Planar Microwave Integrated Phase-Shifter Design with High Purity Ferroelectric Material

Franco De Flaviis, N. G. Alexopoulos, *Fellow, IEEE,* and Oscar M. Stafsudd

***Abstract*—Ferroelectric materials (FEM's) are very attractive because their dielectric constant can be modulated under the effect of an externally applied electric field perpendicular to the direction of propagation of a microwave signal. FEM may be particularly useful for the development of a new family of planar phase shifters which operate up to X-band. The use of FEM in the microwave frequency range has been limited in the past due to the high losses of these materials $\tan\delta = 0.3$ at 3 GHz is typical for commercial $BaTiO_3$ (BTO) and due to the high electric field necessary to bias the structure in order to obtain substantial dielectric constant change. In this paper, how a significant reduction in material losses is possible is demonstrated. This is achieved by using a new sol–gel technique [1] to produce barium modified strontium titanium oxide [$Ba_{1-x}Sr_xTiO_3$ (BST)], which has ferroelectric properties at room temperature. Also demonstrated is how the use of thin ceramics reduces the required bias voltage below 250 V, with almost no power consumption required to induce a change in the dielectric constant. A phase shift of 165° was obtained at 2.4 GHz, with an insertion loss below 3 dB by using a bias voltage of 250 V. Due to the planar geometry and light weight of the device, it can be fully integrated in planar microwave structures.**

I. Introduction

IN THIS PAPER, the use of ferroelectric materials (FEM's) in ceramic form for the realization of a phase shifter operating at 2.4 GHz is designed and tested. The phase-shift capability of FEM results from the fact that if one is below the Curie temperature [2], the dielectric constant of such a material can be modulated under the effect of an electric bias field. Particularly if the electric field is applied perpendicularly to the direction of propagation of the electromagnetic signal, as shown in Fig. 1, the propagation constant ($\beta = 2\pi/\lambda$) of the signal [3] will depend upon the bias field since $\beta = 2\pi\sqrt{\varepsilon_r}/\lambda_0$ and $\varepsilon_r = \varepsilon_r(V_{\rm bias})$. The total wave delay will become a function of the bias field, and, therefore, will produce a phase shift $\Delta\phi = \Delta\beta l$, where l is the length of the line.

The reason why FEM has not been widely used for microwave applications to date is mainly due to the large bias voltage required to change the dielectric constant (typically a waveguide phase shifter based on FEM requires a bias voltage of 2 kV [4]), and to the high losses in the material. Use of a new sol–gel technique for the synthesis of high-quality low-loss barium modified strontium titanium oxide [$Ba_{1-x}Sr_xTiO_3$ (BST)], combined with the use of a thin ceramic structure,

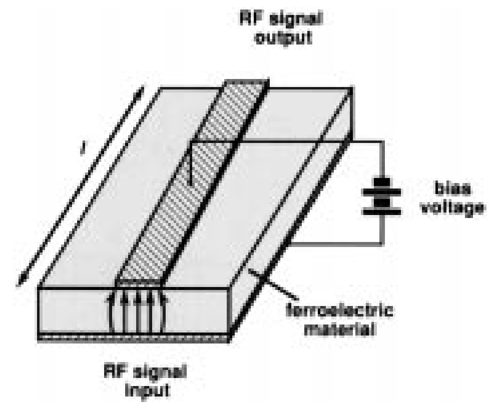

Fig. 1. Schematic structure of ferroelectric-microstrip-based phase shifter.

greatly reduces the insertion loss and the bias voltage. This yields a new family of devices competitive with other types of electrically tunable phase shifters.

Some basics on the new sol–gel technique which have been developed for the production of BST are shown in [1], while a more comprehensive description will be presented in a future publication. For confirmation of the superior quality of this material compared to commercial products, a set of resonant cavity measurement has been performed on different samples of barium strontium titanate with different percentages of strontium. The measurements employ an iris-coupled reaction-type cavity, constructed from standard rectangular waveguide operating in the TE_{101} mode. Results are reported in Table I. Use of this new family of thin ceramics (having thickness between 0.1–0.15 mm) integrated in a microstrip set-up, allows a phase shift greater than 160° with bias voltage below 250 V, and a total power consumption below 1 mW. This result is certainly a major improvement in terms of power requirements compared to ferromagnetic phase shifters, where more than two orders of magnitude of higher power level is required. The reason why an FEM requires less power is due to the fact that the driver energy required to change the properties of the material affects the electrostatic energy, and it is not dissipated. The major advantages of FEM-based phase shifters compared to ferromagnetic phase shifters, are the faster phase-shift capability, the smaller and lighter structure, and the higher the power-handling capability.

