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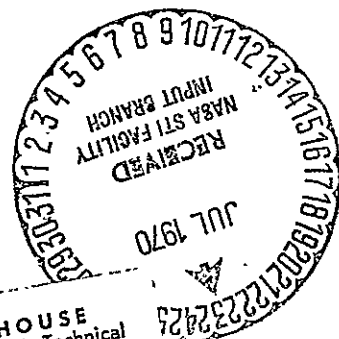
GROUND SIGNAL PROCESSING SYSTEMS, SUMMARY REPORTS ON ANALYSIS, DESIGN AND COST ESTIMATING

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NATIONAL AERONAUTICS AND SPACE ADMINISTRATION

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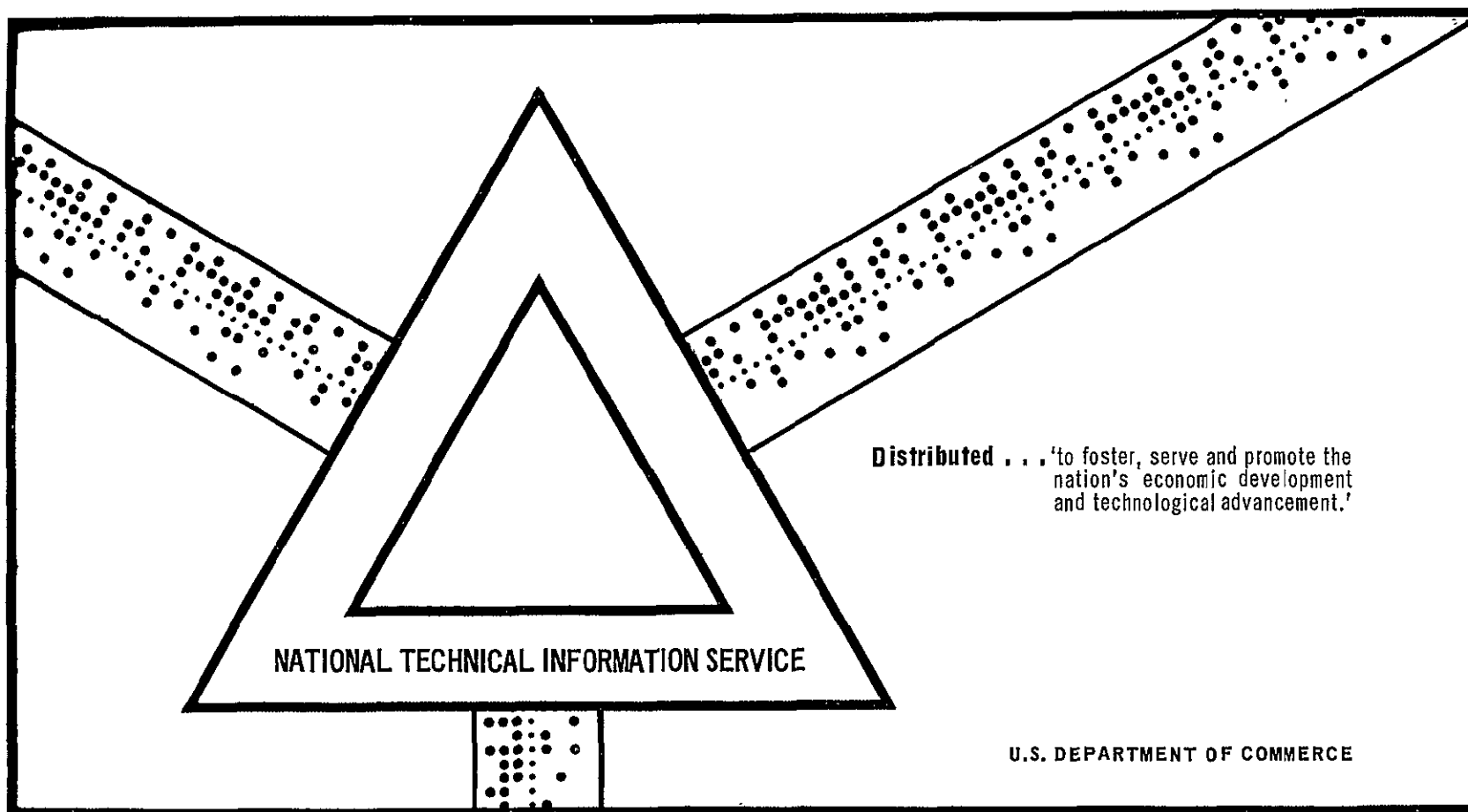
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DESIGN AND COST ESTIMATING**

**PREPARED FOR
NATIONAL AERONAUTICS AND SPACE ADMINISTRATION**

**Under The Technical Management Of
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1.0 INTRODUCTION

The purpose of this study, being performed under NASA Contract NAS 3-11520, is to investigate ground signal processing systems for television satellites as part of a general investigation of satellite television systems currently being conducted by NASA.

The program objective is to determine the practicability of using present state-of-the-art technology to develop low-cost signal processing techniques and devices for receiving low-power color television signals from satellites. Special consideration is to be given to minimizing the complexity and cost of manufacturing and installation while maximizing reliability and maintainability. The practicability of the selected techniques is to be demonstrated by the fabrication of prototype units, which will be tested in television receiving systems.

The program has been divided into two phases and nine major tasks or reporting categories. Phase I, which runs from July 1968 to November 1969, is essentially a study, design and cost determination program. Phase II runs from July 1969 to August 1970 and consists of the design, fabrication, test and delivery of prototype units.

The individual tasks comprising each phase are listed below:

Phase I:

- Task I. Candidate System Selection
- Task II. Subsystem Performance Evaluation
- Task III. Cost Screening
- Task IV. Conceptual Design and Optimization
- Task V-A. Design, Breadboarding, Test and Evaluation
 (initial Portion)*

Phase II:

- Task V-B. Design, Breadboarding, Test and Evaluation (Major Portion)
- Task VI. Production Design, Testing and Performance Evaluation
- Task VII. Fabrication and Acceptance Testing
- Task VIII. Shock and Vibration Testing
- Task IX. Program Management (extends over both phases)

* A portion of Task V was incorporated into Phase I because it was found that breadboarding provided support for cost determination and evaluation of critical performance parameters.

2.0 STATEMENT OF PHASE I PROBLEM AND SCOPE

The purpose of Phase I was to formulate ground signal processing designs which would meet the performance requirements and objectives of the contract while being suitable for low-cost manufacture in quantity. The designs were also required to be fabricated in prototype form for demonstration of the selected techniques.

The major performance requirements were to accept the satellite signals having the composition and power levels given in this section, and to deliver output signals to unmodified color television receivers that would result in picture quality corresponding to 35 dB video signal-to-noise ratio. Tolerable intermodulation distortion was specified as -40 dB. The envelope delay characteristics were the standard FCC specifications for color television. The signal processing units were specified to have gain versus frequency response within ± 1 decibel referred to baseband signals up to 4 MHz.

Approval of these designs by the NASA contracting officers was required before Phase I could be considered complete.

Four transmission formats for sending color television signals from a synchronous orbit satellite are defined in the contract. For convenience these formats will be referred to as systems, A, B, C, and D.

System	Picture Carrier	Video Modulation	Aural Modulation
A	2.25 GHz	AM/VSB - NTSC	FM Subcarrier - 4.5 MHz
B	2.25 GHz	FM, index of 2	4.5 MHz Baseband Subcarrier
C	8.45 GHz	FM, index of 3	4.5 MHz Baseband Subcarrier
D	12.00 GHz	FM, index of 3	4.5 MHz Baseband Subcarrier

System C, the 8.45 GHz system, was deleted in the latter part of the Phase I, when it became apparent that the techniques required were so similar to those of System D, the 12.00 GHz system, that completion of both systems was not justified.

The function of the ground signal processing system will be to receive the satellite signals having the above characteristics, and transform them to high-quality NTSC signals on one of the standard VHF channels for application to a conventional NTSC television receiver. This function is illustrated for FM transmissions in the generic block diagram shown in Figure 2-1. In the case of an AM satellite, only amplification and frequency conversion will be required.

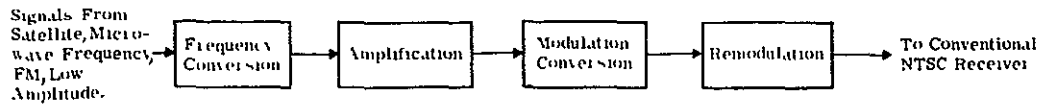


Figure 2-1. Ground Receiving Systems - FM Generic Block Diagram

The requirements and objectives of the ground receiving system are specified in some detail in the contract, NAS 3-11520, and are reproduced in Appendix I of this report. A summary of the major requirements and objectives is shown below, for the four types of ground receiving systems:

<u>System</u>	<u>Frequency, GHz</u>	<u>Nominal Antenna Output Power, dbW</u>	<u>Modulation</u>
A	2.25	-92	AM/VSB, per NTSC specifications.
B	2.25	-107	Composite visual/aural baseband signal is frequency modulated onto an RF carrier, with a modulation index of approximately 2.
D	12.00	-105	Same as "B" except that the modulation index is approximately 3.

3.0 APPROACH

3.1 Objectives

The objective of Phase I is to develop conceptual designs and production cost estimates for converters that meet specific performance requirements and objectives at minimum cost.

Detailed design performance requirements for the converter system are described in Appendix I.

Production cost estimates include direct factory material and labor charges; they do not include development costs, tooling costs, factory overhead and profit, shipping charges for finished converters, distributor's mark-up, retailer's mark-up, or installation costs and maintenance costs. A rule-of-thumb for estimating the retail price from the production cost estimates provided in this report is to double the cited costs. This adjustment will account for all excluded costs except for the amortization of development costs and installation and maintenance costs.

3.2 Basic Technical Approach

The Phase I effort was carried out according to the following procedure:

- 1) The contractual requirements and specifications were interpreted to define functional requirements or basic converter building blocks.
- 2) A survey of applicable component parts and materials was made. Lists of potentially usable component parts, which included general performance capabilities and cost data, were made.
- 3) Candidate system block diagrams were synthesized by matching the data from the first two steps in a manner that produced the required performance.
- 4) Circuit selection was made to meet the functional needs specified in the block diagram.
- 5) Preliminary cost estimates were prepared with the aid of a consumer-product-oriented cost analyst to permit a selection on the basis of cost from alternate configurations. The technical approaches and cost analyses were reviewed by an internal Cost-Performance Review Panel consisting of engineers who were experienced in the production of consumer and commercial electronics products and component parts.

- 6) A study of consumer electronics design, procurement and production procedures was made within General Electric, including field trips to the Radio Receiver, Major Television, and Integrated Circuits production plants.
- 7) Key cost areas were identified as a result of the first cost screening. Vendor contacts were made, to obtain updated performance and cost data on component parts, which could be used in the critical cost-performance areas. Semiconductor devices were obtained and experimentally evaluated to determine their acceptability and to obtain parameter measurements for the detailed circuit design phase.
- 8) Circuit designs were completed on critical functions such as the mixers, local oscillators, IF amplifiers, limiters and discriminators.
- 9) Breadboards were fabricated based on these designs and performance measurements were obtained.
- 10) Design changes and component part substitutions were incorporated in the breadboards to adjust for minor errors in performance and to determine the acceptability of reduced cost component parts. Revised cost estimates were prepared and reviewed by the Cost Performance Review Panel.
- 11) Conceptual designs were completed for all circuits required in the converter systems.
- 12) Final cost analyses were performed for all converter systems.

4.0 RESULTS AND CONCLUSIONS

Designs for all three types of ground receiving systems, specified in Contract No. NAS 3-11520, were formulated. These designs demonstrate that the production of consumer-oriented ground receiving units, meeting the performance requirements and objectives of the contract, is feasible and relatively straightforward.

The prototype units, which were specified in the contract for the demonstration of the practicability of the selected techniques, will be built during the contract. Conceptual designs for these prototype units have been formulated and approved by NASA. Block diagrams of the three types of ground signal processing systems are shown in Figures 4-1 through 4-3. A detailed discussion of these designs, together with more detailed schematic diagrams, is included in Section 5.0.

It is believed that the costs of consumer-oriented ground signal processing units can be held to values that are low enough to be consistent with the purchase requirements of individual consumers. A brief summary of these costs is shown in Table 4-1.

For microwave circuits and devices, e. g., step-recovery diodes, parametric amplifiers, bulk effect devices and tunnel diodes, costs will be largely determined by current market conditions. Thus, whenever it is necessary to estimate the costs of ground signal processing units that use such devices, the prices of these items are based on a best estimate of the market conditions existing at that time.

For lower-frequency circuitry, it was found that transistor-discrete component construction is considerably cheaper than integrated circuit construction: this advantage is likely to be maintained through 1975.

The AM converter, Figure 4-1, consists of a mixer, local oscillator, and IF amplifier, together with the power supply and antenna switch. The FM converters, Figures 4-2 and 4-3, are more complex. Additional stages of IF gain, demodulators and remodulators are required. The circuitry required in these added functions is, however, quite conventional.

Each converter consists of two separate packages, the antenna unit and the indoor unit. The antenna unit contains the mixer, local oscillator and IF amplification in all cases. The remainder of the electronics are contained in the indoor units, with the exception of a power regulator for the X-band local oscillator which is included in that antenna unit.

The use of two separate packages permits the placement of the frequency translation circuitry near the signal source and minimizes signal loss prior to the first amplifier. This configuration enables the converters to meet the noise performance objectives without RF preamplifiers. Also, operating the only

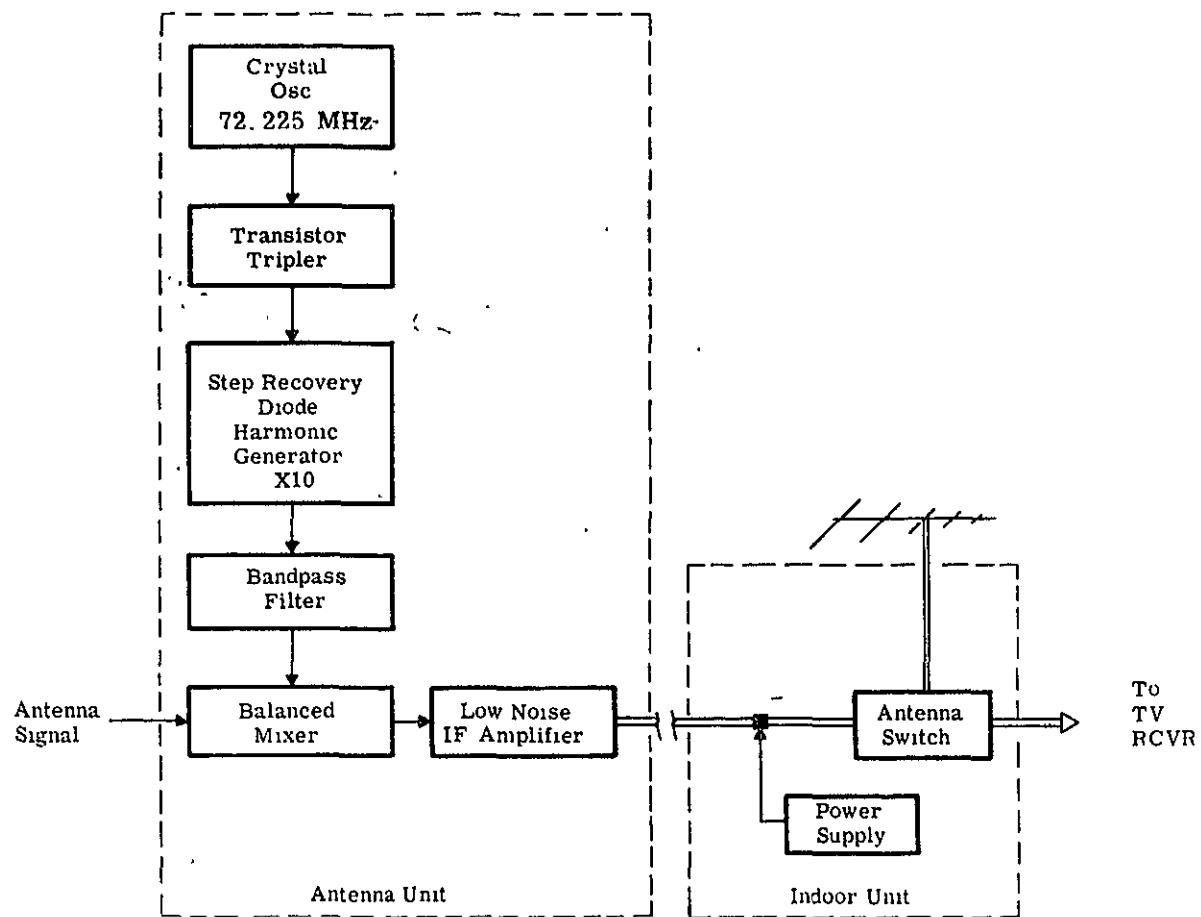


Figure 4-1. 2.25 GHz AM Converter (Block Diagram)

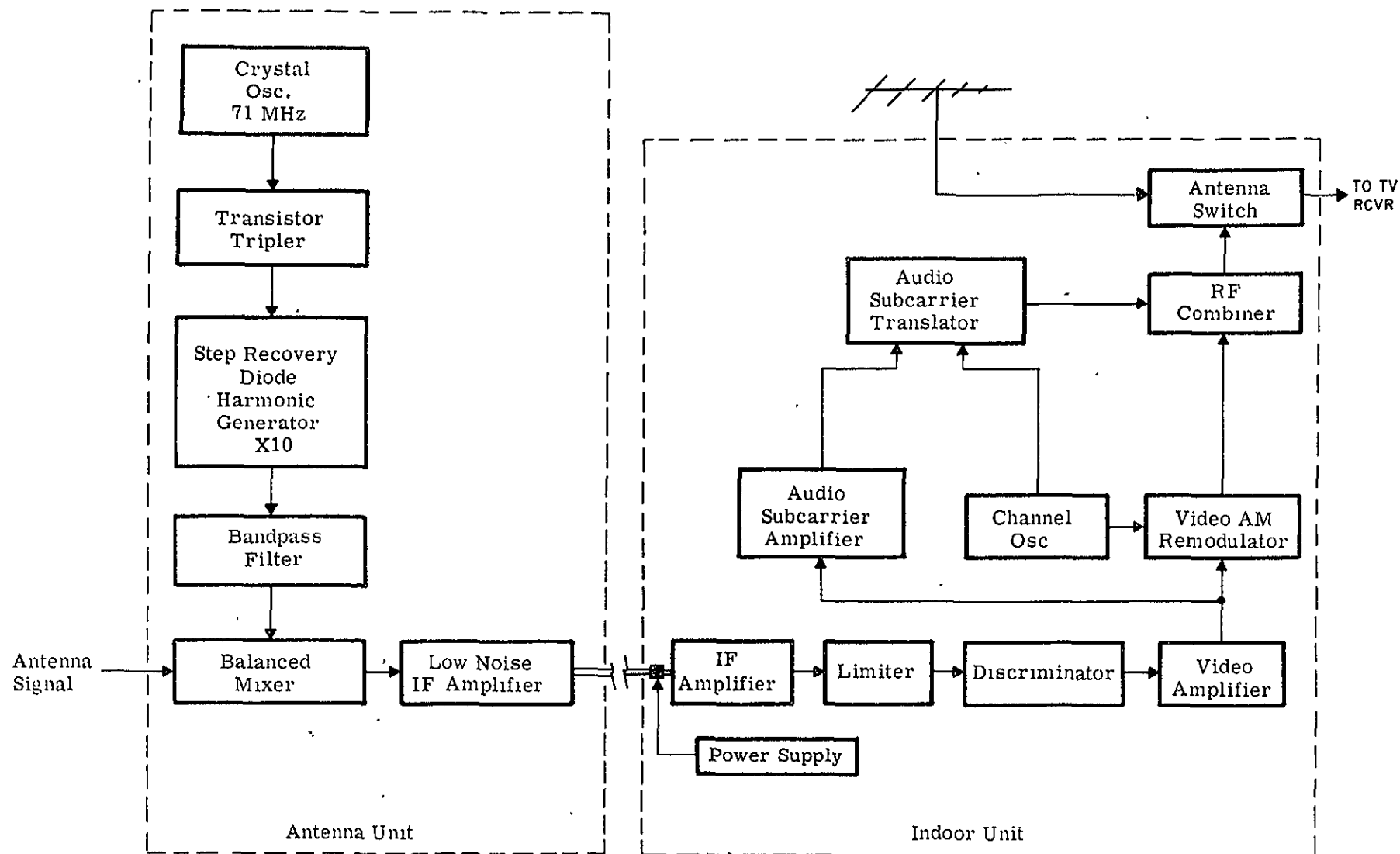


Figure 4-2. 2.25 GHz FM Converter (Block Diagram)

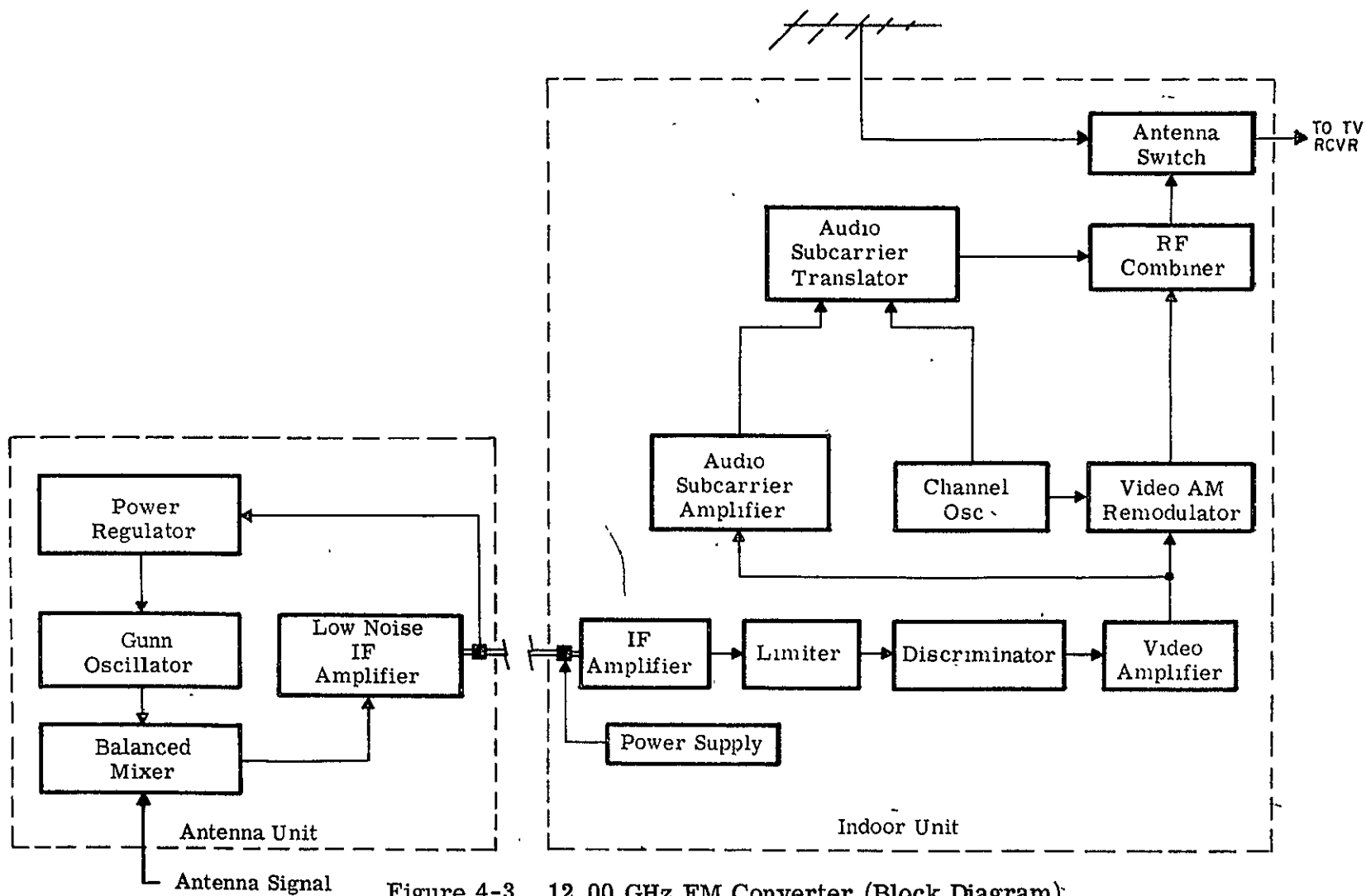


Figure 4-3. 12.00 GHz FM Converter (Block Diagram)

TABLE 4-1
MANUFACTURING COSTS VERSUS QUANTITY, FREQUENCY,
YEAR OF PRODUCTION AND YEAR OF PRODUCTION START

System	Year of Production Start	1969		1972		1975	
		Units/year 10 ³	Units/year 10 ⁶	Units/year 10 ³	Units/year 10 ⁶	Units/year 10 ³	Units/year 10 ⁶
2.25 GHz, AM	1969	29.47	16.51	19.93	10.92	14.73	8.01
	1972	-----	-----	25.53	13.91	19.00	10.30
	1975	-----	-----	-----	-----	25.10	13.53
2.25 GHz, FM	1969	44.38	25.66	31.66	18.13	24.18	13.69
	1972	-----	-----	39.91	22.71	30.48	17.24
	1975	-----	-----	-----	-----	39.38	22.66
12.00 GHz, FM	1969	119.71	47.96	64.03	27.25	37.18	16.29
	1972	-----	-----	70.68	31.15	42.48	20.44
	1975	-----	-----	-----	-----	49.79	24.47

interface between the two packages at the intermediate frequencies results in a minor signal attenuation in the interconnecting cable. The general complexity of the AM converter is about the same as a small transistor radio. The complexity of the FM converters is somewhat more than that of a table model AM/FM radio.

Installation of the indoor unit will require only screw terminal connections which can be accomplished without any special skill. Installation of the antenna unit can be made as a negligible addition to the installation of the antenna. The antenna unit can be attached to the antenna at the factory or service shop. Repair of both units will be by replacement to minimize service calls. Malfunctioning units may or may not be salvaged at the service shop or factory.

5.0 CONVERTER CONFIGURATIONS

The following discussion covers the candidate system designs for four converters and the conceptual designs for three of these.

5.1 Candidate Systems

The performance requirements and objectives for each converter system have been established by several references. The primary reference is the contract work statement from which the technical specifications and references have been extracted and compiled in Appendix I. The secondary references include (1) the NTSC standards for color television signal formats, (2) the FCC regulations regarding color television broadcasting, and (3) an uncompiled set of data concerning the operating characteristics of typical color television receivers designed to operate under the NTSC standards.

5.1.1 2.25 GHz AM Converter System

The 2.25 GHz AM system is required to accept a standard system M AM/VSB signal. The converter-receiver combination should furnish a video signal-to-noise ratio of 35 dB for an antenna output level of -92 DBW. Intermodulation distortion should be not greater than -40 dB and envelope delay as required by the standard FCC specifications for color television.

The main functional requirement of this converter is to translate an AM-VSB signal to a frequency that can be received by an unmodified NTSC receiver. The transmission format is such that there is no requirement for further signal processing.

The block diagram for the AM converter, in its most general form, is shown in Figure 5-1. The main design option is the inclusion of an RF amplifier. The component part survey results reported in Appendix VI, together with the results of noise calculations, were used to assign typical performance requirements and capabilities to the functional blocks. It was determined that the need for RF amplification, to establish the required converter noise figure, was marginal and it was decided to eliminate the RF amplifier in the AM candidate system.

The physical separation of the antenna and TV receiver dictated the need for frequency translation close to the antenna. The cable loss at the microwave frequency could not be tolerated. The mixer and local oscillator functions were, therefore, required at the antenna. The cable loss at the intermediate frequency required the use of either an RF preamplifier or an IF amplifier to establish the desired system noise figure. An IF amplifier was selected on the basis of lower cost and was included in the converter antenna unit.

An indoor unit is required for the power supply and for combining the converter and terrestrial broadcast TV signals prior to presentation to the TV receiver antenna terminals. The candidate system block diagram for the 2.25 GHz AM converter is shown in Figure 5-2.

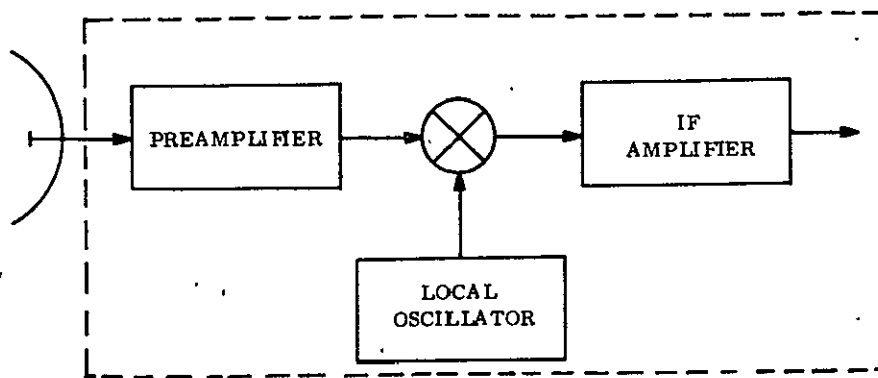


Figure 5-1. 2.25 GHz AM Converter (General Block Diagram)

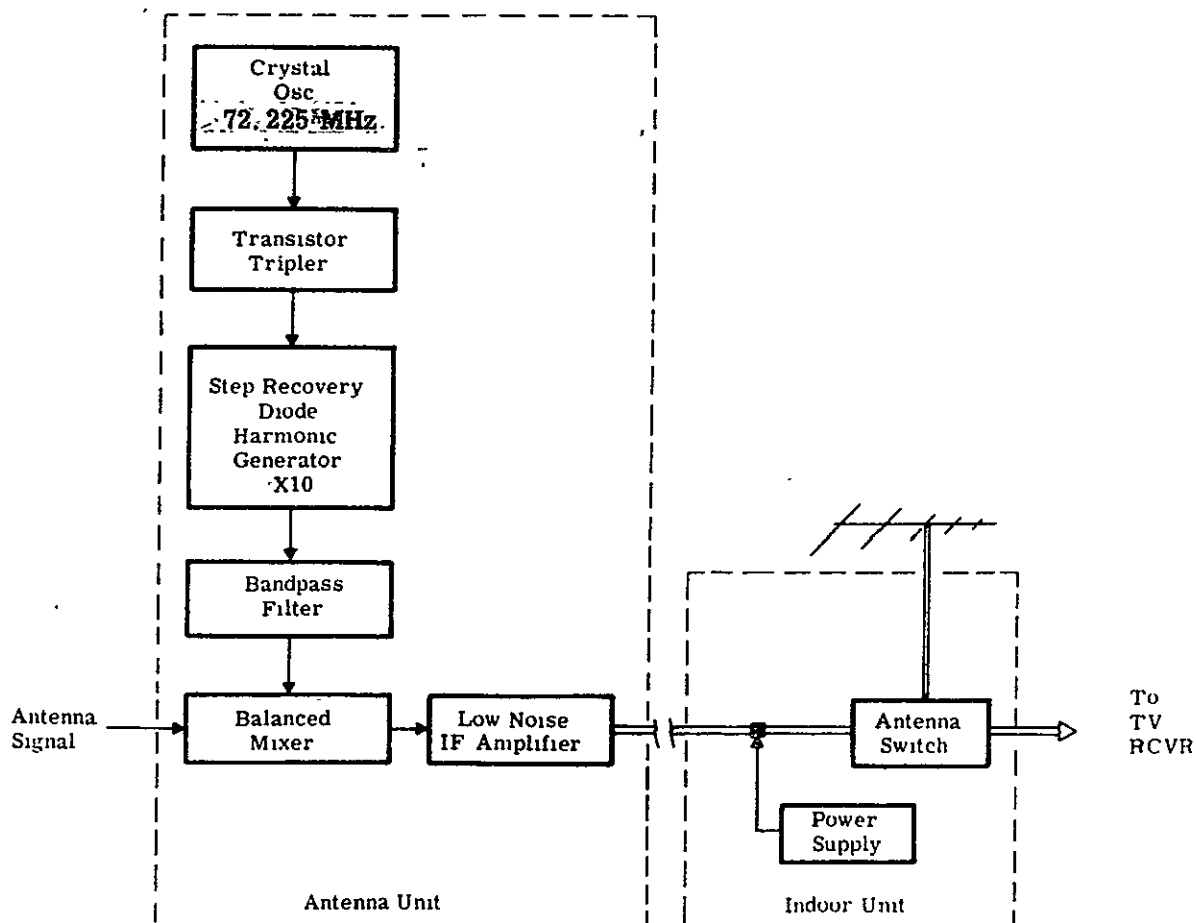


Figure 5-2. 2.25 GHz AM Converter (Block Diagram)

5.1.2 2.25 GHz FM Converter System

The constraints on the 2.25 GHz FM converter system design were as follows: The converter output should be received on an unmodified NTSC receiver. The video output signal-to-noise ratio of the converter-receiver combination should be at least 35 dB for an antenna output level of -107 DBW with a modulation index of 2. Intermodulation and envelope delay specifications were again -40 dB and equivalent to FCC specifications, respectively.

The FM converter requires demodulation and remodulation of the transmitted information. This provides an opportunity to depart from the standard composite video-aural subcarrier structure used in the AM system transmission. Work on different transmission structures was invited by a contract statement which allowed the consideration of alternate satellite transmission formats regarding the placement of the audio subcarrier signal.

An investigation of the FM signal spectrum characteristics was carried out to estimate bandwidth requirements and to define signal properties that might influence the selection of signal-processing options. It was determined that the combination of the color video baseband signal with the audio subcarrier, prior to frequency modulation, would result in potential intermodulation distortion between the audio subcarrier and the video sync tip and the audio subcarrier and the chrominance subcarrier signals. Configurations were considered that would solve this potential problem. Also, parallel analyses were performed to establish the minimum allowable audio subcarrier magnitude required in the satellite frequency modulating signal when using the combined video-audio structure. It was determined that the susceptibility to the intermodulation distortion (previously identified) would be minimal and that the added converter complexity required to process other formats was not warranted.

The baseband signal structure selected consisted of the standard M system composite video signal plus a reduced-amplitude 4.5 MHz aural subcarrier.

The system constraints on output signal-to-noise and modulation index were such that the discriminator is required to operate above threshold to achieve the target output signal-to-noise ratio.

Once the FM transmission format was established, the selection of converter candidate systems for this application was reduced to the choice of the best FM demodulation method. The candidates for this process were conventional as well as frequency feedback discriminators. The feedback discriminator could be used to provide direct conversion from the RF/FM signal to baseband video. This approach would require substantial RF amplification with AGC; therefore, the cost was considered prohibitive. The feedback discriminator could also be applied at an intermediate frequency with the potential of some FM threshold reduction. The combination of a specified modulation index and an available noise margin was such that the threshold improvement was not considered necessary.

The other FM discriminator considered, and selected, was a conventional discriminator preceded by a limiting IF amplifier. The key factors in the applicability of this approach were the available discriminator linearity as a function of the selected intermediate frequency, and the relative cost of gain and low noise figure at different intermediate frequencies. Preliminary tests indicated that adequate discriminator linearity, inexpensive gain, and low IF noise figure could be obtained with an intermediate frequency in the vicinity of 120 MHz. This intermediate frequency selection was checked against the predicted mixer noise performance and it was determined that the FM converter system noise figure could be established within the available noise margin without RF preamplification. If this had not been the case there would have been two options open. First, an RF preamplifier could have been used to establish a lower converter noise figure at additional cost. Second, the intermediate frequency could have been increased to permit more effective image noise suppression in the mixer. The latter option would have resulted in a higher IF noise figure and/or cost. The results of these trade-offs are shown in the 2.25 GHz FM converter candidate system block diagram, Figure 5-3. A separate gain channel is required for the audio subcarrier signal because the transmitted audio subcarrier level was reduced to the minimum level required for audio channel output signal-to-noise of 30 dB. If the video and audio subcarrier signals were processed in a common channel, the audio subcarrier level would not be commensurate with the AM broadcast requirements for intercarrier sound reception in conventional receivers.

The selection of an approach for the remodulation requirement offered few options. The decision to specify minimum audio subcarrier magnitude in the composite video-audio frequency modulating signal required the use of a separate audio subcarrier gain channel in the FM converter signal processor. This requirement opened up a secondary option in the remodulator concerning the NTSC AM signal formulation. The video and audio signals could be combined prior to AM remodulation or by combining the RF outputs of separate video and audio modulators. It was established that conventional NTSC color receivers would function properly when presented with a double sideband AM video signal, as opposed to the vestigial sideband signal that is normally transmitted. No vestigial sideband filter was specified for the video AM remodulation in the converter. It was also decided to use a separate remodulator (translator) to position the audio subcarrier signal in the converter output spectrum. Both of the remodulation functions can and should be derived from a common carrier oscillator. This assures the proper frequency relation between the reconstituted AM carrier and the audio subcarrier, as required for intercarrier sound detection in typical NTSC color television receivers.

The use of separate remodulators in the FM converter eliminates potential intermodulation distortion and permits more precise control of the critical video amplitude modulation requirement concerning percentage modulation at white level. The image sound subcarrier signal resulting from the translation process will be suppressed with a filter.

The power supply and antenna signal combining functions are also required, as in the 2.25 GHz AM system.

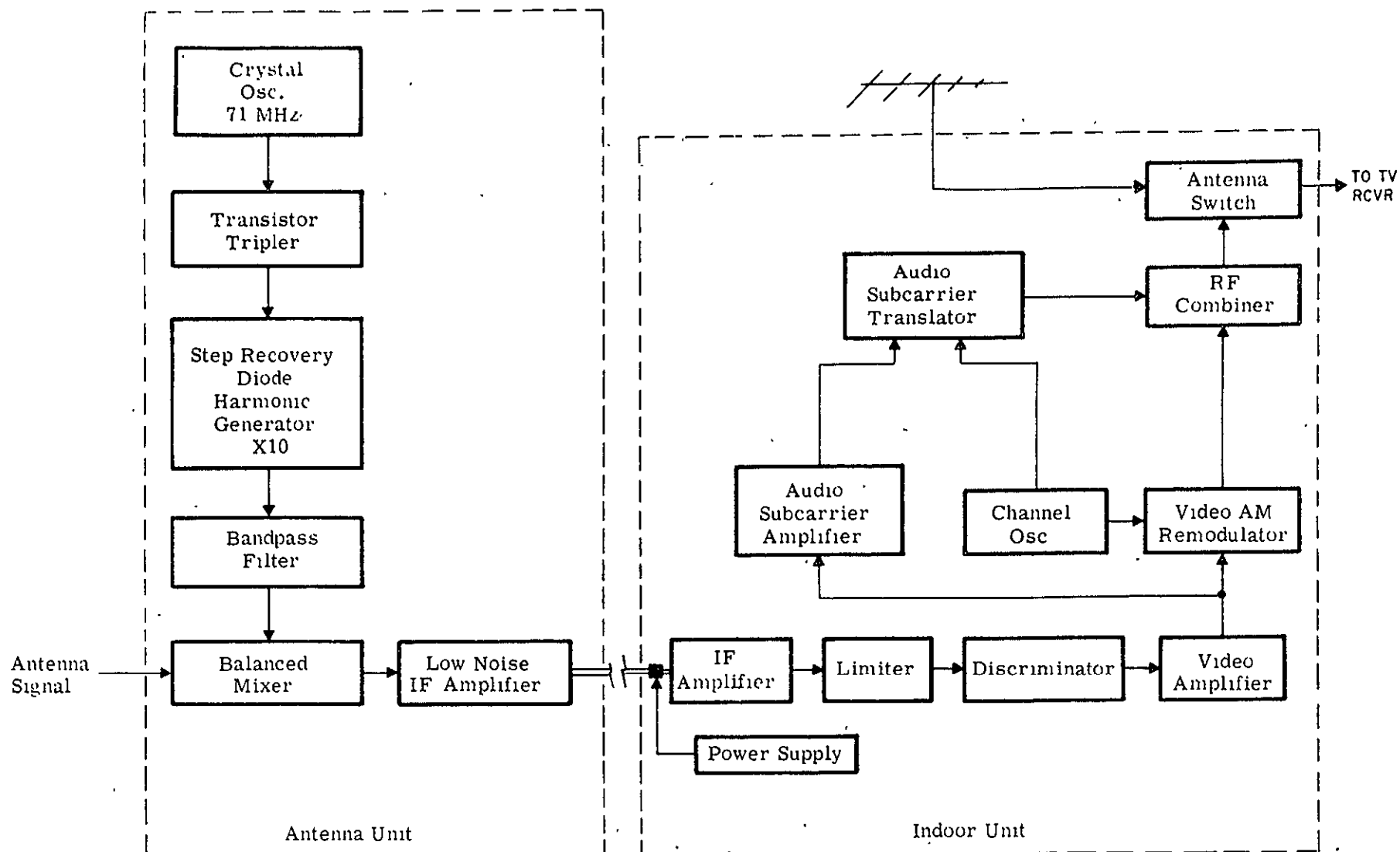


Figure 5-3. 2.25 GHz FM Converter (Block Diagram)

The partitioning of the block diagram functions between the antenna location and the indoor receiver location follows the same logic as in the AM system with regard to the establishment of the receiver system noise figure. The FM converter will require the front-end functions through the first IF amplifier stages at the antenna. The remainder of the IF amplifier, the FM demodulator, and the remodulation functions will be located at the television receiver location, along with the power supply and antenna interconnections. The candidate system block diagram selected for the 2.25 GHz FM converter system is shown in Figure 5-3.

5.1.3 8.45 GHz and 12.00 GHz FM Converter Candidate Systems

There are only three distinctions between the S-band and X-band FM converter system specifications; the obvious carrier frequency difference, an increase in the FM modulation index for the X-band systems to three, and an increase of the antenna output level to -105 dBW for the realization of the 35 dB output signal to noise ratio.

The functional blocks required in the higher-frequency FM converter are similar with the exception that a Gunn oscillator replaces the crystal oscillator and multiplier chain as the local oscillator.

The best choice for the intermediate frequency and FM demodulator is affected by the higher carrier frequency. The separation of the local oscillator and carrier frequencies, on a percentage basis, is reduced by a factor of 3.7 and 5.3, respectively in the 8.45 GHz and 12.0 GHz systems, from the value in the 2.25 GHz FM systems when the same intermediate frequency is retained. This imposes tighter specifications on the mixer and local oscillator designs. The question arises whether the intermediate frequency should be increased in the X-Band FM systems, at additional IF production cost, or whether the same IF should be retained, with a possible increased mixer-local oscillator cost resulting from the more stringent specifications. The decision was made to keep the same intermediate frequency because the cost of maintaining the more stringent fabrication tolerances is less than the increase for semiconductor devices that would result from the use of a significantly higher intermediate frequency.

The block diagrams for all FM converter systems are generically the same. The decision was made by the NASA Project Manager at this point to eliminate the 8.45 GHz system from further consideration and to proceed with the other two. The remaining distinctions between the 2.25 GHz and 12.00 GHz FM system block diagrams lie in the local oscillator and mixer functional requirements and methods of implementation and the need for a slightly increased IF bandwidth in the X-band system as a result of the larger modulation index specified for that converter system. The 12.00 GHz FM converter system block diagram is shown in Figure 5-4.

