Amplitude and Phase of Intermodulation Distortion*

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It is shown that amplitude intermodulation distortion (AIMD) in a feedback amplifier is associated primarily with signal-induced changes in the low-frequency gain of the forward path, whereas phase intermodulation distortion (PIMD) is associated primarily with signal-induced changes in the gain–bandwidth product. The usual nonlinearity in transistor β therefore contributes AIMD but not PIMD. It is shown that both AIMD and PIMD involve intermodulation products at the same frequencies. For comparable nonlinearities the amplitudes of the intermodulation products are roughly equal, but their phases are different by $\pi/2$. An intermodulation specification based on the amplitude of the intermodulation products includes both AIMD and PIMD. In contrast, a specification based on the depth of an amplitude modulation neglects PIMD.

0 INTRODUCTION

Phase intermodulation distortion (PIMD) [1]–[3] is a recently proposed measure for quantifying nonlinear distortion in an amplifier or other network. PIMD is closely related to the familiar two-tone intermodulation distortion (IMD).

In the SMPTE test for IMD, the amplifier input is a combination of a large-amplitude low-frequency sine wave with a small-amplitude high-frequency sine wave. Often the large-amplitude component is 80% of the rated maximum input amplitude, and its frequency is either 50 Hz or 60 Hz. The small-amplitude component is often 20% of the rated maximum input amplitude and therefore falls within the small-signal regime of amplifier operation; its frequency is often 7 kHz. The large sine wave excites nonlinearities in the amplifier, and hence modulates the small-signal transfer function. Because the two input components are independent, the response of the amplifier to the small component (in the presence of the large component) is effectively the response of a linear time-varying network. The SMPTE intermodulation test quantifies nonlinearity by reference to the amplitude modulation of the small-amplitude high-frequency component as it appears at the amplifier output.

In the proposed PIMD test, nonlinearity is quantified by reference to the phase modulation of the small-amplitude high-frequency component as it appears at the amplifier output. In order to emphasize this distinction, the familiar SMPTE test is described throughout the rest of this paper as a test of amplitude intermodulation distortion (AIMD).

It is shown in many places [4] that the depth of amplitude modulation in an AIMD test is related to the amplitude of the modulation sidebands, that is, to the strength of the intermodulation products. What is not so often stressed is that the depth of amplitude modulation depends also on the phase of the sidebands. In fact, both AIMD and PIMD involve intermodulation products at the same frequencies, but the phase relations are such that the standard AIMD test fails completely to detect the intermodulation products associated with PIMD.

Broadly, this paper is concerned with finding the contribution of various types of nonlinearity to the output spectrum, and hence assessing the relative significance of AIMD and PIMD.

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1 AIMD AND PIMD IN IDEALIZED FEEDBACK AMPLIFIERS

Fig. 1 shows a block diagram of an idealized low-pass feedback amplifier, and Fig. 2 shows its gain asymptotes on logarithmic scales. The low-frequency gain of the forward path is G and the time constant of the dominant pole is τ_G. Real amplifiers have additional nondominant poles as suggested by the dotted line in Fig. 2. The (projected) gain–bandwidth product $1/\tau_1$ is

$$\frac{1}{\tau_1} = \frac{G}{\tau_G} \tag{1}$$

and the overall closed-loop gain $A(s)$ is

$$A(s) \equiv \frac{\text{OUTPUT}}{\text{INPUT}} = \frac{G}{1 + GB + sG\tau_1} \simeq \frac{1}{B}\left(\frac{1}{1 + s\tau_1/B}\right) . \tag{2}$$

The loop gain $A_l(s)$ is

$$A_l(s) \equiv \frac{\text{FEEDBACK}}{\text{ERROR}} = \frac{GB}{1 + sG\tau_1} . \tag{3}$$

At complex frequencies whose magnitude is less than the forward-path bandwidth, Eq. (3) can be approximated by

$$A_l(s) \underset{s\tau_G<1}{\rightarrow} GB . \tag{4a}$$

At complex frequencies whose magnitude is greater than the forward-path bandwidth,

$$A_l(s) \underset{s\tau_G>1}{\rightarrow} \frac{B}{s\tau_1} . \tag{4b}$$

The approximate form of Eq. (2) applies if the low-frequency loop gain GB is greater than about 10.

