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MICROWAVE-BASED LOW-COST INSTRUMENT FOR FILM THICKNESS MEASUREMENT

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***Abstract*--The use of inverted microstrip resonators for non-contacting real time thickness measurement of thin liquid or solid films and coatings described earlier [1] has been developed into a low-cost instrument for general laboratory or industrial use. Although microwave-based, the instrument to be described functions without microwave test equipment. Insight into the oscillator behavior was obtained by the use of equivalent circuits and an ABCD matrix technique while several options were considered for the frequency discrimination function. The system successfully determined, the film thicknesses of water, enamel paint, silicone rubber, and copper sheet metal.**

I. INTRODUCTION

This paper treats the development and some microwave aspects of a low cost instrument that uses microwave techniques to measure the thickness of films and coatings in the range 0.1mm and up. It follows in the footsteps of work described earlier, in which a non-contacting microwave technique was used to measure the thickness of a stationary or moving liquid film on a metal plane [1,2]. In that technique a strip conductor is suspended over the film, to form a suspended microstrip line as in Fig. 1. When this line is supplied with a microwave signal of constant frequency, the wavelength of propagation is a function of the thickness of the liquid film, so that, once calibrated, a measurement of the wavelength or the corresponding input impedance translates into the thickness of the liquid film.

In an alternate scheme, also described earlier, a fixed length of strip conductor was suspended over the film and gap-coupled to an input microstrip line, thereby forming an inverted microstrip resonator. The thickness of the film was determined by measuring the resonant frequency of the structure. Both of these methods required external microwave sources and measurement equipment. The most recent method obviates the need for source and measuring equipment. Instead, by coupling the resonator to an amplifier with positive feedback, a film thickness-driven oscillator is formed. This is followed by a discriminator/video detector circuit that essentially converts frequency deviation from a reference to a correlatable dc output voltage, thereby giving a read-out of film thickness. A block diagram of this measurement system, together with the various functions of the components, is seen in Fig. 2.

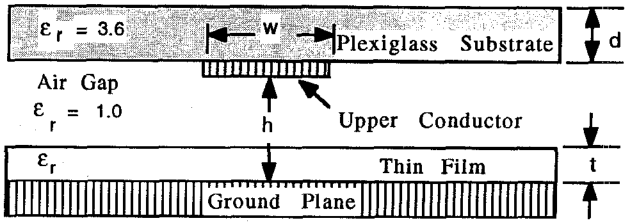
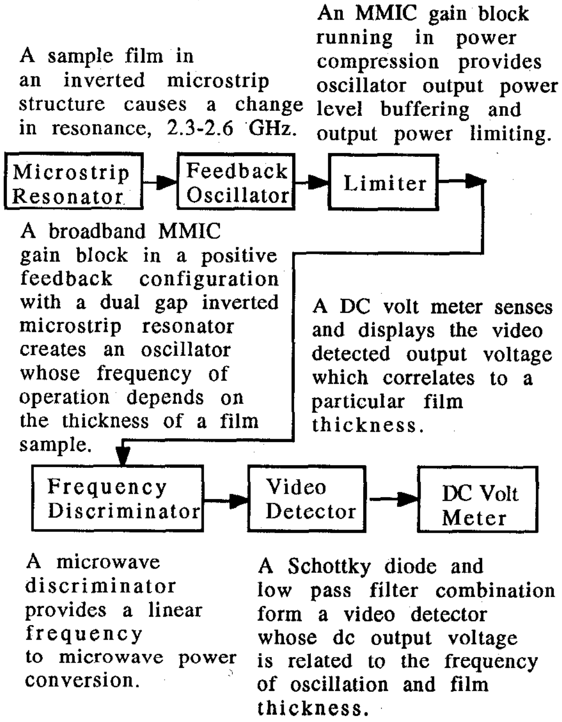


Fig. 1. Inverted microstrip.



A sample film in an inverted microstrip structure causes a change in resonance, 2.3-2.6 GHz.

An MMIC gain block running in power compression provides oscillator output power level buffering and output power limiting.

A broadband MMIC gain block in a positive feedback configuration with a dual gap inverted microstrip resonator creates an oscillator whose frequency of operation depends on the thickness of a film sample.

A DC volt meter senses and displays the video detected output voltage which correlates to a particular film thickness.

A microwave discriminator provides a linear frequency to microwave power conversion.

A Schottky diode and low pass filter combination form a video detector whose dc output voltage is related to the frequency of oscillation and film thickness.

Fig. 2. 2.3-2.6 GHz film measurement system.

