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A06

IMPERIAL COLLEGE LONDON

DEPARTMENT OF ELECTRICAL AND ELECTRONIC ENGINEERING
EXAMINATIONS 2005

MSc and EEE PART IV: MEng and ACGI

Corrected Copy

RADIO FREQUENCY ELECTRONICS

Tuesday, 10 May 10:00 am

Time allowed: 3:00 hours

There are SIX questions on this paper.

Answer FOUR questions.

All questions carry equal marks

Any special instructions for invigilators and information for candidates are on page 1.

Examiners responsible	First Marker(s) :	S. Lucyszyn
	Second Marker(s) :	C. Papavassiliou

1 Consider direct-conversion transmitter architectures.

(a) With the aid of illustrations, briefly explain why direct-conversion transmitter architectures appeal to designers of wireless systems, when compared to superheterodyne transmitter architectures. [3]

(b) What is the major disadvantage of not having an IF stage in direct-conversion transmitters? [3]

(c) Define the term ACLR and discuss how the output power level of the modulator can be adjusted to meet specified ACLR and output noise emissions levels. [3]

(d) Define the term WCDMA and give a typical value for the output carrier power level of a WCDMA basestation. [2]

(e) With a modulator having an output power of +10 dBm and a noise floor of -150 dBm/Hz, the ACLR is -58 dBc. For the required output power at the antenna quoted in 1d), calculate the post-modulation gain and noise power at the antenna within a 1 MHz bandwidth. [3]

(f) With four modulators, each having an output power of -20 dBm and a noise floor of -150 dBm/Hz, the ACLR is -65 dBc. For the required output power at the antenna quoted in 1d), calculate the post-modulation gain and noise power at the antenna within a 1 MHz bandwidth. [3]

(g) If the maximum ACLR and noise power emission levels are set at -60 dBc and -30 dBm, respectively, comment on the performance given in 1(e) and 1(f). Based on the discussion from 1(c), suggest a simple strategy for meeting all the design specifications. [3]

2 (a) Derive, from first principles, the general radar range equation. Assume the following lossless system:

- The transmitter has an input power P_T feeding an antenna having a power gain G_T that is located at a range R_T from the target
- The target has a radar cross-section σ
- The receiver has an input power P_R delivered by an antenna having a power gain G_R that is located at a range R_R from the target

[5]

(b) Two amateur radio enthusiasts, living 1 km apart, decide to communicate with each other at 432 MHz, using the overhead moon as a passive satellite (i.e. a reflecting target). The mean distance to the moon from both enthusiasts is R_M = 381,500 km and the mean diameter of the moon is D_M = 3,500 km:

(i) If the same lossless paraboloidal reflector antennas are used at both locations, calculate the power gain of the antennas if they have a diameter of 9 m.

[4]

(ii) Comment on the resulting beam efficiency for this application if the 3 dB beamwidth of the antennas is 5°.

[2]

(iii) With P_T = 20 dBW, calculate the power at the receiver. Neglect the effects of the earth's atmosphere. As a first order approximation, assume that the moon's radar cross-section can be modelled as a perfectly reflecting flat circular disc.

[3]

(iv) With an antenna temperature T_A = 100 K and a receiver noise temperature T_{RX} = 75 K, calculate the carrier-to-noise power ratio at a receiver having a final IF bandwidth of 7 kHz. Take Boltzmann's constant, $k = 1.38 \times 10^{-23}$ W/Hz/K.

[3]

(v) Calculate the minimum possible propagation delay time between the two ground stations if both antennas suffer from unwanted sidelobes.

[3]

3 An amplifier chain is illustrated in Figure 3.1. All sub-systems are assumed to be perfectly impedance matched.

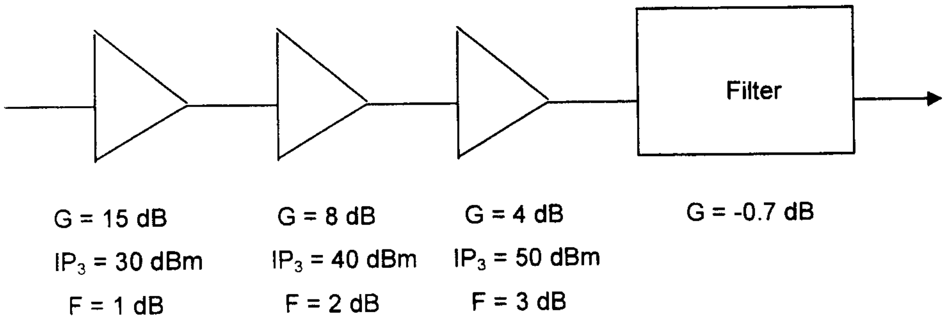


Figure 3.1

For an overall input power of 3 dBm, calculate the following at the output of each sub-system:-

(i) C [2]

(ii) IP_3 [6]

(iii) IMD_3 [3]

(iv) I_3 [3]

(v) Overall F [6]

All variables have their usual meaning.

4 (a) Using simple S-parameter analysis, evaluate the loss in both output power and power gain for a single-ended amplifier failure within:

(i) a single-balanced amplifier [4]

(ii) a double-balanced amplifier [4]

Assume the single-ended amplifiers have an identical voltage transfer function A and they remain perfectly impedance matched.

