Lesson 1: Introduction on Filters and Backgrounds on Distributed Circuits

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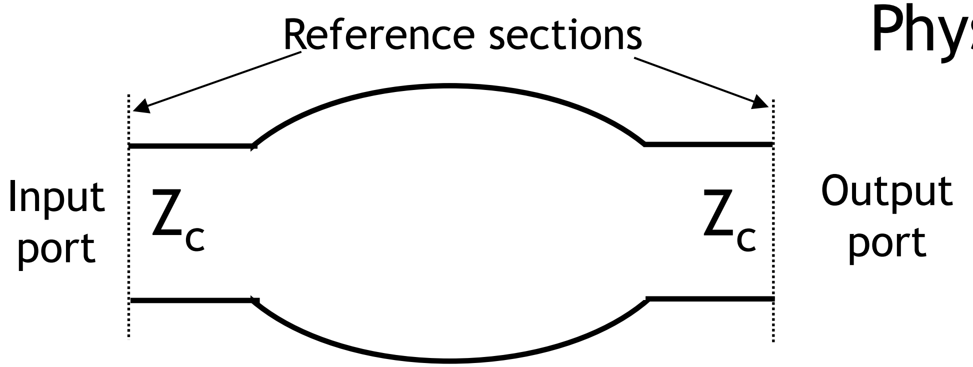
Representation of a physical structure

Physical Structure

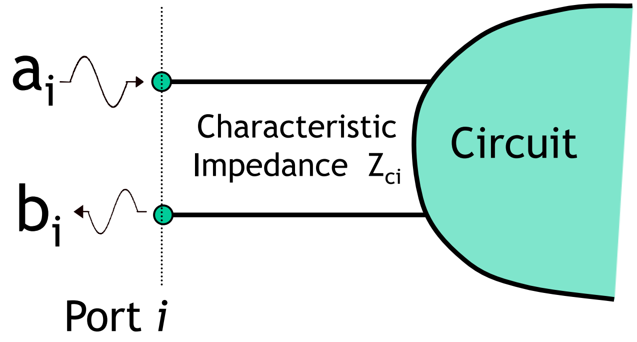


Given a physical structure, the <u>*Scattering Parameters*</u> (**S** matrix) allow a representation of its behavior in terms of the waves propagating on the transmission lines that connect the structure with the external world.

Each line is characterized by its characteristic impedance (Z_c), the phase velocity (v) and a reference section (defined *port*) at which the structure is observed.

The physical structure must be linear to allow the S parameters characterization.

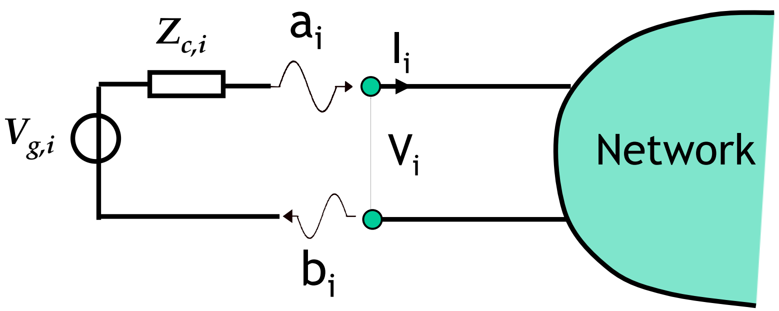
Power Wave definition



(1/2)|a_i|^2: Incident power wave = Available power from a source with impedance Z_{ci}

(1/2)|b_i|^2: Reflected power wave = Difference between the available power and the power absorbed by the port (i.e. flowing into the port)

Definition of conventional *V* and *I*



$$V_{g,i} = Z_{c,i} I_i + V_i$$

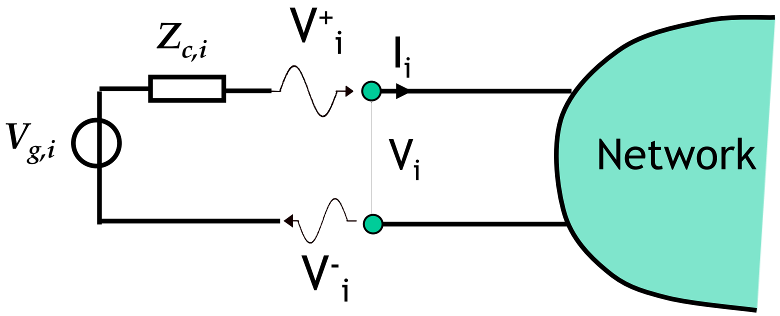
$$P_{IN,i} = \frac{1}{2} \mathrm{Re}\left\{V_i \cdot I_i^*\right\}$$

$$P_{av,i} = \frac{\left|V_{g,i}\right|^2}{8 \cdot \mathrm{Re}\left\{Z_{c,i}\right\}} = \frac{\left|V_i + Z_{c,i} \cdot I_i\right|^2}{8 \cdot \mathrm{Re}\left\{Z_{c,i}\right\}} = \frac{1}{2}\left|a_i\right|^2 \quad \Rightarrow \quad a_i = \frac{V_i + Z_{c,i} \cdot I_i}{2\sqrt{\mathrm{Re}\left\{Z_{c,i}\right\}}} = I_i \frac{Z_i + Z_{c,i}}{2\sqrt{\mathrm{Re}\left\{Z_{c,i}\right\}}}$$

$$P_{IN,i} = \frac{1}{2} \mathrm{Re}\left\{V_i \cdot I_i^*\right\} = \frac{1}{2}\left(\left|a_i\right|^2 - \left|b_i\right|^2\right) \quad \Rightarrow \quad b_i = \frac{V_i - Z_{c,i}^* \cdot I_i}{2\sqrt{\mathrm{Re}\left\{Z_{c,i}\right\}}} = I_i \frac{Z_i - Z_{c,i}^*}{2\sqrt{\mathrm{Re}\left\{Z_{c,i}\right\}}}$$

$$\left(a_i = \frac{I_i + Y_{c,i} \cdot V_i}{2\sqrt{\mathrm{Re}\left\{Y_{c,i}\right\}}}, \quad b_i = \frac{-I_i + Y_{c,i}^* \cdot V_i}{2\sqrt{\mathrm{Re}\left\{Y_{c,i}\right\}}} \right)$$

Power and Voltages waves



$$V_i = V_i^+ + V_i^-, \quad I_i = I_i^+ + I_i^-$$

$$V_i^+ = Z_{c,i} I_i^+, \qquad V_i^- = -Z_{c,i} I_i^-$$

Power waves:

$$a_i = \frac{V_i + Z_{c,i} \cdot I_i}{2\sqrt{\operatorname{Re}\{Z_{c,i}\}}}, \quad b_i = \frac{V_i - Z_{c,i}^* \cdot I_i}{2\sqrt{\operatorname{Re}\{Z_{c,i}\}}}$$

Voltage waves:

$$V_i^+ = \frac{V_i + Z_{c,i} \cdot I_i}{2}, \quad V_i^- = \frac{V_i - Z_{c,i} \cdot I_i}{2}$$

Absence of reflected wave:

Conjugate matching

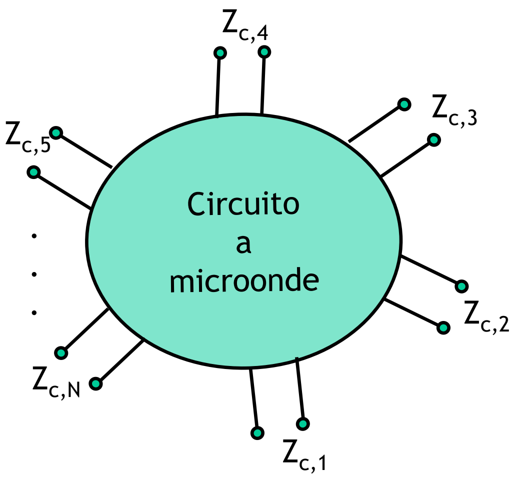
$$Z_i = Z_{c,i}^*$$

Matching

$$Z_i = Z_{c,i}$$

Coincide for $Z_{c,i}$ real

Generalized Scattering Matrix



For a linear circuit, incident and reflected waves are linearly related:

$$b_1 = s_{11}a_1 + s_{12}a_2 + ... + s_{1N}a_N$$
$$b_2 = s_{21}a_1 + s_{22}a_2 + ... + s_{2N}a_N$$
$$..$$
$$b_N = s_{N1}a_1 + s_{N2}a_2 + ... + s_{NN}a_N$$

In matrix form:

$$\underline{\mathbf{b}} = \underline{\underline{\mathbf{S}}} \cdot \underline{\mathbf{a}} \qquad \underline{\underline{\mathbf{S}}} = \begin{pmatrix} s_{11} & \cdots & s_{1N} \\ \vdots & \ddots & \vdots \\ s_{N1} & \cdots & s_{NN} \end{pmatrix}$$

Meaning of S parameters

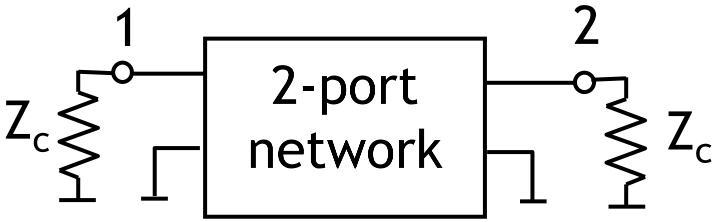
$$s_{ii} = \left.\frac{b_i}{a_i}\right|_{a_{k\neq i}=0}$$

➡ Reflection coefficient at port i when the other ports are connected to their reference impedances Z_{cj} (*matched*)

$$s_{ij} = \left.\frac{b_i}{a_j}\right|_{a_{k\neq j}=0}$$

➡ Transmission coefficient from port j to port i with the other ports matched. Note that $|s_{ij}|^2$ rappresents the transducer power gain between the two ports

S matrix of 2-port passive network



Equivalent Representation

S_{11}, S_{22} = Reflection at the ports
S_{21}, S_{12}=Transmission between the ports

Passive structures imply: $|S_{i,j}|<1$
Moreover, for reciprocal structures: $S_{12}=S_{21}$
Lossless structures implies: $|S_{11}|^2+|S_{12}|^2=1$

In filters application transmission parameters are replaced by their inverse (>1). In decibel it has:

$$\text{Attenuation (dB)} = -20 \log_{10}(|S_{21}|)$$

A very general definition of RF Filters

“A RF filter is a 2-port junction exhibiting a selective frequency behavior in the transmission from the input port to the output port”

