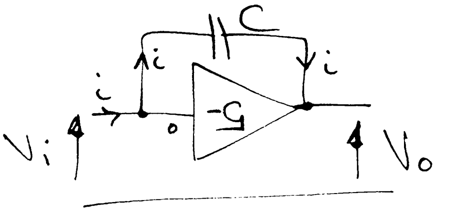
1(a) Consider a general amplifier with capacitance between input and output



$$V_o = -G V_i$$

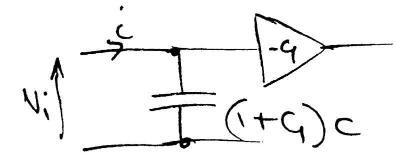
Sum currents at input node:-

$$i = \frac{V_i - V_o}{1/j\omega C} \quad \text{where } V_o = -G V_i$$

$$\therefore \quad i = j\omega C (1+G) V_i = \frac{V_i}{\frac{1}{j\omega(1+G)C}}$$

$\therefore$ equivalent input capacitance to ground $= (1+G) \times C$

$\equiv$



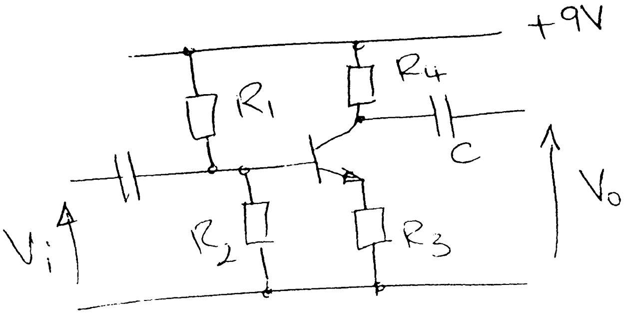
ie: much increased input capacitance

Hence, if the input source has significant impedance, the system acts as a low pass filter. So the

$$f_{-3dB} = \frac{1}{2\pi (1+G) C . R}$$

Miller effect can limit the frequency response of a circuit.

1b). Basic single transistor amplifier



$R_{out} = R_4$

$\text{Gain} = \dfrac{-R_4}{R_3 + r_e}$

$r_e = V_t / I_c$

$V_t = kT/q \simeq 0.025V$

$R_4 = 100\Omega$ for reqd. output impedance

For power gain = 6dB = x4 power ≈ x2 voltage
(i/p and o/p impedances similar)

So select $R_3 = 39\Omega$ say to give >2 x gain.

If $V_{o\,d.c.}$ is +4.5V, then $I_c = 45mA$ and $r_e = 0.56\Omega$

$\therefore$ net gain $= \dfrac{100}{39.6} = 2.51$ (≡ 8 dB)

$V_E = 45 \times 10^{-3} \times 39 = 1.76V$, $V_{BE} = 0.65V$ assumed.

$\therefore V_B = 2.4V$ (base current $\sim \dfrac{45 \times 10^{-3}}{200} = 0.2mA$)

Input impedance $= R_1 || R_2 || h_{fe}(R_3 + r_e) = 75\Omega$ target
($h_{fe}(R_3 + r_e)$ ~ 8kΩ ∴ ignore)

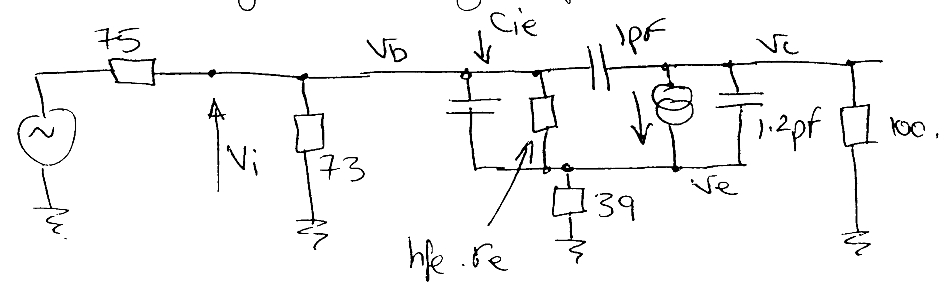
and $2.4 = 9 \cdot \dfrac{R_2}{R_1 + R_2}$ to set V_B

So choosing $R_2 = 120\Omega$, $R_1 = 330\Omega$ ∴ $R_{in} = 88\Omega$
or try $= 100\Omega$ $= 270\Omega$ $= 73\Omega$ ✓ o.k.

