MICROSTRIP LINES

TRANSVERSE ELECTROMAGNETIC TRANSMISSION LINES

One of the most familiar waveguiding structures is the conventional transmission line such as the two-wire line and the coaxial line. The fundamental mode of propagation on a transmission line is essentially a *transverse electromagnetic* (TEM) *wave,* which owns neither electric nor magnetic field in the direction of propagation (1).

An ideal lossless uniform TEM transmission line can be represented by a lumped-circuit and consists of series inductance L and shunt capacitance C, all defined per unit length of the line, as shown in Fig. 1. The inductance L is proportional to the permeability μ of the surrounding medium, and the capacitance C proportional to the permittivity ϵ of the medium. Their values depend on the transverse geometry of the transmission line, and are determined from the electrostatic analysis (1) of the cross-section of the structure that solves a

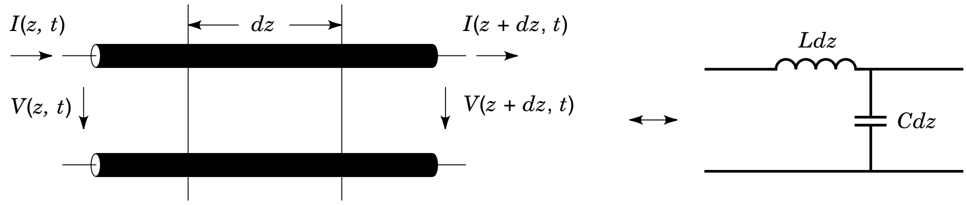

Figure 1. Lumped-circuit representation of an ideal TEM transmission line. L is the series inductance, C the shunt capacitance; both are defined per unit length of the line.

two-dimensional Laplace equation in the medium surrounding the conductors of the transmission line.

The voltage and current waves, expressed by $V = V_0e^{\pm j\beta z}$ and $I = I_0e^{\pm j\beta z}$, along the transmission line are solutions of the *telegraphists*' or *transmission-line equations* (1).

$$\frac{d^2V}{dz^2} + \omega^2LCV = 0 \tag{1}$$

$$\frac{d^2I}{dz^2} + \omega^2LCI = 0 \tag{2}$$

where $\omega = 2\pi f$ (frequency) is the radian frequency, and β is the phase constant.

TEM transmission lines are characterized by line parameters such as phase constant, characteristic impedance, attenuation constant, and so on. Formulas for line parameters of an ideal lossless transmission line are given in what follows (1,2)

Phase constant

$$\beta = \omega\sqrt{LC} = \omega\sqrt{\mu\epsilon} = \omega/v \tag{3}$$

Characteristic impedance

$$Z_0 = \sqrt{\frac{L}{C}} = \frac{1}{vC} \tag{4}$$

Voltage and current along line

$$V(z) = V_0e^{j\beta z}[1 + \Gamma(z)] \tag{5}$$

$$I(z) = \frac{V_0}{Z_0}e^{j\beta z}[1 - \Gamma(z)] \tag{6}$$

Input impedance

$$Z(z) = Z_0\frac{Z_L\cos\beta z + jZ_0\sin\beta z}{Z_0\cos\beta z + jZ_L\sin\beta z} \tag{7}$$

In the above expressions, v is the velocity of TEM waves in the dielectric of line, v_0 is the amplitude of the incident voltage, z_L is the value of the load impedance, and z is the distance along the line from the load end.

STRIPLINES

A stripline, also referred to as a *triplate line*, consists of a conducting strip lying between, and parallel to, two wide conducting ground planes, as shown in Fig. 2. The region between the strip and the planes is filled with a uniform dielectric. Stripline is one of the most commonly used transmission lines for passive *microwave integrated circuits* (MICs). The fundamental mode in a stripline is a TEM mode, and its field distribution is illustrated in Fig. 3.

The line parameters of a stripline can be obtained completely by electrostatic analysis such as the conformal mapping technique (1,3). An approximate expression for the characteristic impedance of a stripline with zero-thickness strip is given by (2,3)

$$Z_0 = \frac{30\pi}{\sqrt{\epsilon_r}}\frac{K'(k)}{K(k)} \tag{8}$$

where ϵ_r is the relative permittivity of the dielectric filled in the stripline, $k = \tanh(\pi w/2h)$, K represents a complete elliptic function of the first kind, and K' its complementary function. The velocity of the TEM mode in a stripline is $v = 1/\sqrt{\mu_0\epsilon_0\epsilon_r}$. Both the wave velocity and characteristic impedance are independent of frequency.