5.2 Circuit Considerations

5.2.1 2.25 GHz AM Circuit Considerations

The next step in the converter design was the selection of circuits and the allocation of performance requirements to the functional blocks identified

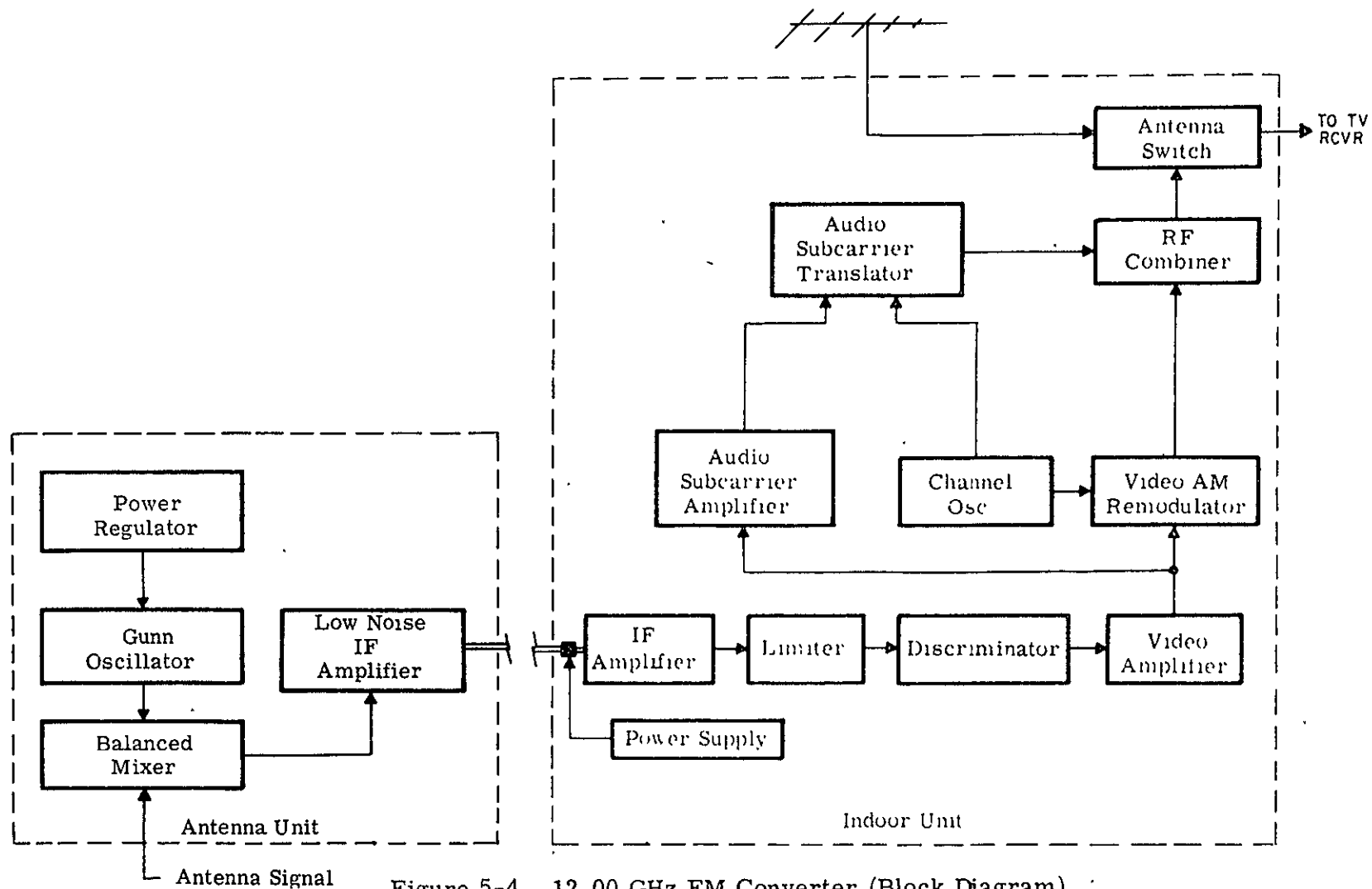


Figure 5-4. 12.00 GHz FM Converter (Block Diagram)

in the candidate systems. The performance parameters for the 2.25 GHz AM converter are listed in Figure 5-5. These values were obtained by matching the available device capabilities with the overall converter requirements. The conceptual design is shown in Figure 5-6.

5.2.1.1 2.25 GHz AM Mixer

The choice of the mixer diode type was simplified by the outstanding cost performance capabilities of the Schottky or hot-carrier diodes. These devices have established themselves as the standard in microwave mixer applications. The low conversion loss and low cost of these diodes cannot be matched by any competing device. The remaining selection concerned the mixer type. The candidates were single diode mixers, balanced mixers and doubly balanced mixers. The single balanced mixer was chosen as the best compromise in performance and total converter cost. The balanced mixer, using a 3 dB hybrid, provides a natural directive coupler for signal and local oscillator power, with good isolation between the two ports. The main disadvantage of the single diode mixer was the problem of local oscillator power insertion. In order to prevent signal power dissipation in the local oscillator insertion port, the coupling between the mixer and the local oscillator would have to be loose. This would require a significant increase in the level of local oscillator power, which is one of the most expensive functions in the converter design.

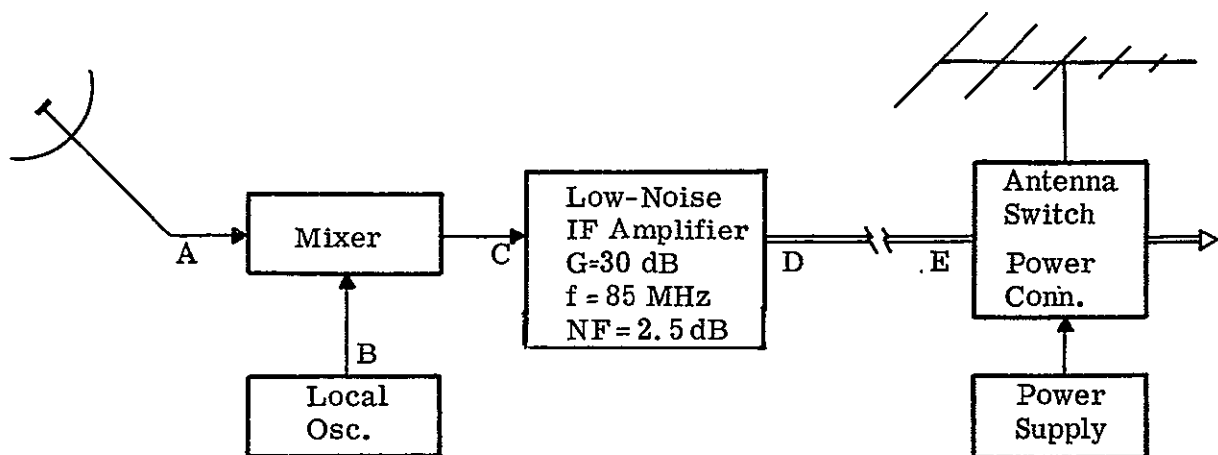
The additional advantage of increased isolation between the local oscillator and signal ports, obtained with a doubly balanced mixer, was not considered necessary. The suppression of local oscillator radiation can be obtained, at a lower cost, with filters.

5.2.1.2 2.25 GHz AM Local Oscillator

The requirements on the 2.25 GHz system local oscillator that are most critical are frequency stability and output power level. The tolerable long-term oscillator frequency drift was determined to be ± 100 kHz. The nominal local oscillator power level required at the mixer is 5 dBm.

Long-term drift in the AM converter local oscillator would appear as an apparent frequency shift of the television receiver local oscillator or the channel being received by the television set. Television receivers use an AFC loop to correct for internal local oscillator frequency drift. This control can also be used to correct for minor frequency shifts in the converter local oscillator. The tolerable drift of the converter output signal, in this situation, is limited by the RF tuning of the television receiver front end rather than by the tracking range of its local oscillator AFC loop. It is estimated that a frequency shift of ± 100 kHz could be tolerated by the VHF tuner RF circuits in a typical television receiver.

The local oscillator power level affects the operation of the mixer, especially if the mixer diodes are operated in a self-biased mode. The effect is caused by a change in the diode impedance with bias level, which in turn changes the mixer noise figure. The optimum local oscillator power level setting is relatively broad and a one-time adjustment of power from the local oscillator is expected to be adequate.



	Minimum			Nominal			Maximum		
	dbw	V/50Ω	V/300Ω	dbw	V/50Ω	V/300Ω	dbw	V/50Ω	V/300Ω
A	-110	22μv		-92	175μv		-80	695μv	
B	-30	.22 v		-25	.38 v		-23	.49v	
C	-116	11μv		-98	88μv		-86	358μv	
D	-86	350μv	850μv	-68	2.75mv	6.75mv	-56	11mv	27mv
E	-87	310μv	750μv	-69	2.5mv	6.1mv	-57	9.8mv	24mv

Figure 5-5. 2.25 GHz AM Converter System Block Diagram with Assigned Signal Levels

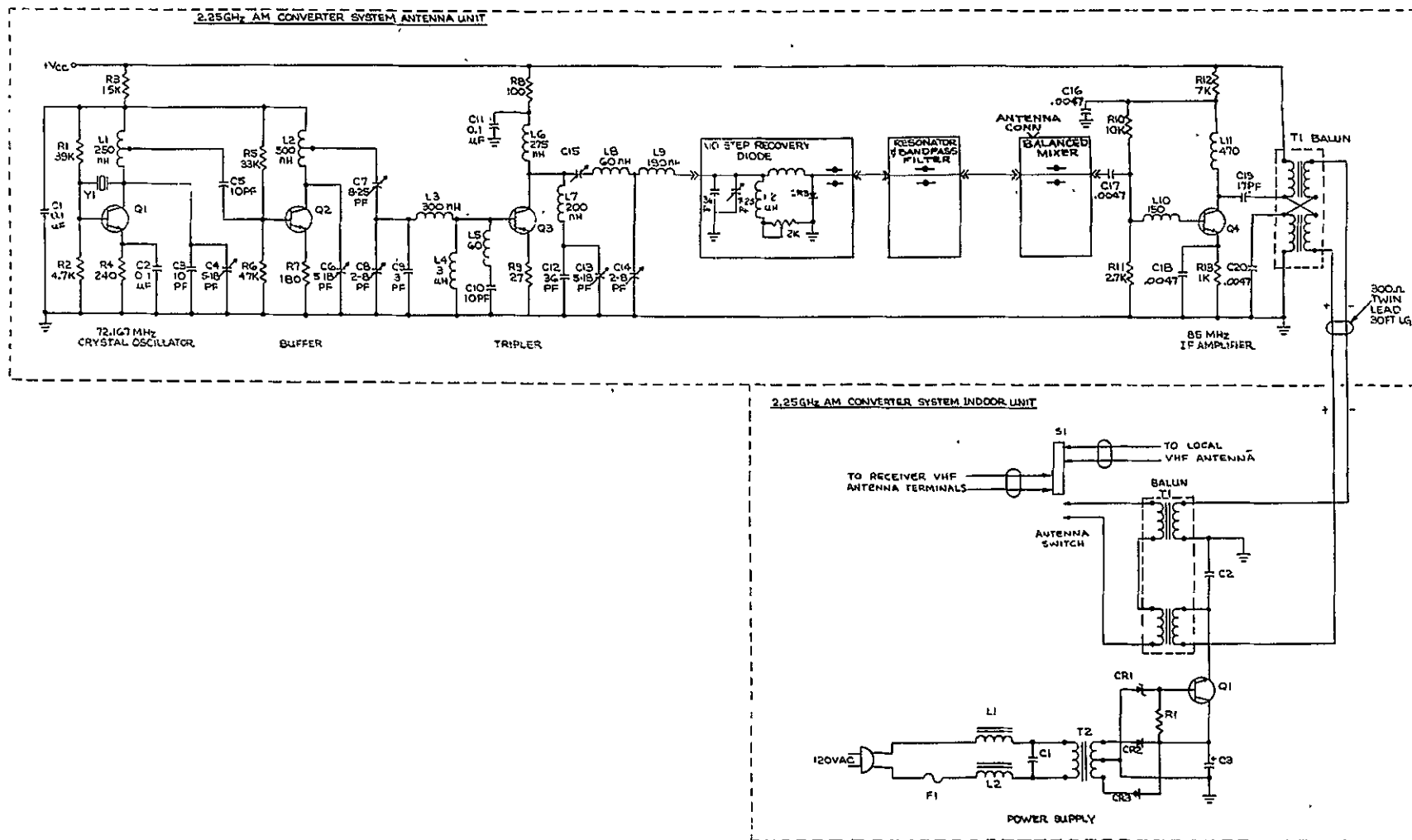


Figure 5-6. 2.25 GHz AM Converter System Conceptual Design

Two methods were considered to obtain the L. O. frequency stability: a crystal-controlled local oscillator and an oscillator with an AFC loop. The crystal oscillator approach requires the use of frequency multiplier circuits because the frequency of inexpensive crystals is limited to approximately 100 MHz at the upper end. On the other hand, the AFC approach requires a high-gain amplifier at the intermediate frequency, a frequency discriminator, and a voltage-controlled oscillator. The initial cost screening indicated that the device cost for a direct oscillator without AFC was greater than the total cost of an oscillator-multiplier chain. Recent price reductions in the microwave transistor market have made the direct transistor oscillator competitive without AFC; however, the added cost of the AFC loop components in the AM system have kept this option out of contention.

Having selected the oscillator-multiplier approach, the problem of optimizing its cost-performance still remained. The crystal oscillator frequency was chosen as high as practical, from the viewpoint of minimizing the crystal cost and interference in the IF passband from the oscillator harmonics. A crystal frequency of 72.167 MHz was selected. A multiplication factor of thirty was therefore required to reach the desired local oscillator frequency of 2165 MHz. This choice of frequency resulted in a converter output within the channel 6 VHF allocation.

The multiplication factor is provided by a transistor tripler, followed by a step-recovery diode circuit that generates maximum output at the tenth harmonic of the drive signal. A tripler was chosen as the intermediate stage after considering the power level, and the cost of the transistors available for this application. Another consideration was the realizability of the filter required to suppress undesired harmonics at the final multiplier output.

The tripler circuit provides three functions: (1) it increases the frequency separation between adjacent harmonics at the final multiplier output, (2) it increases the efficiency of the final multiplier by reducing the multiplication factor, and (3) it provides gain to compensate for the final multiplier conversion loss.

The tripler circuit provides, at least, 10 decibels of power gain in addition to the frequency tripling function. This is net gain between the drive level and third harmonic output level. The power input to the tripler will be in the range of 10 mW. A buffer stage is used between the crystal oscillator and tripler to boost the crystal oscillator output and to prevent frequency pulling during the alignment of the tripler.

The final multiplier uses a step recovery diode (SRD) followed by a bandpass filter. The SRD is the most efficient method of higher-order harmonic power generation that is available. The circuit requirements for this application permit the use of an axial lead diode, which minimizes cost.

The harmonic rejection filter will have an insertion loss of about 3 dB. The power level of the desired harmonic, required at the output of the SRD multiplier is, therefore, in the range of 10 mW. The conversion efficiency of the SRD multiplier is about $100\%/N$, where N is the multiplication factor, in this case 10. The drive level at the tripler output is, therefore, at the 100 mW level. The SRD multiplier is designed to produce a continuous change in output power at the desired harmonic as the drive level is changed. The adjustment in local oscillator power will, therefore, be made at the SRD input.

5.2.1.3 2.25 GHz AM System-IF Amplifier Circuit Considerations

The parameters considered in the selection of the IF amplifier for the AM converter were (1) the intermediate frequency, (2) the IF gain required, (3) the IF bandwidth, (4) the IF amplifier noise figure, (5) the mixer-IF amplifier impedance match, and (6) the IF amplifier device cost. All of these parameters are interrelated. The types of devices considered were discrete bipolar transistors, discrete MOS transistors and integrated-circuit amplifiers.

The low noise figure requirements, of the order of 2.5 dB for the IF amplifier and subsequent components, eliminated integrated circuits from further consideration at this time.

The mixer-IF amplifier impedance match could best be provided with discrete bipolar transistors. This conclusion results from the typical IF impedance of the mixer diodes at the desired operating point. Each diode produced an output IF impedance in the vicinity of 150 ohms. The effective parallel impedance was such that the lower impedance of the bipolar transistors was preferred.

The bandwidth of the IF amplifier in the AM converter system is not critical because the front-end circuits in the television receiver provide adequate selectivity. The converter IF amplifier was, therefore, selected to have a relatively wide bandwidth with respect to single-channel requirements.

The device and circuit selection was, therefore, reduced to the tradeoff of intermediate frequency and gain, and the cost of bipolar transistors that would establish IF noise figures in the range of 2.5 dB.

The gain in the AM converter/IF amplifier combination, which is required to establish the system noise figure, was determined to be 20 dB, assuming a device noise figure of approximately 2 dB. It was found that a single-stage transistor amplifier would fulfill these requirements if the intermediate frequency was selected to coincide with the channel 6 VHF allocation. The cost of the device was forecast in the range of \$.16 per unit, at the annual procurement level of one million units by 1974.

5.2.1.4 2.25 GHz AM Converter Cable and Indoor Unit Consideration

The remaining functional requirements for the AM converter system include the signal lead-in cable and the indoor unit antenna connections and power supply.

The lead-in cable provides a means for supplying the dc power from the indoor power supply to the antenna unit circuits. The choice of twin-lead cable versus coaxial cable was made to minimize the cost, with the realization that the coaxial line may be preferred from a reliability and shielding point of view.

A balun is used to convert the single-ended IF amplifier output to a balanced line configuration at the antenna unit end of the lead-in cable.

This balun also converts the 300-ohm line impedance to 75 ohms at the amplifier output. A balun is also used, at the other end of the twin lead, as a convenient means of coupling the dc power while maintaining a 300-ohm to 300-ohm impedance match that is suitable for the VHF television receiver input.

An antenna switch was selected to connect the standard ground antenna, or the converter output signal, to the television receiver without mutual interference.

The power supply was designed to provide regulation of the dc output voltage, and isolation from the 115 V ac line.

5.2.1.5 2.25 GHz AM Converter Fabrication and Hardware Considerations

Waveguide, stripline and microstripline construction methods were considered for the microwave circuits. These circuits include the mixer and the local oscillator SRD* multiplier and filter. Waveguide fabrication, including cavity structures, was mainly considered for the potential application of bulk effect or transit time devices as local oscillators. Further consideration of this method of construction was dropped once the proposed local oscillator approach was selected. The choice between microstripline (single ground plane) and stripline (double ground plane) construction was made on the basis of the circuit impedance levels that were available in the two construction methods, the radiation susceptibility of the microstripline circuits, and circuit producibility. Practical microstripline circuits require high dielectric constant materials, such as alumina ($\epsilon \approx 9$), to minimize the influence of stray field patterns. The materials that exhibit the desired properties are ceramics or brittle substances, which are not well suited to volume production and assembly techniques. The higher dielectric constant value also requires tighter circuit etch control because circuit dimensions are reduced. The availability of stripline circuit materials is less limited, as indicated in the materials survey lists of Appendix VI. The stripline circuits are physically larger, because the dielectric constants of the materials are generally lower, leading to increased basic material costs. However, circuit tolerances are increased and less expensive etched circuit processing is required. The materials considered for stripline circuits are generally amenable to cold punching, shearing and other volume production processes. Stripline construction, using PPO dielectric, has been selected for the 2.25 GHz microwave circuits. The remainder of the 2.25 GHz AM converter electronics will be fabricated using a lower-grade copper-clad material such as glass-epoxy, G-10.

The preliminary designs for the antenna unit and indoor unit enclosures are discussed in Appendix VI. These designs use low cost in injection molding or extrusion of plastics as the basic fabrication method. The selected plastics can be metallized to provide shielding in the antenna enclosure, or this enclosure can be fabricated from a metal casting at approximately the same cost.

* SRD - Step-recovery-diode.

5.2.1.6 2.25 GHz AM Converter Candidate System Summary

The 2.25 GHz AM system, as shown in Figure 5-6, consists of a singly balanced Schottky diode mixer, a local oscillator consisting of a crystal controlled transistor oscillator and multiplier chain, and a low-noise transistor IF amplifier centered at 85 MHz. These circuits are packaged in an antenna mounted unit. An indoor package is also required which consists of a power supply and antenna interconnection circuitry. The microwave circuits are constructed in stripline form while the lower frequency circuits are fabricated on standard printed circuit boards. The enclosures used in mass production would most likely be case metal, although metallized plastic extrusions or castings may also be used.

5.2.2 2.25 GHz FM Converter Circuit Configurations

The selection of circuits for the 2.25 GHz FM converter was, in some cases, based on criteria that were almost identical to the AM converter. The distinctions in the circuit selection procedure are discussed below. The candidate system block diagram is shown in Figure 5-7. The assigned performance parameters are listed in Table 5-1. The 2.25 GHz FM converter conceptual design is shown in Figures 5-8 and 5-9.

5.2.2.1 2.25 GHz FM Mixer

The performance requirements of the FM and AM S-band mixers were almost identical; therefore, no distinction was required.

5.2.2.2 2.25 GHz FM Converter Local Oscillator Considerations

The main distinction, in the selection of a local oscillator for the FM S-band system, was the inherent existence of the basic components required for an oscillator AFC loop within the FM signal processing chain. This provides an advantage for the application of an AFC controlled oscillator in the FM system versus the AM system. The question of interest, however, is whether the AFC oscillator in the FM system provides a cost-performance improvement over the crystal-controlled oscillator. At the time of the initial cost screening, the advantage was in favor of the crystal oscillator-multiplier approach. As a result of dramatic reductions in S-band transistor prices in the past two months, the advantage is no longer as clear cut. However, the dynamic status of semiconductor costs is such that the cost of the multiplier chain is more likely to be reduced in the future, and the same approach will be used in the 2.25 GHz FM system local oscillator design as was used in the AM converter system.

TABLE 5-1

ASSIGNED SIGNAL LEVELS FOR THE
2.25 GHz FM CONVERTER SYSTEM

	Minimum			Nominal			Maximum		
	dbw	V/50	V/300	dbw	V/50	V/300	dbw	V/50	V/300
A	-112	17.7 μ v		-107	31.5 μ v		-92	177 μ v	
B	-30	0.22v		-25	0.38v		-23	0.49v	
C	-118	8.9 μ v		-113	15.8 μ v		-98	89 μ v	
D	-88	280 μ v	685 μ v	-83	500 μ v	1.23mv	068	2.8mv	6.8mv
E	-89	250 μ v	610 μ v	-84	450 μ v	1.1mv	-69	2.5mv	6.1mv
F	-44	45mv		-39	80mv		-24	450mv	
G	-27	315mv		-22	560mv		-17	1.0v	
H	-17	1.0v		-17	1.0v		-17	1.0v	
I				-39	80mv				
J				-20	700mv				
K				-27	300mv				
L				-27	300mv				
M				-55	13mv				
N				-27	300mv				
O				-20	0.7v				
P				-40	70mv				
Q				-43	50mv				

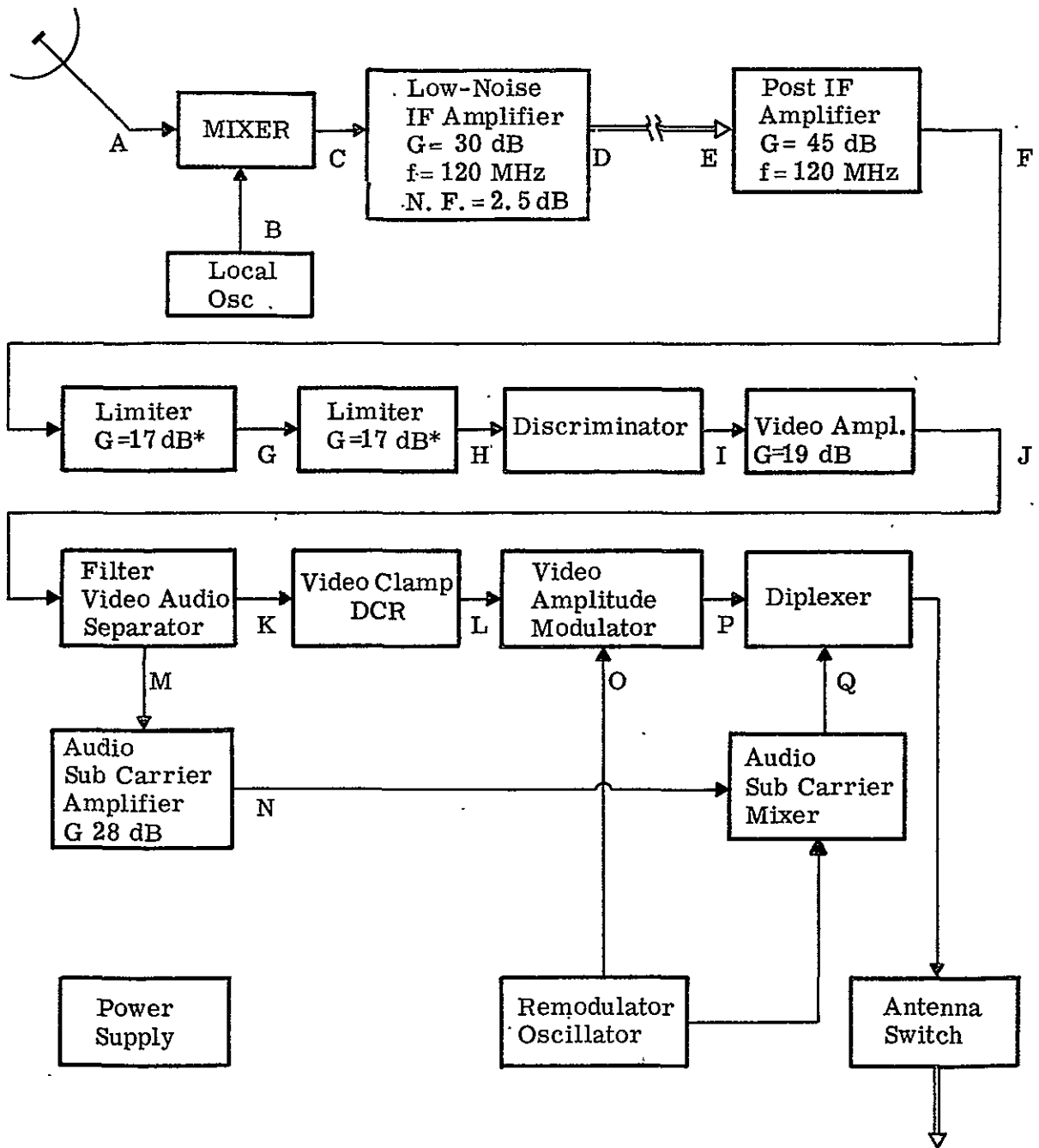


Figure 5-7. Candidate System Block Diagram

* The limiter circuits provide up to 17 dB of gain when the input signal is below the limiting level.

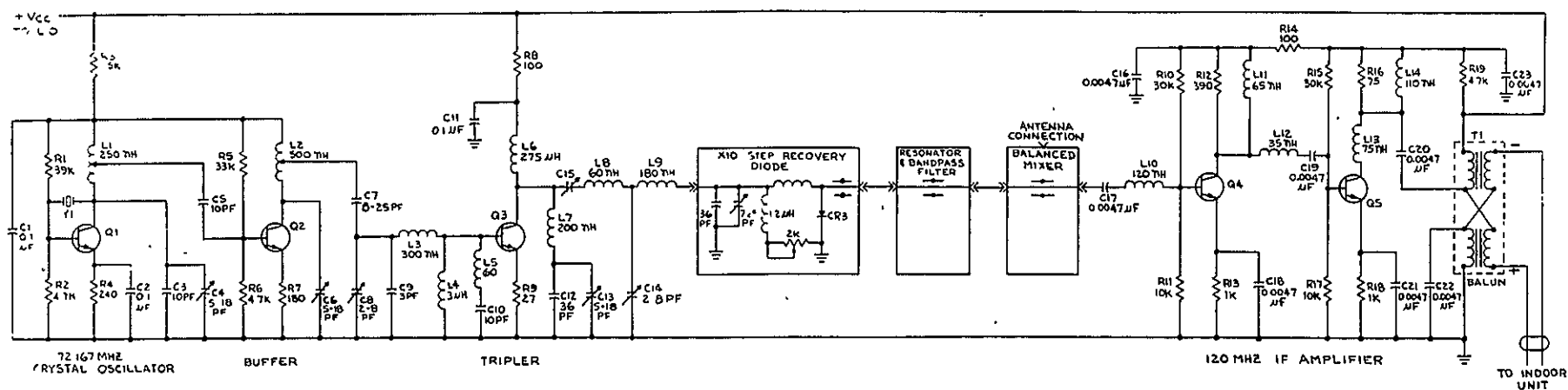


Figure 5-8. 2.25 GHz FM Converter Antenna Unit Conceptual Design

Figure 5-9. FM Converter Systems Indoor Unit Conceptual Design

Another distinction in the FM system local oscillator selection is the stability requirement of the oscillator. The FM system is more tolerant to long-term oscillator drift and less tolerant to short-term local oscillator frequency fluctuations. It was assumed that a local oscillator drift as high as ± 1 MHz, referred to the IF, would be acceptable in the FM systems. This specification permitted the potential use of several local oscillator configurations. The first configuration considered was the crystal controlled oscillator followed by a multiplier chain. This approach was considered for both the S-band and X-band local oscillators and was selected for the lower-frequency converters. The second configuration considered was a temperature-compensated transistor oscillator in the 200 MHz region, with a multiplier chain. In this configuration, the crystal cost could be eliminated and the constraint on the starting frequency imposed by the availability of crystals disappeared. Initial investigations of temperature-compensated oscillators were negative, i. e., suitable long-term stability could not be obtained. It was doubtful that this approach could be used without some thermal environmental control. The third approach considered was a transistor oscillator in the 2-GHz frequency range. This oscillator could be used directly as a local oscillator for the S-band systems and as a starting point for an X-band oscillator-multiplier chain. The investigation of S-band transistors showed that the inexpensive transistors (less than \$1.00 in lots of 10^6) had inadequate gain at 2.13 GHz. The same transistor chips were available in more expensive packages with slightly improved performance (\$6.00 in lots of 10^6). The marginal performance plus the added need for a frequency-determining element or AFC circuits made this approach economically unattractive.

5.2.2.3 2.25 GHz FM Converter IF Amplifier Consideration

An additional consideration in the selection of an IF amplifier for the FM converter as compared to the AM converter is the performance of the FM discriminator. The discriminator bandwidth is established by the specified FM modulation index. The estimated bandwidth for the 2.25 GHz FM discriminator is 30 MHz. The linearity of discriminator circuits, of the tuned variety, is inversely proportional to the ratio of bandwidth to center frequency, as a first approximation. A lower intermediate frequency permits less expensive gain at the cost of discriminator linearity and vice versa. The availability of a low-Q transmission line discriminator resulted in a compromise that led to the selection of 120 MHz as the center frequency for the FM converter IF amplifier.

The same type of bipolar transistor that was used in the AM/IF amplifier provided the performance required in the FM/IF amplifier.

The gain required in the FM/IF amplifier is approximately 75 dB, prior to the limiter stages. The latter stages of this amplifier chain are not constrained by the low noise figure requirements of the first stages. It is possible, from both a noise figure consideration and the availability of devices with the required bandwidth, to use linear integrated circuits for these latter stages. An initial design, using the IC approach, was completed to provide a costing base. It was found that the IC approach was considerably more expensive at this time and through 1975.

The discrete IF amplifier design includes seven transistors. The first two stages are located in the antenna unit to establish the system noise figure; the remaining stages are located in the indoor unit. The performance of several of these transistors has been evaluated and the results are discussed in Section 8.0 of this report.

Selectivity in the IF amplifier chain is provided, mainly, by lumped element filters between IF amplifier sections. The IF amplifier stages cannot be considered as tuned amplifiers in the normal sense. The interstage reactive components are used to compensate the transistor characteristics to obtain maximum flat gain per stage across the desired bandwidth. This compensation does result in an IF type selectivity curve, but the skirt response is uncontrolled.

The antenna lead-in cable is used to connect between the antenna-mounted IF amplifier stages and the indoor unit portion of the IF amplifier. The same type of cable and connections that were described for the 2.25 GHz AM system are used to interconnect the antenna and indoor units of the 2.25 GHz FM system. In the FM system the balun at each end of the twinlead provides a conversion from single-ended 75 ohm impedance to the balanced 300 ohm cable impedance. As before, the dc bias voltage is fed to the antenna unit via the lead-in cable.

5.2.2.4 2.25 GHz FM Converter Limiter and Discriminator Circuit Considerations

The limiter circuits can be considered as extensions of the IF amplifier chain. Their function is to compress any signal level changes that occur in the system prior to the amplitude sensitive discriminator.

Diode limiters and current-limiting differential amplifier stages were considered. The differential amplifier approach was selected, due to its superior high-frequency limiting characteristics and the availability of linear gain in the first limiter stages at low input signal levels. A filter is used at the limiter output to remove the IF signal higher harmonics.

The discriminator circuit could be realized with a feedback discriminator or a tuned circuit discriminator. The feedback discriminator was eliminated when it was shown that an inexpensive transmission-line discriminator met the bandwidth and linearity objectives. This discriminator is discussed in Section 8.0.

5.2.2.5 2.25 GHz FM Converter Remodulator Circuit Considerations

The remodulator circuits are not considered to be critical components in the FM converter design. A conceptual design for the remodulator functions has been prepared but no detailed design effort has been applied in Phase I.

The conceptual design was used as a costing base. The circuitry consists of a baseband amplifier, an audio subcarrier amplifier, the channel oscillator and separate mixers that provide the video AM output and the translated audio signal. The audio and subcarrier video signals are combined at RF.

5.2.2.6 2.25 GHz FM Converter Fabrication and Hardware Considerations

Stripline construction using PPO dielectric was selected for the 2.25 GHz FM converter microwave circuits. Stripline construction was chosen over waveguide or microstrip on the basis of lower materials costs and/or lower fabrication costs. The remainder of the circuits are specified to be built in printed-circuit form using a G-10 glass-epoxy dielectric material. The G-10 material represents a compromise between the higher-cost PPO material which has excellent high-frequency characteristics, and the minimum-cost phenolic or paper-epoxy board materials which have higher dielectric loss at radio frequencies above 10 MHz.

5.2.2.7 GHz FM Converter Candidate System Summary

The 2.25 GHz FM converter system, as shown in Figures 5-8 and 5-9, is equivalent to the 2.25 GHz AM system in the microwave circuitry but requires much more circuitry to perform the demodulation and remodulation functions. Like the 2.25 GHz AM system, it uses a single balanced Schottky diode mixer. The local oscillator consists of a crystal-controlled transistor oscillator, a transistor tripler, and a times-ten step recovery diode multiplier. The local oscillator frequency is slightly lower to produce an IF output centered at 120 MHz. Additional IF gain is required to raise the signal to an acceptable level at the input to the discriminator. The IF amplifier is split, with two stages in the antenna unit and the remainder in the indoor unit. The total antenna unit circuitry, therefore, consists of the mixer, local oscillator, and two stages of IF gain referred to as the IF preamplifier. The remainder of the circuitry is contained in the indoor unit and a twinlead cable provides the interconnection. Two stages of differential limiter/amplifier circuits are placed at the end of the linear IF amplifier to provide a constant-amplitude signal to the discriminator regardless of antenna signal level and IF amplifier gain variations. The discriminator is in a line-type tuned circuit consisting of open and shorted eighth-wavelength transmission lines. The video processor consists of a differential amplifier which converts the balanced discriminator output to a single-ended signal. The audio subcarrier is then separated from the baseband video signal and amplified in a separate channel. The remodulation is accomplished using a common channel oscillator operating at 83.25 MHz (Channel 6). The oscillator output is amplitude modulated with the video signal and is mixed with the audio subcarrier signal to generate RF signals which are combined at the output to form an AM/DSB structure similar to the AM/VSB format used for terrestrial television broadcast.

5.2.3 12.00 GHz Converter Circuit Considerations

The 12.00 GHz FM converter is distinct from the S-band FM converter with respect to the satellite transmission frequency and the frequency modulation index. These distinctions appear as design changes in the mixer and local oscillator only. The mixer requires a higher performance Schottky diode with reduced parasites. A beam-leaded diode was selected. The local oscillator frequency required for the X-band converter is much higher than for the S-band systems and although a similar approach was considered, transistor oscillator and multiplier chain, a bulk device direct oscillator was selected. The 12.00 GHz FM Converter conceptual design is shown in Figures 5-10 and 5-11.

5.2.3.1 12.00 GHz FM Converter Mixer

The X-band mixer design uses the same basic approach as the S-band mixer. The higher frequency requires the use of mixer diodes with reduced parasitic reactances and smaller physical dimensions. The beam-leaded Schottky diode, produced by Hewlett-Packard, was selected for initial investigations. The results using this diode are satisfactory, as reported in Section 8.0.

5.2.3.2 12.00 GHz FM Converter Local Oscillator

The circuit selected for the X-band local oscillator was a Gunn device in a cavity structure. This approach was chosen when the alternatives for producing a stripline oscillator chain were exhausted. Initial work was directed at the design of a crystal-controlled oscillator and multiplier chain. The final multiplier circuit, a step recovery diode, was found to be limited in performance in a stripline version to about 10 GHz. Further effort in this area would have required the use of a cavity structure to minimize losses and to provide adequate tuning capability. Once the need for a cavity structure was established, the simplicity of a direct bulk device oscillator compared to an oscillator-multiplier chain was apparent. The remaining technical question regarding the use of the Gunn device was concerned with oscillator stability, both long-term and short-term. Investigations showed that the short-term instability or incidental FM noise from the Gunn device was within tolerable limits. Further work indicated that the Gunn oscillator could be temperature stabilized against long-term drift with a simple compensating tuning probe. The relative costs of the two approaches were equivalent or slightly favored the Gunn device; therefore, the Gunn oscillator was chosen for the 12.00 GHz FM converter local oscillator.

5.2.3.3 12.00 GHz FM Converter IF and Signal Processor Circuit Selection

The remainder of the circuits selected for the X-band FM circuits are identical to those specified in the S-band FM system. The only exceptions are slightly increased bandwidth requirements in the IF and discriminator circuits which are obtained by modifying the IF tuning and alignment.

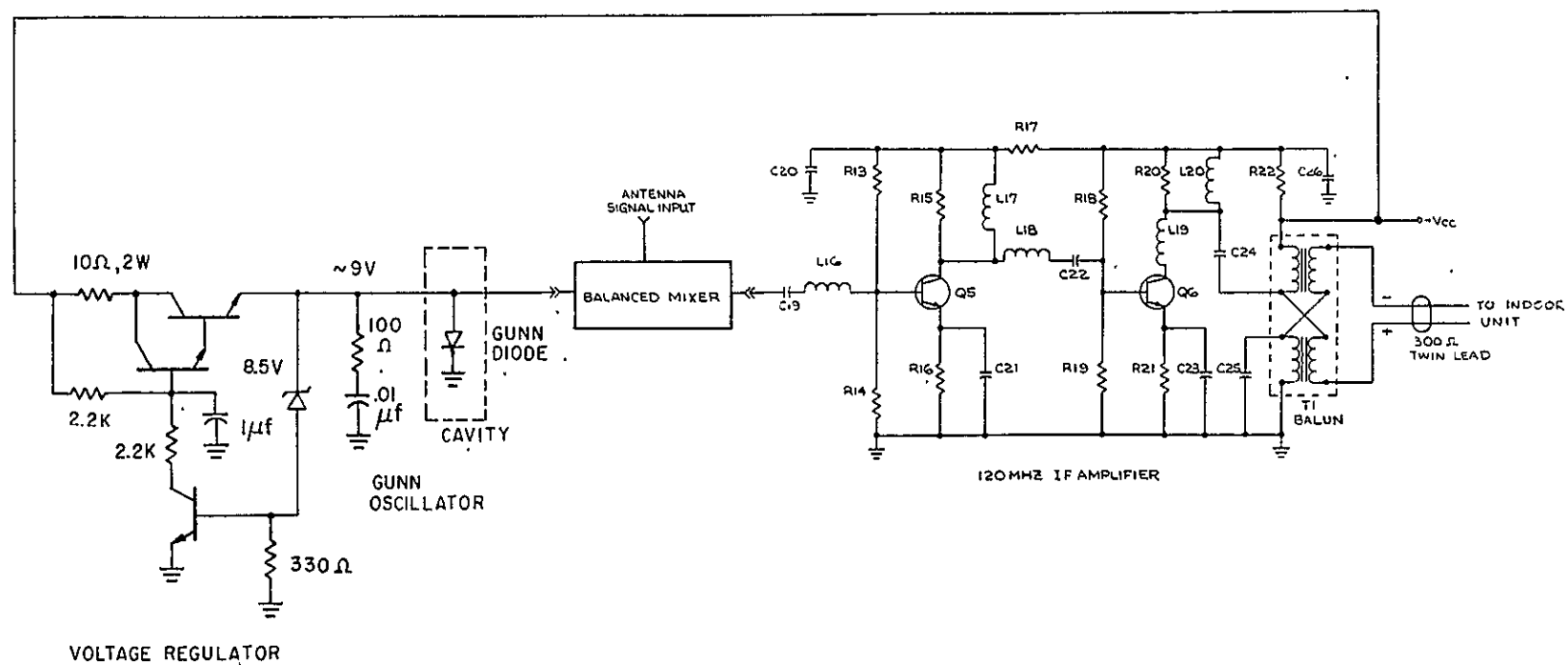


Figure 5-10. 12.0 GHz FM Converter Antenna Unit Conceptual Design

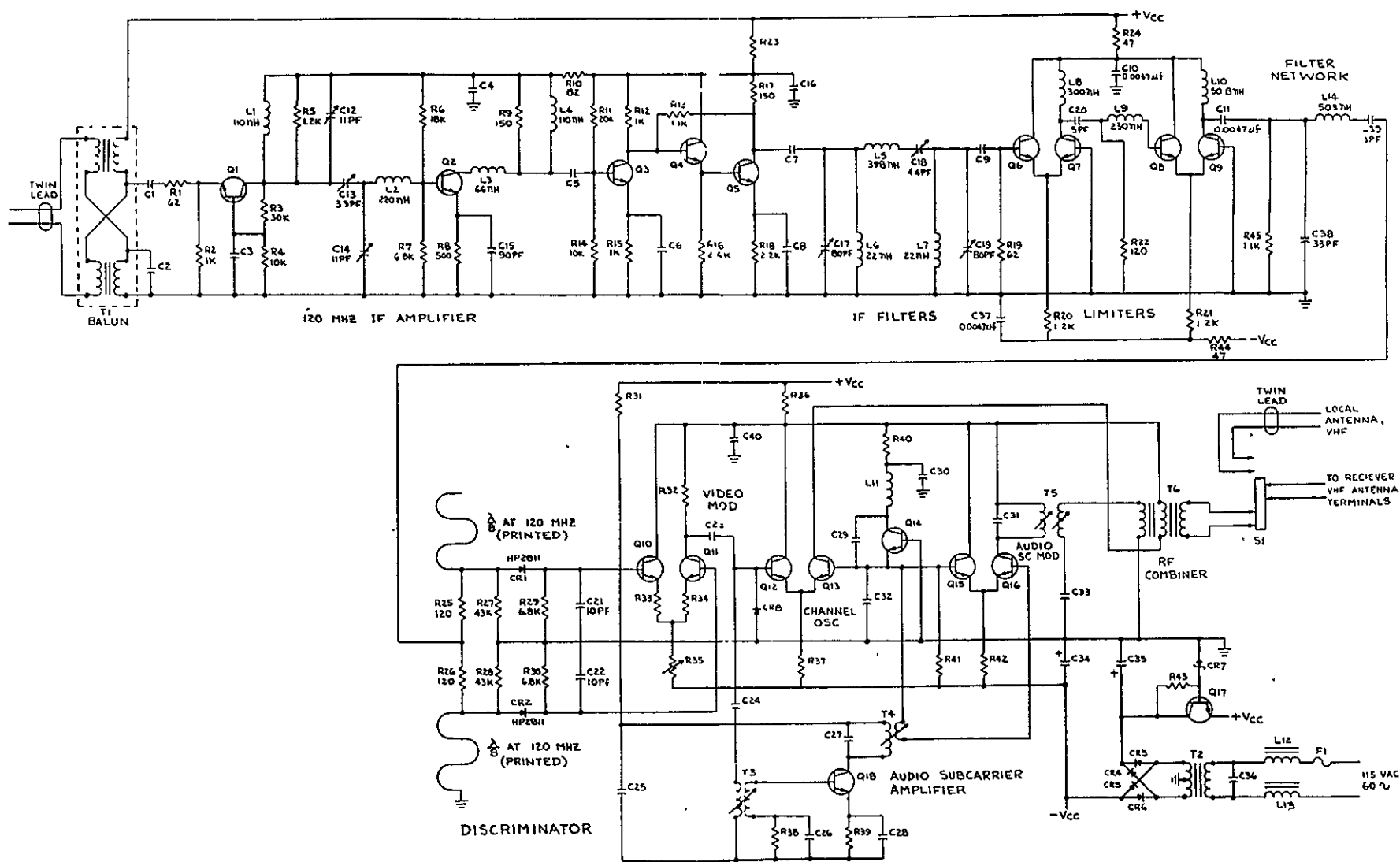


Figure 5-11. FM Converter Systems Indoor Unit Conceptual Design

5.2.3.4 12.00 GHz FM Converter Candidate System Summary

The 12.00 GHz FM converter system, as shown in Figures 5-9 and 5-10, uses the same type of circuits as the 2.25 GHz FM system, with the exception of the local oscillator. The local oscillator in this system is a Gunn oscillator which provides direct energy conversion from dc to microwave. The circuits are partitioned as before with one exception, the relatively high bias current requirement and need for a low-voltage regulator in the antenna unit. This regulator stabilizes the bias and protects the Gunn device from induced transients which may be picked up on the lead-in cable. The remainder of the circuitry — IF amplifiers, limiters, discriminator, video processor and remodulator — are identical to the 2.25 GHz FM circuits with the exception of the additional IF bandwidth of 40 MHz versus 30 MHz.

