

INF 5490

L4: Utfordringer ved RF kretsdesign

Kjøreplan INF5490

- L1: Introduksjon. MEMS i RF
- L2: Fremstilling og virkemåte
- L3: Modellering, design og analyse
- Dagens forelesning:
 - Noen typiske trekk og utfordringer ved RF kretsdesign**

Oversikt over forelesningen

- RF kretsdesign : -> "Multidisiplin"
- Elektromagnetiske bølger
- "Skin depth"
- Passive komponenter ved høye frekvenser
- Transmisjonslinje-teori
- To-port nettverk
- S-parametre
- Filtre
- Q-faktor

RF- og mikrobølge-design er ”multidisiplin”

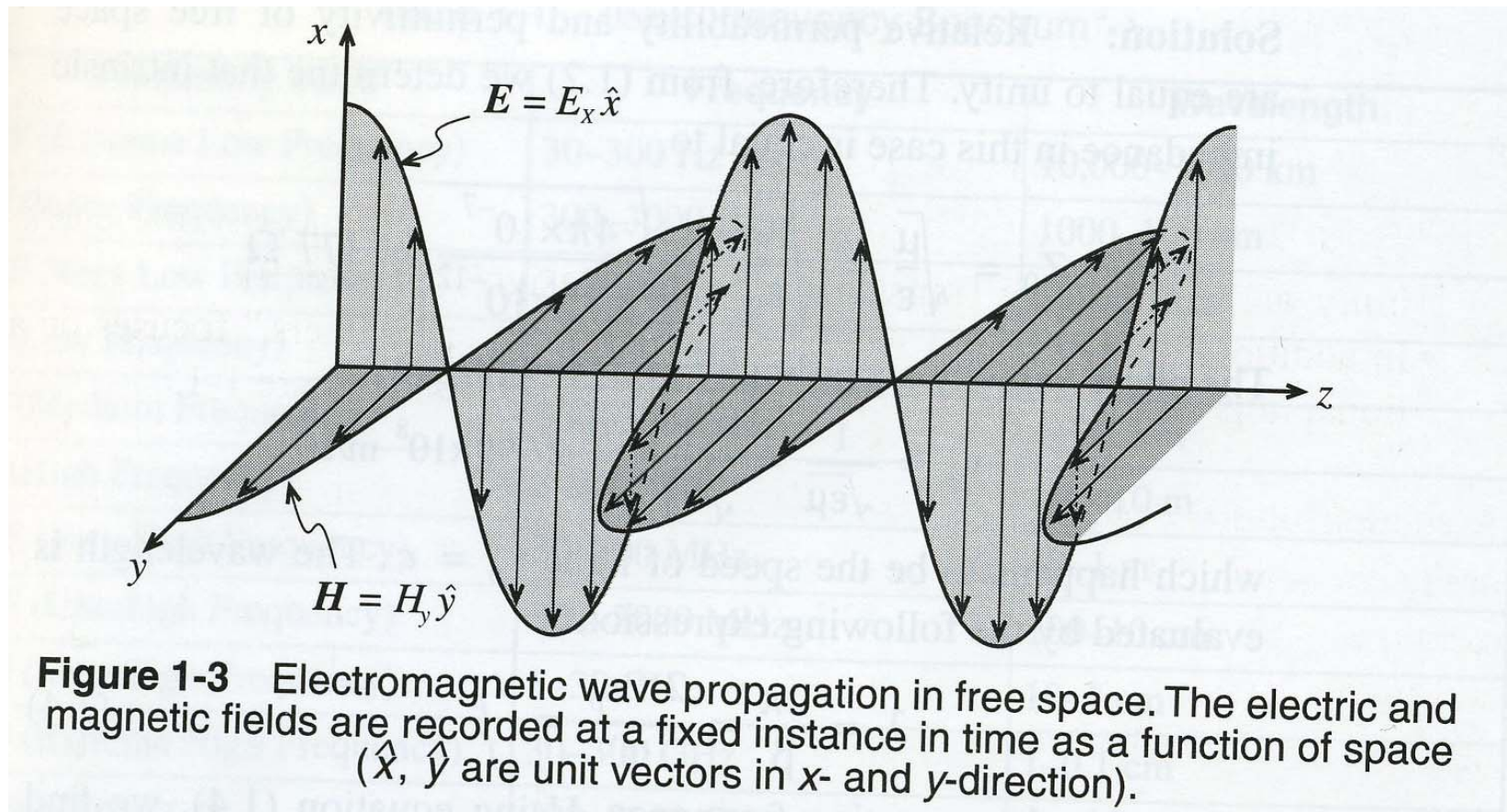
- Teoretisk fundament
 - Elektromagnetisme
 - Signalbehandling
- Teknologiske, praktiske aspekter
 - Krets-teori
 - Kirchhoffs lover for strøm og spenning
- Noe av stoffet i dag overlapper INF5480
 - RF-kretser, teori og design (Tor Fjeldly)
 - Her: → **Sentrale punkter på en forelesning!**

RF kretsdesign

- Hvordan oppfører kretser seg ved høye frekvenser?
- Hvorfor endrer funksjonaliteten til komponentene seg?
- Ved hvilke frekvenser blir vanlig kretsanalyse ugyldig?
- Hva slags ny krets-teori trengs?
- Hvordan kan denne teorien brukes i praksis?
 - → *Figurer og ligninger fra R. Ludwig et al: "RF Circuit Design"*

Elektromagnetiske bølger

- Elektrisk og magnetisk felt



Sentrale bølgeparametre:

Elektrisk felt

$$E_x = E_{0x} \cos(\omega t - \beta z)$$

Magnetfelt

$$H_y = H_{0y} \cos(\omega t - \beta z)$$

Angular frekvens: ω

Propageringskonstant: β

Bølgen gjentar seg når: $\beta \cdot z = 2\pi$

Bølgelengde: $z = \lambda = \frac{2\pi}{\beta}$

Bølgen forplanter seg en avstand λ på en tid $T =$ perioden

Forplantningshastigheten:
(i vakuum: c)

$$v_p \cdot T = \lambda$$

$$v_p = \lambda \cdot \frac{1}{T} = \lambda \cdot f = \frac{2\pi}{\beta} \cdot \frac{\omega}{2\pi} = \frac{\omega}{\beta}$$

Sentrale bølgeparametre, forts.

Ved et gitt sted, dvs. $z = \text{konstant}$, gjentar bølgen seg etter perioden T :

$$\omega T = 2\pi \quad \text{og} \quad \omega = 2\pi / T = 2\pi f$$

der $f = \text{frekvensen}$

Frekvens og bølgelengde

- I vakuum: $\lambda * f = c$
 - Økende frekvens \rightarrow minkende bølgelengde
- Ved høye frekvenser (RF) er bølgelengden sammenlignbar med kretsdimensjonene
 - \rightarrow

Table 1-1 IEEE Frequency Spectrum

Frequency Band	Frequency	Wavelength
ELF (Extreme Low Frequency)	30–300 Hz	10,000–1000 km
VF (Voice Frequency)	300–3000 Hz	1000–100 km
VLF (Very Low Frequency)	3–30 kHz	100–10 km
LF (Low Frequency)	30–300 kHz	10–1 km
MF (Medium Frequency)	300–3000 kHz	1–0.1 km
HF (High Frequency)	3–30 MHz	100–10 m
VHF (Very High Frequency)	30–300 MHz	10–1 m
UHF (Ultrahigh Frequency)	300–3000 MHz	100–10 cm
SHF (Superhigh Frequency)	3–30 GHz	10–1 cm
EHF (Extreme High Frequency)	30–300 GHz	1–0.1 cm
Decimillimeter	300–3000 GHz	1–0.1 mm
P Band	0.23–1 GHz	130–30 cm
L Band	1–2 GHz	30–15 cm
S Band	2–4 GHz	15–7.5 cm
C Band	4–8 GHz	7.5–3.75 cm
X Band	8–12.5 GHz	3.75–2.4 cm
Ku Band	12.5–18 GHz	2.4–1.67 cm
K Band	18–26.5 GHz	1.67–1.13 cm
Ka Band	26.5–40 GHz	1.13–0.75 cm
Millimeter wave	40–300 GHz	7.5–1 mm
Submillimeter wave	300–3000 GHz	1–0.1 mm

To sentrale lover

- **Faradays lov**
 - Varierende **magnetfelt** induserer **strøm**
- **Amperes lov**
 - **Strøm** som flyter setter opp **magnetfelt**

Faradays lov

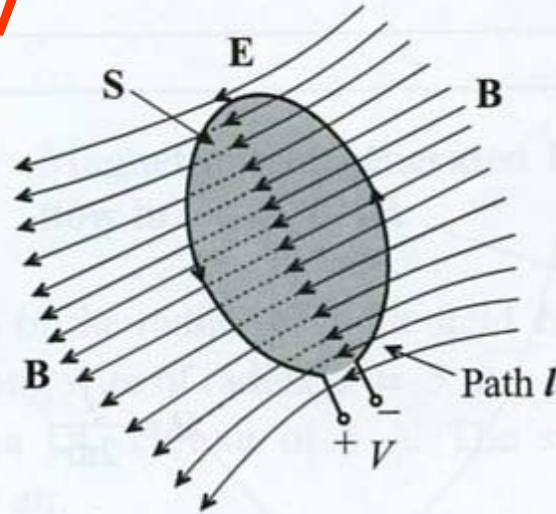


Figure 2-15 The time rate of change of the magnetic flux density induces a voltage.

