

Current mirrors, amplifiers and dumpers

Improving the performance and application of the basic circuit

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The accuracy of a two-transistor current mirror circuit can be greatly improved by the addition of a further two transistors. The resulting four transistor mirror can be used to design simple low-distortion operational-amplifier circuits that produce an output current proportional to either input voltage (v.c.c.s.) or input current (c.c.c.s.). In addition, they make possible the design of "current-dumping" amplifiers where the output current is controlled by a pair of unbiased transistors, operating entirely in Class B with the crossover distortion eliminated by a feedforward amplifier using current mirrors.

The simple two-transistor current mirror in Figure 1 attempts to produce at its output B an identical copy of the input current at A, whilst minimizing unwanted current-voltage interactions. Its operation can be easily understood by considering the input transistor as a collector-base connected diode, driving an output transistor with a matched V_{BE} to produce an identical collector output current. The basic mathematics of its operation were described recently and will not be repeated here¹. Figure 2 shows the symbol often used to signify a current mirror, indicating by an arrow both the polarity of the current and the input side of the mirror. It should be remembered that, due to the circuit topology, the input terminal will always remain at a fixed voltage, in contrast to the output terminal which will take up a voltage determined by the load conditions.

The current transfer ratio I_o/I_{in} , usually termed λ , is normally the most important parameter when using current mirrors. It is obviously desirable that λ should be constant, irrespective of changes in current and output voltage. (Whilst most current mirrors are intended for operation with a unity value of λ they can be designed for other integral values by duplicating transistors accordingly.)

Unfortunately, the performance of the two-transistor mirror is often inadequate, largely due to the high dependence of λ on the values of the transistor parameters in such a simple, uncompensated circuit. It can be shown², by considering basic transistor operation, that the departure from unity current transfer ratio for a two-transistor mirror can be represented by:

where β is the common-emitter current gain, V_{OS} is the difference in base-emitter voltage required to produce identical collector currents, V_T is the thermal voltage $\approx 25mV$, $V_{\Delta Q}$ is the difference in collector-base voltages of the two transistors and $(V_{I})_Q$ is the Early intercept voltage at the operating point Q^* .

$$\lambda_2 = 1 \pm (2/\beta) \pm (V_{OS}/V_T) - V_{\Delta Q}/(V_I)_Q$$

Figure 1 shows the basic two-transistor current mirror circuit. The input current I_{in} flows into terminal A, through the collector-base junction of transistor Tr_1 . The base-emitter junction of Tr_1 is connected to the base-emitter junction of transistor Tr_2 . The output current I_{out} flows out of terminal B, through the collector of Tr_2 . The base of Tr_2 is connected to the emitter of Tr_1 . The emitter of Tr_1 is connected to the emitter of Tr_2 , which is connected to the common terminal C.

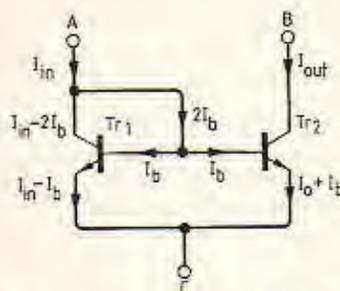


Fig. 1. Basic, two-transistor, n-p-n current mirror

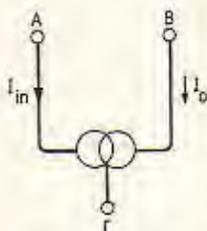
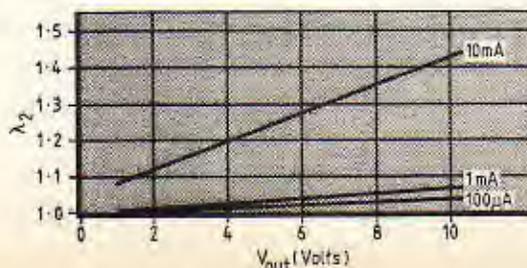


Fig. 2. Shorthand symbol for circuit of Fig. 1.

Fig. 3. Accuracy of current transfer between input and output depends on output voltage and output current. Ratio I_o/I_{in} is plotted here for currents up to 10mA at up to 10V.



