IMPERIAL COLLEGE LONDON

EE4-18
EE9AO6

DEPARTMENT OF ELECTRICAL AND ELECTRONIC ENGINEERING
EXAMINATIONS 2012

MSc and EEE PART IV: MEng and ACGI

Corrected Copy

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RADIO FREQUENCY ELECTRONICS

Tuesday, 15 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) : A.S. Holmes

Special instructions for invigilators

This is a closed book examination.

A Smith chart is to be distributed.

Special instructions for students

This is a closed book examination.

Filter curves and filter tables are attached at the back of this examination paper

A Smith chart is to be distributed.

$$\textit{Boltzmann constant, } k = 1.3806488 \times 10^{-23}\ [J/K]$$

The Questions

1.

a) Draw the block diagram for a single-conversion superhet receiver and, stating its full name, explain the origin of this type of architecture. [2]

b) For the receiver architecture in 1(a):

i) With the help of a spectral illustration, briefly explain why an image rejection filter is needed and where this should be located within the architecture. Also, briefly explain why this filter represents a problem and list 3 possible solutions that can be adopted to alleviate or avoid this problem. [3]

ii) For selecting bands with higher levels of adjacent-band rejection, what simple method is usually adopted? [1]

c) The receiver has an input power level of -50 dBm. With the simplest architecture design, for the type in 1(a) up to the demodulator, it can be assumed that there is only one filter and one amplifier at any given frequency. All amplifiers have a gain of 30 dB, IP_3 of 40 dBm and a noise figure of 1 dB; all filters have a pass band insertion loss of 2 dB and the mixer has a conversion loss of 3 dB and IP_3 of 30 dBm. Ignoring noise contributions from the local oscillator, calculate the following at the output of each sub-system block (where appropriate) and state the main equations used:-

i) C [2]

ii) IP_3 (clearly state any assumptions) [3]

iii) IMD_3 [1]

iv) I_3 [2]

v) Overall F_{RX} [3]

vi) Carrier-to-noise ratio at the output of a 30 kHz bandwidth receiver having its overall system temperature equal to the ambient temperature of 290 K. [3]

All variables have their usual meanings.

2. The Japan Aerospace Exploration Agency (JAXA) is proposing a Microwave-based Space Solar Power System (M-SSPS) for wireless energy transmission, whereby a satellite having large solar panel arrays generates dc power and most of this energy is transmitted to earth via a microwave link having a transmit power of 1 GW at 5.8 GHz.

a) If the geostationary satellite is positioned 36,000 km directly above the earth's 3 km diameter circular receiving target area, calculate the following:

i) The angle subtended by the receiving target at the satellite. [1]

ii) The optimum beam solid angle Ω_M for the satellite's antenna. *Hint: for large aperture antennas having a single pencil beam radiation pattern, this is given by $\Omega_M \approx \theta_E\ \theta_H$, where θ_E and θ_H are the -3 dB beam widths in the E- and H-planes, respectively.* Clearly state any simplifying assumptions. [1]

iii) The directivity of the satellite's antenna. [1]

iv) The ideal effective aperture area of the satellite's antenna. [2]

v) The ideal diameter of the satellite's antenna if it is to be implemented by a conventional circular parabolic reflecting antenna. [1]

b) The International Commission for Non-Ionizing Radiation Protection gives a limit for the maximum safe power density for human exposure of 1 mW/cm^2. Using appropriate calculations, comment on the safety levels at the:

i) Satellite. [3]

ii) Receiving target. [3]

c) If the receiving target area is optimally covered with half-wavelength by half-wavelength sized patch antennas, estimate the maximum number of patch antennas that can fit into the target array. *Hint: convert the circular array into a square array.* [2]

d) Assuming the antennas in 2(c) are 75% efficient, calculate the output power from each. [2]

e) The power from 2(d) is rectified to produce dc power with an efficiency of 40%. Calculate the output dc power from each of the rectifying antennas (know as rectennas) and then calculate the total dc power output form the receiver's target area. [1]

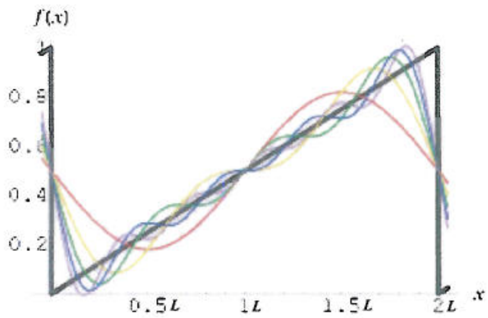
f) The satellite's power combing and antenna subsystem is also 75% efficient. If 2.5 W output transistors are employed, each having an overall dc-to-RF conversion efficiency of 70%, calculate:

i) The total number of transistors needed. [1]

ii) The dc power budget needed to be generated by the solar panel array. [1]

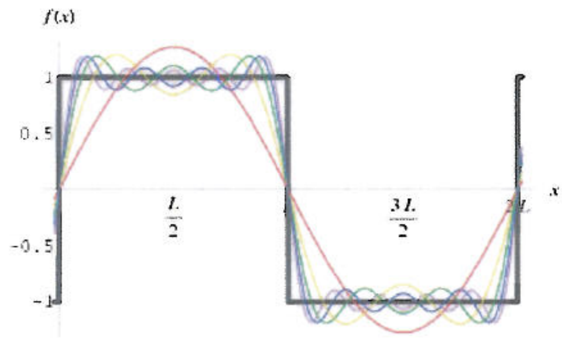
iii) The overall end-to-end dc-to-dc efficiency for the complete system. [1]

3. Saw-tooth generators are found in many analogue applications; producing the sweep voltages used for the horizontal deflection of oscilloscope displays and the tuning voltages for generating FMCW chirp radar signals. The Fourier series for an ideal saw-tooth waveform is given in Figure 3.1, while that for a square waveform is given in Figure 3.2.



$$f(x) = \frac{1}{2} - \frac{1}{\pi}\sum_{n=1}^{\infty} \frac{1}{n}\sin\left(\frac{n\pi x}{L}\right)$$

Figure 3.1 General Fourier series for an ideal saw-tooth waveform



$$f(x) = \frac{4}{\pi}\sum_{n=1,3,5,\ldots}^{\infty} \frac{1}{n}\sin\left(\frac{n\pi x}{L}\right)$$

Figure 3.2 General Fourier series for an ideal square waveform

a) Show mathematically how a saw-tooth generator can be used to realise an approximate square wave.

[3]

b) With the use of six circulators, two fixed attenuators, one fixed phase shifter and a single capacitor, draw the block diagram for a radio frequency circuit that can effectively convert the input from a saw-tooth generator into an approximation of a square wave output (ignore the effects of any impedance mismatches and assume all components to be ideal). Clearly identify all the blocks, with relevant information, and explain how the circuit works.

[10]

c) From the circuit in 3(b), explain why the phase shifter cannot be implemented with a delay line.

[2]

d) From the circuit in 3(b), what is its limitation in constructing the square wave shape?

[2]

e) From the circuit in 3(b), if all the ideal components introduce an insertion loss of 0.5 dB, what would be the overall increase in the insertion loss for the 2-port network at the different frequencies, and what effect will these losses have on the shape of the square wave approximation?

[3]

4.

a) Explain the difference between a duplexer and a diplexer.

[2]

b) What is the most common way of implementing a duplexer in half-duplex and full-duplex mobile phone handsets?

[2]

c) Using the ideal lumped-element *LC* filter synthesis technique to implement a high pass filter and its associated low-pass prototype, design a simple 50 Ω GSM duplexer that can allow the following bands to pass through the duplexer with a 1 dB worst-case pass band attenuation ripple:

TX band: 930 ~ 943 MHz

RX band: 885 ~ 889 MHz

Draw the block diagram for this basic duplexer, indicating the directions of signal flow and any relevant impedances.

Hint: Using the attached filter design tables, choose 7^{th}-order filters with a common cut-off frequency and common frequency ratio, giving approximately 10 dB rejection at the nearest band edge of the other band, and ensure that the one port impedance is double that of the other (e.g. the source termination impedance is $Zs = 2Zo$ and the load termination impedance is $Z_L = Zo = 50\ \Omega$).

[10]

d) Give up to four reasons why the design in 4(c) is not suitable for a real GSM mobile phone handset duplexer.

[4]

e) Calculate ideal worst-case levels of return loss ripple in the pass bands for the ideal filters used in 4(c).

[2]

5. Bi-directional repeaters have been used to amplify and correct for group delay distortions in long-distance full-duplex microwave cable links. They can be realised using circuitry that includes the use of directional couplers.

a) Draw a traditional Rat-Race coupler, numbering all ports and indicating all path lengths and impedances relative to the system's reference impedance *Zo*. In addition, write down the general S-parameter matrix for this component, when it is lossless and symmetrical.

[3]

b) Given two ideal lossless Rat-Race Couplers, two dummy loads and two unilateral amplifiers (one for the forward 'F' and the other for the reverse 'R' directions), design a simple bi-directional repeater for full-duplex applications.

[6]

c) By observation, write down expressions for the overall 2-port S-parameters for the circuit designed in 5(b). Make sure to clearly identify the S-parameters for the different amplifiers. Clearly state any assumptions made.

[5]

d) For the overall 2-port S-parameters derived in 5(c), what practical advantages are there to the voltage wave transmission and reflection coefficients when both amplifiers have identical S-parameters and equal voltage wave reflection coefficients? Assuming lossless couplers, what unilateral power gain would the amplifiers need to have for an overall bilateral power gain of 10 dB?

