Microstrip matching networks

Substrates: Duroid, quartz, alumina, silicon.

Propagation only quasi-TEM since not all E-M field lines are not entirely in substrate; problem at higher freq.

Phase velocity for quasi-TEM: $v_p = \frac{f \varepsilon_0}{\varepsilon_r}$

Characteristic impedance: $Z_0 = \frac{\sqrt{f_0 \varepsilon_0}}{c}$ with $c = \frac{\lambda}{n}$

Wavelength: $\lambda = \frac{c}{f \varepsilon_0}$

Per unit length of the microstrip:

\[
\frac{ff_0\varepsilon}{0\varepsilon} = \frac{(ff_0\varepsilon f)}{c} = f/d\varepsilon = \text{not}
\]

\[
\text{Effective relative dielectric constant of the substrate: } \varepsilon_r
\]

\[
\text{Characteristic impedance: } Z_0 = \frac{\sqrt{ff_0\varepsilon}}{c} = d\varepsilon
\]

\[
\text{Phase velocity for quasi-TEM: } v_p = \frac{ff_0\varepsilon}{\varepsilon_r}
\]

\[
\text{Wavevelngth: } \lambda = \frac{c}{f \varepsilon_0}
\]

\[
\text{Per unit length of the microstrip: }
\]

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\text{Characteristic impedance: } Z_0 = \frac{\sqrt{ff_0\varepsilon}}{c} = d\varepsilon
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\text{Phase velocity for quasi-TEM: } v_p = \frac{ff_0\varepsilon}{\varepsilon_r}
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\text{Wavevelngth: } \lambda = \frac{c}{f \varepsilon_0}
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\text{Per unit length of the microstrip: }
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\[
\text{Phase velocity for quasi-TEM: } v_p = \frac{ff_0\varepsilon}{\varepsilon_r}
\]

\[
\text{Wavevelngth: } \lambda = \frac{c}{f \varepsilon_0}
\]
\[ \left( \frac{\eta}{\mu} - 1 \right) \frac{0.04}{2} + \frac{1}{2} \left( \frac{\mu}{\eta} \right) I \left( \frac{1}{\mu} + 1 \right) \frac{0.02}{1} + \frac{0.02}{I + 1} = \frac{f f_{\varepsilon}}{e} \]

where

\[ \left( \frac{\eta}{\mu} \right) I \left( \frac{0.05}{2} + \frac{M}{\eta} \right) \frac{f f_{\varepsilon}}{0.06} = Z \]

For \[ I \geq \frac{\eta}{M} \]

Assuming zero (negligible) thickness of the strip conductor

\[ Z > \frac{\eta}{t} \]

How to find \( f f_{\varepsilon} \) and \( Z \) preferably in analytical form.
(6) \[
\frac{1}{\sqrt{2}} \left[ \frac{(\eta/M)(1 - \epsilon^2)\epsilon + I}{\epsilon} \right]^{\epsilon/\sqrt{0}} = \gamma
\]

For \( \eta/M > 0.6 \bullet \)

(5) \[
\frac{1}{\sqrt{2}} \left[ \frac{(\eta/M)(1 - \epsilon^2)\epsilon + I}{\epsilon} \right]^{\epsilon/\sqrt{0}} = \gamma
\]

For \( \eta/M < 0.6 \bullet \)

(4) \[
\left( \frac{M}{\eta} I_2 + I \right) \frac{2}{I - \epsilon^2} + \frac{2}{I + \epsilon^2} = ff \epsilon
\]

where

(3) \[
\frac{ff \epsilon^{\sqrt{0}/\nu}}{120 \nu} + \eta/M = \gamma
\]

For \( \eta/M \bullet \)
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\[ (8) \quad \left\{ \left[ \frac{\varepsilon_{r}}{1} - 0.39 + (1 - \varepsilon_{r})\mu_{r} \right] \frac{\varepsilon_{r}}{1 - \varepsilon_{r}} + (1 - \varepsilon_{r})\mu_{r} - 1 - \varepsilon_{r} \right\} \frac{\nu}{\varepsilon_{r}} = \frac{\eta}{M} \]

\[ \frac{\varepsilon_{r} - \nu \varepsilon_{r}^{2}}{\nu \varepsilon_{r}^{2}} = \frac{\eta}{M} \]

For design: eqs. relating \( Z \) with \( \varepsilon_{r} \) and \( M/\eta \).
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For $\frac{h}{W} < \frac{1}{2\nu}$:

\[
\left( \frac{1}{w} \left( \frac{1}{2} + 1 \right) \frac{\nu}{\kappa} \frac{h}{W} \right) + \frac{1}{M} = \frac{1}{W} \frac{1}{M}
\]

and

\[
\left( \frac{1}{w} \frac{1}{Z} \right) \frac{\nu}{\kappa} = B
\]

Non-negligible thickness of conductor included via increased capacitance replace strip width $W$ with $W^\ell$, e.g.

\[
\left( \frac{1}{w} + 1 \right) \frac{1}{Z} \frac{h}{W} + \frac{1}{M} = \frac{1}{W} \frac{1}{M}
\]

where
High impedance, thin substrates $\iff$ less dispersion.

\[ \frac{8 \pi \mu (\text{cm})}{Z} = \frac{df}{f} \left( \frac{df}{f} \right)^2 C + \frac{I}{C - \epsilon} = \left( \frac{Z_{GHz}}{f} \right)^2 \epsilon \]

For $\epsilon$ we have:

\[ \frac{1}{Z} \left( \frac{\mu}{\mu_{cm}} \right)^2 = \left( \frac{Z_{GHz}}{f} \right)^2 \epsilon \]

Dispersion neglected below.