An accurate but simple approximate expression for $K(k)/K'(k)$ is given by

$$\frac{K(k)}{K'(k)} = \begin{cases} \dfrac{\pi}{\ln[2(1+\sqrt{k'})/(1-\sqrt{k'})]} & \text{for } 0 \le k \le 0.707 \\ \dfrac{1}{\pi}\ln[2(1+\sqrt{k})/(1-\sqrt{k})] & \text{for } 0.707 \le k \le 1 \end{cases} \tag{9}$$

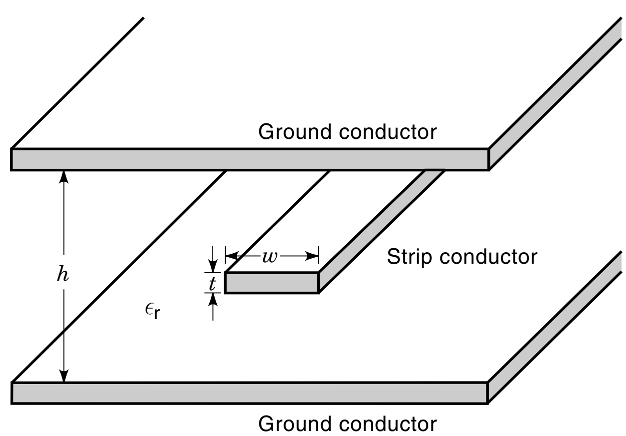

Figure 2. Geometry of a stripline. The strip conductor is sandwiched between two wide parallel conducting ground planes. The region between the strip and the ground planes is filled with a uniform dielectric.

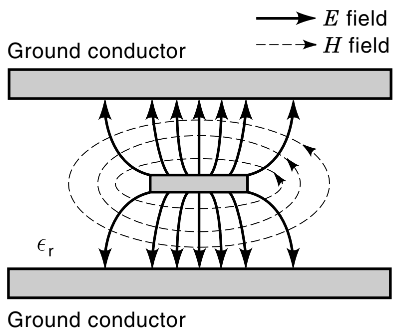

Figure 3. Electromagnetic field distribution in a stripline. The fundamental TEM mode is considered. The electric field goes from the conductor strip to the grounded planes. The magnetic field surrounds the conductor strip.

An approximate expression for the attenuation constant arising from the conductor surface resistance, $R_s = \sqrt{\omega\mu_0/2\sigma}$, is given by (1)

$$\alpha_c = \frac{R_s}{h\eta}\left[\frac{\pi w/h + \ln(4h/\pi t)}{\ln 2 + \pi w/2h}\right] \qquad \text{Nepers/m} \tag{10}$$

where $\eta = \sqrt{\mu_0/\epsilon_0\epsilon_r} = \eta_0/\sqrt{\epsilon_r}$, $\eta_0 = 120\pi$ is the wave impedance in free-space, and t is the thickness of the strip. Equation (10) is valid for $w > 2h$ and $t < h/10$.

The attenuation constant from lossy dielectric medium with $\epsilon = \epsilon' - j\epsilon''$ is expressed by

$$\begin{aligned}\alpha_d &= \mathrm{Re}\big(j\omega\sqrt{\mu_0\epsilon}\,\big) = \mathrm{Re}\big(j\omega\sqrt{\mu_0(\epsilon' - j\epsilon'')}\big)\\ &\approx \frac{\omega\sqrt{\mu_0\epsilon'}}{2}\frac{\epsilon''}{\epsilon'} = \frac{\omega\sqrt{\mu_0\epsilon'}}{2}\tan\delta \qquad \text{Nepers/m}\end{aligned} \tag{11}$$

where $\tan\delta$ is the dielectric loss tangent.

In addition to the dominant TEM mode, higher order *transverse electric* (TE) *modes* and *transverse magnetic* (TM) *modes* can also propagate in a stripline. A TE mode owns magnetic field but no electric field in the direction of propagation. A TM mode contains electric field but no magnetic field in the direction of propagation. The cutoff frequency of the lowest order TE mode is (2)

$$f_c = \frac{15}{h\sqrt{\epsilon_r}}\frac{1}{w/h + \pi/4} \tag{12}$$

where f_c is given in gigahertz, w and h are in centimeters.

