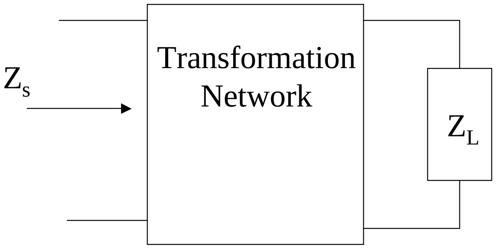
3A - Impedance Transformation and Impedance Matching

References

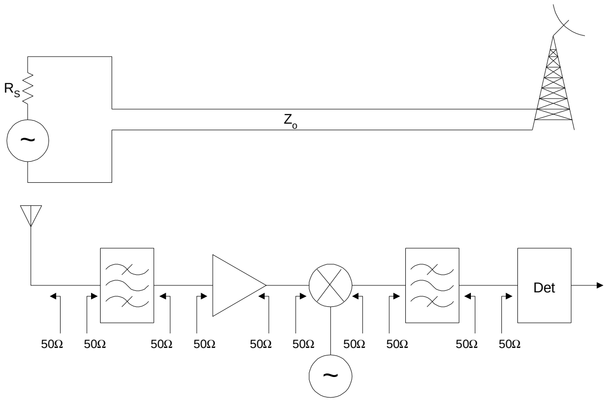
- [1] R. Ludwig, P. Bretchko, "RF circuit design - Theory and applications", 2000 Prentice-Hall.
- [2] D.M. Pozar, "Microwave engineering", 2nd edition, 1998 John-Wiley & Sons.
- [3] R.E. Collin, "Foundation for microwave engineering", 2nd edition, 1992, McGraw-Hill.

Impedance Transformation

- An impedance transformation network is a two-port network that when connected in series with an impedance Z_L at one port, will result in Z_s being seen on another port.
- Z_L is usually not equal to Z_s (otherwise there will be no need for transformation). Z_s is known as the image impedance of Z_L.
- We immediately notice that the transformation network is a 2-port network.



Impedance Transformation and Matching



Why Impedance Tuning is Needed?

- Maximum power is delivered when load is matched to the Tline (assuming generator is matched).
- Impedance matching on sensitive receiver components (antenna, low-noise amplifier etc.) improves the signal-to-noise ratio of the system.
- Impedance matching in a power distribution network (such as antenna array feed network) will reduce amplitude and phase errors.

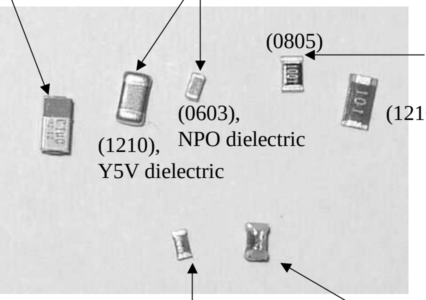
Types of Transformation Network

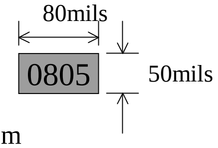
- Single lumped element (either L or C)
- Dual lumped elements (L impedance matching network)
- Triple lumped elements (Pi or T impedance matching network)
- More lumped elements (ladder type)
- Distributed elements (consists of section of Tlines)
- Hybrid - Consists of both Tline and lumped elements

Impedance Transformation Using Lumped Elements

- Lumped components such as surface mounted device (SMD) inductor and capacitor can be easily purchased nowadays.
- SMD capacitors have a range from 0.47pF to greater than 10000pF. With tolerance less than ± 5% and operating temperature between -55°C to 125°C.
- SMD inductors have a range from 1.0nH to greater than 4000nH. With tolerance from ± 5% to± 10%, operating temperature from -40°C to 125°C and Q factor from a minimum of 15 to greater than 45.
- The inductors come in a variety of form, from coil-type, thin-film, to spiral inductors mounted in SMD package. Self-resonance frequency ranges from 200MHz (coil type) for L=2200nH to greater than 5GHz for L<100nH (thin-film).

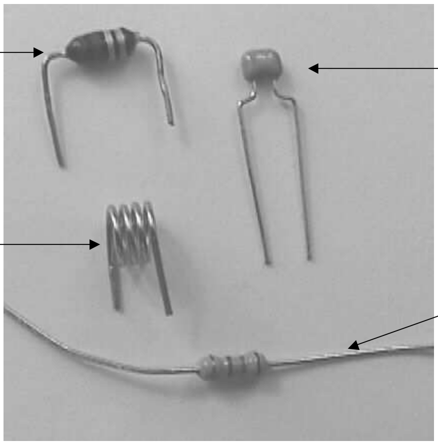
Ultra High Frequencies Passive Components (>250MHz)





Another size from IPC (international package consortium www.ipc.org)
- 0402, even smaller than 0603.
- 1210, bigger than 0805.

Medium Frequency Passive Components (up to 250MHz)



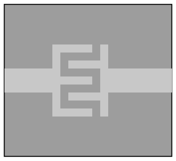
Coil inductor with Ferrite core

Multilayer ceramic capacitor

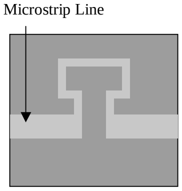
Coil inductor with air core

Carbon/metal film resistor

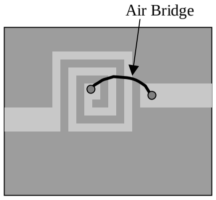
Passive Lumped Components for Incorporation into PCB and other Substrates



