Lecture 17 - Microwave Mixers

Microwave Active Circuit Analysis and Design

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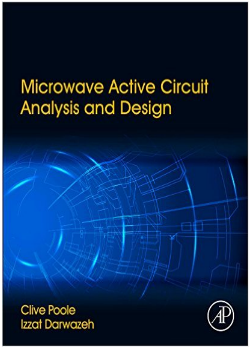

Intended Learning Outcomes

- Knowledge
  - Understand the role and function of mixers in RF systems.
  - Understand various figures of merit used to characterise mixers.
  - Understand the strengths and weaknesses of various mixer topologies.
- Skills
  - Be able to design a single balanced diode passive mixer.
  - Be able to design a double balanced diode passive mixer.
  - Be able to design a gilbert cell mixer, based on BJT or FETs.

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Mixer characterisation

- Conversion gain
- Isolation
- Dynamic range
- Third-order intercept point
- Noise Figure

Conversion gain and isolation

Conversion gain

- The conversion gain (or loss) of a mixer is defined as the ratio of the desired output signal power to the input signal power. In the case of a Downconversion mixer the desired output signal is the IF, and the input signal is the RF. In the case of an upconversion mixer the desired output signal is the RF and the input signal is the IF.
- LO power does not feature in conversion gain calculations.
- Active mixers provide a positive conversion gain, whereas passive mixers do not. Conversion gain is also dependent on impedance matching at the input and output ports (i.e. it depends on power transfer at these ports).
- Although IF signal power does not feature in the conversion gain calculation, the level of the LO will affect the conversion gain so this needs to be specified. Conversion gain of a typical active mixer is approximately +10dB, whilst the conversion loss of a typical passive (diode) mixer is approximately -6dB.

Isolation

- Isolation is a measure of the amount of "leakage" or "feed through" between the mixer ports, especially leakage of the local oscillator signal, as this tends to be the largest of the three.

Dynamic range

Dynamic Range (DR) is defined as the range of input power levels for which the output power is linearly proportional to the input power. The proportionality constant is the conversion loss or gain of the mixer.

- The lower bound of dynamic range is set by the noise floor which defines the *minimum detectable signal* (MDS).
- The 1 dB compression point is commonly used to define the upper bound of of the linear region.
- The system's dynamic range can be defined as the power difference from the MDS to the 1 dB compression point.

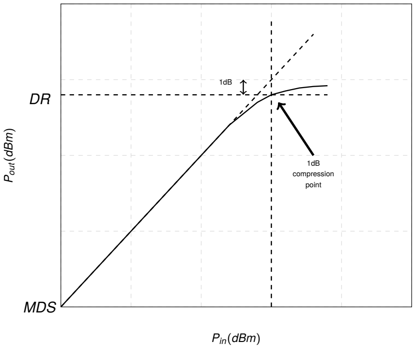

Figure 1 : 1 dB compression point

Third-order intercept point

The third order intercept point (or "IP3") is the input (or output) power level at which the nonlinear intermodulation products caused by the third-order non-linearities are equal to the desired signal.
Two different definitions for intercept points are in use:

1. Based on harmonics : The device is tested using a single input tone. The nonlinear products caused by n^{th} order nonlinearity appear at n times the frequency of the input tone.
2. Based on intermodulation products : The device is fed with two sine tones with a small frequency difference. The n^{th} order intermodulation products then appear at n times the frequency spacing of the input tones. This is the more commonly used approach.

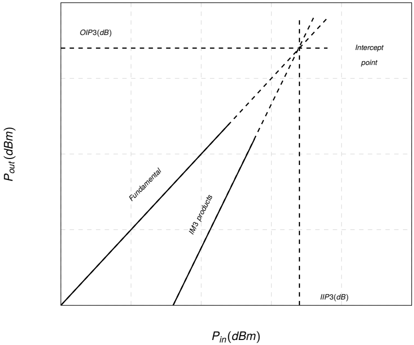

Figure 2 : 3rd order intercept point

Noise Figure

- The same definitions of noise figure that apply to amplifiers may be applied to mixers, provided we take account of the fact that the noise generated by the mixer will be upconverted or downconverted to a different frequency range
- For passive mixers, where there is no gain, only loss, the noise figure is approximately equal to the insertion loss
- Active mixers typically exhibit higher noise figure than passive mixers at comparable linearity, which is as one would expect from when comparing any active circuit with its passive equivalent[3]

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Basic mixer operation

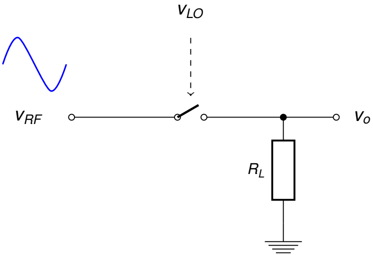

Figure 3 : Single balanced mixer equivalent circuit

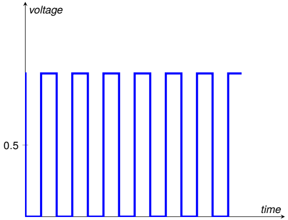

Figure 4 : Single balanced mixer LO switching waveform

$$S(v_{LO}) = \begin{cases} 1 & \text{if } v_{LO} \geq 0 \\ 0 & \text{if } v_{LO} < 0 \end{cases} \quad (1)$$

