Foundations for Microstrip Circuit Design

T. C. EDWARDS

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T. C. EDWARDS

Department of Electrical and Electronic Engineering
Royal Military College of Science
Shrivenham, Swindon, England

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Finally, I should like to draw the attention of the reader to those persons and organizations specifically acknowledged beneath appropriate figures used in the book.

T. C. EDWARDS
Shrivenham, England

Preface

For a variety of technological reasons microwave circuit integration has evolved mainly at the hybrid level, with microstrip transmission lines fulfilling a very wide range of planar circuit requirements. Although active monolithic microwave circuits have recently been reported in the advanced research phase, it seems inevitable that the demand for microstrip circuits will continue to expand.

This book provides foundations for the accurate design of microstrip components and circuits applicable to microwave or high-speed digital subsystems. The text is primarily intended for design engineers and research and development specialists who are active in these areas. It is also likely to prove interesting to lecturers and students on graduate electronics and communication engineering courses.

Whilst most of the techniques discussed lead to expressions which may be calculated using only a simple electronic calculator, the main emphasis throughout the book is on techniques suitable for fast computer-aided design (CAD). This is pursued in the firm belief that, in the near future, almost every individual worker or small group will possess a fairly powerful minicomputer system. A repeat coverage of well-documented theoretical analyses of the microstrip structure has been considered unnecessary and out of place here.

This work is based upon two main recent activities of the writer: teaching a new, graduate design-oriented microwave course and pursuing research involving microstrip lines. The microwave course was given at La Trobe University (Melbourne, Australia) as part of an advanced Communication Engineering Degree, whilst most of the research has been performed at the Royal Military College of Science (Shrivenham, England).

The book is organized into eight chapters, leading through fundamental aspects of microstrip design (principally Chapters 3 and 4) on to circuit applications in Chapter 8, and concluding with Appendix B in which existing CAD programs are summarized.

A basic review on TEM-mode transmission line theory is presented in Chapter 1. This is intended mainly for reference in connection with concepts and/or expressions required in many later chapters. It also serves as an appropriate introductory chapter.

Several structures which can conceivably be incorporated in microwave integrated circuits are described and critically compared in Chapter 2, which also deals with manufacturing processes. This chapter may be used as a source of initial information to assist in the choice of a particular technology other than microstrip.

A large number of significant items are discussed for what is believed to be the first time within a book. These include the following:

Proven-accurate expressions for microstrip analysis or synthesis, under static-TEM or dispersive conditions, and on an isotropic or sapphire substrate (Chapters 3 and 4)

A detailed appraisal of the frequency dependence of characteristic impedance and operating-frequency restrictions (Chapter 4)

Design information on many discontinuities, including short circuits and the transverse slit (Chapter 5)

Accurate design of parallel-coupled lines, special types of coupler, and crosstalk calculations (Chapter 6)

Details regarding several transitions and many measurement procedures (Chapter 7)

Almost all the design information in Chapter 8 and Appendix B.

High-speed digital applications are given in addition to microwave applications. Summaries of design recommendations are provided at the conclusion of appropriate chapters, viz. 3, 4, 5, and 6.

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1

A Basic Review of TEM-Mode Transmission Line Theory

1.1 THE CONCEPT OF RADIO-FREQUENCY TRANSMISSION LINES

A review of important TEM-mode radio-frequency transmission line theory is essential here because microstrip lines can be treated, at least to a first approximation, by assuming a static-TEM situation. We shall see that many significant microstrip design calculations require TEM-mode results from transmission line theory.

When transmission line structures have dimensions which are substantially smaller than the wavelengths of the signals being transmitted, then the structures may be satisfactorily analysed on the basis of line voltages and currents. As the frequency goes higher, and therefore the wavelength becomes smaller, it becomes necessary to set up a complete electromagnetic field solution in order to analyse the line structures. Here we can often assume that small line dimensions prevail and therefore a number of useful results are obtained on a voltage and current basis. The theory thus developed is called distributed circuit theory. Some common TEM-mode line cross-sections which have been fully analysed and are well understood are shown in Fig. 1.1.

