Reflection Oscillator Design Tutorial

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ABSTRACT

This paper discusses the design of a basic reflection oscillator, using a distributed resonator and micro-strip circuit elements. The first design is of a fixed frequency oscillator and the second design uses a varactor for frequency tuning. Throughout the designs, Agilent ADS large and small circuits and simulations are given to verify each design stage.

INTRODUCTION

This tutorial describes the design of a 2GHz reflection oscillator building on the theory from the oscillator basics tutorial. In the first section a simple fixed frequency oscillator is described where the resonator consists of a similar element that resonates with the input of the reflection amplifier. The second design uses a varactor based resonator to allow frequency control of the oscillator to be used within a PLL.

FIXED FREQUENCY NEGATIVE RESISTANCE OSCILLATOR [2,3,4]

For this example we wish to design an oscillator for a fixed frequency of 2GHz. The device selected for this example is the Hewlett-Packard General Purpose Silicon Bipolar Transistor AT-42535.The S-parameters at 2GHz for this device are:

Bias V_{CE} = 8V I_C = 10mA;
S11 = 0.58∠155
S21 = 3.06∠55
S12 = 0.062∠55
S22 =0.48∠-38

From the data it is clear that the device will be stable in a 50ohm system but will it have the ability to oscillate when attached to a resonant load? To answer this we need to calculate the value of K, if it is greater than one then the device will not oscillate into a resonant load and we will need to add feedback to ensure K < 1:

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |D|^2}{2|S_{12}||S_{21}|} > 1$$

$$\text{where } D = S_{11}S_{22} - S_{12}S_{21}$$

We can simulate the FET to determine the stability factor K by using ADS circuit shown resulting in the table of K factor shown in **Table 1**.

freq	StabFact1
1.500GHz	1.120
1.542GHz	1.129
1.583GHz	1.138
1.625GHz	1.147
1.667GHz	1.158
1.708GHz	1.170
1.750GHz	1.183
1.792GHz	1.198
1.833GHz	1.213
1.875GHz	1.230
1.917GHz	1.248
1.958GHz	1.267
2.000GHz	1.288
2.042GHz	1.280
2.083GHz	1.274
2.125GHz	1.269
2.167GHz	1.265
2.208GHz	1.262
2.250GHz	1.260
2.292GHz	1.259
2.333GHz	1.259
2.375GHz	1.260
2.417GHz	1.263
2.458GHz	1.267
2.500GHz	1.272

Table 1 Resulting K factor for the FET when used in common-source mode.

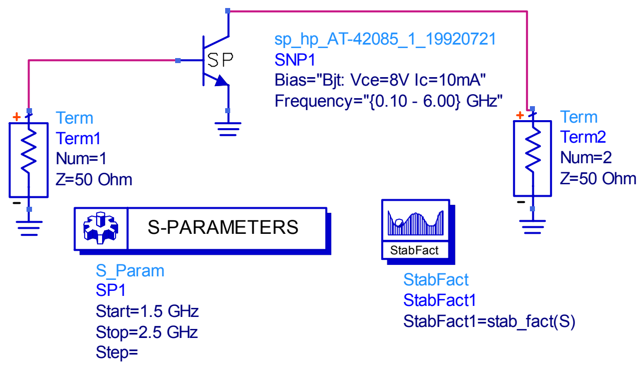


Figure 1 ADS simulation circuit to determine the stability factor K

From **Table 1** we can see the value of K at 2GHz was was **1.28** ie non-conditional stability. Therefore feedback is needed to reduce this below 1 and this is done by adding shunt feedback in the form of an inductor as shown in **Figure 2**.

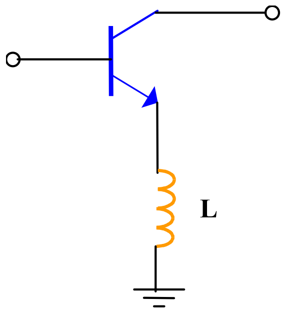


Figure 2 Addition of shunt feedback in the form of an inductor to modify the device S-parameters in the aim of making K less than one.

If we add, for example a 5nH inductor to the emitter, we can analyse the new circuit on a Microwave/RF CAD such as Agilent ADS. Analysis gives us a set of modified S-parameters and a new value of k:

freq	StabFact1
1.500GHz	0.939
2.000GHz	0.850
2.500GHz	0.898

freq	S(1,1)	S(2,1)	S(1,2)	S(2,2)
1.500GHz	0.429 / -11.658	1.646 / 65.202	0.213 / 87.637	0.799 / -18.613
2.000GHz	0.429 / -26.387	1.431 / 57.680	0.314 / 91.284	0.820 / -25.714
2.500GHz	0.418 / -17.443	1.220 / 52.442	0.381 / 80.178	0.789 / -30.357

An associated value of K of 0.85, implies that the device is now conditionally stable and may be liable to oscillation. But the value of K is only just below 1 and the magnitudes of S11 & S22 are still below 1 ie no negative resistance.

