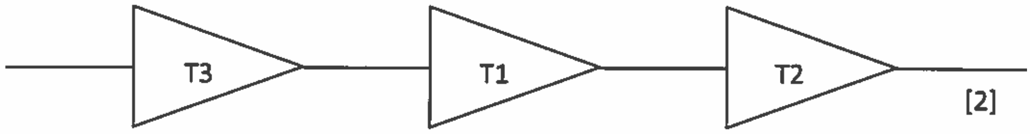
The Solutions for EE4.18 and AO6, 2016

<u>Model answer to Q 1(a): Computed Example</u>



[2]

	T3	T1	T2
i) Power Gain [dB]	15	10	5

Overall Power Gain = G1+G2+G3 = 30 dB

[1]

ii) $P_{OUT}|_{MAXLIN}$ [dBm] +15 = 31.6 mW +25 = 316 mW +30 = 1000 mW

Final Stage $P_{OUT}|_{MAXLIN}$ = +30 dBm

[2]

iii) Basic Efficiency [%] 52.7 52.7 50

Overall Basic Efficiency = Final Stage $P_{OUT}|_{MAXLIN}$/ Sum(P_{DC}) = 1 W / 2.66 W = 37.59 %

[2]

iv) PAE [%] 51.0 47.4 34

Overall PAE = Overall Basic Efficiency *(1 – 1/(Overall Power Gain)) = 37.56

[2]

v) IP_3 [dBm] 40 39.59 38.7

Final Stage IP_3 = 38.7 dBm

[2]

vi) IMD_3 [dBc] 50 29 17.4

Final Stage IMD_3 = 2*(Final Stage IP_3 - Final Stage $P_{OUT}|_{MAXLIN}$) = 17.4 dBc

[2]

vii) P_{DISS} [mW] 29.4 315.6 1316

Overall P_{DISS} = 1.661 W = 32.2 dBm

[2]

<u>Model answer to Q 1(b): Textbook Derivation</u>

$$IMD_3 = \frac{C}{I_3} \quad and \quad IMD_3[dBc] \approx 2 \cdot (IP_3 - C)$$

$$\therefore I_3 \sim \frac{C^3}{IP_3^{\,2}} \quad \therefore \frac{\partial I_3[dBm]}{\partial Pin[dBm]} = 3\frac{\partial C[dBm]}{\partial Pin[dBm]} - 2\frac{\partial IP_3[dBm]}{\partial Pin[dBm]} \quad where, \frac{\partial IP_3[dBm]}{\partial Pin[dBm]} = 0$$

In other words, the third – order intermodulation log power gain slope is three times that of the desired output log power gain slope. Therefore, IMD_3 improves rapidly as input power decreases.

[2]

Model answer to Q 1(c): Computed Example

If the input power drops by 3 dB:

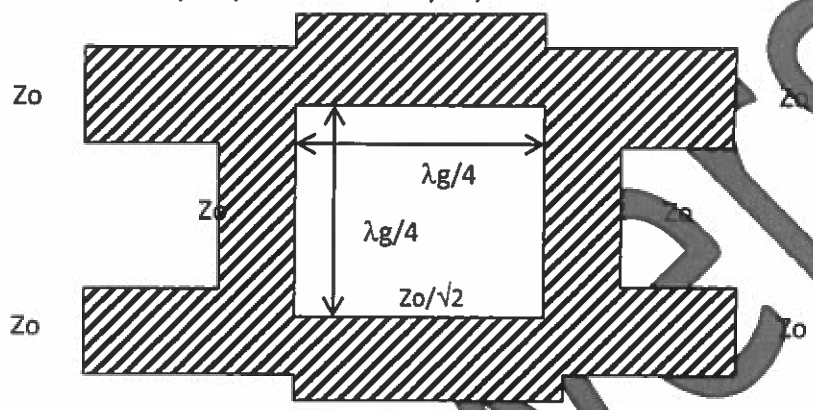
i) P_{OUT} drops by 3 dB [1]

ii) I_3 drops by 9 dB [1]

iii) IMD_3 increases by 6 dB [1]