More detailed and accurate formulas for the stripline parameters, such as the characteristic impedance, attenuation constant, and so on, can be found in Ref. 2.

MICROSTRIP LINES

General Descriptions

Microstrip Geometry and Advantages. A microstrip line (4) is a type of open planar transmission line that consists of a dielectric substrate medium with a ground plane on the lower side and a conducting strip on the upper side. The geometry of a microstrip line is shown in Fig. 4. The substrate provides mechanical rigidity and permits the accurate positioning of the circuitry. The transmission line characteristic parameters, like the phase constant and characteristic impedance, can be determined from the substrate permittivity (ϵ_r) and the geometrical dimensions (strip width w and thickness t, substrate thickness h) in the transverse plane. For this reason, various types of microstrip circuits can be fabricated conveniently with high precisions by employing the simple photolithographic and photo-etching techniques. Use of these techniques at microwave and millimeter wave frequencies has led to the development of hybrid and monolithic microwave integrated circuits (MICs) (5–11).

Microstrip line is now one of the most widely used transmission lines for MICs. Active devices (diodes and transistors), lumped circuit elements (capacitors, resistors, inductors), dielectric resonators, and antennas can be easily incorporated into the circuit. Compared with the traditional bulky and heavy metallic waveguides and coaxial lines, the planar microstrip structures are small size, lightweight, easy for mass production, and inexpensive.

Dielectric Substrate. The properties of the dielectric substrate material affect the overall performance of the microstrip structures. Different substrate materials possess characteristics which may make them better suited for an application. For instance, higher dielectric constant materials are preferred in order to achieve a very compact microwave circuit, while lower dielectric constant materials are required for antenna structures to ensure efficient radiation.

In general, the substrate material parameters, permittivity ϵ and permeability μ, should be homogeneous (independent of position), isotropic (independent of wave propagation direction), and should have low dispersion. The loss tangent should be small to reduce energy dissipation. Furthermore, these parameters should have very small variation with temperature to ensure circuit stability. The substrate thermal conductivity should be high enough to ensure efficient removal of heat from power transistors, attenuators, and loads in high-power applications. In high-power applications, a high breakdown voltage is also desirable. The thermal expansion coefficient of the material should be similar to that of the deposited conductors and housing to withstand temperature fluctuations and improve reliability. The material must allow drilling, cutting, machining, and etching for easy workability and lower production costs. Also important is a good surface

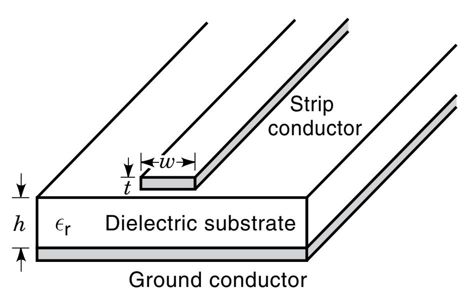

Figure 4. Geometry of a microstrip line. The conducting strip is placed above a dielectric substrate, which is supported on its bottom by a conducting plate.

finish to ensure good conductor adhesion and reduce conductor loss.

A wide variety of dielectric substrates are now commercially available. Characteristics, including mechanical and thermal as well as electronic facets, of a number of representative substrate materials, such as alumina, fused quartz, silicon, gallium arsenide, synthetic, and composite materials are compared in (6–12).

Although the microstrip line appeared first in 1952 (4), the rapid increase of the use of microstrip circuits was seen during the 1960s when high permittivity and low-loss dielectric substrates became available. At the same time, microwave semiconductor devices appeared, and miniature lumped elements (capacitors, resistors, inductors) became available for implantation on plantar circuits. Coupled with steady advances in photolithographic technology, the combination of microstrips, lumped elements, and semiconductors led to the advent of microwave integrated circuits (MICs) (5–11).

Field Configuration and Analysis Methods. The microstrip line is an inhomogeneous transmission line, involving an abrupt dielectric interface between the substrate and the air above it. The electromagnetic fields extend over inhomogeneous regions, partly in the dielectric substrate, and partly in the air, as shown in Fig. 5. Waves propagating along the line cannot be purely TEM, TE, or TM modes, but *hybrid modes* containing both electric and magnetic fields in the transverse and longitudinal directions of propagation (6). This may cause some complication in microstrip analysis and design. However, in many practical situations, the dominant mode of a microstrip line resembles closely a TEM mode. Therefore, it is usually referred to as quasi-TEM mode.

