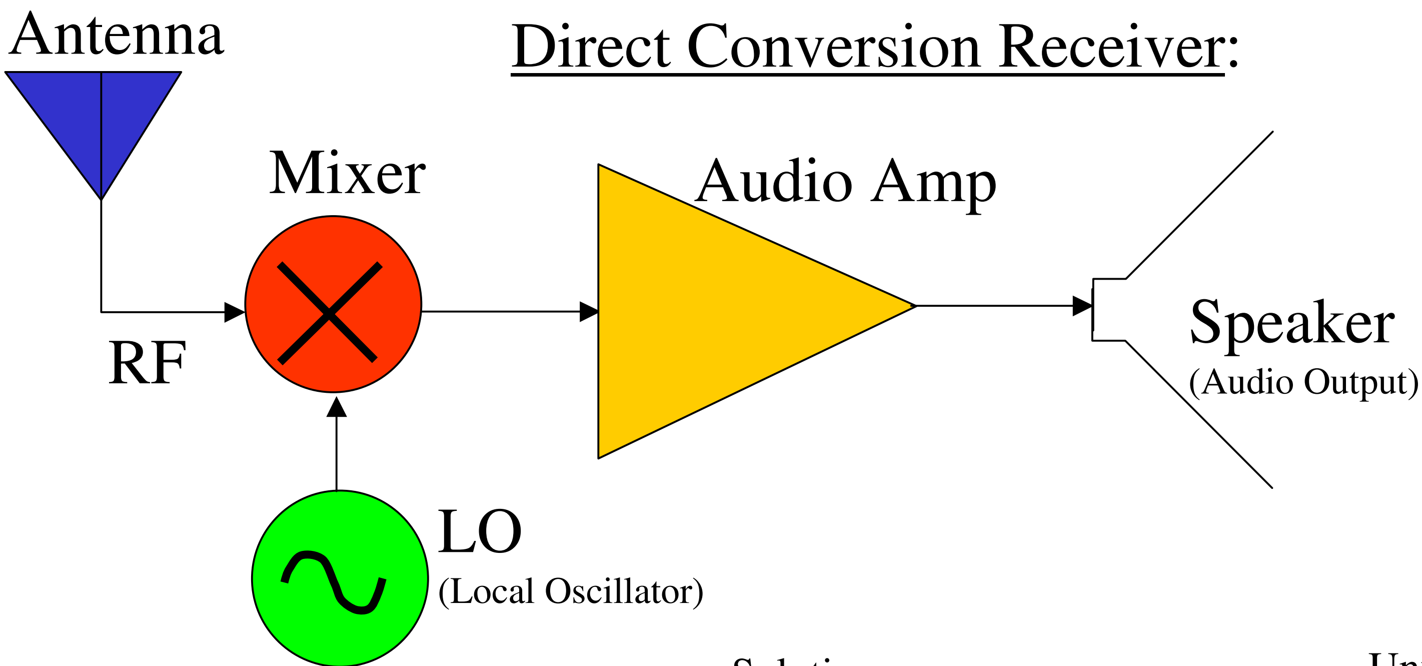
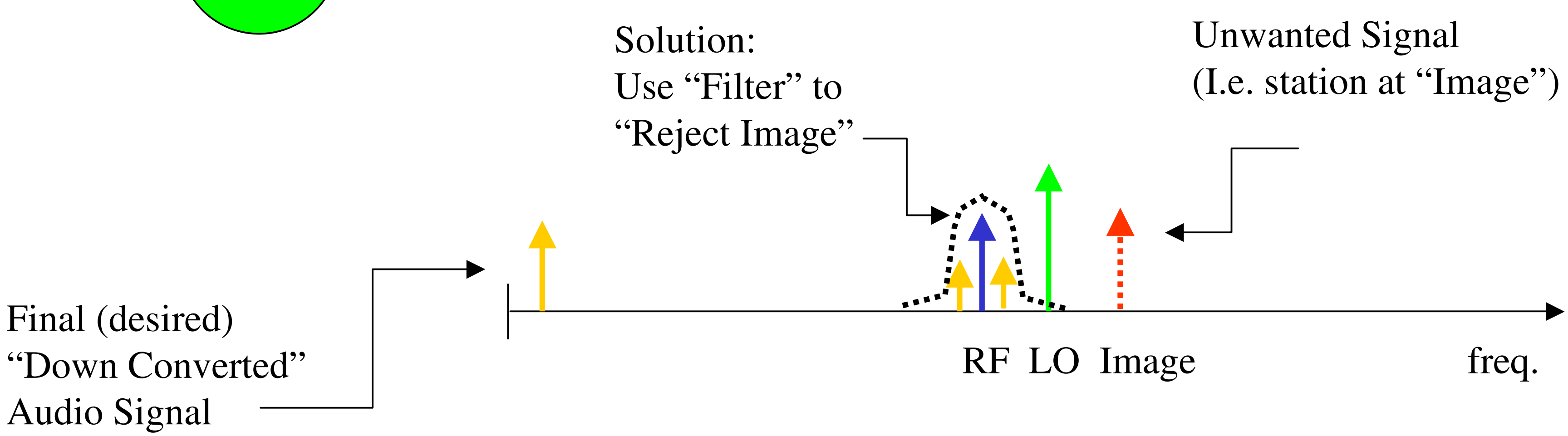
Brief Preview of RF Systems (and architectures)

