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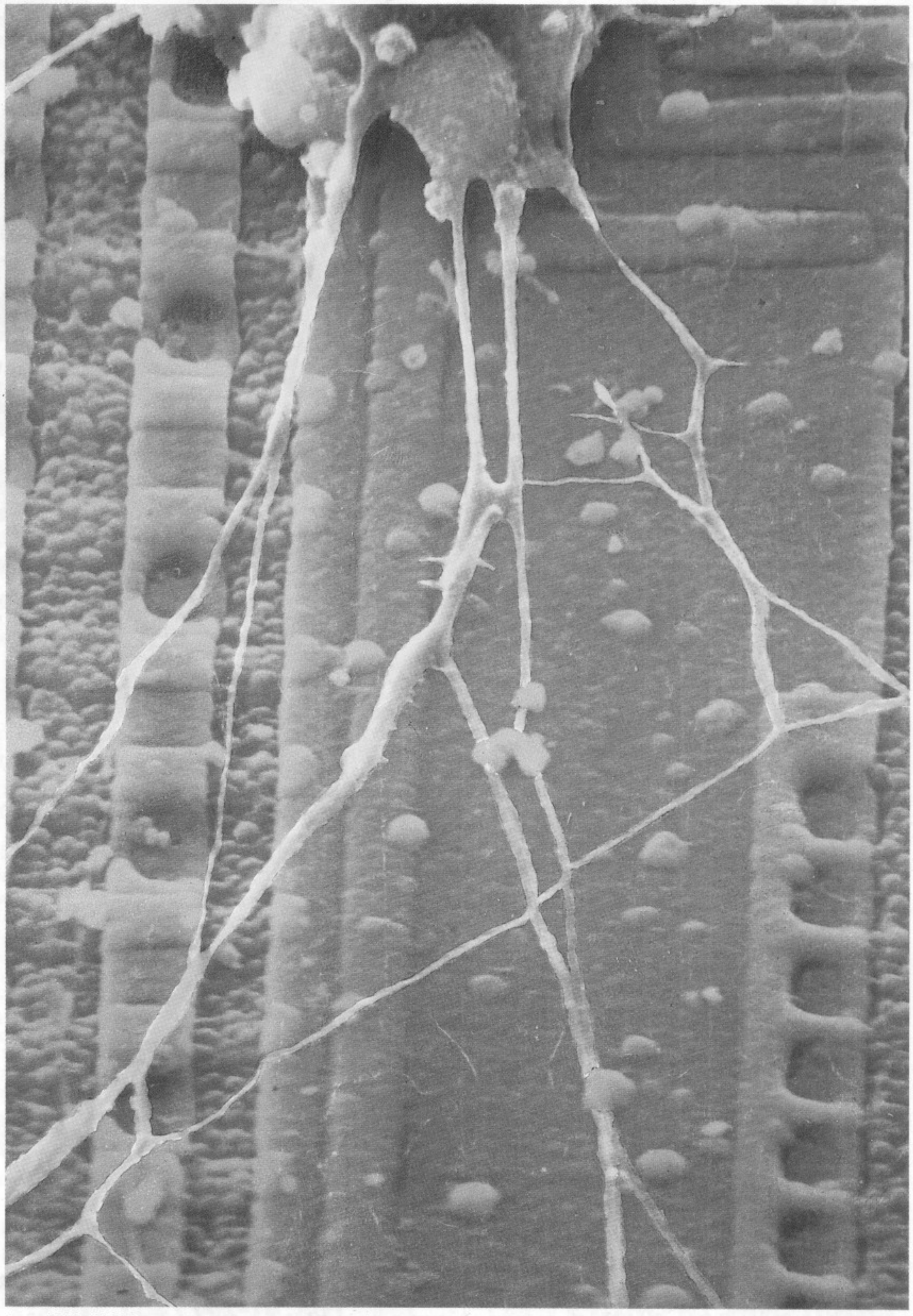

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Crossover distortion in class B amplifiers

Detailed tests on three modes of amplifier operation, including a non-switching class B type, using the same basic circuit produce a few surprises.

E. MARGAN

Ever since the publication of my circuit idea[1] I have received questions as to what extent the circuit was an improvement over the traditional arrangement. Nearly every question emphasized the subjective sensation of distortion reduction achieved[2].

At the time I had measured the circuit performance and found that the circuit definitely posessed certain advantages to justify publication, but enough data was collected to give a precise answer (other than "come and listen for yourselves").

Now I can give some answers regarding electrical performance and offer some hints on what can be perceived. There are still unknowns however and further experiments are encouraged to throw more light on the subject.

