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Intermodulation and mixing

Let $\quad i = c_0 + c_1 v + c_2 v^2 + c_3 v^3 + \cdots$

neglegt the higher order terms

In the case of ~~of~~ having two equal amplitude sinusoidal signals;

$$v = A(\cos w_1 t + \cos w_2 t)$$

i becomes

$$i = c_0 + c_1 A(\cos w_1 t + \cos w_2 t) + c_2 A^2(\cos w_1 t + \cos w_2 t)^2 + c_3 A^3(\cos w_1 t + \cos w_2 t)^3$$

$$= c_0 + c_1 A \cos w_1 t + c_1 A \cos w_2 t + c_2 A^2 [\cos^2 w_1 t + \cos^2 w_2 t + 2\cos w_1 \cos w_2 t] + c_3 A^3 [\cos^3 w_1 t + 3\cos w_1 \cos^2 w_2 t + 3\cos^2 w_1 t \cos w_2 t + \cos^3 w_2 t]$$

$$= c_0 + A c_1 \cos w_1 t + A c_1 \cos w_2 t + c_2 A^2 \left[\frac{1+\cos 2w_1 t}{2} + \frac{1+\cos 2w_2 t}{2} + \frac{2[\cos(w_1-w_2)t + \cos(w_1+w_2)t]}{2} \right] + c_3 A^3 \left[\cos w_1 t \frac{[1+\cos 2w_1 t]}{2} + 3\cos w_1 t \frac{1+\cos 2w_2 t}{2} + 3\frac{1+\cos 2w_1 t}{2}\cos w_2 t + \cos w_2 t \frac{1+\cos 2w_2 t}{2} \right]$$

$$i = C_0 + C_1 A \cos w_1 t + c_1 A \cos w_2 t$$

$$+ \frac{C_2 A^2}{2} + \frac{C_2 A^2}{2} \cos 2w_1 t + \frac{C_2 A^2}{2} + \frac{C_2 A^2}{2} \cos 2w_2 t$$

$$+ C_2 A^2 \cos (w_1 - w_2) t + C_2 A^2 \cos (w_1 + w_2) t$$

$$+ C_3 A^3 \Big[\frac{1}{2} \cos w_1 t + \frac{1}{4} \left[\cos (2w_1 - w_1) t + \cos 3w_1 t \right] + \frac{3}{2} \cos w_1 t$$

$$+ \frac{3}{4} \left[\cos (w_1 - 2w_2) t + \cos (w_1 + 2w_2 t) \right] + \frac{3}{2} \cos w_2 t$$

$$+ \frac{3}{4} \left[\cos (2w_1 - w_2) t + \cos (2w_1 + w_2) t \right] + \frac{1}{2} \cos w_2 t$$

$$+ \frac{1}{4} \left[\cos (2w_2 - w_2) t + \cos 3w_2 t \right] \Big]$$

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rearranging the terms

$$i = c_0 + \frac{C_2A^2}{2} + \frac{C_2A^2}{2}$$

$$+\left(C_1A + \frac{C_3A^3}{2} + \frac{C_3A^3 3}{2} + \frac{C_3A^3}{4}\right)\cos w_1 t$$

$$+\left(C_1A + \frac{3}{2}C_3A^3 + \frac{C_3A^3}{2} + \frac{C_3A^3}{4}\right)\cos w_2 t$$

$$+\frac{C_2A^2}{2}\cos 2w_1 t$$

$$+\frac{C_2A^2}{2}\cos 2w_2 t$$

$$+\frac{C_3A^3}{4}\cos 3w_1 t$$

$$+\frac{C_3A^3}{4}\cos 3w_2 t$$

$$+C_2A^2\cos(w_1-w_2)t + C_2A^2\cos(w_1+w_2)t$$

$$+\frac{3C_3A^3}{4}\cos(2w_2-w_1)t + \frac{3C_3A^3}{4}\cos(2w_2+w_1)t$$

$$+\frac{3C_3A^3}{4}\cos(2w_1-w_2)t + \frac{3C_3A^3}{4}\cos(2w_1+w_2)t$$

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④

Intermodulation in amplifiers

Wideband amplifiers must be extremely linear because $(w_1 - w_2)$, $(2w_2 - w_1)$, $(2w_1 - w_2)$ products all stay in the BW and even the DC term changes by $2\frac{C_2 A^2}{2} = C_2 A^2$ for frequencies nearing the BW.

For lower frequencies all of the terms might fall into the BW creating undesirable effects.

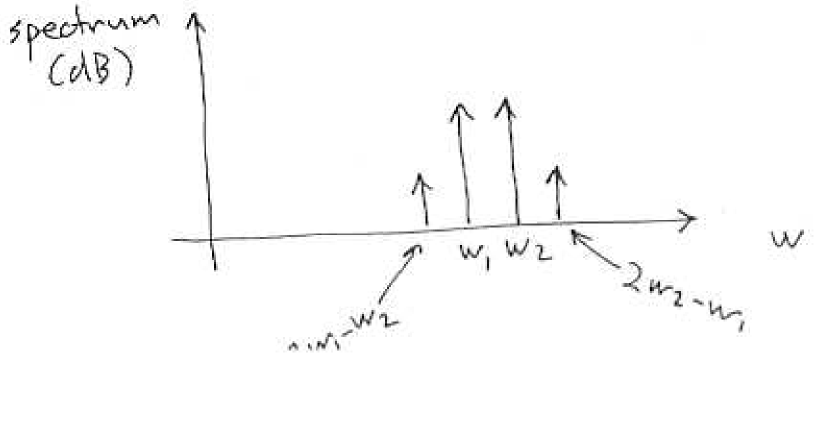
In narrow-band amplifiers the $(w_1 - w_2)$ and $(w_1 + w_2)$ and other terms fall out of the bandwidth except $(2w_2 - w_1)$ and $(2w_1 - w_2)$ terms.

Therefore the output of a nonlinear narrow-band amplifier can be written as:

$$\left[c_1 A + \frac{9}{4} \times \frac{C_3 A^3}{1}\right](\cos w_1 t + \cos w_2 t)$$