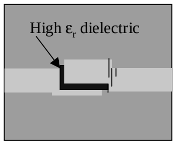
Interdigital Capacitor



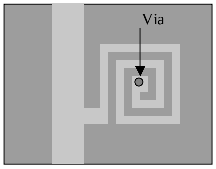
Series Single-Loop Spiral Inductor



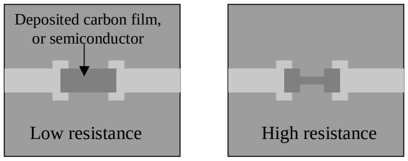
Series Multi-Loop Spiral Inductor



Metal-Insulator-Metal (MIM) Capacitor

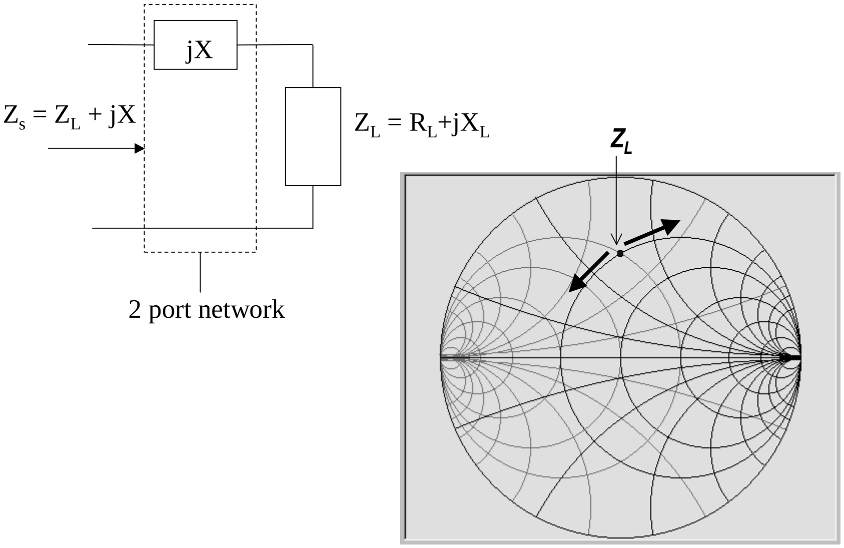


Shunt Multi-Loop Spiral Inductor

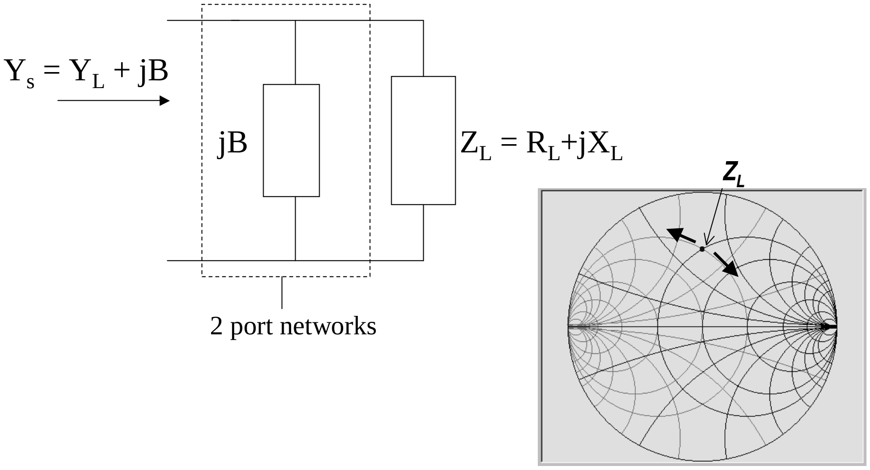


Resistors

Single Lumped Element Transformation Network

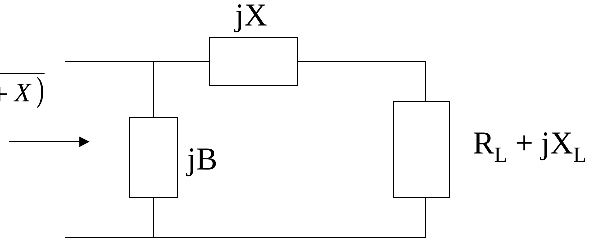


Single Lumped Element Transformation Network Cont...



Dual Lumped Elements Transformation Network

$$Y_s = \frac{1}{Z_s} = jB + \frac{1}{R_L + j\left(X_L + X\right)}$$



If $Z_s = R_s + jX_s$ is given, we could solve for X and B by equating the real and imaginary parts:

This configuration is only applicable for $R_s > R_L$

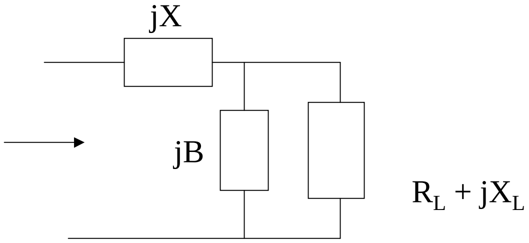
$$X = -X_L \pm \sqrt{R_L\left(R_s - R_L\right) + \frac{R_L}{R_s} X_s^2}$$

$$B = \frac{R_s - R_L}{R_s X_L + R_L X_s + R_s X}$$

(1.1)

Dual Elements Transformation Network Cont...

$$Z_s = jX + \frac{1}{jB + \frac{1}{R_L + jX_L}}$$



If $Z_s = R_s + jX_s$ is given, again we could solve for X and B by equating the real and imaginary parts:

This configuration is only applicable for $R_L > R_s$

$$X = X_s \pm \sqrt{R_s\left(R_L - R_s\right) + \frac{R_s}{R_L} X_L^2}$$

$$B = \frac{R_s - R_L}{R_L X_s + R_s X_L - R_L X}$$

(1.2)

Example 1

- Transform Z_L=100 + j80 to 50 + j40 at 410MHz.

$$X = X_s + \sqrt{R_s\left(R_L - R_s\right) + \frac{R_s}{R_L} X_L^2} = 115.498$$

$R_L > R_s$

$$B = \frac{R_s - R_L}{R_L X_s + R_s X_L - R_L X} = 0.014$$

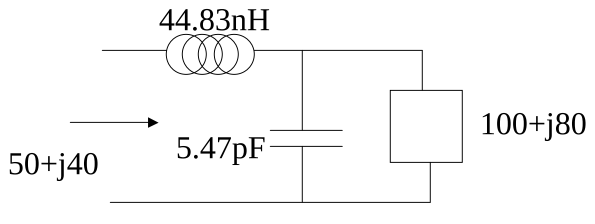
Since X is +ve, an inductor can be used to realize it:

$$L = \frac{X}{2\pi\left(410 \times 10^6\right)} = 44.83nH$$

Since B is +ve, a capacitor can be used to realize it:

$$C = \frac{B}{2\pi\left(410 \times 10^6\right)} = 5.468\,pF$$

Example 1 Cont...



At 410MHz Only!

Exercise 1

- Transform Z_L= 50 + j100 to 300 - j10 at 900MHz using 2 lumped element matching networks.

Example 2

- Repeat Example 1 using Smith chart.

Exercise 2

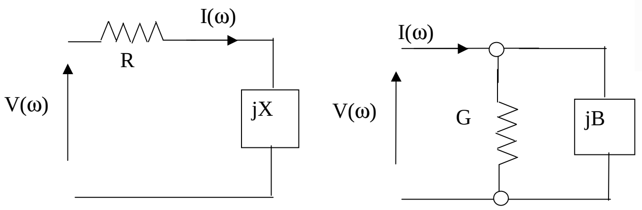
- Repeat Example 2 using Smith chart.

Q Factor

- The Q Factor of a series or parallel impedance is defined by:

$$Q = \omega \frac{[\text{Maximum energy stored}]}{\text{power dissipation}}$$

Taken from the notes of EEN2072, Communication Electronics, F. Kung 2000.

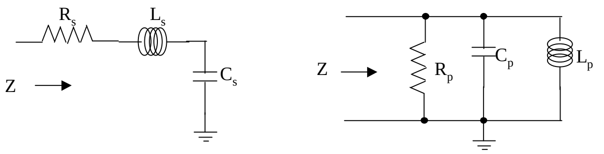


$$Q_s = \frac{|X|}{R} \qquad \textbf{(1.3a)}$$

$$Q_p = \frac{|B|}{G} \qquad \textbf{(1.3b)}$$

Series & Parallel RLC Network

Taken from the notes of EEN2072, Communication Electronics, F. Kung 2000.

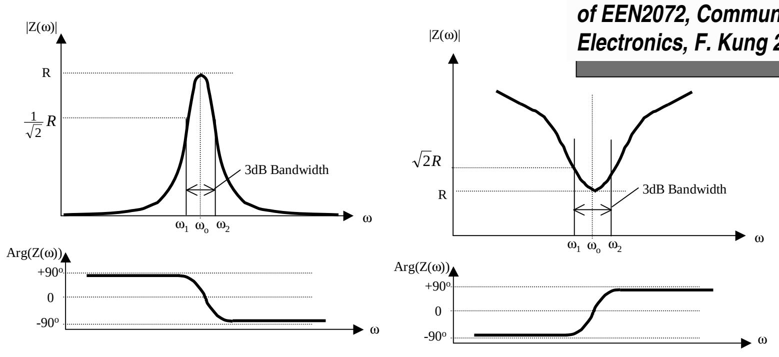


Parameter	Series RLC network	Parallel RLC network
Input impedance	$R_s + j\omega L_s + \frac{1}{j\omega C_s}$	$\left(\frac{1}{R_P} + \frac{1}{j\omega L_P} + j\omega C_P\right)^{-1}$
Resonance frequency	$\omega_o = \frac{1}{\sqrt{L_s C_s}}$	$\omega_o = \frac{1}{\sqrt{L_p C_p}}$
Quality factor, Q at resonance frequency	$Q_s = \frac{\omega_o L_s}{R_s} = \frac{1}{\omega_o R_s C_s}$	$Q_p = \frac{R_P}{\omega_o L_p} = \omega_o R_P C_p$
Bandwidth BW (note that this is just an approximation)	$\frac{\omega_o}{Q_s}$	$\frac{\omega_o}{Q_p}$

(1.4)

Frequency Response of Series & Parallel RLC Network

Taken from the notes of EEN2072, Communication Electronics, F. Kung 2000.