[3]

e) If the lossless Rat-Race couplers were made using a poorly characterised dielectric, what will the risk be to the operation of the repeater designed in 5(b)?

[3]

6. An inter-stage impedance matching network is required between the output of one transistor, having a small-signal output impedance of $15 - j30\ \Omega$, and the input of another transistor, having a small-signal input impedance of $10 + j20\ \Omega$. At a frequency of 2.45 GHz, using a Smith chart or otherwise, you are required to design suitable impedance matching networks to achieve maximum power transfer between these two impedances using the following techniques; draw each circuit and indicate the relevant loci on Smith chart sketches:

a) Resonant method and L-network matching having a shunt capacitor.

[3]

b) Resonant method and L-network matching having a shunt inductor. Give two practical advantages for this solution, when compared to that in 6(a).

[6]

c) Resonant method and series resistor. What is the practical advantage and fundamental problem with this solution?

[3]

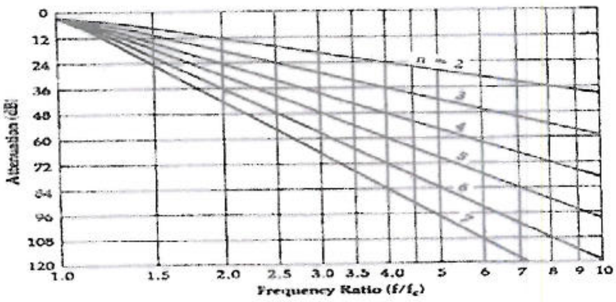
d) Resonant method and quarter-wavelength transformer.

[3]

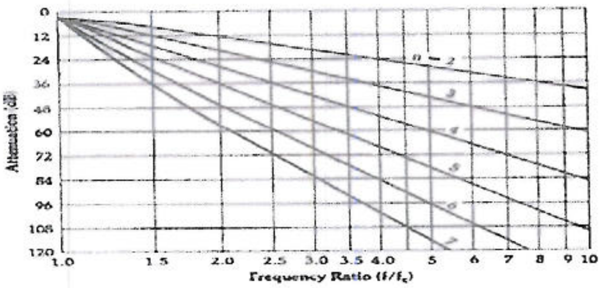
e) With the solution from 6(d), for a microstrip transmission line suspended in air 0.635 mm above the ground plane, use simplified expressions for calculating the physical dimensions of the quarter-wavelength transformer.

[5]

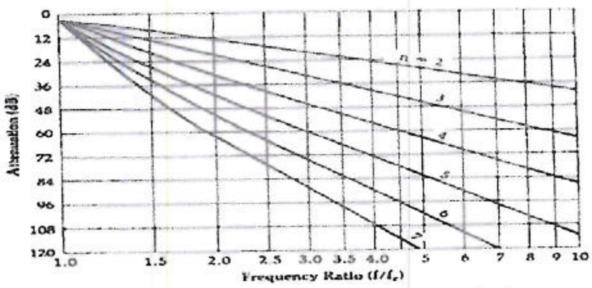
Filter tables



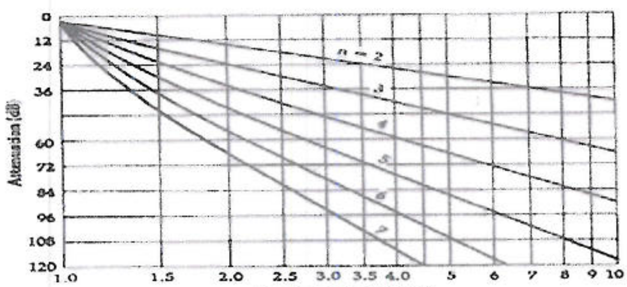
Attenuation characteristics for Butterworth filters.



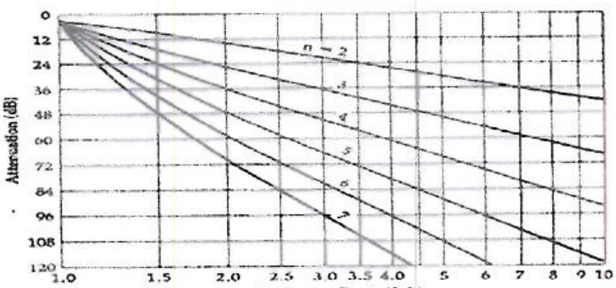
Attenuation characteristics for a Chebyshev filter with 0.01-dB ripple.



Attenuation characteristics for a Chebyshev filter with 0.1-dB ripple.

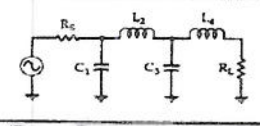


Attenuation characteristics for a Chebyshev filter with 0.5-dB ripple.

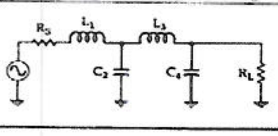


Attenuation characteristics for a Chebyshev filter with 1-dB ripple.

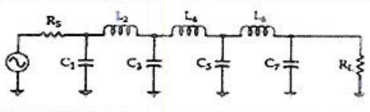
Chebyshev Low-Pass Prototype Element Values for 1.0-dB Ripple



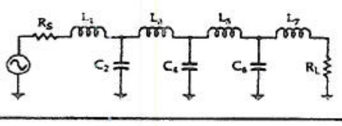
R_S/R_L	C_1	L_2	C_3	L_4
3.000	0.572	3.132		
4.000	0.365	4.600		
8.000	0.157	9.658		
∞	1.213	1.109		
1.000	2.216	1.088	2.216	
0.500	4.431	0.817	2.216	
0.333	6.647	0.726	2.216	
0.250	8.862	0.680	2.216	
0.125	17.725	0.612	2.216	
∞	1.652	1.460	1.108	
3.000	0.653	4.411	0.814	2.535
4.000	0.452	7.083	0.612	2.848
8.000	0.209	17.164	0.428	3.281
∞	1.350	2.010	1.488	1.106
R_L/R_S	L_1	C_2	L_3	C_4



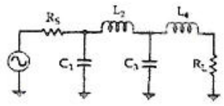
Chebyshev Low-Pass Prototype Element Values for 1.0-dB Ripple



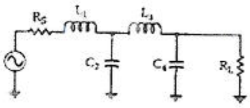
n	R_S/R_L	C_1	L_2	C_3	L_4	C_5	L_6	C_7
5	1.000	2.207	1.128	3.103	1.128	2.207		
	0.500	4.414	0.565	4.653	1.128	2.207		
	0.333	6.622	0.376	6.205	1.128	2.207		
	0.250	8.829	0.282	7.756	1.128	2.207		
	0.125	17.657	0.141	13.961	1.128	2.207		
	∞	1.721	1.645	2.061	1.493	1.103		
6	3.000	0.679	3.873	0.771	4.711	0.969	2.406	
	4.000	0.481	5.644	0.476	7.351	0.849	2.582	
	8.000	0.227	12.310	0.198	16.740	0.726	2.800	
	∞	1.378	2.097	1.690	2.074	1.494	1.102	
7	1.000	2.204	1.131	3.147	1.194	3.147	1.131	2.204
	0.500	4.408	0.566	6.293	0.895	3.147	1.131	2.204
	0.333	6.612	0.377	9.441	0.796	3.147	1.131	2.204
	0.250	8.815	0.283	12.588	0.747	3.147	1.131	2.204
	0.125	17.631	0.141	25.175	0.671	3.147	1.131	2.204
	∞	1.741	1.677	2.155	1.703	2.079	1.494	1.102
n	R_L/R_S	L_1	C_2	L_3	C_4	L_5	C_6	L_7



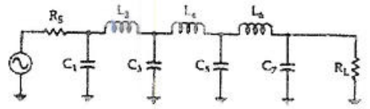
Butterworth Low-Pass Prototype Element Values



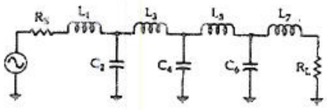
n	R_S/R_L	C_1	L_2	C_3	L_4
2	1.111	1.035	1.835		
	1.250	0.849	2.121		
	1.429	0.697	2.439		
	1.667	0.566	2.828		
	2.000	0.448	3.346		
	2.500	0.342	4.095		
	3.333	0.245	5.313		
	5.000	0.156	7.707		
	10.000	0.074	14.814		
	∞	1.414	0.707		
3	0.900	0.808	1.633	1.599	
	0.800	0.844	1.384	1.926	
	0.700	0.915	1.165	2.277	
	0.600	1.023	0.965	2.702	
	0.500	1.181	0.779	3.261	
	0.400	1.425	0.604	4.064	
	0.300	1.838	0.440	5.363	
	0.200	2.669	0.284	7.910	
	0.100	5.187	0.138	15.455	
	∞	1.500	1.333	0.500	
4	1.111	0.466	1.592	1.744	1.469
	1.250	0.388	1.695	1.511	1.811
	1.429	0.325	1.862	1.291	2.175
	1.667	0.269	2.103	1.082	2.613
	2.000	0.218	2.452	0.883	3.187
	2.500	0.169	2.986	0.691	4.009
	3.333	0.124	3.883	0.507	5.338
	5.000	0.080	5.684	0.331	7.940
	10.000	0.039	11.094	0.162	15.642
	∞	1.531	1.577	1.082	0.383
n	R_L/R_S	L_1	C_2	L_3	C_4



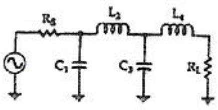
Butterworth Low-Pass Prototype Element Values



