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Short Papers

A Novel Design of an X-Band High-Power Ferrite Phase Shifter

KARL H. HERING

***Abstract*—A nonreciprocal analog-latching phase-shifter design at X-band is described. The device has been operated up to 250 W of average power and 50 kW of peak power. Direct liquid cooling is implemented, pumping FC-78,[1] a dielectric coolant, through the center slot of the ferrite toroid.**

Introduction

A novel design of an X-band ferrite phase shifter is presented. The device is designed to handle high power levels—up to 250 W average and 50 kW peak. The phase shifter is of the nonreciprocal analog latching—a name given to the type of digital phase shifters that consist of one long toroidal bar. In this type of device the phase shift is controlled by the driver signal level, as opposed to the conventional digital ferrite phase shifter, which consists of different lengths of ferrite toroids, each driven into saturation. Internal liquid cooling is used to achieve stable performance of the device even at the high average power levels.

The phase shifter required a thermal design of the configuration, selection of the most applicable ferrite material for high-peak and average-power handling, and a method of implementing cooling of the ferrite bar.

Thermal Design Aspects

Applicable methods of cooling the phase-shifter material are: 1) a cold plate [1], 2) forced-air cooling, 3) conduction cooling using boron nitride in contact with the ferrite [2], or 4) direct liquid cooling, as was chosen for this novel phase-shifter design. Other techniques are limited in either cooling potential or for mechanical reasons. Cooling, using a coldplate, has been used successfully in high-power ferrite phase-shifter designs [1], [2]. In doing so, however, the cross section of the phase shifter becomes large. In many phased-array applications very little space is available between the radiating-element/phase-shifter assemblies. Cooling with boron nitride or similar dielectric material of high thermal conductivity reduces phase shift per unit length and provides only limited cooling capability. Forced-air cooling is simple, but provides very limited cooling and precludes that openings exist for the passage of air. Analog phase-shifter designs have shown direct liquid cooling to be far superior to the other techniques. However, the methods used for

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[1] Trademark of Minnesota Mining and Manufacturing.

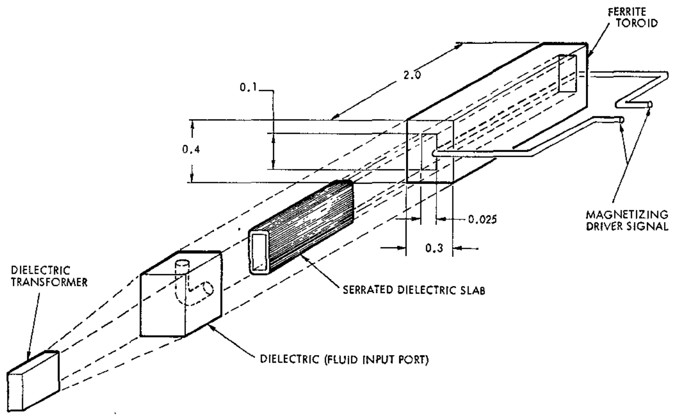

Fig. 1. Direct liquid-cooled X-band analog-latching ferrite phase shifter.

S- and C-band phase shifters [3] are not applicable at X band. In an earlier design for an analog ferrite phase shifter, the bar was encapsulated in a Teflon tube and FC-78[1] was pumped through the device. At X band and for a digital ferrite phase shifter this technique is not readily applicable, because of the small size of the ferrite toroid and the drive wire to be brought out through the Teflon tube.

The technique devised and discussed here is shown in Fig. 1. The FC-78 liquid is pumped through inlet/outlet ports into and through the inside of the ferrite toroid. The parts are epoxy bonded to seal the assembly. The inner slot of the ferrite toroid is filled with a serrated dielectric slab, which is a compromise design to allow the liquid to flow through the toroid and to maximize the phase shift. The requirement for the direct liquid-cooling design is determined by calculating the heat load based on the maximum incident RF average power, the insertion loss of the ferrite materials, and the dimensions of the ferrite toroid. The flow rate and the pressure are given by:

$$F_v = \frac{Q}{64.6\ \Delta t} = \text{flow rate (gal/min)}$$

where Q equals the heat load in watts and Δt the temperature rise in degrees, and

$$P = \frac{2.16 \times 10^{-4} f L \rho Q^2}{d^5} = \text{pressure (lbf/in}^2\text{)}$$

where f is the frictional coefficient, ρ is the density of FC-78, L is the length of the toroid, and d is the equivalent diameter of the coolant path through the toroid. The dielectric coolant FC-78 was chosen because of its good thermal and RF characteristics [3].

