Use of Microelectronic Technology in the Design of a 3.7–4.2-GHz Portable Microwave Repeater

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Abstract—**The use of microelectronic technology in the design of a frequency-agile portable microwave repeater (PMR) provides for light-weight compact construction without sacrifice of performance. The instrument is intended for use as an up and down converter in FM systems and operates at 20-MHz channels in the 3.7–4.2-GHz band with a 70-MHz intermediate frequency (IF). A brief discussion of its design indicates the significance of each microelectronic assembly. Key microelectronic designs include a 4-GHz transistor preamplifier with a 5-dB noise figure, a 4-GHz double-balanced mixer of coplanar construction, and a 4-GHz power amplifier with 1.5 W of output power. Comments are offered regarding the advantages and limitations of microelectronic design. Back-to-back instrument performance yields 0.2-dB flatness and 2-ns group delay over a ±7-MHz band measured at 70 MHz. FM noise performance is discussed.**

I. INTRODUCTION

STANDBY SYSTEMS which may be switched into service in the event of the failure of an operating system are common in microwave communications networks. These standby systems generally occupy a vacant channel in the RF band, and increasing traffic has encouraged the reduction of the number of standby systems so that more channels may be used to carry service. Such a reduction is possible with the use of portable microwave radios that can be substituted for a system which has failed until the operation of that system is restored. The portable microwave repeater (PMR) discussed in this paper is such a radio. It is capable of being tuned to any one of 25 channels, 20 MHz wide, in the 3.7–4.2-GHz band and consists of the following: 1) a dual down-conversion system from any channel in the 3.7–4.2-GHz band to 70 MHz; and 2) a dual up-conversion system from 70 MHz to any of the 3.7–4.2-GHz channels. Frequency modulation and demodulation occur at 70 MHz but are not a part of the PMR. The instrument can serve as a receiver down-conversion system, a transmitter up-conversion system, or a repeater, by interconnection of the receiver 70-MHz output and the transmitter 70-MHz input. Because of the frequency-agility requirement at 3.7–4.2 GHz, the components operating in this band must have minimum amplitude variation with frequency. The entire instrument must achieve flat amplitude response and minimum group delay over any 20-MHz channel. The performance also allows the instrument to be used as an effectively "transparent" down converter or up converter for microwave link tests in substitution measurements or for system alignment.

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The portability requirement encourages the use of microelectronic technology in this application. The small size of the circuits and the inherent short connections allow the final instrument to be portable as well as avoiding performance sacrifices. Those components which employ microelectronic (thin-film) design will be emphasized in this paper.

The design of the complete instrument will be discussed first, followed by greater detail on each of the microcircuit assemblies. Comments will then be offered on the advantages and limitations of the technology. The final section of this paper will deal with the performance of the instrument as an FM radio.

II. INSTRUMENT DESCRIPTION

A photograph of the 4-GHz PMR is shown in Fig. 1(a), and a block diagram is shown in Fig. 1(b). Many of the components of the up converter and the down converter are identical. Those portions of the instrument which employ microelectronic design are highlighted in the block diagram.

Down conversion to 70 MHz is accomplished in two steps. As shown in Fig. 1(b), the FM carrier is down converted to 1430 MHz, filtered, and then down converted to 70 MHz. The first local oscillator (LO) is a YIG-tuned transistor which is phase locked to a synthesized crystal-controlled reference signal. Channel selection is accomplished by tuning this LO to 1430 MHz above the incoming carrier. The second LO is selected by the instrument logic to be either above or below the 1430-MHz intermediate frequency (IF) to allow reversal of the modulation sense. This LO signal is generated by multiplying a crystal-oscillator output signal by either a factor of 7 or 8 and passing the signal through a dual passband filter to develop 1360 MHz and 1500 MHz, respectively.

A thin-film amplifier at 70 MHz provides receiver automatic gain control (AGC). The 70-MHz nine-section arithmetically symmetrical bandpass filter is accurately equalized for group delay using a single-section parabolic equalizer. When the received signal level reaches a minimum threshold, a carrier resupply with pilot signal tones is triggered to provide continuity of the carrier in repeater applications.

As the block diagram indicates, the transmitter 70-MHz input is applied to an amplifier with AGC and then to the 1430-MHz up converter. Selectivity is provided at 1430

