Lecture 12: Microstrip. ADS and LineCalc.

One of the most widely used planar microwave circuit interconnections is microstrip. These are commonly formed by a strip conductor (land) on a dielectric substrate, which is backed by a ground plane (Fig. 3.25a):

![Microstrip Diagram](image)

We will often assume the land has zero thickness, \( t \).

In practical circuits there will often be metallic walls and covers to protect the circuit. We will ignore these effects, as does the text.

Unlike stripline, a microstrip has more than one dielectric in which the EM fields are located (Fig 3.25b):

![Microstrip Fields Diagram](image)

This presents a difficulty. Notice that if the field propagates as a TEM wave, then
\[ v_p = \frac{c_0}{\sqrt{\varepsilon_r}} \]

But which \( \varepsilon_r \) do we use?

The answer is neither because there is actually no purely TEM wave on the microstrip, but something that closely approximates it called a “quasi-TEM” mode. At “lower frequencies,” this mode is almost exactly TEM. Conversely, when the frequency becomes too high, there are appreciable axial components of \( \vec{E} \) and/or \( \vec{H} \) making the mode no longer quasi-TEM. This property leads to dispersive behavior, among other effects.

Numerical and other analysis have been performed on microstrip since approximately 1965. Some techniques, such as the method of moments, produce very accurate numerical solutions to equations derived directly from Maxwell’s equations and incorporate the exact cross-sectional geometry and materials of the microstrip.

From these solutions, simple and quite accurate analytical expressions for \( Z_0 \), \( v_p \), etc. have been developed primarily by curve fitting.

The result of these analyses is that at relatively “low” frequency, the wave propagates as a quasi-TEM mode with an effective relative permittivity, \( \varepsilon_{r,e} \).
The phase velocity and phase constant, respectively, are:

\[ v_p = \frac{c_0}{\sqrt{\varepsilon_{r,e}}} \]  
(3.193), (2)

\[ \beta = k_0 \sqrt{\varepsilon_{r,e}} \]  
(3.194), (3)

as for a typical TEM mode.

In general,

\[ 1 \leq \varepsilon_{r,e} \leq \varepsilon_r \]  
(4)

The upper bound occurs if the entire space above the microstrip has the same permittivity as the substrate, while the lower bound occurs if in this situation the material is chosen to be free space.

The characteristic impedance of the quasi-TEM mode on the microstrip can be approximated as

\[
Z_0 \approx \begin{cases} 
\frac{60}{\sqrt{\varepsilon_{r,e}}} \ln \left( \frac{8d + W}{W / 4d} \right) \left[ \Omega \right] & \frac{W}{d} \leq 1 \\
\frac{120 \pi}{\sqrt{\varepsilon_{r,e}}} \left[ \frac{W}{d} + 1.393 + 0.667 \ln \left( \frac{W}{d} + 1.444 \right) \right] \left[ \Omega \right] & \frac{W}{d} > 1
\end{cases}
\]  
(3.196), (5)
Alternatively, given a desired $Z_0$ and $\varepsilon_r$, the necessary $W/d$ can be computed from (3.197) in the text, and given below.

Again, (1) and (5) were obtained by curve fitting to numerically rigorous solutions. Equation (5) can be accurate to better than 1%.

---

**Example N12.1.** Design a 50-Ω microstrip on Rogers RO4003C laminate with 1/2-oz copper and a standard thickness slightly less than 1 mm.

Referring to the attached RO4003C data sheet from Rogers Corporation, we find that $\varepsilon_r = 3.38 \pm 0.05$ and $d = 0.032"$. We will ignore all losses (dielectric and metallic).

What does “1/2-oz copper” mean? Referring to the attached technical bulletin from the Rogers Corporation, copper foil thickness is more accurately measured through an areal mass. The term “1/2-oz copper” actually means “1/2 oz of copper distributed over a 1-ft$^2$ area.”

For 1-oz copper, $t = 34 \, \mu$m. For 2-oz copper, double this number and for ½-oz copper divide by 2.
We will use (3.197) to compute the required $W/d$ to achieve a 50-$\Omega$ characteristic impedance:

\[
\frac{W}{d} \approx \begin{cases} 
\frac{8e^{A}}{e^{2A} - 2} - \frac{W}{d} \leq 2 \\
\frac{2}{\pi} \left( B - 1 - \ln(2B - 1) + \frac{e_{r} - 1}{2e_{r}} \left[ \ln(B - 1) + 0.39 - \frac{0.61}{e_{r}} \right] \right) & \frac{W}{d} > 2
\end{cases}
\]

(3.197),(6)

To apply this equation, we first need to compute the constants $A$ and $B$:

\[
A = \frac{Z_{0}}{60} \sqrt{\frac{e_{r} + 1}{2}} + \frac{e_{r} - 1}{e_{r} + 1} \left( 0.23 + \frac{0.11}{e_{r}} \right) = 1.376
\]

(7)

\[
B = \frac{377\pi}{2Z_{0} \sqrt{e_{r}}} = 6.442
\]

(8)

Next, we will arbitrarily assume that $W/d < 2$ and use the simpler equation in (6). We find that

\[
\frac{W}{d} = \frac{8e^{1.376}}{e^{2.1.376} - 2} = 2.317.
\]

Is this result less than 2? The answer is no. So, we need to recompute $W/d$ using the bottom equation in (6). We find here that $W/d = 2.316$, which is greater than 2 as assumed.