The β term arises due to the effects of base current in an asymmetrical circuit with the V_{OS} term being due to the mismatch in the transistors' base-emitter voltages. The contribution of the Early intercept voltage is best described as being due to the slope in the transistor I_C vs. V_{CB} characteristics. Of course all these terms are dependent on current or temperature, making a general analytical evaluation quite difficult! Figure 3 illustrates the results obtained when using an RCA CA3096AE transistor array, connected as a two-transistor mirror and operating at currents of 100uA, 1mA and 10mA. Typical values for the n-p-n transistors in the RCA array are: $\beta=200$, $V_{OS}=0.3mV$ and $(V_I)_Q = 100V$, producing error components of around 1%, 1% and 1-5% respectively for the three contributions.

Clearly, the accuracy of the current mirror action for a two-transistor mirror is not very good, degenerating progressively

* The Early intercept voltage is the intercept of the tangent to the I_C vs. V_{CB} curve projected backwards to the $-V_{CB}$ axis. It is therefore dependent on the operating point of the transistor.

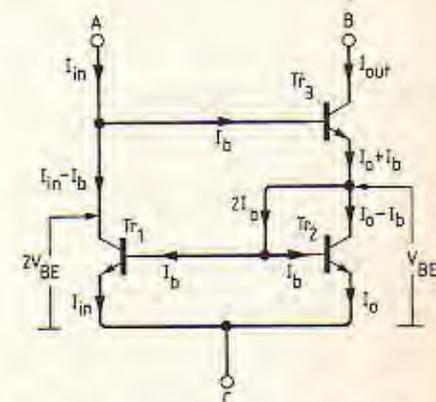


Fig. 4. Addition of Tr_3 helps to isolate Tr_2 from output voltage changes.

above a milliamp. For p-n-p transistors the situation is even worse, because β is very sensitive to collector current for p-n-p planar transistors, falling to extremely low values (≈ 10) at currents above several milliamperes. The uncertainty due to V_{OS} , however, is slightly reduced, since in general p-n-p transistors have tighter V_{BE} matching.

The performance of a two-transistor mirror can be greatly improved by the addition of a third transistor, as in Figure 4, resulting in the standard Wilson current mirror. The third transistor Tr_3 fulfils two roles; the first of which is to buffer Tr_2 from changes in collector voltage and remove to a large extent the voltage sensitive component in the current transfer ratio λ . Changes of collector voltage have much less effect on Tr_3 because it is effectively current driven from its emitter. The second improvement arises from the redistribution of base currents within the circuit, bringing the current-transfer ratio much nearer to unity. Figure 4 shows that, to a second-order approximation, the input and output currents are now equal. In a similar fashion to Equation 1, the current-transfer ratio for a three-transistor mirror can be represented by:

$$\lambda_3 = 1 \pm 2(\Delta\beta/\beta^2) \pm (V_{OS}/V_T) - V_{BE}/(V_1)_{0.7}$$

where $\bar{\beta}$ is the mean of the transistor current gains, $\Delta\beta$ represents the spread of β values for the three transistors and $(V_1)_{0.7}$ is the Early intercept voltage evaluated at a V_{CB} operating point of approximately 0.7V, as this is the difference between the collector voltages of Tr_1 and Tr_2 in a three-transistor mirror circuit. The improvement in the current-transfer ratio in this equation is largely due to a reduced dependence on β and the small voltage difference ($\approx V_{BE}$) between Tr_1 and Tr_2 . A spread of $\pm 20\%$ in current gains for the three transistors in the mirror would produce error components of $\pm 0.2\%$, $\pm 1\%$ and -1% or, overall, approximately $+0$ to -2% tolerance. Texas Instruments have recently introduced monolithic three-transistor Wilson current mirrors exhibiting a current transfer ratio accurate to within 1% of unity up to a milliamp, with a voltage capability of 35V (TL 011). Also, by paralleling transistors within the mirrors they have produced circuits displaying halving, doubling and quadrupling functions (TL 021, TL 012 and TL 014).