6.0 CONSUMER DESIGN PROCEDURES AND PRODUCTION COST ANALYSIS

The second major output of Phase I was the production cost analyses for the converter conceptual designs discussed in Section 5.0. The basic procedures used in obtaining the cost-performance optimized designs were outlined in Section 3.2. The details of the consumer-oriented design philosophy and the costing procedures and results are presented here.

6.1 COST ANALYSES PARAMETERS

The objective of the cost analyses portion of the program was to determine the direct materials and labor costs in the production of the converter systems. Two costing parameters were specified in the contract, annual production volume and year of production. A third parameter was found to be of equivalent importance; i. e., the production costs, in any particular year, were found to depend on the time of initial production. That is, the experience gained via the continued production of a design over several years results in appreciable production cost improvement.

In the initial cost investigation, it was quickly found that a major change in approach and philosophy, from that usually taken in design and development contracts, was required if prices were to be brought down to acceptable levels. This change consisted, essentially, of adopting a consumer-oriented philosophy; i. e., adopting the approach and techniques used by those organizations that are involved in the mass production of low-cost consumer electronics (radios and television sets), as opposed to the usual military or commercial approaches, which involve design and development contracts.

The consumer oriented approach is a part of a hierarchy of quality generally recognized, although not formalized, in electronics production. The hierarchy is as follows:

- Military Electronics - Emphasizes performance reliability, and extensive ruggedness under severe and widely varying environmental conditions. Equipment is designed to military specifications, and both inspection and test are extremely severe, comprehensive, and complete. Examples of military electronics include such items as U. S. Army communications and radar equipment, which must operate with high performance and reliability under battle conditions.

- Commercial Electronics - Emphasizes performance and reliability to only a slightly lesser extent than military electronics, but environmental conditions are more predictable. Examples include commercial broadcast and microwave relay transmitters and receivers, which must run unattended, whose outage time must be very low and which, in some cases, must operate under widely varying weather conditions.

- Consumer Electronics - Designed for home use. Emphasis is on price, although performance and reliability must be acceptable to the consumer. Very low prices are achieved by extreme attention to costs in design, procurement of parts, and large-scale production. About 3% to 6% unacceptable units are anticipated; these are repaired or replaced, at no cost to the consumer, under the manufacturer's warranty. Examples of consumer electronics are home radio and television sets.

The exact ratio of prices between military, commercial and consumer electronics is uncertain and varies with the item. However, the cost difference between military and consumer items, in those cases where a direct comparison is possible, appears to be somewhere between one and two orders of magnitude; that between commercial and consumer items up to about one order of magnitude.

6.2 COST ANALYSES ASSUMPTIONS AND CONSTRAINTS

The cost data represents the direct factory cost in an established production facility. The direct factory cost includes materials and labor costs for assembly and test, but does not include development costs, tooling costs, factory overhead and profit, shipping charges for finished products, distributor's mark up, retailer's mark up, installation costs or maintenance costs. The wholesale cost will be about 1.7 times the factory cost, and the retail cost will be approximately double the direct factory cost.

Factory costs have been estimated for production volumes of 10^3 , 10^4 , 10^5 , and 10^6 per annum. The cost estimates have been prepared to cover the time span between 1969 and 1975 in three-year increments.

Material cost trends were assumed on the basis of historical evidence. The cost trends were assigned to various material categories as a function of the market condition existing at the present time and expected changes in the market condition. "Standard" component parts, such as resistors and capacitors, were assigned a cost trend of +3% per year to reflect an inflationary condition. New semiconductor devices, with as yet unestablished volume markets, were assigned a cost trend of -30% per year on the assumption that significant market penetration would occur within the next six years. Other semiconductor products with partially established markets were assigned a -15% price trend. Established semiconductor products are expected to experience a cost decline of -5% per year.

Labor costs were assumed to follow a +3% per year inflationary trend.

A production cost improvement factor of -10% per year was assumed on the basis of added experience, innovation, component part improvement, and one redesign cycle. This factor can be expected as a result of the continued production of a particular product over a period of years.

6.3 GENERAL COST ANALYSES CONCLUSIONS

Careful use of consumer-oriented design and production techniques will make possible the manufacture, in quantity, of consumer quality ground signal processing systems at prices that customers can afford to pay.

The key cost areas for the converters are the microwave semiconductor components. At the present time the production volumes used in the cost analyses constitute a significant percentage of the total market for the new microwave semiconductor devices. The cost estimates as a function of assumed annual production volume, therefore, decrease markedly. This cost decrease with volume is much less pronounced in the cost estimates for later production years as a result of the assumed market penetration of the key microwave devices.

The most significant total cost trend with time results from the application of the production cost improvement factor.

The converters will be designed and fabricated using discrete semiconductor components, rather than integrated circuits, prior to 1975.

6.4 COST REDUCTION FACTORS

There are four major factors that affect the converter production cost. There are 1) establishment of minimum acceptable performance levels, 2) parts procurement and circuit design practices, 3) production test and warranty procedures, and 4) cost improvement and redesign programs.

6.4.1 Cost-performance Optimization

The definition of performance standards that the public will accept as optimum is difficult. This problem was essentially eliminated from this program by the assignment of performance requirements and objectives in the contract work statement. There still remained some areas for trade-off of performance and cost; however, these areas were found to be more limited than expected.

There is one major distinction between the performance of consumer electronic equipment and that of military or commercial electronics. This distinction is in the area of reliability and maintainability. The reliability of consumer products is less critical because outages affect a smaller number of people and the repair or replacement cycle can be longer. In fact, the economics of repair versus replacement are such that, in most low cost consumer products, the repair cost would be greater than the replacement cost. The ultimate low cost consumer product design and production approaches, therefore, tend to ignore the maintainability aspect in favor of a further production cost decrease.

Other areas, which permitted cost - performance tradeoffs, included the mixer, IF amplifier and local oscillator circuits. The tradeoffs were primarily concerned with the selection of semiconductor devices, once the circuit approaches had been established. Another interesting economic factor became evident in this work. There is a tendency for certain devices to establish a pre-eminence in specific applications with respect to performance. Competition for the market, originally high-profit, state-of-the-art military applications, drives costs down and expands the application base to lower-cost equipments in the commercial and consumer field. The resulting market expansion and further competition often results in the situation that the best

device, from a performance viewpoint, exhibits the lowest cost. Examples of this phenomenon are various families of digital integrated circuits, Schottky mixer diodes, planar process transistors, and many standard component parts such as resistors, and capacitors. The relatively stable cost-performance optima thus established are upset when new device processes are developed or when performance standards are specified which cannot be met by the standard components.

6.4.2 Consumer Parts Procurement

The second factor in consumer product cost reduction is the procurement procedures used in this field. The greatest cost contribution factor in component part procurement is the specification of multiple performance parameters. In any procurement procedure, perhaps the most important single consideration is the limitation of specifications. The specifications placed on component parts determines their costs. The optimum procurement procedure is to place a minimum specification and then screen incoming materials on the basis of those parameters that are most critical in the particular application. For example, a transistor can be purchased by a 2N918 identification for approximately \$1.00 in large quantities. This transistor will provide the advertised performance with over 95% assurance. If, however, a procurement was made for an unspecified transistor that would oscillate at 900 MHz (the actual circuit requirement) its unit cost would be less than \$0.10 in large quantities. More than likely, the vendor would supply 2N918 type transistors which do not meet one of the many 2N918 specifications that were not critical in the oscillator application.

The reject ratio on incoming transistors under this procurement procedure might be as high as 30%, but a substantial material cost saving would be realized by the manufacturer, with little or no sacrifice in performance. In addition, the transistors are procured in extremely large quantities, so that the cost of marketing per transistor is very small. Procurement, in quantities of 20 million or so units, with delivery and payments spread over several months according to an agreed-upon schedule, is a common procedure. Other parts (resistors, capacitors, coils, switches, etc.) are procured in a similar fashion. These procurement techniques account in large part for the extremely low costs of individual parts as shown in Section 6.6. When component parts are obtained in this manner it is often necessary to perform some in-house selection. For example, lower-noise-figure IF amplifier transistors may be screened and used in the first stages of the amplifier chain. Quality control is maintained by sample lot testing of incoming parts, thus increasing the likelihood of product rejects; however, an acceptable product reject rate can be maintained without 100% part inspection.

The qualification of component parts is, generally, an exhaustive procedure in consumer product design and production. Thorough component part evaluation is performed to establish the parameter distributions and lifetime characteristics. These data are used to determine the detailed circuit designs and expected performance variations in the completed products, and parts qualification represents one of the major items in consumer product cost-performance optimization.

6.4.3 Production Line Fabrication and Test

Low-cost assembly is achieved by breaking down the work into small tasks that can be performed by semi-skilled or unskilled labor. The tasks thus defined can be readily learned and performed in a rote fashion. Printed circuit boards are commonly used because they provide a vehicle for holding the component parts through the intermediate assembly stages, permit one-step automatic soldering, and are inexpensive. An assembly-line worker can insert as many as 14 small piece parts in a board at a single station for a direct labor cost of less than five cents. The number of stations used in a line depends on the total parts count of the product and the number of alignment, subassembly and test stations required. An AM/FM portable radio, whose complexity is similar to the TV converters, requires approximately 120 stations for complete assembly, alignment, test and packaging for shipment. The total direct labor cost in the line is of the order of three dollars.

Tests are performed on completed units only in this type of assembly. Some intermediate testing is performed inadvertently, such as in the alignment procedure. A unit is rejected if it cannot be aligned and it may be recycled for rework or the same or on special lines that provide trouble shooting capability.

Environmental tests are performed on a sample basis and these tests are designed to represent the environments encountered in shipment and normal usage. The vibration tests represent typical shipping stresses. Drop tests are performed on table model or hand held items.

The complete cycle of parts qualification, incoming inspection of parts, line tests and environmental tests are designed to maintain a product reject rate by the consumer of below 5%. The reject units are usually replaced under warranty. The number of allowable rejects is intended to be such that the incremental cost of further reduction would be more objectionable to the customers than the annoyance of having to return the sets for repair or replacement.

6.4.4 Cost Improvement in Consumer Production

Production cost improvements are stimulated through design review and production worker incentive programs. The incentives result in increased production line speed and they generate the application of innovative assembly procedures and design changes. The result is an evolutionary change in the product design and assembly, which takes advantage of inventiveness and improvements in component part parameters. The typical cost improvement obtained in a consumer product, which is produced over a number of years, is 5% per year. A further cost improvement is obtained through a redesign

cycle, which permits a greater freedom in the advantageous application of improved component parts and normally results in a 30% cost improvement. These trends have been established in the industry over a period of many years; and although the specific features, which result in cost improvements, cannot be forecast, the cost improvement can be anticipated.

6.5 COST DATA SOURCES

The cost data used in this program were obtained from many sources. A Cost Performance Review Panel consisting of representatives from several General Electric departments served in a consulting capacity. The members of this panel were:

H. Atwood, Manager of Advanced Development Engineering,
Receiver Department
W. Bartley, Manager of Advanced Development Engineering,
Communications Products Department
R. Dome, Senior Consultant, Major Color Television Department
J. Metro, Applications Engineer, Integrated Circuits Center
J. Raper, Manager of Advanced Circuits, Electronics
Laboratory
R. Weisberg, Marketing Manager, Semiconductor Products
Department

This panel provided many varied and valuable inputs to the program. Among the more important inputs were:

- Technical evaluation of designs and design suggestions
- Cost evaluation of designs and cost benchmarks
- Guidelines for television receiver performance
- Consumer product parts procurement procedures and referrals to consumer grade product part vendors.
- Semiconductor product price forecasts
- Guidance through consumer production facilities.

In addition to the panel, the services of an experienced cost analyst were obtained from the product departments. The cost analyst provided data on standard parts, labor, and tooling costs. He also provided price trend data on standard parts and procedures.

Vendor contacts were made in order to obtain cost data on items that were peculiar to the converters and not used in the consumer products. Quotes and price forecasts were obtained on microwave semiconductor products and special transistors. The applications engineers and development engineers of several companies, recognizing the potential market represented by the TV converters, released volume price data and sent evaluation samples of their devices.

The costing of integrated circuits presented a distinct problem in Phase I. The available linear integrated circuits, generally, fall into two categories, 1) special purpose circuits for which large markets exist, and 2) general purpose circuits whose specific application is tailored through the addition of a

significant number of external components. The applicable integrated circuits, the general purpose circuits, have a less competitive position with respect to discrete circuits because of the need for external component parts. None of the available integrated circuit cost data was directly applicable to the converter cost analyses. A special integrated circuit cost study was performed on the basis of tailored circuit requirements and IC production process costs. This analysis is detailed in Appendix IV.

Field trips were made to the General Electric radio and television manufacturing plants to obtain first-hand information concerning consumer-oriented design and assembly procedures. These facilities represent current competitive standards in the industry and are the results of years of cost-performance optimization of the products involved.

6.6 COST ANALYSES RESULTS

The results of the cost analyses are given in Tables 6-1 through 6-12. The 1969 production costs given in these tables are not applicable to any converters that will be sold to the public, because there was no production in 1969. However, the 1969 costs are valuable and were used as a base in cost estimating. They incorporate the most accurate and completely documented data concerning production costs. The 1972 and 1975 cost data were derived from the 1969 data. The earliest likely production date will be 1972. The 1972 and 1975 cost figures can be used in system planning and these costs are summarized in Table 6-0 below.

TABLE 6-0
1972 AND 1975 COST SUMMARY

System	Year of Production Start	1972 Units / Year		1975 Units / Year	
		10 ³	10 ⁶	10 ³	10 ⁶
2.25 GHz AM	1972	25.53	13.91	19.00	10.30
	1975			25.10	13.53
2.25 GHz FM	1972	39.91	22.71	30.48	17.24
	1975			39.38	22.66
12.00 GHz FM	1972	70.68	31.15	42.48	20.22
	1975			49.79	24.47

The costs cited in the tables are representative costs of materials and labor to any original equipment manufacturer in the northeast area. There may be slight variations due to differentials in shipping and labor charges at other locations in the continental United States.

TABLE 6-1

2.25 GHz AM CONVERTER

DATE: JULY 1969

DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1969)

<u>ANTENNA UNIT</u>	<u>ANNUAL PRODUCTION VOLUME</u>			
	10^3	10^4	10^5	10^6
<u>LOCAL OSCILLATOR</u>				
1 HPS 0112 Step Recovery Diode	7.30	5.50	4.80	4.50
2 71 MHz Oscillator Transistors	0.26	0.22	0.18	0.16
1 213 MHz Tripler Transistor	0.60	0.44	0.33	0.25
1 71 MHz Crystal	1.25	1.10	1.00	0.90
9 in ² PPO Circuit Board	1.20	0.70	0.63	0.63
4 in ² G-10 Circuit Board	0.12	0.07	0.06	0.06
R-L-C's	1.59	1.59	1.59	1.59
	12.32	9.62	8.59	8.09
<u>MIXER</u>				
2 HPA 2811 Schottky Diodes	1.20	1.00	0.96	0.86
16 in ² PPO Circuit Board	2.00	1.25	1.14	1.14
1 Amphenol 27-1 Conn.	0.36	0.30	0.27	0.25
	3.56	2.55	2.37	2.25
<u>IF AMPLIFIER</u>				
1 SB4254 Transistor	0.31	0.25	0.20	0.17
4 in ² G-10 Circuit Board	0.12	0.07	0.06	0.06
R-L-C's	0.23	0.23	0.23	0.23
	0.66	0.55	0.49	0.46
<u>MISC. PARTS</u>				
Enclosure	1.20	0.40	0.40	0.40
Hardware and Miscellaneous	1.50	1.25	1.00	1.00
Freight and Spoilage	0.80	0.50	0.40	0.35
	3.50	2.15	1.80	1.75
<u>LABOR, ASSEMBLY AND TEST</u>				
	5.00	2.00	1.50	1.00
Antenna Unit Total Materials and Labor	25.04	16.87	14.75	13.55

TABLE 6-1 (concluded)

<u>INDOOR UNIT</u>	<u>ANNUAL PRODUCTION VOLUME</u>			
	10^3	10^4	10^5	10^6
<u>POWER SUPPLY</u>				
Power Transformer	.40	.40	.40	.40
2 Rectifier Diodes	.38	.28	.22	.20
1 Reference Diode	.19	.14	.11	.10
1 Regulator Transistor	.15	.11	.09	.08
Fuse and R-L-C's	.37	.37	.37	.37
Line Cord	.14	.14	.14	.14
	1.63	1.44	1.33	1.29
<u>MISC. PARTS</u>				
2 Baluns	.10	.09	.09	.09
Antenna Switch	.16	.16	.16	.16
Screw Terminals	.07	.07	.07	.07
30' Twin Lead	.45	.45	.45	.45
Enclosure	1.00	.30	.30	.30
Hardware and Miscellaneous	.40	.25	.25	.25
Freight and Spoilage	.12	.10	.10	.10
	2.30	1.42	1.42	1.42
<u>LABOR ASSEMBLY AND TEST</u>				
	.50	.25	.25	.25
Indoor Unit Total Materials and Labor	4.43	3.11	3.00	2.96
TOTAL COST, 2.25 GHz AM CONVERTER - 1969 FACTORY COSTS	29.47	19.98	17.75	16.51

TABLE 6-2

2.25 GHz AM CONVERTER

DATE: JULY 1969

1969 COSTS BY COMPONENT PART TYPE VERSUS VOLUME

	ANNUAL PRODUCTION VOLUME			
	10^3	10^4	10^5	10^6
<u>SEMICONDUCTORS</u>				
5 TRANSISTORS	1.32	1.02	.80	.66
1 STEP RECOVERY DIODE	7.30	5.50	4.80	4.50
2 SCHOTTKY DIODES	1.20	1.00	.96	.86
3 POWER SUPPLY DIODES	.57	.42	.33	.30
	10.39	7.94	6.89	6.32
<u>PRINTED CIRCUIT BOARDS</u>				
LOCAL OSCILLATOR (PPO, G-10)	1.32	.77	.69	.69
MIXER (PPO)	2.00	1.25	1.14	1.14
IF AMPLIFIER (G-10)	.12	.07	.06	.06
	3.44	2.09	1.89	1.89
<u>R-L-C's</u>				
LOCAL OSCILLATOR	1.59			
IF AMPLIFIER	.23			
POWER SUPPLY	.37			
	2.19	2.19	2.19	2.19
<u>MISC. PARTS</u>				
ENCLOSURES	2.20	.70	.70	.70
CRYSTAL	1.25	1.10	1.00	.90
CABLES AND CONN's	1.28	1.21	1.18	1.16
POWER TRANSFORMER	.40	.40	.40	.40
HARDWARE, MISC., F and S	2.82	2.10	1.75	1.70
	7.95	5.51	5.03	4.86
<u>LABOR</u>				
	5.50	2.25	1.75	1.25
TOTAL MATERIALS	23.97	17.73	16.00	15.26
TOTAL MATERIALS AND LABOR	29.47	19.98	17.75	16.51

TABLE 6-3

2.25 GHz AM CONVERTER

DATE: JULY 1969

PRODUCTION COST ESTIMATES VERSUS TIME

ANNUAL PRODUCTION RATE = 10^3

	AVERAGE ANNUAL COST TREND	PRODUCTION YEAR		
		1969	1972	1975
<u>SEMICONDUCTORS</u>				
5 TRANSISTORS	-5%	1.32	1.14	.92
1 STEP RECOVERY DIODE	-30%	7.30	2.48	.88
2 SCHOTTKY DIODES	-15%	1.20	.62	.38
3 POWER DIODES	-5%	.57	.49	.42
		10.39	4.73	2.60
"STANDARD" PARTS	+3%	13.58	14.80	16.00
LABOR	+3%	5.50	6.00	6.50

AVERAGE COST IMPROVEMENT TREND ON STANDARD PARTS AND LABOR -10%

PRODUCTION COSTS BASED ON
INITIAL PRODUCTION YEARS
(10^3 VOLUME)

INITIAL PRODUCTION YEAR:

1969	29.47	19.93	14.73
1972		25.53	19.00
1975			25.10

TABLE 6-4

2.25 GHz AM CONVERTER

DATE: JULY 1969

PRODUCTION COST ESTIMATES VERSUS TIME

ANNUAL PRODUCTION RATE = 10^6

	AVERAGE ANNUAL COST TREND	PRODUCTION YEAR		
		1969	1972	1975
<u>SEMICONDUCTORS</u>				
5 TRANSISTORS	-5%	.66	.57	.48
1 STEP RECOVERY DIODE	-30%	4.50	1.52	.54
2 SCHOTTKY DIODES	-15%	.86	.45	.27
3 POWER DIODES	-5%	.30	.26	.22
		6.32	2.80	1.51
<u>"STANDARD" PARTS</u>	+3%	8.94	9.75	10.54
<u>LABOR</u>	+3%	1.25	1.36	1.48
AVERAGE COST IMPROVEMENT TREND	-10%			
PRODUCTION COSTS BASED ON INITIAL PRODUCTION YEAR (10 ⁶ VOLUME)				
1969		16.51	10.92	8.01
1972			13.91	10.30
1975				13.53

TABLE 6-5

2.25 GHz FM CONVERTER

DATE: JULY 1969

DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1969)

<u>ANTENNA UNIT</u>		<u>ANNUAL PRODUCTION VOLUME</u>			
<u>LOCAL OSCILLATOR</u>		<u>10³</u>	<u>10⁴</u>	<u>10⁵</u>	<u>10⁶</u>
(SIMILAR TO 2.25 GHz AM)		12.32	9.62	8.59	8.09
<u>MIXER</u>					
(SIMILAR TO 2.25 GHz AM)		3.56	2.55	2.37	2.25
<u>IF AMPLIFIER</u>					
2 SB4254 Transistors		.50	.40	.34	.32
4 in ² G-10 Circuit Board		.12	.07	.06	.06
R-L-C's		.34	.34	.34	.34
		<u>.96</u>	<u>.81</u>	<u>.74</u>	<u>.72</u>
<u>MISC. PARTS</u>					
(SIMILAR TO 2.25 GHz AM)		3.50	2.15	1.80	1.75
LABOR, ASSEMBLY AND TEST		<u>5.25</u>	<u>2.10</u>	<u>1.57</u>	<u>1.05</u>
Antenna Unit Total-Materials and Labor		25.59	17.23	15.07	13.86
<u>INDOOR UNIT</u>					
<u>IF AMPLIFIER</u>					
5 SB4254 Transistors		1.25	1.00	.85	.80
11 in ² G-10 Circuit Board		.30	.17	.15	.15
R-L-C's		.93	.93	.93	.93
		<u>2.48</u>	<u>2.10</u>	<u>1.93</u>	<u>1.88</u>
<u>LIMITER-DISC</u>					
4 SB4254 Transistors		1.00	.80	.68	.64
2 HP 2811 Diodes		1.20	1.00	.96	.86
8 in ² G-10 Circuit Board		.22	.13	.11	.11
R-L-C's		.64	.64	.64	.64
		<u>3.06</u>	<u>2.57</u>	<u>2.39</u>	<u>2.25</u>

TABLE 6-5 (concluded)

REMODULATOR	ANNUAL PRODUCTION VOLUME			
	10^3	10^4	10^5	10^6
8 Transistors	1.52	1.20	.88	.64
1 Clamp Diode	.45	.30	.28	.26
12 in ² G-10 Circuit Board	.32	.19	.16	.16
R-L-C's	1.10	1.10	1.10	1.10
	3.39	2.79	2.42	2.16
<u>POWER SUPPLY</u>				
Power Transformer	.40	.40	.40	.40
4 Rectifier Diodes	.76	.56	.44	.40
1 Reference Diode	.19	.14	.11	.10
1 Regulator Transistor	.15	.11	.09	.08
Fuse and R-L-C's	.37	.37	.37	.37
Line Cord	.14	.14	.14	.14
	2.01	1.72	1.55	1.49
<u>MISC. PARTS</u>				
3 Baluns	.15	.14	.14	.14
30' Twinlead	.45	.45	.45	.45
Antenna Switch	.16	.16	.16	.16
Screw Terminals	.07	.07	.07	.07
Enclosure	1.50	.50	.50	.50
Hardware and Misc.	1.20	.80	.60	.50
Freight and Spoilage	.32	.27	.22	.20
	3.85	2.39	2.14	2.02
LABOR-ASS'Y AND TEST	4.00	2.75	2.00	2.00
INDOOR UNIT TOTAL-MATERIALS AND LABOR	18.79	14.32	12.43	11.80
TOTAL COST, 2.25 GHz FM CONVERTER - 1960 FACTORY COSTS	44.38	31.55	27.50	25.66

TABLE 6-6

2. 25 GHz FM CONVERTER

DATE: JULY 1969

1969 COSTS BY COMPONENT PART TYPE VERSUS VOLUME

	ANNUAL PRODUCTION VOLUME			
	10^3	10^4	10^5	10^6
<u>SEMICONDUCTORS</u>				
23 Transistors	5.28	4.17	3.35	2.89
1 Step Recovery Diode	7.30	5.50	4.80	4.50
5 Schottky Diodes	2.85	2.30	2.20	1.98
5 Power Supply	.95	.70	.55	.50
	16.38	12.67	10.90	9.87
<u>PRINTED CIRCUIT BOARDS</u>				
Local Oscillator (PPO, G-10)	1.32	.77	.69	.69
Mixer (PPO)	2.00	1.25	1.14	1.14
IF Amplifier (2 Parts)(G-10)	.42	.24	.21	.21
Limiter-Disc (G-10)	.22	.13	.11	.11
Remodulator (G-10)	.32	.19	.16	.16
	4.28	2.58	2.31	2.31
<u>R-L-C's</u>				
Local Oscillator	1.59	1.59	1.59	1.59
IF Amplifier	1.27	1.27	1.27	1.27
Limiter-Disc	.64	.64	.64	.64
Remodulator	1.10	1.10	1.10	1.10
Power Supply	.37	.37	.37	.37
	4.97	4.97	4.97	4.97
<u>MISC. PARTS</u>				
Enclosures	2.70	.90	.90	.90
Crystal	1.25	1.10	1.00	.90
Cables and Conn's	1.33	1.26	1.23	1.21
Power Transformer	.40	.40	.40	.40
Hardware, Misc., F and S	3.82	2.82	2.22	2.05
	9.50	6.48	5.75	5.46
<u>TOTAL MATERIALS</u>				
LABOR	35.13	26.70	23.93	22.61
	6.25	4.85	3.57	3.05
<u>TOTAL MATERIAL AND LABOR</u>	44.38	31.55	27.50	25.66

TABLE 6-7

2.25 GHz FM CONVERTER

DATE: JULY 1969

PRODUCTION COST ESTIMATES VERSUS TIME

ANNUAL PRODUCTION RATE = 10^3

	Average Annual Cost Trend	1969	Production Year 1972	1975
<u>SEMICONDUCTORS</u>				
23 Transistors	- 5%	5.28	4.53	3.85
1 Step Recovery Diode	-30%	7.30	2.48	.88
5 Schottky Diodes	-15%	2.85	1.48	.91
5 Power Diodes	- 5%	.95	.82	.69
		16.38	9.31	6.33
Standard Parts	+ 3%	18.75	20.50	22.15
Labor	+ 3%	9.25	10.10	10.90
Cost Improvement	-10%			
Production Costs Based on Initial Production Year (Volume = 10 ³)				
	1969	44.38	31.66	24.18
	1972		39.91	30.48
	1975			39.38

TABLE 6-8

2.25 GHz FM CONVERTER

DATE: JULY 1969

PRODUCTION COST ESTIMATES VERSUS TIME

ANNUAL PRODUCTION VOLUME = 10^6

	Average Annual Cost Trend	1969	Production Year 1972	1975
<u>SEMICONDUCTORS</u>				
23 Transistors	- 5%	2.89	2.49	2.11
1 Step Recovery Diode	-30%	4.50	1.53	.54
5 Schottky Diodes	-15%	1.98	1.03	.63
5 Power Diodes	- 5%	.50	.43	.36
		9.87	5.48	3.64
"Standard" Parts	+ 3%	12.74	13.90	15.02
Labor	+ 3%	3.05	3.33	3.60
Cost Improvement		-10%		
Production Costs Based on Initial Production Year (Volume = 10 ⁶)				
	1969	25.66	18.13	13.69
	1972		22.71	17.24
	1975			22.26

TABLE 6-9

12.00 GHz FM CONVERTER

DATE: JULY 1969

DETAILED COST ESTIMATES VERSUS ANNUAL PRODUCTION VOLUME (1969)

ANTENNA UNIT <u>LOCAL OSCILLATOR</u>	ANNUAL PRODUCTION VOLUME			
	10 ³	10 ⁴	10 ⁵	10 ⁶
1 Gunn Diode	69.00	46.75	30.75	19.40
1 Cavity	6.00	3.25	2.25	2.00
1 Series Regulator Transistors	.50	.40	.38	.38
2 Shunt Regulator Transistors	.21	.18	.18	.18
R-L-C's	.28	.28	.28	.28
P.C.B.	.10	.06	.05	.05
	76.09	50.92	33.89	22.29
<u>MIXER</u>				
2 HPA 2740 Schottky Diodes	14.70	10.70	9.30	8.70
8 in ² PPO Circuit Board	1.00	.63	.57	.57
1 Amphenol 27-1 Connector	.36	.30	.27	.25
	16.06	11.63	10.14	9.52
<u>IF AMPLIFIER</u>				
2 2N4996 Transistors	.50	.40	.34	.32
4 in ² G-10 Circuit Board	.12	.07	.06	.06
R-L-C's	.34	.34	.34	.34
	.96	.81	.74	.72
<u>MISCELLANEOUS PARTS</u>				
Similar to 2.25 GHz AM System	3.50	2.15	1.80	1.75
<u>LABOR ASSEMBLY AND TEST</u>	4.00	2.00	1.34	1.25
ANTENNA UNIT TOTAL * MATERIALS AND LABOR	100.61	67.51	47.91	35.53
<u>INDOOR UNIT (Same as 2.25 GHz FM)</u>				
<u>IF AMPLIFIER</u>	2.48	2.10	1.93	1.88
<u>LIMITER-DISCRIMINATOR</u>	3.06	2.57	2.39	2.25
<u>REMODULATOR</u>	3.39	2.79	2.42	2.16
<u>POWER SUPPLY</u>				
1 Power Transformer	.60	.60	.60	.60
4 Rectifier Diodes	.76	.64	.58	.58
1 Reference Diode	.11	.09	.09	.09
R-L-C's	.71	.67	.67	.67
Fuse, Line Cord	.24	.24	.24	.24
	2.42	2.24	2.18	2.18
<u>MISCELLANEOUS PARTS</u>	3.75	2.33	2.08	1.96
<u>LABOR, ASSEMBLY AND TEST</u>	4.00	2.75	2.00	2.00
INDOOR UNIT TOTAL-MATERIALS AND LABOR	19.20	14.84	13.00	12.40
TOTAL COST, 12.00 GHz FM CONVERTER 1969 FACTORY COSTS	119.71	82.29	60.91	47.96

TABLE 6-10

12.00 GHz FM CONVERTER

DATE: JULY 1969

1969 COST BY COMPONENT PART TYPE VERSUS VOLUME

<u>SEMICONDUCTORS</u>	ANNUAL PRODUCTION VOLUME			
	10 ³	10 ⁴	10 ⁵	10 ⁶
23 Transistors	4.98	3.98	3.31	2.96
2 HPA Schottky Diodes	16.35	12.00	10.54	9.82
4 Rectifier Diodes	.76	.64	.58	.58
1 Reference Diode	.11	.09	.09	.09
1 Gunn Diode	69.00	46.75	30.75	19.40
	91.20	63.46	45.27	32.85
<u>PRINTED CIRCUIT BOARDS</u>				
Mixer (PPO)	1.00	.77	.67	.67
IF Amplifier (G-10)	.42	.24	.21	.21
Power Regulator (G-11)	.10	.06	.05	.05
Limiter-Discriminator (G-10)	.22	.13	.11	.11
Remodulator (G-10)	.32	.19	.16	.16
	2.06	1.25	1.10	1.10
<u>R-L-C's</u>				
Local Oscillator and Power Regulator	.28	.28	.28	.28
IF Amplifier	1.27	1.27	1.27	1.27
Limiter - Discriminator	.64	.64	.64	.64
Remodulator	1.10	1.10	1.10	1.10
Power Supply	.71	.71	.71	.71
	4.00	4.00	4.00	4.00
<u>MISCELLANEOUS PARTS</u>				
Cavity for Gunn Oscillator	6.00	3.25	2.25	2.00
Enclosures	2.70	.90	.90	.90
Cables and Connectors	1.33	1.26	1.23	1.21
Power Transformer	.60	.60	.60	.60
Hardware, Miscellaneous, Freight and Shipping	3.82	2.82	2.22	2.05
TOTAL MATERIALS	111.71	77.54	57.57	44.71
LABOR	8.00	4.75	3.34	3.25
TOTAL - MATERIALS AND LABOR	119.71	82.29	60.91	47.96

TABLE 6-11

12.00 GHz FM CONVERTER

DATE: JULY 1969

PRODUCTION COST ESTIMATES VERSUS TIME

ANNUAL PRODUCTION VOLUME 10^3

	Average Annual Cost Trend	PRODUCTION YEAR		
		1969	1972	1975
<u>SEMICONDUCTORS</u>				
Transistors	- 5 %	4.98	4.28	3.67
Gunn Diode	-22.5%	69.00	32.10	14.90
Cavity	-22.5%	6.00	3.90	1.30
Schottky Diodes	-15 %	1.65	1.00	.62
Schottky Diodes	-30 %	14.70	5.00	1.75
Rectifier Diodes	- 5 %	.76	.65	.56
Reference Diodes	- 5 %	.11	.10	.08
TOTAL - SEMICONDUCTORS		97.20	46.03	22.88
<u>"STANDARD PARTS"</u>	+ 3 %	14.51	15.90	17.35
<u>LABOR</u>	+ 3 %	8.00	8.75	9.56
COST IMPROVEMENT ON STANDARD PARTS AND LABOR		- 10 %		
FACTORY COST BASED ON INITIAL PRODUCTION YEAR				
(VOLUME 10 ³ /YEAR)	YEAR			
	1969	119.71	64.03	37.18
	1972		70.68	42.48
	1975			49.79

TABLE 6-12

12.00 GHz FM CONVERTER

DATE: JULY 1969

PRODUCTION COST ESTIMATES VERSUS TIME
ANNUAL PRODUCTION VOLUME 10^6

	Average Annual Cost Trend	PRODUCTION YEAR		
		1969	1972	1975
<u>SEMICONDUCTORS</u>				
Transistors	- 5 %	2.96	2.54	2.18
Gunn Diode	-22.5%	19.40	9.00	4.20
Cavity	-22.5%	2.00	1.00	.45
Schottky Diodes	-15 %	1.12	.69	.42
Schottky Diodes	-30 %	8.70	2.98	1.02
Rectifier Diodes	- 5 %	.58	.50	.43
Reference Diodes	- 5 %	.09	.08	.06
TOTAL - SEMICONDUCTORS		34.85	16.79	8.76
<u>"STANDARD PARTS"</u>	+ 3 %	9.86	10.80	11.81
<u>LABOR</u>	+ 3 %	3.25	3.56	3.90
COST IMPROVEMENT ON STANDARD PARTS AND LABOR				
FACTORY COSTS BASED ON INITIAL PRODUCTION YEAR (VOLUME 10 ⁶ /YEAR)		YEAR		
		1969	47.96	27.25
		1972		31.15
		1975		24.47

6.7 COST ANALYSIS ERROR ESTIMATES

The following analysis is provided in order to give an indication of the confidence limits of the production cost estimates. The error analysis is performed for each of the three converter systems, using the 1969 cost estimates for an annual production level of a million units. Error analyses for the extrapolated production costs as a function of time are not included here.

The error analysis is performed by assigning likely errors to each cost element in the detailed cost breakdown. Likely errors are assumed in the positive and negative directions. The assumptions are based on judgement regarding the market status of the component parts involved, and the level of circuit or mechanical design established for the components.

The values assumed as likely errors are squared, summed, and the square root of the summed squares is determined. This procedure is followed for the negative and positive error increments independently. The results are the negative and positive composite production cost errors.

The procedure assumes statistical independence in the occurrence of each individual error increment, as well as single-sided normal error distributions.

The error estimate work is summarized in Tables 6-13 through 6-16. The likely negative error is the amount by which the actual costs are less than the estimated costs. The likely positive error is the amount by which the actual costs are greater than the estimated costs.

TABLE 6-13

2.25 GHz AM CONVERTER PRODUCTION COST ERROR ESTIMATES

	Likely Negative Error	Estimated Cost	Likely Positive Error
<u>L.O.</u>	Δ_i		Δ_i
Stop Recovery Diode	1.00	4.50	0.50
Oscillator Transistors	0.02	0.16	0.04
Tripler Transistor	0.05	0.25	0.12
Crystal	0.15	0.90	0.05
PPO, P.C.B.	0.05	0.06	0.01
G-10, P.C.B.	0.01	1.59	0.15
R-L-C's	0.05		
<u>Mixer</u>			
Diodes	0.06	0.86	0.04
PPO, P.C.B.	0.09	1.14	0.09
Antenna Conn.	0.02	0.25	0.25
<u>IF. Ampl.</u>			
Transistor	0.01	0.17	0.01
G-10, P.C.B.	0.01	0.06	0.01
R-L-C's	0.02	0.23	0.07
<u>Misc. Antenna Unit</u>			
Enclosure	0.05	0.40	0.10
Hardware	0.10	1.00	0.05
Freight & Spoilage	0.05	0.35	0.05
<u>Labor, Antenna Unit</u>	0.10	1.00	0.10
Antenna Unit Totals	1.84	13.55	1.69
Antenna Unit $\Sigma \Delta_i^2$	10682		3959
<u>Power Supply</u>			
Transformer	0.05	0.40	0.05
Rectifier Diodes	0.02	0.20	0.06
Reference Diode	0.01	0.10	0.02
Regulator Transistor	0.01	0.08	0.01
R-L-C's	0.01	0.37	0.04
Line Cord	0.01	0.14	0.01

TABLE 6-13 (concluded)

2. 25 GHz AM CONVERTER PRODUCTION COST ERROR ESTIMATES

	Likely Negative Error	Estimated Cost	Likely Positive Error
<u>Misc. Indoor Unit</u>			
2 Baluns	0.01	0.09	0.01
Antenna Switch	0.01	0.16	0.02
Screw Terminals	0.01	0.07	0.01
Lead-in Cable	0.02	0.45	0.25
Enclosure	0.05	0.30	0.10
Hardware	0.02	0.25	0.05
Freight & Spoilage	0.01	0.10	0.01
<u>Labor</u>	0.02	0.25	0.05
Indoor Unit Totals	0.26	2.96	0.69
Antenna Unit Totals	1.84	13.55	1.69
2.25 GHz AM	2.10	16.51	2.38
Indoor Unit $\Sigma \Delta_i^2$	75		765
Antenna Unit $\Sigma \Delta_i^2$	10682		3959
$\Sigma \Delta_i^2$	10757		4724
Total Likely Error $\sqrt{\Sigma \Delta_i^2}$	\$1.03		\$0.69
Estimated Cost		16.51	

TABLE 6-14

2.25 GHz FM CONVERTER PRODUCTION COST ERROR ANALYSIS

	Likely Negative Error	Estimated Cost	Likely Positive Error
<u>L.O.</u> (See 2.25 GHz-AM		8.09	
<u>Mixer</u> (See 2.25 GHz-AM		2.25	
<u>IF Amplifier</u>			
Transistors	0.02	0.32	0.02
G-10 P.C.B.	0.01	0.06	0.01
R-L-C's	0.03	0.34	0.11
Misc. (See 2.25 GHz-AM		1.75	
<u>Labor</u>	0.11	1.05	0.11
Antenna Unit Totals	1.87	13.86	1.75
$\Sigma \Delta_i^2$	10690		4034
<u>Power Supply</u>			
Transformer	0.05	0.40	0.05
Rectifier Diodes	0.04	0.40	0.12
Reference Diodes	0.01	0.10	0.02
Regulator Transistor	0.01	0.08	0.01
R-L-C's	0.01	0.93	0.04
<u>IF Amplifier</u>			
Transistors	0.05	0.80	0.05
G-10, P.C.B.	0.01	0.15	0.01
R-L-C's	0.03	0.93	0.10
<u>Limiter -Disc.</u>			
Transistors	0.04	0.64	0.04
Disc. Diodes	0.12	0.86	0.04
G-10, P.C.B.	0.01	0.11	0.01
R-L-C's	0.02	0.64	0.07
<u>Remodulator</u>			
Transistors	0.08	0.64	0.24
Clamp diode	0.03	0.26	0.03
G-10, P.C.B.	0.01	0.16	0.01
R-L-C's	0.04	1.10	0.12
<u>Misc. Indoor Unit</u>			
3 Baluns	0.01	0.14	0.01
Cable	0.02	0.45	0.25
Antenna Switch	0.01	0.16	0.02
Screw Terminals	0.01	0.07	0.01
Enclosure	0.07	0.50	0.15
Misc Hardware	0.04	0.50	0.05
Freight & Spoilage	0.02	0.20	0.02

TABLE 6-14 (concluded)

2.25 GHz FM CONVERTER PRODUCTION COST ERROR ANALYSIS

	Likely Negative Error	Estimated Cost	Likely Positive Error
<u>Labor</u>	0.20	2.00	0.20
Indoor Unit Totals	0.94	11.80	1.67
Antenna Unit Totals	1.87	13.86	1.75
2.25 GHz FM	2.81	25.66	3.42
Indoor Unit $\Sigma \Delta_i^2$	808		2413
Antenna Unit $\Sigma \Delta_i^2$	10690		4034
$\Sigma \Delta_i^2$	11498		6447
Likely Error $\sqrt{\Sigma \Delta_i^2}$	\$1.07		\$0.81
Estimated Cost		\$25.66	

TABLE 6-15

12.00 GHz FM CONVERTER PRODUCTION COST ERROR ANALYSIS

		Likely Negative Error	Estimated Cost	Likely Positive Error
<u>LOCAL OSCILLATOR</u>				
	Gunn Diode	1.75	19.40	4.50
	Cavity	.25	2.00	.50
	Regulator Transistors	.03	.56	.04
	R-L-C's	.02	.28	.03
	Printed Circuit Board	.01	.05	.01
<u>MIXER</u>				
	Schottky Diodes	.50	8.70	.20
	Printed Circuit Board	.04	.57	.04
	Antenna Connector	.02	.25	.25
<u>IF AMPLIFIER (Antenna Unit Only)</u>				
	Transistors	.02	.32	.02
	Printed Circuit Board	.01	.06	.01
	R-L-C's	.03	.34	.11
<u>MISCELLANEOUS PARTS</u>				
	Enclosure	.05	.40	.10
	Hardware	.10	1.00	.05
	Freight and Spoilage	.05	.35	.05
<u>LABOR</u>				
		.20	1.25	.20
TOTALS ANTENNA UNIT	$\Sigma \Delta$	\$ 3.08	(\$ 35.53)	\$ 6.11
	$\Sigma \Delta^2$	43094		251735
TOTALS INDOOR UNIT (Same as 2.25 (Same as 2.25 GHz FM)				
		\$.94		\$ 1.67
	$\Sigma \Delta^2$	808		2413
TOTALS 12.00 GHz FM CONVERTER	$\Sigma \Delta$	\$ 4.02		\$ 7.78
	$\Sigma \Delta^2$	43902		254148
	$\sqrt{\Sigma \Delta^2}$	\$ 2.10	(\$ 47.96)	\$ 5.04

TABLE 6-16

SUMMARY ERROR ESTIMATES

1969 Production Cost, 10⁶ Annual Production Volume

	Estimated Cost	$\Sigma \Delta_i$	$\sqrt{\Sigma \Delta_i^2}$	$\Sigma \Delta_i$	$\sqrt{\Sigma \Delta_i^2}$
2.25 GHz, AM	\$16.51	2.10	1.03	2.38	0.69
2.25 GHz, FM	25.66	2.81	1.07	3.42	0.81
12.00 GHz FM	54.92	6.50	4.05	5.69	2.18

$\Sigma \Delta_i$ = Sum of all assumed negative errors
 $\sqrt{\Sigma \Delta_i^2}$ = Likely negative error
 $\Sigma \Delta_i$ = Sum of all assumed positive error
 $\sqrt{\Sigma \Delta_i^2}$ = Likely positive error

7.0 PRODUCTION COST ANALYSES THROUGH 1980

The cost analyses presented in Section 6.0 were based on the converter conceptual designs developed in Phase I. The design is expected to go through unspecified modifications and at least one design change over a period of several years' development and production. The cost improvement trend factor in the production cost estimates are based on this design evolution. The validity of these conceptual designs, for cost analyses beyond 1975, is therefore highly suspect.

