High-Power Microwave Transmitter Switch

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A technique for high-speed switching or pulse-modulating the X-band radar transmitter is described in this article. Additional equipment is not needed, and mechanical switches and a high-power modulator are eliminated. Conventional modulation schemes usually result in early failure of the klystron amplifier tubes from thermal stresses; this approach does not have this limitation.

I. Introduction and Summary of Conclusions

A technique for high-speed switching or pulse-modulating the X-band radar transmitter is described in this article. Additional equipment is not needed and mechanical switches and a high-power modulator are eliminated. Conventional modulation schemes usually result in early failure of the klystron amplifier tubes from thermal stresses; this approach does not have this limitation. The conclusions have been verified by tests and the phase stability has been measured (Ref. 1).

II. Modulation Requirements and Associated Problems

Initially, the transmitter will be used for radar ranging of Jovian moons and the rings of Saturn. The experimenter has chosen the following mode of operation. The transmitter will be pulsed on and off, 50% duty cycle, a pulse repetition rate (PRR) between 1 and 30 seconds. This will be continued for one round trip time of flight of the radio signal to the target and back (30 minutes to 2 hours); the transmitter will be switched off and the receiver switched on for one round trip time of flight. This cycle will be continued as long as the target is in view.

During the receive cycle, the receiver will detect the reflected signal plus noise $(S + N)$ alternately with noise only (N), corresponding to the transmitter on-off modulation. The information can then be processed by a computer:

$$(S + N) - N \approx S$$

The scheme is analogous to conventional (Dicke) radiometric techniques.

Long pulse modulation of conventional klystron amplifier tubes subjects them to severe thermal stresses and greatly reduces their operating life. The tube efficiency is

$$100 \times \frac{\text{RF power output}}{\text{DC power in}} \leq 40\%$$

Thus for 200-kW output per tube, the dc input power is 500 kW. With drive off (no RF input to the tube) all of the dc power is dissipated in the collector or anode as heat. With drive on, 200 kW is transmitted through the output window as RF and 300 kW is dissipated as heat in the collector. The collector is a thick-walled forged copper cylinder, approximately 60 cm long and 3 cm thick, with an ID of 10 cm. It has drilled coolant passages to remove the heat. When the drive is off, the electron beam does not disperse uniformly and the maximum power densities are very high in certain areas. Modulation techniques such as switching the drive on and off produce 200-kW power pulses in the collector from a 300-kW base, while switching the dc beam power on and off produces 300-kW collector power pulses. Although the collector is efficiently cooled for CW operation, the interior surface temperature is quite high and has a thermal lag due to gradients through the thick copper. Long pulse modulation can be expected to produce rapid grain growth in the copper and cracks and spalling with ultimate tube failure due to the collector cracking open or copper particles falling into the electron beam and arcing (Ref. 2). This failure mode was also experienced during the first Mars radar experiment with the 100-kW S-band klystron and continued until pulse modulation was discontinued.

Should we elect to proceed with this approach in the face of a warning to expect early tube failure, a high-power modulator must be designed and the high-voltage power supply redesigned. Power supply regulation is obtained by feedback control of the field of the ac generator, which supplies power to the high-voltage transformer. The generator field time constant is the limiting factor and will produce poor rise time and severe overshoots in the dc power pulses.

III. Description of Combining Technique

Economic considerations dictated use of two 250-kW klystrons (Ref. 3) of an existing design rather than sponsorship of the development of a 400-kW unit. Technical feasibility of the latter has been demonstrated by the 1-MW, 8-GHz klystron developed by Varian Associates (Ref. 4), but an operational tube has not been produced.

The outputs of the two klystrons will be combined into a single waveguide by use of a four-port, 90-deg hybrid junction (Ref. 5). An oversimplified description of the combiner is two parallel waveguides connected by a coupling slot or opening in the adjacent walls of the two waveguides (Fig. 1). The design applies for either broad or narrow wall coupling. We have chosen the latter since it simplifies the waveguide configuration for this particular installation.