The microstrip line has been analyzed by numerous workers using various analytical and numerical techniques. As with many other transmission lines, the analysis methods for a microstrip line are aimed at determining the characteristic impedance and propagation constant (phase velocity and attenuation constant). The various methods of microstrip line analysis can be divided into two main groups (6–10). In the first group, which comprises *quasi-static methods*, the nature of the mode of propagation is considered to be pure TEM and the microstrip characteristics are calculated from the electrostatic capacitance of the structure. The quasi-static methods commonly used include the conformal transformation method (13), the variational method (14,15), the finite-difference method, and the integral equation method and others (6–10).

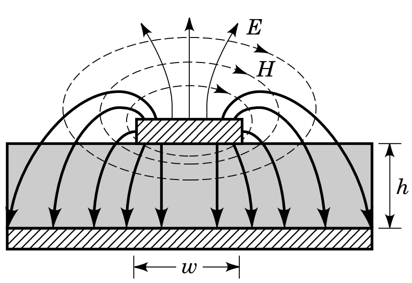

Figure 5. Electromagnetic field distribution in a microstrip line. The fundamental quasi-TEM mode is considered. The electromagnetic fields extend over inhomogeneous regions, partly in the dielectric substrate, and partly in the air.

It is found that this quasi-static analysis is adequate for designing circuits at lower frequencies where the strip width and the substrate thickness are much smaller than the wavelength in the dielectric material.

The methods in the second group are *full-wave approaches* which take into account the hybrid nature of the mode of propagation. They include the integral equation method (16), the spectral domain method (17), and the finite-difference time-domain (FDTD) method and others (6–10,17). The full-wave analysis methods are more rigorous and can predict frequency-dependent variation of the microstrip characteristics. However, they are analytically complex, and usually require large computer memories and long computation time, which may become prohibitive when optimization process is demanded in the design of circuits.

High-Frequency Problems and Quasi-Transverse Electromagnetic Wave Results. Microstrip lines have been extensively used for MICs at frequencies ranging from hundreds of megahertz to tens of gigahertz. At high frequencies, particularly into the millimeter wavelength ranges, conductor ohmic and dielectric losses increase greatly. Surface waves of the dielectric substrate and higher order modes excited at discontinuities start to propagate or to radiate (6–10). The conductor and dielectric losses and radiation reduce the amplitude of a signal propagating along the line and may cause spurious coupling between neighboring parts of a circuit. The simultaneous propagation of several modes, with different velocities, produces a distortion of the signal. The fabrication tolerances become very difficult to meet. These factors prohibit the extensive use of microstrip lines at high frequencies.

For the majority of microstrip lines suitable for MICs, the statically derived results are quite accurate when the frequency is below a few gigahertz. At higher frequencies, up to the limits for the useful operation of microstrip lines, these static results can still be used in conjunction with some quasi-empirical functions in closed formulae to find the variations of microstrip characteristics with frequency. Therefore, results by the static techniques are powerful and significant in the design of microstrip circuits.

Quasi-Transverse Electromagnetic Wave Parameters

Under the quasi-TEM wave approximation, an ideal microstrip line can be represented by the lumped-circuit for a TEM transmission line shown in Fig. 1. The characteristic impedance is expressed by

Characteristic impedance

$$Z_0 = \sqrt{\frac{L}{C}} = \frac{1}{vC} \tag{13}$$

When the substrate of the microstrip line is replaced by air, we have an air-filled line along which the wave will travel at c, the velocity of light in free space. The characteristic impedance of this air-filled line, defined as Z_a, is given by

$$Z_a = \sqrt{\frac{L}{C_a}} = \frac{1}{c \cdot C_a} \tag{14}$$

Combining Eqs. (13) and (14), we get

$$Z_0 = \frac{1}{c\sqrt{C \cdot C_a}} \tag{15}$$

Phase constant

$$\beta = \frac{\omega}{v} = \omega\sqrt{LC} = \frac{\omega}{c}\sqrt{\frac{C}{C_a}} = \beta_0\sqrt{\frac{C}{C_a}} \tag{16}$$

here β_0 is the wavenumber in free space. The normalized phase constant is written as

$$\frac{\beta}{\beta_0} = \frac{c}{v} = \frac{\lambda_0}{\lambda_g} = \sqrt{\frac{C}{C_a}} \tag{17}$$

here λ_0 and λ_g are the wavelengths in free-space and along the microstrip line, respectively. Equations (15) and (17) indicate that the characteristic impedance and phase constant of a microstrip line can be obtained if we can evaluate the capacitance per unit length of the line, with and without the presence of the dielectric substrate.