Model answer to Q 2(a): Bookwork

90° 3dB Directional Coupler (Branch-line Coupler)

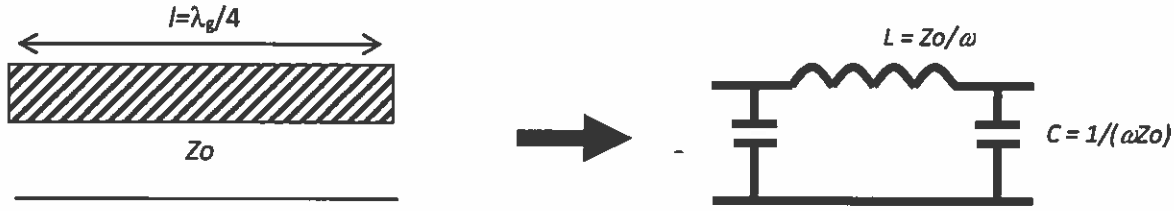


- Works on the interference principle, therefore, narrow fractional bandwidth (15% maximum)
- No bond-wires or isolation resistors required
- Wider tracks make it easier to fabricate and is, therefore, good for lower loss and higher power applications
- Simple design but large
- Meandered lines are possible for lower frequency applications

[5]

Model answer to Q 2(b): Bookwork and Computed Example

The lumped-element equivalent of a $\lambda_g/4$ transmission line is shown below.



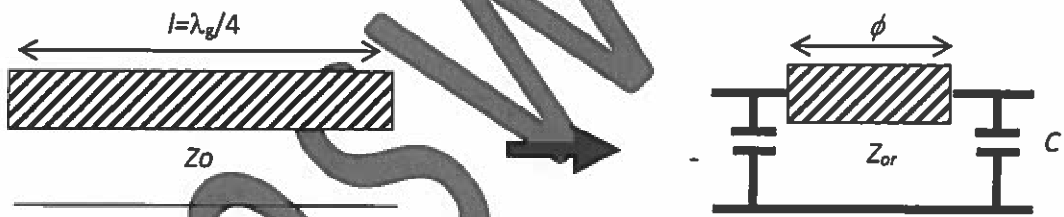
All the previous distributed-element couplers can be transformed into equivalent lumped-element couplers by simply replacing all the $\lambda_g/4$ lengths of transmission lines with the above π-network. Since lumped-element components have a lower Q-factor, when compared to distributed-element components, there is an insertion loss penalty. Also, because this π-network is clearly a low-pass filter, having a cut-off frequency, $f_c = \dfrac{1}{2\pi\sqrt{LC}}$, there is also a bandwidth penalty.

L = 4.42 nH and C = 1.77 pF for the Zo = 50 Ω sections of line
L = 3.13 nH and C = 2.50 pF for the Zo = 35.36 Ω sections of line

[5]

Model answer to Q 2(c): Bookwork and Computed Example

- Lumped-Distributed Couplers



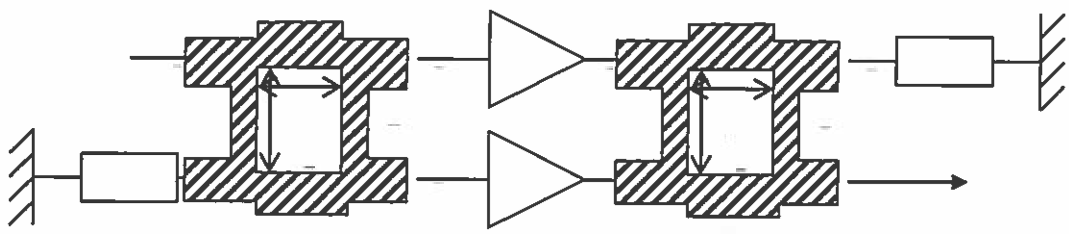
In this 'reduced-size' technique, each $\lambda_g/4$ line is replaced with the above π-network.