II. Phase Shifter Design and Construction

The schematic layout of the electrically tunable microstrip-based phase shifter is shown in Fig. 2 (the matching circuit

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The authors are with the University of California, Los Angeles, Department of Electrical Engineering, Los Angeles, CA 90095 USA.

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TABLE I
LOSS $\tan\delta$ FOR DIFFERENT SAMPLES OF FEM

Sample	*Heat treatment*	*Grain size (μm)*	*Curie temp. (°K)*	*Losses (tanδ)*
$SrTiO_3$	fired 900°C x 1hr	1	37	0.003604
$SrTiO_3$	refired 1200°C x 10hr	3	37	0.00322
$SrTiO_3$	refired 1500°C x 10hr	15	37	0.002957
$SrTiO_3$	refired in O_2 1500°C x 6hr	30	37	0.00208
$Ba_{.2}Sr_{.8}TiO_3$	fired in O_2 1500°C x 10hr	30	105	0.005615
$Ba_{.5}Sr_{.5}TiO_3$	fired in O_2 1500°C x 10hr	30	218	0.0304
$Ba_{.7}Sr_{.3}TiO_3$	fired 1375°C x 10hr	8	280	0.135
$Ba_{.8}Sr_{.2}TiO_3$	fired 1300°C x 1hr	8	324	0.081
$BaTiO_3$	fired 1300°C x 1hr	10	391	0.21
$BaTiO_3$ *commercial*	fired 1300°C x 1hr	9	387	0.32

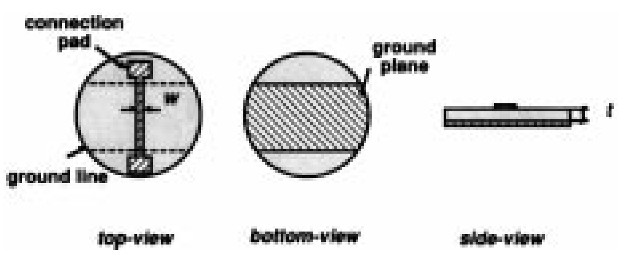

Fig. 2. Detailed layout of the planar ferroelectric phase shifter.

is not shown for clarity). As observed, the active part of the device consists of a microstrip line overlapping the ground plane. The ceramic disk thickness is 0.1 mm, the microstrip line is 50-μm wide, and the total length of the line is 8 mm. These three quantities are the basic parameters for the first phase-shifter design. The total length of the strip will determine the maximum phase shift, which can be obtained for a fixed change of the propagation constant ($\Delta\beta$), associated with the maximum bias voltage applied. The total phase shift ($\Delta\phi$) is given by $\Delta\phi = 2\pi\ell\sqrt{\varepsilon_r}/\lambda_0$ [3] . This means that longer strips will give larger phase shift for a fixed change in the propagation constant ($\Delta\beta$). In practice, it is very difficult to obtain good samples with diameter larger than 15 mm, so here, one compromises for a diameter of 10 mm. The ratio between the microstrip-line width (w) and the substrate thickness (t) will determine the characteristic impedance of the phase shifter (for a given dielectric constant). Because the substrate thickness is of the order of 0.1 mm and the effective dielectric constant of the ceramic is on the order of 600, one needs to choose w as small as possible in order to be able to match the circuit with a 50-Ω system. Widths below 50 μm are not very practical due to the associated high-ohmic resistance and to difficulties in the fabrication process. The thickness of the substrate also determines the required bias field to obtain a desired shift. For example, $BaTiO_3$ (BTO) and BST have a breakdown field strength of the order of 12 MV/m and, therefore, they require an electric-field strength of 2500 kV/m, in order to show a pronounced change in their dielectric constant. For this reason, in order to have a bias voltage below 300 V, a thickness of 0.15 mm or less must be used. A substrate thinner than 0.1 mm is impractical because aside from the fact that it is hard to manage, it also leads to a microstrip-line characteristic impedance which is too low to be useful.

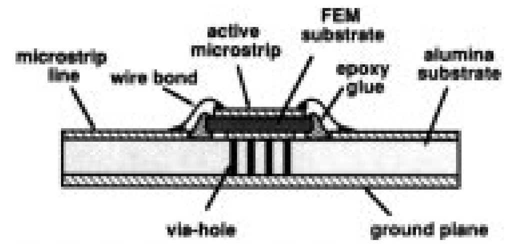

Fig. 3. Mounting schematic of the planar phase shifter on a conventional microstrip transmission line.

The tested ceramic sample was prepared using powder $Ba_{0.8}Sr_{0.2}TiO_3$, which was obtained from the sol–gel process. Subsequently, it was pressed into pellets at a pressure of 2000 kg/cm^2 prior to sintering at 1300 °C for 1 h. After the firing process, the sample is sanded down to 0.1 mm and the faces are polished using diamond wheels. The ground plane and the microstrip line are fabricated, evaporating three layers of metal such as chromium–copper–gold under vacuum to prevent oxidation. The chromium guarantees good cohesion to the ceramic, the copper serves as a buffer layer and makes cohesion between the nichrome and the gold, and the gold ensures electrical conduction. As observed in Fig. 2, the two extremes of the microstrip line have patches which facilitate the connection of the device with the circuit. These patches also allow multiple wire bonds in order to reduce the parasitic inductance associated with the connection. To reduce the parasitic capacitance associated with the larger patches, the ground plane of the sample extends only under the strip as seen in Fig. 2. The sample is mounted using wire bonds on the two ends of the 50-Ω microstrip line printed on alumina substrate. The ground contact is established using multiple via-holes. Epoxy glue is used to hold the sample on the circuit as visible in Fig. 3. If further reduction of parasitics due to the wire bond is desired, the wire bonds can be eliminated by mounting the device upside down. An extensive set of tests to determine the effect of wire bonds on the device performance is carried out in Section III.

III. MEASURED PERFORMANCE

In order to measure the S-parameters of this paper's circuit under different bias conditions, one needs to have suitable bias set-up which doesn't interfere with the microwave signal and avoids voltage leaks in the system. A sketch of the set-up utilized for this measurement is shown in Fig. 4. Use of a quarter-wave high-impedance line and a quarter-wave low-impedance open stub, ensure isolation of the bias from the microwave signal for a reasonably good bandwidth. Additional

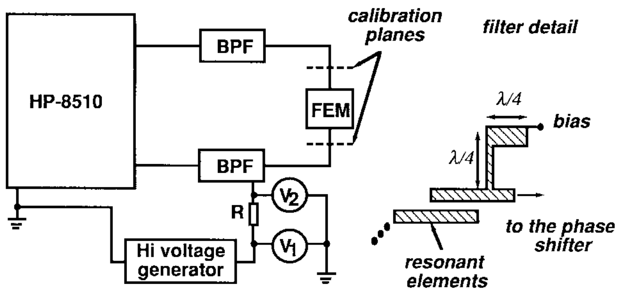

Fig. 4. Measurement set-up and filter details of the hardware used to measure the scattering parameters of the phase shifter.

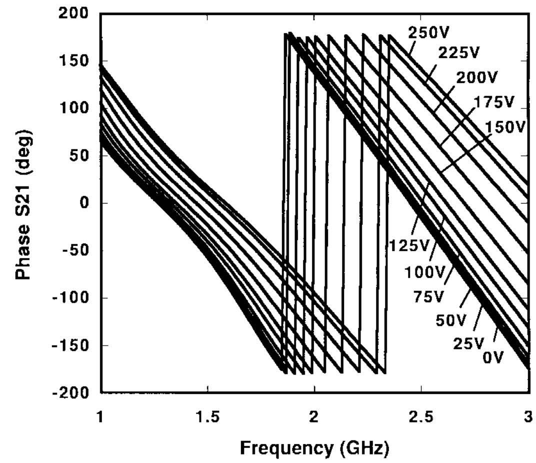

Fig. 5. S_{21} phase measurement for different bias conditions in the range of 1–3 GHz.

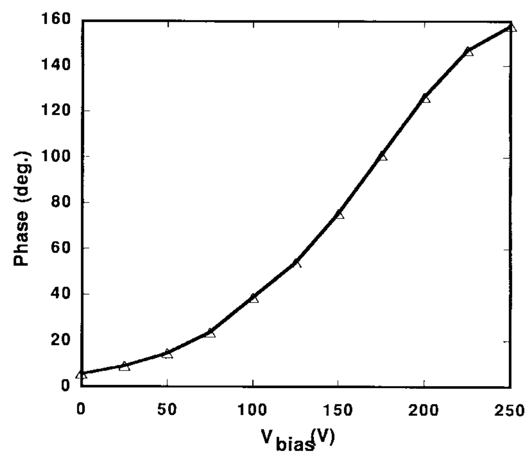

Fig. 6. RF phase shift versus applied bias field.

resistors were used (their value is not important since the circuit has almost no current flowing through) for safety reasons, so that in the case of a bias voltage short (due to breakdown of the ceramic), only limited current will flow through the circuit. To prevent bias voltage leakage in the rest of the circuit, two bandpass filters (BPF) centered at 2.43 GHz were used [5]. The filters were especially designed to withstand high static field, i.e., their corners were rounded and 2-mm-high electrical rigidity glue was deposited in the air gap. The calibration of the network analyzer is done after the filters, so the losses due to the filter are removed. Measurements for the S_{21} phase (in the frequency range between 1–3 GHz) are shown in Fig. 5 for selected bias points. More than 160° phase shift is achieved at 2.43 GHz with bias voltage around 250 V. The correspondence between phase shift and bias voltage is reported in Fig. 6. Very little saturation is reached, since the correspondence between the phase shift and the bias voltage is almost linear up to 250 V (after an initial inertia observed for $V_{\text{bias}} < 60$ V). This implies that further phase shift can be produced by biasing the device harder. Measurement of insertion loss in the same frequency range is shown in Fig. 7. Due to the way the matching circuit was designed, the magnitude of S_{21} having a total change smaller than 2 dB in the frequency range 1.6–3 GHz is observed here. This makes the device particularly suitable for broad-band operations. The insertion loss at 2.43 GHz is below 4 dB with no bias field, and reduces to 2.6 dB when a bias field of 250 V is applied. The corresponding S_{11} measurements are reported in Fig. 8. Particular attention was dedicated to the design of the matching circuit to minimize the variation of insertion losses between biased and unbiased conditions. This goal was successfully achieved with the observation that FEM losses decrease under bias condition [2] . This property can be used by having a very well-matched circuit in the unbiased condition ($S_{11} = -30$ dB) and worse matching under biased condition ($S_{11} = -11$ dB). With this approach, one reduces the maximum total change of insertion loss to 1.6 dB. This concept is essential for a good design of this type of phase shifter, and is well illustrated in Fig. 9, where the dashed line represents the S-parameters for the unbiased circuit, while the solid line corresponds to the biased condition. Another critical parameter for this design is the length of the wires which interconnect the phase shifter with the rest of the circuit. These wires, in fact, introduce a little series resistance and an additional parasitic inductance in series with the microstrip line, which at some critical value, can seriously compromise the matching of the circuit. Three different measurements were performed to find this critical length. The first measurement uses multiple wire bonds (very low inductance) having length around 1 mm on each side of the line; subsequently single wire bonds with increased length from 1 mm up to 2 mm are tested. Results of this test are summarized in Table II. Clearly, when the wire exceeds the length of 1 mm (a total of 2 mm on two sides), the performance degenerates. In this test, the wires used were made of gold, and they have a diameter of 0.0127 mm. Larger diameter also helps to reduce the associated parasitics.

IV. Device Modeling and Parameter Extraction

An initial rough model of the phase shifter can be designed using an HP microwave design simulator. This software allows the modelling of most of the circuit parts, including the parasitic elements and conductor ohmic resistance. Based on the implemented model and on the set of measured data shown in Figs. 5–8, one can extract the dielectric constant value and losses for the device as a function of the bias voltage. This characterization is quite simple since all the components in the circuit can be modeled accurately when the software is

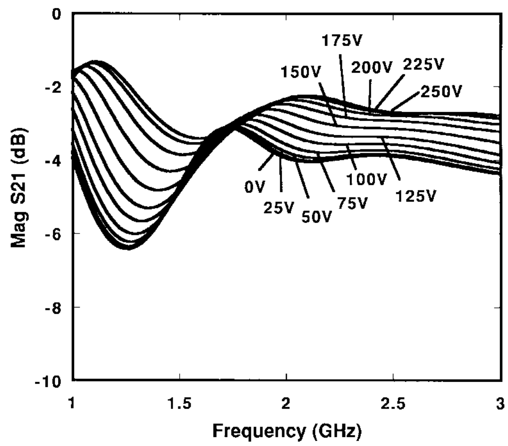

Fig. 7. S_{21} magnitude measurement for different bias conditions in the range of 1–3 GHz.

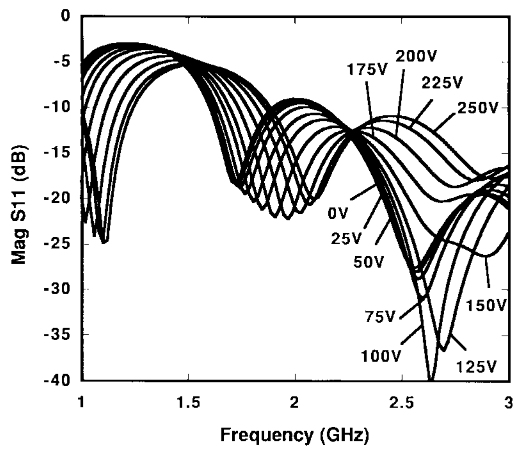

Fig. 8. S_{11} magnitude measurement for different bias conditions in the range of 1–3 GHz.

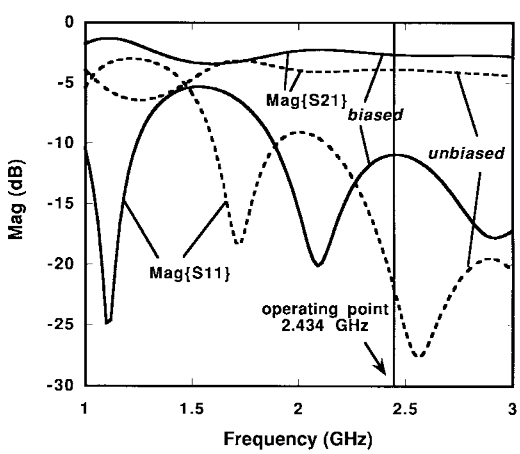

Fig. 9. S-parameter measurement for $V_{\text{bias}} = 0$ V and $V_{\text{bias}} = 250$ V in the range of 1–3 GHz.

used for such a low frequency. In this case, the only unknown parameters are the dielectric constant and the losses of the FEM substrate. Furthermore, the location of the S_{11} ripples or the phase shift of S_{21} is uniquely related to the value of the effective dielectric constant of the substrate, while the value of $\tan\delta$ will determine the average slope of S_{21} versus frequency.

TABLE II
EFFECT OF WIRE BOND LENGTH ON THE PERFORMANCE OF THE PHASE SHIFTER

Wire length (mm)	*Estimated inductance (nH)*	*Matching (dB)*	*Insertion loss (dB)*
1 (*double*)	0.5	-10.98	-2.66
1	1	-14.8	-2.73
1.5	1.5	-8.58	-3.4
2	2	-3.4	-8.58

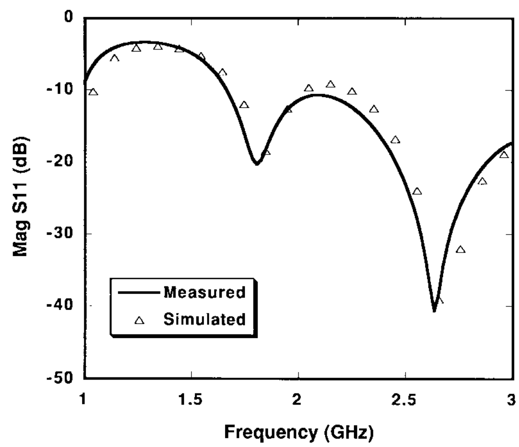

Fig. 10. S_{11} magnitude modeled versus measurement for $V_{\text{bias}} = 100$ V in the range of 1–3 GHz.

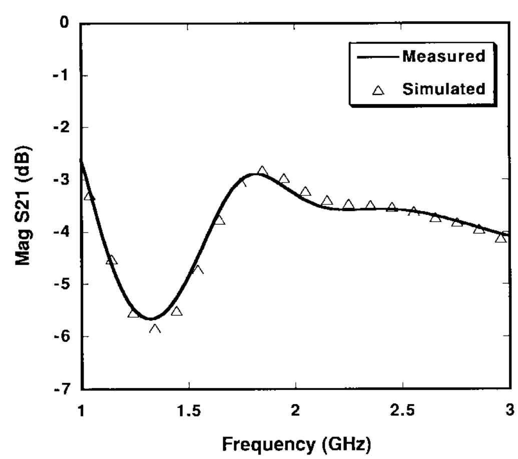

Fig. 11. S_{21} magnitude modeled versus measurement for $V_{\text{bias}} = 100$ V in the range of 1–3 GHz.

So, once one selects the dielectric constant for the substrate based on the phase shift, one just needs to vary the losses until the curves corresponding to the magnitude of S_{11} and S_{21} are matched. An example of parameter fitting at $V_{\text{bias}} = 100$ V is given in Figs. 10–12. Results of the modeling procedure at 2.43 GHz for all bias-voltages' conditions are shown in Figs. 13 and 14. Increase of bias voltage reduces the dielectric constant and losses in the material as expected from [2]. The lowest estimated $\tan\delta = 0.072$ at 2.43 GHz is achieved with bias voltage of 250 V.

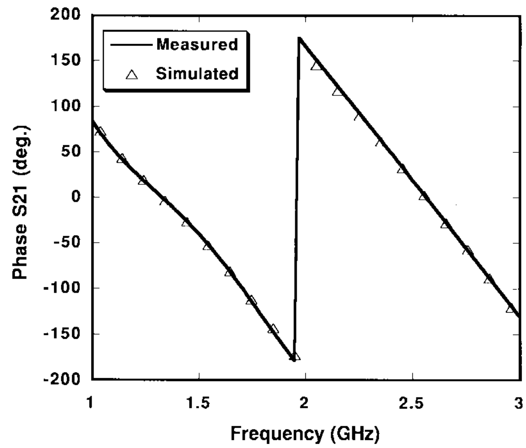

Fig. 12. S_{21} phase modeled versus measurement for $V_{\text{bias}} = 100$ V in the range of 1–3 GHz.

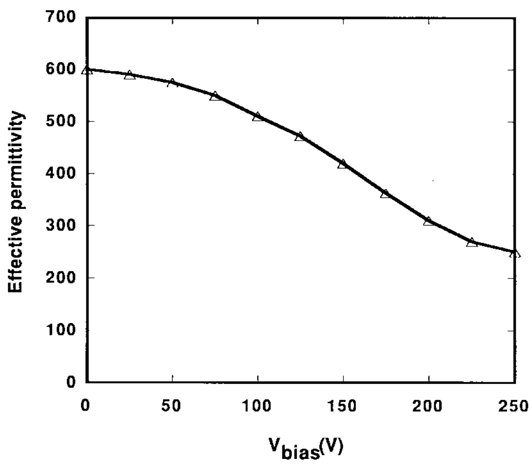

Fig. 13. Effective dielectric constant computed values versus applied bias field.

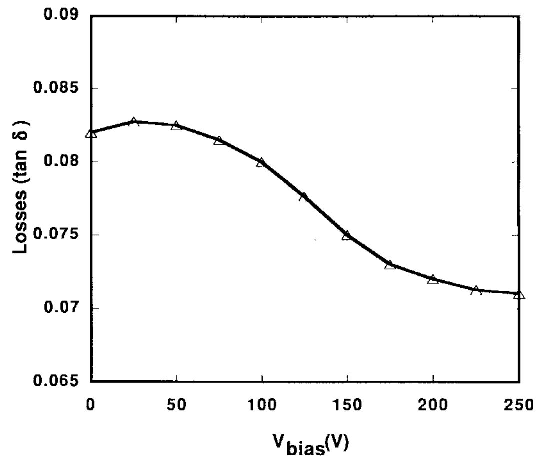

Fig. 14. Computed loss values versus applied bias field.

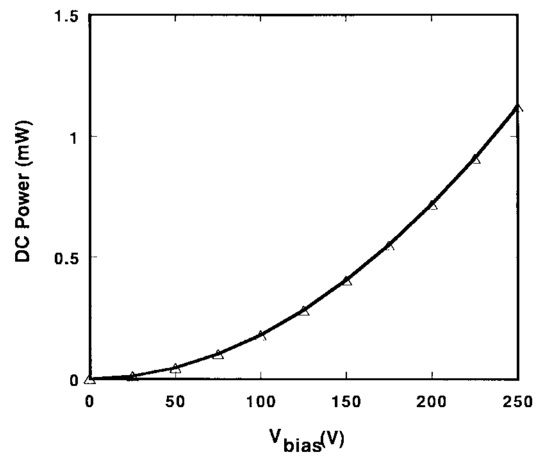

Fig. 15. Dc power requirement versus bias field.

V. Power Requirements and Power-Handling Consideration

Three main considerations need to be made regarding the power requirements of the phase shifter:

1) dc power required to maintain a fixed phase shift;
2) instantaneous power required to change the phase shift from one condition to a new condition;
3) maximum microwave power which can flow in the device without altering its functionality, either due to the occurrence of a nonlinear phenomenon or due to heat dissipation.

The dc resistivity (ρ) measurements of $Ba_{0.8}Sr_{0.2}TiO_3$ give a value of $\rho = 4.3\,10^7$ $\Omega\cdot$ m. Since the cross-sectional area for bias current flow in this sample is $5.03 \cdot 10^{-5}$ m^2, the corresponding dc resistance of the sample is $8.6 \cdot 10^7$ Ω. The dc voltage applied to the phase shifter varies from zero to 250 V, so the maximum power required will be $0.7211 \cdot 10^{-3}$ W. The above calculation shows that in order to maintain a fixed phase, the dc power requirements (less than 1 mW) are extremely small. Therefore, the bias power dissipated and, hence, the heating effects on the characteristics of the FEM are negligible. Results of power measurement on the sample are shown in Fig. 15. The actual total power is slightly higher than the theoretical value, probably due to other current leakage, e.g., in the circuit of the voltmeter or other measurement instruments.