Single balanced mixer output waveform

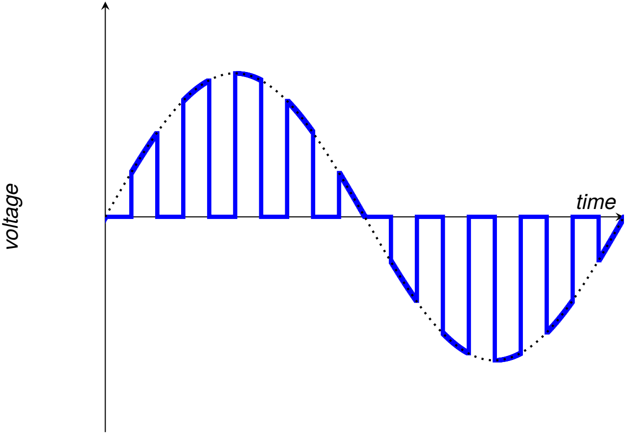

Figure 5 : Single balanced mixer output waveform

Double balanced mixer operation

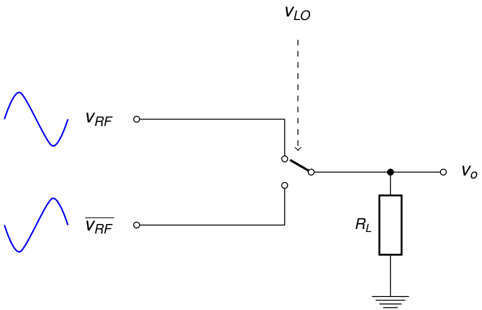

Figure 6 : Double balanced mixer equivalent circuit

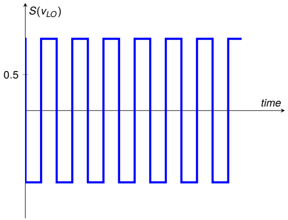

Figure 7 : Double balanced mixer LO switching waveform

$$S(v_{LO}) = \begin{cases} 1 & \text{if } v_{LO} \geq 0 \\ -1 & \text{if } v_{LO} < 0 \end{cases} \qquad (2)$$

Double balanced mixer output waveform

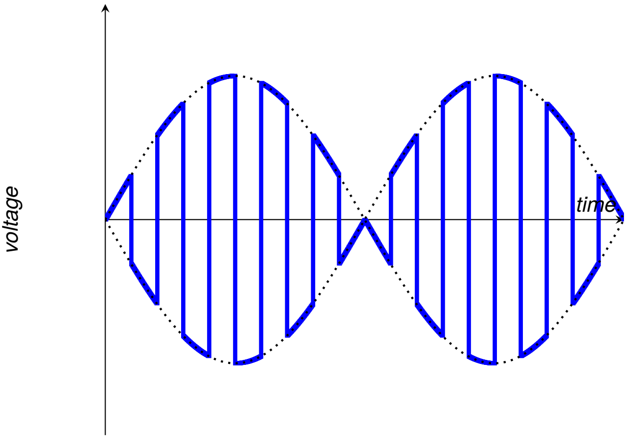

Figure 8 : Double balanced mixer output waveform

Double balanced mixer output waveform

After low pass filtering :

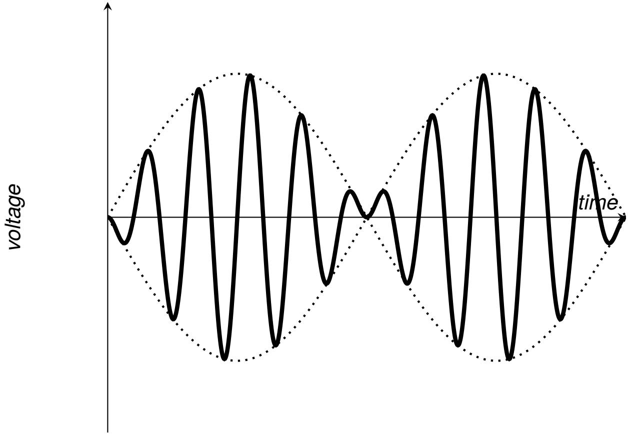

Figure 9 : Double balanced mixer output waveform after filtering

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Non-linear mixer

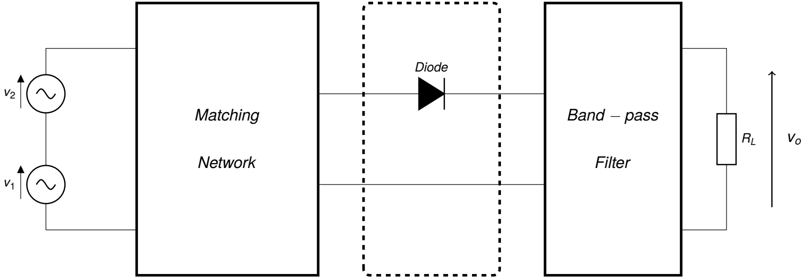

Figure 10 : Mixing using a non-linear device

Non-linear mixer operation

Given the current/voltage characteristic for a semiconductor diode :