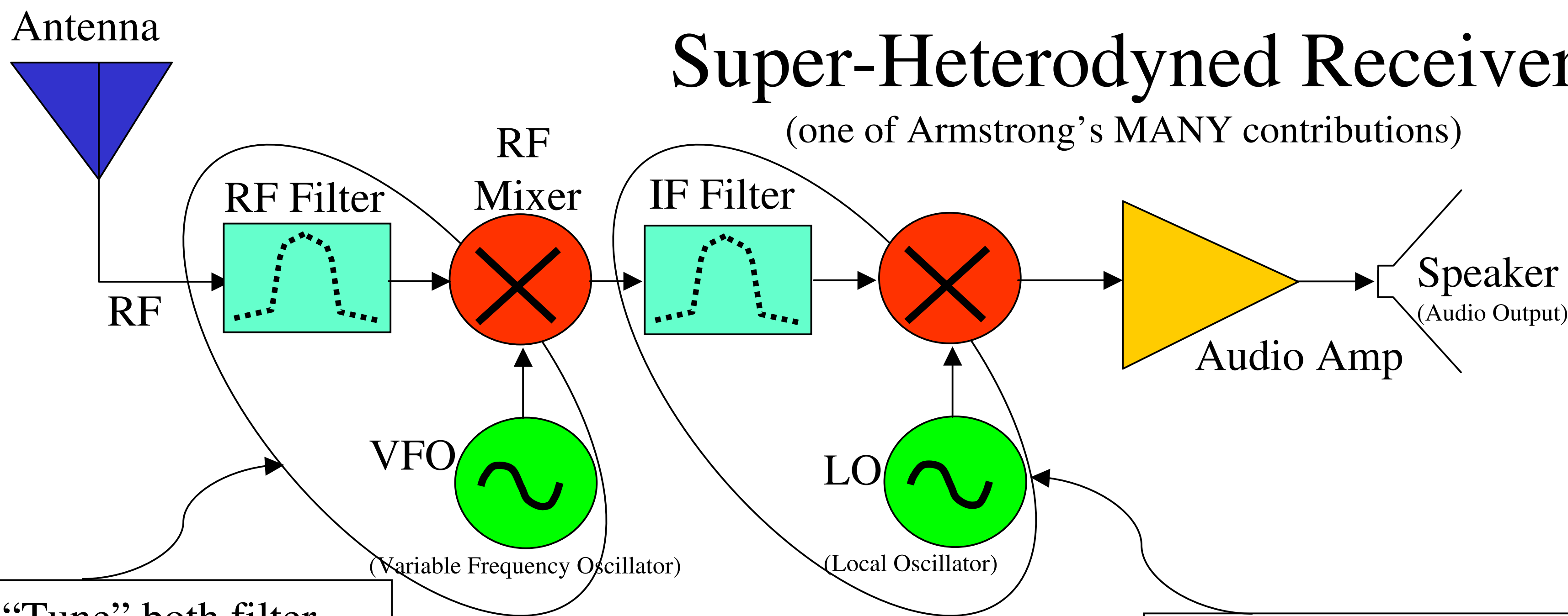
Direct Conversion Receiver:





Super-Heterodyned Receiver

(one of Armstrong's MANY contributions)



"Tune" both filter and VFO at the same time…giving output of mixer ALWAYS at **I**ntermediate **F**req.

IF at *FIXED* Freq. and therefore LO Freq. is also fixed in taking output of IF down to "basedband"

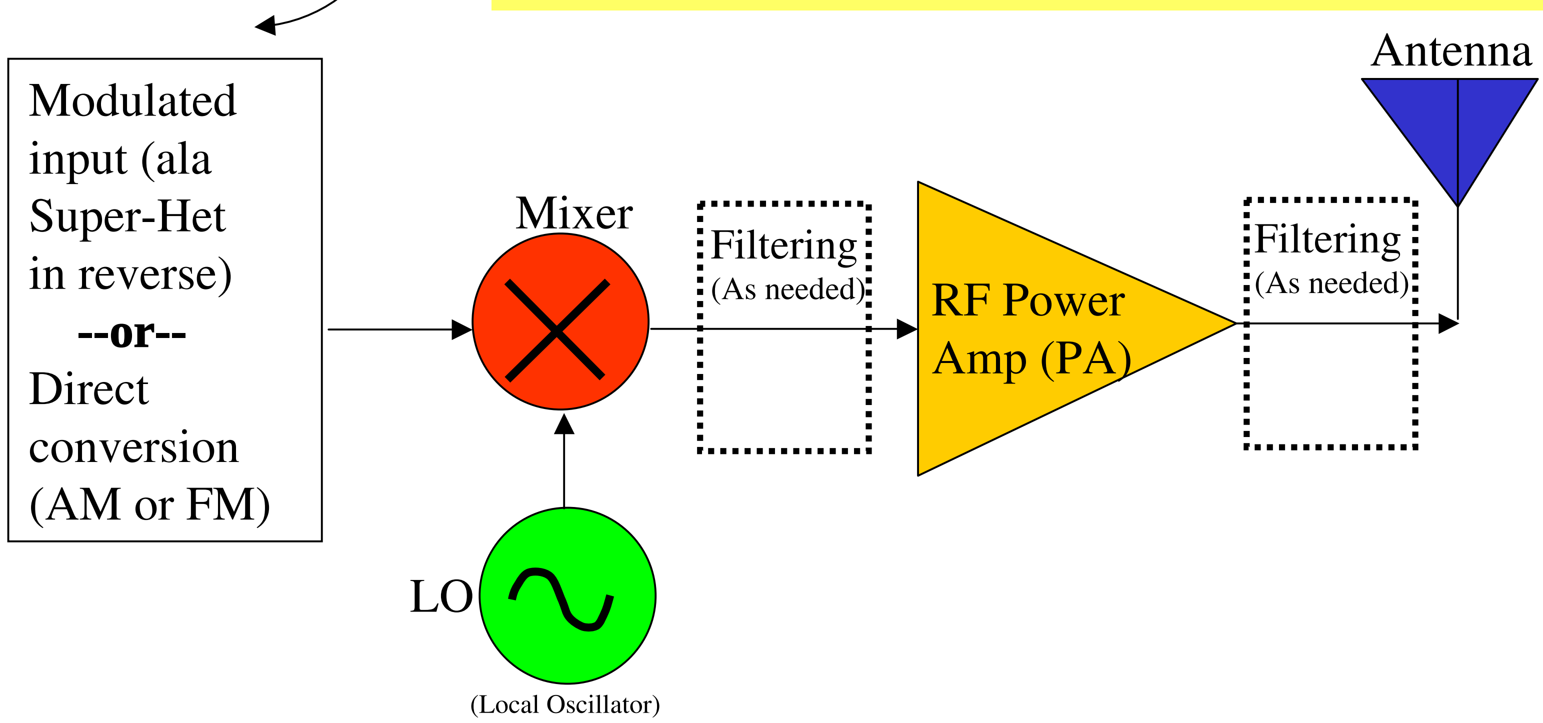
Advantages of Super-Het:

- Tuning/Image Rej. Decoupled from detection
- IF filtering can be optimized and also reused (I.e. both AM and FM bands)
- MUCH easier to work at IF vs. directly at RF (especially at high frequencies)

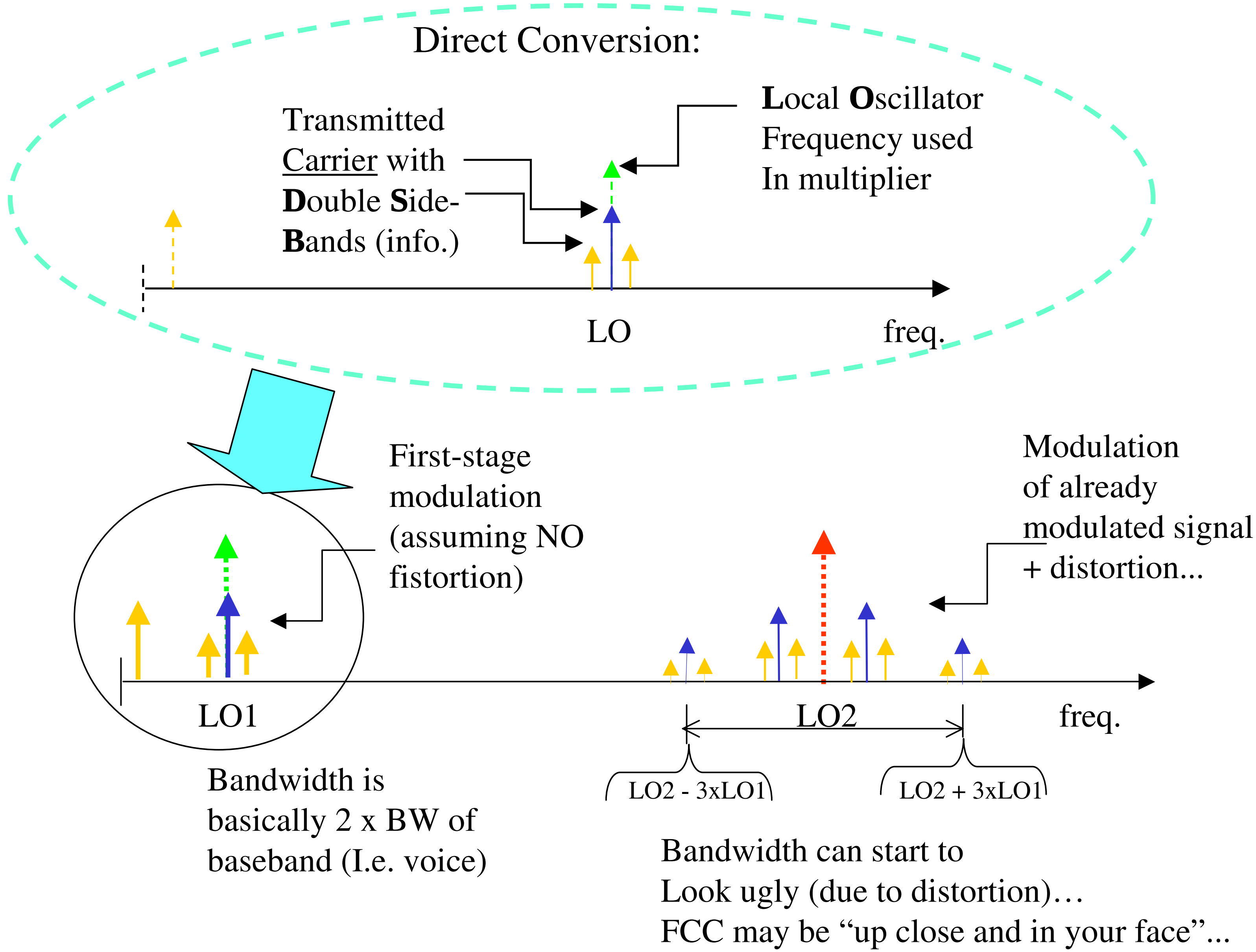
Transmitter Side Issues...

There are LOTS of choices and combinations here:

- Baseband-up-to-10's MHz can be done with DSP!
- AM is really easy (it's below 2MHz anyway)
- FM is also easy…BUT, frequency stability is not (so easy)
- Impedance matching and filtering at RF poses challenges as well...



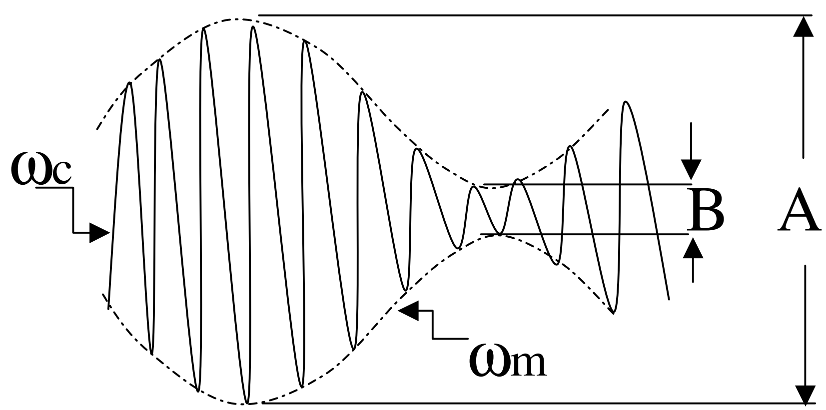
A Quick Look at BW...(including preview of FCC perspective)



Amplitude Modulation (AM) & Circuits

AM Modulation

$$v(t) = \overbrace{\left[v_c + v_m \cos\left(\underbrace{2\pi f_m}_{\omega_m} t\right)\right]}^{v_c(1+m\cos\omega_m t)} \cos \underbrace{2\pi f_c}_{\omega_c} t$$

