"ATTENUATION IN MICROSTRIP TRANSMISSION LINES WITH VERY LOSSY SUBSTRATES"

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Summary

The design and development of microstrip-based attenuators suitable for broad-band applications is described. Expressions are derived for the propagation constant and characteristic impedance of attenuating sections of microstrip. Some practical results are also given.

Introduction

An investigation of microstrip transmission line structures with very lossy substrates is described. The work was in response to an industrial requirement for a cheap and novel attenuator suitable for mass production and capable of providing 30 dB of attenuation from d.c. to 12 GHz. The possibility of replacing a section of the microstrip substrate beneath the top conductor with very lossy conductive material was investigated. There appears to be no previously published theory of this structure in the literature.

Attenuation Theory

First, expressions for the attenuation constant α, phase constant β, and characteristic impedance, Z_0, appropriate to a length of microstrip constructed on a very lossy substrate were obtained in order to evolve a suitable design in which these parameters would be frequency-independent. These quantities are given for a TEM mode by the equations

$$\alpha = R_e(\gamma), \tag{1}$$

$$\beta = I_m(\gamma), \tag{2}$$

$$\gamma = \sqrt{(R + j\omega L)(G + j\omega C)}, \tag{3}$$

and
$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}} \tag{4}$$

where R, L, G, and C are the usual transmission line parameters and ω is the angular frequency. Above about 2 GHz hybrid modes were known to become important [1] but as their effect amounts at the most to only a 16% change over the bandwidth from d.c. to 12 GHz [2] - [4] they were regarded here as of secondary importance to be included as a correction, if necessary, at a later stage.

For low-loss transmission lines the conditions $G \ll \omega C$, $R \ll \omega L$ apply leading to the familiar expressions

$$\alpha = 0,\ \beta = \omega\sqrt{LC} \text{ and } Z_0 = \sqrt{\frac{L}{C}} \tag{5}$$

These conditions are inappropriate for a very lossy line with $G \gg \omega C$. Thus new expressions for α, β, and Z_0 were required. It was shown, for example, that when $G \gg \omega C$, $R \ll \omega L$ that

$$\alpha = \beta = \sqrt{\frac{G\omega L}{2}} \tag{6}$$

while
$$Z_0 = \sqrt{\frac{j\omega L}{G}} \tag{7}$$

The frequency-dependance of these parameters depends upon that of G and L. If the devices are thin compared to a skin depth they will be frequency independent. However, for the lossy substrate we have [5, p.81]

$$G = \frac{\omega\epsilon'' + \sigma}{\epsilon'} C \tag{8}$$

where $\omega\epsilon''$ is the effective conductivity due to the polarisation losses of the conducting substrate of complex dielectric permittivity $\epsilon' - j\epsilon''$, and σ is the real electrical conductivity of the substrate.

Thus, if $\sigma \gg \omega\epsilon''$ we have

$$\alpha = \beta = \sqrt{\frac{\omega LC\sigma}{2\epsilon'}} \tag{9}$$

and
$$Z_0 = \sqrt{\frac{j\omega\epsilon' L}{\sigma C}} \tag{10}$$

so that α, β and Z_0 are each proportional to $\omega^{\frac{1}{2}}$.

Similarly if $\omega\epsilon'' \gg \sigma$, it is easily shown that $\alpha \propto \omega$ while Z_0 will be frequency-independent.

Thus a broadband attenuator cannot be constructed in lossy microstrip.

Equation (9) is of some importance. Previous authors [6] - [8] have stated that when ohmic-losses dominate α is independent of frequency. That conclusion is not true in the case of lossy substrates such as semiconductor materials with conductivities greater than about 0.008 Sm (at 12GHz) For these substrates (9) predicts that α and $\beta \propto \omega^{\frac{1}{2}}$ and this result is expected to be of particular significance for monolithic integrated circuits operating at Gbit rates.

