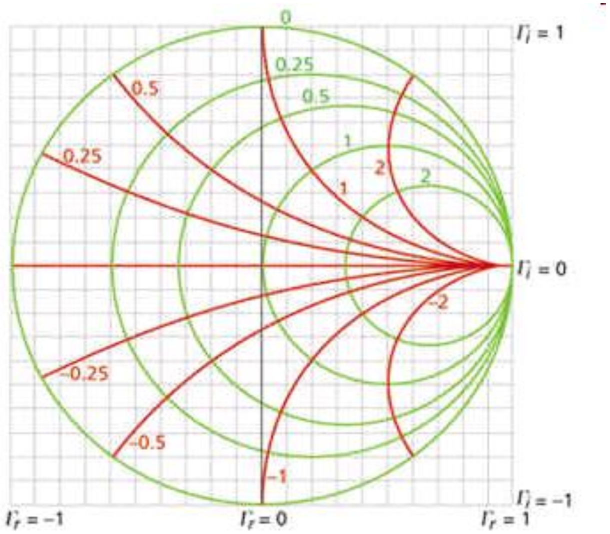
JUAS RF Course 2012

Cavities

RF Theory

Superconducting LEP cavity



The Smith Chart

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Part I: Passive and Active Elements

Section A

- Basics
- Cavity structures
- Equivalent circuit
- Behaviour in time and in frequency domain
- Beam-cavity interaction
- Scaling laws
- Simulation techniques

Section B

- Higher order modes (HOMs)
- Coupling and tuning
- Different forms of cavities
- Voltage breakdown & Multipactor

Section C

- Groups of cavities
- Transmission lines
- Striplines, Microstriplines, Slotlines
- Waveguides
- Active elements
  - Transistors
  - Gridded tubes
  - Klystrons
  - IOTs

Part II: Waves, S-Parameters, Decibels and Smith Chart

Section A

- Forward and backward travelling waves

Section B

- S-Parameters
- The scattering matrix
- Decibels
- Measurement devices and concepts
- Superheterodyne Concept

Section C

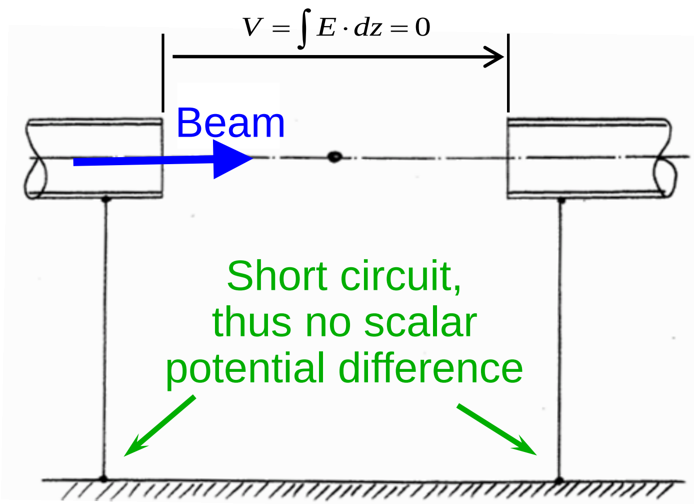
- The Smith Chart
- Navigation in the Smith Chart
- Examples

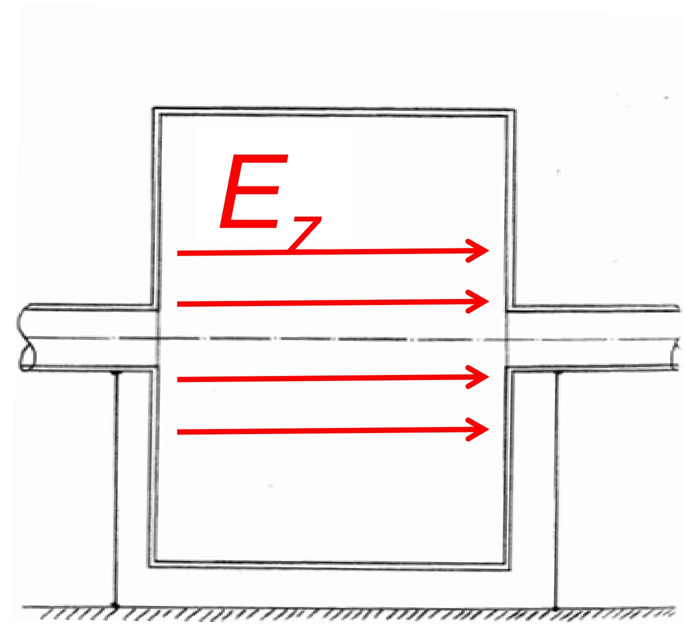
From L and C to a cavity

If you open the beam pipe then both ends are at the same potential

→ put a cavity in there →

Creates E-field for accelerating the particles



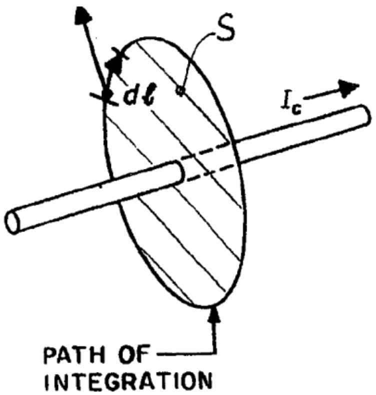


Capacitor at high frequencies,
The Feynman Lectures on Physics

Can the short-circuit be avoided?

Answer: No - but it doesn't bother us at high frequencies.

Maxwell's equations (1)



Ampere's Law:

$$\oint H \cdot dl = I = I_{conduction} + I_{displacement}$$

$$I_{displacement} = \frac{\partial \Phi_D}{\partial t}$$

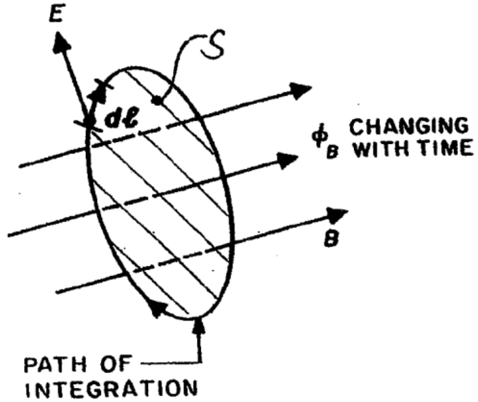
where the electric flux Φ_D is given by

$$\Phi_D = \int_S D \cdot dS = \varepsilon \int_S E \cdot dS,$$

D designating the electric flux density.

$$\nabla \times H = J + \frac{\partial D}{\partial t}$$

with the current density J and the magnetic field H



Faraday's Law:

$$\oint E \cdot dl = -\frac{\partial \Phi_B}{\partial t}$$

with the electric flux Φ_B

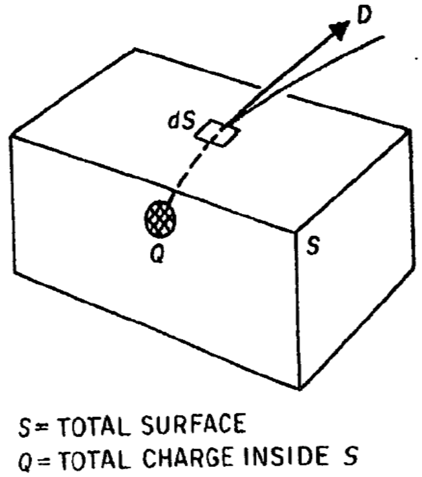
$$\Phi_B = \int_S B dS = \mu \int_S H dS$$

$$\nabla \times E = -\frac{\partial B}{\partial t}$$

with the electric field E and the magnetic field B

scalar vs. vector potential: path of integration makes a difference

Maxwell's equations (2)

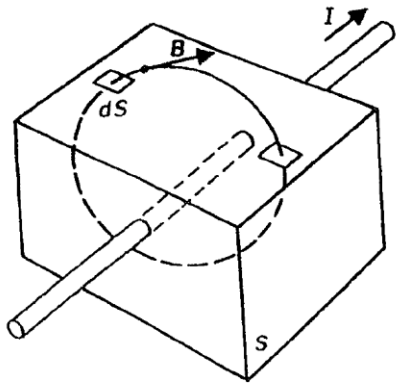


Gauss' Law (Electricity):

$$\int_S D \cdot dS = Q$$

with the electric displacement D

$\nabla \cdot D = \rho$
with the charge density ρ



Gauss' Law (Magnetism):

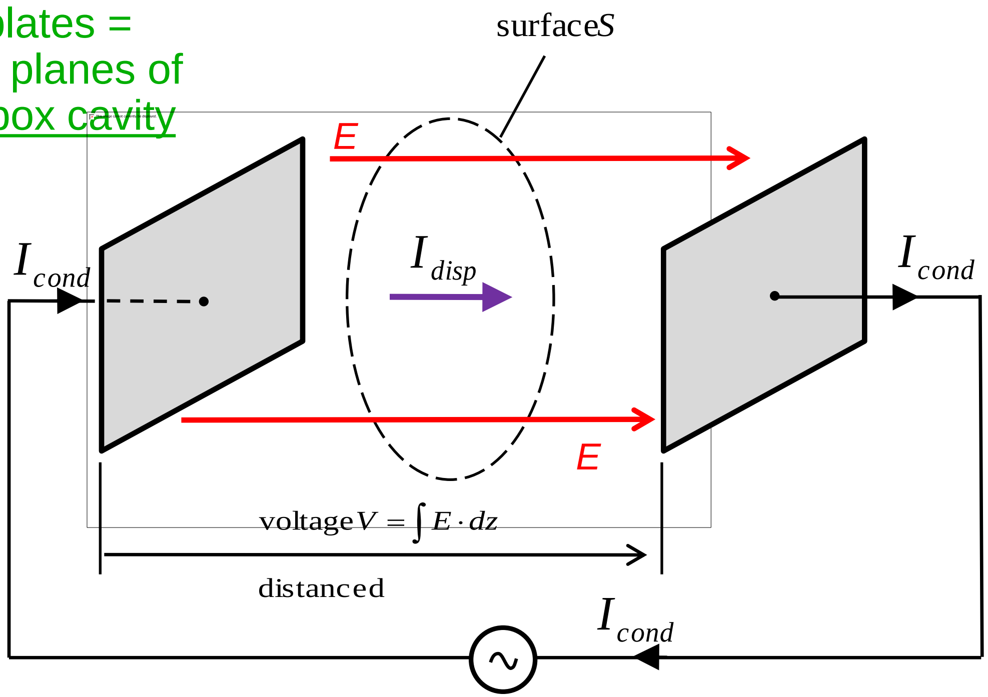
$$\int_S B \cdot dS = 0$$

$\nabla \cdot B = 0$

There are no magnetic charges

Displacement and conduction currents in a simple capacitor

end plates = sides planes of a pillbox cavity



for vacuum and approximately for air:
$\mu = \mu_0 = 4\pi * 10^{-7} = 1.2566 * 10^{-6}$ Vs/(Am)

$\varepsilon = \varepsilon_0 = 8.854 * 10^{-12}$ As/(Vm)

The conduction current continues as displacement current over the capacitor gap

Displacement current in dielectric: $$I_{disp} = \frac{\partial \Phi_D}{\partial t} = \varepsilon \int \frac{\partial E}{\partial t} \partial S = \varepsilon S \frac{\partial E}{\partial t}$$

Conduction current in conductor: $$I_{cond} = \frac{\partial Q}{\partial t} = C \frac{\partial V}{\partial t} = \varepsilon \frac{S}{d} d \frac{\partial E}{\partial t} = \varepsilon S \frac{\partial E}{\partial t}$$

with the electric flux Φ_D and the charge Q.

General Solution for a Rectangular (brick-type) Cavity

When describing field components in a Cartesian coordinates system (assuming a homogeneous and isotropic material in a space charge free volume) with harmonic functions (angular frequency ω) then each Cartesian component needs to fulfill Laplace's equation:

$$\Delta\Psi + k_0^2 \varepsilon_r \mu_r \Psi = 0$$

$$k_0^2 = \omega^2 \varepsilon_0 \mu_0 \qquad k_0 \quad \text{free space wavenumber}$$

$$k_0 = 2\pi / \lambda_0 \qquad \lambda_0 \quad \text{free space wavelength}$$

As a general solution we can use the product ansatz for Ψ

$$\Psi = X(x)Y(y)Z(z)$$

From this one obtains the general solution for Ψ (Ψ may be a vector potential or field)

$$\Psi = \begin{Bmatrix} A\cdot\cos(k_x x) + B\cdot\sin(k_x x) \\ A'\cdot e^{-jk_x x} + B'\cdot e^{jk_x x} \end{Bmatrix} \begin{Bmatrix} C\cdot\cos(k_y y) + D\cdot\sin(k_y y) \\ C'\cdot e^{-jk_y y} + D'\cdot e^{jk_y y} \end{Bmatrix} \begin{Bmatrix} E\cdot\cos(k_z z) + F\cdot\sin(k_z z) \\ E'\cdot e^{-jk_z z} + F'\cdot e^{jk_z z} \end{Bmatrix}$$

standing waves (upper row)

travelling waves (lower row)

with the separation condition

$$(k_x)^2 + (k_y)^2 + (k_z)^2 = (k_0)^2 \varepsilon_r \mu_r$$

see also: G. Dome, RF Theory
Proceeding Oxford CAS, April 91
CERN Yellow Report 92-03, Vol. I

General Solution in Cylindrical Coordinates

As a general solution we can use the product ansatz for Ψ

$$\Psi = R(\rho)F(\varphi)Z(z)$$

From this one obtains the general solution for Ψ (Ψ may be a vector potential or field)

$$\Psi = \begin{Bmatrix} A\cdot \mathrm{J}_{\mathrm{m}}(k_\rho \rho) + B\cdot \mathrm{N}_{\mathrm{m}}(k_\rho \rho) \\ A'\cdot \mathrm{H}_{\mathrm{m}}^{(2)}(k_\rho \rho) + B'\cdot \mathrm{H}_{\mathrm{m}}^{(1)}(k_\rho \rho) \end{Bmatrix} \begin{Bmatrix} C\cdot \cos(m\varphi) + D\cdot \sin(m\varphi) \\ C'\cdot \mathrm{e}^{-\mathrm{j}m\varphi} + D'\cdot \mathrm{e}^{\mathrm{j}m\varphi} \end{Bmatrix} \begin{Bmatrix} E\cdot \cos(k_z z) + F\cdot \sin(k_z z) \\ E'\cdot \mathrm{e}^{-\mathrm{j}k_z z} + F'\cdot \mathrm{e}^{\mathrm{j}k_z z} \end{Bmatrix}$$

standing waves

travelling waves

and the functions

J_{m} ...cylindrical harmonics of the Bessel function of order m

N_{m} ...cylindrical harmonics of the Neumann function of order m

$\mathrm{H}_{\mathrm{m}}^{(1)}$ ...Hankel function of the first kind of order m (outward travelling wave)

$\mathrm{H}_{\mathrm{m}}^{(2)}$ ...Hankel function of the second kind of order m (inward travelling wave)

$$\mathrm{H}_{\mathrm{m}}^{(1)} = \mathrm{J}_{\mathrm{m}} + \mathrm{j}\,\mathrm{N}_{\mathrm{m}}$$

$$\mathrm{H}_{\mathrm{m}}^{(2)} = \mathrm{J}_{\mathrm{m}} - \mathrm{j}\,\mathrm{N}_{\mathrm{m}}$$

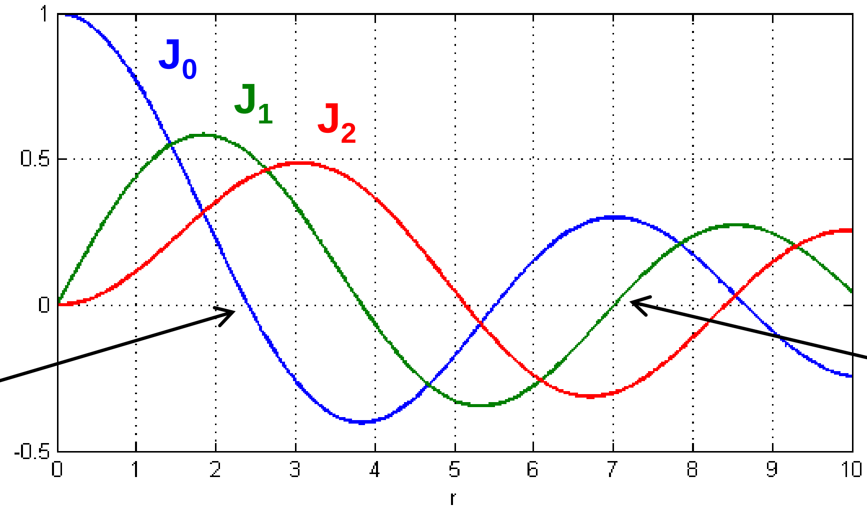
Here the separation condition is

$$\boxed{(k_\rho)^2 + (k_z)^2 = (k_0)^2 \varepsilon_r \mu_r}$$

Hint: the index m indicating the order of the Bessel and Neumann function shows up again in the argument of the sine and cosine for the azimuthal dependency.

Bessel Functions (1)

A nice example of the derivation of a Bessel function is the solution of the cylinder problem of the capacitor given in the Feynman reference (Bessel function via a series expansion).



the term "root" stands for zero-crossing

first root of the Bessel function of 0^{th} order

second root of the Bessel function of 1^{st} order

Comment: For the generalized solution of cylinder symmetrical boundary value problems (e.g. higher order modes on a coaxial resonator) Neumann functions are required. Standing wave patterns are described by Bessel- and Neumann functions respectively, radially travelling waves in terms of Hankel functions.
Hint: Sometimes a Bessel function is called Bessel function of first kind, a Neumann function is Bessel function of second kind, and a Hankel function=Bessel function of third kind.

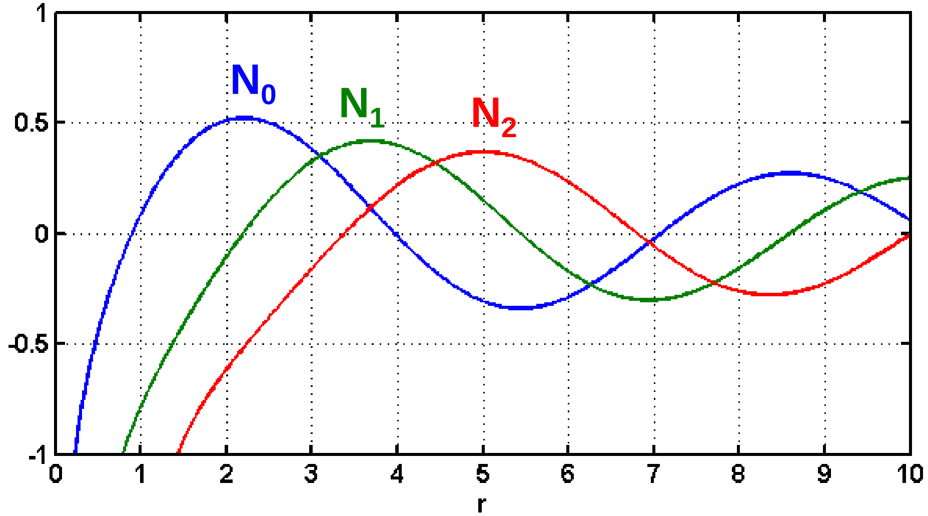
Bessel Functions (2)

Some practical numerical values:

k	$J_0(x)$	$J_1(x)$	$J_2(x)$	$J_3(x)$	$J_4(x)$	$J_5(x)$
1	2.4048	3.8317	5.1356	6.3802	7.5883	8.7715
2	5.5201	7.0156	8.4172	9.7610	11.0647	12.3386
3	8.6537	10.1735	11.6198	13.0152	14.3725	15.7002
4	11.7915	13.3237	14.7960	16.2235	17.6160	18.9801
5	14.9309	16.4706	17.9598	19.4094	20.8269	22.2178

See: http://mathworld.wolfram.com/BesselFunctionZeros.html

Neumann Functions

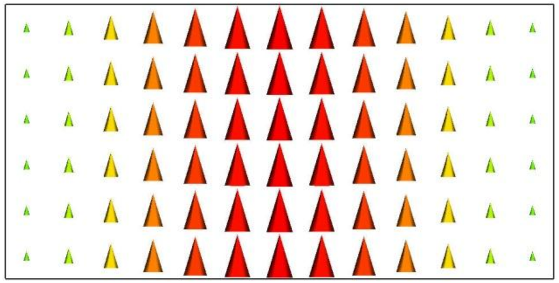


Neumann functions are often also denoted as $Y_m(r)$.

Electromagnetic waves

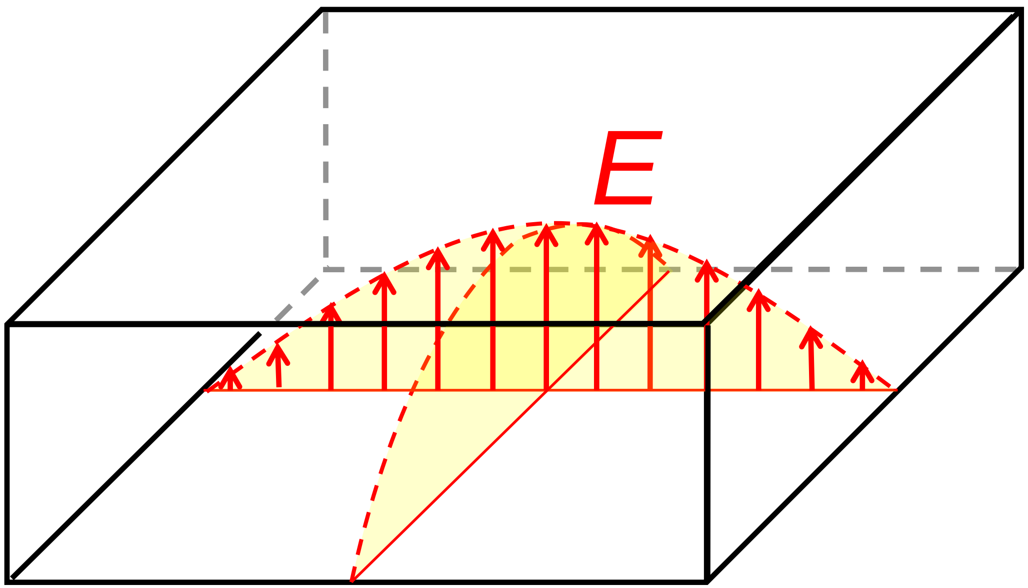
- Propagation of electromagnetic waves inside empty metallic channels is possible: there exist solutions of Maxwell's equations describing waves
- These waves are called waveguide modes
- There exist two types of waves,
  - Transverse electric (TE) modes:
    → the electric field has only transverse components
  - Transverse magnetic (TM) modes:
    → the magnetic field has only transverse components
- Propagate at above a characteristic cut-off frequency

- In a rectangular waveguide, the first mode that can propagate is the TE_{10} mode. The condition for propagation is that half of a wavelength can "fit" into the cross-section => cut-off wavelength **λ_c = 2a**
- The modes are named according to the number of field maxima they have along each dimension. The E field of the TE_{10} mode for instance has 1 maximum along x and 0 maxima along the y axis.
- For circular waveguides, the maxima are counted in the radial and azimuthal direction



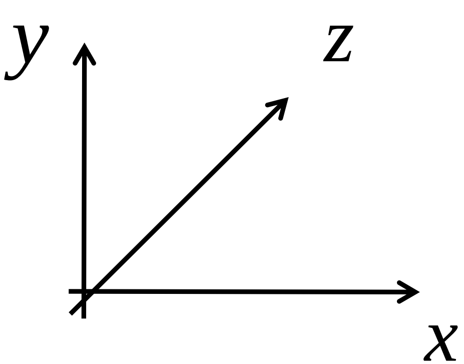
E field of the fundamental TE_{10} mode

Mode Indices in Resonators (1)



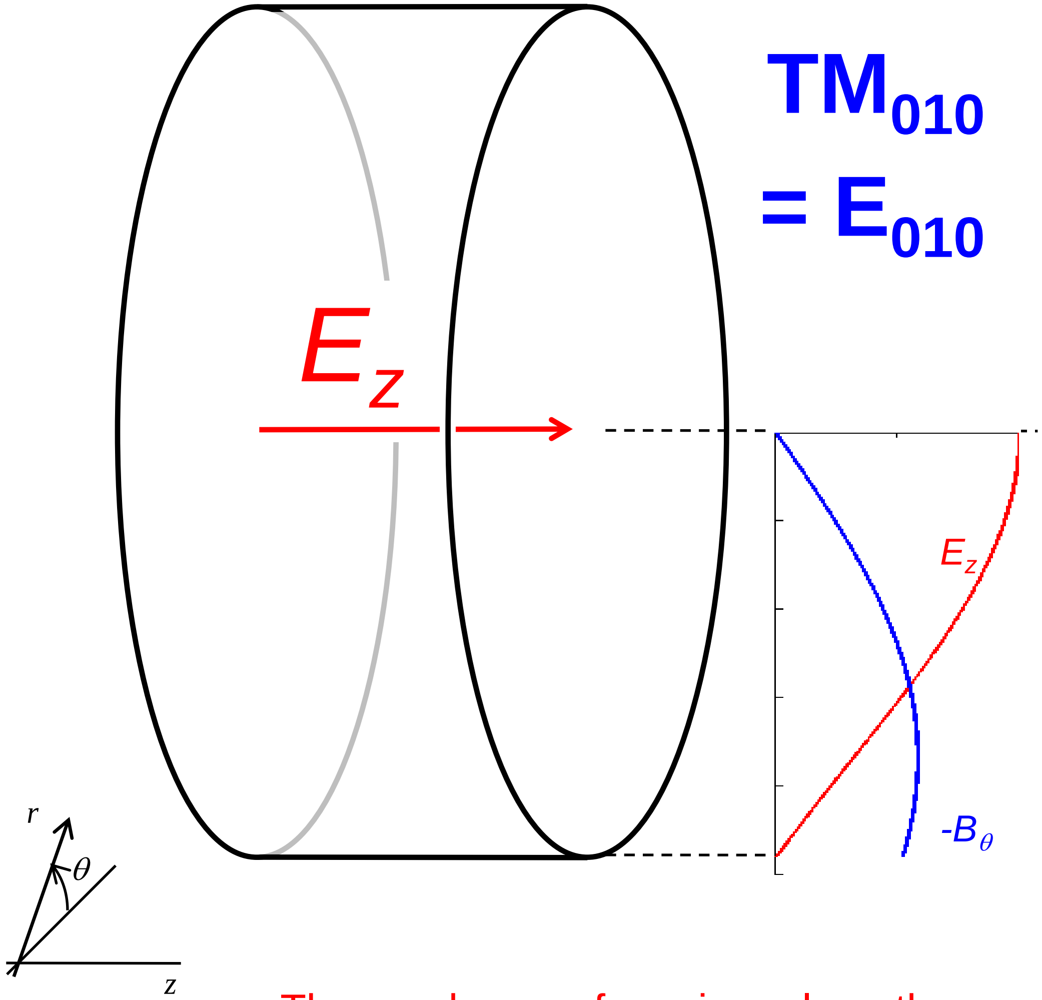
TE_{101}

$=H_{101}$



For a structure in rectangular coordinates the mode indices simply indicate the number of half waves (standing waves) along the respective axis. Here we have one maximum along the x-axis, no maximum in vertical dimension (y-axis), and one maximum along the z-axis. **TE_{101}** corresponds to **TE_{xyz}**

Mode Indices in Resonators (2)



TM_{010} = E_{010}

For a structure in <u>cylindrical coordinates:</u>

The first index is the order of the Bessel function or in general cylindrical function.

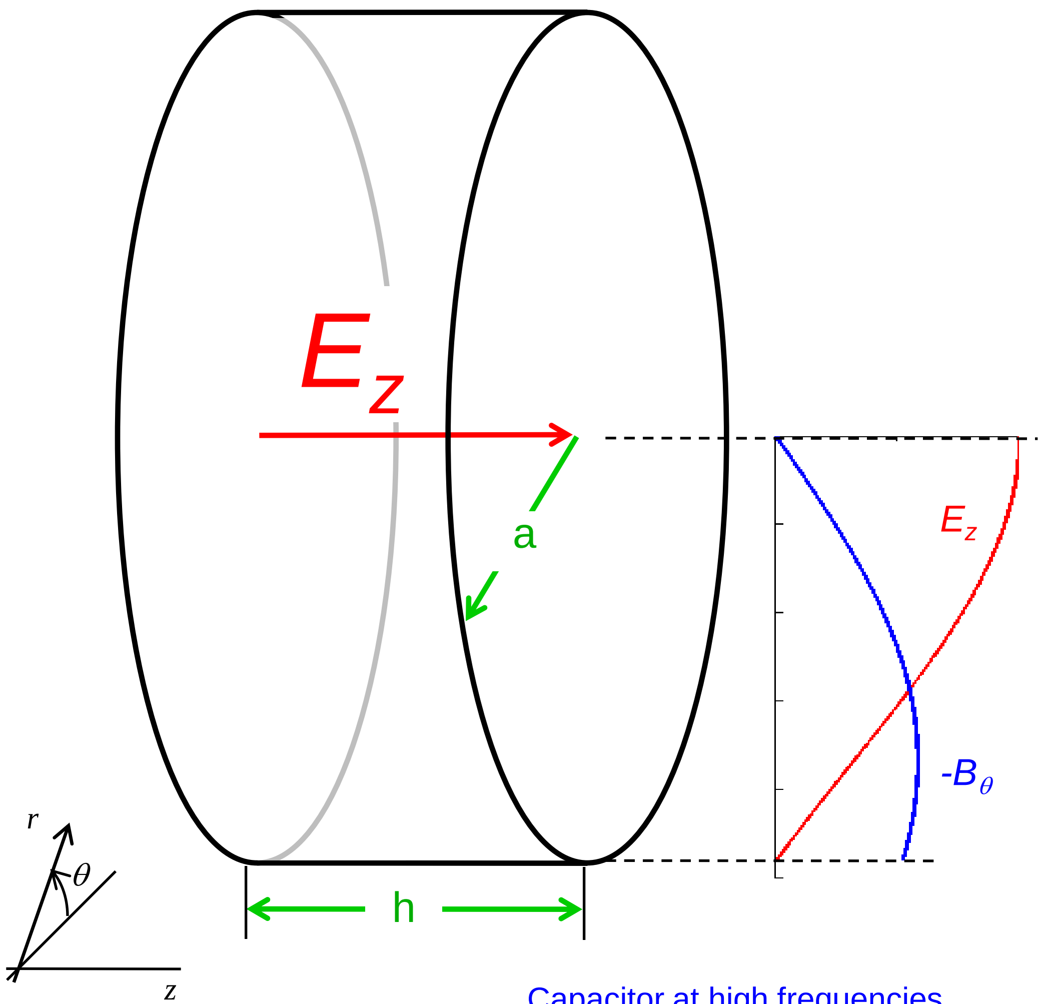
The second index indicates "the root" of the cylindrical function which is the number of zero-crossings.

The third index is the number of half waves (maxima) along the z-axis.

Hint: In an empty pillbox there will be <u>no</u> Neumann function as it has a pole in the center (conservation of energy). However we need Bessel and Neumann functions for higher order modes of coaxial structures.

The number m of maxima along the azimuth is coupled to the order of the Bessel function (see slide on theory).

Fields in a pillbox cavity



Cavity height: h
cavity radius: a

TM_{010} mode resonance
= E_{010} mode resonance for

$$a = 0.383\lambda = 1.53\lambda / 4$$

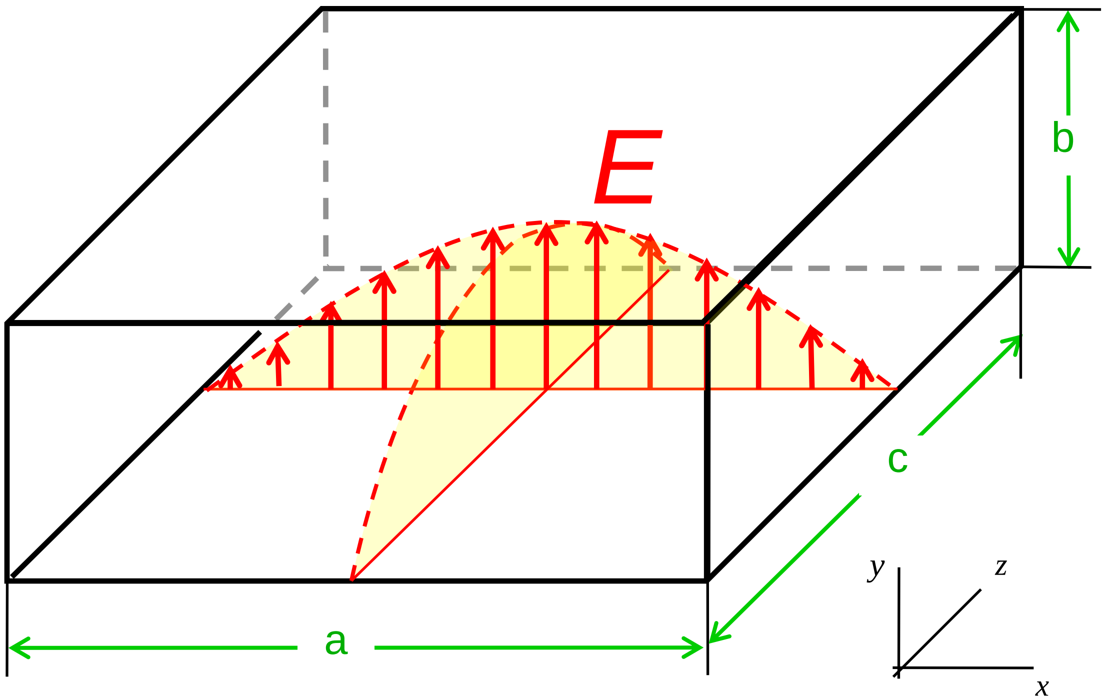
TM_{010} resonance frequency **independent of h**!!!

In the cylindrical geometry the E and H fields are proportional to Bessel functions for the radial dependency.

Capacitor at high frequencies,
The Feynman Lectures on Physics

Common cavity geometries (1)

Square prism H_{101} or TE_{101}



$$\lambda_0 = \frac{2}{\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{c}\right)^2}}$$

$$Q_{H101} = \frac{\lambda_0}{\delta} \underbrace{\frac{b}{2} \frac{\left(a^2+c^2\right)^{3/2}}{c^3(a+2b)+a^3(c+2b)}}_{\text{dimensionless form factor}}$$

$$\text{Skin depth } \delta = \sqrt{\frac{2}{\omega\sigma\mu}}$$

$$\text{with } \omega = 2\pi f$$

Comment: For a brick-shaped cavity (the structure is described in Cartesian coordinates) the E and H fields would be described by sine and cosine distributions. The mode indices indicate the number of half waves along the x-,y-, and z-axis, respectively.