Therefore, we can try a different configuration - it is normal to use a bipolar transistor in the common base configuration as shown in **Figure 3**.

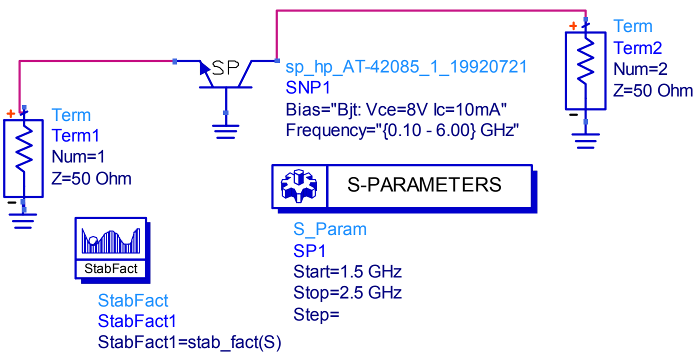


Figure 3 Common-Base configuration to boost negative impedance looking into the emitter and collectors.

Shown in **Table 2** are the new S-Parameters:

freq	StabFact1
1.500GHz	-0.914
2.000GHz	-0.802
2.500GHz	-0.792

freq	S(1,1)	S(2,1)	S(1,2)	S(2,2)
1.500GHz	1.012 / 169.289	2.057 / -29.604	0.062 / 151.800	1.126 / -18.784
2.000GHz	0.956 / 164.470	2.214 / -41.140	0.072 / 128.082	1.230 / -26.579
2.500GHz	1.218 / 158.068	2.313 / -54.156	0.166 / 150.721	1.298 / -35.052

Table 2 Common-Base S-Parameters & K factor with a 5nH inductor connected between the base and ground, now showing >1 magnitudes of S11 & S22 with the K factor < 1.

The value of k is now **- 0.802** ie < 1 conditionally stable, but now the magnitudes of S22 is greater than 1 but still < 1 for S11. If we were to use the device as it stands it would probably oscillate but only after a lot of tuning and probably would not work over a wide temperature range.

Again we can add a small inductance to the base to attempt to raise the magnitudes of S11 & S22 as shown in **Figure 4**.

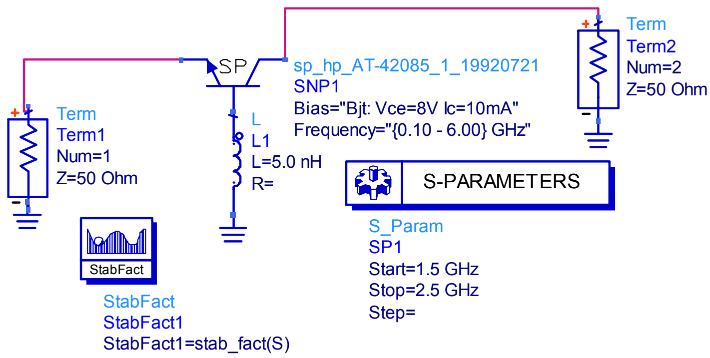


Figure 4 Common-Base configuration to boost negative impedance looking into the emitter and collectors.

The result of the simulation shown in **Figure 4** is shown in the plot of **Figure 5**.

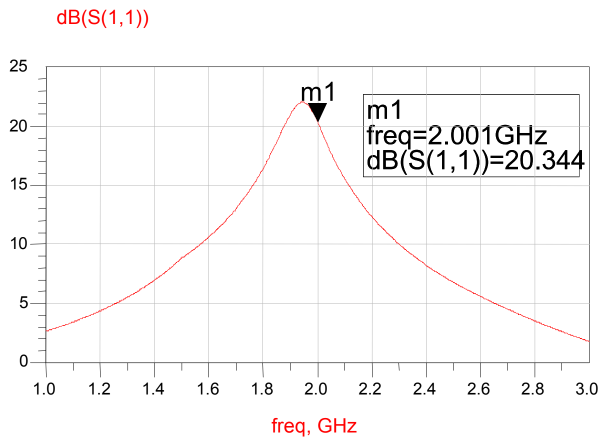


Figure 5 Input return loss plot of the ADS schematic shown in Figure 4, showing a high positive return loss required for the oscillator.

