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LIST OF ILLUSTRATIONS

ERRATA SHEET

Pg. 2 1.2 SPECIFICATIONS -- Scope Outputs

Change "10K∿" to "600∿" (1st line and last line).

Insert the number "1" in front of VRMS in 3rd line.

Pg. 23 4.4-9 -- The first paragraph should read as follows:

"The Low Pass Filter is of 7 pole 500 Hz Butterworth active RC design with three unity voltage gain amplifiers composed of Q42 thru Q47 and the B scope amplifier. Since the absolute value detector outputs DC, which is used to drive the calibrate meter but is not itself a distortion component, the absolute value detector is AC coupled into the LP filter by C35. The three unity voltage gain stages each generate two complex conjugate poles, each pair of successively higher Q. The real valued pole is realized by R150 and C64 of the B scope amplifier."

-- In the second paragraph

Change "four" to "three" in first line

Pg. 23 4.4-11

-- In third paragraph

Insert "R179" after R150. It should read "..feedback resistors R150, R179, and R149 are ..."

-- Include this sentence at the end of 3rd paragraph:

"R179 is adjusted to give a B scope output of 1V at full scale IM levels."

Pg. 26 -- Bottom of 1st column -- in (5), the second line should read "Jour. of Aud. Eng. Soc."

Same change to "Aud." in 2nd line of (6).

Pg. 28 -- In Fig. 5-5 the vertical scale of the chart should be labeled "ms (milli-seconds)." The horizontal scale should be labeled "Hz".

The third paragraph, 10th line should have the number "1" inserted on the blank, and the sentence should end after the word "Volts." The rest of the paragraph should be deleted.

Pg. 29 -- Bottom of 1st column

(9) 2nd line should change "And" to "Aud."

Pg. 41 -- In (8) include this sentence:

"Adjust R179 for a 1V rms output at the B scope output."

Replace "(9)" from present location and put it in front of sentence "Disconnect the frequency..."

Delete the complete (12) paragraph.

Delete numbers "(10)", "(11)", and "(12)".

On Figure 6-3 under Probable Defects column, 3rd box -- delete "Q40".

1.1 GENERAL

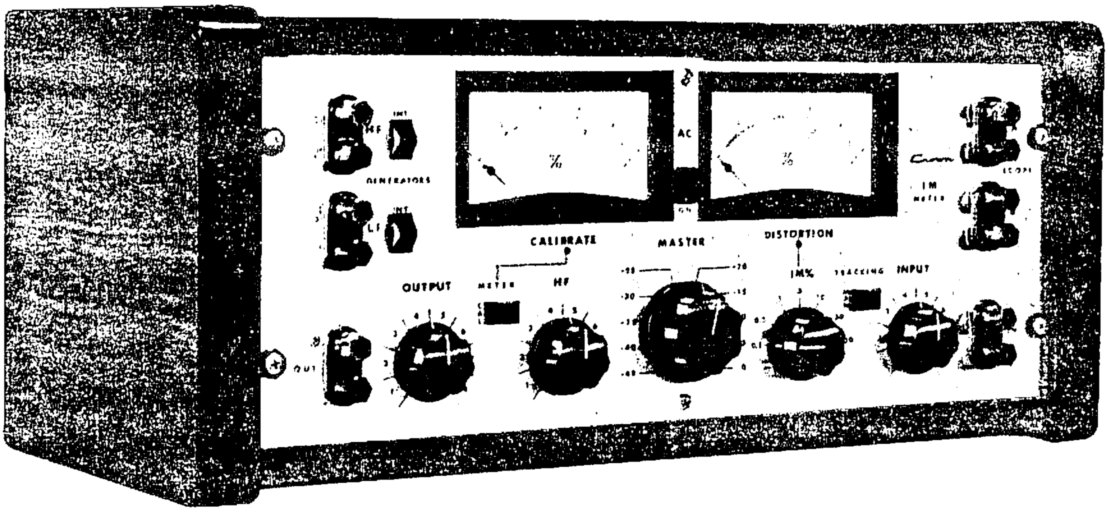


FIG. 1 - 1 IMA INTERMODULATION DISTORTION ANALYZER

The IMA is an intermodulation distortion analyzer of SMPTE type having extremely low residual distortion and high sensitivity. Speed and ease of use have been optimized by employing a master "tracking" attenuator which simultaneously adjusts input and output levels in 5db increments through a 45db range allowing 10 measurements to be made with only one set-up adjustment period. Set-up time has also been minimized by using two meters, one for set-up and one for reading distortion.

These characteristics make the IMA ideally suited for both the design lab and the production line. Spanning 45db in ten steps, (a power ratio of 31,623), ten readings can be taken in less than a minute, all of which can be observed with an instrument residual $\leq$ .005% IM distortion. With a sensitivity of 17mV (HF signal) it is possible to check a power amplifier driving 8Ω (4:1 LF-HF ratio) at 900 micro watts, a power level which is below the lowest crossover notch distortion region in a direct coupled solid state amplifier, (which may be as low as 25mW.)

The circuitry is all silicon solid state employing no inductors except the power transformer. The internal generators, 60Hz and 7KHz, are both Wien bridge FET controlled oscillators. The 60Hz generator being an oscillator eliminates the line voltage LF test signal amplitude regulation-distortion problems that plague all conventional IM meters where the LF signal is derived from the AC mains. FET controlled oscillators do not suffer from microphonics or aging drift as do lamp type ALC systems that are found in many RC bench oscillators.

The test signals are mixed at the virtual ground of an ultra-low distortion op-amp eliminating test generator interactions and critical bridge type mixers while providing inputs common to ground having gain and large output amplitudes. Distortion in mixer amp is typically $<$.001% even with 50V p-p outputs (25V p-p when terminated in 600 Ω). This large output capability along with the high sensitivity of the analyzer allows "tracking" type measurements to be made on unity gain devices.

Both the high pass and the low pass filters are 7 pole Butterworth RC active types having no hum sensitive inductors. The corner frequencies are 2KHz and 500Hz respectively where the filters begin their 42db/octave roll-offs.

All of the circuits employ liberal amounts of negative feedback insuring stability of all vital characteristics.

The power supplies are of the series regulated type providing highly stable + and - 30VDC supplies for all of the circuits.

CROWN guarantees this equipment to perform as specified. CROWN also warrants the components and workmanship of this equipment to be free from defects for a period of 1 year from date of purchase.

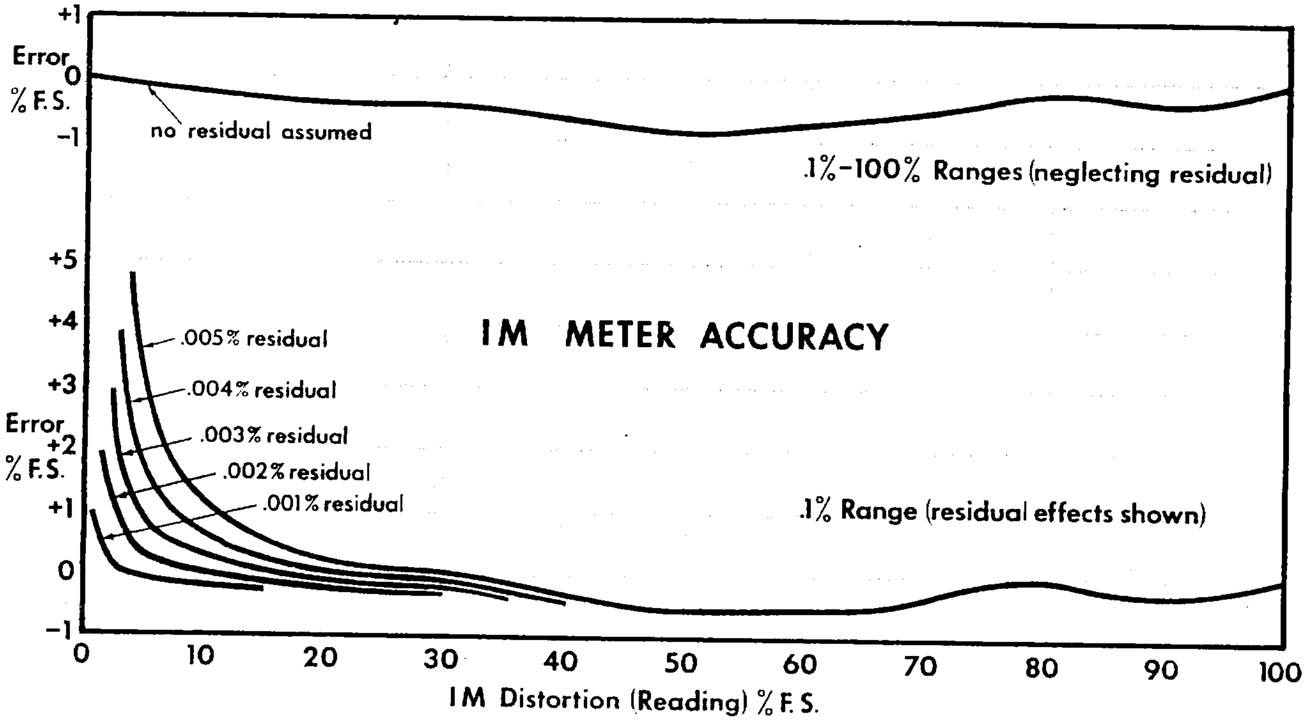
This warranty does not apply to fuses, pilot lamps and/or component or equipment damage due to, misuse, shipping damage, accident, or if the serial number has been defaced, altered, or removed.

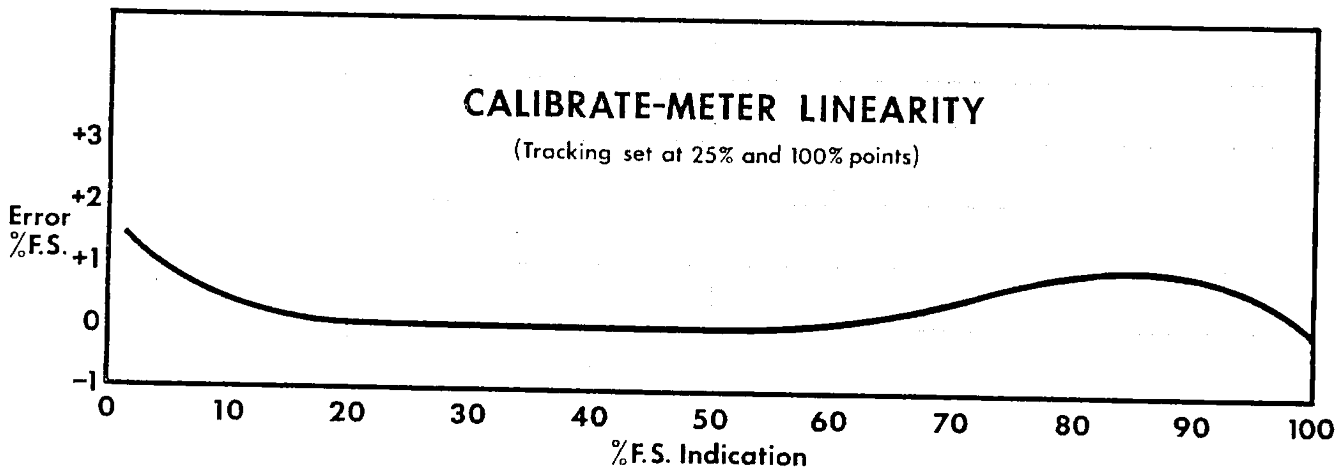
The model 7C enclosure is available to house the rack mounting IMA (See fig. 1-1). The 7C is a rugged walnut finished furniture piece with decorative protective black vinyl end trim. Its dimensions are 8 7/8" H x 21 3/8" W x 10 3/4" D. Its net weight is approximately 13 lbs.

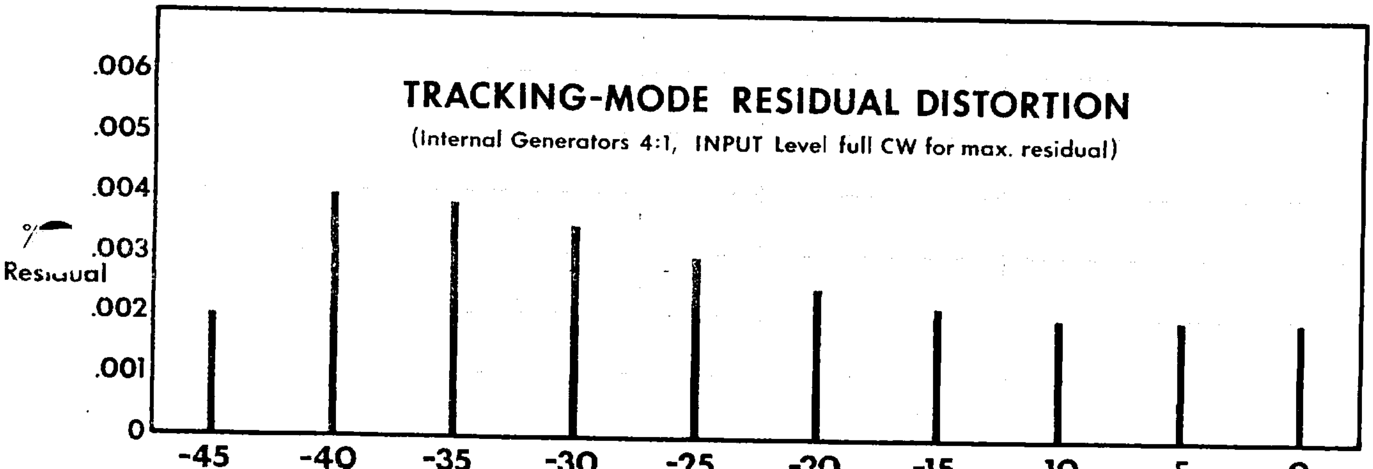
1.2 SPECIFICATIONS

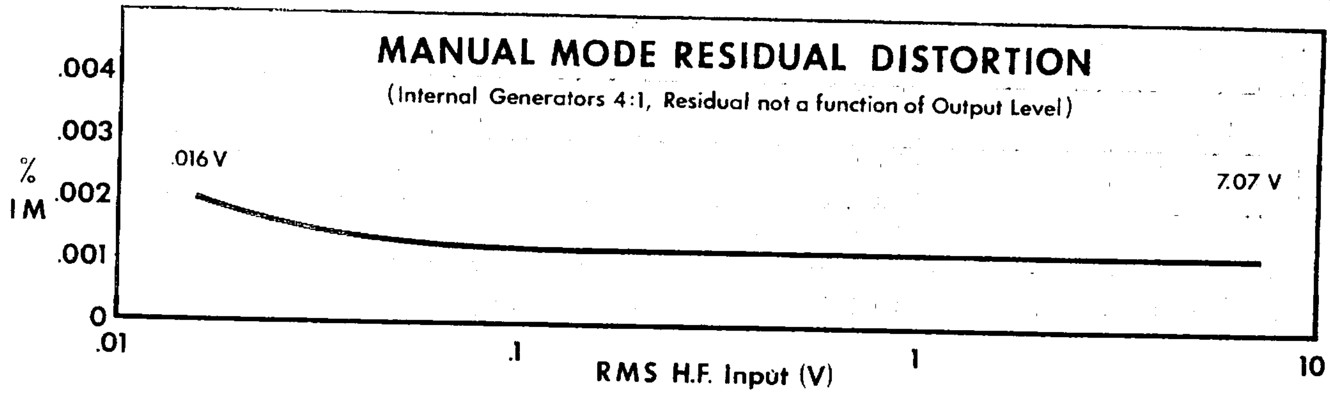
IM Ranges:	0.1, 0.3, 1, 3, 10, 30, 100% F.S. on separate meter.
Residual IM:	Less than .005% with internal generators. (See graphs.)
Demodulation:	Average value sensing (full wave).
Accuracy of IM Scales:	5% of F.S. ±.005% residual. (See graphs).
Range of Tracking:	45db in 5db (±1%) increments. (See graphs for use restrictions.)
Necessary HF Input:	17m V minimum.
Input Impedance:	100K (max) and 45K (min) depending on setting of INPUT level control.
LF/HF Voltage Ratio:	Internal generators continuously adjustable from ∞:1 to 1:1 by HF level on front panel. Ratio is read by reading voltage on CALIBRATE meter with METER switch in the "OUT" position.
LF:	60Hz Low distortion oscillator (< .1% THD) syncronized with AC line. (May be modified for 50Hz operation.) External oscillator input (10-150Hz) provided (one terminal grounded, 5.6K internal impedance).
HF:	7KHz ±1% Low distortion oscillator. External oscillator input (2.5K-20KHz) provided (one terminal grounded, 2.4K-6K internal impedance.)
Meter Laws:	Full wave average value reading.
Output Level:	25V peak (max) for any generator combination internal or external. (12.5 V peak into 600Ω) Level is adjustable by a 40db attenuator (in 10db steps) concentric with 15db variable (fine) control. In the Manual position of the TRACKING switch the MASTER attenuator is cascaded with the 40db OUTPUT attenuator for an additional 45db attenuation range with 5db increments. Unattenuated max. voltage gain from an external generator is 10db min, (inverting).
Output Impedance:	600Ω ±1% for all settings of OUTPUT level controls and MASTER attenuator.
Scope Outputs:	SCOPE A output shows HF envelope at approximately 1.4VRMS and 10KΩ output impedance. May be shorted w-o effecting measurements. SCOPE B output shows demodulated and low-passed IM distortion signal at 1.0 VRMS corresponding to F.S. on the IM meter on any distortion range. Signal is undistorted with 10K output impedance which may be shorted w-o effecting measurements.
Connectors:	Dual binding posts (6) and 3 wire power cord.
Power Requirements:	120VAC ±10% 60Hz. Draws approximately 10W. (May be used 50Hz if 60Hz oscillator sync disconnected or if modified for 50Hz Oscillation.)
Rotary Controls:	OUTPUT level controls, HF level control, MASTER attenuator switch, IM range, INPUT level control (continuously adjustable pot).
Slide Switches:	EXT-INT HF GENERATOR, EXT-INT LF GENERATOR, IN-OUT METER, AC power, MAN-TRK TRACKING- allows INPUT and OUTPUT level controls to remain set when testing over a range of levels using the MASTER attenuator (automatically tracking the two adjustments in 5db increments for 45db).
Semiconductor Complement:	58 bipolar transistors, 13 diodes, 2 junction FETS and one zener diode.
Size:	19" (rack mtg.) width, 7" height, and 8-3/8" depth (from mtg. surface)
Weight:	12 lbs. (25 lbs. with cabinet)
Color:	Bright Anodized 1/8" Scratched Aluminum front panel with black anodized chassis.

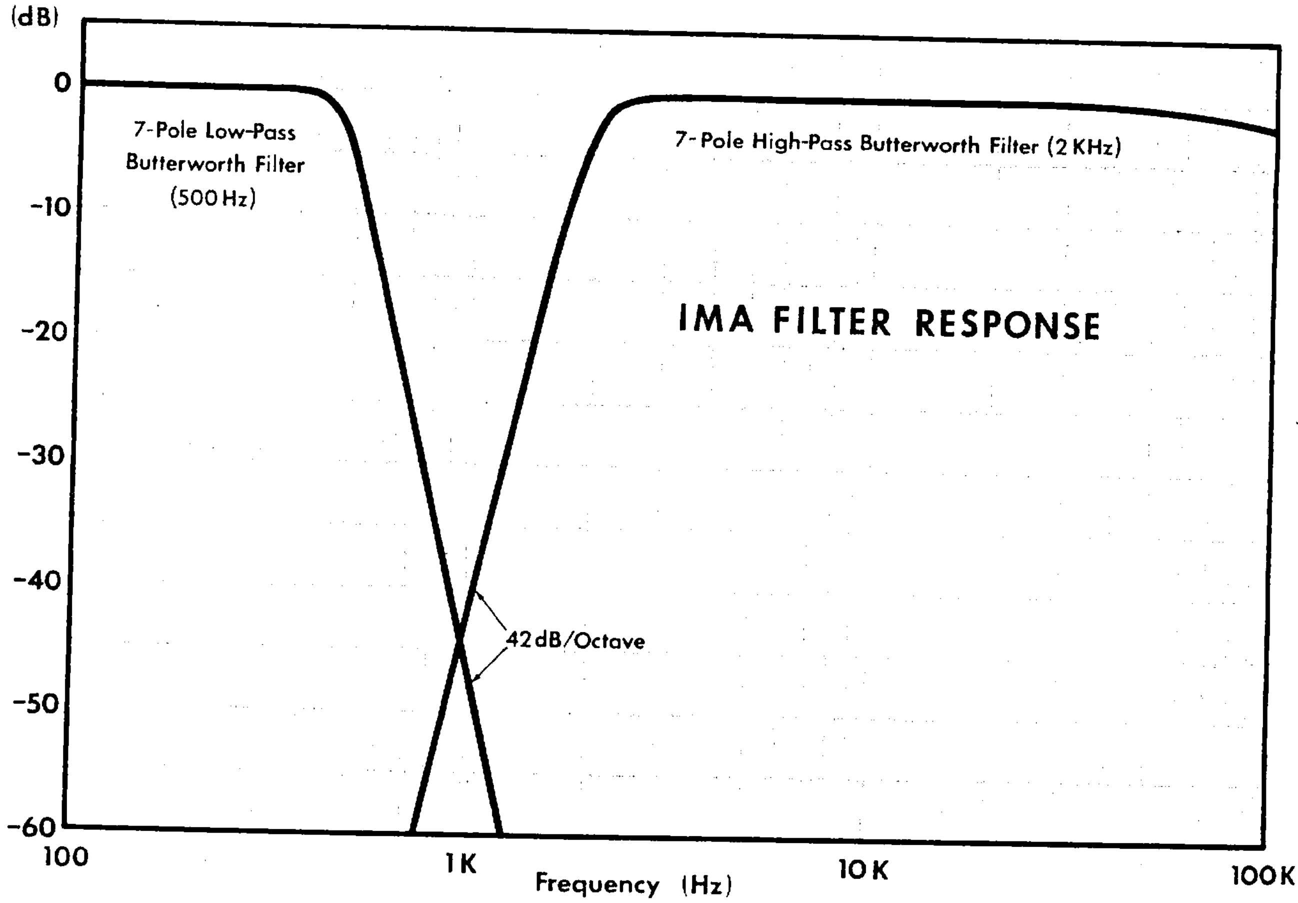
TYPICAL PERFORMANCE GRAPHS

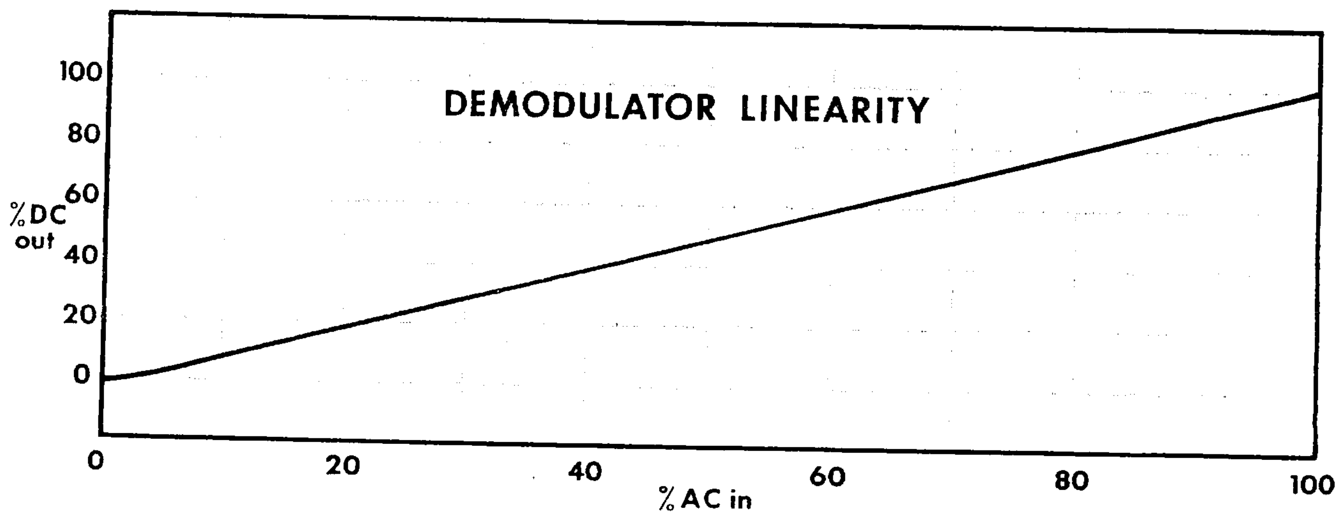


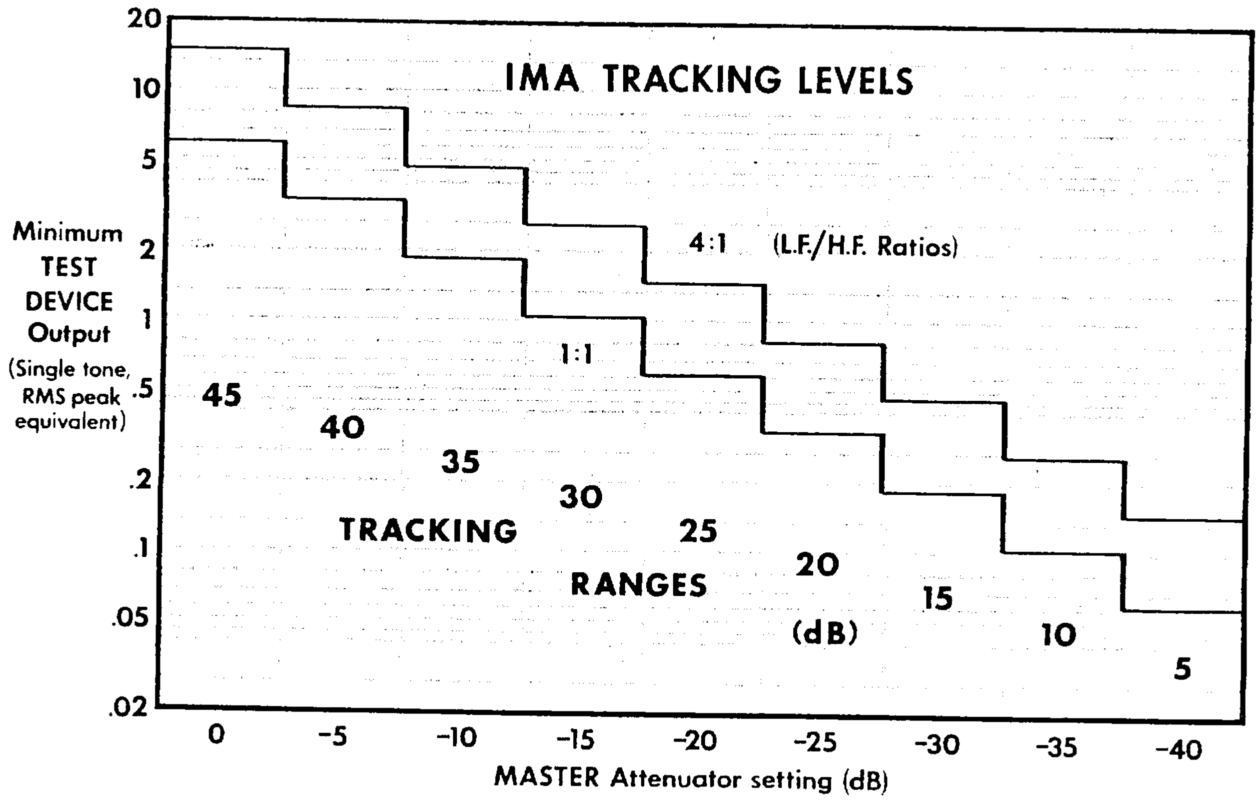





MANUAL MODE RESIDUAL DISTORTION
(Internal Generators 4:1, Residual not a function of Output Level)
.004
.003
% IM .002
.001
0
.016 V
7.07 V
.01
.1
1
10
RMS H.F. Input (V)