For specified demanded gain $1/B$, the small-signal performance of Fig. 1 is almost completely defined by the horizontal asymptote G and the sloping asymptote $1/\tau_1$ in Fig. 2. The nondominant poles have little effect on the response at frequencies inside the closed-loop passband, and may be neglected. Around specified values of G and τ_1 the sensitivities of closed-loop gain to changes in G and τ_1 are [5]

$$S[G]_{\tau_1} \equiv \frac{G}{A}\left(\frac{\partial A}{\partial G}\right)_{\tau_1} = \frac{1}{1 + GB + sG\tau_1} \tag{5a}$$

$$S[\tau_1]_G \equiv \frac{\tau_1}{A}\left(\frac{\partial A}{\partial \tau_1}\right)_G = -\frac{sG\tau_1}{1 + GB + sG\tau_1} . \tag{5b}$$

For small changes in G and τ_1, the resulting change in A can be found from

$$\frac{\delta A}{A} = S[G]_{\tau_1}\frac{\delta G}{G} + S[\tau_1]_G\frac{\delta\tau_1}{\tau_1} . \tag{6}$$

Note that the sensitivities are functions of complex frequency s, and should formally be written as $S[G, s]_{\tau_1}$ and $S[\tau_1, s]_G$. In the special case of a sinusoidal signal $s = j\omega$, the changes in amplitude and phase of the response follow from Eq. (6) as

$$\frac{\delta\{|A(j\omega)|\}}{|A(j\omega)|} = \text{Re}\{S[G, j\omega]_{\tau_1}\}\frac{\delta G}{G} + \text{Re}\{S[\tau_1, j\omega]_G\}\frac{\delta\tau_1}{\tau_1} \tag{7a}$$

$$\delta\{\underline{/A(j\omega)}\} = \text{Im}\{S[G, j\omega]_{\tau_1}\}\frac{\delta G}{G} + \text{Im}\{S[\tau_1, j\omega]_G\}\frac{\delta\tau_1}{\tau_1} . \tag{7b}$$

If we assume the typical constraints of large low-frequency loop gain, that is

$$GB \gg 1 , \tag{8}$$

and frequencies well inside the closed-loop passband, that is

$$\omega \ll B/\tau_1 , \tag{9}$$

Eqs. (5) and (7) reduce to

$$\frac{\delta\{|A(j\omega)|\}}{|A(j\omega)|} \simeq \left(\frac{1}{GB}\right)\frac{\delta G}{G} - \left(\frac{\omega\tau_1}{B}\right)^2\frac{\delta\tau_1}{\tau_1} \tag{10a}$$

$$\delta\{\underline{/A(j\omega)}\} \simeq -\left(\frac{1}{GB}\right)\left(\frac{\omega\tau_1}{B}\right)\frac{\delta G}{G} - \left(\frac{\omega\tau_1}{B}\right)\frac{\delta\tau_1}{\tau_1} \tag{10b}$$

In AIMD and PIMD tests, the large-amplitude low-frequency sine wave excites the various nonlinearities, and is equivalent to producing slow time-varying changes δG and $\delta\tau_1$. The resulting modulations of the amplitude and the phase of the small-signal response to the small-amplitude high-frequency sine wave constitute AIMD and PIMD, respectively. Thus AIMD and PIMD can be calculated from Eqs. (10) if ω is interpreted as the frequency of the small-amplitude sine wave.

