Finetuning Academy

RF & Wireless Matters

RF FUNDAMENTALS

Presented By
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Co-axial transmission line

Characteristic Impedance of Coaxial Line

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

Phase Shift

$$\beta = \frac{\omega}{v_p} = \omega\sqrt{LC} \text{ rad/unit length}$$

Using above, we get

$$Z_0 = \frac{1}{v_p C}$$

Complex Propagation Constant

$$\gamma = \alpha + j\beta = \frac{1}{2}\left(\frac{R}{Z_0} + GZ_0\right) + j\omega\sqrt{LC}$$

Q-Factor

$$\alpha = \alpha_c + \alpha_d$$

Co-axial transmission line

$$\alpha_c = \frac{R}{2Z_0} \text{ Np/unit length}$$

$$\alpha_d = \frac{GZ_0}{2} \text{ Np/unit length}$$

$$Q = \frac{\beta}{2\alpha} = \frac{\beta}{2(\alpha_c + \alpha_d)}$$

$$Q_c = \frac{\beta}{2\alpha_c}$$

$$Q_d = \frac{\beta}{2\alpha_d}$$

$$\frac{1}{Q} = \frac{1}{Q_c} + \frac{1}{Q_d}$$

Strip transmission line

$$v_p = \frac{c}{\sqrt{\epsilon_r}}$$

$$\beta = \frac{\omega}{v_p} = \frac{\omega\sqrt{\epsilon_r}}{c} = k_0\sqrt{\epsilon_r} \text{ rad/unit length}$$

$$Z_0 = \frac{\sqrt{\epsilon_r}}{cC} = \frac{\sqrt{\epsilon_r}}{c\epsilon_r C_0} = \frac{1}{c\sqrt{\epsilon_r}C_0}$$

$$\alpha_d = \frac{\beta}{2}\tan\,\delta = \frac{k_0\sqrt{\epsilon_r}}{2}\tan\,\delta \text{ Np/unit length}$$

or

$$\alpha_d = 4.343\beta\tan\,\delta = 4.343k_0\sqrt{\epsilon_r}\tan\,\delta = 27.3\sqrt{\epsilon_r}\,\frac{\tan\,\delta}{\lambda_0} \text{ dB/unit length}$$

where $\tan\,\delta$ denotes the loss tangent of the dielectric material. In general, the loss tangent $\tan\,\delta$ is also a function of frequency.

Ref: S B Cohn, "Problems in Strip Transmission Lines"

Quasi - TEM Mode

For quasi-TEM modes, the effective dielectric constant ϵ_{re} is defined as follows:

$$\epsilon_{re} = \frac{c^2}{v_p^2}$$

$$\beta = \frac{\omega}{v_p} = \frac{\omega\sqrt{\epsilon_{re}}}{c} = \sqrt{\epsilon_{re}}\,k_0$$

in the quasistatic limit, ϵ_{re} can be assumed to

$$\epsilon_{re} = \frac{C}{C_0}$$

$$Z_0 = \frac{\sqrt{\epsilon_{re}}}{cC} = \frac{1}{c\sqrt{CC_0}} = \frac{1}{c\sqrt{\epsilon_{re}}\,C_0} = \frac{Z_{0a}}{\sqrt{\epsilon_{re}}}$$

$$\alpha_d = 27.3\,\frac{\epsilon_r}{\sqrt{\epsilon_{re}}}\,\frac{(\epsilon_{re}-1)}{(\epsilon_r-1)}\,\frac{\tan\delta}{\lambda_0}\ \text{dB/unit length}$$

$$\alpha_c = \frac{8.686\,R_s}{Z_c W}\ \text{dB/unit length} \qquad R_s = \sqrt{\frac{\omega\mu_0}{2\sigma}}$$

Synthesis of W/h

For $W/h \le 2$

$$\frac{W}{h} = \frac{8\exp(A)}{\exp(2A) - 2}$$

with

$$A = \frac{Z_c}{60}\left\{\frac{\varepsilon_r + 1}{2}\right\}^{0.5} + \frac{\varepsilon_r - 1}{\varepsilon_r + 1}\left\{0.23 + \frac{0.11}{\varepsilon_r}\right\}$$

