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Study of Static Analysis and Losses of Microwave Isolated Single Stripline Structure Using Conformal Mapping Method

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ABSTRACT

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In this paper, we studied about static analysis and losses of microwave isolated single stripline structure using conformal mapping method. Millimeter waves are extensively being used for Radar and Wireless communication. Wireless communication includes wireless computer networks, voice and data networks etc., Transmission lines forms an integral part of these Microwave Integrated Circuits (MICs) and Monolithic Microwave Integrated Circuits (MMICs). One major constraint on the transmission lines for their use in MICs is that they have to be planar. Microstrip lines, slot lines and coplanar structures are used as the fundamental blocks in building these circuits. All these structures are planar, their characteristics being controlled by their dimensions in that plane.

Keywords : MICs, Dispersion, Microwave, Conformal mapping method.
protocol, Reliable communication, Topology. Spur gear assembly, Electromagnetic Induction

I. INTRODUCTION

Over the years various static and dispersion models have been developed for the analysis of striplines. Various models for the analysis of these structures have been consolidated here. We shall begin with a brief introduction on the classic methods employed for the static and dispersion analysis of the various strip line configurations. We shall then proceed with the static and dispersion analysis of the individual striplines mentioned above. We shall conclude with a section on the various design equations that can be used during the stripline design process.

II. Static analysis

Static analysis of striplines involves the analysis of the transmission structure at frequency, f = 0. The analysis is carried out to find the vital parameters of the transmission lines viz., Characterisitc impedance(Z_0), Effective dielectric permittivity (ϵ_{eff}), and the phase velocity (v_p). These three parameters are related to the Capacitance (C) of the structure [1-2]. We know that the wave propagates through a medium with velocity $v = \frac{1}{\sqrt{\mu\epsilon}}$. If the medium is not free space but a uniform dielectric, then the velocity of propagation is given by (1).

$$v = v_0/\sqrt{\epsilon_r} \quad (1)$$

where $v_0 = 1\sqrt{\mu_0\epsilon_0}$ is the free-space velocity and ϵ_r is the relative permit.

$$Z = \frac{V_0}{I_0} = \frac{V_0}{Q_v} = \frac{1}{C_v} \quad (2)$$

where C is the capacitance between the conductors per unit length. The electrostatic capacitance used in independent of the operating frequency and depends only on the static field configuration in the transmission line [3]. If the substrate is thin in terms of wavelength and the strip width is also narrow compared to the wavelength, and high dielectric substrates are used, then static analysis its lef can be enough [4], [5].

Substituting the dielectric by air ($\epsilon_r = 1$), we have

$$Z_a = \frac{1}{C_a v_a} \quad (3)$$

where C_a is the capacitance per unit length of the transmission line with dielectric replaced by air. Dividing (3) and putting $\epsilon_r = C/C_a$ we get

$$Z = \frac{Z_a}{\sqrt{\epsilon_r}} \quad (4)$$

[2] Open transmission lines like Microstrip lines are examples of mixed dielectric problem and as such can't support TEM waves. To aid in the analysis of these structures, a new quantity called *'effective dielectric constant'* is defined under *quasi-static* approximation. This quantity is defined as

$$\epsilon_{eff} = \frac{C}{C_a} \quad (5)$$

The expressions for the phase velocity and characteristic impedance Z then follows

$$v = v_a/\sqrt{\epsilon_{eff}} \quad (6)$$

$$Z = Z_a/\sqrt{\epsilon_{eff}} \quad (7)$$

Propagation constants calculated using (5) - (7) give results accurate enough for most of the practical cases [2].

The electrostatic capacitance is found by the solution of a two-dimensional Laplace Equation

$$\frac{\partial^2\phi}{\partial x^2} + \frac{\partial^2\phi}{\partial y^2} = 0 \quad (8)$$

The solution of (8) would then give us the field within the structure. The total charge can be found out by using Gauss law

$$Q = \epsilon_0\epsilon_r \iint E.dS \quad (9)$$

The integration in (9) being carried out over the entire surface of the transmission structure. The determination of the characteristic impedance Z_0 proceeds with first finding the capacitance C of the transmission line and then using that in (5) to find the effective dielectric constant and then putting it into (6) and (7) to find out phase velocity and characteristic impedance respectively.

We will now discuss the theoretical aspects of the static analysis techniques namely, The Conformal map- ping method, The Variational method, and The Finite Difference method.