If a signal is applied to port 1, a signal of one-half the power will appear at port 3, and one-half will appear at port 4, advanced 90 deg in phase relative to the signal at port 3.

If a second signal of the same power is applied at port 2, -90 deg out of phase with the signal at port 1, the two combine at port 3 with a total power almost equal to the sum of the ports 1 and 2 signals. A small component of output power at port 4 (typically 30 dB less than the power level at 3) is a function of the unbalance in the hybrid.

By changing the phase of the port 2 signal to $+90$ deg with respect to the port 1 signal, the two combine so that most of the power output is at port 4 and the small residue at port 3. Thus a ± 90-deg biphase change in one input signal with respect to the other transfers the signal from one output port to the other. The ratio of the residual power to the combined power (switching isolation) for the ± 90-deg biphase change is shown in Fig. 2 for a typical hybrid as a function of relative gain and phase variations of the two input signals. A mathematical analysis of the hybrid combiner performance is presented in Appendix A.

IV. Description of Transmitter

A functional block diagram of the transmitter as originally planned (Ref. 6) is shown in Fig. 3. The waster load is normally required to absorb only the small component of output power due to the combiner hybrid unbalance. During the initial phasing of the klystrons, the waster load power can be greater, so a 150-kW RF water load was selected. No load with a higher power rating is available. It was planned to employ two such loads, with a power divider for each power amplifier and four loads and three power dividers to absorb the total output power of 400 kW.

To accommodate the new concept of phase-shift switching of the output power, the waster load is required to absorb 400 kW during the interpulse period. The configuration shown in Fig. 4 uses the power amplifier loads with an additional power divider to terminate the combiner instead of the smaller waster load.

References

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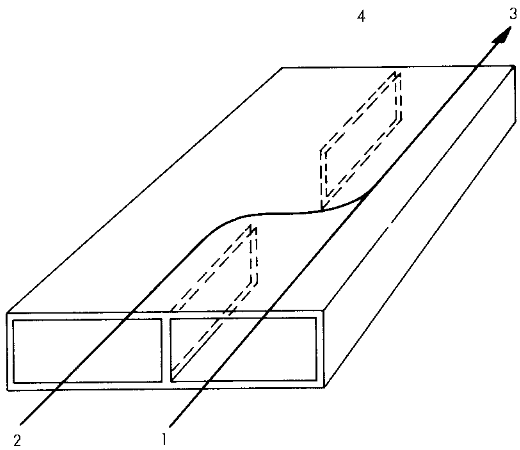


Fig. 1. Simplified diagram of combiner

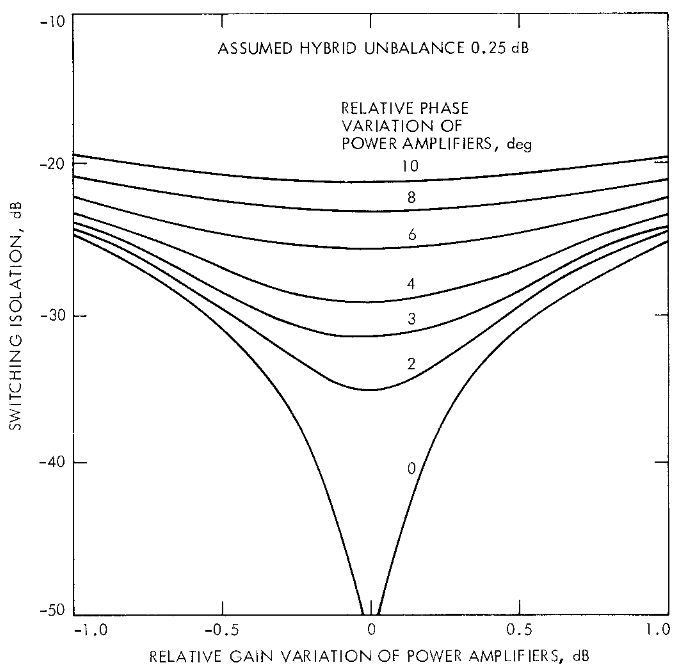


Fig. 2. Hybrid combiner switching isolation (computed)

