio's.

As we found with the LNA, to increase the linearity of the mixer source, degeneration resistors (or inductors) can be added to M2 & M3 sources as shown in **Figure 4**.

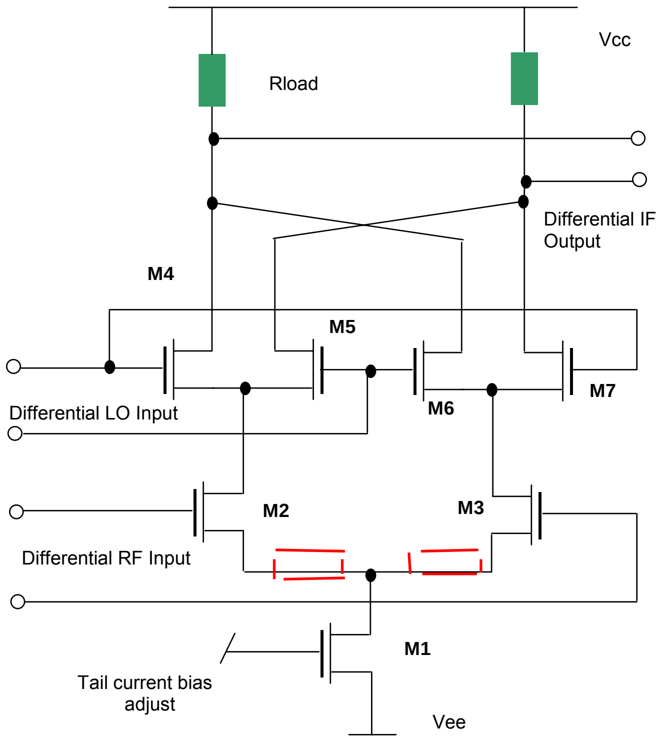


Figure 3 Basic circuit of the Gilbert Cell Double balanced mixer (DBM)

The voltage gain of the mixer with source degeneration is given by:

$$\frac{V_{RF}}{V_{IF}} \approx \frac{2}{\pi}\left(\frac{R_L}{Rs + \frac{1}{gm}}\right)$$

The gain of the first stage is determined by gm given by:

$$gm = \sqrt{\frac{2.K_N.W.I_D}{L}}$$

The voltage overdrive level ie (V_{gs}-V_T) should be set at around **0.2 to 0.4V**. Depending on the current flowing through the 'LNA' RF section we can determine the optimum W/L ratio for the LO switching section.

Other useful design equations are:

$$gm = \frac{2.I_D}{(Vgs - V_T)}$$

$$gm' = \frac{gm}{1+\sigma} \quad \text{Where } \sigma = R_S.gm$$

$$I_D = \frac{K_N.W}{2.L}.(Vgs - V_T)^2$$

The design stages for the basic mixer design with high linearity are:

(1) Select a value for Rs. If we start with the LNA design , Rs will be realized by an inductor Ls.
(2) Using the design equations of section 4(ii) decide on the IM3 value required to calculate the voltage overdrive (Vgs-VT). – This is part of the LNA design process.
(3) Calculate gm and check for compliance of mixer gain with a suitable load resistance.
(4) Determine LO switching W by selecting minimum L allowed and assuming that the overdrive voltage ($Vgs-V_T$) will be between 0.2 and 0.4V.

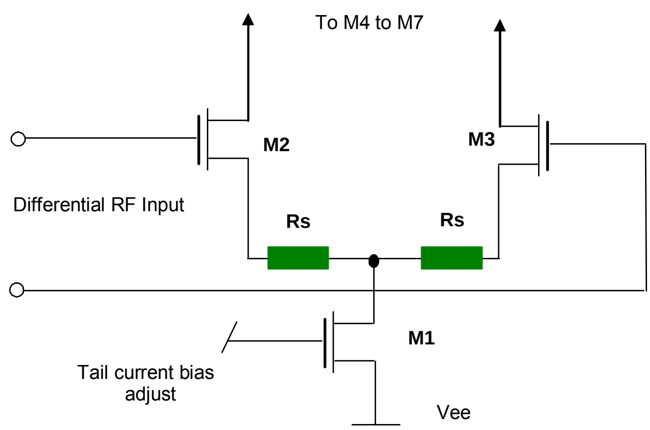


Figure 4 Modified RF input circuit of mixer with source degeneration added for improving the linearity. Note in the LNA design these resistances are realised as loss-less inductors Ls.

5 DESIGN EXAMPLE [1,2,3,4]

This section will describe the design of a Gilbert Cell mixer that will connect to the C-S Differential LNA designed for 'Bluetooth' described in another tutorial

The design criteria for the mixer, is shown in **Table 2-1**.

Parameter	Specification	Units
Frequency	2.45 to 2.85	GHz
Noise Figure (DSB)	< 10	dB
IIM3 Intercept Point (Input)	>20	dBm
Voltage Gain	>8	dB
Power consumption	<100	mW
Source impedance	50	ohms
Load impedance	500	ohms
Voltage Supply	± 2.5	V

Table 2-1 Required specification for the Bluetooth front end Gilbert Cell mixer

We designed the differential LNA to have 50-ohm input and output impedances, hence the input match of the mixer will be 50 ohms also. The 500 ohm output impedance of the mixer is designed to match to an 'off-chip' filter, and is most easily achieved by having 500-ohm mixer load resistors.

We choose an initial Rs value of 10 ohms and using the CMOS14 process results in the following analysis:

Using the spice model data for the Agilent CMOS14 0.5um we have:

L = 0.6um,
μ0 = 433 cm^2/(V*s),
θ = 0.5,
Rs = 10 ohms,
VT = 0.67V
Vsat = 1.73E^5 m/s

First convert numbers to metric format:

μ0 = 433 cm^2/(V*s), = 0.433 m^2/(V*s
L = 0.6um, = 0.6E^{-6}m.

$$\mu_1 \cong 0.433 + 2x0.1x1.73E^5.0.6E^{-6} = 453$$

The equations of section 4(ii) were entered into a spreadsheet, along with a range of Vod from 0.01 to 5V.

From **Table 2-2** we can see that to achieve a minimum IM3 value of 20dBm we require a VOT of 1V. As Vt for this process is 0.67V, the value of Vgs will be (1-0.67) = 0.33V.

Estimation of Input 1dB Compression point and IIIP3 vs Vod

Enter L	**0.6**	**um**	6E-07		
Enter Vsat	**2.40E+07**		2.40E+07		
Enter uo	**4.33E+02**		4.33E-02		
Enter theta	**0.5**				
Enter Rs	**20**	**ohms**	u1 =	**1.44E+01**	

Vod (V)	**1dB Comp** (dBm)	**IIMP3** (dBm)				
0.01	-8.35	1.29	1.33E-03	1.45E-04	1.00250752	1.010055
0.1	2.13	11.77	1.33E-02	1.45E-03	1.02507517	1.102816
0.2	5.65	15.29	2.66E-02	2.89E-03	1.05015035	1.210662
0.3	7.90	17.54	3.99E-02	4.34E-03	1.07522552	1.323538
0.4	9.62	19.26	5.32E-02	5.78E-03	1.10030069	1.441444
0.5	11.05	20.68	6.65E-02	7.23E-03	1.12537587	1.56438
0.6	12.28	21.91	7.98E-02	8.67E-03	1.15045104	1.692346
0.7	13.37	23.00	9.31E-02	1.01E-02	1.17552622	1.825343
0.75	13.87	23.51	9.97E-02	1.08E-02	1.1880638	1.893727

Table 2-2 Calculated 1dB compression point & 3rd order intercept point.