$$\oint \vec{E} \cdot d\vec{l} = -\frac{d}{dt} \iint \vec{B} \cdot d\vec{S}$$

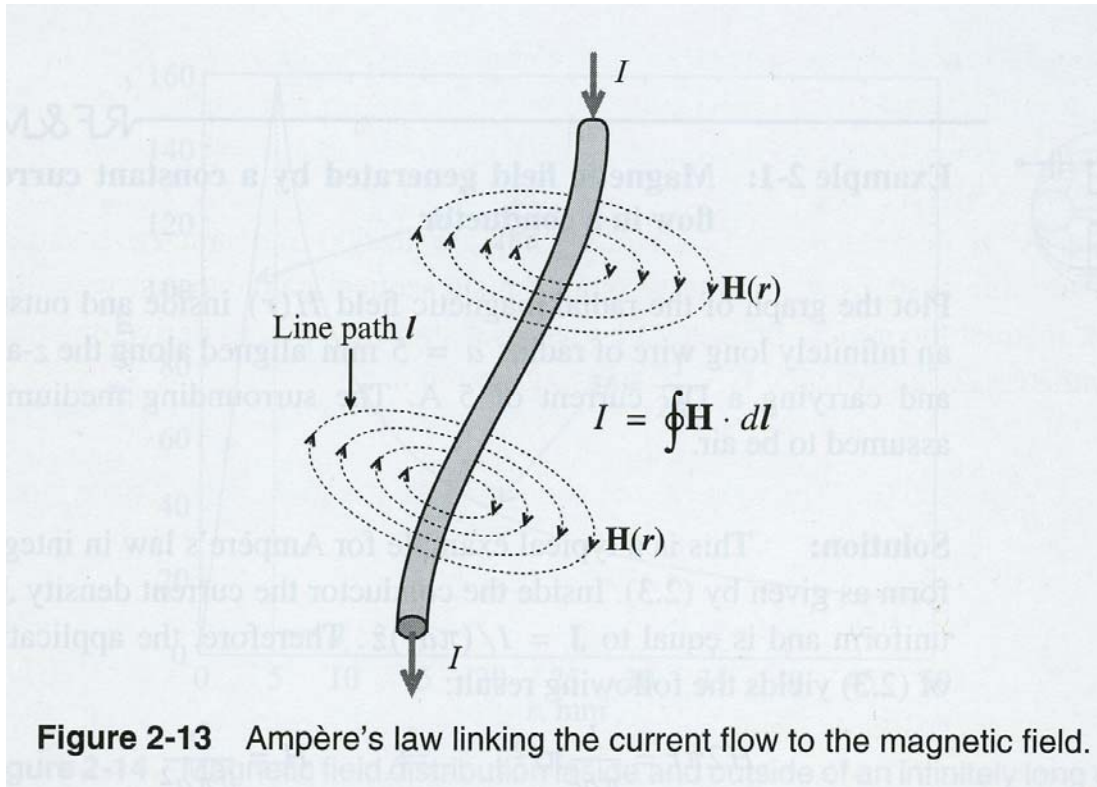
\vec{B} = magnetisk fluks – tetthet

$$\vec{B} = \mu \cdot \vec{H}$$

μ = permeabilitet = $\mu_0 \cdot \mu_r$

\vec{H} = magnetfelt

Amperes lov



$$I = \oint \bar{H} \cdot d\bar{l} = \iint \bar{J} \cdot d\bar{S}$$

”Skin depth”

- Signaloverføring ved økende frekvens
 - **DC** signal:
 - Hele tverrsnittet leder strøm
 - **AC** signal:
 - Varierende strøm induserer et vekslende magnetfelt ([Amperes lov](#))
 - Magnetfeltet er sterkest når radius er liten
 - Størst tidsvariasjon av magnetfeltet når en nærmer seg sentrum
 - Varierende magnetfelt induserer et elektrisk felt ([Faradays lov](#))
 - Det induserte elektriske feltet (som motvirker det opprinnelige) øker i styrke mot midten av ledere

Skin depth, forts.

- Motstanden R øker mot sentrum av lederen
 - Strømmen flyter i **ytterkantene** ved økende frekvens
 - Formel: "skin-depth" →
 - Betegner reduksjon i strømtettheten til 1/e
- Hva betyr denne effekten i praksis? →

$$\delta = (\pi f \mu \sigma_{\text{cond}})^{-1/2}$$

"Skin-depth"

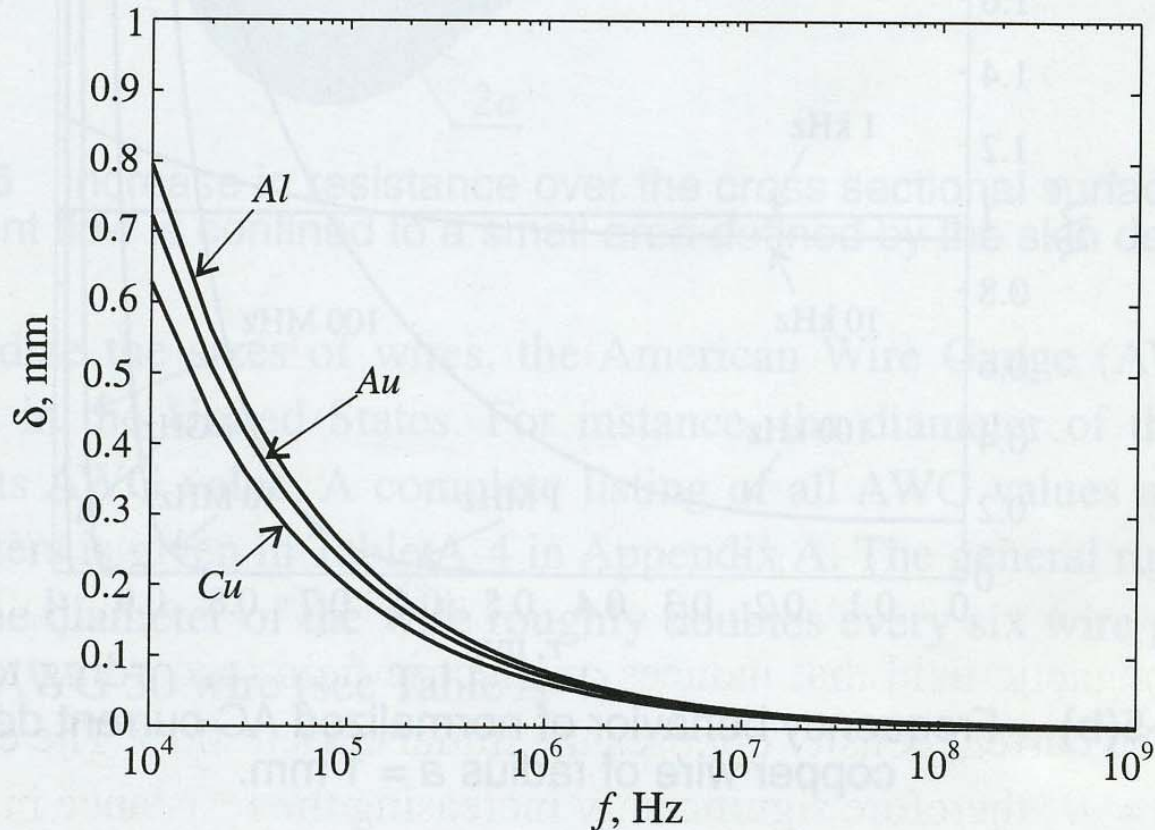


Figure 1-4 Skin depth behavior of copper $\sigma_{Cu} = 64.516 \times 10^6$ S/m, aluminum $\sigma_{Al} = 40.0 \times 10^6$ S/m, and gold $\sigma_{Au} = 48.544 \times 10^6$ S/m.

Strømtetthet ved ulike frekvenser

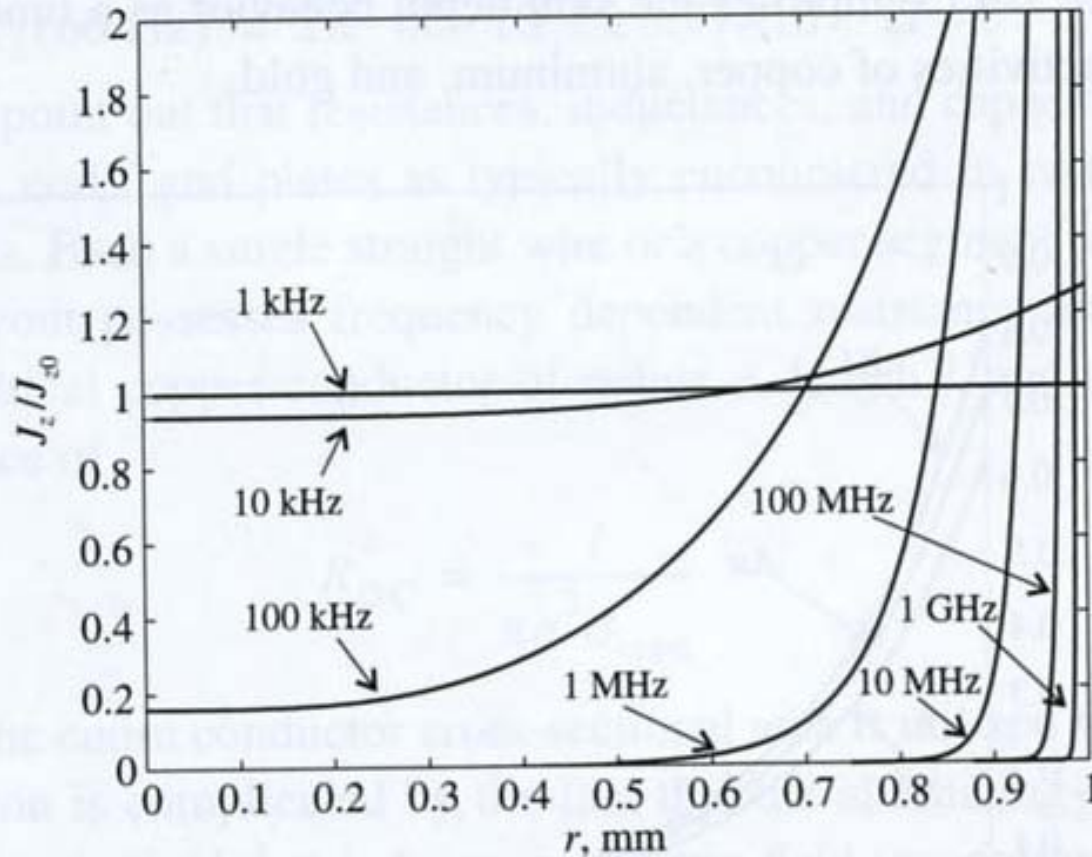


Figure 1-5(b) Frequency behavior of normalized AC current density for a copper wire of radius $a = 1$ mm.

