MICROWAVE FERROELECTRIC DEVICES

Control of reactance in microwave circuits, devices, and systems is a common method by which the response of a microwave circuit such as a filter, resonator, or phase shifter can be tuned. Devices based on a class of voltage-dependent nonlinear dielectrics known as ferroelectrics provide an alternative to semiconductor varactor diodes and ferrimagnetic components which are the most common devices of this type.

After an overview of the relevant materials issues, this article describes microwave devices that exploit the variation of the ferroelectric's permittivity with applied dc electric field. The main feature of these "tunable" microwave devices is the change of their capacitance, impedance, or phase velocity. We will describe varactors, oscillators, tunable filters, and phase shifting devices.

FERROELECTRIC MATERIALS

Even though they do not contain iron, the name ferroelectric was selected because they possess a response to an electric field that, although not the dual, is analogous to a ferromagnetic material's response to a magnetic field. Ferroelectrics are a subgroup of nonlinear dielectrics. The complex permittivity of a ferroelectric material is a function of both the temperature and an applied dc electric field.

Ferroelectric materials possess spontaneous polarization below a temperature referred to as the Curie temperature, at which point they undergo a phase transition. Above the Curie temperature, they are in a paraelectric state where spontaneous polarization disappears, but they still retain a nonlinear dielectric constant with applied electric field. In this article, we will use the term ferroelectric to describe these materials even if they are being operated at temperatures where they are in the paraelectric phase. Unlike ferromagnetic materials, ferroelectric materials in either the ferroelectric or paraelectric state are reciprocal; that is, the transmission coefficient through these devices is the same for different directions of propagation. Until recently, the dielectric losses in ferroelectric materials excluded them from being used at microwave frequencies. However, the continued improvement of ferroelectric materials suitable for use at microwave frequencies has resulted in the design of many microwave devices.

Tunability can be defined for a ferroelectric as the fractional change in the dielectric constant with applied dc bias voltage or

$$\text{Tunability} = \frac{\epsilon_{\text{r,max}} - \epsilon_{\text{r,min}}}{\epsilon_{\text{r,max}}} \tag{1}$$

where $\epsilon_{\text{r,max}}$ is the dielectric constant when no bias voltage is applied, and $\epsilon_{\text{r,min}}$ is the dielectric constant when maximum dc bias is applied. The dielectric constant of a ferroelectric decreases as the bias voltage is increased. Although larger tunability is a desirable feature for most microwave applications, a ferroelectric material with larger tunability usually has a relatively larger dielectric loss. Optimizing material tunability and loss to meet the needs of a particular microwave application remains a challenging task.

Ferroelectrics are inherently broadband. That is, they do not have a low frequency limit like ferrites. Switching time for these materials has been measured to be less than a nanosecond (1), which is sufficient for most microwave applications. Also, these materials are radiation-hardened. Ferroelectrics can be manufactured in bulk, thick-film, and thin-film form.

Like other ceramics, bulk ferroelectric ceramics can handle high peak powers. The limit on the average power is determined by the loss tangent (tan δ) of the ferroelectric. There are many known ferroelectrics, but the most widely used ferroelectric at microwave frequencies is barium strontium titanate, $Ba_{1-x}Sr_xTiO_3$ (BSTO). BSTO with $x = 0.5$ is frequently used for microwave applications since the Curie temperature is well below room temperature yet a reasonable tunability is retained. Bulk ceramics of this composition typically possess relative dielectric constants on the order of 1000 and loss tangents of 0.02 at 10 GHz (2,3). BSTO's Curie temperature can be controlled by varying the barium to strontium ratio. A bulk composite material can be engineered by adding nonferroelectric oxides to BSTO to reduce the dielectric constant and the loss tangent (2,3). For room-temperature operation of these composites in the paraelectric phase, tunability, ϵ_r, and tan δ decrease with decreasing barium content; they also decrease with increasing oxide content. Tunability increases linearly with an increase in bias voltage. Bulk ceramics can be produced using usual ceramic processing techniques.

Thin films are compatible with integrated circuits, and they need lower bias voltages than does bulk material. Thin films can be manufactured by any of the common thin-film deposition techniques, pulsed-laser deposition, sputtering, metal organic chemical vapor deposition (MOCVD), and so on. In thin films, control of the ferroelectric composition and incorporation of doping is also possible to reduce losses at microwave frequencies (4). Thin films can potentially be less costly and easier to manufacture, but they cannot handle high power levels. Thin films of the ferroelectric strontium titanate, $SrTiO_3$ (STO), are used at microwave frequencies because of their compatibility with the high-temperature superconductor (HTS) yttrium barium cuprate (YBCO).

Between bulk and thin-film ferroelectrics lies the realm of thick-film ferroelectrics, which can be produced via tape casting.

Throughout this article, it will be assumed that the ferroelectric is homogeneous and that it is linear with respect to a small, time-varying electric field. The dielectric strength of these ferroelectrics is relatively high. Large dc electric fields (in the range of a few MV/m) can be applied to STO and BSTO before dielectric breakdown occurs. Assuming that a particular ferroelectric composition meets the tunability, ϵ_r and tan δ requirements of the application, the next two sections describe the various issues that need to be addressed before designing a microwave ferroelectric device.

MICROWAVE DEVICE CONSIDERATIONS

Most microwave devices can be categorized according to their physical size in relationship to the wavelength at their upper frequency of operation. Those devices that are very small compared to their operational wavelength are called "electrically small" and can be modeled with discrete circuit components. The term "lumped-element" model is often employed. For devices whose dimensions are larger, it is usually necessary to take into account the frequency-dependent effects. Such devices are described by distributed networks of common circuit elements such as inductors and capacitors. The most common distributed device is the transmission line which is modeled by a ladder network of series inductors and shunt capacitors. As a lumped element, a capacitor which uses ferroelectric material yields a tunable circuit reactance.