So, with this result and $d = 0.032"$, then $W = 2.316 \cdot 0.032" = 0.0741"$. 
A more common unit for width and thickness dimensions in microwave circuits is “mil” where

\[ 1 \text{ mil} = \frac{1}{1000}" = 25.4 \mu\text{m} \]

Therefore,

\[ W = 0.0741" = \frac{74.1}{1000}" = 74.1 \text{ mils} (= 1.88 \text{ mm}). \]

This completes the design of the 50-Ω microstrip.
Measurement of Copper
Thickness by the Weight Method

Thickens determination of rolled and electrodeposited (ED) copper foil should always be performed using the weight method. This method provides far more accuracy than contact thickness gauges. Since the topography of treated foil varies greatly, non-contact methods are recommended.

Copper foil is manufactured and sold by weight. This method fits its original use as a building material, and has remained even after its use in electronic circuits. Today, copper foil is measured in ounces per square foot. For example, "one-ounce copper" weighs one ounce per square foot, and is 0.0014 inches or 35 μm in thickness.

The following describes a test method which is used as a standard procedure in measurement of copper thickness:

An aluminum or G-10 template 12 inches (0.305m) square is used to cut three foot-square samples from the right, center, and left sides of a roll of copper foil. A sharp blade should be used to cut as closely as possible to the edges of the template.

Each copper foil sample is folded in half twice to form a 6 inch (0.152m) square of four layers. This process is repeated to result in a 3 inch (76mm) square of sixteen layers (Figure 1). Using an analytical balance with the sample centered on it, the weight is measured to the nearest 0.001g.

The following formula is used to calculate the thickness of the sample foil:

\[
\text{Thickness} = \frac{A \times C}{B}
\]

Where:
- \(A\) = sample weight (g)
- \(B = 28.375\) g for 1-ounce copper
- \(B = 56.750\) g for 2-ounce copper
- \(C = 0.0014\) inch (35 μm) for 1-ounce copper
- \(C = 0.0028\) inch (70 μm) for 2-ounce copper

The average of the three sample sheets is taken for the average thickness across the roll.

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ROGERS
**ADS and LineCalc**

A considerable part of a microwave design engineer’s time is spent simulating the behavior of a circuit using computer tools.

We will extensively use *Advanced Design System (ADS)* from Agilent Technologies in this course. ADS 2015 can be downloaded and run from your laptop, when on campus.

One useful tool within ADS is *LineCalc*. Given the physical dimensions and material properties, *LineCalc* will accurately compute $Z_0$ and $\varepsilon_{r,e}$ for microstrip as well as for a large number of other planar waveguides. Conversely, it can synthesize a land width given the other parameters such as $Z_0$, $d$, $t$, $\varepsilon_r$, and frequency.

*LineCalc* uses accurate analysis techniques to make these calculations and can include the effect of losses, land thickness, the presence of side and top walls, etc.

Also, *LineCalc* does not assume a TEM mode. Consequently, you may find at “high” frequencies that the microstrip becomes dispersive where $Z_0$ and $\varepsilon_{r,e}$ are functions of frequency.
The effective relative permittivity calculated by LineCalc is $K_{\text{Eff}} = 2.692$, as shown above. Compare this to the value predicted by the approximate expression (1)

$$
\varepsilon_{r,e} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \frac{1}{\sqrt{1 + 12d/W}}
$$

$$
= \frac{3.38 + 1}{2} + \frac{3.38 - 1}{2} \frac{1}{\sqrt{1 + 12 \frac{32}{73.24}}} = 2.666
$$

LineCalc predicts $W = 73.25$ mil. Compare this to Example N12.1 where $W = 74.1$ mils was calculated using the accurate but approximate expression (3.197).
A Designer's Guide To Microstrip Line

Equations, data and conclusions from many sources are compiled to define transmission characteristics. Discussions center on loss, circuit Q, dispersion, dimensional ratios, and modeing.

This is not an announcement of a revolutionary new idea for microstrip circuit design. It is, however, a compendium of the best ideas presented over the past several years on design considerations for this ubiquitous transmission medium.

Microstrip technology is quite mature, offering a superior blend of performance characteristics1 to the designer of microwave integrated circuits (Table 1). Nearly 80 references at the end of this article attest to the number of investigators who have published design formulas for microstrip. So today, the problem is not that the circuit designer lacks information concerning this transmission medium, but that too much information is available, scattered throughout too many journals. What follows is an attempt to review the most useful formulas and conclusions, and arrange them in a topical, easy-to-follow order.

But first, understand the difference between microstrip and other forms of MIC transmission lines that are often erroneously referred to as “microstrip.” By definition, a microstrip transmission line consists of a strip conductor and a ground plane separated by a dielectric medium (Fig. 1a). The dielectric material serves as a structural substrate upon which the thin-film metal conductors are deposited. Conduccors are usually gold or copper.

