X-Band Phased Array Antenna Using Ferroelectric $(Ba,Sr)TiO_3$ Coplanar Waveguide Phase Shifter

Seung Eon Moon, Han Cheol Ryu, Min Hwan Kwak, Young Tae Kim, Su-Jae Lee, and Kwang-Yong Kang

A phased array antenna was fabricated using four-element ferroelectric phase shifters with a coplanar waveguide (CPW) transmission line structure based on a $Ba_{0.6}Sr_{0.4}TiO_3$(BST)/MgO structure. Epitaxial BST films were deposited on MgO (001) substrates by pulsed laser deposition. To attain the large differential phase shift and small losses for a ferroelectric CPW phase shifter, an impedance-matching-part adding technique between the effective transmission line and connecting cable was used. The return loss and insertion loss for this technique-adapted BST CPW device were improved with respect to those for a normal BST CPW device. For an X-band phased array antenna system consisting of ferroelectric BST CPW phase shifters, power divider, dc block, patch antenna, and programmed dc power, the steering beam could be tilted by 15° in either direction.

Keywords: Ferroelectric phase shifter, impedance matching, phased array antenna.

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Seung Eon Moon (phone: + 82 42 860 5603, email: semoon@etri.re.kr), Han Cheol Ryu (email: hcryoo@etri.re.kr), Min Hwan Kwak (email: mhkwak@etri.re.kr), Young Tae Kim (email: ytkim@etri.re.kr), Su-Jae Lee (email: leesujae@etri.re.kr), and Kwang-Yong Kang (email: kykang@etri.re.kr) are with the Basic Research Laboratory, ETRI, Daejeon, Korea.

I. Introduction

The core element of a phased array antenna system, the phase shifter, is to control the time variation of a signal wave. Superposing all the time variation of the phase shifters, the beam of an antenna can be tilted electronically or mechanically. This system has been studied for various fields, for example, wireless and satellite communications and military radar systems [1], [2].

Depending on the type of electronic control medium or mechanism adopted, these phase shifters can be categorized as ferrite, semiconductor, or ferroelectric phase shifters. For a ferromagnetic phase shifter, it is hard to control speedy beam scanning because of the slow response time, and a bulky element size and high power consumption are other weak points. For a semiconductor phase shifter, large loss in the high frequency range and a low transmission power capability are shortcoming points [3]-[5].

Due to its small size, simple process for making the device, low power consumption, low production cost, high transmission power capability, fast response time, and possibility for electric tuning because of the feasibility to change the dielectric constant, ferroelectric material is one of the promising candidates for a microwave tunable device, as well as in other application fields such as dynamic random access memory, non-volatile ferroelectric random access memory, piezoelectric devices, and pyroelectric devices. Moreover, analog or digital control is possible for a ferroelectric phase shifter due to the possibility of electric continuous tuning.

In the viewpoint of material for microwave ferroelectric tunable devices, a large dielectric constant variation and low dielectric loss characteristics are required, and many studies have been done [6]-[11]. Also, there are many designs for ferroelectric phase shifters [12]-[19]. Among them, the coplanar waveguide (CPW) transmission line has been studied for a long time and can be recommended because of both a simple structure and low production cost due to its simple device making process. However, in the ferroelectric CPW phase shifter, it is hard to get both an impedance matched design and a large differential phase shift under the dc bias variation because the wider gap design, which is generally more feasible in a 50 ohm matched ferroelectric CPW transmission line, is incompatible with a high electric field for large dielectric tunability. Therefore, a somewhat large return loss and an additional insertion loss may be expected for the ferroelectric CPW phase shifter because of the impedance mismatch.

In this paper, we report on the results for the improved microwave loss characteristics of a ferroelectric CPW phase shifter maintaining the phase tuning property. Using the power divider, bias tee, and microstrip patch antenna, the X-band 1 × 4 phased array antenna was assembled and the measured performance of the system reported.