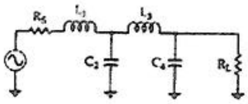
n	R_S/R_L	C_1	L_2	C_3	L_4	C_5	L_6	C_7
5	0.900	0.442	1.027	1.910	1.756	1.389		
	0.800	0.470	0.866	2.061	1.544	1.738		
	0.700	0.517	0.731	2.285	1.333	2.108		
	0.600	0.588	0.609	2.600	1.126	2.552		
	0.500	0.686	0.496	3.051	0.924	3.133		
	0.400	0.838	0.388	3.736	0.727	3.965		
	0.300	1.094	0.285	4.884	0.537	5.307		
	0.200	1.608	0.186	7.185	0.352	7.935		
	0.100	3.512	0.091	14.095	0.173	15.710		
	∞	1.545	1.694	1.382	0.894	0.309		
6	1.111	0.289	1.040	1.322	2.054	1.744	1.335	
	1.250	0.245	1.116	1.126	2.239	1.550	1.688	
	1.429	0.207	1.236	0.957	2.499	1.346	2.062	
	1.667	0.173	1.407	0.801	2.858	1.143	2.509	
	2.000	0.141	1.653	0.654	3.369	0.942	3.094	
	2.500	0.111	2.028	0.514	4.141	0.745	3.931	
	3.333	0.082	2.658	0.379	5.433	0.552	5.280	
	5.000	0.054	3.917	0.248	8.020	0.363	7.922	
	10.000	0.026	7.705	0.122	15.786	0.179	15.738	
	∞	1.553	1.759	1.553	1.202	0.758	0.259	
7	0.900	0.299	0.711	1.404	1.489	2.125	1.727	1.296
	0.800	0.322	0.606	1.517	1.278	2.334	1.546	1.652
	0.700	0.357	0.515	1.688	1.091	2.618	1.350	2.028
	0.600	0.408	0.432	1.928	0.917	3.005	1.150	2.477
	0.500	0.480	0.354	2.273	0.751	3.553	0.951	3.064
	0.400	0.590	0.278	2.795	0.592	4.380	0.754	3.904
	0.300	0.775	0.206	3.671	0.437	5.761	0.560	5.258
	0.200	1.145	0.135	5.427	0.287	8.526	0.369	7.908
	0.100	2.257	0.067	10.700	0.142	16.822	0.182	15.748
	∞	1.558	1.799	1.659	1.397	1.055	0.656	0.223
n	R_L/R_S	L_1	C_2	L_3	C_4	L_5	C_6	L_7



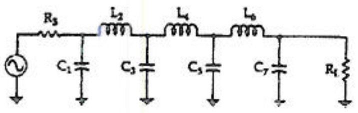
Chebyshev Low-Pass Element Values for 0.01-dB Ripple



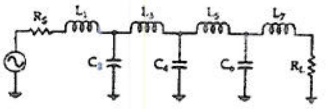
n	R_S/R_L	C_1	L_2	C_3	L_4
2	1.101	1.347	1.483		
	1.111	1.247	1.595		
	1.250	0.943	1.997		
	1.429	0.759	2.344		
	1.667	0.609	2.750		
	2.000	0.479	3.277		
	2.500	0.363	4.033		
	3.333	0.259	5.255		
	5.000	0.164	7.650		
	10.000	0.078	14.749		
	∞	1.412	0.742		
3	1.000	1.181	1.821	1.181	
	0.900	1.092	1.660	1.480	
	0.800	1.097	1.443	1.806	
	0.700	1.160	1.228	2.165	
	0.600	1.274	1.024	2.598	
	0.500	1.452	0.829	3.164	
	0.400	1.734	0.645	3.974	
	0.300	2.216	0.470	5.280	
	0.200	3.193	0.305	7.834	
	0.100	6.141	0.148	15.390	
	∞	1.501	1.433	0.591	
4	1.100	0.950	1.938	1.761	1.046
	1.111	0.854	1.946	1.744	1.165
	1.250	0.618	2.075	1.542	1.617
	1.429	0.495	2.279	1.334	2.008
	1.667	0.398	2.571	1.128	2.461
	2.000	0.316	2.994	0.926	3.045
	2.500	0.242	3.641	0.729	3.875
	3.333	0.174	4.727	0.538	5.209
	5.000	0.112	6.910	0.352	7.813
	10.000	0.054	13.469	0.173	15.510
	∞	1.529	1.691	1.312	0.523
n	R_L/R_S	L_1	C_2	L_3	C_4



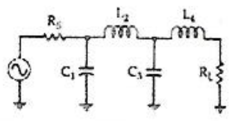
Chebyshev Low-Pass Element Values for 0.01-dB Ripple



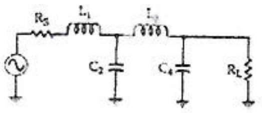
n	R_S/R_L	C_1	L_2	C_3	L_4	C_5	L_6	C_7
5	1.000	0.977	1.685	2.037	1.685	0.977		
	0.900	0.880	1.456	2.174	1.641	1.274		
	0.800	0.877	1.235	2.379	1.499	1.607		
	0.700	0.926	1.040	2.658	1.323	1.977		
	0.600	1.019	0.863	3.041	1.135	2.424		
	0.500	1.166	0.699	3.584	0.942	3.009		
	0.400	1.398	0.544	4.403	0.749	3.845		
	0.300	1.797	0.398	5.772	0.557	5.193		
	0.200	2.604	0.259	8.514	0.368	7.826		
	0.100	5.041	0.127	16.741	0.182	15.613		
	∞	1.547	1.705	1.645	1.237	0.488		
6	1.101	0.851	1.796	1.841	2.027	1.631	0.937	
	1.111	0.760	1.782	1.775	2.094	1.638	1.053	
	1.250	0.545	1.864	1.489	2.403	1.507	1.504	
	1.429	0.436	2.038	1.266	2.735	1.332	1.899	
	1.667	0.351	2.298	1.061	3.167	1.145	2.357	
	2.000	0.279	2.678	0.867	3.768	0.954	2.948	
	2.500	0.214	3.261	0.682	4.667	0.761	3.790	
	3.333	0.155	4.245	0.503	6.163	0.568	5.143	
	5.000	0.100	6.223	0.330	9.151	0.376	7.785	
	10.000	0.048	12.171	0.162	18.105	0.187	15.595	
	∞	1.551	1.847	1.790	1.598	1.190	0.469	
7	1.000	0.913	1.595	2.002	1.870	2.002	1.595	0.913
	0.900	0.816	1.362	2.089	1.722	2.202	1.581	1.206
	0.800	0.811	1.150	2.262	1.525	2.465	1.464	1.538
	0.700	0.857	0.967	2.516	1.323	2.802	1.307	1.910
	0.600	0.943	0.803	2.872	1.124	3.250	1.131	2.359
	0.500	1.080	0.650	3.382	0.928	3.875	0.947	2.948
	0.400	1.297	0.507	4.156	0.735	4.812	0.758	3.790
	0.300	1.669	0.372	5.454	0.546	6.370	0.568	5.148
	0.200	2.242	0.242	8.057	0.360	9.484	0.378	7.802
	0.100	4.701	0.119	15.872	0.178	18.818	0.188	15.652
	∞	1.559	1.867	1.866	1.765	1.563	1.161	0.456
n	R_L/R_S	L_1	C_2	L_3	C_4	L_5	C_6	L_7



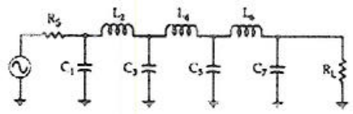
Chebyshev Low-Pass Prototype Element Values for 0.1-dB Ripple



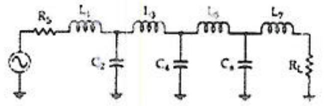
n	R_S/R_L	C_1	L_2	C_3	L_4
2	1.355	1.209	1.638		
	1.429	0.977	1.982		
	1.667	0.733	2.480		
	2.000	0.580	3.054		
	2.500	0.417	3.827		
	3.333	0.293	5.050		
	5.000	0.184	7.426		
	10.000	0.087	14.433		
	∞	1.391	0.819		
3	1.000	1.433	1.594	1.433	
	0.900	1.426	1.494	1.622	
	0.800	1.451	1.356	1.871	
	0.700	1.521	1.193	2.190	
	0.600	1.648	1.017	2.603	
	0.500	1.853	0.838	3.159	
	0.400	2.186	0.660	3.968	
	0.300	2.763	0.486	5.279	
	0.200	3.942	0.317	7.850	
	0.100	7.512	0.155	15.466	
	∞	1.513	1.510	0.716	
4	1.355	0.992	2.148	1.585	1.341
	1.429	0.779	2.348	1.429	1.700
	1.667	0.576	2.730	1.185	2.243
	2.000	0.440	3.227	0.967	2.856
	2.500	0.329	3.961	0.760	3.698
	3.333	0.233	5.178	0.580	5.030
	5.000	0.148	7.607	0.367	7.614
	10.000	0.070	14.887	0.180	15.230
	∞	1.511	1.768	1.455	0.673
n	R_L/R_S	L_1	C_2	L_3	C_4



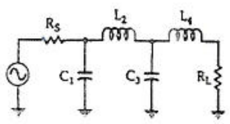
Chebyshev Low-Pass Prototype Element Values for 0.1-dB Ripple