MEASUREMENT TECHNIQUES

Historically, crossover distortion was the first distortion mechanism encountered in transistorized audio amplifiers ("transistor sound"). It was considerably reduced by employing the now common bias technique[3] and it is surprising how little attention it has received in literature since. By the discovery and explanation of other distortion mechanisms it eventually faded into the background until the famous feedback vs feedforward error-correction debate, followed by the subjective evaluation debate. Although subjective evaluation was recognised to produce statistically unusable results[4], there remained an impression that not everything could be measured to correlate with the descriptions of what has been perceived by the "golden eared" ones. In fact there are many works that stress the output stage non-linearities as the major source of problems (refs 5,7,8,10,12).

Various methods have been developed to evaluate amplifier distortion. Most of them use steady-state signals to aid analysis. Several forms of distortion, however, have their origin in conditions that are variable by definition ("What have sine waves to do with music?").

Having experimented with different methods I finally decided that the 'subtraction' method[6] could offer most in flexibility as it requires no precision signal reference and can be used with both steady-state and transient signals.

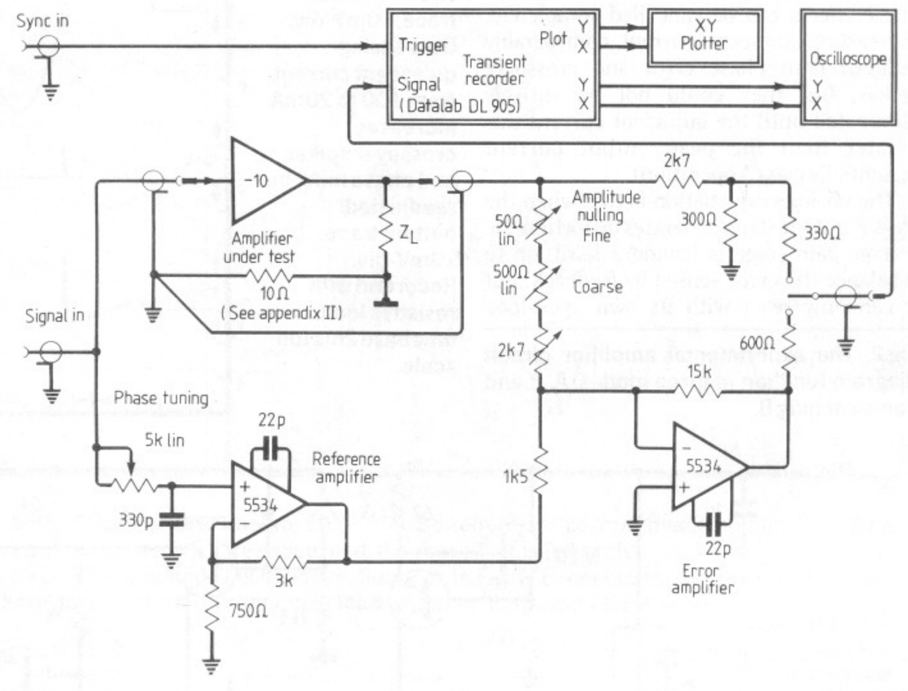


Fig.1. The "subtraction" test set-up due to Baxandall can be used with both steady-state and transient signals and does not require a precision reference.

Fig. 1 shows the test circuit, while Fig. 2 shows the experimental amplifier built to enable comparative measurements between standard class A and class B circuits and the circuit proposed[1] which will be referred to as class NSB (non-switching B). The circuit shown in Fig. 2 has the advantage of using the same devices in all the three modes, thus enabling direct comparison of test results. The front-end was built around a 5534 operational amplifier with open-loop gain of 60dB, unity-gain compensated by 22pF, a 20dB closed-loop gain, 40dB overall feedback and 200kHz closed-loop roll-off. The experimental amplifier performance is compared to another 5534 as its basic performance was considered acceptable by power amplifier standards. This amplifier has also a high-frequency single-pole network which matches its high-frequency roll-off and phase to the experimental amplifier. The output signals of both amplifiers are summed together by a resistive network. Being of opposite phase, the output signals are effectively subtracted, leaving only noise and distortion at the nulling point. A further 5534 is used to prevent nulling point loading by other test equipment and to amplify the error signal (×10) to increase the level as required by the input sensitivity of the test equipment.

Using a sine-wave generator at first to test the class B performance it was noted that if the level of distortion was changed the high-frequency single-pole control had to be readjusted to give minimum output from the subtraction amplifier (Fig. 3). To investigate the pattern of the change the generator was replaced by the circuit shown in Fig. 4, consisting of a square-wave generator driving two tunable band-pass filters both with independent Q-factor adjustment. At high Q settings the filters produce exponentially decaying sine waves simulating a real-life transient (Fig. 5, top trace). Such a signal offers the advantage of looking simultaneously at both distorted and undistorted amplifier response and make direct comparison.

