Model answer to Q 1(a): Extended Bookwork and Discussions in Class

a) Qualitatively, describe what can be heard when an amplitude modulation radio receiver is not tuned to any broadcasting station and switched from UHF to VHF to HF. Justify possible reasons for what is heard, in terms of the external source and sources within the receiver. For full marks, quote any appropriate equation to support your answer.

Table 1-1 IEEE Frequency Spectrum

Frequency Band	Frequency	Wavelength
ELF (Extreme Low Frequency)	30–300 Hz	10,000–1000 km
VF (Voice Frequency)	300–3000 Hz	1000–100 km
VLF (Very Low Frequency)	3–30 kHz	100–10 km
LF (Low Frequency)	30–300 kHz	10–1 km
MF (Medium Frequency)	300–3000 kHz	1–0.1 km
HF (High Frequency)	3–30 MHz	100–10 m
VHF (Very High Frequency)	30–300 MHz	10–1 m
UHF (Ultrahigh Frequency)	300–3000 MHz	100–10 cm
SHF (Superhigh Frequency)	3–30 GHz	10–1 cm
EHF (Extreme High Frequency)	30–300 GHz	1–0.1 cm
Decimillimeter	300–3000 GHz	1–0.1 mm
P Band	0.23–1 GHz	130–30 cm
L Band	1–2 GHz	30–15 cm
S Band	2–4 GHz	15–7.5 cm
C Band	4–8 GHz	7.5–3.75 cm
X Band	8–12.5 GHz	3.75–2.4 cm
Ku Band	12.5–18 GHz	2.4–1.67 cm
K Band	18–26.5 GHz	1.67–1.13 cm
Ka Band	26.5–40 GHz	1.13–0.75 cm
Millimeter wave	40–300 GHz	7.5–1 mm
Submillimeter wave	300–3000 GHz	1–0.1 mm

It can be seen that as frequency decreases from UHF to VHF to HF (i.e. 3 GHz down to 3 MHz), the galactic noise increases. As a result, the listener will hear the background noise increase. This is evident from the equation for the available input noise power into the receiver;

$$Ni = kT_A B$$

where T_A is the antenna noise temperature, which includes the contribution from the sky temperature. Therefore, assuming a constant bandwidth B into the detector as frequency changes, the input noise power will increase with a decrease in frequency. In addition, flicker (or 1/f) noise is generated within all active semiconductor devices. As a result, internal noise generation within the receiver increases with a decrease in frequency.

[4]

Model answer to Q 1(b): Discussions in Class

b) State the most common domestic application at Point B on Figure 1.1, indicating its approximate frequency, and briefly explain possible implementation trade-offs at such frequencies. For full marks, quote an appropriate equation to support your answer.

The mobile phone operating at ~900 MHz (e.g. GSM900 in Europe) is the most common application at this frequency. At this frequency, it is very challenging to implement low loss components using distributed transmission lines, since wavelength is inversely proportional to frequency. As a result, high quality factor lumped-elements are employed, because they can be made extremely compact, for mobile phone handsets. On the other hand, since mobile phone base stations have less constraint on size, transmission line components are used; especially when judicious use of high dielectric constant materials can be employed to shrink size down – with size-scaling roughly proportional to the square root of the dielectric constant.

[4]

Model answer to Q 1(c): Discussions in Class

c) State the most common domestic application at Point C on Figure 1.1. For this application, explain why the uplink frequency is greater than the downlink frequency. For full marks, quote an appropriate equation to support your answer.

The satellite television receive-only (TVRO) service operates at approximately 10 GHz. The relationship between the size of an antenna and its gain is given below.

$$\text{Effective Aperture of the Recieving Antenna, } A_{RX} \sim G_{RX} \cdot \frac{\lambda_o^2}{4\pi}$$

On a satellite, its uplink and downlink antennas may well have the same aperture size. As a result, as its operating frequency increases so does the gain. However, as gain increases the half-power beam width reduces. Therefore, a higher frequency satellite uplink results in a higher gain pencil beam that is idea for pinpointing the uplink ground station, while also minimizing the risk of co-channel interference with other services. Moreover, the lower frequency satellite downlink results in a wider beam width antenna that has a greater footprint for wider geographical broadcasting coverage.