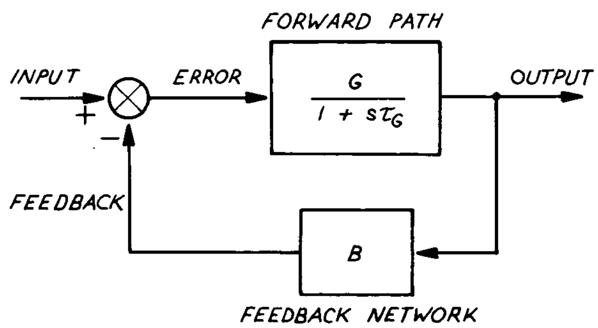

Fig. 1. Block diagram of an idealized low-pass feedback amplifier.

2 BEHAVIOR OF AIMD AND PIMD

The behavior of AIMD, calculated from Eq. (10a), is well known. We remark that, at low enough values of ω, the component of AIMD associated with nonlinearity in G must dominate over the component associated with nonlinearity in τ_1. Depending on details, this situation probably pertains in audio power amplifiers over the audible frequency band.

For comparable nonlinearities in G and τ_1 (that is, for roughly equal values of $\delta G/G$ and $\delta\tau_1/\tau_1$), PIMD calculated from Eq. (10b) is completely dominated by the nonlinearity in τ_1.

These results have an obvious physical interpretation. Referring to Fig. 2, the magnitude of the in-band closed-loop gain is determined largely by its horizontal asymptote and, for specified demanded gain $1/B$, this asymptote is determined by the forward-path low-frequency gain G. The phase of the closed-loop response at any frequency is determined by the ratio of this frequency to the frequency of the intersection of the horizontal and sloping asymptotes, that is, by the gain–bandwidth product $1/\tau_1$ if $1/B$ is specified. Thus G has a dominating influence on the magnitude of the closed-loop gain, and $1/\tau_1$ has a dominating influence on the phase of the closed-loop gain. Broadly speaking, therefore, *it is signal-induced changes in the forward-path horizontal asymptote (that is, nonlinearity in G) that dominate in the generation of AIMD, and signal-induced changes in the sloping asymptote (that is, nonlinearity in τ_1) that dominate in the generation of PIMD.*

Notice that the coefficients $1/GB$ and $\omega\tau_1/B$ in Eqs. (10) are the reciprocals of the low-frequency and high-frequency loop gains [Eqs. (4)]. For specified nonlinearities $\delta G/G$ and $\delta\tau_1/\tau_1$, both AIMD and PIMD decrease monotonically as the loop gain is increased. Any interpretation of [1]–[3] as implying that increased feedback leads to increased PIMD is plainly wrong for the realistic boundary conditions (8) and (9). The obvious way to reduce both AIMD and PIMD in an amplifier is to increase the loop gain. It is especially advantageous to increase the high-frequency loop gain by means of increasing the gain–bandwidth product $1/\tau_1$. Bode's emphasis [6] on gain–bandwidth product in low-distortion amplifiers is as valid today as it was 40 years ago.

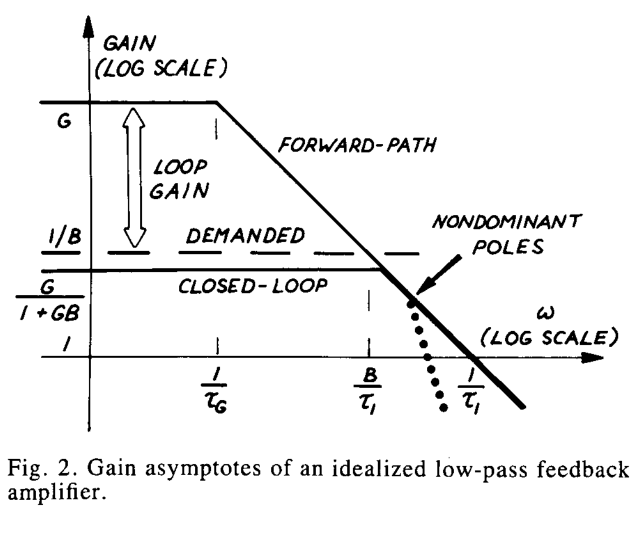

Fig. 2. Gain asymptotes of an idealized low-pass feedback amplifier.

3 RELATIONS BETWEEN AIMD, PIMD, AND TIME DELAY

A general relation [6] exists between the amplitude response and the phase response of a minimum-phase network to a sinusoidal input. For example, it is well known that if the amplitude response is changing at a rate asymptotic to $20n$ dB per decade, the phase shift is asymptotic to $n\pi/2$ radian.

Both the amplitude and the phase of the response of Fig. 1 are uniquely related to the closed-loop asymptotes in Fig. 2. These asymptotes are in turn uniquely related to the forward-path asymptotes and, as a consequence, any change in either the amplitude or the phase of the closed-loop response is uniquely related to the changes in the forward-path asymptotes. The changes in amplitude and phase of the closed-loop response of Fig. 1 are therefore related one to the other, and there must be a relation between AIMD and PIMD calculated from Eqs. (10).

There is, in fact, a general relation between AIMD and PIMD such that, given a complete specification of either one, it is possible to calculate the other. *A specification of PIMD for an amplifier therefore contains no information that is not present in a complete specification of AIMD.* (These last sentences may seem surprising in view of the italicized sentence in Section 2; note that the calculation of PIMD from AIMD requires a complete specification of the latter.)

If the sensitivity of any small-signal transfer function $A(s)$ to changes in any parameter $(\cdot)$ is written $S[\cdot, s]$, then the changes in amplitude and phase of $A(j\omega)$ in response to a small fractional change $\delta(\cdot)/(\cdot)$ are

$$\frac{\delta\{|A(j\omega)|\}}{|A(j\omega)|} = \mathrm{Re}\{S[\cdot, j\omega]\} \frac{\delta(\cdot)}{(\cdot)} \qquad (11a)$$

$$\delta\{\underline{/A(j\omega)}\} = \mathrm{Im}\{S[\cdot, j\omega]\} \frac{\delta(\cdot)}{(\cdot)} . \qquad (11b)$$

Provided $S[\cdot, s]$ is analytic in the right half of the s plane, then, via the Hilbert transformation,

$$\mathrm{Im}\{S[\cdot, j\omega]\} = \frac{2\omega}{\pi} \int_0^\infty \frac{\mathrm{Re}\{S[\cdot, jw]\}}{w^2 - \omega^2}\, dw . \qquad (12)$$

The imaginary part of any sensitivity function (hence PIMD) is completely prescribed if the real part (hence AIMD) is known. The requirement that $S[\cdot, s]$ should be analytic in the right half of the s plane is no practical restriction for a feedback amplifier: the same requirement applies if a Nyquist diagram is to predict stability correctly [6].

Cordell [7] has expressed PIMD as a change in a kind of time delay, equivalent to

$$\delta\tau_p = -\frac{1}{\omega}\left(\delta\{\underline{/A(j\omega)}\}\right) = -\left(\frac{\mathrm{Im}\{S[\cdot, j\omega]\}}{\omega}\right)\frac{\delta(\cdot)}{(\cdot)} . \quad (13)$$

By this definition, $\delta\tau_p$ is the change in phase delay. The propagation delay of a signal through an amplifier or other network is more usually associated with group delay than with phase delay. A more useful concept might therefore be the change in group delay:

$$\delta\tau_g = -\frac{d}{d\omega}\left(\delta\{\underline{/A(j\omega)}\}\right) = -\left[\frac{d}{d\omega}\left(\mathrm{Im}\{S[\cdot, j\omega]\}\right)\right]\frac{\delta(\cdot)}{(\cdot)} . \quad (14)$$

Within the range of validity of Eq. (10b), the changes in phase delay and group delay are equal (although this is not the case in general). Cordell's numerical results and his comments thereon are therefore correct, but it is suggested that considerations of group delay should provide their basis.

4 NULL PIMD

The two terms in Eq. (10b) have the same variation with frequency ω. It is therefore possible in theory to obtain a broadband null in PIMD via a cancellation.

From Eqs. (5a), (5b), and (7b) an exact relation corresponding to Eq. (10b) is

$$\delta\{\underline{/A(j\omega)}\} = -\frac{\omega G\tau_1}{(1 + GB)^2 + (\omega G\tau_1)^2} \times \left[\frac{\delta G}{G} + (1 + GB)\frac{\delta\tau_1}{\tau_1}\right] . \quad (15)$$

Hence for null PIMD the changes in G and τ_1 must satisfy

$$\frac{\delta G/G}{\delta\tau_1/\tau_1} \underset{\mathrm{PIMD}\to 0}{=} -(1 + GB) . \quad (16)$$

Like all cancellation phenomena, the null is sensitive to small departures from the ideal relations. Notice that the null condition depends strongly on the low-frequency loop gain GB.

A special case of interest is that in which B is set to zero and all feedback is removed. From Eq. (16) the null condition becomes

$$\frac{\delta G}{G} \underset{\substack{B=0\\ \mathrm{PIMD}\to 0}}{=} -\frac{\delta\tau_1}{\tau_1} . \quad (17)$$

Evidently an amplifier without feedback generates zero PIMD if the nonlinearity is such that the fractional changes in forward-path low-frequency gain and gain–bandwidth product are equal. Expressed differently, an amplifier without feedback generates zero PIMD if the forward-path pole τ_G is invariant.

Application of feedback to an amplifier that satisfies the null condition (17) results in finite closed-loop PIMD. It is apparently this special case which has led to conjecture [2] that negative feedback converts AIMD into PIMD. The reality is that any amplifier with or without feedback can have zero PIMD if Eq. (16) is satisfied, but any change in the feedback factor B always destroys an existing null. One could equally conjecture that, if an amplifier satisfies Eq. (16) for $B \neq 0$, then *removal* of feedback converts AIMD into PIMD. The special null condition at $B = 0$ is just one point in a continuum.

5 SPECTRA OF AIMD AND PIMD

When a carrier $\cos\omega_c t$ is amplitude modulated to a depth k_a by $\cos\omega_m t$, the resulting waveform is

$$f_a(t) = \cos\omega_c t + \tfrac{1}{2}k_a\cos(\omega_c + \omega_m)t + \tfrac{1}{2}k_a\cos(\omega_c - \omega_m)t . \quad (18)$$

When the carrier is phase modulated with peak excursion k_p radian, then, provided $k_p < 0.5$, the resulting waveform is

$$f_p(t) = \cos\omega_c t + \tfrac{1}{2}k_p\sin(\omega_c + \omega_m)t + \tfrac{1}{2}k_p\sin(\omega_c - \omega_m)t + \text{very small terms} . \quad (19)$$

Thus the spectra of amplitude-modulated and phase-modulated signals contain the same frequency components. The amplitudes of the sidebands depend on the depths of modulation; their phases differ by $\pi/2$.

In intermodulation tests, δG and $\delta\tau_1$ are periodic at the frequency of the large-amplitude low-frequency component of input. This frequency (or its harmonics) corresponds to ω_m if Eqs. (18) and (19) are interpreted as the output spectra in AIMD and PIMD tests. The small-amplitude high-frequency component of input corresponds, of course, to ω_c. Thus *the spectra associated with AIMD and PIMD contain components at the same frequencies but of different phase*. A matter of some importance is the relative amplitude of the sidebands associated with AIMD and PIMD.