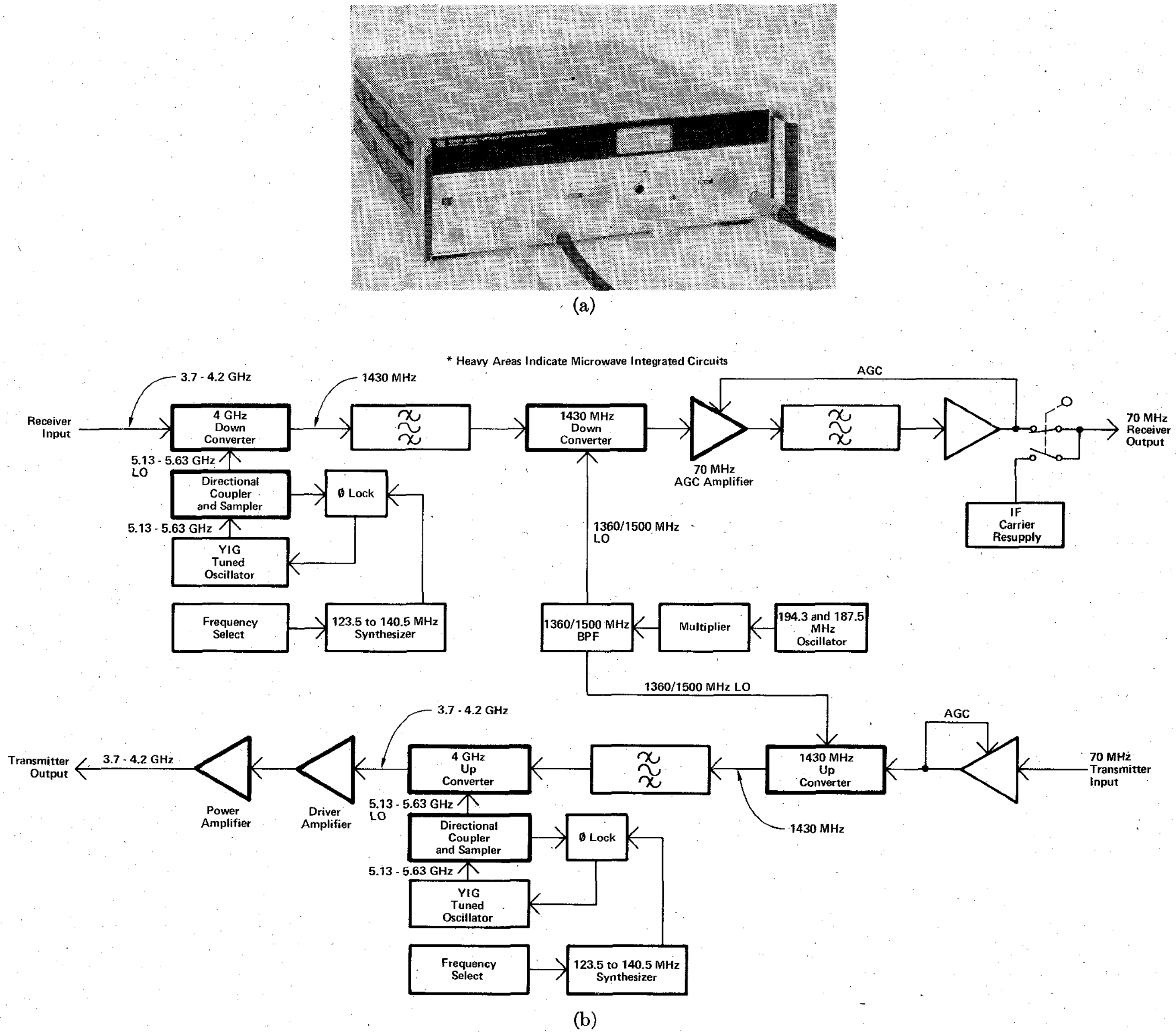


Fig. 1. (a) Photograph of the PMR with cables connected to the 4-GHz ports and with receiver and transmitter 70-MHz ports interconnected. The instrument is 5.25 in high. (b) Block diagram of the PMR with microcircuit assemblies highlighted.

MHz and the signal is applied to the 4-GHz up converter. Receiver and transmitter share the 1500/1360-MHz LO; the transmitter 5.13–5.63-GHz phase-locked LO is identical to that of the receiver. Transmitter output power is developed by two amplifier assemblies providing more than 1 W at the instrument output.

The instrument design includes various alarm circuits to indicate an unlocked condition of an LO, a triggering of the carrier resupply, or a low-transmitter output level. Muting of the transmitter occurs under an out-of-lock condition. A meter indicates input and output levels of both receiver and transmitter.

III. MICROELECTRONIC COMPONENTS

A. General

Computer optimization was used in the design of all components. Circuit modeling included lumped-element and transmission-line parameters; equivalent circuits, derived from S parameter data, represented the diodes and transistors.

Most of the circuits are fabricated on sapphire substrates, providing a smooth surface for the deposition of thin-film resistors and capacitors. 50-Ω/square tantulum nitride material is used to form resistors; capacitors are composed of interdigitated chrome–gold conductors over tantalum pentoxide and quartz layers. Whenever possible, low-cost alumina is used for conductor-only circuits. Circulators are fabricated on garnet substrates. Tuning is accomplished by making and breaking bonds to allow adjustment of capacitance, inductance, resistance, and line length. The circuits are designed to interface with 50 Ω of characteristic impedance.

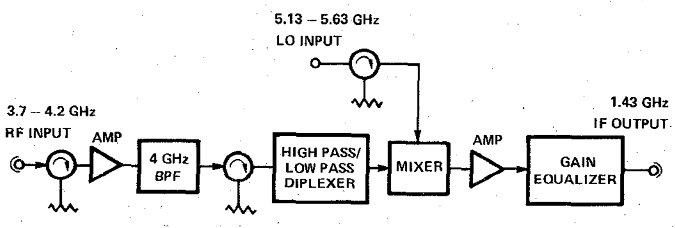
Individual substrates are placed on a baseplate held in place by mounting hardware and wire bonded to establish interconnections. Hermetically sealed feedthroughs in the baseplate provide inputs and outputs for RF and dc, and costs are minimized by performing all machining and welding on one surface. The entire assembly is hermetically sealed with an extruded cover which is bonded to the baseplate. Examples of subassemblies are given in the discussion that follows and are shown in the various figures.

B. 4-GHz Down Converter

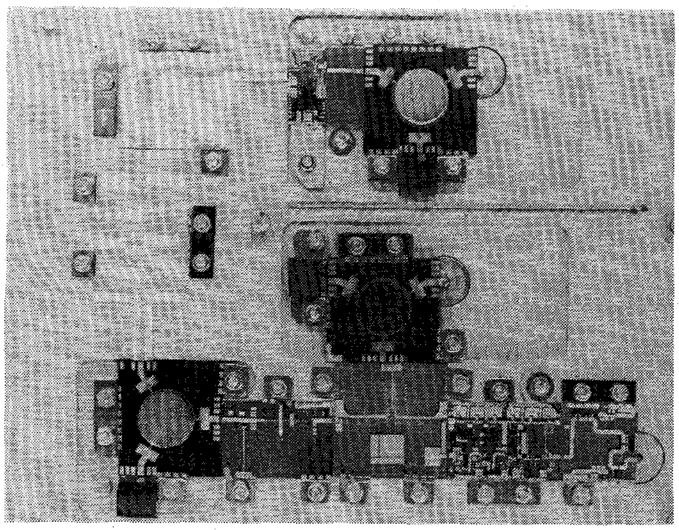
A block diagram of the 4-GHz down converter is shown in Fig. 2(a), and a photograph of the microcircuits mounted on a baseplate is shown in Fig. 2(b). The 4-GHz input signal is applied to a circulator which matches the input (1.2:1). The signal is then delivered to the 4-GHz preamplifier. The interface between the preamplifier and the bandpass filter is tuned to avoid gain ripples. Selectivity of the 3.7–4.2-GHz band is established by the filter. The circulator following the filter avoids interface problems with the mixer. High-frequency signal components emanating from the mixer are terminated in a complementary diplexer [1] formed from six-section foreshortened high-pass and low-pass filters. The high-pass filter is terminated in 50 Ω and includes the converter image frequency in its passband. The mixer is double balanced and has an IF centered at 1430 MHz with a ±20-MHz bandwidth. The IF output from the mixer is applied to a three-stage 1430-MHz preamplifier which has a 20-dB gain and a 4-dB noise figure. Gain level is set at 1430 MHz in this assembly using cascaded π attenuators of 8, 4, 2, and 1 dB. IF slope is set on the same substrate using a parallel *LC* resonant circuit (chip capacitor in shunt with a transmission line) coupled to the transmission line through a 150-Ω resistor. The slope is set to be positive or negative by tuning the transmission-line length so that resonance is above or below 1430 MHz. Conversion-gain flatness versus frequency with the IF output frequency fixed is shown in Fig. 2(c).