[4]

Model answer to Q 1(d): Discussions in Class

d) Briefly explain the reason for the peak in atmospheric noise at Point D in Figure 1.1 and state the approximate frequency of this peak.

This peak is the result of atmospheric attenuation due to resonance absorption from water molecules at ~22 GHz.

[2]

Model answer to Q 1(e): Discussions in Class

e) State the most common application at Point E on Figure 1.1, indicating its approximate frequency, and briefly explain why this application is in this band.

Local multi-point distribution systems (LMDS) operate in frequency bands between 24 and 38 GHz, because there is a low atmospheric attenuation window in this part of the frequency spectrum; providing better signal-to-noise performance. Also, at this relatively high frequency, small aperture antennas give high gain pencil beam performance.

[2]

Model answer to Q 1(f): Discussions in Class

f) Briefly explain the reason for the broad peak in atmospheric noise at Point F in Figure 1.1 and state the approximate centre frequency of this broad peak. State the most common domestic application and why it is found in this band.

This peak is the result of atmospheric attenuation due to resonance absorption from oxygen molecules at ~60 GHz. Indoor wireless local area networks operate in this band, due to the naturally occurring high suppression of co-channel interference.

[2]

Model answer to Q 1(g): Discussions in Class

g) State the most common application at Point G on Figure 1.1, indicating its approximate frequency, and briefly explain why this application is in this band.

This represents the 94 GHz military band, operating in a low atmospheric attenuation window. It is used for systems that need high gain and very narrow pencil beam antennas, such as high performance radars and imaging applications.

[2]

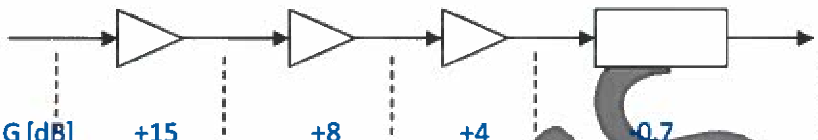
Model answer to Q 2: Computed Example

Using the following general equations:

$$C = P_{OUT}(fo) = G_1 G_2 P_{IN} \quad and \quad IP_3 = \left(IP_3|_1\, G_2\right) \| \, IP_3|_2$$

$$IMD_3 = \frac{C}{I_3} \sim \left(\frac{IP_3}{C}\right)^2 \equiv 2\left(IP_3[dBm] - C[dBm]\right)[dBc]$$

$$F = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1 G_2} + \frac{F_4 - 1}{G_1 G_2 G_3} + \ldots$$

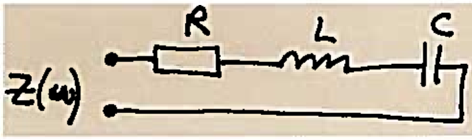


G [dB]	+15	+8	+4	-0.7
IP_3 [dBm]	+30	+40	+50	
F [dB]	+1	+2	+3	

Pout [dBm]	+3	+18	+26	+30	+29.3	[2]
IP_3 [dBm]		+30	+35.9	+39.5	+38.8	[6]
IMD_3 [dBc]		+24	+19.8	+19	+19	[3]
I_3 [dBm]		-6	+6.2	+11	+10.3	[3]
F [dB]	+1.1					[6]

Model answer to Q 3(a): Standard Derivation

a) Draw the circuit diagram for a series *RLC* tuned circuit, identifying all the elements, and write down the expression for its total impedance. Assume that the element values are all frequency invariant.



$$Z(\omega) = R + j\left(\omega L - \frac{1}{\omega C}\right)$$

[3]

Model answer to Q 3(b): New Derivation

b) From the total impedance in 3(a), and given the general mathematical identity in equation (3.1), derive an expression for the associated differential-phase group delay for all frequencies.

$$\frac{\partial}{\partial x}\{tan^{-1}(u)\} = \frac{1}{1+u^2}\frac{\partial u}{\partial x} \qquad (3.1)$$

$$\frac{\partial \angle Z(\omega)}{\partial \omega} = \frac{1}{1+\frac{\left(\omega L - \frac{1}{\omega C}\right)^2}{R^2}} \cdot \frac{1}{R}\left(L + \frac{1}{\omega^2 C}\right)$$

DIFFERENTIAL-PHASE GROUP DELAY:

$$\tau(\omega) = \frac{-\partial \angle Z(\omega)}{\partial \omega} = \frac{-R\left(L + \frac{1}{\omega^2 C}\right)}{R^2 + \left(\omega L - \frac{1}{\omega C}\right)^2}$$

There should be a plus instead of a minus in the above proof!!!!