$$I = I_S\left(e^{qV_D/k_BT} - 1\right) \quad (3)$$

where V_D is the voltage across the diode, I is the forward current and I_S is the reverse saturation current. We can replace the exponential function e^x by the following equivalent Taylor series[5] :

$$e^x = \sum_{n=0}^{\infty} \frac{x^n}{n!} \quad (4)$$

For small values of x, (4) can be approximated by :

$$e^x - 1 \approx x + \frac{x^2}{2} \quad (5)$$

The sum of the two input voltage signals $v_1 + v_2$ is applied to a diode, and the output voltage, v_o is proportional to the current through the diode. Ignoring DC terms, the output signal voltage will be of the form :

$$v_o = (v_1 + v_2) + \frac{1}{2}(v_1 + v_2)^2 + \ldots \quad (6)$$

which can be rewritten as :

$$v_o = (v_1 + v_2) + \frac{1}{2}v_1^2 + v_1 v_2 + \frac{1}{2}v_2^2 \quad (7)$$

Double-balanced diode mixer

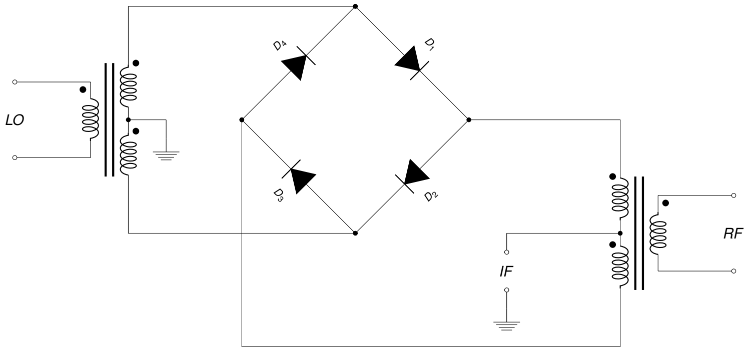

Figure 11 : Double balanced diode mixer

Double-balanced diode mixer : positive L.O. cycle

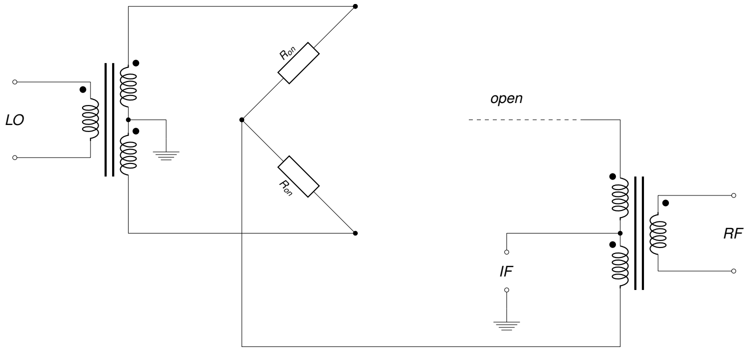

Figure 12 : Double balanced diode mixer model : positive LO cycle

Double-balanced diode mixer : negative L.O. cycle

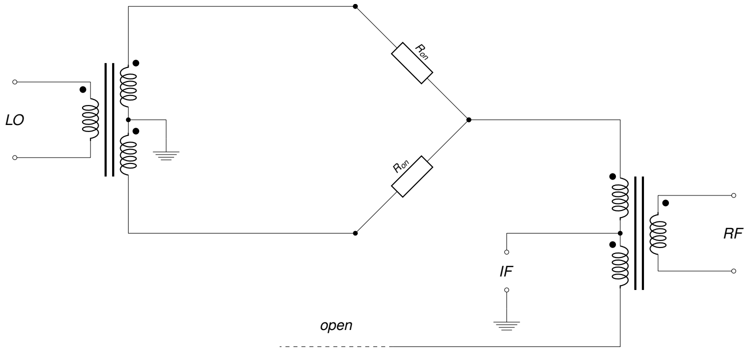

Figure 13 : Double balanced diode mixer model : negative LO cycle

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Active mixer building blocks

- The V/I conversion function is usually carried out by a voltage controlled current source.
- The mixing function consists of multiplying one signal by another in the time domain using a current multiplier.
- The I/V conversion function can be carried out simply by a means of a load resistor.