The *effective permittivity* ϵ_e is defined as the capacitance ratio

$$\epsilon_e = \frac{C}{C_a} \tag{18}$$

From Eq. (17) we immediately get

$$\epsilon_e = \left(\frac{c}{v}\right)^2 \tag{19}$$

$$\beta = \sqrt{\epsilon_e} \cdot \beta_0 \tag{20}$$

The effective permittivity has a physical meaning that the original inhomogeneous microstrip line is replaced by an equivalent homogeneous line with conductors having exactly the same geometry (w, h, t), surrounded by a single homogeneous dielectric of effective permittivity ϵ_e.

From Eqs. (17) and (18), we also have a formula

$$\lambda_g = \frac{\lambda_0}{\sqrt{\epsilon_e}} \tag{21}$$

Formulas for Quasi-Transverse Electromagnetic Wave Parameters. As stated previously, closed formulas are of significant importance in the design of microstrip line circuits. Various workers have reported formulas for microstrip calculations (6–10). These formulas may be classified into two types, one for the analysis purpose, and the other for the synthesis purpose. When the geometrical and material parameters (w, h, t, ϵ_r) are known, we use the analysis formulas to evaluate the line's electrical characteristics ϵ_e, Z_0, and λ_g. Conversely, when Z_0 and ϵ_r are given and we want to specify the width-height ratio w/h of the microstrip line, we employ the synthesis formulas.

Analysis Formulas (w/h and ϵ_r Given). Very accurate formulas derived by Hammerstad and Jensen (18) are provided below.

Effective Permittivity ϵ_e

$$\epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2}\left(1 + 10\frac{h}{w}\right)^{-a(w/h)\cdot b(\epsilon_r)} \tag{22}$$

with

$$\begin{aligned} a\left(\frac{w}{h}\right) &= 1 + \frac{1}{49}\ln\left\{\frac{(w/h)^4 + [w/(52h)]^2}{(w/h)^4 + 0.432}\right\} \\ &\quad + \frac{1}{18.7}\ln\left[1 + \left(\frac{w}{18.1h}\right)^3\right] \end{aligned} \tag{23}$$

$$b(\epsilon_r) = 0.564\left(\frac{\epsilon_r - 0.9}{\epsilon_r + 3}\right)^{0.053} \tag{24}$$

Characteristic Impedance Z_0

$$Z_0 = \frac{\eta_0}{2\pi\sqrt{\epsilon_e}} \cdot \ln\left[F_1\left(\frac{w}{h}\right)\frac{h}{w} + \sqrt{1 + \left(\frac{2h}{w}\right)^2}\right] \tag{25}$$

with

$$F_1\left(\frac{w}{h}\right) = 6 + (2\pi - 6)\exp\left[-\left(30.666\frac{h}{w}\right)^{0.7528}\right] \tag{26}$$

where $\eta_0 = 120\pi$ is the wave impedance in free-space. The accuracy of these expressions is better than 0.01% for $w/h \le 1$ and 0.03% for $w/h \le 1000$.

Synthesis Formulas (Z_0 and ϵ_r Given) (7). For narrow strips (when $Z_0 > (44 - 2\epsilon_r)\ \Omega$):

$$\frac{w}{h} = \left(\frac{\exp(A)}{8} - \frac{1}{4\exp(A)}\right)^{-1} \tag{27}$$

where

$$A = \frac{Z_0\sqrt{2(\epsilon_r + 1)}}{119.9} + \frac{1}{2}\left(\frac{\epsilon_r - 1}{\epsilon_r + 1}\right)\left(\ln\frac{\pi}{2} + \frac{1}{\epsilon_r}\ln\frac{4}{\pi}\right)$$

