VLA TECHNICAL REPORT NO. 41

MODULE T3

IF-TO-BASEBAND CONVERTER

J. W. Archer

January 1980

TABLE OF CONTENTS

1.0 FUNCTION

The IF-to-baseband converter (T3) module is located in D-Rack in the Central Electronics Room. It forms an integral part of the baseband subsystem for the Very Large Array spectral line receiver. It is designed to interface with the IF combiner (T2) module, which provides the input IF signals to the T3 module. The four 50-MHz bandwidth IF signals, in the frequency range 1.3 to 1.7 GHz are transmitted from an antenna to the Central Electronics Room, modulated onto a millimetre wavelength carrier, via the circular waveguide communication system. There they are demodulated in the modem (T1) module and fed to the T2 module for distribution to other locations for further processing.

The T3 module converts the composite L-band IF signal to four baseband outputs which correspond to a selected segment of each IF passband. These outputs are fed to four separate baseband filter modules (T4 module). Fixed frequency 1200 MHz and 1800 MHz local oscillator signals are supplied to the T3 module from the central LO filter (L14). Variable frequency UHF local oscillator signals are supplied via a distribution network from the Fluke synthesizers through the variable frequency driver module (L18).

2.0 THEORY OF OPERATION

A block diagram of the IF-to-baseband converter (T3) module is shown in Figure 2-1. The composite IF signal at input J21 possesses the spectral distribution shown in the insert of Figure 2-1. The input signal is first amplified by amplifier AR1, then divided into two signals of equal amplitude by the 3 dB power divider PD1. Each component is attenuated, for impedance matching purposes, by attenuators AT1 and AT2, before being fed to mixers M1, M2. The net gain from the amplifier input to each mixer is nominally -0.5 dB, with a typical value for the amplifier gain being 8.5 dB and attenuator -6 dB.

The IF signals are converted by the mixers to the UHF range by mixing with either an 1800 MHz or 1200 MHz fixed frequency local oscillator signal. The result is that, at the output of mixer M1, the A channel IF passband (1.3 to 1.35 GHz) now corresponds to a range of frequencies 100 to 150 MHz and the B channel IF passband (1.4 to 1.45 GHz) corresponds to 200 to 250 MHz. Similarly, at the output of mixer M2, the C channel IF passband (1.55 to 1.60 GHz) corresponds to the inverted frequency range 250 to 200 MHz and the D channel IF passband (1.65 to 1.70 GHz) corresponds to 150 to 100 MHz. The two mixer outputs are further split by a pair of two-way power divider PD2, PD3 to provide four converted IF output signals. All of the above functions are performed in the L-band converter submodule.

It is desirable to be able to select any portion of each IF passband for conversion to baseband and subsequent spectral analysis at a bandwidth selected by commands to the T4 module. For this reason, four variable frequency LO signals are provided to the T3 module from the master LO system, with frequencies selectable in either the 100-150 MHz or 200-250 MHz range. So that only the desired sideband of the UHF LO signal is selected, the IF-to-baseband mixer must exhibit substantial rejection of the unwanted image band. For image signals, within the IF band, falling in the desired baseband bandwidth relative to the frequency of the tunable LO, the

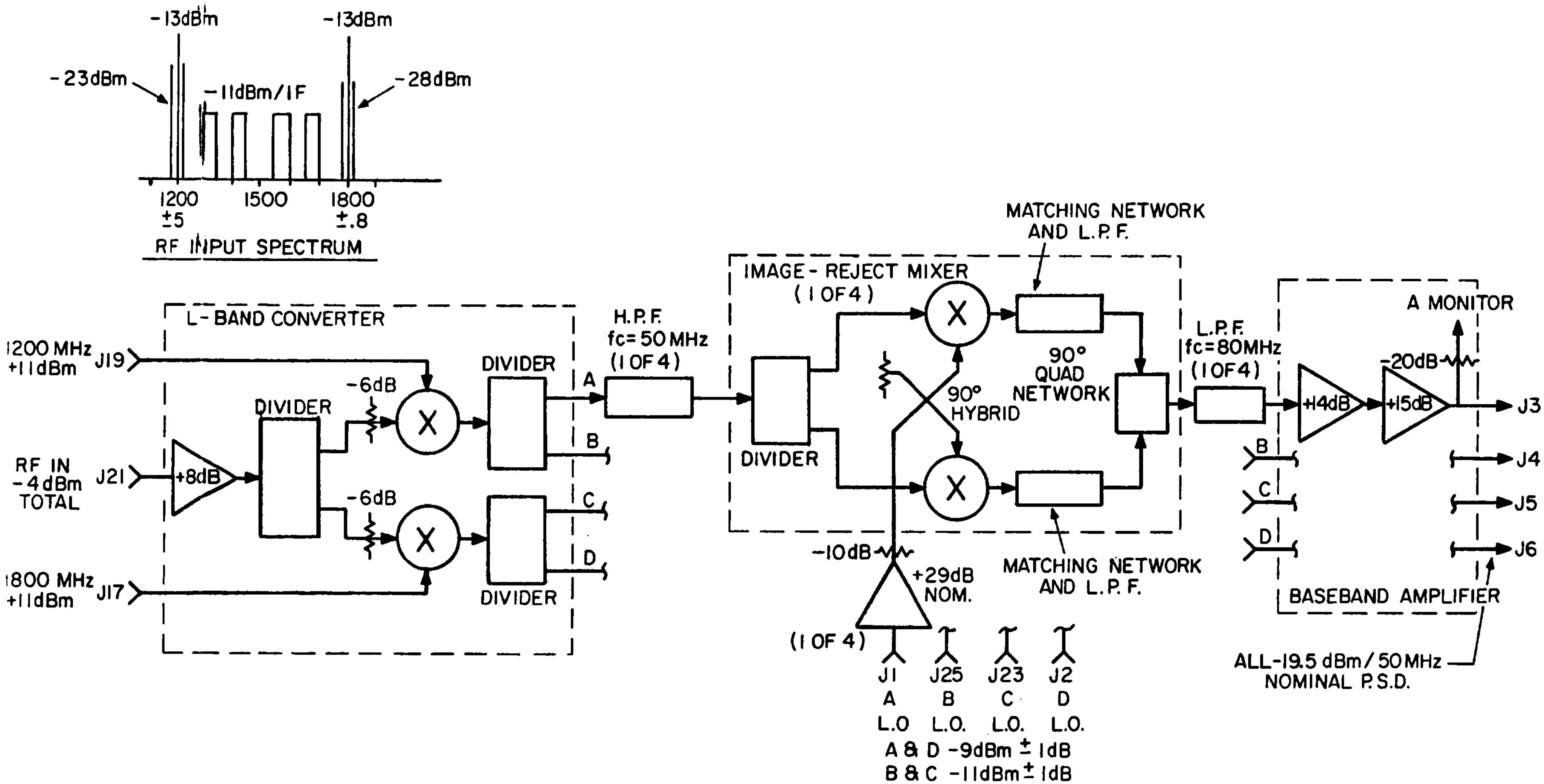


FIGURE 2-1: BLOCK DIAGRAM OF THE IF-TO-BASEBAND CONVERTER (T3) MODULE

rejection must exceed 30 dB to result in less than 1% error in the synthesized spectral distribution produced by the digital spectral processor.

Mixers IRM1, IRM2, IRM3, IRM4 exhibit the desired IF-to-baseband conversion characteristics, IRM1 and IRM2 being lower sideband reject mixers and IRM3 and IRM4 being upper sideband reject mixers in order to maintain the correct sense in frequency relative to the composite input IF spectrum. Typical mixer conversion loss is -27.5 dB. The mixers are preceded by five-section high-pass filters with a 3 dB cutoff frequency of 50 MHz. These filters aid in reducing the level of low-frequency spurious signals resulting from the first conversion process, which, if fed through to the baseband output of the module, would cause spurious responses in the final synthesized spectrum.

UHF local oscillator signals for the image reject mixers are fed to the module via inputs J1, J2 [-9 dBm (±1 dB)] and J23, J25 [-11 dBm (±1 dB)]. The amplifiers AP1, AP2, AP3, AP4, followed by 10 dB attenuators AT3, AT4, AT5, AT6 provide a nominal input level at each mixer LO port of +10 dBm. The amplifiers exhibit a nominal gain of 30 dB. The attenuators are used to improve the impedance match seen by the mixers at each LO port.

The baseband output from each mixer is fed via a low-pass filter to a broadband, low noise amplifier. The four-section low-pass filters, with a 3 dB cutoff frequency of 80 MHz, aid in reducing the power level of feedthrough UHF local oscillator signals reaching the baseband amplifiers. The possibility of the amplifier stages being overdriven by these out-of-band signals is thereby avoided. The baseband amplifier outputs connect via rear panel jacks J3, J4, J5, J6 to separate T4 modules. The baseband amplifier gain is nominally 29 dB per channel, resulting in a nominal output power spectral density of -19.5 dBm/50 MHz. Front panel baseband output monitor points are also provided, at a nominal level of -40 dBm/50 MHz.

3.0 SUBMODULE DESCRIPTION

3.1 L-band Converter Submodule

The L-band converter submodule functions as a signal frequency converter and power divider, to provide outputs of the correct frequency and power level for the image reject mixers. It comprises an integrated microstripline package utilizing Keene Corporation Diclad 527 circuit board material with ¼" aluminum on one side and 1 oz. copper on the other. The dielectric constant of the material is 2.5±.05 and the substrate thickness used in this application is 0.030"±.004. Figure 3-1 shows the block diagram for the submodule and Figure 3-2 is a photograph of the assembly. The composite L-band IF input signal is first amplified by an Avantek UTO-2023 amplifier. This amplifier must have a high 1 dB compression point for random noise signals, since the total input power is nominally -4 dBm and the required gain to overcome losses in the subsequent power divider and attenuators is 8 dB, resulting in a total output power of +4 dBm. The UTO-2023, with minimum CW 1 dB compression point of +14 dBm, provides adequate performance, allowing for substantial departure of the input power from nominal value. Input return loss is better than -10 dB and the amplifier noise figure is typically 8 dB resulting in a nominal noise figure for the complete submodule of 11 dB.

The power divider, which follows the amplifier, is implemented directly in microstripline. It is a compensated, in-line power divider design, with a center frequency of 1.5 GHz, a nominal 1.2:1 VSWR bandwidth of 900 MHz and a typical output port isolation of greater than 20 dB over this frequency range. Microstripline impedances and dimensional data for the power divider are given in Figure 3-3. The 100 Ω series output termination is implemented using a chip resistor.

In order to provide an improved impedance match between the power divider and the mixer, drop-in, thin film chip attenuators are inserted in the 50 Ω microstripline between the two

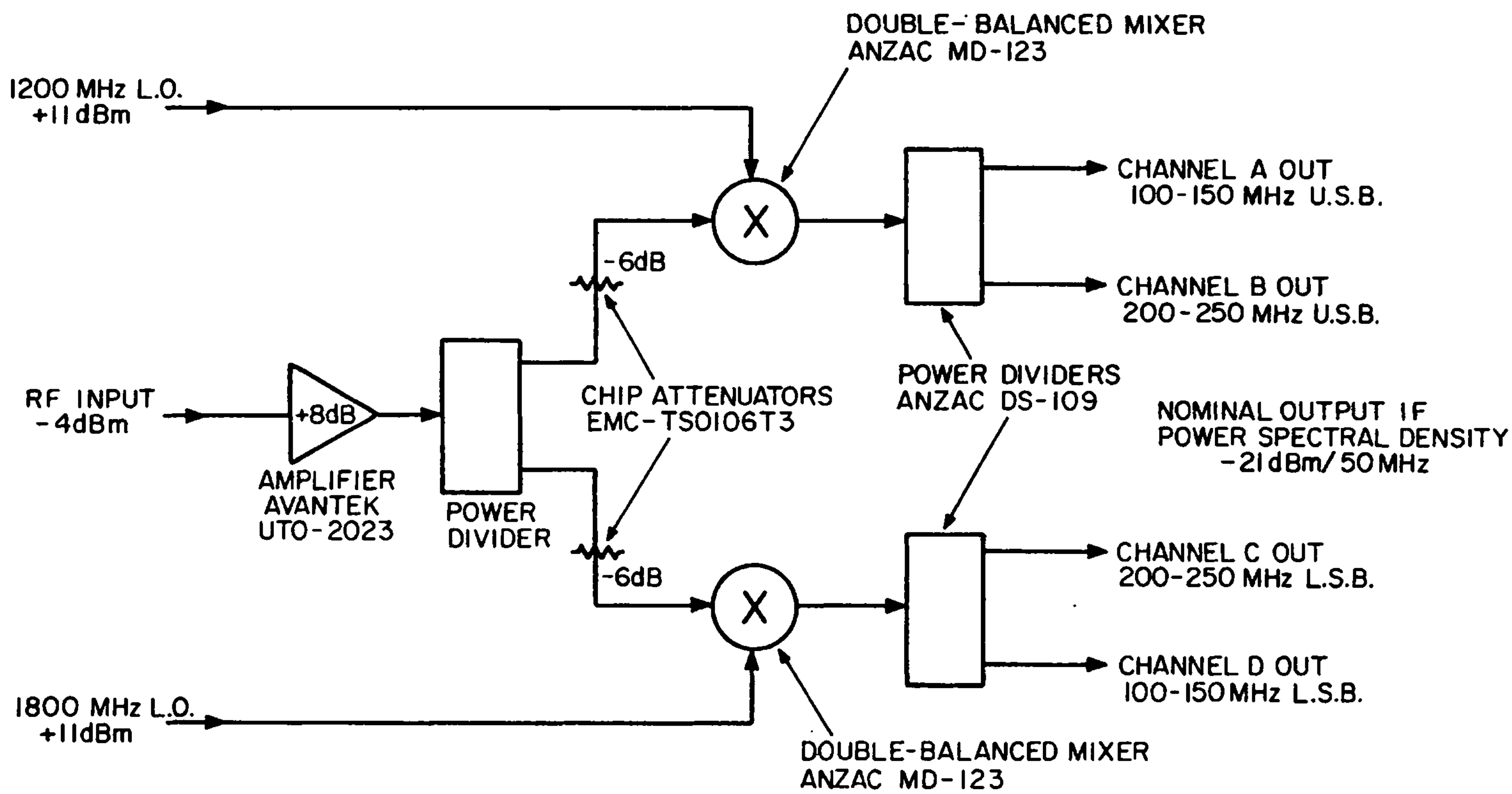


FIGURE 3-1: BLOCK DIAGRAM FOR THE SUBMODULE

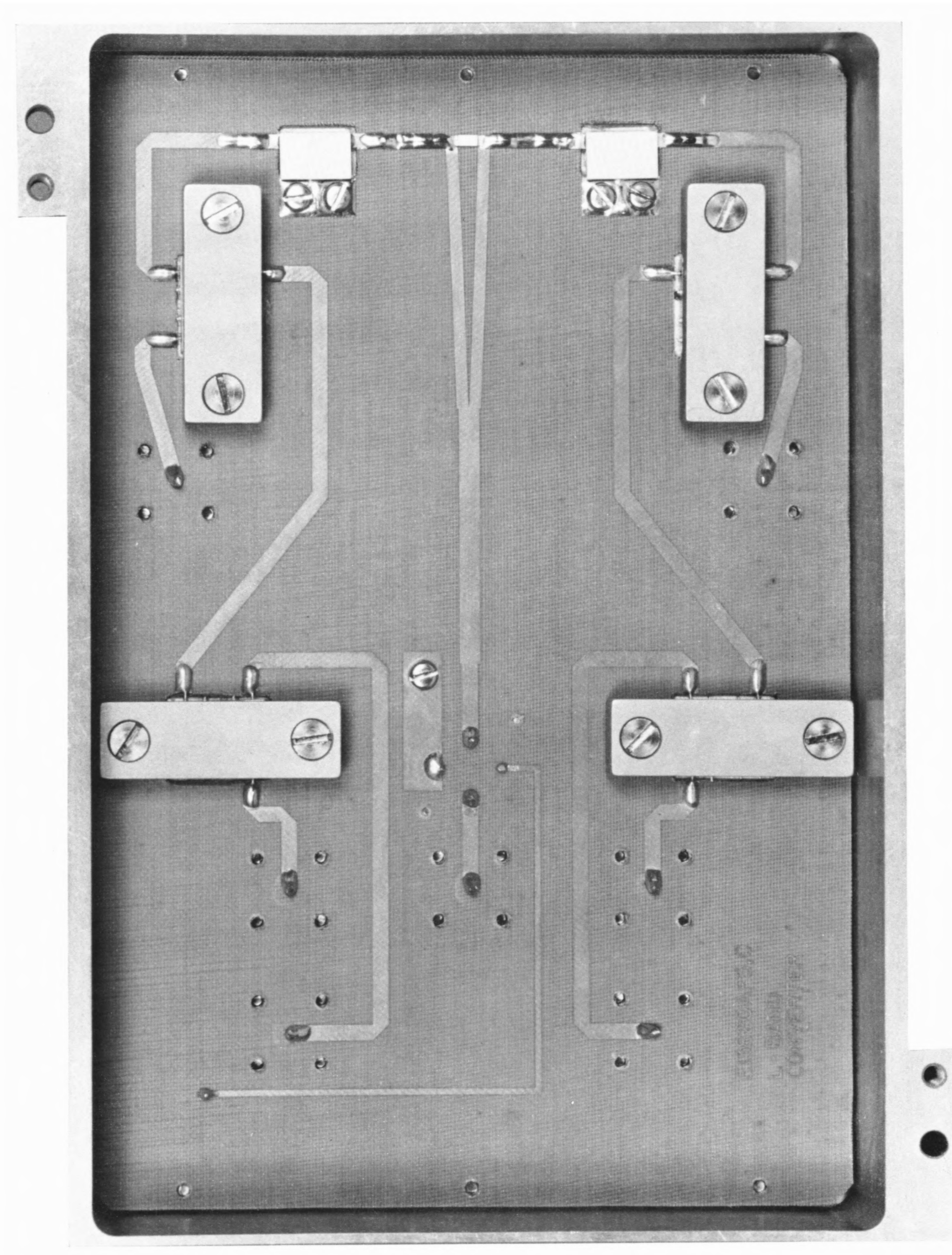


FIGURE 3-2

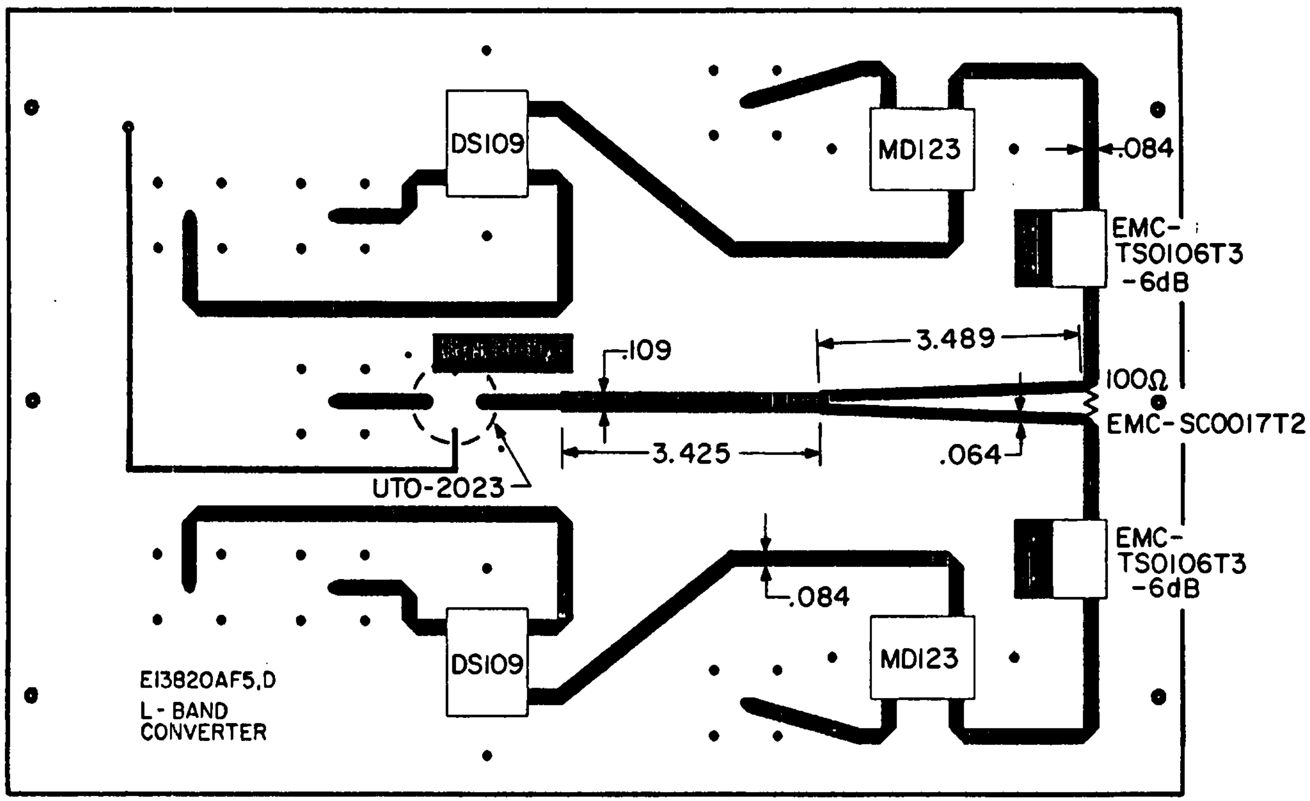


FIGURE 3-3: MICROSTRIPLINE IMPEDANCES AND DIMENSIONAL DATA FOR THE POWER DIVIDER

devices. The mixers are Anzac Type MD123 in a flatpack case. These devices exhibit very flat conversion loss response and low intermodulation distortion over the 1 to 2 GHz frequency band. Typical conversion loss is -6.5 dB over this range at a nominal LO power level of +11 dBm. The IF output from each mixer is divided between two output ports using Anzac DS109 power dividers, also packaged in a flatpack case.

The total insertion loss of the L-band converter submodule from composite IF input to any of the four output ports is nominally -10 dB with a passband variation of less than 0.1 dB/ 50 MHz. The resulting nominal power spectral density at any image reject mixer input is -21 dBm/50 MHz.

3.2 Image Reject Mixers

Figure 3-4 is a photograph of a typical mixer assembly. Only the circuit details for the baseband quadrature network amplifier will be outlined in this Section. For further information concerning overall mixer design principles and operation theory refer to Appendix A.

Figure 3-5 shows a schematic diagram of the baseband quadrature network. The amplifier section is a two-stage design employing a common base transistor connection followed by a common collector stage. The nominal input impedance is 3 Ω over the frequency range dc to greater than 50 MHz and the overall power gain is 10 dB. Gain flatness depends upon the adjustment of the feedback compensation capacitor, C16. With the aid of this adjustment, the amplitude response of the complete image reject mixer can be made flat to within ±.15 dB over the baseband frequency range of 186 kHz to 50 MHz.

3.3 Local Oscillator Power Amplifiers

These devices are designed to provide 29 dB (A & D) or 31 dB (B & C) gain (±.2 dB) over the frequency range 100-250 MHz, delivering an output power of +20 dBm (±.1 dB) to the LO

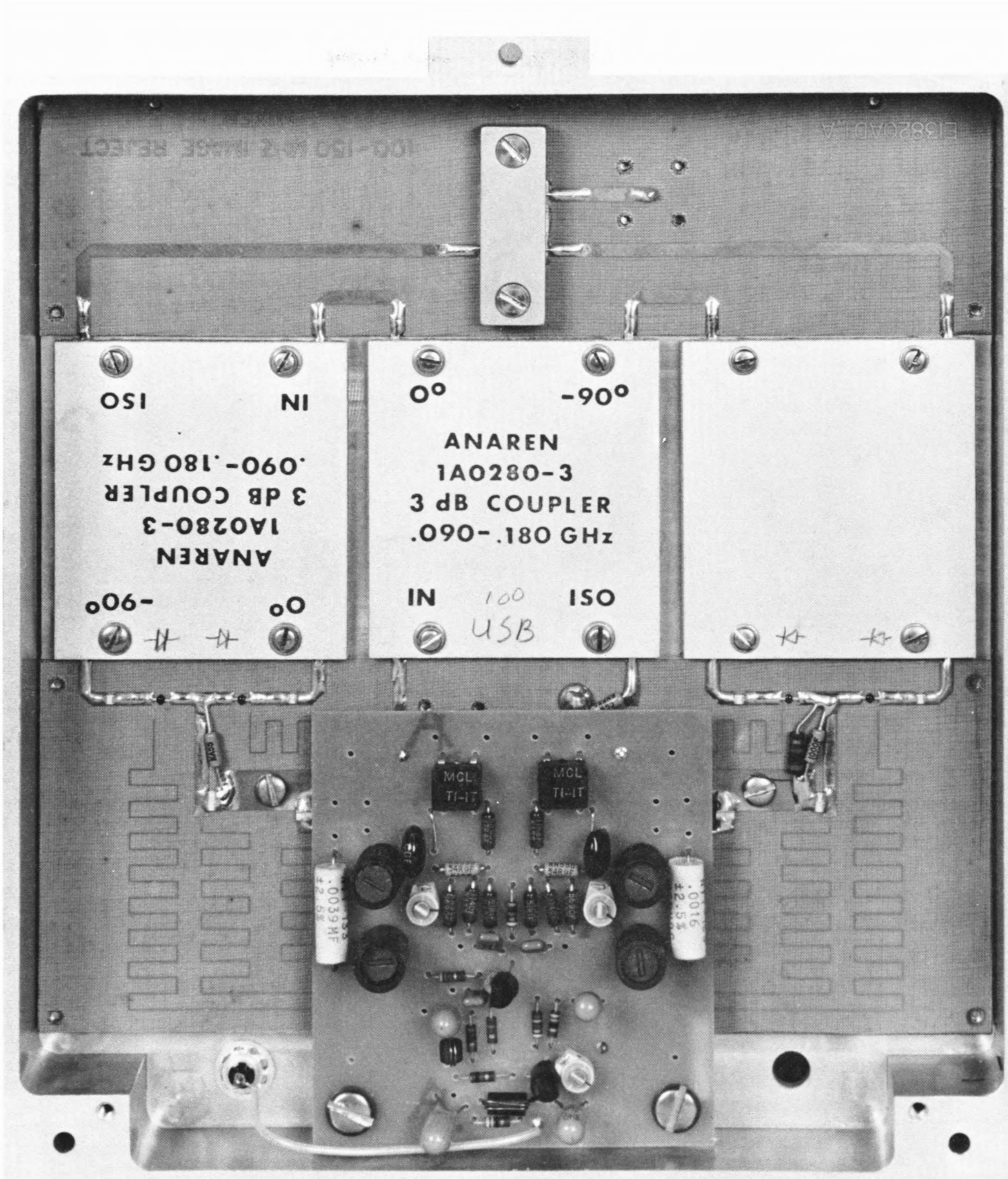
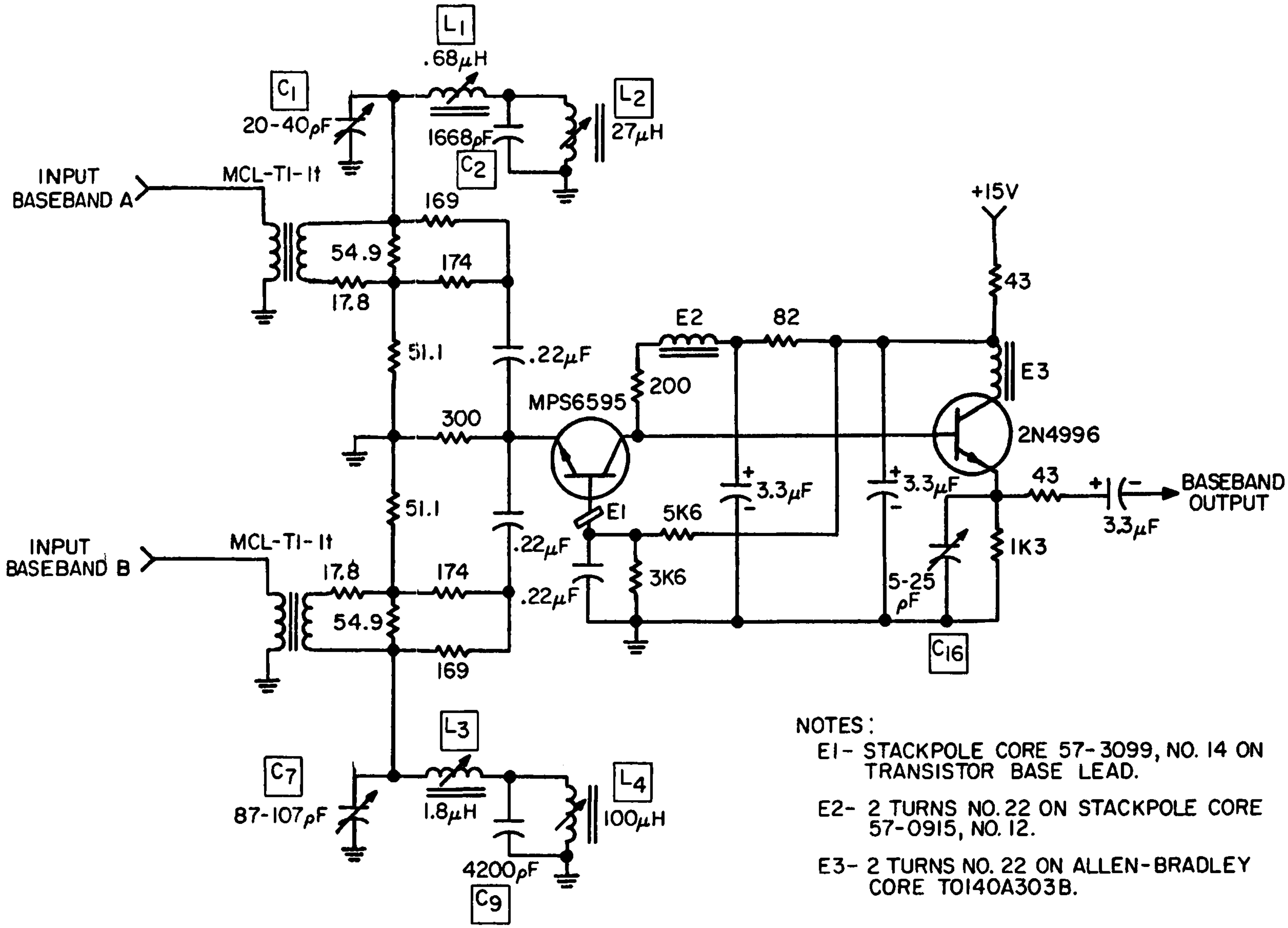

ANAREN
1A0280-3
3 dB COUPLER
.090-.180 GHz
0°
-90°
IN
ISO
100
USB
MCL
T1-1T

FIGURE 3-4



NOTES:

E1- STACKPOLE CORE 57-3099, NO. 14 ON TRANSISTOR BASE LEAD.

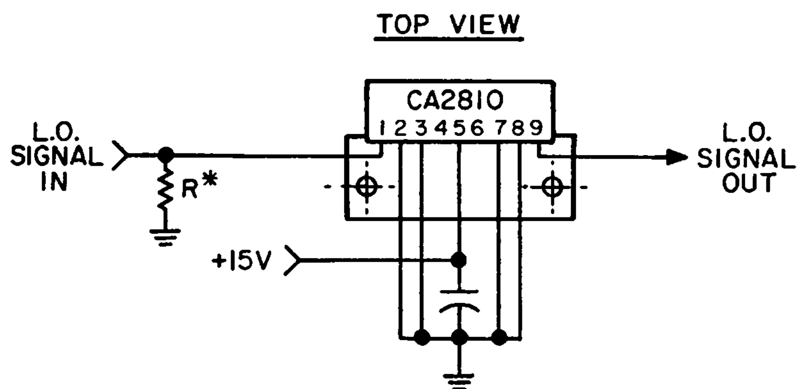
E2- 2 TURNS NO. 22 ON STACKPOLE CORE 57-0915, NO. 12.

E3- 2 TURNS NO. 22 ON ALLEN-BRADLEY CORE T0140A303B.

FIGURE 3-5: SCHEMATIC DIAGRAM OF THE BASEBAND QUADRATURE NETWORK port attenuators connected to the image reject mixer. The nominal input power level is shown in Figure 3-1. A single TRW Type CA2810 device is used for each power amplifier, with connections as shown in the schematic diagram of Figure 3-6.