A simplified schematic of the 3.7–4.2-GHz preamplifier is shown in Fig. 3(a). The amplifier, visible in the photograph of Fig. 2(b), achieves a 5.0-dB noise figure and has a 13-dB gain with typical flatness versus frequency as shown in Fig. 3(b). The transistors are silicon bipolar devices. Noise figure and gain flatness have been optimized by the computer techniques mentioned previously.

The 4-GHz mixer circuit is represented by the diagram of Fig. 4(a). It is fabricated on a 0.025-in-thick sapphire substrate; the diode quad is a classical ring array and is driven from the LO port through a quarter-wavelength balun as described by Marchand [2]. The tight coupling of the two lines forming the balun approximates a similar



(a)



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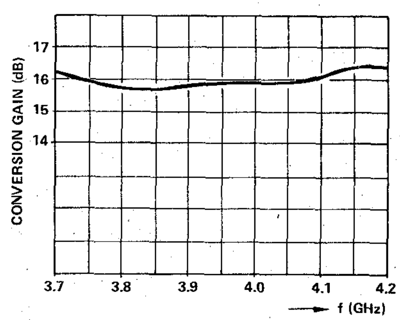
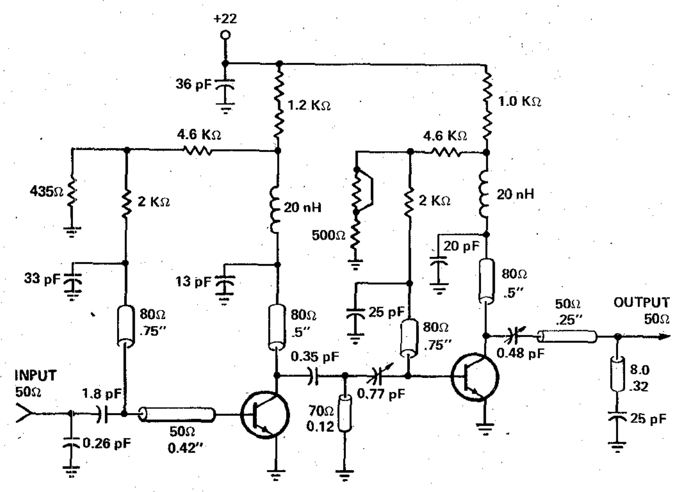
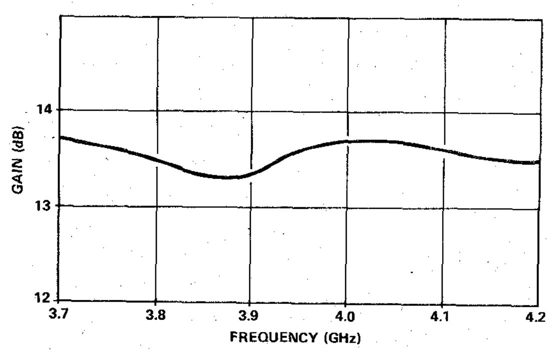


(c)

Fig. 2. 4-GHz down converter. (a) Block diagram. (b) Photograph of the circuit assembly. (c) Conversion gain over the 3.7–4.2-GHz band with IF fixed at 1430 MHz.



(a)



(b)

Fig. 3. 4-GHz preamplifier. (a) Schematic diagram. Transmission lines are described by characteristic impedance and equivalent length for $\epsilon_r = 1$. (b) Gain versus frequency of a typical unit.

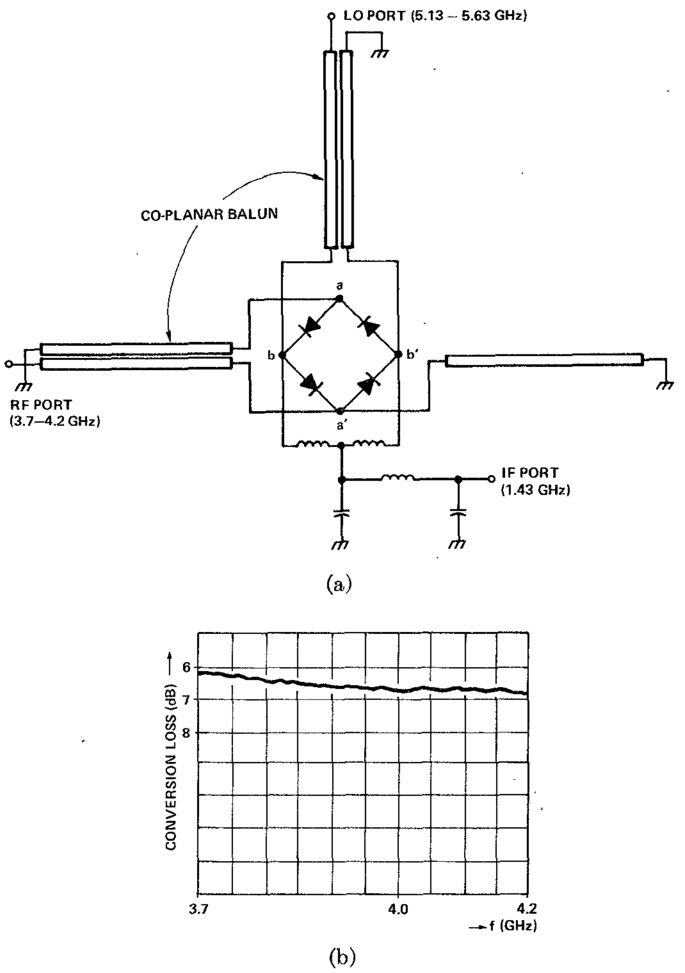


Fig. 4. 4-GHz mixer. (a) Schematic diagram. (b) Conversion loss versus frequency for a typical unit with IF fixed at 1430 MHz.