If the assumption is made that the design concept will be fixed through 1980, a cost forecast would indicate a nearly asymptotic basic materials and labor cost trend, modified by continual cost improvement as a result of the production learning curve. The net result, under these circumstances, would be an annual cost improvement of less than 5% per year after 1975.

It is not expected that the situation with respect to this product will be that stable. The following changes are foreseen:

- Additional development in device performance capabilities and methods of fabrication.
- Changes in the converter system design constraints.
- Basic changes in the converter design approaches as a result of the above two items.

The production cost analyses presented in Section 6.0 are believed to be quite accurate. Estimates were not presented beyond the year 1975 in an attempt to preserve the integrity of the data presented. A discussion of anticipated performance and cost trends, presented in Appendix V, amplifies the reasoning behind the statements presented above.

8.0 PRELIMINARY COMPONENT PARTS AND BREADBOARD EVALUATIONS

A survey of applicable devices was performed to provide a basis, with respect to operational parameters, for preliminary system configuration. The critical components identified were:

- RF Amplifiers
- Mixers
- Local Oscillators
- IF Amplifiers
- FM Discriminators

Candidate components and their gross performance parameters are shown in Tables I through X of Appendix VI. The tables of devices contain device prices available in October 1968. These figures were used to rank device costs.

The initial work plan for Phase I did not include extensive breadboard construction and evaluation. It became apparent as the program progressed that more extensive breadboard work would be required to meet the Phase I objectives and the contract work plan was amended to extend the Phase I schedule to include this effort.

The purposes for which the component parts and breadboards were evaluated are:

- . Selection of component parts and materials
- . Demonstration of circuit feasibility and evaluation of circuit performance capabilities
- . Definition of functional component designs to provide a detailed costing base
- . Execution of cost-performance tradeoffs regarding alternate component part selection
- . Evaluation of fabrication techniques and materials
- . Empirical design optimization of stripline circuits.

The circuits and components that were selected for breadboarding were those that were classified as critical with respect to performance and/or cost. The critical performance circuits are those which establish the converter noise figure. In general, they are also high on the relative cost list. The most critical cost circuits are those in the local oscillator chains.

The circuits that were breadboarded are:

- S-Band balanced mixers
- 120 MHz IF amplifiers for the FM converter consisting of: a two-stage, low-noise amplifier for the antenna unit, a five-stage amplifier for the indoor unit, the interconnecting twinlead cable and balun assembly, and the amplifier filters.
- S-band local oscillator chain composed of a crystal controlled oscillator and buffer amplifier, a transistor tripler including idler circuits, a times-10 step recovery diode harmonic generator, and two bandpass filters for spurious multiplier harmonic rejection.
- Two temperature-compensated oscillators for possible replacement of the crystal oscillator in the FM S-band converter.
- A limiter-discriminator combination for the FM signal processor.
- X-band balanced mixers.
- X-band local oscillator chain consisting of: a crystal oscillator and buffer, a transistor tripler with idler circuits, a times-twenty step recovery diode harmonic generator, and two bandpass filters for spurious multiplier harmonic suppression.
- X-Band local oscillator using a cavity mounted Gunn diode.

8.1 Breadboard Relation to Conceptual Designs

The breadboard assemblies were selected to represent generic operation of circuits where multiple applications existed. For example, the S-band mixer can be used for both of the 2.25 GHz systems. Also the similarity of the AM and FM low noise amplifier designs did not warrant the separate evaluation of the single stage IF amplifier used in the AM system. The S-band local oscillator chain was not duplicated for the two S-band systems, and the FM system IF amplifier and limiter-discriminator breadboards were designed to function in the wider bandwidth, X-band FM system.

8.2 Breadboard Problems

One general problem was encountered in the fabrication of the stripline breadboards. Extreme warping of the Rexolite material selected for initial fabrication was encountered in the processing of the double clad laminate. Attempts were made to alleviate the problem with both pre-processing and post-processing annealing with unsatisfactory results. A similar problem was encountered with the Zytron-G, PPO material. The problem was circumvented by selecting General Electric pre-annealed PPO material. This material still exhibited some warping when a significant portion of one side of the double-clad board was removed in the etching process. The extent of the warping was such that the etched boards could be readily assembled with clamping plates to hold the stripline circuits in alignment.

8.3 Component Part and Breadboard Evaluation Results

The evaluation of component parts outside the breadboards included measurements of transistors, integrated circuits and Schottky diodes for the mixers.

Complex gain versus frequency measurements were made on PA7600 and LM271 integrated circuit samples to determine their applicability for IF amplifiers and remodulators in the FM systems. The results were used to obtain a conceptual design of the IF amplifier but breadboard circuits were not constructed.

The S-parameters of thirty SB4254 transistors have been measured to obtain average values for IF amplifier and limiter designs. These averages are listed in Table 8.1.

TABLE 8.1

TYPICAL VALUE OF S-PARAMETERS
FOR SB-4254 AT 10 v; 2 mA

Freq. (MHz)	S ₁₁		S ₂₁		S ₁₂		S ₂₂	
	Mag.	Angle Degrees	Mag.	Angle Degrees	Mag.	Angle Degrees	Mag.	Angle Degrees
100	.719	-48.2	4.77	129.5	.017	65.5	.939	-8.0
120	.679	-54.9	4.36	122.5	.019	78.1	.925	-8.9
140	.642	-59.5	4.06	116.8	.020	63.3	.915	-9.8

A group of HPA 2900 diodes, including the ones used in the mixers, are compared in Figure 8-1 according to their V-I characteristics. The spread of voltage at a given current is about $\pm 10\%$, with most diodes falling in a $\pm 5\%$ spread. It appears that the dc characteristic alone is insufficient to make a good selection. A supplementary measure is proposed as follows. A jig similar to the actual mixer will be built. The two diodes will be replaced by a resistor and diode under test. The conversion loss will be measured by comparing with a calibrated signal switched in and out at the IF amplifier output. The noise output with zero microwave signal input will also be read. These two values will permit the selection of good diode pairs.

In addition to the device evaluations performed outside the breadboards, several performance parameters of the completed breadboard assemblies were obtained.

8.3.1 S-band Mixer Breadboard Evaluation

The S-band mixer consists of a stripline branch type directional coupler and two Schottky diodes mounted in opposition with respect to input ports. Stripline structure was chosen for ease of fabrication, better RF shielding, better dimension control and overall low cost.

The pattern of the stripline structure is shown in Figure 8-2. On the left side of the diode sites two sections of open-short branches are provided for IF return while presenting high impedances to the RF and LO signals.

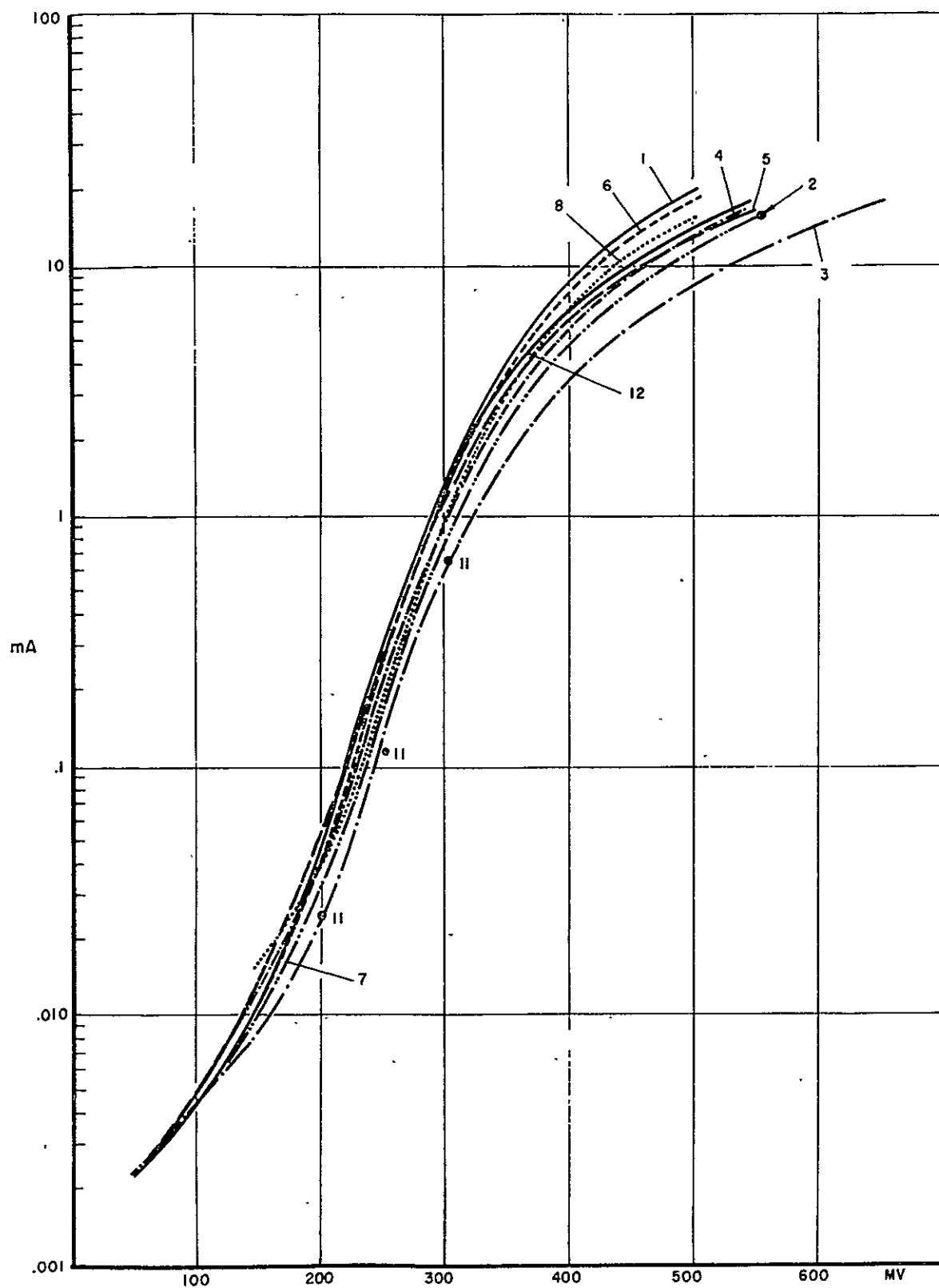


Figure 8-1. HPA 2900 Diodes - Forward I-V Characteristics

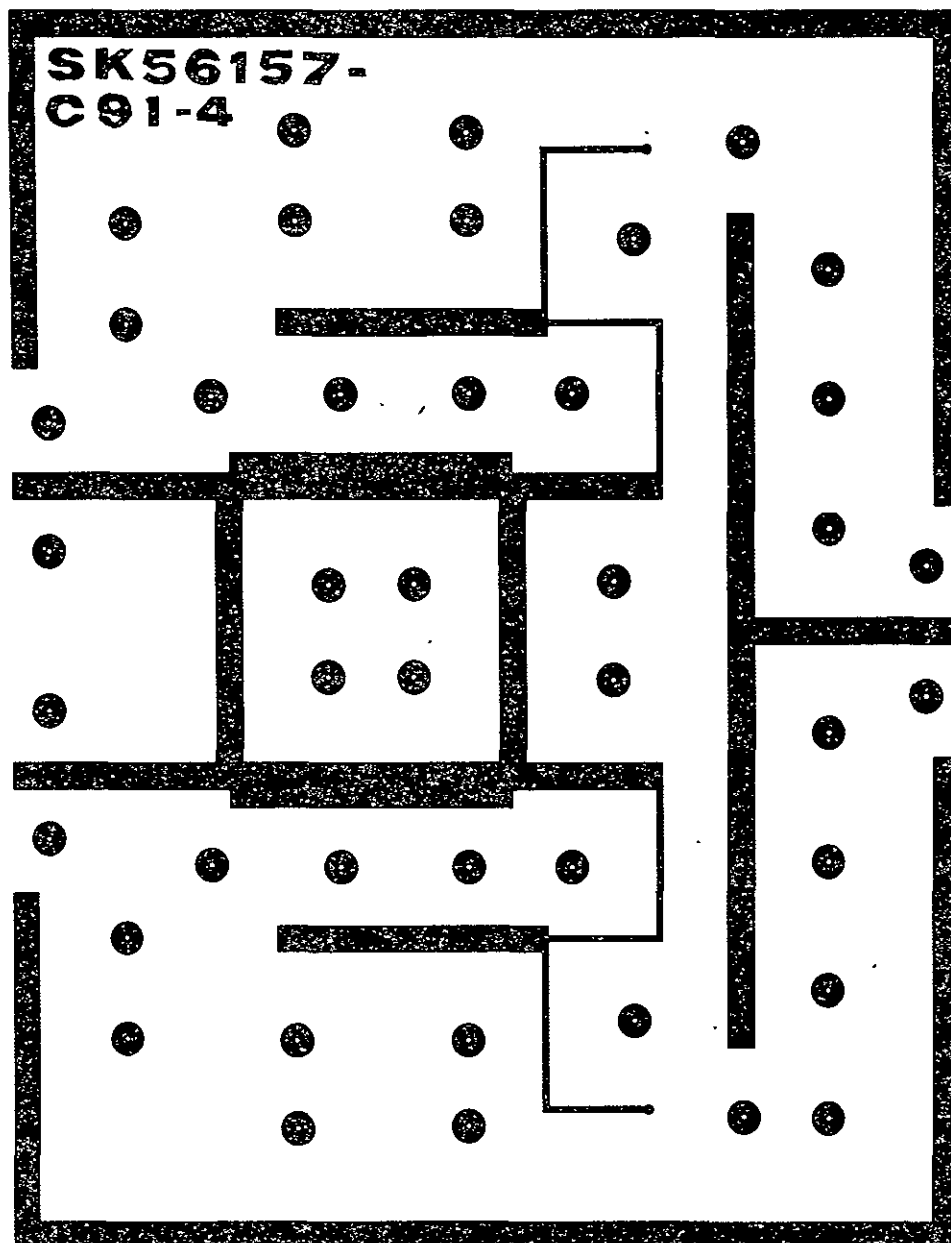


Figure 8-2. S-band Mixer

The RF shorts behind the diodes are provided with open end 1/4-wave sections properly corrected for fringe capacitance. The combined parallel output is connected to the IF amplifier.

The mixer was evaluated in a setup incorporating an AIL Hot and Cold Noise source generator and a 30 MHz precision receiver. The 2.25 GHz balanced mixer and the associated IF antenna unit, as well as the IF ground unit, were tested together. The NF measurements obtained correspond to a DSB reading of the complete assembly. The SSB NF is assumed to be 3 dB worse; i. e., assuming that the noise power is doubled and the conversion loss for the signal and the image are the same.

The overall NF measurements of the mixer-IF assembly are plotted in Figure 8-3 where HPA 2418 and Schottky HPA 2900 diodes are compared. It is observed that the NF difference between them is less than 1 dB when operated at a local oscillator level of 2 to 6 dBm. The NF attained with the HPA 2900 is less than 6.5 dB DSB, or less than 9.5 dB, SSB, which is within the desired noise margin range.

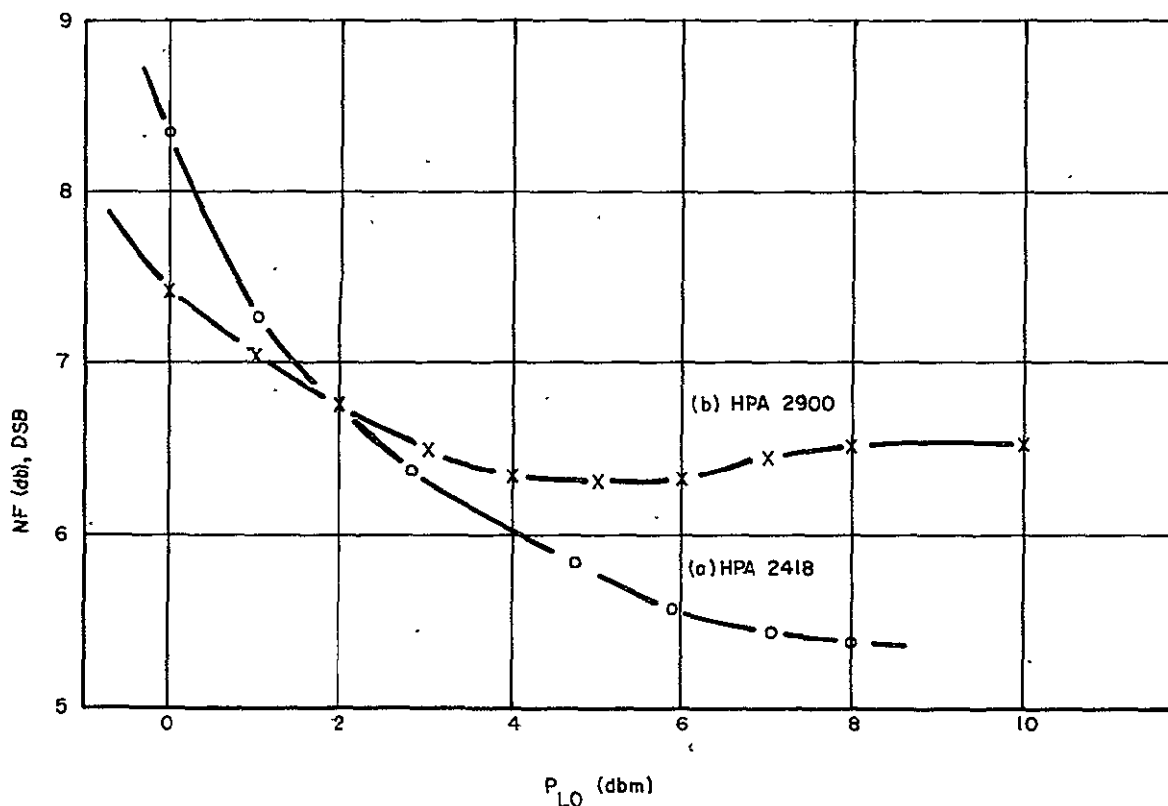


Figure 8-3. S-band Converter Noise Figure vs. Local Oscillator Power Level

The properties of the hybrid with the diode as load in a dynamic condition are evaluated with regard to the isolation between the various ports and are shown in Figures 8-4, 8-5, and 8-6. The characteristics are shown to be a function of LO power. For a local oscillator power between 4 to 6 dBm the isolation values are:

Between	Isolation
IF and LO ports	~10 dB
Signal and LO	-15 dB

Somewhat better isolations are expected with further design refinements. The relative conversion loss is shown in Figure 8-7. The conversion loss is better for the lower side band (i.e., signal below oscillator frequency) than for the upper side band. The initial mixer was not optimum for the desired upper sideband operation due to the partially empirical nature of stripline circuit design. Additional optimization will be required in Phase II.

A second breadboard of S-band mixer with Rexolite 2200 material was also evaluated. The results are comparable to the first.

The mixer unit and the IF amplifier antenna unit was subjected to temperature changes. The results are nearly satisfactory. The NF variation over the frequency band as well as over the temperature range is more than the noise margin allowed. The fault seems to lie more on the amplifier side rather than the mixer side. Changes in the IF amplifier circuit are expected to correct this. The measured NF of the complete S-band system is tabulated in Table 8-2. These results are neither typical nor final; improvements are expected. The 2.25 GHz mixer and IF preamplifier breadboards are shown in Figure 8-8.

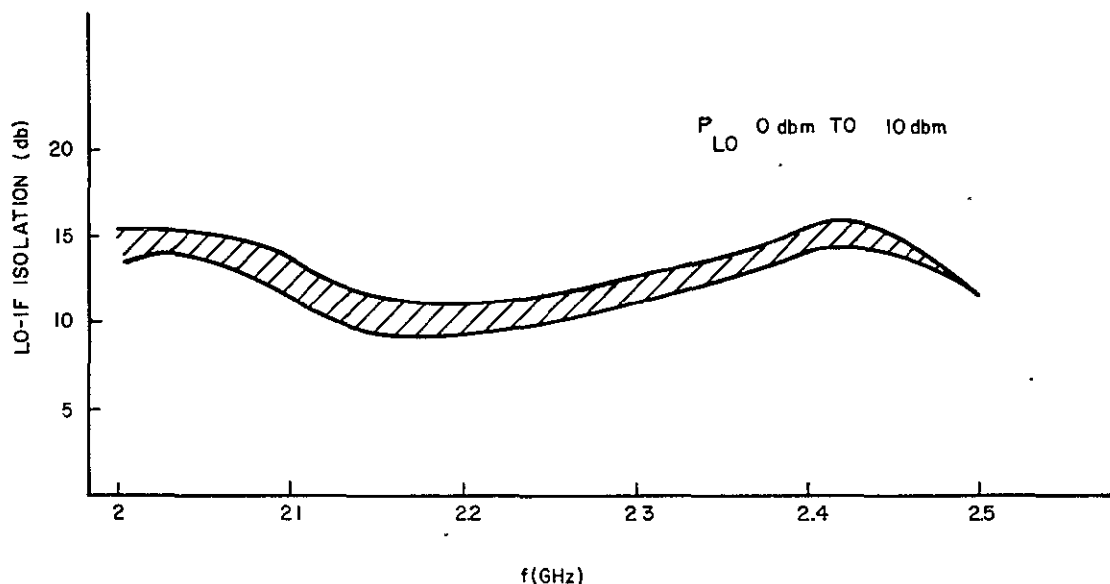


Figure 8-4. LO-IF Isolation vs. Frequency for S-band Mixer

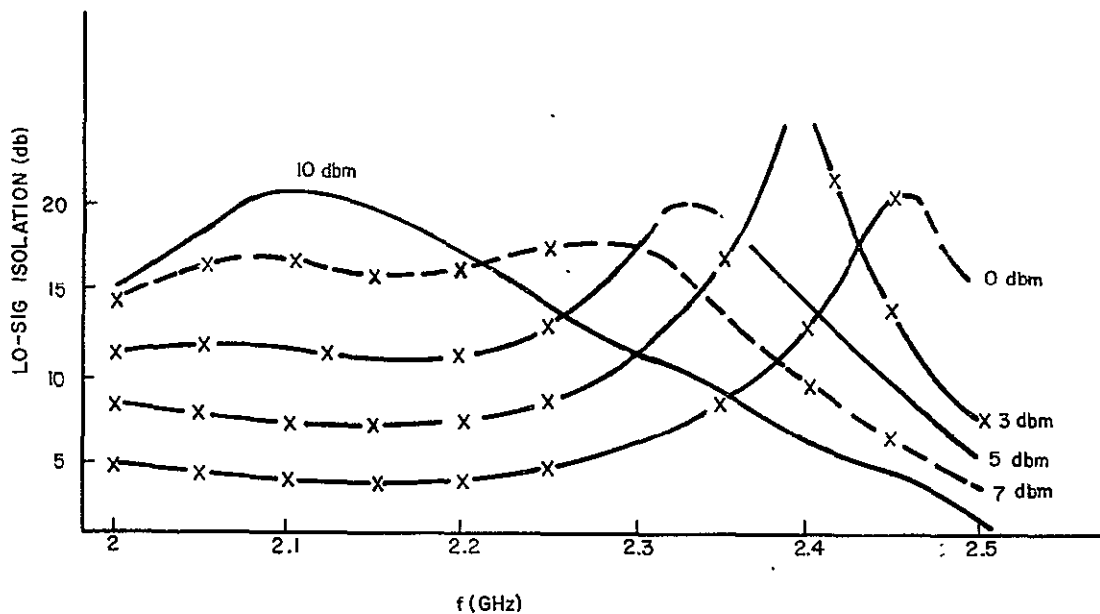


Figure 8-5. LO-SIG Isolation vs. Frequency for S-band Mixer

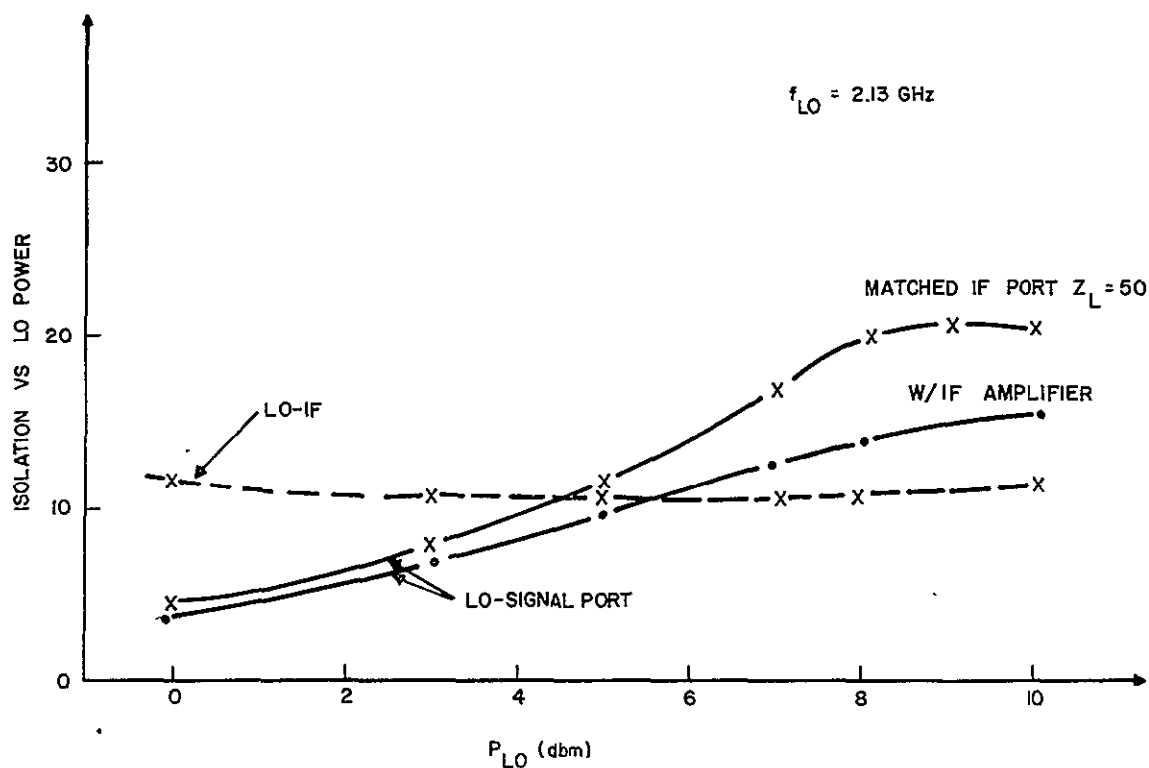


Figure 8-6. Isolation vs. LO power for S-band Mixer

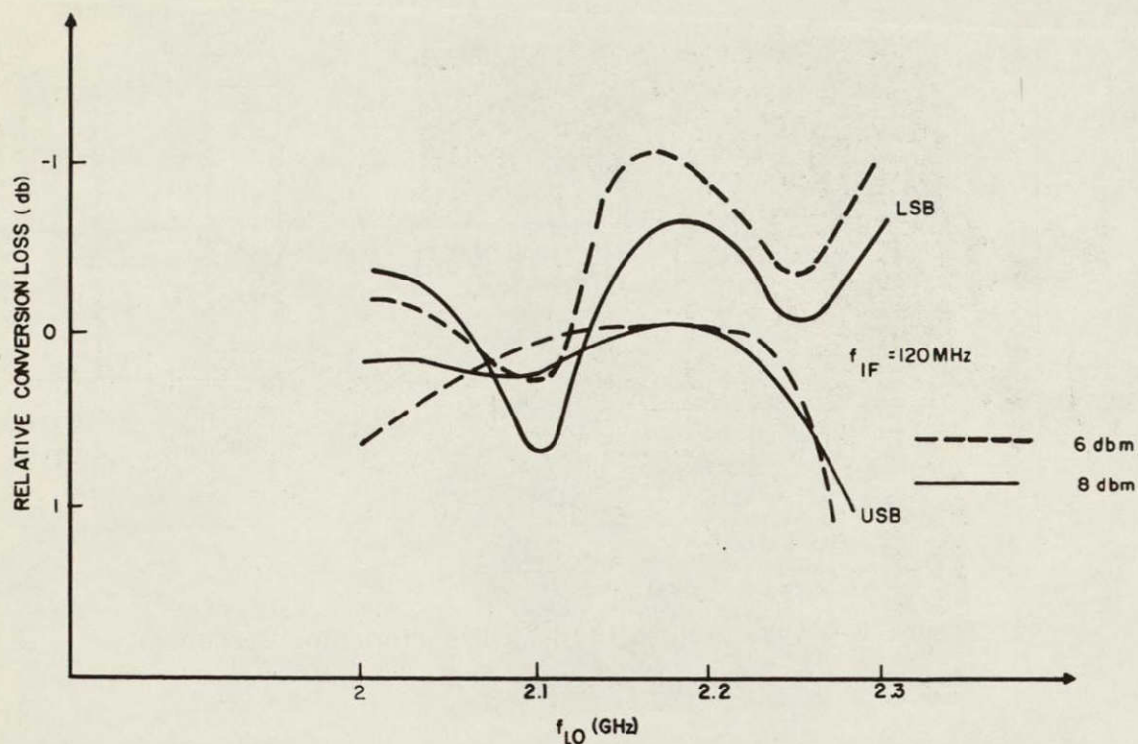


Figure 8-7. Conversion Loss vs. LO Frequency for S-band Mixer

TABLE 8-2

DSB NF OF THE S-BAND FM BREADBOARD
(MIXER AND IF)
OVER THE TEMPERATURE RANGE

TEMP °F	IF FREQUENCY (MHz)		
	110	120	130
135	7.6 dB	8.1 dB	7.0 dB
120	6.9	7.5	7.6
100	7.2	6.8	6.8
75	6.4	6.4	6.2
50	6.9	6.2	6.5
25	6.9	6.3	6.7
0	6.4	6.7	6.4
-20	6.7	6.7	6.9
-40	6.4	7.4	7.3

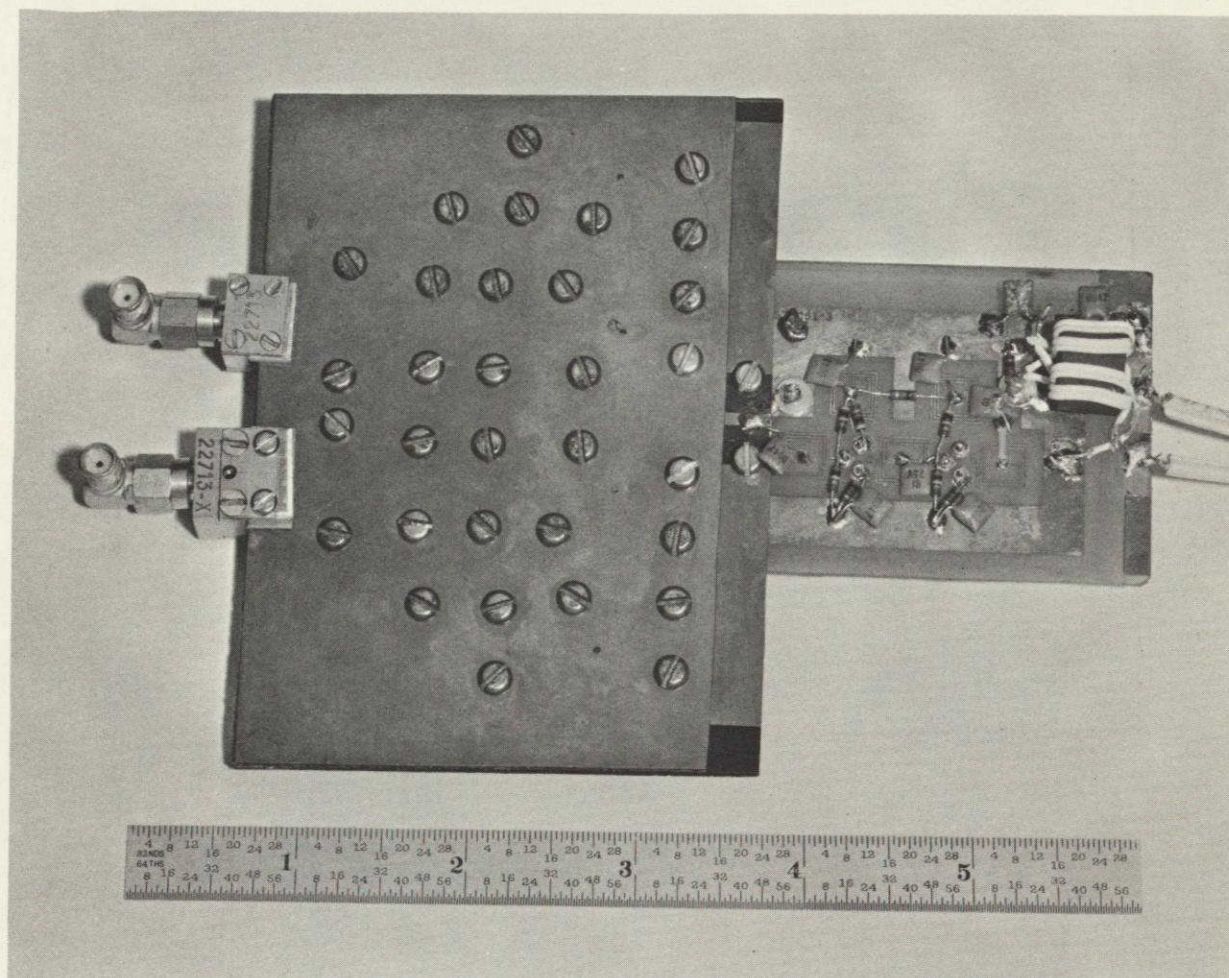


Figure 8-8. S-Band Mixer and IF Preamplifier Breadboard

8.3.2 S-band Local Oscillator Breadboard Evaluation

Breadboards were constructed for the local oscillator circuits used in the 2.25 GHz FM converter. A crystal controlled oscillator, 71 MHz, with a buffer stage was used as the reference frequency source.

The breadboard circuit for the oscillator chain is shown in Figure 8-9

The 71 MHz crystal oscillator and buffer characteristics are measured as follows:

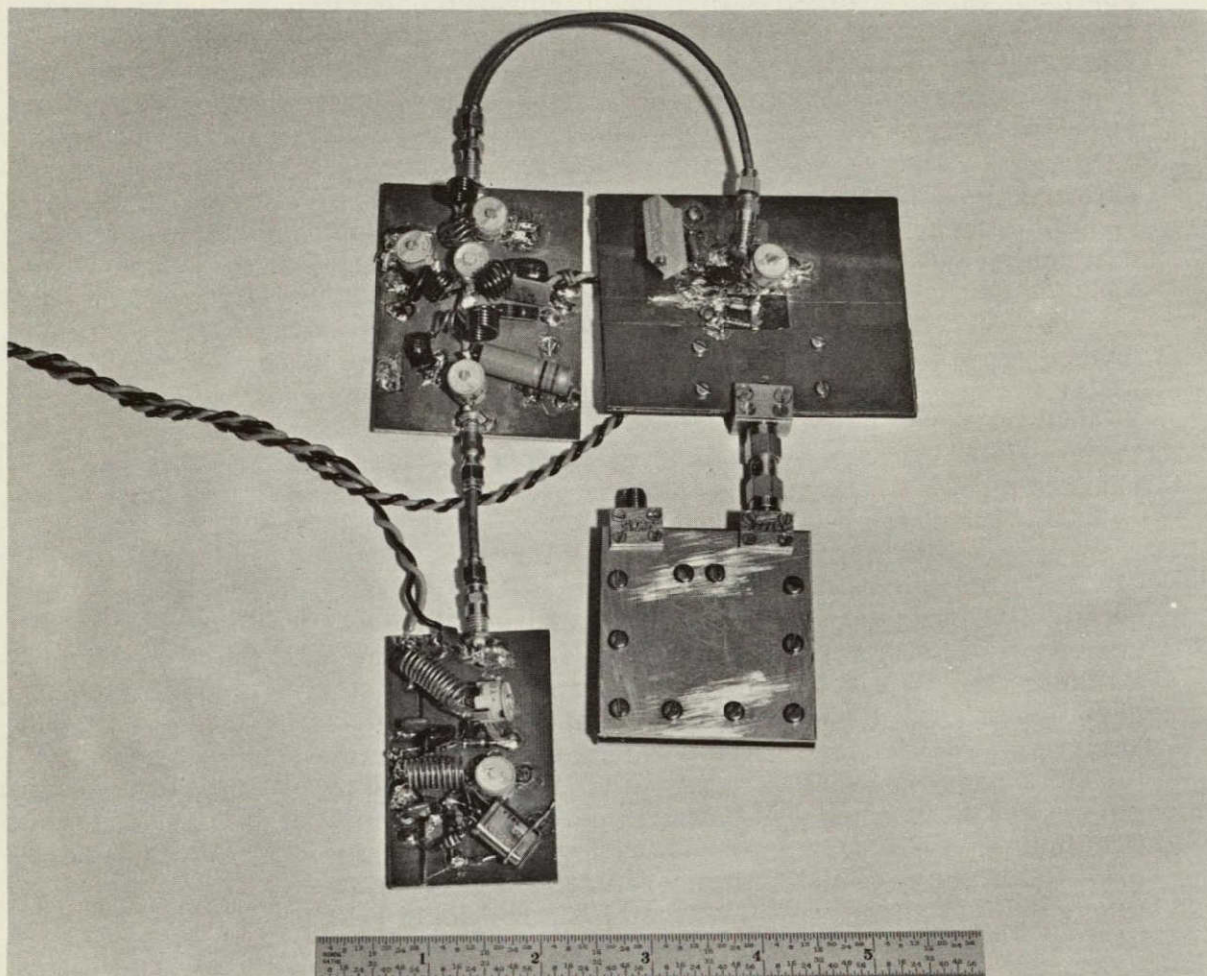
$$V_{CC} = 12V, I_C \text{ (total)} = 15 \text{ mA}$$

Power output - 20 mW.

Stability; $\Delta f = 2.3 \text{ kHz}$ at 71 MHz over temperature range of -50°C to $+60^{\circ}\text{C}$.

Self Starting

Second Harmonic Power = -45 dB (with respect to fundamental)



- | | |
|-------------|---|
| Lower Left | 71 MHz Crystal Oscillator and Buffer |
| Upper Left | 71 MHz to 213 MHz Tripler |
| Upper Right | Times 10 Step Recovery Diode Multiplier |
| Lower Right | Quarter-Wave Section Bandpass Filter |

Figure 8-9. 2.25 GHz FM Converter Local Oscillator Chain Breadboard

Efforts were made to remove the crystal from the oscillator. The approach was first directed to a temperature-compensated oscillator at 213 MHz ($3 \times f_{e0}$). Both Colpitts and Clapp configurations were evaluated. It was found that the "usual" advantage of the Clapp configuration did not exist at this frequency; the controllable small inductances (necessary when increasing the effective capacitance across collector-emitter) were just as much a problem as if the transistor's collector capacitance were an appreciable portion of the external capacitance, which is the case in the Colpitts configuration. Best results were achieved with zero T.C. capacitors (Mepco) in the Colpitts circuit configuration. The maximum frequency deviation, over the temperature range of -40°C to $+55^{\circ}\text{C}$, was ± 1.8 MHz at 213 MHz. This was about 2:1 worse than desired. The results appeared to be limited by the $+30\text{ppm}/^{\circ}\text{C}$ best tolerance capacitors and uncontrolled temperature effects attributed to the transistor package as opposed to the chip itself (in this case an epoxy case on the 2N4996 transistor). When collector voltages other than 12 volts were applied to the oscillator circuit the temperature coefficient, f versus temperature, changed sign. The output power of the oscillator varied about 10% over the -40°C to $+55^{\circ}\text{C}$ range.

Attempts were then made to realize the stability requirements at 71 MHz. This is not as attractive, of course, since a tripler would again be necessary, whereas the 213 MHz oscillator would eliminate the necessity for a tripler. Results at 71 MHz, as expected, were more favorable. A frequency deviation of about 65 kHz, over the temperature range of -20°C to $+55^{\circ}\text{C}$, was achieved. This frequency-temperature curve was again $\pm$ slope, hence, further compensation is very difficult. It is reasoned that transistor packaging effects are again causing the low temperature problems. It is concluded that, with further engineering, the required stability ($+33$ kHz at 71 MHz) can be achieved over the complete temperature range. It is questioned, however, whether the elimination of the crystal control would result in a less expensive oscillator in light of the manufacturing problems associated with temperature compensation.

Transistor Tripler

The fundamental power out, was measured at -50 dB with respect to power out at 213 MHz. Second harmonic power is -20 dB down at output, indicating a modification is in order to further reduce this component. No significant problems are anticipated, however.

The power out of this stage is running about 50 mW which is somewhat less than desired to drive the X10 multiplier. It may be necessary to add a stage of gain, or increase the power out of the buffer amplifier. These steps are being investigated.

