A Low-Loss *Ku*-Band Monolithic Analog Phase Shifter

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***Abstract*—A GaAs monolithic Ku-band analog phase shifter integrating a 90° branch line coupler with planar varactor diodes has been fabricated for the first time. A phase shift of 109°±3° with an insertion loss of 1.8±0.3 dB was measured from 16 to 18 GHz. A 360° phase shifter with 4.2±0.9 dB insertion loss was realized in the same frequency range by connecting three phase-shifter chips in series. To our knowledge, this is the lowest insertion loss obtained by a 360° *Ku*-band phase shifter using monolithic circuits. In addition, hyperabrupt varactors using nonuniform doping profiles increased the phase shift by more than 30° and produced a more linear dependence of phase shift on control voltage.**

I. Introduction

PHASED-ARRAY radar systems need a phase shifter in each module for the beam scanning required to search and track multiple targets. It is therefore important that a phase shifter designed for a monolithic GaAs transmit–receive (TR) module occupy a small area and use a minimum number of passive and active components. An analog phase shifter is advantageous in this respect because it can provide a given phase shift with less area and fewer devices than a digital phase shifter.

A number of hybrid analog phase shifters have been developed [1]–[3]. However, there has been only one monolithic analog phase shifter reported [4]. In this phase shifter, the Schottky-barrier gate capacitance of a GaAs MESFET was used as the varactor, and a phase shift of 105° with 2.5±0.5 dB insertion loss was measured at *X*-band. In order to minimize the series resistance between the gate (anode) and the source (cathode), critical photolithography or additional ion implantation was needed. Furthermore, it is difficult to extend this approach to higher frequencies due to high gate resistance and the effects of distributed elements along the gate fingers.

In this work, we report a low-loss *Ku*-band monolithic analog phase shifter using planar varactor diodes and a 90° branch-line coupler. The fabrication of this circuit is very simple and a high yield is expected. In addition, since GaAs monolithic circuits consisting of planar Schottky diodes and a 90° coupler have been demonstrated to be suitable for *Ka*-band applications [5], we believe that this approach should be easily extendable to higher frequencies.

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II. Analog Phase Shifter Design

Our goal was to build a low-loss GaAs monolithic analog phase shifter with minimum variations in phase shift between 16 GHz and 18 GHz. Fig. 1(a) shows the circuit diagram of the analog phase shifter and Fig. 1(b) is a scanning electron microscope (SEM) micrograph of the finished chip. The chip size is 3.1 mm×2.4 mm and it is 0.127 mm thick. The phase-shifter circuit consists of a 90° branch line coupler, 50-Ω microstrip lines connecting the coupler and the varactors, tuning inductors, and a dc bias line. The lengths of the 50-Ω microstrip line and the tuning inductor are l and L, respectively. A 90° branch-line coupler was chosen because of its low loss and simplicity. The varactor diodes are connected to the 0° and −90° ports of the branch-line coupler. Since the phase shift is determined by the reactance presented at these ports, it is controlled by varying the capacitance of the varactor diodes [6].

The planar varactor diode occupies a smaller area than a MESFET, and does not require ion implantation or critical photolithography to lower its series resistance, which is a dominant factor in determining the insertion loss of the phase shifter. The phase shift which can be produced is proportional to the capacitance change of the varactor diodes. However, the thicker n-type epitaxial layer required to support the larger excursions in depletion regions will contribute to a higher series resistance. The choice of the doping concentration consequently involves a tradeoff among breakdown voltage, series resistance, and the rate of the capacitance change with respect to the bias voltage. In our design, we used 5000 Å of n-type layer with uniform doping concentration of 8×10^{16} cm^{-3}. This layer should have a 4.2:1 capacitance ratio between 0 V and reverse breakdown if there is no parasitic capacitance. However, in reality the smallest capacitance, and the largest phase shift attainable, are limited by the parasitic capacitance. In this work we chose the diode area such that the capacitance at zero bias is 0.35 pF. Including the parasitic