To produce a broadband attenuator the conditions $G \gg \omega C$, $R \gg \omega L$ must be satisfied. Then

$$\gamma = \sqrt{RG} \tag{11}$$

so that
$$\alpha = \sqrt{RG}, \tag{12}$$

$$\beta = 0, \tag{13}$$

and
$$Z_0 = \sqrt{\frac{R}{G}} \tag{14}$$

If the device is thinner than a skin depth R and G will be frequency independent. We then have for $\sigma \gg \omega\epsilon''$

$$\alpha = \sqrt{\frac{RC\sigma}{\epsilon'}} \tag{15}$$

and
$$Z_0 = \sqrt{\frac{R\epsilon'}{C\sigma}} \tag{16}$$

both expressions being frequency-independent as required.

However, if $\omega\epsilon'' \gg \sigma$ both α and Z_0 are frequency-dependent. Thus for a broadband attenuator it is required that $G \gg \omega C$, $R \gg \omega L$ and $\sigma \gg \omega\epsilon''$. A structure which may satisfy these conditions is shown in Fig. 1. Note that the top conductor is replaced at the lossy section by lossy material to ensure $R \gg \omega L$. In the presence of the top conductor $R \ll \omega L$. The line parameters R, G may be obtained from a consideration of Fig. 1 as

$$R = \frac{1}{\sigma W(h + t)}\ \Omega\, m^{-1} \tag{17}$$

$$G = \frac{\sigma W}{(h + t)} \text{ Mho m}^{-1} \qquad (18)$$

where σ is the conductivity of the lossy material. Thus

$$\alpha = \sqrt{RG} = \frac{1}{(h + t)} \text{ nepers m}^{-1} \qquad (19)$$

and

$$Z_0 = \sqrt{\frac{R}{G}} = \frac{1}{\sigma W}\ \Omega \qquad (20)$$

Both expressions are independent of frequency provided the height of the lossy material is less than a skin depth and $\sigma \gg \omega\epsilon''$.

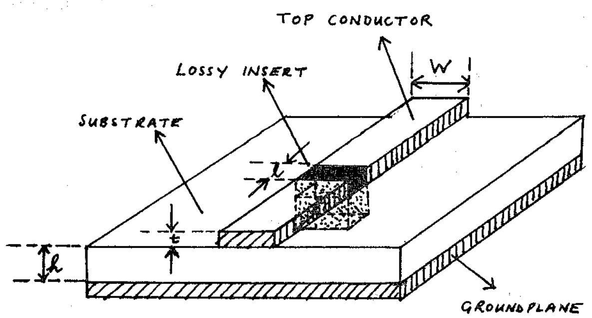

Fig. 1 Broadband Attenuator Structure

Fabrication and Measurement Procedure

The attenuation of the lossy microstrip and of the attenuators was determined from the measured scattering parameters s_{11}, s_{22} and s_{21}. Double-sided copper-clad printed circuit board was used in this initial study. The top conductor was defined by etching and then a short section removed at the intended location of the lossy insert. A hole of the correct dimensions was cut into the substrate material and was then filled with a suitable conductive material. For the very-lossy microstrip experiments a top conductor was then soldered back into place. The lines designated as attenuators were left "topless." The attenuation, A(dB), was obtained from the equation

$$A(\text{dB}) = 10 \log_{10} \frac{|s_{21}|^2}{(1 - |s_{11}|^2)(1 - |s_{22}|^2)} \qquad (21)$$

