The Solutions for EE4.18 and AO6, 2018

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

i) Signal/noise ratio becomes more problematic as frequency decreases below 1 GHz, because of the increase in galactic noise. As a result, more signal power is required to maintain a good signal/noise ratio.

[2]

ii) The frequency between 1 GHz and 10 GHz is convenient for commercial exploitation because both galactic and atmospheric noise are very low. Also, cheap lumped-element passive components and high gain active components can be used in this frequency band.

[2]

iii) The water absorption peaks are found at 22 GHz and 183 GHz; while the oxygen absorption peaks are found at 60 GHz and 119 GHz.

[2]

iv) At 38 GHz and 94 GHz there are troughs in the atmospheric attenuation curves. As a result, Line-of-Sight communication links (having wide bandwidths) and directional radar systems (having highly directional antennas) are used in these bands, respectively.

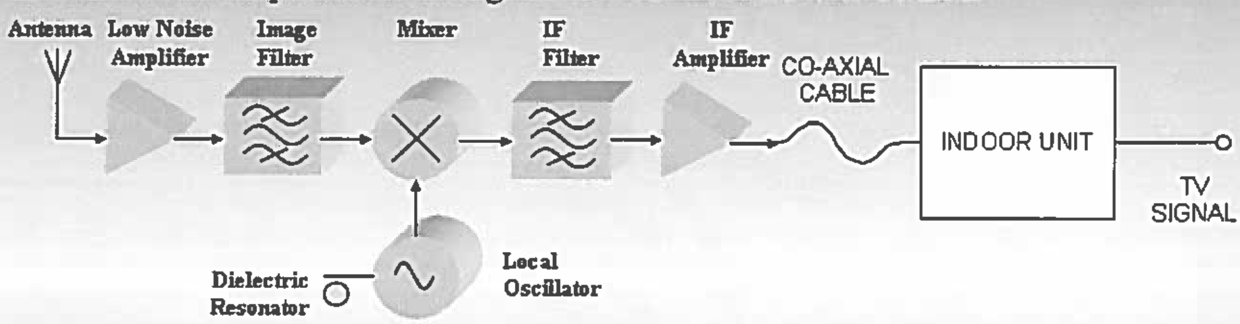
[2]

v) The 60 GHz band has very high atmospheric attenuation. As a result, it is ideal for indoor wireless local area network applications and inter-satellite communication links. The potential interference to other nearby services operating in the same band can be minimized by the inherent drop in signal strength, as distance increases.

[2]

Model answer to Q 1(b): Bookwork

Consider the simplified block diagram of a satellite TV receiver's LNB



- *LNA (NF < 1 dB):* low noise transistors are not standard & matching circuits are lossy
- *RF Filter (50 dB Image Rejection):* poor Q-factors, ∴ poor selectivity & rejection
- *Mixer:* double-balanced mixers can be large, ∴ expensive
- *Oscillator:* high-Q off-chip resonator is required for stability/low phase noise
- *IF Filter:* low frequency components are very large, ∴ expensive
- *IF Amplifier:* cheaper to implement in silicon technology

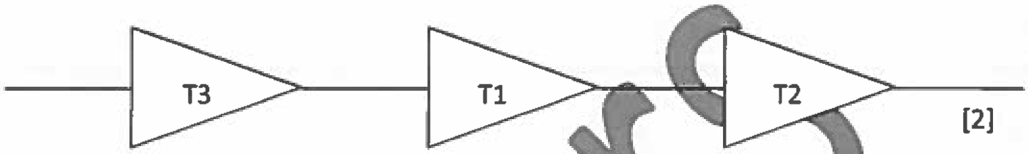
[10]

Model answer to Q 2(a & b): Computed Example

An RF signal of 1 mW is input to a power amplifier with an output of 1 W. Design a suitable power amplifier to give the best overall performance using the following transistor stages.