3.4 Baseband Amplifiers

The baseband amplifier board comprises four separate two-stage amplifier sections, one for each baseband output channel. Each individual section is made up of a cascade of one Avantek GPD 461 and one Avantek GPD 462 amplifier, with external coupling capacitors connected as shown in Figure 3-7. Overall gain is typically 28 dB, flat to within ±0.1 dB in the frequency range 186 kHz to 50 MHz. A resistor divider network connected between the GPD 462 and ground provides an impedance matched baseband level monitor output which is attenuated approximately -20 dB relative to the actual baseband output signal.



*NOTES: R IS SELECTED TO ADJUST GAIN OF LO POWER SO THAT POWER AT IRM IS + 10dBm (±.1dB) FOR NOMINAL LO POWER AT REAR PANEL OMQ CONNECTORS.

FIGURE 3-6: SCHEMATIC OF TRW TYPE CA2810 DEVICE CONNECTIONS

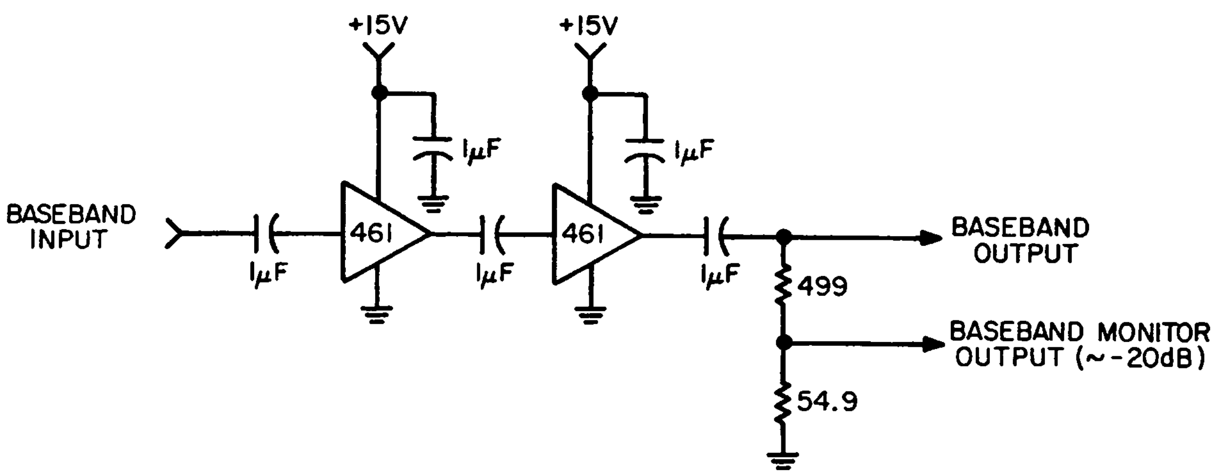


FIGURE 3-7: SCHEMATIC OF EXTERNAL COUPLING CAPACITORS CONNECTIONS

4.0 SPECIFICATIONS AND ADJUSTMENTS

4.1 Performance Specifications - T3 Module

The following specifications are realistic performance goals for the production T3 module derived from measurements on the prototype and early production modules and from system performance requirements.

4.1.1 Input/output signals

Composite IF Input Signal	
Total Power	- 4 dBm (±3 dB)
IF Passband Power Spectral Density	-11 dBm/50 MHz (±3 dB)
LO Input Signals	
1200 MHz	+11.0 dBm (±1.5 dB)
1800 MHz	+11.0 dBm (±1.5 dB)
100-150 MHz	- 9.0 dBm (±1.0 dB)
200-250 MHz	-11.0 dBm (±1.0 dB)
Baseband Output Signal	
Power Spectral Density (Nominal)	-19.5 dBm/50 MHz
Front Panel Connectors (Nominal)	-40.0 dBm/50 MHz

4.1.2 Worst-case unit-to-unit T3 gain variations

(Fixed Frequency)

Submodule	Total Nominal Gain (dB)	Worst Case Unit-to-Unit Variation (dB)
L-band Converter	-10.0	±0.5
Image Reject Mixer	-27.5	±1.0
Baseband Amplifier	+29.0	±0.5
TOTAL	- 8.5	±2.0

Hence, the expected worst case variation from antenna to antenna in baseband output power spectral density is ±5.0 dB.

4.1.3 Worst-case unit-to-unit variations in differential amplitude and phase response

Single Module Amplitude Response Variation (186 kHz to 50 MHz)	<±.4 dB
Two Module Differential Amplitude Response Variation (186 kHz to 50 MHz)	<±.2 dB
Two Module Differential Phase Response Variation (neglecting a fixed delay offset) (186 kHz to 50 MHz)	<±2°
Change in Differential Amplitude Response at a fixed baseband frequency as a function of differential module temperature	<±.01 dB/°C
Change in Differential Phase Response at a fixed baseband frequency as a function of differential module temperature	<±.08°/°C

4.2 Comparative Testing of Modules

Figure 4-1 indicates the procedures to be used for comparative testing of the IF-to-baseband converter modules. All signals for module and submodule testing are provided by the "T3 module test set" described in more detail in Appendix B. With the modules connected to the test fixture, select the appropriate "IF input" signal frequency range for the appropriate channel in which measurements are to be made. Connect a frequency counter to the "IF input monitor" jack on the test set. Connect the correctly terminated 50 Ω probe tees and probes of the HP Vector Voltmeter to the appropriate "baseband output" jacks on the test set. For baseband frequencies of less than

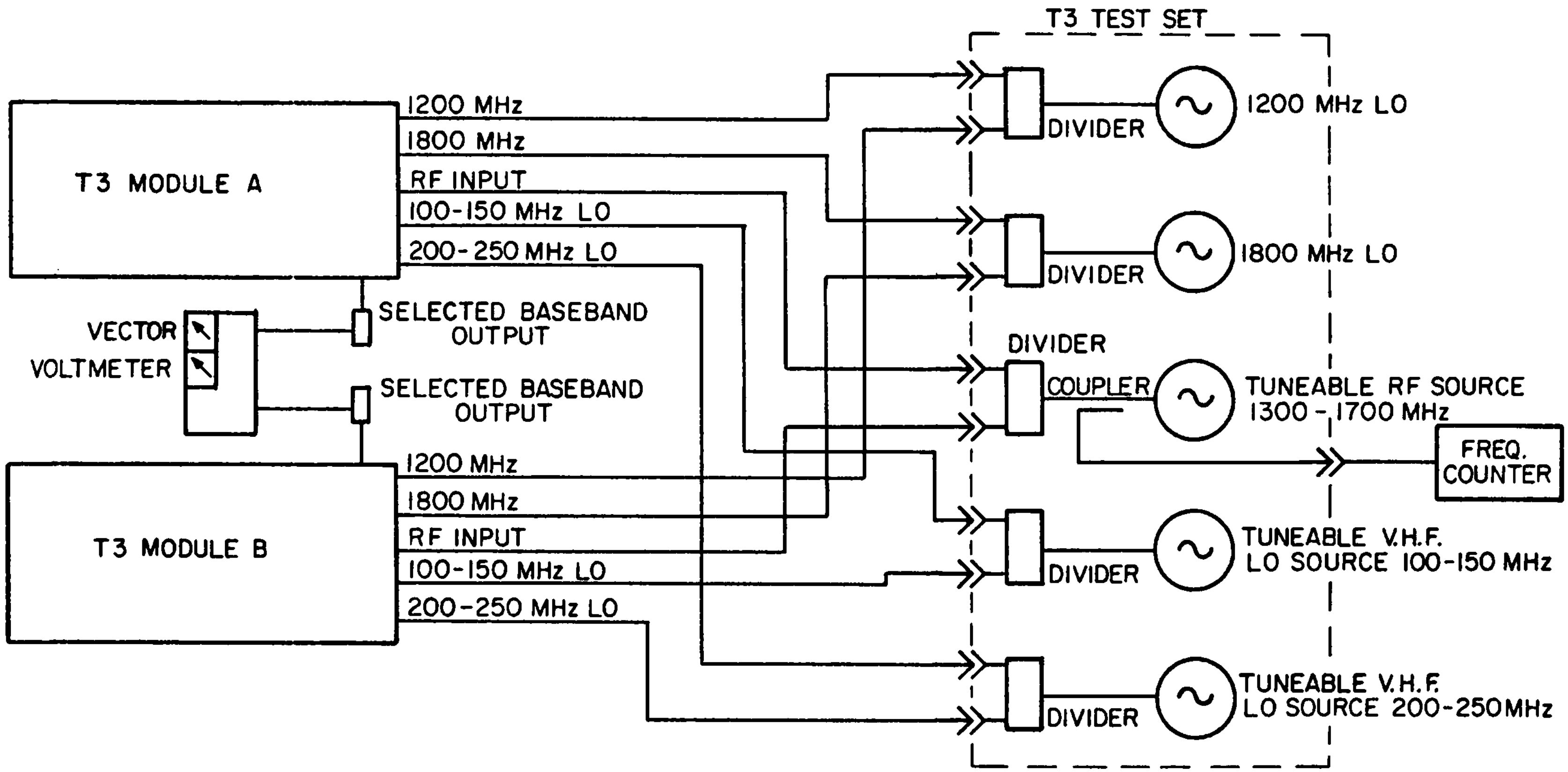


FIGURE 4-1: PROCEDURES TO BE USED FOR COMPARATIVE TESTING OF THE IF-TO-BASEBAND CONVERTER MODULES

2 MHz, the Vector Voltmeter does not give reliable phase readings. In this frequency range, the Wiltron Phase Meter should be used. Set the 100-150 MHz variable LO frequency synthesizer to 100 MHz for A channel measurements or 150 MHz for D channel measurements. Set the 200-250 MHz synthesizer to 200 for B channel measurements or 250 MHz for C channel measurements.

By varying the RF input frequency, point-by-point plots can now be made of the differential baseband amplitude and phase response of the modules under test. Measurements at 2 MHz (above baseband frequency of 2 MHz) or 200 kHz (below this frequency) increments in baseband frequency should give sufficient resolution. Typical results of such comparative measurements on prototype modules are shown in Figure 4-2.

For differential response measurements at differing module temperatures, the extender cables provided with the test set may be used to enable one unit to be placed in the temperature controlled oven. Complete baseband frequency response measurements should be taken at intervals of 10°C in differential temperature for a range of temperature differences -10°C to +20°C. These measurements should be repeated through at least one complete thermal cycle to ensure repeatability. At least one hour thermal stabilization time should be allowed, after each change in temperature, before measurements are taken. Typical temperature coefficient measurements are shown in Figure 4-3, for two prototype modules.

It is suggested that comparative tests as outlined above be carried out on a sample of between 10% and 20% of the production T3 modules, randomly selected. This should be sufficient to check for significant systematic departures from nominal performance.

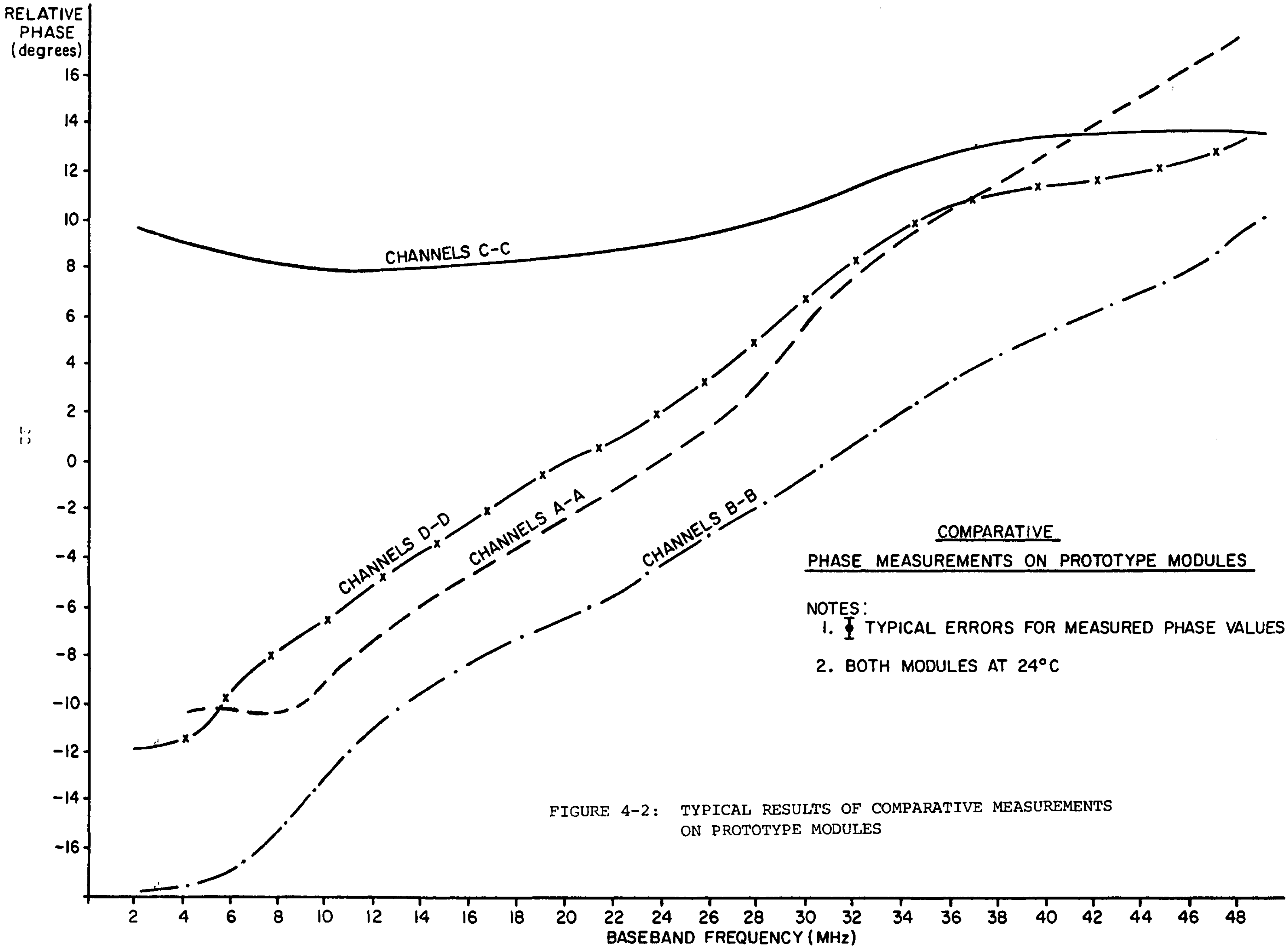


FIGURE 4-2: TYPICAL RESULTS OF COMPARATIVE MEASUREMENTS ON PROTOTYPE MODULES

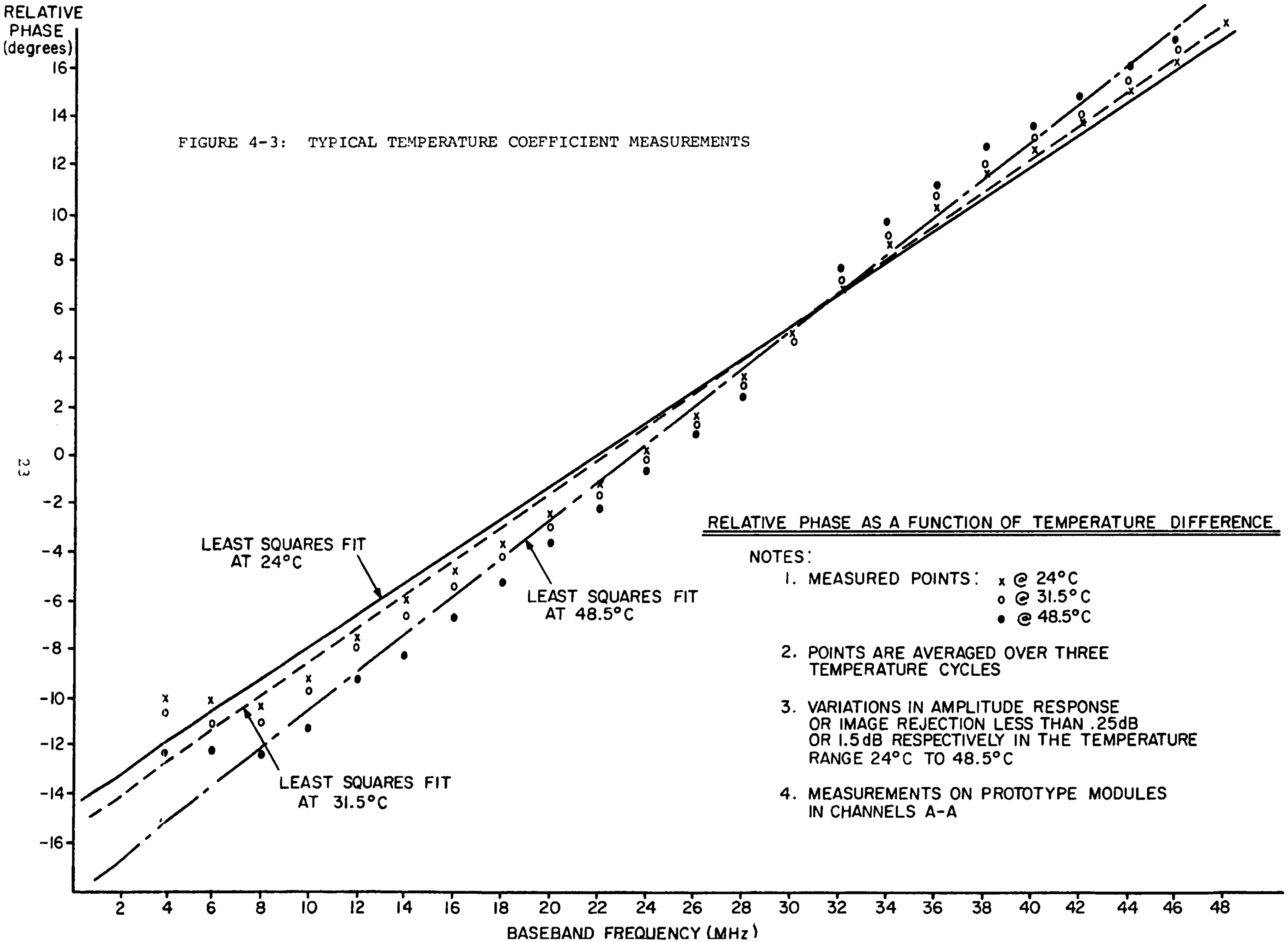


FIGURE 4-3: TYPICAL TEMPERATURE COEFFICIENT MEASUREMENTS

4.3 Image Reject Mixers - Specifications and Adjustment Procedures

The performance of the four image reject mixers in the T3 module may be specified in terms of the following parameters.

4.3.1 With the LO frequency for a given mixer set so that the full 50 MHz RF band is converted to the baseband (186 kHz to 50 MHz), and a LO power level of +10 dBm (±1 dB), the conversion loss should be 27.5±1.0 dB. The maximum allowable variation in conversion loss across the baseband range is ±0.2 dB.

4.3.2 With an arbitrary LO frequency within the specified range for a given mixer, and a LO power level of +10 dBm (±1 dB), the rejection of the undesired image response should meet the following criteria (refer to Figure 4-4):

a) For image signals within the RF passband and separated from the LO signal frequency by less than the maximum available baseband bandwidth, the image rejection should exceed 30 dB.
b) For image signals outside this range, the image rejection should exceed 20 dB.

4.3.3 A given mixer should meet the following additional specifications within its operating frequency range:

a) Return loss at the RF or LO ports should exceed 20 dB.
b) Isolation between RF and baseband or LO and baseband should exceed 40 dB.
c) Isolation between RF and LO ports should exceed 15 dB.

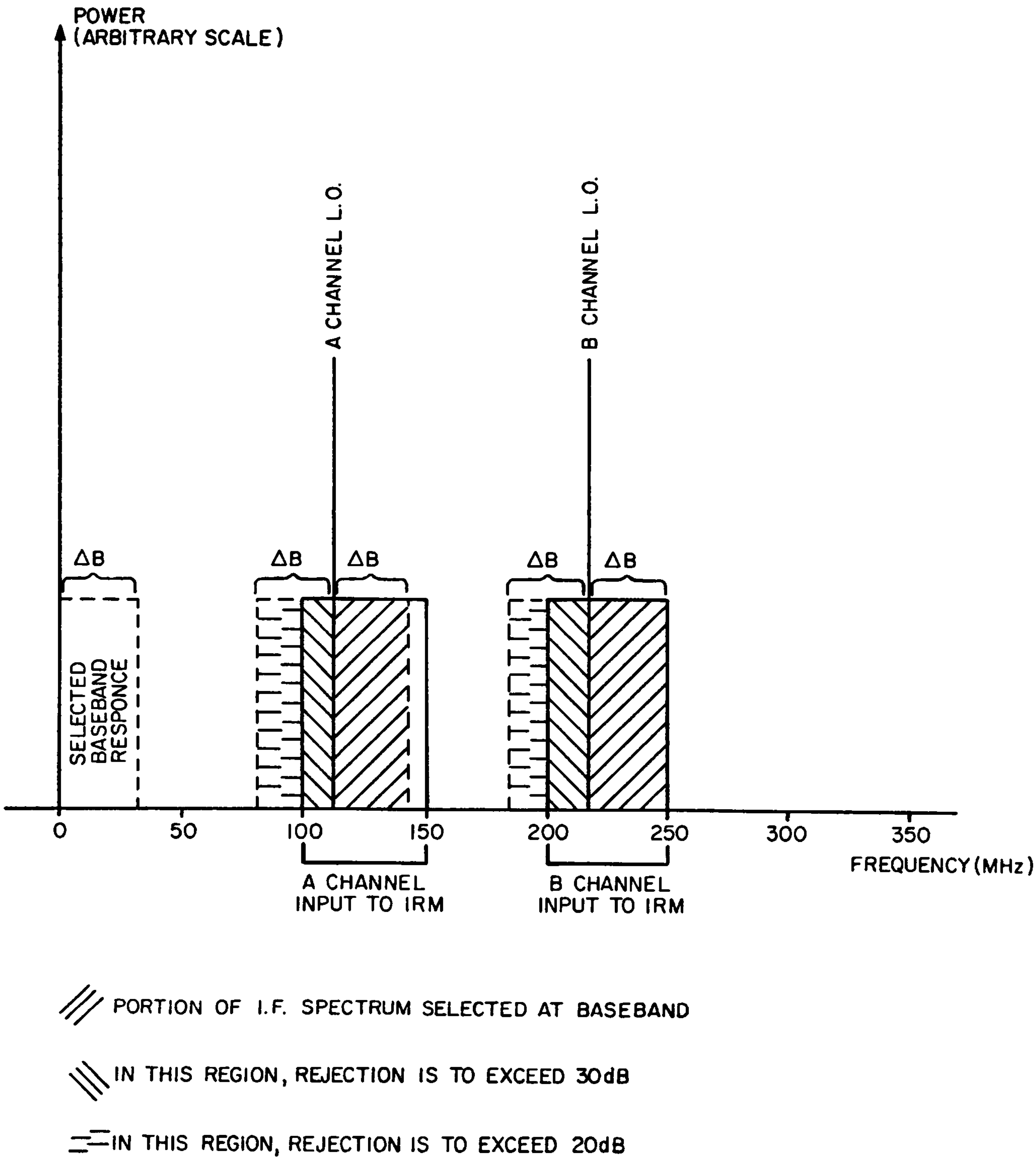


FIGURE 4-4: CRITERIA FOR REJECTION OF UNDESIRED IMAGE RESPONSE

d) The mixer noise figure should not exceed its conversion loss by more than 8 dB.

e) The 1 dB gain compression level, referred to the RF input, should exceed +7 dBm.

f) The third order intercept point for mixer two-tone intermodulation products, referred to the RF input, should exceed +16 dBm.

g) The second order intercept point for quadrature network amplifier two-tone intermodulation products, referred to the RF input, should exceed +30 dBm.

An approach to the alignment of the image reject mixers is outlined below. The mixer should be connected, in the manner shown in Figure 4-5, to the "T3 module test set". A Tektronix 7L13 Spectrum Analyzer should be connected to the "mixer test baseband output" jack. Typical spectrum analyzer settings for preliminary mixer performance analysis are:

Freq. Span/Div.	5 MHz
Center Freq.	25 MHz
Resolution	300 kHz
Reference Level	-10 dBm
Vertical Scale	10 dB/Div. Log.
Time Base Source	Ext.
Time Base Mode	Norm.
Display Mode	Store
Persistence	-30° CW from Max.

STEP 1: Adjust the LO synthesizer so that the frequency is at the appropriate edge of the range for a given mixer to display the complete 50 MHz bandwidth desired sideband response (i.e., 100 MHz or 200 MHz for a LSB reject mixer, 150 MHz or 250 MHz for an

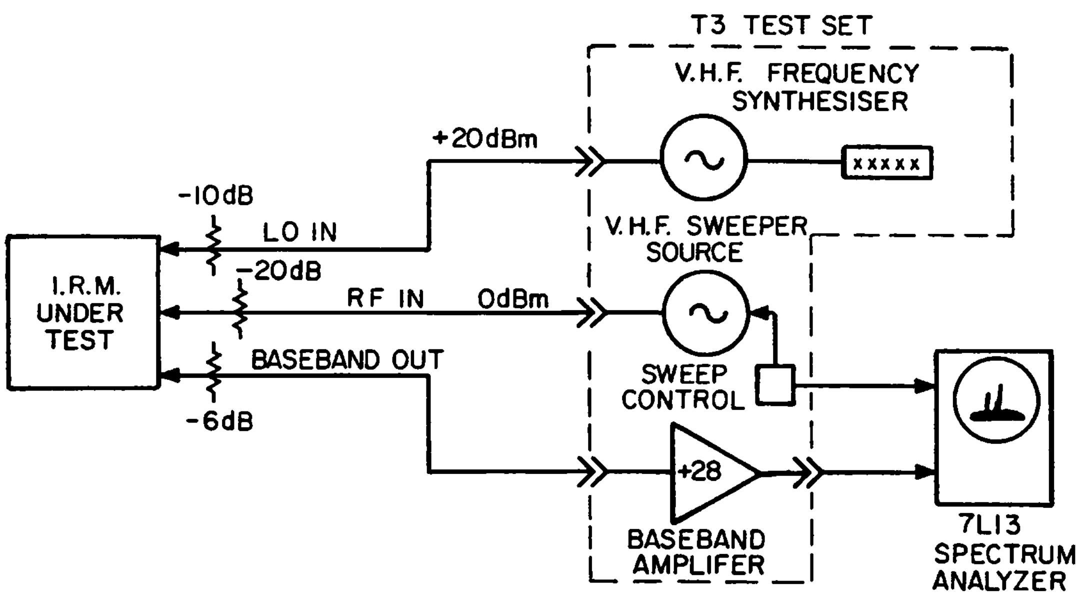


FIGURE 4-5: MIXER CONNECTIONS TO T3 MODULE TEST SET

USB reject mixer). Adjust the test set sweep controls to cover the desired IF band (i.e., for a 100-150 MHz mixer, sweep center frequency should be 125 MHz with a Δf of ± 25 MHz - for a 200-250 MHz mixer, sweep center frequency should be 225 MHz with a Δf of ± 25 MHz). This may best be done with the aid of the spectrum analyzer connected to the "mixer IF sweep out" jack, by simply adjusting the center frequency control on the analyzer to display the desired region of the spectrum.

STEP 2: Reconnect the spectrum analyzer to the "mixer test baseband output" output. The displayed output enables mixer conversion loss and conversion loss flatness to be determined. Note the mixer output power level displayed - this is to be used as a reference for image rejection measurements. Change the LO frequency by 50 MHz, setting it to the opposite end of the range. The spectrum analyzer now displays the rejected image response for this LO frequency. As a first step towards optimizing this response, adjustments are made to the inductors L1 and L3 and the capacitors C1 and C7 of the baseband quadrature network, in order to reduce the level of unwanted image signal to within specification, relative to the desired sideband reference level obtained earlier. The image rejection is affected by each component over a different portion of the baseband spectrum - these ranges are approximately

Component	Frequency Range Affected (MHz)
L3	2-10
L1	5-25
C7	10-35
C1	30-50

All adjustments interact to some extent and it is therefore necessary to repeat the adjustment procedure several times to achieve an optimum response.

STEP 3: Next, change the spectrum analyzer center frequency to 1 MHz, the Freq. Span/Div. to 200 kHz and the resolution to 30 kHz. Adjust the sweep center frequency to display the undesired image response only. Adjust inductances L2 and L4 until the image rejection between 200 kHz and 2 MHz is within the specified limits.

STEP 4: Having made the above adjustments and obtained the best response, the amplitude balance of the mixer quadrature outputs should be adjusted to give optimum rejection. The most significant effect of this compensation, in general, is for baseband frequencies above 20 MHz. Place a variable resistor (0-5 kΩ, miniature) in parallel with one of the 100 Ω matching resistors (refer to Figure 3-5) - if adjustment of the variable resistor degrades the response, transfer it to the complementary resistor - and adjust till the best rejection is obtained. If no improvement is noted, remove the resistor. If improvement is obtained, measure the value of the shunt resistor and replace with a fixed 1% metal film (RN55) resistor of closest value.

STEP 5: The next step is to vary the LO frequency, in 5 MHz increments, across the range specified for the mixer, while adjusting the IF sweep center frequency to maintain a display of the rejected image response. The inductance and capacitance settings (L1, L3, C1, C7) should now be adjusted (<u>in small increments only</u>!!) so that greater than the specified minimum image rejection can be obtained for each LO frequency setting, over the full baseband frequency range.

STEP 6: Having optimized the image rejection performance, the conversion loss response for the desired sideband should be rechecked as in STEP 1. Response flatness can be optimized by adjusting the feedback compensation capacitor, C16, in the quadrature network amplifier.

With a little practice, this procedure can easily be learned and optimized. The whole adjustment should take no longer than one hour per mixer once the operator has gained the necessary experience in the method. Measurement of the remaining mixer parameters is straightforward using the "T3 module test set", and will not be outlined in detail here. It should only be necessary to check the specified performance data, other than image rejection and conversion loss response, on a limited sample of the production image reject mixers.

4.4 Spurious Correlated Signals at the Baseband Output of the T3 Module

At the correlator input, the minimum detectable level for pairs of corrected spurious signals (in different antenna systems) can be written as:

$$\sigma = S_o\sqrt{B/T}$$

where $B = B_0/N$

S_o is the noise power spectral density

B is the correlator channel bandwidth

B_0 is the baseband filter bandwidth

N is the number of channels resolved by the correlator

T is the integrating time.

At the output of the T3 module, the noise power spectral density is nominally -19.5 dBm/50 MHz or 2.2441×10^{-13} W Hz^{-1}. Since the power spectral density at this point is bandwidth independent, the worst case interference by spurious correlated signals occurs for the narrowest channel bandwidth (0.2 MHz, 256 channels resolved)

$$\sigma_{min} \text{ (worst case) } = 3.0178 \times 10^{-14} \text{ W } (-105.2 \text{ dBm})$$

for an integration time of 12 hours. Hence, no spurious correlated signal in the baseband frequency range should be greater in amplitude than -105.2 dBm at the T3 module output.

There exist two principal sources of spurious correlated signals. The first is the coupling of DCS system signals, typically at frequencies below 500 kHz, into the baseband system via the rack power wiring. Without the use of a power line filter, these spurious signals exhibit maximum levels at the baseband output of the T3 module as shown in the following table.

Spur Frequency	Level (Measured ±3 dB)	Source in D-Rack
9.8 kHz	-96 dBm	Local Buffer
11.1 kHz	-86 dBm	Data Tap
88.8 kHz	-97 dBm	Data Tap

With the use of the simple L-C filter shown in Figure 4-6, these levels were reduced to less than the maximum tolerable spurious signal level.