Transmission lines can usually be described by their phase velocity and characteristic impedance, which are given by

$$v_\text{p} = \frac{1}{\sqrt{LC}} \tag{2}$$

and

$$Z_\text{c} = \sqrt{\frac{L}{C}} \tag{3}$$

respectively, where L is the inductance per unit length and C is the capacitance per unit length. By introducing a material whose dielectric constant, ϵ_r, is controlled (or tuned) by a direct-current (dc) bias voltage, the phase velocity and characteristic impedance can be varied by changing capacitance.

An issue to consider when choosing a microwave device topology is the power-handling requirements. For high-power applications, the number of device topologies that are appropriate is limited. Further discussion is offered in the next section.

In any nonlinear material, device, or system, another important practical consideration is the strength of signals generated at other than the desired frequency. Since most systems are bandwidth-limited, the most troublesome condition arises when two desired signals, f_1 and f_2, both within the passband produce signal at frequencies $2f_1 - f_2$ and $2f_2 - f_1$, also within the passband. A plot of the signal strengths of f_1, f_2, $2f_1 - f_2$, and $2f_2 - f_1$ is often used to determine the third-order intercept point (IP3), which is an important figure of merit.

Since ferroelectrics have a high dielectric constant, the circuits that employ these materials tend to have very low impedance. Therefore, impedance matching is also another major issue to be addressed when using ferroelectrics. A consequence of a voltage-dependent capacitance being utilized to tune the phase velocity of a transmission line as given by Eq. (2) is that the characteristic impedance of the transmission line is also tuned per Eq. (3). This further complicates the impedance matching problem.

MICROWAVE GUIDING STRUCTURES

At microwave frequencies, ferroelectrics can be introduced into many different types of rectilinear structures that are used to guide electromagnetic waves. These guiding structures include parallel plate and rectangular waveguides, which can be loaded (or filled) with ferroelectric material. There are also many planar structures that use the ferroelectric material as a tunable substrate, like microstrip, slotline, coplanar strip, and coplanar waveguide. Each structure has a different set of advantages and disadvantages. For applications where ferroelectrics are used to provide bias-dependent propagation properties, it is convenient to divide the guiding structures into two categories: (1) geometries that can handle high microwave power but require large bias voltages (parallel plate and rectangular waveguide) and (2) those which are compatible with small microwave power levels and require only modest bias voltages (planar structures). Note that the ferroelectric permittivity is a function of the electric field. In a planar structure, the bias voltage is applied across a thinner ferroelectric, and so smaller bias voltage will produce the same variation of the permittivity as a larger bias voltage (which creates a similar electric field intensity) would produce in a parallel plate or rectangular waveguide.

Planar waveguiding structures contain the metallization defining the waveguiding structure delineated on one or more plane. Often this metallization layer is on the top surface of a dielectric substrate. Hence, these geometries are compatible with photolithographic processing. Since the metallization delineating the waveguide is at the interface of two regions (usually dielectric), the guided wave is propagating such that a portion of the field is in each region. The choice of the correct planar transmission line is determined by many factors including (1) orientation of the bias field and microwave field with the ferroelectric region, (2) thickness of the ferroelectric material, and (3) compatibility with other circuit elements. Planar structures are compatible with ferroelectric thin films and with semiconductors for microwave monolithic integrated circuits. However, planar structures require a dc block to isolate the radio frequency (RF) from the high dc voltage that is used to tune the ferroelectric permittivity (5).

Microstrip

The most common planar transmission line is microstrip. As shown in Fig. 1, both the bias and the dominant mode microwave electric field are oriented primarily normal to the interface. It should be noted that although microstrip is the most widely used planar transmission line, when using thin-film ferroelectrics deposited on a substrate with the delineated metal layer on top as shown in the Fig. 1(b), the high dielectric constant of the ferroelectric results in decreased tuning efficiency. This can be seen by considering the capacitance per unit length to be a series combination of the ferroelectric capacitor and the substrate capacitor. The capacitance contribution from the thin-film ferroelectric is much smaller than that from the substrate, and the tunability of the phase velocity and the characteristic impedance of the dominant mode are reduced accordingly.

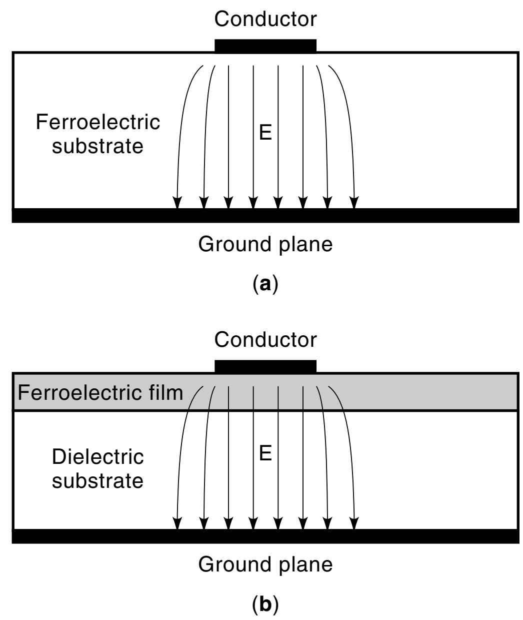

Figure 1. Microstrip planar transmission line cross section with (a) a homegeneous bulk substrate and (b) a thin film on a bulk substrate. The electric field between the metal strip on the top surface and the metal ground plane is primarily normal to the surface.

Tunable filters employing parallel-coupled microstrip resonators do not suffer from this inefficiency since the coupled-line mode (6) possesses a significant electric field component parallel to the surface. This coupled-line mode is similar to the coplanar strip transmission line (discussed below) with a ground plane. Practical design equations for microstrip on layered dielectric substrates are based on a quasi-static analysis (7).