By trying to get minimum output from the error amplifier it was noted that if the phase was nulled in the high signal level region where crossover spikes occured there remained considerable phase error in the low-level region where no switching distortion was being produced (Fig. 5, middle trace). And *vice versa*: if the phase was nulled in the undistorted region a lot of uncompensated phase error appeared, broken by the crossover spikes (Fig. 5, bottom trace). This means that only when the output signal amplitude falls below the level of transistor cut-off both amplitude and phase effects can be cancelled completely. Increasing quiescent current considerably reduced both phase error and crossover spikes, but they could not be entirely eliminated until the quiescent current was greater than the peak output current (essentially class A operation).

The obvious explanation is that when the class B output stage generates distortion the voltage gain stage is having a hard job to rebalance the error sensed by feedback, but it can only react with its own open-loop bandwidth and gain[7]. Also, the output impedance, being not near-zero under distortion conditions, forms an attenuator, together with the load which becomes part of the feedback loop. This is confirmed by Fig. 6 which shows error increase under capacitive load.

The degree and nature of distortion in class B configuration was a real surprise for it was expected that high distortion would show up at levels where the inactive output device becomes unbiased and eventually reverse biased (Fig. 7, top trace). That is also why it was expected that class NSB bias mode (Fig. 7, bottom trace) could be a better solution. In fact, in Fig. 8, where voltages across the output emitter resistances are recorded, the distortion threshold is reached when the output level falls below 100mV (on 4 ohm load, 0.4 ohm emitter resistances and 100 mA quiescent current).

It was also found that distortion falls as the voltage gain stage bandwidth rises, which was expected[8]. This throws a bit more light to schemes of alternative frequency compensation networks[9] and local error correction techniques[10,11]. While those methods reduce the errors considerably, the proposed circuit eliminates them in principle, enabling the overall feedback to be always effective. Comparing the recorded class B performance with class NSB recorded in Figs 9 & 10 shows the distortion generated with class NSB operation is very close to the noise floor. These figures are the same as can be achieved with class A

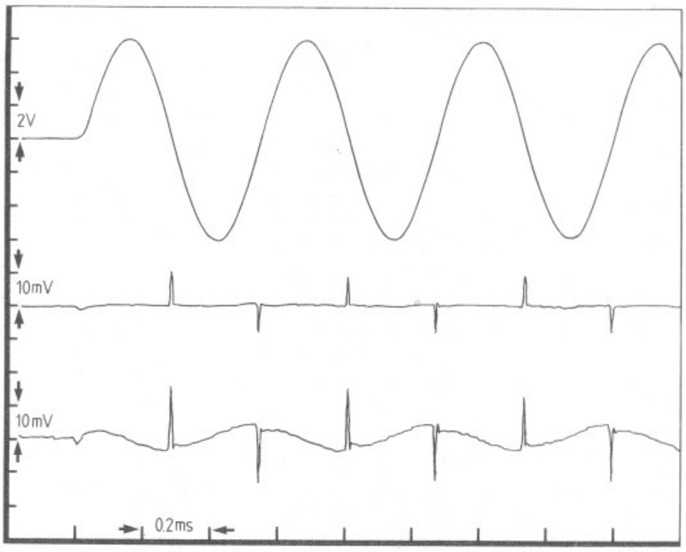


Fig.3. Class B operation: Output signal, top trace, 2V/div. Error signal, middle trace, 10mV/div. Decreasing quiescent current from 100 to 20mA increases crossover spikes and phase must be readjusted: bottom trace, 10mV/div. Recorded with resistive load, timebase 2ms full scale.