C's must be small impedance @ 600MHz, say <1Ω ⇒ $C > \dfrac{1}{2\pi . 600 \times 10^6}$ = 265pF.

∴ 1nF ok.

1(c) Small signal model of input :-



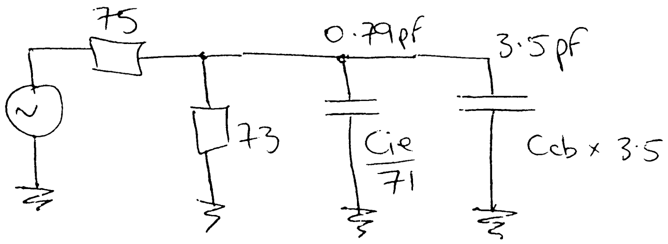
$$C_{ie} = \frac{1}{2\pi r_e f_t} = 56.8\ pF$$

$r_e = 0.56$, $f_t = 5\times10^9$

To refer these to gnd.
$G = -2.5$, so the Miller effect on C_{cb} = $\times 3.5$

Also, as $V_e = \frac{R_3}{(r_e + R_3)} \cdot V_i = 0.986 V_i$ then b-e impedances are multiplied by $\frac{1}{(1-0.986)} = \times 71$

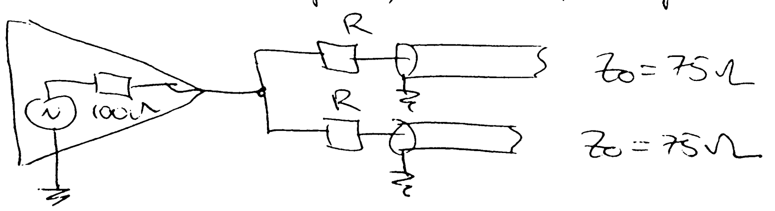
So, input circuit becomes :-



$\therefore$ $C' = 4.3\ pF$
$R' = 37\ \Omega$

So $f_{-3dB} = \frac{1}{2\pi R'C'} = 1.0\ GHz$ ✓ o.k. for 600MHz.

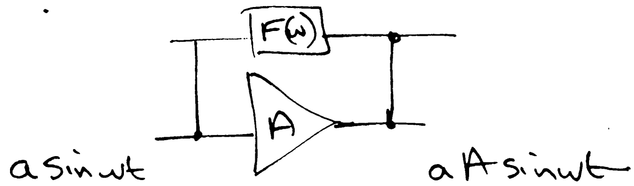
d) Amplifier has 100Ω o/p impedance. To provide 2 identical matched outputs, use a pair of resistors :-



For each line to see 75Ω $[(75+R)\|100] + R = 75$

$\therefore R = 25\Omega$

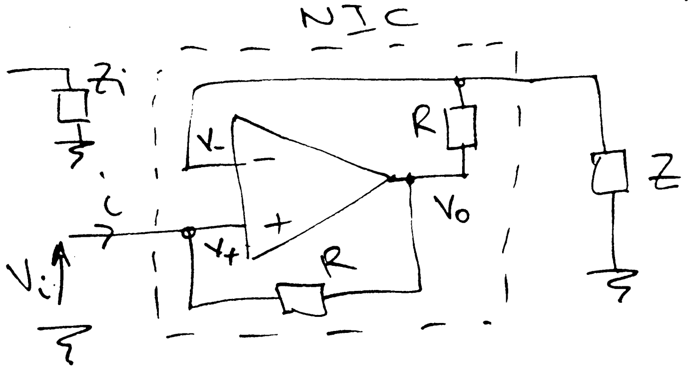
2(a).