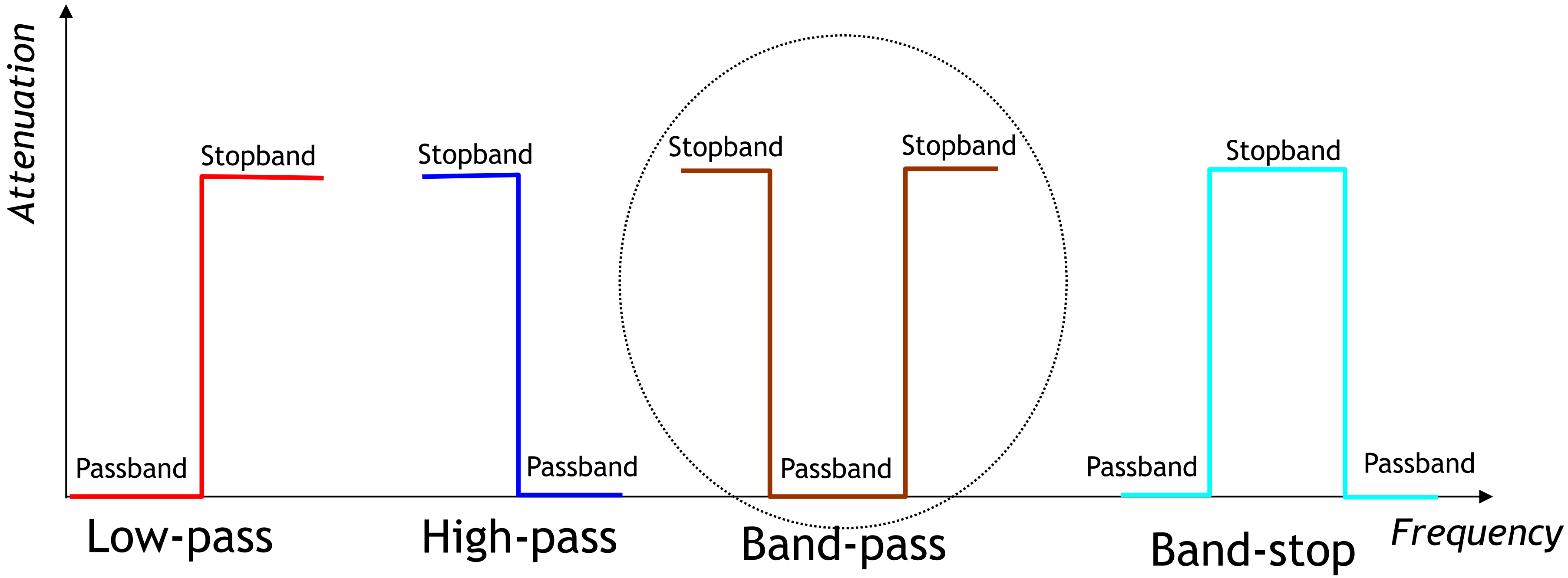
Passband(s): Frequency band(s) which is transferred from input to output without attenuation (ideally)

Stopband(s): Frequency band(s) where the transmission is blocked. In general, the signal rejection is obtained <u>by reflection</u>

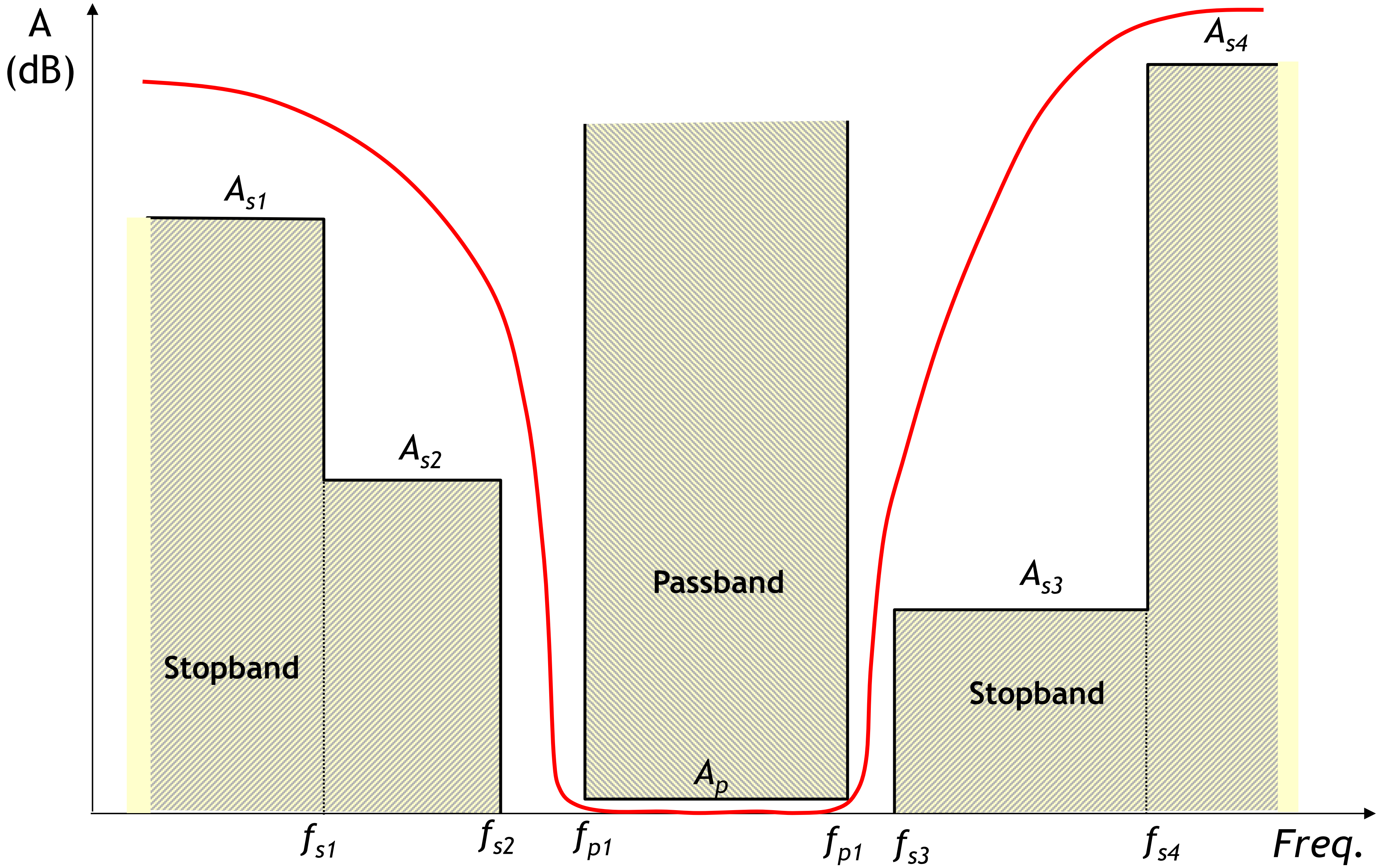
Basic Classification

Filter are classified according to the number and location of passbands and stopbands. The most basic classification considers 4 filters classes:

Low-pass, High-pass, Band-pass, Band-stop



Specification of filter requirements: the Attenuation Mask



Attenuation and matching

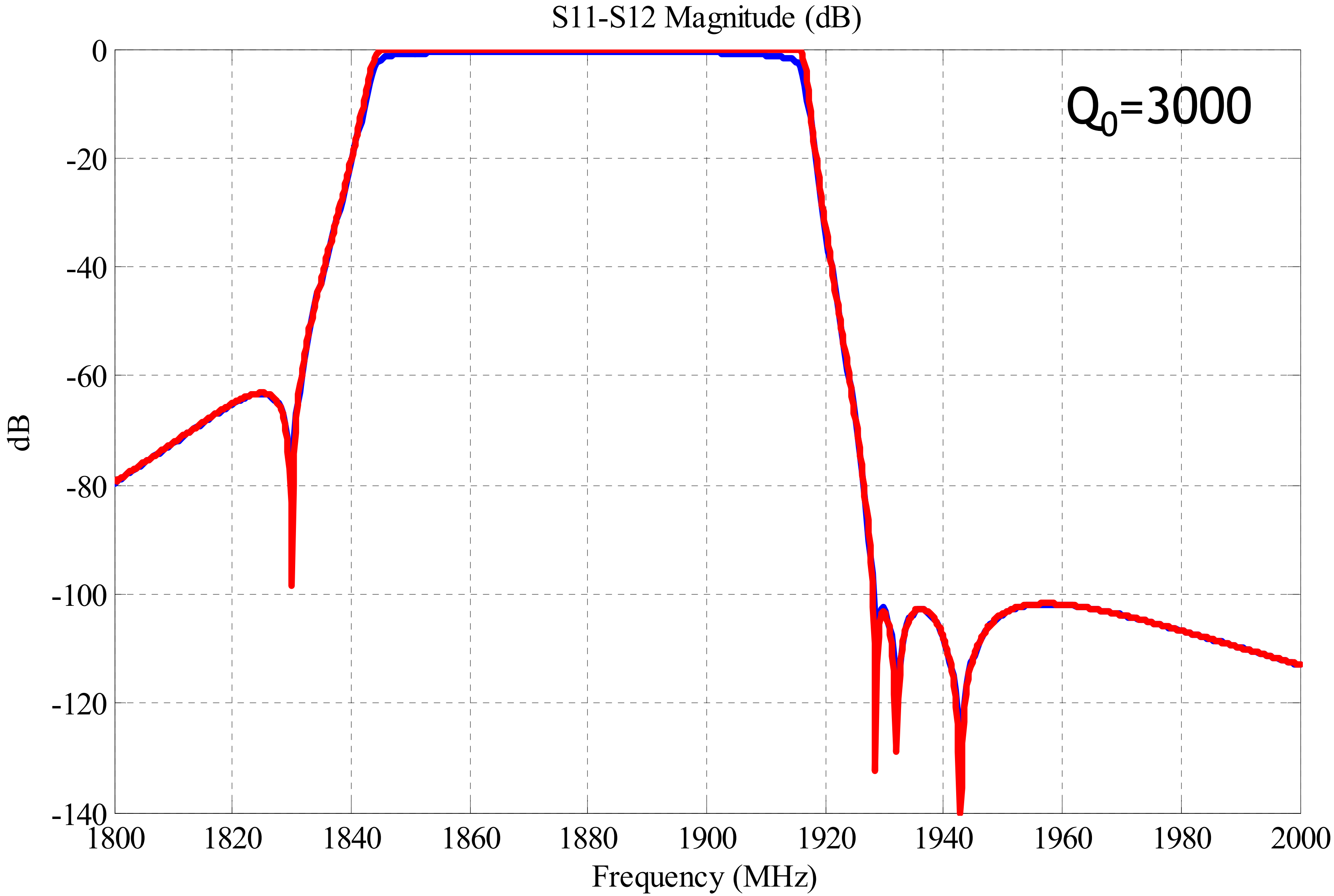
Ideally, a filter is a lossless network. In this case the attenuation in the passband is produced only by the not perfect match at input/output ports.