Parallel RLC
R⇧ Q⇧
R⇩ Q⇩

Series RLC
R⇧ Q⇩
R⇩ Q⇧

Poles and Zeros of Series and Parallel RLC Network

Extra !

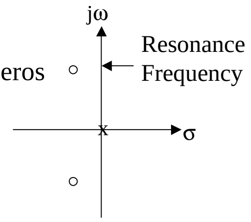
For series RLC: $Z(\omega)=R+j\omega L+\frac{1}{j\omega C}$

Resonance frequency is the frequency where input impedance to a passive RLC network becomes real.

$$=\frac{1+j\omega RC-\omega^2 LC}{j\omega C}$$

← 2 complex conjugate zeros

← 1 pole on jω axis

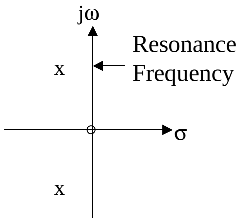


For parallel RLC: $Z(\omega)=\left(\frac{1}{R}+\frac{1}{j\omega L}+j\omega C\right)^{-1}$

1 zero on jω axis →

$$=\frac{j\omega RL}{R-\omega^2 RLC+j\omega L}$$

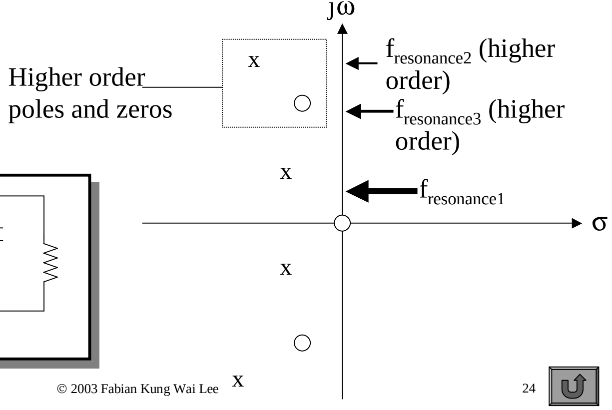
2 complex conjugate poles

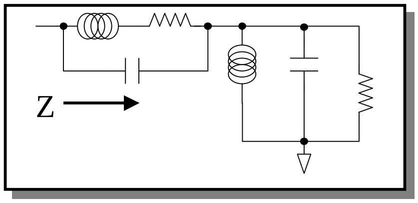


Resonance Frequency of Higher Order Systems

Extra !

- For a system with more than one L and C, there will be higher order poles and zeros. These will distort the location of the fundamental resonance frequency of the network and introduce higher order resonance frequencies.

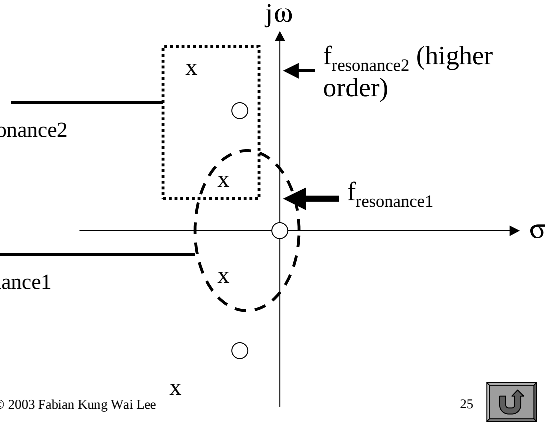




Extra !

Resonance Frequency of Higher Order Systems Cont...

- Since each resonance frequency is still due to the dominant poles and zeros, the concept of Q factor with regards to 3dB bandwidth can still be applied to higher order network.

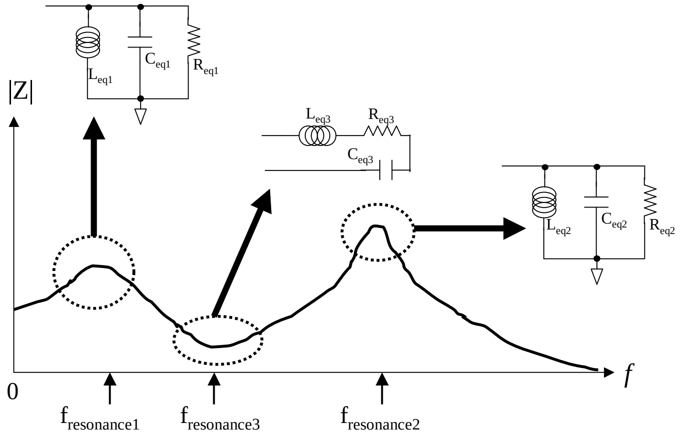


Dominant poles and zero contributing to $f_{resonance2}$

Dominant poles and zero contributing to $f_{resonance1}$

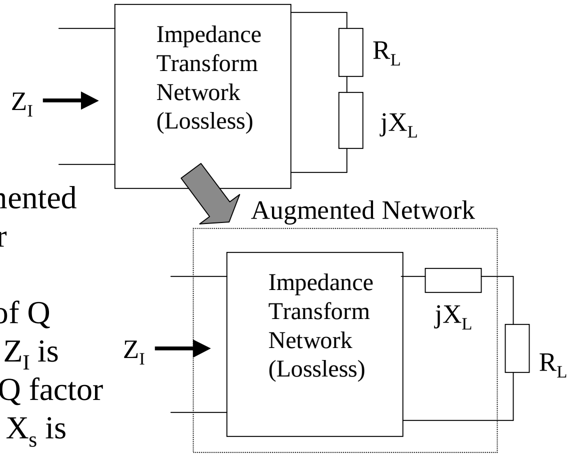
Extra !

Resonance Frequency of Higher Order Systems Cont...



Extra !

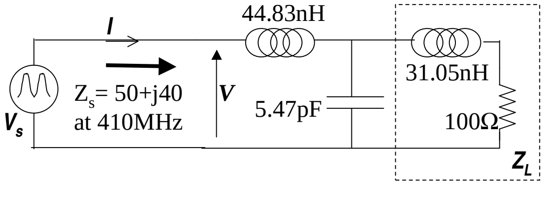
Impedance Transformation Network as a Resonating Network



If $Z_I = R_s$, then the augmented network is actually under resonance during normal operation. The concept of Q factor can be applied. If Z_I is complex, the concept of Q factor can still be applied if the X_s is small.