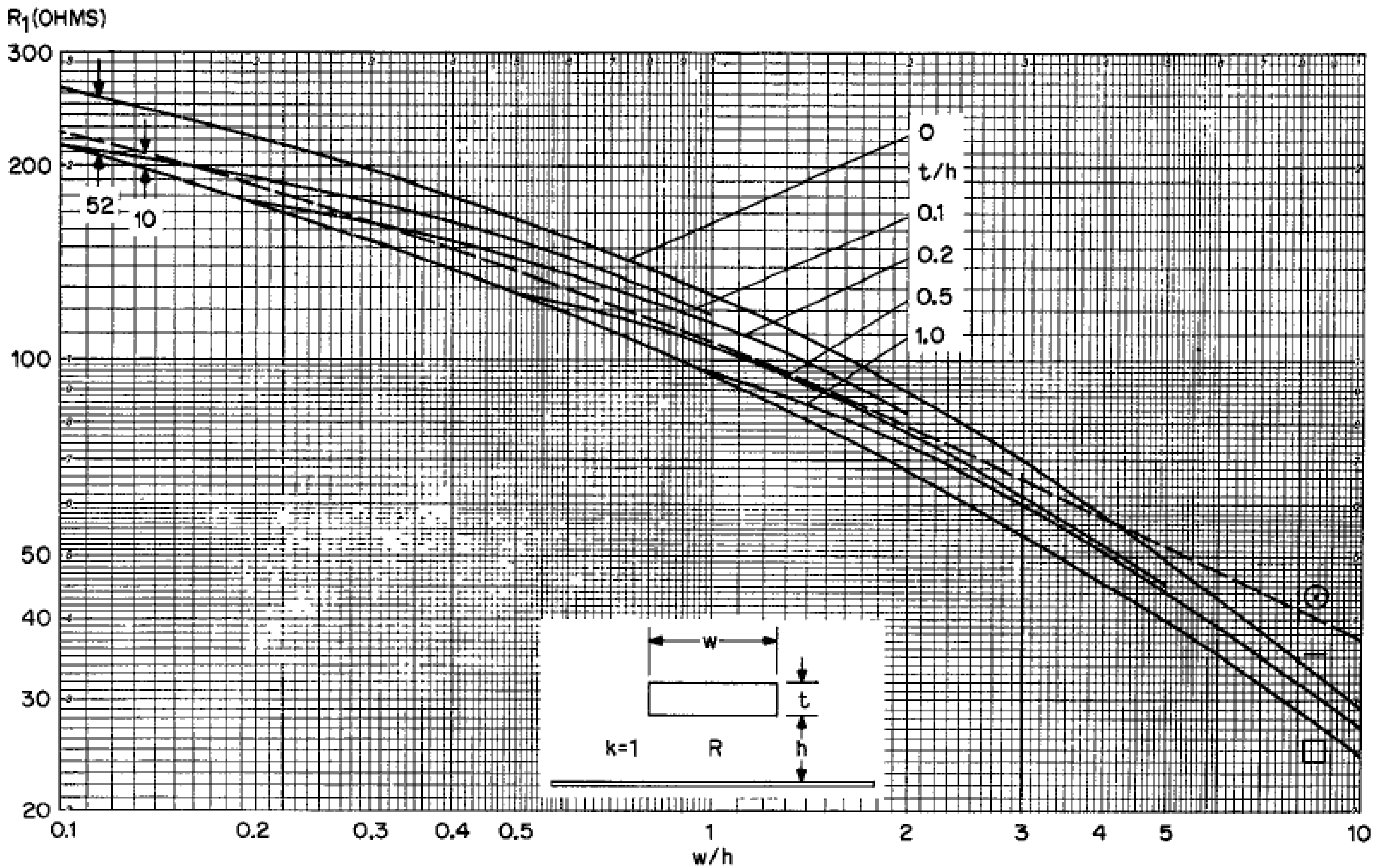
and for $W/h \ge 2$

$$\frac{W}{h} = \frac{2}{\pi}\left\{(B-1) - \ln(2B-1) + \frac{\varepsilon_r - 1}{2\varepsilon_r}\left[\ln(B-1) + 0.39 - \frac{0.61}{\varepsilon_r}\right]\right\}$$

with

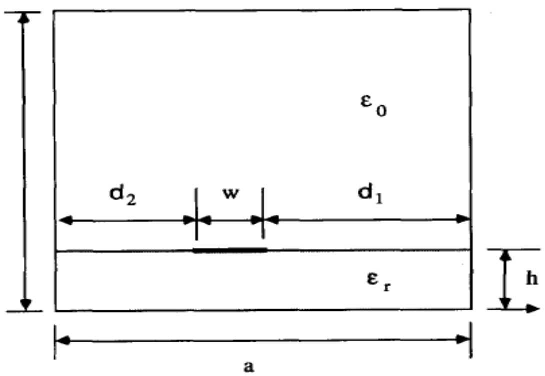
$$B = \frac{60\pi^2}{Z_c\sqrt{\varepsilon_r}}$$

Effect of Strip thickness



Shielding effect in u-strip line

$w = h = 0.635$ mm and $\epsilon_r = 9.7$



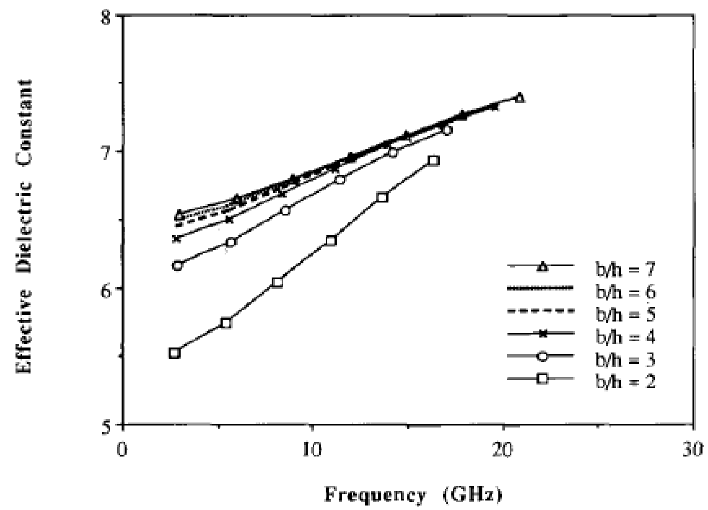


Fig. 5. The effects of lowering the top cover for fixed $a/h = 10$ (i.e., $d_1/h = d_2/h = 4.5$).