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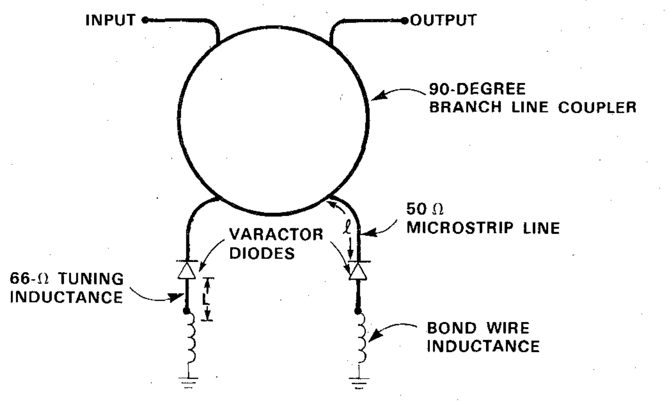
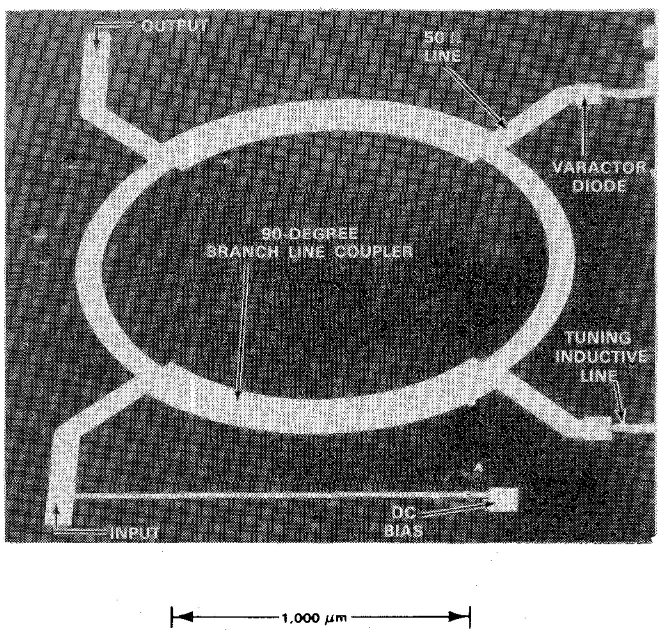


(a)



(b)

Fig. 1. (a) Circuit diagram of the analog phase shifter. (b) SEM micrograph of the finished chip.

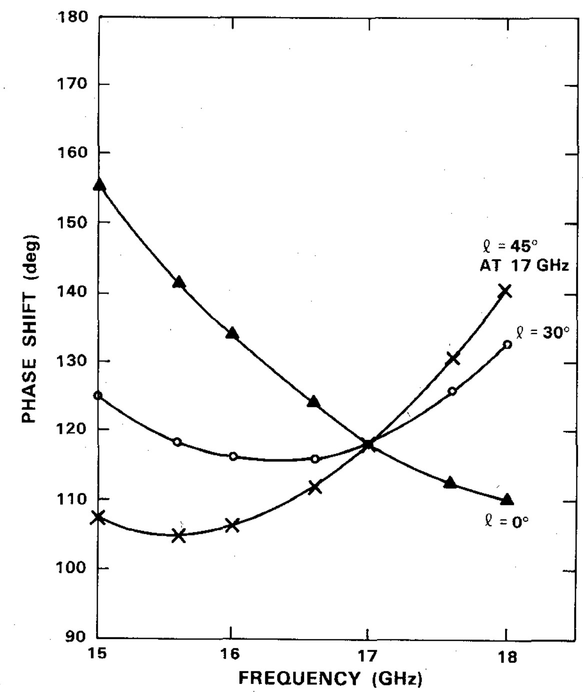


Fig. 2. Calculated phase shift for various lengths of the 50-Ω line in the circuit. The length is expressed as the electrical length at 17 GHz. An ideal 17-GHz 90° hybrid is assumed and the phase shift is defined as the difference of the insertion phase for varactor capacitance values of 0.35 pF and 0.1 pF. A 175-μm length of 66-Ω microstrip line and 0.2 nH of bond wire inductance are also included.

capacitance, a 3.5 : 1 capacitance ratio is expected for the varactor diode.