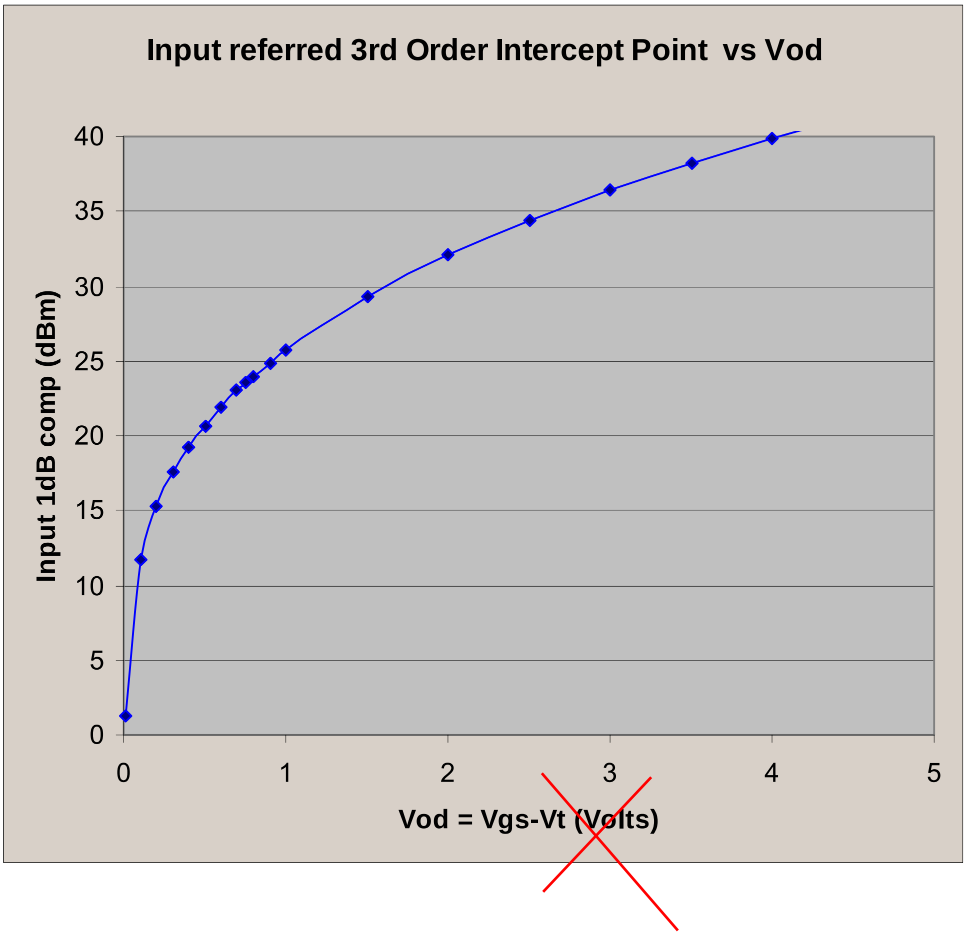


Figure 5 Plot of Input referred 3rd order intercept point (IIM3 dBm) with voltage overdrive (Vgs-VT)

With voltage gain goal of 7.5dB we can calculate gm [2]

$$\text{Conversion Gain C.G} \approx \frac{2}{\pi}\left(\frac{R_L}{Rs+\frac{1}{gm}}\right) \quad \text{--- (1)}$$

We can re-arrange to get gm for our voltage gain of 7.5dB ie

$$\text{Convert dB to conversion gain} =$$

$$10^{\frac{7.5}{10}} = 5.62$$

$$\therefore$$

$$5.62 \approx \frac{2}{\pi}\left(\frac{500}{10+\frac{1}{gm}}\right)$$

Alternatively we could derive gm assuming we wish to design for minimum noise as in the case of the LNA. To find the optimum gate width (W) for low noise we can use the following expression [3]:

$$W_{opt} \approx \frac{1}{3\omega.L.C_{ox}.R_{gen}}$$

Where Rgen is the resistance of the source connected to the mixer input – typically we design this for 50 ohms.

$$Cox = \frac{Eox}{TOX} = \frac{3.46 * 8.85E^{-12}}{9.06E^{-9}} = 3.37E^{-3}$$

In this case W would be:

$$W_{opt} \approx \frac{1}{3x2.5E^{9}x2\pi\ x0.6E^{-6}x3.37E^{-3}x50}$$

$$W_{opt} \approx 315\text{um}$$

For our example however, we will be designing the mixer to have a specific gain and output load impedance, therefore rearranging the earlier equation (1) we can obtain gm ie

$$gm = \left[\frac{2}{\pi}.\frac{R_L}{Vgain} - Rs\right]^{-1}$$

$$gm = \left[\frac{2}{\pi}.\frac{500}{5.62} - 10\right]^{-1} = 0.0199$$

Now with gm calculated we can W (Assuming we take the minimum gate length to be 0.6um and assume a current of 3mA) ie.

$$gm^2 = 2.Kp.\frac{W}{L}.I_{DS} \text{ re - arrange to get W}$$

$$W = \frac{gm^2.L}{2.Kp.I_{DS}} = \frac{(0.0199)^2 x\ 0.6}{2 \times 171E^{-6} \times 3E^{-3}}$$

$$W \sim 231 gm = \frac{2.I_D}{(Vgs - V_T)}$$

For a first run simulation we can assume all gate widths to be 231, however we need to ensure that the switching FETs are driven by a gate overdrive value of between 200mV to 400mV.

$$gm = \frac{2.I_D}{(Vgs - V_T)}$$

$$gm = \frac{2x3E^{-3}}{(0.3)} = 0.02 \quad \text{now find W from}$$

$$gm^2 = 2.Kp.\frac{W}{L}.I_{DS}$$

$$W = \frac{gm^2.L}{2.Kp.I_{DS}} = \frac{(0.02)^2 x\ 0.6}{2 \times 171E^{-6} \times 3E^{-3}} \Rightarrow W \sim 233$$

We can now run our first simulation by assuming all devices have W = 233um and L = 0.6um, tail current of 6mA, source resistors of 10 ohms and load resistors of 500 ohms.

5.1 BALUN DESIGN [5]

For the purposes of the simulation we need to convert the differential inputs and outputs of the mixer to single ended source and load impedances. The device that achieves this balanced to un-balanced transformation is known as a **'Balun'**.

A typical balun circuit is shown in **Figure 6**, where the 'Zin' block allows the input impedance to be measured, resulting in the correct value of 50 ohms.

In this case the impedance transformation is 1:4 ie 50 ohms to a differential impedance of 200 ohms.

If we wish to add a load to each port then they will be half this at 100 ohms – see example

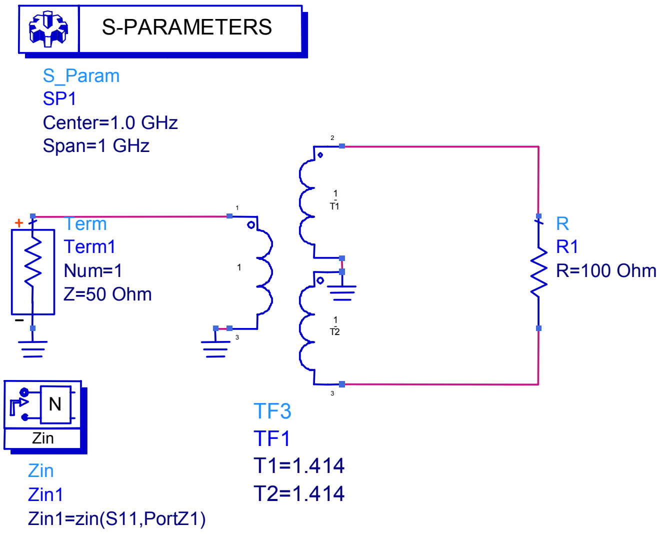


Figure 6 Schematic diagram of a balun transformation. In this example the grounded center tap does not effect the transformation. The 'Zin' block allows the input impedance to be measured and will result in the correct value of 50 ohms.

The general equation for balun transformer is:

$$\frac{Z_{OUT}}{Z_{IN}} = \left(\frac{N_{sec}}{N_{prim}}\right)^2$$

Where Nsec = Number of turns on secondary winding & Nprim = number of turns on the primary winding

Example: We have designed our differential mixer inputs to be 100ohms. Determine the turns ratio for the balun to match to a source impedance of 50ohms.

As we said earlier the 'differential' impedance will be twice that required for each of the output ports, so we need to transform from 50ohms to 200 ohms ie 1:4.

$$\frac{200}{50} = 4 = \left(\frac{N_{sec}}{N_{prim}}\right)^2 \therefore \frac{N_{sec}}{N_{prim}} = 2$$

The circuit of the balun for ADS simulations is shown in **Figure 7**.