II. Experiment

Epitaxial $(Ba_{0.6}Sr_{0.4})TiO_3$ (BST) films were deposited onto MgO substrates by a pulsed laser deposition method. A focused pulse laser from a Kr:F eximer gas laser (approximately 2.5 J/mm^2) transfers the materials of stoichiometric BST to the heated substrate attached on the heater. The oxygen pressure in the deposition chamber was fixed at 200 mTorr, while the substrate temperature was maintained at 750 ^oC. The laser repetition rate was 5 Hz and the deposition rate was 0.033 nm/pulse. The structural properties of BST films were characterized by X-ray diffraction (XRD) θ–2θ and ϕ scans using a Rigaku X-ray diffractometer equipped with a Cu K_α radiator source and a 4-circle X-ray diffractometer. The thickness of the deposited BST films was about 400 nm, which was confirmed by a cross-sectional scanning electron microscope.

To investigate the dielectric properties of BST films on an MgO substrate in the microwave region, two kinds of devices were made. One was an interdigital (IDT) capacitor and the other was a CPW transmission line. A 2 μm thick Au/Cr metal electrode was patterned on the BST/MgO substrate by photolithography and a dry etching method. The gap size between the finger lines, the finger line width, and the number of fingers for the IDT device were 3 to 10 μm, 5 to 20 μm, and 3 to 7, respectively. In the CPW device, the gap between the signal line and ground, the width, and the transmission line length were 5 to 20 μm, 5 to 20 μm, and 3 to 14 mm, respectively.

Microwave performances of these devices were measured using an HP 8510C vector network analyzer at a frequency range of 0.5 to 35 GHz, and the measurement equipment was calibrated up to the microwave probes to minimize measurement uncertainty using a commercial standard calibration pad. A dc bias field between the finger lines, or in the gap to modulate the dielectric constant of BST film, was applied through a bias tee to protect the vector network analyzer.

In the BST IDT capacitors, the dielectric properties of the BST film on an MgO substrate were calculated by applying the equivalent circuit model and conformal mapping method to the measured reflection scattering parameter [20]. Based on the measured dielectric constant of BST films on MgO substrates, the impedance of the BST CPW phase shifter was calculated by the modified conformal mapping method.

The bias tee for the X-band application was designed using a bandpass filter with quarter-wavelength and open-circuited stubs, simulated with commercial electromagnetic software, and made using general photolithography for the Au/MgO structure. The microstrip patch antenna was designed using commercial software and made for a 20 mil-thick Duroid RT 5880 substrate (ε_r = 2.2). The microwave performance of the bias tee and the resonant frequency of each patch were measured using the vector network analyzer. The beam scanning patterns of a phased array antenna system, which was assembled by combining power divider, bias tee, ferroelectric CPW phase shifters, and patch antenna were measured using an NSI 255L near-field measurement system.

III. Results and Discussion

Figure 1 shows an X-ray diffraction (XRD) pattern of the BST thin film deposited on an MgO (001) substrate. Only (00*l*) peaks appeared in the XRD pattern, implying that the BST thin film was well aligned along the (001) direction. And the full-width at half-maximum (FWHM) of the (002) BST peak in the rocking curve was 0.55^o. The XRD ϕ scan of the {112} reflections of BST film on the MgO substrate is shown in the inset of Fig. 1. For (001) BST films, these peaks are expected to be seen in the XRD ϕ scan by interplanar angles calculation, $\alpha = 35.3°, \beta = 90° \quad \left((112), (\bar{1}12), (1\bar{1}2), (\bar{1}\bar{1}2)\right)$, coinciding with the experimental results and indicating an alignment of the BST <001> and <100> parallel with the <001> and <100> of the substrate, respectively. The ϕ scan (pole figure) in the inset of Fig. 1 also supports that the BST thin film is (001)-oriented on MgO (001) substrates.

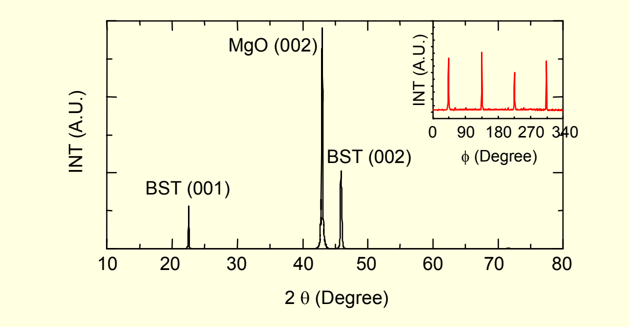

Fig. 1. X-ray θ-2θ diffraction scan patterns of (001)-oriented BST films deposited on MgO (001) substrates. The inset figure shows the phi-scan plots of the {112} reflections for the samples.