Coplanar Waveguide, Coplanar Strip, and Slotline

Other planar transmission lines such as coplanar waveguide, coplanar strip, and slotline have two or more conductors on the patterned surface. Hence the electric fields of both the dominant microwave mode and the bias are tangential to the substrate surface as shown in Fig. 2. The dominant mode of these planar transmission lines can be efficiently tuned with a bias field whether a bulk ferroelectric substrate or a thin-film ferroelectric on bulk dielectric substrate is employed. In the later case, good tunability is retained since the thin-film capacitance and substrate capacitance are in parallel. Design equations for a coplanar waveguide on multilayered dielectric

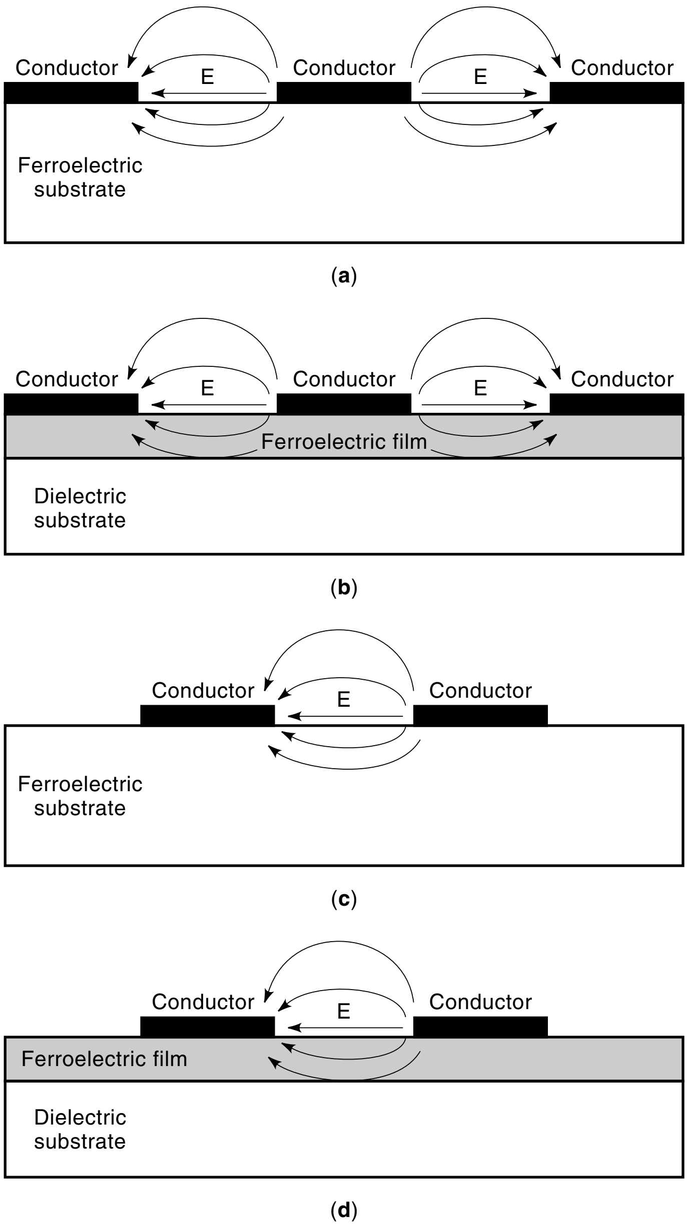

Figure 2. Coplanar waveguide (CPW) transmission line cross section with (a) a homegeneous bulk substrate and (b) a thin film on a bulk substrate and coplanar strip (CPS) transmission line cross section with (c) a homegeneous bulk substrate and (d) a thin film on a bulk substrate. The electric field between the metal strips on the top surface is primarily parallel to the surface.

substrates are available (7). Similar analyses using a partial-capacitance conformal-mapping approach can be applied to other geometries to account for the ferroelectric thin film.

For the coplanar waveguide, the narrower the gaps between the center conductor and the ground planes, the higher the electric field intensity (and tunability) for a given bias voltage. Although coplanar waveguide is one of the simplest transmission lines, the microwave current density is sharply peaked at the edges of the strips causing large conductor losses. The problem is enhanced by the high dielectric constant and small thickness of the ferroelectric (8).

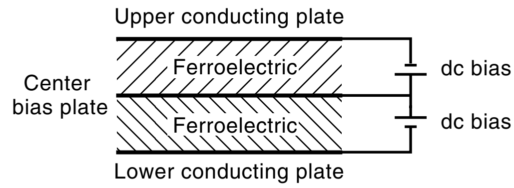

Figure 3. Parallel plate waveguide phase shifter (propagation into or out of paper).

Parallel Plate and Rectangular Waveguide

A parallel-plate waveguide is a two-conductor guiding structure that supports transverse electromagnetic (TEM) waves. Thus, the electric and magnetic field are orthogonal to each other and to the direction of propagation. Figure 3 shows how this type of waveguide can be loaded with a ferroelectric medium to provide a variable phase velocity, which is given by

$$\nu_{\mathrm{p}} = \frac{1}{\sqrt{\mu\epsilon}} \tag{4}$$

where μ and ϵ are the permeability and permittivity of the ferroelectric. Both the dc and the RF electric field are vertical. The ferroelectric is bifurcated with an electrode that is used to apply the dc bias with respect to the grounded waveguide walls.