Since field lines between the strip and the ground plane are not contained entirely in the substrate (Fig. 1b), the propagating mode along the strip is not purely transverse electromagnetic (TEM) but quasi-TEM. Assuming the quasi-TEM mode of propagation, the phase velocity in microstrip is given by

\[ V_p = \frac{c}{\sqrt{\varepsilon_{ef}}} \]

where \( c \) is the velocity of light, and \( \varepsilon_{ef} \) is the effective dielectric constant of the substrate material. The effective dielectric constant is lower than the relative dielectric constant, \( \varepsilon_r \), of the substrate, and takes into account external fields.

The wavelength, \( \lambda_p \), in microstrip line is given by

\[ \lambda_p = \frac{V_p}{f} \]

where \( V_p \) is given by Eq. 1 and \( f \) is frequency.

The characteristic impedance of the transmission line is given by

\[ Z_0 = \frac{1}{V_p} \frac{\lambda_p}{C} \]

where \( C \) is the capacitance per unit length of the line.

The analysis for the evaluation of \( V_p \) and \( C \) based on the quasi-TEM mode is fairly accurate at lower microwave frequencies. At higher frequencies, the ratio of longitudinal-to-transverse electromagnetic components becomes significant and the propagating mode can no longer be considered quasi-TEM. Analysis of this "hybrid mode" is far more rigorous.

Closed-form expressions are vital.

Early attempts to characterize the performance of microstrip lines were based on the quasi-TEM model. Various approximate methods such as conformal mappings, integral transforms, and computer-aided design (CAD) have been developed. The accuracy of these methods depends on the strip width-to-height ratio (\( \frac{w}{h} \)) and the frequency of operation. For practical applications, it is often desirable to have a simple expression that gives a good approximation of the performance of microstrip lines.

Table 1: Comparison of popular transmission media

<table>
<thead>
<tr>
<th>Transmission Mode</th>
<th>Microstrip</th>
<th>Stripline</th>
<th>Coaxial</th>
<th>Waveguide</th>
</tr>
</thead>
<tbody>
<tr>
<td>Line losses</td>
<td>high</td>
<td>high</td>
<td>medium</td>
<td>low</td>
</tr>
<tr>
<td>Inserted Q</td>
<td>low</td>
<td>low</td>
<td>medium</td>
<td>high</td>
</tr>
<tr>
<td>Power capability</td>
<td>poor</td>
<td>fair</td>
<td>good</td>
<td>poor</td>
</tr>
<tr>
<td>Isolation between</td>
<td>poor</td>
<td>fair</td>
<td>good</td>
<td>poor</td>
</tr>
<tr>
<td>Circuit</td>
<td>easy</td>
<td>very easy</td>
<td>easy</td>
<td>easy</td>
</tr>
<tr>
<td>Bandwidth</td>
<td>large</td>
<td>large</td>
<td>large</td>
<td>small</td>
</tr>
<tr>
<td>Minimization</td>
<td>excellent</td>
<td>very good</td>
<td>poor</td>
<td>poor</td>
</tr>
<tr>
<td>Mass and weight</td>
<td>large</td>
<td>medium</td>
<td>large</td>
<td>large</td>
</tr>
<tr>
<td>Realization of</td>
<td>very good</td>
<td>fair</td>
<td>poor</td>
<td>poor</td>
</tr>
<tr>
<td>passive circuitry</td>
<td>very good</td>
<td>very good</td>
<td>very good</td>
<td>very good</td>
</tr>
<tr>
<td>Integration with</td>
<td>very good</td>
<td>fair</td>
<td>poor</td>
<td>poor</td>
</tr>
<tr>
<td>strip devices</td>
<td>very good</td>
<td>very good</td>
<td>very good</td>
<td>poor</td>
</tr>
<tr>
<td>Integration with</td>
<td>very good</td>
<td>very good</td>
<td>very good</td>
<td>very good</td>
</tr>
<tr>
<td>lumped elements</td>
<td>very good</td>
<td>very good</td>
<td>very good</td>
<td>poor</td>
</tr>
</tbody>
</table>

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MICROWAVES • May, 1977
The closed form expressions for \( Z_0 \) and \( \varepsilon_{wr} \) have been reported by Wheeler\(^8\), Schneider\(^9\) and Hammerstad\(^4\). Wheeler and Hammerstad have also given an expression for \( W/h \) in terms of \( Z_0 \) and \( \varepsilon_{wr} \). For practical range of microstrip lines (0.05 \( \leq W/h \leq 20 \) and \( \varepsilon_r \leq 16 \)) Hammerstad reported that his expressions are more accurate than earlier work, and fall within ± 1 per cent of Wheeler's numerical results. His expressions, which are based on the work of Wheeler and Schneider, include useful relationships defining both characteristic impedance and effective dielectric constant:

For \( W/h \leq 1 \),

\[
Z_0 = \frac{60}{\sqrt{\varepsilon_{wr}}} \ln \left( \frac{8}{\varepsilon_r} + 0.25 \frac{W}{h} \right) \tag{4}
\]

where:

\[
\varepsilon_{wr} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left[ 1 + 12 \left( \frac{W}{h} \right)^2 \right] + 0.06 \left( \frac{W}{h} \right)^2 \tag{5}
\]