The parallel line structure of Fig. 1.1(a) is rarely used for carrying radio-frequency signals because, being unshielded, it radiates energy and is easily disturbed by any nearby objects. Notice that both the parallel and the coaxial lines need only have dielectric supports at intervals along their lengths to maintain correct separation of the conductors. Furthermore, these supports may

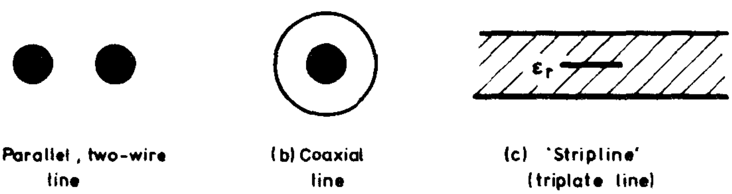

Fig. 1.1 Cross-sections of some common TEM-mode transmission lines

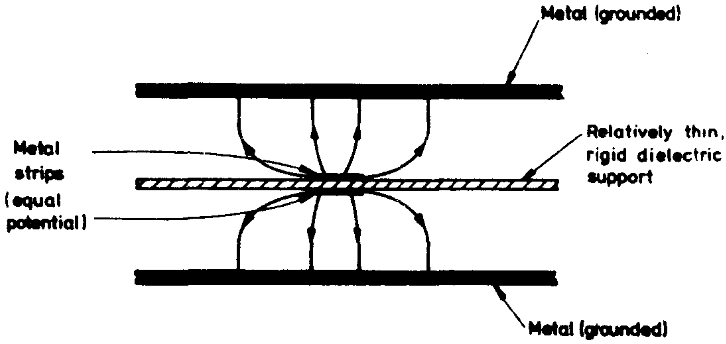

Fig. 1.2 High-Q triplate stripline, showing electric field

be filamentary, so that the 'dielectric filling' can consist almost entirely of air and the dielectric losses are low. On the other hand, the stripline inherently requires a solid, low-loss, dielectric filling.

Electric and magnetic field patterns associated with the TEM mode of propagation are well known, and the electric field is shown here in Fig. 1.2 for a high-Q triplate stripline.

The remainder of this chapter is concerned with definitions and results from transmission line theory.

1.2 PRIMARY TRANSMISSION LINE CONSTANTS

In this chapter we are concerned only with uniform transmission lines. Regardless of the actual structure, such lines are shown schematically as in Fig 1.3(a). *Primary* constants are indicated in Fig. 1.3(b) and can be defined as follows:

Resistance along the line $= R$	all specified 'per loop metre'
Inductance along the line $= L$	
Conductance shunting the line $= G$	
Capacitance shunting the line $= C$	

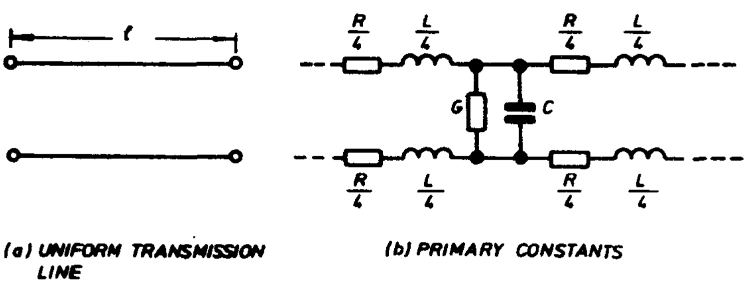

Fig. 1.3 The uniform transmission line: primary constants

The 'per loop metre' specification means that each quantity is determined on a 'go and return' basis. For example, to find the total value of R per loop metre for a coaxial line we must add the resistance of one metre of the inner conductor to that of the outer conductor. In most radio-frequency transmission lines the effects due to L and C tend to dominate, because of the relatively high inductive reactance and capacitive susceptance respectively. In such cases we refer to 'loss-free' lines, although in practice some information about R or G may be necessary to determine actual power losses. The 'loss-free' concept is just a useful and good approximation.