In an AIMD test, the depth of amplitude modulation follows from Eq. (10a) as

$$k_a \to \frac{\delta\{|A(j\omega_c)|\}}{|A(j\omega_c)|} \simeq \left(\frac{1}{GB}\right)\frac{\Delta G}{G} + \left(\frac{\omega_c\tau_1}{B}\right)^2\frac{\Delta\tau_1}{\tau_1} \quad (20)$$

where ΔG and $\Delta\tau_1$ are the peak amplitudes of the time-varying components of δG and $\delta\tau_1$ at frequency ω_m. In a PIMD test, the peak phase excursion follows from Eq. (10b) as

$$k_p \to \delta\{\underline{/A(j\omega_c)}\} \simeq \frac{1}{GB}\left(\frac{\omega_c\tau_1}{B}\right)\frac{\Delta G}{G} + \left(\frac{\omega_c\tau_1}{B}\right)\frac{\Delta\tau_1}{\tau_1} . \quad (21)$$

Thus the relative amplitudes of the sidebands at

$\omega_c \pm \omega_m$ associated with AIMD and PIMD are:

$$\frac{\text{PIMD}}{\text{AIMD}} = \frac{k_p}{k_a} \simeq \frac{\frac{1}{GB}\left(\frac{\omega_c\tau_1}{B}\right)\frac{\Delta G}{G} + \left(\frac{\omega_c\tau_1}{B}\right)\frac{\Delta\tau_1}{\tau_1}}{\left(\frac{1}{GB}\right)\frac{\Delta G}{G} + \left(\frac{\omega_c\tau_1}{B}\right)^2\frac{\Delta\tau_1}{\tau_1}} \; . \tag{22}$$

Three cases of Eq. (22) are relevant. If $\Delta G/G \gg \Delta\tau_1/\tau_1$, then AIMD dominates:

$$\frac{\text{PIMD}}{\text{AIMD}} \rightarrow \frac{\omega_c\tau_1}{B} \ll 1 \; . \tag{23a}$$

If $\Delta G/G \ll \Delta\tau_1/\tau_1$, then PIMD dominates:

$$\frac{\text{PIMD}}{\text{AIMD}} \rightarrow \frac{B}{\omega_c\tau_1} \gg 1 \; . \tag{23b}$$

If $\Delta G/G \simeq \Delta\tau_1/\tau_1$, then AIMD dominates if ω_c is below the forward-path pole, and PIMD dominates if ω_c is above the forward-path pole:

$$\frac{\text{PIMD}}{\text{AIMD}} \rightarrow \omega_c G\tau_1 \; . \tag{23c}$$

6 AIMD AND PIMD IN A REAL AMPLIFIER [8]

Fig. 3 shows in outline the topology of many audio power amplifiers and monolithic operational amplifiers. For this topology the asymptotes in Fig. 2 correspond to

$$G \simeq G_{T1}\,\beta_2\,\beta_3\,R_L \tag{24}$$

$$B = \frac{R_{F1}}{R_{F1} + R_{F2}} \tag{25}$$

$$\tau_1 \simeq \frac{C}{G_{T1}} \tag{26}$$

where G_{T1} is the transfer conductance of the first stage, and β_2 and β_3 are the current amplification factors of the transistors (probably Darlingtons) in the second and third stages. The lag compensating capacitor C may include a significant component from the collector–base capacitance C_{ob} of the second-stage transistor.

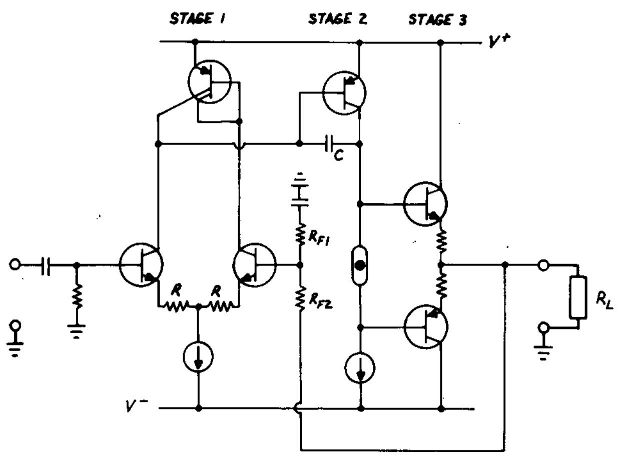

Fig. 3. Popular outline circuit topology for an audio power amplifier.