[5]

Model answer to Q 3(c): Bookwork

c) For the element values in 3(a), write down the well-known expressions for:

i) Ideal lossless resonance frequency, ω_I.

$$\omega_I = \frac{1}{\sqrt{LC}}$$

[1]

ii) Unloaded quality factor at ω_I.

$$Q_U(\omega_I) = \frac{\omega_I L}{R}$$

[1]

Model answer to Q 3(d): New Derivation

d) Using the results in 3(b) and 3(c), prove that the unloaded quality factor at ω_I is directly proportional to the ratio of differential-phase group delay and the time period of the cycle.

$$\tau(\omega_I) = -2\frac{L}{R} \Rightarrow \frac{Q_U(\omega_I)}{\omega_I}(-2)$$

$$\therefore \quad Q_U(\omega_I) = \frac{\omega_I}{2}\,|\tau(\omega_I)|$$

$$\text{BUT} \quad \omega_I = 2\pi f_I = \frac{2\pi}{T_I}$$

$$\therefore \quad Q_U(\omega_I) = \pi \cdot \frac{|\tau(\omega_I)|}{T_I}$$

[4]

Model answer to Q 3(e): New Derivation

e) Given an n^{th}-stage BPF, constructed using lossless LC tuned circuits, the maximum theoretical insertion phase variation is $n\pi$ (as frequency increases from dc to infinity).

i) Derive a qualitative expression for the differential-phase group delay, in terms of the order of the filter and -3 dB bandwidth.

$$\Delta\phi(\text{DC TO }\infty) = n\pi \qquad \therefore \Delta\phi(\omega_1 \text{ TO } \omega_2) < n\pi$$

$$\tau \approx -\frac{\Delta\phi}{\Delta\omega} \lesssim \frac{-n\pi}{(\omega_2-\omega_1)} = -\frac{n\pi}{\Delta\omega|_{3dB}}$$

[2]

ii) Using the rough approximation in 3(e)(i), show how the loaded quality factor at centre frequency ω_0 is directly proportional to the ratio of differential-phase group delay and the time period of the cycle.

$$Q_L(\omega_0) = \frac{\omega_0}{(\omega_2-\omega_1)} \qquad \therefore \quad Q_L(\omega_0) \gtrsim \frac{\omega_0\,|\tau|}{n\pi}$$

$$\omega_0 = 2\pi f_0 = 2\pi/T_0 \qquad \therefore Q_L(\omega_0) \gtrsim \frac{2}{n}\cdot\frac{|\tau|}{T_0}$$

[2]

iii) Comment on the similarity of the expression in 3(e)(ii) for quality factor and that derived in 3(d).

$$\therefore Q_L(\omega_0)\big|_{n=1} \gtrsim 2\cdot\frac{|\tau|}{T_0} \qquad \text{c.f.} \quad Q_U(\omega_I) = \pi\cdot\frac{|\tau|}{T_I}$$

It can be clearly seen that the filter's qualitative expression for Q-factor is close to the exact expression for a single resonator.

[2]

Model answer to Q 4(a): Discussions in Class

a) Compare and contrast lumped-element impedance matching over the use of distributed-elements, in terms of frequency performance. How does this affect their role in dc biasing networks?

Lumped-element components are attractive because of their small size and their smooth frequency characteristic in their normal operating frequency range. For this reason they are used extensively in the RF electronics industry. However, as frequency increases they exhibit resonances, which make them unusable as frequency approaches the first self-resonant frequency. In contrast, distributed-element transmission lines can be used at high frequencies, but are not desirable at low frequencies. This is because their role depends on their electrical length, which results in a physical length that is inversely proportional to frequency. Therefore, as frequency decreases they can become prohibitively long.

