A TEN MEGAWATT RADAR DUPLEXER

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The high power duplexer array illustrated in Figure 1 consists of two balanced hybrid duplexers. Each duplexer is rated at 5 megawatts. The high power transmission is divided equally into two paths. Each path furnishes power to half an orthogonal-feed radar antenna which radiates 10 megawatts in a circular pattern. During high power transmission, the diode driver simultaneously supply forward bias to all T-R diodes. Since the conducting diodes offer a short-circuit across the collinear ports of the high power hybrids, the hybrids are able to route most of the power into the antenna. The leakage signals which escape the T-R short-circuits are terminated harmlessly in the dummy loads which are located on the low power hybrids. Total high power isolation is the sum of the T-R and hybrid isolations. Total high power transmission loss is the sum of the hybrid line and T-R short-circuit losses for each parallel path.

During reception, the driver units simultaneously supply a reverse bias voltage to all T-R diodes. The diodes become a high-impedance parallel R-C circuit across the coaxial transmission path leading toward the receiver. During this interval, the transmitter is quiescent. The low level antenna signal is divided equally in each high power hybrid, passed through the T-R mounts and then recombined in the low level hybrids. The receive signals can be delivered either to separate receivers as shown or can be combined once more for single channel operation. For protection of the receiver against excessive power due to fault leakage or nearby high power interference, low level limiters are included in the receiver lines. The total receive loss includes the sum of the losses due to the high level hybrid, T-R mount, low level hybrid, receiver limiter and interconnecting cables. Of particular concern is the dissipative and reflective losses of the diodes. The dissipative loss is essentially flat over a 10% bandwidth. The shunt resistance, R_p, of the diodes varies as the inverse square of frequency. The reflective losses are due to diode leakage capacitance and have a significant effect on receive bandwidth. This reactance is cancelled completely (with shunt inductive tuning) but only at a single frequency. A small ratio of diode reactance to line impedance severely limits the total number of diodes

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which can be used in parallel. A further limit is imposed on the high power capability which depends upon a large number of diodes. Some benefits can be gained by reducing the transmission line impedance offered to the diodes.

The high power dissipation varies as the inverse square of the quantity of diodes, $Pd = (4 P_i Rs)/n^2 Z_o$, whereas the insertion loss varies only directly with the quantity, $L_R = (4.35nZ_o)/R_p$. If the line impedance is reduced by a factor of four and the quantity of diodes is increased by a factor of two, the insertion loss is reduced by a factor of two but the power capability remains the same. The technique is sometimes useful but was not required for the subject duplexer.

The duplexer switching action is synchronized with the radar timing system. The T-R is placed in conduction just prior to the high power transmission and is reverse biased just after transmission. Although the diodes required the relatively high value of 600 volts of reverse bias, a suitable driver transistor was selected which had a 1000 volt breakdown rating. The driver circuit design was carefully developed to reduce the cross-over magnitude of collector voltage and current. This was accomplished by the use of a uni-directional current retarding circuit. On "turn-on", the current lags the voltage. On "turn-off", the voltage lags the current.

The common practice at low frequency of using diodes as switching elements requires that the currents and voltages for control be at least equal in magnitude to the peak AC values of transmission. This is a requirement because the usual diode switching analysis is based on a model which is a relatively efficient rectifier. An efficient rectifier, because it responds fast, is inefficient for bilateral energy control.

When allowable switching times are significantly greater than the radio frequency period, improved control efficiency can be achieved by virtue of control energy storage. Fortunately, silicon P-N junction diodes exhibit a charge storage phenomenon of the class required. The characteristic is particularly apparent in P-I-N diodes when used at frequencies above 100 Megahertz. Although most semiconductor rectifiers of the common power supply type exhibit this property, the P-I-N diode has been especially tailored for the specific application.

