3B1 Crib 2023

1. (a)

Norm to 50Ω -> 5+2j (B1) Plot and read off smith chart $0.71\angle 8°$

n.b an analytic solution is also possible and equally valid.

(b)

Assume fringing fields extend by thickness of board.

$$C = \frac{(w+2d)\epsilon_o\epsilon_r}{d}$$

$$Z_0 = \sqrt{\frac{L}{C}}, v = \frac{1}{\sqrt{LC}} = \frac{c_o}{\sqrt{\epsilon_r}},$$

$$Z_0 = \frac{1}{vC} = \frac{\sqrt{\epsilon_r}}{c_0}\frac{d}{(w+2d)\epsilon_o\epsilon_r}$$

$$(w+2d)\epsilon_o\epsilon_r c_0 Z_0 = \sqrt{\epsilon_r}d$$

$$(w) = \frac{\sqrt{\epsilon_r}d}{\epsilon_o\epsilon_r c_0 Z_0} - 2d$$

W=2.7mm

c)

See Smith chart. Start at B1. Rotate around centre to unit R circle (B2) clkwise towards generator.

Track length 0.5 λ-(0.239λ+0.188λ) = 0.449λ

λ= 3×10^8/(250×10^6*sqrt(4.2) = 58.5cm

so track length is 262mm.

Required reactance is -2j × 50, -> 6.4pF

Reduce with higher ϵ_r, shunt C, or series L.

(ii) ϵ_r will change electrical length and the characteristic impedance. Z_0 is now 62.5Ω. Need to switch smith chart to this characteristic impedance.

250+100j becomes 4+1.6j (C1)

λ= 3e8/(250e6*sqrt(2.69)=73.2cm, so the line is 0.358 λ long.

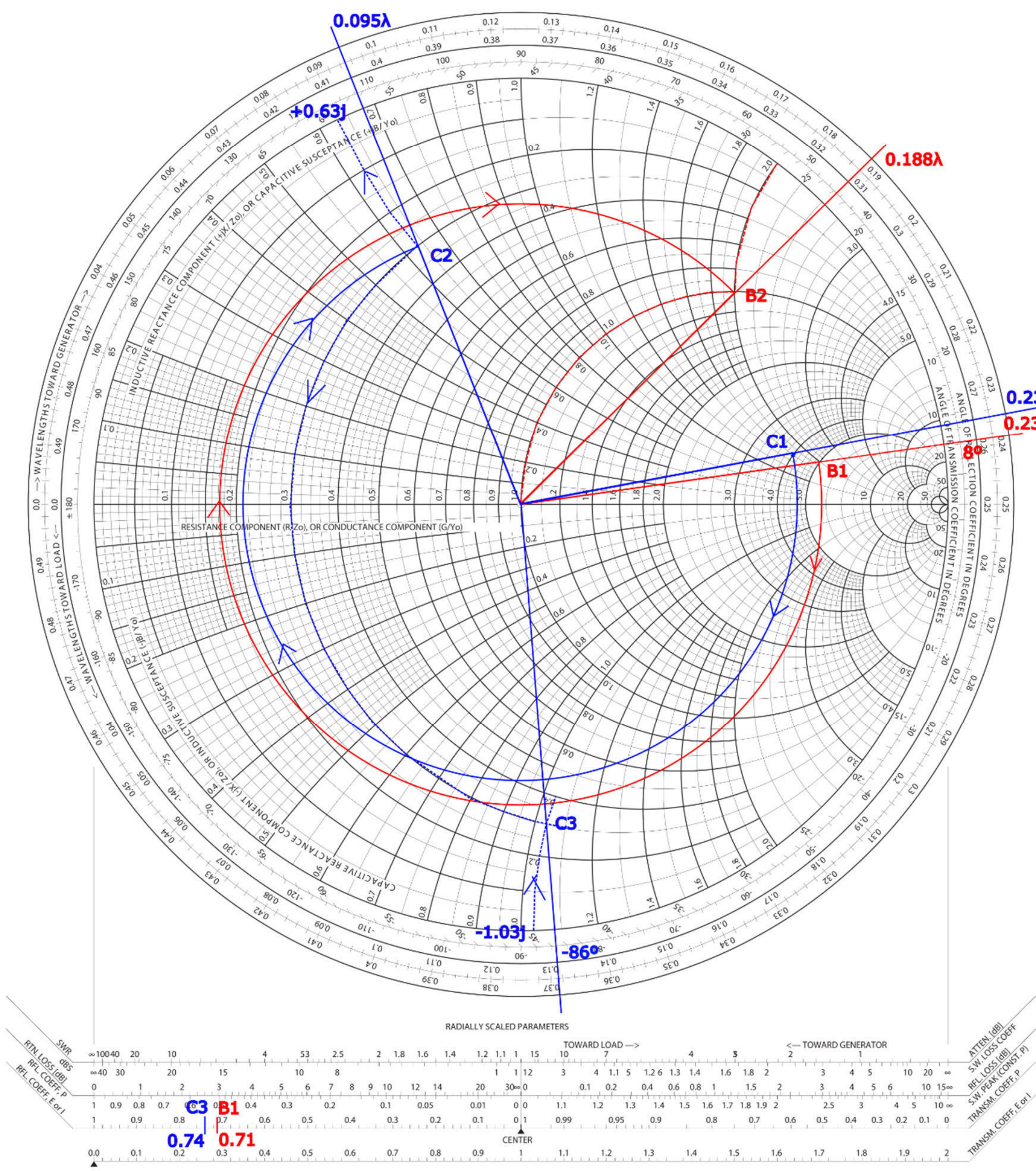
Starting at 0.237 λ this takes us to (0.237+0.0.358)-0.5= 0.095 λ (point C2 on smith)