For wide strips (when $Z_0 < (44 - 2\epsilon_r)\ \Omega$):

$$\begin{aligned} \frac{w}{h} &= \frac{2}{\pi}[(B - 1) - \ln(2B - 1)] \\ &\quad + \frac{\epsilon_r - 1}{\pi\epsilon_r}\left[\ln(B - 1) + 0.293 - \frac{0.517}{\epsilon_r}\right] \end{aligned} \tag{28}$$

where

$$B = \frac{59.95\pi^2}{Z_0\sqrt{\epsilon_r}}$$

Strip Thickness Correction. In correcting the above results, ϵ_e and Z_0 given by Eqs. (22) and (25), for nonzero strip thickness t, a corrected strip width w_e/h is defined as follows (18):

$$\begin{aligned} \frac{w_e}{h} &= \frac{w}{h} + \frac{1}{2\pi}\frac{t}{h}\left[1 + \frac{1}{\cosh(\sqrt{\epsilon_r - 1})}\right] \\ &\quad \cdot \ln\left[1 + \frac{4h\exp(1)}{t\coth^2(\sqrt{6.517w/h})}\right] \end{aligned} \tag{29}$$

With this corrected strip width, the effect of strip thickness on ϵ_e and Z_0 of microstrip lines can be included in the Eqs. (22) and (25). We have therefore

$$\epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2}\left(1 + 10\frac{h}{w_e}\right)^{-a(w_e/h)\cdot b(\epsilon_r)} \tag{30}$$

$$Z_0 = \frac{\eta_0}{2\pi\sqrt{\epsilon_e}} \cdot \ln\left[F_1\left(\frac{w_e}{h}\right)\frac{h}{w_e} + \sqrt{1 + \left(\frac{2h}{w_e}\right)^2}\right] \tag{31}$$

where $\eta_0 = 120\pi$ is the wave impedance in free-space. In the above expressions, the functions $a(w_e/h)$, $b(\epsilon_r)$, and $F_1(w_e/h)$ are defined in Eqs. (23), (24), and (26), respectively, with the normalized strip width w/h replaced by the corrected normalized strip width w_e/h. It is observed that the effect of the strip thickness on ϵ_e and Z_0 is insignificant for small values of t/h. However, the effect of strip thickness is significant on conductor loss in the microstrip line.

Effect of Dispersion. The effect of frequency (dispersion) on ϵ_e and Z_0 has been described in a number of publications. The accurate expressions in (18) for $Z_0(f)$ and in (19) for $\epsilon_e(f)$ are

$$Z_0(f) = Z_0 \cdot \frac{\epsilon_e(f) - 1}{\epsilon_e - 1} \cdot \sqrt{\frac{\epsilon_e}{\epsilon_e(f)}} \tag{32}$$

$$\epsilon_e(f) = \epsilon_r - \frac{\epsilon_r - \epsilon_e}{1 + (f/f_{50})^m} \tag{33}$$

where

$$f_{50} = \frac{f_{\mathrm{k,TM_0}}}{0.75 + [0.75 - (0.332/\epsilon_r^{1.73})]w/h}$$

$$f_{\mathrm{k,TM_0}} = \frac{c \cdot \tan^{-1}\left(\epsilon_r\sqrt{\dfrac{\epsilon_e - 1}{\epsilon_r - \epsilon_e}}\right)}{2\pi h\sqrt{\epsilon_r - \epsilon_e}}$$

$$m = m_0 m_c$$

$$m_0 = 1 + \frac{1}{1 + \sqrt{w/h}} + 0.32\left(\frac{1}{1 + \sqrt{w/h}}\right)^3$$

$$m_c = \begin{cases} 1 + \dfrac{1.4}{1 + w/h}\left[0.15 - 0.235\exp\left(-\dfrac{0.45f}{f_{50}}\right)\right] & \text{for } w/h \le 0.7 \\ 1 & \text{for } w/h > 0.7 \end{cases}$$

Z_0, ϵ_e are the quasi-TEM wave values obtained earlier, and c is the velocity of light in free space.

Effect of Enclosure. Most microstrip circuit applications require a metallic enclosure for hermetic sealing, mechanical strength, electromagnetic shielding, mounting connectors, and ease of handling. Because the fringing electromagnetic fields are prematurely terminated on the enclosure walls, both the top cover and side walls tend to lower impedance and effective dielectric constant. Equations for evaluating the effect of the top cover and side walls on Z_0 and ϵ_e are provided in Refs. 6 and 7.