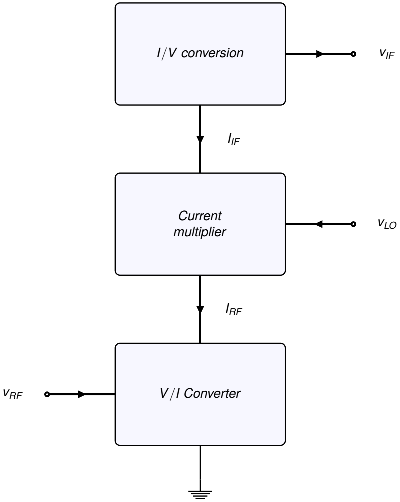

Figure 14 : Active mixer building blocks

Single-balanced active mixer using BJT

- The current source transistor, Q_3, sets the total current in the upper two transistors, which then compete for a fraction of this current.
- The load resistors, R_L, convert the collector currents into voltages. The gain of a differential pair depends on the transconductance, g_m, of the two transistors which, in turn, depends on their collector current (according to $g_m = I_C/V_T$).
- The circuit of figure 15 can therefore be used as a mixer when the current in Q_3, is modulated by the RF signal voltage.
- We apply the local oscillator signal as a differential voltage to the bases of Q_1 and Q_2, and take the IF signal as a differential voltage across the collectors of Q_1 and Q_2.
- The degenerative feedback resistance, R_e, is added to the current source to improve its linearity.

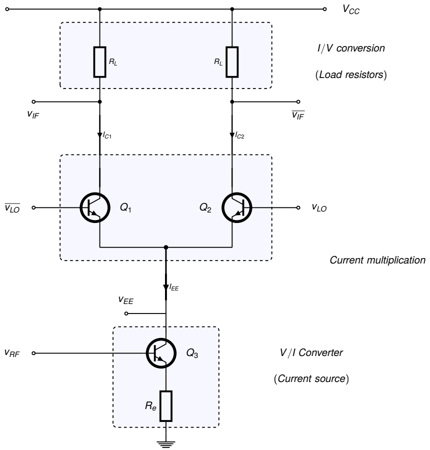

Figure 15 : Single-balanced active mixer using BJTs

BJT Single balanced mixer analysis

Assuming the two transistors Q_1 and Q_2 in figure 15 have identical voltage-current characteristics, the relationships between the small signal collector currents and base-emitter voltages of Q_1 and Q_2 are as follows :

$$i_{C1} = I_S \, e^{(v_{BE1}/V_T)} \tag{8}$$

$$i_{C2} = I_S \, e^{(v_{BE2}/V_T)} \tag{9}$$

Where I_S is the saturation current and V_T is the thermal voltage defined by $V_T = kT/q \approx 26mV$ at room temperature (around T=290 K).
Alternatively we can write :

$$v_{BE1} = V_T \, \ln\left(\frac{i_{C1}}{I_S}\right) \tag{10}$$

$$v_{BE2} = V_T \, \ln\left(\frac{i_{C2}}{I_S}\right) \tag{11}$$

The voltage at the common emitter point, v_{EE}, in figure 15 is given by :

$$v_{EE} = \overline{v_{LO}} - v_{BE1} = v_{LO} - v_{BE2} \tag{12}$$

From (12) we can write :

$$v_{LO} - \overline{v_{LO}} = v_{BE2} - v_{BE1} \tag{13}$$

BJT Single balanced mixer analysis

$v_{LO} - \overline{v_{LO}}$ is the differential local oscillator input signal voltage, Δv_{LO}, so applying (10) we can write :

$$\Delta v_{LO} = v_{LO} - \overline{v_{LO}} = V_T \ \ln\left(\frac{i_{C2}}{I_S}\right) - V_T \ \ln\left(\frac{i_{C1}}{i_S}\right) \tag{14}$$

$$\Delta v_{LO} = V_T \ \ln\left(\frac{i_{C2}}{i_{C1}}\right) \tag{15}$$

Therefore :

$$\left(\frac{i_{C1}}{i_{C2}}\right) = e^{(-\Delta v_{LO}/V_T)} \tag{16}$$

For most microwave transistors the current gain is high, therefore we can usually ignore the base current, i.e. $i_E \approx i_C$. Therefore from figure 15 we can write :

$$i_{EE} = i_{C1} + i_{C2} \tag{17}$$

Combining (16) and (17) we can now write the collector signal currents of Q_1 and Q_2 in figure 15 in terms of Δv_{LO} and I_{EE} as :

$$i_{C1} = \frac{i_{EE}}{1 + e^{(\Delta v_{LO}/V_T)}} \tag{18}$$

$$i_{C2} = \frac{i_{EE}}{1 + e^{(-\Delta v_{LO}/V_T)}} \tag{19}$$

BJT Single balanced mixer analysis

The difference between the two collector currents can now be written as follows :

$$\Delta i_{IF_{12}} = i_{C1} - i_{C2} \tag{20}$$

$$= I_{EE}\left(\frac{1}{1+e^{(\Delta v_{LO}/V_T)}} - \frac{1}{1+e^{(-\Delta v_{LO}/V_T)}}\right) \tag{21}$$