Rectangular waveguides are popular in the microwave region. They are single-conductor guiding structures that confine the electromagnetic wave in the interior of the waveguide. Typically, the waveguide is operated in the dominant TE_{10} mode. Figure 4 shows how a ferroelectric can be used in this type of waveguide to provide a variable phase velocity, which is given by

$$\nu_{\mathrm{p}} = \frac{1}{\sqrt{\mu\epsilon}\sqrt{1-\left(\frac{\lambda}{\lambda_{\mathrm{c}}}\right)^2}} \tag{5}$$

where λ_c is the cutoff wavelength. Again, both dc and RF electric field are vertical, and the ferroelectric is bifurcated by an electrode. Unlike the parallel plate waveguide (which has no sidewalls), rectangular waveguide has sidewalls. Therefore, a slot needs to be cut into a sidewall to connect the electrode to a dc power supply. The area of the slot opening must be small to prevent the microwave energy from leaking out of the slot.

APPLICATIONS

In this section, we will describe several applications of ferroelectrics at microwave frequencies. These include varactors,

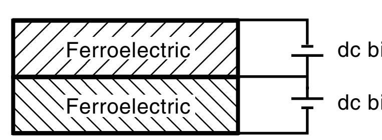

Figure 4. Rectangular waveguide phase shifter (propagation into or out of paper).

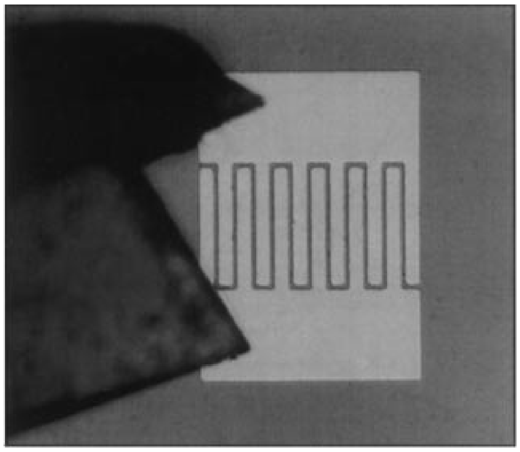

Figure 5. Photograph of a typical interdigitated capacitor on a thin-film ferroelectric covered substrate. The gap between fingers in the metal electrodes is 6 μm. A microwave probe is shown contacting the device from the left.

voltage-controlled oscillators (VCOs), tunable filters, and phase shifting devices.

Varactor

Varactors are variable-reactance circuit elements. They are used in switching or modulation of a microwave signal, for the generation of harmonics in an applied microwave signal, and in the mixing of two microwave signals of different frequency. As a discrete tunable capacitor, ferroelectric-based capacitors are applicable in a number of microwave circuits including VCOs, tunable filters, and oscillators. Parallel-plate configurations have not been successfully implemented (due to high required processing temperatures) in producing a high-quality ferroelectric thin film on a low-surface-resistance metal. Interdigitated capacitors (9), where the metal electrodes are deposited on top of the ferroelectric (either thin-film or thick-film) or bulk substrate, have proven to be a more practical option. A typical interdigitated capacitor is shown in Fig. 5. Although strongly dependent on the ferroelectric material involved, Figs. 6 and 7 show the level of performance

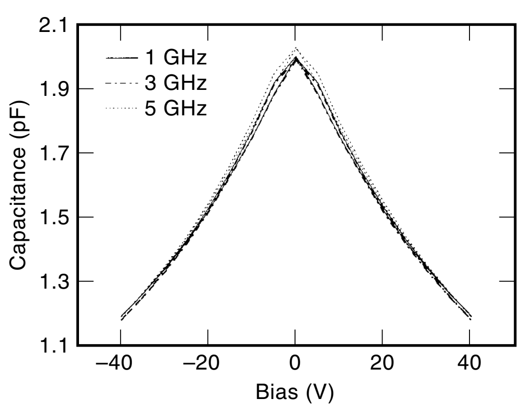

Figure 6. Capacitance versus bias voltage of a $Sr_{0.5}Ba_{0.5}TiO_3$ thin-film interdigitated capacitor on an MgO substrate for frequency values of 1, 3, and 5 GHz. The data represents bias swept from −40 V to 40 V and back to −40 V.

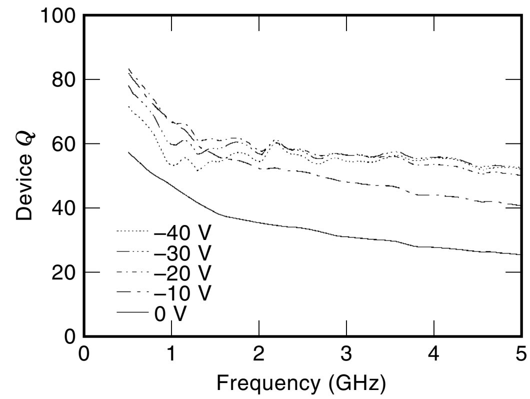

Figure 7. Interdigitated capacitor quality factor as a function of frequency for a $Sr_{0.5}Ba_{0.5}TiO_3$ thin-film on an MgO substrate with bias voltages of 0 V, −10 V, −20 V, −30 V, and −40 V.

available from thin-film ferroelectric interdigitated capacitor technology. Losses in ferroelectric interdigitated capacitors arise from the losses in the ferroelectric material as denoted by the dielectric loss tangent and from resistive losses in the metal electrodes. Using a low-surface-resistance metal such as silver or (in some cases) superconductors minimizes the electrode loss component, and in most cases the unloaded quality factor of the ferroelectric varactor is given by

$$Q_{\mathrm{U}} = \frac{1}{\tan\delta} \tag{6}$$

Design and modeling of layered interdigitated lumped element capacitors is based on a conformal mapping approach (10).

