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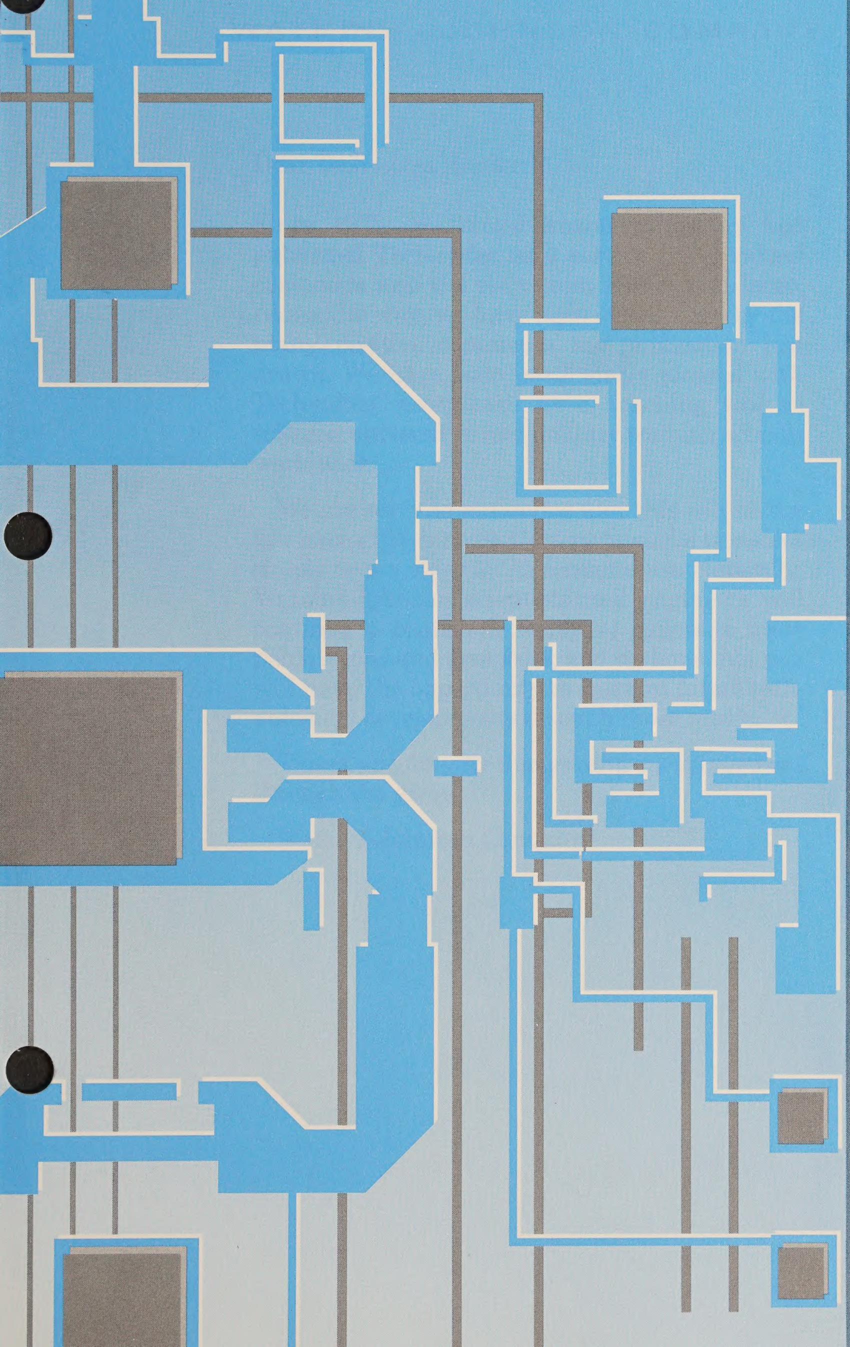
GaAs MMIC Voltage-Controlled Phase Shifter



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Demand is increasing for high-resolution phase shifters for use in small tactical phased arrays, low spurious-signal frequency translators, and some FMCW radars. This paper describes an analog MMIC phase-shift circuit that can be swept continuously through 105 degrees while amplitude is held constant. Thus, phase and amplitude resolution are dependent only on analog control voltage resolution. The phase-shifter is suitable for operation over a minimum 10-percent bandwidth anywhere from 8 to 12 GHz. Die size is 2.0 by 1.7 mm.