Which can be rewritten as :

$$\Delta i_{IF_{12}} = i_{EE}\left(\frac{e^{(-\Delta v_{LO}/2V_T)} - e^{(\Delta v_{LO}/2V_T)}}{e^{(-\Delta v_{LO}/2V_T)} + e^{(\Delta v_{LO}/2V_T)}}\right) \tag{22}$$

Equation (22) can be more neatly expressed by using the definition of the hyperbolic tangent (tanh) function, which is defined as :

$$\tanh(x) = \frac{e^{x} - e^{-x}}{e^{x} + e^{-x}} \tag{23}$$

We can now rewrite (20) as :

$$\Delta i_{IF_{12}} = i_{EE}\tanh\left(\frac{\Delta v_{LO}}{2V_T}\right) \tag{24}$$

BJT Single balanced mixer analysis

Now, if we assume that the bias current I_{EE} is modulated by v_{RF}, as illustrated in figure 15, we can replace the simple term I_{EE} in (24) by $(I_{EE_o} + g_{m_3} v_{RF})$ where I_{EE_o} is the quiescent DC bias current and g_{m_3} is the transconductance of the current source transistor, Q_3. We can now write :

$$\Delta i_{IF_{12}} = (I_{EE_o} + g_{m_3} v_{RF}) \tanh\left(\frac{\Delta v_{LO}}{2V_T}\right) \tag{25}$$

Which can be expanded to :

$$\Delta i_{IF_{12}} = I_{EE_o} \tanh\left(\frac{\Delta v_{LO}}{2V_T}\right) + g_{m_3} v_{RF} \tanh\left(\frac{\Delta v_{LO}}{2V_T}\right) \tag{26}$$

Consider the Maclaurin series expansion of $\tanh(x)$, as follows[5] :

$$\tanh(x) = x - \frac{1}{3}x^3 + \frac{2}{15}x^5 - \frac{17}{315}x^7 + \frac{62}{2835}x^9 \cdots \tag{27}$$

BJT Single balanced mixer analysis

From (27) we can show that $\tanh(x) \approx x$ for small values of x (i.e. for values of x up to $x = 0.5$ we have $\tanh(x)/x > 0.92$). So we can approximate the tanh in (26) on the assumption that v_{RF} and v_{LO} are small (i.e. less than $V_T = 26\ mV$ at room temperature). We therefore have :

$$\boxed{\Delta i_{IF_{12}} \approx I_{EE_o}\left(\frac{\Delta v_{LO}}{2V_T}\right) + g_{m_3}\left(\frac{\Delta v_{LO}\ v_{RF}}{2V_T}\right)} \quad (28)$$

The first term in equation (28) represents the Local Oscillator leakage component, which is proportional to the DC bias current, i_{EE_o}. The second term contains the product term we are interested in, namely ($v_{RF}\ \Delta v_{LO}$).

Active double balanced mixer - the Gilbert Cell

- The deficiencies of the single-balanced mixer can be overcome by adopting a double balanced design, and using a balanced RF signal feed.
- This can be achieved by combining two single balanced circuits, one being driven by v_{RF} and the other driven by its inverse, $\overline{v_{RF}}$. The balanced LO inputs and IF outputs are combined by connecting the respective nodes together.
- The resultant circuit is known as a *Gilbert cell*[1] the basic topology of which is shown in figure 16.

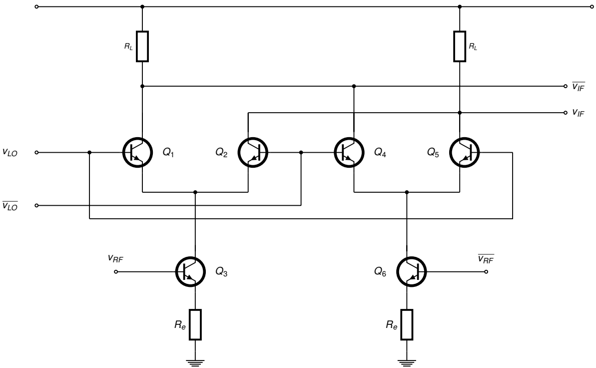

Figure 16 : Gilbert cell mixer topology

Active double balanced mixer - the Gilbert Cell

The second transistor pair in figure 16, Q_4, Q_5, has a signal response similar to the original pair, Q_1, Q_2, but 180^o out of phase, as this pair is being driven by an inverted version of the LO signal (relative to that driving Q_1 and Q_2). By analogy with equation (28) we can therefore write:

$$\overline{\Delta i_{IF_{45}}} \approx I_{EE}\left(\frac{\Delta v_{LO}}{2V_T}\right) - g_{m_6} v_{RF}\left(\frac{\Delta v_{LO}}{2V_T}\right) \tag{29}$$

Where, in this case, $\Delta i_{IF_{45}} = i_{C4} - i_{C5}$
If we ensure that all the transistors being used have identical characteristics, so the g_m values are the same in all cases and specifically $g_{m_3} = g_{m_6}$, the differential IF output current of the circuit of figure 16 is given by:

$$\Delta i_{IF} = \Delta i_{IF_{12}} - \overline{\Delta i_{IF_{45}}} \tag{30}$$

$$\approx 2 g_m v_{RF}\left(\frac{\Delta v_{LO}}{2V_T}\right) = \left(\frac{g_m}{V_T}\right) v_{LO} \cdot v_{RF} \tag{31}$$

The above subtraction of $\overline{i_{IF_{45}}}$ from $i_{IF_{12}}$ has the effect, to a first order approximation, of cancelling the common term containing the unmodulated v_{LO} signal, leaving only the desired product term $v_{LO} \cdot v_{RF}$. Thus, the IF feed-through component that was present at the output of the single balanced mixer described by (26) has been removed. This is the primary benefit of using the double balanced topology in figure 16.