Voltage-Controlled Oscillator

Resonant circuits are used in oscillators and tunable filters. An oscillator provides a sinusoidal signal, and it is used as a source of microwave energy. The oscillator output should be clean (noiseless) and stable (frequency and power should not change with time). A high-Q resonator is used in an oscillator circuit to obtain good frequency stability and low noise. The tunable capacitance of a ferroelectric-based capacitor is particularly applicable to a class of devices called a voltage-controlled oscillator (VCO). The bias-controlled change in reactance varies the oscillation frequency of an active element such as a transistor. Although there are many different oscillator topologies the designer can choose from, to first order the oscillation frequency can be varied in proportion to the square root of the bias-dependent capacitance. In principle, there is no difference in the design of a VCO using a ferroelectric capacitor and a semiconductor varactor diode (11). A VCO employing a ferroelectric tunable ring resonator has demonstrated 3% frequency tunability at 17 GHz (12). Phase noise is one of the primary limitations of any VCO application. Although several mechanisms contribute to phase noise, in many cases the Q factor of the tunable element is the limiting factor. As can be seen from Leeson's formula, we have (11)

$$\mathscr{L}(f_{\mathrm{m}}) = \frac{1}{2}\left[1 + \frac{1}{f_{\mathrm{m}}^2}\left(\frac{f}{2Q_{\mathrm{L}}}\right)^2\right]\frac{FkT}{P_{\mathrm{avs}}}\left(1 + \frac{f_{\mathrm{c}}}{f_{\mathrm{m}}}\right) \quad \text{(dBc/Hz)} \tag{7}$$

where f_m is the offset frequency, f is the oscillation frequency P_{avs} is the power level, F is the noise figure, f_c is the $1/f$ noise corner frequency, and Q_L is the loaded quality factor which is related to the unloaded device quality factor Q_U by

$$Q_U = \frac{1}{\dfrac{1}{Q_L} - \dfrac{1}{Q_{ext}}} \tag{8}$$

where Q_{EXT} is the external quality factor. As can be seen in Leeson's formula, the quality factor of the variable capacitor, which is, to first order, the reciprocal of the dielectric loss tangent, has a major impact on the phase noise of the VCO.

Tunable Filter

A filter is any device or circuit which exhibits frequency selectivity; that is, the amplitude and phase of the output signal are functions of frequency. A simple example is a bandpass filter where, ideally, all frequencies in a certain range are passed without change to the signal, whereas at any other frequency no signal appears at the output. There are many different ways to realize a microwave filter (13). Many of these rely on resonators which are coupled together in a carefully controlled fashion to realize the desired filter transfer function. Tunability of the filter transfer function can be acheived with ferroelectrics (14,15). It has been demonstrated that the center frequency of a microwave filter can be tuned by approximately 10% using ferroelectrics. In practice, there are many filter topologies that lend themselves to tuning with ferroelectrics. Conceptually the simplest to envision is tuning the center frequency of a bandpass filter composed of coupled half-wavelength resonators by varying the phase velocity of the resonant elements and hence their resonant frequency. From Eq. (2) it can be seen that the phase velocity is inversely proportional to the square root of the capacitance. Since many filter topologies rely on capacitive coupling of resonators, utilizing tunable coupling between resonators allows the design of tunable bandwidth filters.

Phased Array Antenna

Phased array antennas can steer transmitted and received signals without mechanically rotating the antenna. Each radiating element of a phased array is normally connected to a phase shifter and a driver, which determines the phase of the signal at each element to form a beam at the desired angle. The most commonly used phase shifters are ferrite and diode phase shifters. Ferrite phase shifters are preferred at microwave frequencies, but they are expensive. The cost of a phased array mainly depends on the cost of phase shifters and drivers, and thus lower-cost phase shifting devices need to be developed to make the phased array antenna affordable for more applications. In this section, three different applications of ferroelectrics to phased array antennas will be described.

Ferroelectric Lens Antenna. The cost of a phased array depends mainly on the cost of phase shifters and drivers. A typical array may have several thousand elements as well as several thousand phase shifters and drivers; hence, it is very expensive. Therefore, reducing the cost and complexity of the phase shifters, drivers, and controls is an important consideration in the design of phased arrays. The ferroelectric lens phased array uniquely incorporates bulk phase shifting (2,3,16,17); the array does not contain individual phase shifters but rather uses ferroelectric material. This will reduce the number of phase shifters from $(n \times m)$ to $(n + m)$, where n is the number of columns and m is the number of rows in a phased array. The number of phase shifter drivers and phase shifter controls is also significantly reduced by using row–column beam steering. The ferroelectric lens has the advantages of small lens thickness, high power-handling capability, and simple beam-steering controls, and it uses very low power to control the phase shift. Thus, it leads to low-cost phased arrays. However, it should be noted that the use of row–column steering may limit the level of side lobes that can be achieved.

Description of Ferroelectric Lens and Its Operation. The ferroelectric lens is shown in Fig. 8; each column of the lens is a

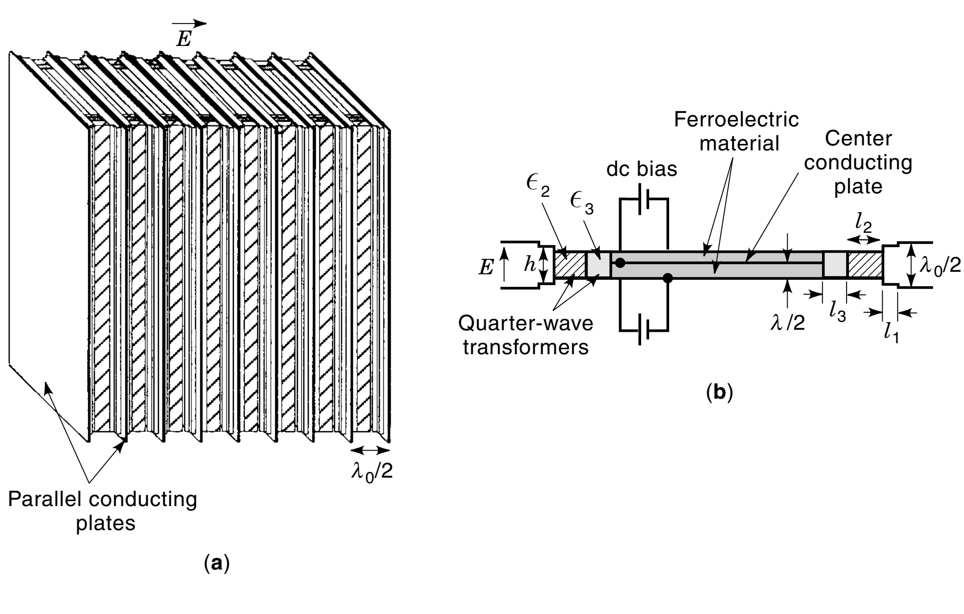

Figure 8. Ferroelectric lens.