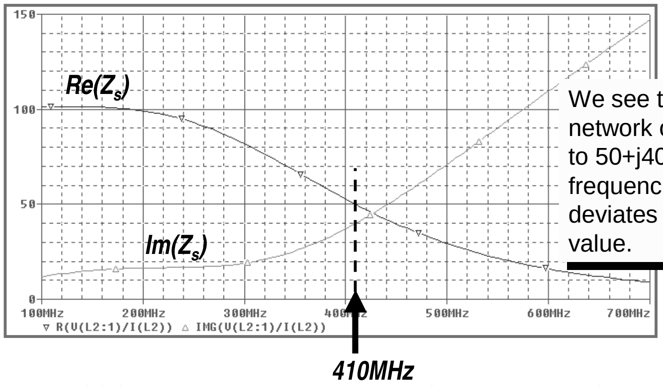
Bandwidth of the Matching Network

- Suppose in Example 1 the load Z_L is actually given by an inductor in series with a resistor, so that at 410MHz we obtain Z_L = 100+j80.



- We input the above schematic in a circuit simulator (PSPICE) and run a frequency sweep (change the frequency of the source V_s while measure I & V) from 100MHz to 800MHz.

Bandwidth of the Matching Network Cont...

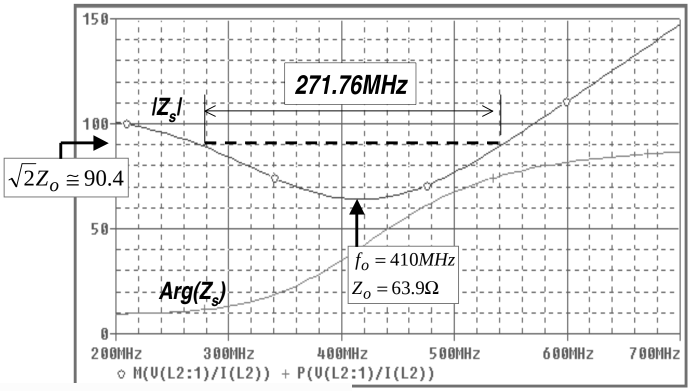


We see that the matching network only transform Z_L to 50+j40 at 410MHz, at other frequencies, Z_s gradually deviates from the desired value.

- Within a range of frequencies near to the operating frequency f_o=410MHz, $Z_s=R_s + jX_s$ is quite near the desired value. We will call this range of frequency the bandwidth (BW) of the transformation network.

Bandwidth of the Matching Network Cont...

- To examine this closer, we plot Z_s in terms of its magnitude and phase.

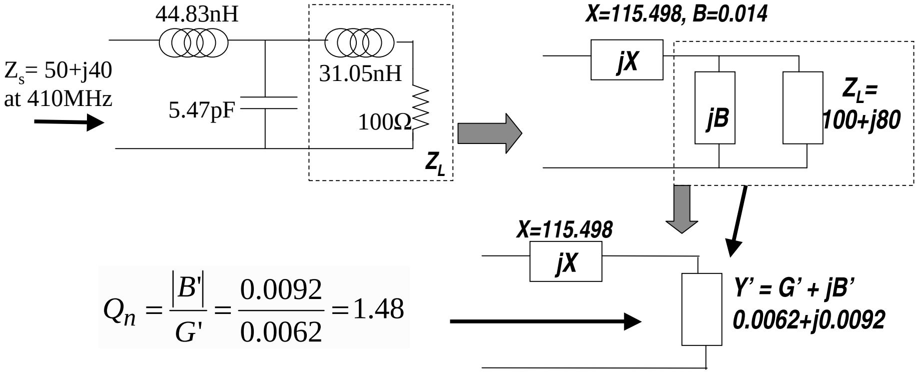


Following the theory of series RLC network, we define the 3dB BW as the range of freq. Where $|Z_s|$ is less than $\sqrt{2}Z_o$, where Z_o is the magnitude of the impedance at the operating freq. f_o = 410MHz.
We see that the 'measured' BW is:
BW = 271.76MHz

$|Z_s|$ and Arg(Z_s) is very close to the pattern of series RLC circuit near operating frequency f_o

Bandwidth of the Matching Network Cont...

- Now consider the circuit of Example 1 again. We could compute a quantity known as the Nodal Q factor, Q_n as follows:



$$Q_n = \frac{|B'|}{G'} = \frac{0.0092}{0.0062} = 1.48$$

Bandwidth of the Matching Network Cont...

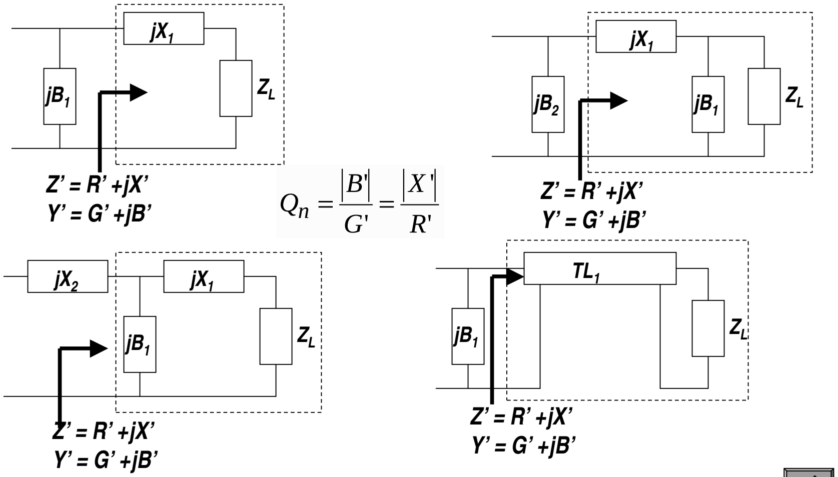
- We could calculate the BW of the system using the equation in (1.4): $BW \cong \frac{f_o}{Q_n} = \frac{410MHz}{1.48} = 277MHz$

271.76MHz

- Surprisingly this is quite near the measured value using simulation. Both measured and calculated BW using this method will match even closer if Z_s is real, or X_s=0. This applies to all lumped element transformation network as well (3 elements or more).
- When X_s is not 0, there is an error, the larger $|X_s|$, the greater the error. However this does illustrate that we could in general compare the BW of various transformation network merely by calculating Q_n.
- High Q_n denotes narrow BW, low Q_n denotes wide BW.

Nodal Q Factor, Q_n

- Q_n for a few favorite transformation networks.



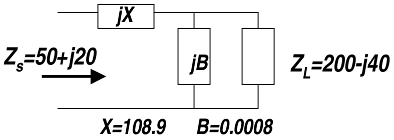
$$Q_n = \frac{|B'|}{G'} = \frac{|X'|}{R'}$$

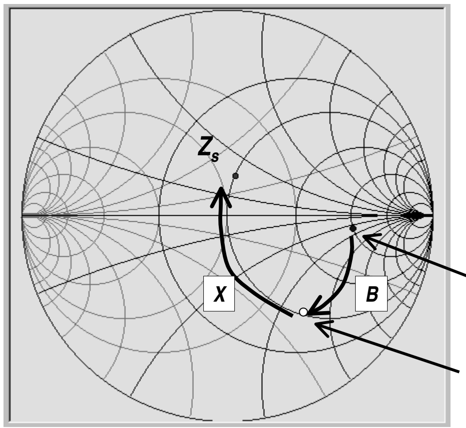
Nodal Q Factor, Q_n Cont...

- The previous slides only illustrate the concept of using nodal Q factor to estimate and compare bandwidth between transformation networks heuristically. A more formal argument and derivation can be found from various materials:
  - R. Ludwig, P. Bretchko, “RF circuit design - Theory and applications”, 2000, Prentice-Hall.
  - J.R. Smith,”Modern communication circuits”, 2nd edition 1998, McGraw-Hill.
  - EEN3096 (Communication Electronics), year 2000 of MMU.
  - Unpublished works of F. Kung, 2003.