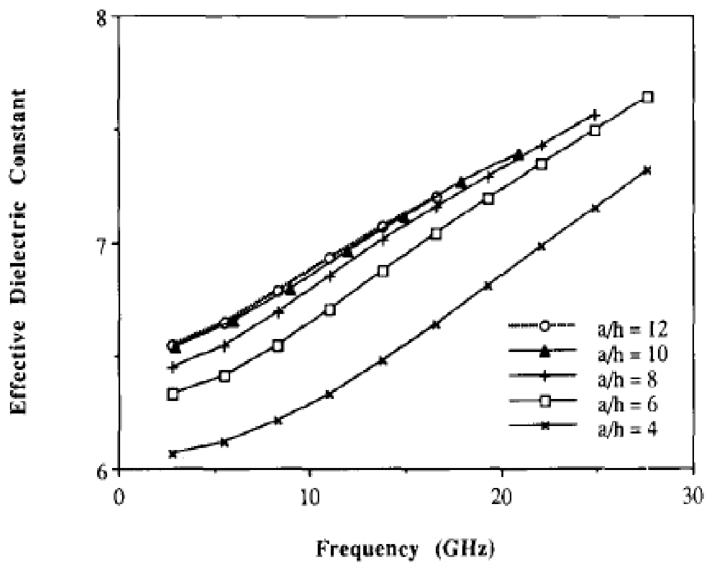
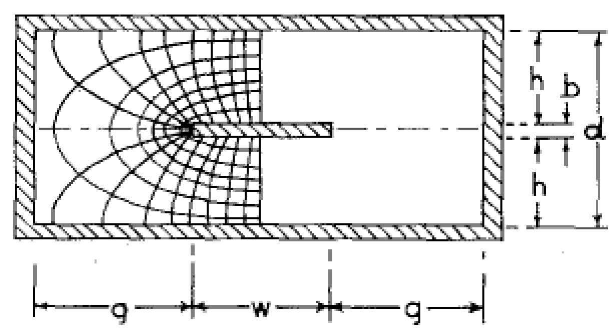


Fig. 6. The effects of closing in the two sidewalls with the normalized sidewall-to-strip spacing $d_1/h = d_2/h = (a/h - w/h)/2$. The top-cover height is fixed at $b/h = 7$.



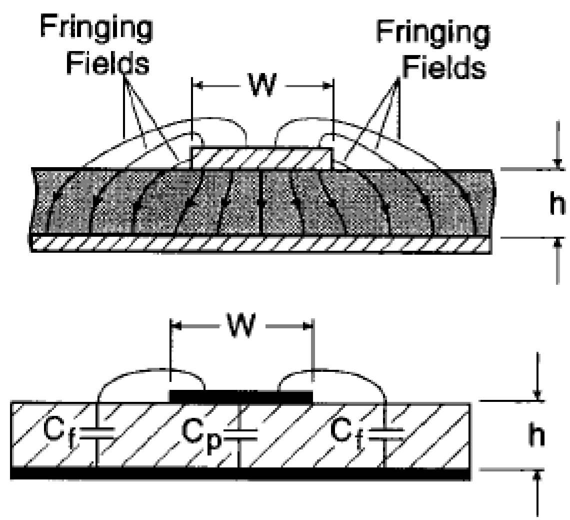
RULE OF THUMB: b>7*h and a>10*h

Capacitance of Micro-strip line

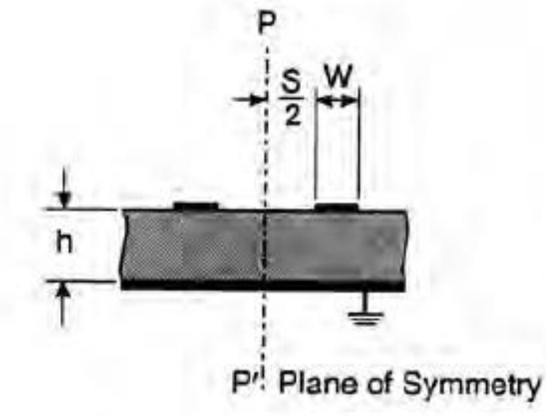
$$C = C_p + 2C_f$$

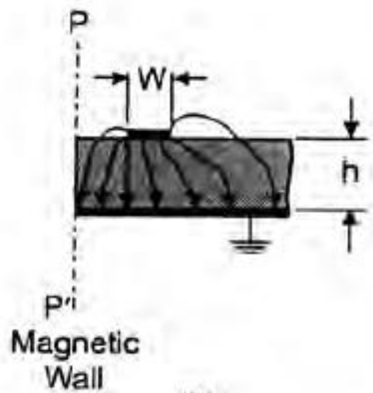
where

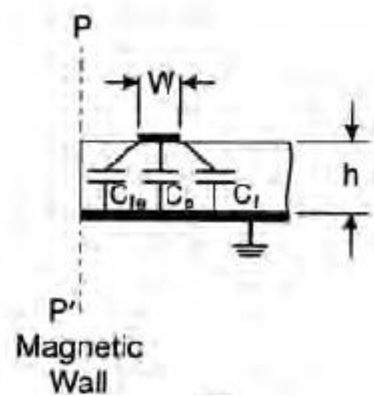
$$C_p = \frac{\epsilon_0 \epsilon_r W}{h}$$