IMA FILTER RESPONSE
(dB)
0
-10
-20
-30
-40
-50
-60
7-Pole Low-Pass Butterworth Filter (500 Hz)
7-Pole High-Pass Butterworth Filter (2 KHz)
42 dB/Octave
100
1 K
10 K
100K
Frequency (Hz)


DEMODULATOR LINEARITY
100
80
60
40
20
0
%DC out
0
20
40
60
80
100
%AC in



Section 2

2.1 UNPACKING

As soon as the analyzer shipment is received, please inspect for any damage incurred in transit. Since the unit was carefully inspected and tested at the factory, it left the factory unmarred. If damage is found, notify the transportation company immediately. Only the consignee may institute a claim with the carrier for damage during shipment. However, CROWN will cooperate fully in such an event. Be sure to save the carton as evidence of damage for the shipper's inspection.

Even if the unit arrived in perfect condition as most do, it may be advantageous to save the packing materials. They will prove valuable in preventing damage should there ever be occasion to transport or ship unit. Never ship the unit by mounting it rigidly to the shipping container, as this provides no shock absorption.

2.2 MOUNTING AND ENVIRONMENT

The IMA is designed on a standard rack mounting format (See fig. 2-1). It may be rendered free standing (bench mount) by placing it in a suitable 7" x 19" opening cabinet. The model 7C cabinet is available for this purpose.

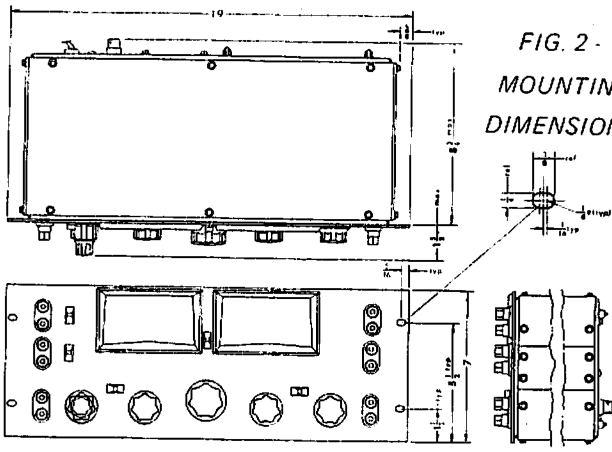


FIG. 2 - 1 MOUNTING DIMENSIONS

The environment should be vibration free if very low residual distortion levels are desired. Vibration modulates level control settings and stray capacitances associated with the input attenuators' wiring—thereby generating an artificially high residual distortion level.

Care should be taken to avoid magnetic fields which have harmonic components which lie in a frequency band of 1KHz around the high frequency test frequency. These fields (from fluorescent light ballasts, power transformers coupled oscillators, etc.) may be picked up in the test leads despite the analyzer's relative immunity to such fields.

2.3 CONNECTING CABLES

The major problem associated with ultra-low distortion testing is pollution which may be of three basic origins.

1. *External fields (usually magnetic) which have high frequency components in the detection band of the analyzer* are picked in the test cables polluting either the test signal and/or the output of the device under test. Such fields are frequently found around power transformers and may be difficult to avoid especially if the offender is in the device under test. The following hookup cabling (fig. 2-2) is recommended to minimize such problems.

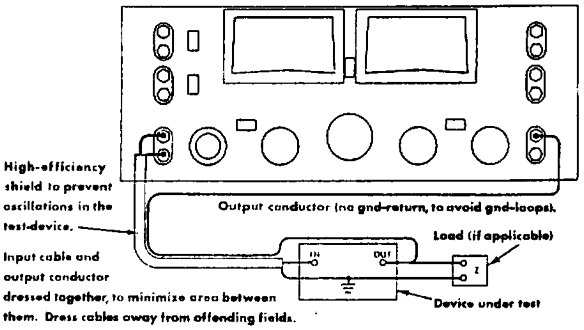


FIG. 2 - 2 TEST CABLE CONSTRUCTION

If the device under test is grounded it will probably be necessary to disconnect the chassis to AC mains ground of the IMA to prevent a ground loop. This can easily be accomplished by placing a 3wire to 2wire adapter plug on the AC power plug.

2. *Bad Connections* joining the device under tests to loads. This problem is most obnoxious on power amplifiers where large currents are flowing through connectors. The metal-to-metal interfaces of the connectors if not tightly joined will act as a non-linear resistances distorting the signal going to the load and the devices output signal in cases where its source impedance is not infinitesimal. For this reason it is always preferred to connect the analyzer signal input lead directly to the device's output and not anywhere along the cable going to the load. In severe cases soldering of test cables to the device under test may be warranted.

3. *R F* signals if sufficiently large may be partially demodulated by the input amplification circuitry causing high residual distortion. Defective circuitry under test containing RF oscillations in the output will be detected and indicate higher than normal distortion levels. Such a characteristic makes an IM meter an especially potent production line QC instrument.

3.1 CONTROLS

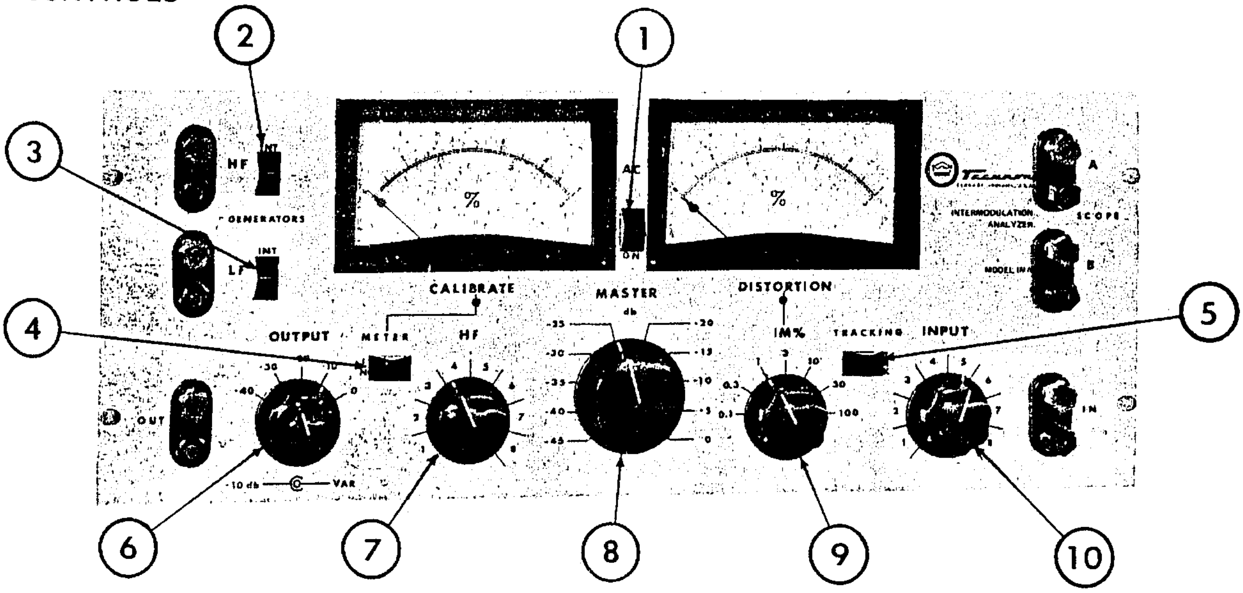
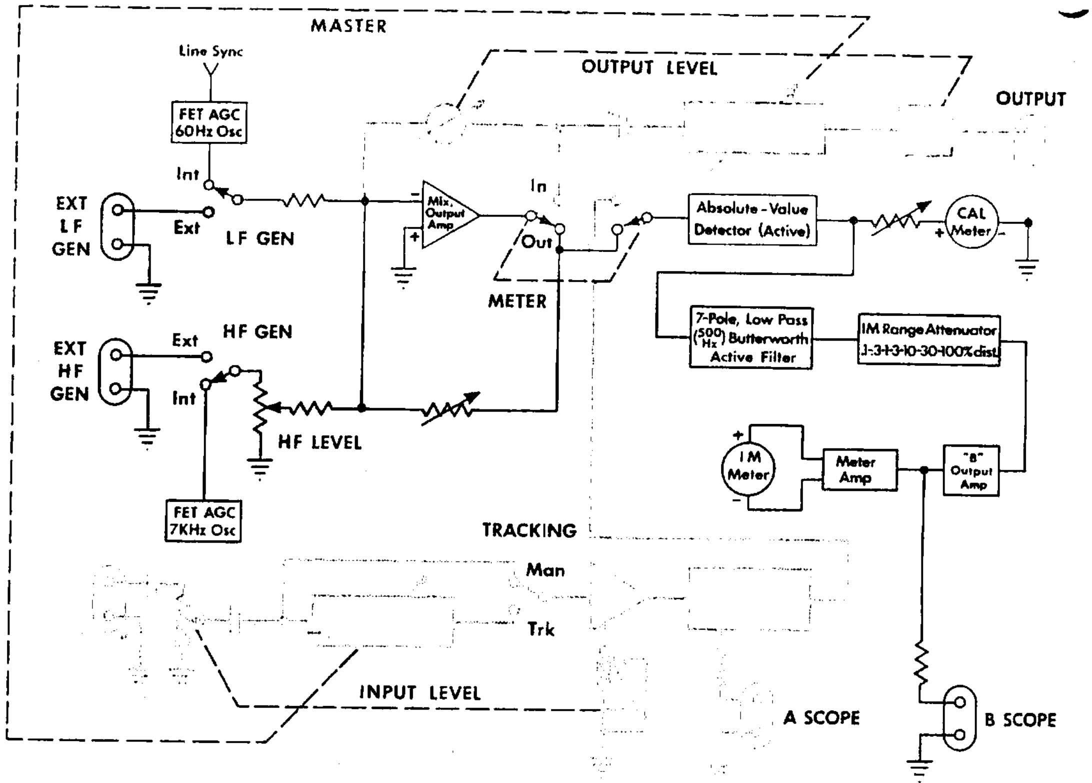


FIG. 3-1 CONTROLS

1. AC power switch-in the up position power is applied and meters will light.
2. INTernal - EXTernal High Frequency GENERATOR switch-in the INTernal position the internal 7KHz oscillator is connected and in the EXTernal position the generator is disconnected and turned off and connection is made to the adjacent pair of binding posts to allow use of an external generator.
3. INTernal - EXTernal Low Frequency GENERATOR switch-in the INTernal position the internal 60Hz oscillator is connected and in the EXTernal position the internal oscillator is disconnected and connection is made to the adjacent pair of binding posts to allow use of an external generator.
4. INput - OUTput METER switch - selects function of calibrate meter. In INput position (see fig. 3-2b) the meter reads the average amplitude of the high frequency input signal (100% or full scale is calibrated to read IM distortion). In the OUTput position the calibrated meter reads the output amplitude of the test generators (see fig. 3-2a) at the output of the mixer amplifier. (Note that the OUTPUT level controls and the output signals are defeated in this mode.)
5. MANual - TRacKing TRACKING switch - is used to select the mode of IM measuring preferred. In the MANual mode the unit functions as a conventional IM meter (see fig. 3-2a) where the OUTPUT and INPUT level controls must be readjusted for every measurement taken. In the TRacKing mode the input and output levels are simultaneously adjusted (see fig. 3-2c) to maintain the calibrated condition which is set upon the first measurement to be made.
6. OUTPUT level controls - The large outer knob attenuates the output in 10db steps through 40db while the inner knob serves as a vernier level adjust having at least 15db of range.
7. High Frequency level control - adjusts the amount of high frequency signal in the output without affecting the low frequency signal gain.
8. MASTER attenuator - attenuates the output level while simultaneously increasing the gain in the TRacKing mode of operation (see fig. 3-2c). Serves as an output level attenuator with 5db steps through 45db when in the MANual mode of operation.
9. IM range control - selects the full scale distortion sensitivity of the IM distortion meter. Ranges of 0.1%, 0.3%, 1%, 3%, 10%, 30%, and 100% IM distortion may be selected.
10. INPUT level control - adjusts the gain of the input amplifier and the input signal attenuation to allow a wide range of input voltages to be adjustable to the calibrate level in order to read distortion.

A. METER switch in OUTPUT position.



B. METER switch in IN position.

TRACKING switch in MANual position.

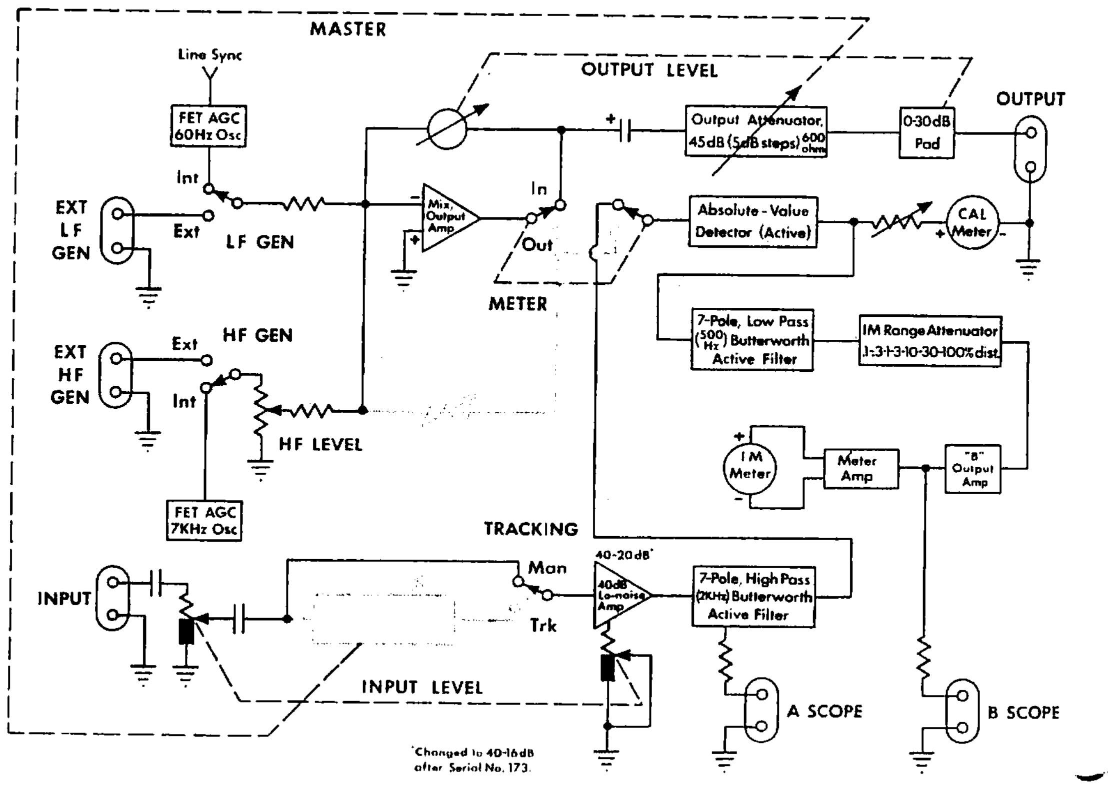


FIG. 3-2 BLOCK DIAGRAMS

C. *As B except*

TRACKING switch

in TRacKing position.

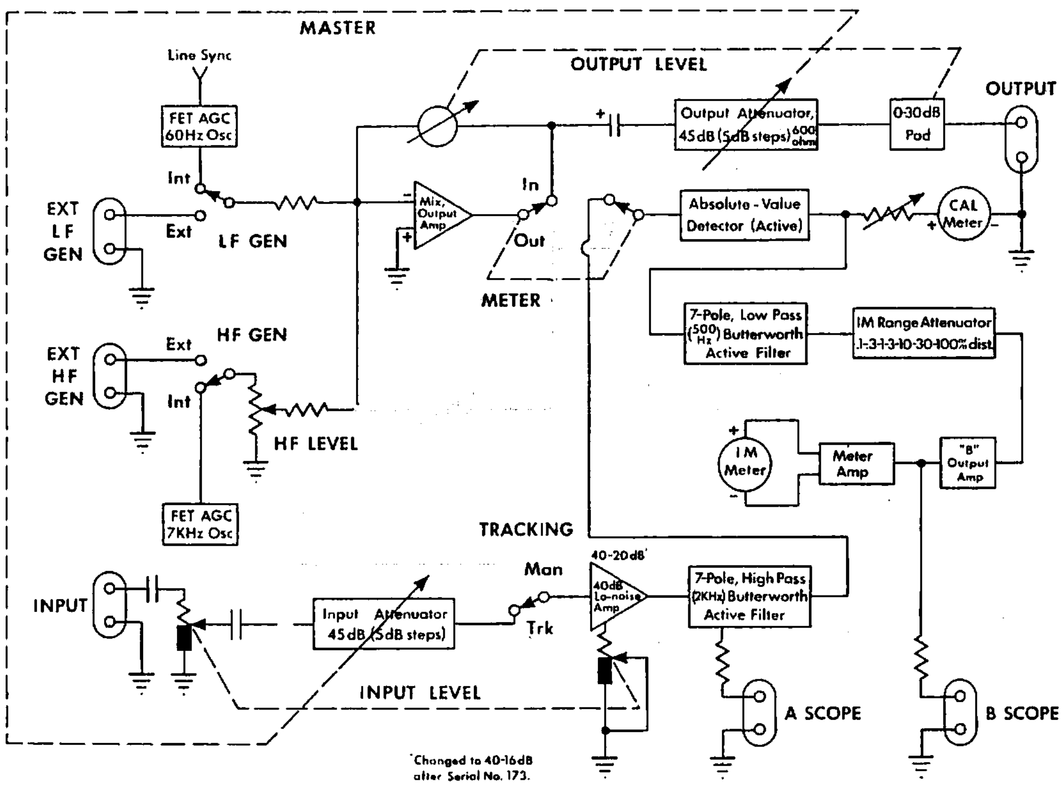


FIG. 3-2 BLOCK DIAGRAMS (concl.)

3.2 OPERATION

After plugging in the analyzer and turning on the AC power, the first step in using the analyzer is to set up the LF:HF generator ratio as desired for the tests to be performed (see section 5). Proceed as follows:

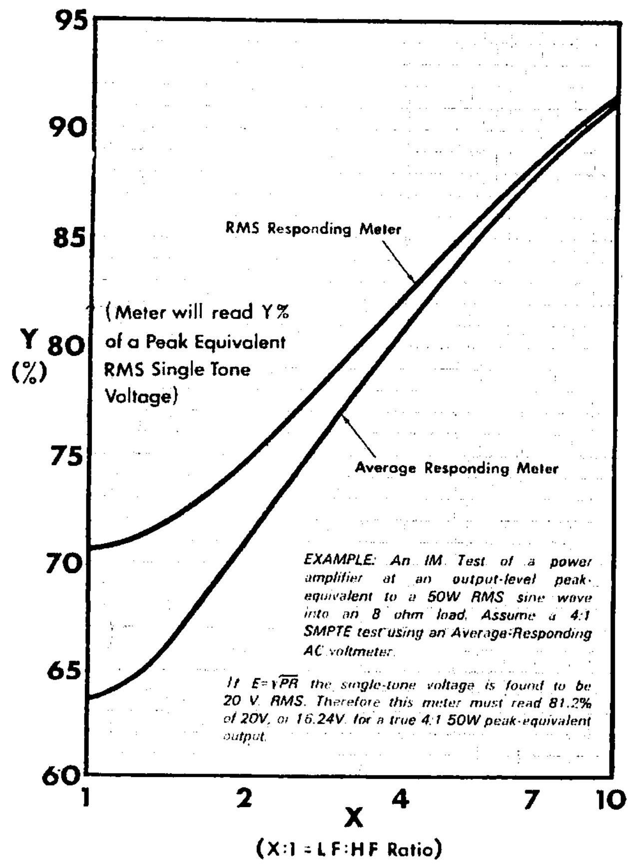
1. *Set the METER switch to OUTput.*
2. *Remove any HF input.* This can be done in a variety of ways depending upon the circumstances. If no external HF generator is being used setting the INT - EXT HF GENERATOR switch to EXTernal will suffice. If an external HF generator is being used its output may be turned off or disconnected. In any case turning the HF level control completely CCW will remove the HF input.
3. *The CALIBRATE meter should now read 100%,* f.s., if the internal low frequency generator is being used (100% adj. is internal). If an external LF generator is being used the external generator's output should be adjusted to give full scale deflection on the meter.

   NOTE: A 65 % IM meter reading is normal, and indicates proper calibration (see section 6.1.3).
4. *Remove the LF input.* This may be accomplished in a similar manner as was removing the HF input in 2 except that there is no LF level control.
5. *Apply the HF input.* If the internal oscillator is desired place the INT - EXT HF GEN switch in the INTernal position. If not, connect the external oscillator to the HF input and place the HF GENERATOR switch in the EXTernal position.
6. *Adjust the HF level control for the desired LF-HF Ratio.* If, for example, the desired ratio is 4:1 set the HF level such that the meter reads 25% of full scale. The rule is: $\text{Ratio} = \frac{\text{LF \% of scale}}{\text{HF \% of scale}}$ For 1:1 the HF level would then be set to give a full scale deflection.