- Phase shift around loop = 0
- Loop gain = 1

(b). Consider a negative impedance converter. (NIC):

≡



$$i = \frac{V_i - V_o}{R} \quad \text{———(1)}$$

$$\frac{V_o - V_-}{R} = \frac{V_-}{Z} \quad \text{———(2)}$$

$V_+ = V_i = V_-$ for ideal op-amp

$\therefore \; V_o = V_i\,(1 + R/Z)$ from (2)

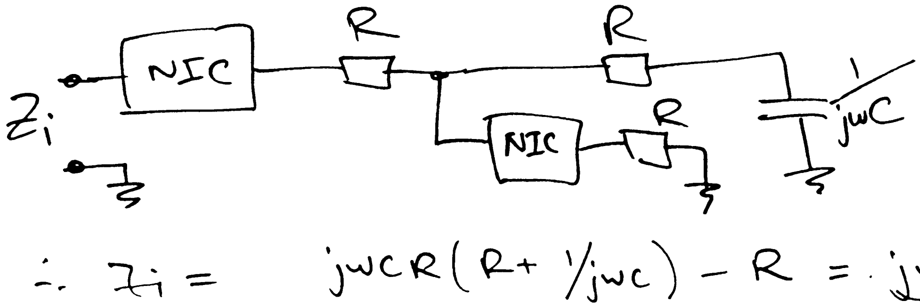
and $\therefore \; i = \dfrac{V_i - V_i - V_i R/Z}{R}$ sub into (1)

$$\therefore \; i = -\frac{V_i}{Z}$$

$$\therefore \; Z_i = -Z$$

Chaining these together:

$$Z_i = -\left[\frac{-R\,(R + 1/j\omega C)}{1/j\omega C} + R\right]$$



$\therefore \; Z_i = j\omega CR\,(R + 1/j\omega C) - R = j\omega CR^2$

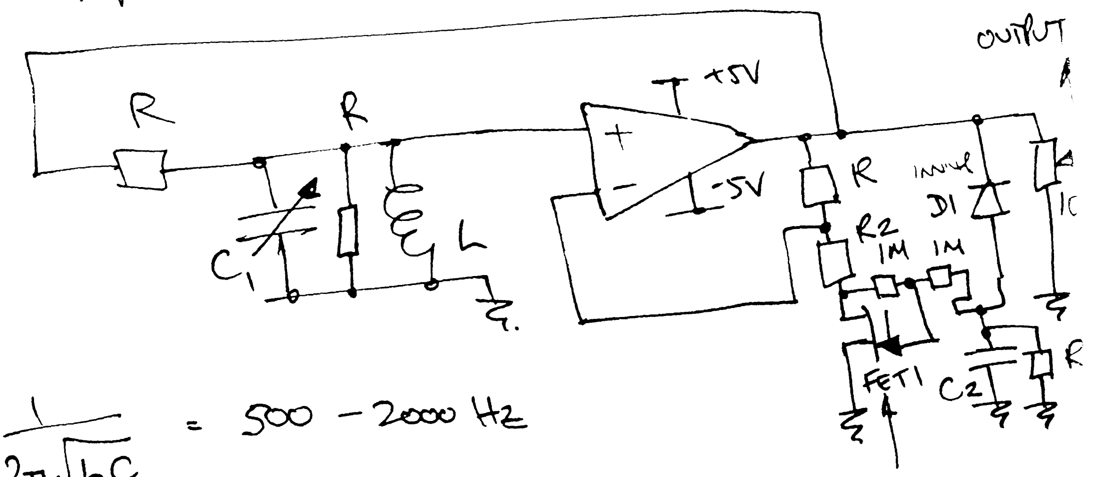
≡ to inductor $L = CR^2$

2 (b) cont.d.

- Synthesised inductors only work where the op-amp behaves as such so, there are current, frequency, voltage and slew-rate limits above which the device model fails.

- synthesised inductors can be more compact, low cost and less prone to interference and can be easily varied (change R or C). and do not suffer non-linearity of magnetic core material.

(c)



$$f_{res} = \frac{1}{2\pi\sqrt{LC}} = 500 - 2000\ \text{Hz}$$

depletion mode FET: -ve voltage turns off.

Choose R's = 10kΩ nominal

Inductor synthesised as in part (b) with

$C = 10nF$ and $R = 10k\Omega \Rightarrow L = 1H$

$\therefore C_1 = 101\ nF$ (max.) to $6.3\ nF$ (min.)