Operation of the phase shifter can be summarized as follows: The input signal is divided into two branches in which phase is separated by a nominal 120 degrees using lumped element high- and low-pass filtering as lead and lag networks. MESFET-based pi-attenuators control the two signal-vector amplitudes, which are then recombined with an in-phase combiner. The MESFET attenuators, each combined with a simple off-chip control circuit, provide a very temperature-stable linear attenuation (in dB) versus control voltage. This leads to a phase shifter that has low sensitivity to temperature-induced variations and requires only two control voltages.

Design

The objective was to design an X-Band MMIC phase shifter which would be bidirectional, provide phase resolution of better than 5 degrees (equivalent to 6-bit accuracy), and have potential for phase resolution of 2 to 3 degrees.

It was anticipated that this phase shifter would be cascaded with fixed 180 and 90 degree differential phase bits and that each of these could have maximum errors of ±5 degrees over 8 to 12 GHz. In addition, MESFET attenuator phase was expected to be affected to some degree by both amplitude setting and temperature. Finally, some extra tolerance was desired for design uncertainty and process variations. Thus, to insure that a full 360 degrees of continuous phase control could be obtained, the high-pass and low-pass filters that form the lead/lag network of the analog phase shifter were designed to provide 120 degrees of differential phase between branches.

A block diagram of the basic microwave circuit selected is presented in Figure 1. Figure 2 is a microphotograph of the resulting MMIC phase shifter.

Combiner/Divider Network

A lumped element "L" network [1] design is used for the power divider and combiner. The advantages this network has over a lumped-element Wilkinson network are 1) slightly wider isolation bandwidth, 2) a 28-percent lower value for the series inductors, which reduces loss and size, and 3) only two capacitors are required, instead of three. The complete divider occupies only 0.29 mm^2. Coupled spirals designed after Swanson [2] form the inductors. Mutual coupling for non-adjacent, as well as adjacent, lines is taken into account. The result is a simple lumped-element model in which parasitic shunt capacitances of the spiral inductor model can be absorbed in the low-pass divider network design. Capacitors are formed with a silicon-nitride process.

Filter Network

The low-pass and high-pass filters that follow the power divider/combiner are lumped-element designs using the same type of inductors and capacitors. The low-pass filter is a pi-network and the high-pass filter is a T-network.