II. SYSTEM COMPONENTS AND OPERATION

The principle underlying the change in resonant frequency with film thickness has been given earlier[1,2]. In summary, this resonant frequency f_r is given by

$$f_r = n\,c\,/\,(2\,L \cdot \sqrt{\varepsilon_{eff}}). \qquad \text{Eq.(1)}$$

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where n is an integer, c is the velocity of light, and L is the length of the microstrip line, and

$$\varepsilon_{eff} = C_\lambda / C_l = 1/[(t/h)(\varepsilon_0/\varepsilon_f - 1) + 1] \;. \qquad \text{Eq.(2)}$$

Here, C_λ = capacitance per unit length of the microstrip line filled with a dielectric;

C_l = the capacitance of the same microstrip line without dielectric.

As a practical matter, Equation (1) must be altered to account for the capacitive coupling and field fringing effects at the coupling gaps. One way to account for these capacitive effects is to assume that the line has been increased by a small incremental length ΔL making the effective length of the resonator L' [3]. Thus,

$$L' = L + \Delta L \;. \qquad \text{Eq.(3)}$$

$$f_r = n\,c \,/\, (2[L + \Delta L] \cdot \sqrt{\varepsilon_{eff}}\,). \qquad \text{Eq.(4)}$$

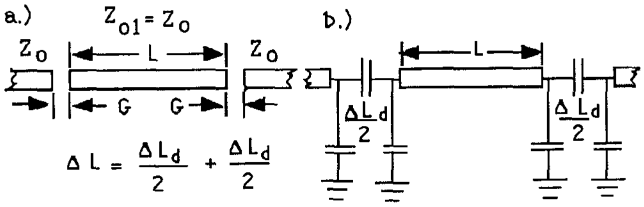


Fig. 3. a.) Double gap resonator and b.) equivalent circuit [4].

Olyphant and Ball [3] present a useful technique for experimentally determining the effective permittivity of the resonator as well as the incremental ΔL using a set of 'n' half-wave length resonators and rearranging Equation (4) into the form of a linear equation in (f/n) whose slope is ($-\Delta L$) and whose intercept is $y = f_r(L/n)$ at $(f_r/n) = 0$ when the resonator is assumed to be of semi-infinite length.

$$f_r\,(L/n) = c\,/2\sqrt{\varepsilon_{eff}} - \Delta L(f_r/n) \quad (\text{Hz}\cdot\text{m}). \qquad \text{Eq.(5)}$$

This technique was used in the design.

The resonant frequency deviation Δf_r from f_0 with no film in the resonator can be expressed as:

$$\Delta f_r = f_0 - f_r, \qquad \text{Eq. (6)}$$

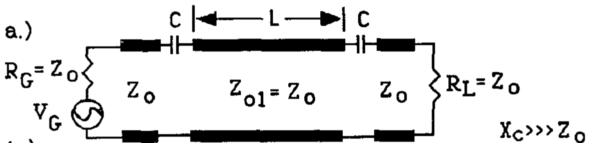
By combining these earlier equations, an expression for the change in the frequency of resonance can be obtained.

$$f_r = \frac{nc}{2L'}\left[(t/h)\left(\frac{\varepsilon_0}{\varepsilon_{film}} - 1\right) + 1\right]^{1/2} \qquad \text{Eq.(7)}$$

where $(t/h) < 0.7$. The frequency deviation can be approximated by

$$\Delta f_r = f_0 - f_r \approx - (nc/4L')(t/h)[(\varepsilon_0/\varepsilon_{eff}) - 1] \;. \qquad \text{Eq.(8)]}$$

Rizzi [4] offers a more analytical approach to the design and modeling of double gap microstrip resonators by use of ABCD parameters from which he derives expressions for resonance and the loaded Q, Q_L. (See Fig. 4.)



b.)

$$\begin{bmatrix} \bar{A} & \bar{B} \\ \bar{C} & \bar{D} \end{bmatrix} = \begin{bmatrix} 1 & -j\bar{X}_c \\ 0 & 1 \end{bmatrix} = \begin{bmatrix} \cos\phi & j\sin\phi \\ j\sin\phi & \cos\phi \end{bmatrix} \begin{bmatrix} 1 & -j\bar{X}_c \\ 0 & 1 \end{bmatrix} \qquad \bar{X}_c \equiv \frac{X_c}{Z_0} \quad \phi = \beta L$$

$$\begin{bmatrix} \bar{A} & \bar{B} \\ \bar{C} & \bar{D} \end{bmatrix} = \begin{bmatrix} (\cos\phi + \bar{X}_c \sin\phi) & j\{(1-\bar{X}_c^2)\sin\phi - 2\bar{X}_c\cos\phi\} \\ j\sin\phi & (\cos\phi + \bar{X}_c\sin\phi) \end{bmatrix}$$

Fig. 4. a.) Equivalent circuit and b.) normalized ABCD matrix for a double gap resonator [4].