When the top cover and side walls are placed in close vicinity of the microstrip line, the line parameters may vary significantly. It is recommended to leave at least ten times the substrate thickness between the circuit and the cover (11).

A metal box is actually a microwave cavity, which resonates at some particular frequencies corresponding to its resonant modes. The size of the box should be chosen in such a way that the signal frequency does not coincide with the resonant frequencies. When this cannot be done, the resonant modes can be damped by placing absorbing materials at carefully selected locations.

Attenuation. The attenuation constant of a transmission line is usually defined as

$$\alpha \approx \frac{P_{\mathrm{loss}}}{2P(z)} = \frac{\text{Power loss per unit length}}{2(\text{Power transmitted})} \tag{34}$$

Attenuation in microstrip lines is caused by three loss components: conductor loss, dielectric loss, and radiation loss. They are discussed briefly next.

Ohmic Losses. Within the conductors, these result from the finite conductivity σ of the metal. The following approximate expression (9) is found sufficient in most situations:

$$\alpha_c \approx 8.686\frac{R_s}{wZ_0} \qquad \text{dB/m} \tag{35}$$

where $R_s = \sqrt{\omega\mu/2\sigma}$ is the metal wall surface resistance.

Dielectric Losses. Produced by the energy dissipated within the substrate, these are proportional to its dielectric loss tangent $\tan\delta$ (9)

$$\alpha_d \approx 27.3 \cdot \frac{\epsilon_e - 1}{\epsilon_r - 1} \cdot \frac{\epsilon_r}{\sqrt{\epsilon_e}} \cdot \frac{f}{c} \cdot \tan\delta \qquad \text{dB/m} \tag{36}$$

where c is the velocity of light in free space, and f is the frequency. The dielectric loss due to the substrate is normally very small compared with the conductor loss. However, the dielectric loss in silicon substrates (used for monolithic MICs) is usually of the same order or even larger than the conductor loss because of the large loss tangent $\tan\delta$ of the silicon wafers.

Radiation Losses. An infinite straight microstrip line propagating the dominant quasi-TEM mode does not radiate. However, at every discontinuity, higher order modes are excited, some of which will radiate part of the power.

Frequency Range of Operation

Like any other transmission line, microstrip lines cannot be utilized above a certain upper frequency limit. The maximum frequency of operation of a microstrip line is limited due to several factors such as effects of dispersion, excitation of higher order modes, and radiation losses.

The frequency at which significant coupling occurs between the quasi-TEM mode and the lowest order TM surface wave is given by (20)

$$f_T = \frac{300\tan^{-1}(\epsilon_r)}{\pi h\sqrt{2\epsilon_r - 2}} \tag{37}$$

where f_T is in gigahertz and h is in millimeters.

For a sufficiently wide microstrip line, a transverse-resonant mode can exist which can also couple strongly to the quasi-TEM microstrip mode. By employing the transverse-resonance equivalent circuit model, and taking into account the microstrip side-fringing effect, the cut-off frequency of the transverse-resonant mode can be easily derived (20) as follows:

$$f_c \approx \frac{300}{\sqrt{\epsilon_r}(2w + 0.8h)} \tag{38}$$

where f_c is in gigahertz and w and h are in millimeters.

For a microstrip line, radiation losses become significant at frequencies higher than (9)

$$f = 2.14 \cdot \frac{(\epsilon_r)^{0.25}}{h} \tag{39}$$

where f is in gigahertz and h is in millimeters.

Power Handling Capability

Although microstrip lines are not as well suited for high-power application as are waveguides or coaxial lines of comparable cross section, they can be used for some medium-power applications. A 50 Ω microstrip on a 25 mil thick alumina substrate can handle a few kilowatts of power (6).

The power handling capacity of a microstrip line, like that of any other dielectric filled transmission line, is limited by dielectric breakdown and by heating caused by ohmic and dielectric losses. An increase in the temperature due to conductor and dielectric losses limits the average power handling capability of the microstrip line, while the breakdown between the strip conductor and ground plane limits the peak power handling capability.

