60W NDFL Amp

To accompany the article on nested differentiating feedback loops, here is a practical amplifier design, presented as a module, with very low distortion. Design by Edward M. Cherry.

THIS AMPLIFIER will perhaps be of most interest to home constructors who want to rebuild an existing system and upgrade its performance without the expense of new major components. The power output transistors employed are the well-known types MJ802 and MJ4502 which have been around for several years and have proved their reliability. Indeed, the whole design is mature and home constructors should have no difficulty in making it work.

Grounding

In any amplifier where the basic distortion has been reduced to a few parts per million, several distortion mechanisms not ordinarily considered may become significant. One such mechanism is associated with currents circulating in the ground leads and power-supply wiring.

Figure 1 explains the origin of this distortion. The current in each power transistor of a class B stage is a half-wave rectified version of the output. The two currents, drawn alternatively from the positive and negative supplies, are equivalent to a circulating full-wave rectified current and this is basically an even-harmonic distortion of the signal output. If there is any mutual inductance between the power-supply wiring (including the
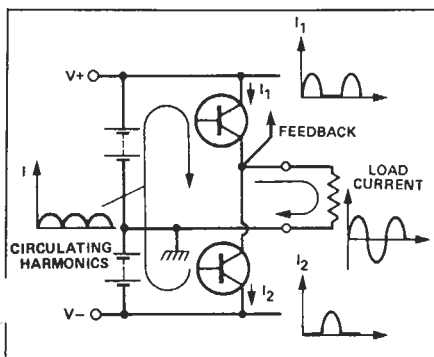

Figure 1 Circulating even-harmonic current in a Class-B output stage.

grounds) and the signal wiring (also including the grounds), then an even-harmonic distortion is induced in the amplifier and feedback is powerless to correct it.

The circuit board has been laid out so as to minimise this effect. The areas enclosed by some tracks are critical, and home constructors making their own PCBs are cautioned to follow the layout exactly, using the accompanying foil pattern.

Note that the circuit uses three distinct ground symbols.

a) is the *quiet ground* track on the circuit board (one per channel).

b) is the *noisy ground* track on the circuit board (one per channel).

c) is the metal chassis ground (there are six connections to the chassis in total).

Each channel is connected to chassis ground at two points. The input socket is connected to the chassis (rather than insulated from it), the input lead from socket to circuit board is shielded, and the quiet ground track is connected to chassis ground at the input socket via the screen. Similarly, the ground output terminal is screwed into the chassis, the leads from the circuit board to the output terminals are a twisted pair and the noisy ground track is connected to chassis ground at the output terminals via the ground output lead. The remaining two connections to chassis are in the power supply (Fig. 5).

Note that a 10 ohm resistor, R31, links the quiet and noisy ground tracks. This resistor is short circuited at low frequencies by the input shielding and neutral output wiring to chassis ground. However, the resistor takes over at high frequencies where wiring inductance become significant.

The 15uH filter inductors in the supply rails are also for suppressing circulating currents (R6 and R7 represent the winding resistances of L1 and L2).

This amplifier employs only two nested differentiating feedback loops and its distortion is not down to the ultimate limit. The benefit of including the filter inductors is therefore marginal. The author is not blessed with 'golden ears' and cannot hear the effect of removing the filters, although the difference is clear-

ly measurable. The filters should certainly be included in amplifiers that use three or more NDFLs. As the inductors must be home-made, and therefore cost nothing but time, and as they do make a measurable (if small) improvement, most home constructors will probably wish to include them. Winding data is given in Table 1.

The precise values of inductance and resistance are not important — ±50% is good enough — but do not use the 1.25 mm wire from L3 as something like 0.1 ohm series resistance is essential. For a similar reason, do not parallel the 470uF bypass capacitors C9 and C10 with high-frequency types. Brass or steel mounting screws are perfectly satisfactory for the filter inductors, as linearity is not important.

Critical Components

The majority of the components in this amplifier are not critical. Almost any small-signal diodes will do, such as the 1N914 and 1N4148. Q1 and Q2 should be high-gain, low-noise types — BC109 and BC549 are among the cheapest available. The others could be almost any small signal types: BC107 and BC547 are readily available NPN types, the BC177 and BC557 are suitable PNPs. The driver and output transistors should be the types shown: TIP29C and TIP30C for the drivers, MJ802 and MJ4502 for the power transistors. The biasing transistor, Q11, could be any NPN in a TO-126 pack that can be mounted on the heatsink: the TIP29C is a readily available type that would suit.

Unless the contrary is indicated on the Parts List, resistors can be standard ½ W types and the capacitors can be the lowest available working voltage. A few components, however, do require special mention. A feedback amplifier cannot be more linear than its feedback network, so the various components that constitute the feedback network should have small voltage coefficients.
Specifically:
a) The overall feedback resisitors R11 and R12 should be high-stability types, such as metal oxide or metal film;
b) C4,C6 and C8 should be NPO ceramics, not high-K types (NPO means negative-positive zero, a low-K capacitor with a very low temperature coefficient; metallised plate ceramics, for example. Silvered mica capacitors are also suitable);
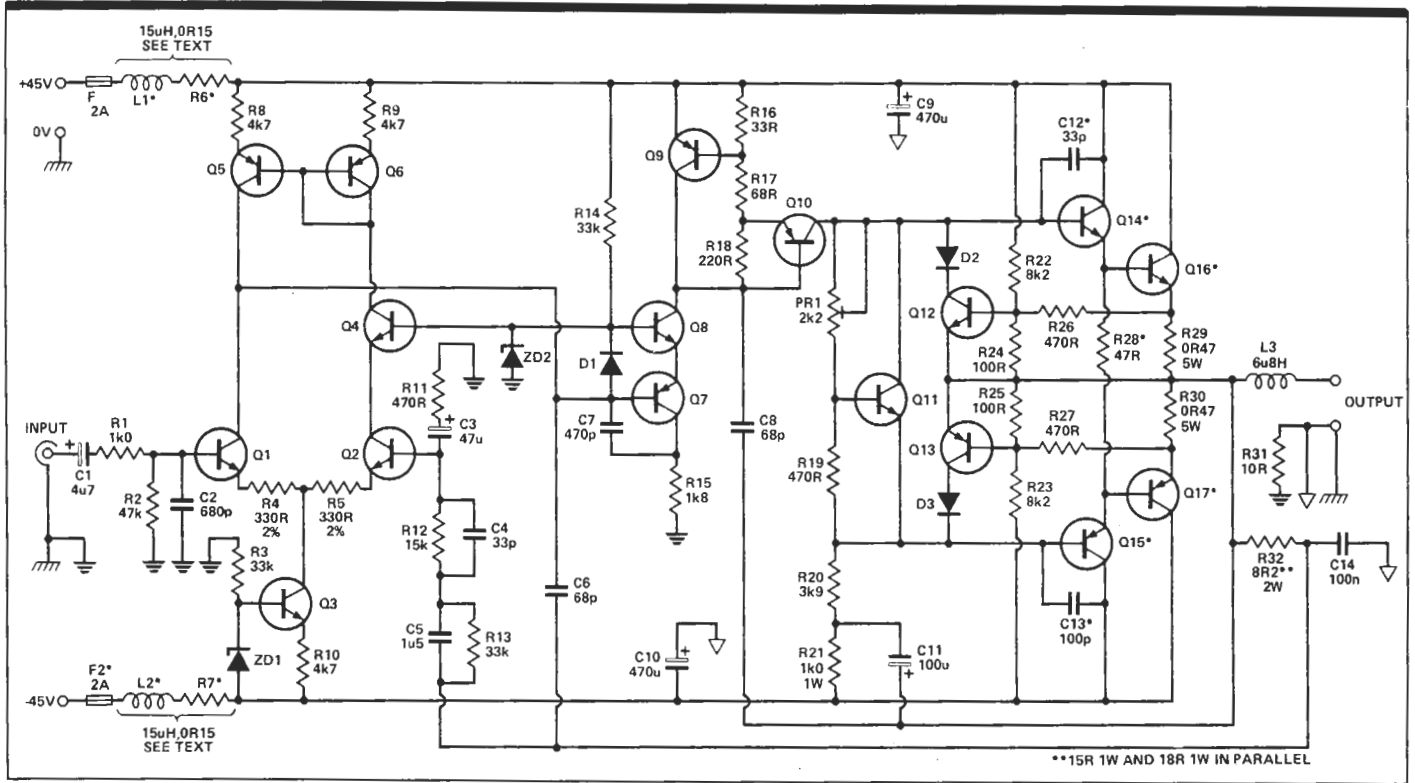c) C5 and C14 should be polycarbonate, polystyrene or polypropylene types, but not polyester (eg. mylar types);

Figure 2 Circuit diagram of the 60 W power amp. Components marked with a single asterisk are not mounted on the PCB.

d) C3 should be an ordinary cheap aluminum electrolytic, definitely not one of the relatively expensive resin-dipped tantalum types (this is not a misprint!)

The 6u8 H inductor (L3) needs to be home-made. Winding data is given in Table 1. The bobbin should be mounted on the circuit board with a nylon screw; brass or steel must not be used, because of non-linear eddy current losses.
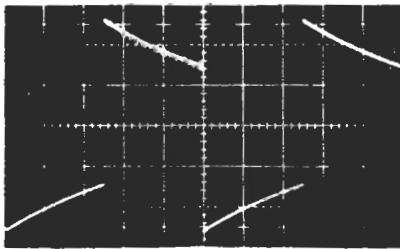

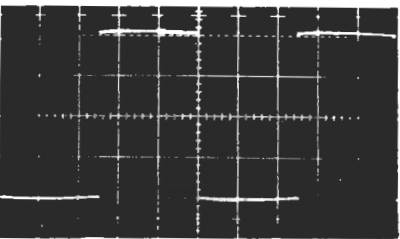**Figure 3a** Square wave response of the amp without group-delay compensation.

Figure 3b Square wave response of the amp with group-delay compensation — note the improvement over Fig. 3a.

Construction

Assembly of the PCB is quite straightforward. It is probably best to commence by soldering all the resistors in place. Note that R32 could be either a 2 W type (not common) or two 1 W resistors (15R and 18R) in parallel. Note that the emitter ballast resistors of Q16 and Q17 (R29 and R30) should have very low inductance and if you have trouble with high frequency instability, these resistors are likely to be the culprit. The best solution may be several carbon resistors in parallel. Mount R29 and R30 a few millimeters above the board.

Assemble the diodes next, making sure you get them all the right way round. Install the links next. Follow with the capacitors. Note that C5 and C14 must be polycarbonate types and C4,6 and 8 must be NPO ceramics. None of the other ceramic capacitors should be hi-K types, as mentioned earlier. When mounting C9 and C11, see that there is three or four millimetres between the capacitor body and the adjacent 5 W resistors (R29 and R30)

The transistors may be mounted now. See that each is oriented correctly. Wind L3 next and mount it on the board. Details are given in Table 1. It is not necessary to strictly follow the former dimensions given, but the inductance needs to be close to 6u8 H and wound from 1.25 mm wire at least, for low resistance.

Assembly of the components mounted to the heatsink comes next. The heat-
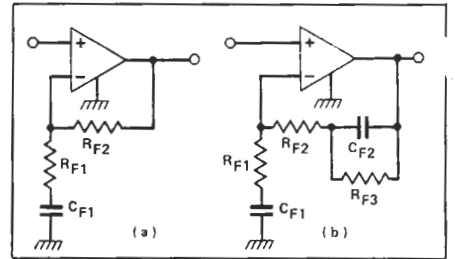

Figure 4 Circuit for compensating low frequency group delay: (a) basic uncompensated circuit; (b) compensated circuit.
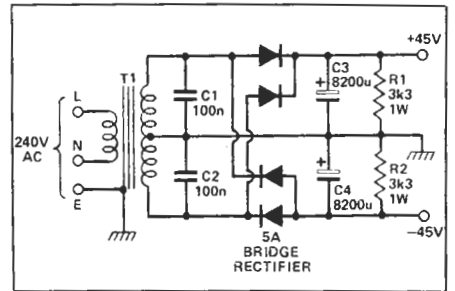
Figure 5 Suggested PSU for the amplifier.

sinks in the original were a standard type sold by many companies. Each heatsink has a thermal resistance to ambient of about 1°C/W, and other types could, of course, be substituted. The specified thermal resistance permits continuous operation at full power: smaller heatsinks (up to 2°C/W) could be substituted if the amplifier is to be used only for domestic sound reproduction. Use one heatsink per channel.

Figure 6 Showing the general technique for connecting inputs, outputs and grounds to a stereo pair of modules.
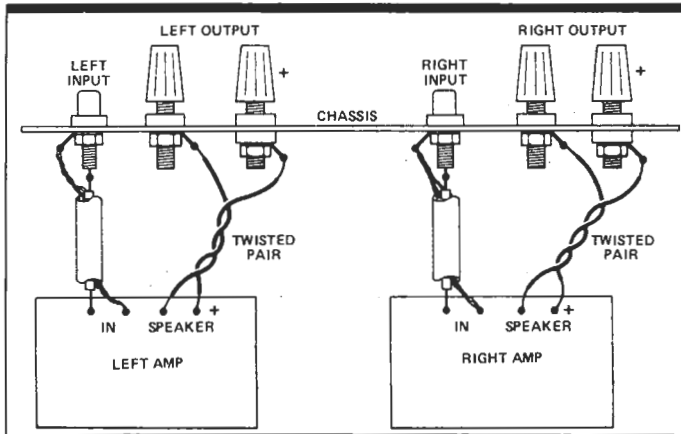
Harmonic	Rated output 21V9 60 W	-20 dB 2V19 600 mW
2nd	115 ppm	40 ppm
3rd	100	25
4th	32	15
5th	40	9

Harmonics higher than the 3rd are ultrasonic and hence inaudible.

TABLE 1

Formers
If a suitable type is not at hand, these may be turned from 25 mm diameter polystyrene rod to give 12 mm internal bobbin diameter with 7.5 mm winding space between cheeks.
Wire & Winding L1,2
Take two 1680 mm lengths of 0.75 mm diameter enamelled copper wire and wind onto each former leaving 20 mm or so lead length at start and finish.
Wire & winding L3
Take a 1190 mm length of 1.25 mm diameter enamelled copper wire and wind it onto the former. Leave 20 mm or so lead length at start and finish.