$$Z_{or} = \frac{Zo}{Sin\phi} \quad and \quad C = \frac{Cos\phi}{\omega Zo}$$

With ϕ = 45°,
Z_{or} = 70.7 Ω and C = 1.25 pF for the Zo = 50 Ω sections of line
Z_{or} = 50 Ω and C = 1.77 pF for the Zo = 35.36 Ω sections of line

[5]

Model answer to Q 2(d): Solution given in class



[5]

Model answer to Q 3(a): Bookwork

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 3(b): Bookwork

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 3(c): Computed example

(i) An inductor is connected in series with the output of the transistor. If it has an inductance of L = +7/(2πfo) = 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 √(5*50) = 15.81 Ω, to be inserted in series with the inductor.

[4]

(ii) An inductor is connected in series with the output of the transistor. If it has an inductance of L = +7/(2πfo) = 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 3(d): Bookwork

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 4(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]

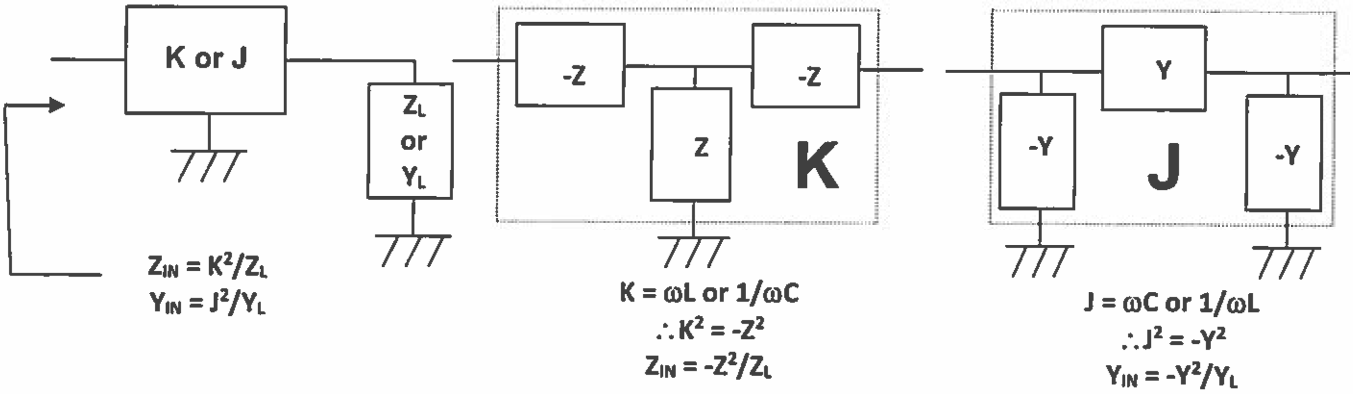
Model answer to Q 4(b): Bookwork

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 λg/4 section of transmission line can be used to perform impedance inversion:

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

$Y_{IN} = J^2/Y_L$ Admittance inversion constant, J = Yo with λg/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.



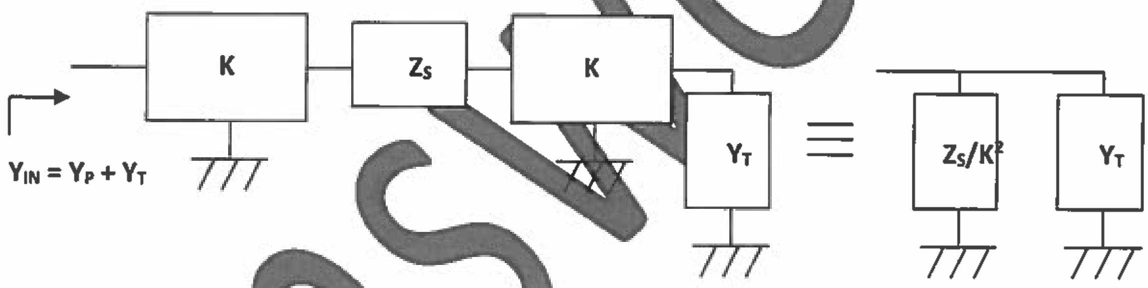
Impedance/Admittance Inverter

T-Impedance network (K-Inverter)