(b) Derive an expression for the insertion loss of a non-ideal 3 dB coupler having an equal power coupling of C [dB]. [4]

(c) Derive an expression for the overall power gain and η_{ADD} for a balanced amplifier having input and output coupler losses of α_i and α_o, respectively. [4]

(d) A single-ended amplifier has a power gain of 10 dB and a basic efficiency of 25%. The input and output couplers have an insertion loss of 0.5 dB. Using the equations derived from 4(c), calculate the overall power gain and PAE. [4]

5 (a) Calculate the loaded-Q factor for a resonant circuit that has a centre frequency at f_o = 1.8 GHz and 3 dB bandwidth BW_{3dB} = 200 MHz. [2]

(b) Calculate both the loaded and unloaded quality factors for a series *L-C-R* resonator having an inductance of 80 nH, combined series loss resistance of 1 Ω, resonant frequency of 1.8 GHz and terminated in a load impedance Z_o = 50 Ω. [3]

(c) Draw a simplified block diagram for a typical RF feedback oscillator. What are the two main conditions for oscillation to occur? [2]

(d) For an oscillator that is designed around a series *L-C-R* resonator, define a criteria for minimum phase noise in terms of the quality factor of the resonator. Calculate the resonator's optimum value of load impedance for a combined series loss resistance of 1 Ω. Also, calculate the resulting insertion loss for this resonator. [4]

(e) Given an inverting amplifier, having input and output impedances of 50 Ω:

(i) Design the resonator with its integral *L-C* impedance transformers for a 1.8 GHz low noise oscillator. The resonator is based around a discrete 80 nH inductor having a series loss resistance of 1 Ω. [3]

(ii) Comment on the practicality of the series capacitance value calculated for the resonator. [2]

(iii) What should the insertion phase of this resonator be at resonance and will it meet the requirement for oscillation? [1]

(iv) What is the minimum theoretical power gain of the inverting amplifier at resonance? [1]

(f) Given the choice of silicon or GaAs, which of these semiconductor would give a better phase noise performance and why? [2]

6 (a) Explain why filters having sharp frequency roll-off characteristics require large components to achieve low insertion losses.

[4]

(b) Explain why impedance and admittance inverters are required for realising practical narrow bandwidth filters. In addition, with the use of simple block diagrams, explain how these inverters work.

[6]

(c) Redesign the 1.8 GHz resonator topology shown in Figure 6.1, by employing all capacitive admittance inverters, so that the series tuned circuit can be replaced with a shunt parallel tuned circuit.

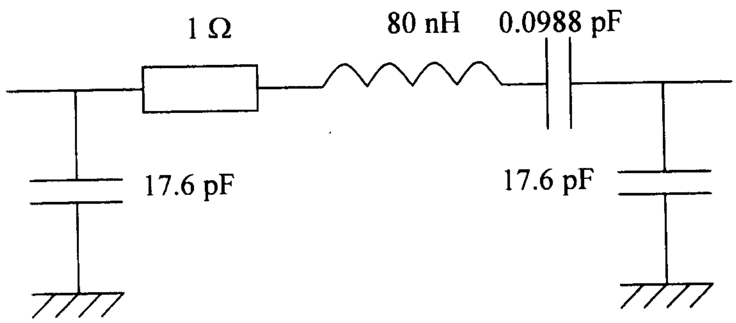


Figure 6.1

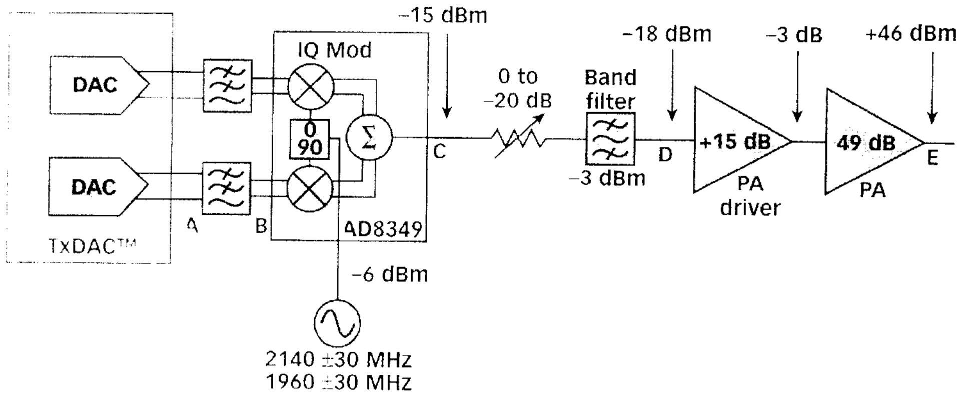
[8]

(d) What effect on the insertion phase does the all capacitance admittance inverter have?

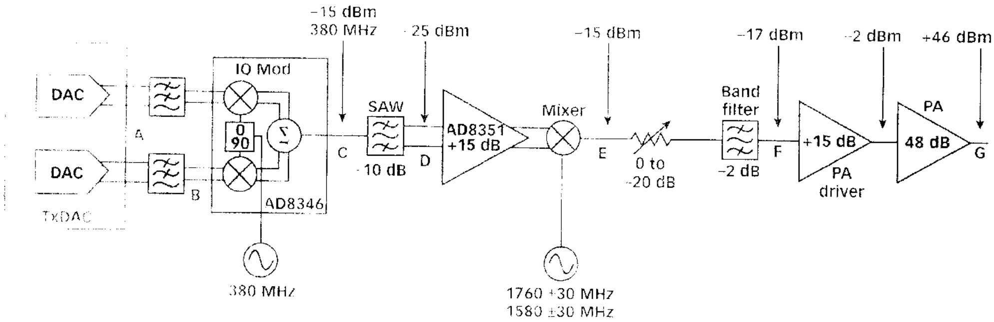
[2]

Model answer to Q 1(a): Bookwork

DIRECT-CONVERSION TRANSMITTER



SUPERHETERODYNE TRANSMITTER



Direct-conversion transmitters appeal to designers of wireless systems for their simplicity and low cost. There is no need for an IF stage and their associated mixers, local oscillator and filtering.