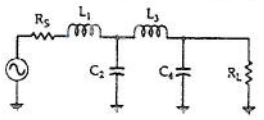
n	R_S/R_L	C_1	L_2	C_3	L_4	C_5	L_6	C_7
5	1.000	1.301	1.556	2.241	1.556	1.301		
	0.900	1.285	1.433	2.380	1.488	1.488		
	0.800	1.300	1.282	2.582	1.382	1.738		
	0.700	1.358	1.117	2.868	1.244	2.062		
	0.600	1.470	0.947	3.269	1.085	2.484		
	0.500	1.654	0.778	3.845	0.913	3.055		
	0.400	1.954	0.612	4.720	0.733	3.886		
	0.300	2.477	0.451	6.196	0.550	5.237		
	0.200	3.546	0.295	9.127	0.366	7.869		
	0.100	6.787	0.115	17.957	0.182	15.745		
	∞	1.561	1.807	1.766	1.417	0.651		
6	1.355	0.942	2.080	1.659	2.247	1.534	1.277	
	1.429	0.735	2.249	1.454	2.544	1.405	1.629	
	1.667	0.542	2.600	1.183	3.064	1.185	2.174	
	2.000	0.414	3.068	0.958	3.712	0.979	2.794	
	2.500	0.310	3.765	0.749	4.651	0.778	3.645	
	3.333	0.220	4.927	0.551	6.195	0.580	4.996	
	5.000	0.139	7.250	0.361	9.261	0.384	7.618	
	10.000	0.067	14.220	0.178	18.427	0.190	15.350	
	∞	1.534	1.884	1.831	1.749	1.394	0.638	
7	1.000	1.262	1.520	2.239	1.680	2.239	1.520	1.262
	0.900	1.242	1.305	2.361	1.578	2.397	1.459	1.447
	0.800	1.255	1.245	2.548	1.443	2.624	1.362	1.697
	0.700	1.310	1.083	2.819	1.283	2.942	1.233	2.021
	0.600	1.417	0.917	3.205	1.209	3.384	1.081	2.444
	0.500	1.595	0.753	3.764	0.928	4.015	0.914	3.018
	0.400	1.885	0.593	4.618	0.742	4.970	0.738	3.855
	0.300	2.392	0.437	6.054	0.556	6.569	0.557	5.217
	0.200	3.428	0.286	8.937	0.369	9.770	0.372	7.890
	0.100	6.570	0.141	17.603	0.184	19.376	0.186	15.813
	∞	1.575	1.858	1.921	1.827	1.734	1.379	0.631
n	R_L/R_S	L_1	C_2	L_3	C_4	L_5	C_6	L_7



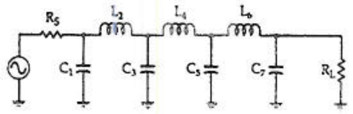
Chebyshev Low-Pass Prototype Element Values for 0.5-dB Ripple



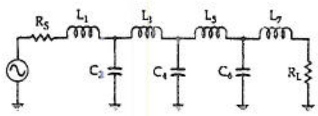
n	R_S/R_L	C_1	L_2	C_3	L_4
2	1.984	0.983	1.950		
	2.000	0.909	2.103		
	2.500	0.564	3.165		
	3.333	0.375	4.411		
	5.000	0.228	6.700		
	10.000	0.105	13.322		
	∞	1.307	0.975		
3	1.000	1.864	1.280	1.834	
	0.900	1.918	1.209	2.026	
	0.800	1.997	1.120	2.237	
	0.700	2.114	1.015	2.517	
	0.500	2.557	0.759	3.436	
	0.400	2.985	0.615	4.242	
	0.300	3.729	0.463	5.576	
	0.200	5.254	0.309	8.225	
	0.100	9.890	0.153	16.118	
	∞	1.572	1.518	0.932	
4	1.984	0.920	2.586	1.304	1.826
	2.000	0.845	2.720	1.238	1.985
	2.500	0.516	3.766	0.869	3.121
	3.333	0.344	5.120	0.621	4.480
	5.000	0.210	7.708	0.400	6.987
	10.000	0.098	15.352	0.194	14.262
	∞	1.436	1.889	1.521	0.913
n	R_L/R_S	L_1	C_2	L_3	C_4



Chebyshev Low-Pass Prototype Element Values for 0.5-dB Ripple



n	R_S/R_L	C_1	L_2	C_3	L_4	C_5	L_6	C_7
5	1.000	1.807	1.303	2.691	1.303	1.807		
	0.900	1.854	1.222	2.849	1.238	1.970		
	0.800	1.926	1.126	3.060	1.157	2.185		
	0.700	2.035	1.015	3.353	1.058	2.470		
	0.600	2.200	0.890	3.765	0.942	2.861		
	0.500	2.457	0.754	4.367	0.810	3.414		
	0.400	2.870	0.609	5.296	0.684	4.245		
	0.300	3.588	0.459	6.871	0.508	5.625		
	0.200	5.064	0.306	10.054	0.343	8.367		
	0.100	9.556	0.153	19.647	0.173	16.574		
	∞	1.630	1.740	1.922	1.514	0.903		
6	1.984	0.905	2.577	1.368	2.713	1.299	1.796	
	2.000	0.830	2.704	1.291	2.872	1.237	1.956	
	2.500	0.506	3.722	0.890	4.109	0.881	3.103	
	3.333	0.337	5.055	0.632	5.699	0.635	4.481	
	5.000	0.206	7.615	0.406	8.732	0.412	7.031	
	10.000	0.096	15.186	0.197	17.681	0.202	14.433	
7	1.000	1.790	1.296	2.718	1.385	2.718	1.296	1.790
	0.900	1.835	1.215	2.869	1.308	2.883	1.234	1.953
	0.800	1.905	1.118	3.076	1.215	3.107	1.155	2.168
	0.700	2.011	1.007	3.364	1.105	3.416	1.058	2.455
	0.600	2.174	0.882	3.772	0.979	3.852	0.944	2.848
	0.500	2.428	0.747	4.370	0.838	2.289	0.814	3.405
	0.400	2.835	0.604	5.295	0.685	5.470	0.669	4.243
	0.300	3.546	0.455	6.867	0.522	7.134	0.513	5.635
	0.200	5.007	0.303	10.049	0.352	10.496	0.348	8.404
	0.100	9.456	0.151	19.649	0.178	20.631	0.176	16.665
	∞	1.646	1.777	2.031	1.789	1.924	1.503	0.895
n	R_L/R_S	L_1	C_2	L_3	C_4	L_5	C_6	L_7



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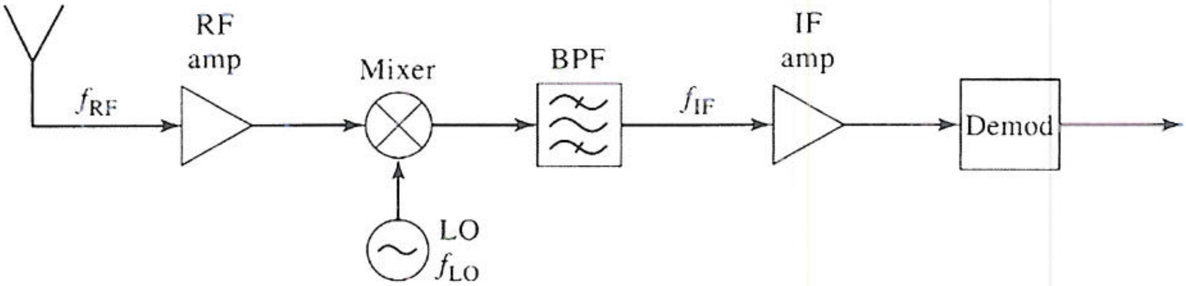
The Solutions for E4.18, 2012

Model answer to Q 1(a): Bookwork

Draw the block diagram for a single-conversion superhet receiver and, stating its full name, explain the origin of this type of architecture.



Superhet is an abbreviation for Super-sonic Heterodyne (or Superheterodyne for short) and refers to the fact that the IF frequency is much higher than audio frequencies.

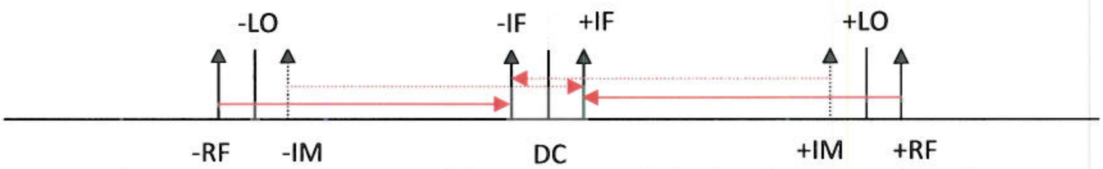
[2]

Model answer to Q 1(b): Extended Discussions

(i) With the help of a spectral illustration, briefly explain why an image rejection filter is needed and where this should be located within the architecture. Also, briefly explain why this filter represents a problem and list 3 possible solutions that can be adopted to alleviate or avoid this problem.

Image rejection filter should be inserted in the RF path, before the mixer.

Without this filter, with reference to the spectral illustration should below, the +RF that is translated down into the +IF will be combined with any negative spectral image (-IM) that is translated up to the +IF.



This filter represents a problem because if the local intermediate frequency (IF) is too low then a very expensive high-order filter is needed in order to sufficiently reject the close-to-RF image frequency.

Three possible ways to alleviate or avoid the problem include: (1) using a double-superheterodyne architecture, which will allow a reduction in the filter order by having a higher first IF; (2) use a zero-IF architecture, which does not need an IR filter; and (3) employ and image rejection mixer, which does not need an IR filter.