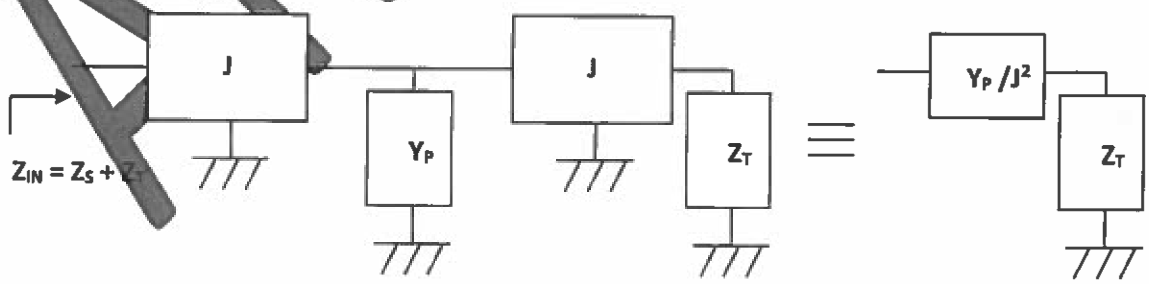
π-Admittance network (J-Inverter)

- 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 having 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 having two admittance inverters.



[6]

<u>Model answer to Q 4(c): Computed Example</u>

A series *R-L-C* circuit has the following impedance:

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

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

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

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

$$J = \omega C$$

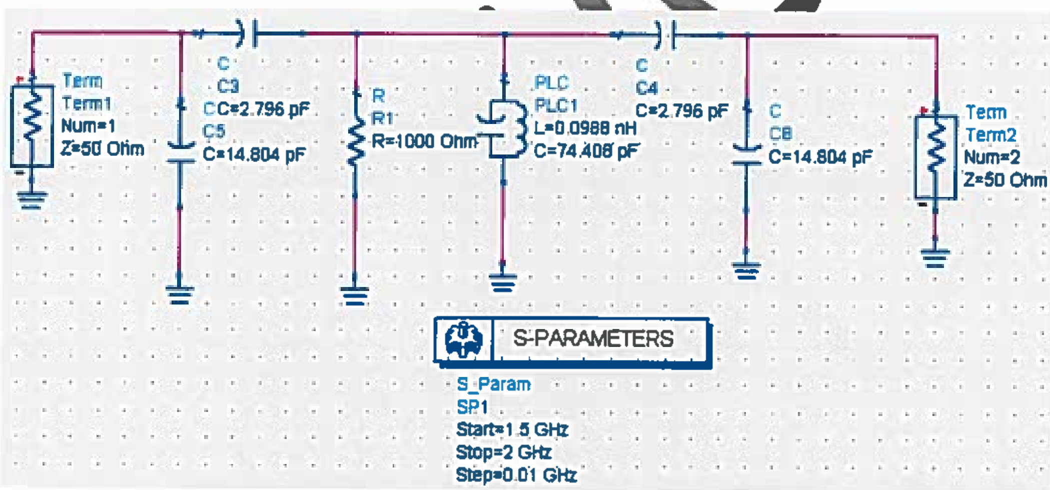
$$Y_P = R_S(\omega C)^2 + j\omega L_S(\omega C)^2 + \frac{(\omega C)^2}{j\omega C_S}$$