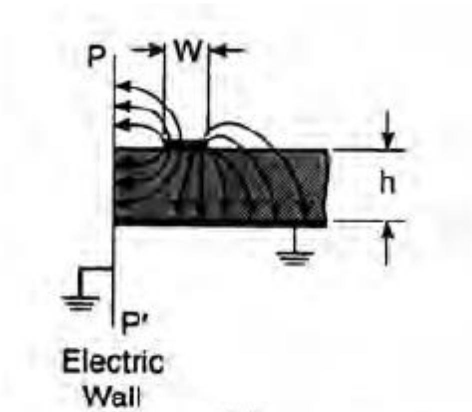
Capacitance of coupled line

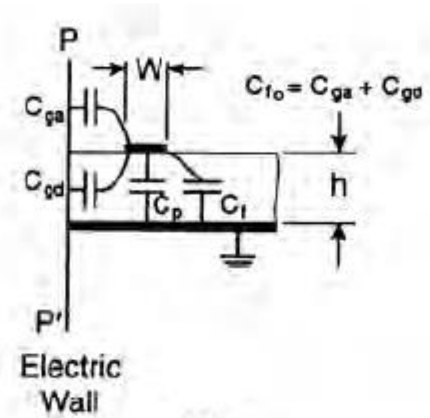






Symmetrical coupled line





$$Z_{0e} = \frac{1}{v_{pe} C_e} = \frac{\omega}{\beta_e C_e} \qquad v_{pe} = \frac{c}{\sqrt{\epsilon_{ree}}}$$

and

$$Z_{0o} = \frac{1}{v_{po} C_o} = \frac{\omega}{\beta_o C_o} \qquad v_{po} = \frac{c}{\sqrt{\epsilon_{reo}}}$$

Symmetrical coupled line

$$\epsilon_{ree} = \frac{C_e}{C_{0e}} \qquad \epsilon_{reo} = \frac{C_o}{C_{0o}}$$

$$v_{po} > v_{pe}$$

Poor coupler directivity, spurious response at 2f0 in parallel coupled line filters

$$C_{ab} = \frac{1}{4\pi f_0 Z_{0o} \tan\, \theta_0}$$

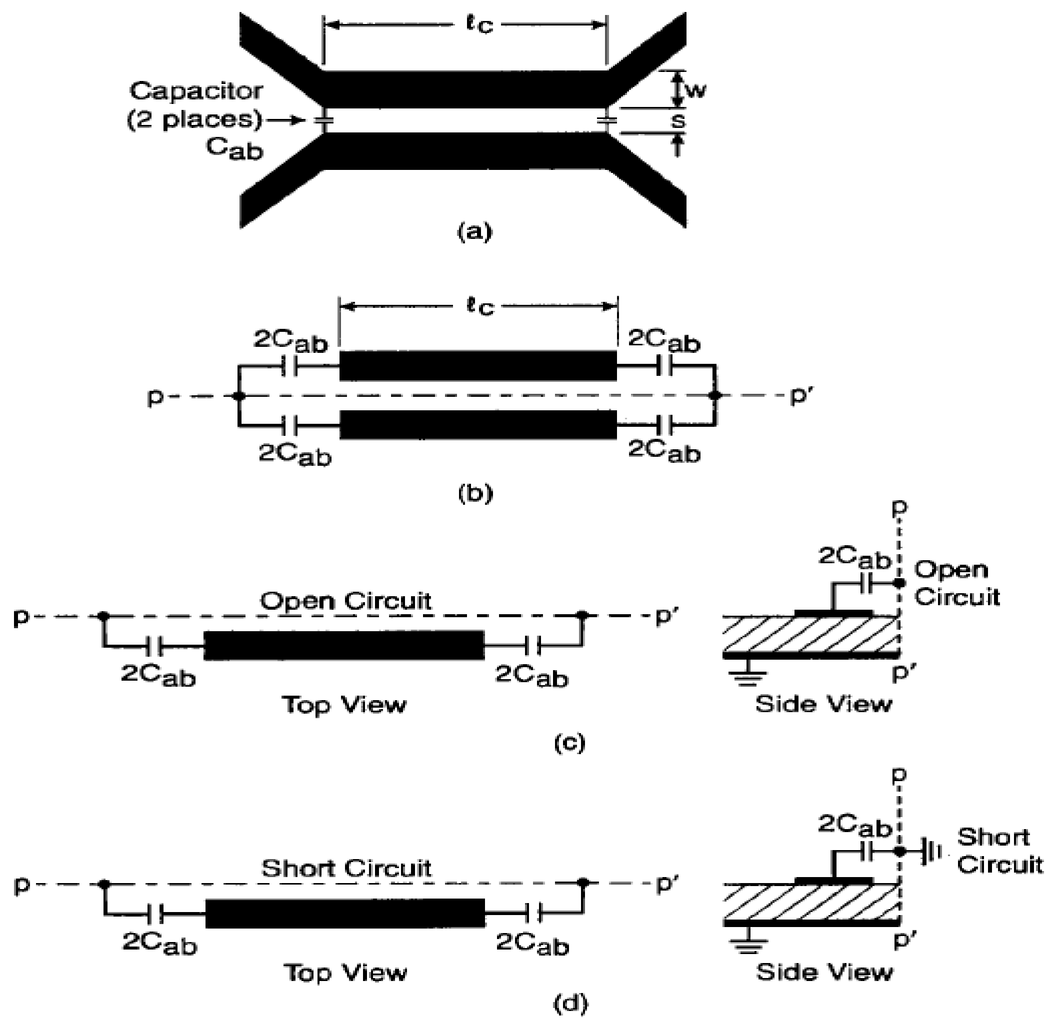
where

$$\theta_0 = \frac{\pi}{2}\sqrt{\frac{\epsilon_{reo}}{\epsilon_{ree}}}\ \text{rad}$$

