INVESTIGATION OF THE STABILITY FACTORS
OF THE TPX 1069 THICK FILM AMPLIFIER

BY

JOHN J. VIC
HEARING AID DEVELOPMENT

SYDNEY, AUSTRALIA
Contents

1. FFA Test Jig

2. Measurements of
   a) Track Resistance
   b) Current Importance of FFA vs. Frequency
   c) Genre
   d) Noise output vs. supply and load conditions

3. Modelling the AC Performance
   a) Simple Model
   b) Computer Model
   c) Discussion of Results

4. Proposed Solution to the HF Instability

5. Low Frequency Instability

6. Conclusion

Appendices
Abstract

An investigation was initiated by the H.A.E. section on an apparent H.F. instability of the TFX 1059, which also give rise to increased quiescent current being drawn. However, as the examination proceeded, several extra problems were uncovered, the more serious of these being a "flicker" distortion evident at low frequency and almost full drive of the amplifier. Associated with this problem was quite severe 2nd harmonic distortion at the output, seemingly indicating lack of current drive to charge the coupling capacitor, and finally, current spike superimposed on the supply line with an apparent "resonance" type maximum at 2kHz when supplied from a zero dm voltage supply.
1. **The Thick FIlm Amplifier Jig**

As the amplifier chip has a high gain-bandwidth product \( (7 \times 10^6) \), it is critical that the surrounding circuitry's performance is carefully defined. For example, a track resistance of about 100Ω produces a signal level comparable to that of the input level if it is driven from the output of the amplifier.

There were 3 jigs made, with only the final one being successful in obtaining at least a stable output by using XP techniques to reduce the effects of mutual coupling between various tracks. Since I was investigating (at least initially) a high frequency instability that manifest itself as a "cook's comb" superimposed on top of the basic sine wave, there was good justification for using XP techniques. The circuit is etched into a double sided board, with the unetched side acting as a ground plane. Coupling between input and output circuitry is minimized by use of earth leads separating the two, and supply resistance is minimized by the use of heavy earth and positive supply rails, as well as keeping track lengths to a minimum.

Despite all the precautions, some TPA displayed the H.F. instability at maximum output with the tone control set to flat.

2. **Measurements**

   a) **Track Resistance**

   The track resistance on the thick film and the p.c. board were measured using a HP 4192A digital voltmeter set up on the 4 wire resistance testing mode.

![Circuit Diagram]

\[ R = \text{contact resistance} - \text{wire resistance} \]

\[ R_t = \text{track resistance} \]

\[ Z = \text{load impedance of amplifier} \]
Most critical are the values of $R$ (contact resistance + wire resistance) of pins 7, 2, 1, and 4 as in other parts, they are swamped by the external resistors.

Measurements made on the value of $R$ indicated a typical value of a 10m ohms (5m ohms minimum and 20m ohms maximum).

$R_m$ was the track resistance of the longest path track and was found to lie in the range 20-27m ohms.

Measurements made on the resistances associated with the tracks internal to the TPA showed that they came within specifications of the conductive paints used.

b) Output Impedance

In this I used a Standard test method - no load voltage vs on load voltage. The load in this case was series combination of the coupling capacitor and a resistor (15kΩ). The graph attached shows a negative resistance anomaly at very low frequencies despite the effect of taking several readings and obtaining average values. This seemed to point to possible instability problem, which, as will be shown later in this report, was verified.

The output resistance varies from 2.0ohms minimum to 1.2ohms maximum at 5kHz, with typical value being about 1ohm.

Measurements were carried using a Fluke 8060A DMM with an input impedance of 1MΩ, and frequency response from 40Hz to 20kHz at the -5dB points.