Capacitor needs renormalising to 62.5Ω, so is -1.6j

Now move around const R circle by -1.6j. Start at 0.3+0.63j so end with 0.3-1.03j. point C3.

V reflection = $0.74\angle -86°$

A popular question with a range of answers. In (a) a common mistake was reading from the wrong axis of the Smith Chart. Most could get the width in (b) a common omission was the assumption that the fringing fields expand by the board thickness. The straight forward Smith chart in the 1st part of (c) was well answered. The 2nd part, most realised that the change in ϵ_r would give a change in the electrical length, but many missed that the impedance also changes.


0.095λ
+0.63j
C2
0.188λ
B2
0.235λ
0.239λ
C1
B1
8°
C3
-1.03j
-86°
RESISTANCE COMPONENT (R/Zo), OR CONDUCTANCE COMPONENT (G/Yo)
INDUCTIVE REACTANCE COMPONENT (+jX/Zo), OR CAPACITIVE SUSCEPTANCE (+jB/Yo)
CAPACITIVE REACTANCE COMPONENT (-jX/Zo), OR INDUCTIVE SUSCEPTANCE (-jB/Yo)
—> WAVELENGTHS TOWARD GENERATOR —>
<— WAVELENGTHS TOWARD LOAD <—
ANGLE OF REFLECTION COEFFICIENT IN DEGREES
ANGLE OF TRANSMISSION COEFFICIENT IN DEGREES
RADIALLY SCALED PARAMETERS
TOWARD LOAD —>
<— TOWARD GENERATOR
SWR
dBS
RTN. LOSS [dB]
RFL. COEFF, P
RFL. COEFF, E or I
ATTEN. [dB]
S.W. LOSS COEFF
RFL. LOSS [dB]
S.W. PEAK (CONST. P)
TRANSM. COEFF, P
TRANSM. COEFF, E or I
CENTER
C3
B1
0.74
0.71

2.

a) i) efficiency = directivity / gain = -2dB = 63.1%

efficiency = r_rad/(r_rad+r_ohmic)

r_rad+r_ohmic = 120

r_rad = 120×63.1 = 75.7 Ω

(ii) Use gain to account for losses.