In the real world the filter components introduce **dissipation**, which becomes the main contribution to passband attenuation.

Passband Losses affect very few (in general) the port matching

In addition to the attenuation mask (which refer to overall attenuation) must be then specified also the matching requirements at the ports

Effect produced by losses on attenuation shape



Requirements on transmission phase

- To have a distortion-less two port, the phase of transmission parameter should be linear in the passband
- Often, the requirement is given on the *group delay*, which is defined as the derivative of transmission phase with respect the radian frequency. The group delay should be ideally constant in the passband
- Phase linearity is generally assessed a posteriori, after the attenuation requirements are satisfied.
- If the requirements on phase linearity is not verified, a *phase equalizer* may be required.
- In alternative, complex transmission zeros can be introduced in the response for phase equalization

Approaches to the design of microwave filters

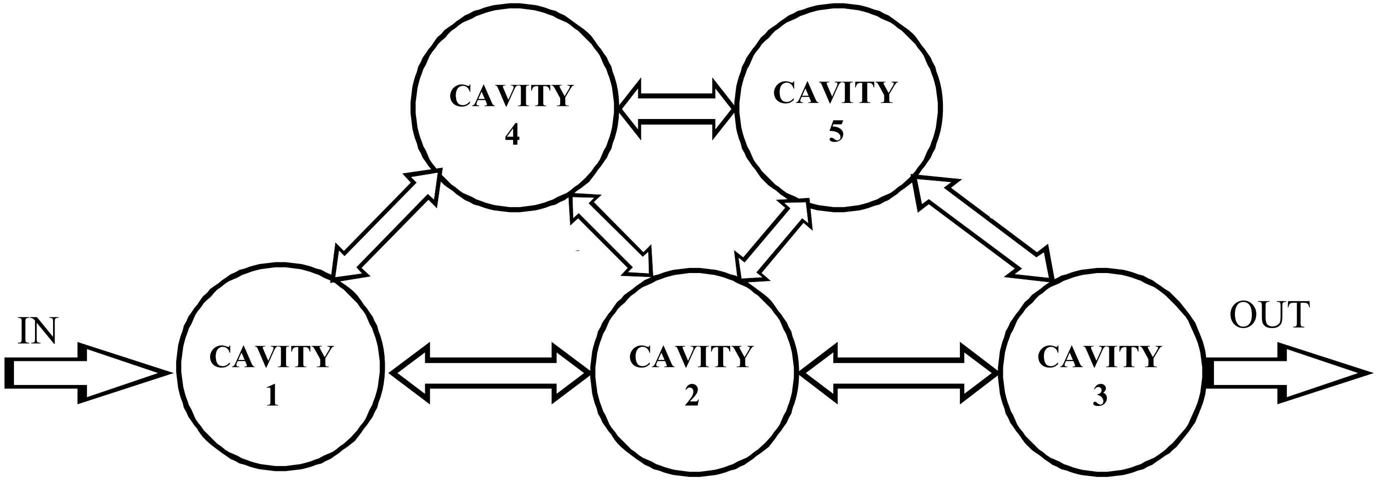
- Image parameters methods (old technique based on attenuation produced by basic blocks, which are suitably interconnected).
- Synthesis of networks composed by commensurate transmission line sections and stubs (suited for broadband filters; poses strong bounds on the configuration of the filter structure)
- **Equivalent Synthesis method,** based on the equivalence between the real (distributed components) structure and a lumped component filter network (suitable for narrow and moderate bandwidth passband filters. The most used technique today

Advantages of the Equivalent Synthesis Method

- Precise control of the curve representing the filter response
- The synthesis is developed in the lumped-element world, where well established, analytical and numerical procedures are available
- The results of the synthesis can be expressed by means of universal parameters which maintain the same meaning both for lumped-element circuit and distributed microwave networks
- A first-order dimensioning of the physical structure can be easily performed using these universal parameters

Recalls on Microwave Circuits for Filter Design

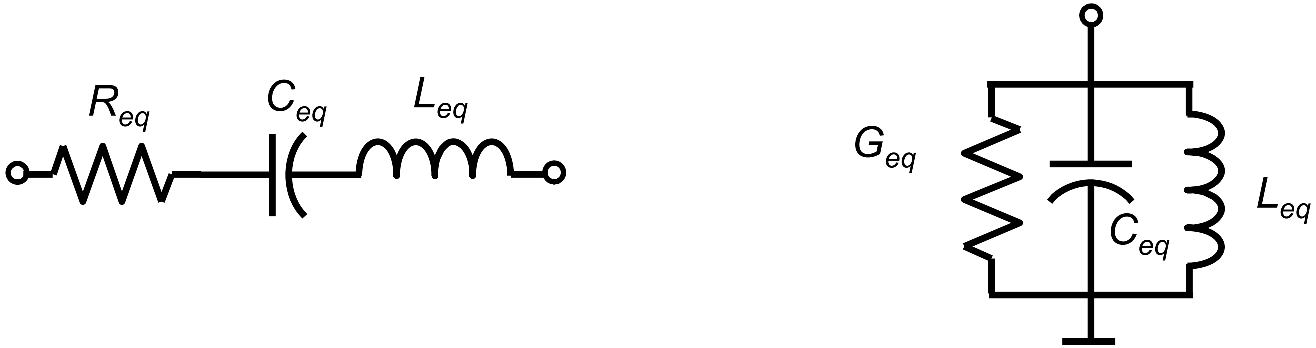
General physical structure of a microwave filter



Basic components:
- Cavities (resonating on a specific mode)
- Coupling structures (represented as ⟺)

Not always cavities and couplings can be identified separately

Basic equivalent circuit of the cavity

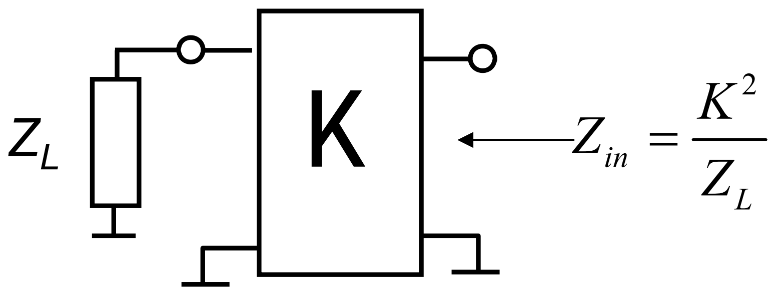


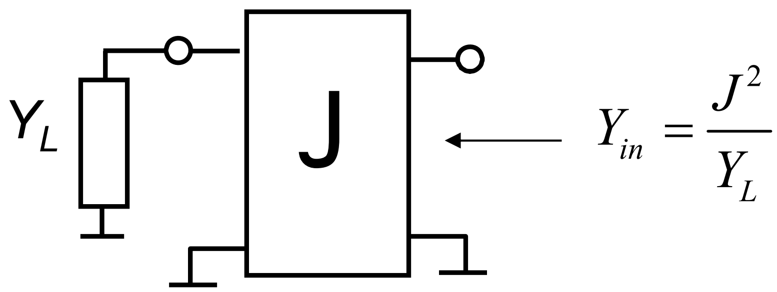
Series or parallel resonator

<u>Characteristic parameters</u>:
- Mode Resonant frequency (f_0)
- Equivalent slope parameter (L_{eq}, C_{eq})
- Loss parameter (R_{eq}, G_{eq})

Basic equivalent circuit of the coupling

Couplings are generally modeled with impedance or admittance inverters





The inverter has two basic functions:
- Operate as impedance transformer
- Change the nature of the load

The inverter is an ideal component that can only be approximated by real elements in a limited frequency band

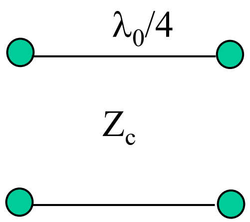
S parameters of the Impedance Inverter

$$S_{11} = S_{22} = \frac{\frac{K^2}{Z_0} - Z_0}{\frac{K^2}{Z_0} + Z_0} = \frac{K^2 - Z_0^2}{K^2 + Z_0^2} \qquad \text{Real (positive or negative)}$$

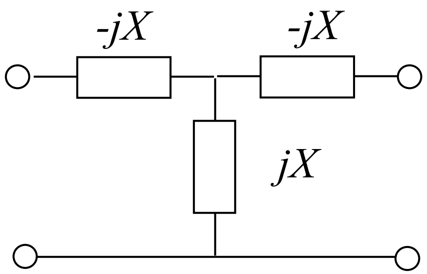
$$\phi_{11} - \phi_{12} = \pm\frac{\pi}{2} \quad \Rightarrow \quad \phi_{12} = \pm\frac{\pi}{2}$$

The impedance inverter is a symmetrical and reciprocal 2-port network. It behaves as an ideal $\pi/2$ phase shifter.
Imposing the lossless condition and the K value it is not sufficient to identify univocally the network (the ± sign of ϕ_{12} remains undetermined)

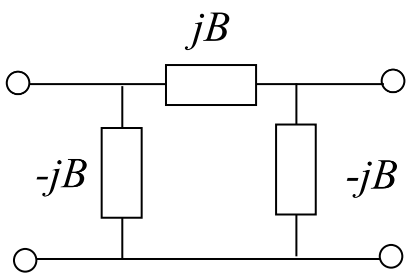
Equivalent circuit for the impedance inverter