   NOTE: The Calibrate meter has been tracked at the factory to give precise reading at 25% and 100%. Typical meter error is shown in the graphs following chapter 1.2.
7. With the LF-HF ratio adjustments now complete *switch the METER switch to the INput* position.
8. The test signal is now prepared for application to the device to be tested. Only its level need be adjusted. Depending on which is desired, *adjust the OUTput level for the specified input or output voltage on the device being tested.*

NOTE: Most IM measurements are specified in terms of RMS peak equivalent single tone sinusoidal measurements. To interpret a given meter reading the type of meter being used must be known. The most common type of audio voltmeter is an average responding meter whose scale or digital readout reads RMS equivalent voltage of a sinusoid. Figure 3-3 shows both a table and graph indicating the appropriate scale factor by which the desired RMS equivalent sinusoidal voltage must be multiplied to get the correct meter indication for a given LF-HF ratio. Also shown are the correction factors for a "true" RMS responding meter.

FIG. 3-3 METER CORRECTIONS FOR EQUIVALENT SINGLE TONE PEAK VOLTAGES



Devices having very high voltage gains such as sensitive microphone preamplifiers may require an external attenuator to adequately reduce the test signal output level. The IMA is compatible with all standard 600Ω audio bench attenuators.

9. *Select the desired mode of IM measuring, MANual or TRacKing* on the TRACKING switch. If the manual mode is desired proceed to step 10.

If the TRacKing mode is desired the test device must have sufficient output for the maximum desired CW MASTER attenuator position. These sensitivities are shown in fig. 3-4. Also the voltage gain must be greater than -1.3db for a high input impedance device, and greater for lower input impedance devices (proportional to loading of 600Ω generator impedance). Device must be capable of this much or more RMS voltage.

Devices not having a voltage gain that is roughly independent of their input signal amplitude cannot properly employ the TRacKing mode. Typical of such devices would be compressors, expanders, non-linear wave shapers, etc.

Available Tracking Range (db)	Max. CW Position on MASTER	HF Output ONLY	4:1 total Output	1:1 total Output
45	0	3.03	15.1	6.06
40	-5	1.7	8.5	3.4
35	-10	.955	4.78	1.91
30	-15	.537	2.69	1.074
25	-20	.303	1.51	.606
20	-25	.17	.85	.34
15	-30	.0955	.478	.191
10	-35	.0537	.269	.1074
5	-40	.0303	.151	.0606

FIG. 3-4 TRACKING MODE CAPABILITIES IN TERMS OF MAX. TEST DEVICE OUTPUT VOLTAGE

If step 8 was not executed at the correct setting of the MASTER attenuator due to not having anticipated the requirements of the tracking mode repeat step 8 at the correct MASTER attenuator position.

10. *Adjust the INPUT level control* for a full scale deflection on the CALIBRATE meter.

11. *Set the IM% distortion range switch* to the range that gives the most convenient display. This reading is the distortion being metered.

NOTE: Pinning of the meter is not detrimental -- overtravel is necessary to measure impulse-type waveforms and to give accurate IM readings.

12. *If the TRacKing mode is being employed switch to other desired output levels* (attenuates output by turning CCW) and repeat 11 if the distortion reading is off scale or sufficiently down scale to allow increasing the full scale sensitivity.

4.1 INTRODUCTION

It has been widely recognized for many years that single tone distortion analyses (harmonic) are an inadequate indicator of audible distortions that are generated in sound recording and reproducing processes. There are 5 basic reasons why this is so:

1. Harmonics are not always aurally offensive. In fact they may even be esthetically pleasing.
2. Single sinusoids while they may be convenient to generate, bear little resemblance to typical program materials and do not reveal the complex interactions that occur between widely differing parts of the frequency spectrum in most audio electronic equipment.
3. The correlation between quantity and audibility of distortion products[(1)] is not implicit in a THD (notch type analyzer) measurement which is the universally accepted type.
4. THD measurements may easily be polluted by frequency instabilities as is the case with wow and flutter in mechanical recording and reproducing mechanisms.
5. THD measurements frequently measure noise in a system rather than distortion.

From the equipment manufacturer's and designer's view point harmonic distortion measurements are undesirable for all the foregoing reasons as well as:

1. Present day harmonic distortion analyzing devices have residual distortion levels much greater than some devices to be tested. Quality control in production and design requires an accurate analysis of distortion in any quality audio product.
2. Harmonic distortion measurements are tedious usually requiring a converging sequence of fundamental nulling operations which are time consuming and highly prone to error.
3. Proper evaluation of a product for complete harmonic analysis is a two parametered system requiring distortion to be recorded at various frequencies and signal levels which greatly increases the number of measurements necessary.
4. Failure to evaluate complex interactions between different parts of the spectrum may cause a serious defect to be completely overlooked.
5. *Good* harmonic distortion testing systems are too expensive to allow their use in all manufacturing testing, and sales and servicing areas where they would be needed.

So why then haven't single tone testing methods been totally discarded? Three basic reasons have retained their use: conceptual simplicity; lack of adequate competition from other types of analyzers in terms of cost, quality, and utility; and the Hi-Fi "numbers game". In high fidelity industries it is more appealing to talk harmonic distortion figures because they are typically 2 to 6 times lower than 4:1 ratio intermodulation measurements.

To improve upon these deficiencies a system where two tones are applied to the test device simultaneously is used. Non-linearities in the test device then give rise to both harmonics and sum and difference frequencies. The sum and difference frequencies are audibly dischordant — the IM analyzer measures these more audibly-obnoxious parts.

While two frequencies are far fewer than the vast number of audible frequencies that would be found in, for example a symphony, they do in a simple way provide a test of complex interactions that transpire in many devices in such a circumstance. SMPTE[(2)] type measurements employ a 4:1 low frequency to high frequency voltage ratio where the low frequency may typically be 60Hz and the high frequency 7KHz. Such a frequency ratio gives a highly realistic portrait of musically encountered situations.

The 4:1 SMPTE LF-HF ratio gives rise to an inherent sensitivity[(3)] to high order distortion products in the amplifiers transfer function which gives a pseudo-weighting to the reading which in turn gives better agreement with listening tests.

Frequency instabilities have little effect on IM measurements where the associated filters are broad compared to the instantaneous errors.

SMPTE IM measurements essentially measure a comparatively narrow band of frequencies around the high frequency test frequency which serves to reject low frequency noise such as power supply hum, l/f noise components, etc.

The CROWN IMA performs SMPTE measurements with an ease and sensitivity never before obtained in an IM analyzer. Not only have semiconductors replaced the vacuum tubes of previous designs but changes have been made in circuit techniques to derive many performance advantages available with modern circuit designs.

(1) Shorter, D.E.L. "The Influence of High order products in non-linear distortion" Electronics Engineering (April 1950) p. 152.

(2) Society of Motion Picture and Television Engineers

(3) Callendar, M. V., and S. Matthews "Relations between amplitudes of harmonics and intermodulation frequencies." Electronic Eng. p. 230 (1951).

4.2 BASIC CONCEPTS

In SMPTE type IM measurements two sinusoids of widely differing frequencies, say 60Hz and 7KHz, are summed in a 4:1 voltage ratio respectively, to generate the test signal.

When this signal is applied to a non-linear device the low frequency will modulate the high frequency generating in effect modulation sidebands about the high frequency in a similar fashion as in AM modulated radio wave (see fig. 4-1). This produces frequencies of 6940Hz and 7060Hz (60Hz sidebands) centered about the 7KHz (carrier).

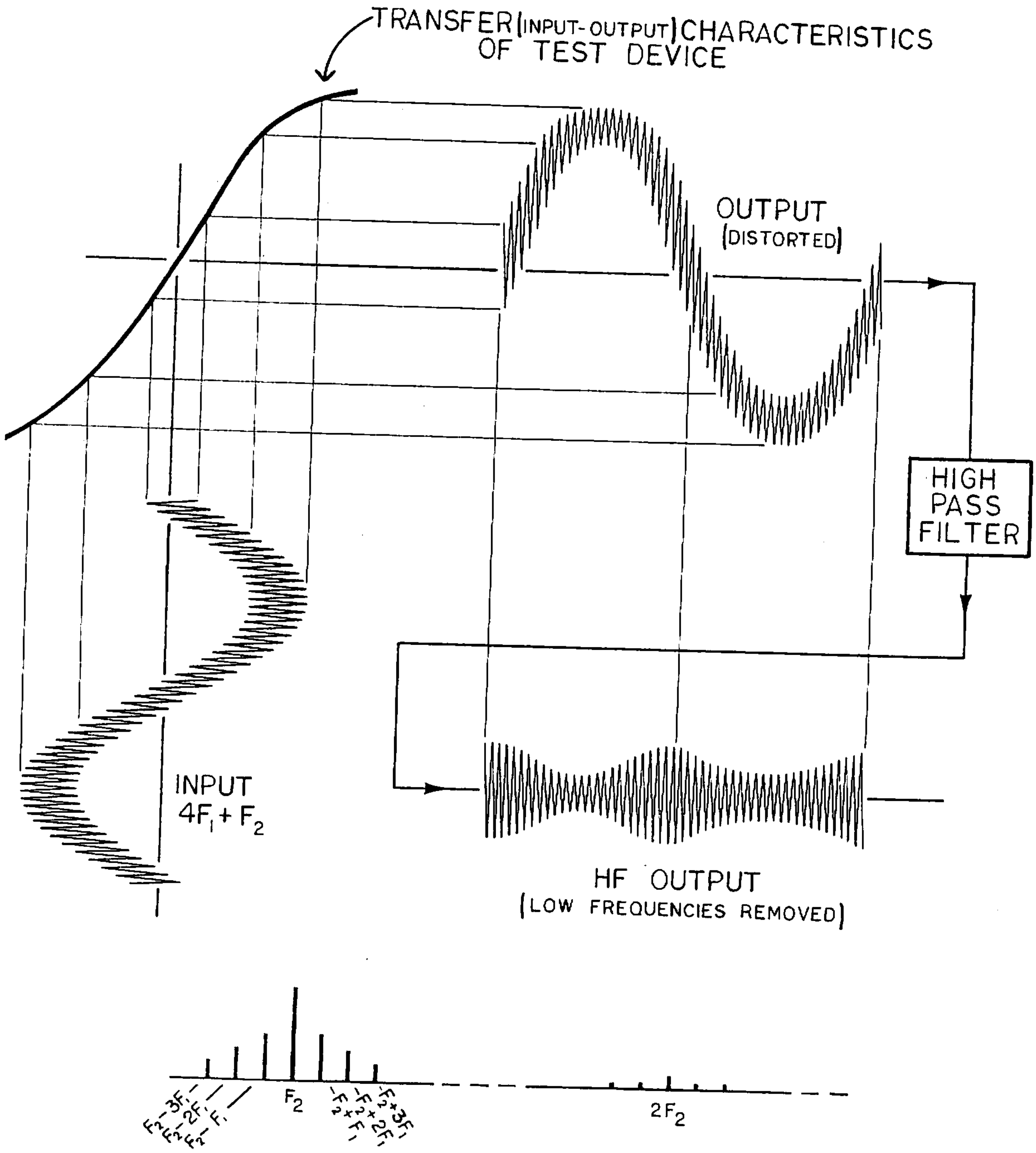
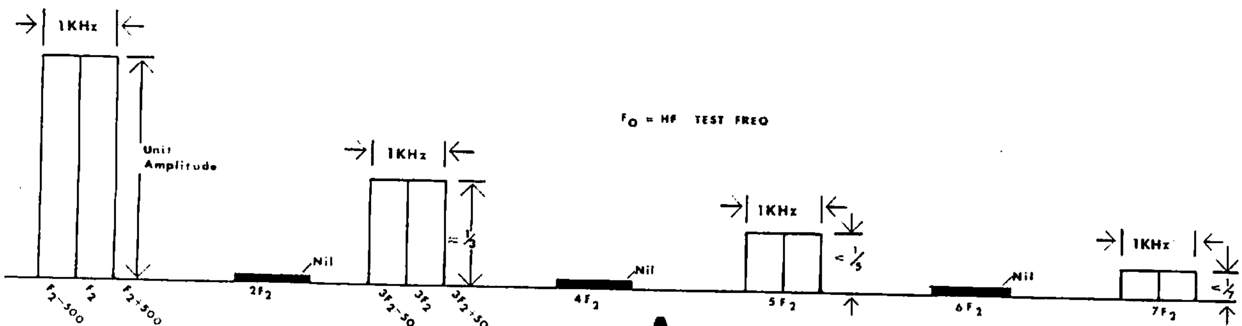


FIG. 4 - 1 DISTORTION GENERATION

These are not the only intermodulation products (sidebands) for most non-linear devices are not linear modulators any more than they are linear devices. Therefore the result is that harmonics of 60Hz appear as sidebands to 7KHz, i.e. 7K ±60Hz, 7K ±120Hz, 7K ±180Hz, 7K ±240Hz, etc. Harmonics of these frequencies and IM products associated with the harmonics of 7KHz are also generated such as 14KHz ±60Hz, 14KHz ±120Hz, 21KHz ±60Hz, etc.

The dominant components generated and measured will be those centered around 7KHz. The measuring techniques used in the IMA (and all competitive SMPTE devices) automatically suppresses components associated with the even order harmonics of the high frequency test signal. Components associated with the odd order harmonics are inherently suppressed by their harmonic number in the measuring process. Due to the dominance of the 7KHz products these factors will not cause a measurable error in all practical situations.

The range of intermodulation (distortion product) frequencies to which the IMA responds is basically the bandwidth of an internal low pass filter mirror imaged about the high frequency test signal frequency (Fig. 4-2a). The low-pass filter is 500Hz wide and nearly rectangular which with a 7KHz high frequency test signal creates a response which extends from 6.5KHz to 7.5KHz with a very narrow notch (< 6Hz wide and infinitely deep) at the high frequency (7KHz). The responses around the odd order harmonics of the high frequency (7KHz) will be the same 1KHz width but reduced in amplitude by the harmonic number and a small amount of ultrasonic roll-off in the IMA input amplifier. Deviations of the demodulator from an ideal absolute value detector also reduces the odd order responses and increases even order responses. A table of typical normalized sensitivities for a 7KHz high frequency signal is shown in Fig. 4-2b. Fig. 4-2c shows an intuitive explanation of why the even order harmonics do not indicate while the odd order do but weakly. Inasmuch as the harmonics are too weak as compared to the fundamental to cause zero crossings they undergo partial cancellation. In the case of even orders the cancellation is total.

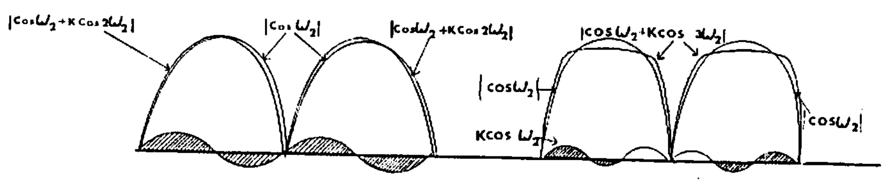


A

Typ Response Array for 7KHz

Resp Order	Norm Sens
x1	1.0
x2	.024
x3	.33
x4	.023
x5	.19
x6	.021
x7	.13
x8	.019
x9	.097

B



Contribution to average value of output by shaded areas is zero.
Zero crossings of Cos ω_2 reverse phase of higher terms every half cycle of ω_2.

C

DC & HARMONIC FREQUENCY

	DC	$\cos \omega_1$	$\cos 2\omega_1$	$\cos 3\omega_1$	$\cos 4\omega_1$	$\cos 5\omega_1$	$\cos 6\omega_1$
K_o	K_o						
$K_1 E$		$K_1 E_1$					
$K_2 E^2$	$\frac{K_2}{2}\left[E_1^2+E_2^2\right]$		$\frac{K_2}{2} E_1^2$				
$K_3 E^3$		$\frac{3K_3}{4}\left[E_1^3+2E_1E_2^2\right]$		$\frac{K_3}{4} E_1^3$			
$K_4 E^4$	$\frac{3K_4}{8}\left[E_1^4+4E_1^2E_2^2+E_2^4\right]$		$\frac{K_4}{2}\left[E_1^4+3E_1^2E_2^2\right]$		$\frac{K_5}{8} E_1^4$		
$K_5 E^5$		$\frac{K_5}{16}\left[10E^5+30E_1 \cdot E_2^4+60E_1^3 \cdot E_2^2\right]$		$\frac{5K_5}{16}\left[E_1^5+4E_1^3E_2^2\right]$		$\frac{K_4}{16} E_1^5$	
$K_6 E^6$	$\frac{K_6}{32}\left[90\left(E_1^4E_2^2+E_1^2 \cdot E_2^4\right)+10\left(E_1^6+E_2^6\right)\right]$		$\frac{K_6}{32}\left[15E_1^6+120E_1^4E_2^2+90E_1^2E_2^4\right]$		$\frac{K_6}{32}\left[6E_1^6+30E_1^4 \cdot E_2^2\right]$		$\frac{K_6}{32} E_1^6$

	$\cos \omega_2$	$\cos 2\omega_2$	$\cos 3\omega_2$	$\cos 4\omega_2$	$\cos 5\omega_2$	$\cos 6\omega_2$
K_o						
$K_1 E$	$K_1 E_2$					
$K_2 E^2$		$\frac{K_2}{2} E_2^2$				
$K_3 E^3$	$\frac{3K_3}{4}\left[E_2^3+2E_1^2E_2\right]$		$\frac{K_3}{4} E_2^3$			
$K_4 E^4$		$\frac{K_4}{2}\left[3E_1^2E_2^2+E_2^4\right]$		$\frac{K_4}{8} E_2^4$		
$K_5 E^5$	$\frac{K_5}{16}\left[10E_2^5+30E_1^4 \cdot E_2+60E_1^2E_2^3\right]$		$\frac{5K_5}{16}\left[E_2^5+4E_1^2E_2^3\right]$		$\frac{K_5}{16} E_2^5$	
$K_6 E^6$		$\frac{K_6}{32}\left[15E_2^6+90E_1^4 \cdot E_2^2+120E_1^2 \cdot E_2^4\right]$		$\frac{K_6}{32}\left[6E_2^6+30E_1^2E_2^4\right]$		$\frac{K_6}{32} E_2^6$

Any device that can be characterized as having a unique transfer function that is not a function of time or frequency may be modeled by a transfer function that is a simple power series: Let

$$E_{OUT} = \sum_{L=0}^{\infty} K_L E_{IN}^{L}$$

AND LET $E_{IN} = E_1 \cos \omega_1 T + E_2 \cos \omega_2 T$

WHICH MAY BE EXPANDED BY USING THE FACT

THAT $$(X+Y)^L = \sum_{J=0}^{L} \frac{L!}{J!(L-J)!} (X^{L-J} Y^J) \equiv \sum_{J=0}^{L} \binom{L}{J} (X^{L-J} Y^J)$$

FORMS SUCH AS $\frac{L!}{J!(L-J)!}$ SHALL BE WRITTEN AS $\binom{L}{J}$

AND ALSO

$$\cos^N \omega = \frac{1}{2^N} \sum_{i=0}^{N} \binom{N}{i} \cos (N-2i)\omega$$

THEN $E_{OUT} =$

$$\sum_{L=0}^{\infty} K_L \left\{ \sum_{J=0}^{L} \binom{L}{J} \left[\left(\frac{E_1^{L-J}}{2^{L-J}} \sum_{i=0}^{L-J} \binom{L-J}{i} \cos(L-J-2i)\omega_1 T \right) \left(\frac{E_2^J}{2^J} \sum_{M=0}^{J} \binom{J}{M} \cos(J-2M)\omega_2 T \right) \right] \right\}$$

WHICH SIMPLIFYING AND USING THE RELATIONSHIP $\cos X \cos Y = \frac{1}{2} \left[\cos(X+Y) + \cos(X-Y) \right]$

$$= \sum_{L=0}^{\infty} \frac{K_L}{2^{L+1}} \left\{ \sum_{J=0}^{L} \binom{L}{J} \left[E_1^{L-J} E_2^J \left(\sum_{M=0}^{J} \sum_{i=0}^{L-J} \binom{L-J}{i} \binom{J}{M} \left(\cos[(L-J-2i)\omega_1 + (J-2M)\omega_2] T + \cos[(L-J-2i)\omega_1 - (J-2M)\omega_2] T \right) \right) \right] \right\}$$

THE FOLLOWING IS A TABLE OBTAINED BY COMPUTING THE PRECEEDING WHERE $K_L = 0$ FOR $L \geq 7$.

INTERMODULATION FREQUENCY

	COS $(2\omega_1 \pm 3\omega_2)$	COS $(2\omega_1 \pm 4\omega_2)$	COS $(3\omega_1 \pm \omega_2)$	COS $(3\omega_1 \pm 2\omega_2)$	COS $(3\omega_1 \pm 3\omega_2)$	COS $(4\omega_1 \pm \omega_2)$	COS $(4\omega_1 \pm 2\omega_2)$	COS $(5\omega_1 \pm \omega_2)$
K_0								
K_2E^2								
K_3E^3								
K_4E^4			$\frac{K_4}{2}E_1^3 E_2$					
K_5E^5	$\frac{5K_5}{8}E_1^2 E_2^3$			$\frac{5K_5}{8}E_1^3 E_2^2$		$\frac{5K_5}{16} E_1^4 E_2$		
K_6E^6		$\frac{K_6}{32}[15E_1^2 E_2^4]$	$\frac{K_6}{32}[60E_1^3 E_2^3 \cdot +30E_1^5 E_2]$		$\frac{K_6}{32}[20E_1^3 E_1^3]$		$\frac{K_6}{32}[15E_1^4 E_2^2]$	$\frac{K_6}{32}[6E_1^5 E_2]$

INTERMODULATION FREQUENCY

	$\text{Cos}(\omega_1 \pm \omega_2)$	$\text{Cos}(\omega_1 \pm 2\omega_2)$	$\text{Cos}(\omega_1 \pm 3\omega_2)$	$\text{Cos}(\omega_1 \pm 4\omega_2)$	$\text{Cos}(\omega_1 \pm 5\omega_2)$	$\text{Cos}(2\omega_1 \pm \omega_2)$	$\text{Cos}(2\omega_1 \pm 2\omega_2)$
K E							
K_2E^2	$K_2E_1E_2$						
K_3E^3		$\frac{3K_3}{4}E_1 E_2^2$				$\frac{3K_3}{4}E_1^2 E_2$	
K_4E^4	$\frac{3K_4}{2}[E_1^3E_2 + E_1E_2^3]$		$\frac{K_4}{2}E_1 E_2^3$				$\frac{3K_4}{4}E_1^2 E_2^2$
K_5E^5		$\frac{10K_5}{16}[2E_1E_2^4 + 3E_1^3E_2^2]$		$\frac{5K_5}{16}E_1 E_2^4$		$\frac{10K_5}{16}[2E_1^4 E_2 + 3E_1^2 \cdot E_2^3]$	
K_6E^6	$\frac{K_6}{32}[180E_1^3E_2^3 + 60E_1 \cdot E_2^5 + 60E_1^5E_2]$		$\frac{K_6}{32}[60E_1^3 E_2^3 + 30E_1 \cdot E_2^5]$		$\frac{K_6}{32}6E_1E_2^5$		$\frac{K_6}{32}[60E_1^4E_2^2 + 60E_1^2 \cdot E_1^4]$

The conclusions to be drawn from such an expansion are:

1. Only frequencies of the form $p\omega_1 + q\omega_2$ are generated, where p and q are integers.
2. If the highest order term in the power series is of exponent N, then no frequencies for which $|p| + |q| > N$ will be generated.
3. All frequencies for which $|p| + |q| = N$ will be present except in the special case of perfect cancellation of all the associated coefficients of a respective frequency. For this to happen some associated K's must be positive, and some negative.
4. The parity of $|p| + |q|$ is that of the power term which produces the component. (For even n, $|p| + |q|$ is even, for odd n, $|p| + |q|$ is odd).
5. The rapid convergence to zero of Kn vs n does not imply as rapid a convergence of the total distortion products vs n.
6. All of the sum and difference frequency pairs are of the same amplitude and polarity. ($\cos(\omega_1+\omega_2)$ + and $\cos(\omega_1-\omega_2)$ + are a sum and difference frequency pair.)
7. If any Kn of same parity n are of opposite polarity there may exist at least one maxima and/or minima for any order of distortion vs |E out| for $|E_1| + |E_2| < \infty$

Point by point this means to the user:

1. If ω_2 is the high frequency the measured IM products will be of the form $p\omega_1$. Any frequencies found other than $p\omega_1$ constitute pollution of measurements or lack of conformity to the assumptions initially made about the test device.
2. If frequencies are found $|p| + |q| > N$ then n was prematurely truncated in the modeling.
3. Failure to pass a high order of $p\omega_1$ through the low-pass filter following the IM meter demodulator may allow the measured distortion to be erroneously low, even zero.
4. Perfectly balanced devices with transfer functions symmetrical about zero (K_n(n even)=0) will display only odd parity $|p| + |q|$ distortion products.
5. Even small amounts of high order terms (discontinuities such as crossover notch distortion in amplifiers) can cause large amounts of total distortion.
6. All of the high frequency intermodulation components around $|q| = 1$ where $|p|\omega_1$ is within the low pass filters bandwidth will be measured without cancellation when using a modulation type meter such as the IMA. Each individual sideband can be measured by using a wave analyzer at the B SCOPE output and measuring $|p|\omega$, and dividing by two (all measurements at B Scope must be normalized.) This assumes that the components of $|q| = 3,5,7$, etc. will not be substantially entering the reading which is usually the case due to their usually reduced amplitude and the analyzers inherently lower response at higher orders of ω_2.
7. The presence of notable discontinuities in the transfer function will generate a maximum of distortion by appropriately selecting the test level. For the case of an amplifier with crossover notch distortion a maximum of distortion will be generated at some low test level that is of similar dimensions as the notch region on the transfer function.