Peak radiated power is 10dBm+21dB =31dBm = 1.2589W

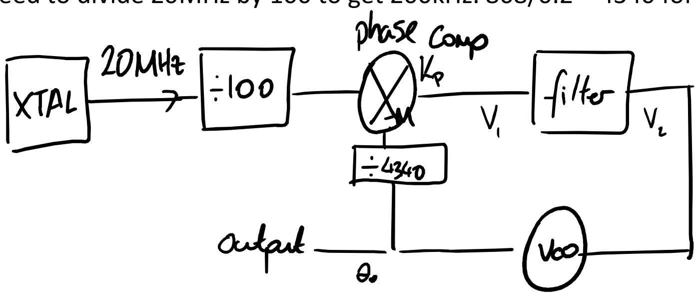
S=1.2589/(4×π×10^2)=0.001 W/m^2

Dipole has gain of 2.15dB

Ae = 0.0156m^2

Power = Ae×S = 15.6uW

b) i) Need to divide 20MHz by 100 to get 200kHz. 868/0.2 = 4340 for phase comparator



Phase comparator compares $\frac{\theta_{VCO}}{M}$ and $\frac{\theta_{XTAL}}{N}$

ii) Call V_1 Vin and V_2 Vout for the filter.

$$\frac{d\theta}{dt} = j\omega\theta = K_o V_2$$

$$V_1 = K_p(\theta_{\text{ref}} - \theta_0/M)$$

$$V_1 = -V_2 \frac{R_1 + R_1 R_2 j\omega C}{R_2}$$

$$K_p(\theta_{\text{ref}} - \theta_0/M) = -\frac{j\omega\theta}{K_f}\left(\frac{R_1 + R_1 R_2 j\omega C}{R_2}\right)$$

$$= -\frac{j\omega R_1 \theta}{K_o R_2} + \frac{R_1 \omega^2 \theta C}{K_o}$$

Use that $-\omega\theta = \dot{\theta}$ and $\omega^2\theta = \ddot{\theta}$. Compare to Mech databook.

$$\omega_n^2 = M\frac{R_1 C}{K_o K_p}$$

$$\frac{2c}{\omega_n} = M\frac{R_1}{K_o K_p R_2}$$

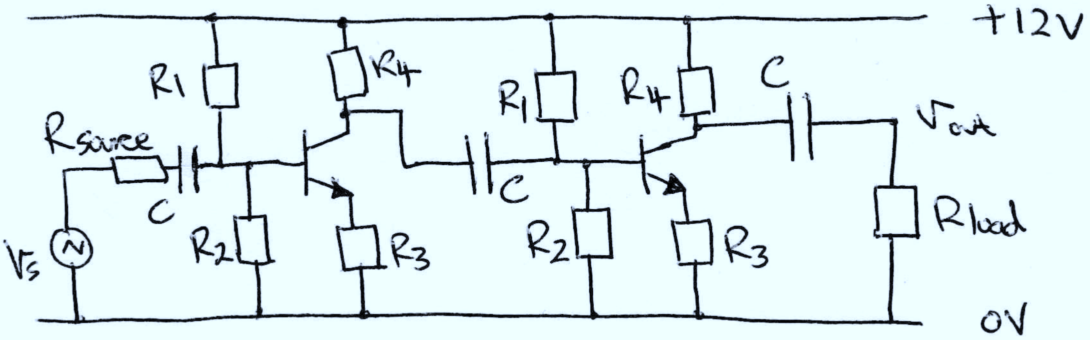
$$c = \frac{\omega_n M R_1}{2K_o K_p R_2} = \frac{M R_1}{2K_o K_p R_2}\sqrt{\frac{M R_1 C}{K_o K_p}} = \frac{1}{2R_2}\sqrt{\frac{M R_1}{C K_o K_p}}$$

(c)

At input side of the Wilkinson coupler, the transformed outputs appear in parallel. So need to match to 100 ohm to make a 50 ohm input. Therefore the lamda/4 sections should be sqrt (100×75) = 86.6 ohms. A 150 ohm resistor between the outputs completes the Wilkinson.

A few very good answers along with some very poor ones. In part (a) a worrying number struggled with the conversion of gains from dB to linear in part (i). Part (ii) was mostly well answered although many double counted the antenna efficiency by using antenna gain and then multiplying by efficiency. (There is a small ambiguity as to whether a 50% matching loss occurs on both sides, both answers were accepted). In (b) most realised that a divider is needed but omitted this from the loop analysis.