X10 Multiplier

Power in at 213 MHz - 100 mW

Power reflected - 15 mW

Power output at 2130 MHz - +4 dBm (including filter)

Both the HPA 0114 and 0112 diodes have been investigated in this multiplier. Originally the 0112 was more expensive, however recent price changes resulted in the 0112 selling for approximately \$10.90 as compared to about \$15.95 for the 0114.

2130 MHz Filter

The five-section, quarter wave, alternately shorted stripline filter shown in Figure 8-10 will be used. The half wave, open line, filter version, Figure 8-11, was evaluated and found equivalent, except for size.

The selected $\lambda/4$ filter, Figure 8-10, utilizes .080 diameter screws to achieve the shorted lines to the ground plane. An attempt was made to provide shorting by metallizing back in the signal line surface to let the standard stripline mounting screws provide the grounding. The results showed a severe degradation in VSWR and insertion loss.

12 GHz Mixer Breadboard Evaluation

The 12 GHz mixer design is shown in Figure 8-12. The hybrid structure was evaluated separately. At the time, difficulty was encountered with regard to dielectric (1422 Rexolite) breakage. Since the PPO material is electrically so close to it, the subsequent mixer was made with PPO boards (1/32" thick, 2 oz. double copper clad) without changing the pattern dimensions for the hybrid. The signal port has two alternate inputs, one through a second order Butterworth filter, the other straight through. Similarly the oscillator port has a 3rd order filter path and a straight-through path. The hybrid was designed to have a 60 ohm characteristic impedance to reduce the TEE junction ambiguity and to better match the diode impedance. Therefore, the signal and the oscillator paths are provided with a quarter-wave section to match for 50 ohms, which is convenient for instrumentation. The alternate inputs are mechanically cut for each individual experiment. For example, the filters are evaluated separately by cutting off the connections to the hybrid.

In the oscillator port a shunt quarter-wave section, shorted to ground, is provided for IF return. The physical dimension of the hybrid at the IF frequency is negligible. The presence of this branch in the oscillator arm should affect the signal very little because of the high isolation of the hybrid. This arrangement is preferable to that in which the equivalent branches were attached to the immediate neighborhood of the diodes.

The HPA 2740 diodes were attached to the stripline by ultrasonic bonding. To facilitate the bonding process, a gold plating step was added. The additional cost of this operation in a production environment would be less than \$0.10 per board. RF returns for the diodes are provided by the 1/4-wave sections following the diodes. The total loading for the IF is less than 1.5 pf, which is negligible compared to the capacitance of the diodes themselves.

The filters were evaluated and found to have a center frequency that was 1% above the desired value, a loss of 3 dB in the center band, and a f_{LO} response at ± 600 MHz of -37 dB. A modification has been made since, but no data has been obtained. The signal port input has VSWR values of 1.6, 1.38, 1.32 for oscillator powers of 0 dBm, 3 dBm, and 5 dBm, respectively.

Noise figure measurements, including the IF amplifier, were undertaken using various methods. The noise figure is about 11.0 ± 1 dB at an oscillator power level of 5 dBm. Reducing the LO power to 3 dBm, the NF degrades by about 0.5 dB.

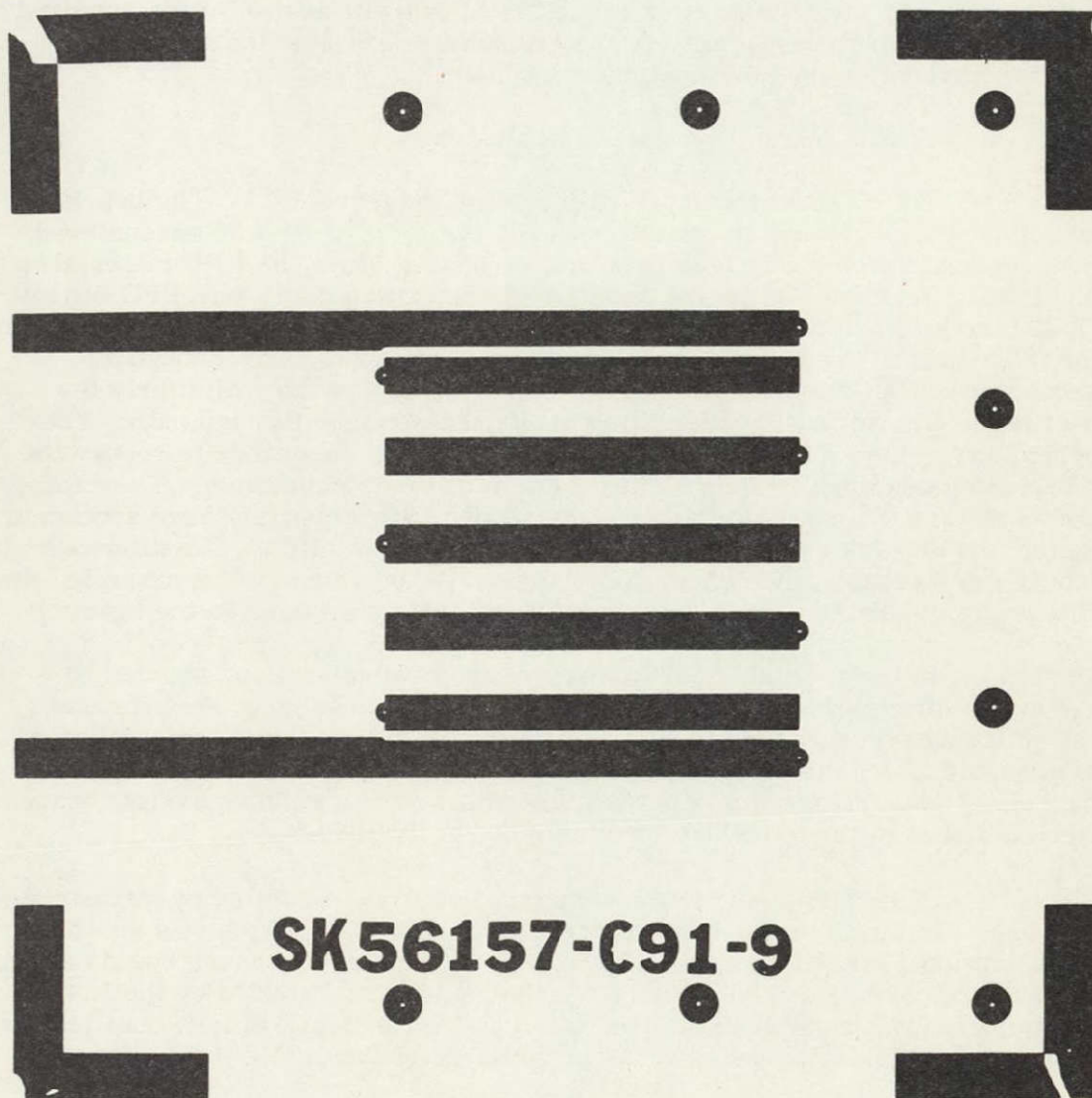


Figure 8-10. 2.13 GHz Local Oscillator Bandpass Filter using Quarter-wave Shorted Sections

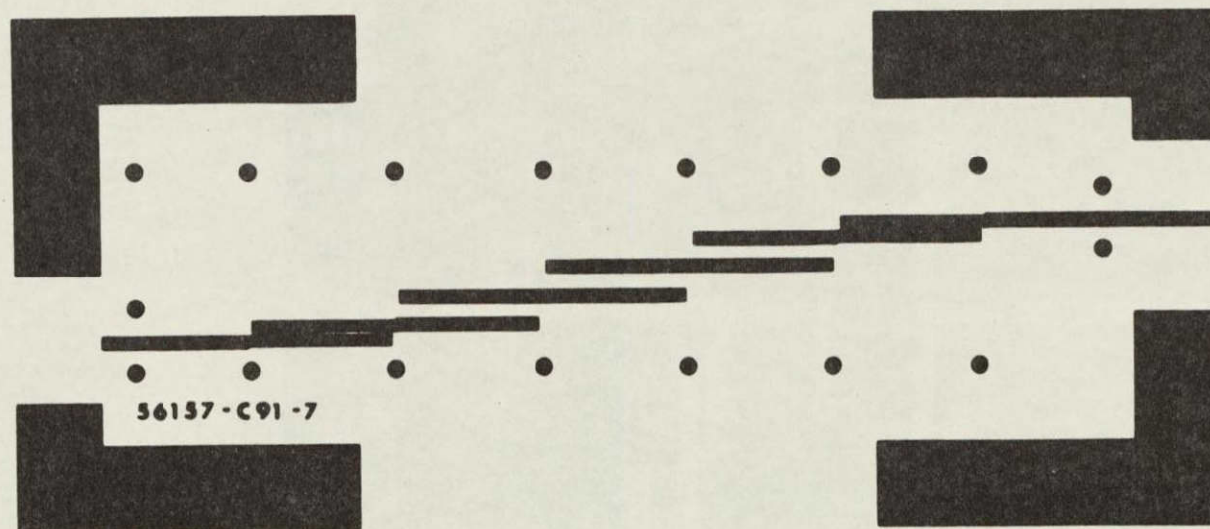


Figure 8-11. 2.13 GHz Local Oscillator Bandpass Filter using Half-wave Sections

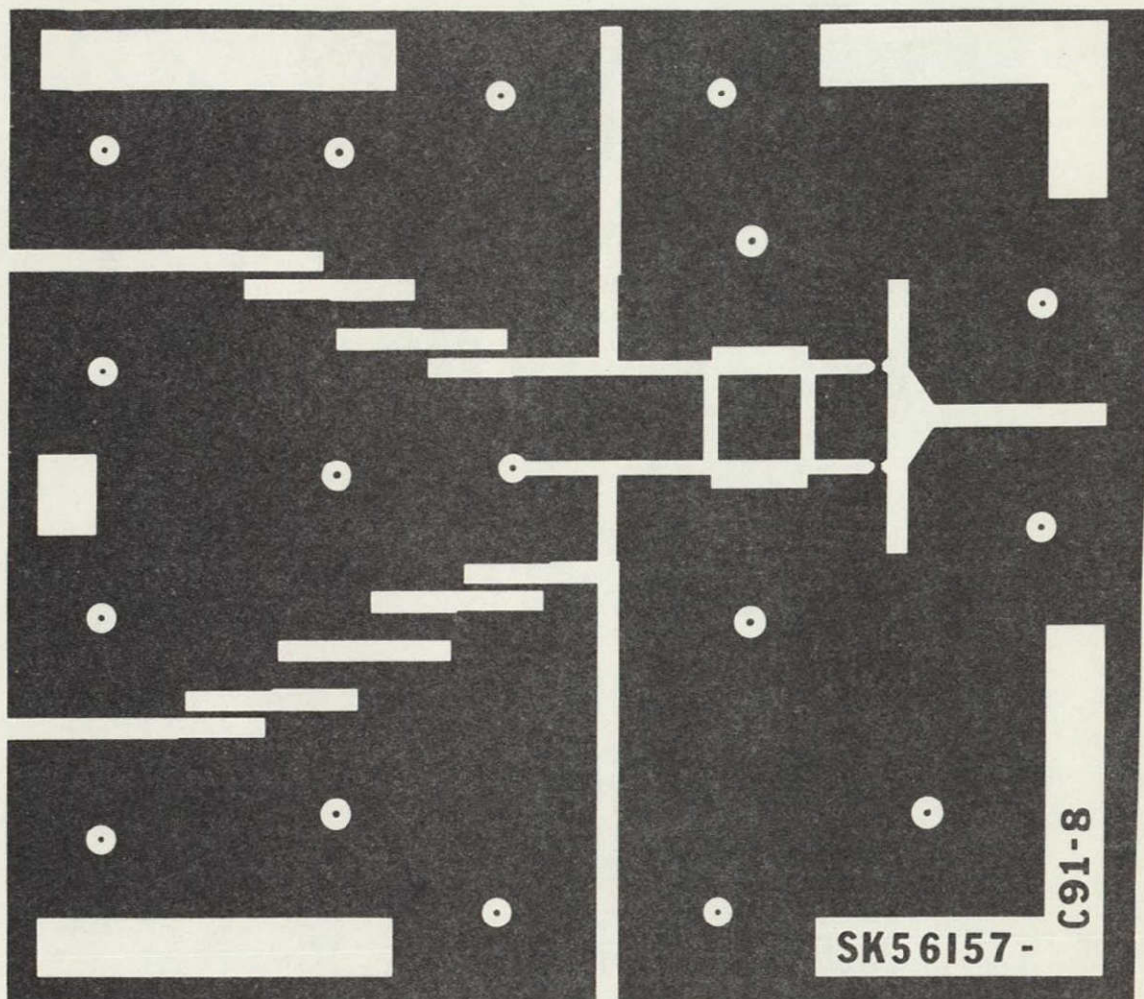


Figure 8-12. 12.00 GHz Mixer Layout

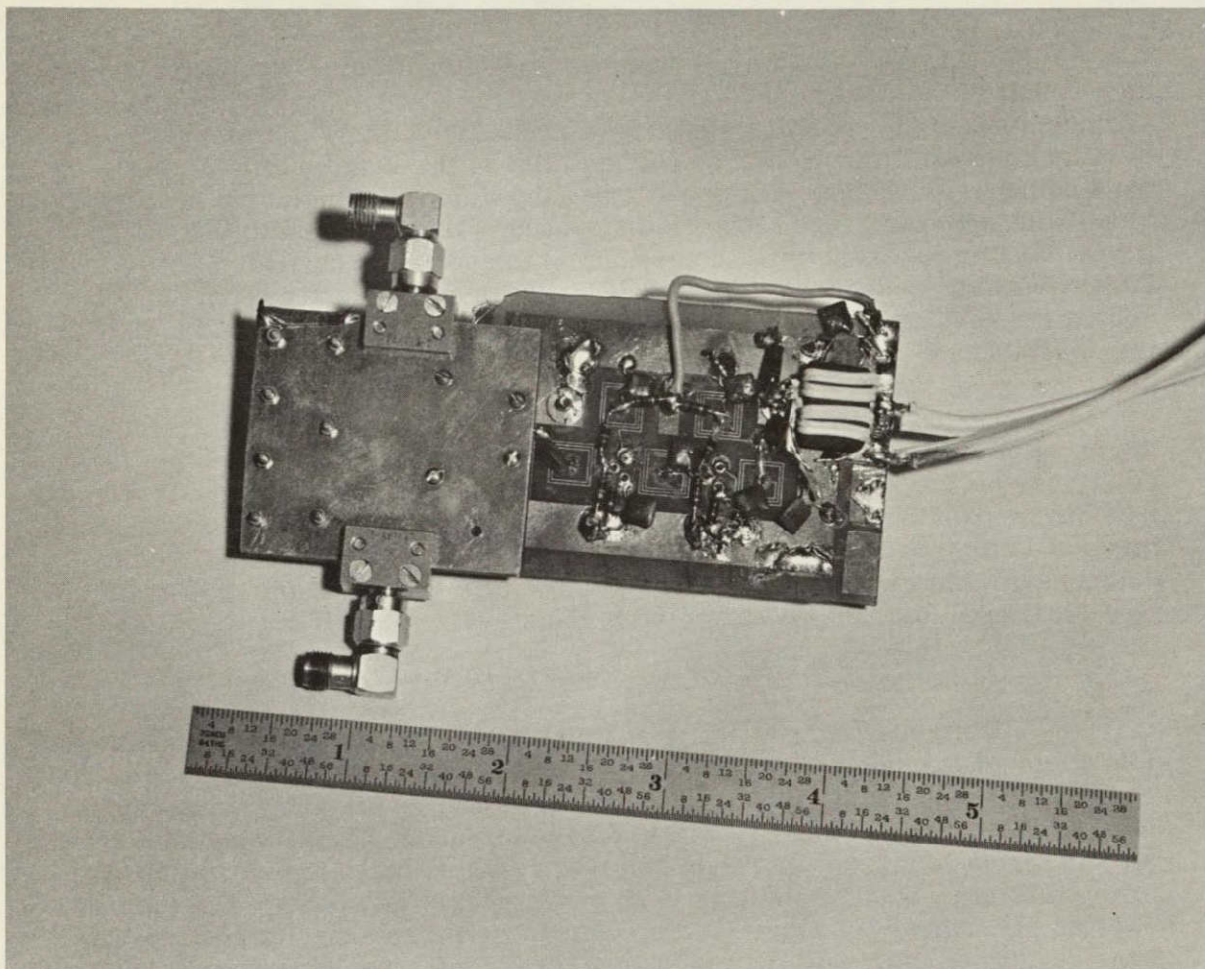


Figure 8-13. 12 GHz Mixer and IF Preamplifier Breadboard Circuits

The diodes used in the first experiment are randomly selected. The physical size of the diode, 8 mils in diameter - pigtail 13 mils long, renders handling very difficult. A special jig will be required for evaluation or matching. Image rejection was not provided in the first design. Conceivably, some improvement will be attained by lowering the conversion efficiency of the image frequency, thus reducing the 3 dB loss of signal-to-noise ratio caused by the image noise. The oscillator shunt branch will be changed to obtain a higher impedance characteristic and the impedance match, between the mixer and the IF amplifier, will be improved. A picture of the 12 GHz mixer and the IF preamplifier breadboards is shown in Figure 8. 13.

12.00 GHz Local Oscillator Breadboard Evaluation

A Varian VSX-9001 Gunn oscillator was obtained on loan for evaluation as an X-band local oscillator. The unit obtained was designed for operation around 9.8 GHz, however, power could be obtained at frequencies up to 11.00 GHz with mechanical tuning. The output power level was in excess of 25 milliwatts with a bias of 9 volts at 250 ma. The spurious FM output was evaluated using a spectrum analyzer and by observing the output of the breadboard discriminator designed for the FM systems. In the latter experiment the Gunn oscillator and a laboratory X-Band source were used as inputs to the X-Band system mixer to provide an IF output of 120 MHz. The spurious FM observed on the spectrum analyzer was in the range of 100 KHz rms. The discriminator output was noise free with no discernible evidence of spurious FM.

A temperature cycle was used to determine the oscillator detuning as a result of the cavity thermal expansion coefficient. The results indicated an uncompensated frequency stability of about 200 KHz/°C. This was an order of magnitude worse than the desired stability of the local oscillator in the X-band converter. An alumina probe was substituted for the tuning probe provided with the cavity and a modified mounting was fabricated to lengthen the probe. With this assembly a compensated temperature coefficient of less than 20 KHz/°C was observed. This procedure was subsequently repeated with equivalent results using a Gunn oscillator design for operation at 11.88 GHz, the desired local oscillator frequency for the 12.00 GHz FM converter.

A breadboard cavity was constructed to incorporate the tuning probe modification and a change from iris to capacitor-probe output coupling. This cavity has been operated at the desired frequency and power level. The capacitive probe provides smooth power level adjustments with negligible effect on the oscillator frequency. Final adjustments are required to complete the compensating probe design. A filter, of the type used with the multiplexer type local oscillator selected for the S-Band systems, will not be required.

120 MHz IF Amplifier Breadboard Evaluation

With due regard to gain and noise requirements, together with considerations of low cost, the transistor 2N4252 and, subsequently, the plastic version, SB4254, has been chosen for the IF amplifiers. The manufacturer specified the NF as, typically, 2.5 to 3 dB up to about 150 MHz and a current gain, in terms of h_{fe} , as 15 dB or better. The spot noise figure and the S-parameters of up to 50 units have been measured. The spot NF's of the units, near 130 MHz, have a mean of 2.5 dB. It is therefore expected that the mix of the regular product will give proper yield for the antenna and ground IF units. The S-parameters measured on 30 units are averaged arithmetically and are given in Table 8-3. A set of computer programs was written, so that the circuit configurations can be evaluated by iteration.

A performance print-out of a design of the antenna unit is given in Table 8-3. The circuit element values are given in the schematic diagram, Figure 5-8, which is reproduced on the following page as Figure 8-14.

Two breadboards of the antenna unit IF preamplifiers have been made per an earlier design incorporating printed coils. Each has been evaluated with the S-Band and X-Band mixers. The overall performances are presented in the discussion of the mixers. Some improvement such as (1) optimizing the current and voltage to yield the best NF, and (2) input matching over the band, is still required. The IF preamplifier printed circuit board layout is shown in Figure 8-15.

The twinlead used to connect the antenna and ground units is satisfactory in terms of matching, loss, and dc supply capability. Nevertheless, disturbances of the converter breadboards have been observed as a result of line loading due to nearby conductive objects. If this is a serious problem in the final operating environment the twinlead can be readily replaced with coaxial line at a cost increment of about one dollar. The cable terminations, baluns, can operate with a line impedance of either 300 ohms (twinlead) or 75 ohms (coax).

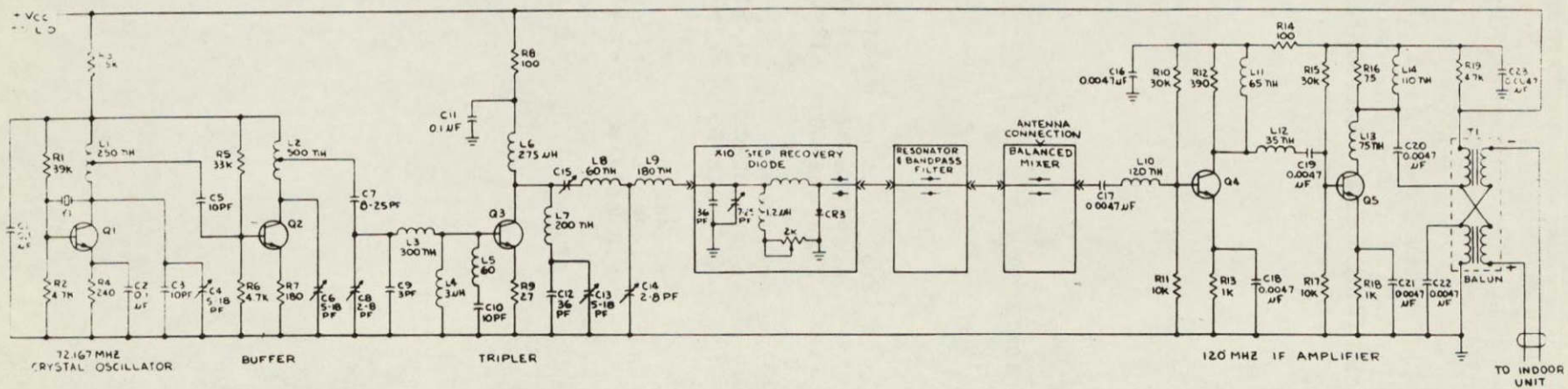


Figure 8-14. 2.25 GHz FM Converter Antenna Unit Conceptual Design

TABLE 8-3
IF AMPLIFIER CALCULATION
(Antenna Unit)

AMPLFR 16:11 SS FRI 06/13/69

NO. OF STAGES

? 2				
FREQ MHZ	S11	S12	S21	S22
100	-.10658	8.69276E-04	24.3227	.243518
J	-.392377	2.20140E-03	-12.2379	.55545
POWER GAIN DB.=	28.7003			
PHASE OF S21 IN DEGREES=		-26.7064		
120	-.186922	-1.07383E-03	-1.25378	.453425
J	.105005	3.74921E-05	-27.7334	.342927
POWER GAIN DB.=	28.8689			
PHASE OF S21 IN DEGREES=		87.4027		
140	-5.40451E-02	-1.40027E-03	-18.7424	.324022
J	.400768	2.21662E-03	-13.5501	.18746
POWER GAIN DB.=	27.2826			
PHASE OF S21 IN DEGREES=		35.8621		

SK56157-C91-3

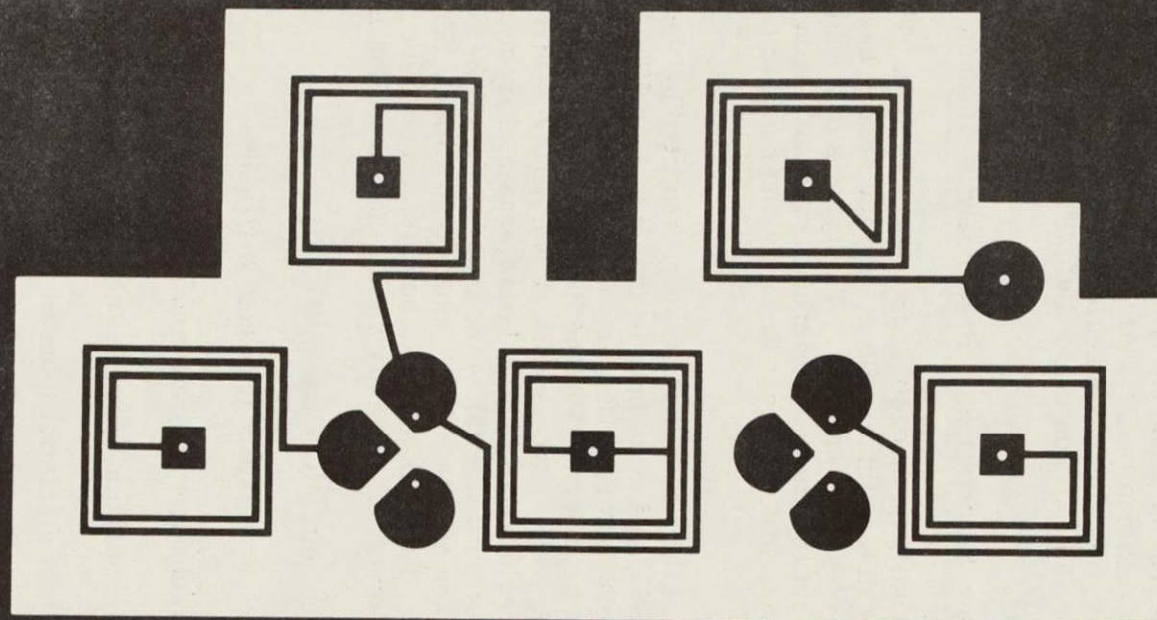


Figure 8-15. 120 MHz IF Preamplifier Printed Circuit Board Layout

The ground unit of the IF amplifier consists of one common base stage, one interstage double-tuned coupling, one common emitter stage, one broadband feedback amplifier and a filter. The total gain is about 45 dB over a band of 40 MHz. The phase delay is about 40° to 400° over the band and is fairly linear. The final performance, with the FM discriminator, will tell whether the present degree of linearity is sufficient. Because of the low Q nature of the suggested discriminator we believe the phase characteristic of IF is not as critical as when other conventional FM discriminators are used. A picture of the Indoor Unit portion of the 120 MHz IF amplifier is shown in Figure 8-16.

FM Limiter-Discriminator Breadboard Evaluation

Two discriminator breadboards were evaluated. The first breadboard consisted of the discriminator alone, and the second breadboard included the IF limiters and a post-limiter filter. The results of the first breadboard evaluation indicated that the basic discriminator circuit, using transmission lines as tuned elements, had sufficient linearity and bandwidth for this application.

A review of the line discriminator, reported by Panter^{*} in 1948 and C. W. Lee^{**} in 1963, led us to believe that it was applicable in the frequency range of concern. For an IF frequency of 120 MHz, a $1/8$ -wavelength transmission line can be fabricated in the form of etched microstrip line on a suitable substrate. According to our data, a No. 30 AWG wire, or its equivalent on a $1/16$ " single-clad-epoxy-glass board, has a characteristic impedance of 125 ohms. The discriminator takes the form of Figure 8-17, where two 125 ohm resistors, balanced to within 2%, and two $\lambda/8$ lines, one open and one short, form a bridge. The peak detected outputs, through diodes D-1 and D-2, are balanced video outputs. It is easily shown that the output is given by

$$\text{or} \quad V_{\text{out}} = V_{\text{in}} (\sin \theta (f) - \cos \theta (f)) \cdot \eta$$

$$\text{where} \quad V_{\text{out}} = I_{\text{in}} \cdot Z_0 \cdot (\sin \theta (f) - \cos \theta (f)) \cdot \eta$$

η is the diode detection efficiency

θ is the propagation phase of the lines

Z_0 is the characteristic impedance of the transmission lines

For $\theta = 45^{\circ}$, corresponding to the center frequency, the output is zero. For an IF of 120 MHz and a deviation of ± 20 MHz, the θ value will be linearly varied through 45 ($1 \pm 20/120$) $\approx 45 \pm 7.5$ degrees. The plot of $(\sin \theta - \cos \theta)$ vs.

^{*} P. F. Panter, "Modulation, Noise and Spectral Analysis," McGraw Hill, 1965 (Referred to ITT Report, 1948).

^{**} C. W. Lee, "Super Wide Band FM Line Discriminator," IEEE, 51, pp. 1675-1676, November, 1963.

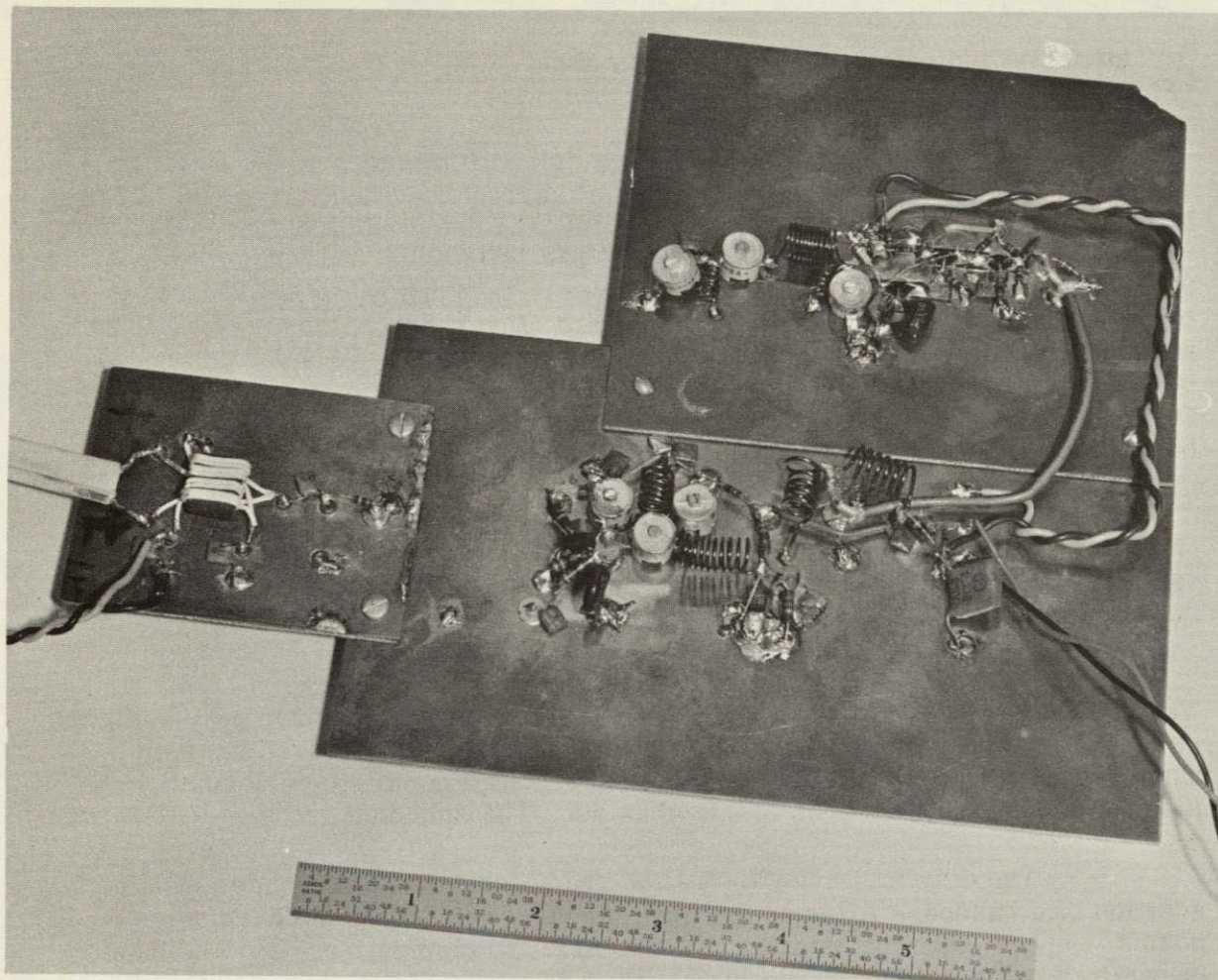


Figure 8-16. 120 MHz IF Amplifier Breadboard of FM Indoor Unit

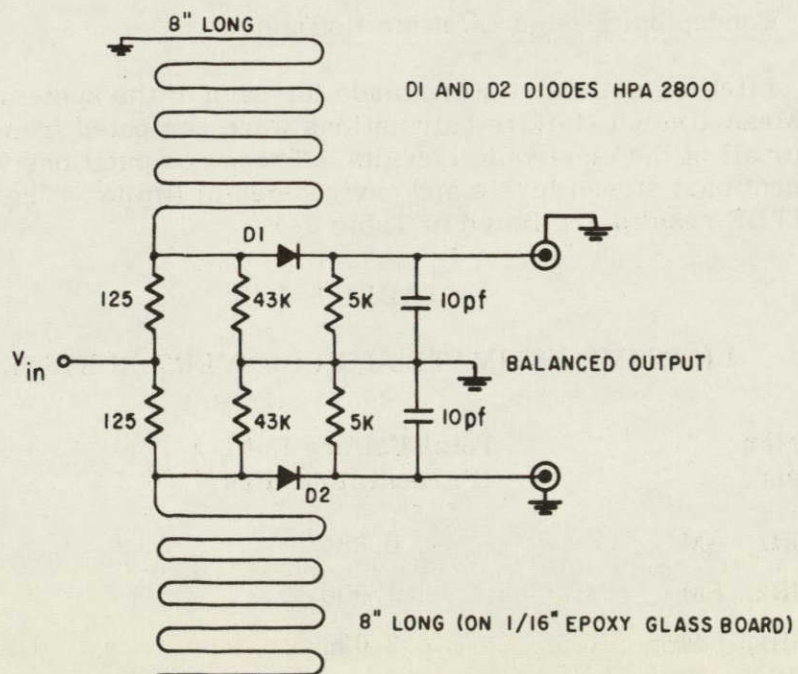


Figure 8-17. Line Type Discriminator

θ (degrees) indicates that the output is about 17 % of input voltage, if $\eta = 1$ (Figure 8-18).

The linearity measured on the original breadboard, as shown in Figure 8-19, is estimated to be better than $\pm 2\%$ over the frequency band from 100 to 140 MHz. For an input voltage of 220 Mv (rms), the detected dc output is 24 Mv for the band from 100 to 140 MHz, or 0.6 Mv/MHz. Since the input impedance to the signal is equal to the line characteristic impedance, the input current is less than 2 mA. In other words, for a 2 mA IF current, a video output of 24 mV is obtained with ± 20 MHz deviation. The efficiency of conversion could be improved by raising the signal amplitude to improve the diode efficiency. Our experience in etched line assures us that this configuration is adaptable to batch fabrication at a low cost.

If alumina substrates are used instead of epoxy glass boards, the transmission line layout can be reduced to $1/3" \times 1/3"$. This would be an advantage if the integration of other circuits can be consolidated within a simple compatible structure.

The second discriminator breadboard evaluation is not complete. The preliminary results are as follows. The limiter stages have been designed using S-parameter data to equalize the response over the desired bandwidth. The observed limiting action is as desired although complete performance measurements have not been obtained as yet. The schematic is shown in Figure 8-20.

The only major configuration change that has been made is the addition of a bandpass filter at the limiter output to remove the harmonic components generated in the limiting process prior to discrimination. The bandpass filter being used is shown in Figure 8-20.

A complete evaluation of the combined limiter-filter discriminator combination is expected to indicate that additional design modifications will be required. A picture of the second limiter-discriminator breadboard is shown in Figure 8-21.

Conceptual Design Lifetime Estimates

Lifetime estimates were made for each of the system conceptual designs. Mean-time-to-failure calculations were computed from MIL-Handbook 217 for all of the electronic circuits. These computations were based on the component part stress levels and environmental limits. The total failure rate and MTBF results are listed in Table 8-4.

TABLE 8-4

LIFETIME ESTIMATES FOR CONVERTER SYSTEMS

Converter System	Total Failure Rate, λ_T (Percent/1000 Hrs.)	MTBF (Hours)
2.25 GHz, AM	0.884	117,000
2.25 GHz, FM	1.900	52,600
12.00 GHz, FM	2.026	49,300