The second source of spurious signals at the baseband output is the feedthrough of low-frequency mixing products from the first conversion stage in the L-band converter submodule. The 5 MHz sidebands on the 1200 MHz carrier and the 500 kHz sidebands on the 1800 MHz carrier of the composite IF input

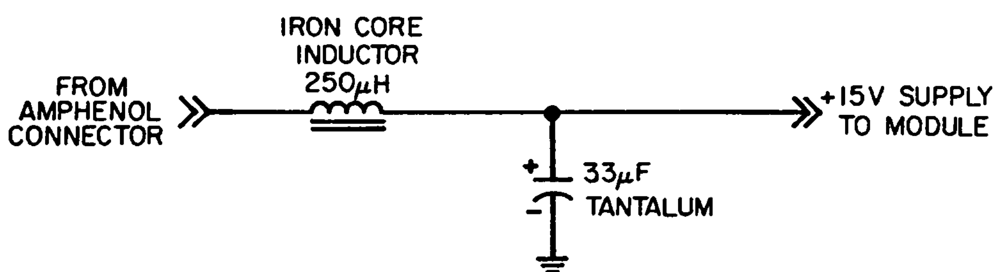


FIGURE 4-6: L-C FILTER

signal generate 5 MHz spurs in the A and B channel outputs and 500 kHz spurs in the C and D channel outputs. The expected level of these spurs, without the high-pass filters shown in Figure 2-1 inserted, can be predicted as follows:

Power in 1200 or 1800 MHz carrier at MD123 input	-12 dBm
Power in 5 MHz sideband	-27 dBm
Power in 500 kHz sideband	-27 dBm
Power out of MD123 at 5 MHz (neglecting LO phasing)	-25.5 dBm
Power out of MD123 at 500 kHz (neglecting LO phasing)	-30.5 dBm
TOTAL loss at 5 MHz or 500 kHz from MD123 to baseband output	-47 dB
So predicted T3 output level of 5 MHz spur	-72.5 dBm
and predicted T3 output level of 500 kHz spur	-77.5 dBm

These figures neglect an additional factor of uncertain magnitude (possibly -20 dB extra rejection) due to LO phasing effects at the MD123 and the effects of Walsh function phase switching on the correlator output.

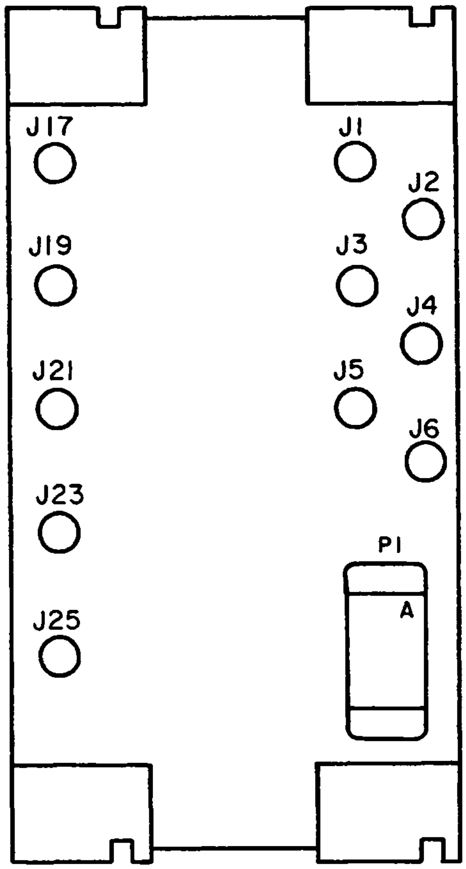
The measured value of the magnitude for these spurs at the T3 output without filtering are as follows:

5 MHz spur:	-73 dBm (±2 dB)
500 kHz spur:	-79 dBm (±3 dB)

The insertion of high-pass filters between the L-band converter and the image reject mixers reduces the level of these spurs by at least 70 dB.

5.0 INPUT/OUTPUT CONNECTIONS

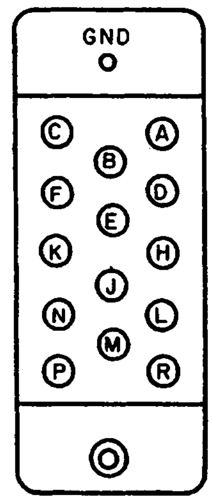
Refer to Figure 5-1 for location of input/output jacks and AMP-connector pin numbers. The input/output connections are as shown in the figure.



TRIPLE WIDE MODULE

(REAR VIEW)

CONN	FUNCTION
J1	100-150 MHz LOCAL OSC A
J2	100-150 MHz LOCAL OSC D
J3	CHANNEL A OUT
J4	CHANNEL B OUT
J5	CHANNEL C OUT
J6	CHANNEL D OUT
J7-J16	NOT USED
J17	1800 MHz LO
J19	1200 MHz LO
J21	RF INPUT (FROM T2)
J23	200-250 MHz LOCAL OSC C
J25	200-250 MHz LOCAL OSC B
J18,20	NOT USED
J22,24	NOT USED



P1 (REAR VIEW)

P1		
PIN	FUNCTION	WIRE COLOR
A	+28V ✳	——
B	+15V	RED
C	+5V ✳	——
D	GND	BLACK
E	-15V ✳	
F		
H		
J		
K		
L		
M		
N		
P		
R		

✳ INDICATES A FUNCTION NOT FOUND IN THIS MODULE

FIGURE 5-1

6.0 DOCUMENTATION - BOM, ARTWORK, DRAWING REFERENCES

REVISIONS

REV	DATE	DRAWN BY	APPRV'D BY	DESCRIPTION

DRAWN BY	DATE 8/14/79
DESIGNED BY	DATE
APPROVED BY	DATE 8/15/K.K.

NEXT ASSY	USED ON

NATIONAL RADIO ASTRONOMY OBSERVATORY
SOCORRO, NEW MEXICO 87801

V L A

PROJECT T3

TITLE BASEBAND CONVERTER "C" DWG

DWG NO. A13820C4 SHEET 1 OF 4

ASSEMBLY NAME T3 - BASEBAND CONVERTER			SERIES/MODEL	USED ON
DRAWING NO.	**REV.**	**TITLE**		**NOTES**
A13820Z03	A	BASEBAND CONVERTER - BOM		
D13820P03	B	- ASSY		
C13820EØ8	E	- SCHEMATIC		
A13820W14	–	- REAR PANEL CONN.		
A13820Z13	A	LO POWER AMPLIFIER - BOM		
D13820P13	B	- ASSY		
C13820M27	E	- HOUSING		
C13820M37	A	- SPACER		
A13820Z29	–	- BOM		
C13820M36	–	- DRILL JIG		
C13820P28	–	- DRILL JIG ASSY		
A13820Z18	A	BASEBAND AMPLIFIER - BOM		
C13820P11	A	- ASSY		
C13820M24	–	- R.F. SHIELD		
C13820M23	–	- DRILL DWG.		
C13820AB15	A	- P.C.B. ARTWORK		
A13820Z19	–	L - BAND CONVERTER - BOM		
D13820P18	B	- ASSY		
D13820M11	C	- HOUSING		
B13820M22	B	- CLAMP		
D13820MØ9	D	- STRIPLINE DRILL DWG		
E13820AF5	D	- STRIPLINE ARTWORK		

NOTES:

GENERAL USE ITEM

DWG. NO.

SHEET

REV

ASSEMBLY NAME T3 - BASEBAND CONVERTER | SERIES/MODEL — | USED ON —

DRAWING NO.	REV.	TITLE	NOTES
A13820Z14	—	IMAGE REJECT MIXER (100-150 USB) - BOM	
A13820Z15	—	IMAGE REJECT MIXER (100-150 LSB) - BOM	
D13820P16	—	- ASSY	
D13820M1Ø	D	- HOUSING	
D13820MØ4	B	- STRIPLINE DRILL DWG	
E13820AD1	A	IMAGE REJECT MIXER (100-150 MHz) - ARTWORK	
A13820Z16	—	IMAGE REJECT MIXER (200-250 USB) - BOM	
A13820Z17	—	IMAGE REJECT MIXER (200-250 LSB) - BOM	
D13820P15	—	- ASSY	
D13820M10	D	- HOUSING	REF
D13820M04	B	- STRIPLINE DRILL DWG	REF
E13820AD2	B	IMAGE REJECT MIXER (200-250 (MHz) - ARTWORK	
A13820Z12	—	BASEBAND QUADRATURE NETWORK - BOM	
D13820P12	B	- ASSY	
C13820M26	A	- DRILL DWG	
C13820AD19	A	- ARTWORK	
D13820M15	E	BAR SUPPORT, TOP & BOTTOM	
D13820M16	A	PARTITION PLATE	
D13820M13	A	PANEL, FRONT	
C13820M17	A	PANEL, REAR	
D13820MØ6	B	SIDE PLATES	

NOTES:

1. GENERAL USE ITEM

DWG. NO. A1382001

SHEET 3 of 4

REV.

ASSEMBLY NAME T3 - BASEBAND CONVERTER		SERIES/MODEL	USED ON
DRAWING NO. (1–18)	**REV.**	**TITLE**	**NOTES**
013050104	-	GUIDES	1.

DWG. NO. 1 [illegible]

SHEET 4 of 4

REV.

NOTES:

1. GENERAL USE ITEM

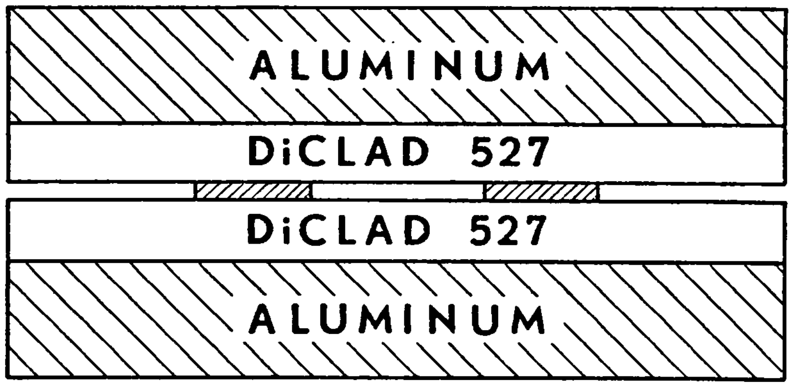
7.0 COMPONENT INFORMATION - MANUFACTURERS DATA SHEETS

KEENE
CORPORATION
CHASE-FOSTER DIVISION

P.O. BOX 4305, EAST PROVIDENCE, R.I. 02914 • TELEPHONE (401) 434-2340
P.O. BOX 700, NEWARK, DELAWARE 19711 • TELEPHONE (302) 738-9150

ALUMINUM CLAD DICLAD 527.
TECHNICAL BULLETIN 602
April 1977

ALUMINUM CLAD DICLAD 527 FOR STRIPLINE AND MICROSTRIP CIRCUITS



Description:

DiClad 527 P.T.F.E./Glass laminate, bonded to heavy gauge aluminum. The bonding is accomplished by a very thin fluorocarbon film which does not change the apparent dielectric constant of the laminate. The opposite side of the laminate may be left unclad or clad with standard E.D. copper. For Dielectric material specifications see Bulletins 600 and 620.

Application:

Aluminum clad DiClad 527 may be used as a unitized dielectric/ground plane/outer case.

Advantages:

Lower Production Costs — fewer pieces to handle and fewer assembly steps required.
Better End Item Quality — since the dielectric and the aluminum are bonded during lamination, there are no random air gaps between them.
Mechanical Strength for Small Components — the cost of machining cases for small components may be prohibitive. Punching both the aluminum and the dielectric and using eyelet or rivet assembly may increase profits.

Sheet Size:

16" ±1" x 36" ±1"

Price and Availability:

Due to the specialized nature of this product it is priced and manufactured on a per order basis.
~~Please contact Chase-Foster direct.~~

KEENE
CORPORATION
CHASE-FOSTER DIVISION

P.O. BOX 4305, EAST PROVIDENCE, R.I. 02914 • TELEPHONE (401) 434-2340
P.O. BOX 700, NEWARK, DELAWARE 19711 • TELEPHONE (302) 738-9150

PRODUCT INFORMATION
BULLETIN #600
DI-CLAD 522 & 527
ISSUED MARCH 1978
Ref. Price List PL-2172

LAMINATED MICROWAVE DIELECTRIC MATERIAL

Product Definition:

Di-Clad is a lamination of PTFE coated glass fabric with copper cladding on one or two surfaces. The composite utilizes the highest grade and purity of base materials in conjunction with high purity electro-deposited copper. This construction results in a controlled reproducible dielectric constant. In addition, a clean room environment and high quality standards ensure closely controlled thickness tolerances. These features, in conjunction with a low dissipation factor, make Di-Clad an excellent product for microwave stripline application.

Applications:

Di-Clad is used in such systems as F.A.A Radars, phased array antennas, Satellite communication equipment, collision avoidance systems, intrusion sensing devices, microwave test equipment and mobile communications systems. Typical circuits utilizing Di-Clad in these systems include power dividers, mixers, I.F. amplifiers, filters, parallel line and capacitive couplers and antennas.

Features:

There are two grades available, 522 and 527. The most significant difference between these two grades is dissipation factor; with a Dielectric Constant of 2.5 Grade 522 has a typical dissipation factor of .0010 at 1 MHZ and .0025 at X band. With a dielectric constant of 2.5, Grade 527 has a typical dissipation factor of .0019 at X band (8.2 to 12.4 GHZ.)

Dielectric Constant: Both grades are available with a nominal dielectric constant in the range of 2.4 to 2.6. The standard tolerance on the nominal value is ± .04 for grade 527 and ± .05 for grade 522. When ordering the value and tolerance should be specified. Closer tolerances can be obtained on special request.

Thickness: 1/32" to 1". Base laminate thickness of .004" to 1/32" are available and defined in Bulletin #620.
Thickness Tolerances: Di-Clad 522 and 527 can be supplied to the Class 4 tolerance of MIL-P-13949E. See tables on Page 2.
Sheet Size: Standard sheets are supplied as 16"x36" or 36"x48" to tolerances of ± 1" on length and width. Panels are available cut from the sheet size.

U.S. Government Specs.: Di-Clad 522 is qualified to the Copper Clad Laminates Spec. MIL-P-13949E Type FL-GT. The base material also qualifies to Spec. MIL-P-19161 Type GTE. Grade 527 is qualified to MIL-P-13949E type GX.

Clading: The standard grades utilizes Electro Deposited Copper; however, other clading is available such as aluminum or rolled copper.

Plating: Di-Clad PTFE laminates have excellent properties for all electro plating operations, including through holes and gold plating processes.
See Bulletin #630

Bondable Products:

1) DiClad can be supplied with an unclad surface which is chemically etched for bonding.
2) DiClad can be provided with an after etch surface that is suitable for multi layer applications.

STANDARD GRADES

Grade No.	Military Grade Base	Military Grade: Copper Clad	Description	Sheet Size & Thickness Range
522	MIL-P-19161 Type GTE	MIL-P-13949 Type GT	Outstanding Properties Controlled Dielectric Constant; Minimum Moisture Sensitivity Fine Weave	16"x36"; 36"x48" 1/32" to 1.000"
527	—	MIL-P-13949 Type GX	Controlled Dielectric Constant & Low Dissipation Factor	16"x36"; 36"x48" 1/32" to 1.000"

TOLERANCES — COPPER CLAD LAMINATES — ALL GRADES
(INCHES PLUS OR MINUS)

Tolerances: Di-Clad 522 and Di-Clad 527 can be supplied with thickness tolerances as close or closer than specified in Spec. MIL-P-13949E. Standard for these grades are listed below.

COPPER — ONE OR BOTH SIDES AND 1 OUNCE OR 2 OUNCE

Note these tolerances apply to overall thickness of sheet including copper foil.

Overall Thickness Incl. Copper Foil	Chase-Foster Normal	MIL-P-13949E Class 1	Chase-Foster Close	MIL-P-13949E Class 2	Chase-Foster Extra Close	MIL-P-13949E Class 3
1/32"	±.0065"	(±.0065")	±.003"	(±.0040")	±.002"	(±.0020")
3/64"	±.007"		±.003"		±.002"	
1/16"	±.0075"	(±.0075")	±.004"	(±.0050")	±.002"	(±.0030")
3/32"	±.009"	(±.0090")	±.005"	(±.0070")	±.003"	(±.0040")
1/8"	±.012"	(±.012")	±.006"	(±.0090")	±.004"	(±.0050")
1/4"	±.022"	(±.022")	±.012"	(±.012")	±.006"	(±.006")

Di-Clad 522 and 527 can be supplied to thickness tolerances shown in Spec. MIL-P-13949E Class 4 for microwave applications.

COPPER — ONE OR BOTH SIDES AND 1 OUNCE OR 2 OUNCE

Tolerances for material in this category are based on sheet thickness exclusive of Copper Foil.

Sheet Thickness Exclusive of Copper Foil	Tolerance	Sheet Thickness Exclusive of Copper Foil	Tolerance
.030"	±.002"	.090"	±.003"
.047"	±.002"	.120"	±.0035"
.060	±.002"	.240"	±.004"

THICKNESS: At least 90% of the area of the sheet shall be within the tolerance given and at no point shall the thickness vary from the nominal by a value greater than 125 per cent of the specified tolerance.

SHEET TYPICAL PROPERTIES — MECHANICAL*
(1/16 THICK)

	Test	Condition	522	527	MIL-P-13949 522 & 527
Continuous Operating Temperature			500°F	500°F	
Tensile Strength, psi Lengthwise	D-229	A	19,000	—	—
Crosswise			15,000	—	—
Flexural Strength, psi Lengthwise	Method 1031	A Fed-STD-406	19,000	15,500	15,000
Crosswise	Method 1031	Fed-STD-406	16,000	12,000	10,000
Flexural Modulus, psi x 10^6 Lengthwise	D-229 ASTM	A	1.2	—	—
Crosswise			1.0	—	—
Compressive Strength, psi	D-229 ASTM	A	20,000	—	—
Izod Impact str., Flatwise lb. per in. notch Lengthwise	D-229 ASTM	E-48/50	15	—	—
Crosswise			13	—	—
Rockwell Hardness, M. Scale	D-229 ASTM	A	30	—	—
Specific Gravity		A	2.2	2.2	—
Water Absorption, %	Method 1031	FED-STD-406 E1/105 + D24/23	.01	.01	.10
Peel Strength lb. per in. of width 1 oz.	NEMA LP-1 10-12	After 20 sec. at 500°F Solder Float	12	12	8.0 Ave. 6.5 Min.
2 oz.	"	"	16	16	10.0 Ave. 8.5 Min.
1 or 2 oz.	MIL-P-13949E	(at elevated temp.)	10	10	2.0 Min.
Solder Blister Resistance (sec.)	MIL-P-13949E	500°F	60+	60+	20.0 Min.
Flammability, sec. to extinguish first ignition	UL-94	A	0	0	—
second ignition			0	0	—
Linear Coefficient of Thermal Expansion in/in/°C	(25°-150°C)				
Lengthwise			$9\text{-}10 \times 10^{-6}$	$9\text{-}10 \times 16^{-6}$	
Crosswise			$9\text{-}10 \times 10^{-6}$	$9\text{-}10 \times 16^{-6}$	

*Except for peel strength and hot solder properties, mechanical properties shown are for plain laminate, or after etching away of copper if copper-clad. For electrical properties, same except for insulation resistance.

DI-CLAD SHEETS

TYPICAL PROPERTIES — ELECTRICAL*
1/16" THICK

	Spec./ Standard	Condition	522	527	MIL-P-13949E[2]
Insulation Resistance Meghoms	ASTM EIA-comb.	C96/35/90	100,000	100,000	—
Dielectric Strength Parrallel, Stepwise kv	Method 4031 FED-STD-406	A D-48/50	70 45	70 45	20
Dielectric Strength Perpendicular, short time vpm	D-229 ASTM	A	800	800	
Volume Resistivity megohms, cm	Method 4041 FED-STD-406	MIL-P-13949E	10^{6+}	10^{6+}	10^{6}
Surface Resistance meghoms	Method 4041 FED-STD-406	MIL-P-13949E	10^{4}+	10^{4}+	10^{4}
ARC Resistance seconds	Method 4011 FED-STD-406	A	180+	180+	180
Dielectric Constant 1 MHZ[1] X BAND	MIL-P-13949E		 2.5	 2.5 2.5	522 527 2.8 Max. 2.4-2.6
Dissipation Factor 1 MHZ[1] X Band	MIL-P-13949E		(Typical) .0010	— (Typical) .0019	.005 Max. .0022 Max.

[1]Dielectric constant available in range of 2.4 to 2.6 with a standard tolerence of ±.05 on grade 522 and ±.04 on grade 5.7. Closer tolerances available on special request.

[2]When MIL-P-13949E is specified on an order, values will be material requirements.

All Statements contained in this data sheet are based on tests we believe to be reliable but cannot guarantee. The only obligation of the manufacturer or seller will be to replace such material proved to be defective. Neither manufacturer or seller are liable for any injury, loss or damage direct or consequential, resulting from the use of or the inability to use the product. It is the responsibility of the user to determine the applicability of the product for his intended use, and the user assumes all risk and liability whatsoever in connection therewith.

No statement or suggestion not contained in this data sheet will have any force or effect.

WARP AND TWIST TOLERANCES
ALL GRADES

Data shown below is same as standards listed in Military Spec. MIL-P-13949.

Total Variation, Maximum, Per Cent (on basis of 36" dimension).

Thickness Inches	CLASS A Copper 1 side	Copper 1 side	CLASS B Copper 2 sides	Copper 2 sides
1/32	12%	5%	10%	2%
3/64	10%	5%	5%	1%
1/16	10%	5%	5%	1%
3/32	8%	3%	5%	1%
1/8	8%	3%	5%	1%
5/32	5%	1.5%	5%	1%
3/16	5%	1.5%	5%	1%
7/32	5%	1.5%	5%	1%
1/4	5%	1.5%	5%	1%

For nominal thicknesses not shown, the warp or twist of the next greater thickness will apply. Test Method ASTM D-709.

PITS AND DENTS
ALL GRADES

Data shown below is same as standards listed in Military Spec. MIL-P-13949.

Class A — The maximum total point count for pits and dents combined, in any 12"x12" area of surface shall be less than 30. The point system is listed below.

Class B — There shall be no pits with a dimension greater than .015". Pits with the longest dimension greater than .005" shall not exceed 3 per sq. ft.

Longest Dimension (Inches)	Point Value Class A	Class B
0.005 to 0.010 inclusive	1	1
0.011 to 0.015 inclusive	2	2
0.016 to 0.020 inclusive	2	Not Allowed
0.021 to 0.030 inclusive	4	Not Allowed
0.031 to 0.040 inclusive	7	Not Allowed
Over 0.040	30	Not Allowed

COPPER FOIL

Electrolytic 1 and 2 oz. copper foil of printed circuit quality is standard. Other weights including ½, 3 and 5 oz. and rolled copper foil are available on request.

THICKNESS TOLERANCES OF FOIL

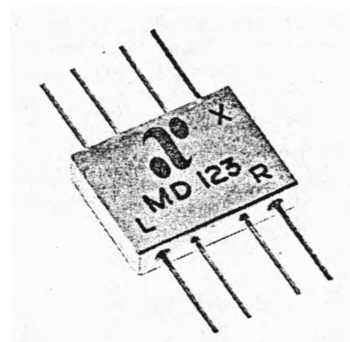
NOMINAL WEIGHT OZ. PER SQ. FT.	NOMINAL THICKNESS, Inch	THICKNESS TOLERANCE, Inch PLUS	MINUS
1	0.0014	0.0002	0.0002
2	0.0028	0.0003	0.0003
3	0.0042	0.0004	0.0004
4	0.0056	0.0006	0.0006
5	0.0070	0.0007	0.0007

SCRATCHES

Any number of scratches less than 140 microinches deep are permitted. For full size sheets, the number of scratches, 4 inches in maximum length over 140 microinches deep and with a maximum depth of 1/5 of copper foil thickness, shall not exceed five per square foot, except within 1/4 inch of the edge, where deeper scratches are permitted. For cut sheets, there shall be no scratches over 140 microinches deep, except within 1/4 inch of the edge, where deeper scratches are permitted.

SUBMINIATURE FLATPACK DOUBLE-BALANCED MIXER

10 MHz - 3 GHz



FEATURES

- Low Profile Flat-Pack — only .125" high
- 10 — 3000 MHz
- 7 db Conversion Loss Typical
- Rugged — designed to meet MIL-E-5400
- Specifications Guaranteed —54°C to +125°C
- Low Cost

GUARANTEED SPECIFICATIONS

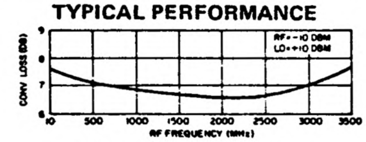
Frequency Range:			
R (RF) Port	10 MHz - 3 GHz		
L (LO) Port	10 MHz - 3 GHz		
X (IF) Port	DC - 3 GHz		
Conversion Loss:			
(10 MHz - 3 GHz)*	8 db Max.		
Isolation (db min.):	LO-RF	LO-IF	RF-IF
10-500 MHz	25	20	20
500 MHz - 1 GHz	30	25	25
1 - 3 GHz	25	25	20
LO Power Required:	+7 to +20 dbm (all specs. guaranteed with +10 to +13 dbm)		
Input Power Total:	400 mW Max. @ 25°C (derate linearly to +125°C @ 3.2 mW/°C)		
X Port Input Current:	50 mW Max.		
Operating Temperature Range:	-55°C to +125°C		
D.C. Polarity:	Positive — with two in-phase signals applied to LO and RF Ports		
Impedance:	50 ohms (all ports)		

*Specified from 10 MHz to 2 GHz for IF frequency. IF port response down 4 db at DC, down 3 db at 3 GHz (see IF PORT BANDWIDTH graph).

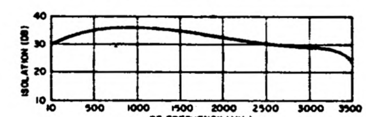
TYPICAL PERFORMANCE

Noise Figure:	Within 1 db of conversion loss
1 db Compression Level:	+7 dbm (with +10 dbm LO)
1 db Desensitization Level:	+5 dbm (with +10 dbm LO)
Two-Tone IM Ratio	95 db with LO = +10 dbm
3rd order (with two -30 dbm RF signals):	85 db with LO = +20 dbm
Third Order Intercept Point:	+17.5 dbm with +10 dbm LO
	+22.5 dbm with +20 dbm LO
Environmental:	Designed to meet MIL-E-5400 and MIL-STD-202

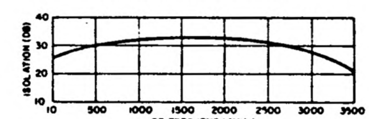
TYPICAL PERFORMANCE



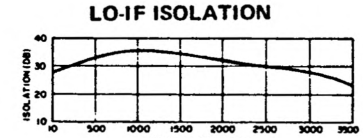
CONVERSION LOSS



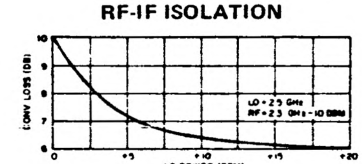
LO-RF ISOLATION



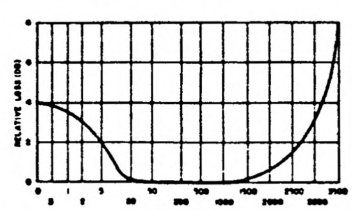
LO-IF ISOLATION



RF-IF ISOLATION



CONVERSION LOSS/LO POWER



IF PORT BANDWIDTH

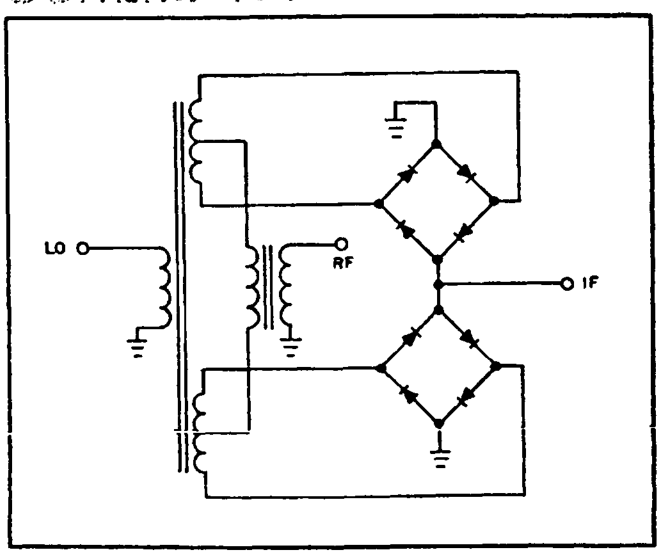
DESCRIPTION

The Model MD-123 is a high performance, subminiature double-balanced mixer using well-matched, low noise hot carrier diodes. The L and R ports have a bandwidth of 10 - 3000 MHz while the X ports has a bandwidth of DC to 3000 MHz. Inputs to any two ports will produce the sum and difference frequencies at the third port with a minimum of undesired modulation products. The double-balanced mixer may be used as a frequency converter, phase detector, double sideband suppressed carrier modulator, pulse modulator, frequency doubler, or a voltage/current variable attenuator.

Advanced Anzac broadband ferrite transformer techniques achieve a typical mixer noise figure of less than 7 db and an LO-RF isolation of better than 30 db up to 1 GHz. Guaranteed performance is achieved with an LO input power of +10 dbm, although the mixer may be used with LO inputs ranging from +7 to +20 dbm. Precision balanced circuits provide two-tone third order IM ratios of better than 85 db with -30 dbm input tones.

The ultra-low profile flat-pack, measuring only .125" high, allows much denser electronic equipment packaging. The hermetically sealed, metal package provides RFI shielding; and the rugged case and internal construction will withstand severe environmental conditions. The flat-pack will accommodate convenient stripline or dual in-line mounting, and the leads may be soldered or welded.

SCHEMATIC

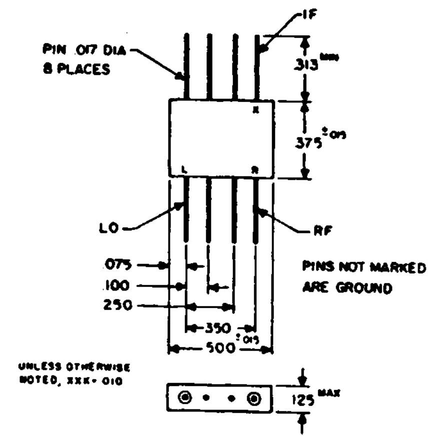


ORDERING INFORMATION

Please specify Model No. when ordering.

Model MD-123:	$175.00 (1-5 Qty.)
Availability:	Stock to 3 weeks
Terms:	Net 30, f.o.b. factory

MECHANICAL DATA



Mounting Area	.19 sq in	1.2 sq cm
Volume	.025 cu in	.39 cu cm
Weight (approx.)	.09 ounces	2.5 grams

Case Material:	Cold Rolled Steel
Finish:	Case - Gold electroplated per MIL-G-45204 type I, Class I Cover - Cold Rolled Steel plated with electroless nickel per MIL-C-26074B Class I, 0.00012 inch thick
Leads (Weldable and Solderable):	Thru leads — Dumet, type D per MIL-STD-1276B Ground Leads — No. 52 alloy, type F per MIL-STD-1276B Gold Plated per MIL-G-45204, type I, Class I

TYPICAL SPURIOUS RESPONSE CHART[1]

HARMONICS OF F_{rf}		F_{LO}	$2F_{LO}$	$3F_{LO}$	$4F_{LO}$	$5F_{LO}$	$6F_{LO}$	$7F_{LO}$	$8F_{LO}$
$7F_{rf}$	99	95	100	90	95	99	100	85	95
$6F_{rf}$	97	90	95	90	97	95	100	90	100
$5F_{rf}$	85	80	85	70	90	67	92	65	85
$4F_{rf}$	85	80	95	85	95	85	95	85	90
$3F_{rf}$	70	60	65	70	70	55	75	55	75
$2F_{rf}$	60	70	70	65	80	60	73	70	98
F_{rf}	30	0	33	13	35	35	42	35	45
	$F_{rf} \pm F_{LO}$	35	40	46	54	54	55	56	55

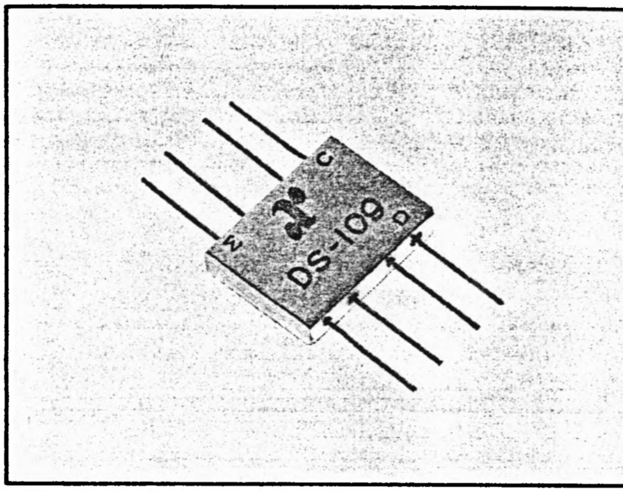
Spurious responses caused by internal harmonic generation and mixing of the input signals are shown above. The mixing products are referenced in dB below the desired $F_{LO} \pm F_{rf}$ output or 0 level at F_{if}. This performance can be typically attained with F_{LO} and F_{rf} at approximately 100 MHz, F_{LO} at + 10 dbm (10 mW) and F_{rf} at −10 dbM (0.1 mW) using broadband resistive terminations at all ports.