$$+ \frac{3}{4} C_3 A^3 \left[\cos(2w_1 - w_2)t + \cos(2w_2 - w_1)t\right]$$

The spectrum of the output of the amplifier is given below



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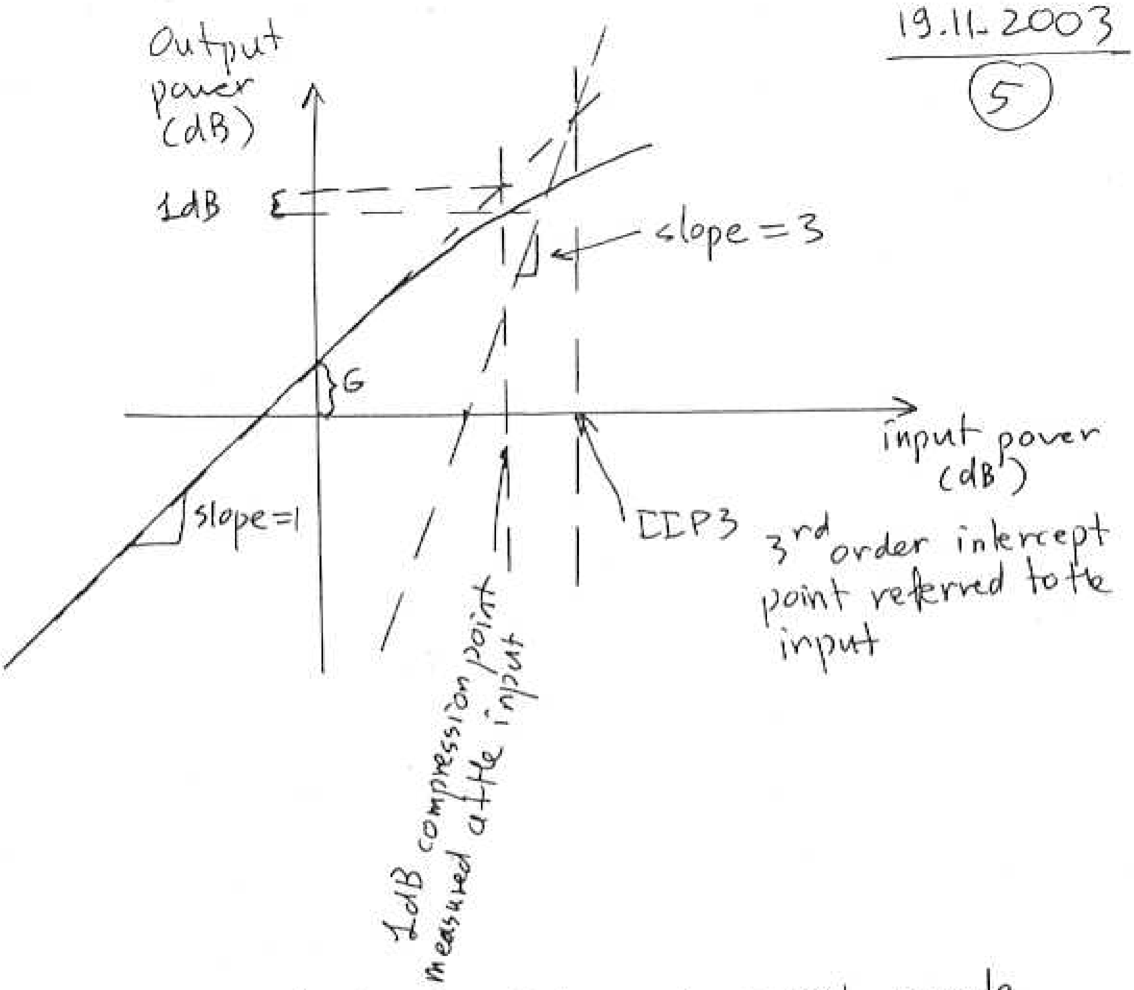

In order to find the intercept point equate the equations

$$\left| c_1 A + \cancel{\frac{9}{4} c_3 A^3} \right| = \left| \frac{3}{4} c_3 A^3 \right|$$

slope of the gain line is represented by this term, therefore delete the other term

$$|c_1| \cancel{A} = \frac{3}{4} |c_3| A^{\cancel{3}\,2}$$

$$A^2 = \left| \frac{c_1}{c_3} \right| \frac{4}{3} \Rightarrow A = \left(\left| \frac{c_1}{c_3} \right| \frac{4}{3} \right)^{1/2}$$

is the intercept point expressed in the input voltage amplitude. The input power created by A is

$$\frac{A_{IIP3}^2}{2R_s} = \frac{A^2}{2R_s} = \frac{2}{3R_s} \left| \frac{c_1}{c_3} \right|$$ (the 3rd order intercept point referred to the input)

3-point method for estimating IIP3

You can use SPICE in the transient simulation mode and then compare the spectral components. This method can be used, but care must be taken to minimize the computational noise and equally spaced small time steps must be used to avoid irregularities at the output spectrum. Then this is paid in the form of long simulation times and large output files. Instead incremental gain can be calculated at 3 different levels to compute IIP3.

$$i = c_0 + c_1 v + c_2 v^2 + c_3 v^3$$

the incremental gain is given by $\frac{di}{dv}$

$$g(v) = \frac{di}{dv} = c_1 + 2c_2 v + 3c_3 v^2$$

$$g(0) = c_1$$

$$g(V) = c_1 + 2c_2 V + 3c_3 V^2 = g(0) + 2c_2 V + 3c_3 V^2$$

$$g(-V) = c_1 - 2c_2 V + 3c_3 V^2 = g(0) - 2c_2 V + 3c_3 V^2$$

$$\Rightarrow \quad g(V) + g(-V) = 2g(0) + 6c_3 V^2 \Rightarrow$$

$$c_3 = \frac{g(V) + g(-V) - 2g(0)}{6V^2}$$ and

$$g(V) - g(-V) = 4c_2 V \Rightarrow$$

$$c_2 = \frac{g(V) - g(-V)}{4V}$$

$$IIP3 = \frac{2}{3}\left|\frac{c_1}{c_3}\right|\frac{1}{R_s}$$

$$= \frac{2}{3R_s}\cdot\left|\frac{g(0)\cdot 6V^2}{g(V)+g(-V)-2g(0)}\right|$$

$$= \frac{4V^2}{R_s}\left|\frac{g(0)}{g(V)+g(-V)-2g(0)}\right|$$

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7-A

Intermodulation & mixing

Let $i = c_0 + c_1 v + c_2 v^2 + c_3 v^3 + \cdots$

if the input v is expressed as

$$v = B\cos w_1 t + D\cos w_2 t$$

1

1 1

1 2 1

1 3 3 1

i becomes only for the cubic term

$$i = c_3 (B\cos w_1 t + D\cos w_2 t)^3$$

$$= c_3 \left[B^3 \cos^3 w_1 t + 3B^2 D \cos^2 w_1 t \cos w_2 t + 3B D^2 \cos^2 w_2 t \cos w_1 t + D^3 \cos^3 w_2 t \right]$$

$$i = c_3 B^3 \cos^3 w_1 t \quad 3B^2 D\left[\tfrac{1}{2} + \tfrac{1}{2}\cos 2w_1 t\right]\cos w_2 t$$ ← intermodulation terms

$$c_3 D^3 \cos^3 w_2 t \quad 3BD^2\left[\tfrac{1}{2} + \tfrac{1}{2}\cos 2w_2 t\right]\cos w_1 t$$

not needed for intermodulation

$$i = 3B^2 D \cancel{\tfrac{1}{2}} \cos w_2 t + \frac{3B^2 D}{2}\cos 2w_1 t \cos w_2 t$$

not needed

$$+ 3BD^2 \cancel{\tfrac{1}{2}} \cos w_1 t + \frac{3BD^2}{2}\cos 2w_2 t \cos w_1 t$$

$$i = \frac{3B^2 D}{2} \cdot \frac{1}{2}\left[\cos(2w_1 - w_2)t + \cos(2w_1 + w_2)t\right]$$