	$P_{OUT}\|_{MAXLIN}$ [dBm]	P_{DC} [mW]	IP_3 [dBm]
Stage 1	25	600	40
Stage 2	30	2000	40
Stage 3	15	60	40

Table 2.1 Transistor Stage Specifications.
(Note that all the transistors have a perfect impedance matching)



[2]

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]

Model answer to Q 2(c): Textbook Derivation

From first principles, prove that the 3rd order intermodulation log-power gain slope is three times that of the desired output log-power slope.

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

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

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 2(c): Computed Example

In linear operation, if the overall input power drops by 3 dB, what happens to the following:

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 3(a): New Application of Theory

The Michelson interferometer, shown in Figure 3.1, can be analysed as a general two-port network. By inspection of Figure 3.1, write down equations for the effective forward voltage-wave transmission coefficient (S_{21}) and input voltage-wave reflection coefficient (S_{11}) for this passive and reciprocal network. Clearly define all variables used. Hints, the electrical path lengths can be represented by $(k_o dx)$, where $k_o = 2\pi/\lambda$, λ is the wavelength for a monochromatic input source, integer x identifies a particular path and the beam splitter is both symmetrical and reciprocal.

The effective input voltage wave transmission coefficient is given by:

$$\begin{aligned} S_{21} &= \exp(-jk_o d1)\rho_s \exp(-jk_o d3)\rho_m \exp(-jk_o d3)\tau_s \exp(-jk_o d2) \\ &+ \exp(-jk_o d1)\tau_s \exp(-jk_o d4)\rho_m \exp(-jk_o d4)\rho_s \exp(-jk_o d2) \end{aligned}$$

ρ_s = **voltage wave reflection coefficient of the beam splitter**
τ_s = **voltage wave transmission coefficient of the beam splitter**
ρ_m = **voltage wave reflection coefficient of a mirror**

$$\begin{aligned} S_{11} &= \exp(-jk_o d1)\rho_s \exp(-jk_o d3)\rho_m \exp(-jk_o d3)\rho_s \\ &+ \exp(-jk_o d1)\tau_s \exp(-jk_o d4)\rho_m \exp(-jk_o d4)\tau_s \end{aligned}$$

[6]

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

Given: $d1 = d2 = \lambda \quad \therefore \exp(-jk_o d1) = \exp(-jk_o d2) = \exp(-j2\pi) = 1$

And the mirrors are made from perfectly conducting metal, $\therefore \rho_m = -1$

$$S_{21} = -\rho_s \tau_s [\exp(-j2k_o d3) + \exp(-j2k_o d4)]$$

$$S_{11} = -\rho_s^{\,2} \exp(-j2k_o d3) - \tau_s^{\,2} \exp(-j2k_o d4)$$

[3]

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

For this interferometer to function properly, an ideal beam splitter must reflect 50% of any incident power and allow the rest of the power to be transmitted through without attenuation.

(i) Since 50% of the power is reflected and 50% is transmitted through the ideal beam splitter then it must have the following solutions for its forward voltage wave transmission coefficient and input voltage wave reflection coefficient:

$$\tau_s = \rho_s \exp(\pm j\pi/2) = \frac{\exp[j(\angle\rho_s \pm \pi/2)]}{\sqrt{2}} \quad e.g.\ \tau_s = \pm\frac{j}{\sqrt{2}}$$

$$\rho_s = \tau_s \exp(\pm j\pi/2) = \frac{\exp[j(\angle\tau_s \pm \pi/2)]}{\sqrt{2}} \quad e.g.\ \rho_s = \pm\frac{1}{\sqrt{2}}$$

[4]

(ii) For a lossless 2-port network, the beam splitter must satisfy the following in order to obey the conservation of energy principle:

$|\tau_s|^2 + |\rho_s|^2 = 1$ and this is confirmed!