The resulting modified S-parameters and K factor are given in **Table 3**.

freq	StabFact1
1.500GHz	-0.856
2.000GHz	-0.587
2.500GHz	-0.429

freq	S(1,1)	S(2,1)	S(1,2)	S(2,2)
1.500GHz	2.763 / 155.978	3.906 / -45.647	0.762 / 163.487	1.860 / -29.955
2.000GHz	10.475 / 45.778	13.207 / -171....	4.471 / 60.654	5.210 / -149.885
2.500GHz	2.183 / 19.125	1.859 / 149.421	1.240 / 32.610	0.516 / -167.596

Table 3 Common-Base S-Parameters & K factor with a 5nH inductor connected between the base and ground, now showing >1 magnitudes of S11 & S22 with the K factor < 1

The value of K is -0.58 ie conditionally stable. Therefore we have a device that is capable of oscillating when presented with a resonant load on the emitter.

If we connect a component Γ_G ie on the emitter port, with a value of $\sim 1\angle\text{-}45$ then this should resonate the input. We can use a 50ohm load termination Γ_L as the value of S22' is > 1. If it was < 1 then we would have to modify the circuit configuration and/or vary the value of the base inductance until both $|S_{11}|$ and $|S_{22}|$ are > 1.

We can therefore, attach a capacitor to the input of the device to resonate it. Looking on a Smith chart a pure reactance of -45 degrees gives a corresponding reactance of 2.375*50 ohms (at 2GHz).

Therefore, the capacitor required will be:

$$C = \frac{1}{2\pi f X_c} = \frac{1}{2\pi * 2E9 * 2.375 * 50} = \boxed{0.67\text{pF}}$$

We now have the values required for the basic oscillator an ADS S-parameter simulation using the 'osctest' will confirm whether oscillation will occur. The simulation is shown in **Figure 6**.

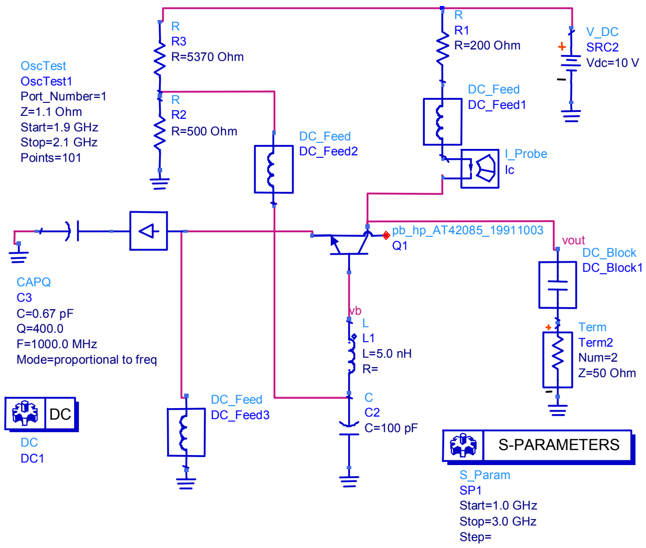


Figure 6 ADS S-Parameter simulation of the basic oscillator – see text for description of the simulation features.

The S-parameter model has been replaced with a non-linear model (Spice model) and as a result the DC bias circuit is needed to ensure the bipolar transistor is biased to 10mA with a Vce of 8V. The 200-ohm resistor is designed to drop 4V to give a Vce of 8V. The emitter is grounded via the DC_feed which is a block to all RF. The base inductor is RF grounded via the 100pF capacitor and this is where the base voltage bias is applied. The base resistors are set to give a 10mA collector current.

The 'Oscport' block will output the closed loop magnitude and phase of the circuit over the frequency range specified within the Oscport block. **Note** the magnitude has to be > 1 and have zero phase at the required frequency. The result of the simulation is shown in **Figure 7**.

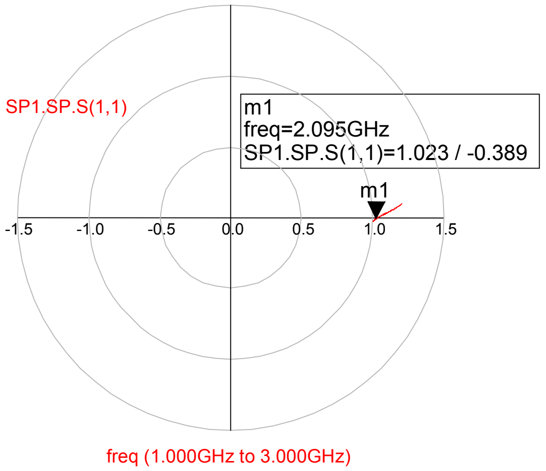


Figure 7 Polar magnitude/phase result of the S-Paramter simulation using the 'Oscport' block shown in Figure 6.