[1] RF harmonic referenced to RF input signal — LO harmonic referenced to LO input signal.

Printed in U.S.A.

FLATPACK IN-PHASE (ISO-T) 2-WAY POWER DIVIDER

10 - 500 MHz



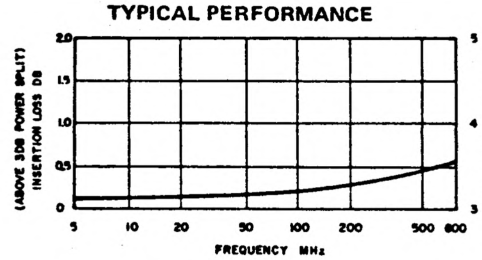
FEATURES

- Low Profile — Only 0.125" High
- Metal-Housed, RFI Shielded
- Rugged — Designed to meet MIL-E-5400
- 10-500 MHz Frequency Range
- Hermetically Sealed Case
- Low Cost

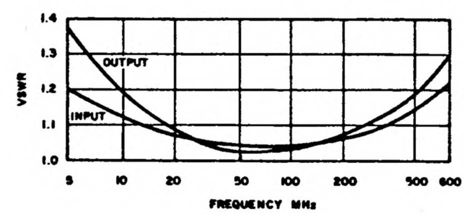
GUARANTEED SPECIFICATIONS

Frequency Range:	10-500 MHz
Insertion Loss: (above 3 db split)	0.5 db Max.
Isolation - Ports C to D:	
10-25 MHz	25 db Min.
25-500 MHz	30 db Min.
VSWR:	1.3 : 1 Max.
Impedance:	50 ohms (all ports)
Amplitude Balance:	0.1 db Max.
Phase Balance:	1.0° Max.
Matched Power Rating:	1 Watt Max.
Internal Load Dissipation:	0.05 Watt Max.
Operating Temperature Range:	-55°C to +85°C

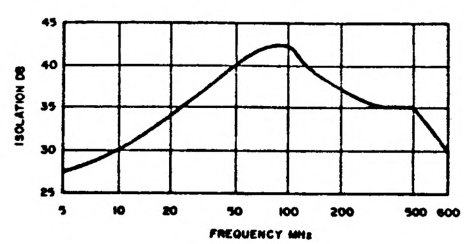
TYPICAL PERFORMANCE



INSERTION LOSS



VSWR



ISOLATION

DESCRIPTION

The Model DS-109 ISO-T 2-way power divider is a 10-500 MHz subminiature hybrid junction with the delta port internally terminated. Signals fed into the input (sum) port are equally divided, in phase, to the two output (C and D) ports. Conversely, signals fed into the side (C and D) ports are vectorially added at the sum port

The ultra-low profile flatpack (only 0.125" high) provides improved performance with increased reliability, and very importantly, is ideal for high-density electronic circuit packaging. Metal-housed, hermetically sealed, and RFI shielded, the DS-109 is ruggedly constructed to withstand severe environmental conditions, while at the same time lending itself to convenient stripline or microstrip mounting. The leads of this device can be readily soldered or welded.

ENVIRONMENTAL

This Device Has Been Designed to Meet the Following Environmental and Physical Conditions of MIL-STD-202:

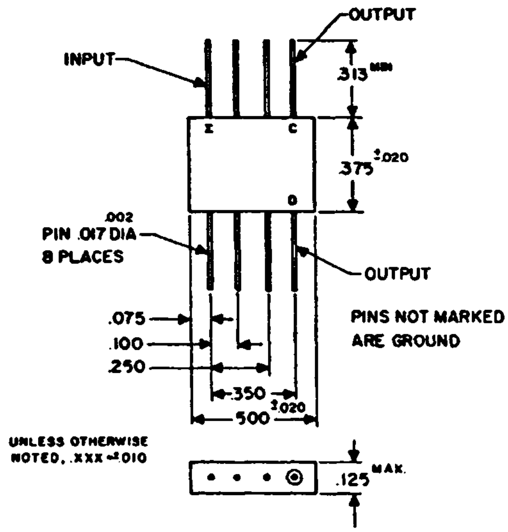
Thermal Shock:	Method 107, Test Condition A -55°C to 85°C, 30 minutes at temperature extremes, 5 cycles
Humidity:	Method 103, Test Condition B (96 hours)
Barometric Pressure:	Method 105, Test Condition D 100,000 feet
Moisture Resistance:	Method 106
Life Test:	Method 108, Test Condition B (250 hours)
Seal Test:	Method 112, Test Condition B (Gross Leak, 10^{-5} atm cc/sec.)
Vibration:	Method 204, Test Condition B 10-2,000 Hz, 15 G peak
High Impact Shock:	Method 207
Solderability:	Method 208
Other Applicable Specifications:	MIL-E-5400

ORDERING INFORMATION

Please specify Model No. when ordering.

Model DS-109:	$19.00 (1-50 Qty.)
Availability:	Stock
Terms:	Net 30, f.o.b. factory

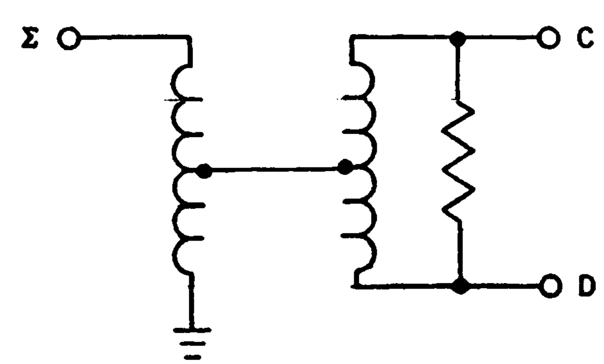
MECHANICAL DATA



Mounting Area:	.19 sq.in.	1.2 sq.cm.
Volume:	.025 cu.in.	.39 cu.cm.
Weight (approx.)	.09 ounces	2.5 grams

Case Material:	Cold Rolled Steel
Finish:	Case - Gold electroplated per MIL-G-45204 type I, Class I Cover - Cold Rolled Steel plated with electroless nickel per MIL-C-26074B Class I, 0.00012 inch thick
Leads (Weldable and Solderable):	Thru leads — Dumet, type D per MIL-STD-1276B Ground Leads — No. 52 alloy, type F per MIL-STD-1276B Gold Plated per MIL-G-45204, type I, Class I

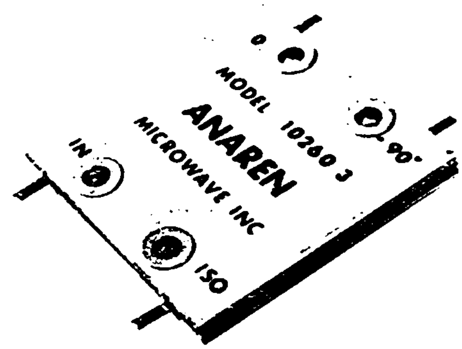
SCHEMATIC



Printed in U.S.A.

Anaren

Hybrid Couplers
3 dB, 90°
Ultra-Miniature



Features

For installation directly in microstrip circuits
Largest selection of frequency ranges and case styles
Lowest cost
Smallest size, lightest weight
Laminated stripline construction
High isolation with low VSWR
Rugged aluminum cases
Meets MIL-E-5400 Class 3 requirements
Custom designs available

Applications

Inexpensive power dividers and combiners
Low cost balanced amplifier designs
Matrix amplifiers
Voltage variable PIN diode attenuators
Balanced mixers and modulators
Switching networks
Balanced detectors
Antenna feed networks
Phase shifters and comparators

Description

This ultra-miniature series of 3 dB, 90° hybrid couplers is available in more than 40 standard models and 17 case styles to cover the frequency range 30 MHz to 4.2 GHz. The popular 225-400 MHz band is covered by 6 standard case styles. Custom designs are available in all frequency ranges.

Anaren ultra-miniature, 3 dB, 90° hybrid couplers are well suited for a variety of applications: power ~~dividers and~~ combiners; balanced amplifier circuits with low input and output VSWR; matrix amplifiers; voltage variable PIN diode attenuators; balanced mixers and modulators; switching networks; balanced detectors; antenna feed networks; phase shifters and comparators.

There are a number of techniques available for constructing microwave quadrature hybrids, but Anaren uses the backward wave 3 dB, 90° hybrid coupler in stripline form. This stripline version is smaller, provides better performance and can cover wider bandwidths than other types. The single-section, backward wave, 3 dB hybrid allows octave bandwidth coverage and multi-section versions can easily be designed to cover multi-octave and decade bandwidths.

All Anaren ultra-miniature couplers are printed on stable teflon-glass substrates using using shielded stripline techniques. They are laminated under heat and pressure using a low loss dielectric bonding compound. The package assures high reliability and is capable of withstanding extreme environmental stress.

The couplers are designed to mate with 30 mil microstrip ~~and mounting holes are provided for attaching~~ the coupler ground plane surface to the mounting base ground plane. A reliable electrical ground contact is necessary for optimum performance.

The Anaren ultra-miniature coupler is a reciprocal four-port network.

An input signal applied to any port (port 1, for example) will divide equally to the two opposite ports (3 and 4) with port 2 remaining isolated. The voltage at port 4 lags the voltage at port 3 by 90°. This phase quadrature relationship is independent of frequency and is the unique property which makes the 90° coupler so versatile.

See pages 57, 83 for additional information.

Electrical Specifications

Model No.	Frequency (GHz)	Isolation Min/Typ (dB)	VSWR Max/Typ	Insertion Loss Max (dB)	Amplitude Balance Max (dB)	Phase Balance Max (deg)	Case Style
10230-3	.030-.076	17/20	1.20/1.10	0.35	±0.75	±1.5	102
10270-3	.030-.076	20/25	1.20/1.10	0.35	±0.75	±1.5	103
1A0270-3	.040-.080	20/25	1.20/1.10	0.25	±0.50	±1.5	103
1A0230-3	.054-.088	20/22	1.20/1.10	0.35	±0.50	±1.5	102
1A0262-3	.055-.110	20/22	1.20/1.10	0.25	±0.50	±1.5	102
10261-3	.0625-.125	20/27	1.20/1.10	0.35	±0.50	±1.5	104
1C0261-3	.070-.140	20/25	1.20/1.10	0.35	±0.50	±1.5	104
1A0280-3	.090-.180	20/27	1.20/1.10	0.25	±0.50	±1.5	104
1H0280-3	.090-.180	18/25	1.20/1.10	0.25	±0.50	±1.5	116
1B0261-3	.100-.200	20/25	1.20/1.10	0.25	±0.50	±1.5	104
1G0280-3	.100-.160	25/27	1.20/1.10	0.25	±0.30	±1.5	104
1J0280-3	.100-.160	20/27	1.20/1.10	0.25	±0.30	±1.5	116
10920-3	.100-.400	20/25	1.20/1.10	0.35	±0.45	±1.5	117
1A0920-3	.100-.500	16/20	1.35/1.15	0.50	±0.60	±2.0	117A
10280-3	.116-.150	20/27	1.20/1.10	0.25	±0.30	±1.5	104
10262-3	.125-.250	20/27	1.20/1.10	0.25	±0.50	±1.5	105
1J0263-3	.220-.460	20/25	1.20/1.10	0.25	±0.60	±1.5	106
10260-3	.225-.400	20/25	1.20/1.10	0.25	±0.50	±1.5	106
1A0260-3	.225-.400	20/25	1.20/1.10	0.25	±0.50	±1.5	107
1B0260-3	.225-.400	20/25	1.20/1.10	0.25	±0.50	±1.5	108
10360-3	.225-.400	17/22	1.25/1.15	0.25	±0.50	±2.0	109
1R0260-3	.225-.400	20/25	1.20/1.10	0.25	±0.50	±1.5	110
1S0260-3	.225-.400	20/25	1.20/1.10	0.25	±0.50	±1.5	106
1H0263-3	.250-.500	20/22	1.20/1.10	0.25	±0.50	±1.5	110
10263-3	.250-.500	20/25	1.20/1.10	0.25	±0.50	±1.5	106
10223-3	.250-1.0	20/23	1.30/1.15	0.50	±0.75	±1.5	101
1D0263-3	.300-.550	20/25	1.20/1.10	0.25	±0.50	±1.5	106
1E0263-3	.350-.650	20/25	1.25/1.10	0.25	±0.50	±1.5	106
1A0263-3	.400-.600	20/25	1.20/1.10	0.25	±0.50	±1.5	106
1B0263-3	.400-.700	20/25	1.20/1.10	0.25	±0.50	±1.5	108
1H0264-3	.440-.880	20/25	1.20/1.10	0.25	±0.50	±1.5	111A
10264-3	.500-1.0	20/25	1.20/1.10	0.25	±0.50	±1.5	111
1B0264-3	.500-1.0	20/25	1.20/1.10	0.25	±0.50	±1.5	112
1A0264-3	.600-1.2	20/25	1.20/1.10	0.25	±0.50	±1.5	111
10330-3	.700-1.4	20/25	1.25/1.10	0.25	±0.50	±1.5	113
10890-3	.950-1.225	20/25	1.25/1.15	0.25	±0.30	±1.5	112
1D0265-3	.96-1.215	20/25	1.25/1.10	0.25	±0.30	±1.5	111
10265-3	1.0-2.0	20/24	1.25/1.10	0.25	±0.50	±1.5	113
1B0265-3	1.0-2.0	20/24	1.25/1.15	0.25	±0.50	±1.5	112
1E0320-3	1.3-2.6	20/25	1.30/1.15	0.25	±0.50	±2.0	113
10320-3	1.7-2.5	20/23	1.30/1.20	0.25	±0.50	±1.5	113
1B0320-3	1.7-2.5	20/25	1.30/1.15	0.25	±0.50	±1.5	112
1B0266-3	1.7-3.4	18/21	1.30/1.20	0.25	±0.50	±1.5	114
10266-3	2.0-4.0	18/21	1.30/1.20	0.25	±0.50	±2.0	114
1A0266-3	2.1-4.2	18/21	1.30/1.20	0.25	±0.50	±2.0	114

Power: 200 W average (except 10230-3 is 100 W) 1 kW peak.

Power rating applies when solder tab/coupler interface has been conformally coated to eliminate voltage breakdown

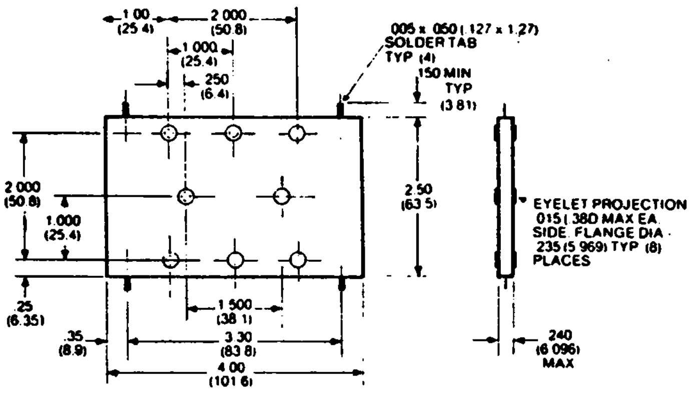
Nominal Impedance: 50 ohms, non-reactive.

These specifications apply for tests performed with properly designed microstrip test fixtures.

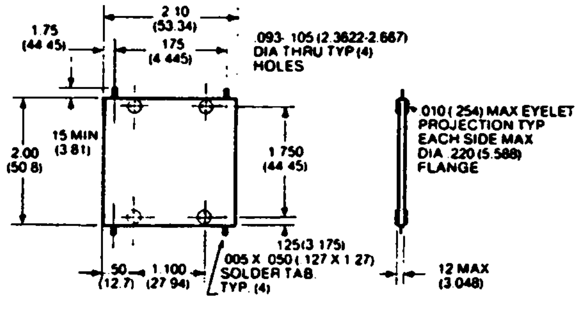
Couplers meet environmental requirements of MIL-E-5400, Class 3; as applicable.

Specifications subject to change without notice

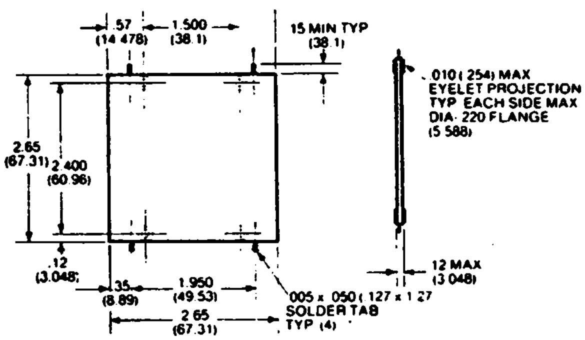
Mechanical Specifications



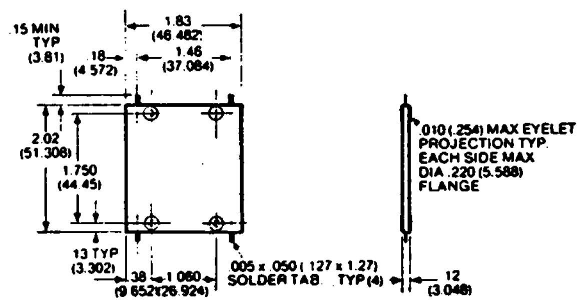
Case style 101 WEIGHT: 3.3 (94)



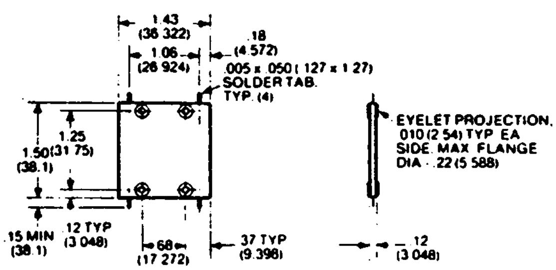
Case style 102 WEIGHT: .69 (19)



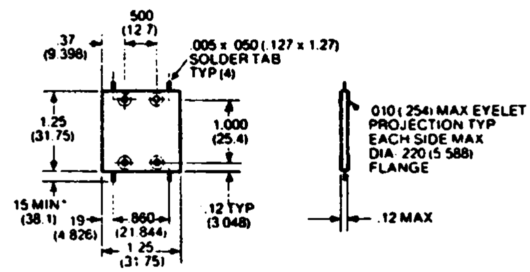
Case style 103 WEIGHT: 1.16 (33)



Case style 104 WEIGHT: .61 (17)

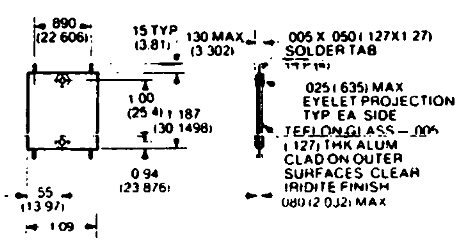


Case style 105 WEIGHT .35 (10)

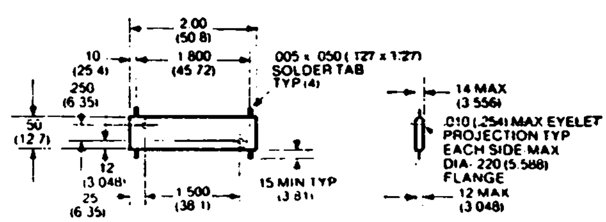


*Except 1S0260-3 is 0.50 (12.7)

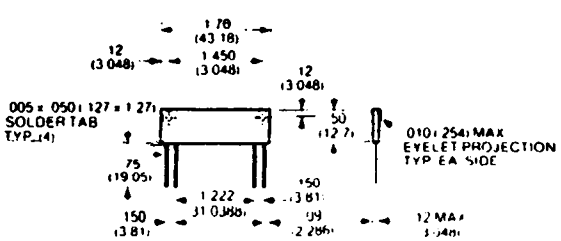
Case style 106 WEIGHT .10 (3)



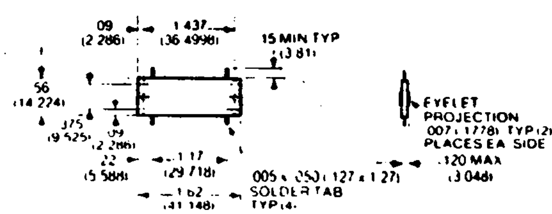
Case style 107 WEIGHT .17 (5)



Case style 108 WEIGHT .17 (5)



Case style 109 WEIGHT .14 (4)

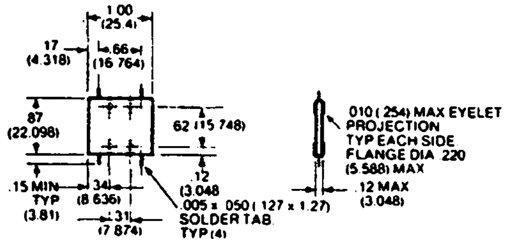


Case style 110 WEIGHT .11 (4)

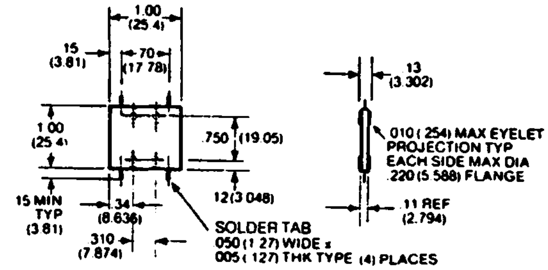
Contact Anaren for latest outline details.

Specifications subject to change without notice

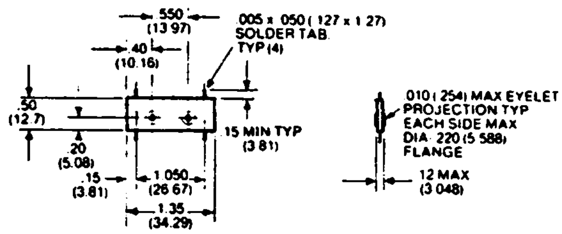
Mechanical Specifications (continued)



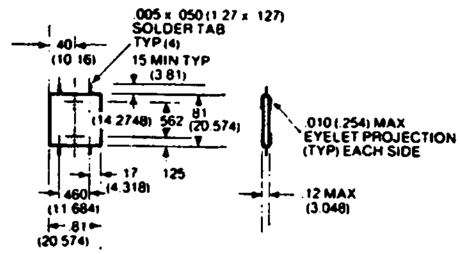
Case style 111 WEIGHT: .87 (25)



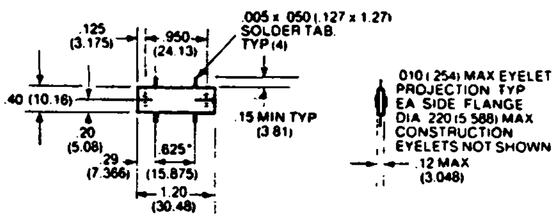
Case style 111A WEIGHT: .17 (5)



Case style 112 WEIGHT: .11 (3)

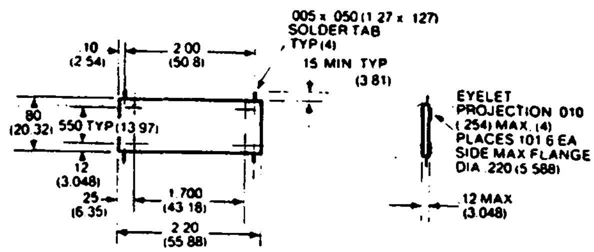


Case style 113 WEIGHT: .11 (3)

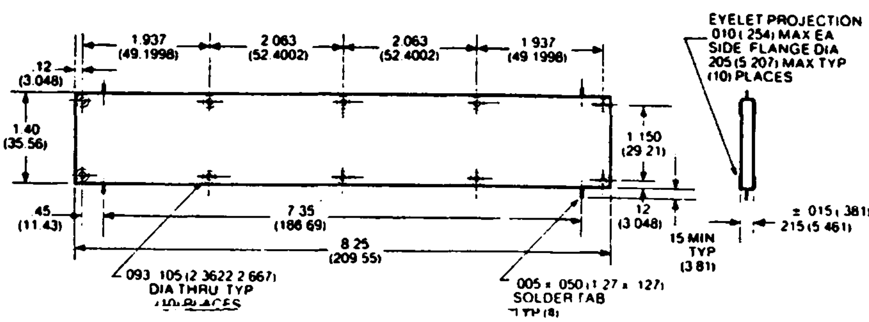


*Except 1A0266-3 is .595 (15.133)

Case style 114 WEIGHT: .06 (2)



Case style 116 WEIGHT: .26 (7)



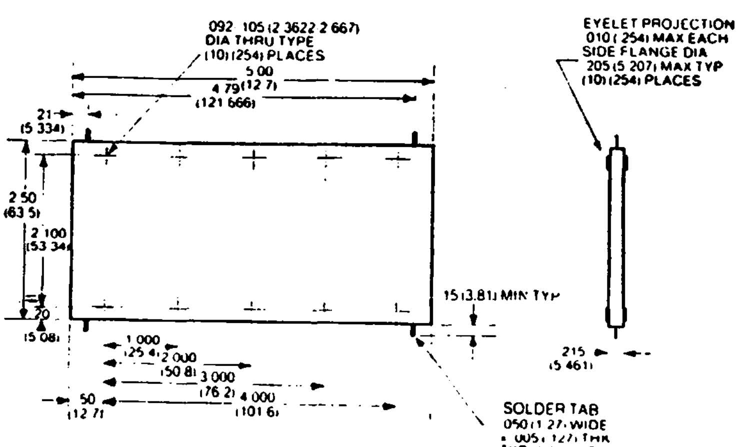
Case style 117 WEIGHT: 3.8 (108)

Contact Anaren for latest outline details.

Dimensions are in inches and (mm.)
~Weights are in oz. and (gr.)

Case material: 6061-T6 Aluminum (except 1A0260-3 is .005 (.127) thick alum. clad)
Solder Tabs: Silver plated copper (solderability per MIL-STD-20, Method 208)
Marking per MIL-STD-130
Finish: Clear IRIDITE per MIL-C-5541A, type 1, Grade C Class III
Mounting holes are 0.099 ± .006 (2.5 ± .15) dia

Specifications subject to change without notice



Case style 117A WEIGHT: 4.1 (116)

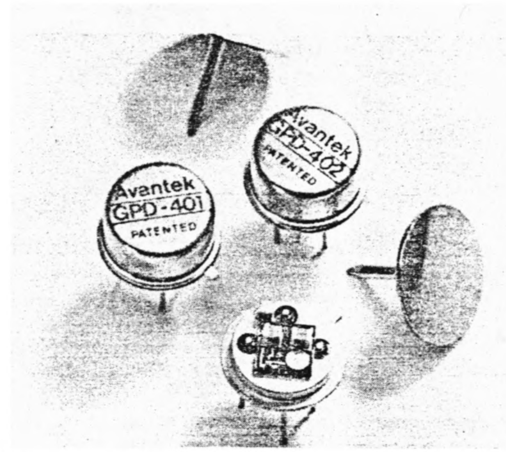
Avantek

DATA SHEET

GPD-400 Series
Patented*

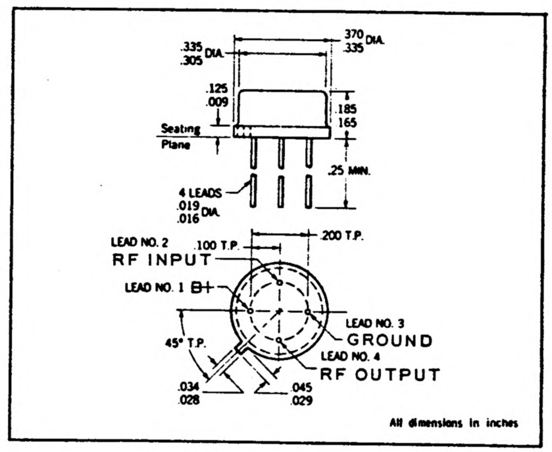
Thin Film Amplifier
5 to 400 MHz

Miniature Transistor Amplifier



FEATURES

- Low Cost
- Cascadable
- Low Profile TO-12 (4-leaded TO-5) Package
- Thin Film Sapphire Construction
- Over 6 Octaves of Amplifier Bandwidth
- Avantek Silicon Transistor Chips



DESCRIPTION

The GPD is a complete transistor amplifier, ready to operate in a microstrip circuit upon application of DC voltage. Packaged in a miniature TO-12 transistor package, the Avantek GPD serves as a completely cascadable amplifier, without bandwidth shrinkage, from 5 to 400 MHz. The low frequency response of the GPD-460 series may be set arbitrarily low by selection of external series input and output capacitors, and the DC bypass capacitor.

The Avantek GPD is an entirely new kind of basic device, designed to provide the circuit engineer major savings in both time and money. Various gain and power output choices are available to permit the user to cascade modules to meet the performance characteristics required in his equipment design. Small size, excellent performance, ready availability and substantial cost savings in equipment manufacture and parts handling are significant advantages that can be gained over standard discrete component methods of manufacture by the use of GPD amplifiers. The costly and time-consuming problems accompanying in-house amplifier design, construction and testing can be totally avoided by inserting GPD's, either singly or cascaded, into a system circuit.

The Avantek GPD is a wideband, single-stage unit of gain, featuring flat response across its greater-than-six-octave bandwidth. The tiny GPD modular amplifier is made with highly reliable sapphire substrates, Avantek microwave transistor chips, thin film circuits, thin film resistors and chip capacitors. All the complex circuitry is encapsulated inside the tiny TO-12 package. The using engineer is spared the normal frustrating RF design problems — impedance matching networks, feedback loops, biasing and stabilization elements.

APPLICATIONS:

The GPD-400 Series amplifier is designed for applications requiring very broadband amplifiers, preamplifiers, isolation amplifiers, and IF amplifiers. The patented circuit design of the GPD permits cascading of units to achieve gain up to any desired level without interstage matching when cascaded in 50-ohm systems. The specified band edges (5 to 400 MHz) are not 3 dB points, but are the points between which the specified gain performance is guaranteed. The low frequency response of the GPD-460 units may be set as close to DC as required (but not DC, for DC response see the UTD-561).

*U.S. Patent 3493882

INSTALLATION AND OPERATING INSTRUCTIONS:

Installation of the GPD amplifier is similar to the installation of any standard semi-conductor product in a TO-8 or TO-5 package. A clamp is provided to secure the GPD firmly to the ground plane. This step insures positive contact between the GPD package and the ground plane so that no problems with VSWR or oscillation in a multi-stage system will be encountered.

The GPD amplifier is designed for use in a 50-ohm microstrip system. It can be used in other impedance systems, but performance may be degraded.

The microwave transistor used in the GPD must be protected from current surges which may be generated by energy storage in system capacitances. Always remove bias voltages from the GPD before inserting or removing the unit under test.

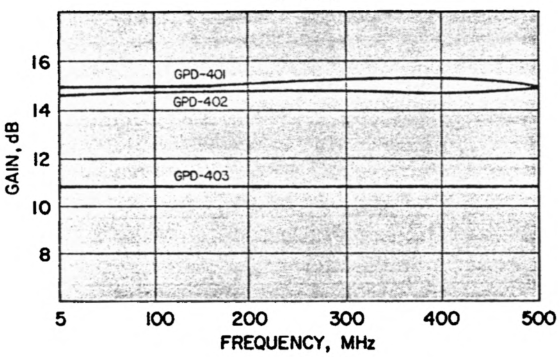
The use of a high-pass filter and/or pad is recommended at the output of gas-discharge-tube noise sources. This protects the transistor in the amplifier from possible high-level ignition-pulse transients which may appear at the RF output ports of these generators (see appropriate manufacturer's literature for further details).

The amplifiers may be stored at temperatures from -65°C to +200°C. The transistors are silicon and all metallization is gold. The operating case temperature is specified at +71°C (+160°F). The amplifiers will operate reliably at temperatures through +125°C (+257°F) although an external heat sink should be used, particularly on the GPD-403.