The statement has been made that the IMA is relatively immune to frequency modulation of the high frequency carrier as is characteristic of electro-mechanical devices such as turntables, tape recorders where wow and flutter are present and wide range single radiator cone loudspeakers where doppler effects may be evidenced.

Large amounts of FM will however cause amplitude modulation due to small amplitude response curvatures of the IM analyzers high pass filter. This effect does not make a significant contribution for the aforementioned devices for their actual distortion will easily dominate this effect. FM would not be troublesome unless the high frequency test signal was very near the high pass filter cut-off frequency, say 2.5 to 3KHz.

4.3 FUNCTIONAL DESCRIPTION (See fig. 4-3)

4.3-1 The test generators (LF and HF) of the IMA are both Wien bridge oscillators eliminating any level or distortion dependency on the AC line voltage as is common to all conventional analyzers which use a 60Hz LF signal taken from the power line (via transformer). If the LF signal is distorted average responding meters will give an incorrect interpretation of the test level (peak voltage) which can induce substantial errors when testing power devices. For this reason the 60Hz oscillator is designed to have < .1% THD. The oscillator is syncronized with the AC line to prevent distortion "beats" when a test device's distortion is related to the AC line frequency. The 7KHz oscillator has a fast automatic level control (ALC) to suppress any amplitude modulations of the oscillator as these would be detected as IM distortion. Field effect transistors (FETs) were used in the ALC circuits of both oscillators due to their inherent speed, simplicity, stability, and freedom from microphonics.

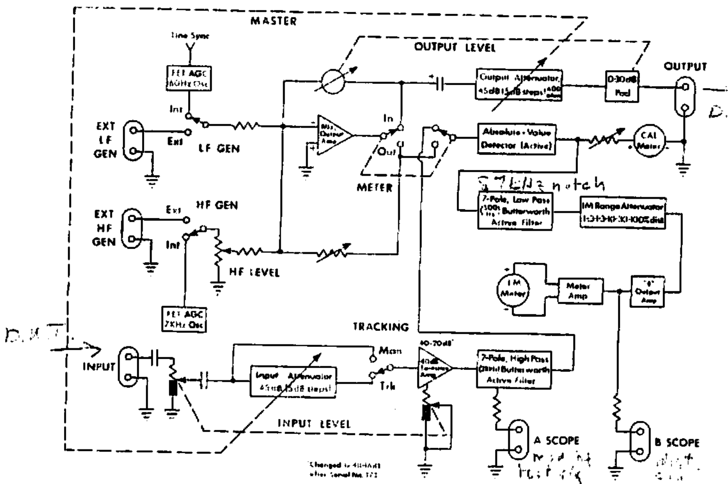


FIG. 4 - 3 IMA BLOCK DIAGRAM

4.3-2 Provision is made for inputting external generators in place of either internal oscillator. Such would be desired if frequencies other than 60Hz and/or 7KHz were desired. Great care should be used in selecting the external HF oscillator as most commercial test oscillators contain substantial amplitude variances!

4.3-3 An ultra-low distortion operational amplifier is used to mix the test generators and provide voltage and power gain to drive the analyzer's output. Mixing is done at the virtual ground (see fig. 4-4) which eliminates any chance of currents from one test generator flowing into the other and modulating it. The concept of a virtual ground stems from the fact that if voltages were present at the inverting (-) input an extremely large voltage would appear at the amplifiers output (overload) due to its very high voltage gain. The result is a mixing scheme having over 100db of LF to HF generator isolation with less than .001% distortion at 50V P-P output.

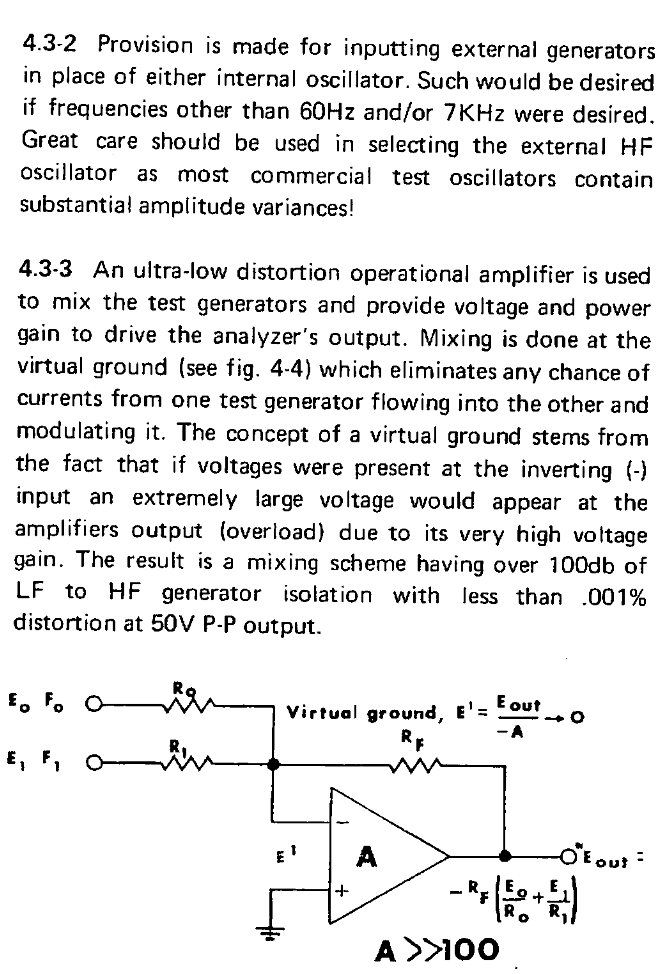


FIG. 4 - 4 OP-AMP MIXER

*Note: Since $e' \simeq 0$ and the amplifier impedance $\gg 0$, the currents in R_o, R_1, and R_f must sum to zero at the virtual ground, i.e. $-\frac{E_{out}}{R_f} = \frac{E_o}{R_o} + \frac{E_1}{R_1}$

4.3-4 The signal once mixed passes on to the METER IN-OUT switch where it may either be routed to the attenuators or to the absolute value detector which drives the calibrate meter. When in the IN position the signal passes on to the MASTER attenuator which is a 600Ω ladder attenuator having a 45db range with 5db increments. The MASTER attenuator is followed by the coarse OUTPUT level which is a 600Ω L-pad with 40db range in 10db increments.

4.3-5 The input signal upon entering the analyzer passes through a capacitor to the input attenuator to the input attenuators. The first attenuator is a specially tapered cermet* potentiometer which is ganged with a second section which controls the input amplifier's voltage gain. Turning this from the fully CW position, the second section first attenuates the input amplifier and for the second of the rotation (to fully CCW) the first section begins potentiometrically attenuating the input signal. The output of the first attenuator (pot section) passes through a second capacitor to the input attenuator section of the MASTER attenuator which is a precision potentiometric attenuator with 45db range and 5db increments. The MASTER attenuator is wired such that its attenuation increases in 5db steps when its output attenuator section decreases attenuation in 5db steps. This maintains a constant "test loop" gain (where the device under test has constant gain characteristics) for all settings of the MASTER attenuator. Thereby as many as 10 tests can be quickly and easily executed in this "TRacKing" mode with only one initial set-up. The "TRacKing" mode may be defeated with the TRACKING switch, which selects either the input (MAN) or output (TRK) of the MASTER input attenuator to be fed to the input amplifier.

4.3-6 The input amplifier is of special low-noise, low distortion, high input impedance design using a cascode input amplifier stage. Low noise and low distortion are of great importance since both produce signals which are detected as IM distortion. The dominant source of noise is thermal originating in the resistors of the input attenuators. This noise is at maximum in the -40db position of the MASTER attenuator when in the TRacKing mode. In the position the noise figure of the input amplifier is less than 1db. It is because of thermal noise that the INPUT level attenuator first decreases the input amplifier gain before allowing its potentiometric attenuator to take effect (which would further increase the thermal noise).

High input impedance is essential in the input amplifier to avoid introducing tracking mode attenuation errors in the MASTER input attenuator. Negative feedback coupled with the bootstrapped cascode input produces an input impedance so large as to be negligible compared to the stray capacitances in the attenuator system.

4.3-7 The output of the input amplifier is fed to a high pass filter for removal of all the low frequency part of the test signal's components. This high pass filter is of Butterworth type (maximum amplitude flatness design) to insure a minimum of spectral distortion with an optimum of usable bandwidth. Seven poles characterize this filter giving 42db/ octave attenuation to low frequencies starting at 2KHz. The filter is of RC active[4] design eliminating any hum sensitive inductors such as are typically found in IM analyzers. The A Scope output is taken after the sixth pole of the filter where it is at a convenient level to output. The A Scope output gives a convenient display of the modulated high-frequency test signal (see section 5). The seventh pole of the filter furnishes approximately 12db of gain as well as filtering. Its output is routed to the METER [illegible]h through which it drives the absolute value detector w[illegible]n in the INput position.

4.3-8 The absolute value detector in effect detects or demodulates the high frequency signal from the high pass filter. The operational amplifier circuitry performs the operation of producing the absolute value of the input voltage (full wave rectification). This output is used to drive a calibration meter as its DC or average value is the average value of the high frequency signal to which the % distortion test is referenced. The calibrate meter then indicates the calibrate level for testing (100%, full scale). The inherently flat to DC response of the absolute value detector allows it to be used along with calibrate meter to set up the voltage amplitude ratios when the METER switch is set to OUTput.

4.3-9 The output of the absolute value detector is connected to a low pass filter where the intermodulation components (low frequencies) are separated from the high frequency signal. This low pass filter is of 7pole Butterworth type with a 500Hz cutoff frequency. Butterworth response is desired to allow maximum amplitude accuracy of all significant intermodulation signals with excellent high frequency rejection. Since any [illegible] picked up in such a filter is measured as distortion, the [illegible] of inductorless active RC filters[4] gives the IMA unusual immunity to external fields. Four cascaded filter stages which are AC coupled to the absolute value detector form the active filter which attenuates at a rate of 42db/octave above 500Hz. The shape of the apparent notch in the IM distortion spectrum that occurs at the HF test frequency is governed by the RC coupling into the low pass filter and in following AC coupled stages.

4.3-10 The output of the low pass filter is AC coupled to the IM range attenuator which attenuates the signal for all ranges of .3% or more full scale IM distortion sensitivity. All attenuator resistors are 1% precision units.

4.3-11 The range attenuator is followed by a stable low noise AC amplifier which amplifies the distortion signal to a level appropriate for driving the meter amplifier and for external observation at the B Scope output terminals (See Section 5). The signals at the B Scope output will be proportional to the full scale meter reading.

4.3-12 The meter amplifier is a high feedback current sensing type which provides an extremely linear input voltage to meter current conversion while providing the proper drive for critical damping of the meter movement.

4.3-13 The power supplies ($\pm$30VDC) are series regulated with a single zener reference and a single adjustment which simultaneously adjusts both supplies to the same amplitude. Supply ripple and line voltage variations are highly suppressed by high internal feedback.

4.4 CIRCUIT ANALYSIS (See Schematics, Section 7)

4.4-1 The LF (60Hz) Oscillator Q1 thru Q4 makes up a differential input amplifier which is used in a Wien bridge configuration (see fig. 4-5).

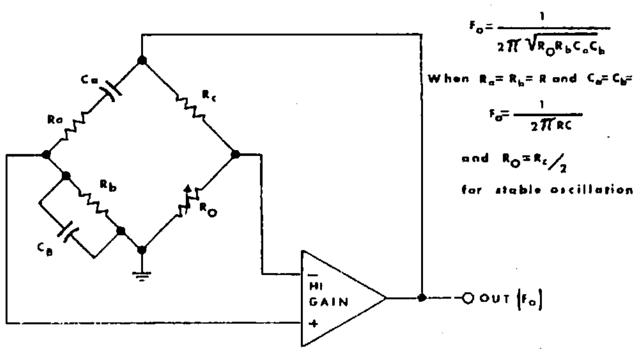


FIG. 4 - 5 WIEN BRIDGE OSCILLATOR

(4) Kincaid, Russel "RC Filter Design by the Numbers" The Electronic Engineer p57-64 Oct. 1968.

R_1 is used to trim the oscillation frequency to exactly 60Hz where proper syncronization with the line frequency can be obtained. The sync signal is derived from the 6.3VAC winding of the power transformer via R11, C3, and R12. C3 is used to remove AC noise and line frequency harmonics from the sync signal to keep oscillator noise and distortion to a minimum.

The function of the ALC system is to regulate the output level of the oscillator by in effect adjusting R_D. High stability is required of this system because of the syncronization signal which may be of unsteady amplitude due to AC line voltage variations. The block diagram of the ALC system is shown in Fig. 4-6.

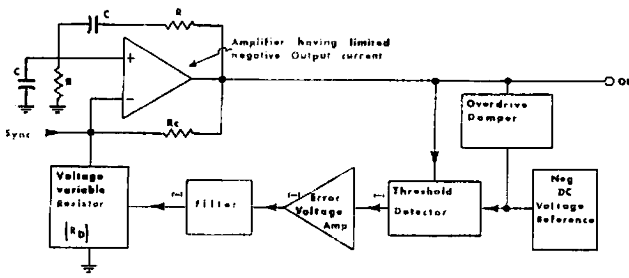


FIG. 4 - 6 OSCILLATOR ALC SYSTEM

The voltage variable resistor used to vary R_D is an N-channel junction FET,Q6 in series with R8 and R9 and in parallel with R13. One half of the drain voltage is fed back to the gate via R14 and R15, which improves the resistance vs voltage linearity(5) and renders it symmetrical about zero volts (Q6 has in interchangeable Source-Drain geometry). The effect is to minimize oscillator distortion and eliminate even-order distortion products.

Applying a negative voltage to the gate of Q6 increases the effective resistance, thus increasing R_D and diminishing the propensity to oscillate.

C4 stores the negative control voltage which is fed to the gate of Q6 and effectively removes all 60Hz AC signals from the control amplifier Q5's output.

Q5 functions as the error voltage amplifier, threshold detector and overdrive damper. Whenever the negative peak output voltage of the oscillator becomes excessive the base-emitter junction of Q5 will become forward biased causing a current to flow from the oscillator output (R7) to C4, R14, and R15. Since a relatively small increase in the voltage applied to Q5 beyond the potential where B-E conduction begins will cause a rather large current to flow into the high Z (at DC) network of C4, R14, and R15, Q5 can be accurately described as a high gain ALC servo. The damping effect of Q5 is apparent whenever the oscillation becomes so large that the oscillator amplifier clips due to delivering all its available negative output current to the instantaneously low impedance of Q5's forward biased emitter. Such is the case at turn-on and when severe sync level changes occur abruptly. Output level convergence is greatly hastened by this feature. Elsewise the oscillator settling time with such a high gain ALC servo would be intolerable.

R16 and R17 are highly stable metal film resistors which divide the regulated -30VDC supply voltage to produce the reference voltage for Q5. D1 isolates Q5 from C4 during the positive going half cycle of the oscillator output. When the B-E emitter junction is back biased to where avalanche current flows into R16 and R17 the C-B junction of Q5 would be forward biased were it not for D1.

4.4-2 The HF (7KHz) Oscillator The HF oscillator is identical in operation to the LF oscillator with exceptions.

The ALC system is speeded up to allow cycle to cycle compensation for system gain variations that would produce modulation noise in the 7KHz signal. This is simply done by making C8 small (C4 in LF osc.).

There is no syncronization thus there are no components corresponding to R11, C3, and R12 of the LF oscillator.

The -30VDC supply is filtered by C9 and R33 to prevent small hum and noise signals on the supply from modulating the oscillator.

The regenerative or frequency determining feedback loop of the 7KHz oscillator is switched open by the HF GENERATOR switch when in the EXTernal position. This shuts off the 7KHz oscillator thereby preventing "beats" with an external oscillator which could pollute the measurements.

4.4-3 The Op Amp Mixer Q13 and Q14 form an NPN differential common emitter pair which in turn drive Q15 and Q16 a PNP differential pair. The base of Q13 is the inverting input of the amplifier. The non-inverting amplifier input (the base of Q14) is grounded.

The output of Q16 (collector) is used to drive Q17 emitter follower which drives Q18, the final voltage amplifier. Q18 is driven in a grounded base configuration to obtain a large gain bandwidth and a large undistorted voltage swing.

Q19 forms a constant current source for the collector of Q18. R48, R49, R50, and R51 determine the bias point of Q19 and thereby Q18.

Q20 and Q21 are the output power amplifiers which provide current gain for the voltage output of Q18. D3 and D4 cause a bias current to flow thru Q20, R52, R53, and Q21 which eliminates any tendency of Q20 and Q21 to show crossover notch distortion.

C10, C11, C12, and C13 are used to stabilize the amplifier to allow the application of high feedback. R39, R40, and R41 form the main feedback loop which is connected when the METER switch (S6A) in the INput position. R37 and R38 become the feedback loop when the METER switch is in the OUTput position. R41 is paralleled by R40 to produce a 50Vp-p amplifier output when the internal generators are mixed 4:1 and R41 is set to maximum resistance. This is done to prevent unintentional overloading of the amplifier by setting the OUTPUT (vernier) level too high. The amplifier may be overloaded v[illegible] using a 1:1 ratio if the level is not held to 50Vp-p or less at the amplifier's output.

Since R41 (OUTPUT level variable) is a wire wound control and has finite resolution a minimum feedback resistance is formed by R39 such that the minimum resistance increment of R40 is less than 1% of R39, i.e. OUTPUT level is settable to 1% resolution under the worst condition. An added benefit is that amplifier feedback stabilization is simplified by not allowing the feedback to become total.

4.4-4 The Output Attenuators (step)

The MASTER attenuator is of 600 10 position, 45db, unbalanced type design having 5db attenuation per step and using 1% precision resistors. R54 and R55 are higher dissipation types than R56-72 due to the large voltages that the op-amp mixer may deliver to the input of the attenuator. Any output of the attenuator is capable of withstanding the output amplifiers maximum output signal with the output shorted for an indefinite period of time. The switch (SW1) is of star wheel double roller arm detent type for extremely long life (over 250,000 cycles).

The OUTPUT level control is a 5 position, 40db, L-pad having 10db steps and a 600 ohm output impedance. Both legs of the L-pad are aligned to an accuracy of 1% by F[illegible] hru R80. This attenuator may also be shorted without any possible damage. The switch (SW2) is of ball detent type.

The output of these cascaded attenuators then goes to J3 the front panel OUT binding posts.

4.4-5 The Input Attenuators

The input signal having been input to J6, (the IN binding posts), travels via C16 to potentiometer R81A the front section of the INPUT level control. The output of R81A is coupled thru C17 to the input of the input attenuator section of the MASTER attenuator (S1B). C16 and C17 remove much of the low frequency test signal before it reaches the input amplifier, making its signal amplification less likely to suffer from IM distortion.

The input attenuator of the MASTER control is a 100K potentiometric design with 5db steps which are in reverse attenuation order as the output attenuator section. The resistors of the input attenuator R82 thru R91 are all 1% precision types. Potentiometric design is used because output resistance vs. attenuation must be minimized for minimum thermal noise.

The output and input of SW1B are routed to the TRACKING switch SW7 where the MANual or TRacKing mode of analyzer operation may be selected. SW7 routes the input of SW1B to the input amplifier in the MANual mode or the output of SW1B to the input amplifier in the TRacKing mode.

4.4-6 The Input Amplifier Q22 and Q23 form a cascode input amplifier which in turn drives Q24 a common emitter voltage amplifier. Q24 drives Q25 an emitter follower which gives a low output impedance while presenting a large load impedance to the collector of Q24 to maximize its voltage gain. The result is a large open loop voltage gain which when fed back to Q22 via R100 creates a very linear negative feedback amplifier with extremely high input impedance.

The bias to Q22 flows thru R93 which is bootstrapped by C19, R94 and R81B to eliminate signal loading in the bias circuit. R81B, the rear section of the INPUT level control, increases resistance when turned CCW thereby decreasing the amplifier gain to minimize system noise.

R92, in series with the amplifiers input, is provided to protect the amplifier from overload damage. A film type resistor was chosen since such types do not drastically drop in resistance when greatly overheated.

The resistance elements of the feedback path R94 and R100 are film types and R81B is Cermet* for high gain stability vs. temperature.

R95, R96 and C20 serve to bias Q23 and couple its base to the cascode pairs common, the emitter of Q22. The ratio of these two resistors, the V_{BE} of Q23, and the power supply

*T.M. CTS

voltage determine the Vce of Q22 which is nearly independent of the incoming signal potential. R97 and the V_{BE} of Q24 serve to determine the collector current bias point of Q22.

C21 is used to stabilize the amplifier-feedback system.

4.4-7 The High-Pass Filter

Six of the seven poles of this 2KHz Butterworth 7 pole active RC filter are formed by three unity voltage gain amplifiers composed of Q26 thru Q31. Each amplifier generates a pair of complex conjugate poles. In the first amplifier for example C22, C23, R101, and R102 determine the pole locations. In each successive amplifier stage the poles are located closer to the $j\omega$ axis. This pole ordering technique minimizes the maximum signal amplitude generated in the filter array. If high Q poles were to be placed first in the cascade array the filter dynamics would be reduced near cutoff.

Each amplifier uses a PNP-NPN reverse amplified emitter follower to produce a high input impedance, low output impedance amplifier. with a voltage gain very close to unity.

Capacitors, C24 and C29, are placed across the base-emitter junctions of Q27 and Q31 respectively to provide stability at RF frequencies.

The A SCOPE output is taken after the third amplifier (sixth pole) and coupled thru R110 to J4 a dual binding post on the front panel. J4 may be shorted and the resistance of R110 will not load the filter enough to cause loading or overload in the filters.

The seventh pole (real valued) is formed by C30 and R111 whose output goes to a buffer amplifier Q32 thru Q35 having approximately 12db voltage gain. The output of this amplifier is directly coupled to the absolute value detector. This is necessary because the absolute value detector has a non-linear input impedance which would create zero stability problems if a capacitor were used for input de-coupling.

With DC coupling between these two circuit blocks it now is imperative that the buffer amp have low DC output drift. To minimize such, total DC feedback is allowed by C31 and Q32 and Q33 form a differential input pair for offsetting temperature-voltage drifts.

The output of the differential pair is taken at the collector of Q32 where it is impressed across R113 and the base emitter junction of Q34, a common emitter voltage amplifier which drives Q35 an emitter follower. The result is a high open loop voltage gain which when negative feedback is applied via R115, R114, and C31 creates a very gain stable amplifier. R114, and R115 are film resistors to insure high gain-temperature stability. C57 is used to insure stability.

4.4-8 The Absolute Value Detector is a DC operational amplifier composed of Q36 thru Q39 which is coupled with a non-linear network to create such a response.

It is functionally described by Fig. 4-7.