The average power handling capability of microstrip lines is determined by the temperature rise of the conductor strip and the dielectric substrate. The transmission line losses, the thermal conductivity of the substrate material, the surface area of the strip, and the temperature of the medium surrounding the microstrip line are the main factors determining the average power handling capability of microstrip lines. Therefore, dielectric substrates with low-loss tangent and larger thermal conductivity will enhance the average power handling capability of microstrip lines.

The peak voltage that can be applied without causing dielectric breakdown determines the peak power handling capability of microstrip lines. Thick substrate can support higher voltages. Therefore, low impedance lines and lines on thick substrates have generally larger peak power handling capability.

Other Types of Microstrip Lines

There are several derivatives of microstrip lines being used in MICs. These include inverted and suspended microstrip lines, a multilayered microstrip, a thin film microstrip, and a valley microstrip line. These structures are briefly described in Refs. 6 and 7.

OTHER TOPICS

In actual microstrip circuits, various types of transmission line discontinuities, such as open ends, gaps, steps in width, bends, T-junctions, and cross-junctions are frequently encountered. In the design of microstrip circuits, a complete understanding of the characteristics of microstrip discontinuities included in the circuits is necessary. At low frequencies, discontinuities can be represented by lumped-element equivalent circuits based on quasi-static models. However, at high frequencies, a more rigorous characterization of frequency-dependent parameters like the scattering parameters is required. Various methods including quasi-static and full-wave analysis methods for characterizing microstrip discontinuities are described in Refs. 6–10 and 17.

Microstrip line is now the most widely used structure for microwave systems in radar and communications purposes. Examples of passive circuits include filters, impedance transformers, hybrids, couplers, power dividers/combiners, delay lines, baluns, and circulators. Amplifiers, oscillators, and mixers employing solid state devices (diodes and transistors) constitute the other class. Microstrip-based antennas have also found wide applications (21,22,11). Descriptions of the analysis and design of passive and active microstrip circuits and their applications in MICs and monolithic MICs (MMICs) are contained in numerous publications (6–10). For further knowledge of the design and fabrication process of MICs and MMICs, see Refs. 23 and 24.

There have been a number of sophisticated microwave computer aided design (CAD) packages available on the market. They can be used for analysis purpose (the user describes a structure and determines its electrical response) and synthesis purpose (the software determines a physical structure meeting a special electrical behavior). A survey of available CAD softwares for microwave circuits is given in Refs. 7 and 9.

In addition to the microstrip lines, two other types of planar transmission lines are also widely used today in various microwave systems. They are the *slot lines* and *coplanar lines* (coplanar waveguides and coplanar strips). Hybrid combination of these planar lines with microstrip lines allows flexibility of circuit design and improves the performance of some circuit functions. Coplanar lines have received particular attention in recent years due to their many appealing properties. These include low dispersion, high flexibility in the design of characteristic impedance, and ease of connecting shunt and series lumped elements, or active devices without using via holes. Consequently, coplanar lines are used commonly in MMICs in conjunction with semiconductor devices which are also coplanar in nature. Descriptions of the analysis and design of slotline and coplanar line circuits can be found in Ref. 6 for further study.

It is worth mentioning finally that the major portion of this article is devoted to the analysis and design aspects of the microstrip line with applications to the microwave integrated circuits. However, high-speed digital circuits include, just like microwave integrated circuits, active and passive circuit elements interconnected by sections of strip and other microstrip transmission lines having a wide range of characteristic impedances. The speed of digital circuits has been steadily increased over the past decade; hence, the inductive effects on interconnect line-performance due to line-inductance that were previously assumed to be insignificant may be important. Moreover, the resulting junctions and discontinuities, and the electromagnetic coupling between the interconnects, contribute to reflections, signal delay and distortion, and crosstalk which can degrade the circuit and system performance. The same problems are encountered in printed circuit boards, single and multichip modules, and other packages. Therefore, a high-speed circuit design and simulation must incorporate these interconnections. The analysis of the microstrip line in this article applies to the vast area of electrical interconnections for printed circuit boards, single and multichip modules, and other packages. Examples of these additional applications can be found in Refs. 25–29, listed at the end of this article.

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MICRO-SYSTEMS TECHNOLOGY (MST). See MICROMACHINED DEVICES AND FABRICATION TECHNOLOGIES.

MICROTECHNOLOGY. See NEUROTECHNOLOGY.