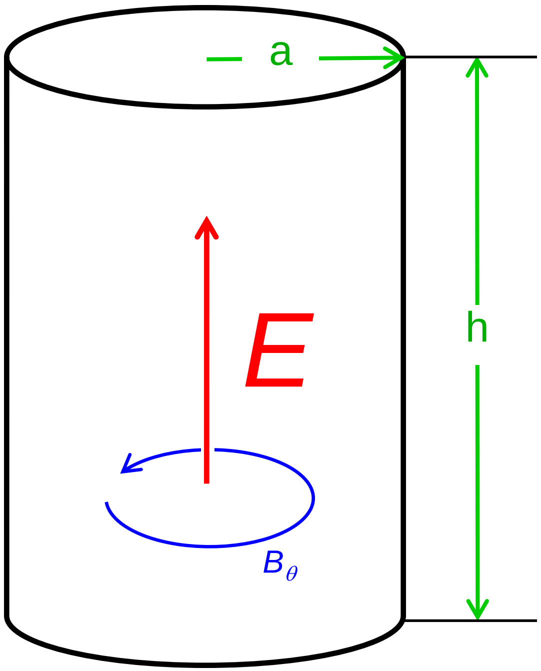
this simplifies in the case a=c:

$$\lambda_0 = \sqrt{2}a$$

$$Q = \frac{1}{\delta}\frac{ab}{a+2b}$$

Common cavity geometries (2)

Circular cylinder: E_{010}, = TM_{010}



$$\lambda_0 = 2.61a$$

$$Q = \left(0.383\frac{\lambda_0}{\delta}\right)\left[1+\left(0.383\frac{\lambda_0}{h}\right)\right]^{-1}$$

$$= 0.383\lambda_0/\delta\left[1+\frac{a}{h}\right]^{-1} = \frac{a}{\delta}\left[1+\frac{a}{h}\right]^{-1}$$

$R/Q \approx 185\,h/a$ for not too big ratios of h/a[1]

Note: h denotes the **full** height of the cavity

In some cases and also in certain numerical codes, h stands for the half height

1: formula uses Linac definition and includes time transit factor

R/Q for cavities

The full formula for calculating the R/Q value of a cavity is

$$\frac{R}{Q} = \frac{4\eta}{\chi_{01}^{3}\pi J_1^{2}(\chi_{01})} \frac{\sin^{2}(\frac{\chi_{01}}{2}\frac{h}{a})}{\frac{h}{a}}$$

see lecture: RF cavities, E. Jensen, Varna CAS 2010

with

$$\eta = \sqrt{\frac{\mu_0}{\varepsilon_0}} = \sqrt{\mu_0^2 c_0^2} = 4\pi \times 10^{-7} \times 3 \times 10^{8} = 377\Omega$$

$\chi_{01} = 2.4048$ (First zero of the Bessel function of 0th order)

$$J_1(\chi_{01}) = 0.5192$$

This leads to

$$\frac{R}{Q} = 128 \frac{\sin^{2}(1.2024\frac{h}{a})}{\frac{h}{a}}$$

The sinus can be approximated by sinx = x (for small values of x) leading to

$$\frac{R}{Q} \approx 128 \frac{(1.2024\frac{h}{a})^{2}}{\frac{h}{a}} = 185\frac{h}{a}$$

Common cavity geometries (3)

Circular cylinder:

H_{011}

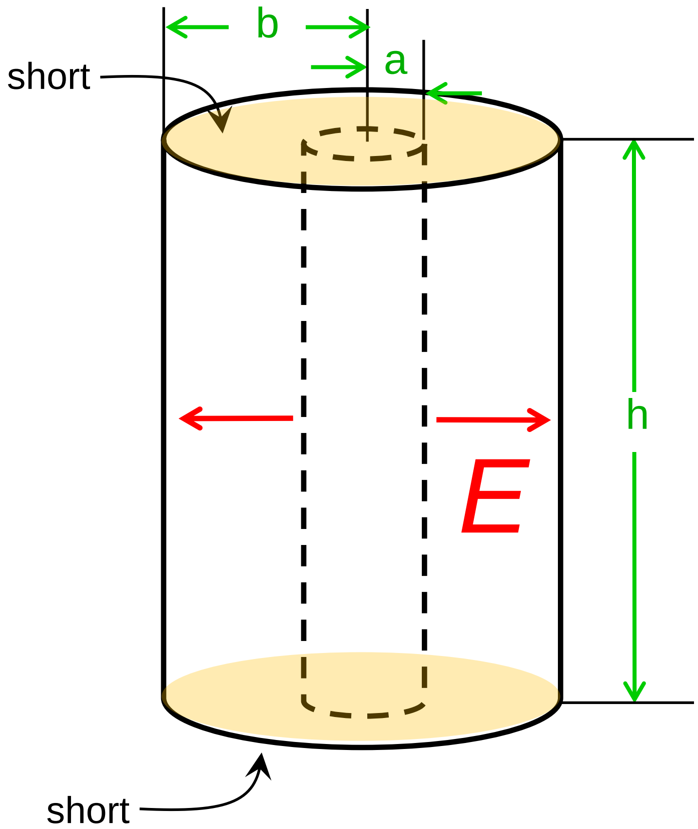
$$Q = 0.61\frac{\lambda_0}{\delta}\frac{\left[1+0.17\left(\frac{2a}{h}\right)^2\right]^{3/2}}{1+0.17\left(\frac{2a}{h}\right)^3}$$

H_{111}

$$Q = 0.206\frac{\lambda_0}{\delta}\frac{\left[1+0.73\left(\frac{2a}{h}\right)^2\right]^{3/2}}{1+0.22\left(\frac{2a}{h}\right)^2+0.51\left(\frac{2a}{h}\right)^3}$$

Common cavity geometries (4)

Coaxial TEM



$$\lambda_0 = 2h \text{ or } h = \lambda_0 / 2$$

$$Q = \frac{\lambda_0}{\delta} \frac{1}{4 + \frac{h}{b} \cdot \frac{1 + b/a}{\ln(b/a)}}$$

$$\text{Optimum } Q \text{ for } (b/a) = 3.6 \quad (Z_0 = 77\Omega)$$

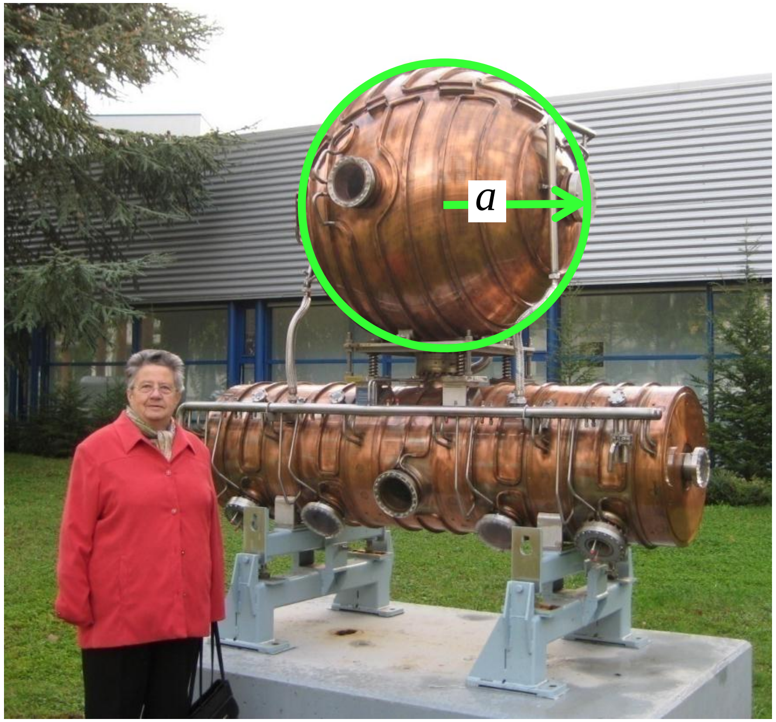
$$Q_{optimum} = \frac{\lambda_0}{\delta} \frac{1}{4 + 7.2\frac{h}{b}}$$

Coaxial line with minimum loss
→ slide TEM transmission lines (3)

Taken from S. Saad et.al.,
Microwave Engineers' Handbook, Volume I, p.180

Common cavity geometries (5)

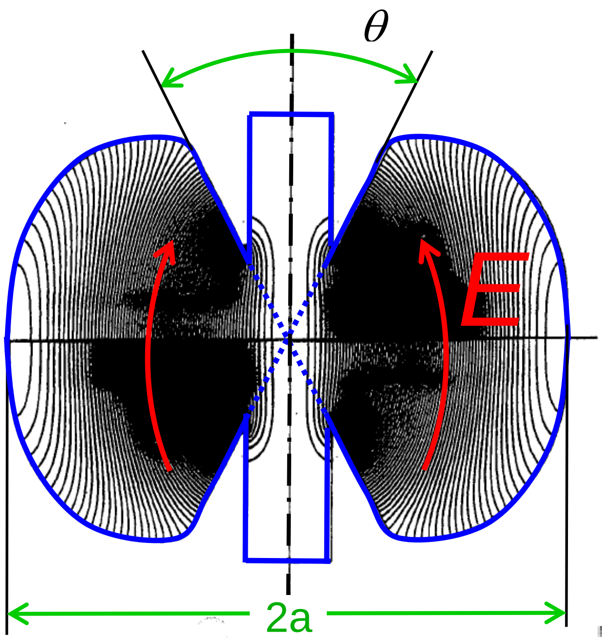
Sphere



"Energy storage in LEP"

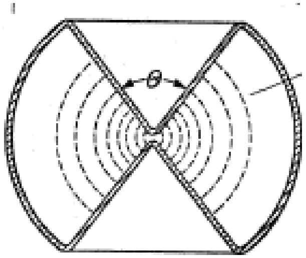
$$\lambda_0 = 2.28a$$
$$Q = 0.318(\lambda / \delta)$$

Sphere with cones



"Nose cone cavity"

$$\lambda_0 = 4a \rightarrow a = \lambda_0 / 4$$
$$\text{Optimum } Q \text{ for } \theta = 34^\circ$$
$$Q_{opt,34^\circ} = 0.1095(\lambda / \delta)$$

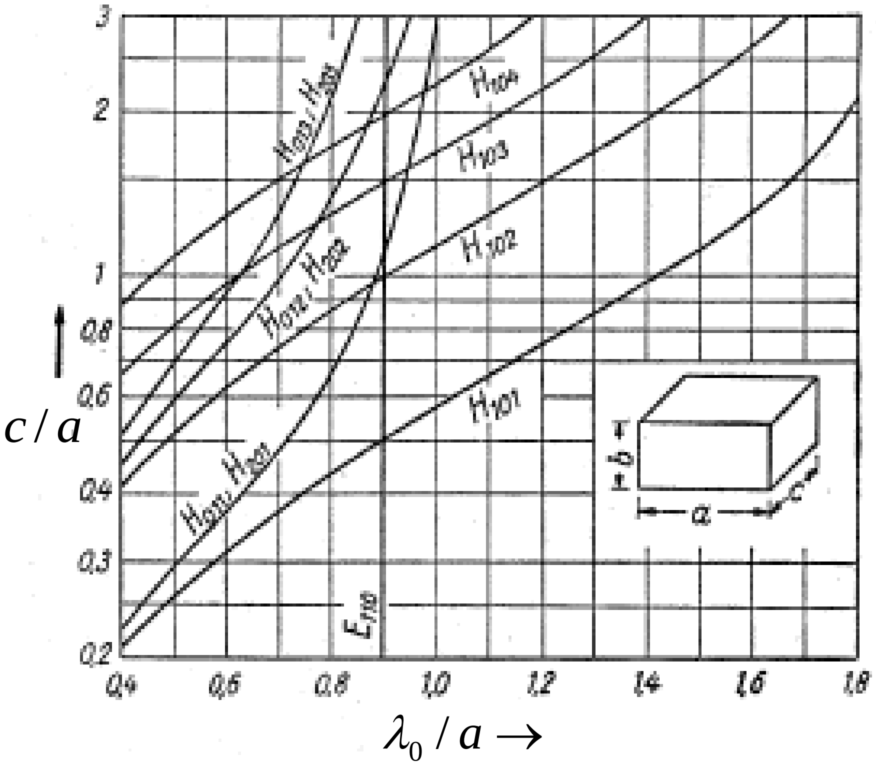


the tips of the cone don't touch

a spherical "λ/4-resonator"

Mode chart of a brick-shaped cavity

Reprinted from Meinke, H. and Gundlach, F. W., *Taschenbuch der Hochfrequenztechnik*,S.471 Erste Auflage, Springer-Verlag, Berlin (1968) and *Techniques of Microwave Measurements* by Carol G. Montgomery, 1st ed., 1947; by permission, McGraw-Hill Book Co., N. Y.



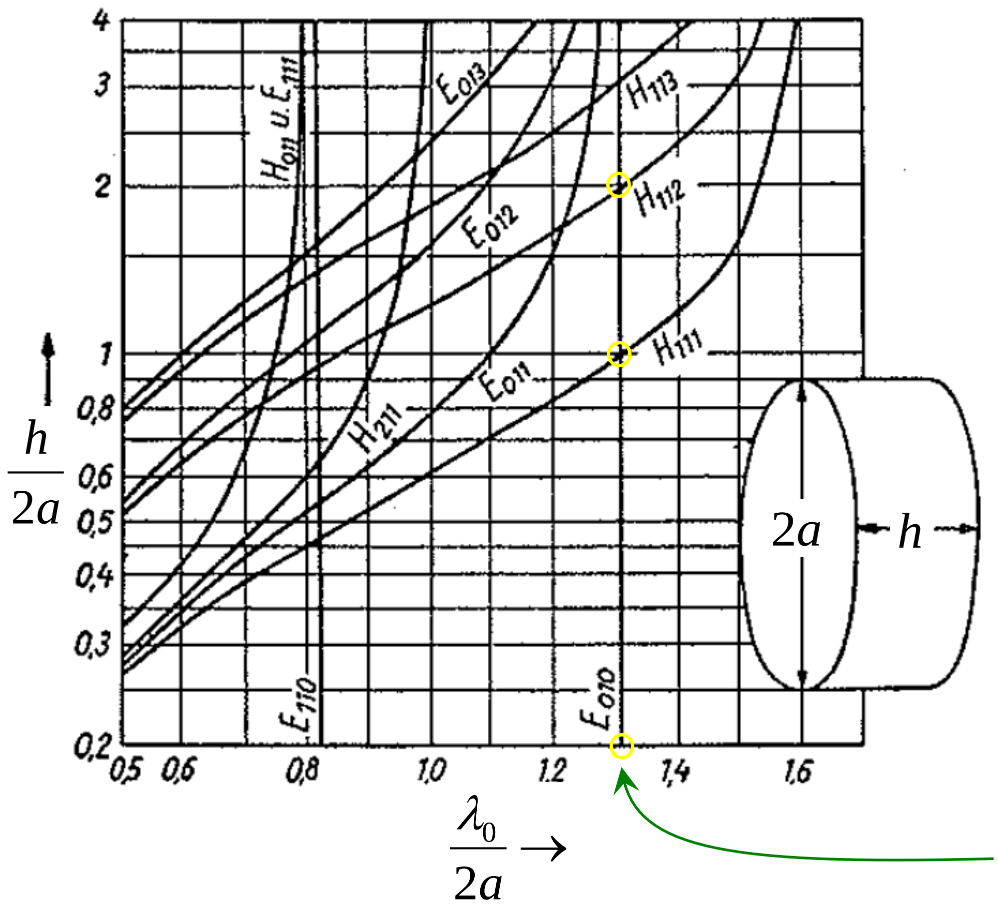
The resonant wavelength of the **H_{mnp}** resonance calculates as

$$\lambda_0 = \frac{2}{\sqrt{\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{c}\right)^2}}$$

for a E_{mn} or a H_{mn} wave with p half waves along the c-direction.

Mode chart of a Pillbox cavity – Version 1

Reprinted from Meinke, H. and Gundlach, F. W., *Taschenbuch der Hochfrequenztechnik*,S.471 Erste Auflage, Springer-Verlag, Berlin (1968) and *Techniques of Microwave Measurements* by Carol G. Montgomery, 1st ed., 1947; by permission, McGraw-Hill Book Co., N. Y.



Cylindrical cavity with **radius a, height = h** and **resonant wavelength λ_0.**
H stands for TE and
E for TM modes.

Example:
E_{010}: $\lambda_0 \approx 2.6a$
H_{111}: $h \approx 2a$
H_{112}: $h \approx 4a$

Mode chart of a Pillbox cavity – Version 2

Reprinted from Meinke, H. and Gundlach, F. W., *Taschenbuch der Hochfrequenztechnik*,S.471
Erste Auflage, Springer-Verlag, Berlin (1968) and
Techniques of Microwave Measurements by Carol G. Montgomery,
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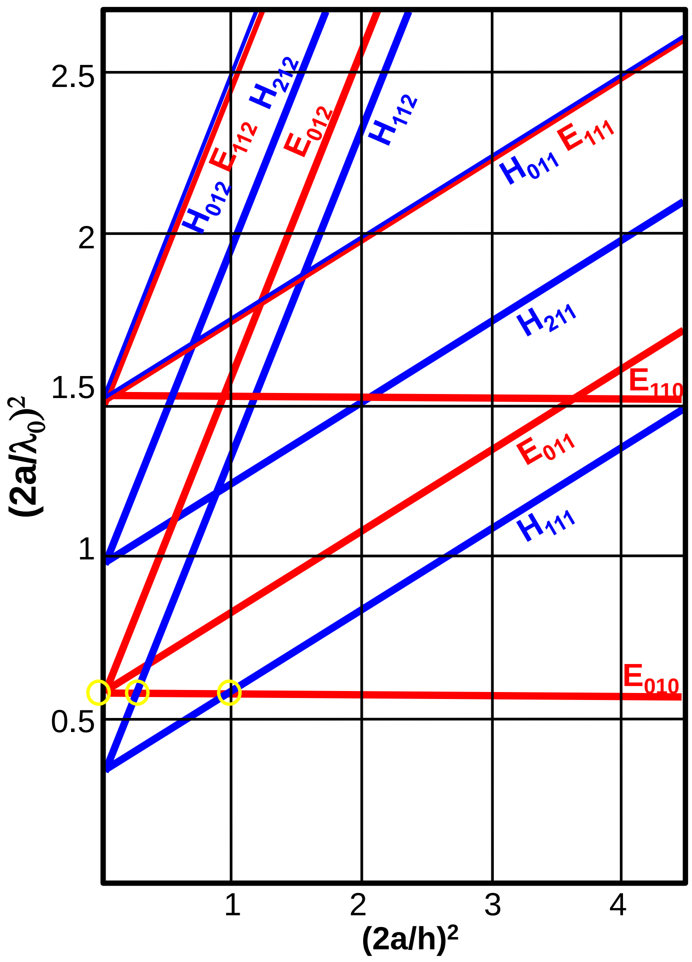
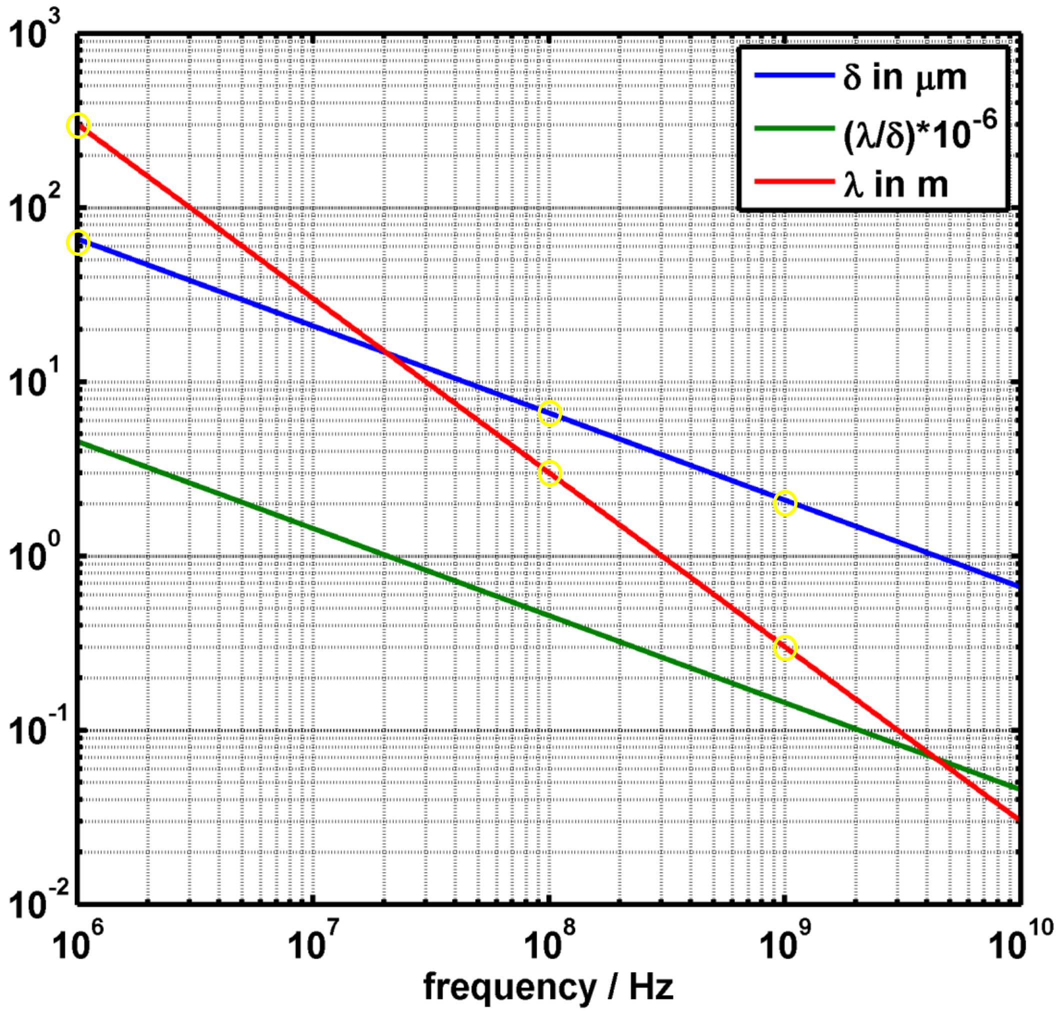


Cylindrical cavity with **radius a, height = h** and **resonant wavelength λ_0.**
H stands for TE and
E for TM modes.

Example:
E_{010}: $(2a/\lambda_0)^2 \approx 0.6$ → $\lambda_0 \approx 2.6a$
H_{111}: $h \approx 2a$
H_{112}: $h \approx 4a$

Skin-effect and scaling laws for copper

Reprinted from Meinke, H. and Gundlach, F. W., *Taschenbuch der Hochfrequenztechnik*, Dritte Auflage, Springer-Verlag, Berlin (1968) and *Techniques of Microwave Measurements* by Carol G. Montgomery, 1st ed., 1947; by permission, McGraw-Hill Book Co., N. Y.



Plotted are the wavelength λ in [m], the skin depth δ in [μm] and the ratio $(\lambda/\delta) * 10^6$ for copper.
Conductivity of copper: $\sigma = 58 * 10^6$ S/m

Skin-effect graph, plot for copper

$$\delta = \sqrt{\frac{1}{\pi f \mu \sigma}}$$

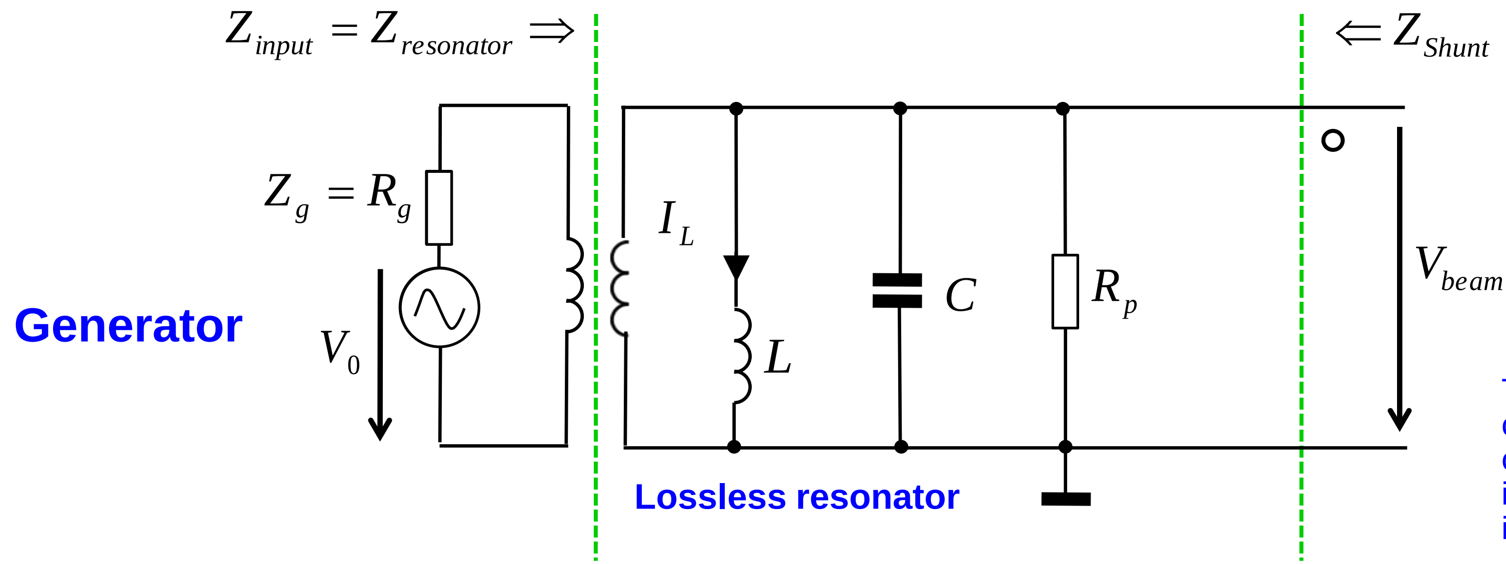
$$\lambda = \frac{c}{f}$$

Examples:

f	λ	δ
1 GHz	0.3 m	2 μm
100 MHz	3 m	6.6 μm
1 MHz	300 m	66 μm
50 Hz	6 000 km	9.3 mm

$\mu_0 = 4\pi \cdot 10^{-7}$ A/N^2
$\varepsilon_0 = 8.854187 \cdot 10^{-12}$ F/m
$c = 299\,792\,458$ m/s $\approx 3 \cdot 10^8$ m/s

Equivalent circuit (1)



The generator may deliver power to R_p and to the beam, but also the beam can deliver power to R_p and Rg!

The beam is usually considered as a current source with infinite source impedance.

$$R_p = \text{resistor representing the losses of the parallel RLC equivalent circuit}$$

We have Resonance condition, when $\omega L = \dfrac{1}{\omega C}$

→ Resonance frequency: $\omega_{res} = 2\pi f_{res} = \dfrac{1}{\sqrt{LC}} \Rightarrow \boxed{f_{res} = \dfrac{1}{2\pi}\dfrac{1}{\sqrt{LC}}}$

Equivalent circuit (2)

- Characteristic impedance "R upon Q"

(R/Q) is independent of Q and a pure geometry factor for any cavity or resonator! This formula assumes a HOMOGENEOUS field in the capacitor !

$$X = \frac{R}{Q} = \omega_{res} L = \frac{1}{\omega_{res} C} = \sqrt{L/C}$$

- Stored energy at resonance

$$W = \frac{CV^2}{2} = \frac{LI_L^2}{2}$$

- Dissipated power

$$P = \frac{V^2}{2R}$$

- Q-factor

$$Q = \frac{R}{X} = \frac{\omega_{res} W}{P}$$

W ... stored energy

P ... dissipated power

- Shunt impedance (circuit definition)

$$R = \frac{V^2}{2P}$$

- Tuning sensitivity

$$\frac{\Delta f}{f} = -\frac{1}{2}\frac{\Delta C}{C}$$

- Coupling parameter (shunt impedance over generator or feeder impedance Z)

$$k^2 = \frac{R}{R_{input}}$$

The Quality Factor (1)

- The quality (Q) factor of a resonant circuit is defined as the ratio of the stored energy W over the energy dissipated P in one cycle.

$$Q = \frac{\omega_{res} W}{P}$$

- The Q factor can be given as
  - Q_0: Unloaded Q factor of the unperturbed system, e.g. a closed cavity
  - Q_L: Loaded Q factor with measurement circuits etc connected
  - Q_{ext}: External Q factor of the measurement circuits etc

- These Q factors are related by

$$\frac{1}{Q_L} = \frac{1}{Q_0} + \frac{1}{Q_{ext}}$$

The Quality Factor (2)

- Q as defined in a Circuit Theory Textbook:

$$Q = \frac{\omega_{res} L}{R}$$

- Q as defined in a Field Theory Textbook:

$$Q = 2\pi \frac{\text{energy stored in the resonator}}{\text{energy dissipated per cycle}}$$

- Q as defined in an optoelectronics Textbook:

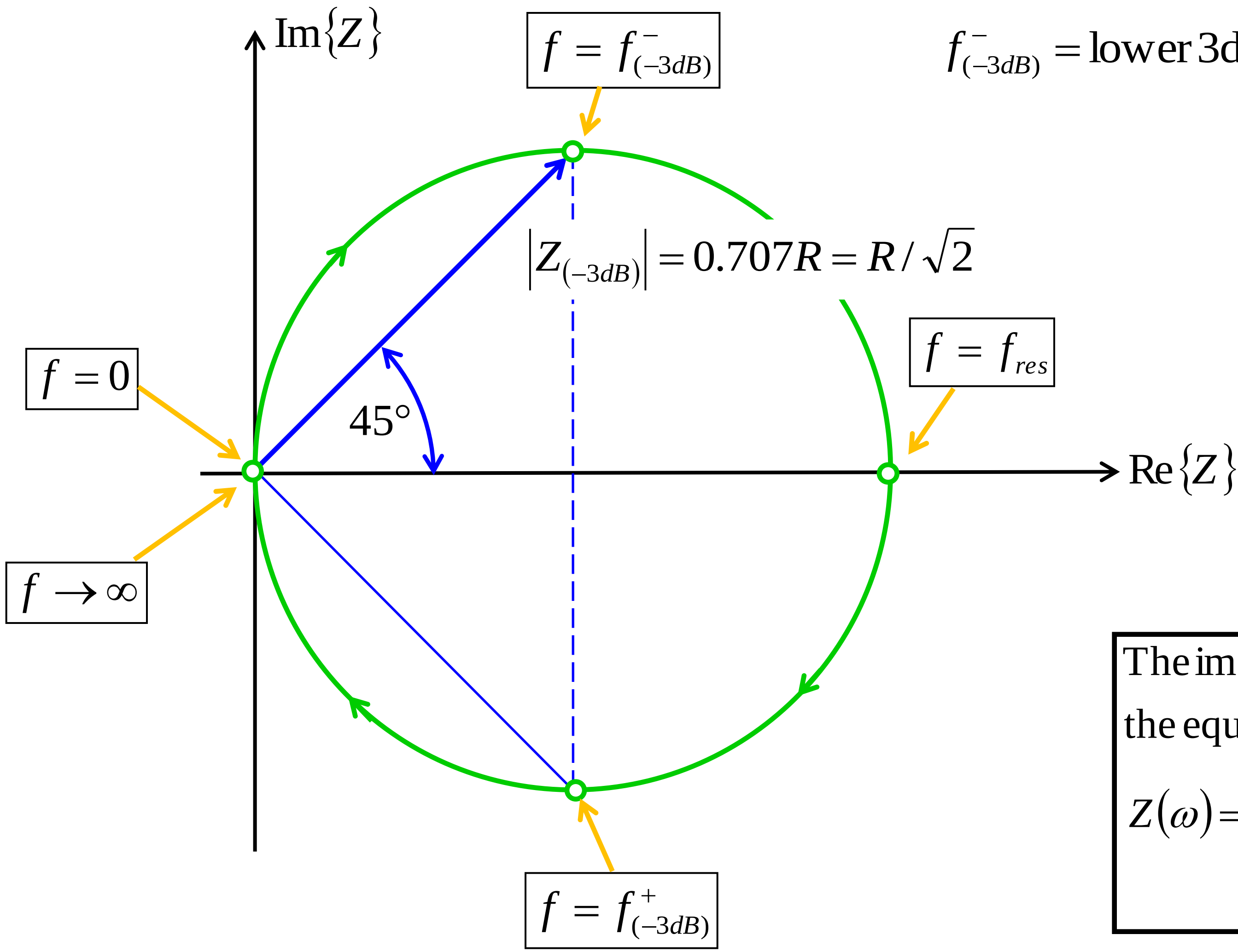
$$Q = \frac{\nu_0}{\nu_{1/2}}$$

$$\nu_0 = \text{the resonant frequency}$$

$$\nu_{1/2} = \text{"full-width at half power maximum" (FWHM)}$$

Reprinted from Madhu S. Gupta, Educator's Corner, *IEEE MicroWave Magazine*, Volume 11, Number 1, p. 48-59, February 2010.
See also: G. Dome, CAS (CERN Accelerator School), Oxford, June 1992

Input Impedance: Z-plane



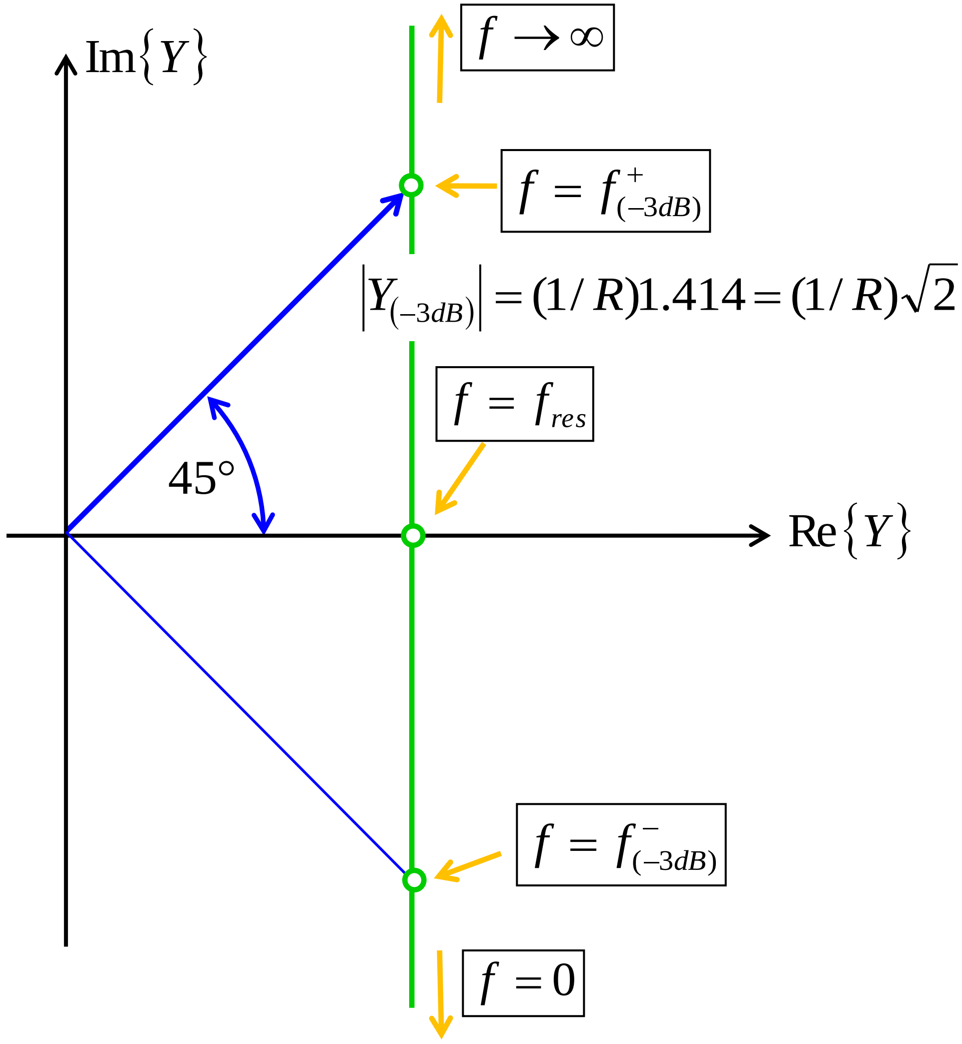
$f^-_{(-3dB)} = \text{lower 3dB point}$

The impedance Z for the equivalent circuit is :

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

$f^+_{(-3dB)} = \text{upper 3dB point}$

Input Admittance: Y-plane



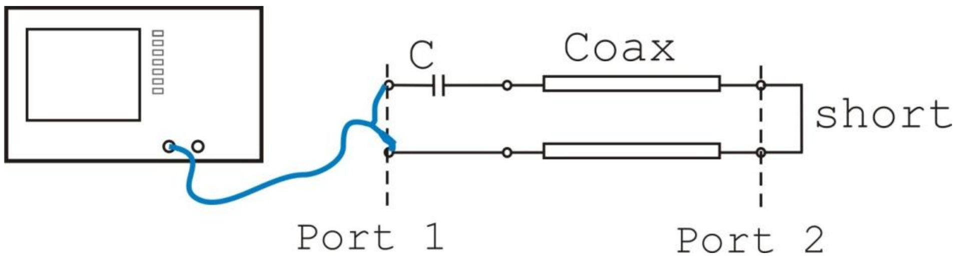
Evaluating the admittance Y for the equivalent circuit we get

$$Y = \frac{1}{Z} = \frac{1}{R} + j\omega C + \frac{1}{j\omega L}$$

$$= \frac{1}{R} + j(\omega C - \frac{1}{\omega L})$$

$$= \frac{1}{R} + j\frac{1}{R/Q}(\frac{f}{f_{res}} - \frac{f_{res}}{f})$$

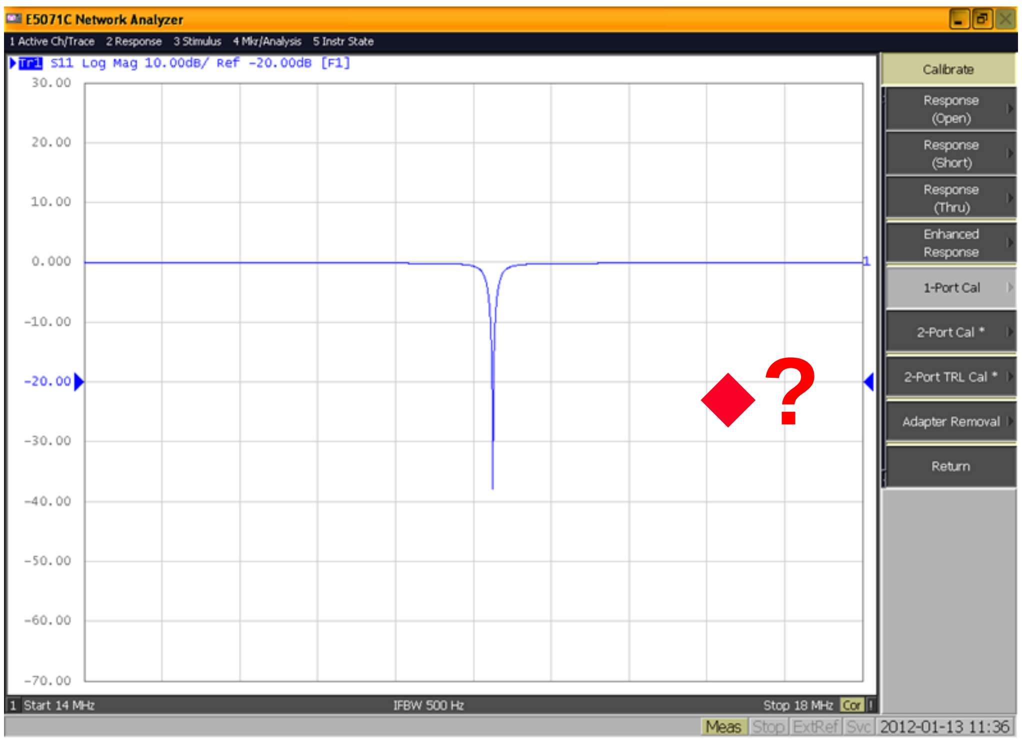
Example: Measurement of Q with VNA in Reflection



- ◆**But how?**
- ◆**This is the recipe:**

→ **Get the resonance frequency and read out the 3dB-points**

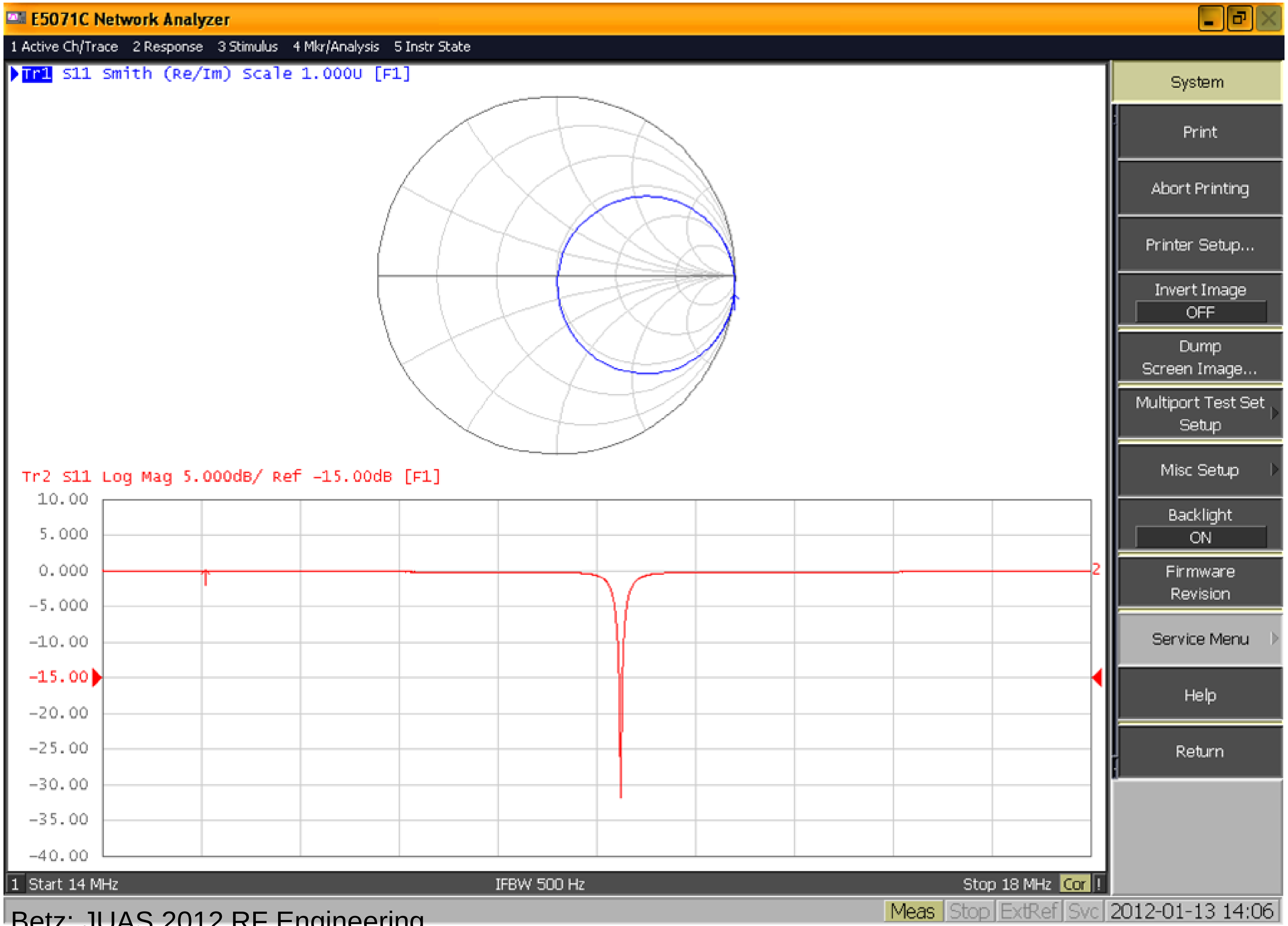
→ **Calculate $Q = f_{res}/\Delta f$.**



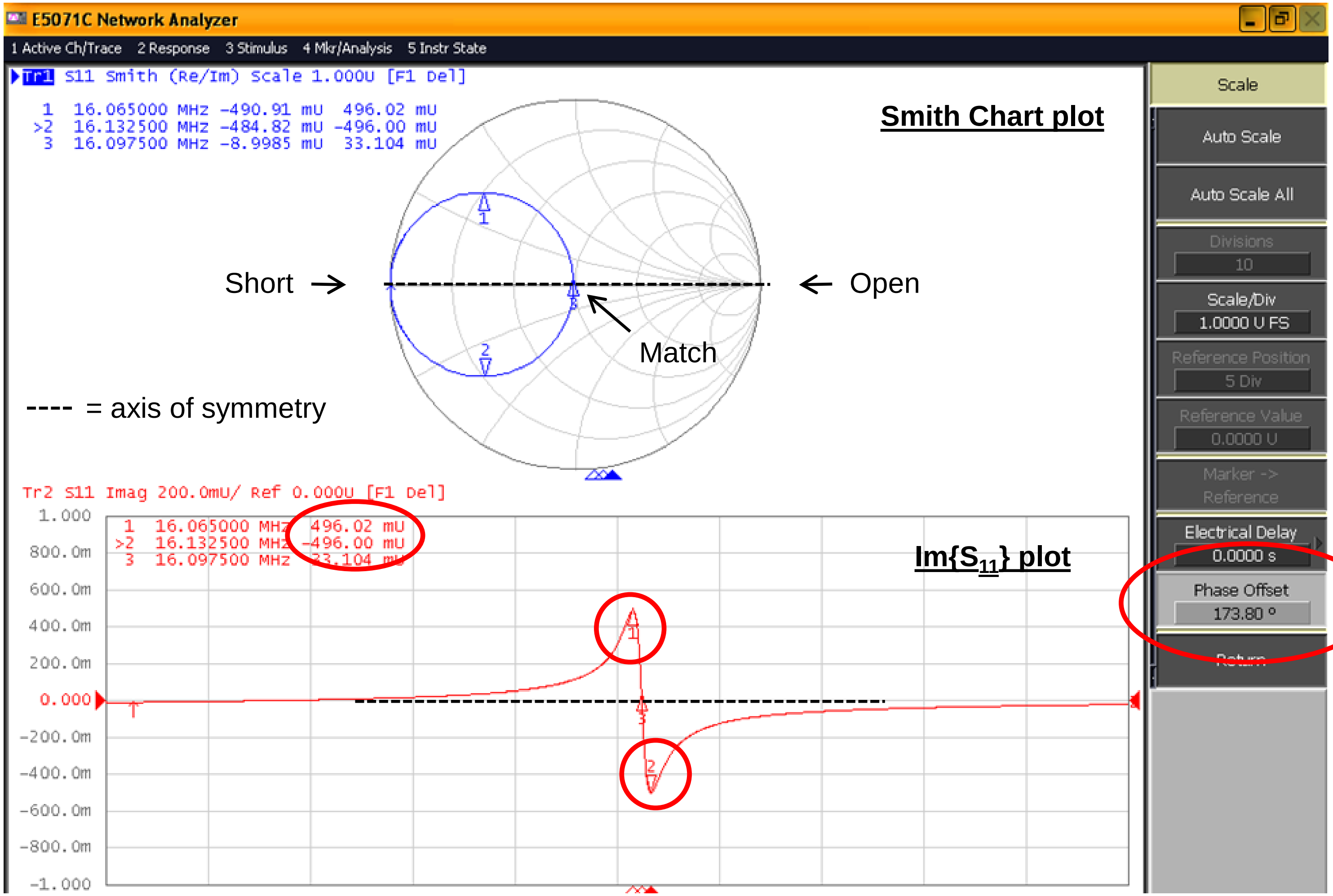
◆**Ooops? Not so straightforward?**

◆This is "our" recipe: Determine Q in the Smith-chart!