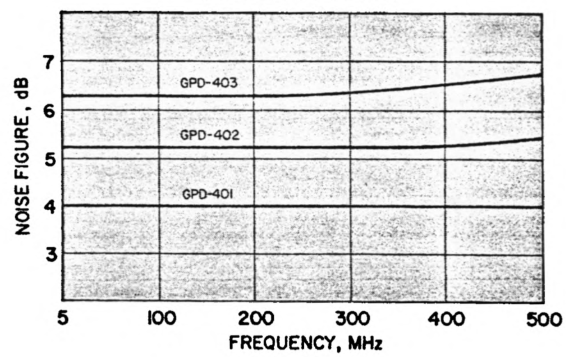
More information concerning applications and use of the GPD amplifier is available from Avantek. Write for the Applications Bulletin *Designing With GPD Amplifiers.*

TYPICAL PERFORMANCE

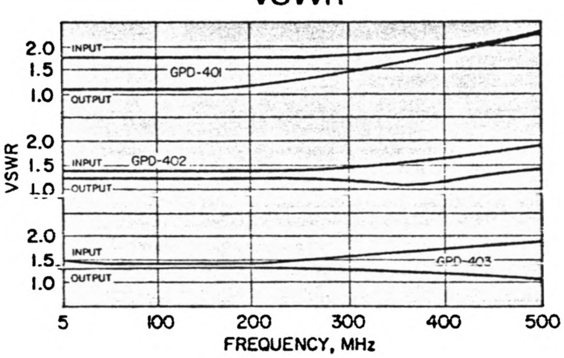
GAIN



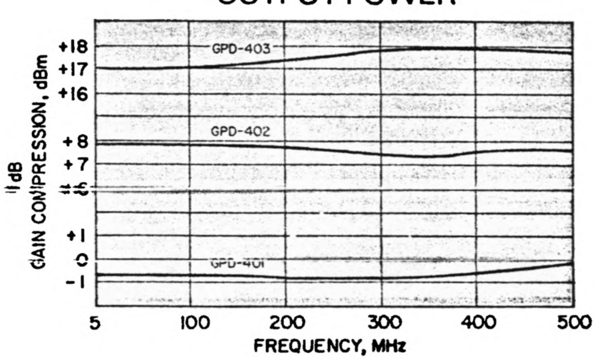
NOISE FIGURE



VSWR



OUTPUT POWER

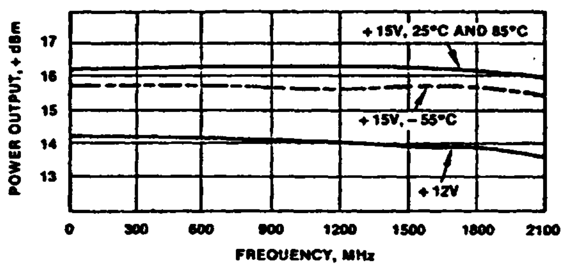


GUARANTEED SPECIFICATIONS

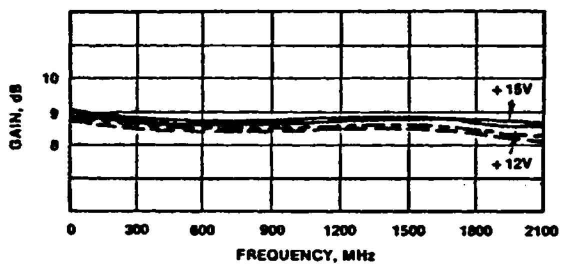
Model	Frequency Response (MHz) Minimum	Gain (dB) Minimum	Flatness (dB) Typical	Noise Figure (dB) Typical	Reverse Isolation (dB) Typical	Power Output for 1 dB Gain Compression (dBm) Typical	Avantek Intercept Point for IM Products (dBm) Typical	VSWR (50 - ohms) Typical In	Out	Input Power Volts DC	Current (mA) Typical	Storage Temperature (°C)	Weight (grams)
GPD 401	5-400	13	1.0	4.5	-20	-2	+8	2.0	2.0	15	10	-65 to +200	1.0
GPD 461	Same as GPD 401, except three external capacitors are required to establish low frequency roll-off.												
GPD 402	5-400	13	1.0	6.0	-20	+6	+18	2.0	2.0	15	24	-65 to +200	1.0
GPD-462	Same as GPD-402, except three external capacitors are required to establish low frequency roll-off.												
GPD-403	5-400	9	1.0	7.5	-20	+15	+26	2.0	2.0	24	65	-65 to +200	1.0
GPD-463	Same as GPD-403, except three external capacitors are required to establish low frequency roll off												

Typical Performance — UTO-2023

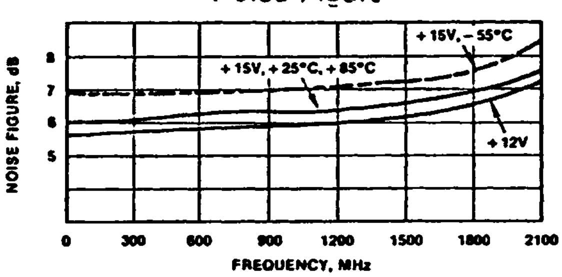
Power Output



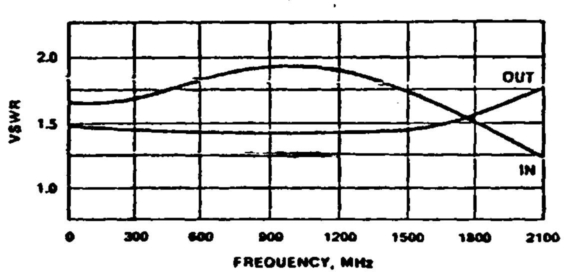
Gain



Noise Figure



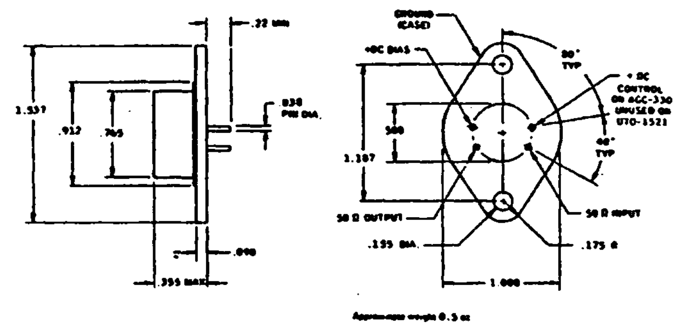
VSWR



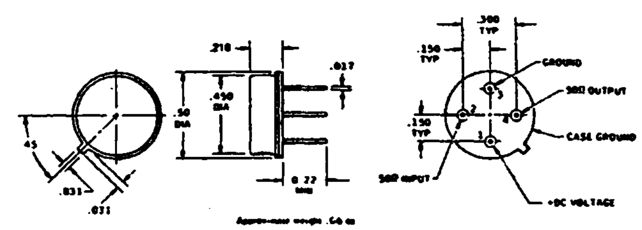
Case Drawings

Unless otherwise specified all dimensions in inches.
Tolerances: .XX = .02
.XXX = .01

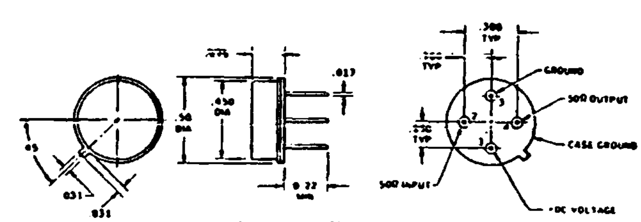
TO-3



TO-8T



TO-8U



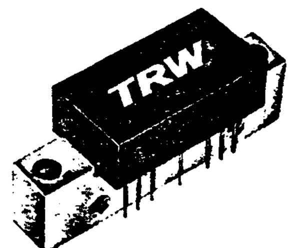
Key: +25°C ______ ; +85°C— — — —; −55°C—·—·—

TRW RF SEMICONDUCTORS

DATA SHEET NO. 582 (8/75) 7.5M

CA2810 Wide Bandwidth Linear Hybrid Amplifier

- Power Output, 800mW
- 33dB Gain
- 400mW PEP @ -32dB IMD
- Instantaneous Bandwidth, 10-350MHz
- Low Noise Figure, 4.5dB



CA

The CA2810 is a high-reliability thin-film hybrid amplifier utilizing an all gold metalization system. Units are designed for wide bandwidth linear operation in 50 to 100 ohm systems. This hybrid provides excellent gain stability with temperature and very low distortion due to push-pull amplifier circuitry. This module is recommended for wide bandwidth, low noise and linear applications.

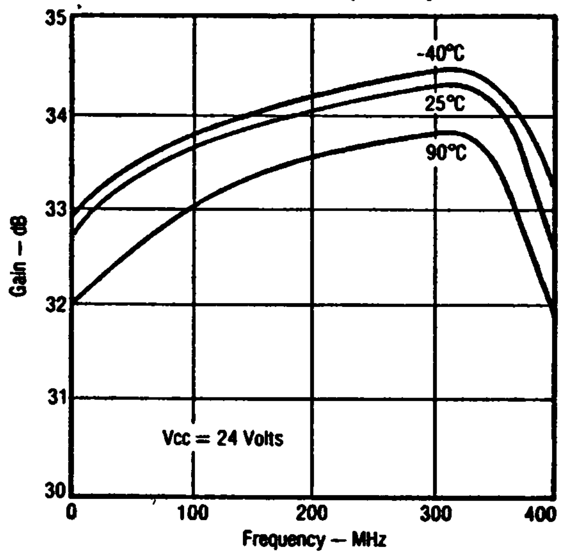
Absolute Maximum Ratings

Vcc	RF Power Input	Storage Temperature	Operating Temperature
28 Volts	+5dBm	-40°C to +100°C	-20°C to +90°C

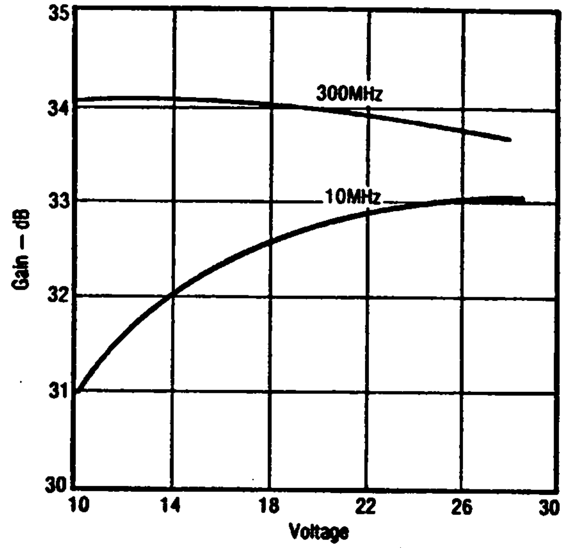
Electrical Characteristics for 50Ω Systems (T_{CASE} = 25°C and 24V)

Symbol	Characteristic	Conditions	Value
P_G	Power Gain	f = 50MHz	33 ±1dB
NF	Noise Figure, Broadband	f = 60MHz f = 300MHz	4.5dB Typ 8.0dB Max
I_{TO}	Third Order Intercept, See Figure 1	f_1 = 300MHz	+43dBm Typ
VSWR	Input/Output VSWR for 50Ω Systems	f = 10-350MHz	2:1 Typ
I_{CC}	Supply Current	24V	330mA Max
P_o	Power Output – 1dB Compression	f = 200MHz	800mW
P_{RI}	Reverse Isolation	f = 10-350MHz	40dB Typ
F_R	Frequency Response	f = 30-300MHz f = 10-350MHz	±1.0dB Max ±1.5dB Max
d_{SO}	Second Harmonic Distortion	~~Tone at 10mW~~ f_{2H} = 10-300MHz	-66dB Typ
PEP	Peak Envelope Power for Two Tone Distortion Test See Figure 1	f = 10-300MHz at -32dB	400mW Typ

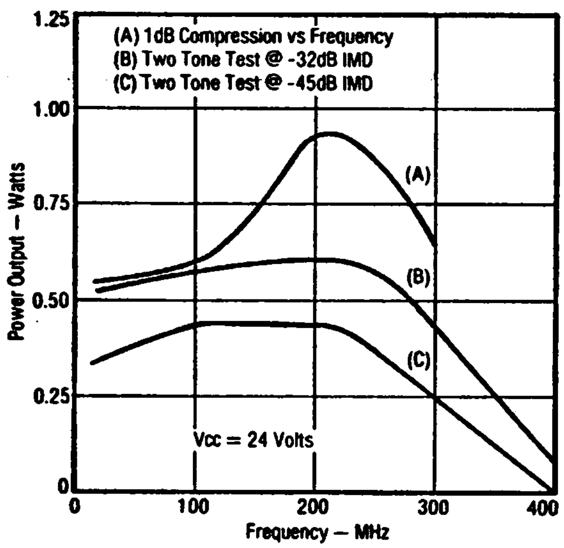
Gain vs Frequency



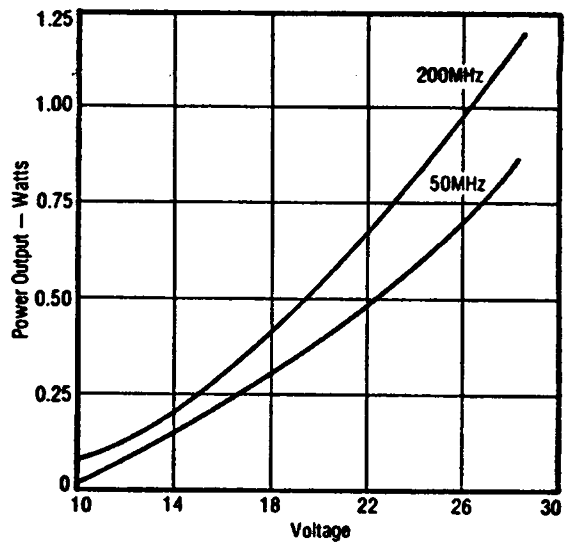
Gain vs Voltage



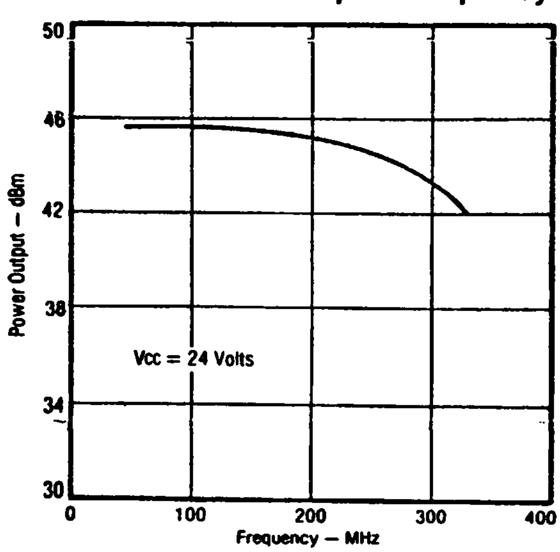
Power Output vs Frequency



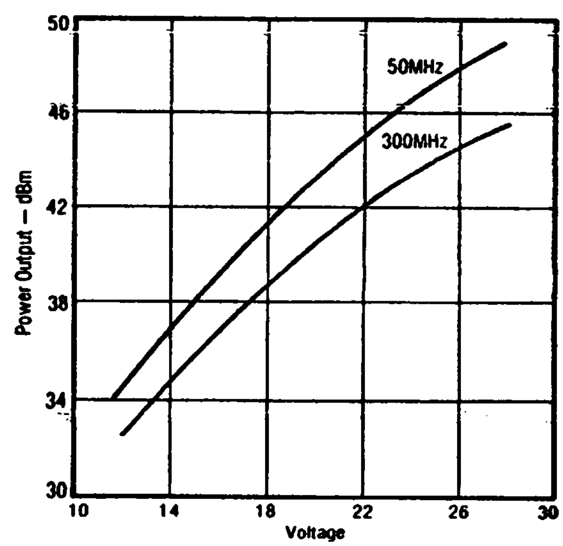
1dB Compression vs Voltage



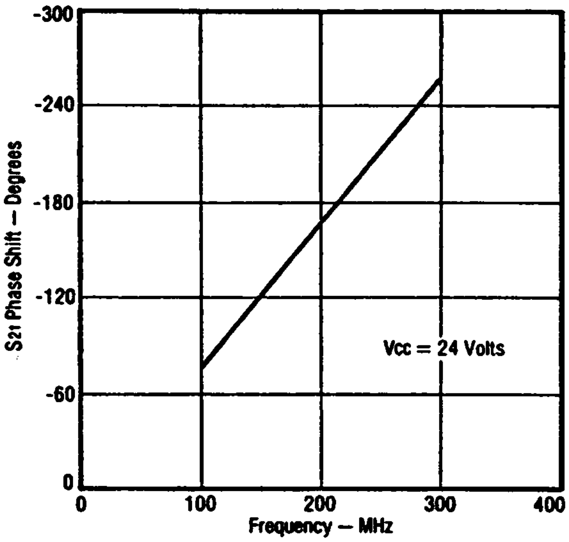
Third Order Intercept vs Frequency



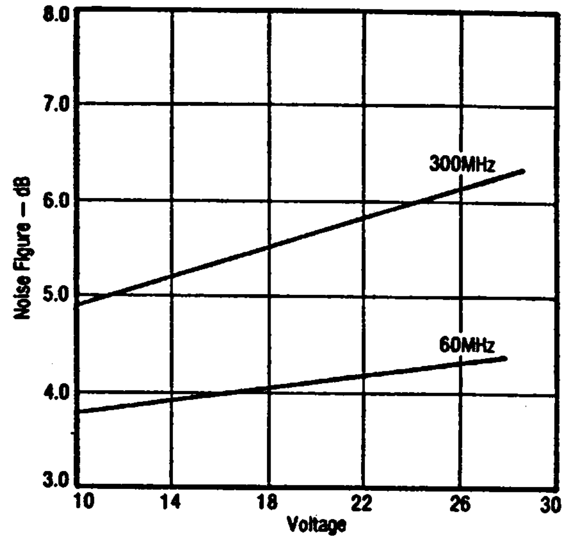
Third Order Intercept vs Voltage



Phase Shift vs Frequency

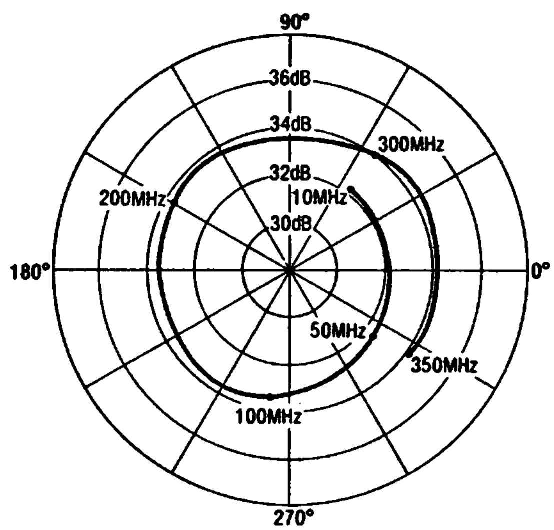


Noise Figure vs Voltage

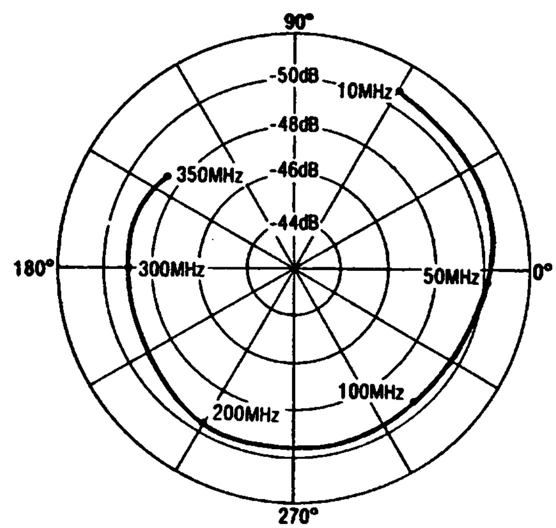


S-Parameters
Vcc = 24V, Zo = 50Ω

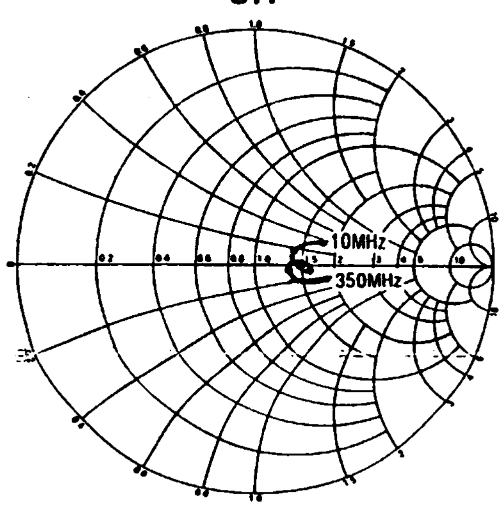
S21



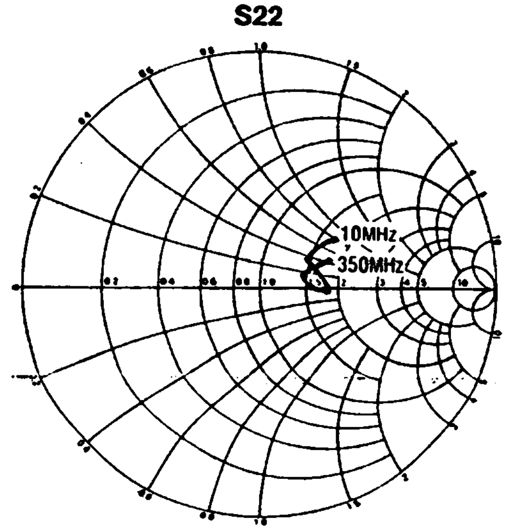
S12



S11



S22



CA Package Outline

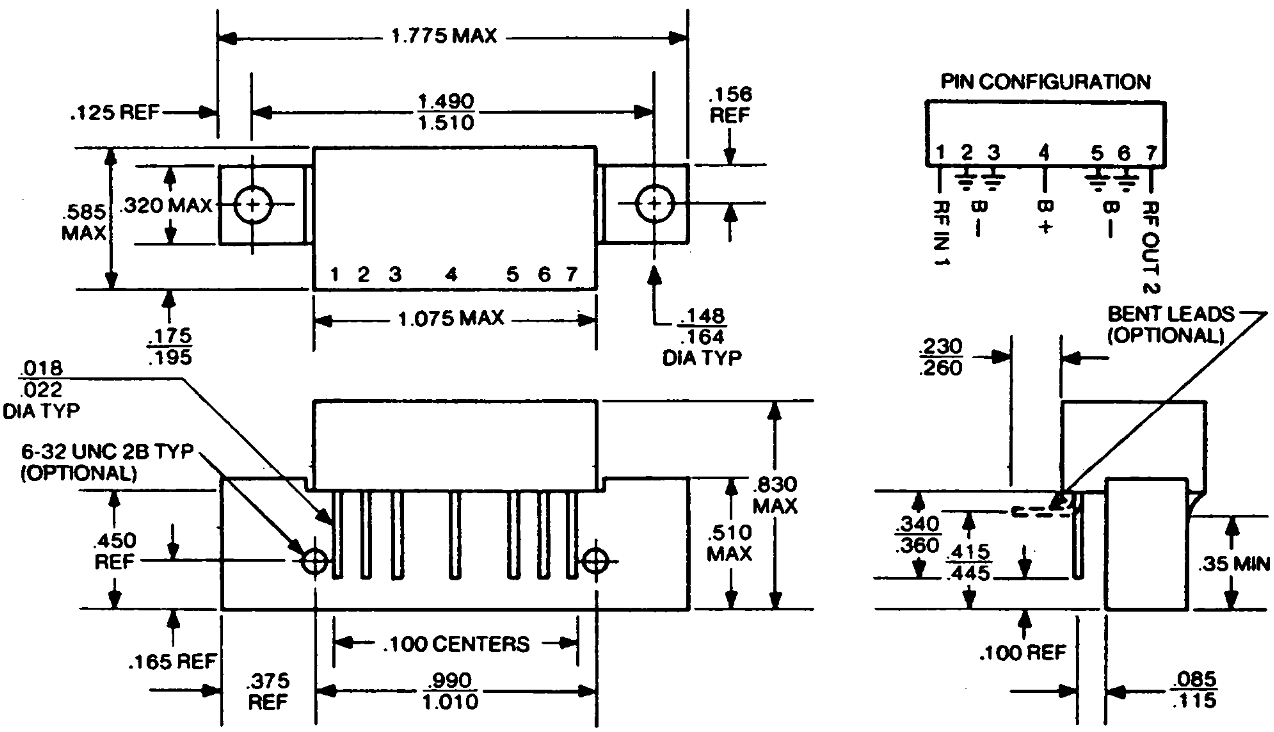
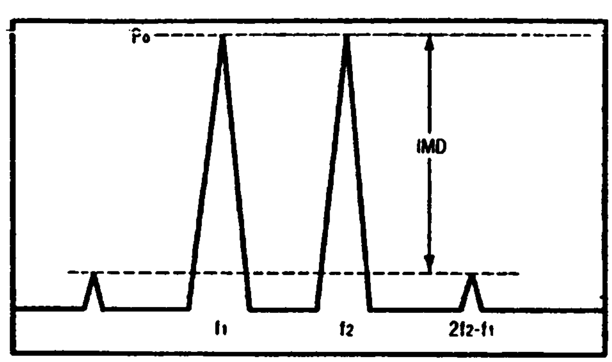