Example 3

- Transform the load Z_L = 200 - j40 to 50 + j20 at 2.4GHz. Find the nodal Q factor and estimate the bandwidth of the circuit. Use Smith chart to aid the design.





$$L = \frac{108.9}{2\pi(2.4\times10^9)} = 7.22nH$$

$$C = \frac{0.008}{2\pi(2.4\times10^9)} = 0.53pF$$

$$Q_n = \frac{89.23}{50.58} = 1.764$$

$$BW = \frac{2.4GHz}{1.764} = 1.36GHz$$


Z

R: 50.58

X: -89.23

Constant Q_n Circles

- Q_n depends on the point location on the Smith chart. We could joint all points on the Smith chart giving a similar Q_n to form a curve or locus. It happens that this locus is a circle, known as Constant Q_n circles.
- The center and radius for the circles can be derived as follows.
- From section Section 2.2 on Smith chart:

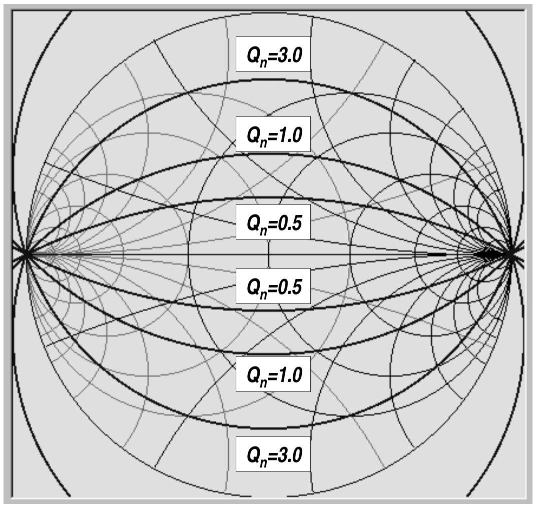
$$r + jx = \frac{1+U+jV}{1-U-jV} = \underset{r}{\boxed{\frac{1-U^2-V^2}{(1-U)^2+V^2}}} + j\underset{x}{\boxed{\frac{2V}{(1-U)^2+V^2}}}$$

$$Q_n = \frac{|x|}{r} = \frac{2|V|}{1-U^2-V^2} \longrightarrow U^2 + \left(V \pm \frac{1}{Q_n}\right)^2 = 1 + \frac{1}{Q_n^2}$$

$$\Gamma_{center} = 0 \mp j\frac{1}{Q_n}$$

$$Radius = \sqrt{1+\frac{1}{Q_n{}^2}} \quad \textbf{(1.5)}$$

Constant Q_n Circles Cont...



Q_n	Radius	$1/Q_n$
0.5	2.2360	2.000
1.0	1.4142	1.000
2.0	1.1180	0.500
3.0	1.0541	0.333
5.0	1.0198	0.200

Limitation of 2 Lumped Elements Network

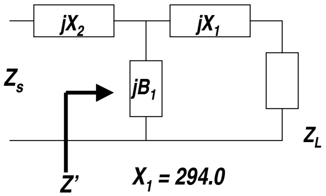
- By now it is obvious of the limitation of the 2 elements network. For instance in Example 3 there are only two ways to transform Z_L=200 - j40 to Z_s=50 + j20.
- Therefore we cannot control the nodal Q factor of 2 elements network, it is determined by the values of Z_L and Z_s.
- Using an extra element, we have extra degree of freedom and we can control the value of Q_n in addition to performing impedance transformation/matching. This is the advantage of using the T or Pi networks.

Three or More Lumped Elements Transformation Network

- For more than 3 lumped elements, analytical method such as shown in previous slides is very cumbersome to apply.
- It is more easier to perform 3 elements transformation network design with the aid of Smith Chart.
- As oppose to 2 elements network, 3 or more elements network do not suffer from blind spot. It can transform any passive load Z_L to any required impedance value.

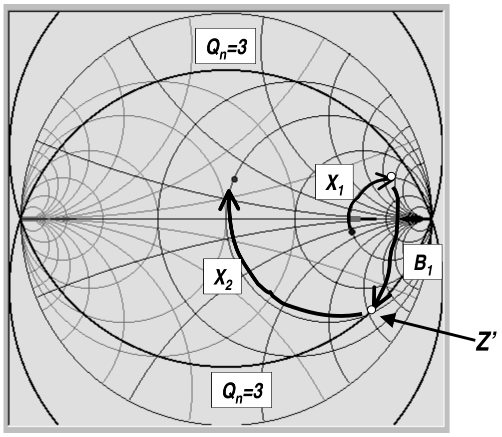
Example 4

- Repeat Example 3 using 3 elements transformation network, either T or Pi, with the aid of Smith chart. It is required that Q_n be equal to 3. (Z_L=200-j40, Z_s=50+j20).



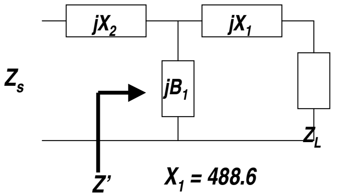
$X_1 = 294.0$
$B_1 = 0.0083$
$X_2 = 174.6$

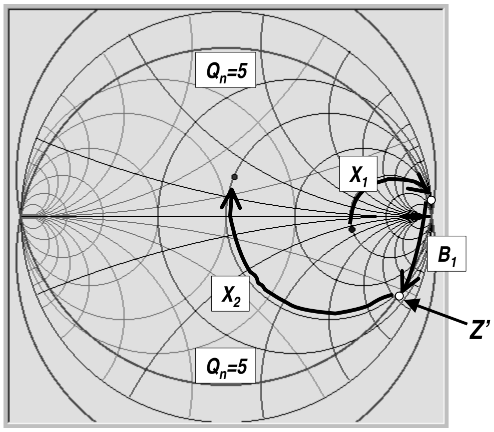
L_1= 19.5nH
C_1 = 0.55pF
L_2 = 11.58nH



Example 5

- Repeat Example 4 using 3 elements transformation network, either T or Pi, with the aid of Smith chart. It is required that Q_n be equal to 5. (Z_L=200-j40, Z_s=50+j20).

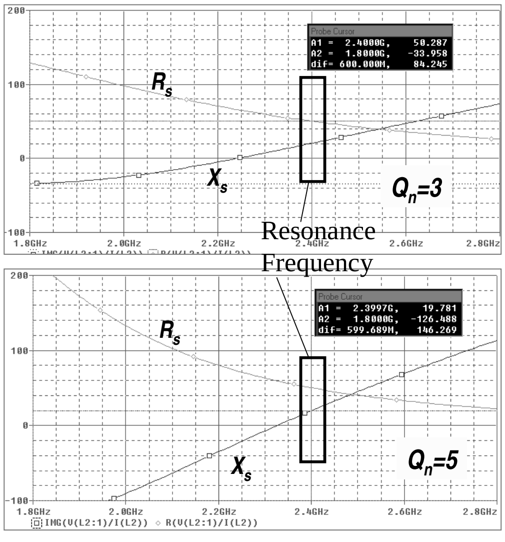




$X_1 = 488.6$
$B_1 = 0.00584$
$X_2 = 261.0$

L_1= 32.4nH
C_1 = 0.387pF
L_2 = 17.3nH

Z_s Versus f from Simulation with PSPICE



```
A1 =   2.4000G,     50.287
A2 =   1.8000G,    -33.958
dif= 600.000M,      84.245
```

```
A1 =   2.3997G,     19.781
A2 =   1.8000G,   -126.488
dif= 599.689M,     146.269
```

Both circuits from Example 4 and 5 are fed into PSPICE. AC simulation is run from 1.8GHz to 2.8GHz and the results are compared. It is seen that the T network with higher nodal Q factor has narrower BW, characterized by more rapid deviation from f_o = 2.4GHz.