Active double balanced mixer - the Gilbert Cell

- Figure 17 shows a practical Gilbert Cell implementation such as would be implemented in MMIC form. The common bias current, I_{EE}, is set by the fixed current source Q_7.
- The gain of the two differential amplifiers, formed of Q_1 & Q_2 and Q_4 & Q_5, is controlled by modulating the emitter bias current via the transistors Q_3 and Q_6.
- In normal operation the Local Oscillator signal is applied differentially to the bases of Q_1 and Q_5 (positive phase) and Q_2 and Q_4 (negative phase), whilst the RF signal is applied differentially to the bases of Q_3 (positive phase) and Q_6 (negative phase).
- The IF signal is taken differentially from the collectors of the two upper transistor pairs.
- Another enhancement applied to the Gilbert Cell in figure 17 is the addition of the emitter degeneration resistors, R_E. These are used in practical circuits to improve linearity at the expense of some conversion gain[2].
- The effect of R_E is to reduce the transconductance of the lower transistors, Q_3 and Q_6 by a factor of $1/(1 + g_m R_E)$, due to the action of local feedback.

Active double balanced mixer - the Gilbert Cell

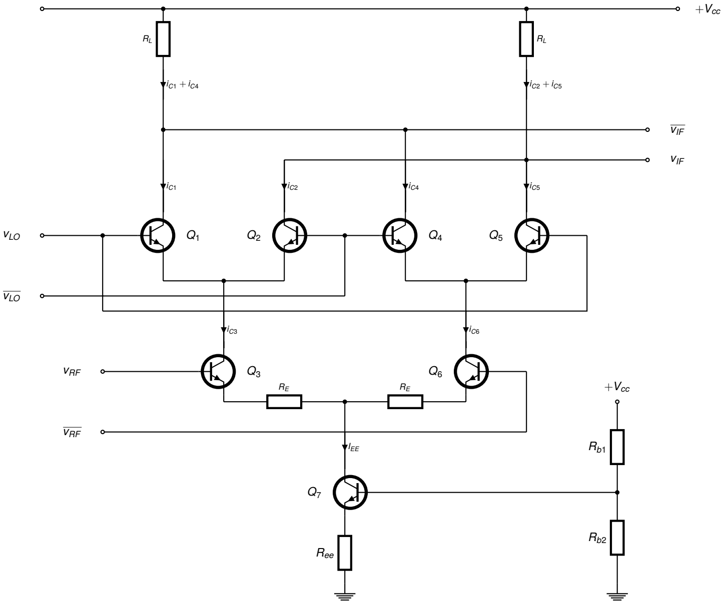

Figure 17 : Gilbert cell mixer implementation using BJT

Active double balanced mixer - the Gilbert Cell

The collector signal currents of the upper pair transistors illustrated in figure 17, once again ignoring DC bias and base currents, are as follows :

$$i_{C1} = \frac{i_{C3}}{1 + e^{-v_{LO}/V_T}} \quad (32)$$

$$i_{C2} = \frac{i_{C3}}{1 + e^{v_{LO}/V_T}} \quad (33)$$

$$i_{C4} = \frac{i_{C6}}{1 + e^{v_{LO}/V_T}} \quad (34)$$

$$i_{C5} = \frac{i_{C6}}{1 + e^{-v_{LO}/V_T}} \quad (35)$$

For the lower pair transistors, Q_3 and Q_6, we can write :

$$i_{C3} = \frac{I_{EE}}{1 + e^{-v_{RF}/V_T}} \quad (36)$$

$$i_{C6} = \frac{I_{EE}}{1 + e^{v_{RF}/V_T}} \quad (37)$$

Active double balanced mixer - the Gilbert Cell

Noting that, in the case of the lower transistors, I_{EE} is the fixed bias current provided by Q_7. Combining (32) through (37), we obtain expressions for the collector currents I_{C1}, I_{C2}, I_{C4}, and I_{C5} in terms of input signal voltages v_{RF} and v_{LO}.