To a first order, the projected gain–bandwidth product $1/\tau_1$ is independent of β_2 and β_3. Therefore, from the commonsense reasoning in Section 2, nonlinearity in β_2 and β_3 should introduce AIMD but not PIMD. Cordell [7] has confirmed this experimentally, and [8] contains a formal derivation of the result. Signal-induced changes in C are the most likely source of changes in the gain–bandwidth product, and therefore, from Section 2, nonlinearity in second-stage C_{ob} is the most likely practical source of PIMD. Again, [8] confirms this commonsense prediction.

Changes in G_{T1} change both asymptotes and, in fact, correspond to the null defined by Eq. (17): to a first order, nonlinearity in G_{T1} produces no open-loop PIMD. Nonlinearity in G_{T1} also corresponds to the situation defined by Eq. (23c): AIMD dominates the closed-loop spectrum if ω_c is below the forward-path pole, PIMD dominates if ω_c is above the forward-path pole.

Nonlinearity in G_{T1} is most usually associated with transient IMD [5], [9]–[11]. Such nonlinearity is very unlikely to be excited in the SMPTE 2-tone IMD test, and therefore is very unlikely to contribute significantly to either AIMD or PIMD.[1]

7 CONCLUSIONS

This paper gives a commonsense justification for the fact that the only nonlinearity which is likely to contribute to PIMD in a typical amplifier (Fig. 3) is the collector–base capacitance of the second-stage transistor. Nonlinearities in transistor current amplification factors do not produce PIMD. Cherry [8] confirms these predictions formally, and also shows that the nonlinearity in transistor mutual conductance (that is, the exponential V_B–I_C characteristic) does not produce PIMD.

In an intermodulation test, the amplitude of any modulation sideband certainly affects its audibility and hence affects the subjective distortion sensation. The phase of a sideband may or may not affect its audibility (according to the listener's opinion regarding the au-

[1] R. R. Cordell has pointed out to the author that there are at least two situations in which this statement could be wrong. Both involve circuits in which the first-stage emitter degeneration resistors are so small that G_{T1} depends strongly on the first-stage emitter current.

The first case is where the first-stage tail current source is approximated by a simple resistor. The common-mode component of the input and feedback voltages then results in significant changes in operating current, and the variation of G_{T1} could be as much as ±5%. This mechanism is operative at all input signal frequencies, so the nonlinearity could certainly be excited in an SMPTE test.

The second case is where the combined gain $\beta_2\beta_3$ of the second and third stages is so small that the low-frequency signal current in the first stage is a significant proportion of the quiescent current, and the familiar tanh nonlinearity of a long-tailed pair is excited. Such an amplifier will most likely suffer from gross hard transient IMD, in addition to PIMD in an SMPTE test.

dibility of phase shifts), but the effect of a change in phase is expected to be small compared with the effect of a change in amplitude. Therefore, for equal modulation parameters k_a and k_p, AIMD and PIMD should be roughly equally audible.

There would be no need for a distinction between AIMD and PIMD in audio amplifier testing if the alternative specification [4] of SMPTE two-tone IMD were adopted as standard. Referring to the output spectrum shown in Fig. 4, define

$$\mathrm{IMD}_2 \equiv \frac{A_{1+1} + A_{1-1}}{A_0} \times 100\% \quad (27a)$$

$$\mathrm{IMD}_3 \equiv \frac{A_{1+2} + A_{1-2}}{A_0} \times 100\% \ . \quad (27b)$$

These definitions are, of course, entirely compatible with all existing standards and equipment, provided it is recognized that the majority of contemporary IMD test sets measure only AIMD. Even this is probably not a serious limitation for, as shown experimentally and theoretically [7], [8], PIMD in real amplifiers tends to be small compared with AIMD.