Results and Discussion

Fig. 2 shows plots of attenuation versus frequency, f, for microstrip with a section of very lossy substrate. Interpretation of these results was facilitated by plotting $\log_{10}A(\text{dB})$ versus $\log_{10} f$. These curves indicated that at low frequencies $\alpha \propto f^{0.1}$, while at higher frequencies the variation was between $\alpha \propto f^{0.8}$ and $\alpha \propto f^{1.7}$. According to (9) $\alpha \propto f^{\frac{1}{2}}$ was expected, assuming that $G \gg \omega C$ and $R \ll \omega L$. The results were explained on the assumption that σ was too small so that $\omega\epsilon'' + \sigma < \omega\epsilon'$. This would mean $G \lesssim \omega C$. A detailed analysis showed that α could then be frequency-independent while at higher frequencies $\alpha \propto f$, thus explaining why $\alpha \propto f^{0.1}$ and $\alpha \propto f^{0.8}$ was observed. The case $\alpha \propto f^{1.7}$ represents the well-known phenomenon of dielectric heating [9] and Pucel et al [1] confirmed that for a non-conducting substrate $\alpha \propto f$. The result $\alpha \propto f^{1.7}$ was attributed to the existence of eddy currents in the small conducting particles of the lossy substrate since eddy current losses are $\propto f^2$ at low frequencies and $f^{1.5}$ at high frequencies [9]. It was concluded that the proposed theory of very lossy microstrip was correct.

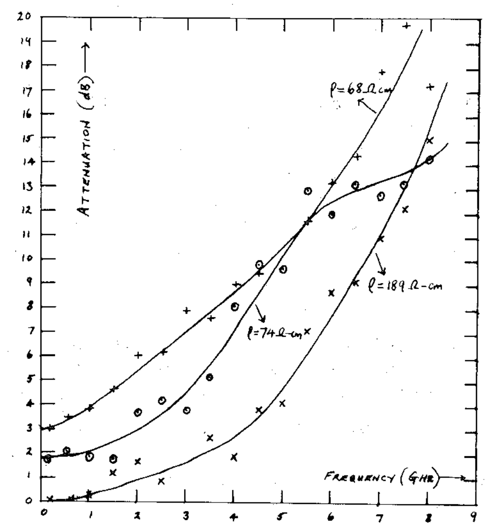

Fig. 2 Attenuation versus Frequency for Microstrip with Lossy Insert

The plots of α and Z_0 versus frequency for the prototype attenuators were analysed in great detail for frequencies up to 3 GHz only. It was possible to show that the increase in α with frequency, Fig. 3, occurred because the condition $R \gg \omega L$ was not satisfied. One reason for this was because it had been assumed, after Luskow [10], that the surface layer of the lossy insert of thickness t would serve as the top conductor. Actually R was found to be associated with the full thickness of the lossy insert (t+h), and this is reflected in the correct design equations (17)-(20). A second reason was that L was enhanced due to the presence of eddy currents in the lossy insert. Thus the value of L was deduced as 17.0 μHm^{-1} compared with about 314 nHm^{-1} calculated from Wheeler's equation (8), [11].

An attempt was made to increase R by reducing σ. To keep Z_0 constant G was increased by using a reduced height structure, Fig.4. An improved attenuator characteristic resulted (Fig.3) but the mismatch increased with frequency, possibly because of the reduced height section.

A third attenuator was designed with $\sigma \gg \omega\epsilon'$ to ensure $G \gg \omega C$. Also R and G were calculated from (17) and (18). This attenuator exhibited

