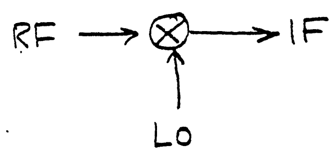
Mixer

A mixer is a non-linear device which produces signals at the sum and difference of the input frequencies. Basically, any non-linear device can serve as a mixer but diodes, FET's and BJT's are the most common.

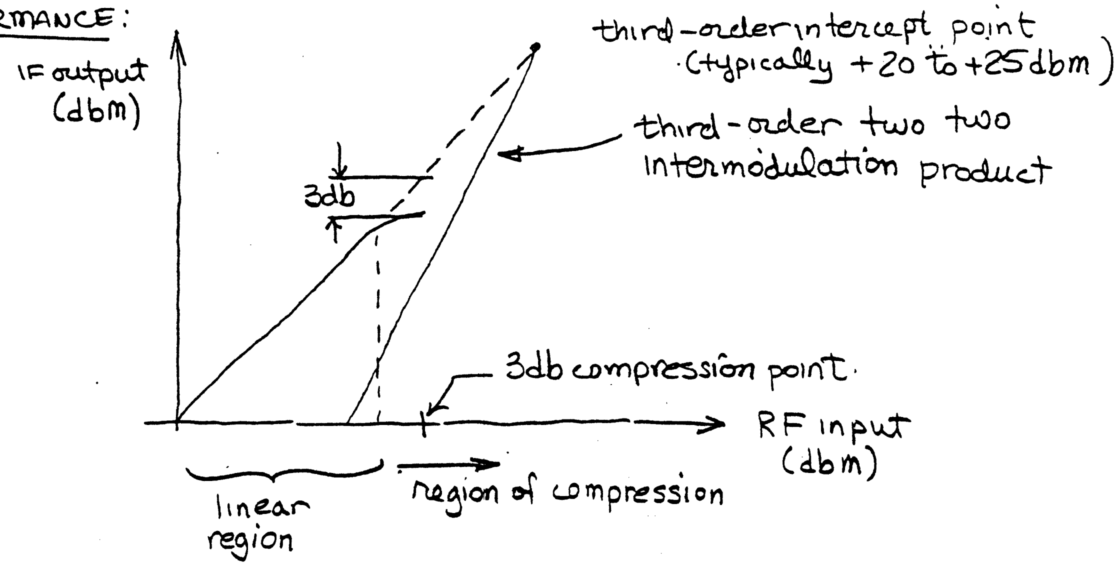
BASIC MIXER THEORY & DEFINITIONS

Schematically, we can write a mixer as



The large signal is usually known as the local oscillator (LO). The other input signal is the r.f. signal which, in receivers, is usually quite low in amplitude. The mixer output is an intermediate (IF) frequency. This is usually a single frequency with a relatively narrow bandwidth for receiver applications.

MIXER PERFORMANCE:



The 3-db compression point marks the upper end of r.f. signal strength for good mixer operation. A common mixer measure is the third-order, two tone intermodulation product. A third order IM product is a sum/difference of the 2nd harmonic of one signal and another signal, i.e.

$$2f_{RF} \pm f_{LO}$$

$$\text{or} \quad 2f_{LO} \pm f_{RF}$$

Components of output current in a mixer with the characteristics

$$i_0 = aV_1 + bV_1^2 + cV_1^3$$

and with inputs:

case A $\quad V_1 \cos\omega_1 t$

case B $\quad V_1 \cos\omega_1 t + V_2 \cos\omega_2 t$

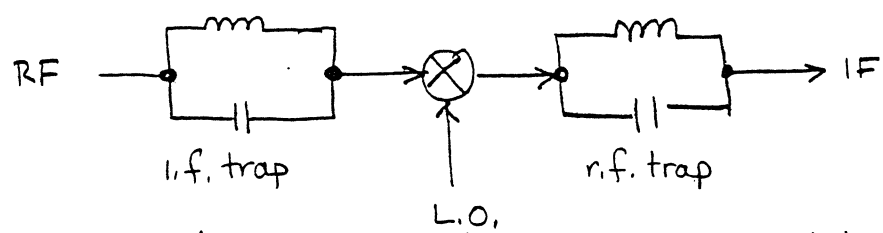
	First order	Second order	Third order
case A	$aV_1 \cos\omega_1 t$	$\frac{b}{2}V_1^2 + \frac{b}{2}V_1^2\cos 2\omega_1 t$	$\frac{3}{4}cV_1^3\cos\omega_1 t + \frac{c}{4}V_1^3\cos 3\omega_1 t$
case B	$aV_1\cos\omega_1 t$ $+aV_2\cos\omega_2 t$	$\frac{b}{2}(V_1^2+V_2^2)$ $+\frac{b}{2}V_1^2\cos 2\omega_1 t$ $+\frac{b}{2}V_2^2\cos 2\omega_2 t$ $+bV_1V_2\cos(\omega_1+\omega_2)t$ $+bV_1V_2\cos(\omega_1-\omega_2)t$	$\left(\frac{3}{4}cV_1^3+\frac{3c}{2}V_1V_2^2\right)\cos\omega_1 t$ $+\left(\frac{3}{4}cV_2^3+\frac{3c}{2}V_1^2V_2\right)\cos\omega_2 t$ $\frac{c}{4}V_1^3\cos 3\omega_1 t+\frac{c}{4}V_2^3\cos 3\omega_2 t$ $\frac{3c}{4}V_1^2V_2\left[\cos(2\omega_1+\omega_2)t + \cos(2\omega_1-\omega_2)t\right]$ $+\frac{3}{4}cV_1V_2^2\left[\cos(2\omega_2+\omega_1)t + \cos(2\omega_2-\omega_1)t\right]$
	these are the first order terms.	these are the second order terms	these are the third order terms