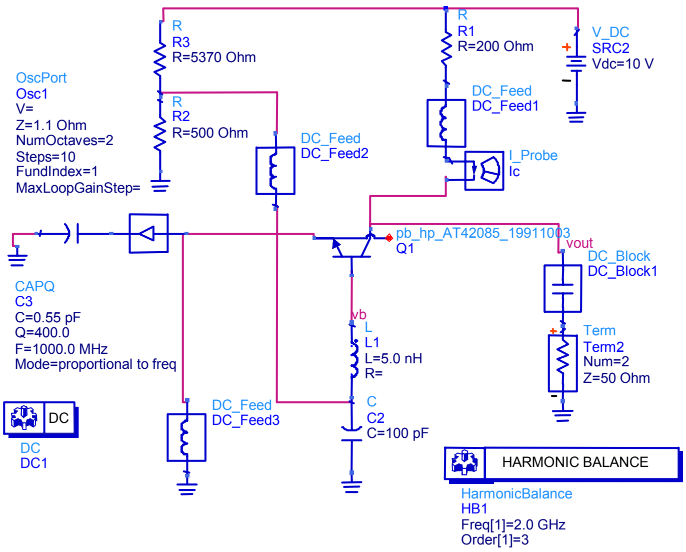


Figure 8 Harmonic balance ADS simulation of the basic oscillator design.

The S-parameter simulation shows that we do achieve a magnitude of >1 at zero phase at 2GHz. Therefore the next step is to perform a harmonic balance simulation to predict the output power spectrum and phase noise performance. The Harmonic balance ADS simulation is shown in **Figure 8**.

The Harmonic balance simulator box is set to the following values:

Freq: 2000MHz , order 3.

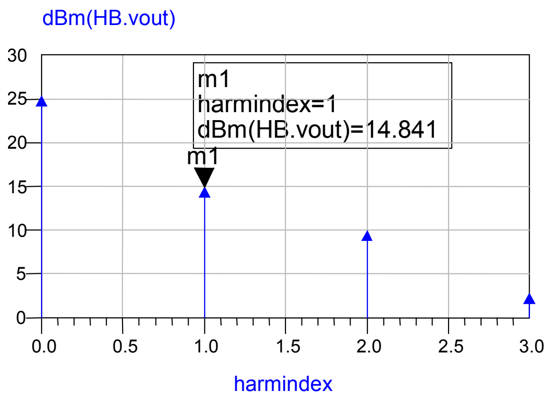
Noise(1): Select Log, Start 100Hz, Stop 10MHz, Points/decade 10, Select Include FM noise.

Noise(2) Select "vout", Select Non linear Noise & Oscillator

Osc Osc Port name "Osc1"

All other parameters can be left to their default values.

The resulting frequency spectrum is shown in **Figure 9** with the corresponding phase noise plot shown in **Figure 10**.



harmindex	HB.freq
0	0.0000 Hz
1	2.003GHz
2	4.006GHz
3	6.010GHz

Figure 9 Oscillator output power spectrum. The frequency of the fundamental is given by index 1 which has a frequency of 2.003GHz and an output power of 14.8dBm. Note to give the correct frequency the tuning capacitor on the bipolar emitter junction was slightly lowered from 0.67pF to 0.55pF.

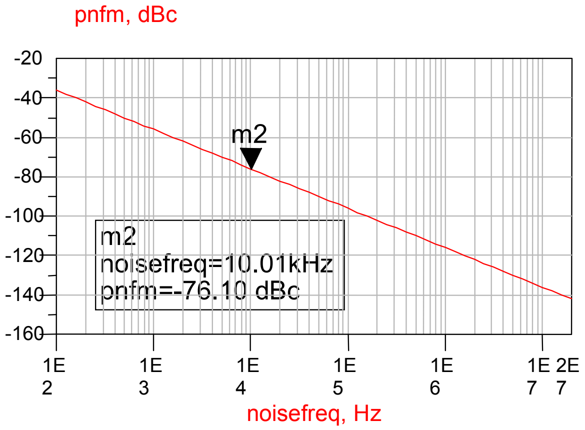


Figure 10 Phase noise prediction of the basic oscillator design.

The next stage of the design is to replace some of the circuit elements with micro-strip lines, specifically the tuning capacitor, shunt inductor and RF bias circuits (as shown by the DC_Feeds).

(1) Calculate 5nH inductor as an open-circuit stub.

$$\text{Reactance } X_L = 2\pi.f.L = 2\pi.2.0E^9.5.0E^{-9} = 62.8\Omega$$

$$X.L = Zo.\tan\theta + 90 \ \ \text{(OPEN Circuit STUB) \&}$$

$$X.L = Zo.\tan\theta \ \ \text{(SHORT Circuit Stub)}$$

$$Where\, \theta < 90^0 \left(90^0 \frac{\lambda_{op}}{4} \right)$$

Rearrange to get θ ie

$$\theta = \tan^{-1}\left(\frac{X_L}{Zo}\right) + 90 \ = \ 51.5 + 90 = 142 \text{ degrees}$$

(2) Calculate 0.55pF capacitor as an open-circuit stub.