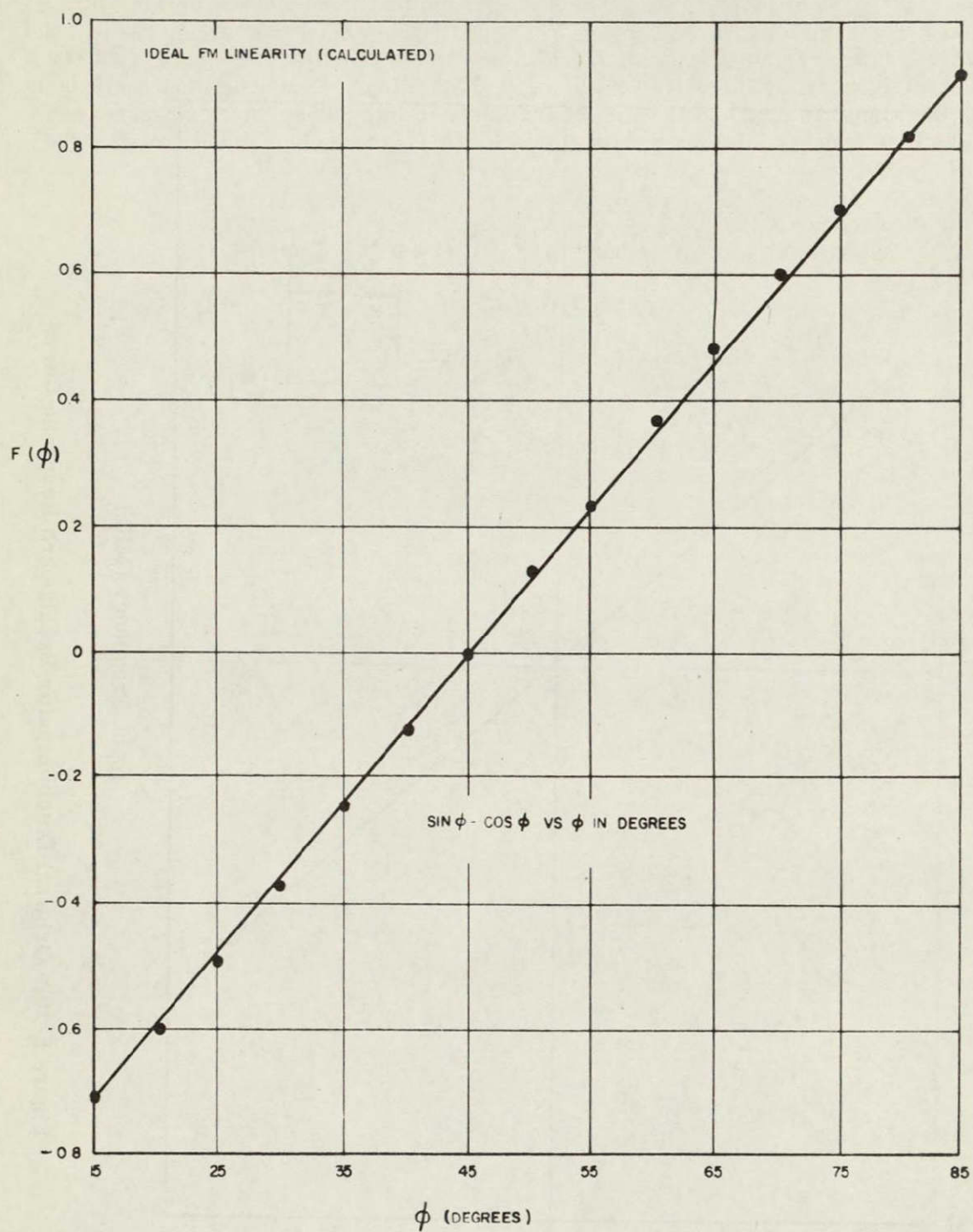


Figure 8-18. Ideal Line Type Discriminator Output vs. Phase

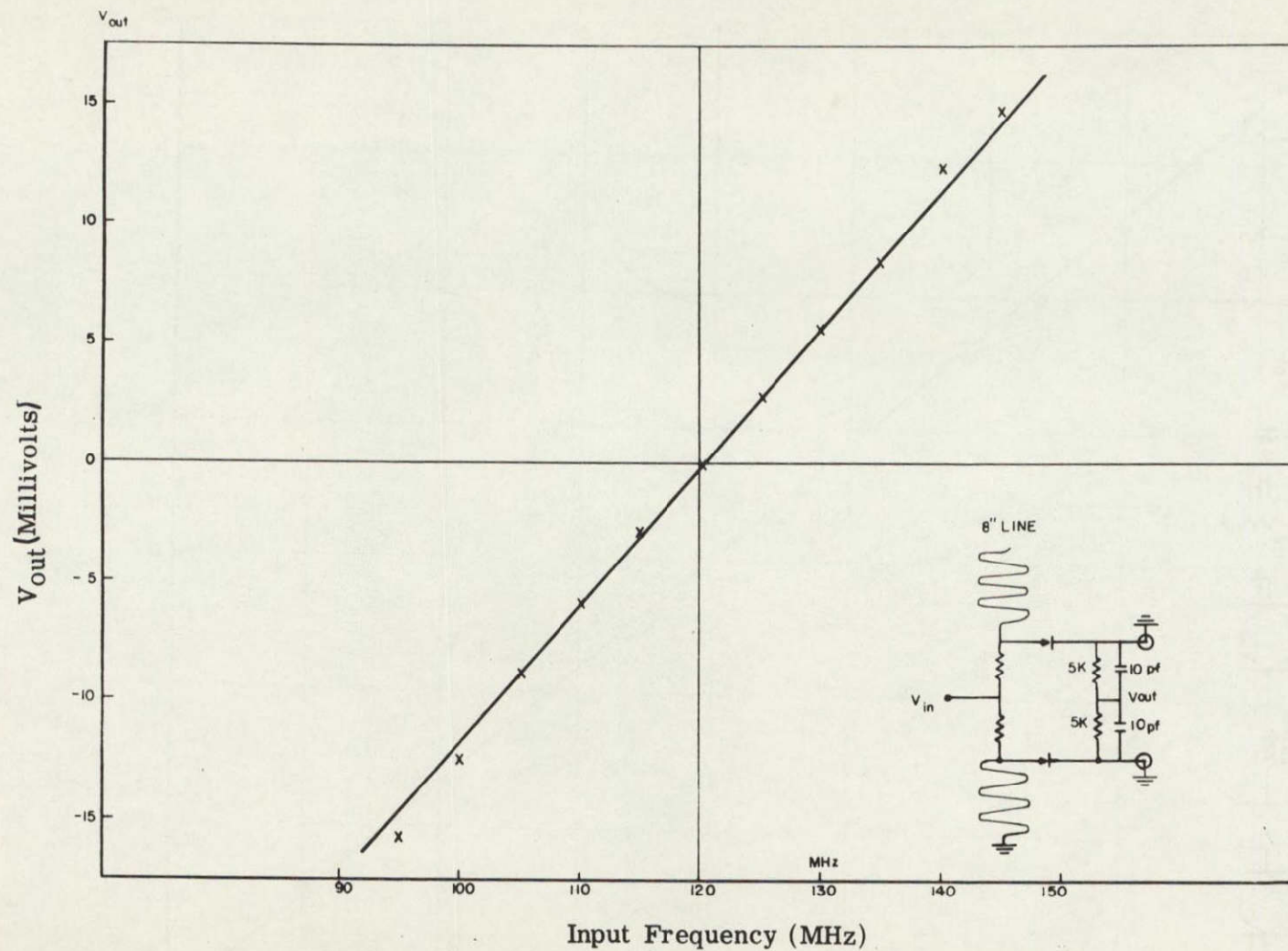


Figure 8-19. Original Discriminator Breadboard Response Curve

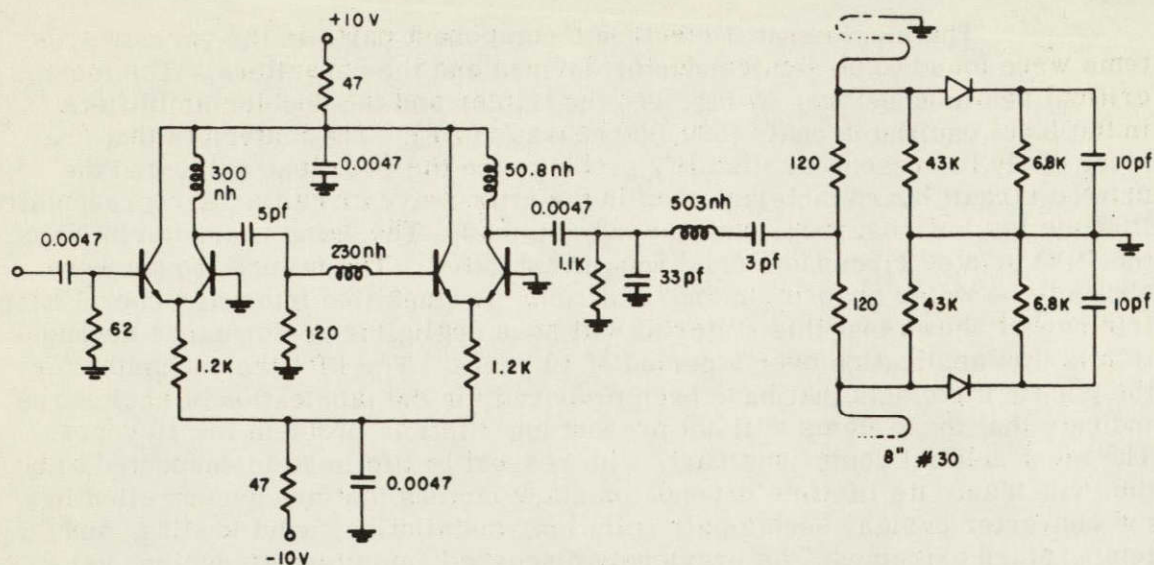


Figure 8-20. Combined Limiter-Discriminator Circuit

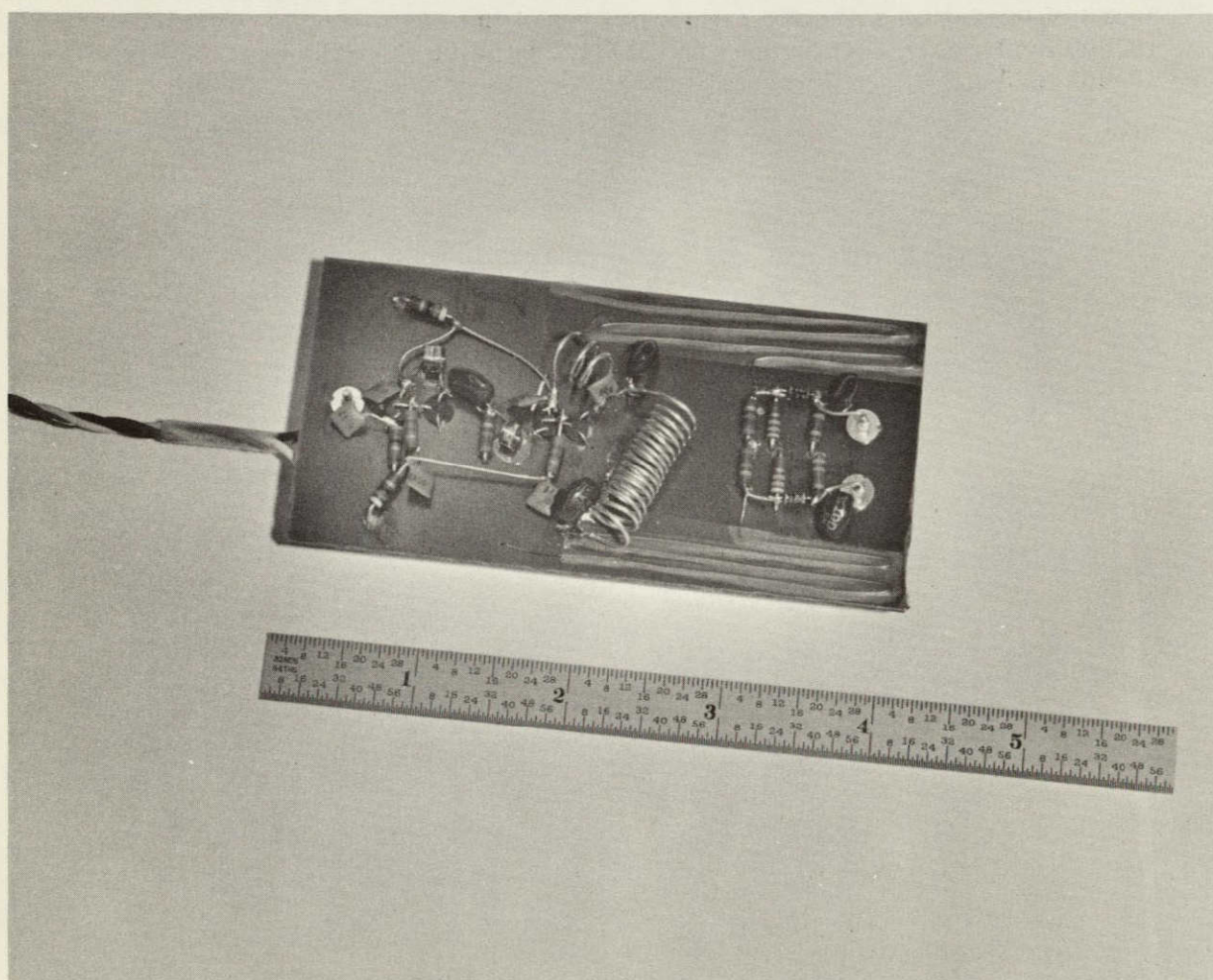


Figure 8-21. Limiter-Discriminator Breadboard

The most critical electronic component parts in the various systems were found to be semiconductor devices and the capacitors. The most critical semiconductors, by far, are the tripler and the doubler amplifiers in the local oscillator chain (500,000 hours MTBF). The materials that are most likely to present a reliability problem are the twin-lead cable and the printed circuit board material used in the microwave circuits. A representative lifetime for low cost twin-lead was not obtained. The aging characteristics of the PPO printed circuit material was investigated. The failure modes were shown to be water absorption and oxidation. A classified internal General Electric report shows that this material will have negligible performance degradation in this application over a period of 10 years. The lifetime estimates for the plastic materials that have been proposed for the fabrication of enclosures indicate that these items will not present any lifetime problem for 10 years. The most critical component part, with respect to lifetime, is suspected to be the twin-lead. Its lifetime depends on many factors that are uncontrolled in the converter design, such as air pollution, installation, wind loading, and temperature extremes. As previously discussed, an alternate design, using a coaxial lead-in, has been considered and the circuit designs can be readily adapted to its use.

Lifetime data on Gunn oscillators was obtained from Varian Associates. Sixteen oscillators currently on life test have accumulated over 100,000 hours of operation without a failure. Five of the devices have operated over 10,000 hours each. The computed MTBF at the 90% confidence level is 48,682 hours.

APPENDIX I

CONTRACTUAL PERFORMANCE REQUIREMENTS

The performance requirements that the ground system processing units must meet, and the additional performance objectives that it is desirable that they meet, are described in Sections 5.0 and 6.0, Exhibit A, Contract NAS3-11520. These sections are reproduced in this Appendix.

The requirements of the 8.45 GHz system, which was deleted by a contract modification in the latter part of Phase I, have been retained here in order to illustrate the similarity of the 8.45 GHz system to the 12.00 GHz system.

Some portions of Sections 5.0 and 6.0 of Exhibit A apply directly only to Phase II. e.g., environmental test requirements. These portions are shown here because it was necessary to take the environmental requirements into consideration from the beginning of the study; therefore, they were relevant to Phase I.

5.0 DESIGN REQUIREMENTS

For the purpose of this program the following design requirements shall apply:

5.1 Method of Transmission from Satellite to Ground Receiver

5.1.1 2.25 GHz, VSB modulation

5.1.2 2.25 GHz, FM modulation with modulation index of 2

5.1.3 8.45 GHz, FM modulation with a modulation index of 3

5.1.4 12.00 GHz FM modulation with a modulation index of 3

The modulation index is defined as one-half the carrier frequency deviation caused by the peak-to-peak video signal divided by 4.5 MHz. Where the peak-to-peak video signal ranged from the white level to the sync peak level.

5.2 Unit Location and Site

The units shall be located only at:

5.2.1 The antenna

5.2.2 The receiver

5.2.3 Part at the antenna and part at the receiver

The signal must be fed from the antenna to the receiver by means of cable or twin lead. The length of the cable shall be assumed to be 30 feet.

5.3 Color Television Receiver

A conventional NTSC compatible color television receiver shall be a part of all receiving systems. Signal processing units shall be the means of making the received signals compatible with the conventional receiver.

The conventional television receiving system shall be defined as follows:

- 5.3.1 In the VHF range the conventional receiver shall have an 8.0 ± 0.5 decibel noise figure.
- 5.3.2 In the UHF range the conventional receiver shall have 13.0 ± 1.0 decibel noise figure.
- 5.3.3 The thirty (30) feet of cable or twin lead from the antenna to the receiver shall have a loss of 2.0 ± 0.2 decibels in the VHF range and 5.0 ± 0.2 decibels in the UHF range.

5.4 Source of Power

The power supply shall be at the television receiver and be capable of supplying power through the high frequency down lead to any antenna mounted unit if possible. Either DC power or AC power shall be employed.

5.5 Safety

All systems shall be inherently safe, so that there is no electrical hazard to the user.

5.6 Unit Cost

- 5.6.1 Unit cost shall be defined as estimated manufacturing cost for units when mass produced and shall include:
 - 1. The cost of the signal processing unit.
 - 2. The cost of cable or twin lead and power cable, if required.
- 5.6.2 The unit cost shall not include:
 - 1. Installation cost
 - 2. Maintenance cost

5.7 Types of Systems

Linear amplifiers, parametric down converters, mixers, and FM to AM converters shall all be considered. FM systems with and without deemphasis shall both be analyzed.

5.8 Color Television Signal Format

- 5.8.1 The transmitted signal format for the 2.25 GHz, VSB modulation system shall be that adopted by the Federal Communications Commission (Code of Federal Regulations, Title 47, Sections 73.682 and 73.687 Revised as of January 1, 1966).
- 5.8.2 The transmitted signal for the three FM systems shall be obtained by frequency modulating the carrier signal with a composite signal formed by the sum of the color video signal as defined by CFR, Title 47, Section 73.682, and the audio signal FM modulated onto a 4.5 MHz subcarrier. Any pre-emphasis, envelope delay, or other transmitter characteristics required by CFR, Title 47, Section 73.687 and which compensate for television receiver characteristics, shall be included in a network which pre-conditions the composite signal before modulating the RF carrier. This pre-conditioning network shall also compensate for envelope delay of the signal processing unit.
- 5.8.3 Alternate locations of the sound subcarrier are to be considered if they lead to improved design or lower cost.
- 5.8.4 The VSB filter in the converter may be eliminated, provided that it is demonstrated that the resulting DSB signal does not noticeably degrade the performance of the TV receiver.

5.9 Antenna and Cable

The antenna design shall not be considered in this investigation. However, the signal processing unit shall have a connector for attaching the antenna cable or twin lead.

5.10 Antenna Output

Antenna output power levels are specified in dbw, decibels referred to 1 watt.

5.10.1 At 2.25 gigahertz:

- 1. The antenna output shall be assumed to be in the range -110 to -80 dbw for vestigial sideband transmission.
- 2. The antenna output shall be assumed to be in the range -112 to -92 dbw for FM transmission.

5.10.2 At 8.45 gigahertz the antenna output shall be assumed to be in the range of -91 to -111 dbw.

5.10.3 At 12.00 gigahertz the antenna output shall be assumed to be in the range of -90 to -111 dbw.

5.11 Antenna Output Impedance

The antenna output impedance shall be balanced 300 ohms, unbalanced 75 ohms or unbalanced 50 ohms, whichever matches the cable or twin lead. The VSWR (voltage standing wave ratio) of the antenna shall be assumed to be in the range 1.0 to 1.2.

5.12 Weight and Size

The antenna and/or mounted units must be small and light enough to permit antenna mounting and/or receiver mounting at the rear of the television receiver as applicable. The units shall be readily attachable to existing equipment without modification.

6.0 DESIGN OBJECTIVES

If it appears that any objective cannot be achieved, (for a specific ground signal processing system) the Contractor shall notify the Contracting Officer with a recommended course of action, shall perform no further work on that system until the Contracting Officer directs a resumption of work on that system in accordance with the "Changes" clause of the contract. It is anticipated that work will be resumed within two weeks from receipt of notification.

6.1 Performance

The characteristics of the signal processing unit shall be such that the performance of the signal processing unit together with the television receiver channel shall not be appreciably degraded from the performance of the television receiver channel alone.

6.1.1 Intermodulation

Intermodulation can take place between the ground, picture, and color signals causing distortion. The signal to intermodulation distortion ratio shall be greater than 40 decibels in both video and sound signals.

6.1.2 Bandwidth and Envelope Delay

The gain versus frequency characteristic of the signal processing unit shall not cause more than ± 1 decibels gain variation in the video signal for video frequencies less than 4.0 megahertz. The signal processing unit shall cause an envelope delay in the video band within the following specifications. The envelope delay relative to the average envelope delay between 0.05 and 0.20 MHz shall be zero

microseconds up to 3.0 MHz and then linearly decreasing to 4.18 MHz so as to be equal to ± 0.08 microseconds at 3.58 MHz. The tolerance shall increase linearly from ± 0.02 microseconds at 3.58 MHz to ± 0.04 microseconds at 2.0 MHz where the tolerance remains constant down to 0.2 MHz and the tolerance shall increase linearly to ± 0.04 microseconds at 4.18 MHz from that at 3.58 MHz.

The transmitted FM signal, including aural subcarrier, shall be 30 MHz bandwidth at 2.25 gigahertz and 40 MHz bandwidth at 8.45 gigahertz and 12.00 gigahertz.

6.1.3 Signal to Noise Ratio

The signal to noise ratio is defined as the ratio, in decibels, of the peak-to-peak amplitude of the picture signal (see Figure I-1) to the r.m.s. amplitude of the noise within the 4.5 MHz bandwidth of the video signal.

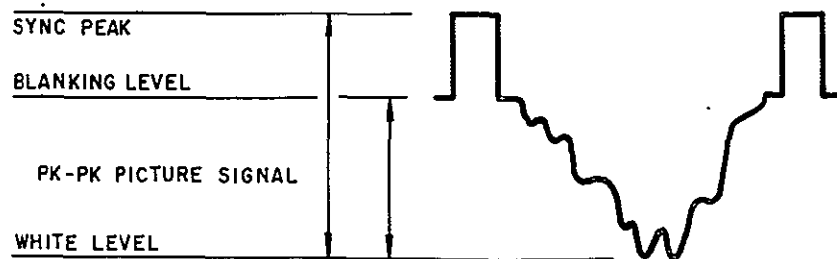


Figure I-1. Video Signal

The signal to noise ratio measured at the output of the final detector shall be greater than 35 dB for the following antenna output power levels:

1. -92 dbw for 2.25 gigahertz carrier, vestigial side-band modulation.
2. -107 dbw for 2.25 gigahertz carrier, frequency modulation with modulation index equals 2.
3. -106 dbw for 8.45 gigahertz carrier, frequency modulation with modulation index equals 3.
4. -105 dbw for 12.00 gigahertz carrier, frequency modulation with modulation index equals 3.

6.2 Effect on Conventional Television Reception

- 6.2.1 The connection of the signal processing unit to the conventional television receiving system (5.3) shall not decrease the output signal to noise ratio by more than 2.0 dB for normal television signals in the VHF and UHF ranges.
- 6.2.2 It is desirable and a design objective that the signal processing unit be constructed so that when it is used with the conventional television receiving system (5.3), tuning from satellite transmitted signals, to conventional VHF signals, to conventional UHF signals may be performed with no antenna switching. It is considered less desirable, but permissible to use antenna switching. If antenna switching is used, it shall be done at the receiver and the materials and fabrication costs charged to the signal processing unit.
- 6.2.3 The frequency band chosen for transmitting the television signals from the signal processing unit to the conventional television receiver shall not include the frequencies of television channels 3, 5, 8, 23, 25 and 61, (the television channels in the Cleveland area). Similarly, the frequencies of the television channels in the area of the Contractor's plant shall not be used. The frequency band used shall be chosen on the basis of technical and design consideration.

6.3 Environment

6.3.1 Antenna Mounted Units

1. Shall perform as specified at temperatures from -40°F to +130°F.
2. Shall be subjected to 3 g r.m.s. sine wave vibrations from 10 to 500 hertz, (10 minute sweep time) and 2.0 g r.m.s. sine wave vibration from 500 to 2,000 hertz, (4 minute sweep time), along each axis. Shall be subjected to 4 g r.m.s. random vibration from 10 to 2,000 hertz for two minutes on each axis.
3. Shall be subjected to 20 g peak shock for up to 10 milliseconds on each axis.

6.3.2 Receiver Mounted Units

1. Shall perform as specified at temperatures from 30°F to 140°F.

2. Shall be subjected to 1.5 g r.m.s. sine wave vibrations from 10 to 500 hertz (10 minute sweep time) and 1.0 g r.m.s. sine wave vibrations from 500 to 2,000 hertz, (4 minute sweep time), along each axis. Shall be subjected to 2 g r.m.s. random vibration from 10 to 2,000 hertz for two minutes on each axis.
3. Shall be subjected to 20 g peak shock for up to 10 milliseconds on each axis.

6.3.3 Failure During Tests

Units that fail during the shock and vibration tests are not required to be repaired or replaced. However, notification of the failure of any unit is to be received by the Contracting Officer within one day of the failure. Failure shall also be reported in the next monthly report to NASA. The report shall include the nature of the failure, the suspected cause of failure, test conditions at time of failure, recommended design changes (if any), and pictures where appropriate.

6.4 Spurious Radiation

Radiation of spurious signals from the ground signal processing unit and associated equipment shall conform to FCC regulations and current industry standards.

APPENDIX II

ANALYSIS OF VISUAL/AURAL FREQUENCY MODULATION

The object of this analysis is to provide a quantitative basis for the selection of satellite signal modulation parameters and converter performance parameters in the FM converter systems. The satellite transmission includes the video and the audio information required for television. The manner in which these signals are formulated, i. e., the manner in which the satellite transmitter is modulated, affects the bandwidth requirements of the ground converter and the likelihood of undesirable signal intermodulation in the system. Three options have been identified for the composition of the satellite transmission as applicable to these converter systems. The most straightforward approach would be the use of a composite base-band video signal plus audio subcarrier to deviate the satellite transmitter. The audio subcarrier amplitude would be chosen to produce an audio subcarrier amplitude, after FM demodulation and AM remodulation, that would meet the FCC television broadcast standards, which specifies the audio subcarrier power to be ten to fifty percent of the visual carrier sync peak power. This means the audio subcarrier peak amplitude relative to carrier at sync peak would fall between 0.316 and 0.707.

The second option would specify a lower audio subcarrier to sync peak ratio prior to modulation of the satellite transmitter. This approach requires a separate gain channel in the ground converter for the audio subcarrier signal in order to re-establish the desired video-audio output power ratio in the remodulated signal. This approach offers the additional advantage of permitting improved dc restoration on the video sync pulse in the converter video channel prior to AM remodulation. This clamping is not encumbered by the audio subcarrier signal that would be present in a common amplifier system as described in the first option. This approach, using the dc restoration, permits the output of the discriminator to be ac coupled while maintaining the proper modulation index at critical portions of the video signal such as the white level. Furthermore, with the discriminator output ac coupled, the remodulation circuits are more tolerant to local oscillator drift in the converter front end.

The third option covers all methods wherein a separate carrier is transmitted from the satellite to carry the audio information. These schemes are complicated by the narrowband characteristic of the audio signal (deviation of ± 25 kHz) which would be impossible to process without the establishment of an intercarrier relationship with some portion of the FM video signal. Without this technique the audio FM would be destroyed by the incidental FM in the microwave transmitters and local oscillators. One advantage of the latter approach would be the protection against audio-video intermodulation in the

converter processing. A version of the latter scheme is used in terrestrial television relay via microwave links, the exception being that a completely different link is used for audio transmission.

The second option offers the most advantages for the satellite converter systems but has the potential disadvantage of introducing audio-video inter-modulation at both the satellite transmitter and the converter circuits through the discriminator. The minimization of this potential problem can be realized by providing adequate converter bandwidth while using the minimum audio sub-carrier amplitude (in the satellite modulating signal) that will produce a satisfactory audio signal to noise ratio in the television receiver.

This analysis therefore is directed first at defining the minimum audio subcarrier amplitude that will produce an audio output signal to noise ratio of 40 dB with a 30% test signal modulation at the satellite transmitter. The second part of the analysis provides estimates of RF bandwidth required for FM color television signals with various color saturation levels and with different audio modulation parameters.

Audio output specifications for the television converter are assumed as follows. The output signal to noise ratio should be at least 40 dB when the carrier is modulated to 30% of the maximum deviation limits with a test signal of 400 Hz. The standard audio pre-emphasis prescribed for television audio modulation will be used with a time constant of 75 μ sec. The audio IF bandwidth for the television receiver is 200 KHz. The audio output signal to noise ratio objective will be met when the converter discriminator is operating at a carrier to noise level of 11 dB rather than the expected nominal 13 dB. This objective is used as a concession toward accordance with the television accepted practice wherein good quality audio is received for any perceptible grade of video signal.

Audio Subcarrier Modulation Parameters

Audio bandwidth 15 KHz

Pre-emphasis time constant 75 μ sec = $\frac{1}{2\pi \times 2100}$

Deviation ± 25 KHz maximum for 100% modulation

± 7.5 KHz maximum for 30% modulation

The FM noise improvement, I_{FM} , in the audio channel through the use of pre-emphasis is assumed to be equivalent to reducing the audio post discriminator bandwidth from 15 KHz to 2.1 KHz. This factor is accounted for in the television receiver detector by substituting in the standard FM improvement formula.

$$I_{FM} = \frac{3}{2} (m^2) \frac{\text{Prediscriminator Noise Bandwidth}}{\text{Post Discriminator Noise Bandwidth}}$$

The modulation index, m, under test conditions is defined as

$$m = \frac{\Delta_f \text{ peak caused by audio subcarrier}}{\text{Maximum Audio Frequency}}$$

$$m = \frac{7.5 \text{ KHz}}{15 \text{ KHz}} = 0.5$$

$$I_{FM} = \frac{3}{2} (0.5)^2 \frac{200 \text{ KHz}}{2.1 \text{ KHz}}$$

$$I_{FM} = \frac{3}{2.1} \cdot 0.25 \cdot 100 = 35.7$$

$$I_{FM_{dB}} = 10 \log_{10}(35.7) = 15.5 \text{ dB}$$

It can be seen that the test signal output signal-to-noise ratio objective for the television audio channel will be exceeded as long as the television receiver audio demodulator is operating with an input S/N of the 40 dB - 15.5 dB or 24.5 dB. This signal to noise ratio represents the value required at the output of the converter discriminator in the audio subcarrier amplifier bandwidth. It is assumed that the remodulation process performed in the converter and the subsequent "reception" of the converter output by the television receiver will cause no further signal to noise degradation of the audio subcarrier.

Knowing the output signal to noise ratio of the audio subcarrier at the converter discriminator allows the computation of the effective modulation index of the audio subcarrier signal on the satellite transmitter. A modified version of the FM improvement formula is used to account for the fact that the audio subcarrier represents energy at the upper end of the discriminator output bandwidth where the FM noise reduction is less effective. The FM improvement of the audio subcarrier (a modulating signal on the satellite carrier) when detected in the converter discriminator is a function of the effective modulation index of the subcarrier.

$$I_{FM_{dB}} = 10 \log_{10} \left[\frac{1}{2} \cdot (M_{SC})^2 \cdot \frac{\text{Converter IF Bandwidth}}{\text{Audio Subcarrier Amplifier Bandwidth}} \right]$$

A previous assumption was made that the minimum carrier to noise ratio at the converter discriminator input was 11 dB. The FM improvement through the converter discriminator for the audio subcarrier is therefore

$$I_{FM_{SC}} = C/N \text{ TV Disc.} - C/N \text{ Converter Disc.} = 2.45 \text{ dB} - 11 \text{ dB} = 13.5 \text{ dB}$$

The preceding relationship can therefore be solved to determine the effective audio subcarrier modulation index on the satellite transmitter and hence the minimum audio subcarrier amplitude with respect to the video signal at the satellite modulator.

$$M_{SC}^2 = 10 \frac{13.5}{10} \cdot 2 \left[\frac{200 \text{ KHz}}{30 \text{ MHz}} \right] = 0.298$$

$$M_{SC} = 0.545 = \frac{\Delta f \text{ peak}}{4.5 \text{ MHz}} \quad (\text{main carrier})$$

$\Delta f = 2.45 \text{ MHz}$ peak deviation of main carrier by audio subcarrier

The defined deviation by the video structure is as follows:

$$M_r = 2 = \frac{\Delta f \text{ peak-to-peak}}{2 \cdot 4.5 \text{ MHz}} \quad (\text{white to sync tip})$$

$\Delta f = 18 \text{ MHz}$ (white to sync tip)

Since white is defined as 0.125 sync tip,

$$\Delta f \text{ (zero to sync tip)} = \frac{18}{0.875} \text{ MHz} = 20.6 \text{ MHz}$$

The ratio of the peak audio subcarrier to the sync tip amplitude is therefore,

$$r = \frac{2.45 \text{ MHz}}{20.6 \text{ MHz}} = 0.119 \text{ or } \approx 12\%$$

Notice that this estimate is based on the assumption that the total converter RF bandwidth is 30 MHz, and that the FM deviation is linear or constant as a function of frequency.

The value of effective modulation index for the audio subcarrier varies inversely as the square root of the assumed RF bandwidth of the converter; therefore its sensitivity to an error in the bandwidth is small.

The 12.00 GHz system represents a different case because of the different video modulation index. The assumed RF bandwidth is 40 MHz as will be shown in subsequent analyses. The audio subcarrier C/N at the TV receiver discriminator is the same as in the previous analysis because none of the parameters are changed in the TV receiver. This value is 24.5 dB. The FM improvement realized with respect to the audio subcarrier in the converter discrimination will be modified

$$I_{FM_{SC}} = \frac{1}{2} \cdot (m_{SC})^2 \cdot \frac{RF \text{ Bandwidth}}{200 \text{ KHz}}$$

$$m_{SC}^2 = 10^{1.35} \times 2 \cdot \frac{0.2}{40} = 0.224$$

$$m_{SC} = 0.473$$

$$m_{SC} = 0.473 = \frac{\Delta F \text{ (peak audio subcarrier)}}{4.5 \text{ MHz}}$$

$$\Delta F_{SC} = 4.5 \text{ MHz} \cdot 0.473 = 212 \text{ MHz}$$

The deviation from the video signal is defined as $3 = \frac{\Delta F \text{ (white to sync tip)}}{2 \times 4.5 \text{ MHz}}$

$$\Delta F_{\text{video}} = 27 \text{ MHz for white to sync tip}$$

$$\Delta F_{\text{video}} = \frac{27 \text{ MHz}}{0.875} \text{ for full sync tip amplitude}$$

The ratio of peak audio subcarrier to sync tip is therefore

$$\frac{2.12 \text{ MHz}}{30.9 \text{ MHz}} = 0.069 \approx 7\%$$

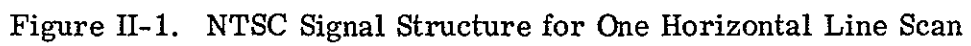
The magnitude of the audio subcarrier relative to the sync tip amplitude is therefore less when the wider video deviation is used.

RF BANDWIDTH ESTIMATES

The next step in the analysis is to determine the deviation limits resulting from the various signal components that make up the composite video and audio subcarrier waveform that modulates the satellite carrier. Following this, the significant "sideband" signal components from the resulting FM can be determined, permitting the estimation of the total RF bandwidth requirements under different sets of ground rules.

Full Gamut NTSC Color Transmission, 100% Saturation.

The total video signal limits for 100% saturated colors cover a range of 1.059 times the sync tip amplitude. The color subcarrier amplitudes, as referred to the sync tip amplitude, are as follows: (see Figures II-1 and II-2).



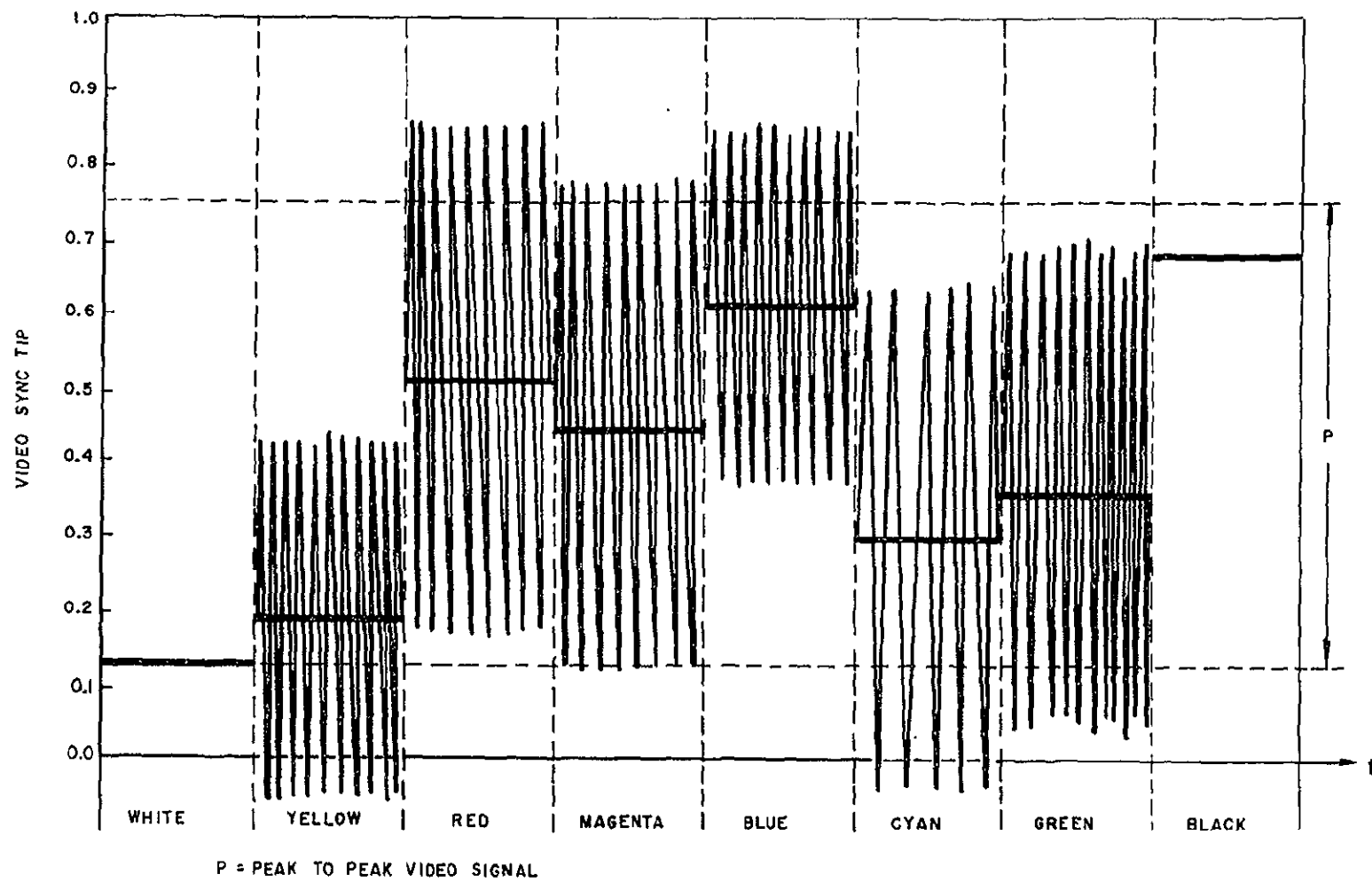


Figure II-2. Video Structure for One Horizontal Line of a Color Bar Signal

One of the video signal structures is illustrated in Figure II-1. This curve represents one side of the RF envelope of an amplitude modulated TV broadcast received by NTSC color receivers when the audio signal is not included.

Figure II-2 illustrates the color subcarrier detail that would exist within the dashed rectangle of Figure II-1 when a 100% saturated color bar line is transmitted. The horizontal dashed lines correspond to white and blanking levels at 0.125 and 0.750 sync tip amplitudes, respectively. Notice that the color subcarrier exceeds the zero reference in the yellow and cyan color bars. Table II-1 contains data concerning various amplitudes of luminance and color subcarrier signals for both 100% and 75% color saturation signals. The first column (Reference Level) refers to the luminance components of the various color bars. The second and fourth columns, C_{100} and C_{75} , are the peak color subcarrier amplitudes at 100% and 75% saturation. The third and fifth columns are the effective modulation indices for the color bars referred to the deviation of the satellite FM transmitter.

These modulation indices are used to compute the significant Bessel function components in the resultant FM signals. Columns six and seven, JN_{100} and JN_{75} , contain the highest significant Bessel function orders or components subject to the constraint that the energy contained in these components is greater than one percent of the unmodulated carrier power on an instantaneous basis. The last column contains the relative deviation of the satellite carrier from the reference position (reference is zero on Figures II-1 and II-2) due to the luminance component of the video signal. These frequencies are considered as pseudo carrier center frequencies during each color bar transmission and are the quasi steady-state frequencies about which the color subcarrier deviates the satellite carrier. They are also referred to as the carrier "idle frequencies" during color bar transmission.

TABLE II-1

	Ref. Level	C_{100}	M_{100}	C_{75}	M_{75}	JN_{100}	JN_{75}	f_c (MHz)
Yellow	0.187	0.246	1.40	0.185	1.05	2	2	3.85
Red	0.510	0.348	1.97	0.261	1.48	3	2	10.50
Magenta	0.448	0.324	1.84	0.243	1.38	3	2	9.23
Blue	0.612	0.246	1.40	0.185	1.05	2	2	12.60
Cyan	0.289	0.348	1.97	0.261	1.48	3	2	5.95
Green	0.352	0.324	1.84	0.243	1.38	3	2	7.25
Color Burst	0.750	0.125	0.71	-----	-----	1	-	15.45
White	0.125	0.000	0.000	-----	-----	-	-	2.58

The carrier idle frequencies during the color bar transmission are used as baselines in determining the positions of the Bessel function components of interest. These frequencies are referenced to zero video level. There are only two significant color bars that affect the bandwidth requirements on the "white" side of the spectrum; these are yellow and cyan. The second order Bessel function for the yellow color bar exists at,

$$-2 \times 3.58 \text{ MHz} + 3.85 \text{ MHz} = -3.31 \text{ MHz}$$

beyond the zero reference.

At the other edge of the spectrum, the first order Bessel function resulting from the color reference burst exists at,

$$1 \times 3.58 \text{ MHz} + 15.45 \text{ MHz} = 19.03 \text{ MHz}.$$

This places the significant color burst sideband inside the deviation limit caused by the sync tip. A "sideband" distribution of 1 MHz is, therefore, used to define the RF spectrum limit beyond the sync tip deviation for sync tip definition.

Thus, without allowing for the audio subcarrier, the following spectrum energies are defined.

On the "white" side, the furthestmost significant component from the idle frequency is the third order Bessel function resulting from the cyan color bar, and it exists at -4.79 MHz. On the other extreme, the sync tip signal structure requires 1 MHz of bandwidth beyond the peak deviation caused by the sync tip. The total bandwidth required without audio is, therefore

Deviation from Zero to Sync Tip	20.6 MHz
Cyan "Sideband"	4.79 MHz
Sync Tip "Sideband"	<u>1.00 MHz</u>
	26.39 MHz

When the audio subcarrier is added there is a need for further carrier deviation. The deviation caused by the audio subcarrier is added to the deviation resulting from the composite video waveform on each end of the spectrum. The most significant sideband components must be reanalyzed. On the sync tip end, the first order Bessel function resulting from the audio-carrier modulation must be accommodated to prevent intermodulation distortion.

On the "white" end, the audio carrier deviation is added to the peak deviation caused by the color subcarrier need for 100% saturated color bars for yellow or cyan. The sideband structure in this case consists of both the color subcarrier harmonics and the audio subcarrier first harmonic.

An effective modulation index can be computed $m_{\text{eff}} = (m_1^2 + m_2^2)^{1/2}$ where m_1 and m_2 are the modulation indices for the color and audio subcarriers, respectively.

$$m_{\text{eff}} = ((1.97)^2 + (0.545)^2)^{1/2}$$

$$= 2.04$$

For this level of modulation, the third order Bessel function is significant and should be preserved for the highest modulating frequency (4.5 MHz). This results in a "white" side spectrum out to $-3 \times 4.5 \text{ MHz} + 5.95 \text{ MHz} = -7.55 \text{ MHz}$. The total RF spectrum with the audio subcarrier is therefore estimated as

Zero to sync tip deviation	20.6 MHz
Audio sideband on sync tip	4.5 MHz
Audio-chrominance sideband on white end	<u>7.5 MHz</u>
	32.65 MHz

When this is compared to the spectrum required without the audio sub-carrier (26.39 MHz) the cost in bandwidth or carrier power level is,

$$\frac{(32.65 - 26.39) \text{ MHz}}{26.39 \text{ MHz}} = 0.237$$

in the worst case analysis.

As a practical matter some restriction of the bandwidth could be imposed. Quantitative analysis of the results is not presented. If no further bandwidth allocation was made at the sync tip edge of the RF spectrum, the audio sub-carrier would "beat" with the sync pulse in the receiver and would produce frequency components such as 60 Hz and 15 kHz in the audio output and 4.5 MHz in the video output. All of these signals could be suppressed with minor difficulty.

When the spectrum is constrained on the other end, a beat of 920 kHz, which cannot be adequately suppressed, is produced in video output. This results from the difference frequency between the audio and color subcarriers (4.50 MHz - 3.58 MHz). The magnitude of the beat is determined by the bandwidth restriction and the color saturation level. It is suggested that the bandwidth constraint on this end of the spectrum be determined experimentally.

Another matter of practical importance is the fact that the RETMA transmission standards for color television limit the video waveform excursions and, therefore, reduce the color saturation obtainable. At the same time these standards would reduce the FM carrier deviation.

The most significant difference in using RETMA modulation definitions is a reduction in the color subcarrier maximum modulation index.

The modified effective modulation index for the audio plus chrominance subcarrier combination is 0.91.

The significant order Bessel functions are up to J_2 . Thus, the sideband energy would extend to,

$$-2 \times 4.5 \text{ MHz} + 4.85 \text{ MHz} = -4.15 \text{ MHz}.$$

The total bandwidth estimate for the RETMA modulation standard (worst case) is,

Zero to Sync Tip Deviation	20.6 MHz
J_1 Audio Sideband on Sync Tip	4.5 MHz
J_2 Composite Sideband on "White" Side	<u>4.15 MHz</u>
	29.25 MHz

As before, this could be reduced 3.5 MHz on the sync tip end, with the accompanying audio subcarrier and sync tip intermodulation to,

$$(29.25 - 3.50) \text{ MHz} = 25.75 \text{ MHz}.$$

Further limitation on the "white" end of the spectrum is not considered practical but it can be investigated experimentally.

The RETMA modulation bandwidth estimates without audio are as follows:

Zero to Sync Tip Derivation	20.6 MHz
Sync Tip Sideband	1.0 MHz
J_2 White End Sideband	<u>2.31 MHz</u>
	23.91 MHz

$$\begin{array}{l} \text{Audio Signal "Cost"} \\ \text{With RETMA Modulation} \\ \text{Standards (Worst Case)} \end{array} = \frac{29.25 - 23.91}{23.91} = 22.4\%$$

SUMMARY

The estimated RF bandwidth requirements for the satellite television FM transmission, with a modulation index* of two, are:

	<u>Worst Case</u>	<u>Practical</u>	<u>No Audio</u>
Full NTSC Color 100% Saturation (MHz)	32.65	29.15	26.39
RETMA Standards 75% Saturation (MHz)	29.25	25.75	23.91

The practical bandwidth estimate will result in some intermodulation between the audio subcarrier and the sync pulses. These estimates are for the case where the video modulation index is defined as 2; that is, the 2.25 GHz FM system. A 15 kHz audio bandwidth is assumed. An analysis was made assuming a 10 kHz audio bandwidth, which resulted in an RF bandwidth reduction of less than one percent.

For the 12.00 GHz FM converter system, which will use an FM modulation index of three, a similar analysis produced the following RF bandwidth estimates:

	<u>Worst Case</u>	<u>Practical</u>	<u>No Audio</u>
Full NTSC Color 100% Saturation (MHz)	44.4	40.9	37.2
RETMA Standards 75% Saturation (MHz)	39.9	36.4	33.6

* Appendix I Section 5.1.4.

APPENDIX III

CONSUMER ELECTRONICS PRODUCTION

1.0 General

The consumer-oriented philosophy of electronics production and its consequences in consumer production techniques has been discussed in Section 6.0. This Appendix describes specific examples, radio and TV production respectively, to further illustrate the philosophy and techniques involved, and to indicate how low unit costs can be achieved.

2.0 Radio Set Manufacture

The manufacture of radio sets was selected as an example because their level of complexity is equivalent to that of ground signal processing units. Except for the absence of microwave circuits, their level of design and manufacturing difficulty is also about the same. Radio sets use transistor-discrete part techniques almost entirely.

The General Electric Radio Receiver Department (RRD) was selected as representative of a radio set manufacturer, and their philosophy and techniques were subjected to study.

The major topics of interest were:

Specifications

Design Procedures

Production Procedures

Manufacturing Cost Breakdowns and Sales Markups (Examples)

These items are discussed, in detail, below.

2.1 Specifications

The term "specification" is used by consumer electronics manufacturers; documents called specifications are issued, and in one sense the term is intended for the same purpose as military and commercial specifications, i.e., to insure that the equipment referred to meets the requirements. However, the emphasis is drastically different. Consumer electronics specifications are primarily for two purposes, to insure that the equipment:

- 1) Survives shipping
- 2) Performs adequately in the environment for which it was intended; usually, the home.

Examples of consumer specification requirements are shown below. The shock tests for a radio are:

- 1) Table model: 36" drop test in its package, on each of the 6 sides and 4 corners.
- 2) Portable: 30" drop test, unpackaged, in operation, on each of the 6 sides and 4 corners. The radio must continue to play, and the case must not crack. Minor scratches and chips are permitted.

The vibration or shake test is about 1/2 inch amplitude at 200 rpm for 1 hour. This test is meant to simulate the transportation environment, i. e., truck or train.

The operating temperature range for home use is normally specified to be 20°F to 120°F. If a specific climate is selected the temperature range can be restricted and design savings will result. In particular, it was pointed out that in specifying the antenna unit for the TV Converter, the temperature requirements for India would be different from those for Alaska and two different designs would be used. It is quite simple to meet any specific temperature at little or no cost change. This could be done quite inexpensively, as opposed to the relatively high cost of making a single converter that would operate in both places.

Humidity and other characteristics are treated similarly; the requirements are given to insure that the equipment will work in the intended environment. Usually numbers are given, but sometimes the requirement is merely that the set shall work in the specified environment.