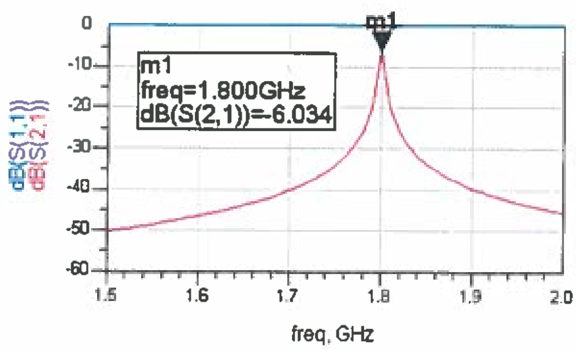
$$\therefore R_P = \frac{1}{R_S(\omega C)^2} \quad ; \quad C_P = L_S(\omega C)^2 \quad ; \quad L_P = \frac{C_S}{(\omega C)^2}$$

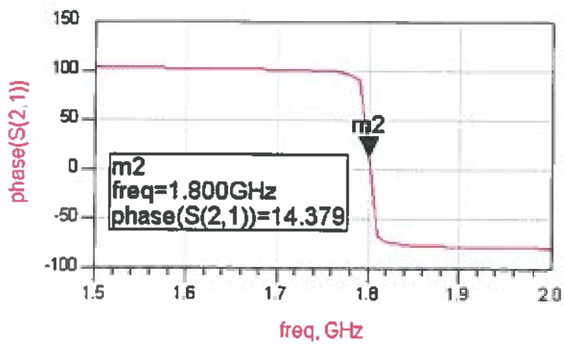
If we choose:

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

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







[8]

Model answer to Q 4(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]

Model answer to Q 5(a): New application of theory

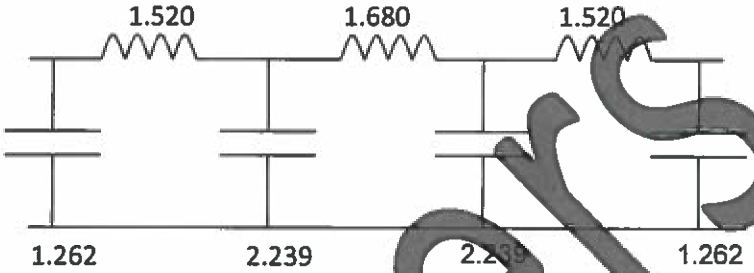
Given a square wave signal generator, with a clock frequency that can vary between DC and 1000 MHz, set the output to 600 MHz. Using a band-pass filter having a centre-frequency set to the third harmonic of the signal generator, the desired 1800 MHz sinusoidal signal can be extracted. Alternatively, a high-pass filter can be used, however, the 5th and 7th etc., harmonics will also be present but at much lower power levels.

[4]

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

From graphs provided, for Chebyshev filters with a 0.1 dB ripple, a 7th order filter is required to achieve an out-of-band rejection of > 70 dB with an fc/f ratio of 1500/600 = 2.5.

From tables provided, with $R_S = R_L = Zo = 50$, the prototype low-pass filter and associated coefficients are given below:



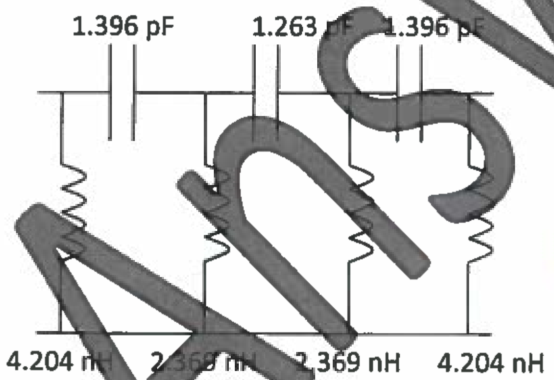
High-pass de-normalising:

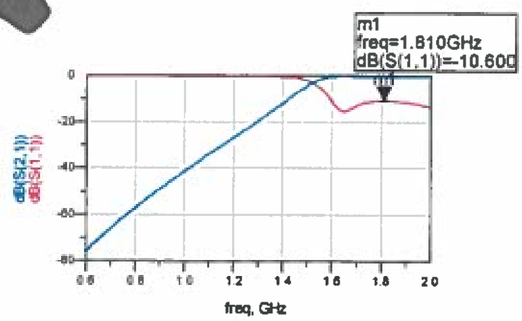
Shunt inductor:

$$L_P = \frac{R_L}{2\pi f_c L n}$$

Series Capacitor:

$$C_S = \frac{1}{2\pi f_c C n R_L}$$





The slight deviation in the results from those predicted by theory are due to limited component tolerances used within the simulations.

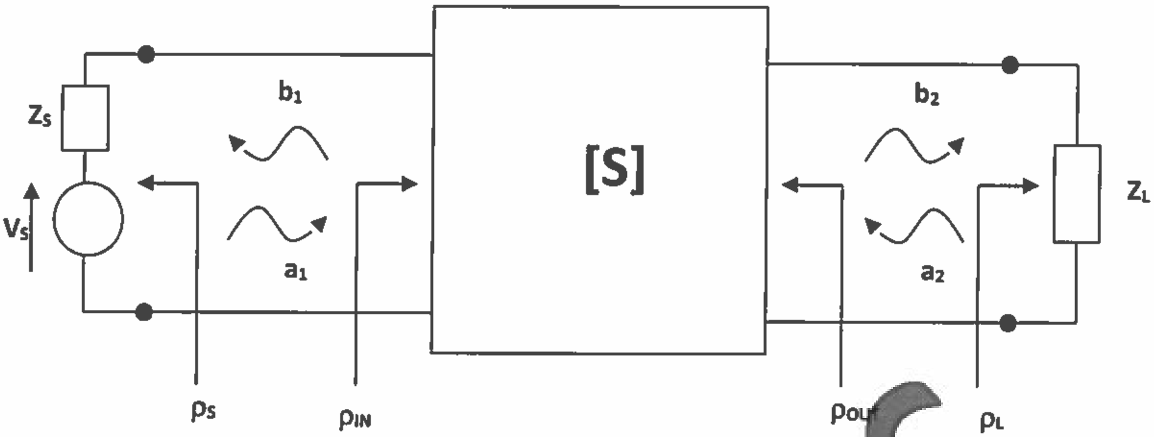
[10]

Model answer to Q 5(c): New application of theory

The worst-case level of return loss within the pass band is 10 log[1- antilog(-0.1/10)] = -16.43 dB. This will have no effect within the pass band. However, well-below the pass band the return loss will be 0 dB. At the fundamental frequency of 600 MHz, all the signal will be reflected back by the filter and the square wave signal generator may not work properly (if at all). One possible solution is to insert a circulator between the square-wave generator and the filter, however, passive circulators tend to be narrow band in nature. However, a more practical solution is to use a balance topology for the filters, whereby two identical filters are embedded between two identical ultra-broadband 3 dB quadrature directional couples. This approach is inherently impedance matched across the bandwidth of the couplers.

[6]

Model answer to Q 6(a): Bookwork



$$\textit{Incident Wave, } a=\frac{V_+}{\sqrt{Zo}}=\frac{1}{2}\left(\frac{V(0)}{\sqrt{Zo}}+I(0)\sqrt{Zo}\right)=+I_+\sqrt{Zo}$$

$$\textit{Reflected Wave, } b=\frac{V_-}{\sqrt{Zo}}=\frac{1}{2}\left(\frac{V(0)}{\sqrt{Zo}}-I(0)\sqrt{Zo}\right)=-I_-\sqrt{Zo}$$

$$\textit{therefore,}\quad V(0)=\sqrt{Zo}(a+b)=(V_++V_-)\quad \textit{and}\quad I(0)=\frac{(a-b)}{\sqrt{Zo}}\equiv(I_++I_-)$$