$$\omega C = 2\pi.2E^9.0.55E^{-12} \ = 0.069$$

$$\omega C = \frac{\tan\theta}{Zo} \ \therefore \tan^{-1}(Xc.Zo)$$

$$\theta = 19 \text{ degrees}$$

$$\text{Where } \theta < 90^0 \left(90^0 \frac{\lambda_{op}}{4} \right)$$

(3) DC_Feed

Use 90 degree 90 ohm lines.
(4) DC_Capacitor shunts

Use 90 degree 20 ohm lines

The ADS model was updated to use ‘ideal’ micro-strip lines and open-circuit stubs as shown in **Figure 11**. These ideal components can be easily converted to real micro-strip lines by running ‘Linecalc’ in ADS to allow design of the oscillator layout.

The following section will now deal with a variable frequency oscillator.

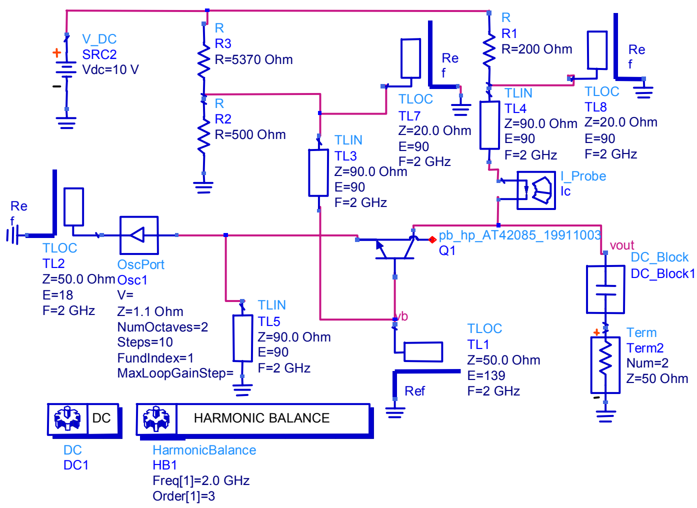


Figure 11 ADS schematic of the reflection oscillator with ‘ideal’ microstrip RF bias lines, capacitors and inductors fitted.

VARIABLE FREQUENCY NEGATIVE RESISTANCE OSCILLATOR [2,3,4]

In the following section the design of a varactor controlled variable frequency negative resistance oscillator is given. This design example will use the same device and RF circuitry used in the fixed frequency version.

In order to pick a suitable varactor for the design we need to decide on the tuning bandwidth. This particular requirement is for a tuning bandwidth of 20MHz/V over the tuning range of 1 to 10V. To give us a bit of margin to cover temperature effects etc we pick a bandwidth of ~ 25MHz/V. The varactor should have minimal case parasitics so we opt for a SMT package device.

The equivalent circuit of a typical varactor together with the package parasitics is shown in **Figure 12**.

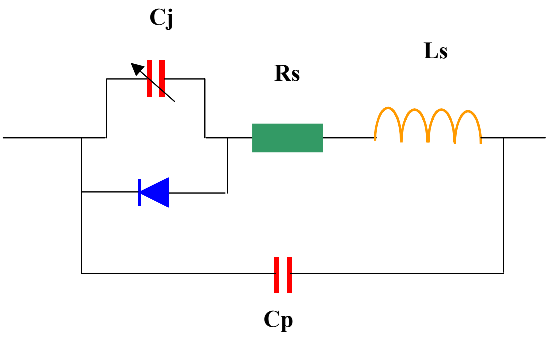


Figure 12: Typical equivalent circuit of a varactor diode. The series inductor Ls and parallel capacitor Cp are package parasitics, typical values are Cp ~ 0.1pF and Ls = 1.5nH.

The general equation for calculating the capacitance of the varactor is:

$$C_V = \frac{C_J}{\left(1+\frac{V_R}{V_J}\right)^{\gamma}} + C_P$$

where :

C_V = diode capacitance

V_R = applied voltage,

V_J = junction contact potential (~ 0.7V)

γ = Capacitance exponent

C_V = diode capacitance

For the Macom MA46506 varactor diode, is 1.0pF at 4V with a Gamma of $\gamma = 0.5$.

At 1V, CV = 1.66pF & at 10V, CV = 0.66pF

From the fixed frequency oscillator we can calculate the inductance of the input of the reflection oscillator as:

$$L = \frac{\left(\frac{1}{2.\pi.f}\right)^2}{C} = \frac{\left(\frac{1}{2.\pi.2E^9}\right)^2}{0.55E^{-12}} = 11.5nH$$

By rearranging the varactor equation we can find CJ by using the data CV=1.0pF at 4V.

This will give a value of CJ ~ 2.39pF

The easiest way to determine the value of the coupling capacitor is to generate a spreadsheet and enter values of Varactor coupling capacitor, Max & Min voltage, Gamma, CJ and resonator inductance as shown in **Table 4**.