structure in a coaxial transmission line with the grounded-coaxial outer conductor replaced by a microstrip line. The coupling is further increased by removing the metallization on the back side of the substrate and providing a cavity in the baseplate below. The even-mode impedance of the transmission-line pair is greater than 120 Ω and the odd-mode impedance is 60 Ω, providing impedance match to the diode quad. The RF signal is applied to a similar balun, also described by Marchand for the coaxial case. A shorted $\lambda/4$ stub is added to accomplish improved current balance versus frequency [2]. This stub and the grounded $\lambda/4$ section of the RF transmission-line pair provides low impedance to ground for the IF. The IF output connections to the diode quad can then be made at the LO-connection points [points b and b' in Fig. 4(a)]. Note that the grounded section of the LO-input transmission-line pair provides a high impedance to the IF output at points b and b', and the symmetrical $\lambda/4$ sections of the RF-input balun provide a low impedance at points a and a'. The requirement of high impedance from points b and b' to ground, however, disallows operation at low IF's. The IF voltage appears at points b and b', referenced to ground, and can be coupled off as shown. LO current is forbidden from flowing through the IF path between points b and b' by setting this path to $\lambda/2$ at the LO frequency. The IF is effectively center tapped as shown, and the three-section low-pass structure is included to further reduce LO and RF leakage as well as to provide impedance matching to 50 Ω. This mixer design has the following characteristics which make it suitable for this application: 1) high IF with minimum RF bypass capacitance resulting in flat conversion efficiency versus frequency; 2) broad-band RF and LO circuitry avoiding high-frequency (LO harmonic and sum frequency) resonances, which often cause intolerable amplitude-flatness problems in the frequency-conversion system; and 3) thin-film coplanar construction allowing minimum cost and convenient interfacing to other components.

The mixer diodes are beam-lead Schottky-barrier type with a 0.25-pF maximum capacitance at zero bias. The mixer substrate is shown mounted on the 4-GHz-converter baseplate assembly in the photograph shown in Fig. 2(b). A plot of typical conversion loss versus frequency with the IF fixed at 1430 MHz is shown in Fig. 4(b). Table I summarizes the mixer performance.

The 4-GHz bandpass filter and the high-pass/low-pass diplexer are visible in the photograph of Fig. 2(b). The bandpass filter is a five-section design composed of $\lambda/4$ short-circuited stubs interconnected with $\lambda/4$ transmission lines [3]. The diplexer mentioned in the preceding discussion consists of a six-section high-pass filter (terminated) and a six-section low-pass filter (signal path), each foreshortened to five sections. The low pass consists of high-impedance sections and short ($<\lambda/4$) open-circuited stubs of low impedance. The high pass consists of silicon dioxide chip capacitors in the series arms and transmission lines of 90-Ω characteristic impedance for the shunt arms. Tuning of the high pass is accomplished by breaking bonds to change the effective length of the shunt arms. The diplexer crossover frequency is at 5.4 GHz, and VSWR at the mixer-interface port is 1.2:1 over the RF band and the image-frequency band. Image-signal leakage through the low-pass arm of the diplexer is less than −17 dB. The diplexer provides isolation of the mixer from circuitry at the mixer RF input for frequencies above 6.5 GHz, allowing the circulator to be optimized for the 3.7–4.2-GHz and 5.13–5.63-GHz bands.

The circulators are used to provide RF- and LO-input match and isolation between the mixer and the bandpass filter. A thin-film microstrip 50-Ω load terminates the third port of each circulator. The circulators exhibit 17-dB isolation and less than 0.7-dB insertion loss over the 3.6–5.7-GHz band. The substrate is G1002,[1] a garnet material with a saturation magnetization of $4\pi M_s = 1000$ G. Design was accomplished using the theory of Fay and Comstock [4], as modified for microstrip by Masse [5]. The substrate material is 0.075 in thick. Impedance matching to 50 Ω is accomplished using high/low-impedance sections, eliminating the need for a $\lambda/4$ transformer [6]. The

[1] Supplied by Trans-Tech, Inc., Gaithersburg, Md.

TABLE I
TYPICAL 4-GHz MIXER PERFORMANCE

Conversion Loss		6.6 dB
Return Loss	LO	8.5 dB
	RF	9.5 dB
	IF	11 dB
To RF Isolation		22 dB

magnet is inserted into the baseplate from the underside; a mild steel pole piece is bonded to the substrate to provide concentration of the magnetic field.

Conversion flatness of the 4-GHz down converter versus frequency with fixed IF is shown in Fig. 2(c). The design allows adjustment of gain to 16 $\pm$ 1 dB. The noise figure is 8.0 dB and conversion flatness over any 20-MHz channel in the 3.7–4.2-GHz band is 0.1 dB. LO power of +10 dBm is required to drive the converter, and return losses of -20 dB are achieved at all ports.

Leakage of the LO signal from the RF-input port of the 4-GHz down converter is -80 dBm. In the microstrip mode, this level of isolation is achieved by the combined attenuation of the circulators, the bandpass filter, and the $|S_{12}|$ of the preamplifier. Radiated spurious-mode leakage, however, would be intolerable without shielding of the circuits. This is accomplished by using an interior cover with waveguide below cutoff-type feedthroughs electrically located at the input and output ports of the bandpass filter. Energy in the microstrip mode is carried through the waveguide section by the transmission lines on the narrow substrates visible in the photograph of Fig. 2(b).

The 4-GHz up converter, identifiable in the block diagram of Fig. 1(b), is identical to the 4-GHz down converter with amplifiers omitted and the direction of circulation reversed for the circulators (polarity of magnets reversed). Conversion flatness over the 3.7–4.2-GHz band is 0.5 dB.