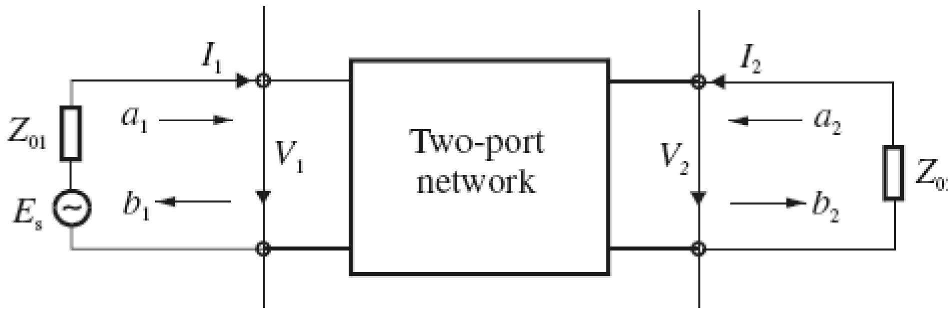
The physical length of a capacitor compensated quarter-wave coupler is given

$$l_c = \frac{\frac{\pi}{2} - \tan^{-1}\left(\pi f_0 C_{ab} Z_{0e}\right)}{k_0\sqrt{\epsilon_{ree}}}$$

Compensation for unequal phase velocities



S Parameters



$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \end{bmatrix}$$

For an N-Port Network

$$\begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_M \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} & \dots & S_{1M} \\ S_{21} & S_{22} & \dots & S_{2M} \\ \vdots & \vdots & \dots & \vdots \\ S_{M1} & S_{M2} & \dots & S_{MM} \end{bmatrix} \cdot \begin{bmatrix} a_1 \\ a_2 \\ \vdots \\ a_M \end{bmatrix}$$

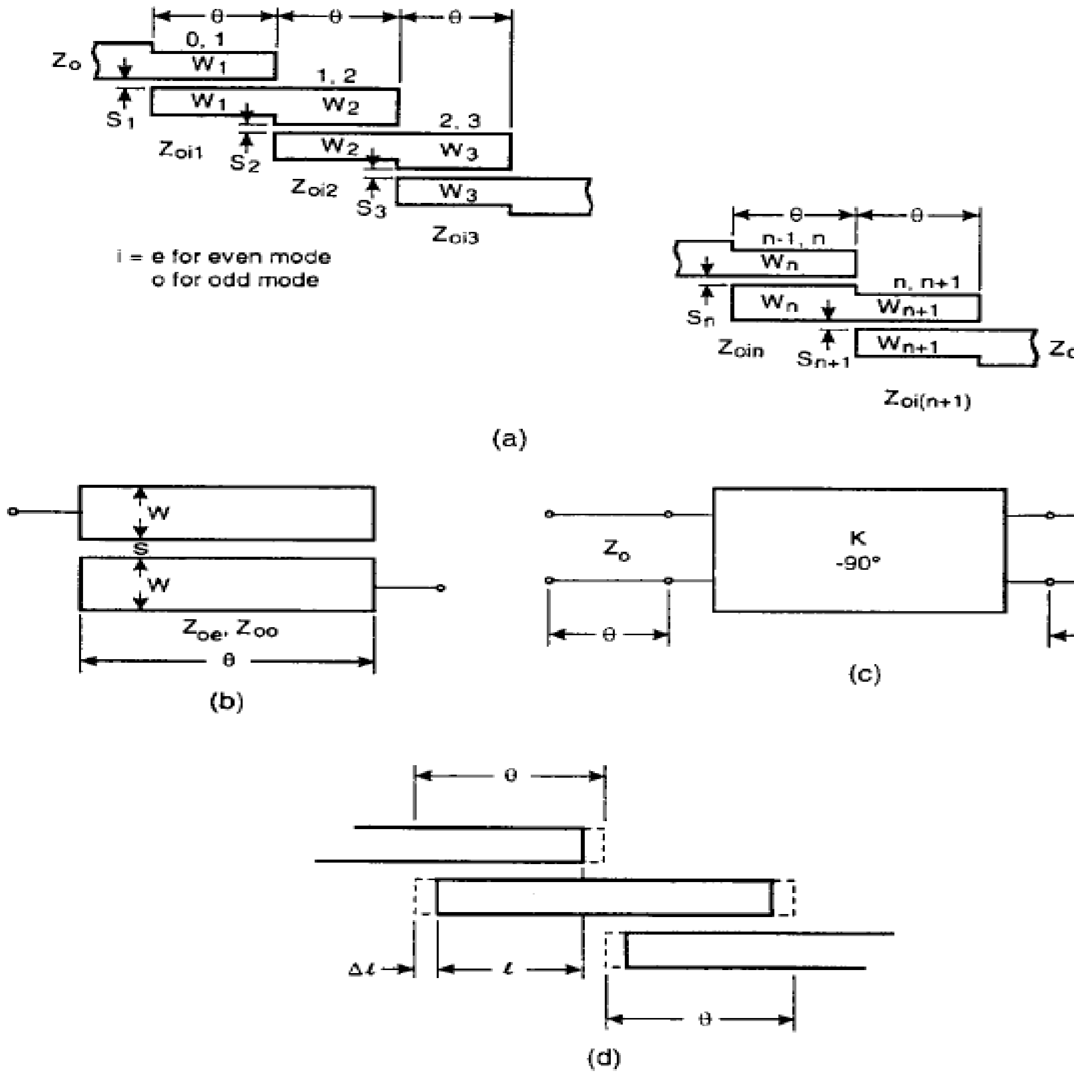
For a reciprocal network, $S_{ij} = S_{ji}$ and $[S]$ is a symmetrical matrix such that

$$[S]^t = [S]$$

where the superscript t denotes the transpose of matrix. For a lossless passive network,

$$[S]^t[S]^* = [U]$$

Edge Coupled Filter



$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} \cos\theta & jZ_0 \sin\theta \\ \dfrac{j\sin\theta}{Z_0} & \cos\theta \end{bmatrix} \begin{bmatrix} 0 & -jK \\ -\dfrac{j}{K} & 0 \end{bmatrix} \begin{bmatrix} \cos\theta & jZ_0 \sin\theta \\ \dfrac{j\sin\theta}{Z_0} & \cos\theta \end{bmatrix}$$