From the entered value of Capactance at 4V and Gamma we can calculate CJ by rearranging the varactor formula using VR as 4V and VJ as 0.7V. Then by using CJ and entering the two control voltages, we can calculate the maximum and minimum varactor capacitances.
We then assume that the coupling capacitor will be in series with the varactor and therefore, with different coupling capacitors we can calculate the new max & min varactor capacitances. By entering the inductor we can calculate the max & min resonator frequencies and tuning bandwidth.

From **Table 4** we can see that if we pick a varactor coupling capacitor of 0.4pF then we should achieve our tuning bandwidth (with margin) of ~ 25MHz/V.
Note also that we need a total inductance of 30nH including the inductance of 11.5nH looking into the reflection amplifier. Therefore an inductance of 18.5nH needs to be initially added in series with the varactor.

Enter Cap at 4V	1	pF	2.591192776		
Gamma	0.5				
Enter Vlow	1	V			
Enter Vhigh	10	V			
Max Cap	1.66	pF	1.66274E-12		
Min Cap	0.66	pF	6.62761E-13		
Enter Tuning range	9	V			
Enter Total Resonator Ind	30	nH	0.00000003		
	Varactor Network				
Varactor coupling Cap (pF)	Max C	Min C	Min Freq (MHz)	Max Freq (MHz)	Tuning Range (MHz/V)
0.1	0.09432701	0.08689	2992	3117	13.93
0.2	0.17852625	0.153637	2175	2344	18.84
0.3	**0.25414573**	**0.206519**	**1823**	**2022**	**22.14**
0.4	0.32243326	0.249449	1618	1840	24.62
0.5	0.38440588	0.284994	1482	1721	26.58
1	0.62444697	0.398591	1163	1455	32.51
2	0.90792132	0.4978	964	1302	37.56
3	1.06980435	0.542837	888	1247	39.86
4	1.17451257	0.568557	848	1219	41.20
5	1.24778982	0.585192	823	1201	42.07
6	1.30194146	0.596834	805	1189	**42.68**

Table 4 Showing the resulting tuning ranges for the varactor and coupling capacitor combinations. To achieve ~ 25MHz/V we use a series capacitor of 0.4pF in series with the varactor. Note also that we need a total inductance of 30nH including the inductance of 11.5nH looking into the reflection amplifier.

These resonator components were added to the harmonic balance ADS schematic as shown in **Figure 13**. Note the inductor has been reduced to adjust the frequency to 2GHz at the mid control voltage (ie Cvaractor = 1pF).

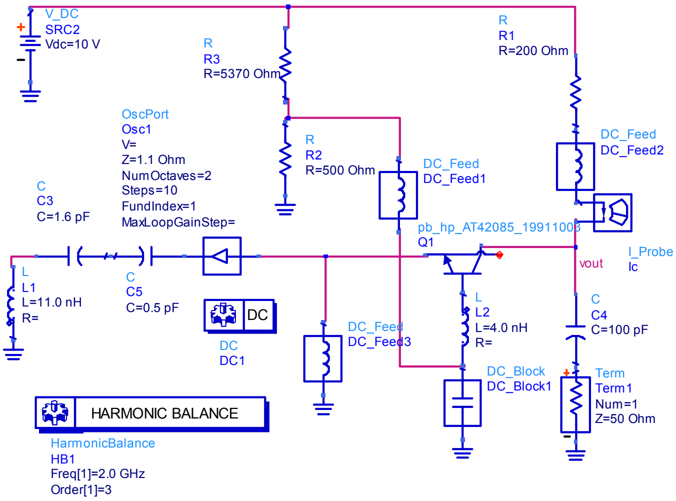
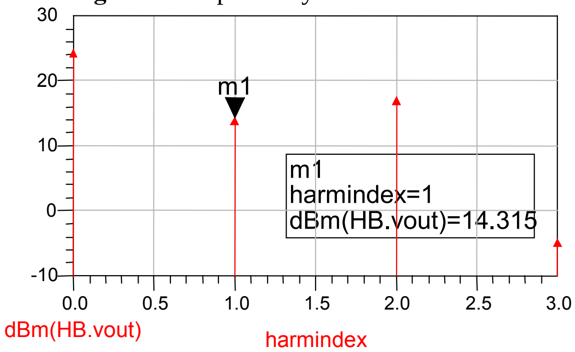


Figure 13 ADS schematic of the Voltage Controlled Oscillator (VCO). The varactor is represented by C3 which is set to 0.66pF (for 10V control voltage) and 1.66pF (for 10V control voltage).