The propagation of a wave along the line is characterized by the (complex) *propagation coefficient* γ:

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)} \tag{1.1}$$

or

$$\gamma = \alpha + j\beta \tag{1.2}$$

where α = *attenuation coefficient*, Nepers per metre
β = *phase-change coefficient*, degrees, or radians, per metre

At sufficiently high radio frequencies, eqns (1.1) and (1.2) yield the important result:

$$\beta = \omega\sqrt{LC} \tag{1.3}$$

1.3 SECONDARY CONSTANTS FOR TRANSMISSION LINES

In one complete wavelength λ_g along the line the travelling wave must experience 2π radians phase shift. Thus

$$\beta\lambda_g = 2\pi$$

or

$$\beta = \frac{2\pi}{\lambda_g} \tag{1.4}$$

(Hence β is also called the 'wave number'.)

From the relationship for the velocity $v_p = f\lambda_g$ we can also write

$$v_p = \frac{\omega}{\beta} \qquad \text{m/s} \tag{1.5}$$

By using eqn (1.3) we also obtain

$$v_p = \frac{1}{\sqrt{LC}} \tag{1.6}$$

The velocity of propagation v_p is also given in terms of the absolute permeability μ and permittivity ε of the medium through which the wave passes:

$$v_p = \frac{1}{\sqrt{\mu\varepsilon}} = \frac{c}{\sqrt{\mu_r\varepsilon_r}} \tag{1.7}$$

where $c = 2.99793 \times 10^8$ m/s, the velocity of light in free space—often approximated to 3×10^8 m/s. In coaxial lines designed for microwave applications the velocity is very close to c because there is so little dielectric filling, only well-separated spacers. (This is *not* true for coaxial cables, whose foam filling reduces the velocity.)

Most lines do not possess any ferromagnetic materials and thus have $\mu_r = 1$. For stripline (Fig. 1.1c) the filling consists of a uniform dielectric and eqn (1.7) then gives the velocity as

$$v_p = \frac{c}{\sqrt{\varepsilon_r}} \tag{1.8}$$

We note that the wave is slowed by the dielectric medium.

By noting that $c = f\lambda_0$ and $v = f\lambda_g$ we can use eqn (1.8) to show that

$$\lambda_g = \frac{\lambda_0}{\sqrt{\varepsilon_r}} \tag{1.9}$$

(The subscript 'g' means 'guide'; we can look upon the transmission line as a 'guiding' structure.) This expression, eqn (1.9), is particularly significant. It explicitly states that the wavelength is reduced according to the square root of the relative permittivity of the material in which the line is embedded. For example, a coaxial transmission line is filled with alumina ($\varepsilon_r \simeq 9$) and excited at 10 GHz. The *free-space* wavelength would be 3 cm, but the wavelength in the coaxial line is only $3/\sqrt{9} = 1$ cm. Thus, distributed components become shorter and occupy less space, if high-permittivity materials can be used. This has important implications for microstrip.

The final secondary constant to be discussed is the characteristic impedance Z_0. This is given generally by

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + J\omega C}}$$

At high radio frequencies this simplifies to

$$Z_0 = \sqrt{\frac{L}{C}} \tag{1.10}$$

which is a *purely resistive* result.

For coaxial lines two values are commonly encountered:

At v.h.f.: $Z_0 = 75\,\Omega$ usually

At u.h.f. and s.h.f. (microwave frequencies): $Z_0 = 50\,\Omega$ usually

Changing Z_0 requires a change in the dimensions of the line. Specifically, in the case of coaxial lines, it requires changing the ratio of the radii applicable to the inner and outer conductors. Dimensional alterations are much more easily carried out with stripline or its derivatives and therefore such structures can be fabricated for a wide range of characteristic impedances (typically 20 to 150 Ω for stripline).

It is not often appreciated that eqn (1.10) can be reexpressed, with the aid of eqn (1.6), in the following alternative forms:

$$Z_0 = v_p L \tag{1.11a}$$

or

$$Z_0 = \frac{1}{v_p C} \tag{1.11b}$$

Equation (1.11b) is particularly useful in establishing some fundamental microstrip parameters.