Edge Coupled Filter

$$\frac{Z_0}{K_{01}} = \sqrt{\frac{\pi \Delta f}{2\omega_1 g_0 g_1}}$$

$$\frac{Z_0}{K_{j,j+1}} = \frac{\pi \Delta f}{2\omega_1 \sqrt{g_j g_{j+1}}}, \; j = 1 \text{ to } n-1$$

$$\frac{Z_0}{K_{n,n+1}} = \sqrt{\frac{\pi \Delta f}{2\omega_1 g_n g_{n+1}}}$$

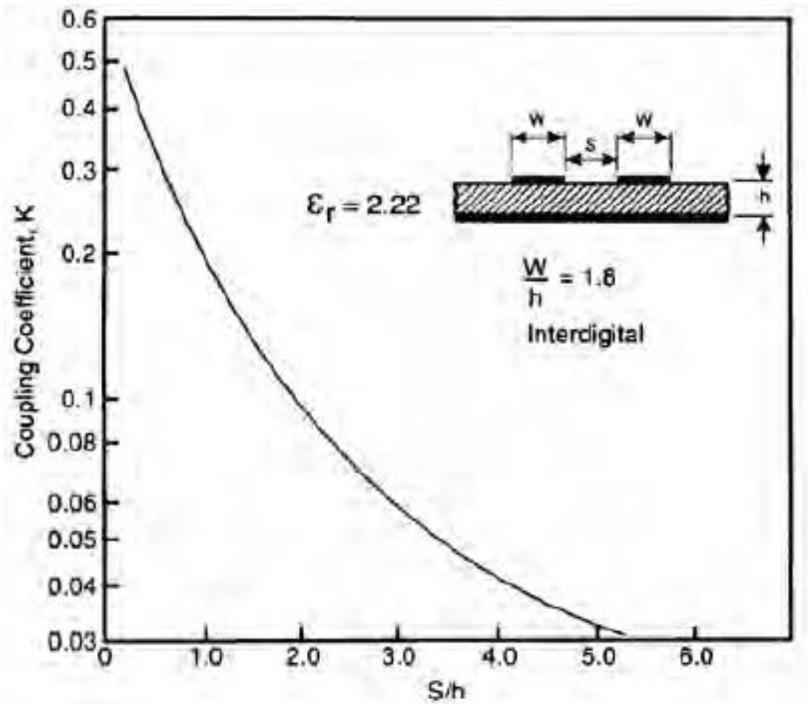
$$\frac{(Z_{0e})_{j+1}}{Z_0} = 1 + \frac{Z_0}{K_{j,j+1}} + \left(\frac{Z_0}{K_{j,j+1}}\right)^2, \; j = 0 \text{ to } n$$

$$\frac{(Z_{0o})_{j+1}}{Z_0} = 1 - \frac{Z_0}{K_{j,j+1}} + \left(\frac{Z_0}{K_{j,j+1}}\right)^2, \; j = 0 \text{ to } n$$

An approximate value for the physical length is obtained from average of even and odd mode velocities, i.e.

$$\theta = \frac{2\pi}{\lambda} l = \frac{2\pi}{\lambda_0} \frac{\sqrt{\epsilon_{ree}} + \sqrt{\epsilon_{reo}}}{2} l = \frac{\pi}{2}$$

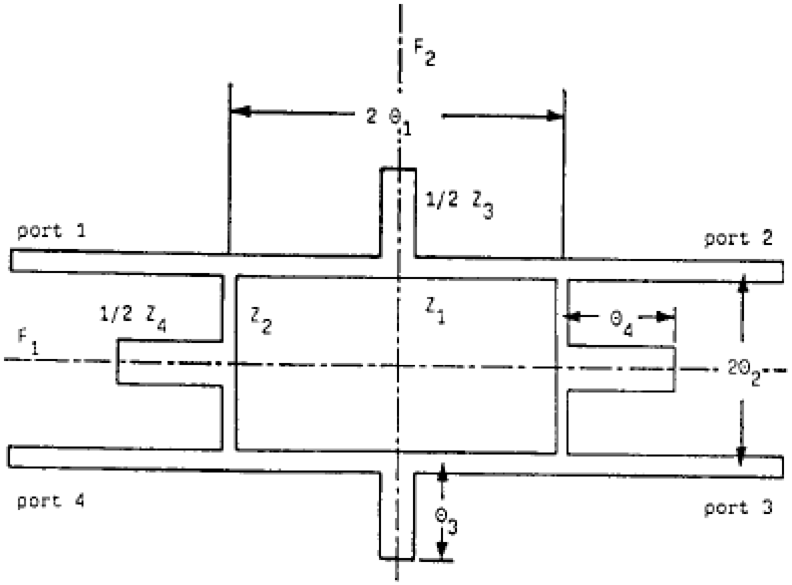
Edge Coupled Filter



$$\text{Midband Insertion Loss} = \frac{\alpha \lambda_o}{10w} \Sigma g_i$$

where α is the loss in dB/inch, λ_o is measured in inches, and w is the fractional BW