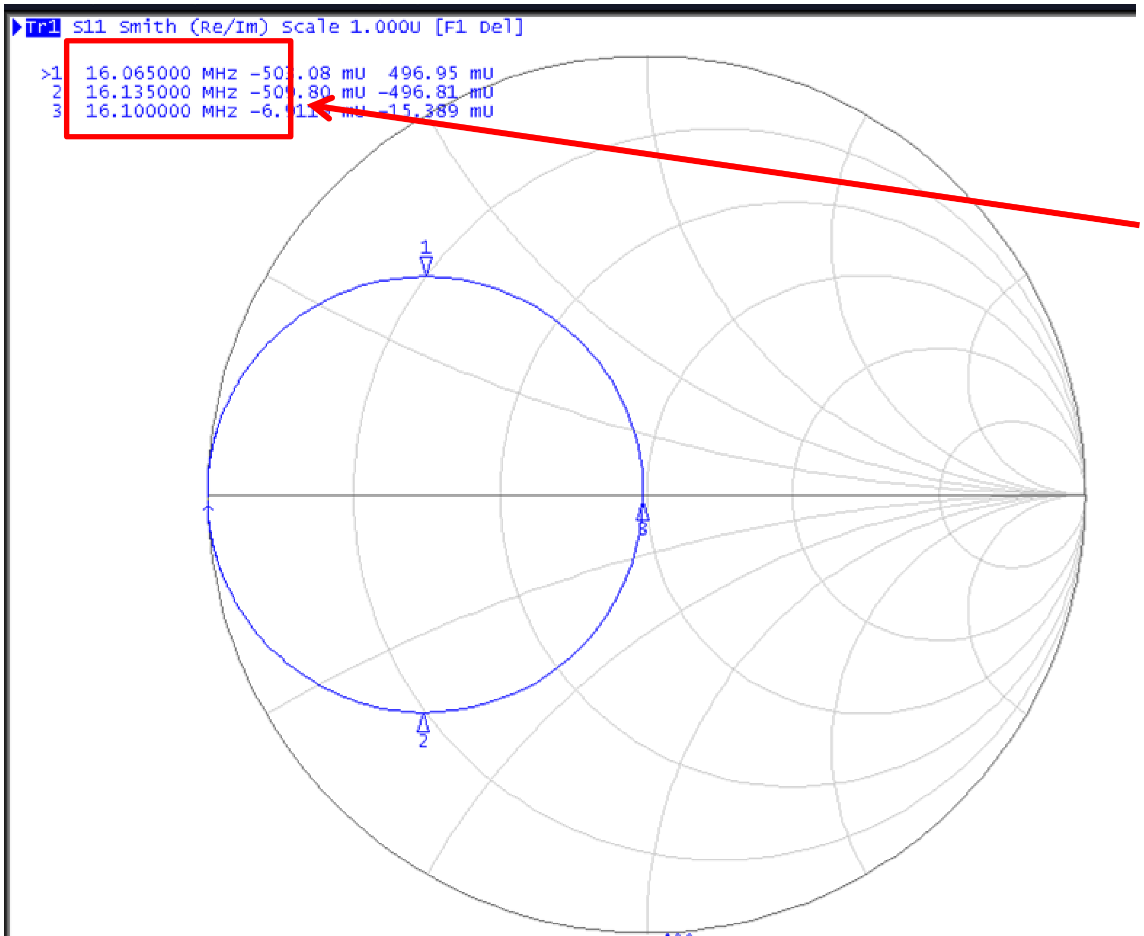
1. Put your network analyzer in Smith-chart mode.



2. Move your graph in the Smith-chart to the so-called "**detuned short position**".
3. For this, you display the **imaginary part** of S11 and change the **phase offset** so that the graph is **symmetric to the abscissa**.
   Hint: Put Markers on the plot to make sure that your graph is symmetric



3. Use **markers** in the Smith-chart to read out the frequencies at points **(1)** and **(2)**, in the upper and lower halves of the circle, this is the **minimum and maximum of Im{S_{11}}**.
4. Calculate the difference in frequency Δf, this is the 3 dB bandwidth of the loaded cavity.
5. Read-out the resonant frequency f_{res} at point **(3)**
4. Now your formulae will give you the loaded Q: $Q_L = f_{res}/\Delta f$.



Here:

Q_L= 16.1 MHz / 70 kHz

Q_L = 230

Note:

to determine the unloaded Q_0, the condition for placing the markers in Step 3 is:
Im{S_{11}} = Re{S_{11}}

All other steps stay the same.

3 dB bandwidth

In the Z-plane (= impedance) |Z| reduces to 0.707 to the value at resonance.

The real part of Z becomes 50% of the real part of that at resonance.

The phase deviates +- 45 degrees from the phase at resonance.

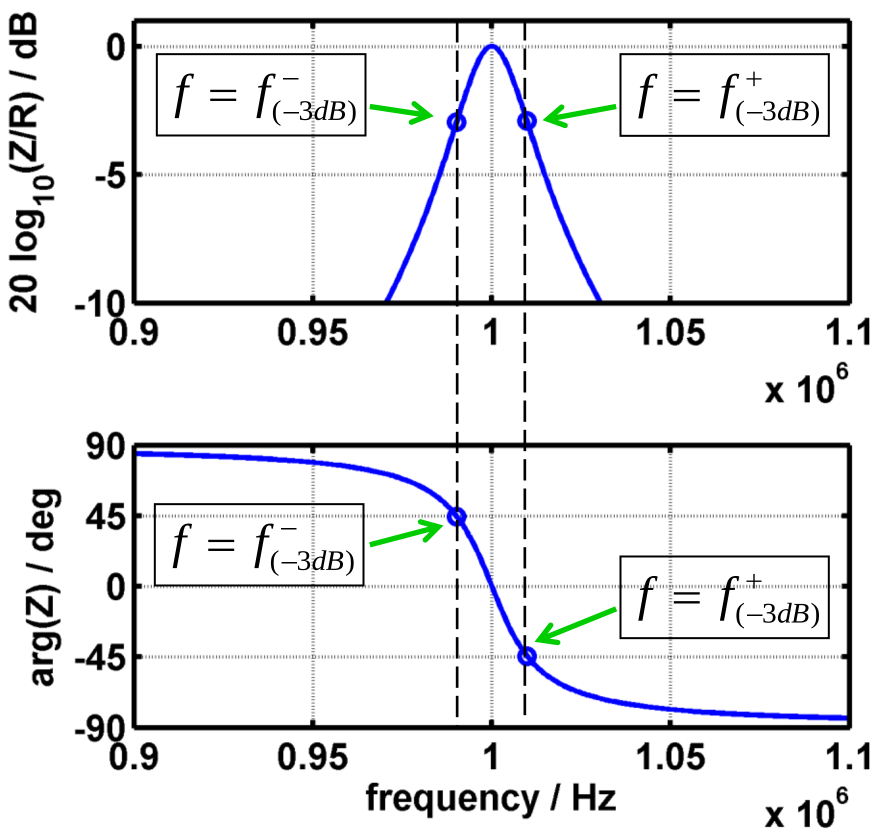
0.707 in voltage = unit voltage – 3dB (decibel)

0.707 in voltage = 50% in power since power ~ V^2

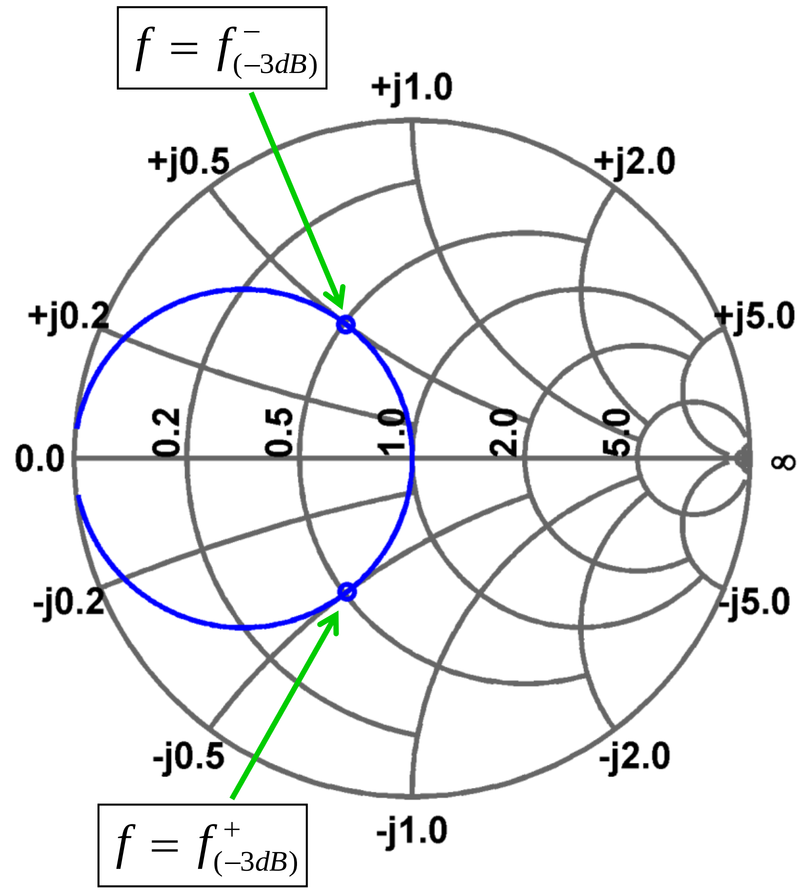
The Q factor of a resonance peak or dip can be calculated from the center frequency f_{res} and the 3 dB bandwidth $\Delta f = f^{+}_{(-3dB)} - f^{-}_{(-3dB)}$ as $Q = f_{res} / \Delta f$.

Simulated data

Cavity response vs. frequency
Top: logarithmic magnitude
Bottom: phase



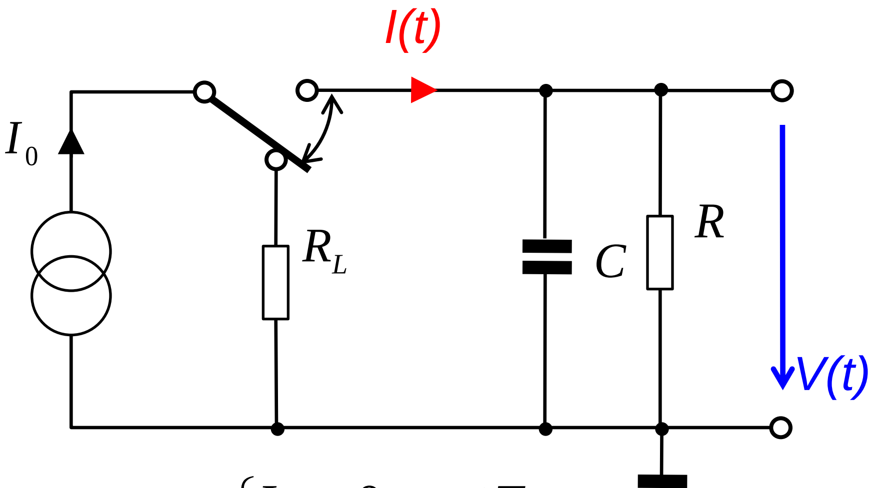
Data represented in Smith Chart



Shown is the model of a Ferrite loaded cavity:
R=200 kΩ, Q=50, f_{res}=1 MHz

Decibels and Smith Chart are discussed in detail in Part II.

Transients on an RC-Element (1)

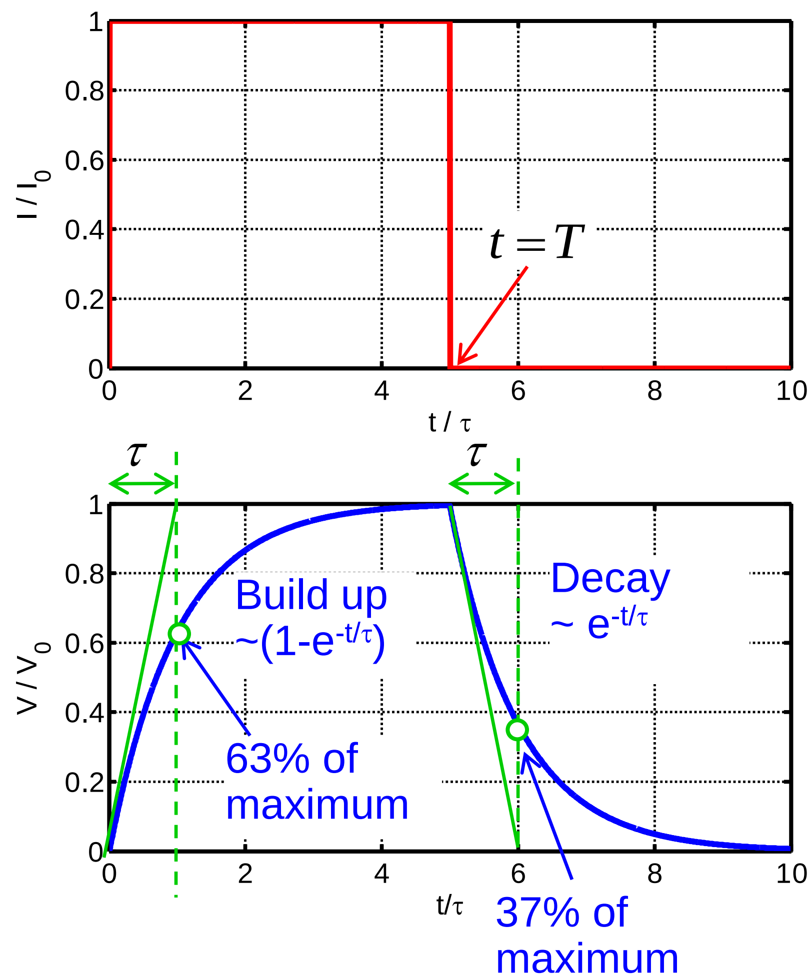


$$I(t) = \begin{cases} I_0 & 0 < t \le T \\ 0 & \text{otherwise} \end{cases}$$

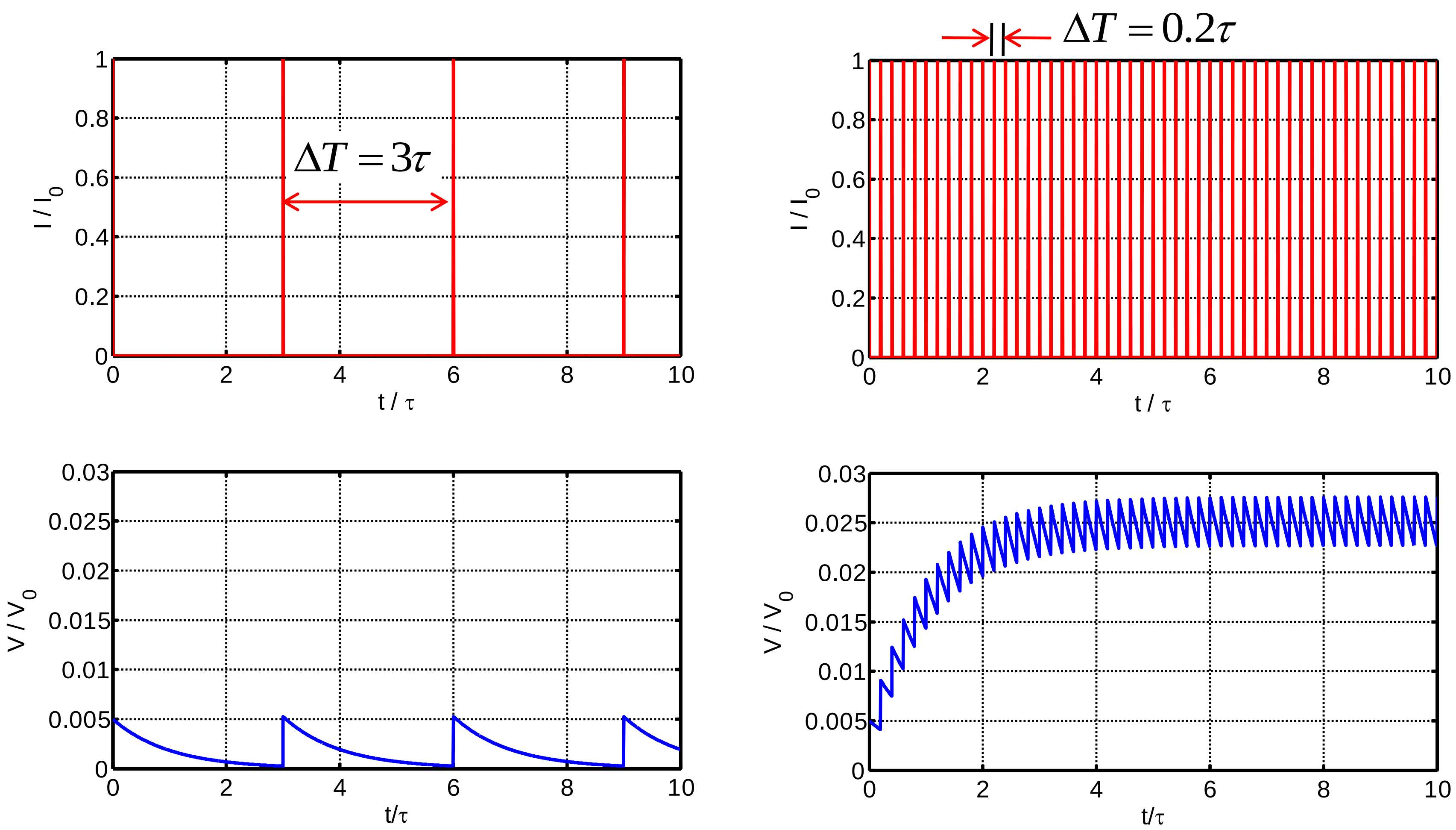
A voltage source would not work here! Explain why.

$\tau = RC \ldots$ time constant

$V_0 = I_0 R \ldots$ maximum voltage

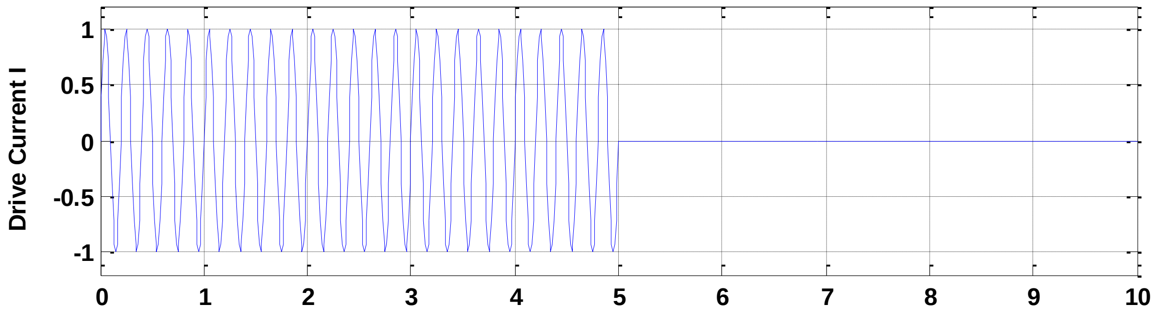


Transients on an RC-Element (2)



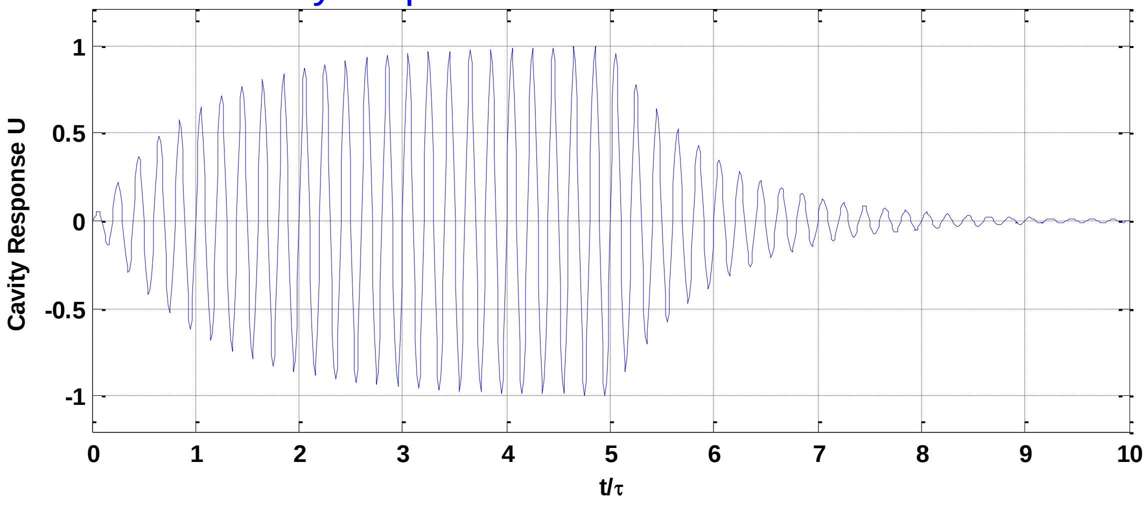
Response of a tuned cavity to sinusoidal drive current (1)

Drive current I



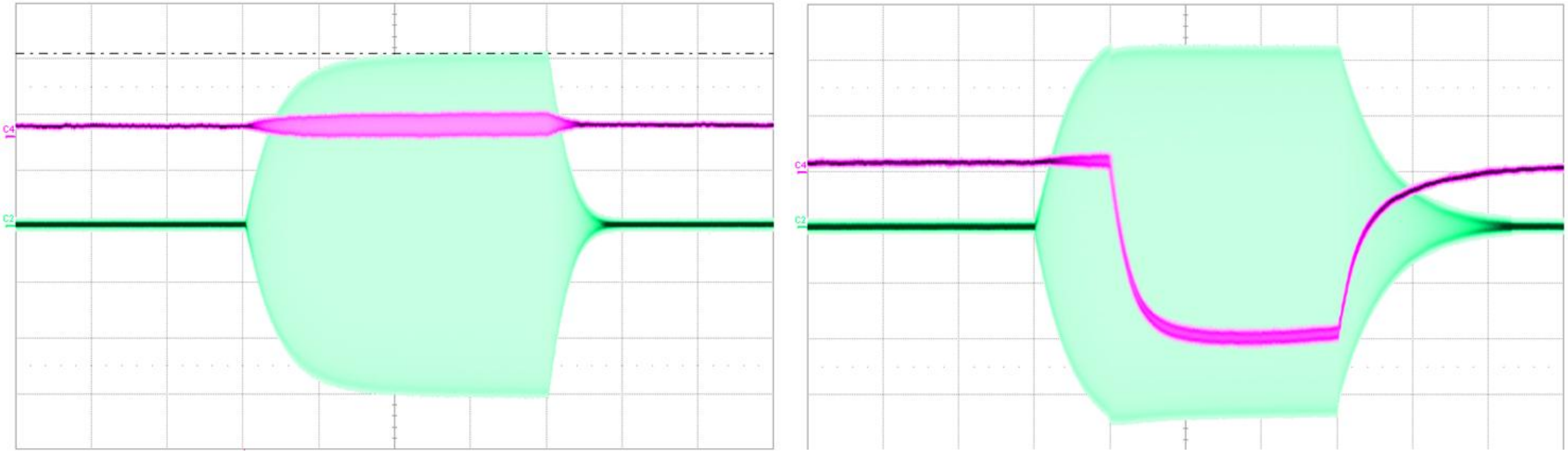
In the first moment, the cavity acts like a capacitor, as seen from the generator (compare equivalent circuit). The RF is therefore short-circuited

Cavity response U



In the stationary regime, the inductive (ωL) and capacitive reactances ($1/(\omega C)$) cancel (operation at resonance frequency!). All the power goes into the shunt impedance R => no more power reflected, at least for a matched generator...

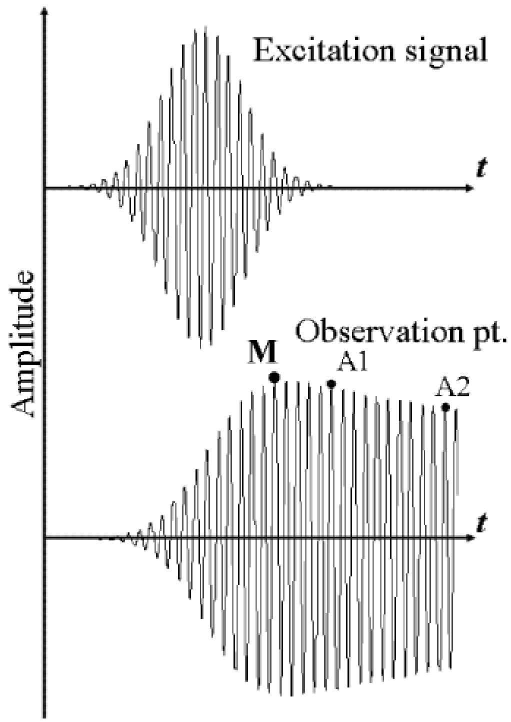
Measured time domain response of a cavity

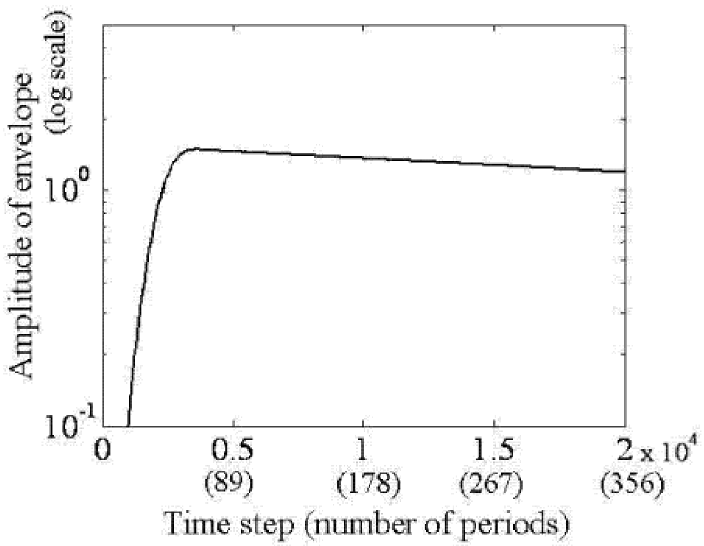


◆ Cavity E field (green trace) and electron probe signal (red trace) with and without multipacting. 200 μs RF burst duration.

see: O. Heid, T Hughes, COMPACT SOLID STATE DIRECT DRIVE RF LINAC EXPERIMENTAL PROGRAM, IPAC Kyoto, 2010

Numerically calculated response of a cavity in the time domain





$$E = E_o \sin\omega_c t \cdot e^{-\left(\frac{t-T}{T}\right)^2}$$

$$Q = \frac{\omega_r (t_2 - t_1)}{2\ln\left(\frac{A_1}{A_2}\right)},$$

see: I. Awai, Y. Zhang, T. Ishida, Unified calculation of microwave resonator parameters, IEEE 2007

Response of a tuned cavity to sinusoidal drive current (2)

Differential equation of the envelope

(shown without derivation):

$$\dot{V} = \frac{1}{2C} = (I - \frac{V}{Z}) = \frac{1}{2ZC}(IZ - V)$$

$\dot{V}, V, I, Z$ are complex quantities, evaluated at the stimulus (drive) frequency.

For a tuned cavity all quantities become real. In particular $Z = R$, therefore

$$\dot{V} = \frac{1}{2RC}(IR - V)$$

$$\rightarrow \text{time constant becomes}$$

$$\underline{\underline{\tau}} = 2RC = 2\frac{R}{Q}QC = \frac{2Q}{\omega_0} = \underline{\underline{\frac{Q}{\pi f}}} = \frac{QT}{\pi}$$

"Q over π periods"

V... envelope amplitude
C... cavity capacitance
I... drive current
Z... cavity impedance
R... real part of cavity impedance

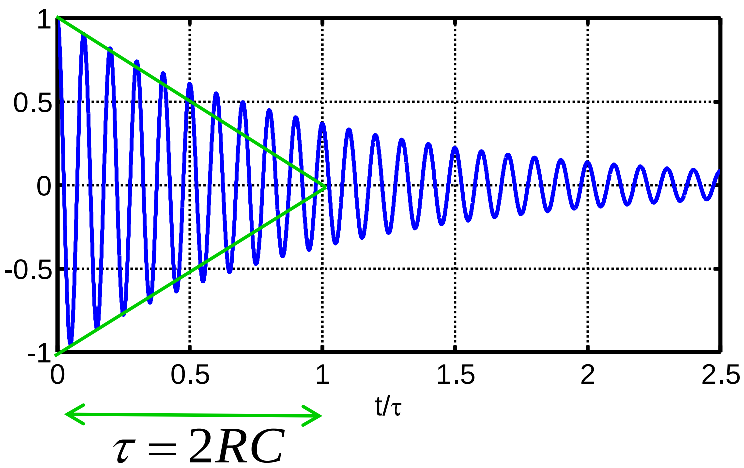
This τ value refers to the 1/e decay of the field in the cavity. Sometimes one finds τ_w referring to the energy with $2\tau_w = \tau$.

The voltage (or current) decreases to 1/e of the initial value within the time τ.

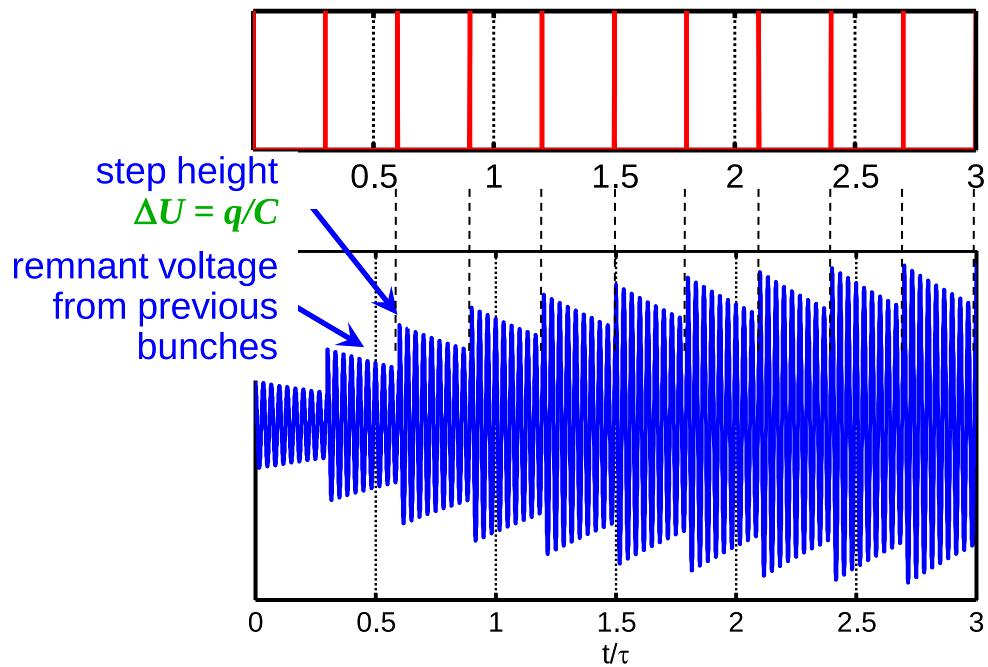
see also: H. Klein, Basic concepts I
Proceeding Oxford CAS, April 91
CERN Yellow Report 92-03, Vol. I

Beam-cavity interaction (1)

Cavity response in time domain $c(t)$ from one very short bunch



Bunched beam $b(t)$ with bunch length t_b, bunch spacing T and beam current I_0



Resulting response for bunched beam obtained by convolution of the bunch sequence with the cavity response $r(t) = b(t) \otimes c(t)$

Condition that the induced signals in the cavity add up:

cavity resonant frequency f_{res} must be an integer multiple of bunch frequency $1/T$

Beam-cavity interaction (2)

For a quantitative evaluation the worst case is considered with the induced signals adding up in phase.

Two approaches:

- Equilibrium condition: Voltage drop between two bunch passages compensated by newly induced voltage

$$U_{end} e^{-T/\tau} = U_{step} = U_{end} - \frac{q}{C} \Rightarrow \underline{U_{end} = \frac{q}{C} \frac{1}{1 - e^{-T/\tau}}}$$

- Summing up individual stimuli

$$U_{end} = \frac{q}{C}(1 + e^{-T/\tau} + e^{-2T/\tau} + \ldots) = \frac{q}{C} \frac{1}{1 - e^{-T/\tau}}$$

Approximation for $T/\tau << 1$:

$$1 - e^{-T/\tau} = 1 - (1 - T/\tau + \ldots) \approx T/\tau$$

$$\underline{U_{end}} = \frac{q}{C} \frac{1}{T/\tau} = \frac{q}{C} \frac{2RC}{T} = 2R \frac{q}{T} = \underline{2RI_0}$$

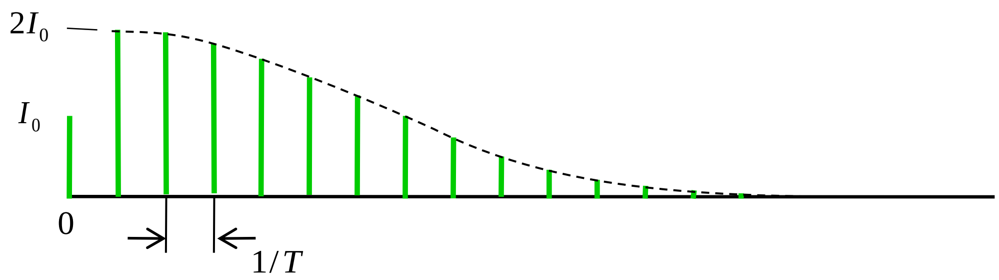
where I_0 is the mean beam current.

Beam-cavity interaction in Frequency domain

- Frequency domain

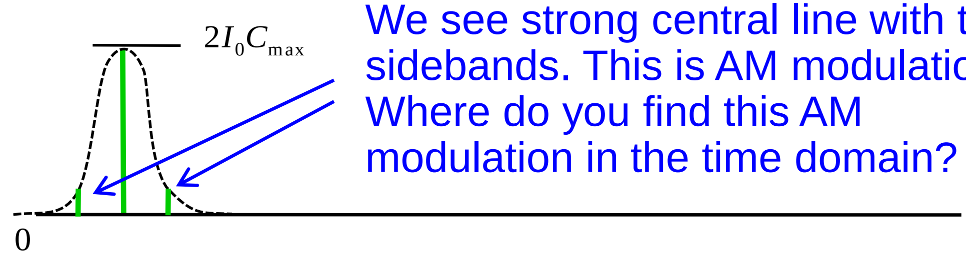
beam spectrum *B(f)*

$$B(f) = 2I_0 \frac{\sin(\pi t_b f)}{\pi t_b f}$$



cavity response *C(f)*

Resulting spectrum obtained by multiplication
*R(f) = B(f) * C(f)*



We see strong central line with two sidebands. This is AM modulation. Where do you find this AM modulation in the time domain?

Typical parameters for different cavity technologies

Cavity type	***R/Q***	***Q***	***R***
Ferrite loaded cavity (low frequency, rapid cycling)	4 kΩ	50	200 kΩ
Room temperature copper cavity (type 1 with nose cone)	192 Ω	$30 * 10^3$	5.75 MΩ
Superconducting cavity (type 2 with large iris)	50 Ω	$1 * 10^{10}$	500 GΩ

Different definitions of the shunt impedance *r*

Four different parameters
=> confusion can be maximized by using 2^4 = 16 different definitions...

Example: Pillbox cavity
r = 3.3 MΩ
L = 0.2 m
cos(φ) = 0.866
transit-time factor T = 0.756
(defined later, see transit time factor slides!)

Linac and electrical definition most often used.

Linac definition:

$$P = \frac{\hat{U}^2}{R} \text{ with the } \underline{\text{peak}} \text{ voltage } \hat{U}$$

Electrical (or circuit) definition for circular machines uses the effective voltage *U* => factor 2

$$P = \frac{\hat{U}^2}{2R}$$

Full correct shunt impedance designation	cos(φ) included	T included	L included	LINAC definition	Value
r (electrical def.)	**0**	**0**	**0**	**0**	**3.3 MΩ**
R (Linac def.)	**0**	**0**	**0**	**1**	**6.6MΩ**
r/L	0	0	1	0	16.5 MΩ/m
R/L (effective shunt impedanceZ)	0	0	1	1	33.0 MΩ/m
rT^2 (electrical def. with T)	**0**	**1**	**0**	**0**	**1.88 MΩ**
RT^2 (Linac def. with T)	**0**	**1**	**0**	**1**	**3.77 MΩ**
rT^2/L	0	1	1	0	2.86 MΩ/m
RT^2/L	0	1	1	1	5.72 MΩ/m
r cos$^{2(}$φ)	1	0	0	0	2.47 MΩ
R cos$^{2(}$φ)	1	0	0	1	4.95 MΩ
r cos$^{2(}$φ)/L	1	0	1	0	12.37 MΩ/m
R cos$^{2(}$φ)/L	1	0	1	1	24.75 *MΩ*/m
$rT^2 \cos^2(\varphi)$	1	1	0	0	1.41 MΩ
$RT^2 \cos^2(\varphi)$	1	1	0	1	2.83 MΩ
$rT^2 \cos^2(\varphi)/L$	1	1	1	0	7.07 MΩ/m
$RT^2 \cos^2(\varphi)/L$	1	1	1	1	14.14 *MΩ/m*

Electromagnetic scaling laws

A cavity of a given geometry can be scaled using three rules:

- The ratio of any cavity dimension to λ is constant. To put it another way, all cavity dimensions are inversely proportional to frequency
- Characteristic impedance R/Q = const.
- $Q * \delta / \lambda$ = const.