Compared to the uniformly doped material, a hyperabrupt varactor with a modified doping profile has a faster capacitance change rate with the bias voltage. As a result, a phase shifter using hyperabrupt varactors will have a larger total phase shift and a higher phase-shift resolution. In addition, a proper choice of doping profile can provide a more linear phase shift with the bias [7], [8]. However, the main reason for using a hyperabrupt varactor here is to increase the total phase shift. Instead of a continuously graded doping profile, three epitaxial layers with different doping concentrations were grown by molecular beam epitaxy (MBE) for simplicity.

A section of 50-Ω line is used between the coupler and the varactor to minimize the phase-shift variations over the bandwidth. The calculated phase shift for different lengths of the 50-Ω line is shown in Fig. 2. In this calculation, we assumed values of 0.35 pF and 0.1 pF for the 0-V and breakdown-voltage varactor capacitances, respectively, 0.2 nH for bond wire inductance, and a 17-μm length of 66-Ω microstrip line for the tuning inductance. In the absence of the 50-Ω line, indicated as $l = 0^0$ in Fig. 2, the phase shift decreases monotonically from 16 GHz to 18 GHz. As the length of this line increases, the phase-shift variation at lower frequencies is reduced and a phase-shift minimum occurs close to the design center frequency. If the length increases further, the phase shift increases monotonically with frequency in the range measured. In this work, a length of 30° at 17 GHz was chosen in order to keep the phase-shift variation with frequency small. As shown in Fig. 2, the phase-shift variation between 16 GHz and 18 GHz was $\pm 8°$.

The tuning inductance affects the total phase shift as well as its variation over the bandwidth. A section of 66-Ω microstrip line was used as tuning inductance. The phase shift is plotted in Fig. 3 for various lengths of inductive lines. The amount of phase shift increases with the length of this tuning inductor. However, for lengths exceeding approximately 85 μm, the variation also increases with the line length. For instance, between 16 GHz and 18 GHz the variations are $\pm 13°$ for the 350-μm line and $\pm 8°$ for the 175-μm line. Therefore, a 175-μm line was chosen as a compromise, providing moderate total phase shift and an acceptable phase-shift variation. Note that the variation of phase shift with frequency is dependent on the lengths of both the 66-Ω tuning inductor and the 50-Ω microstrip connecting line. However, the absolute value of the phase shift at the design center frequency is independent of the length of the 50-Ω connecting line.

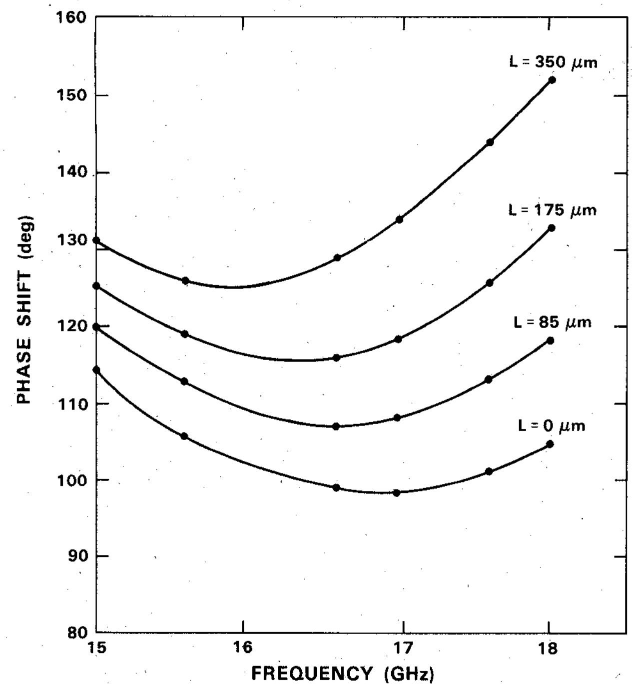


Fig. 3. Calculated phase shift for different lengths of the tuning inductor. The impedance of the inductive line is 66 Ω and the electrical length of the 50-Ω line is 30° at 17 GHz. The rest of the parameters are the same as those used in Fig. 2.