For \( W/h \geq 1 \),

\[
Z_0 = \frac{120}{\sqrt{\varepsilon_{wr}}} \ln \left( \frac{W/h + 1.888 + 0.007 \ln (W/h + 1.444)}{2} \right) \tag{6}
\]

where:

\[
\varepsilon_{wr} = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 1 + 12 \frac{W}{h} \right) \tag{7}
\]

Hammerstad notes that the maximum relative error in \( \varepsilon_{wr} \) and \( Z_0 \) is less than ± 0.2 per cent and 0.8 per cent, for 0.05 \( \leq W/h \leq 20 \) and \( \varepsilon_r \leq 16 \). His expressions for \( W/h \) in terms of \( Z_0 \) and \( \varepsilon_r \) are:

For \( W/h \leq 2 \),

\[
\frac{W}{h} = \frac{8 \exp \left( \frac{A}{2} \right)}{\exp \left( 2A \right) - 2} \tag{8}
\]

For \( W/h \geq 2 \),

\[
\frac{W}{h} = \frac{2}{\pi} \left[ B - \ln \left( 2B - 1 \right) \right] \tag{9}
\]

where:

\[
A = \frac{Z_0}{60} \sqrt{\frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2} \left( 0.23 + 0.11/\varepsilon_r \right)}
\]

\[
B = \frac{371}{2} \frac{\varepsilon_r + 1}{\sqrt{\varepsilon_r}}
\]

These expressions provide the same accuracy as Eqs. 4, 5, 6 and 7.

The results discussed above assume a two-dimensional strip conductor. But in practice, the strip is three-dimensional—its thickness must be considered. However, when \( W/h \leq 0.006 \), \( \varepsilon_r \leq 10 \), and 0.1 \( \leq W/h \leq 5 \), the agreement between experimental and theoretical \((W/h=0)\) result is excellent.\(^{10}\)

The zero-thickness \((W/h=0)\) formula given above can also be modified to consider the thickness of the strip.\(^4\) When the strip width, \( W \), is replaced by an effective strip width,

\[ W_e \text{ Expressions for } W_e \text{ are:} \]

For \( W/h \geq 1/2 \),

\[
W_e = \frac{W}{h} + \frac{1}{\ln \left( 1 + \frac{2h}{W} \right)} \tag{10}
\]

For \( W/h \leq 1/2 \),

\[
W_e = \frac{W}{h} + \frac{1}{\ln \left( 1 + \frac{4hW}{L} \right)} \tag{11}
\]

Additional restrictions for applying Eqs. 10 and 11 are \( t \geq h \) and \( t < W/2 \). Typical strip thickness varies from 0.0002 to 0.0008 inch for metalized alumina substrate, and from 0.001 to 0.003 inch for low-dielectric substrates.

Most microwave integrated circuit applications require a metallic enclosure for hermetic sealing, strength, electromagnetic shielding and ease of handling. The effect of the top and side walls on the microstrip characteristics has been studied using numerical methods.\(^{11,12,13}\) The conclusion: packaging tends to lower impedance and effective dielectric constant. This is because the fringing flux lines are prematurely terminated, which increases the density of field lines in air. But when the ratio of the distance between the lower and upper walls to substrate thickness is larger than five, and the side wall spacing is five times the strip width, the effect of enclosure is negligible on microstrip characteristics.

**Consider dispersion at higher frequencies**

The formulas for characteristic impedance and effective dielectric constant presented thus far have been based on a quasi-TEM mode of propagation. At lower frequencies, this is a good static approximation of a dynamic structure. However, at higher frequencies the effective dielectric constant and characteristic impedance of a microstrip line begins to change as frequency, increases making the transmission line dispersive.\(^{14}\) This dispersive characteristic is due to propagation of hybrid modes.

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\(^{1}\) M. E. May, 1977

\(^{2}\) M. E. May, 1977

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**2. Dispersive effects raise the effective dielectric constant (e\(_{w0}\)) slightly as frequency is increased.**
A DESIGNER'S GUIDE TO MICROSTRIP

The frequency dependence of the effective dielectric constant describes the influence of dispersion on the phase velocity, whereas the frequency dependence of the effective width describes the influence of the dispersion on the characteristic impedance. The phase velocity in microstrip line decreases with increasing frequency, hence \( v \) increases with frequency. The characteristic impedance of microstrip line increases with frequency, so the effective width must decrease with frequency.

Fortunately, changes in \( v \) and \( Z \) with frequency are very small. However, the frequency below which dispersion effects may be neglected is given by the relation

\[
f_c \text{(GHz)} = 0.3 \sqrt{\frac{Z_o}{K}} \frac{h}{c} \quad (\text{h in cm})
\]

Equation 12 shows that \( f_c \) is higher for high-impedance lines on thin substrates.

The numerical analysis for dispersion in shielded as well as open microstrip transmission lines has been treated extensively.\(^{16-29}\) These analyses, however, require extensive computations and fail to provide insight into the dominant physical phenomenon at work. For these reasons, the numerical approach is not convenient for microstrip circuit design and therefore, not discussed in this paper.

Both experimental and empirical attempts to describe microstrip dispersion have also been reported.\(^{30-34}\) The empirical expressions reported for dispersion are limited in terms of applicability and suffer from inadequate theoretical foundation. Analytical formulas for dispersion which agree closely with experimental and numerical results have appeared just recently.