[3]

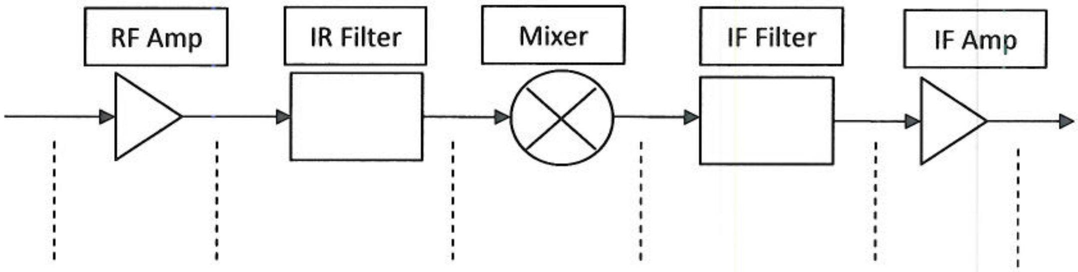
(ii) For selecting bands with higher levels of adjacent-band rejection, what simple method is usually can be adopted.

Selectivity is improved by having much cheaper high order band-pass filters located in the IF.

[1]

Model answer to Q 1(c): Calculated Application of Theory

The receiver has an input power level of -50 dBm. With the simplest of architecture design for the type in 1(a), up to the demodulator, it can be assumed that there is only one filter and one amplifier at any given frequency. All amplifiers have a gain of 30 dB, IP3 of 40 dBm and noise figure of 1 dB; all filters having a pass band insertion loss of 2 dB and the mixer has a conversion loss of 3 dB and IP3 of 30 dBm. Ignoring noise contributions from the local oscillator, calculate the following at the output of each sub-system block (where appropriate) and state the main equations used



G [dB]	+30	-2	-6	-2	+30
IP_3 [dBm]	+40	+100	+30	+100	+40
F [dB]	+1	+2	+6	+2	+1

+100 are assumed values and this may change for different students!

Pout [dBm]	-50	-20	-22	-28	-30	0	[2]
IP_3 [dBm]		+40	+38	+27.88	+25.88	+39.89	[3]
IMD_3 [dBc]		120	120	111.76	111.76	79.78	[2]
I_3 [dBm]		-140	-142	-139.76	-141.76	-79.78	[1]
F_{RX} [dB]	(1.259 +0.000585 +0.00472 + 0.00369 +0.00259) = 1.271 = 1.04 dB						[3]

The following main equations should be used:

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

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

$$I_3 = (IMD_3[dBc] - C\,[dBm])[dBm]$$

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

If the System Temperature is the same as the ambient temperature of To = 290 K (which would be rather unusual) then (kToB) = -129.2 dBm the output noise temperature would be: $No = G_{RX} F_{RX} (kT_o B)$ = -78.16 dBm, giving a signal-to-noise ratio at the output of the receiver So/No = 78.16 dB.

[3]

Model answer to Q 2(a): Calculated Application of Theory

If the geostationary satellite is positioned 36,000 km directly above the earth's 3 km diameter circular receiving target area, calculate the following:

i) The angle subtended by the receiving target at the satellite.

$$\psi = 2tan^{-1}(1.5/36000) = 83.333x10^{-6}\ radians$$

[1]

ii) The optimum beam solid angle for the satellite's antenna. *Hint: for large aperture antennas having a single pencil beam radiation pattern, this is given by $\Omega \approx \theta_E\ \theta_H$, where θ_E and θ_H are the -3 dB beam widths in the E- and H-planes, respectively.* Clearly state any simplifying assumptions.

$$\Omega_M \approx \theta_E\ \theta_H = \theta^2 \text{ assumed } = \psi^2 = 6.944x10^{-9}$$

[1]

iii) The directivity of the satellite's antenna.

$$\text{Directivity, } D_{TX} = \frac{4\pi}{\Omega} = 1.80956x10^9 = 92.6\ dBi$$

[1]

iv) The ideal effective aperture area of the satellite's antenna.

$$\text{Free-space wavelength, } \lambda_o = 5.17\ cm$$

$$D_{TX} = \frac{4\pi}{{\lambda_o}^2} A_{TX}\ \therefore A_{TX} = 385.26x10^3\ m^2$$

[2]

v) The ideal diameter of the satellite's antenna if it is to be implemented by a conventional circular parabolic reflecting antenna.

$$A_{TX} = \pi\left(\frac{d_{TX}}{2}\right)^2\quad \therefore d_{TX} = 700\ m$$

[1]

Model answer to Q 2(b): Calculated Application of Theory

The International Commission for Non-Ionizing Radiation Protection gives a limit for the maximum safe power density for human exposure of 1 mW/cm^2. Using appropriate calculations, comment on the safety levels at the:

i) Satellite.

Power density = 1 GW/$385.26x10^3\ m^2$ = 260 mW/cm^2

This is 260 times the maximum safe level,

but in space nobody can hear you scream and so this is not so much of a problem!

[3]

ii) Receiving target.

EIRP = $P_{TX}\ G_{TX}$ = 182.58 dBW

Spreading loss = $1/(4\ \pi\ Range^2)$ = -162.12 dB/m^2

Power density = 20.46 dBW/m^2 = 11 mW/cm^2

This is 11 times the maximum safe level

and, therefore, could pose a significant heath hazard with long-term exposure!

[3]

Model answer to Q 2(c): Application of new theory

If the receiving target area is optimally covered with half-wavelength by half-wavelength sized patch antennas, estimate the maximum number of patch antennas that can fit into the target array. *Hint: convert the circular array into a square array.*

$$A_{RX} = \pi \left(\frac{d_{RX}}{2}\right)^2 \quad where\ d_{RX} = 3\ km$$

$$A_{RX} = 7.069\ x10^6\ m^2\ and\ each\ side\ is\ 265{,}868\ cm\ long$$

Therefore, it is possible to fit about 102,850 x 102,850 = 10,578,122,500

patches into the target area

[2]

Model answer to Q 2(d): Calculated Application of Theory

Assuming the antennas in 2(c) are 75% efficient, calculate output power from each.

Area of patch is 2.585 cm x 2.585 cm = 6.682 cm^2

Therefore, 11 mW/cm^2 x 6.682 cm^2 x 0.75 = 55.128 mW from each patch antenna

[2]

Model answer to Q 2(e): Calculated Application of Theory

The power from 2(d) is rectified to produce dc power with an efficiency of 40%, calculated the output dc power from each of the rectifying antennas (know as rectennas) and then calculate the total dc power output form the receiver's target area.

55.128 mW x 0.4 = 22.05 mW output dc power from each rectenna

[1]

Model answer to Q 2(f): Calculated Application of Theory

The satellite's power combing and antenna subsystem is also 75% efficient. If 2.5 W output transistors are employed, each having an overall dc-to-RF conversion efficiency of 70%, calculate the:

i) Total number of transistors needed.

1 GW/0.75 = 1.3333 GW needs to be output from the transistors.

1.333333 GW/2.5 W = 534 $x10^6$ transistors are needed

[1]

ii) Dc power needed to be generated by the solar panel array.

1.333333 GW/0.7 = 1.904757 GW dc power generated by the satellite

[1]

iii) Overall end-to-end dc-to-dc efficiency for the complete system.

1.904757 GW/233.25 MW = 12.2% efficiency

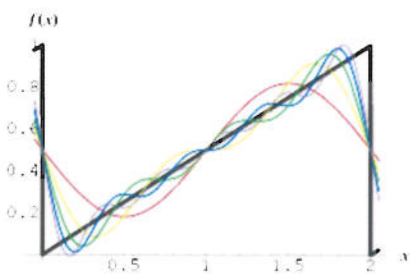
[1]

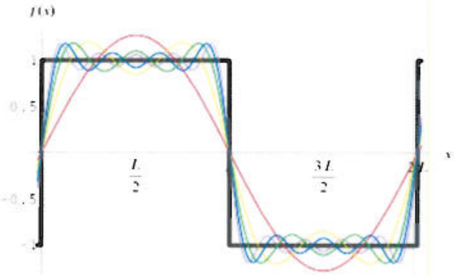
Model answer to Q 3(a): Application of New Theory

Show mathematically how a saw-tooth generator can be used to realise an approximate square wave.

The Fourier series for a saw-tooth and square wave are:

$$f(x) = \frac{1}{2} - \frac{1}{\pi}\sum_{n=1}^{\infty}\frac{1}{n}\sin\left(\frac{n\pi x}{L}\right) \qquad f(x) = \frac{4}{\pi}\sum_{n=1,3,5\ldots}^{\infty}\frac{1}{n}\sin\left(\frac{n\pi x}{L}\right).$$





The saw-tooth wave generator has the following Fourier series

$$f(t) = \frac{1}{2} - \frac{1}{2}\left(Sin\omega_o t + \frac{Sin2\omega_o t}{2} + \frac{Sin3\omega_o t}{3} + \frac{Sin4\omega_o t}{4} + \frac{Sin5\omega_o t}{5} + \frac{Sin6\omega_o t}{6}\ldots\right)$$

