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TH-3 Microwave Radio System:

Microwave Integrated Circuits

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Microwave integrated circuits have been employed in the Bell System for the first time in the TH-3 radio system. These circuits, using strip transmission lines, provide numerous functions previously achieved with assemblies of individual waveguide components. By utilizing the modern technologies of photolithography, precision die casting, ceramic processing, and tantalum-gold thin films, these circuits provide advantages in size, fabrication, bay "layout," performance, and cost.

I. INTRODUCTION

A series of four high-performance, functionally integrated, microwave circuits has been developed for use in the 6-GHz TH-3 radio system. The circuit designs integrate numerous microwave functions, operating mostly in a strip transmission line medium, into single packages. They replace assemblies of numerous waveguide components used in the design of earlier radio relay systems. The ability to design each function for its specific application and match it to its neighbor allows high performance standards. The resulting reduction in interconnections improves the reliability. Considerable space and cost saving is realized since many common mechanical requirements are satisfied by the same housing. The housings, made of aluminum, are die cast which accounts for a considerable portion of the cost saving. The stripline center conductor and thin film resistors are created by photochemical techniques from a tantalum and gold film. The use of ceramic substrates and thin film technology is important to the achievement of long-term reliability. The inherent precision and low cost of these techniques provide dimensional control and reproducibility and contribute to the overall cost saving.

The circuits, three of which are used in each TH-3 radio bay, are introduced in a preceding paper[1] by A. Hamori and R. M. Jensen. There are two versions of the carrier distribution circuit. One is for use in a repeater station bay, the other for use in a main station bay. Generally, they perform in the same way in that they distribute microwave carrier power, and provide input and output access to the modulators. Monitoring, testing, and isolation features are included at several points throughout the circuits. The traveling-wave tube input and output circuits operate in the modulated signal path. They provide low transmission loss with negligible slope, high isolation, monitoring, testing, attenuating, filtering, and tuning features within their respective packages. Figures 1 and 2 are photographs of the repeater station carrier distribution circuit and the traveling-wave tube input circuit, schematic drawings are given in Section IV.

Many of the functions integrated into the four circuits are created around one or more circulators as the basic element. A total of seven circulators is included in the repeater station version of the carrier distribution circuit, while the main station version includes six circulators. The traveling-wave tube input and output circuits have two circulators each. All circuits incorporate the necessary transitions between stripline and the transmission modes of the various connecting components. These involve coaxial, waveguide, reduced-height waveguide, and dc connections.

The normal ambient temperature ranges over which the circuits perform are: 80 ± 20°F for the carrier distribution circuits, 90 ± 20°F for the traveling-wave tube input circuit, and 95 ± 20°F for the traveling-wave tube output circuit. The elevated ambient around the TWT circuits is due to the power dissipation of the TWT power amplifier.

II. CIRCUIT DESIGN

2.1 *Circulator and Stripline*

Figures 3a and b show a cross-section view and a top view of the circulator and stripline. The typical circulator admittance as a function of frequency is shown in Fig. 3c. The ground plane spacing is 0.125 inch. A 99.5-percent-pure alumina substrate, 0.024 inch thick, is suspended centrally between the ground planes. The center conductor pattern located on one surface of the substrate is nominally 0.0003 inch thick, about 10 skin depths at 6000 MHz. Since the medium is not homogeneous,[2] the energy does not propagate purely in a TEM mode. There are field components in the direction of