These analytical expressions, by Getinger\(^{35}\) and Carlin,\(^{36}\) are very similar. But the results given by Getinger are closer to experimental as well as numerical results. The dispersion in \( \epsilon_d(f) \) is given by

\[
\epsilon_d(f) = \epsilon_e - 
\frac{\epsilon_e - \epsilon_0}{1 + (f/f_c)^2}
\]

where:

\[
f_c = \frac{Z_o}{8 \pi h}
\]
\[G = 0.6 + 0.009 Z_o\]

Here, frequency, \( f \) is in GHz and substrate thickness, \( h \), in cm. It can be seen from Eq. 13 that for \( f \gg f_c \), \( \epsilon_d(f) \approx \epsilon_0 \). In other words, high-impedance lines on thin substrates are less dispersive. Experimental and theoretical results for various microstrip geometries are compared in Fig. 2. There is a close agreement between the calculated values (Eq. 13) and experimental values.

Although many researchers have attempted to describe the effect of dispersion on \( \epsilon_d \), there are fewer analyses which deal with frequency dependent behavior of \( Z \). Krage and Haddad\(^{38}\) and Knorr and Tufekcioglu\(^{39}\), for example, describe the increase in characteristic impedance with frequency by numerical analyses. The agreement between the results given by these two analyses is reasonably good.

Recently, closed-form expressions for \( Z \) (f) based on a parallel-plate model of microstrip line have been reported.\(^{40}\) These expressions are:

\[
Z(f) = \frac{377 h}{W_{eff}(f) \sqrt{\epsilon_d(f)}}
\]

The effective width, \( W_{eff}(f) \), is given by

\[
W_{eff}(f) = W + \frac{W_{eff}(0) - W}{1 + (f/f_c)^2}
\]

where \( W_{eff}(0) \) is obtained from Eq. 14 when \( f \rightarrow 0 \).

The variation in characteristic impedance with frequency is shown in Fig. 2. The solid curve is arrived at using Eqs. 14 and 15, while the dotted curve is the one reported by Knorr and Tufekcioglu.\(^{39}\) The increase in \( Z(f) \) for \( h = 10 \) and \( W/h = 1 \) is only 4 per cent from DC to 10 GHz, which is quite small. This change cannot be confirmed experimentally since, at 10 GHz, transitions pose a considerable problem in accurate measurement. Therefore, the effect of dispersion on \( Z \) can be generally neglected.

Two mechanisms contribute to loss.

1. Attenuation constant, \( a \), is one of the most important characteristics of any transmission line. There are two sources of dissipative losses in a microstrip circuit: conductor loss and substrate dielectric loss.

Assuming a uniform current distribution across strip width and ground planes, conductor loss may be approximated as:

\[
a_c = \frac{3.88 \times 10^{-3}}{Z_o} \text{dB/cm}
\]

where \( \mu_r \) is the free space permeability and \( a \) is the conductivity of the microstrip material.

2. Surface resistivity, \( R_s \), for the conductor is given by

\[
R_s = \frac{d_{avg}}{\mu_0 \mu_r}
\]

where \( d_{avg} \) is the free space permittivity.

It should be noted that this simple expression for conductor loss is valid only for very wide strip widths (\( W/h \approx 1 \)). There is a very large discrepancy between the experimental data and this expression at practical values of \( W/h \). However, Eq. 16 can be brought closer to reality by considering the nonuniform current distribution. Expressions for the conductor loss derived by Pucel\(^{43}\) in this manner are very accurate.

For \( W/h \leq 1/2 \),

\[
a_c = \frac{3.88 \times 10^{-3}}{2 \times Z_0} \left( 1 + \frac{1}{W_{eff}} \frac{h}{t} \frac{h}{W_{eff}} \right)
\]

(Continued on p. 176)

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For $1/2 < W/h < 2$,

$$\alpha_r = \frac{8.98 R_e \cdot P \cdot Q}{2 + \frac{Z_{sh}}{h}}$$

For $W/h > 2$,

$$\alpha_r = \frac{8.98 R_e \cdot Q \cdot \left( \frac{W}{h} + 2 \ln \left( 2 + \frac{W}{h} \right) \right)}{2h}$$

$$\alpha_r = \frac{8.98 R_e \cdot Q \cdot \left( \frac{W}{h} + \frac{W}{\sqrt{h}} \right)}{2h}$$

where:

$$P = 1 - \left( \frac{W}{h} \right)^2$$

$$Q = 1 + \frac{h}{W} + \frac{h}{W} \left( \ln \frac{2h}{h} - \frac{1}{h} \right)$$

For a fixed characteristic impedance, conductor loss decreases inversely with substrate thickness and increases with the square root of frequency.

The dielectric loss increase in microstrip line is an important parameter when microwave circuits requiring small attenuation are considered. Welsh and Pratth and Schneider derived expressions for the attenuation constant for a dielectric with loss tangent, given below:

$$\alpha_d = 27.3 \frac{h}{\lambda_0} \left( \frac{\tan \delta}{\epsilon_0} \right)^{\frac{1}{\epsilon_0 - 1}} \frac{\epsilon_0 - 1}{\epsilon_0 - 1}$$

where $\lambda_0$ is the free space wavelength.