The simulation was run with the min & max varactor capacitances resulting in the plots shown in **Figure 14** and **Figure 15** respectively.



harmindex	HB.freq
0	0.0000 Hz
1	1.897GHz
2	3.793GHz

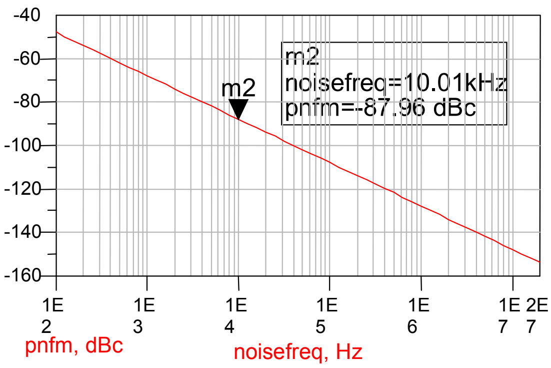
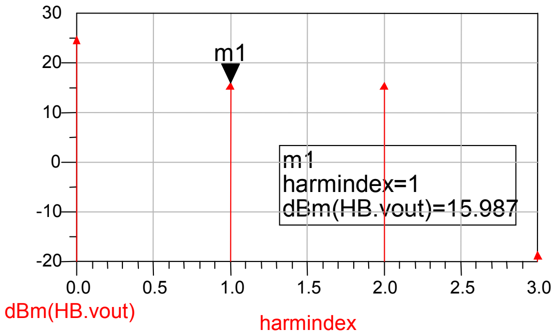


Figure 14 Simulation plots of the ADS schematic shown in Figure 13, showing output power, frequency & phase noise performance with the varactor set to 1V.



harmindex	HB.freq
0	0.0000 Hz
1	2.131GHz
2	4.262GHz

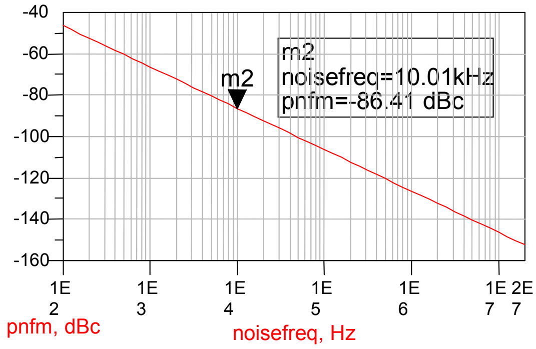


Figure 15 Simulation plots of the ADS schematic shown in Figure 13, showing output power, frequency & phase noise performance with the varactor set to 10V.

The simulations show a frequency range of 1897MHz to 2131MHz, resulting in an approximate (as the characteristic is not linear) tuning bandwidth of 26MHz/V. Note to bring the VCO on frequency the inductor in the resonator was decreased to 11nH.

Also all other RF tuning elements have been replaced with ‘ideal’ micro-strip lines & stubs **[1]**. It is best to replace one element at a time and re-check that the simulation still runs correctly. The final circuit is shown in **Figure 16**.