usually $c \ll b$

In general m,n product outputs at

$$nf_{LO} \pm mf_{RF}$$

considerations in mixer design depend upon device choice, i.e. stability and gain. If we use an active mixer, i.e. one which exhibits gain, then that device can be unstable – just like an amplifier, and we must utilize the Linvill and Stern criteria to get reasonable C and K values by choosing proper terminations.



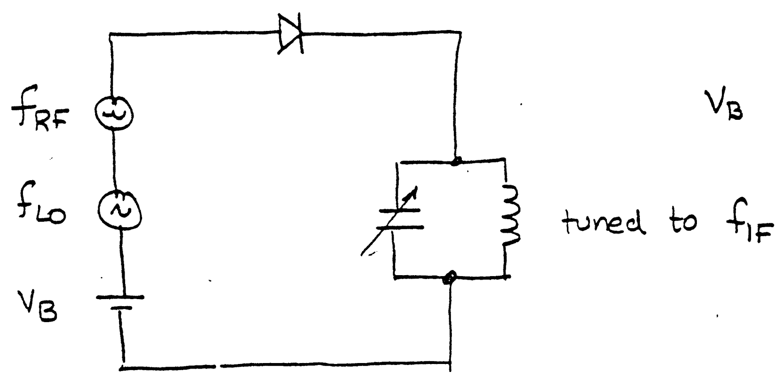
One way of stopping mixer oscillation not covered by the stability criteria is to use a parallel resonant circuit as a "trap" at the r.f. input and i.f. output of the mixer. The r.f. input trap is tuned to reject (or "trap") the i.f. frequencies; the r.f. trap in the i.f. output "traps" (or rejects) r.f. in the i.f. output. This approach eliminates r.f. feedback between the i.f. out and r.f. input ports.

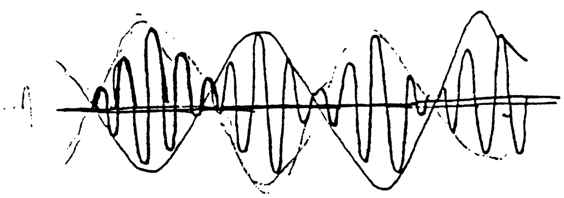
If you use an active device you will be concerned with the conversion gain which will be defined to be

$$\text{conversion gain} = \frac{\text{IF power output}}{\text{RF power input}}$$

In general, a BJT mixer will have a conversion gain equal to that of a conventional amplifier.

simple diode mixer

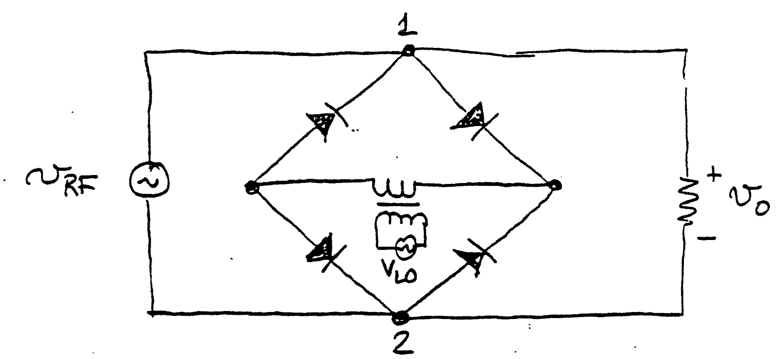




mixer output consists of positive sum of voltages.

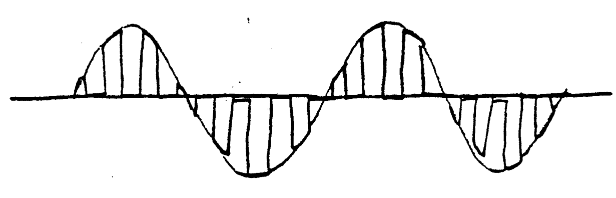
In this type of mixer there is NO gain, only loss. There also is no isolation between the RF and LO ports which can lead to radiation and feed back problems.