In designing the ferroelectric CPW phase shifter, not only the geometrical dimensions but also the physical values are important factors as they determine the characteristic impedance of the device to influence its loss characteristics. To know the characteristic impedance of the CPW phase shifter based on a BST/MgO structure, the dimension of the electrode and the thickness and dielectric constants of the BST film and MgO substrate should be known.

Although the geometric dimension can be easily measured, the microwave dielectric properties of the BST film on an MgO substrate can not be obtained directly as in the low frequency measurement case because the wavelength of measurement frequency is not much larger than the device size. The one method is an application of both an equivalent circuit model and conformal mapping to the measured 1-port reflection scattering parameters S_{11} for the BST IDT capacitor, which are plotted in Fig. 2(a) over the frequency region of 0.5 to 35 GHz with 0 and 40 V dc biases. This shows that the device still behaves like a capacitor for all measured frequency regions, since the measured S_{11} are in the lower hemisphere of the Smith chart. Therefore, we extracted the capacitance and quality factor of the BST IDT capacitor from the S_{11} using a parallel resistor-capacitor model. The capacitance of the device is shown in Fig. 2(b) as a function of frequency under external dc bias voltage from 0 to 40 V with 10 V steps. The capacitance variation due to soft mode hardening and the quality factor of the BST IDT capacitor with no dc bias were about 50 % and 20, respectively. These results demonstrate that the BST film is a promising material for the application of tunable microwave devices at room temperature.

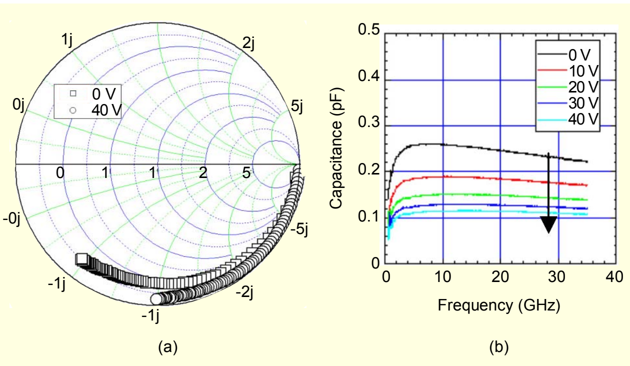

Fig. 2. (a) Smith chart of reflection scattering parameter S_{11} for the IDT device based on (001) oriented BST film. The longer and shorter lines correspond to the S_{11} for the IDT with the applied dc bias of 0 and 40 V, respectively. (b) The measured capacitance as a function of frequency with dc bias from 0 to 40 V for the interdigital capacitor device.

Relative dielectric constants of the BST film, ε_{BST}, were extracted from the capacitances of the BST IDT capacitor using the conformal-mapping model. In the capacitance of the BST IDT device based on MgO substrate, the dielectric constants of the substrate, film, and air have the following relation,

$$C = q_{MgO}\varepsilon_{MgO} + q_{BST}\varepsilon_{BST} + q_{air}\varepsilon_{air}\,, \tag{1}$$

where C is the capacitance of the BST IDT device and q_i' correspond to the factors for the MgO substrate, BST film, and air, respectively, which depend on the geometry of the device. Introducing both the geometrical dimensions and physical values of the constituent materials into (1) for the BST IDT device, the calculated dielectric constant of the BST film was about 1000 at 10 GHz with no dc bias.