[2]

(iii)

$$S_{21} = \pm\frac{j}{2}\exp(-j2k_o d3)[1 + \exp(+jk_o\delta)]$$

$$S_{11} = \pm\frac{1}{2}\exp(-j2k_o d3)[1 - \exp(+jk_o\delta)]$$

[2]

(iv) For a lossless 2-port network, the interferometer must satisfy the following in order to obey the conservation of energy principle:

$|S_{21}|^2 + |S_{11}|^2 = 1$ and this is confirmed, since:

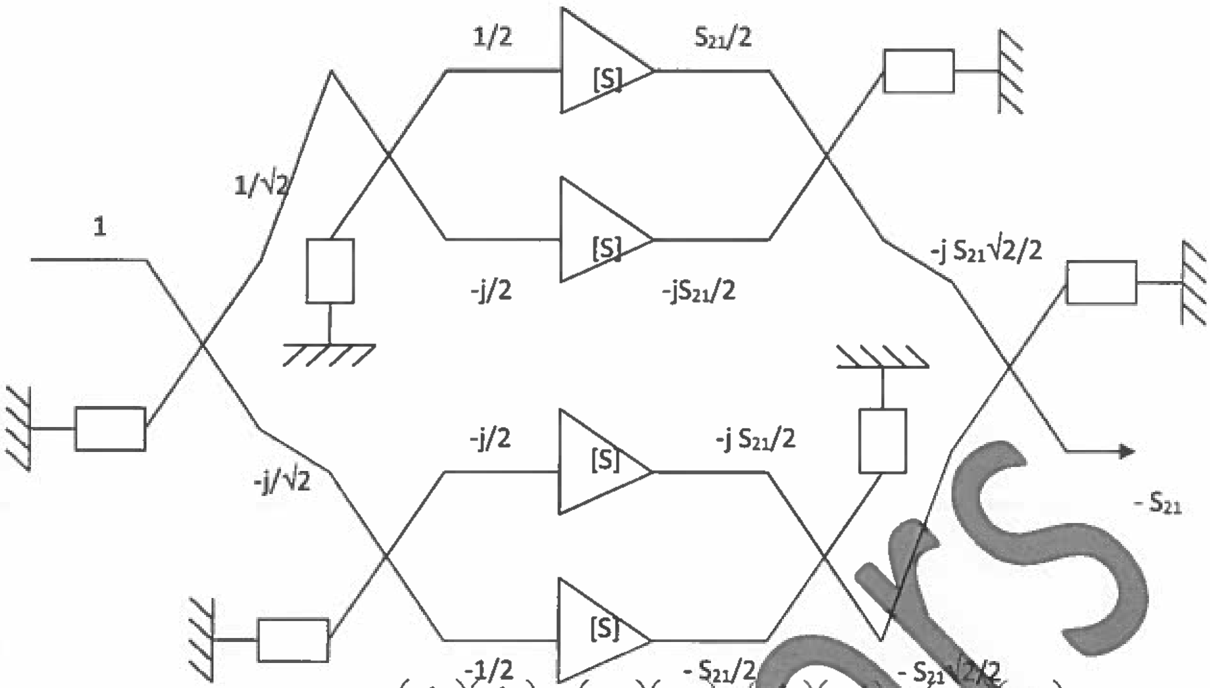
$|S_{21}| = 1$ and $|S_{11}| = 0$ when $\delta = n\lambda$

$|S_{21}| = 0$ and $|S_{11}| = 1$ when $\delta = (n + 1/2)\lambda$

[3]

Model answer to Q 4(a): New application of theory

Draw the topology of a double-balanced amplifier. If 3 dB quadrature couplers are used in conjunction with identical non-ideal single-ended amplifiers, use S-parameter analysis to determine expressions for the overall insertion gain and input return loss. Assume the couplers are perfectly matched to the reference impedance *Zo* and the interconnections between the main components are ideal.



$$S_{21}|_{overall}=\left(\frac{1}{\sqrt{2}}\right)\left(\frac{1}{\sqrt{2}}\right)S_{21}\left(\frac{-j}{\sqrt{2}}\right)\left(\frac{-j}{\sqrt{2}}\right)+\left(\frac{1}{\sqrt{2}}\right)\left(\frac{-j}{\sqrt{2}}\right)S_{21}\left(\frac{-1}{\sqrt{2}}\right)\left(\frac{-j}{\sqrt{2}}\right)+$$