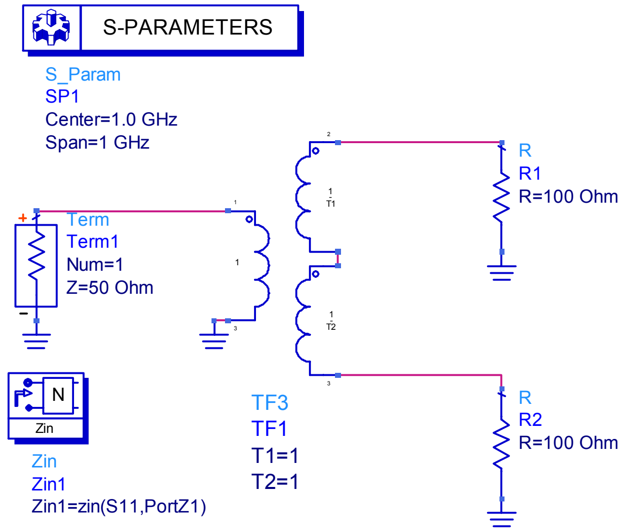


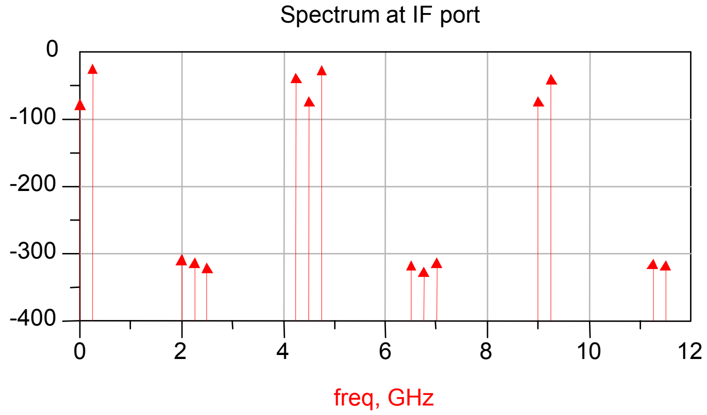
Figure 7 ADS example simulation of the 1:4 balun. Note we will still get a 50 ohms input impedance is we disconnect the 100 ohm loads and connect a 200 ohm impedance across the output of the balun.

6 SIMULATIONS [6]

The first simulation shown in **Figure 10** shows a harmonic balance simulation of the mixer to determine conversion gain, DBB & SSB noise figure.

IF_spectrum

Eqn IF_spectrum=dBm(HB.Vout)



Eqn ConversionGain=dBm(mix(HB.Vout,{-1,1}))-RF_pwr

freq	ConversionGain
250.0MHz	10.29

noisefreq	NFdsb	NFssb
250.0MHz	5.965	9.281

freq	Mix	
	Mix(1)	Mix(2)
0.0000 Hz	0	0
250.0MHz	-1	1
2.000GHz	2	-1
2.250GHz	1	0
2.500GHz	0	1
4.250GHz	3	-1
4.500GHz	2	0
4.750GHz	1	1
6.500GHz	4	-1
6.750GHz	3	0
7.000GHz	2	1
9.000GHz	4	0
9.250GHz	3	1
11.25GHz	5	0
11.50GHz	4	1

Figure 8 Results from the simulation of Figure 10, showing the output carrier spectrum, conversion gain & noise figure, together with the mix table of output carriers (explained more in the text). Note that for our example we have combined the differential outputs using a 'Balun' and thus we have 3dB more conversion gain ~ 10dB where we have designed for a single differential stage RF gain of 7.5dB. Also note that the double sideband noise figure is ~3dB worse than the single sideband value, as both the IF and 'Image' carriers are taken into account on the DSB noise figure.

In the simulation of **Figure 10**, balun (balanced to un-balanced) transformers have been used to provide single ended inputs and outputs and will thus combine the differential IF signals to give a 3dB increase in the differential conversion gain ie 10dB (With the mixer designed to give a differential gain of 7.5dB).

The next ADS harmonic balance simulation (**Figure 12**) is set up to sweep the local oscillator input power and plot the resulting DSB noise figure and conversion gain. This simulation is useful when optimizing the LO switching circuit (ie W and V overdrive).

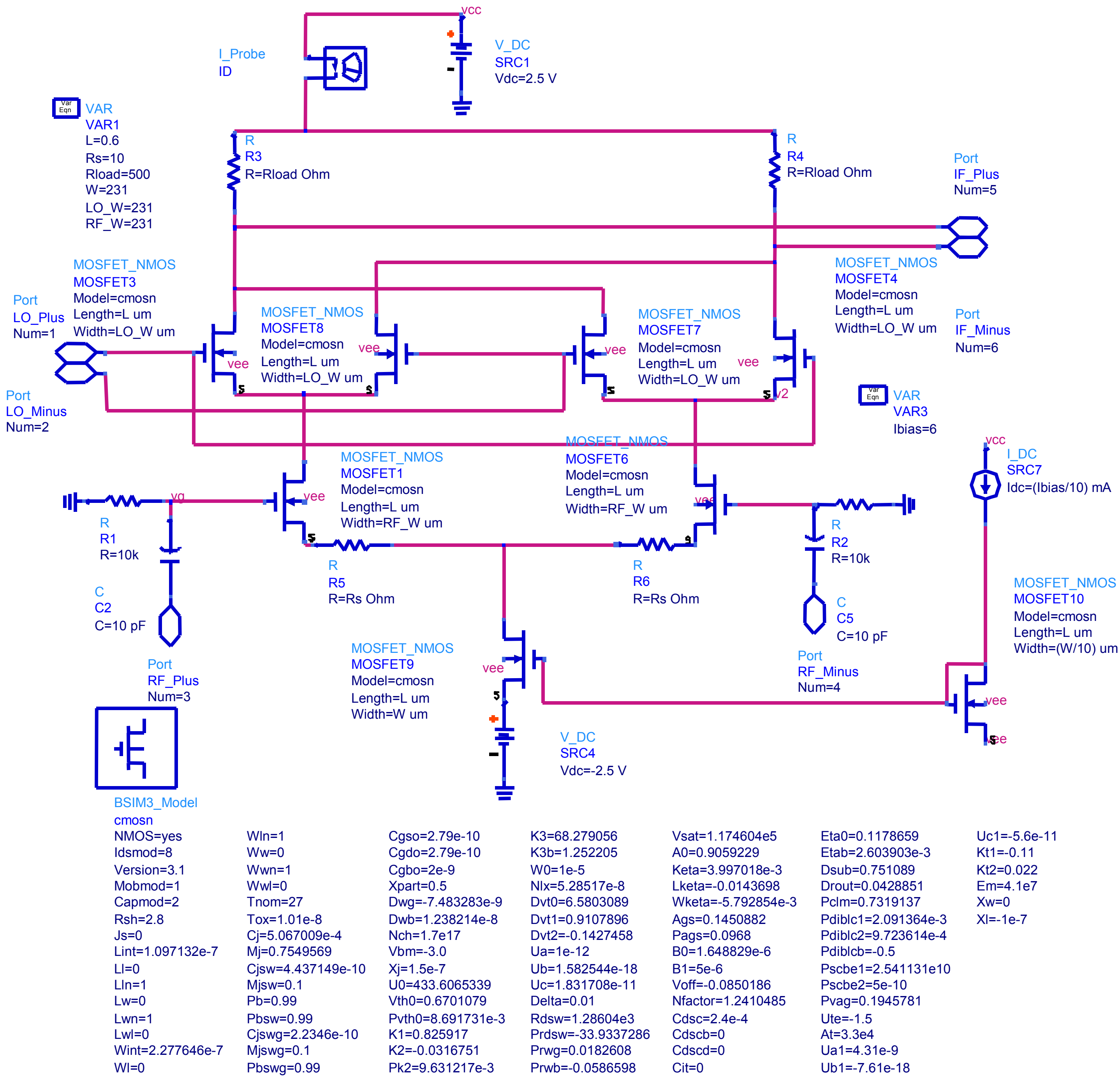


Figure 9 ADS schematic of the Gilbert Cell Mixer circuit. For this simulation the gates of the RF MOSFETS 1 & 6 are grounded (via 10K resistors that block any RF signal) and to ensure correct bias the tail MOSFET 9 is connected to a negative supply. The tail current is set by the 'bias' variable – in this case the controlling MOSFET 10 has a width = W/10, hence the bias = bias/10. The RF, LO and IF ports are terminated with a port so that the circuit will appear as a 6 port device when used in simulations as a sub-circuit..