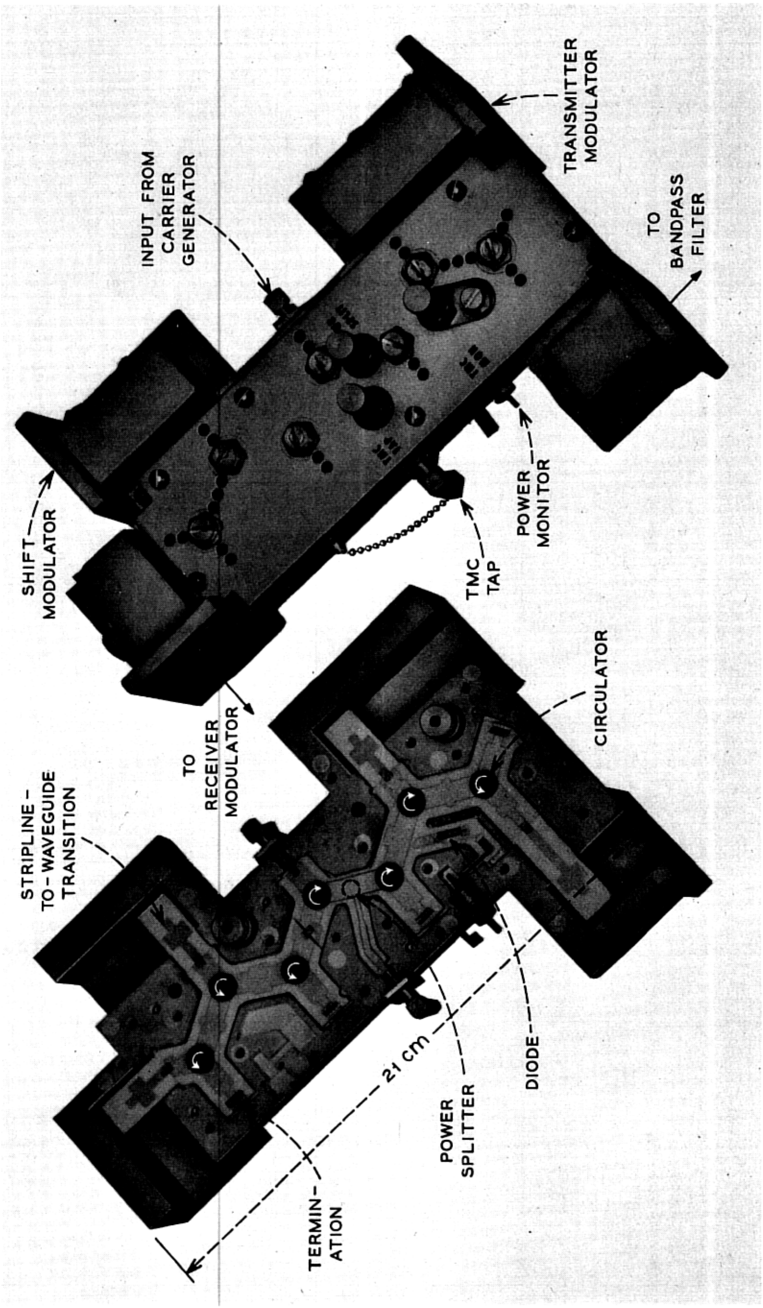


Fig. 1—Open and closed views of the repeater station carrier distribution circuit.

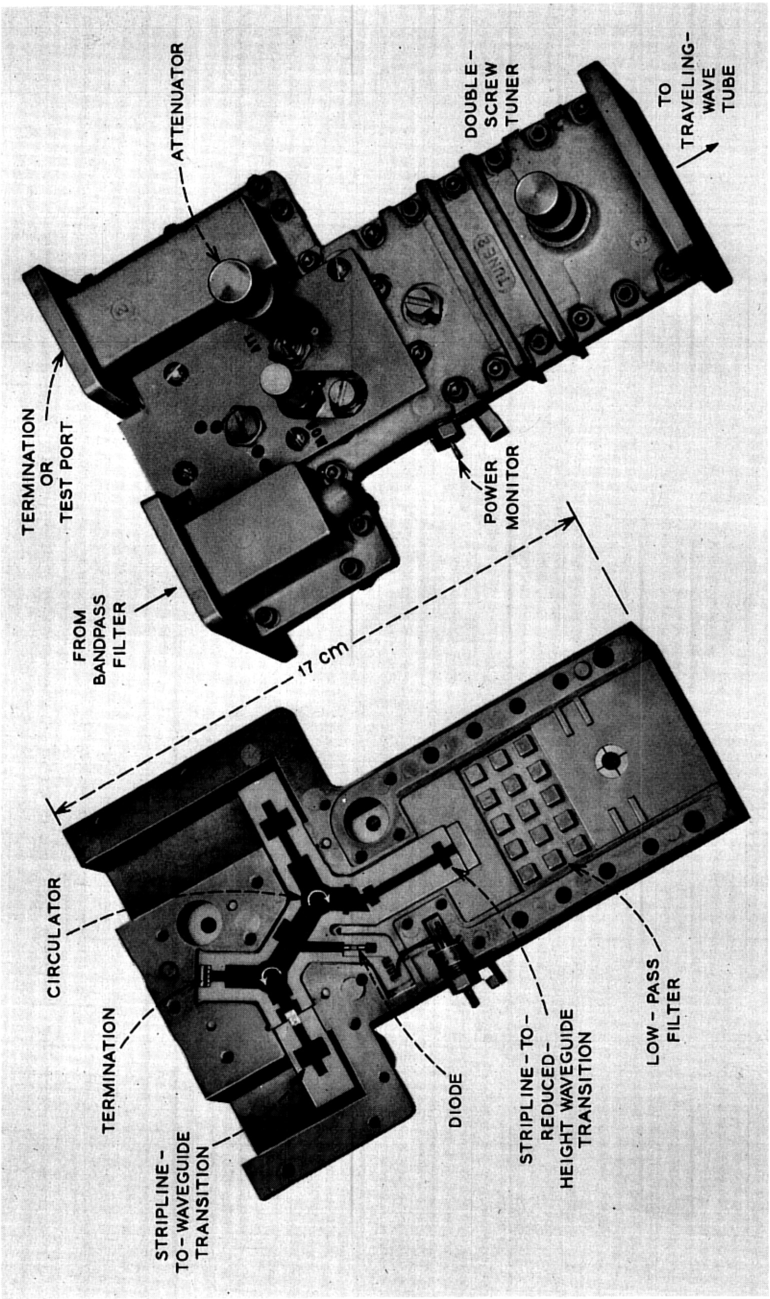


Fig. 2—Open and closed views of the traveling-wave tube input circuit.

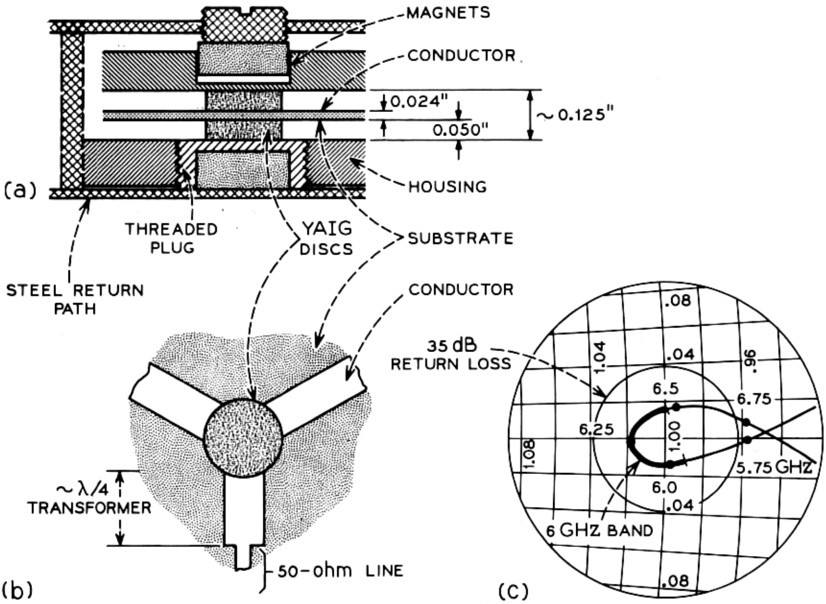


Fig. 3—6-GHz circulator design. (a) Cross-section view (b) Top view of circulator junction. (c) Typical tuned circulator admittance (Smith chart).