The square wave needs the following Fourier series

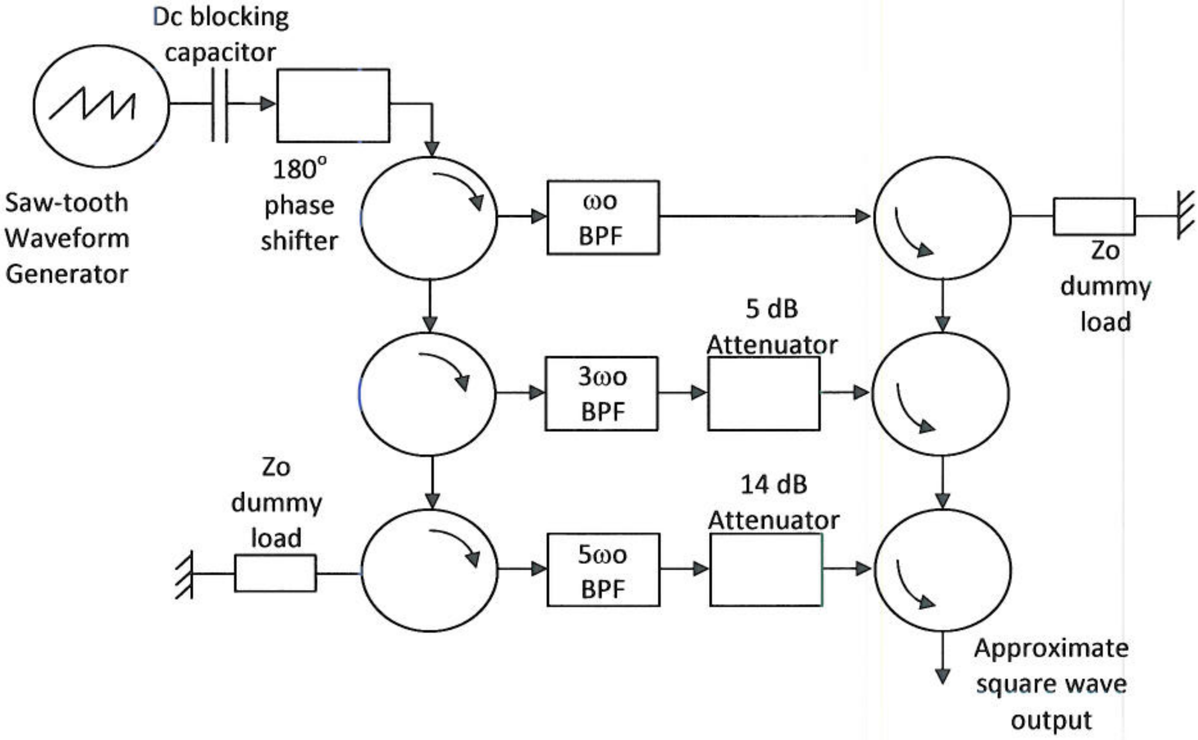
$$f(t) = \frac{4}{\pi}\left(Sin\omega_o t + \frac{Sin3\omega_o t}{3} + \frac{Sin5\omega_o t}{5} + \frac{Sin7\omega_o t}{7}\ldots\right)$$

Therefore, we need to remove the ½ dc component and all the even harmonics for the saw-tooth waveform. One can also remove the phase inversion with a phase shifter.

[3]

Model answer to Q 3(b): Application of New Theory

With the use of six circulators, two fixed attenuators, one fixed phase shifter and a single capacitor, draw the block diagram for a radio frequency circuit that can effectively convert the input from a saw-tooth generator into an approximation of a square wave output (ignore the effects of any impedance mismatches and assume all components to be ideal). Clearly identify all the blocks, with relevant information, and explain how the circuit works.



The dc blocking capacitor removes the dc component from the saw-tooth signal. The phase then inverts the phase of all the harmonics. It is assumed that there are no other issues of phase to consider (not a good assumption in practice). Since all the lossless circulators have infinite bandwidth the top left one allows ωo through the top branch, but all the other harmonics are reflected back into the circulator from the perfect reflection of the ωo band-pass filter (BPF), so on and so forth. The filters then select the harmonics needed to construct the square waveform, but only after they have been suitably attenuated and combined by the right-hand multiplexing column of circulators.

[10]

Model answer to Q 3(c): Application of Bookwork

From the circuit in 3(b), explain why the phase shifter cannot be implemented with a delay line.

A delay line is an example of a true time shifter, whereby the group delay is constant for all frequencies, but the insertion phase decreases linearly with an increase in frequency. For this application, one needs a 108° phase shift at all the harmonic frequencies and, therefore, a true phase shifter is needed.

[2]

Model answer to Q 3(d): Application of New Theory

From the circuit in 3(b), what is its limitation in contracting the square wave shape?

Only the 1^{st}, 3^{rd} and 5^{th} harmonics have been used to construct the square wave. Therefore, the shape of the waveform will exhibit distortion in the form of ripples.

[2]

Model answer to Q 3(e): Application of New Theory

From the circuit in 3(b), if all the ideal components introduce an insertion loss of 0.5 dB, what would be the overall increase in the insertion loss for the 2-port network at the different frequencies, and what effect will these losses have on the shape of the square wave approximation.

When the capacitor, circulator and filter now include a 0.5 dB insertion loss, there will be an introduction of an overall increase in the 2-port insertion loss of 4.5 dB for all the frequency components and, therefore, there will be no further distortion in the shape of the waveform.

[3]

Model answer to Q 4(a): Bookwork

Explain the difference between a duplexer and a diplexer.

A DUPLEXER is a 3-port network that allows the transceiver to use the same antenna.. A DIPLEXER is a 3-port filter that either separates or combines signals from 2 different frequency bands.

[2]

Model answer to Q 4(b): Bookwork

What is the most common way of implementing a duplexer in half-duplex and full-duplex mobile phone handsets?

In a half-duplex mobile phone handset, the duplexer is implemented with a transmit/receive (T/R) module, which normally employs a T/R switch (e.g. SP2T). In a full-duplex mobile phone handset, the duplexer can be as simple as a circulator in low-power applications or filters in high isolation applications (e.g. W-CDMA)

[2]

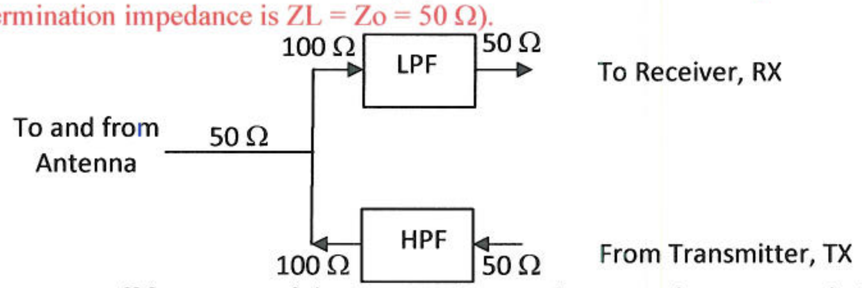
Model answer to Q 4(c): Calculated Application of Theory

Using the ideal lumped-element *LC* filter synthesis technique to implement a high pass filter and it associated low-pass prototype, design a simple 50 Ω GSM duplexer that can allow the following bands to pass through the duplexer with a 1 dB worst-case pass band attenuation ripple:

TX band: 930 ~ 943 MHz and RX band: 885 ~ 889 MHz

Draw the block diagram for this basic duplexer, indicating the directions of signal flow and any relevant impedances.

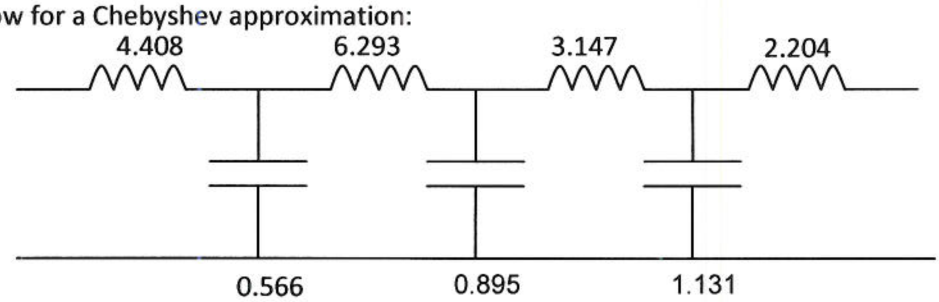
Hint: Choose 7th-order filters with a common cut-off frequency and common frequency ratio, giving approximately 10 dB rejection at the nearest band edge of the other band, and ensure that the one port impedance is double that of the other (e.g. the source termination impedance is Zs = 2Zo and the load termination impedance is ZL = Zo = 50 Ω).



First choose a common cut-off frequency of the geometric mean between the upper end of the RX band and the lower cut-off frequency of the TX band (i.e. fc = sqrt(889 MHz x 930 MHz) = 909.27 MHz. Also, the frequency ratio for both filters is f/fc = 930/909.27 = 909.27/889 =1.023.

From graphs provided, for the Chebyshev low-pass filter with a 1 dB ripple, a 7^{th} order filter is required to achieve an out-of-band rejection of ~ 10 dB and fc/f ratio of 1.023.