Figure 1. Intermodulation Test

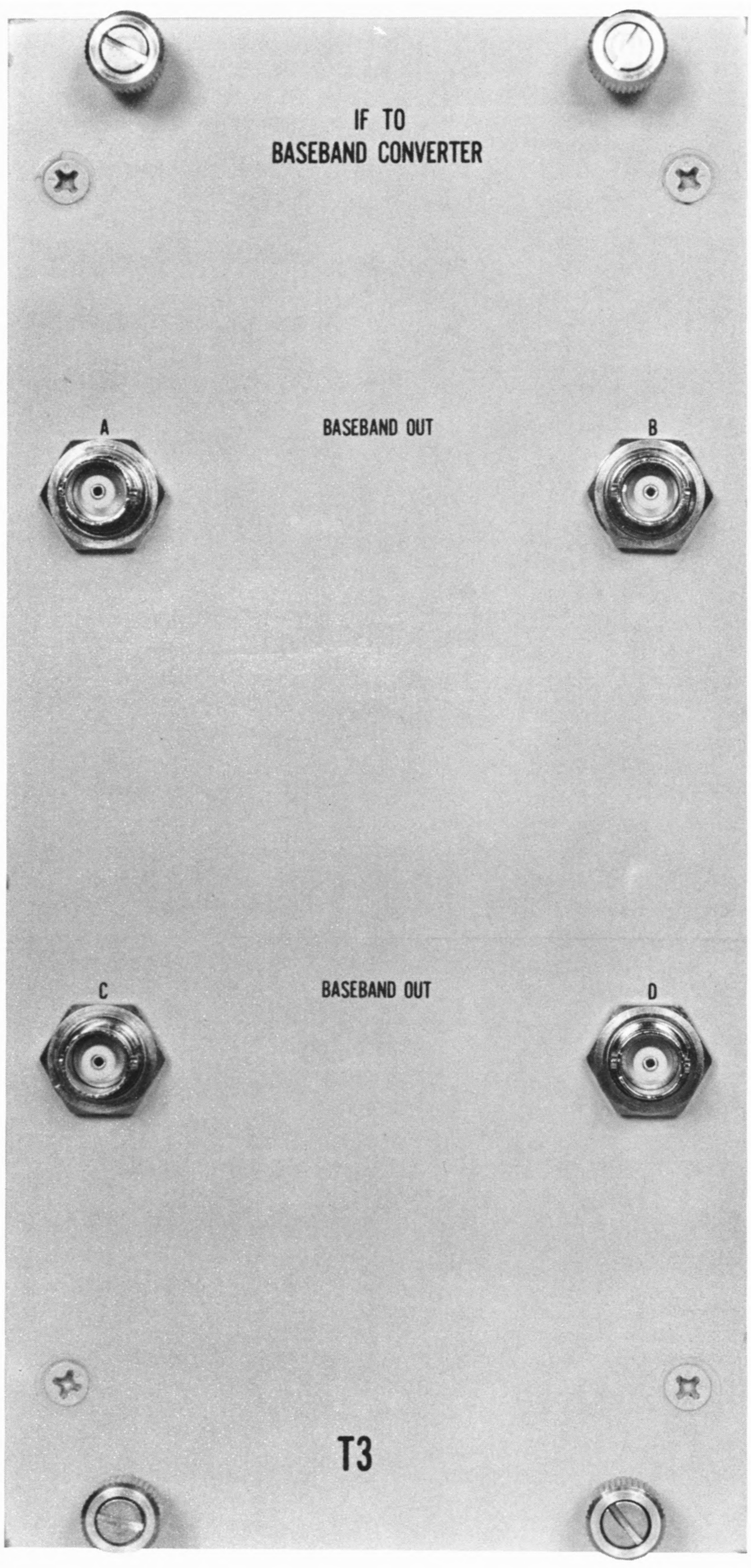


$$ITO = Po + \frac{IMD}{2} \text{ @ } IMD > 60dB$$

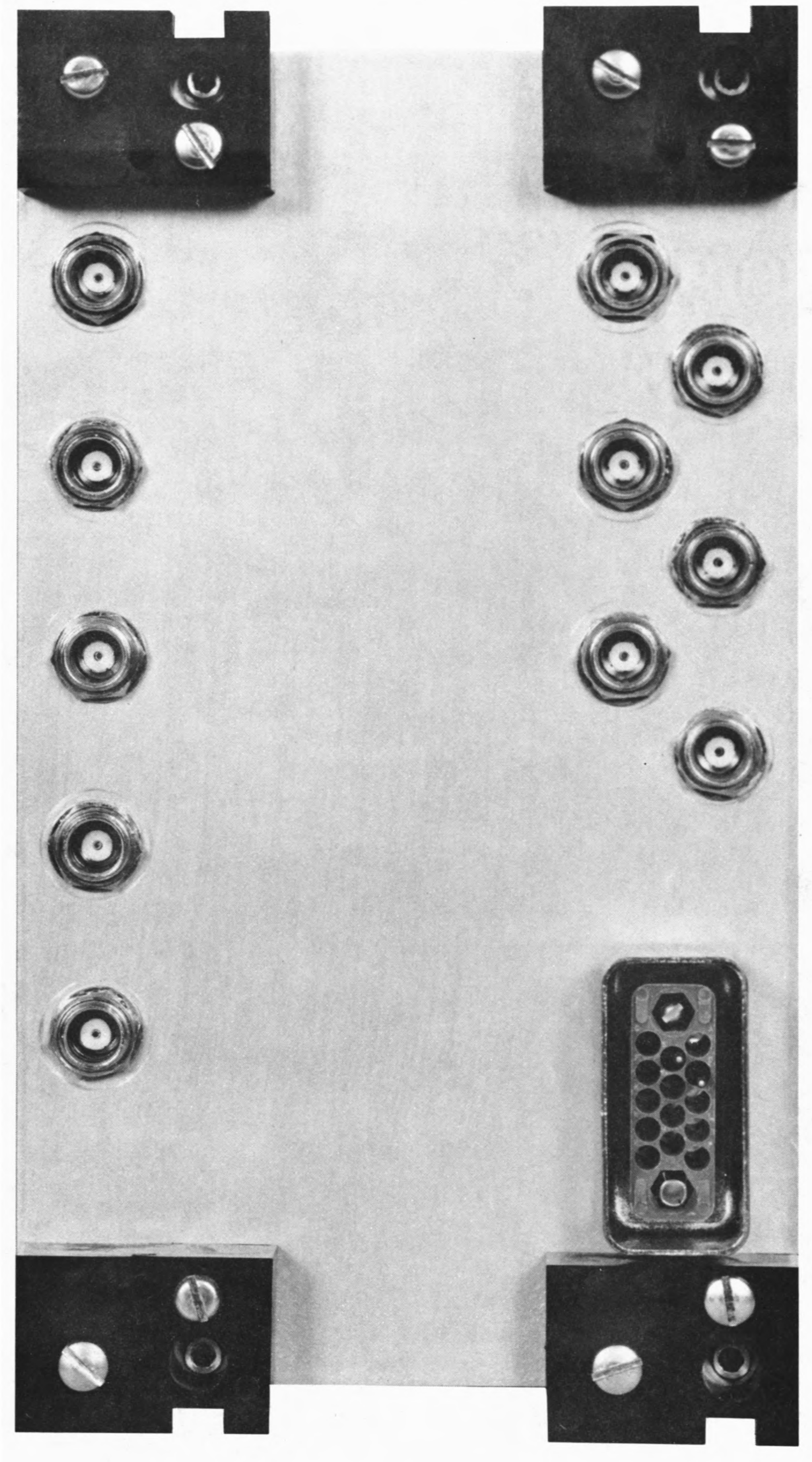
$$PEP = Po \text{ @ } IMD = -32dB$$

PRINTED IN U S A

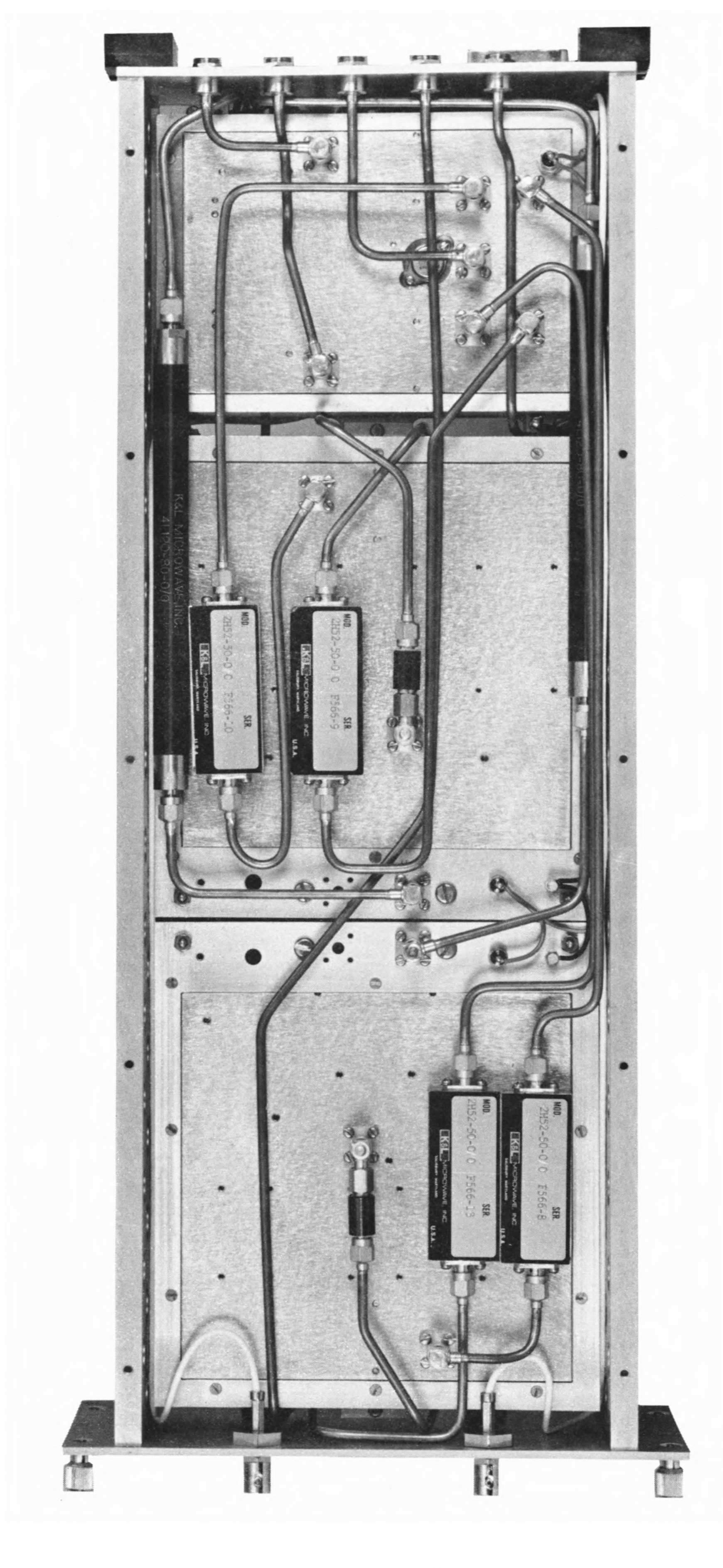
8.0 PHOTOGRAPHS OF A TYPICAL MODULE



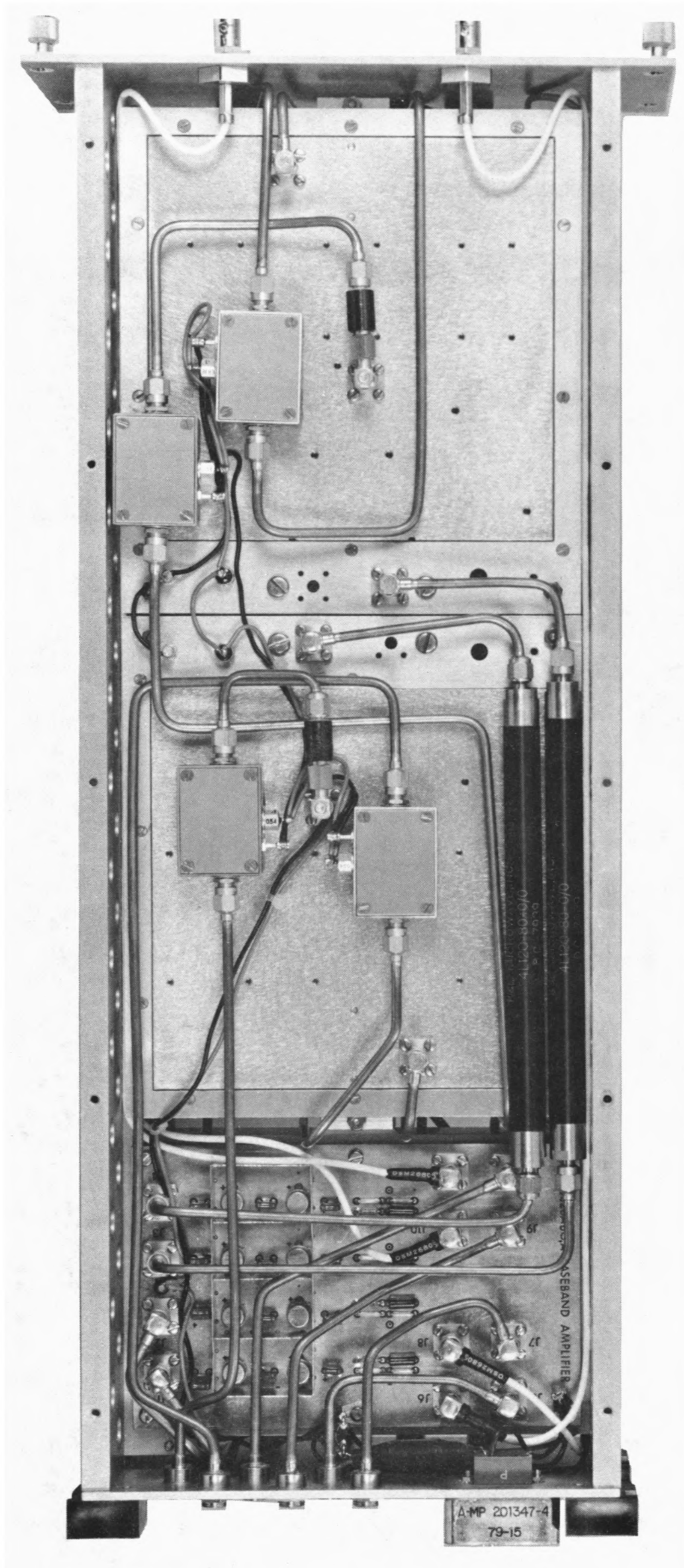
FRONT VIEW



REAR VIEW



RIGHT SIDE VIEW



LEFT SIDE VIEW

APPENDIX A

SUBMITTED TO IEEE FOR PUBLICATION DECEMBER 1979

A BROADBAND UHF MIXER EXHIBITING HIGH IMAGE REJECTION OVER A MULTIDECADE BASEBAND FREQUENCY RANGE

J. W. ARCHER

National Radio Astronomy Observatory*
Socorro, New Mexico

J. GRANLUND

National Radio Astronomy Observatory*
Charlottesville, Virginia

R. MAUZY

National Radio Astronomy Observatory*
Green Bank, West Virginia

ABSTRACT

Image reject frequency converters operating over greater than octave RF and LO input bandwidths in the UHF frequency range are described. The devices, designed for RF to video conversion, exhibit very flat conversion loss response and greater than 34 dB rejection of the undesired sideband at any baseband frequency between 186 kHz and 50 MHz.

*The National Radio Astronomy Observatory is operated by Associated Universities, Incorporated, under contract with the National Science Foundation.

INTRODUCTION

The image rejection mixers described herein were designed for use in the IF-to-baseband conversion subsystem of the Very Large Array [Heeschen (1975)], a complex radio astronomical Fourier synthesis array being constructed in central New Mexico. Mixers which reject either the upper or lower sideband were required. This paper will discuss only the results for lower-frequency sideband rejection; the performance and design of the higher-frequency sideband rejection mixer is similar in most respects.

The mixers operate with local oscillator (LO) signals tunable over either of the ranges f_L = 100 to 150 MHz or 200 to 250 MHz with the desired RF input sideband extending from f_L to 150 MHz or f_L to 250 MHz, respectively. They were required to exhibit a conversion loss response, for the upper sideband, uniform to within 0.25 dB for any combination of RF and LO signal frequencies within the specified ranges, such that the baseband output signal frequency lies between 186 kHz and 50 MHz. Rejection of at least 30 dB was required for the lower-frequency sideband, over an IF band of either 50-150 MHz or 150-250 MHz at any LO frequency within the specified range and at any baseband frequency in the range 186 kHz to 50 MHz [D'Addario (1978)]. The conversion loss of the complete mixer was specified to be less than 30 dB and its noise figure was not to exceed the conversion loss by more than 8 dB. At a nominal IF input power spectral density of -20 dBm/50 MHz, all spurious intermodulation products were to be at least 50 dB below the upper sideband baseband response.

BASIC PRINCIPLES

The principle of operation of image rejection mixers was originally described by [Hartley (1928)]. Such a device is designed to convert only the desired sideband of the RF input signal to a specified output frequency, while suppressing the conversion of the unwanted portion of the RF spectrum. This attribute is a highly desirable characteristic of mixers used for IF-to-baseband conversion in cross-correlating spectrometers for radio astronomical applications, since it permits the selective conversion and spectral analysis of any segment of the IF passband of the receiving system by simply altering the LO frequency and video bandwidth.

The schematic of Figure 1 illustrates the operation of an image rejection mixer. The quadrature hybrid supplies local oscillator signals, differing in phase by $\pi/2$ radians, to each of the mixers. Phase shift through the A and B portions of the phase difference network differ by $\pi/2$ at all baseband frequencies. The waveform expressions suppose that the A network has no phase shift and the B network a $\pi/2$ phase lag, but design of the network is greatly simplified by requiring only that a $\pi/2$ relative phase difference be maintained.

In a practical mixer, phase and amplitude errors between the ~~quadrature baseband channels give rise to incomplete rejection of the~~ unwanted sideband. Consider a mixer with quadrature baseband outputs with phase error $\phi(\omega)$ providing inputs to a 90° phase difference network with inherent quadrature error $\theta(\omega)$. The image rejection may be written as:

$$R(\omega) = \{ \frac{E_A{}^2 + E_B{}^2 + 2E_AE_B \cos [\phi(\omega) + \theta(\omega)]}{E_A{}^2 + E_B{}^2 - 2E_AE_B \cos [\phi(\omega) + \theta(\omega)]} \}^{\frac{1}{2}}$$

THE BASEBAND QUADRATURE NETWORK

Given a pair of coherent input signals, the baseband quadrature network must transform those signals to produce outputs which differ in phase by $\pi/2 \pm \delta$, over a specified range of frequencies. Given that the quadrature mixer phase errors are specified, the allowable peak phase error (δ) is determined by the minimum required image rejection for the complete mixer assembly. The phase shift can be realized physically with a pair of all-pass networks. Each network of the pair has a zero in the right half of the complex frequency plane, corresponding to each pole in the left half, in order to obtain the desired characteristics of flat amplitude response and monotonically decreasing phase shift with increasing frequency.

The phase difference between the outputs of a pair of networks may be made to ripple about $\pi/2$ over a specified frequency range, since the differential transfer function [$H_A(s)/H_B(s)$, where H_A, H_B are the transfer functions of the individual networks] contains both poles and zeros in the right half plane. For an N-pole differential transfer function, there exist N specific frequencies at which exact phase quadrature exists between the outputs of the individual networks. With a sufficient number of poles it is possible to approximate a $\pi/2$ phase difference over a given frequency range to within any desired tolerance.

The pole locations of the differential transfer functions are unique if a minimal deviation, equal-ripple phase response is specified, for a given frequency range and number of poles. The equi-ripple condition is satisfied by transfer functions which belong to a set of polynomial functions described by Cauer (1933). The unique Cauer solution can be obtained by techniques based on the use of elliptic functions [Darlington (1950), Weaver (1954), Bedrosian (1960)], or, alternatively, the pole-zero locations for certain even numbers of poles can be found by algebraic means, as described by Albersheim and Shirley (1969). For the present design, an elliptic function solution, a modification of the method of Darlington (1950), is used to derive the corner frequencies and the coefficients of the transfer function polynomials (Granlund 1979).

Networks with the desired rational all-pass transfer function may be realized in a number of different ways. However, the two-source approach illustrated in Figure 2(a) can be shown to result in a simpler topology with more readily synthesized components than other methods. The network shown in Figure 2(a) represents a candidate for either the A or B networks described above, the pair of voltage sources delivering outputs proportional to one of the quadrature mixer outputs.

The voltage gain (for $\alpha = 1$) may be written as:

$$\frac{v}{e} = \frac{y/g - 1}{y/g + 1}$$

The network will be an all-pass, unit-gain device if, and only if, the phase of y/g is $\pm\pi/2$, implying that y be a pure susceptance, jb, and g a pure conductance. The all-pass nature of the circuit requires that

$$\frac{v}{e} = \prod_j \frac{s-s_j}{s+s_j} = \frac{N(s)}{D(s)}$$

where the transfer function polynomials N(s), D(s) may be written in terms of a pair of odd and even polynomials, $D_o(s)$, $D_e(s)$ respectively

$$N(s) = D_e(s) - D_o(s)$$

$$D(s) = D_e(s) + D_o(s)$$

Then

$$\frac{v}{e} = \frac{D_e/D_o - 1}{D_e/D_o + 1}$$

and so

$$y = g(\frac{D_e}{D_o})$$

It follows that, since D(s) is a Hurwitz polynomial, y is a susceptance function having poles and zeros that alternate along the jω axis. The simplest development for the susceptance jb is a ladder connection of shunt capacitors and series inductors as indicated in Figure 2(b). The susceptances and reactances of the components comprising the network, normalized by the fixed conductance g, may be derived by a continued fraction expansion of the polynomial describing the desired network admittance.

Finite source impedance for the voltage sources may be accommodated by requiring less than unity gain for the all-pass network. In this case, the gain can be written

$$\frac{v}{e} = K \prod_j \frac{s-s_j}{s+s_j} \; ; \; K < 1$$

which results in the relationship

$$y = g\left(\frac{1-K}{1+K}\right) + \cfrac{1}{\cfrac{1}{\left(\frac{4Kg}{1-K^2}\right)} + \cfrac{1}{\left[\frac{4Kg}{(1+K)^2}\right] \cdot \frac{D_e}{D_o}}}$$

This susceptance may be realized in the form shown in Figure 2(c), where the complete network has been transformed to result in a topology in which conductances are associated with both voltage sources.

The discussion so far has not considered the effects of losses in the inductors of the ladder on network performance. It can be shown (Granlund 1979) that, if the coils are represented by a combination of a pure inductance and a series resistance, representing skin effect and core losses, then, to a first approximation, the amplitude response of the network has a first-order dependence on coil Q, whereas the phase response possess only a second-order dependence.

A study of these effects for typical components in an eight-pole network indicates that although the gain of the network may increase by up to 0.2 dB from the lower to the upper edge of the design band, the phase error due to losses alone will be several orders of magnitude smaller than the design specification. Furthermore, loss induced phase errors may be compensated for by suitably adjusting the design values of the elements in the reactance networks.

The design presented here uses an eight-pole, equi-ripple approximation to the desired transfer function over the band of frequencies 186 kHz to 50 MHz. The theoretical peak phase deviation from quadrature within this frequency range is 0.014 radian and the corresponding minimum image rejection, assuming an ideal quadrature mixer, is -43.1 dB. Figure 3 shows the theoretical image rejection of the complete mixer assembly as a function of baseband frequency. Normalized component values for the A and B networks are listed in Table I for a number of baseband frequency ranges. The allowable tolerances on these values have been investigated in terms of the degradation of the theoretical image rejection of the complete mixer. It can be shown that deviations of 1% from nominal value, for a given component, can result in up to 3 dB degradation in image rejection, when all other components in the network are held at their nominal value. Because such close tolerances are required, it is desirable to include adjustable elements in the practical quadrature network, which can be tuned to optimize performance.

In practice, each mixer baseband port will be represented by a single voltage source in combination with a known internal series resistance. Therefore, a network topology is required which allows each reactive network to be suitably connected to a two-terminal source while maintaining the basic electrical characteristics of the circuit of Figure 2(c). The resistive embedding network shown in Figure 4(a) is a realization which will serve the dual purpose of two-terminal source interfacing and signal combining after the appropriate phase shift has been introduced. The resistance values must be chosen in such a way as to correctly normalize the reactance networks, given the finite source impedances, while maintaining the all-pass nature of the complete network.

In general, the topology is such that the following configurations are not possible:

i) One source terminal, one output terminal and one reactance network terminal common, allowing them to be grounded.

ii) A circuit employing a two-terminal source, with one source terminal and one output terminal common.

From Figure 4(a), it is clear, however, that one terminal of each reactance network and one output terminal are common and, hence, may be grounded. The power combining function is passive, the sum power being delivered to a common load resistance.

The basic design relationships for the embedding network can be derived for each half of the network in Figure 4(a). By applying a series of transformations, an equivalent circuit with topology similar to that of Figure 2(a) can be derived. The equivalent circuit is shown in Figure 4(b). The signal gain may be written as

$$\frac{v}{e} = \frac{R\, R_3\, (Z_o - r_c)}{R_a R_b \Sigma_1} \cdot \frac{\frac{R_2 R_4}{R_3} - jX}{Z_o + jX}$$

where

$$\Sigma_1 = R_1+R_3+R_4 \qquad R_a = \frac{R_1R_3}{\Sigma_1} + R_2 \qquad R_b = \frac{R_3R_4}{\Sigma_1} + R$$

$$Z_c = r_c + jX = \frac{R_1R_4}{\Sigma_1} + jX \qquad Z_o = Z_c + \frac{R_aR_b}{R_a+R_b} - jX$$

The condition for an all-pass response is

$$Z_o = \frac{R_2R_4}{R_3}$$

which is realizable, since if R, R_1, R_3, R_4 have positive, finite, nonzero values, R_2 is given by

$$R_2 = \frac{(b-1) \pm \sqrt{(b-1)^2 + 4ac\ (b+1+ac)}}{2a}$$

where

$$a = \frac{R_1R_3}{\Sigma} \qquad b = \frac{R_3}{R_4} \qquad c = \frac{1}{R_b}$$

The second term in the radical is positive, so the radical exceeds $|b-1|$ and exactly one finite, positive, nonzero value exists for R_2.

The magnitude of the gain, with the all-pass condition in effect may be simplified to

$$\left|\frac{v}{e}\right| = \frac{\mp RR_3}{R_3(R_1+R_4) + (R+R_2)\ \Sigma_1}$$

The most useful measure of circuit performance is the ratio of the power delivered to the load R to the power available from one of the mixers. This power gain may be written as

$$G_s = 4\left(\frac{Z_1}{R}\right)\left(\frac{R_6}{Z_1+R_5+R_6}\right)^2 \left|\frac{v}{e}\right|^2$$

A problem arises with the passive interconnection of Figure 4(a). Variation of X_A with frequency can influence the value of Z_o seen by the B network, and vice versa. If the all-pass condition has been met, with iterative impedance Z_o (i.e., the impedance seen at the terminals of jX_B, when jX_A is replaced by Z_o), then a crosstalk gain may be defined. If jX_B is replaced by Z_o and jX_A by a voltage source of internal impedance Z_o, the voltage gain from the terminals of jX_A to the terminals of jX_B may be written as

$$G_c = \frac{2RR_2R_3}{(R_2+R_3)[(R_2+R_3)R_4 + 2R(R_3+R_4)]}$$

The crosstalk modifies the frequency dependence of the differential transfer function and hence can cause degradation of the performance of the quadrature network. The simplest method for controlling the interaction is to make the diagonal resistances large enough, and the others small enough, so that sufficient attenuation from the terminals of X_A to the terminals of X_B is provided to reduce the crosstalk to an acceptable level. It can be seen from Figure 3 that crosstalk attenuation approximately equal to the desired theoretical image rejection, with an ideal quadrature mixer, results in adequate suppression of the interaction. Unfortunately, one-half of the additional attenuation will appear as signal insertion loss through the quadrature network.

The degree of impedance mismatch between the mixer baseband ports and the phase shift network should also be considered. The mixers would be expected to operate optimally when loaded by a constant, nonreactive baseband port termination. However, the reactances X_A and X_B vary from short- to open-circuit a number of times across the operating frequency range, which could result in significant variations of phase shift network input impedance with frequency. On the impedance plane, the input impedance describes a circular locus with changing frequency, becoming perfectly resistive when X_A or X_B are open or short-circuited. The maximum and minimum values of these resistances, corresponding to the open or short-circuit condition, respectively, are given by:

$$R_{max} = R_4 + \frac{R_3(R_2+R)}{R_3+R_2+R} \qquad R_{min} = \frac{R_2(R_3R_4+R_3R+R_4R)}{R_4R_2+RR_2+R_3R_4+R_3R+R_4R}$$

For a mixer with characteristic impedance Z_1, two conditions must be met for correct operation of the circuit of Figure 4. Firstly, the iterative impedance seen at the source terminals must be equal to Z_1, i.e.,

$$Z_1 = \left[\left(R_5 + \frac{R_6R_{min}}{R_6+R_{min}}\right)\left(R_5 + \frac{R_6R_{max}}{R_6+R_{max}}\right)\right]^{\frac{1}{2}}$$

and, secondly, the resistance R_1 is given by

$$R_1 = \frac{R_6(R_5+Z_1)}{R_6+R_5+Z_1}$$

These conditions allow the values of R_5 and R_6 to be determined.

Since the bilinear transformation from the impedance plane onto the reflection coefficient domain maps circles into circles, the maximum and minimum VSWR presented to the mixer by the quadrature network may readily be calculated, once the values of R_5 and R_6 have been determined.

The design of the resistive embedding network must, therefore, represent a trade-off between crosstalk minimization, insertion loss minimization and input VSWR reduction. The free variables in the design procedure are seen to be the normalizing impedance for the reactance networks and the mixer output resistance. Given the maximum allowable cross-coupling and input VSWR, resistance values are selected to give a normalizing impedance which minimizes the desired sideband signal insertion loss of the complete quadrature network, while maintaining the required all-pass transfer function. The solution uses computer based, iterative minimization of the upper sideband transmission loss, constrained by the desired maximum cross-coupling and minimum input return loss. Table II indicates the component values and insertion loss for the present design for several of the cases that were analyzed, with a maximum specified crosstalk gain of -43 dB and a maximum specified VSWR of 1.25:1. It is seen that a normalizing impedance of 50 Ω, with a mixer source impedance of 50 Ω, results in a satisfactory and physically realizable compromise, with an available power gain from mixer to quadrature network output of -22.85 dB.

MIXER WITH QUADRATURE BASEBAND OUTPUTS

Since the peak deviation from phase quadrature due to the eight-pole reactive phase shift network is 0.014 radians, the peak allowable amplitude and phase quadrature errors between mixer baseband outputs over the frequency range 186 kHz to 50 MHz are seen to be 0.5 dB and 0.021 radian respectively, for an image rejection of at least 30 dB. These requirements constrain the approach taken in the design of the quadrature mixer.

The phase relationship between the mixer baseband outputs is determined primarily by the phase and amplitude balance of the RF and LO hybrids, as well as by the differences between the transfer functions of the individual mixers. To achieve a high degree of phase and amplitude balance at the outputs of a given hybrid, all ports must be terminated in well-matched loads. This implies, for the present design, that each mixer must exhibit high return loss at both RF and LO inputs and must have an amplitude and phase response well-matched to that of its counterpart.

In the mixers described here, the quadrature hybrids are implemented as meander-line, broadside-coupled directional couplers in stripline. This approach results in a compact hybrid with superior phase and amplitude balance, when compared with lumped element designs, over the specified octave LO and RF bandwidths.

A balanced mixer with high return loss at both LO and RF ports can be fabricated with the aid of a quadrature hybrid and a pair of Schottky Barrier diodes, as shown circled in Figure 5. For the single balanced mixer, let the dynamic reflection coefficients of the diodes be given by Γ_1, Γ_2 at the in-phase and quadrature ports, respectively. The voltage reflection coefficient at either the LO or RF input port, with the other input terminated, is given by

$$\rho_{in} = \pm\tfrac{1}{2}[\Gamma_2 - \Gamma_1]$$

The RF to LO port isolation may also be written as

$$I_{LR} = \tfrac{1}{2}j[\Gamma_2 + \Gamma_1]$$

If the dynamic impedances of the diodes are closely matched, the input return loss at either LO or RF port is large. However, the LO-to-RF isolation will be poor unless matching networks are provided to transform the diode impedance to a value close to the characteristic impedance.

In the present design, the functions of high-pass filter and impedance transformer are combined in a single network inserted between diode and quadrature hybrid. The high-pass filter presents a low impedance termination to the diode at all baseband frequencies. The network is a single shunt inductor, series capacitor design, the shunt inductor serving the dual purpose of providing a dc return path for each diode. If matched diode pairs ($\Delta Z_{dyn} < 15\ \Omega$), with nominal dynamic impedance of 300 Ω resistive at 200 MHz, 1 mW incident LO drive, and identical matching networks are used, the reflected signal seen by each port of the hybrid will be well balanced in amplitude and phase with a VSWR of less than 2:1 over the specified RF range. Figure 6 illustrates the frequency dependence of the impedance (Z_{Dn}) presented to the hybrid and also demonstrates that the impedance (Z_{D1}) seen by the diode is large at baseband frequencies. Given these matching network qualities, it is easily deduced that the VSWR at the mixer LO or RF port is expected to be less than 1.03:1 and that the LO-to-RF isolation should be greater than 15 dB over the full RF frequency range.

Design of the baseband circuit of the mixer is complicated by two factors. Firstly, the RF and LO frequencies are closely spaced and, secondly, a baseband bandwidth of greater than two decades is required. A low-pass filter is needed in the baseband output circuit in order to isolate the fed-through LO signal and to properly terminate the mixer diodes at RF/LO frequencies. The mixers described here use a three-pole Chebyshev response filter, normalized to the nominal baseband output resistance of the mixer (50 Ω), which provides a minimum of 15 dB LO and RF-to-baseband isolation.

IMPLEMENTATION AND TESTING OF A PRACTICAL IMAGE REJECT MIXER

A combination of stripline and microstrip techniques is employed in the implementation of the quadrature mixer. The three quadrature hybrids are stripline devices, whereas all interconnections between components are microstripline on an aluminum-backed, Teflon-fiberglass dielectric material. The inductors for the filters and matching networks are synthesized by precisely determined lengths of high impedance microstrip line. All capacitors in these circuits are high Q ceramic chip capacitors. The use of these techniques results in repeatable inductance and capacitance values while minimizing the effects of parasitic reactances on mixer performance. When combined with the matched pairs of HP 5082-2208 Schottky diodes, employed in the manner shown in Figure 5, the approach taken in fabrication results in a mixer with exceptional phase quadrature accuracy and amplitude balance between the baseband ports.

Measured peak deviation from quadrature at any combination of LO, RF and baseband frequencies within the specified design ranges was less than 0.013 radian. Measured peak amplitude unbalance was 0.25 dB, with less than 0.20 dB variation in conversion loss across the full baseband bandwidth for any LO-RF combination within specification. The nominal measured conversion loss of the quadrature mixer alone, from the RF port to either baseband port, was -12.2 dB. The foregoing measurements were made with +10 dBm LO drive at -20 dBm RF input power.

The baseband quadrature network operates optimally when the driving source impedance is 50 Ω resistive. However, each of the single balanced mixers will have a typical output resistance lying within the range 125 to 200 ohms depending upon the particular diodes used. Impedance matching between the quadrature mixer and the baseband quadrature network is achieved over the multidecade baseband range with the aid of resistive, minimum-loss, impedance matching networks (nominal impedance ratio 3:1). The resultant driving impedance for the baseband quadrature network will then be in the range 49.1 to 51.47 ohms with a total insertion loss between the mixer and the network input of 9.86 to 11.36 dB respectively. However, for a given mixer, since matched diodes are employed, the difference in source resistance between the quadrature outputs will be less than .7 ohms and the difference in insertion loss less than .5 dB.

The baseband quadrature network is constructed on a separate printed circuit board mounted inside the mixer housing, and wired directly to the quadrature mixer outputs. In order to maintain reasonable overall conversion loss for the complete mixer assembly, the passive phase shift network is followed by a two-stage wideband transistor amplifier. The amplifier is designed to have a nominal input impedance of 3 ohms (±.2 ohms tolerance) which varies by less than .1 ohm across the baseband range of 186 kHz to 50 MHz.

The amplifier also exhibits low-noise figure, multidecade bandwidth and sufficient gain to overcome the effects of the resistive embedding network which precedes it.

Tests on more than thirty completed image reject mixers indicate excellent, repeatable performance. Figures 7 and 8 show typical measured image rejection as a function of baseband frequency for a range of LO frequencies and as a function of LO frequency for different baseband frequencies. Figure 7 also demonstrates the relationship between conversion loss and baseband frequency. Image rejection of greater than -34 dB has been achieved in every unit tested to date. Conversion loss was less than -27 dB and was uniform to within ±0.15 dB across the full baseband bandwidth. LO-to-baseband and RF-to-baseband isolation have been found to exceed 15 dB relative to the conversion loss for frequencies in the range 100-150 MHz or 200-250 MHz. For RF frequencies in the range 186 kHz to 50 MHz, where direct feedthrough of spurious signals may cause problems, isolation is at least 40 dB greater than the in-band conversion loss. Return loss at either RF or LO port was measured to be greater than 25 dB within the design frequency range for any mixer.

The maximum excess noise figure, relative to the conversion loss, for any mixer tested was 7.0 dB, within the specified limits. Finally, measurements of two-tone intermodulation products indicate that:

i) The third order intercept point for mixer products is typically +18 dBm at the RF input.

ii) The second order intercept point for wideband amplifier products is typically +33 dBm at the RF input.

One decibel gain compression typically occurs with an RF input level of +7 dBm.

CONCLUSION

An image reject RF-to-video converter design has been described which provides repeatable performance over octave RF and LO, and multidecade baseband, bandwidths. Greater than 34 dB image rejection is typically achieved, with conversion loss for the wanted sideband being less than 27 dB and uniform to within ±0.15 dB across the baseband 186 kHz to 50 MHz. More than one hundred of the devices are presently being incorporated into the VLA IF-to-baseband conversion subsystem.

REFERENCES

Albersheim, W. J. and Shirley, F. R. "COMPUTATION METHODS FOR BROADBAND 90° PHASE DIFFERENCE NETWORKS", IEEE Trans. Circuit Theory, CT-16, pp. 189-196, May 1969.

Bedrosian, J. "NORMALIZED DESIGN OF 90° PHASE DIFFERENCE NETWORKS", IRE Trans. Circuit Theory, CT-7, pp. 128-136, June 1960.

Cauer, W. "EIN INTERPOLATIONS PROBLEM MIT FUNCTIONEN MIT POSITIVEM REALTEIL", Math. Zeit., 38, pp. 1-44, 1933.

D'Addario, L. R. "DESIGN CONSIDERATIONS FOR IF RECEIVER SUBSYSTEM", NRAO/VLA Electronics Memorandum No. 175, July 1978.

Darlington, S. "REALIZATION OF CONSTANT PHASE DIFFERENCE", B.S.T.J., 29, pp. 94-104, January 1950.

Granlund, J. NRAO Internal Memorandum, 1979.

Hartley, R. V. L. U.S. Patent 1,666,206, April 1928, Modulation Systems.

Heeschen, D. S. "THE VERY LARGE ARRAY", Sky and Telescope, 49, pp. 334-351, June 1975.

Weaver, D. "DESIGN OF RC WIDEBAND 90° PHASE-DIFFERENCE NETWORKS", PROC. IRE, 42, pp. 671-676, April 1954.