3 (a)



R_1, R_2: base bias resistors – set base voltage for $V_c = \frac{V_s}{2}$ and set i/p imped.
R_3 : negative feedback to set stage gain
R_4 : output resistance, with gain $\simeq -R_4/(R_3+r_e)$
C : coupling capacitors to pass signal frequencies but block d.c. bias voltages between stages

(b) $R_4 = 50\Omega$ for matched output impedance
To determine gain: 33dB = ×44.7 ⇒ ×6.7 per stage with 2 extra coupling stages compared to source ⇒ load case, we have a required gain of ×13.3 per stage

$\therefore R_3 + r_e = \frac{50}{13.3} = 3.76\,\Omega$

with $R_4 = 50\Omega$ and $V_c = 6V$ d.c. (½ supply), $I_c = 0.12A$

$\therefore r_e = \frac{0.025}{I_e} = 0.21\Omega \quad \therefore R_3 = 3.55\Omega$ (say 3.3Ω std.).

Choose $R_2 = 1.5 \times 50 = 75\Omega$ and set $V_B = 0.65 + 0.12 \times 3.3$
$= 1.05$ V +10% optional
$\simeq 1.1$ V

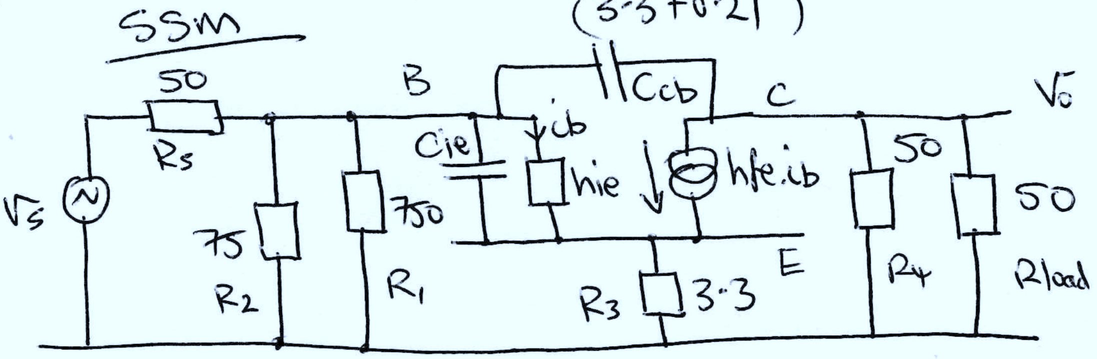
$\therefore R_1 = 750\Omega$

C should have small impedance <1Ω @ 860MHz so with $C = 10nF$, $|Z| \sim 0.02\Omega$. Input imped. = 75‖700‖750 = 62Ω o.k.

(c) For -3dB roll-off, we need value for C_{ie} and analyse small-signal model: $f_t = \frac{1}{2\pi C_{ie} r_e} = 22 \times 10^9$ with $r_e = 0.21\Omega$ ⇒ $C_{ie} = 34.4$ pF

3 contd. (c) Emitter gain $= \frac{3.3}{(3.3+0.21)} = 0.94$

SSM



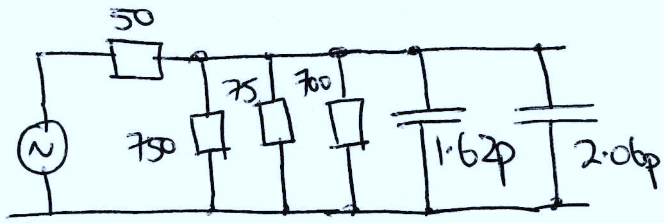
$h_{ie} = h_{fe}.r_e = 42\Omega$. Refer input side to grounded values for B-E equivalent components: $C_{ie} \times (1-0.94) \Rightarrow 2.06\,pF$

and B-C —"— —"— $h_{ie} \times \frac{1}{(1-0.94)} \Rightarrow 700\Omega$

with loaded gain $= -\frac{50/2}{(3.3+0.21)} = -7.12$

$C_{cb} \Rightarrow (1+7.12) \times 0.2 = 1.62\,pF$

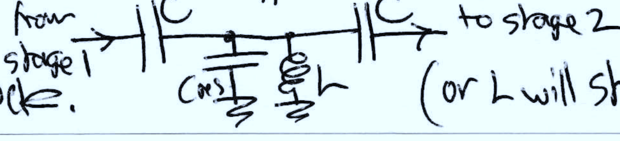
Hence input cct becomes:



$-3dB$ roll-off $= \frac{1}{2\pi R'C'}$ $\quad R' = 27.7\Omega \quad C' = 3.68\,pF$

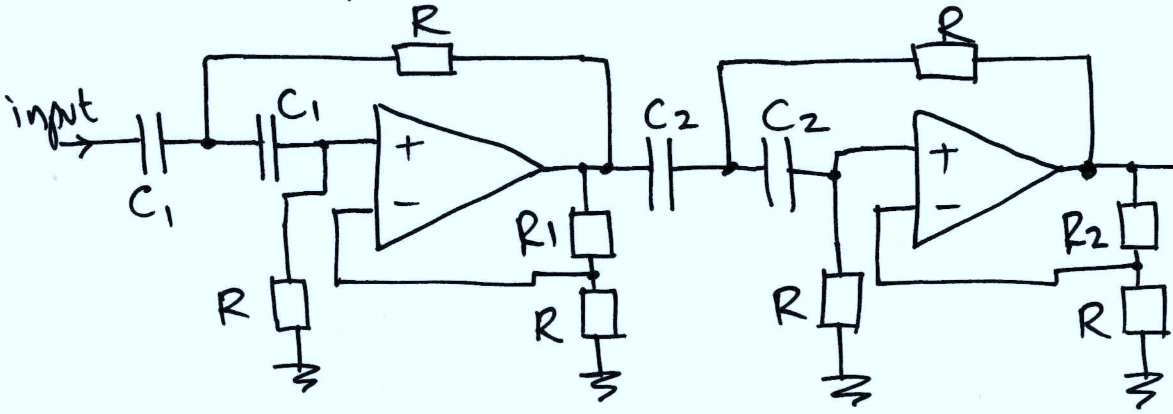
$= 1.56\,GHz$ ∴ fine for 260MHz operation, even with 2-stages cascaded. As 3dB is twice operating freq. then 2 stages cascaded will give ≈ −3dB @ operating freq. Could decrease R_3 to compensate (by ≈ 10% if reqd).

(d) Use series LC between stages. $Q = 5 = \frac{\omega L}{r}$ (where $\omega = 2\pi f$) where $r = (50+62)$ in place of central coupling capacitor and $C_{res} = \frac{1}{\omega^2 L}$ but C_{res} is small and L large: 0.33pF and 104 nH respectively. Use parallel LC with $Q = 5 = \frac{R}{\omega L}$ and $R = 50 \| 62 = 27.7\Omega$ then $L = 1.03\,nH$ and $C = 33\,pF$



but 2 coupling caps reqd. to d.c. block. (or L will short bias to gnd.)

4(a)(i) High-pass VCVS circuit : 0·2Hz roll-on



$R = 100k\Omega$

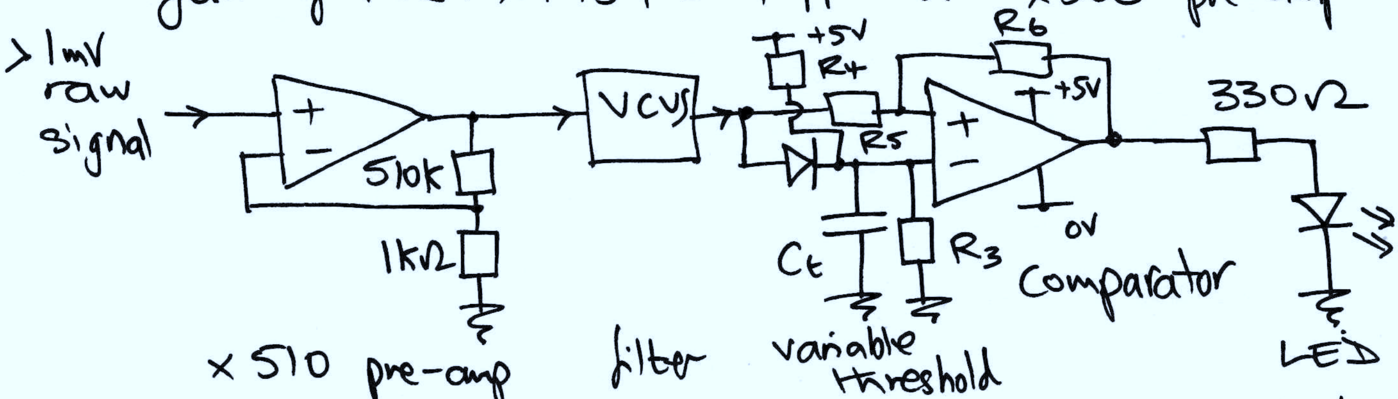
Use Bessel filter for pulse shape retention – important in [illegible] application. Downside is reduced sharpness in low freq. cut-off. $f_{-3dB} = \frac{f_n}{2\pi RC}$ for high-pass