The skin depth δ is given by

$$\delta = \sqrt{\frac{2}{\omega\sigma\mu}}$$

with the conductivity σ ,the permeability μ, and the angular frequency $\omega=2\pi f$.

Note that it is proportional to $\frac{1}{\sqrt{f\sigma}}$

For instance, in copper ($\sigma_{copper} = 5.8*10^7$ S/m) the skin depth is ≈9 mm at 50 Hz, while it decreases to ≈ 2 μm at 1 GHz.

Scaling of a pillbox-type cavity

Starting point: SUPERFISH simulation results for a cavity of a given geometry with copper walls. Parameters: f = 3030 MHz, Q_0 = 9625 and R = 631 kΩ

Question: What are the characteristic parameters (Q, R/Q, λ) of a cavity of similar shape, that operates at a frequency of 814.5 MHz, built with steel walls?
(σ_{copper} = 58 MS/m, here we assume $\sigma_{steel} \approx$ 2 MS/m)

Answer: For the first cavity we find
Skin depth σ_1 = 1.195 μm,
Resonant wavelength $\lambda_1 = c/f_1$ = 98.97 mm,
$Q_1 * \sigma_1 / \lambda_1 = 0.1162$

For the larger steel cavity all dimensions have to be scaled by the inverse frequency ratio f_1/f_2, which gives a factor of 3030/814.5 = 3.72
=> $\underline{\lambda_2}$ = 3.72 λ_1 = 368 mm

The characteristic impedance remains unchanged.
$R_2/Q_2 = R_1/Q_1 = 632 * 10^3 / 9625 = \underline{65.56\ \Omega}$

The skin depth for steel at 814.5 MHz is σ_2 = 12.5 μm.
Using $Q_1 * \sigma_1 / \lambda_1 = Q_2 * \sigma_2 / \lambda_2$ we find $\underline{Q_2 = 3420}$

Finally, the shunt impedance gets
$\underline{R_2} = (R_1/Q_1) * Q_2 = 65.56 * 3420 = \underline{224\ k\Omega}$

Simulation Tools

- Poisson Superfish (poisson equation; poisson = fish in french)
- Microwave Studio, Mafia (Maxwell's finite integration algorithm), http://www.cst.com
- Ansoft HFSS (High frequency structure simulator), http://www.ansoft.com
- GdfidL ("Gitter drauf fertig ist die Laube" – no joke, really true!)

Simulation Techniques (1)

- Frequency domain analysis
  - CST Microwave Studio 2009, HFSS 12.0
  - Uses a tetrahedral mesh
  - Maxwell's equations solved in frequency domain for *one frequency point at a time*
  - Frequency sweeps take very long time, very powerful PC or computer cluster needed!
  - Applications: quite universal

- Time domain analysis
  - Microwave Studio
  - Space is discretized by a rectangular mesh
  - Excitation of structure with time domain pulse
  - Transformation to frequency domain by Fourier Transform => entire frequency range with only one run => fast!!!
  - Bad convergence for resonant structures, since pulse does not decay fast
  - Applications: Waveguide transitions, connectors, antennas, but no resonant structures such as cavities!!!

Simulation Techniques (2)

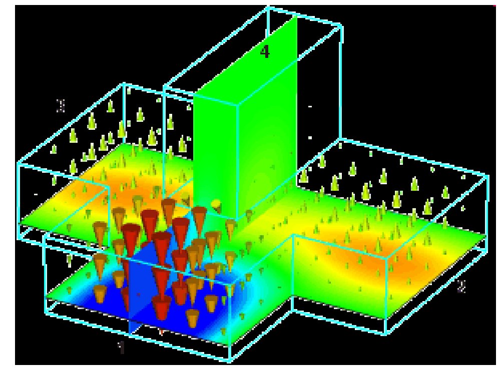
- Eigenmode analysis
  - Microwave Studio, Mafia, HFSS, Superfish …
  - Allows to calculate eigenmodes of resonant structures
  - Used for instance to determine resonant frequencies of cavities, including higher order modes

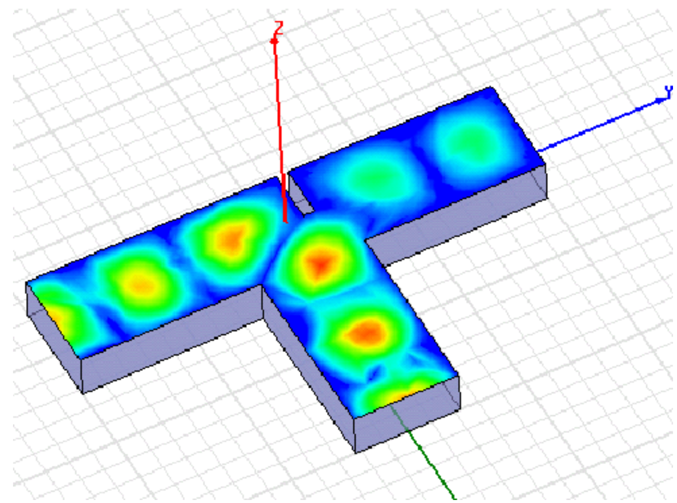
The Mesh

- Space discretized by a mesh
- Mesh width in the order of a tenth of the wavelength in the material
- Successive mesh refinement to improve precision
- Expert systems or user determine critical regions where mesh needs to be denser
- Magic T shown below: Roughly 10^5 mesh cells and a few seconds to minutes simulation time on a present-day PC

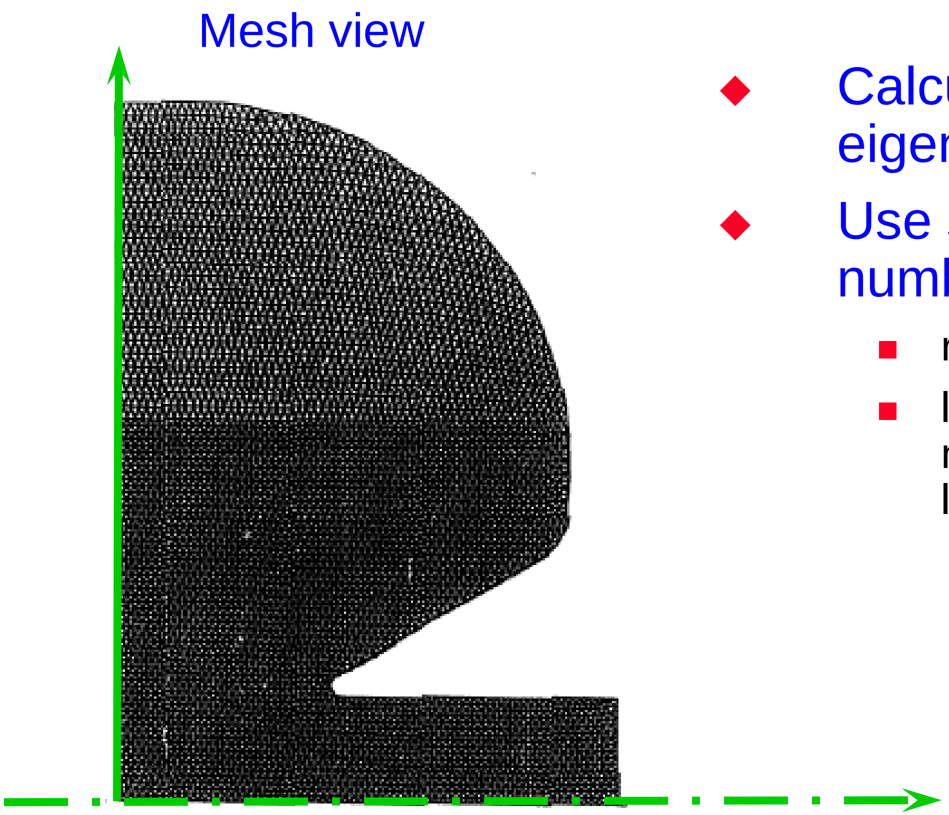
3D Simulation examples

- A Magic T with Microwave Studio 4.3
  - Arrows show the E field of the TE_{10} mode
  - Power goes in at the front port
  - How much power gets out by the other ports?
- A T-junction with HFSS 9.0
  - Junction with conducting iris
  - Magnitude of TE_{10} electric field





Superfish: 2 ½ D simulation

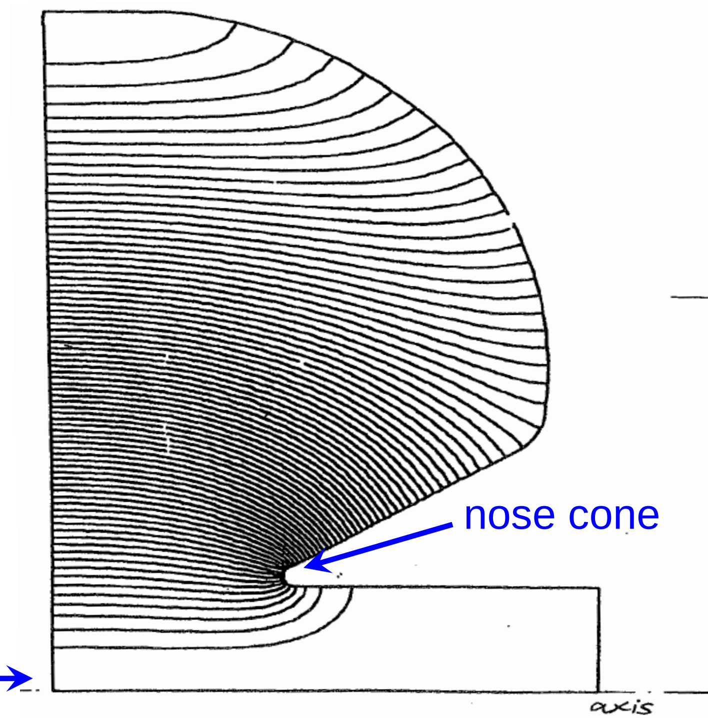


- Calculate resonant modes using eigenmode analysis
- Use symmetries to reduce the number of mesh points!!!
  - rotational symmetry around axis
  - left-right symmetry by defining metallic boundary (electric field lines perpendicular to this plane)

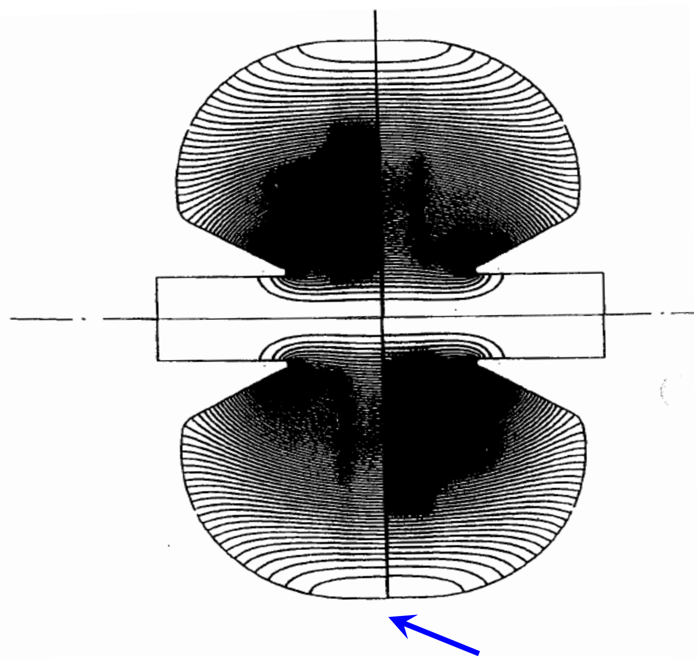
Field pattern

type-1 cavity with all symmetries exploited

entire type-1 cavity



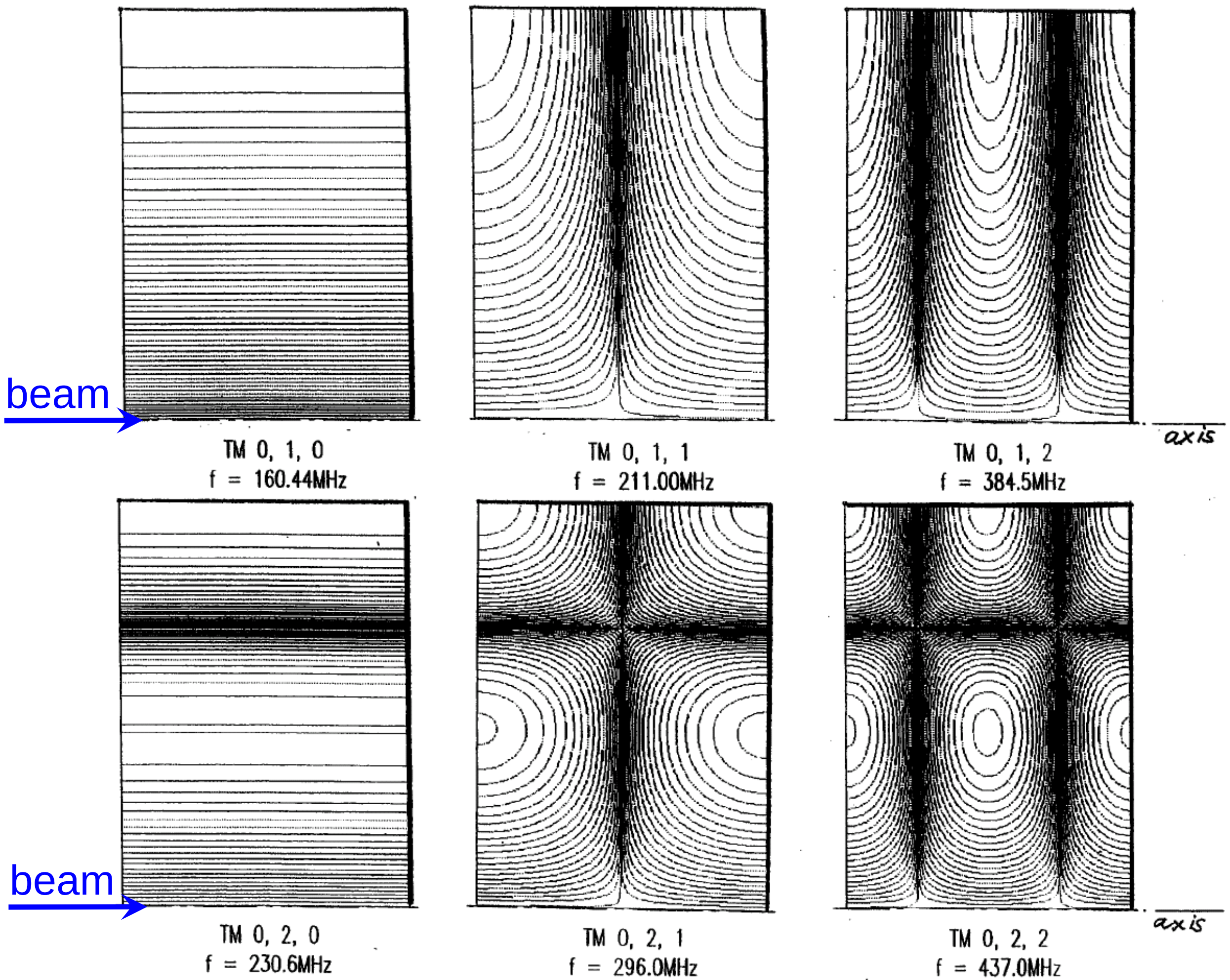
beam



equatorial conducting plane

- The electric field lines are plotted

Accelerating modes in a pillbox cavity

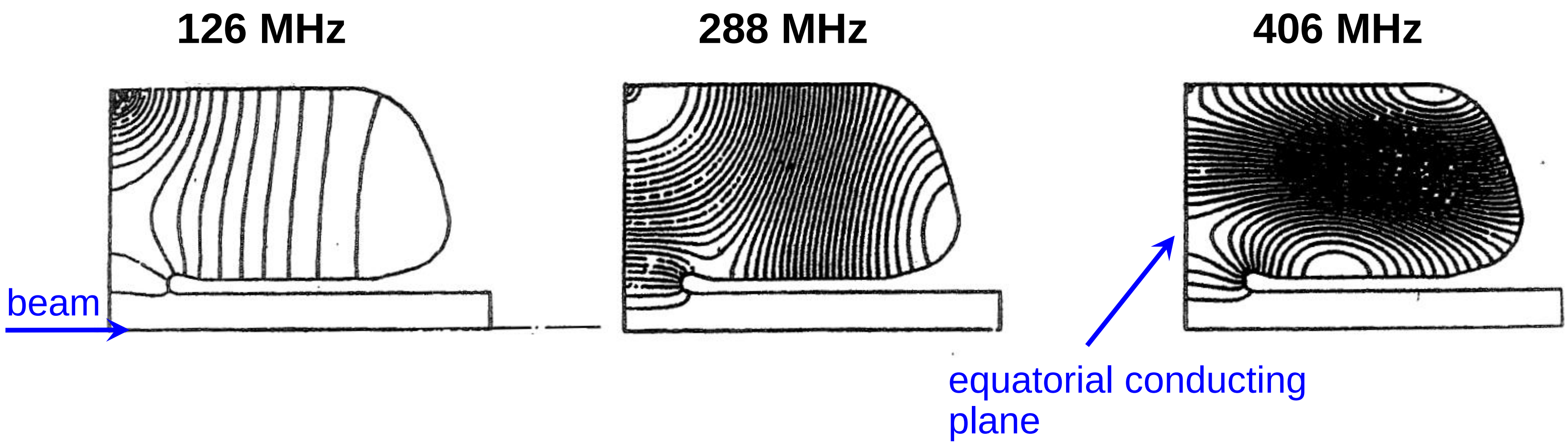


Rotationally symmetric modes of type TM_{0mn}, that is, 0 maxima over the azimuth (0 to π), m maxima in radius and n maxima along the beam axis.

Why are these modes accelerating modes?

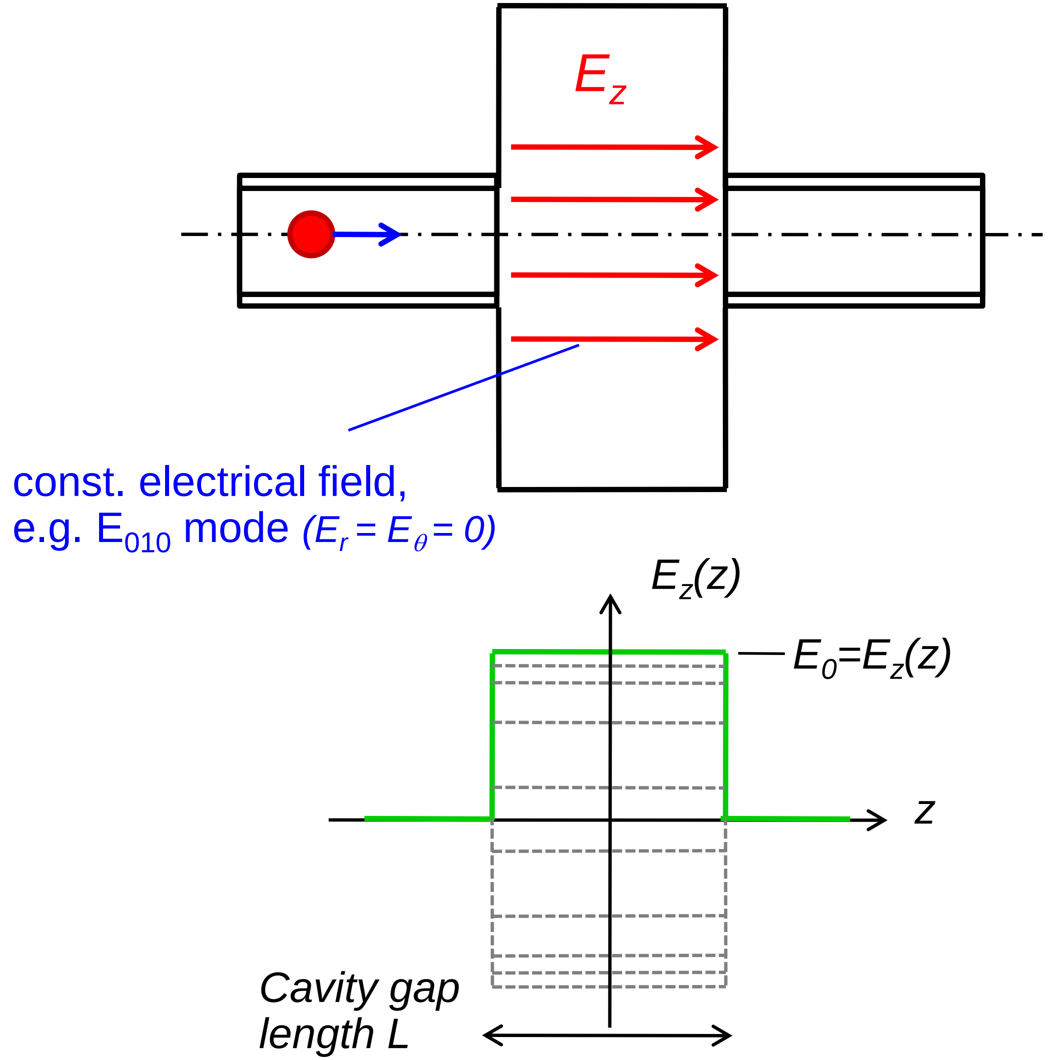
What would deflecting modes look like?

Higher order modes (HOMs)



Higher order modes in a 100-MHz cavity.
All these modes are TM type modes. This is due to the boundary condition: electric wall in equatorial plane.
references: G. Rogner, CERN report SPS/SME/Note 86-65

Transit time factor (1)



The "voltage" in a cavity along the particle trajectory (which coincides with the axis of the cavity) is given by the integral along this path for a fixed moment in time:

$$V = \int_L E_z(z)\, dz$$

But: the field in the cavity is varying in time:

$$E_z(z,t) = E_z(z)\, f(t)$$
$$= E_z(z) \cos(\omega t + \varphi)$$

Thus, the field seen by the particle is

$$V = E_0 \int_{-L/2}^{L/2} \cos(\omega t + \varphi)\, dz$$

Transit time factor (2)

The transit time factor describes the amount of the supplied RF-energy that is effectively used to accelerate the traversing particle.

$$T = \frac{V}{\hat{V}}$$

$V \ldots$ voltage seen by a particle

$\hat{V} \ldots$ reference voltage

→ relative loss in accelerating voltage

Usually, as a reference the moment of time is taken when the longitudinal field strength of the cavity is at its maximum, i.e. cos(φ)=1. A particle with infinite velocity passing through the cavity at this moment would see

$$\hat{V} = E_0 L$$

Now the particle is sampling this field with a finite velocity. This velocity is given by $v = \beta c$. The resulting transit time factor returns therefore as

$$T = \sin\left(\frac{L}{2}\frac{\omega}{\beta c}\right) \Big/ \left(\frac{L}{2}\frac{\omega}{\beta c}\right)$$

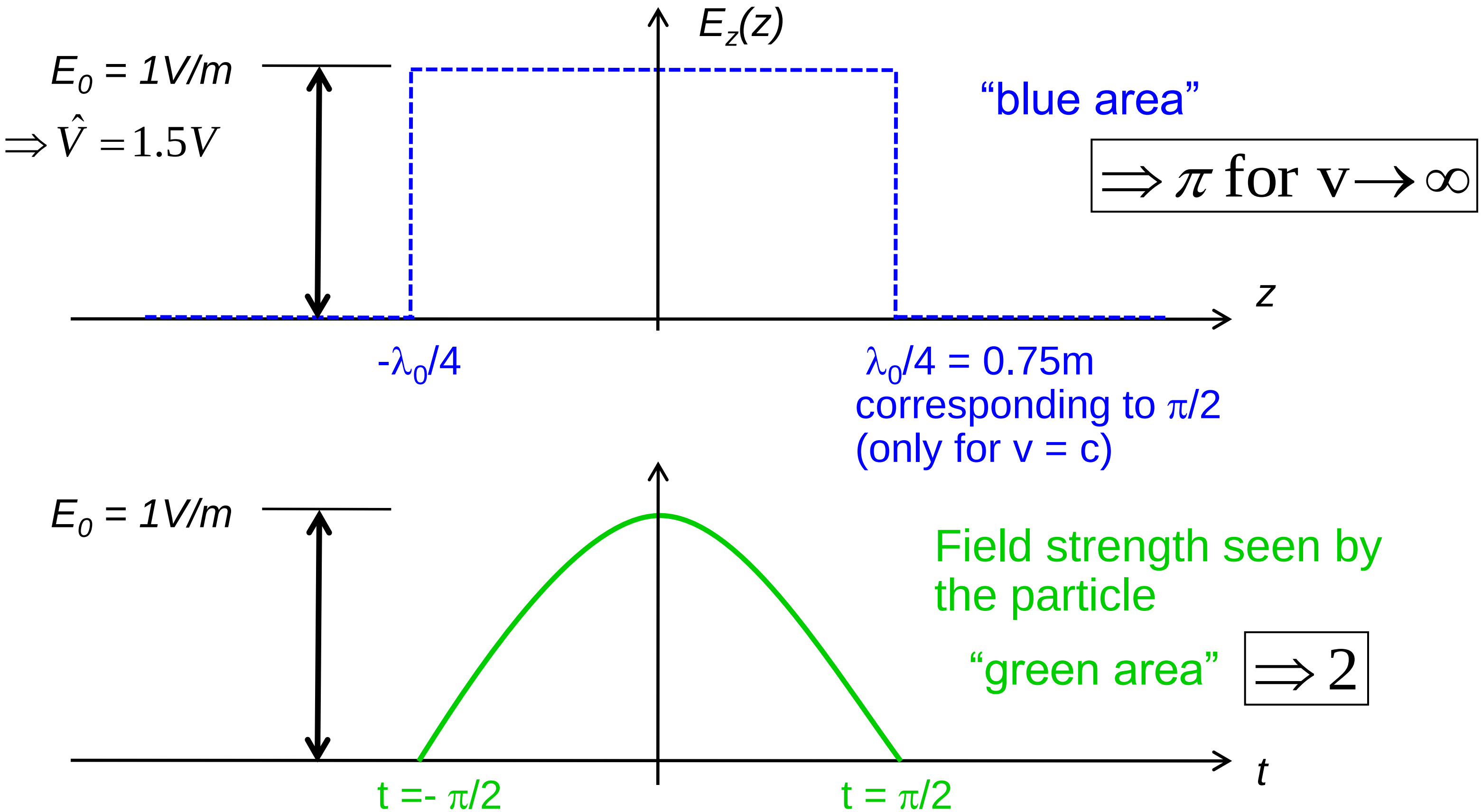
Transit time factor, p.565f. ,Alexander Wu Chao, Handbook of Accelerator Physics and Engineering

Transit time factor (3)

Example: Cavity gap length $L = \lambda_0/2$

λ_0 = 1m corresponding to f = 300MHz

particle velocity $v = c$ or $\beta = 1$



Acceleration

We have "slow" particles with β significantly below 1. They become faster when they gain energy and in a circular accelerator with fixed radius we must tune the cavity (increase its resonance frequency).

When already highly relativistic particles become accelerated (gaining momentum) they cannot become significantly faster as they are already very close to c, but they become heavier. Here we can see very nicely the conversion of energy into mass. In this case no or little tuning of the resonance frequency of the cavity is required. It is sufficient to move the frequency of the RF generator within the 3dB bandwidth of the cavity.

Fast tuning (fast cycling machines) can only be done electronically and is implemented in most cases by varying the inductance via the effective μ of a ferrite.

Tuning of cavities (1)

Slater's perturbation theorem: $$\frac{\Delta f}{f} = -\frac{1}{2}\frac{\Delta W}{W}$$

with W designating the energy stored in the cavity

- Inductive tuner
  - In regions of high magnetic field
  - increases resonant frequency ($\Delta W < 0$)

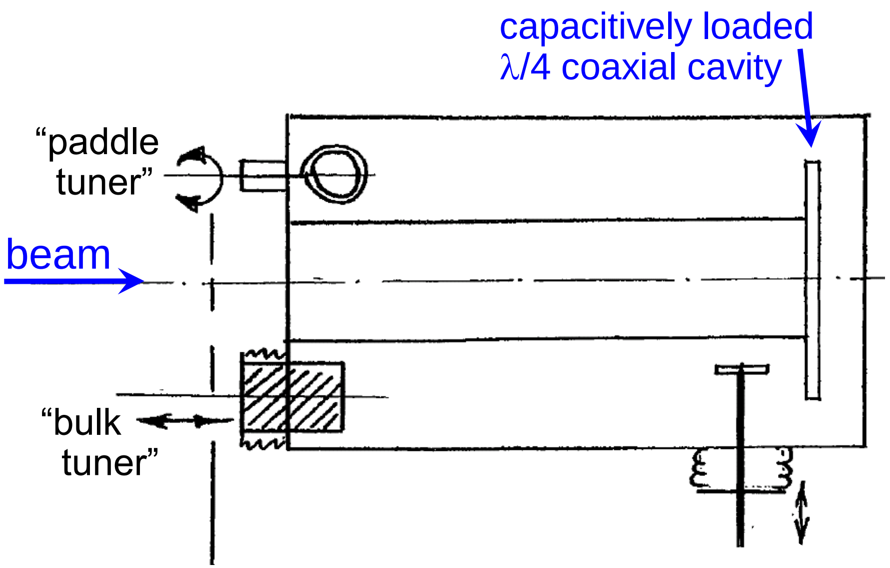
$$\Delta W = -\frac{LI^2}{2}; \quad dW = -\frac{\mu_0 \mu_r H^2}{2} dV$$

- Capacitive tuner
  - In regions of high electric field
  - decreases resonant frequency ($\Delta W > 0$)

$$\Delta W = \frac{CI^2}{2}; \quad dW = \frac{\varepsilon_0 \varepsilon_r E^2}{2} dV$$

Tuning of cavities (2)

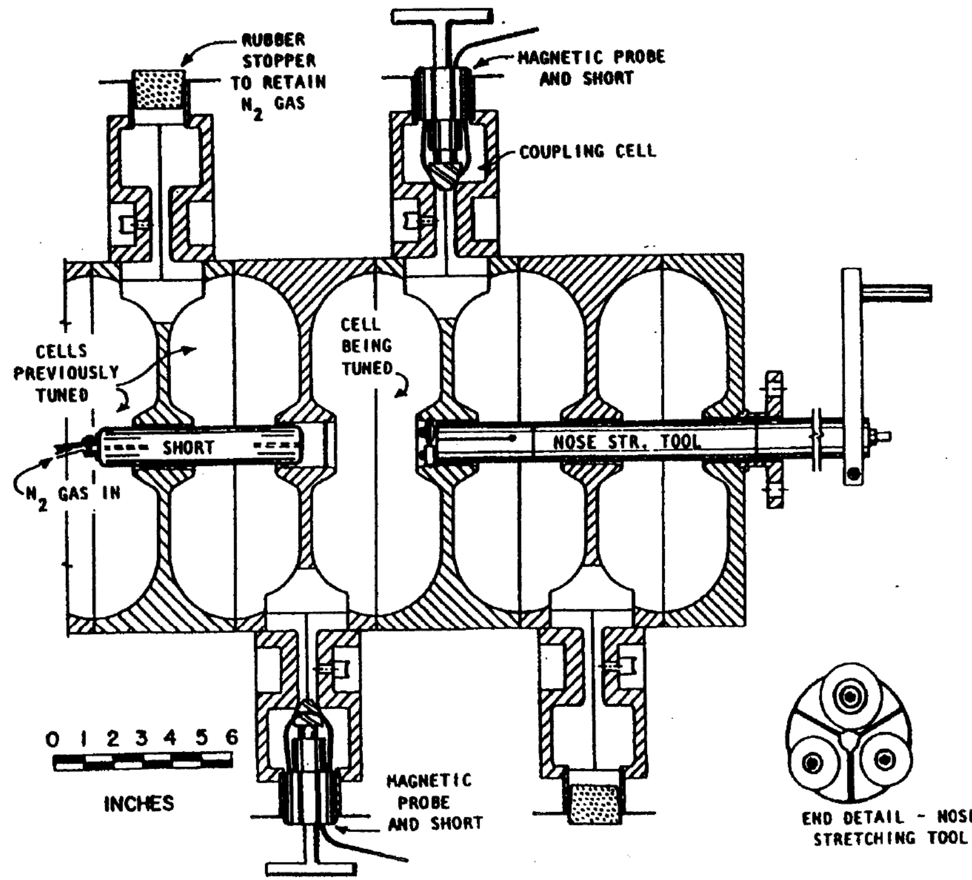
- Operational tuning
  - paddle tuner
  - bulk tuner



- Tuning by global deformation
  - mechanically (SC cavities)
  - thermally

Source of right image: G.R Swain et al., "Cavity tuning for the LAMPF 805 MHz Linac", 1972 Linac Conference, p. 242

- Initial tuning during manufacture
  - Bumping, dummy tuners
  - modification of joint sizes or plating thickness

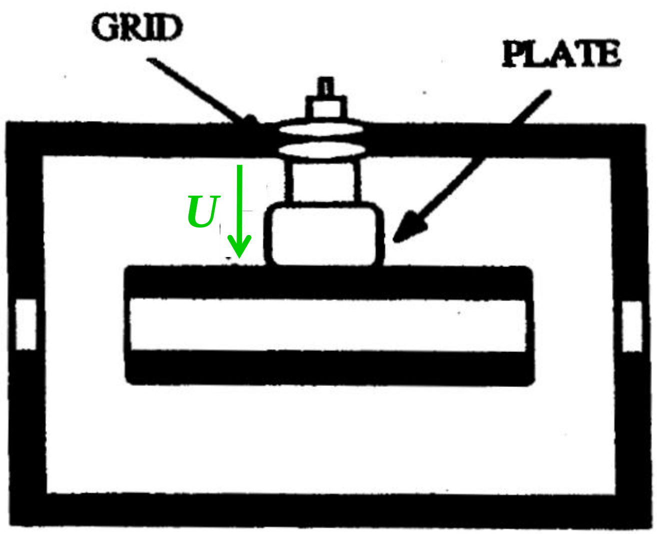


Coupling cavities to the outside world (1)

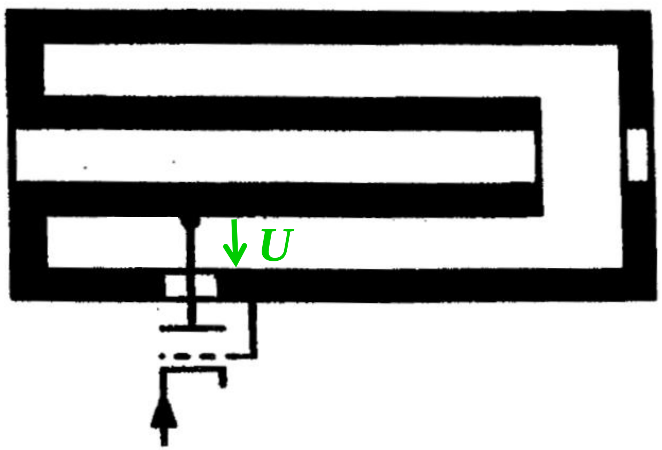
- Direct coupling (DC coupling)

  Generator (tube) has to "see" a certain voltage U

"Wideroe" or $\lambda/2$ structure



basic $\lambda/4$ - resonator



Source: M. Puglisi: "Conventional RF cavity design"
CAS "RF engineering for Particle Accelerators", CERN 92-03, Vol. 1

Coupling cavities to the outside world (2)

- Inductive coupling

Generator requirement:

$$U = \sqrt{2PZ}$$

P ... required power

Z ... optimum load resistance

Induced voltage in loop:

$$U = \mu_0 \frac{d}{dt} \int_S H ds$$

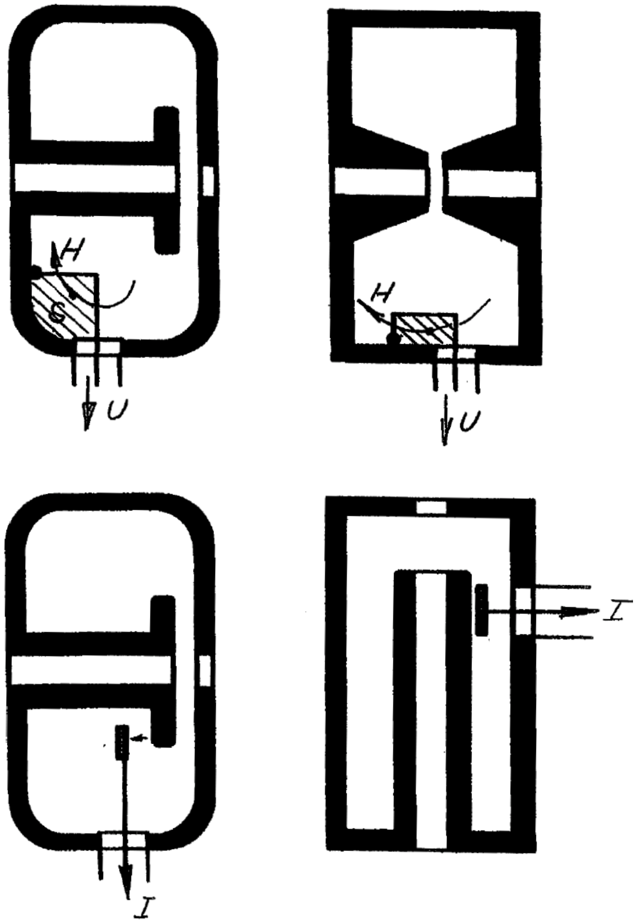
- Capacitive coupling

Generator requirement:

$$I = \sqrt{2P/Z}$$

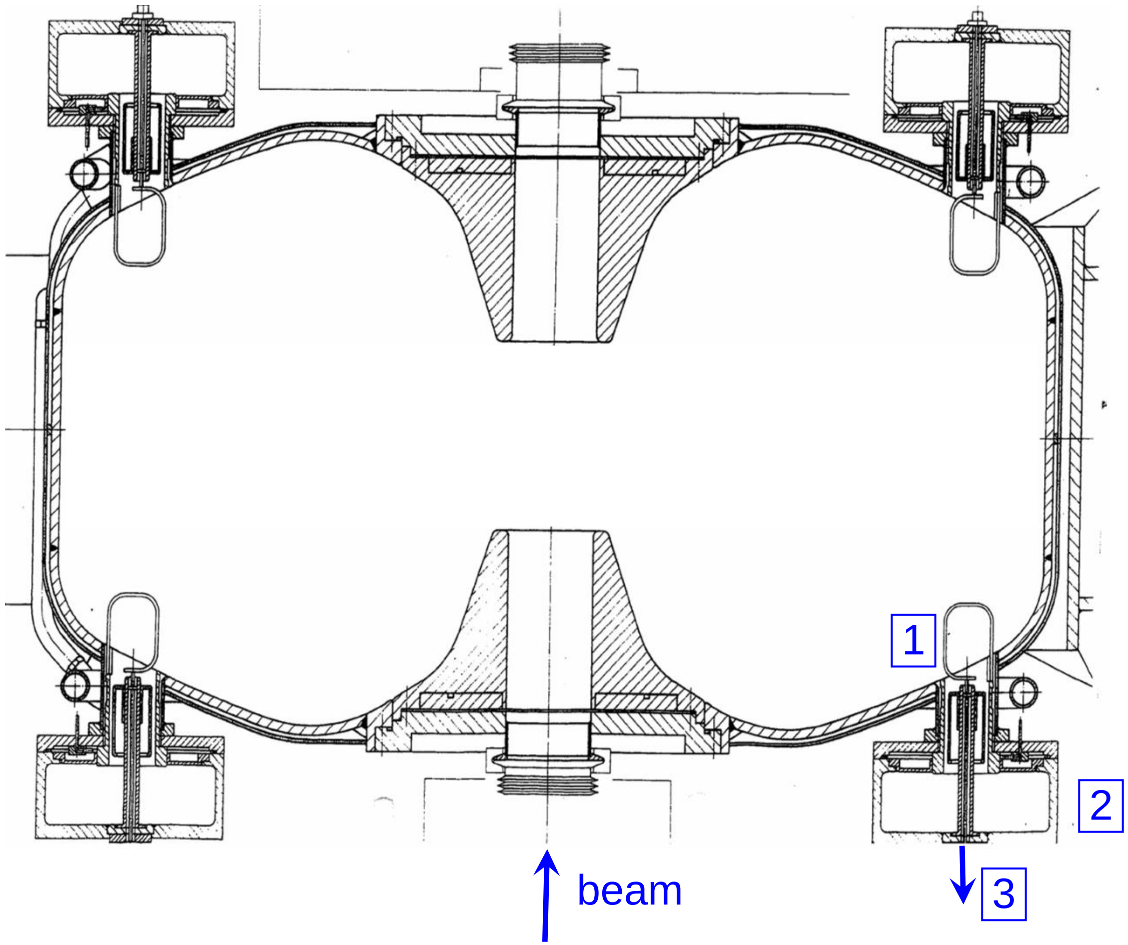
Induced displacement current

$$U = \varepsilon_0 \frac{d}{dt} \int_S E ds$$



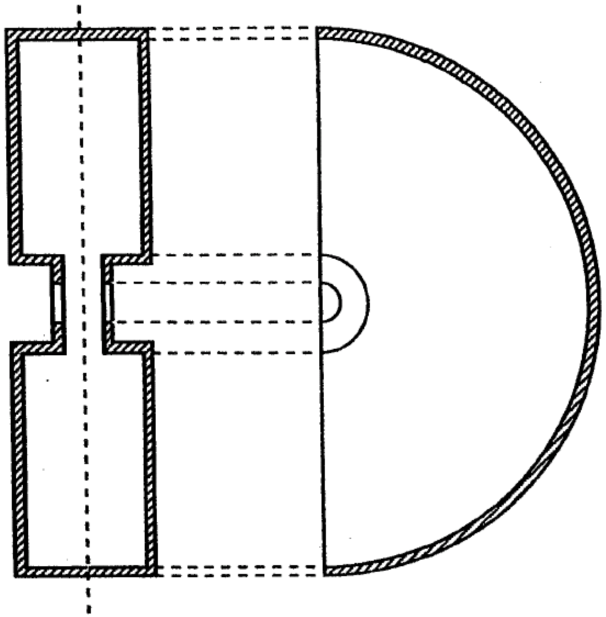
General example

A single-cell configuration: 114-MHz room temperature cavity of CERN PS. Type I profile with nose-cone to optimize shunt impedance