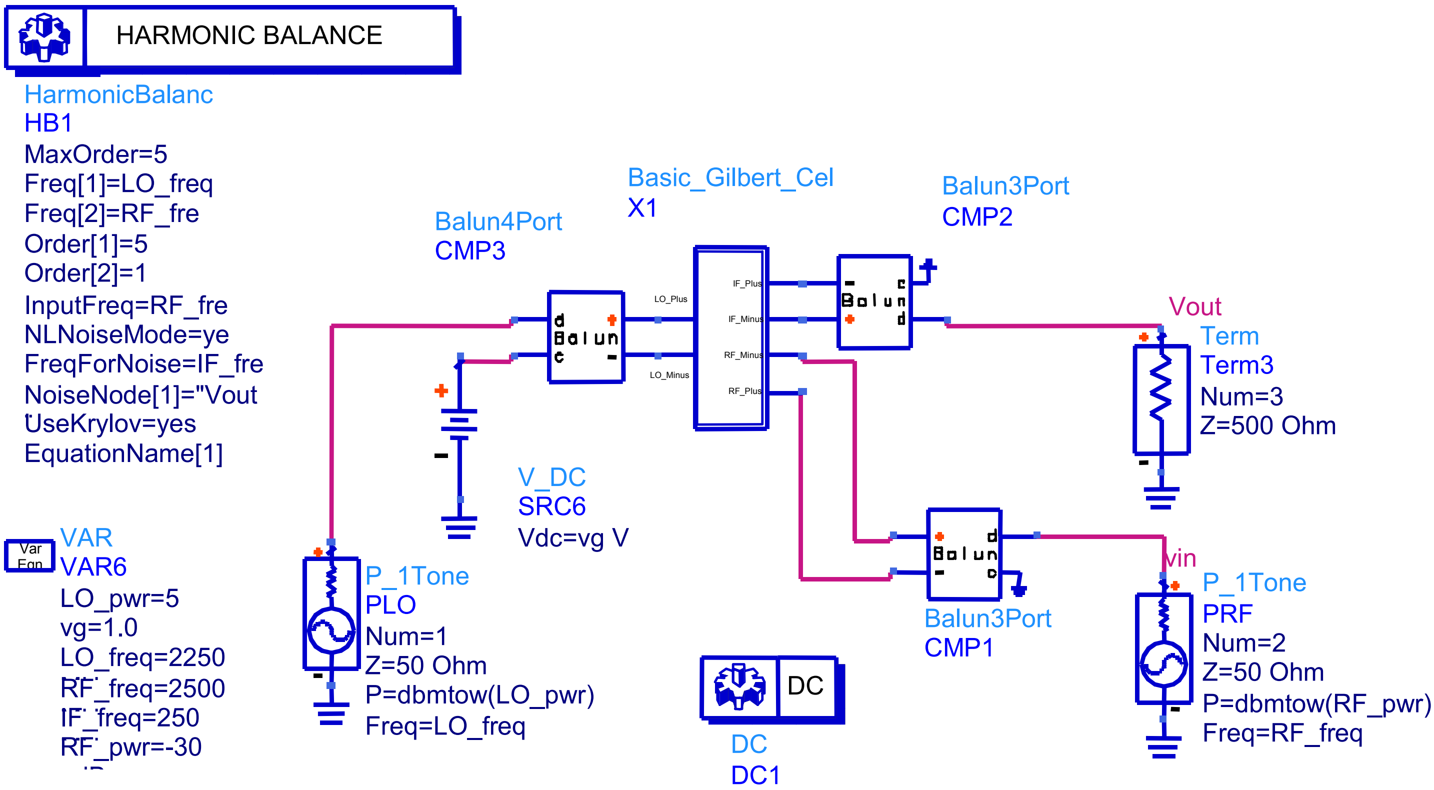
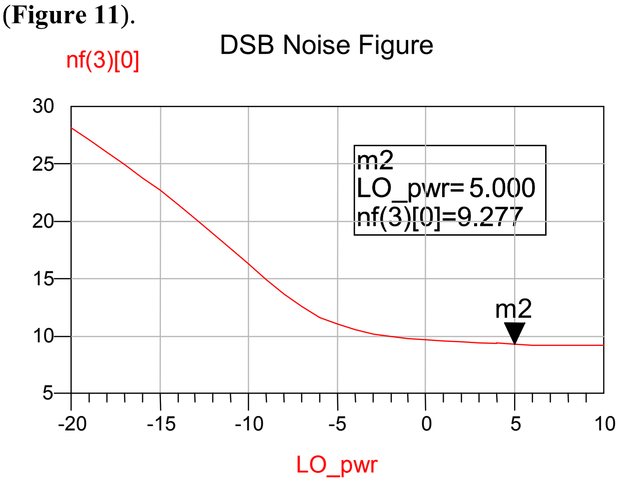


Figure 10 Harmonic balance simulation of the Gilbert Cell Mixer. As the inputs and outputs are differential balun transformers have been added to convert to single ended inputs and outputs. The 500-ohm load Term3 correctly terminates the mixer 500-ohm output impedance. The RF frequency is set to 2500MHz (RF_freq), Local oscillator frequency to 2250MHz (LO_freq) , resulting in an IF frequency of 250MHz (IF_Freq).. For correct switching of the LO transistors the variable vg needs to be set to 1V – running the simulation this gives Vgs across the switching transistor of ~ 1V.

The resulting plots of DSB noise figure and conversion gain vs local oscillator input power are shown in (**Figure 11**).



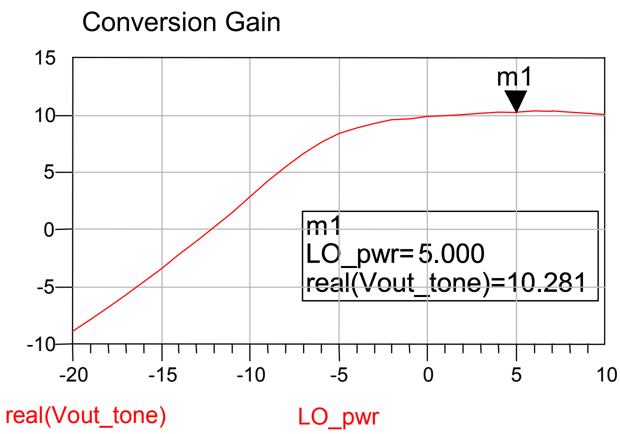


Figure 11 Results from the simulation shown in Figure 12, showing DSB noise figure and conversion gain vs LO drive level.

freq	Mix
LO_pwr=-20.000, Mix(1)	
0.0000 Hz	0
250.0MHz	0
2.000GHz	1
2.250GHz	1
2.500GHz	1
4.250GHz	2
4.500GHz	2
4.750GHz	2
6.500GHz	3
6.750GHz	3
7.000GHz	3
8.750GHz	4
9.000GHz	4
9.250GHz	4
11.00GHz	5
11.25GHz	5
11.50GHz	5
LO_pwr=-20.000, Mix(2)	
0.0000 Hz	0
250.0MHz	1
2.000GHz	-1
2.250GHz	0
2.500GHz	1
4.250GHz	-1
4.500GHz	0
4.750GHz	1
6.500GHz	-1
6.750GHz	0
7.000GHz	1

Table 2-3 From the simulation of Figure 12 a table was generated giving values of mix(1) and mix(2) and their associated frequencies. Thus if we select vout with and index of {0,1} we will be selecting the IF frequency at 250MHz.

The plots of **Figure 11** show us that to maintain optimum conversion gain (and minimum DSB noise figure) we need an LO drive power in excess of 0dBm.

For the 3rd ADS simulation we want to simulate the gain compression characteristics of the mixer. The simulation for this is shown in **Figure 15** with the resulting plots shown in **Figure 13**. The top plot shows the graphical gain compression characteristic. Here we have a plot of linear gain shown in blue (Line) and a plot of the compressing IF gain response. As the input RF is increased the IF gain begins to decrease as the mixer saturates. Where the IF gain deviates by 1dB from the linear gain is the gain compression point. Markers 1 & 2 differ by 1dB for the same input power of –4.5dBm. The output power at gain compression given by marker 3 is 4.382dBm.

The variable '**inpwr**' is generated from the 'Gain compression simulation box giving a slightly higher input power level of –3.189dBm. To calculate the output power at gain compression we add the linear gain (Gain variable) and subtract 1dB (as the IF plot has dropped 1dB from the linear gain value).