$$i_{C1} = \frac{I_{EE}}{\left[1 + e^{-v_{LO}/V_T}\right]\left[1 + e^{-v_{RF}/V_T}\right]} \tag{38}$$

$$i_{C2} = \frac{I_{EE}}{\left[1 + e^{-v_{RF}/V_T}\right]\left[1 + e^{v_{LO}/V_T}\right]} \tag{39}$$

$$i_{C4} = \frac{I_{EE}}{\left[1 + e^{v_{LO}/V_T}\right]\left[1 + e^{v_{RF}/V_T}\right]} \tag{40}$$

$$i_{C5} = \frac{I_{EE}}{\left[1 + e^{v_{RF}/V_T}\right]\left[1 + e^{-v_{LO}/V_T}\right]} \tag{41}$$

Active double balanced mixer - the Gilbert Cell

With reference to figure 17 the differential output current is given by :

$$\Delta i_{IF} = (i_{C1} + i_{C4}) - (i_{C2} + i_{C5}) \tag{42}$$

Which can be rearranged as :

$$\Delta i_{IF} = (i_{C1} - i_{C5}) - (i_{C2} - i_{C4}) \tag{43}$$

Applying equations (38) to (43) and using the definition of tanh, we can rewrite (43) as :

$$\Delta i_{IF} = I_{EE} \tanh\left(\frac{v_{LO}}{2V_T}\right) \cdot \tanh\left(\frac{v_{RF}}{2V_T}\right) \tag{44}$$

The differential output current is converted into a differential voltage by the load resistors, R_L, so we can write :

$$v_{IF} = I_{EE}R_L \tanh\left(\frac{v_{LO}}{2V_T}\right) \cdot \tanh\left(\frac{v_{RF}}{2V_T}\right) \tag{45}$$

Active double balanced mixer - the Gilbert Cell

If we again assume that both v_{RF} and v_{LO} are small (i.e. less than V_T), we can apply the approximation $\tanh(x) \approx x$ to (45) and thereby obtain:

$$v_{IF} \approx \left(\frac{I_{EE}R_L}{4V_T^2}\right) v_{LO}\, v_{RF} \tag{46}$$

Equation (46) means that, for small enough signals, the differential output IF voltage is directly proportional to the product of the RF and LO input signal voltages. Equation (46) is true irrespective of the polarities of v_{RF} and v_{LO}.

The output of a Gilbert cell is therefore a true four quadrant multiplication of the differential base voltages of the LO and RF inputs. For this reason the circuit, although applied as a mixer in this context, is often simply referred to as a *Gilbert cell multiplier*.

Gilbert Cell Operating Modes

In practice there are three distinct operating modes for the Gilbert cell, according to the magnitudes of v_{RF} and v_{LO} relative to V_T, as follows :

1. If both v_{RF} and v_{LO} are much less than V_T, then the hyperbolic tangent function is approximately linear and the circuit behaves as a true four quadrant analogue voltage multiplier, as per (46). The input voltage range can be extended by adding 'predistortion' circuits at the inputs, which have an approximately $tanh^{-1}$ characteristic. This technique is sometimes used at lower frequencies but is not common at microwave frequencies[1].
2. If one of the input voltages significantly exceeds V_T then one of the transistor pairs will be driven into saturation and will behave like on/off switches. This is effectively equivalent to multiplying the other input signal by a square wave. This mode of operation is quite common in downconversion mixers where the IF signal is a lot larger than the RF signal. The 'squaring' of the LO signal in this mode is not a problem as the information content of the RF signal is preserved.
3. If both of the input voltages significantly exceed V_T then all of the transistors are operating as switches. This mode is sometimes employed when the mixer is being used as a phase detector, as the phase relationship between the two input signals is preserved, even though any information contained in the signal amplitudes will have been lost.

Conversion gain of the Gilbert Cell

Consider (45) for the case of a downconversion mixer, where the RF signal is of a small amplitude. Using the approximation $\tanh(x) \approx x$ for $x << 1$, we can write :

$$v_{IF} = \frac{I_{EE} R_L v_{RF}}{2V_T} \tanh\left(\frac{v_{LO}}{2V_T}\right) \tag{47}$$

Where I_{EE} is the DC bias current through Q_7. We will simplify the analysis by firstly considering the absence of emitter degeneration resistors, i.e. we set $R_E = 0$. We can then define the transconductance of the bias transistor, Q_7, in figure 17 as :

$$g_{m_7} = \frac{I_{EE}}{V_T} \tag{48}$$

We now substitute (48) into (47) to give :

$$v_{IF} = \tanh\left(\frac{v_{LO}}{2V_T}\right) \frac{g_{m_7} R_L v_{RF}}{2} \tag{49}$$

which gives the conversion gain of the circuit of figure 17 as :

$$\frac{v_{IF}}{v_{RF}} = \tanh\left(\frac{v_{LO}}{2V_T}\right) \frac{g_{m_7} R_L}{2} \tag{50}$$