High frequencies, quasi-TEM not valid and $Z$ increases with freq. while $\mu$ decreases. Also, functions of freq. at high frequencies $\iff$ $\epsilon$ decreases with freq. $Z$ and $\mu$ are $\propto$ freq.

Example 2.5.1: see table 2.5.4. Use CAD tools (MDS)
Expressions for dispersion of $Z^0$ are also available. Another problem: losses. Both dielectric and ohmic. For dielectric substrates, dielectric losses are normally smaller than conductor losses.

Further complication: quality factor $Q$. Calculated from:

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

where $Q_c$, $Q_d$, and $Q_r$ are the conductor loss, radiation loss, and loss due to conductor loss, respectively. Quality factor $Q$ also affects the radiation loss, so that the total loss can be expressed as:

$$\text{DB}(\alpha) = \frac{\alpha}{2} \approx 27.3 \text{ dB} \text{ or } Q = \frac{P_{\text{in}}}{P_{\text{out}}} \approx 8.686 \text{ or } \frac{P_{\text{in}}}{P_{\text{out}}} = Q \Leftrightarrow$$

$$\frac{\alpha}{\nu} = \emptyset \text{ and } \alpha = \text{ the total loss.}$$

(14) \[ \frac{H_0^0}{Z} = \emptyset, \quad \frac{p \lambda C}{\nu} = p \emptyset, \quad \frac{c \lambda C}{\nu} = c \emptyset \]

(13) \[ \frac{\emptyset + p \emptyset + c \emptyset}{I} = \emptyset I \]
Microstrip matching networks

Uses of microstrip lines:

- Quarter-wave transformer: can transform 50 Ohm resistor to any value of resistance.
- Quarter-wave microstrip can transform 50 Ohm resistor into any value of impedance.
- Microstrip + short/open circuited stubs can transform short and open circuited stubs.
- Series tr. L lines:
- Uses of microstrip tr. Lines:
Microstrip matching networks

Using short/open-circuited stubs and strips. Lines: see Fig. 2.5.6.

For design procedure, see Fig. 2.5.7. General idea: use shunt stubs.

Alternatively, match any load to 50 Ohms. See Fig. 2.5.6 for procedure.

Using short/open-circuited stubs and strips. Lines: see Fig. 2.5.7.

Using shunt/series stubs. Lines: see Fig. 2.5.8.

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(Fig. 2.5.11) Adjust the length of the stub to get the final \( B_m \). (Fig. 2.5.11) But another option: use any practical value of \( Z_{02} \) (i.e. \( \lambda_0 \neq I/B_m \)).

\[ V_{in} \]

Procedure: For \( B_m \) produced by \( \lambda/8 \) of O-C stub.

\[ B_m = \frac{\lambda}{2} \]

or \( \frac{\lambda}{4} \) stub, or \( \lambda/8 \) of O-C stub. (Fig. 2.5.10)

\[ \text{For } B > 0, \text{ use } \lambda/8 \text{ C stub or } 3\lambda/8 \text{ O-C stub. (Fig. 2.5.10).} \]

Use: \( Z_{01} \) or \( Z_{02} \) (Fig. 2.5.9). Then add appropriate B in shunt to get appropriate B produced by \( \lambda/8 \) of O-C stub. (Fig. 2.5.9)

\[ \text{For } B_m \text{ produced by } \frac{\lambda}{4} \text{ C stub, use } \frac{\lambda}{2} \text{ or } \frac{\lambda}{4} \text{ transformer for } 50 \text{ Ohm. } \]

Another option: use \( \lambda/4 \) transformer in series + \( \lambda/8 \) or \( 3\lambda/8 \) S-C stubs.
Example 2.5.2 iii

Values for source and load L's that give a good match are: L = 0.6147160°, Lₗ = 0.682297°. (fig. 2.5.12)

Switch to Z (or Y) S-chart and locate Yₛ and Yₗ

2.8 - j1.9 and Yₗ = 0.4 - j1.05

= Yₛ and 25° corresponding to these L's. Use ZY chart: L = Yₛ

Point A

circle to intersection of that circle with SWR circle for Yₛ

Start from I = Iₛ = Yₛ (I = 50 Ω); go on const. conductance

do source side first: fig. 2.5.5 a)

Switch to Z (or Y) S-chart and locate Yₛ and Yₗ

Values for source and load L's that give a good match are: L =
Microstrip matching networks