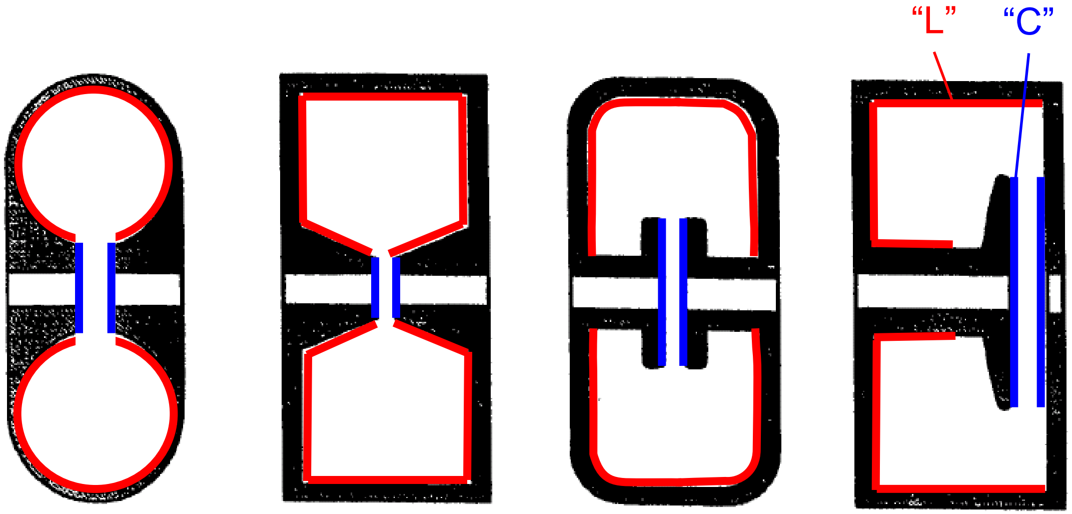


1: higher order mode (HOM) coupling loop which serves for eliminating beam-induced power
2: HOM filter
3: HOM power guided towards load and dissipated

Different forms of the pillbox cavity



Cross section of a radial cavity



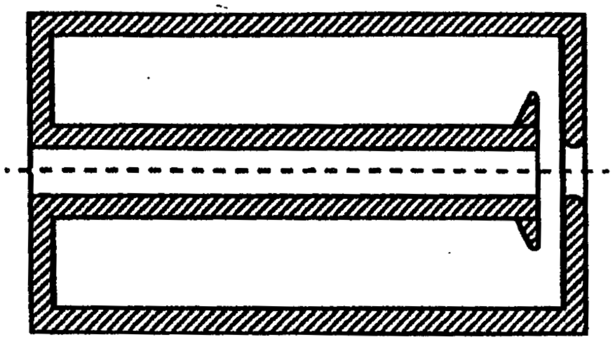
nose-cone cavities

disc-loaded cavities

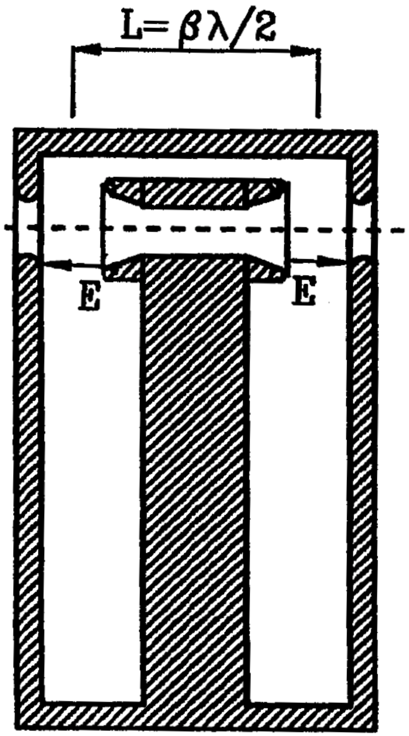
Four different cross sections of fundamentally similar cavities. In spite of their similarity they have been given different names...

Source: M. Puglisi: “Conventional RF cavity design”
CAS "RF engineering for Particle Accelerators", CERN 92-03, Vol. 1

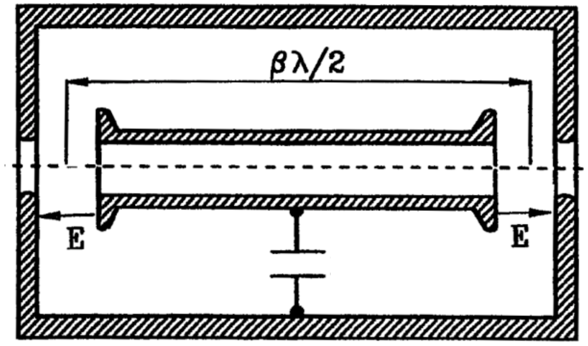
The coaxial (TEM-mode) cavity



basic λ/4 - resonator



modified λ/4 - resonator for acceleration in βλ/2-mode



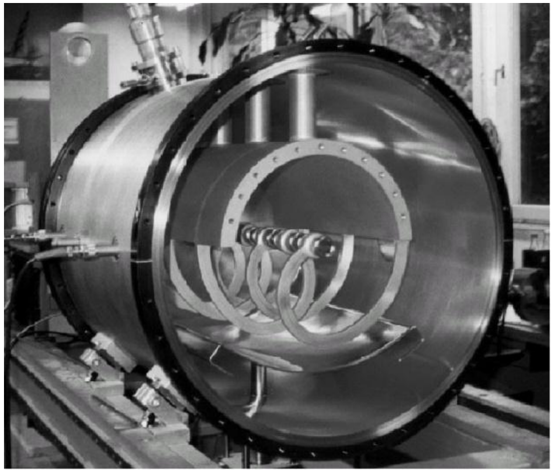
“Wideroe” or λ/2 structure

works for β=0.2-0.4
does not work for β=1 as we are in a quasi-static case

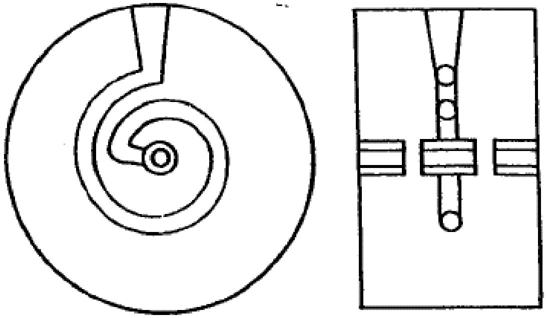
Source: M. Puglisi: “Conventional RF cavity design”
CAS "RF engineering for Particle Accelerators", CERN 92-03, Vol. 1

Spiral resonators

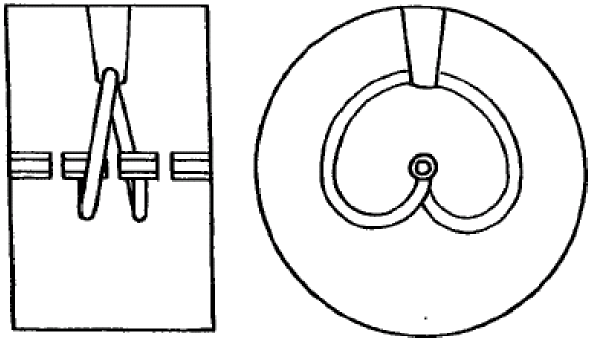
- For small β relatively low RF frequencies have to be used
- Drift tubes are mounted on λ/4 lines acting as λ/4 resonators (will be treated in second part of lectures)
- Long λ/4 lines coiled up to make structure smaller



A β = 5.4 % power resonator



Spiral resonator.

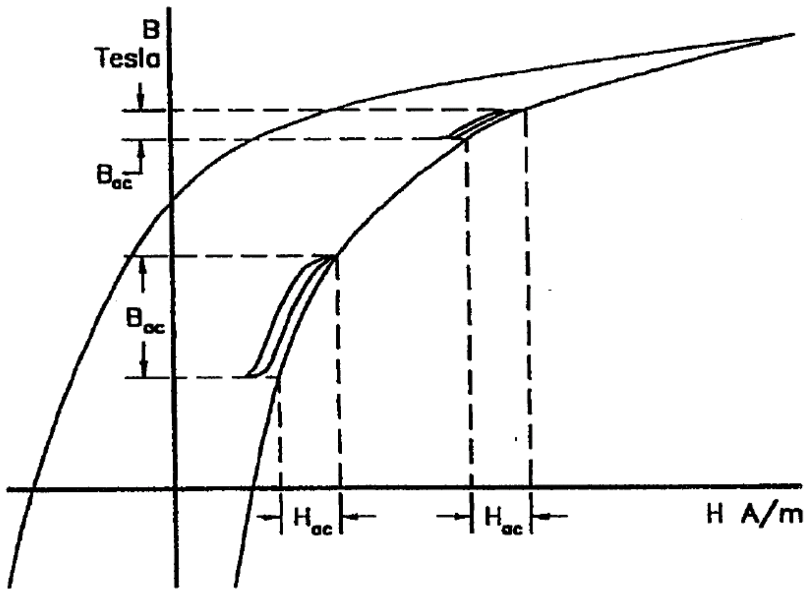


Split-ring resonator.

Ferrite loaded cavities (1)

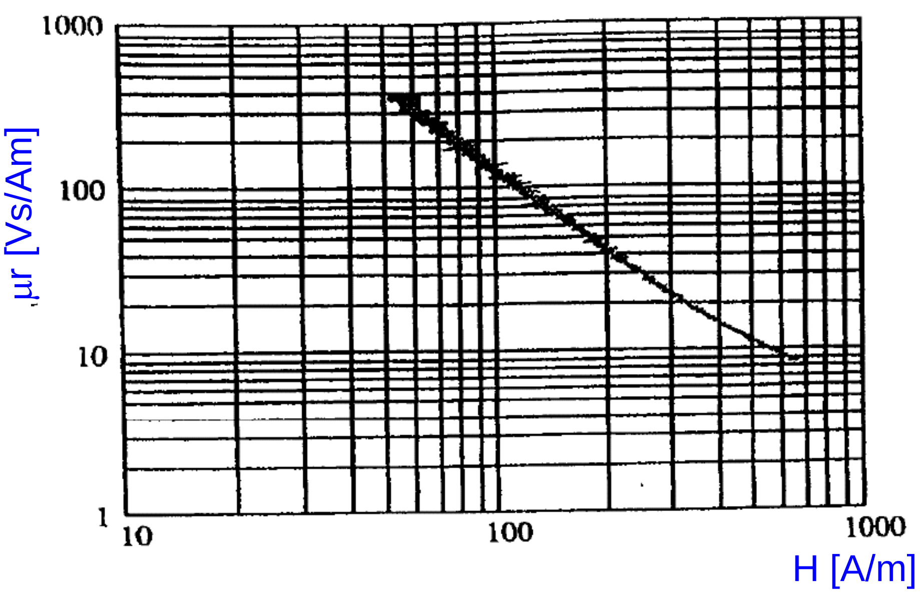
- Tuning possible by choosing an appropriate static or slowly varying magnetic bias field => differential μ adjustable. Bias field and RF field are parallel.

B versus ***H*** with added AC field



A period of the AC field corresponds to one round in one of the little hysteresis loops.

μ_r versus ***H***

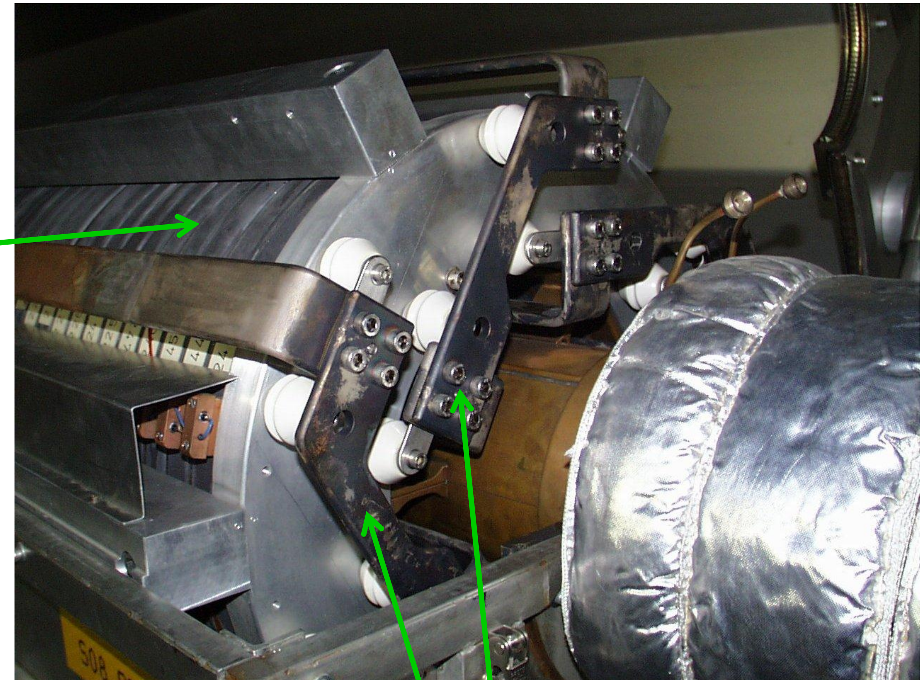


Bias ***B*** field parallel to RF magnetic field in above plot

Ferrite loaded cavities (2)

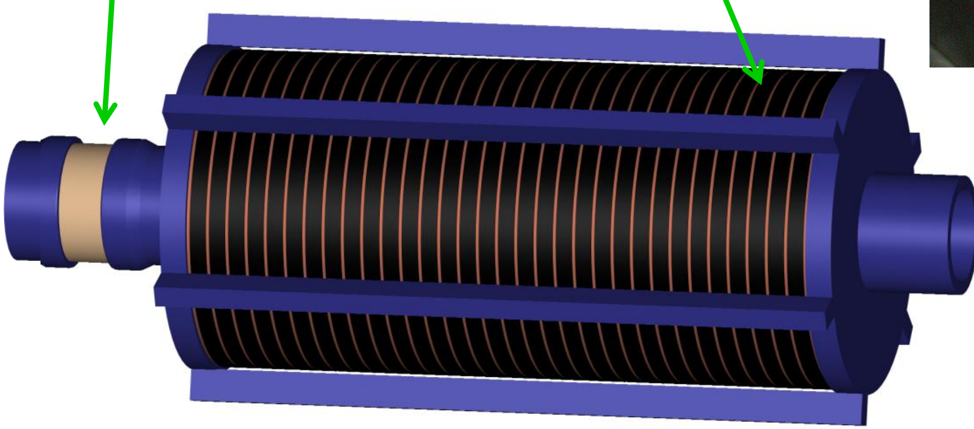
This is essentially a λ/4 cavity with magnetically variable length.

Ferrite toroidal discs, interleaved with copper sheets on "equipotential lines" for cooling.



Picture: K. Kasper (GSI)

ceramic window



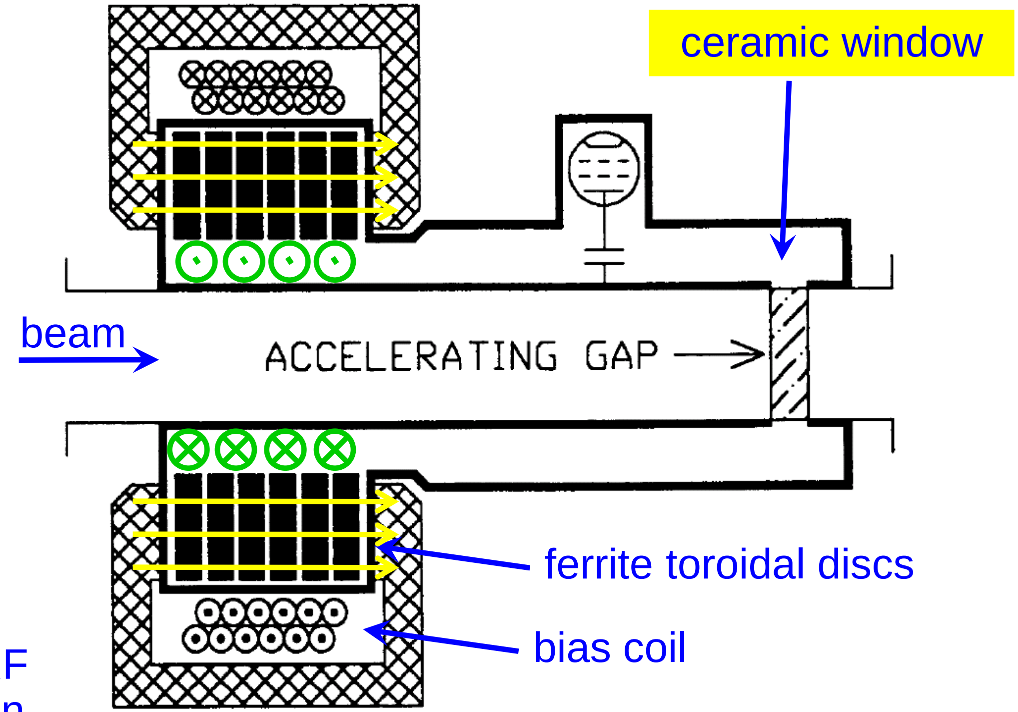
Bus bars to supply the DC bias, the **DC field is parallel (azimuthally)** to the RF field.

Source: H. Damerau (CERN), private communication
Cavity in the SIS (SchwerIonen Synchrotron)
at GSI, frequency range from 0.8 – 5.4 MHz

Ferrite loaded cavities (3)

- Ferrite loading makes line electrically longer => cavity size can be reduced
- Bias ***B*** field in ferrite orthogonal to RF magnetic field
- tunable between 46 and 61 MHz by variable bias field

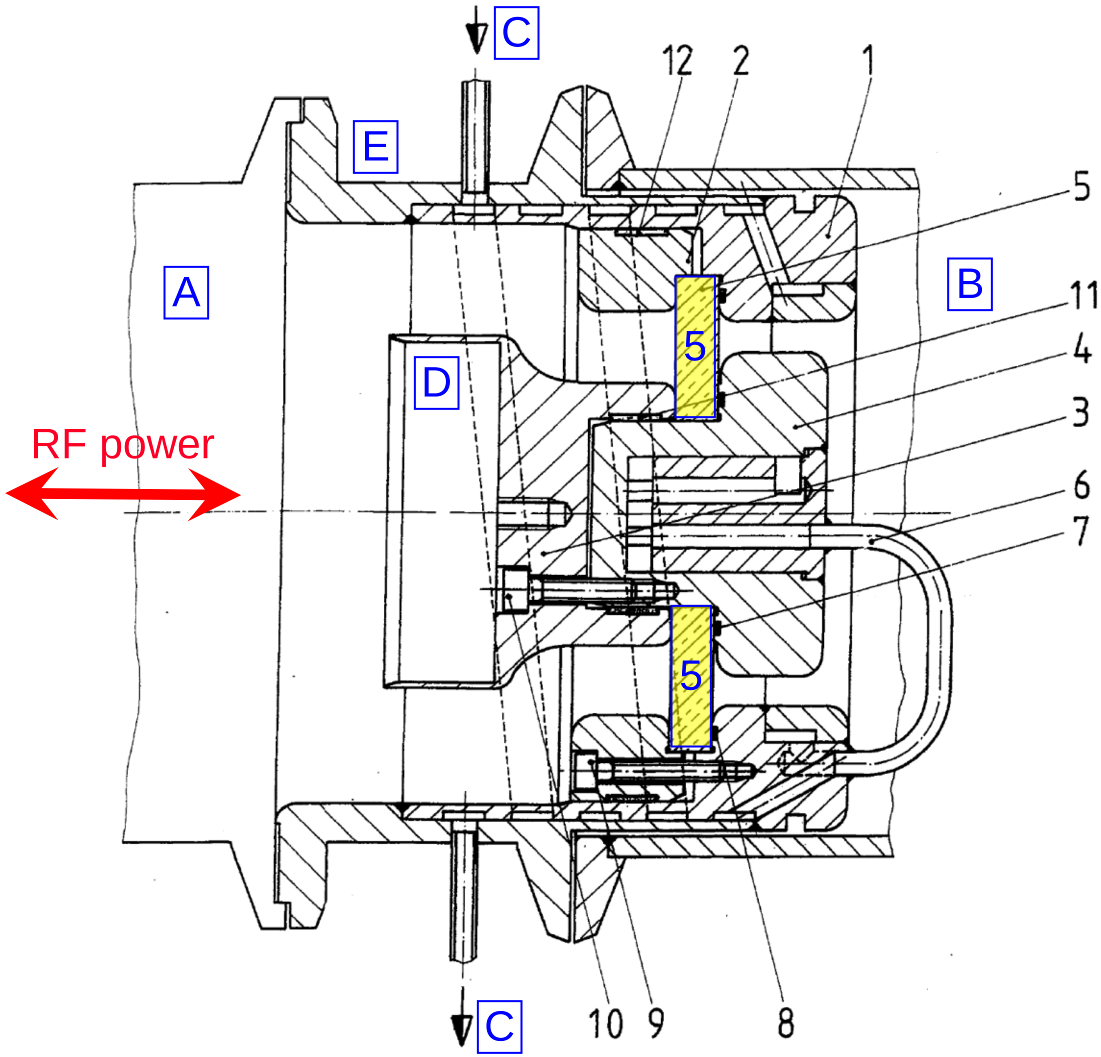
Tunable cavity for TRIUMF (Three Universities Meson Facility, Vancouver) designed by LANL (Los Alamos).



In this mode of operation of orthogonal bias, the DC field is orthogonal to the RF field. The μ of the ferrite can be varied in a more efficient way as compared to parallel magnetic bias.

From: ISK Gardner: "Ferrite dominated cavities" CERN 92-03, Vol. II [2]

RF window



An RF window for a 114 MHz LEP cavity

On which side is the vacuum?

How does the structure continue on the left side?

5: Ceramic disc
6: Coupling loop
7: Vacuum seal

A: Pressurized side (air)
B: Cavity side (vacuum)
C: Cooling water ducts
D: Inner conductor
E: Outer conductor

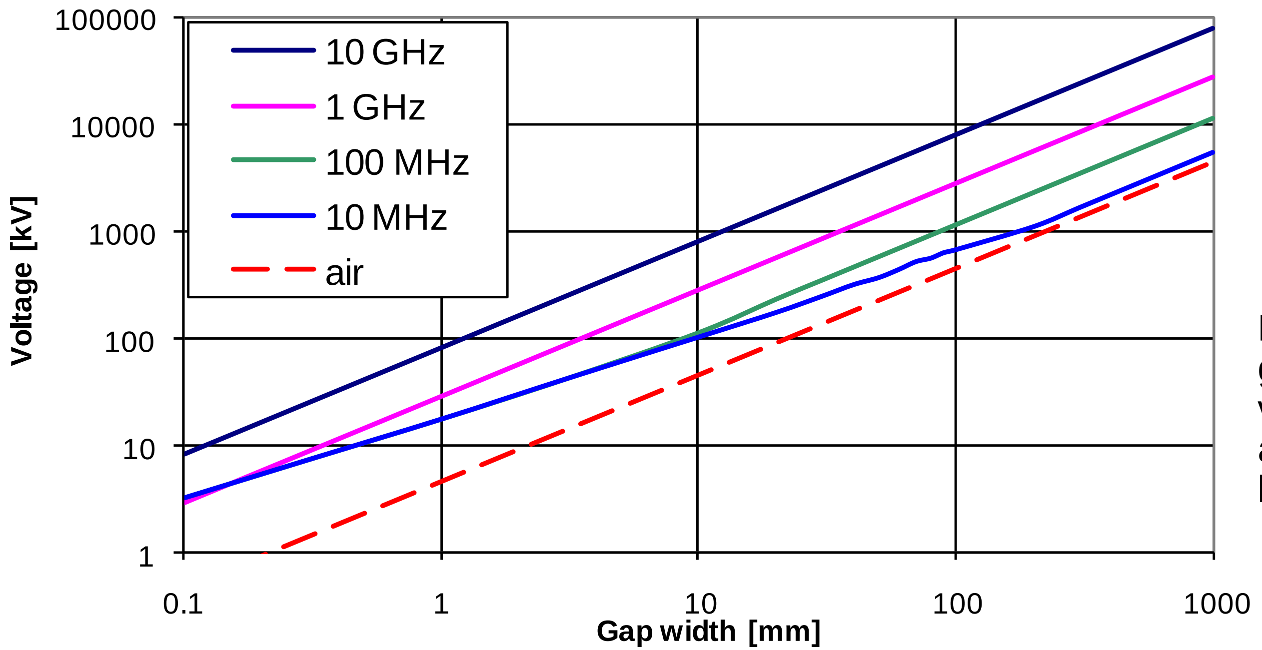
"Kilpatrick" voltage breakdown (1)

The maximum E field achievable is limited by a process known as RF **breakdown**.
The Breakdown voltage is given by

$$W \cdot E^2 \cdot e^{\frac{-1.7\cdot 10^5}{E}} = 1.8\cdot 10^{14}$$

where W [eV] is the impact energy of the electrons and E the electric field [V/m] (W = E * gap width for DC, W < E * gap width for RF).

- High power effect
- Destructive!!
- Breakdown voltage proportional to square root of frequency



Breakdown voltage given in plot for vacuum (solid lines) and in air (dashed line)

Ref.: W. D. Kilpatrick, Criterion for vacuum sparking designed to include both Rf and DC, The Review of Scientific Instruments, Vol. 28, No. 10, 1957

"Kilpatrick" voltage breakdown (2)

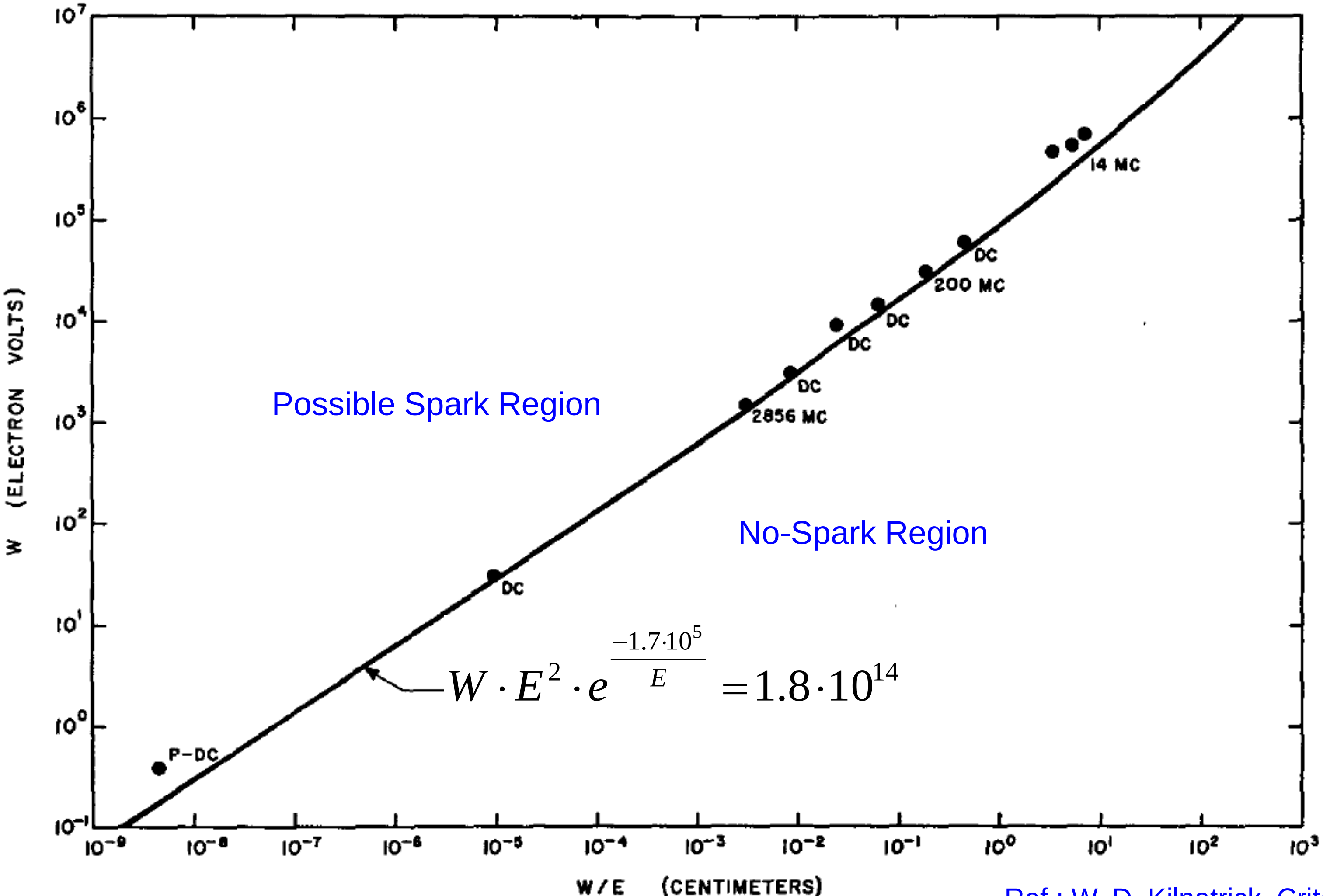
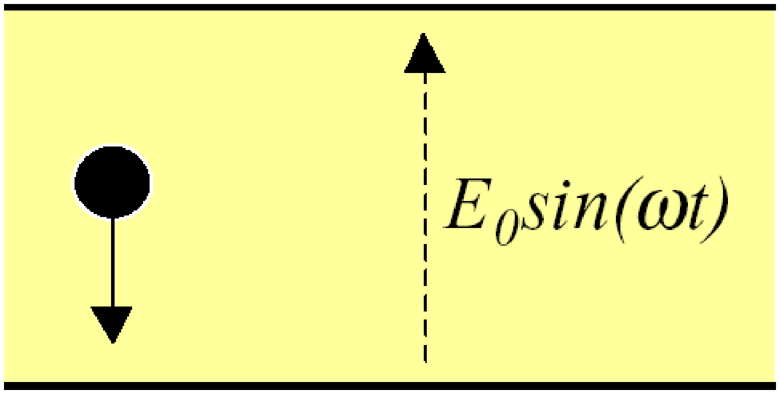
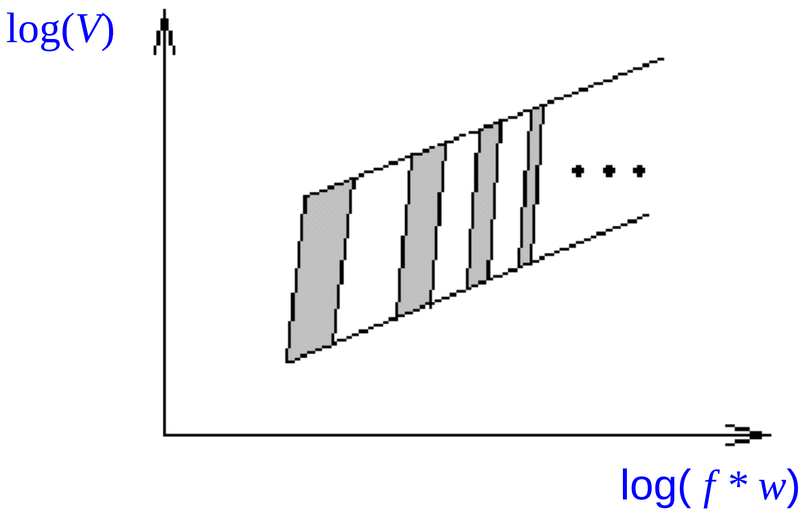


FIG. 1. W is the maximum ion energy at the cathode, in electron volts. For dc, W corresponds to the applied voltage, and W/E is the gap spacing for plane parallel fields. For rf, W is a function of frequency and gap (see text).