3. The Conformal Mapping method

A mapping $w = f(z)$ defined on a domain D is called conformal at $z = z_0$ if the angle between any two curves in D intersecting at z_0 is preserved by f. Such a mapping is known as *Conformal mapping (a.k.a Angle-preserving mapping).* If $f(z)$ is analytic in domain D and $f'(z_0) \neq 0$, then f is conformal at $z = z_0$. The criterion for analyticity is that: if $u(x, y)$ and $v(x, y)$ satisfy the Cauchy-Reimann equations $\frac{\partial\mu}{\partial x} = \frac{\partial v}{\partial y}$ and $\frac{\partial\mu}{\partial y} = \frac{\partial v}{\partial x}$ at all points in domain D, then the function $f(z) = u(x,y) + jv(x,y)$ is analytic everywhere in D provided *u(x,y)* and *v(x,y)* are continious and has first-order partial derivatives.

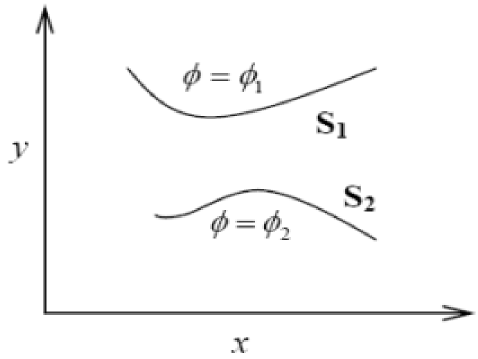

Fig. 1. A system of curves

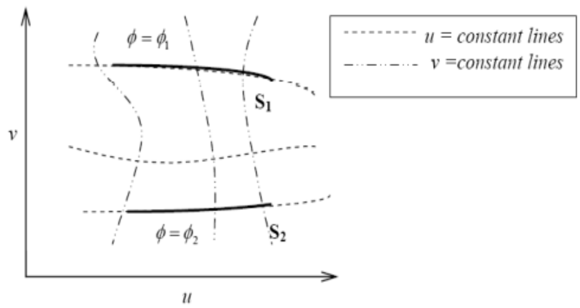

Fig. 2. The system of curves after conformal mapping

The reason we can apply conformal mapping to solve Laplace equations follow directly from a result in Complex analysis, which states that: If f is an analytic function that maps from a domain D to D' and if W is harmonic in D' , then the real valued function $w(x, y) = W(f(z))$ is harmonic in in D [6].

[3], [2] Consider the solution of a two-dimensional Laplace equation $\Delta^2\phi = 0$ for the system of conductors shown in Fig 1 with the boundary conditions as $\phi = \phi_1$ on S_1 and $\phi = \phi_2$ on S_2. The principle behind conformal mapping approach is to transform the given system of conductors to a different complex plane where it may be easier to solve the given laplace equation. This technique gives an upper bound on Z_0. Let us consider a conformal transformation W given by (10)

$$W = F(z) = F(x + jy) = u + jv \qquad (10)$$

where

$$u = u(x, y)$$

and

$$v = v(x, y)$$

Assuming a Transverse Electromagnetic Mode (TEM), we shall define the gradient operator as ∇_t

$$\nabla_t u = \frac{\partial u}{\partial x}\frac{a_x}{h_1} + \frac{\partial u}{\partial y}\frac{a_y}{h_2} \qquad (11)$$

and

$$\nabla_t v = \frac{\partial v}{\partial x}\frac{a_x}{h_1} + \frac{\partial v}{\partial y}\frac{a_y}{h_2} \qquad (12)$$

The scale factors h_1 and h_2 are given by

$$\frac{1}{h_1^2} = \left(\frac{\partial u}{\partial x}\right)^2 + \left(\frac{\partial u}{\partial y}\right)^2$$

$$\frac{1}{h_2^2} = \left(\frac{\partial u}{\partial x}\right)^2 + \left(\frac{\partial v}{\partial y}\right)^2$$

$$\frac{1}{h_1^2} = \frac{1}{h_2^2} = \frac{1}{h_2} = \left|\frac{dw}{dz}\right|^2$$

where the last step directly follows from the Cauchy-Reimann equations. Laplace equation in uv plane is given by

$$\frac{\partial}{\partial u}\frac{h_2}{h_1}\frac{\partial \phi}{\partial u} + \frac{\partial}{\partial v}\frac{h_1}{h_2}\frac{\partial \phi}{\partial v} = \frac{\partial^2\phi}{\partial u^2} + \frac{\partial}{\partial v^2} = 0 \qquad (13)$$

The above result shows that the potential function ϕ satisfies the same Laplace equations in uv co-ordinate systems. The same result has been shown in a slightly different way in [7].