In the conformal mapping method, the impedance of the BST CPW transmission line based on the MgO substrate is defined as follows:

$$Z_0 = \frac{30\pi}{\sqrt{\varepsilon_{eff}}}\frac{K(k_0')}{K(k_0)}, \tag{2}$$

and

$$\varepsilon_{eff} = q_{BST}\varepsilon_{BST} + q_{MgO}\varepsilon_{MgO} + q_{air}\varepsilon_{air}\,,$$

$$q_{BST} = \frac{K(k_1)K(k_0')}{2K(k_1')K(k_0)}, \tag{3}$$

where the effective dielectric constant ε_{eff} is a value considering all neighboring material, BST film, MgO substrate, and air. The K function is a complete elliptic integral and filling factors k_i and k_i' are defined as follows:

$$k_0 = \frac{w}{w+2g}, \quad k_0' = \sqrt{1-k_0{}^2},$$

$$k_1 = \frac{\sinh(\frac{\pi w}{4t})}{\sinh(\frac{\pi(w+2g)}{4t})}, \quad k_1' = \sqrt{1-k_1{}^2}, \tag{4}$$

where w is the width and g is the gap of the CPW device.

The calculated geometry dependent characteristic impedance using a conformal mapping method for a ferroelectric BST CPW transmission line is shown in Fig. 3, where the thickness of the BST film and MgO substrate were 400 nm and 0.5 mm, respectively, and the dielectric constant of the BST and MgO were 1000 and 9.6, respectively. The thickness of the BST film was chosen from the thickness-dependent saturation of the dielectric constant variation. And the dielectric constant of the BST film was determined as described in the above paragraph. The thickness of the MgO substrate was chosen because of the handy convenience, and the dielectric constant of the MgO substrate can be found elsewhere. In the figure, the wider the gap size is, the larger the impedance of the ferroelectric BST CPW transmission line. But there is a limit to increase the gap size because of the applied electric field dependent dielectric tunability. And the narrower the width size is, the larger the characteristic impedance of the ferroelectric BST CPW transmission line. Note that, there is also a limit to decrease the width size because of the line-width limit in the general photolithography and electric field clouding effect. Therefore, we chose the width and gap dimensions as 10 and 10 μm, respectively, though the characteristic impedance of the ferroelectric CPW phase shifter is smaller than 50 ohm.

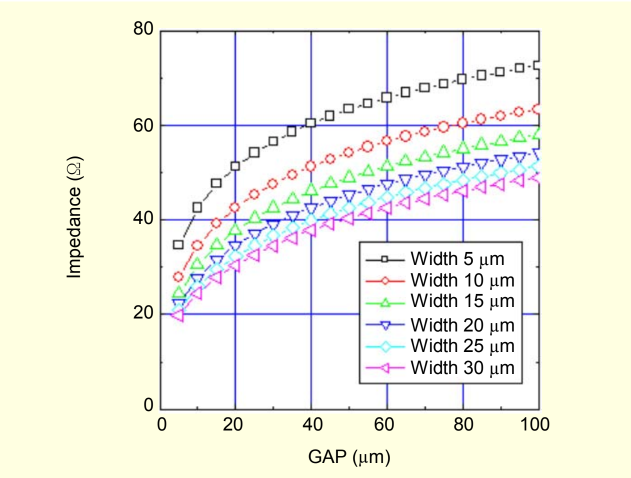

Fig. 3. The calculated characteristic impedance of a coplanar waveguide transmission line as a function of the width and gap based on a BST/MgO structure.

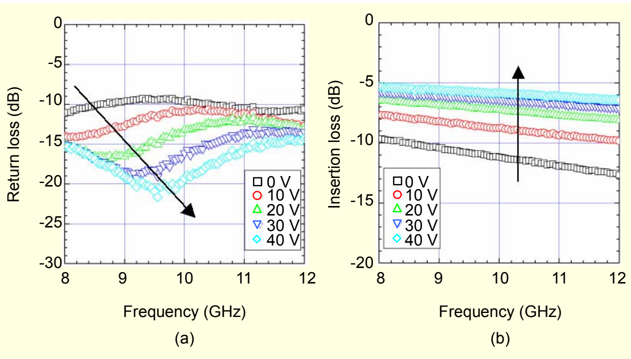

Fig. 4. Frequency dependent (a) return loss and (b) insertion loss for normal ferroelectric BST coplanar waveguide phase shifter with a variation of 40 V dc bias.