Conversion gain of the Gilbert Cell

For a downconversion mixer we expect the RF input to be of small amplitude and we need its treatment to be as linear as possible, so as to preserve the information content[4]. The handling of the LO input, on the other hand, need not be linear, since the LO is of known amplitude and frequency and therefore has no information content. Distortion of the LO signal is of no consequence, and so we may choose the LO signal amplitude so as to maximise conversion efficiency. In fact, the LO input is usually designed to switch the upper transistor quad so that for half the cycle Q_1 and Q_5 are 'on' and taking all of the current i_{C3} and i_{C6}. For the other half of the LO cycle, Q_1 and Q_5 are 'off' and Q_2 and Q_4 are on, so all of i_{C3} and i_{C6} flows through these respective transistors. In other words, for a switching Gilbert Cell where $v_{LO} >> 2V_T$, then (50) can be approximated as :

$$\frac{v_{IF}}{v_{RF}} = u(v_{LO}) \left(\frac{g_m R_L}{2} \right) \tag{51}$$

where

$$u(v_{LO}) = \begin{cases} 1 & \text{if } v_{LO} \geq 0 \\ -1 & \text{if } v_{LO} < 0 \end{cases} \tag{52}$$

We can represent the square wave function $u(v_{LO})$ by its Fourier series expansion:

$$u(v_{LO}) = \frac{4}{\pi} \left(\sin(\omega_{LO} t) + \frac{1}{3}\sin(3\omega_{LO} t) + \frac{1}{5}\sin(5\omega_{LO} t) + \cdots \right) \tag{53}$$

Conversion gain of the Gilbert Cell

The conversion gain of a downconversion mixer is defined as the ratio of IF signal amplitude to RF signal amplitude. We are therefore only interested in the first term in (53), which has an amplitude of $\pi/4$. The conversion gain given by (51) now becomes :

$$\frac{v_{RF}}{v_{IF}} \approx \frac{2}{\pi} g_m R_L \tag{54}$$

We now consider the effect of finite values of R_E, which has the effect of reducing the transconductance of the lower transistors, Q_3 and Q_6 by a factor of $(1 + g_m R_E)$. With finite R_E the conversion gain given by (50) now becomes :

$$\frac{v_{IF}}{v_{RF}} = -\tanh\left(\frac{v_{LO}}{2V_T}\right)\left(\frac{g_m R_L}{1 + g_m R_E}\right) \tag{55}$$

Employing the function $u(v_{LO})$ defined above yields:

$$\frac{v_{IF}}{v_{RF}} = u(v_{LO})\left(\frac{g_m R_L}{1 + g_m R_E}\right) \tag{56}$$

The voltage gain of the downconversion mixer with the emitter degeneration resistors added is given by :

$$\frac{V_{RF}}{V_{IF}} \approx \frac{2}{\pi}\left(\frac{g_m R_L}{1 + g_m R_E}\right) \tag{57}$$

The FET Gilbert Cell

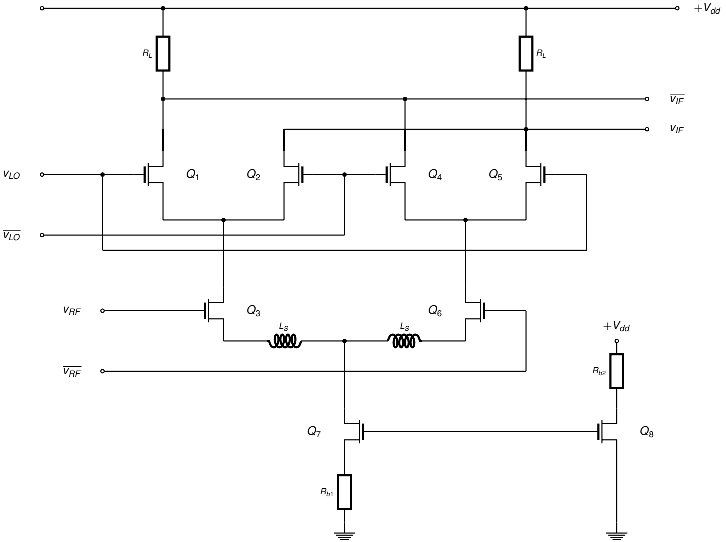

Figure 18 : Integrated FET Gilbert cell mixer

References

B. Gilbert.

The multi-tanh principle: a tutorial overview.

Solid-State Circuits, IEEE Journal of, 33(1):2–17, January 1998.

Rok-Hee Lee, Jae-Young Lee, Sang-Hun Lee, B. Shrestha, Sun-Jin Kim, G.P. Kennedy, Chan Hyeong Park, Nam-Young Kim, and Sang-Hoon Cheon.

Circuit techniques to improve the linearity of an up-conversion double balanced mixer with an active balun using InGaP/GaAs HBT technology.

In *Microwave Conference Proceedings, 2005. APMC 2005. Asia-Pacific Conference Proceedings*, volume 2, pages 4 pp.–, December 2005.

S.A. Maas.

Microwave mixers.

The Artech House Microwave Library. Artech House, Incorporated, 1993.

J.W.M. Rogers and C. Plett.

Radio Frequency Integrated Circuit Design.

Artech House Microwave Library. Artech House, Incorporated, 2014.

K.A. Stroud and D.J. Booth.

Advanced Engineering Mathematics.

Palgrave Macmillan Limited, 2011.