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A Theoretical Model for Thin Film Ferroelectric Coupled Microstripline Phase Shifters

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A Theoretical Model for Thin Film Ferroelectric Coupled Microstripline Phase Shifters

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Abstract—Novel microwave phase shifters consisting of coupled microstriplines on thin ferroelectric films have been demonstrated recently. A theoretical model useful for predicting the propagation characteristics (insertion phase shift, dielectric loss, impedance, and bandwidth) is presented here. The model is based on a variational solution for line capacitance and coupled strip transmission line theory.

Index Terms—Ferroelectricity, Phase shifter circuits

I. Introduction

Recently, microwave phase shifters have been demonstrated that used thin ferroelectric films and superconducting or normal metal coupled microstrip lines that also served as biasing electrodes [1,2]. The insertion loss of these Ku- and K-band phase shifters was substantially better than their semiconductor microwave integrated circuit counterparts. More than 400 of continuous phase shift with about 4 dB loss at 77 K was obtained using $YBa_2Cu_3O_{7-\delta}$ electrodes and the incipient ferroelectric SrTiO3, sequentially deposited onto 0.25 mm thick $LaAlO_3$ substrates. An insertion loss of about 5 dB was obtained from a similar design using gold electrodes and Ba_xSr_{1-x} TiO_3 films on $LaAlO_3$ at room temperature. The films were grown by pulsed laser ablation.

These low loss phase shifters enable a new type of phased array antenna called a ferroelectric reflectarray [3,4]. A reflectarray antenna combines the best features of a gimbaled parabolic dish (i.e., high efficiency and low cost) and a direct radiating array (e.g., vibration-free scanning). It consists of a surface of printed elements illuminated by a radiating feed. The energy from the feed can be re-radiated to form a cophasal beam. In the past, reflectarray antennas have been implemented using spiral elements interconnected with diodes to achieve far field phase shift [5]. However, their performance has been limited by the losses of the phase shifting elements and the finite number of phase shifting bits. Fixed beam printed microstrip patch reflectarrays have also been reported [6]. But a viable technique for including variable phase shift with printed radiators was elusive. The phase shifters analyzed here hold promise for this application because they are compact, low-loss, and can be fabricated lithographically on the same surface as the radiating element.

II. Variational Formulation of Line Capacitance

The designs are based on a series of coupled microstriplines interconnected with short sections of nominally 50 Ω microstrip. Bias up to 400 V is applied to the sections via printed bias-tees consisting of a quarter-wave radial stub in series with a very high impedance quarter-wave microstrip. A sketch of the cross-section is shown in Fig. 1. By concentrating the fields in the odd mode, the phase shift per unit length is maximized and by using the film in thin film form the effects of high loss tangent are minimized. The amount of phase shift can be increased by cascading coupled line sections. Though methods for calculating the propagation parameters of coupled transmission lines are well known, coupled lines on stratified substrates are difficult to analyze. And the high permittivity of the ferroelectric layer causes unacceptably long computation time by full-wave electromagnetic simulators because the geometry must be fractured into many thousands of cells. The multi-layer structure is analyzed here using a computationally efficient variational method to calculate the complex propagation constant and characteristic impedance. The method is quite general and can be used for multiple layers of various dielectrics or other types of transmission lines. For example, a multi-layer microstrip can be analyzed by allowing the strip spacing (s) to become much greater than the substrate thickness (h) or strip width (w).

In the case of cascaded coupled lines increasing phase shift occurs at the expense of bandwidth since the structure

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resembles a multi-pole filter. Changing the dielectric constant of the ferroelectric film to change insertion phase also modifies the pass band characteristic, resulting in a net bandwidth of ≈10%. The bandwidth compression phenomenon is shown in Fig. 2 for an 8-section $Ba_{0.60}Sr_{0.40}TiO_3$ on 0.3 mm thick MgO phase shifter [10]. Roll-off at the high frequency end is attributed to the bias tees. The impedance matrix of the cascade network can be derived by traditional coupled line theory using the superposition of even and odd mode excitation [7]. Then an equivalent S-parameter model can be extracted and used to predict the pass-band characteristics of the phase shifter.

Line capacitance (C) can be calculated by adapting the quasi-TEM variational expression from Koul and Bhat [8]

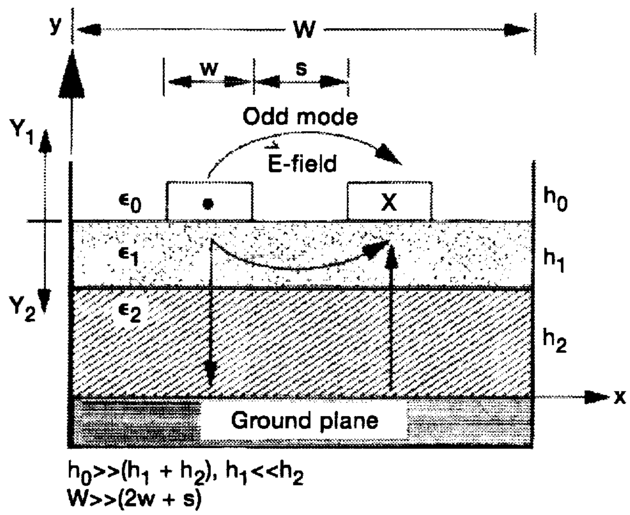

Figure 1.—Cross section of a coupled microstripline phase shifter. The ferroelectric film and substrand thickness are h_1 and h_2, respectively. The X denotes electric current directions. Maximum voltage coupling occurs when the line length (l) is $\lambda_g/4$.

and using the transverse transmission line method of Crampagne, Ahmadpanah and Guiraud [9].

$$1/C=\left(1/q^2\right)\int_s \rho(x,y)\phi(x,y)dl \qquad (1)$$

where $q = \int_s \rho(x,y)dl$ and $\phi(x,y)$ is the potential distribution which satisfies Poisson's equation and can be expressed in terms of the admittances ($Y_{1,2}$) at the charge plane ($y = h_1 + h_2$) looking in the positive and negative y directions. According to Crampagne, et al., the admittance can be expressed in terms of the dielectric constant of each layer. The charge distribution $\rho(x,y)$ for the even and odd mode excitations was assumed to have the form:

$$\rho(x,y)_{e,o} = 1/w\left\{1 + A_{e,o}\left|(2/w)(x-(W-s-w)/2\right|^3\right\},$$

$$(W-s)/2 - w < x < (W-s)/2 \qquad (2)$$

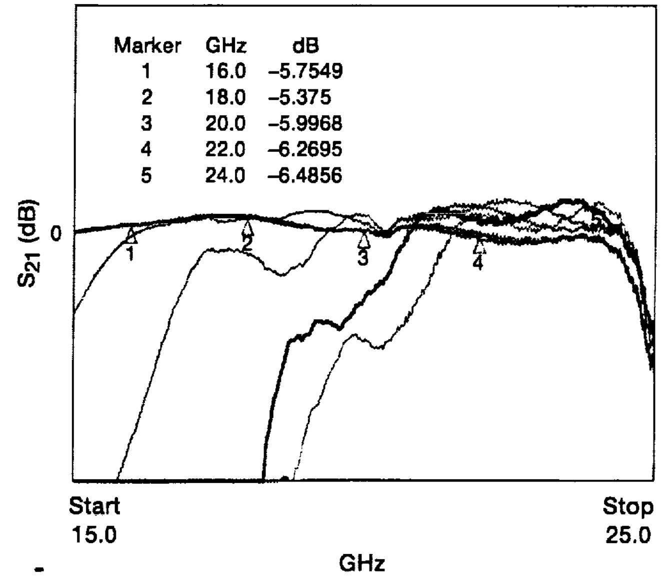

Figure 2.—Bandpass characteristics of an 8-section coupled microstripline phase shifter using a 400 nm $Ba_xSr_{1-x}TiO_3$ film on 0.3 mm thick MgO. Bias voltages from left to right are 0, 25, 50, 100, and 150V. Coupled lines are 350 μm long and 30μm wide separated by 7.5 μm.

where $A_{e,o}$ are constants derived by maximizing the even (e) and odd (o) mode capacitance. That is, a trial function that maximizes capacitance yields the most accurate result. It is an attribute of the variational method that the trial function for the charge distribution does not have to be known precisely a priori in order to evaluate capacitance. The expression for A_o is given in (3) to (6).

The variational approach is only valid for electrically thin stratified substrates. But since the coupling to surface waves represents an operational limit, it is appropriate for practical microwave applications.

$$A_o = \sum_n \frac{(4\cdot M(n)-L(n))\cdot L(n)\cdot g(n)}{(4\cdot M(n)-L(n))\cdot M(n)\cdot g(n)} \qquad (3)$$

where

$$L(n)=\sin\left[\frac{n\cdot\pi}{2\cdot W}\cdot(W-s-w)\right]\cdot\sin\left(\frac{n\cdot\pi\cdot w}{2\cdot W}\right) \qquad (4)$$

and

$$g(n)=\frac{4}{n\cdot\pi\cdot Y(n)}\cdot\left(\frac{2\cdot W}{n\cdot\pi\cdot w}\right)^2 \qquad (5)$$

and

$$M(n)=\left(\frac{2\cdot W}{n\cdot\pi\cdot w}\right)^3\cdot\sin\left[\frac{n\cdot\pi}{2\cdot W}\cdot(W-s-w)\right]\cdot\left[3\cdot\left[\left(\frac{n\cdot\pi\cdot w}{2\cdot W}\right)^2-2\right]\right.$$

$$\left.\times\cos\left(\frac{n\cdot\pi\cdot w}{2\cdot W}\right)+\left(\frac{n\cdot\pi\cdot w}{2\cdot W}\right)\cdot\left[\left(\frac{n\cdot\pi\cdot w}{2\cdot W}\right)^2-6\right]\cdot\sin\left(\frac{n\cdot\pi\cdot w}{2\cdot W}\right)+6\right]$$

(6)

Here, n is summed over all odd integers. A summation from 1 to 1999 was adequate for convergence. A_e is derived similarly, by summing over all even integers.

In the even mode currents in the strips are equal in amplitude and flowing in the same direction. In the odd mode currents in the strips are equal in amplitude but flow in opposite directions. So, Z_{oe} is the characteristic impedance of one strip to ground with equal currents in the same direction and Z_{oo} is the characteristic impedance of one strip to ground with equal currents in opposite directions. The microstrip mode exists when s >> h,w and $Z_{oe} = Z_{oo}$. The thin FE film is most effective when the phase velocity is dominated by the odd mode fields. The propagation constant is given by:

$$\beta = \omega / v_p = (\pi / \lambda_o)\left(\varepsilon_{EVEN}{}^{1/2} + \varepsilon_{ODD}{}^{1/2}\right) \qquad (7)$$

where

$$\varepsilon_{EVEN} = C_E / C_{Eair} \quad \text{and} \quad \varepsilon_{ODD} = C_O / C_{Oair} \qquad (8)$$

and C_{Eair} and C_{Oair} are obtained by replacing all dielectrics with air (i.e., $\varepsilon_r = 1$).

The admittance at the charge plane, corresponding to Fig. 1, is easily shown to be:

$$Y(n) = \varepsilon_0 \cdot \left[\varepsilon_1 \cdot \left[\frac{\varepsilon_2 \cdot \coth\left(\frac{n \cdot \pi \cdot h_2}{W}\right) \cdot \coth\left(\frac{n \cdot \pi \cdot h_1}{W}\right) + \varepsilon_1}{\varepsilon_1 \cdot \coth\left(\frac{n \cdot \pi \cdot h_1}{W}\right) + \varepsilon_2 \cdot \coth\left(\frac{n \cdot \pi \cdot h_2}{W}\right)} \right] + \coth\left(\frac{n \cdot \pi \cdot h_0}{W}\right) \right] \qquad (9)$$

A more general formulation for the admittance consisting of an additional dielectric layer, say air, with thickness h_3 between the host substrate and the ground plane can be expressed as:

$$Y(n) = \varepsilon_0 \cdot \coth\left(n \cdot \pi \cdot \frac{h_0}{W}\right) + \varepsilon_0 \cdot \varepsilon_1 \cdot \frac{\left[\varepsilon_2 \cdot \varepsilon_0 \cdot \frac{\left(\coth\left(n \cdot \pi \cdot \frac{h_3}{W}\right) + \varepsilon_2 \cdot \tanh\left(n \cdot \pi \cdot \frac{h_2}{W}\right)\right)}{\left(\varepsilon_2 + \coth\left(n \cdot \pi \cdot \frac{h_3}{W}\right) \cdot \tanh\left(n \cdot \pi \cdot \frac{h_2}{W}\right)\right)} + \varepsilon_0 \cdot \varepsilon_1 \cdot \tanh\left(n \cdot \pi \cdot \frac{h_1}{W}\right) \right]}{\left[\varepsilon_0 \cdot \varepsilon_1 + \varepsilon_2 \cdot \varepsilon_0 \cdot \frac{\left(\coth\left(n \cdot \pi \cdot \frac{h_3}{W}\right) + \varepsilon_2 \cdot \tanh\left(n \cdot \pi \cdot \frac{h_2}{W}\right)\right)}{\left(\varepsilon_2 + \coth\left(n \cdot \pi \cdot \frac{h_3}{W}\right) \cdot \tanh\left(n \cdot \pi \cdot \frac{h_2}{W}\right)\right)} \cdot \tanh\left(n \cdot \pi \cdot \frac{h_1}{W}\right) \right]} \qquad (10)$$

Finally, the odd mode capacitance becomes:

$$C_O = \frac{\left(1 + \frac{A_o}{4}\right)^2}{\sum_n g(n) \cdot (L(n) + A_o \cdot M(n))^2} \qquad (11)$$

And the even mode capacitance is similarly obtained, by summing over even integers with A_e replacing A_o. For a TEM transmission line, the characteristic impedance is obtained as $Zo = [(CCair)^{1/2}c]^{-1}$ where c is the speed of light in vacuum.

In general, minimum attenuation is obtained when the effect of the ground plane loss is minimized (i.e., in the odd or balanced mode current flows into one strip and returns through the other). Maximum attenuation occurs in the even or unbalanced mode when equal currents flow into both strips and return through ground.

A comparison between the quasi-TEM approximation and a full-wave electromagnetic simulation is given in Table I for microstrip. The spacing s was allowed to increase just until the even and odd mode capacitance was equivalent. Choosing arbitrarily large values for s yields anomalous results.

TABLE I.—MODELED DATA FOR A 2 μM FERROELECTRIC LAYER ON .25 mm THICK $LaAlO_3$, Zo = 50 Ohms, s >> w,h (MICROSTRIP MODE).

Er Ferroelectric Layer	*Eeff (Sonnet ™)*	*Eeff (Variational)*
300	18.76	18.43
600	21.34	21.00
900	23.49	23.09
1200	25.41	24.93
1500	27.18	26.59
1800	28.84	28.12

The quasi-TEM solution runs on a Pentium II machine in about 10 sec regardless of the value entered for the ferroelectric film dielectric constant. The same calculation on a commercial electromagnetic simulator using finite element techniques can take several hours because the geometry must be fractured into thousands of cells for these very high dielectric constants.

For coupled strip lines using the superposition of the even and odd modes:

$$Z_{11} = Z_{22} = -j/2(Z_{Oe} + Z_{Oo})\cot\theta(V) \qquad (12)$$

and

$$Z_{13} = Z_{31} = -j/2(Z_{Oe} - Z_{Oo})\csc\theta(V) \qquad (13)$$

Where θ(V) is the voltage dependent electrical length of the coupled lines obtained from (7) and the physical length of the coupled region (i.e., θ(V) = β(V)l). The characteristic impedance can be expressed as the geometric mean of the even and odd mode impedance such that:

$$Z_O = (Z_{Oe} Z_{Oo})^{1/2} \qquad (14)$$

Equation (14) is strictly valid only for pure TEM propagation and ignores frequency dependence. However, for a practical geometry with moderate coupling, like that considered here, the expression is appropriate. To facilitate the calculation, the Z-parameters are converted into ABCD or chain parameters for the cascaded sections. The conversion is:

$$A_f = \frac{Z11_f}{Z21_f} \qquad (15)$$

$$B_f = \frac{Z11_f \cdot Z22_f - Z12_f \cdot Z21_f}{Z21_f} \qquad (16)$$

$$C_f = \frac{1}{Z21_f} \qquad (17)$$

$$D_f = \frac{Z22_f}{Z21_f} \qquad (18)$$

where the indices are used to show frequency dependence. If the short intervening sections of microstrip line (between each coupled line section) are assumed to be of zero length the overall two-port chain matrix becomes:

$$\begin{pmatrix} a_i & b_i \\ c_i & d_i \end{pmatrix} = \begin{pmatrix} A_i & B_i \\ C_i & D_i \end{pmatrix} \cdot \begin{pmatrix} A_i & B_i \\ C_i & D_i \end{pmatrix} \cdot \begin{pmatrix} A_i & B_i \\ C_i & D_i \end{pmatrix} \cdot \begin{pmatrix} A_i & B_i \\ C_i & D_i \end{pmatrix}$$
$$\times \begin{pmatrix} A_i & B_i \\ C_i & D_i \end{pmatrix} \cdot \begin{pmatrix} A_i & B_i \\ C_i & D_i \end{pmatrix} \cdot \begin{pmatrix} A_i & B_i \\ C_i & D_i \end{pmatrix} \cdot \begin{pmatrix} A_i & B_i \\ C_i & D_i \end{pmatrix} \qquad (19)$$

Where the indices i and f are intended to be interchangeable. Finally, the two-port S-parameters of, in this case an 8-section phase shifter, can be converted back according to:

$$s11_i = \frac{a_i + \frac{b_i}{Z_o} - (c_i \cdot Z_o) - d_i}{a_i + \frac{b_i}{Z_o} + c_i \cdot Z_o + d_i} \qquad (20)$$

$$s21_i = \frac{2}{a_i + \frac{b_i}{Z_o} + c_i \cdot Z_o + d_i} \qquad (21)$$

The predicted bandpass characteristic for a $Ba_xSr_{1-x}TiO_3$ film on 0.25 mm thick $LaAlO_3$ is shown in Fig. 3 and is in good agreement with experiment.

The ferroelectric layer thickness is crucial to performance. In principle the phase shift for a 2 μm thick film is 2.2 times greater than that of a 0.5 μm film. However maintaining the crystalline quality of the pulse laser ablated $Ba_xSr_{1-x}TiO_3$ films past a thickness of 0.5 μm or so has proven to be difficult [10].

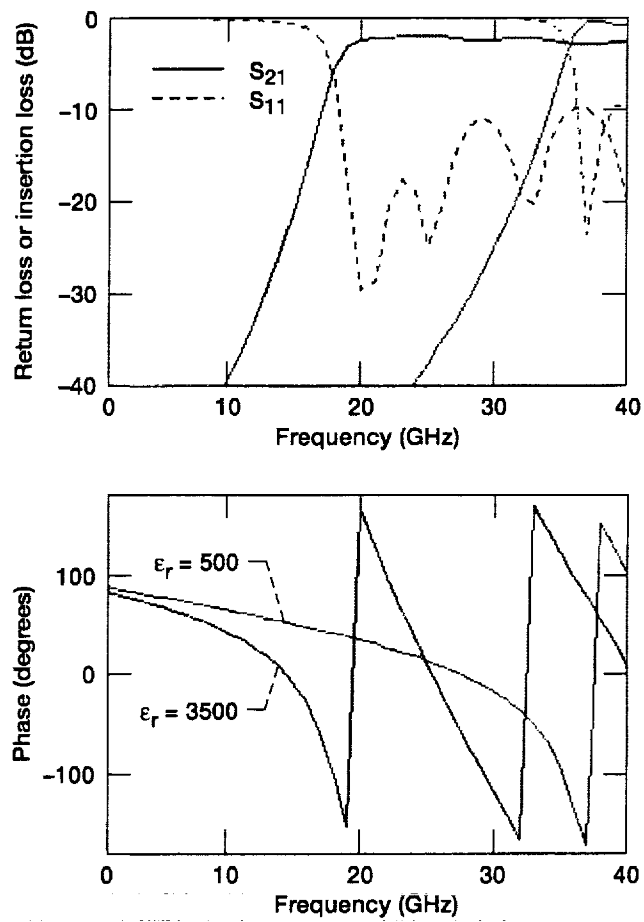

Figure 3.—Theoretical calculation for the bandpass characteristic of an 8-section coupled microstripline phase shifter using a 0.5 μm $Ba_xSr_{1-x}TiO_3$ film on 0.3 mm thick MgO. The coupled length was 350 μm, w = 30 μm, and s = 10 μm. The permittivity of the film was tuned from 3500 (1–j0.05) (no bias) to 500 (1–j0.005) (maximum bias).

III. Conclusions

The phase shifters analyzed here hold promise for reflectarray applications because they are compact, low-loss, and can be lithographed on the same surface as the radiating element. A key advantage of this technology is the relatively large feature size. Active devices at the frequencies of interest here would necessitate submicron gate length GaAs FETs. The finest feature size associated with the coupled line phase shifters is the electrode separation s, typically ≈ 10 μm. Whereas the GaAs FET performance is largely dictated by transconductance and hence carrier transit time across the gate region, the coupled line phase shifters are static devices. The electrode gap separation determines the degree of electromagnetic coupling and the dc potential required to tune the film. It is clear from the experimental and modeled data that the inherent dielectric loss of epitaxial ferroelectric films isn't necessarily devastating insofar as microwave device performance is concerned. Indeed the loss tangent of a thin dielectric film ($h_1 \leq 2$ μm) on a good substrate ($\tan\delta \leq 0.001$) can deteriorate substantially ($\tan\delta \leq 0.05$) before the insertion loss of the structures presented here is compromised. If the $\tan\delta$ of the ferroelectric film could be maintained at 0.005 or less, the films contribution to total loss would be essentially negligible except for the mismatch it introduces as it is tuned. But even a $\tan\delta$ of <0.05 is manageable and can result in a 3 dB loss phase shifter that would enable a practical scanning reflectarray antenna. A theoretical model for these devices has been developed and has been shown to be in good agreement with experiment and electromagnetic simulators. Both the insertion phase shift and pass-band characteristics can be closely approximated.

IV. References

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