$$+ \frac{3BD^2}{2} \cdot \frac{1}{2}\left[\cos(2w_2 - w_1)t + \cos(2w_2 + w_1)t\right]$$

higher frequency terms ⇒ not needed

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7-B

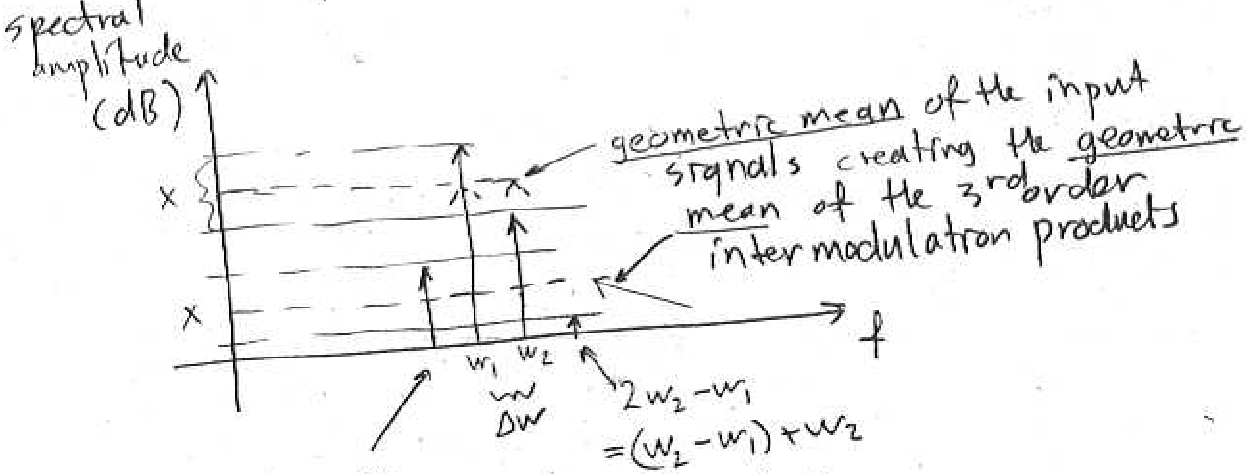
Therefore the intermodulation products created by unequal amplitude signals are given by

$$i = \frac{3}{4} B^2 D \cos(2w_1 - w_2)t + \frac{3}{4} B D^2 \cos(2w_2 - w_1)t$$

so the ratio of the intermodulation products is given by

$$\frac{\cancel{\frac{3}{4}} B^{\cancel{2}} \cancel{D}}{\cancel{\frac{3}{4}} \cancel{B} D^{\cancel{2}}} = \frac{B}{D}$$

At the spectrum they will correspond to



$2w_2 - w_1$
$= (w_2 - w_1) + w_2$
$= w_2 + \Delta w$

$2w_1 - w_2$
$= (w_1 - w_2) + w_1$
$= w_1 - \Delta w$

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Spurious Free Dynamic Range

The dynamic range of an amplifier (or any device) is defined as the range defined by the maximum and minimum signals that can be handled by the amplifier (device). In electronics it is defined as the maximum signal without distorsion (or tolerable distorsion) and the noise level. It is usually expressed as the ratio of the two levels in decibels.

As far as the third-order intermodulation is concerned, the dynamic range can be defined as the range defined by; the maximum input level (of each of the two signals creating intermodulation) which creates 3rd order intermodulation products equal to the output noise level and the minimum input signal level which creates a signal level equal to the noise level at the output.
(The definition referred to the input)
(The horizontal distance marked at the figure as SFDR)

or

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The range defined by; (with the assumption of linear gain)

The output signal created by the input signal level (along with a signal of equal amplitude) which also creates 3rd order intermodulation product equal to the output noise level and the output noise level (The definition referred to the output and it is depicted as the vertical distance marked as SFDR at the figure).

In order to find the input signal level at which the output noise level equals to the IP3 output level, let us write a few equations.

Output noise = F + G + Nin

Output power = Sin + G

Output IP3 at IIP3 = IIP3 + G

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IP3 output power at any S_{in} $= (IIP3 + G) + (S_{in} - IIP3) \times 3$

$= -2 \times (IIP3) + G + 3 \times S_{in}$

Equating IP3 power to output noise for a given S_{in}

$F + G + N_{in} = -2 \times IIP3 + G + 3 S_{in_c} \Rightarrow$

$S_{in_c} = \dfrac{F + N_{in} + 2 \times IIP3}{3}$ S_{in_c} = critical i/p signal level for SFDR

SFDR (Spurious free dynamic range)

$= \underbrace{S_{in_c} + G}_{\text{output signal of the SFDR point}} - \underbrace{(N_i + F + G)}_{\text{output noise}}$ (at the output)

$= \underbrace{S_{in_c}}_{\text{input signal}} - \underbrace{(N_i + F)}_{\text{output noise referred to the input}} \Rightarrow$

$$SFDR = \frac{F + N_{in} + 2 IIP3}{3} - N_i - F$$

$$= \frac{2}{3}\left[IIP3 - \underbrace{(F + N_i)}_{\text{output noise referred to the input}}\right]$$