For $\epsilon_0 > 0$,

$$\alpha_d = 4.34 \frac{h}{\sqrt{\epsilon_0 (\tan \delta)}}$$

where $\epsilon_0$ is the free space permittivity.

Dielectric losses are normally very small compared with conductor losses for dielectric substrates. Dielectric losses in silicon substrates, however, are usually in the same order as, or larger than conductor losses. The reason for this is that resistivities higher than few hundred ohm-cm are difficult to maintain for Si. However, higher resistivity can be maintained in GaAs, and hence the losses are less.

Figure 4 compares the total loss for 50-ohm microstrip lines on silicon, GaAs, and quartz. It is obvious from the figure that silicon MICs are more lossy than GaAs circuits will have more loss than ceramic circuit because semiconductor substrates are thinner. Lines on quartz have the least loss.

**Quality factor depends on substrate**

The quality factor, $Q$, of a microstrip line can be related to the total losses in the line by:

$$Q_r = \frac{\beta}{2 \alpha_r}$$

where $Q_r$ is the total resonator. $\alpha_r$ is the total loss in the resonator and $\beta = 2\pi/\lambda_0$. Microstrip line $Q$s are lower than the $Q$s of coaxial or waveguide transmission line.

When the losses in a resonant line are considered (such as $\lambda/2$ or $\lambda/4$ resonators) another loss factor, $\alpha_r$, due to radiation at the discontinuities must also be considered. The corresponding radiation $Q$ factor is given by:

$$Q_r = \frac{2\pi}{4\pi} \frac{Z_0 h (\sinh \lambda_0 h)}{2(h/\lambda_0) \rho}$$

where

$$\rho = \frac{\sinh \left( \frac{\pi}{\lambda_0 h} \right)}{1 + \frac{\sinh \left( \frac{\pi}{\lambda_0 h} \right)}{\cosh \left( \frac{\pi}{\lambda_0 h} \right)}}$$

Note that the effect of dispersion is considered, as described by Eq. 13. The total $Q$ of the resonator can be expressed by

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

Here, $Q_c$ and $Q_r$ are the quality factors corresponding to conductor, dielectric and radiation losses, respectively. Finally, the circuit quality factor, $Q_c$, can be defined as

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

The variation of $Q_r$, $Q_c$, and $Q_r$ with frequency for a quarter-wave resonator on GaAs, alumina and quartz substrates is shown in Fig. 5. A quarter-wave, 50-ohm resonator on 25-mil-thick alumina substrate has a $Q_r$ of about 240 at 2 GHz and 550 at 10 GHz, whereas $Q_c$ is 230 at 2 GHz and nearly 160 at 10 GHz. This is due to the fact that radiation losses are higher than conductor and dielectric losses at higher frequencies. A quarter-wave, 50-ohm resonator on 10-mil GaAs substrate has $Q_c$ of about 82 at 2 GHz and nearly 150 at 10 GHz. This is due to the fact that radiation losses are smaller than conductor and dielectric losses for thin substrates at higher frequencies. Thus, the commonly accepted rule for high-Q microstrip circuits using thick substrates does not apply due to high radiation losses incurred under this condition.

**Moding limits high-frequency operation**

Maximum frequency of operation in microstrip line is limited by the excitation of spurious modes in the form of

(Continued on p. 109)
surface waves and transverse resonances. Surface waves are TM and TE modes which propagate across a dielectric substrate with ground plane. The frequency at which significant coupling occurs between the quasi-TEM mode and the lowest-order surface wave mode is given by:

$$f_s = \frac{c}{2 \times h} \sqrt{\frac{\varepsilon_r - \varepsilon_t - 1}{\varepsilon_t - 1}} \cdot \tan^{-1}(\varepsilon_t)$$  \hspace{1cm} (22)

For \(\varepsilon_r > 10\), Eq. 23 reduces to:

$$f_s (\text{GHz}) = \frac{10.6}{h \sqrt{\varepsilon_r}}$$  \hspace{1cm} (24)

Cutoff frequency, \(f_c\), decreases when either the substrate thickness or dielectric constant is increased.

Thus three limitations—maximum substrate thickness, minimum \(Q\) and surface wave excitation—define a region of useful microstrip line operation (see Fig. 6). From the usable region of Fig. 6, one obtains a range of substrate thickness which should be used for microstrip line circuits.

For \(\varepsilon_r = 2.7\), this range is:

- \(0.25 < h < 1.3\) cm @ 2 GHz
- \(0.01 < h < 0.36\) cm @ 10 GHz
- \(0.01 < h < 0.17\) cm @ 20 GHz

In addition to the conductor and dielectric losses (\(Q_s\)), the maximum \(Q\) of microstrip is also limited by radiation losses from discontinuities. When radiation losses are taken into account for calculation of maximum \(Q\), the plots in Fig. 6 are slightly modified. But, if packaging and circuit design techniques are employed to reduce radiation losses, the curves in Fig. 6 will remain valid.