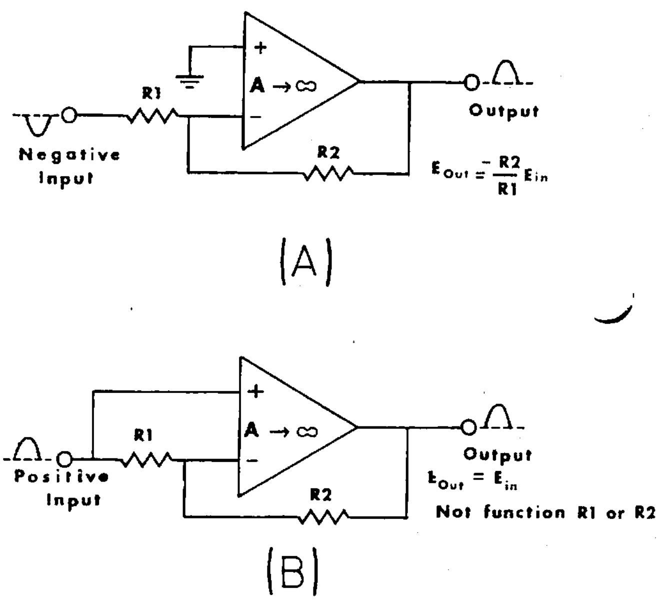


FIG. 4 - 7 ABSOLUTE VALUE DETECTOR

Diodes D5 and D6, depending on the input signal's polarity, will route the incoming signal thru D5 to the non-inverting amplifier input if Ein is positive (Fig. b). If Ein is negative the non-inverting input will remain grounded via D6 and R118. The result is as depicted in Fig. 4-7.

The feedback resistors R1 and R2 of Fig. 4-7 are R119, R120, and R121 in the analyzer. Adjusting R120 adjusts the R2/R1 ratio to produce equal gain for both input signal polarities.

Q36 and Q37 form a differential input pair which in drive Q38 a common emitter amplifier. The output of Q38 drives Q39 an emitter follower. C32 and C33 are used to provide feedback stabilization. The result is a very gain stable IM demodulator with flat to DC response and a low output impedance.

4.4-9 **The Low Pass Filter** is of 7 pole 500Hz Butterworth active RC design with four unity voltage gain amplifiers composed of Q40 thru Q47. Since the absolute value detector outputs DC which is used to drive the calibrate meter but is not itself a distortion component the absolute value detector is AC coupled into the LP filter by C35. The real valued pole is realized first by R129 and C36 which isolated from the following filter section by Q40 and Q41. The following three stages each generate two complex conjugate poles each pair of successively higher Q.

Each of the four amplifiers uses a PNP-NPN reverse amplified emitter follower to produce a high input impedance, low output impedance amplifier with a voltage gain very close to unity.

The output of the LP filter is AC coupled to the IM% range by C43.

4.4-10 **The IM Range Attenuator** is a precision potentiometric attenuator which attenuates the output of the LP fi o the B scope amplifier for all distortion ranges above 0.1% full scale.

The resistors of the attenuator R140 thru R146 are all 1% precision types.

The switch SW3 is of star wheel double roller arm detent type for extremely long rotational life (over 250,000 cycles).

4.4-11 The B Scope Amplifier is a low noise 4 transistor (Q48 thru Q51) amplifier. A differential input pair (Q48 and Q49) is used as opposed to a single transistor input stage because of its generally superior overload recovery characteristics. Overload recovery is important as this characteristic limits the reading time for measurements when using the tracking mode of operation.

Q48's output drives Q50 a common emitter amplifier which in turn drives Q51 an emitter follower output stage. The result is a high open loop gain amplifier that gives high gain stability with feedback connected for ≈ 48db gain.

The feedback resistors R150 and R149 are film types for high temperature-gain stability. C44 allows total DC feedback to occur resulting in a low DC offset at the amplifier output.

The B SCOPE output is taken thru R152 from the amplifier output. C45 is placed across the output jack J5 to eliminate any high frequency crosstalk from appearing at the B SCOPE terminals.

4.4-12 The Meter Amplifier is a three transistor (Q52 thru Q54) operational amplifier that has the meter movement placed in the feedback loop for highly linear average value AC voltage to DC current conversion. Operationally it is depicted by Fig. 4-8.

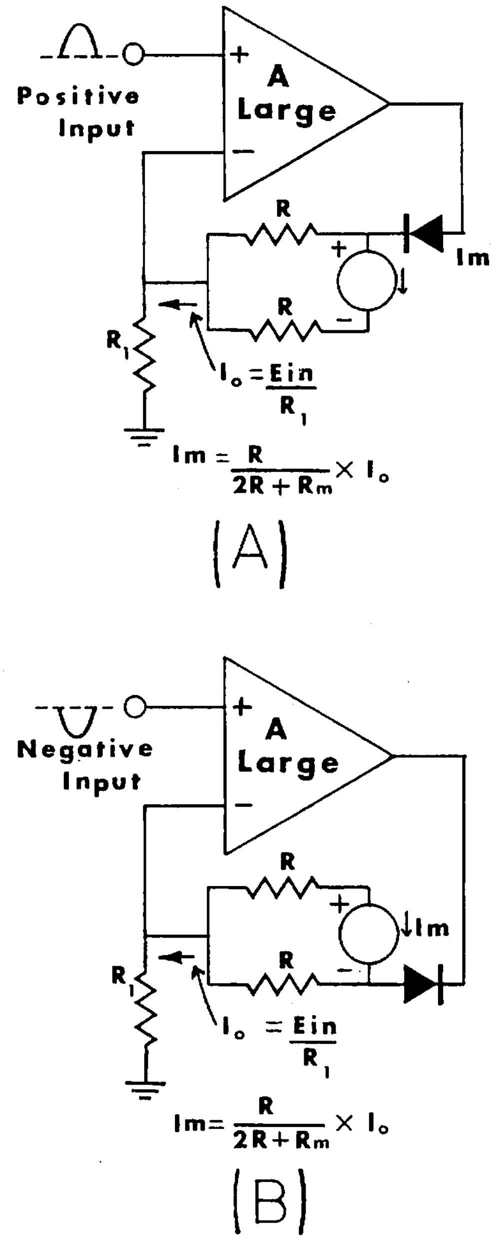


FIG. 4-8 THE METER AMPLIFIER

Fig. 4-8(a) shows the current path for a positive input which creates a positive current in the meter. Some of the amplifier output current is lost to the bridge resistors depicted as R and this creates the $\frac{R}{2R+R_M}$ term in the equations for I_M. Note that I_M is positive in both cases (a) and (b). Substituting for Io this means that $I_M = \frac{R}{2R+R_M} \cdot \frac{Ein}{R_1}$ which indicates I_M to be a linear function of the amplitude of Ein and scalable by adjusting R_1.

In the meter amplifier R157 and R158 form the bridge resistors with R155 and R156 making up R1. C45 is used to allow total DC feedback to eliminate any chance of a DC input signal causing a residual meter reading. With no input signal the bias current flows thru D7 and the meter resistor bridge to the base of Q53. This bias current will deflect the meter slightly and for this reason the meter must be mechanically zeroed with power applied to the analyzer. C47 is used to remove transients from the meter leads that might crosstalk into surrounding circuits. D7 and D8 are low leakage silicon diodes which have low reverse leakage currents and thereby render valid the model shown in Fig. 4-8.

R159 and R160 are selected such that the meter is inherently protected against overload by virtue of limited amplifier (Q54) drive.

Long term calibration stability is assured by using film resistors for R156 thru R158 and a Cermet* trimmer adjust.

4.4-13 The Power Supplies ($\pm$30VDC) are both of the full wave rectifier type each followed by a series voltage regulator. (See fig. 7-2)

In a series regulator, transistors (Q55 and Q58) are used as a variable series resistance. These resistances are varied by control amplifiers Q56, Q57, Q59 and Q60. The control amplifiers measure the difference between a predetermined fraction of the output voltage and a DC reference voltage. The control amplifiers act so as to reduce this difference to zero.

The positive 30volt supply compares the voltage of the reference diode D13 + the V_{BE} of Q57 to a fraction of the output voltage determine by the setting of R166 part of the voltage divider network of R165, R166, and R167.

The negative supply uses the positive supply as its reference by coupling its divider network (R171 and R172 1% precision resistors) to the positive supply. The output of the negative supply is thereby simultaneously adjusted by adjusting R166 of the positive supply.

Diodes D9 thru D12 are the rectifiers and C48 and C52 are the supply capacitors. C49 and C53 serve to bootstr- the regulator amplifiers increasing their dynamic ga- and filtering out supply AC ripple. D13 is powered thru R164 from the +30 regulated supply to prevent its voltage being varied by the line voltage. R161 and R168 limit the short circuit current should the supplies be momentarily shorted. The supplies will not survive a sustained shorted condition.

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5.1 AMPLIFIER TESTING

Much of the published literature on IM distortion has been weighted heavily in favor of electronic amplifier testing as its basic usage.

In spite of all the published literature it does not seem that the relation of IM distortion to amplifier testing problems has been clearly stated. Some authors have used such poorly chosen terms as "SMPTE IM testing is basically a measure of an amplifier's low frequency distortion" and the like. This is not so! But rather SMPTE IM testing reveals the sensitivity of an amplifier's high frequency characteristics to a low frequency stress.

In most practical systems there is a strong correlation between high frequency distortion and IM distortion although not usually as straight forward as the result of the expansion of chapter 4.2 given for the case of a power series representation of the transfer function. The following is an example of a system having both high and low frequency distortion but no SMPTE IM distortion, Fig. 5-1.

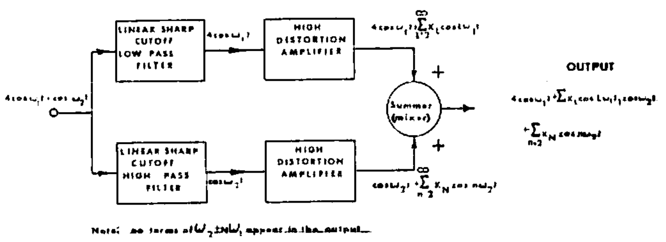


FIG. 5 - 1 A DISTORTED AMPLIFYING SYSTEM HAVING NO SMPTE IM

An example of a practical system resembling Fig. 5-1 would be a parallel filter designed graphic equalizer. Such a device would be better tested by harmonic methods. However the input and output amplifiers would submit to SMPTE testing.

Fortunately most amplifying systems and especially solid state power amplifiers are particularly amenable to SMPTE IM testing. This is because both the high and low frequency signals follow concurrent signal paths and are incident upon the same non-linear mechanisms although not always equally affected by these mechanisms. Also the distortion terms generated by a mechanism such as crossover notch which is very common in solid-state designs are of high order and generate large amounts of SMPTE IM allowing such defects to be readily detected.

Much discussion is contained in the available literature[1-6] on the relations between harmonic and intermodulation distortion components. The computations therein performed are applicable only to the case of a device having a unique frequency independent transfer function that can be represented by a power series. Departures from this ideal are due to a multitude of causes, i.e.:

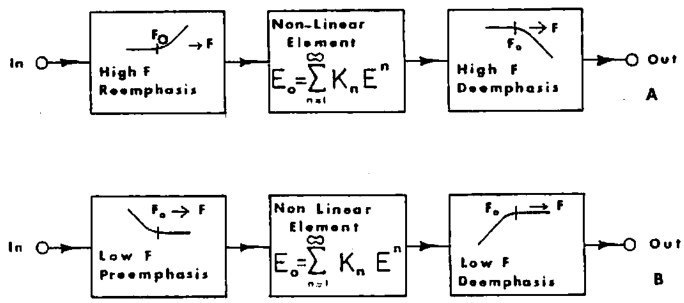


FIG. 5 - 2 FLAT SYSTEMS WITH FREQUENCY DEPENDENT DISTORTION

I. Frequency dependent distortion mechanisms: such as transformers where large low frequency signals saturate a ferro-magnetic core and reduce its efficiency, power transistors which change internal junction temperature with application of the low frequency and thereby change gain, bias conditions, and speed, slow power transistors which change bias conditions, and gain due to excess heating at high frequencies, amplifier stages which require frequency dependent signal amplitudes which cause preceding stages to deliver overload (distorted) signals, switching transients in class B or AB solid state power designs, etc.

II. Frequency dependent gain in parts of the amplifier system: For example Fig. 5-2(a) will evidence a reduced amount of IM distortion while 5-2(b) will

(1) M. V. Callendar and S. Matthews "Relations Between Amplitudes of Harmonics and Intermodulation Frequencies" Electronic Engineering (June 1951) p. 230.

(2) W. I. Warren and W. R. Hewlett "An Analysis of the Intermodulation Method of Distortion Measurement" Proc. IRE (April 1948) p. 457.

(3) B. Oliver "Distortion and Intermodulation" H.P. Application Note No. 15

(4) E. W. Berth-Jones "Intermodulation Distortion - Its Significance and Measurement" Jour. Brit. IRE (Jan. 1953) p. 57.

evidence an increased amount of distortion as based upon comparison to the distortion that would be generated by the non-linear element without frequency sensitive elements cascaded.[7] The case of Fig. 5-2(a) would be more sensitive if the low frequency : high frequency ratio were increased greatly to increase the amount of low frequency incident on the non-linear device. The 4:1 ratio of SMPTE IM measurements was chosen because it produced ample sensitivity to most situations and this example is an exception. Most clearly the equivalent ratios of any two such tones musically sounded may have ratios ranging from 100:1 to 1:100; therefore there should be no esthetic objection to increasing the ratio. Practical examples of the foregoing systems would be in various recording processes and in domestic FM and TV sound transmission-reception systems.

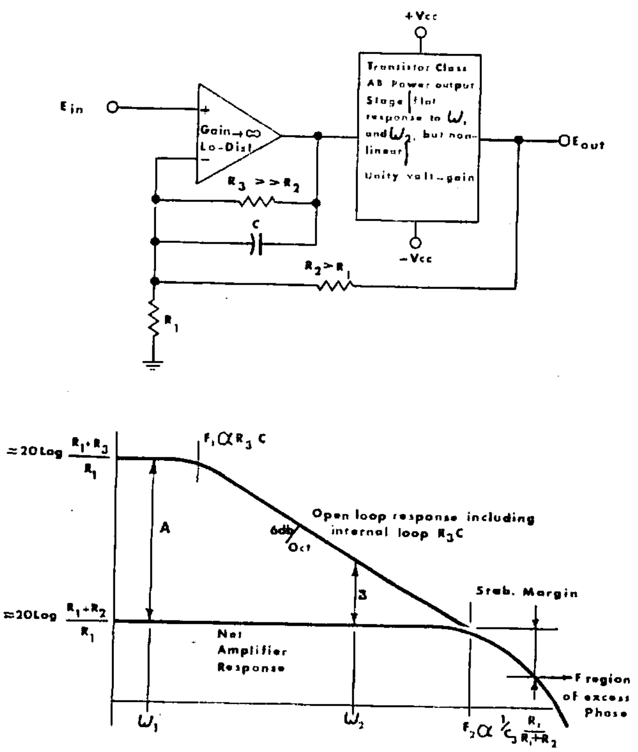


FIG. 5-3 TYPICAL POWER AMPLIFIER FEEDBACK STRUCTURE

III. Frequency dependent negative feedback loops: Optimal designs of all feedback systems are limited by the maximum amount of high frequency feedback that may be applied without resulting in instability. This is due to inherent phase lags caused by reactances and time delays in the input-output responses of all physical devices. For this reason the feedback amplitude must be reduced below the critical level unity at high frequencies where the collective phase lag is 180 degrees and more.[8] A typical example is Fig. 5-3 of a solid-state power amplifier where the primary cause of phase lag is the reduced gain-bandwidth and large time delay characteristics of the output power transistors. The Bode plot is shown to illustrate how the internal feedback loop of R_3 and C have reduced the high frequency gain and thus provided a stable condition under the application of the flat response feedback loop of R2 and R1. It is important to note that while the overall amplifier has a flat response the feedback factor is much greater at ω_1 resulting in nearly tota elimination of low frequency distortion. If SMPTE IM distortion were truly "a measure of low frequency distortion" varying R3, and thus the feedback factor (A) at ω_1, would vary the IM; but not so! The IM will be independent of the value of R3 but as expected the low frequency harmonic distortion (negligible) will be decreased by maximizing A or removing R3.

On the other hand varying C or the feedback factor[B] at ω_2 will affect the SMPTE IM distortion. Increasing B will proportionately decrease the IM. Unfortunately B is limited by the original amplifier stability problem. The slope of the open loop response often cannot be appreciably increased beyond 6db/octave due to the additional undesirable phase lag that would be introduced by such operations. If the amplifier designer has already optimized B the only further method to reduce the IM is to work on the power output stage.

From this example it is clear that SMPTE IM testing from a design standpoint in today's designs is more appropriately termed a "low frequency stressed method of evaluating high frequency distortion."

(5) A. Block "Measurement of Non-Linear Distortion" Jour. of And. Eng. Soc. (Jan. 1953) Vol. 1, No. 1, p. 62.

(6) D. E. O'N. Waddington "Intermodulation Distortion Measurement" Jour. of And. Eng. Soc. (July 1964) vol. 12, No. 3, p. 221.

(7) J. S. Aagaard "An Improved Method for the Measurement of Non-Linear Audio Distortion" IRE Trans. Aud. (Aug. 6, 1958) p. 121.

(8) H. W. Bode "Network Analysis and Feedback Amplifier Design" D. Van Nostrand Princeton, N.J. (1945).

In the previous example the amplitude of the $\omega_2+n\omega_1$ distortion components will be slightly greater than the $\omega_2-n\omega_1$ distortion components due to the decreasing feedback at the higher frequencies; however the resulting total will still be indicative of the SMPTE IM distortion at ω_2 for all analytical purposes. Increasing ω_2 will clearly increase the IM distortion. As long as the essential distortion components lie within the IMA's 500Hz low pass filter response range the total distortion will be independent of ω_1's frequency. (In some designs bias conditions in the output stage may be substantially altered by thermal cycling caused by ω_1 which will cause the IM to be sensitive to the frequency of ω_1. Also if the output stage proper is improperly stabilized high frequency oscillations may appear which typically creates sensitivity to ω_1.)

IV. Oscillations: There is no simple relationship between the amount or frequency of an oscillation and the IM reading it will cause. Oscillations can cause IM distortion inside the analyzer as well as in the device under test. Inasmuch as oscillations are not acceptable in properly functioning amplifiers such increased sensitivity is a virtue and not a vice. Sometimes an IM meter will prove to be a more effective oscillation detector than a wide band oscilloscope at high signal levels.

RF oscillations are not unusual in class AB or B direct coupled transistor outputs having high negative feedback implicit in their circuit topologies. Stability in such devices is complicated by proportionally large ranges of voltages and currents applied to the semiconductor devices which greatly varies the high frequency parameters of each device. Thermal cycling at the ω_1 and $2\omega_1$ frequencies may further aggravate the tendency to oscillate.

V. Noise: As in all other types of measurements SMPTE distortion is ultimately limited by noise. Such effects are substantially less than in the case of THD (notch type) distortion measurements due to the substantially reduced bandwidth that characterizes the IMA ($\approx$ 1KHz as compared to hundreds of KHz in notch type analyzers).

Present day circuit designs have frequently produced distortion levels that are easily confused with noise if distortion analyzing devices do not give high noise rejection.

The noise bandwidth of the IMA is depicted in Fig. 4-2, Analyzer Responses. If the noise is random and white across the entire spectrum of $n\omega_2$ the total detected noise ($\bar{e}_t$) can be computed as:

$\bar{e}_t=\bar{e}_1 \sqrt{B}$ where $\bar{e}_1$ is the rms noise voltage $/\sqrt{\text{hertz}}$ and B is the effective total bandwidth resulting from all the harmonic of ω_2 responses of the analyzer.

$B = \sqrt{(1)^2 + (1/3)^2 + (1/5)^2 + (1/7)^2 + (1/n)^2} \times$ 1KHz where n is odd. Even order harmonics of ω_2 are not detected.

Note that B converges rapidly as n increases with the first three terms forming the result, i.e.

$B = 1.1\text{KHz}$

If the noise spectrum is not white over $n\omega_2$ but is essentially white in each band of $n\omega_2$, B may be computed by using an appropriate weighting factor for each 1/n component in the root of the sum of the squares computation for B.

The effects of noise upon a measurement is always to increase the apparent distortion. If an increased immunity to noise is necessary a wave analyzer can be used at the "B" scope terminals where the true distortion components ($n\omega_1$) may be separately measured and summed: (Summation by root of the sum of the squares will yield true RMS distortion.)

Further comments on noise are found in sections 5.2 & 5.3.

When testing amplifiers it is instructive to view both the signal being output from the test device and the B scope signal simultaneously. A dual trace oscilloscope in the chopped display mode or alternate display mode with a trigger signal derived from ω_1 is most convenient. (A single trace scope can be used if it is sync locked or triggered with an external source of ω_1) Figure 5-4 shows a typical scope display of such.

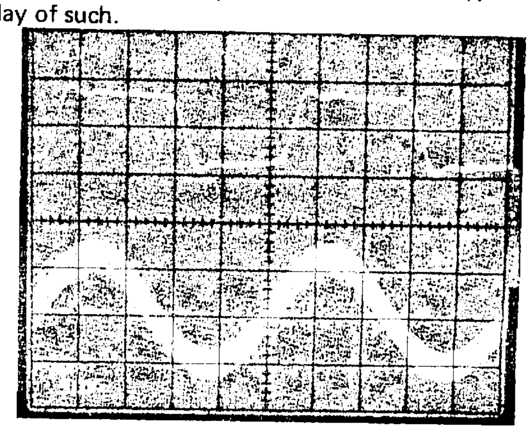


FIG. 5 - 4 SCOPE DISPLAY SHOWING "B" SCOPE

The "B" Scope display shown indicates that the gain of the test device is slightly greater during the positive going half of its output voltage as evidenced by the positive polarity associated with this time period. The negative output half cycle indicates reduced gain associated with this polarity of output.

Note that the "B" Scope output is slightly time delayed with respect to the device output signal. This is due to the phase response properties of the 500Hz low pass filter. A typical plot of the LP filters time delay vs. frequency is shown in Fig. 5-5.

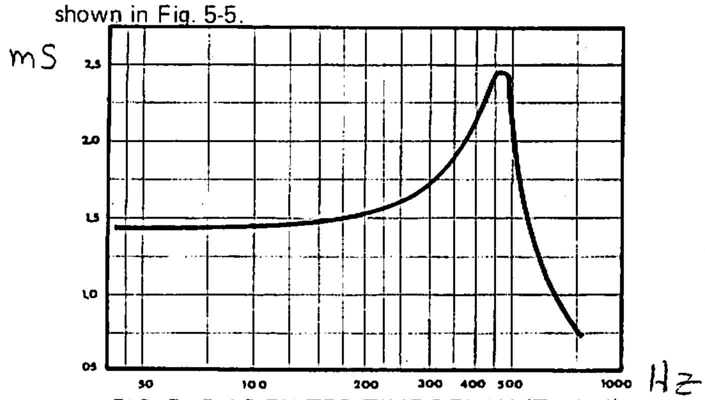


FIG. 5 - 5 LP FILTER TIME DELAY (Typical)

Observing the "B" scope output will allow the operator to readily discern the presence of noise in the measurement. If a measurement must be made and the signal to noise cannot be readily improved by improving the test device, a wave analyzer may be used at the "B" scope output. Such a technique allows high resolution and ability to identify and measure the distortion terms. If a true RMS distortion measurement is required it may readily be computed from the data. The open circuit voltage at "B" scope corresponding to full scale on the meter is 40 Volts for this unit serial number 717 . All distortion readings from the "B" scope terminals should be normalized using this factor and the associated IM range sensitivity.

The "B" Scope signal may be used to indicate the second derivatives of the ω_2 transfer function (incremental gain change) vs Eout if the following oscilloscope X-Y display is executed, Fig. 5-6.

A large LF:HF ratio such as 10:1 would probably be preferred in this test to provide increased resolution. The special signal source for the X display is necessitated by the need to phase shift this signal in order to compensate for phase distortion in the amplifier under test and to compensate for the time delay of the LP filter. Other low frequency test frequencies (other than line frequency) may be used if an appropriate phase shifter is constructed. The line cord phasing and phasing control should be adjusted such that the extreme right hand scope deflection coincides with the time of peak positive output of the device under test. The display will then show high frequency average gain deviance vertically coinciding with low frequency output voltage horizontally.