Specifications for components — e.g., parts — are usually restricted to one parameter, and a variation is permitted, to allow for the vendor's normal distribution. The vendor is not ordinarily required to conduct elaborate tests nor make guarantees. His output is sampled on about a one unit per 1000 basis, and accepted or rejected. The vendor then might, for example, conduct automatic tests on one parameter; this can be done for a fraction of a cent per unit. (If for any reason a manual test is required, the cost is about one cent per unit). If the vendor changes his production procedures, he notifies the manufacturer and the acceptability of the resulting product is evaluated.

With this specification procedure, it is expected that 5% to 6% of the radios produced will fail in some manner and will be returned for replacement under warranty. Efforts are under way to reduce this to 2.5% to 3%.

This method of specification appears to be highly successful. Satisfactory radios have been produced and sold on a large scale in the United States, Canada, West European countries, South America, and Southern Asia, including Viet Nam and many other countries.

2.2 Design Procedures

Since the problem of cost is so important in consumer electronics, a major portion of the design approach is devoted to it. Some of the key considerations are, minimizing the:

- Number of parts used
- Costs, quality and precision of parts
- Assembly costs

Minimizing the number of parts is a circuit design problem; i. e., to use the minimum number of circuits and the minimum number of parts per circuit. Tradeoffs are, of course, involved; e. g., fewer stages of higher gain versus more stages of lower gain. Attention is also given to the size and material of the printed circuit boards, the fasteners used in assembly, the soldering technique, the mechanical layout for assembly (e. g., room for the parts), the cabinet or enclosure, knobs and other miscellaneous hardware.

To minimize the costs of electronic parts, circuits are designed to allow for statistical deviation of electrical and mechanical characteristics, as determined by actual test. Tradeoffs here may involve assembly costs, number of stages, and engineering design costs, as functions of parts quality. Costs of cases are minimized by materials, molding techniques, etc.

Assembly costs are affected by line speed, number of stations, quality of labor required, number of rejects, etc., and thus in turn by the number and mechanical quality of parts, mechanical layout, and the number and difficulty of tests performed on production items.

Typically, the tests required may include:

- Drop tests - about 6 per 1000 units.
- Electrical Tests and Adjustments
 - Alignment - By relatively unskilled labor, applying a standard signal and manipulating screws in pre-determined sequence to obtain a pre-determined pattern on an oscilloscope.
 - Measurement of sensitivity and power output; also applying pre-determined signals to obtain outputs on pre-determined scales on a meter or an oscilloscope.

- Tests of Unit in Case - crude tests - apply a pre-determined signal and obtain a go/no go result, mainly to determine that the set was not damaged while being put in its case.

2.3 Production Procedures

The assembly of a radio begins with introduction of a printed circuit board onto the assembly line, a conveyor which moves past various stations. At each station, a worker, usually a girl, inserts a number of parts into the board. When all the parts have been inserted, the parts are dip soldered automatically by a flat plate which lifts molten solder from a vat as the board moves across it. The board has been pre-treated so that solder adheres only to the desired areas. The board is inspected, the excess solder trimmed off, and any parts which go on the bottom are soldered in by hand. At successive stations speakers, etc., are attached, the set is aligned and tested and then placed in its cabinet.

There are 80 to 90 people on a given line, producing from 1000 to 5000 sets per day. With part-time production from supplemental lines, total production in a 240-day year may be as high as 1.6 million.

Once a line has reached maturity - i. e., initial learning completed - the cost per radio decreases about 4% to 5% per year, in spite of increasing labor pay scales. This is due to improvements in parts, circuits and assembly techniques.

2.4 Cost Breakdowns

The cost breakdown of a typical radio, list price approximately \$22.60, is shown below as an example.

Front end to Detector	\$6.00 (Materials, \$4.00; Labor, \$2.00)
Audio Section	\$2.00
Speaker	\$0.30
Back of Cabinet	\$0.05
Knobs	\$0.10
Labor	\$0.30
Labor Overhead	\$0.70
Engineering Writeoff	\$0.25

Manufacturing Cost	\$9.75
Cost to Wholesaler	\$14.30
(Manufacturer's Markup 30%)	

Cost to Retailer	\$15.90
(Wholesaler's Markup 10%)	

Cost to Customer -	\$22.60
List price - Dealer's Markup 30%	

Miscellaneous costs of interest include:

- Labor, \$0.03 to \$0.04 per minute
- Resistor, \$0.007 per unit
- Pretuned IF Coil, \$0.05 per unit
- Fixed Capacitor, \$0.01 per unit
- Transistor, \$0.04 per unit
- Variable Tuning Capacitor, \$0.40 per unit
- Knobs, \$0.01 - \$0.06 per unit
- Switch, \$0.02 - \$0.06 per unit

Given an FM radio, it costs about \$2.00 to add AM. Therefore, at present low-cost FM-only radios are not often made.

The radio whose costs are shown above (a small AM/FM receiver) is more complex than an AM converter but probably slightly less complex than an FM converter.

APPENDIX IV

INTEGRATED CIRCUITS

Integrated circuits are normally defined as circuits fabricated through sequential batch processes, including photolithography, diffusion, etching, etc., wherein circuits consisting of active and passive parts are constructed on specially prepared substrates. The definition has generally been expanded to include compatible thin and thick film technologies.

The advantages of integrated circuits include small size (of minor importance in the signal processing units) reliability, and potentially low individual circuit costs. Disadvantages include limited adjustment capability poor noise characteristics, an inability to incorporate significant inductive reactances, and the inclusion of undesirable parasitics.

At the beginning of the present study (Contract NAS3-11520) it was anticipated that integrated circuits would prove to be less expensive in quantity production, than conventional techniques but, as discussed in Sections 3.0 and 5.0, this did not prove to be the case. The reasons were discussed briefly in those sections. In this Appendix, the capabilities, limitations and costs of integrated circuits are discussed in more detail.

The cost effectiveness of linear integrated circuits in the converter designs is limited by two factors. First, the realization of tuned gain or selectivity in linear integrated circuits, using low cost planar processing techniques, has not been developed. The available linear integrated circuits, suitable for converter IF applications, consist of direct coupled or R-C coupled wideband gain functions. The filter function required in the applications under consideration must be provided with hybrid components, which represent additional processing or fabrication steps. Second, the designs of the linear IC's have been influenced by the need for a wide base of application, hence the IC's are seldom tailored for a specific application. This general purpose design approach also results in the need for external component parts to match the IC to the particular application.

The analysis of integrated circuit costs yield factors and related characteristics in the remainder of this Appendix comes largely from material prepared for integrated circuit courses by Integrated Circuit Engineering Corporation, Phoenix, Arizona. This material was supplemented by vendor performance specification data and circuit descriptions.

A. IC PRODUCTION COST ESTIMATES

The following derivation has been produced to supplement the Integrated Circuit Costing via vendor quotes. The prices obtained from vendors reflect several indeterminate factors that result in the quotation of component value rather than cost. The procedure outlined in this report will correspond to a vendor's calculation of internal production cost and adders.

The actual production cost of integrated circuits, in volumes greater than 10^6 units, is predominantly dependent on process yields and unit packaging costs. In addition, the yields are dependent on

- circuit dimensional tolerance
- circuit area and complexity
- tolerance of circuit performance

The IC production costs are definable in three categories.

- I. The costs accrued during the batch fabrication or wafer processes,
- II. The costs accrued on an individual chip or circuit basis,
- III. General overhead costs of the production facility and organization.

Yield factors can be assigned to each process or group of processes. These yield factors represent the ratio of circuits surviving a process to the number of good units entering a process. Based on this definition, the composite yield is the product of the individual yields and the percentage survival at any point can be computed from the yield factors of the processes completed to that point. It is important to realize that the yield factors are statistically based on prior experience and are peculiar to individual circuit designs. The yield factors are also subject to fluctuations that are process control oriented as well as human factors oriented. Yield factors will be assumed for this analysis. They will be subject to revision as more pertinent yield data is obtained for particular circuits.

B. COST FACTORS

The per chip or unit circuit cost analysis must start by determining wafer processing costs. These costs are apportioned among the marketed chips from that wafer. Costs will be added on a per unit basis for the second cost category mentioned. Overhead costs, which are partially volume oriented, will then be determined.

1. Category I Costs

The batch fabrication costs are relatively stable both in time and across vendors. Some variation exists depending on the size wafer being processed.

Nominal Wafer Material and Labor Costs for Wafer Processing

	Wafer Diameter	<u>2"</u>	<u>3"</u>
C_{WM} Wafer Material Cost		\$10.	\$15.
C_{DP} Diffusion Processes Cost		15.	15.
* C_{TF} Thin Film Processing Cost		5.	5.
C_{WT} Wafer Probe Test Cost		5.	10.
C_{WD} Wafer Dicing Cost		2.	3.

* Compatible thin film technology may or may not be used to incorporate resistors and capacitors on the chip. The cost is added where applicable.

In addition to material and labor costs in the batch fabrication, there is the charge to equipment amortization or return on capital invested in production line equipment. An average value for return is 0.30 of the investment annually. A million dollar investment in production equipment will provide a wafer throughput of 250 per week. This is a nominal value for current stock circuits. This is equivalent to 80 dollars per wafer processed in a year. The charge per wafer processes is therefore,

$$C_R = \frac{0.3}{\text{year}} \times \frac{\$80.00}{\text{wafer/year}} = \$24.00/\text{wafer}$$

The total Category I costs per wafer processed (nominal values) are summarized in the following table:

WAFER PROCESSING COST ESTIMATES			
	Wafer Diameter	<u>2"</u>	<u>3"</u>
Without Thin Film		\$56.	\$67.
With Thin Film		\$61.	\$72.

The per chip costs for Category I depend on the number of chips processed per wafer and the total production yields.

The number of chips processed per wafer is the ratio of useable wafer area to die area.

$$N_W = \frac{A_W}{A_D}$$

The assumed useable areas for 2 and 3 inch wafers are 2.25 and 4.5 square inches respectively.

2. Category II Costs

After wafer test and dicing, only the useable chips are retained. The testing to this point is not complete and only obvious rejects are eliminated to reduce subsequent packaging and test costs. The inclusion of chips that will not meet final performance specifications is reflected in the final electrical test yield factor.

Typical unit costs for Category II are listed below.

C_{DB}	Die Bond Cost	\$0.02
C_P	Package Costs	0.12
	a. Plastic Dual in Line (14 lead)	0.12
	b. Flip Chip* (14 lead)	0.12
	c. Beam Lead* (14 lead)	0.12
	d. Ceramic Flat Pak (14 lead)	0.35
	e. Ceramic Dual in Line (14 lead)	0.35
	f. Super Flat Pak (40 lead)	1.00
C_{WB}	Wire Bond Cost	
	For packages, a, d, e	0.12
	For packages b, c	0.04
	For packages f	0.40
C_{ET}^{**}	Electrical Test and Sealing Cost	0.07

* The lower package and bond costs for flip chip and beam lead options are currently offset by increased wafer processing costs. The addition of more automation in the lead connection will tend to reduce the bond cost differential to the point where the combined package and connection costs per 14 lead unit will approach \$0.16.

** The electrical test and sealing cost is dependent on test complexity. Analog circuit tests will tend to be more costly if the performance margin is tight.

3. Category III Costs

The per unit charges in Category III reflect the integrated circuit design and production start-up costs. The present engineering charges for a typical circuit range between \$50,000.00 and \$100,000.00. This charge is distributed across the expected total production of the circuit in question. For a 10^6 production volume, this represents a \$0.05 to \$0.10 per unit cost. This cost represents one of the limiting factors in small lot circuit fabrication.

The industry is applying automated design and mask fabrication procedures to reduce this charge. It is expected that the cost of design and line start-up for conventional circuits will be reduced to the vicinity of \$5,000.00-\$10,000.00 per circuit within the next four years.

Novel circuits or the application of integrated circuit fabrication to new areas will remain expensive because the automated procedures will rely on a catalog of proven techniques to be effective.

Additional costs in Category III are represented as percentage adders to cover general management, facilities and marketing costs and to cover profit and taxes. The overhead factor, K_2 , is assumed to be 0.30 of all identified costs. The profit and tax factor assumed is 0.25 of all costs plus overhead. The resultant figure represents a fair selling price for the circuits assuming the yield factors are accurate.

The analysis of cost is no guarantee that a vendor will be willing to produce devices for the prices arrived at.

C. YIELD FACTORS

Yield factors estimates are required to determine unit costs. All three cost categories defined are yield dependent.

1. Category I Yield Factors

Y_1 Diffusion Processes

Y_2 Thin Film Processes

Y_3 Wafer Probe Test

Y_4 Wafer Dicing

The total Category I yield is

$$Y_I = Y_1 \times Y_2 \times Y_3 \times Y_4$$

If no thin film process is used, a unit yield factor for Y_2 is inserted.

The yield factors will fluctuate widely and will generally be known only by the manufacturer. Y_1 , Y_2 , and Y_3 will be highly dependent on circuit design, i. e., the complexity or area of the chip, dimensional tolerances, and the circuit performance tolerances.

These yield factors are also dependent on operator proficiency and care, and can be controlled to a large extent by the manufacturer.

2. Category II Yield Factors

Y_5 Die Packaging

Y_6 Sealing and Electrical Test

$$Y_{II} = Y_5 \times Y_6$$

The electrical test yield depends predominantly on the performance tolerance of the circuit and the effectiveness of the wafer probe test in eliminating faulty dice.

The total yield, Y_T , is the product $Y_I \times Y_{II}$.

D. UNIT SELLING PRICE CALCULATION FOR MONOLITHIC SEMICONDUCTORS

The selling price for semiconductor devices and circuits that are fabricated through compatible monolithic processes can be estimated with the following formula:

$$P = \left[\frac{A_D}{A_W} \left(\frac{C_I}{Y_I Y_{II}} \right) + \frac{C_{II}}{Y_{II}} + C_{III} \right] (1 + K_2)(1 + K_3)$$

A_D = Area of component dice

A_W = Useable area of wafer

C_I = Category I processing costs (Table I)

C_{II} = Category II processing costs (Table II)

C_{III} = Category III design amortization cost

$$C_{III} = \frac{C_D}{N}$$

C_D = Design and production start-up costs for a particular circuit

N = Total production or amortization base for the design and start-up costs

Y_I = Category I process yield

Y_{II} = Category II process yield

K_2 = Overhead factor (nominally 0.3)

K_3 = Profit and tax factor (nominally 0.25)

The parameters listed above must be selected judiciously to provide an accurate price estimate. The area and cost parameters are quite well established in the industry. The chip area is a function of circuit design and a large portion of analog circuit area is used in the fabrication of monolithic resistors and capacitors. Guide lines are available for estimating chip area from a circuit diagram and are tabulated below. These values are nominal and represent a current industry standard.

1. Component Part Areas for Estimation of Chip Area

Transistors	Area (Sq. Mils)
NPN	60 to 20
PNP	75 to 30
MOS (200 μ mho gm)	6.5

Diodes	50 to 20
--------	----------

Capacitors	$\mu\mu\text{f/sq mil}$
SiO ₂	0.25
Junction	0.2
MOS	0.3
* Thin Film	0.3

Resistors (0.5 mil line width obtainable)

Base Diffusion	200-800 Ω /sq.
Emitter Diffusion	2-5 Ω /sq.
MOS	20 sq. mils
Pinch	30 sq. mils

* Thin Film	40-500 Ω /sq.
-------------	----------------------

2. Pads and Metallization Area

Ball Bond Pad	20 to 35 sq mils each
Wedge Bond Pad	10 to 20 sq mils each
Metallization Runs	1 mil centers
Beam Leads	12 mils per die side

Scribe Lanes - 2 mil width and at least 3.5 mils from any active junctions.

Scribe lanes are not required for beam lead devices.

* Extra chip area is not necessarily required for thin film components fabrication on passivated circuits.

E. SELLING PRICE ESTIMATES FOR SEMICONDUCTORS

The semiconductor devices and integrated circuits specified in the FM converter system have been costed on the basis on current lists prices and expected reduction in these prices with time. These price estimates will be supplemented in this report through the procedure described in the previous sections.

1. PA7600 Amplifier

The PA7600 circuit consists of six transistors (one used as a zener diode) and ten resistors. The total chip area can be estimated on the following basis:

	Unit Area (Sq. Mils)	Total Area (Sq. Mils)
Transistors 6	20	120
Resistors 10	15	150
Bonding Pads 11	15	<u>165</u>
		435

Allowing for metallization and scribe lanes, the estimated chip size is 25×25 mils.

The estimate chip area is 625 sq. mils (A_D).

It is assumed that this circuit could be processed on a 2" diameter wafer with a useable area of 2.25×10^6 sq. mils (A_W). The number of chips per wafer is

$$N = \frac{A_W}{A_D} = 3600$$

The C_I costs from Table I is \$56.00 per wafer. The estimated yield, Y_I , for wafer processing is 0.5.

The C_{II} costs, assuming a plastic dual in line package, from Table II is \$0.33. The estimated packaging yield, Y_{II} , is 0.40.

The development cost C_{III} is assumed to be completely amortized.

The overhead and profit factors are assumed to be 0.30 and 0.25 respectively.

The computed selling price is

$$P = \left[\frac{A_D}{A_W} \left(\frac{C_I}{Y_I \times Y_{II}} \right) + \frac{C_{II}}{Y_{II}} + 0 \right] (1 + 0.30)(1 + 0.25)$$

$$P = \frac{1}{3600} \times \frac{\$56.00}{0.5 \times 0.4} + \frac{\$0.33}{0.4} \quad (1.625)$$

$$P = \$0.078 + \$0.825 \quad (1.625)$$

$$P = \$1.47$$

The current small volume list price is \$11.00 per unit.

2. LM271 Amplifier

The LM271 circuit consists of three transistors, three diodes, and two resistors.

The chip area, A_D , is known to be 400 sq. mils.

$$A_W = 2.25 \times 10^6$$

$$C_I = \$56.$$

$$Y_I = 0.6$$

$$C_{II} = \$0.33$$

$$Y_{II} = 0.5$$

$$C_{III} = 0$$

$$K_2 = 0.30$$

$$K_3 = 0.25$$

$$P = \left[\frac{A_D}{A_W} \left(\frac{\$56.}{0.6 \times 0.5} \right) + \frac{0.33}{0.5} + 0 \right] \quad (1.625)$$

$$P = (\$0.033 + \$0.66) \quad (1.625)$$

$$P = \$1.14$$

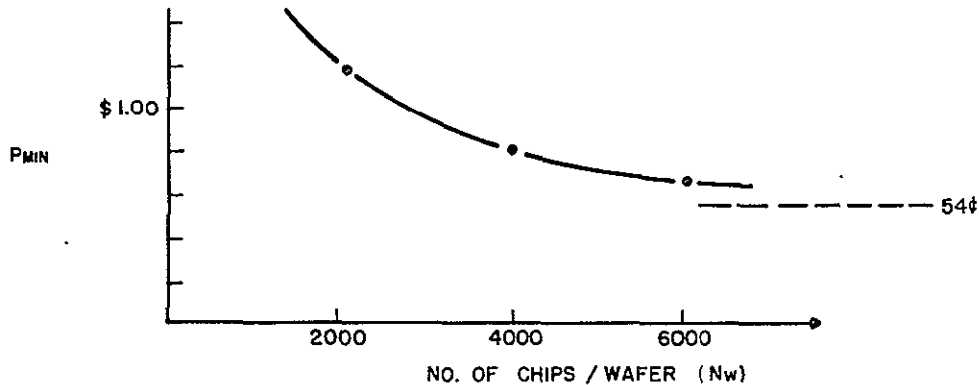
The current small quantity list price is \$4.50

3. AMELCO 710

The small quantity list prices for these circuits are already in the range of price one would obtain using this analysis. It is of interest to consider a minimum selling price based on 100% yields.

$$P_{\text{MIN}} = \left(\frac{C_I}{N_W} + C_{II} \right) (1.625)$$

$$= \left(\frac{\$56.}{N_W} + \$0.33 \right) 1.625$$



The packaging costs are predominant for all small chips having high yield factors. The industry has been attempting to transfer these packaging costs to the batch processing; for example, through the use of beam leads and flip-chip approaches. An increase in processing costs and a reduction of Y_I result, and, for the beam lead approach, the chip area increases. The reduction of C_{II} costs to \$0.23 however reduces the limiting unit cost for 100% yield to about \$0.40. At the present time, the costs are comparable considering yield factors.

A similar effort at cost reduction in the currently convention lead bonding techniques are expected to reduce the C_{II} costs to a similar value to keep them competitive. These developments, plus improvement of yield factors, will result in IC prices for 5000 sq. mil circuits that are less than \$1.00 per unit.

The stock circuits specified in the IC design for the FM signal processor should be obtainable on a guaranteed order of 10^6 units or more for the following prices.

PA 7600	2 at \$1.50	\$3.00
LM 271	3 at \$1.15	\$3.45
CA 3020	1 at \$1.00	\$1.00
CA 3028	3 at \$1.00	\$3.00
AM 710	2 at \$1.00	\$2.00
		<u>\$12.45</u>

In five years, these prices could be expected to reach the following:

PA7600	2 at \$0.75	\$1.50
LM271	3 at \$0.70	\$2.10
CA3020	1 at \$0.60	\$0.60
CA3028	3 at \$0.60	\$1.80
AM710	2 at \$0.60	\$1.20
		\$7.20

Design of new circuits to meet the functional requirements of the signal processor through the combination of existing designs can result in a further cost reduction. An estimated price for IC's to perform the FM converter signal processing is \$6.00 in 1974. This redesign will also result in the elimination of discrete R-C and L components for a nominal cost reduction. The corresponding materials cost for the FM signal processor semiconductors in discrete form is \$3.52 at the present time in volume. With expected semiconductor price decreases, this will reduce to \$2.26 by 1974. At the present time the IC's require about the same quantity of R-L-C components externally as the discrete semiconductor.

The prices for discrete transistors and diodes can be computed using this analysis. Microwave transistor prices will be limited on the lower end by packaging costs as yield factors increase for small geometry devices.

APPENDIX V

ANTICIPATED PERFORMANCE AND COST IMPROVEMENTS

A. GENERAL

The prospects for long-range improvements in cost and performance of the ground signal processing systems were discussed briefly in Section 6.0 of this report. The situation will be described in more detail in this Appendix.

As discussed in the report, there is a strong tendency for the materials costs for present designs to converge toward a constant value in about 1975. By that time, the semiconductor components parts for lower frequencies, and microwave circuitry, will be well established and will have accrued major cost reductions. Cost reductions for materials beyond that point will be of the order of only 3% to 5% per year. A major reservation regarding the above should be noted. The prediction assumes the establishment of competitive high volume markets for the devices specified in the conceptual designs. It may therefore be concluded, with fairly high probability, that ground signal conversion units manufactured to the designs described in this report will decrease only slightly in cost (of the order of 15% to 25%) between 1975 and 1980. This decrease represents the cumulative effects of reductions in manufacturing costs due to improvements in component parts circuit design, equipment assembly, etc.

However, the use of present designs, by 1980, is by no means certain. There exists a reasonable probability, based partly on historical trends and partly on advanced technology, which is expected to become available in the next few years, that major breakthroughs will occur that will make possible the use of radically different and much cheaper designs by 1980.

The most obvious possibilities for major breakthroughs are in the areas of integrated circuitry and modification of signal configurations. At present it appears that the wide ranges of signal amplitude required in the signal processing preclude the use of a single-chip unit, because of possible feedback within the chip. Also, tunable filters or frequency determining elements external to the chip are required and these drive the cost up. However, these are problems that could probably be overcome by a fundamental reapproach to the design at an IC unit, provided sufficient market existed to justify the high development cost. Further, it may be expected that the basic costs of linear integrated circuits will continue to decrease fairly rapidly as manufacturing techniques improve.

Modifications to the present signal configurations may include such approaches as the use of modified AM signals, restructuring of the baseband for FM transmission, or both.

B. PROJECTION OF PRESENT DESIGN APPROACHES

1. Low Frequency Circuitry

All circuitry and parts beyond the mixer (IF amplifier, limiters, discriminators and remodulators) will, by the 1973-1975 time period, be well established high volume devices. As can be seen from the cost tables of Section 6.0, most of them already are. Unit parts costs are very low; for example, transistors are of the order of \$0.10 each, resistors are less than \$0.01 each. Further reductions will be of the order of 3% to 5% per year. There appears to be no possibility of major breakthroughs, such as would be required to produce major reductions in costs, in conventional discrete circuitry.

2. Microwave Circuitry

In spite of the novelty of producing consumer grade microwave equipment, low cost designs have been established. However, the major portion of the converter cost is associated with the microwave components. The higher cost items are; mixer diodes, step recovery diodes, and the stripline materials and processing. There remains a certain amount of reservation about the step recovery diode. However, there is no technical reason prohibiting the development of techniques for very low cost manufacture, and it is anticipated that very low cost diodes (less than \$1.00 each at S-Band and \$4.00 to \$5.00 each at X-Band) will be available by 1975. The further reduction in price of the Gunn diode appears to be the only significant exception to the cost convergence of the present design. Between 1975 and 1980, this diode may be reduced in price by as much as \$15.00, with a corresponding reduction in the cost of the X-Band signal processing unit.

C. FUNDAMENTAL CHANGES

1. General

Major cost reductions below the projected 1975 costs, between 1975 and 1980, depend upon breakthroughs that will permit radical changes in design, manufacture and therefore costs of the signal processing units.

Prediction of such breakthrough is, of necessity, much less accurate and certain than the extrapolations that constitute the basis of the cost estimates of the present designs, and which extend to about 1975. Therefore, the remainder of this Appendix will be somewhat speculative. However, it is considered necessary and desirable to discuss the problem and make tentative predictions, in order to fulfill the intent of the contract and present some idea however tenuous, of what may be expected to happen.

As described previously the two most apparent possibilities for major breakthroughs are in integrated circuits and in the use of new signal configurations.

2. Integrated Circuits

Integrated circuits are not cost competitive in present signal processing units because of the costs of the required external tunable filters and other controls and also because of the high processing costs. A prediction of the exact manner in which the limitations of integrated circuits can be circumvented is not yet possible. However, it will be assumed here that, by beginning with the basic signal processing unit functions and designing a complex IC unit, the problems can be identified, isolated and solved, so that a one- or two-chip unit can be developed. It will further be assumed that this development can be performed by 1975, so that by that date the IC unit can be regarded as an established device, and that its cost, at that time, will be approximately the same as that of conventional units. These assumptions are based on historical trends of the last few years; major breakthroughs, such as the transistor, tunnel diode, parametric amplifier, and digital IC's have occurred on comparable time scales. If it is further assumed that development costs can be neglected, i. e., are not to be included in this program, then over the five-year period from 1975 to 1980, a decrease in costs of about 5% per year, using IC's, is reasonable, and costs in 1980 will be from 20% to about 25% of those in 1975. Again it must be emphasized that these projections are highly uncertain and have very little technical documentation. However, equally spectacular changes and cost improvements have occurred frequently in the past.

3. Modified Signal Configurations

One major possibility of the post-1975 time period is that spectrum will become so valuable, and the cost of satellite power will decline sufficiently, so that the use of AM may be justified. It is not anticipated that power will become cheap; merely that it will become significantly less expensive than it is now.

In that case, the detailed AM configuration might be modified from present NTSC standards, involving, for example, modifications of the present sync pulse, in order to somewhat decrease the required peak satellite power.

4. Additional Cost Reduction Possibilities

Additional cost reduction can result from:

- . Integration of the indoor unit into the television set
- . Integration of the antenna unit into the antenna feed

D. POST-1975 COST ESTIMATES

On the basis of the discussions of Section C, a series of assumptions regarding signal processing unit design and cost has been set up, and are listed

below. On the basis of these assumptions, a projected cost reduction schedule is shown in Table V-1 for FM circuits only. The assumptions are:

- . IC's will be developed by 1975 and will provide a 5%/year cost reduction in the IF and remodulator circuitry
- . Incorporation into the TV set beginning in 1975 will reduce the signal processing unit cost by about 30%.

The circuitry adaptable to IC's (IF and demodulator/remodulator) and the associated labor account for about 45% of the present converter costs; the remainder — mainly RF circuitry, packaging and miscellaneous hardware — the other 55%.

Incorporation of the converter into the TV set will be a subject of study during Phase II of the present program, with a cost reduction objective of about 50%.

The computations for cost reductions in Table V-1 are as follows for any date:

$$C_{IC} = 0.45C_{1975} \times (0.95)^{\text{Date}-1975} \times (0.7)$$

$$C_R = 0.55C_{1975} \times (0.95)^{\text{Date}-1975} \times (0.7)$$

$$C_T = C_{IC} + C_R$$

Therefore:

$$C_T = C_{1975} \times (0.95)^{\text{Date}-1975} (0.7)$$

where

$$C_{IC} = \text{Cost of integrated circuits}$$

$$C_R = \text{Cost of remainder of converter}$$

$$C_T = \text{Total cost of converter}$$

TABLE V-1
ESTIMATED COST REDUCTIONS, 1975-1980

<u>Year</u>	<u>Speculative Relative Costs Percent</u>	<u>Speculative Costs, 2.25 GHz FM, Production Rate of 10⁶ Units/Year</u>
1975	100.0	\$13.69
1976	66.5	9.10
1977	63.4	8.66
1978	59.6	8.15
1979	57.3	7.85
1980	54.6	7.46

APPENDIX VI

COMPONENT PARTS, MATERIALS AND FABRICATION TECHNIQUES

A. ENCLOSURES

1. Antenna Unit

The conceptual design of the antenna unit enclosure, shown in Figure VI-1, consists of a plastic or metal housing and cover. The inside of the housing is approximately $5" \times 6" \times 1\text{-}3/4"$ with a dividing wall, to form two compartments and four bosses to support and protect the antenna unit circuit boards. There are two holes in the housing for weathertight RF connections. The cover is fastened to the housing with two screws and makes a weathertight joint with the housing, as shown in the detail. Adequate mechanical protection for radio-frequency interference and solar radiation will be provided. At the present time, it is assumed that the antenna unit will be fastened to a round antenna mast as shown in the installation detail. Two screws will draw the bracket tight around the mast to the housing and hold the enclosure to the antenna mast. Different size brackets may be provided to fit the different diameter antenna masts. Fabrication techniques depend on the quantity of enclosures that are to be fabricated. With reference to Figure VI-1, it can be seen that the enclosure design lends itself to metal casting or to injection molding process as used in fabricating thermoplastics.

2. Indoor Unit

The design philosophy behind this concept is to realize a simple, attractive, economical unit. Simplicity of mechanical design is the key to keeping costs down and packaging time to a minimum. A few simple parts, loose tolerances and common materials are the factors to consider. Appearance is an important asset, because most people will have only visual contact with it. The design should conform, in every aspect, to the most economical fabrication process using available materials; but it should also be strong enough to withstand human handling.

Based on the above constraints, a preliminary design is shown in Figure VI-2. The mechanical design includes the chassis, with integrated control panel, and the cover. The chassis is well suited for packaging several different circuit boards with standard fasteners. It is cut to length from extruded plastic channel, with the required holes and slots for switches and connections. Four standard size rubber legs may be bonded or mechanically attached to the chassis, or it could be fastened to any flat plane if so desired.

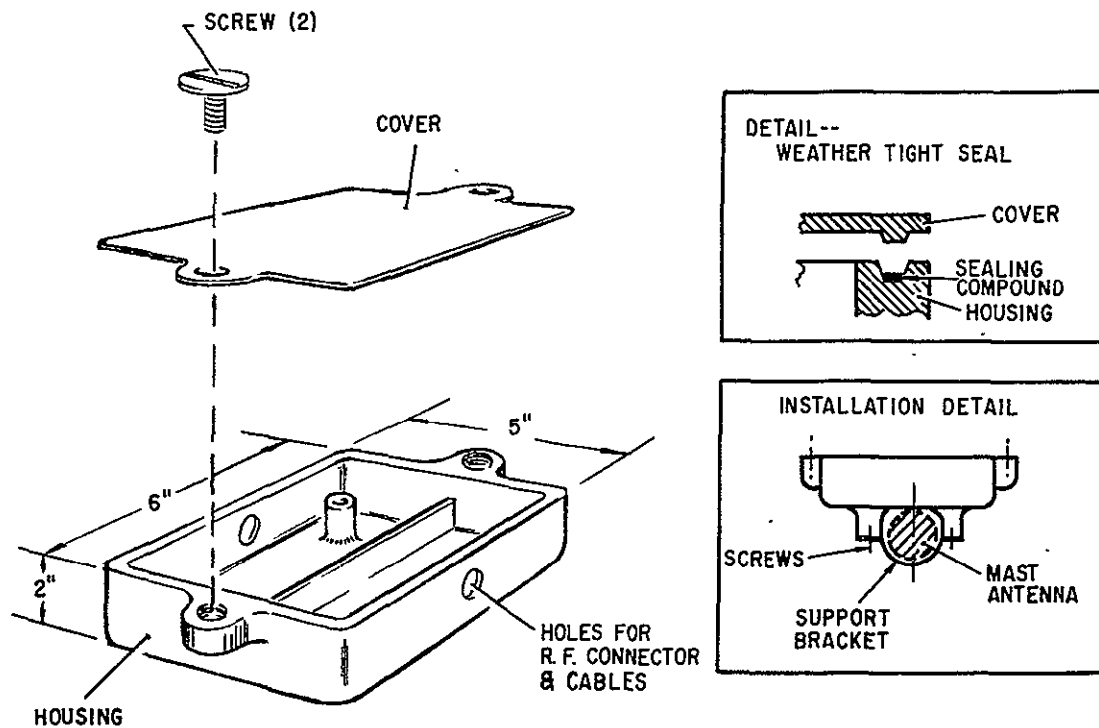


Figure VI-1. Enclosure, Preliminary Design Antenna Unit

The cover is fastened to the chassis with four out-of-sight screws and provides structural integrity and appearance. It is cut to length as shown, providing a 1/4" edge over the control panel as a protection to the control switch or cable connections.

Both parts are cut from prefinished standard structural type shapes, extruded from plastic. The extrusion is considered to be the most economical fabrication technique available for thermoplastics. Extrusion does not change the properties of the plastic significantly, but is limited to a continuous length of constant cross-section parts, as shown for this design. Moderate modifications of the form, as it emerges from the extruder, is used as a means of fabrication. The strength properties of an extruded part are suited to our application.

B. R. F. CONNECTOR

Figure VI-3 shows the design of the R. F. connector as proposed by an outside vendor (Sealectro Company). It is a bulkhead mount, solder-type connector and can be machined from hexagonal aluminum alloy stock. The inner conductor and contact strip would be made of beryllium-copper and gold plated to provide a better connection. As a bulkhead jack coaxial connector, it is

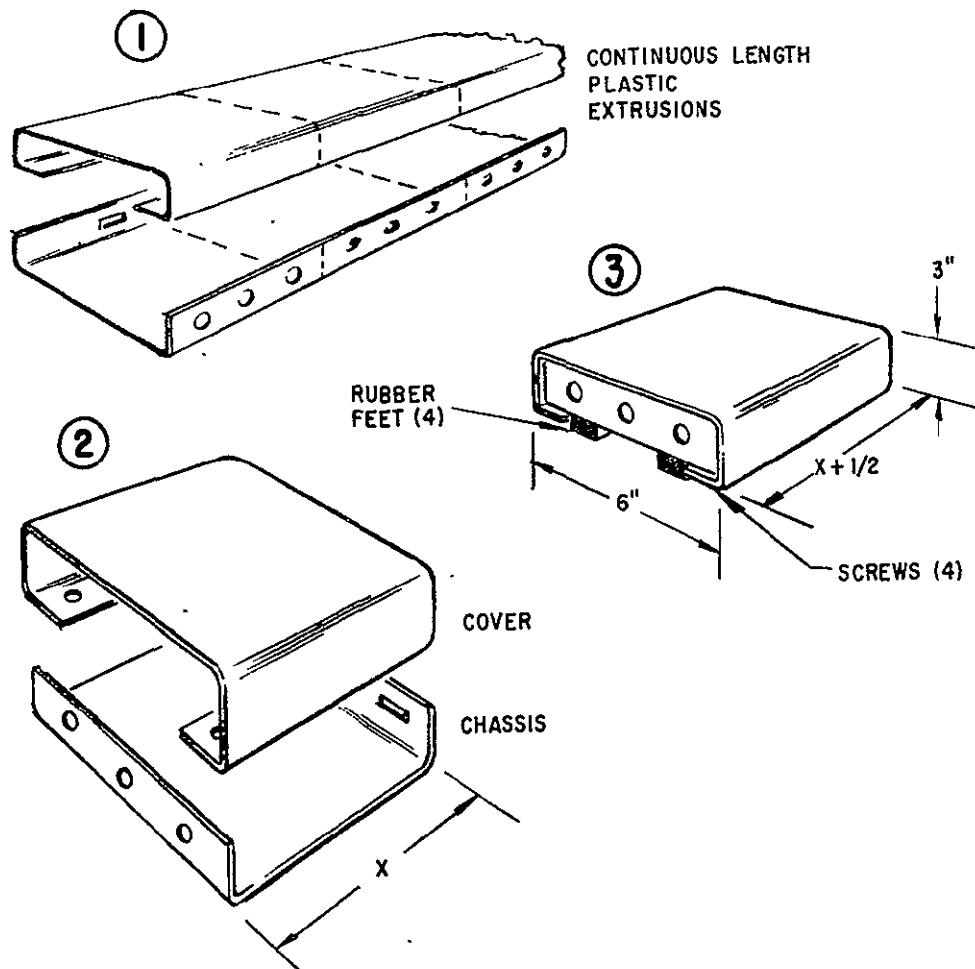


Figure VI-2. Enclosure, Preliminary Design Indoor Unit

intended to be mounted through a hole in the wall of the box and fastened with a locknut and washer on the outside and with a gasket on the inside to make a weathertight joint.

There is a mating, crimp-type plug on the antenna cable, sealed with shrink tubing. It will screw onto the bulkhead mounted connector by a coupling nut and engages the inner conductor. The connection is sealed by a gasket inside the cable plug.

The coaxial connector design will be considered and all the trade-offs, such as materials and functional requirements versus manufacturing and assembly costs, will be discussed with the NASA Project Manager.

C. SURVEY OF COPPER CLAD LAMINATES FOR MICROWAVE STRIPLINE

Accent in this survey is given to the cost of the laminates to provide the microwave designer with a starting point in the framework of this contract. Detailed information of electrical, physical and chemical properties is available

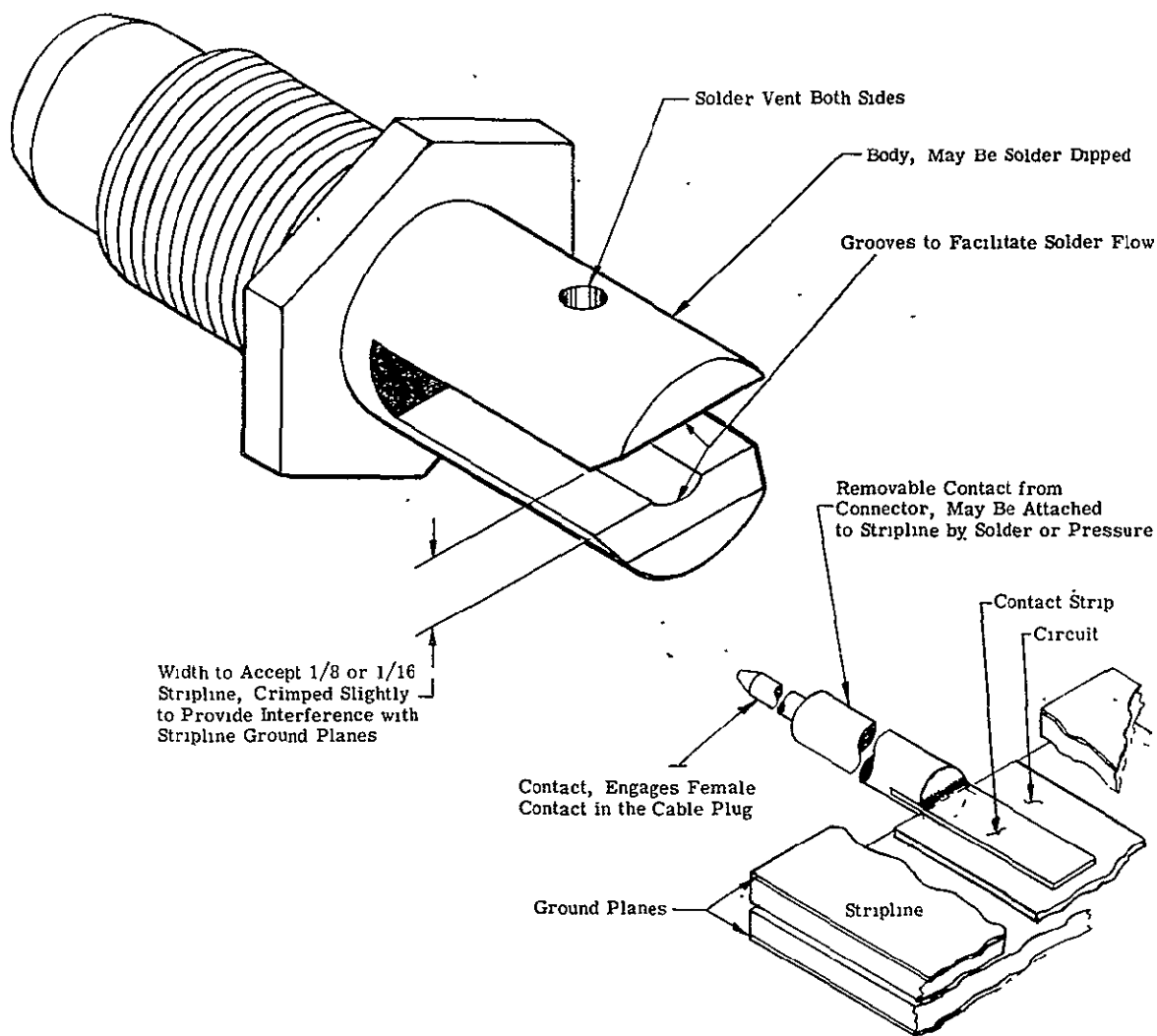


Figure VI-3. RF Connector, Antenna to Mixer, Conceptual Design

for the materials listed and should be checked in order to understand what each dielectric material can and cannot do. There are many laminates available to suit nearly every design need, but the trick is, of course, to choose the right one for the job, considering the cost and performance. The combination of dielectric constant and dissipation factor may narrow the allowable choice. Possible selections will be reduced still further by the required temperature range. Final selection from the remaining materials involves an appropriate trade-off among physical characteristics and cost. Because, hopefully, we do not need extremely good dimensional stability or a very wide temperature range, material investigated is with a dielectric constant from 2.1 to 2.7. However, within this narrow field there is a variety of mechanical, electrical and economical characteristics to consider. The survey is summarized in Table VI-1, which is comprised of 4 sheets.

As the survey shows, the least expensive dielectric laminates are the irradiated Polyolefins approximately \$0.04 per square inch and the reinforced Polystyrenes approximately \$0.05 per square inch, followed by PPO (Polyphenylene Oxide) and a glass reinforced Teflon.

Irradiated Polyolefins could be described as a punchable, chemically resistant isotropic material with a very low dissipation factor. Its major disadvantages are poor dimensional stability and low temperature range.

Cross-linked Polystyrene is a thermoset, cross-linked plastic filler or binder, which is isotropic and uniform over wide temperature and frequency ranges. It has good physical characteristics and good thermal stability.

Polyphenylene Oxide (PPO) has greater dimensional stability and temperature range than irradiated polyolefin but has higher dissipation factor and poor chemical resistance. Being unreinforced, it is isotropic.

Glass-reinforced Teflon has good physical properties, good thermal match, good chemical resistance and good processing properties. Its principal drawbacks are relatively higher dissipation factor and anisotropy.