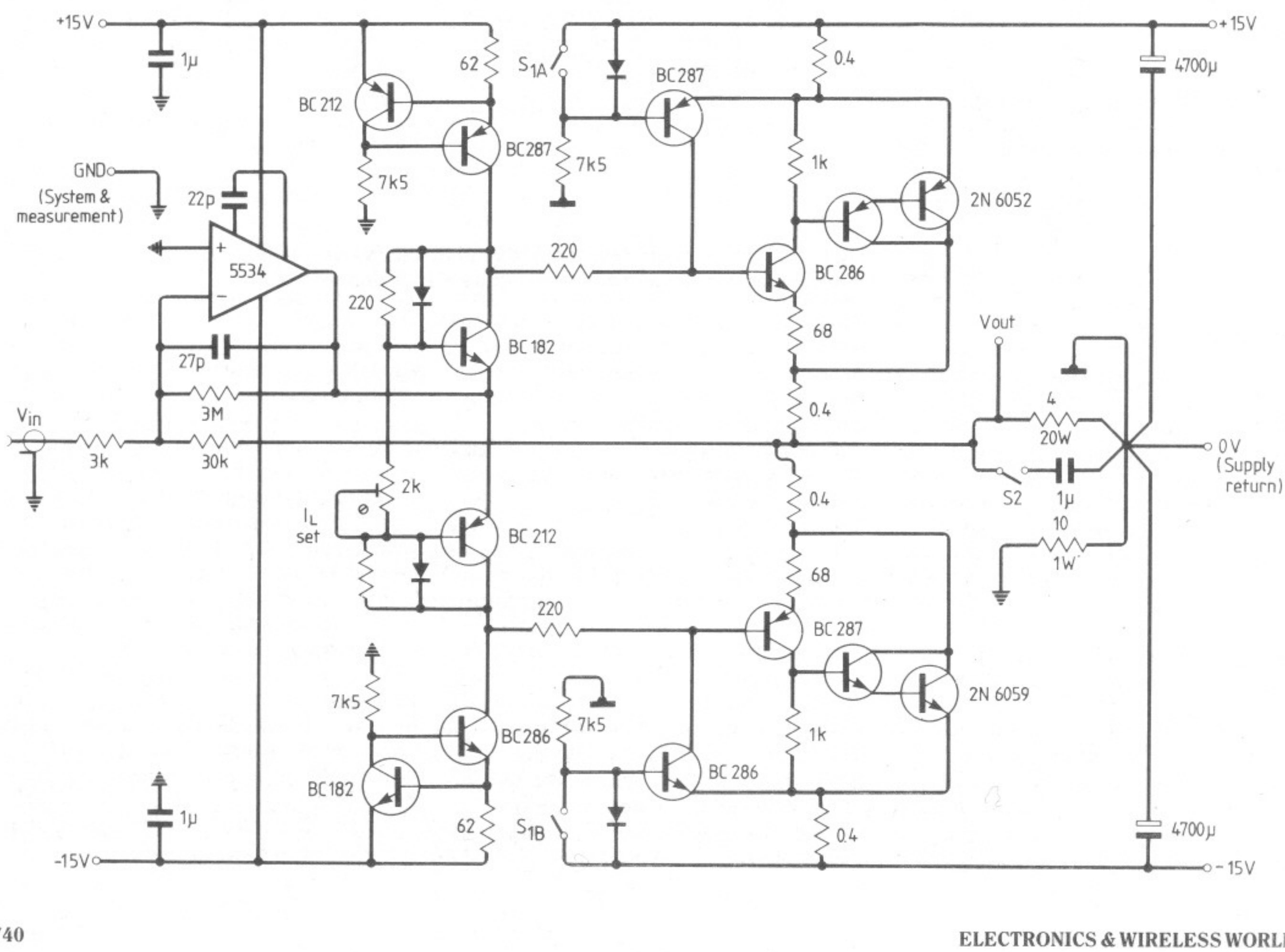


Fig.2. The experimental amplifier circuit diagram function in three modes, A, B and non-switching B.

operation. Such performance speaks for itself.

The proposed class NSB circuit has several distinct advantages over similar circuits presented before[12]. First, it uses only negative feedback (in contrast to positive or combined positive and negative in similar circuits) to sense and prevent switching off (Fig. 7, bottom trace). Secondly, the quiescent current is sensed directly, thus no thermal feedback is needed to achieve thermal stability. Third, thermal stability of the circuit does not rely mainly on high value of emitter degeneration resistance, so those resistances can be made small (less than 0.1 ohm) and so improve output impedance linearity in dependance of output current.

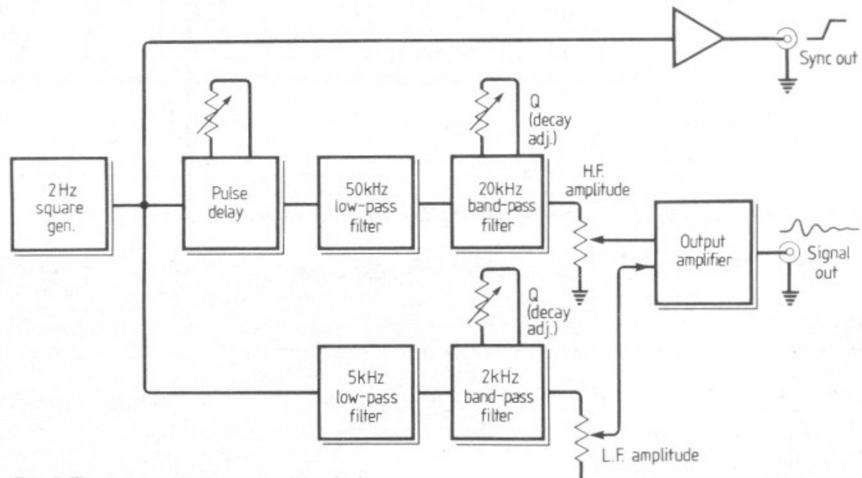
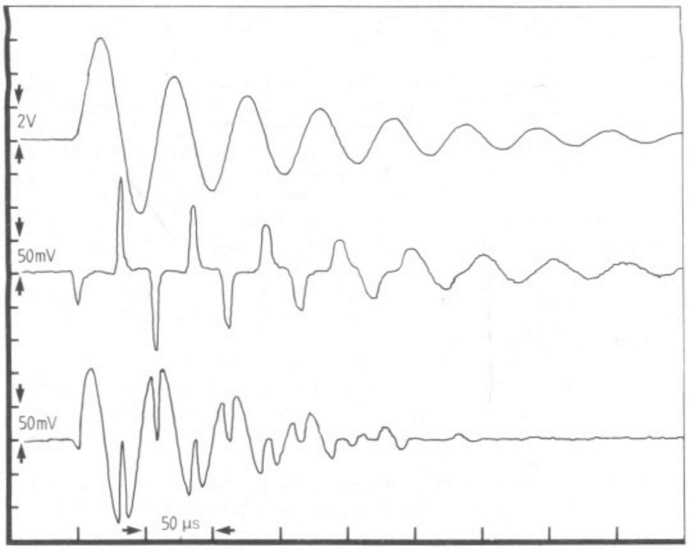