The microwave performance of the normal BST CPW phase shifter with 10 μm gap, 10 μm width, and 9 mm transmission line length to attain 180° differential phase shift is shown in Fig. 4. A periodic resonance could be observed in the frequency dependent return loss and is due to the coupling of the inductance from the center signal transmission line and capacitance between the center signal line and the side ground line. As the dc bias is applied, the dielectric constant of the ferroelectric film is decreased due to the soft mode hardening, and eventually the resonance moves to a higher frequency as shown in Fig. 4(a). The insertion loss is proportional to the measured frequency in the ferroelectric CPW device, as shown in Fig. 4(b), because the shortened transmitted wavelength with increasing frequency may experience a longer effective transmission line. The insertion loss is decreased as the applied dc bias is increased because the dielectric loss in the ferroelectric thin film is reduced for several reasons [21]. The return and insertion losses were about 9.7 and 11.2 dB at 10 GHz with no dc bias, respectively.

The somewhat large return and insertion losses are believed to be due to an impedance mismatch as mentioned above. To improve the loss characteristics, it is necessary to increase the characteristic impedance of the ferroelectric BST CPW phase shifter. As shown in (2), there are two ways to do this. The first is to decrease the complete elliptic integral, and the other is to decrease the effective dielectric constant of the device. In the former case, the larger gap size with respect to the width size is necessary to increase the characteristic impedance of the ferroelectric BST CPW phase shifter, but it needs the larger applied electric field to acquire the same phase tuning as in the smaller gap case, so there are some limits for this kind device to embody a commercial system. In the latter case, increasing the gap size, decreasing the thickness, or reducing the dielectric constant of the BST film in the structure is necessary to decrease the effective dielectric constant of the device. Also, note that it is difficult to decrease both the thickness below the critical thickness and the dielectric constant of the BST film below the critical limit due to deterioration of the dielectric properties of the BST film such as lower tunability or lower differential phase shift [22].

The flexibility to change the dimension of the CPW transmission line and the dielectric constant of the BST film in the ferroelectric phase shifter were somewhat limited, so the other method was tried. The core work is to add the impedance matching part at both ends of the CPW transmission line. Therefore, there are two parts in the BST CPW phase shifter, an impedance matching part and effective transmission line part. The structure of the impedance matching part was basically a CPW transmission line, which was comprised of a center signal line and two neighboring ground lines. The impedance of the input section in the impedance matching part was about 50 ohm, the dimension of which was determined by considering both the dielectric constant and thickness of the BST film and MgO substrate, as well as the connection bead dimension in the packaging metal box. The dimensions of the impedance matching part were calculated and simulated using the conformal mapping method and high frequency electromagnetic simulation tool. And the 50 ohm ferroelectric CPW dimension in the impedance matching part was 210 μm in width and 130 μm in gap size. In the impedance matching part, the width of the center line was decreased to that of the signal line in the effective transmission line part, and the gap size between the center line and the ground was decreased to that in the effective transmission line part.

Figure 5 shows the microwave performance of the ferroelectric BST CPW impedance matched transmission line and the phase shifter with added impedance matching part. As shown in Figs. 5(a) and 5(b), the return loss is below 20 dB from 1 GHz to 35 GHz, and the insertion loss is about 0.4 dB at 10 GHz for a 3 mm BST CPW transmission line. Excepting the existence of the impedance matching part, the dimensions and the constituent material conditions for the BST CPW phase shifter with added impedance matching part were the same as those of the device mentioned in Fig. 4. The total length of this ferroelectric CPW phase shifter was 14 mm, and the effective transmission line length was 9 mm. The differential phase shift was about 70 ^{o} with a 40 V dc bias variation at 10 GHz. In Figs. 5(c) and 5(d), the frequency-dependent return loss performances are somewhat different from those shown in Fig. 4 for a normal ferroelectric CPW phase shifter due to the existence of the impedance matching part. Please note that the total envelope of the return loss in the X-band region was lowered wholly, and the value was improved about 5 dB considering the highest point. Comparing the measured results in Fig. 4(b) with those in Fig. 5(d), the insertion loss was also improved by about 1 dB due to the improvement of the return loss for this technique-adapted BST CPW phase shifter. The return and insertion losses were about 20 and 10 dB with no dc bias, respectively, and the differential phase was about 180 ^{o} with the 150 V dc bias variation at 10 GHz. Though the performances of the ferroelectric BST CPW phase shifter are not better than those of other designs such as coplanar-strip, microstrip, and transmission types [9], [23], and [24], the ferroelectric CPW phase shifter has the merits of a simple process and low production cost.