Passive komponenter ved høje frekvenser

- Ekvivalentkrets for resistor

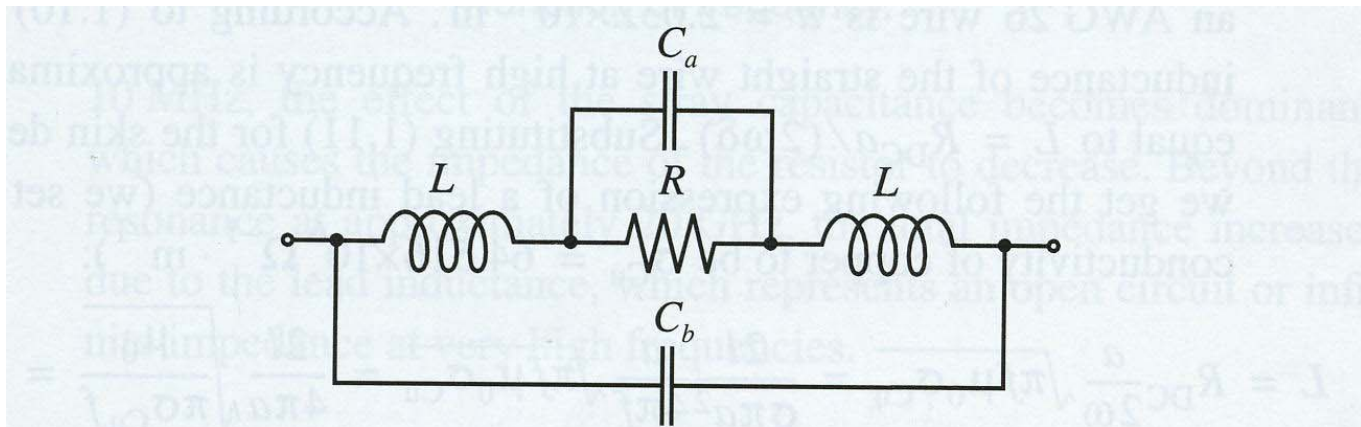
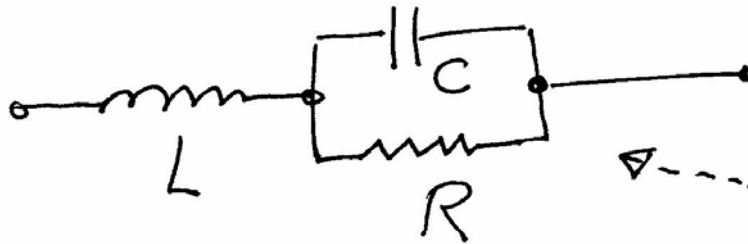


Figure 1-8 Electric equivalent circuit representation of the resistor.

Beregning av resistor-impedans

Forenklet modell:

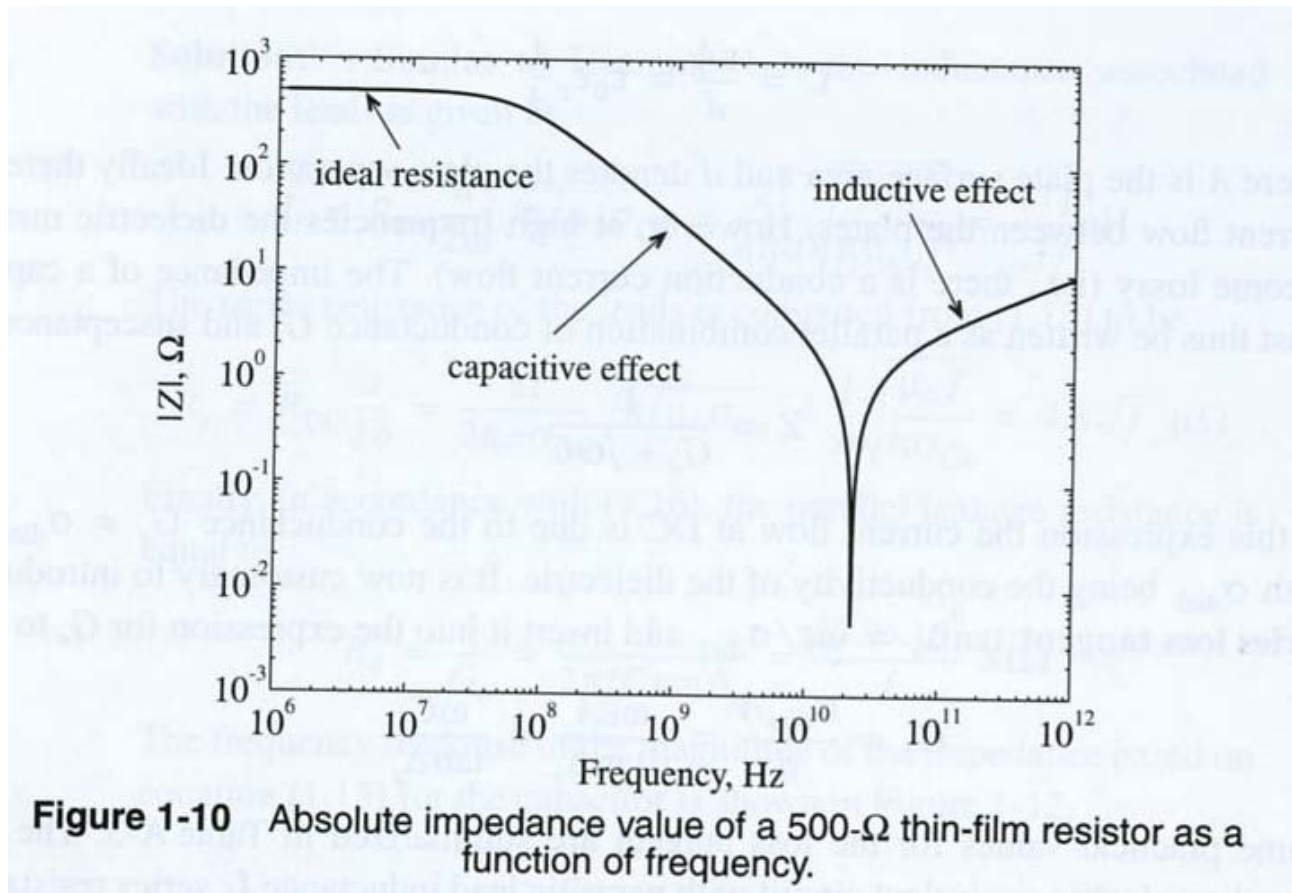


$$G = \frac{1}{R} + sC$$

$$z = sL + \frac{1}{\frac{1}{R} + sC} = sL + \frac{R}{1 + sRC}$$

$$z(j\omega) = j\omega L + \frac{R}{1 + j\omega RC}$$

Impedans som funksjon av f



Drøfting :

$$z(j\omega) \rightarrow R, \text{ når } \omega \rightarrow 0$$

$$z(j\omega) \rightarrow j\omega L, \text{ når } \omega \rightarrow \infty$$

Resonans når leddene er motsatt like store

$$sL = -\frac{R}{1 + sRC}$$

$$LRCs^2 + Ls + R = 0$$

$$s^2 + \frac{1}{RC}s + \frac{1}{LC} = 0$$

$$s = -\frac{1}{2RC} \pm j\sqrt{\frac{1}{LC} - \frac{1}{4R^2C^2}}$$

Høyfrekvens kapasitans

- Ekvivalenteskjema

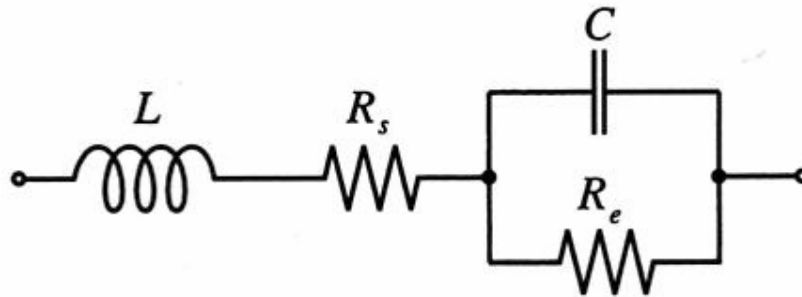


Figure 1-11 Electric equivalent circuit for a high-frequency capacitor.

Impedans som funksjon av f

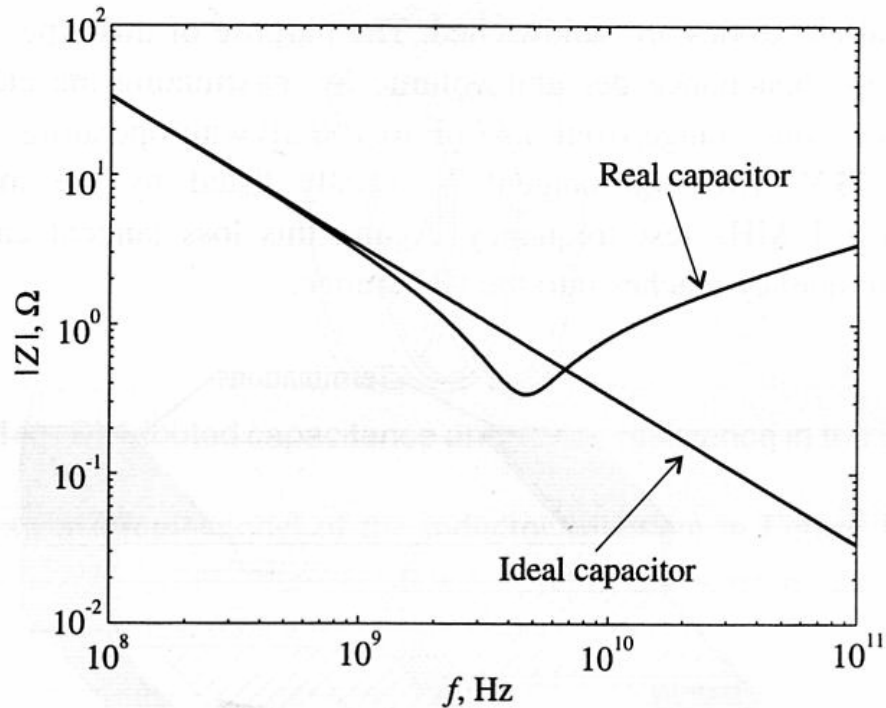


Figure 1-12 Absolute value of the capacitor impedance as a function of frequency.