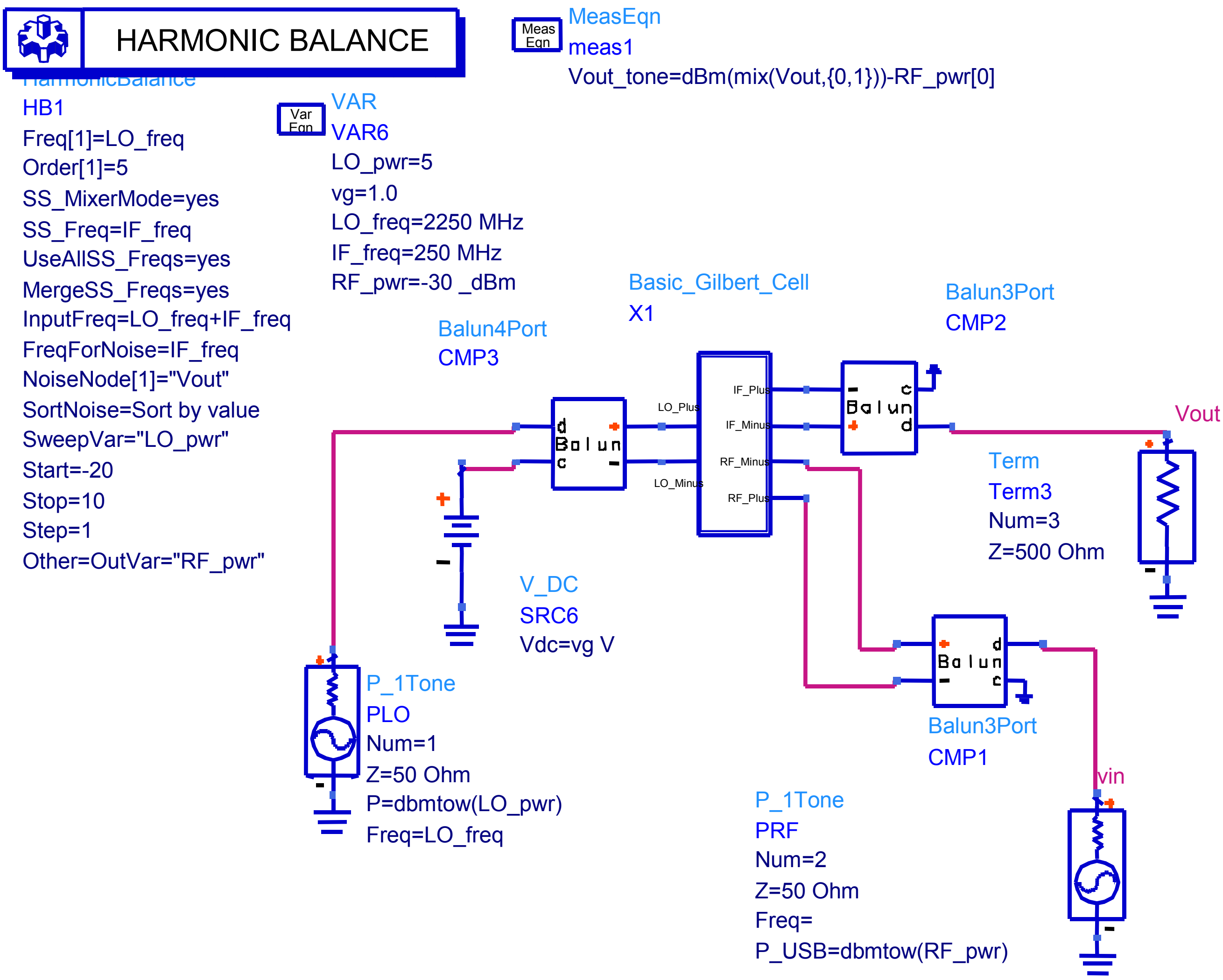
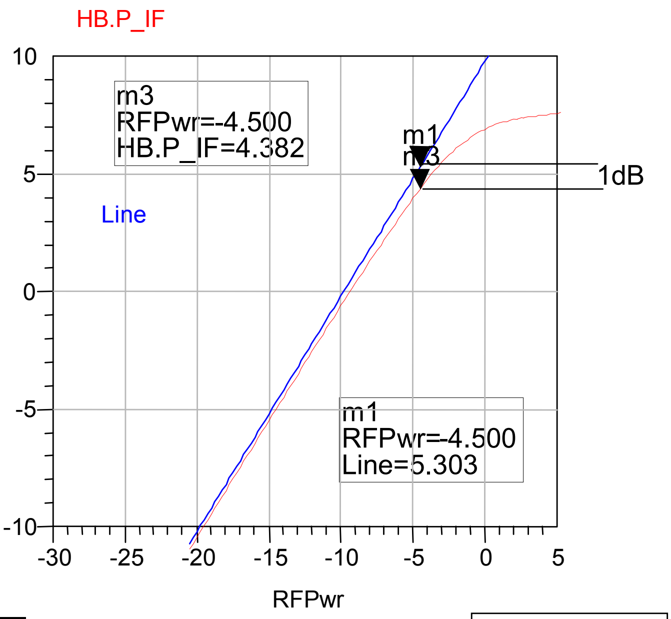


Figure 12 This ADS Harmonic simulation is setup to sweep the Local oscillator power (LO_pwr) from –20dBm to +10dBm in 1dBm steps. The RF input frequency = LO_freq (2250MHz) + IF_freq (250MHz) = 2500MHz (Set to a power (RF_pwr) of –30dBm. This simulation is slightly different to the one shown in Figure 10 in that upper sideband is selected (by setting the power variable P_USB in 'PRF one tone source). In the Harmonic simulator box, setting SS_Mixer_Mode to 'yes' causes the simulator to solve for all small-signal mixer sidebands. This default option requires more memory but delivers more accurate results. The measurement equation Vout_tone selects the IF frequency (in dBm) using the mix function then substracts the RF_pwr (at fundamental frequency) to give conversion gain. In order to select the IF frequency at 250MHz we need to know what the mix terms are, the first value 0 is mix(1) and the second value 1 is mix(2) if you plot a table of mix(1) and mix(2) you will find this equates to an output carrier at 250MHz – this is shown in Table 2-3.



Eqn Line=HB1.HB1.HB.RFPwr+Gain[0]

Gain[0]
9.803

Eqn Gain=10*log10(HB.P_IF-HB1.HB1.HB.RFPwr)

Output_Gain_Comp	inpwr
5.613	-3.189 dBm

Eqn Output_Gain_Comp=Gain-1+inpwr[0]

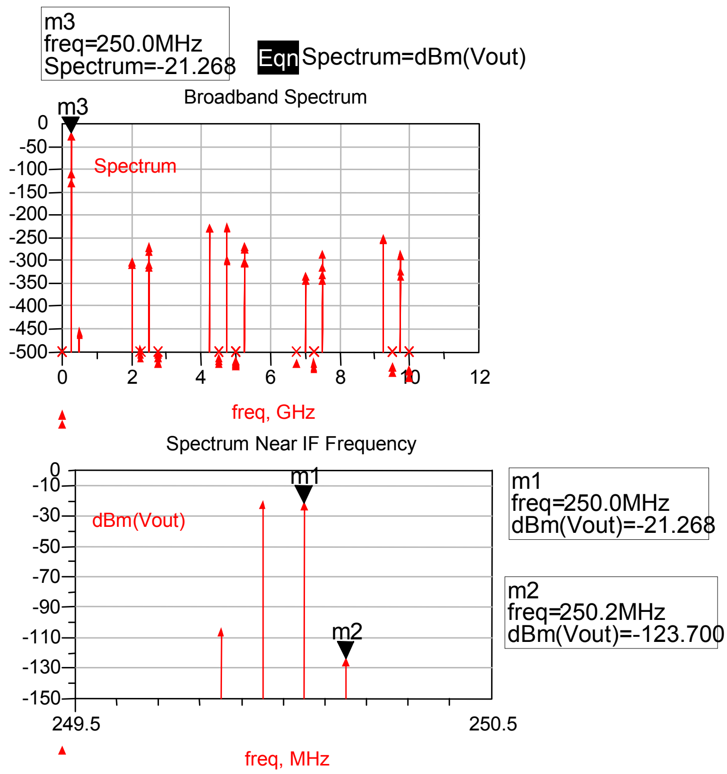
Figure 13 The gain compression from the simulation shown in Figure 15. The Line equation generates a linear gain line shown in blue. The red plot shows the non-linear IF gain compression characteristic. Markers1 & 3 are placed at the same input power where the two lines diverge by 1dB. The input power at this point is defined by marker3 at –4.5dBm. Using the 'Gain compression' result – variable 'inpwr' yields a slightly higher value of –3.19dBm. The output power where gain compression occurs is simply inpwr plus the linear gain –1dB (as the 'linear' gain at this point has dropped by 1dB).