At resonance, C_1 and L impedances cancel, therefore a gain of 2 is required in the op-amp for an overall loop gain of unity. Hence choose $R_2 = 6.8k\Omega$ say. then for start-up the loop gain > 1 (as FET has low resistance channel).
Amplitude control is set by the demodulator D1, C2, R controlling the FET gate. Larger oscillations increase the FET impedance

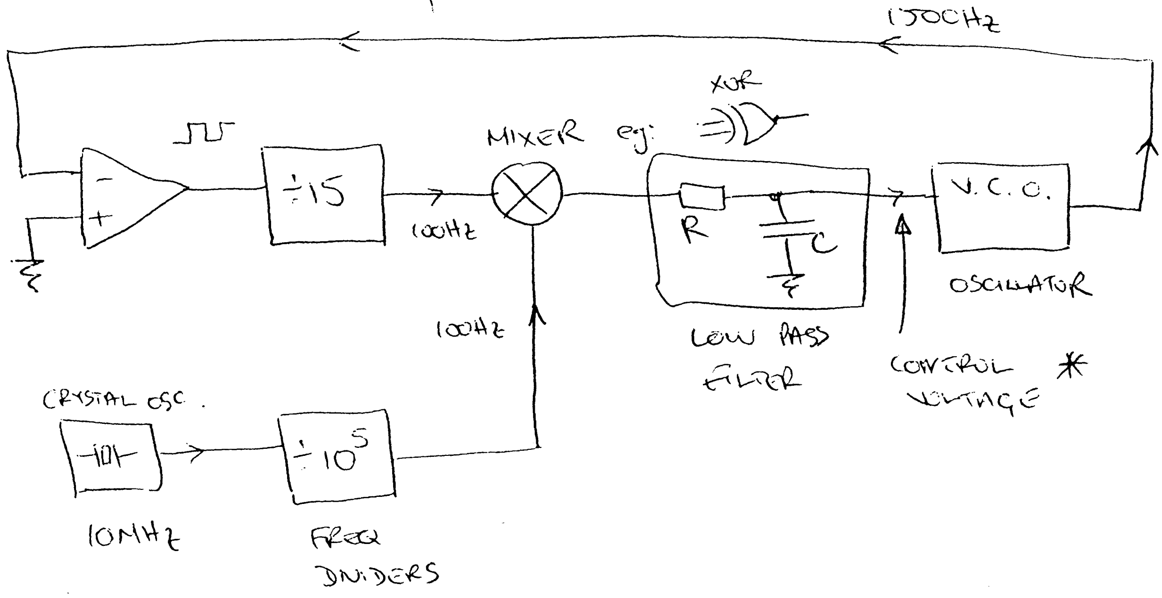
2(c) Contd.

and the op-amp gain falls. It will settle at several V amplitude. The output potentiometer sets the final output amplitude. The C_2R time constant must be long for low sine-wave distortion eg: 1s ∴ $C_2 = 100\mu F$.

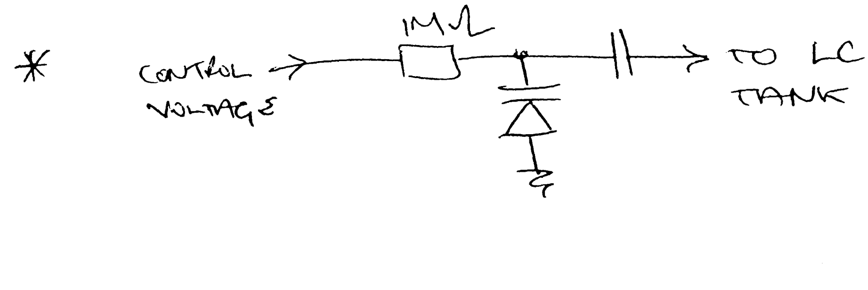
The 2 × 1MΩ gate resistors reduce the FET distortion (by adding $V_{ds}/2$ to the gate voltage to cancel the non-linear term).

d) To accurately lock the oscillator frequency to 1.5kHz, with reference to a 10MHz crystal oscillator - we need to divide both down to the same frequency eg: oscillator ÷ 15 = 100Hz
reference ÷ 10^5 = 100Hz.