Calculations were done on a Texas Instruments TI58 using the following programme:

```
lbl A : STD Ind 0 Op20 INY SBH
lbl B : STD Ind 6 Op67 INY SRR
lbl C : STD 27 (SBH SUM X (RCL IND 0 + RCL Ins : -1))
        Op 20 Op 21 INYSR
lbl E : 2 STD 0
        15 STD 0
lbl SUM: (15.1X X (X X RCL 27 X 15 RCL 6) ^2 1/x) INVSR
```

Registers : Contents
0, 1   Counters
2-12   $V_{OL}$ (open circuit voltages) vector
15-26  $\mathbf{V_{LOAD}}$ vector
27     Frequency, Hz

Description of the Programme

A, B - used to store $(V_{OL})$, $(V_{LOAD})$ arrays.
C - Input a frequency, press C, result is the output impedance (magnitude) at that frequency.
E - is used to initialize counter registers.
sum - calculates the magnitude of load impedance at particular frequency.
c) **Gain Check**

The AWM 1460 chip uses a series of transconductance amplifier stages that drive the final class B output stage.

The following measurements were taken, and are tabulated on graphs 2, 3, and 4.

(i) gain from input of I.C. to volume control pin
(ii) gain from volume control to power output
(iii) overall gain from input to output.

It must be pointed out that the amplifiers checked were stabilised for r.f. instability by introducing a h.f. rolloff at the input stage of the thin film amplifier.

The overall gain was found to be within specification if allowance for the h.f. rolloff is considered.

The gain from input to volume control and the gain from volume control pin to power output also performed according to the simple model of the transconductance stages given by AWA microelectronics division.

d) **Noise performance versus supply voltage and load variations**

Again, using the basic test jig, and a Rukus 8000A 3MM, the noise performance was readily measured. The input was terminated by a 3K3 resistor, and output loads were 15.40hm, 450hms (purely resistive) and a Tibbetts 8212X24 earphone. The noise voltage was measured across these using the 8000A 3MM.

It was found that minimum noise occurs when the supply voltage is between 2 and 2.5V, and the 15 resistor at the output. This is hardly surprising as the noise voltage produced across a low valued resistor will be less assuming constant noise power output from the amplifier.

Taking the supply voltage below 1.7V resulted in oscillation at approximately 5kHz, no matter what load is applied. On the other extreme, when the supply voltage exceeds 4.2V, breakdown occurs within the semiconductor material in the I.C. and heavy currents are drawn (> 20mA).

Since the test equipment was different to that used by AWM, corrections had to be allowed for comparing the two noise figures.

Define:

- source resistance $R_s = 3.3k$
- signal gain (voltage) $G_s = 4000$
- bandwidth of measurement $B = 1955$ Hz
- input voltage $V_i$
- output voltage $V_o$
- load resistance $R_L$
- power gain of device $A_p$
- noise power at input $P_{n1}$
- noise power at output $P_{n0}$
- noise voltage at output $V_{n0}$
Stage I Gain

Measured between test filaments 60 x 50

Gain, \( \frac{V_i}{V_o} \)

Note: Input resistance has been changed - see text
Then it follows that:

\[
\text{Signal power in} = P_{\text{in}} = \frac{V_0^2}{R_s}
\]

\[
\text{Signal power out} = P_{\text{out}} = \frac{V_o^2}{R_o}
\]

\[
\text{Power gain} = \frac{P_o}{P_{\text{in}}} = \left(\frac{V_o}{V_i}\right)^2 \times \frac{R_o}{R_s}
\]

Noise input power due to resistor \( P_{\text{in}} = 4kT\beta \), and Noise output power is

\[
P_{\text{out}} = \frac{V_{\text{no}}^2}{R_L}
\]

Noise factor, \( NF = 10 \log_{10} \left( \frac{P_{\text{out}}}{P_{\text{in}}} \right) \)

i.e.,

\[
NF = 10 \log_{10} \left( \frac{4kT\beta C_V V_{\text{no}}}{V_o^2} \right)
\]

Typical noise voltage out \( V_{\text{no}} = 5\text{mV rms (measured)} \) and so

\[
NF = 12\text{dB} \quad \text{for } B = 19955 \quad \#
\]