Høyfrekvens induktans

- Ekvivalentskjema

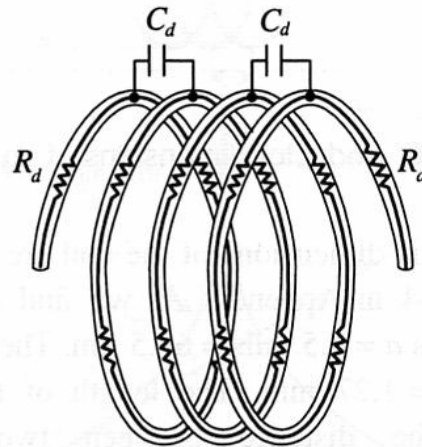


Figure 1-14 Distributed capacitance and series resistance in the inductor coil.

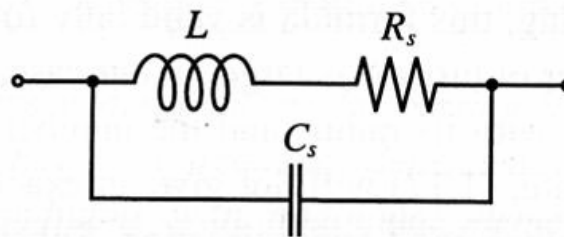


Figure 1-15 Equivalent circuit of the high-frequency inductor.

Impedans som funksjon av f

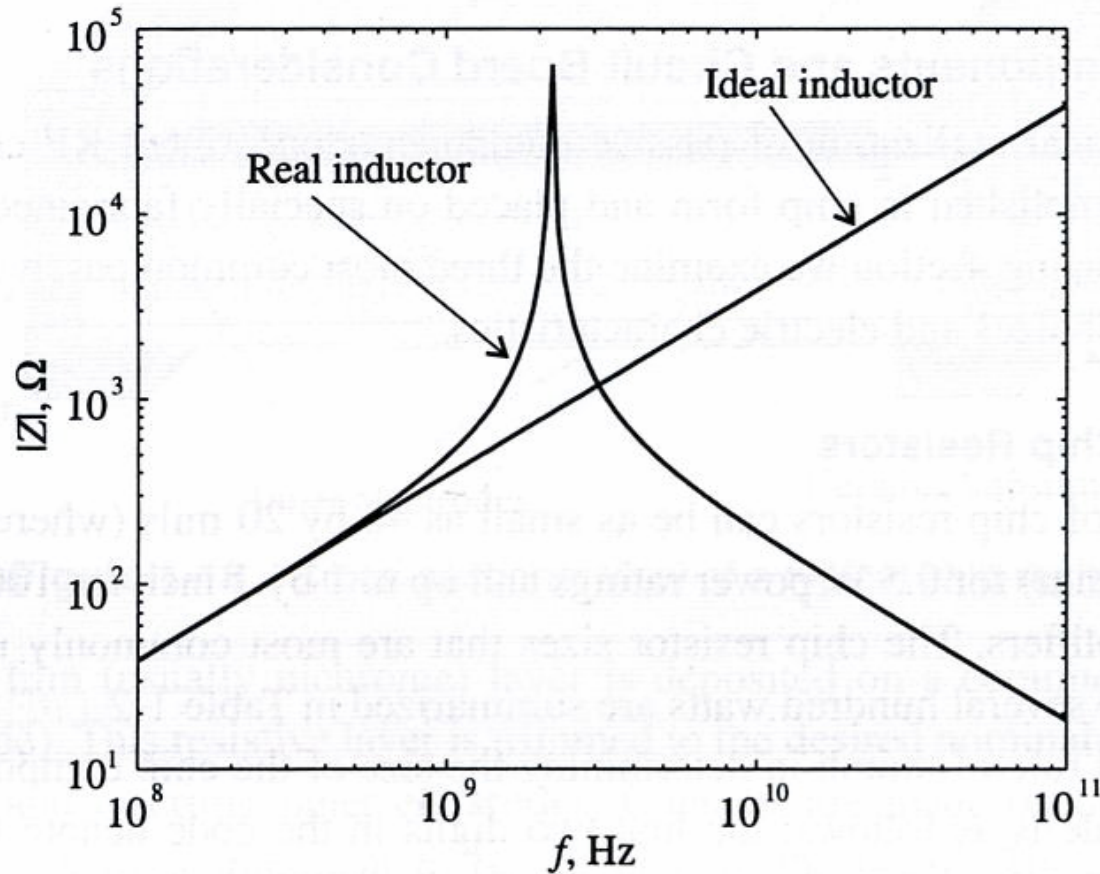


Figure 1-17 Frequency response of the impedance of an RFC.

Transmisjonslinje-teori

- Frekvens øker \rightarrow bølgelengde avtar (λ)
- Når λ blir sammenlignbar med komponentstørrelsen, vil en oppleve et **spenningsfall over komponenten!!**
 - Strøm og spenning har ikke en konstant verdi
- Spenning og strøm oppfører seg som **bølger** som utbrer seg i ledere og komponenter
 - De har derfor en verdi som avhenger av hvor en måler \rightarrow
 - Signalene må utbres ved **transmisjonslinjer**
 - Må ta hensyn til **refleksjoner, karakteristiske impedanser**

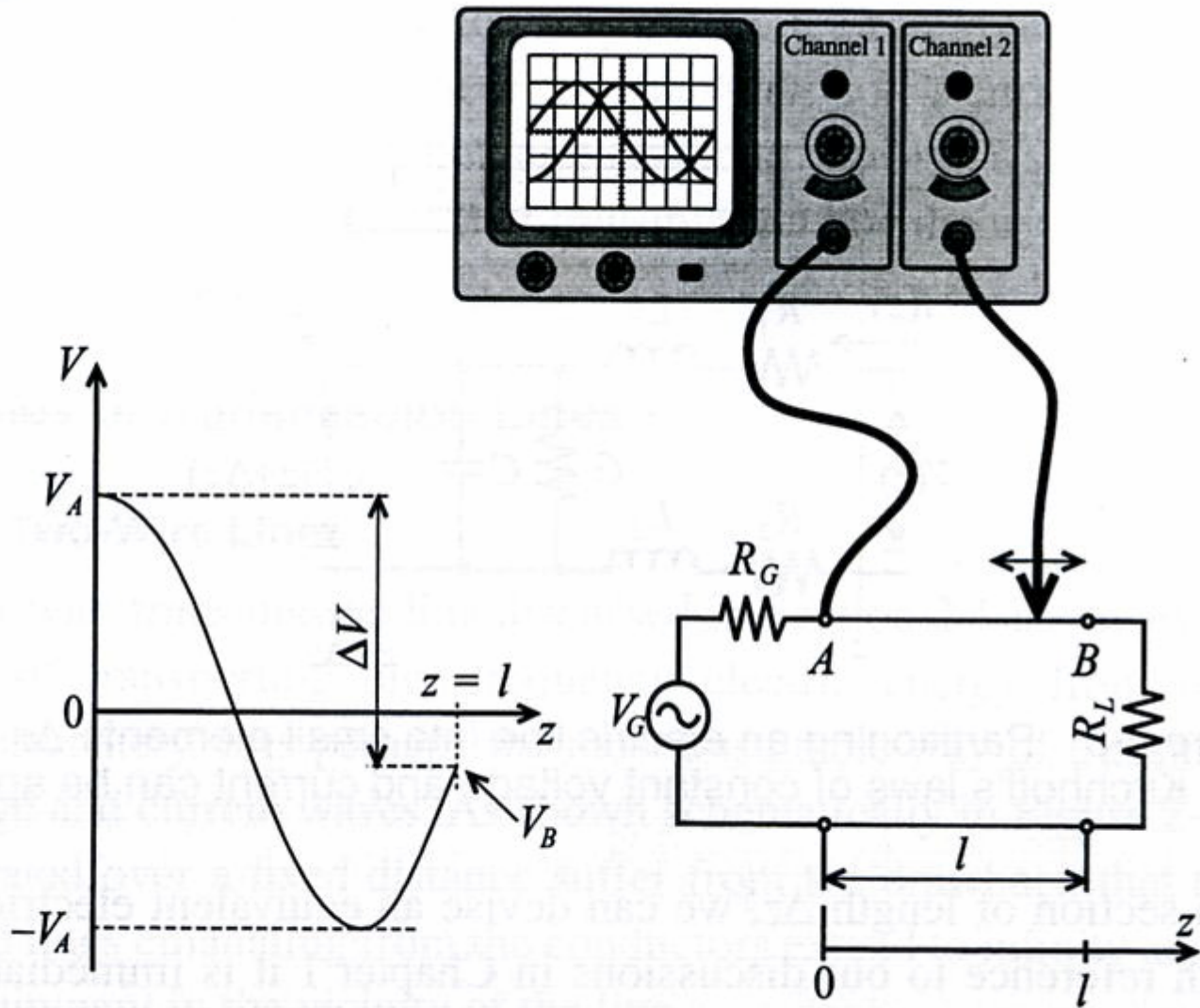


Figure 2-2 Amplitude measurements of 10 GHz voltage signal at the beginning (location A) and somewhere in between a wire connecting load to source.

Transmisjonslinje

- En leder kan modelleres som en transmisjonslinje

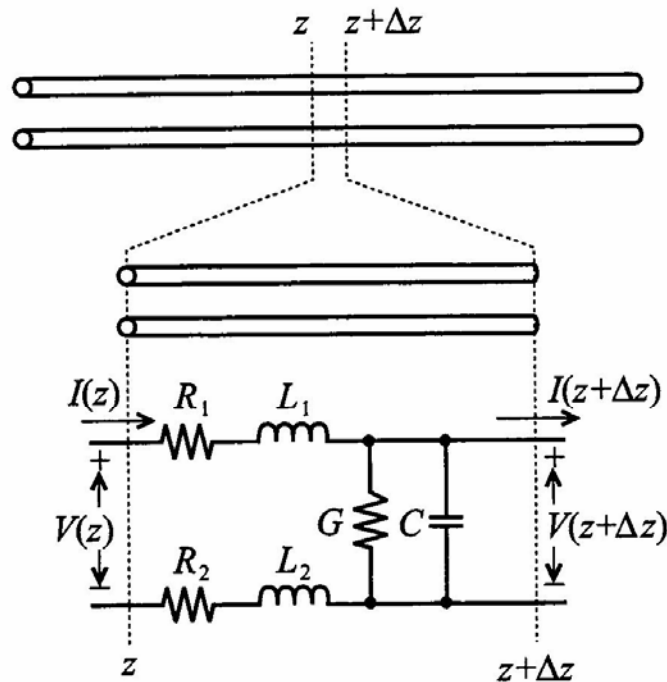


Figure 2-3 Partitioning an electric line into small elements Δz over which Kirchhoff's laws of constant voltage and current can be applied.