$$\therefore 0.2 = \frac{1.432}{2\pi . 10^5 . C_1} = \frac{1.606}{2\pi . 10^5 . C_2}$$

$\therefore C_1 = 11.4\,\mu F$, $C_2 = 12.8\,\mu F$

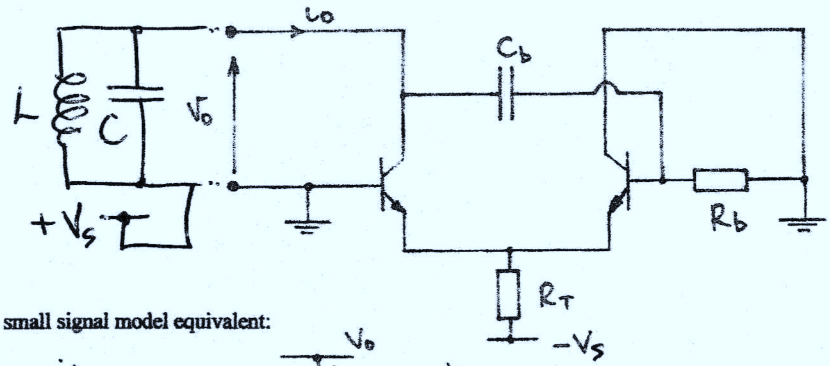
$R_1 = (A_1 - 1)R = 8.4k\Omega$, $R_2 = (A_2 - 1)R = 75.9k\Omega$

(ii) Need to amplify signals by ~1000's including VCVS gain of $1.084 \times 1.759 = 1.91$ $\therefore \approx \times 500$ pre-amp.

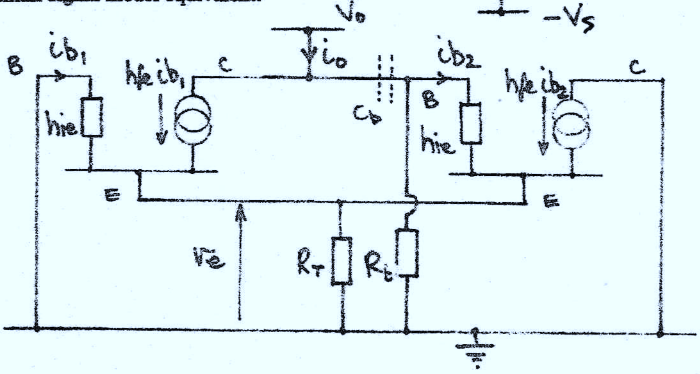


Filter output amplitude >1V pulses. Ct holds threshold for say 10s $\therefore C_t = 1000\mu F$, $R_3 = 10k\Omega$, $R_4 = 100k\Omega$ for 0.5V threshold. Larger pulses will increase threshold. R_5 and R_6 give switching hysteresis $\simeq 0.1V$ say $\therefore R_5 = 1k\Omega$, $R_6 = 47k\Omega$

4 (b)



The small signal model equivalent:



① $i_{b1} = \frac{-V_e}{h_{ie}}$, ② $i_{b2} = \frac{V_o - V_e}{h_{ie}}$, ③ $V_e \approx R_T h_{fe}(i_{b1} + i_{b2})$,