single balanced diode mixer

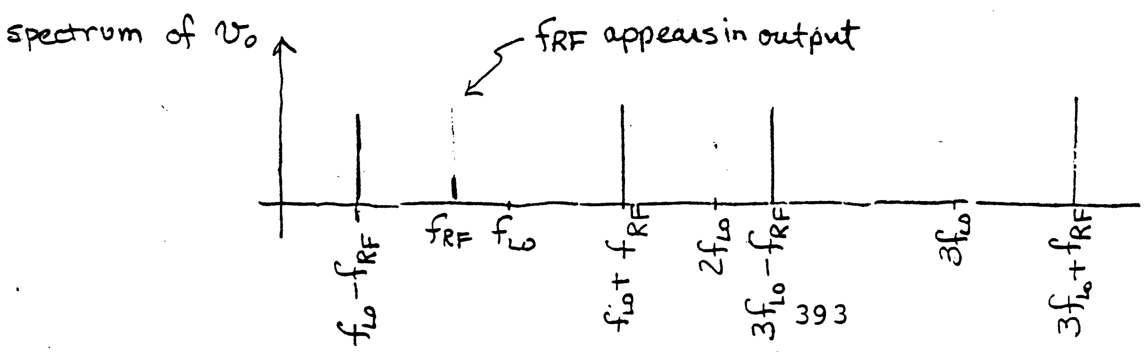


$v_{RF} \ll v_{LO}$

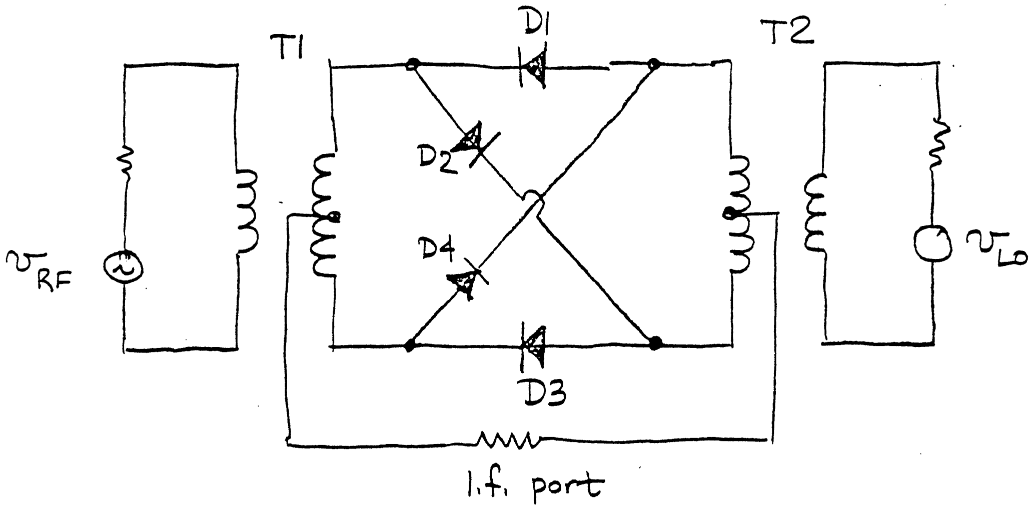
v_{LO} turns the diodes on and off acting to short v_{RF} between 1 and 2.



This gives NO components at f_{LO} or its odd harmonics at the output, hence, single balanced mixer, which simplifies the filter problem

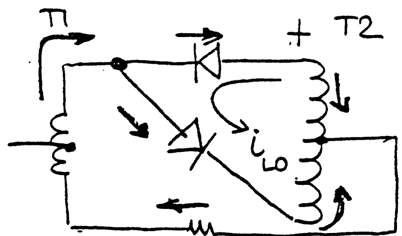


double balanced mixer



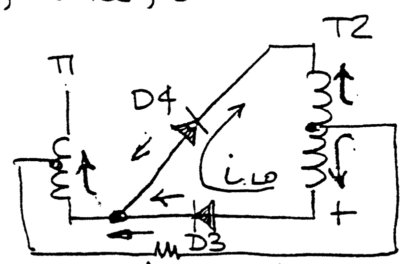
$v_{LO} \gg v_{RF}$

There is no r.f. or l.o. in mixer output; hence, double balanced.



top of T2 is positive
turning D1, D2 on

r.f. signal flows thru top of T1, thru D1 and D2 into T2 thru T2's centertap back to the centertap of T1. Note no i_{LO} in r.f. transformer T1. r.f. currents in i.f. transformer T2 cancel.

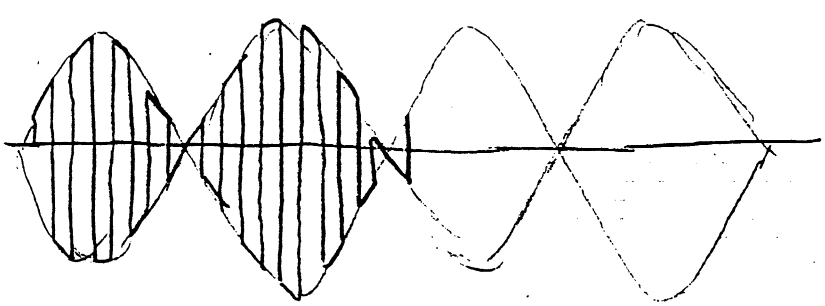


in other di[...] due to phas[...] of T1

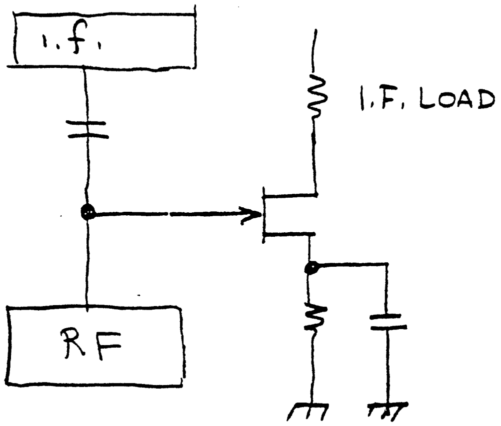
bottom of T2 is positive
turning D3, D4 on.