The 4th and final ADS simulation calculates the 3rd Order intercept point of the mixer. From the gain compression simulation we can estimate the IM3 intercept point to be ~ 10dB above the Output Gain compression point – in this case 10 + 4.382 = 14.38dBm (or 15.61dBm if you take the other calculation).

Again there is another useful simulation tool that will directly calculate the IM3 of the mixer ie the **IP3out** function.

So far we have designed the mixer to achieve our specification in terms of DSB noise figure, gain, output impedance. But there appears to be some difference in the calculated IM3 and the simulated IM3. We know that the resistor Rs has an effect on the gain compression and IM3 (due to de-generation) in which gain is exchanged for linearity [3].

The simulated results agrees with the predicted IIM3 shown in **Table 2-2**, however the resulting IIM3 and gain compression point not only depends on the voltage overdrive of the input transistors but also (to a lesser extent) the switching transistors. To achieve the required linearity the bias current was increased from 6mA to 9mA. This of course will degrade the DSB noise figure slightly and was now found to be 10.3dB (from 9.2dB).



Eqn TOIoutput=1.5*m1-0.5*m2

TOIoutput
29.948

ConvGain	Input_IM3	Output_IM3
8.732	21.216	29.948

Figure 14 Plot of the resulting IM3 simulation shown in Figure 16. The first plots show all the mixing products and harmonics attenuated by the IF band-pass filter. The second plot shows the two input carriers together with their IM3 products, resulting in an intercept point of ~ 30dBm using a graphical method. The table shows the direct calculation of the Intercept point using the IP3out simulation box in conjunction with the Harmonic balance simulator.

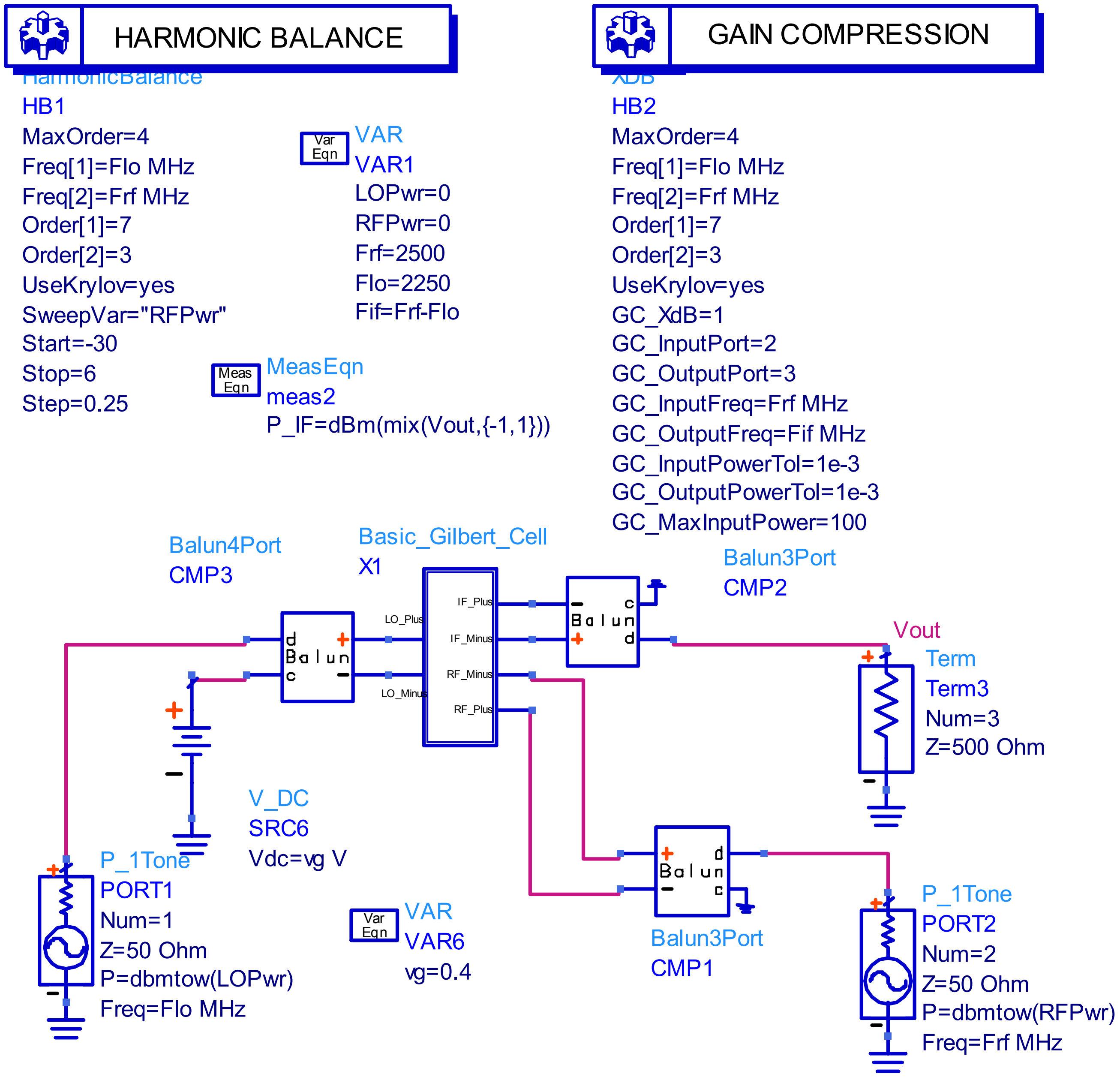


Figure 15 ADS simulation to derive the gain compression characteristic of the mixer. The P_IF function produces and ideal straight line of linear gain. Thus when the compressed gain of the mixer is added to the graph, the point at which the two lines diverge by 1dB is the gain compression point. To simply the process of finding the gain compression point the 'Gain Compression' simulation block is added to the schematic. The two Input frequencies are declared ie RF & LO, together with the output frequency Fif. The ports have also to be defined, in this case the input port (RF) is port2, and the output port (IF) is port 3.