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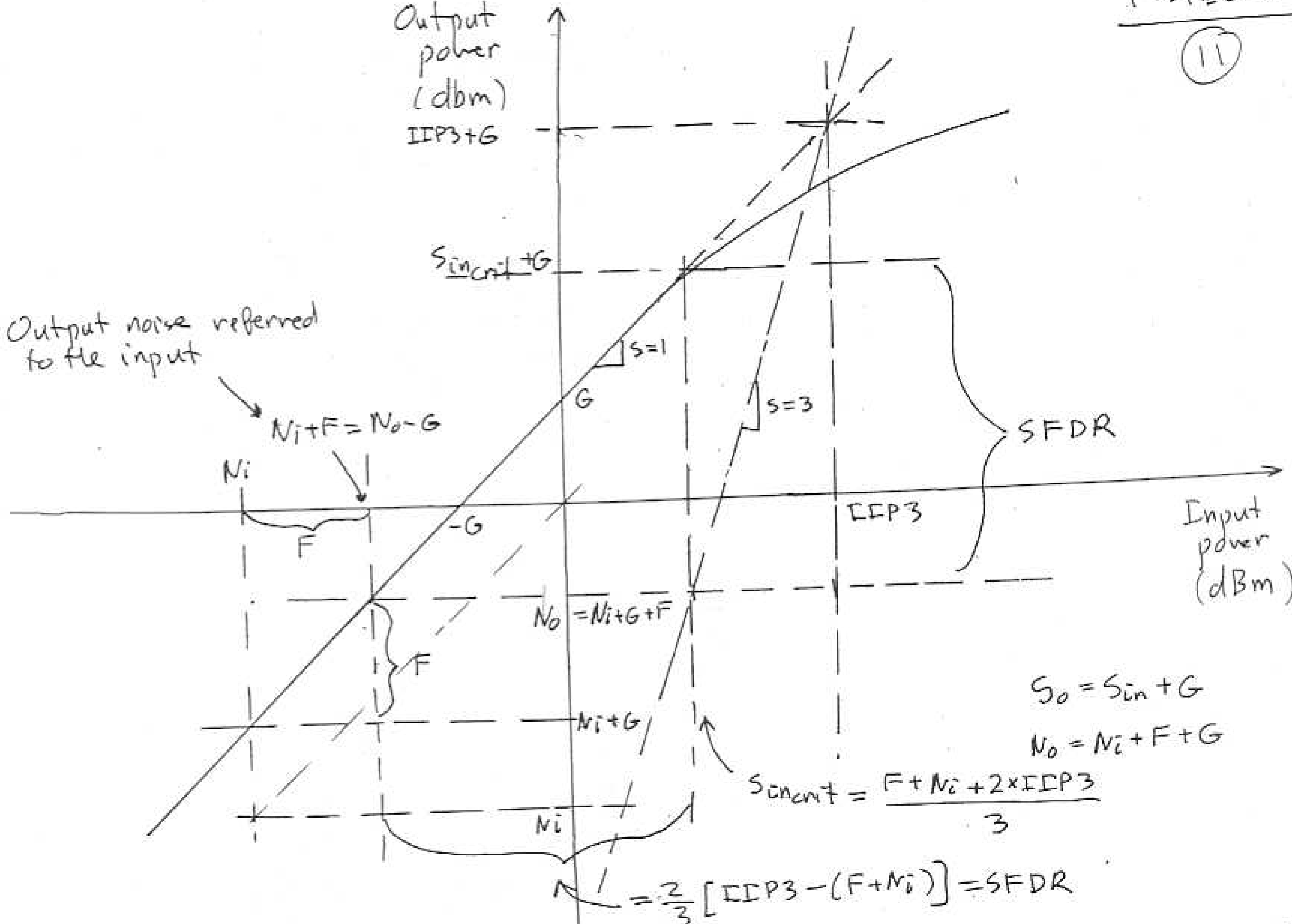
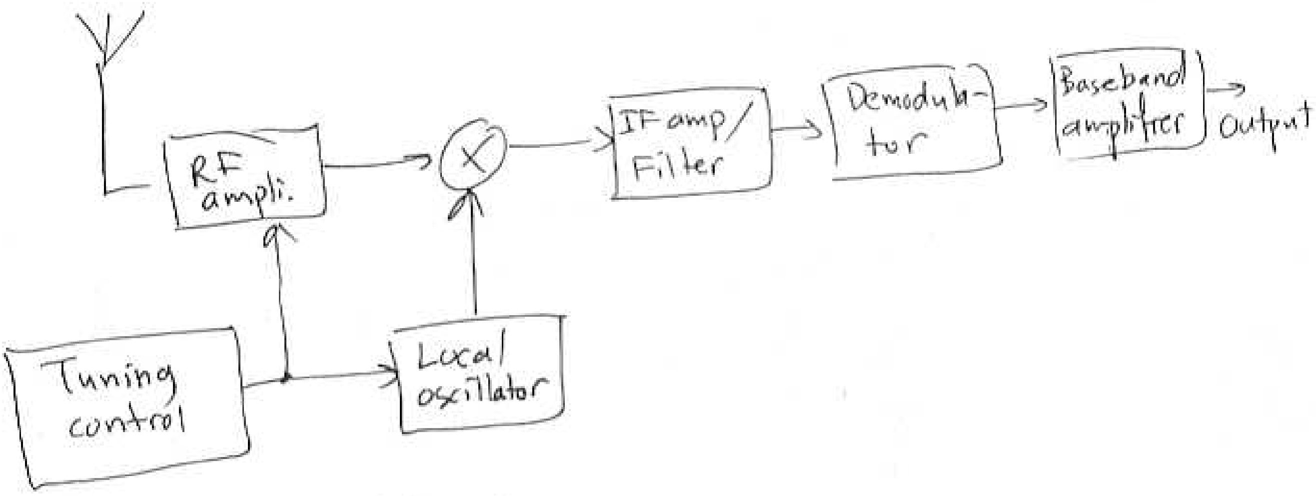


Typical output power vs input power graph for an amplifier

MIXERS



The superheterodyne receiver.

The basic idea is to select a band of frequencies (usually as narrow as possible) by the use of RF amplifier which is also a Low Noise Amplifier.

Then shift the communication band of frequencies (the channel) to the IF filter/amplifier BW and select it by the use of IF filter. The o/p of the IF amplifier/filter is then demodulated and fed to the baseband amplifier. The tunable Local oscillator lets us to use constant frequency IF filter thereby letting us to have better filter characteristics. From IF onwards the receiver chain operates at constant frequencies, therefore their designs can also be optimized better and in an easier way.

The gain of the RF amplifier is optimized in such a way that the output after the mixer is high enough to let us use an ordinary amplifier as far as the noise is concerned. But, it is also made as low as possible to minimize intermodulation. Actually the design of the RF amplifier is a joint optimization of the filter characteristics, noise figure and intermodulation. If it is to be used in a portable system, then power consumption plays an important role in the optimisation.

Another function of the RF amplifier/filter is the image rejection. While the RF input frequency is translated to IF by the mixer, the output of the RF amplifier at the image frequency ($2 \times IF$ away from the desired RF input) also falls into the IF band. Therefore the only protection against image reception in simple heterodyne receivers is the attenuation of the RF filter/amplifier at that frequency.

When designing a superheterodyne receiver, the RF, IF frequency bands and the tunable LO range is chosen such that major intermodulation products does not fall into the IF band as far as possible. Especially the 3rd order intermodulation products create a lot of headache. Harmonics of RF and LO frequencies must certainly be avoided for the IF band. If harmonics of IF fall into the RF band, then the system can oscillate. These are some examples of the pitfalls of superhet design. And by no means they are complete.

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Noise figure (SSB vs. DSB) of Mixers

The noise which falls into the IF band comes from two seperate parts of the spectrum. LO + IF band and LO-IF band. The signal comes from only one of the bands. The usual mixer usage case where the signal comes only from one of the bands the noise figure is called the single sideband noise figure. In the case where the signal comes from both of the bands along with the noise, it is called the DSB noise figure.

The noise figure of the mixer is not solely an inherent or intrinsic property of the mixer. It also depends on the noise characteristics of the L.O. drive. A sharp LO drive is needed to get the most out of the mixer as far as noise is concerned.

Square-law mixers

If we could only have an ideal square-law device represented by

$$i = a_0 + a_1 v + a_2 v^2$$

and the rest of the terms were really equal to zero. Then if we simply add 2 signals as:

$$v = v_{RF} \cos w_{RF} t + v_{LO} \cos(w_{LO} t)$$

and process it by our device, then the output would have been

$$i = a_0 + a_1 v_{RF} \cos w_{RF} t + a_1 v_{LO} \cos w_{LO} t$$

$$+ a_2 \Big[v_{RP}^2 \cos^2 w_{RF} t + v_{LO}^2 \cos^2 w_{BO} t + \underbrace{2 v_{RF} v_{LO} \cos w_{RF} t \cos w_{LO} t}_{\text{multiplication of the 2 inputs}} \Big]$$

Filtering out the rest of terms, we could then have

$$i = \cancel{2} v_{RF} v_{LO} a_2 \left[\frac{\cos(w_{LO} - w_{RF})t + \cos(w_{LO} + w_{RF})t}{\cancel{2}} \right]$$

which acts as a mixer without intermodulation distorsion.