Fig.4. Test signal generator block diagram.



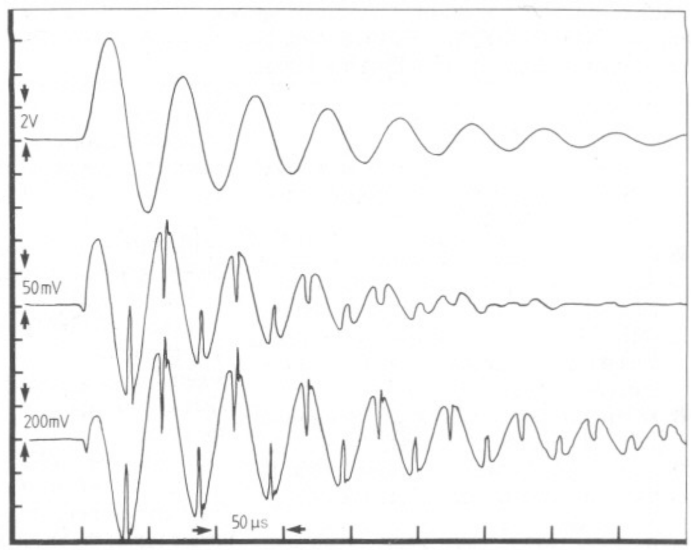
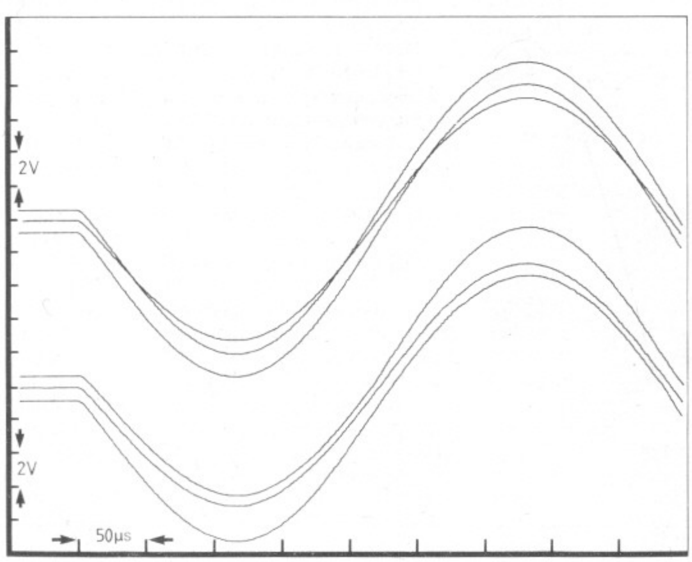


Fig.5. With exponentially decaying sine wave both distorted and undistorted class B response can be compared simultaneously. Top trace: test amplifier output, 2V/div. Middle trace: phase nulled in distorted region shows uncompensated phase in undistorted region, 50mV/div. Bottom trace: correct phase nulling, 50mV/div. Resistive load, timebase 0.5ms full scale.

Fig.6, above. Same as in Fig. 5, except bottom trace recorded with capacitive load, showing increased phase error.