propagation. Hence, the medium is dispersive and the propagation constant is a function of frequency and dimensions. This effect, however, is extremely small and may be ignored for practical purposes, and a TEM mode is assumed. The center conductor width required for a characteristic impedance of 50 ohms at 6000 MHz is 0.114 inch. The effective dielectric constant resulting from the presence of the high-dielectric alumina ($\epsilon_r \approx 9.5$) in a small portion of the field region is approximately 1.7 for a 50-ohm line.

The substrate is supported midway between the ground planes by discs of aluminum substituted yttrium iron garnet 0.050 inch thick bonded to both sides of the substrate. The two 0.050-inch-thick discs, plus the 0.024-inch-thick substrate, 0.0003-inch-thick conductor, and two bonds of 0.0002 inch each, stack up to nearly the ground plane spacing. The small difference and the variations due to mechanical tolerances are taken up by tightening a threaded plug located in one ground plane at each circulator junction. Each plug is adjusted individually to contact the circulator discs, thus assuring that no air gap exists. An air gap as small as one thousandth of an inch would change the impedance of the circulator junction and seriously affect the performance. The dimensions of the discs are adjusted to create a resonant junction at the frequency band of interest. The center conductor pattern on the substrate includes quarter-wave transformers to provide suitable impedance matches into the resonant junction.

The junction impedance is approximately 15 ohms. A quarter-wave impedance transformer of 28 ohms transforms the junction impedance to 50 ohms. Each of the functions included in the integrated circuits was developed individually in its own structure and matched to 50 ohms. After integration the performance is optimized by further tailoring the design to improve interface matches between neighbors. The transformer impedance and length are intentionally chosen to be incorrect for an optimum impedance match. This allows for the incorporation of tuning screws into the matching scheme where performance demands are high. Since mechanical tolerances and ferrimagnetic material properties are difficult and expensive to control to the degree required for consistently high circulator performance, two tuning screws may provide orthogonal fine tuning of the circulator admittance trace so that it may be tuned to the center of the Smith chart. One screw, the ratio screw, effectively changes the transformer admittance and provides fine tuning along the conductance axis. The other screw, the length screw, adjusts the effective length of the transformer and provides tuning along the susceptance coordinate. When such fine tuning is not required, the screws are not included and a correction is made in the center conductor geometry.

The YAIG material upon which the circulator design is based has a saturation magnetization ($4\pi M_s$) of 1160 gauss. This material, which is used in other system components, meets the criteria for avoiding low field losses. The 1300-oersted magnetic biasing field is provided by a pair of barium ferrite permanent magnets at each circulator junction. A threaded screw moves one of the magnets to open or close the air gap in the magnetic circuit and provide tunability for the magnetic flux density at the circulator junction. Tunability is required since the permanent magnet nonuniformity can be 10 percent. There are also variations due to mechanical tolerances and ferrimagnetic materials properties. Each circulator is functionally tuned for best operation by adjusting the magnetic field tuning screw.

A study was made of the temperature dependence of the circulator performance. Generally, the circulator performance degrades at the extremes of the $\pm 20°F$ operating temperature range. The degradation is primarily due to changes in the biasing field produced by the permanent magnets. Rather than compensate the magnetic circuit to eliminate the degradation, a margin of performance was added to the room temperature performance objectives. Where isolators and impedance matches are involved, the performance objective was 30 dB reverse loss, or 30 dB return loss per circulator. In order to provide some additional margin to allow for degraded performance over the range of operating temperatures and manufacturing tolerances, the design objective was set at 35–40 dB per circulator. Forward transmission loss is approximately 0.1 dB per pass.

2.2 *Isolator*

The isolation function included in all four circuits is obtained by terminating one arm of a circulator with a thin film resistor in the manner shown in Fig. 4a. The terminating resistor is fabricated by photo-chemical techniques from a film of tantalum nitride. The resistive film is essentially a short section of high-loss transmission