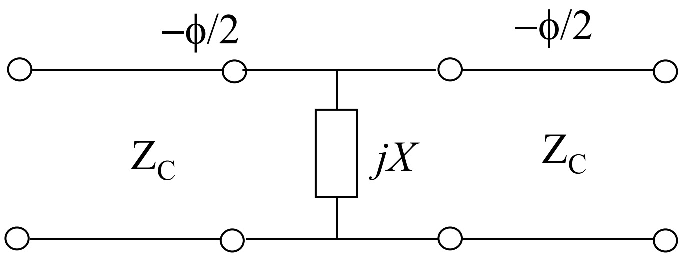
K=Zc



K=X



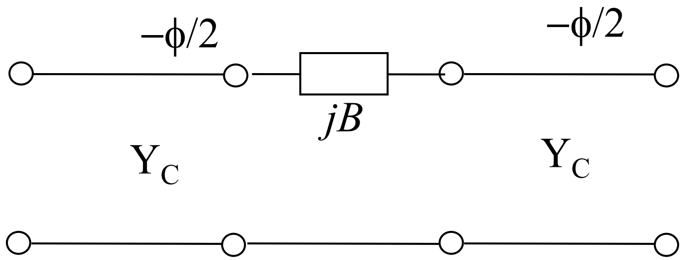
J=1/K=B



$$K = \left| Z_C \tan\left(\phi/2\right) \right|$$

$$\frac{\phi}{2} = \frac{1}{2}\tan^{-1}\left(\frac{2X}{Z_C}\right)$$

$$\frac{X}{Z_C} = \frac{K/Z_C}{1-\left(K/Z_C\right)^2}$$

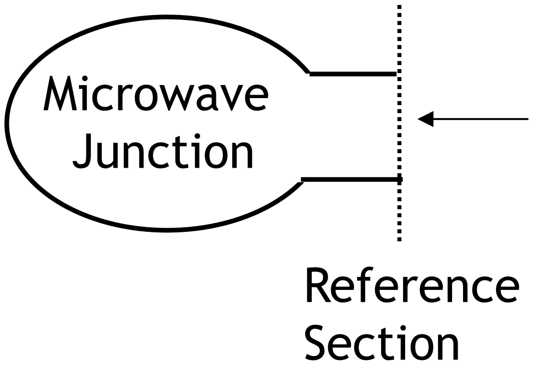


$$J = \left| Y_C \tan\left(\phi/2\right) \right|$$

$$\frac{\phi}{2} = \frac{1}{2}\tan^{-1}\left(\frac{2B}{Y_C}\right)$$

$$\frac{B}{Y_C} = \frac{J/Y_C}{1-\left(J/Y_C\right)^2}$$

Modeling of a microwave junction with a lumped equivalent circuit (1-port)



$Z_J = R_J(\omega) + jX_J(\omega) \approx Z_J = R_J(\omega_0) + jX_J(\omega)$

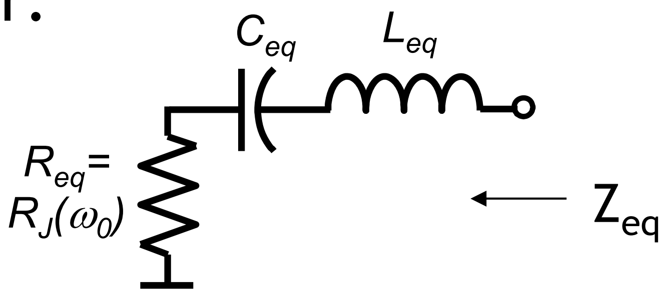
Goal: model of the junction in a defined frequency range B around f_0 ($B \ll f_0$)

The model is represented by an equivalent lumped-element impedance $Z_{eq} = R_{eq} + X_{eq}$ obtained by imposing:

- Constant real part: $R_{eq} = R_J(\omega_0)$
- Same value at ω_0 of the imaginary part: $X_{eq}(\omega_0) = X_J(\omega_0)$
- Same value of derivative at ω_0: $\partial X_{eq}(\omega_0)/\partial\omega = \partial X_J(\omega_0)/\partial\omega$

Equivalent impedance Z_{eq}

The most elementary network exhibiting Z_{eq} is a series resonator:



$$Z_{eq} = R_{eq} + j\left(\omega L_{eq} - \frac{1}{\omega C_{eq}}\right)$$

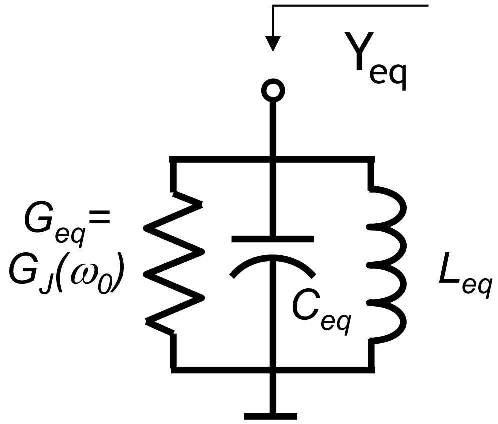
Imposing the previous conditions:

$$X_{eq}(\omega_0) = \omega_0 L_{eq} - \frac{1}{\omega_0 C_{eq}} = X_J(\omega_0)$$

$$\left.\frac{\partial X_{eq}(\omega)}{\partial \omega}\right|_{\omega=\omega_0} = L_{eq} + \frac{1}{\omega_0^2 C_{eq}} = \left.\frac{\partial X_J(\omega)}{\partial \omega}\right|_{\omega=\omega_0}$$

$$\boxed{L_{eq} = \frac{1}{2}\left[\left.\frac{\partial X_J(\omega)}{\partial \omega}\right|_{\omega=\omega_0} + \frac{X_J(\omega_0)}{\omega_0}\right], \quad \frac{1}{\omega_0^2 C_{eq}} = \frac{1}{2}\left[\left.\frac{\partial X_J(\omega)}{\partial \omega}\right|_{\omega=\omega_0} - \frac{X_J(\omega_0)}{\omega_0}\right]}$$

Equations for the Equivalent Admittance



$$Y_{eq} = G_{eq} + j\left(\omega C_{eq} - \frac{1}{\omega L_{eq}}\right)$$

Imposing the previous conditions:

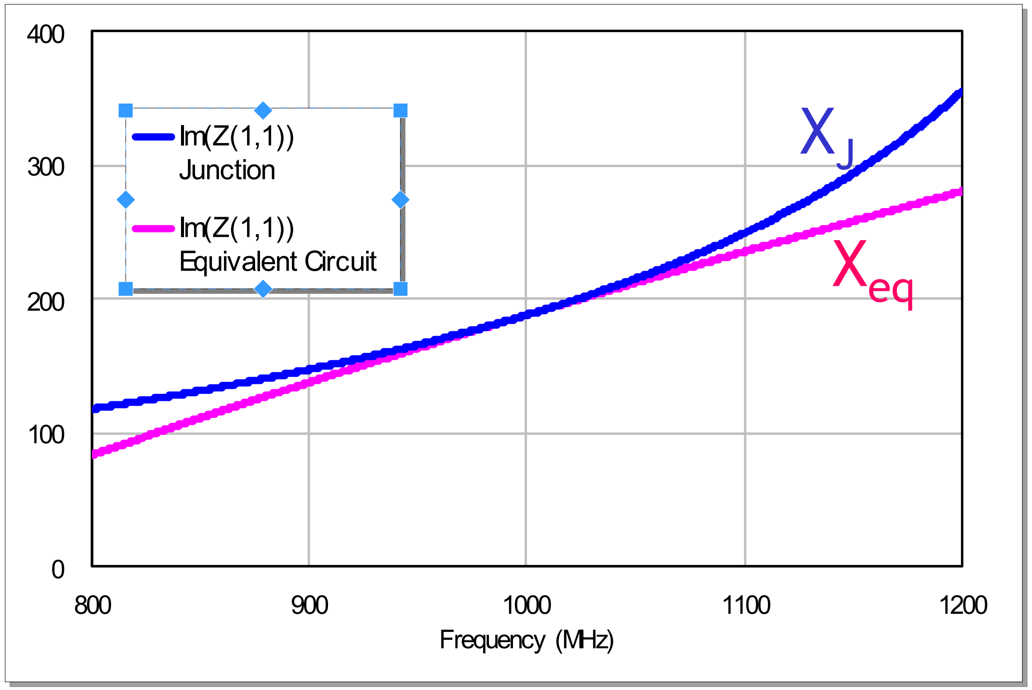
$$B_{eq}(\omega_0) = \omega_0 C_{eq} - \frac{1}{\omega_0 L_{eq}} = B_J(\omega_0)$$

$$\left.\frac{\partial B_{eq}(\omega)}{\partial \omega}\right|_{\omega=\omega_0} = C_{eq} + \frac{1}{\omega_0^2 L_{eq}} = \left.\frac{\partial B_J(\omega)}{\partial \omega}\right|_{\omega=\omega_0}$$

$$C_{eq} = \frac{1}{2}\left[\left.\frac{\partial B_J(\omega)}{\partial \omega}\right|_{\omega=\omega_0} + \frac{B_J(\omega_0)}{\omega_0}\right], \qquad \frac{1}{\omega_0^2 L_{eq}} = \frac{1}{2}\left[\left.\frac{\partial B_J(\omega)}{\partial \omega}\right|_{\omega=\omega_0} - \frac{B_J(\omega_0)}{\omega_0}\right]$$

Remark on the equivalent network

- The equivalence between the microwave junction and the lumped-element equivalent circuit is exact only at f_0
- The deviation between Z_J (Y_J) and Z_{eq} (Y_{eq}) becomes larger and larger with the increase of B



f_0=1000 MHz

$X(f_0)$=188.4 Ω

Special case: Resonant Junction (Cavity)