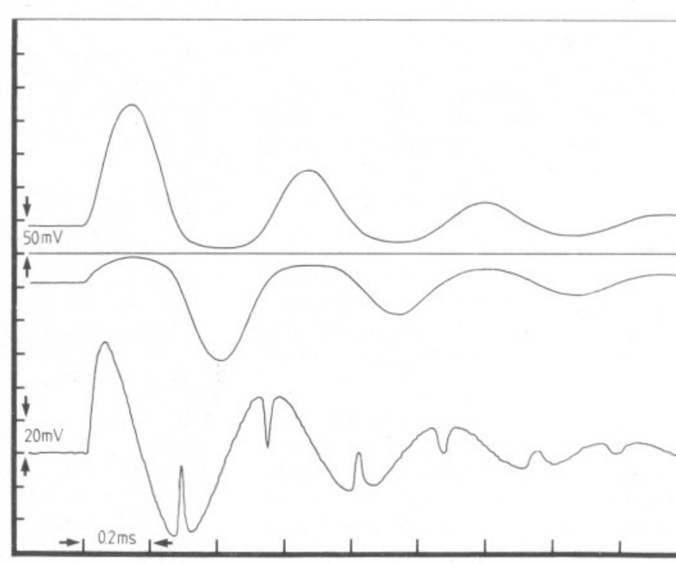


Fig.7. In class B operation the inactive output device becomes reverse biased: the output voltage is compared to the bias voltage under quiescent condition and with input signal applied, top traces, 2V/div. No such condition is allowed in class NSB operation, bottom traces, 2V/div. Resistive load, timebase 0.5ms full scale.

Fig.8. Class B output transistor currents (recorded as voltages across the emitter resistors) with zero current level shown for comparison. Vertical sensitivity 50mV/div., R(e) = 0.4ohm, R(L) = 4.0ohm, I(q) = 100mA. Bottom trace shows the error signal at 20mV/div. Time base: 2ms full scale.

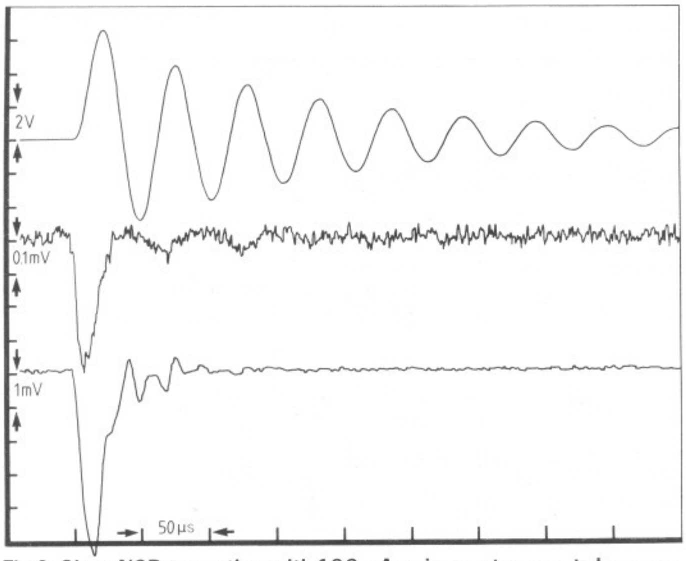


Fig.9. Class NSB operation with 100mA quiescent current shows no trace of phase error and no crossover spikes. Top trace: 2V/div., middle trace: 0.1mV/div., resistive load, bottom trace: 1.0mV/div., capacitive load, timebase: 0.5ms full scale.

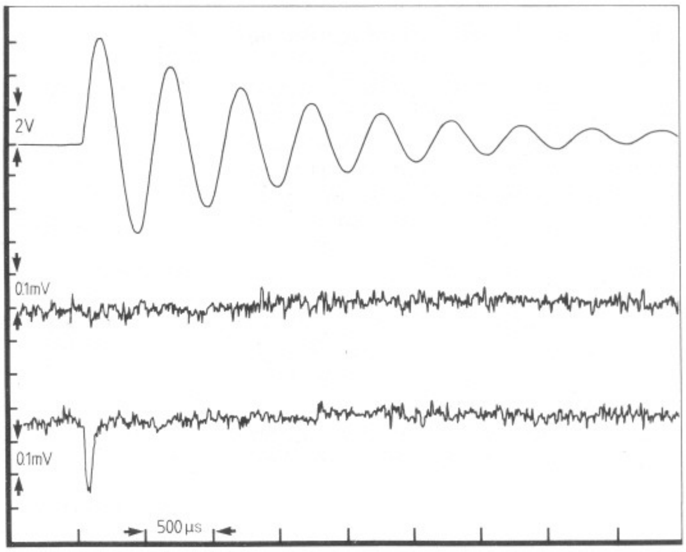


Fig.10. Same as Fig.9 except bottom trace 0.1mV/div. recorded with capacitive load; timebase 5ms full scale.

CONCLUSIONS

Although further investigation is needed several conclusions can be readily drawn from the data presented:

- Class B amplifier generates crossover distortion until the output signal current falls below the level determinated by the ratio of load impedance to emitter degeneration resistances and quiescent current setting.
- When crossover spikes are present a phase error is also generated[7].
- Phase error is inversely proportional to the open-loop bandwidth of the voltage gain stage.
- Phase error is also dependent on the ratio of the amplifier output impedance to the load impedance.
- The envelope of the phase error signal stays in fixed proportion to the output signal envelope until a threshold is reached and it suddenly disappears (switching phase modulation).
- In complex signals the individual components are differently affected: the higher the frequency, the greater the phase error. This, and the previous point mean that phase coherence is lost during that part of a musical signal which bears dominant localization and 'definition' information.
- A cost and bias level compromise combined with thermal stability requirement has forced many designers of commercially equipment to underbias the output stage (for comparison, see ref. 8).

As a consequence, comparing amplifiers of different design shows differences that depend on open-loop bandwidth and gain as well as on output impedance and quiescent current setting. Also, using the same loudspeaker load doesn't guarantee freedom from load-induced differences.

Regarding audibility of the described performance bear in mind that many subjective evaluation sessions have reported objections which could be attributed to lost phase coherence. Unfortunately, I have no means of performing a well-controlled listening session; someone with more experience in this field is invited to contribute. Of course, when listening to some digitally recorded piano I could definitely express my preference for classes A and NSB performance, even though not belonging to the "golden eared" category, but the opinion of a single person (and a strongly biased one who also knows what to listen for) can hardly have statistical meaning.

In fact, the phase errors recorded are of the order of 0.05 degrees at middle frequencies increasing up to several degrees at the top of the audio range, measured with a resistive load. Reactive loads and/or reduced open-loop bandwidth produces even greater phase error but to standardize measurement a reference reactive load and bandwidth are required to be defined.

But equally important, a statistically meaningful definition of the audibility threshold to switching phase modulation would be welcome. Only in regard to this threshold can the data presented here undergo relevant evaluation.

Finally, it has been demonstrated that both class A and class NSB are free from the effect described, thus highlighting the inherent quality of the NSB principle as a solution for crossover distortion.

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Appendix 1 - Phase error calculation

The error signal undergoes phase modulation during the first rising edge of the output signal, Fig 5, whereupon a fixed phase relationship is established. If we neglect the exponential amplitude decay term as it is present in both the input and output signal as well as in the error signal, and label the input signal as sine, then the error signal is clearly a cosine. Looking at the amplitude nulling network under the correct nulling condition:

reference signal	$x = A \sin \omega t$
output signal	$y = B \sin (\omega t + e)$
error signal	$z = C \cos \omega t$

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OPTIMUM RESONATOR IMPEDANCE

The oscillator phase noise is inversely proportional to the resonator power, which in turn is inversely proportional to the resonator impedance, which should thus be made as low as possible. There is, however, a bound below which decreasing the resonator impedance does not improve any more or even deteriorates the oscillator c.n.r. This is caused by several factors:

- At too low a value of the resonator impedance the oscillator transistors can no longer provide sufficient output power.
- The l/f corner frequency of transistors increases with drain or collector current.
- Due to the base resistance of bipolar transistors, the equivalent input noise voltage cannot be decreased below a certain threshold and will even increase when the collector current gets too high.

Then there is the practical problem of decreasing the resonator impedance while keeping a sufficient electronic tuning range because the capacitance variation of v.h.f. varactors is limited. The impedance could be transformed downward by coupling the resonator via a tap to the oscillator circuit. This increases, however, the r.f. voltage to the varactors and so the a.m.-to-p.m. conversion.

CIRCUIT DIAGRAM

The circuit diagram of Fig.5 needs little explanation. Oscillation frequency is determined by L_3, C_3, C_4, D_1 and D_2. Resistors R_1 and R_2 provide low frequency feedback to reduce the l/f noise and the input offset voltage of the differential amplifier. Their optimum value is a compromise between l.f. feedback and amplitude control range. Coupling capacitor C_2 has a low impedance at the oscillation frequency, but forms practically an open circuit at low frequencies. Components L_1, L_2 and C_5 prevent the r.f. noise of Tr_5 entering the oscillator and C_6 shorts the transistor l.f. noise. The low-frequency decoupling capacitors are connected to the positive supply voltage to prevent modulation of the drain gate capacitances by noise on the power supply.

Capacitor values of C_1 and C_4 are of the order of some pF. They balance the voltages across the gate source junctions of the differential amplifier and across the varactors, to reduce a.m.-to-p.m. conversion. Besides, C_1 improves the frequency response of the amplifier.

To change frequency by a large amount L_3, R_1 and R_2 are the only components that need to be replaced. If the electronic tuning sensitivity is not critical, the oscillator can be tuned with C_3 over more than one octave.

PHASE NOISE MEASUREMENTS

The phase noise of the oscillator of Fig.5 was measured using several types of transistors in the differential amplifier. The place of the tap on the resonator coil and the amplifier tail current were experimentally optimized for lowest phase noise. The results are shown in the table.