$$\left(\frac{-j}{\sqrt{2}}\right)\left(\frac{1}{\sqrt{2}}\right)S_{21}\left(\frac{-j}{\sqrt{2}}\right)\left(\frac{1}{\sqrt{2}}\right)+\left(\frac{-j}{\sqrt{2}}\right)\left(\frac{-j}{\sqrt{2}}\right)S_{21}\left(\frac{1}{\sqrt{2}}\right)\left(\frac{1}{\sqrt{2}}\right)=-S_{21}$$

$$Insertion\ Loss = 10\log\left\{|S_{21}|^2\right\}$$

$$S_{11}|_{overall}=\left(\frac{1}{\sqrt{2}}\right)\left(\frac{1}{\sqrt{2}}\right)S_{11}\left(\frac{1}{\sqrt{2}}\right)\left(\frac{1}{\sqrt{2}}\right)+\left(\frac{1}{\sqrt{2}}\right)\left(\frac{-j}{\sqrt{2}}\right)S_{11}\left(\frac{-j}{\sqrt{2}}\right)\left(\frac{1}{\sqrt{2}}\right)+$$

$$\left(\frac{-j}{\sqrt{2}}\right)\left(\frac{1}{\sqrt{2}}\right)S_{11}\left(\frac{1}{\sqrt{2}}\right)\left(\frac{-j}{\sqrt{2}}\right)+\left(\frac{-j}{\sqrt{2}}\right)\left(\frac{-j}{\sqrt{2}}\right)S_{11}\left(\frac{-j}{\sqrt{2}}\right)\left(\frac{-j}{\sqrt{2}}\right)=0$$

$$\text{Return } Loss = 10\log\left\{|0|^2\right\}\rightarrow -\infty$$

[10]

<ins>Model answer to Q 4(b): Discussions in Class</ins>

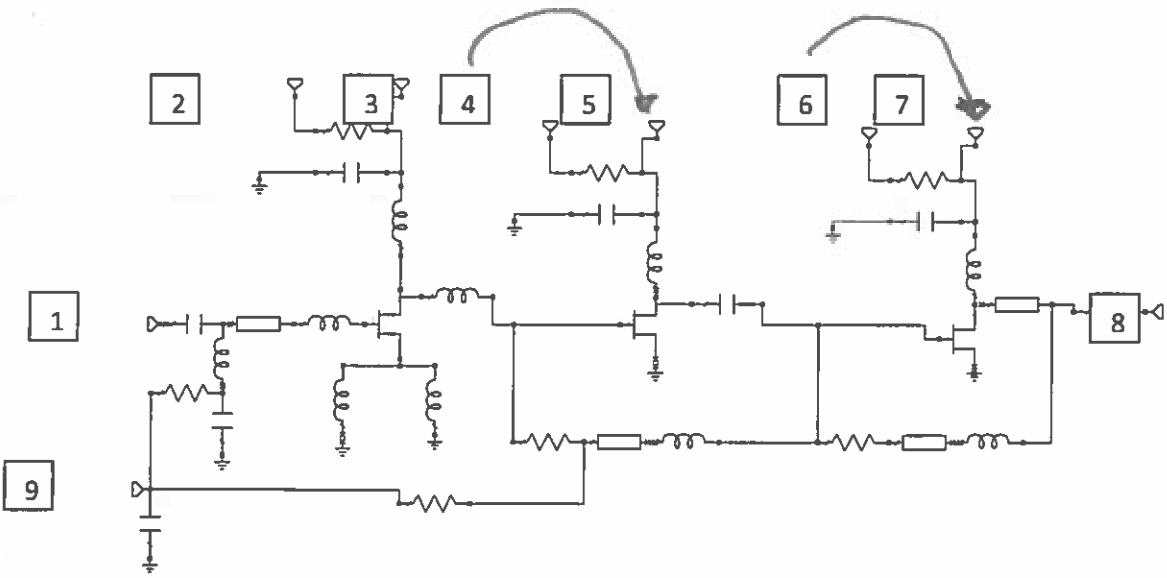
For the topology in 4(a), if the working single-ended amplifiers have a forward voltage wave transmission coefficient of $S_{21} = |10|\angle 35°$, determine the overall insertion gain and input return loss if one of the amplifiers fails, such that $S_{21} = 0$. Assume that there is no change in the input or output impedances of the failed transistor. What is the main application of this topology and what are its advantages and disadvantages when compared to a single-ended amplifier?