r.f. signal current flows thru bottom of T1 thru D3, D4 from centertap of T2. No i_{LO} in T1, no net i_{RF} in T2.

does not remain one-sided

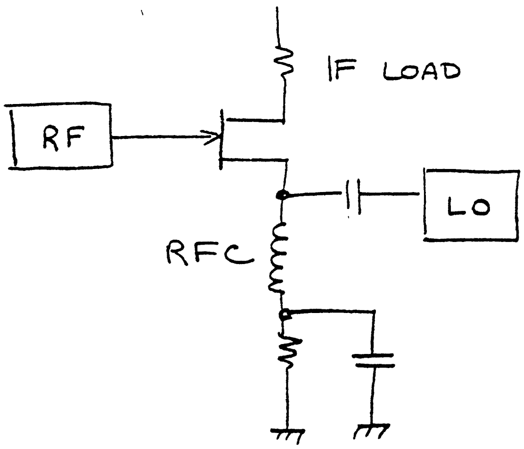


FET mixers. (all single ended)

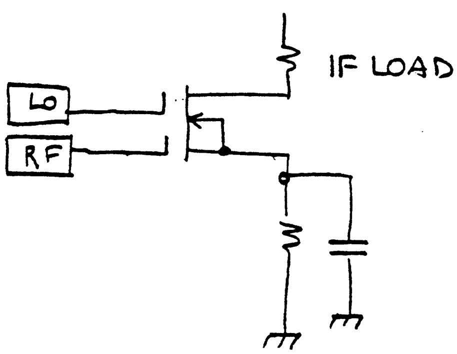


just like original diode mixer
— poor isolation

but
— conversion gain
— approximate square law device



— better isolation
— need more LO power since lower impedance at source than gate

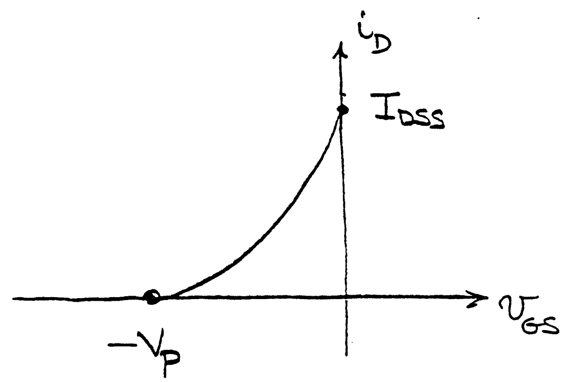


— good isolation
— lower gain

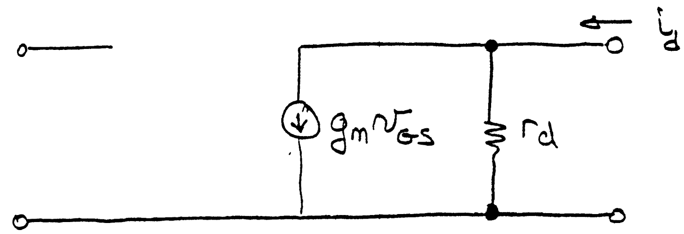
can do all these with BJT's as well.

n-channel JFET mixer:

$$i_D = I_{DSS}\left(1 - \frac{v_{GS}}{V_p}\right)^2$$



in small signal model



$$i_d = g_m v_{GS} + \frac{V_{DS}}{r_d} \approx g_m v_{GS}$$

since r_d very large.

look at g_m.

$$g_m = \frac{\partial i_D}{\partial v_{GS}} = \frac{\partial}{\partial v_{GS}}\left[I_{DSS}\left(1 - \frac{v_{GS}}{V_p}\right)^2\right] = I_{DSS}(2)\left(1 - \frac{v_{GS}}{V_p}\right)\left(-\frac{1}{V_p}\right)$$

$$= \underbrace{-\frac{2I_{DSS}}{V_p}}_{g_{mo}}\left(1 - \frac{v_{GS}}{V_p}\right) = g_{mo}\left(1 - \frac{v_{GS}}{V_p}\right)$$

g_{mo} = value of g_m at $v_{GS} = 0$

g_m is also a function of v_{GS}.

$$\Rightarrow \quad i_d \approx g_{mo}\left(1 - \frac{v_{GS}}{V_p}\right) v_{GS} = \underbrace{g_{mo} v_{GS}}_{\text{first order term}} - \underbrace{\frac{g_{mo}}{V_p} v_{GS}^2}_{\text{second order term}}$$

define a conversion gain $g_c = \dfrac{\text{IF current out}}{\text{RF voltage in}}$

let $v_{GS} = \underbrace{V_{GS}}_{\text{d.c. term}} + \underbrace{V_{LO}\cos\omega_{LO}t}_{\text{local oscillator}} + \underbrace{V_{RF}\cos\omega_{RF}t}_{\text{r.f. input}}$