TABLE I

NORMALIZED LADDER NETWORK COMPONENT VALUES

(Eight Pole Transfer Function)

		A - Network				B - Network			
Frequency Range Hz	Peak Phase Error (Radians)	C_1 (Farads)	C_2 (Farads)	L_1 (Henries)	L_2 (Henries)	C_1 (Farads)	C_2 (Farads)	L_1 (Henries)	L_2 (Henries)
2	1.174E-8	8.357E-2	1.561E-1	4.737E-2	5.408E-1	2.154E-2	2.780E-1	9.026E-2	1.425E+0
5	7.230E-6	4.780E-3	9.969E-2	2.863E-2	3.680E-1	1.277E-2	1.830E-1	5.592E-2	1.004E+0
10	8.930E-5	2.999E-3	7.139E-2	1.917E-2	2.844E-1	8.344E-3	1.359E-1	3.858E-2	8.062E-1
50	2.320E-3	9.133E-4	3.345E-2	7.158E-3	1.701E-1	2.858E-3	7.201E-2	1.597E-2	5.387E-1
100	5.500E-3	5.294E-4	2.430E-2	4.602E-3	1.400E-1	1.751E-3	5.579E-2	1.086E-2	4.676E-1
500	2.220E-2	1.416E-4	1.172E-2	1.606E-3	9.295E-2	5.358E-4	3.178E-2	4.417E-3	3.532E-1
1000	3.430E-2	7.891E-5	8.599E-3	1.012E-3	7.894E-2	3.168E-4	2.517E-2	2.993E-3	3.180E-1

NOTES:

i) The frequency range for the tabulated data is from 1 Hz to the number listed. To scale the ladder network component data to other frequencies (range held constant), divide the values by the scale factor.

ii) The tabulated data was calculated for a normalizing impedance of unity. To scale the component values to a different normalizing impedance, divide the capacitance values, and multiply the inductance values by the scale factor.

TABLE II

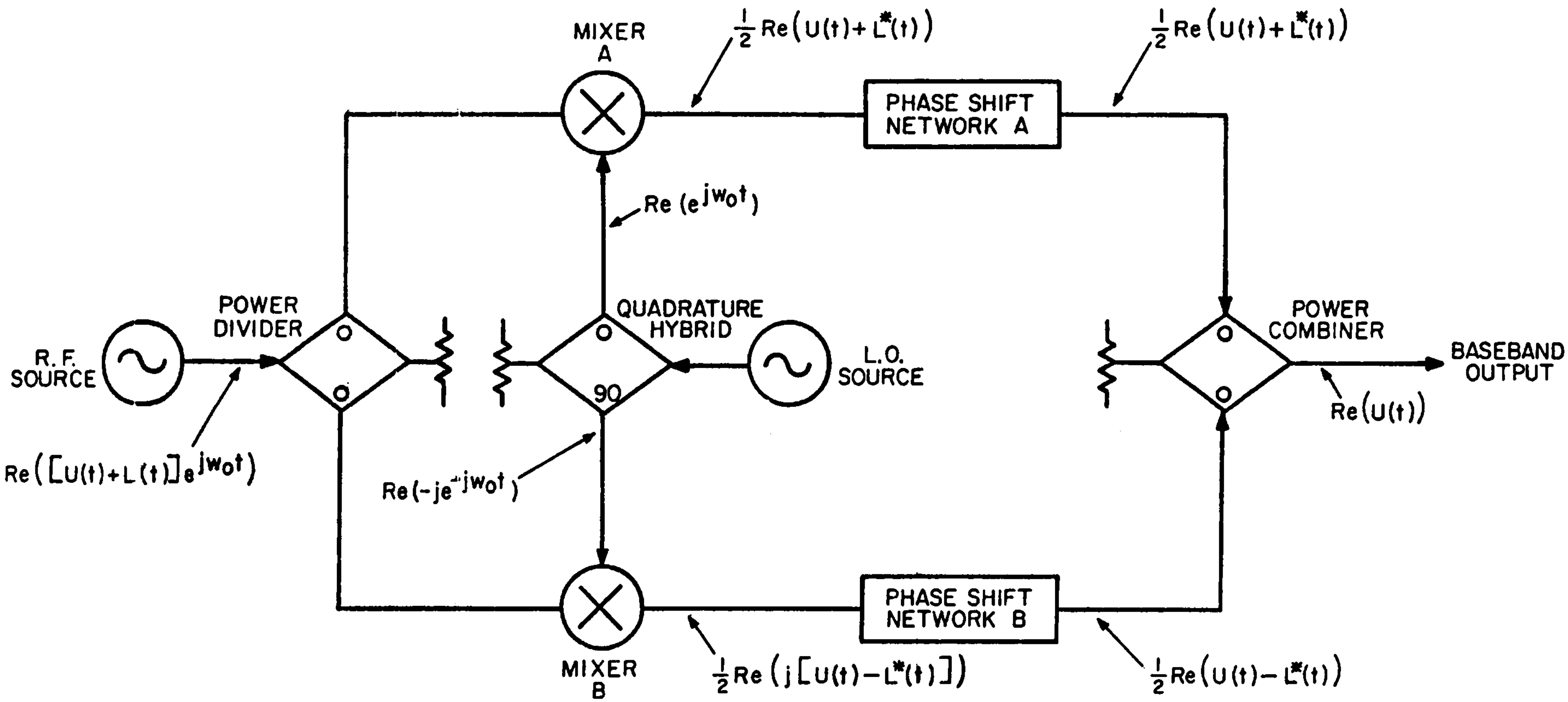
EMBEDDING NETWORK COMPONENT VALUES

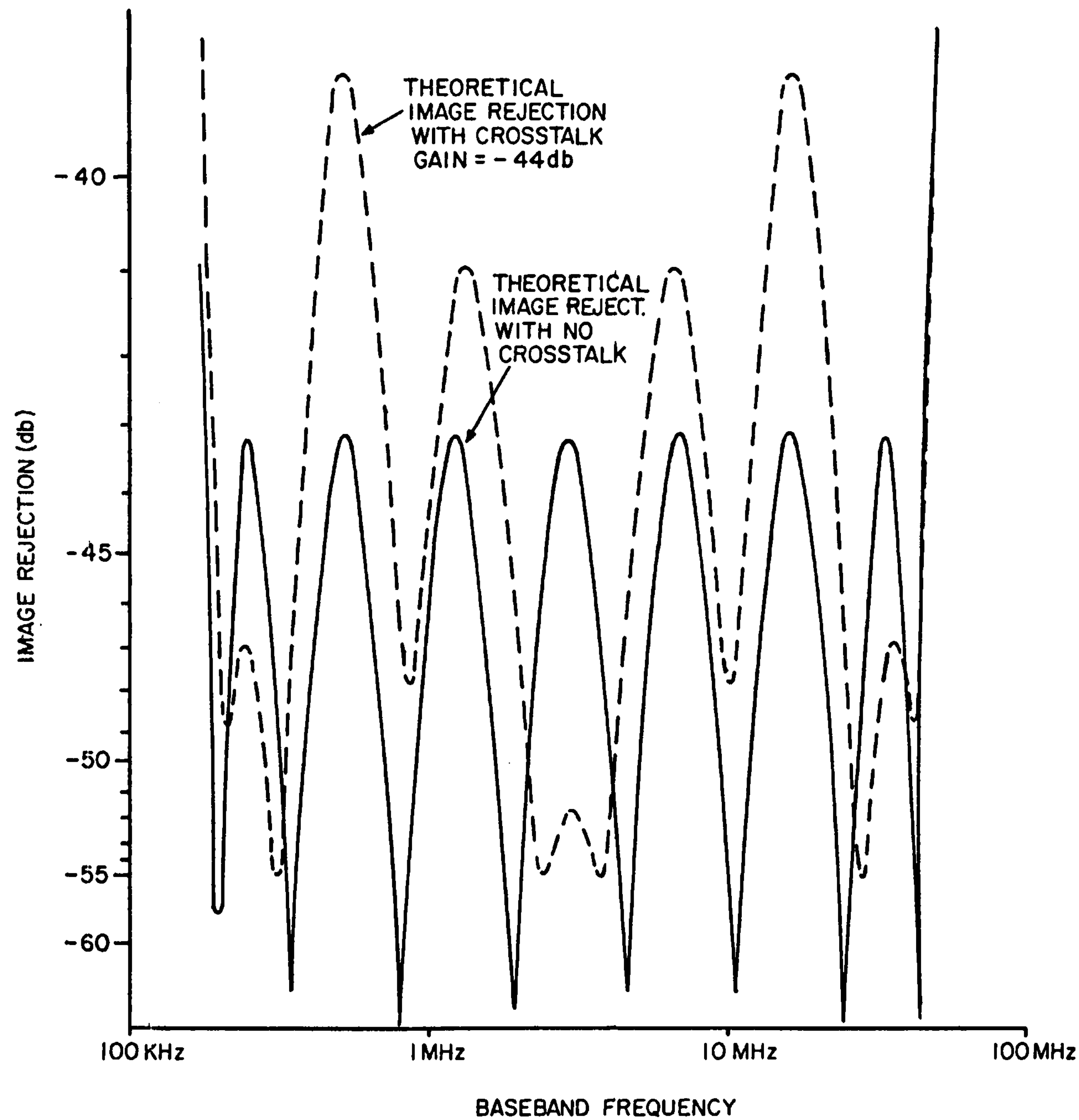
Mixer Output Impedance (Ohms)	Normalizing Impedance (Ohms)	VSWR Presented To Mixer	Signal Gain (dB)	R_2 (Ohms)	R_3 (Ohms)	R_4 (Ohms)	R_5 (Ohms)	R_6 (Ohms)	
603.95	201.915	1.250	-22.664	901.55	1086.45	243.32	0	∞	[See Note i)]
50	100	1.127	-25.945	189.81	473.22	249.32	0	65.94	
50	50	1.250	-22.849	51.64	172.36	166.90	17.86	54.56	[See Note ii)]
12.5	50	1.075	-27.245	81.88	275.29	168.12	0	14.25	

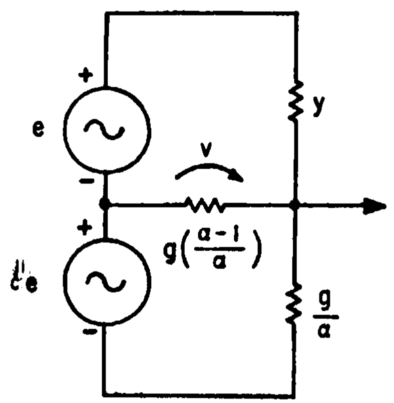
NOTES: i) In this case the constraints were minimizing crosstalk gain and maximizing signal gain, without restriction on mixer impedance. The results show the maximum attainable signal gain for a crosstalk gain of -43dB.

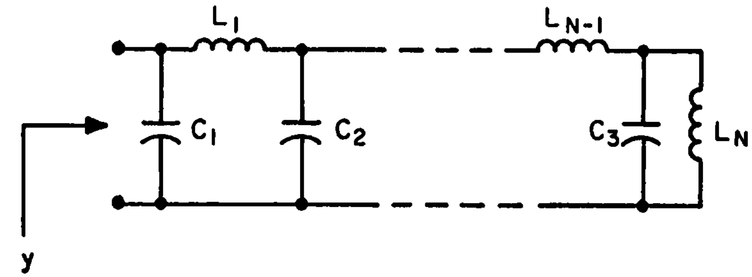
ii) Several other cases were analysed with mixer output impedance and required matching VSWR specified. Of those listed here, the second case results in a good compromise with only .185 dB less gain than the optimum.

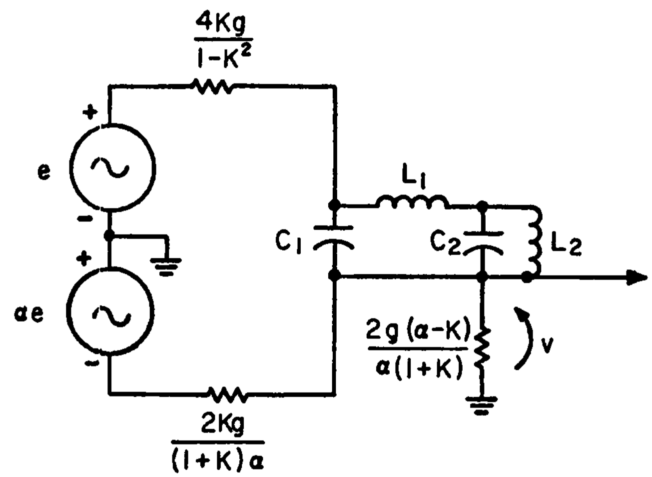
FIGURES

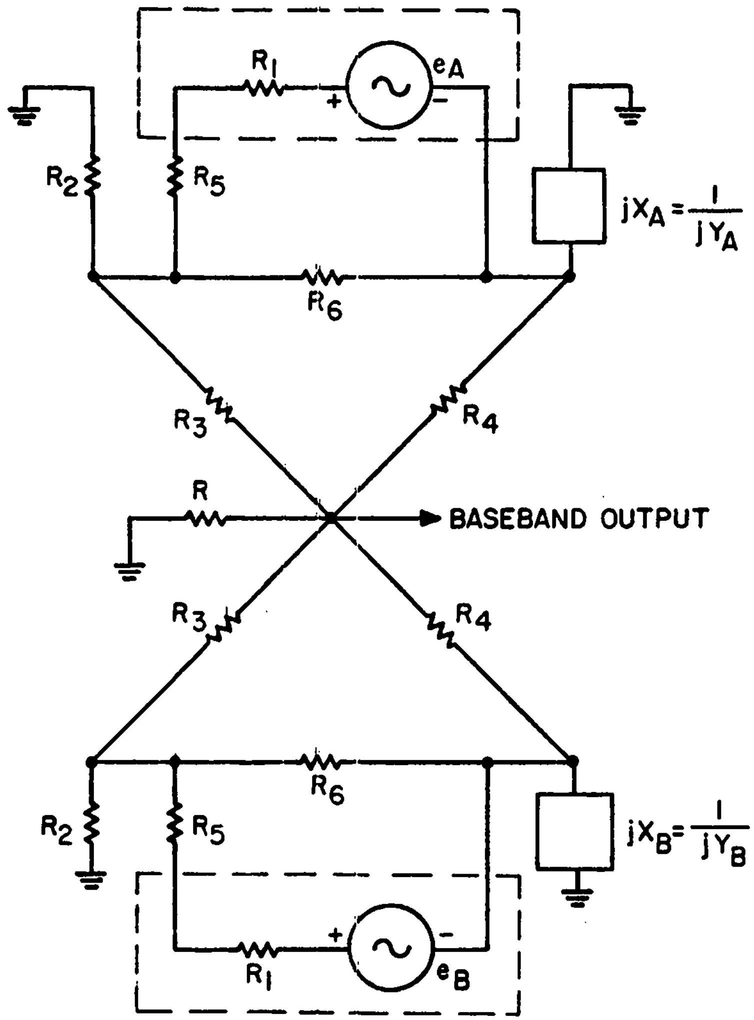

MIXER A
$\frac{1}{2}$ Re(U(t) + L*(t))
$\frac{1}{2}$ Re(U(t) + L*(t))
PHASE SHIFT NETWORK A
Re (e^{jw_0t})
POWER DIVIDER
QUADRATURE HYBRID
POWER COMBINER
R.F. SOURCE
L.O. SOURCE
BASEBAND OUTPUT
90
Re(U(t))
Re([U(t)+L(t)]e^{jw_0t})
Re(-je^{jw_0t})
PHASE SHIFT NETWORK B
MIXER B
$\frac{1}{2}$ Re (j[U(t) − L*(t)])
$\frac{1}{2}$ Re(U(t) − L*(t))

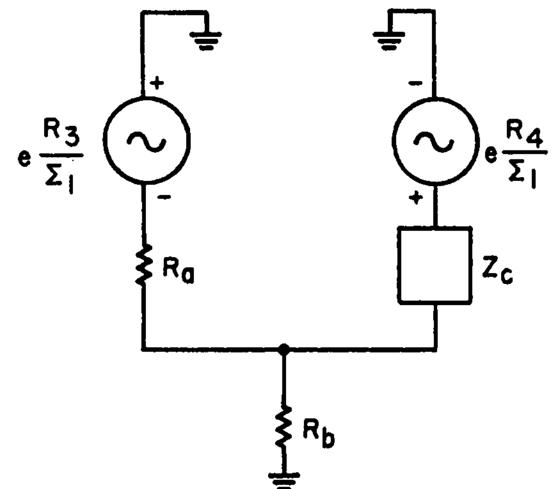

THEORETICAL
IMAGE REJECTION
WITH CROSSTALK
GAIN = -44db
THEORETICAL
IMAGE REJECT.
WITH NO
CROSSTALK
IMAGE REJECTION (db)
-40
-45
-50
-55
-60
100 KHz
1 MHz
10 MHz
100 MHz
BASEBAND FREQUENCY

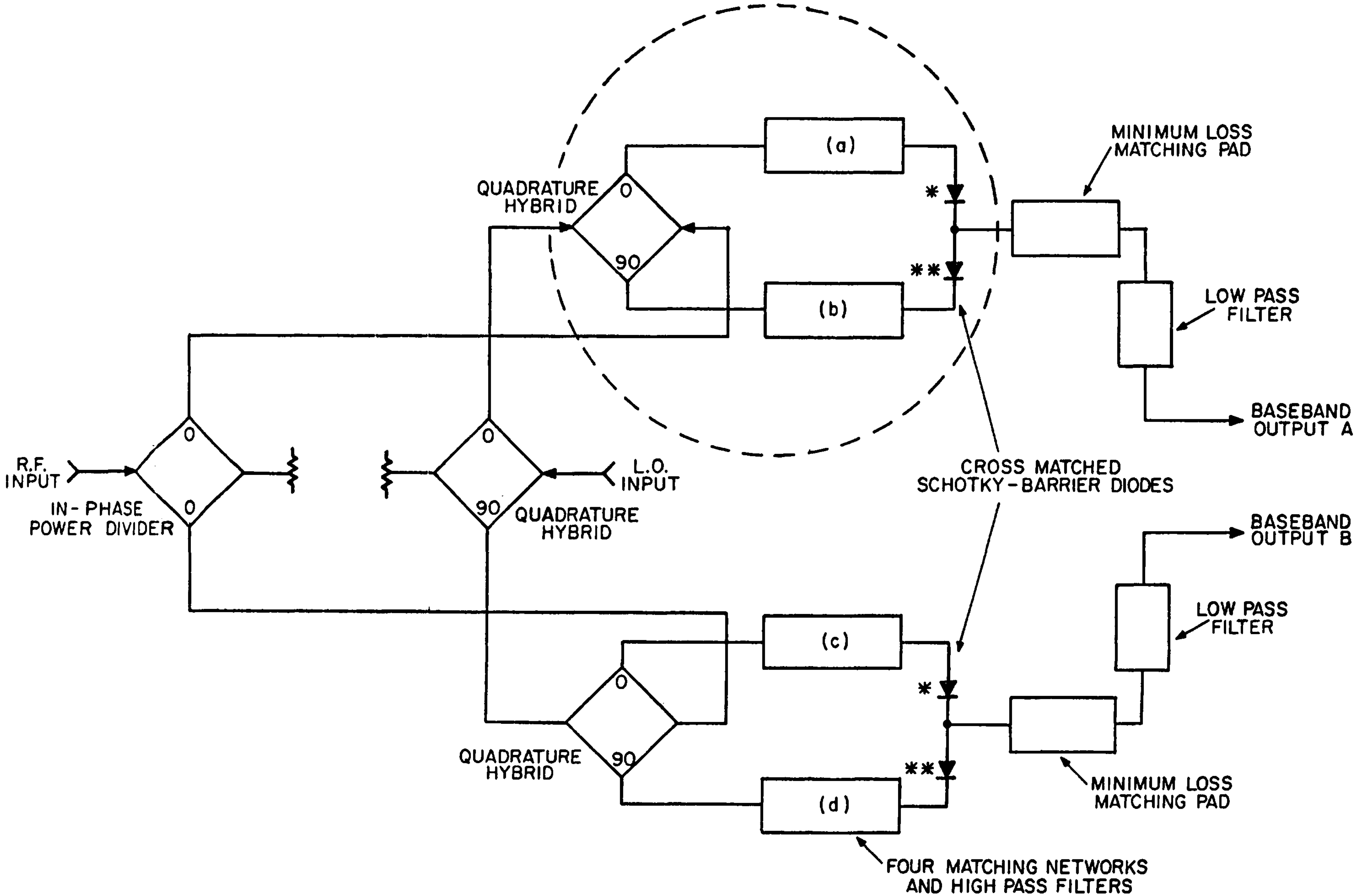

e
y
v
$g\left(\frac{\alpha-1}{\alpha}\right)$
αe
$\frac{g}{\alpha}$


L_1
L_{N-1}
C_1
C_2
C_3
L_N
y

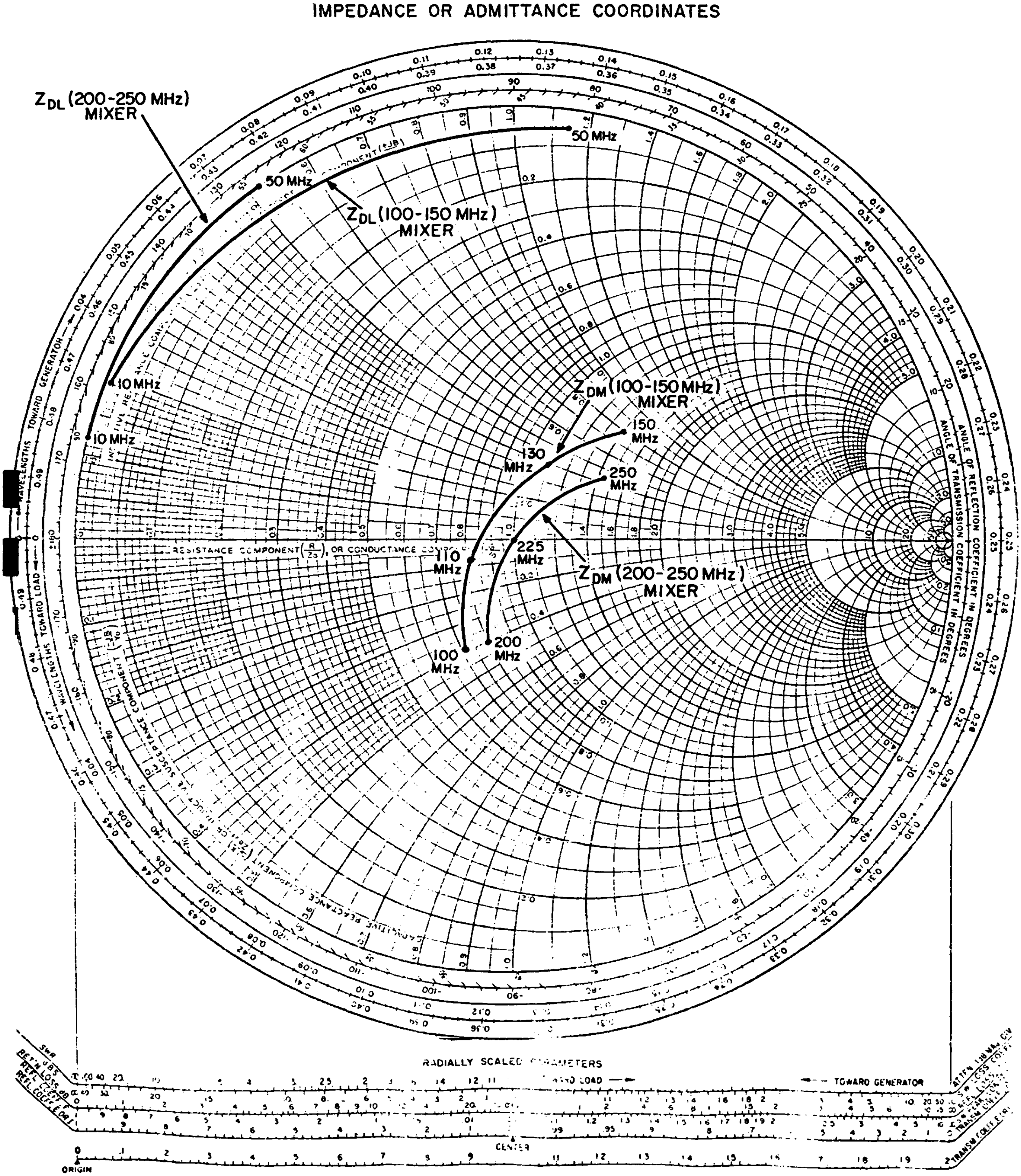

$\frac{4Kg}{1-K^2}$
e
L_1
C_1
C_2
L_2
αe
$\frac{2g(\alpha-K)}{\alpha(1+K)}$
v
$\frac{2Kg}{(1+K)\alpha}$


R1
eA
+
−
R2
R5
jXA = 1/jYA
R6
R3
R4
R
BASEBAND OUTPUT
R3
R4
R6
R2
R5
jXB = 1/jYB
+
−
R1
eB


+
−
e R3/Σ1
−
+
e R4/Σ1
Ra
Zc
Rb


MINIMUM LOSS MATCHING PAD
(a)
QUADRATURE HYBRID
0
90
*
**
(b)
LOW PASS FILTER
BASEBAND OUTPUT A
R.F. INPUT
0
0
IN-PHASE POWER DIVIDER
0
90
QUADRATURE HYBRID
L.O. INPUT
CROSS MATCHED SCHOTKY-BARRIER DIODES
BASEBAND OUTPUT B
LOW PASS FILTER
(c)
0
*
QUADRATURE HYBRID
90
**
(d)
MINIMUM LOSS MATCHING PAD
FOUR MATCHING NETWORKS AND HIGH PASS FILTERS

NAME	TITLE	DWG. NO.
SMITH CHART FORM 82-BSPR (9-66)	KAY ELECTRIC COMPANY, PINE BROOK, N.J. ©1966 PRINTED IN U.S.A.	DATE



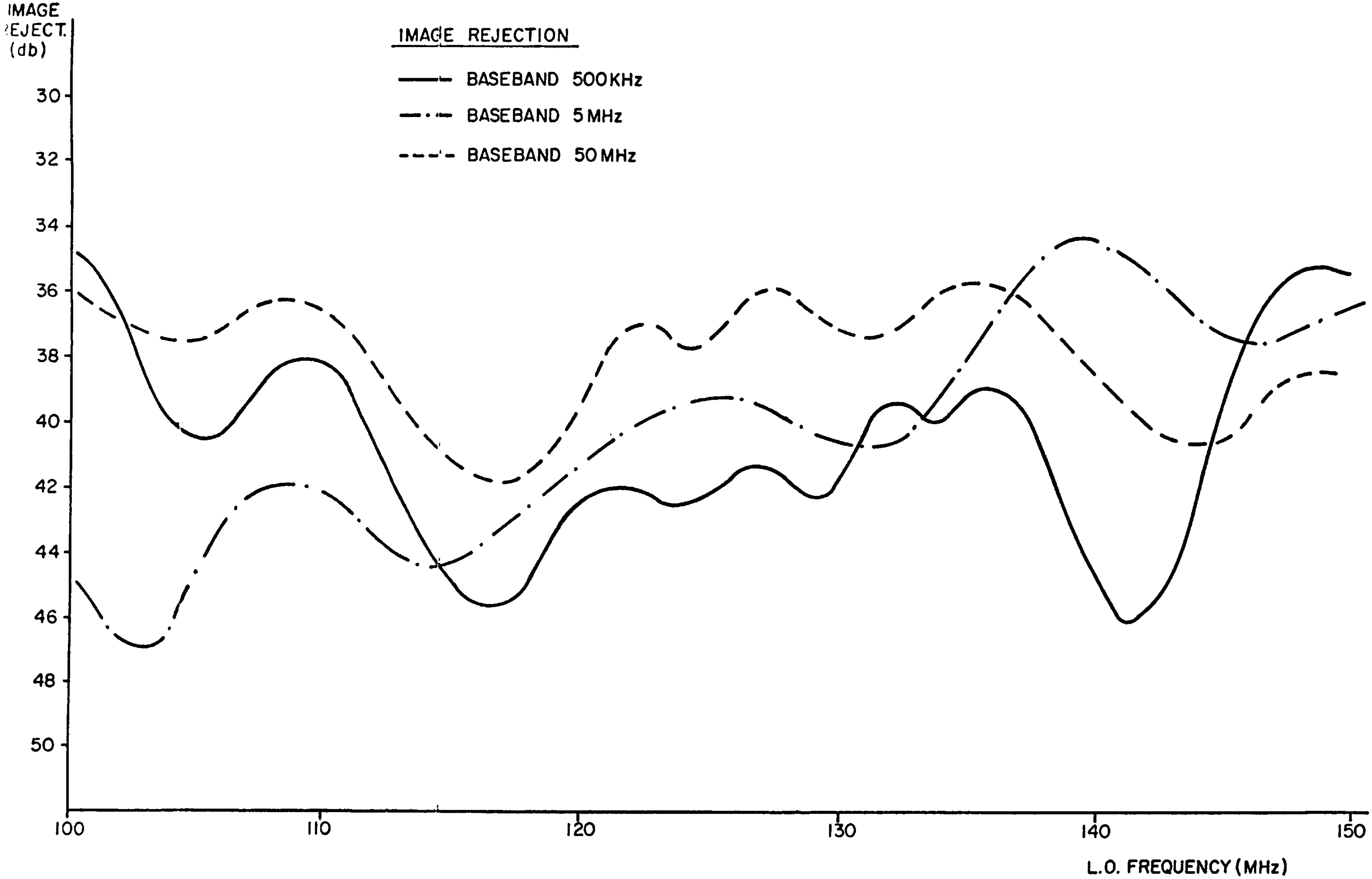


IMAGE
REJECT.
(db)
IMAGE REJECTION
BASEBAND 500KHz
BASEBAND 5MHz
BASEBAND 50MHz
30
32
34
36
38
40
42
44
46
48
50
100
110
120
130
140
150
L.O. FREQUENCY (MHz)

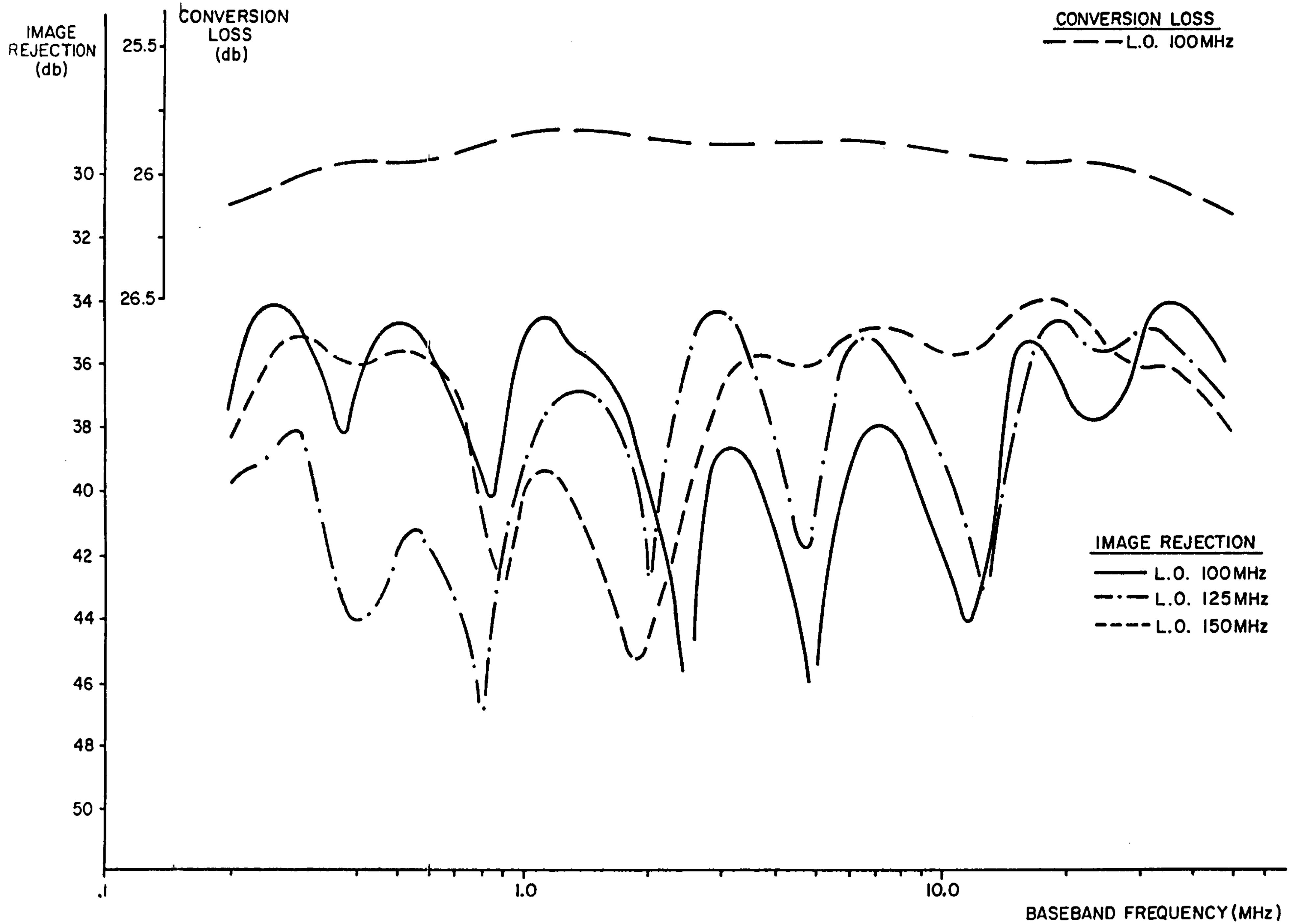


IMAGE REJECTION (db)
CONVERSION LOSS (db)
25.5
26
26.5
30
32
34
36
38
40
42
44
46
48
50
CONVERSION LOSS
L.O. 100MHz
IMAGE REJECTION
L.O. 100MHz
L.O. 125MHz
L.O. 150MHz
.1
1.0
10.0
BASEBAND FREQUENCY (MHz)

APPENDIX B
T3 MODULE TEST SET

B.1 DESCRIPTIONS

Figure B1 shows a block diagram of the T3 module test set. The basic function of the unit is to facilitate testing of either the complete T3 module, or individual submodules by providing all the necessary RF signals.