The circuit of the bias tee basically consisted of an inductor, capacitor, and balun based on the microstrip line, and the role is to protect the measurement system from the applied dc voltage. The microwave performances of the designed bias tee were simulated using the electromagnetic high frequency software, and the schematic figure is shown in Fig. 6(a). The simulated results for this device are shown in Fig. 6(b). The return and

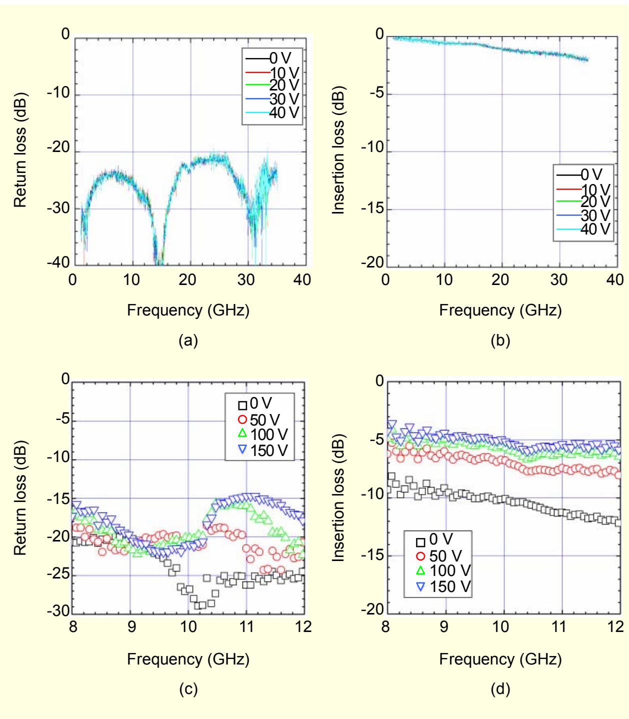

Fig. 5. Frequency dependent (a) return loss and (b) insertion loss for ferroelectric BST coplanar waveguide transmission line with a 210 μm width and 130 μm gap size, and (c) return loss and (d) insertion loss for ferroelectric BST coplanar waveguide phase shifter with added impedance matching part with a variation of 40 V dc bias.

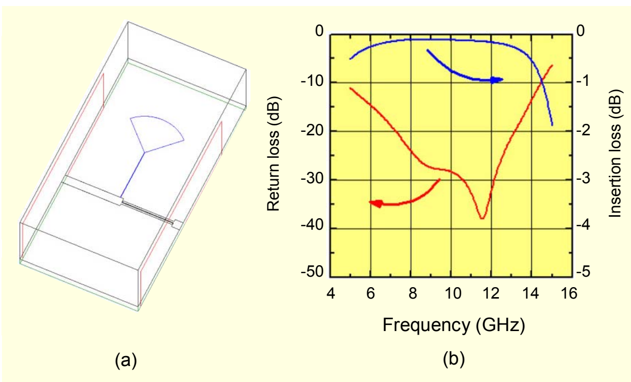

Fig. 6. (a) A schematic figure and (b) simulated results of using electromagnetic high frequency software for the designed bias tee.

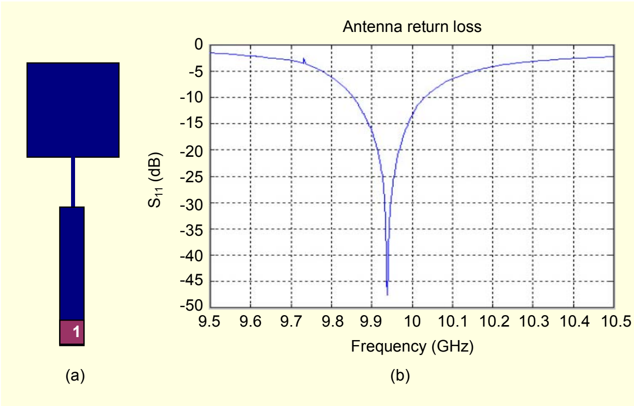

Fig. 7. (a) A schematic figure and (b) measured return loss characteristics of the designed microstrip patch antenna.