The main application of this amplifier is power combining, since the output power is ideally a factor of 4 greater than that of the single-ended amplifier. If one of the single-ended amplifiers fails then $S_{21}|_{overall} = 3S_{21}/4$ and Insertion loss is $10\log\{(3|S_{21}|/4)^2\} = 17.5$ dB, i.e. a drop of 2.5 dB from a fully working amplifier. The input return loss should not change if the impedance of the failed amplifier doesn't change and is, therefore, still minus infinity. Therefore, this type of power combining amplifier is useful because it provides redundancy in the case of failure and also ideal port impedance matching. The disadvantages of this topology is that it requires 4 identical single-ended amplifiers. Also, practical losses in the couplers result in a direct loss in power gain and output efficiency will be significantly reduced.

[10]

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

Draw the equivalent circuit model for the MMIC shown in Figure 5.1, and mark the RF and DC bias ports with the corresponding probe pad numbers shown. Describe the type of amplifier circuit. Hint: if you are uncertain about a component then state any assumptions used.



This is a three-stage low noise amplifier with parallel feedback in the last two stages.

[10]

Model answer to Q 5(b): Bookwork and Discussions in Class

Briefly describe the different range of component technologies used for the transistors, inductors and capacitors, and also state the advantages and disadvantages of these technologies. State what compromises have to be made with the design of MMICs, when compared to HMICs.

This amplifier used metal-semiconductor field-effect transistor (MESFET) technologies. This is the workhorse of microwave GaAs ICs. They have good general performance in gain, power-added efficiency, output power and noise. However, they do not have superior performance over HEMT devices. The inductors are all of the planar 2D spiral type. This given high inductance per unit area but poor noise and bandwidth performance. The capacitors are all of the metal-insulator-metal type. These can give high capacitance values per unit area, but are prone to low yield due to pinhole short circuits.

Most MMIC devices have to be tailored to volume production and tend not to give state-of-the-art performance. This introduces compromises for the MMIC designer: (i) this can be a serious problem for LNA and PA design; (ii) special devices (e.g. Gunn diodes, PIN switches, and hyperabrupt varactor diodes) are rarely used in MMIC processes; (iii) FET switch is a poor substitute for the PIN diode and the HEMT millimetre-wave; (iv) oscillator will have a low output power compared with a Gunn diode; (v) most compromises can be absorbed into the specifications of the system design, and good communications between the circuit designer and systems designer are very beneficial to the final product

[5]

Model answer to Q 5(c): Bookwork and Discussions in Class

Briefly comment as to why the complexity of the full equivalent circuit model is much more than the circuit derived in 5(a) and explain why circuit modelling alone is not sufficient if a significant reduction in the chip area is required.

In practice, most of the components identifies in 5(a) should be represented by many equivalent circuit model elements. As a result, while there are only 3 MESFETs, 8 capacitors, 10 spiral inductors, 7 resistors, 4 microstrip interconnections identifies, the circuit could be represented by hundreds of individual parasitic elements. In order to reduce the chip size the components have to be placed closer together. However, by doing this, the electromagnetic coupling between components cause adverse interactions that can only be modelled sufficiently using 3D electromagnetics simulation packages.

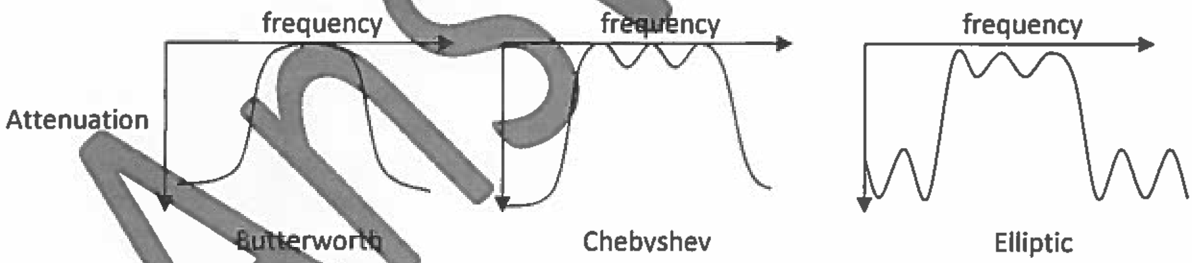
[5]