Kan deles opp i infinitesimale sub- enheter

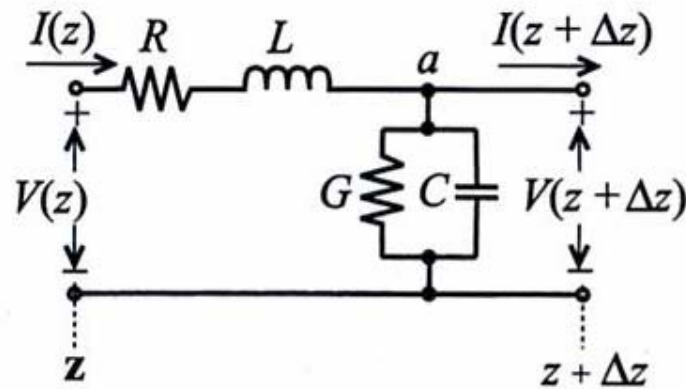


Figure 2-17 Segment of a transmission line with voltage loop and current node.

Benytte Kirchhoffs lover

- Gir 2 koblede 1.ordens diff-ligninger

$$(R + j\omega L)I(z)\Delta z + V(z + \Delta z) = V(z) \quad (2.26)$$

$$\lim_{\Delta z \rightarrow 0} \left(-\frac{V(z + \Delta z) - V(z)}{\Delta z} \right) = -\frac{dV(z)}{dz} = (R + j\omega L)I(z) \quad (2.27)$$

$$-\frac{dV(z)}{dz} = (R + j\omega L)I(z) \quad (2.28)$$

$$I(z) - V(z + \Delta z)(G + j\omega C)\Delta z = I(z + \Delta z) \quad (2.29)$$

$$\lim_{\Delta z \rightarrow 0} \frac{I(z + \Delta z) - I(z)}{\Delta z} = \frac{dI(z)}{dz} = -(G + j\omega C)V(z) \quad (2.30)$$

$$\frac{d^2 V(z)}{dz^2} - k^2 V(z) = 0 \quad (2.31)$$

$$k = k_r + jk_i = \sqrt{(R + j\omega L)(G + j\omega C)} \quad (2.32)$$

$$\frac{d^2 I(z)}{dz^2} - k^2 I(z) = 0 \quad (2.33)$$

Løsning: 2 bølger

- Løsningen er bølger i **positiv** og **negativ** retning

$$V(z) = V^+ e^{-kz} + V^- e^{+kz} \quad (2.34)$$

$$I(z) = I^+ e^{-kz} + I^- e^{+kz} \quad (2.35)$$

$$I(z) = \frac{k}{(R + j\omega L)} (V^+ e^{-kz} - V^- e^{+kz}) \quad (2.36) \quad (\text{Jmfr.2.27})$$

Karakteristisk linje-impedans: $Z_0 = \frac{V^+}{I^+} = -\frac{V^-}{I^-}$

$$Z_0 = \frac{(R + j\omega L)}{k} = \sqrt{\frac{(R + j\omega L)}{(G + j\omega C)}} \quad (2.37)$$

Impedans ved **tapsfri** transmisjonslinje

$$Z_0 = \sqrt{L/C}$$

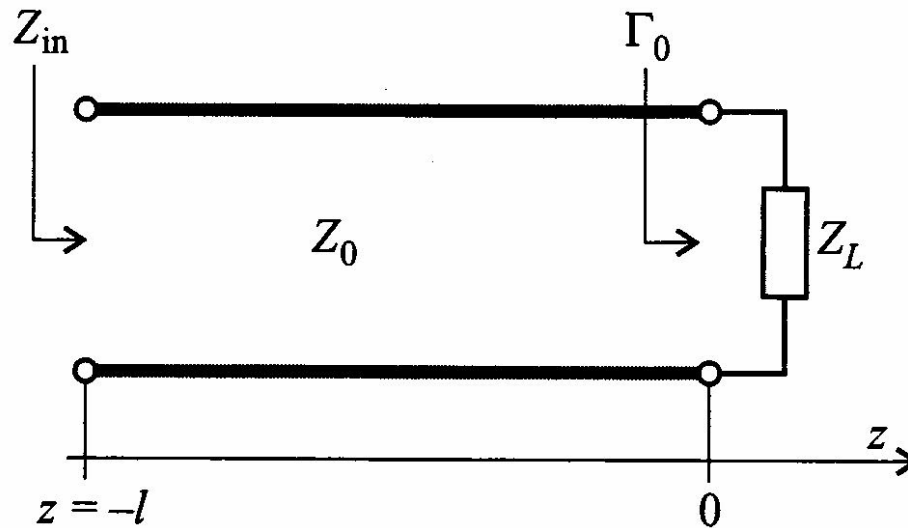


Figure 2-23 Terminated transmission line at location $z = 0$.

Refleksjon

- Hvordan hindre refleksjoner og sørge for god signalforplantning?
- Definisjon av **refleksjonskoeffisient** →

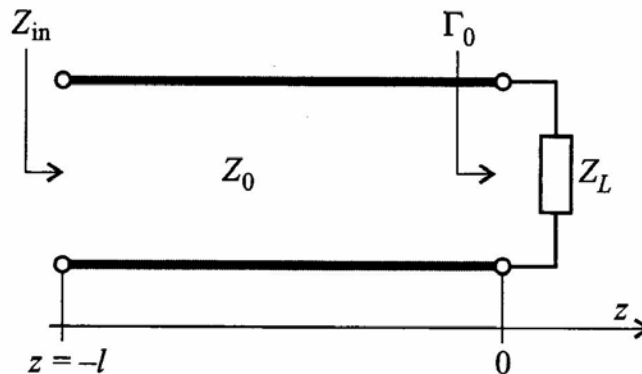


Figure 2-23 Terminated transmission line at location $z = 0$.

Refleksjonskoeffisient

$$\Gamma_0 = \frac{V^-}{V^+} \quad \leftarrow \text{definisjon av refleksjonskoeffisient i } z = 0$$

$$V(z) = V^+ (e^{-kz} + \Gamma_0 \cdot e^{+kz})$$

$$I(z) = \frac{V^+}{Z_0} (e^{-kz} - \Gamma_0 \cdot e^{+kz})$$

Impedans i $z = 0$:

$$Z(0) = \frac{V(0)}{I(0)} = Z_0 \frac{1 + \Gamma_0}{1 - \Gamma_0} = Z_L \quad = \text{lastimpedansen}$$

$$\Gamma_0 = \frac{Z_L - Z_0}{Z_L + Z_0}$$

Ulike termineringer

$$\Gamma_0 = \frac{Z_L - Z_0}{Z_L + Z_0}$$

Åpen linje

→ refleksjon med lik polaritet

$$Z_L = \infty \Rightarrow \Gamma_0 = 1$$

Kortslutning

→ refleksjon med invers polaritet

$$Z_L = 0 \Rightarrow \Gamma_0 = -1$$

Ingen refleksjon når:

$$Z_0 = Z_L \Rightarrow \Gamma_0 = 0$$

→ "MATCHING"

Stående bølger

- Kortsluttet krets gir stående bølger ($Z_L = 0$)

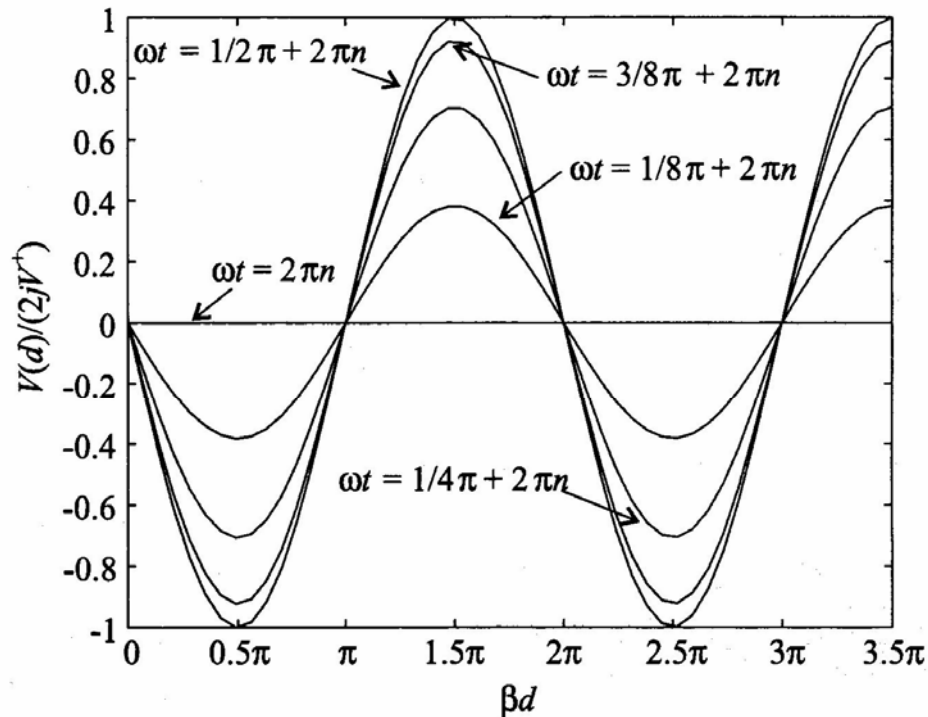


Figure 2-25 Standing wave pattern for various instances of time.