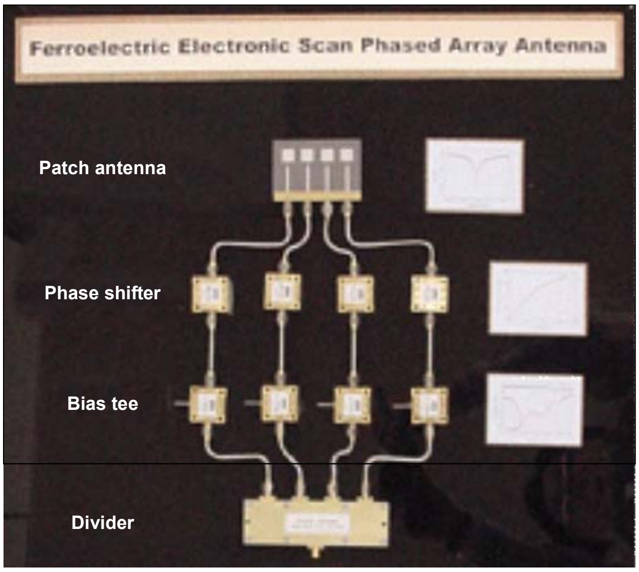

Fig. 8. A picture of phased array antenna system consisting of power divider, bias tees, ferroelectric phase shifters, and patch antennas.

insertion losses were below 20 dB and 0.2 dB in the X-band region, respectively, and the measured results based on the Au/MgO structure were on the whole similar to the above simulation results.

The microstrip patch antenna was designed and simulated by using high frequency software. Its shape and measured return loss characteristics are shown in Fig. 7. The shift in resonant frequency was much smaller, about 1%, with respect to the simulated results, and the measured loss characteristics and band width were adequate for the experiment.

The constituents of the phased array antenna were power divider, bias tees, ferroelectric phase shifters, and patch antennas as shown in Fig. 8, and electrical power source. In the phased array antenna system, the measuring signal was divided in the commercial power divider and transmitted to each bias tee, ferroelectric phase shifter, and the radiator. The superposed beam, which was radiated from each patch antenna, could be controlled electronically by each signal modulation in each phase shifter under the programmed dc bias.

Using a computer-aided program, the predetermined applied voltage set was operated on the four ferroelectric BST phase shifters. The direction of the superposed phased array antenna beam could be analog/digital controlled by the predetermined applied voltage sets. In this work, using the above mentioned ferroelectric BST phase shifters, the differential phase shift between the phase shifters was 0^{o}, 20^{o}, 40^{o}, and 60^{o}. The radiation patterns of the 1×4 linear phased array antenna with ferroelectric phase shifter array are shown in Fig. 9. The beam of the phased array antenna system could be tilted maximally about 15^{o} in one direction at 10 GHz.

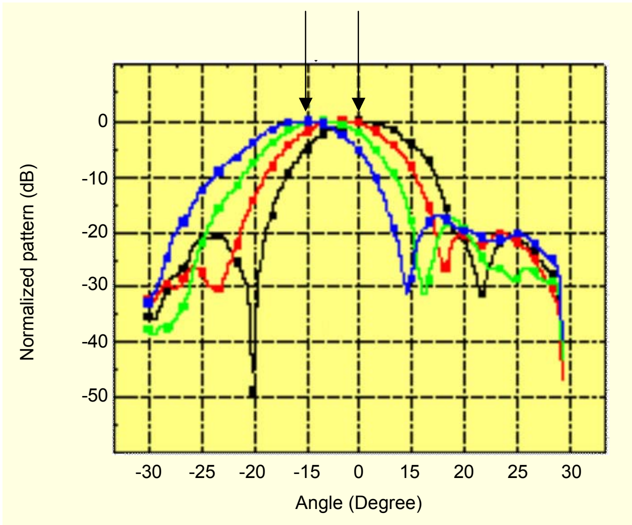

Fig. 9. Beam pattern of phased array antenna using ferroelectric BST coplanar waveguide phase shifter with a variation of dc bias. The differential phase shift between the phase shifters was 0°, 20°, 40°, and 60°, and the beam of the phased array antenna system could be tilted maximally about 15° in one direction at 10 GHz.