④ $i_o = h_{fe} i_{b1} + i_{b2} + V_o/R_b$ subst. ① and ② into ③ :-

$$V_e = -R_T h_{fe} \frac{V_e}{h_{ie}} + R_T h_{fe} \frac{V_o}{h_{ie}} - R_T h_{fe} \frac{V_e}{h_{ie}}$$

(small cf. as $R_T \gg h_{ie}$)

$$\therefore V_e \left(\cancel{1} + \frac{2 R_T h_{fe}}{h_{ie}}\right) = R_T \frac{h_{fe} V_o}{h_{ie}} \quad \therefore V_e \approx \frac{V_o}{2} \quad ⑤$$

subst. ① and ② into ④ and subs. for V_e using ⑤

$$\therefore i_o = -\frac{h_{fe} V_o}{2 h_{ie}} + \frac{\cancel{V_o}}{\cancel{2 h_{ie}}} + \frac{V_o}{R_b}$$

(small cf.)

$$\therefore i_o = V_o \left(\frac{1}{R_b} - \frac{h_{fe}}{2 h_{ie}}\right) \text{ and as } \frac{h_{ie}}{h_{fe}} = r_e,$$

$$Z_o = \frac{V_o}{i_o} = \left(\frac{1}{R_b} + \frac{1}{-2 r_e}\right)^{-1} \Rightarrow \underline{Z_o = R_b // -2 r_e}$$

$$f_{res} = \frac{1}{2\pi\sqrt{LC}} = 1.58\,\text{GHz}$$

with $L = 1\,\text{nH}$

$C = 10.1\,\text{pF}$

Parasitic // loss R

$= Q\omega L$

$= Q\, 2\pi f_{res} L$

$= 199\,\Omega$

Load R

$= 100\,\Omega$

$\therefore$ -ve resistance

$< 100 // 199\,\Omega$

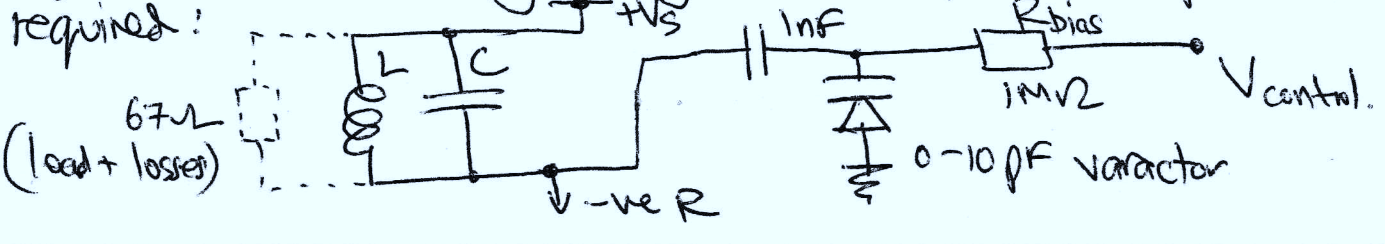
$< 67\,\Omega$

$\therefore$ choose $-R = -50\,\Omega$ $\therefore r_e = 25\,\Omega = \frac{0.025}{I_c}$

Current in $R_T \Rightarrow \frac{(3 - 0.65)\,V}{R_T\,\Omega} = 2 \times I_c = 2\,\text{mA}$

$\therefore R_T = 1.18\,\text{k}\Omega$

To tune, add varactor across LC circuit. To reduce f_{res} by by 40% means increasing C by almost 100%, hence ~10 pF varactor required:



Examiner’s comments

Q1 and Q2 – comments in crib text

Q3 RF amplifier

A very popular question with good attempts on the whole. The 2-stage amplifier design was well answered, although the gain was sometimes incorrect by a factor of 2 either way. The frequency response was also quite well attempted in many cases, although the unloaded gain was occasionally considered rather than the loaded value. The resonant filter section at the end attracted a number of attempts of rather variable quality.

Q4 VCVS filters and oscillator

The VCVS filter section was quite straightforward and well attempted in most cases, with a correct choice of filter type and values in many cases although a Chebyshev filter would have been a poor choice given the importance of pulse shape. The circuit design was less well answered in general, the best attempts included a variable gain section or amplitude tracking threshold. The negative impedance oscillator was generally well attempted.