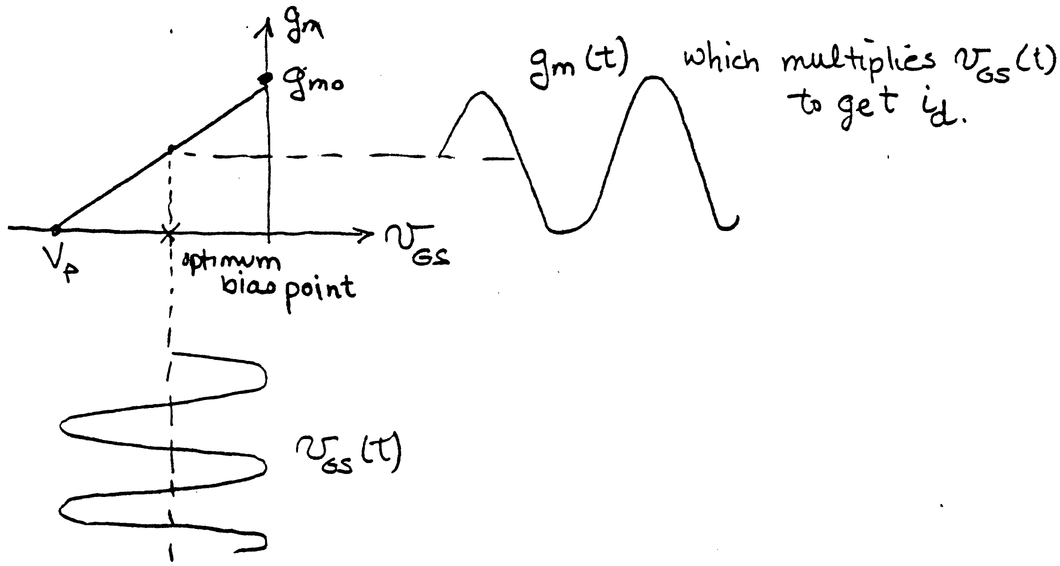
2nd order term.

$$\frac{g_{mo}}{|V_p|}\left[V_{LO}V_{RF}\cos\omega_{LO}t\cos\omega_{RF}t\right]$$

$$\frac{2g_{mo}}{|V_p|}\left[\frac{1}{2}\cos(\omega_{RF}-\omega_{LO})t+\frac{1}{2}\cos(\omega_{RF}+\omega_{LO})t\right]$$

$$\therefore\quad g_c = \frac{\frac{2g_{mo}}{|V_p|}V_{LO}V_{RF}\,\frac{1}{2}\,\cancel{\cos(\omega_{RF}-\omega_{LO})t}}{V_{RF}\,\cancel{\cos\omega_{RF}t}} = \frac{g_{mo}V_{LO}}{|V_p|}$$

Note that g_c is nothing more than the slope of the $g_m - v_{GS}$ curve and can be used to graphically examine mixer operation. The larger the slope – the larger the gain.



mixer design example

to design a 30 MHz mixer for a 5 MHz i.f. output. The l.o. is at 35 MHz.

The solution is a BJT mixer with both r.f. and L.O. injected into the base. We will use a 2N2221A transistor biased at $I_C = 2mA$, $V_{CE} = 10$ volts.

2N2221A $\quad (I_C = 2mA,\ V_{CE} = 10 \text{ volts})$

$$(y_{11})_{ce} = 6.25 + j9.5 \text{ m℧} \quad @\ 30\text{ MHz}$$

$$(y_{22})_{ce} = 0.027 + j0.28 \text{ m℧} \quad @\ 5\text{ MHz}$$

since both inputs are near 30 MHz, i.e. 30 and 35 MHz, we can get by with using the 30 MHz parameters for both.

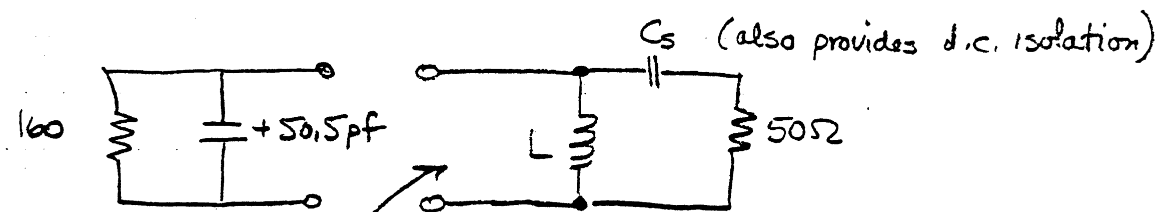
The transistor input @ 30 MHz looks like a resistor in parallel with a capacitor.

$$(y_{11})_{ce} = (g_{11})_{ce} + (jb_{11})_{ce}$$

$$(g_{11})_{ce} \Rightarrow 160\Omega \text{ resistor}$$

$$+(b_{11})_{ce} \Rightarrow j\omega C = j9.5\times10^{-3} \quad \therefore\ C = \frac{9.5\times10^{-3}}{2\pi(30\times10^{6})} = +50.5\text{p}$$

For a 50Ω local oscillator we could use an L-network.