IV. Conclusion

We have designed, fabricated, and tested a CPW phase shifter based on BST film for electronic scan phased array antenna application. The return and insertion loss of the ferroelectric CPW phase shifter were improved using an impedance-matching-part adding technique. The concept of beam scanning characteristics of an X-band 1×4 phased array antenna using four-element ferroelectric BST coplanar waveguide phase shifters has been proven. We have demonstrated that the main beam of the phased array antenna could be continuously scanned by an analogous variation of the applied dc voltage. The steering beam could be tilted by 15 ^{o} in either direction. From the experimental results, the ferroelectric phase shifter can be a next-generation phase shifter for an electronic scan phased array antenna.

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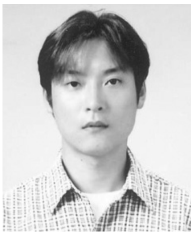
Seung Eon Moon received the BS, MS and PhD degrees in physics from Seoul National University, Seoul, Korea in 1990, 1994 and 2000. Since 2000, he has been working for Electronics and Telecommunications Research Institute (ETRI) in the development of microwave tunable devices based on ferroelectric thin films. His current research activities are in the application of a phased array antenna based on ferroelectric microwave/millimeter components, and the development of high density data storage and intelligent fusion sensor systems.



Han Cheol Ryu received the BS degree in electronic engineering from Hanyang University, Seoul, Korea in 2000 and the MS degree from Pohang University of Science and Technology, Pohang, Korea, in 2002. Since 2002, he has been working for ETRI in the development of ferroelectric microwave devices. His current research activities are in the development of antennas and microwave/millimeter tunable devices.

Min Hwan Kwak received the BS and MS degrees in electronic materials engineering and the PhD degree in electrical engineering from Gyeongsang National University, Jinju, Korea, in 1995, 1998 and 2002. He joined ETRI in 2002. He is involved in the development of high-Tc superconducting (HTS) thin film and microwave devices and subsystems. His current research interests are in the development of ferroelectric and HTS microwave/millimeter tunable devices and subsystems.

Young Tae Kim received the BS and MS degrees in electrical engineering and the PhD degree in electronic engineering from Soonchunhyang University, Asan, Korea, in 1998, 2000 and 2004. Since 2004, he has been working for ETRI in the development of ferroelectric microwave devices. His current research activities are in the development of ferroelectric microwave/millimeter tunable devices and subsystems.

Su-Jae Lee received the BS degree in physics from Kyungsung University, Korea, in 1982, and the MS and PhD degrees in physics from Pusan National University, Busan, Korea, in 1982 and 1997. He joined ETRI in 1997, where he worked in the development of microwave tunable dielectric thin films and tunable devices for next generation wireless communication. During 2000 to 2005, he was a project leader on the development of a ferroelectric phase shifter MMIC for a phased array system, and is currently a Senior Researcher in the Basic Research Lab at ETRI. His research interests include the development of new multifunctional nano-oxide dielectrics for the creation of next-generation nano-oxide electronic devices and nano-sensors.

Kwang-Yong Kang received the BS degree from Seoul National University in 1975 and the PhD degree in solid-state physics from Pusan National University in 1988. He joined ETRI in 1989, where he was part of the Basic Research Department, involved in the development of HTS thin films and microwave/millimeterwave devices. His research interests include metal-insulator transition devices and HTS/ferroelectric microwave devices. He has recently explored THz technologies. He is a Member of IEEE MTT and OSA of USA and JJAP of Japan. He is also a Member of IEEK and KPS of Korea. Currently, he is a Director of the Terminal Components Research Department of the Basic Research Laboratory, ETRI.