"The Influence of Transistor and Circuit Design Philosophy on the Performance of CATV Amplifiers"

G.G. Luettgenau &
P. Rebeles

TRW Semiconductors
14520 Aviation Blvd.,
Lawndale, California 90260

Abstract

This paper discusses the characteristics of CATV transistors and circuits with respect to intermodulation. Individual distortion mechanisms are located and described by means of a precision computer program. Practical measurement techniques are evaluated.

Introduction

The major performance - limiting factors in CATV systems are noise and distortion, resulting mainly from the nature and imperfections of the bipolar transistor.

In the early days of CATV extensive testing and screening of transistor candidates by the manufacturer and user was the only way to determine the suitability of a device for this demanding application. Little was known about the mechanisms contributing to undesirable traits, let alone about their control and possible elimination.

This situation has changed. The sales volume for CATV-transistors has reached a level high enough to allow and justify the expenditure of sizeable R&D funds in the areas of device study and improvement. The development of new mathematical tools and the availability of powerful computers have resulted in a much better understanding of noise and distortion mechanisms. New manufacturing techniques like ion implantation, diffused ballast resistors, gold-metallization, etc., are available to implement the new-found knowledge. Therefore, the much improved characteristics of todays CATV transistors are the result of specific, targeted design and tightly controlled manufacturing.

Analytical Methods

When trying to improve a transistor, one will soon realize that there are certain objectionable properties which cannot be eliminated because they are founded in the very mechanisms which make a transistor what it is, and that there are others which can be manipulated. For instance, the exponential V-I characteristic of the base-emitter diode and the random nature of the processes of diffusion and re-combination are unalterable. On the other hand, one can do something about base-widening and squeezing (Kirk and Early Effects), one can control avalanche phenomena, parasitic resistances and non-linear capacitances.

The first step in an effort to understand the complicated situation calls for a mathematical analysis. In recent years several methods have been proposed and used with varying degrees of success. Most often, non-linear mechanisms involving reactive elements (having memory) are represented by means of a Volterra series.[1)] The method is applicable to the most sophisticated transistor models. It yields accurate results as long as the device is operated at low signal levels. Another method, thought to be more efficient by some,[2)] consists of perturbing a model of any degree of complexity to find an equivalent circuit. The linear portions of the device may be represented by the familiar PI-circuit. Non-linear effects are described by the addition of controlled current sources shunting input and output terminals of the PI-circuit.

Yet another way is to refrain from the use of series-representations or small signal approximations completely and to perform an analysis similar to that used to investigate the transient response of transistors. The efficiency of modern computers permits the use of extremely small integration times at a reasonable cost, resulting in a high degree of accuracy. The advantages of this approach are:

1. There is no "small-signal" limitation.
2. Several input signals, e.g., a two-or three-tone test signal, may be applied to the transistor model simultaneously. An automatic Fourier analysis performed on the output will yield detailed information in the frequency domain.

A program of this kind, which may be said to operate in "real-time" was written. The input Questionnaire is shown in the appendix. One may specify all pertinent transistor characteristics, bias points, signal level, as well as circuit parameters, including negative feed-back. All frequencies must be multiples of 10 MHz. Up to five inputs may be used. The program will automatically calculate all harmonic frequencies, 2nd order beats, intermodulation - and triple-beats. Desired and spurious signals listed in dB-magnitude and phase are shown in the output table in the Appendix.

The resolution of the output wave-form is good enough to show spurious responses in the microvolt range. This means that the "computer-noise-level" is at about -60 dBmV.

The program may also be used to calculate the performance of amplifier chains. This works as follows: The output wave-shape of the first stage is Fourier-analyzed and stored along with the amplifier gain at each frequency of interest. Each frequency component is then divided by the respective amplifier gain. After that the individual components are combined to form the time-varying input signal for the next amplifier stage. The cascading feature allows the study of distortion build-up.