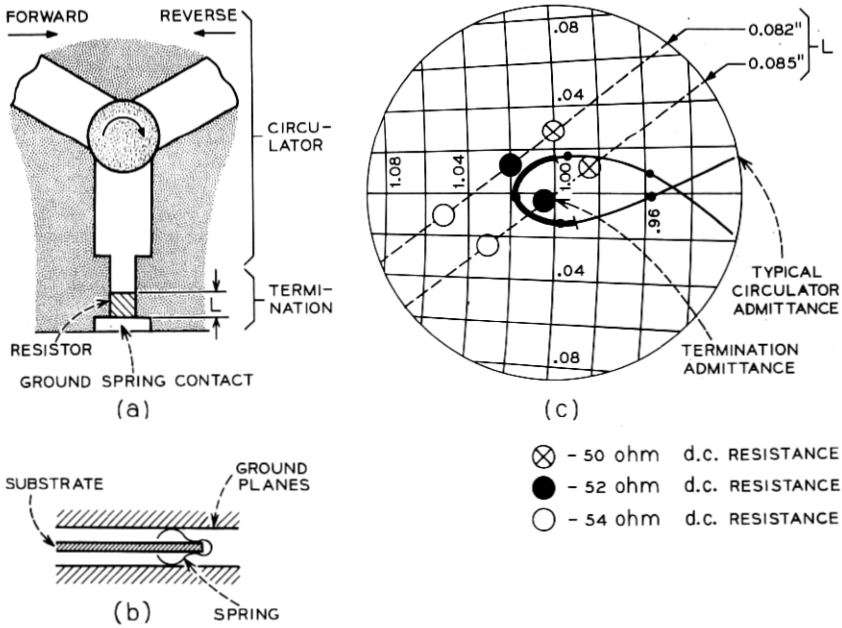


Fig. 4—6-GHz isolator design. (a) Top view of isolator without termination grounding spring. (b) Cross-section view of termination grounding spring. (c) Admittance characteristics of circulator and terminating resistor. The spot size represents the range of admittance variation over the 6-GHz frequency band for each resistor length and dc resistance value indicated.

line. It is connected to the housing through a metal spring clip which slips over the edge of the substrate and contacts a conductor island and the ground planes simultaneously, as shown in Fig. 4b. Thus, each terminating resistor is located close to an edge of the substrate. Various lengths and dc resistances of the resistor were studied. A resistor having a length of 0.085 inch and dc resistance of 52 ohms was chosen as the best termination. This result was based upon the position of the admittance trace on the 1/50 mho admittance chart as shown in Fig. 4c. The spot size represents the range of admittance variation over the 5925–6425-MHz frequency band for the indicated lengths and resistances. Also shown is the typical circulator admittance. It is desired that these two admittances be complex conjugates. The greatest deviation from this relationship is equal to a 0.02 difference in VSWR. This predicts a return loss, and a corresponding reverse loss, for the isolator of 40 dB. The reverse loss of an isolator is determined by the admittance match at the circulator-termination interface. Actual performance of single-stage isolators of this type is: reverse loss consistently in excess of 35 dB and forward loss approximately 0.1 dB.

The power dissipation in the resistor should be limited to 0.5 watt for short time periods and 0.1 watt for extended or continuous periods. These limitations are based on the permanent change in resistance due to elevated temperatures. Tests[3] on resistors show that for 0.1-watt continuous power dissipation, or 10 watts per square inch of film, the change in resistance should be less than 1.0 percent after 20 years. Generally, the limitation on power dissipation is of concern only in the TWT output circuit. Here the signal has been amplified to a power of 10 watts and, should the output of the circuit be severely mismatched, excessive power could be dissipated in the termination of the output isolator.

2.3 *Power Splitter*

The power splitting function[4] used in the repeater station carrier distribution circuit to divide the transmitter microwave carrier (TMC) between the shift modulator and transmitter modulator is shown schematically in Fig. 5a. A tuning screw is located at the common arm junction between two circulators. Various amounts of shunt capacitance may be added to the otherwise matched transmission line by adjusting the screw. Mismatches approaching a short circuit (1–2 dB return loss) can be achieved causing significant portions of the incident power to be reflected. The reflected power is circulated to

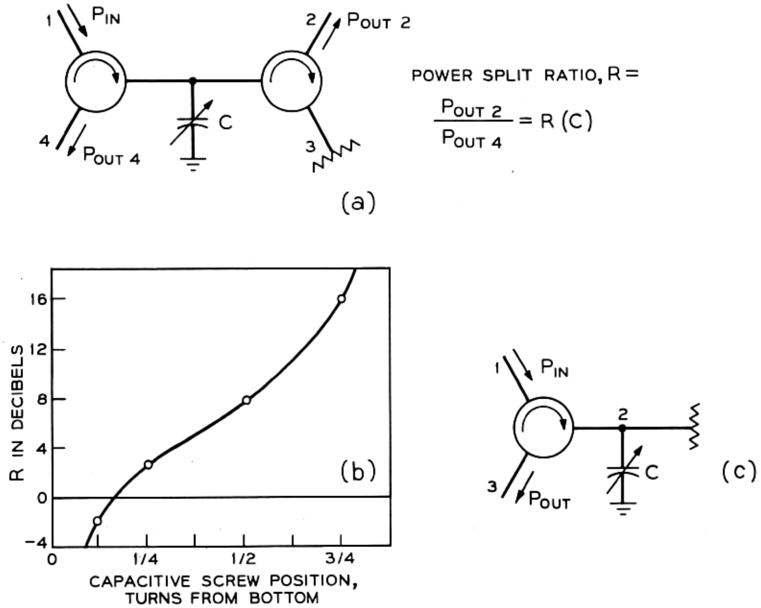


Fig. 5—Variable power splitter and attenuator. (a) Power splitter circuit. (b) Power split ratio vs capacitive screw position. (c) Transmission type attenuator circuit, adjustable from 5 to 30 dB.

port 4, and the power transmitted past the mismatch goes to port 2. Hence, the input power is divided. Note that the input and output impedance matches are not dependent upon the power dividing mismatch. Figure 5b shows the variation of the ratio of the divided powers as a function of the screw position for a frequency in the center of the band.

If the tuning screw reactance is assumed to be purely capacitive and is located at a point on the common arm junction, it is clear that the mismatch will be a function of frequency. The capacitive reactance varies inversely with frequency; hence, the power division ratio will vary as a function of frequency across the 6-GHz band. The ratio varies by 1 dB or more.