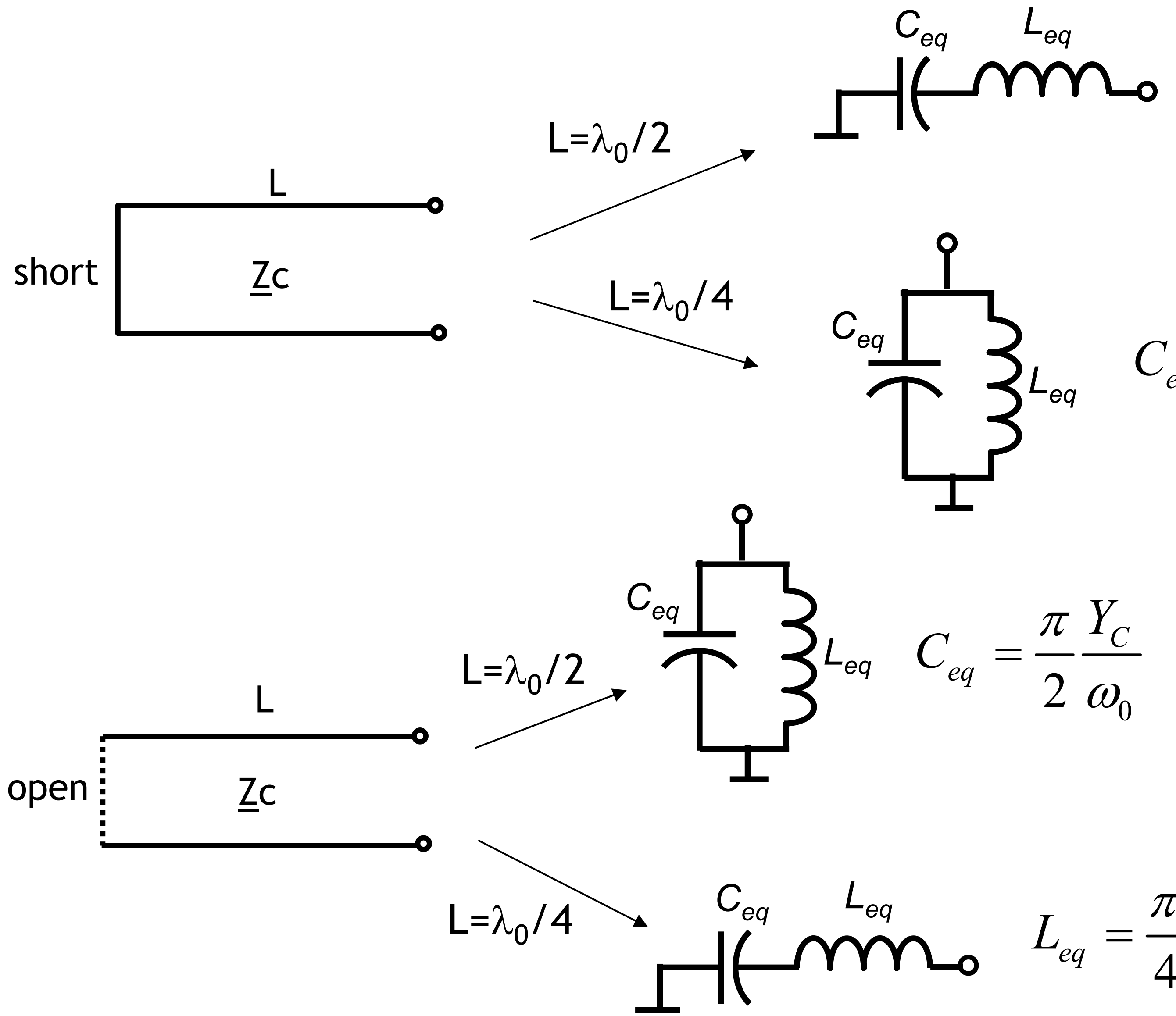
In case of resonant junctions ($X_J(\omega_0)=0$, $B_J(\omega_0)=0$), the previous equations become:

$$L_{eq} = \frac{1}{2}\frac{\partial X_J(\omega)}{\partial \omega}\bigg|_{\omega=\omega_0} \qquad C_{eq} = \frac{1}{\omega_0^2 L_{eq}}$$

$$C_{eq} = \frac{1}{2}\frac{\partial B_J(\omega)}{\partial \omega}\bigg|_{\omega=\omega_0} \qquad L_{eq} = \frac{1}{\omega_0^2 C_{eq}}$$

These equations define an equivalent circuit for a large class of **cavity resonators.**
The type of resonator depends on the **resonant mode** and on the **reference section**

Example: TEM cavity realized with short-circuited/ open-circuited transmission line



Short, $L=\lambda_0/2$:

$$L_{eq} = \frac{\pi}{2}\frac{Z_C}{\omega_0}$$

Short, $L=\lambda_0/4$:

$$C_{eq} = \frac{\pi}{4}\frac{Y_C}{\omega_0}$$

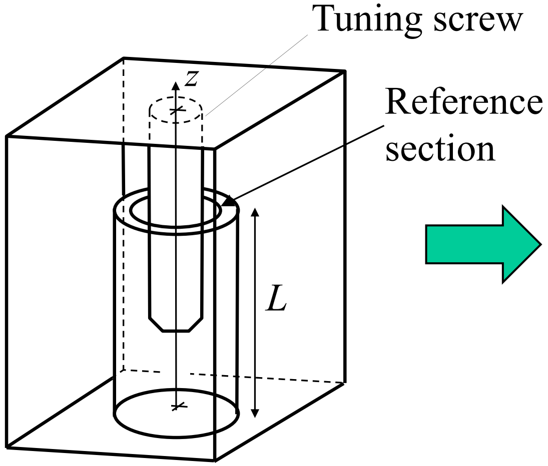
Open, $L=\lambda_0/2$:

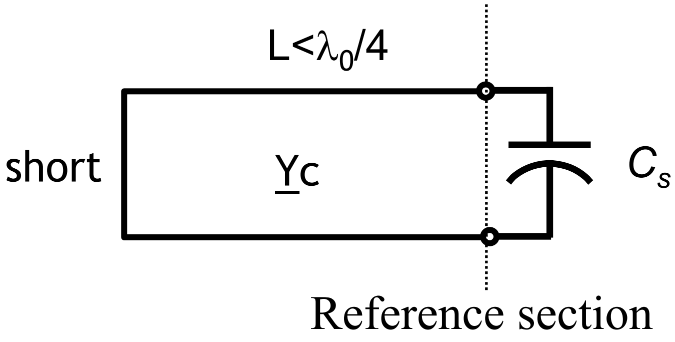
$$C_{eq} = \frac{\pi}{2}\frac{Y_C}{\omega_0}$$

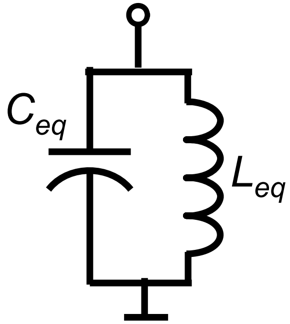
Open, $L=\lambda_0/4$:

$$L_{eq} = \frac{\pi}{4}\frac{Z_C}{\omega_0}$$

Capacity-loaded coaxial resonator







$$B_{tot} = \omega C_s - Y_C \cot(\beta L)$$

Resonance condition:

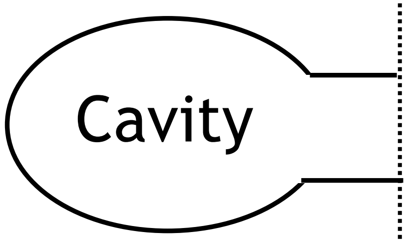
$$B_{tot}(\omega_0) = 0 \quad \Rightarrow \quad \omega_0 C_s = Y_C \cot(\beta_0 L)$$

Equivalent capacitance:

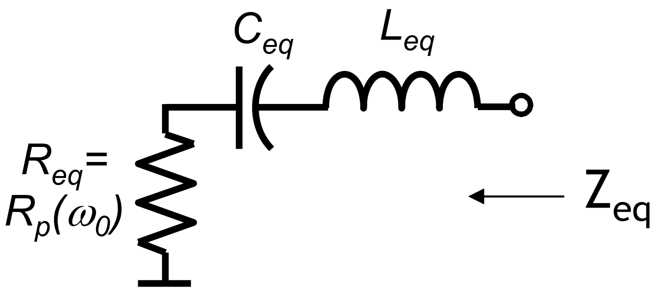
$$\omega_0 C_{eq} = \frac{1}{2} \frac{\partial B_{tot}(\omega)}{\partial \omega}\bigg|_{\omega=\omega_0} = \frac{1}{2} Y_C \left[\cot(\beta_0 L) + \frac{\beta_0 L}{\sin^2(\beta_0 L)} \right] =$$

$$= \frac{1}{2} Y_C \cot(\beta_0 L) \left[1 + \frac{2\beta_0 L}{\sin(2\beta_0 L)} \right] \quad \overset{L=\lambda_0/8}{\Rightarrow} \quad \omega_0 C_{eq} = \frac{Y_C}{2}\left[1 + \frac{\pi}{2}\right]$$

Losses in the cavity: the unloaded Q



$$Q_0 = \omega_0 \frac{\text{Energy stored in the junction}}{\text{Power dissipated in the junction}}$$

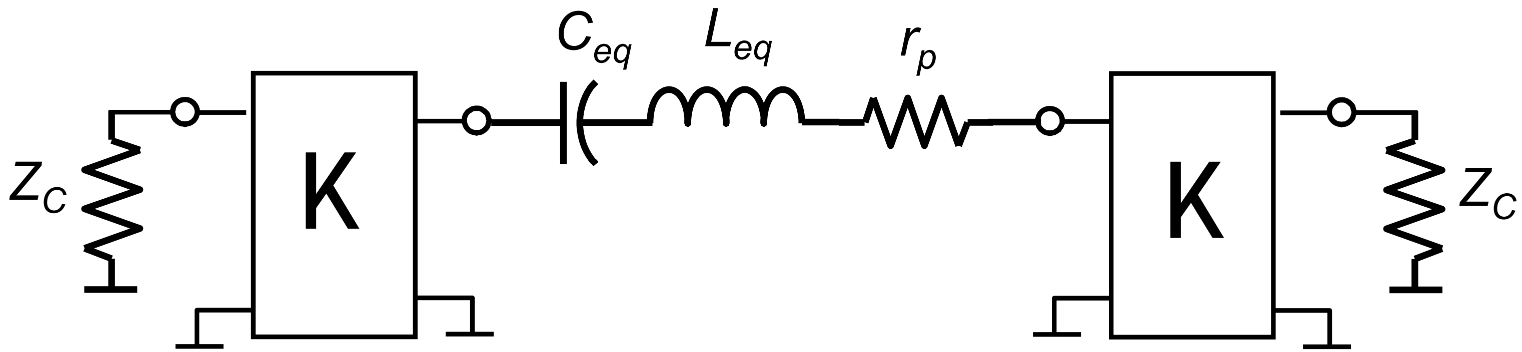


$$Q_0 = \frac{\omega_0 L_{eq}}{R_p} \qquad R_p = R_J + R_\varepsilon$$

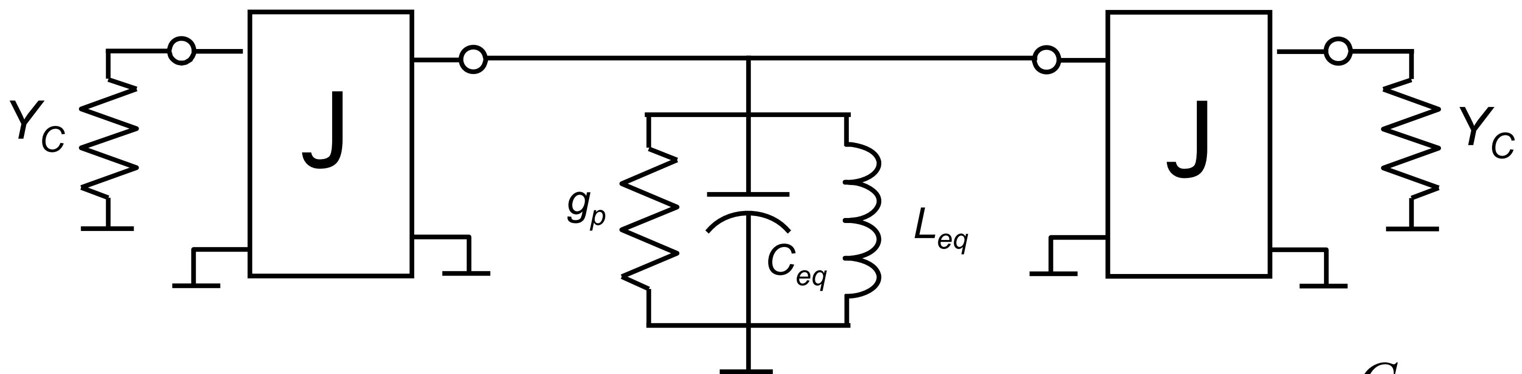
R_J is the sum of two terms: Rp, due to thefinite conductivity of metallic walls and Re to to medium dissipation (dielectric losses). Is has then:

$$\frac{1}{Q_0} = \frac{1}{Q_{0J}} + \frac{1}{Q_{0\varepsilon}}$$

Modeling of a cavity coupled to loads



$$Q_L = \frac{\omega_0 L_{eq}}{2K^2/Z_C} = \frac{f_0}{B_{3dB}} \qquad \left(\text{for } r_p << K^2/Z_C\right) \qquad Q_0 = \frac{\omega_0 L_{eq}}{r_p}$$



$$Q_L = \frac{\omega_0 C_{eq}}{J^2/Y_C} = \frac{f_0}{B_{3dB}} \qquad \left(\text{for } g_p << J^2/Y_C\right) \qquad Q_0 = \frac{\omega_0 C_{eq}}{g_p}$$

Parameters of loaded cavities

The loaded Q (QL): determines

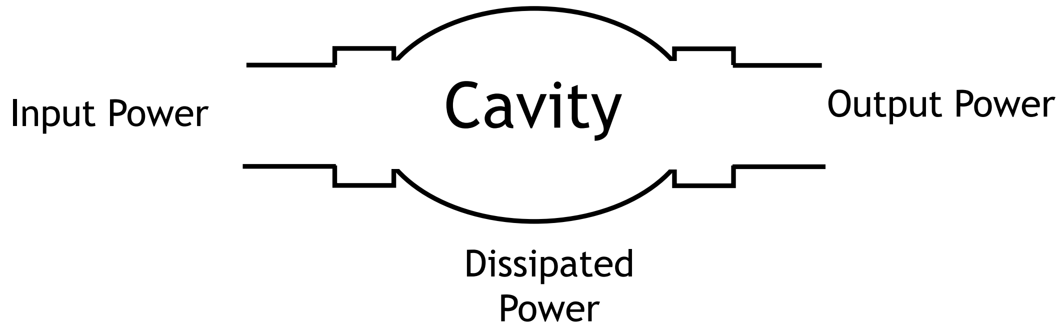
- The 3 dB bandwidth ($B_{3dB}=f_0/Q_L$). For a given cavity (L_{eq} or C_{eq}), a value for K (J) can be evaluated for obtaining the desired B_{3dB}:

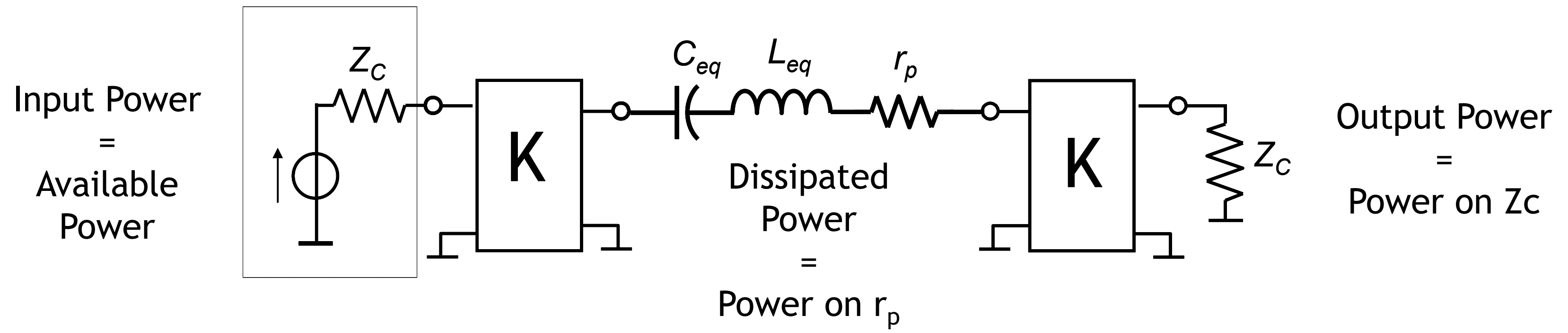
$$K = \sqrt{\frac{Z_C \cdot \omega_0 L_{eq}}{2Q_L}}, \qquad J = \sqrt{\frac{Y_C \cdot \omega_0 C_{eq}}{2Q_L}}$$

- The matching band at the input (output) port. For small losses, it can be shown that the bandwidth B_Γ for a given value Γ of the input reflection coefficient is related to B_{3dB} as follows:

$$B_\Gamma \simeq |\Gamma| \cdot B_{3dB}$$

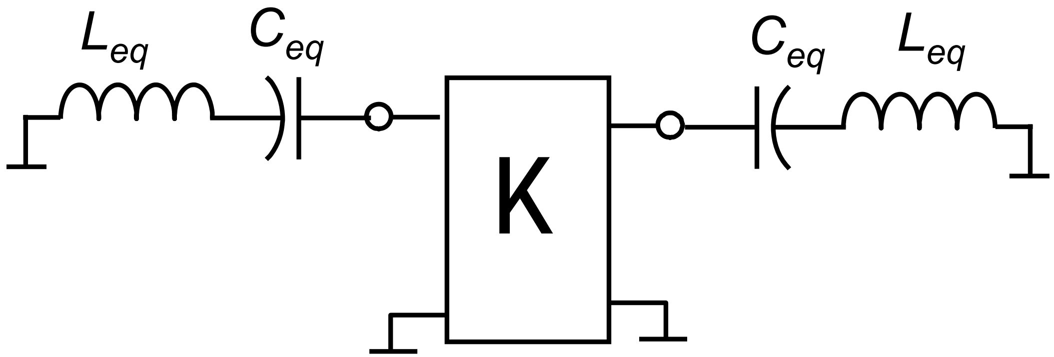
Power transmitted to load (transducer attenuation)





$$A_{dB} = 20\log\left(1 + \frac{Q_L}{Q_0}\right)$$

Coupling of two cavities: equivalent circuit



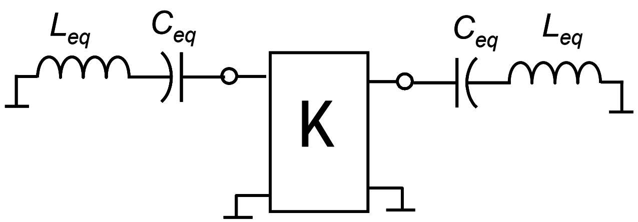
Characteristic parameter: the **coupling coefficient**

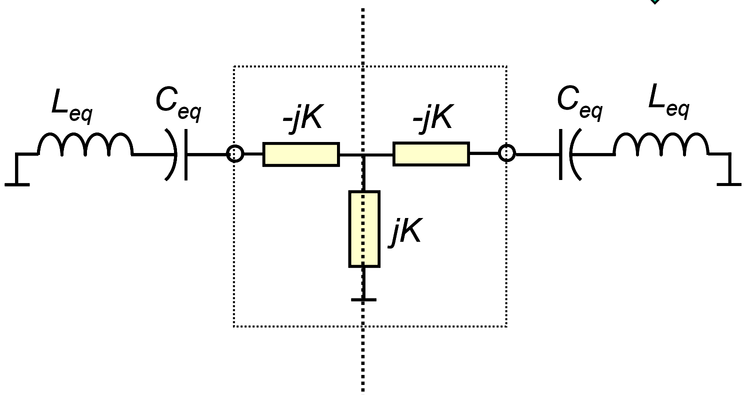
$$k = \frac{K}{\omega_0 L_{eq}} \qquad \left(= \frac{J}{\omega_0 C_{eq}} \text{ for shunt resonators} \right)$$

k determines the coupling bandwidth of the two resonators (the larger is *k* the larger is the bandwidth)

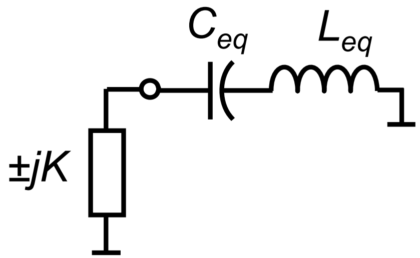
Evaluation of *k* from resonances

- Even and odd resonances of two coupled resonators:





Short/Open



Short→-jK
Open→+jK

$$f_o = f_0 \left[\sqrt{\left[\left(\frac{K}{2\omega_0 L_{eq}} \right)^2 + 1 \right]} - \frac{K}{2\omega_0 L_{eq}} \right]$$

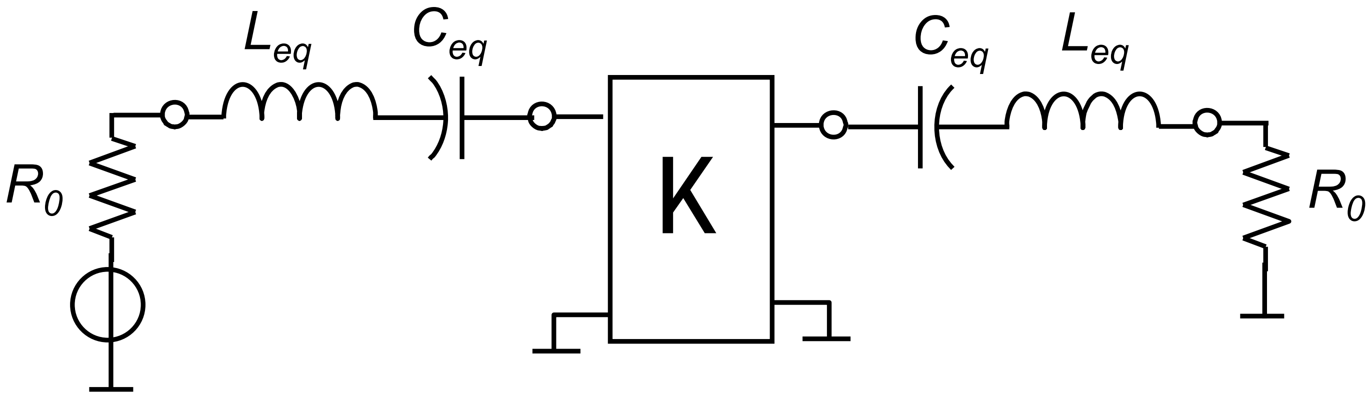
Odd Resonance (-jK)

$$f_e = f_0 \left[\sqrt{\left[\left(\frac{K}{2\omega_0 L_{eq}} \right)^2 + 1 \right]} + \frac{K}{2\omega_0 L_{eq}} \right]$$