Broadband Branch Line Coupler



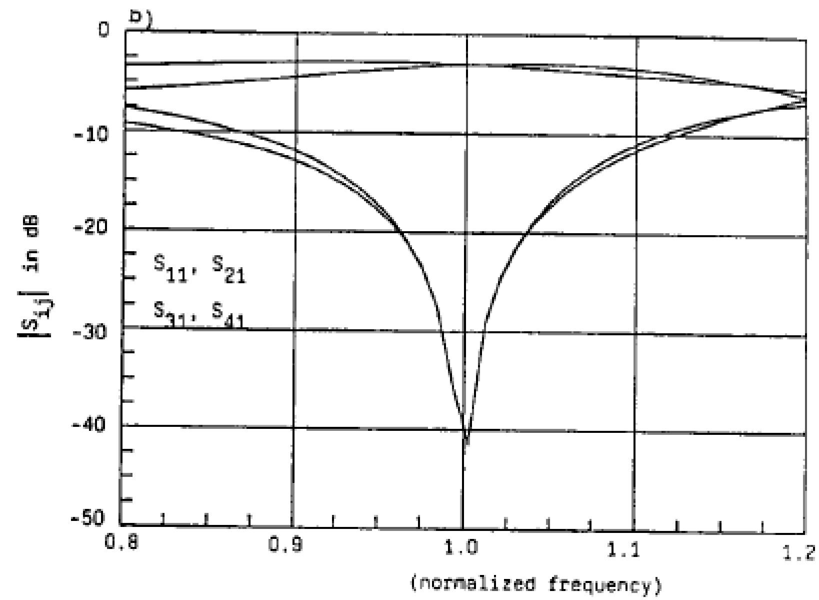
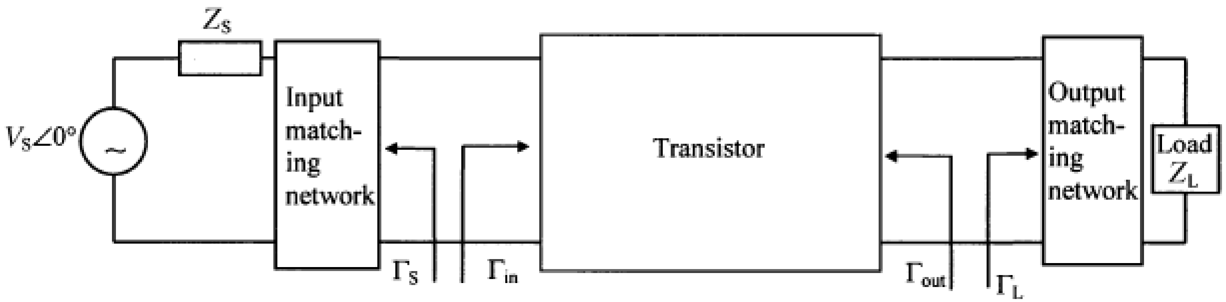


Fig. 5. a) Frequency behaviour of the eigenvalues of a coupler pictured in Fig. 2. All characteristic impedances are set to 70.7Ω. The calculations yield an optimum solution of θ_1 = 26.5°, θ_2 = 35.3°, θ_3 = 56.3° and θ_4 = 35.3° electrical length.
b) Associated S-Parameters

Ref: 1990 IEEE MTT-S Digest

Two Port Gain and Stability



$$\text{Transducer power gain,}\quad G_T = \frac{P_L}{P_{avs}} = \frac{\text{Power delivered to the load}}{\text{Power available from the source}}$$

$$\text{Operating power gain,}\quad G_P = \frac{P_L}{P_{in}} = \frac{\text{Power delivered to the load}}{\text{Power input to the network}}$$

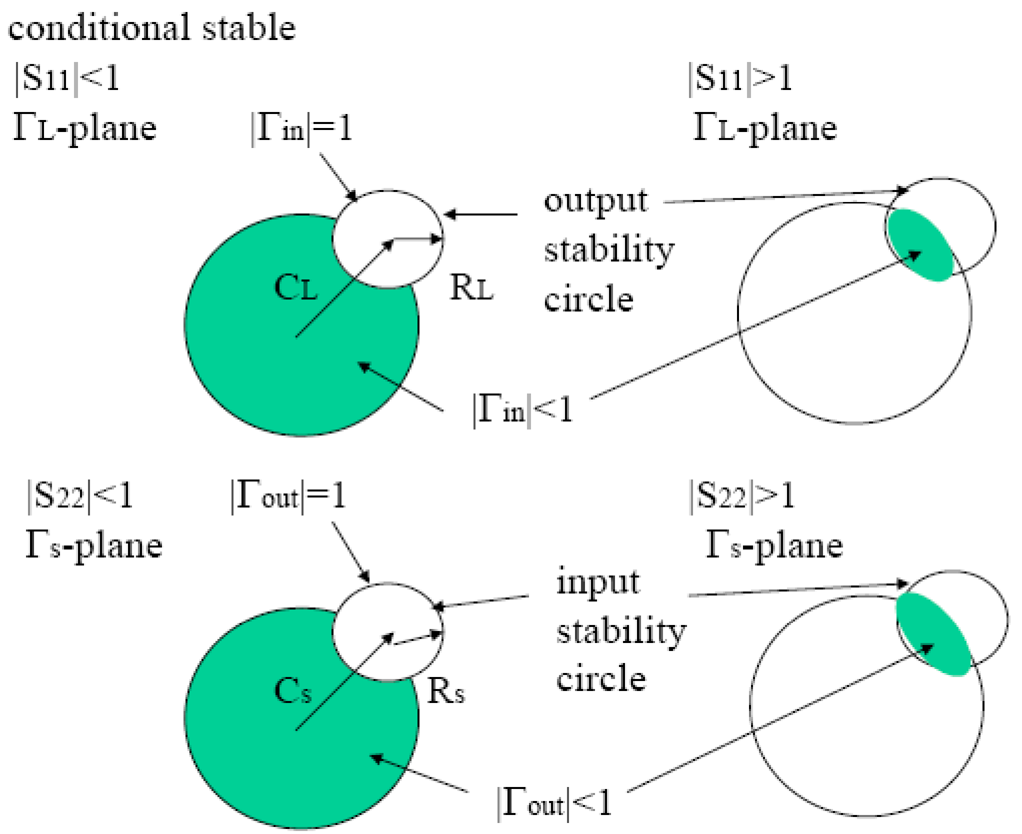
$$\text{Available power gain,}\quad G_A = \frac{P_{AVN}}{P_{avs}} = \frac{\text{Power available from the network}}{\text{Power available from the source}}$$