Exercise 3

- Repeat Example 5 using 3 elements T transformation network, with the aid of Smith chart. It is required that Q_n be equal to 1. (Z_L=200-j40, Z_s=50+j20). Can you synthesize the T network ? Suggest a solution to this.

Exercise 4

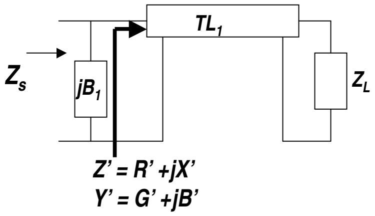
- Repeat Example 5 using 3 elements Pi transformation network, with the aid of Smith chart. It is required that Q_n be equal to 3. (Z_L=200-j40, Z_s=50+j20). Can this impedance transformation be realized ? Discuss the result.

Pros & Cons of Lumped Element Network

- Lumped element network is compact, small in size.
- Suitable for use up to frequency of 2.5GHz.
- Not every values of inductance and capacitance are available.
- Stability, value changes with temperature.
- Tolerance of components.
- Difficult to tune.
- Higher cost.

Distributed Transformation Network

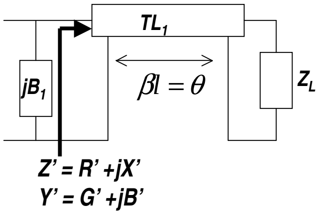
- Single Stub transformation network.

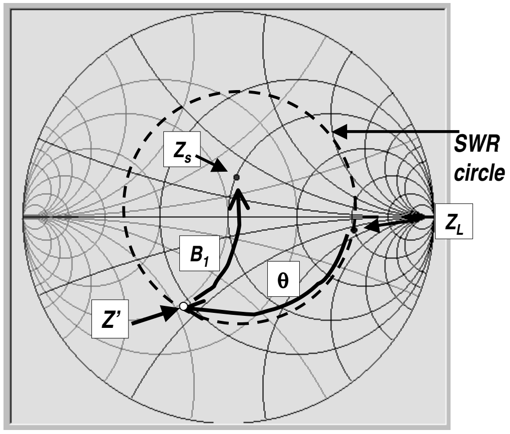


- jB can be implemented using a Tline with open/short circuit at one end. Can also use lumped elements such as SMD capacitors. In this case the network is known as hybrid network.
- No blind spot.

Example 6

- Transform the load $Z_L = 200 - j40$ to $50 + j20$ at 2.4GHz. Find the nodal Q factor and estimate the bandwidth of the circuit. Use Smith chart to aid the design. Synthesize the circuit.





$$\theta = \beta l = 1.013$$
$$B = -0.0356$$

Example 6 Cont...

- Use a microstrip line to implement the circuit, Z_c = 50Ohm. Dielectric constant = 4.7, and d=1.6mm.
- Step 1 - Synthesize Tline.
- From Example 5, Section 3.0, we see that the required W must be 2.88mm.

$$\varepsilon_{eff} = 3.55$$
$$\beta = \omega\sqrt{\varepsilon_o \varepsilon_{eff}\,\mu}$$
$$= 2\pi\left(2.4\times10^9\right)\sqrt{3.55\varepsilon_o\mu_o} = 94.77$$
$$l = \frac{\theta}{\beta} = \frac{1.013}{94.77} = 0.011\,\text{m} = 1.1\text{cm}$$

Example 6 Cont...

- Step 2 - Synthesize jB_1.
- We can use an inductor for B_1:

$$L = \frac{1}{2\pi\left(2.4\times10^{9}\right)\cdot 0.0356} = 1.863nH$$

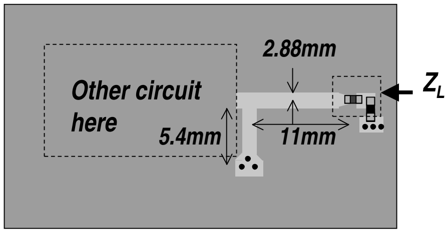
- Or we can use another short circuit Tline to generate B_1:

$$Z_{in}(l) = jZ_c \tan(\beta l) = \frac{1}{-jB} = j\left(\frac{1}{B}\right)$$

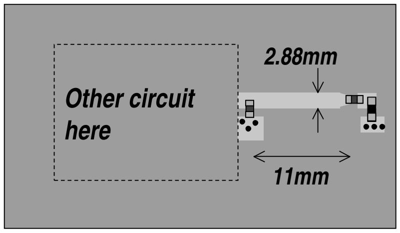
$$l = \frac{1}{\beta}\tan^{-1}\left(\frac{1}{Z_c B}\right) = \frac{1}{94.77}\tan^{-1}\left(\frac{1}{0.0356\times50}\right) = 0.0054\,\mathrm{m}$$

Example 6 Cont...

- Thus the final circuit...

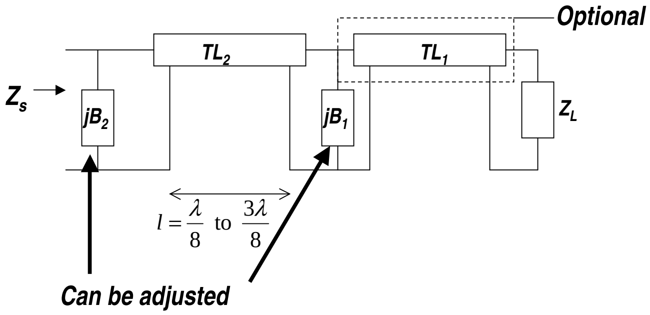


Or...

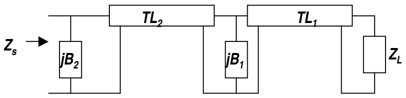


Double-Stub Distributed Network

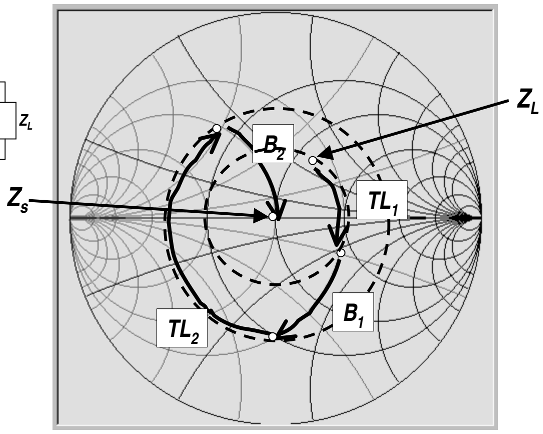
- The single-stub network suffers from the disadvantage of requiring a variable length of Tline between the load and the stub. This may not be a problem for fixed transformation network, but would pose some difficulty if an adjustable tuning network is desired.
- To overcome this disadvantage a double-stub transformation network is used.



Double-Stub Distributed Network Cont...