8 ACKNOWLEDGMENT

Dr. G. K. Cambrell, of the Department of Electrical Engineering, Monash University, showed Eq. (12) to the author.

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APPENDIX

PRELIMINARY EXPERIMENTS

AIMD and PIMD can be generated artificially, as shown in Fig. 5, and some preliminary experiments have been conducted to assess the relative audibility.

For very low modulating frequencies (1 or 2 Hz), AIMD and PIMD sound noticeably different; PIMD is difficult to describe, whereas AIMD sounds as intuition would suggest—a slowly pulsating tone. The carrier frequency and the depth of modulation seem irrelevant, although most of the experiments were conducted with carriers in the range from a few hundred hertz to a few kilohertz and with modulation parameters $k_a = k_p$ in the range of 0.1–0.3.

At higher modulating frequencies, AIMD and PIMD sound very much the same. None of the subjects could

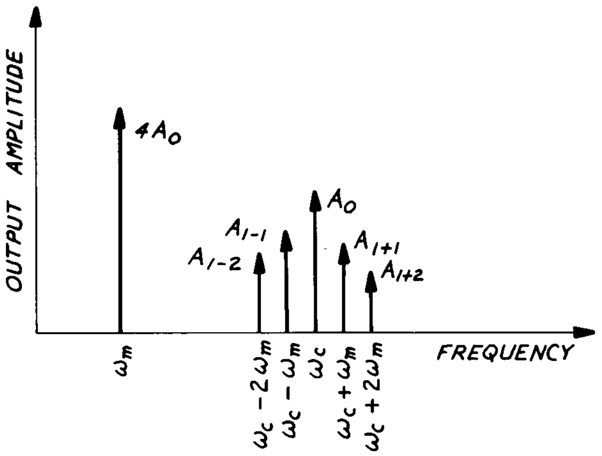

Fig. 4. Partial output spectrum for a two-tone IMD test.

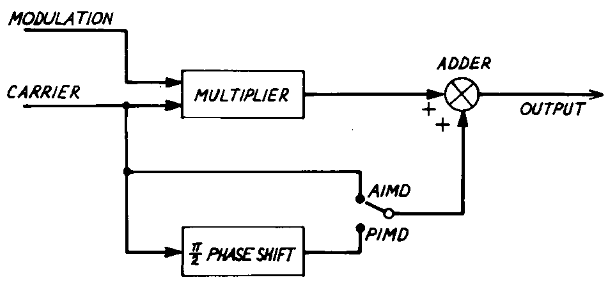

Fig. 5. Artificial generation of AIMD and PIMD.

reliably distinguish between AIMD and PIMD when the modulating frequency was either 50 or 60 Hz. AIMD and PIMD are difficult to distinguish when the modulation is 10 Hz.

Subjective testing is notoriously difficult to carry out. It should be stressed that the above results are not a definitive demonstration that AIMD and PIMD at realistic audio frequencies are indistinguishable. Rather, the experiments demonstrate that AIMD and PIMD are, in fact, ". . . roughly equally audible. . . ." The case for adopting Eqs. (27) as the definition of IMD is strengthened.

THE AUTHOR

Edward Moore Cherry was born in Melbourne, Australia, in 1936. The University of Melbourne awarded him bachelor's and master's degrees in science in 1957 and 1959, respectively, and a Ph.D. in electrical engineering in 1962. From 1963 until 1965 Cherry was a member of the technical staff of Bell Telephone Laboratories, New Jersey, and in 1966 he was a temporary research associate of the United Kingdom Atomic Energy Authority, Harwell. In 1967 he returned to Australia to take up an appointment with Monash University, where he is presently associate professor of electrical engineering.

In 1973 for his sabbatical, Cherry returned to Bell Laboratories on a Fulbright scholarship. Professor Cherry's research interest is in electronic hardware; he has published one book and over 50 papers in the field. He is a fellow of the Institution of Radio and Electronics Engineers, and has twice been awarded that Institution's annual medal for the most meritorious paper published in its Proceedings.