The energy stored in the electrostatic field is given by

$$W_e = \frac{1}{2_\epsilon}\iint\left[\left(\frac{\partial \phi}{\partial x}\right)^2 + \left(\frac{\partial \phi}{\partial y}\right)^2\right]dxdy$$

$$= \frac{1}{2_\epsilon}\iint|\partial t\phi|^2 dxdy$$

$$= \frac{1}{2_\epsilon}\iint\left[\frac{1}{h_1^2}\left(\frac{\partial \phi}{\partial u}\right)^2 + \frac{1}{h_2^2}\left(\frac{\partial \phi}{\partial v}\right)^2\right]h_1\,\mathrm{dx}\,h_2 dy$$

$$= \frac{1}{2_\epsilon}\iint\left[\left(\frac{\partial \phi}{\partial u}\right)^2 + \left(\frac{\partial \phi}{\partial v}\right)^2\right]dudv$$

$$= 1/2C(\phi_2 - \phi_1)^2 \qquad (14)$$

The last equation shows that capacitance C is same in the conformal mapped domain.

Situations may sometimes arise where a particular type of a stripline can be considered as a ploygon. This polygon can then the conformal mapped to the upper half of a plane by the use of Schwarz-Christoffel Transformation. The transformation is formulated as [6]: *Let f(z) be a function that is analytic in the plane y>0 having the derivative*

$$f'(z) = A(z-x_1)^{a_1/\pi-1}(z-x_2)^{a_2/\pi-1}\dots(z-x_{n2})^{a_n/\pi-1} \quad (15)$$

where $x_1 < x_2 < \dots < x_n$ and each α_i satisfied $0 < \alpha_i < 2\pi$. Then $f(z)$ maps the upper half of the plane $y > 0$ to a polygon with its interior angles $a_{1,}a_2 \dots\dots a_n$ Successive application of this transformations would sometimes be needed to end up with the desired configuration. Detailed explanation on Schwarz-Christoffel Transformation and the method to obtain the functions for such a mapping is outlined in [6]. The steps followed while using the conformal mapping approach shall be outlined during the analysis of Micro stripline.

4. Losses

Two kinds of losses are associated with striplines. Conductor and dielectric losses. The total loss of the stripline in questions is then the combination of these two i.e., the total loss α is the sum of Dielectric loss (α_d) and the conductor loss (α_c)

$$a = \alpha_c + \alpha_d$$

The loss is attributed to the finite conductivity of the strip and the ground planes and lossy dielectric as substrates.

The techniques used to simplify the expressions for the propagation constants and Zₒ. The expression for the propagation constant γ is given by (12)

$$\gamma = \sqrt{(R+j\omega L)(G+j\omega C)} \quad (16)$$

where R,G, L and C are the values of the distributed elements of the transmission line

$$\gamma \sqrt{(j\omega L)(j\omega C)\left(1+\frac{R}{j\omega L}\right)\left(1+\frac{R}{j\omega C}\right)}$$

$$= j\omega\sqrt{LC}\sqrt{1-j\left(\frac{R}{\omega L}+\frac{G}{\omega C}\right)-\frac{RG}{\omega^2 LC}} \quad (17)$$

using Taylor's expansion for $\sqrt{1+x}$ we can write (17) as

$$\gamma \approx j\omega\sqrt{LC}\left[1-\frac{j}{2}\left(\frac{R}{\omega L}+\frac{G}{\omega C}\right)\right] \quad (18)$$

separating the real and imaginary parts we have,

$$\alpha \approx \frac{1}{2}\left(R\sqrt{\frac{C}{L}}+G\sqrt{\frac{L}{C}}\right) = \frac{1}{2}\left(\frac{R}{Z_o}+GZ_o\right) \quad (19)$$

and

$$\beta = \omega\sqrt{LC} \quad (20)$$

Characteristic impedance, Z_o is approximated as

$$Z_o = \sqrt{\frac{R+j\omega L}{G+j\omega C}} \approx \sqrt{\frac{L}{C}} \quad (21)$$