Now, since

\[
NF_2 = 10 \log_{10} \left( \frac{B_1}{B_2} \right) \quad \text{Re} \left( \frac{1}{\text{Re} + \text{Im}} \right) \quad \left[ 10 \log_{10} \left( \frac{NF_2}{10} \right) \right]
\]

and

\[
B_1 = 19955 \quad B_2 = 707 \quad R_{s2} = 51 \quad \text{NF}_2 = 6\text{dB}
\]

then

\[
NF_2 = 18.74\text{dB} \quad \text{corrected for difference in bandwidth and source resistance. This shows that the measured noise figure is within specification.}
\]

3. Modelling the AC performance of the TTA.

Modelling was done in two separate stages, a simple minded approach and a more detailed analysis using the WANG computer to perform a node analysis on a model that included feedback loops.

a) Simple minded approach.

By using the block diagram given by AWN alone, and neglecting the effects of internal feedback paths within these blocks, the following simple equivalent circuit is obtained for mid to HF analysis.
Transfer functions are:

\[ T_1(s) = \frac{V_2}{V_8} = \frac{R_1}{SC \cdot R_L s + (R_2 + R_3)} \]

\[ T_2(s) = \frac{R_2}{R_c + R_v + R_L} \cdot \frac{SC \cdot R_v + 1}{SC \cdot RV / (R_2 + R_3) + 1} \]

\[ T_3(s) = \frac{(SC \cdot R_P (R_H + 1) R_v)}{SC \cdot R_L (R_H + R_L) + 1} \]

These are graphed on the following Bode plots.

More complete analysis (see appendix A4) was then done using the Wang. Included in this model were expressions yielding both DC and AC feedback loops around the complete amplifier. Parameters were adjusted until they had an overall gain of 60dB (for convenience), and the amplifier analyzed at both LP and HF ends. Since the computer analysis was done on closed loop response, and no peaking of amplitude was evident anywhere that would indicate potential instabilities, the amplifier model was judged not accurate enough.

The reasons for this inaccuracy are:

1) The oversimplifications made on the circuit due to lack of space on the machine. If an analysis programme such as SPICE was used, then the entire circuit could be modelled down to device level, thereby enabling one to see exactly what was happening.

2) The bias circuitry was totally ignored in the model.

3) The supply resistance was left out.

The approach therefore was abandoned, and bench measurements made.

4. A Proposed solution to the HF instability

a) ANW, on request from my section, made up some canned versions of the ANW 1460 using a ten pin TO-5 package. An extra lead was bonded onto the collector of the predriver transistor Q29 that would enable us to put in a dominant pole on the output stage and roll off much before 80kHz break point of the power stage. We aimed for a corner frequency of 10kHz. However, it was found that introducing my type of capacitor lead to the amplifier not operating at all!

A large capacitor (1uF chip type) was then introduced by accident onto the bias bypass point, and, sure enough, this killed the HF oscillations; it also increased the noise output from the amplifier, going up from 5mV to 12.6mV rms.

b) A compromise had to be made on the stability/noise tradeoff. Several types of chip capacitors were used in the place of the 1uF, and the following results were obtained.
Gain Plot of Transfer Function

\[ A = \frac{-R_1}{R_1 + R_2} \cdot \frac{1}{sC_1(R_2 + sC_2) + 1} + \frac{R_3}{R_1 + R_2} \cdot \frac{1}{sC_3(R_2 + sC_4) + 1} \]

\[ + \frac{5C_5 + s}{5C_5(R_2 + sC_6) + 1} \cdot R_6 \]

Gain, dB

-60 \( \text{dB} \)
-30
-20
-10
-0
0
10
20
30
40
50
60

\[ R_1 = 10k \]
\[ R_2 = 8.5k \]
\[ C_2 = 10\mu F \]
\[ R_3 = 20k \]
\[ R_4 = 1\mu F \]
\[ R_5 = 100\mu F \]
\[ C_5 = 1.5\mu F \]
\[ C_6 = 0.47\mu F \]
<table>
<thead>
<tr>
<th>Capacitor Value (microfarads)</th>
<th>Noise out (Bandwidth = 20kHz, voltage in RMS millivolts)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.07</td>
<td>9.3</td>
</tr>
<tr>
<td>0.085</td>
<td>6.5</td>
</tr>
<tr>
<td>0.066</td>
<td>6.0</td>
</tr>
<tr>
<td>0.0418</td>
<td>5.9</td>
</tr>
<tr>
<td>0.0232</td>
<td>6.4</td>
</tr>
<tr>
<td>0.0170</td>
<td>5.9</td>
</tr>
<tr>
<td>less than 0.015</td>
<td>oscillations - not noise</td>
</tr>
</tbody>
</table>

It was decided to use a minimum value of 0.042 as the bias bypass capacitor in all future aids.