Study Results

The greatest benefit derived so far has been insight into the behavior of individual distortion mechanisms. Table 1 shows 2nd and 3rd order distortion for a transistor having no distortion mechanisms other than the non-linearity of the base-emitter diode. Notice the 3rd order null and phase-reversal at

$$Rg = 1/2\ R_{in} = \frac{26mV}{I_e}(1 + \beta) = 13\ \Omega$$

Table 2 shows the influence of the avalanche voltage, which determines the output characteristics of the transistor. Observing the phase of the distortion products, one sees at once the possibility of caneling one mechanism by another through the proper choice of circuit impedances and device characteristics.

Table 1. Distortion vs. Source Resistance

Rg	2nd Order		Triple-Beat	
Ω	dB	ϕ	dB	ϕ
6.5	-19.85	180°	-45.56	180°
13.0	-21.43	180°	- ∞	-
26.0	-23.89	180°	-48.1	0°
75.0	-29.53	180°	-45.7	0°
1kΩ	-49.44	180°	-61.3	0°

V_{out} = 3 x 120 dBμV

R_c = 75Ω

I_e = 75mA

Common - Emitter

$1 + \beta = 75$

Table 2. Distortion vs. Avalanche Voltage

V_A	2nd Order		Triple-Beat	
Volts	dB	ϕ	dB	ϕ
50V	-42.32	0°	-57.37	180°
75V	-52.56	0°	-68.67	180°
100V	-62.00	0°	-76.44	180°

V_{out} = 3 x 120 dBμV

R_c = 75Ω

I_e = 75 mA

V_{cc} = 24 Volts

V_c = 18.375 Volts

Common - Emitter

Usually, it is not possible to minimize both 2nd and 3rd order distortion simultaneously when only two mechanisms are at play. There is a third major source of distortion, which is the result of the change of transistor current gain as a function of I_C. This dependency can be described by:

$$\beta = \frac{\beta\ peak}{1 + A \cdot \ln^2 \frac{Ic}{I_{c\ peak}}}$$

where A is a shape factor, typical for a specific transistor type. At a given value of operating collector current, one may readily determine the percent-change of β per decade I_C. Shown in Tables 3 and 4 are distortions as a function of negative and positive β slopes. The reader is invited to attempt to balance the various undesired outputs and thus take a first step towards becoming a CATV-transistor designer.

Table 3. Distortion vs. Negative β - Slope

Slope	2nd Order		Triple Beat	
	dB	ϕ	dB	ϕ
-6.7%	-41	0°	-60.2	180°
-3.3%	-46.3	0°	-66.3	180°
-1.33%	-53.8	0°	-74.4	180°

V_{out} = 120 dBμV

I_c = 75 mA

β = 75

Table 4. Distortion vs. Positive β - Slope

Slope	2nd Order		Triple Beat	
	dB	ϕ	dB	ϕ
+6.7%	-46.34	180°	-59.5	0°
+3.3%	-53.95	180°	-65.8	0°
+1.33%	-6115	180°	-74.0	0°

V_{out} = 120 dBμV

I_c = 75 mA

β = 75

Good Nulls - Bad Nulls

Not long ago, a major transistor manufacturer claimed for his product, that all non-linearities, except the one of the E-B diode, had been suppressed to insignificant levels. If this was a true statement (which it was not) the transistor would not be optimum. In order to achieve the degree of linearity compatible with CATV System design goals, compensation must be used. Once this philosophy is adopted, of course, there is always the possibility of obtaining perfect cancellations or nulls at specific operating conditions. For a long time devices having nulls were eyed with great suspicion and often rejected as potentially ill-behaved.

Obviously, since the magnitude of the spurious outputs due to the various distortion mechanism are bias voltage or current dependent, circuits with insufficient stabilization of the operating point, may show a shift of null-locations. Fortunately, understanding the cause points to the remedy. Much more dangerous however, are situations in which the location of the null is signal level dependent. Such nulls are typically very sharp, which indicates precarious balancing between strong distortions. In these cases higher order products are no longer negligible, resulting in a deviation from the "2 for 1"-law. Mathematical analysis entailing series-representations in which terms above the third are dropped, are not suitable to investigate the amplitude dependence of nulls. No such limitations exist for the large-signal computer program described in this paper.