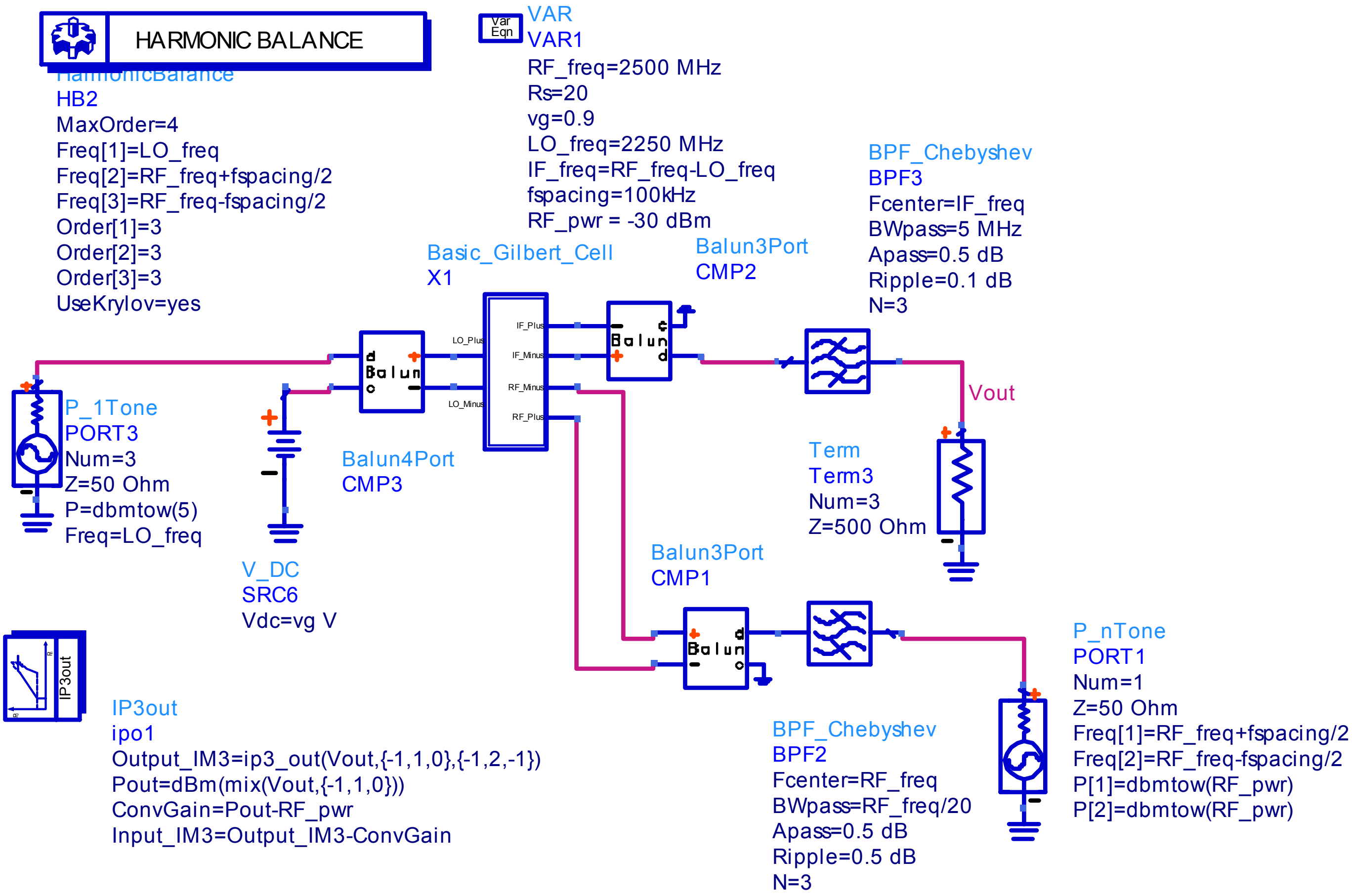


Figure 16 ADS simulation to determine the 3rd order intercept point of the mixer. In this simulation two RF carriers are set to one 50kHz below the RF_freq of 2500MHz and the other one 50kHz above the RF_freq. The spacing of the carriers set by fspacing is therefore 100kHz. The clever part of the simulation is the IP3out where the IM3 output intercept point (dBm) is calculated from the two IF frequencies at 250MHz + 50KHz and –50KHz. Note the filter is added to the IF to filter out the effects of RF breakthrough.

Luckily noise and IIM3 are fairly independent of each other as IIM3 is not particularly sensitive to L and W, so important for noise optimization is not a variable of IIM3!

It is of course possible to improve the linearity of the mixer by adding voltage follower buffer stages on each output, this will have the added advantage in that the frequency response of the mixer will not be altered by the following stage input capacitance (assumimg inductive loading!).

We can now compare our simulation results to that of the specification. The table below (**Table 4**) shows the original key parameters.

Parameter	Specification	Predicted	Units
Frequency	2.45 to 2.85	2.45 to 2.85	GHz
Noise Figure	< 10	10.3	dB
IM3 Intercept Point (Intput)	>20	21.2	dBm
Voltage Gain	>8	8.2	dB
Power consumption	<100	90	mW
Source impedance	50	50	ohms
Load impedance	500	500	ohms
Voltage Supply	± 2.5	± 2.5	V

Table 4 Predicted performance of the mixer against the original specification.

7 DESIGN COMPLETION

The design so far, generally meets the specification given and the design process has shown how gain, noise figure and linearity are interdependent and thus trade offs need to be made at the beginning of the design process. In particular the fact reducing the W/L ratio increase the overdrive voltage and hence linearity but reduces the gain and increases the noise figure!

There are a number of things that can be done to improve the design including replacing Rs with a lossless inductor, using separate tail current sources.

7.1 LOSSLESS SOURCE DEGENERATION

We saw how the introduction of the source resistor Rs greatly improved the linearity of the mixer. However, using a resistor will cause a voltage drop and hence an needed increase in the voltage rails to ensure the devices remain in saturation. Another important problem with using a resistor is the noise it will generate.

The solution is to use a lossless component – an inductor.

There are a number of issues when using an inductor Ls one being the circuit will now be more frequency dependant. A second issue is that inductors used in CMOS tend to have low Quality (Q) factors – typically 2 to 3. Thirdly the addition of the source inductor will have a significant effect on the input impedance (Re) and will need to be checked to ensure the impedance is not negative (which may cause instability!).

We can set up a S-parameter simulation to alter the value of Ls to determine the input resistive impedance. To ensure the simulation will work the switching transistors are biased on using two voltage sources.

The S-parameter simulation is shown in **Figure 18** with the resulting impedance plot shown in **Figure 17**.

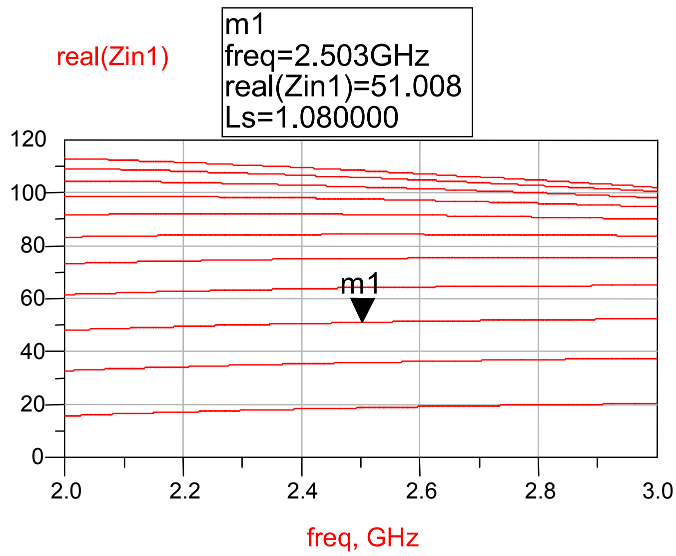


Figure 17 Input impedance result from the simulation shown in Figure 18. The value of the source de-generating inductor (Ls) has been varied between 0.5nH and 5nH. We see that with an inductance of ~1.1nH the input impedance is 50 ohms.

With a source de-generation inductor of value 1.1nH results in an input impedance of ~ 50 ohms (Rs =18ohms). If 100 ohm input impedance is required (If connecting to a balun) then Ls = 3nH.

The simulation shows that the mixer circuit is stable as the value of Re is positive and **NOT** negative (the resulting input return loss S11 is negative as it should be)