Lumped-element components are used to implement a low-pass filter for DC biasing networks. Unfortunately, due to undesirable resonances within the components, it is important to make sure that there is sufficient out-of-band isolation above the filter's cut-off frequency, to avoid the RF signal path "seeing" the power supply. At higher frequencies, quarter-wave transformers are employed in biasing networks, with the use of distributed-element lines.

[4]

Model answer to Q 4(b): Bookwork

b) With lumped-element impedance matching, what terminating impedance conditions are best suited for L-match, π-match and T-match networks?

L-Match – Most useful with one low impedance and one high impedance terminations,

$R_{MIN} < R_{INTERMEDIATE} < R_{MAX}$

π-Match – Most useful with both high impedance terminations (e.g. low frequency valves)

$R_{INTERMEDIATE} < R_{MIN}$

T-Match – Most useful with both low impedance terminations (e.g. transistors)

$R_{MAX} < R_{INTERMEDIATE}$

[4]

Model answer to Q 4(c): Computed example

c) A 2 GHz narrow-band amplifier has an output impedance of (5 – j7) Ω and must be matched to a system impedance of 50 Ω. Design simple matching circuits to achieve maximum power transfer:

i) With the use of one lumped-element component and one distributed-element component.

(i) An inductor is connected in series with the output of the transistor. If it has an inductance of $L = +7/(2\pi fo) = 0.557$ nH then it will resonate-out the reactive component of the transistor's output impedance. All that is needed now is a quarter-wavelength impedance transformer, with a characteristic impedance of $\sqrt{(5*50)} = 15.81\ \Omega$, to be inserted in series with the inductor.

[4]

ii) With the use of two lumped-element components.

(ii) An inductor is connected in series with the output of the transistor. If it has an inductance of $L = +7/(2\pi fo) = 0.557$ nH then it will resonate-out the reactive component of the transistor's output impedance. Now an L-network is cascaded with the inductor.

$$Loaded\ Q-factor,\ Q_L = \sqrt{\frac{Zo}{R_{OUT}} - 1} = 3$$

$$Q_S = \frac{|X_S|}{R_{OUT}} \equiv Q_L \quad \therefore L_S = \frac{3x5}{2x\pi x2x10^9} = 1.194nH$$

$$Q_P = \frac{R_G}{|X_P|} \equiv Q_L \quad \therefore C_P = \frac{3}{50x2x\pi x2x10^9} = 4.775pF$$

In order to reduce the component count, L_S is combined with L to give a combined inductance of 1.751 nH.

[4]

Model answer to Q 4(d): Discussions in Class

d) With MMIC technology, discuss the difficulties implementing lumped-element and distributed-element components. How can micromachining technologies help to overcome some of the problems with monolithic implementations? How does this affect their role in filter networks?

With both lumped-element and distributed-element components, monolithic technology has problems with the conductor dimensions being very small. As a result of the limited cross-sectional areas, these components exhibit high current densities and poor unloaded Q-factors. Therefore, they are not suitable for implementing low loss or high selectivity filters. In addition, with lumped-element components, their values are limited because of unwanted shunt capacitive parasitics that result in low self-resonant frequencies. Also, their physical structures begin to exhibit distributed-element characteristics at higher frequencies.

Micromachining technology allows lumped-element and distributed-element components to be realised without the use of dielectrics. As a result, they will not suffer from dielectric losses. Moreover, without a dielectric, their size increases and, thus, it may be possible to reduce the current densities within the conductors. This can significantly increase the Q-factors of the components and, thus, make them more attractive for realising low loss and high selectivity filters.

[4]

Model answer to Q 5(a): Discussions in Class

a) Using S-parameters, what are the levels of insertion loss and return loss at the -3 dB cut-off frequency for a lossless filter? You are asked to design a filter with a maximum pass band return loss level of -6.868 dB. What will be the worst-case pass band insertion loss ripple for a lossless band-pass filter?