[3]

Model answer to Q 1(b): Bookwork

A traditional superheterodyne transmitter has an IF stage, where the filters are more selective than at the RF stage. This allows filtering of broadband noise, images and spurious components.

[3]

Model answer to Q 1(c): Bookwork

ACLR = acceptable adjacent channel leakage (power) ratio. Power in the adjacent channels is dominated by spectral leakage from the modulator and not from its noise floor. The goal of the transmitter chain designer is to select a modulator output level that provides a minimum ACLR, while also satisfying system noise requirements. For example, if the modulator's output power level is increased then less post-modulation gain is required and, therefore the output noise power emission levels are reduced. However, if the modulator's output level is too close to the compression point then the ACLR degrades to unacceptable levels.

[3]

Model answer to Q 1(d): Bookwork

For a wideband code division multiple access (WCDMA) system, a basestation typically transmits at carrier power levels to +46 dBm (40W).

[2]

Model answer to Q 1(e): Computed Example

Post-modulation gain = +46 dBm – (-10) dBm = 56 dB

Noise (dBm/1 MHz) = -150 dBm/Hz + 10 log(1 MHz) + 56 dB = -34 dBm

[3]

Model answer to Q 1(f): Computed Example

Post-modulation gain = +46 dBm – (-20 + 6) dBm = 60 dB

Noise (dBm/1 MHz) = -150 dBm/Hz + 10 log(1 MHz) + 60 dB = -30 dBm

[3]

Model answer to Q 1(g): Extension of Theory

From 1e) the output noise emission requirement is met with a safety margin of 4 dB, but the ACLR does not meet the specification by 2 dB. The solution is to decrease the modulator's output power by up to 4 dBm. This will increase the post-modulator gain and, therefore, the output noise emission, but by moving away from compression the ACLR should improve to an acceptable level.

From 1f) the output noise emission requirement is met (with no safety margin), but the ACLR easily meet the specification with a safety margin of 5 dB. The solution is to increase the modulator's output power by a couple of dBm. This will decrease the post-modulator gain and, therefore, provide a safety margin for the output noise emission. By moving towards compression the ACLR will degrade, but it should be kept within acceptable limits.

[3]

Model answer to Q 2(a): Bookwork

At the target the power density is given by:

$$PD_{TARGET} = \left(\frac{P_T}{4\pi R_T^2}\right) G_T \quad [W/m^2]$$

Power captured and reradiated isotropically by the target, $P_{TARGET} = \sigma PD_{TARGET} \quad [W]$

At the receiver the power density is given by:

$$PD_{RX} = \left(\frac{P_{TARGET}}{4\pi R_T^2}\right) \quad [W/m^2]$$

Power at the input to the receiver:

$$P_R = A_{RX} PD_{TXT} \quad [W]$$

The effective aperture of the receiving antenna, $A_{RX} = \dfrac{\lambda_o^2}{4\pi} D_o \quad [m^2]$

Where $G_R = \eta Do$ and with a lossless antenna its efficiency, $\eta = 100\%$

$$\therefore P_R = \left(\frac{\lambda_o^2}{4\pi}\right) G_R \left(\frac{1}{4\pi R_R^2}\right) \sigma \left(\frac{P_T}{4\pi R_T^2}\right) G_T$$

$$\therefore \left(\frac{P_R}{P_T}\right) = \left(\frac{\lambda_o}{4\pi R_T R_R}\right)^2 \frac{G_T \sigma G_R}{4\pi} \quad \text{Radar Range Equation}$$

When the transmitter and receiver share a common antenna:

$$\left(\frac{P_R}{P_T}\right) \rightarrow \left(\frac{\lambda_o G}{4\pi R^2}\right)^2 \frac{\sigma}{4\pi}$$

[5]

Model answer to Q 2(b)(i): Computed Example

The effective aperture for an ideal lossless paraboloidal reflector antenna is given by:

$$A = \frac{\pi D^2}{4} = 63.62 \quad m^2$$

$$\Omega = \frac{\lambda_o^2}{A} = 7.6x10^{-3} \quad with \quad \lambda_o = 0.694\, m$$

$$Directivity,\ D_o = \frac{4\pi}{\Omega} = 1{,}658$$

$$Power\ Gain,\ G \rightarrow D_o = 32.2\ dBi$$

[4]

Model answer to Q 2(b)(ii): Computed Example

With a pencil beam radiation pattern (i.e. having a large aperture and low sidelobes): $\Omega \approx \theta_E \theta_H$, Where θ_E and θ_H are the -3 dB beamwidths in the E- and H-planes, respectively.

$$\therefore \theta \approx \sqrt{\Omega} = 0.087\ radians \equiv 5^o$$

To an observer on the earth, the moon subtends an angle of 2ϕ: where

$$2\phi = 2\tan^{-1}\left(\frac{3.5/2}{381.5 + 3.5/2}\right) = 0.523^o$$

Therefore, the beam efficiency will be very small because the angle subtended by the target is an order of magnitude smaller than the beamwidth of the antenna.