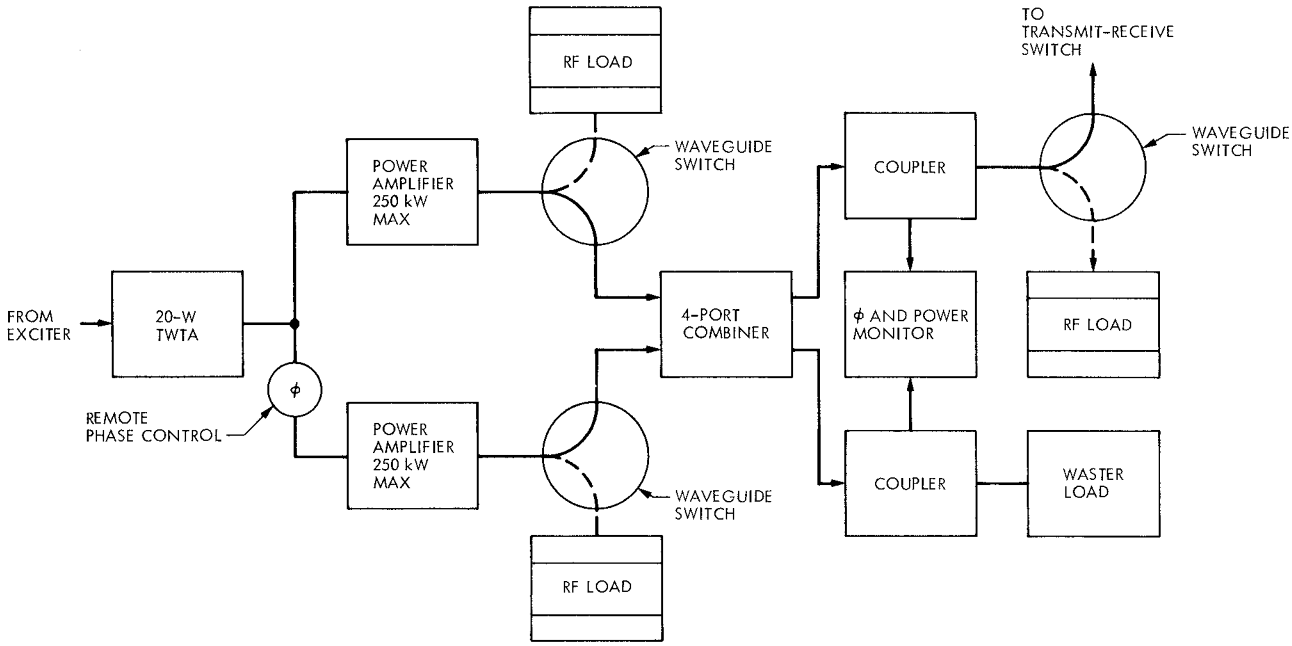


Fig. 3. 400-kW X-band radar transmitter

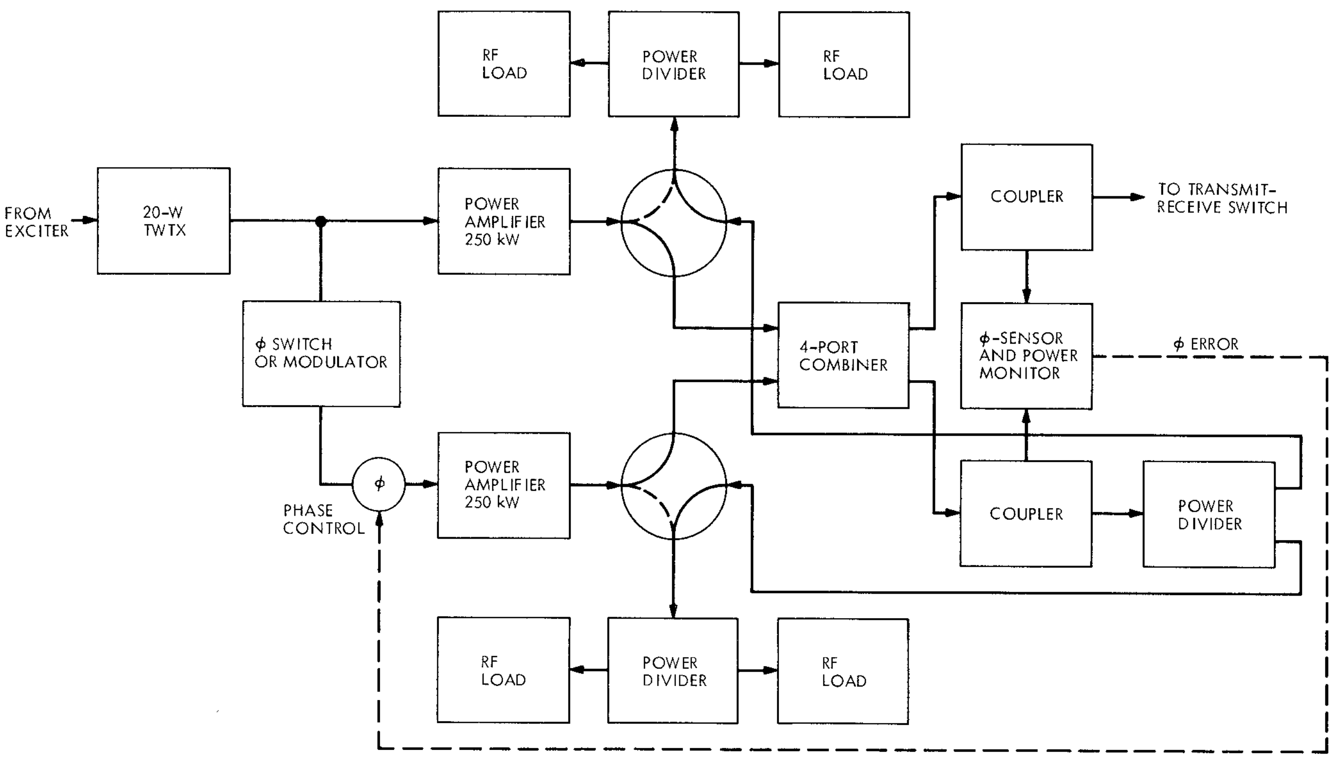


Fig. 4. RF power switching using phase shifting technique

Appendix

The performance of a hybrid combiner used as a high-power switch can be analyzed by employing the scattering matrix of the four-port hybrid. The hybrid, shown schematically in Fig. A-1, is assumed to be lossless, and its scattering matrix is therefore unitary. The matrix elements can be represented as given by Eq. (A-1), assuming symmetry.

$$S = \begin{bmatrix} A_1\epsilon^{j\phi_1} & A_2\epsilon^{j\phi_2} & A_3\epsilon^{j\phi_3} & A_4\epsilon^{j\phi_4} \\ A_2\epsilon^{j\phi_2} & A_1\epsilon^{j\phi_1} & A_4\epsilon^{j\phi_4} & A_3\epsilon^{j\phi_3} \\ A_3\epsilon^{j\phi_3} & A_4\epsilon^{j\phi_4} & A_1\epsilon^{j\phi_1} & A_2\epsilon^{j\phi_2} \\ A_4\epsilon^{j\phi_4} & A_3\epsilon^{j\phi_3} & A_2\epsilon^{j\phi_2} & A_1\epsilon^{j\phi_1} \end{bmatrix} \tag{A-1}$$

The A elements are real. The following bounds may be stated for a practical hybrid following the analysis in Ref. 5.