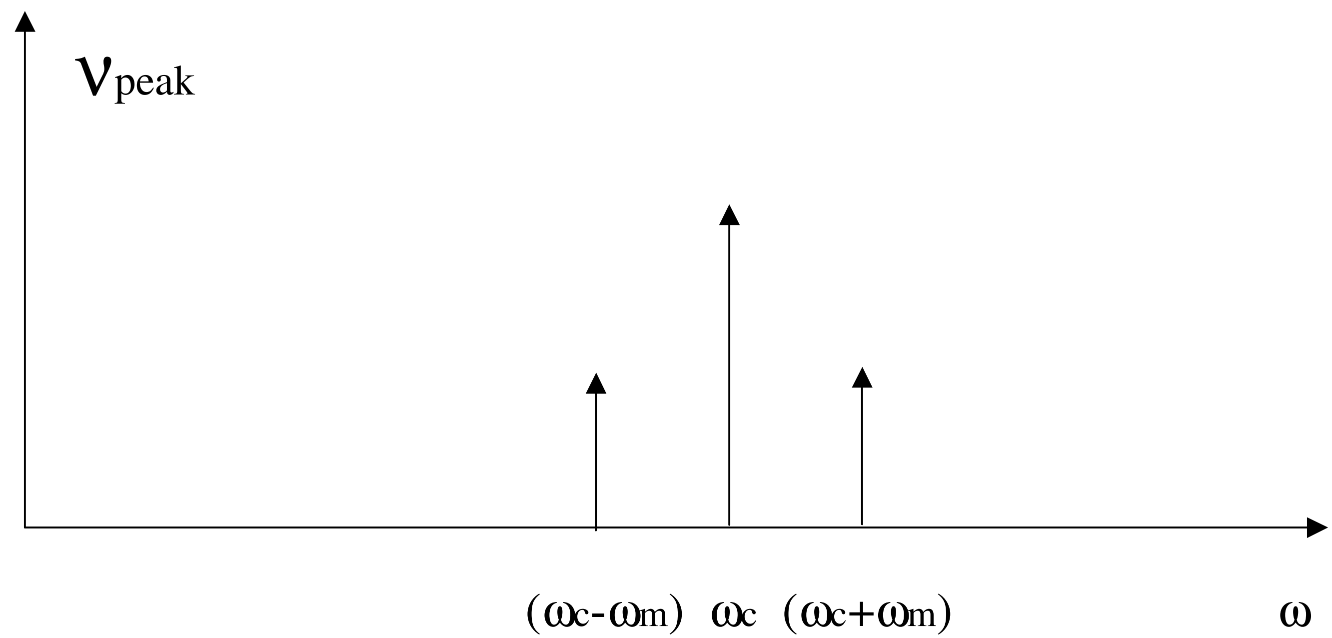


$$m \equiv \frac{v_m}{v_c} = \frac{A-B}{A+B}$$

$$v_{(t)} = v_c \cos\omega_c t + v_m \cos\omega_m t \cdot \cos\omega_c t$$

Using Law of Cosines

$$v(t) = v_c \cos\omega_c t + \frac{v_m}{2}\cos(\omega_c - \omega_m)t + \frac{v_m}{2}\cos(\omega_c + \omega_m)t$$



This is the ideal situation...

How do we get modulation in reality and what are limitations?

Circuits to realize AM:

A few simple cautions (about AM) ---

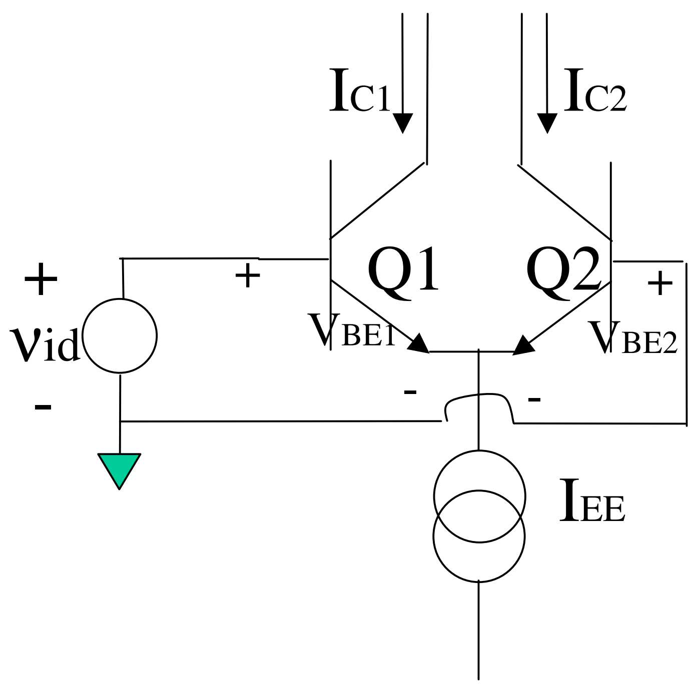
a) $m \leq 1$ otherwise information is lost

b) law of Cosine => ideal multiplication…

c) lets start here from circuits point of view

For high frequencies and at impedance levels often used, differential circuits with matched impedance lines are very convenient…