C. 1430-MHz Down Converter

This down converter beats the 1430-MHz IF carrier with a LO at either 1360 MHz or 1500 MHz to produce the 70-MHz carrier. The LO frequency can be above or below 1430 MHz to allow reversal of modulation sense. The assembly includes a three-stage preamplifier (identical to the IF amplifier of the 4-GHz down converter); equalizer attenuators for impedance matching, gain-level adjustment, and gain-slope adjustment; a three-stage LO amplifier; and a single-balanced mixer. The single-balanced mixer uses two Schottky-barrier-diode chips and a hybrid-ring directional coupler with coupling-arm impedances optimized to allow a more broad-band operation than that normally achievable in a "rat-race" hybrid. This widened bandwidth is necessitated by the two LO frequencies. In optimizing the mixer, the coupler developed by Pon [7] is carried one stop further by allowing computer optimization which adjusts transmission-line impedances independently to achieve broader bandwidth. The low-pass structure at the IF port provides impedance matching and additional rejection of LO and RF. The mixer has a 5.5-dB conversion loss and an 18-dB LO to RF isolation.

The 1430-MHz up converter is identical to the down converter with the preamplifier substrate reversed. In this case, the 70-MHz signal is fed into the converter at the mixer IF port, up converted to 1430 MHz with the LO frequency selectable [see Fig. 1(b)], and amplified.

D. 70-MHz AGC Amplifier

A photograph of this amplifier appears in Fig. 5(a). The amplifier consists of four two-stage gain blocks separated by a total of three attenuating stages. The schematic diagram of Fig. 5(b) represents one amplifier–attenuator combination. Each p-i-n-diode attenuator is capable of 20-dB attenuation. The AGC voltage detected at the receiver output is converted into three sequential voltages which activate the attenuators. For very low input signals, all attenuators have minimum loss, and the AGC amplifier assembly exhibits 55-dB gain. As the input increases, the attenuator nearest the output is activated. When the input has increased nearly 20 dB, the second attenuator from the last is activated. The attenuator closest to the input is activated last, and the effect of AGC on system thermal noise is minimized. The AGC range is 55 dB. Attenuation linearity versus the AGC voltage detected at the receiver output is shown in Fig. 5(c). The amplifier assembly achieves a 5.5-dB noise figure at the maximum gain. This design is very broad band, exhibiting less than 1.0-dB gain variation to 500 MHz. Harmonic distortion is less than -40 dB at the nominal input level of -15 dBm.

E. Sampler Assembly for Phase-Locked Loop

The directional coupler and sampler assembly shown in the receiver and transmitter phase-locked loops of Fig. 1(a) consist of a sampler substrate, a substrate including a 10-dB coupler and a bandpass filter, and a narrow interconnecting substrate which passes through a waveguide below the cutoff section similar to that discussed previously in Section III-B. The coupler/filter substrate is visible in the photograph of Fig. 6; the bandpass filter consists of four $\lambda/2$ sections coupled by interdigital capacitors [3]. The sampler substrate samples the 5.13–5.63-GHz LO at the synthesizer reference-oscillator rate. Suppression to -90 dBm of harmonics of the reference signal leading to the 4-GHz converter is accomplished by the bandpass filter and an interior cover which includes the waveguide feedthrough.

F. 4-GHz Power Amplifier

A photograph of the circuit assembly for the power amplifier is shown in Fig. 7(a). The basic substrate of this amplifier is a gain block consisting of two transistors in parallel. Each gain block has 4-dB compressed (class A) gain. As can be discerned from Fig. 7(a), the input signal to the power amplifier is split in a quadrature hybrid, amplified by two gain blocks, split again, amplified by two gain

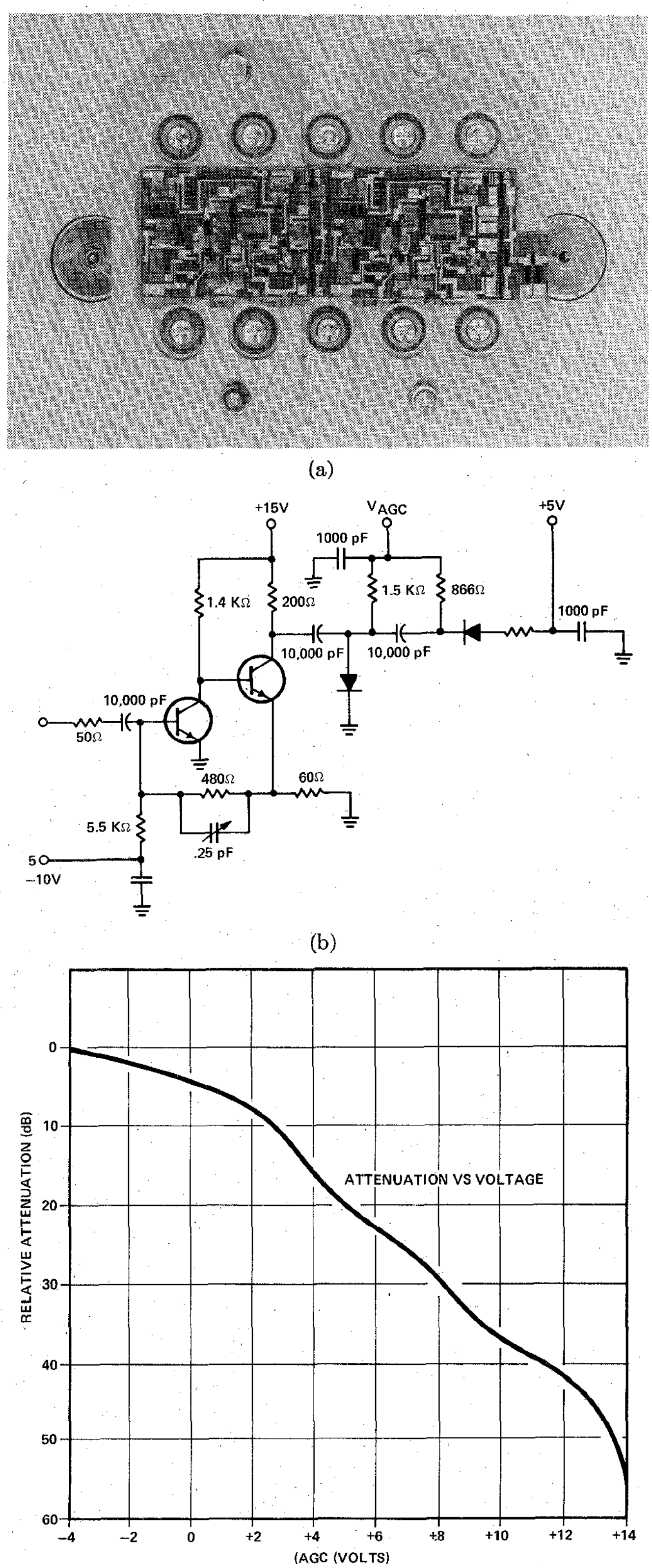


Fig. 5. 70-MHz AGC amplifier. (a) Photograph of the circuit assembly. (b) Schematic diagram. (c) Attenuation curve versus detected output voltage [see Fig. 1(b)].