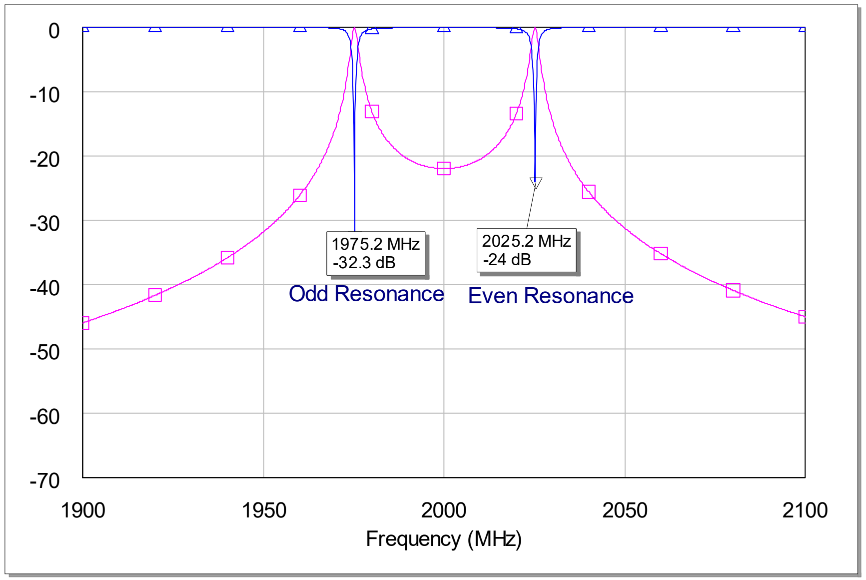
Even Resonance (+jK)

$$f_0 = \sqrt{f_e \cdot f_o}, \qquad k = \frac{f_e - f_o}{f_0}$$

Practical evaluation of *k*



$\omega_0 L_{eq} >> R_0$ ➡ Input/output coupling ≈0

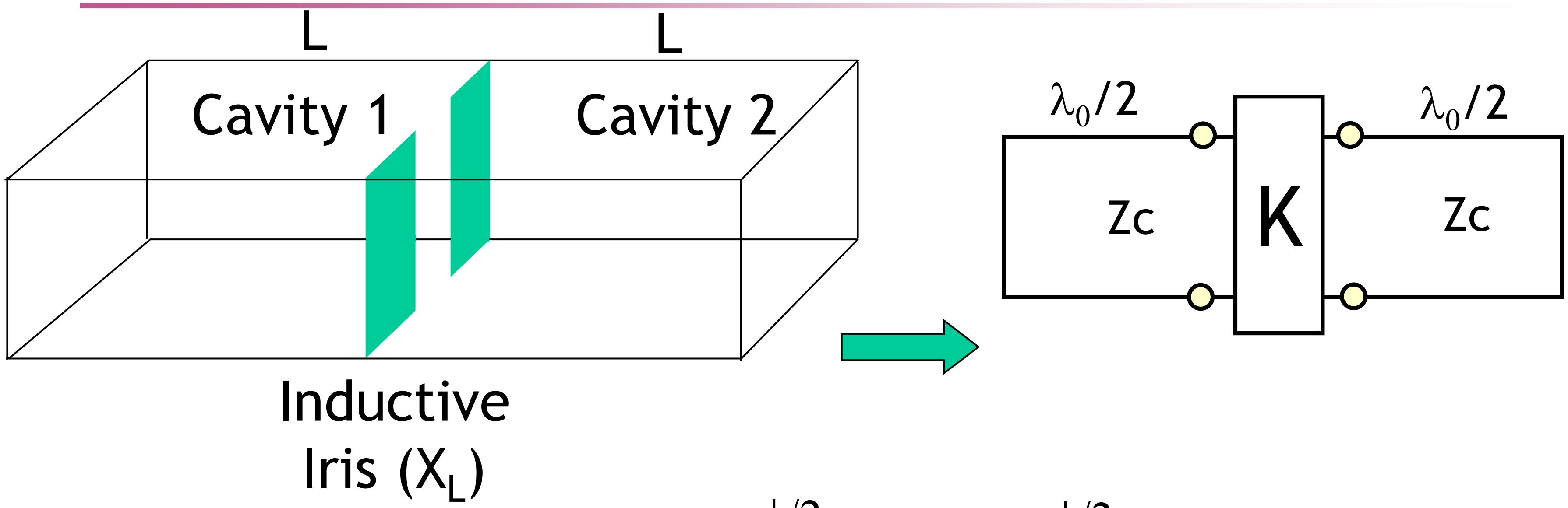


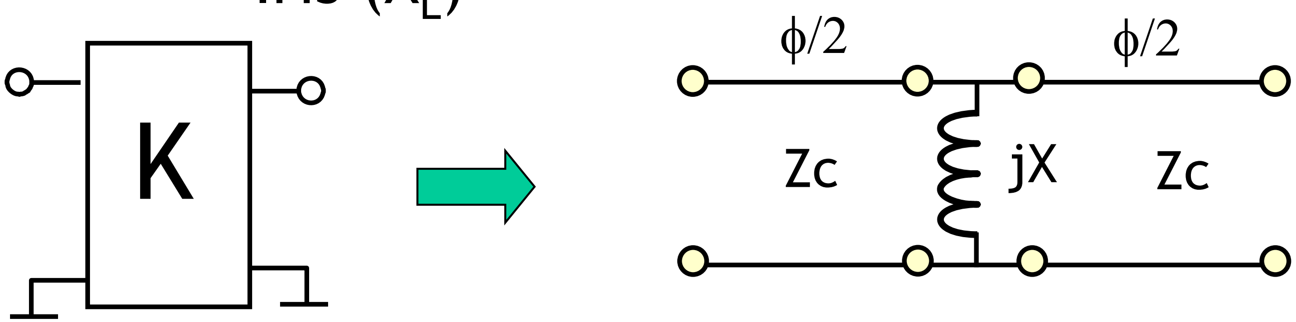
X_{eq}=10, R0=0.01
f_e=2025.2 MHz
f_o=1975.2

$$f_0 = \sqrt{f_e \cdot f_o} = 2000MHz$$

$$k = \frac{f_e - f_o}{f_0} = 0.025$$

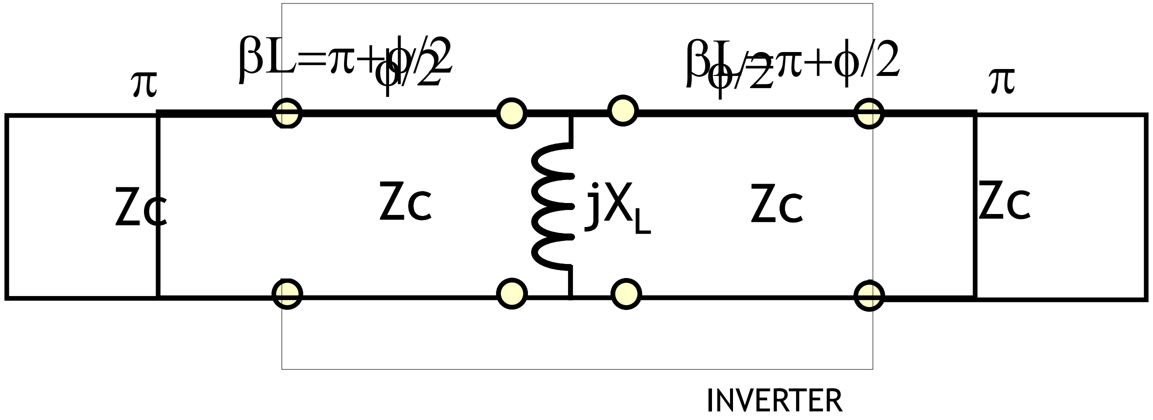
Coupled cavities: example 1





$$X = \frac{K}{1-\left(\frac{K}{Z_C}\right)^2}, \quad \phi = -\tan^{-1}\left(2X\right)$$

example 1 (cont.)

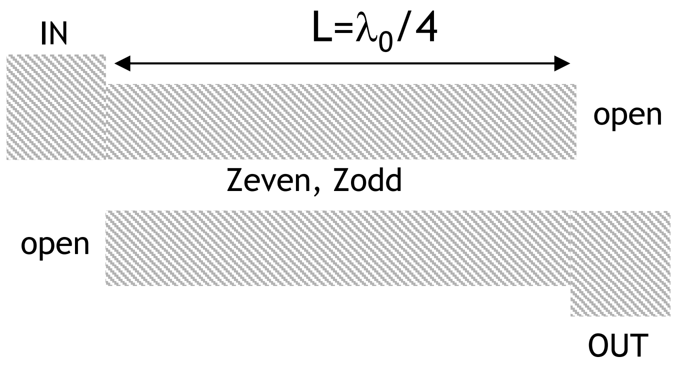


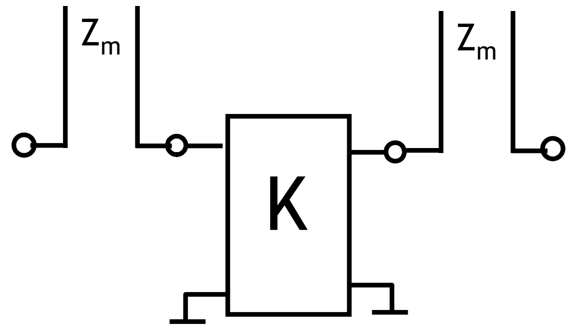
$$\beta L = \pi + \frac{\phi}{2} = \pi - \tan^{-1}(2X_L) \qquad X_L = \frac{K}{1-\left(\frac{K}{Z_C}\right)^2} \approx K \quad (\text{for } K << Z_C)$$

Equiv. reactance: $\omega_0 L_{eq} = \frac{\pi}{2} Z_C \left(\frac{\lambda_g}{\lambda_0}\right)^2$