Hence, lets revisit the Emitter Couples Pair (ECP) from 113



Basic BJT "Laws":

$$I_c = I_s e^{V_{BE}/V_T}$$

$$V_T = kT/q$$

$$V_{BE} = \frac{kT}{q}\ln\left(I_c/I_s\right)$$

$$I_{c1} = I_s e^{V_{BEI}/V_T}; \quad I_{c2} = I_s e^{V_{BE2}/V_T}$$

$$v_{id} = V_{BEI} - V_{BE2} = V_T \ln\left(\frac{I_{c1}}{I_s}\right) - V_T \ln\left(\frac{I_{c2}}{I_s}\right)$$

$$= V_T \ln\left(I_{c1}/I_{c2} \right)$$

$$I_{c1} = I_{c2} e^{\left(v_{id}/V_T \right)} \quad or \quad I_{c2} = I_{c1} e^{\left(-v_{id}/V_T \right)}$$

$$I_{c1} + I_{c2} = I_{EE}$$

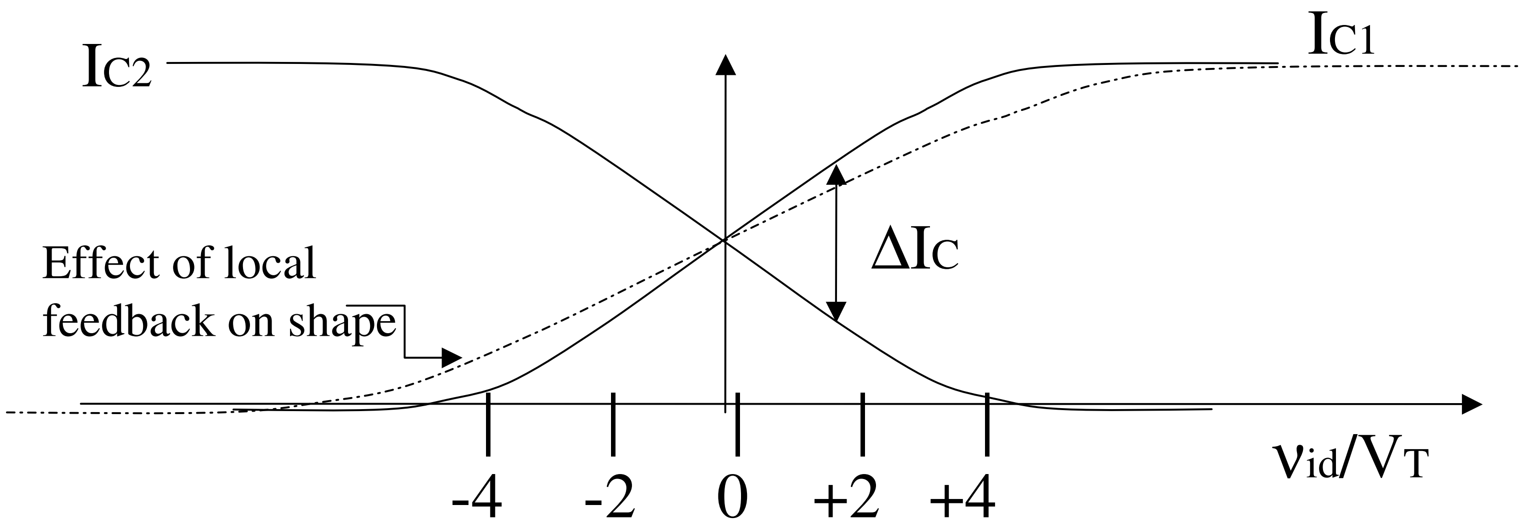
Now, for a bit of hacking to get the transfer function.

$$I_{c1} = I_{EE} - I_{c2} = \frac{I_{EE}}{1 + e^{-vid/V_T}}$$

$$I_{c2} = I_{EE} - I_{c1} = \frac{I_{EE}}{1 + e^{+vid/V_T}}$$

$$\Delta I_c = I_{c1} - I_{c2} = I_{EE}\left(\frac{1}{1 + e^{-vid/V_T}} - \frac{1}{1 + e^{+vid/V_T}}\right)$$

$$= I_{EE} \tanh\left({vid}/{2V_T} \right)$$



Comments: a) Hyperbolic Tan function gives symmetry about v_{id}=0, b) "linear" for about $1V_T$ (then "distortion" as we'll discuss below), c) local feedback can stretch linearity if desired.

<u>Small Signal Regime</u>

$$\tanh {}^{v_{id}}\!/_{2V_T} \approx \frac{v_{id}}{2V_T}$$

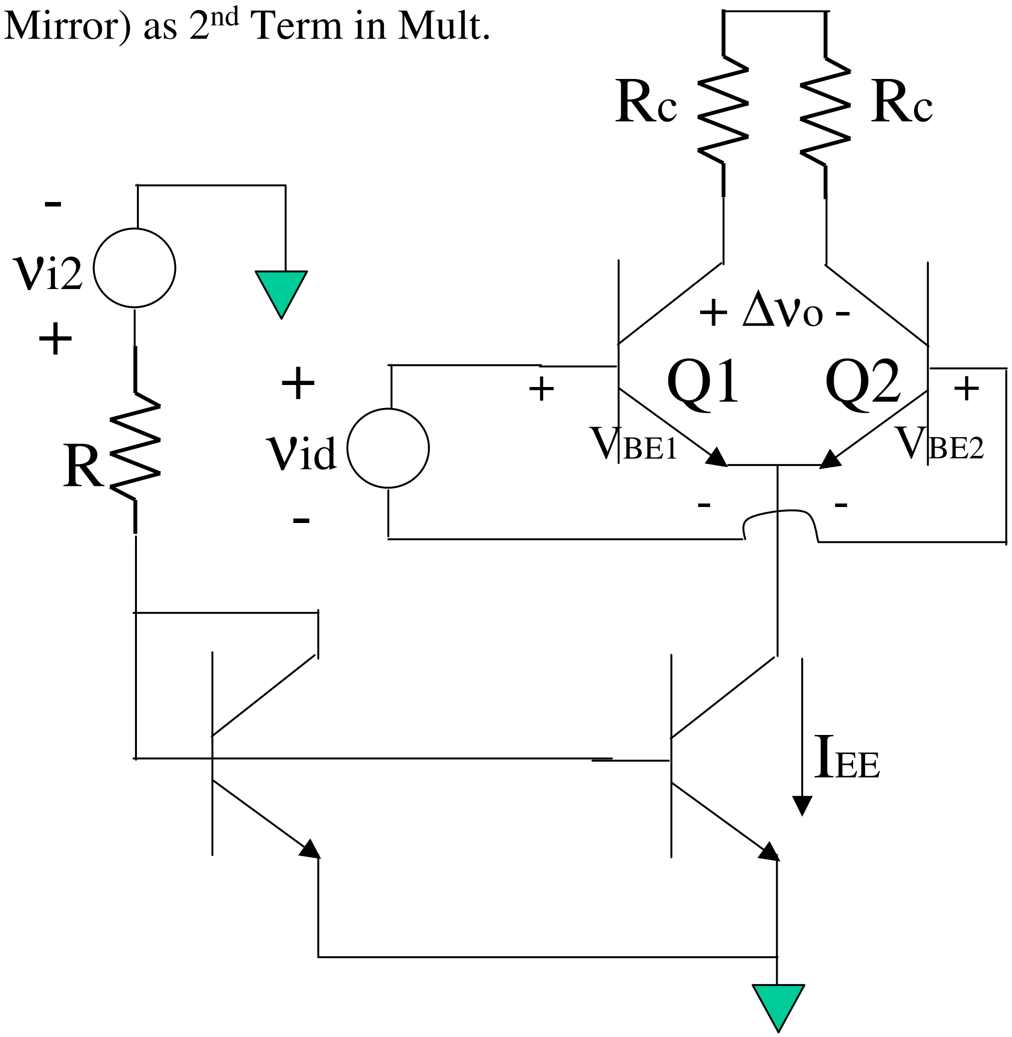
$$\Delta I_c \approx I_{EE}\frac{v_{id}}{2V_T} \quad and \quad I_{c1} = I_{c2} = {}^{I_{EE}}\!/_{2} = I_c$$