[2]

Model answer to Q 2(b)(iii): Computed Example

If we assume that the moon's radar cross-section can be modelled as a perfectly reflecting flat circular disc:

$$\sigma = \frac{\pi D_M^2}{4} = 9.62x10^{12} \quad m^2$$

Therefore, the power at the receiver is given by:

$$P_R = \left(\frac{\lambda_o D_o}{4\pi R^2}\right)^2 \frac{\sigma P_T}{4\pi} = 3x10^{-17} \quad W$$

[3]

Model answer to Q 2(b)(iv): Computed Example

System equivalent noise temperature at the receiver, $T_S = T_A + T_{RX} = 175$ K

System Input Noise Power, $N = k\, T_S\, B = 1.69x10^{-17}$ W

Therefore, the carrier-to-noise power ratio at a receiver, $C/N = P_R/N = 1.775 = 2.5$ dB

[3]

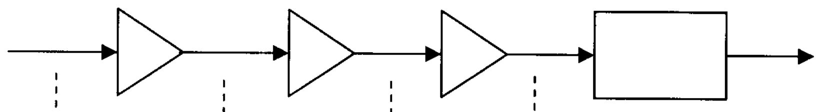
Model answer to Q 2(b)(v): Computed Example

If both antennas suffer from unwanted sidelobes then it is theoretically possible that the sensitive receiver could detect the high power transmitter through unwanted coupling of the sidelobes of both antennas. In this case, the minimum possible delay will be given from:

Delay = separation distance between earth stations / speed of light = 3.33 microseconds.

[3]

Model answer to Q 3: Computed Example



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]

Using the following general equations:

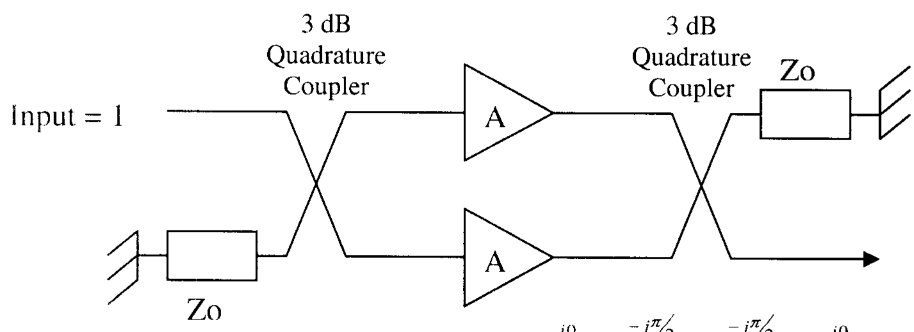
$$C = P_{OUT}(fo) = G_1 G_2 P_{IN} \quad and \quad IP_3 = \left(IP_3|_1\, G_2\right) \| IP_3|_2$$

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

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

Model answer to Q 4(a): Computed Example

(i) For a single-balanced Amplifier:



$$\text{Output} = \frac{e^{j0}}{\sqrt{2}} A \frac{e^{-j\pi/2}}{\sqrt{2}} + \frac{e^{-j\pi/2}}{\sqrt{2}} A \frac{e^{j0}}{\sqrt{2}} = Ae^{-j\pi/2}$$

Under normal operation:

$$S_{21} = Ae^{-j\pi/2} \qquad Power\ Gain = |S_{21}|^2 = A^2$$

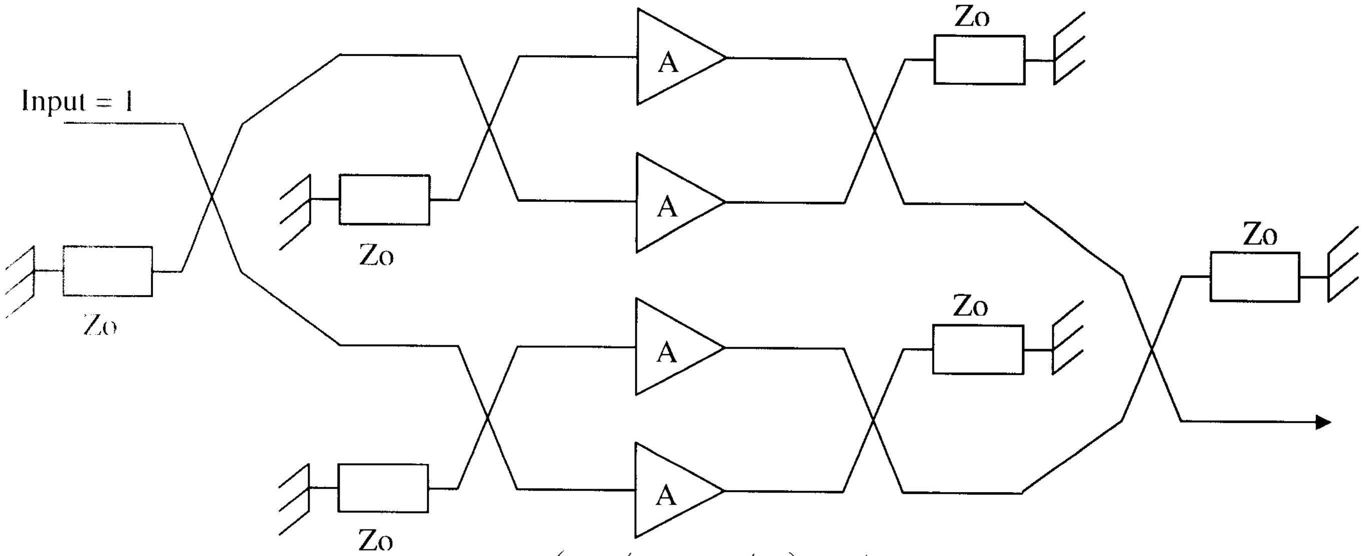
If one of the transistor/amplifiers fails:

$$S_{21} = 0 + \frac{e^{-j\pi/2}}{\sqrt{2}} A \frac{1}{\sqrt{2}} = \frac{A}{2} e^{-j\pi/2} \qquad Power\ Gain = |S_{21}|^2 = \frac{A^2}{4}$$

Therefore, there is a drop in power gain by 6 dB. Also, since the power from one single-ended amplifier is lost there is a 3 dB drop in output power. However, the power from the functioning single-ended amplifier is fed into the output 3 dB quadrature coupler. This results in another 3 dB power loss. Therefore, the overall output power lost is 6 dB.