2.4 *Attenuator*

Variable attenuation functions are included in the circuits at three places. Each is of slightly different construction in order to satisfy the requirements of the particular application. However, they are similar in operation. The attenuator is created by combining a power splitter and a terminating resistor–as shown in Fig. 5c. A circulator-mismatch combination divides the incident power as previously described and a terminating resistor is added to dissipate either the transmitted or reflected portion of the power. The transmission type (port 2 terminated) shown in Fig. 5c is more suitable where high ranges of attenuation are required, e.g., 5–30 dB. The reflection type (port 3 terminated) is more suitable for low ranges of attenuation, e.g., 0–5 dB. The attenuator included in the traveling-wave tube input circuit (see Section IV) is somewhat different from these two. It is basically a transmission type. However, it is specially constructed to work in a low attenuation range. Since the circuit is operating in the modulated signal path, the attenuator loss must be relatively insensitive to frequency and it also must be capable of a very low minimum attenuation, 0.1 dB or less. Further, its location in the system makes it a desirable point at which to inject or detect test signals during bay testing. Therefore, the termination is an external, removable, waveguide termination rather than a thin film termination. Removal of the waveguide termination makes the port available for testing, provided the mismatching element of the attenuator can be removed or minimized. A coaxial tuning screw was used. It consists of a metal cylinder sliding inside an axial hole in a metal screw body. The two are separated by a dielectric sleeve several thousandths of an inch thick. The cylinder contacts the center conductor and is spring-loaded to protrude from the screw body. As the screw body is turned, the overlap of the two elements varies to change the area of the coupling capacitance. The close spacing and large areas possible with this design allow a greater range of capacitance variation than that available from the limited area and larger spacing of a plane screw body as used in the power splitter. Two mechanical stops are included in the construction of the screw. When the screw is set fully clockwise the capacitance is high ($\approx$ 50 pF) and the return loss is approximately 0.1 dB. When the screw is set fully counterclockwise the capacitance is reduced to the value necessary to create greater than 20 dB return loss. Hence, the attenuator has a continuous adjustment range from 0.1 dB to 20 dB. The loss slope is less than 0.1 dB across the 5925–6425-MHz frequency band in the low attenuation range. The screw is housed in a collet type locking screw so that it may be locked in position, or adjusted against mechanical drag.

2.5 *Power Monitoring*

Power monitoring features are included at several points throughout the circuits. The power on the main signal path is sampled and detected. The current output from the detector is indicated on a panel meter where it may be visually monitored for changes. In some cases, alarm circuits indicate failure of some critical component of the system if the current drops below a specified amount.

Generally, sampling is achieved by placing a transmission line close to the main signal path and capacitively coupling a small amount of power to the line. Different sizes of coupling gaps are used depending on the power on the main signal path. A large-diameter tuning screw is located near the coupling gap to allow fine adjustment of the coupling.

The sampled signal is detected by a point contact silicon diode. The operating current for the meter monitor is 14 microamperes into a 6000-ohm load. For the cases where alarm circuits are used the current is 50 microamperes into 1100 ohms. The diode sensitivity is such that these currents are achieved with −6 dBm and 0 dBm of incident power, respectively. Adequate sensitivity is important since the power coupled from the main signal path to the diode may increase the transmission loss through the integrated circuit along the main signal path.

2.6 *Transmitter Microwave Carrier Tap*

A test set used to measure the FM noise figure of the receiver requires a carrier signal to drive its modulator at the same frequency as the transmitter microwave carrier. This signal is provided by the TMC tap, thus eliminating the need for an additional signal generator at each station. A sample of the TMC, approximately +10 dBm, is available at a miniature coaxial connector, which is essentially an open circuit when not in use. The access port is decoupled from the main signal path by nominally 10.7 dB. The coupling is adjustable, and may be set and locked at the desired level during operation.

Since a decoupling of approximately 10 dB is desired, the TMC power must be divided at the coupling junction such that one-tenth the power is coupled to the TMC tap. The impedance looking toward the tap should be approximately 10 times that of the main signal path. Two quarter-wave transformers are used to transform the 50-ohm connecting cable impedance to 500 ohms at the coupling junction while assuring that during periods when the test signal is not utilized, the open circuit of the connector is transformed to the coupling junction as an open circuit. Calculations show that the impedances of the two sections need only be in the proper ratio to achieve the desired coupling. Characteristic impedances of 25 and 75 ohms were chosen, as these are readily producible. A tuning screw midway along one section provides adjustability of the coupling.

2.7 *Low-Pass Filter and Double-Screw Tuner*

A low-pass filter and double-screw tuner are included in the reduced-height waveguide section of the traveling-wave tube input and output circuits. The filter and the two irises for the tuner can be cast into the housing at little additional cost.

2.8 *Connectors and Transducers*

In addition to the previously mentioned functions, the circuits provide coaxial-to-stripline, waveguide-to-stripline, and reduced-height waveguide-to-stripline transitions that allow interconnection of the integrated circuits with other components in the radio bay.

The coaxial-to-stripline transitions are created by simply providing a mechanical interface since no transmission mode conversion is necessary. To provide the mechanical interface, the center conductor of the miniature connector is extended, by means of a pin, through a bore in the housing. The bore is dimensioned so that it forms a 50-ohm air line with the extending pin. The end of the pin is flattened so that it may be overlapped onto the conductor on the substrate and soldered in position. Greater than 20 dB return loss, which is adequate for the application, is achieved in this manner.