$$\textit{Power Incident At Port, } P_+=|a|^2$$

$$\textit{Power Reflected By Port, } P_-=|b|^2=|\rho|^2|a|^2$$

$$\textit{Power Delivered to Port Termination Impedance, } P=(P_+-P_-)=\left(|a|^2-|b|^2\right)=|a|^2\left(1-|\rho|^2\right)$$

$$\text{Input Voltage Wave Reflection Coefficient, } S_{11}=\left.\frac{b_1}{a_1}\right|_{a_2=0}$$

$$\text{Forward Voltage Wave Transmission Coefficient, } S_{21}=\left.\frac{b_2}{a_1}\right|_{a_2=0}$$

$$\text{Insertion Phase}=\forall S_{21}$$

Reverse Voltage Wave Transmission Coefficient, $\mathbf{S_{12}} = \left.\dfrac{b_1}{a_2}\right|_{a_1=0}$

Output Voltage Wave Reflection Coefficient, $\mathbf{S_{22}} = \left.\dfrac{b_2}{a_2}\right|_{a_1=0}$

$$\textit{Input Return Loss, } RL_{IN} = 10\log|S_{11}|^2$$

$$\left.\begin{array}{r}\textit{Insertion Power Gain (Loss), } G_I \\ \textit{Forward Transducer Power Gain (Loss), } G_{FT}\end{array}\right\} = 10\log|S_{21}|^2 \quad \textit{when } Z_S = Z_L = Zo$$

$$\left.\begin{array}{r}\textit{Reverse Power Isolation, IS} \\ \textit{Reverse Transducer Power Gain (Loss), } G_{RT}\end{array}\right\} = 10\log|S_{12}|^2 \quad \textit{when } Z_S = Z_L = Zo$$

$$\textit{Output Return Loss, } RL_{OUT} = 10\log|S_{22}|^2$$

[5]

<ins>Model answer to Q 6(b): Computed Example</ins>

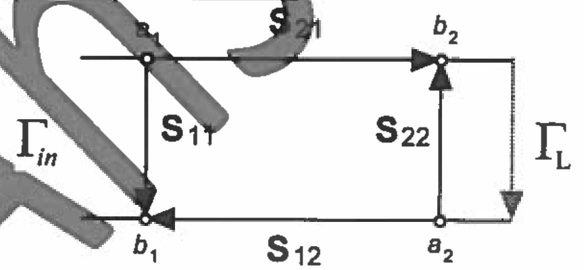
(i) This is a lossless transmission line of length 2λg, with perfect impedance matching to the reference impedance.

(ii) This is an isolator. The insertion loss in the forward direction is 0 dB and 23.1 dB in the reverse direction. Again, perfect port impedance matching is achieved.

(iii) This is an amplifier with forward power gain of 19.7 dB, reverse isolation of 10.45 dB. If the output is matched, the input return loss is 20 dB. If the input is matched, the output return loss is 16.5 dB.

[5]

<ins>Model answer to Q 6(c): Textbook Derivation</ins>



$$b_1 = S_{11}a_1 + S_{12}(\Gamma_L b_2)$$
$$b_2 = S_{21}a_1 + S_{22}(\Gamma_L b_2)$$
$$\therefore b_2 = (1 - \Gamma_L S_{22}) = S_{21}a_1$$
$$\therefore \Gamma_{in} = S_{11} + \frac{S_{21}S_{12}\Gamma_L}{1 - S_{22}\Gamma_L}$$

[6]

<ins>Model answer to Q 6(d): Textbook Example</ins>

If $|\Gamma_{in}| \le 1$, the circuit will be unconditionally stable. Since $S_{11} = S_{22} = 0$, $|\Gamma_{in}| \le S_{21}S_{12} = 0.07e^{-j\pi/2}$ and $|\Gamma_{in}| = 0.07/2 << 1$, i.e. unconditionally stable!

[4]