Although it is not possible to give a quantitative estimate of the amount of driver power dissipated in the ferroelectric material since a knowledge of the large signal properties of FEM's is currently not available, some considerations are possible: the phase shifter as viewed from the terminals of the driving source is essentially a nonlinear parallel plate capacitor. It can be shown by an idealized calculation [6] that the energy transferred in changing the voltage from 0 to 250 V, thereby changing phase by 160°, is about 0.003 J. The instantaneous current requirements are a function of the speed with which the phase shifting is performed. Although the peak power capability of the driver is high, it is at least one order of magnitude smaller than that required for comparable frequency and speed ferrite phase shifters. One of the advantages of these phase shifters is that very little energy supplied by the driver to effect a rapid change in phase is dissipated in the FEM. The bulk of the energy transferred is used to change the energy stored in the electrostatic field of the capacitor or mechanically stored in the ceramic (these FEM's are also piezoelectric) [2]. The power-handling capability of the phase shifter would be limited by the following factors:

1) RF voltage level which can be sustained without breakdown;
2) tolerable temperature rise due to the attenuation of the transmitted RF power;
3) material response to large amplitude RF signal.

In general, in a microwave system employing microstrip lines, the coaxial connectors rather than the microstrip set the ultimate limit to peak power. In this case, since the substrate is very thin, one needs to consider both possibilities to obtain the lower breakdown voltage. The substrate used has a thickness of 0.1 mm, and a maximum breakdown voltage for a pulsed signal of 8.5 MV/m. This corresponds to a maximum tolerable voltage of 850 V. The coaxial connector sets a limit because of air breakdown, (the breakdown electric-field strength for dry air is 3 MV/m and the internal difference in the radii of a 50-Ω connector is approximately 1.5 mm); thus, the breakdown voltage is 4500 V, and the lower limit is determined by the substrate, and corresponds to 850 V. With any transmission line having characteristic impedance Z_0 and maximum breakdown voltage $V_{\rm bk}$, the peak power allowable is given by

$$P_{\max} = \frac{V_{\rm bk}^2}{2Z_0}. \tag{1}$$

In this case, the characteristic impedance of the microstrip line is below 10 Ω, so the maximum corresponding transmitted power is about 36 kW. Although this is just a theoretical value, it shows that the breakdown voltage is not a limiting factor, even for a thickness of 0.1 mm.

The temperature rise due to conductor and dielectric losses is well treated in [7]. The expression for the temperature rise above ambient is

$$\Delta T = \frac{0.2303h}{K}\left[\frac{\alpha_c}{w_{\rm eff}} + \frac{\alpha_d}{2w_{\rm eff}(f)}\right] \quad {}^\circ\text{C/W} \tag{2}$$

where α_c and α_d are the conductor and dielectric losses, respectively, (in decibels per meter), $w_{\rm eff}$ and $w_{\rm eff}(f)$ are the effective microstrip widths, and K is the thermal conductivity of the substrate. If one considers a rise θ above ambient, a microstrip line could carry a maximum average power given by

$$P_{\max} = \frac{\theta}{\Delta T}. \tag{3}$$

For this sample, one estimates $\Delta T = 0.04$ °C/W and the characteristic impedance is below 10 Ω. If one considers a rise in the temperature of $\theta = 10^\circ$ above ambient, the total power handled ranges up to 1 kW.

The effects of a large amplitude RF voltage on the properties of FEM's are largely unknown, and would have to be determined through experiment. Although only limited information is available, it is likely that large RF fields will cause nonlinear behavior, e.g., harmonic generation and mixing.

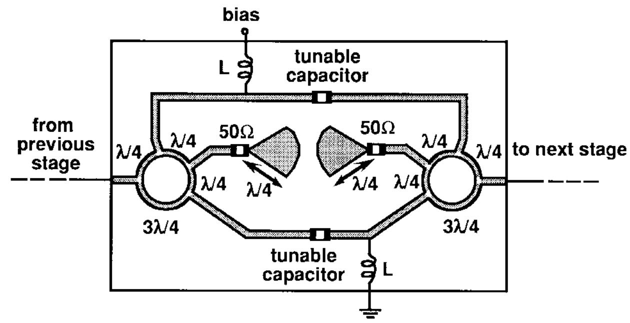

Fig. 16. Possible microwave network to combine the phase shift obtained by single capacitors.

In conclusion, the real limiting factors to power-handling capability are the heat dissipation and the manifestation of nonlinear effects. No particular attention is focused to these phenomena for powers below 500 W.