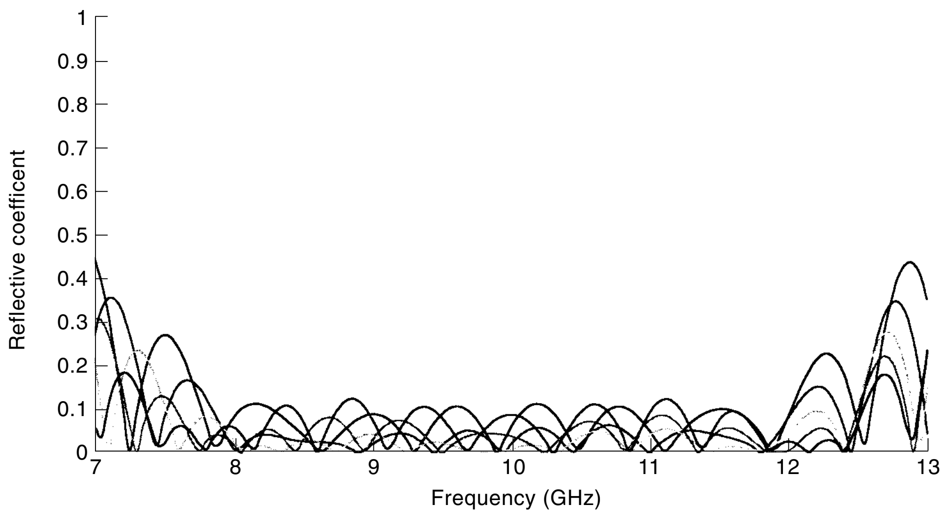

Figure 9. Theoretical reflection coefficient for ϵ_r to 80 to 120.

set of conducting parallel plates that are loaded with bulk ferroelectric material. The material is bifurcated by a center conducting plate that is used to apply the dc bias voltage to the ferroelectric. The separation between the parallel plates at the input and output end is $\lambda_0/2$, where λ_0 is the free space wavelength. Since only the TEM mode is desired, the separation between the parallel conducting plates is reduced to avoid higher-order mode propagation in the dielectric loaded section of the waveguide. Specifically, the separation between the center bias plate and either conducting plate is less than $\lambda/2$, where λ is the wavelength in the ferroelectric. Quarter-wave dielectric impedance transformers are used to match the empty waveguide to the ferroelectric loaded waveguide.

For scanning applications, a phase shifting device must provide 360° differential phase shift. The amount of the ferroelectric material needed (in the direction of propagation) to obtain 360° differential phase shift is (16)

$$t = \frac{\lambda_0}{\sqrt{\epsilon_{\mathrm{r,max}}} - \sqrt{\epsilon_{\mathrm{r,min}}}} = \frac{\lambda_0}{\sqrt{\epsilon_{\mathrm{r,max}}}\,[1 - \sqrt{1 - \mathrm{tunability}}]} \tag{9}$$

where $\epsilon_{\mathrm{r,max}}$ is the dielectric constant when no bias voltage is applied, and $\epsilon_{\mathrm{r,min}}$ is the dielectric constant when maximum dc bias is applied. Tunability is the fractional change in the dielectric constant as defined earlier. Thus, the thickness of the ferroelectric material needed is a function of the dielectric constant and the tunability of the ferroelectric, and the wavelength. Also, it can be shown that in order to obtain 360° phase shift, the dielectric loss through the ferroelectric is (16)

$$\alpha(\mathrm{dB}) = \frac{27.3 \tan\delta}{1 - \sqrt{1 - \mathrm{tunability}}} \tag{10}$$

It may be noted that the lens loss is independent of the ferroelectric permittivity and depends only on its loss tangent and tunability.

In general, the ferroelectrics with higher dielectric constant offer higher tunability, which is desired to reduce the lens thickness. However, matching the lens to free space is easier for smaller ϵ_r. Therefore, a compromise is needed between reducing the lens thickness (to reduce overall lens size) and achieving reasonable impedance match to reduce reflections from the lens surface. For a typical value of $\epsilon_r \sim 100$, it is possible to obtain a tunability of 20%, which results in a reasonable lens thickness of $\sim\lambda_0$ (e.g., 3 cm at 10 GHz). From Eq. (10), it can be seen that the tan δ must be less than 0.005 to limit the lens loss to less than 1 dB. The existing ferroelectric materials are a bit more lossy (tan δ = 0.008 at 10 GHz).