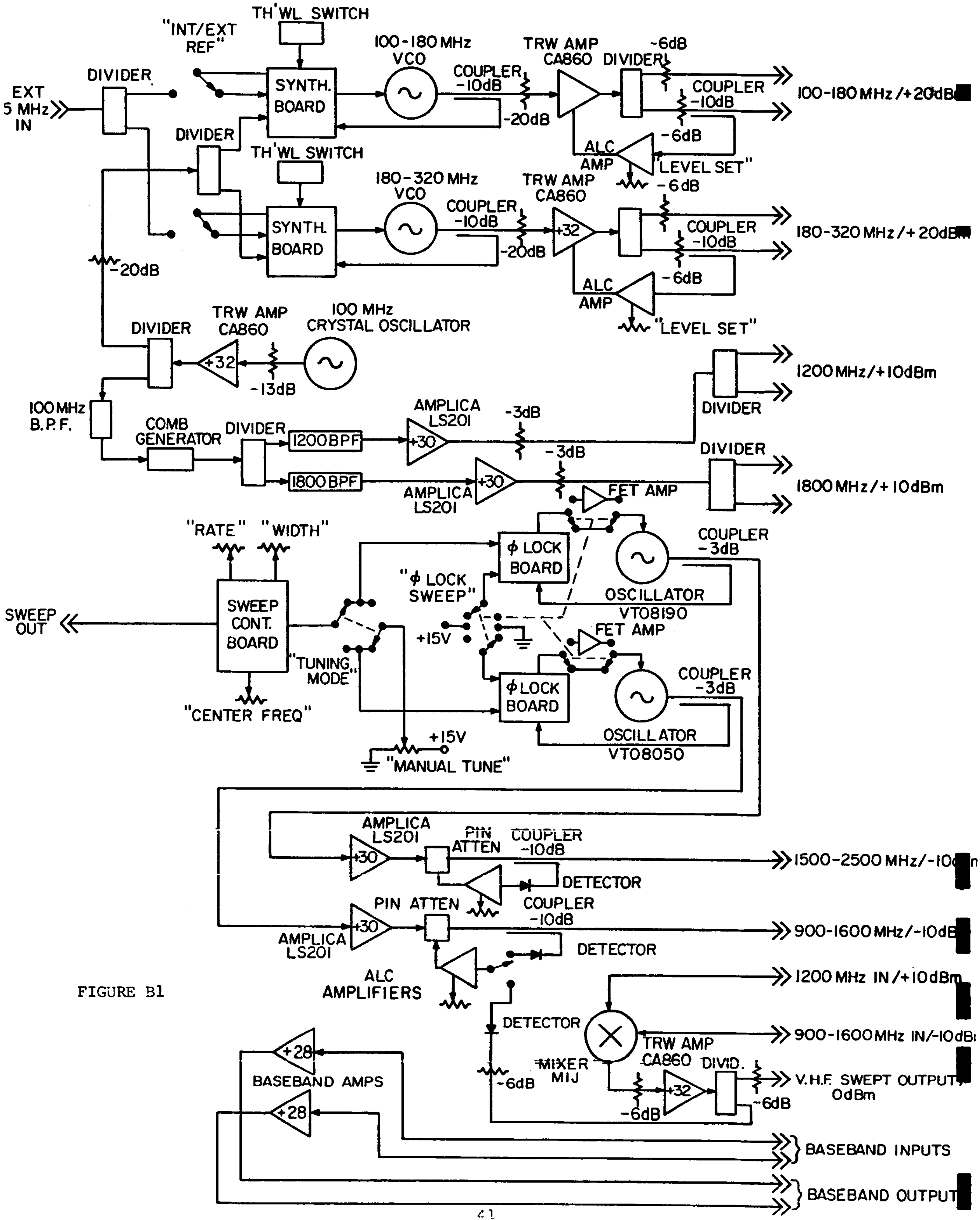
The local oscillator signals required for T3 testing are:

i) 100-150 MHz variable frequency at -15 dBm (output at +20 dBm).
ii) 200-250 MHz variable frequency at -15 dBm (output at +20 dBm).
iii) 1200 MHz fixed frequency at +10 dBm.
iv) 1800 MHz fixed frequency at +10 dBm.

In the test set all of these outputs are provided, phase locked to an internal 100 MHz crystal oscillator reference source. The UHF local oscillators may also be locked, independently of the L-band LO's, to an external 5 MHz reference (+3 dBm±3 dB) if so desired.

The UHF local oscillator signals are provided by phase locked fundamental transistor oscillators - the digital phase lock loop is set up so as to operate as a frequency synthesizer, with output frequency selectable in 10 kHz steps by means of a front panel thumbwheel switch. Maximum operating frequency ranges for the synthesizer oscillators are 100-180 MHz or 180-320 MHz. Each VCO output is amplified by a buffer amplifier with 32 dB gain and then split by a two-way power divider. Nominal output power is +20 dBm, held constant to within ±0.5 dB by an ALC loop which controls the gain of the output amplifier.

The 1200 MHz and 1800 MHz L-band LO signals, are amplified signals selected by filters from a comb spectrum. The comb is generated from the 100 MHz reference signal. Connection is made to the front panel via filters and amplifiers, which provide clean spectrum



TH'WL SWITCH
"INT/EXT REF"
EXT 5 MHz IN
DIVIDER
SYNTH. BOARD
100-180 MHz VCO
COUPLER -10dB
-20dB
TRW AMP CA860
-6dB
DIVIDER
COUPLER -10dB
-6dB
100-180 MHz/+20dB
ALC AMP
"LEVEL SET"
-6dB
DIVIDER
TH'WL SWITCH
SYNTH. BOARD
180-320 MHz VCO
COUPLER -10dB
-20dB
TRW AMP CA860
+32
COUPLER -10dB
-6dB
180-320 MHz/+20dB
ALC AMP
"LEVEL SET"
-20dB
DIVIDER
TRW AMP CA860
+32
-13dB
100 MHz CRYSTAL OSCILLATOR
100 MHz B.P.F.
COMB GENERATOR
DIVIDER
1200 BPF
1800 BPF
AMPLICA LS201
+30
-3dB
DIVIDER
1200 MHz/+10dBm
AMPLICA LS201
+30
-3dB
DIVIDER
1800 MHz/+10dBm
FET AMP
"RATE"
"WIDTH"
SWEEP OUT
SWEEP CONT. BOARD
"CENTER FREQ"
"TUNING MODE"
"ϕ LOCK SWEEP"
+15V
ϕ LOCK BOARD
OSCILLATOR VTO8190
COUPLER -3dB
FET AMP
ϕ LOCK BOARD
OSCILLATOR VTO8050
COUPLER -3dB
+15V
"MANUAL TUNE"
AMPLICA LS201
+30
PIN ATTEN
COUPLER -10dB
DETECTOR
1500-2500 MHz/-10
PIN ATTEN
COUPLER -10dB
+30
AMPLICA LS201
DETECTOR
900-1600 MHz/-10dB
ALC AMPLIFIERS
1200 MHz IN/+10dBm
DETECTOR
900-1600 MHz IN/-10dB
MIXER MIJ
TRW AMP CA860
DIVID.
+32
-6dB
-6dB
-6dB
V.H.F. SWEPT OUTPUT 0dBm
+28
+28
BASEBAND AMPS
BASEBAND INPUTS
BASEBAND OUTPUT

FIGURE B1

1200 MHz and 1800 MHz LO signals at power levels of +10 dBm (±0.5 dB) for general T3 module and submodule testing.

Power-leveled, swept frequency L-band oscillators are also provided. Sweep center frequency, frequency deviation and sweep rate are independently adjustable. A zero-to-ten volt sawtooth output is also provided as a sweep drive for a Tektronix 7L13 spectrum analyzer, to enable the analyzer to be synchronized to the T3 test set sweeper source.

The maximum range of selectable center frequencies is 900 MHz to 2500 MHz using the two oscillators. Nominal output power is -10 dBm (±0.25 dB), with leveling to ±0.1 dB for frequency deviations of less than ±50 MHz provided by ALC loops acting on PIN diode attenuators. The L-band oscillators may also be phase locked to an external reference signal (+10 dBm). The phase locked loops are analogue loops with a built-in automatic search and lock facility.

The lower-frequency sweeper output may be used to generate a UHF swept frequency signal with center frequency varying from 50 MHz to 450 MHz, using the mixer shown in the schematic. For this mode of operation the sweeper output (center frequency in the range 1250 MHz to 1650 MHz) is connected to the RF port of the mixer and the phase locked 1200 MHz LO signal is fed to the LO port. ALC loop control is transferred to the UHF output as shown in the block diagram. The UHF output level is 0 dBm±0.5 dB, with less than ±0.1 dB variation for sweep frequency deviations of less than ±50 MHz when the ALC loop is operational.

~~Finally, a pair of broadband amplifier channels, with 29 dB~~ nominal gain, is provided, with front panel input/output access. These amplifiers (identical in basic configuration to the T3 module baseband amplifier) facilitate image reject mixer testing by providing greater sensitivity for image rejection measurements.

B.2 CIRCUIT DETAILS

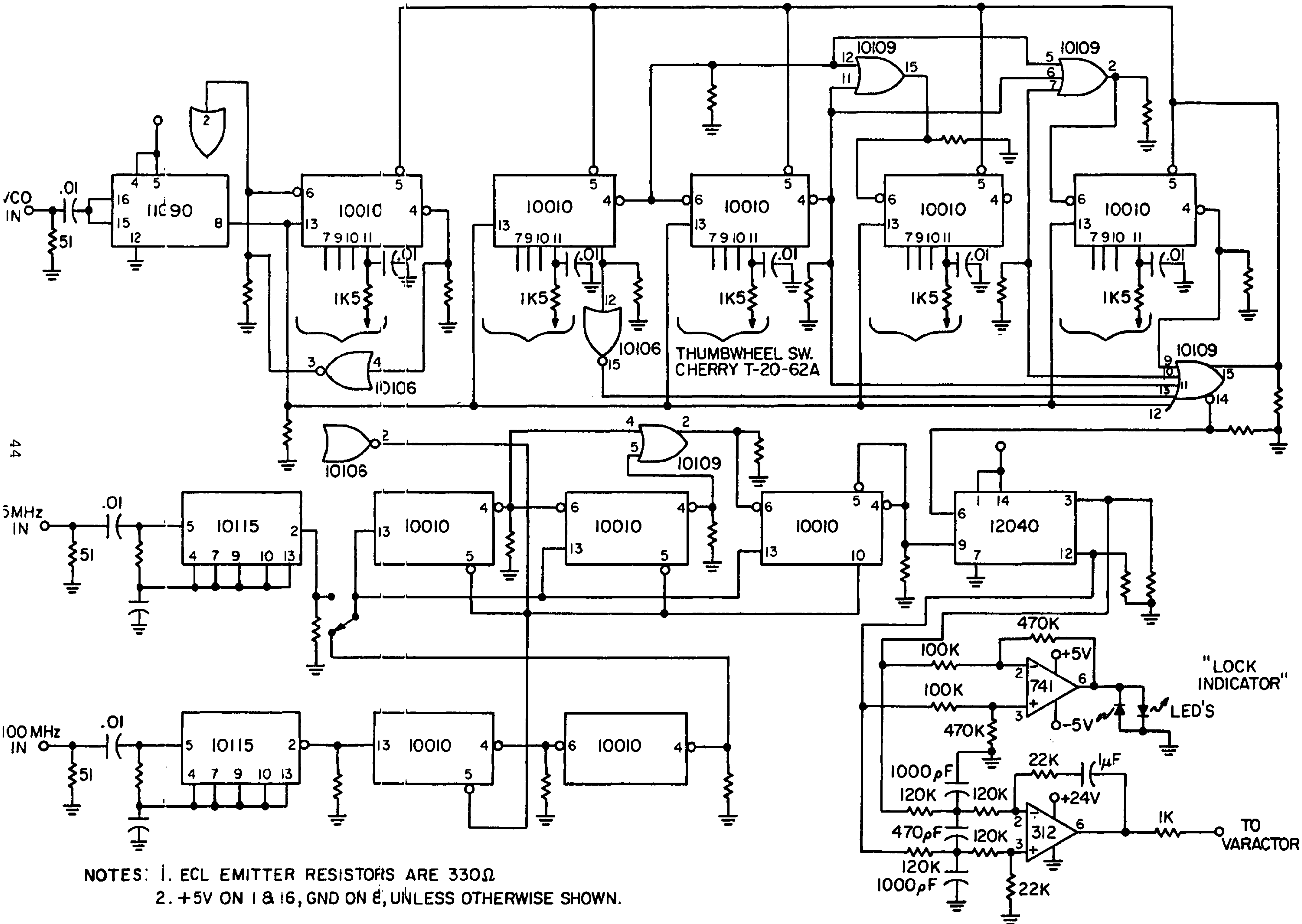
B.2.1 UHF Frequency Synthesizer

A circuit diagram for the synthesizer digital board is shown in Figure B2. Figure B1 shows a detailed block diagram of the synthesizer system. The VCO is an octave bandwidth, fundamental transistor oscillator, with varactor tuning, made by Techtrol Inc. Minimum output power is +17 dBm, with a variation across the tuning range of ±1.25 dB for the 100-180 MHz unit, Model VCO171. Minimum output power for the 180-320 MHz oscillator, Model VCO172, is +15.5 dBm with a tuning range variation of ±1.25 dB.

A portion of the RF signal from these oscillators is fed to the digital circuitry from a -10 dB coupler in the output line. There, an ECL prescaler (11C90) followed by five programmable decade dividers (10010) form a divide by N counter, with division ratio selectable by the setting of a BCD encoded thumbwheel switch. The switch gives a direct reading of the selected frequency in MHz. The minimum frequency increment is 10 kHz.

A second input to the digital board provides a 5 MHz reference signal. It can either be derived from a precision external source or by frequency division from the internal 100 MHz clock oscillator. The mode is selectable by a front panel switch. The reference 5 MHz is divided in frequency by a factor of 500 through a series of 10010 counters. This reference signal and the divided-down VCO output are applied to an ECL digital phase frequency detector, which produces a pulse train whose mark-space-ratio is dependent on the relation phase of the two input signals.

A high gain loop integrator follow, with characteristics that result in a second-order closed loop transfer function for the system ($\omega_n \sim 2\pi \times 10^3$ rad/sec, $\zeta \sim 6.0$). Circuitry to provide an indication of a phase locked condition for the synthesizer loop is also provided.



NOTES: 1. ECL EMITTER RESISTORS ARE 330Ω
2. +5V ON 1 & 16, GND ON 8, UNLESS OTHERWISE SHOWN.

F[redacted]E B[redacted]

The VCO output is fed via the -10 dB coupler, an attenuator and a low-pass filter to a gain controlled amplifier (TRW Type CA860), with 32 dB maximum gain. The gain control circuit is shown in Figure B3, as part of the ALC loop amplifier schematic.

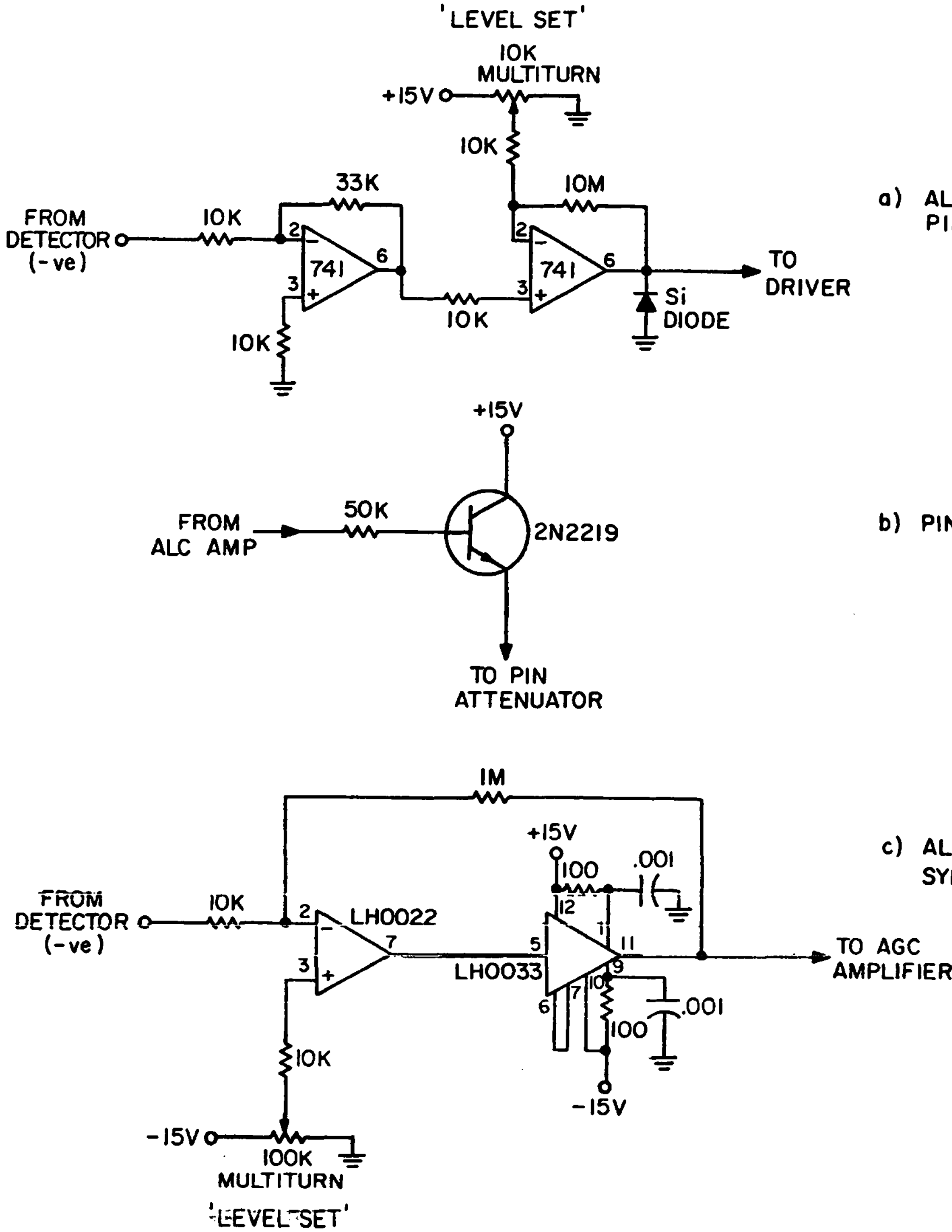
B.2.2 Reference Oscillator, Comb Generator and Filters

The primary frequency reference for the T3 module test set is a 100 MHz crystal oscillator made by Vectron Laboratories. Frequency accuracy is ±.001%, stability as a function of temperature ±.002% from 0°C to 50°C, aging rate 3×10^{-8}/day and output power +7 dBm in 50 Ω. The output is attenuated, before being fed to a power amplifier and power divider of a similar nature to that used as the synthesizer output buffer. Power from the divider is +23 dBm for both outputs. One output signal is used as a reference for the synthesizer, the other drives a step recovery diode comb generator (HP 5082-33002A). The 1200 MHz and 1800 MHz components of the comb are selected by band-pass filters and amplified to give a nominal power level for the reference signals of +10 dBm.

B.2.3 L-Band Sweep and Phase Lock Oscillators

As shown in Figure B4, the phase lock oscillators are analogue phase locked loops of conventional design. The VCO's are Avantek VTO8090 and VTO8150 varactor-tuned transistor MIC oscillators, respectively.

A portion of each VCO output signal is fed via a -10 dB coupler to the phase detector circuit. The phase detector comprises in-phase and quadrature channels, implemented with the aid of a quadrature hybird (Anaren 10265-3), a microstrip in-line power divider and two high performance double-balanced mixers. The output of the in-phase detector is fed to the loop integrator, which has characteristics which result in the following closed loop parameters: $\omega_n \approx 6.3 \times 10^4$ rad/sec, $\zeta \sim 0.707$.



a) ALC AMPLIFIER FOR PIN DIODE

b) PIN DIODE DRIVER

c) ALC AMPLIFIER FOR SYNTHESISER AGC (CA860)

FIGURE B3

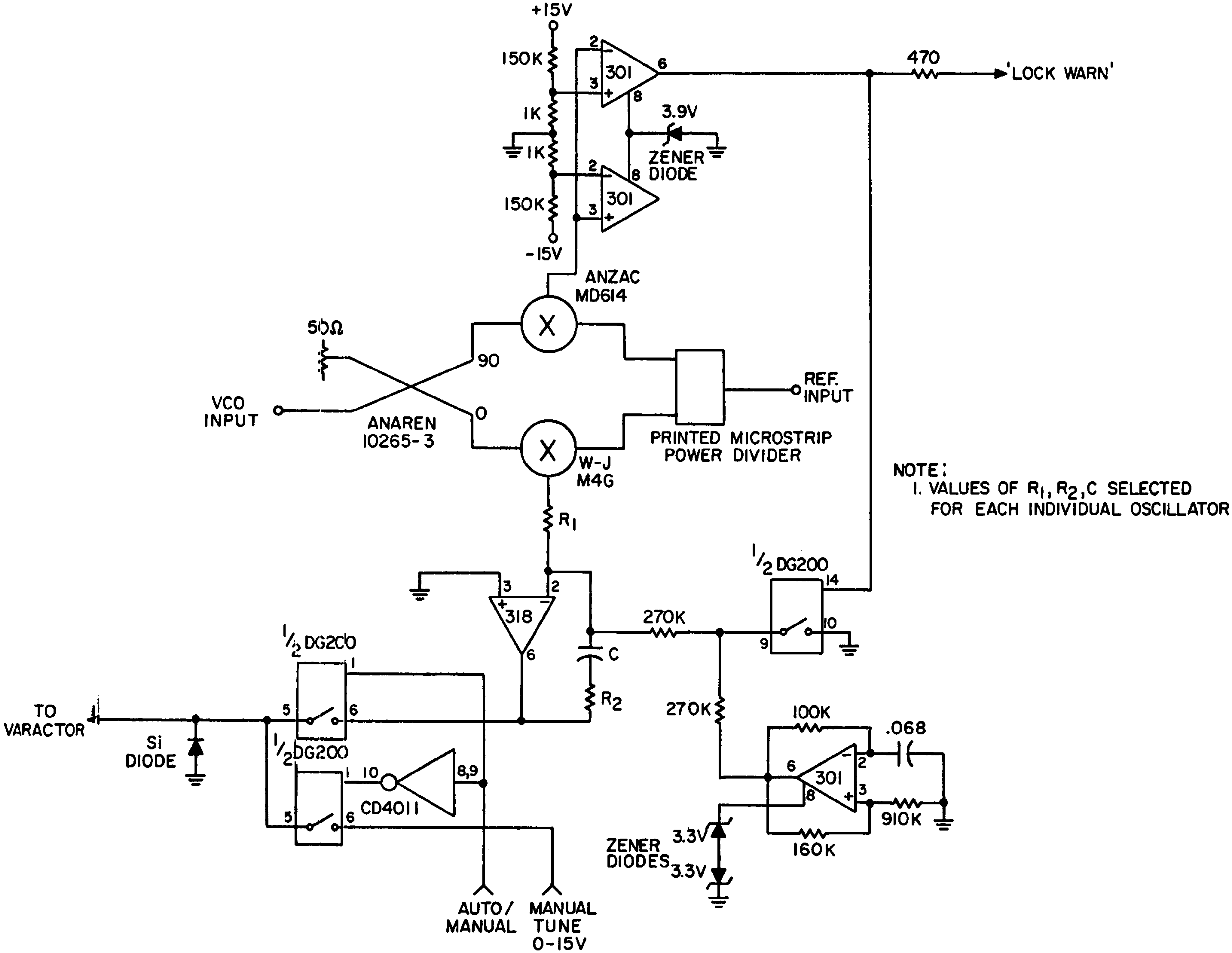

+15V
150K
301
1K
1K
150K
-15V
470
'LOCK WARN'
3.9V
ZENER DIODE
ANZAC MD614
50Ω
90
0
VCO INPUT
ANAREN 10265-3
REF. INPUT
PRINTED MICROSTRIP POWER DIVIDER
W-J M4G
NOTE:
1. VALUES OF R1, R2, C SELECTED FOR EACH INDIVIDUAL OSCILLATOR
R1
1/2 DG200
318
C
R2
270K
270K
100K
.068
910K
160K
TO VARACTOR
Si DIODE
CD4011
ZENER DIODES
3.3V
3.3V
AUTO/ MANUAL
MANUAL TUNE 0-15V

FIGURE B-4

The quadrature phase detector output is fed to a threshold detector, the output of which is high when the absolute value of the quadrature phase detector output is less than 0.1 volts.

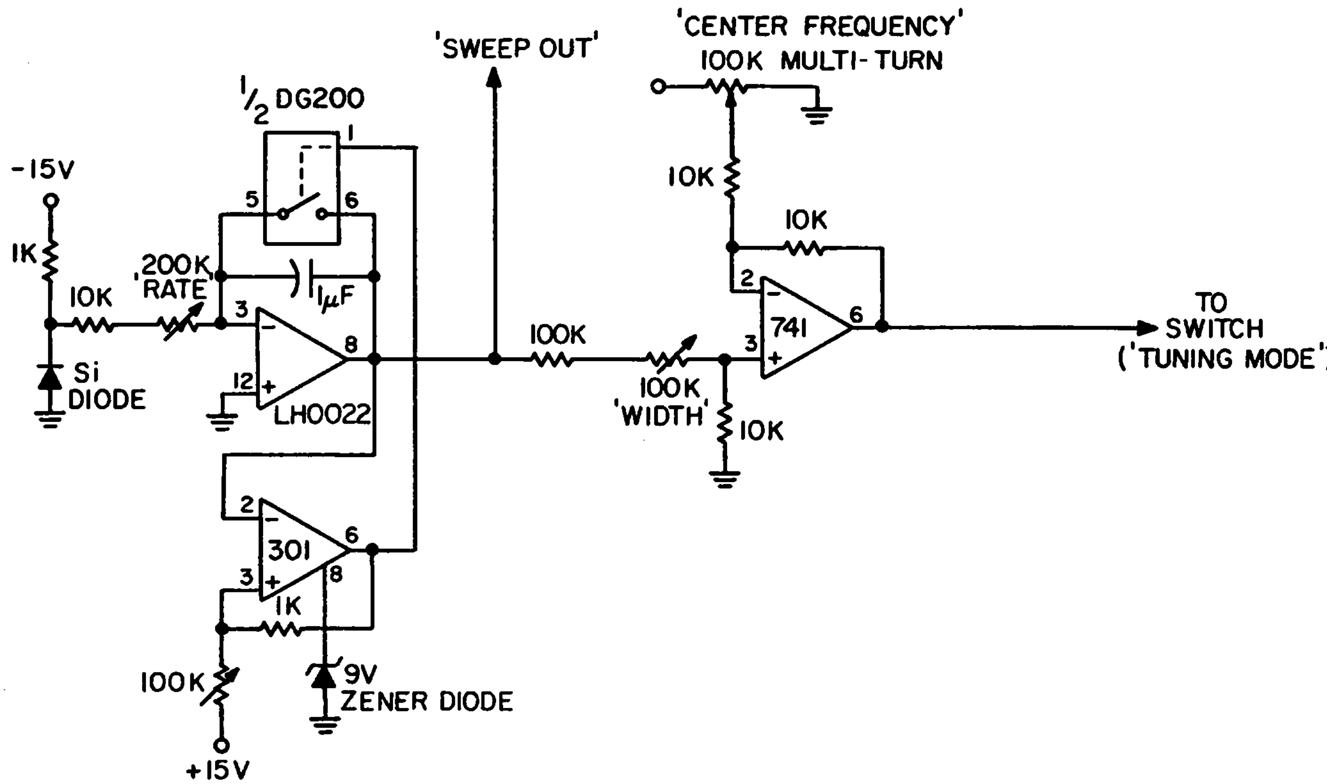
With the loop initially open (IC5a closed, IC4 open), the quadrature phase detector output is less than the threshold, switch IC5b is open and the lock warn output is high. Under these conditions, a square wave of period 60 msec, and peak-to-peak amplitude ±4 volts, output from IC7 causes the output of the integrator IC3 to ramp up and down between the supply rails at a rate of approximately 1.0 volt/msec. When the loop is closed, by opening IC5a and closing IC4, the VCO is swept through its tuning range by the ramp at the output of IC3. When lock is acquired, the quadrature phase detector output increases above the threshold, sending the lock warn output low and closing switch IC5b, thus removing the sweep signal.

Sweep tuning is possible with the loop in the open condition by applying a positive voltage between 0 and 15 V to the "manual tune input". Auto-phase-lock to manual operation switching is achieved by applying a voltage to the "auto/man" input - 0 volts results in locked operation, +15 V results in the open loop or manual condition.

The Avantek VCO's give nominal output power of +13 dBM for the VTO8090 and +10 dBm for the VTO8150. The output signals pass through the -10 dB couplers, amplifiers and ALC attenuators to the front panel. Nominal output power levels, at the front panel, for both the VTO8090 and the VTO8150 are -10 dBm.

B.2.4 Swept Frequency Signal Source Driver

Figure B5 is a schematic of the sweep drive circuitry. A sawtooth waveform of 0 to 10 volt amplitude is generated by the combined integrator (IC1) and threshold detector (IC2). The repetition rate may be varied by adjusting the variable resistor R_{rate}. The following amplifier stage (IC3) provides variable output offset facilities and variable gain, enabling the center


'SWEEP OUT'
'CENTER FREQUENCY'
100K MULTI-TURN
1/2 DG200
-15V
1K
10K
200K
'RATE'
1μF
Si
DIODE
LH0022
100K
'WIDTH'
741
TO
SWITCH
('TUNING MODE')
301
9V
ZENER DIODE
+15V

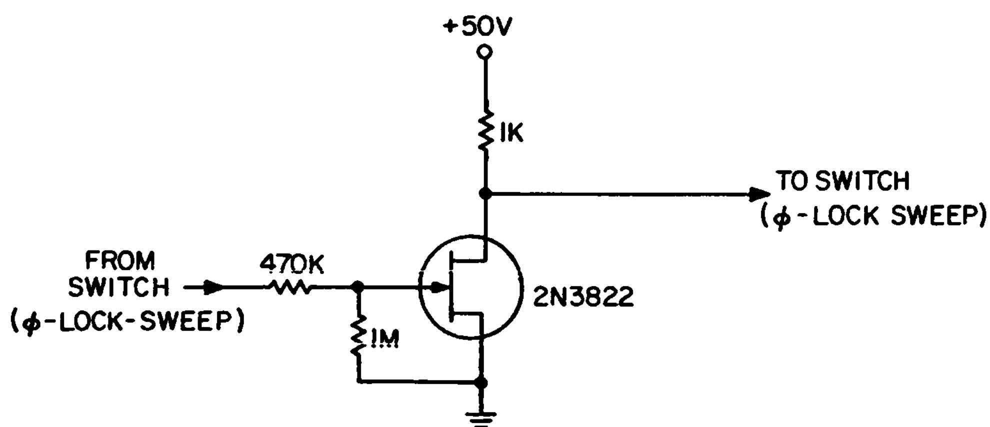

+50V
1K
TO SWITCH
(φ-LOCK SWEEP)
FROM
SWITCH
(φ-LOCK-SWEEP)
470K
1M
2N3822

FIGURE B5

frequency of the VCO sweep and the frequency deviation to be independently adjusted. The FET output amplifier provides sufficient voltage gain to allow the VCO to be tuned through its full frequency range.

B.2.5 ALC Amplifier

Figure B3 shows a schematic diagram of the ALC amplifier used both for the control of the sweeper oscillator output level and for synthesizer oscillator output level control.