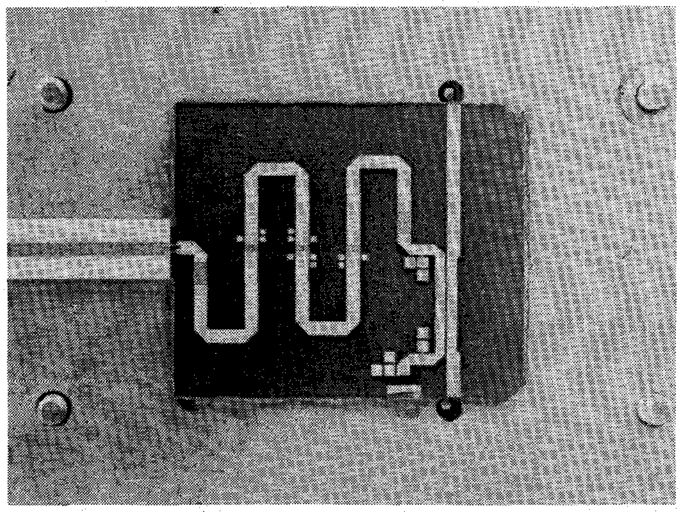


Fig. 6. Photograph of coupler filter for sampler assembly.

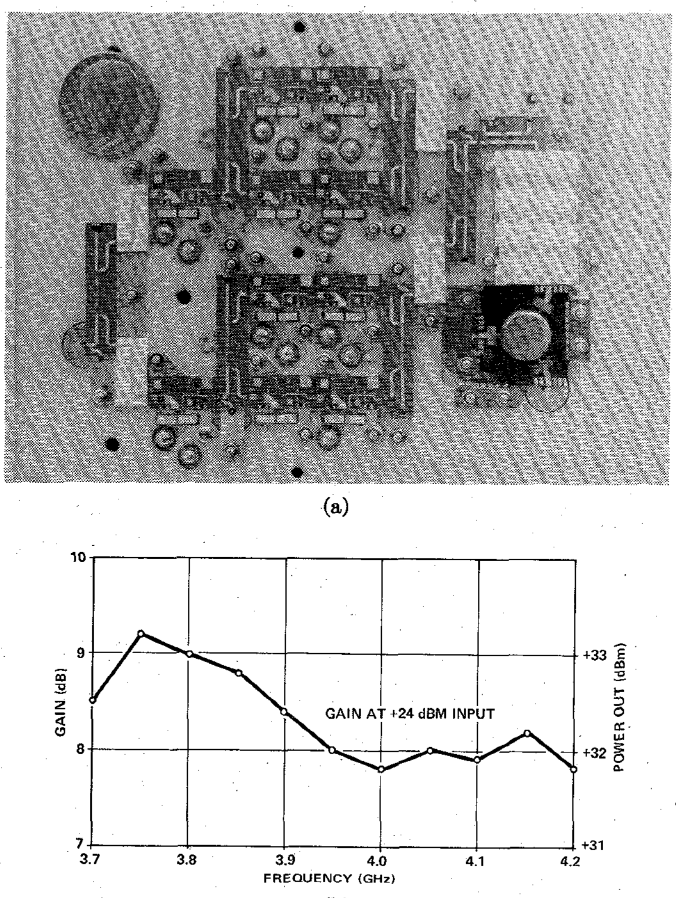


Fig. 7. 4-GHz power amplifier. (a) Photograph of circuit assembly. (b) Plot of gain and output power versus frequency.

blocks in tandem, and the four outputs are recombined in quadrature hybrids. Typical gain and output power versus frequency are plotted in Fig. 7(b). Interdigitated lines are used in the coupler to achieve sufficiently tight coupling. The circulator provides for output match and isolation. The three-section bandpass filter is of similar construction to the aforementioned filter of the 4-GHz down converter. The open-circuit stubs visible between the resonators suppress third harmonics.

G. 3.7–4.2-GHz Driver Amplifier

The driver amplifier uses two preamplifier substrates followed by five power-amplifier gain blocks in cascade to develop 48-dB gain at +25-dBm output power. The variation of output over the band is less than 1 dB.

H. Comments on the Use of Microcircuits—Advantages and Limitations

Over the past decade microelectronic technology in microwave design has met with considerable enthusiasm. Both manufacturers and users of microwave equipment have found these concepts which promise smaller size and potentially lower cost circuitry to be quite exciting. The advantages of microcircuit design are real enough, but so are the limitations.

Microelectronics circuitry does, in fact, allow signal-processing functions at microwave frequencies to be realized in relatively small-size packages. Minimizing the number of interconnecting cables provides for lower cost and avoids the accompanying performance degradation. Signal-processing functions can be realized at lower cost than with most other design approaches. Because only one package is required for many components, e.g., the 4-GHz converter, another cost advantage is realized. The design can allow for the heat sinking of high-power solid-state devices to the baseplate without the insulating effect of intermediate packages. Tunability of microcircuits is accomplished by designing in structures which are stepwise tunable. Tuning is convenient because the circuitry can be accessed without affecting performance. Lumped elements can be approximated, solid-state devices are easily mounted to substrates, and the microelectronic approach is compatible with the increasingly powerful techniques of computer-aided design.