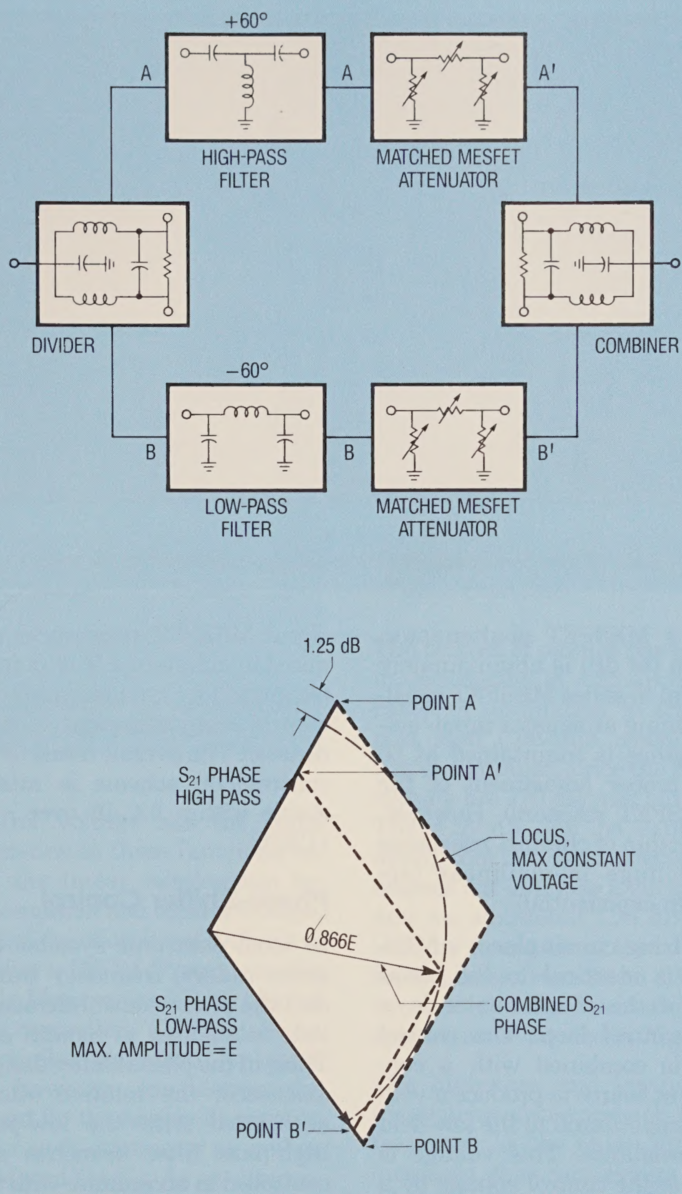

Figure 1. MMIC analog differential phase shifter.

Attenuators

The key to achieving high resolution phase control is a well-controlled temperature-stable attenuator design. The linearizing technique by Fisher and Dobkin [3] fulfills this requirement by producing an attenuation (in dB) versus control-voltage curve that is linear and inherently temperature compen-

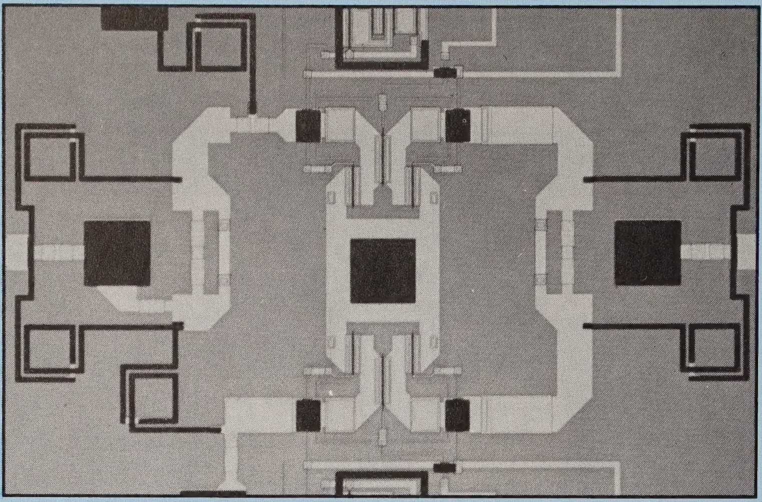

Figure 2. Microphotograph of analog voltage-controlled phase shifter.

sated. In a MESFET pi-attenuator, attenuation (in dB) is approximately proportional to series MESFET resistance (assuming attenuator input/output impedance is maintained at 50 ohms by proper adjustment of the shunt MESFET resistors). However, the relationship of channel resistance to gate voltage is nonlinear (approaches an exponential).

The linearizing circuit places a MESFET that is identical to the series MESFET of the rf attenuator in a feedback control loop. The control MESFET is combined with a constant-current source to produce a voltage drop proportional to the low-field channel resistance. This voltage is compared to the control voltage by a differential amplifier which then adjusts the rf attenuator MESFET gate voltage such that channel resistance is linearly changed. The most important result of the relationship between control voltage and channel resistance, established by the feedback control, is a variable attenuation that is independent of temperature. The rf attenuator shunt MESFET resistances are optimized to maintain a 50-ohm input/output match by an additional feedback control loop referenced to a 50-ohm resistor. The overall result of the two-control-loop scheme is attenuation stable within 0.4 dB over −55°C to +125°C.

Phase-Shifter Control

In most electronic systems (phased-array radars, frequency translators, etc.) the phase-control-element amplitude is required to remain constant. Thus, in the phase-shifter design being discussed, the relative attenuation associated with the low-pass and high-pass filter branches must be controlled in accordance with the functions given in equations 1 and 2. These relations are derived from the vector diagram shown in Figure 1.

$$A_L = \left| \, 20 \log \left(\cos\left(\Phi - 30^\circ\right)\right)^{-1} \right| \quad [1]$$

$$A_H = \left| \, 20 \log \left(\sin \Phi\right)^{-1} \right| \quad [2]$$

Zero relative phase ($\Phi = 0$) is defined as the condition where the high-pass filter branch attenuation is maximized. Figure 3 is a plot of these relationships.

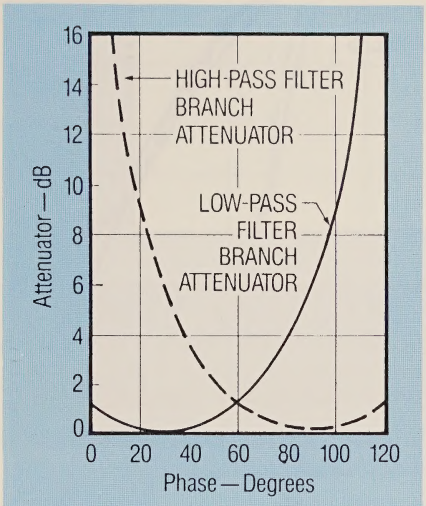

Figure 3. Attenuation functions required to maintain constant amplitude.

The control voltage has the same characteristics as these functions because of the linear relationship between attenuation and control voltage. This control function is not difficult to obtain using a PROM (Programmable Read Only Memory) and DAC (digital to analog controller). This may be an overly complex approach for some systems, however, considering the requirement for the analog linearizing circuits for each attenuator as well as the PROM and DAC's. As an alternative, the exponential relationship between the series MESFET attenuation and gate control voltage could be used so that the cosine functions shown in Figure 3 could be linearized (with break points required at 30 degrees on the low pass side and at 90 degrees on the high pass side.) However, a temperature-stabilization scheme would be required to substitute for the feedback attenuator control system presented in the "Attenuator" section. The data presented in the "Results" section were obtained using the feedback-linearized attenuator design for this reason.

The requirement that the circuit be bidirectional dictated that only passive circuit elements could be used. The consequence of this requirement was that high insertion loss had to be accepted. The factors that contribute to the loss and their estimated effects are:

Power Division Circuit	3.5 dB
Filters	0.5 dB
Variable Attenuator, Min.	2.0 dB
Power Combiner	4.8 dB

The process of passive power recombination results in 4.25 dB of loss, because of dissipation in the isolated port load (100-ohm resistor) of the power combiner, since the combining signals are 120 degrees out of phase. The 4.25 dB can be explained by noting that two vectors recombined 90 degrees out of phase incur a 3 dB loss and an additional 1.25 dB insertion loss by increasing the differential fixed phase from 90 to 120 degrees because the two vectors begin to oppose, as illustrated in Figure 1.

The high-pass and low-pass cutoff frequencies also approach the center frequency as the phase difference is increased, thus constricting the bandwidth and increasing insertion loss another 0.5 dB or more.

Results

Power dividers were included separately on the GaAs wafer for evaluation. The broadband performance of the "L" lumped-element divider mea-

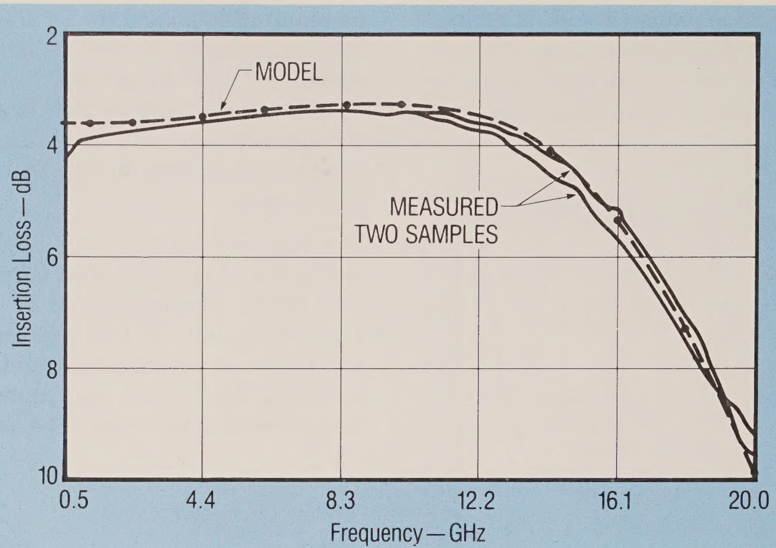

Figure 4a. "L" type lumped-element power divider insertion loss (including power division).

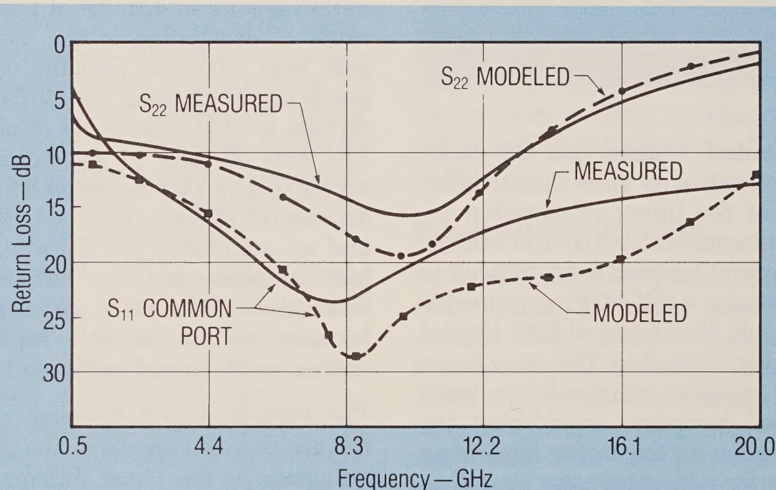

Figure 4b. Power divider measured and modeled return loss.

sured with on-wafer RF probing is shown in Figures 4a and 4b, along with the computer-predicted performance.

A power divider combined with the high- and low-pass filter networks was also included on wafer. Measured insertion loss and return loss for the high-pass filter side (common port was input), along with computer model predictions, are shown in Figures 5a and 5b. Similar data for the low-pass filter

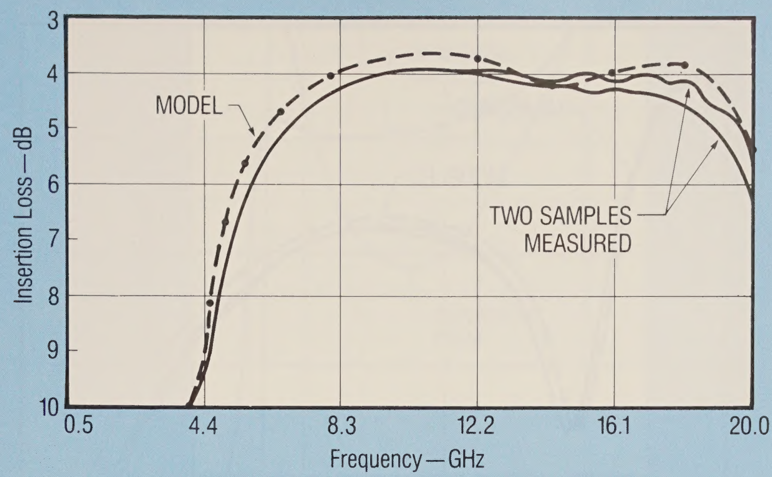

Figure 5a. Measured and modeled insertion loss for two sample power divider high-pass branches (including power division loss).

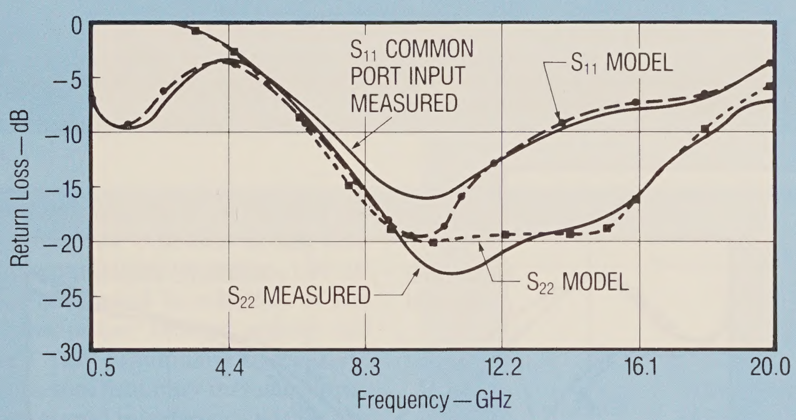

Figure 5b. Measured and modeled input and output return loss for power divider high-pass branch.

side is given in Figures 6a and 6b. Correlation with the computer model was acceptable, thus validating the model, including the coupled spiral. The insertion loss resonance at 4.5 GHz did not materially affect the performance in the design band, so the cause of the resonance was not investigated. The measured differential phase between the high- and low-pass filter branches was 120 ±3 degrees from 8 to 12 GHz, as shown in Figure 7. Differential loss between the filter branches was less than 0.5 dB.

Zero relative phase can be defined as the condition where the high-pass filter branch attenuation is maximized. If the high-pass filter branch attenuation is decreased to a minimum, phase

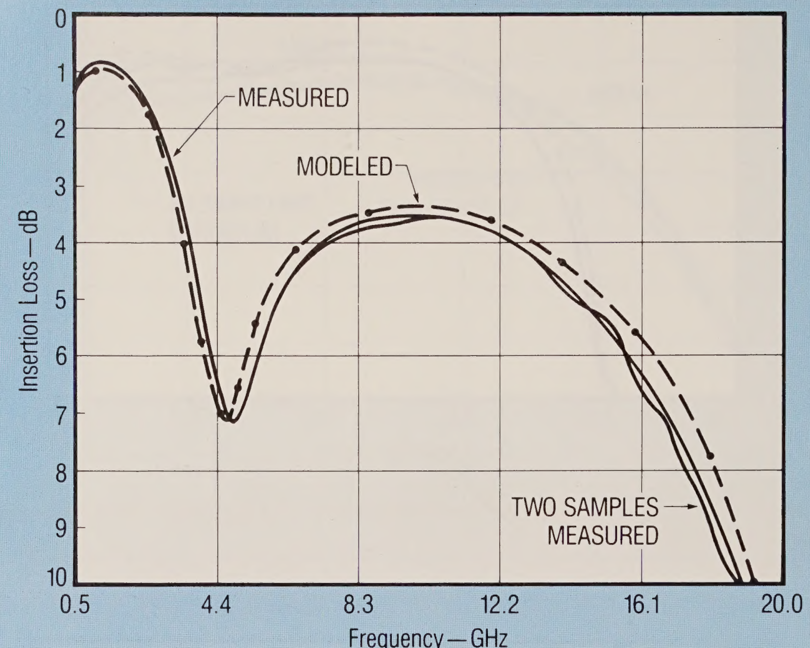

Figure 6a. Measured and modeled insertion loss for two sample power divider low-pass branches (including power division loss).

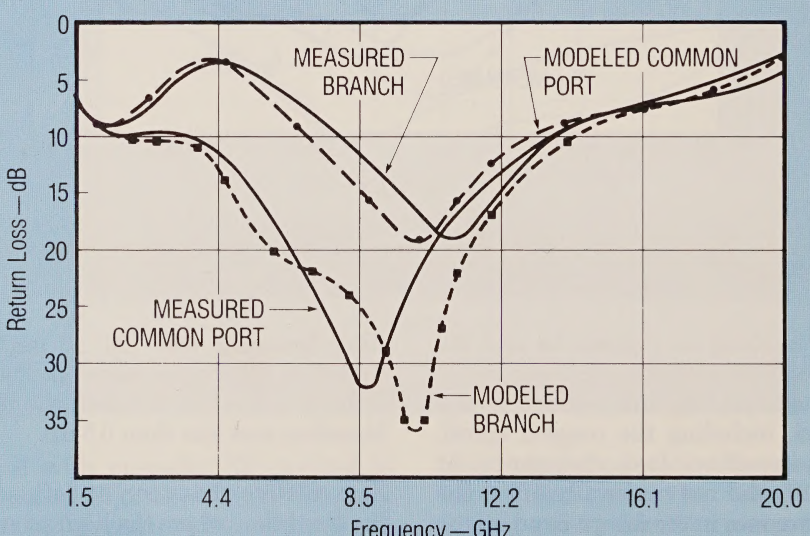

Figure 6b. Measured and modeled return loss for power divider low-pass branch.

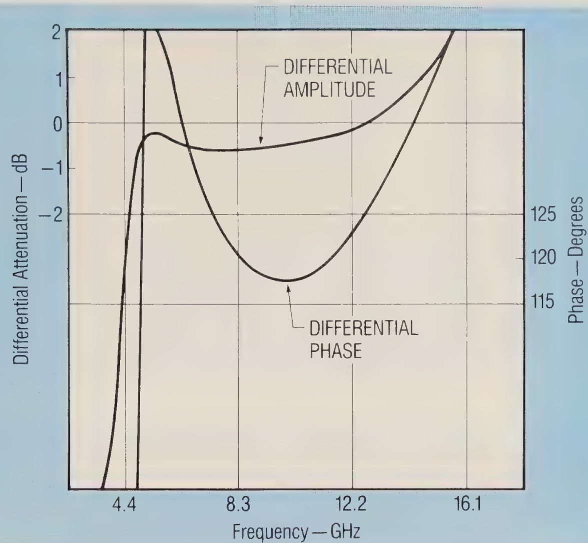

Figure 7. Differential phase and amplitude for power dividers and filters.

is advanced approximately 60 degrees. Phase advances up to a total of 120 degrees as the low-pass filter branch attenuation is increased to a maximum.

Amplitude variations of up to 4 dB were measured with this simple control scheme. Of this variation, 1.25 dB can be attributed to addition of the two fixed-phase vectors spaced 120 degrees (as previously discussed). Other factors that may increase variation are signal interference within the power dividers/combiners (because of less than ideal isolation) and effects of input and output mismatches. Unfortunately, RF wafer probing of divider isolation was not possible. Test fixture measurement of input and output return loss (which includes the effects of bond wires), was between 10 and 14 dB at 10 GHz, for all phases.

A constant amplitude can be obtained within 10-percent bandwidth segments for all phases, at selected frequencies between 8 and 12 GHz, by controlling the attenuators essentially as described by equations 1 and 2. The control voltages plotted in Figure 8 were

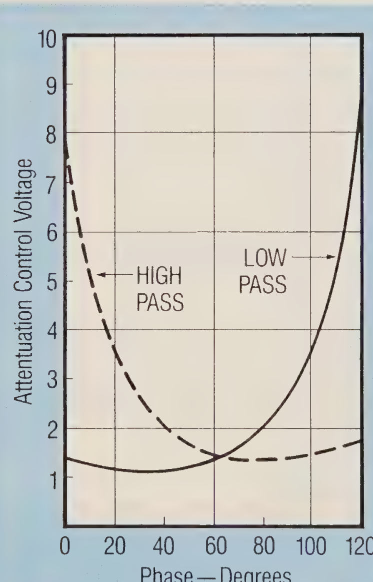

Figure 8. Attenuator control required to maintain amplitude constant as in Figure 9.

obtained by keeping the amplitude constant at 10 GHz as phase was changed over 120 degrees. These curves follow the theoretical functions shown in Figure 3 quite closely. The left side of Figure 9 shows amplitude variation obtained over the 8- to 12-GHz band as phase is changed in 5-degree increments (shown on the right side of Figure 9).

Amplitude and phase as a function of frequency at −40°C, 25°C, and +70°C are plotted in Figure 10 for constant control voltage. The plots shown were taken at a relative phase setting of 60 degrees. They represent the worst case observed over the 120-degree phase range. Amplitude variation is less than ±0.5 dB and phase variation less than ±5 degrees. Amplitude and phase variation can be further improved by finer programming of the control voltage.

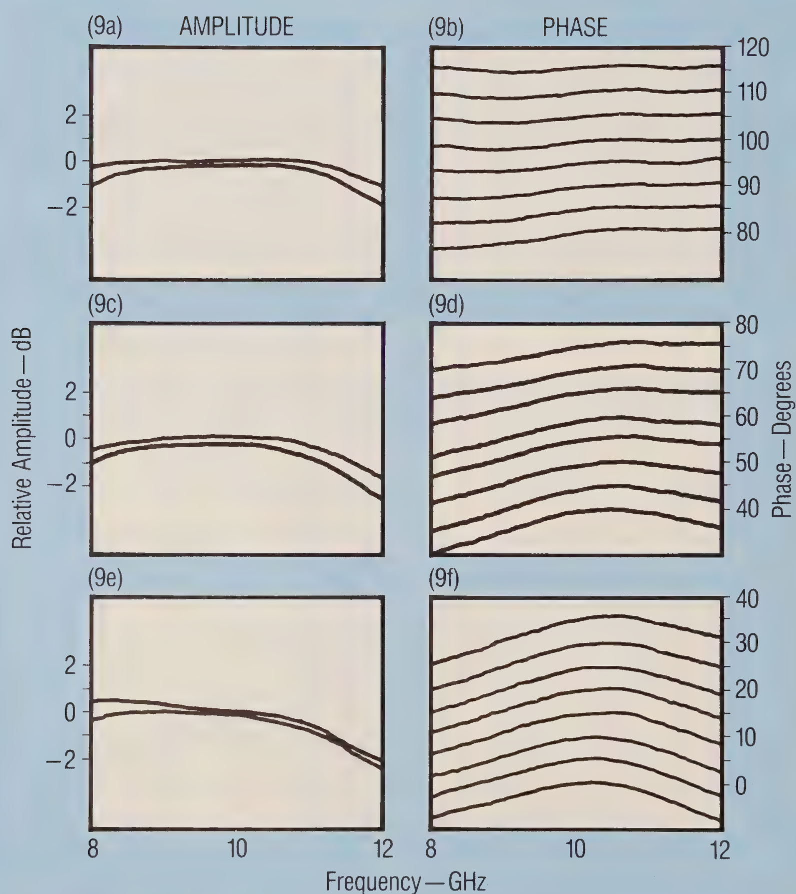

Figure 9. Amplitude held constant at 10 GHz, left side, while phase is varied in 5° increments, right side.

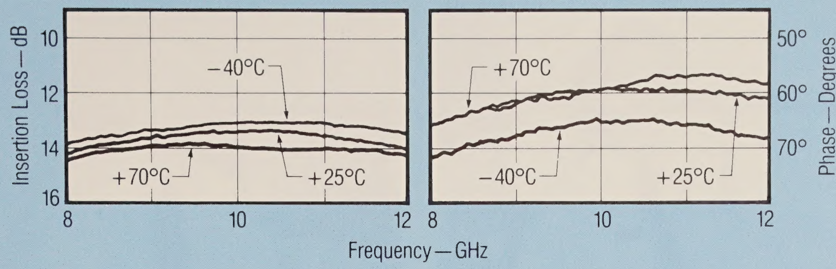

Figure 10. Insertion loss and phase at 3 temperatures.

Conclusion

Results from the development of an X-Band analog bidirectional phase shifter have been presented. A very good theoretical model for a coupled spiral inductor enabled the implementation of a VHF lumped-element "L" type power divider to a X-Band GaAs MMIC. High resolution phase and amplitude control as well as excellent temperature stability were obtained by using a unique temperature-compensated MESFET pi-attenuator for control in a vector modulator scheme. The final result was a temperature-stable phase shifter with which amplitude could be maintained constant as phase was shifted by up to 120 degrees in increments as small as one degree, and which was stable over −40°C to +70°C.

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3. Fisher, D. and D. Dobkin, "A Temperature-Compensated Linearizing Technique for MMIC Attenuators Utilizing GaAs MESFETS as Voltage-Variable Resistors," *1990 IEEE MTT-S Symposium Digest,* pp. 781-784.

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