RF-kretser

- En høyfrekvens-krets kan betraktes som
 - en samling av et endelig antall **transmisjonslinje-seksjoner** forbundet med **diskrete aktive** og **passive** komponenter

To-port nettverk

- Fordelaktig med **to-port-beskrivelse**
 - Kretser kan deles opp i enkle bestanddeler
 - to-porter
 - Kan benyttes for å forenkle analysen av sammensatte nettverk
- Ulike typer to-porter
 - **Z, Y, h-matrix**
 - Hver har ulike egenskaper ved sammenkobling
 - $Z \rightarrow$ serie, $Y \rightarrow$ parallell, hybrid \rightarrow blanding
 - Figur \rightarrow

Multiport-nettverk

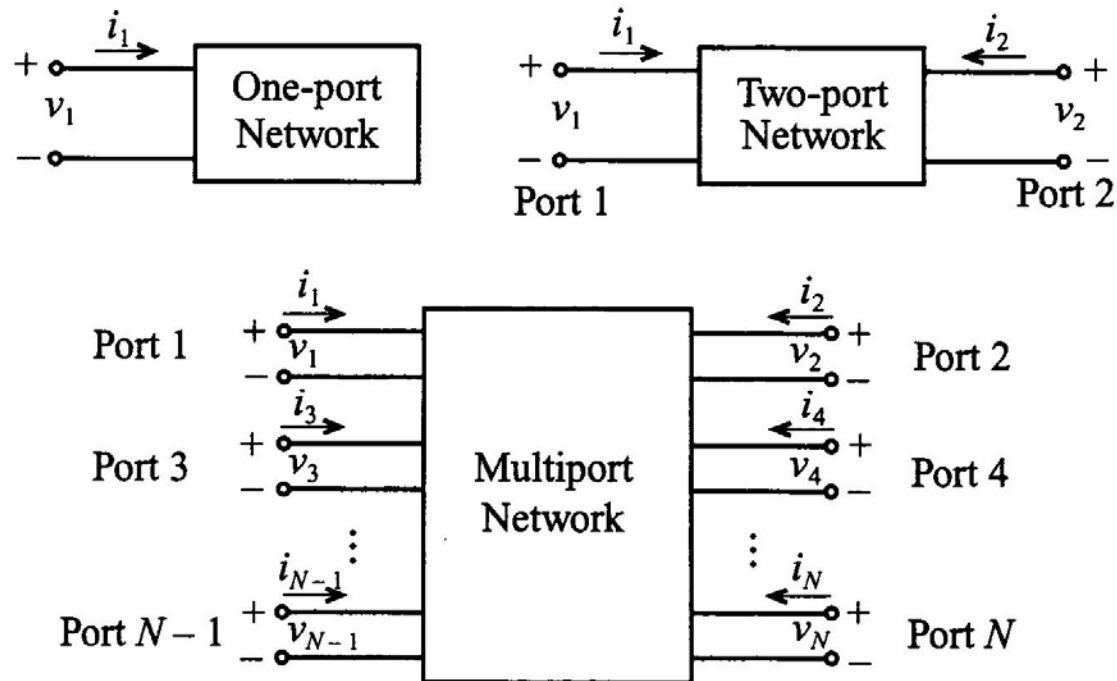


Figure 4-1 Basic voltage and current definitions for single- and multiport network.

Eks. Z-matrise

$$\begin{Bmatrix} v_1 \\ v_2 \\ \vdots \\ v_N \end{Bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} & \cdots & Z_{1N} \\ Z_{21} & Z_{22} & \cdots & Z_{2N} \\ \vdots & \vdots & \ddots & \vdots \\ Z_{N1} & Z_{N2} & \cdots & Z_{NN} \end{bmatrix} \begin{Bmatrix} i_1 \\ i_2 \\ \vdots \\ i_N \end{Bmatrix} \quad (4.2)$$

$$\{\mathbf{V}\} = [\mathbf{Z}]\{\mathbf{I}\} \quad (4.3)$$

ABCD nettverk

$$\begin{Bmatrix} v_1 \\ i_1 \end{Bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{Bmatrix} v_2 \\ -i_2 \end{Bmatrix} \quad (4.10)$$

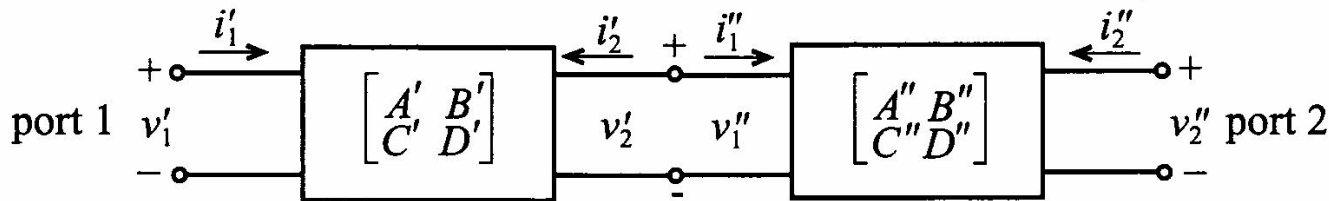


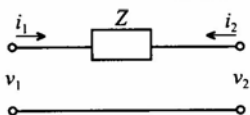
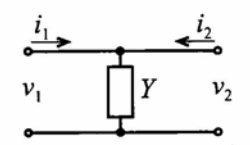
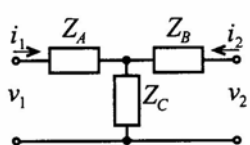
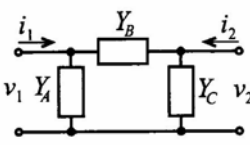
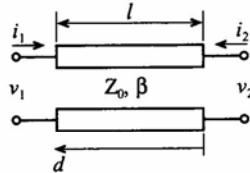
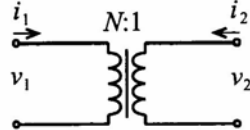
Figure 4-9 Cascading two networks.

$$\begin{aligned} \begin{Bmatrix} v_1 \\ i_1 \end{Bmatrix} &= \begin{Bmatrix} v_1' \\ i_1' \end{Bmatrix} = \begin{bmatrix} A' & B' \\ C' & D' \end{bmatrix} \begin{Bmatrix} v_2' \\ -i_2' \end{Bmatrix} = \begin{bmatrix} A' & B' \\ C' & D' \end{bmatrix} \begin{Bmatrix} v_1'' \\ i_1'' \end{Bmatrix} \\ &= \begin{bmatrix} A' & B' \\ C' & D' \end{bmatrix} \begin{bmatrix} A'' & B'' \\ C'' & D'' \end{bmatrix} \begin{Bmatrix} v_2'' \\ -i_2'' \end{Bmatrix} \end{aligned} \quad (4.21)$$

Velegnet for kaskadekobling

ABCD-parametre for "nyttige" 2-porter

Table 4-1 ABCD-Parameters of Some Useful Two-Port Circuits.

Circuit	ABCD-Parameters	
	$A = 1$ $C = 0$	$B = Z$ $D = 1$
	$A = 1$ $C = Y$	$B = 0$ $D = 1$
	$A = 1 + \frac{Z_A}{Z_C}$ $C = \frac{1}{Z_C}$	$B = Z_A + Z_B + \frac{Z_A Z_B}{Z_C}$ $D = 1 + \frac{Z_B}{Z_C}$
	$A = 1 + \frac{Y_B}{Y_C}$ $C = Y_A + Y_B + \frac{Y_A Y_B}{Y_C}$	$B = \frac{1}{Y_C}$ $D = 1 + \frac{Y_A}{Y_C}$
	$A = \cos \beta l$ $C = \frac{j \sin \beta l}{Z_0}$	$B = j Z_0 \sin \beta l$ $D = \cos \beta l$
	$A = N$ $C = 0$	$B = 0$ $D = \frac{1}{N}$

Konvertering mellom ulike 2-port realiseringer

Table 4-2 Conversion between Different Network Representations

	[Z]	[Y]	[h]	[ABCD]
[Z]	Z_{11} Z_{12} Z_{21} Z_{22}	$\frac{Z_{22}}{\Delta Z}$ $\frac{Z_{12}}{-\Delta Z}$ $\frac{Z_{21}}{-\Delta Z}$ $\frac{Z_{11}}{\Delta Z}$	$\frac{\Delta Z}{Z_{22}}$ $\frac{Z_{12}}{Z_{22}}$ $\frac{Z_{21}}{-Z_{22}}$ $\frac{1}{Z_{22}}$	$\frac{Z_{11}}{Z_{21}}$ $\frac{\Delta Z}{Z_{21}}$ $\frac{1}{Z_{21}}$ $\frac{Z_{22}}{Z_{21}}$
[Y]	$\frac{Y_{22}}{\Delta Y}$ $\frac{Y_{12}}{-\Delta Y}$ $\frac{Y_{21}}{-\Delta Y}$ $\frac{Y_{11}}{\Delta Y}$	Y_{11} Y_{12} Y_{21} Y_{22}	$\frac{1}{Y_{11}}$ $\frac{Y_{12}}{-Y_{11}}$ $\frac{Y_{21}}{Y_{11}}$ $\frac{\Delta Y}{Y_{11}}$	$\frac{Y_{22}}{-Y_{21}}$ $\frac{1}{-Y_{21}}$ $\frac{\Delta Y}{-Y_{21}}$ $\frac{Y_{11}}{-Y_{21}}$
[h]	$\frac{\Delta h}{h_{22}}$ $\frac{h_{12}}{h_{22}}$ $\frac{h_{21}}{-h_{22}}$ $\frac{1}{h_{22}}$	$\frac{1}{h_{11}}$ $\frac{h_{12}}{h_{11}}$ $\frac{h_{21}}{h_{11}}$ $\frac{\Delta h}{h_{11}}$	h_{11} h_{12} h_{21} h_{22}	$\frac{\Delta h}{h_{21}}$ $\frac{h_{11}}{h_{21}}$ $\frac{h_{22}}{h_{21}}$ $\frac{1}{h_{21}}$
[ABCD]	$\frac{A}{C}$ $\frac{\Delta ABCD}{C}$ $\frac{1}{C}$ $\frac{D}{C}$	$\frac{D}{B}$ $\frac{\Delta ABCD}{B}$ $\frac{1}{-B}$ $\frac{A}{B}$	$\frac{B}{D}$ $\frac{\Delta ABCD}{D}$ $\frac{1}{-D}$ $\frac{C}{D}$	A B C D

S-parametre

- 2-port benyttes for definisjon av S-parametre
- "Power waves" defineres som

$$a_n = \frac{1}{2\sqrt{Z_0}}(V_n + Z_0 I_n) \quad (4.36a)$$

$$b_n = \frac{1}{2\sqrt{Z_0}}(V_n - Z_0 I_n) \quad (4.36b)$$

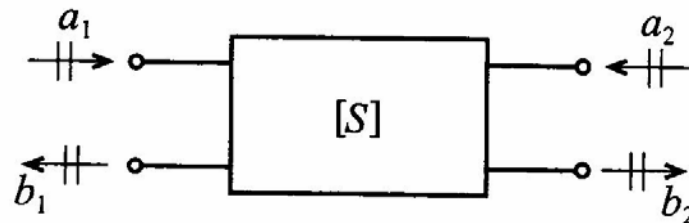


Figure 4-14 Convention used to define S-parameters for a two-port network.