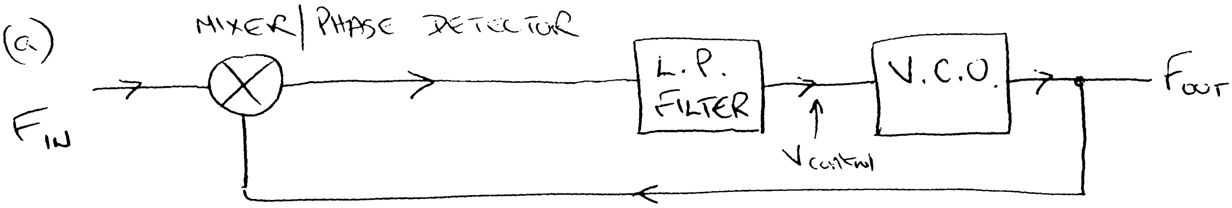
Use a phase-locked-loop (PLL) :-



To make oscillator tuneable, put varactor across the LC tank:-



3(a)

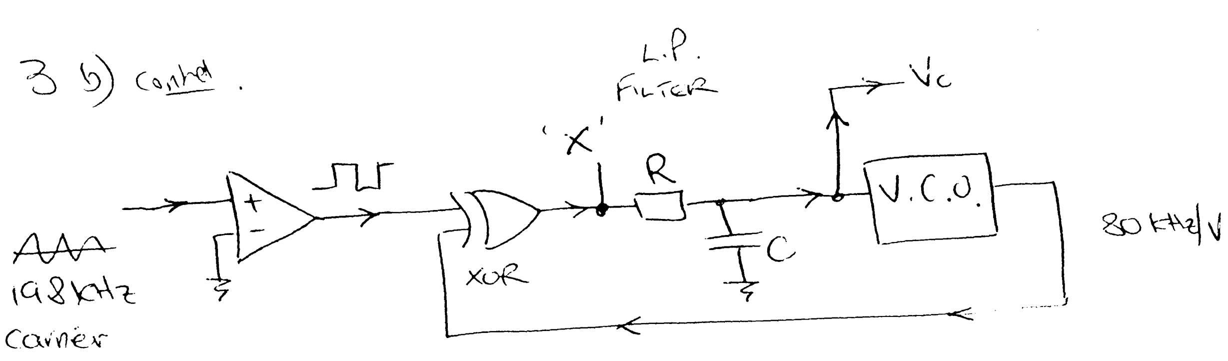


- A phase detector compares the phase/frequency of the input with the voltage controlled oscillator output
- If the freq's are different, the phase detector outputs the sum & difference frequencies. The (lower) diff. freq. passes through the low pass filter and this modulates the V.C.O. output - putting sidebands onto its spectrum
- one of the sidebands is equal in freq. to the input, Fin
- Hence the phase detector produces a d.c. component pushing the V.C.O. to match Fin ie: the loop locks.
- Once locked, the V.C.O. follows Fin, but has a 'flywheel' effect - not following fast changes or drop-outs in Fin. The loop 'inertia' and stability are set by the loop gain and L.P. filter bandwidth.

(b)

In order to extract the phase-keyed data from the radio carrier, we shall use a PLL to create a stable version of the carrier and look for the phase shifts with respect to this. We can work with square wave signals ∴ use an XOR gate as the phase detector.

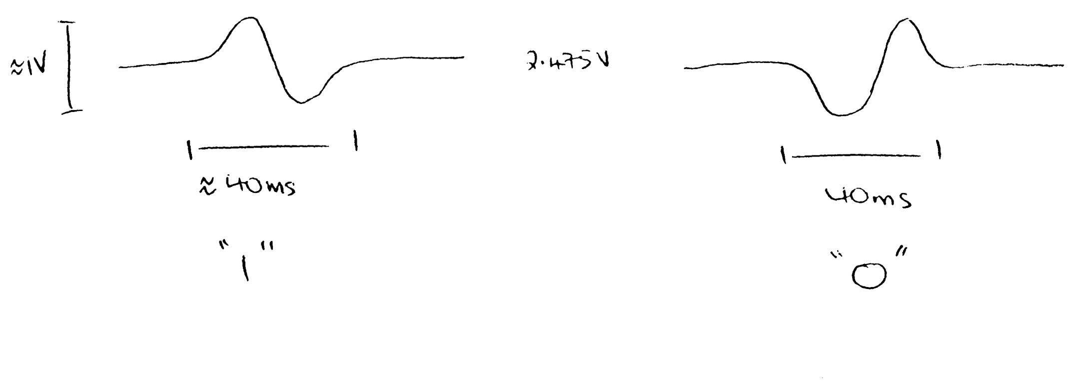
3 b) contd.