The limitations of the microelectronic approach must be considered. Tightly coupled structures can be very difficult to achieve without introducing new problems that require a compromise of performance or cost. The baluns of the 4-GHz converter exemplify such a difficulty. However, through design innovation such problems are continually solved. Very high and very low characteristic impedances (greater than 100 or less than 30 Ω) are often more difficult to achieve as compared to coaxial or stripline structures. Narrow high-impedance transmission lines and thinner substrates require careful attention to tolerances. As evidenced by the microcircuit module designs discussed in this paper, higher order moding in the enclosures is an ever-present problem when large-level signal differences (greater than 30 dB) exist or low-level leakage is important. Such moding can cause undesirable feedback in amplifiers, resulting in instability. Passive components exhibit lower Q, and consequently higher insertion losses, than their coaxial, stripline, or waveguide counterparts. Another limitation of microcircuit technology is the requirement for specialized equipment to allow fabrication, assembly, and interconnection of the substrates. For example, sophisticated bonding machines are required.

In summary, microelectronic technology offers an additional tool for the microwave designer, but is not a panacea for design problems. Cost advantages are achievable on a function-for-function replacement basis when compared to other design techniques. Performance advantages are realized when those parameters which involve interconnections (i.e., phase and amplitude ripple) are involved. When low insertion loss or high Q (e.g., sharp filter cutoff) are required, microcircuit design should be avoided. The designer must consider all techniques for a given requirement and should not assume that microelectronic techniques offer the optimum approach.

IV. INSTRUMENT PERFORMANCE

As discussed in Section I, the 4-GHz PMR is a maintenance radio designed to operate as a temporary replacement for a microwave receiver bay or transmitter bay, or both. Much of the characterization of the PMR is in Bell System terminology [8]. The significant figures of merit for the PMR are the following: 1) noise performance in decibels above a reference level with C message weighting and measured at zero signal level (dBrnCO) (or, alternatively, noise power ratio in decibels) measured at baseband and including a Western Electric Type 4A FMT–FMR, and 2) group delay and amplitude flatness measured at 70 MHz. Both of the preceding measurements are performed with the PMR-transmitter output signal attenuated to the nominal input level of −28 dBm and applied to the receiver input. The measurements are then performed "back-to-back" using the 70-MHz ports to interface with the 4A terminal or a microwave-link analyzer.

The noise loading setup is shown in Fig. 8(a). A white-noise spectrum is low-pass limited, applied to a bank of stopband filters, and used as the input signal to the modulator (4A FMT) which provides the 70-MHz FM signal for input to the PMR. The output signal from the PMR receiver is applied to the FM demodulator, and the noise in each of the test slots is measured. Additional explanation of this measurement is found in [8].

The PMR receiver achieves an 8-dB noise figure and a 0.2-dB maximum amplitude variation over any 20-MHz band. Applying a carrier in the 3.7–4.2-GHz band at nominal level (−28 dBm) and FM demodulating the 70-MHz output carrier yields baseband fluctuation noise at −64 dBmO maximum from 300 kHz to 10 MHz. No spurious signals are visible above the thermal noise level.

The transmitter achieves a 1.5-W output with a 0.2-dB flatness over any 20-MHz band. By appropriate filtering (1430-MHz bandpass ahead of the 4-GHz converter input and 5.13–5.63-GHz bandpass in the sampler assembly), spurious signals are suppressed to 70 dB below the carrier.

Typical curves are shown in Fig. 8(b) of dBrnCO versus drive level to the 4A FMT for the setup shown in Fig. 8(a). Noise measurements are made for a 3-kHz slot with 1500-channel loading and Western Electric Type 457K

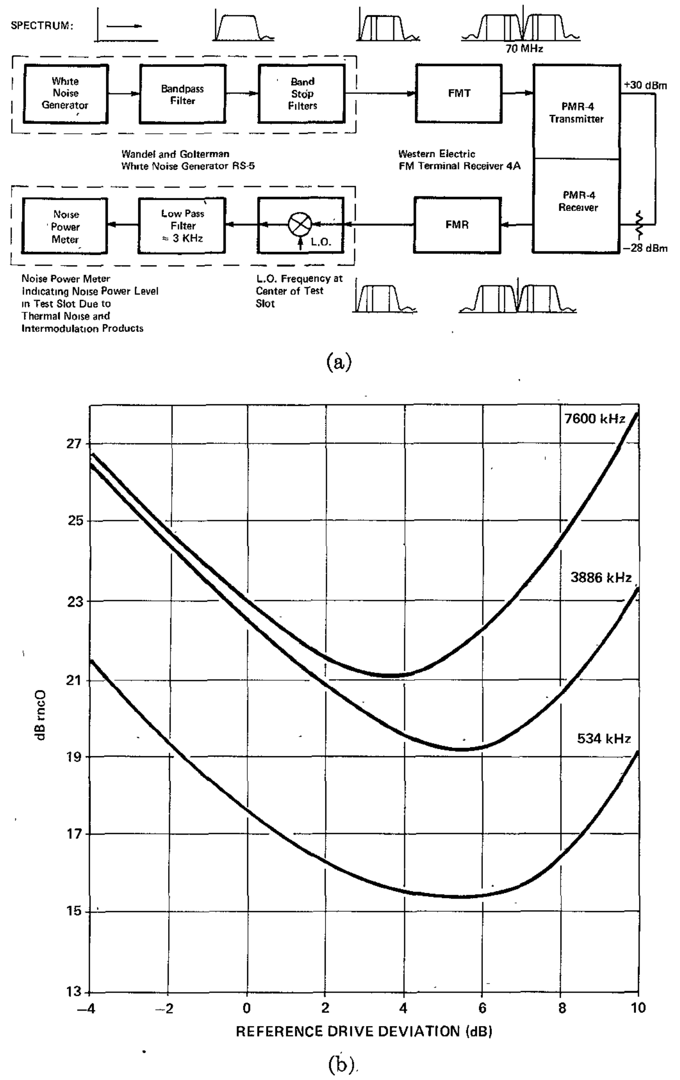


Fig. 8. FM noise-power ratio measurements. (a) Diagram of test setup. (b) Typical curves of noise power in dBrnCO versus baseband reference drive.

preemphasis. Test-tone deviation (0 TLP) is defined as 86.2-kHz rms/channel deviation. As indicated in Fig. 8(b), the nominal operating point of the PMR is thermally limited. This is due to the low FM-intermodulation distortion resulting from the broad-band nature of the instrument.