Definisjon av S-parametrene

- Beregninger viser at effekten (power) blir:

$$P_n = \frac{1}{2} \operatorname{Re}\{V_n I_n^*\} = \frac{1}{2} (|a_n|^2 - |b_n|^2)$$

S-parametre

$$\begin{Bmatrix} b_1 \\ b_2 \end{Bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{Bmatrix} a_1 \\ a_2 \end{Bmatrix} \quad (4.41)$$

Hva hver enkelt S-parameter betyr

$$S_{11} = \left. \frac{b_1}{a_1} \right|_{a_2=0} \equiv \frac{\text{reflected power wave at port 1}}{\text{incident power wave at port 1}} \quad (4.42a)$$

$$S_{21} = \left. \frac{b_2}{a_1} \right|_{a_2=0} \equiv \frac{\text{transmitted power wave at port 2}}{\text{incident power wave at port 1}} \quad (4.42b)$$

$$S_{22} = \left. \frac{b_2}{a_2} \right|_{a_1=0} \equiv \frac{\text{reflected power wave at port 2}}{\text{incident power wave at port 2}} \quad (4.42c)$$

$$S_{12} = \left. \frac{b_1}{a_2} \right|_{a_1=0} \equiv \frac{\text{transmitted power wave at port 1}}{\text{incident power wave at port 2}} \quad (4.42d)$$

Måling av S-parametre

- S-parametrene måles når linjene er **terminert** med sin **karakteristiske impedans**

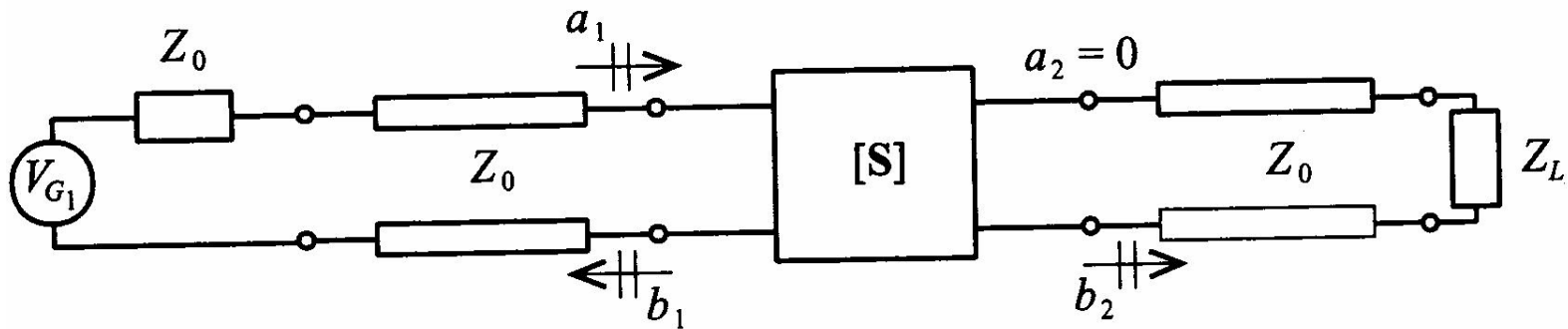


Figure 4-15 Measurement of S_{11} and S_{21} by matching the line impedance Z_0 at port 2 through a corresponding load impedance $Z_L = Z_0$.

Filtre

- Ulike filtertyper

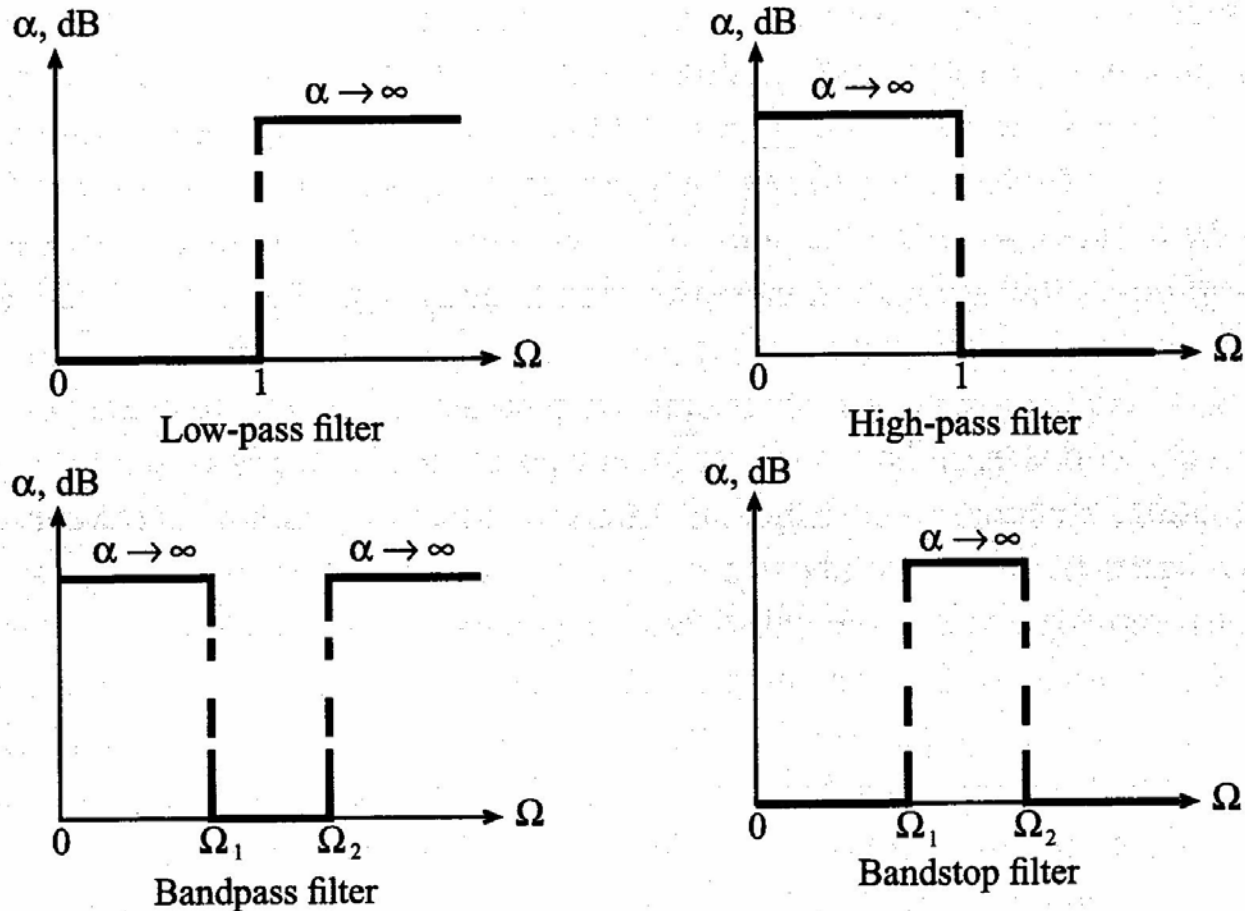


Figure 5-1 Four basic filter types.

Eks. på 3 ulike filtertyper

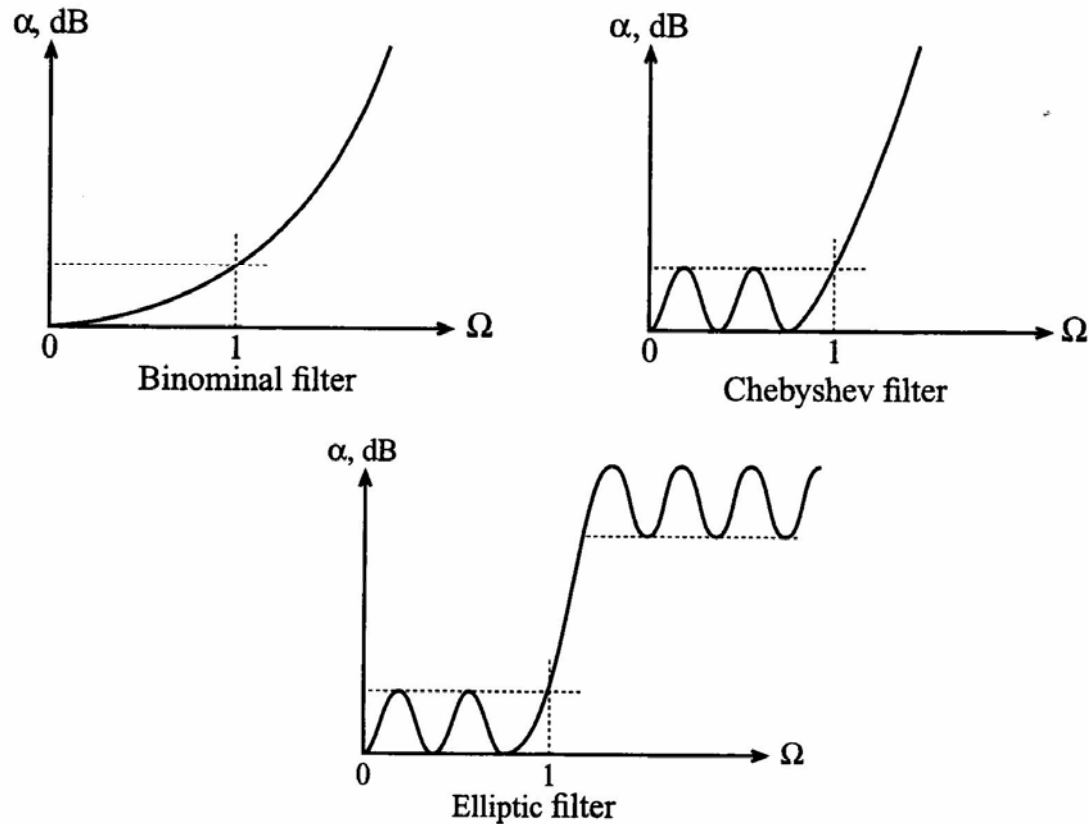


Figure 5-2 Actual attenuation profile for three types of low-pass filters.