The Table shows that the lowest phase noise

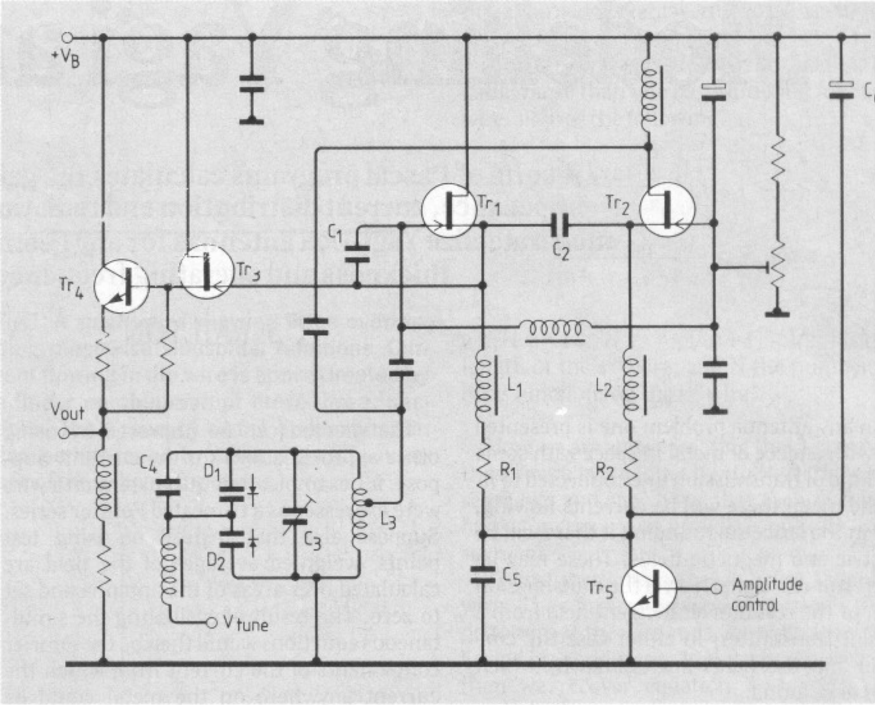

Fig.5. To change frequency by a logic current, the only components to be charged are L_3 and R_1, R_2. As it stands, it can be tuned over an octave with C_3.

can be obtained with junction fets. Besides, the phase noise of an oscillator with fets turned out to be less sensitive to the place of the resonator tap and the amplifier tail current.

TABLE 1. Lowest phase noise obtained at oscillation frequency of 100MHz at 5kHz from the carrier for several types of transistors in the differential amplifier.

Transistor	Remarks	$\mathcal{L}$(5kHz) (dBc/Hz)
2N3823	v.h.f junction fet	−116
BF198	v.h.f. bipolar	−106
CA3127	v.h.f. bipolar array	−105
BCY59C	low l/f noise bipolar	−91
BFR90	low noise wideband	−80
BFQ69	low noise wideband	−70

The phase noise of the oscillator with the 2N3823, BF198 and CA3127 was close to the value estimated with equations 2 and 3. The high noise of the BCY59C is probably due to its high base resistance. The extremely high phase noise obtained with the BFR90 and BFQ69 is caused by a.m.-to-p.m. conversion of amplitude noise generated by l/f noise. By increasing the speed of the amplitude control circuit, the phase noise was reduced to −90dBc/Hz.

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A. P. Dekker is now with Nokia Telecommunications, Espoo, Finland, having completed this work at the Dr Neher Laboratories of the Dutch PTT, Leidschendam.

Non-switching class B amplification

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Under nulling condition, $y = x + z$, and after elementary trigonometry:

$$B^2 = A^2 + C^2$$

$$e = \arctan (C/A).$$

Now as only the output and the error signal are recorded, substitute $A^2 = B^2 - C^2$. So finally

$$e = \arctan \frac{C}{\sqrt{B^2 - C^2}}$$

Appendix 2 – Experimental amplifier

To eliminate influence from other distortion mechanisms as much as possible careful circuit design and layout were needed. Separate power supplies for the voltage gain stage and power stage were used and high current ground was separated from the signal ground (see ref. 13). Test equipment ground was connected to the input signal ground except for Fig. 8 recording, where the whole test set-up was floating and the test amplifier "live" output was connected to the transient recorder ground.

The voltage gain stage was designed around a high performance operational amplifier to simplify control over the open-loop bandwidth and gain, which were chosen to represent typical values found in modern power amplifiers.

Bias control was designed to be variable over a wide range to achieve requirements for class B and class A bias. The circuit must behave as a symmetrical low-impedance voltage source for the output stage. Class NSB operation was achieved by opening the contacts of S1a and S1b and adjusting the quiescent current to the same value as the one chosen for class B operation. Class A quiescent current was set to three amps.