Distortion at high frequencies

In addition to the mechanisms described so far, there are two more sources of distortion: Emitter and Collector Junction-capacitance. As the frequency of operation is raised into the VHF range, these mechanisms dominate, especially with respect to second order performance. In order to display its detrimental effect, the emitter capacitance need not be non-linear at all. It acts by shunting input current past the base-emitter junction, thereby modifying the instantaneous current gain. The collector capacitance, although a negative-feedback element, is non-linear and therefore at once suspect.

In order to determine the magnitudes of distortion, a model which had no deficiencies except some emitter capacitance, was operated under the following conditions:

I_e = 100 mA

R_ℓ = 75Ω

V_{out} = 120 dBμV

f = 10 MHz

C_e = 39 pf

The amplitudes of the 2nd and 3rd harmonics were:

2nd = -45.18 dB

3rd = -67.20 dB

Subsequently, C_e was removed and some collector capacitance installed instead. Increasing the value of C_c, a point was found at which

C_c = 0.283 pf

C_e = 0 pf

2nd Harm. = -45.67 dB

3rd Harm. = -71.10 dB

Adding the emitter capacitance, the situation was:

C_c = 0.283 pf

C_e = 39 pf

2nd Harm. = -85.18 dB

3rd. Harm. = -63.54 dB

The second harmonic distortion was nearly cancelled out, while the 3rd harmonic components added up on a voltage basis. The following additional observations were made:

*Emitter Distortion

Change	Result
$C_e \rightarrow C_e/2$	6 dB better
$I_e \rightarrow 2 \cdot I_e$	12 dB better
$R_\ell \rightarrow 2 \cdot R_\ell$	6 dB better

*Collector Distortion

Change	Result
$I_e \rightarrow 2 \cdot I_e$	no change
$C_c \rightarrow C_c/2$	6 dB better
$R_\ell \rightarrow 2 \cdot R_\ell$	6 dB worse

Feed-back

All examples mentioned so far, pertained to basic amplifiers without negative feed-back. Practical circuits employ heavy shunt and series feed-back. Through most of the operating frequency range the transistor gain decreases 6dB/Octave. Feed-back results in gain-linearization Spurious signals are reduced by about the same amount as the gain is reduced by means of negative feed-back. The net effect is an increase in distortion by about 6 dB for every octave-increase in signal frequencies. A new generation of CATV-transistors with F_t values of 5 GHz is now available. These devices allow the application of stronger and more uniform feed-back throughout the VHF range resulting in much improved performance at the high frequency end. Negative feed-back is not always without problems. In one experiment, a transistor operating at a point of perfect 3rd order cancellation, was subjected to negative feed-back. While the effect on second order outputs was beneficial, as expected, there was now an output on frequencies normally occupied by 3rd order products. The explanation lies in the fact that feed-back allows 2nd order products to get back to the amplifier input and mix with the original drive signal. A similar effect exists in cascades. Amplifiers which by themselves may have only 2nd order distortion, result in cascades with apparent 3rd order by-products. In this respect push-pull amplifiers show a decided advantage over single-ended circuits.

Distortion Testing

Traditionally, cross-modulation has received most attention in the U.S. Later other 3rd order tests, such as triple beat, gained importance. Often the lack of correlation between cross-modulation and triple beat was puzzling, if not frustrating. The only consolation (?) was that cross-modulation values were always better than theoretically possible. The explanation of this phenomenon is easily seen, if one examines the equation for the gain of a feed-back amplifier.

$$\text{Voltage Gain} = \frac{1}{\frac{1}{\alpha} - \beta}$$

were α is the transistor gain and β the attenuation of the feed-back network. At high frequencies, α lags about 90 degrees while β is largely resistive. Cross-modulation results from the modulation of α by the envelope of an interfering signal.