Ulike betegnelser, begreper

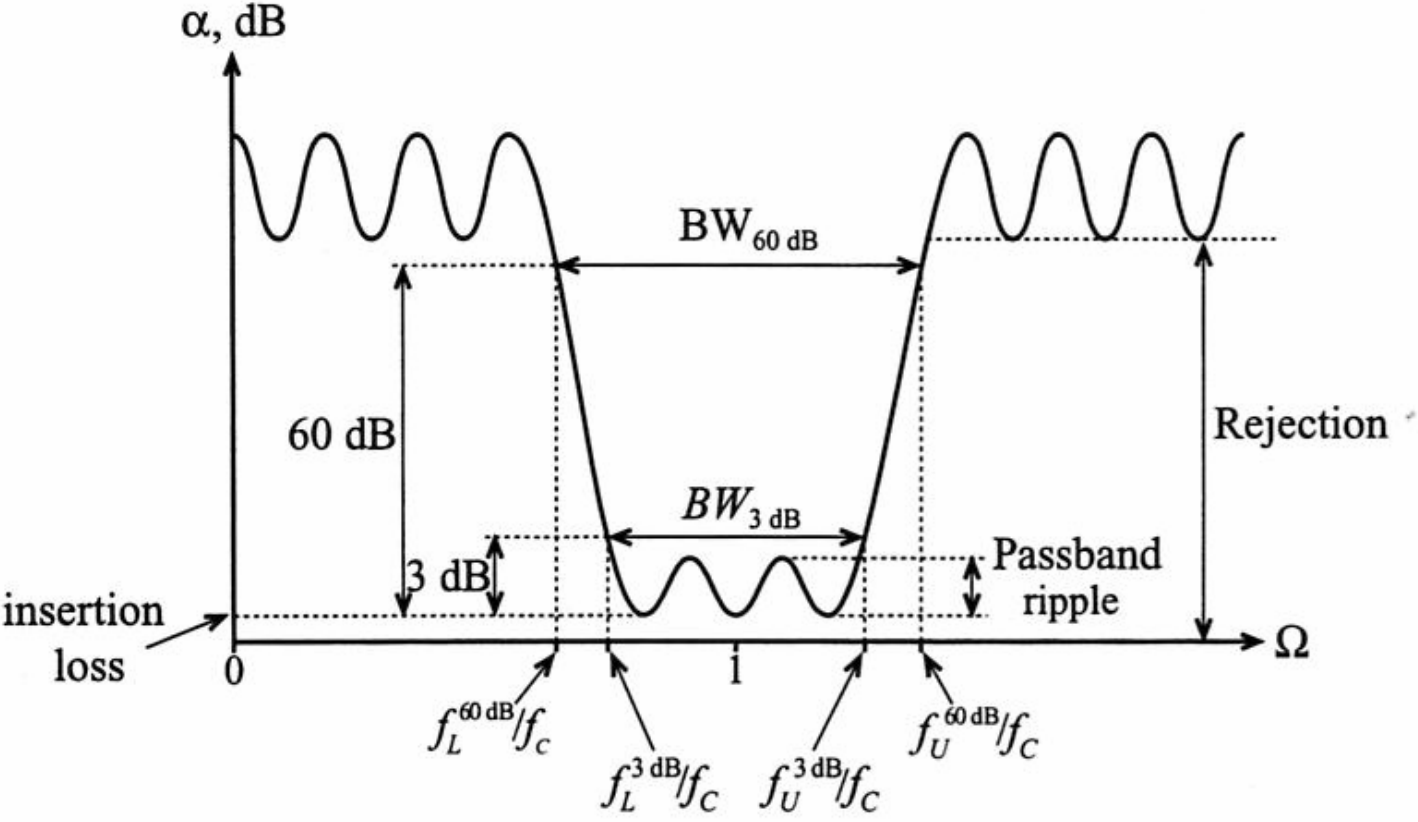


Figure 5-3 Generic attenuation profile for a bandpass filter.

Q-faktor

- Definisjon av **Q-faktor**

$$Q = \omega \left. \frac{\text{average stored energy}}{\text{energy loss per cycle}} \right|_{\omega = \omega_c} = \omega \left. \frac{\text{average stored energy}}{\text{power loss}} \right|_{\omega = \omega_c} = \omega \left. \frac{W_{\text{stored}}}{P_{\text{loss}}} \right|_{\omega = \omega_c} \quad (5.4)$$

- Det finnes ulike definisjoner av Q-faktor
 - Definisjonene er ekvivalente

$$Q_{LD} = \frac{f_c}{f_U^{3\text{dB}} - f_L^{3\text{dB}}} \equiv \frac{f_c}{BW^{3\text{dB}}}$$

Unloaded – loaded Q

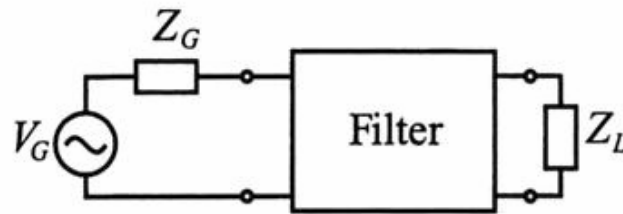


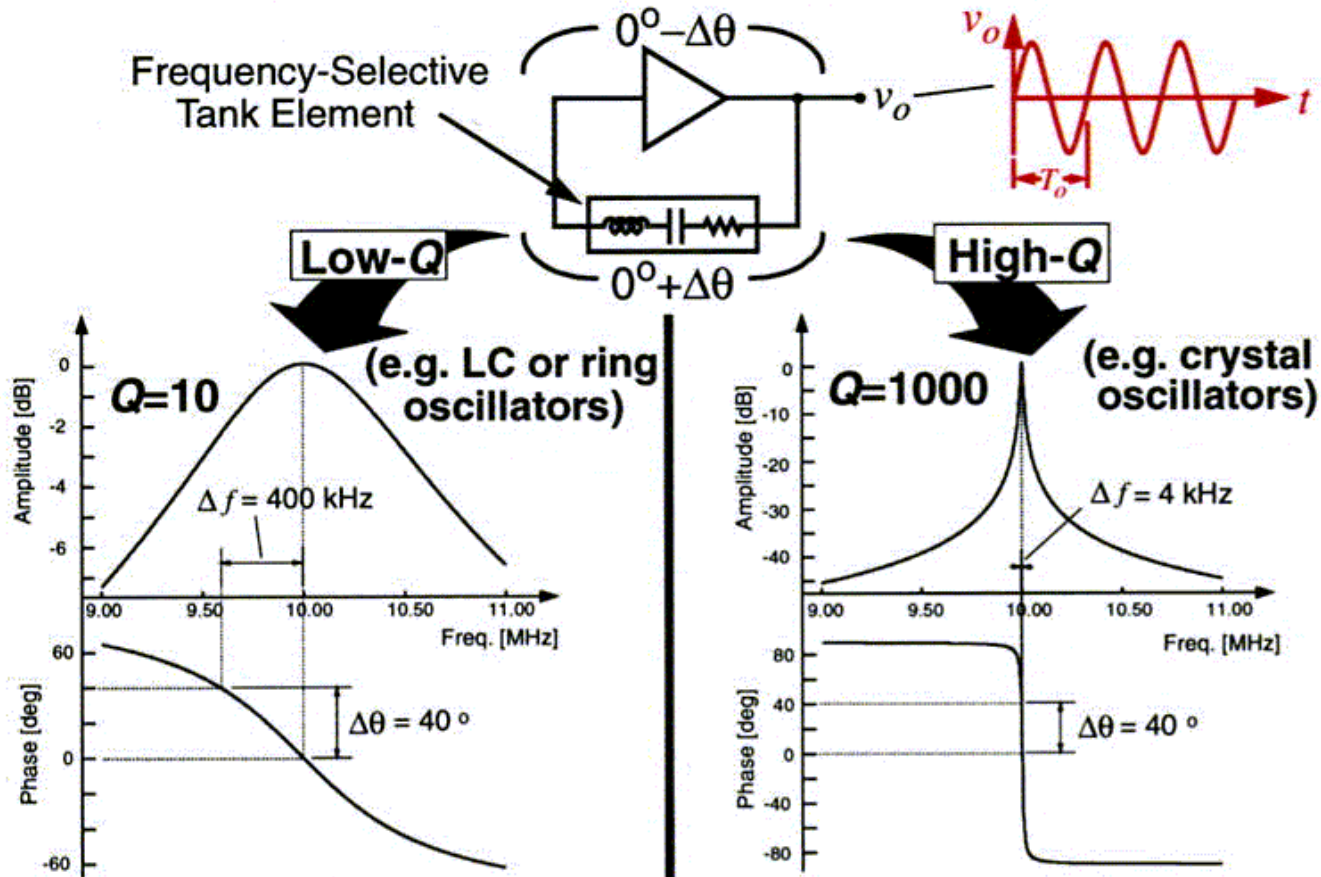
Figure 5-4 Filter as a two-port network connected to an RF source and load.

$$\frac{1}{Q_{LD}} = \frac{1}{\omega} \left(\frac{\text{power loss in filter}}{\text{average stored energy}} \right) \Bigg|_{\omega = \omega_r} + \frac{1}{\omega} \left(\frac{\text{power loss in load}}{\text{average stored energy}} \right) \Bigg|_{\omega = \omega_r} \quad (5.5)$$

$$\frac{1}{Q_{LD}} = \frac{1}{Q_F} + \frac{1}{Q_E}$$

Hva Q-faktoren betyr i praksis for stabilitet

Achieving High Oscillator Stability



- High tank $Q \Rightarrow$ high frequency stability