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

With the use of simple illustrations for the attenuation against frequency curves, describe the differences between Butterworth, Chebyshev and Elliptical-function filters. Also, comment on the group delay characteristics for these filters.

Butterworth filters have a maximally flat frequency response in the passband. Chebyshev filters exhibit ripples in the passband only; the number depends on the filter order. Neither the Butterworth nor Chebyshev filters have ripples in the stopband. Elliptical filters have the sharpest roll-off attenuation characteristics of all three types, with the Chebyshev having sharper roll-off than the Butterworth. The elliptic filter exhibits ripples in the stopband (transmission zeros) as well as ripples in the passband (transmission poles).

Chebyshev and elliptic filters have higher group delays than the Butterworth, particularly near the 3 dB cut-off frequency and should be avoided in applications where pulse distortion is critical



[5]

Model answer to Q 6(b): Computed Example

Given prototype low-pass filter attenuation curves and tables for the corresponding normalised element values (see attached sheets), design an L-C lumped-element band-pass filter that meets the following specifications:

Centre Frequency, f_o	500 MHz
3 dB Bandwidth, B	50 MHz
Attenuation Bandwidth	100 MHz
Pass-Band Ripple (Peak-to-Peak)	0.1 dB
Stop-Band Attenuation	45 dB
Input Impedance, R_{IN}	100 Ω
Output Impedance, R_{OUT}	50 Ω

Determine f/fc = B/BW_{STOP} = 2.

From curves: Butterworth with 0.0~~0~~1 dB ripples need >7^{th} order. → where? can't find curve

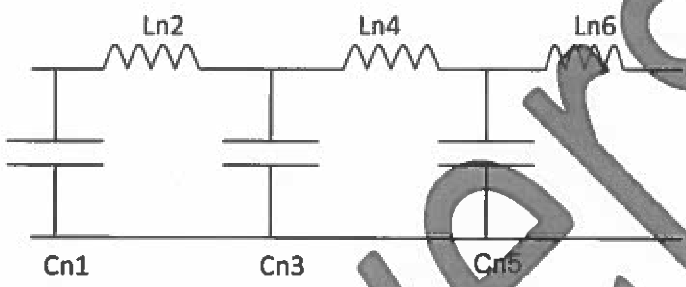
Chebyshev with 0.01 dB ripples need 6^{th} order for 46 dB rejection having f/fc = 2.

Chebyshev with 0.10 dB ripples need 6^{th} order for 50 dB rejection having f/fc = 2.

This last option is best, as it has good out-of-band attenuation margin.

For Rin/Rout = Rs/R_L = 2, the normalised prototype low pass filter coefficients are:

Cn1	Ln2	Cn3	Ln4	Cn5	Ln6
0.414	3.068	0.958	3.712	0.979	2.794



To de-normalize shunt capacitors and inductors:

$$C_p = \frac{Cn}{2\pi f B R_L} \quad L_P = \frac{R_L B}{2\pi f_c^2 Cn}$$

To de-normalize series inductors and capacitors:

$$L_s = \frac{Ln R_L}{2\pi B} \quad C_S = \frac{B}{2\pi f_o^2 Ln R_L}$$

Cp1 = 26.36 pF   Lp1 = 3.84 nH
Cs2 = 0.21 pF   Ls2 = 488.29 nH
Cp3 = 61.08 pF   Lp3 = 1.66 nH
Cs4 = 0.71 pF   Ls4 = 590.78 nH
Cp5 = 62.33 pF   Lp5 = 1.63 nH
Cs6 = 0.23 pF   Ls6 = 444.68 nH

[15]