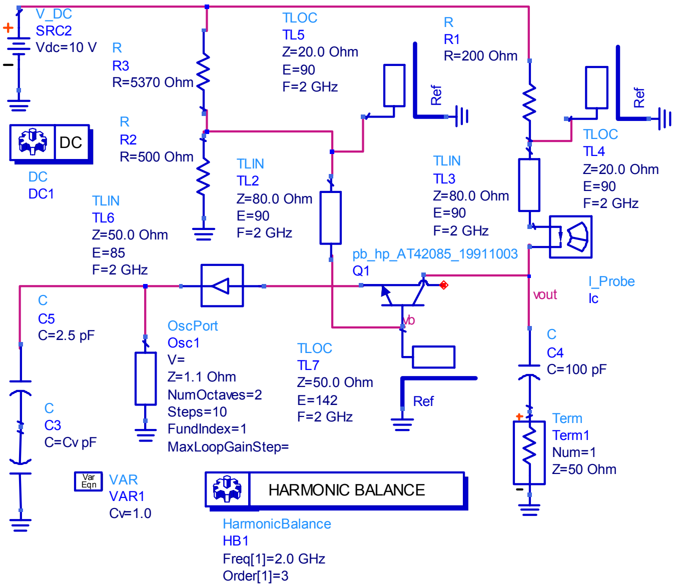


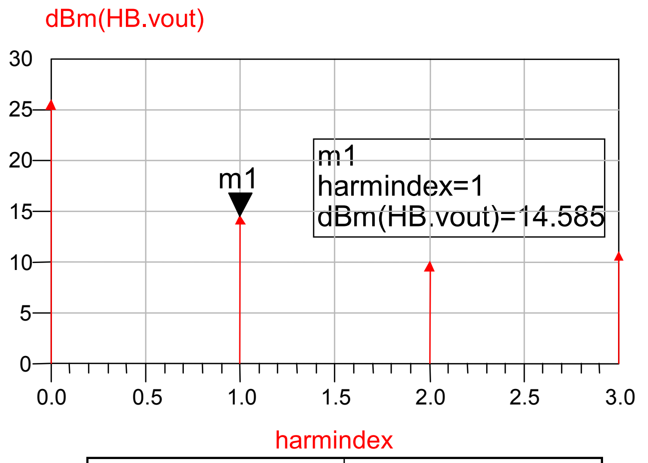
Figure 16 Final VCO ADS schematic. The lumped element inductors have been replaced with 'ideal' micro-strip lines. The varactor is represented by the 1pF capacitor (which equates to the mid-tuning voltage). In operation the open-circuit stub on the resonator would be trimmed/extended to center the oscillating frequency. The open-circuit stub on the bipolar base would be trimmed/extended to ensure reliable oscillation - especially over temperature.

The simulation was re-run first with a varactor capacitance of 0.3pF (equating to a control voltage of 1V) and then re-run a second time, with a capacitor value of 3.33pF (equating to a control voltage of 10). The results of frequency and output power are given in **Table 5**.

Note that micro-strip transmission lines are inherently narrow-band, therefore the addition of the 'ideal' transmission lines to the design in place of the lumped elements have considerably narrowed the tuning bandwidth of the oscillator. To widen the tuning bandwidth of the oscillator, the varactor coupling capacitor C5 was increased to 2.5pF.

Control voltage	Frequency (MHz)	Power(dBm
1V	2095	14.7
4V	1994	14.5
10V	1893	11.5

Table 5 Summary of frequency simulations with the varactor control set to max and min voltages (ie 10V & 1V). This results in a tuning bandwidth of ~ 22.2MHz/V. This bandwidth has been achieved by increasing the varactor coupling capacitor to 2.5pF.



harmindex	HB.freq
0	0.0000 Hz
1	1.994GHz
2	3.989GHz

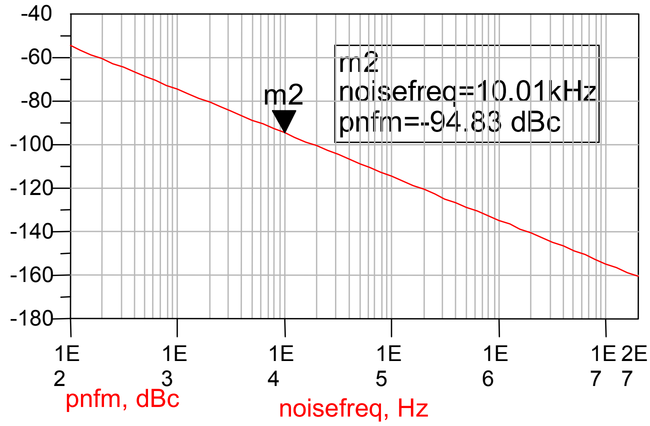


Figure 17 Phase noise, output power, frequency and Collector current results from the ADS schematic shown in Figure 16 – with the varactor set to mid-range ie 1pF.

The simulation showed that the low frequency power has dropped off by 3dB. The design as it stands does not have any tuning on the output that could be added to provide some power levelling by load mis-match. The length of stub TL1 can adjusted to ensure that oscillation occurs at about 2GHz, and a small amount of frequency adjustment can be mad by adding tuning to the shorted stub TL6. To ease alignment (but make the design slightly more complication) TL6 can be replaced with an open-circuit stub (with an additional length of 90 degrees), noting however that a DC return for the bipolar emitter is required, that can be provided by an inductive quarter-wave length of transmission line.

The basic oscillator can now be designed using ideal transmission lines replaced by 'real' micro-strip lines according to the substrate material dielectric constant (Er) and thickness (H) used **[1].** When generating a VCO layout, consideration must be given to the component solder pads on the electrical design. These pads are in themselves transmission lines and will add electrical length that must be taken into account.

CONCLUSION

This paper discussed the design of a basic fixed-frequency reflection oscillator and a variable frequency varactor controlled version, both using a distributed resonator and micro-strip circuit elements. Both oscillators were designed to operate at 2GHz with the variable frequency version designed to have a tuning bandwidth of ~ 20MHz/V. Throughout the design process staring with the basic oscillator (using ideal lumped elements) to the more complex final design (using 'ideal' micro-strip elements), Agilent ADS large and small simulations were given at each stage to verify circuit performance. It was noted that substituting 'ideal' micro-strip elements for the lumped RF bias and resonator circuits considerable reduced the tuning bandwidth of the oscillator. This was overcome in this design by increasing the varactor coupling capacitor. Another possibility would be to use a larger capacitance swing varactor ie one with a larger Gamma eg MaCom MA46H071 (Gamma 0.75) or even the MA46H201 (Gamma 1.25).

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