160Ω, $-j9.5$m℧ for conjugate matching

Use KBR formulas

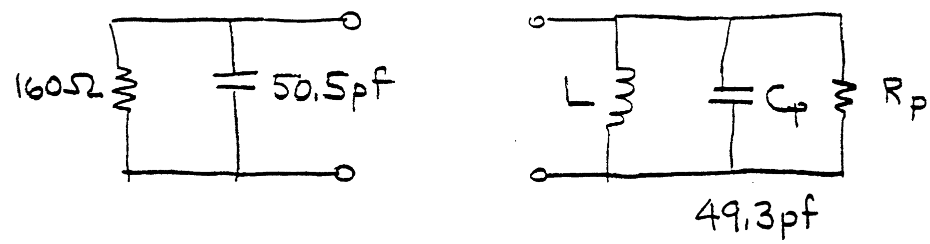
$$Q_t^2 = \frac{R_t}{R} - 1 = \frac{160}{50} - 1 = 2.2 \quad \text{(assumes match)}$$

$$Q_t = 1.48$$

$$C_s = \frac{1}{\omega_o R_t Q_t} = \frac{1}{2\pi(30\times10^6)(50)(1.48)} = 71.7\times10^{-12}$$

convert to parallel to compute L

$$C_p = C_s\left(\frac{Q_t^2}{Q_t^2+1}\right) = 71.7\text{pf}\ \frac{2.2}{2.2+1} = 49.3\text{ pf}$$

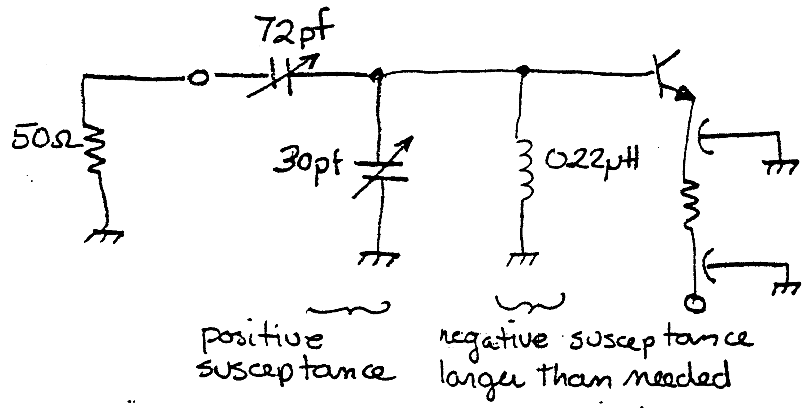


To get resonance then

$$L = \frac{1}{\omega_o^2 C_{TOT}} = \frac{1}{(2\pi \times 30 \times 10^6)^2 (99.8 \times 10^{-12})} = 0.28\,\mu H$$

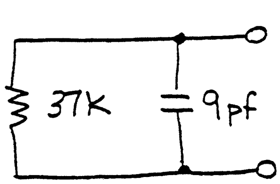
This is a small but reasonable value.

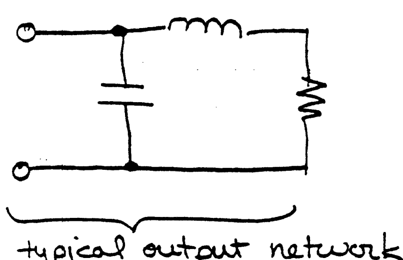
For a practical circuit we pick 0.22 μH as the nearest standard L value and add a negative susceptance (a 30 pf variable capacitor) to tune its reactance.



Analyze the output circuit in a similar manner.

$$(y_{22})_{ce} = (0.027 + j0.28)\ m℧$$





$$Q_t^2 = \frac{R_t}{R} - 1 = \frac{37K}{50} - 1 = 739 \qquad Q_t = 27.2$$

$$Q_t = \frac{\omega_o L}{R} \quad \text{or} \quad L = \frac{Q_t R}{\omega_o} = \frac{(27.2)(50)}{2\pi(5\times10^6)} = 43.3\,\mu H$$

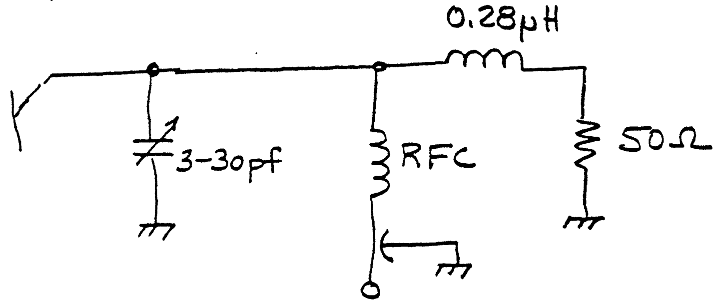
Convert to series circuit to get proper C.

$$L_p = \frac{Q_t^2+1}{Q_t^2} L_s = \frac{739+1}{739} L_s \approx L_s = 43.3\,\mu H$$