Ref.: W. D. Kilpatrick, Criterion for vacuum sparking designed to include both Rf and DC, The Review of Scientific Instruments, Vol. 28, No. 10, 1957

Multipactor (1)





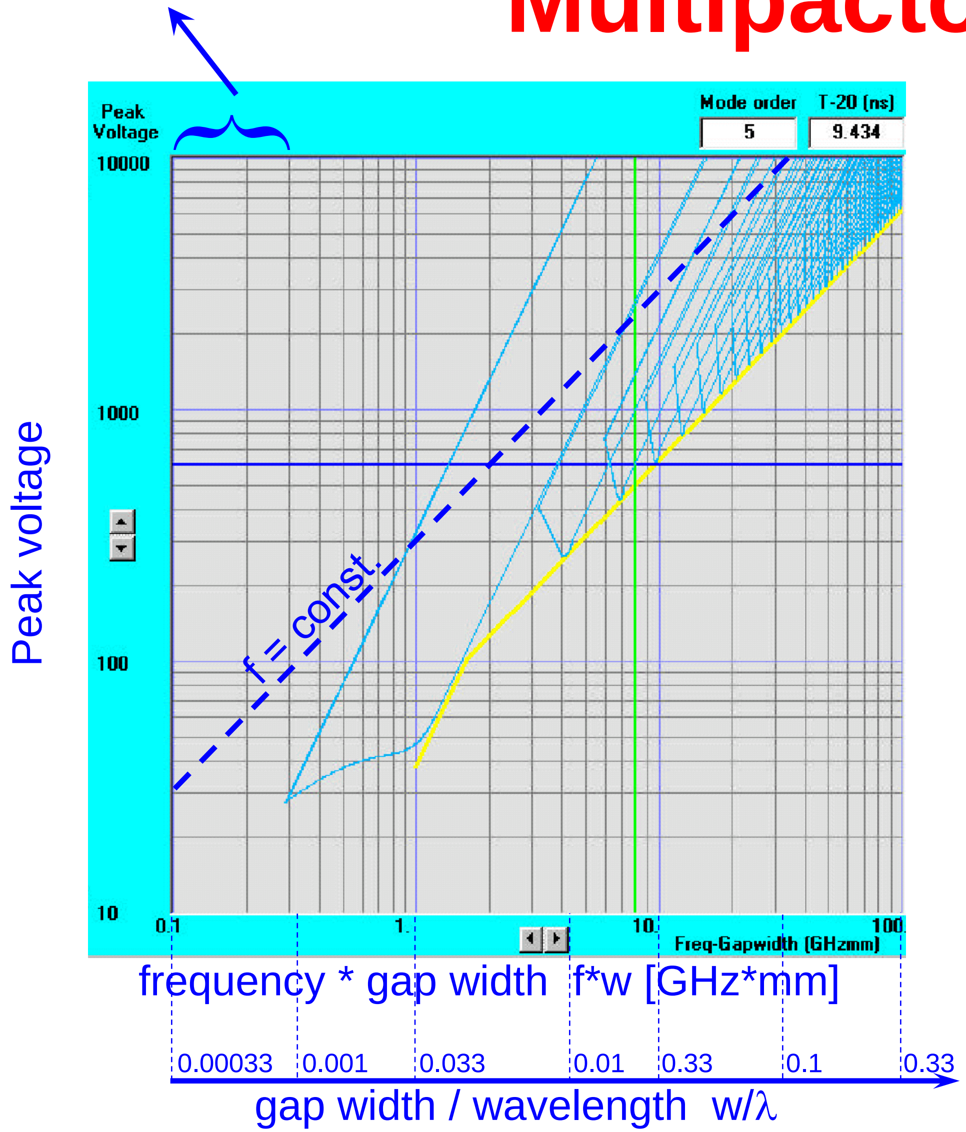
Basically,

- Electrons get accelerated in an electric field
- When they hit the wall, secondary electrons are freed
- If the electric field changes sign as it is the case for RF fields, the secondary electrons will eventually see an accelerating field
- Therefore, at least for some distinct frequency bands and accelerating voltages, resonance effects can be expected

V ... gap voltage
f ... RF frequency
w ... gap width

No multipactor for very small gap width or very high frequencies

Multipactor (2)

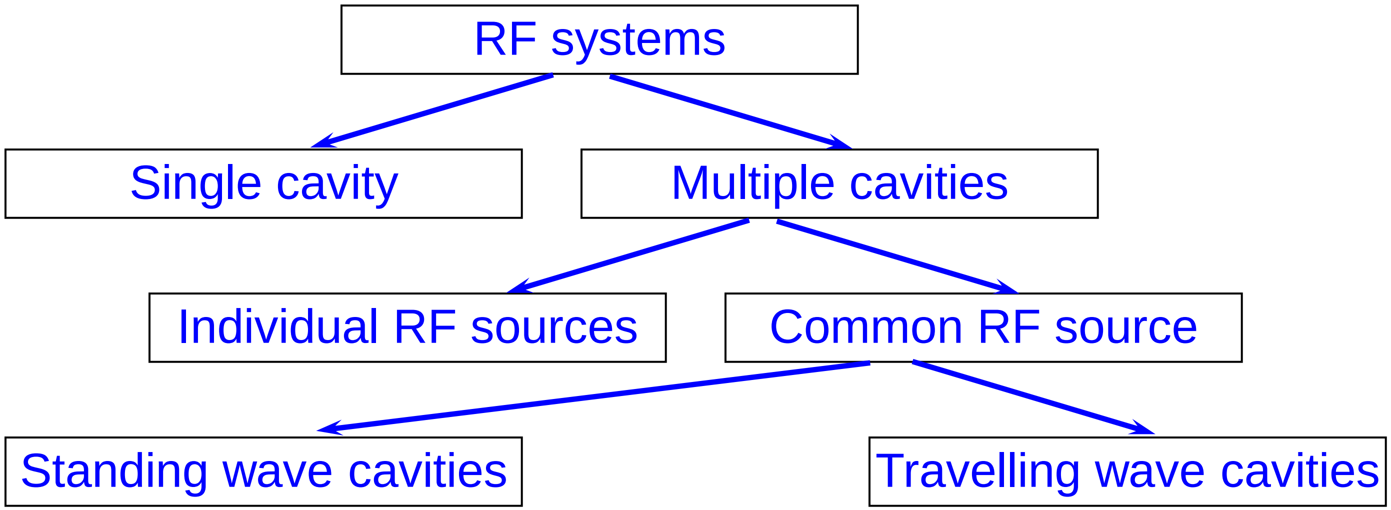


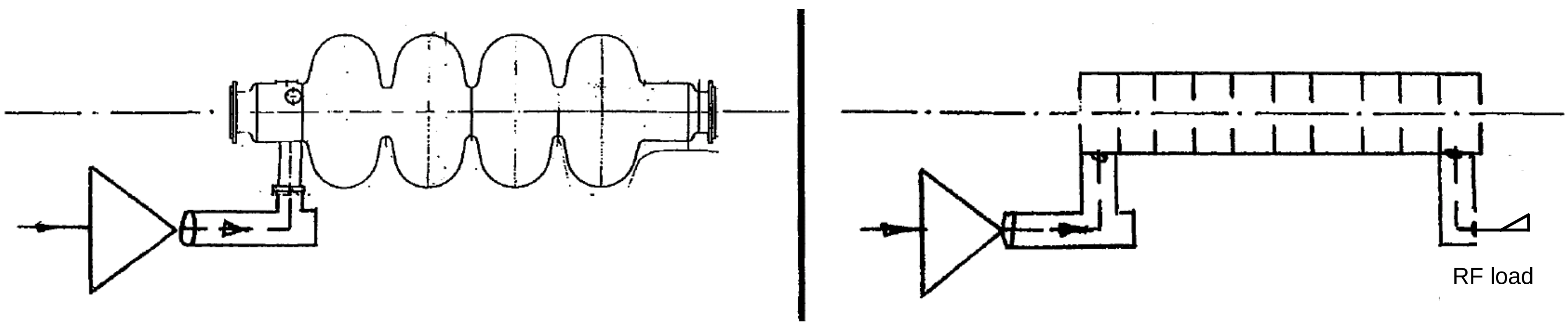
More formally,

- Multipactor is a resonant avalanche discharge, typically a low-power effect.
- The basic "two-point" resonance condition is met if the time of flight of an electron between electrodes equals an odd number of RF half cycles
- Other necessary condition: The coefficient of secondary electron emission must be larger than 1. This corresponds to an energy range between 50 eV and 5000 eV for copper surfaces

Multipactor calculator available at http://www.estec.esa.nl/multipac/

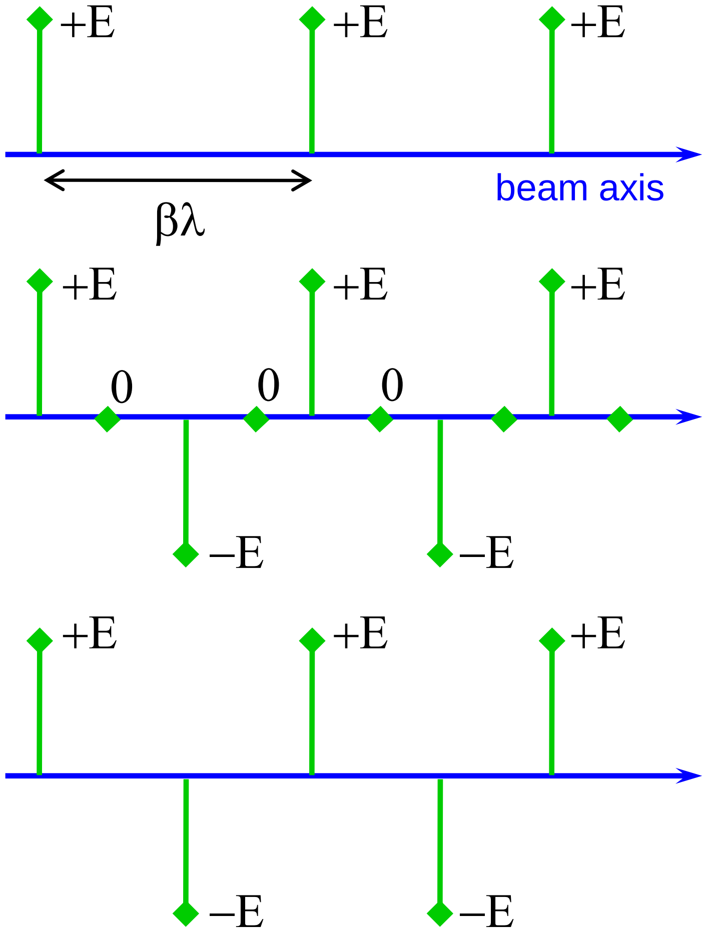
RF systems


RF systems
Single cavity
Multiple cavities
Individual RF sources
Common RF source
Standing wave cavities
Travelling wave cavities


RF load

Standing wave cavities

- The only possible phase differences between the SW fields in lossless cells are 0° or 180°.
- For N cells there are N possible longitudinal modes. Practically used modes:
  - 0° (zero mode): Gap distance $\beta\lambda$ with $\beta = v/c$. Structures: Alvarez Drift Tube Line (DTL)
  - 90° ($\pi/2$ mode): Distance active cell to coupling cell $\beta\lambda/4$ or $\beta\lambda/2$. Structures: Side coupled, Disk and Washer
  - 180° (π mode): Gap distance $\beta\lambda/2$. Structures: Wideroe, superconducting cavities (LHC, TESLA), Interdigital (IH)

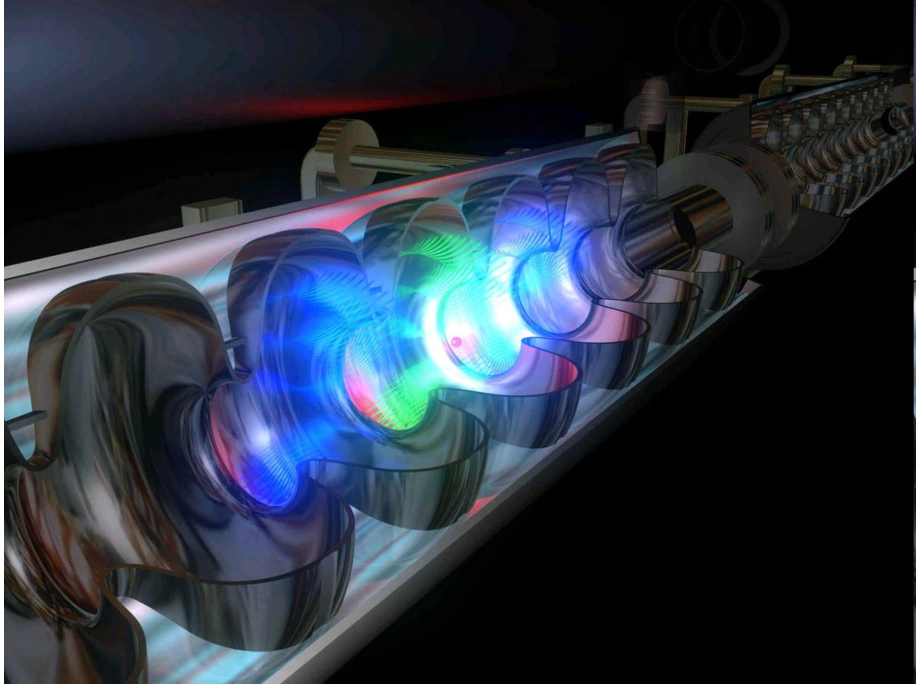


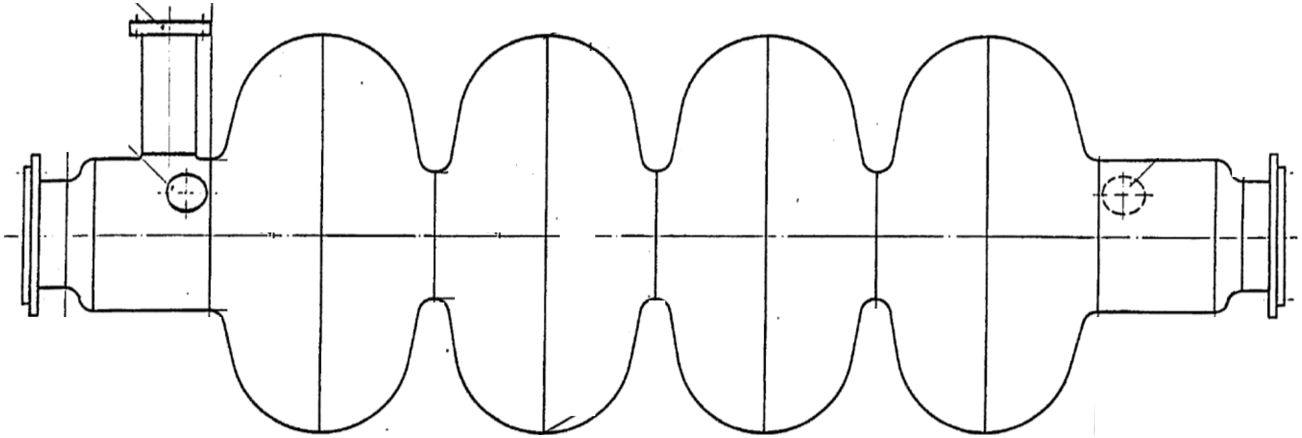
Travelling wave cavities

- Phase difference between adjacent cells can be chosen arbitrarily to assure synchronism with beam. Values around $2\pi/3$ give best compromise between structure length, group velocity (filling time) and overall dissipation.
- Without beam loading, almost the full input power is dissipated in the absorber.
- The field decay due to attenuation of the structure can be taken into account by designing "constant gradient“ rather than constant geometry structures.
- There exists a specific amount of beam loading for which all RF power is transmitted to the beam, resulting in zero power dissipated in the absorber =>"fully loaded" structure.
- Repercussion of beam loading and structure transients on generator is minimised.
- Structures
- Loaded waveguide (generally used in electron linacs, e.g. CERN LIL, CLIC ...), Parallel bar

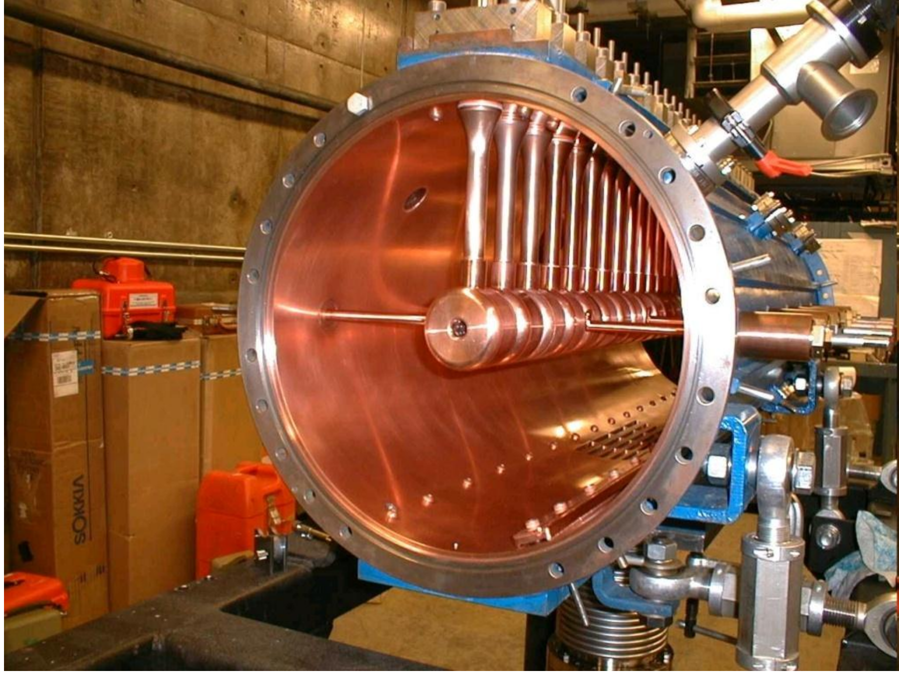
Examples (1)

- Standing wave cavity in multicell configuration
- This superconducting cavity was used in CERN's LEP
- "Type II" profile without nose-cone to avoid multipactor and reduce r/Q

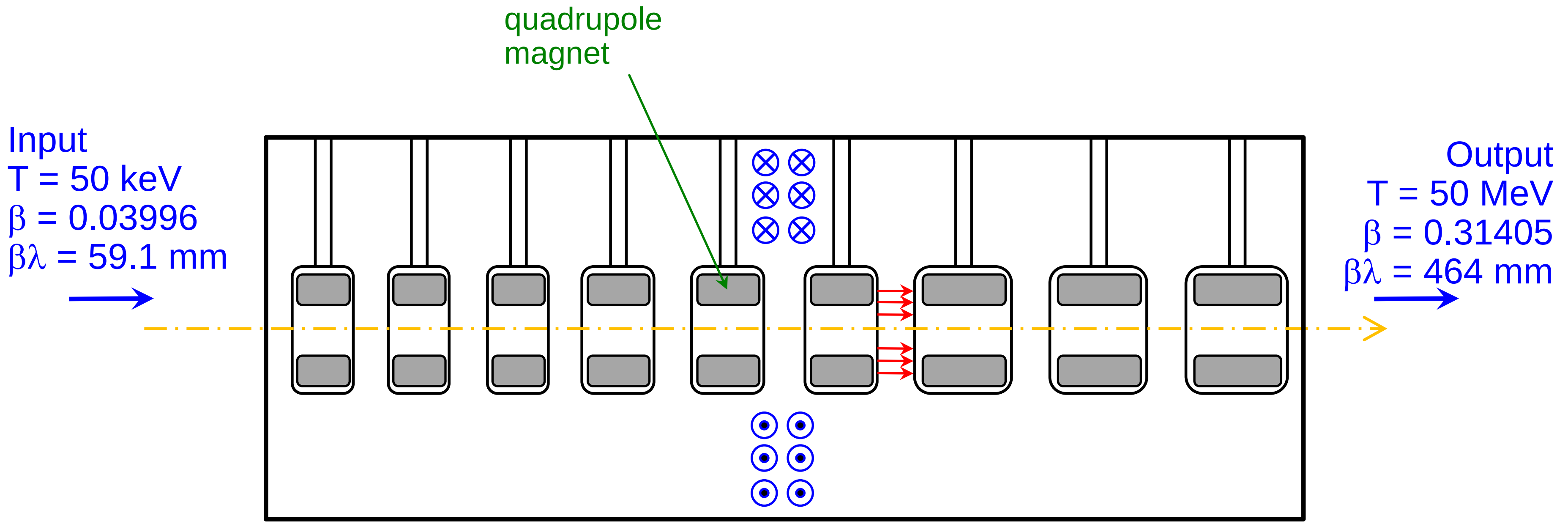




Examples (2)

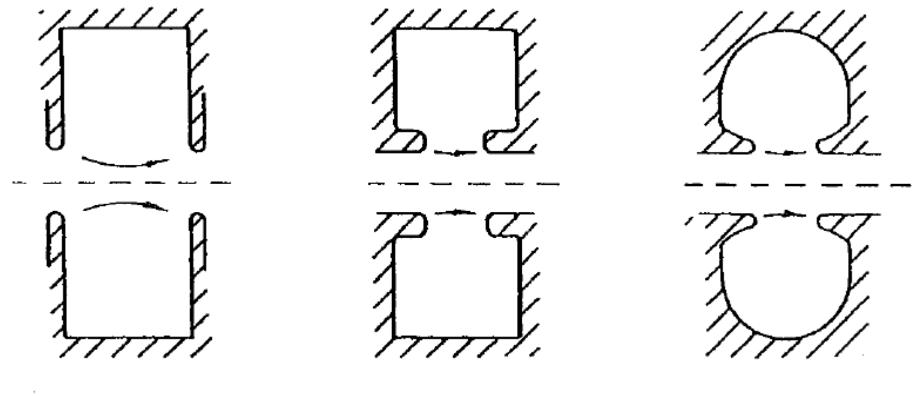


- Travelling wave cavity
- Below: An ALVAREZ structure (Drift tubes with interposed quadrupole magnets), used in the CERN 50 MeV Proton LINAC
  Frequency: 202.56 MHz

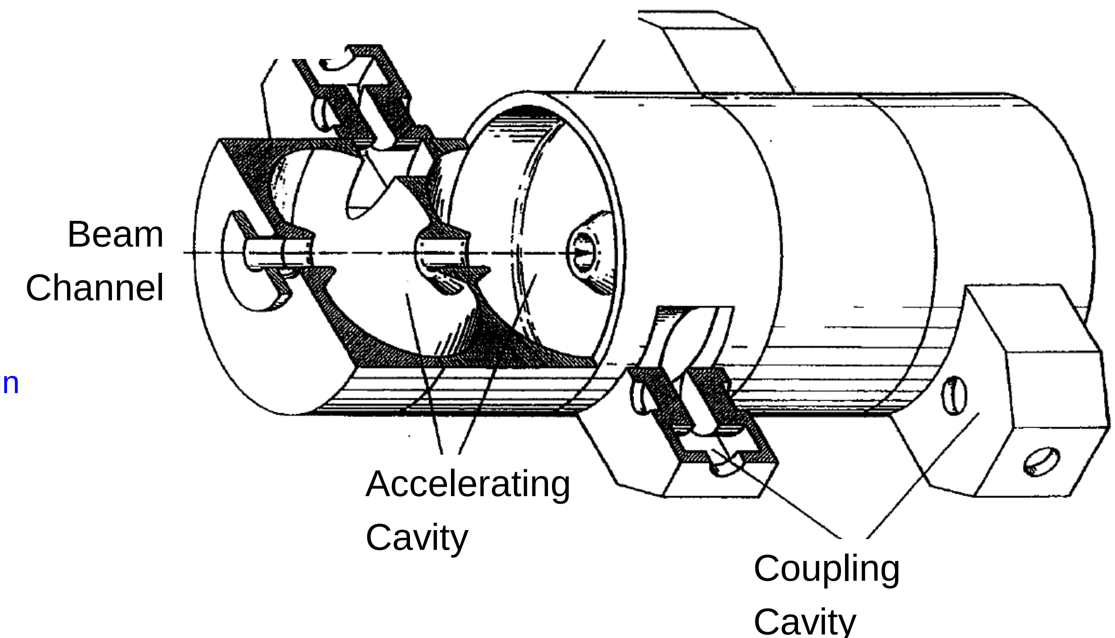


Side coupled structures

- Cavity geometry changes to optimize shunt impedance
- Higher shunt impedance => higher accelerating gradient



- Side coupled cavity configuration for optimum shunt impedance

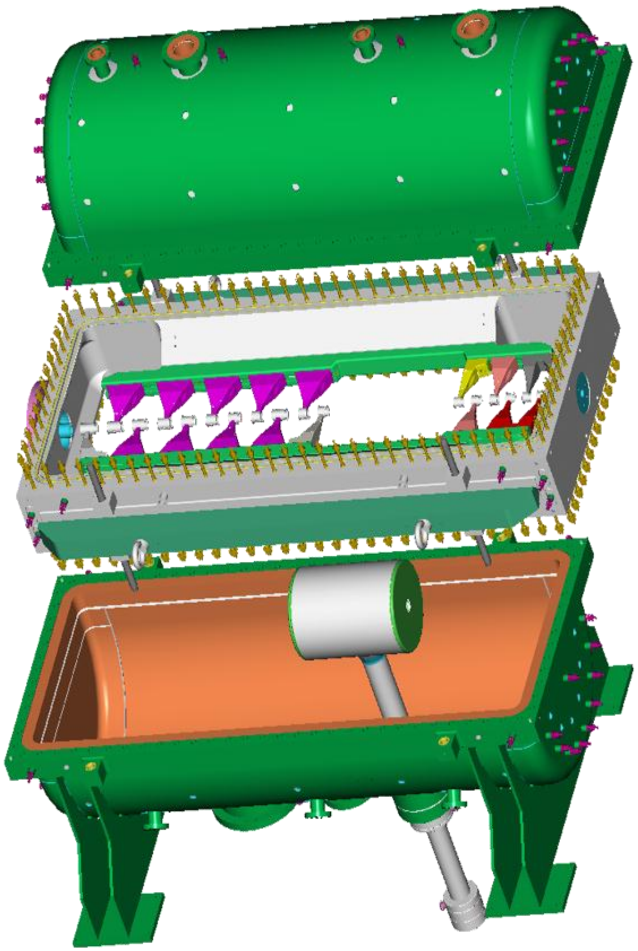


Source: E.A. Knapp and W Shlaer: Design and initial performance of a 20MeV high-current side-coupled cavity electron accelerator, 1968 Linac Conference Proceedings, p. 635 to 649

The IH structure

- IH stands for Interdigital H mode
- Interleaved fingers “adapt” the deformed H (TE) mode that is usually deflecting
- Inside the resonator tank cylindrical cavity drift tubes of varying length (matching the ion velocity) are mounted alternating on opposite sides. The magnetic field lines are parallel to the beam axis and the induced currents flow azimuthally on the wall, creating electric fields of alternating direction between the drift tubes. This field forces the ions forward.
- The big pot is necessary for transverse focusing.

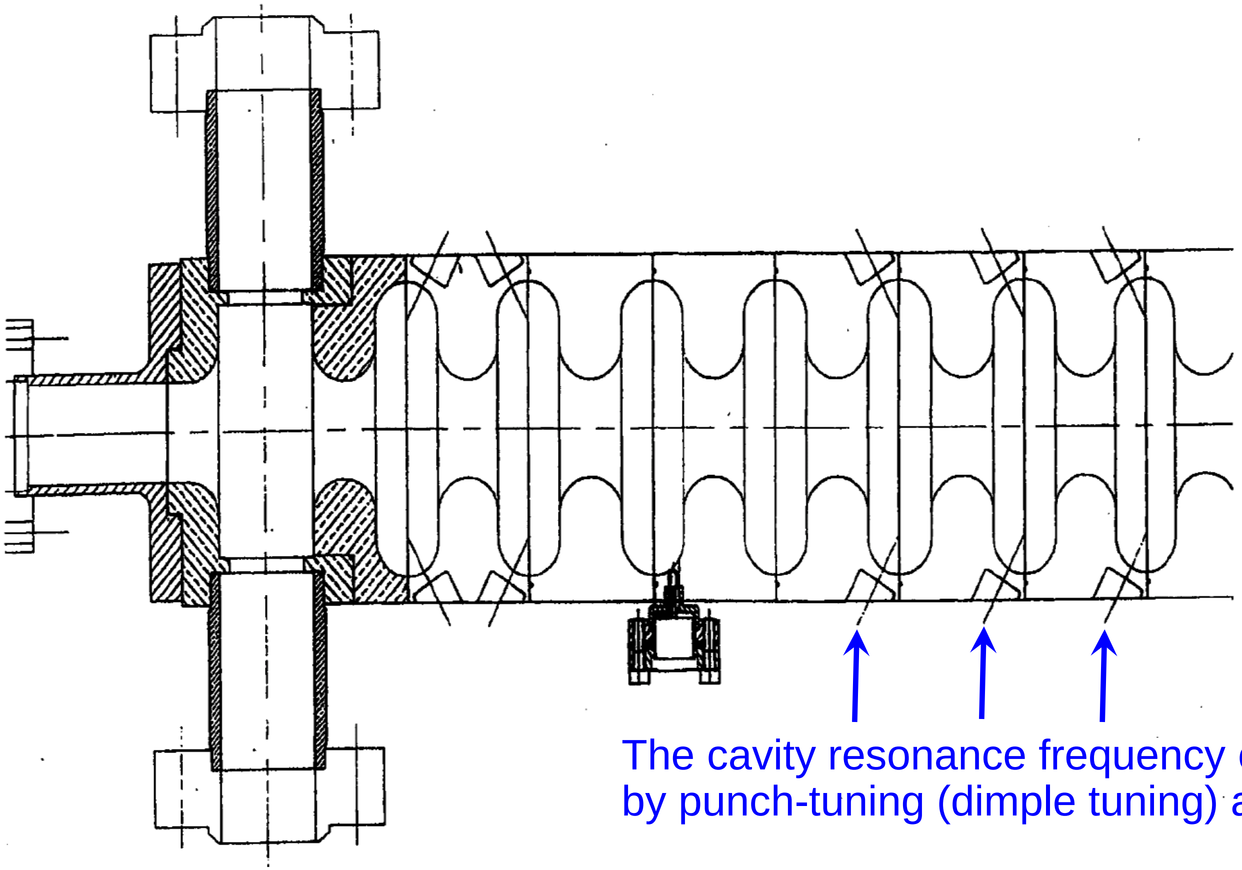
Properties of the structure on the right:
E_{in} = 300 keV, E_{out} = 1.1-1.2 MeV
Electrode voltage V_{eff} = 4.05 MV
Tank length L = 1.5 m
Number of gaps = 20
Peak power consumption W_{peak} = 36 kW



Source: fy.chalmers.se/subatom/f2bfw/poster97_ps_pic/ihstructure.ppt

A disc loaded waveguide structure

CERN LIL (Linear Injector for LEP), operating frequency 2.98 GHz



The cavity resonance frequency can be changed by punch-tuning (dimple tuning) at these spots.

TEM transmission lines (1)

Transverse electric modes (TEM) can propagate on any structure with at least two conductors

Given a structure with

C' ... capacitance per unit length [F/m]

L' ... inductance per unit length [H/m]

It then follows

characteristic impedance $Z = \frac{V_{\text{wave}}}{I_{\text{wave}}} = \sqrt{\frac{L'}{C'}}$ $\quad [Z] = \Omega$

velocity of propagation $v = \frac{1}{\sqrt{L'C'}} = \frac{c_0}{\sqrt{\mu_r \varepsilon_r}}$ $\quad [v] = m/s$

If $\mu_r = \varepsilon_r = 1$ (vacuum or approximately air), then the velocity of propagation is equal to the velocity of light $c_0 = 2.998\ \text{E}8\ m/s$

TEM transmission lines (2)

Formulae for the characteristic impedance Z can be found in many textbooks (e.g. "Reference Data for Radio Engineers" or others). From a known Z the values for C' and L' can be deduced by

$$\left.\begin{aligned} C' &= \frac{1}{vZ} \\ L' &= \frac{Z}{v} \end{aligned}\right\} \text{ for "normal" cable } (\mu_r = 1)$$

$$C' = \frac{100\sqrt{\varepsilon_r}}{3Z} \qquad [C'] = pF/cm$$

$$L' = \frac{\sqrt{\varepsilon_r}}{30} Z \qquad [L'] = nH/cm$$

$$\text{For coaxial cables:} \quad Z = \sqrt{\frac{\mu_r}{\varepsilon_r}}\, 60 \ln\left(\frac{R}{r}\right)$$

TEM transmission lines (3)

Coaxial cable with minimum loss:

$$\alpha_R = \frac{R_S\sqrt{\varepsilon_r}}{Z_0 D}\cdot\frac{1+\frac{D}{d}}{\ln\left(\frac{D}{d}\right)}$$

R_S Surface resistance

$$R_S = \frac{\rho}{\delta} = \frac{1}{\sigma\delta}$$

$$f_\alpha = \alpha_R\frac{Z_0 D}{R_S\sqrt{\varepsilon_r}} = \frac{1+\frac{D}{d}}{\ln\left(\frac{D}{d}\right)}$$

$$Z_L = \sqrt{\frac{\mu_0}{\varepsilon_0}}\,\frac{\ln\left(\frac{D}{d}\right)}{2\pi\sqrt{\varepsilon_r}}$$

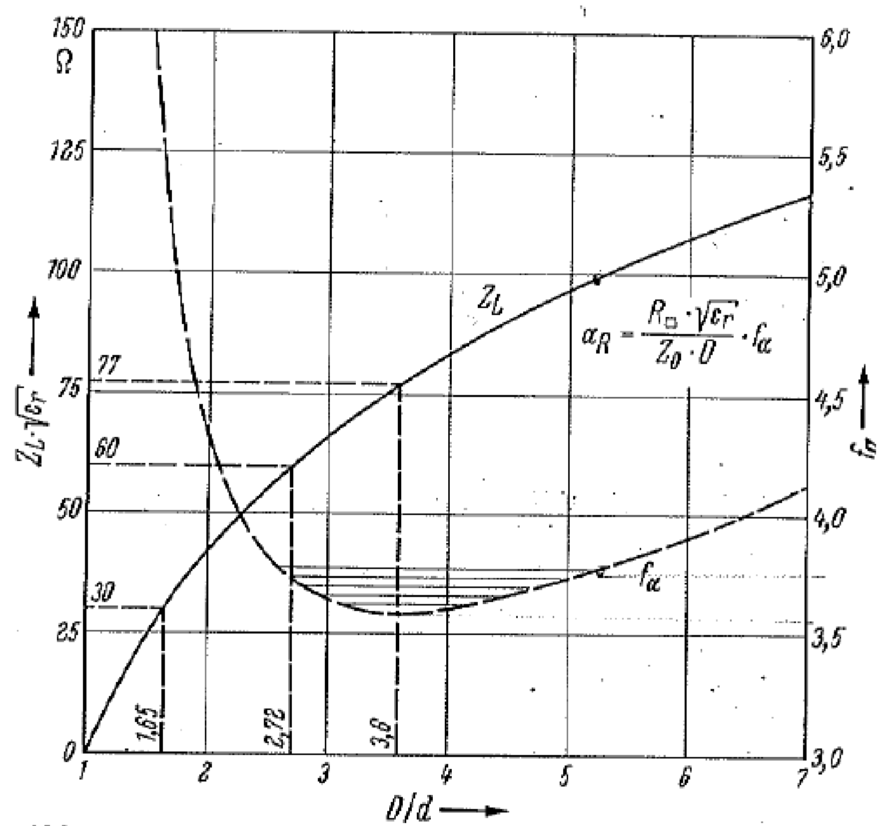


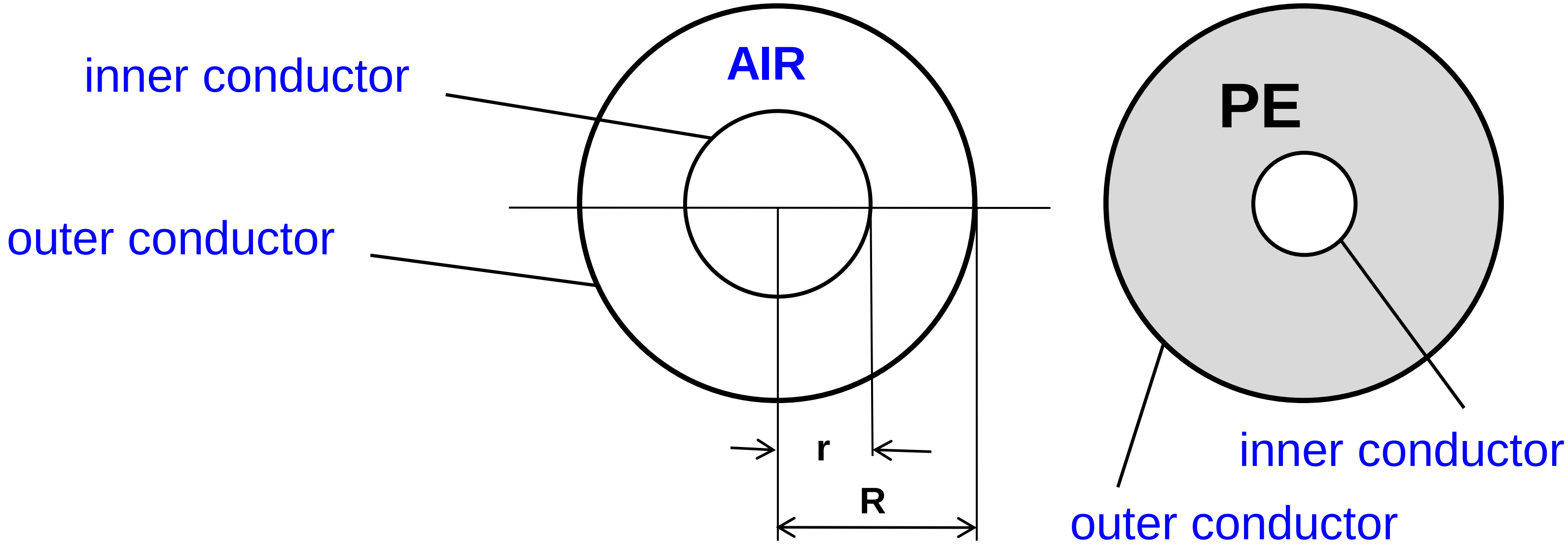
Abb. 4.6/2. Wellenwiderstand und Dämpfung α_R eines Koaxialkabels in Abhängigkeit vom Durchmesserverhältnis D/d. In dem eingezeichneten Toleranzfeld bedeutet eine Linie jeweils 1 % Abweichung vom Optimum

Reprinted from O.Zinke, H.Brunswig,
Lehrbuch der Hochfrequenztechnik, p.222

TEM transmission lines (4)

Applied to 50-Ohm-lines (the impedance mostly used) one finds

	Vacuum or air	Polyethylene (PE)
ε_r	1	2.26
v (m/sec)	$3E8 = c_0$	$0.665\ c_0$
L' (nH/m)	166.7	250.6
C' (pF/m)	66.7	100.2
R/r	2.30	3.50



Transmission lines (1)

- **Coaxial lines**
  - frequency range: 0...10 GHz
  - largest practical size: 350 mm for outer conductor, 150 mm for the inner conductor
  - power rating: for CW operation at 200 MHz: 1 MW
  - low-pass line, upper frequency limit given by moding
  - relatively high attenuation
  - power limited by inner conductor (high field => thermal load)
  - in general easier to handle than waveguides
- **Waveguides**
  - frequency range 0.32...325 GHz (standard guides)
  - largest practical size: 590 mm x 298 mm
  - power rating: 150 MW peak at 310 MHz
  - low attenuation
  - bandpass, low frequency cut-off determined by dimension

Transmission lines (2)

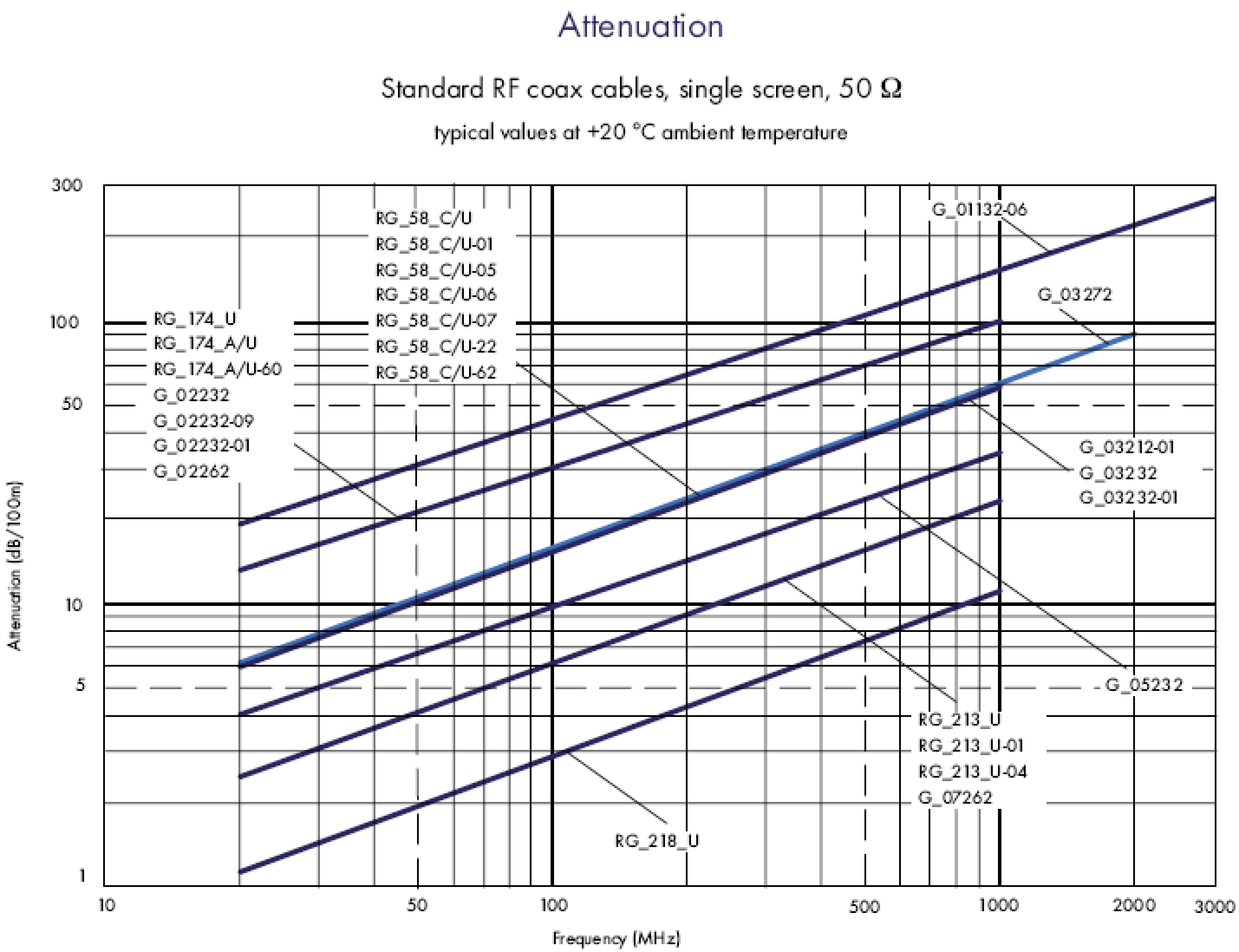
Standard RF coax cables — single screen — 50 Ω