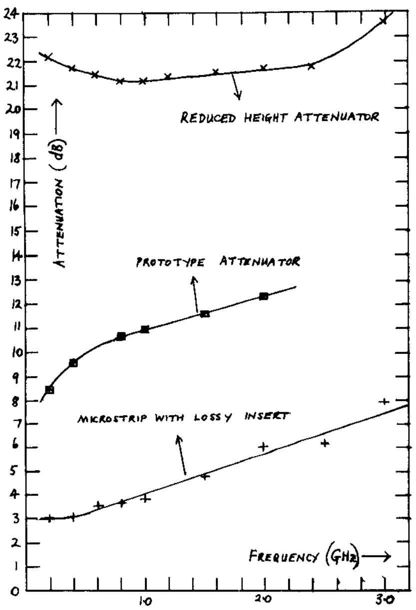

Fig.3 Attenuation versus Frequency for Different Structures Investigated

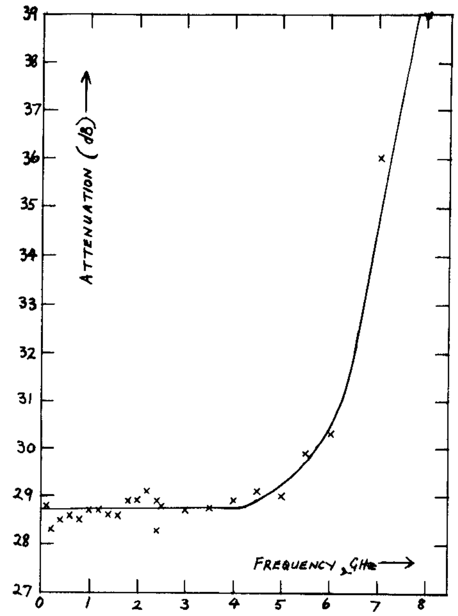

Fig.5 Attenuation versus Frequency for Third Attenuator

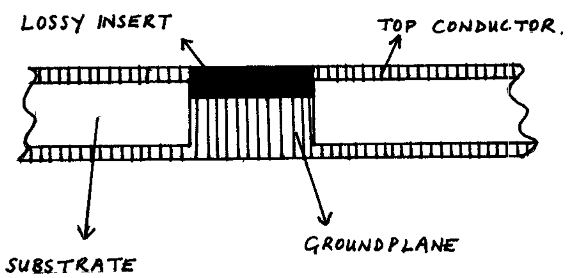

Fig.4 Reduced Height Attenuator

broadband attenuation behaviour, Fig.5. The increase in attenuation above 5 GHz was possibly due to the skin effect since σ was inadvertently made too high thus reducing the skin depth δ . Since $\alpha \propto (h+t)^{-1}$ and $(h+t)^{+1}$ should then become δ, α then increases as $f^{\frac{1}{2}}$. Efforts are being made to improve upon this result.

Conclusion

This approach to broadband attenuator design has been successful and further development is being undertaken. The important result that the attenuation and phase constants of microstrip interconnections in monolithic integrated circuits will be strongly frequency-dependent was discovered. This is additional to the normal dispersive behaviour of microstrip.

Acknowledgements

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References

1. R.A.Pucel, D.J.Massé, and C.P. Hartwig "Losses in Microstrip" IEEE Trans. Microwave Theory and Tech., vol.MTT-16, No 6,pp 342-350, June 1968
2. W.J.Getsinger, "Microwave Dispersion Model" IEEE Trans. Microwave Theory Tech., vol.MTT-21, No 1,pp 34-39, January 1973
3. R.D.Owens, "Predicted Frequency Dependence of Microstrip Characteristic Impedance Using the Planar Waveguide Model" Electronics Letters, vol,12, No 11, pp 269-270, May 1976
4. H.J.Carlin "A Simplified Circuit Model for Micro strip" IEEE Trans. Microwave Theory Tech.,vol.MTT-21, pp 589-591, September 1973
5. R.E.Collin "Foundation for Microwave Engineering" McGraw-Hill, 1966
6. T.M.Hyltin,"Microstrip Transmission on Semiconductor Dielectrics", IEEE Trans. Microwave Theory Tech.,vol. MTT-13, pp 777-781, November 1965
7. J.D. Welch and H.J. Pratt "Losses in Microstrip Transmission Systems for Integrated Microwave Circuits" NEREM RECORD, vol.8,pp 100-1, 1966
8. I.J.Bahl and D.K.Trivedi, "A Designers Guide to Microstrip Line" Microwaves, pp 174-182, 1977
9. D.J.Fink "Standard Handbook for Electrical Engineers" 10th Ed., McGraw-Hill, p2-21
10. A.A.Luskow, "Precision Attenuators for Microwave Frequencies" Marconi Instrumentation, vol.14, No 4,pp 81-86, 1974
11. H.A.Wheeler, "Transmission Line Properties of Parallel Strips Separated by a Dielectric Sheet", IEEE Trans Microwave Theory Tech.,vol.MTT-13, pp 172-185, March 1965