$$gm \equiv {}^{I_c}\!/_{V_T}$$

$$then \qquad \Delta I_{c1} = g_m v_{id}$$

But, let's not stop here…   (i.e. what we've done in 113)

I_{EE} (Via Current Mirror) as 2^{nd} Term in Mult.



$$I_{EE} = \frac{v_{i2} - V_{BE}}{R} \quad assume \qquad v_{i2} = v_{i2}(dc) + v_{i2}(ac)$$

$$= \frac{v_{i2}(dc) - V_{BE}}{R} + \frac{v_{i2}(ac)}{R}\cos(\omega_m t)$$

$$I_{EE} = \frac{v_c}{R} + \frac{v_m}{R}\cos(\omega_m t) \text{ where } v_c \equiv v_{i2}(dc) - V_{BE}$$

$$v_m \equiv v_{i2}(ac)$$

Now, using this expression for I_{EE} back in ΔI_c expression

$$\Delta I_c = I_{EE}\tanh\left(v_{id} \big/ 2V_T \right) = \frac{1}{R}\left(v_c + v_m\cos\omega_m t\right)\tanh\left(\frac{v_{id}}{2V_T}\right)$$

Adding Rc's (see above figure) and looking at $\Delta \upsilon_o$

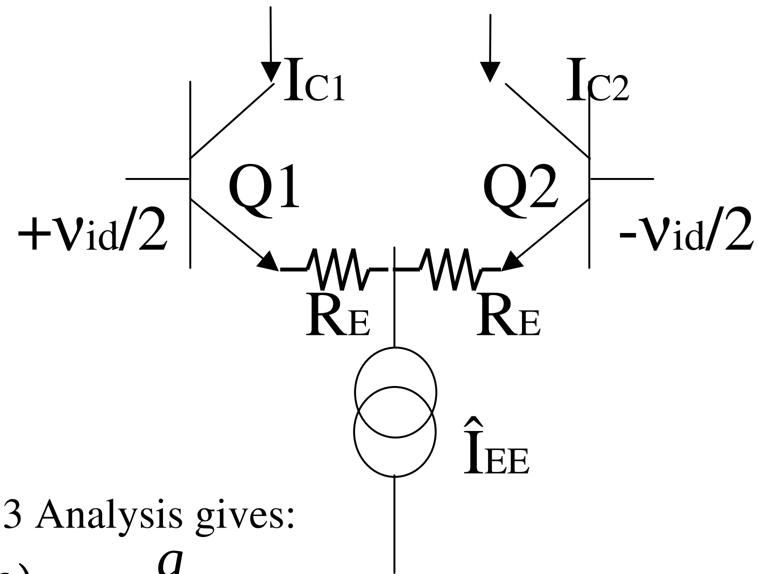
$$\Delta v_o = -\Delta I_c R_c = -\frac{R_c}{R}\left(v_c + v_m\cos\omega_m t\right)\tanh\left(\frac{v_{id}}{2V_T}\right)$$

Again, using <u>small signal regime</u> ($v_{id} << V_T$)

$$\Delta v_o = \frac{-R_c}{2V_T R}\left(v_c + v_m\cos\omega_m t\right)\left(\bar{v}_{id}\cos\omega_c t\right)$$

$$note : v_{id} \equiv \bar{v}_{id}\cos\omega_c t$$

Although tanh formulation was used, with small signal approx. ($v_{id} << V_T$), one can <u>slightly</u> improve dynamic range by local feed back with R_E's on the diff pair.



Limitation:

Basic EE113 Analysis gives:

$$g_m(eff) = \frac{g_m}{1 + g_m R_E}$$

$$\text{If } g_m R_E >> 1, \text{ then:}$$

$$g_m(eff) \simeq \frac{1}{R_E}$$

$$\therefore \Delta I_{c1} \simeq \frac{\frac{v_{id}}{2}}{R_E} \quad \Delta I_{c2} \simeq \frac{-\frac{v_{id}}{2}}{R_E}$$