A waveguide-to-stripline transition[5] was developed having greater than 30 dB return loss over the 6-GHz band. The substrate and center conductor are extended through a channel to the broad wall of the waveguide. This extension has a characteristic impedance of 50 ohms. The channel ends at the waveguide wall but the substrate and conductor continue into the waveguide to form an E-field coupling probe as shown in Fig. 6a. The conductor pattern and probe position were varied to optimize experimentally the impedance match relative to a 50-ohm standard. After the incorporation of the transition into the integrated circuits, minor modifications were made to improve the complex conjugate impedance match between the transition and a circulator. Figure 6b shows the typical waveguide port admittance trace obtained from this combination. A special version of this design is used for the bandpass filter port on both carrier distribution circuits. An extra half wavelength of 50-ohm line is included between the circulator and the waveguide. This increases the physical separation between the BPF port and the transmitter modulator port and provides space for the connecting components.

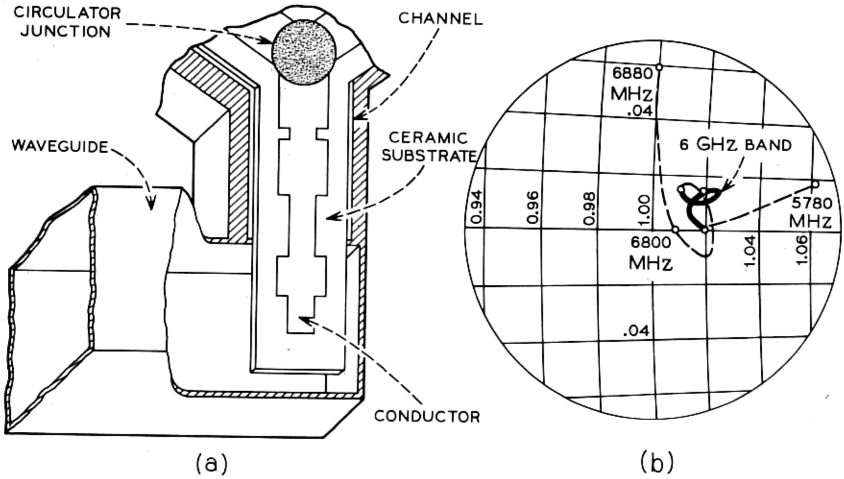


Fig. 6—Stripline-to-waveguide transition. (a) Cut-away view of transition integrated with circulator. (b) Typical waveguide port admittance.

A reduced-height waveguide-to-stripline transition was developed for use in the TWT input and output circuits. In a manner similar to the full-height design, the substrate is extended into the waveguide. However, the substrate enters through the end wall of the guide as shown in Fig. 7. The conductor is shorted to the broad wall to form an H-field coupling loop. The transition has return loss capability of greater than 30 dB.

III. CIRCUIT FABRICATION

3.1 *Substrates*

The substrate material is 99.5-percent-pure aluminum oxide. It is specially processed to a fine grain size before firing. The material was chosen for its low loss tangent and compatibility with processing methods. The fine grain structure allows a finer surface finish to be achieved. A surface finish of 16 microinches, or better, is desirable in order to control the sheet resistivity of the sputtered tantalum nitride film. Excessive surface roughness results in a higher resistivity than expected. For the generation of precision microwave terminating resistors, the objective is usually a low value of resistivity to allow subsequent anodic trimming to the desired resistance.

The shape of the substrates is quite irregular as shown in the photo-

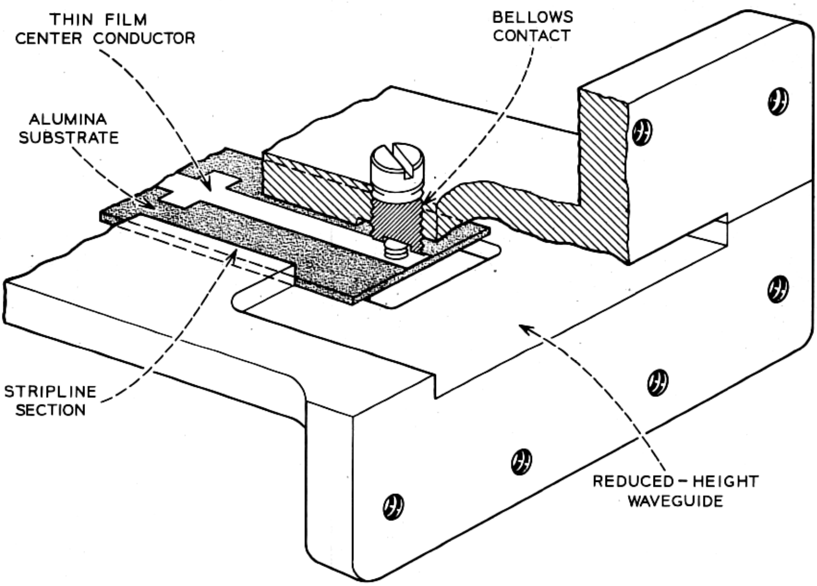


Fig. 7—Stripline-to-reduced-height waveguide transition for interconnection to the traveling-wave tube.

graphs, Figs. 1 and 2. Initially, the center conductor patterns were created on simple rectangular substrates which were placed into housings having large open rectangular areas. This approach was successful in other microwave integrated circuit designs[4,6] at lower frequencies. However, at 6000 MHz, and where isolations as high as 68 dB between ports are required, considerable difficulty was encountered. In this case, the dimensions of the open areas within the housing were large enough to support the propagation of energy through waveguide type modes. This provided shunt paths around the isolators. Consequently, isolations between ports in excess of 40 to 50 dB were not possible. To eliminate this problem, the metal walls in the housings were moved closer to the center conductor pattern to create a channel suitable for the TEM stripline mode, but below the cutoff dimension for the waveguide modes. Channel widths of 0.570 inch were used. Isolations between ports in excess of 80 dB were observed with this configuration. A second consideration also becomes important in the determination of the substrate shape. With the metal housing walls moved closer to the center conductor as described, the dimension of the open areas near the circuit junctions was close to a resonant dimension in some cases. This allowed cavity type modes to be sustained, affecting couplings, transmission losses, etc. Additional relocation of the metal walls in the troublesome areas shifted the resonant frequencies of these modes out of the band of interest. These two considerations resulted in complex substrate and housing-channel geometries.