Phased Array Configurations Using Ferroelectric Lens for Two-Dimensional Scanning. The ferroelectric lens offers electronic scanning in one plane. The lens proposed here can be fed by a nonscanning planar array, like a slotted waveguide array. A combination of slotted waveguide array with phase shifters and the lens proposed here can be used as a phased array that can scan in two planes. A space feed can be used with the combination of two lenses proposed here (with a polarization

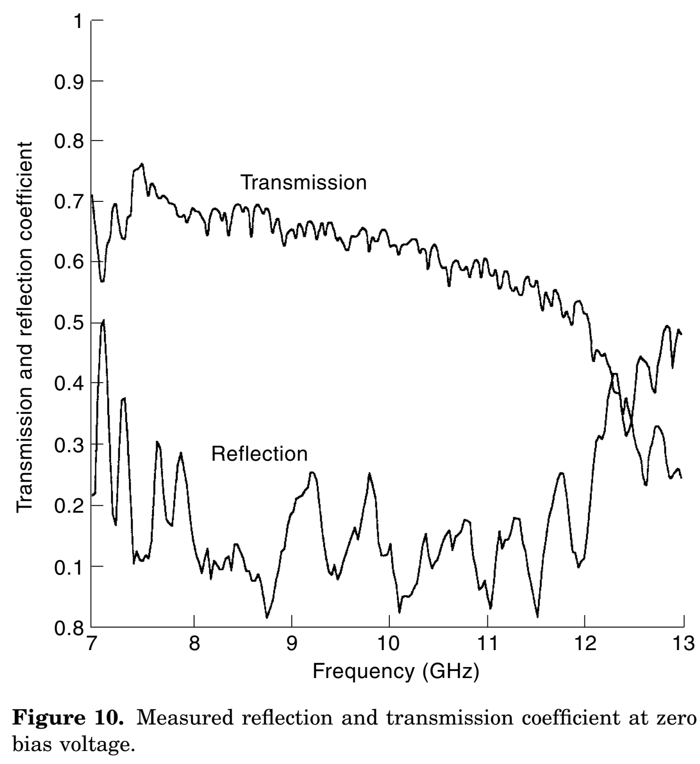

Figure 10. Measured reflection and transmission coefficient at zero bias voltage.

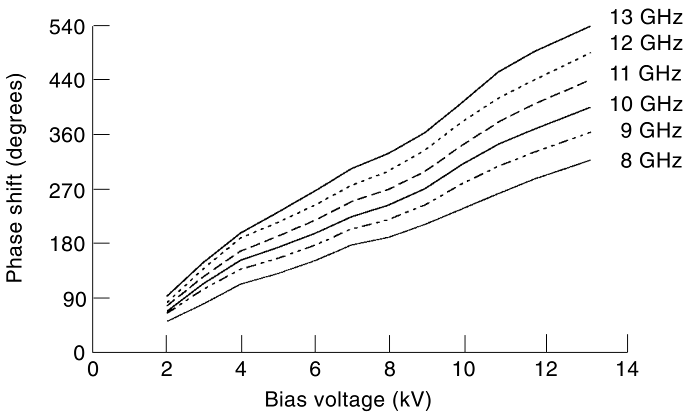

Figure 11. Measured differential phase shift.

rotator in between) to scan the beam in two planes. The details of these alternatives are discussed elsewhere (16,17).

Theoretical and Experimental Results. For the theoretical analysis of the ferroelectric lens, an individual section between two conducting parallel plates of the lens can be considered as one column of a phased array. The column can be analyzed as a two-dimensional (2-D) parallel-plate waveguide with electric field of the TEM mode normal to the plates as shown in Fig. 8. A matching network was designed using mode matching technique assuming that the dielectric constant of the ferroelectric varies from 120 to 80 (33% tunability) over a frequency range of 8 GHz to 12 GHz (40% bandwidth) and that $\lambda_0 = 0.8$ in. and $\lambda = 0.1$ in. The computed results are shown in Fig. 9. The matching network parameters are $l_1 = 0.2956$ in., $l_2 = 0.1860$ in., $l_3 = 0.0505$ in., $h = 0.2345$ in., $\epsilon_2 = 2.54$, and $\epsilon_3 = 35$ (see Fig. 8).

Experiments were performed with the ferroelectric composition $Ba_{0.55}Sr_{0.45}TiO_3$ with 60% oxide. This material offered a good compromise among ϵ_r, tan δ, and tunability. At 10 GHz, for this composition, $\epsilon_r = 100$ and tan $\delta = 0.0079$. The ferroelectrics were 1 in. long (in the direction of propagation), 0.05 in. high and 5 in. ($\sim 4\lambda_0$ at 10 GHz) wide. Figure 10 shows the measured transmission and reflection coefficients at zero bias. The reflection coefficient is sufficiently small over a wide frequency band as the theory had predicted in Fig. 9. Figure 10 also shows that the loss increases with frequency; this is due to two reasons. First, tan δ increases with frequency, which is expected for ceramics; second, the electrical length (in terms of wavelengths) of the ferroelectric in the direction of propagation increases with frequency since the physical length is kept constant (1 in.).

Figure 11 shows the measured phase shift as a function of the bias voltage for various frequencies. As expected, the phase shift increases linearly with frequency because the electrical length of the material increases with frequency. Since ferroelectrics are good insulators, the dc current requirements are very low. For example, at 10 kV bias voltage, the dc current drawn was 0.05 mA, and thus the dc power dissipated is only 0.5 W. The bias voltage can be reduced by further bifurcating the ferroelectrics (using interdigital electrodes).

Figure 12 shows the reflection coefficients as a function of frequency for various bias voltages. The standing wave ratio (SWR) is less than 2 for frequency range of 8 GHz to 12 GHz as theoretically predicted earlier (see Fig. 9).