To get to pt. 1 we need change in susceptance of

\[ \Delta Y = j \tan \left( \frac{\theta}{2} \right) \]

\[ Y = j \tan \theta \]

Same procedure for \( Y_L \): I. Intersection of SWR circle for \( Y_L \) with constant conductance circle = pt. B. 2. For that length required, read off directly from a \( n \)-strip in series. Length required? Read off directly from 

From A to \( Y \) ? Go on constant \( Y \) circle (SWR circle), i.e., put

\[ \theta = \tan \left( \frac{\nu}{2} \right) \]

Shunt capacitance

Last part is again series tr. Line short circuited stub. Last part is again series tr. Line

\[ \Delta Y = j \cot \theta \]

Differece? Susceptance needed is inductive.
Microstrip matching networks

Short-circuiting: use bypass capacitors $C_B$ (50 to 500 pF).

1. Coupling capacitors $C_A$ to prevent DC from going to source.
2. Providing DC bias to base and collector. For collector biasing simple RF choke will do (0.077 A line is 5-C). For base DC goes directly to base terminal and needs to present an O-C.
3. S-chart: $T = 0.13 A$. Series &-strip stays the same.
4. From two, e.g., for $L$ use $L = 0.0095$ instead of $L = 0.14$. 
Modification: symmetrize the stubs. Two O-C or S-C stubs in shunt must give the same susceptance $\pm$ divide the original by

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2.53 which gives \( \lambda = 21.6 \text{ cm} \)

\[ \lambda = \frac{c}{f} \]

\( c = 300 \text{ cm} \text{s}^{-1} \) and \( f = 1 \text{ GHz} \). From Fig. 2.5.1 which gives \( \lambda = 70 \text{ cm} \text{s}^{-1} \) and \( f = 1 \text{ GHz} \). To find physical length.

Final result: \( W = 2.42 \text{ mm} \), \( h = 1.91 \text{ mm} \).

- For \( Eg. \) use eq. 2.5.7 (why do we need it?)
- Use eq. 2.5.11 for more accurate calculation of \( W/h \)

\[ W \approx \frac{h}{\lambda} \]

From Fig. 2.5.2, for \( Z_0 = 50 \Omega \) and \( \epsilon_r = 2.23 \), we have \( h = 0.784 \text{ mm} \).

Geometry of \( h\)-strip:

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(16) $j\theta' \cot \theta_0 Z' - = (\theta - S)^{\nu_0} \theta^\prime \tan 0 Z' = (\theta - S)^{\nu_0} Z$
(15) $j\theta' \tan 0 Z' = (\theta - \theta)^{\nu_0} \theta^\prime \cot 0 Z' - = (\theta - \theta)^{\nu_0} Z$

Reminder:

For imaginary part: use O-C stub

This provides the real part of the required final $\theta$.

This provides the real part of the required final $\theta$.

$$j\theta' \tan 0 = 0 \theta^\prime \cot 0 \theta = 4/\sqrt{\theta}$$

For imaginary part: use O-C stub

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Design no. 2: Use $\theta$-strip lines with different $\theta$.

Use $\theta$-strip lines with different $\theta$.
Weneed negative image part of $\frac{\gamma}{8}$ so that parallel combination produces $\frac{3\gamma}{8}$ but double the $Z_0$. From above eqs., we can use O-C stub with $l = \frac{\gamma}{8}$, or S-C stub with $l = \frac{\gamma}{8}$. We need negative image part of $\gamma$ in both cases. From above eqs.,

$$\frac{26.32 \text{ and } 47.6 \text{ Ohm.}}{47.6 \Omega \text{ transformer on } 50 \text{ Ohm load}}$$

Modification: use balanced stubs. Keep the length the same.

$$\frac{\gamma}{4}$$

produces required susceptance $-j\beta 0.021 = 1.0 / 0.021$.

Another O-C stub with $\frac{3\gamma}{8}$ and $Z_0 = 0 \alpha / 1 = \frac{0 \gamma}{1} = 0 \gamma / 0.021$.

Same procedure for load for O-C: $Z_0 = 0 \alpha / 1 = \frac{0 \gamma}{1} = \frac{0 \gamma}{0.021}$.

For O-C: $Z_0 = 0 \alpha / 1 = \frac{0 \gamma}{1} = \frac{0 \gamma}{0.021}$.

Both look like a shunt inductor with $\frac{\gamma}{4}$, so we can use O-C stub with $l = \frac{\gamma}{8}$, or S-C stub with $l = \frac{\gamma}{8}$.
substitute $e$ and $W/\lambda$.

changing the substrate: $l = \frac{W}{f}$, where $e$ depends on the substrate.

Note on changing the substrate: $l = \frac{W}{f}$, where $e$ depends on the substrate.

reverse engineering: Find $I$ for given circuit.

• 2.5.5 - reverse engineering:

i.e., don't use it for general loads.

quarter-wavelength transformer works only with real loads, not that quarter-wavelength transformer works only with real loads.

• 2.5.4 - using $V/4$ and transmission line (Fig. 2.5.18) :

• 2.5.3 - using different $Z_0$ and general load:

More examples: