N-BIT L-BAND PHASE SHIFTERS FOR PHASED ARRAYS

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ABSTRACT

Phase shifters required for electronic beamsteering in phased arrays are typically built with a small number of discrete bits. At the lower microwave radar frequencies, a large number of bits is desired to both reduce sidelobes and perform adaptive beamforming. Continuously variable phase shifters with infinite resolution have been built at L-band (1.3 GHz). Two phase shifters are reported on: a 360° unit which is a hybrid of all GaAs components, and a fully monolithic 90° shifter made in silicon-on-sapphire.

Introduction

Developments in both solid state microwave power devices and monolithic microwave integrated circuits (MMIC's) have increased interest in phased array radar systems which are composed of radiating elements directly driven by active transmit/receive modules. These modules typically consist of transmit and receive amplifiers, a phase shifter, transmit/receive switching, a limiter, and a circulator. Some form of receive weighting is normally employed for antenna sidelobe reduction, often in a separate manifold. Additionally, subarray or individual element weighting may be employed to adaptively place nulls in the antenna pattern for jammer and clutter rejection.

In the case of large arrays of elements, as would be employed in large ground based radars such as PAVE PAWS[1] at UHF, or in higher frequency airborne radars,[2] adequate sidelobes can be obtained by the use of 3- or 4-bit phase shifters. However, for applications such as Airborne Early Warning Radar in which frequency is low for long range propagation, and a limited aperture restricts the number of elements to ten's, it is found that sidelobe levels of -35 to -45 dB will require 6- or 7-bit phase shifters. Continuously variable, or n-bit, phase shifters would allow phase change with a resolution equal to the number of bits in the D/A used to control them.

Applications

Three applications which need n-bit phase shifters are discussed: low sidelobe arrays with few elements; adaptive nulling; and adaptive modules. Figure 1 shows the sidelobe levels attributed to random phase and amplitude errors as a function of the number of elements. The equivalent mean square phase error of a phase shifter is given by

$$\text{mean squared phase error, } \overline{\delta^2} = \frac{\pi^2}{3(2^{2B})}$$

where B is the number of bits. Since the phase shifter is usually employed to not only steer the beam but to also correct, in situ, the large phase errors[1,3] present in solid state modules, such phase arrays will typically require 6- or 7-bit phase shifters. The smallest bit of a 7-bit shifter corresponds to 2.8°, and presents very difficult tolerance problems in a discrete mechanization.

The requirement for number of bits is further increased if adaptive arrays which perform the weighting at RF are considered. Many adaptive systems have been configured with the concept in mind that beamforming would take place either at IF, or digitally with quantization

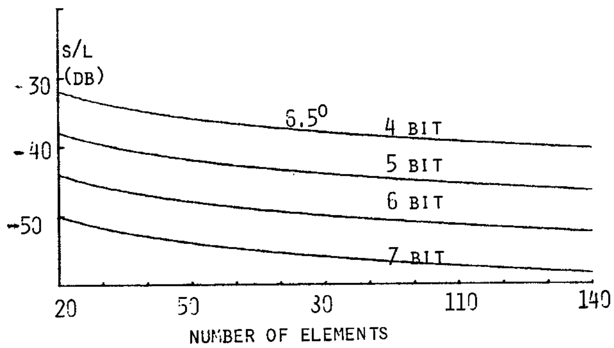

FIGURE I SIDELOBES DUE TO RANDOM ERRORS

at each element. While these approaches may be acceptable in some configurations, recent work aimed at achieving nulls in a fully adaptive UHF system[4] has shown the advantages of beamforming at RF to broaden the RF bandwidth over which nulls are achievable. Beamforming at RF avoids mismatches between IF filter bandpasses, mixer spurious products, and quantization irregularities such as accuracy match, time jitter, and number of quantization levels (Figure 2). Weights in an RF weighting system consist of phase and amplitude commands which are D/A converted in the active module to control phase and gain at the element. The study[4] has shown that although A/D's on each of the inputs to the adaptive processor may consist of only 9 bits, the D/A converter will require 11 to 14 bits depending on the jamming to noise ratio. Discrete implementation of a phase and gain control for this resolution would clearly not be feasible.

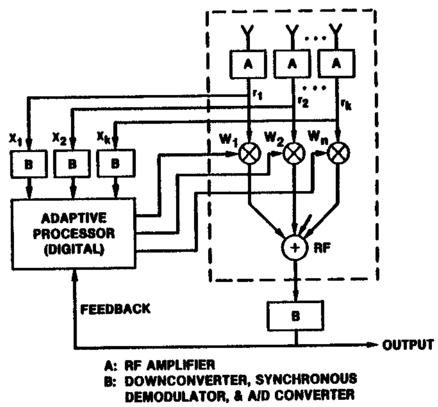

Figure 2. RF Adaptive Weighting

Adaptive modules have been suggested as a means to correct gain and phase errors in themodule, thus reducing the need for much testing and yielding a very reproducible transmit/receive element. Such a module would employ a phase and gain measurement within the module, modest computational capability, and a feedback to adaptively correct tolerances in the module. Figure 3 illustrates the concept for such a module. A continuously variable phase and gain control could improve the reproducibility of such a module to the number of bits available in the A/D.

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Phase Shifter Design, Fabrication, and Test Results

Hybrid phase shifter designs which provide continuously variable phase and amplitude control have been built at C- and Ku-band.[5,6] Under two contracts, the Naval Air Development Center is developing components which will result in a monolithic phase and gain controller at L-band (1.3 GHz). The work reported here represents the product of the first phase of these contracts. The conceptual design is the same in both efforts: 4 quadrature vectors are summed to achieve 360° of phase control. Adjacent vectors are summed pairwise in accordance with the quadrant of the phase shift desired. The amplitude of each of the quadrature vectors is determined by the phase and amplitude desired. Variable gain amplifiers for each of the vectors not desired are turned off. The specifications for overall operation are shown in Table I.

TABLE I: N-BIT PHASE SHIFTER REQUIREMENTS

Frequency (GHz)	1.3 $\pm$ 5%
Phase Shift	0° to 360°
Accuracy	$\pm$ 1° in 20 MHz
Input VSWR	$\leq$ 1.5:1
Power Handling	100 mW
Temperature Sensitivity	0.2°/°C
Gain Control	20 dB

Raytheon Company, Wayland, MA, constructed a 360° phase shifter using GaAs dual gate FET's and passive circuits built as they would be made on a monolithic substrate but connected together in a hybrid form. General Electric Company, Syracuse, NY, built a 90° phase shifter which was fully monolithic using silicon-on-sapphire (SOS) MESFET's as attenuators. Both efforts resulted in 10 phase shifters.

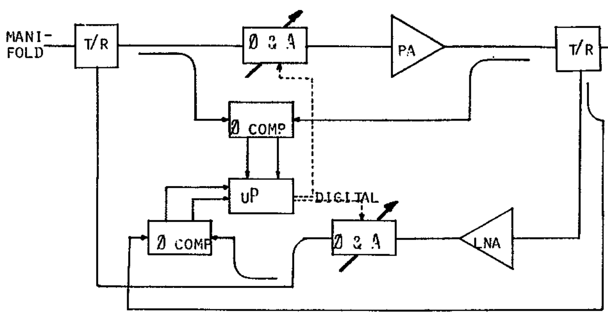

FIGURE 3 ADAPTIVE MODULE

GaAs 360° Hybrid

Figure 4 shows the detailed circuit configuration used to implement the 360° hybrid. The gain control required for each of the vectors is implemented using identical dual gate FET's. A fifth FET is used to compensate for losses in the phase shifter and to provide a variable gain control for the whole unit. The FET is an interdigitated power FET with sixteen 100-micron gate width channels. The device is fabricated on a three-layer, vapor phase, epitaxially grown structure consisting of a contact layer (0.2 um), an active layer ($n \sim 10^{17}$ cm^{-3}, $t \geq 0.4$ um), and a low doped buffer layer ($n < 10^{14}$ cm^{-3}, $t \sim 2$ um). The gate fingers are nominally 1-micron long and are separated by nominal 2-micron spacing. Gate metallization is Ti/Pt/Au. The device provides a +27 dBm output at 1 dB gain compression, and a small signal gain of over 18 dB. The amplifier circuit was synthesized using inductive-tee matching networks, and fabricated on an 18-mil thick sapphire substrate. The amplifier has > 45 dB gain control. The 20 dB of gain control at the high end of the range exhibits less than 2° phase shift, thus

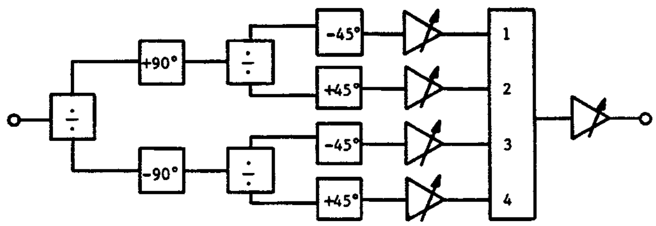

Figure 4 Lumped Element Quadrature Shifter

allowing the final (gain control) amplifier to be added (Figure 5) with minimum amounts of compensation required. At lower gains the phase change with gate voltage reaches 30°/V.

The power dividers and combiners were also fabricated on sapphire substrates. The goal of monolithic design at L-band led to a lumped element circuit approach with negative, high pass and positive, low pass insertion phase networks. The four-way input power divider is constructed as a two-way divider followed by $\pm$ 90° insertion phase networks, with each output followed by $\pm$ 45° insertion phase networks. Though the insertion phase of each network varies with frequency, by using a negative frequency the phase slopes can be designed to be equal over reasonable bandwidths. Differential losses of 0.3 dB between ports were achieved with differential phase changes over a 10% bandwidth of less than 2°. The insertion loss was 2.5 dB worst case. Figure 5 shows the completed and packaged phase shifter. Eleven substrates are mounted in a brass housing measuring 1.69" x 0.85" x 0.312".

Data was taken using an automatic network analyzer to demonstrate the ability to reach any phase angle from 0° to 360° with the constraint of constant output. While it might seem that the limit of dynamic range in each FET should limit the resolution attainable at settings where one vector predominates, the change of phase in the FET amplifiers at low gain states means that in fact resolution will be limited to D/A resolution. Figure 6 shows the variation of phase over a 40-MHz bandwidth for a given voltage setting and the gain variation with respect to a nominal 3.8 dB gain. Experimentation with varying the input power indicated RF biasing effects occurring at powers well below saturation which may limit application of these devices in an open loop configuration. Variations to 6° over the input power range from 0 to + 10 dBm were observed.

SOS Monolithic 90° Phase Shifter

A fully monolithic phase shifter covering a single quadrant was built using a 17-mil SOS substrate. In this case, MESFET's were used as attenuators in the two outputs of a quadrature hybrid. Figure 7 shows the monolithic circuit consisting of an input split into quadrature components by an interdigitated Lange-type hybrid using air bridge crossovers. To minimize use of substrate, it was necessary to fold the interdigitated lines. The phase relationship between the outputs was 90° $\pm$ 3°, and loss across a 1-2 GHz band was less than 1.3 dB. At each output of the hybrid are 5000-um MESFET's shunt mounted to ground. Attenuation is controlled by gate bias, but was limited to about a 12-dB range. Dual gate FET's were not used due to the difficulty of manufacture in silicon and the desire to implement monolithic phase shifters in the less expensive and proven silicon technology if possible. With reference to Figure 4, it can be seen that if the leakage in one vector is significant, it will not be possible to reach a full 90° range. To compensate, an extra line length of about 30° was added. The resulting vectors are combined in a Wilkinson combiner.

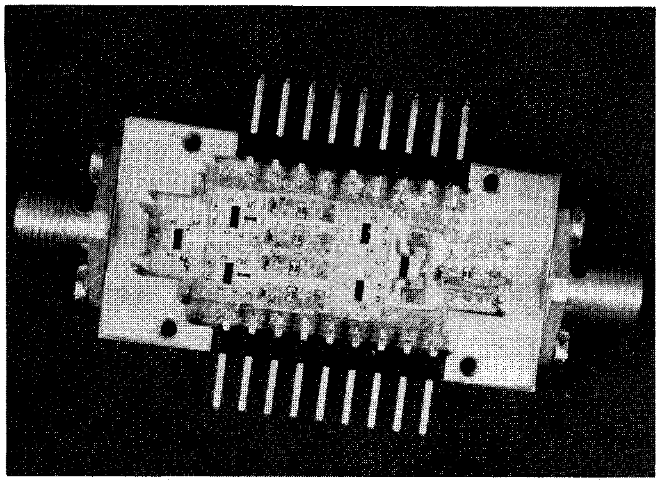

Figure 5 360° Phase Shifter

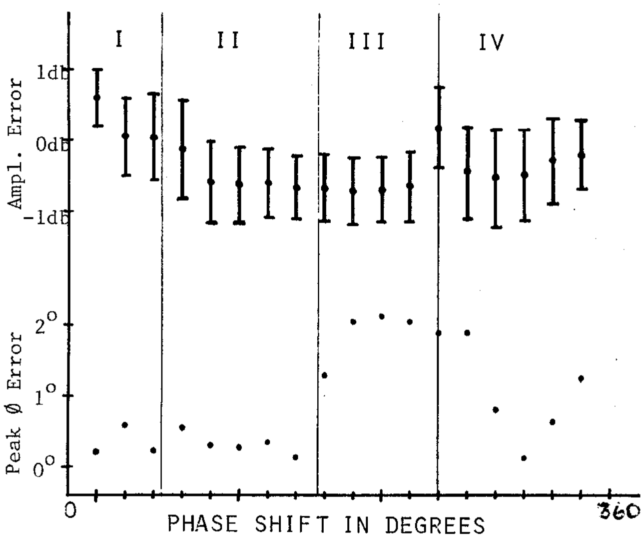

Figure 6 Phase and Gain Errors

Figure 8 shows the results for one of the shifters plotted against a normalized control parameter which linearizes the phase. Insertion loss was high and varied more widely than expected. Wilkinson implementation was discovered to contain a 1400-ohm resistor rather than a 100-ohm resistor due to process errors. Futhermore, the > 20° variation in insertion phase of the FET causes excessive loss variation when combined with the small range of attenuation adjustments. If reproducible devices can be made, the balanced configuration of Figure 4 will tend to reduce this effect.

Conclusions

Continuously variable phase shifters have been implemented in both GaAs and SOS for L-band operation. Resolution obtainable would appear to exceed requirements, but repeatability and AM/PM conversion effects cast doubt on the feasibility of using these circuits in the most stringent applications without some form of feedback. Implementation studies have started to consider what problems might be encountered in obtaining 360° phase shift with discrete 180° and 90° bits to replace three quadrants of the vector diagram of Figure 4.

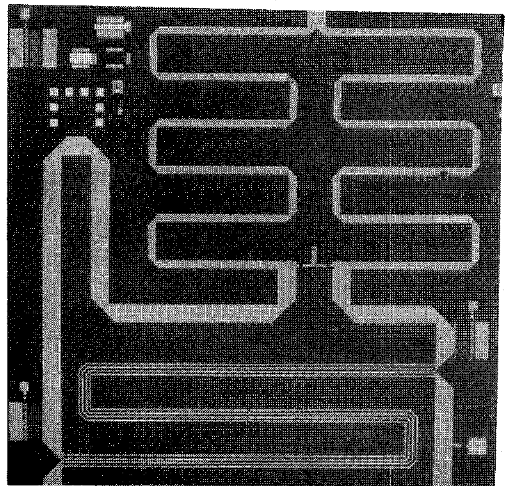

Figure 7 90° Monolithic Phase Shifter

Acknowledgement

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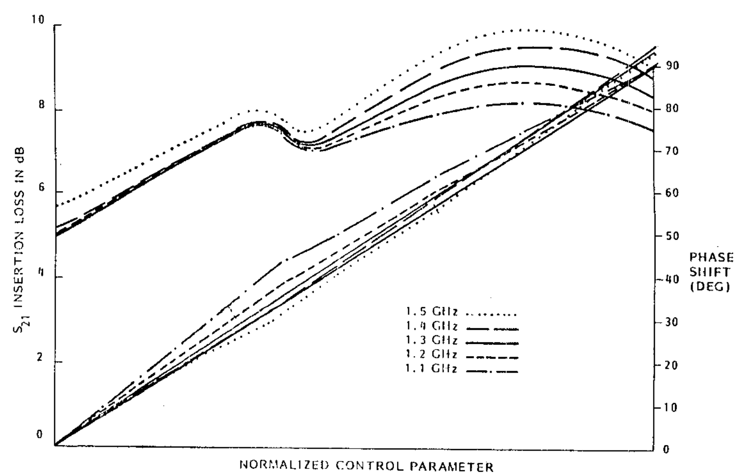

Figure 8 Insertion Loss and Phase Shift