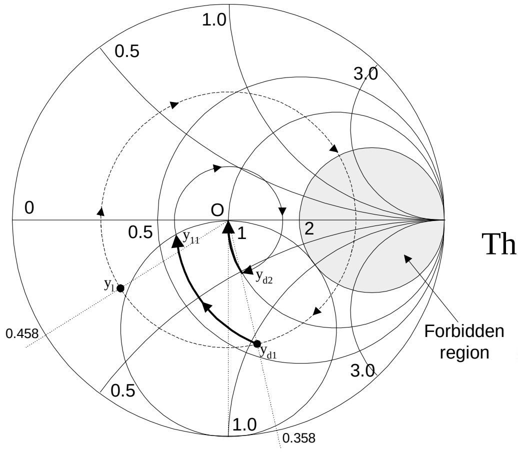


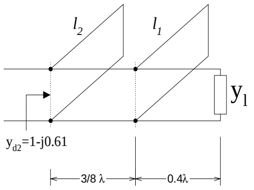
Suppose we want to transform Z_L to Z_s=50Ohm.

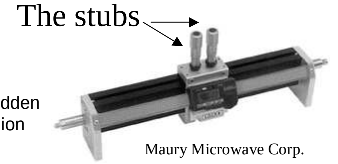


Double-stub Matching Cont...

- Double-stub matching using waveguide:





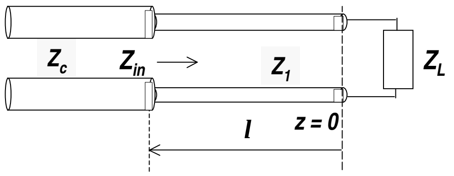


Maury Microwave Corp.

Quarter Wave Transformer

- A quarter wave transformer is a simple and useful circuit for matching a real load impedance to a transmission line. An additional feature is that it can be extended to multisection design for broader bandwidth.
- Consider a terminated lossless Tline again, using (1.7) of "2 - Microwave Network Analysis" and letting $l = \frac{\lambda}{4}$:

$$l = \frac{\lambda}{4} \quad \Rightarrow \quad \beta l = \frac{2\pi}{\lambda} \cdot \frac{\lambda}{4} = \frac{\pi}{2}$$



$$Z_{in}(l) = Z_1 \frac{Z_L + jZ_1 \tan\left(\frac{\pi}{2}\right)}{Z_1 + jZ_L \tan\left(\frac{\pi}{2}\right)} = \frac{Z_1^{\ 2}}{Z_L}$$

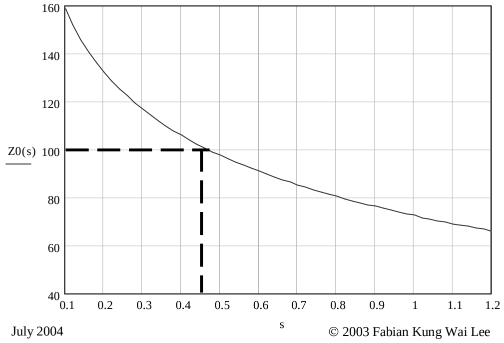
$$\Rightarrow \quad Z_{in}(l) = \frac{Z_1^{\ 2}}{Z_L} \quad \textbf{(1.6)} \qquad \Rightarrow \quad Z_c = \frac{Z_1^{\ 2}}{Z_L}$$

Example 7

- Design a quarter wave transformer to transform a 200Ω load into 50Ω at 2.4GHz using a microstrip line constructed on a dielectric with dielectric constant 4.2 and thickness 1.6mm.

$$Z_1 = \sqrt{Z_c R_L} = \sqrt{50 \times 200} = 100$$

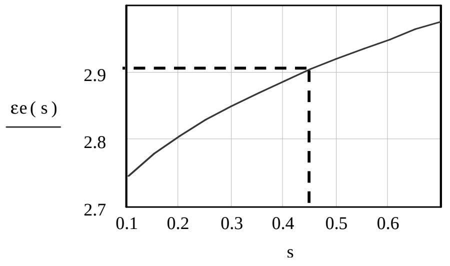
Using the microstrip design equations of “1 - Advance Transmission Line”:



$$\frac{w}{h} = 0.45$$

$$w = 0.45 \times 1.6 = 0.72\text{mm}$$

Example 7 Cont...



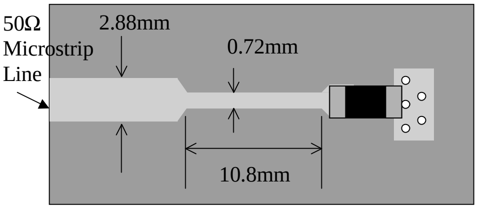
$$\varepsilon_e = 2.91$$

$$v_p = \frac{1}{\sqrt{\varepsilon_e \varepsilon_o \mu_o}} = 1.75743 \times 10^8$$

$$\beta = \frac{\omega}{v_p} = 85.81$$

$$\beta l = \frac{\pi}{2}$$ For quarter wavelength

$$l = \frac{\pi}{2\beta} = 0.0183 = 10.8\text{mm}$$



Quarter Wave Transformer Cont...

- One drawback of the quarter wave transformer is that it can only match a real load impedance, a complex load impedance can always be transformed to a real impedance.
- At the operating frequency f_o , the electrical length of the matching section is $\lambda_o/4$. But at other frequencies the length is different, so a perfect match is no longer obtained. So the quarter wave transformer has a limited bandwidth, like other transformation methods.
- Writing Z_{in} as: $t = \tan\theta, \quad \theta = \beta l$

$$Z_{in} = Z_1 \frac{Z_L + jZ_1 t}{Z_1 + jZ_L t} \qquad \textbf{(1.7)}$$

Extra ! BW of Quarter Wave Transformer

- Using (1.6) and (1.7): $\Gamma = \frac{Z_{in} - Z_c}{Z_{in} + Z_c} = \frac{Z_L - Z_c}{Z_L + Z_c + j2t\sqrt{Z_c Z_L}}$

$$|\Gamma| = \frac{|Z_L - Z_c|}{\left[(Z_L + Z_c)^2 + 4t^2 Z_o Z_L\right]^{1/2}}$$

$$= \frac{1}{\left\{((Z_L + Z_c)/(Z_L - Z_c))^2 + \left[4t^2 Z_c Z_L/(Z_L - Z_c)^2\right]\right\}^{1/2}}$$

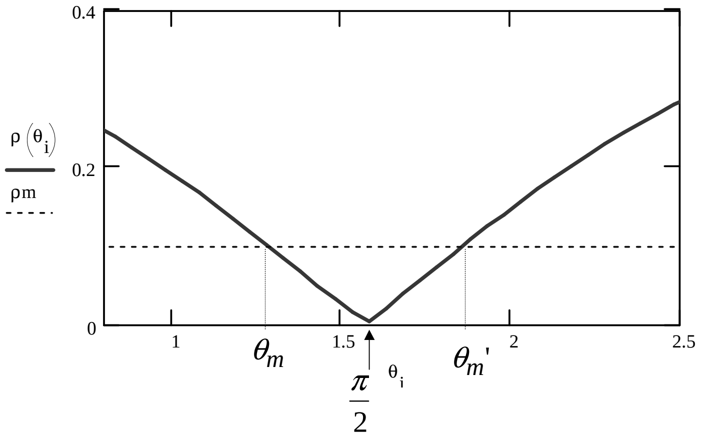
$$= \frac{1}{\left\{1 + \left[4Z_c Z_L/(Z_L - Z_c)^2\right]\sec^2\theta\right\}^{1/2}} \qquad \textbf{(1.8a)}$$

- For frequency near f_o , $l \cong \lambda_o/4$, $\sec^2\theta >> 1$, and this simplifies to:

$$|\Gamma| = \rho \cong \frac{|Z_L - Z_c|}{2\sqrt{Z_c Z_L}}|\cos\theta| \qquad \textbf{(1.8b)}$$

BW of Quarter Wave Transformer Cont...