Relate fabrication tolerances to properties

High-impedance lines, or lines on thin substrates of high dielectric constant material require very narrow conducting (Continued on p. 152)

5. The total \(Q\) of a resonator is determined by considering circuit (\(Q_c\)) and radiation (\(Q_r\)) quality factors.

6. Maximum substrate thickness, minimum \(Q\) and surface-wave modes border a useful operating region.
A DESIGNER'S GUIDE TO MICROWAVE STRIPES.

Fabrication tolerances naturally impose a limit on the accuracy of such lines. Bash and Trivedi, for example, have studied the effect of dimensional tolerances on $\epsilon_r$ and $Z_0$, with the conclusion that the maximum per cent error in $\epsilon_r$ is $0.16 \times 10^{-1}$, whereas $0.05 \times 10^{-1}$ for $Z_0$. A 1 per cent deviation in strip width of a 50-ohm line on alumina substrate gives rise to a 0.4 per cent change in $\epsilon_r$, which introduces a maximum error of 0.1 per cent in phase velocity as well as characteristic impedance.

The variation of $(\Delta \epsilon_r)/(\epsilon_r \times 10^{-1})$ with $X$ is plotted in Fig. 7. Note that the error increases with $X$ for $X > 1$, the percentage error in $Z_0$ being less than $0.5 X/10^{-1}$ whereas for $X < 1$, the error is less than $0.5 X/10^{-1}$. Microstrip lines on semiconductor substrates use substrate thicknesses of 10 mil or less. Since a 50-ohm line on silicon substrate ($X = 12.0$) requires a 0.4 per cent deviation of 7.4 mil for a typical substrate thickness of 10 mil. When the deviation in the strip width is 0.5 mil, the error in $Z_0$ is about 0.1 per cent. This indicates that 50-microstrip lines on semiconductor substrates ($X > 10$) require dimensional tolerances in the order of 0.5 mil for satisfactory performance (i.e., $\epsilon_r < 1 + X/10^{-1} < 1.0$).

Some final notes on power handling.

The peak power handling capability of microstrip line is poor. Although limited primarily by the sharp edges of the strip conductor, in some cases, connectors or launchers decide the power handling capability of the microstrip line. One additional factor which may significantly limit the power handling capability is the effect of internal mismatches. However, it has reported successful operation up to 10 kW at 2-b and 4-kW at X-band. Peak power data is not readily available and difficult to calculate accurately.

The average power capability of microstrip line is influenced by the temperature rise of the strip conductor and the supporting substrate. It is, therefore, related to the loss tangent and thermal conductivity of the substrate (low loss tangent and large thermal conductivity will increase the average power capability of microstrip lines). It has been reported that a 50-ohm alumina line (50-mil) can handle 100 W of CW power.

References


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RO4000® hydrocarbon ceramic laminates are designed to offer superior high frequency performance and low cost circuit fabrication. The result is a low loss material which can be fabricated using standard epoxy/glass (FR-4) processes offered at competitive prices.

The selection of laminates typically available to designers is significantly reduced once operational frequencies increase to 500 MHz and above. RO4000 material possesses the properties needed by designers of RF microwave circuits and matching networks and controlled impedance transmission lines. Low dielectric loss allows RO4000 series material to be used in many applications where higher operating frequencies limit the use of conventional circuit board laminates. The temperature coefficient of dielectric constant is among the lowest of any circuit board material (Chart 1), and the dielectric constant is stable over a broad frequency range (Chart 2). This makes it an ideal substrate for broadband applications.

RO4000 material's thermal coefficient of expansion (CTE) provides several key benefits to the circuit designer. The expansion coefficient of RO4000 material is similar to that of copper which allows the material to exhibit excellent dimensional stability, a property needed for mixed dielectric multilayer board constructions. The low Z-axis CTE of RO4000 laminates provides reliable plated through-hole quality, even in severe thermal shock applications. RO4000 series material has a Tg of >280°C (536°F) so its expansion characteristics remain stable over the entire range of circuit processing temperatures.

RO4000 series laminates can easily be fabricated into printed circuit boards using standard FR-4 circuit board processing techniques. Unlike PTFE based high performance materials, RO4000 series laminates do not require specialized via preparation processes such as sodium etch. This material is a rigid, thermoset laminate that is capable of being processed by automated handling systems and scrubbing equipment used for copper surface preparation.

RO4003™ laminates are currently offered in various configurations utilizing both 1080 and 1674 glass fabric styles, with all configurations meeting the same laminate electrical performance specification. Specifically designed as a drop-in replacement for the RO4003C™ material, RO4350B™ laminates utilize RoHS compliant flame-retardant technology for applications requiring UL 94V-0 certification. These materials conform to the requirements of IPC-4103, slash sheet /10 for RO4003C and /11 for RO4350B materials.
Chart 1. RO4000® Series Materials
Dielectric Constant vs. Temperature