If $\left|\frac{1}{\alpha}\right| << |\beta|$ and, as stated before:

$$\text{Gain} = \frac{1}{j\frac{1}{\alpha} - \beta}$$, then a change of

α will have little effect on the magnitude of the gain, but will change the gain phase. The result is that an unmodulated carrier passed through the amplifier is not amplitude - but rather phase (cross) modulated by the interfering signal.

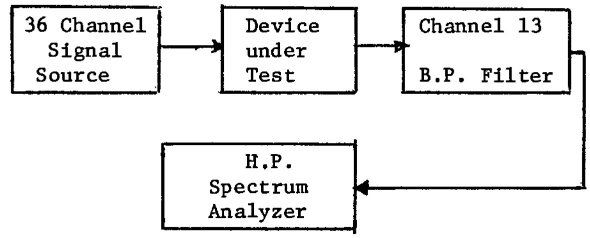
Perfect conversion exists if

$$\angle\, 1-\alpha\beta = -90^{o}$$

Many CATV amplifiers operate under conditions where, with or without intent and knowledge, considerably AM to PM conversion takes place. The authors built one amplifier in which the β-phase could be manipulated by a small trimmer capacitor. Adjusting the phase caused all cross-modulation to disappear from a previously poor off-the-air channel 13 picture.

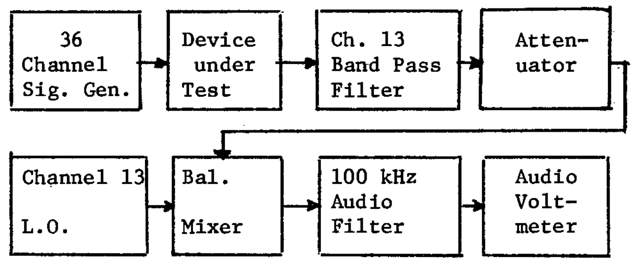
Therefore, since X-mod is heavily circuit dependent, it is not a good quality indicator for transistors. Triple-beat or similar tests in which only amplitude and not phase is of significance, are preferred. Presently en vogue is a test in which the combined spurious power caused by triple-beating 35 carriers is measured in an empty channel 13. Since some 360 components are involved, the spurious power has noise character and may be treated as such. Following are two test setups.

T-B Noise Test using Spectrum Analyzer



Sweep Width = 0
IF Bandwidth = 50 kHz
Video Bandwidth = 10 Hz
Tune-in channel 13 = 0 dB Reference
Switch off channel 13 = dB Beat Noise
Add 2.5 dB to reading to account for noise-like signal and log-amplifier characteristics.

T-B Noise Test using Balanced Mixer



Tune-channel 13, using L.O.
Use Attenuator to establish reference on Audio Voltmeter. Be careful not to overload balanced mixer.
Turn-off channel 13 signal source.
Read Beat-Noise on Voltmeter add 1.05 dB to reading, if voltmeter is average-reading.

References:

1. S. NARAYANAN "Applications of Volterra Series to Intermodulation Distortion Analysis of Transistor Feed-back Amplifiers," IEEE Transactions on Circuit Theory, CT-17, November '70, pp 518-527.

2. R.M. - M. CHEN, C.F. HEMPSTEAD, Y.L. KUO, M.L. LIOU, R.P. SNICER, and E.D. WALSH "Role of Computing and Precision Measurements," The Bell System Technical Journal, December 1974, pp 2249-2267.