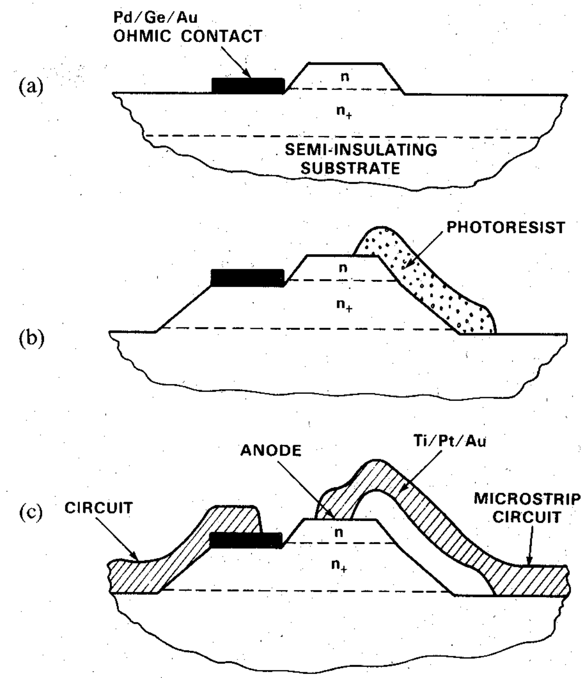


Fig. 4. The process steps of the monolithic analog phase shifter. (a) Ohmic contact. (b) Mesa etch and air bridge definition. (c) Electron-beam evaporated Ti/Pt/Au for anode, air bridge, and microstrip circuit.

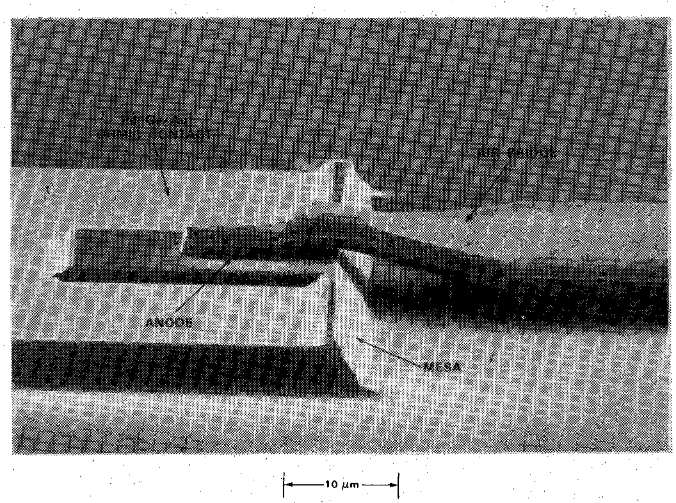


Fig. 5. An SEM micrograph of the planar varactor diode in the phase shifter. The nonalloyed Pd/Ge/Au ohmic contact has very smooth surface.

III. Device and Circuit Fabrication

GaAs epitaxial layers grown by MBE were used for the varactor diodes. A 2-µm n^+ layer doped to 1×10^{18} cm^{-3} was first grown on a semi-insulating substrate, followed by a 5000-Å n-type layer with 8×10^{16} cm^{-3} doping concentration. In addition to the uniformly doped n-type layer, a quasi-hyperabrupt material was also used. In this material, three epitaxial layers, 3000 Å of 1×10^{16} cm^{-3}, 4000 Å of 3×10^{16} cm^{-3}, and 4000 Å of 8×10^{16} cm^{-3}, were grown sequentially on top of the n^+ layer.