H+S type	Item no.	Curves see page	Center conductor [1] Design	Mat.	Dim. mm	Dielectric [2] Mat.	Dim. mm	Screen 1 [3] Mat.	Dim. mm	Cover %	Screen 2 [4] Dim. mm	Cover %	Jacket [5] Mat.	Dim. mm	Colour	Weight kg/ 100 m	Operating voltage kV	Max. operation frequency	Cable group* crimp	clamp
G_03212-01	22610095		Strand-19	CuSn	0.90	PE	2.95	CuSn	3.60	95	-	-	PUR[1]	4.95	black	3.60	2.5	1	U7	U7
RG_58_C/U	22510015		Strand-19	CuSn	0.90	PE	2.95	CuSn	3.60	96	-	-	PVC2(LM)	4.95	black	3.70	2.5	1	U7	U7
RG_58_C/U-01	22510350		Strand-19	CuSn	0.90	PE	2.95	CuSn	3.60	96	-	-	PVC2(LM)	4.95	black	3.70	2.5	1	U7	U7
RG_58_C/U-05	22511239		Strand-19	CuSn	0.90	PE	2.95	CuSn	3.60	96	-	-	PVC2(LM)	4.95	blue	3.70	2.5	1	U7	U7
RG_58_C/U-06	22510017		Strand-19	CuSn	0.90	PE	2.95	CuSn	3.60	96	-	-	PVC	4.95	black	3.70	2.5	1	U7	U7
RG_58_C/U-07	22511244		Strand-19	CuSn	0.90	PE	2.95	CuSn	3.60	96	-	-	PVC2(LM)	4.95	grey	3.70	2.5	1	U7	U7
RG_58_C/U-22	22511607		Strand-19	CuSn	0.90	PE	2.95	CuSn	3.60	96	-	-	PVC2(LM)	4.95	red	3.70	2.5	1	U7	U7
RG_58_C/U-62[b]	23024284		Strand-19	CuAg	0.90	PE	2.95	CuAg	3.60	96	-	-	PVC(UL)	4.95	black	3.70	2.5	1	U7	U7
G_03232	22510128	5 and 6	Strand-7	Cu	0.95	PE	2.95	Cu	3.60	95	-	-	PVC	5.00	black	3.70	2.5	1	U7	U7
G_03262-1	22512108		Strand-7	CuSn	0.90	PE	2.95	CuSn	3.60	96	-	-	LSFH[1]	4.95	black	3.90	2.5	1	U7	U7
G_03272	22511434		Strand-7	Cu	0.95	PE	2.95	Cu	3.60	95	-	-	PE[1]	5.00	black	3.50	2.5	2	U7	U7
G_05232	22510176		Strand-7	Cu	1.50	PE	4.80	Cu	5.60	92	-	-	PVC2(LM)	7.40	black	7.70	3.5	1	-	U19
RG_213_U	22510052		Strand-7	Cu	2.25	PE	7.25	Cu	8.10	96	-	-	PVC2(LM)	10.30	black	15.30	5.0	1	U29	U28
RG_213_U-01[a]	22510053		Strand-7	Cu	2.25	PE	7.24	Cu	8.10	96	-	-	PVC2(LM)	10.30	black	15.30	5.0	1	U29	U28
RG_213_U-04	22510055		Strand-7	Cu	2.25	PE	7.25	Cu	8.10	96	-	-	PVC	10.30	black	15.30	5.0	1	U29	U28
G_07262	22511836		Strand-7	Cu	2.25	PE	7.28	Cu	8.10	96	-	-	LSFH[1]	10.30	black	15.30	5.0	1	U29	U28
RG_218_U	22510066		Wire	Cu	5.00	PE	17.30	Cu	18.40	96	-	-	PVC2(LM)	22.10	black	66.90	11.0	1	-	U44

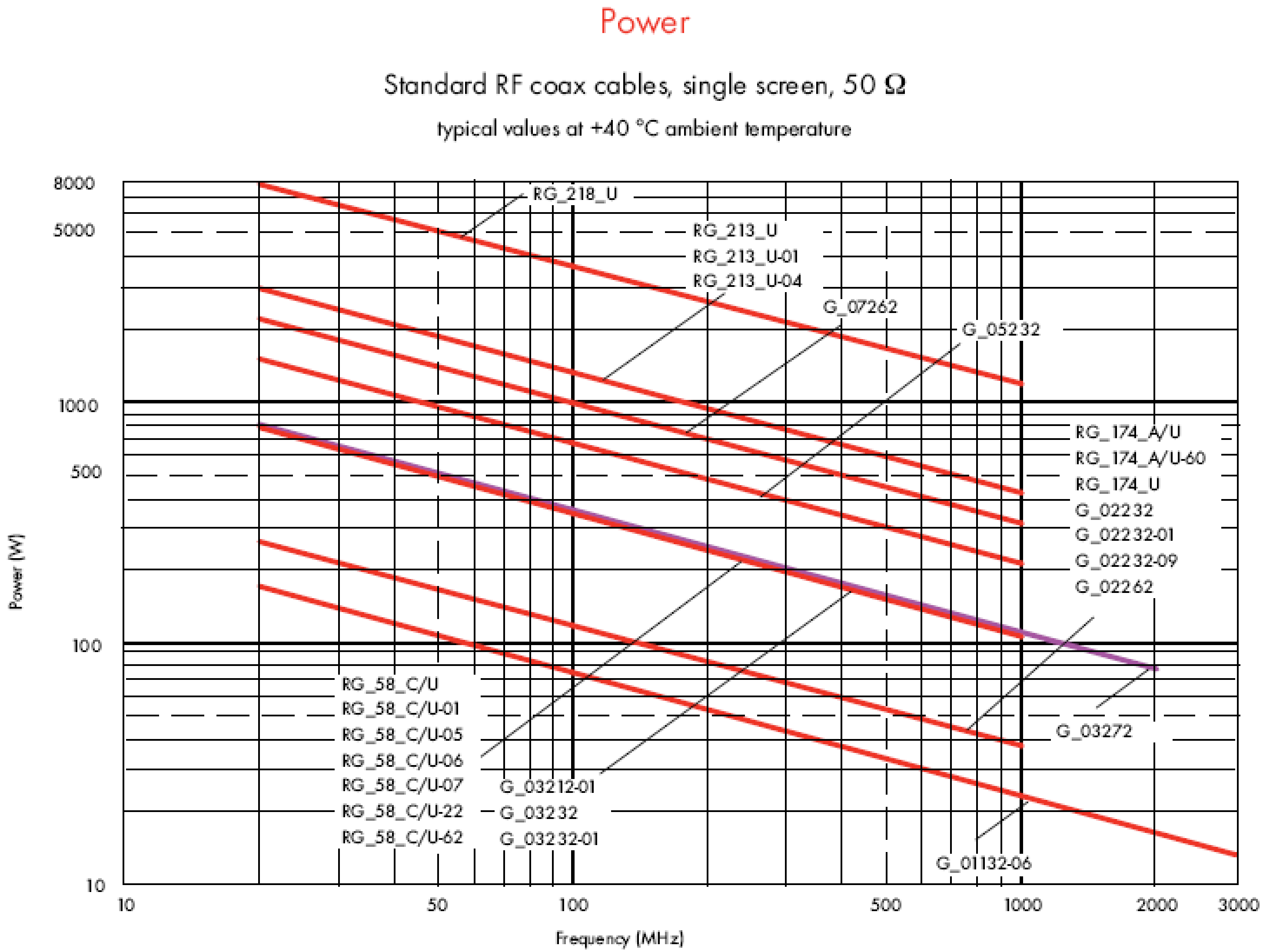
* for suitable connectors

[a] precision type: impedance 50 ± 1 Ω
[b] UL recognised (see UL types page 117)
[1] Low Smoke Free of Halogen (LSFH) acc. waste electrical and electronic equipment (WEEE) and restriction of the use of certain hazardous substances (RoHS) directive.

Transmission lines (3)



Transmission lines (4)



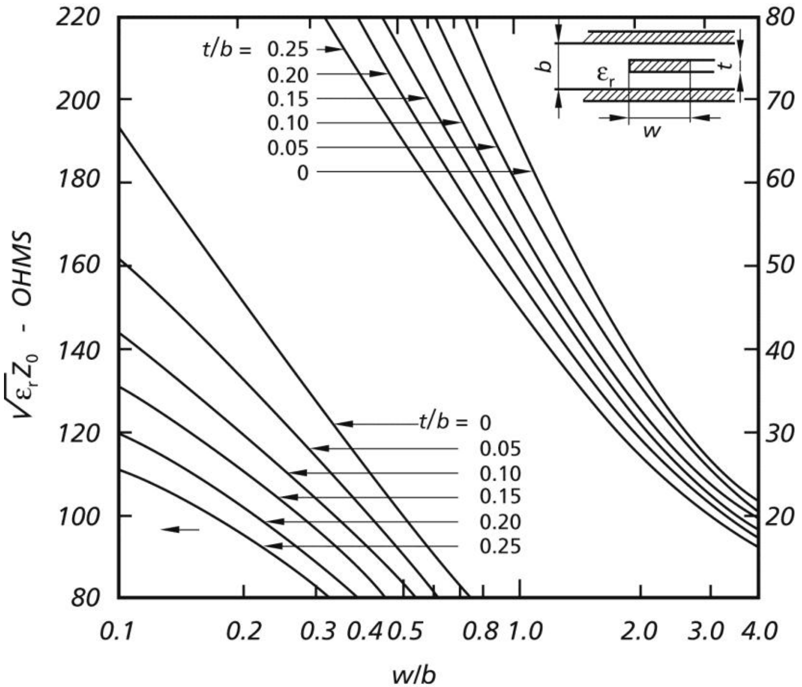
Striplines (1)

A stripline is a **flat conductor** between a top and bottom ground plane. The space around this conductor is filled with a homogeneous dielectric material. This line propagates a pure TEM mode. With the static capacity per unit length, **C'**, the static inductance per unit length, **L'**, the relative permittivity of the dielectric, ε_r and the speed of light **c** the characteristic impedance Z_0 of the line is given by

$$Z_0 = \sqrt{\frac{L'}{C'}}$$

$$v_{ph} = \frac{c}{\sqrt{\varepsilon_r}} = \frac{1}{\sqrt{L'C'}}$$

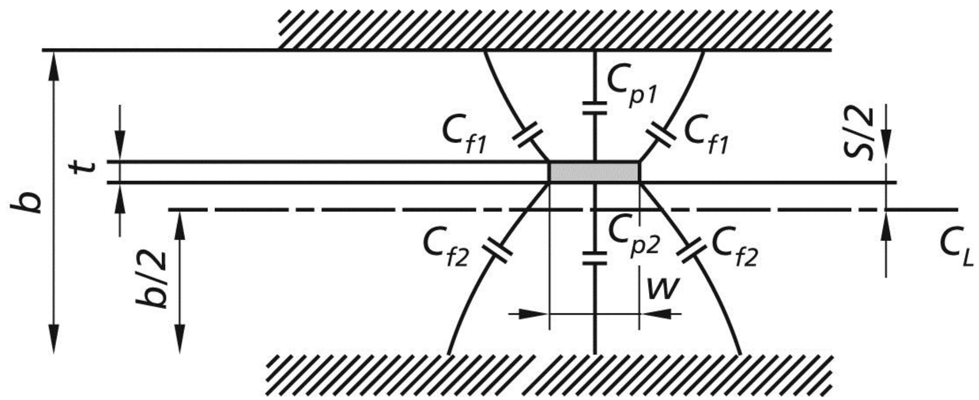
$$Z_0 = \sqrt{\varepsilon_r}\,\frac{1}{C'c}$$



Characteristic impedance of striplines

Striplines (2)

For a mathematical treatment, the effect of the fringing fields may be described in terms of static capacities. The total capacity is the sum of the principal and fringe capacities Cp and Cf.



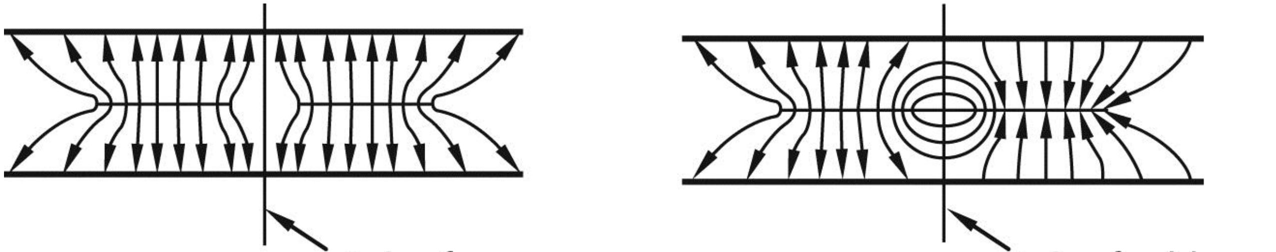
$$C_{tot} = C_{p1} + C_{p2} + 2C_{f1} + 2C_{f2}$$

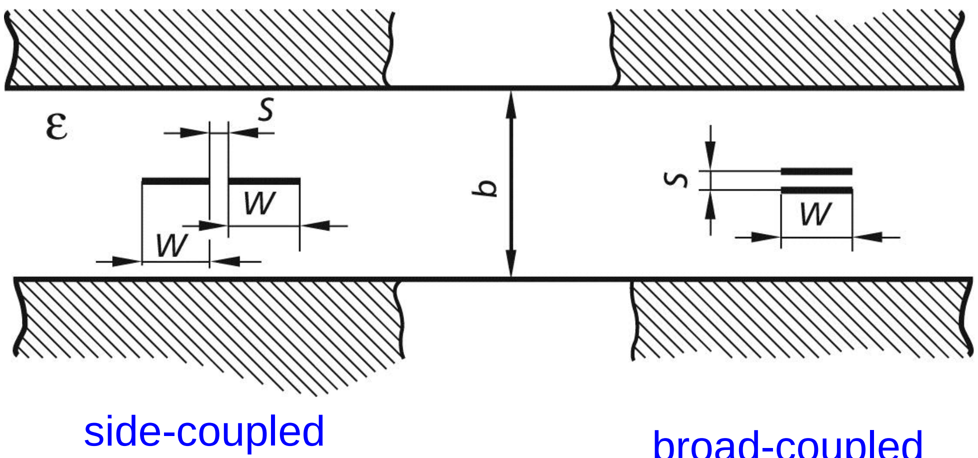
Cf stands for fringe field capacity,

Cp stands for principal capacity

Striplines (3)

Coupled striplines (in odd and even mode):





$$Z_{0,even} = \frac{1}{\sqrt{\varepsilon_r}} \cdot \frac{94.15\,\Omega}{\frac{w}{b} + \frac{\ln 2}{\pi} + \frac{1}{\pi}\ln\left(1 + \tanh\left(\frac{\pi}{2}\frac{s}{b}\right)\right)}$$

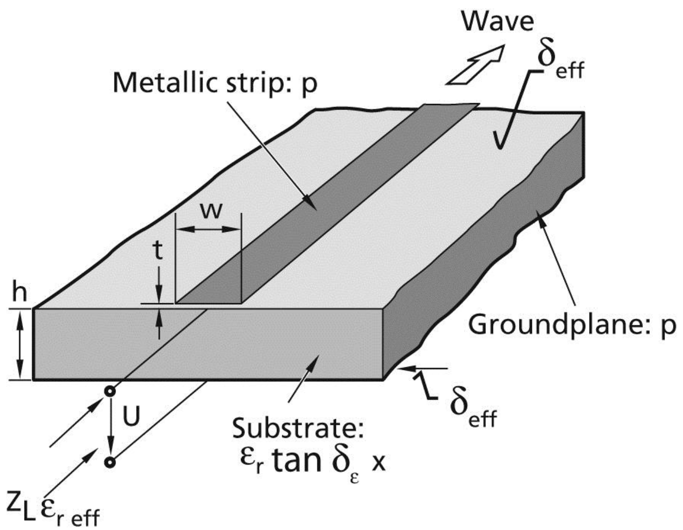
$$Z_{0,odd} = \frac{1}{\sqrt{\varepsilon_r}} \cdot \frac{94.15\,\Omega}{\frac{w}{b} + \frac{\ln 2}{\pi} + \frac{1}{\pi}\ln\left(1 + \coth\left(\frac{\pi}{2}\frac{s}{b}\right)\right)}$$

This formula for side-coupled structure only.

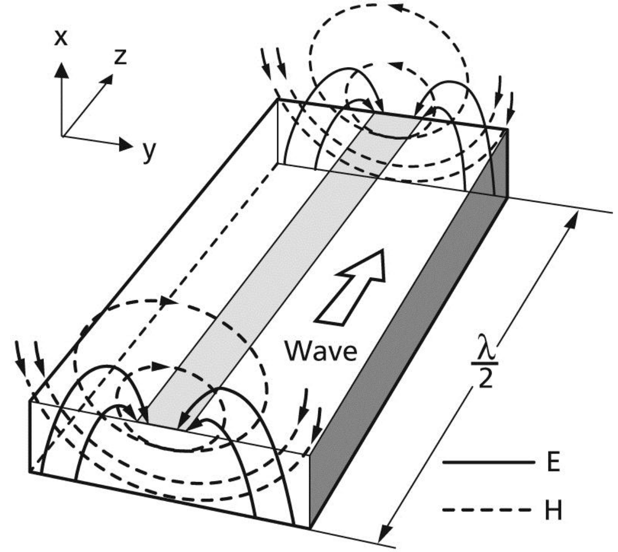
Microstriplines (1)

A microstripline may be visualized as a stripline with the top cover and the top dielectric layer taken away. It is thus an asymmetric open structure, and only part of its cross section is filled with a dielectric material. Since there is a transversely inhomogeneous dielectric, only a quasi-TEM wave exists. This has several implications such as a frequency-dependent characteristic impedance and a considerable dispersion.

(a) Mechanical construction



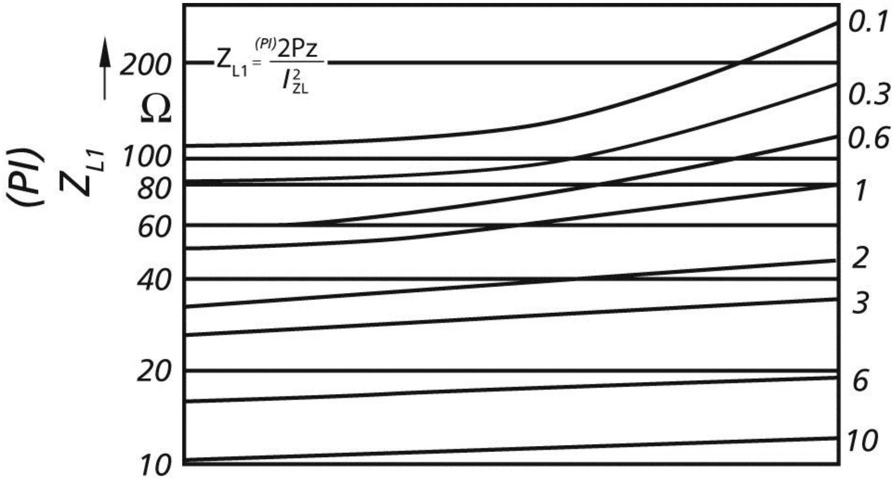
(b) Static field approximation



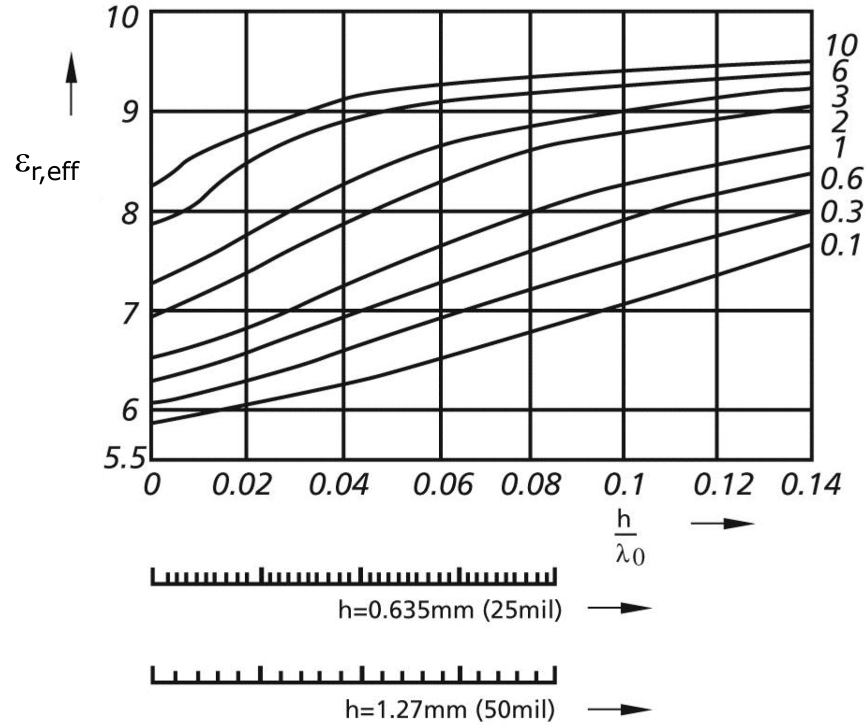
Note: Quasi-TEM wave due to different dielectric constants in different parts of the cross-section. We do get longitudinal field components.

Microstriplines (2)

Frequency-dependent characteristic impedance

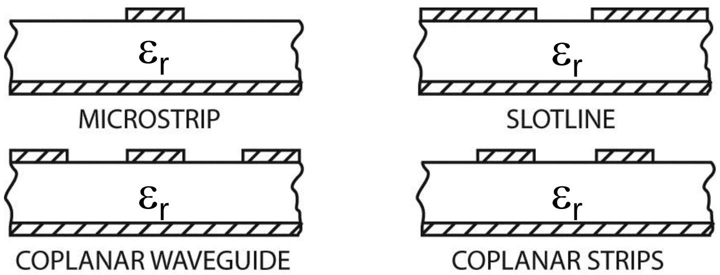


Effective permittivity

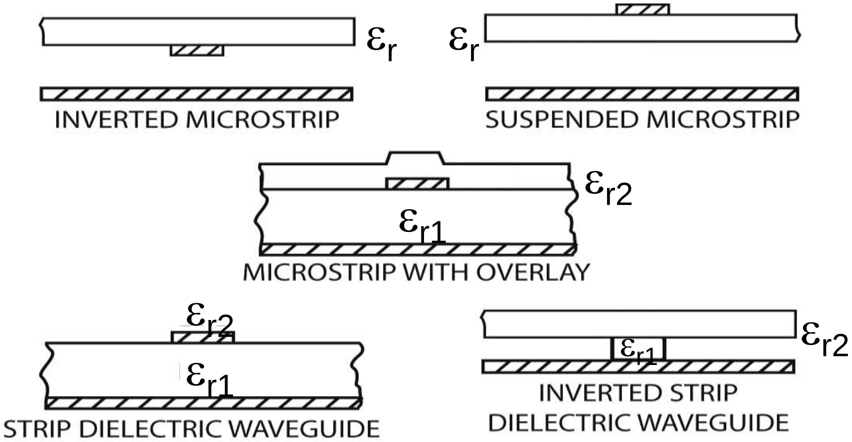


Microstriplines (3)

Planar transmission lines used in MIC (microwave integrated circuits)



Various transmission lines derived from microstrip



Slotlines (1)

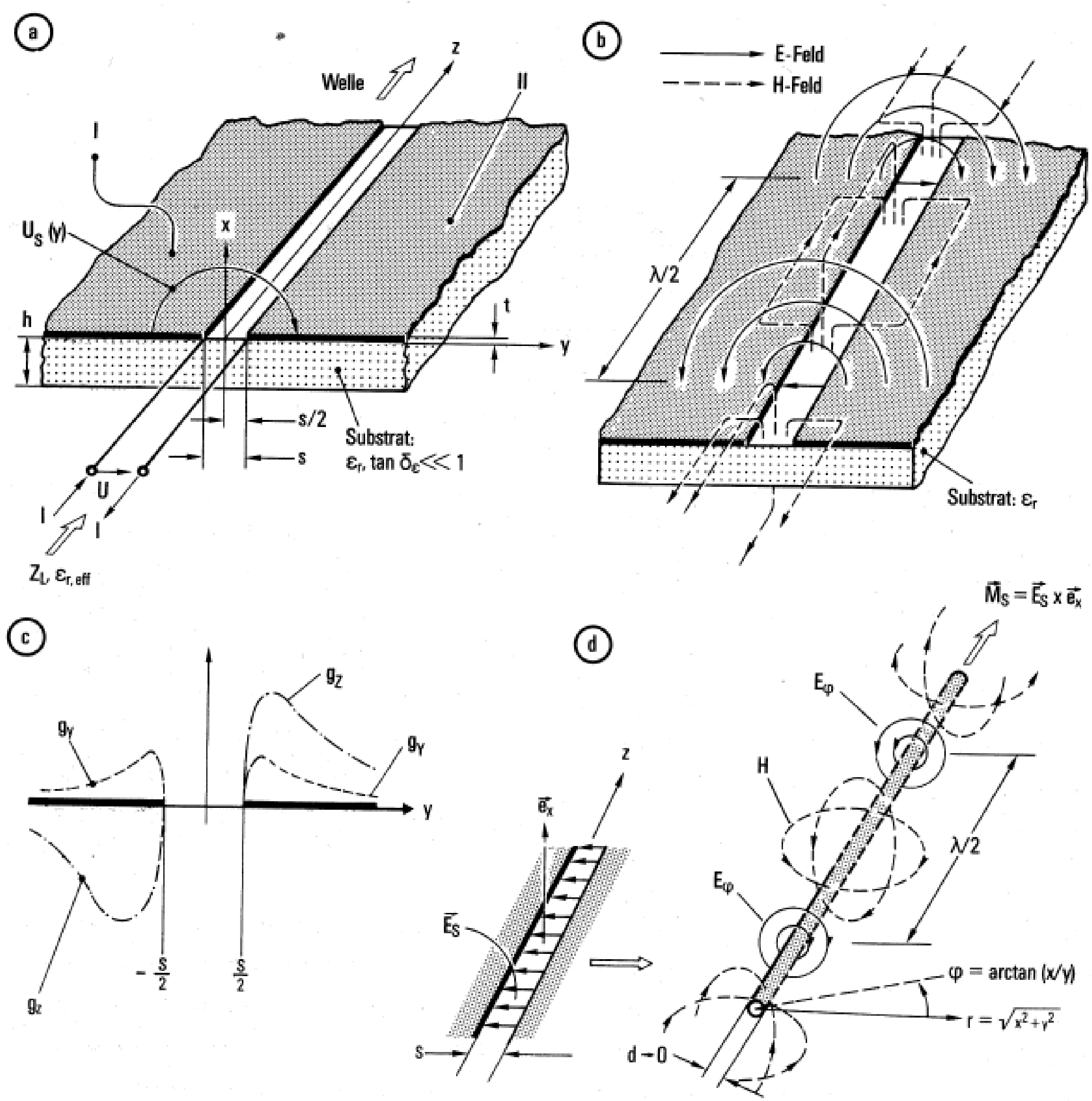
The slotline may be considered as the dual structure of the microstrip. It is essentially a slot in the metallization of a dielectric substrate. The characteristic impedance and the effective dielectric constant exhibit similar dispersion properties to those of the microstrip line.

(a) Mechanical construction

(b) Field pattern
(TE approximation)

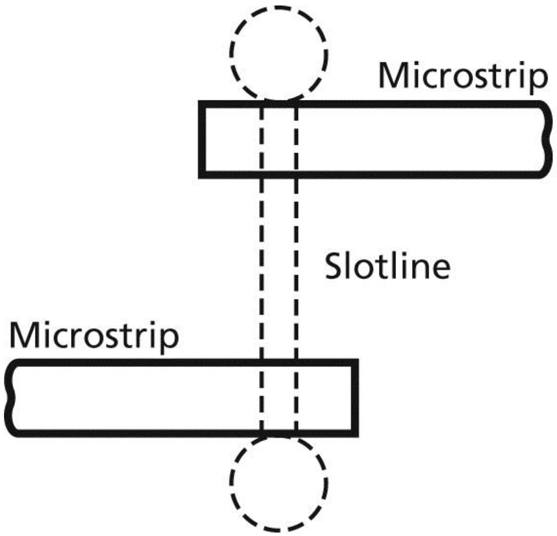
(c) Longitudinal and transverse current densities

(d) Magnetic line current model.



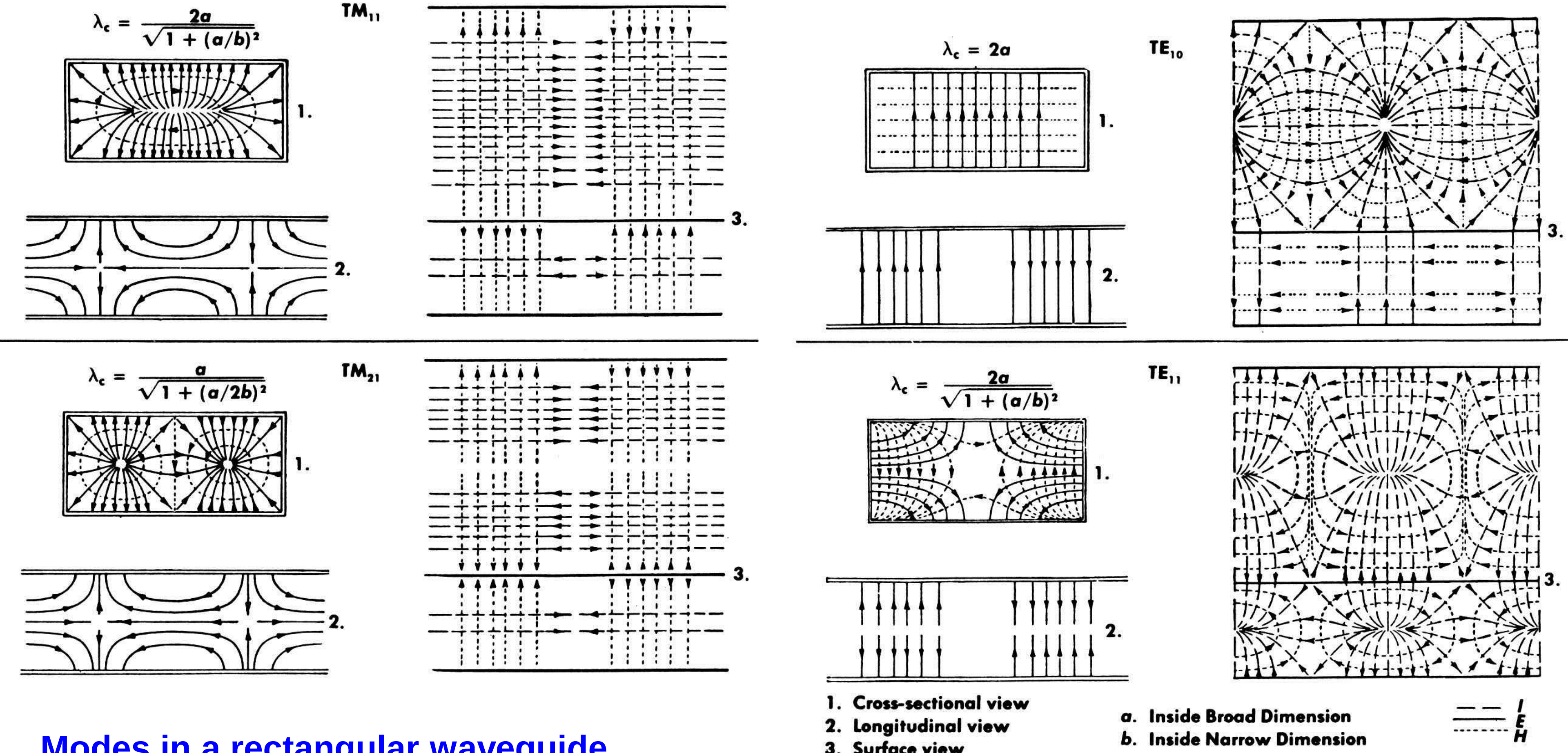
Slotlines (2)

A broadband (decade bandwidth) pulse inverter. Assuming the upper microstrip to be the input, the signal leaving the circuit on the lower microstrip is inverted since this microstrip ends on the opposite side of the slotline compared to the input.



Two microstrip-slotline transitions connected back to back for 180° phase change.

Rectangular waveguide



Modes in a rectangular waveguide with dimensions a and b.
solid lines: E field, dotted lines: H field

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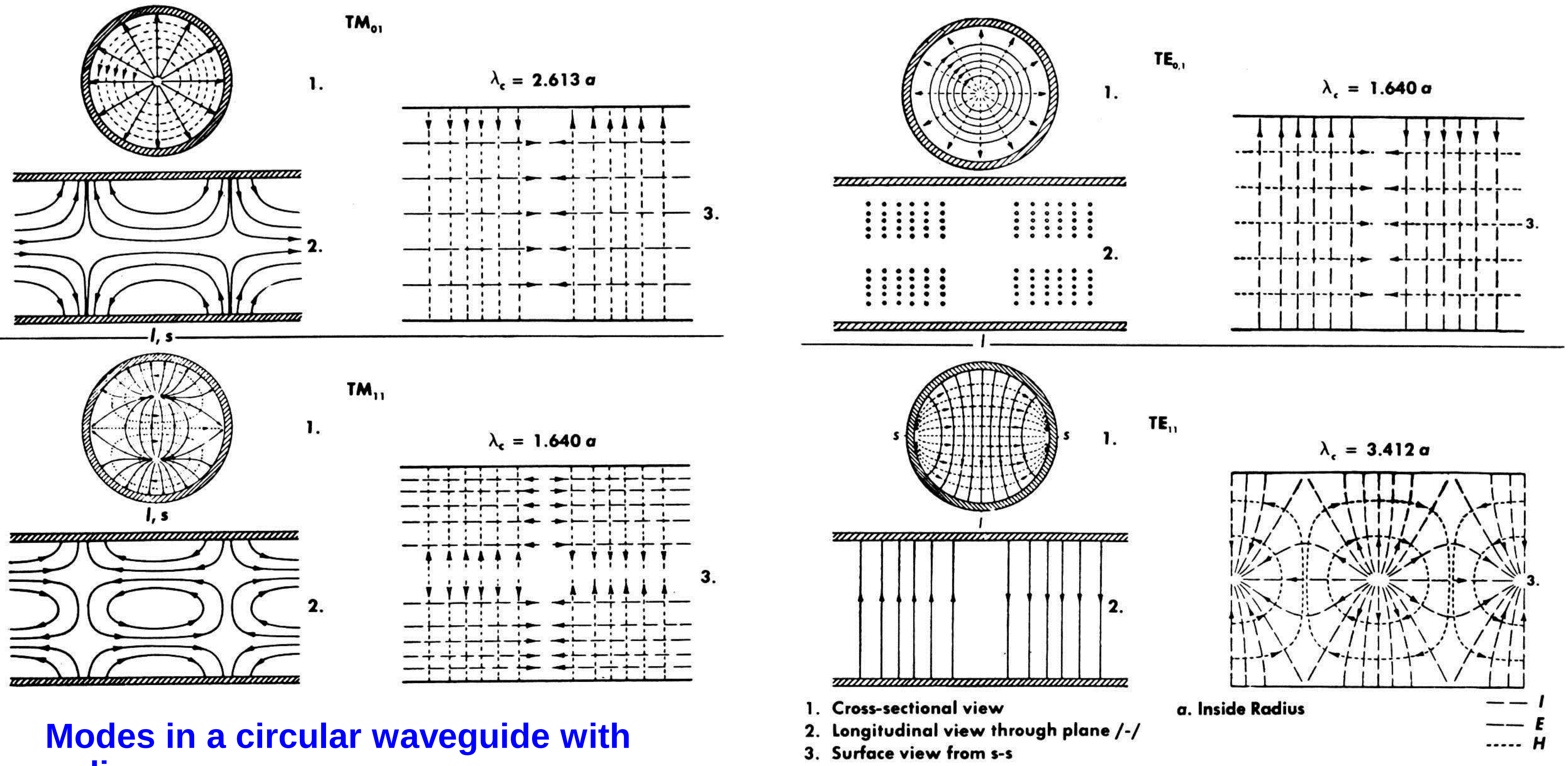
Circular waveguide



Modes in a circular waveguide with radius a
solid lines: E field, dotted lines: H field
Please note the similarity to the pillbox cavity!

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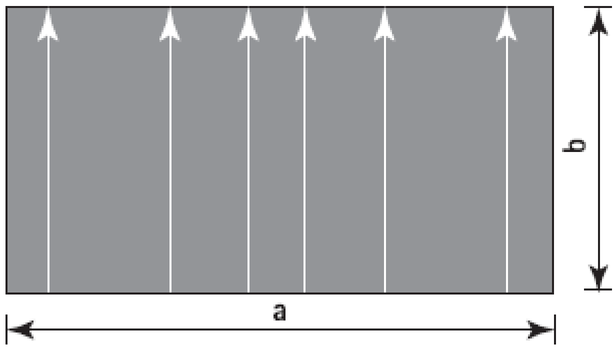
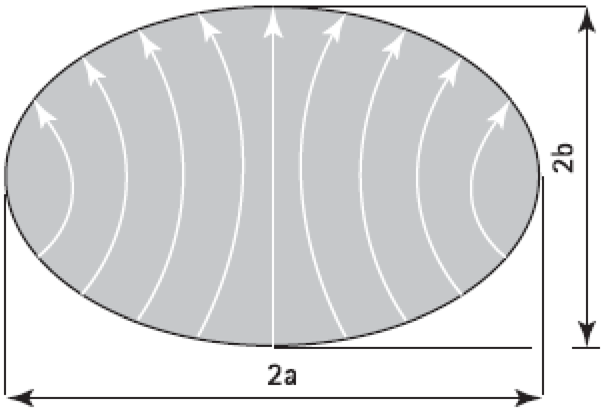
Elliptical waveguide (1)

E field lines for TE_{10} mode



E field lines for TE_{c11} mode



The small index c stands for the polarization and refers to sine (s) or cosine (c).

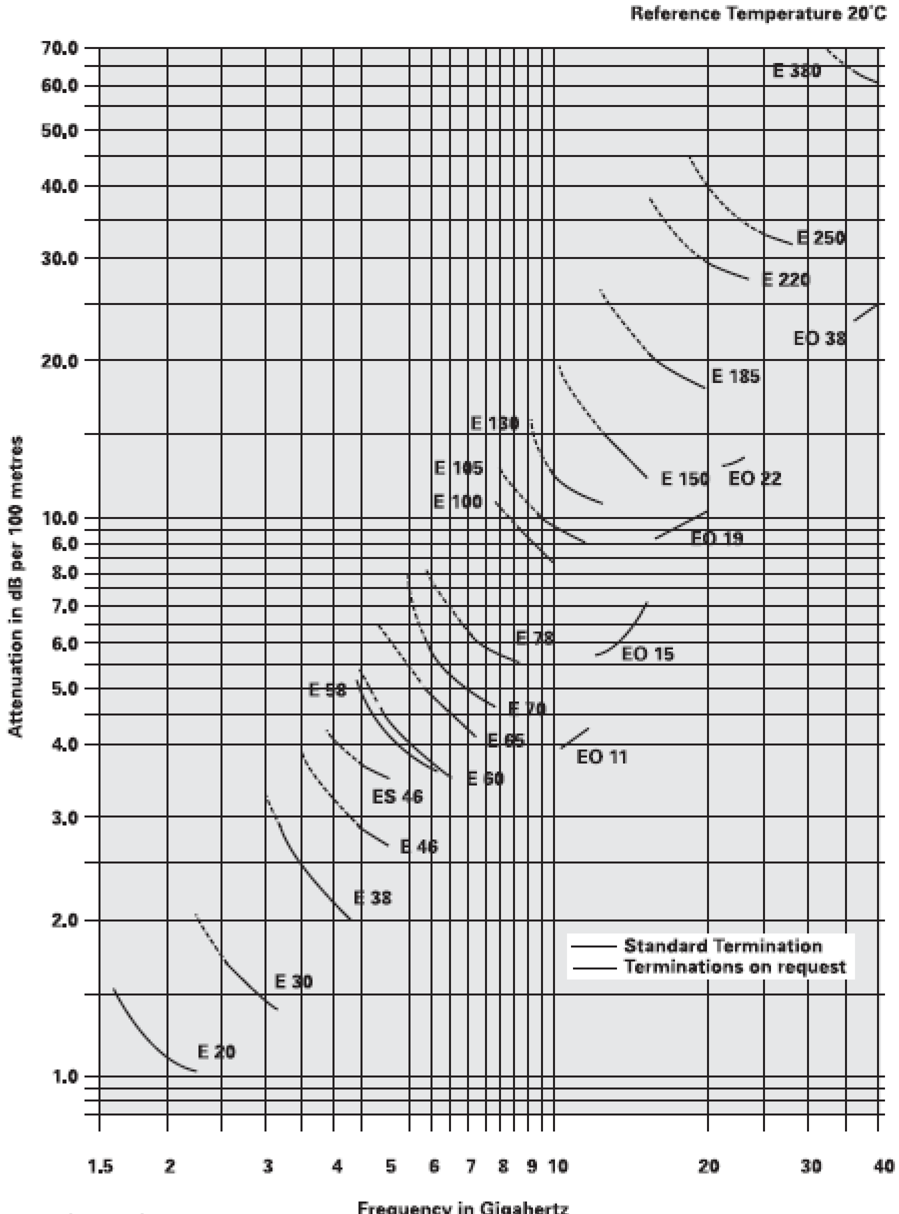
Reprinted from Flexwell, RFS Datasheet *Elliptical Waveguides*

The cut-off wavelengths of the various modes that can propagate in an elliptical waveguide can be found analytically using rather complicated methods or numerically.