APPENDIX

```
OWN OUT. FREQS. OR CALC. (O,C) ? C
0  NOS. OF IN. FREQS.  ? 3
0  INPUT FREQS.  ? 70E6,110E6,120E6
0  AMP. REF. TO VOUT (DB)  ? 0,-6,-6
1  SUPPLY VOLT.  ? 24
2  AVALAN. VOLT.  ? 1E20
3  REF. OUT VOLT.(DBUV)  ? 110
4  LOAD RESISTANCE  ? 75
5  SOURCE RESISTANCE  ? 75
6  BETA OR ALP. OF TRAN.  ? 80
7  MAXIMUM BETA OR ALPHA  ? 80.1
8  IE FOR MAX. BETA/ALP.  ? 150E-3
9  EMITTER CURRENT  ? 90E-3
10 TRANSIT FREQUENCY  ? 5E9
11 EMITTER JUNC. CAP.  ? 20E-12
12 COLLECTOR CAP. AT VCC  ? .2E-12
13 FB. CUTOFF FREQ. # 1  ? 1E20
14 FB. CUTOFF FREQ. # 2  ? 1E20
15 FEEDBACK FACTOR  ? 0
16 NOS. OF CYCLES  ? 3
17 NOS. OF STAGES  ? 2
NEED VOUT BETWEEN STAGES  ? Y
P. PULL OR S. ENDED STAGES (P/S)  ? S
```

First Stage Output

DC. CURRENT = 8.8980E-02 AMPS.

FREQ.(MHZ)	123MIT	LEVEL(DB)	PHASE(DEG)	GAIN(DB)
10	100	-42.82	167.2	35.51
20	10	-68.58	321.3	34.98
30	10	-70.43	312.7	34.21
40	100	-42.19	133.7	33.33
50	100	-42.65	126.6	32.41
60	1	-70.09	299.1	31.50
70	100000	-.01	125.7	30.62
80	1	-71.33	294.9	29.79
100	10	-83.62	293.4	28.27
110	100000	-6.01	116.2	27.58
120	100000	-6.01	114.8	26.93
130	10	-83.74	291.7	26.32
140	10000	-63.52	101.4	25.75
150	10	-82.22	285.7	25.20
160	1	-75.75	283.9	24.69

First Stage Output continued

170	10	-82.02	283.8	24.20
180	100	-60.16	125.2	23.74
190	100	-59.42	127.3	23.30
210	1000	-87.98	264.7	22.48
220	10000	-68.20	130.6	22.09
230	100	-61.53	130.8	21.72
240	10000	-66.90	130.7	21.36
250	10	-82.92	266.9	21.02
260	10	-82.64	267.1	20.69
290	10	-87.37	267.8	19.77
300	1	-80.94	268.0	19.48
310	10	-86.63	268.5	19.21
330	1000	-104.49	273.6	18.67
340	10	-91.27	269.1	18.42
350	10	-90.83	269.4	18.17
360	1000	-102.58	276.2	17.93

Second Stage Output

DC. CURRENT = 8.8980E-02 AMPS.

FREQ.(MHZ)	123MIT	LEVEL(DB)	PHASE(DEG)	GAIN(DB)
10	100	-57.67	251.5	35.51
20	10	-61.70	113.7	34.98
30	10	-63.03	100.0	34.21
40	100	-47.93	199.3	33.33
50	100	-47.17	188.3	32.41
60	1	-62.76	71.1	31.50
70	100000	-.01	251.4	30.62
80	1	-64.41	59.2	29.79
100	10	-75.45	51.2	28.27
110	100000	-6.01	232.4	27.58
120	100000	-6.01	229.5	26.93
130	10	-76.60	40.4	26.32

Second Stage Output continued

140	10000	-66.62	284.1	25.75
150	10	-75.39	31.9	25.20
160	1	-69.40	31.8	24.69
170	10	-75.35	29.8	24.20
180	100	-61.95	301.3	23.74
190	100	-61.03	302.4	23.30
210	1000	-84.45	327.8	22.48
220	10000	-68.70	301.1	22.09
230	100	-61.86	300.4	21.72
240	10000	-67.06	299.6	21.36
250	10	-80.40	325.7	21.02
260	10	-80.14	325.2	20.69
290	10	-85.58	323.4	19.77
300	1	-79.43	323.7	19.48
310	10	-85.19	324.0	19.21
330	1000	-101.22	287.9	18.67
340	10	-90.63	323.5	18.42
350	10	-90.48	324.7	18.17
360	1000	-102.31	292.9	17.93

```
WANT ANOTHER RUN  ? N
END.

   54.138 /   317.506 /        272
```