In spite of the irregular shape the ceramics industry is capable of supplying finished substrates at a reasonable cost. The shape is die cut from a ribbon of "green" ceramic near the finished thickness. All dimensions are oversize to allow for shrinkage during the firing process. After firing, the substrates are ground to the required thickness and surface finish. Several outside edges of the substrates are also ground to dimension so that they may subsequently be used as referencing edges for registering the pattern relative to the housing.

There are several practical considerations that limit the maximum substrate dimension to approximately four inches. The photographic process capability used in creating the pattern masks is limited to four inches. Also, the substrate manufacturer has similar limitations on size due to process capability and breakage due to handling. Breakage during pattern creation must also be considered. Hence, the conductor patterns are divided into sections, each on its own substrate. The substrate conductor patterns are interconnected with solder-clad ribbon, using solder reflow techniques, at assembly into the housing.

3.2 *Conductor Pattern*

The stripline center conductor geometry is a primary performance-influencing parameter. The width and thickness of the conductor pattern determine the impedance of a transmission line when it is suspended in a medium between two fixed ground planes. Since impedance matching between the various functions is crucial to functional integration, it is essential to control these factors for accuracy and reproducibility. Various linewidths between 0.010 and 0.270 inch are used. These are produced within approximately two-mil tolerances. The photo-chemical techniques lend themselves well to provide the necessary dimensional control at low cost. A suitably cleaned substrate receives several layers of sputtered or evaporated materials. These include: tantalum nitride subsequently used to create the necessary resistors, titanium to provide adherence for following films, palladium[7] to insure long-term adherence of the conductor layers, and gold to prevent oxidation and provide a conductive surface which can subsequently be electroplated to greater thickness.

The conductor pattern and resistor areas are generated from a substrate prepared as described. The undesired materials are selectively etched away using the appropriate photo-chemical techniques. The materials left form the conductor and resistor patterns. The patterns are registered on the substrate by including referencing marks on precision masks to align with the previously mentioned reference edges of the substrates. These reference edges are subsequently positioned against corresponding edges in the housing at assembly to provide proper registration between pattern and housing. The conductor pattern is electroplated with a heavy layer of copper followed by a gold flash to a total thickness of approximately 0.3 mil. The final gold flash prevents oxidation of the conductor surface. The resistor areas are thermally stabilized and trim anodized to proper value at the appropriate stage of processing. The transmission line terminating resistors are 52 ohms. Current limiting resistors in the diode monitor circuits are 1000 ohms.

3.3 *Patterned Substrate Assembly*

The patterned substrate has numerous additional parts added to it before it is ready for assembly into a housing. Circulator discs are bonded to each side of the substrate at each circulator conductor junction. The discs are optically located concentric to the geometric center of the pattern. The center is identified by a cross-hair index mark which is generated along with the pattern. The cross hairs are included in the precision mask which is cut at ten times oversize and reduced to minimize variations due to cutting tolerances. The discs are cemented in position. A fixture is used to align the disc on the unpatterned side of the substrate. At assembly, the brass plug in the housing is turned down against the disc-substrate assembly, placing the stack in compression.

3.4 *Housings*

A precision machined, hardened steel die is used to cast aluminum housings in large quantities to finished dimensions. Where tolerances are tight, or where surfaces must be carefully mated, machining stock is provided for finish cuts with machine tools. Many holes and bores are cast into the housing. Additional reaming and tapping of holes is also required. A small draft is necessary to facilitate release of the part from the die. The effects of this draft and any dimensional variations were determined. These required compensation by changes in the center conductor pattern dimensions or by modifications in the steel die. This was particularly effective in the area of the waveguide transitions where the location of the probe conductor relative to the waveguide walls is critical. Aluminum housings can be made available in production quantities by the casting and finish machining approach at a fraction of the cost of equivalent machined housings.

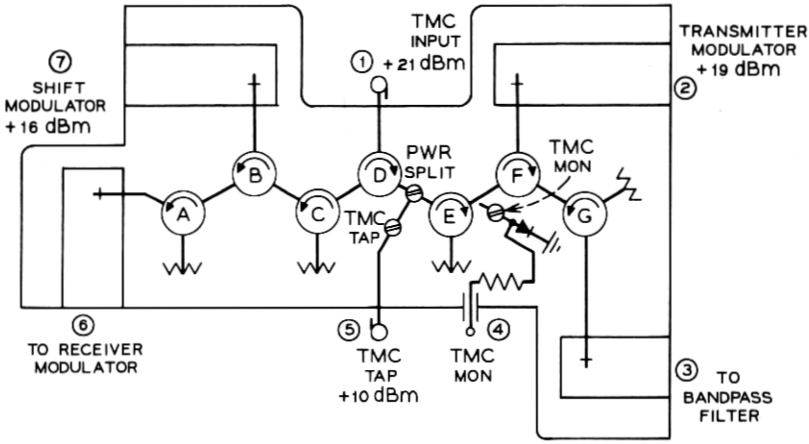
Due to the similarity between the repeater station version and main station version of the carrier distribution circuit, the same cast housing is used. Slight machining variations, e.g., the size and location of some holes, and the removal of one waveguide port make the one casting suitable for both circuits. Similarly, a single casting serves for both the traveling-wave tube input and output circuits with minor machining variations. The RF leakage power outside the housing was measured for a large number of assembled circuits. The leakage power was consistently low enough to avoid the generation of interfering tones due to multiple leakage paths within a bay.