From the conservation of energy, for a lossless two-port network: $|S21|^2 = 1 - |S11|^2$

At the –3 dB cut-off frequency, $|S21|^2 = 0.5$ and, therefore, $|S11|^2 = 0.5$
In other words both the insertion loss and return loss levels are -3 dB.

For a maximum return loss level or –6.868 dB, the worst-case insertion loss level is 1 dB.

[5]

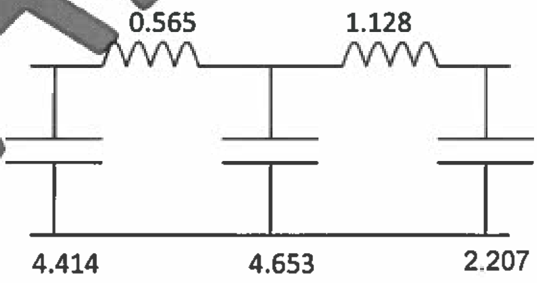
Model answer to Q 5(b): Computed Example

b) Using the worst-case level of ripple calculated in 5(a), but this time for the stop band return loss, design a lumped-element *LC* band stop filter to meet the following specifications:

Lower pass band -3 dB cut-off frequency:	540 MHz
Upper pass band -3 dB cut-off frequency:	660 MHz
Stop band bandwidth:	60 MHz
Band stop attenuation:	> 45 dB
Source impedance, *Zs*:	50 Ω
Load impedance, Z_L:	100 Ω

-3 dB cut-off frequencies are at f_{P1} = 540 MHz and f_{P2} = 660 MHz
Centre frequency, fo = $\sqrt{(f_{P1} f_{P2})}$
Pass band bandwidth, Bp = $f_{P1} - f_{P2}$ = 120 MHz
Stop band bandwidth, Bs = 60 MHz
f/fc = Bp/Bs = 2

From the attenuation curves, the 5th order 1 dB ripple Chebyshev filter with $Rs/R_L = 0.5$ meets the specification with a stop band attenuation margin of 3 dB. The normalised values for the low-pass prototype is given below:

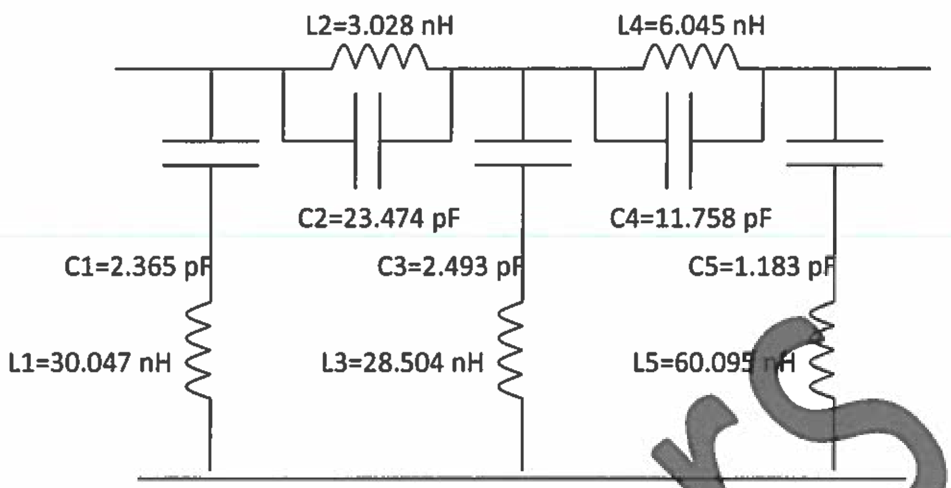


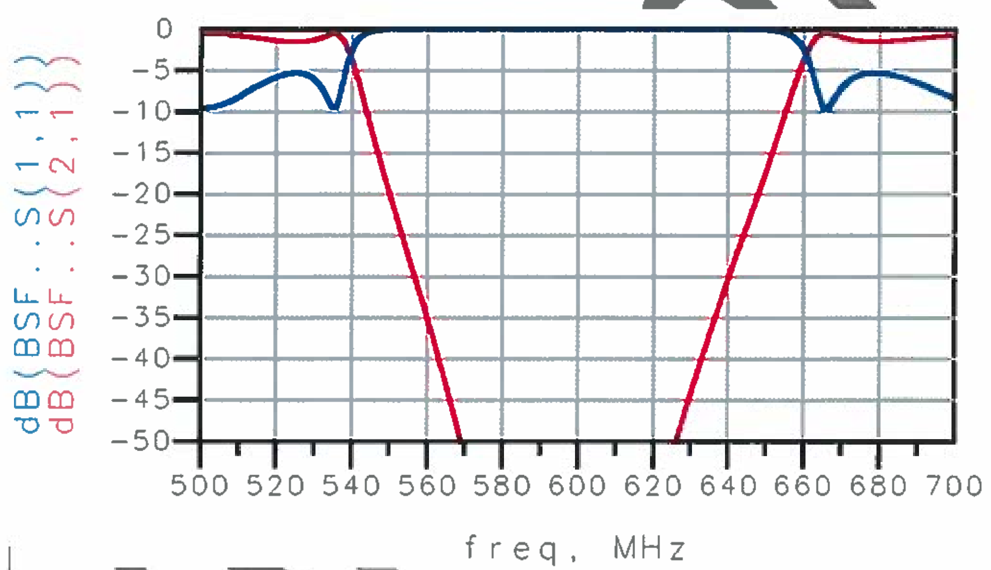
For a band stop filter, the capacitors are replaced by a series *LC* resonators and the inductors are replaced by parallel *LC* resonators. The un-normalized shunt connected series tuned circuit element values are:

$$Cs = \frac{BpCn}{2\pi f_o^2 R_L} \quad and \quad Ls = \frac{R_L}{2\pi Bp L_n}$$

The un-normalized series connected parallel tuned circuit element values are:

$$Cp = \frac{1}{2\pi BpCnR_L} \quad and \quad Lp = \frac{BpLnR_L}{2\pi f_o^2}$$





The slight deviations in pass band levels are due to rounding errors in the component values (3 decimal places).

[10]

<u>Model answer to Q 5(c): Discussions in Class</u>

c) Using S-parameters, define differential-phase group delay and explain the general relationship between its frequency response and that of sharp cut-off insertion loss characteristics.

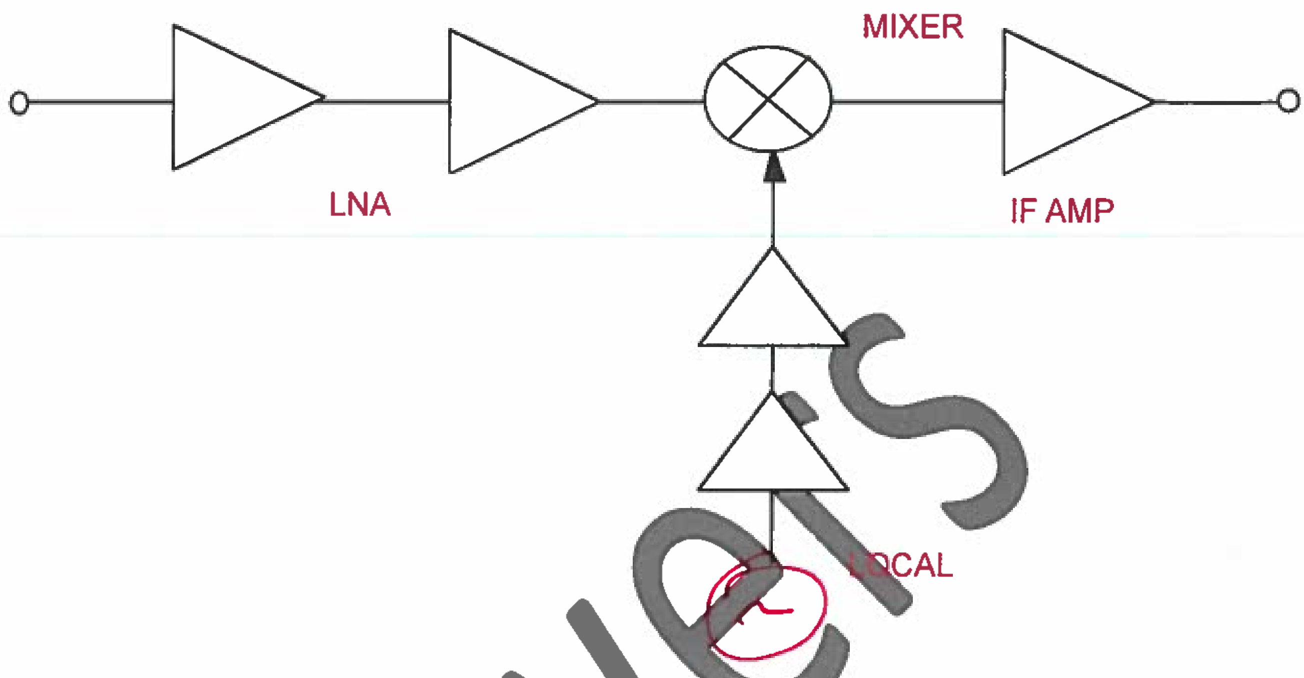
Differential-phase group delay is defined as $\tau = -\partial \angle S21 / \partial \omega$