For a minimum isolation of 30 dB between ports 1 and 2,

$$A_1 \cong A_2 \leqq 0.0316 \tag{A-2}$$

For a maximum unbalance of 0.25 dB between ports 3 and 4,

$$A_3 = KA_4 \tag{A-3}$$

where $0.972 \leqq K \leqq 1.09$, and

$$\phi_3 = \phi_4 - (90^\circ + \delta) \tag{A-4}$$

where $-0.1^\circ \leqq \delta \leqq 0.1^\circ$

By proper assignment of reference planes, ϕ_3 may be set to 0. Hence,

$$\phi_3 = 0, \phi_4 = 90^\circ + \delta \tag{A-5}$$

Since the hybrid's matrix is unitary,

$$A_1^2 + A_2^2 + (KA_4)^2 + A_4^2 = 1$$

or

$$A_4 = \sqrt{\frac{1 - (A_1)^2 - (A_2)^2}{1 + K^2}} \approx \sqrt{\frac{1}{1 + K^2}} \tag{A-6}$$

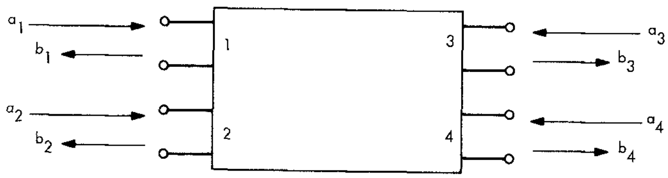
The hybrid is utilized with two power amplifiers in the circuit given in Fig. A-2. It is assumed that an initial gain and phase adjustment has been made to power amplifier 2 which sets its output to $KE_1\epsilon^{-j\delta}$. In the water load position (+90 deg), this adjustment nulls the signal at port 3 for the gain G and phase θ variable parameters equal to unity and zero, respectively. The voltage scattered from the four ports is described in Fig. A-2 as functions of the relative gain G and phase θ variations and the hybrid parameters discussed above.

The subject of particular interest is the ratio of power impressed on the antenna for the two biphase conditions. This switching isolation ratio R is expressed by

$$R = \frac{b_3 b_3^* (+90^\circ)}{b_3 b_3^* (-90^\circ)}$$

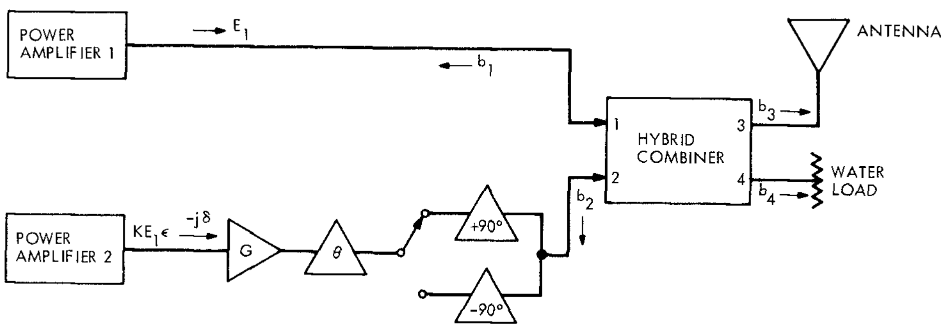
or

$$R = \frac{1 + G^2 - 2G\cos\theta}{1 + G^2 + 2G\cos\theta} \tag{A-7}$$



$$[b] = [S] \times [a]$$

Fig. A-1. Four port hybrid



$$b_1 = E_1 A_1 \left[\epsilon^{j\phi_1} \pm jKG\epsilon^{j(\phi_2 - \delta + \theta)}\right]$$

$$b_2 = E_1 A_1 \left[\epsilon^{j\phi_2} \pm jKG\epsilon^{j(\phi_1 - \delta + \theta)}\right]$$

$$b_3 = E_1 \frac{K}{\sqrt{1 + K^2}} \left[1 \mp G\epsilon^{j\theta}\right]$$

$$b_4 = jE_1 \frac{1}{\sqrt{1 + K^2}} \left[1 \pm K^2 G\epsilon^{j(\theta - 2\delta)}\right]$$

Fig. A-2. Combiner circuit