There is no significant difference in cost between one ounce and 2 ounce copper clads or single and double sided boards. Some vendors quote both the same, some show variation only of a fraction of a cent per square inch. 1/16 inch thick laminates are widely used in the industry.

The known manufacturers were contacted and asked for cost estimates. Eight of these ten responded. Most of the responses are in the form of quotes and price lists.

In addition to the microwave stripline laminates, the following product survey (sheet 3) shows also information on copper clad textolite for a printed circuit application. Item 9 is a phenolic paper base laminate normally used in television industry, with an approximate price of $\frac{1}{2}$ ¢ per one square inch. As a rough rule, phenolic paper laminate may be used at 120°C; however, it tends to soften somewhat at temperatures about 100°C. It generally is not attacked by common solvents such as alcohol, ether, or the petroleum products, and resists low concentrations of mineral acids. It is attacked by the oxidizing acids, particularly at higher concentrations and elevated temperatures and by strong alkalies.

Item 10, glass-epoxy laminate (G-10) is also used primarily in electronic applications at room temperature. The glass fabric-base grade has good heat resistance, plus very high tensile and impact strength. Fabric base grades, in general, are tougher than paper-base grades. Copper clad laminates cost approximately one to two cents per square inch.

Sheet 4 of the product survey shows the ultra-thin copper clads for possible use in our IF filter design. Three manufacturers were contacted and verbal and formal quotes obtained. However, the prices after 90 days are to

be considered budgetary and subject to adjustment upward or downward depending on market conditions. The prices range for items 10 through 13 from 1¢ per square inch to 1.8¢ per square inch with different dielectric constants.

TABLE VI-1

 1 OF 1
 CON. NO. NAS 3-11529
COPPER CLAD LAMINATES FOR MICROWAVE STRIPLINE
PRODUCT SURVEY1/23/69
JL

LINE	MANUFACTURER	TRADE NAME	DIELECTRIC CONSTANT & VARIATION W/ T & FREQ.	DISSIPATION FACTOR & VARIATION W/ T & FREQ.	SHEET THICKNESS TOL.	MATERIAL DESCRIPTION	COST PER IN.	REMARKS
1.0	ROGERS COMPANY ROGERS, CONNECTICUT	RT/DUROID 5870	2.35±.05 THRU 10 ⁴ TO 10 ¹⁰ HZ	.0008 THRU 10 ⁴ TO 10 ¹⁰ HZ	±7%	NON-WOVEN GLASS MICROFIBER-REINFORCED TEFLON DBLE. SIDED, 2 OZ. COPPER, 1/16 THK.	25.0	SIZE OF SHEETS 16"x21"=2.33 SQ. FT. QTY. 1000 SQ. FT. & OVER
1.1	"	"	"	"	"	SINGLE SIDED, 1/16 THK., 2 OZ. COPPER CLAD	22.0	"
2.0	CUSTOM MATERIALS, INC. ALPHA INDUSTRIAL PARK CHELMSFORD, MASS. 01824	CUSTOM POLY CR	2.62 THRU -40°F TO +130°F S & X BAND	0.001 THRU -40°F TO +130°F S & X BAND	STD. TOL. ±.004 IN. GOOD DIMENSIONAL STABILITY	CROSS REINFORCED POLY- STYRENE, 1/16 THK., 2 OZ. COPPER ONE SIDE	5.49	QTY: 7000 SQ. FT. SHEET. 24"x36"
2.1	"	"	"	"	"	1/16 THK., 2 OZ. COPPER, TWO SIDES	5.65	"
2.2	"	CUSTOM POLY CP	2.32 THRU -40°F TO +130°F S-BAND	.0004 THRU -40°F TO +130°F S-BAND	STD. TOL. ±.002 IN. POOR DIMENSIONAL STABILITY	IRRADIATED POLYOLEFIN, 1/16 THK., 2 OZ. COPPER, ONE SIDE	5.2	"
2.4	"	"	"	"	"	1/16 THK., 2 OZ. COPPER, TWO SIDES	5.38	"
3.0	ELECTRONIZED CHEMICAL CO. SOUTH BEDFORD STREET BURLINGTON, MASS. 01803	POLYGUIDE #165	2.320±.005 AT 1300 MHZ	.0002±.00005 AT 1 MHZ	DIELECTRIC THICKNESS: .062±.002, COPPER 2 OZ. .0028± 10%	IRRADIATED (MODIFIED) POLY-OLEFIN, 1/16 THK., 2 OZ., COPPER ONE AND TWO SIDES	4.1	IN QTY OF: 5000 SQ. FT., SHEET SIZE: 24"x36"= 6 SQ. FT.
4.0	POLYMER COMPANY ENG. PRODUCTS DIVISION READING, PENN. 19603	Z-TRON "G"	2.55 THRU 10GHZ & 2.53-2.56 @ 8.5GHZ @ -40°F TO +130°F	.0005 @ -40°F & 60HZ .0003 @ 130°F & 60HZ .0011 @ 70°F & 3GHZ .0016 @ 77°F & 8.5GHZ	DIELECTRIC THICKNESS: .062±.001 STD. 2 OZ. COPPER .0028"	(POLYPHENYLENE OXIDE) PPO 1/16 THK., 1 OZ. OR 2 OZ., COPPER CLAD, BOTH SIDES	12.5	SHEET SIZE: 16"x24", QTY: 100 TO 150 SQ. FT.
5.0	TELLITE COMPANY 200 SO. JEFFERSON STREET ORANGE, NEW JERSEY 07050	TELLITE 3 B	2.32±.05 UNDER ALL CONDITIONS. 2.32±.01 TO 4.3 GHZ	.001 UNDER ALL CONDITIONS. .00015 TO 10 GHZ	DIELECTRIC THICKNESS: .062±.002" STD. 2 OZ. COPPER: .0028	IRRADIATED POLYOLEFIN, 1/16 THK., 2 OZ. COPPER, BOTH SIDES	4.2	SHEET SIZE: 22 1/2"x32"= 5 SQ. FT. QTY: 1000 SQ. FT. +
5.1	"	"	"	"	"	1/16 THK., 1 OZ. COPPER, BOTH SIDES	3.9	"
5.2	"	"	"	"	"	1/32 THK., 2 OZ. COPPER, BOTH SIDES	4.8	"
5.3	"	"	"	"	"	3/32 THK., 2 OZ. COPPER, BOTH SIDES	4.6	"
5.4	"	"	"	"	"	3/32 THK., 1 OZ. COPPER BOTH SIDES	4.2	"
6.0	THE BUDD COMPANY POLYCHEM DIVISION 119 ANN STREET HARTFORD, CONN. 06103	DI-CLAD 522;108 TK; 112TK	2.5 (TEST PER ASTM METHOD D-229) 2.6±.05 (CONTROLLED)	.0008 (TEST PER ASTM METHOD D-229)	±.0085 NORMAL ±.004 CLOSE ±.002 EXTRA CLOSE 112T & 112TK CLOSE TOL. IS ±.005; 112TK EXTRA CLOSE TOL. IS ±.005;	1/16 THK., 2 OZ. COPPER, ONE SIDE; TEFLON GLASS CLOTH, FINE WEAVE;	11.2	QTY: 5000 SQ. FT. SHEET SIZE: 17"x38"

TABLE VI-1 (Continued)

SHEET 2 OF 4
CONTRACT: NAS 3-11520COPPER CLAD LAMINATES FOR MICROWAVE STRIPLINE
PRODUCT SURVEY

ITEM	MANUFACTURER	TRADE NAME	DIELECTRIC CONSTANT & VARIATION W/T & FREQ.	DISSIPATION FACTOR & VARIATION W/T & FREQ.	SHEET THICKNESS TOL.	MATERIAL DESCRIPTION	COST PER IN*	REMARKS
6.1	THE BUDD COMPANY POLYCHEM DIVISION 119 ANN STREET HARTFORD, CONN. 06103	DI-CLAD 522# 108TK & 112TK	2.5 (TEST PER ASTM METHOD D-229) 2.6±.05 (CONTROLLED)	.0008 (TEST PER ASTM METHOD D-229)	±.0085 NORMAL ±.004 CLOSE ±.002 EXTRA CLOSE, 112T & 112TK CLOSE TOL. IS ±.006; 112TK EXTRA CLOSE TOL. IS ±.005;	1/16 THK., 2 OZ. COPPER, TWO SIDES	11.8	QTY: 5000 SQ. FT. SHEET SIZE 17"x38"
6.2	"	DI-CLAD 108T	2.6 (ASTM METHOD D-229)	"	"	1/16 THK., 2 OZ. COPPER, ONE SIDE	9.9	"
6.3	"	"	"	"	"	1/16 THK., 2 OZ. COPPER, TWO SIDES	10.5	"
6.4	"	DI-CLAD 112T	"	"	"	1/16 THK., 2 OZ. COPPER, ONE SIDE	9.1	"
6.5	"	"	"	"	"	1/16 THK., 2 OZ. COPPER, TWO SIDES	9.7	"
7.0	AMERICAN ENKA COMPANY BRAND-REX DIVISION WILLIMANTIC, CONN. 06226	REXOLITE 1422	2.53 AT 10MHZ - 10GHZ THRU -60° TO +140°C OVER 1MHZ TO 10GHZ UNCHANGED.	.00012 @ 1MHZ @ 25°C .00025 @ 10MHZ @ 25°C .00066 @ 10GHZ @ 25°C	±.003 DIELECTRIC; ±.0007/- .0003 COPPER	CROSS-LINKED STYRENE COPOLYMER 1/16 THK., 2 OZ. COPPER, TWO SIDES. (NO ONE S. AVAIL)	8.3	QTY: 10,000 SQ. FT., SHEET SIZE: 34"x36" (8 SQ. FT.) FOR 12GHZ APPL.
7.1	"	REXOLITE 2200	2.62 @ 10MHZ-10GHZ MORE STABLE THAN 1422	.0004 @ 1MHZ @ 25°C .0005 @ 10MHZ @ 25°C .0014 @ 10GHZ @ 25°C	"	GLASS REINFORCED CROSS-LINKED STYRENE COPOLYMER 1/16 THK., 2 OZ. COPPER, SINGLE & DOUBLE SIDED	5.5	SHEET SIZE: 36"x36"=9 SQ. FT., SUGGEST FOR 2.5GHZ APPL
8.0	GENERAL ELECTRIC COMPANY LAMINATED PRODUCTS DEPT. COSHOCOTON, OHIO 43812	TEXTOLITE 11711	2.54 @ 140°F 2.55 @ 72°F 2.55 @ -21°F	.0004 @ 140°F .0005 @ 72°F .0010 @ -21°F	±.004 STD TOL. *±.002 IMPROVED TOL.	POLYPHENYLENE OXIDE (PPO) 1/16 THK., 1 OZ. COPPER, ONE SIDE	7.1	QTY: 100,000 SQ. FT., DELIV; OVER 5 YEARS PRIOR NOTIFICA- TION: 18 MO. INTERNAL PRICE
8.1	"	"	"	"	"	1/16 THK., 2 OZ. COPPER, ONE SIDE	7.3 *10.1	"
8.2	"	"	"	"	"	1/16 THK., 1 OZ. COPPER, TWO SIDES	7.3	"
8.3	"	"	"	"	"	1/16 THK., 2 OZ. COPPER, TWO SIDES	7.6 *10.3	"
8.4	"	"	"	"	"	1/32 THK., 1 OZ. COPPER, ONE SIDE	6.7	"
8.5	"	"	"	"	"	1/32 THK., 2 OZ. COPPER, ONE SIDE	7.0 *9.5	"
8.6	"	"	"	"	"	1/32 THK., 1 OZ. COPPER, TWO SIDES	7.0	"

TABLE VI-1 (Continued)

SHEET 3 OF 4
CONTRACT: NAS 3-11520

COPPER CLAD LAMINATES FOR MICROWAVE STRIPLINE
PRODUCT SURVEY

ITEM	MANUFACTURER	TRADE NAME	DIELECTRIC CONSTANT & VARIATION W/T & FREQ.	DISSIPATION FACTOR & VARIATION W/T & FREQ.	SHEET THICKNESS TOL.	MATERIAL DESCRIPTION	COST \$ PER IN ²	REMARKS
9.7	GENERAL ELECTRIC COMPANY LAMINATED PRODUCTS DEPT. COSHOCKTON, OHIO 43812	TEXTOLITE 11711	2.54 @ 140°F 2.55 @ 72°F 2.55 @ -21°F	.0004 @ 140°F .0005 @ 72°F .0010 @ -21°F	±.004 STD TOL. *±.002 IMPROVED TOL.	1/32 THK., 2 OZ. COPPER, TWO SIDES	7.4 *9.7	"
8.8	"	"	"	"	"	1/16 THK., 1 OZ. COPPER, ONE SIDE & .0025 THK., ALUM. OTHER SIDE	7.3	"
9.0	GENERAL ELECTRIC COMPANY LAMINATED PRODUCTS DEPT. COSHOCKTON, OHIO 43812	TEXTOLITE 11571 & 11666	4.1 @ 1MC AS RECEIVED. 4.3 @ 1MC AFTER 24 HR. IN WATER @ 23°C	.032 @ 1MC AS RECEIVED .034 @ 1MC AFTER 24 HR. IN WATER @ 23°C	±.006 STD. TOL. ±.005 CLASS II ±.003 CLASS III	PHENOLIC PAPER, 1/16 THK., 2 OZ. COPPER, ONE SIDE. (#11571BK2)	.56	QTY: 10,000 SQ. FT., INTERNAL PRICE USED IN T.V. MFG.
10.0	GENERAL ELECTRIC COMPANY LAMINATED PRODUCTS DEPT. COSHOCKTON, OHIO 43812	TEXTOLITE 11558 (NEMAGRADE G-10)	4.5 @ 1MC AS RECEIVED 4.6 @ 1MC AFTER 24 HR. IN WATER @ 23°C.	.021 @ 1MC AS RECEIVED .022 @ 1MC AFTER 24 HR. IN WATER @ 23°C.	±.0075 STD TOL. ±.005 CLASS II ±.003 CLASS III	EPOXY GLASS CLOTH 1/16 THK., 2 OZ. COPPER, TWO SIDES, (#11558BL2)	1.44	QTY: 10,000 SQ. FT. INTERNAL PRICE
10.1	"	"	"	"	"	1/16 THK., 2 OZ. COPPER, ONE SIDE (#11558BK2)	1.26	"

TABLE VI-1 (Continued)

SHEET 4 OF 4

ULTRA THIN COPPER CLADS
PRODUCT SURVEY

ITEM	MANUFACTURER	TRADE NAME	DIELECTRIC COSTANT & VARIATION W/T & FREQ.	DISSIPATION FACTOR & VARIATION W/T & FREQ.	SHEET THICKNESS TOL.	MATERIAL DESCRIPTION	COST \$ PER IN.	REMARKS
11.0	THIokol CHEMICAL CO. PANELYTE INDUSTRIAL DIV. TRENTON, NEW JERSEY	PANELYTE 161-EL	D-24/23 AT ONE MC, 1/32" THK. LAMINATE; 4.8 AVERAGE	D-24/23 AT ONE MC, 1/32" THK. LAMINATE; .025 AVERAGE	DIELECTRIC: .0025"±.0007 COPPER: .0028"±.0003"	EPOXY-GLASS LAMINATE, 1 GIEL G-10 .0025" THK. BASE WITH 2 OZ. COPPER, TWO SIDES	1.0	QTY: 10,000 SQ. FT.
12.0	CIRCUIT MATERIALS CO. DIV. OF DODGE INDUSTRIES PRINCETON, NEW JERSEY	CMC 200	3.25	.0021 AT ONE MC	.006" OVERALL NO TOL. AVAILABLE	CMC 200 COPPER-MYLAR LAMINATE POLYESTER FILM .002" THK. BASE WITH 2 OZ. COPPER, TWO SIDES	1.25	QTY: 7000 SQ. FT.
13.0	RIEDEL PAPER CO. INDUSTRIAL PRODUCTS DIV. MATTHEWS, NORTH CAROLINA	MONOTHERM, MKC-232	4.0 AT ONE MC	.026 AT ONE MC	DIELECTRIC: .003"±10% COPPER: .0028"±.00075 MEASURED THICKNESS: .0093"±.0001" OVERALL	POLYIMIDE FILM, 3 MIL. "KAPTON"; 2 OZ. COPPER, TWO SIDES	1.8	QTY: 10,000 SQ. FT.

D. MICROWAVE CIRCUIT FABRICATION

Three distinctive fabrication techniques exist for the microwave circuits required in the converters; stripline, microstripline, and waveguide.

Microstriplines advantages are: its production is amenable to photo mask and etch processes, high dielectric constant materials are available for substantial size reduction in the actual circuit. A main disadvantage is the susceptibility to surrounding structure and the shielding must be spaced a relatively large distance from the substrate or circuit to prevent field distortions and performance changes.

Stripline has the same basic fabrication advantages, however, its symmetrical ground plane structure makes assembly more difficult and the available dielectric materials that are compatible with practical assembly techniques do not have as high dielectric constants. The planar geometry of the stripline circuit will therefore be greater than the microstrip equivalent. The total packaging volume will likely be smaller with stripline because it is less susceptible to adjacent structures as a result of the double ground plane symmetry.

Waveguide circuits rely on air dielectric and are, therefore, the largest. S-Band waveguide is nominally $1.5" \times 3.0"$ in cross-section. Mixer circuits with the fractional wave length stubs and couplers become quite bulky at S-Band. At X-Band, the waveguide cross-section is approximately $0.5" \times 1.0"$. A factor of $(3)^3$ is achieved in volume by X-Band over S-Band as a rough approximation. The size of X-Band circuits in waveguide is practical for this system. Other advantages associated with waveguide are: it is self-shielding, it is self-supporting, it can provide support for the active devices, and it can provide for a direct antenna interface without cable connectors. The main disadvantage of waveguide is the cost of manufacture. There is no consumer oriented production of waveguide structures that has cost reduction as its prime motivation.

An independent factor concerning the choice of waveguide versus microstripline or stripline is the potential application of bulk effect devices (such as impatt and Gunn effect) as local oscillator power sources. These devices can provide power at the desired frequencies directly and they are also voltage tuneable over reasonable ranges. One draw back with the bulk effect devices is the need for high Q circuits to "clean-up" the signal. The waveguide structure may be used to provide this filter function.

Stripline and Microstrip Design Trade-offs

The use of stripline or microstripline permits the flexibility of designing for various circuit impedance levels within practical constraints.

Figure VI-4 shows the relationship between linewidth and characteristic impedance for stripline with a dielectric constant ~ 2.5 . Curves are shown for different values of the groundplane spacing.

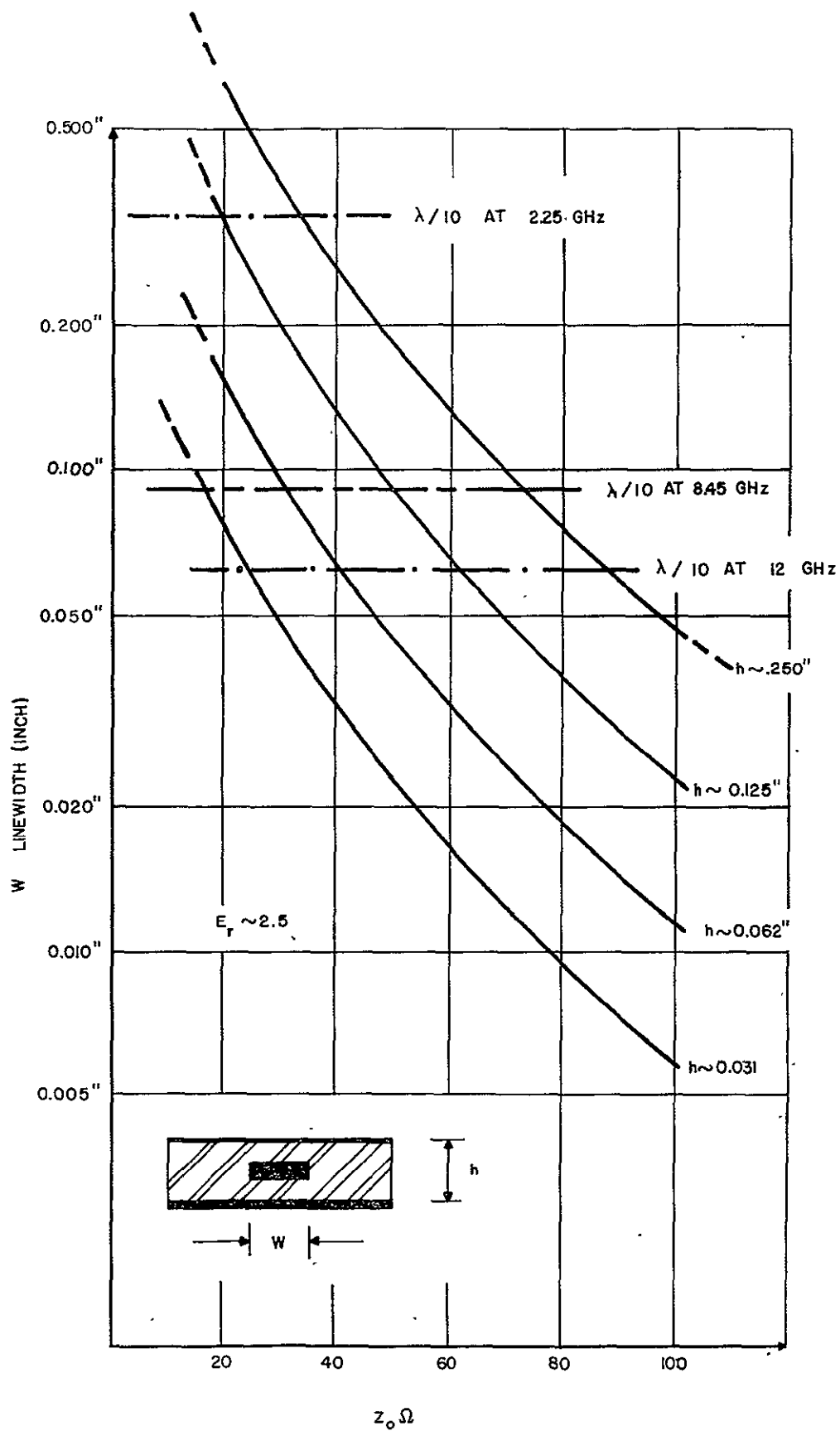


Figure VI-4. Stripline Linewidth vs. Impedance and Dielectric Thickness

Limitations on Linewidth

When the linewidth becomes comparable to the wavelength, higher order modes can propagate, and the transmission line is no longer useful. In the present case, the limit is chosen arbitrarily as one tenth of a wavelength. The minimum linewidth is dependent upon what tolerances are considered technically and economically feasible. The curves aid the selection of the optimum ground plane spacing for a given stripline component at a given frequency of operation.

The stripline configurations use a line width of about 40 mils on 1/32" (0.031") thick PPO board material to achieve a 50 ohm line impedance. A comparable microstripline fabricated on 20 mil thick alumina would require a line width of about 20 mils. Assuming a comparable etch accuracy, the PPO material is about one-half as sensitive to etch errors. A comparable relationship exists with respect to linear line lengths. The ratio of line lengths in PPO and alumina is equal to the square root of the inverse ratio of dielectric constants.

$$\frac{L_{\text{PPO}}}{L_{\text{Alumina}}} = \sqrt{\frac{\epsilon_r \text{ Alumina}}{\epsilon_r \text{ PPO}}} = \sqrt{\frac{9}{2.55}} = 1.88$$

The effects of etch tolerances and board dielectric thickness on line impedance for 1/16" and 1/32" PPO stripline circuits (Z = 50 ohm) are shown in Table VI-0. The impedance versus line width is shown in Figure VI-5, for two values of PPO board of thicknesses near 1/16".

Figure VI-6 shows a similar relationship for microstripline on ceramic ($\epsilon = 9$).

E. COMPONENT PART SURVEY

The following tables of device performance parameters and prices were compiled at the start of Phase I to provide a basis for component part selection and cost screening. The cost data included represents low volume procurement levels as of the fall of 1968.

TABLE VI-0

EFFECTS OF ETCH TOLERANCES AND BOARD DIELECTRIC THICKNESSES

Board Thickness (50 Ω line)	W (Mils)	$\frac{\Delta z}{\Delta W}$ (Ω /mil)	$\frac{\Delta z}{\Delta H}$ (Ω /mil)
1/16"	89	0.35	1.8
1/32"	38.5	0.87	9.1

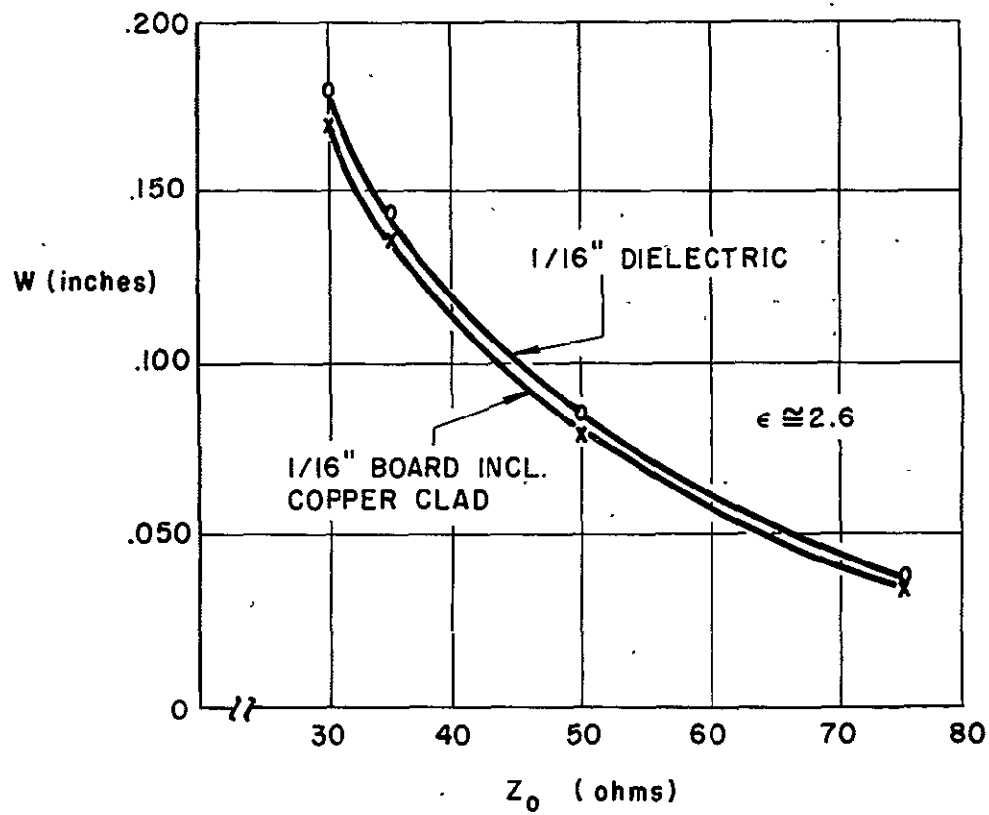


Figure VI-5. Stripline Linewidth versus Characteristic Impedance

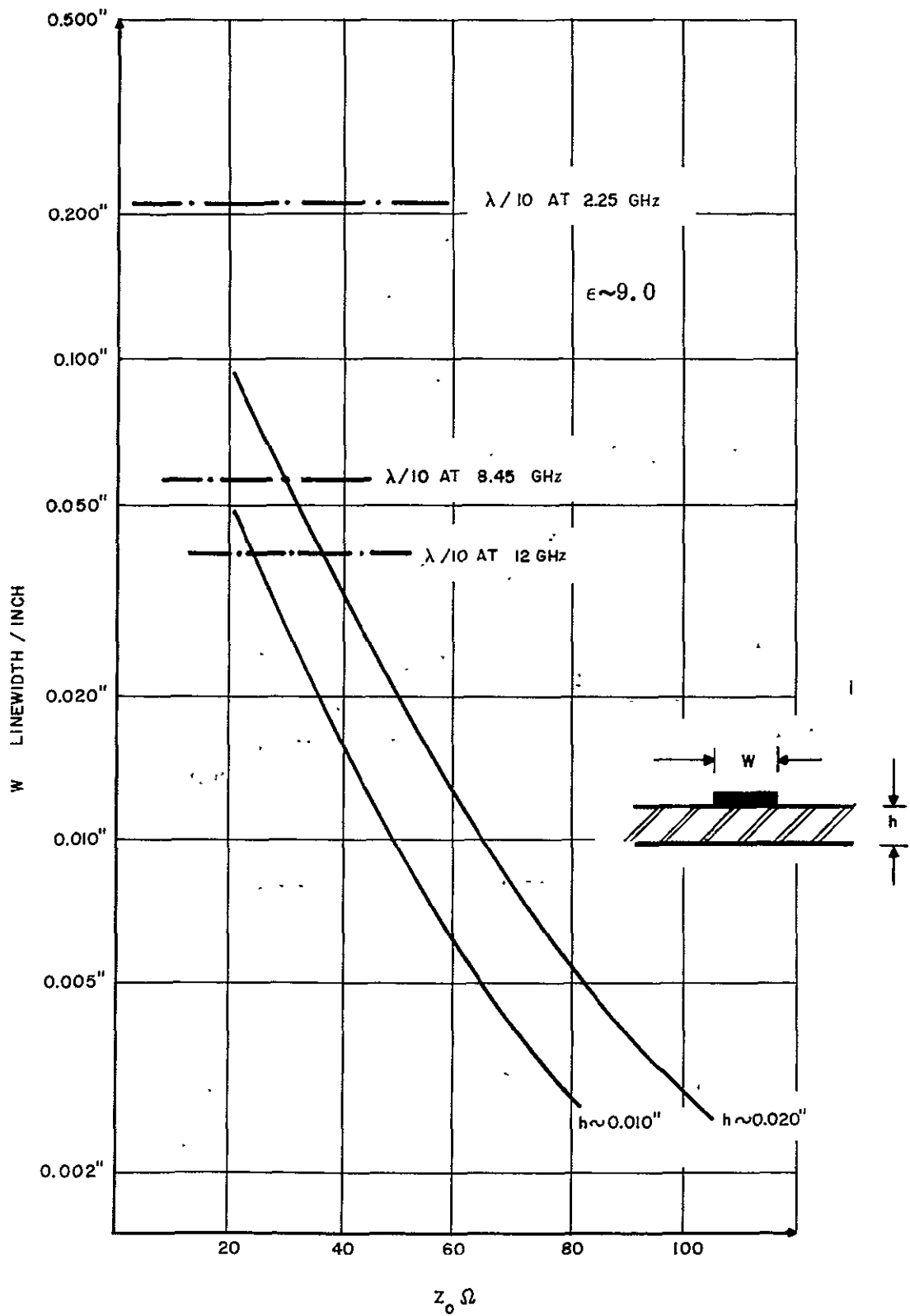


Figure VI-6. Microstripline Linewidth vs. Impedance and Dielectric Thickness

TABLES OF RELEVANT DEVICES

<u>FUNCTION</u>	<u>DEVICE CATEGORY</u>	<u>TABLE NO.</u>
<u>Preamplifier</u>		
2.25 GHz	Transistors, tunnel Diodes	I + II
8.45	Tunnel Diodes	III
12.00	Tunnel Diodes	IV
<u>Mixer</u>		
2.25 GHz	Schottky Diodes	V
8.45	Schottky Diodes	VI
12.0	Schottky Diodes	VII
<u>Local Oscillator</u>		
2.25 GHz	Transistor, Diodes	VIII, IX
8.45	Transistor, Diodes	VIII, IX
12.	Transistor, Diodes	VIII, IX
<u>Video Amp.</u>		
	IC's	
	Transistors	X
<u>Modulators</u>		
	IC's, Transistors	X
<u>Miscellaneous</u>		
	Transistors, Diodes	X

TABLE I

Preamplifier 2.25 GHzTransistors

<u>Manufacturer</u>	<u>Type No.</u>	<u>f_t GHz</u>	<u>NF at f_t GHz</u>	<u>Expected Gain</u>	<u>Price</u>	<u>Delivery</u>
Texas Instruments	TIXM 103	1.8	4.5dB at 1.5	~6dB	\$50	30 days
	TIXM 105	2.2	4.5 at 2.25	~6dB		45 days
	TIXM 106	2.2	4.5 at 2.25	~6dB	\$82.50	
KMC	KD5201		4.5-6dB - 2GHz			
	KD5525	2	2.5 at 1GHz	8dB	\$90	1 week
	KD5526	2	4dB at 1GHz	14dB	\$50	2 days
Nippon Electric Co.	V415	3.8	4.8dB at 2GHz	5.4dB	?	Stock
	V417	4.5	Not available	6.5dB	?	Stock

TABLE II

Preamplifier: 2.25 GHzTunnel Diodes

Manufacturer	Material	Type No.	Rest. Cutoff Frequency	C_j	R_s	Exp. Min. Device Noise Figure	Max. Gain BW Product	Price		
								1-9	10-99	100-up
General Electric	Ge	TD 401	5 GHz	2pF	3 Ω	5 dB	1.150 GHz	\$45	\$30	\$26
	Ge	TD 402	10 GHz	1pF	4 Ω	4.3 dB	2.30 GHz	\$56.50	\$37.50	\$25
	GaSb	MS1504A	6 GHz	1.5pF	4 Ω	3.7 dB	1.55 GHz	\$125	\$93-83	\$75
	GaSb	MS1507A	12 GHz	.6pF	6 Ω	3.4 dB	3.8 GHz	\$225	\$168 150	\$135
Microstate	Ge	MS1005A	6 GHz	1pF	3 Ω	4.8 dB	2.3 GHz	\$36	\$24- 21	\$18
	Ge	MS1007A	10 GHz	.6pF	4 Ω	4.3 dB	3.8 GHz	\$48	\$32- 28	\$24

TABLE III

Preamplifier: 8.45 GHzTunnel Diodes

Manufacturer	Material	Type. No.	Rest. Cutoff Frequency	C_j	R_s	Exp. Min. Device NF	Max. Gain BW Product	Price			
								1-9	10-24	25-99	100-up
General Electric	Ge	TD 405	25 GHz	.36pF	6 Ω	4.7 dB	6.4 GHz	\$127	\$85	\$85	\$57
	Ge	TD 407	40 GHz	.18pF	6 Ω	4.3 dB	12.8 GHz	\$270	\$180	\$180	\$160
Microstate	Ge	MS1022A	25 GHz	.4 pF	5 Ω	4.7 dB	5.75 GHz	\$141	\$94	\$84	\$70
	Ge	MS1208A	40 GHz	.25pF	7 Ω	4.6 dB	9.2 GHz	\$420	\$281	\$252	\$210

TABLE IV

Preamplifier: 12 GHzTunnel Diodes

Manufacturer	Material	Type No.	Rest. Cutoff Frequency	C_j	R_s	Exp. Min. Device NF	Max. Gain BW Product	Price			
								1-9	10-24	25-99	100-up
General Electric	Ge	TD 405	25 GHz	.36pF	6Ω	5.3 dB	6.4 GHz	\$127	\$85	\$85	\$57
	Ge	TD 407	40 GHz	.18pF	6Ω	4.6 dB	12.8 GHz	\$270	\$180	\$180	\$160
Microstate	Ge	MS1022A	25 GHz	.4pF	5Ω	5.4 dB	5.75 GHz	\$141	\$94	\$84	\$70
	Ge	MS1208A	40 GHz	.25pF	7Ω	4.9 dB	9.2 GHz	\$420	\$281	\$252	\$210

MIXER DEVICES

TABLE V

Schottky Diodes: 2.25 GHz

Manufacturer	Approximate Conversion Loss Claimed by Manufacturer				
	4 dB	4.5 dB	5 dB	5.5 dB	6 dB and up
Microstate		MS1651 x	MS1650 x		
Philco			L8200		
MA	MA4853		MA4852		MA4851 MA4A401 MA4A402
Sylvania	D5503C	D5503B			
H. P. A.		2401	2418	2351	2900
T. I.					

TABLE VI

8.45 GHz

Manufacturer	Approximate Conversion Loss Claimed by Manufacturer			
	4.5 dB	5 dB	5.5 dB	6 dB
MA	MA4859B	MA4858	MA4857	MA4856
Sylvania		D5506C	D5506B	
H. P.		2606	2607	2608
T. I.	TIxV12	TIxV13	TIxV14	TIxV15
Philco-Ford				L8202
Microstate	MS1651 x	MS1650 x		

TABLE VII

Schottky Diodes: 12 GHz

Manufacturer	Approximate Conversion Loss Claimed by Manufacturer				
	5 dB	5.5 dB	6 dB	6.5 dB	7 dB and up
Microstate	MS1651 x	MS1650 x			
MA	MA4861H	MA4861 G	MA4861F	MA4861E	
Sylvania		D5507-C	D5507 B	D5507 A	D5507
HPA					2740

TABLE VIII

<u>Components Designations</u>	<u>Application</u>	<u>Performance Guide</u>	<u>Unit Cost</u>	<u>Remarks</u>
<u>Transistors:</u>				
2N3866 (RCA)	Oscillators	600 MHz, Power Out 1 Watt	\$2.00	TO-5
SE5020 (Fairchild)	IF Amp.	200 MHz, 20 dB gain 3.3 dB NF		
2N5397 (Siliconix)	IF Amp.; (1st stage)	450 MHz, 18 dB gain	\$5.50	N-channel
	Modulator	2.5 dB NF		Silicon FET
MEM554 (Gen. Inst.)	IF Amp.; Modulator	200 MHz, 18 dB 3.0 dB N. F.	\$1.98	MTOS
3N140 (RCA)	IF Amp.	200 MHz: Gain 19 dB NF 5 dB	\$1.62	MOSFET Dual Gate
MFE 3006 (Motor.)	IF Amp.	100 MHz: Gain 20 dB NF 4.5 dB	\$1.50	MOSFET
2N3823 (TI)	IF Amp.	100 MHz: Gain 11 dB NF 2.5	\$4.00	JFET
K2104-07 (KMC)	IF Amp.	400 MHz: Gain 14 dB NF 2.5 dB	\$25.00	BIPOLAR
2N3571/72 (TI)	IF Amp.	450 MHz: NF 4-6 dB	\$7.00	BIPOLAR
2N4254 (TI)	IF Amp. Oscillator amp., Limiter	120 MHz: Gain 15 dB NF 2-3 dB	\$1.00	BIPOLAR
2N5135 (FC)	dc Regulator	dc High Gain	\$0.13	NPN
2N4957 (Motor.)	IF Amp., L. O. (S band)	450 MHz, NF 2.6 dB Gain 35 dB, 120 MHz	\$13.00	PNP Silicon
MT1961 (FC)	S Band Osc., IF Amp.	5 GHz Max. Osc., 450 MHz: Gain 12 dB, NF 2.3-3.5 dB		

TABLE IX

<u>Component Designations</u>	<u>Applications</u>	<u>Performance Guide</u>	<u>Unit Cost</u>	<u>Remarks</u>
Diodes:				
5082-0114 (HPA)	Multiplier to 2.13 GHz	T min. 125 ns. t_t 400 ps., 10 pf.	\$2.00	SRD
5082-0153 (HPA) -0253	Multiplier to 12 GHz	T min. 20 ns. t_t 150 ps., 1.1 pf.	Not Available	SRD
5082-2302 (HPA)	Detection	$C_t(0)$ 1 pf., 1 mA., 0.4V.	\$5.00	Schottky
5082-2800 (HPA)	Video Detection	$C_t(0)$ 2 pf., 15 mA., $V_{F2} = 1$ volt 1 mA., $V_{F1} = 0.4V$.	\$<1.00	Schottky

TABLE X

<u>Component Designations</u>	<u>Applications</u>	<u>Performance Guide</u>	<u>Unit Cost</u>	<u>Remarks</u>
IC:				
3028A (RCA)	Limiter, Discriminator	DC to 120 MHz	\$1.47	Diff. Amp. or Cascade
PA-7600 (Philco)	IF Amp. or Video	43 dB gain at (65 MHz) NF 8 dB BW	\$11.80	
PA-7713 (Philco)	IF Amp.	200 MHz, 12 dB Gain	\$9.10	
MC-1110 (Motorola)	IF Amp.	26 dB at 100 MHz 4.0 dB NF	\$37.50	
MC-15500 (Motorola)	IF Amp.	Gain Bandwidth 2 GHz NF 5 dB at 60 MHz	\$4.25	
NS LM 171/LM 271 (National S. C.)	Limiter; Modulator	100 MHz; PG 24.6 dB (CC-CB) 27.5 (CE-CB)	\$4.50	

APPENDIX VII

GLOSSARY OF SEMICONDUCTOR TECHNOLOGY TERMS

Active Device. A device displaying gain and control.

Beam-lead Bonding. A free-bonding technique in which thick, gold extensions of the thin-film terminals of the semiconductor devices and circuits are electro-formed so they extend beyond the edges of the chips. The resulting strips may be fed into automatic bonding machines much like a roll of postage stamps. The devices would be clipped apart as they were bonded unless, for example, the circuit called for transistors with common electrode connections.

Chip. A small rectangular, square, or round die containing all components and interconnections of a circuit.

Circulator. A nonreciprocal wave-guide element, i.e., an element which propagates a wave in one direction with negligible attenuation but which will almost completely absorb the reflected wave which is simultaneously traveling in the opposite direction.

Field-Effect Resistor. A resistor formed by the channel region of a junction field-effect structure. Such resistors have a nonlinear V-I characteristic.

Field-Effect Transistor. Simply a bar of semiconductor material whose resistance is modulated by varying the effective cross-sectional area of the bar by electrical means.

Flat Pack. A flat, rectangular-shaped, integrated-circuit package with coplanar leads.

Flip Chip. A face-bonding technique in which the thin-film terminals of an integrated circuit are bonded directly to the thin films of the substrate.

Gunn Diode. A bulk effect semiconductor device that exhibits a negative resistance effect when biased. The device is capable of direct conversion of DC power to microwave signals.

Hybrid Circuit. A circuit which combines the thin-film and the active components by semiconductor techniques. The active devices are attached to the thin-film passive component by a suitable bonding process.

IC. Abbreviation for integrated circuit.

Idler Frequency. In a parametric amplifier, the difference between the pump frequency and the signal frequency.

Integrated Circuit. A circuit which combines the previously separate disciplines of materials, devices, and circuits into a single discipline.

Microelectronics. The entire body of the electronic art which is connected with, or applied to the realization of electric circuits, subsystems, or the entire system for extremely small electronic parts. (From E. Keonjian (ed.), Microelectronics, McGraw-Hill Book Company, New York, 1963).

Microstripline. An unbalanced structure consisting of a transmission line separated from a single groundplane by a dielectric material. A relatively high dielectric constant material is normally used $\epsilon_r > 8$ to confine the field patterns to the volume between the transmission lines and the groundplane.

Monolithic Circuit. A circuit consisting of a semiconductor substrate into which all of the parts are fabricated by diffusion and epitaxial processes.

MOS. Abbreviation for metal-oxide semiconductor.

Parametric Amplifier. A class of amplifying and frequency converting devices which utilizes the properties of nonlinear (time-varying) reactances.

Passive Substrate. A substrate which may serve as physical support and heat sink to an integrated circuit, but which exhibits no characteristics of an active device.

Pump. A signal source in a parametric amplifier.

Semiconductor Integrated Circuit. See monolithic circuit.

Stripline. A balanced structure using a centered transmission line with dielectric between the center conductor and a ground plane on each side.

Substrate. The physical material upon which an integrated circuit is fabricated. Used primarily for mechanical support, it may also serve a useful thermal or electrical function.

Thin-Film. A coating up to a thickness of about $10^4 \text{ \AA}$.

Thin-Film Circuit. A circuit which consists of a passive substrate on which parts are deposited in the form of pattern films of conductive or non-conductive materials.

Varactor. One of several trade names (Microwave Associates, Inc.) used to indicate an especially processed single-crystal silicon diode which exploits the phenomenon of voltage-sensitive capacitance.