[4]

(ii) For a double-balanced Amplifier:



$$\text{Output} = \left(\frac{e^{-j\pi/2}}{2\sqrt{2}}A + \frac{e^{-j\pi/2}}{2\sqrt{2}}A\right)\frac{e^{-j\pi/2}}{\sqrt{2}} + \left(\frac{e^{-j\pi}}{2\sqrt{2}}A + \frac{e^{-j\pi}}{2\sqrt{2}}A\right)\frac{e^{-j0}}{\sqrt{2}} = Ae^{-j\pi}$$

Under normal operation:

$S_{21} = Ae^{-j\pi}$ $\quad$ $Power\ Gain = |S_{21}|^2 = A^2$

If one of the transistor/amplifiers fails:

$S_{21} = 0 + A\frac{e^{-j\pi}}{4} + A\frac{e^{-j\pi}}{4} + A\frac{e^{-j\pi}}{4} = \frac{3e^{-j\pi}}{4}A$ $\quad$ $Power\ Gain = |S_{21}|^2 = \frac{9A^2}{16}$

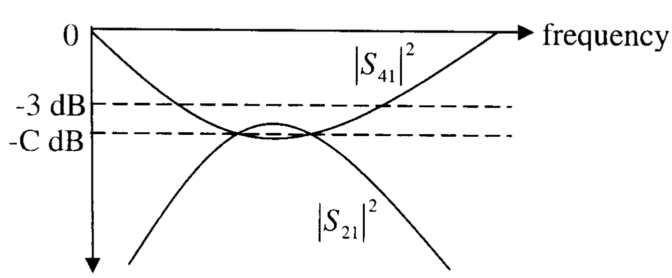
Therefore, there is a drop in power gain by 2.5 dB. Thus, the overall output power lost is 2.5 dB.

[4]

Model answer to Q 4(b): Computed Example

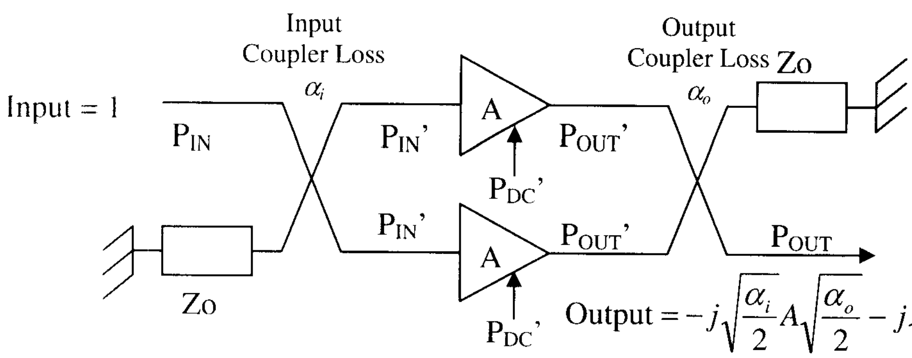
For a 4-port coupler, Insertion Loss, $\alpha = |S_{21}|^2 + |S_{41}|^2$

A 3 dB coupler is regarded as having equal power at both the coupled and direct ports,
i.e. $|S_{21}|^2 = |S_{41}|^2 \equiv C^2 \quad \therefore \alpha = 2C^2 \quad \Rightarrow C\,[dB] + 3\,dB$



[4]

Model answer to Q 4(c): Computed Example



$$\text{Output} = -j\sqrt{\frac{\alpha_i}{2}}A\sqrt{\frac{\alpha_o}{2}} - j\sqrt{\frac{\alpha_i}{2}}A\sqrt{\frac{\alpha_o}{2}} = -j\sqrt{\alpha_i}A\sqrt{\alpha_o}$$

$$Power\ Gain, G = |S_{21}|^2 = \alpha_i A^2 \alpha_o \quad \Rightarrow \quad \alpha_i[dB] + 20\log\{A\}[dB] + \alpha_o[dB]$$

$$With\ lossless\ couplers, G' = |A|^2$$

$$P_{IN}' = \alpha_i \frac{P_{IN}}{2} \quad ; \quad P_{OUT}' = G'P_{IN}' \quad ; \quad P_{OUT} = 2\alpha_o P_{OUT}'$$

$$\therefore G = \frac{P_{OUT}}{P_{IN}} = \frac{2\alpha_o P_{OUT}'}{2 P_{IN}'/\alpha_i} = \alpha_i G' \alpha_o$$

$$\eta' = \frac{P_{OUT}'}{P_{DC}'} \quad ; \quad \eta_{ADD} = \frac{P_{OUT} - P_{IN}}{P_{DC}} = \frac{2\alpha_o P_{OUT}' - 2 P_{IN}'/\alpha_i}{2P_{DC}'} = \eta'\left(\alpha_o - \frac{1}{\alpha_i G'}\right)$$

[4]

Model answer to Q 4(d): Computed Example

G' = 10 dB = 10; η' = 25%; $\alpha_i = \alpha_o$ = -0.5 dB = 0.89125

Therefore, G = 7.943 = 9 dB and η_{ADD} = η'(0.779) = 19.5%

[4]

Model answer to Q 5(a): Computed Example

Loaded-Q factor for the resonant circuit is simply:

Q_L = (f_o = 1.8 GHz)/(BW_{3dB} = 200 MHz) = 9

[2]

Model answer to Q 5(b): Computed Example

Loaded-Q factor for the resonant *R-L-C* resonator is:

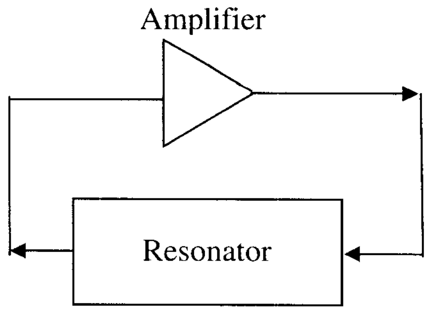
$Q_L = (\omega L)/(R_{LOSS} + 2\,Z_o) = 9$

Unloaded-Q factor for the resonant *R-L-C* resonator is:

$Q_U = (\omega L)/(R_{LOSS}) = 905$

[3]

Model answer to Q 5(c): Bookwork



For oscillation to occur, the closed loop gain must be greater than unity and the total phase around the loop must be an integer multiple of 360°.

[2]

Model answer to Q 5(d): Computed Example

For an oscillator that is designed around a series *L-C-R* resonator, a criteria for minimum phase noise, in terms of the quality factor of the resonator, is that the loaded-Q/unloaded-Q ratio be equal to 0.5.

$Q_L/Q_U = R_{LOSS} / (R_{LOSS} + 2\, Z_{LOAD}) = 0.5$

Therefore, the optimum value of $Z_{LOAD} = 0.5\ \Omega$

$$|S_{21}(\omega_o)| = \left(1 - \frac{Q_L}{Q_o}\right) \equiv 0.5 \quad \therefore Insertion\ Loss = 6\ dB$$

[4]

Model answer to Q 5(e): Computed Example and Bookwork

For the identical *L-C* impedance transformers:

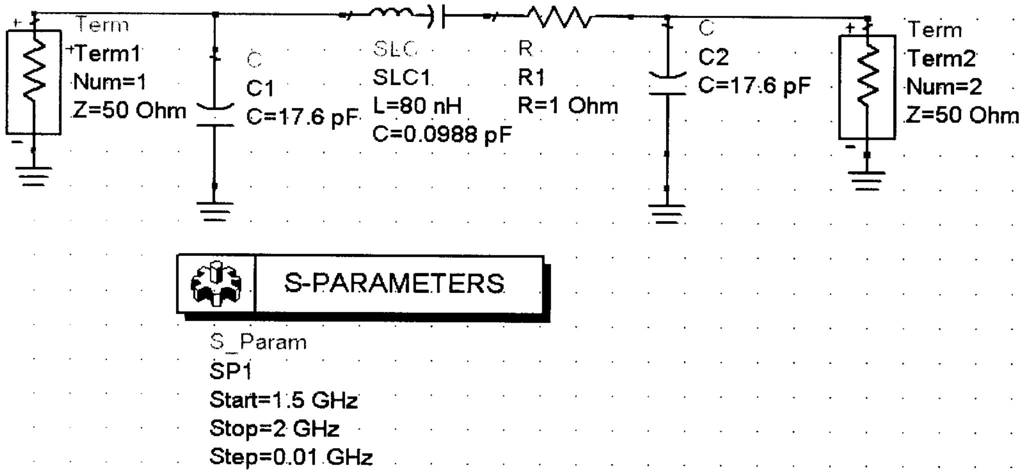
$$Q_S = Q_P = \sqrt{\left(\frac{R_P = 50}{R_S = 0.5} - 1\right)} = 9.95$$

$$Q_S = \frac{|X_S|}{R_S} = \frac{\omega L_s}{R_S} \quad \therefore L_s = 0.44\ nH$$

$$Q_P = \frac{R_P}{|X_P|} = \omega C_s R_P \quad \therefore C_P = 17.6\ pF$$

Therefore, with a total series inductance of 80 nH, the resonator's inductance will be (80 – 2 x 0.44) = 79.12 nH. For resonance, the corresponding series capacitance will be:

$$C = \frac{1}{(79.12\ nH)\,\omega^2} = 98.8\ fF$$



[3]

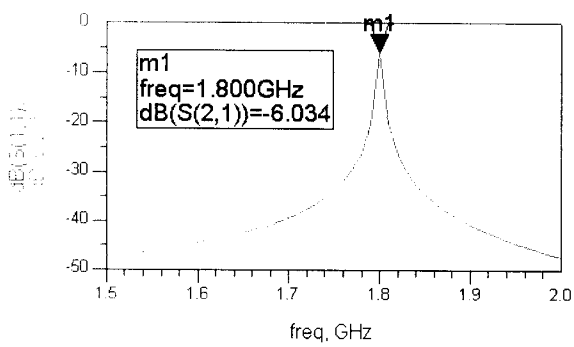
The series inductance values of the impedance transformers are very low, when compared to the value for the resonator. Also the series capacitance for the resonator is very low when compared to the shunt capacitances of the impedance transformers. This is because of the very large impedance transformation ratio of 100. In practice, it would be very difficult to realise a 98.8 fF capacitance.

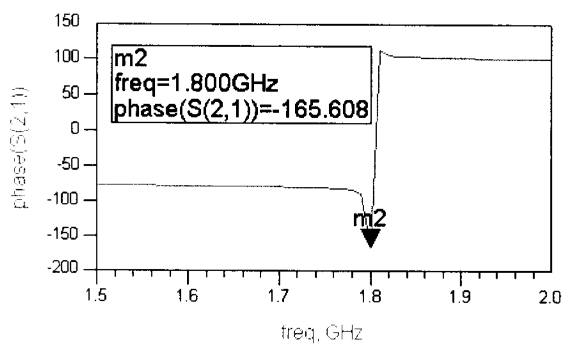
[2]

At resonance, the insertion phase should be -180°. With the inverting amplifier, this will give a loop insertion phase of 360°, which meets one of the two main conditions for oscillation.