IV. CIRCUIT PERFORMANCE

The performance requirements for each circuit are determined by system needs and device capabilities. The isolation requirements are derived from the carrier-to-interference ratio objectives established for the radio bay. Transmission loss is minimized to that which the design is capable of meeting. Any additional loss adds to the microwave generator power output requirements. Impedance match requirements are based on the allowable ripple amplitudes and mismatch losses.

Specific performance requirements for the circuits are discussed in a preceding paper[1] by A. Hamori and R. M. Jensen. The performance of two of the integrated circuits will be described to show the performance capability achieved. Figure 8 lists the important room temperature performance limits over the frequency band of 5925–6425 MHz for a repeater station carrier distribution circuit. Figure 9 shows the traveling-wave tube input circuit.

The realization of some of the isolations is through combinations of isolators and common arm matches. For example, in Fig. 8 the isolation from port 7 to port 1 is due to the impedance match on the common arm between circulators "A" and "B" and the isolator at position "C." Any power reflected from the common arm match passes through circulator "B" to circulators "C" and "D" to port 1. Portions of the modulated signal power returning from the transmitter modulator at port 2 may be reflected from the common arm junction between circu-



IMPEDANCE MATCHES		ISOLATIONS AND LOSSES		
PORT	RETURN LOSS (MIN. dB)	FROM PORT	TO PORT	dB
1	23	2	7	60 MIN.
2	30	6	1	68 MIN.
3	32	7	1	68 MIN.
6	28	1	2	3 MAX.
7	28	2	3	0.5 MAX.
		7	6	0.5 MAX.

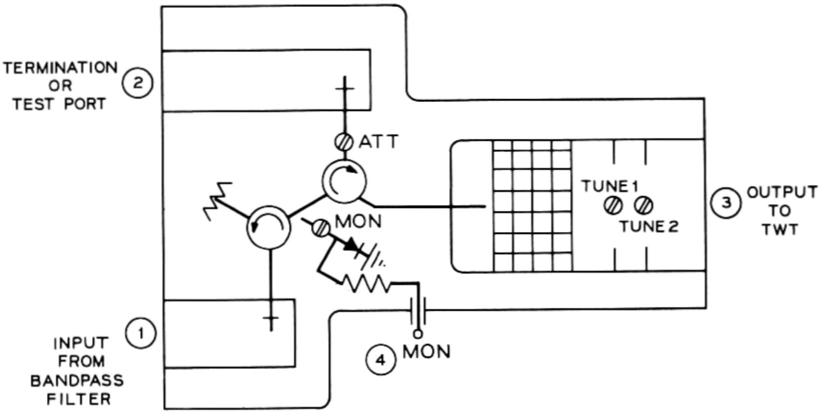
ADJUSTMENTS

PWR SPLIT – SET FOR +3 dB POWER AT PORT 2 RELATIVE TO PORT 7.
TMC TAP – SET FOR –11 dB PORT 5 RELATIVE TO PORT 1.
TMC MON – SET FOR 14 μA. INTO 6-kohm METER LOAD.

Fig. 8—Performance requirements of repeater station carrier distribution circuit.

lators "F" and "G." It then passes through circulator "F" to the line sampled by the power monitor. Since the sampling is nondirective, the diode will detect the sum of the modulated signal power and the desired carrier power. Therefore, tuning screws are included in these common arms so that the matches may be finely adjusted to minimize the reflected signals.

In cases where the common arm match does not enter into a required isolation, circulator tuning screws are not used and an adequate match is achieved by adding fixed tabs to the center conductor geometry. In addition to common arms, several isolators are tabbed where the



IMPEDANCE MATCHES		ISOLATIONS AND LOSSES		
PORT	RETURN LOSS (MIN. dB)	FROM PORT	TO PORT	dB
1	32	1	3	1* MAX.
2	20	2	3	0.5** MAX.
3	—	3	1	30 MIN.

ADJUSTMENTS

ATT – CONTINUOUSLY VARIES LOSS, PORT 1 TO 3. *FULLY CLOCKWISE, **FULLY COUNTER CLOCKWISE.

MON – SET FOR 14 μA. INTO 6-kohm METER LOAD.

TUNE 1, TUNE 2 – ADJUSTED TO OPTIMIZE INPUT MATCH TO TWT.

Fig. 9—Performance requirements of traveling-wave tube input circuit.

performance demands are not extreme. An isolator using tabs rather than screws is consistently capable of 25 dB reverse loss, or greater. This simplifies the adjusting procedure at the expense of finely tuned performances.

The main station carrier distribution circuit has similar performance to the repeater station version. High isolation between the transmitter and receiver circuits is readily achieved since the two are physically separated by a metal housing insert which separates the channels of the two circuits. Isolation in excess of 80 dB is inherent.

V. CONCLUSIONS

The use of microwave integrated circuits in the TH-3 radio system represents a significant advance in the incorporation of modern technology into the design of radio bays. By utilizing photolithography, precision die casting, ceramic processing, and tantalum-gold thin films, these circuits provide advantages in size, fabrication, bay "layout," performance, and cost. A carrier distribution circuit constructed with conventional waveguide components would have ten times the weight, fifteen times the volume, and would require three times the number of RF interconnections. The ability to fabricate and test each function to work best with its neighbor results in well-controlled superior overall performance. The reduced number of interconnections improves the reliability and contributes to the cost savings.

The established advantages of this approach have led to the design and application of similar circuits for the 4-GHz TD-3 radio relay system.

VI. ACKNOWLEDGMENTS

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