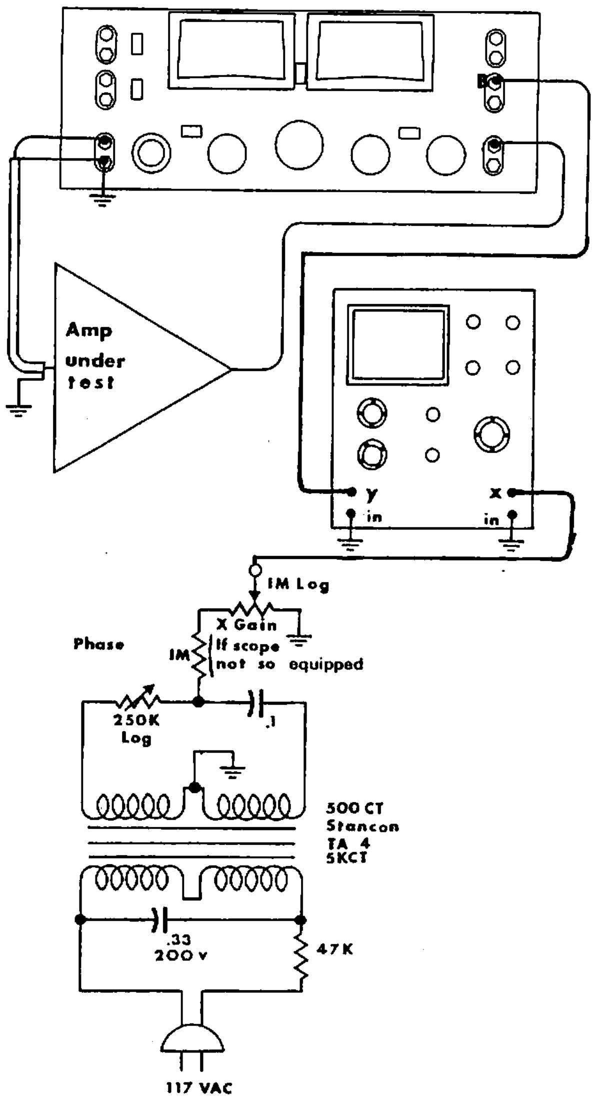


FIG. 5 - 6 TRANSFER FUNCTION SCOPE DISPLAY

The "A" scope output may be used for oscillographic distortion analysis in high distortion devices where distortion of the high frequency envelope becomes visible. Such a method is LeBel's[9] oscillographic method. This method uses the "A" scope output on the oscilloscope's vertical display while the horizontal is syncronized with the low frequency test signal. The distortion is computed by arithmetically summing the modulation percentages (baseline referenced) of all visible "notches" or areas of diminished peak output on the display. All notches of the upper and lower halves of the display are each added to form the total which could obviously be much greater than 100%. The stated reference shows the following empirically determined graph, Fig. 5-7, which will roughly relate this total to per cent IM as read by the meter. The relation there shown is not totally quantitative, as implied in the reference, as can be evidenced by Fig. 5-8.

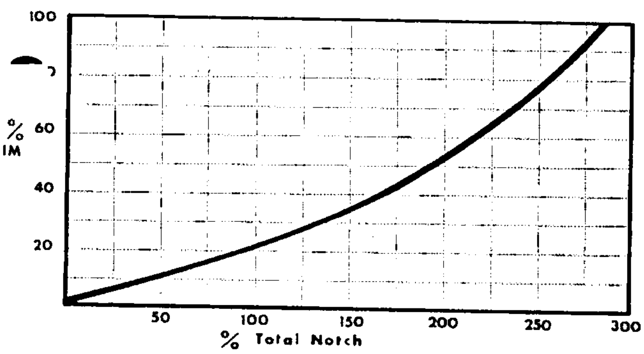


FIG. 5 - 7 CONVERSION GRAPH FOR LE BEL'S METHOD

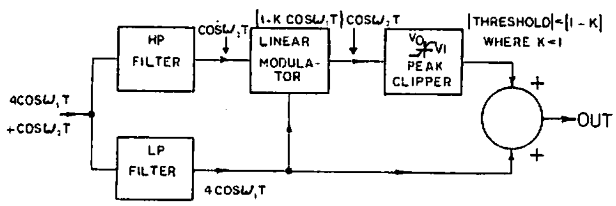


FIG. 5 - 8 A HIGH DISTORTION SYSTEM HAVING NO PEAK HF ENVELOPE DISTORTION

Viewing the "A" scope output in cases where the measurement is being polluted by power line frequency coming from the test device or test hookup often proves instructive. Often in low distortion (but polluted) readings a line frequency transient can be viewed on the "A" scope signal by careful inspection.

One of the prime virtues of the IMA is the facility with which measurements can be taken in the tracking mode of operation.

The tracking mode of operation was designed with 5db steps which allows 2 points per cycle to be plotted when plotting % distortion vs log output watts on semilog paper. (4 points per cycle when plotting log output volts on semilog paper.) All test points will be equally spaced (distance-wise) along the log output scale. Fig. 5-9 is an example of such a plot.

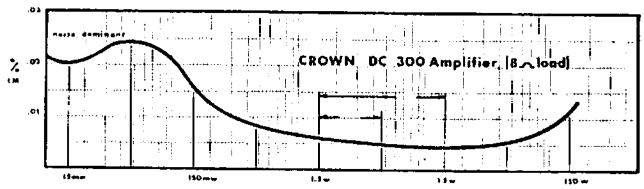


FIG. 5 - 9 TYPICAL AMPLIFIER DISTORTION PLOT LAYOUT

If more test points (data) are desired the output level can be shifted 2.5db and 10 more points can be plotted, each point in power being half way between previous adjacent test points.

The aforementioned sensitivity of the IMA to cross over notch distortion in class AB amplifiers can be evidenced in the following graph. The amplifier plotted was reputed to have < .07% IM at 100 W out and < .3% at IW out. While this did not prove to be true the most important deficiency was hidden by the choice of test points. Note the rise to a whopping 8.6% at 31.6 mw. The commercial practice of testing IM distortion at 1W and above is clearly negligent. For a 100W amplifier 1W constitutes only a 20db decrease in level. Wide range musical program material frequently has a dynamic range of 60db. While it is not necessary to test an amplifier over this wide a range it is necessary to test over all of the range that could possibly show crossover notch distortion. Experience has shown that most crossover notch distortions are found in the 10mw to 1W power range for direct coupled class AB solid state amplifiers driving 4 to 16 loads.

9) C. J. LeBel's "A New Method of Measuring and Analyzing Intermodulation" And. Eng. (July 1951) p. 18.

Crossover notch has been the primary villain that has given solid-state amplifiers a black eye. The previously adequate practice of measuring distortion from 1W upwards was then eagerly adopted with improbity. If such a product is to be properly and honestly evaluated IM distortion must be measured and from 10mw and upwards.

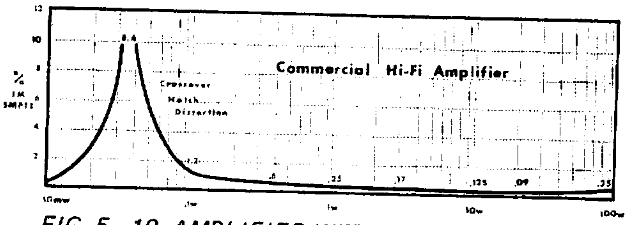


FIG. 5 - 10 AMPLIFIER WITH CROSSOVER NOTCH DISTORTION

The IMA has the unprecedented sensitivity to measure down to 900 micro Watts (into 8Ω). No conceivable notch distortion would be below this level in a direct coupled amplifier.

5.2 PRODUCTION USAGE

The extreme speed of distortion measuring makes the IMA a natural choice for production line distortion testing.

Once initial setup has been made only minor adjustment is necessary to accommodate production variances in the device under test. In the case of amplifiers (flat response) being tested tests are conducted at prescribed output levels which correspond to a fixed setting on the INPUT level control. In such a circumstance only the OUTPUT level control would need adjustment from unit to unit and only so much as the overall device gain varied. The CALIBRATE meter can be used as the output level meter in most situations (providing distortion is moderate and the LF:HF ratio at the test device output does not show variance). Otherwise a peak reading meter or oscilloscope would be needed to establish the output level. In many circumstances the IMA setup (OUTPUT level adjustment) may give ample indication of gain deviance to eliminate a special gain test for the devices under test.

Some applications may be characterized by requiring a fixed input signal level while requiring compensation for gain variance with the IMA INPUT level control. An example of this would be amplifier high frequency noise figure testing as discussed in section 5-3. If gain variances are small an amplifier under distortion test (at low levels) may be tested for high frequency noise (hiss) by shutting off the LF generator and reading noise. Thus the IMA may eliminate special tedious high frequency noise measuring testing.

Settling time of the IMA (after overloading the "B" scope amplifier and meter amplifier) are the basic deterrent to unlimited testing speed. Overload recovery is complete after approximately four seconds for the lowest distortion levels. At higher distortion levels the time is less. This speed however is commensurate with human recording rates. For instance, if an operator is in the tracking mode of operation if he first makes a reading, next changes level (1 or more steps) and then records (writes down) the previous reading, the IMA will be ready for taking the next reading by the time the former reading has been recorded.

5.3 OTHER USAGES

5.3-1 Amplifier Noise Figure Testing

As mentioned in section 5.1 noise influences the IMA in a predictable manner. Therefore if all causes of intermodulation are removed from the device under test, as can usually be accomplished by removing the low frequency part of the test signal, only noise will be measured. The noise that is measured will be mostly hiss and gathered in a highly rectangular band of 1KHz bandwidth around the high frequency test signal plus some subdued responses as depicted in Fig. 4-2. If the higher order responses are not desired they may be removed by filtering. Figure 5-11 shows a filter designed for use with a 7KHz test frequency and a test amplifier whose gain would not be adversely affected by a 5 KΩ load.

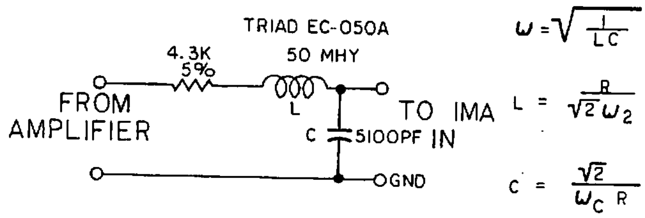
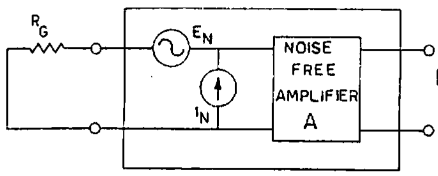


FIG. 5 - 11 LP NOISE FILTER (2 pole Butterworth)

Thus the use of the IMA gives a form of spot noise measurement with a very rectangular 1KHz bandwidth. With a 1KHz bandwidth averaging time is acceptable for many production testing purposes. Only for high accuracy applications will a slower metering system be needed. Such can easily be adended by using the "B" scope terminals. If desirable the IM meter may be heavily damped by placing a 100mfd or larger capacitor across the meter terminals of M2.

Any amplifier may be characterized by the following noise model, Fig. 5-12.



$$E_{OUT} = A\left[\overline{E_N^2} + 2\delta R_G\sqrt{\overline{E_N^2}\cdot\overline{I_N^2}} + R_G\,\overline{I_N^2} + 4KTR_GB\right]^{1/2}$$

FIG. 5 - 12 MODEL OF A NOISY AMPLIFIER

The first three terms in the expression for $\overline{e_{out}}$ are the excess noise of the amplifier expressed in terms of fictitious generators placed at the input. The factor δ is the cross-correlation coefficient of $\bar{e}_n$ and $\bar{i}_n$. Components of e_n and i_n that are of common internal origin will have a non-zero δ ranging from -1 to +1. Components of e_n and i_n having no common origin have a δ of zero. The latter ter ›ntaining the 4KTRgB is the thermal noise generated by the source resistance Rg. K is Boltzmann's constant = 1.38 X 10^{-23} joule/degree Kelvin, T is the temperature in degrees Kelvin and B is the measuring bandwidth in hertz. For the IMA B is 1KHz or $\approx$ 1.1KHz if the total noise is white to five times the HF test frequency and not filtered out of the measurement (see section 5.1). $\bar{e}_n$ and $\bar{t}_n$ are here taken to be white and averaged over B.

To evaluate the performance of the amplifier with such an Rg it is given a noise figure which is defined as:

$$\text{N.F.(db)} = 20\log\frac{e_{out}\,(\text{thermal}) + e_{out}\,(\text{due to } e_n \text{ and } i_n)}{\overline{e_{out}}\,(\text{thermal})}$$

The noise figure for a noiseless amplifier would then be Odb, and would be read by reading the db scale of a voltmeter measuring noise and subtracting the component (calculated) of $\overline{e_{out}}$ (thermal) from the reading.

Evaluation of noisy amplifiers with the IMA is simplified by the following:

I. Most devices have a spectrum that is white over each or all response frequencies of the IMA and A, $\bar{e}_n$, and i_n are not functions of ω but constants.

II. ‿e form of the computation for bandwidth is ..mplified by the fact that the frequency responses are highly rectangular eliminating the need for special correction factors due to non-rectangular filter responses.

III. The factor A is implicitly measured while executing the test eliminating the two step testing needed for conventional noise tests. Production line variances in A are compensated by adjusting the INPUT level control for the calibrate condition (f.s. on calibrate meter).

IV. The test frequency may be quickly changed by varying the frequency of an external oscillator (when using external HF GENERATOR). Note that B is not affected by this operation.

The quality of the external oscillator (low amplitude modulation components) is very important when using external HF source.

V. Gain instability (2-500Hz) is heavily penalized in this test even though it may not actually constitute a large quantity of noise in the form of $\bar{e}_n$ or $\bar{i}_n$ in the band being measured. The common causes of such effects such as "the popcorn effect" in semiconductors where a bipolar device will randomly alternate between two states of h_{fe} will produce high IM products. Ordinarily a number of conventional noise vs. f tests might be necessary to detect this noise generating defect.

The "popcorn" effect may also be detected by an oscilloscope and a trained eye watching the noise signal (HF off) where it will appear as a number of random period "square" waves in the noise signal.

Rg should be selected to typify normal operating conditions for the amplifier. If typical conditions vary widely Rg can be switched and multiple readings taken. For bipolar transistors an Rg $<$ 100Ω will measure $\bar{e}_n$, while IMΩ would measure i_n. Reactances should be minimized around Rg if absolute accuracy of test is desired.

On the other hand Rg can be substituted by Z g if the - plifier application requires. The noise computation then contains Z g in the excess noise terms and Z g (real part) in the thermal noise expression. In general Z g (real part) must be determined by direct measurement at the analyzer test frequency. For example a magnetic phonograph cartridge has a Z g (real part) at 7KHz that is much greater than the DC resistance that would be measured by an ohmeter. Its Z g (real part) will take on a maximum at the frequency of self resonance where the system capacitance resonates with the cartridge inductance. Z g (real part) may be in the 100's of Kilohms at this frequency which is indicative of a great deal more thermal noise (hiss) than would be predicted by several hundred ohms of DC resistance. Fig. 5-13 is provided for ease of computing $\bar{e}_{Rg}$ (thermal).

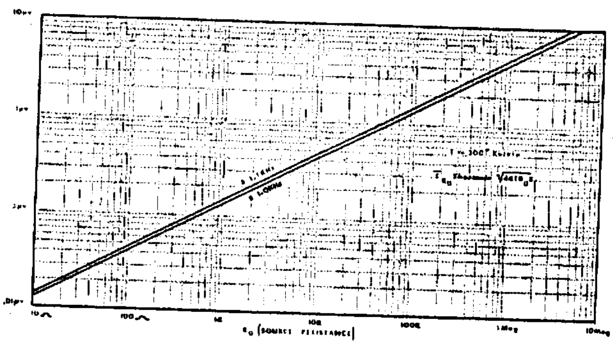


FIG. 5 - 13 THERMAL NOISE vs. SOURCE RESISTANCE

The following example (Fig. 5-14) is provided to illustrate the typical set-up and computations involved.

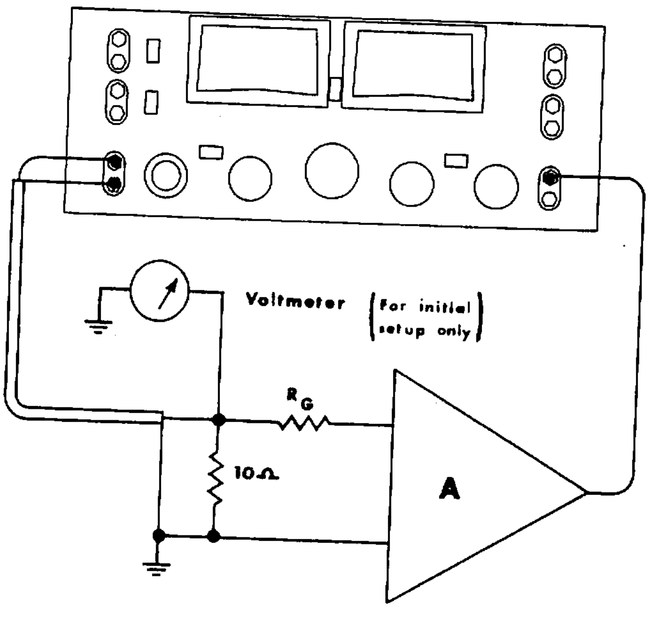


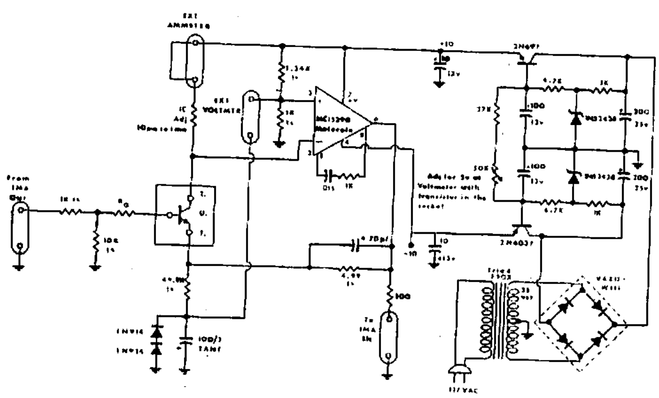
FIG. 5 - 14 AMPLIFIER NOISE TEST EXAMPLE

Let's assume A to be $\approx$100 or $\approx$ to 40db; then for 20mV at the IMA input (enough to test in MANual mode) the amplifier input will need to be adjusted to 200 micro Volts. If the external voltmeter cannot measure 200 micro Volts the IMA output attenuators can be set for 20mV and then attenuated 40 more db which is an accurate method since they are precision attenuators. Choosing Rg = 10K and looking upon figure 5-13 we find this represents .43 micro Volts of noise at B = 1.1KHz (A is assumed to evidence a flat or white noise output past 50KHz, otherwise the filter of Fig. 5-11 should have been placed in the output signal line and B = 1KHz read). This noise signal represents .215% of the amplifier's input test signal (.43/200). In other words if the amplifier under test were perfect the IM meter would read .215%. The amount that it reads greater than this is the excess noise or in db, the noise figure. If the meter read 0.3% the noise figure would be 2.9db ($20 \log \frac{.3}{.215}$).

Now if this test amplifier were to be substituted by a similar amplifier the only setup adjustment necessary would be to readjust the INPUT level control of the IMA for the calibrated condition (100%) on the CALIBRATE meter, since amplifier noise measurements are referenced by the signal input to the amplifier.

5.3-2 Transistor Noise Figure Testing

Since most good amplifier designs are limited by the noise figure of their bipolar input transistors it is often desirable to pretest the transistors before placing them in an amplifier. Figure 5-15 shows a typical test jig for such a pretest operation. The jig is designed for testing NPN bipolar transistors at 5 volts Vce.



Ic may be adjusted (1 μA to 1mA) by varying the transistor's collector supply resistor which should be a low noise metal film or wire-wound type resistor. Ic as measured at the external meter terminals will typically be

200nA greater than the actual due to the MC1539's input bias demands. Typically this resistor would be equal to $\frac{5.6}{Ic+2x10^{-7}}$.

To avoid Vce errors due to voltage drop across Rg, that is $\frac{IcRg}{h_{fe}}$ should be held to .5Volts or less.

Measurement proceeds in the same manner as in the example of section 5.3-1 except a 40.1db preattenuator is built into the input of the jig. All input signals should be divided by this factor to find the true amplifier input level.

In this circumstance changing transistors should not require any change of the IMA's setup. If the test level were to indicate low it would be indicative of a bad transistor under test. The system feedback is very high and the measured noise should be solely from the transistor under test (TUT) and would typically be white over the measurement range.

5.3-3 Tape Drop-out Evaluation

In magnetic tape recording rapid fluctuations of recording sensitivity due to problems in the recording media are referred to as drop-outs. Those changes in sensitivity which fall within the low-pass filters response range may be metered by reading the modulation inherent in the playback of a recorded tone (single frequency).

The frequency range of the IMA for such measurements may be extended down to 1KHz by using the instrument with the METER switch in the OUTput position with the tape recorder feeding the EXTernal HF generator input (No LF generator on). The HF level should be adjusted for full scale on the calibrate meter. Any hum or low frequencies will now be metered as modulation therefore the use of this mode of operation is limited to low-noise and/or high modulation noise applications.

Since drop-out phenomena contains a great abundance of low frequencies a highly damped meter reading the "B" scope output will probably be found necessary.

Figure 5-16 shows a plot of tape drop-out vs. recorder frequency.

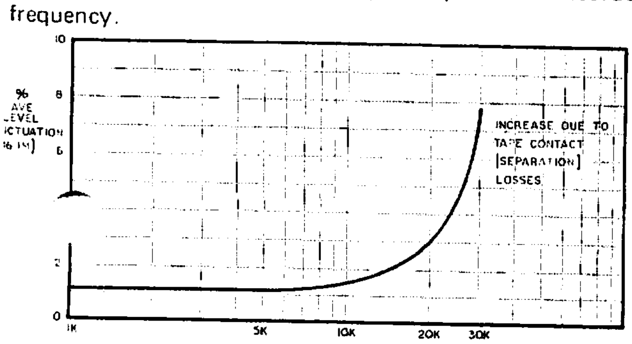


FIG. 5 - 16 TAPE DROP-OUT vs. FREQUENCY

5.3-4 Testing Oscillator Amplitude Stability

Electronics oscillators may be tested for amplitude stability (3-500Hz instabilities). By inputting the oscillators output into the EXTernal HF generator input, switching the calibrate METER switch to OUTput (no LF source) and adjusting for full scale deflection of the calibrate meter, frequencies down to 2KHz may be tested without any ill effects from oscillator FM increasing the residual. (FM is converted to AM if acted on by the HP filter near is cutoff frequency.) This method does require that the oscillator not have substantial hum or other such low frequencies in its output that would be metered. If this proves to be so the conventional way of feeding the signal thru the IMA input (and thru the HP filter) must be used.

Fig. 5-17 shows a very stable oscillator plotted for both IMA input methods. The standard method shows the residual FM of this otherwise very stable oscillator.

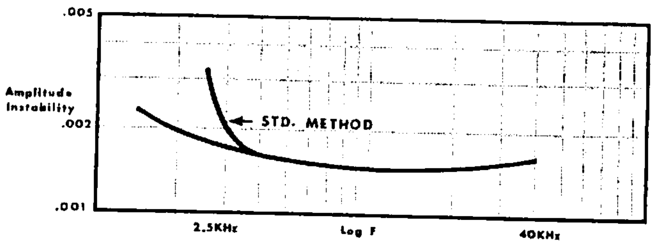


FIG. 5 - 17 OSCILLATOR AMPLITUDE STABILITY vs. FREQUENCY

5.3-5 Magnetic Tape Recorder Distortion Testing

It is particularly enlightening to test tape recorders for SMPTE intermodulation distortion since partial optimization is often available to the user in the form of a bias adjustment and because tape recorders are highly band limited devices. (Devices that have sharp high frequency cut-off characteristics will not evidence harmonic distortion if the distortion components are above the cut-off frequency.)

Fig. 5-18 shows a plot of IM vs. frequency run at a peak equivalent playback level of Odb (NAB) for several settings of bias. Figure 5-19 shows a plot of IM vs. level for several settings of bias. Note that the ultimate residual is limited by tape drop-out phenomena, (Fig. 5-16); therefore Figs. 5-18, 19 were plotted by using a wave analyzer set to 120Hz the dominant distortion component seen at the B scope output.