$$\Delta I_c = \Delta I_{c1} - \Delta I_{c2} \simeq \frac{v_{id}}{R_E}$$

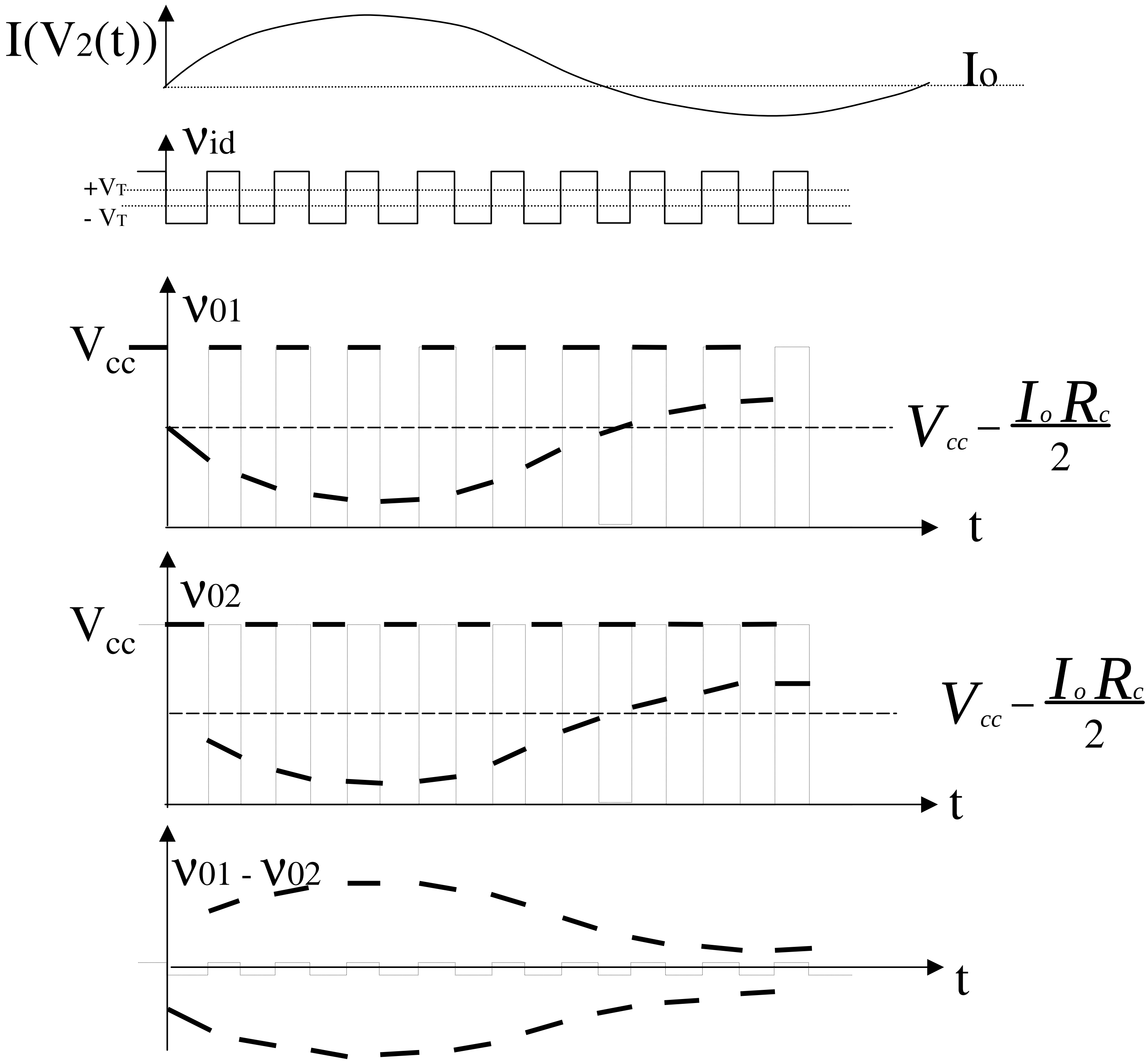
But! There's no I_{EE} left (ala $g_m = I_{EE}/(2V_T)$) and hence multiplication is gone which is not what we wanted. Hence, $g_m R_E \approx 1$ could increase dynamic range but not kill mult.

Looking at the other extreme, where v_{id} drives Q_1 and Q_2 into a "switching mode" behavior, we find that a quite acceptable "modulated" signal is realized. Consider voltage waveforms, rather than ΔI_c:

$$v_{o1} = V_{cc} - I_{c1}(t) R_c$$

$$v_{o2} = V_{cc} - I_{c2}(t) R_c$$

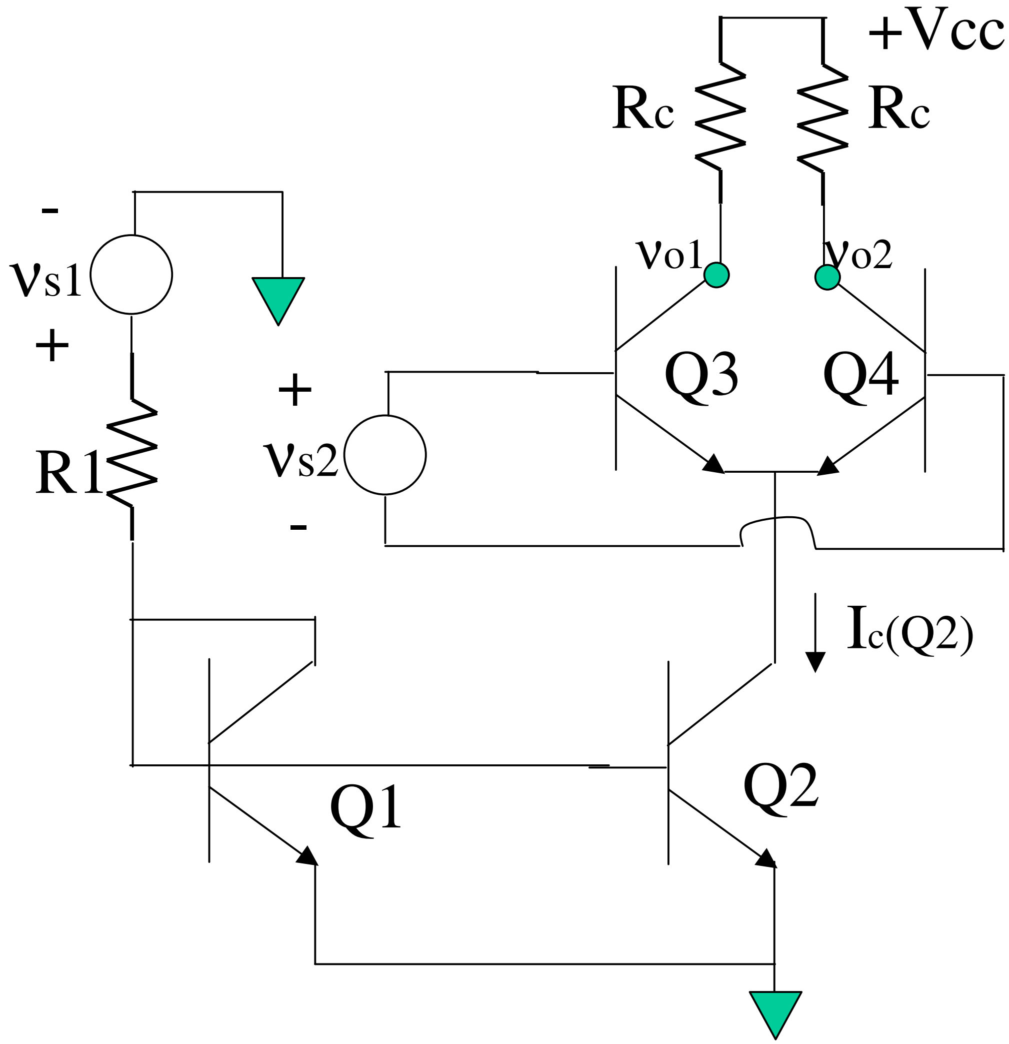
The following set of figures show how the "tail current" $I(V_2(t))$ gets switched between collectors C1 and C2 which gave the step-wise jumps downward for V_{cc}