[1]

The minimum theoretical power gain of the inverting amplifier should be 6 dB at resonance, in order to meet the other condition for oscillation.





[1]

Model answer to Q 5(c): Bookwork

Silicon transistors have much lower flicker (1/f) noise than GaAs transistors. One of the main reasons is because the silicon nitride passivation layer results in fewer trapped energy states. Therefore, silicon transistors can be used to realise oscillators that have much lower phase noise characteristics.

[2]

Model answer to Q 6(a): Bookwork

Designing high-performance narrow fractional bandwidth filters is more difficult than designing for wider bandwidths. The reason is that more components are needed to get steeper roll-off characteristics. As a result, energy stays within the filter for longer and so the group delay of the filter increases. Moreover, the longer the energy stays within the filter the more of it will be dissipated as unwanted heat and so the pass-band attenuation increases. For this reason, the components must increase in size, in order to reduce current densities and, thus, reduce the associated wasted power dissipated within the components' materials. Therefore, using passive realization technologies, narrow fractional bandwidth filters have to be relatively large.

[4]

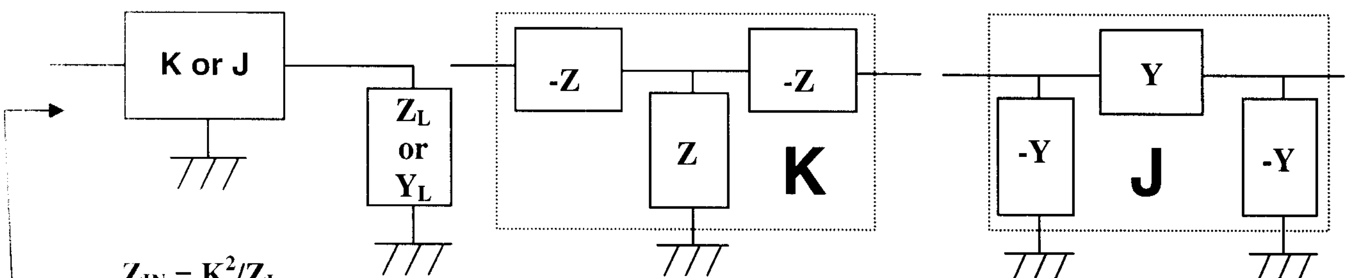
Model answer to Q 6(b): Bookwork

In addition to size, as the fractional bandwidth of a filter gets smaller, the required range of L and C components values increases. In practice, it may not be possible to achieve such a high range from a discrete component. For this reason, impedance and/or admittance inverters are used to convert the component values that are available into those that are not normally available. It has already been shown that a $\lambda g/4$ section of transmission line can be used to perform impedance inversion:

$\mathbf{Z_{IN} = K^2/Z_L}$ Impedance inversion constant, K = Zo with $\lambda g/4$ transmission line

$\mathbf{Y_{IN} = J^2/Y_L}$ Admittance inversion constant, J = Yo with $\lambda g/4$ transmission line

In addition, lumped elements can be used (i.e. either all inductive or all capacitive), as shown below. Note that negative reactances or susceptances are meant to be absorbed by neighbouring positive reactances or susceptances.



$\mathbf{Z_{IN} = K^2/Z_L}$
$\mathbf{Y_{IN} = J^2/Y_L}$

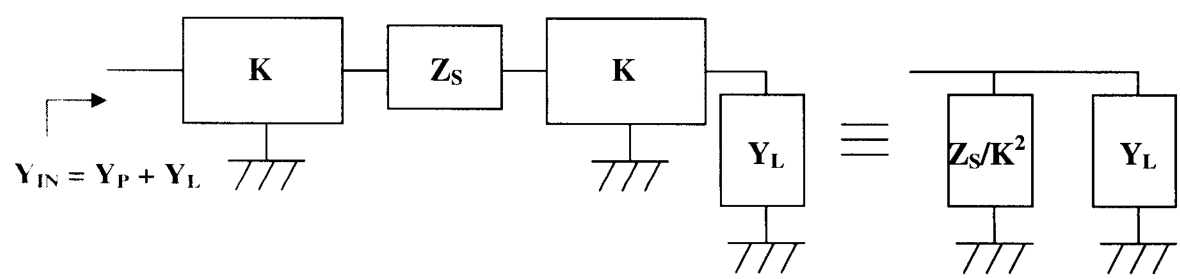
$\mathbf{K = \omega L \text{ or } 1/\omega C}$
$\mathbf{\therefore K^2 = -Z^2}$
$\mathbf{Z_{IN} = -Z^2/Z_L}$

$\mathbf{J = \omega C \text{ or } 1/\omega L}$
$\mathbf{\therefore J^2 = -Y^2}$
$\mathbf{Y_{IN} = -Y^2/Y_L}$