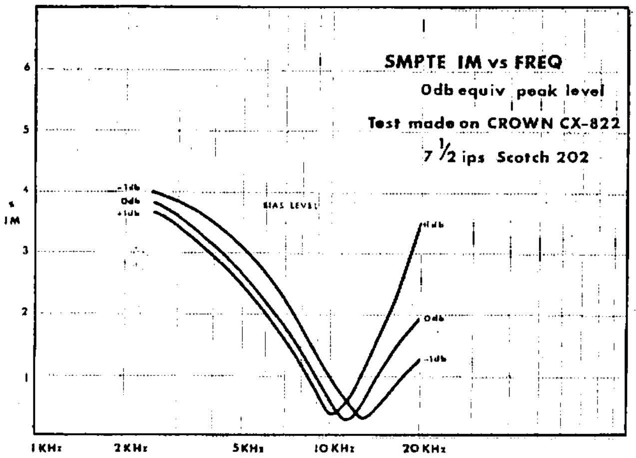


FIG. 5 - 18 TAPE RECORDER IM vs. FREQUENCY

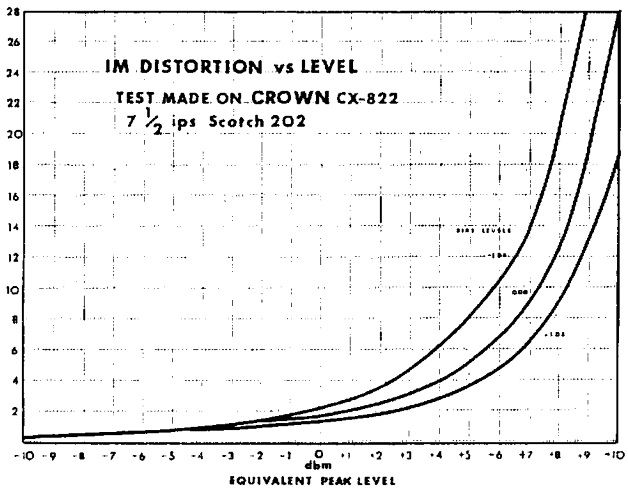


FIG. 5 - 19 TAPE RECORDER IM vs. LEVEL

Viewing the distortion components at B scope should evidence only even order harmonics of 60Hz and drop-out noise. The presence of 60Hz or odd order harmonics of 60Hz would indicate even order non-linearities which are not properties of the tape as it possesses only odd order non-linearities. Such distortion could be caused by magnetized heads, distorted bias wave form, or electronic distortion.

5.3-6 Phonograph Stylus Wear and Tracking

To test a phonograph stylus(10) IM distortion it is first necessary to use a pre-recorded IM test record. The following is a list of applicable test records:

1. Cook Laboratories, 101 2nd St., Stamford Conn. "Stereo IM Dist. Test, Series 300".
2. CBS Laboratories, High Ridge Rd., Stamford, Conn. "Sq. Wave, Tracking, and IM Test record, STR 110".

Note that a number of these records use a 600Hz low frequency test signal. Since the IMA's LP filter is at 500Hz only the first order IM product will be metered which may impair the overall accuracy of the measurement. Fortunately conditions such as mistracking and worn stylus tips usually create large amounts of distortion which is usually easily measured and/or viewable at the A scope output. The use of LeBel's oscillographic method as discussed in section 5.1 may therefore prove useful.

The reference cited by Roys shows IM testing to be a particularly potent method for analyzing not only reproducing distortions due to flatted stylus tips and distortions inherent in the geometries of the recording-reproducing cycle but also the condition of the master used for making the subsequent pressings. Polishing of the master was found to generate large amounts of IM distortion but not harmonic distortion!

The reason for mistracking causing IM should be obvious. If the low frequency can cause the stylus tip to momentarily lose contact with the groove walls the high frequency will be momentarily interrupted reducing its level and thus generating IM.

5.3-7 Testing Electro-Mechanical Transducers

Electromechanical transducers such as loudspeakers and microphones may be tested for intermodulation if a reciprocal process of high linearity can be constructed. For instance if a loudspeaker system is to be tested a microphone must be used to reconvert the acoustical signal to an electrical signal for analysis. Obviously to use a microphone that is more non-linear than the speaker system would be to render all readings meaningless.

A useful technique which is sometimes applicable is to vary the signal incident upon one transducer without greatly disturbing the other. If the total distortion is essentially unaffected by such a process the transducer that underwent the signal level change is probably not responsible for the measured distortion. In the case of the loudspeaker tests if moving the microphone away from the speaker (on axis) does not modify the measured distortion then the microphone is probably not contributing to the distortion (care must be used to avoid cancellation effects near the HF test frequency due to reverberant and/or multiple source signals).

10. H. E. Roys "IM Distortion Analysis as Applied to Disk Recording and Reproducing Equipment" Proc. I.R.E. (Oct. 1947) p. 149.

For loudspeaker testing condenser microphones of the pressure sensitive type will usually give more than adequate linearity in sound pressure fields of reasonable intensity. High intensity sound fields will ultimately displace the diaphragm sufficiently to evidence IM due to the sensitivity being an inverse function of the diaphragm spacing.

When testing loudspeakers and microphones varying the test frequencies may be very instructive due to the effects of mechanical displacements vs. frequency and of crossovers in the systems.

IM testing of moving coil cone type woofers is primarily a measure of linearity of the magnetic flux field coupling the voice coil. Harmonic distortion tests on such a device may be very misleading concerning this particular parameter.

The testing of high linearity devices may very likely require linear filtering techniques coupled with the test transducers. For instance the example of loudspeaker testing might require removal of the low frequency from being incident upon the microphone. The simplest method of achieving such an effect would be to form a low frequency reverberant sound pressure null (on Axis) and place the microphone in this null. If worse comes to worst acoustic filters can be constructed using Helm holtz resonators (11).

5.3-8 Miscellaneous

I. Optical Film Recording[12] [13] [14] - IM meter is used to optimize the developing and register.

II. CC1F Type IM Measurement[15] [16] [17] - The CC1F method differs from the SMPTE method in that two closely spaced high frequencies are mixed in a 1:1 ratio and applied to the device under test. The device output is then passed thru a bandpass filter that responds only to the difference frequency between the two high frequency test frequencies. This method is only useful for detecting distortion generated by devices having a nonsymmetrical transfer function. While the IMA will not measure CC1F distortion it is capable of performing the two tone signal mixing process with ultra low distortion.

(11) L. E. Kinsler and A. R. Frey "Fundamentals of Acoustics" (book) Chapt. 8 John Wiley 1962.

(12) J. G. Frayne and R. R. Scoville "Variable Density Recording" Jour. SMPE (June 1939).

(13) G. W. Read and R. R. Scoville "An Improved Intermodulation Measuring System". Jour. SMPE (Feb. 1948).

(14) J. K. Hilliard "Distortion Tests by the Intermodulation Method" Proc. I.R.E. (Dec. 1941) p. 614.

(15) H. H. Scott "Audible Audio Distortion" Elec. (Jan. 1945) p.126.

(16) E. W. Berth-Jones "Intermodulation Testing" Wireless World (June 1951) p. 233.

(17) D. E. O'N Waddington "Intermodulation Distortion Measurement" Jour. Aud. Eng. Soc. (July 1964) Vol. 12, No. 3, p.221.

5.1 ROUTINE CALIBRATION CHECKS

5.1.1 Introduction

The IMA is particularly unique in that it features many self-testing features. Even the distortion metering system can be quickly calibration checked from front panel controls with no external equipment. The following steps should be performed in order.

5.1.2 Preliminary Steps

Plug in the line cord and turn the instrument on. The meter lights which illuminate the panel meters should come on. The meter indications may be either on or off scale, depending on the settings of the front panel controls. Proceed to preset the front panel controls as follows:

HF GENERATOR switch - EXTernal

LF GENERATOR switch - EXTernal

OUTPUT level 10DB - Odb
VARiable - fully CW

METER switch - OUTput

... level - 12 o'clock

MASTER attenuator - Odb

IM% - 100%

TRACKING switch - MANual

INPUT level - fully CCW

The meter reading should now read nearly zero on the CALIBRATE meter and zero on the DISTORTION meter. Do *not* mechanically zero the CALIBRATE meter because its zero adjustment is used to track the meter. See section .2.8. The DISTORTION meter should now be mechanically zeroed with a screwdriver thru the front panel access hole if it does not show zero.

.1.3 Test for Oscillator Operation and IM Calibration

) Set the LF GENERATOR switch to INTernal. The CALIBRATE meter should read full scale, 100%, indicating normal LF oscillator operation. The DISTORTION meter should read % indicating correct calibration of the distortion metering system.

(2) Set the LF GENERATOR switch to EXTernal and the HF GENERATOR switch to INTernal. The CALIBRATE meter deflection should now be adjustable from the previously observed meter zero to in excess of full scale indicating that the HF oscillator is operating. Adjust the HF level to read 25% deflection on the meter. Use care not to change this adjustment throughout the following tests.

6.1.4 Output Attenuator Checks

(1) Connect a wire (preferably shielded) between the OUTput binding posts and the INput binding posts. Set the METER switch to the INput position and adjust the INPUT level control for a full scale (100%) deflection on the CALIBRATE meter. Switch the OUTPUT level (10DB) control down 10db to the -10db position. The CALIBRATE meter should now read at or very close to 31.6%. Now readjust the INPUT level for a 100% deflection and turn down the OUTPUT level (10DB) another 10db (-20db position). The meter should again measure 31.6%. Repeat this process of adjusting the INPUT level, adjusting the OUTPUT down 10db, and reading the meter until all positions of the OUTPUT level (10DB) control have been tested.

(2) Return the OUTPUT level (10DB) control to the Odb position and adjust the INPUT level control for full scale (100%) deflection on the CALIBRATE meter. Turn the OUTPUT level (VARiable) control fully CCW. The CALIBRATE meter should now read 18% or less indicating that the OUTPUT VARiable has at least 15db of range. Reset the VARiable control to fully CW.

(3) Set the MASTER attenuator to the -5db position. The CALIBRATE meter should now read very nearly 56.2%. Set the MASTER attenuator to the -10db position. The CALIBRATE meter should now read 31.6% as evidenced when testing the OUTPUT level (10DB) control. Restore the CALIBRATE meter to 100% and continue the process of checking the MASTER attenuator 5 and 10db per INPUT level setting in a manner similar to testing the OUTPUT level (10DB) control. Upon completion of testing the MASTER attenuator turn the INPUT level fully CCW and then return the MASTER attenuator to the Odb position.

6.1.5 Input Attenuator Tracking and Residual Distortion Tests

Set the LF GENERATOR switch to INTernal, the TRACKING switch to TRacKing, the IM% switch to the 0.1% range, and the INPUT level control fully CW. Adjust the OUTPUT level (VARiable) for full scale (100%) deflection of the CALIBRATE meter. Rotate the MASTER attenuator thru its positions. The CALIBRATE meter should not deviate more than ±2% around a nominal full scale setting which may be slightly different from the initial full scale adjustment. The DISTORTION meter after it has recovered from the large transient caused by changing the MASTER attenuator setting should not read more than .005% distortion. Read the meter by reading the general downswing of the meter. (Occasional transients—characteristic of the thermal-noise signal causing most of the observed residual distortion — may cause occasional meter deflections in excess of .005%. High RF noise or line voltage transients may artificially increase the residual. For this reason the wire coupling the INput and OUTput terminals should be shielded.) The worst meter reading should be obtained at the -40db position of the MASTER attenuator where thermal noise will be at a maximum.

Having completed these checks the routine calibration check procedure is now complete.

6.2 COMPLETE CALIBRATION

6.2.1 Introduction

The IMA is a reliable instrument, and as such, should give long trouble-free service. However, if routine calibration tests indicate any discrepancies or as a complete assurance that component aging is not degrading performance the following calibration procedure should be carried out.

Because most of the adjustments interact, the entire calibration procedure should be performed in sequence. Any decision to make only a partial calibration should be reserved to someone having sufficient knowledge of the instrument to determine where interaction is a problem and where it is not. The location of the various adjustments and test points is indicated in Figs. 7-3 and 7-4.

It is recommended that sections 3 and 4 of the manual be read before attempting calibration. Only those knowledgeable in servicing electronic instruments should attempt to repair or calibrate the unit.

CROWN maintains a factory-repair department for those customers not possessing the necessary personnel or test equipment to maintain the instrument.

6.2.2 Equipment Needed

Name	Required Characteristics	Recommended Type
Oscilloscope	DC-10MHz	Tektronix 561B
Plug-in	DC-10MHz	Tektronix 3A6
Plug-in	Time Base	Tektronix 2B67
10X Prod	12pf input cap.	Tektronix P6006
AC Meter	1% f.s. accuracy	HP 400F
DC meter	2% f.s. accuracy	Triplett 601
Freq. Counter	.1% accuracy	Atec 1130A
Audio f. Sine Generator	Good amplitude stability .5% distortion	Wavetek 130's
Misc. Cables	To interconnect IMA with above equipment	
Alignment Screwdriver	Plastic to reduce danger of overtorquing, etc.	GC 8727

6.2.3 Preliminary Steps

(1) If the unit is mounted in a cabinet such as a 7-C cabinet remove the cabinet.

(2) Plug in the instrument, turn it on, and allow at least 30 min. of warm-up time before beginning any st[illegible] other than preliminary.

(3) Set the front-panel controls as follows:

HF GENERATOR	INT.
LF GENERATOR	EXT.
OUTPUT 10db	Odb
VAR	full CW
METER	IN
HF	12 o'clock
MASTER	Odb
IM%	100
TRACKING	MAN
INPUT	full CCW

(4) Adjust the DISTORTION meter mechanical zero to read zero. (Do not tamper with the CALIBRATE meter zero - see 6.2-8.)

6.2.4 Power Supplies

(1) Remove the top cover of the IMA by removing the 1[illegible] retaining screws.

2) Connect the DC voltmeter to ground and to jumper number 3. Adjust R166 (on the power supply board) for +30 VDC on jumper number 3. (See fig. 6-1).

3) Connect the DC voltmeter to jumper number 1. The voltage should read -30 to -31VDC.

6.2.5 HF Oscillator

1) Connect the frequency counter to the OUTput terminals.

2) Adjust R19 for 7KHz. Major adjustments of R19 may require readjusting R25 (7KHz level) to maintain oscillation, or to avoid oscillator-amplifier overload. Increasing the frequency by decreasing R19 tends to reduce the amplitude of oscillation and decreasing frequency vice versa.

3) Adjust R25 for -2.5VDC at point B. If the voltage is found greatly negative step 2 may require repeating as the amplifier was overloaded.

4) Switch the HF GENERATOR switch to EXT. The oscillator calibration is complete.

6.2.6 LF Oscillator

(1) Connect the oscilloscope to the OUTput terminals and syncronize (trigger) the timebase from the AC line. Set the LF GENERATOR switch to INT.

(2) Unplug the sync wire from the 6.3VAC and connect it to the adjacent ground terminal. This may cause oscillation to cease requiring adjustment of R8.

(3) Adjust R1 for a stationary sinusoidal 60Hz display on the oscilloscope. R8 may require retouching similarly as in step 2 of the HF oscillator adjustments.

(4) Reconnect the sync wire to the 6.3VAC.

(5) Adjust R8 for -3.5VDC at point A.

Modifications:

If the LF oscillator must be free running (not line synced) such as would be required when running with 50Hz power, yet desiring a 60Hz LF generator, use the following procedure.

(1) Set the LF GENERATOR switch to INT.

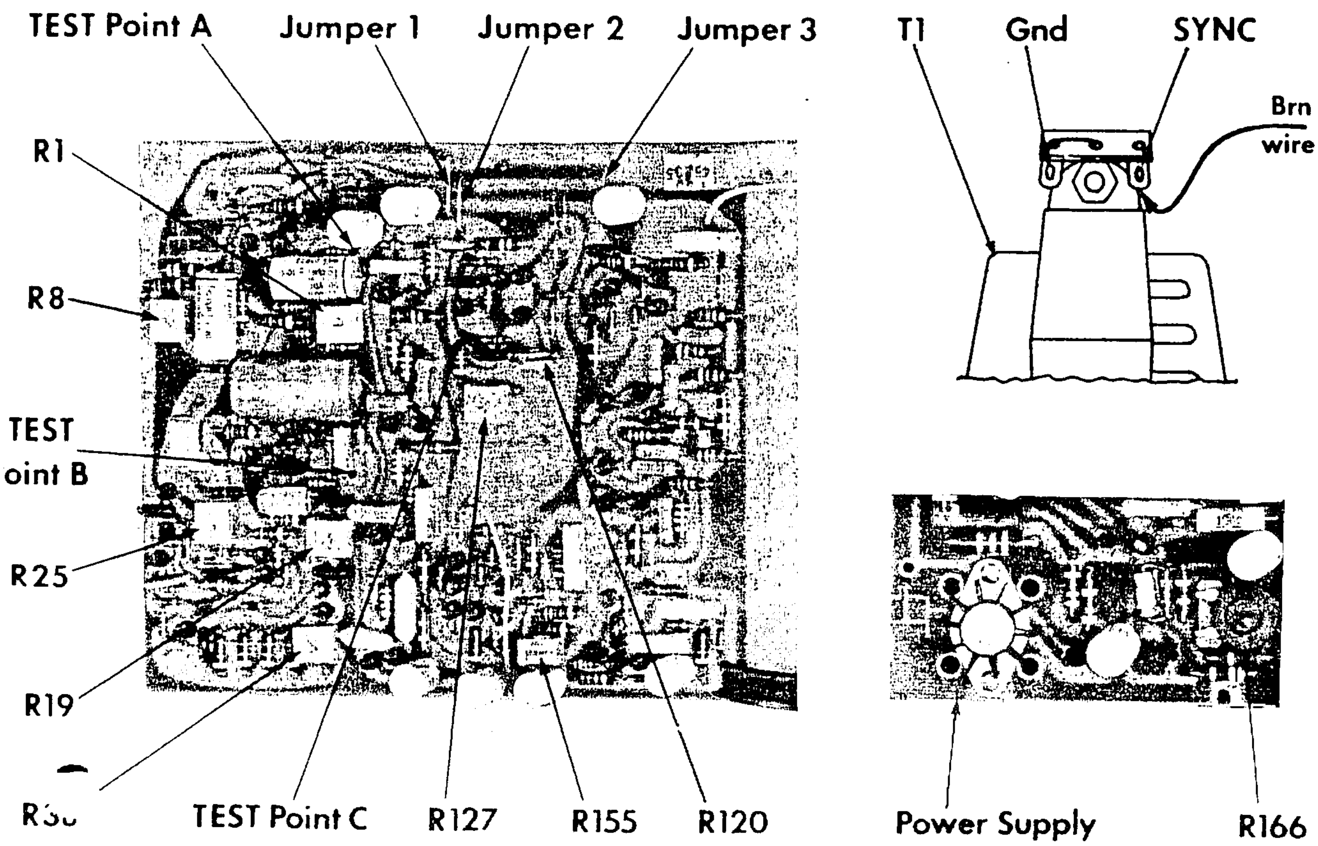


FIG. 6-1 LOCATIONS OF ADJUSTMENTS

(2) Connect a frequency or period measuring counter to the OUTput terminals that is capable of resolving 60Hz with sufficient accuracy. If such an instrument is not available omit steps 2,4.

(3) Unplug the sync wire from the 6.3 VAC and connect it to the adjacent ground terminal. This may cause oscillation to cease requiring readjustment of R8.

(4) Adjust R 1 for 60Hz on the counter. R8 may require retouching similarly as in step 2 of the HF oscillator adjustments.

(5) Adjust R8 for -3.5VDC at point A.

If the LF oscillator must be 50Hz line synced the oscillator frequency can be lowered by changing R3 to a 15K 5% film resistor and changing R2 to a 12K 5% film resistor (reuse R3, 12K 5% film). Use the same procedure for set-up as for a 60Hz line synced unit except read 60Hz as 50Hz.

6.2.7 OUTPUT Level Trim

(1) Connect the AC voltmeter to the OUTPUT terminals. Turn the OUTPUT level controls fully CW.

(2) The meter should read 14.14VRMS. If it does not essentially read thus it will be necessary to retrim R41 with a new R40. (This adjustment is to provide 50Vp-p SMPTE output and does not ordinarily require precision for most users therefore this section of calibration may be disregarded.)

If it is desired to replace R40 use only film type resistors, do not overheat R41, and above all do not short the op amp mixer's output (yellow wire) to ground.

6.2.8 Calibrate Meter Gain and Tracking - 45db ([illegible])

(1) Connect the OUTput terminals to the INput terminals and connect the AC voltmeter across the INput terminals.

(2) Set the LF GENERATOR switch to EXT and the HF GENERATOR switch to INT.

(3) Adjust the OUTPUT level controls for a 16mV signal on the voltmeter.

(4) Turn the INput level control fully CW.

(5) Connect the scope with 10X low cap prod to the output of the absolute value detector (point C). Adjust R120 for equal positive peak voltages of adjacent half cycles of the 7KHz signal.

(6) Adjust R127 for a full scale (100%) reading on the CALIBRATE meter.

(7) Adjust the OUTPUT level controls for a 4mV signal on the voltmeter.

(8) Adjust the CALIBRATE meter for a 25% deflection by adjusting the mechanical zero of the meter.

(9) Repeat steps 3,6,7, and 8 in sequence until no further adjustment of R127 or the CALIBRATE meter zero is required. Disconnect the test equipment and cables as the CALIBRATE meter is now calibrated.

6.2.9 OUTput Meter Adjust

(1) Set the HF GENERATOR switch to EXT, the LF GENERATOR switch to INT, and the METER switch to OUT.

(2) Adjust R38 for a full scale (100%) deflection on the CALIBRATE meter.

6.2.10 IM Calibrate

(1) Set the LF GENERATOR switch to EXT, the HF GENERATOR switch to INT, the METER switch to IN, and the IM% switch to 100.

(2) Connect the audio oscillator to the LF Generator terminals and adjust its frequency to approximately 7KHz. Parallel the AC voltmeter with the oscillator's output.

(3) Connect the frequency counter to the "B" scope terminals and connect the oscilloscope to the OUTput terminals.

(4) Adjust the external oscillator's level to produce a pattern that appears as shown in figure 6-2. Adjust the level carefully to obtain the optimum null point.

Nulls

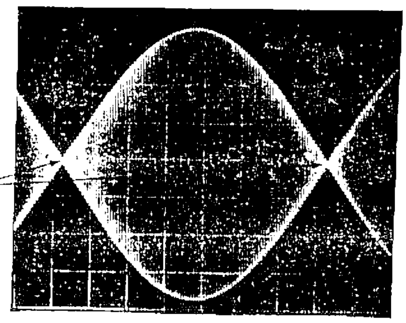


FIG. 6-2 IM GENERATOR SETUP

(5) Connect the OUTput terminals to the INput terminals and adjust the INPUT level control for 100% on the calibrate meter.

(6) Adjust the external oscillator frequency slightly to produce a 60Hz beat (read 60Hz on the frequency counter).

(7) Recheck the null point as set in step 4. Read the AC voltmeter. Divide this reading by ten and adjust the oscillator level to this new voltage. The consequence of this operation is a precision synthetic IM source of exactly 10%.

(8) Set the IM% switch to the 10% range. Adjust the INPUT level for 100% on the CALIBRATE meter.

(9) Adjust R155 for a full scale (10% distortion) reading on the DISTORTION meter.

(10) Disconnect the frequency counter from the "B" scope terminals and connect the AC voltmeter to the "B" scope terminals.

(11) Read and record the voltage indicated on the meter. This voltage is the "B" scope voltage as recorded in section 1.2. Remove the previous entry and record the new reading. The IMA is now completely calibrated.

6.2.11 Finish Check

Execute tests 6.1-4, and 6.1-5 of the Routine Calibration Check procedure.

6.3 TROUBLESHOOTING

6.3.1 Introduction

Although the IMA is a highly reliable instrument, component deterioration or failure could cause the instrument to no longer meet specifications, or become inoperative. Should this happen, recalibration per section 6.2 will usually restore proper operation. If the malfunction is a result of component failure it is unlikely that recalibration will be to any avail. The following information is intended to serve as a supplement to sections 4.3 and 4.4 from which most servicing would be intuitively obvious to anyone sufficiently familiar with the IMA to be qualified to service the instrument.

Once the faulty circuit is located (or before), the user should contact CROWN for advice on the relative merits of repairing it himself or returning it to CROWN for factory service. If the unit is still under warranty it is particularly advantageous to the customer to contact CROWN before attempting repair to avoid invalidating the warranty.