The fabrication steps are shown in Fig. 4. The n^+ layer in the ohmic contact region was first exposed by chemically etching the n-type epitaxial layers on top. Then Pd (300 Å)/Ge (400 Å)/Au (3400 Å) metallizations were defined by lift-off. The contact metallizations were sintered at 450°C for 30 s to form the ohmic contact. This recently developed nonalloyed contact [9] has low contact resistance and very smooth surface morphology, which is particularly important for devices with small area. A mixture of hydrochloric acid, hydrogen peroxide, and water was used to etch deep mesas for diode isolation. This mixture produced a sloped side wall to avoid any possible discontinuity in the metallizations or photoresist at the edge of the mesa. An air bridge was used to connect the anode of the diode and the microstrip circuits on the semi-insulating substrate. The bridge was first defined using photoresist. Then the 2.5-µm-thick Ti/Pt/Au metallization for the anode of the diode, microstrip circuit, and the air bridge was electron-beam evaporated and defined by PMMA and AZ-1470 double-layer resist lift-off. We have fabricated comparable air bridges using an electroplating technique, but have found that the evaporated air bridge has better dimensional control and the processing is simpler. Because the same metallization is used for anode, air bridge, and circuit, only four masks are needed. This simple fabrication process is the key for high yield and reproducibility. Fig. 5 is an SEM micrograph of the varactor diode showing a featureless Pd/Ge/Au ohmic contact.

IV. Measured Results

The varactor diodes fabricated on the uniformly doped epitaxial layer had a reverse breakdown voltage of 16 V and a series resistance of less than 4 Ω. The average capacitance was 0.34 pF at zero bias and 0.11 pF at reverse breakdown. The capacitance ratio was approximately 3 : 1,

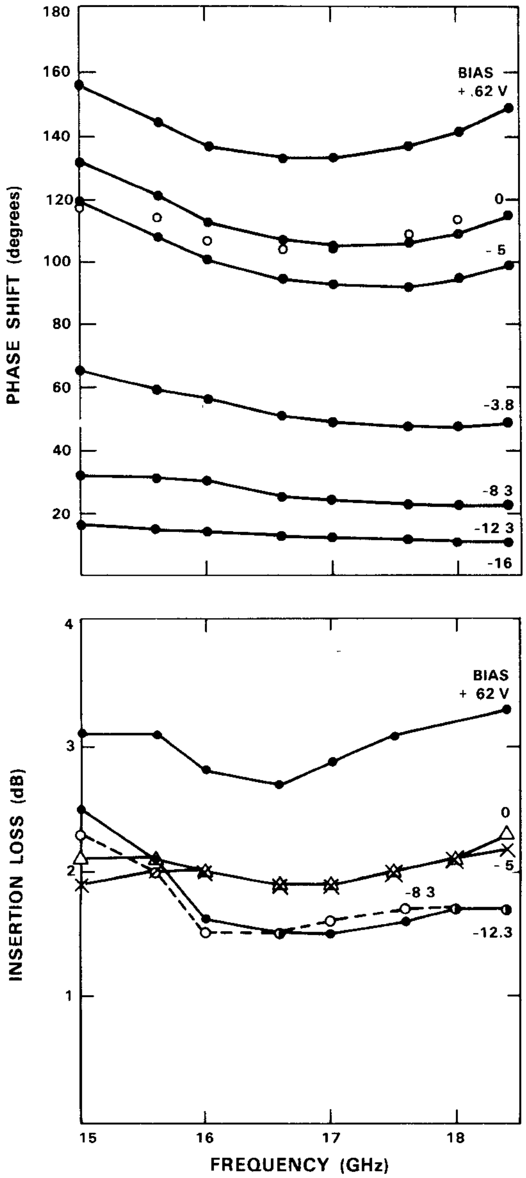


Fig. 6. Measured phase shift and insertion loss of the phase shifter fabricated on the uniformly doped material. The open circles are the calculated phase shift and the insertion loss includes 0.5 dB of test-fixture loss.