There are two main techniques used to analyze the two types of losses mentioned above

Perturbation method [12]: The technique avoids the determination of the transmission line parameters L, C, R, and G. It uses the field equations to determine the losses assuming that there will be little changes between the field of a lossy line and that of a loss-less line, thus justifying its name. The method considers the power flow along a transmission line with attenuation α (in the absence of reflections), which is given by

$$P(z) = P_o e^{-2\alpha z} \quad (22)$$

where P_o si the power at z=0 plane. The power loss per unit length can be represented as a derivative

$$P_l = \frac{-\partial P}{\partial z} = 2\alpha P_o e^{-2\alpha z} = 2\alpha P(z) \quad (23)$$

Attenuation constant α is now defined using (22) and (23) as

$$\alpha = \frac{P_l(z)}{2P(z)} = \frac{P_l(z=0)}{2P_o} \quad (24)$$

The crux of the problem, as we can see, is the computation of the power (P_o) and the power loss (P_l). Due to the generality of this technique, this method is applicable for both conductor and dielectric losses. P_o is in general given by the Poynting theorem.

$$P_o = \frac{1}{2}Re\int_s E \times H^* ds \quad (25)$$

The Ohmic conductor loss can be written as

$$P_t = \frac{R_s}{2}\int_s |J_s|^2 \int_s |H_t|^2 \mathrm{ds} \quad (26)$$

The Dielectric power loss is

$$P_t = \frac{\omega\epsilon''}{2}\int_V |E|^2 dv \quad (27)$$

where ϵ'' is imaginary part of the dielectric permittivity

$$\epsilon = \epsilon' + j\epsilon'' = \epsilon(1 + j\tan\ \delta)$$

where tan δ is the loss tangent of the dielectric. Depending on the expression used for P_t in (27), the attenuation constant will refer to either the conductor loss or the dielectric loss.

Wheeler's incremental inductance rule [12], [13]:

The rule gives the effective resistance due to the skin effect and begins with inductance calcuations. The method holds good for all types of metallic surfaces for which skin of metallic surfaces for which skin effect plays a significant role. The only constraint being that the thickness of the conductor should be great compared to the skin depth (at least twice). The incremental rule can be stated as: the effective resistance in any circuit is equal to the change of reactance due to the penetration of magnetic field into the metal surfaces as would be caused by the surface receeding to a depth of $\delta/2$. As we can see that the field cannot penetrate a perfect conductor and hence the losses in the conductor is due to their *non-infinite* conductivity. The power loss into a cross section S of a perfect conductor is

$$P_t = \frac{R_S}{2}\int_S |J_S|^2 ds = \frac{R_S}{2}\int_C |H_t|^2 \mathrm{dl W/m^2} \qquad (28)$$

The above itegral refers to the power loss per unit length, the contour integral being carried out across the two conductors. The inductance per unit length is given by

$$L = \frac{\mu}{|I|^2}\int_S |H|^2 ds \qquad (29)$$

The expression for L assumes a lossless conductor. However, there will be penetration of the field and thus H will be no longer 0 inside the surface. This will add an incremental inductance ΔL to L. Knowing that the mean depth of current inside the conductor is $\delta/2$we have

$$\Delta L = \frac{\mu_0 \delta}{2|I|^2}\int_C |H_t|^2 ds \qquad (30)$$

Putting (29) in (30), we have

$$P_t = \frac{|I|^2 \omega \Delta L}{2} W/m \qquad (31)$$

Using (31) and proceeding further, we have the final expression for α as (12)

$$\alpha_c = \frac{R_s}{2Z_0\eta}\frac{dZ_0}{dl} \qquad (32)$$

where $\eta = \sqrt{\mu_0/\epsilon}$ is the intrinsic impedance of the dielectric. As can be observed from our argument, that this rule is applicable only for the evaluation of conductor losses.

5. Conclusions

A brief survey of the various techniques used in Static and Dispersion analysis of various Microstrip line configurations has been done. Wherever possible closed-form expressions have been provided to aid in CAD of these microstrip lines. Dispersion analysis for several microstrip lines are absent as closed-form expressions were not available for these. Dispersive effects on Dielectric and Conductor losses has also been omitted in the discussion. All the static equations given here were tested against various other transmission line calculators available. The equations were implemented as a part of the Microwave transmission line tool written by me.

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