Three small components are mounted on the heatsink adjacent to the transistors to keep certain leads short: R28, C12 and C13. Construction is very much simplified if a 4-way terminal strip is installed under one of the collector mounting bolts of Q16 and a 5-way strip under one of Q17's mounting bolts. Figure 8 shows details.

The collector and emitter leads from each power transistor to the circuit board should be twisted. The base leads to Q14 and Q15 could be twisted in with the corresponding collector and emitter leads (although this is not necessary) and the base lead of Q11 can be kept separate. Note that all transistors must be insulated from the heatsink. Note also that the TIP30C specified for Q10 needs its leads dressed to fit the board.

Quiescent current in the power transistors should be set to 40-60 mA by PR1. *Be warned* that this quiescent current is almost zero until PR1 is about three-quarters of its maximum resistance, after which the current increases very rapidly; be sure that PR1 is set to *minimum resistance* when the amplifier is turned on or the first time.

A convenient way to check the quiescent current is by means of the voltage drop across R29 and R30; this should be 40-60 mV (total) for zero signal input to the amplifier.
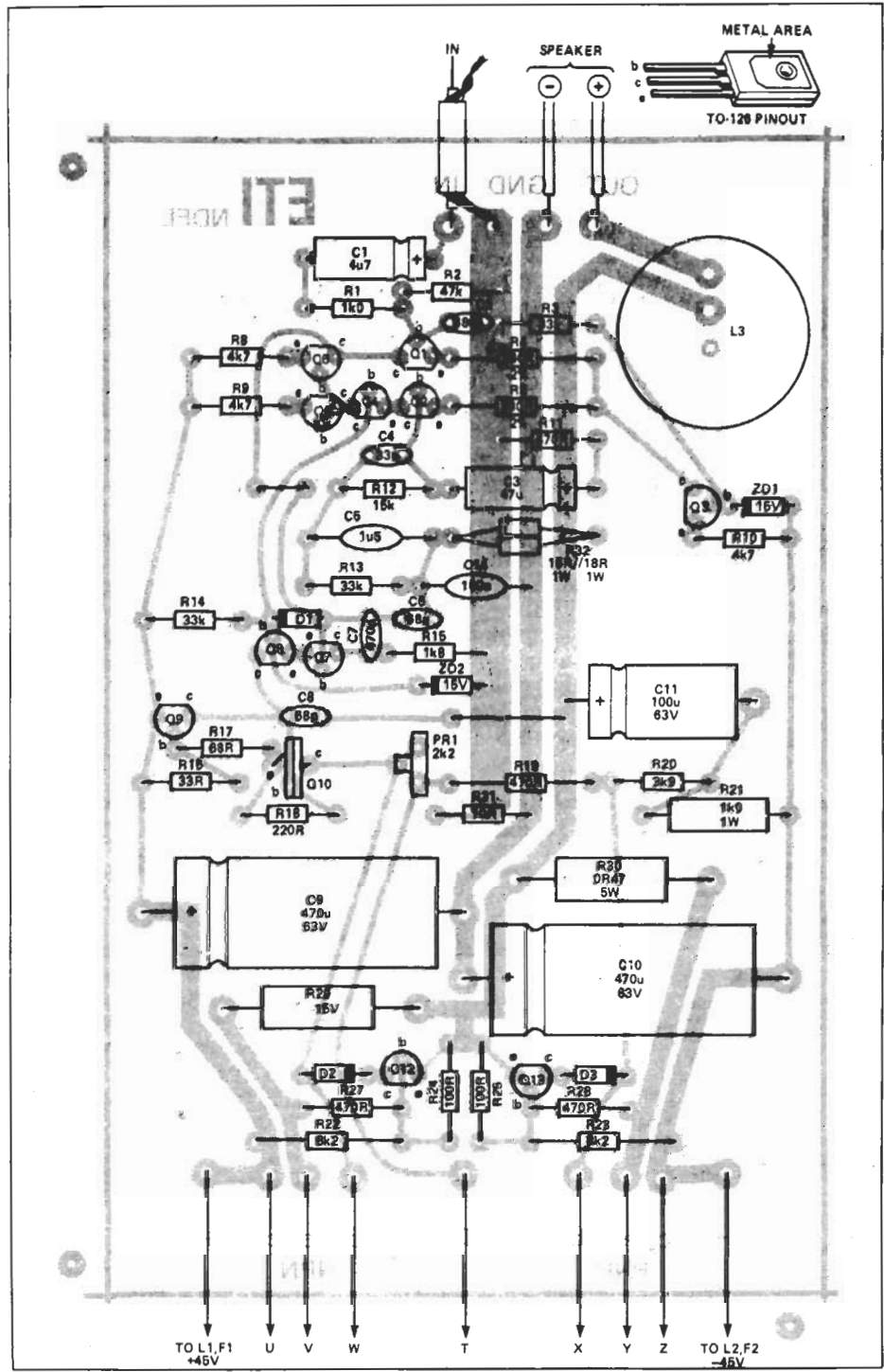

Figure 7 Component overlay for the power amplifier.

Resistors

(all ½W, 5% except where stated)

R1	1k0
R2	47k
R3,13,14	33k
R4,5	330R 2%
R6,7	see text
R8-10	4k7
R11	470R metal oxide or metal film
R12	15k metal oxide or metal film
R15	1k8
R16	33R
R17	68R
R18	220R
R19,26,27	470R
R20	3k9
R21	1k0, 1W
R22,23	8k2
R24,25	100R
R28	47R
R29,30	0R47, 5 W
R31	10R
R32	8R2, 2 W or 15R//18R, each 1 W

Potentiometer

PR1	2k2 miniature vertical preset

Capacitors

C1	4u7 axial electrolytic
C2	680pF ceramic
C3	47uF axial electrolytic
C4	33pF 100 V NPO ceramic
C5	1u5 polycarbonate
C6,8	68pF 100 V NPO ceramic
C7	470pF ceramic
C9,10	470uF 63 V axial electrolytic
C11	100uF 63 V axial electrolytic
C12,13	33pF 100 V ceramic
C14	100nF 100 V polycarbonate

Inductors

L1,2	15uH (see text and Table 1)
L3	6u8 H (see TAble 1)

Semiconductors

Q1,2	BC109, BC549 etc.
Q3,4,8,12	BC107, BC547 etc.
Q5-7,9,13	BC177, BC557 etc.
Q11,14	TIP29C
Q10,15	TIP30C
Q16	MJ802
Q17	MJ4502
D1-3	1N4148, 1N914, etc.
ZD1,2	15 V 400 mW zener

Miscellaneous

F1,2	2 A standard fuse

PCB; one 4-way and one 5-way terminal strip; heatsink to suit (see text); PCB stakes; bobbins for inductors; wire, etc.

HOW IT WORKS

Figure 2 is the complete circuit of one channel of the amplifier; equations referred to in the explanation refer to last month's feature. The circuit is clearly based on Fig. 10 (in the theory article), with major parameters

$$1/B = 32.9$$
$$r_x = 800 \text{ nS}$$

The value of B is set by the overall feedback resistors R11 and R12 (470R and 15k — see Equation 1). r_x is set by:

a) R4 and R5 (33R) plus C6 and C8 (68p) in conjunction with the chosen value of B (see Equation 13);
b) R15 and C7 (1k8 and 470p — see Equation 14);
c) R32 and C14 (8R2 nd 100n) plus the 8 ohm nominal load and L3 (6u8 H);
d) R12 and C4 (15k and 33p) via the other constants in Equation 15.

The first stage requires little comment. Q1 and Q2 operate at 1.5 mA each, Q3 is a current source, Q4 is a common-base stage to equalise the quiescent voltages on Q1 and Q2; Q5 and Q6 constitute a current mirror. R1 and C2 form a 200 kHz low-pass filter against RF interference.

The current amplifier operates at 3 mA, set by R18, and it incorporates a catching diode (D1) to accelerate recovery from overdrive. The pre-driver, Q10, operates at 8 mA; Q9 protects the stage against damagingly large currents under fault conditions. Driver quiescent current is 25 mA, set by R28.

Transistors Q12 and Q13 provide short-term protection for the power transistors. Short-circuit current is limited to about 4 A, and peak signal current is limited to 7 A. Long-term protection is provided by 2 A fuses in each supply rail; these should be 'ordinary' types, rather than delay or quick-blow. In the unlikely event of transistor failure, these fuses limit the loudspeaker current to 2 A, corresponding to 32 W into 8 ohms.

The common alternative of a single fuse in the loudspeaker lead is less satisfactory: it provides less protection for the amplifier; it provides less protection for the loudspeaker as the fuse must be rated to carry the full signal current, and it introduces distortion on large-amplitude, low-frequency signals.

Low Frequency Compensation

A feature of Fig. 2 not discussed so far is a low-frequency compensating circuit, R13 and C5.

Amplifiers of the basic circuit topology of Fig. 2 (theory article) have a group delay which is different for different signal frequencies. Some frequencies take longer or shorter times than others to pass through the amplifier. High-frequency group delay in NDFL amplifiers can be corrected, as described last month, by a small capacitor in the feedback network (See Equation 15). Errors in low-frequency group delay, in both Figures 2 and 10 (theory article) are associated with the input coupling capacitor and the capacitor in series with R_{F1}. Low-frequency square-wave inputs are reproduced with a 'tilt' as in Fig. 3a.

One approach to this problem is to use a truly direct-coupled amplifier, with no capacitors in series with the signal path; commercial audio power amplifiers of this type appeared in the 1970s. Unfortunately, such amplifiers are prone to drift. A significant DC voltage may appear at the output even when there is no input. Although it is possible to reduce drift in a power amplifier to an acceptable level, it is not possible with today's technology to build a system that is truly direct-coupled from pick-up input, through the RIAA network and the power amplifier.

In the last few years a generation of amplifiers has appeared which include some form of servo amplifier to correct the drift. All circuits known to the author re-introduce the problem of group delay, albeit in a lesser form.

The approach adopted in the design is to retain the coupling capacitors and thereby eliminate drift, but include a group-delay correcting circuit. Figure 4 shows the outline. Group delay is optimally compensated if:

$$R_{F3} = 2R_{RF2}, \quad (16)$$
$$R_{F2}C_{F2} = R_{F1}C_{F1} \quad (17)$$

Figure 3b shows the improvement in square-wave response.

Low-frequency group-delay compensation could well be included in audio power amplifiers and pre-amplifiers other than NDFL types.
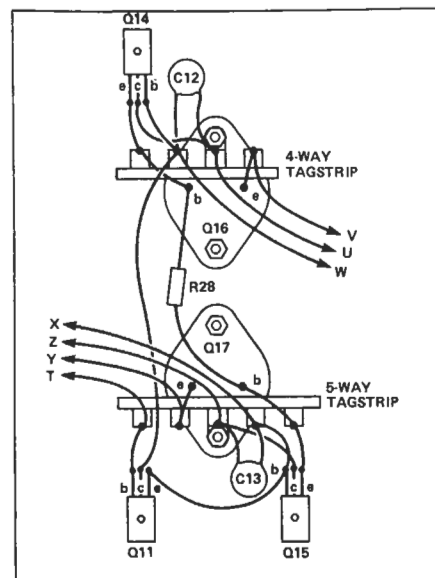

Figure 8 Wiring diagram for the components mounted on the heatsink.

HARMONIC ANALYSIS AT 1kHz

Harmonic	Rated output 21V9 60 W	-20 dB 2V19 600 mW
2nd	19 ppm	5 ppm
3rd	14	3.5
4th	2.5	2.5
5th	3.0	1.5
6th	1	1
7th	1.8	1.8
8th	1	1
9th	1.0	1
10th	1.8	1

Notice how the harmonics drop away at small signal amplitude. In this regard a class-B NDFL amplifier is more like a conventional class-A amplifier than a class-B amplifier.

1 ppm = 0.0001%

OUT   IN

ETI NDFL

− c e b PNP T NPN b e c +
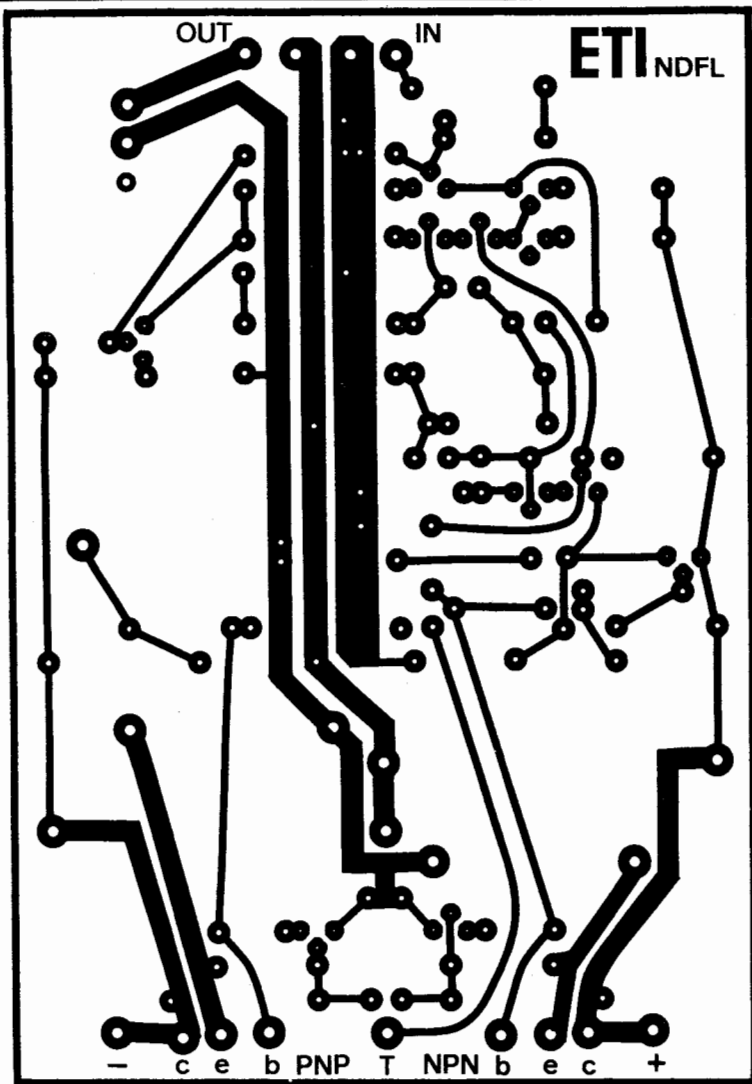
Designing NDFL Amps

The use of nested differentiating feedback loops (NDFLs) is a new technique for reducing audible-frequency distortion in an amplifier to a vanishingly low level. As the name implies, NDFLs rely on negative feedback, but they use it in a new way. Edward M. Cherry explains the theory involved.

IN ORDER TO understand just how far the new NDFL technique can improve an amplifier, we first need to know the fundamental limits to the reduction of distortion that can be achieved with conventional techniques. To begin with, we survey familiar negative-feedback theory.

Figure 1 is a block diagram of an amplifier with negative feedback. In this diagram, the forward path corresponds to the amplifier before feedback is applied, and its gain is traditionally designated by the Greek letter u. The feedback network returns a fraction B of the output to the input circuit, where it is in some way subtracted from the true input to provide the actual input to the forward path.

In many practical amplifiers, the subtraction is accomplished by applying the input and feedback signals to the two inputs of a balanced differential first stage of the forward path. Figure 2 is an outline practical circuit. In this circuit the feedback factor B is the attenuation of the network comprising R_{F1} and R_{F2}

$$\beta = \frac{R_{F1}}{R_{F1} + R_{F2}}$$

A typical value for an audio power amplifier might be 1/20. The forward-path gain u in Fig. 2 corresponds to gain from input to output when the audible network is removed. A typical value for a simple audio power amplifier might be 1000.

For Fig. 1, the overall closed-loop gain A is given precisely by

$$A = \frac{\text{Output}}{\text{Input}} = \frac{\mu}{1 + \mu\beta}$$

he quantity uB is called the loop gain. Physically, loop gain is the gain that would be observed if the feedback 'loop' in Fig. 1 was cut at some point, a signal was injected into one side of the cut, and

the resulting signal at the other side of the cut was measured.

If the values of u and B are such that loop gain is small compared with unity, the closed-loop gain is very nearly equal to the forward path gain (that is, the gain without feedback)

$$A \to \mu$$
$$\mu\beta \ll 1$$

However, if loop gain is large compared with unity, the closed-loop gain approaches the reciprocal of the feedback factor and becomes almost independent of the forward-path gain

$$A \to 1/\beta$$
$$\mu\beta \gg 1$$

The quantity $1/B$ is often called the demanded gain, as it is the value the overall closed-loop gain would take in ideal circumstances.

As a numerical example, if we substitute the above values $u = 1000$ and $B = 1/20$ into Equation 2, the gain of our 'typical' audio power amplifier works out as $A = 19.6$. The approximate Equation 4 predicts A   20, within 2% of the correct answer.
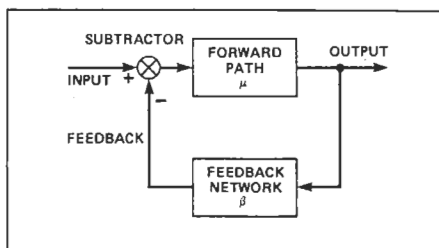
Fig. 1 Block diagram of a feedback amplifier.

The quantity $1 + uB$ occurs often in feedback theory. It is called the return difference F.

$$F = 1 + \mu\beta$$

Physically, return difference has the significance

$$F = \frac{\text{forward-path gain}}{\text{closed-loop gain}}$$

For values of loop gain greater than about 10, loop gain and return difference are almost equal — in our 'typical' example the values are 50 and 51 respectively.

Simplified treatments of feedback theory show that, if the distortion

generated in the forward path (that is, the amplifier without feedback) at a particular output signal amplitude is D_u, then the resulting closed-loop distortion D_A at the same output signal amplitude is

$$D_A = D_\mu/F$$

Distortion is improved when feedback is applied to an amplifier by a factor equal to the return difference. In our 'typical' amplifier, $F = 51$; if the distortion without feedback happened to be 10%, then feedback should reduce the distortion to 0.196%.

More rigorous treatments of feedback theory show that Equation 7 is no more than a poor approximation to the truth. In the first place, real amplifiers are far more complicated than Fig. 1 suggests, because several different feedback paths (not all intentional!) can be identified. For example, the collector-base capacitances of transistors inevitably provide some unintended feedback at high frequencies. There is a very real problem in interpreting just what loop gain and return difference mean when there is more than one feedback loop. Once the correct interpretation is established, return difference invariably turns out to be a function of frequency, and the reduction of distortion corresponding to Equation 7 depends on the value of return difference at the frequency of the distortion, not the frequency of the input. Feedback therefore, does not reduce all distortion components equally.

Finally, it is found that the closed-loop distortion of an amplifier can contain new components that were not present in the distortion that existed in the forward path before feedback was applied. These new distortion components initially increase as loop gain is increased, but they fall away again towards zero as loop gain is made large.

Despite all these complications, the fact remains that adequate negative feedback, properly applied, does reduce distortion. Why, then, do amplifier designers not simply apply some arbitrarily large amount of feedback and reduce amplifier distortion to the vanishing point?

TIM, IIM, PIM, . . .

In the last 10 years or so, readers of audio magazines have been made aware of a conjecture that goes something like this:
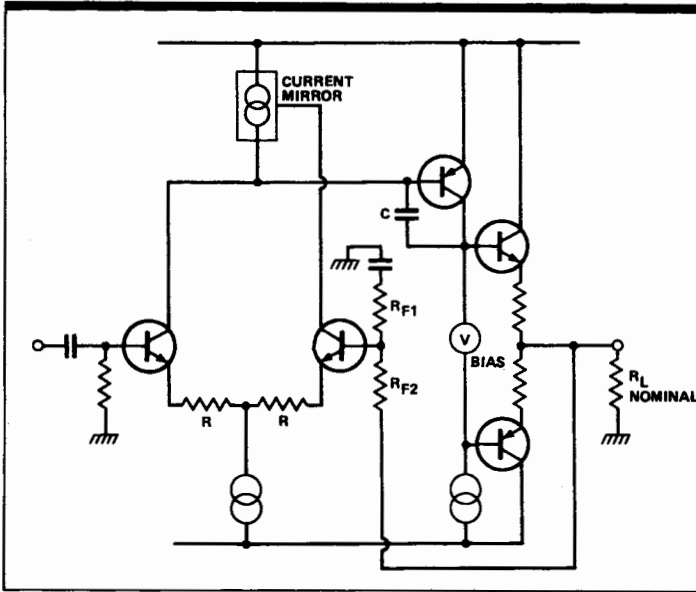
Fig. 2 Outline circuit of an audio power amplifier.

"Harmonic distortion and the usual intermodulation distortion decrease with increasing feedback. Transient intermodulation distortion (TIM) increases with increasing feedback, and is approximately directly proportional to the feedback. Therefore, there is an optimum value for the feedback at which the subjective distortion sensation is least. This optimum feedback is unlikely to exceed about 20 dB."

More recently, there has been conjecture that heavy overall feedback should be applied with caution if interface intermodulation distortion (IIM) is to be avoided. An amplifier should provide a low open-loop output impedance so that the need for feedback-generated loudspeaker damping is minimised.

There has also been conjecture that negative feedback, which reduces the usual intermodulation distortion, may increase phase intermodulation distortion (PIM) by converting amplitude non-linearities into phase non-linearities.

Unequivocally, none of these conjectures has any basis in the new NDFL amplifiers. As an aside, there is a substantial body of opinion that none of these conjectures has any basis, full stop.

Instability And Oscillation

A fundamental limit to the amount of feedback that can be applied to an amplifier is set by the onset of instability and oscillation.

If the magnitudes of the forward-path gain and demanded gain of the idealised Fig. 1 are plotted versus angular frequency w (in radian/second) on logarithmic scales, the resulting graph looks something like Fig. 3. The 3 dB bandwidth of the amplifier without feedback is $1/r_u$, and the gain-bandwidth product (at which gain drops to unity) is $1/r_1$.

Because the graph is on logarithmic scales, the separation between the curves of forward-path gain and demanded gain is the loop gain (remember then, to divide two numbers, you subtract their logarithms; if you divide u by $1/B$, you get uB). The magnitude of loop gain falls to unity at the frequency $1/r_x$ where the curves intersect and their separation is zero (remember that the logarithm of unity is zero).

By a similar argument, return difference is the separation between the curves of forward-path gain and closed-loop gain, as indicated in Fig. 3.

We could make a similar graph to Fig. 3, showing the phases of u and $1/B$. Again, the phase of loop gain would turn out to be the separation between the two curves. However, there is a remarkable piece of mathematics due to Bode, who used a transformation evolved by Hilbert (1862-1943), which shows that there is a relation between the magnitude and phase of the response of any linear system. Subject to some qualifications, our proposed graph of the phases is completely predictable from Fig. 3 and contains no new information.

As an example, many readers will know that, if the forward-path in Figs. 1 and 3 has a high frequency cut-off rate variously described as single pole, 20

dB/decade, or 6 dB/octave, then its phase shift is 45° at the 3 dB cut-off frequency $1/r_u$, and is asymptotic to 90° at very high frequencies.

In 1932, Nyquist applied a theorem which dates back to Cauchy (1789-1857) to drive the condition for a feedback amplifier to be stable and free from oscillation. If a polar plot is made of the magnitude and phase of return difference as frequency is varied, a vaguely 'snail-shaped' curve results. Such a polar plot is called a Nyquist diagram. Subject again to some qualifications, the stability criterion for a feedback amplifier is that its polar plot of return difference should not enclose the origin. Figure 4 shows one example each of a stable situation and an unstable situation.

Because the phase of return difference can be predicted from Fig. 3 via Bode's result, a Nyquist diagram can also be constructed from Fig. 3 and the onset of instability can be predicted. In 1945 Bode showed that Nyquist's criterion could in fact be expressed in terms of the gradients of the curves in Fig. 3, thereby eliminating the work of finding the phase explicitly and plotting the Nyquist diagram. Bode's exact rule is complicated, but a useful paraphrase is

"If in graphs such as Fig. 3 the separation between the forward-path gain and demanded gain decreases toward zero at a rate not exceeding 30 dB/decade, the amplifier is unlikely to oscillate."

This paraphrase makes no allowance for the tolerances on components. It assumes, in effect, that everything about the forward path is well known and constant. In the audio context, the paraphrase takes no cognizance of the fact that the capacitance of the leads that connect an amplifier and loudspeaker is anything but well known. A more conservative rule, appllicable to the audio context, is therefore

In graphs such as Fig. 3, the separation between the forward-path gain and demanded gain should not decrease towards zero at a rate exceeding 20 dB/decade."

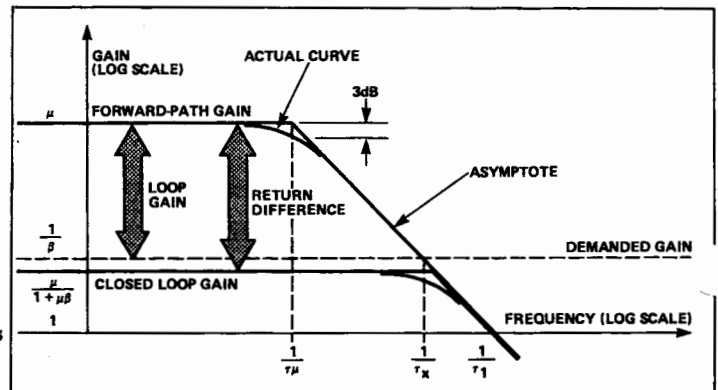The practical consequence is that the forward path of an audio amplifier with

Fig. 3 Logarithmic plots of gain versus frequency for Fig. 1.

conventional resistive feedback should have a single dominant pole which sets the fall-off of gain at frequencies above $1/r_u$. The second and subsequent poles should lie at frequencies substantially above $1/r_x$ (the frequency where the separation reaches zero), because each pole contributes a 20 dB/decade downwards slope to the graph of forward-gain path.

Maximum Available Feedback

In Fig. 2, the first stage is a differential amplifier with a current mirror at its output; the input and feedback signals are applied to the two bases to perform the subtraction process of Fig. 1. The second stage provides a large voltage gain, and the lag compensating capacitor C provides the dominant pole of the forward path corresponding to $1/r_u$ in Fig. 3. The third
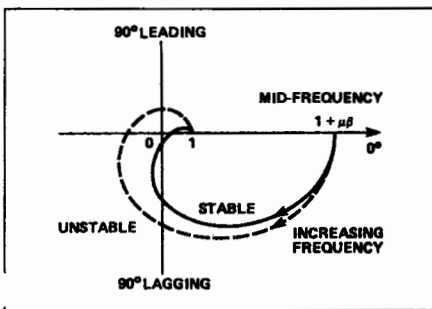
Fig. 4 Nyquist's stability criterion. The curves are polar plots of return difference for changing frequency.

stage is a complementary class-B emitter follower whose function is to transfer the output voltage from the second stage to the loudspeaker load. In practice, the transistors in the second and third stages are often Darlingtons, and the input transistors are often replaced by FETs.

In any similar amplifier, there is at least one pole associated with the finite transit time of electrons through each transistor. The transit time for typical small-signal transistors is a fraction of a nanosecond, but for power transistors of the ubiquitous 2N3055 class the transit time may be as long as a few tenths of a microsecond. Thus, the output stage of Fig. 2 may have a pole in the vicinity of 1 MHz.

As we saw in the previous section, the unity-loop-gain frquency $1/r_x$ in Fig. 3

must be substantially less than the frequency of all poles except the dominant pole $1/r_u$ if an amplifier is to be stable. If the power transistors are of the 3055 class, then no matter how fast the other transistors may be, there is going to be one pole at about 1 MHz. Therefore, $1/r_x$ must be chosen to correspond to something like 200 kHz. Even with more modern power transistors, $1/r_x$ is restricted to about 1 MHz. The art of designing a stable power amplifier involves choosing the lag compensating capacitor C such that $1/X_X$ is appropriate to the transistors actually used.

The geometry of Fig. 3 is such that, no matter how u, B and r_u are separately chosen, the return difference F(w) at any angular frequency w cannot exceed

$$F(\omega) \leqslant 1/\omega\tau_x$$

Thus, if $1/r_x$ is designed to correspond to 200 kHz, return difference at 20 kHz cannot exceed 10 (= 20 dB), and cannot exceed 200 (= 46 dB) at 1 kHz. An amplifier that boasts 80 dB of feedback (F = 10,000 at low frequencies) must have $1/r_u$ corresponding to about 20 Hz; return difference must begin falling above 20 Hz, and the former values at 1 kHz and 20 kHz (46 dB and 20 dB) still apply.

Returning now to Equation 7, the effectiveness of feedback in reducing distortion is set by the frequency of the distortion, not the frequency of the input. The audible frequency range is generally reckoned to extend to about 20 kHz and, with the foregoing constraints, return difference at this frequency cannot exceed 10. Remembering that 20 kHz is the third harmonic of 6.667 kHz, we see that feedback cannot reduce offensive odd-harmonic distortion of mid-treble input signals by more than a factor of 10. Remembering too that 20 kHz is the seventh harmonic of 2.857 kHz, we see that feedback cannot reduce crossover distortion of mid-range input signals by more than a factor of 10.

Until recently there has been no way around this problem except to increase the unity-loop-gain frequency $1/r_x$, and this demands that the frequencies of the transistor poles must be increased if stability is to be preserved. Fragile, expensive power transistors, with narrow bases to achieve short transit times, become mandatory.

The NDFL Approach

There is, however, another solution to the stability problem. If the forward-path gain has two dominant poles, so that its gain falls at 40 dB/decade, the rate of closure between the graphs of forward-path gain and demanded gain would still be 20 dB/decade provided the demanded gain itself were to fall at 20 dB/decade. In essentials, this requires that the usual frequency-independent resistive feedback factor B should be replaced by something having a frequency dependence of the form wr_F (remember that the demanded gain is the reciprocal of the feedback factor). Mathematicians tell us that a linearly rising frequency response corresponds to differentiation with respect to time and, in hardware terms, a capacitive feedback network will perform just this action.

Figure 5 shows the outline of an amplifier incorporating nested differentiating feedback loops. Notice first that the forward path has been separated into a number of stages, whose mid-frequency gains are u_1 to u_N respectively. The variable s is what mathematicians call complex frequency; for sinusoidal signals its magnitude is equal to the angular frequency w of the sinusoid. Factors of the form $(1 + sr_x$ represent a frequency response that rises proportional to frequency above the frequency $1/r_x$ — that is, they represent a zero. Similarly, factors of the form $1/(1 + sr)$ represent a frequency response that falls inversely proportional to frequency above the frequency $1/r_0$ — that is they represent a pole. Thus, the stages in Fig. 5 have special frequency responses: all stages except the first have a pole at $1/r$, and all except the first and last two have a zero at $1/r_x$.

Notice also that there are differentiating feedback networks, each denoted by ST_F, linking the output back to various points in the forward path. The resulting feedback loops are arranged one inside another, like a nest of Chinese boxes — hence the name nested differentiating feedback loops.

The amplifier is completed by an overall resistive feedback network B.

If we removed all the feedback from Fig. 5, the forward-path gain would be shown in Fig. 6: constant up to the frequency $1/T_0$, then falling at an N—1)-pole rate (20(N—1) dB/decade) up
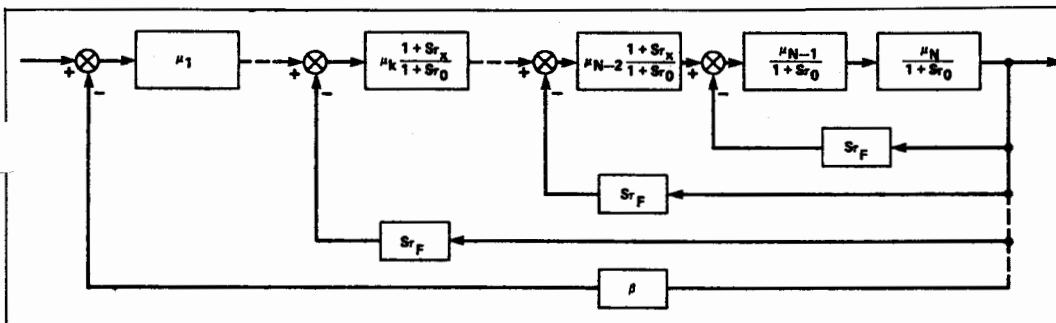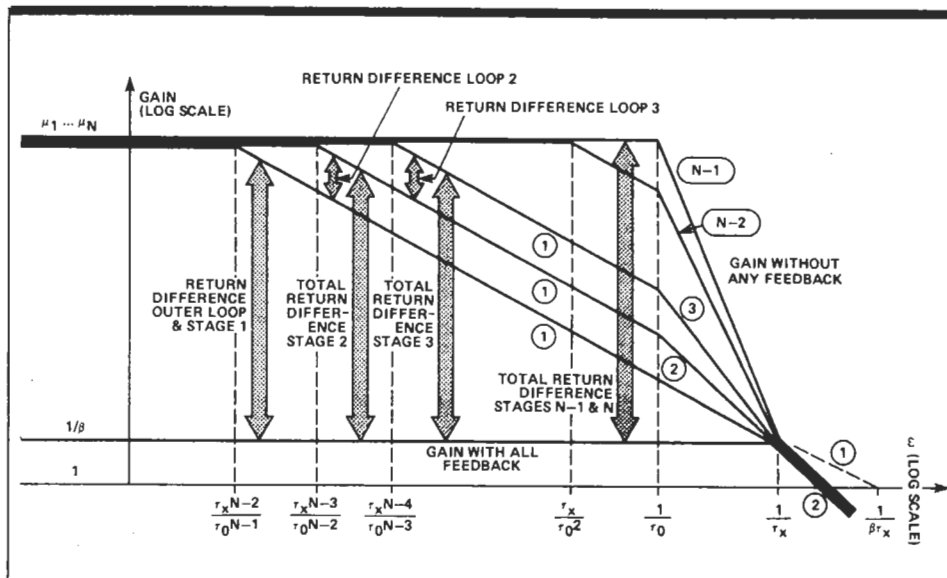

Fig. 5 Block diagram of an NDFL amplifier.

Fig. 9 Complete plots of gain versus frequency for Fig. 5.

Outline Practical Circuit

Figure 10 shows how an amplifier of the basic topology of Fig. 2 can be modified to include two NDFLs.

Notice first that the lag compensating capacitor, C, in the penultimate stge of Fig. 2 has been removed in Fig. 10. In its place are two capacitors (C) linking the output back to various points in the forward path. These capacitors are the feedback networks of the nested differentiating feedback loops.

The output stage has been changed to include a modified form of Thiele's load-stabilising network. Some form of LRC filter is required to locate one of the poles correctly, and with the circuit shown we get double value from the components.

The input stage itself is unchanged, but an inexpensive small capacitor in the overall feedback network B can be used to correct the group delay and improve the reproduction of transient waveforms.

Another essential addition is an amplifying stage between the two nested differentiating feedback factors. This rather peculiar circuit (which dates back to Rush in 1964) seems largely to have been forgotten. It uses one NPN transistor and one PNP to provide a well-defined gain (13).

As already suggested, once the demanded gain $1/B$ and the critical frequency $1/T_X$ are chosen, the circuit almost designs itself. The equations are:

$$\frac{R_{F1}}{R_{F1} + R_{F2}} = \beta,$$
$$RC = \beta\tau_x,$$
$$R_Y C_Y = \tau_x,$$
$$\tau_L = (\sqrt{3} - 1)\tau_x.$$

All stage gains and poles and zeros automatically look after themselves.

Figure 11(a) shows the 5 kHz square-wave response of Fig. 10 as built from 5%-tolerance resistors, 20%-tolerance

capacitors, and unselected production transistors. Evidently the circuit is 'designable'; Equations 12-15 really do predict component values for good transient response.

A nice feature of the modified Thiele circuit in Fig. 10 is that, when the load is made capacitive (a well-known source of high-frequency oscillation in amplifiers), the voltage waveform at the FEEDBACK POINT is the waveform the amplifier would have delivered into its nominal resistance load. Figures 11(b) and (c) illustrate this; the violent ringing in Fig. 11(b) is simply an LC resonance between the filter inductor and the load capacitance, and is in no way indicative of approaching instability.

Figure 12 shows details of the 1 kHz sinusoidal response under overdrive conditions. Note the quick, clean recovery.

An amplifier has been built in which the circuit can be switched from Fig. 2 to Fig. 10, to illustrate the improvement in performance of adding two NDFLs. Figure 13 compares the measured third-harmonic distortions of 1 kHz. Notice how the distortion of Fig. 10 drops away to below three parts per million at small signal amplitudes. Such behaviour is more typical of class-A amplifiers than class-B amplifiers, and may account for the clean sound of NDFL amplifiers.

Crossover distortion associated with incorrect bias of the output stage is one of the most audibly annoying forms of distortion. Audio amplifiers based on Fig. 2 sometimes have a type of crossover distortion that does not show up in normal measurements. Correct biasing of the output stage relies on close tracking of the thermally-compensated biasing device and the power transistors. At best the biasing device can be thermally bonded to the power transistor case. More usually it is bonded to the heatsink, but there is no way it can simultaneously sense the actual junction temperatures of all the power transistors. Under rapidly-fluctuating dynamic signal conditions, the junction
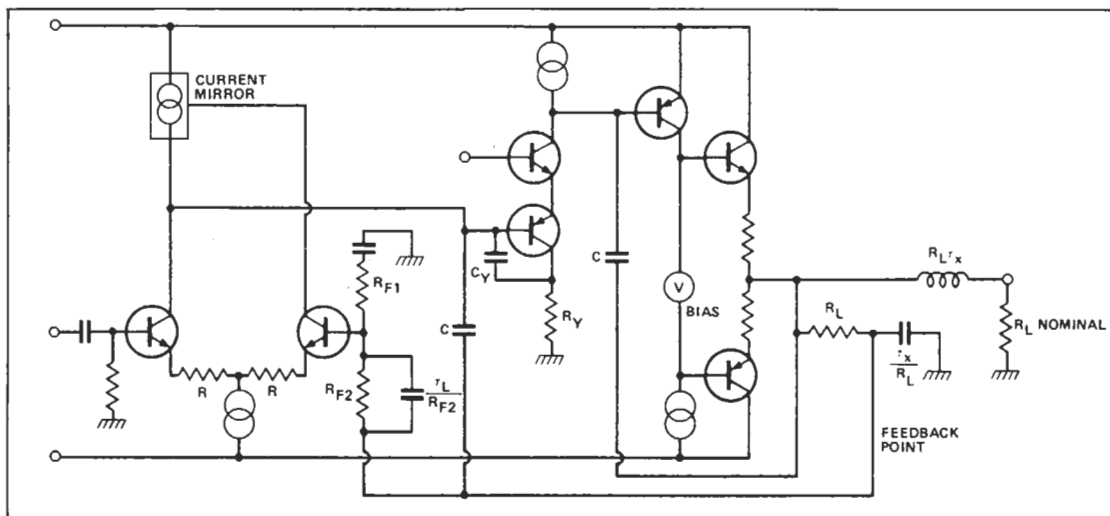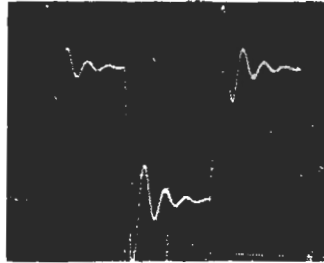

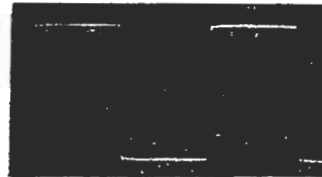Fig. 10 Outline circuit for an NDFL amplifier.

(a) 8 ohm resistance load.

(c) waveform at feedback point for (b).

(b) 8 ohm and 2uF parallel load.

Fig. 11 5 kHz square wave response of Fig. 10.

Fig. 12 Detail of output waveform from Fig. 10 under overdrive.

temperatures may be wildly different from each other and from the case or heatsink temperatures, and therefore the biasing may be wrong.

Figure 14 compares the static crossover distortion of Figs. 2 and 10 when the bias is deliberately set 0V5 too low. Dynamic mistracking of the biasing circuit should not introduce audible crossover distortion in an NDFL amplifier.

One final point. The NDFL technique maximises the return difference (and hence minimises distortion components) at frequencies up to $1/T_O$ Above this frequency the return difference falls away rapidly, and distortion rises. Choosing $1/T_O$ to correspond to 20 kHz minimises audible-frequency distortion, but does not minimise ultrasonic distortion.

For example, a common specification for audio power amplifiers is their THD at 20 kHz. The harmonics of 20 kHz lie at 40 kHz, 60 kHz, 80 kHz, and so on. All are ultrasonic (and hence inaudible) and the NDFL technique does not minimise them. A measurement of THD at 20 kHz may therefore give a quite misleading indication of an NDFL amplifier's audible per-

Fig. 14 2 kHz crossover distortion when bias is set wrongly.
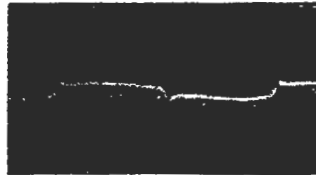
(a) Fig. 2 (conventional amplifier).
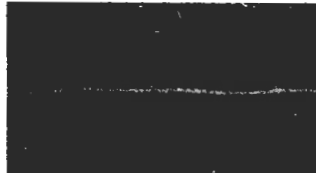
(b) Fig. 10 (NDFL amplifier).

formance. Valid objective tests include the SMPTE and CCIF tests for two-tone intermodulation distortion, the proposed IEC test for TIM, Cordell's proposed three-tone test for TIM and the pro-
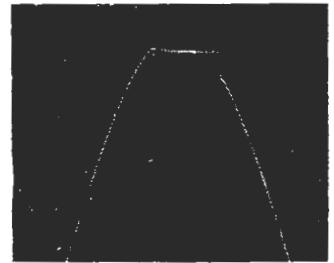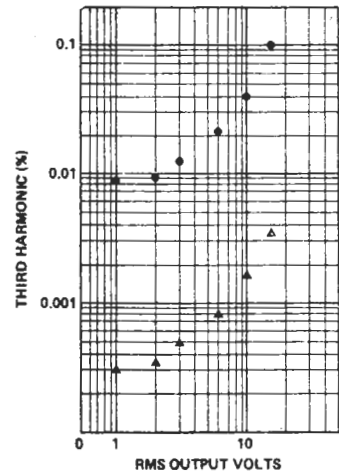

● FIGURE 2 (CONVENTIONAL AMPLIFIER)
▲ FIGURE 10 (NDFL AMPLIFIER)

Fig. 13 1 kHz harmonic distortion.

posed test for input-output intermodulation distortion IOD. The distinguishing feature of all these tests is that they measure the distortion at audible frequencies. **ETI**