Traveling Wave Antenna. Another type of phased array antenna that also uses bulk phase shifting is a traveling wave antenna, as shown in Fig. 13. The antenna is a slab of ferroelectric material with conducting strips on the top side of the slab and a ground plane on the bottom. This type of antenna is well-suited for millimeter-wave applications when a low-loss dielectric (not a ferroelectric) is used as a substrate, and frequency variation is used to scan the antenna beam electronically. Instead of changing the frequency, the dielectric constant of the ferroelectric substrate can be changed to electronically scan the antenna beam in the E plane (17–19). It can be shown (18) that the radiation angle of the antenna beam is given by

$$\theta = \sin^{-1}\lambda_0\left(\frac{1}{\lambda_g} - \frac{1}{d}\right) \tag{11}$$

where λ_0 and λ_g are the free space and guide wavelength, respectively, and d is the spacing between the conducting

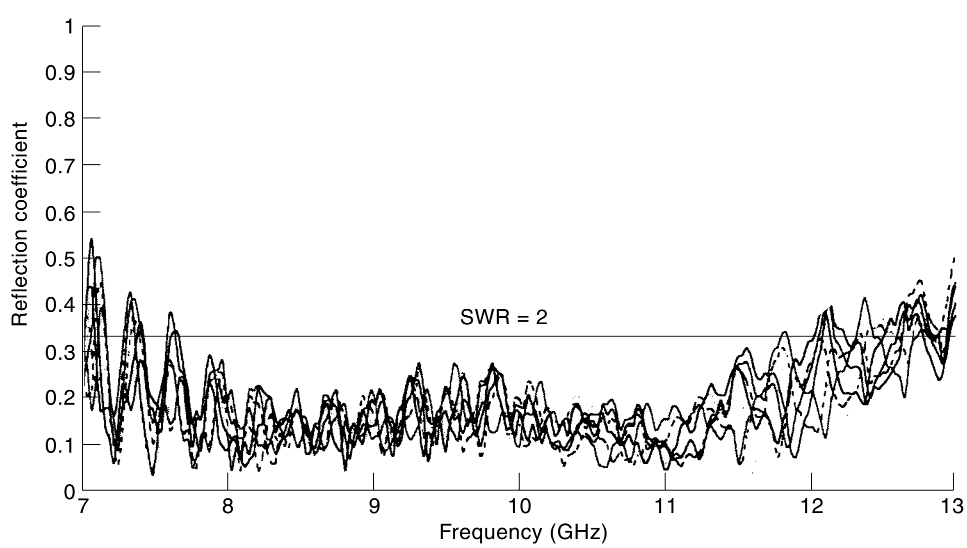

Figure 12. Measured reflection coefficient for various bias voltages (0 to 13.5 kV).

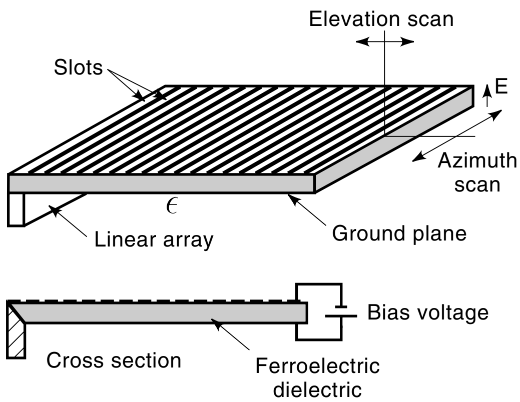

Figure 13. Ferroelectric traveling-wave antenna.

strips. The guide wavelength and thus the scan angle changes as the dielectric constant of the substrate changes. The main advantage of this type of antenna is that only a single dc power supply is needed to scan the beam, and it is a very simple structure. However, this antenna has a major disadvantage that makes it quite impractical. Since the physical size of most microwave antennas is at least a few wavelengths (if not a few tens of wavelengths), the loss that the electromagnetic wave would suffer as it travels down the antenna is enormous. Also, the instantaneous bandwidth of this antenna is very small because it is a frequency scan antenna. That is, the beam pointing direction changes as the frequency changes. Like the ferroelectric lens, the traveling wave antenna offers electronic beam scanning in one plane. Electronic scanning in the other plane (azimuth plane in Fig. 13) can be achieved with phase shifters in the linear array feed for this antenna.

Discrete Phase Shifter. Ferroelectrics have also been proposed for discrete phase shifter applications at microwave frequencies. There are several advantages of using ferroelectrics over ferrites in phase shifters. First, since ferroelectrics are voltage-driven devices, the dc control power requirements are small. However, unlike latching ferrite phase shifters that only require current pulses, the bias voltage needs to be applied to the ferroelectrics during the entire transmit and/or receive cycle. Second, ferroelectrics provide reciprocal phase shift. Third, the high dielectric constant of the ferroelectric has the effect of decreasing the overall size of the phase shifter. At the present time, however, the dielectric loss in ferroelectrics is higher than that in ferrites at microwave frequencies.

The basic design equations for a discrete phase shifter are the same as those for the ferroelectric lens. For the same electric field applied in the lens, the discrete phase shifter should provide similar phase shift using the same ferroelectric. Phase shifters have been designed using ferroelectric-loaded rectangular waveguides as well as planar transmission lines, like microstrip and coplanar waveguide, on a ferroelectric substrate (5,20,21).

For the rectangular waveguide, impedance matching techniques similar to the ones used in the ferroelectric lens can be applied. For the microstrip line, several impedance matching techniques have been tried (5) including quarterwave transformers, open circuit stubs and radial stubs. For the coplanar waveguide, usually the lines are tapered to provide a 50 Ω impedance (21).

High Temperature Superconductors and Ferroelectrics

The discovery of high temperature superconductors (HTS) has generated many tunable device designs (22) using both thin films and bulk ferroelectrics. One of the main incentives is the promise of the low conductor loss associated with HTS. In addition, both STO and BSTO are closely lattice matched with the HTS yttrium barium cuprate, $YBa_2Cu_3O_{7-\delta}$ (YBCO), meaning that the ferroelectric and YBCO can be epitaxially grown on top of each other to form multilayer thin film structures. This has been done with STO and YBCO. To take advantage of the low microwave losses, operation must be below the critical temperature for YBCO. Most of the research has been done at 77 K, the liquid nitrogen boiling temperature. In BSTO, the barium to strontium ratio can be adjusted so that BSTO can be operated in the paraelectric phase at 77 K, and STO remains paraelectric down to the lowest temperatures.

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MICROWAVE FILTERS. See DIELECTRIC RESONATOR FILTERS.