Group delay and flatness data are shown in Fig. 9(a) and (b). The curve of Fig. 9(a) is applicable to the use of the PMR for 1500-channel loading (7380-kHz maximum baseband frequency). The data in Fig. 9(b) are taken with the 70-MHz bandpass filter removed and single-section parabolic equalization applied to the 70-MHz output, demonstrating the potential use of the PMR for wide basebands.

By changing levels and injecting carriers at various points in the receiver system, it is possible to assign noise-power contributions to various mechanisms in the instrument. The significant contributors are the LO's, the thermal noise in the receiver, and the thermal noise in the transmitter. The noise powers are customarily referred to

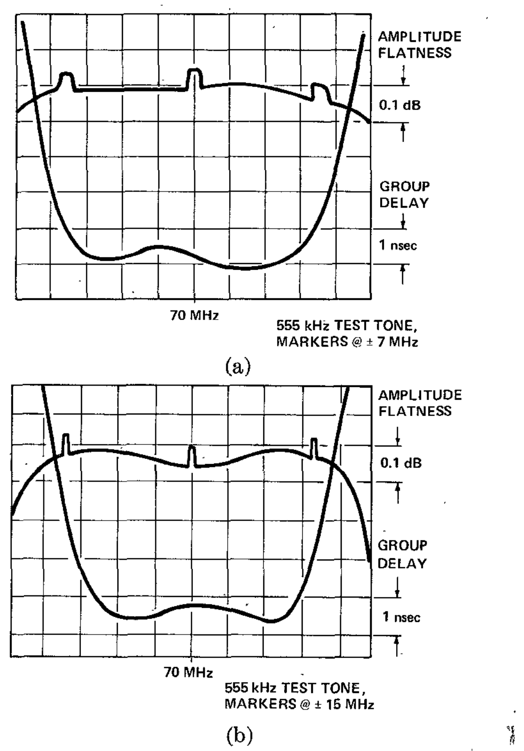


Fig. 9. Typical per channel amplitude flatness and group delay. (a) Response for a ±10-MHz channel. (b) Response for a ±15-MHz channel with 70-MHz bandpass filter removed and additional group-delay equalization provided.

TABLE II
MEASURED FM-FLUCTUATION NOISE OF PMR IN A FLAT 3-kHz SLOT WITH −30 dBm AT THE RECEIVER INPUT

	Baseband Slot:	534	3886	7600 KHz
Receiver:				
	Thermal Noise	13 pw	153 pw	210 pw
	5 GHz LO	9 pw	2 pw	1 pw
	1360 / 1500 MHz LO	9 pw	30 pw	36 pw
	Receiver TOTAL	31 pw	185 pw	247 pw
Transmitter	TOTAL	25 pw	46 pw	61 pw
PMR Back-to-back total Fluctuation Noise		56 pw	231 pw	308 pw

some 0-dB point of the baseband signal (0 TLP), and usually expressed in picowatts for a flat 3-kHz-wide bandwidth [8]. Table II summarizes the noise contributions. The relatively high contribution of the second LO is not uncommon for a crystal multiplier. The 4-GHz YIG-tuned oscillators exhibit very low noise contribution, indicating their usefulness in this application.

ACKNOWLEDGMENT

Design of the 4-GHz PMR represents the combined efforts of a team, too many in number to mention herein. The author wishes to thank R. Benson, J. Hall, H. Halverson, L. Niitani, V. Roland, R. F. Rawson, and N. Sugihara who have contributed directly to this writing. In particular, the author also wishes to thank E. J. Guertin of AT&T Long Lines.

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Reliability Testing of Microwave Transistors for Array-Radar Applications

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***Abstract*—Solid-state array radar is of great current interest because of the inherent reliability of solid-state devices and the concomitant promise for improvement in system reliability. However, no extensive reliability base has been established for solid-state devices employed under radar operating requirements. In this paper some of the important factors bearing a device reliability are treated. Accelerated life tests under RF conditions are presented for L-band power transistors. Preliminary life-test and failure-analysis data are also presented with recommendations on how the information can be used by the radar systems designer.**

I. INTRODUCTION

SEMICONDUCTOR devices are inherently more reliable than vacuum tubes because no materials are consumed during operation and high g applications can be satisfied. The diffusion profiles are stable for hundreds of years even at severe operating conditions; however, reactions between metal contacts and Si or SiO_2, manufacturing defects, oxide-stability problems, lack of adequate designed-in ruggedness, and other similar factors can limit the reliability of these devices. Failure rates of 0.001 percent/1000 h have been established for some devices. The success of the Minuteman missile program is an excellent example of solid-state reliability.

Solid-state radar requirements, however, impose unique stresses for devices. The devices must be qualified for these special operating stresses in order to establish confidence that premature system failure will not occur.

Bipolar microwave power transistors, similar to that shown in Fig. 1, are often employed in the class C amplifiers that comprise transmitter subsystems. These transistors are multicell devices which are subjected to severe stress (depending on power-level requirements) during pulsed-radar operation. For this reason, it is believed that these particular components are the critical elements in achieving long system life. This paper will outline efforts to establish reliability of L-band power transistors for phased-array radar applications.

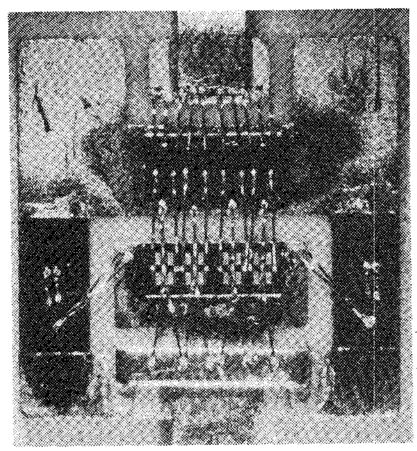

Fig. 1. Typical L-band bipolar microwave transistor.

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II. FACTORS AFFECTING RELIABILITY

There are three major classes of failures for any system. These are illustrated in Fig. 2 and consist of: 1) infant mortality—failures which can often be traced to defects; 2) random failures; and 3) wearout failures which inclued