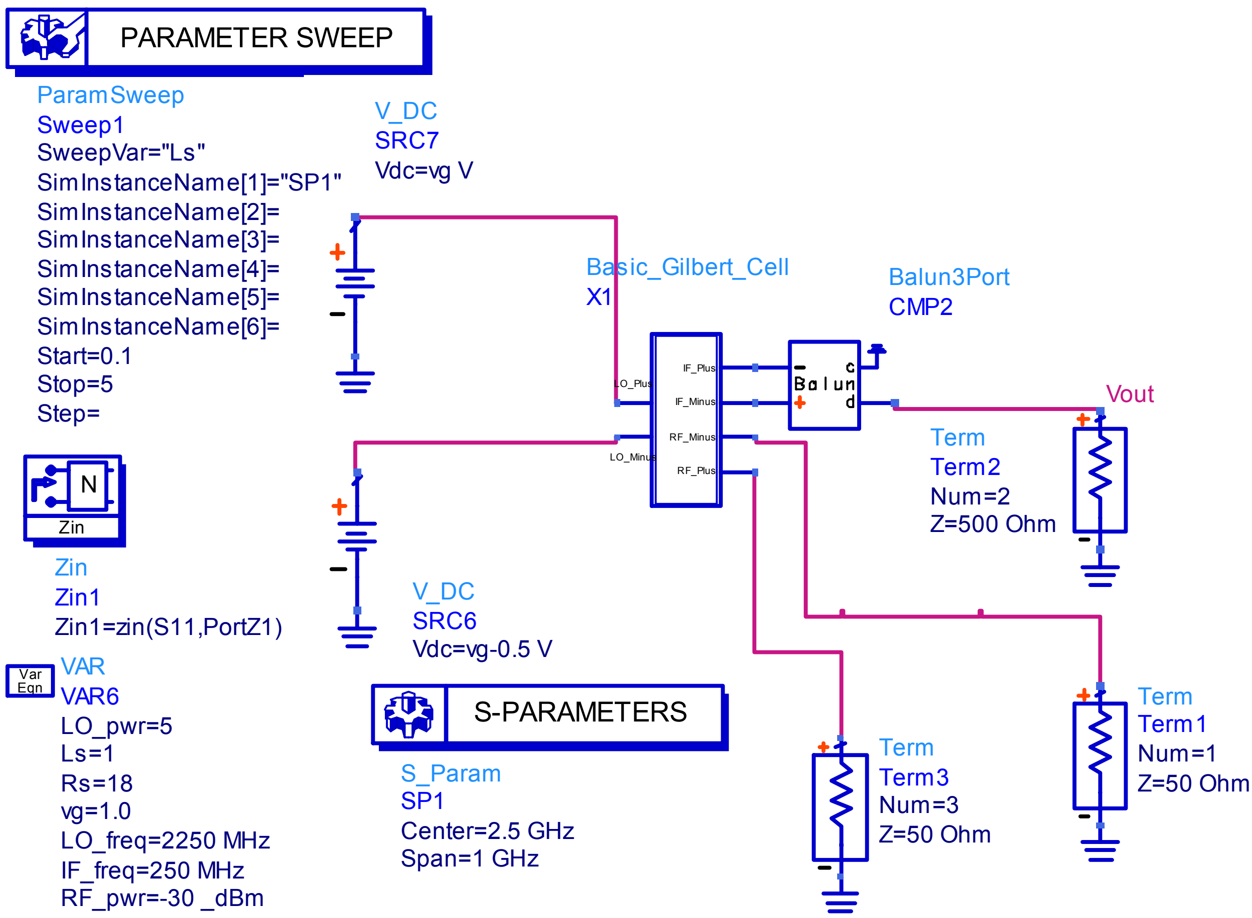


Figure 18 S-Parameter simulation to determine the effect of varying the source de-generation inductance (Ls) on the input impedance (Re). The Parameter sweep box is set to vary the variable Rs from 0.1 to 5nH and applies to the simulator 'SP1'. This is to allow an optimum value of Ls to be used for an input match ie Ls = 1nH for 50 ohms and Ls=3nH for 100 ohms

When we add an inductor we are in fact adding an inductor with an added series resistance dependant on the Q (Quality factor) of the inductor:

$$Q = \frac{\omega . L}{2 . R}$$

For example at 2.5GHz the series R will be ~ 23ohms for a 5nH inductor.

Therefore, when simulating the circuit with a source degenerating inductor we use the **INDQ** model if we know the unloaded Q of the inductor (Depends on the topography & process) or if we wish to determine the Q required we can use an ideal inductor in series with a resistor (**SRL** series resistor inductor model).

7.2 ALTERNATIVE DE-GENERATION SCHEME

The design to date, has used resistors followed by inductors to provide source de-generation in order to meet the given linearity specification. The source resistor (Rs) was substituted with a source inductor (Ls) to improve the voltage swing of the mixer. There is however, another topography that uses a de-generating source resistor but doesn't degrade the voltage swing [7]. And the circuit is shown in **Figure 19**. In this configuration there is one de-generating resistor connected between the two sources, together with a separate current source. (Note that for best performance theses current sources need to be well matched).

Each of the two current sources set the current through each side of the differential stage and are set to **Itail x 2**.
As the potential at each source should be the same (hence the need for well-matched current sources) there will be no voltage drop across the de-generating resistor (Rs), thus no degrading the voltage swing of the mixer.

8 CONCLUSION

This tutorial has described the design process of a MOS Gilbert cell (differential) mixer with ADS simulations given to predict the various circuit parameters of gain, noise figure and linearity
The design process depends on the applications key driving parameter, where the mixer could be designed for optimized Gain/Output load, DSB Noise figure or Linearity.
As with the differential LNA, the design equations for predicting Intercept point & Compression point where given and showed that this parameter could be optimized, by using source de-generation.

In order to ensure a stable design (ie with no risk of oscillation) it was highlighted that the input impedance should be checked to ensure the input impedance is positive and not negative.
Finally, an alternative design scheme was given to allow a source resistor to be used without degrading the voltage swing capability of the mixer, by using two separate current sources in lieu of the single 'tail' current source in the conventional design.

9 REFERENCES

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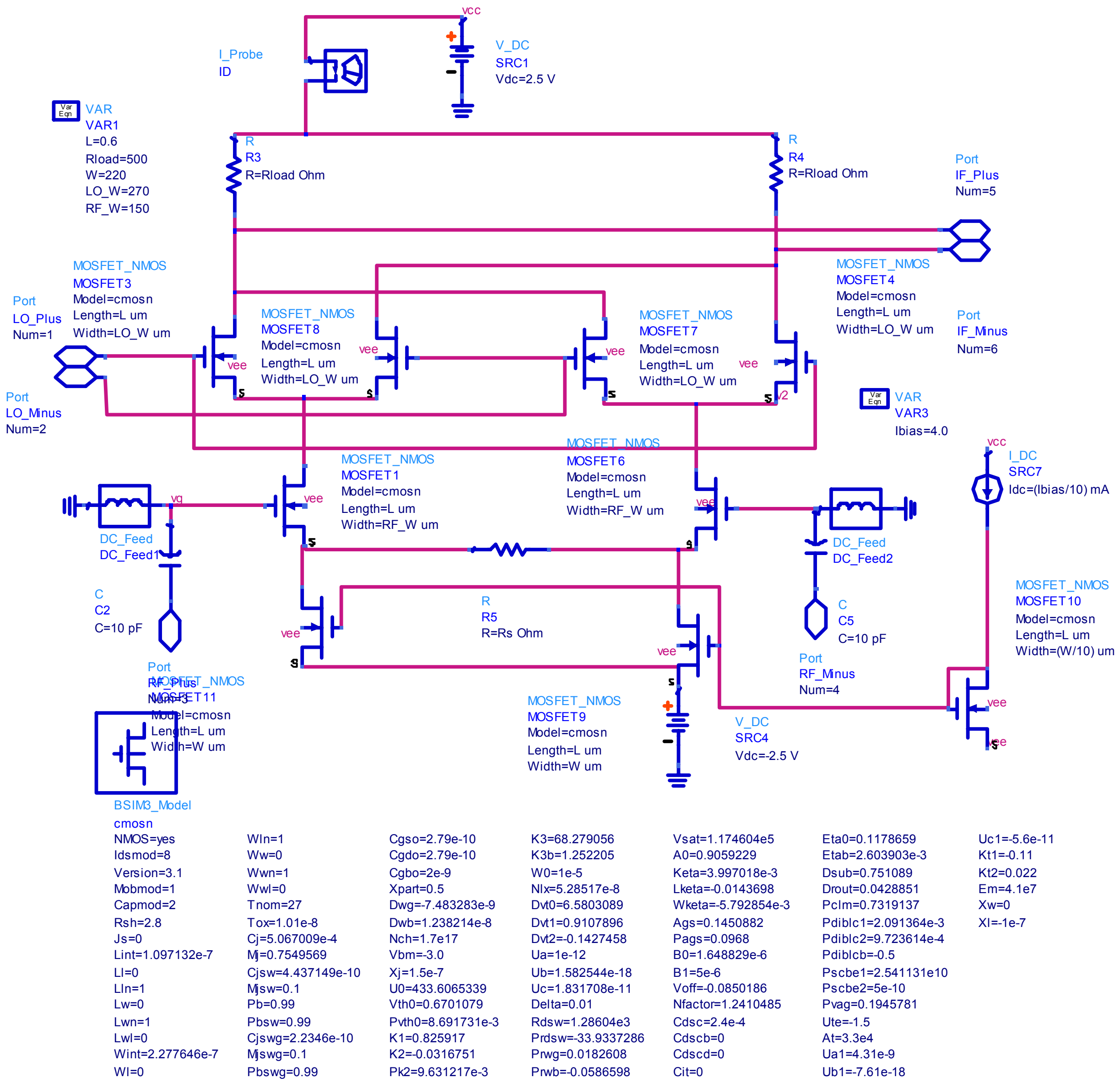

Figure 19 Gilbert Cell mixer with source de-degeneration (to improve linearity) but does not degrade the voltage swing of the mixer. The two 'tail' current sources need to be well matched for best performance.