The conversion gain

The amplitude of one of the mixer products is

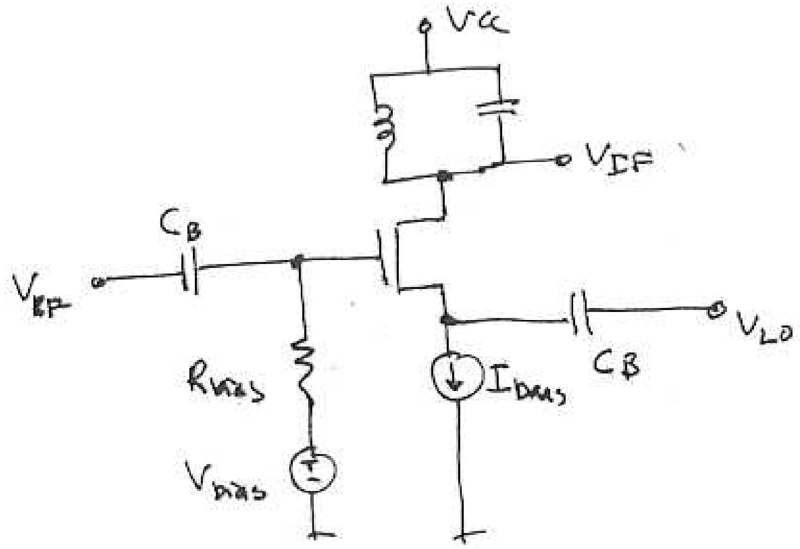
$v_{RF} v_{LO} a_2$

The ratio of this amplitude to the input amplitude is called the conversion gain

$$G_{conv.} = \frac{\cancel{v_{RF}} v_{LO} a_2}{\cancel{v_{RF}}} = v_{LO} a_2$$

In MOSFET's, if the device is long enough and biased appropriately, then the drain current can be expressed as

$$i_d = \frac{\mu C_{ox} W}{2L} (V_{gs} - V_T)^2$$



In the circuit V_{gs} is given by

$$V_{gs} = V'_{bias} + V_{RF} - V_{LO}$$

Substituting into the equation while keeping V_{bias} high enough to keep the FET in the square-law region

$$i_d = \frac{\mu C_{ox} W}{2L} \left[V'_{bias} + V_{RF} - V_{LO} - V_T \right]^2$$

organizing and filtering out the unnecessary terms,

$$i_d = \frac{\mu C_{ox} W}{2L} \left[(V_{RF} - V_{LO}) + (V'_{bias} - V_T) \right]^2$$

$$= \frac{\mu C_{ox} W}{2L} \left[\underbrace{(V_{RF} - V_{LO})^2}_{\text{square-law term}} + \underbrace{2(V_{RF} - V_{LO})(V'_{bias} - V_T)}_{\text{linear terms}} + \underbrace{(V'_{bias} - V_T)^2}_{\text{DC term}} \right]$$

$$= \frac{\mu C_{ox} W}{2L} \left[V_{RF}^2 - 2 V_{RF} V_{LO} - V_{LO}^2 \right]$$

these are also filtered out because they produce D-C + 2w terms

$$i_{cross} = \frac{\mu C_{ox} W}{2L} 2 V_{RF} V_{LO} \cos w_{RF} t \cos w_{LO} t$$

$$i_{cross} = \frac{\mu C_{ox} W}{2L}\left[\cos(\omega_{RF}-\omega_{LO})t + \cos(\omega_{RF}+\omega_{LO})t\right] V_{RF} V_{LO}$$

Therefore $G_{conversion} = G_c$

$$G_c = \frac{\mu C_{ox} W}{2L} \frac{\cancel{V_{RF}} V_{LO}}{\cancel{V_{RF}}} = \frac{\mu C_{ox} W V_{LO}}{2L}$$

Notice that the conversion gain is independent of bias as long as the FET stays in the square-law region. But the conversion gain is still dependent on V_{LO}.

In this configuration we are nearer to the ideal mixer. This mixer has a linear input-output characteristics when converting from one frequency to the other. If it is used within its limits, it can be considered a linear device although it uses a very basic nonlinearity.

In the figure which biases the gate voltage through V_{bias} is meaningless because the actual D.C. gate-to-source voltage is determined by the MOSFET itself depending on I_{BIAS}. It is needed for proper operation of the I_{BIAS} circuity and may be for adjusting the gate-to-drain voltage. That is why I used V'_{bias} instead of V_{bias} because it is different than V_{bias}.

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Single bipolar transistor

$$i_c \cong I_s \, e^{v_{be}/V_T}$$

let $v_{be} = v_{in} + v_{bias}$ — v_{in} being the AC component

$$i_c \cong I_s \, e^{v_{in}/V_T} \cdot e^{v_{bias}/V_T} \qquad V_T = \frac{kT}{q} = 26\,mV$$

$$\Rightarrow i_c = \underbrace{I_s \, e^{v_{bias}/V_T}}_{I_C} \cdot e^{v_{in}/V_T}$$

I_c being the D.C. current flowing

$$i_c \cong I_c \, e^{v_{in}/V_T}$$

expanding e^{v_{in}/V_T} into power series and taking the terms up to the square term (the rest are related to the intermodulation)

$$i_c \cong I_C \left[1 + \frac{v_{in}}{V_T} + \frac{1}{2}\left(\frac{v_{in}}{V_T}\right)^2 \right]$$

filtering out the linear terms

$$i_c \cong \frac{I_c}{2V_T^2}\left(v_{RF}\cos w_{RF}t + V_{LO}\cos w_{LO}t\right)^2$$

$$= \frac{I_c}{2V_T^2}\left\{ 2v_{RF}V_{LO}\left[\frac{\cos(w_{RF}+w_{LO})t + \cos(w_{RF}-w_{LO})t}{2}\right] + \cdots \right]$$

$$\Rightarrow G_C = \frac{I_c}{2V_T^2}\,\frac{\cancel{v_{RF}}V_{LO}}{\cancel{V_{RF}}} = \frac{I_c V_{LO}}{V_T \times 2V_T}$$

but $\frac{I_c}{V_T} = g_m \Rightarrow G_C = \frac{g_m V_{LO}}{2V_T}$ where V_T is the thermal voltage $= 26\,mV$