5. LF Instability

The LF instability detected (at near clipping of the device) is quite closely tied in with the d.c. loop and value of battery resistance used. If a zero-ohm (true voltage) supply is used, this effect is minimal, with amplifiers showing very few impulsive bursts. However, increasing the resistance measured to simulate the typical battery values (6 ohms at d.c.) results in the impulsive bursts occurring at approx. 10-20Hz into a load of 300ohm (purely resistive).

By varying the d.c. battery bypass capacitor from 55 to 47 to 100uF resulted in a change in the frequency of the tone bursts. For example, using 47uF bypass and a 900ohms load forces the impulsive burst frequency (IBF) to be 166Hz, while using a 100uF bypass and 800ohms load forces the IBF = 56Hz.

Using a load of 300ohms and 47uF bypass results in IBF = 10-20Hz varying from 4.0 to 4.0.

Unfortunately, this particular problem is the most serious, as no easy solutions can be found. There is a need to bypass the supply with very high value, low e.e.r. electrolytic cap, which there is no room for.

Another problem that is evident is the lack of sufficient drive capability at the low frequency and after the breakpoint of the load and output capacitor. The distortion here is quite bad, with output waveforms (voltage) across the loading looking like half-wave rectified sine waves. This particular problem is evident in the TPA that display the "Flicker" distortion the most, whereas ones not displaying it all, also don't have so much LF distortion. This one again points to the D.C. loop, with which there is very little control.

In order to ensure the film amps used in the production section were quite good (i.e. not displaying the impulsive burst instability) a test set up was made up for the Quality Assurance Department. It is described in the note "Flicker" Distortion Testing, but a brief description is warranted here.
Assuming an input frequency of 250Hz, and input level to the TPA such that it is below that of level of clipping, then the flicker component comes in, it produces a very wide bandwidth of frequencies being largely impulsive. Now, if the resulting signal is bandpass filtered about 1.09kHz, a tone burst of 1.09kHz results.

This enables the operator to pick up any "flicker" distortion quite easily, as the good amplifiers have no audible output. Once in clipping, naturally the 1.09kHz tone is heard quite clearly.

Yet another interesting problem was noticed when a zero ohm supply was used. Now, although the flicker is non-existent to a large extent, a very large current spike occurs in the supply leads when operating at about 2kHz. It seems to be a resonance effect of some type - it is present right down to 200Hz although much smaller value. This problem was considered to be largely insignificant as the presence of supply resistance causes the current spike to be lost altogether.
Conclusion

Although the AWM1460 chip is capable of delivering very clean, high power outputs, the neglect (or underestimation) of the battery impedance lead to a rejection rate of 90% due to the "flicker" distortion. The evaluation of battery impedance is a very controversial one, and will be subject to investigation by the H.A.B. section in the future. Problems also could arise from the present lack of dynamic testing of devices - the "flicker" distortion is only in evidence when near full power output (wat clipping) and this will also be reduced in magnitude hopefully by a series of dynamic tests that are also being proposed by the H.A.B. section.
Appendices

A1: ANM 1460 Hearing Aid Amplifier Circuit
A2: TFA 1069 Hybrid Circuit
A3: Battery Resistance of ES76 Sine Onde Cell
A4: Wang Computer Analysis of TFA 1069 Model
Average Battery Resistance of E576E
Silicon Oxide cell

Load conditions: 15.9 mA d.c.
Number tested: 6, random selection from 8 batches
Closed Loop H.F. Response