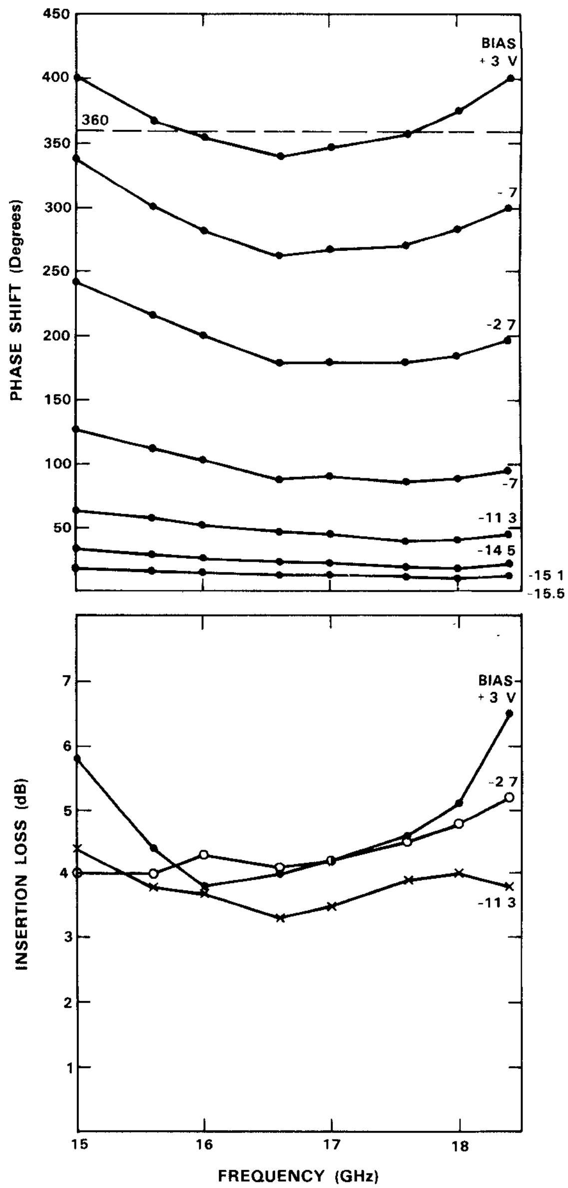


Fig. 7. Measured phase shift and insertion loss of the 360° phase shifter which was realized by cascading three phase shifter chips.

which was very close to the predicted value. From an analysis of the capacitance of different sizes of diodes, a parasitic capacitance of approximately 0.03 pF was calculated. The intrinsic capacitance ratio inferred using this parasitic capacitance is 3.9:1.

Measured phase shift and insertion loss are shown in Fig. 6. The phase measured at 16 V reverse bias was used as the reference. From 0 to 16 V reverse bias, a phase shift of $109° \pm 3°$ was obtained between 16 GHz and 18 GHz. If a forward bias of 0.62 V is applied, the total phase shift increases to $138° \pm 3°$, but the insertion loss also increases rapidly. In Fig. 6, the open circles in the phase-shift plot are phase shifts calculated using 0.34 pF and 0.11 pF as the zero-bias and 16-V varactor capacitances, respectively. Satisfactory agreement, within 5°, was obtained from 16 to 18 GHz. For reverse biases, the measured insertion losses were 1.8 ± 0.3 dB. All the insertion losses reported here include approximately 0.5 dB of test-fixture loss. Increasing the input power to 20 dBm resulted in no measurable change in phase shift and insertion loss for any bias voltage. The average VSWR was less than 1.3:1 for reverse-biased diodes and the worst case was 1.6:1.

Because both the input and output impedances are 50 Ω, the phase shifters can be cascaded directly. A 180° phase shifter was realized by connecting two such phase-shifter chips in series. From 0 to 16 V reverse bias, a phase shift of $183° \pm 7°$ with an insertion loss of 3.2 ± 0.6 dB was measured from 16 GHz to 18 GHz. This is consistent with the phase shift and insertion loss of the individual chip, and only one dc control voltage is required. A full 360° phase shifter was constructed in the same manner by cascading three chips. As shown in Fig. 7, the measured phase shift was $359° \pm 17°$ from 16 GHz to 18 GHz and the insertion loss was 4.2 ± 0.9 dB. To the best of our knowledge, this is the lowest insertion loss reported for a *Ku*-band monolithic 360° phase shifter using active solid-state devices.