The resonance condition can be found by using the ABCD matrix to find an expression for the insertion loss L_I and solving for the minimum loss condition at resonance i.e. $L_I = 0$ where $\phi_r = \beta L$.

$$L_I = 10 \log\left[1 + \bar{X}_c^2\left(\cos\phi + \frac{\bar{X}_c}{2}\sin\phi\right)^2\right] \qquad \text{Eq.(9)}$$

$$\phi_r = \pi - \arctan\left[\frac{2}{\bar{X}_{cr}}\right] = \frac{2\pi L}{\lambda_r} \qquad \text{Eq.(10)}$$

where X_{cr} is the value of the gap capacitive reactance at resonance and $\lambda_r = V_p/f_r$. The frequency of resonance f_r can be expressed as:

$$f_r = \frac{c\,\phi_r}{2\pi L\sqrt{\varepsilon_{eff}}} \qquad \text{Eq.(11)}$$

An expression for the loaded Q of the resonator also derived from the ABCD matrix insertion loss expression is found to be useful for estimating the value of the capacitance C needed for resonance by choosing a loaded Q from a desired resonator 3 dB bandwidth around a desired f_r and assuming that the value of ϕ_r is near π and $X_{cr} >>> Zo$.

$$Q_L = \phi_r \frac{\bar{X}_{cr}}{4}\sqrt{\bar{X}_{cr}^2 + 4} = \frac{f_r}{3\text{ dB Bandwidth}} \qquad \text{Eq.(12)}$$

Use of this procedure, however, requires knowledge of or the ability to simulate the capacitance-gap characteristic of the microstrip gap in order to obtain the proper gap spacing. Equation (10) can be used to calculate the correct upper conductor line length, L.

III. THE SYSTEM

In the system block diagram shown in Fig. 2, the inverted microstrip resonator is used to form a film thickness driven oscillator whose output is subsequently gain buffered and power limited to remove any output detector voltage variation due to changes in signal amplitude rather than frequency. A frequency discriminator and video detector combination then converts oscillator frequency variation due to film thickness inputs to a correlatable dc output voltage.

After considering the use of both single gap and double gap microstrip resonators for formation of an oscillator, the double gap resonator used in a feedback configuration with an MMIC gain block appeared to be less complex than a negative resistance oscillator and single gap resonator combination [5]. The basic equations required for a feedback oscillator may be expressed as: [6]

$$\phi_A + \phi_R + \phi_C = 2\pi n \qquad \text{Eq.(13)}$$
where $n = 0, 1, 2, \ldots$

$$G_A - L_R - L_C > 0 \text{ dB} \qquad \text{Eq.(14)}$$

Equation (13) states that the sum of the insertion phases, ϕ_A, ϕ_R, and ϕ_C, around the feedback loop of the oscillator must be a multiple of 2π. ϕ_A, ϕ_R, and ϕ_C represent the insertion phases at resonance of the amplifier gain block, frequency selective resonator, and feedback circuitry respectively. Equation (14) states that the "open loop" small signal gain at resonance, $G = G_A - L_R - L_C$, around the feedback path must be greater than unity, or 0 dB. G_A represents the gain of the amplifier, while L_R and L_C represent the loss of the resonator and feedback circuitry respectively.

By using a second broadband MMIC gain block running in power compression for power limiting, the oscillator output power level was held to around +13dBm and the input power level was buffered such that the oscillator input power level can vary from +2 to +13dBm and the output power level will remain at a constant +13dBm.

The purpose of the microwave discriminator and video detector combination is to linearly convert an input frequency to a correlatable output voltage. As seen in Fig. 5, four different kinds of frequency discriminator schemes were considered.

The first combining technique[7] (Fig.5a.) makes use of the special characteristics of the "Magic tee". A major drawback to the use of the Magic tee discriminator is its size and cost. A microstrip TEM version of this circuit using 180° Rat Race hybrids may provide an alternative, however.

The delay line-phase detector method using differential voltages generated by square law detectors is borrowed from polar display applications such as those used in network analyzers to display phase relationships. Again, this method is overly complex as the phase discriminator itself is formed by the combination of three 90° hybrids and one 180° hybrid.

The third type of frequency discriminator considered (Fig. 5c.) also uses the equal phase splitter and delay line technique to create a frequency dependent phase difference but mixes the delay and reference signal to produce a dc voltage.

The final frequency discrimination technique considered and the one which was used is the simplest conceptually and the most likely to consume the least space with the least expense. This technique simply makes use of the frequency response nature of a network. Since all that is really required for frequency discrimination is to create a linear conversion between some input frequency and an output voltage, any network, filter or input diode match which is sufficiently linear for the task will do. The network of Fig. 5d with a $\lambda/4$ delay line performs just such a function in combination with a biased Schottky barrier diode running in a peak detected mode capable of generating several volts dc.