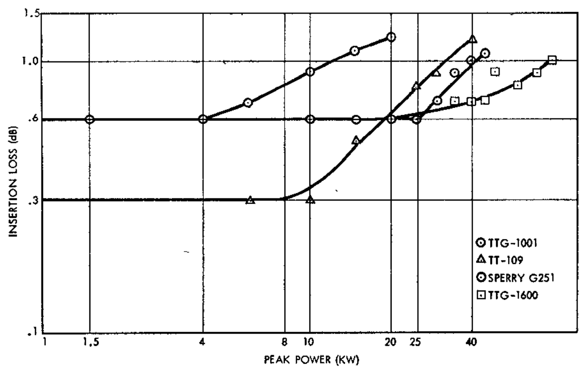

Fig. 2. Peak-power handling of several ferrimagnetic materials.

TABLE I
FIGURE OF MERIT
(Deg/dB)

Ferrimagnetic Material	Dielectric Constants of Material in the Center Slot			
	$K=1$	$K=15$	$K=20$	$K=24$
TTG-1600	323	487	334	675
TT-109	421	597	327	
TTG-1001	323	425	314	

The design of the coolant path represents a compromise between electrical and thermal design objectives. The flow path must provide sufficient surface area to be in contact with as much of the ferrite surface as possible. However, to obtain maximum phase shift per unit length the slot should be filled with high dielectric constant material. A design study was made of serrated dielectric slabs with different dielectric constants to maximize the figure of merit (phase shift per insertion loss). Results are shown in Table I. The figure of merit (degrees/decibel) is given for three ferrimagnetic materials that were evaluated in detail. The variation of the figure of merit versus dielectric constant is due to higher losses in the dielectric of $K=20$.

The coolant is brought into and out of the ferrite toroid through dielectric blocks with right-angle openings, as shown in Fig. 1. The blocks are made from high dielectric constant material similar to the dielectric properties of the ferrite. Conventional RF design [2] of dielectric transformers is used to match the assembly and obtain a low VSWR over the 8.5–9.5 GHz band.

Several ferrite materials were investigated as to their peak power-handling capability. At the same time, the variation of the magnetization versus temperature was considered in the selection. Such information is readily available from the manufacturer, but can also be evaluated using a hysteresis loop tracer and a thermally controlled environment. The garnet materials, such as TTG-1001[2] and TTG-1660, are best suited for the development of a phase shifter handling high average power. Data of peak-power handling are presented in Fig. 2. The expansion resulting from an increase in temperature of ferrimagnetic material, which is restricted by a waveguide, creates magnetostriction effects. Fig. 3 shows the variation of B_r/B_{r0} as a function of pressure (lbf/in^2). The ratio B_r/B_{r0} is a measure of the percentage change of the phase shift due to pressure where B_r and B_{r0} equal the remanence induction with and without pressure on the toroid. TTG-1600 is seen to be less sensitive to magnetostriction than TTG-101. Tests on ferrite materials such as TT1-414 and TT1-109 showed no magnetostriction, as has been reported elsewhere [5]. However, ferrite materials have lower handling capability of peak and average power. On the basis of the characteristics of low magnetostriction, good figure of merit, and excellent high-power handling, TTG-1600 was selected for the toroid material in the final model.

[2] TT stands for Trans-Tech Inc.