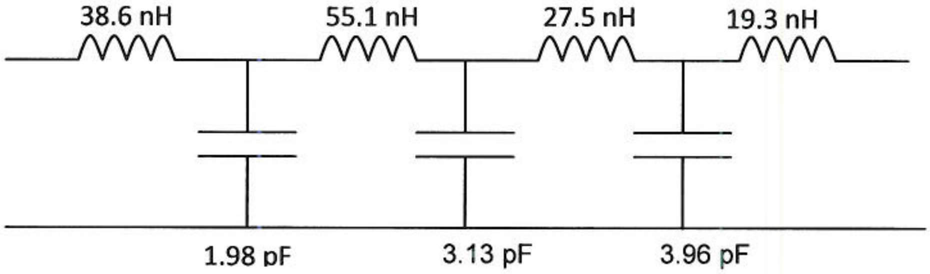
From tables provided, with $R_S = 2\,R_L = 2Zo = 100$, the prototype low-pass filter and associated coefficients are given below for a Chebyshev approximation:

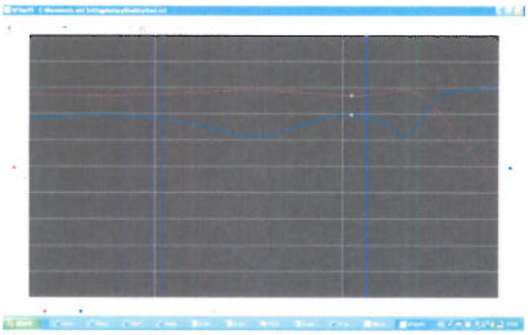


Low-pass de-normalising:

Series inductor: Shunt Capacitor:

$$L_S = \frac{R_L}{2\pi f_c} Ln \qquad C_P = \frac{Cn}{2\pi f_c R_L}$$

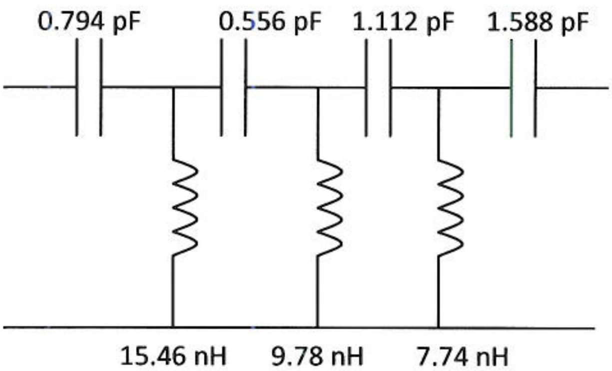


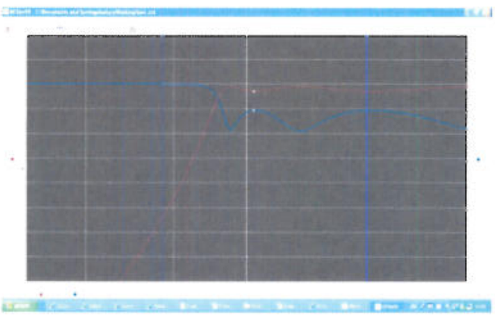


High-pass de-normalising:

Shunt inductor: Series Capacitor:

$$L_P = \frac{R_L}{2\pi f_c Ln} \qquad C_S = \frac{1}{2\pi f_c Cn R_L}$$





The deviation in the results from those predicted by theory are due to limited component tolerances used within the simulations, giving rise to a 1.5 dB ripple instead of the desired 1 dB.

[10]

Model answer to Q 4(d): New Interpretation of Theory

Explain reasons why the design in 4(c) is not ideal for either the TX and RX bands.

First of all, 7^{th} order Chebyshev approximations for this very small frequency ratio only gives about 10 out-of-band rejection at the other band's nearest band edge frequency. This should be about 100 dB of rejection in practical mobile phone handset applications.

Also, with the transmitter, a BPF should be used instead of the HPF, to stop any harmonics from radiating out of the antenna.

Moreover, with the receiver, a BPF should be used instead of the LPF in order to minimise the noise bandwidth and, therefore, the noise power (as well as any interfering signals) entering into the receiver.

Finally, since the band-pass ripples are up 1 dB, the reflected power at each filter will effect the overall performance of the duplexer in both the TX and RX bands.

[4]

Model answer to Q 4(e): Calculated Application of Theory

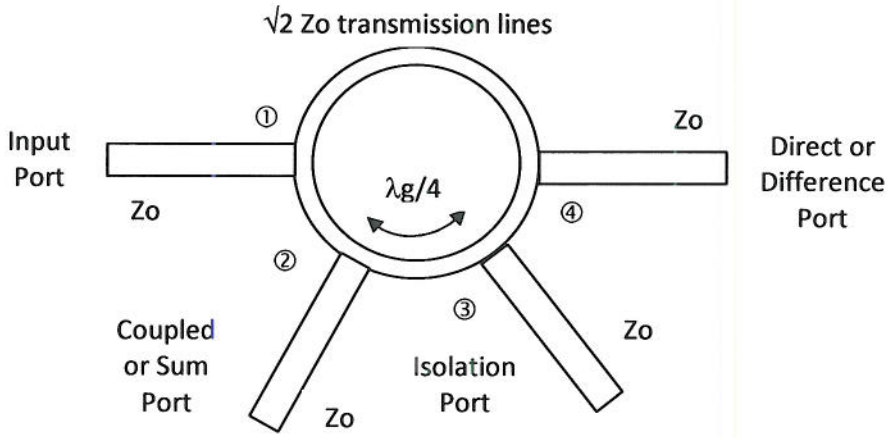
Calculate the worst-case levels of return loss ripple in the pass bands for the ideal filters used in 4(c).

From the principle of conservation of energy, the worst-case return loss for a lossless LC filter will be $|S_{11}|^2 = 1-|S_{21}|^2 = -6.868$ dB, since the $|S_{21}|^2 = 0.79432$ for a 1 dB ripple.

[2]

Model answer to Q 5(a): Bookwork

Draw a traditional Rat-Race coupler, numbering all ports and indicating all path lengths and impedances relative to the system's reference impedance *Zo*. In addition, write down the general S-parameter matrix for this component, and when it is lossless and symmetrical.

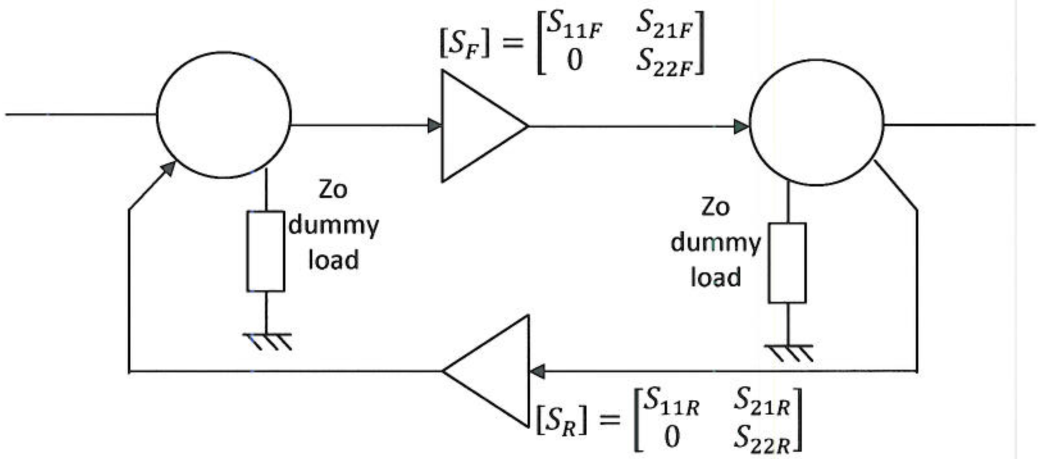


$$[S] = \begin{bmatrix} S_{11} & S_{12} & S_{13} & S_{14} \\ S_{21} & S_{22} & S_{23} & S_{24} \\ S_{31} & S_{32} & S_{33} & S_{34} \\ S_{41} & S_{42} & S_{43} & S_{44} \end{bmatrix} \rightarrow \begin{bmatrix} 0 & -j/\sqrt{2} & 0 & j/\sqrt{2} \\ -j/\sqrt{2} & 0 & j/\sqrt{2} & 0 \\ 0 & j/\sqrt{2} & 0 & -j/\sqrt{2} \\ j/\sqrt{2} & 0 & -j/\sqrt{2} & 0 \end{bmatrix}$$

[3]

Model answer to Q 5(b): Application of New Theory

Given two ideal lossless Rat-Race Couplers, two dummy loads and two unilateral amplifiers (one for the forward 'F' and the other for the reverse 'R' directions), design a simple bi-directional repeater for full-duplex applications.



[6]

Model answer to Q 5(c): Application of New Theory

By observation, write down expressions for the overall 2-port S-parameters for the circuit designed in 5(b). Make sure to clearly identify the S-parameters for the different amplifiers. Clearly state any assumptions made.

The couplers are assumed to be ideal, lossless and symmetrical.

$$S_{21} = \frac{-j}{\sqrt{2}} S_{21F} \frac{-j}{\sqrt{2}} = -\frac{S_{21F}}{2}$$
$$S_{12} = \frac{-j}{\sqrt{2}} S_{21R} \frac{-j}{\sqrt{2}} = -\frac{S_{21R}}{2}$$
$$S_{11} = \left(\frac{j}{\sqrt{2}} S_{11F} \frac{j}{\sqrt{2}}\right) + \left(\frac{-j}{\sqrt{2}} S_{22R} \frac{-j}{\sqrt{2}}\right) = \left(\frac{-S_{11F}}{2}\right) + \left(\frac{-S_{22R}}{2}\right)$$
$$S_{22} = \left(\frac{-j}{\sqrt{2}} S_{11R} \frac{-j}{\sqrt{2}}\right) + \left(\frac{j}{\sqrt{2}} S_{22F} \frac{j}{\sqrt{2}}\right) = \left(\frac{-S_{11R}}{2}\right) + \left(\frac{-S_{22F}}{2}\right)$$

[5]

Model answer to Q 5(d): Application of New Theory

What practical advantages to the voltage wave transmission and reflection coefficients, for the overall 2-port S-parameters derived in 5(c), when both amplifiers have identical S-parameters and with equal voltage wave reflection coefficients? What unilateral power gain would the amplifiers need to have for an overall bilateral power gain of 10 dB?

The voltage wave transmission coefficients will be identical in both directions and so the 2-port network can be interchanged.

The voltage wave reflection coefficients will be zero at both ports and so is perfectly impedance matched.