$$G_T = \frac{1-|\Gamma_S|^2}{|1-S_{11}\Gamma_S|^2}\cdot|S_{21}|^2\cdot\frac{1-|\Gamma_L|^2}{|1-\Gamma_{out}\Gamma_L|^2}$$

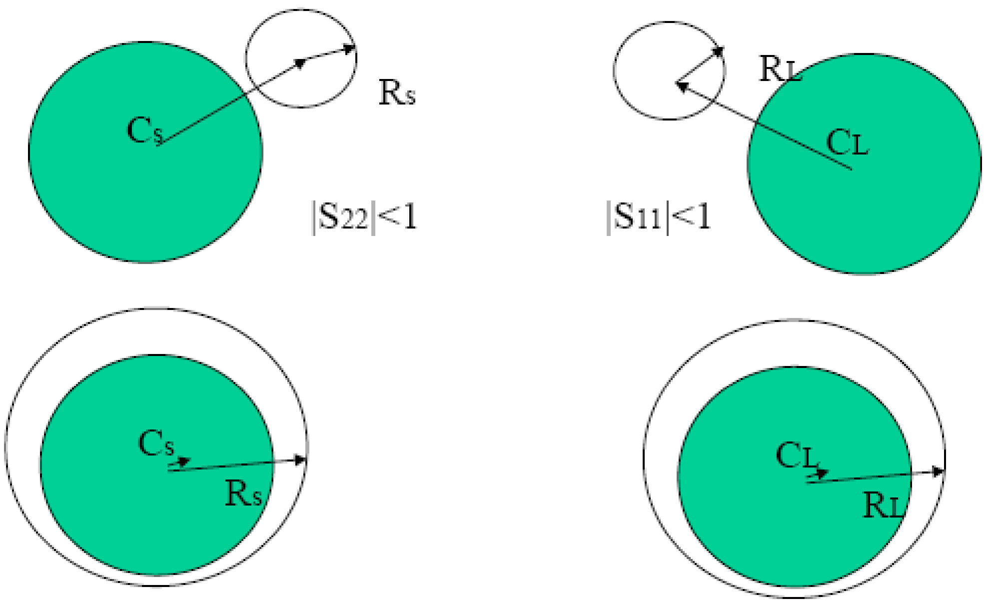
$$\Gamma_{in} = f(\Gamma_L) = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L},$$

$$\Gamma_{out} = g(\Gamma_S) = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}.$$

Conditional Stability



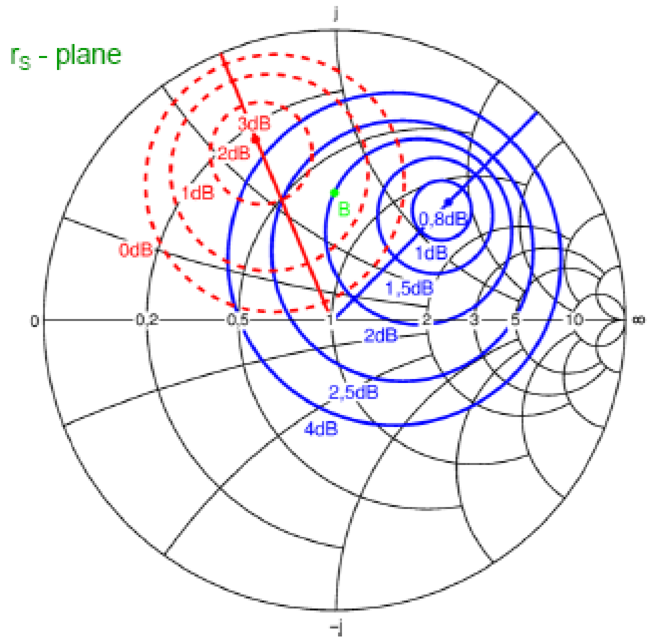
Unconditional Stability



$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1$$

$$|\Delta| = |S_{11}S_{22} - S_{12}S_{21}| < 1$$

Gain, NF Circles



Gain: ------------

NF: ------------