Guest Editorial:
'The Impossible Void'
Walt Jung

In this issue *TAA* begins what is hoped to be an on going series on distortion and/or imperfections in audio amplifiers. Fundamental to this series (as opposed to so many previous ones of the ho-hum variety) is the goal of *concrete* electrical-audible correlation. I feel (and I'm sure many of you agree) that it is time someone really sat down and gave this serious consideration, and *TAA* is the most logical frame of reference for such a potential new insight.

As an audio magazine, *TAA* is neither the glossy "never-bad review" consumer slick, nor is it the cultist "be damned how it measures" subjective only publication. Both of these two extremes are unrealistic, in different senses of the word.

On the surface, the mass circulation hi-fi publications would seem to be much better equipped technically to investigate such matters as electrical/audible correlations, yet the absence of any such work is obvious. Everyone seems to be content to continue saying, "Oh yes, there is still much that is unknown in that area."

On the other hand, the far out, so-called "high end" crowd tend to belittle measurements as virtually meaningless. Typical are such comments as, "Specifications are nearly worthless, as they do not yet relate to musical quality." *Why* do the equipments which sound musical sound that way? *That* is the burning issue.

Are specifications useful, or aren't they? It has always been my firm opinion that specs are worthwhile, as are subjective listening tests. But a yawning void divides the two, and it is here that the work is needed. Not much changes if members of each camp continue merely to throw stones at each other.

The subjective reviewers need to make more measurements, the measurement types need to do more listening. In other words, *let's get together on the common problem.*

This series will attempt the "impossible": bridging this gap. And it is really not impossible, it just takes the right attitude, insight, equipment, and a fair amount of sweat. I think the article which begins in this issue will demonstrate that the two areas can be brought together, at least on one common point.

The subject of this first installment is *slewing induced distortion* which I call SID. If that sounds remote, please note that SID includes transient inter-modulation (TIM) and many other distortion buzzwords so much in vogue these days. I believe SID is a broader and more penetrating view of the distortion phenomenon than has been presented to date, and I hope this article begins to vindicate the IC op amp in the minds of those who have heretofore been convinced of its alleged "inferiority."

However, this article has other ramifications which go far beyond the use of IC op amps. The concept of SID can, in general, be extended to include all audio amplifiers which use feedback. That takes in a lot of territory: tubes, transistors, and whatever! The material will, I think, provide ample food for thought for many if not all *TAA* readers.

Many things will doubtless evolve from this article, but one I feel should be stressed here is the importance of more complete electrical testing of audio amplifiers in these areas. The articles will demonstrate that audible defects in audio amplifiers can be tied to slew rate- -but how many amplifiers are specified for this parameter? And how many equipment reviews routinely test for it? The answers are simply too few, to say the least.

All of the above is not to say that SID is the only source of bad sound in amplifiers today. But it is a major one, and one which is far from being universally appreciated. The subject is not altogether a simple one, either, as to understand it you must be able mentally to separate behavior of a feedback amplifier under transient or HF signal conditions, for both small and large signal excursions.

However, although it may be less than crystal clear at first, we trust this first article will give you a beginning grasp of the phenomenon. SID is not at all mystical or nebulous and can be measured quite directly and repeatably. Neither does it occur exclusively during transient conditions, by any means. I hope you enjoy this and succeeding articles and derive useful information from them- -information you can relate to your own experiences.

This material first appeared within The Audio Amateur*, issue 1/1977, page 3*

Slewing Induced Distortion in Audio Amplifiers: Part I

by WALT JUNG
Contributing Editor

THIS STUDY BEGAN as a general investigation into forms of transient IM distortion in op amps in general, and in IC op amps in particular. Since so very much has been written in the past few years on transient intermodulation distortion (TIM) in audio amplifiers[1-11, 15, 17]--most particularly the power amplifier--my original object was to seek some answers to which I could attach numbers *and correlate these with what I heard.*

Many of these writings damn IC op amps as the bane of the quality sound we all seek and treasure. On the face of some of it, one should dismiss the use of op amps in an audio signal path as something abhorrent, to be avoided at all costs. And indeed, some of the correspondence I receive doesn't just suggest this, it virtually demands it. I'm sure we have all read more than one equipment review which has mentioned "IC" or "transistor" sound, the harsh, hard, gritty stuff that grates the sensitivities.

But, if we sit back and reflect on the overall recording-to-reproduction system (of which our own end is only a part), we can see some obvious inconsistencies. We all know a recorded signal goes through many, many amplifiers before it reaches our ears and most of them are beyond our control. Consider an obviously well recorded example of today's releases and I think we can all agree the sound *can* be very good. And solid state amplifiers are used almost exclusively in the recording process.

Many console manufacturers use design concepts based largely on op amps, of either IC or modular variety. Some have excellent track records, while others do not. So great a number of companies use IC op amps in their products that the sheer numbers as well as the design differences (even if all details were available) would prevent us pinpointing which of these sound good and which sound bad.

Some attempts have been made to identify the "solid state" sound, to use an overworked and undefined term, most notably the Hamm paper[5] which appeared in the AES *Journal*. Hamm condemns solid state amplifiers by alleging that they sound hard when overloaded, due to their generation of an almost purely odd harmonic distortion product structure. While it is not my aim in this particular installment to get into the issues of the sound-during-clipping phenomenon, I hope to deal with them in the near future.

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One point which Hamm's studies stress, however (and one which is probably familiar to us all), is that odd harmonic distortion in audio amplifiers is painfully obvious to the ear. This point ties in quite well with the subject of this article, the control of the slewing distortion mechanism, which produces odd harmonic distortion, *inherently.*

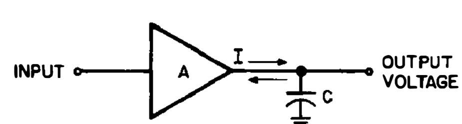
I hope the results of this study will clarify the use of IC op amps in audio, to a point that the reader will believe they don't automatically generate problems, can be used with confidence, and are highly predictable in both measured results and *listening quality.* Since the published correlation of measured results with sound quality is so woefully lacking in the audio field, I suspect any progress at all in this area will be more than welcome.

A General Look at Slew Rate

Slew rate limiting can occur at virtually any point in the audio chain, but is most likely to occur at points of maximum voltage swing, where the required rate of change is greatest. The limitation comes about due to a fundamental voltage/current relationship in capacitors as illustrated in block form in Fig. A.

Here an audio amplifier is represented by the symbol A. Capacitor C, which in practice could be either an integral part of the amplifier or an external load capacitance, is electrically connected across the output of the amplifier. Thus it sees the full output volt-

FIG. A



SR = slew rate = maximum output voltage rate of change, in V/µS (µsec.) or V/S
in terms of circuit parameters,
$SR = I/C$ where I is capacitor charging current (amp) and C is capacitance being charged (Farads). Yields SR in terms of V/S (divide by 10^6 for V/µS).
Example: I - 1mA = 1×10^{-3},
$C = 0.01\mu F = 1 \times 10^{-8}$
$SR = 0.1 V/\mu S$

Fig.A: General representation of slew rate limiting.

FIG. B

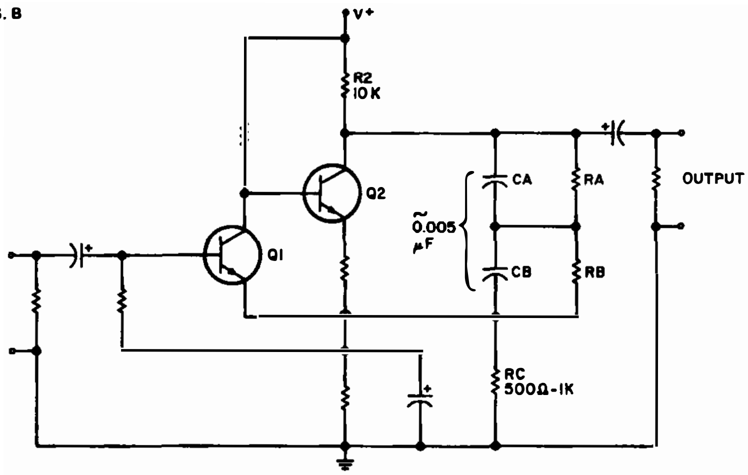
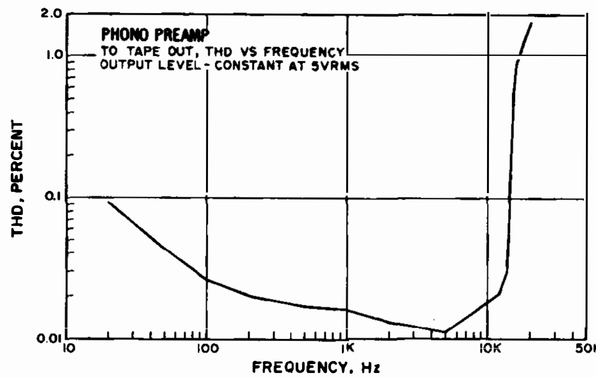


Fig.B: Classic two transistor RIAA preamp.

FIG. C



age swing. It is a fundamental circuit relationship (*regardless* of the type of active devices used) that the maximum output current available from this amplifier will determine the maximum rate of voltage change which can appear across this capacitor.

This can be stated mathematically quite simply. For a current I, the rate of change, or slew rate (abbreviated SR), is simply SR = I/C. With I in Amperes and C in Farads, SR is in units of Volts per second (V/S). More commonly, SR is given in V/µS, as 10V/µS. 10^7V/S would be equivalent to 10V/µS.

A constant current into a fixed value capacitor will result in a linear, or ramp-like waveform of voltage. Audio signals are not ramps, or triangular waveforms, to be sure, but for a sine wave the maximum rate of change occurs at the zero crossings. This factor is the basis of the so-called "full power bandwidth" (abbreviated fp) which relates SR and a maximum full amplitude sine wave signal. This relationship is simply

$$fp = \frac{SR}{2\pi Eop}$$

where Eop is the peak output voltage. Thus the two parameters are directly related, and slew rate can be expressed in terms of fp as

$$SR = 2\pi Eopfp$$

We will do well to remember that fp is by definition the beginning of *complete* slew rate limiting, and generally will be accompanied by 1-3% THD. The desired output sine wave under slew rate limited conditions will in actuality more nearly resemble a triangular wave, due to the rate limiting effect.

The above is about as complete a discussion as you will be able to find in many references on the subject, as if to imply that is all there is to it. Nothing could be further from the truth. In fact, slew rate, or *slewing induced distortion, is the single largest distortion mechanism in solid state audio amplifiers today*.

This is a little appreciated fact, and is evident by both the dearth of published material on it in audio literature, and the number of product specifications which neither recognize nor define it. Further, recent comments to this writer by individuals seemingly knowledgeable in audio even indicates some confusion among professionals about how to measure slew rate. Adding to the confusion, a rash of recent articles speak of transient IM distortion as though it were something entirely separate from slewing induced distortion when in actuality it is not, in many cases. I hope this article will clear up some of this confusion and provide the reader with a convincing overview of the magnitude of slew induced distortion problems in audio.

As I suggested earlier, many discussions of amplifier slewing rate imply (or flatly state) that it is a mechanism which suddenly produces distortion when the full power bandwidth point is reached. This is simply not true, unless you accept the premise that distortion is only significant when it reaches 1%. In reality, the approach of the slewing rate limit of an amplifier can produce easily measurable and significant distortion products at frequencies as low as 1/10 or 1/5 of the full power bandwidth. I have personally observed this phenomenon on many, many samples of different IC op amp types, as well as more conventional preamp circuits and power amplifiers.

Many people evidently see no reason to get excited over the slew rate flap; and simply do not believe it is a major audible defect. I can recall one design engineer (who claimed to be an audiophile) who said any distortion above 10kHz was not worth worrying over, as it was by definition inaudible. This is naive optimism.

Consider the presence of two high level, HF tones with a close frequency spacing. Their LF intermodulation product pops out down in the bass or mid frequency region, and is *highly* audible. And this is exactly the way slewing induced distortion works on two-tone IM tests, producing strong LF intermodulation components. You should also consider brief HF transients as well as tones. If the slewing rate of an amplifier is pushed (not even exceeded, necessarily) even momentarily, intermodulation will occur and will probably be audible. It can also occur (indeed, is most likely) with high level *supersonic* signals which will cross-modulate down into the audible region.

These audio circuit ills can be dealt with in many ways, which we will discuss later in this series. First, however, let's examine cases which give rise to slewing problems that I hope will give you a deeper appreciation of the severity of the problem. In other words, if you are still unconvinced, let me cite some graphic examples to make a believer of you.

One case in point is the familiar RIAA phono preamp stage. I would venture a guess that a great many of them suffer (at least potentially, if not in fact) from slewing induced distortion, simply because they cannot fully charge their own equalization capacitors. Fig. B, a simplified schematic of the classic two-transistor feedback pair as applied to RIAA phono preamp use, will demonstrate this.

At high frequencies, equalization feedback capacitors CA and CB appear as a single equivalent capacitance to ground in series with RC. A typical (lumped) value of capacitance for CA-CB will be in the range of 0.005µF, with RC in the range of 500 Ohms to 1k. Collector load resistor RL may be about 10k.

Much is written about phono preamp overload phenomena, to the point that supply voltages are being run at 30 to 40V, to handle high cartridges outputs, with the object of yielding a 1kHz output of 10V RMS or so. *[The Technics SU 9600 reportedly uses 136V in its preamp. See* Wireless World, *Nov. '76, p.41.--Ed.]* Here we uncover the inconsistency of this thinking, however. If such a stage must handle 10V at 1kHz, is it not reasonable to expect it also to do so at 20kHz? I would think so, but it is just about impossible with typical circuit values. For instance, a 10V RMS level is 14V peak, and at 20kHz the required slew rate is

$$\begin{aligned} SR &= 2\pi Eopfp \\ &= 2\pi(14)(20000) \\ &= 1{,}750{,}000 \text{ V/S (or 1.75 V/µS)} \end{aligned}$$

since SR = I/C, and using a C of 0.005µF the required charging current is

$$\begin{aligned} I &= (SR)(C) \\ &= (1.75 \times 10^6)(5 \times 10^{-9}) \\ &= 8.75\text{mA} \end{aligned}$$

Now 8.75mA by itself may not appear to be an insurmountable limit, but this stage will be generating about 1% distortion at 20kHz for this level of current. If, as a safety factor, we raise the current by a factor of 5, the Q2 stage will be running at over 50mA which will get the distortion down, but certainly creates other problems, as Q2 will be dissipating nearly a Watt. I believe it should be obvious that the circuit values of Fig. B will not allow 10V RMS @ 20kHz.

One might be tempted to reduce capacitor size to gain relief, but this will in turn raise resistance proportionately, creating noise problems. In a limit-

FIG. D

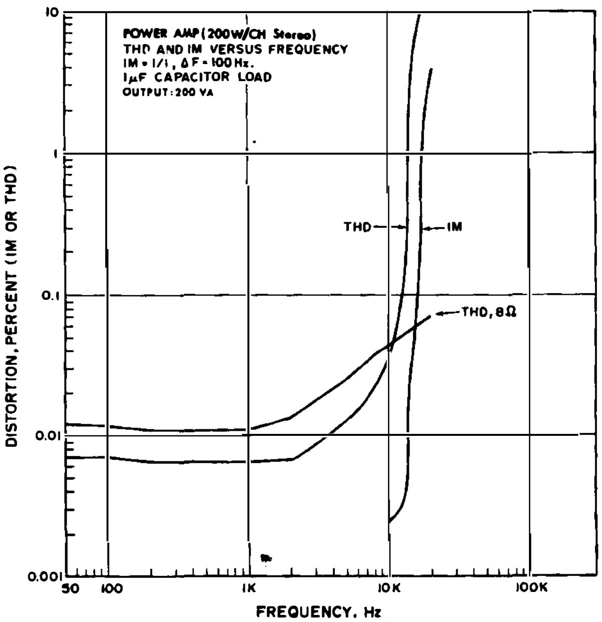


FIG. E

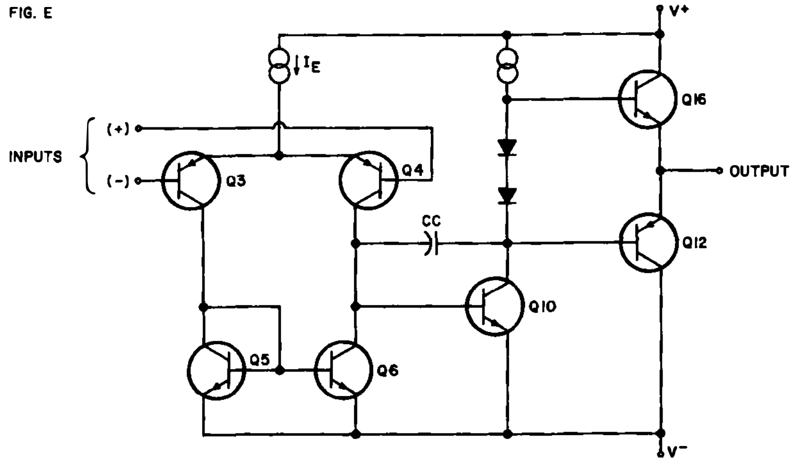


Fig.E: Simplified IC op amp (301 or 741).

ed look into this problem I have noted a great many examples of this type of circuit which suffer from the same fundamental ill. Graphically the distortion generated by such a stage is shown in Fig. C, the THD performance of a phono preamp circuit in a currently popular preamp. This circuit is severely slew limited, and although capable of 10V @ 1kHz would only produce 5V @ 20kHz, the level at which the data was taken. The measured slew rate was 0.65V/μS, which agrees reasonably well with the bias current and capacitance values.

My point here is that although the amplifier circuit *without* capacitor loading may have a high (and adequate) slew rate by itself, it simply cannot supply sufficient output current to drive the RIAA feedback network to full output at high frequencies. The result is gross distortion at these frequencies, in terms of both THD and IM.

The example in Fig. C is by no means an isolated case; a great many widely publicized "high performance" IC phono preamps suffer from similar ills. Viewed in this light, it is somewhat ludicrous to tout a preamp circuit for high overload, high undistorted output capability at a frequency of 1kHz, if it cannot also produce a similar output at 20kHz. As we will see in the course of this discussion, one of the yardsticks which can detect the presence of slew induced distortion is a high resolution measurement of THD through the audio band up to 100kHz or so.

If the distortion level at full voltage output at 20kHz is low (on the order of 0.01%), and within a factor of 2 or so of the 1kHz distortion, the amplifier is likely to be free of slew induced distortion, and thus of the accompanying intermodulation effects. On the other hand, if the amplifier cannot produce full voltage undistorted output at 20kHz, it is *highly* likely to be suffering from slew induced distortion.

Before departing the subject of phono preamps and their susceptibility to this type of distortion, I should mention that Holman's paper, "New Factors in Phonograph Preamplifier Design,"[16] *(see also* The BAS Speaker, *Nov., Dec. '75 and Jan. 76.--Ed.)* gives an excellent discussion of testing methods for phono preamps to detect this form of distortion. This paper is required reading for the audiophile.

The signal path is full of such things as line amplifiers, tone amplifiers, and equalizers. This generally very broad area of signal processing can utilize many different forms of circuit technology, but some of the most efficient realizations come about with the use of IC op amps. A major portion of this article will deal with a study of slew distortion in IC op amps, with conceptual results which are equally applicable to preamp, line amp, or power amp--indeed *any* audio amplifier. First, however, let me comment on slewing in the output stage of the power amplifier.

The final power output stage of amplifiers is acutely sensitive to slewing induced distortion, for several reasons. First, the voltage swing is at its maximum level, because of the high powers involved. A 200 Watt into 8Ω amplifier, for instance, must swing about 56 Volts peak for full output. In terms of the full power relationship, this would require a slew rate of 7V/μS for 20kHz reproduction. Since this rate would produce a 1% distortion at this frequency, a slew rate for this power level should obviously be several times 7V/μS.

Ideally, a slew rate which will result in minimum slew induced distortion is on the order of 0.5 to 1V/μS *per peak output Volt.* In the case of the 200 Watt amplifier this would imply a 56V/μS slew rate, which gives you an idea of the severity of the problem. (The rationale behind the 1V/μS per peak Volt is explained in detail below.)

Given a power amplifier which could slew at a rate on the order of 50V/μS (which is no mean feat, by the way), it should ideally be capable of this slew rate *into varying loads.* In practice it is one thing to build an amplifier which can slew at 50V/μS into a resistor, and quite another thing to do so into a reactive load.

Considering a 1μF capacitor as an example, a 50V/μS slew rate would require an output current of 50 Amperes. It is quite easy to see how difficult this becomes when the accompanying high voltages which must simultaneously be handled are considered. Practical amplifiers under these conditions will slew at rates closer to 10V/μS.

Thus, we should clearly understand that a power amplifier can be slew limited from one of two sources: either its own internal compensation capacitance(s) or from reactive loading which causes the protection circuitry to activate, so limiting the output slew rate. *Either* of these two conditions will severely distort high frequency waveforms causing serious intermodulation products, which will result in poor reproduction. I do not ask you to accept these statements as factual on faith; they may be readily demonstrated by measured data.

Fig. D is a graph of measured distortion, in the form of both THD and two tone 1:1 HF IM. These are plotted in percentages as a function of driving frequency. This is also justified in the IM case, as the difference frequency of the two tones is less than 100Hz (1% or less). The conditions are a 1μF pure reactive load and a constant 200VA output voltage level. The amplifier is a popular 200 Watt per channel unit. Slew rate into this capacitance was measured at 5V/μS which corresponds to an fp of 14.2kHz.

In the case of THD, we may note that it is 2% at the fp point but has begun to rise far below this frequency. I include also a reference curve of an 8Ω resistive load THD at a 200 Watt level which by contrast is less than 0.1% even out to 20kHz. Some of the gentle upward slope in this curve is probably slew limiting due to the amplifier's internal compensation, although some output stage crossover conduction spikes are also present. Note that both curves break away from the LF plateau at a fairly low frequency, about 2kHz.

The IM curve gives convincing evidence for the slew induced distortion argument, demonstrating that low frequency products at levels of several per cent can be produced simply due to slew limiting. Here the two-tone pair is swept in frequency from 10 to 20kHz, and the general shape of the IM distortion rise is remarkably similar to the THD rise, although less sensitive. This example indicates a correlation between the two methods of testing for this particular distortion mechanism. Further evidence of this correlation is indicated in the op amp data to follow.

The two-tone 1:1 HF IM test is a particularly appropriate one for power amplifier tests, for several reasons. Although not quite as straightforward as a THD test it is still relatively simple to implement compared to the sine/square method or noise transfer test, and gives reasonable sensitivity to this distortion mechanism.

Extended range THD tests to 100kHz or more at full output voltage level with high resolution equipment can also reliably indicate SID (and consequently TIM) but are not desirable in power amps for two reasons. First, they will cause a very high stress on the output stage due to storage time effects, with even possible destruction in some cases.

Second, as pointed out by others,[12, 17] the harmonic distortion figures measured will most likely be erroneous, due to the natural rolloff with frequency of the amplifier. Two-tone IM tests up to 20 (or 30) kHz do not stress the amplifier output transistors nearly as much, and since the products being measured are reflected downward in the audio spectrum there is no loss of accuracy due to rolloff.

The two-tone technique is not new, but has not been used to any substantial degree in the U.S., particularly in amplifier testing. I advocate the adoption of the 1:1 two-tone swept HF IM test as a standard technique for audio amplifiers, at both power output and signal processing levels. Future equipment tests in this publication will utilize the technique, and work will also be initiated on an instrument suitable for home construction.

Some comments are appropriate here on the overall problem insofar as power amps are concerned. The more insight we gain about the power amplifier slewing problem, the more staggering the situation appears to be. Some of these problems, for example the high fidelity drive into reactive loads, seem almost insoluble with presently available technology. This article cannot really hope to completely address the power amplifier problem, and will not attempt to. What we hope for is an overview of the mechanism of slewing induced distortion in its general form, and a fairly definitive picture of how it can be measured (and controlled) in low level amplifiers, particularly IC op amps.

This article is the first of an ongoing series on audio amplifier distortion; future installments may perhaps more completely treat the power amplifier case. Many of the points and measurement techniques we will discuss are equally applicable to low level and power amplifiers, and as the narrative progresses this will be underscored.

The Op Amp Slew Limiting Mechanism

Like so many other things, IC op amps are used in audio in ways that are good and bad. The outcome depends upon both the user's viewpoint and his/her level of understanding. Incomplete understanding of the problems which arise in effectively applying IC op amps to audio is in itself understandable: the range of available devices is staggering. Yet, certain general principles govern effective use of these devices, and designers should at least understand these fundamentals. I attempted to examine some of these problems in my AES paper,[23] which I later expanded into one chapter of my first book[24] and the audio volume derived from it.[25] However, this material is no longer adequate, for two major reasons.

First, many new, improved devices have appeared since they were published, and second, slew induced distortion warrants a much more extensive discussion. SID in audio circuits (particularly in op amps) is probably the *only* major distortion mechanism, if the design is a reasonable one. This may sound startling, but the cases to which this statement applies are probably more numerous than many people suspect. I am quite sure this type of distortion causes many audio circuits to sound bad, due to the nature of their distortion products.

However (at least in IC op amps) SID can be dealt with, using appropriate design techniques and fairly simple test procedures, even with a minimum of equipment. Op amp slewing rate problems can indeed give rise to TIM, but I suspect that an amplifier truly free of SID will never have TIM. However, I am not sure that a so-called "TIM-free" amplifier cannot have slewing problems. If you consider the slew rate into reactive loads, the amplifier will, of course, generate IM as I have shown above.

In the testing part of this study, I examined IC op amp slewing by closely measuring the behavior of a large number of devices. Out of this I developed a predictive analysis technique showing whether a given device would be free of SID for a given application, as well as several general criteria for slewing specifications, for devices as well as for circuits.

Slew induced distortion comes about because of the nature of an op amp's design. It is a phenomenon that can probably never be completely eliminated, but it can certainly be minimized to manageable proportions. To understand the basic mechanism, refer to Fig. E, a much simplified diagram of a 301 or 741 IC op amp (this general circuit is equally applicable to many power amplifier designs as well).

This amplifier consists of an input differential pair Q3-Q4 fed by a constant current source, I_E. The Q3-Q4 outputs are fed into a current mirror comprised of Q5-Q6, which converts the differential output to single-ended form at the base of Q10. Q10 is the second voltage gain stage of the amplifier, and its output collector voltage swing is buffered by transistors Q12-Q16 before appearing at the output terminal.

This amplifier has an overall low frequency voltage gain of 100dB or more, and is compensated for unity gain stability by Cc, a Miller integrating capacitor connected around voltage gain stage Q10. This capacitor causes the voltage gain of Q10 to decrease at high frequencies, yielding the necessary gain/phase characteristics for stability.

Because of the very high open loop gain and the necessity for a stable closed loop under feedback conditions, the presence of Cc is a necessary requirement, at least for a general purpose op amp.

However, the presence of that frequency compensation capacitor has a very serious effect on the amplifier's speed, most notably its slewing rate, or output voltage rate of change ability. Examining the diagram you can observe that the right terminal of Cc sees essentially the full amplifier output voltage (Q12-Q16 being unity gain buffers). The voltage rate of change across this capacitor (which is, in fact, the amplifier's slew rate) is determined by the current into it, which can only come from Q4 or Q6. The maximum current Q4 can deliver is I_E, under an input condition where Q3 is fully off; the maximum current Q6 can sink is again I_E, when Q4 is fully off (by virtue of the current mirror Q5, Q6). Thus the peak current into the capacitor is I_E, for either charge or discharge.

With simple capacitor current/voltage relationships, we can express the voltage across Cc, which is the circuit's slew rate.

$$SR = \frac{\Delta Eo}{\Delta t} = \frac{I_E}{Cc} \text{ Volts/second}$$

In a 301 or 741 amplifier, I_E is about 15µA, and Cc is 30pF. Therefore the slew rate is 0.5 Volts per microsecond. This means the amplifier can execute a full scale output swing from +10V to -10V (20V) in 40µS.

In terms of sine wave output signals, there is the aforementioned equation which relates SR to the full power (maximum p-p voltage) sine wave output frequency, fp. In the case of typical op amps, the specified peak output swing is 10V peak, for devices we will be discussing. For the 0.5V/µS slew rate mentioned above, the corresponding fp is then

$$fp = \frac{SR}{2\pi Eop}$$

$$\approx 8\text{kHz}$$

Doesn't sound too encouraging, does it? Actually in terms of a 20kHz full power bandwidth, the SR would be 1.25V/µS.

Now we may look more deeply into this slew rate limiting mechanism involving the amplifier's input stage. In practice op amps are intended to be used in conditions of an ideal 50-50 current balance in the input stage, a state where I_E splits equally between Q3 and Q4. Under such a condition the differential input voltage is *near zero*, recalling one of our fundamental axioms (see reference 24, chapter 1).

Under changing conditions of common mode input voltage, frequency, output level, loading, etc., this condition *must constantly be maintained, if the device is to function as an op amp*. If the input stage is *not* balanced, the input voltage is by definition not zero, *therefore it is not operating as an op amp*. This condition can (and will) occur when an input signal is applied which exceeds the amplifier's slewing ability. However, it can also happen to a lesser degree, for signals which require rates of change just below the slew rate. This would also be an example of an unbalanced condition (but less than 100%).

In the case of a square wave input with a rise time faster than the slew rate, the input stage will toggle back and forth between conduction states of Q3 and Q4, as the loop attempts to follow the fast square wave. Fig. Fa illustrates this effect. Note that during the slewing rate interval where the output is ramping to a new level, very large input differential voltages occur. Again, by definition, the amplifier is not functioning as an op amp during this slewing interval.

The effect of a sine wave input is shown in Fig. Fb. In the normal input/output example, the input voltage is reproduced as a reasonable replica (here on a 1-1 basis). But, when a frequency and amplitude combination is applied which exceeds the slewing ability of the circuit, the output becomes triangular rather than sinusoidal in shape, as you can see by the superimposed (dotted) proper waveform shape. If we raise the frequency, the amplitude would fall even further, and become more triangular in shape.

One highly interesting aspect of slew rate limitation in op amps is not always obvious, particularly if you just eyeball an output waveform on a scope. At the full power frequency, fp, THD will be on the order of about 1% (the exact number varying, due to different factors). You can see this 1% distortion rather easily by watching for the ramp-like slopes on an output sine wave.

But what happens below fp? The distortion does not just go away suddenly as you lower frequency. By contrast it can be significant at frequencies as low as 1/10 of fp. This is the real reason behind this article. SID can be subtle, and present in almost every op amp circuit.

If we examine Fig. E again and recall the statements concerning the balance of Q3-Q4, we can see that under slewing conditions Q3 and Q4 are operating far from ideally. Even at a frequency just below fp Q3 and Q4 will by necessity be swinging far beyond the 50/50 balance point. In fact they will alternate between near 100% and 0% conduction of I_E at the slew limiting frequency. With this situation, Q3 and Q4 will be generating *gross* amounts of odd order distortion, predominantly third harmonic. This is the best way to identify SID in THD measurements: watch for the appearance of third harmonic components of distortion.

FIG. G

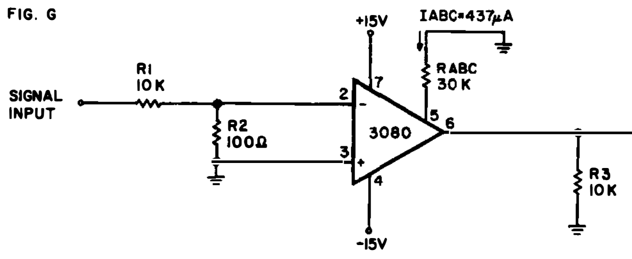


Fig.G: Differential input stage model to illustrate distortion generated.

V_O (RMS)	ΔI_O (% I_{ABC})	THD (%)
3.	97	10
1.7	55	2.7
0.95	31	0.83
0.53	17	0.26
0.3	9.7	0.09
0.17	5.5	0.07

To illustrate the seriousness of the distortion generated in a differential pair as a function of current swing, I devised the test circuit of Fig. G. This simple circuit uses a 3080 OTA as a differential pair model which delivers an output current into R3, a fixed load resistor. The current set up in R_{ABC} is I_{ABC}, which for our purposes is analogous to I_E of Fig. E. The output current of a 3080 is $\pm I_{ABC}$, which is similar to the $\pm I_E$ output current of Q3-Q4 in Fig. E.

A signal input applied to the circuit at various levels allows the output to be measured from near full scale output current swings, downward. The results demonstrate the problem impressively. This should clearly illustrate the potential non-linearities of an op amp as it approaches its slewing rate limit. Even relatively small imbalances in the differential pair produce appreciable distortion.

The slew rate of a given amplifier *is in no way directly altered by feedback.* Slew rate is a parameter *independent of feedback,* and will measure the same whether tested open or closed loop. Actually to be completely correct, slew rate can only be measured open loop, since by definition causing an amplifier to slew by excitation with a fast step will open the loop momentarily, during the slewing interval. But feedback, regardless of whether it is 100% or 1%, will not change the basic slew rate, since slew rate is determined by I/C relations.

While Fig. G has illustrated the non-linearity of the differential stage open loop, the matter can also be demonstrated in a different manner, quite simply, and with little equipment.

Fig. H is a plot of the actual p-p input error voltage of a 741 op amp, operated in a unity gain inverter. This particular connnection is a convenient one, as the error voltage appears between the summing point and ground, thus is easily observed.

Although this error voltage is ideally zero, in a practical amplifier with a 6dB per octave open loop rolloff it will rise 6dB/octave with frequency. This rise will be quite predictable at frequencies below the slew limiting point.

In Fig. H the rise may be observed from 500Hz to about 7kHz, where the voltage begins to rise much faster. This rise signifies the onset of slew limiting, as the loop is forced to create much larger error signals to swing the input stage to greater percentages of output, in charging the compensation capacitor.

This test is not very sensitive, nor does it yield quantitative data. It does, however, quite simply demonstrate the non-linearity and abrupt deviation from predicted behavior associated with slew limiting.

Although specific designs of op amp input stages vary widely and take on many forms different from the simple one shown, a great many of them use bipolar input stages without emitter degeneration. It is this type of input stage which is most susceptible to the non-linear, voltage in/current out problem which causes high distortion under slewing conditions. The input transistors can be either NPN or PNP, but if undegenerated they will be highly non-linear away from balance, or during slewing.

To raise slew rate in an op amp, either I_E must be raised or Cc lowered. If the op amp is an externally compensated type, and is to be used at a high gain, Cc can be reduced, which raises slew rate in direct proportion. However, such a solution is not always desirable, or possible. For instance, if the op amp is used at unity gain, slew rate must be increased by other means.

FIG. Fa

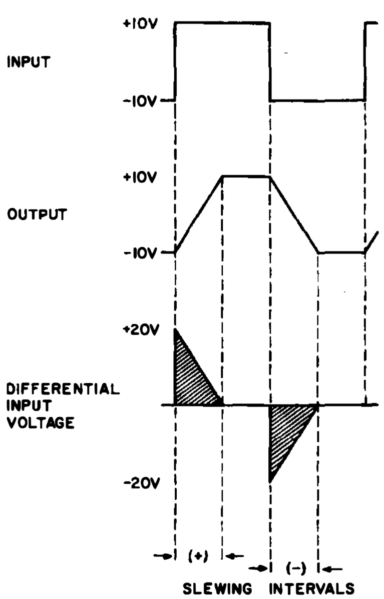


Fig.Fa: Response of op amp to a square wave input.

FIG. Fb

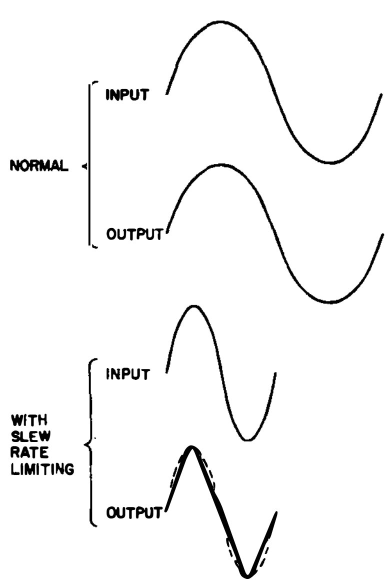
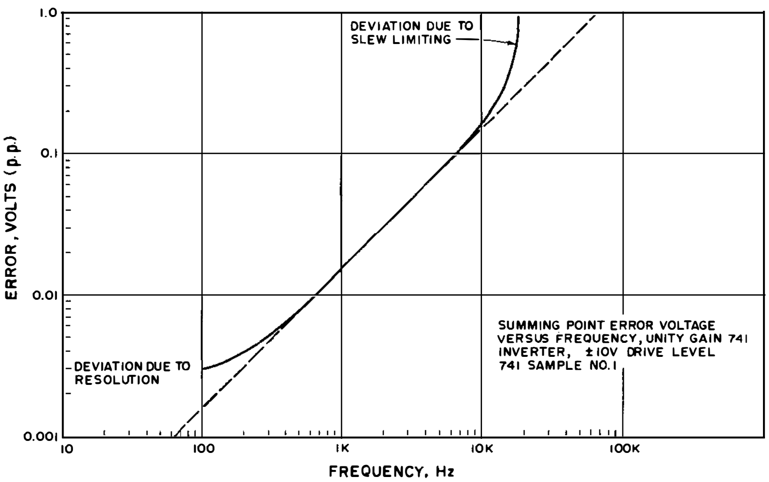


Fig.Fb: Response of op amp to a sine wave input below and above rate limiting.

FIG. H



Some op amp designs attack the problem by various means of increasing I_E, which allows more current to charge Cc. This is usually accompanied by the use of emitter degeneration in the differential pair, to lower transconductance. This scheme is illustrated (in much simplified form) in Figs. Ia1 and Ia2. RE1 and RE2 are the emitter degeneration resistors, and Q1-Q2 the input differential pair. Since the use of RE1 and RE2 lowers the stage gain, I_E can be raised, and thus slew rate is raised.

Another method uses input stage devices with lower basic transconductance: FET differential pairs, for instance, illustrated in Fig. Ib.

In Fig. Ib, a P channel JFET pair is the input setup, while in Ib2 PMOS devices are shown. Both techniques lower input stage gain directly, because FETs have lower basic transconductance than do bipolars. Thus I_E can be raised, increasing slew rate.

Another technique uses "slew enhancement" which dynamically increases I_E during the slew interval only, as illustrated by Fig. Ic.

Here Q1-Q2 are the differential amplifier pair which operate more or less conventionally, *for small signals*. At high slew rates additional current is forced by the cross-coupled arrangement, which "enhances" or raises slew rate. The performance of all of these means of slew rate improvement is discussed in detail in the testing phase of this study, coming next issue.

PREPRINTS ANYONE?

The editors of TAA are seriously considering offering preprints of the four articles in Contributing Editor Walter Jung's series, "Slewing Induced Distortion in Audio Amplifiers". Part two: Phase I Testing, Part 3: Phase II Testing and Part 4: Listening Tests along with this first part will, we estimate, run to approximately 40 pages of text and charts. For those who want to peruse this important series as a whole ahead of 1977 publication dates, the cost will be $16.50 postpaid.

Those who are interested please make checks payable to The Audio Amateur. If sufficient orders are on hand by April 1, 1977 we will proceed with the project. If not, checks will be returned. The preprints, if produced, will be mailed first class about April 20. Send orders to: The Audio Amateur - Preprints
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FIG. Ia1: NPN INPUTS

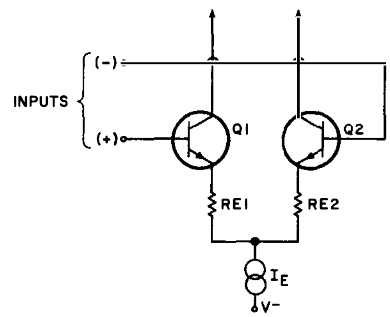

FIG. Ia2: PNP INPUTS

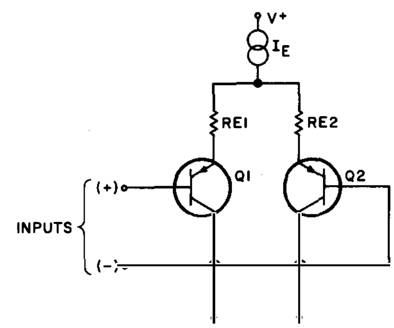

Fig.Ia: Emitter degeneration.

FIG. Ib1: P CHANNEL J FETS

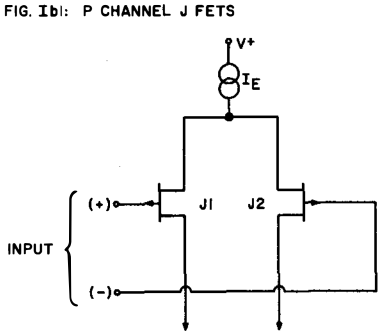

FIG. Ib2: PMOS

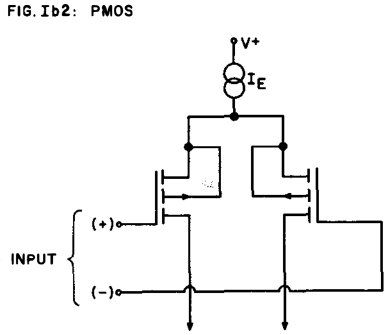

Fig.Ib: Lower transconductance input devices.

FIG. Ic

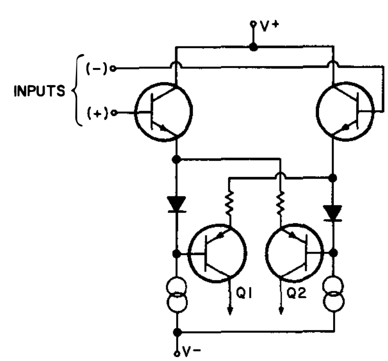

Fig.Ic: Slew enhancement.

THE NOT-QUITE-PASSIVE RADIATOR

spectrum. He designed a simple circuit (which varies daily, so far, and is apparently a long way from optimum) which applies maximum damping to the secondary radiator (shunts the voice coil) until the amplifier power applied to the primary radiator exceeds a pre-set point. He then raises the resistance across the voice coil to permit greater cone movement, thus providing a transient signal which is the sum of the output of the *two* cones.

A similar approach assumes that the first element of any transient is positive: that is, the wavefront is positive. If a diode/capacitor circuit is applied to the voice coil of the secondary radiator, it is possible to permit free cone excursion on the "downstroke," with limited rate of travel on the return, thus utilizing the secondary radiator only a small percentage of the time. Audible radiation would take place only when frequency was low enough to permit coupling between the two radiators, and then only when the signal was in the positive (ascending wavefront) portion of the curve. Strong percussives, then, might be reproduced with considerably more realism than otherwise possible.

It has been pointed out that the susceptibility of any speaker to this application can be easily tested; an important point, as some very low efficiency drivers are not suitable. Simply rap the magnet assembly while observing the cone, first with an open voice coil and then with a short (a 25¢ coin reaches most terminals). The difference should be quite significant.

To reduce duplicating experimentation, keep a log and work on only one of the several variables at a time. Remember a nearly infinite number exists of combinations of components which can be added to the circuit governing (or "reporting on") the operation of the not-quite-passive radiator.

If you decide to try this technique, good luck, be persistent, and please share your information.

THE FOLDED, STAPLED BASS HORN
Continued from page 13

the Calrad and Radio Shack 5" speakers, as they are inefficient and yield poor bass response. The speaker chamber can be modified to contain a pair of 8" speakers (cut off an inch of the cardboard at the throat for extra depth if necessary) or six 4" speakers, if desired.

This enclosure is designed to be both a quality low frequency speaker and a high quality midrange unit as well, with a useful range up to 3,000 Hz without the bother and expense of an additional crossover. The horn itself does not radiate much energy above 200 to 250 Hz: it's sole purpose is to couple the low-frequency mechanical energy of the small drivers to the surrounding air. The horn part of this enclosure is driven by the back side of the speakers and works to restore the low frequency energy which is normally lost through poor coupling.

The midrange frequencies are directly radiated from the front of the four speakers, which constitute a small array. Several developed mathematical models help explain why an array can be superior, in the midrange frequencies at least, to just one of the speakers by itself. Whatever the ultimate reason for this may be, the fact remains that speaker arrays are quite effective.

The 5" drivers are lightweight and rigid enough to respond accurately at midrange frequencies; however, this is certainly not the case at higher frequencies. The cones are simply neither lightweight enough or inertia-free to follow the amplifier signal faithfully at the higher frequencies, and should be crossed over to a high quality tweeter at a maximum upper limit of 3,000 Hz. Even casual listening can detect offensive intermodulation distortion in the higher frequencies by the severe ringing when playing a very complex source such as choral music. A single low-pass crossover coil however, will limit the speakers' response to the midrange frequencies where they perform quite well. (See next issue for a crossover circuit.--Ed.) The grouping of the four speakers becomes directional at frequencies above 1500 Hz, so that the ideal crossover would be below that frequency.

Any good tweeter or tweeter-midrange will help out in the high end, although the best complement by far for this horn is a version of the Heil driver which you can construct for around $25. The details will appear in the next issue.

The speakers will be ideally suited to a lower power amplifier than what is commonly in use today, as the enclosure increases the efficiency of the speakers tremendously. A twenty watt amplifier should be sufficient to drive the speakers to very high volume levels while still having enough reserve to do permanent damage. The enclosure works best in a corner, with the mouth about nine inches from the wall, or if no corner is available, the mouth should be facing a wall 6½ inches away. This placement will best utilize the wall or corner as an extension of the horn, further improving low frequency coupling.

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Slewing Induced Distortion: Part 2a

Phase I: Total Harmonic Distortion Tests for SID

by Walt Jung
Contributing Editor

AT THE OUTSET of this study we really did not know what form of test methods we would need to completely identify, isolate and quantify slewing induced distortion (SID). Granted, it is readily observable on an oscilloscope (when gross), but to what degree will a percentage of "x" distortion be related to so many V/µS? Or will total harmonic distortion (THD) and intermodulation (IM) tests be shown to be related to slew rate? To each other?

Now that seems like a long time ago, actually. I have observed a pattern all through the course of my testing which correlates slew rate to measured THD in every instance where slewing is apparent. Hardly an isolated phenomenon, SID effects could be observed in all but one or two cases out of the several dozen IC types tested. Once I gained experience with the testing techniques, performance of devices which were tested late in the study could be reliably predicted from examining their specs.

As you will see when we get to phase II tests, a further correlation can be made between slew rate/THD tests and two tone 1/1HF IM tests. Once again, as I gained experience device performance could also be reliably predicted for IM tests, from either the THD results or the data sheet spec (if sufficiently detailed).

Although it's perhaps somewhat premature, I noticed in all cases that the THD tests seemed to stress an amplifier much more vigorously. I gathered similar data from both tests in relation to slew rate, but the THD tests were much more demanding of an amplifier. THD showed up faults more readily, and gave generally higher percentages of distortion. This is somewhat surprising, but probably most welcome to many, because the THD test method is so much simpler to use.

For the above reasons the bulk of the data I gathered in my study (and that presented here) is based on THD. It is clear to me that SID can be reliably detected by THD analysis, and that the results correlate well with other methods.

Block Diagram

The THD test system is relatively simple and consists of the equipment of Fig.I-1. There are two signal sources: a function generator supplying fast rise time square waves, and a low distortion sine wave source. The square wave source is used to observe the slew rate of the unit under test (U.U.T.), in conjunction with the oscilloscope.

The sine wave source provides the high purity sine wave for THD tests, and may or may not also contain the THD analyzer. Although in concept other gear could as well be used, the Sound Technology equipment conveniently supplies the required resolution, range, and accuracy in a single package.

Pertinent details of test set-up specs are as follows: the sine wave source and analyzer should have a residual distortion of ≈0.002% over the range of 100Hz to 10kHz, and as low as feasible up to 100kHz. The object is to make high resolution, extended range THD measurements at *full output voltage levels*, from 100Hz to 100kHz. The set-up must have extraneous or spuriously induced noise residuals of less than 100dB referred to full scale, for good repeatability and a high confidence factor.

The distortion products from the analyzer must also be made available for monitoring, by one channel of a dual trace scope, during THD tests. This enables the characteristic third harmonic distortion rise to be identified, pinpointing SID.

The square wave source should generate a variable frequency in the range of 1 to 100kHz, but will most often be used at 10kHz. It must produce ±10V square waves into a 50Ω termination with a rise and fall time of less than 100nS, preferably 50. The square wave should be free of overshoots, ringing or other fidelity shortcomings, as it will be used to measure devices under test. I used a Heath IG-1271 function generator, but many others are also satisfactory.

To begin a THD measurement on a device, we must first measure for slew rate using the square wave source and scope. Now at this point we must make quite clear what is done, and just how slew rate is measured in detail. Even if it seems simple, bear with me. We have pitfalls to avoid, as well as important subtleties to note.

First of all, a measurement of slew rate must be just that, a measurement of the amplifier's output in a *slew limited operating mode*. At low output levels the output waveform from an amplifier should resemble Fig.I-2a, which shows exponential rise and fall (characteristics of an RC time constant). In the low signal level range (1-2V) this shape will be observed, but as we approach full output, the output waveform will taken on a linear rise and fall time, with ramplike slopes, as in Fig.I-2b.

This slew rate limited condition is best measured at full rated voltage swing; for standard op amps this equals ±10V. The peak amplitude swings must not be allowed to clip, as this may invalidate the reading(s). If there is any doubt whatsoever concerning possible clipping (clipping of a square wave is not always obvious), switch the function generator momentarily to triangle or sine, waveforms with peaks which will readily display clipping.

Now to be accurate in measurements, the scope (plus probe, if used) must be a wideband model, with less than 50nS risetime and 10mHz or more of bandwidth. It must also have an accurately calibrated time base (as well as vertical deflection), since slew rate is being measured as voltage change per unit of time.

In the course of testing op amps, many different slewing waveforms will be noted. The waveform of Fig.I-2b is ideal and is drawn here for reference (if you find one, let me know). Note that both the up (+) and down (-) ramps of the waveform are smooth, constant slopes and precisely symmetrical. There are no aberrations such as overshoots or ringing.

In practice you more often see waveforms such as Figs.I-2c, I-2d, and I-2e. There are all undesirable for various reasons. Exactly why is covered below under the discussion of various devices.

In making a slew rate measurement,

FIG. I-1

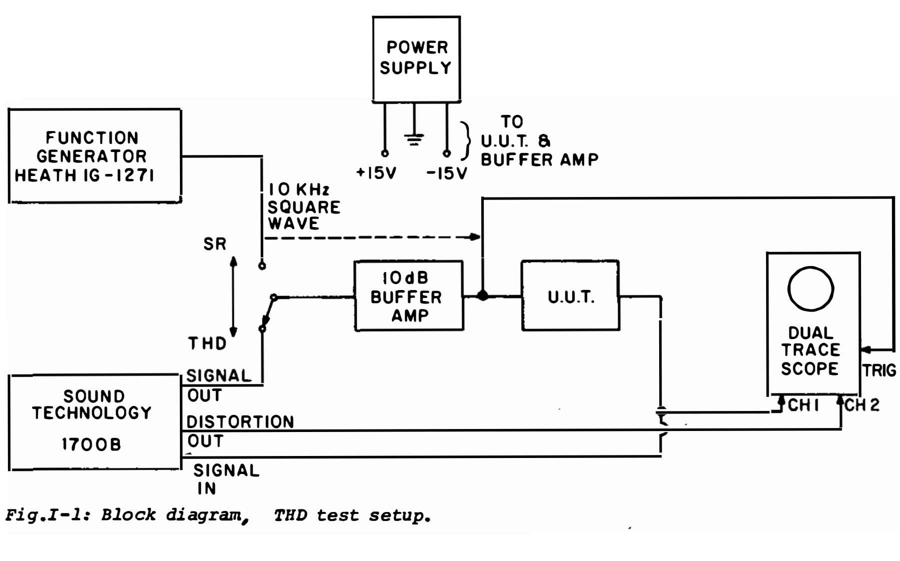


Fig.I-1: Block diagram, THD test setup.

FIG. I-2

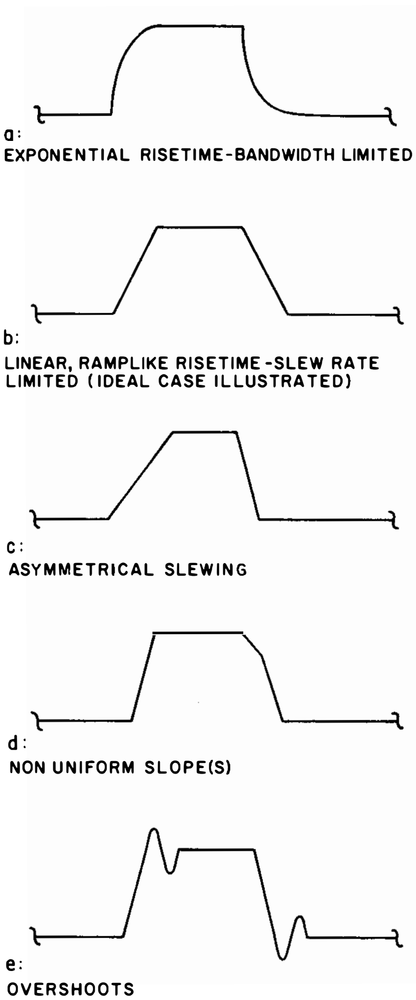
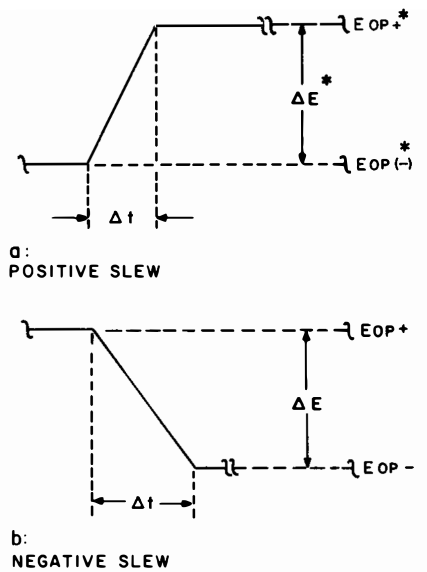


Fig.I-2: Visual waveform differences, and aberrations.

FIG. I-3



$*\Delta E = Eop^{+} - Eop^{-}$

$SR = \frac{\Delta E}{\Delta t}$

Example: $Eop^{+} = +10V$

$Eop^{-} = -10V$

$\Delta t = 10\mu S$

$SR = \frac{10V-(-10V)}{10\mu S}$

$= \frac{20V}{10\mu S} = 2V/\mu S$

Since Δt is measured from Eop^{-} to Eop^{+}, this is positive slew, then

$SR(+) = 2V/\mu S$

For negative slew, use same basic expressions, but label accordingly.

Example: $\Delta V = 20V$, $\Delta t = 15\mu S$

$SR(-) = \frac{20V}{15\mu S} = 1.3V/\mu S$

Fig.I-3: Measuring slew rate.

carefully adjust the starting position of the waveform so that the beginning of the slewing interval is aligned with the left graticule marker (see Fig.I-3a). Then adjust the signal level until the vertical deflection is exactly 20V p-p. Slow down the time base to ensure you measure the 20V between the *flat* portion of the square wave, and not any peaks, dips, or other bumps which can occur near the (+) (-) transitions.

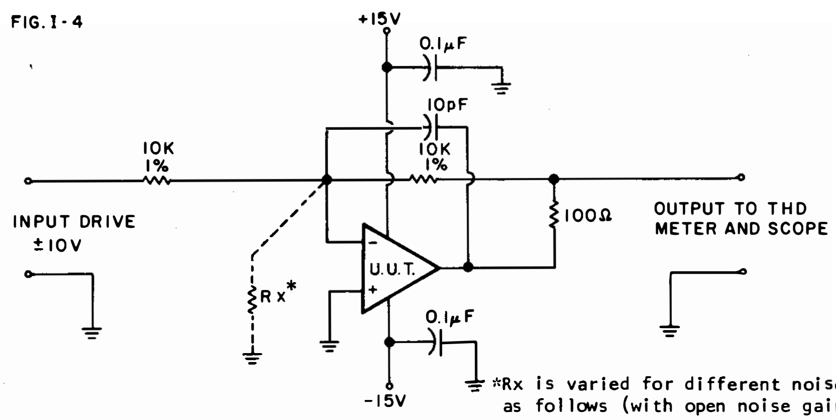
With the waveform adjusted in amplitude for the exact rated output, then again speed up the time base for a convenient display of the slewing ramp (for instance, 1µS/div) and recheck the beginning of the slew interval for the horizontal calibration point. Then read the slewing time interval, Δt, as the time from slew beginning to the crossing of the peak voltage.

You may have some difficulty if there are overshoots or other distortions. In these cases extrapolate the end point of the slew rate to where it would first cross the peak deflection. This may be necessary at either the beginning or end of the slew interval.

Measure both (+) and (-) slew rates (which may well be different) and note their rates in V/µS. Two examples are shown in Figs.I-3a and I-3b; in cases where they are different, note this accordingly.

The unity gain test circuit of Fig. I-4 was used in all but a few cases for measuring slew rate. Slew rate is measured with Rx open, in a unity gain inverting mode with x1 (unity gain) compensation for the device being tested (if an externally compensated unit).

FIG. I-4



*Rx is varied for different noise gains as follows (with open noise gain is 2).

Noise Gain	Rx
3	10K
5	3.3K
12	1K
102	100Ω
1000	10Ω

Fig.I-4: Test circuit for slew induced distortion.

Thus, the slew rate given with a devices data is the actually *measured* (not data sheet) slew rate. Supply voltages are ±15V within 0.1%, unless otherwise noted. Special cases of compensation are noted on the data which follow, as well as corresponding slew rates.

From the block diagram (Fig.1) you can observe that the drive to the U.U.T. comes through the buffer amp. Normally this creates no problem in measuring slew rate, as the 318 is one of the fastest slewing devices available (over 60V/µS on the unit used) and will thus create little additional error if the slew rate of the U.U.T. is less than 15 or 20V/µS. In the case of a high speed device (>20V/µS), the square wave can be applied directly to the U.U.T. at a ±10V level (shown dotted).

The buffer amplifier, shown in Fig.I-5, is necessary to elevate the output level of the Sound Technology 1700B up to ±10V so as to be capable of driving unity gain op amp circuits to full output. Its 10dB gain gives some reserve output capability beyond ±10V (or 7V RMS).

The A1 device is of critical importance, as any THD or noise generated in this stage will be seen as an ultimate resolution limit, if greater than that of the Sound Technology. The device used here must outperform all others being tested.

Fig.I-6 illustrates the performance of the 318 device used for the buffer amp in all tests. The lower curve here is the basic residual THD of my 1700B, from 100Hz to 100kHz. The rise in THD above 30kHz is due to SID in the 1700B's oscillator, by the way.

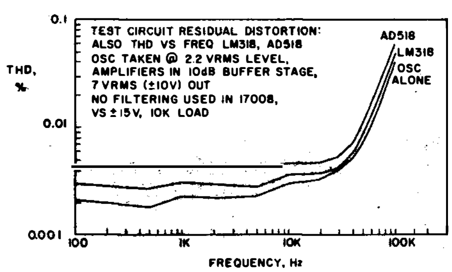
A 318 used in the buffer essentially duplicates this curve, but with slightly higher THD readings, due mostly to the *noise* of the device, not actual distortion. The 318 actually has less distortion than the oscillator, but since they cannot be separated, the composite curve becomes the new reference residual distortion level. The 518 was also evaluated in this circuit and yielded results comparable to the 318 in terms of speed and low distortion, but with slightly higher noise, as reflected in its curve. Either device is usable for the buffer function, but the 318 is preferred for its lower noise floor and thus greater potential resolution.

As noted in the conditions, I used no filtering in the THD measurements, due to the 100kHz range. Quite often wideband noise limited the ultimate accuracy of THD measurements at low frequencies, but for the sake of consistency all measurements are wideband, even in view of this apparent sacrifice. Ultimate THD below 1kHz is hardly likely to be SID, so this is not a serious compromise.

The importance to SID detection of the extended range (1kHz through 100kHz) cannot be over-emphasized. This measurement *must* be done to lend any validity to conclusions. Single frequency spot THD measurements, which have been reported as "uncorrelated," are a gross oversight at best, and are simply not comprehensible as an objective analysis. The goal is to paint a picture of rate-sensitive distortion, and extended range THD is the means.

As you can note from the Fig.I-4 U.U.T. test circuit, we use the inverting mode at unity gain unless otherwise specified, and at the device's full rated (±10V) output. The use of inverting

FIG I-6



mode eliminates contamination of measurements due to common mode distortion which appears in the non-inverting connection. The unity gain compensation gives the worst possible case for slew rate, the one most likely to generate SID.

One might argue that the high feedback of the unity gain connection can mask the level of SID, but in practice this is not the case. As it turns out, if slewing distortion is present it becomes woefully apparent without great efforts at detection. In certain cases, higher stage gains are set up by reducing the feedback around the U.U.T. by means of Rx. Noise gains higher than that of two for the standard case are illustrated in the table.

Test circuit input and feedback resistors are 10k, to minimize the loading of both the buffer and the U.U.T. If additional distortion were to be generated due to U.U.T. output stage non-linearity or loading effects, it would be difficult to separate from SID. Thus minimal loading is used, to maximize sensitivity to SID (only) detection.

I am certainly aware that both the inverting-only connection and the minimal loading stipulations for this test are somewhat unrealistic, in a practical sense. However, as the two forms of distortion which we eliminate by these are to be covered in a separate study, I believe these procedures are much more definitive in isolating SID. Indeed, to test otherwise it would be difficult (if not impossible) to positively identify the distortion source(s).

FIG.I-5

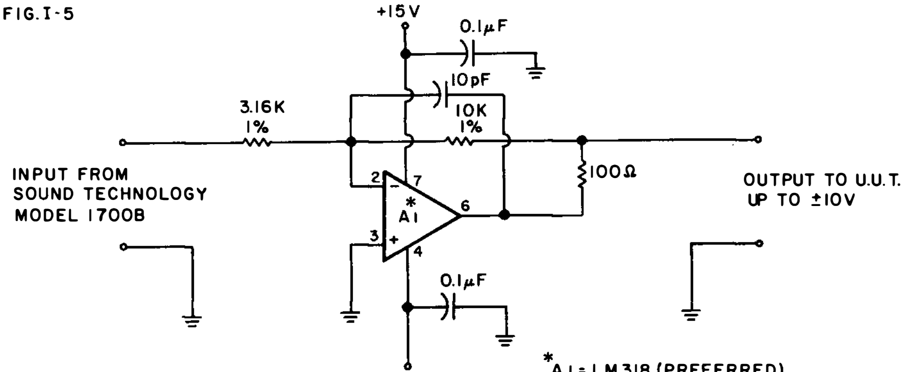


Fig.I-5: Buffer amplifier.

Test Results

A good way to begin discussing test results is to examine slew induced THD at various levels in a popular IC type (the 741). See Fig.I-7.

Since slew rate is a measure of the output rate of change, it follows that either frequency or peak amplitude can be used as the variable (with the other held constant) to examine its nature. Fig.I-7 shows the peak amplitude and frequency relationship and the resultant THD, for a 0.5V/μS slew rate device.

As might be expected, the 7V RMS (or ±10V) output level curve shows the worst distortion, as it occurs lowest in frequency. At lower levels of 2V and 1V RMS the distortion curve retains the same general shape, but is pushed upward in frequency. This curve shape is a classic one, and also one which we can immediately recognize as resulting from SID. It will recur repeatedly throughout this study, for almost every device examined.

The curves are so similar in shape that you can predict the 1% THD intercept point from the ratio of amplitudes, almost exactly. For instance, the 7V RMS curve crosses at 8kHz, the 2V RMS curve at 25kHz. These ratios (7/2 and 25/8) are nearly the same. This also holds true for the 2V and 1V RMS curves, as well as many others, as we will see in due course.

It might be argued that reducing the output level of a low slew rate device (such as the 741) will allow it to be used in audio circuits, such as for example the 1V RMS curve (a typical line level). However, the device can still generate serious distortion on signal peaks at high frequencies, as is evident by the 2V curve and will be further demonstrated by the IM tests. Bear in mind also that this is a unity gain condition: higher gain (more practical) circuits blacken the picture much more severely.

Fig.I-8 shows the variability of slew rate and resultant THD for samples from three different 741 vendors. The slew rates are 0.5, 0.8, and 1V/μS; the resulting curves are similar to those of Fig.I-7 in general, although in this case all are taken at ±10V out. We must expect such variability in IC op amps.

A dramatic demonstration of how SID can limit audio frequency performance is contained in the data of Fig.I-9. These curves represent measured performance of the 0.5V/μS slew rate 741 sample, which is a device close to the "typical" slew rate spec. These data demonstrate just how such a device would perform in typical higher gain circuitry.

The lower of the three curves (#1) shows the same data as contained in Fig. I-7 (for 7V RMS), repeated here for reference. The second curve is for a noise gain of 12, while the third curve is for a noise gain of 102.

While curve #1 is certainly poor performance if considered alone, it is "good," relatively speaking. The op amp feedback mechanism here is attempting to reduce the strong third harmonic being generated due to slewing.

Curve ##2 and 3, which are taken with progressively less feedback, therefore show much higher distortion, due to less correction. These two curves, particularly #3, show performance which can hardly be considered adequate for any audio use of reasonable quality.

Note the data are only plotted up to 8kHz; distortion would be even worse above 10kHz. Curve #3 appears to indicate a decrease in distortion above 5kHz. This is misleading, since the distortion is still being generated, but the harmonic product percentage is reduced by the rolloff of the amplifier, due to its 3dB bandwidth of 10kHz for this feedback condition. In reality distortion is even worse than is evident here.

While this example is rather gross in terms of performance, it is not at all uncommon to see 741 (or comparable slew rate capability devices) specified for audio use. This should only be done if the output signal levels are maintained to *very* small peak levels perhaps a factor of 10 (or more) below the conditions of Fig.I-9 and if the required bandwidth(s) are narrow and/or the gains low (high feedback factors).

FIG. I-7

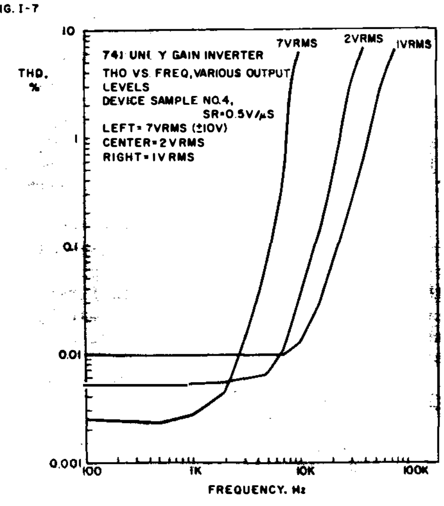
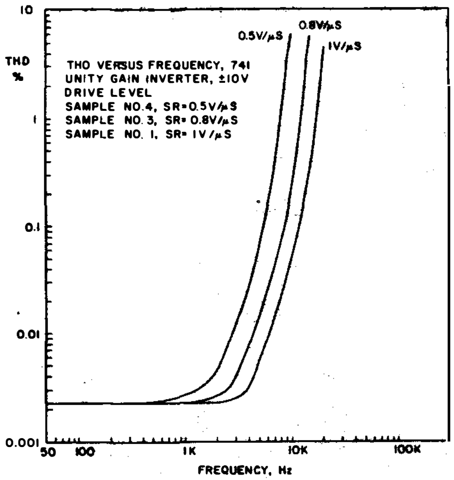


FIG. I-8



In short, it would be much safer (and wiser) to relegate the 741 and other low slew rate (<1V/μS) devices to voice grade and other low accuracy, poor fidelity applications where its shortcomings are less of a hazard, and true op amp predictability is not a prerequisite.

One method of improving an op amp's slewing capability (and consequently its final distortion parameters) is to *custom compensate* it for the exact working gain to be used. As should be painfully obvious, not only does a low slew rate in an internally compensated unit create large distortions at low gains, the situation becomes progressively worse at higher stage gains, as shown by the 741 example.

It may come as a surprise to some, but slewing rate restrictions in an amplifier compensated for unity gain conditions can be such an overriding limitation that in many cases the net situation can be improved by operating at a *higher* gain (and higher slew rate), with *less* overall feedback. This fact is demonstrated using three different IC amplifiers as examples, beginning with Fig.I-10. The data derived from extended range THD tests on these amplifiers also illustrate several other factors of distortion performance. The implications are highly inportant for drawing correct conclusions in a given situation as to the nature of distortion observed.

The first example is the 709, a noto--riously slow amplifier with unity gain compensation, as its slew rate is only 0.2V/ S. This is even slower than a 741, and the 0.2V/μS (x1 comp.) curve (A) of Fig.I-10 demonstrates just how severe the slew induced distortion is at only a few kHz. However, being an externally compensated unit, the 709 can be adjusted for optimum compensation, that is, compensation to match the actual working gain.

With x10 compensation, curve B, the device slews at about 2V/μS, and the beginning of the almost vertical slope of this curve (which indicates severe SID) is pushed out to about 20kHz. Note the absence of an obvious rise in low frequency distortion, as the real LF distortion below 1kHz is masked by noise. This curve's slope from 1kHz to 20kHz is more gradual, compared to the slew limited area above. In this region the feedback loop is attempting to correct other distortions in the device, most of which is the crossover distortion due to its class B output stage. The gentle upward slope is due to the open loop bandwidth rolloff of the 709.

Curve C, x100 compensation and noise gain, illustrates performance which is not dominated by slew limiting at all, but reflects further the open loop bandwidth rolloff, and higher rise in distortion due to progressively less feedback.

My point is that although the feedback has been reduced by a factor of 100 from the first to third curves, we see no corresponding x100 increase in distortion. To the contrary, distortion is actually reduced due to slewing improvements over most of the range, and at low frequencies where slewing is not a factor, rise is less than a factor of three-to-one in degradation.

An amplifier such as the 709, if skillfully applied, can be an effective performer due to its excellent maintenance of gain-bandwidth with differing compensations. Unfortunately, the 709's class B output stage generates quite a bit of crossover distortion, particularly at high frequencies, and this can be one of its ultimate limitations. Newer devices free of this defect can be more effective.

The 301A can also be custom compensated, and in some ways is more attractive, as it only requires a single compensation capacitor where the 709 needs three components. You can go only so far with "opening up" the bandwidth, however, because the unity gain capacitor is only 30pF. Generally speaking, the larger this capacitor, the more you can do with it by way of reducing its value, thus increasing bandwidths and slew rate.

FIG. I-9

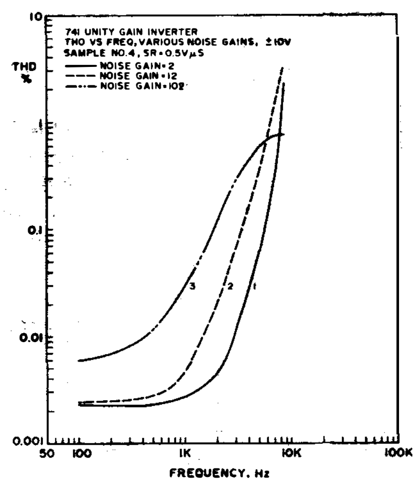


FIG. I-10

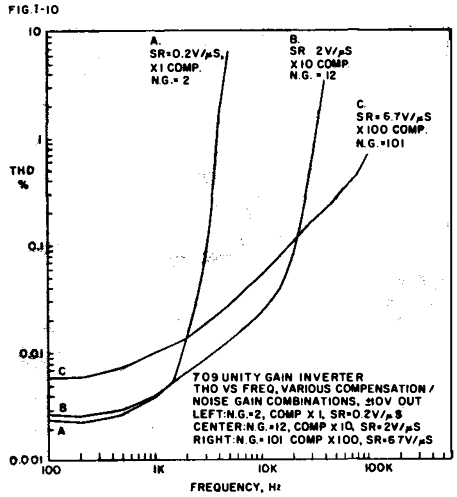
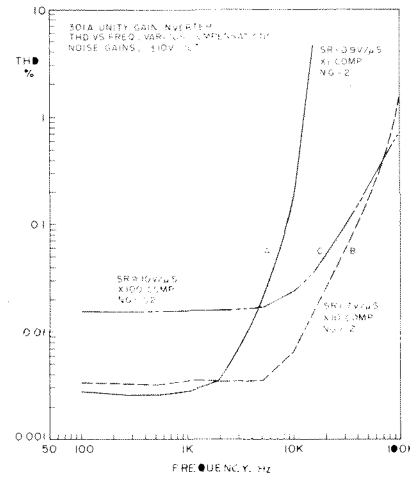


FIG. I-11



With the 301A sample plotted in Fig.I-11, the unity gain slew rate was 0.9V/μS, which shows the characteristic slew rate limited rise beginning at a few kHz. With x10 compensation, curve B, this effect is pushed upward in frequency where it shows similar shape, the 7V/μS curve. Note the low frequency distortion is still quite low, and there is no rise at middle frequencies as with the 709. Curve C, x100 compensation, is not a stock trick for the 301A; it was done in this case by using a 1pF or so twisted hookup wire "gimmick" (four turns of insulated #22 about ½" long) wired between pins 1 and 8. Also, pin 5 of the IC must be snipped off at the body to reduce its positive feedback, which will otherwise limit stability.

These data show somewhat higher distortion at low frequencies (actually noise) but without gross slew limiting, just a bandwidth rolloff increase in distortion. It is apparently greater than 6dB/octave at HF, due to lack of feedback to suppress the remaining slew induced distortion. For this device, both the x10 and x100 compensation performance are reasonable for modest accuracy applications, certainly a far cry from a 741 or 709!

The NE540 power driver amp excels at high gains because of an inherently high slew rate and bandwidth. (Don't confuse it with the *AD540*, a different device I discuss later.) This unit is an excellent example of an IC with good audio characteristics, and also aptly illustrates the effectiveness of custom compensation. See Fig.I-12.

Compensated for a x10 gain, the NE540 has a slew rate of 4V/μS. Its performance for this condition is not too spectacular (but not totally unreasonable either) and it becomes slew limited at 40kHz. For the x100 compensation, however, the slew rate is more than 20V/μS and, as the curve shows, there is no slew limiting whatsoever, just a bandwidth rolloff related distortion rise above 10kHz. The NE540 can also operate *uncompensated*, at a gain of 60dB. At this level distortion is of course easily measurable, about 0.15% below 10kHz. Still, this is very respectable performance in view of the high gain, and the device will slew at 100V/µS for this connection.

To give further insight and understanding of the interaction of slewing and gain rolloff related distortion, Fig.I-13 is an open loop plot of the NE-540's gain and THD. At low frequencies, the open loop gain is 93dB, and the open loop bandwidth is slightly more than 10kHz. Open loop distortion is quite flat, at about 3.3%. Above 5kHz gain rolloff invalidates measurements.

If we compare these findings to those in Fig.I-12, the bandwidth rolloff at 10kHz in Fig.I-13 corresponds to the upturn in THD at the same frequency, in Fig.I-12's gain of 1000 curve. Further, the distortion measures 0.15% which is a ratio of 22 with regard to the open loop distortion of 3.3%. With feedback, the NE540's open loop gain drops 6dB (to 87dB), due to the loading effect of the 10K feedback resistor on the device's open loop 10K output impedance. Thus the actual feedback is 87 - 60dB, or 27dB, a ratio which is almost exactly equal to the measured distortion reduction.

This point may seem belabored to some, but I stress it here because in so many instances you *cannot* predict closed loop distortion as readily, as for instance in the gross effects caused by slew limiting, which can confuse the issue to the point of frustration.

From this information you could also extrapolate the true THD at low gains; at the x100 compensation the measured 0.015% also agrees well. The x10 compensation does not precisely agree, as it predicts a distortion of 0.0018% and 0.003% was measured. The minor difference is due to noise and measurement resolution limitations.

An effective distortion reduction technique with an amplifier whose bandwidth can be "opened up" (such as the NE540) is what is popularly called "input compensation." This is nothing more than an RC network across the amplifier input terminals which forces the feedback loop to a high gain level, *at high frequencies only*. It does *not* (as some authors have implied) "roll off" the input signal, as the network is applied differentially, between the (+) and (-) terminals.

In simple terms, this technique allows a very high loop gain at low frequencies, and a correspondingly high slew rate. The slew rate is, in fact, one which would accompany the compensation appropriate for gain level of the network, such as 40dB.

To illustrate the effectiveness of this technique, I applied input compensation to the NE540 (with values chosen as outlined in my *IC Op Amp Cookbook*,[24] p.285). This is a x100 stage insofar as compensation goes, but *unity gain* for signals. Therefore at low frequencies almost the full open loop gain is available for feedback (minus only 6dB). Here the feedback would be 81dB, thus the distortion should be 0.0003% or 3ppM (parts per million). It measures a great deal higher, of course, due to the noise components generated by the large HF gain. Much of this noise is out of band, however, and therefore not audible (although measurable). If you examine the distortion curve for this operation you can see that it curves upward above 10k, the point where the feedback is diminishing, thus allowing distortion to rise.

FIG. I-12

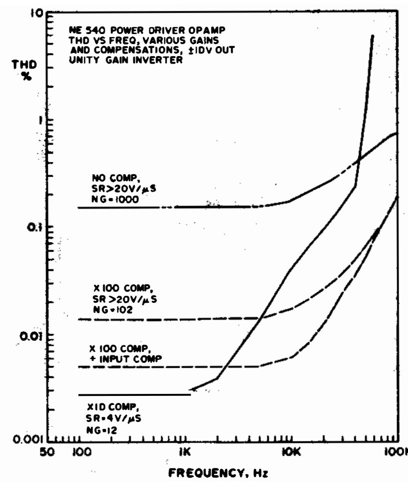


FIG. I-13

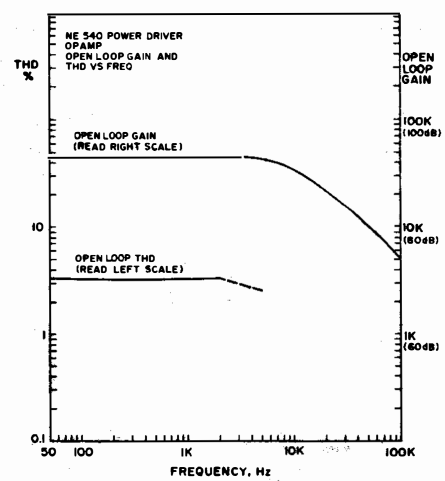
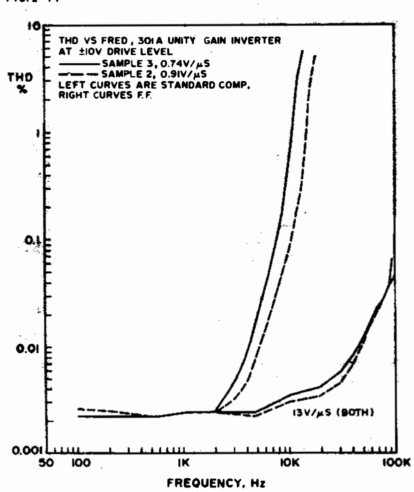


FIG. I-14



Input compensation is applicable to (and effective with) any adjustable compensation amplifier which can (or must be) compensated for high noise gains. This includes the devices just discussed as well as many others, and also some of the newer "decompensated" op amps. A decompensated op amp is simply one which is stable at some minimum gain *higher* than unity, such as x3, x5, x10, etc. Several examples of this type of op amp are shown in the data which follow, and

FIG. I-15a

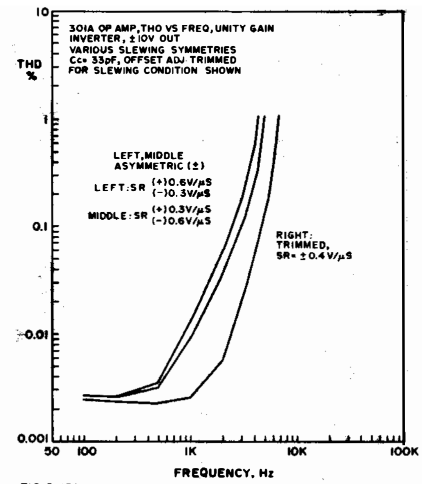
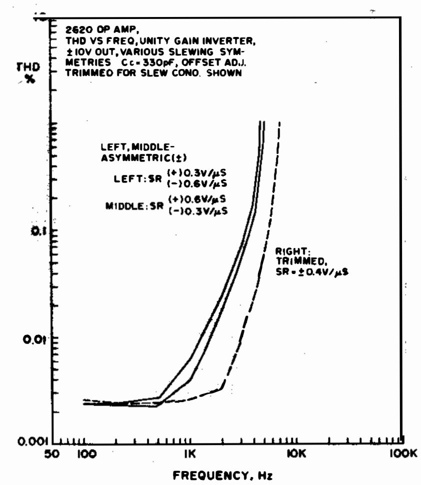


FIG. I-15b



they are generally capable of much higher slew rates.

As we saw with the 741 data, the range of slew rate variability is fairly wide (as is the consequent audio performance) among products of various manufacturers of a given device. Even a specific manufacturer's device will have tolerance variations in a single lot and more substantial variations between different "runs."

For the 301A, as compensated for unity gain, the vendor-to-vendor variation is illustrated by Fig.I-14. The two samples have slew rates of 0.7V/µS and 0.9V/µS. Again this generates the characteristic slew limited distortion curve, slightly separated due to these differences in slew rate.

However, the 301A is unique in that it can also be operated in a *feedforward* mode, in inverting (only) applications. This type of operation yields a much higher slew rate, specified as 10V/µS. The devices tested here both had slew rates (in the feedforward mode) of 13V/µS. This is sufficient to allow truly exceptional performance, as the THD is very low, being only slightly in excess of the source residual.

This class of performance is what we should all be seeking for high quality audio circuits. Use of the 301A for this operation is particularly attractive because of the circuit simplicity, and of course the unit's basic economy. It is highly suitable to summers, mixers, buf-

FIG I-16

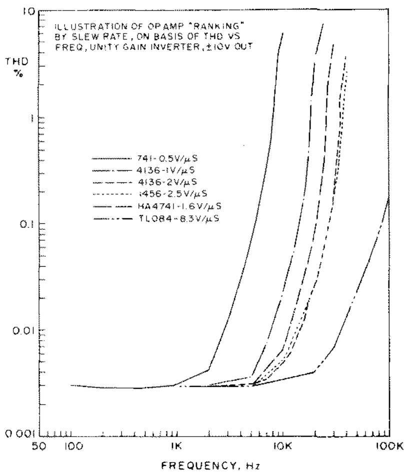


FIG I-17a

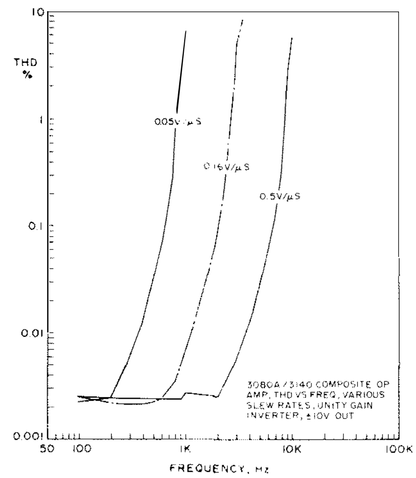
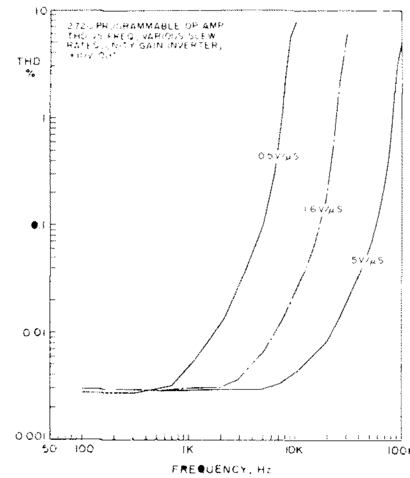


FIG I-17b



fers, multiple feedback filters, and other inverting type audio circuits.

The 301A serves well to demonstrate another aspect of the slewing induced distortion problem, in Fig.I-15. The degree of slewing distortion generated in a given op amp is closely related to the *symmetry* of the positive and negative slewing rates. Ideally, they should be as nearly identical as possible, to produce predictable results. Usually, most IC op amps are fairly good in this regard, but we cannot assume it to be a general rule for two reasons. First, manufacturers rarely specify slewing rate as more than a typical parameter, and they never specify it for symmetry. So you really have no control over it whatsoever, other than your own knowledge of a given manufacturer's product.

Should a device have a basic "built-in" asymmetry, it will never yield as low a distortion as even a slower device which possesses good symmetry. One prime example of this is the 356 op amp which slews with a 2/1 asymmetry (which can also vary). Although it is a reasonably fast device with a slew rate of 15V/μS, it is easily bettered by several slower devices (as will be shown).

To demonstrate the sensitivity to slew rate symmetry, I set up a 301A in a x1 compensated test circuit, but with a variable DC bias current injected into the input stage's current mirror, pins 1 and 5. It has been generally observed that op amps which employ a current mirror in the first-to-second stage interface (see Issue #1, 1977 series, p.6, Fig.E) can often be *trimmed* for slew rate, by altering the static DC bias in the current mirror sides.

If slewing is to be completely symmetrical the current gain of the mirror stage should be unity, so as to deliver equal charge and discharge currents to the compensation capacitor. With the 301A, for instance, the slew rate can not only be trimmed to a nominal (+) and (-) match, but symmetry can also be altered to a range of 2/1 or 1/2 (or even more).

Fig.I-15a shows the result of this in terms of THD. Note that the two asymmetrical slewing cases show a much higher distortion than the symmetrical case; at 2kHz, for instance, the difference is as much as an order of magnitude. Also the distortion breaks away from the residual noise level at a much lower frequency for asymmetrical slewing.

A similar test on a completely different op amp shows results which are remarkably similar. In Fig.I-15b a 2620 was compensated with a relatively high capacitance (330pF) to yield a basic slew rate identical to the 301A case, and symmetric/asymmetric data taken. The general character of performance is virtually identical to the 301A.

With either of these devices (and others which show this (±) slew trim capability) a distinct dip can be observed in THD as the slew rate is trimmed for symmetry. Generally, this can be accomplished via the device's offset trim network, but there may in some cases be minor differences. Note, however, that this method of trimming slew rate will not necessarily result in lowest DC offset; in fact, it will most likely *compromise* the DC offset.

Why the slew rate needs to be symmetrical for lowest distortion may be intuitively appreciated if the overall feedback loop is considered for signals in the slewing region. If slew rate is asymmetrical, the feedback loop (in attempting to maintain a constant output voltage) will force the input stage into an unequal (+) and (-) deviation from balance. An asymmetric slew rate may also be viewed as unequal positive and negative going gains, which result in unbalance. If we monitor the summing point with an oscilloscope during asymmetric slewing we can see that the error voltage will have a parabolic waveshape.

Asymmetric slewing produces even order harmonic distortion, whereas symmetric slewing produces odd harmonic products. Watch the distortion products during slew trim and you'll see the even order ones go to minimum as slewing becomes symmetrical. Symmetric slewing optimizes utilization of the input stage's transconductance curve, so as to generate the least objectionable (non-zero, but minimum) combination of distortion products as a result of slewing conditions.

Although slew trimming of an amplifier is impractical for general use, the point is the sensitivity of this parameter to symmetry. In evaluating an amplifier, pay close attention to the symmetry of the slewing, as well as the basic rate. This caution applies to *all* audio amplifiers (not just op amps) and the testing of this parameter should be a standard procedure for audio circuit evaluation.

It may have already occurred to the reader that, using these insights, amplifiers could be *ranked* according to their slew rates. To some extent this is true, but it is not a completely all-inclusive statement, due to other factors.

Generally, if the slew rate being considered is 2V/μS or less the statement about ranking is true. In this range, bandwidth is not a great factor at all, as demonstrated by Fig.I-16.

The first four curves here are for different devices, and their major separating factor is slew rate. The 0.5V/μS 741 curve is for reference; following this are a 1V/μS 4136 curve, a 1.6V/μS HA4741 curve (not to be confused with the MC4741), and a 2V/μS 4136. Clearly these devices at least can be ranked by slew rate alone as far as performance is concerned.

The 4136 samples are an interesting point; the 1V/μS unit is an original source unit (an RC4136) selected for a close-to-spec. slew rate. The 2V/μS unit is an XR4136, a second source improved 4136, a design which features significantly higher slew rates. The HA4741's spec. is 1.6V/μS so it is right on the money, as being typical. This is, incidentally, the same spec as the XR4136.

Another method of demonstrating how slew rate can completely dominate the distortion characteristic of an op amp is to generate a *family* of curves which show varying degrees of distortion versus frequency, for various slew rates using a single device, with fixed unity gain compensation. This effect is shown in Fig.I-17, actually the same sort of information for two different op amps, which have different slew rate capabilities, but can be made to overlap. The two devices chosen are programmable op amps, a device operational feature which allows slew rate to be selected by appropriate bias, and a family of THD vs. frequency curves generated (for each device). Input stages of both devices are undegenerated bipolar types.

Fig.I-17a shows three distortion curves for a 3080, set up for slew rates of 0.05, 0.15, and 0.5V/μS, respectively. For this test I selected a 3080A device from several for symmetrical slew rate, and used a 3140 unity gain follower to buffer the high impedance output node of the 3080A. (For details see my *IC Op Amp Cookbook*,[24] p.448.)

Interestingly enough, these three curves are, to all intents and purposes, *identical* in character, and are also precisely separated in frequency by the ratio of slew rates chosen, 10dB in this case. Since the 3080 is a linearly programmable device, one can generate an infinite number of these curves for it. Thus within the bounds of device performance, the designer may choose his limiting THD curve by adjusting slew rate.

The 2720 operates similarly but has a decade greater slew rate capability. Fig.I-17b traces its performance for adjusted slew rates of 0.5, 1.6, and 5V/µS. The curves are very similar to those of the 3080A, and they also occur at 10dB frequency separations, again in direct proportion to the slew rate(s). The two 0.5V/µS curves in I-17a and I-17b are decidedly similar, though measured from two different devices. Data from these two amplifiers for these conditions show that slew rate is the primary determining factor in producing distortion. It also indicates that only when a device has a slew rate of several V/µS does THD become very low (≤0.01%) across the whole audio bandwidth. The inescapable conclusion: slew rate is one of the major criteria of device performance for audio applications.

We may now assume that if slew rate is sufficiently high, distortion will be negligibly low, but this reasoning does not apply in every case. Fig.I-18 demonstrates that we must qualify a general statement of this nature in terms of specific device characteristics, as there are exceptions where distortion performance cannot be totally predicted from specified slew rate.

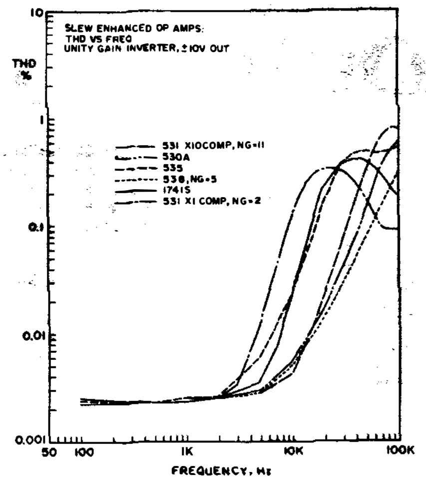
These data show similarly based distortion performance for "slew enhanced" op amps, typified by the 531, 1741S, 535, and 538. Such slew enhanced op amps utilize a class B input stage to dynamically increase the charging current of the amplifier's compensation capacitance. While this technique prevents total slew limiting and the resulting several percent distortion, it is not a panacea for low audio distortion. The class B mechanism, however, is only active at relatively high input levels; thus the *low level* SID of a slew enhanced op amp will be quite similar to an unenhanced device with a comparable (unboosted) slew rate.

You can observe this factor first in the data for the 531, which exhibit an initial distortion rise as the onset of complete slew limiting is approached. The device plotted here, for instance, appears to be similar to a conventional 1V/µS device in the shape of its initial distortion rise, i.e., the distortion below a 0.1% level.

Distortion rises with increasing output rates of change, until the class B current boost mechanism is actuated; then it levels off, even dropping somewhat. As frequency is increased further it remains substantially the same (but not constant). These behavior characteristics can be noted in the data of both the 531 and the 535, as well as the 1741S. At audio frequencies, the 535 and 1741S are nearly identical.

As a demonstration that distortion performance cannot be accurately predicted with a slew-enhanced device: the 531 has a 30V/µS slew rate specification while the 535 and 1741S specs. are 15V/µS and 12V/µS. Yet in terms of measured performance the 535 and 1741S are both appreciably better. The reason? Their

FIG. I-18



basic (unenhanced) slew rate is faster than the 531 (at least for the samples tested here) and this parameter determines distortion performance below the class B trigger level.

Fig.I-18 also plots the 538, a decompensated version of the 535, suitable for working gains of five or more. Since slew rate can be improved by the factor of compensation reduction, the 538 also features a x5 greater basic slew rate. As a result it will exhibit a low distortion characteristic to higher frequencies than will a 535, evident in the 535's curve.

The increase in frequency for a given distortion level is not exactly five times (as might be initially assumed), because the device must be used in a x5 gain condition which results, of course, in less feedback, and subsequently a somewhat higher measured distortion. Nevertheless the 538 is a substantial improvement over the 535, and is therefore a more attractive device.

An interesting proof of the validity of the higher (unenhanced) slew rate's value is shown in the performance of the 531 for a x10 compensation. This curve (Fig.I-18) is moved outward in frequency due to the smaller compensation capacitance and thus higher basic slew rate. This re-emphasizes the points made above in the discussion of variable compensation devices.

The 530A is one of the better performing devices shown in Fig.I-18, a more recent design which features improved input stage linearity and greater speed. Its performance is on a par with the x10 compensated 531 or the decompensated 538, but the 530A is a unity gain stable device.

As a group, slew enhanced devices can be effective in reducing slewing induced distortion, but on an absolute basis not as much as devices which increase slew rate by more direct means. The class B mechanism, in itself a non-linear effect, can (potentially at least) be a source of additional distortion. The most attractive devices in this group are the decompensated 538 which is reasonably clean at gains of five, the 531 (an externally compensated unit) which can be custom compensated for high stage gains, and the 530A, a faster, more linear unity gain stable device. Any of these units will be much cleaner if applied following the basic ground rules which minimize the approach of slew limiting.

The IM testing data which follow will give further insight into this conclusion, which is also supported by the outcome of the IM tests. However, while performance of unity gain slew enhanced devices can be exceeded by many other devices which are basically faster, keep in mind that even a slew enhanced device such as a x1 compensated 531 or a 535 is better than a 741 (or other <1V/µS slew rate device), since they avoid gross slew limiting.

Decreasing the transconductance of the input stage is a highly effective means of increasing a device's basic slew rate. A number of circuit techniques will accomplish this goal, but two of them yield outstanding measurable results in audio application.

The most direct method substitutes lower transconductance input stage devices, for example P channel JFET's or PMOS units, operating differentially and drain loaded. A bit later we'll discuss the performance of three designs which use this approach.

Another technique, which can be employed in virtually any op amp, uses emitter degeneration resistors. Designers have employed it in a large number of devices, the 318 and 518 being the most outstanding (fastest) examples. Emitter degeneration in a bipolar amplifier has drawbacks, however. The increased emitter current requires buffering to maintain a low DC input current, thus Darlington pairs are often used when it is applied. For audio, this has the serious drawback of adding additional undesirable noise sources. And emitter degeneration resistors themselves also add noise, so an emitter degenerated amplifier is by nature noisier than one which is not. In audio this type of device is therefore used only in higher level circuitry (preferably).

In op amp design history, one early use of emitter degeneration was "first generation" FET input IC op amps as typified by the 740, 8007, AD540, and NE536. This class of device typically uses a P channel JFET pair operated as source followers, which in turn drive an emitter degenerated bipolar PNP differential pair. Slewing rates achieved by this design are approximately 6V/µS, which is an order of magnitude better than "741" type performance.

Although the bandwidth of these amplifiers is the same as that of a 741, or 1MHz typical, the slewing rate improvement alone is sufficient to allow dramatic improvement in distortion performance. This is a point which should be underscored: an improvement in slew rate in a case of slew limiting can effect an improvement far beyond that of an equivalent bandwidth improvement, were slew rate to be held constant.

This is clearly evident in the THD vs. frequency performance of this class of devices, when compared with 741 or other similar slew rate units. Fig.I-19 plots the 8007, the AD540 (not to be confused with the NE540, discussed above), and the NE536. The specified typical slew rate for all these units is 6V/µS, but production variations yield samples above and below this mean. The performance of these samples could again almost be ranked by slew rate, at least at high frequencies.

Both the 8007 and the AD540 exhibit the classic slew limited THD curve, but the NE536, due to its higher slew rate, is stopped short of tracing out the classic pattern at 100kHz (due to equipment limits).

In the case of the NE536, the HF distortion is sufficiently low in this instance to warrant comparison with the equipment residual; and in fact near the 100kHz it is only slightly above the residual THD curve. At lower frequencies we see an anomaly, however, as both the 8007 and the AD540 have lower THD in the 2K to 20K range than the NE536. The differences are subtle, but measurable, and at this writing cannot be accounted for.

In perspective, any of these three amplifiers offers excellent performance, with THD below 0.01% to frequencies even above 20kHz. An amplifier which can achieve this kind of performance is in general quite good, and can be recommended without serious reservations, at least insofar as slew induced distortion is concerned. Again, this conclusion will be further supported by the IM test results to follow.

More recently, designers have introduced FET input amplifiers which use low transconductance FET input *differential pairs* to achieve higher slew rates. These may be termed "second generation" FET input op amps, as their design objectives seek to conquer the drawbacks of first generation designs. However, their overall audio performance is something of a mixed blessing. The three basic design approaches are the 356-357 series, which uses a P channel JFET input stage; the 3140, which uses a PMOS input stage; and the TL084, which uses a P channel JFET input/remainder bipolar type design.

The 356 has a 5MHz bandwidth and a 15V/µS slew rate, while the 357, a decompensated (minimum gain x5) version, has a 25MHz bandwidth and a 75V/µS slew rate. The 3140 has 4.5MHz bandwidth and a 9V/µS slew rate spec. The TL084 (a quad unit) has a 3MHz bandwidth and a 7V/µS slew rate. Although all these amplifiers appear to be eminently suitable for audio, in practice they are not and we must add qualifiers in at least two cases.

The 356/357 design has an inherent slew rate asymmetry built into it, slewing faster on negative slopes than positive. This gives rise to higher than "ideal case" distortion (for an equivalent slew rate, symmetrical amplifier). Easily measurable, at least in the case of the 356, this flaw causes an otherwise reasonably good amplifier to achieve less than its full potential. The 357, because it slews much faster, moves this "higher than ideal" distortion upward in frequency. While this minimizes the asymmetry defect, the 357 is also limited to gains of five or more, which makes it not quite as effective as the 356 for general purpose use.

The 3140 also shows asymmetrical slew, faster (+) than (-). The (-) slew rate is limited by a speed problem in the output stage which will produce a sudden 1% or more rise in THD at 30-40kHz, while very low below these frequencies. This problem is *not inherent* in the 3140's voltage amplification stages, however, and the output stage can be "fooled" by forcing a class A pulldown current to the V- rail, which removes the restriction and makes the slew rate nominally symmetrical. Operated in this way the data indicate it is one of the higher performance devices tested for SID.

In Fig.I-20 the reader can compare these factors on a common basis. The symmetrically slewing 3140 outperforms the 356, although the 356 has a higher slew rate. The 3140 is actually fairly close to the equipment residual over this range.

The TL084 is free of the problems of slew asymmetry; in fact, on the sample plotted none was measurable. The TL084 produced further evidence of the validity of the slew symmetry criterion, yielding higher performance than the faster 356. It even slightly betters the 3140 in the 10-30kHz region. It did appear to have a higher residual noise level at lower frequencies, however, which may limit its use for low level stages. From other than the noise standpoint, the TL084 appears to be the best of the currently available general purpose quad amplifier types, as it conquers the major limitation of slew rate and has a higher than average bandwidth.

We can make some further comparison with the first generation FET amplifiers, for instance the 3140 and the NE536, which slewed at about the same speed. The 3140 shows lower distortion, which would follow, as it has wider bandwidth and thus more corrective feedback. This would follow logically, since given two amplifiers with similar (symmetrical) slew rates, the wider bandwidth device will perform better, if we are comparing two devices whose condition is not limited. If we compare the TL084 with the NE536, we see the TL084 has less distortion in the 20-30kHz region since it has higher feedback than the NE536. Further up the frequency scale, where the TL084 begins to slew limit, the NE536 has the edge due to its faster slew rate.

The "dielectrically isolated" class of amplifiers perform quite well both in slew rate and bandwidth. Dielectric isolation allows very high op amp speeds, generally an order of magnitude higher in both bandwidth and slew rate. Impressive performance results when designers combine this technique with a good symmetrical slew rate.

Fig.I-21 illustrates THD performance of two dielectrically isolated op amp devices, the 2620 and 2525. Both are externally compensated, high slew rate, wide bandwidth devices. The 2620, however, can be opened up to a higher bandwidth, and will have slightly greater feedback for a given gain setting than the 2525, as the performance data reveal.

Compensated for unity gain, the 2620 sample slewed at 5.7V/µS but, as the data show, has excellent THD performance in the audio region, even though slew limited at just below 100kHz. This underscores the value of the extra bandwidth (8MHz here), as the 1MHz AD540 slewing at 5.7V/µS does not do as well (see Fig.I-19).

With no compensation, the 2620 is stable at gains of five or more and, as the noise gain of five plot shows, THD which is still low, without apparent slew limiting. Indeed, this curve is only slightly above the residual of the Sound Technology 1700B, whose oscillator uses a x1 compensated 2620 type device. The distortion rise towards 100kHz is the approach of slew limiting.

The 2525 performance shows that slew rate alone will not do the job completely; high feedback is also needed to achieve the lowest distortion. This device slews much faster than the 2620 (over 60V/µS measured) but has somewhat less feedback for the same gain setting. It also appears to have slightly higher open loop distortion, which may be a limiting factor.

Perhaps now the reader can begin to appreciate that we are speaking of differences in superlatives--either device for any condition shown is quite good, and will yield excellent low distortion performance.

A most interesting general class of op

FIG. I-19

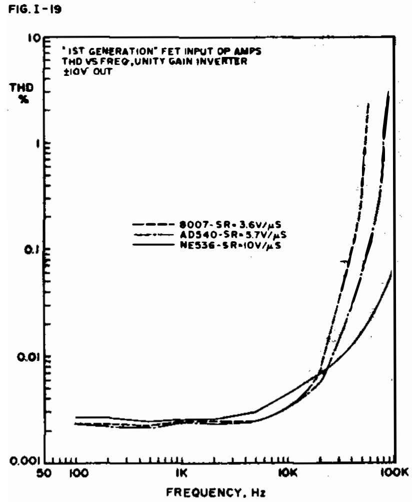


FIG. I-20

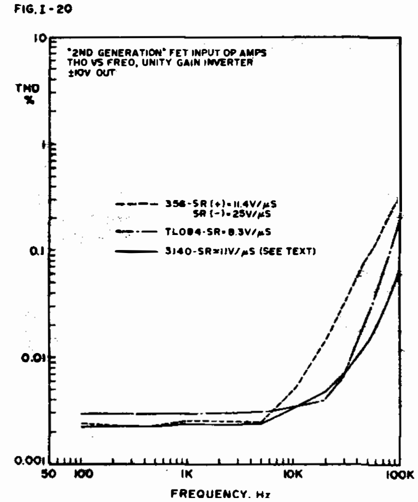
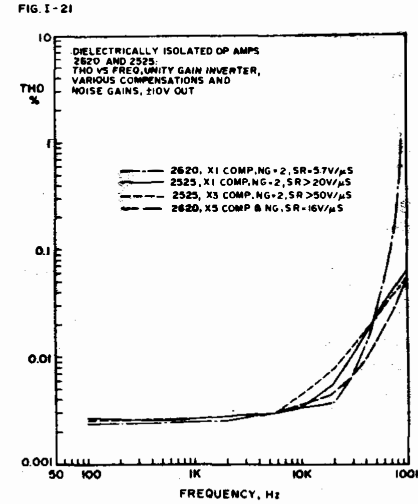


FIG. I-21



amps use *feedforward* configurations to achieve high bandwidths and slew rates. This technique usually bypasses the frequency response limitations of the PNP transistors, effectively extending audio performance, as shown in Fig.I-22.

We have already examined the 301A in a feedforward mode and found it one of the superior performers. Like the 301A, the OP-01 also operates in a feedforward mode. This unit slews at about 20V/µS with good symmetry. Its performance data (Fig.I-22) are somewhat disconcerting, as they indicate a *lower* THD than the source in the HF region. This was double-checked, but really did measure as shown. There is little doubt that the device is a good performer, but no ready explanation can be offered for the lower-than-source THD above 50kHz. One possibility is near equal amplitude but *out of phase* distortion components, which could cause cancellation of input THD.

The 301A and OP-01 must both be used inverting only to achieve high speeds. However, some designs use a differential input stage, followed by a feedforward second stage, notably the 318 and the TDA1034.

We have already discussed the 318, which is our reference amplifier. The TDA1034 is a new device, unique both in that it was specifically designed for audio applications and in its combination of performance characteristics.

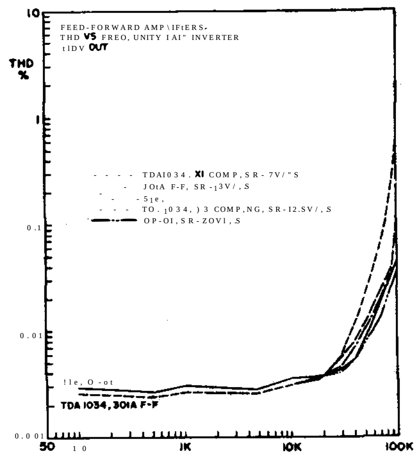
With unity gain compensation the TDA-1034 slews at 7V/µS (Fig.I-22). This causes slew limiting to be approached at 100kHz, but it still performs excellently in the audio range. With a minimum compensation of three, its performance very closely approaches the residual curve, with a slew rate of about 13V/µS. Not only does the TDA1034 possess high bandwidth and slew rate, but it also has a high current output stage, and an extremely low-noise input stage. It is the only IC op amp device which can be generally recommended for any audio use with no qualifications(at this writing.)

While other devices are higher speed (at unity gain), the differences are subtle and not at all overriding for audio range performance. It serves not only as an example of a good audio design, but also as a testimonial that good audio performance characteristics are indeed possible on an IC chip. This is one instance where the U.S., with its generally commanding superiority in IC design, has been upstaged by European technology. U.S. firms doubtless will soon be second-sourcing the TDA1034 design. It has all the marks of a winner.

Fig.I-23 is our final note on the overriding effects of slew rate on audio distortion. I used the 318, our *highest* performance device (the buffer amp circuit) because it is free from SID. However, *any* amplifier can also slew limit, due to capacitive loading (as in the case of the RIAA preamp example). Again the mechanism is the basic I/C relationship. Except for capacitive loading cases the I is the devices' output current limit, and C is the load capacitance.

For the 318 the (+) output current is about 20mA, for (-) it is about 40mA. With a C_L of 0.01µF this yields slew rates of about 2 and 4V/µS respectively. The result is a familiar slew limited THD curve, with 1% THD at about 35kHz. For a C_L of 0.1µF this curve is moved down by a factor of ten.

FIG.I-22



These data imply simply that slew rate is an ever-present potential source of distortion in any audio amplifier, even the best of them. It is a basic circuit phenomenon unique not only to IC op amps but to *any form of circuit technology*, regardless of feedback factors or other often discussed misconceptions as to its cause. In other words, if you go back to basics, it all makes sense (or should!).

This completes our discussion of the THD testing of IC op amps, with what I hope are some firm conclusions in the minds of the readers. The next phase of our work utilizes two-tone 1:1 HF IM tests, which follows as Phase II.

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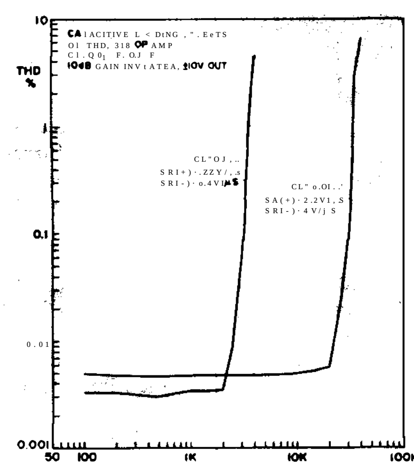
FIG.I-23



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Slewing Induced Distortion: Part 2b

Phase II: Two-tone Intermodulation Distortion Tests for SID

by Walt Jung
Contributing Editor

THE TWO TONE, 1:1 high frequency IM tests (hereafter referred to simply as IM tests) were performed on some, but not all, the IC op amps tested in the Phase I THD tests. One main objective at the outset was to determine whether or not a clearly demonstrable correlation exists between slew rate and two tone IM tests; and also between the IM and THD data.

Once this objective was met, I tested more selectively, generally those devices which had proved superior in THD tests, with data of certain units selected to illustrate the pattern. Overall the results correlate quite well, as devices which perform well in THD tests also do well in IM tests. I discovered many subtleties during the tests, however, which provide further information and food for thought.

Our IM test set was especially assembled for this study, and can most honestly be described as a breadboard. We spent considerable time, however, in ensuring that the measurements were reasonably valid and repeatable. A self-calibration feature was built into the test setup as a means of guaranteeing this, and a fair degree of effort expended with the lash-up in minimizing noise and spurious error sources. Although a breadboard, the setup does allow repeatable measurements, and consistent results. Therefore, while the test setup may lack some desirable features (mainly from the standpoint of convenience) it does yield results which are valid in my estimation.

The test setup consists of the equipment shown in Fig.II-1. Some of this is standard lab equipment, but a fair portion was built specifically for the task. In basic concept, the two tone HF IM test linearly mixes two sine wave sources with a close frequency spacing, and applies the composite signal to the U.U.T. (unit under test). The U.U.T. output will contain (ideally) only the original two tones, and no intermodulation products. If IM is present in the test device, it will appear as sum and difference frequencies. In the case here we are interested in the difference frequency only, which was maintained (for convenience, mostly) at 100Hz.

A low pass filter with sufficient HF rejection can be used to separate the original tones from the IM components, and the resultant IM measured rather simply with an AC voltmeter (suitably calibrated).

Factors critical to the success of this method are the mixing stage, which (if active) must not generate any IM of its own. The low pass filter must have a very sharp cutoff between the Δf frequency and the *lowest* two tone frequency used, otherwise the leakage of the tones through the filter will contaminate the readings as residual IM.

FIG. II-1

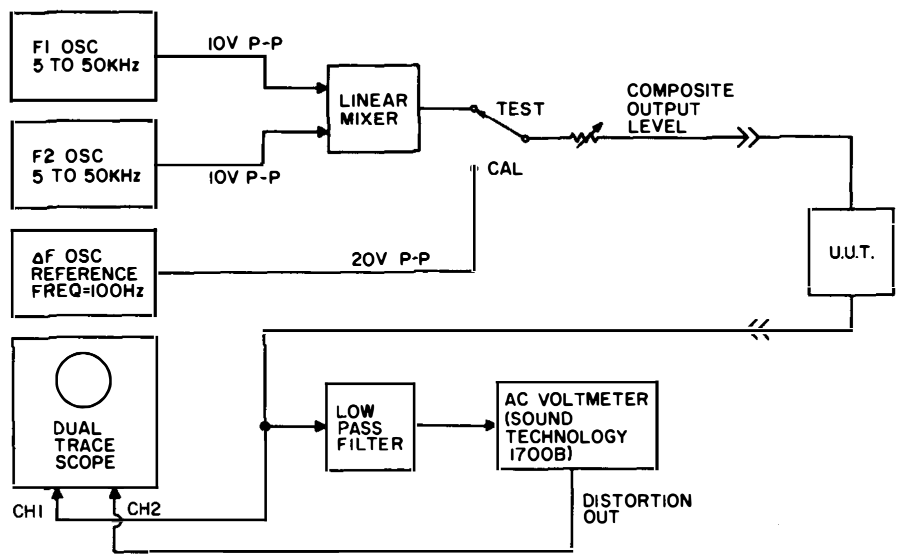
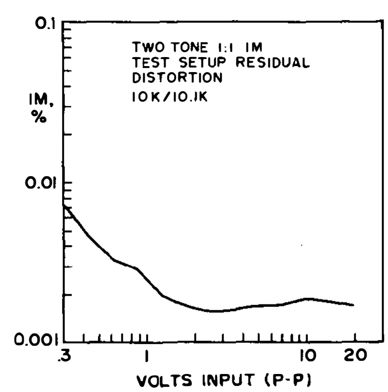


Fig.II-1: Block diagram , two-tone IM test setup.

In the test setup here, a four-pole Chebyshev filter design was used with a rejection of ≥80dB at 5kHz, allowing a basic limiting resolution of 0.01% at 5kHz and greater above. Shielding, screening, and appropriate grounding must also be used to eliminate hum and other noise components as contamination sources. Finally, both the f1 and f2 oscillators must be very clean in terms of hum and low frequency disturbances within the passband of the LP filter.

Before taking any readings, I did various tests to demonstrate a minimum of errors due to the above factors. In summary, our test set can be demonstrated as free of residual IM to the extent of at least 0.01% (or less), at levels as low as 100mV RMS at 10kHz (worst case). This factor is shown by the residual distortion plot of Fig.II-2.

FIG. II-2



Operation of the test set is relatively simple. The two tones f1 and f2 are manually set to the desired frequencies, and the resulting envelope adjusted (by trimming either f1 or f2) for the desired Δf repetition rate. Both f1 and f2 levels are 10V p-p. When mixed this yields a 20V p-p composite waveform. This can be scaled downward as desired, by the COMPOSITE OUTPUT LEVEL. The Δf reference oscillator is also set to a 20V p-p level, and the output can be switched between either Δf, or f1+f2.

For calibration, the Δf oscillator is fed through the U.U.T. and the low pass filter, which has unity gain at 100Hz. By monitoring the output of the U.U.T. with the scope, its drive is then adjusted for the desired p-p output @ the Δf frequency. Simultaneously, the AC voltmeter is adjusted for a 100% full scale reading. To read 80dB below 100mV RMS, a high resolution voltmeter is required; in the test set here the Sound Technology 1700B was used for metering and the f1 oscillator.

With a 100% reference level at 100Hz established through the U.U.T. and filters, the setup is calibrated and now ready for an IM tests. We accomplish this by switching to the f1-f2 source, and increasing the voltmeter sensitivity until we obtain a reading. At the same time, the DISTORTION OUT signal is monitored to observe the IM. The level of the IM is read in either dB or as % of full scale, with the 100% reference level being the 20V p-p (or lower, as adjusted) f1+f2 signal.

The U.U.T. is operated again here in a unity gain inverter with (unless otherwise noted) unity gain compensation. The test circuit is a duplicate of Fig.I-4 (page 9).

Test Results

As I mentioned, one of my first IM testing goals was to establish a correlation between slew rate and the resulting IM, as I did in the THD tests. Fig. II-3 is one test which served as the first indicator of this.

Here two previously tested op amps were selected for an *identical bandwidth* but with close to a *10/1 difference in* slew rate. One is a 741 which slewed at 0.5V/µS, the other the AD540 which slewed at 5.7V/µS. The results are quite dramatic, indicating that the higher slew rate device generates almost negligible amounts of IM at either 10 or 20 kHz, over the range of levels shown.

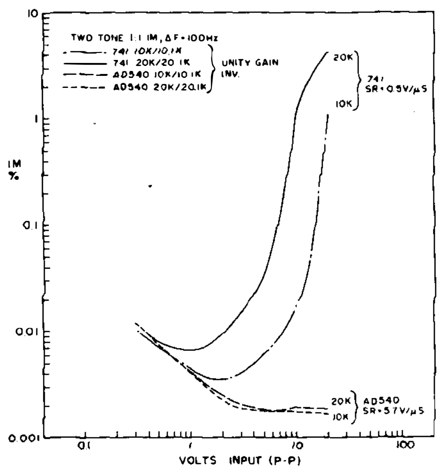
In distinct contrast, the 741 generates large amounts of IM at either 10 or 20kHz, particularly at high levels. The 20kHz curve is higher in level for a given frequency, or, viewed another way, for a given percentage of distortion the level difference between 20 and 10kHz is a factor of two. This same sort of pattern is generated in the THD tests, indicating a relationship to the output voltage rate of change, or slew rate. It is not surprising that the 741 performs poorly in this test, but what is interesting is the apparent close relation of IM percentage to the voltage rate of change as presented by the two tone IM signal.

Our next test was a similar sort at 20kHz, on a device with adjustable slew rate, the 2720. Fig.II-4 plots the results of this test for output levels of 100mV to 7V RMS (±10V) of output. At a low slew rate setting a great deal of IM is generated, but it decreases progressively for greater slew rates. These curves are apparently similar (or would be if complete), with equidistant spacings, but since the lower two cannot be traced completely to dynamic range limitations, a different sort of measurement perspective is needed, one which stresses the device's voltage rate-of-change tracking fidelity more completely.

Fig.II-5 shows the same device in a different form of IM test, full output level (±10V), with a frequency sweep of 5 to 50kHz. With this form of data the performance relationships become more evident. Here the 0.5V/µS and 1.6V/µS curves are virtually identical in shape and separated in frequency (at the 1% distortion level) by a factor consistent with the ratio of the slew rates. The 5V/µS curve appears to start on the same trajectory, but is limited by the upper sweep frequency limit of 50kHz. In the data range which is present, its form also resembles the 1.6V/µS curve.

Now we are able to make an extremely interesting observation, if we compare these data with the 2720's THD data (Fig.I-17b, p.13), as we notice the family of curves are quite similar in shape. If we take the 1.6V/µS curves as similar condition *examples*, we can see that not only are they similar in general shape, but the data points *almost coincide*, with the IM data being slightly less sensitive. With this information, the correlation seems to be established without major reservations.

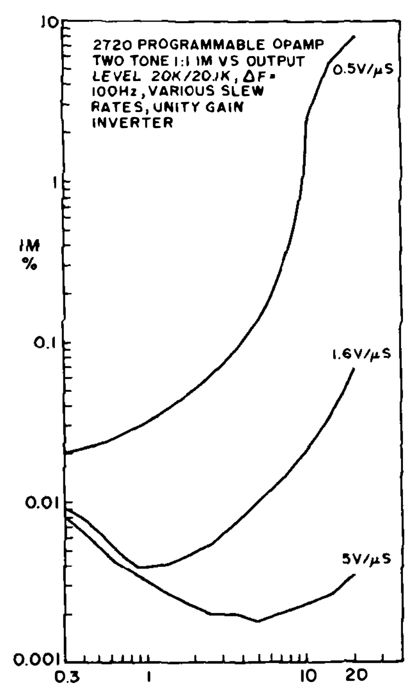
FIG.II-3



If the data do support the IM/THD correlation, is it possible for us to rank amplifiers by slew rate in terms of IM? In fact, we can do so, within reasonable limits, as Fig.II-6 shows. These data, which are also swept IM from 5k to 50k at full voltage output, begin with the 0.5V/µS 741 for reference. If it was not apparent from the slew rate data, perhaps these IM data will establish that a low slew rate device such as this is simply not adequate for high (or even moderate) quality use.

Next in performance comes the *4136 sample, which slews at 1V/µS. Although* it is a different device from the 741, the 1% IM frequency intercept is roughly twice that of a 741, or the ratio of slew rates. Lower down on this curve we can note that the 1V/µS 4136 curve is roughly similar to the 535 slew enhanced device, indicating that the 535 would possibly slew in the 1V/µS range were it unenhanced. In other respects, the 535 data cannot be compared directly or ranked by its slew rate.

FIG.II-4



Next in terms of performance is a 2V/µS 709 which is very closely paralleled but somewhat bettered by a 2.5V/µS 1456. Again (for the 709) the 1% intercept is four to five times that of the 741, roughly the ratio of slew rates.

Going further, we see the 356, the asymmetric slewing device. Close to it but definitely superior in this instance is a x100 comp 709, which slews at 6.7V/µS. Here is a case which clearly reiterates the necessity of symmetrical slewing, as the 709 is operating with 40dB less feedback, but generates less IM. An even better demonstration is the 8007 which slews at 3.6V/µS, but is far superior to the 356 in IM performance.

Between the 536 and the AD540 we have an anomaly as far as slew rate ranking is concerned, but as you may recall, the 536 behaved somewhat *peculiarly* in one regard in its THD list. The AD540 is a quite attractive unit, with less than 0.005% IM all the way out to 50kHz. Also interesting is the slew enhanced 531 @ x10 comp, which again emphasizes the value of high natural slew rate.

The lowest extremity of the curves is the residual distortion level of the test setup, over the 5k-50kHz range. Below 10kHz, as may be noted, the leakage through the filter becomes the limit of performance measurement for most of the devices, so comparisons in this range should be taken with a grain of salt.

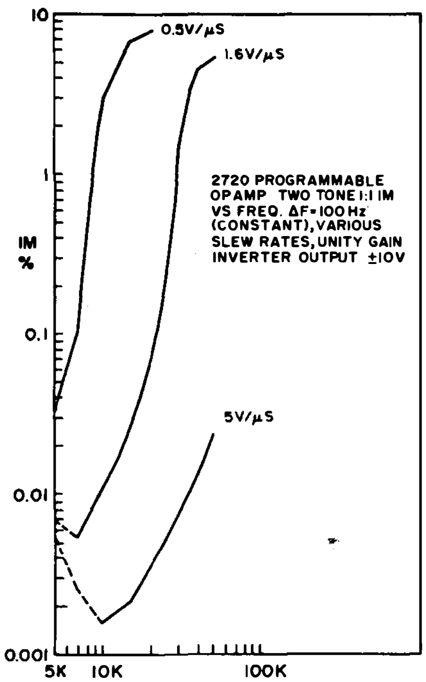
Above 10kHz, the residual is less than 0.002% and, as is noted, a great number of devices simply sat down in this area in terms of IM: namely, the 318, 1034, 2620, OP-01, 301A (FF), 2525, 3140, and TL084. However, THD tests separated most of these devices in terms of performance, although admittedly their differences are of a somewhat subtle nature. This indicates that of the two tests THD is the more strenuous and demanding of a device, and will serve as a more complete indicator of ultimate performance, at least between the types of tests as performed here.

The IM tests would perhaps be more complete and revealing if continued to 100kHz, and with resolution extended below 10kHz. I hope such equipment improvements can be made within the near future. However, the overall data indicate the THD tests to be more sensitive, and definite (similar) patterns emerge from both tests.

As I mentioned, performance of several devices could generally be predicted from slew rate alone. The TL084, for instance, was observed to slew very symmetrically, and at 7V/µS. With the device's 3MHz bandwidth, this is almost a sure guarantee of good IM performance, as in fact it did indeed turn out. I hazarded a similar observation on the OP-01 prior to tests, on the basis of its slew rate and bandwidth: my prediction was borne out by performance measurement.

Ideally, if a strong correlation is established between slew rate and THD/IM test results, it might be possible to predict qualitatively a given device's performance just by oscilloscope observation of its slewing behavior. This obviously simplifies testing and proves

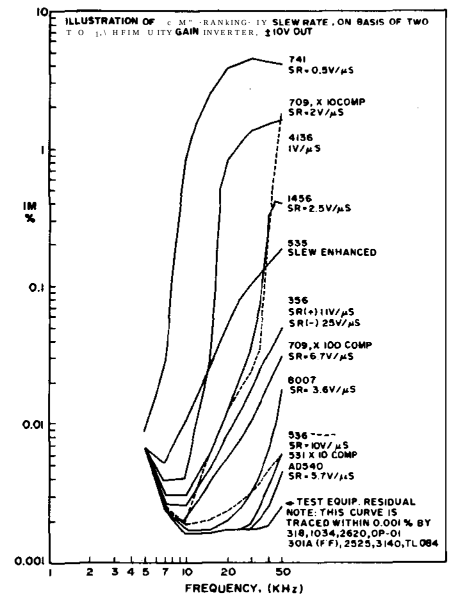
FIG. II-5



true on the TL084 and the OP-01.

Unfortunately, however, this procedural assumption could be dangerous for at least two reasons. One is the behavior

FIG. II-6



of slew enhanced devices which, as we saw, appear to be superlative in slew rate terms, but are less than superior in THD/IM tests. Reason two is simply one of prudence and/or conservativism.

It is unwise to make all-inclusive statements such as, "If it slews fast and symmetrically, it will measure well." While I observed this to be true in about 90% of the cases tested, I also saw exceptions. These may or may not be slew rate related, but since they are yet undefined they cannot be ignored.

In general, however, it can be stated that an amplifier which slews fast and symmetrically will be a good performer, and as such will eliminate the major sources of SID. Since this form of distortion has been shown to be overwhelmingly predominant in most IC op amps, it can hardly be argued that a fast and symmetrically slewing device is *not* a significant improvement.

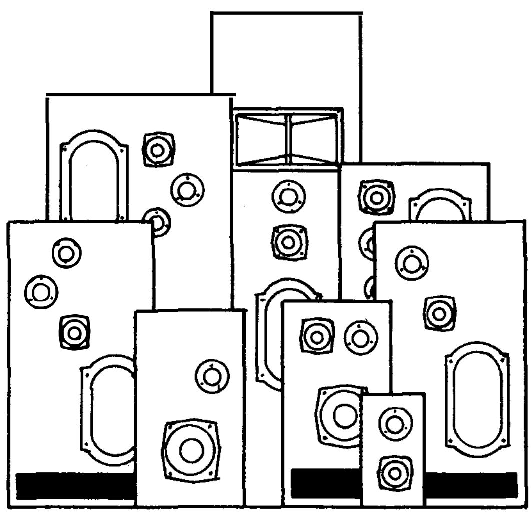
As a final argument for the rigor and completeness of the THD and IM tests, we found other forms of distortion in several devices, although they were admittedly small by comparison. A device could be entirely free of SID, for instance, and yet have non-linearities in its output stage. These would never show up by doing an analysis of slew rate and bandwidth, but could easily place a limit on the device's ultimate performance.

More than one example of such an IC op amp exists, and I measured some in this study but deleted them from the discussion for such reasons. But you'd never discover such anomalies without making the measurements, and this is but one way in which those of us searching for the truth can be fooled.

Next time we'll move to Phase III, testing for SID by the TIM method.

Please turn to page 36 for the references for this segment.

SPEAKER KITS



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Slewing Induced Distortion: Part 3

Phase III: Transient Intermodulation Tests for SID

by Walt Jung, Contributing Editor;
Mark L. Stephens, Signetics Corp.; and
Craig Todd, Dolby Labs.

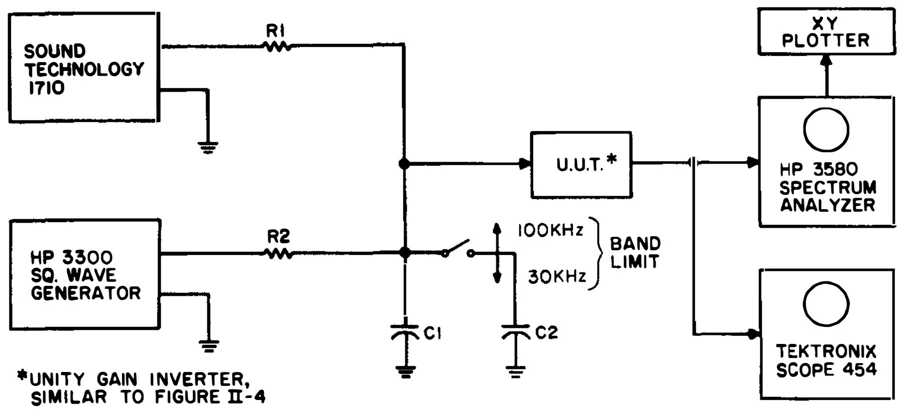
IN THIS PHASE of testing, the test method described in reference 17 was implemented to investigate transient intermodulation in various common op amp circuits. Our test set-up is shown in Fig.III-1. A Sound Technology 1710 was our low distortion sine wave source and a Hewlett-Packard 3300 provided the highly symmetrical square waves. The signals were passively mixed via R1 and R2 to prevent possible TIM generation in the applied signal source.

Capacitors C1 and C2 provided the necessary band limiting of the square wave to 30kHz and 100kHz (-3dB point of filter) depending on the switch position. The fundamental frequency of the square wave was 3.18kHz and the sine wave 15kHz. The square wave and sine wave were mixed in a 4:1 square-to-sine ratio, based on peak-to-peak readings on an oscilloscope. We used an HP 35B0 spectrum analyzer in conjunction with an X-Y plotter to measure and record the resultant output spectrum of the unit under test (U.U.T.), from 5Hz to 20kHz.

The peak-to-peak test level at the device's output was varied according to the capability of the U.U.T. The most severe test was 20Vpp amplitude, with 100kHz band limiting; the mildest, 2.5Vpp, with 30kHz limiting. To simplify testing we used only four levels: 20Vpp, 10Vpp, 5Vpp, and 2.5Vpp. The combination of these four test levels and two filter conditions yielded a total of eight possible input signal combinations, of varying slew rate intensity. The equivalent slew rate is calculated for each possible input combination in Table III-1. As you can see from the chart, the most severe input slew rate to the U.U.T. is 10.2V/μS, and the mildest 0.4V/μS.

Before we present the test data, a few comments about the test method are in order. The sine-square test is only one

FIG.III-1



*UNITY GAIN INVERTER, SIMILAR TO FIGURE II-4

of several ways of measuring transient intermodulation distortion. Other ways are high frequency harmonic distortion, two-tone intermodulation (above), and noise transfer techniques.[14] However, the sine-square method has the distinct advantage of producing the greatest slew rate stress on an amplifier using frequency components *within the audio band.* The other tests usually require out of band fundamentals to generate measurable amounts of distortion. The better the amplifier, the more important this distinction becomes in evaluating its performance.

Unfortunately, the sine-square test method has a serious problem, which became apparent after evaluating some of the best op amps. The problem concerns amplifier distortion products which are coincident with the even order distortion products of the square wave generator. Theoretically, a square wave should consist only of odd order harmonics of the fundamental frequency. Almost every generator has a very slight asymmetry in its square wave output which creates small but definitely measurable amounts of even-order distortion.

Table III-1

Resulting Slew Rates for Amplitude(s)/Band Limit(s)

	Square Wave Component		Sine Wave Component		Composite Output	
	Amplitude (Vpp)	SR* (V/μS)	Amplitude (Vpp)	SR** (V/μS)	Amplitude (Vpp)	SR (V/μS)
30kHz band limit	16	3.016	4	0.188	20	3.2
	8	1.508	2	0.094	10	1.6
	4	0.754	1	0.047	5	0.8
	2	0.377	0.5	0.0235	2.5	0.4
100kHz band limit	16	10.05	4	0.188	20	10.2
	8	5.025	2	0.094	10	5.1
	4	2.512	1	0.047	5	2.55
	2	1.256	0.5	0.0235	2.5	1.27

* SR square = $\frac{Vpp}{RC}$, RC = filter time constant (μS)

** SR sine = $2\pi(15kHz)\ Vpp/2 = 0.047\ Vpp$

Typical amounts for a general purpose square wave generator are 50 to 60dB down from the fundamental. Thus, if one measures a very good amplifier that has only small even-order distortion products falling on the square wave harmonics, the true distortion of such a device would be masked by the generator and therefore unmeasurable. An experimenter might then erroneously conclude that the amplifier is free from transient intermodulation distortion, whereas actually the amplifier is producing small amounts of distortion below the measurement threshold.

One might suggest that any amplifier producing distortion products coincident with the square wave harmonics should also produce other readily measurable intermodulation products of comparable magnitude. This simply is not the case and can be easily demonstrated, by testing a 356 or a 530A. Both these amplifiers show only even-order square wave products, even under the severest slew rate test.

To accurately measure these two devices, we require a square wave generator with even-order products down at least 90dB. In this test series, we obtained this result by carefully adjusting the symmetry of our square wave generator at periodic intervals. Only when we reduced the generator's even-order distortion did we begin to see differences between the best op amp circuits, that typically had only even-order distortion products.

The magnitude of these even-order products for the best circuits were from 0 dB to 6dB greater than the generator residuals, and in many cases required detailed comparison of the input and output spectra over several runs to verify that the products were in fact actually there. Figs.III-2 and III-3 show the input and output spectra for the 530A, which demonstrate the phenomenon.

The even-order harmonic problem with the sine-square test can produce what appears to be anomalous behavior in certain circuits. If a square wave generator with even-order products of -70 to -80dB is used, is it actually possible to see *less* even-order distortion on the output of a device than on the input? At first this seems ridiculous, but in fact it is quite understandable if you realize that the distortion components from the device under test can add (out of phase) to the generator components to reduce the output products. This happens frequently with good op amp circuits, if the generator products are on the same order as the device's distortion products.

In general, the sine-square test is valuable for measuring transient intermodulation distortion because it places the greatest slew rate stress on an amplifier, with in-band fundamentals. However, it is an extremely difficult test to implement since it requires an ultra-symmetrical square wave source that is stable with time and temperature. It is further flawed by the fact that most good amplifier circuits seem to show only even-order products that fall on the square wave harmonics, and these products are independent of the sine wave component. Thus one could use a square wave test only (see Holman[16]) and obtain the same distortion results. These difficulties place a limit on the usefulness of the sine-square test for isolating transient distortion.

Test Results

A number of op amps from our Phase I tests as well as others were evaluated with the sine-square test, and the results appear in the Table III-2 performance summary. All were hooked up as unity gain inverters except the decompensated units which had to be operated at higher gains, for stability reasons. These exceptions are noted under test conditions in the summary. The number specified for percentage DIM* was calculated by taking the magnitude of all distortion products, summing them by squares and then normalizing the square root of the sum to the magnitude of the sine wave component.

This normalization method seemed a little ridiculous for those units that showed only even-order distortion products falling on the square wave harmonics, since removing the sine wave entirely had no effect on the products. Why normalize distortion to a fundamental that frequently does not contribute measurably to the distortion products? In addition, the square wave contains most of the power and thus it seems more reasonable to normalize to the square wave fundamental. However, for the sake of consistency we adhered to the originally proposed definition.[17]

* *Otala uses DIM and TIM interchangably. Both refer to the same test and mechanism. DIM is "dynamic IM" as opposed to normal IM, which he thinks is static.*

CRAIG TODD has been active in amateur radio and audio since early high school. He received a B.S. in Physics from Caltech in 1970 and then went to work for the USC Geology Dept. designing electronic instrumentation. He became associated with the pro audio field when he designed and built a professional 12 bit digital audio delay line. In 1974 he joined Signetics Corp. where he worked on the LM381 preamp family of circuits and designed the NE570 Compandor chip. He is now with Dolby Laboratories in San Francisco.

Craig has written and presented several papers for AES and IEEE conferences, and is a member of both organizations.

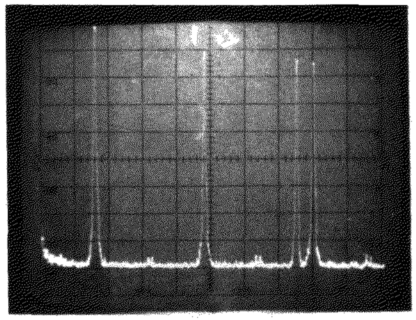


Fig.III-2: 530A, 20Vpp, 100kHz. Input spectrum as applied to U.U.T.

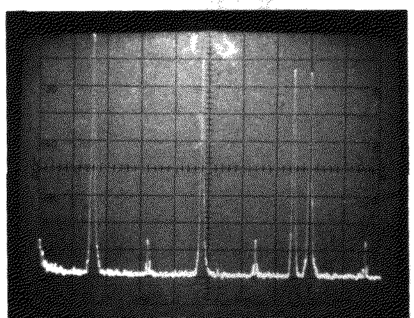


Fig.III-3: 530A, 20Vpp, 100kHz. Output spectrum.

Of all the devices tested, several had transient intermodulation components within 1dB of the residual distortion in the generators at the severest test condition, 20Vpp with 100kHz filtering. These were:

OP-01	NE536
HA2525	TDA1034 (@ x3 compensation)
TL084	LM318 (518)
	HA2620 (@ x5 compensation)

Since a small amount of noise or variation in the square wave generator harmonic components could create a 1dB difference between input and output spectra, we decided to use this as the resolution limit of the measurement system and characterize all the devices as having exceptional transient distortion performance. In Table III-2, some of these "select" devices are specified to have better percent DIM than the others. This is because these op amps exhibit extremely small differences, and the data reflect what we measured. We feel though, that these differences are within the resolution of the measurement system.

Although the Table III-2 data are fairly self-explanatory, some comments are appropriate. Devices which were tested in Phase I THD and Phase II two-tone IM tests are listed first, in the general order in which they appeared in Phase I. These are the identical devices so direct comparisons can be made. Some additional units available to co-authors Stephens and Todd are also listed; namely the 4558, and the 356 and 357 (at the end).

For the higher performance devices mentioned above we often performed only a single test, for example 20V, 100kHz. If a unit performed well at this level it was, of course, not necessary to test at reduced slew rates. Some devices, such as the TDA1034 and 2620, showed some TIM at this level with x1 compensation, but negligible amounts with reduced compensation. Others, such as the 301A, improved markedly if used with high SR compensation (such as feed-forward).

An even larger group of devices passed the 20V test at 30kHz, which is more realistic in terms of a practical audio bandwidth. In general, a device will show serious TIM when the signal slew rate exceeds the device slew rate. The details of this relationship are covered below.

Fig.III-4 clearly demonstrates the relationship between transient intermodulation distortion (DIM) and device slew rate capability. This graph shows percentage DIM vs. device slew rate for all types of devices under one standard test condition. The maximum slew rate of the input sine-square for this case is 3.2V/µS. Thus, the device would have to have a slew rate of at least this much to pass the waveform with unmeasurable distortion. The graph shows that distortion rises above the resolution level around 6.5V/µS, which is roughly twice the slew rate of the input waveform.

This indicates that "on the average" a device must have at least twice the slew rate of the input signals to pass them with negligible distortion. As the slew rate capability of the devices falls below 6.5V/µS the graph is seen to rise linearly to very high amounts of distortion. A "best" straight line drawn through the data points turns out to have a slope of 3:1 on the logarithmic coordinates. This indicates that DIM varies as the third power of the ratio of the input slew rate to the device slew rate. A simple equation expressing this relationship would be

$$\%\ \text{DIM} = K\left(\frac{\text{SR of signal (output)}}{\text{SR of device}}\right)^3$$

where K = 0.16% for our data.

This relationship is extremely impor-

MARK L. STEPHENS was born in Salt Lake City, Utah, in Nov. 29, 1947. As a teenager he experimented with vacuum tube circuits and subsequently developed a strong interest in audio as a result of building his own amateur radio equipment. His interest in audio led him to further experimentation and construction of a hi-fi system and, eventually, to a career in electrical engineering. Mr. Stephens received the B.S.E.E. degree from Carnegie Institute of Technology in 1969 and the M.S.E.E. from the University of California at Berkeley in 1974. During his graduate work he specialized in integrated circuit design and non-linear circuit analysis techniques.

Mr. Stephens has worked in instrumentation design for Hewlett-Packard Company and is now doing integrated circuit design for consumer electronics at Signetics Corporation. He has written several published works, holds one United States patent, and has three others pending.

tant to audio designers as it indicates how transient intermodulation varies with the input signal levels.

Note in the Fig.III-4 data that some devices do not fit the characteristic straight line relationship between distortion and slew rate. These are grouped to the right of the line and generally show excessive distortion for their high slew rate capability. With the exception of the Bi-FET devices (356, 357), all these are slew-enhanced op amps.

They feature an input transconductance that varies with level to produce rapid slew rates for large signals. Unfortunately, the changing transconductance gives rise to a crossover type of distortion mechanism. Since for small signals their slew rate capability is low, they begin to produce distortion for relatively slow waveforms. As the speed and amplitude of the input is increased, the performance of the device gets better, and it is more capable of producing the required output. Thus at high slew rate inputs, the distortion doesn't increase, it merely remains the same percentage as it was under low slew rate conditions.

We found that under varying input slew rate waveforms, the output spectra of the slew-enhanced devices remained fairly constant; only the relative magnitudes of the individual distortion products varied up and down. Increasing the input slew rate caused some distortion terms to increase and some to decrease--but the magnitude remained fairly constant under these conditions. It is interesting to compare this behavior with the leveling off of THD observed in our phase I THD tests.

The Bi-FET devices also failed to fit the characteristic straight line, but they suffer from a different type of problem from the slew-enhanced circuits. The Bi-FETs showed only even-order distortion falling on the square wave harmonics. They produced no other intermodulation products as did the slew-enhanced devices. The Bi-FET devices seem to alter the symmetry of the waveform, indicating some kind of lop-sided nonlinearity. This theory is supported by the basic slew rate of the 356, which is 11V/µS positive and 27V/µS negative. The problem experienced by the Bi-FETs is

Table III-2

Device	Conditions (unity gain unless noted)	SR, device (V/µS)	SR, signal (V/µS)	DIM (%)	Comments
741	10V, 30k	0.5	1.6	13.6	
	5V, 30k	0.5	0.8	1.1	
	2.5V, 30k	0.5	0.4	0.15	
	1.25V, 30k	0.5	0.2	0.1	
	0.625V, 30k	0.5	0.1	0.06	Sq. wave harmonics
709	x10 G & C 20V, 30k	2.5	3.2	0.75	
	10V	2.5	1.6	<0.04	
301A	x1 G & C 10V, 30k	(+)0.95 (-)0.8	1.6	1.06	
	x10G & C 20V	(+)7.5 (-)5.0	3.2	0.07	
	10V, 30k	(+)7.5 (-)5.0	1.6	<0.04	
	20V, 100k	(+)7.5 (-)5.0	10.2	0.09	
4136	20V, 30k	1	3.2	8.5	
	10V	1	1.6	0.63	
	5V	1	0.8	0.14	
4558	20V, 30k	(+)1.5 (-)1.45	3.2	5.2	
	10V	(+)1.5 (-)1.45	1.6	0.3	
	5V	(+)1.5 (-)1.45	0.8	<0.08	
1456	20V, 100k	2.5	10.2	2.03	
	20V, 30k		3.2	0.31	
	10V	2.5	1.6	0.09	
2720	20V, 30k	0.5	3.2	31.9	
	20V, 30k	1.6	3.2	1.28	
	20V, 30k	5	3.2	0.02	
531	20V, 100k	(+)20 (-)18.7	10.2	0.4	
	20V, 30k	(+)20 (-)18.7	3.2	0.4	
	10V	(+)20 (-)18.7	1.6	0.43	
	5V	(+)20 (-)18.7	0.8	<0.08	
535	20V, 100k	(+)20 (-)25	10.2	0.2	
	20V, 30k	(+)20 (-)25	3.2	0.19	
1741S	20V, 100k	(+)16 (-)20	10.2	0.11	
	20V, 30k	(+)16 (-)20	3.2	0.14	
	10V	(+)16 (-)20	1.6	0.14	
538	x10 gain 20V, 100k	(+)30 (-)40	10.2	0.35	
	20V, 30k	(+)30 (-)40	3.2	0.15	
	10V	(+)30 (-)40	1.6	0.15	

Table III-2

Device	Conditions (unity gain unless noted)	SR, device (V/µS)	SR, signal (V/µS)	DIM (%)	Comments
530	20V, 100k	27.5	10.2	0.08	
	20V, 30k	27.5	3.2	0.08	
	10V, 30k	27.5	1.6	<0.04	
530A	20V, 100k	(+)45 (-)50	10.2	0.14	
	20V, 30k	(+)45 (-)50	3.2	<0.02	
8007	20V, 100k	3.6	10.2	1.01	
	20V, 30k	3.6	3.2	0.06	
AD540	20V, 100k	5.7	10.2	0.26	
	20V, 30k	5.7	3.2	0.04	
536	20V, 100k	10	10.2	<0.02	
356	20V, 100k	(+)11.5 (-)25	10.2	0.14	Sq. wave har. only
3140	20V, 100k	(+)9.5 (-)5	10.2	0.52	(-) transitions ragged
	20V, 30k	(+)9.5 (-)5	3.2	0.04	Sq. wave har. only
	20V, 100k (3.3k to -15V)	(+)9.5 (-)12.5	10.2	0.04	Sq. wave har. only
TL084	20V, 100k	8.3	10.2	0.02	Sq. wave har. only
2620	x1 comp 20V, 100k	5.7	10.2	0.2	
	20V, 30k	5.7	3.2	<0.02	
	x10 G & C 20V, 100k	65	10.2	<0.02	
2525	x1 comp 20V, 100k	20	10.2	0.02	Sq. wave har. only
301A	Feedforward 20V, 100k	(+)12.5 (-)50	10.2	0.06	Irregular waveforms
OP-01	20V, 100k	20	10.2	0.02	Sq. wave har. only
318	20V, 100k	75	10.2	0.022	
TDA1034	x1 comp 20V, 100k	7	10.2	0.2	
	20V, 30k	7	3.2	<0.02	
	x3 G & C 20V, 100k	16	10.2	<0.02	
356	20V, 100k	(+)11.5 (-)27.5	10.2	0.15	Sq. wave components only (for all 356s and 357s)
	20V, 30k	(+)11.5 (-)27.5	3.2	0.09	
	10V	(+)11.5 (-)27.5	1.6	0.1	
357	20V, 100k	(+)65 (-)75	10.2	0.1	
	20V, 30k	(+)65 (-)75	3.2	0.08	
	10V	(+)65 (-)75	1.6	<0.04	

not inherent in all FET op amps, by any means. The 536, for example, had DIM levels below the resolution of our measurement equipment.

Some Perspective on TIM

The current literature (references 3, 4, 6, 8, 9) generally says that three amplifier design factors cause transient intermodulation distortion:

1. A small signal open loop bandwidth within the audio range, i.e., lower than 20kHz.
2. Large amounts of negative feedback (such as typically found in op amp circuits).
3. Significant transient signal errors, produced by poor input stage linearity, and ultimately, limited overload margins.

As it has been written, factors one and two internally stress an amplifier by driving it into large signal overload conditions, and as a result factor three then predominates, generating the intermodulation distortion. In the discussion which follows, we will attempt to show that the first two are not sufficient in themselves to produce this form of distortion in the devices we measured by the recommended sine-square method.[17] Factor three, which effects TIM, is characterized by the device's slew rate.

Factor three is, in our view, just a different way of phrasing the premise which has been the undercurrent of this entire study, namely that deviation from input stage balance produces large distortions in an op amp, that is, during slewing. If slewing is avoided, then these distortions are not produced, and this can be shown to be true, *with relative independence with regard to the first two factors.* It is of paramount importance that these points be appreciated, as factors one and two in themselves, if taken at face value, rule out use of op amp devices entirely in quality audio.

Few currently available IC op amps have have open loop bandwidths greater than 20kHz, which would suggest that *any* device chosen for an audio circuit will produce TIM. This will be shown to be a fallacy, since a great many IC op amps show TIM below residual levels (>90dB down, 100kHz BL).

Factor two suggests it should be impossible to produce low TIM in high feedback factor designs, such as normally typified by op amp circuits* This will also be shown to be a fallacy, as a great many IC op amp circuits show TIM below residual levels, with near 100% feedback, or feedback which is 90dB or greater at low frequencies.

Factors one and two are examined closely in the light of measured TIM on specific device examples in Figs.III-5-7.

Fig.III-5 exhibits the TIM spectra for two different devices under the same test condition of 10Vpp output, with a 30kHz band limit. The top curve represents a 741, and the bottom curve a 536.

* *The precise effect of negative feedback on distortion is described in: Jung, W., Stephens, M., Todd, C., "Slewing Induced Distortion and its Effect on Audio Amplifier Performance", AES Preprint, 57th Convention, May 1977.*

FIG. III-4

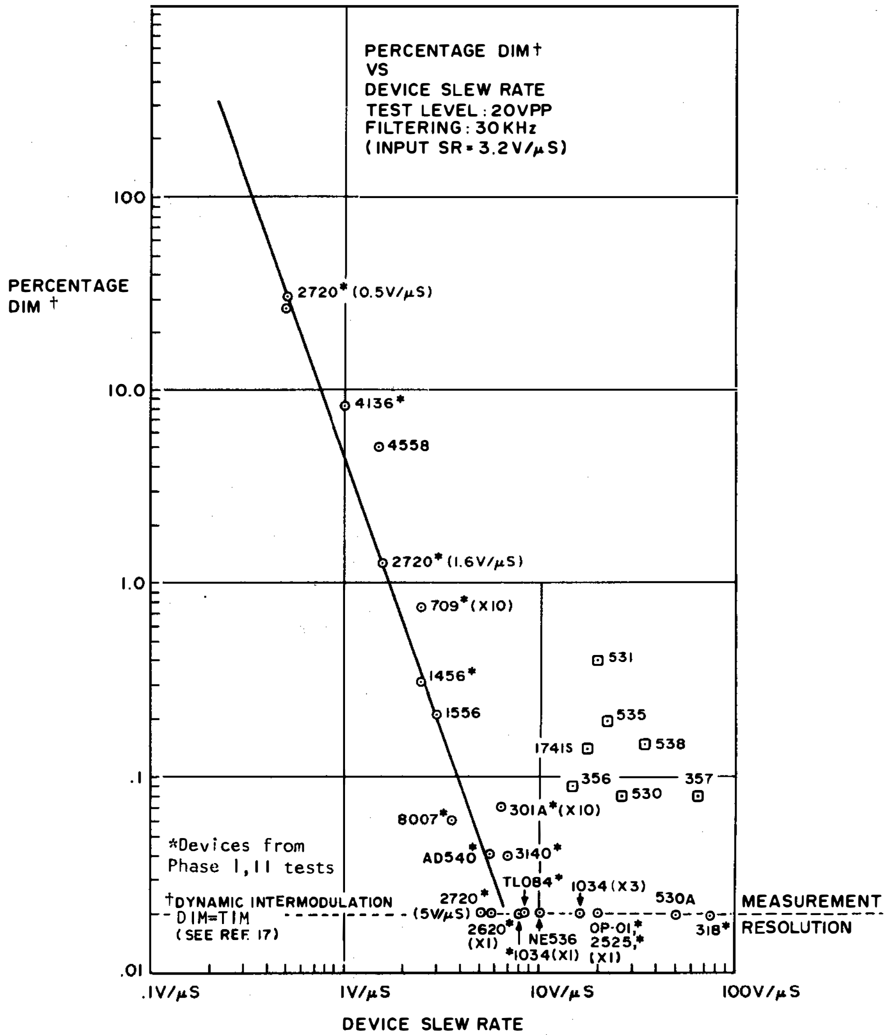


FIG. III-5

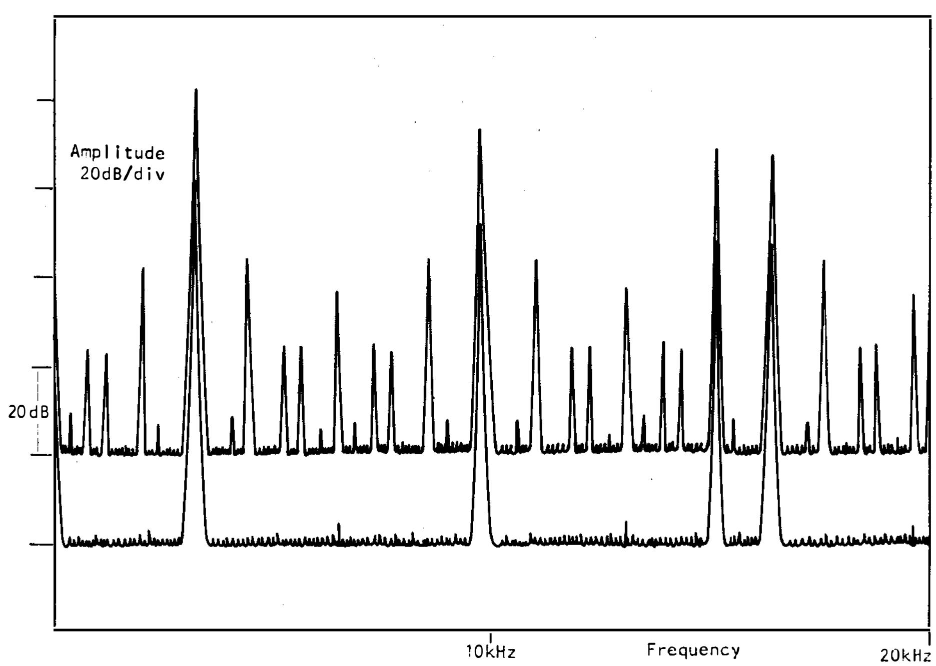


Fig. III-5. Comparison of TIM Performance, Different SR devices - Unity Gain Inverter 10Vpp, 30kHz BL. Top 741 SR = .8V/µs; Bottom 536 SR = 10V/µs.

These two devices are specifically chosen for near equivalent bandwidths, but an approximate order of magnitude difference in slew rates. Test conditions are identical, with a noise gain of two which represents approximately 100dB of negative feedback.

The 741 produces gross amounts of transient distortion, while the 536 is completely clean. This particular test case contradicts factors one and two, as the 536 which performs superlatively has an open loop bandwidth on the order of 20Hz, and a feedback of nearly 100dB. The results indicate strongly that factor three, as characterized by the high slew rate of this device, is responsible.

Another test which points to a similar conclusion is shown in Fig.III-6. Here we examine the TIM performance of a 2720 device, for 20Vpp out, 30kHz BL, with various slew rates programmed. The slew rates are 0.5, 1.6, and 5V/µS respectively, and are represented by the upper, middle, and lower traces in order. Clearly the increase in slew rate is accompanied by a corresponding improvement in TIM performance, as was true in THD and two-tone IM observed previously for this device. Feedback is high (NG = 2) in all three cases. While bandwidth of the 2720 changes with different programming currents, as does slew rate, at all times the open loop bandwidth is 100Hz or less. This device's results also point to slew rate as the determinative performance criterion.

A test which isolates slew rate alone as the major variable *in a single device* is shown in Fig.III-7. Here we see TIM performance of a 301A, for two different levels of compensation and feedback. In the upper trace TIM performance is relatively poor, the conditions are unity gain compensation and a noise gain of two. The lower trace shows essentially zero TIM, and the conditions are a x10 compensation and noise gain.

In an externally compensated op amp such as the 301A, the loop gain and gain bandwidth product remain the same when the device is compensated proportional to the noise gain, as here. Thus the only major variable of this case is the slew rate, which increases by an order of magnitude. The results speak for themselves, isolating factor three as the culprit.

Having isolated and demonstrated the singular importance of the third factor, our next logical step is how we may accurately characterize it. We have found in our study that the easiest and most direct measure of an amplifier's high frequency large signal non-linearities and overload margins is its inherent slew rate. In general, design innovations that improve slew rate also reduce transient intermodulation distortion (as was true for THD and two-tone IM). This is true for all devices tested, including the slew-enhanced circuits.

Perhaps now it is appropriate to comment in perspeetive on how factors one and two could arise as design criteria. In most amplifiers open loop bandwidth and slew rate have a direct link; therefore a design which extends open loop bandwidth to above 20kHz would also increase slew rate proportionally, and so reduce TIM. It might therefore be concluded that such a bandwidth is a fundamentally necessary criterion (if the slew rate relationship was not completely understood).

Also, designs which use less feedback require smaller compensation capacitances for stabilization. A natural consequence of the smaller capacitance is a higher slew rate, which results in less TIM. Therefore it could also be concluded (unjustifiably) that less feedback is a necessary condition for low TIM.

The results of this study clearly contradict factors one and two, at least to the extent that they should be qualified in terms of their relationship to slew rate. A modified form of these factors might be: 1) "More open loop bandwidth is better, but for the most part to the extent that it improves factor three;" 2) "Less feedback *can* be better, if it results in an increase in slew rate."

Both of these factors in their original form have been shown by many different examples of results to be incomplete performance criteria in themselves. They can (and often have been) misapplied to prejudge certain types of audio cir-

FIG. III-6

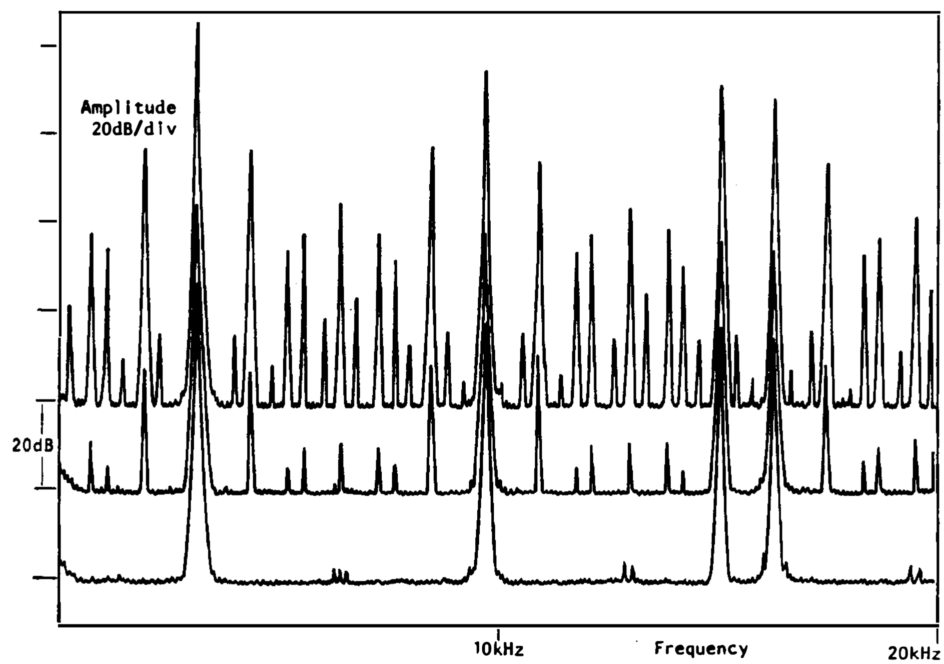


Fig. III-6. Comparison of TIM Performance, Adjustable SR device, Unity Gain Inverter, 20Vpp 30kHz BL. 2720 Op Amp, Top SR = .5V/µs; M SR = 1.6V/µs; B SR = 5V/µs.

FIG. III-7

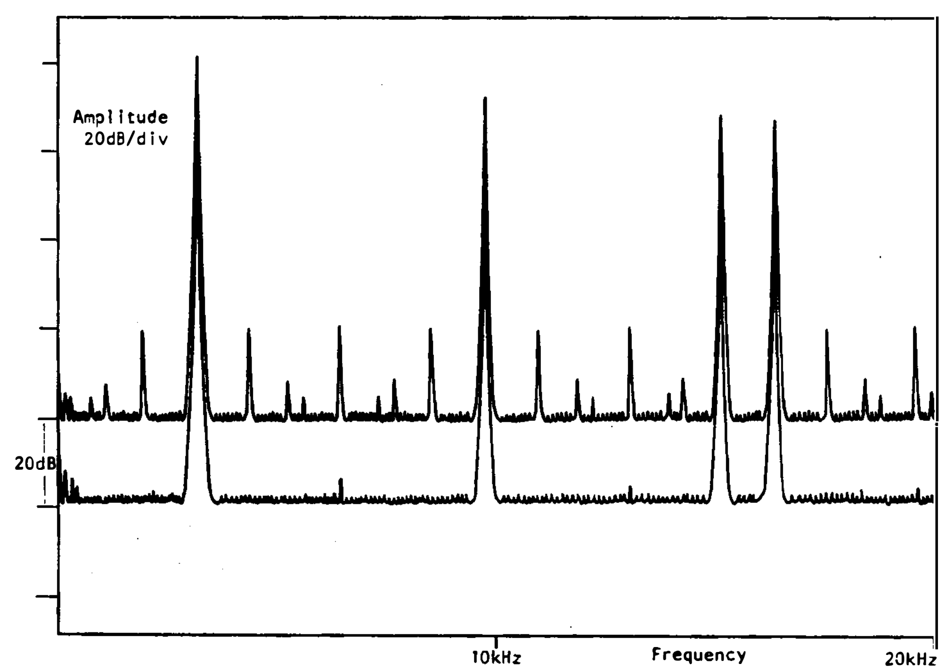


Fig. III-7. Comparison of TIM Performance, Custom Compensated Device, Unity Gain Inverter, 10Vpp, 30kHz BL. 301A Op Amp, Top X1 Comp, SR = .9V/µs; Bot. X10 Comp, SR = 7V/µs.

cuits, most often IC op amps.

The major and over-riding criterion is factor three, which is related to and inseparable from the amplifier's slew rate. Thus it is most correct to speak in terms of slew rate to characterize large signal high frequency and transient performance dynamics, as slew rate is by far the most significant limit to audio circuit fidelity. This study has shown this to be true, regardless of how one chooses to measure this fidelity--THD, two-tone IM, or TIM, or (to follow) listening.

In summary, a general relationship exists between device slew rate capability and transient intermodulation distortion. This relationship, for most devices, has been shown to have a simple cubic form. That is, transient intermodulation increases as the third power of the signal input slew rate for a given device.

Exceptions to this relationship do exist and emphasize the fact it is not just slew rate capability in a circuit which is important but exactly how the slew rate is achieved. Nevertheless, with the exception of the noted anomalous units, the relationship has wide applicability to most op amp circuits.

One of the most important factors which should be appreciated concerning TIM behavior in op amps is its relationship to output level. This can be graphically shown by resulting TIM output spectra for both high and low levels, and summarized by a plot of RMS TIM vs. input level.

FIG. III-8a

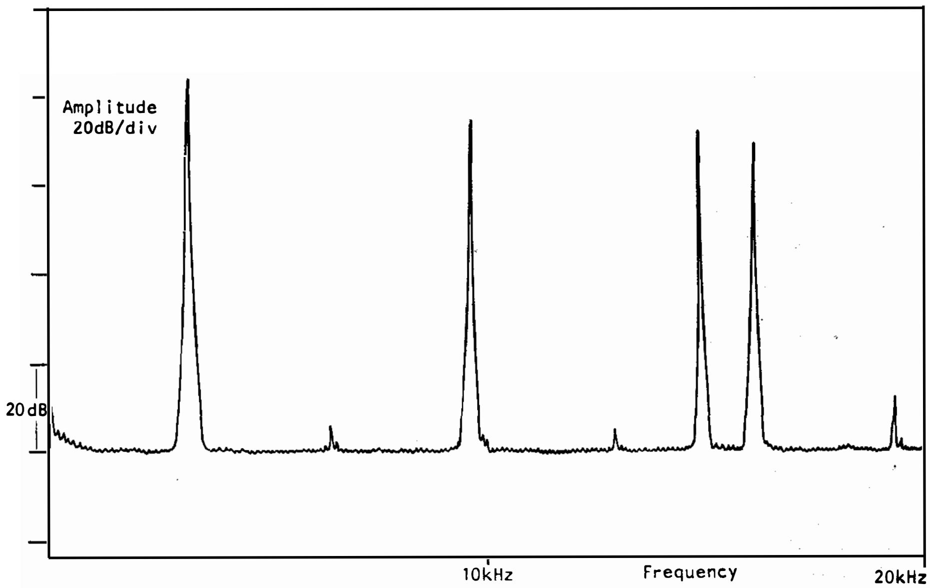
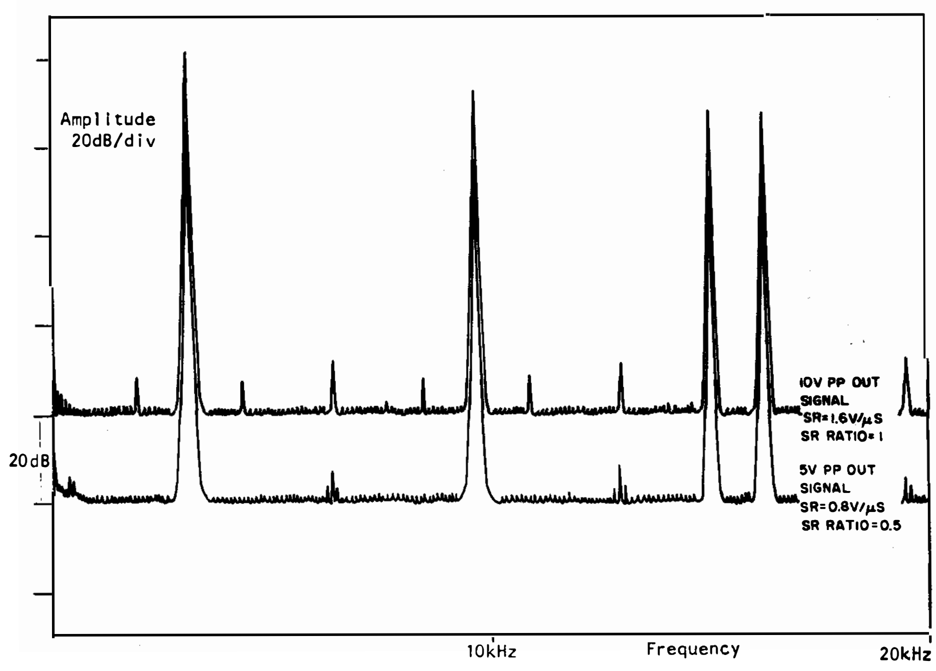


*Fig. III-8(a). TIM Performance, Unity Gain Inverter, 1.25Vpp, 30kHz BL, **741 Op Amp.** Signal/Device SR = .4.*

FIG. III-8b

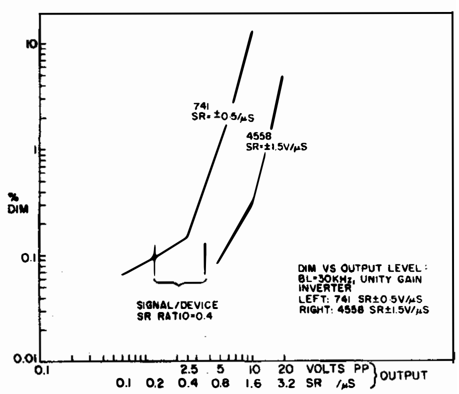


In Fig.III-8a we shown a 741 TIM spectrum for the conditions of 1.25Vpp and a 30kHz BL. It is quite apparent there is no TIM being generated at this level, whereas the same device at a 10Vpp level (Fig.III-5) was in serious trouble. This underscores the fact that it is not just slew rate, but SR in relation to the signal (output) SR which causes difficulty--in other words, how much the device's inherent slew rate is being taxed by the output signal. In the case here, the signal SR is 0.2V/µS, and the device 0.5V. This suggests a 0.4 signal/device slew rate ratio as a minimum for this case of a 30kHz BL, which agrees with Fig.III-4.

This conclusion is further reinforced by Fig.III-8b, an offset TIM spectrum for a 4558 op amp for 10 and 5Vpp, 30kHz conditions. In the 10V upper trace we can see that weak TIM is being generated (0.3% RMS). The signal slew rate for this condition is 1.6V/µS, roughly equal to the device's slew rate. In the lower trace which is for 5V, corresponding to 0.8V/µS, there is no TIM, again supporting the 0.4 signal/device slew rate ratio criterion (for TIM).

The RMS TIM data for these two devices are plotted vs. the output level in Fig. III-9. In this form, the data show how TIM rises rapidly as this ratio is exceeded. Actually, for both devices, the

FIG. III-9



TIM at an output slew rate below the device slew rate by a factor of 2.5 is masked by the residual of test setup.

We glean an interesting point from this: TIM at even a 1:1 ratio of signal/device slew rates is only on the order of 0.3%. In fact, if we examine the device performance summary given in Table III-2, we find that almost all devices show RMS distortion levels of 0.3% or less for signal/device slew ratios of 0.75 or less.

From these data it would seem fair to conclude that a signal/device slew rate ratio of .4 or less would be conservative, as it yields negligible distortion (≤0.1%) in all cases examined. Ratios lower than 0.4 are, of course, more conservative in terms of TIM performance. Later we will show that a safety factor of an even *greater* ratio of slew rates (<0.4) is necessary to satisfy other total distortion criteria which are *more stringent* than TIM performance alone.

Test Methods Correlation

We have now performed three different electrical tests on an identical group of IC op amps, and shown in general that all three sets of results can be linked to slew rate. It remains now to bring the existing data together in a form which can produce an assessment of which method (or methods) are preferred, and why.

Fig.III-10 is a general indicator of how the three different forms of SID are sensitive to device output amplitude. Here THD, two-tone IM, and TIM distortion data are plotted vs. output drive level. It is relatively easy to see that the curves have a definite similarity, as they all possess a common rapid rise in distortion with increasing level. Although the data shown here are for a 0.5V/µS SR 741 device, similar results could be shown for higher speed units. This figure demonstrates the common link of all three forms of distortion to slew rate, which should serve as an overall perspective.

A more complete appreciation for the three types of distortion and their common link to slew rate may be obtained from Fig.III-11. Here we use the same device and performance data, but in normalized form, with a common reference of signal/device *SR ratio*. This shows exactly and comparatively how each form of distortion behaves as a device is exercised with various signal/device SR ratios.

Again we note the very similar rapid rise in distortion. In general we can see that all three forms of distortion show this rise in magnitude as the SR ratio approaches unity. There are, however, great differences in sensitivity to the exact ratio among the different test methods.

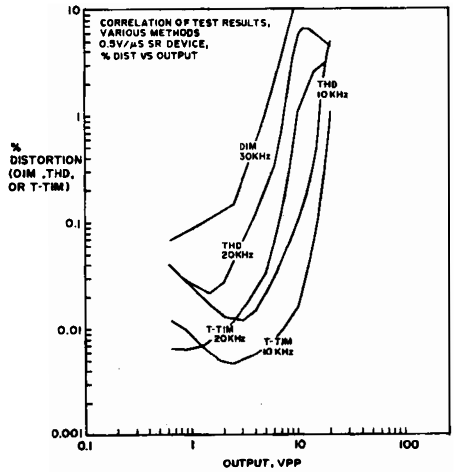
The largest dynamic range of distortion resolution is shown by THD, which is also most sensitive, showing the highest unity SR ratio percentage. This indicates not only that the THD method is the most sensitive means of SID detection near the unity SR ratio point, but it can also detect SID further down from this reference--as shown here almost three orders of magnitude further down.

The two-tone IM method is not nearly as sensitive as THD around the unity ratio, but does possess reasonable dynamic range below this reference.

The TIM method has a unity ratio sensitivity comparable to two-tone IM, but quite poor dynamic range below this reference. Its unique ability is its power to illustrate gross percentages of distortion at SR ratios *higher* than unity, a feature which is of little or no practical value since devices would hardly be operated in such a manner.

Thus we see that not only are the three methods correlated, they demonstrate different sensitivities and dynamic ranges of SID detection. This latter factor is highly important to any audio circuit/equipment evaluation work. Other factors are of course involved in selecting a particular method, such as equipment availability, convenience, and potential error factors, to name a few.

FIG. III-10



Of the three methods the THD method (swept, full voltage level, 100Hz-100kHz) is generally the best choice, except for the case of power amplifier stress and bandwidth related errors. For low level processing the method has few drawbacks, as bandwidths are often greater than 100kHz.

The two-tone HF IM method appears to be most suited to the power amplifier problem as it can be used with in-band fundamentals to detect SID. This circumvents the bandwidth/stress problems.

The TIM method is interesting, but has a number of serious drawbacks. The burden of square wave fidelity is one of these; another is the requirement for a relatively high aggregate of sophisticated test gear. Data reduction to the form of an RMS distortion figure is extremely tedious, a weakness seemingly inherent in the method.

On the other hand, the method does yield an immediate and obvious qualitative dynamic check of distortion performance from its spectrum display. However, it might be argued that simpler methods (such as the Holman[16]) produce similar dynamic results, with a near equal density of in-band components.

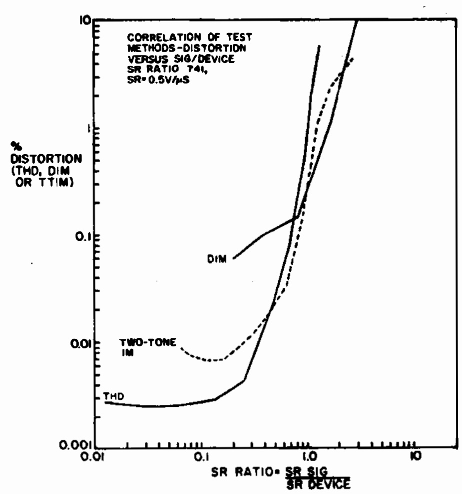
A further demonstration of the sensitivity of the THD method of evaluating an amplifier for SID is shown in Table III-3. Here are charted the results of nine devices found to be outstanding in all three types of tests. Clearly devices can show substantial variation in THD and yet be inseparable in two-tone IM or TIM results. This is even more evident when we compare the respective device THD curves against two-tone IM or TIM data.

Thus it seems more than reasonable to conclude that the THD sweep method, when applicable, is the most exhaustive of the three used in these tests. Any of the devices listed can be applied in confidence that they will be free from SID (when used as indicated).

We earlier showed that a simple oscillographic measure of slew rate and its symmetry serves as a minimum qualitative check, which can ultimately be related to SID (in its various forms). In the absence of anomalous behavior (such as some slew-enhanced units, and the Bi-FET devices), this check can be extrapolated directly as a general quality measure.

We suggest that a full rated output level slew rate measurement be adopted as a standard industry measurement technique, in conjunction with THD and two-tone IM as SID yardsticks. In the case of power amplifiers, the slew rate check should be done both with resistive and capacitive loads; this is also true for the THD/two-tone tests.

FIG. III-11



Device (operated as noted)	THD (20V, 10DkHz)	Two-tone IM (20V, 50k)	DIM (20V, 100k)
301A (feedforward)	0.12	0.002	0.06
536	0.067	0.006	0.02
3140 (with pulldown)	0.063	0.002	0.04
TL084	0.18	0.002	0.04
2525	0.062	0.002	0.02
2620 (x5 comp)	0.052	0.002	0.02
318	0.048	0.002	0.02
OP-01	0.037	0.002	0.02
TDA1034 (x3 comp)	0.045	0.002	0.02

Table III-3

PREPRINTS: The entire series of four articles are available in booklet form from TAA's circulation office for $16.50 ea.

ACKNOWLEDGMENTS

The following manufacturers assisted in this study by providing sample devices for test, and their cooperation is gratefully acknowledged. For specific device information, contact them directly (mentioning this article).

Analog Devices Semiconductor, 829 Woburn St., Wilmington, MA 01887 — AD518, AD540

Fairchild Semiconductor, 464 Ellis St., Mountain View, CA 94040 — uA741, uA709

Harris Semicondictor, P.O. Box 883, Melbourne, FL 32901 — HA2625 (commercial 2620), HA2525, HA2725 (commercial 2720), HA4741

Intersil, 10900 N. Tantau Ave., Cupertino, CA 95014 — ICL8007

Motorola Semiconductor, Box 20924, Phoenix, AZ 85036 — MC1456, MC1741S

Concluded on page 57

they understand the issue, they will not vote for S.864 *or any similar legislation.* Don't say "if S.864 is passed" or anything of the kind--come on strong about the irrationality of that whole approach to RFI.

JUNG: SID, Part 3
Continued from page 28

National Semiconductor 2900 Semiconductor Dr. Santa Clara, CA 95051	LM301A LM318 LM356, 357
Precision Monolithics 1500 Space Park Dr. Santa Clara, CA 95050	OP-01
Raytheon Semiconductor 350 Ellis St. Mountain View, CA 94040	RC4136 RC4558
RCA Solid State Division Route 202 Somerville, NJ 08876	CA3140 CA3080
Signetics 811 E. Arques Ave. Sunnyvale, CA 94086	TDA1034 (5534) NE 536 NE531, 535, 538, 530A NE 540, 541
Texas Instruments Dallas, TX 75222	TL084

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TEST METHODS:

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14. McClain, E.F. Jr., "Intermodulation Distortion Produced by Out-of-Band Program Components" Journal of the AES, Vol 24,#2 March, 1976.
15. Jelsing, T., "Causes and Elimination of TID," AES preprint #A-5, AES Zurich Convention, March 1976.
16. Holman, T., "New Factors in Photograph Preamplifier Design," *Journal* of the AES, Vol.24 #4, May 1976.
17. Leinonen, E., M. Otala, & J. Curl, "Method for Measuring Transient Intermodulation Distortion (TIM)," AES preprint #1185, Fall Convention, 1976.

Finally, I'd like to address an issue not covered in your editorial --RFI in equipment other than home entertainment equipment.

Perhaps the most disturbing prospect is the specter of interference in medical-electronic devices, such as cardiac pace-makers. At least one CB magazine has noted that high powered CB installations using linear amps may interrupt the operation of a pacemaker and thus prove fatal. (I don't know what data was published in that article, but I do know that there is data available that indicates such a possibility....)

Of course, the implications extend far beyond the pacemaker. Medical electronics is a growing field. Experiments involving the implantation of electrodes in the human brain to provide inputs for artificial sensory devices have been startlingly successful. The functional artificial eye and ear may not be far off. Of course, if the bandwidth required for good resolution in television pictures is any indication of what will be required for the artificial eye, the whole concept may not be feasible in the kind of RF jungle that CBers are creating today.

That should put the legal aspects of RFI in a new light, and, hopefully, make us less ready to accept the presence of "CB in our environment".

Jack Hannold
Elwyn, PA 19063

COAXIALS AND EFFICIENCY

A COUPLE OF WORDS about the "high efficiency speaker system" (Issue #3, 1976, p.37) seem in order. First, the matter of coaxial speakers. One objection to side-by-side mounting raised was the non-coincident arrival time of acoustic pulses. In general, the most the human ear can distinguish is 2 milliseconds. This corresponds to worst possible conditions with woofer and tweeter 60cm. apart. This is easily incorporated into speaker design, thus is unimportant.

With regard to general clarity of sound, see the *Journal of the Audio Engineering Society,* April, 1976. Paul Klipsch did distortion analysis of coaxial vs. discrete components. His results clearly show the first distortion component to be 13dB lower in discrete components (20cm. apart) than in comparable coaxial systems. The results show discrete woofers and tweeters give audibly less distortion than coax.

It seems also that in calculating the efficiency of the system, the passive crossover was neglected. This is a mistake. The Altec crossover typically wastes (according to their own literature) about 50% of

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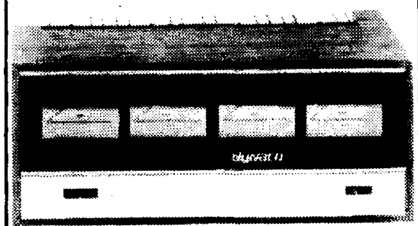
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Slewing Induced Distortion: Part 4

Phase IV: Listening Tests for SID

by Walt Jung
Contributing Editor

WE COME NOW to the acid test or, in popular parlance, where the rubber meets the road: listening. If all the electrical tests we've made in this series are significant, some definite patterns should evolve in listening tests of the op amps studied for SID. If we find no pattern in listening, all the above work will probably be rejected as meaningless by the ultra purist, "subjective only" advocates.

Fortunately, as we will see, the answers from the listening tests not only correlate with the measurements, *they correlate well*. But before we get into that, let's look first at how the tests were done.

Test Setup

For this testing phase, I did a fair amount of preliminary work to prepare my system for an easy and repeatable "objective" subjective test (if there is such a thing). In the general interest of simplicity and practicality, these listening tests were all done in mono, evaluating one IC at a time.

Fig.IV-1 is the block diagram of the system. Both channels of my reference stereo power amp (Ampzilla) are fed in parallel from a single channel of a Dyna PAT-5 preamp. This guarantees a balanced mono signal from each speaker, which are Magnepan MG-II's. All sources were operated either in a mono mode (tuner) or strapped in parallel for mono, in the case of tape and phono. Thus only one signal channel of the PAT-5 is used for the tests.

The PAT-5 channel I actually used was one specially modified for these tests to eliminate all traces of SID. Details of the modification are to be described in a future article (and embodied in a kit), but in general consisted of wholesale replacement of circuitry, with the use of the highest performance ICs (in terms of SID) selected from the THD and IM tests. This modification resulted in a full output level (7V) THD of less than 0.002%, across the full audio band.

Listening tests on this modified version of the PAT-5 opened up new dimensions in musicality, and even an increased depth and spaciousness, even though operation is strictly mono. I doubt if valid listening tests could be performed for SID without this type of modification, as quite a bit of SID was originally present in both the phono and high level circuitry. Unfortunately, this factor will complicate the exact duplication of my tests by other readers, at least for the present time.

The audible separation and identification of SID in an experimental audio circuit will be very difficult or impossible, unless the test system is already low in SID. In other words, you can't easily A-B a before/after switch for SID if you have a large measure of it in your present preamp or power amp. However, if you have a first-rate system, with a smooth clean upper range, it should be possible.

An audible verification of this is a completely free and unrestrained high end which will reproduce transients with an exact and detailed naturalness. High level cymbals and/or traps should sound effortless, with crisp reproduction. Upper register violins should be smooth, sweet, and warm, with no traces of edge or hardness whatsoever.

You'll know what I mean by this if violins elicit a "goose bump" effect when reproduced on your system. It is this sort of quality which is needed for reference in comparing various ICs. This of course assumes your recordings and/or other sources to be top-notch.

Given the above, you can compare various ICs for SID effects in an appropriate listening test circuit. In my case I connected the listening test circuit between the TAPE OUT jack of the PAT-5 and the TAPE 2 "in" jack. The PAT-5 is operated with all input sources except TAPE 2, with the monitor switch on TAPE 2.

In this manner the INPUT/MONITOR push button switch can be used to select between the input reference signal (A) and the signal through the test circuit (B) which appears as the Tape 2 monitor signal. Other preamps (if used) have similar facilities. Thus the tape monitor switch is used as an A-B select between the original source and the version passed through the test circuit.

The listening test circuit is shown in Fig.IV-2, and is especially designed for maximum sensitivity to SID. The first amplifier A1 is a 6 to 20dB gain stage (gain set by R3), which scales up the signal from its normal line level to one which will drive the U.U.T. to near full voltage output in normal operation. The device used for A1 is a 318, selected for its virtually zero SID and wide bandwidth.

I operated the U.U.T. in the unity gain inverting mode, for the reasons previously cited. Its output, which is greater than the input by the gain factor of the A1 stage, is then scaled down to the original level by R9/R10-R36. The output signal across R10 is equal to the original input level, within the allowance of resistor tolerances. For A-B testing, levels should be matched, either by the use of matched pairs for the like resistors noted, or by trimming one resistor (such as R9).

In my circuit I matched levels within ±0.1dB by trimming R9. C3 removes any DC offset generated by the test circuit. Exclusive of the U.U.T., distortion through this circuit is less than 0.004% and consists mostly of noise. Thus there is fair assurance that what you will be listening to is actually distortion in the U.U.T., not in the remainder of the circuit. This was also audibly verified by using a 318 in the U.U.T. position.

In use the circuit is fairly simple, but an oscilloscope is a handy operational aid, and can also visually indicate SID. With the circuit connected and operating, check first for correct 1:1 signal reproduction across R10-R38. Input levels and/or the use of gain control R3 should be adjusted for an amplitude @ A1-6 of close to 20Vpp, but below clipping. Avoid clipping, which confuses the distortion issue. Now, depending on the U.U.T. in service, you may or may not be able to hear distortion.

You can gain some familiarity with the sound of SID by purposely using a very low slew rate device, and playing a high level, high frequency selection (with suitable quality). One method of ensuring this is to use first a 301A overcompensated with 330pF. By using the scope to monitor the summing point of the U.U.T. on a sensitivity of about 50mV/division, you should see increasingly large voltage levels coinciding with the HF passages which trigger slew limiting.

This can be tested audibly, by switching the output of the test circuit to DISTORTION PRODUCTS, which allows you to listen to these error components. When SID is generated you will hear a hard, gritty, and grating distorted sound that

FIG. IV-1

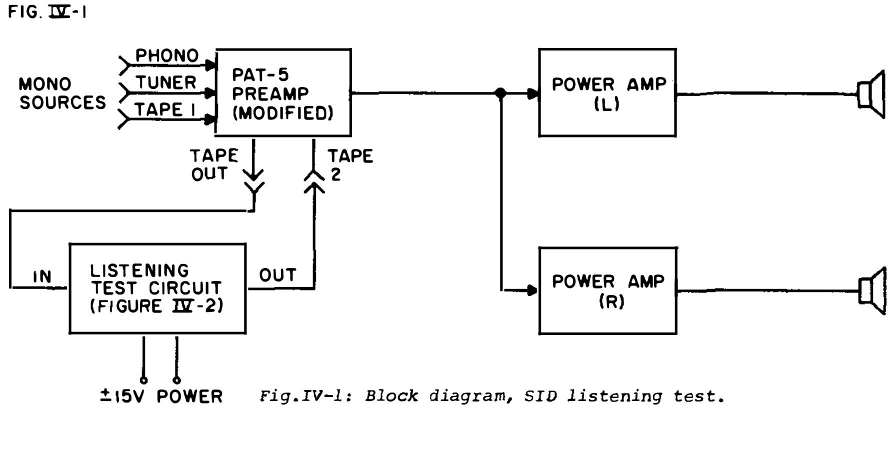
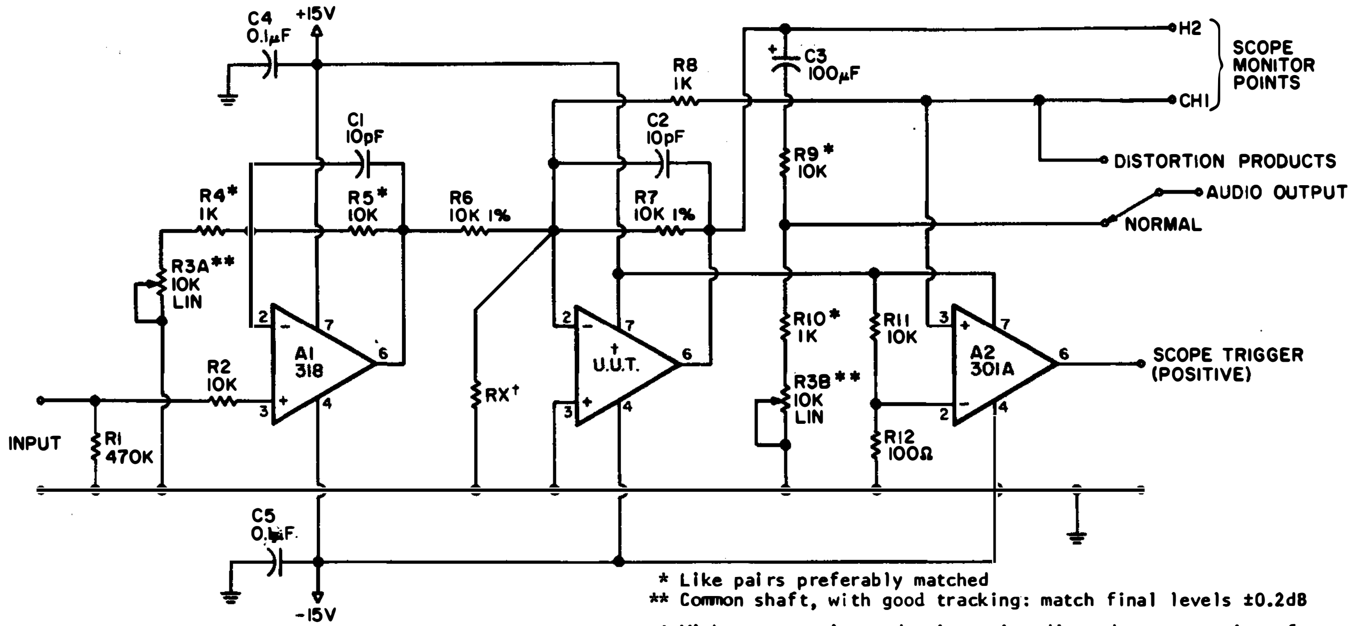


Fig.IV-1: Block diagram, SID listening test.

FIG. IV-2



* Like pairs preferably matched
** Common shaft, with good tracking: match final levels ±0.2dB
† With compensation and noise gain adjusted as appropriate for test

Fig.IV-2: Circuit for SID listening tests.

is unmistakable. You now have a reference for what to listen for within the overall sound picture. Listening to the summing point in effect scales up the distortion, and isolates it for scrutiny. It will never be this bad in a normal frame of things, but this test will give you an audible perspective of its nature.

At this slew rate setting, SID is relatively easy to generate and detect. By switching the monitor point back to NORMAL, distortion should be readily apparent with most program material, particularly high level passages with high frequencies. To verify that you are hearing true SID, rotate R3 to a lower gain setting, which will result in less output voltage from the U.U.T. The harsh distortion will disappear, although the overall level remains the same (assuming the R3 a-b sections track well). This demonstrates how SID is level dependent, being worst at the highest voltages.

With your ears thus "calibrated," you are ready for analytical listening tests. This is best done in steps, beginning with a low slew rate device such as the 741 or a 709 with x1 compensation. Careful listening to HF components of program material should reveal edginess on crescendos and peaks. The scope should also serve as a verification of this, by displaying bursts of fuzz at the summing point which coincide with the SID.

By using the scope in a dual trace mode, with the second channel monitoring the U.U.T. output, watch for the maximum levels which can show SID. With repeated listening, you should be able to A-B the source and the U.U.T. to verify whether the SID is actually coming from the U.U.T. or the source.

I find many terms describe the sound of SID. Interestingly, the subjective reviewers have used a great many of them for years to describe audible defects. Yet no one has directly linked these colorations to measurable circuit parameters. Part of this difficulty is in finding the right relationship; an even greater one is to quantify the audible degradation as well as the electrical measurements, once a relation has been established. At present, I'll admit I am a lot worse at quantifying the subjective experience of SID than the electrically measured form.

Yet audible gradations do seem to evolve from the listening experience. The audible degradations detected in these listening tests did not appear at all to be characteristic of the particular device type; rather they were related to the device's slew rate capability.

Several factors are important to the overall success of the listening tests. One of these is the correct choice of program material. It should emphasize high frequencies, the range of 2-3kHz and up. The levels should preferably be as high as possible and feature extended solo passages: violin concertos fill this bill nicely, if the recording is first-rate. One of my most useful test records is the Bruch concerto (see listing), which features a final movement which can demonstrate (expose) SID better than a lot of others.

Bluegrass material is useful but requires more care. Selections which feature well recorded solo fiddles can be almost as useful as a violin concerto. Plucked string instruments such as guitar, banjo, and mandolin are also useful indicators of transient quality, but they must be *acoustical* pickups only. A good test record for this type of material is the Mike Auldridge album listed; it is well recorded and outstanding in its dynamics.

Rock and typical pop music is the least useful, and in general should not be considered for evaluating SID. So much distortion and compression is built into these recordings that you'll never be able to separate things properly.

The Sheffield direct cut discs are outstanding in unrestricted dynamics, the best I know of to demonstrate what transients should sound like. In commercial recordings, this is about as close as you can get to studio sound. The Thelma Houston/Pressure Cooker release listed is useful for its spectacular drum transients and muted brass, which will pinpoint SID.

In the listening tests our procedure was to plug a device into the U.U.T. socket and listen carefully to the HF program content. Generally, I used only the HF range for comparison, giving little attention to lows. A first level, or most sensitive comparison could be made on one of the more controlled and "steady state" musical selections such as the Bruch (best) or Mike Auldridge cuts (next best).

I carefully monitored the solo violin in the extended high level passages for differences between A and B states. This would show up first as a barely discernible dulling of the string tone, or slight loss of warmth and sweetness, a vanishing of the "air."

These differences were so subtle they would probably not be detectable at all without an A-B comparison. I also no-

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SELECTED TEST RECORDINGS

*1 *Bruch and Sibelius violin concertos--Zino Francescatti, Schippers/Bernstein, New York Philharmonic,* COL MS 6731
2 *Orff: Carmina Burana--M.T. Thomas, Cleveland Orchestra,* COL MX 33172
3 *Mahler: 8th Symphony--Solti, Chicago Symphony,* London OSA-1295
4 *Lincoln Mayorga & Distinguished Colleagues, Vol.III,* Sheffield LAB-2
*5 *Thelma Houston/Pressure Cooker, "I've Got the Music for Me,"* Sheffield LAB-2
*6 *Mike Auldridge, "Blues and Bluegrass"* Takoma D-041
7 *Linda Ronstadt, "Hasten Down the Wind,"* Asylum TE-1072
8 *Linda Ronstadt, "Prisoner in Disguise,"* Asylum TE-1045

** Particularly useful*

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ticed that this first slight amount of degradation (level B) would not be detected at lower U.U.T. output voltage levels; only at near maximum output could it be observed.

The next detectable level of audible deterioration (level C) was a more apparent dulling of strings, and loss of warmth. The sound of strings is now tending toward dryness. This is a difference of degree, beyond the just detectable. Overall HF response is still generally satisfactory, and the audible coloration is slight. This level may possibly be audible in a straight listening test, if the listener is familiar with the musical selection.

Level D is one of marginal quality, with noticeable losses of string tone, warmth, and dimensioning. This is the first really serious level of audible defects, and can be noted by instruments which seem to "collapse" in their sonic image in level B, as opposed to a three-dimensional image which projects in level A.

I noted this in various instruments such as violins, banjo, mandolin, and traps. Instruments tend to sound "covered up" or recessed into the overall sound. Massed strings begin to take on an edge with high levels, and instruments generally blend as a homogenous source rather than concerted individual ones. Live voice reproduction sounds constricted or restrained, with a lack of natural quality.

The final categorized defect (level E) is recognizable on any sort of program material, and need not be sought after--it is apparent. Coloration and distortion of highs is obvious, and may also be accompanied by grit, fuzz, etc. In this severe case, the defect seems to affect voices to a greater degree, being noticeable on live speakers, particularly on sibilants. A speaking voice under these conditions sounds quite constricted or restrained, as if in a form of HF crossover distortion. The loss of naturalness is immediately apparent when A/B compared, but also obvious even without. HF musical transients may be almost completely subdued to the point of loss, with their residual manifestations being smeared. It is this quality level which was purposely induced initially.

Naturalness is the one general term which comes to my mind as most apt for judging the perspective of audible SID. If an amplifier sounds completely free and natural for any form of source material, regardless of difficulty, it is probably free of audible SID. The various levels of deterioration just described (B-D) all remove degrees of natural quality, to level E where the lack of naturalness is grossly apparent. For levels B and C, the degradations seem to be subtractions or losses from the original sonic image. With level D, these losses are increased, but with added distortions such as "edginess," "grit," or "constriction."

I suspect (but don't know of a simple way of proving it as yet) that levels B and C can be equated to the approach of slew limiting, as was evident in THD tests as the initial rapid climb in distortion. Level D seems to be this degree, plus perhaps occasional spillovers into complete slew limiting where the rasping of edginess and grit takes place. The transients sound as though they are covered up.

Level E is slew limiting for a great percentage of the time. This may be generally confirmed by the observation of ample bursts of summing point voltage, which indicates the open loop condition of slew limiting. I have made no effort at all to classify defect levels worse than level E, since this is already an intolerable degree of distortion.

To confirm or deny my thesis about the differentiation of quality levels B-C and D-E, and whether or not actual slew rate limiting was being triggered, I added the A2 comparator stage to the listening circuit (Fig.IV-2). I did so after I had already categorized the majority of the ICs for sound quality. This circuit simply triggers the scope very positively when a slewing condition is present in the U.U.T.

In operation, the (-) input of A2 is biased at +150mV by R11-R12. When a U.U.T. summing point voltage deviation occurs greater than this level, A2's output goes positive. This level change can be used to trigger the scope externally, so as to coincide the start of the sweep with a slewing interval when present. This allows positive identification and verification of slewing.

The A2 comparator's only weakness is that it responds just as well to *clipping* of the U.U.T. However, clipping is easily recognized, as the U.U.T. output (CH2) will be at a negative saturation level as the sweep starts. Conversely, a true slewing interval will start at some more positive level near zero, and ramp negative. You can (with careful observation) actually *measure* a device's slew rate under program conditions using this technique.

Using the comparator to detect slewing conditions and carefully setting the drive level to the U.U.T., I made a number of listening/electrical tests to

Table IV-1
Listening test results, referred to full output of ±10V)

SID category	I Deterioration			II Gross distortion	
Quality level	A	B	C	D	E
Audible character	No differences detected for any program material	Just discernible softening, loss of sweetness	Further softening, somewhat dry, generally satisfactory with slight loss of dimension	Colorations apparent, loss of dimension, "covered" sound, dulled transients, constricted, edge begins	Coloration and distortion obvious, more constricted covered sound, transients smeared, grit, edginess, fuzz
Associated slew rates	>4V/µS	2-4V/µS	1-2V/µS	0.5-1V/µS	<0.5V/µS
Samples tested	318, 518 TDA1034	1456 4136 (2V/µS)	4136 (1V/µS)	741	
	2625 2525				
	8007 NE536 AD540 3140		4741 356*		
	NE541 (x100 comp) NE540 (x100 comp) TL084 OP-01	NE541* (x10 comp) NE540* (x10 comp)			
	530A	530			
			535 538* 1741S		
	531 (x10 comp)		531 (x1 comp)		
	2720 (5V/µS) 301A (x10 comp) 301A (x100 comp) 301A (FF)		2720 (1.6V/µS)	2720 (0.5V/µS) 301A (x1 comp)	2720 (0.16V/µS)
		709 (x10 comp)			709 (x1 comp)

* *Audible ranking possibly due to factors other than slew rate*

verify the presence of slew limiting, on what material, and how often. The results are indeed interesting and I am sure will stimulate lively reader discussions.

Generally, once the device possessed a slew rate setting of 1V/µS (or more), there was never any gross limiting observed. Clipping of the U.U.T. occasionally triggered the scope, but since this could be easily identified, it was dismissed.

This general picture did not change appreciably with a device slew rate near 0.5V/µS, but some slew limiting did take place on the most difficult material on hand, the Bruch concerto. However, to reach a U.U.T. output level which would occasionally slew the U.U.T. on true HF signals, the overall level was such that frequent clipping was taking place on peaks.

This indicates one of two things. Either we need a more demanding test work, or it is unrealistic to expect 8-10kHz and up levels to be at or near peak program levels. Since the latter is more probable, this indicates that actual slew limiting would be relatively infrequent for 0.5V/µS to 1V/µS devices (although certainly still possible).

Below 0.5V/µS and at about 0.05 to 0.1V/µS frequent slew limiting takes place. On a rich HF range recording such as the Bruch, this is overwhelmingly evident, and the output waveform can readily be observed to contain the triangular slew limited components. Even on more ordinary program material slew limiting is not at all infrequent. We should expect this, since the fp of a 0.1V/µS device is only 1.6kHz which is near the center of the audio spectrum energy distribution, and thus wide open for slew overloading.

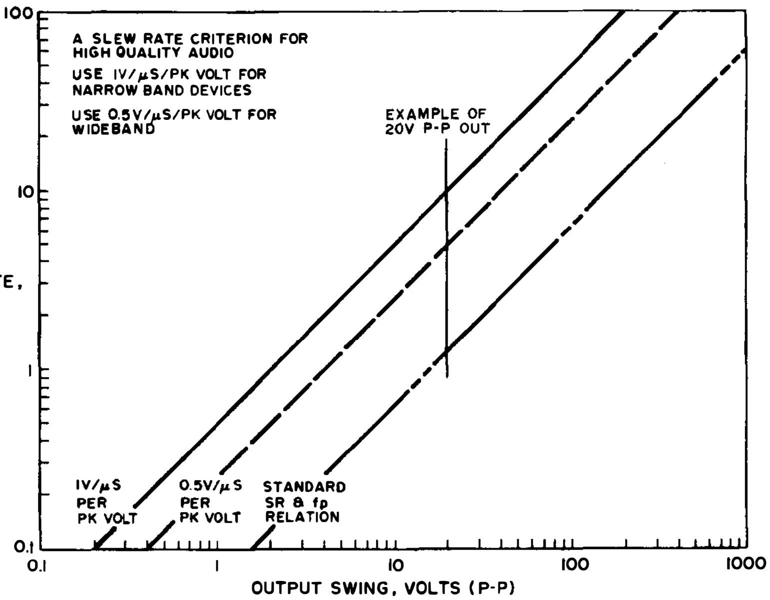
To bring these tests into a comprehensive focus, two broad but distinct categories of audible distortion seem to be associated with SID. At low levels of SID, the audible effects are a general loss of naturalness and dulling of detail. I will call this category I SID, a general deterioration, and it encompasses quality levels B and C and overlaps D.

A more serious form of audible distortion due to SID is associated with complete slew limiting, the condition when an amplifier is called upon to deliver a rate of rise in excess of its slewing ability. Gross distortion is evident in this case, apparent by fuzz, grit, and harsh reproduction on signal peaks. This I will call category II SID, and it encompasses quality levels D and E.

Level D brackets SID categories I and II, and depending upon the specific slew rate and program material may produce results of either. A low slew rate device such as a 0.5V/µS unit may occasionally slew limit and produce category II SID on certain program material, but not consistently. It will generally be producing category I SID (to its worst degree) accompanied by the associated dulled HF range.

Table IV-1 summarizes my listening test findings very nicely, and assigns an associated slew rate range for each quality level. I made these quality level and slew rate range judgments during the listening tests, using the devices shown. I will not say my results are immutable, but I do feel they illustrate a very definite pattern. This pattern is simply that higher slew rate devices generally sound better, and if sufficient slew rate is not present, a device can sound disastrously bad (category II).

FIG. IV-3



A less concrete result is the exact differentiation of quality levels B-D and I submit these results as one subjective listener to the situation. Others may hear things a little differently, but I do believe my point has been made that slew rate can be linked directly to audible quality, and in different gradations.

This thesis is experimentally supported by several devices whose quality level could be changed by an adjustment in slew rate: i.e., the 301A, 709, 4136 (different samples) and the 2720 which could be spotted at various levels.

Several devices on the IV-1 chart do not fit the general pattern of audible quality proportional to slew rate. These are starred, indicating that their ranking is probably due to other factors. For example, the NE540 and 541, with x10 compensation, slewed at 4V/µS or more. However, their sound for this condition was not of A level quality (indistinguishable from source), it was more of a level B type. This may be due to the rise in THD for these devices which approaches 0.1% @ 20kHz. When compensated for a x100 condition, they both became A level quality, and their distortion is lower for this condition.

The slew enhanced devices present some unique listening experiences. One of my initial curiosities concerning these units was whether or not the effects of the class AB input stage mechanism can be heard; the results make me believe they can. The 535, 1741S and x1 comp 531 all sound about the same, of a general C quality although perhaps bordering on D level. One can hear the beginnings of an edge and a definite loss of imaging.

The 538 units presented problems in evaluation, as two of the three samples exhibited a curious parasitic instability. The one unit which behaved more or less normally seemed to be better than the 535, 531 (x1) and 1741S, but it still was not altogether what I thought it should be. It seemed to be of B quality, but because the results were inconsistent in two of three cases, it is listed in IV-1 as a C.

The 530 was the best performer of the slew enhanced units, and sounded only slightly less than A quality. This unit showed lower THD than any of the others, which indicates that the ear is sensitive to really quite low levels of distortion in the 10-20kHz range.

The 356, on the basis of its slew rate and THD, one would expect to do fairly well. However, it was noticeably less than transparent, losing dimension and yielding a dry sort of sound; hence its C rating. Some of this may be the device's basic asymmetry, as noted above. This underscores the necessity of more research into this aspect of IC performance.

All these anomalies indicate that the ear can indeed perceive ***very*** small levels of distortion, down to well below 0.1%, perhaps as low as 0.01 or 0.02%. It may be that what we are hearing is not exclusively distortion, but other interrelated results which cannot as yet be completely pinpointed.

A New Slew Rate Criterion

This information, in conjunction with the results of the electrical tests, can be used as the foundation of a *new slew rate criterion*. It should be one based on the requirements necessary for high quality audio in a 20kHz bandwidth, and may be used as a predictive tool in circuit design or evaluation. I present it in graph form in Fig. IV-3.

To use this nomograph, only the required amplifier p-p output voltage need be known; the graph will then tell you the required slew rate in V/µS. As an example, an op amp intended to deliver its rated 20V swing is plotted (vertical line). This line intersects the standard fp relation (shown dotted, for reference only) at a level of 1.25V/µS. As has been shown, this is inadequate for high quality use; it is included here only for perspective.

The remaining two lines correspond to the new slew rate criterion, and represent slew rates of 0.5 and 1V/µS per peak output volt. The 0.5V/µS curve may be viewed as a minimum objective, the 1V/µS as a conservative one. The example of the op amp would thus require a 5V/µS (minimum) or 10V/µS device slew rate, as indicated by the intersection of the vertical line at these points.

Since two range extremes of the criterion are presented, the question arises: which should a designer use? The 1V/µS per peak output volt is the most conservative, and provides allowance for other error factors. It should preferably be used for instances where an excess of bandwidth is *not* present. The 0.5/µS curve can be used for wide band devices, which will have higher feedback factors and are therefore " more forgiving."

This criterion is justified on the basis of both the electrical and the listening tests. In fact, the latter would indicate a 0.5V/µS rating per peak output volt to be adequate.

As a final perspective on the listening tests, the reader should appreciate that since they were all done at maximum amplifier output level, they are as pessimistic as can be. In practice, this means any given device will perform proportionally better as the output levels are lowered. This should be an exact linear relationship, that is, a 2/1 performance increase, if you halve the output swing. In terms of the quality level brackets, this means a device will move to the left, or improve, in quality level. Since the brackets are intended to be roughly binary weightings (at least from B-D), a device could possibly move more than a single quality level.

Exactly how much improvement will result should be taken with a grain of salt, however, since what we are dealing with here are peak program levels, not nicely defined sine wave amplitudes. Regardless of level, however, the relative rankings of slew rate (and thus the associated devices) will still hold, which is simply to say "faster is better" except for the special cases as noted.

You can also take the new slew rate criterion, of course, and extend it *upward* in level to include power amplifiers. The Table IV-1 data, which are based on ±10V output levels, would in this case be optimistic. Of course the ICs tested here do not generally operate at higher voltage levels (with the exception of the 541, a power driver), but the slew rates associated with different quality levels should be almost directly proportional. Hence, a ±50V output swing should require five times the slew rates shown for the respective quality levels to attain comparable qualities.

At such an operating level, the crossover from level D to E would be 2.5V/µS, and A level quality would occur at around 20VµS. I can't say definitely that all this will prove out, as I haven't run as many tests on power amps as on ICs. I do know, however, that a faster, symmetrically slewing power amp sounds better, just as ICs do (see my Dyna 400 review in Issue #2, 1977,p. 48).

A System Slew Rate Perspective

With all this information we can now assemble a slew rate/signal flow diagram for the entire audio system. The Fig.IV-4 diagram applies the new criterion to an audio system on a logical, stage-by-stage basis, and shows the required slew rate of each stage. Although this drawing is somewhat hypothetical, and the exact numbers may vary in an individual case, the general principles of the relationship will hold for any set of values.

Defining the slew rate requirements for the components of a system begins with the power amplifier, which is first specified in terms of slew rate from its rated output voltage. Note the correct key terms here is *voltage*, not power. It is peak voltage swing which determines the required slew rate. To use a popular example, assume a 200 Watt into 8 Ohm amplifier. This results in a 112 Volt p-p output voltage for rated power. From Fig.IV-3, this requires a slew rate of 50V/µS (actually 56V/µS, but rounded off for purposes of illustration).

If the power amp is to be the weakest or limiting link of the system, all preceding stage slew rates should be in excess of the figure predicted by this step, when related to their individual levels.

To determine the slew rate required of the previous stage, the power amplifier slew rate must be *referred to its own input*, by dividing by the stage voltage gain to give equal basis comparison. In Fig.IV-4, assuming a 50V/µS power amp slew rate and a voltage gain of 25 (typical), this power amplifier's "input referred" slew rate would be 2V/µS. This means that even in the case of an "infinitely" fast power amp, a 2V/µS input slew rate from the preceding source would result in no more than a 50V/µS final output level slew rate.

Since an "infinitely fast" power amp does not exist, the driving source should have a slew rate in excess of the power amp's input referred slew rate, to avoid deteriorating the latter's final slew rate to less than 50V/µS (or whatever figure is appropriate).

My rule of thumb used here is a x5 ratio, but even greater multipliers will yield more conservative results. Therefore, in this case the preamp's final stage should possess a 10V/µS or more slew rate, so as not to deteriorate the power amp's 2V/µS input referred slew rate.

You may extend this rationale to all other fixed voltage gain stages, such as the output stage of the preamp (line amp block), which will typically run a gain of 10 (20dB). For an output slew rate of 10V/µS this stage will then have an input referred slew rate of 1V/µS. Again, using the multiplier of 5, all input sources should have a slew rate of 5V/µS so as not to deteriorate this slew rate and thus the final system slew rate.

Working toward the system input we find the master volume control which can pass preamp (or other source) voltage levels near maximum output, regardless of the final playback level. We may therefore justifiably regard the preamp in terms of its rated output voltage for slew rate, rather than referring it to the input referred slew rate of the line output stage. We can do this by using the new criteria of Fig.IV-3. For a preamp with ±10V output, the minimum slew rate would be 5V/µS, as noted in the example case here.

Further points should be dealt with on a system basis, such as out-of-band rolloffs to prevent possible supersonic slew limiting or IM generation. Typically such measures appear at the power amp input in the form of a passive RC network, although in some cases they may be appropriate at other points. Also, tone control and/or equalization HF boosts should be carefully considered on a system basis, as they can require slew rates in excess of that implied from the above, if the boosts affect frequencies above audibility.

Conclusion

After three series of electrical tests and one listening test on some 100 ICs, many things have been learned about audio performance feedback amplifiers as related to slew rate. However, as in many such instances of research, the important thing is to separate the wheat from the chaff. In plain words, the wheat follows.

The slew rate, or large signal voltage rate of change ability of an audio amplifier to be used in feedback or "op amp" type circuits is of *major* importance. This fact is demonstrated by both electrically measured and audible results on a large number of amplifiers of widely differing designs. In all cases where slewing rate limitations cause deviation from an ideal balanced input stage state, easily measurable distortions result.

Contrary to many previously published comments, this form of distortion can be identified and quantified (most effectively) by simple THD tests. Except for some specialized cases, the results of THD performance tests can be directly related to the slew rate of a given amplifier. This may also be stated for other forms of test, such as two-tone HF IM, and the sine/square technique, but those tests generally seem to be less sensitive to detection of SID.

Listening tests of these same amplifiers yield results which correlate surprisingly well, to the degree that audible sound quality may also be ranked by slew rate in a large percentage of cases. Listening test results reveal two major categories of distortion due to slew

FIG. IV-4

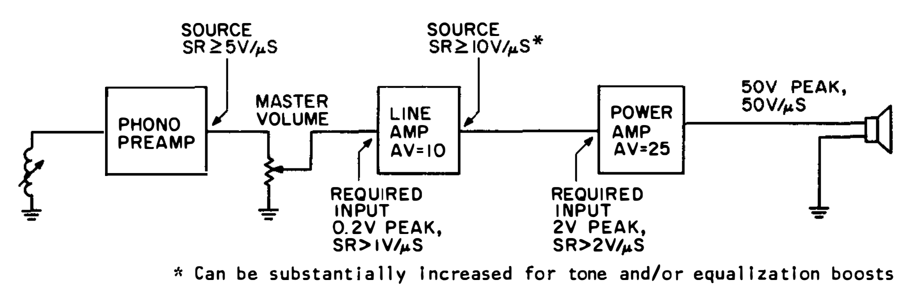


Fig.IV-4: Audio system slew rate/signal level flow rate: category I, a general deterioriation, and category II, a gross distortion. In electrical terms these categories are associated with (I) the approach of the device's slew rate, and (II) the exceeding of this slew rate.

Interestingly, the subjective impressions of the sounds of slew limiting have been used in print in many instances for some years, yet no one has defined this link in clear and unambiguous terms, let alone quantitative ones.

Slew limiting effects in both electrical and audible performance degradations can be avoided by applying a new slew rate criterion, to wit: "The circuit, including all possible loading conditions, should possess a slew rate of 0.5V/µS (minimum) to 1V/µS (conservative) *per peak output volt*." If this criterion is satisfied, electrically measurable distortions due to slew limiting will generally border on the unmeasurable, or THD below 0.01% of full scale (below 20kHz), as well as correspondingly low IM and TIM.

Audibly, the device performance will be such that slew limiting effects will be undetectable in a full level A-B test on the most difficult program material. The only qualifications for applying this slew rate criterion are that the device slew rate be symmetrical and that the input stage be of a class A design.

This criterion can be misapplied, if care is not taken to limit device input bandwidth to a normal 20kHz. Such a misapplication could result, for instance, when out-of-band components cause slew limiting. A more general rule for guaranteed satisfaction is to maintain the signal/device slew rate ratio at 0.25 (or less), a re-phrasing of the new slew rate criterion. This may be accomplished by appropriate passive band limiting to define signal upper slew rate limits.

A major implication of this study is the demonstrated necessity for both product specifications which include slew rate and standard test methods which recognize it. Although the solid state audio age has been with us for more than a decade, there is no present general recognition or appreciation of this distortion phenomenon. This is regrettable, as it is much more significant than most of the performance parameters typically specified and tested in audio gear. We hope this study will have a beneficial effect on this situation.

The above is a nutshell summary of the outcome of the SID study. We are aware that many of the general conclusions and specific points of this study are in conflict with some of the previously published works on TIM. No doubt this will lead to controversy, but then perhaps a clearing of the air is in order.

I have endeavored in every phase of the testing to be as objective as possible, while presenting an ample amount of reliable data which establish certain points. I feel confident that the electrical tests will generally be accepted as valid, but the listening tests (since they were done through my ears) may justify further test cases. I would welcome these in the interest of overall validity, and encourage readers to duplicate the tests on their own.

I welcome reader comments. However, please write directly to the appropriate manufacturer to find sources for a specific IC. Send along your comments and/or criticism, especially as to future research into the general subject of amplifier distortions. *[If you expect a reply, do please leave space on letters for comment, and always enclose a stamped, addressed envelope.--Ed.]* *[For a full listing of manufacturers and their addresses, see pp. 28 and 57 of TAA, Issue 3, 1977.]*

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AUDIO RESEARCH RE-WORKS DYNA ST-70

Continued from page 11

filament supplies were added, the builder could doubtless use EL34s in the finished unit.

The enterprising will note that the constant current source of the D76A is a single 6FQ7 tube and two resistors per channel and that DC balance would not be too difficult to add to the bias system. A delayed regulated high voltage power supply of higher current rating with separate filament supplies (possibly DC) would be the natural perfectionist's version of this unit.

THE SOUND

While we have not compared the ST-70-C3 to the D76A, we have had the chance to compare it to several solid state units. We preferred it to all save the Williamson Twin 20 Mark II. The sound of the two is remarkably similar--both very pleasing, concise, and without any grit or edginess.

We lived with the unit for many months and it has that hallmark of all good power amps, it really seems to have no "character" of its own. FM from the music stations runs along satisfyingly by the hour and suddenly we are dazzled by a live broadcast of the Boston Symphony when the sound takes on an attention arresting, nourishing and satisfying character. Side by side channel comparisons with amps of similar power make it evident that the ST-70-C3 is a very elegant device for reproducing music. Modest by today's power standards no doubt, but a quite welcome and valued addition to our system.

We trust that those who elect custom options in building versions of this unit will share results with other readers through the letters column. Those who know of good sources for more difficult to find parts are encouraged to share that data.

All in all, the ST-70-C3 project ought to provide a lot of new experience for tube buffs and perhaps a few will become converts to the cult. We found it an exceptionally satisfying experience.

SPECIAL NOTICE

OLD COLONY SOUND LAB will act as sole supplier and warranty agent for the Audio Research ST-70-C3. Audio Research will not answer any mail or phone queries about the product, its construction or maintenance.

preprint no. 1252 (G-7)

SLEWING INDUCED DISTORTION AND IT'S EFFECT ON AUDIO AMPLIFIER PERFORMANCE--WITH CORRELATED MEASUREMENT/LISTENING RESULTS

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presented at the
57th Convention
May 10 - 13, 1977
Los Angeles

AES

AN AUDIO ENGINEERING SOCIETY PREPRINT

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Abstract

There has been a great deal of material in the literature in recent years on transient intermodulation distortion (TIM) as a major distortion mechanism in audio amplifiers, particularly IC op amps. A detailed study of high level high frequency performance of op amps involving over 100 different device samples reveals the true distortion mechanism to be slew induced distortion (SID), with TIM actually being only one particular manifestation of SID.

The study demonstrates a direct correlation between device slewing rate and THD, two tone IM, and TIM test results, as well as listening tests. The results allow not only predictable electrical and audible results of feedback amplifiers based on slew rate behaviour, but also dispel several popular myths involving open loop bandwidth and feedback factors as design criteria.

Some major implications of this study are a new slew rate criteria for high quality audio circuit performance, the nature of various op amp slewing behaviour patterns, the audible nature of SID as correlated to actual slew rate, and the necessity for industry recognition of slew rate in both equipment specs and testing methods.

Introduction

The frequent lack of correlation between an amplifier's measured versus listening performance is well known. This leads to the immediate conclusion that relevant measurements are not being performed. Transient Intermodulation Distortion (TIM) [3,4,27] has been advanced as a distortion mechanism which could be partially responsible for this lack of correlation, and yet elude common measurements.

Most work on TIM has dealt with clipping of an amplifier's internal stages, which produces gross slew limiting on the amplifier output. Of much greater interest is the performance of an amplifier properly operated below its slew limit, and that is what most of this paper is concerned with. It will be shown that Slew Induced Distortion (SID) is the major distortion mechanism in most present day amplifiers. Measurements of this distortion will be presented and compared with calculations of its magnitude. It will be shown that an amplifier's Slew Rate (SR) and Gain Bandwidth product (GxBW or ω_u) are its most important specifications for audio performance. Some design quidelines will be given to allow designers to use and design amplifiers to avoid this type of distortion.

Data from three types of distortion tests will be presented. These are Total Harmonic Distortion (THD), Two Tone Difference Inter-Modulation Distortion (IM), and the recently proposed test for TIM.[16] Some of the relative merits of these measurement techniques will be discussed and it will be shown that, where applicable, THD is the optimum technique. It will become obvious that low frequency distortion tests such as 1 khz THD or 60 hz, 7 khz IM tests are useless for detecting SID. It will also become obvious that I.C. op amps, viewed with suspicion by some, are capable of superlative performance when properly operated below Slew Rate.

The Slew Induced Distortion Mechanism

It is important to understand the dominant distortion mechanism of an amplifier. We will mostly deal with operational amplifier circuits, but since most present day power amps are of similar design the discussion and data will be relevant to them as well.

Fig. 1a is an idealized model of our basic amplifier. Its input stage is a voltage to current converter or transconductance stage, characterized by the parameter g_m. The output current of this stage is simply

$$\Delta i = g_m \Delta V \qquad (1)$$

The second stage of our amplifier is an integrator with an output voltage

$$V_o = \frac{1}{C}\int \Delta i dt = \frac{g_m}{C}\int \Delta V\, dt \qquad (2)$$

The resistor R is responsible for the finite D.C. gain of the amplifier. At low frequencies the open loop gain is

$$A_o = g_m R \qquad (3)$$

The open loop frequency response begins dropping (Fig. 1b) at a frequency

$$\omega_o = \frac{1}{RC} \qquad (4)$$

Since for audio circuits we have no interest in the amplifier gain at D.C., it is much more convenient to neglect R (as in equation 2) and work with the unity gain bandwidth which, due to the integrator's -6dB / octave response is equal to the gain bandwidth product.

$$\omega_u = A(\omega) \times \omega = A_o\omega_o = g_m R \times \frac{1}{RC} = \frac{g_m}{C} \qquad (5)$$

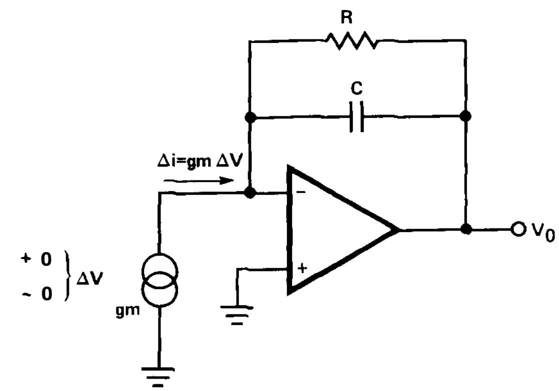


FIGURE 1a AMPLIFIER MODEL

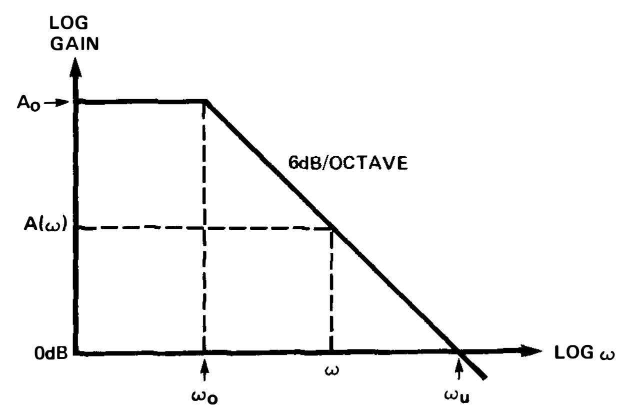


FIGURE 1b FREQUENCY RESPONSE

Refering to equation 2, we have

$$V_o = \omega_u \int \Delta V \, dt \qquad (6)$$

Thus for an amplifier with a 6 dB /octave frequency response, the amplifier can be characterized simply by its unity gain bandwidth or gain bandwidth product. Our next step is to examine the differential input voltage as a function of the output voltage. Differentiating equation 6 we have

$$\Delta V = \frac{1}{\omega_u} \frac{dV_o}{dt} \qquad (7)$$

This important result shows us that the instantaneous differential input voltage of an amplifier is proportional to the slope (or slew) of the output, with $1/\omega_u$ as the constant of proportionality.

If we now look at a real amplifier we will understand what SID is. Fig. 2a is a very simple real amplifier which will serve to demonstrate this. Q1 and Q2 are the differential input pair and Q3, Q4 form a current mirror. This stage is our transconductance amplifier with a transconductance of

$$g_m = \frac{I_k}{2V_t} = \frac{q}{2}\frac{I_k}{kt} \qquad (8)$$

Q5 with its current source load IA and capacitor C forms our integrator. We will neglect the finite d.c. gain produced by R. Ideally Δi is

$$\Delta i = g_m \Delta V = I_k \frac{\Delta V}{2V_t} \qquad (9)$$

but this is only true when ΔV is small. The exact expression for this input stage is

$$\Delta i = I_k \operatorname{Tanh} \frac{\Delta V}{2V_t} \qquad (10)$$

Our transconductance stage is not linear and thus will produce distortion when ΔV is large. Equations 9 and 10 are plotted in Fig. 2b.

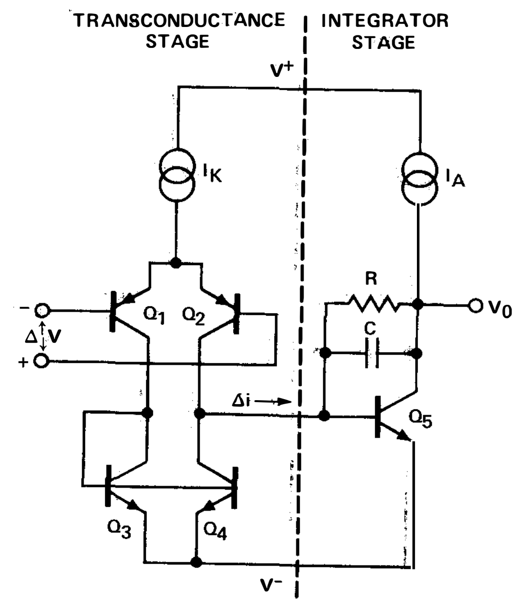


FIGURE 2a REAL AMPLIFIER

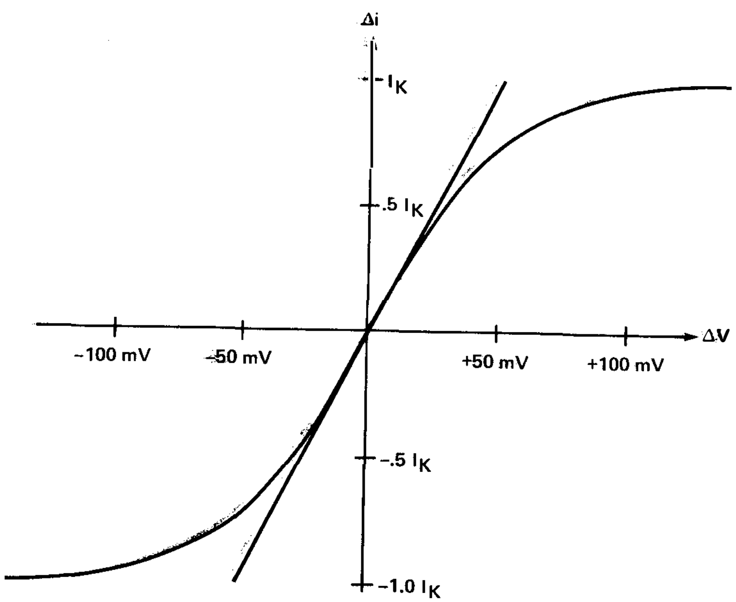


FIGURE 2b TRANSCONDUCTANCE NONLINEARITY

The maximum output current from our input stage is Ik. This determines the maximum rate of change of Vo which is the maximum slew rate of our amp.

$$S.R._{max} = \frac{Ik}{C} \qquad (11)$$

How close we are working to the S.R. max is simply

$$\frac{S.R.}{S.R._{max}} = \frac{\Delta i}{Ik} \qquad (12)$$

This ratio is easily measurable from outside the amplifier with a differentiator

$$\frac{\Delta i}{Ik} = \frac{1}{S.R._{max}} \frac{dVo}{dt} \qquad (13)$$

A glance at Fig. 2b tells us that operating with a $\Delta i/Ik$ ratio $> .25$ will produce some obvious distortion. This is equivalent to saying that operation at a greater than 25% of the maximum slew rate will produce distortion. This distortion depends solely on the rate of change of the output signal, hence the term "Slew Induced Distortion."

So far we have been talking only of the amplifier with no mention of feedback. We have been discussing the open loop performance. Amplifiers are rarely used open loop so we must turn our attention to the effects of feedback on amplifier performance. An important point to keep in mind as we discuss feedback is that feedback networks are placed around an amplifier and have <u>no effect</u> on its internal performance. Feedback will not effect the validity of any of the equations developed above.

As is well known, feedback will reduce distortion. Let's take a qualitative look at how this happens. A simple feedback network has been placed around our amplifier in Fig. 3. The differential input voltage is

$$\Delta V = \frac{Vin\, R_2 + Vo\, R_1}{R_1 + R_2} \qquad (14)$$

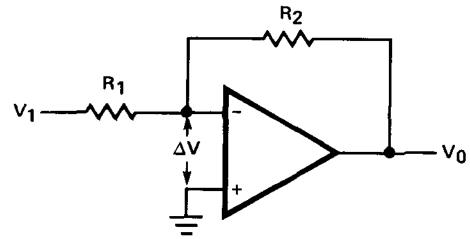
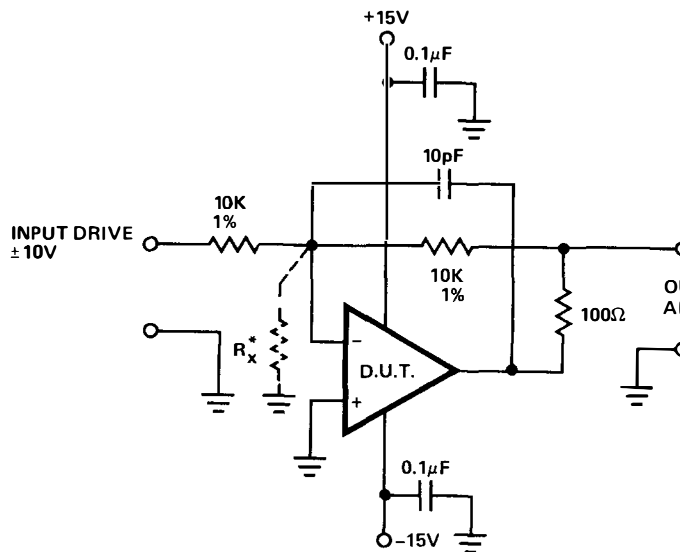


FIGURE 3 AMPLIFIER WITH FEEDBACK



*R_X IS VARIED FOR DIFFERENT NOISE GAINS AS FOLLOWS (WITH R_X OPEN NOISE GAIN IS 2)

NOISE GAIN	R_X
3	10K
5	3.3K
12	1K
101	100Ω
1000	10Ω

FIGURE 5 TEST CIRCUIT FOR SLEW INDUCED DISTORTION

This is the error voltage which we would like to be zero, but will be non-zero if Vo contains a gain or phase error, or distortion. If we operate the amplifier near its slew limit, we know that the amplifier will be very non-linear. The feedback will reduce the non-linearity from Vin to Vout, but it still exists from ΔV to Vout. If the feedback is doing its job and producing a relatively clean signal at Vout, then it follows that the signal ΔV must be distorted. The distortion must be of the proper magnitude and phase to compensate for the amplifier's internal nonlinearity. It is instructive to look at some of these waveforms as shown in Fig. 4. These are pictures of a 748 op amp compensated to unity gain by 30 pf and operated as shown in Fig. 3. The amplifier had the following performance:

$$f_u = \frac{\omega_u}{2\pi} \approx 1.5 \text{ Mhz}$$

$$\text{S.R.} = \begin{array}{l} +\ .97 \text{ V}/\mu\text{s} \\ -\ .91 \text{ V}/\mu\text{s} \end{array}$$

The amplifier was operated at full output swing of 20 V p-p. Two frequencies were used, 12.7 Khz and 19.1 Khz. At 20 v p-p these frequencies produce slew rates of $\pm$.8 V /µs and $\pm$1.2 V /µs respectively. These two frequencies were applied to the closed loop amplifier for gains of 1 and 10. For either gain, the output was a visibly clean sine wave for the 12.7 Khz, $\pm$.8 V /µs signal. The 19.1 Khz, $\pm$1.2 V/µs signal drove the amp into slew limiting, and this is shown in Fig. 4b. The output slewing waveform was visibly the same for either gain. Fig. 4F summarizes the photos.

The important point to see is that the op amp input, ΔV, becomes highly distorted in an attempt to linearize the response of the closed loop amplifier. As the maximum slew rate is exceeded this process breaks down and the error voltage goes wild. Operation at lower gain (more feedback) yields lower

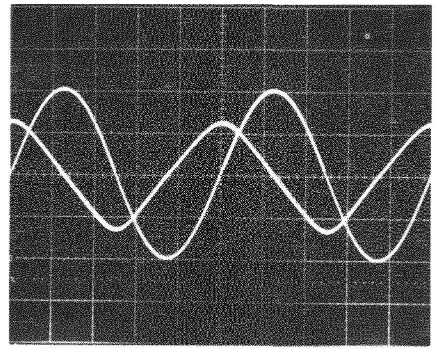


Fig. 4a

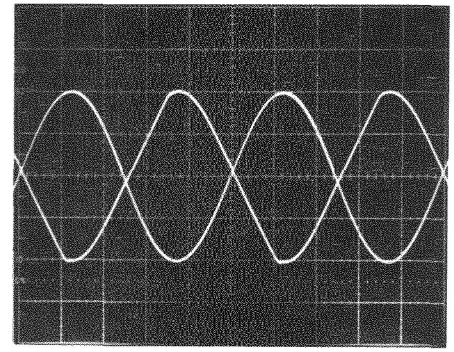


Fig. 4b

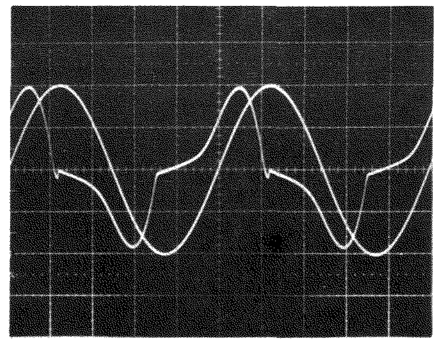


Fig. 4c

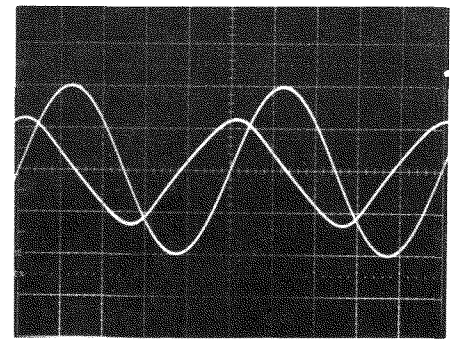


Fig. 4d

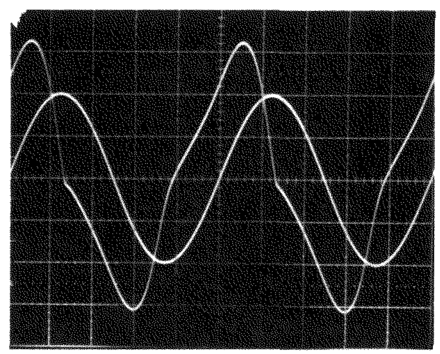


Fig. 4e

Fig.	Gain	S.R.	THD	Trace 1	Trace 2
4a	1	$\pm$.8	$<$.05	Vin 5v/div	Δ V .1v/div
4b	1	$\pm$1.2	3.5	"	Vout 5v/div.
4c	1	$\pm$1.2	3.5	"	Δ V .5V/div.
4d	10	$\pm$.8	.56	"	Δ V .1V/div.
4e	10	$\pm$1.2	4.1	"	"

Fig. 4f

Equivalent Slew Rate of Sine Wave = $2\pi f V_o$

Definition: $f_p = \frac{\text{Device Slew Rate}}{2\pi V_o}$

distortion operation, and allows low distortion operation closer to the slew rate limit.

There is nothing particularly unique about slew-induced distortion in audio amplifiers. It can be measured, calculated, and improved upon by using standard techniques that have been available for some time. The only elusive aspect of this form of distortion is that rather than occuring on a peak magnitude (like clipping), it occurs on the rising or falling edge of the waveform. This is due to the fact that the dominant non-linearity in the circuit, the transconductance of the input stage, is followed by an integrating stage. Thus in Fig. 1, if the transconductance stage were overloaded and producing clipped square waves of current output, the integrating stage would transform these square waves into triangle waves at the output. The triangle wave is the ultimate example of gross slewing distortion.

Although slew limiting is most often encountered in amplifiers due to internal I/C relations such as described above, it can also occur due to output current/load capacitance rate limiting, with the end effect being similar. This type of slew limiting can occur in equalized pre-amps which cannot adequately charge frequency shaping capacitors, or power amplifiers which cannot drive capacitive loads due to protection circuitry.

The distortion products produced by SID are measurable either by methods of THD, two tone HF IM, or TIM[16], and in all cases they become significant as the amplifier's inherent slew rate is approached.

Test Methods for SID characterization

A major objective of this study was to develop a reliable and predictive test method for the presence of SID. This objective was not only met, but was done for three different means of measurement, all of which correlate well with

each other, with calculations, and finally, with listening results. The different methods are discussed below.

THD Tests

It has been previously reported that THD testing methods are insensitive to the detection of TIM distortion.[16] In actuality this is only true for spot frequency THD tests. A full output voltage level THD sweep test from 100 HZ to 100kHZ has been found to be the most sensitive test to detect SID in op amps, as it exercises the output rate of change tracking fidelity to a high degree. Unfortunately this form of test is not always directly applicable to power amps, but it is an excellent one for IC op amps.

To implement this test, some important restrictions must be placed on the test circuit. The test configuration must operate in the inverting mode, to eliminate common mode distortion effects which exist when an op amp is operated non-inverting. The magnitude of these effects in some designs can approach that of SID, therefore a non-inverting test is incapable of separating these two components. Similarly, output stage non-linearity must also be minimized by careful restriction of loading to 10K or more. These precautions assure us that we are measuring SID. Distortion produced by poor common mode rejection and output loading should be evaluated separately and are not the subject of this study.

A test circuit which is suitable for SID tests is shown in Figure 5. It is a unity gain inverter, with compensation adjusted for unity gain, except for special cases as noted. Input-output signal levels are full rated voltage swings of $\pm$10V (7VRMS), except as noted.

The device under test (D.U.T.) is operated in this circuit, and the first test made is a check for its actual slew rate. For a given device the actual slew rate can vary markedly from the data sheet value, therefore results can only be correlated by actual measurement, using a fast rise square wave source. Ideally slewing should be symmetric, so the measurement should take note of both (+) and (-) slew rates. After the S.R. test, measurements of swept THD can proceed.

THD data on a 741 IC op amp with a 0.5V/µS SR is shown in Figure 6. This data indicates in the full output curve a characteristic sharp rise from the LF residual level, to a 1% level at the 8kHZ fp frequency, this occurring within only 2 octaves. For lower output levels such as for 2V and 1V RMS, the 1% frequency is proportionally higher, in fact by the ratio of amplitudes. In all three cases the characteristic sharp rise in distortion can be noted as the SR is being approached.

SID, improves considerably for higher slew rate devices, or compensation conditions which result in higher slew rates. In Figure 7, THD data on a 301A amplifier is shown for various compensation/gain conditions with all data referred to a 7VRMS output level.

The first curve is for unity gain compensation, where the SR is 0.9V/µS; the behaviour is similar to but slightly better than the 741 for similar conditions. For the x10 compensation curve, the resulting slew rate is 7V/µS and the performance is much better, with slew limiting not reached until 90kHZ. The improvement is due to the x10 improvement in Gain-Bandwidth product and slew rate.

The third curve is for a x100 compensation/gain, and here slew limiting is not at all evident, as the rise in THD is 6dB/octave, or bandwidth related.

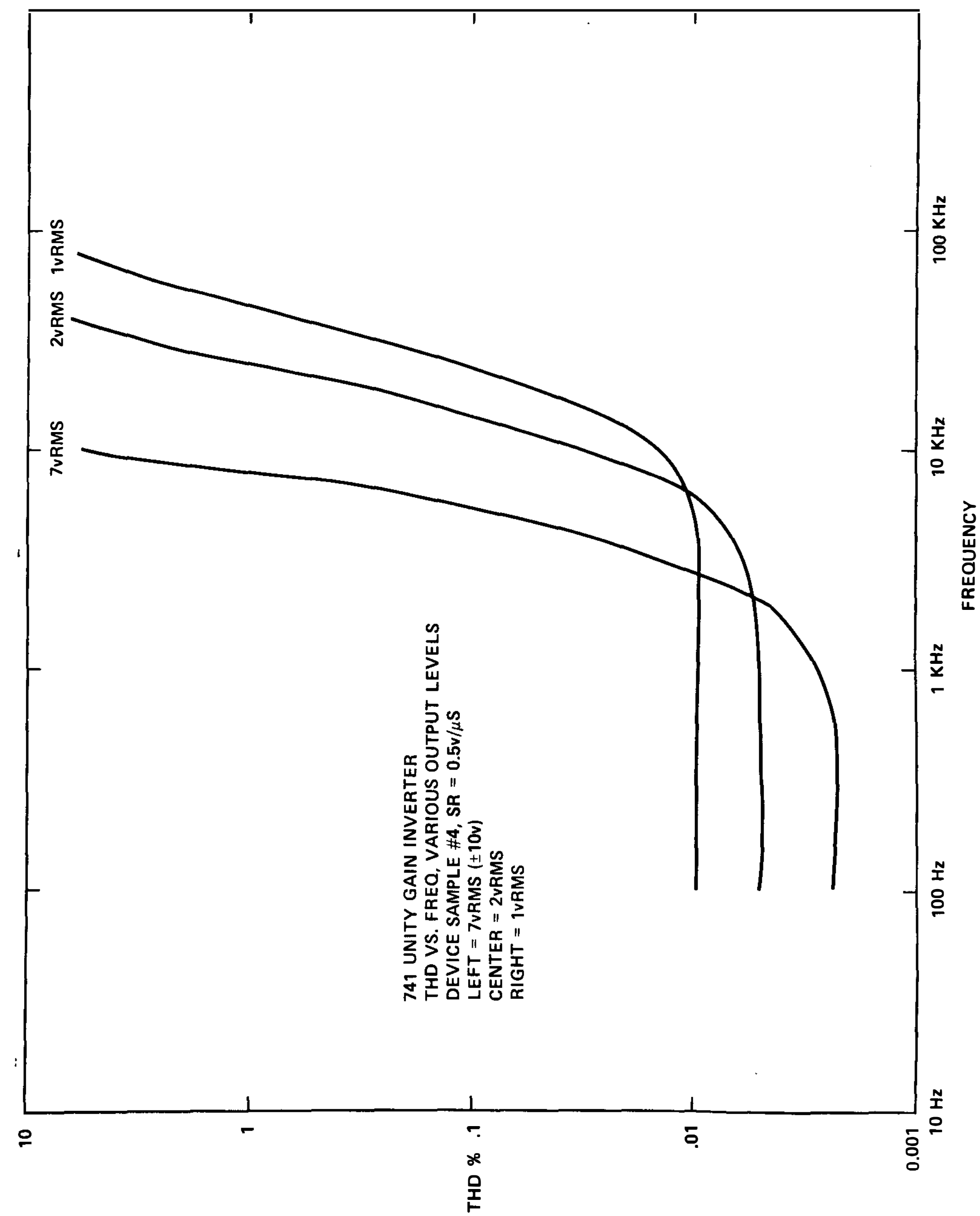


FIGURE 6

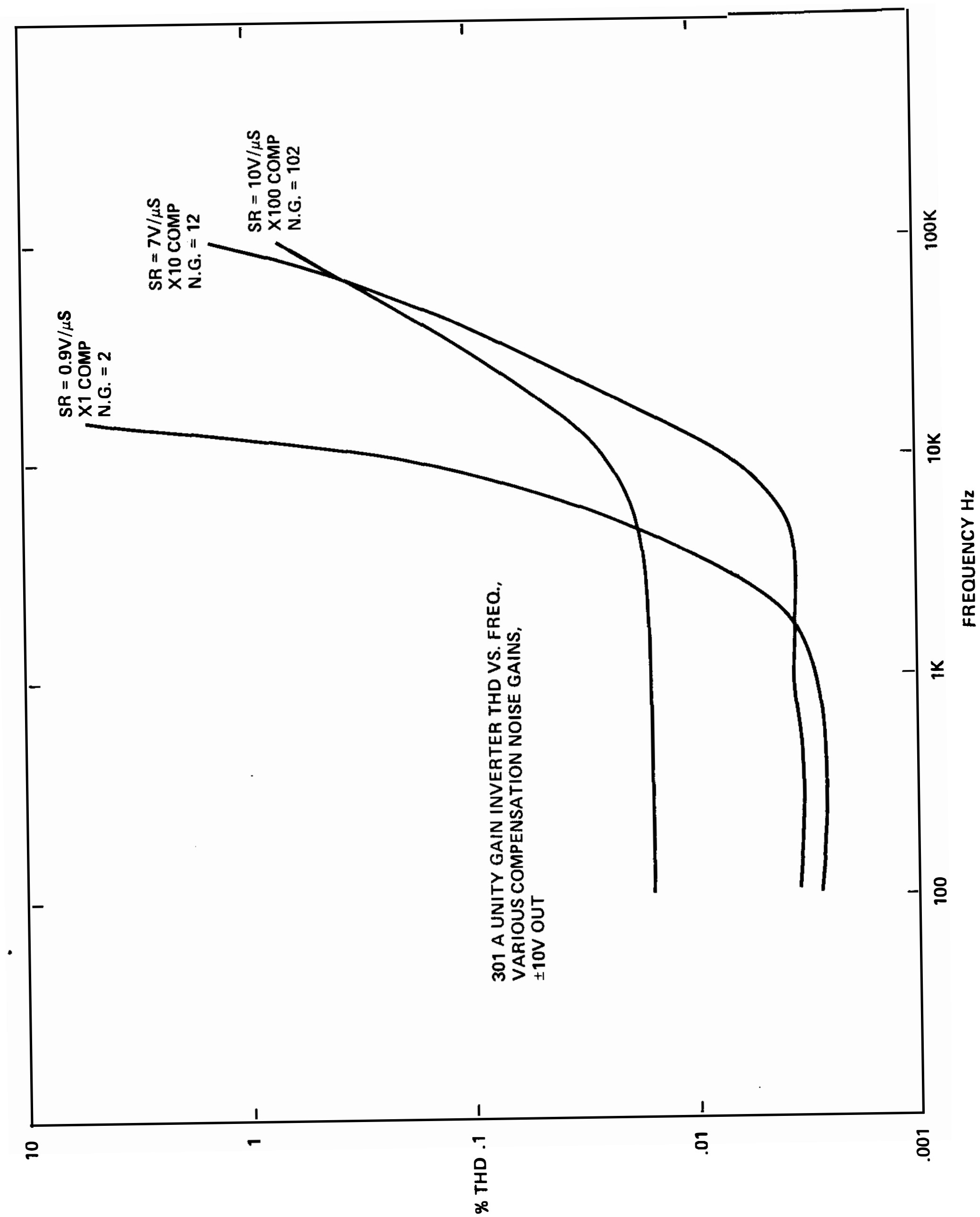


FIGURE 7

Slewing symmetry has a pronounced effect on SID, and SID will only be minimized when the (+) and (-) slew rates are equal. In some IC devices, particularly those which use current mirrors, slew symmetry can be trimmed, which demonstrates this effect as shown in Figure 8.

Here the THD performance of a 301A op amp with a trimmed SR of 0.4V/μS is plotted, and the data indicates an fp of 6.7kHZ which agrees with the theory. For assymmetric slewing however, the distortion generated is much higher and the break point occurs much lower in frequency. This sort of behaviour can be noted in many amplifiers, and those in which slewing is inherently assymetric will not yield as low a distortion as even slower devices which are symmetric. Assymmetric slewing is caused by an assymmetrical transconductance curve and leads to much 2nd harmonic distortion. The 2nd harmonic will rise in amplitude before the 3rd does and is thus detectable at lower levels.

Slew Rate and THD

An interesting demonstration of the effectiveness of slew rate improvement on THD is contained in Figure 9. This data is for the 2720, a programmable IC op amp, where SR can be adjusted via a bias terminal. Shown here is the resulting THD for SR of 0.5, 1.6 and 5V/μS respectively. As can be readily noted, the resulting performance improves directly as SR is increased.

Since the previous examples have indicated a quality of performance directly tied to slew rate, it might seem fair to assume that a very high slew rate is sufficient in itself to achieve this quality. This is not completely the case however, as shown by Figure 10.

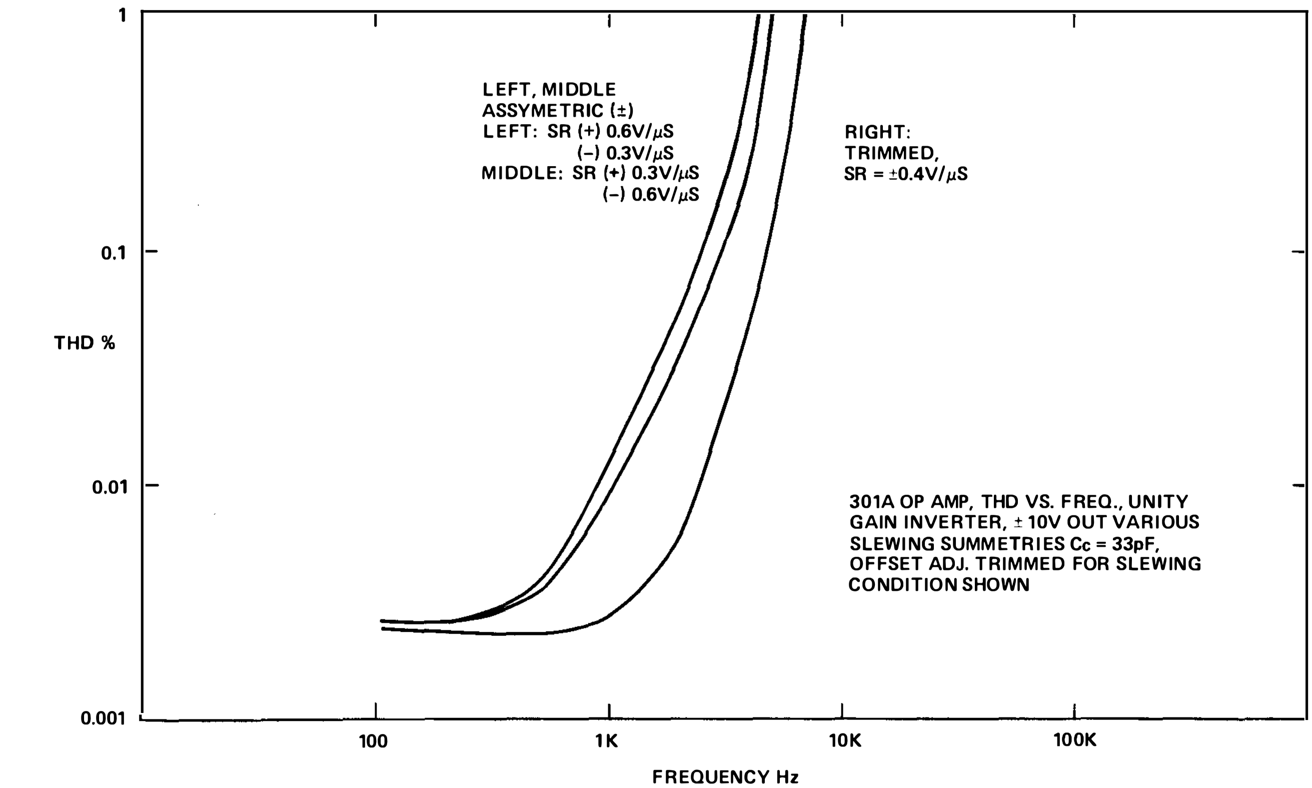


FIGURE 8

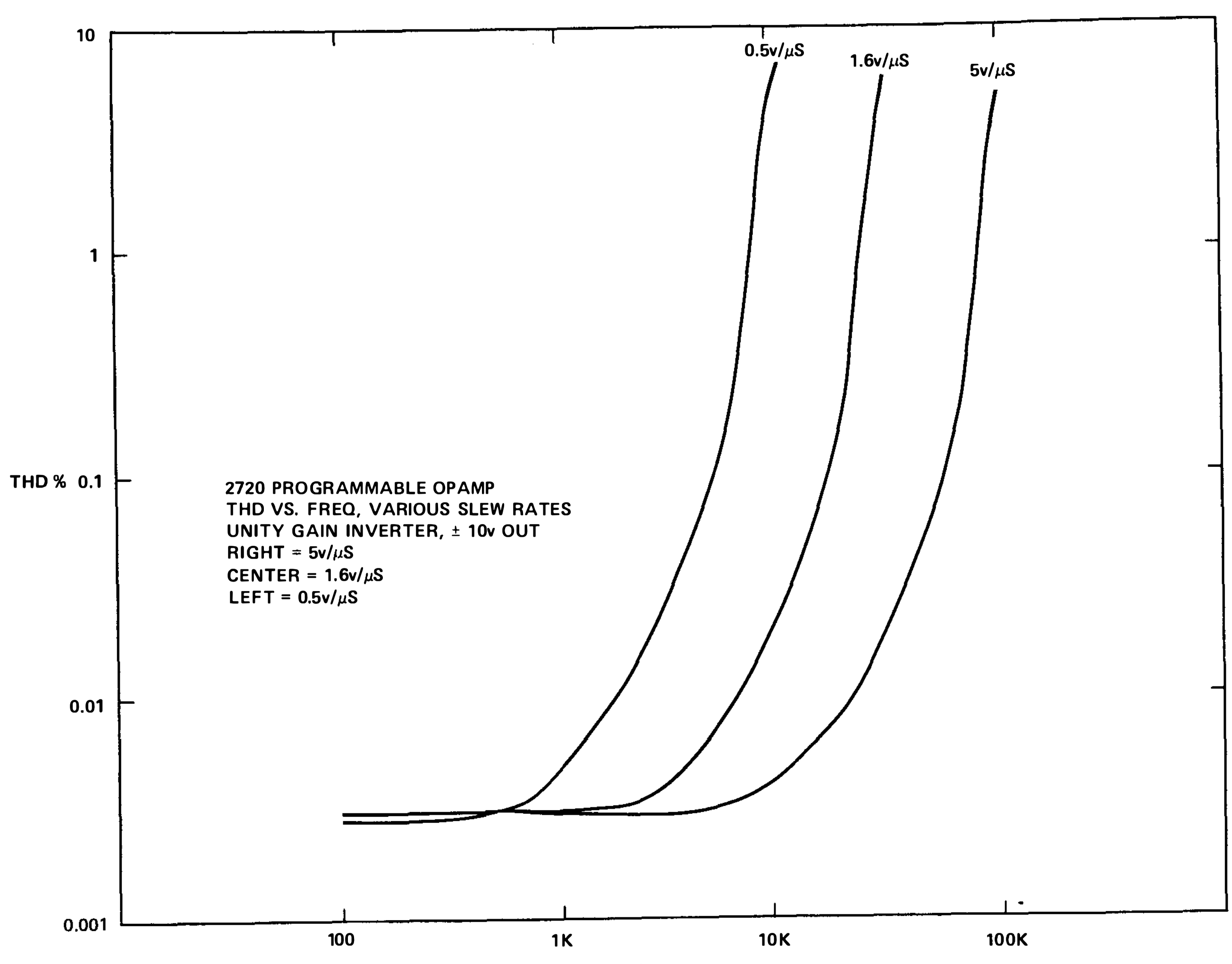


FIGURE 9

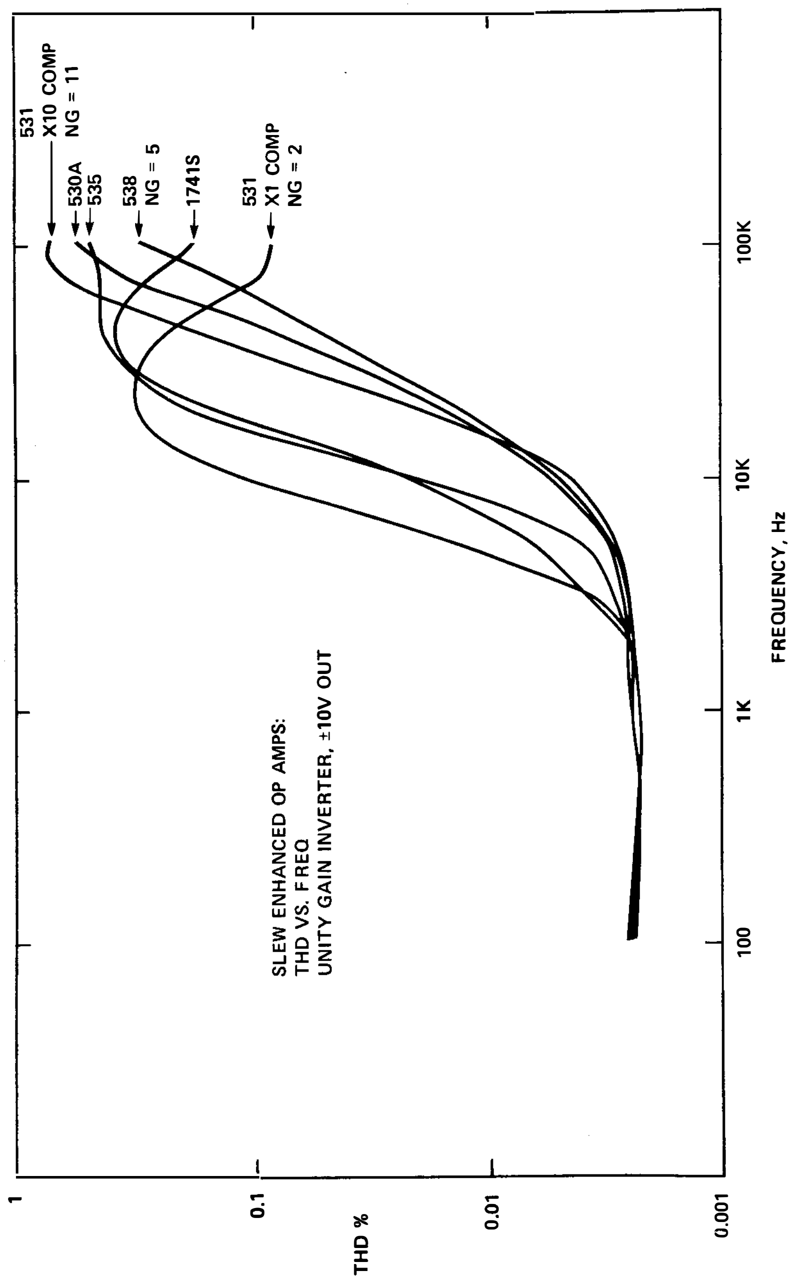


FIGURE 10

This data is THD performance for a class of op amps known as "slew enhanced" types. This form of op amp uses a class B (or AB) input stage to dynamically increase the output current, and thus SR.

In terms of performance, slew enhanced units generally show a low level distortion performance like a conventional op amp up to a point, but complete slew limiting is prevented. The data reflects this, but also shows substantial differences in performance for the various devices tested. Highest performers are those units which show the best low level linearity, and highest GxBW; ie. the x10 531, the 530A and the 538.

At this point, data has been shown which reflects the key behaviour patterns observed in the group of IC samples tested. In general, if the device slew rate is 5V/μS or more, is symmetrical, and does not use slew enhancement, the THD performance will be superlative. This will be evidenced by a THD of 0.01% or less up to 20kHZ, and for the best devices, 0.1% or less up to 100kHZ. Of those tested the best devices in the above terms were: NE5534 (equivalent to TDA 1034) 536, 318, 518, TL084, 3140, 2620, 2525, 301A (feed forward) and the OP-01, Nearly as good were the AD540 and 8007. The common characteristic of all of these amplifiers is their high slew rates; all are 5V/uS or more.

Two-tone HF IM Tests

The second series of tests conducted on the sample group of IC op amps was HF two-tone difference IM, hereafter called simply IM. This type of test also shows SID, as evidenced by IM, to be governed by amplifier slew rate. For these test a 1:1 mixed high frequency tone pair at full output level is swept from 10kHZ to 50kHZ. The difference frequency is maintained @100HZ. All tests were performed in the test ciruit of Figure 5.

Figure 11shows data which indicates the relationship of IM performance and SR. This data was taken with the 2720 programmable op amp, with slew rates of 0.5,1.6 and 5V/µS, conditions similar to Figure 9.

The nature of the IM performance behaviour strongly resembles the data based on THD, showing a similar rise as slew limiting is approached. This behaviour pattern is a characteristic one of IM, just as it is for THD.

Figure 12 shows a composite plot of IM performance for a variety of different IC op amps. The highest performance devices here show data which is at the equipment residual level, while the others show quality generally proportional to slew rate. The notable exceptions to this are the 535, a high speed slew enhanced type, and the 356 an assymetric slewing unit. Both units have high slew rates, but the method of achieving it prevents optimum linearity.

The data from the IM tests follow the same general pattern as THD based data. It is less sensitive, though, due to the fact that it measures even order products and the amplifiers usually (if perfectly symmetrical) generate odd order. This test is quite effective in pinpointing amplifiers which are assymmetrical such as the LF356. A two tone IM test to measure odd order products (2f1-f2) would yield more useful data on the symmetrical devices.

TIM Tests

A selected sampling of devices which had undergone the THD and IM tests were then subjected to the TIM tests as outlined in reference 16. Like the previous tests, the test circuit of Figure 5 was used in these tests. Our results do not directly correlate with those of Reference 16 because we are operating the amplifier with no common mode swing in order to isolate the SID distortion from common mode distortion. Figure 13 summarizes the results of these measurements, for full level tests performed with a 30kHZ square wave band limit.

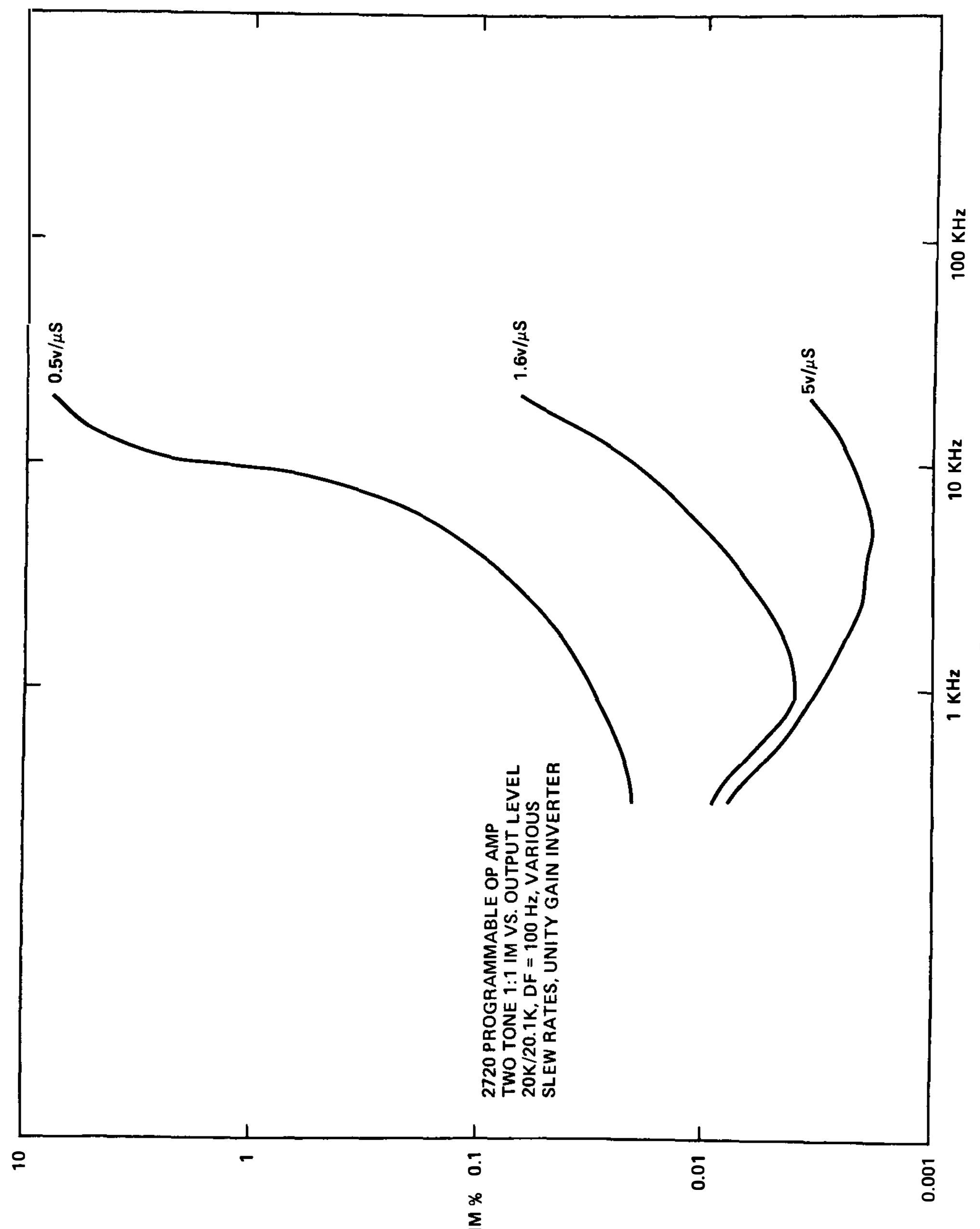


FIGURE 11

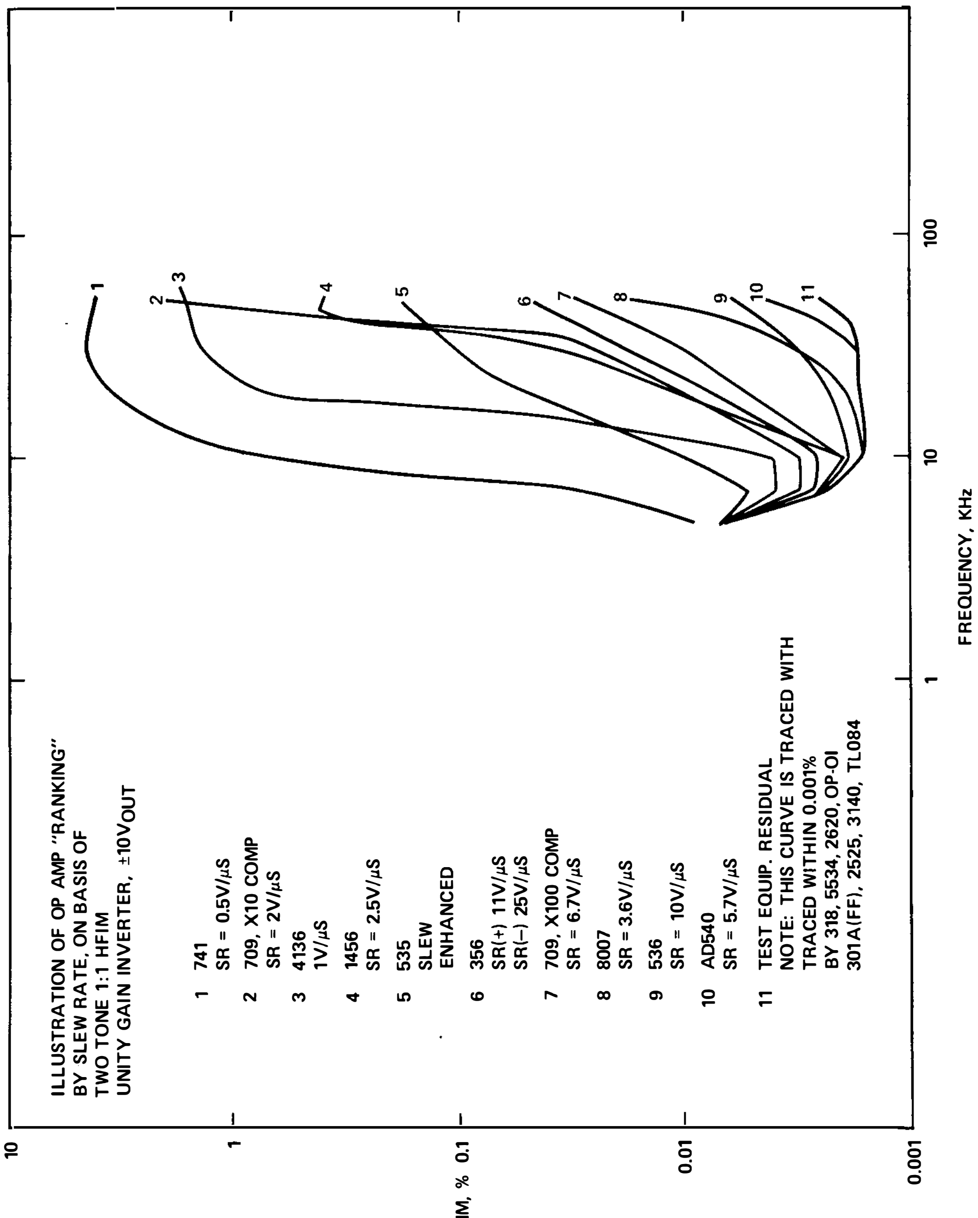


FIGURE 12

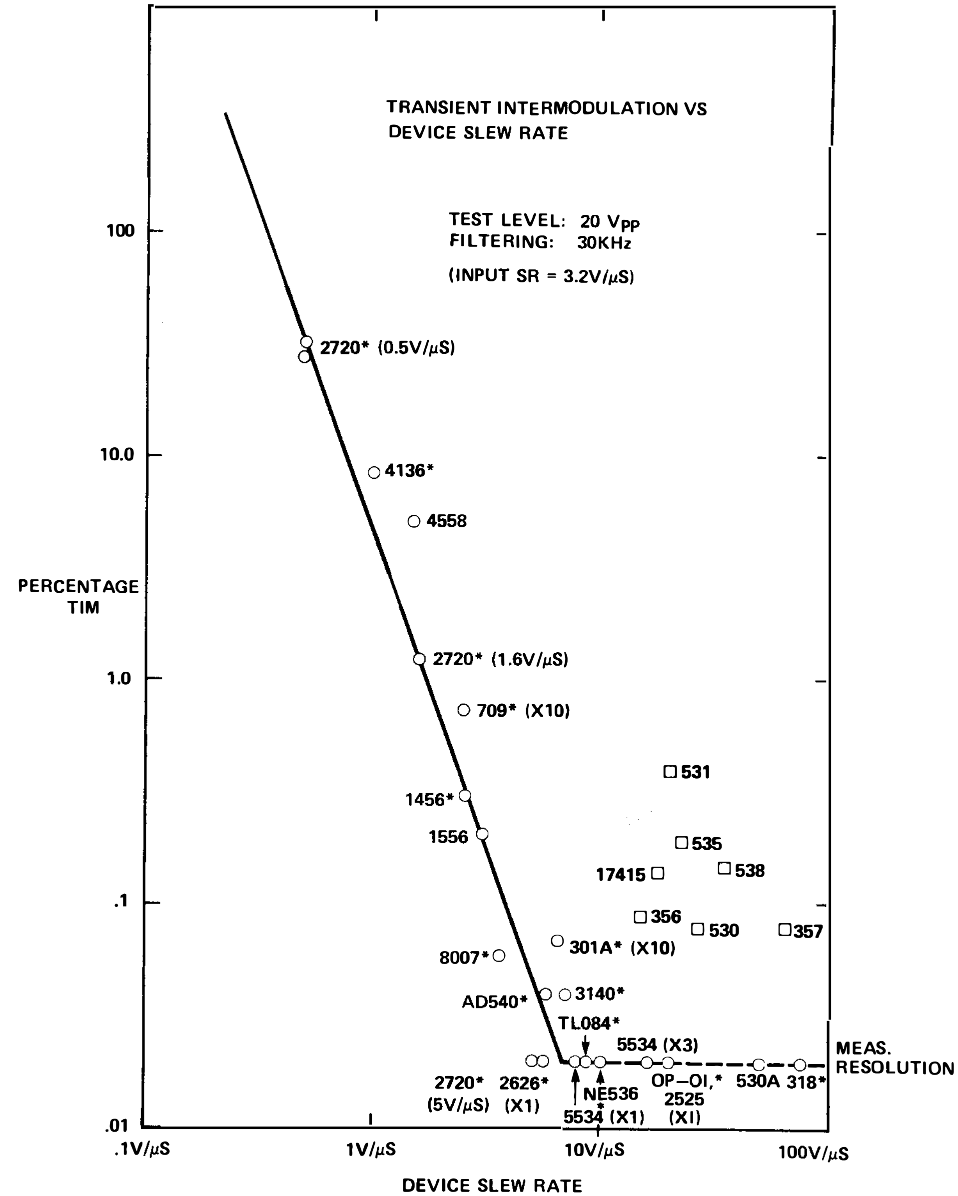


FIGURE 13

The relationship between transient (or dynamic) intermodulation distortion (DIM) and device slew rate capability is clearly exposed by the graph in Figure 13. This graph shows percentage DIM versus device slew rate for all types of devices under one standard test condition. The maximum slew rate of the input sine-square for this case is 3.2V/µS. Thus, the device would have to have a slew rate of at least this much to pass the waveform with unmeasurable distortion. The graph shows that distortion rises above the resolution level around 6.5V/µS, which is roughly twice the slew rate of the input waveform. This indicates that "on the average" a device must have at least twice the slew rate of the input signals to pass them with negligible distortion. As the slew rate capability of the devices falls below 6.5V/µS the graph is seen to rise linearly to very high amounts of distortion. A "best" straight line drawn through the data points turns out to have a slope of 3:1 on the logarithmic coordinates. This indicates that DIM varies as the third power of the ratio of the input slew rate to the device slew rate. A simple equation that expresses this relationship would be

$$\%\ DIM = K\left[\frac{SR\ of\ signal}{SR\ of\ device}\right]^3 \qquad (15)$$

where K = 0.16% for our data

This relationship is extremely important to audio designers as it indicates how transient intermodulation varies with the input signal levels.

It should be noted from Figure 13 that there are devices that do not fit the characteristic straight line relationship between distortion and slew rate. These devices are grouped to the right of the line and generally show excessive distortion for their high slew rate capability. With the exception of the Bi-FET devices (356,357), all of these are slew-enhanced op amps. They feature an input transconductance that varies with level to produce rapid slew rates for large signals.

Unfortunately, the changing transconductance gives rise to a crossover type of distortion mechanism. Since, for small signals their slew rate capability is low, they begin to produce distortion for relatively slow waveforms. As the speed and amplitude of the input is increased, the performance of the device gets better, and it is more capable of producing the required output. Thus at high slew rate inputs, the distortion doesn't increase, it merely remains the same percentage as it was under low slew rate conditions. We found that under varying input slew rate waveforms, the output spectrum of the slew-enhanced devices remained fairly constant, only the relative magnitudes of the individual distortion products varied up and down. Increasing the input slew rate caused some distortion terms to increase, and some to decrease, but the magnitude remained fairly constant. It is interesting to compare this behaviour with the leveling off of THD observed in the THD tests.

The BiFET devices also did not fit on the characteristic straight line, but they suffer from a different type of problem than the slew-enhanced circuits. The Bi-FET's only showed even order distortion falling on the square wave harmonics. No other intermodulation products were produced as the slew-enhanced devices did. The Bi-FET devices seem to alter the symetry of the waveform, indicating that some kind of lop-sided non-linearity is in action. This theory is supported by the basic slew rate of the 356 which is 11 V/µS positive and 27V/µS negative. The problem experienced by the Bi-FETs is not inherent in all FET op amps, by any means. The 536, an older design, had DIM levels below the resolution of our measurement equipment.

Devices which are capable of differing slew rates, such as the 2720 and 301A, show TIM performance which improves as slew rate is increased. To examine the effects of open loop bandwidth and the degree of feedback as design criterions for low TIM, several specific tests were performed. The results of these are the spectrum plots shown in figures 14, 15.

Figure 14 shows comparative performance for two different op amps for conditions of 10V output and a 30kHZ band limit. The 0.8V/uS device (a 741) clearly shows strong TIM, but the 10V/µS device (a 536) shows a spectrum which is indistinguishable from the input. Open loop bandwidth of both devices is less than 20hZ, feedback is nearly 100dB at low frequencies, and GxBW is 1 mhz.

Figure 15 shows a performance comparison for 20V, 30kHZ band limit conditions, with slew rates adjusted to 0.5,1.6 and 5V/µS using the 2720 device. It is clear that TIM is reduced as the slew rate is increased. For these conditions, device open loop 3dB bandwidth is for all cases less than 200HZ, and feedback is nearly 100dB at low frequencies.

It is apparent from these two tests and others made that the TIM test performance is strongly effected by slew rate, just as is THD and IM. There is no directly measurable or obvious sensitivity to open loop bandwidth. Gain bandwidth product and loop gain (feedback) effect TIM performance, as they do THD and IM in that they effect how close to slew limit one can work before distortion rises.

A further demonstration of how TIM behaves similar to THD and IM performance is contained in Figure 16. This data is based on the common condition of a 30kHZ band limit but with TIM plotted versus output amplitude. To show the similarity, two different slew rate devices are used, 0.5 and 1.5V/µS. At low signal levels TIM is at a very low level; as the output signal level is increased, TIM shows a rapid rise, similar in behaviour to THD and IM.

Comparison of Tests

If these three test methods are compared on a common base, it is possible to see a definite pattern in their behaviour. This is shown in Figure 17. For this

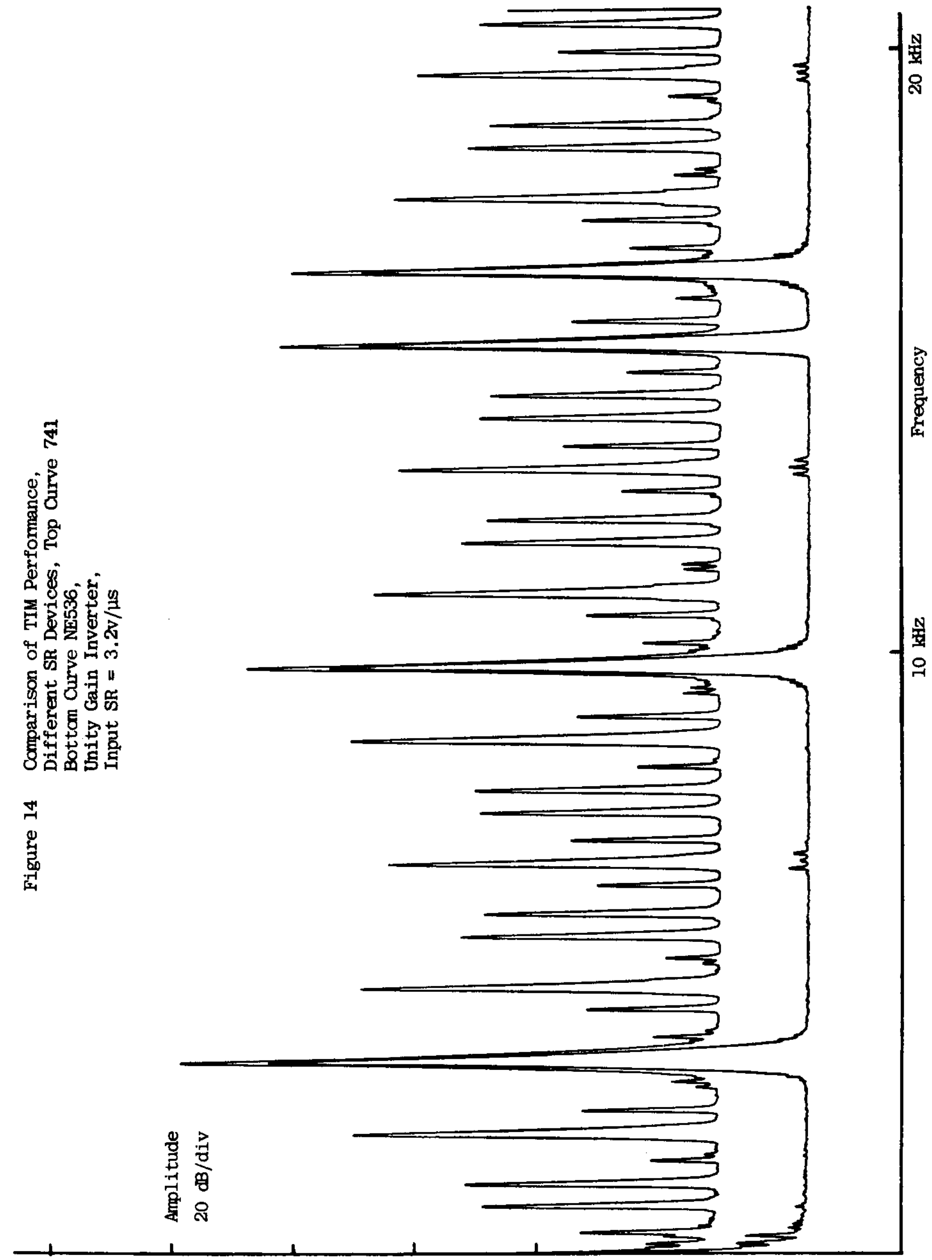
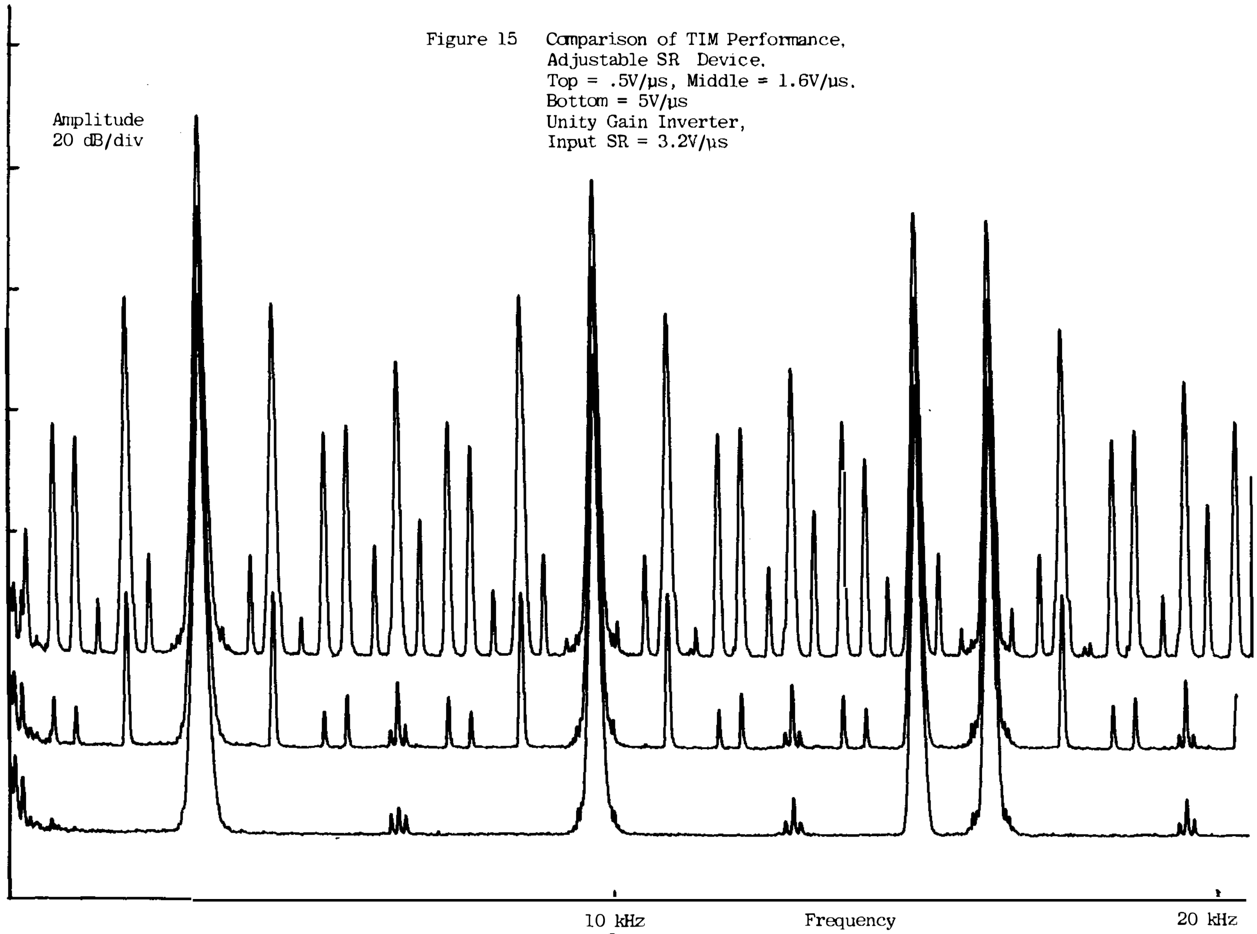


Figure 14 Comparison of TIM Performance, Different SR Devices, Top Curve 741
Bottom Curve NE536,
Unity Gain Inverter,
Input SR = 3.2v/µs

Figure 15 Comparison of TIM Performance, Adjustable SR Device. Top = .5V/µs, Middle = 1.6V/µs, Bottom = 5V/µs Unity Gain Inverter, Input SR = 3.2V/µs



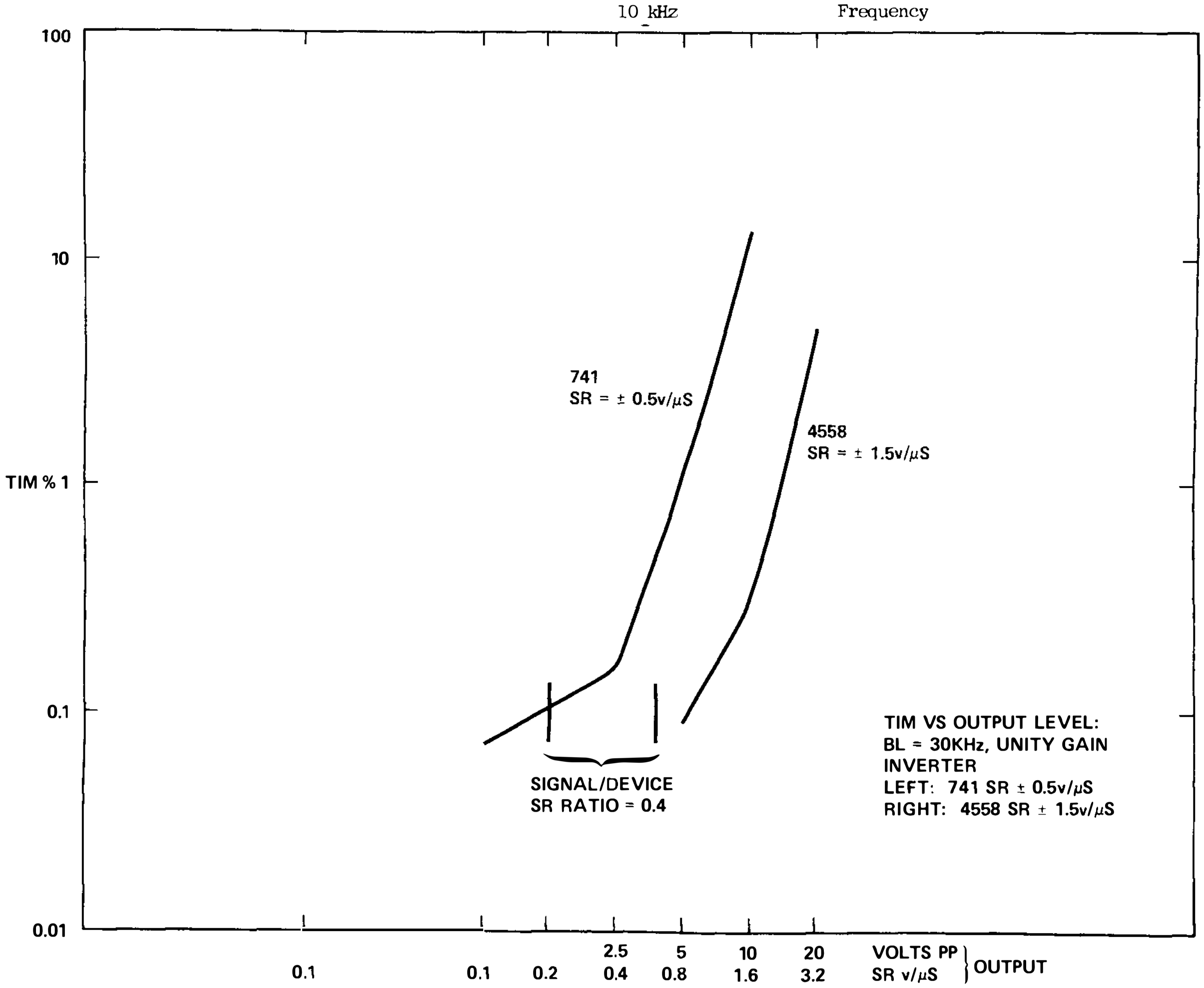


FIGURE 16

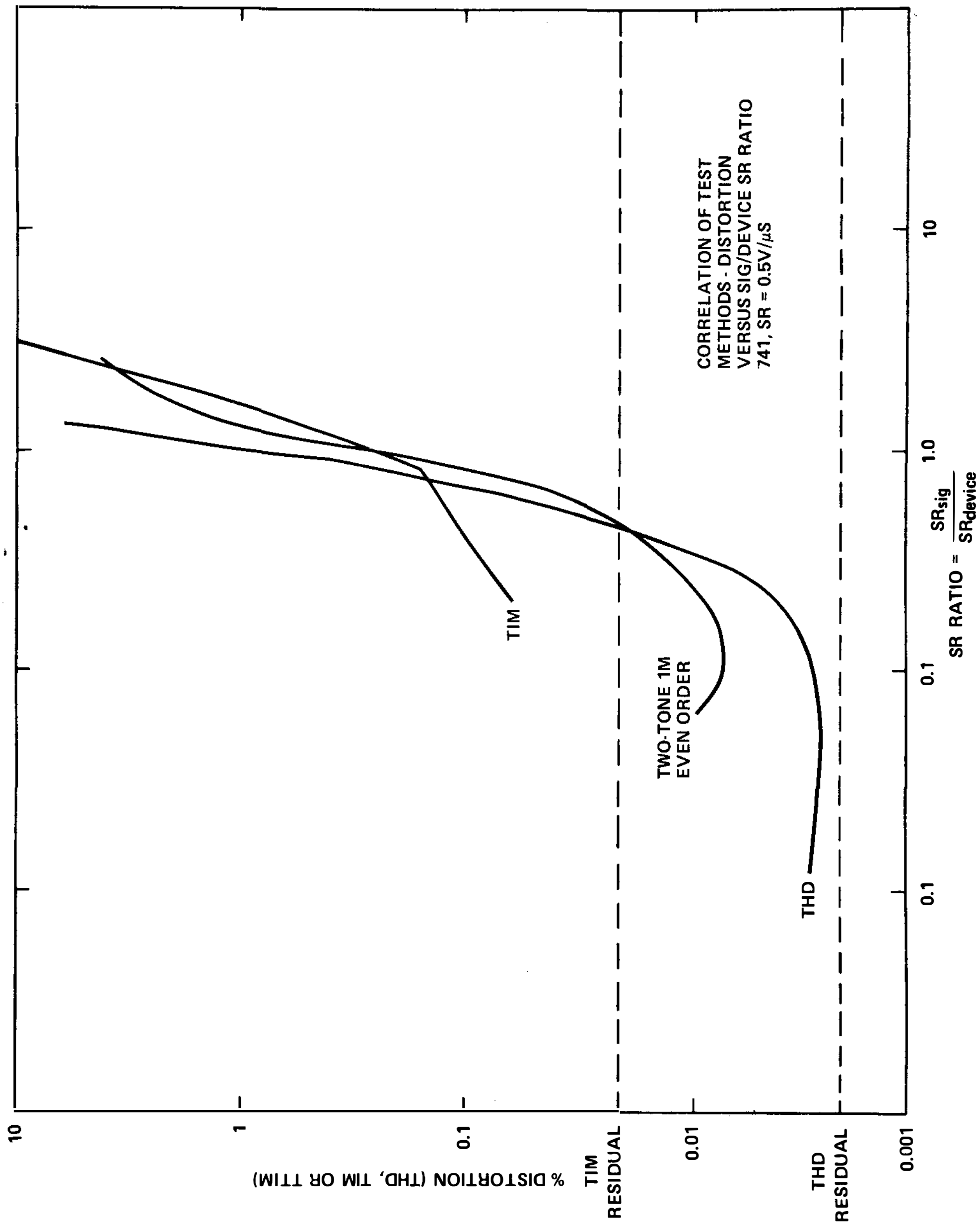


FIGURE 17

figure, the horizontal axis is normalized in terms of the ratio of the signal slew rate to that of the device. By this means it is possible to see just how the various forms of distortion behave as the device slew is taxed, and also to indicate the relative sensitivity of the three test methods.

The THD method shows the widest dynamic range of the 3 methods, and gives the highest percentage distortion at unity slew rate (1%.) The anomalous slope for the TIM test is due to our detection of some 2nd order low level non-linearities in the 741 tested. This produced a 2nd harmonic of the square wave which we were able to detect in the output spectrum. Since the TIM distortion number is normalized to the 15khz sine wave amplitude, and the square wave amplitude is 12 dB larger, the distortion shows up a factor of 4 larger than it should. Our experiences showed that it was very difficult to detect SID with the sine-square TIM test at slew rates under 1/2 of the maximum.

Unfortunately, there is a serious problem with the sine-square test method, which became apparent after evaluating some of the best op amp circuits. The problem concerns amplifier distortion products which are coincident with the even order distortion products of the square wave generator. Theoretically, a square wave should consist only of odd order harmonics of the fundamental frequency. Practically, every generator has a very slight assymetry in its square wave output, which creates small but definitely measureable amounts of even order distortion. Typical amounts for a general purpose square wave generator are 50 to 60 db down from the fundamental. Thus, if one was measuring a very good amplifier that had only small even order distortion products falling on the square wave harmonics, the true distortion of such a case would be masked by the generator and therefore unmeasureable. The conclusion might then be erroneously drawn that the amplifier was free from transient intermodulation distortion, when actually the amplifier was producing small amounts of distortion below the threshold of measurement.

One might question at this point that any amplifier that would produce distortion products coincident with the square wave harmonics should also produce other intermodulation products of comparable magnitude, that could be readily measured. This simply is not the case and can be easily demonstrated, by testing a 356 or an 530A. Both of these amplifiers show only even order square wave products, even at the severest slew rate test. To accurately measure these two devices, a square wave generator with even order products down at least 90db is required. In this series of tests, this was obtained by carefully adjusting the symmetry of our square wave generator at periodic intervals. Only when we reduced the generator's even order distortion did we begin to see differences between the best op amp circuits, that typically had only even order distortion products. The magnitude of these even order products for the best circuits were from 0 db to 6db greater than the generator residuals, and in many cases required detailed comparison of the input and output spectrum over several runs to verify that the products were in fact actually there.

The two tone difference IM test is much more sensitive to even order distortion than the sine-square test. Where it was difficult to detect distortion in the 356 with the TIM test, the IM test found it easily (Fig. 12.) It is probable that a two tone IM test set up to look for odd order products would show superiority for finding odd order distortion products. The main attraction of the TIM test is that it allows a quick qualitative look at an amplifier's performance; if that is, one has a spectrum analyzer handy.

THD evolves as the most desirable test method. It is sensitive and equipment is common. But, when a limited bandwidth circuit is being evaluated, one must use some form of IM test. It appears that maximum sensitivity can be obtained with the use of only two tones to isolate a given product.

A very simple test to determine if an amplifier is approaching slew rate is to look at the phase shift through the amplifier as a function of amplitude. The phase shift of an amplifier should not depend on amplitude, only frequency. However, when an amplifier approaches slew limiting the phase shift will change [29]. Implementation of this test is simple. Pass the highest frequency of interest through the amplifier and monitor the phase shift as the amplitude is raised. If the phase changes at large amplitudes, then SID is being produced. To date no work has been done to evaluate the sensitivity of this test method for audio amplifiers operated below the slew limit.

From this data it can be concluded that SID is a relevant factor and easily measurable evidence of it is produced beginning (for low GxBW devices) at as low as 20% of the devices slew rate, or at a slew rate ratio of .2. High frequency THD is a simple method of measuring SID.

Calculation of Slew Induced Distortion

Thus far, little has been said in the literature about how to calculate slew-induced or transient intermodulation distortion. This is no doubt due to the complexity of the problem, especially handling the frequency dependence of the amplifier stages and the incorporation of feedback. There is however, a straight forward technique that can be used to find closed form expressions for every possible harmonic or intermodulation distortion component. The technique involves forming a Volterra Series to characterize the output as a function of some input variable [(28)]. The coefficients of the Volterra Series can then be used to find the magnitude and phase of all distortion products. This technique has been widely used to predict distortion in radio frequency circuits with a high degree of accuracy.

Unfortunately, it takes more time and space to explain the technique itself than it does its application to a given problem. For this reason, we have decided not to present a full analysis at this time. However, with appropriate assumptions and simplifications, many useful features of the Volterra Series' technique can be used to find approximate expressions for SID. These are conceptually easier to understand and are quite accurate for relatively small distortion conditions.

Consider a 741 type operational amplifier, which can be broken down into two basic stages, an input transconductance amplifier, and an integrating amplifier. These are shown in Fig. 18.

The transconductance stage is assumed to be the dominant non-linearity and consists of a symetrical saturating type of characteristic which is independent of frequency.

The non-linear characteristic (formed by a double differential pair) is modelled as a current source output Δi, for an input differential voltage ΔV, and can be represented by equation (16)

$$\Delta i = Ik \tanh\left[\frac{\Delta V}{4Vt}\right] \qquad (16)$$

where $Vt = \frac{KT}{q} \cong 26$ MV at 300 ^{O}K

Ik = bias current of stage

The graph of equation (1) is shown in Fig. 19.

Equation 16 and Fig. 19 differ from equation 10 and Fig. 26 in our previous example because the 741 input stage has a pair of transistors on each side. Equation (16) in its present form will not allow closed form expressions for distortion. It must be expressed as a truncated power series with variable ΔV, to complete the calculations. This is shown in equation (17).

$$\tanh x = x - \frac{x^3}{3} + \dots + \dots \qquad (17)$$

Thus combining (16) and (17) we have

$$\Delta i = Ik \tanh\left[\frac{\Delta V}{4V_t}\right] \cong Ik\left[\left(\frac{\Delta V}{4V_t}\right) - \left(\frac{\Delta V}{4V_t}\right)^3 \frac{1}{3} + \dots\right] \qquad (18)$$

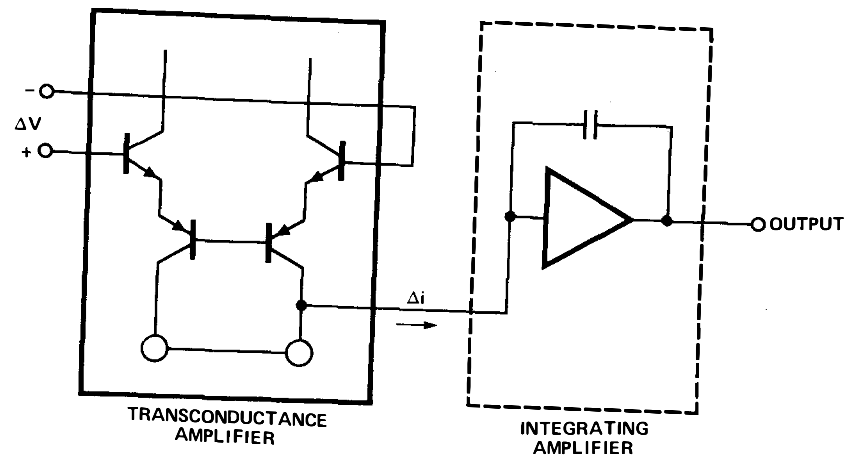


FIGURE 18

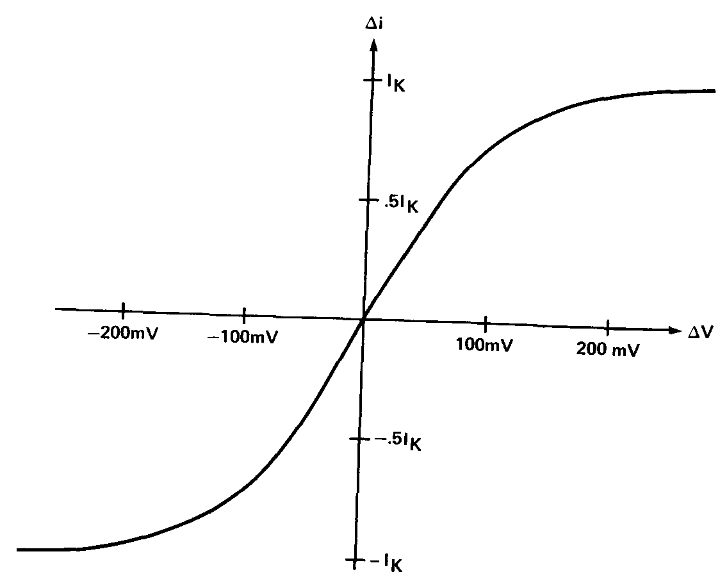


FIGURE 19

The first term in the power series is the desired linear component, and the cubic term (and other higher order terms) form undesirable distortion products. Distortion will eventually be calculated from (18) after making some additional necessary assumptions.

The second stage in the 741, the integrator, is assumed to be ideal, and have a gain characteristic G(f) which is proportional to 1/f. This is expressed by (19)

$$G(f) = \frac{K_2}{f} \tag{19}$$

The $\pi/2$ phase shift is neglected in (19) since only the magnitude is needed for the distortion calculation. The constant K_2 is determined by the overall gain of the composite amplifier which must be approximately unity at a frequency of 1 MegHz to make our circuit model represent the performance of a real 741 type op amp.

The actual gain characteristic of a 741 op amp is summarized by the Bode plot in Figure 20. For most audio frequency calculations, it is convenient to neglect the low frequency pole at 10 Hz, and to assume infinite dc gain and a constant gain-bandwidth product. This has a negligible effect on calculations since distortion is only affected by the magnitude of the gain bandwidth product. The open loop gain for this approximation is specified by equation (20)

$$\text{open loop gain} = \frac{V_{out}}{\Delta V} = \frac{10^6}{f} \tag{20}$$

By combining equations (20), (19) and (18), the constant K_2 can be expressed in more familiar terms. At a frequency of 1 MegHz we have:

$$\frac{V_{out}}{\Delta V} = 1 = \begin{bmatrix}\text{gain of} \\ \text{transconductance} \\ \text{stage}\end{bmatrix} \begin{bmatrix}\text{gain of} \\ \text{integrator}\end{bmatrix}$$

$$1 = \frac{Ik}{4V_t} \left[\frac{K_2}{10^6}\right] \tag{21}$$

$$K_2 = \frac{4V_t}{Ik} \times 10^6 \tag{22}$$

$$\text{And thus } G(f) = \frac{4V_t \times 10^6}{I_k} \times \frac{1}{f} \tag{23}$$

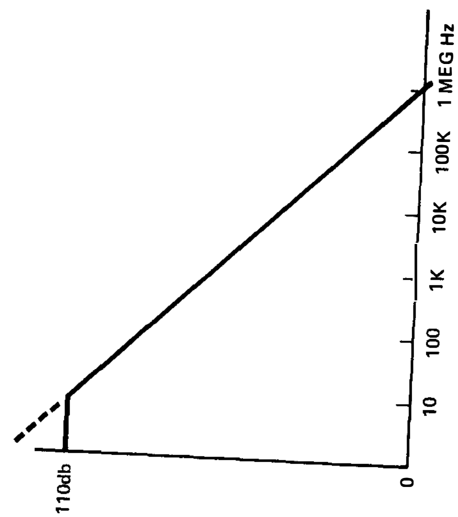


FIGURE 20

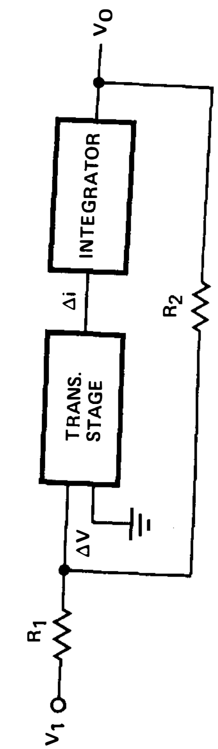


FIGURE 21

The 741 type op-amp that has been developed thus far, is now placed in an inverting gain configuration with resistive feedback components. The feedback network is assumed to be linear and independent of frequency. The circuit used for distortion calculations is shown in Fig. 21.

In this circuit, a feedback factor H can be specified as a function of R_1 and R_2

$$H = \frac{R_1}{R_1 + R_2} \qquad (24)$$

Since the closed loop gain is equal to R_2/R_1, we have

$$H = \frac{R_1}{R_1 + R_2} = \frac{1}{1 + |G|} \qquad (25)$$

For inverting gains of 1, 10, and 100 the factor H is 1/2, 1/11 and 1/101, respectively.

Additional assumptions that must be made to simplify calculations are:

1) Small distortion conditions exist (<1%.) This enables a power series expansion of the transconductance non-linearity.
2) The distortion only consists of odd order products because of symetry, and because of 1) the distortion is dominated by third order terms.
3) The distortion is reduced by the magnitude of the loop gain at the frequency of the distortion product.

A harmonic distortion analysis will be developed here to compare with measured data, although an intermodulation analysis could also have been pursued. The final result will solve for harmonic distortion (which is dominated by the third harmonic) as a function of output voltage level, frequency, and feedback factor (or closed loop gain.)

The following method will be used to solve for harmonic distortion. First, an output level Vo, and frequency f will be specified. Then using (20), ΔV will be calculated and used in (18) to find open loop distortion. Finally the loop gain will be computed and used to predict the closed loop distortion.

For a sinusoidal output voltage of Vo cos $2\pi ft$, we can compute ΔV from (20)

$$\Delta V = \frac{Vo \cos 2\ ft}{(10^6/f)} \qquad (26)$$

If this ΔV is substituted into (18) and simplified, the resulting equation will show an open loop distortion ratio of:

$$\frac{\text{magnitude of 3rd harmonic}}{\text{magnitude of fundamental}} = \frac{\left(\frac{V}{4V_t}\right)^2}{12} = \begin{matrix}\text{Distortion} \\ \text{(open loop)}\end{matrix} = \frac{1}{12}\left(\frac{Vo\ f}{4V_t \times 10^6}\right)^2 \qquad (27)$$

The open loop distortion is reduced by the loop gain at the third harmonic frequency, 3f, and by the integrator frequency response which attenuates the third harmonic by a factor of 3. The loop gain at frequency 3f is

$$\text{loop gain} = \left(\frac{I_k}{4V_t}\right) \times \left(\frac{4V_t \times 10^6}{I_k\ 3f}\right) \times H = \frac{10^6}{3f} H \qquad (28)$$

Therefore the closed loop distortion is

$$\begin{matrix}\text{distortion} \\ \text{(closed loop)}\end{matrix} = \frac{\text{distortion (open loop)}}{\text{loop gain}} = \frac{1}{3}\left[\frac{\frac{1}{12}\left(\frac{Vo\ f}{4V_t \times 10^6}\right)^2}{\left(\frac{10^6\ H}{3f}\right)}\right] \qquad (29)$$

$$\text{THD (3rd)} = \frac{Vo^2\ f^3}{12(4V_t)^2\ H \times 10^{18}} = \frac{Vo^2\ f^3}{1.29 \times 10^{17}\ H} \qquad (30)$$

Equation (30) shows that harmonic distortion should vary directly with the cube of the input frequency, directly with the square of output voltage, and inversely with the feedback factor H. In order to test the accuracy of this equation, calculated data for distortion was compared directly with measured THD data from a 741 amplifier. Figures 22, 23 and 24 compare calculated and measured distortion for a constant amplitude, swept frequency test condition, for three values of feedback factor H. Figure 25 compares calculated and measured distortion for a constant frequency, swept amplitude test condition, also for three values of feedback factor. The agreement is generally good and is excellent for the swept frequency tests. At lower distortion levels, the agreement deteriorates due to measurement resolution limits. At higher distortion, the agreement deteriorates due to large distortion conditions; that is, the fundamental assumptions in developing the calculation are violated. The anomalous behavior of the G = 100 test results are due to a low closed loop bandwidth of 10kHZ, and the absence of loop gain at these frequencies. Figure 25 also indicates same form of crossover distortion that dominates at low signal levels, and masks the true distortion characteristics. It should be clear from all the figures that increasing feedback reduces distortion.

The demonstrated accuracy of equation (30) in predicting harmonic distortion in a 741 amplifier, leads to some powerful conclusions concerning slew-induced distortion.

1) It means that slew-induced distortion can be modelled and calculated using straight-forward harmonic distortion techniques.
2) It emphasizes that there is nothing new, unique, or mysterious about slew-induced or transient intermodulation distortion.

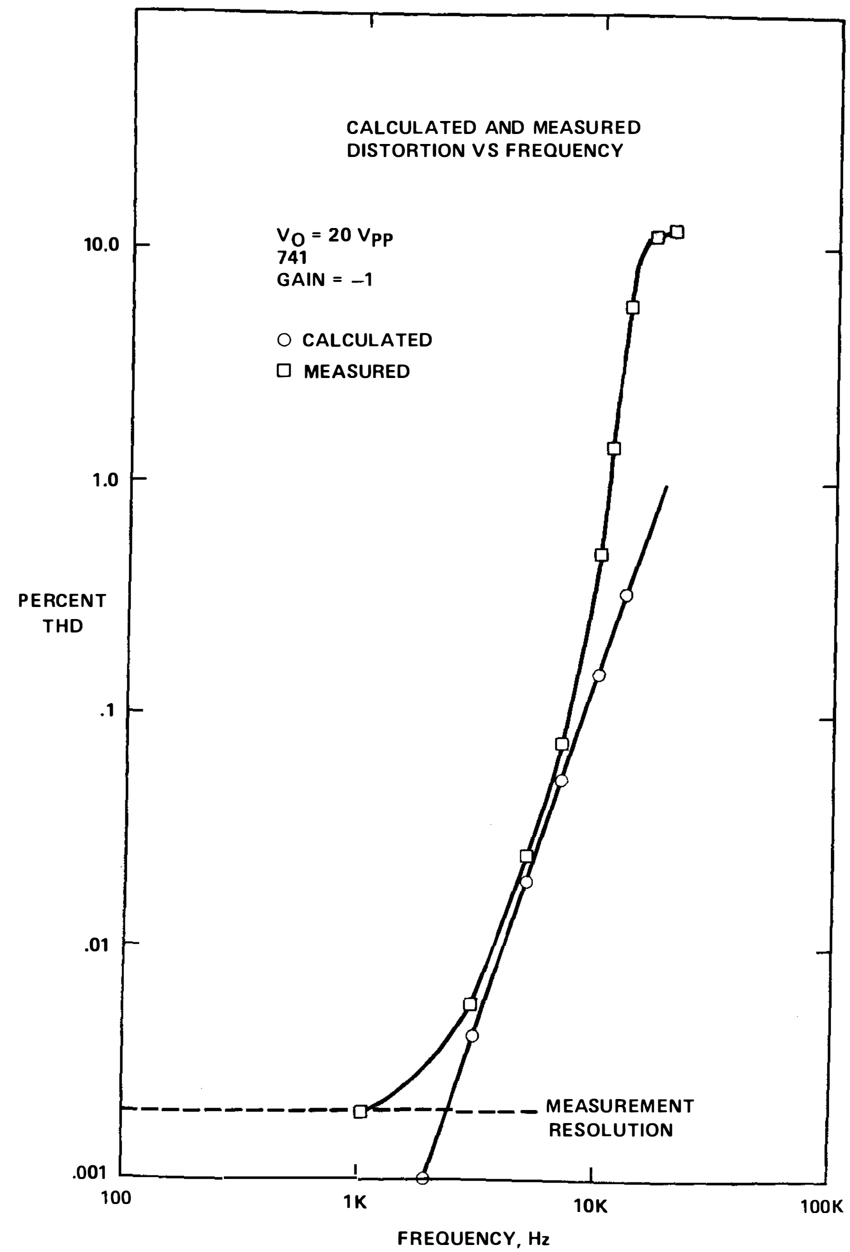


FIGURE 22

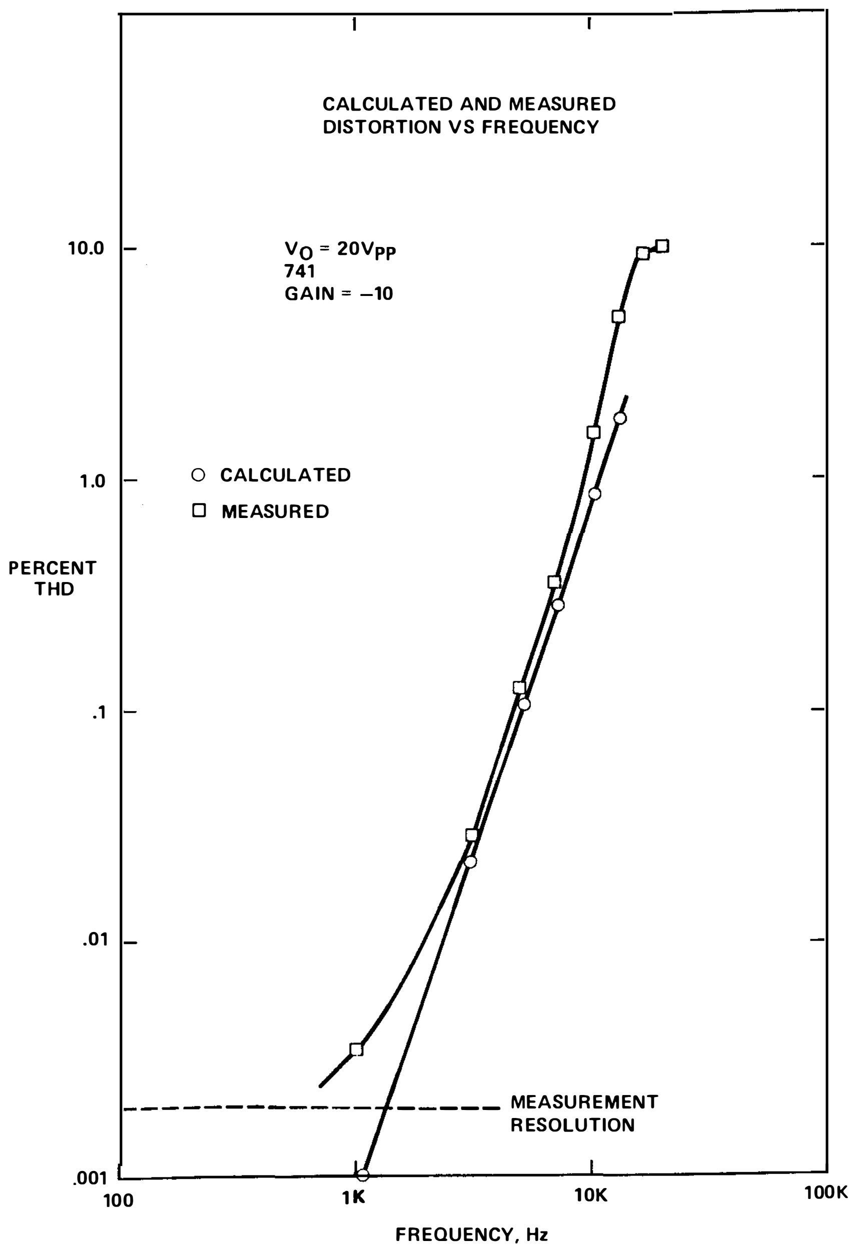


FIGURE 23

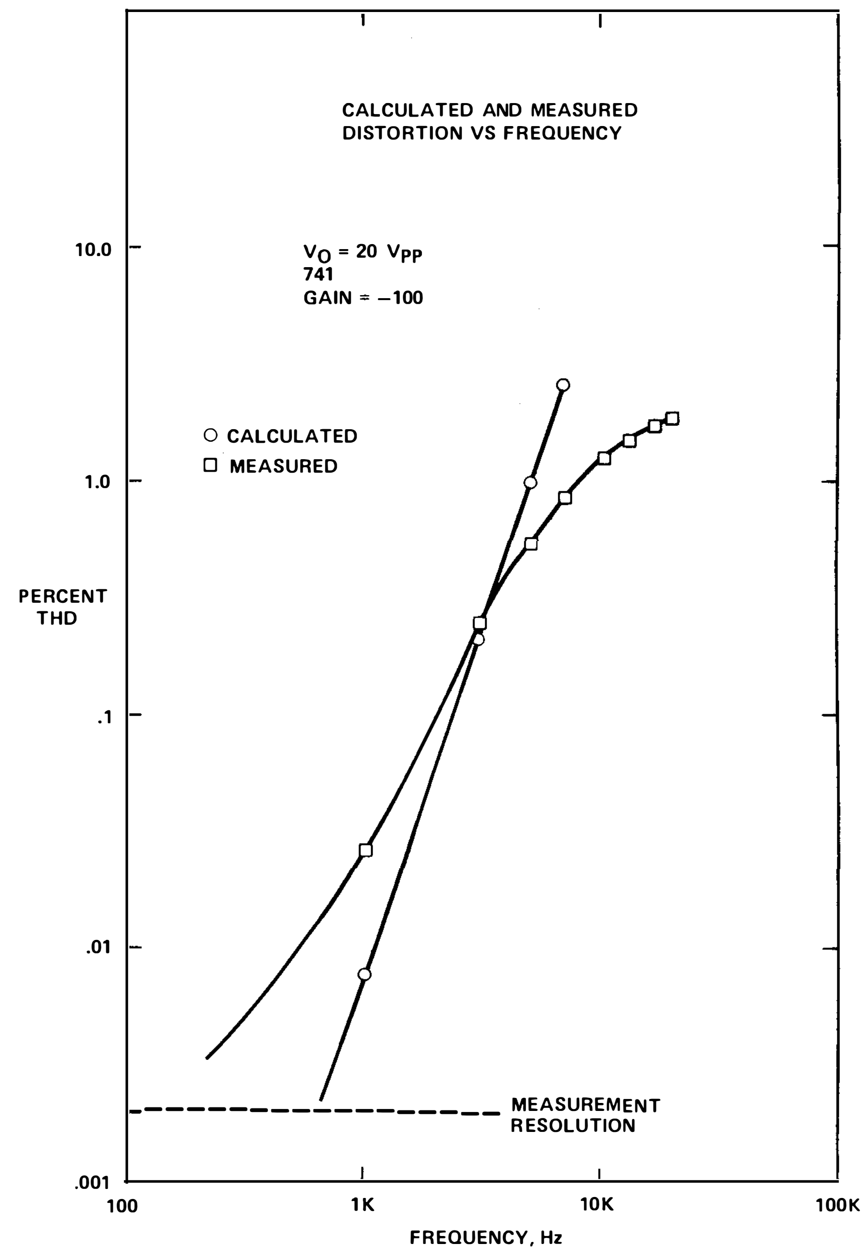


FIGURE 24

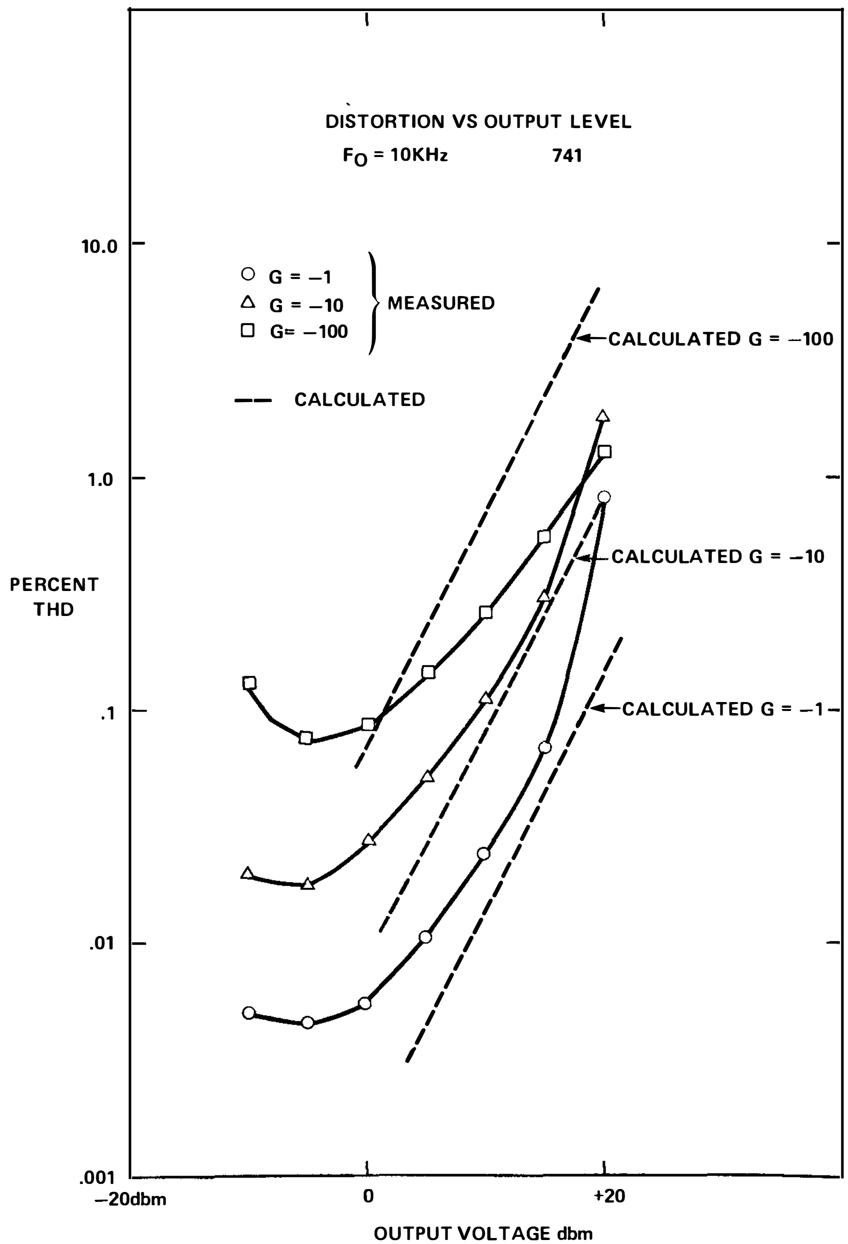


FIGURE 25

3) It shows that slew-induced distortion is increased by the sharpness of the non-linearity and decreased by higher gain-bandwidth products and larger feedback.
4) It demonstrates that since the slew rate of a constant amplitude sine wave is proportional to its frequency, that slew-induced distortion (or transient intermodulation distortion) should vary as the cube of the input slew rate. This is confirmed by the data in Figures 13 and 17, that show the variation of TIM with slew rate is a cubic relationship.
5) It indicates that since distortion can be predicted up to 85% of the intrinsic slew-rate limit of the device, that slew-induced distortion (or transient intermodulation distortion) is inextricably tied to the devices' slew limit. Those factors which cause signals to tax the slew capability of the amplifier, also increase distortion.
6) It shows that increasing a device's slew capability, without adding additional non-linearities, will reduce slew-induced and transient intermodulation distortion.

Present TIM theory suggests that feedback increases distortion. Our measurements and calculations show that, at least for signal conditions below the slew rate limit, that feedback reduces distortion. Actually the truth lies somewhere in between.

Increasing feedback reduces distortion provided the amplifier is operating below its slew-limit. For signal conditions above the slew rate limit, Figs. 22, 23, 24 show that more feedback will increase distortion. TIM measurements show this same pattern of feedback improving distortion below the slew limit and degrading it above the slew limit.

Listening Tests

These IC's were auditioned in a listening test to assess the degree of correlation between the various forms of electrical distortion and audible defects. These tests were done in mono, in an inverting configuration similar to Figure 2. To sensitize the test for SID however, the test device was preceeded by a preamp to drive it to near full scale output with program material. The output was then scaled down and level matched with the original input to within 0.2dB. A-B tests we then conducted on each IC to determine audible degradation.

The results of this test indicate that not only can SID be detected audibly, but that the ear is sensitive to very low levels of distortion. The results of these tests are summarized in Table 1, which also indicates the relative quality weighting.

"A" level quality is that indistinguishable from the source on the most difficult high frequency program material. In general devices of over 4V/µS slew rates fit into this category. Exceptions were some (but not all) slew enhanced devices, and the assymetric devices.

There are two broad categories of audible SID, one which can be associated with the approach of slew limiting, Category I; and one in which slew limiting actually occurs, Category II. The audible characteristic of the two are deterioration, and gross distortion, respectively.

Category II distortion will occur relatively infrequently on normal program material if the device slew rate is above 0.5V/uS. However, Category I distortion is possible in many instances and adjectives used to describe it have been seen in print often.

Design Guidelines

Some sensible design guidelines begin to emerge from this work. The primary one is speed-faster amplifiers are generally better. There are two

Table 1
Listening test results (referred to full output of ±10V)

Category of SID		I Deterioration		II Gross distortion	
Quality level	A	B	C	D	E
Audible character	No differences detected for any program material	Just discernible softening, loss of sweetness	Further softening, somewhat dry, generally satisfactory with slight loss of dimension	Colorations apparent, loss of dimension, "covered" sounds, dulled transients, constriction, edge begins	Coloration and distortion obvious, more constricted covered sound, transients smeared, grit edginess, fuzz
Associated slew rates	>4V/uS	2-4V/uS	1-2V/uS	0.5-1V/uS	<0.5V/uS
Samples Tested	318, 518	1456	1741S	741	
	TDA1034(5534)		356*		
	2625		4741		
	2525		535		
	8007		538*		
	NE536				
	AD540				
	3140				
	TL084				
	OP-01				
	530A	530			
	NE541(x100Comp)	NE541*(x10Comp)			
	NE540(x100Comp)	NE540*(x10Comp)			
	531(x10Comp)		531(x1Comp)		
	2720(5V/uS		2720(1.6V/uS)	2720(0.5V/uS)	2720(0.16V/uS
	301A(x10,x100,FP)			301A(x1Comp)	
		4136(2V/uS)	4136(1V/uS)		709(x1Comp)
		709(x10Comp)			

*Audible ranking possibly due to factors other than slew rate aspects to speed: bandwidth and slew rate. In general they tend to go up together. It can be confidently stated that raising an amplifier's GxBW, or ωu is desirable. The reason is that at any given frequency (neglecting D.C. and very low frequencies) the loop gain of the amplifier will be higher and more feedback related distortion reduction will take place, which lets one work closer to the slew rate limit.

It has also become apparent that higher slew rate is better, but some caution is required. Since slew rate is determined by the dynamic range of the non-linear input transconductance amplifier, it is important that high slew rate not be achieved at the expense of linearity. Some devices do achieve high slew rate at the expense of linearity. The slew-enhanced devices such as the 535 and the assymmetrically slewing 356 are examples of this. These devices are inherently incapable of performing as well as devices with more linear input stages. Emitter degeneration is an example of a technique that allows higher slew rate while at the same time linearizing the input stage. The 318 is a good example of this type of amplifier. FET input types are also excellent, provided they are symmetrical. A good example is the 536. For the same transconductance, FET input stages are linear over a much larger range than bipolar input stages.

To restate these design criteria, we primarily want an amplifier which is linear for large input signal (ΔV) levels. This gives us high slew rate and low open loop distortion. Secondly, we would like this amplifier to have as high a unity gain bandwidth (GxBW or ωu) as possible, so that when we apply feedback, the loop gain will be as high as possible for distortion reduction. The loop gain determines how close we can operate to slew limiting before

distortion begins to rise.

Some previously discussed design criteria for low TIM 3,4,7,9 such as the use of low feedback, low open loop (D.C.) gain, and a high open loop pole frequency (ωo) have no basis in fact or theory. More feedback increases the loop gain and reduces distortion. The location of the open loop pole (ωo) is of little significance to audio designers, and it does not have to be placed at a frequency above 20 KHz for superlative performance to be obtained. Our measurements support these statements.

From all of the above, it seems appropriate to adopt a new form of slew rate criteria for audio circuits. From the four series of tests (THD, IM, TIM, listening) this would be a criterion which specifies a minimum slew rate with regard to the maximum output voltage level in use. The criterion is:

> "The circuit, including all possible loading conditions, should posses a slew rate of 0.5V/µs (minimum) to 1V/µs (conservative) per peak output volt."

Application of this simple criteria will result in negligible SID, either electrically or audibly if the slew rate is symmetrical (± 20%) and the input stage has a smooth transfer characteristic (unlike the slew enhanced types.)

If large signals outside of the audio band are expected, it is wise to provide filtering to keep these signals out of the amplifier loop.[11] Otherwise these signals may cause the amplifier to approach its slew limit and generate SID.

Conclusions

One major result of this study is a much more clear assessment of the true behaviour pattern of operational amplifiers. Distortion has been analyzed qualitatively, quantitatively, and theoretically. This information can be applied by designers without the fear of violating arbitrarily contrived design rules relating to open loop D.C. gain or feedback factors.

Another major result is that tests for slew rate and SID are not only appropriate for audio gear, they are absolutely essential. Hopefully these terms will soon appear in both product specifications and test methods, rather than such terms as TIM, which are not only misunderstood, but technically incomplete and creating a great deal of confusion in the popular press[9,32,33,34] with but a rare example of understanding [31].

There is nothing "transient" about SID. It will occur continuously with steady high frequency tones. The fact that this distortion occurs in musical transients is due to the nature of the signal, and not the distortion mechanism.

As a final point, we feel there is still much to be learned about distortion mechanisms, measurement techniques, and perception. We consider this study but one step in that direction.

Acknowledgement:

This paper was based in part on the article series, "Slew Induced Distortion in Audio Amplifiers," published by The Audio Amateur, Box 176, Peterborough, New Hampshire 03458.

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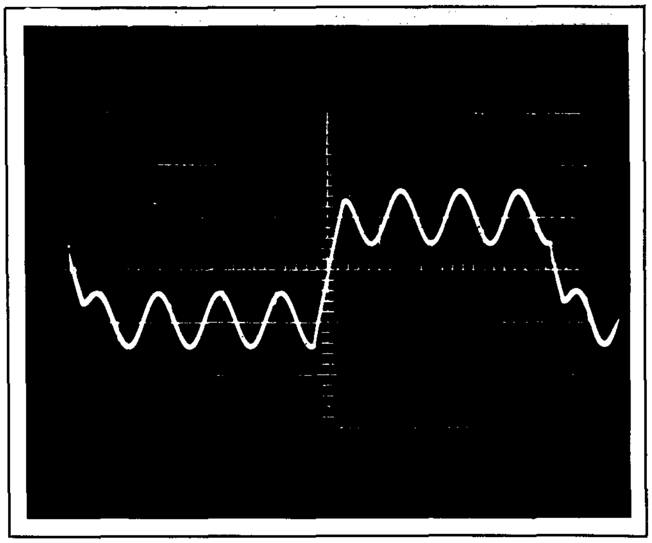
An Overview Of SID and TIM

Walter G. Jung, Mark L. Stephens, and Craig C. Todd

PART I



In this series of articles we hope to shed useful light on the high-frequency performace of amplifiers. Modern operational amplifiers and circuits of similar topology have an inherent Slew Rate (SR) limit, and they will produce distortion as the output Signal Slope (SS) approaches this limit. We refer to this distortion as Slew Induced Distortion (SID). If an amplifier is driven into slew-rate limiting gross distortion will be produced. This is analogus to driving an amplifier into amplitude clipping, which also produces gross distortion. The distortion produced by driving an amplifier towards slew-rate limiting has also been described as Transient Intermodulation distortion (TIM) [3,8,9,17,18,51,56].

Until recently [33,34] there has not been a thorough study of this distortion. Therefore, this series is intended to be a comprehensive overview and explanation of SID. We will explain how and when SID is produced by an amplifier, and measurement techniques for and typical measurements of this distortion will be described. The results of a listening test for SID will be discussed, and the results of a theoretical calculation of SID in a 741 op amp will be shown and compared with measurements. Some reasonable design criteria will also be reviewed. Above all, we will attempt to give a good overall perspective of this subject so that the reader will be able to judge its relevance to his or her own situation.

Before discussing how SID occurs within amplifiers, it is necessary and appropriate to first consider how the slew rate itself is related to an audio signal. A sine-wave audio signal has definite and measurable parameters, namely its amplitude and frequency. However, a somewhat more subtle parameter (and one germane to this issue) is the *slope* of the signal, as is determined by its amplitude and frequency. A simple relationship which defines the signal slope (SS) of a sine wave is the equation

$$SS = 2\pi Vpf \qquad (1)$$

where Vp is the peak signal voltage, and f its frequency.

Sometimes this equation may be seen written in terms of slew rate (SR) [20,21,22,29,30,54], however we wish to clarify the point here that signals *in themselves* have no inherent slew limit, or maximum allowable slope, as do amplifiers. Therefore, we will use the terminology of SS to describe the slope of a sine-wave (or other) signal and SR to describe the slew rate of an amplifier. Note that this is an important distinction, as an amplifier has a defined SR, which is (by very definition) its maximum output-voltage rate of change, or slope, as set by its design. It is a defining performance limit for that amplifier, just as power output is (or any other basic performance parameter, for that matter).

The reader should note that this equation may be manipulated into an expression in terms of a frequency (f), for a given signal slope and peak voltage; for instance:

$$f = SS/2\pi Vp. \qquad (2)$$

When the relation is thus used, and the particular SS under discussion is the slew rate limit of a given amplifier and Vop its peak output voltage, it would appear as

$$fp = SR/2\pi Vop. \qquad (3)$$

This expression yields a power bandwidth, fp, which is determined by the amplifier SR and the peak output voltage, Vop. Generally, fp is understood to be the bandwidth for a 1 percent THD limit. Note that fp is directly proportional to SR and inversely proportional to Vop. The practical significance of this is that high output-voltage amplifiers require more SR to maintain a given distortionless bandwidth.

Also, an important distinction to be made is that power bandwidth defines an entirely different form of bandwidth than does the more familiar small-signal bandwidth, and the two terms should *never* be confused. Exceeding the power bandwidth of an amplifier causes gross distortion; exceeding its small-signal bandwidth results only in a frequency response rolloff [37].

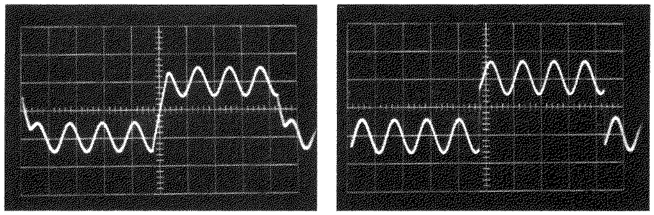
SID and TIM
Which is Which and What Do They Mean?

Unfortunately, many of the popular explanations serve to confuse rather than clarify the issue, and this short preparatory discussion will, we hope, clarify some of these points to the reader.

"TIM" stands, of course, for *transient intermodulation distortion,* sometimes called simply "transient distortion." If this name is taken in a literal sense, it implies a distortion

Portions of this article are adapted from "Slewing Induced Distortion in Audio Amplifiers" by the authors in *The Audio Amateur,* Feb., 1977 (P.O. Box 176, Peterborough, N.H. 03458), part of an article series which is available in book form. Portions were also adapted from the authors' article "Slewing Induced Distortion — Its Effect on Audio Amplifier Performance, with Correlated Listening Results," Audio Engineering Society Preprint No. 1252 from the May, 1977, convention. (See bibliography references nos. 33 and 34.) **© Copyright 1979 by Walter G. Jung, Mark L. Stephens, and Craig C. Todd.**

Fig. 1 — Mixed square/sine output from amplifiers with and without TIM. General conditions: 5-kHz square wave and 40-kHz sine wave. Fig. 1a — Strong TIM, sine wave missing on waveform transitions, slewing evident. (Scale: 10 V/div.) Fig. 1b — Little or no TIM, waveform is a linear sum of sine and square waves. (Scale: 1 V/div.)



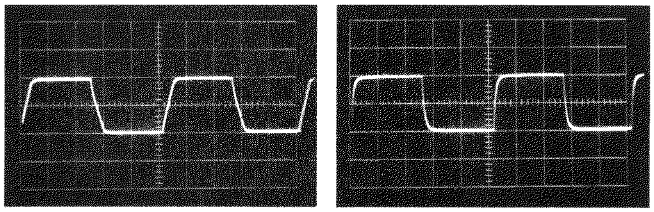


Fig. 2 — Amplifier square-wave responses with and without slew limiting. Fig. 2a — Slew limiting (10 kHz, 10 V p-p). Fig. 2b — No slew limiting (10 kHz, 1 V p-p).

mechanism which produces intermodulation when subjected to transients. A point to be noted is that if the term were understood literally, this would imply transients of *both high* and *low* frequencies *and/or* high or low operating levels. In other words, *all* transients.

In actual practice, however, transient IM occurs only for signals with simultaneous high level and high frequencies — not lower levels or lower frequencies. The key parameter of such signals is that they are characterized by *high signal slopes,* not just high frequencies or high levels. Neither high frequencies nor high levels in themselves *necessarily* result in distortion, unless their combination is such that a high effective SS is produced.

High SS waveforms are not confined solely to transient waveforms. It just so happens that musical signals which exhibit high signal slopes more often are transient in nature — a fortissimo cymbal clash, for instance. Thus, TIM is probably a descriptive term for the distortion as it occurs on *musical* waveforms, but the term is not totally descriptive of the distortion mechanism itself [33,34,44,52].

TIM is actually generated when the SS approaches or exceeds the amplifier SR. Thus, a more easily understood term as to what actually happens would be one which relates *both* to SS and SR. In an amplifier, distortion is produced when the output voltage SS approaches or attempts to exceed the SR, as the amplifier limits (clips) for such a circumstance. This can happen for either transient or steady-state signals [33, 34, 52] if they have a sufficiently high SS. Thus we feel a more descriptive term to describe the mechanism is Slew Induced Distortion [33, 58] as it is distortion induced either by the onset of or actual slewing. Other descriptive variations of this terminology are seen in print, such as "slew rate distortion" and "slewing distortion," and mean essentially the same thing [11].

Effect of Excessive Signal Slope On Amplifier Performance

A demonstration of the sensitivity of amplifiers to SS is contained in the two waveform photos of Fig. 1. Figure 1a shows a mixed square/sine wave signal combination, where the level and risetime of the square wave are such that the SS is greater than the amplifier SR. For this particular output voltage, then, slew limiting is produced on the square-wave edges, causing the momentary disappearance of the sine wave. Note in particular the square wave transition in the center of the screen. This is, of course, a strong case of TIM, which is induced by the condition of slewing.

In 1b, the same signal is shown at a reduced level, and, as can be noted, the slew limiting is gone, as the waveform indicates simply a *linear* sum of the sine and square wave. The point being made here is that the distortion is not being caused so much by the transient as it is by the high SS (in Fig. 1a). Thus, it should be appreciated (in a qualitative sense) that SID (or TIM) is a distortion which is *level sensitive* in terms of both amplitude and frequency (since both affect SS).

This factor is demonstrated in another way by the square-wave response photos of Fig. 2. In Fig. 2a, a 10V p-p square wave is shown, and, as can be noted, the amplifier is slewing, as evident by the linear rising and falling waveform edges. In 2b, the waveform is at a lower level, and here the square wave is reproduced without slew limiting. This is evident by the *exponential* shape of the waveform edges, which is an indicator that the amplifier is operating linearly [15,36,37]. It is in actuality operating as a low-pass filter, as is defined by its small signal bandwidth, fc.

A square wave passed through a single-pole filter will exhibit the general waveform shape of Fig. 2b, and such a waveform at the output of an amplifier is a qualitative indicator that no slew limiting is present. At progressively higher voltage-output levels, slew limiting may set in (as in 2a), and the waveform then takes on the ramp-like slopes [15,37,50,63].

This is incidentally an excellent check to make on an amplifier if possible, increasing output square waves. If the exponential waveshape holds true for increases in level up to the rated output, the amplifier is behaving *optimally,* as it cannot be made to slew for any realistic signal conditions [11,43]. For this to be true, the power bandwidth must be *greater* than the small-signal bandwidth [45] which in turn says that the amplifier is guaranteed free from internal overload due to excessive SS. An amplifier can be designed for a defined small-signal bandwidth either by use of an input low-pass filter or appropriate feedback connections to constrain output SS below the SR. Further details of this from a

Fig. 3 — Inter-relationship of amplifier response, feedback, and SID. Fc is the small signal bandwidth which varies for different gains. Fp is amplifier full-power bandwidth which is independent of gain (for a given output level).

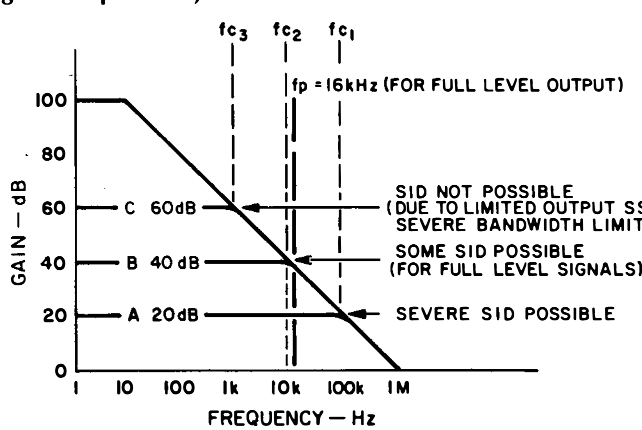
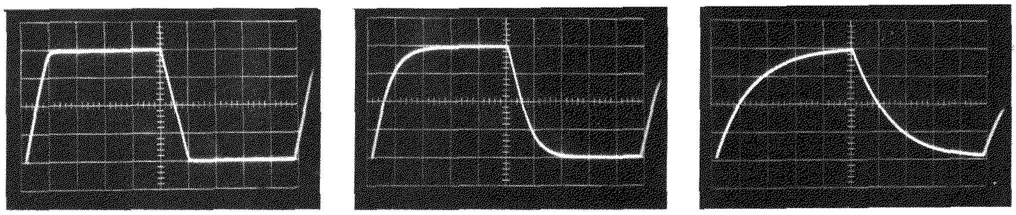


Fig. 4 — Relative relationship of fc and fp, and the resulting effect on SID. General conditions: 5 kHz square wave, 20 V p-p.

Fig. 4a — $Fc_1 > fp$; slewing evident.

Fig. 4b — $Fc_2 = fp$; some slewing on highest SS.

Fig. 4c — $Fc_3 < fp$; no slewing evident.



design standpoint are contained in several references [11,43,45] and are also discussed later on.

The Effect of Feedback on SID

One of the popular explanations for the cause of TIM and SID is said to be excessive negative feedback used around audio amplifiers [3,4,6,7,8,9,10,13,48]. In fact, this appears to be one of the more volatile parts of the issue, even to the extreme that already there have appeared statements in the literature calling for maximum feedback factors on the order of 12 dB and amplifiers advertised as having "zero feedback." The general argument advanced is that increasing negative feedback increases the susceptibility to TIM, and optimum feedback factors are said to be on the order of 30 to 40 dB.

It is interesting to consider how changes in feedback will affect the performance of an amplifier. There are certain aspects of the "less feedback is better" school of thought which have definite merit, but the *entire* situation must be considered for a true and complete perspective.

Consider a fixed gain-bandwidth amplifier open-loop response, as illustrated in Fig. 3. This amplifier has a unity-gain frequency of 1 MHz (such as a 741) and a full-power bandwidth of 10 kHz (at full output). Suppose we examine its susceptibility to SID for gains of 20, 40, and 60 dB, and at full output level. The small signal bandwidth (fc) for these three conditions will be 100 kHz, 10 kHz and 1 kHz, respectively [30]. However, for each condition of feedback, the full-power frequency (fp) remains at 10 kHz. Then, for the 20-dB (heavy feedback) gain condition SID is definitely possible, for output frequencies of 10 to 100 kHz. For 40 dB of gain, fc is equal to fp, and slight SID is possible. For 60 dB of gain, fc is less than fp, so SID is not possible.

A demonstration of this is contained in the photos of Fig. 4, taken from an IC op amp operating fairly close to the conditions of Fig. 3. For this device fp is 17 kHz, and Fig. 4a shows a square wave for the condition where fc is greater than fp; slewing is evident. In 4b, fc is equal to fp, and some slewing is noticeable at the *initial* rise of the square wave where SS is highest. In 4c, fc is less than fp and no slewing is evident. In all three instances, the experiment follows what the Bode diagram predicts.

The reason that slewing is not evident for the high-gain, low-feedback condition is because the amplifier output SS is severely curtailed, due to the very low small-signal bandwidth. This is another demonstration of the point made above that slewing can be prevented by making fc less than fp. For a fixed gain-bandwidth amplifier, as just demonstrated, this generally says that less feedback can prevent or reduce susceptibility to TIM or SID, as it reduces fc in relation to fp, or lowers the output SS in relation to amplifier SR. This is however hardly the *optimum* manner to arrive at this objective, as it will most certainly result in a generally noisier and more distorted amplifier, as well as possibly insufficient bandwidth. If fc is to be maintained less than fp, it should be done by another method, obviously.

Another view on the "less feedback is better" argument is to consider an amplifier which is compensated (optimally) for a higher gain (less feedback) condition. Due to fundamental feedback stability criteria, such an amplifier will have proportionally *less* compensation capacitance necessary. The smaller capacitance for less feedback then allows a higher SR to be realized by the amplifier, and so it is less susceptible to TIM or SID, as it can now handle greater SS waveforms linearly. In this case, the improvement is an indirect result of less feedback, a point which should be appreciated fully — *it also results because the SR is raised.*

These points are somewhat subtle, and we do appreciate that a fair amount of semantics are involved in the discussion which accompanies this issue. There are, however, several key points which are clear and should be made.

Since the limited SR is the cause of the distortion, it follows that design means which improve amplifier SR will lower distortion as a general result. (While this is generally true, there are notable exceptions, such as slew enhanced devices, which will be discussed later.) Feedback is certainly involved in the overall issue, but intimations that there is a fixed magical upper limit to feedback factors have no sound engineering basis to our knowledge. Given sufficient SR (and an otherwise linear amplifier), there is no inherent reason why 60 to 80 dB of feedback is not allowable [33,45,47,52]. The ultimate stability limit will, in practice, confine it to less than this as a natural consequence of usable gain-bandwidths, at least at audio frequencies.

Another part of the semantics issue comes to play with the argument that less feedback in combination with a more linear open-loop characteristic is desirable towards prevention of TIM. Essentially this is true, because without a high degree of overall feedback, less compensation (if any) is needed, and SR goes up as a result. However, *local* feedback around a stage *is still feedback,* and if bipolar transistors are used, it hardly seems possible to get truly excellent open-loop linearity *without* a lot of feedback, since their voltage transfer is basically exponential. So the argument should perhaps be oriented towards a closer definition of *what kind of feedback,* as well as its degree.

To get back to the more conventional amplifier, the point has been made that it is SR which is the fundamental predictor of SID (and/or TIM), and amplifier improvements which increase SR generally lower SID (and TIM).

The remaining low TIM criteria, wide open-loop amplifier bandwidth, involves semantics also. Taken literally, an open-loop bandwidth of 20 kHz (as commonly specified) [1,2,3,4,6,7,8,10,14] will be interpreted to mean 20 kHz *small signal bandwidth.* What is *really* important is a 20 kHz (or more) *power* bandwidth, which will minimize or eliminate slew limiting [33,34,39,45,52].

Amplifiers can be designed for 20 kHz (or more) open-loop bandwidths, but often with a severe penalty of low-frequency linearity and gain accuracy [40,45]. By results from several different forms of tests, there appears to be no fundamental necessity for a wide open-loop small-signal bandwidth, given a power bandwidth sufficient to eliminate slew limiting. Several specific test results discussed later on clearly demonstrate this point.

Fig. 5a — Amplifier model.

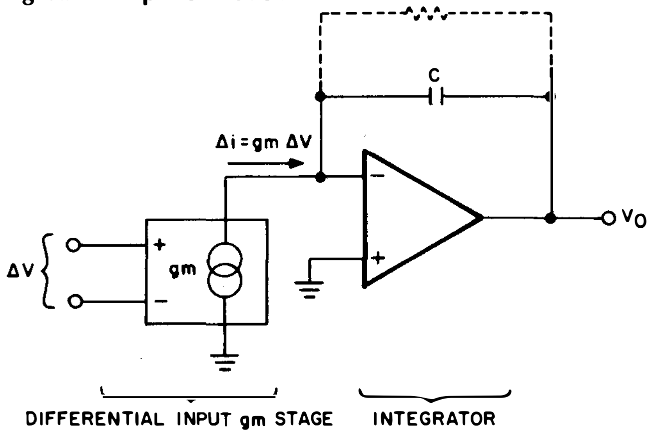
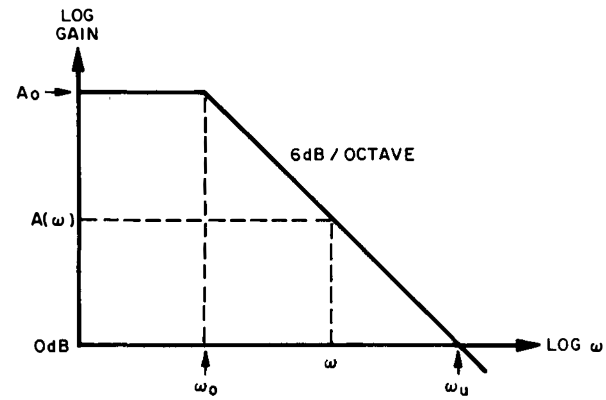


Fig. 5b — Amplifier frequency response.



Analysis of The Slew-Induced Distortion Mechanism

It is of fundamental importance to understand the various distortion sources in amplifiers, such as the SID mechanism of interest here. In this discussion we will mostly deal with operational amplifier circuits, but since many present-day power amps are of similar topology and are subject to similar physical laws, the discussion and data will be relevant to them as well.

Figures 5a is an idealized model of a typical operational amplifier [20,21,22,24]. Its input stage is a voltage-to-current converter or transconductance stage, characterized by the parameter g_m. The output current of this stage (Δi) is simply

$$\Delta i_{(t)} = g_m \Delta V_{(t)}. \tag{4}$$

The second stage of the amplifier is an integrator, with an output voltage (Vo)

$$Vo_{(t)} = \frac{g_m}{C}\int \Delta V_{(t)} dt. \tag{5}$$

The resistor R is responsible for the finite d.c. gain of the amplifier. At low frequencies the open-loop gain is

$$Ao = g_m R. \tag{6}$$

The open-loop frequency response begins dropping (Fig. 5b) at a frequency

$$\omega o = 1/RC. \tag{7}$$

Since for audio circuits we have no great interest in the amplifier gain at d.c., it is much more convenient to neglect R (as in equation 5) and work with the unity gain bandwidth (ωu) which, due to the integrator's -6 dB/octave response, is equal to the gain bandwidth product.

$$\omega u = A_{(\omega)} \times \omega = A_o\, \omega_o = g_m/C. \tag{8}$$

Referring to equation 5, we have

$$Vo_{(t)} = \omega u \int \Delta V_{(t)} dt. \tag{9}$$

Thus, for an amplifier with a six dB/octave frequency response, the amplifier can be characterized simply by its unity-gain bandwidth or gain-bandwidth product. Our next step is to examine the differential input voltage as a function of the output voltage. Differentiating equation 9 we have

$$\Delta V_{(t)} = \frac{1}{\omega u} \frac{dVo_{(t)}}{dt}. \tag{10}$$

This highly important result clearly shows us that the instantaneous differential input voltage of an amplifier is directly proportional to the slope of the output voltage, with $1/\omega u$ as the constant of proportionality.

If we now look at an actual amplifier, we will understand what SID really is. Figure 6a is a very simple real amplifier which will serve to demonstrate this. Q1 and Q2 are the differential input pair, and Q3-Q4 form a current mirror. This Q1-Q4 stage is our transconductance amplifier with a trans conductance of

$$g_m = I_k/2V_T \tag{11}$$

where $V_T = K_T/q$ (26 mV at room temperature). Q5, with its current source load I_A, forms our integrator, in concert with C. We will neglect the finite d.c. gain produced by R,

Fig. 6a — Real amplifier.

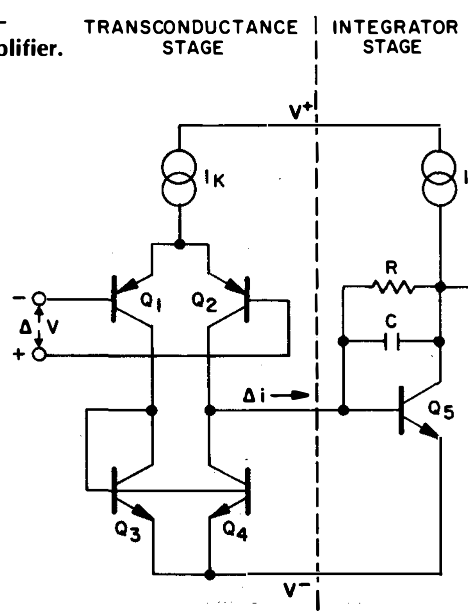


Fig. 6b — Transconductance nonlinearity.

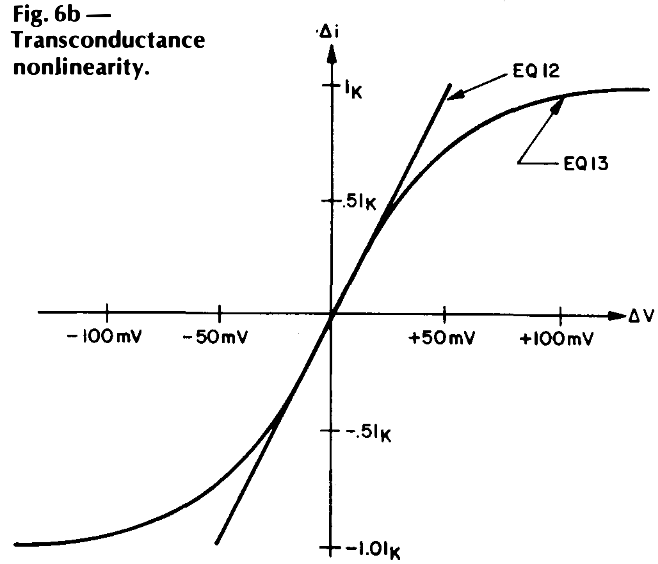
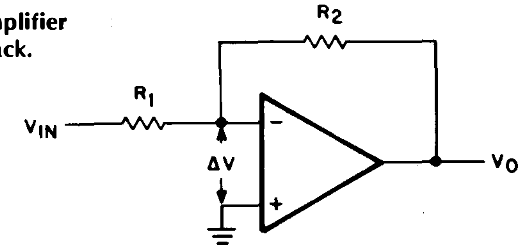


Fig. 7 — Amplifier with feedback.



inasmuch as it has no bearing on ωu (see above). Ideally the g_m stage output current (Δi) is

$$\Delta i_{(t)} = g_m \Delta V_{(t)} = I_k(\Delta V_{(t)})/2V_T. \tag{12}$$

However, this is only true when ΔV is small. The exact transfer expression for this input stage is [23].

$$\Delta i_{(t)} = I_k \tanh(\Delta V_{(t)}/2V_T). \tag{13}$$

As this expression shows, the transconductance stage is linear only for small signals, and thus will produce distortion for high output currents, when ΔV is large. Equations 12 and 13 are plotted in Fig. 6b and illustrate this point more clearly.

The maximum output current (limit) from our input stage is I_k. This determines the maximum rate of change of V_o, which is the slew rate of our amplifier. This is simply

$$SR = I_k/C. \tag{14}$$

How close we are working to the SR is

$$SS_{output}/SR = \Delta i/I_k. \tag{15}$$

This relation is one important and useful, as will be seen. The ratio SS/SR we will here define as the *slew rate ratio* (SR ratio), which relates the output SS to the amplifier SR.

This ratio is easily measurable from outside the amplifier with a differentiator,

$$\Delta i/I_k = (1/SR)(dV_o/dt). \tag{16}$$

Figure 6b graphically tells us that operating with a SR ratio >0.25 (or $\Delta i > 0.25 I_k$) will produce some obvious distortion. This is equivalent to saying that operation at greater than 25 percent of the amplifier's SR will produce distortion. This distortion depends solely on the SS of the output, hence our use of the term "Slew Induced Distortion." The amplifier is producing distortion by being forced towards its SR limit; the distortion is slew induced.

So far we have been talking only of the amplifier with no mention of feedback and we have been discussing the open-loop performance. Amplifiers are rarely used open loop, so we must turn our attention to the effects of feedback on amplifier performance. An important point to keep in mind as we discuss feedback is that feedback networks are placed around an amplifier and have no direct effect on its *internal* performance. Feedback alone will not effect the validity of any of the equations developed above. It will, however, under certain signal conditions, cause these relationships to be taxed, creating a SID-producing situation. This statement will become more clear with subsequent discussions (if not already so from the preliminary discussion).

As is well known, feedback reduces distortion. Let's take a qualitative look at how this happens. A simple feedback network has been placed around our amplifier in Fig. 7. The differential input voltage is

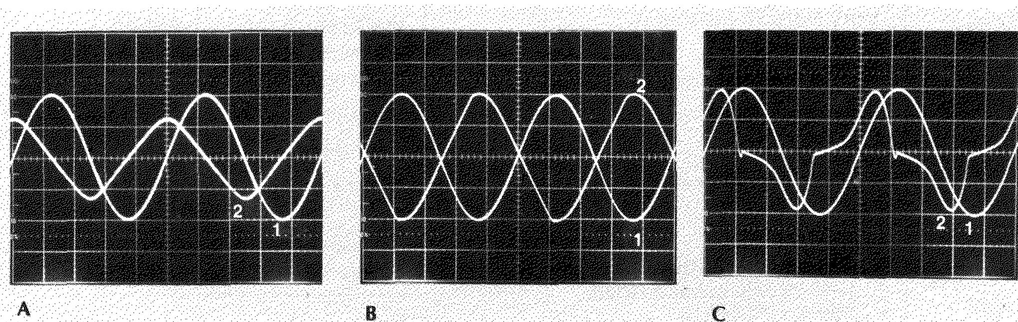
$$\Delta V = (V_{in}R_2 + V_0 R_1)/(R_1 + R_2). \tag{17}$$

This is the error voltage which we would like to be zero, but it will be non-zero if V_o contains a gain or phase error, or distortion. If we operate the amplifier near its slew limit, we know that the amplifier transfer characteristic is very non-linear (see 6b). The feedback will reduce this non-linearity from V_{in} to V_{out}, but it will necessarily still exist from ΔV to V_{out}. If the feedback is doing its job and producing a relatively clean signal at V_{out}, then it follows that the signal ΔV must be distorted. The distortion of ΔV must be of the proper magnitude and phase to compensate for the amplifier's internal nonlinearity, if it is in reality reducing distortion. A qualitative insight of this is contained in the waveforms shown in Fig. 8. These are pictures of the performance of a 748 op amp, compensated to unity gain by 30 pF and operated as shown in Fig. 7. The amplifier had the following performance (measured before the experiment):

$f_t = \omega_t/2\pi$

$= 1.5$ MHz

$SR = +0.97, -0.91$ V/ μS.



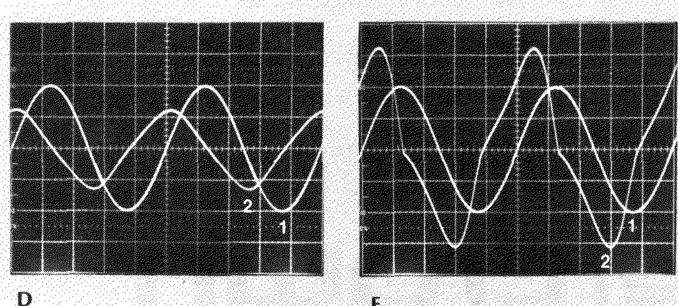


Fig. 8 — A 748 op-amp operating under various conditions detailed in the table.

Table I — Operating conditions for 748 op-amp.

Fig.	Gain	SS.V/μS.	THD,%	Trace 1	Trace 2
8a	1	±0.8	0.05	V in 5 V/div.	ΔV 0.1 V/div.
8b	1	±1.2	3.5	V in 5 V/div.	V out 5 V/div.
8c	1	±1.2	3.5	V in 5 V/div.	ΔV 0.5 V/div.
8d	10	±0.8	0.56	V in 5 V/div.	ΔV 0.1 V/div.
8e	10	±1.2	4.1	V in 5 V/div.	ΔV 0.1 V/div.

The amplifier was operated at its full rated output swing of 20 V p-p. Two test frequencies were used, 12.7 kHz and 19.1 kHz. At 20V p-p (10V peak) these frequencies produced signal slopes of ±0.8 V/ μS and ±1.2 V/ μS respectively. These two frequencies were applied to the closed-loop amplifier, for signal gains of 1 and 10. For either gain condition, the output was a visibly clean sine wave for the 12.7 kHz, ±0.8 V/ μS signal (not shown). However, the 19.1 kHz, ±1.2 V/ μS signal drove the amplifier into slew limiting, and this is shown in Fig. 8b. The output slewing waveform was visibly the same for either gain. Table I summarizes and identifies the conditions and results shown.

The important point to note from this is that the op-amp input, ΔV, becomes highly distorted in an attempt to linearize the response of the closed-loop amplifier. In 8a and 8d, for example, ΔV is just beginning to become non-linear, but is still relatively low in level. As the maximum slew rate is exceeded, this process breaks down and the error voltage abruptly increases, as can be noted in 8c and 8e (note the different scale factors for ΔV). Operation at the lower gains (more feedback) yields lower distortion operation, and allows low-distortion operation closer to the slew rate limit.

There is nothing particularly unique about SID in audio amplifiers. It can be measured, calculated, and improved upon by using standard techniques that have been available for some time [57]. The only elusive aspect of this form of distortion is that rather than occurring on a peak magnitude (like clipping), it occurs on the rising or falling edge of the waveform, when the SS approaches or exceeds the amplifier SR. This is due to the fact that the dominant non-linearity in the circuit, the transconductance of the input stage, is followed by an integrating stage. Thus in Fig. 5, if the transconductance stage were overloaded and producing clipped square waves of current output, the integrating stage would transform these square waves into triangle waves at the output. The triangle wave is the ultimate example of gross slewing distortion, and its presence is a visible verification that the amplifier is operating open loop during the slew interval(s).

Although slew limiting is most often encountered in amplifiers due to *internal* IC relations, such as have been just described, it can also occur due to output-current/load-capacitance rate limiting, with the end effect being similar [33,34]. This type of slew limiting can occur for example in RIAA-equalized preamps which cannot adequately charge frequency-shaping capacitors [33,41] or power amplifiers which cannot drive capacitive loads due to protection circuitry [33].

The distortion products produced by SID are measurable either by methods of THD [16], two-tone high-frequency IM, or TIM [14,33,34,51], and in all cases they become significant as the amplifier's inherent SR is approached by the output signal slope.

Representative results from these test methods are discussed in Part II of this series. In this next installment, sample data from different types of distortion tests are presented consisting of total harmonic distortion (THD), two-tone-difference intermodulation distortion (IM), and the recently proposed test for TIM [18]. Some of the relative merits of these measurement techniques will be discussed, and it will be seen that while they are all useful to the detection of this distortion, there are differences in sensitivity and practicality between them. Generally speaking, low-frequency distortion tests such as 1-kHz THD, or 60-Hz/7-kHz (SMPTE) IM tests are useless for detecting SID, since the signal slope is not sufficiently high. An interesting outcome is that IC op amps, long viewed with suspicion by many, are actually capable of truly superlative performance when properly operated below their slew-rate (SR) limit.

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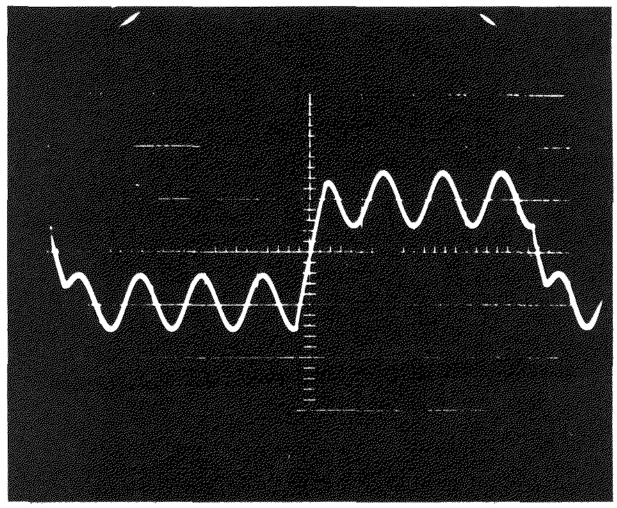
An Overview Of
SID and TIM

Walter G. Jung, Mark L. Stephens, and Craig C. Todd

Part II—Testing



A desirable object for the study of SID is to develop a reliable and predictive test method (or methods) for the presence of this distortion. With our studies, this objective was generally met, and good correlation was observed among several different means of measurement and theoretical calculations [33, 34]. These electrical test methods also appear to correlate roughly with listening tests made on the same devices. The results of these different tests made on a wide variety of IC op amps are described in this part.

THD Tests

It has been often reported that THD test methods are generally insensitive to the detection of TIM distortion [8, 9, 18, 49, 64]. In actuality this is only true for insufficient (or fixed) signal slopes, i.e. when SS<SR. This factor will be demonstrated in the discussion below. A 1-kHz *spot-frequency* THD test is an example of a test which is (typically) either too low in SS, or if fixed in level, not dynamic at all. An example of a *dynamic* test in terms of SS is one which moves the SS of the test signal *up to and through* the amplifier SR; i.e. where the SR ratio is forced to reach and pass unity. It must do this, of course, without amplitude clipping of the output signal, which implies a swept frequency test.

In practice, a relatively straightforward means of exercising an amplifier for SID (or TIM) is to apply a low frequency (about 100 Hz) signal at full rated output voltage, and then sweep the frequency upward until a sudden rise in distortion is noted [16, 33], the 1 percent distortion level coinciding with the amplifier's full power bandwidth.

In op amps, a full output-voltage level sweep test for THD from 100 Hz to 100 kHz has been found to be a sensitive and easily applied test to detect SID, as it exercises the output signal slope-tracking fidelity to a high degree. Unfortunately, this form of test is not always directly applicable to power amps, but it is an excellent one for IC op amps. Reasons which can defeat its validity for some equipment are limited signal bandwidth, which masks true distortion products, and in power amps, output stage stress. For wideband, low-level stages, it can be an excellent test. However, the SID distortion mechanism cannot always be conveniently isolated and quantified, simply because one does not always have direct access and control over amplifier configuration and/or operating condition (s).

It *is* possible to isolate SID from other distortion sources when the test configuration can be completely controlled. For instance, when testing op amps, this can be accomplished by placing some important restrictions on the test circuit [33, 34, 35, 37, 38]. The test configuration should operate in the inverting mode, to eliminate the common-mode distortion effects which exist when an op amp is operated non-inverting [37, 38]. The magnitude of these effects in some designs can approach that of SID, therefore a non-inverting test is simply incapable of discriminating these two components. Similarly, output stage non-linearity should also be minimized by careful restriction of loading, to 10K or more. These precautions assure us that we are truly measuring *only* SID and not other additional distortions such as those produced by poor common-mode rejection or output loading. These distortion mechanisms should be evaluated separately [31, 38] and are not the subject of this study. Failure to make certain the test conditions are free from these distortions can lead to questionable results.

A test circuit which takes these points into consideration and is suitable for SID tests is shown in Fig. 9. It is a unity-gain inverter, with the device's frequency compensation adjusted for unity gain, except for special cases as noted. Input-output signal levels are full rated-voltage swings of ±10V or 7V rms (except as noted), which generally maximizes the output SS. The heavy feedback condition maximizes sensitivity to slewing distortions, since it maximizes the potential error voltage.

The device under test (DUT) is operated in this circuit, and a check is made for its actual slew rate. Note that for a given device, the actual slew rate can vary from the data sheet value, therefore results can only be correlated by actual measurement. Ideally, slewing should be symmetric, so the measurement should take note of both plus and minus slew rates. After the SR test, measurements can proceed.

Portions of this article are adapted from "Slewing Induced Distortion in Audio Amplifiers" by the authors in **The Audio Amateur**, Feb., 1977 (P.O. Box 176, Peterborough, N.H. 03458), part of an article series which is available in book form. Portions were also adapted from the authors' article "Slewing Induced Distortion — Its Effect on Audio Amplifier Performance, with Correlated Listening Results," Audio Engineering Society Preprint No. 1252 from the May, 1977, convention. (See bibliography references nos. 33 and 34.) © Copyright 1979 by Walter G. Jung, Mark L. Stephens, and Craig C. Todd.

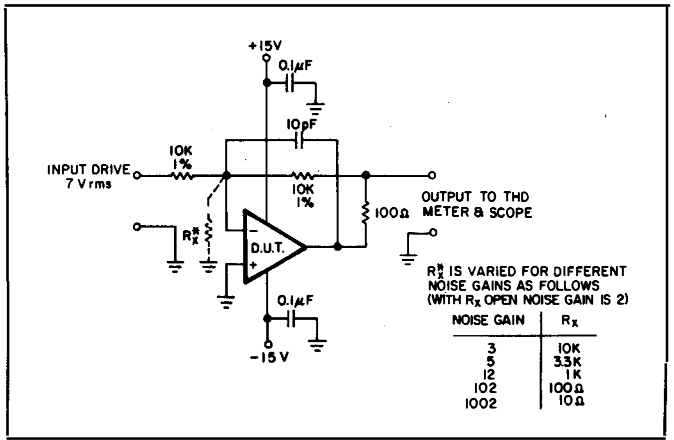


Fig. 9 — Test circuit for slew induced distortion.

Representative THD performance data on a common 741 IC op amp with a 0.5V/μS SR is shown in Fig. 10. These data indicate in the full output-level curve a characteristic sharp rise from the low frequency (LF) residual level to a 1 percent THD level at 8 kHz (fp), this occurring within only two octaves. For lower output levels, such as for 2V and 1V rms, the 1 percent THD frequency is proportionally higher, in fact by the ratio of amplitudes. In all three cases, the characteristic sharp rise in distortion can be noted as the device's SR is approached by the SS. The 1 percent THD point is reached when the SS becomes equal to the fixed device SR. This can be noted as a relatively constant SS for the three different 1 percent THD intercept points, as is evidenced by the different frequencies at which this point is reached for different levels.

SID improves considerably for higher SR devices or compensation conditions which result in higher device slew rates. In Fig. 11, THD data on a 301A amplifier is shown for various compensation/gain conditions, with all data referred to a 7V rms output level.

The first curve (left) is for unity-gain compensation, where the device SR is 0.9V/μS; the behavior is similar to but slightly better than the 741 for similar conditions. For the X10 compensation curve, the resulting slew rate is 7V/μS, and the performance is much better, with slew limiting not reached until 90 kHz. The improvement is largely due to the X10 improvement in SR and gain-bandwidth product, without any major penalty in LF distortion or noise.

The third curve is for a X100 compensation/gain, and here slew limiting is not at all evident.

Slewing *symmetry* has a pronounced effect on SID, and SID will only be minimized when the plus and minus slew rates are equal. In some IC amplifier devices, particularly those which use current mirrors, slew symmetry can be trimmed to demonstrate this effect, as shown in Fig. 12.

Here the THD performance of a 301A op amp with an SR of 0.4V/μS (when trimmed) is plotted, and the data indicate an fp of 6.7kHz, which agrees with the theory. For asymmetric slewing, however, the distortion generated is higher, and the break point occurs lower in frequency. This sort of behavior can be noted in many amplifiers, and those in which slewing is inherently asymmetric will not yield as low a distortion as devices which are symmetric.

An aspect of SR asymmetry which illustrates why inverting mode operation is recommended for SID characterization is demonstrated in Fig. 13 and 14. Figure 13 is the full-scale (20 V p-p) slewing response of a 301A amplifier, and, as can be noted, there is a marked difference between plus and minus slopes [21, 22, 37, 38]. This same amplifier was used in the inverting-mode pattern of Fig. 4a (Part I), where it was seen to be nominally symmetric.

Slewing differences between inverting and non-inverting input operating modes show up in THD tests, as is demonstrated by Fig. 14. This data is for the same amplifier operated at unity gain, with curve A for inverting mode, curve B non-inverting. Note that the fp is lower in curve B and distortion much higher at lower frequencies than curve A. This general pattern can also be seen in other devices as well [38].

An interesting demonstration of the effectiveness of SR improvement on THD is contained in Fig. 15. This data is for the 2725, a *programmable* IC op amp, where the device SR can be adjusted via a bias terminal. Shown here is the resulting THD for SRs of 0.5, 1.6 and 5V/μS respectively. As can be readily noted, the resulting performance improves directly as SR is increased.

Since the previous performance examples have indicated that quality is generally directly tied to slew rate, it might seem fair to assume that a very high slew rate is sufficient in itself to achieve this quality. However, this is not completely the case, as is shown by Fig. 16. These data are THD performance for a class of op amps known as "slew enhanced" types [20]. This form of op amp uses a class B (or AB) input stage to dynamically alter (increase) the output current (I_k) and thus boost SR for high SS conditions.

In terms of THD, slew-enchanced units generally show a low SS distortion performance much like a conventional op

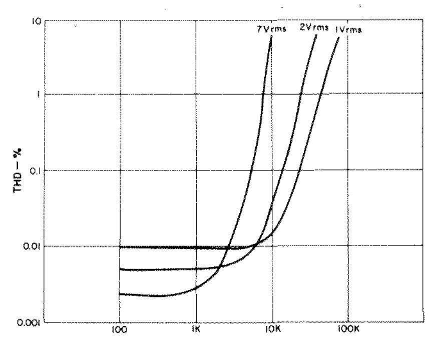


Fig. 10 — THD vs. frequency for a 741 op amp operated as a unity gain inverter at various output levels. Device slew rate is 0.5 V/μS.

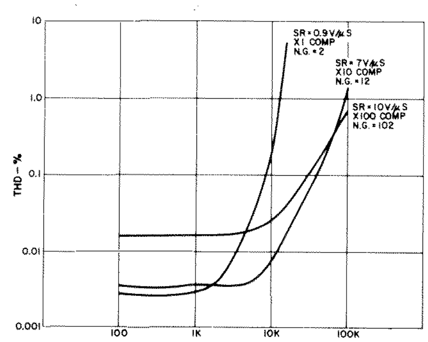


Fig. 11 — THD vs. frequency for a 301A op amp operated as a unity gain inverter at various compensation/ noise gain conditions. Output is 7 V rms.

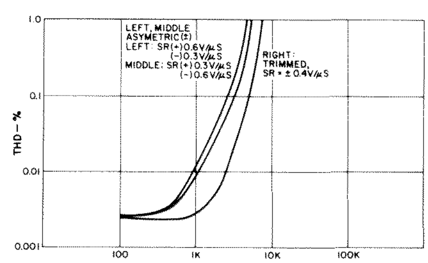


Fig. 12 — THD vs. frequency for a 301A op amp operated as a unity gain inverter with various slewing symmetries, Cc is 33 pF, output is 7 V rms. Offset adjustment is trimmed for slewing condition shown.

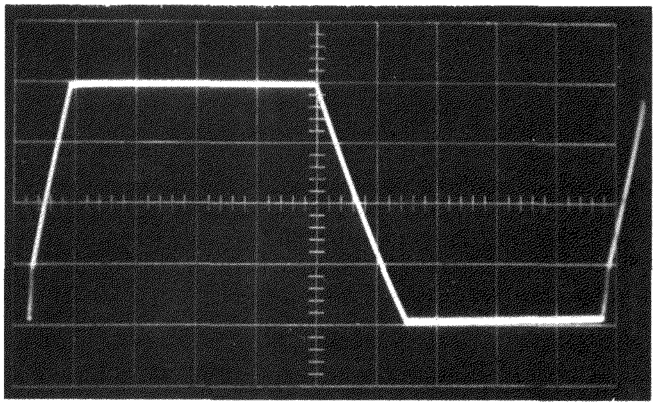


Fig. 13 — Asymmetric slewing due to common mode asymmetry in a 301A op amp operated as a unity gain follower; 20 V p-p.

amp up to a point, but complete slew limiting is prevented. The data reflects this, showing a general initial rise, then a leveling off in THD. It also shows substantial differences in the performance of the various devices tested. Highest performers are those units which show the combination of good low-level linearity concurrent with high GxBW, e.g. the x10 531, the 530A, and the 538.

As a final example of THD performance, the data of Fig. 17 indicate what effect an adjustment in SR *independent* of small-signal bandwidth (or feedback) has on SID. For this test, a 318 op amp is used in the circuit shown. The 318 has a very high SR of 50V/μS with a gain bandwidth of 40MHz, and its performance is sufficiently high (curve A) that the THD measured is essentially the residual of the analyzer used [31, 33, 34, 38].

As the test circuit shows, the current sources and load C_L constitute a slew limit mechanism which can be used to experimentally alter SR, independent of both feedback and amplifier bandwidth. Curve B indicates THD for a 6.7V/μS condition, C for a 0.5V/μS condition. Note that the fp for C is 8 kHz, as equation 3 predicts.

This test indicates two things; one that SR is a good general indicator or predictor of high frequency distortion for high SS waveforms. Second, it indicates a pattern of distortion rise in curve C much more sudden than any previously noted. This indicates that the heavy feedback (for the 318, as used here) is successful in suppressing the typical two octave rate of rise noted in other patterns above [33, 53, 60].

At this point, THD performance data has been shown which reflects the key behavior patterns observed in the group of IC samples tested. From this, it can in general be noted (for these tests) that if the device slew rate is 5V/μS or more, is symmetrical, and does not use nonlinear slew enhancement, the THD performance can be superlative. This will be evidenced by a THD of 0.01 percent or less up to 20 kHz (a 20-kHz *distortion-free* bandwidth) with an fp of 80 kHz or more. For the best devices, THD can be 0.1 percent or less up to 100 kHz. Of those tested, the best devices in the above terms were: NE5534 (equivalent to TDA 1034), 536, 318, 518, the TL080 and TL070 series, 3140, 2625, 2525, 301A

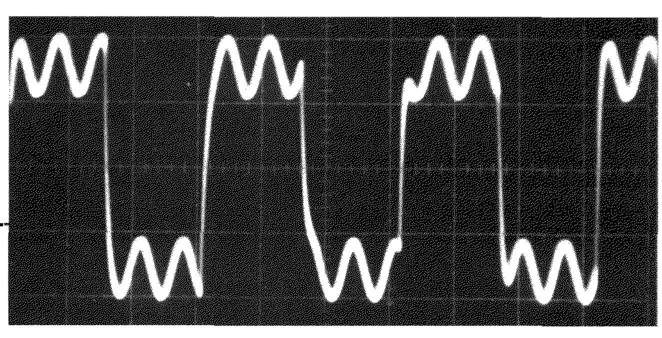
Sine-Square Test

A combined sine-square wave IM test has been proposed by Leinonen, Otala, and Curl as a method of measuring TIM [18]. For this test, the signal is a 3.18-kHz square wave, which has been filtered with a simple one-pole, low-pass R-C filter, at either 30 or 100 kHz, and combined with a 15-kHz sine wave of one quarter the peak-to-peak amplitude of the square wave. The resulting square-wave signal component has a very high slope, which is in theory actually limited only by the low-pass filter. As can be appreciated from this factor, this test has the capability of stressing an amplifier to a high degree for non-linearities related to signal slope and/or slew rate. Figure B-1 is an oscilloscope photo of a 30-kHz band-limited signal (DIM 30).

The output spectrum of the amplifier under test is analyzed for intermodulation products generated by non-linear mixing of the sine and square waves. The rms sum of these products relative to the amplitude of the *15-kHz sine* wave is defined as the percentage distortion. This definition of the test does not include the residual distortion products of the square-wave source (which are not a result of the intermodulation under examination). As typically occurring in practice, these spurious products are the even-order harmonics of the square wave (which, of course, should ideally be absent).

If a very high-quality square-wave generator is used, for example with even-order harmonics 90 dB down from the fundamental, even-order distortion resulting from *amplifier*

Fig. B-1 — Time domain representation of DIM-30 test signal.



asymmetry is measurable. In the tests of this study, this type of distortion was included, as sometimes it was the *only* distortion present in the output spectrum.

It should be noted that a test signal as defined above has a very wide spectrum. For example, even though the square wave is low-pass filtered at 30 kHz, there is still significant energy present up to several hundred kHz.

A very interesting and inherent property of an ideal square wave (with no band limit) is that every individual harmonic of the Fourier series comprising the square wave contributes the same amount to the resulting slope of the square-wave transition. This is because the amplitudes of the harmonics fall in exact proportion to their rise in frequency, which makes the slope constant for increasing harmonics. Thus, it should be intuitively appreciated that an unfiltered square wave constitutes an extreme test in terms of signal slope. In the ideal case, for a fundamental frequency/amplitude combination resulting in a slope of "x" V/μS, the composite slope will be infinite; in a practical case of "n" harmonics, the slope will be (n + 1) (x) V/μS.

This pattern of constant signal-slope contribution per harmonic is not *strongly* ameliorated by a simple 30-kHz single-pole filter, such as is used in the sine-square test. As a result, a very high percentage of the signal slope is contributed by ultrasonic energy. As a specific case in point, every odd harmonic comprising a 16V p-p, 3.18-kHz unfiltered square wave contributes 0.16V/μS to its slope. When passed through the 30-kHz filter, there will be five square wave harmonics below 30 kHz (f1, 3 f1, 5 f1, 7 f1, 9 f1). These components will contribute slopes of 0.16, 0.15, 0.14, 0.13 and 0.12 V/μS, respectively, to the composite test signal slope, while the 15-kHz, 4V p-p sine-wave signal (f2) has a slope of 0.19V/μS. This sums up to a slope contribution of 0.9V/μS for those test signal components below 30 kHz.

The total composite test signal slope for these conditions has a slope of over 3 V/μS. It is therefore clear that over two-thirds of the test signal slope is contributed by the square-

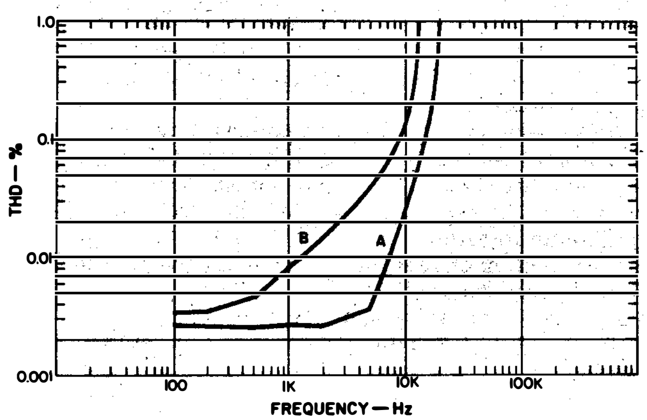


Fig. 14 — THD vs. frequency for a 301A op amp at 7 V rms output in inverting[(A)] and noninverting[(B)] modes.

(feed-forward), and the OP-01. Nearly as good were the AD540 and 8007. The common characteristic of all of these amplifiers is their high slew rate and input stage linearity. (No ranking is implied, and other types may be capable of such performance.)

Two-Tone HF IM Tests

A second series of tests conducted on this sample group of IC op amps was HF two-tone difference IM, hereafter called simply IM. This type of test also shows SID, as evidenced by IM, to be generally governed by amplifier slew rate. For this

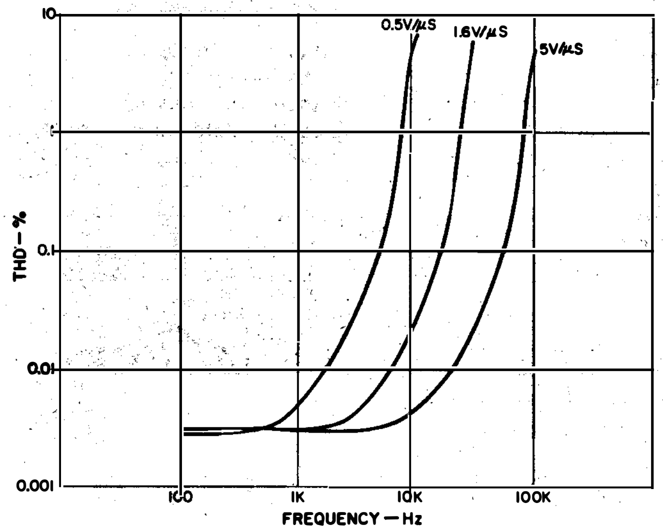


Fig. 15 — THD vs. frequency for a 2725 programmable op amp operated as a unity gain inverter at 7 V rms output at various slew rates.

test a one-to-one mixed, high-frequency tone pair at full output level is swept from 10 kHz to 50 kHz. The difference frequency is maintained at 100 Hz. All tests were performed in the test circuit of Fig. 9.

Figure 18 shows some data which indicate the general relationship of IM performance and SR. These data were taken

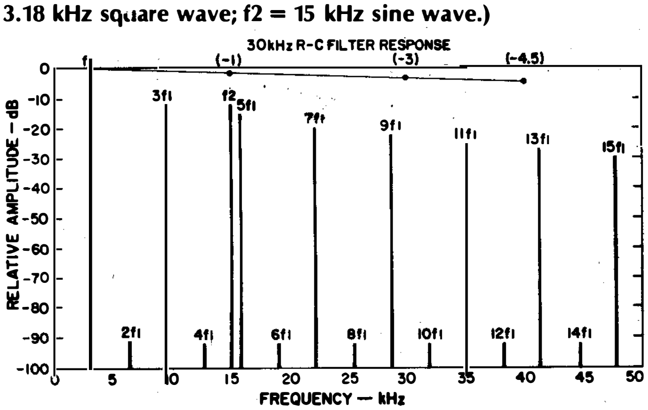
wave harmonics *above* 30 kHz, which are not completely filtered. Obviously, this form of test cannot be construed as an "in-band" test, as the bulk of the energy distribution in terms of signal slope is concentrated in the *ultrasonic* region of the spectrum.

The above points are graphically illustrated in Figs. B-2 and B-3. Figure B-2 is a simple spectral distribution plot of a sine-square 30-kHz band-limit test. This shows the relative amplitude of the individual signal components as they appear at the input to an amplifier being tested. The 30-kHz filter response is also shown for reference, superimposed above the spectral lines of the signal.

The spectrum, as shown here, very closely resembles the conditions used in our sine-square tests, for the 30-kHz case. As can be noted, the non-ideal even-order products are approximately -90 dB with respect to the fundamental.

The plot of Fig. B-2 is simplistic in the sense that it gives no real appreciation for what is required of the amplifier in terms of SS capability or SR. The graph of Fig. B-3 is intended to convey this.

Fig. B-2 — Spectrum of sine-square test signal. (f1 = 3.18 kHz square wave; f2 = 15 kHz sine wave.)



This figure is simply a graph (or graphs) of the signal slope which results for 30-kHz and 100-kHz band-limit conditions versus p-p operating level. For the case discussed above, the example of a 20V p-p level and 30-kHz band limit is plotted, and the resulting SS is 3.2V/μS, as noted. Were the band limit 100 kHz, the SS would be over 10V/μS (for the same operating level). From the simple relationship shown, an SS can be calculated for any operating level for either case of filtering.

An important point to be noted is the fact that this relationship applies to *voltage* swing, and it can apply to either preamps (at lower levels) or power amps (at the higher levels). It has an indirect link to power output (since power is a function of load impedance as well as voltage).

Finally, as will be noted from the discussions of the tests in the text, a given amplifier should have an SR *greater* than the SS generated by a particular test condition. The 3.2V/μS SS case, for example, would require an amplifier with an SR in excess of 3.2V/μS, for distortionless reproduction.

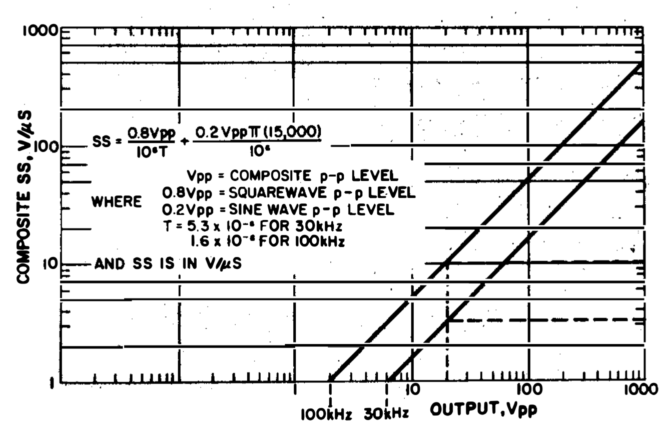


Fig. B-3 — Sine-square test signal slope.

with the 2725 programmable op amp, with slew rates of 0.5, 1.6 and 5V/ µS (condition similar to Fig. 15).

The nature of the IM performance behavior with respect to increasing SS strongly resembles the data based on THD, showing a similar rise as the amplifier SR is approached. This behavior pattern is a characteristic one of IM [14, 33, 34, 51], just as it is for THD. The rise in IM (dotted) at low SS reflects the equipment residual.

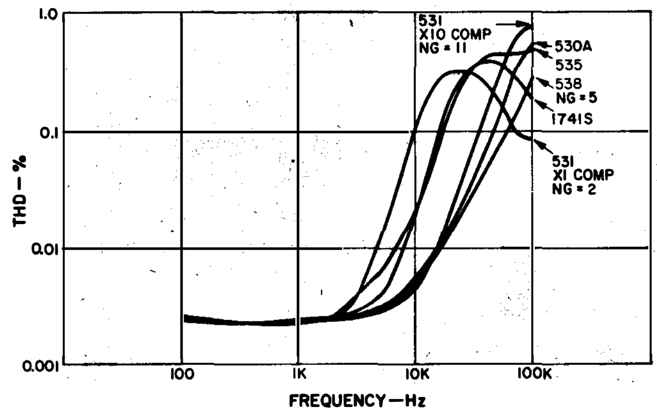


Fig. 16 — THD vs frequency for various slew-enhanced op amps operated as unity gain inverters at 7 V rms output.

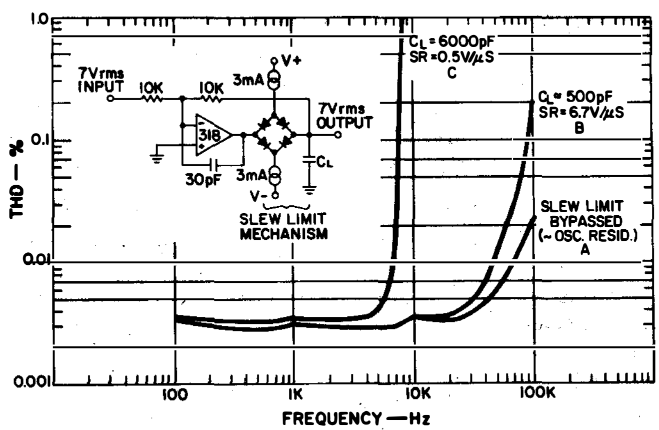


Fig. 17 — THD vs. frequency for a 318 op amp with an artificially induced signal path slew limit.

Fig. 18 — Two tone IM (mixed 1:1) vs. frequency (Δf = 100 Hz, constant) for a 2725 programmable op amp operated as a unity gain inverter at ±10 V output for various slew rates.

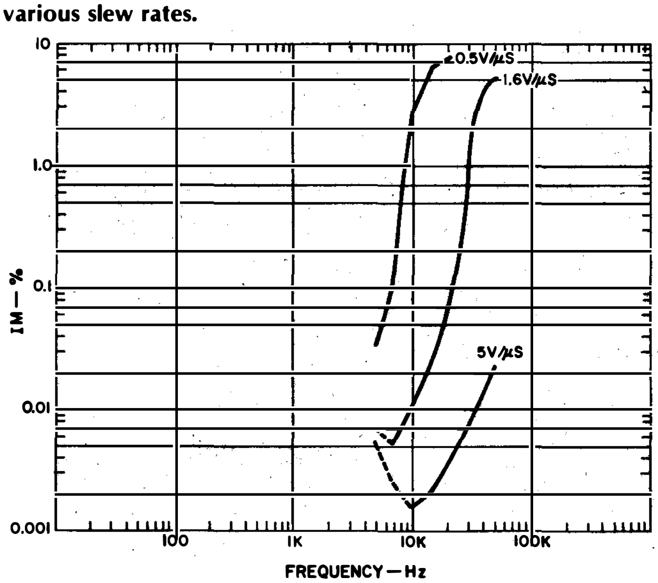


Figure 19 shows a composite plot of IM performance by this method, for a variety of different IC op amps. The highest performance devices here show IM distortion at the equipment residual level, while the others show quality generally proportional to slew rate. The notable exceptions to this pattern are the 535, a high-speed slew-enhanced type, and the 356, an asymmetric-slewing unit with appreciable second-order distortion. Each unit has a high slew rate, but the exact method of achieving it prevents optimum linearity from being realized.

The data from the IM tests follow the same general pattern as THD-based data in terms of distortion rise for SR ratios approaching unity. It is less sensitive, though, due to the fact that it measures even-order products and the amplifiers usually (if perfectly symmetrical) generate odd order. This test is quite effective in pinpointing amplifiers which have inherent transfer asymmetries (and thus even-order distortion), such as the 356 type. A two-tone IM test to measure odd-order products (2 f1 - f2) would yield more useful data on the symmetrical devices.

It should be noted that an IM test such as this can be more useful for band-limited amplifiers, as it can measure IM products folded downward to lower frequencies by the HF tone pair.

Sine-Square Tests

A selected sampling of devices which had undergone the THD and IM tests were subjected to the sine-square tests as outlined in reference 18 and described in the sidebar. Like the previous THD and IM tests, the test circuit of Fig. 9 was used. Our results do not directly correlate with those of reference 18, because we are operating the amplifier with no common-mode swing (inverting mode) in order to isolate SID from common-mode distortion. Figure 20 summarizes the results of these measurements, for full output-level tests performed with a 30-kHz square-wave band limit.

The general relationship between Dynamic Intermodulation Distortion (DIM after the terminology of reference 18) and device SR capability is shown by the graph in Fig. 20. This graph shows percentage DIM versus device SR, for all types of devices under one standard test condition. The maximum SS of the input sine-square signal for this case is 3.2V/µS. Thus, a given device would require an SR of at least this much to pass the waveform without gross distortion. This graph shows that distortion rises above the residual level at around a device SR of 6.5V/µS, which is roughly twice the SS of the input waveform.

This is an important and useful indicator; on the average, a device must have an SR capability of twice the input signal slope to pass signals with negligible distortion. As the SR capability of the test devices falls below 6.5V/µS, the graph is seen to rise linearly to very high amounts of distortion. A best straight line drawn through the data points turns out to have a slope of 3:1 on the logarithmic coordinates. This indicates that DIM varies as the third power of the ratio of the input SS to the device SR. A simple equation that expresses this relationship is

$$\%DIM = K\,(SS/SR)^3 \qquad (18)$$

where K = 0.16 percent for our data.

This relationship is quite a valuable one to audio designers, as it indicates how DIM varies with SR ratio. A very interesting observation which can be made from Fig. 20 is that the DIM test is relatively insensitive to distortion detection, when the SR ratio is less than 0.5. It will, of course, measure gross IM levels for conditions of SS>SR, but this is hardly a practical mode of amplifier operation. Since the distortion mechanism being analyzed by a given test method is the inter-relationship of SS and amplifier SR, the point must be made that an optimum test method should show usable results over a wide range of conditions.

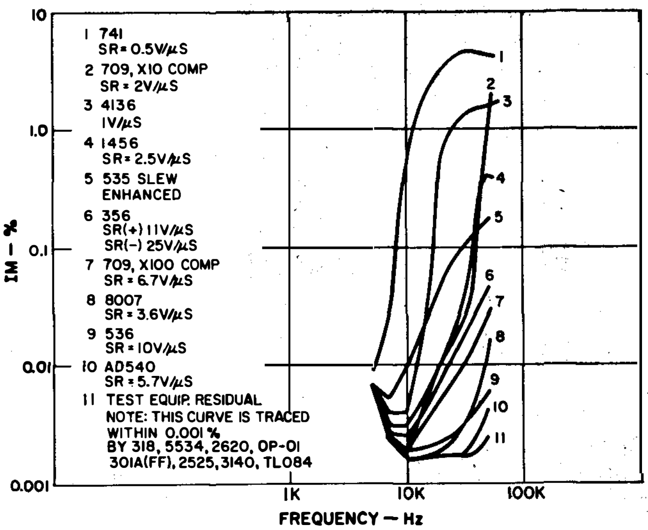


Fig. 19 — Ranking of op amps operated as unity gain inverters ±10 V output, by slew rate on basis of two tone (1:1) high frequency IM.

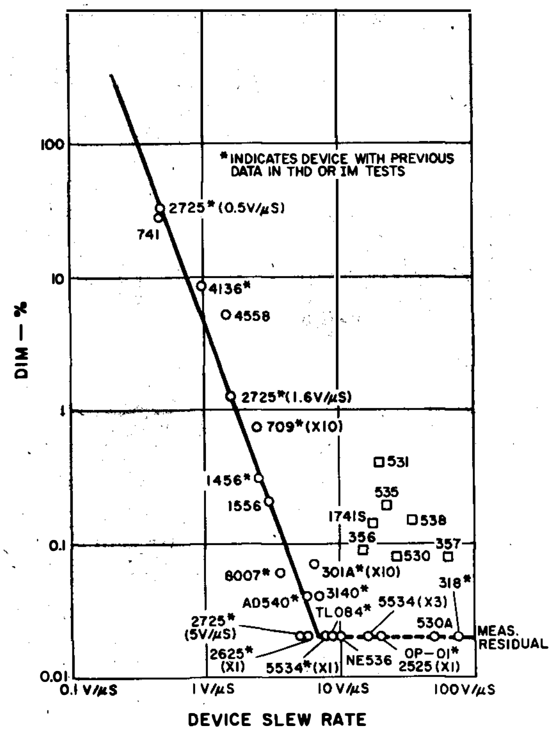


Fig. 20 — DIM vs. device slew rate. Test level is 20 V p-p, filtering is at 30 kHz, input SS is 3.2 V/μS. (*Indicates device with previous data in THD or IM tests.)

It should also be noted from Fig. 20 that there are devices that *do not* fit the characteristic relationship between distortion and slew rate. These devices are grouped to the right of the line and show excessive distortion for their high slew-rate capability (compared to the general trend). With the exception of the 356 and 357 devices, all of these op amps are slew-enhanced units. They feature an input-stage transconductance that varies with level to produce rapid slew rates for large signals. Unfortunately, the changing input-stage transconductance of these devices (a non-linearity), gives rise to a crossover type of distortion mechanism. Since, for small signals their SR capability is low, they begin to produce distortion for relatively low SS waveforms. As the SS of the input is increased, the slew capability of the device increases, and it is more capable of producing the required output. Thus, at high SS inputs, the distortion doesn't increase, it merely remains the same percentage as it was under low SS conditions.

We found that under varying input SS waveforms, the output spectrum of the slew-enhanced devices remained fairly constant; only the relative magnitudes of the individual distortion products varied up and down. Increasing the input SS caused some distortion terms to increase and some to decrease, but the magnitude remained fairly constant. It is interesting to compare this behavior with the leveling off of THD observed in the THD tests at high SS conditions.

The 356 and 357 devices also did not fit on the characteristic straight line, but they suffer from a different type of problem than do the slew-enhanced circuits. These units showed only even-order distortion falling on the square-wave harmonics; no other intermodulation products were produced (as did the slew-enhanced devices). The 356 and 357 devices seem to alter the symmetry of the waveform, indicating that an asymmetric nonlinearity is in action. This theory is supported by other forms of tests (for example, references 31 and 38). It should be understood that the problem experienced by these particular devices is not inherent in all Bi-FETs, or even other FET op amps, by any means. The 536, an older design, had DIM levels below the resolution of our measurement equipment. Also, the TL080 (and TL071) FET device families are capable of high performance for these tests, as is the LF351 and other devices of the same families.

Devices which are capable of differing slew rates, such as the 2725 and 301A, show DIM performance which improves as device SR is increased. In an experiment to examine the effects of open-loop bandwidth [4, 6, 7, 8, 9, 10, 14] and the degree of feedback as design criteria for low DIM, several

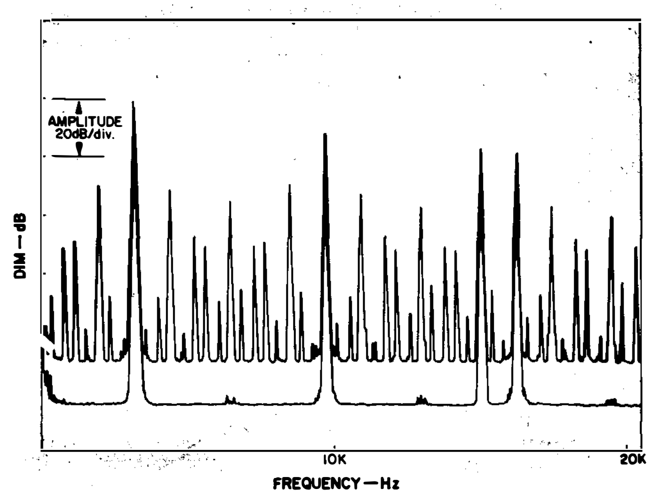


Fig. 21 — Comparison of DIM performance of two devices with different slew rates, both operated as unity gain inverters. Top curve is 741, bottom is NE536; input signal slope is 3.2 V/μS. Spectrum analyzer sweep, 0-20 kHz linear.

specific tests were performed. The results of these are the spectrum plots shown in Figs. 21 and 22.

Figure 21 shows comparative DIM performance for two different op amps for conditions of a 10V p-p output and a 30-kHz band limit (SS=1.6V/μS). The 0.8V/μS device (a 741) clearly shows strong DIM, but the 10V/μS device (a 536) shows a spectrum which is indistinguishable from the input. Open-loop bandwidth of both devices is less than 20Hz, feedback is nearly 100 dB at low frequencies, and gain-bandwidth is 1 MHz.

Figure 22 shows a performance comparison for 20V, 30 kHz (3.2 V/μS SS) band limit conditions, with slew rates adjusted to 0.5, 1.6, and 5V/μS, using the 2725 device. It is clear that DIM is reduced as the SR is increased above that of the SS (or

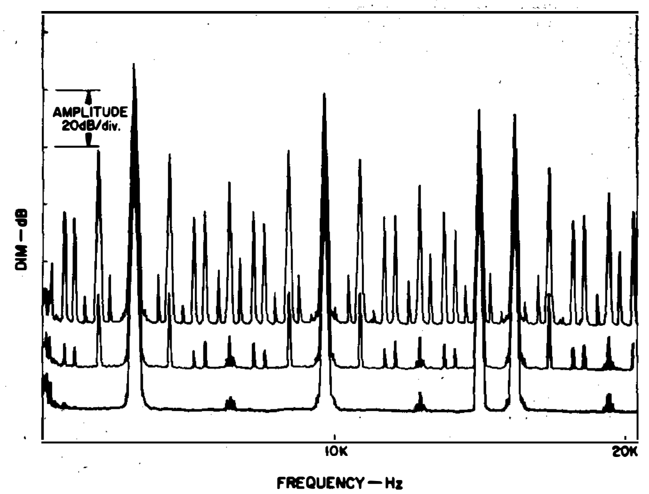


Fig. 22 — Comparison of DIM performance for a 2725 adjustable slew rate op amp operated as a unity gain inverter with input signal slope of 3.2V/μS. Top curve is SR of 0.5 V/μS; middle is 1.6 V/μS; bottom is 5 V/μS.

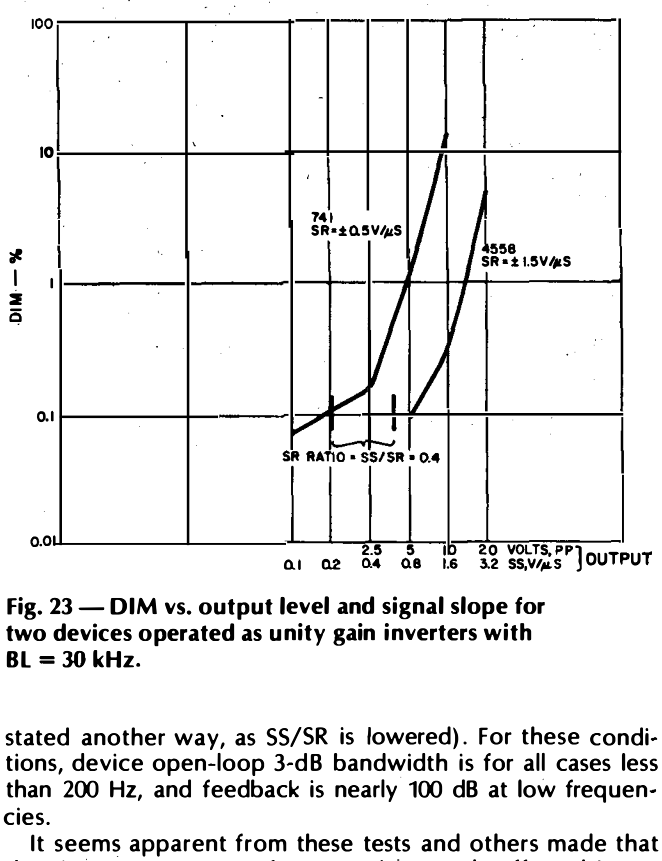


Fig. 23 — DIM vs. output level and signal slope for two devices operated as unity gain inverters with BL = 30 kHz.

stated another way, as SS/SR is lowered). For these conditions, device open-loop 3-dB bandwidth is for all cases less than 200 Hz, and feedback is nearly 100 dB at low frequencies.

It seems apparent from these tests and others made that the sine-square test performance is strongly affected by SR just as are THD and IM. There appears to be no directly measurable or obvious sensitivity to open-loop bandwidth. Gain-bandwidth product and loop gain affect DIM performance, as they do THD and IM, in that they affect how close to slew limit one can work before distortion rises.

A further demonstration of how DIM behaves in a manner similar to THD and IM performance is contained in Fig. 23. These data are based on the common condition of a 30-kHz band limit, but with DIM plotted versus output amplitude. To show the similarity, two different SR devices are used, 0.5 and 1.5V/μS. At low signal levels DIM is at a very low level; as the output signal level is increased, DIM shows a rapid rise, similar in behavior to THD and IM, as the SR ratio approaches unity.

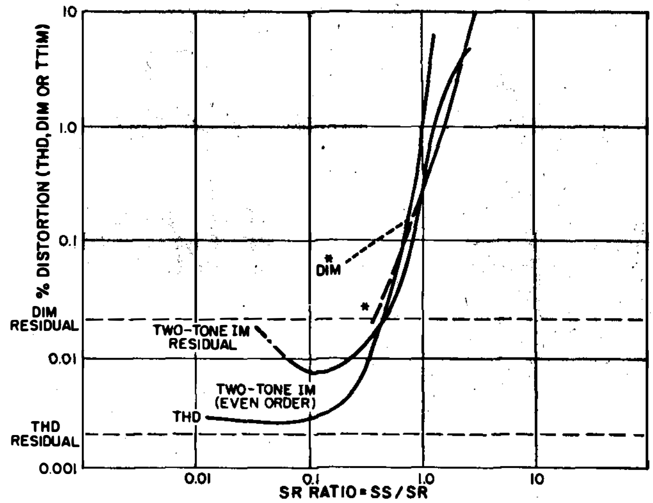


Fig. 24 — Correlation of test methods, various forms of distortion vs. signal/device SR ratio for a 741 with SR = 0.5 V/μS. (*Dotted line on the DIM curve results if the even order harmonics of the square wave are counted in the distortion computation. This is very difficult to do since an extremely high quality square wave generator is required. The proponents of this test recommend ignoring these components and measuring only the distortion resulting from the intermodulation between the sine and square waves, in which case a pattern similar to the dashed line results.)

Comparison of Test Methods

If the three test methods used are compared on a common base, it is possible to see a definite common pattern in their behavior, which is done in Fig. 24, where the horizontal axis is normalized in terms of the ratio of the SS to the SR of the device. By this means, it is possible to see just how the various forms of distortion behave as the device slew capability is taxed, and also to indicate the relative sensitivity of the three test methods.

The THD method shows the widest dynamic range of the three methods and gives the highest percentage distortion at a unity slew-rate ratio, 1 percent. The anomalous low-level slope for the TIM test is due to our detection of some second-order low-level nonlinearities in the 741 tested. This produced a second harmonic of the square wave which we were able to detect in the ouput spectrum. Since the TIM distortion number is normalized to the 15-kHz sine-wave amplitude, and the square-wave fundamental amplitude is 12 dB larger, the distortion shows up a factor of four larger than it should. Our experiences with the equipment available for these tests was that it was very difficult to detect SID with the sine-square test at signal slopes less than ½ that of the device SR (see Fig. 20).

Unfortunately, there is a serious problem with the sine-square test method that is not totally equipment related, one which became apparent after evaluating some of the best op-amp circuits. The problem concerns amplifier distortion products which are coincident with the even-order distortion products of the square-wave generator. Theoretically, a square wave should consist only of odd-order harmonics of the fundamental frequency. Practically, every generator will have some slight asymmetry in its square-wave output, which creates small but definitely measurable amounts of even-order distortion. Typical amounts for a general purpose lab square-wave generator are 50 to 60 dB down from the fundamental. Thus, if one were measuring a very good amplifier that had only low-level distortion products falling on the even-order square-wave harmonics, the true distortion of such a case would be masked by the generator, and therefore unmeasurable. The conclusion could then be erroneously drawn that the amplifier was free from transient intermodulation distortion, when in fact the amplifier was producing small amounts of distortion below the threshold of measurement.

One might point out that an amplifier producing distortion products coincident with the square-wave harmonics should also produce other intermodulation products of comparable magnitude, ones that could be readily measured. This simply was not the case in our tests and can be easily demonstrated by testing an asymmetric device such as a 356 or a 530A. Both of these amplifiers show the pattern of only even-order square-wave products, even at the most severe SS TIM test (10V/µS). To accurately measure these two devices, a square-wave generator with even-order products down at least 90 dB is required. In our series of tests, this was realized by carefully monitoring and adjusting the symmetry of our square-wave generator at periodic intervals. Only when the generator's even-order distortion was reduced to these low levels did we begin to see differences between the best op-amp circuits that typically had only even-order distortion products. The magnitudes of these even-order products for the best circuits were as low as only 0 to 6 dB greater than the generator residuals, and in many cases required detailed comparison of the input and output spectrum over several runs to verify that the products were, in fact, actually there.

The two-tone difference IM test is much more sensitive to even-order distortion than the sine-square test. For example, where it was difficult to detect distortion in the 356 with the sine-square TIM test, the IM test found it easily (Fig. 19). It is possible that a two-tone IM test designed to look for odd-order products would show superiority for finding odd-order distortion products. The main attraction of the TIM test is that it allows a quick *qualitative* look at an amplifier's performance.

THD evolves as a very desirable test method, as it is not only sensitive, but equipment for it is common. However, when a limited bandwidth circuit is being evaluated, some form of IM test becomes necessary.

Listening Tests

IC op amps from the group subjected to the above electrical tests were auditioned in a listening test [33] to assess the degree of correlation between the various forms of electrical distortion and audible defects. These tests were done in mono, in an inverting test amplifier configuration similar to Fig. 9. To sensitize the test for SID, however, the test device was preceded by a preamp to drive it to near full-scale output (and so, maximum SS) with program material. The full-scale output was then scaled down and level matched with the original input to within ±0.2 dB. A-B tests were then conducted on each IC to determine audible degradation. Source material was a variety of phonograph records, using a moving-magnet cartridge.

The results of this test indicate that not only can SID be detected audibly, but also suggest that the ear is apparently sensitive to levels of distortion lower than 1 percent. The results of these tests are summarized in Table II, which also indicates the relative quality weighting.

Before discussing these results, it is highly important that the reader appreciate the basic fact that these listening tests and the quality levels they indicate for a given SR are referred to 10V peak levels. One cannot generally assume these quality levels as absolute, as operation of given (fixed SR) device at other output levels will change the working SR ratio. As a necessary result, distortion will change accordingly, i.e. improve in going to *lower* levels or degrade in going to *higher* levels (for those devices capable of higher levels).

The above effects are, of course, simply due to the level-dependent property of SID; it is worst at highest SS or highest SR ratios. It is for this reason that the operating parameters associated with each test device are given here in several different terms, so as to avoid confusion. What the reader is most interested in, of course, is what parameters of a device are necessary to achieve a given quality level.

In terms of the reproduction observed, "A" level quality is that indistinguishable from the source on the most difficult high frequency program material. In general, devices of over 4V/µS slew rates fit into this category. Exceptions were some (but not all) slew-enhanced devices and the asymmetric devices. Quality levels B and C are degradations of a somewhat subtle nature, as noted. Quality level E and portions of D are distorted in a sense which is gross or obvious.

There appear to be two broad categories of audible SID, one which can be associated with the approach of slew limiting, Category I, and one in which slew limiting actually occurs, Category II.

Category II distortion will occur relatively infrequently on normal program material if the device slew rate is above 0.5V/µS. However, Category I distortion is possible in many instances, and adjectives used to describe it have often been seen in print.

Since the quality levels just described are for the devices and associated slew rates *operated at ±10V output levels*, some means of relating this to more general conditions is desirable. If the SR for each quality level is divided by the operating voltage level, it can be normalized to a required SR/V figure. This is simply the SR required per peak volt of output to attain a given quality level. For example, "A" quality level was observed for devices which achieved 0.4V/µS/V (or more) performance. This requires a 4V/µS device for 10V (peak) operation or a 0.4V/µS device for 1V (peak) operation.

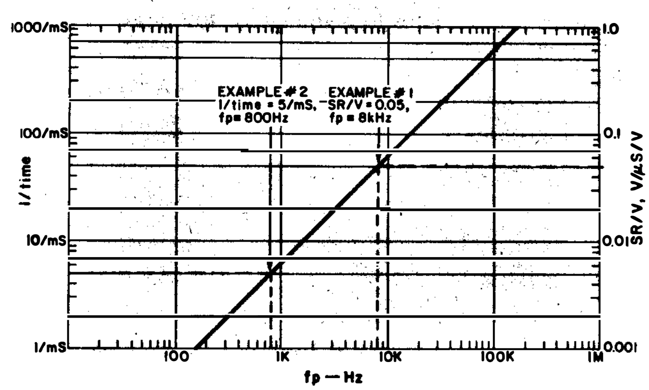
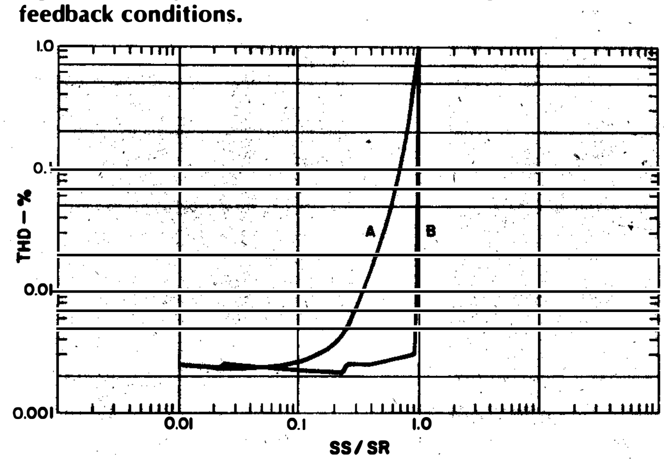


Fig. 25 — Relationship between SR/V (1/time) and fp.

Fig. 26 — Comparison of SID for low and high feedback conditions.



The parameter SR/V is related to a power bandwidth, which can be calculated as

$$fp=(SR/V)(10^6/2\pi) \qquad (19)$$

where fp is in Hz, and SR/V is in V/μS/V.

As can be calculated from (19) or the table, a 0.4V/μS/V SR/V level corresponds to a 64 kHz power bandwidth.

Some authors [62] have expressed the parameter of SR/V in units of 1/time (which may or may not appear to be simplified to the reader). A power bandwidth can also be calculated from this parameter as

$$fp=(\text{"x"}/mS)\ 1/(2\pi) \qquad (20)$$

where fp is in Hz, and 1/time is in 1/mS (x is the variable). These two relationships (19, 20) are graphically summarized in Fig. 25, with either parameter as an input.

The general observation which can be made from Table II is that obviously distorted reproduction begins to be noticed (level E) at an SR/V level of 0.05V/μS/V, or a power bandwidth of less than 8 kHz. This is plotted as Example 1 in Fig. 24. Other researchers conducting listening tests have arrived at a corresponding distortion level threshold in terms of 1/time, at a level of 5/mS [62] (or 0.005 V/μS/V), which equates to an 800-Hz power bandwidth (example #2).

The level dependence of SID has caused much confusion as to where and when a given SR is a limiting factor. The reader should understand that a 5V/μS amplifier SR (for example) will most likely not be a limitation for a preamp output, but may be critically so for a power amplifier. The difference is in the voltage swings the two types of amplifiers are called upon to produce without distortion.

As an illustrative example, if we assume a 1.5V power amp sensitivity for full output, this equates to roughly a 2V peak level from the preamp. To produce a 2V peak level in terms of the highest performance of Table II, the 0.4V/μS/V guideline implies a device SR of 0.4x2=0.8V/μS. This level of performance is met by many devices, for example the popular 4558 or 4559, at 1.5V/μS.

For the power amplifier, if we assume an example of 100W into 8 ohms, this equates to a voltage of 40V peak. Applying the highest performance level again of 0.4V/μS/V, the SR required is 16V/μS or more.

Hopefully, the above discussion illustrates how SR *must* be related to operating voltage level to predict quality. It should be appreciated that an SR number quotation *by itself* is relatively meaningless, if it is not related to operating level.

Also, another point which should be made is that performance simply does not continue to dramatically increase with greater SR, once sufficient SR has been obtained. For a preamp output, for example, if an SR of 5V/μS is more than sufficient to meet any possible operating condition, 50V/μS may not improve operation in practice and may well represent a meaningless numbers race.

As pointed out in reference 33, the listening tests of this research are basically the subjective observations of one individual and should not be construed as a result applicable to all situations.

Summary of Test Results

Some sensible guidelines for amplifier selection now begin to emerge from this series of tests. The primary one is speed, which is to say that faster amplifiers are generally better. There are two basic aspects to speed, bandwidth and slew rate, and in general they tend to go up together. It can be stated that raising an amplifier's gain-bandwidth product (or unity-gain frequency) is usually desirable [27, 28, 29, 30, 31]. The reason is that at any given frequency (neglecting d.c. and very low frequencies) the loop gain of the amplifier will be higher and more feedback-related distortion reduction will take place, which lets one work closer to the device SR (and, of course, reduces other distortions as well).

It has also become apparent that higher SR is generally better, even for *equal* bandwidth, but some caution is required here. Since slew rate is determined by the dynamic range of the (usually) nonlinear input transconductance amplifier, it is important that high slew rate *not* be achieved at the expense of linearity.

The slew-enhanced devices, such as the 535 and the asymmetric 356, are examples of amplifiers which violate this

Table II—Listening test results (referred to full output of ±10 V).

Category of SID		I—Deterioration		II—Gross Distortion	
Quality Level (1)	**A**	**B**	**C**	**D**	**E**
Audible Character	No differences detected for any program material.	Just discernible softening, loss of sweetness.	Further softening, somewhat dry, generally satisfactory with slight loss of dimension.	Colorations apparent, loss of dimension, "covered" sounds, dulled transients, constriction, edge begins.	Coloration and distortion obvious, more constricted covered sound, transients smeared, grit, edginess, fuzz.
Associated Parameters (2)					
SR	>4V/μS	2-4V/μS	1-2V/μS	0.5-1V/μS	<0.5V/μS
SR/V	>0.4V/μS/V	0.2-0.4V/μS/V	0.1-0.2V/μS/V	0.05-0.1V/μS/V	<0.05V/μS/V
fp	>64kHz	32-64 kHz	16-32 kHz	8-16 kHz	<8 kHz
Samples Tested	318, AD518	1456	1741S	741	
	NE5534 (TDA1034)		356*		
	2625		4741		
	2525				
	8007				
	NE536		NE535		
	AD540		NE538*		
	3140				
	TL084				
	OP-01				
	NE530A	NE530			
	NE541 (x100)	NE541* (x10)			
	NE540 (x100)	NE540* (x10)			
	NE531 (x10)		NE531 (x1)		
	2720 (5V/μS)		2720 (1.6V/μS)	2720 (0.5V/μS)	2720 (0.16V/μS)
	301A (x10, x100, or FF comp)			301A (x1)	
		4136 (2V/μS)	4136 (1V/μS)		
		709 (x10)			709 (x1)

Notes: *Audible ranking here is possibly due to factors other than SR.
1: Listing of various devices within *columns* is not a ranking. Character in column "D" is generally in category I, but may at times fall into category II.
2: Some prefer expressing SR/V in units of 1/time. As point of reference, 1V/μS/V=1000/mS.

premise. These devices are inherently incapable of performing as well as devices with more linear overall transfer characteristics. Emitter degeneration used in an input stage is an excellent example of a technique that allows higher SR [20, 21, 24, 47], while at the same time linearizing the input stage and extending its dynamic range. The 1456 or 318 is a good example of this type of amplifier; FET differential input types, which by their very nature have low transconductance, are also excellent (provided they are symmetrical). A good example of this type of topology is the TL080, TL071, or the LF351 series.

To restate these design criteria, we primarily want an amplifier which is linear for large input signal (ΔV) levels, and importantly, one which can deliver relatively large currents to the compensation capacitance [24]. This gives us high SR and a highly linear input dynamic range, which allows large error signals. Secondly, we would like this amplifier to have as high a unity-gain bandwidth as possible, so that when we apply feedback, the HF loop gain will be as high as possible for distortion reduction. The loop gain determines how close we can operate to slew limiting before distortion begins to rise (as it inevitably will, in a practical circuit).

Some previously discussed design criteria for low TIM, such as the use of low open-loop (d.c.) gain and a high open-loop pole frequency (ω_0), do not appear to be fundamentally necessary conditions for low TIM [33, 34, 45, 47], given an SR > SS.

From the above considerations, it seems useful to suggest a new form of SR criteria for audio circuits. From the four series of tests (THD, IM, sine-square, and listening), this would be a criterion which specifies a minimum SR with regard to the maximum output voltage level in use. Our criterion is "The circuit, including all possible loading conditions, should possess a (symmetrical and unenhanced) slew rate of 0.5V/μS (minimum) to 1V/μS (conservative) per peak output volt." Application of this simple criterion will result in negligible SID, either electrically or audibly, if the slew rate is symmetrical ($\pm$ 20 percent) and the input stage has a linear transfer characteristic (constant transconductance, unlike slew-enhanced types).

Inasmuch as the above criterion is a stringent one, and in view of some conservative operating conditions, some qualifiers could be added. In general, this criterion specifies an fp of 80 kHz, which is four times the generally accepted audio bandwidth of 20kHz. The reasoning behind this is the rise in distortion with the onset of slewing, sometimes described in the literature as "soft TIM" [7, 8]. Figure 26 illustrates this effect in curve A, which is the THD performance of a 741 on a normalized scale of SR ratio. As can be noted, appreciable distortion exists at ratios as low as 0.25 (or two octaves below fp) [33, 60].

Curve B is for a heavily fed-back amplifier, using a high gain bandwidth IC. As can be noted, distortion is at measurement residual levels right up to the point of actual slew limit (this has been referred to as "hard TIM" in the literature).

Obviously, in the case of B, less derating is necessary, since there is virtually zero distortion until actual slew limit. However, inasmuch as most practical amplifiers will show some distortion prior to slew limit, the 80 kHz fp is intended to guarantee a 20-kHz distortion-free bandwidth (for all output levels). Of course, for less than high performance uses, the criterion can be derated as the user sees fit.

This criterion is perhaps most applicable to amplifiers where the user does not have total control over performance parameters, such as SR, bandwidth, and input dynamic range (and/or linearity). For such applications, it can be useful as a guiding selection criterion, for example with IC types.

When one has design freedom from the ground up, and can optimize *all* operating parameters, other design approaches can be more useful. These are discussed in the final part of this series, which includes a design process to guarantee non-slew-limited performance.

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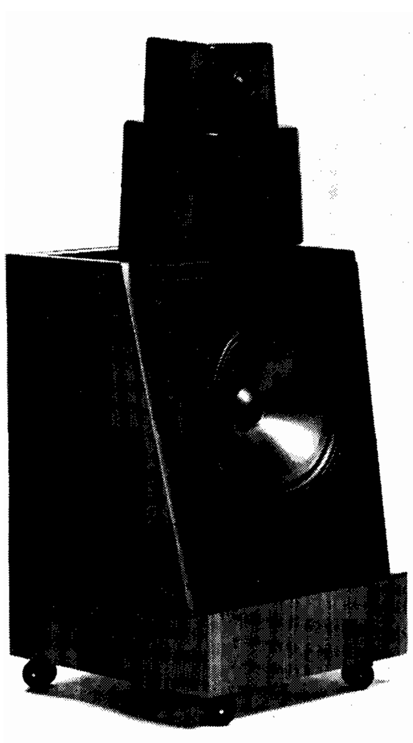

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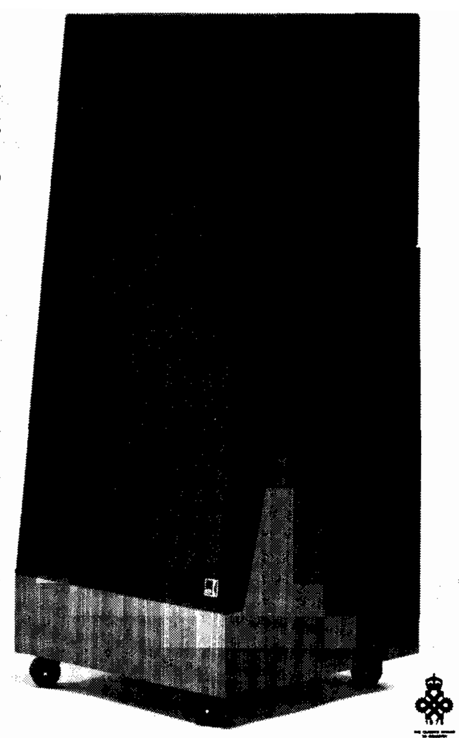

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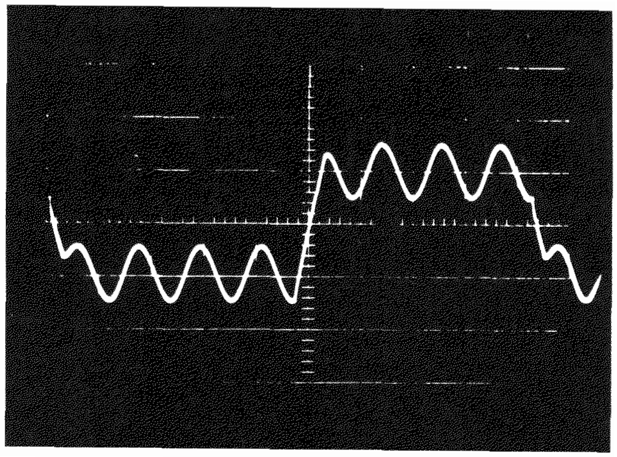
An Overview Of SID and TIM

Walter G. Jung, Mark L. Stephens, and Craig C. Todd

Part III — Analysis and Design of Amplifiers for Minimum SID

Calculation of Slew Induced Distortion

Thus far, little has been said in the literature about how to calculate slew induced or transient intermodulation distortion. This is, no doubt, due to the complexity of the problem, especially handling the frequency dependence of the amplifier stages and the incorporation of feedback. There is, however, a straightforward technique that can be used to find closed-form expressions for every possible harmonic or intermodulation distortion component. The technique involves forming a Volterra series to characterize the output as a function of some input variable [57]. The coefficients of the Volterra series can then be used to find the magnitude and phase of all distortion products. This technique has been widely used to predict distortion in radio frequency circuits with a high degree of accuracy.



Unfortunately, it takes more time and space to explain the technique itself than it does its application to a given problem. For this reason, we have not included a full analysis within the article and direct the interested reader to the reference cited. However, with appropriate assumptions and simplifications, many useful features of the Volterra series technique can be used to find approximate expressions for SID. These are conceptually easier to understand and are quite accurate for relatively small distortion conditions.

Consider a 741-type operational amplifier, which can be broken down into two basic stages, an input transconductance amplifier and an integrating amplifier. These are shown in Fig. 27. The transconductance stage is assumed to be the dominant nonlinearity and consists of a symmetrical saturating-type characteristic which is independent of frequency. The nonlinear characteristic (formed by a double differential pair) is modeled as a current source output Δi, for an input differential voltage ΔV, and can be represented by

$$\Delta i = I_K \tanh \left[\frac{\Delta V}{4V_T}\right] \tag{21}$$

where $V_T = KT/q$ or approximately 26 mV at 300° K and I_K = the bias current of the stage. The graph of equation (21) is shown in Fig. 28.

Equation 21 and Fig. 28 differ from equation 13 and Fig. 6b in our previous example of Part I, because the 741 input stage has a *pair* of transistors on each side. Equation (21) in its present form will not allow closed-form expressions for distortion. It must be expressed as a truncated power series with variable ΔV to complete the calculations, and this is shown in equation (22).

$$\tanh x = x - \frac{x^3}{3} + \ldots + \ldots . \tag{22}$$

Thus combining (21) and (22) we have

$$\Delta i = I_k \tanh \left[\frac{\Delta V}{4V_T}\right] \cong I_K \left[\left(\frac{\Delta V}{4V_T}\right) - \left(\frac{\Delta V}{4V_T}\right)^3 \frac{1}{3} + \ldots\right]. \tag{23}$$

The first term in the power series is the desired linear component, and the cubic term (and other higher order terms) form undesirable distortion products. Distortion will eventually be calculated from (23) after making some additional necessary assumptions.

The second stage in the 741, the integrator, is assumed to be ideal and has a gain characteristic G(f) which is proportional to 1/f. This is expressed by

$$G(f) = K_2/f. \tag{24}$$

There is a $\pi/2$ phase shift in (24) which has been neglected. The reason for this will become evident as the calculation progresses.

The constant K_2 is determined by the overall gain of the composite amplifier, which must be approximately unity at a frequency of 1 MHz to make our circuit model represent the performance of a real 741-type op amp.

The actual gain characteristic of a 741 op amp is summarized by the Bode plot in Fig. 29. For most audio-frequency calculations, it is convenient to neglect the low frequency pole at 10 Hz and to assume infinite d.c. gain and a constant, gain-bandwidth product. This has a negligible effect on calculations, since it will be shown that the distortion is determined by the available loop gain at high frequencies.

The open loop gain for this approximation is specified by

$$\text{open loop gain} = \frac{V_{out}}{\Delta V} = \frac{10^6}{f}. \tag{25}$$

Portions of this article are adapted from "Slewing Induced Distortion in Audio Amplifiers" by the authors in *The Audio Amateur*, Feb., 1977 (P.O. Box 176, Peterborough, N.H. 03458), part of an article series which is available in book form. Portions were also adapted from the authors' article "Slewing Induced Distortion — Its Effect on Audio Amplifier Performance, with Correlated Listening Results," Audio Engineering Society Preprint No. 1252 from the May, 1977, convention. (See bibliography references nos. 33 and 34.) **©Copyright 1979 by Walter G. Jung, Mark L. Stephens, and Craig C. Todd.**

By combining equations (23), (24) and (25), the constant K_2 can be expressed in more familiar terms. At a frequency of 1 MHz we have:

$$V_{out}/\Delta V = 1 = \left[\text{gain of transconductance stage}\right]\left[\text{gain of integrator}\right]$$

$$1 = \frac{I_k}{4V_T}\left[\frac{K_2}{10^6}\right] \qquad (26)$$

$$K_2 = \frac{4V_T}{I_K} \times 10^6. \qquad (27)$$

$$\text{And thus } G(f) = \frac{4V_T \times 10^6}{I_K} \times \frac{1}{f}. \qquad (28)$$

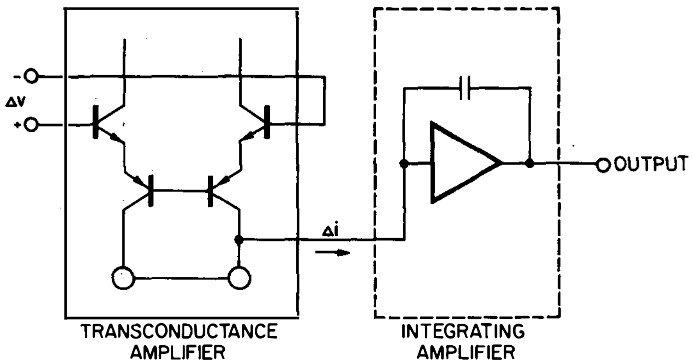


Fig. 27 — Two-stage model of an op amp.

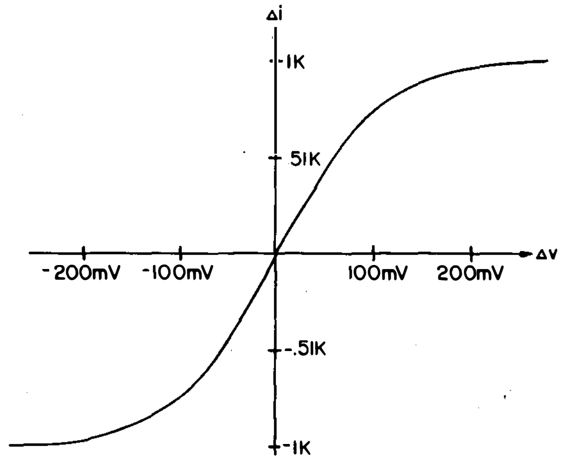


Fig. 28 — Transfer characteristics of a transconductance amplifier.

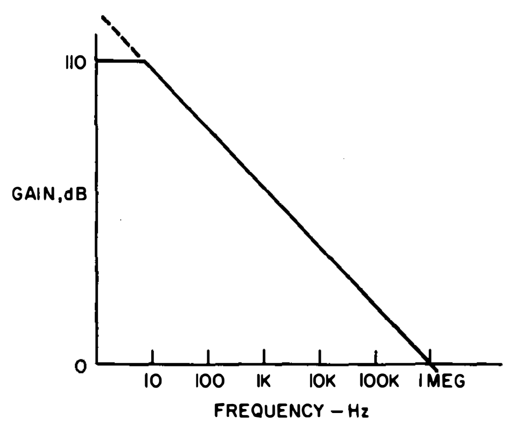


Fig. 29 — Gain-frequency characteristics for a 741 op amp.

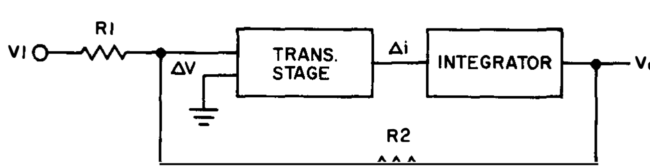


Fig. 30 — Model amplifier with feedback applied.

The 741-type op amp that has been developed thus far is now placed in an inverting gain configuration with resistive feedback components. The feedback network is assumed to be linear and independent of frequency. The circuit used for distortion calculations is modeled in Fig. 30. In this circuit, a feedback factor ß can be specified as a function of R_1 and R_2

$$ß = R_1/(R_1+R_2). \qquad (29)$$

Since the closed loop gain G is equal to R_2/R_1, we have

$$ß = \frac{R_1}{(R_1+R_2)} = \frac{1}{(1+|G|)}. \qquad (30)$$

For inverting gains of 1, 10, and 100 the factor ß is 1/2, 1/11 and 1/101, respectively.

Additional assumptions that must be made to simplify calculations are:

1) Small distortion conditions exist (<1%). This enables a power series expansion of the transconductance nonlinearity.

2) The distortion consists of only odd-order products because of symmetry, and, because of 1), the distortion is dominated by third-order terms.

3) The distortion is reduced by the magnitude of the factor (1 + loop gain), at the frequency of the distortion product. It is further assumed that loop gain is much greater than 1, so that distortion is reduced by approximately the magnitude of the loop gain. Any phase shift in the loop gain can therefore be neglected.

A harmonic distortion analysis will be developed here to compare with measured data, although an intermodulation analysis could also have been pursued. The final result will solve for harmonic distortion (which is dominated by the third harmonic) as a function of output voltage level, frequency, and feedback factor (or closed loop gain).

The following method will be used to solve for harmonic distortion. First, an output level V_o and frequency f will be specified. Then using (25), ΔV will be calculated and used in (23) to find open-loop distortion. Finally the loop gain will be computed and used to predict the closed-loop distortion.

For a sinusoidal output voltage of $V_o \cos 2\pi ft$, we can compute ΔV from (25)

$$\Delta V = \frac{[V_o \cos (2\pi f)t]}{(10^6/f)} \qquad (31)$$

If this ΔV is substituted into (23) and simplified, the resulting equation will show an open-loop distortion ratio of:

$$\frac{\text{magnitude of 3rd harmonic}}{\text{magnitude of fundamental}} = \frac{\left(\frac{\Delta V}{4V_T}\right)^2}{12} =$$

$$\text{Distortion (open loop)} = 1/12\left(\frac{V_o f}{4V_T \times 10^6}\right)^2 \qquad (32)$$

The open-loop distortion is reduced by the loop gain at the third harmonic frequency, 3f, and by the integrator frequency response which attenuates the third harmonic by a factor of 3. The loop gain at frequency 3f is

$$\text{loop gain} = \left(\frac{I_K}{4V_T}\right) \times \left(\frac{4V_T \times 10^6}{I_K\, 3f}\right) \times \beta = \frac{10^6}{3f}\,\beta. \qquad (33)$$

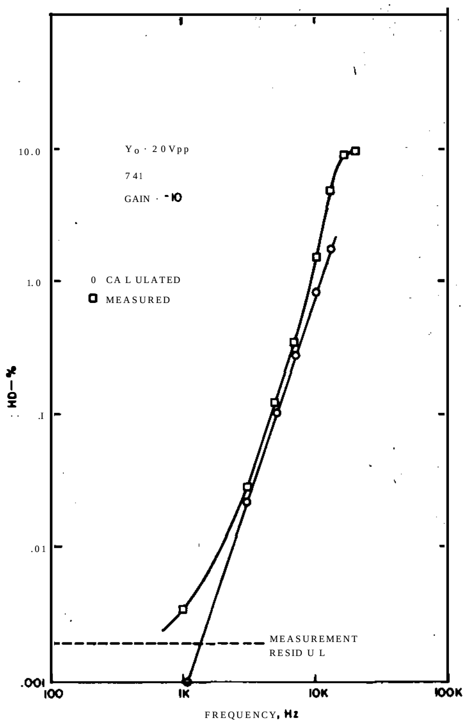


Fig. 31 — Calculated and measured distortion vs. frequency for a 741 at a gain of -1.

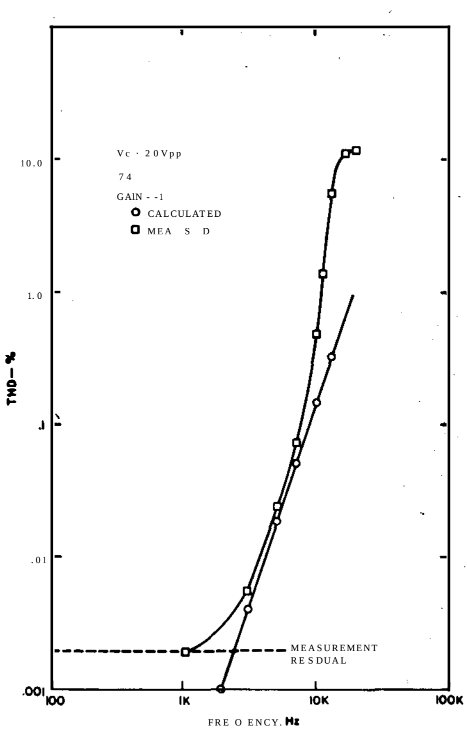


Fig. 32 — Calculated and measured distortion vs. frequency for a 741 at a gain of -10.

Therefore the closed loop distortion is

$$\frac{\text{distortion}}{\text{(closed loop)}} = \frac{\text{distortion (open loop)}}{\text{loop gain}} =$$

$$= \frac{1}{3}\left[\frac{\frac{1}{12}\left(\frac{V_o f}{4V_T \times 10^6}\right)^2}{\left(\frac{10^6\beta}{3f}\right)}\right] \qquad (34)$$

$$\text{THD(3rd)} = \frac{V_o^2 f^3}{12(4V_T)^2\beta \times 10^{18}} = \frac{V_o^2 f^3}{1.29 \times 10^{17}\beta}\,. \qquad (35)$$

Equation (35) shows that harmonic distortion should vary directly with the cube of the input frequency, directly with the square of output voltage, and inversely with the feedback factor, β. In order to test the accuracy of this equation, calculated data for distortion was compared directly with measured THD data from a 741 amplifier. Figures 31, 32 and 33 compare calculated and measured distortion for a constant-amplitude, swept-frequency test condition for three values of feedback factor, β. Figure 34 compares calculated and measured distortion for a constant-frequency, swept-amplitude test condition, also for three values of feedback factor. The agreement is generally good and is excellent for the swept frequency tests. At lower distortion levels, the agreement deteriorates due to the noise floor of the distortion analyzer.

At higher distortion levels, the agreement deteriorates due to large distortion conditions, that is, the fundamental assumptions in developing the calculation are violated. The anomalous behavior of the G = 100 test results is due to the loop gain falling below unity at 10 kHz, which also violates a basic assumption of the calculation. Figure 34 indicates an additional crossover type of distortion that dominates at low signal levels and masks the true distortion characteristics. It should be clear from all the figures that increasing feedback *reduces* distortion.

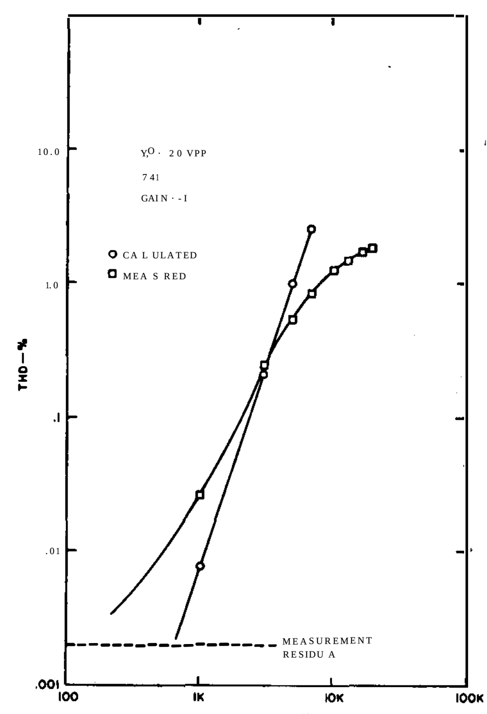


Fig. 33 — Calculated and measured distortion vs. frequency for a 741 at a gain of -100.

Equation (35) was developed specifically for the 741 op amp, which has a unity gain frequency (f_T) equal to approximately 1 MHz and a differential input stage consisting of four bipolar devices. A more generalized equation can also be developed which allows f_T to be a variable and which permits the number of input devices (n) to vary. This equation is

$$\text{HD (3rd)} = \frac{Vo^2}{12[n \cdot V_T]^2 ß} \left[\frac{f}{f_T}\right]^3 \qquad (36)$$

where n = number of bipolar devices (2, 4, 6, ...), V_T = KT/q = 26 mV at 300° K, ß = feedback factor, f_T = unity-gain frequency, Vo = output voltage, and f = frequency of fundamental.

Equation (36) reveals some characteristics of SID which were not evident from equation (35). First, it can be seen that increasing n reduces the distortion. This is due to a reduction in the curvature of the input transconductance curve (i.e. less change in g_m for the same current change) as n increases. Unfortunately, practical limitations usually require n to be 2 or 4 at most, so increasing n has a limited usefulness in reducing SID. Second, equation (36) shows the strong effect of the unity gain frequency on SID. Increasing f_T by a factor of 3 results in a distortion reduction of almost 30 dB! Clearly, f_T is a highly important parameter in improving SID. However, it is important to make the close connection between f_T and the SR limit. As pointed out by Solomon [21, 24] and others, for the 741-type circuit topology with bipolar input devices, f_T is proportional to the SR limit. This relationship is shown below

$$SR = 2\pi f_T(4V_T). \qquad (37)$$

Therefore, improving f_T produces a proportionate improvement in the SR limit, which reduces SID.

Results of SID Calculation And Comparison with Measurements

The demonstrated accuracy of (35) and the generalized form in (36) in predicting harmonic distortion in a 741 amplifier leads to some useful conclusions concerning slew induced distortion.

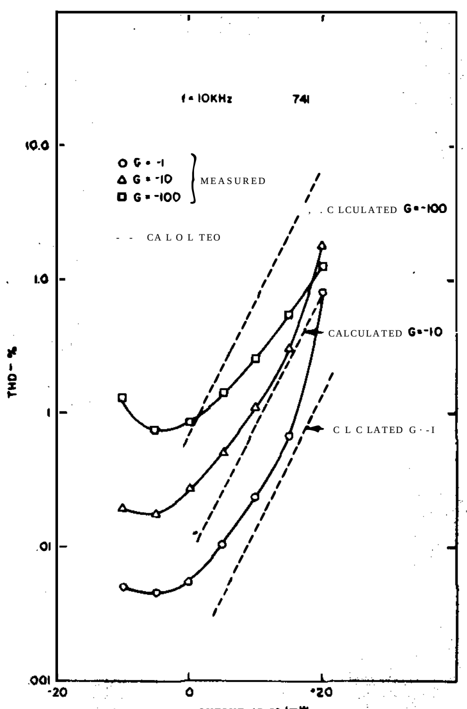


Fig. 34 — Distortion vs. output level for a 741 at various gain levels.

1) It means that slew induced distortion can be modeled and calculated with closed-form expressions, based on Volterra series principles.

2) It shows that slew induced distortion is increased by the input signal slope (SS) and the sharpness of the transconductance curve. It also shows that SID is decreased by more feedback and by a higher gain-bandwidth product.

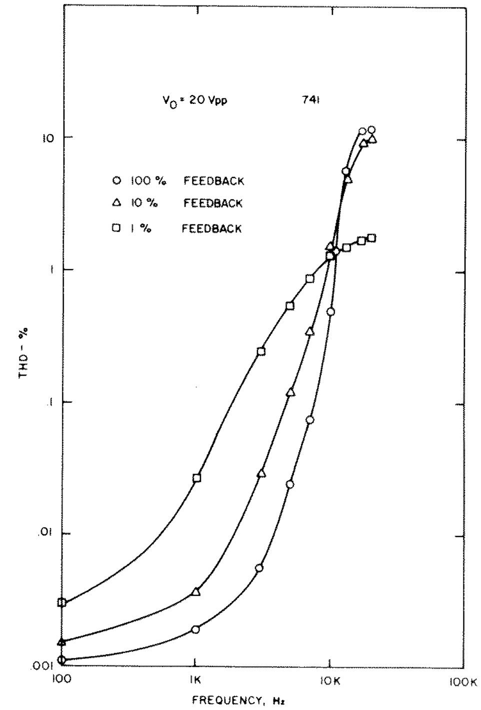


Fig. 35 — Distortion vs. frequency for a 741 at various feedback conditions.

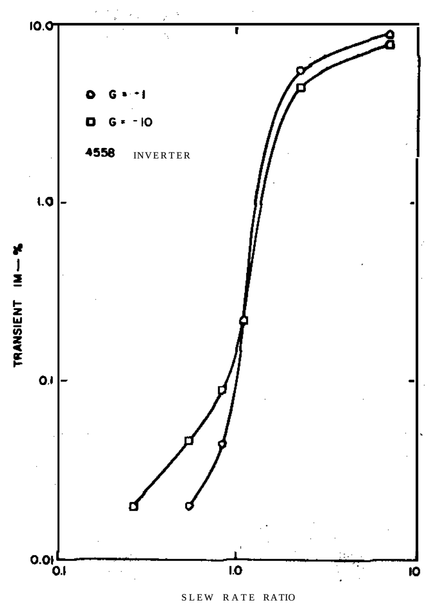


Fig. 36 — Transient intermodulation distortion vs. slew rate ratio for a 4558, operated inverting at two gain levels.

3) It demonstrates that since the slope of a constant amplitude sine wave is proportional to its frequency, that SID (or DIM, as in the sine-square test) should vary as the *cube* of the input SS. This is confirmed by the data in Figs. 20 and 24 that show the variation of DIM with SS is a cubic relationship.

4) It shows that increasing a device's slew capability, without adding additional nonlinearities (like slew enhancement), will reduce slew induced distortion.

The Effect of Feedback for SS>SR

Present TIM theory suggests that feedback increases distortion. Our measurements and calculations show that, at least for signal conditions *below* the slew rate limit (SR ratio <1), that feedback reduces distortion. The overall effect of feedback on distortion (for a constant slew rate capability) is shown by our data to depend on whether the SS is less than or greater than the SR limit. For SS<SR, increasing feedback reduces distortion. For SS>SR, increasing feedback increases distortion. There is a crossover point around SS=SR where feedback has a minor effect on distortion. These trends are evident in the THD plot of Fig. 35 and the sine-square (TIM) plot of Fig. 36. It should be remembered, however, that for distortion-free performance the SS must be less than the SR, and if this criterion is met, feedback can generally be relied on for distortion reduction. Operating an amplifier with the SS>SR is simply not a realistic consideration for high-fidelity reproduction. Some discussion and experiments of the next section will clarify these points further.

Designing for Minimal SID or TIM

We have now reached a point where the factors which govern the behavior of the SID mechanism have been discussed in principle. However, the discussion thus far has been largely focused on behavior as viewed from outside an amplifier or how to characterize it in terms of SID.

Perhaps more important is how to design an amplifier from the ground up for minimum susceptibility to SID or TIM. This section focuses on these aspects of the situation and develops techniques which can be used to predictably model circuit performance.

We will begin the discussion by returning to a two-stage amplifier model, shown in Fig. 37, which is similar in many regards to Fig. 5a of Part I or to Fig. 27 above. This two-stage circuit will now be used to develop a general topology which can be used to model amplifier performance and also dramatically illustrate the TIM and SID phenomenon.

A circuit topology similar to Fig. 37 was described 10 years ago in a classic paper by Solomon [24] et al. This paper contained a number of defining behavioral relations, which are not only historically important, but are also applicable to amplifiers of this type in general [47, 64].

A basic point which should be appreciated with regard to this two-stage amplifier is that one can actually design it to yield a given overall gain-bandwidth for an *infinite* set of combinations of stage 1 and stage 2 gains. The key question is, does it matter whether stage 1 or stage 2 furnishes the bulk of the gain? For herein lies the answer to the entire TIM and SID problem. In other words, how should the gain be partitioned between the two stages for best overall performance? Before we plunge into the equations which govern this, perhaps some discussion would be helpful towards insight.

We have already established by (14) that the SR which will be seen at Vo is set by I_K and C1. However, we also know that to increase SR we cannot just arbitrarily increase I_K or decrease C1, because of stability reasons. We must also decrease g_m simultaneously with either of these measures to maintain stability. In general, a lower g_m implies less gain in stage 1, i.e. the stage can accept greater input error signals ΔV before the saturation which results in TIM and/or SID is reached. Thus, it can be said that *to maximize SR in a given bandwidth, the stage preceding the integrator of a two-stage amplifier design such as this must have a low g_m and high I_K.*

Solomon expressed this as a *low* g_m/I_k ratio in [24] and [21], and it has also been expressed as a *high* I_k/g_m ratio by Gray and Meyer in [22]. The latter form allows an expression to be written which directly describes the amplifier's maximum input-voltage capability or dynamic range. This is the voltage which, when exceeded, will result in slewing. It is simply

$$Vth = \frac{I_K}{g_m}. \tag{38}$$

Others have termed this the input-voltage dynamic range [50]; however, the meaning is similar.

A greater application for how these relationships function may be obtained by examining two representative IC amplifiers with dissimilar Vth's. These types are used as examples because they are externally compensated and readily available. This allows convenient experimental duplication. A 301A amplifier (or 741, as noted above) has a g_m of

$$g_m = \frac{I_K}{4V_T}. \tag{39}$$

This equation can be expressed in terms of I_K/g_m or Vth, as

$$V_{th(301A)} = \frac{I_K}{g_m} = 4V_T. \tag{40}$$

Since V_T = 26 mV at room temperature, a useful approximation of (40) is

$$V_{th(301A)} \cong 0.104 \text{ V}. \tag{41}$$

Thus, a peak input voltage of 104 mV to a 301A (or 741) will cause it to slew.

To turn to another amplifier type, a representative FET input device is the TL070 (or TL080) which has a g_m of approximately

$$g_{m(070)} \cong 1.5\, I_K. \tag{42}$$

If this relation is expressed in terms of Vth, it becomes

$$V_{th(070)} = \frac{I_k}{g_m} = 0.67 \text{ V}. \tag{43}$$

As can be noted by comparison of (41) and (43), the TL070 FET achieves a Vth more than six times that of the 301A bipolar for similar conditions, This, for a comparable bandwidth, produces a higher SR. For example, for a 1 MHz bandwidth condition, the TL070 has a 4.3 V/ μS SR (Cc = 47 pF), while the 301A is only 0.67 V/ μS (Cc = 30 pF) [31].

For the amplifier model under discussion, a relationship can also be drawn between Vth, SR, and gain-bandwidth product (GBP) similar to that expressed in [24]. We use the more general GBP, rather than f_T, since GBP can often exceed f_T. Also, f_T is usually taken to mean the unity-gain crossover frequency and implies unity-gain stability. This is not always a requisite. An expression for SR in these terms is

$$SR = \frac{(V_{th}\, 2\pi\, GBP)}{10^6} \tag{44}$$

where SR is in V/ μS, Vth in volts, and GPB is the gain-bandwidth product (Hz) *at audio frequencies.* (The relevance of this equation to the subject of TIM and SID is fundamental. Although first described by Solomon and others [24], the authors would like to document that this relationship's importance to *audio amplifier* performance has been previously noted in letters to the A.E.S. by R. Cordell of Bell Labs, 9/77, and B. Olsson of Xelex AB, 11/77 and 2/79.)

This relationship clearly demonstrates that SR is directly proportional to Vth and GBP for this model. However, the caution should be extended that it does not apply universally. Two particular exceptions are some feed-forward amplifiers and slew-enhanced circuits such as IC type 531. In the case of a feed-forward type, such as the 5534, Vth is not

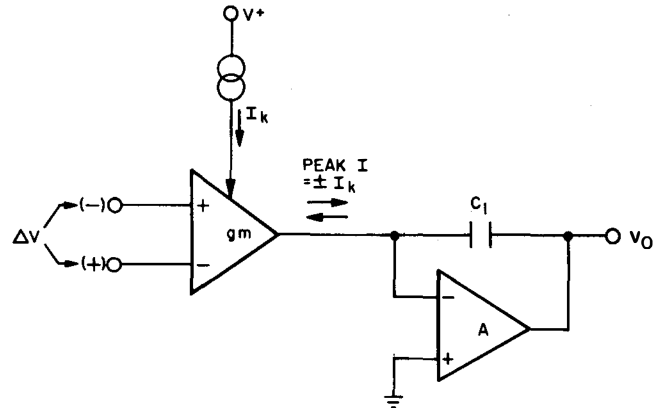
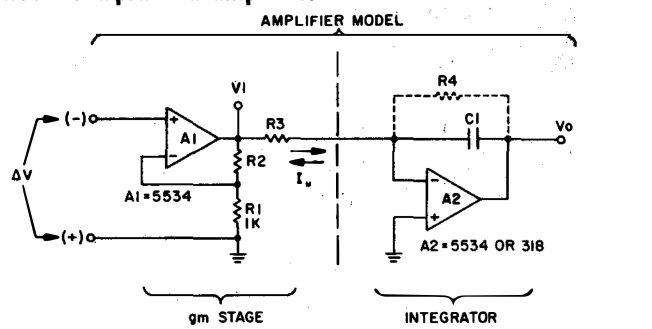


Fig. 37 — Two-stage transconductance-integrator model of a practical amplifier.



$$gm = \frac{\Delta i}{\Delta V} = \left(\frac{R1+R2}{R1}\right)\left(\frac{1}{R3}\right) \quad (45)$$

$$I_M = \frac{V1(max)}{R3} \quad (46)$$

$$Vth = \left(V1(max)\right)\left(\frac{R1}{R1+R2}\right) \quad (47)$$

$$SR = \frac{I_M}{C1} = \frac{V1(max)}{R3\ C1} \quad (48)$$

$$fp = \frac{SR}{2\pi Vop} = \left(\frac{V1(max)}{R3\ C1}\right)\left(\frac{1}{2\pi Vop}\right) \quad (49)$$

$$GBP = \frac{gm}{2\pi C1} \overset{(50)}{=} \left(\frac{R1+R2}{R1}\right)\left(\frac{1}{R3}\right)\left(\frac{1}{2\pi C1}\right) \quad (51)$$

$$fo = \frac{1}{2\pi R4\ C1} \quad (52)$$

Fig. 38 — Synthesized two-stage amplifier model.

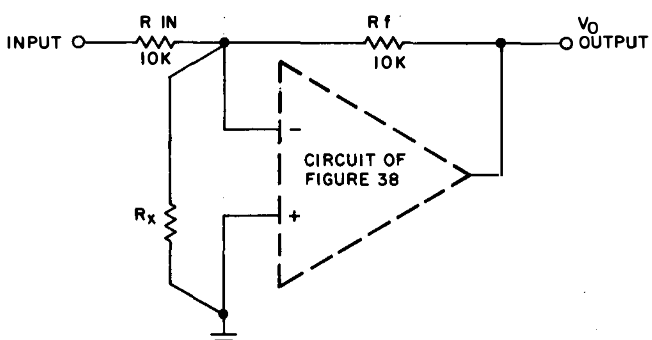


Fig. 39 — Test circuit for synthesized model.

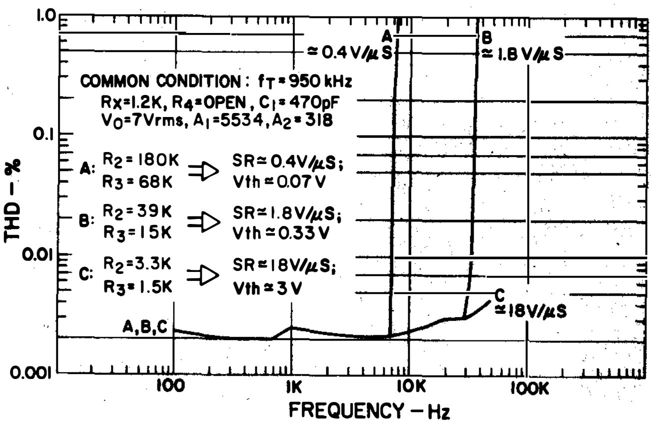


Fig. 40 — THD vs. frequency of the synthesized 741 op-amp model for various rate slew conditions (Test 1).

altogether a straightforward predictor of SR, as its Vth of 52mV and f_T = 10 MHz predicts only a 3 V/ μS SR. However, the GBP of this device is actually 22 MHz at audio frequencies — if this figure is used in (44), the SR predicted is 7 V/ μS, which agrees reasonably well. An important point is also that one must *not* be misled into the belief that slew-enhanced devices, which can show large voltages for Vth, lead directly to quality results. As has been shown previously, such amplifiers must be treated individually, as their dynamic input nonlinearities makes them special cases.

The relationship described by (44), while certainly an important one, can be erroneously misinterpreted. For example, it should not be interpreted to mean that *only* a very high Vth is fundamentally the route to high SR and thus low TIM. As (44) clearly shows, raising GBP (where allowable) achieves a similar result, and a practical example is the amplifier compensated for a higher noise gain (and thus GBP), such as the 301A of Fig. 11 (Part II). Such an example illustrates a *low* Vth device (the 301A) achieving a *high* SR. Another example is the 5534, a high GBP device, but with a very low Vth, only 52 mV! And, it should be noted, sufficient GBP must be present to result in a useful final closed-loop bandwidth.

The important thing to be remembered for this relationship is not totally Vth or GBP in absolute terms, but their *interrelationship,* which in many cases can be manipulated to achieve a high SR. The concept of a high Vth is, of course, most important when one is attempting to maximize SR *with a given GBP,* for as (44) shows, it is the only way it can be done with this type of circuit topology.

Experiments Which Demonstrate The Principle

A very cogent demonstration of the just described relationships can be made by synthesizing a two-stage amplifier model and subjecting it to various feedback and open-loop performance combination.

The circuit used for a series of these experiments is shown in Fig. 38 and is actually composed of two local-feedback IC op amps, which together comprise the model. A1 performs the function of a g_m input stage, converting the input voltage ΔV into a proportional current in R3. A2 performs the function of the integrator. Actual devices used for the experiments to be described were the 5534 for A1 and either a 5534 or 318 for A2. The devices used must, of course, have an inherent SR in excess of that which will be demanded by the model's operating conditions, as well as low distortion. These factors, combined with the local feedback, yield an amplifier with virtually ideal characteristics (even without overall feedback) as any nonlinearities are strongly suppressed.

A series of performance defining equations are included in the figure, and these can be manipulated with a great degree of freedom (another reason for using a model such as this, in fact). Some comment on these relations is in order before they are put to use, though.

Transconductance of the A1 stage is defined as

$$g_m = \left(\frac{R_1+R_2}{R_1}\right)\left(\frac{1}{R_3}\right) \quad (45)$$

Maximum output current, (I_m), is defined simply by the clipping voltage limit of A1, V1 (max) divided by R3, or

$$I_m = \frac{V_{1(max)}}{R_3}. \quad (46)$$

I_m, the peak output circuit, is analogous to I_K of Fig. 37, in that it sets the SR. It is slightly different in this case, due to more design freedom.

It is important to note that these two relationships are not exactly equivalent to those associated with Fig. 37. For example, the g_m of Fig. 38 can be set independent of I_m (if desired), and I_m can be set independent of g_m (if desired). This extra flexibility and the use of a voltage amplifier to produce V1

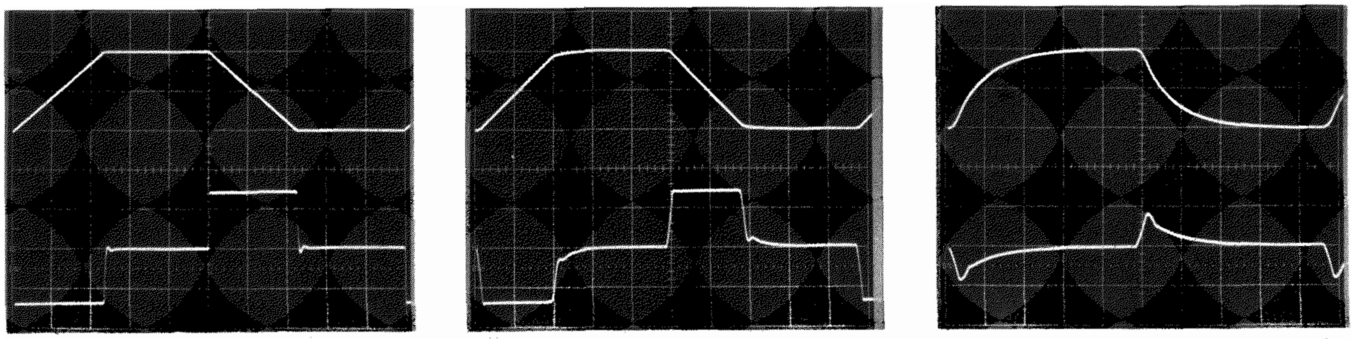


Fig. 41 — Transient performance of synthesized op-amp model with various slew rates to a 20-V p-p square wave of various frequencies. Top traces are outputs, bottom traces error voltages. A, SR = 0.4 V/ μS, 5 kHz; B, SR = 1.8 V/ μS, 20 kHz; C, SR = 18 V/ μS, 50 kHz. (Scales: All, 10 V/cm; A, 20 μS/cm; B,5 μS/cm; C, 2 μS/cm.)

yields a direct monitor of the conditions in the input stage. A standard g_m input stage does not allow voltage monitoring of error signals.

Because of the above, Vth in this circuit is simply

$$Vth = V_{1(max)} \frac{R_1}{R_1+R_2}. \quad (47)$$

As this expression shows, Vth is simply the output overload voltage of A1, divided by the gain set by R1-R2.

The remaining performance equations are simply derived from combinations of others; as the figure shows

$$SR = \frac{V_{1(max)}}{R_3C_1} \quad (48)$$

$$fp = \left(\frac{V_{1(max)}}{R_3C_1}\right) \left(\frac{1}{2\pi V_{op}}\right). \quad (49)$$

Gain bandwidth product (GBP) follows from (8)

$$GBP = \frac{g_m}{2\pi C_1}. \quad (50)$$

Substituting g_m as described by (45), this becomes

$$GBP = \left(\frac{R_1+R_2}{R_1}\right) \left(\frac{1}{R_3}\right) \left(\frac{1}{2\pi C_1}\right). \quad (51)$$

Open loop bandwidth (in the presence of R4) is

$$fo = \frac{1}{2\pi R_4C_1}. \quad (52)$$

Without R4, it is reasonable to regard A2 as a near-ideal integrator, in which case fo is well below the audio range for practical values of C1, and the gain-bandwidth product is constant throughout the audio range, as set by (51).

Test Results

The first test (Test 1) performed on the model was to synthesize a standard 741 op amp in terms of GBP and manipulate it for differing SR. The results should show very linear behavior up to fp, and a hard limit or sudden distortion rise as slew limiting is reached. Conditions were set up for a unity signal gain inverter, with a noise gain of 20 dB, using the test circuit of Fig. 39.

Figure 40 shows the results of Test 1 for a THD swept-frequency test, at an output of 7V rms. Conditions A, B, and C are approximately 0.4, 1.8, and 18 V/μ S respectively. The different circuit conditions to yield these SR are noted. As should be noted, since GBP and the feedback conditions are identical for all three of these tests, the only variables are SR and Vth.

As can be noted from the A and B curves, these conditions produce a sudden distortion increase when the SS of the test signal equals the amplifier SR. The high SR of condition C prevents the limit from being reached, for any test condition. Note that Vth increases, going from A to C, in the same proportion as SR.

For a case of transient signal condition, the photos of Fig. 41 show how this same amplifier behaves for the three conditions set down in Fig. 40, but with a different method of measurement.

Figure 41A shows waveforms for the "A" test condition (SR = 0.4 V/ μS) for a signal condition of a 5-kHz, 20-V p-p square-wave input. The top trace shows the Vo waveform, which clearly resembles a 741 – type response (31, 38), changing 20 V in over 40 μS. Inside the loop, the error voltage V1 is shown at the bottom. Here it is seen that V1 saturates negative, then positive, for the corresponding (+) and (-) slew intervals, respectively. It is clear from this photo that the slewing evident in Vo is a result of saturation in V1.

Figure 41B shows waveforms for the "B" test condition (SR = 1.8 V/ μS), with a 20-kHz, 20-V p-p square-wave input. At the top, the Vo waveform shows that slewing is present, as is evident by the linear (+) and (-) slopes. This is confirmed by the V1 waveform, which again indicates saturation of the 1st stage for these corresponding times. This is similar to Fig. 41A, but the difference is that for this higher SR condition, the slewing intervals are simply shorter (note scale factor differences — do not be misled by same *general* waveshape).

Figure 41C is very interesting, because it demonstrates that a sufficiently high SR and Vth can completely prevent saturation of the first stage and maintain operation within the small signal region entirely. Conditions of these photos are an SR of 18 V/ μS. However, the feedback conditions described above in conjunction with the 20-dB noise gain result in an amplifier closed-loop, small-signal bandwidth of 95 kHz. This in turn is equivalent to a single-pole, low-pass filter with a time constant of 1.7 μS. For a 20-V p-p output from this filter (the amplifier), the maximum signal slope is 12 V/ μS.

For Fig. 41C, the signal input is a 20-V p-p 50-kHz square wave, and it can be noted that there is no slewing evident in Vo. The waveform is exponential in shape with a risetime of about 4 μS — consistent with the small signal relationships.

That slewing is not present is also confirmed by V1, which shows that the error voltage remains below the clipping level. Note that the highest amplitudes of V1 occur at the peak SS of Vo or at the transition points of the square wave.

This particular test confirms in another way the point made in Part I of this series, that slewing can be prevented by maintaining the amplifier small-signal bandwidth at a lower frequency than the power bandwidth. In 41C, fc is 95 kHz, but fp is 290 kHz, and no slewing is evident.

With this same model, experiments were also conducted to examine the sensitivity of the amplifier to open-loop bandwidth (fo). Test two conditions were commonly set up as described in Fig. 42, which resulted in an SR of 1.3 V/ μS and an f_T of 540 kHz. For this test circuit with R4 present at

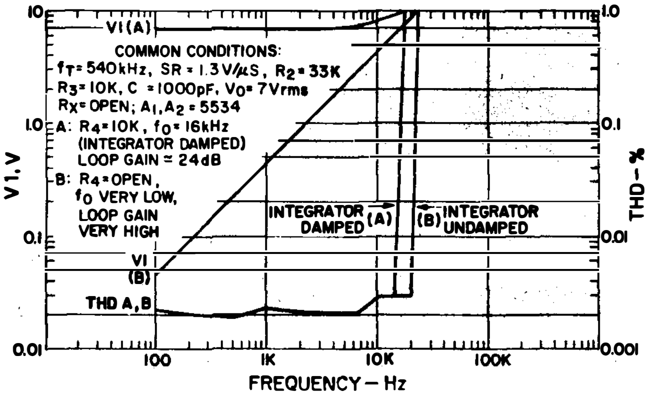


Fig. 42 — THD and error voltage vs. frequency of the synthesized 741 op amp model for different open-loop gain conditions (Test 2).

10K, fo becomes 16 kHz and the open-loop gain is 30 dB. With Rx open, the feedback is then 24 dB. With R4 open, the circuit becomes a classic op amp, with a very high open-loop gain and fo very low. Note, however, that *GBP* remains *unchanged for either condition.*

For condition A, where R4 is 10K, THD curve A indicates that slew limiting is reached at 18 kHz. V1 (A) is a plot of the rms error voltage versus frequency. Since it is essentially flat with frequency, it is testimonial to the wide open-loop bandwidth. Note that V1 increases to its clip level at 18 kHz, coincident with the slew rate limit point.

The B condition shows corresponding results with R4 removed, and the most obvious difference is the (apparent) increase in fp. Error voltage V1 (B) now increases 6 dB per octave with frequency, the inverse of the integrator's gain rolloff — what is necessary to maintain a flat output versus frequency for the overall circuit.

The *apparent* increase in SR for condition B is not an increase for this condition, but rather reflects a less than potential maximum SR *for the A condition.* This is so because the 10K resistor loading the integrator absorbs a portion of the charging current available to Cl for slewing.

These points are also brought out in the square-wave photos of Fig. 43. This shows response of the circuit of Fig. 42 to a 5-kHz, 20-V p-p square wave for conditions A and B.

For these test conditions, the transient performance is shown in Fig. 43. The slewing in Vo shown in 43A shows a quasi-linear ramp or a combination of ramp and exponential waveform caused by R4. Since R4 constrains the open-loop gain to a relatively low value, this is also reflected in the large error voltage shown in V1 (bottom).

The voltage V1 is clipped for the slew intervals (as expected) but also shows a very large potential (10 V) for the steady = state waveform positions. This excessive error voltage reflects a relatively large gain error for this circuit.

Figure 43B shows the Vo and V1 response for the same input drive but with R4 removed or condition B. Note that in 43B the slewing intervals are shorter and linear, as would be expected due to the constant and larger C1 charging current available. The error voltage shown in V1 is much lower in the steady-state periods, reflecting the increased gain available in the integrator. The low gain error is also reflected (more subtly) in the greater amplitude in Vo, compared to Fig. 43A.

This test indicates that, by both THD and transient response tests, there is no inherent advantage to a wide open-loop, small-signal bandwidth. By contrast, there are definite disadvantages to the constraint such operation can place on amplifier characteristics, such as limited LF loop gain and also some sacrifice in SR. And, while it is not apparent from this particular experiment, loading an integrator stage in a conventional amplifier will usually degrade the open-loop distortion characteristics.

Predicting A Non-Slew-Limited Response

We are now at a point where the information developed can be merged into a set of relationships useful in designing a *non-slew-limited* amplifier or an amplifier which is free of SID and TIM, by definition. This evolves in a fairly straightforward manner from the relations just discussed.

A non-slew-limited amplifier is simply one which *cannot* be made to slew for any signal input level below that which causes *amplitude* clipping. Input waveform shape is unrestricted and may include all waveshapes up to and including square waves. The square wave (as discussed in the sine-square box of Part II) is the most rigorous test to which an amplifier can be subjected because of its very high SS (infinite, for an ideal square wave). Therefore, if an amplifier can be proven to be free of slewing distortion for a square-wave test for all signal amplitudes in its linear range, it is by definition non-slew limited and will be largely free from SID or TIM problems.

All amplifiers will have by design a small-signal bandwidth, fc. This bandwidth will either be determined by the feedback configuration or an input pre-filter. The amplifier will then band limit a square-wave input signal to a bandwidth of fc. For simplicity at this point, we will assume this to be a single pole rolloff. For such a filter response it can be shown [33, 67, 70] that the signal slope of the resulting band–limited–output square wave is

$$SS_{(sq)} = \frac{2\pi V_{pp} f_c}{10^6} \tag{53}$$

where Vpp is the peak-to-peak amplitude at the filter output, fc is the small-signal bandwidth, and SS is in V/µS.

That this signal slope is much higher than a sine wave at fc (passed through the same filter) can be shown by the relation of the two slopes. A sine wave at fc will be down by 3 dB

A

B

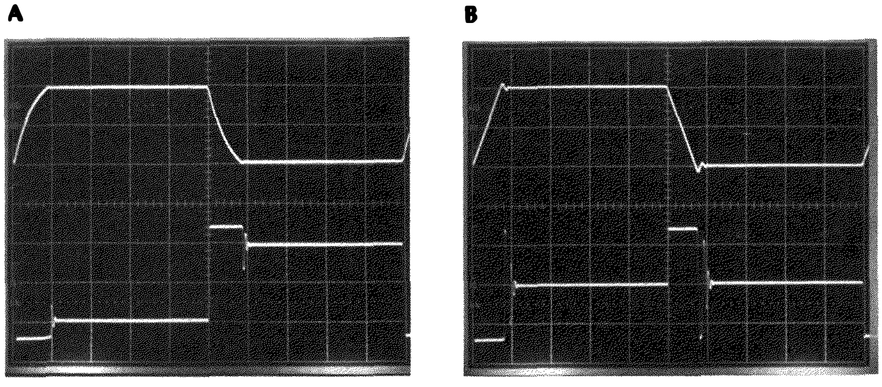
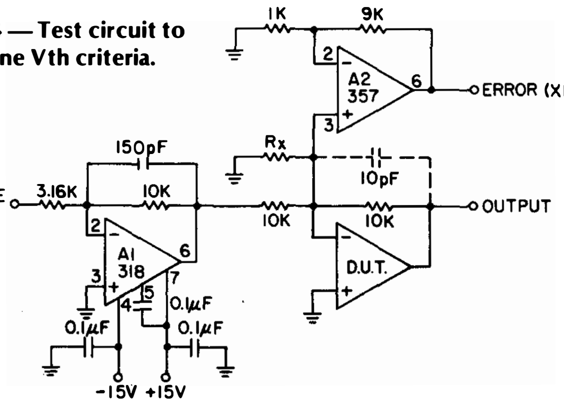


Fig. 43 — Transient performance of synthesized op- amp model with different open-loop gains to a 20-V p-p, 5-kHz square wave. Top traces are outputs, bottom traces error voltages. A, R4 = 10k; B, R4 = open. (Scales: 10 V/cm, 20 µS/cm.)

Fig. 44 — Test circuit to examine Vth criteria.



in amplitude, which can be expressed by modifying equation (1) by multiplying it by $\sqrt{2}/2$, yielding

$$SS_{(sine)} = \frac{\sqrt{2}\pi V p f c}{10^6}. \quad (54)$$

Equations (53) and (54) can be combined to show their ratios as

$$SS_{(sq)} = 2\sqrt{2}\, SS_{(sine)}. \quad (55)$$

Since this is nearly three times the signal slope of a sine wave at the frequency fc, it is clearly a more rigorous test. That it is the most rigorous test comes from the fact that the SS of the unfiltered square wave is infinite. It is clear then that an amplifier which passes a square-wave test without nonlinear distortion appearing in the output tends to be an optimum design. The question now arises, how can this be guaranteed?

We already know that to guarantee freedom from slew limiting we must, as a minimum, guarantee that the amplifier SR is in excess of the output SS for all possible signal conditions. For the non-slew-limited amplifier, this will encompass the signal slopes of square waves up to the rated output. We can set up a criterion to provide this with only a few parameters. Initially, let us consider a conventional feedback amplifier which follows the relationships discussed for Vth, SR, and GBP. By general feedback theory, we can express the bandwidth of this amplifier as

$$fc = GBP\ \beta \quad (56)$$

where fc is the small signal bandwidth, GBP is its gain-bandwidth product, and ß the feedback factor. For this initial part of the discussion we will assume no other filtering, and the amplifier alone determines the bandwidth, as just outlined.

To guarantee no slew limiting, we desire that SR ≥ SS. To provide this, we can write an inequality, substituting the appropriate equivalents for SR and SS, as they pertain to this amplifier. SR is as described by (44), and SS by (53). The inequality is

$$\frac{2\pi\, Vth\, GBP}{10^6} \geq \frac{2\pi\, Vpp\, fc}{10^6}. \quad (57)$$

With simplification, we can express this in terms of Vth as

$$Vth \geq \frac{Vpp\ fc}{GBP} \quad (58)$$

Equation (58) gives us an expression for a minimum Vth, but we can further simplify it by substituting (56), which yields

$$Vth \geq Vpp\beta. \quad (59)$$

The rather simple appearance of this expression may hide its rather profound implications. Since Vppß is in fact equal to the peak-to-peak input voltage, this relationship states that Vth should be in excess of the maximum pp input amplitude. In other words, the input stage (alone) will not overload when driven with a full-scale input signal [47, 67].

That the criterion works can be illustrated with some data just presented. In test 1, condition C it was observed that the experimental amplifier did not slew limit when subjected to a full-amplitude square-wave input. For condition C, Vth was 3V and the SR was 18 V/ μS. If a minimum Vth is calculated from (59) for this amplifier, it is found to be 2V. Therefore condition C satisfies (59), since 3V>2V.

On the other hand, if condition B is examined, it will be noted that Vth is only 0.33 V, and slew limiting *did* occur (Fig. 41B). Here the criterion was violated; i.e., 0.33V< 2V.

Another example, more in the line of a real amplifier, was the variable-feedback amplifier from Part I, discussed in Figs. 3 and 4. If Figs. 4a, 4b and 4c are re-examined, it will be noted that slew limiting is evident in condition A and some in B. Condition C is a non-slew-limited case.

Since the gains in this case were 20, 40 and 60 dB, respectively, ß is correspondingly 0.1, 0.01, and 0.001. As the output level is 20 V p-p in all cases, it can be noted that conditions A and B violate the minimum Vth criterion, which says that Vpp should be less than the 301A's Vth of 0.104 V. In condition C, the criterion is satisfied, and no slew limiting is evident.

It may already have occurred to some readers that this criterion is a most restrictive one, as it dictates *very low feedback factors* to eliminate slew limiting *in the case of low Vth amplifier stages.* Inasmuch as all directly coupled, undegenerated bipolar-transistor differential-amplifier pairs have a Vth of 0.052, this can quite logically explain TIM and slew limiting possibilities in power amplifiers, where Vpp may be upwards of 70 V.

It is interesting to plug typical power amplifier numbers into the relationship of (59) to see what results. A 100 W-into-8-ohm amplifier with a gain of 20 (26 dB) has a Vpp of 80 V and a ß of 0.05, which results in a required Vth of 4 V . . . clearly many times in excess of the 0.052 V resulting when an undegenerated bipolar differential pair is used in the input stage.

As a historical comment, the vacuum tube, still favored by many, has a Vth on the order of 3 V, for a typically used type

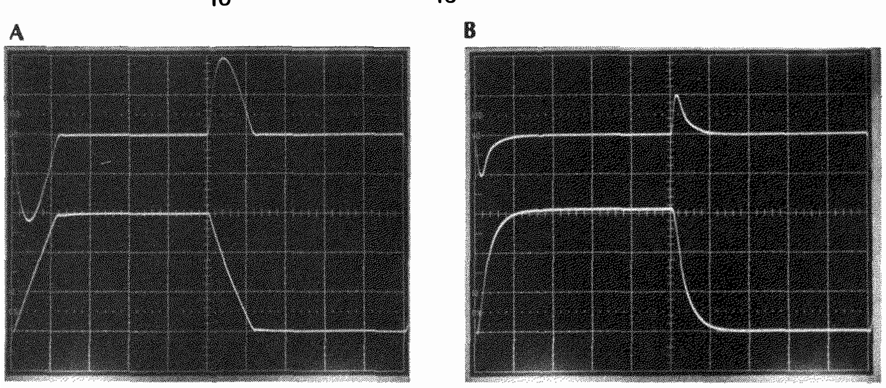


Fig. 45 — Transient response of a 301A, operated inverting with unity-gain compensation, Cc = 33pF, to 20-kHz square wave filtered at 100 kHz. Top traces are error voltages, bottom traces outputs. A, slew- limited response, and B, non-slew-limited response. (Scales: 5 μS/cm both; A, 0.5 V/cm top, 2 V/cm bottom; B, 0.1 V/cm top, 0.5 V/cm bottom.)

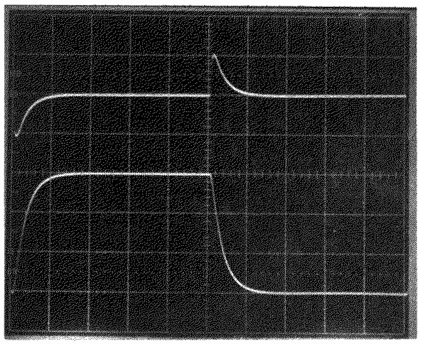


Fig. 46 — Transient response of a 301A, operated inverting and adjusted for slew suppression, Rx = 1.2 k, Cc = 5 pF. Top trace is error voltage, bottom trace is output. (Scales: 5 μ S/cm for both; 0.05 V/cm top, 2 V/cm bottom.)

such as the 12AU7. Viewed in this light, it is quite easy to see why a vacuum-tube design is much less susceptible to SID type problems; not only did they have less feedback in general, but they could also easily accommodate much larger inputs without first-stage clipping (47).

Viewed in just the above light, it is rather easy to conclude that the transistor audio power amplifier cannot be made to work. If, for example, we were to manipulate ß to satisfy (59) for the 100 W amplifier, using a Vth of 0.05, ß becomes 0.000625 (or less), which corresponds to a gain of more than 60 dB! While this probably is a completely impractical signal gain, it is possible to use special compensation "tricks" such as input compensation [25, 31, 63], which provide a low ß, but at elevated frequencies (above the audio range).

Of much greater interest are practical techniques which can be used to design an amp for no SID, *without* having heavy restrictions placed on the feedback loop. This can be done by separating the filtering and amplification functions, so that each can be optimized separately.

If an amplifier is preceded by a low-pass input filter with a cutoff frequency of fc, the filter-plus-amplifier combination can control the output signal slope with relative independence of the feedback factor. There are still restraints upon the Vth (or Vpp) of the amplifier, however, they are lessened to a great degree.

For this discussion it is assumed that the amplifier operates linearly and its own natural cutoff frequency, as determined by (56), is sufficiently higher than that of the input filter so as to cause negligible interaction. For such a linearly operated system, the peak-to-peak output of the input filter can be scaled by the gain of the amplifier, and the SS resulting at the output is of the same form as (53), but the relevant Vpp is the rated output of *the amplifier*.

If we now write an inequality such that the amplifier SR is to be maintained greater than the output SS, it follows the initial development form to (58), which is repeated here

$$Vth \geq \frac{Vpp\ fc}{GBP} \tag{58}$$

Written thus, it can be seen that as fc is lowered and GBP raised, the Vth required can be lowered. Within certain constraints, this allows considerably more design freedom. Like the previous relationship, this is one best understood by examining some performance which illustrates it functioning.

For an amplifier where Vth and GBP are fixed (as the 301A example of Part 1, Figs. 3 and 4), the only relief from the slewing problem is to decrease feedback in accordance with (59) until the criterion for Vth is satisfied. However, when we have control over GBP, we can manipulate things effectively to minimize slewing problems as we can by changing Vth.

A test circuit which can be used to demonstrate the relationship of (58) is shown in Fig. 44. Here A1 and the associated components form a 100-kHz single-pole filter, which drives the D.U.T., connected in an inverting circuit. This allows direct observation of the error voltage, thus this monitor shows directly when Vth is exceeded. The error voltage of the D.U.T. is buffered by A2, a high-speed FET amplifier, which furnishes a voltage gain of 10 to aid observation of low error voltages without loading the summing point. Rx is used to adjust the feedback of the D.U.T. test amplifier. A small (10 pF) feedback capacitor is used to minimize HF phase errors (which can obscure detection of slewing near threshold).

To check the validity of (58), a hypothetical amplifier stage was set up to pass a 6-V p-p output signal, after being filtered by the 100-kHz input filter. (Such a stage, for example, could represent the last stage of a preamplifier, and the numbers quoted are reasonable design figures.) A 301A compensated for unity gain with a resulting SR of 1 V/ μS and GBP of 1.5 MHz was used, with Rx open. The results for this device are shown in Fig. 45.

The bottom trace of this photo, 45A, is the output, which as can be noted is severely slew limited for the 6-V p-p level. The error voltage (top) is 1 V peak in level, well in excess of Vth, a confirmation that slewing is present in the output.

If (58) is an accurate predictor of slew suppression, it should be possible to adjust this stage to a point where slewing is not present.

If (58) is rewritten in terms of Vpp, as

$$Vpp \leq \frac{Vth\ GBP}{fc} \tag{60}$$

we should be able to calculate a Vpp below which this is true, for this circuit. Equation (60), with the substitution of the appropriate conditions, indicate that slewing should disappear below 1.5 V p-p, the level where Vth is 0.1 V.

A photo for these conditions (displayed similarly) is shown

A B

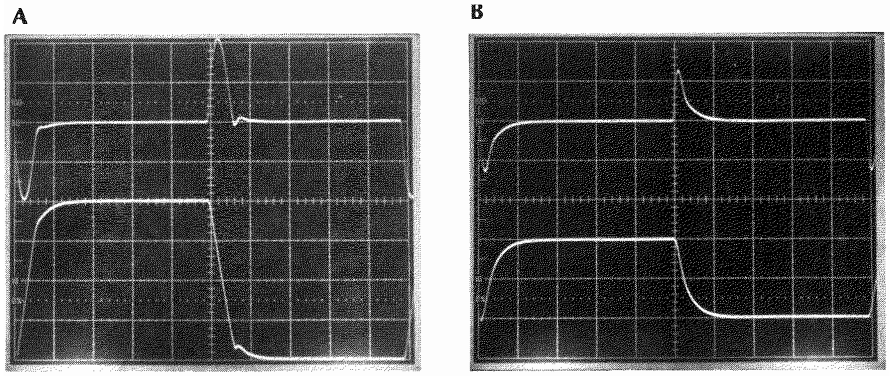


Fig. 47 — Transient response of TL070, operated inverting with unity-gain compensation, Cc = 33pF, GBP = 1.5 MHz, Vth = 0.67 V, to 20-kHz square wave filtered at 100 kHz. Top traces are error voltages, bottom traces outputs. A, slew-limited response; B, non- slew-limited response. (Scales: 5 μS/cm both; out puts both at 5 V/cm; error voltage, A, 1 V/cm, B, 0.5 V/cm.)

in 45B. As the output level and Vth indicate, slewing is just barely discernible in the output waveform (bottom). For levels below 1.5 V it will be absent; above 1.5 V it will appear with increasing degree, with increasing amplitude.

Equation 60 can also be used to adjust GBP to a point where higher output levels are possible without slewing. With the same 301A compensated with 5pF, its GBP became 10 MHz, which should allow the 6 V p-p output to be realized. For stability, Rx must become 1.2 K for this test.

The results, shown in Fig. 46, indicate that a 6-V output is realized without slewing. As can be noted, the error voltage is under 0.1 V (top) for this condition, indicating that operation is conservatively below the slew limit level. Equation 60 actually predicts a 10 V p-p output before slew limiting is reached.

Another demonstration of how the relationships of (58) and (60) operate is possible by using an amplifier with a radically different Vth to see if it predictably follows a similar pattern. This was done for a TL070 device, which for a similar compensation capacitance of 33 pF also has a 1.5 MHz GBP. However, due to its higher Vth of 0.67 V, the SR for this device and condition is 6.7 V/μS. As should be noted, these conditions produce a test amplifier with 6.7 times the Vth and SR over the 301A.

Figure 47A shows the output/error voltages for the TL070 compensated as noted for a 20-V p-p output. Slewing is evident in the output (bottom) and indicated by the 2-V peak error voltage (top) which is in excess of Vth. Equation (60) predicts that slew limiting should disappear below a 10-V p-p output, which is shown in 47B. Note that the error voltage is just over 0.6-V peak, and slewing is just barely noticeable in the output (bottom).

If this amplifier is adjusted for a higher GBP, as was done in the 301A case in Fig. 46, it shows a similar improvement. For this 10-MHz GBP condition, the output predicted by (60) would be 67 volts p-p or in excess of the supplies. The results at a 20-V level are shown in Fig. 48, and there is no slewing detectable at all.

It should be noted that these two examples do indeed demonstrate similar adherence to the relationship described. If the results are compared for conditions where the error voltage is at the Vth level, for example Figs. 45B and 47B, it can be noted that although the two output levels produced are different (due to different SR and Vth), *the error voltages are of a similar percentage of the output* or about 6.7 percent. This demonstrates that it is, indeed, possible to satisfy a common criteria (SR>SS) by different means, with similar errors by the different routes taken.

Another way of stating this is to rephrase an earlier statement, that Vth *in itself* is not a single totally important parameter—it is important to this subject to the extent it affects SR and input overload. The relationships set down in (58) and (60) are somewhat deceptive in this regard, as they do not contain an SR term. However, it should be remembered that these two relationships are *fundamentally based on an SR criterion* and, as such, contain terms which are useful towards manipulating or maximizing SR. In a very broad perspective, it should also be understood that it is incomplete to imply that input dynamic range, Vth, or other similar conceptual terms describe the entire situation in terms of a no-slew-limit guarantee, for they do not. As the experiments just described have demonstrated, even a low Vth amplifier can be effectively used. If its operating conditions are set up to provide an SR>SS, the obvious slewing distortion can be suppressed.

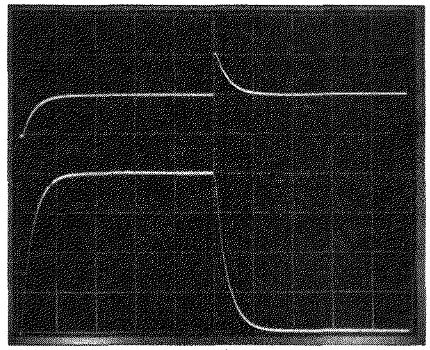


Fig. 48 — Transient response of a TL070, operated inverting and adjusted for slew suppression, Cc = 5 pF, Rx = 1.2 k. Top trace is error voltage at 0.1 V/cm, bottom trace is output voltage at 5 V/cm.

There is a great deal more which can be said about specific amplifier operating conditions and methods of suppressing SID by guaranteeing SR>SS. Unfortunately, however, the scope of all of these factors might be a complete article or series in itself. Therefore, we will limit comments on these points to the highlights at this time.

What the relationships just discussed show is that when the output of an amplifier stage is, by design, purposely confined to signal slopes less than the SR of that amplifier, the amp will not slew limit. Further, if SR is maintained greater than SS for all output levels up to (or above) the clipping level, the amplifier will not slew limit for any input level below clipping.

While this was demonstrated with a model consisting of a separate input filter followed by the amplifier under test, it also holds true when the filter is integral to the amplifier, i.e. the amplifier is an active LP filter. An amplifier can, in fact, be designed in this manner for slew suppression, as described by Leach [10]. However, the conversion of an amplifier to an integrator at high frequencies will usually result in more compensation being necessary for stability, hence there can often be little net improvement for this approach.

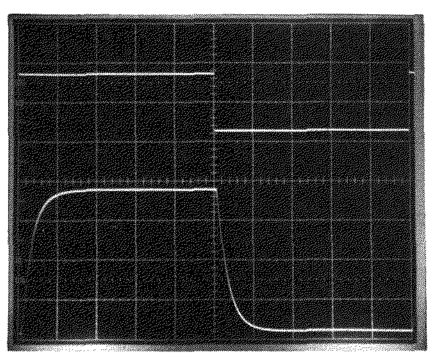


Fig. 49 — Transient response of a non-slew-limited amplifier design, loaded with 8 ohms, to a 10-kHz square wave. Top trace is input at 2 V/cm, bottom trace is output at 20 V/cm.

In practice, effective control and design freedom are also realized when the slope limiting filter is placed before the amplifier. This allows reduced compensation and a high SR in the amplifier, with complete control of maximum signal slope by means as simple as a single RC input section.

An example of a power amplifier design based on these principles is described in reference [71], and it is worth noting that a commercial design [72] following these principles has received some good marks from audiophiles and subjective reviewers. To illustrate the point that this amplifier is indeed a non-slew-limited design, a full-level output (80 W) square wave from it is shown in Fig. 49, along with the input square wave. It is clear that the response is small-signal-bandwidth limited only, and the 6 μS risetime does not, in fact, vary as a function of level.

The design techniques and experimental data described above for reduction of SS by prefiltering at the amplifier input have all been based upon single-pole, low-pass filters. While this type of filter has been shown to be quite effective for control of SS, and thus prevention of slew limiting, more sophisticated filter techniques are even more effective in reducing SS.

It has been shown [12, 66, 67, 70] that higher order filters are even more effective for reduction of SS, compared to a simple first-order type, for a given cutoff frequency. There are, of course, trade-offs to be made in comparing one to the other, considering the higher performance against the increased complexity. Also, the damping of the filter must be considered, as well as its frequency. However, the increased complexity of a second-order filter really depends on exactly how it is realized and may not in fact be prohibitive. For example, Leach has shown in [66] how the amplifier itself can be used as the active portion of the prefilter, without undue stability constraints, in what appears to be a practical and attractive topology. Further, in [70] it is shown that a second-order Bessel LP filter alignment will produce approximately 1/2 the SS of a single-pole filter for otherwise similar conditions. Unfortunately, time did not permit detailed experimentation with these techniques for this article, but they appear to have significant merit towards the reduction of SID effects.

Generally, the above discussion describes two alternate means which can be used to design a non-slew-limited amplifier and thus prevent SID and TIM. A logical question which may be raised is, do they yield equal results in auditioning? While we do not at this point have subjective response data to definitively answer this question, informal listening tests by one author (W.J.) tend to favor circuit topologies which are designed from a standpoint of equation (59), using linearized input stages, such as FET or degenerated bipolar devices. As time progresses, it is hoped that further listening tests will more clearly define the optimum choice between the two approaches.

Conclusions

In this article we have attempted to cover a quite broad topic from a multiplicity of viewpoints, in both discussion and analysis. These different techniques of analysis all indicate a common pattern of distortion in feedback audio amplifiers, which is a function of the ratio of signal slope to amplifier slew rate, a dimensionless parameter we define as SR ratio. When the SR ratio is less than unity, this distortion is suppressed; when greater than unity, strong nonlinear distortion products appear, which are subjectively objectionable.

Control of this distortion, which we call SID, can be achieved by maintaining linear amplifier behavior, with an SR greater than the highest SS, or stated in terms of SR ratio, an SR ratio less than unity. Since SS is both frequency and amplitude dependent, it follows that greater SR in an audio amplifier is required for higher voltage output stages, where the SS is highest.

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Control of this distortion can be exercised by appropriate selection of amplifier type, by specifying an SR sufficient to the application. In design, it can be achieved by providing a sufficiently conservative SR (on the order of 0.5 V to 1V/ μS per peak output volt) or by designing for a non-slew-limited response. A non-slew-limited amplifier has an inherent SR greater than its maximum possible output SS and will therefore never slew for any input signal, including square waves, or its SR ratio is guaranteed <1. It is characterized by frequency response which is small-signal-bandwidth limited, for any output below its clipping level. As such it has no major nonlinear distortion products due to slewing effects. Such an amplifier is also said to be TIM free and may be described in this context as well.

It is recognized that there is considerable controversy on the relative importance of TIM and SID, their audibility, and some of the relevant design criteria. For this discussion, it is not our purpose to dwell excessively on the relative importance of SID, its audibility or other factors which are often subject to opinionated views. What we seek to do is describe means to quantify and control this distortion mechanism, and basically this is the only main point being addressed.

The existence of the distortion mechanism is, of course, *not* a subject of debate, and like other distortions in amplifiers, knowledge of its behavior patterns is valuable to either the circuit designer or the informed user of audio equipment. We would, however, like to express caution with respect to certain alarmist commentary, for example those to the effect that low TIM or minimal SID is the magic elixir of quality audio. While this distortion source is quite important, so are many others. Once sufficient linearity and slew rate have been provided in a design, there may actually be little gained by boosting SR further (to far beyond that necessary). The *optimu* audio amplifier is best designed with *all* contributions to audible defects given proper perspective.

We appear to currently be immersed in a specifications race on the part of some manufacturers in this regard, which is not only unfortunate for the confusion it spreads (as to what is most important), but doubly so from the standpoint that if nonlinear techniques are being used to achieve high SR numbers, the user can actually pay a penalty in *higher* distortion!

Another specifications race practice appears to be the quotation of amplifier maximum output SS *for small signal condition* as its specified SR. If an amplifier is operating linearly in non-slew-limited conditions, the output SS for a fixed signal will linearly follow the output level, and at no point will it reach the true amplifier SR, which is, in fact, a limit. It is therefore erroneous or misleading to quote a maximum SS as an SR in such a case, as the *true* SR limit is never reached. In our opinion, while such an amplifier has real merit, it might more clearly and suitably be described in such terms as "maximum linearly reproduced SS" or the qualifier added that it is a true non-slew-limited design, as described in the text. Using the terminology of SR implies that the amplifier *can* be made to slew; if, in fact, it *cannot* be made to slew this should be clearly stated, for it is a point which distinguishes the design.

(In Part II on page 44 in July under "Comparison of Test Methods," we made the statement that the squarewave's *fundamental* amplitude was 12 dB larger than the sine wave's. The square wave *itself* is 12 dB larger in amplitude, as described in the sidebar.)

We hope this discussion has served to bring together some of the various issues involved so as to create a new perspective for the reader. We recognize that some of the points made in this article have been made elsewhere and acknowledge the work of previous authors. We believe that the extensive bibliography will be helpful to the reader to appreciate this material, and to tie older data in with the new material presented within this article.

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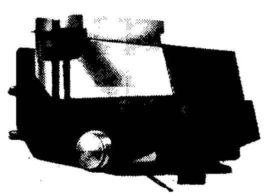

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