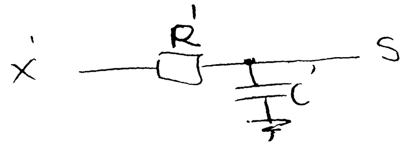
- In the steady state, $V_c = \frac{198}{80} = 2.475$ V d.c.
- For an XOR with 5V supply, the phase sensitivity = $5/_{180}$ V deg^{-1} hence ± 22.5° gives ± 0.63 V swing in the d.c. component at 'X'.
- Choose R & C such that there is very little change in Vc as the phase shift keying (PSK) occurs i.e. we want a steady signal at the XOR to compare the carrier to.
- The frequency of the PSK data is approx. $\frac{1}{2 \times 20\,ms} = 25$ Hz, so if we choose RC to give a cut-off at say 1 Hz, then the PLL flywheel effect will be significant.

  $\therefore \frac{1}{2\pi RC} = 1$ with R = 100 kΩ, C = 1.5 μF

- To extract the PSK data, we need to look at the signal at 'X'. If we filter this with a low pass filter at ~ f_{data} then, we shall see the d.c. component of the XOR output carrying the data :-



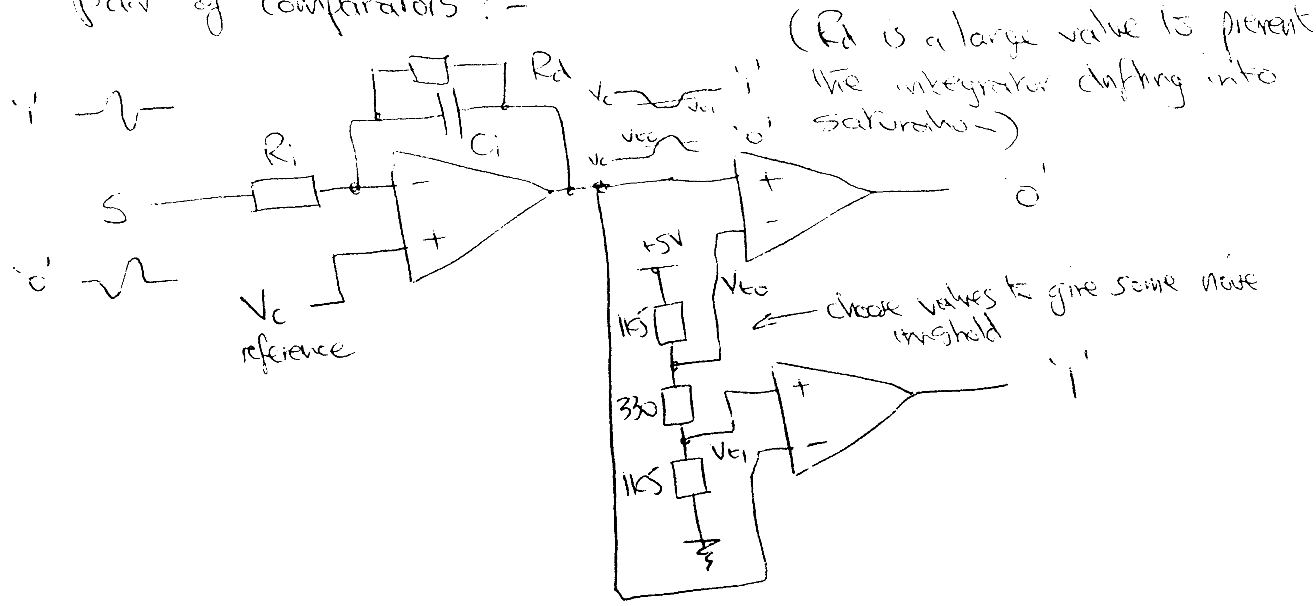
3 (b) contd.