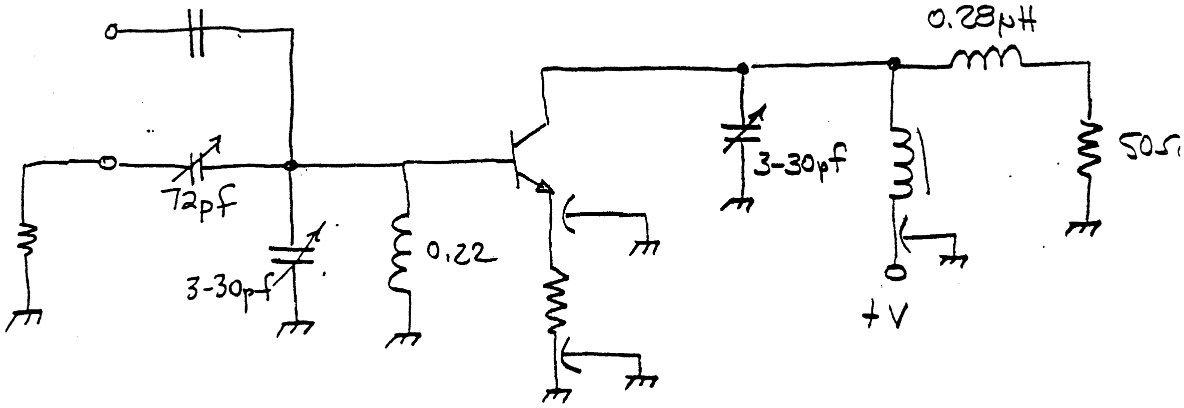
to resonate $C = \dfrac{1}{\omega_o^2 L} = \dfrac{1}{(2\pi\times5\times10^6)^2(43.3\times10^{-6})}$

$= 23.4$ pf.

Add a parallel capacitor $C = 23.4 - 9 = 14.4$ pf at output.



total circuit



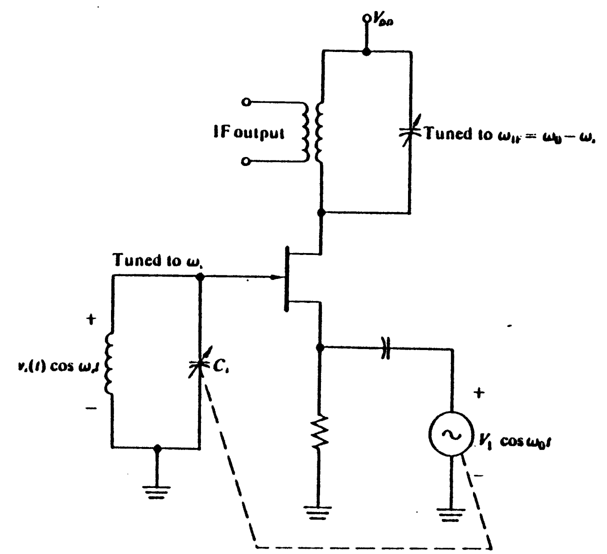


Fig. 7.4-5 FET mixer.

7.5 SEMICONDUCTOR CONVERTER CIRCUITS

Though a separate oscillator and a separate mixer circuit can normally be designed so that each does its own job best, it is possible to combine the two functions in a single active device. This combination is known as a converter. High-quality receivers usually keep these two functions separate, whereas most mass-produced receivers combine them.

Figure 7.5-1 shows a typical bipolar transistor converter stage. (A two-power-supply version is shown to reduce the complexity of the drawing slightly. The reader should have no difficulty in visualizing this circuit in a single-power-supply form.)

In such a circuit we design the oscillator circuit to give the driving level the value of $x = qV_1/kT$ desired for the mixer operation. In designing the oscillator we would like to be able to neglect the input signal circuit and the output IF circuit. Normally one does neglect them, and then checks up by showing that the voltage drops across these impedances caused by the calculated currents are truly negligible.

As in all mixing situations, one must worry about oscillator amplitude variation across the band, signal and oscillator circuit tracking, and mixer interference and distortion problems. As usual, compromises will be necessary. For example, the amplitude stability of the circuit may be improved by increasing the amplitude of

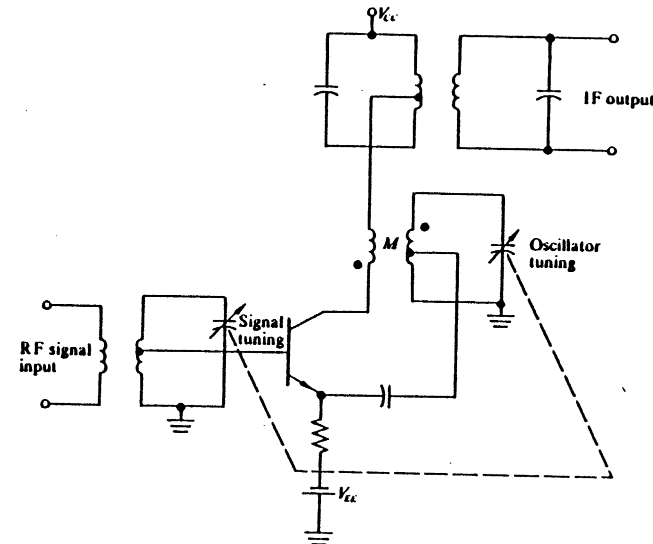


Fig. 7.5-1 Bipolar transistor converter.

the oscillations, while the interfering output terms from oscillator harmonics or the difficulties with excessive oscillator voltage in either the IF or the input circuit are all minimized by decreasing the amplitudes of the oscillations to their smallest possible values.

Any of the previous mixer circuits may be turned into a converter by combining it with the appropriate oscillator circuit from Chapter 6.

Example 7.5-1 For the converter shown in Fig. 7.5-2, find an expression for $v_o(t)$.