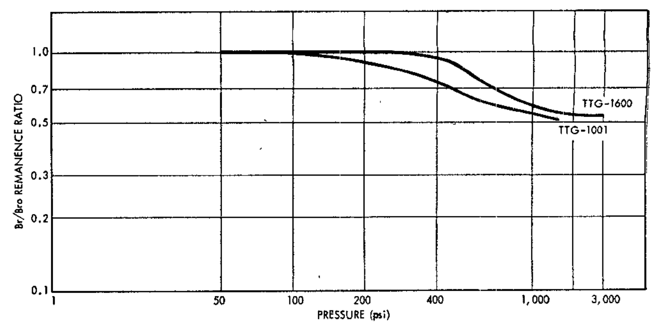

Fig. 3. Change in remanence induction with pressure.

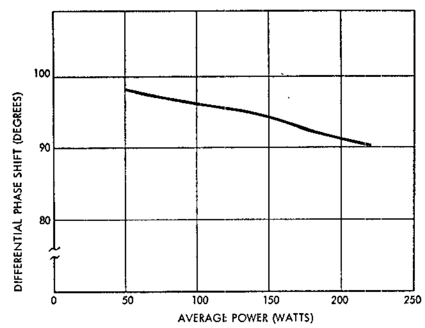

Fig. 4. Differential phase shift versus RF input power.

TABLE II
X-BAND PHASE-SHIFTER PERFORMANCE

Material	TTG-1600
Frequency	8.5–9.5 GHz
VSWR	<1.35:1
Average power	250 W
Peak power	50 kW
Figure of merit	450 deg/dB
Bar dimensions	0.300×2.00×0.400 in (width by length by height)
Slot width	0.025 in wide
Dielectric in the slot	$K=24$ serrated
Cooling transformer	$K=15$ 0.300×0.250×0.400 in (width by length by height)
Matching transformer	$K=4$ 0.250×0.200×0.400 in (width by length by height)
Weight	0.1 lb (ferrite bar/waveguide, no flanges)

When the phase-shifter design was satisfactorily completed at low power, the device was fully assembled for high average-power test. Results are shown in Fig. 4. The device was operated up to 250-W average power successfully. The data of Fig. 4 were taken at the center frequency of the band at 9 GHz. No effort was made to achieve differential phase shift independent of frequency. This feasibility has been demonstrated elsewhere [6], [7].

PERFORMANCE SUMMARY

The *X*-band phase shifter has the characteristics summarized in Table II. The device has good performance characteristics. Such a small compact device finds application in high-power radar and communication systems where fast beam steering on the order of 10 μs is required.

ACKNOWLEDGMENT

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Wide-Band Phase Locking and Phase Shifting Using Feedback Control of Oscillators

DAVID RUBIN

***Abstract*—Two YIG-tuned Gunn oscillators[1] have been frequency locked over a 3-GHz range in X-band. Injected power into the locked oscillator was approximately 10 dB below its output power. Analog frequency-tracking circuitry was used together with phase-comparator feedback to achieve output phases which remained within $\pm 10°$ over a 1.2 GHz bandwidth. Controlled phase shifting was obtained by applying dc voltages within the feedback loop.**

In order for oscillators to phase track one another over large frequency ranges, it is first necessary that their free-running frequencies be brought to within the appropriate range [1]

$$\Delta f \leq \frac{f_o}{2Q_o}\sqrt{P_i/P_o} \tag{1}$$

where $\Delta f = |f_i - f_o|$ is the frequency difference between the master and slave oscillators without signal injection, P_i is the power injected into the slave oscillator cavity, P_o is the slave oscillator output power, and Q_o is the loaded Q of the slave cavity. The above formula holds for $P_i \ll P_o$. Even with oscillators and magnetic structures built to extremely close mechanical tolerances, with the same current through the two coils their free-running frequencies may differ as much as 200 MHz and have linearity variations of ± 15 MHz throughout the X-band region. Frequency-locking measurements at 10 GHz, wherein the locking range of the slave oscillator is measured as a function of the P_i/P_o ratio, gave a Q_o of 450, indicating that the slave frequency must be held to within 3.5 MHz of the master frequency for locking to occur with $P_i/P_o = 0.1$.