$S_o = \frac{1}{2\pi R'C'}$ $\therefore R' = 10k\Omega$

$C' = 330nF$

- To reconstruct the original 'i's and 'o's from the L.P. filtered signal at 'x', this can be done in a number of ways using comparators, logic and/or analogue circuits. For example, we can use an integrator followed by a pair of comparators :-



- set integrator gain to give reasonable output signals, so if the input, S, swings by 0·5V for 20ms and $R_i = 10k\Omega$ say, then to get 0·5V at the integrator output:

$$i \times t = \frac{20 \times 0{\cdot}5 \times 10^{-3}}{10^4} = C_i \times \Delta V = 0{\cdot}5 \times C_i \quad \therefore \quad C_i = 2\mu F$$

(and set $R_d = 1M\Omega$ say)

- The comparator thresholds are set so that only a FSK signal causes the integrator output to cross the thresholds.

(• Note, because an integrator is a L.P. filter, we could omit R' and C' in this implementation.)

4c) contd.

$\therefore$ at 60kHz, $\delta = \sqrt{\left(\frac{2}{2\pi . 60\times10^3 . 4\pi\times10^{-7} . 3{\cdot}33\times10^7}\right)}$

$= 3{\cdot}56\times10^{-4}$ m (0·36 mm)

Since, the wire is only 0·25 mm dia., we shall assume the full section carries the current and ignore the skin effect.

$\therefore$ Resistance of 1m Cu wire, $R_0 = \frac{\rho l}{A} = \frac{\rho l}{\pi r^2}$

$\therefore R_0 = 0{\cdot}611\ \Omega$

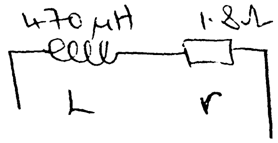
Radiation efficiency $= \left(\frac{R_e}{R_0 + R_e}\right) \times 100\%$

$= 0{\cdot}005\%$

(or less for triang. current distn. by factor of 4)

This is a low efficiency, so might need more amplifier gain - hence cost saving may not be worth it!

4(d)

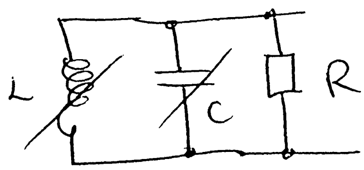


at 60kHz, $\omega L = 177\ \Omega$

$\therefore Q = \frac{\omega L}{r} = \frac{177}{1{\cdot}8} = 98{\cdot}4$

Bandwidth $= f_0/Q = \frac{60\times10^3}{98{\cdot}4} \simeq 610$ Hz

$f_0 = \frac{1}{2\pi\sqrt{LC}} = 60\times10^3$ $\therefore$ with $L = 470\ \mu H$ $\Rightarrow C = 15$ nF



For parallel resonance, with same Q

$98{\cdot}4 = \frac{R}{\omega L}$ $\therefore R = 17{\cdot}4\ k\Omega$

4 (a) 15×10^3 W @ 60 kHz

$c = f\lambda = 3\times10^8$ m/s $\therefore$ $\lambda = 5000$ m

Power density, $P = \dfrac{G \times 15\times10^3}{4\pi R^2}$ — Gain, $G = 1.5$; Range, $R = 102\times10^3$ m

$\therefore$ $P = 1.72\times10^{-7}$ W/m²

$\underline{P} = \frac{1}{2}\,\underline{E}\times\underline{H}^*$, $|P| = \frac{1}{2}H^2\eta$ where $\eta_0 = \sqrt{\mu_0/\epsilon_0} = 120\pi$

$\therefore$ $H = 3\times10^{-5}$ A/m ($E = 11$ mV/m)

(b) Antenna eqn. $G = \dfrac{4\pi A_e}{\lambda^2}$

-76 dB $\equiv 2.5\times10^{-8}$ liner $\therefore$ $A_e = 0.05$ m²

$\therefore$ power into 250Ω matched load $= 8.6$ nW $= V^2/R$

$\therefore$ $v = 1.5$ mV$_{rms}$ amplitude

(c) Radiation resistance, $R_e \simeq 80\pi^2\left(\dfrac{1}{5000}\right)^2$ Ω ← v. short $\therefore$ assume ideal dipole

$R_e = 3.16\times10^{-5}$ Ω

(or could ÷4 for triangular current distn)

Skin depth $\delta = \sqrt{2/\omega\mu\sigma}$

Taking resistivity of copper wire $\rho = 3\times10^{-8}$ Ωm

$\therefore$ $\sigma = 3.33\times10^{7}$ S/m