Extra !



BW of Quarter Wave Transformer Cont...

Extra !

- If we set a maximum value, ρ_m , of the reflection coefficient magnitude that can be tolerated, putting this into (1.8a) and solve for θ_m :

$$\cos\theta_m = \frac{\rho_m}{\sqrt{1-\rho_m^{\,2}}}\cdot\frac{2\sqrt{Z_c Z_L}}{|Z_L - Z_c|}$$

- Assuming TEM or quasi-TEM mode:

$$\theta_m = \beta l = \overset{\beta}{\boxed{\frac{2\pi f_m}{v_p}}}\;\overset{\text{Quarter wavelength}}{\boxed{\frac{v_p}{4 f_o}}} = \frac{\pi f_m}{2 f_o}$$

$$\Rightarrow f_m = \frac{2\theta_m f_o}{\pi}$$

- And the bandwidth is given by:

$$\Delta f = 2|f_o - f_m|$$

$$= 2 f_o - \frac{4}{\pi}\cos^{-1}\left[\frac{\rho_m}{\sqrt{1-\rho_m^{\,2}}}\cdot\frac{2\sqrt{Z_L Z_c}}{|Z_L - Z_c|}\right] \qquad \textbf{(1.8c)}$$

Final Note on Quarter Wave Transformer

- In the previous analysis the reactance associated with the discontinuities must be taken into account.
- Proper compensation technique must be used.

Example 8

- Design a single-section quarter wave transformer to match a 10Ohm load to a 50Ohm Tline, at f_o=2.4GHz. Determine the bandwidth for which VSWR<1.3. Use the microstrip line of Example 6 to realize it.

$$Z_1 = \sqrt{50.10} = 22.361$$

$$\rho_m = \frac{VSWR - 1}{VSWR + 1} = 0.13$$

From example 6

$$\downarrow$$

$$\beta\big|_{2.4\text{GHz}} = 94.77 \Rightarrow \lambda\big|_{2.4\text{GHz}} = 2\pi / \beta = 6.6cm \qquad \frac{\lambda\big|_{2.4\text{GHz}}}{4} = 1.7cm$$

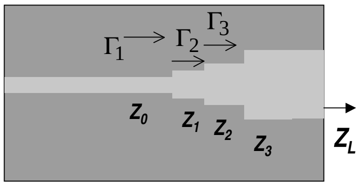
$$\left[2 - \frac{4}{\pi}\cdot \text{acos}\left[\frac{\rho\text{m}}{\sqrt{1-\rho\text{m}^2}}\cdot\frac{\left(2\cdot\sqrt{50\cdot 10}\right)}{|10-50|}\right]\right]\cdot\text{fo} = 4.511\times 10^8 \Rightarrow \Delta f = 451.2MHz$$

Pros & Cons of Distributed Network

- Easy to fabricate and incorporate into microwave circuit. Utilize the PCB itself.
- Cheap and stable, good tolerance if mechanical tolerance is properly controlled.
- Easier to tune than lumped element network.
- Modern manufacturing facilities use LASER to trim the transmission line dimension during tuning.
- At low frequency, the length of the Tline can be unwieldy large.

Increasing Bandwidth of Distributing Matching Network

- For applications requiring more bandwidth than a single quarter wave section can provide, multi-section transformers can be used.



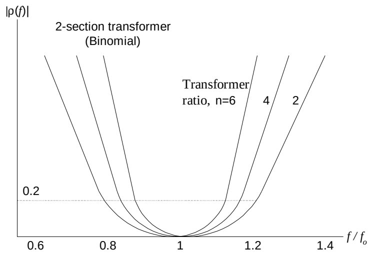
Z_n must increase or decrease monotonically
Z_L must be real.

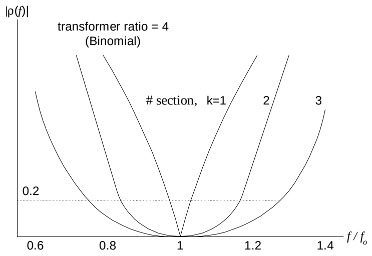
The theory of multisection transformer is beyond the time frame of this course. Interested students please refer to Section 5.10-12 of reference [3].

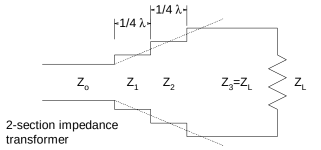
- We can synthesize any desired reflection coefficient response as a function of frequency, by properly choosing Γ_n and using enough sections.

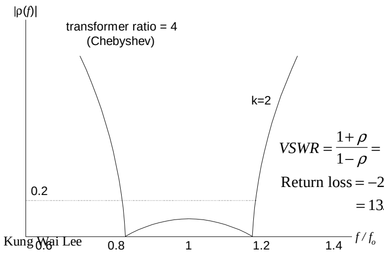
Extra !

Multisection Quarter-Wave Transformer



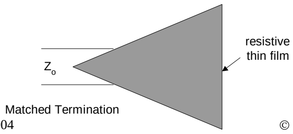






$$VSWR = \frac{1+\rho}{1-\rho} = 1.5$$

$$\text{Return loss} = -20\log\rho$$
$$= 13.98dB$$



Extra !

Binomial and Chebyshev Transformers

Binomial Transformer

- ***impedance of consecutive 1/4 wave lines are proportional to binomial coefficients***
- ***gives maximally flat passband characteristic***

Chebyshev Transformer

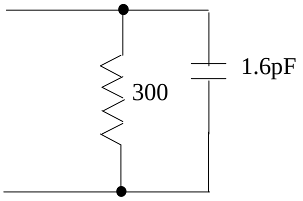
- ***wider bandwidth than Binomial Transformer for the same number of 1/4 wave sections***
- ***ripple over passband***

Tapered Transition

- ***characteristic impedance varies continuously in a smooth fashion***
- ***taper length of 0.5 - 1.5 wavelength required***

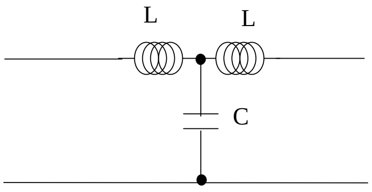
Question 1 (16 marks)

- Consider the parallel RC network below. Design a 2-element lumped network that will transform the RC network into 50Ω at 900MHz.
- Also determine the nodal Q factor and estimate the operating bandwidth of the transformation network.



Question 2 (14 marks)

- A T network is shown below. Derive, in terms of ω, L and C:
- (a) the ABCD matrix of the network.
- (b) the S matrix of the network, take $Z_{o1}=Z_{o2}=Z_o$ to be 50.



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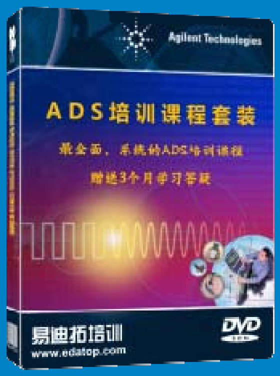

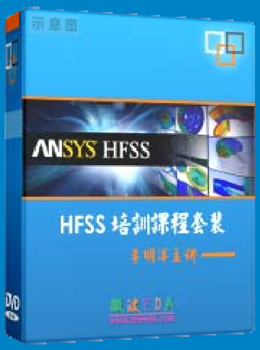

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