In addition to the main frequency-controlling current, which was series connected to both YIG coils,[1] a small additional current was added to one of the coils. Without injection locking, the latter current was varied until the two oscillator frequencies were approximately the same, as noted on a spectrum analyzer. This additional current was noted at incremental frequencies throughout X band, first with frequency increasing, then with frequency decreasing. The currents were different for the two cases due to hysteresis of the magnetic circuit. Fig. 1 shows the additional currents needed for frequency tracking, and also shows a current synthesized by the circuit of Fig. 2.

For the frequency-tracking circuit, one connection serves as both input and output. The input voltage is taken from the slave-oscillator YIG coil and rises linearly with frequency. Operational amplifiers 1 and 2 serve as voltage follower and inverter, respectively, and do not load the YIG coils. At frequencies below 7.5 GHz, all diodes are held below breakdown, and a constant current enters operational amplifier number 3 through a resistor connected to $-V_s$. As the frequency is raised, each diode in turn breaks down, first removing, then adding current to the input of operational amplifier number 3. The input voltage at which a diode breaks down and the amount of current conducted through a diode, is dependent on the two resistors connected to it. The current entering operational amplifier number 3, and hence its output voltage, is thereby made nonlinear with respect to the input frequency.

Operational amplifier number 4 is called a Howland pump [2]. It delivers an output current proportional to the voltage input, i.e., even though the load itself is a resistance in series with a voltage source. The delivered current is shown as the solid line in Fig. 1. With this circuit the YIG oscillators could track to 2 GHz. Also note that a similar method could be used for tracking varactor-tuned oscillators, wherein the input would be a voltage proportional to frequency, and the nonlinear output voltage would go to a varactor diode instead of being returned to the input. In the latter case, the Howland current pump would be unnecessary.

Fig. 3 shows a block diagram including the feedback method used. The phase comparator used is a 3-dB quadrature coupler with input and normally isolated terminals connected to the input RF sources and output terminals connected to diode detectors. The detected voltage difference is proportional to the sine of the input phase difference. This provides the input for a dc amplifier and pump circuit. This in turn feeds current to the slave-oscillator YIG coil, which tends to adjust its free-running frequency until the inputs to the phase comparator are in phase. If a dc reference voltage E_Φ is connected to one of the terminals of the latter differential amplifier, the feedback circuitry will adjust the current until the output voltage of the phase comparator provides approximately the same voltage to the other terminal of the difference amplifier.

Hence one could adjust the gain of the pump circuit so that if 1 V at reference terminal E_Φ represented $+90°$ phase shift, one could set a reference voltage E_Φ' so that $\sin \Phi = E_\Phi'$ and get a phase range that in practice will be less than $\pm 90°$. For accuracy over wide bandwidths, power levelers should be used so that P_i/P_o is constant. For the experiments to be described the power output of both YIG oscillators varied similarly over the band so that $P_i/P_o = 0.1$ over most of the frequency range.

With feedback the locking range at any given frequency is increased [3]. Likewise a given phase change due to the differences in the free-running frequencies is reduced. This can be seen from the following argument.

Let $\Delta\phi$ and $\Delta\phi'$ be the phase differences between YIG oscillator outputs with and without feedback as seen at the input terminals of the phase comparator. Corresponding to $\Delta\phi$ and $\Delta\phi'$ are the free-running slave-oscillator frequencies f_o and f_o'. These frequencies are linearly dependent on the YIG coil current, so that $f_o = c_1 I_o + c_2$ and $f_o' = c_1(I_o + I_f) + c_2$, where I_f is the current fed back from the phase comparator circuit. The latter is proportional to the sine of the phase difference, i.e., $I_f = c_3 \sin \Delta\phi' \approx c_3 \Delta\phi'$ for small phase differences. The Adler formula [1] becomes, for small angles,

$$\Delta\phi \approx c_4(f_o - f_i) \tag{2}$$

$$\Delta\phi' \approx c_4(f_o' - f_i) \tag{3}$$

where

$$c_4 \approx \frac{2Q_o}{f_i}\sqrt{\frac{P_o}{P_i}} \tag{4}$$

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[1] Watkins-Johnson Model 5008-2.