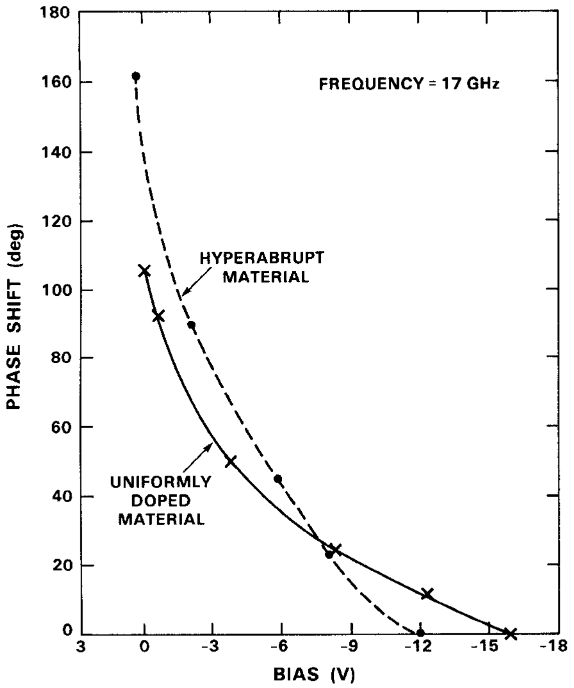


Fig. 8. Measured phase shift as a function of the dc bias at 17 GHz. The solid line is for the phase shifter fabricated on a uniformly doped material and the broken line is for the hyperabrupt material.

The hyperabrupt diodes with a modified doping profile have larger capacitance variation for the same amount of bias change. As a result, larger phase shift for the same dc control bias range are expected for phase shifters incorporating the hyperabrupt varactor diodes. Analog phase shifters identical to those previously described were fabricated on the material with the multiple epitaxial layers described in the previous section. We also reduced the mesa to one half of its original size to minimize the effects of the parasitic capacitance. At zero bias, the diode capacitance was 0.29 pF and a capacitance ratio of 6.3:1 was obtained between zero and 12 V reverse bias. The series resistance was 3.7 Ω, which is similar to the uniform-profile varactor diodes we fabricated. As expected, a significantly larger phase shift was measured. From 16 GHz to 18 GHz, a phase shift of $166° \pm 6°$ was measured between -12 and $+0.5$ V, which is the largest forward bias one can apply without increasing the insertion loss. The measured insertion loss was 1.9 ± 0.6 dB for the same bias range. From 0 V to -12 V, a phase shift of $146 \pm 8°$ was measured, which is an increase of more than 30° compared to the phase shifter using regular varactor diodes. Using hyperabrupt varactors does provide a more linear phase shift than that with a regular varactor with uniform doping profile, as illustrated in Fig. 8. Although linear phase shift is not required in this work, a more linear phase shift response means a more precise phase control.

V. Summary and Conclusions

We have fabricated an analog phase shifter at *Ku*-band with greatly simplified processing. The performance is accurately predicted by the theoretical calculations. The use of the evaporated air bridge greatly simplified the process and improved the dimensional control. The Pd/Ge/Au metallizations enabled us to obtain a very uniform ohmic contact which can be further scaled down in area to reduce the parasitic capacitance. Using regular varactor diodes with uniformly doped epitaxial layer, a full 360° phase shifter was realized by cascading three phase-shifter chips. From 16 GHz to 18 GHz, the measured phase shift and the insertion loss were $359° \pm 17°$ and 4.2 ± 0.9 dB, respectively. The phase shifter with hyperabrupt varactors had a larger phase shift with better linearity as a function of the bias voltage. It is feasible to construct a 360° phase shifter using two such chips with approximately 3 dB of insertion loss. Because of the high-frequency capability of the varactor diodes and the simplicity of the design and fabrication, we feel that the analog phase shifter approach reported here can be readily applied to higher frequencies, such as the millimeter-wave frequency range.

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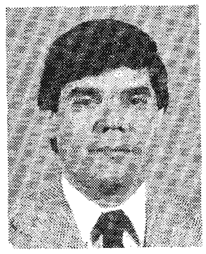

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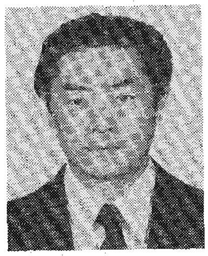

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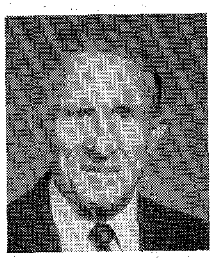

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