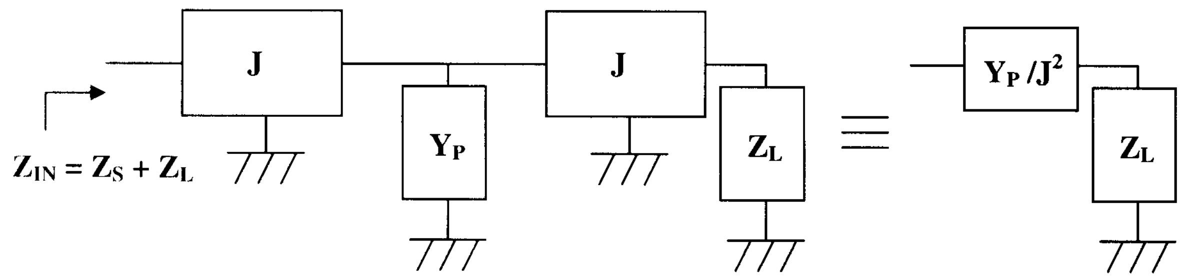
Impedance/Admittance Inverter | T-Impedance network (K-Inverter) | π-Admittance network (J-Inverter)

- Two identical inverters connected in cascade represents a zero inversion
- A series (shunt) element placed between two identical inverters appears as a shunt (series) element

As an example, a shunt connected parallel L-C tuned circuit can be "synthesized" from a series connected series L-C tuned circuit using two impedance inverters.



As an example, a series connected series L-C tuned circuit can be "synthesized" from a shunt connected parallel L-C tuned circuit using two admittance inverters.



[6]

Model answer to Q 6(c): Computed Example

For a series *R-L-C* circuit the corresponding impedance is:

$$Z_S = R_S + j\omega L_S + \frac{1}{j\omega C_S}$$

This can be converted into a shunted parallel R-L-C tuned circuit by using J-inverters. The corresponding admittance is:

$$Z_P = G_P + j\omega C_P + \frac{1}{j\omega L_P} \equiv \frac{J^2}{Y_S} = Z_S J^2$$

Using discrete capacitance values to realise an admittance inverter with a –C/+C/–C π-network:

$$J = \omega C$$

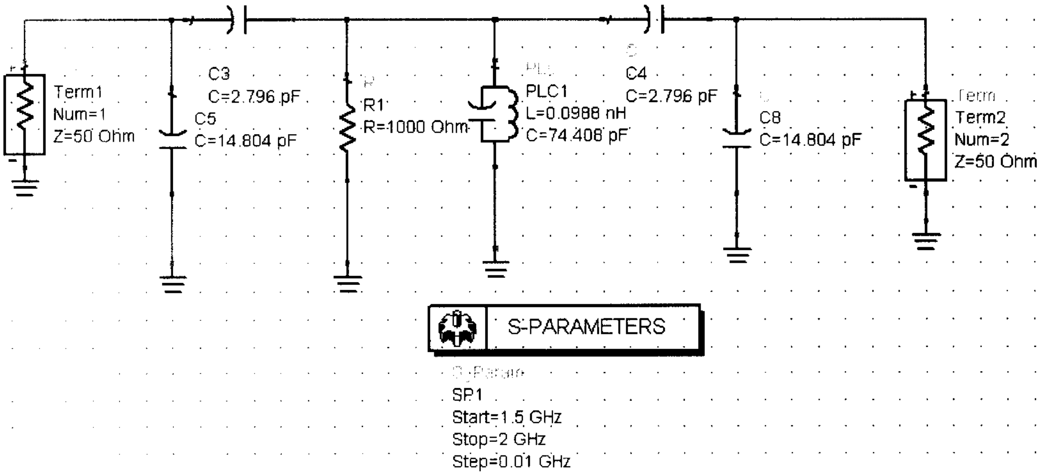
$$Y_P = R_S(\omega C)^2 + j\omega L_S(\omega C)^2 + \frac{(\omega C)^2}{j\omega C_S}$$

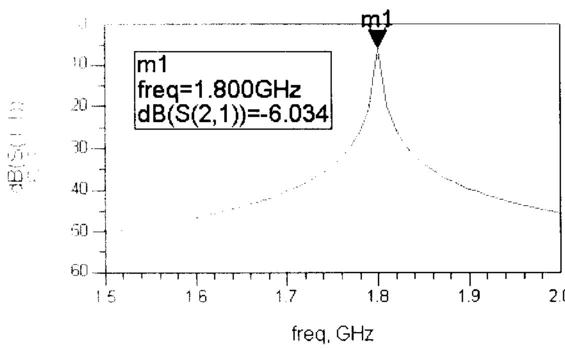
$$\therefore R_P = \frac{1}{R_S(\omega C)^2} \quad ; \quad C_P = L_S(\omega C)^2 \quad ; \quad L_P = \frac{C_S}{(\omega C)^2}$$

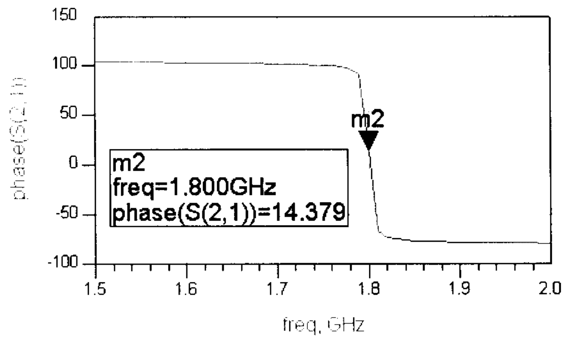
If we choose:

$$(\omega C)^2 = 1x10^{-3} \quad \therefore C = 2.796\ pF \quad for\ f = 1.8\ GHz$$

$$\therefore R_P = 1{,}000\ \Omega \quad ; \quad C_P = 80\ pF \quad ; \quad L_P = 0.0988\ nH$$







[8]

Model answer to Q 6(d): Extended Theory

Each J-inverter introduces 90° of insertion phase and, therefore, a pair of J-inverters will introduce 180° of insertion phase.

[2]