Coupling coefficient: $k = \frac{K}{\omega_0 L_{eq}} = \frac{2}{\pi} \frac{X_L}{Z_C} \left(\frac{\lambda_g}{\lambda_0}\right)^2$

Resonators with coupled lines: Interdigital





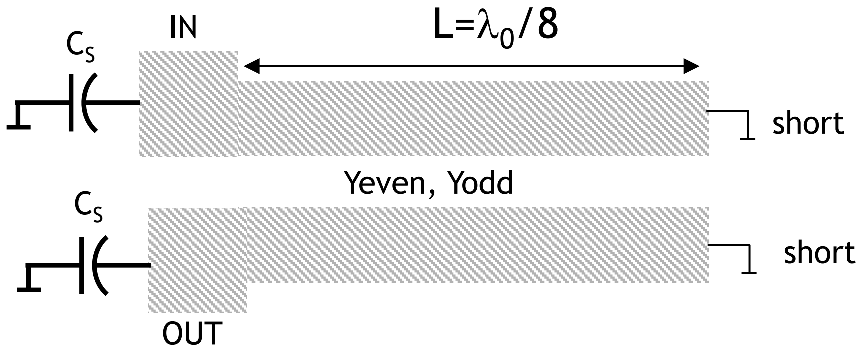
$$Z_m = \frac{1}{2}\left(Z_{even} + Z_{odd}\right)$$

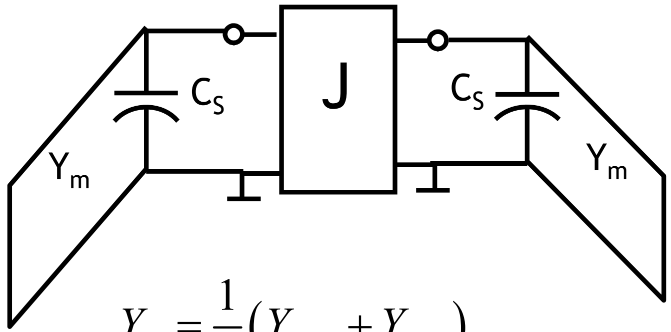
$$K = \frac{1}{2}\left(Z_{even} - Z_{odd}\right)\frac{1}{\sin\left(\beta L\right)}$$

$$\omega_0 L_{eq} = \frac{\pi}{4} Z_m$$

$$k = \frac{K}{\omega_0 L_{eq}} = \frac{4}{\pi}\left(\frac{Z_{even} - Z_{odd}}{Z_{even} + Z_{odd}}\right) = \frac{4}{\pi} m$$

Resonators with coupled lines: Comb





$$Y_m = \frac{1}{2}\left(Y_{even} + Y_{odd}\right)$$

$$J = \frac{1}{2}\left(Y_{even} - Y_{odd}\right)\frac{1}{\tan\left(\beta L\right)}$$

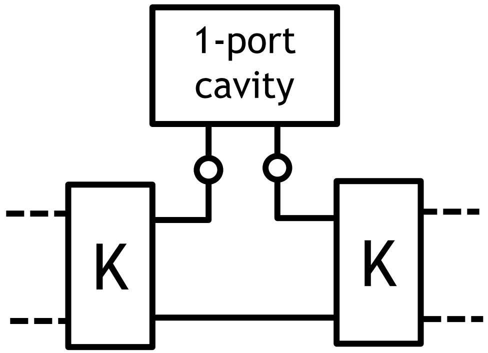
$$\omega_0 C_{eq} = \frac{Y_m}{2}\left[1 + \frac{\pi}{2}\right],$$

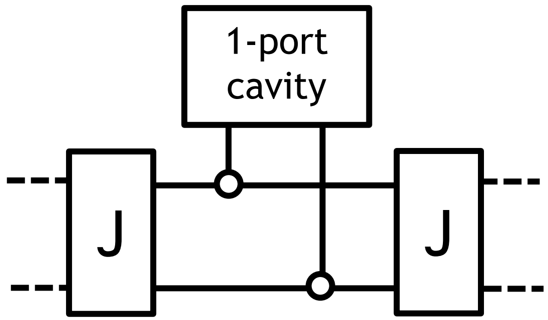
$$k = \frac{J}{\omega_0 C_{eq}} = \frac{2}{\pi + 2}\left(\frac{Y_{odd} - Y_{even}}{Y_{even} + Y_{odd}}\right) = \frac{2m}{\pi + 2}$$

Note: $$\frac{Y_{odd} - Y_{even}}{Y_{even} + Y_{odd}} = \frac{Z_{even} - Z_{odd}}{Z_{even} + Z_{odd}} = m$$

1-port and 2-port Cavities

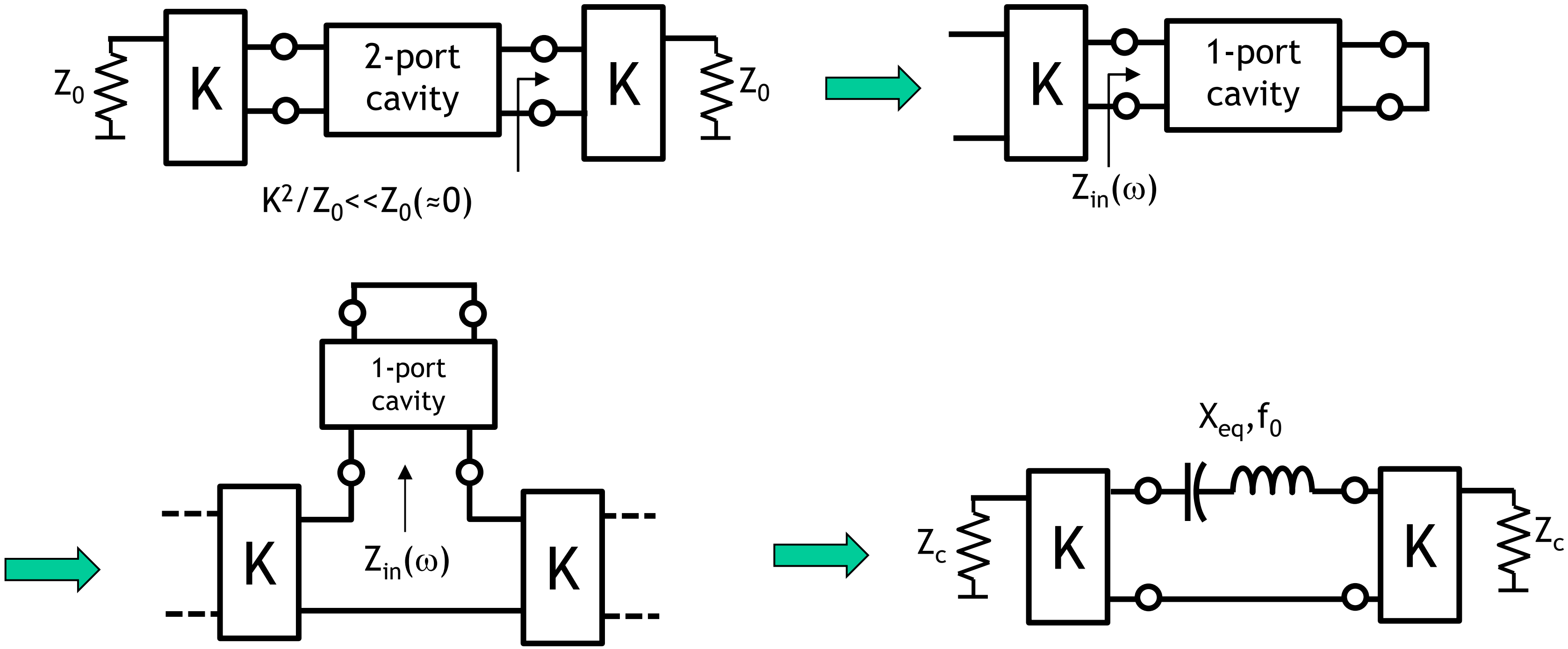
Previous results refer to 1-port cavities that can be represented with a series or shunt resonator:





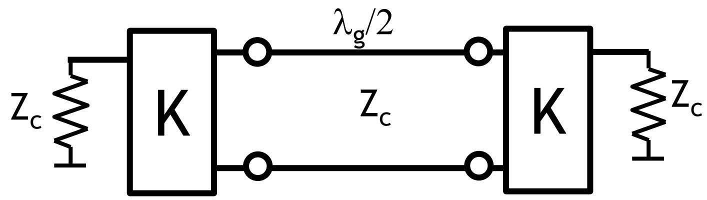
In case of 2-port cavities the equivalence should be imposed on all the elements of S or Z or Y matrix.
It is however possible an approximation allowing to still refer to 1-port resonator

1-port equivalent circuit of 2-port cavities

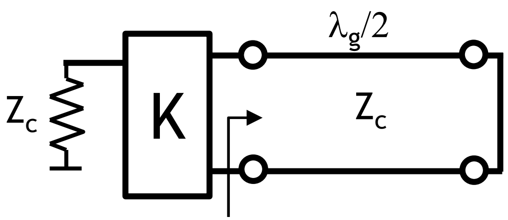


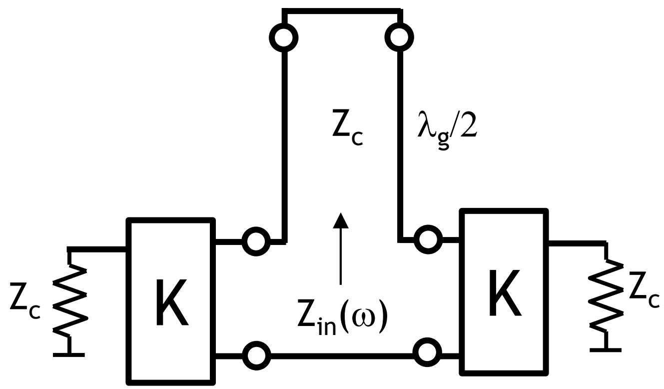
In general, this is acceptable for narrowband filters ($B/f_0 << 1$), but represents a further approximation that is added to the approximation of the cavity with a lumped resonator

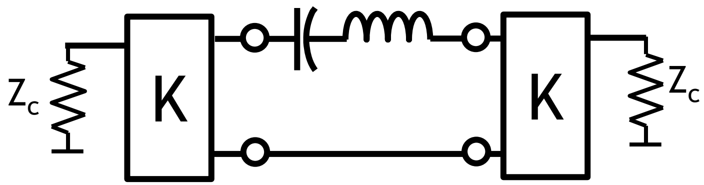
Example: Transmission line cavity



Assuming $K^2/Z_c \ll 1$:







$$X_{eq} = \omega_0 L_{eq} = \frac{\pi}{2} Z_C$$