VI. Recommendation for the Design of Large Phase Shifters

The phase shifter tested in this paper is capable of a phase shift of 160° under the highest bias condition. In some applications like radar systems, higher phase shift may be required (up to 360°). Some attention to efficiently achieve this performance is given in this section. As a guideline, if a total phase shift higher than 120° is desired, use of multiple-stage phase shifters is recommended. In this manner, one reduces global insertion losses, makes the design of the matching circuit easier, and limits the maximum bias voltage. Clearly, each stage of the phase shifter must be properly isolated using a microwave network. As an example, a two-stage phase shifter using tunable capacitors (each having phase shift capability of 7.5°) as shifting elements is designed. The concept can be applied to microstrip transmission lines as well. The problem in the use of FEM tunable capacitors as shifting elements is the fact that they will not be capable of producing a phase shift larger than 90°. In this case, the capacitor will operate as an open circuit and no transmission will occur. Thus, a clear limit in capacitor-usable phase shift is about 30°. The combination of two capacitors in a two rat-race microwave network is shown in Fig. 16. Capacitors are connected to the isolated arms of two rat-race devices which add the total signal (and the phase shift) at the output stage. Based on a similar concept, an isolated power splitter or circulators can be used instead of the rat-race. Theoretical results simulated using HP microwave design software (MDS) are shown in Table III. Values are obtained for a capacitance change of about 30% of its nominal value, measured for BTO capacitors at 2.4 GHz [8]. The results reported for the S_{11}- and S_{21}-parameters are the worst case between the two possible bias conditions. Clearly, the advantage of the combined circulators, power splitter, and rat-race over the two series capacitor is observed.

Another possible solution in order to increase the total phase shift by a factor of two, consists of the use of an impedance transformer with a circulator, as shown in Fig. 17. The impedance transformer (a 1/4 transmission line, for ex-

TABLE III
COMPARISON BETWEEN DIFFERENT TOPOLOGIES TO COMBINE TWO PHASE SHIFTERS

Configuration	*S11 (dB)*	*S21 (dB)*	*Δ Phase(S21) (deg)*
One capacitor	-6.31	-1.16	7.84
Two capacitors	-2.26	-3.91	12.4
One rat-race	-9.16	-0.59	7.56
Two rat-race	-18.66	-0.117	14.69
One pow. split.	-8.38	-0.683	7.71
Two pow. split.	–10.1	-0.451	12.32

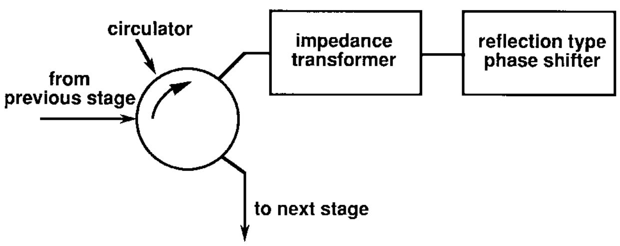

Fig. 17. Reflection-based phase shifter.

ample) must be designed such that the maximum advantage in terms of phase shift can be achieved for a minimum change of the tunable device.

VII. CONCLUSION

A novel sol–gel technique for the production of low-loss FEM for microwave applications has been introduced. New design for planar phase-shifter circuits has been presented and a design methodology to obtain a low-loss broad-band phase shifter operating at 2.43 GHz has been discussed and implemented. Measured results show net improvement over existing ferroelectric phase shifters, in terms of reduction of required bias voltage, broad-band capability, and reduction of loss. Use of planar structure devices allows the integration of this new type of phase shifter with conventional microwave circuits. Recommendations for the implementation of a multiple-stage phase shifter are also given.

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Franco De Flaviis was born in Teramo, Italy, in 1963. He received the degree in electronics engineering from the University of Ancona, Ancona, Italy, in 1990. He received the M.S. degree in electrical engineering from the Department of Electrical Engineering, at the University of California, Los Angeles (UCLA), in 1994. Since 1993 he has been working toward the Ph.D. degree in the same department.

In 1991 he worked at Alcatel as a Researcher specializing in the area of microwave-mixer design. In 1992 he was a Visiting Researcher at UCLA, working on low intermodulation mixers. He is presently working in the field of numerical techniques for the analysis and design of 3-D microwave structures, with particular emphasis on time-domain approach techniques. His research has compassed both theoretical and experimental studies of microwave mixers and circuits, and the synthesis of low-loss ferroelectric material for phase-shifter design to be employed in scan-beam antennas systems.

N. G. Alexopoulos (S'68–M'69–SM'82–F'87), photograph and biography not available at the time of publication.

Oscar M. Stafsudd received the M.S. and Ph.D. degrees in physics from the University of California, Los Angeles (UCLA), in 1961 and 1967, respectively.

In 1960, he was a member of the staff of the Standards and Research Department of Atomics International (later, the Rockwell Science Center), where he specialized in optical properties of materials and crystal growth. In 1964, he joined the research staff of the Hughes Research Laboratories, Malibu, CA, where he continued research in crystal growth, optical and properties of materials and laser physics. In 1967, he joined the Department of Electrical Engineering, UCLA, as an Assistant Professor and continues working there as a full Professor. His research interests are in quantum and solid-state electronics.