I(V2(t))
Io
vid
+VT
- VT
v01
Vcc
$V_{cc} - \frac{I_o R_c}{2}$
t
v02
Vcc
$V_{cc} - \frac{I_o R_c}{2}$
t
v01 - v02
t

A Quick Review (EE113 Final Exam 1991)

Consider the circuit shown below where Q1 and Q2 provide both biasing and an ac signal current to the differential pair. The transistor pair Q3 - Q4 provides a differential stage with a separate ac signal input (the circuit has two different ac signal inputs).



Assume that all V_{BE}'s for biasing are the same and given the following input to the current amplifier (mirror):

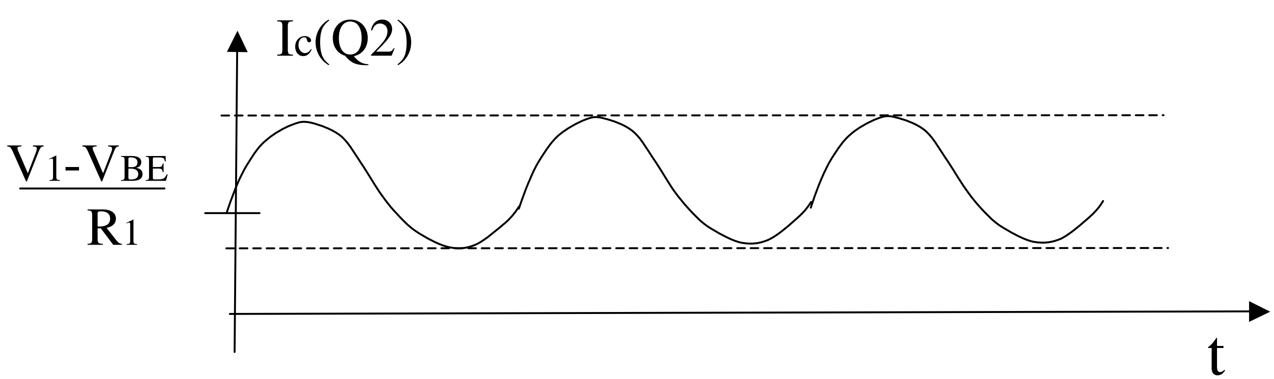
$$V_{s1} = V_1 + \frac{V_1}{2} \sin \omega_1 t$$

(Q) Write an expression for $I_c(Q2)$ and sketch its value as a function of time.

$$I_c(Q1) = \frac{V_{s1} - V_{BE}}{R_1}$$

Current mirror gives $I_c(Q1)=I_c(Q2)$

$$I_c(Q2) = \frac{V_1 - V_{BE}}{R_1} + \frac{V_1 - V_{BE}}{2R_1} \sin \omega_1 t$$



(Q) Initially assuming that $V_{s2} = 0$ (inputs grounded), find the dc and ac signals at V_{o1} and V_{o2}.

$$I_c(Q3) = I_c(Q4) = \frac{\alpha I_c(Q2)}{2}$$

$$V_{o1} = V_{o2} = V_{cc} - \frac{\alpha I_c(Q2) R_c}{2}$$

$$= \underbrace{V_{cc} - \frac{\alpha R_c}{2R_1}(V_1 - V_{BE})}_{dc} - \underbrace{\frac{\alpha R_c}{4R_1}(V_1 - V_{BE})\sin\omega_1 t}_{ac}$$

(Q) Now apply the following small signal input to the differential pair:

$$V_{s2} = V_2 \sin\omega_2 t$$

Write an expression for the ac signal V_{o1}. HINT: Since the bias current is not constant, instead of using a constant g_m, use

$$g_m = \frac{kT}{q} I_c(t)$$

$$v_{o1} = -\frac{gmR_c}{2} v_{s2}$$

$$= -\frac{kT}{2q} R_c \left[\underbrace{\frac{\alpha(V_1 - V_{BE})}{2R_1}}_{I_{c3}} + \underbrace{\frac{\alpha(V_1 - V_{BE})}{4R_1}}_{I_{c3}/2} \sin\omega_1 t \right] V_2 \sin\omega_2 t$$

$$V_{o1} = \frac{-kT}{2q} I_{c3} R_c \left[1 + \frac{1}{2}\sin\omega_1 t \right] v_2 \sin\omega_2 t$$

Assuming $\omega_2 = \frac{\omega_1}{10}$ sketch the Fourier Spectrum and suggest what such a circuit might be useful for. (this is for you to do now…)