Solution. If at the oscillator frequency $\omega_0 = 1/\sqrt{L_3C_3} = 1.5 \times 10^7$ rad/sec the impedance of the input-tuned circuit is an effective short circuit compared with the base-emitter impedance of the transistor, then the oscillation amplitude and frequency for the converter may be found by grounding the base of the transistor and employing the results of Section 6.4. Specifically,

$$\omega_0 = 1.5 \times 10^7 \text{ rad/sec},$$

$$\frac{G_m(x)}{g_{mQ}} \approx \frac{G_L}{ng_{mQ}} = \frac{G_L L_3}{g_{mQ} M_{34}} = 0.19,$$

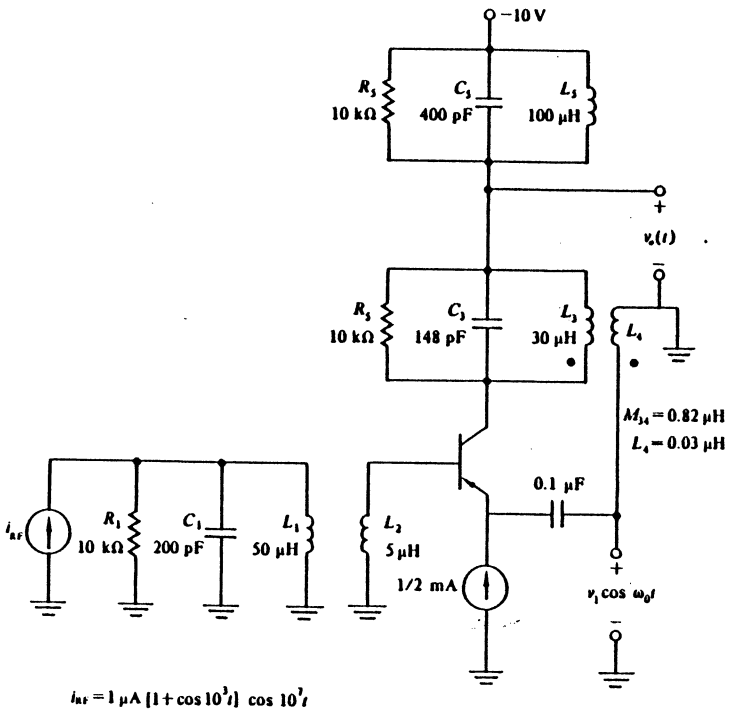


Figure 7.5-2

and from Fig. 4.5-6 $x = 10 (V_1 = 260$ mV). In addition, since the loop gain is much less than unity at any other frequency, spurious oscillations at the RF or IF frequencies are not possible.

If we also assume that the input RF-tuned circuit is not loaded by the transistor, then the transistor base voltage is given by

$$v_B(t) = \frac{i_{RF} R_1 M_{12}}{L_1} = 1.57 \text{ mV}(1 + \cos 10^3 t) \cos 10^7 t,$$

since the bandwidth of the input-tuned circuit ($BW_{RF} = 1/R_1 C_1 = 5 \times 10^5$ rad/sec) is sufficient to pass i_{RF} undistorted.

Now with the aid of Eq. (7.2-13), we obtain the IF component of collector

current in the form

$$i_{IF}(t) = -\alpha v_B(t) G_c = -\alpha v_B(t) g_{in} I_1(x)/I_0(x).$$

This IF component of collector current is extracted by the output-tuned circuit ($\omega_{IF} = 1/\sqrt{L_5 C_5} = 5 \times 10^6$ rad/sec, $BW_{IF} = 2.5 \times 10^6$ rad/sec) to yield

$$v_o(t) = (-10 \text{ V}) - R_5 i_{IF}(t),$$

which with $x = 10$ reduces to

$$v_o(t) = (-10 \text{ V}) - (286 \text{ mV})(1 + \cos 10^3 t) \cos 5 \times 10^6 t.$$

To check the assumptions made in obtaining $v_o(t)$, we first obtain $|Z_B(j\omega_0)|$, where $Z_B(p)$ is the impedance of the input circuit "seen" at the base of the transistor. At ω_0, R_1 may be neglected compared with L_1 and C_1 to yield

$$|Z_B(j\omega_0)| = \left| \omega_0 L_2 (1 - k_{12}^2) - \omega_0 \left(\frac{L_2}{M_{12}} \right)^2 \left[\frac{L_1}{(\omega_0/\omega_{RF})^2 - 1} \right] \right|$$

$$\approx 48 \ \Omega,$$

which is indeed negligible compared with $(1 + \beta)/g_{in} = 5.2$ kΩ, which is the transistor impedance at its base terminal. Hence our calculations of the oscillation frequency and amplitude neglecting the input circuit is justified.

With the techniques of Section 5.5, the equivalent "linear loading" of the transistor on the input-tuned circuit is readily shown to be

$$G_{NL} = \frac{I_{RF}}{V_{RF}} = g_{in}/(1 + \beta);$$

G_{NL} may be reflected to the input of the RF-tuned circuit as $G_{NL}(M_{12}/L_1)^2 = 1/208$ kΩ; hence as a first approximation the transistor loading may be neglected. Actually the transistor in this example does decrease the impedance of the input-tuned circuit by 5%, and thus the IF component of $v_o(t)$ will be 5% below the value previously calculated.