The stored charge due to a forward control current, I_o, and material life-time, τ, is $Q_s = I_o\tau$. The charge removed by a peak RF current, $\dot{I}_{rf}$, at a frequency, f_{rf}, is $Q_r(\dot{I}_{rf} - I_o)/\pi f_{rf}$. By letting $Q_r = Q_s$ and rearranging, the control gain can be expressed as $\dot{I}_{rf}/I_o = \pi\tau f_{rf} + 1$. Obviously, when $\tau << 1/f_{rf}$, $\tau \to 0$ and the current gain is unity. But when $\tau >> 1/f_{rf}$ at VHF and higher frequencies, the control gain can be very great indeed. For example, a rectifier normally rated at 2 or 3 amps of continuous forward current but actually operated at 1 ampere of peak forward bias and having a material lifetime of 1 microsecond should be able to control a current pulse of 470 peak amperes at a frequency of 150 Megahertz. In a $Z_o = 50$ ohm shunt TEM switching circuit, a single diode should be able to switch a peak power of more than one megawatt. In practice, the actual ability to switch such high power depends upon the transient and average temperatures of the diode. The power capability is directly related to heat capacity and is a trade-off with junction capacitance and switching speed of the diode. The heat capacitance is directly proportional to the junction area for a given thickness. So is the electrical capacitance. Since reception bandwidth is usually difficult to achieve in the shunt TEM because of junction capacitance, a direct conflict exists between power capability and bandwidth. Modern process and fabrication techniques have produced optimum diodes quite adequate for this application up through C band. The Unitrode UM1090, for example, has a peak temperature rating of 300° Centigrade. The diode dice is fairly large having a diameter of 0.09 inch. The single-ended, steady state temperature resistance is 3.5°C/watt. The transient temperature impedance for a 1000 microsecond transmission is 0.35°C/watt. The forward bias impedance at 150 MHz. is typically 0.1 + j0.3 ohms at 1 ampere of bias. The reverse bias parallel equivalent resistance, R_p, is 15,000 ohms or greater at -600 volts and the parallel capacitance is less than 3.0 pfd. Sixteen of these diodes should have a peak power capability, based on a 100°C transient rise, of almost 10 megawatts/T-R mount for a 1000 microsecond pulse or about 4 times the rating of the subject T-R mount. Allowing a peak temperature rise of 200°C, the array should be capable of a 10 percent duty at these power levels or about 10 times the subject rating. The diode has been successfully stressed at Continental on a single basis at 100,000 watts peak power with a 30 microsecond pulse at a 0.001 duty. For a 16 diode array, this would indicate a capability of at least 6.4 megawatts/ T-R mount.

An illustration of the T-R mount is shown in Figure 2. The typical results of a 16 diode array are listed in Table II. The diodes were arranged, as illustrated, in a ring within the abrupt-step transition. The transition is employed to transform a 6" coaxial line to an "N" connector for low-level output. Bias is applied through the isolated tuning stub to the center conductor. All diodes are biased directly in parallel. This technique is workable if the diodes are carefully selected for a nominal distribution of forward current at a saturation voltage of 0.9 volts or more. The spread of currents usually is well within 20 percent. Since the RF resistance of the diode approaches the parasitic impedance value, RF current distribution is not difficult to achieve in a large array.

The method of grossly driving the diodes is compatible with obtaining superior heat sinking to at least one end of the diodes. It also provides a minimum value of diode series inductance. At 150 Megahertz the diodes do not need series tuning in order to get good high power isolation over a broad band of frequencies. Also, diodes can be inserted with removable slugs. Efficient and economical replacement is achieved.

Two basic duplexers of the type described in this paper have operated successfully for more than four months at two sites which have widely different environments. Several hundred hours of successful operation have been logged.