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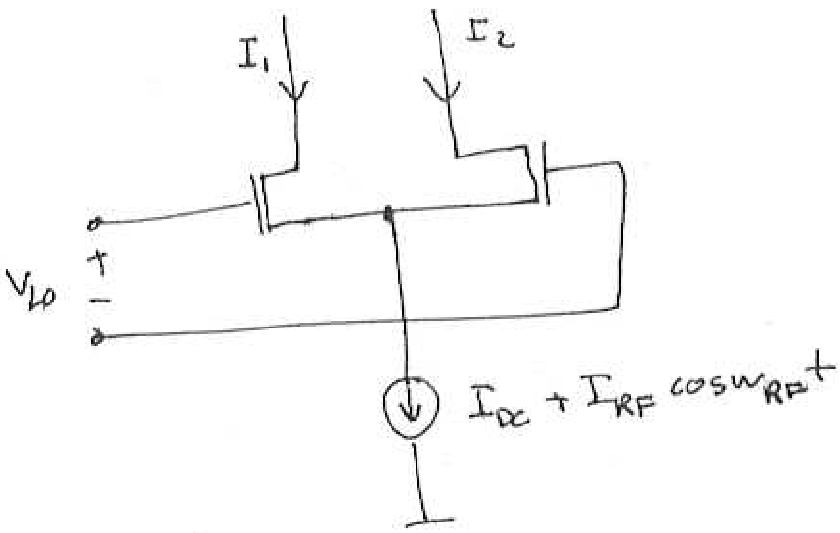
since at the linear term
the coefficient is I_C/V_T which determines
$g_m \Rightarrow g_m = \frac{I_C}{V_T}$ obviously

Multiplier based mixers

Single balanced mixers



If small enough signals are applied the output is like a square law mixer where the output amplitude is proportional with both V_{LO} and I_{RF}. But if V_{LO} is high enough that it effectively swithes the drain currents on and off.
The output being $I_1 - I_2 = i_o(t)$

$$i_o(t) = \underbrace{\text{sgn}\left[\cos w_{LO} t\right]}_{\text{odd harmonics + fundamental}} \underbrace{\left[I_{bias} + I_{RF} \cos w_{RF} t\right]}_{\text{DC + rf current to prevent nonlinearity}}$$

In contrast to square-law and BJT mixers, neither the fundamental, nor the harmonics of the RF drive exist at the output of the single balanced mixer.

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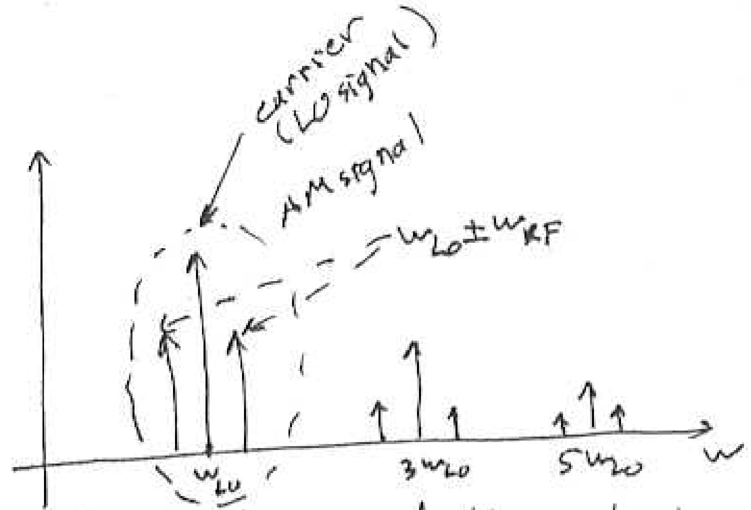
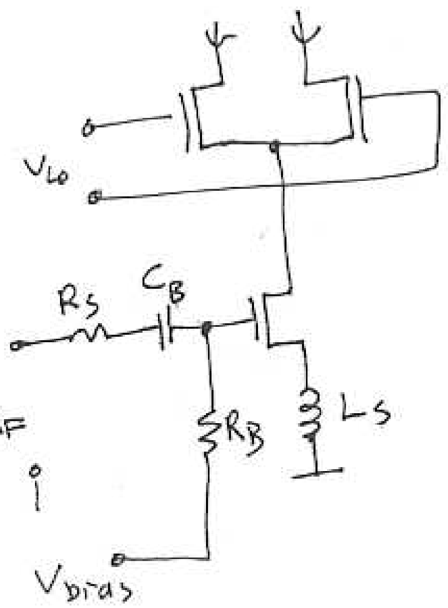


the spectrum of the output current looks like above

The linearity of the mixer depends on the linearity of the current source. For this reason some kind of source degeneration is needed. Resistive degeneration can be used but it adds noise. Inductive degeneration is preferred because of noise like an LNA and it also suppresses higher harmonics. An example is given below.



Here L_S creates linearization and R_B is high enough to add less noise and not to load v_{RF}. V_{BIAS} this time is effective in determining the operating point.

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The fundamental component of

$sgn[\cos\omega_{LO}t]$

has the amplitude $\frac{4}{\pi}$. Therefore the mixer products of the fundamental can be written as

$$i_{of}(t) = \frac{4}{\pi}\cos\omega_{LO} + x\, I_{RF}\cos\omega_{RF}t$$

$$= \frac{4}{\pi}\cdot\frac{I_{RF}}{2}\left[\cos(\omega_{LO}t+\omega_{RF})t + \cos(\omega_{LO}-\omega_{RF})t\right]$$

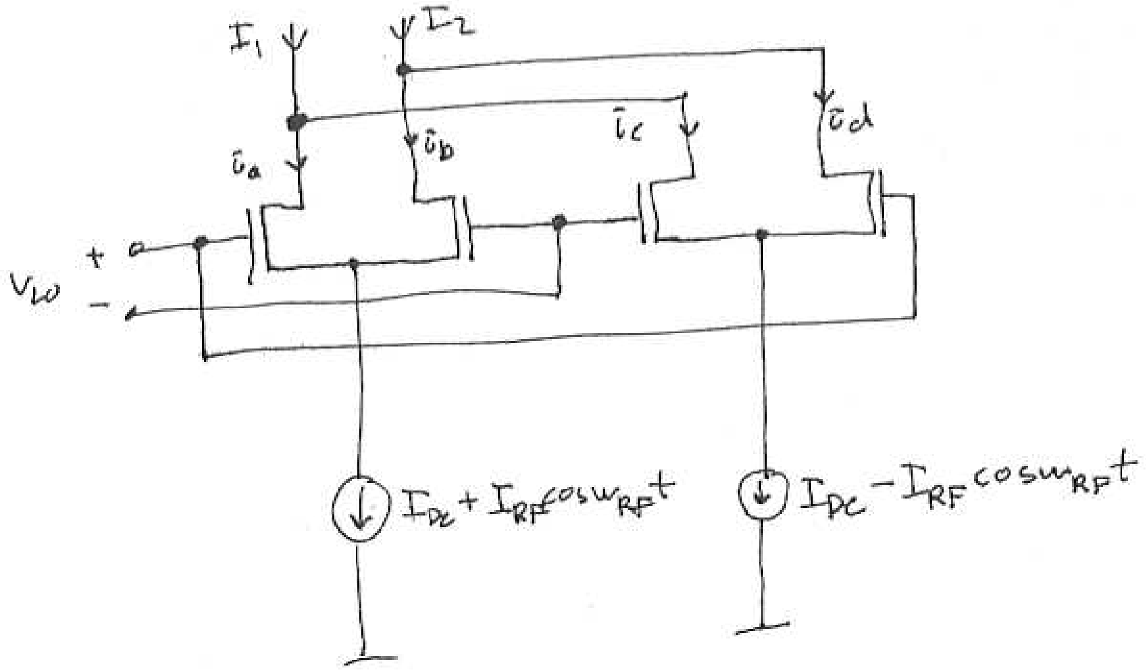
substituting $I_{RF} = g_m v_{RF}$ to include the gain of the current converter in the equation

$$i_{of} = \frac{2}{\pi} g_m v_{RF}\left[\cos(\omega_{LO}+\omega_{RF})t + \cos(\omega_{LO}-\omega_{RF})t\right]$$

$\Rightarrow$ $G_c = \frac{2g_m}{\pi}$ is obtained for the conversion gain.