Elliptical waveguide (2)

Typical attenuation values for flexible elliptical waveguides:

Rigid rectangular cross-section waveguides are rather seldom used in industry.



EO stands for **overmoded waveguide**

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Elliptical waveguide (3)

Datasheet

WVG. TYPE	OPER. FREQ. GHz	CUT OFF FREQ. GHz	MAX. VSWR/ RETURN LOSS, dB	ATTENUATION dB/100m (ft) IN THE OPERATING FREQUENCY BAND			AVG. POWER kW MID BAND	GROUP VELOCITY %c MID BAND	GROUP DELAY ns/100m (ft) MID BAND
				LOW BAND	MID BAND	HIGH BAND			
E30	2.7 - 3.1	1.8	1.128/24.4	1.61 (0.49)	1.49 (0.45)	1.4 (0.43)	30.37	78.4	425.4 (129.7)
E38	3.6 - 4.2	2.4	1.15/23.1	2.37 (0.72)	2.20 (0.67)	2.08 (0.63)	16.27	78.8	423.2 (129.0)
EP38	3.6 - 4.2	2.4	1.083/28.0	2.37 (0.72)	2.20 (0.67)	2.08 (0.63)	16.27	78.8	423.2 (129.0)
E46	4.4 - 5.0	2.88	1.15/23.1	2.92 (0.89)	2.80 (0.85)	2.73 (0.83)	10.93	79.0	422.1 (128.7)
EP46	4.4 - 5.0	2.88	1.083/28.0	2.92 (0.89)	2.80 (0.85)	2.73 (0.83)	10.93	79.0	422.1 (128.7)
ES46	4.4 - 5.0	3.08	1.15/23.1	3.69 (1.12)	3.55 (1.08)	3.49 (1.06)	8.39	75.5	441.6 (134.6)
ESP46	4.4 - 5.0	3.08	1.073/29.1	3.69 (1.12)	3.55 (1.08)	3.49 (1.06)	8.39	75.5	441.6 (134.6)
EP58	4.4 - 6.2	3.56	1.083/28.0	5.10 (1.55)	3.96 (1.21)	3.60 (1.10)	6.54	74.1	450.3 (137.2)
E60	5.6 - 6.425	3.65	1.15/23.1	4.15 (1.27)	3.95 (1.20)	3.80 (1.16)	7.24	79.4	420.3 (128.1)
EP60	5.6 - 6.425	3.65	1.062/30.5	4.15 (1.27)	3.95 (1.20)	3.80 (1.16)	7.24	79.4	420.3 (128.1)
E65	5.9 - 7.125	4.01	1.15/23.1	4.9 (1.50)	4.5 (1.37)	4.25 (1.30)	5.26	78.7	423.8 (129.2)
EP65	5.9 - 7.125	4.01	1.062/30.5	4.9 (1.50)	4.5 (1.37)	4.25 (1.30)	5.26	78.7	423.8 (129.2)
EP70	6.4 - 7.75	4.34	1.062/30.5	5.5 (1.68)	5.0 (1.52)	4.8 (1.46)	4.65	79.1	421.5 (128.5)
E78	7.1 - 8.5	4.72	1.15/23.1	6.2 (1.89)	5.8 (1.77)	5.6 (1.71)	3.67	79.6	419.0 (127.7)
EP78	7.1 - 8.5	4.72	1.062/30.5	6.2 (1.89)	5.8 (1.77)	5.6 (1.71)	3.67	79.6	419.0 (127.7)
EP100	9.0 - 10.0	6.43	1.105/26.0	9.5 (2.90)	8.9 (2.71)	8.4 (2.56)	1.91	73.6	453.1 (138.1)
E105	10.0 - 11.7	6.49	1.15/23.1	9.6 (2.92)	9.2 (2.79)	8.9 (2.71)	1.77	79.9	417.3 (127.2)
EP105	10.0 - 11.7	6.49	1.062/30.5	9.6 (2.92)	9.2 (2.79)	8.9 (2.71)	1.77	79.9	417.3 (127.2)
E130	10.7 - 13.25	7.43	1.15/23.1	12.6 (3.84)	11.5 (3.52)	11.1 (3.39)	1.22	78.5	424.8 (129.5)
EP130	10.7 - 13.25	7.43	1.083/28.0	12.6 (3.84)	11.5 (3.52)	11.1 (3.39)	1.22	78.5	424.8 (129.5)
E150	13.4 - 15.35	8.64	1.15/23.1	14.6 (4.44)	14.0 (4.26)	13.7 (4.16)	0.88	79.7	418.6 (127.6)
EP150	13.4 - 15.35	8.64	1.083/28.0	14.6 (4.44)	14.0 (4.26)	13.7 (4.16)	0.88	79.7	418.6 (127.6)
E185	17.3 - 19.7	11.06	1.15/23.1	20.3 (6.17)	19.4 (5.92)	18.9 (5.75)	0.51	80.2	416.1 (126.8)
EP185	17.3 - 19.7	11.06	1.083/28.0	20.3 (6.17)	19.4 (5.92)	18.9 (5.75)	0.51	80.2	416.1 (126.8)
E220	21.2 - 23.6	13.36	1.105/26.0	28.8 (8.77)	28.3 (8.63)	28.1 (8.56)	0.31	80.3	415.6 (126.7)
E250	24.25 - 26.5	15.06	1.15/23.1	33.2 (10.1)	32.4 (9.88)	32.0 (9.75)	0.31	80.5	414.2 (126.3)
E300	27.5 - 33.4	19.05	1.15/23.1	50.0 (15.2)	46.0 (14.0)	44.4 (13.5)	0.14	78.1	427.1 (130.2)
E380	37.0 - 39.5	23.45	1.15/23.1	61.3 (18.7)	60.7 (18.5)	60.0 (18.3)	0.09	79.1	421.9 (128.6)

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Amplifiers (1)

- **Semiconductors**
  - Bipolar transistors
  - Field effect transistors
  - many others
- Frequency range: 0...100 GHz
- Power range: from close to thermal noise level to many kW
- High reliability, but lifetime not infinite (thermal fatigue, metal migration, etc.)
- Often unforgiving, failure is normally definitive
- Inherently low-voltage, high current devices compared to tubes
- Low to medium gain

Amplifiers (2)

- **Gridded Tubes (electron tubes)**
- Frequency range: 0...0.5 GHz (tetrodes), 0...3 GHz (triodes)
- Power range:
  - for CW (continuous wave) up to 30 MHz: 1 MW
  - at 300 MHz: 200 kW
  - pulsed at 200 MHz: 4 MW
- Medium reliability, lifetime cathode limited to 5000...40000 hours
- Relatively robust
- Inherently medium to high voltage, low current devices
- Density modulated
- High gain at low frequencies, medium gain at high frequencies

Amplifiers (3)

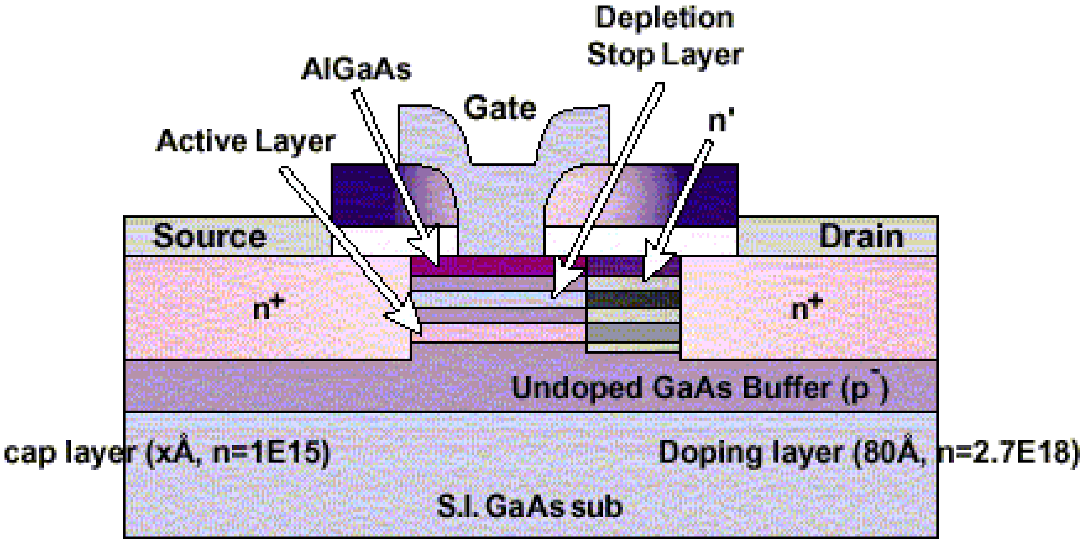
- **Klystrons**
- Frequency range: 0.3...10 GHz
- Power range:
  - CW at 350 MHz: 1 MW
  - pulsed at 3 GHz: 30 MW
- Medium reliability, lifetime cathode limited
- Needs expert care
- Inherently very high voltage device
- Velocity modulated
- Very high gain (≈40 to 60 dB, about 10 dB per passive resonator)
- Tend to be noisy (acoustically and electrically)

- **Others**
- Travelling wave tubes, magnetrons (Microwave ovens!!), Gyrotrons
- 2-beam accelerators (CLIC)

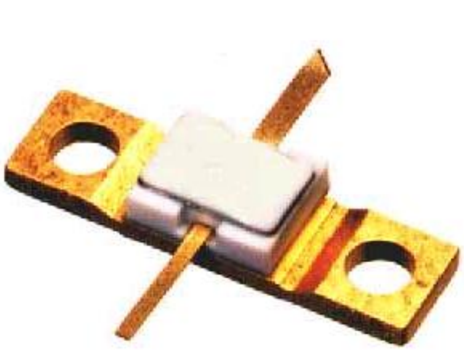
Transistors (1)

Example: a field effect transitor (FET)

Structure of an advanced
pulse-doped MESFET



High Power and Low Distortion
GaAs FET



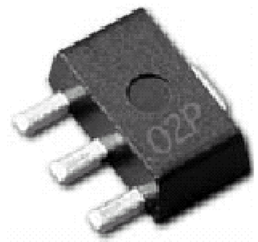
Transistors (2)

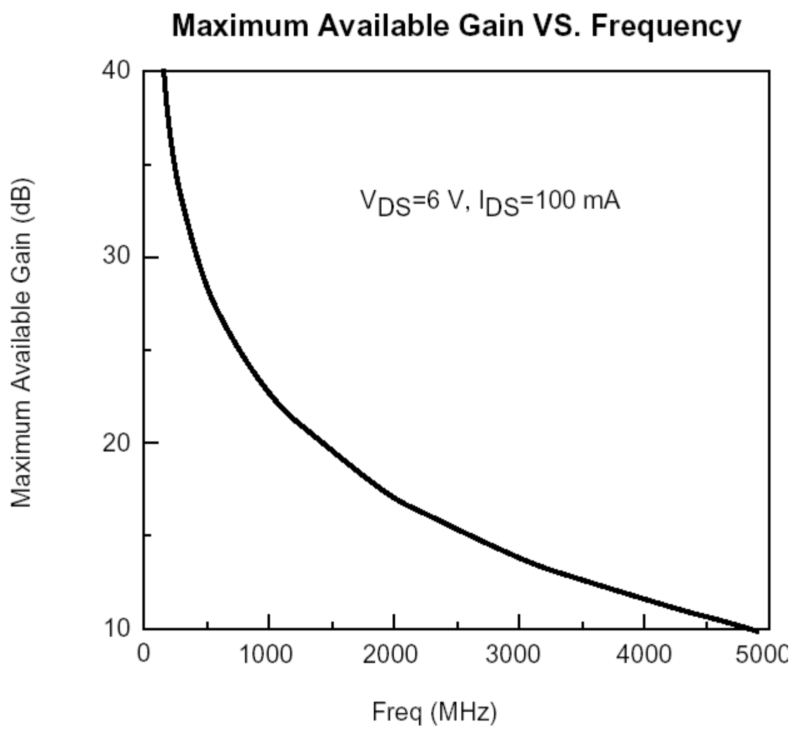
A typical data sheet of a Medium Power GaAs FET

- Up to 2.5 GHz frequency band
- Beyond 22 dBm output power
- Low distortion characteristics
- Low power consumption
- High power gain
- Low-cost plastic mold package
- Low thermal resistance lead

Applications

- Driver amplifier preceding final power amplifier for DECT

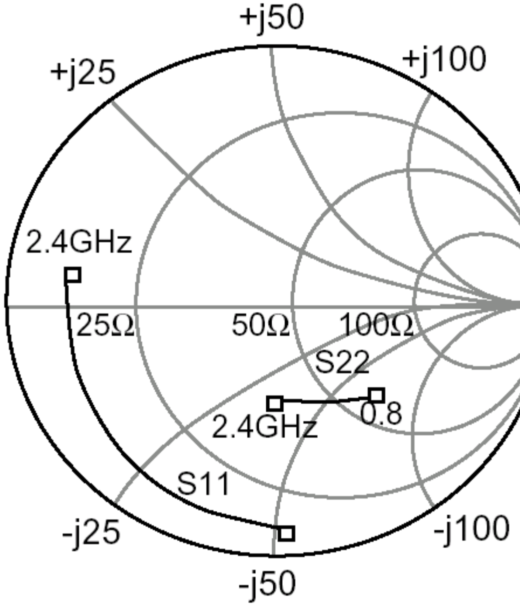




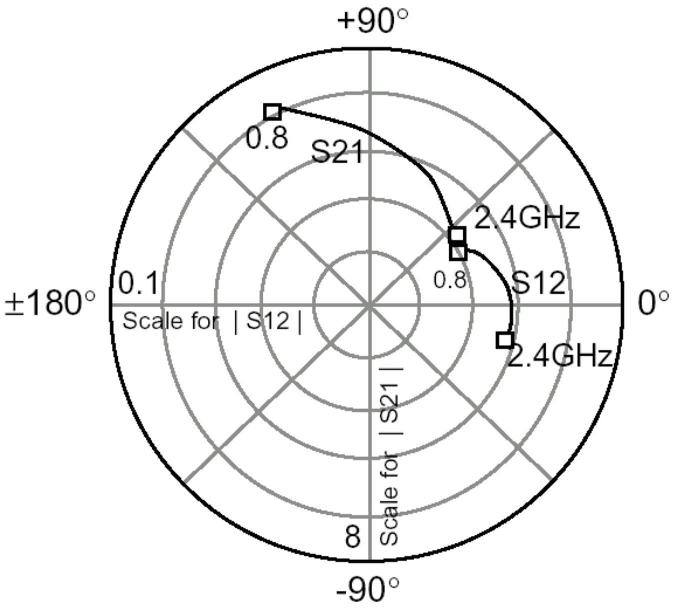
Transistors (3)

Transistor scattering parameters
They will be covered in detail in the second part of this lecture...

The input and output reflection S_{11} and S_{22}

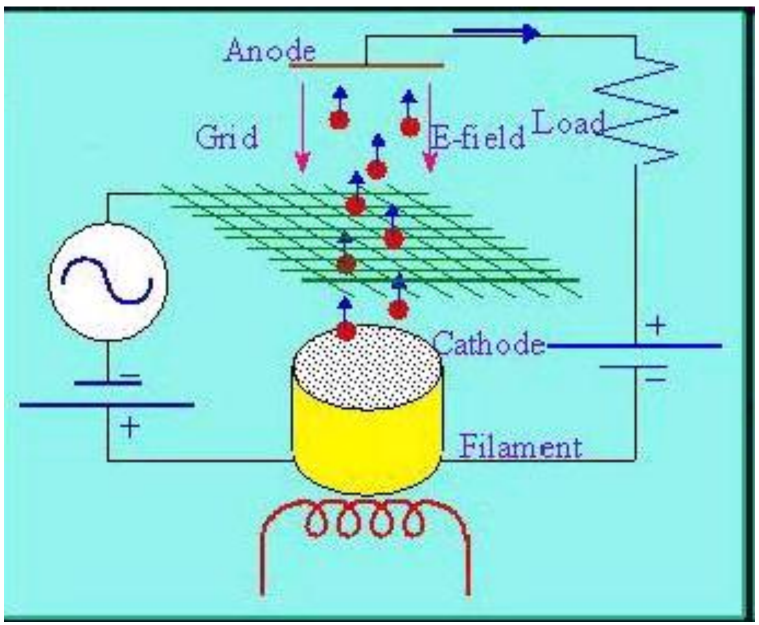


The forward transmission S_{21} and the backward transmission S_{12}

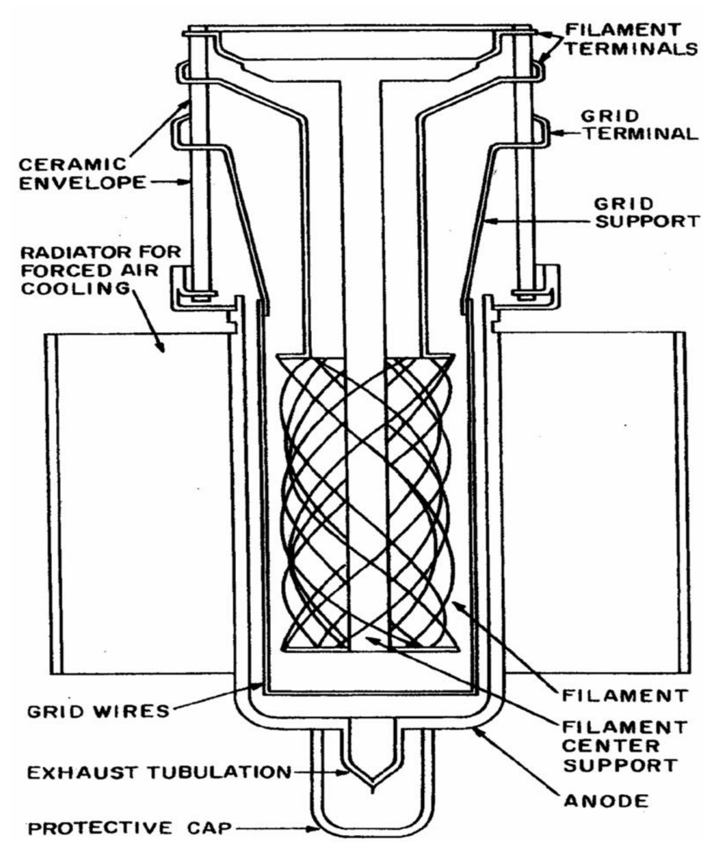


Gridded tubes

- Filament burns off electrons
- acceleration in DC field
- density modulation by grid
- => voltage controlled current source



A medium power external anode transmitting tube



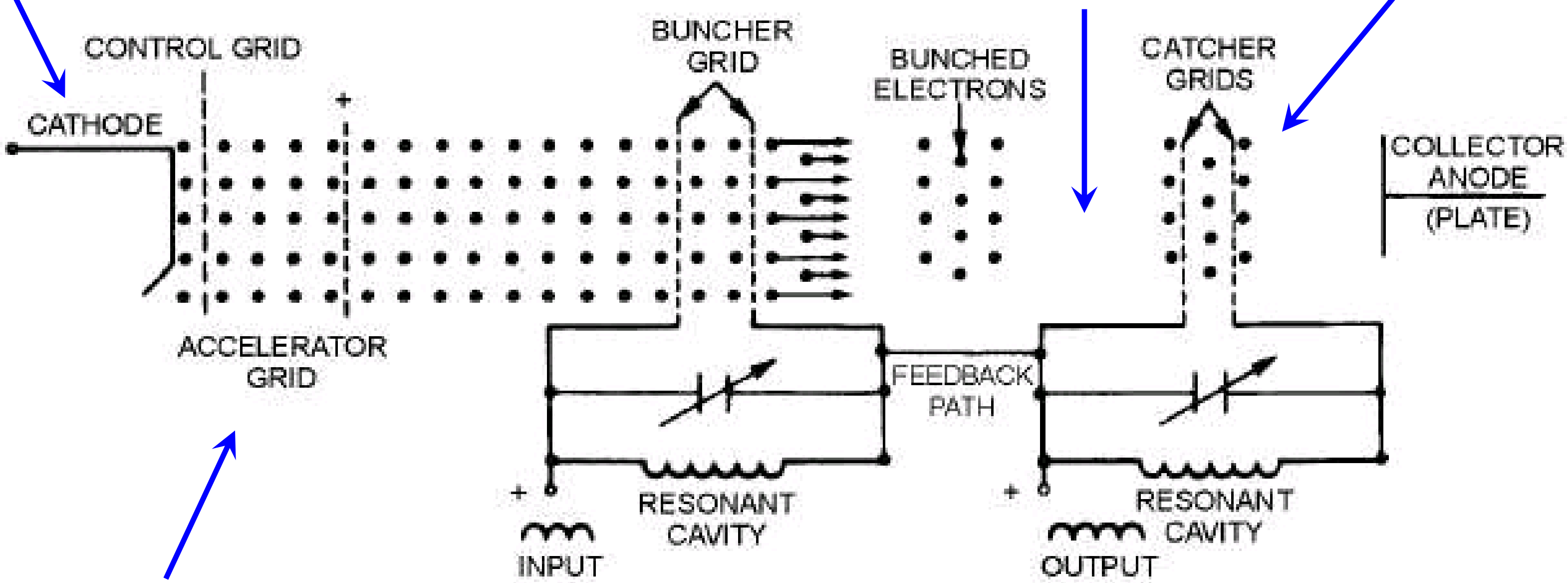
Klystrons (1)

Filament (electron gun) emits electrons

velocity modulation in first cavity

extraction of high intensity RF field in second cavity

drift space



acceleration in DC field

RF input

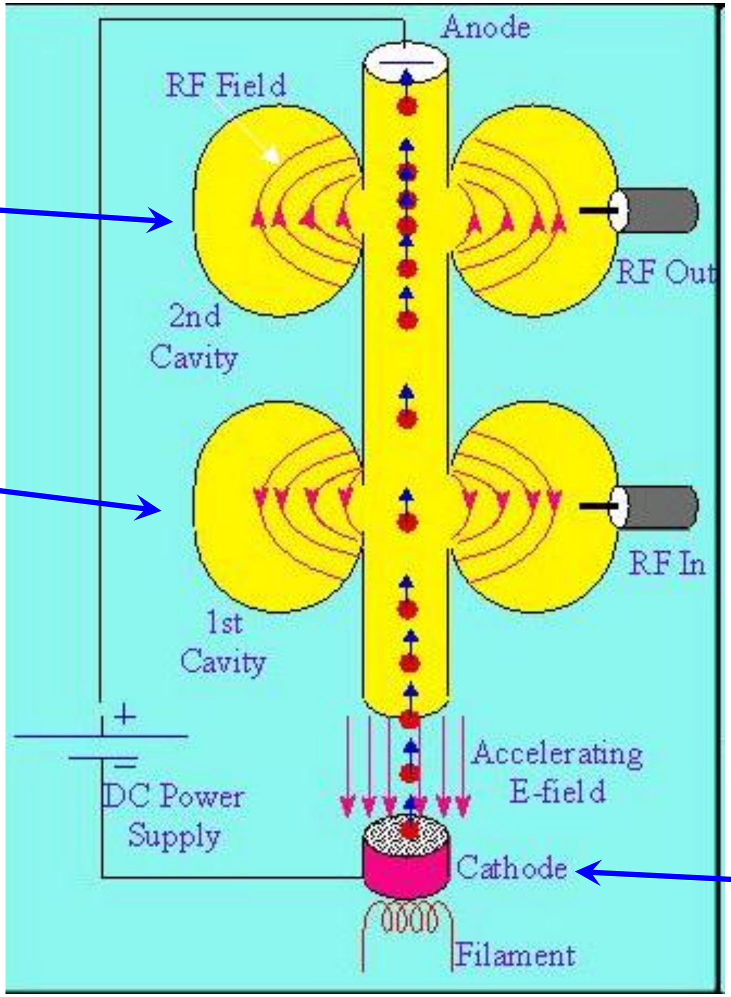
RF output

Klystrons (2)

extraction of high intensity RF field in second cavity

bunching cavity

accelerating DC field



RF output

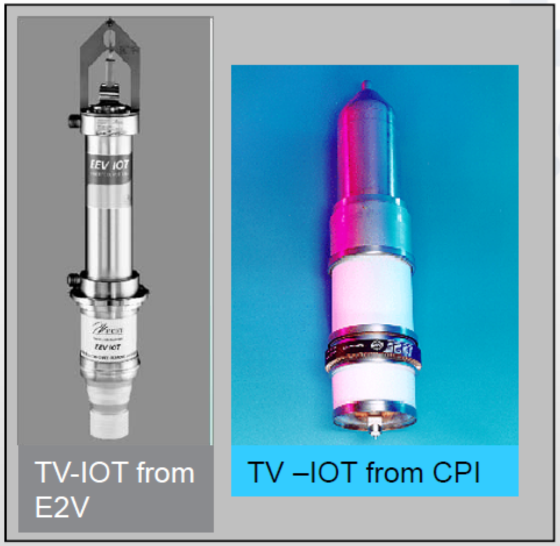
RF input

electron source (heated)

IOT – Inductive Output Tubes (1)

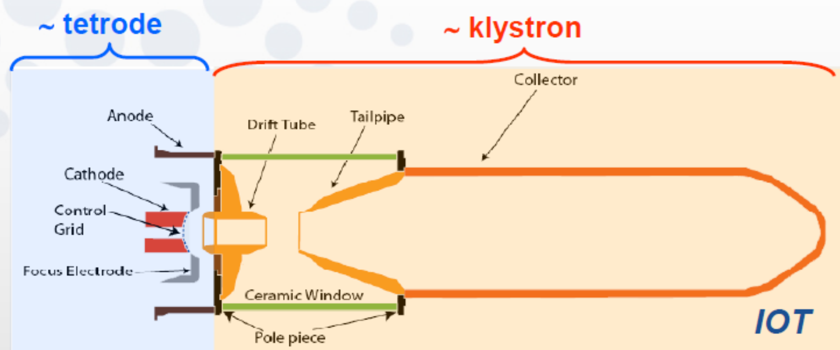
ESRF A Light for Science

IOT - Inductive Output Tubes or klystrodes



TV-IOT from E2V

TV –IOT from CPI



[www.cpii.com/eimac/PDF/Theory.PDF]

- Often with external in-air cavities allowing easy IOT exchange

- TV IOT: typically 60 kW at 460 – 860 MHz
- IOT developed for accelerators [Thales, CPI]:
  - 80 kW CW at 470 – 760 MHz
  - $\eta \approx 70\%$ ☞ *operation in class B*
  - Intrinsic low Gain = 20 … 22 dB $\Rightarrow P_{in} = 1$ kW
  - Compact, external cavity ⇒ easy to handle
  - BUT: low unit power ⇒ power combiners

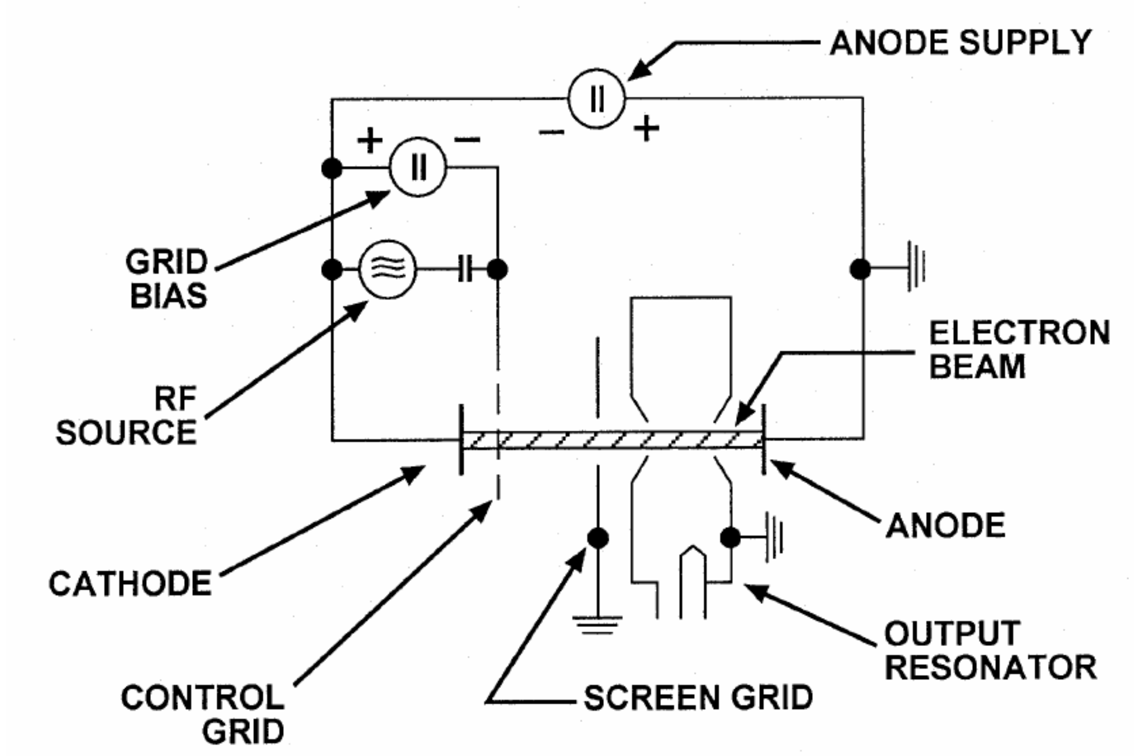
- 1.3 GHz IOT for cw X-FEL Linacs & ERLs
  - 16…20 kW
  - $\eta \approx 55$ to 65% [Thales, CPI, E2V]
  - No adequate klystron on the market
  - Superiority of IOTs:
    - ☞ Higher efficiency
    - ☞ Less amplitude & phase sensitivity to HV ripples
    - ☞ No collector overheating after loss of drive
    - ☞ Expected lower costs

European Synchrotron Radiation Facility SRF'2009 – 18th Sept. 2009 Tutorial: RF Power Sources J. Jacob, slide 15

History of the IOT: www.bext.com/iot.htm

IOT – Inductive Output Tubes (2)

IOT RF Power Sources for Pulsed and CW Linacs



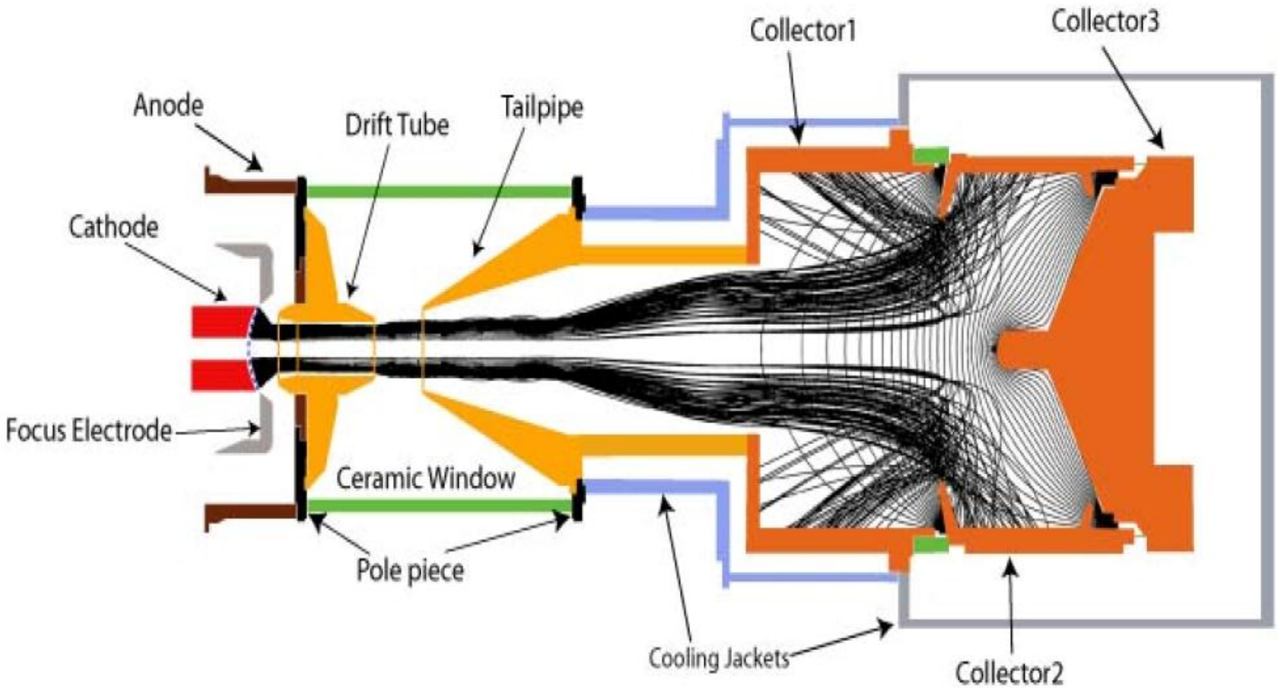
An IOT is a simple device

IOT – Inductive Output Tubes (3)

Communications & Power Industries
Eimac division

IOT RF Power Sources for Pulsed and CW Linacs

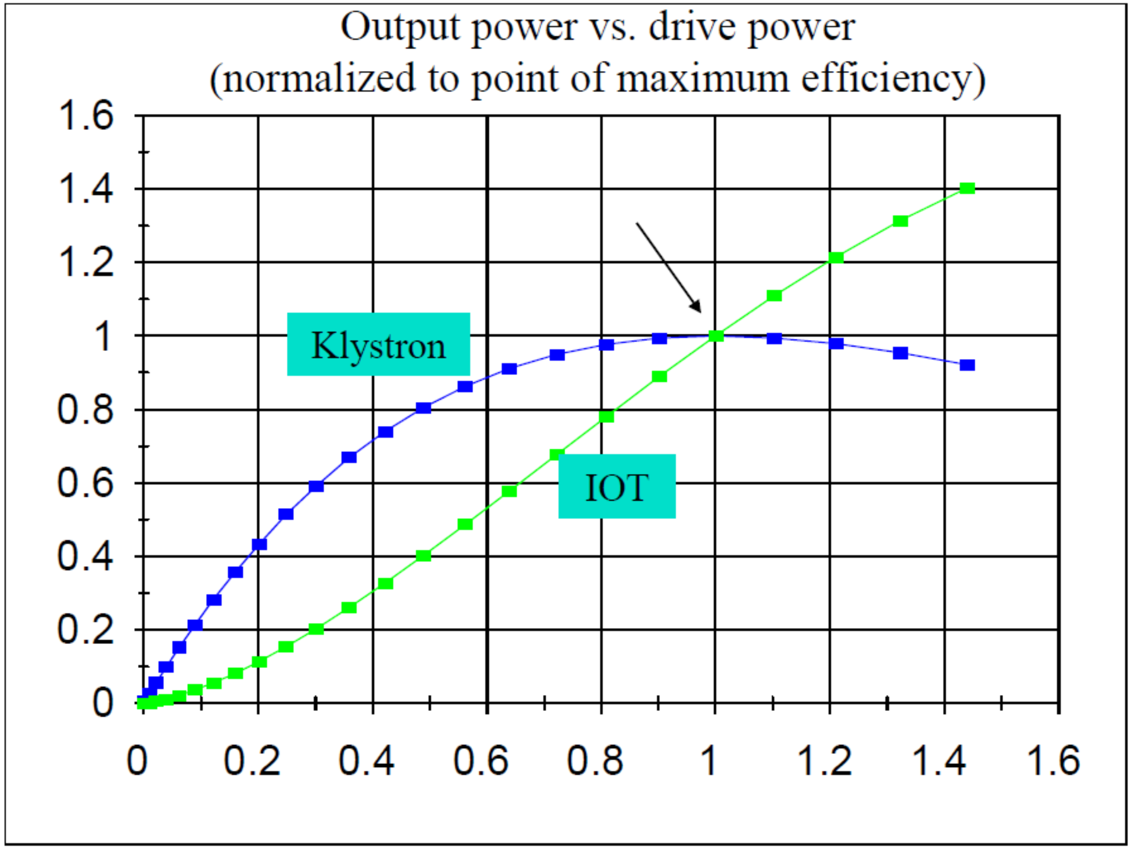
This computer graph shows the equipotential lines inside the collector assembly and the resulting distribution of the spent electron beam.



IOT – Inductive Output Tubes (4)

Communications & Power Industries
Eimac division

IOT RF Power Sources for Pulsed and CW Linacs



Comparison of amplifier characteristics: IOT vs. klystron

Comparison of solid state and vacuum technology for RF power generation (1986)

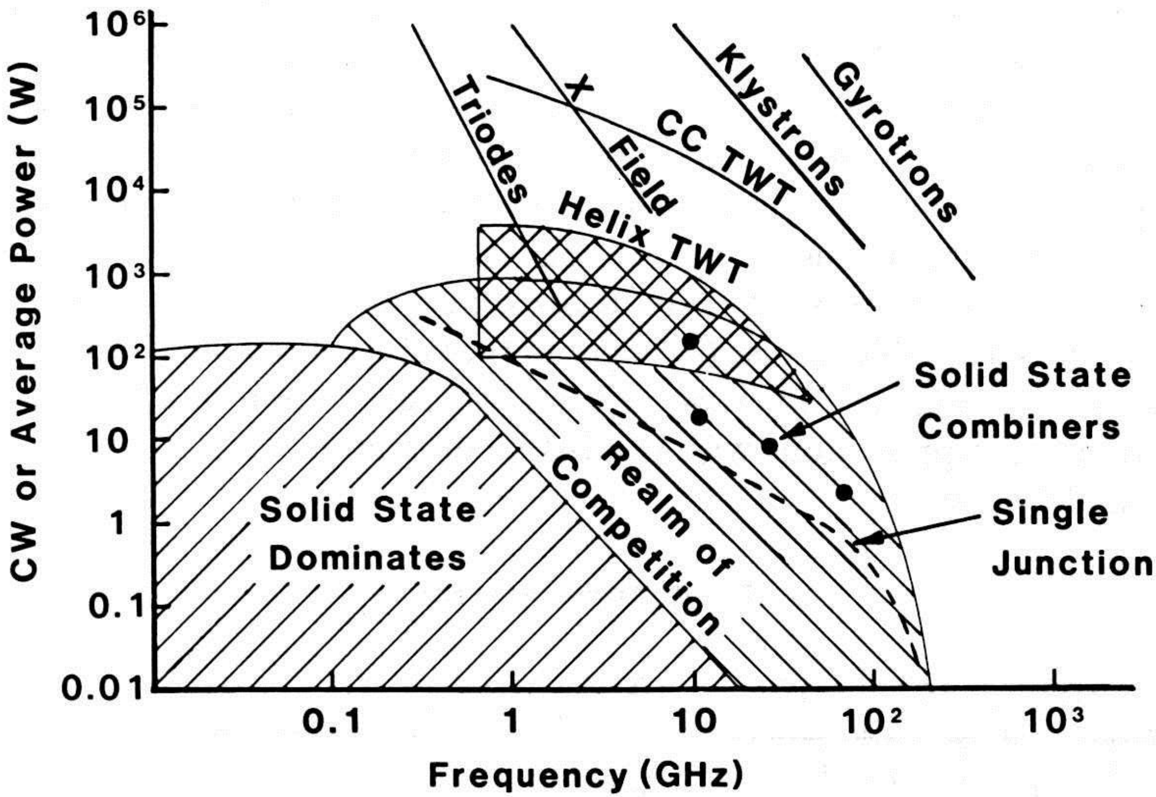
Solid state devices move steadily up in frequency

Abbreviations:

X Field: crossed field, especially magnetrons

TWT: Travelling wave tubes

CC TWT: coupled cavity TWT



Ref.: Gilmour, A S, *Microwave Tubes*, Artech House, 1986

Comparison of solid state and vacuum technology for RF power generation (2009)