The Quasi Complimentary Output Stage
A Topology History and Review

1. History and Review

In the quasi-complimentary audio amplifier output stage, an identical pair of output transistors or MOSFET's is used in the upper and lower halves, usually driven by a complimentary pair of driver transistors. The two halves of the output stage are therefore neither symmetrical nor mirror images of each other; the circuitry of the two halves is quite different. This critical factor classifies the topology and dictates certain benefits but performance limitations.

In the early stages of audio amplifier development in the early 1960's there was little choice but to use the quasi-complimentary format. PNP silicon power transistors of ruggedness comparable to contemporary NPN's were simply not available. Interestingly, the Lin design of 1956 [1], a major breakthrough in transistor amplifier design, (in that although a push-pull design it did away with the earlier requirement for input and output transformers), was an all-*PNP* quasi-complimentary design. At the time only germanium transistors, and PNP types, were readily available. Germanium devices have a much more gradual turn on characteristic than silicon, so crossover distortion problems that were to dominate later judgement of this topology, were not of early concern.

As silicon transistors supplanted germanium and PNP technology now remained weaker than NPN for quite some time, NPN quasi-complimentary output stages followed through the early 1960's. But by 1968 the quasi-complimentary topology had been all but totally discredited, as suitable silicon PNP transistors became available. Quasi-complimentary design was then quickly overshadowed by full complementary configurations.

Dilley and later Bailey discussed [2] the fundamental problem that differing input impedances of the two halves of the output stage inevitably gave greater overall distortion than a complimentary symmetry output stage and even the earlier transformer coupled configuration. In the upper half of the output stage the input impedance is that of two base-emitter junctions, (Q1 and Q3 in **Figure 1**), while in the lower half the input signal is applied to just one, (Q2, **Figure 1**). This resulted in marked asymmetry between input impedances of the two halves, notably at low output currents, giving rise to serious crossover distortion. Both halves exhibited quite different transfer characteristic curves, so simple resistive equalisation between the two was never going to be possible, nor was suitable biasing to remove the crossover region problem a viable option.

While regarding quasi-complimentary as having better output stage bandwidth and linearity than transformer driven designs, Haas indicated [3] much poorer tolerance of quiescent current bias to power supply variation than a full complimentary structure.

Poor distortion performance of such early transistor audio amplifiers occurred at low rather than high volume, where much listening is in practice done. This was the reverse of valve amplifier performance, with which many people were more familiar. This contributed significantly to re-emergence of supposedly superseded valve technology and use of Class A. The counter culture continues to this day, despite all the advances in Class AB transistor technology since the late '60's [4].

Meanwhile, in 1968 the Acoustical Manufacturing Company announced [5] their "Quad 303" amplifier development of the quasi-complimentary design, containing output triples instead of just the conventional driver and output devices. The triple topology addressed the crossover distortion problem by adding sufficient gain such as to make both halves of the output stage act very much more as pure emitter followers, appearing closely matched.

Other designers, such as Dilley [6], Visch [7] and Stevens [8], proposed improvements to the basic circuit. But much more critical to the development of the design was the earlier introduction of a modification by first Shaw [9] and then in alternate form by Baxandall [10], coming to be known as the "Baxandall diode"; (D1 of **Figure 1**). The diode is intended to mimic the same paired Vbe drop and impedance characteristic in the lower output half of the output stage as the paired base-emitter junctions occurring already in the top half. This yields radical improvement, and distortion characteristics claimed to be almost on a par with complimentary symmetry emitter follower designs, (and better than some FET topologies [11]), and of quite acceptable degree, broadly speaking. Some analysis by calculation was contributed by Blundell [12].

Hood later added a capacitor across the Baxandall diode [13]. This was accepted with little or no later review or analysis, presumably on a "the more the merrier" basis, and a huge variety of values have been variously applied [14]. By the time of such developments however, and from this time onward, the inherent and obvious symmetry advantages of using matched NPN and PNP output pairs vastly overshadowed the quasi-complimentary design, which consequently faded from the limelight.

Discussing the advantages of full complimentary output stages four years later in 1974, noted US designer James Bongiorno, (of S.A.E. and later G.A.S., both companies producing amplifiers in the 200W plus range in the mid '70's), commented adversely on the quasi-complimentary configuration [15]. *"... the classic cross-over notch cannot be eliminated with this type of output stage design ... Another difficult problem associated with quasi-complimentary output stages is their high frequency instability, which is more commonly described as common mode conduction or latch-up in the output stage itself. This problem is due to phase shifts within the output stage itself and is almost impossible to cure."* Technical Notes para. 4, following, discusses common mode conduction. This and the other problems Bongiorno refers to are however not solely specific to quasi-complimentary designs.

Reasons for considering the quasi-complimentary approach in variations including but not limited to that with the Baxandall diode, remain. High power NPN's remain cheaper and more commonly available, and the possibility of a price discount due to doubled quantity of a single device also exists. Where such price advantage is critical and some, albeit minimal, decrease in distortion performance can be acceptable, the quasi-complimentary design should be at least considered, and there are several practical examples existing [16].

It has also been suggested [17] that use of identical type output transistors in preference to compliments can lessen certain distortions. PNP transistors tend to have a lower ft than equivalent NPN's, due to differing majority carriers between the two transistor forms. This means that at high frequencies supposedly complimentary output transistors are anything but, and high frequency asymmetry worsens crossover and other distortion performance.

Quasi-complimentary FET designs have of course also been proposed, [18], [19] and at least one quasi-complimentary valve design too [20]! Complimentary high power MOSFET pairs remain less than common, less than ideally matched in characteristics, and certainly expensive - much more so than BJT's even after some years of development. Cheaper high Vds N-channel MOSFET's, perhaps intended for switching applications, purchased in higher quantity and without having to consider matching to a compliment, may well confer advantage. Crossover distortion performance remains notably poorer than conventional configurations however. The basic circuit configuration remains badly asymmetrical, and no improvement is offered by Baxandall Diode type modification. However, it may be considered that such reduction in performance is again acceptable from a circuit conferring simplicity and economy.

Today use of quasi-complimentary output stages is avoided where possible but certainly not unknown, especially in regard to high power professional use amplifiers and even with high end Hi-Fi [21]. There remain cost advantages to the use of quasi-complimentary stages, specifically where high power output is sought, and some reduction in distortion performance is acceptable. Possibly there may be some technical advantages too; output transistors of identical type and closely matched parameters such as gain-bandwidth may help minimise distortion at higher audio frequencies [22].

Much development of the basic quasi-complimentary output stage has continued in the field of IC manufacture. For example, the well known and ubiquitous Sanyo "STK" series of audio power amplifier IC's included several [23] using the quasi-complimentary topology, in devices designed in the 1970's and produced well into the 1980's. Many of the Sanken hybrid power IC's provide other examples [24]. In IC manufacture the use of all NPN output stage continued to provide significant advantage, but carried with it difficulties in regard to the PNP lower half driver, specific to the nature of IC fabrication. PNP IC fabricated devices suffered from fT an order of magnitude or more less than their NPN counterparts, had significantly lower gain, introduced unwanted phase shift, had lower current handling capability and required excessive chip area. "As a consequence of these deficiencies, IC designers avoid using the lateral PNP transistor wherever possible, and continually seek alternatives to it." [25]

In a 1974 and 1975 circuit designs, Hitachi [26] and RCA [27] IC designers addressed the problem of distortion caused by widely differing gains of PNP compared to NPN transistors when fabricated in IC form, (refer to **Figure 4** and **Figure 5** for the circuit and operation summary). Not only is the gain different between NPN's and PNP's but the gain of the PNP lower half driver, (in the conventional circuit), varies significantly with collector-emitter voltage, which itself varies substantially during each cycle of conduction in normal operation. The result is even order harmonic distortion in the output signal. Use of feedback to try

and clean this up results in new issues to be dealt with, including stability and the need for more gain and more gain stages. A MOSFET can be used, instead of the lower half PNP driver, to avoid the gain variation with collector-emitter voltage problem [28]. Special bias for the IC fabricated MOSFET is then required, or it can't be made to produce signal swing acceptably close to the supply rail.

Schade of RCA demonstrated [29] quasi-complimentary with series connected outputs and drivers to permit use of higher supply voltages in discrete designs, and again facilitate (driver) manufacture in IC form. National Semiconductor's 1976 IC circuit by Russell [30] et al. used a P-channel JFET lower driver for signal inversion, with a bias circuit allowing the upper NPN output device to provide a feed-forward boost to the FET on high frequency negative signal excursions, to hold up overall high frequency response. Refer to Figures 6 for the circuit and operation summary.

The National Semiconductor LM12 TO-3 pack Power Op Amp IC of 1986 is capable of 80W into 4Ω with only 0.01% distortion. The manufacturer's Application Note [31] discusses the advantages of the design, and the problems with conventional quasi-complimentary topology that were overcome. "*It is stable with all reactive loads and does not have the spurious-oscillation problems observed with the familiar quasi-complimentary amplifier.   Controlled high-frequency response is a significant advantage of this design, especially when compared to standard quasi-complimentary. The frequency compensation, capacitive loading, asymmetrical response and cross-over distortion problems often encountered with the configuration are conspicuously absent*". Refer to Figure 3 following for a more detailed discussion of the LM12 output stage circuit.

US manufacturer QSC [32] released in 1997 an addition to their substantial range of "Powerlight" public address amplifiers; the Powerlight 8.0PFC; 4kW per channel into 2Ω, 8kW into 4Ω, and all in a three rack unit case. Many special techniques are used to achieve this, including a switch mode power supply, with power factor correction, and a four step Class H supply rail arrangement. The output stages use high power N-channel MOSFET's in a quasi-complimentary full-grounded bridge output circuit, according to the company's literature. It can be expected that substantial effort will have gone into development of the basic topology in regard to crossover distortion, by the manufacturers of high power amplifiers such as QSC.

The Cyberlogic NC-412 amplifier is another example. Rated at 1200W into 4Ω, (by 4 channels), uses all N-channel vertical MOSFET output stages. Its specification for THD is <0.085% @ full rated power into 4Ω, 1kHz. Rating THD at full power is likely to minimise the figure; as a percentage THD is likely to be somewhat greater at lower levels, and worse at higher frequencies. In any case THD distortion less than a tenth of this figure might be technically practical and expected from a comparably powered full complimentary output stage design. In practice however, large professional amplifiers with full complimentary Class AB output stages and THD+IM up to 0.1% are common; the QSC USA400, 900 and 1310 range for instance.

2. Technical Notes

2.1 Parallel Resistor

The value of resistor in parallel with the Baxandall Diode, (R6 in **Figure 1**), is usually made the same as that in the collector circuit of the lower half driver transistor, (R8 in **Figure 1**). The value does not appear to be critical [33].

2.2 Parallel Capacitor

Regularly seen as a 10nF~22nF [34] capacitor, (C1 of **Figure 1**), in parallel with the Baxandall Diode, is a further minor modification by Hood [35]. The value does not appear to be too critical. The intent it seems is to simulate the effect of the missing PNP output transistor base-emitter capacitance [36]. "*The capacitor thus gives another small improvement in output stage symmetry*" [37] but some authorities [38] it would appear, do not regard it worthy of inclusion, even when focusing on distortion performance of such circuits.

2.3 Drive Source Impedances

A consideration and comparison of the drive source impedance characteristics of the two halves of the quasi-complimentary output stage [39] suggests that less than complete drive symmetry still exists despite

the Baxandall diode etc. It would appear that this has little or no practical effect in view of other components within the stage being deliberately placed, (in conventional designs), to impose earlier slew rate or frequency response limitations for stability purposes.

2.4 Common Mode Conduction

Drive current problems can occur in output stages as output device gain falls with increasing frequency. Where an output device can no longer turn OFF fast enough to follow the driving waveform, feedback around the amplifier as a whole acts to compensate and turns ON the other output transistor. Common mode conduction can then occur. With both top and bottom halves of the output stage both on at the same time, conduction is then between the positive and negative power supply rails. For a given signal output amplitude, power supply current then can be shown to increase with frequency. Overall efficiency correspondingly decreases, because common mode current does not contribute to useful output current to the load.

Faster devices with better switching speed, or active reverse biasing methods involving interconnection of the output transistor bases, or minimal Rbe, (R2 and R8 in **Figure 1**), are necessary to overcome such problems. "*Because of the lack of symmetry, there is no easy way to achieve reverse biasing for both outputs in a quasi-complimentary amplifier.*" [40] The only options therefore are firstly to ensure that output devices with relatively high switching speeds are used. This may mean a trade-off in device parameters such as ruggedness, lower dc current gain, higher base-emitter voltages and higher saturation voltage. Secondly, Rbe should be low to decrease switch off time. This however increases drive requirements and driver dissipation.

2.5 Output Emitter and Collector Resistors

For greater power output it is necessary to parallel additional output transistors, (Q5 and Q6 in **Figure 2**). To aid equal current sharing each needs its own emitter resistor [41]. Conventional circuits often do not include an emitter resistor for a single lower half output transistor at all although it should be there, (R9 in **Figure 1**, R15 and R16 in **Figure 2**). The emitter resister assists thermal stability, (as well as current sharing with parallel devices), at some minor expense with regard to efficiency. The resistor in the collector circuit of the lower half output transistor, (R5 in **Figure 1**, R9 in **Figure 2**), does not serve this purpose. This resistor needs to be present to help make the lower half of the output stage mirror the top half, and while totally necessary also acts at the expense of efficiency.

3. Example Circuits

3.1 Example Quasi Complimentary Output Stage, Figure 3

National Semiconductor's LM12 Monolithic Power Op Amp, simplified schematic of the output stage. "*The high-frequency path around the PNP in the quasi-complimentary output stage causes it to behave like a true complimentary amplifier.*" [42]

Operation

The design features the provision of a high frequency signal path around the conventional inverting PNP, (Q15), of the quasi-complimentary output stage. This path is through Q9.

The biasing circuit for Q9 includes constant current source I4, the base-emitter diode of Q7, and R3. 500μA across R3, 400Ω, gives 200mV, which is mirrored across R6. R6 current is thereby set at 200mV/200Ω = 1mA. Q8 and Q9's collector currents sum to this 1mA, and each operate about a median 500μA. In the quiescent state 500μA from constant current source I5 flows through Q8. When the top half of the output is active I5 drives output Q16 via R5 and current in Q8 is reduced. When the lower half is active I5 plus base current from Q15 flows through Q8.

When the top half of the output stage is active, inverting Q8 is being driven such that less collector current is flowing in it, its collector voltage is rising and Q16 is conducting. At this time, decreased collector-emitter current in Q8 means the R6 current is increasingly supplied by Q9. In other words, Q9 conducts more

heavily. Q8's collector voltage rise reverse biases Q15 and Q17 so although Q9 is being biased to conduct more heavily Q15 restricts the opportunity for Q9 collector-emitter current to flow. There therefore comes the concern that Q9 will saturate and both Q9 and Q17 will therefore be slow to come on when the drive changes from sourcing to sinking output current.

Clamp diode D1 keeps Q9 out of saturation, by allowing Q9 collector-emitter current to continue to flow in this situation. In addition, Q17 is kept at the threshold of conduction. Both can quickly commence conduction when drive and output reverse polarity. "*Keeping Q17 from turning off completely reduces high-frequency cross-over distortion.*"

When the lower half of the output stage is active, Q8 is being driven such that increased collector current is flowing in it, its collector voltage is lowered and Q16 is off.   Q15 and hence Q17 are forward biased and conducting. Increased conduction in Q8 means the reverse for Q9, as before; Q9's conduction is decreased. This is the mode of operation of the output stage when sinking load current, at low frequencies. At higher frequencies the operation becomes different.

At higher frequencies the gain of the inverting PNP, Q15, drops and less drive current therefore flows to Q17 than it should. A path through Q9 takes over however.   Q9 effectively receives drive to its emitter from Q8. Q9's base voltage is fixed by the I4, Q7, R3 bias network. So Q9's collector rises and Q17 continues to get forward drive despite Q15's gain fall-off. Drive is maintained at high frequencies [43].

In summary, at high frequencies and on alternate half cycles, first the drive path is through invertor Q8 and then emitter follower Q16, and then second the drive path is through emitter follower Q8 and invertor Q17. Depending on whether drive is coming from its collector or emitter, Q8 act as both inverter and then emitter follower. "*The AC equivalent circuits are about the same for the two paths*".

Careful design consideration has gone into the positioning of compensation capacitor C1 to minimise high frequency crossover distortion while avoiding spurious oscillations. "*Controlled high frequency response is a significant advantage of this design especially when compared to standard quasi-complimentary*".

Q12, Q13 and Q14 are a "boost" circuit that increases the low frequency gain of the output emitter follower Q16, to equalise the gain of the upper and lower halves of the output stage. An inherent "problem" of the conventional quasi-complimentary output stage is of course that the upper NPN, (Q16 in this case), is configured as an emitter follower with zero voltage gain, while the lower output transistor, (Q17 in this case), has the speaker impedance in its collector circuit acting as a load. It therefore has voltage gain, albeit very low. The boost circuit provides low frequency voltage gain to the upper half to equalise with that inherent for Q17.

R5 in the boost circuit is kept as low as possible such that "*the boost is effectively bypassed at high frequencies. The boost is, in fact, more stable than an extra follower.*"

3.2 Example Quasi Complimentary Output Stage, Figure 4

Output stage intended for ease of IC fabrication, by Y. Sakamoto, S. Iguchi of Hitachi, U.S. Patent 3,813,606, May 28, 1974. Q6 and Q7 can also be NPN's, further aiding IC manufacture; it is simply necessary that both be of the same conductivity type.

Operation

Q6 and Q7 act together as an inverting amplifier with unity gain; i.e. as an inverting buffer. Effectively 100% local negative feedback occurs between the collector and base of Q7. By providing the necessary signal inversion for the lower half of the quasi-complimentary output stage, but not of itself contributing to the overall gain of the lower half, the Q6~Q7 circuit enables the gains of the upper and lower halves of the output stage to be closely matched. The Q10 x Q11 gain can be well matched to the Q8 x Q9 gain.

The Patent discussion indicates that IC fabricated PNP's were lateral types with gain 5~15, poorly matched to vertical type NPN's with gain 50~70. Distortion due to the gain imbalance is the result in a conventional circuit, and stability improvements were claimed as well.

3.3 Example Quasi Complimentary Output Stage, Figure 5

Output stage intended for ease of IC fabrication, by M. B. Knight and N. Caldwell, U.S. Patent 3,863,169, Jan. 28, 1975. The current mirror Q25, R24, Q27 and R26, may be replaced by two FET's performing the same function.

Operation

Q15 and the current mirror Q25, R24, Q27 and R26 act to replace the inverting PNP found in the lower half of a conventional circuit. Q15 is in common-base configuration, and has substantially unity current gain. On negative signal excursions the "input" voltage level to Q15 base falls, and Q15 is forward biased and conducts. Q15 collector current flows through R24 and Q25, and is mirrored in R26, Q27. In the process inversion has occurred; the R26, Q27 current now drives Q30 and Q2. The lower NPN outputs can be closely matched to the upper NPN outputs, Q6 and Q1.

The gain of the current mirror, Q25, R24, Q27 and R26, depends on the ratio of R24 and R26 and matching of Q25 and Q27 base-emitter junctions. Usually these are equal and the current mirror gain is -1. The point is that the gain of the current mirror is substantially independent of the gains of Q25 and Q27, the PNP devices used in the circuit. The symmetry of current gains gives the amplifier low even order harmonic distortion.

3.4 Example Quasi Complimentary Output Stage, Figure 6

Output stage intended for IC fabrication, by R. W. Russell and K. M. Black of National Semiconductor, U.S. Patent 3,974,456, Aug. 10, 1976. The circuit maintains a bias on Q21 such that Q21 has some limited conduction through both positive and negative signal excursions. For high frequency negative excursions this continuous forward bias allows signal to pass from input to output via Q21's base emitter, around slower responding P channel JFET Q32. In this way high frequency response is maintained for negative excursion signals, that otherwise would be lost. Overall loop stability and bandwidth is improved.

Operation

Q21 forms a feed-forward loop around FET Q32 at high frequencies for negative signal excursions. Q21 is biased for continuous conduction, to a limited degree. The method of establishing this bias is as follows.

For a JFET, current Idss flows with no forward bias applied, and with gate and source shorted together. JFET's are commonly used as constant current sources utilising this performance. Here, under quiescent conditions, Idss of Q23 is mirrored through Q28, establishing the same current equal to Idss through matched JFET Q32. Thus Q32's gate-source voltage becomes zero, following that of Q23, where gate and source are shorted together.

With Q32 gate-source voltage being zero, the base-emitter voltage drop across Q21 becomes equal to the forward voltage drop across diode D18. D18 and Q21 are manufactured with closely matched junction characteristics.

The current through D18 is that through Q23, (which is Idss), plus the current through constant current sink I(31). The current through Q21 is equal to this and is the current through Q32, (which is Idss), plus an amount equal to I(31), which flows through Q34.

For positive signal excursions the signal path is as expected, via Q16 and Q21. For low frequency negative excursions the signal path is via D18, Q32 and Q34. The voltage at the cathode end of D18 becomes more negative, Q32 becomes forward biased and increased conduction in Q34 results. JFET Q32 provides the necessary signal inversion to drive NPN Q34.

For high frequency negative signal excursions Q32 responds too slowly. But as detailed above, Q21 has some limited forward bias and conduction, even during negative signal excursions. This conduction means that signal is passed forward from Q21's base to emitter now. The high frequency negative signal excursion

causes Q21 conduction to be decreased; Q34 current - which is Q21 current plus current being sunk from the load - remains unchanged, as drive from Q32 hasn't responded to the signal quickly enough. The result is a change in the balance of the two currents that make up that through Q34; decreased Q21 current means increased current sunk from the load. Q21 has provided a feed-forward "boost" to Q32 for high frequency negative signal excursions, thereby extending overall high frequency performance.

4. Notes and References

[1] "Quasi-Complimentary Transistor Amplifier", H.C. Lin, Electronics, September 1956, pp.173~175, and U.S. Patent 2,896,029, "Semiconductor Amplifier Circuits", July 21, 1959, Hung Chang Lin, RCA Corporation.

[2] "A New Solid State Amplifier for Studio and Broadcast Use", W. L. Dilley, JAES, October 1965, Volume 13, No. 4, pp332~344 and "30-Watt High Fidelity Amplifier", A.R. Bailey, Wireless World, May 1968, pp.94~98.

[3] "Design Factors and Considerations in Full Complimentary Symmetry Audio Power Amplifiers", G.C. Haas, JAES, July 1968, Volume 16, No. 3.

[4] "The Art of Linear Electronics", J. Linsley Hood, Butterworth Heinemann, 1993, ISBN: 0750637463, pp.157.

[5] "Low Distortion Class B Output", Acoustical Manufacturing Co., Wireless World, April 1968, p.67.

[6] "A New Solid State Amplifier for Studio and Broadcast Use", W. L. Dilley, JAES, October 1965, Volume 13, No. 4, pp332~344, discussing development of a line level rather than output amplifier. The design process remains of interest however. Dilley uses a phase splitter pre-driver with upper and lower half drives taken from collector and emitter respectively. All NPN driver and output upper and lower halves are then identical. Marked reduction in distortion was claimed. To then achieve DC coupling an alternate configuration used a conventional VAS feeding two PNP drivers in a differential pair configuration, each in turn driving an NPN output as collector load.

[7] Wireless World, N.M. Visch, April 1975, p.166. The unusual Visch circuit endeavours to control the output stage quiescent current, independently of adjustment or temperature, using a constant current source. This apparently produces a circuit superior to the Quad triple by virtue of the absence of "take over distortion", an ill-defined term.

[8] Hi-Fi News and Record Review, G. A. Stevens, February 1971, pp.252~253.

[9] "Quasi-Complimentary Output Stage Modification", I. M. Shaw, Wireless World, June 1969, pp.265~266. Shaw's simple, cheap diode modification gave distortion reduction on his comparable test amplifiers from 1.5~2% for a conventional circuit at low power, to 0.5~0.6%; a major step forward.

[10] "Symmetry in a Class B", P.J. Baxandall, Letters, Wireless World, September 1969, pp.416~417. The text demonstrates clearly, through transfer characteristic curve plots, the benefit of the added diode.

Both Baxandall and Shaw add the diode such as to recreate a symmetry of Vbe junction voltage drops between upper and lower halves of the output stage. The topologies they apply the modification to are slightly different, so Baxandall's diode appears at first sight to be in quite a different position to Shaw's. In Baxandall's circuit the driver emitter resistor is returned from the driver emitter to the output transistor collector. In Shaw's circuit the driver emitter resistor is returned to the amplifier output. The two connection points differ by the low value "quasi-emitter-resistor" in the lower half output transistor's collector circuit.

Although Baxandall is in no way disparaging of the earlier Shaw's work, it is possible to read into Baxandall's article a piqued tone. Baxandall refers to having already annunciated the concept "in a recent London lecture to the B.K.S.T.S." , without suggesting a date or referencing a conference paper or some-such. Presumably Wireless World readers at the time would have been familiar with who or what the " B.K.S.T.S." was! Baxandall then goes on to show little if any measured performance difference to Shaw's idea, while consistently slightly misquoting Shaw's driver resistor arrangement.

On balance it would appear to be fair to say that both Baxandall and to a lesser degree Hood merely added to an idea first published by Shaw earlier, in sufficient detail and distinction to have given his name to the new circuit. Peter Baxandall was by far the figure of greater renown however, and such is the way of the world!

[11] "Audio Power Amplifier Design Handbook", D. Self, Newnes, 1996, ISBN 0750627883, p.101.

[12] "New Approach to Transistor Circuit Analysis", A. J. Blundell, Wireless World, June (p.287~292) and July (p.329~ 334), 1971.

[13] 75W Amplifier, Hi-Fi News and Record Review, J. Linsley Hood, November 1972, pp.2120~2123, and "Solid State Audio Power", J. Linsley Hood, Electronics World and Wireless World, November 1989, p.1044.

[14] Typically 10~20nF; 220nF in "An Approach to Audio Amplifier Design", J.R. Stuart, Wireless World, September 1973, p.444 seems a little excessive.

[15] "High Voltage Amp Design", James Bongiorno, Audio, February 1974, p.30.

[16] "Playmaster Twin Twenty Five Stereo Amplifier", Leo Simpson, Electronics Australia, April 1976, pp.36~41, May 1976, pp.56~67 and p.125, September 1976, pp.61~63, using 2N3055's as output devices, the circuit appearing again in "Guitar Amplifier for Practice Sessions", John Clarke, Electronics Australia, March 1981, pp.56~64.

[17] "Solid State Audio Power", J. Linsley Hood, Electronics World and Wireless World, November 1989, pp.1042~1048, December 1989, pp.1164~1168, January 1990, pp.16~21.

[18] A PNP BJT is added to an N-channel FET in the lower half to act in place of a P-channel FET, ref. "Sound MOSFET Design", D. Self, Electronics World and Wireless World, September 1990, pp.760~764.

[19] In "Better Audio From Non-Compliments?", Electronics World and Wireless World, December 1994, pp.988~992, Bengt Olsson puts forward an unusual two rather than three stage design, somewhat unfavourably reviewed in "Few Compliments for Non-Compliments", D. Self, Electronics World and Wireless World, September 1995, pp.762~765. Refer also to Anthony Holton's directly Olsson derivative "N-Channel" amplifier project, featured on his content heavy web site.

[20] Linear Technology, Application Note 18, "Power Gain Stages for Monolithic Amplifiers", March 1986, Jim Williams, p.8. While not intended for audio applications specifically, this circuit shows what is possible, and the problems inherent in trying to construct the lower half of the stage with valves. Performance suffers, as is shown by the same drive circuit with full complimentary output, at p.6.

[21] D. Sweeny and S. Mantz, in their review "An Informal History of Amplifiers", Audio, June 1988, pp.46~55, "*such schemes are still advocated by high end designers John Bedini and Bascom King*".

[22] "Solid State Audio Power", J. Linsley Hood, Electronics World and Wireless World, November 1989, p.1044.

[23] At a rough estimate as many as half the "STK" audio power amplifier IC's listed in, (for example), the Sanyo 1985 Semiconductor Handbook "Thick Film Hybrid Integrated Circuits" have quasi rather than full complimentary outputs.

[24] SI-1000G series (SI-1020G, SI-1030G and SI-1050G), Sanken Bulletin No. QA-08B, Sept. 1980. THD is specified as 0.5% max. measured at full output. The comparable Sanyo STK077, STK080, STK084, (also 20W, 30W and 50W respectively), are rated at 0.2%~0.3% THD, (Pout = 0.1W to Pout(max), f = 20Hz~20kHz).

[25] "Bipolar-Transistor IC Technology - What's Happening Now?", M. V. Hoover, RCA Solid State Division, IEEE Transactions on Consumer Electronics, Vol. CE-23, No. 4, Nov. 1977, pp.505 ~517. Hoover describes the "vertical" NPN as the "prime mover" used in bipolar transistor IC's, the requirement for a suitable complimentary "vertical" PNP, but the limited suitability of PNP "verticals". These "*can only be used in emitter follower circuits*" due to their mode of design in IC form. The "lateral" PNP is the form of PNP used in IC's and its disadvantages are as quoted. It remains clear however that even as emitter followers, "vertical" PNP's remain substantially less than desirable than alternate arrangements using NPN's. Output stages, (where as an emitter follower the "vertical" PNP might be supposed to find suitable application),

using just NPN's remained popular with IC designers, and the subject of much state of the art engineering development through the 1970's.

[26] "Transistor Circuit", U.S. Patent 3,813,606, May 28, 1974, Y. Sakamoto, S. Iguchi, Hitachi.

[27] "Composite Transistor Circuit", U. S. Patent 3,863,169, Jan. 28 1975, M. B. Knight, N. Caldwell, "Bias Circuitry for Stacked Transistor Power Amplifier Stages", U.S. Patents 3,886,466 and 3,887,880, May 27 1975 by C. F. Wheatley, Jr. and June 3 1975 by A. J. Leidich, all of RCA Corp.

[28] "Bipolar-Transistor IC Technology - What's Happening Now?", M. V. Hoover, RCA Solid State Division, IEEE Transactions on Consumer Electronics, Vol. CE-23, No. 4, Nov. 1977, p. 508, Fig. 10 showing a 1973 development, the RCA CA3100 op amp.

[29] "Transistor Series Amplifier", U.S. Patent 3,887,878, June 3, 1975, O. H. Schade, RCA Corp. The somewhat unusual circuit, (Fig. 3 of the Patent document), comprises a phase splitter after Wheatley, RCA Corp., U.S. Patent 3,573,645, April 6, 1971 "Phase Splitting Amplifier", and series connected outputs and drivers. The phase splitting method used leads to a problem maintaining drive bias on high negative signal excursions, and an additional circuit is included to counter this.

[30] "Amplifier Output Stage", U.S. Patent 3,974,456, Aug. 10, 1976, R. W. Russell and K. M. Black, National Semiconductor Corp.

[31] National Semiconductor, Application Note 446B, "A Monolithic Power Op Amp", October 1987, R. J. Widlar and M. Yamatake, covers the operation of the actual IC internal circuit. Application Note 446, "A 150W IC Op Amp Simplifies Design of Power Circuits", March 1986, R. J. Widlar and M. Yamatake, focuses on application and use of the IC. The "LM12 80W Operational Amplifier" data sheet, March 1995, appears to describe a quite different device, lower powered and using a full complimentary output stage.

[32] QSC Audio Products Inc., Costa Mesa, California.

[33] "Audio Power Amplifier Design Handbook", D. Self, Newnes, 1996, ISBN 0750627883, p.101.

[34] 220nF in "An Approach to Audio Amplifier Design", J.R. Stuart, Wireless World, September 1973, p.444 seems a little excessive.

[35] 75W Amplifier, Hi-Fi News and Record Review, J. Linsley Hood, November 1972, pp.2120~2123.

[36] "Solid State Audio Power", J. Linsley Hood, Electronics World and Wireless World, November 1989, p.1044.

[37] "Playmaster Twin Twenty-Five", Leo Simpson, Electronics Australia, April 1976, p.41, and quoted verbatim in Guitar Amplifier For Practice Sessions", John Clarke, Electronics Australia, March 1981,p57.

Simpson quotes Hood's Hi-Fi News and Record Review, Nov. 1972 reference as source material, but then goes on to give an explanation for the function of the capacitor which is at variance with Hood's explanation, certainly as given in the Solid State Audio Power, Nov. 1989 reference. To quote Simpson, "*The 0.022mF capacitor across* [the Baxandall diode] *compensates for the load capacitance in the collector circuit of the* [lower half - PNP - driver transistor] *due to the Miller capacitance of the base-collector junction of the* [lower half output -NPN- transistor]." According to Hood himself, the capacitor is there to "*simulate the effect of the* [missing PNP] *output transistor base/emitter capacitance*".

[38] "Audio Power Amplifier Design Handbook", D. Self, Newnes, 1996, ISBN 0750627883, p.100.

[39] "Playmaster Twin 25", Leo Simpson, Electronics Australia, September 1976, pp.61~62.

[40] "Power Bandwidth Limitations in Audio Amplifiers", J. Alves, IEEE Trans. On Broadcast and TV, March 1973, pp.79~86.

[41] In their June 1980 issue Electronics Australia produced their Playmaster 300W amplifier, by John Clarke and Leo Simpson. It used three MJ15003/ MJ15004 output pairs in complementary symmetry, but contained provision for an all NPN quasi-complimentary version by virtue of a series of circuit configuration altering PCB links; an excellent concept. Performance data comparing the two versions, although referred to in the text, was omitted in the case of the quasi-complimentary version; it would have been interesting had

it been present! So too were appropriate emitter resisters for current sharing, requiring a major correction (Electronics Australia, August 1980, p.141), involving additional links, track cutting and additional components.

[42] Refer Figure 2, National Semiconductor Application Note 446B, October 1987, R.J. Widlar and M. Yamatake, "A Monolithic Power Op Amp".

[43] At very high frequencies ("above 2.5MHz") Q9 appears as simply its base-emitter diode, C2 shunting base-collector. In this condition the Q9 base-emitter diode can be seen to provide no more than a 0.6V level shift to Q8 emitter drive passing directly to Q17 base.