Double-balanced mixers

In order to cancel out the fundamental and the harmonics of the LO drive from the o/p spectrum (i.e. to approach more to the ideal multiplier) along with the cancelled fundamental and harmonics of the RF drive as in the single-balanced mixer, the multiplier structure is repeated and connected in such a way that the sidebands are added up, but the LO components are cancelled as shown below.



$$i_o(t) = I_1 - I_2$$

$$i_a - i_b = sgn[\cos w_{LO}t][I_{DC} + I_{RF}\cos w_{RF}t]$$

$$i_c - i_d = -sgn[\cos w_{LO}t][I_{DC} - I_{RF}\cos w_{RF}t]$$

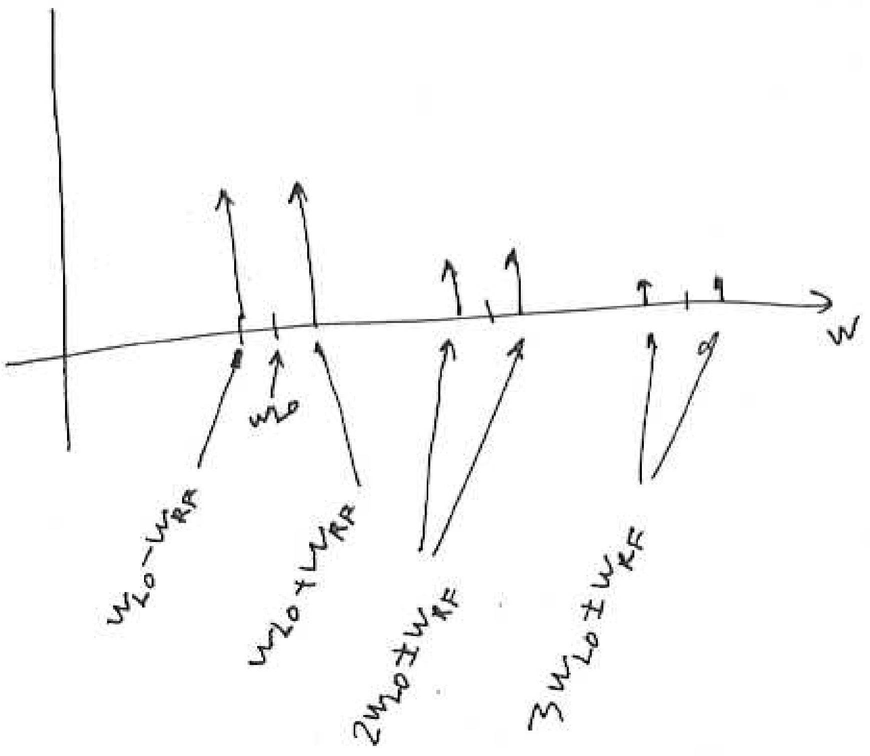
$$i_c - i_d = sgn[\cos w_{LO}t][-I_{DC} + I_{RF}\cos w_{RF}t]$$

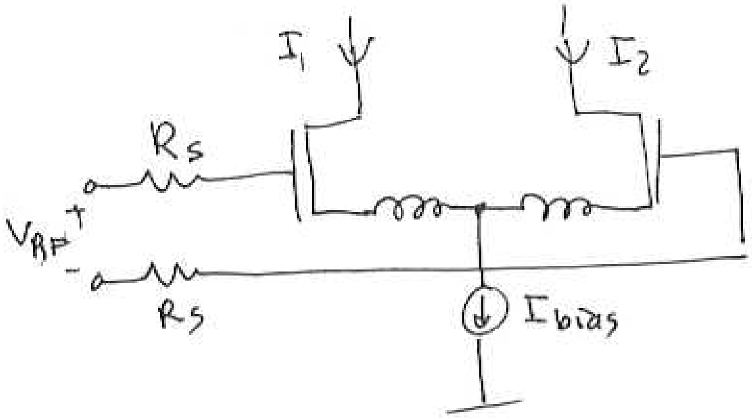
$$i_o(t) = I_1 - I_2 = i_a + i_c - i_b - i_d = (i_a - i_b) + (i_c - i_d)$$

$$= \mathrm{sgn}[\cos w_{LO} t]\left[\cancel{I_{DC}} + I_{RF}\cos w_{RF} t - \cancel{I_{DC}} + I_{RF}\cos w_{RF} t\right]$$

$$= 2\,\mathrm{sgn}[\cos w_{LO} t]\; I_{RF}\cos w_{RF} t$$

We still have the harmonics of LO drive in the multiplication producing the harmonic sidebands. Therefore the spectrum looks like:

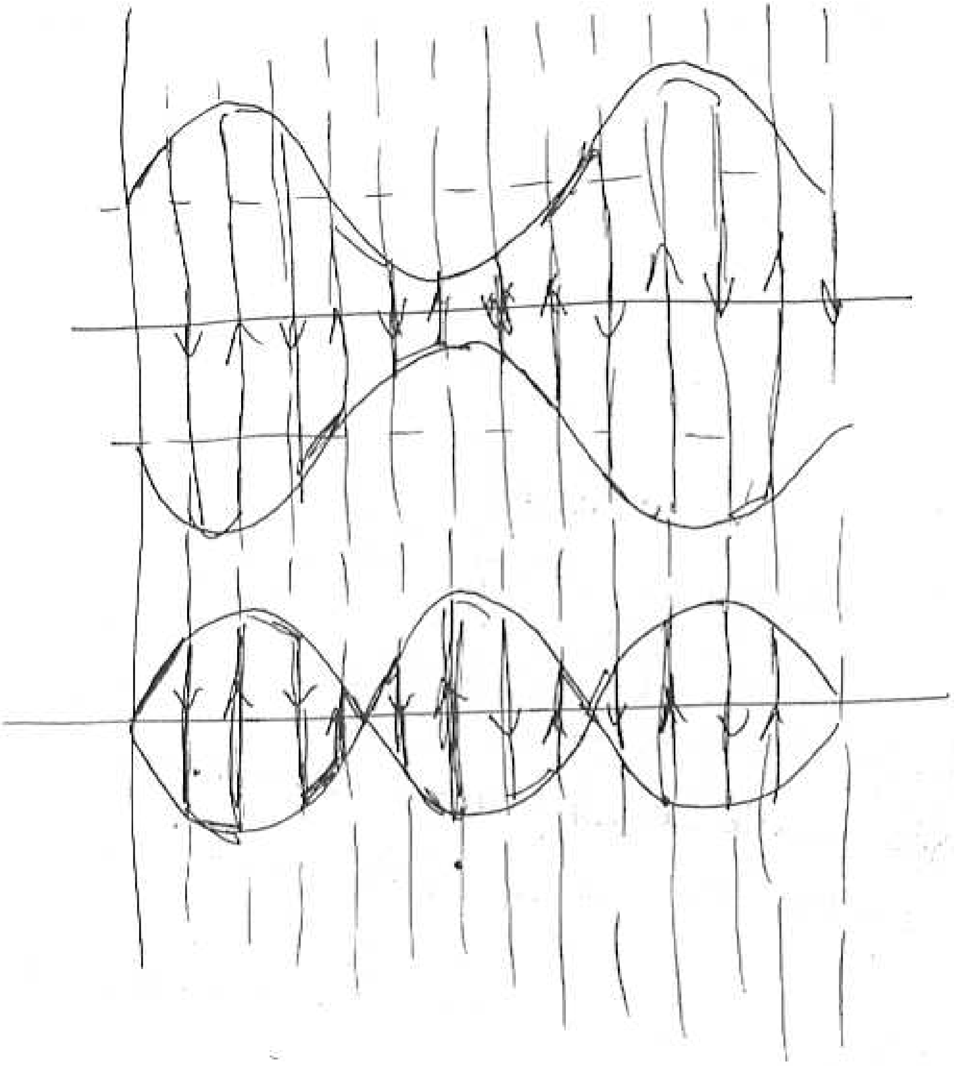




Linearized differential RF transconductor for double-balanced mixer.

Noise in current mode mixers can be quite high due to different reasons. Practical mixers of this type can exhibit SSB noise figures at least 10dB with values ranging to 15dB.

Linearity of the mixers can be treated in a similar manner to amplifiers, defining IIP3 the same way



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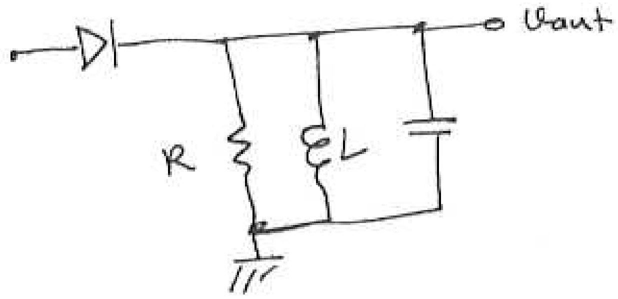
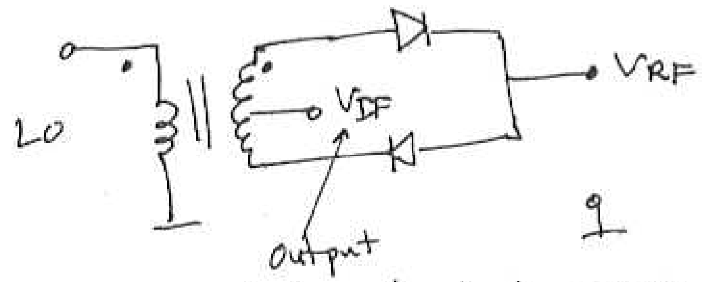
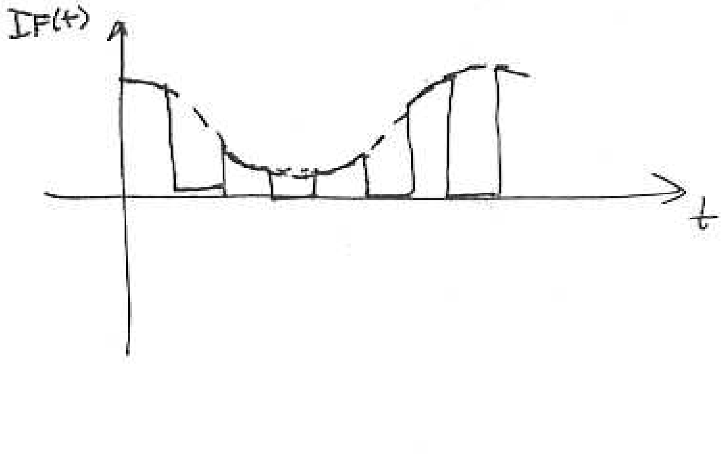


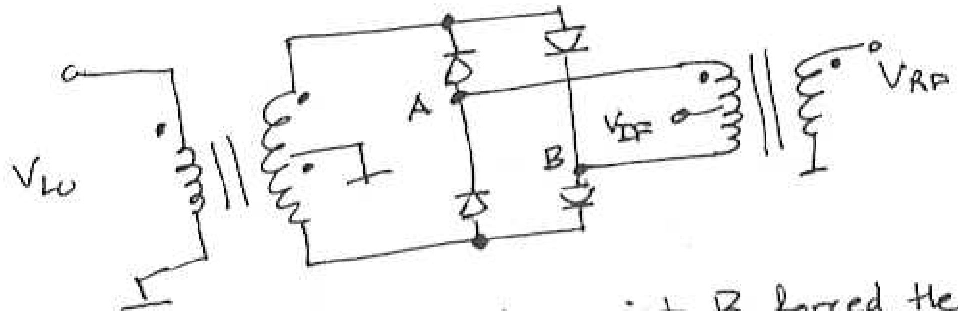
Single diode mixer choosing the right frequency component making use of the tank circuit



Single balanced diode mixer; if LO is sufficiently large RF is conducted to IF (at the positive half cycle). At the negative half cycle RF is cut off from IF. Therefore the output looks like:



Double-balanced mixer



At the positive cycle point B forced the potential of the ground because the LO drive is symmetric around the ground. Therefore conducting the RF voltage to IF. At the negative cycle of the LO, this time point A is forced to the ground potential, therefore connecting negative RF to voltage to the IF port, effectively multiplying the RF port voltage with $\mathrm{sgn}[\cos \omega_{LO} t]$. Therefore the resulting operation is double-balanced mixing.

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Intermodulation in mixers (again)

The intermodulation products and definitions and terms stated for an amplifier are also used for mixers, too.

Look back at 19.11.2003/3

When designing a circuit containing a mixer keep away from $2\omega_1, 3\omega_1, 4\omega_1, 2\omega_2, 3\omega_2, 4\omega_2$

$2\omega_1 \pm \omega_2$, $2\omega_2 \pm \omega_1$ at least try to keep away from $\pm n\omega_1 \pm m\omega_2$ as possible as you can.

Terminate all three ports of the mixer with narrow-band terminations inorder to minimize frequency interactions and D.C. terms introduced by self mixing. Terminating them with characteristic impedance at all frequencies might look plausable. But, it also increases chances of spurious pick-ups at various frequencies.