Initial Symptom	Test	Symptom of Test	Probable Defects
High Risidual Dist. (Noise at "B" Scope).	Residual for MAN and TRK modes for all MASTER positions.	Distortion highest at -40 in TRK.	Q22 noisy or circuit-board contaminated near Q22.
		Distortion highest at -45 in TRK.	RF is interfering with IMA.
		Distortion not dependent on MASTER or INPUT Level setting.	7 KHz oscillator defective; DC voltage at TEST POINT B too low; Q7 or Q12 noisy; R25 erratic; Analyzer section noisy; Q36, Q37, Q40, Q42, Q44, Q46 noisy.
High Residual Dist. (Hum at "B" Scope).	Residual with and without LF signal, and for various settings of OUTPUT Level (Var.).	Distortion not dependent on LF signal.	Power Supply defective, C9 open, or IMA interfered with by powerfull external hum-field.
		Distortion dependent on LF signal and on OUTPUT Level.	Oscillation in mixer; C10 thru C14 defective; C55 or C56 open; R41 defective (voltage coefficient shows high IM just above minimum gain).
		Distortion dependent on LF signal but not OUTPUT Level.	Oscillation in INPUT Amplifier or High-Pass Filter; C21, C24, C29, C55, C56, C57 defective.
Excessive mis-tr[illegible]g in TRK mode.	Par. 6.1.4(3)	OK	Mis-dressed wiring between SW1B and S7. R82 thru R91 defective.
		Not OK	R54 thru R72 defective.

FIG. 6-3 IMA TROUBLESHOOTING CHART

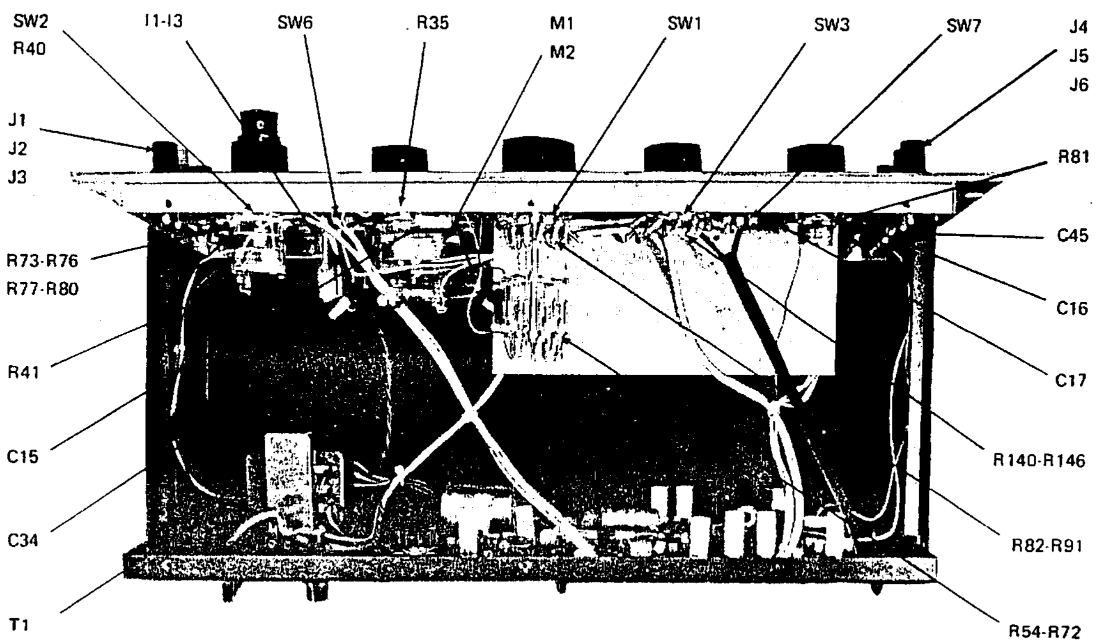
IMA PARTS LIST and ASSEMBLY

Schematic Ref.	CROWN Part No.	Description
CAPACITORS		
C15	1679	225 Mfd/30V elec.
C16, C17	1693	.0039 Pacer
C34, C45	2938	.1 Mfd. mylar
CONNECTORS		
J1, J2, J3, J4 J5, J6	2823A	Dual binding-post
CONTROLS		
R35	D 3097-9	5K linear pot.
R41	7856	25K, 5W lin. WW
R81	D 3439-3 ~~3101A~~	100 K - 2K dual pot.
SW1	3098	10 pos, 2 pole rot.
SW2	3100	5 pos, 2 pole rot.
SW3	3099A	7 pos, rotary
SW4, SW5, SW6 SW7, SW8	2668	DPDT Slide
FUSE		
F1	1758	¼ a. MDL
LAMPS		
11,12,13	1866	6.3V bayonet
METERS		
M1, M2	1987A	5" Hoyt, 200 μa.
RESISTORS		
R40		Selected
R54	3140	600 ohm 2½ W. 1%
R55	3153	1.07K, ¾ W., 1%
R56	3168	7.68K film, 1%
R57, R59, R61 R63, R65, R67 R69, R71	3111	732 ohm film, 1%
R58, R60, R62 R64, R66, R68, R70	3116	2.15K film, 1%

Schematic Ref.	CROWN Part No.	Description
RESISTORS (Cont.)		
R72	3113	931 ohm film, 1%
R73	3114	1.3K film, 1%
R74	3119	4.12K film, 1%
R75	3124	13K film, 1%
R76	3127	41.2K film, 1%
R77	3105	210 ohm film, 1%
R78	2990	43 ohm ½ W., 5%
R79	3057	18 ohm ¼ W.
R80	3109	604 ohm film, 1%
R82	3128	44.2K. film, 1%
R83	3126	24.9K film, 1%
R84	3125	14K film, 1%
R85	3123	7.87K film, 1%
R86	3120	4.42K. film, 1%
R87	3118	2.49K. film, 1%
R88	3115	1.4K. film, 1%
R89	3112	787 ohm film, 1%
R90	3107	442 ohm film, 1%
R91	3108	562 ohm film, 1%
R140	3122	6.81K. film, 1%
R141	3117	2.37K. film, 1%
R142	3110	681 ohm film, 1%
R143	3106	237 ohm film, 1%
R144	3104	68.1 ohm film, 1%
R145	3103	23.7 ohm film, 1%
R146	3102	10.2 ohm film, 1%
TRANSFORMER		
T1	1656A	117V pri./70 VCT 6.3V sec's.

MISCELLANEOUS

CROWN Part No.	Description	CROWN Part No.	Description
17837	Sub Panel	1262	Lamp Socket
7838	Front Panel	1903	Meter Bezel
7836	Back Chassis	2346	Black Toggle Cap.
2757	#8 x ⅝ Sht. Mtl. scr.	3157	Inner Con. knob
		3158	Outer Con. knob
1221	Fuseholder, HMB	1577	Large knob
2824A	Power Cord Cover	1622	Knob



IMA ASSEMBLY (Bottom View)

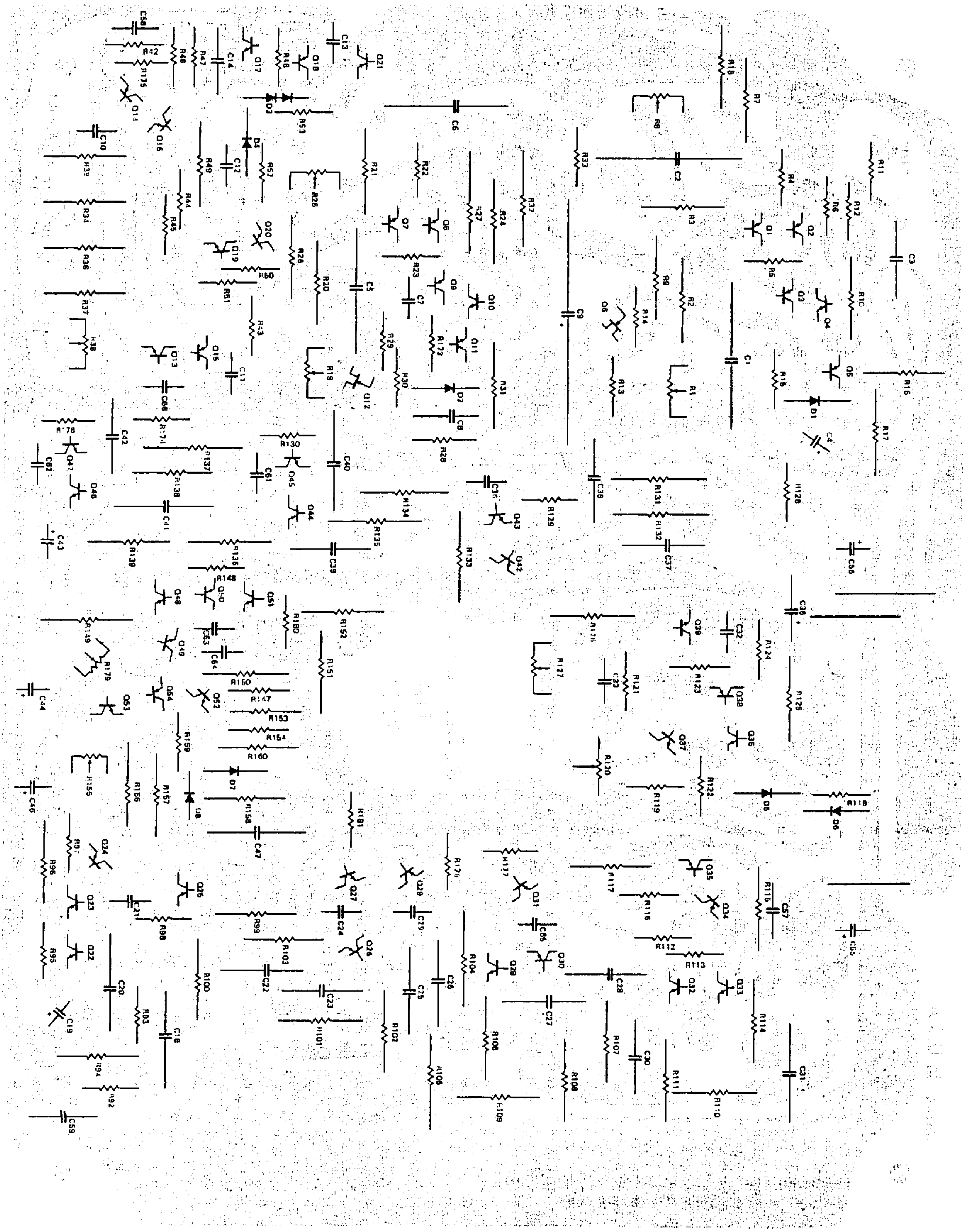
IMA MAIN-BOARD PARTS LIST

Schematic Ref.	CROWN Part No.	Description
CAPACITORS		
C1, C2	3142	.22 MF polycarbonate
C3, C39	2977	.047 MF Filmatic Mylar
C4, C19, C55, C56	2868	5 MF 30V Vertical
C5, C6	C3141-6	2200 PF polystyrene
C7, C10, C21	2821	10 PF mica
C8	C3063-2	.0082 MF Filmatic Mylar
C9	C 4147-2 ~~1679~~	225 MF 30V
C11, C13, C66	3411	200 PF Mica
C12	3409	47 PF Mica
C14, C59	1751	.01 MF disc ceramic
C18, C20, C31, C47	2938	.1 MF Filmatic Mylar
C22, C23, C25, C26, C27, C28	1693	.0039 MF Pacer
C24, C29, C36, C61, C62, C63, C65	3410	100 PF Mica
C30	3178	.0047 MF Filmatic Mylar
C32, C57	2820	5 PF Mica
C33	2342	27 PF Mica
C35	1750	1 MF Tantalum
[illegible]	1665	.015 MF Filmatic Mylar
[illegible]	3288	.012 Filmatic Mylar
C40	2198	.018 MF Pacer
C41	3218	.22 MF Filmatic Mylar
C42	3161	0.1 MF Filmatic Mylar
C43	3679	25 MF 35V Vertical
C44, C46	3729	100 MF 16V Vertical
C58	2241	.002 MF Ceramic disc.
C64	2288	.001 MF Ceramic disc.
DIODES		
D1, D2, D7, D8	3181	1N4148
D3	3039	MZ2361
D4, D5, D6	3447	1N270
RESISTORS		
R1, R38	3092	5K Cermet Pot
R2, R17, R21, R31, R100	2343	10K 1/2 Watt 1% Film
R3	3132	12K 1/2 Watt 5% Film
R4, R22, R112, R122, R147	3002	270K 1/4 Watt 5% Film
R5, R23	2633	18K 1/4 Watt 10%
R6, R43, R44	2632	15K 1/4 Watt 10%
R7, R24	1059	3.9K 1/2 Watt 10%
R8, R25, R179	3668	500 ohm Cermet Pot
R9, R26	3144	1.3K 1/2 Watt 5% Film
R10, R18, R27, R157, R158	3146	3.3K 1/2 Watt 5% Film
R11, R12, R128, R153	2883	100K 1/4 Watt 10%
R13, R47, R48, R50	2626	470 ohm 1/4 Watt 10%
R15, R29, R30	2882	56K 1/4 Watt 10%
[illegible], R32	3195	27K 1/2 Watt 1% Film
R19	3094	2.5K Cermet Pot
R20	3148	9.1K 1/2 Watt 5% Film
R28, R175	2873	180 ohm 1/4 Watt 10%
R33	2872	100 ohm 1/4 Watt 10%

Schematic Ref.	CROWN Part No.	Description
RESISTORS (Cont.)		
R34, R36, R37, R114	3147	5.6K 1/2 Watt 5% Film
R39	3145	2.2K 1/2 Watt 5% Film
R42	3139	75K 1/4 Watt 5% Film
R45, R51, R174	2627	1K 1/4 Watt 10%
R46	2876	1.5K 1/4 Watt 10%
R49	1064	15K 1/2 Watt 10%
R52, R53	1011	47 ohm 1/4 Watt 10%
R92	3152	2.2K 1/4 Watt 5% Film
R93, R95	2887	1.8M 1/4 Watt 10%
R94	2350	100 ohm 1/2 Watt 5% Film
R96	3052	22M 1/2 Watt 10%
R97	2879	39K 1/4 Watt 10%
R98, R116, R124, R159	2628	2.2K 1/4 Watt 10%
R99, R103, R106, R109, R117, R133, R136, R139, R151, R160	1042	5.6K 1/2 Watt 10%
R101, R111, R115, R126	3134	18K 1/2 Watt 5% Film
R102	3135	22K 1/2 Watt 5% Film
R104	3132	12K 1/2 Watt 5% Film
R105	3137	33K 1/2 Watt 5% Film
R107	3129	4.3K 1/2 Watt 5% Film
R108	3138	91K 1/2 Watt 5% Film
R110, R152	3109	604 ohm 1/2 Watt 1% Film
R113, R119, R121, R123, R148	2878	12K 1/4 Watt 10%
R118	2877	8.2K 1/4 Watt 10%
R120	1681	5K Vertical PC Pot
R125	1035	10K 1/2 Watt 10%
R127	3093	10K Cermet Pot
R129, R130, R173, R176, R177, R178, R180, R181	2631	10K 1/4 Watt 10%
R131, R132	3136	24K 1/2 Watt 5% Film
R134, R135	3131	11K 1/2 Watt 5% Film
R137, R138	3130	6.8K 1/2 Watt 5% Film
R149, R156	3194	1K 1/2 Watt 1% Film
R150	3196	330K 1/2 Watt 1% Film
R154	2885	270K 1/4 Watt 10%
R155	3091	1K Cermet Pot
TRANSISTORS		
Q1, Q2, Q4, Q5, Q7, Q8, Q10, Q11, Q17, Q18, Q20, Q25, Q27, Q29, Q31, Q35, Q39, Q43, Q45, Q47, Q51	2961	2N3859 A Selected
Q3, Q9, Q15, Q16, Q19, Q21, Q24, Q26, Q28, Q30, Q34, Q38, Q42, Q44, Q46, Q50, Q54	3786	PN4250 A Selected
Q6, Q12	3053	2N5459 F.E.T.
Q13, Q14, Q22, Q23, Q32, Q33, Q36, Q37, Q48, Q49, Q52, Q53	2962	TZ-81 Selected

MI 241A

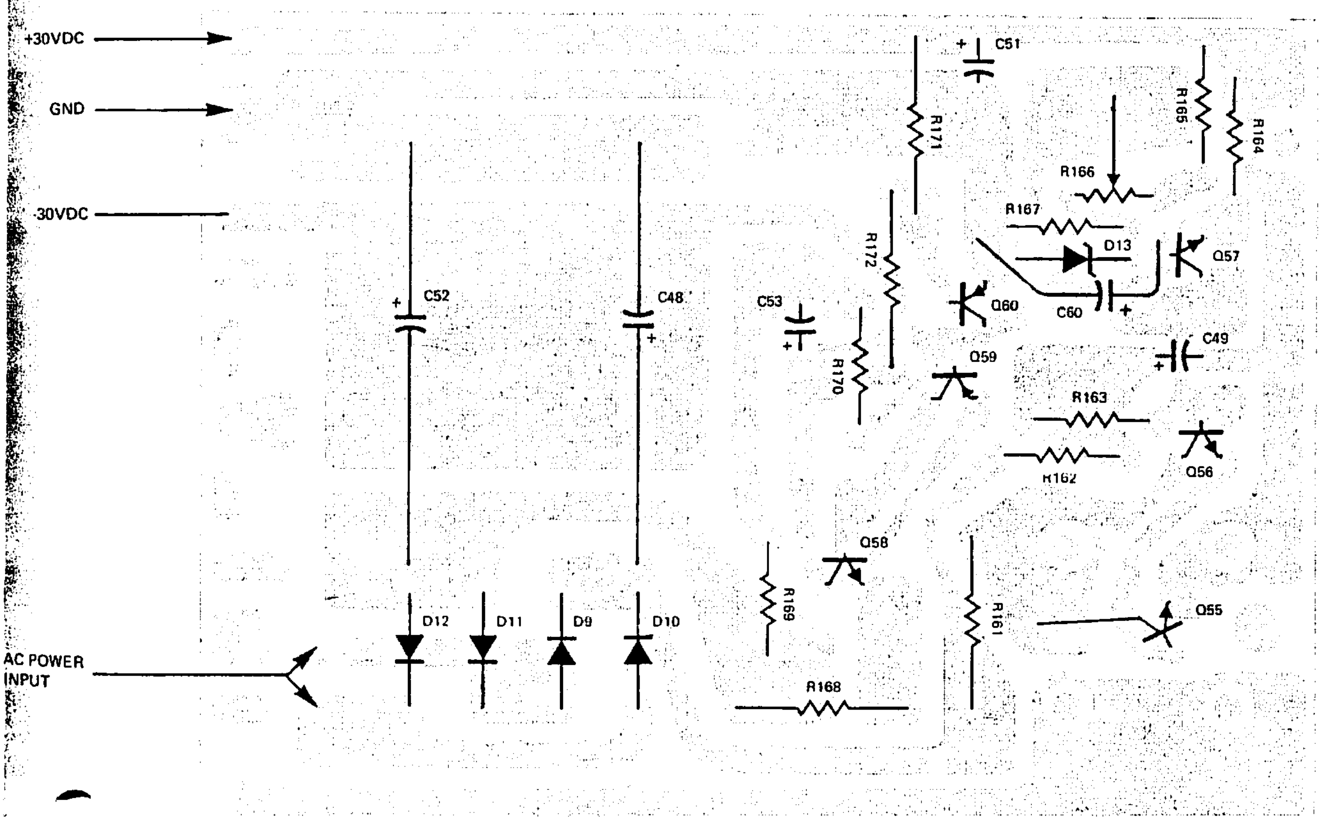
IMA MAIN BOARD
(Component Side)



POWER SUPPLY PARTS LIST and ASSEMBLY

Schematic Ref.	CROWN Part No.	Description
CAPACITORS		
C48, C52	1085A	500/55
C49, C53	2869	100/12 (vert.)
C51	2870	25/35 (vert.)
C60	1678A	5/70
DIODES		
D9, D10, D11 D12	2851	IN4003 recti.
D13	3277	IN968B
RESISTORS		
R161	2355	5.6 ohm 1 W.
R162, R169	1639	6.8K ½ W.
R163, R170	1640	4.7K ½ W.
R164	1051	3.3K ½ W.
R165	1076	1.5K ½ W.
R166	2067	2K pot. (horz.)
R167	1067	2.7K ½ W.

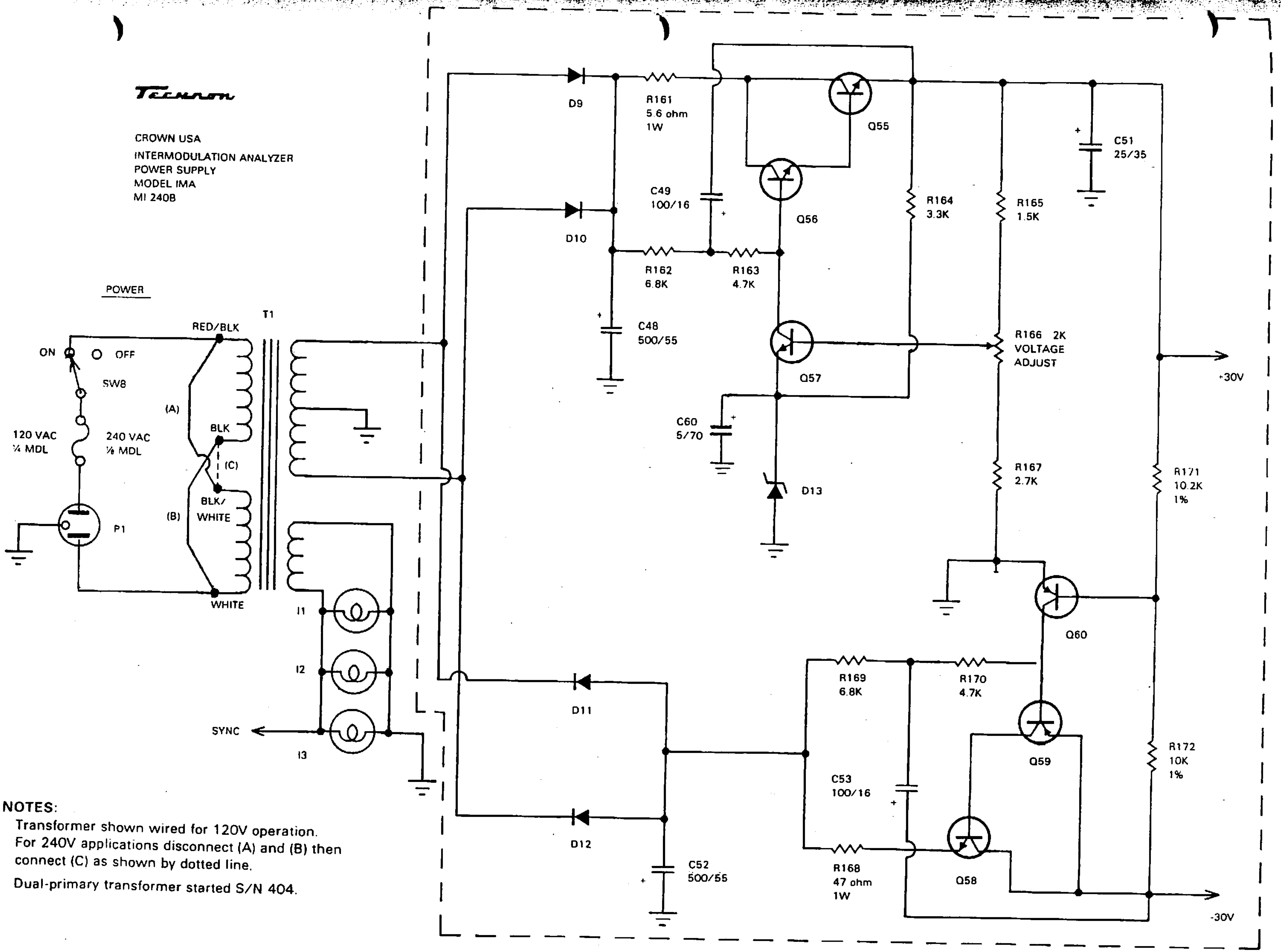
Schematic Ref.	CROWN Part No.	Description
RESISTORS (Cont.)		
R168	1073	47 ohm 1 W.
R171	2344	10.2K ½ W. 1%
R172	2343	10K ½ W. 1%
TRANSISTORS		
Q55	2976	2N3054
Q56, Q57	2961	2N3859A (Sprague)
Q58	2975	2N2102
Q59, Q60	3786	PN4250 A Selected
MISCELLANEOUS		
	1250	Transistor pads
	3175	2225B therma sink
	1824	#4 lockwasher
	1827	4/40 x ¼ Rd. hd. screw
	1938	4/40 hex nut
	2635	NF213 heat sink
	7857	Power supply board



POWER SUPPLY BOARD ASSEMBLY
MI-242
(Component Side)

Techron

CROWN USA
INTERMODULATION ANALYZER
POWER SUPPLY
MODEL IMA
MI 240B



NOTES:
Transformer shown wired for 120V operation.
For 240V applications disconnect (A) and (B) then connect (C) as shown by dotted line.
Dual-primary transformer started S/N 404.