The unilateral power gain needs to be 16 dB, with lossless couplers, since each coupler effectively attenduates the signal by 3 dB.

[3]

Model answer to Q 5(e): Application of New Theory

If the lossless Rat-Race couplers were made using a poorly characterised dielectric, what is the risk be to the operation of the repeater designed in 5(b)?

The wrong value of relative permittivity in the dielectric constant of the transmission lines used to realise the couplers will change the effective path lengths of the signals around the coupler. As a result, at a fixed operational frequency, the isolation port of the coupler may exhibit finite isolation and, therefore, the resulting leakage will create a feedback path around the circuit that will cause instability and oscillation.

[3]

Model answer to Q 6(a): Calculated Application of Theory

An inter-stage impedance matching network is required between one transistor, having a small-signal output impedance of 15 – j30 Ω, and the input of another transistor, having a small-signal input impedance of 10 + j20 Ω. At a frequency of 2.45 GHz, you are required to design suitable impedance matching networks to achieve maximum power transfer between these two impedances using the following techniques; draw each circuit and indicate the relevant loci on Smith chart sketches:

Resonant method and L-network matching using a shunt capacitor.

The +j20 Ω inductive load reactance is resonated out with a 3.248 pF capacitor.

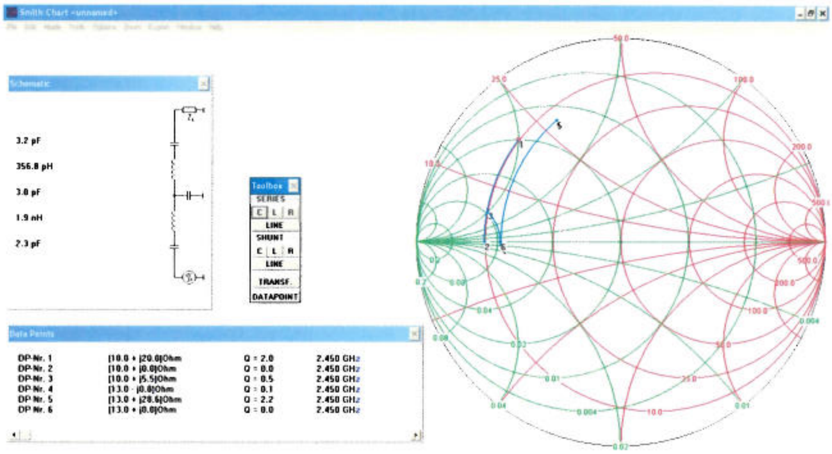
The –j30 Ω capacitive source reactance is resonated out with a 1.949 nH inductance.

The next step is to match the 10 Ω load resistance to the 15 Ω source resistance:

$$Q = \sqrt{\frac{15}{10} - 1} = \frac{1}{\sqrt{2}} \quad \text{and } \omega_o = 2\pi \; 2.45x10^9$$

$$Q = Q_s = \frac{\omega_o L_s}{10} = \frac{1}{\sqrt{2}} \qquad \therefore L_s = 0.459\, nH$$

$$Q = Q_p = 15\omega_o C_p = \frac{1}{\sqrt{2}} \qquad \therefore C_p = 3.062\, pF$$



[3]

Model answer to Q 6(b): Calculated Application of Theory

Resonant method and L-network matching using a shunt inductor. Give two practical advantages for this solution, when compared to that in 6(a).

The +j20 Ω inductive load reactance is resonated out with a 3.248 pF capacitor.

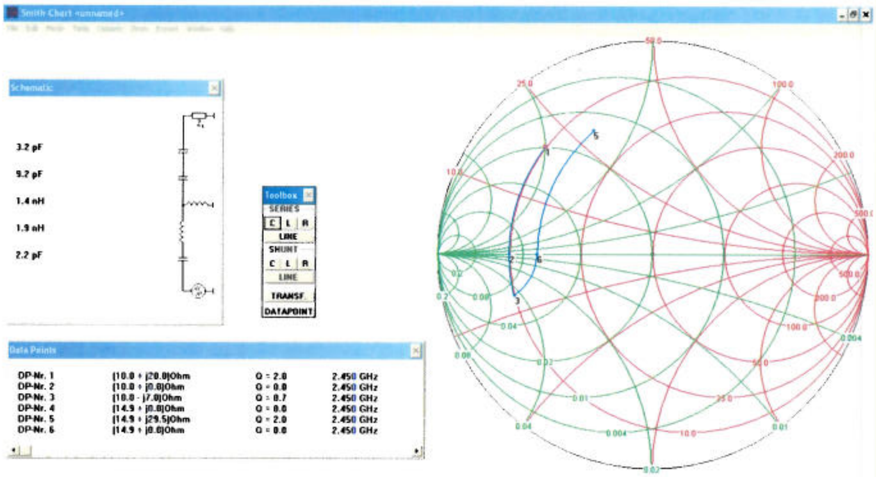
The –j30 Ω capacitive source reactance is resonated out with a 1.949 nH inductance.

The next step is to match the 10 Ω load resistance to the 15 Ω source resistance:

$$Q = \sqrt{\frac{15}{10} - 1} = \frac{1}{\sqrt{2}} \quad \text{and } \omega_o = 2\pi \; 2.45x10^9$$

$$Q = Q_s = \frac{1}{10\omega_o C_s} = \frac{1}{\sqrt{2}} \qquad \therefore C_s = 9.187\, pF$$

$$Q = Q_p = \frac{15}{15\, \omega_o L_p} = \frac{1}{\sqrt{2}} \qquad \therefore L_p = 1.378\, nH$$



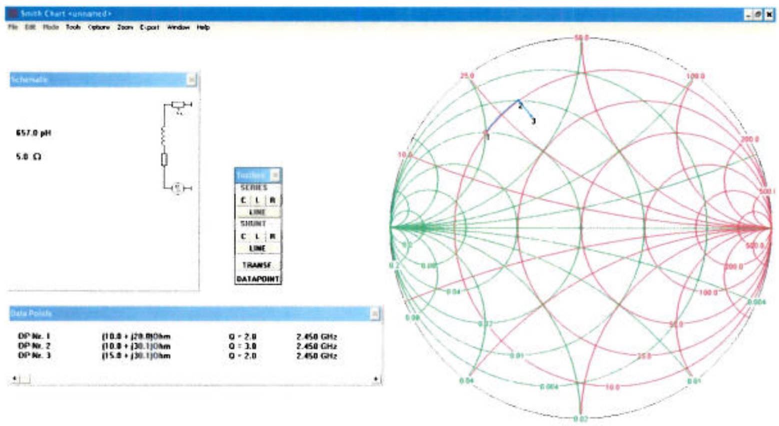
The first advantage is that the inductance associated with via holes can be absorbed into the shunt inductance. In addition, the series capacitor can be combined with the series capacitor used to resonate out the load inductance. This will reduce the overall component count of the matching network by 25%.

[6]

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

Resonant method and series resistor. What is the practical advantage and fundamental problem with this solution?

In principle, the matching network can consist of a series inductance of 0.650 nH (giving a reactance of +j10 Ω) and a series resistance of +j5 Ω. This means that there are only 2 matching network components. BUT, while the output of the source transistor exhibits maximum power transfer, the lossy matching network absorbs power and, therefore, reduced the maximum available power that can be input into the load transistor.



[3]

Model answer to Q 6(d): Calculated Application of Theory

Resonant method and quarter-wavelength transformer.

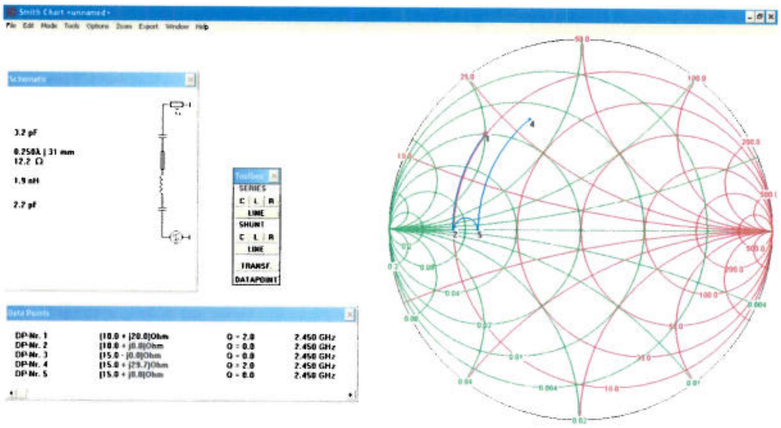
The +j20 Ω inductive load reactance is resonated out with a 3.248 pF capacitor.

The –j30 Ω capacitive source reactance is resonated out with a 1.949 nH inductance.

The next step is to match the 10 Ω load resistance to the 15 Ω source resistance:

The characteristic impedance of the quarter-wavelength transformer is

$$Z_{oTX} = \sqrt{10\ x\ 15} = 12.247\ \Omega$$



[3]

Model answer to Q 6(e): Calculated Application of Theory

With the solution 6(d), for a microstrip transmission line suspended in air 0.635 mm above the ground plane, use simplified expressions for calculating the dimensions of the quarter-wavelength transformer.

Length of the microstrip transformer, l, is $l = \frac{\lambda_0}{4} = \frac{c/f_0}{4} = 30.6\ mm$

Width of the microstrip transformer , W, is:

$$Z_{oTX} \sim \eta_o \left(\frac{h}{W}\right) = 377 \left(\frac{0.365}{W\ [mm]}\right) = 12.247\ \Omega$$
$$\therefore W \sim 19.55\ mm$$

[5]