Chart 2: RO4000 Series Materials
Dielectric Constant vs. Frequency

Chart 3: Microstrip Insertion Loss
(0.030” Dielectric Thickness)
<table>
<thead>
<tr>
<th>Property</th>
<th>Typical Value</th>
<th>Direction</th>
<th>Units</th>
<th>Condition</th>
<th>Test Method</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Dielectric Constant, ( \varepsilon_r )</strong> (Process specification)</td>
<td>3.38 ±0.05</td>
<td>Z</td>
<td>--</td>
<td>10 GHz/23°C</td>
<td>IPC-TM-650 2.5.5.5</td>
</tr>
<tr>
<td><strong>Dielectric Constant, ( \varepsilon_r )</strong> (Recommended for use in circuit design)</td>
<td>3.55</td>
<td>Z</td>
<td>--</td>
<td>10 GHz/23°C</td>
<td>2.5.5.6 Full Sheet Resonance</td>
</tr>
<tr>
<td>Dissipation Factor, ( \tan \delta )</td>
<td>0.0027 - 0.0031</td>
<td>Z</td>
<td>ppm/°C</td>
<td>-50°C to 150°C</td>
<td>IPC-TM-650 2.5.5.5</td>
</tr>
<tr>
<td>Volume Resistivity</td>
<td>1.7 X 10¹⁰</td>
<td>MΩ•cm</td>
<td>COND A</td>
<td>2.5.17.1</td>
<td></td>
</tr>
<tr>
<td>Surface Resistivity</td>
<td>4.2 X 10⁹</td>
<td>Ω</td>
<td>COND A</td>
<td>2.5.17.1</td>
<td></td>
</tr>
<tr>
<td>Electrical Strength</td>
<td>31.2 (780)</td>
<td>KV/mm (V/mil)</td>
<td>0.51mm (0.020&quot;)</td>
<td>IPC-TM-650 2.5.6.2</td>
<td></td>
</tr>
<tr>
<td>Tensile Modulus</td>
<td>26,889 (3900)</td>
<td>MPa (kpsi)</td>
<td>RT</td>
<td>ASTM D638</td>
<td></td>
</tr>
<tr>
<td>Tensile Strength</td>
<td>141 (20.4)</td>
<td>MPa (kpsi)</td>
<td>RT</td>
<td>ASTM D638</td>
<td></td>
</tr>
<tr>
<td>Flexural Strength</td>
<td>276 (40)</td>
<td>MPa (kpsi)</td>
<td>RT</td>
<td>IPC-TM-650 2.4.4</td>
<td></td>
</tr>
<tr>
<td>Dimensional Stability</td>
<td>&lt;0.3 &lt;0.5</td>
<td>mm/m (mils/inch)</td>
<td>after etch +E2/150°C</td>
<td>IPC-TM-650 2.4.39A</td>
<td></td>
</tr>
<tr>
<td>Coefficient of Thermal Expansion</td>
<td>11 14 46</td>
<td>ppm/°C</td>
<td>-55 to 288°C</td>
<td>IPC-TM-650 2.1.41</td>
<td></td>
</tr>
<tr>
<td>Tg</td>
<td>&gt;280 &gt;280</td>
<td>°C</td>
<td>DSC</td>
<td>IPC-TM-650 2.4.24</td>
<td></td>
</tr>
<tr>
<td>Td</td>
<td>425 390</td>
<td>°C</td>
<td>TGA</td>
<td>ASTM D3850</td>
<td></td>
</tr>
<tr>
<td>Thermal Conductivity</td>
<td>0.71 0.69</td>
<td>W/m/°K</td>
<td>80°C</td>
<td>ASTM C518</td>
<td></td>
</tr>
<tr>
<td>Moisture Absorption</td>
<td>0.06 0.06</td>
<td>%</td>
<td>48 hrs immersion @ 0.060&quot; @ sample temperature 50°C</td>
<td>ASTM D570</td>
<td></td>
</tr>
<tr>
<td>Density</td>
<td>1.79 1.86</td>
<td>gm/cm³</td>
<td>23°C</td>
<td>ASTM D792</td>
<td></td>
</tr>
<tr>
<td>Copper Peel Strength</td>
<td>1.05 (6.0)</td>
<td>N/mm (pli)</td>
<td>after solder float 1 oz EDC Foil</td>
<td>IPC-TM-650 2.4.8</td>
<td></td>
</tr>
<tr>
<td>Flammability</td>
<td>N/A ^V-0</td>
<td></td>
<td>UL 94</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Typical values are a representation of an average value for the population of the property. For specification values contact Rogers Corporation.

RO4000 LoPro laminate uses a modified version of RO4000 resin system to bond reverse treated foil. Values shown above are RO4000 laminates with out the addition of the LoPro resin. For double-sided board, the LoPro foil results in a thickness increase of approximately 0.0007" (0.018mm) and the DK is approximately 2.4. Therefore, effective DK is highly dependent on core thickness.

Prolonged exposure in an oxidative environment may cause changes to the dielectric properties of hydrocarbon based materials. The rate of change increases at higher temperatures and is highly dependent on the circuit design. Although Rogers' high frequency materials have been used successfully in innumerable applications and reports of oxidation resulting in performance problems are extremely rare, Rogers recommends that the customer evaluate each material and design combination to determine fitness for use over the entire life of the end product.
The information in this data sheet is intended to assist you in designing with Rogers' circuit material laminates. It is not intended to and does not create any warranties express or implied, including any warranty of merchantability or fitness for a particular purpose or that the results shown on this data sheet will be achieved by a user for a particular purpose. The user should determine the suitability of Rogers' circuit material laminates for each application.

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