With sharp cut-off frequencies, high order filters are needed. This means that there are large numbers of passive filter components, where electromagnetic energy is exchanged between them. As a result the signal stays within the filter longer than for filters with a less sharp insertion loss cut off and, thus, the differential-phase group delay is higher. Also, more energy is dissipated in the components and, therefore, insertion loss is higher unless larger (higher Q-factor) components are employed.

[5]

Model answer to Q 6(a): Extended Bookwork and Discussions in Class

a) Draw the high level subsystems block diagram from what can be deduced from Figure 6.1. An active transistor should be represented as a discrete amplifier stage.



[4]

Model answer to Q 6(b): Bookwork

b) State the type of receiver that Figure 6.1 represents and list its general characteristics.

Single-Conversion Superheterodyne Receiver

* Image rejection filter should also be included !!!
* Super-sonic Heterodyne (or Superhet for short) refers to the fact that the IF frequency is much higher than audio frequencies
* Selectivity is improved by sharper IF band-pass filters
* Increased complexity and, therefore, cost

[4]

Model answer to Q 6(c): Discussions in Class

c) What type of mixer is used and how is it configured to operate? Briefly explain what this design is attractive for an MMIC.

This is a single-ended resistive mixer. The LO is applied to the gate and the RF applied to the drain electrodes of a cold-FET. The IF is taken from the drain electrode. Low conversion loss (typically 13 dB for a LO power of 2 dBm at $Vgs = 0$ V) can be improved with an optimum gate-source bias voltage, with good linearity performance. For a monolithic microwave integrated circuit (MMIC), space is at a premium. As a result, a single-ended solution is preferred over more complicated topologies that require more expensive real estate. The use of a 3-terminal device, over a discrete diode allows extra flexibility in performance offered with the use of a bias voltage to control intrinsic parasitic components.

[4]

Model answer to Q 6(d): Discussions in Class

d) If the LO needs the use of a dielectric resonator, where would this resonator be located in practical applications and what is the reasoning for this?

The dielectric resonator is normally located off the chip. The reason is that MMMIC technology only offers 2D patterning and of lossy dielectric materials. As a result, high Q-factor resonator puck resonators cannot be integrated.

[2]

Model answer to Q 6(e): Discussions in Class

e) In general, what is the rule-of-thumb power level needed for an LO signal at the input of a general mixer used in a receiver and justify the reason for this level?

In general, the local oscillator should present a power level of ~0 dBm to the mixer within the receiver. When compared to the relatively very low RF input signal power levels, the much higher LO power level acts as a pump that dominates the behaviour of the nonlinear intrinsic parasitic(s) that performs the mixing action.

[2]

Model answer to Q 6(f): Bookwork

f) For a more complete receiver, what subsystem block is missing from the front end? Explain why this is needed and list three possible solutions to alleviate the problem.

[4]