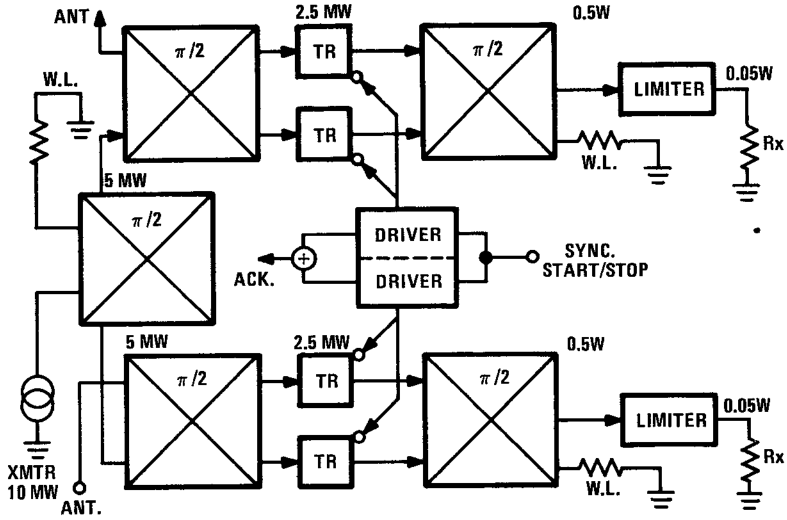

FIGURE 1. FUNCTIONAL DIAGRAM OF 10 MEGAWATT DUPLEXER

TUNABLE CENTER FREQUENCY	130 - 170 MHz
INSTANTANEOUS BANDWIDTH	10%
PULSE WIDTH (MAX.)	1000 MICROSEC.
PEAK POWER	12 MW
AVERAGE POWER	120 KW
ISOLATION	>80 dB
INSERTION LOSS (XMT.)	<0.2 dB
INSERTION LOSS (RCVE.)	<1.0 dB
VSWR (XMT. OR RCVE)	1.2:1
TURN-ON SPEED (T_D+T_R)	<2.0 MICROSEC.
RECOVERY SPEED ($T_D+T_f+T_s$)	<16.0 MICROSEC.
COOLING	WATER-AIR
NOISE RATIO TO 2.5 dB N.F. RCVR	<1.05:1

TABLE I. ELECTRICAL PERFORMANCE CHARACTERISTICS OF DUPLEXER X6-2

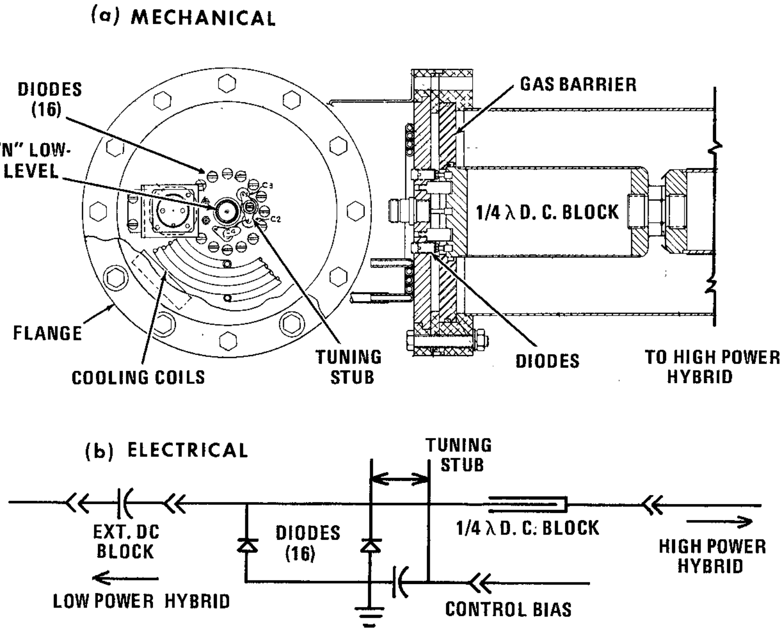

FIGURE 2. SIMPLIFIED MECHANICAL AND ELECTRICAL DIAGRAM OF T-R MOUNT

I_f = I AMP/DIODE @ 0.9 VOLT	$I_o = 1 \times 10^{-6}$ AMP @ 50°C AND -600V.
R_t = 0.15 Ω / DIODE	X_{LS} = 0.3Ω @ 150 MHz/DIODE
R_p = 15KΩ @ 150 mhz./DIODE	X_{cp} = 352 Ω @ 150 MHz/DIODE
L_T = 60. dB	L_R = 0.35 dB
P_{dp} = 117 WATTS/DIODE	P_{da} = 1.4 WATTS/DIODE
ΔT_p = 41°C	ΔT_a = 5.6°C

TABLE II. ELECTRICAL CHARACTERISTICS OF DIODES IN T-R MOUNT

X6-1