The entire prototype system comprising the resonator, oscillator, limiter, and frequency discriminator-video detector has an estimated materials cost of under $10 U.S. and was placed on a 9x18x1cm Plexiglass board.

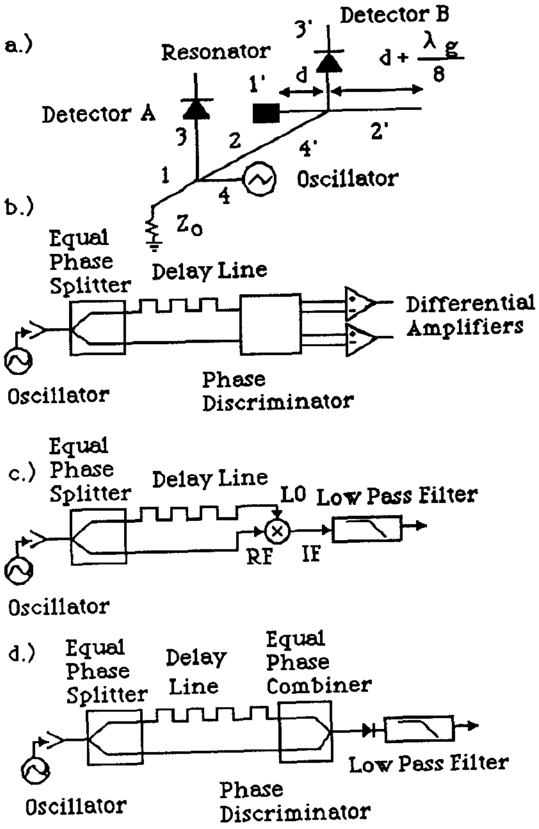


Fig. 5. Frequency discriminators a.) hybrid, b.) delay line - phase detector, c.) delay line mixer, d.) frequency response.

IV. EXPERIMENTAL RESULTS

The system successfully determined the thicknesses of a variety of dielectric materials including water, enamel paint, and silicone rubber with thicknesses under 2mm as well as sheet copper. Samples sizes were of uniform shape and no smaller than (2.54 cm x 1.27 cm).

It is noted that for copper the frequency of resonance increases with thickness rather than decreases, as it does for a dielectric. This is because the inductance per unit length along the resonating line decreases faster than the capacitance per unit length along the line increases as the thickness is increased. It should also be noted that the system was retuned towards the low frequency end of the operating band during this measurement for copper so that results could be obtained within the band of the system operation.

V. CONCLUSION

The ability to measure the thickness of various films and coatings in a non-contacting, non-destructive, real time, and inexpensive manner could find a number of applications. In manufacturing, for example, the instrument described here could be used to monitor the application of curing paints, coating, or films in static or real time feedback control systems. Since the resonator's frequency of resonance is also dependent

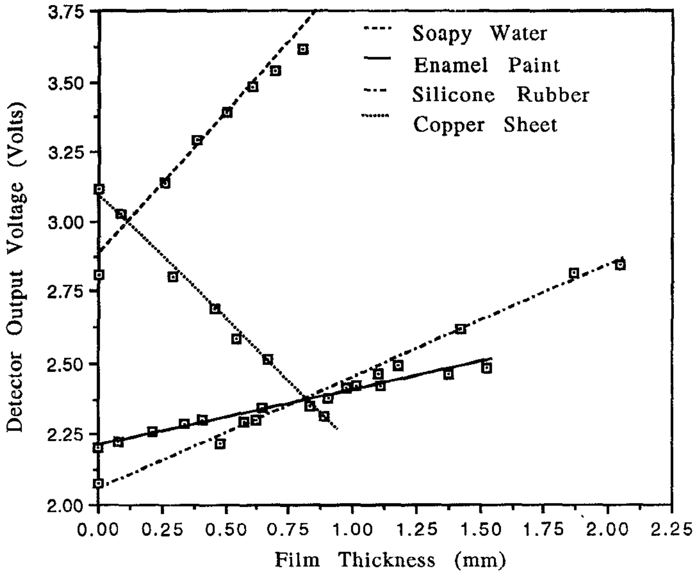


Fig. 6. Film thickness data , W=3.6 mm, G=1.52 mm, h=2.87 mm, d=2.31 mm, and L=47.2 mm.

on the relative permittivity of the sample, it might also be used to monitor the processing of materials which undergo changes in permittivity or thickness as chemical or physical properties change during manufacture. Curing plastics, binding materials, or solidifying liquids and pastes might provide likely candidates for use.

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