1.4 TRANSMISSION LINE IMPEDANCES

The input impedance to a transmission line varies with the distance progressed along the line. Therefore a suitable distance notation is required, which is shown in Fig. 1.4. Many standard textbooks derive the expressions given as eqns (1.12) to (1.17) here.

By setting down the distance limits into expressions for the voltage and current at *any* point along the line we obtain, for the load impedance Z_L, the following expression for the input impedance at $z = 0$:

$$Z_{in} = Z_0\left(\frac{Z_L \cosh \gamma l + Z_0 \sinh \gamma l}{Z_0 \cosh \gamma l + Z_L \sinh \gamma l}\right) \tag{1.12}$$

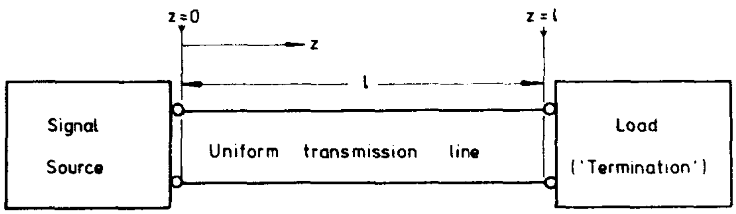

Fig. 1.4 Notation applicable to a general, uniform line

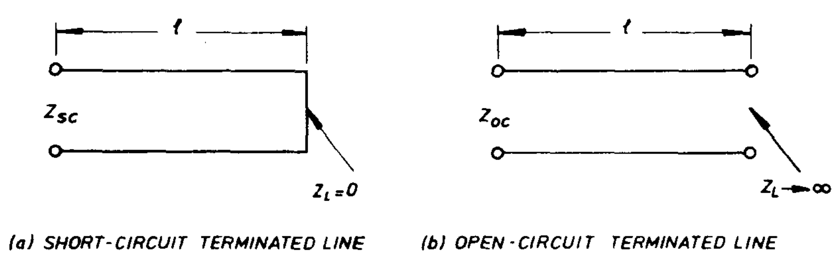

Fig. 1.5 Short-circuit and open-circuit terminated lines

This expression is completely general. For loss-free lines the result simplifies to

$$Z_{in} = Z_0\left(\frac{Z_L + jZ_0 \tan \beta l}{Z_0 + jZ_L \tan \beta l}\right) \tag{1.13}$$

The input impedances for lines having short-circuit or open-circuit terminations (see Fig. 1.5) follow from eqn (1.13), and these impedances can now be easily written down.

For the short-circuit case $Z_L = 0$ and eqn (1.13) gives

$$Z_{sc} = jZ_0 \tan \beta l \tag{1.14}$$

For the open-circuit case $Z_L \rightarrow \infty$ and eqn (1.13) gives

$$Z_{oc} = -jZ_0 \cot \beta l \tag{1.15}$$

Short-circuit or open-circuit terminated lines are very useful structures in a wide variety of circuits.

By choosing the correct lengths (l) inductive or capacitive circuit elements are automatically realized in transmission line form, as can be seen from eqns (1.14) and (1.15). There are many applications in resonators, filters, and matching and coupling networks.

When the line is terminated in a load impedance exactly equal to the characteristic impedance (Z_0) of the line itself it is perfectly broad-band matched. The denominator is equal to the numerator on the right-hand parentheses of eqn (1.12) and hence

$$Z_{in} = Z_0$$

This completely matched condition is often used in test and measurement procedures.

1.5 REFLECTION AND VOLTAGE STANDING-WAVE RATIO

In all cases, except the completely matched condition, the load termination reflects some of the energy originally sent down the line. Interference between the incident and reflected waves, travelling at the same velocity but in opposite directions, causes a 'standing-wave' field pattern to be set up.

The voltage reflection coefficient Γ (the ratio of reflected to incident voltage at the load) is given by

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0} \tag{1.16}$$

The ratio of maximum to minimum amplitude of the standing wave is called the voltage standing-wave ratio r (or 'vswr'), given by

$$r = \frac{1 + |\Gamma|}{1 - |\Gamma|} \tag{1.17}$$