Further improvements in mirror performance can be obtained by the introduction of a fourth transistor to equalize the collector voltages of Tr_1 and Tr_2 , as shown in Fig. 5. Note that the same symbol can be used to represent current mirrors, irrespective of the number of transistors used. The only errors remaining now are due to finite β and base-emitter voltage differences, giving:

$$\lambda_4 = 1 \pm 2(\Delta\beta/\beta^2) \pm (V_{OS}/V_T)$$

producing, typically, for the CA3096AE array:

$$\lambda_4 = 1 \pm 0.1\% \pm 1.0\% \approx 1 \pm 1\%$$

A comparison between the three- and four-transistor mirrors is given in Fig. 6.

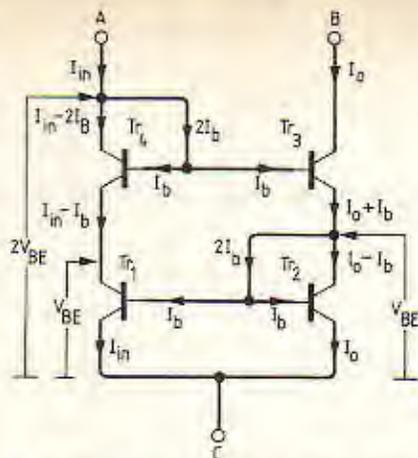


Fig. 5. Fourth transistor equalizes input and output collector voltages, further reducing unbalancing influences.

RCA CA3096AE transistor array			
I_{in}	V_o	Transfer ratio	
		λ_3	λ_4
100 μ A	2V	0.995	1.001
	10V	0.996	1.001
1 mA	2V	0.990	0.999
	10V	0.991	1.000
10 mA	2V	0.886	0.991
	10V	0.890	0.994

Fig. 6. Table shows improvement in tolerance to current and voltage variations between circuit of Fig. 4 and that of Fig. 5.

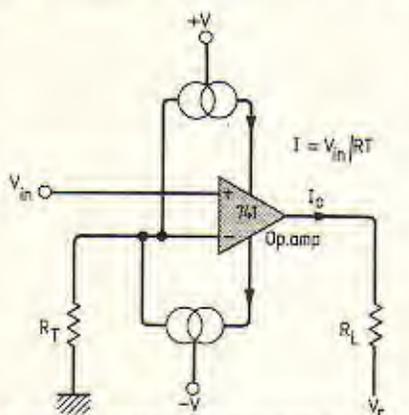


Fig. 7. Voltage-controlled current source. I_o is proportional to V_{in} .

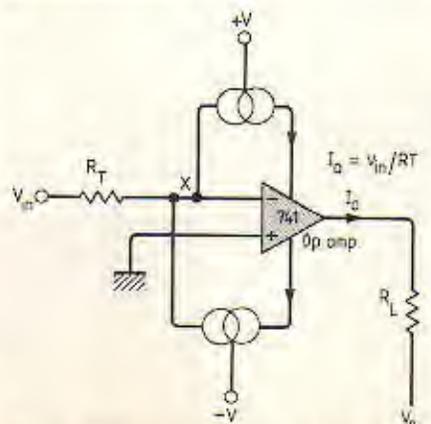


Fig. 8. Shunt feedback instead of the series type in Fig. 7 produces inverting v.c.c.s.

The two sets of results were taken from the circuit of Fig. 5, with the currents measured directly by 4½-digit digital meters. Transistor Tr_4 was then shorted out to obtain the results for a three-transistor mirror. In both cases it can be seen that the current-transfer ratios are held very constant against output voltage changes. The removal of the Early intercept voltage error component (approximately -1%) from the four-transistor circuit is evident. In addition, the current transfer ratio is maintained to higher current levels because of the increased β buffering action with the four-transistor mirror. At 10mA it is still within 1% of unity, whereas the three-transistor version has fallen to approximately 90%. These factors make the four-transistor modified Wilson mirror the best choice for circuit designs, both discrete and monolithic. For precision circuits MAT 01 AH matched transistor pairs (Precision Monolithics) can be used for Tr_1 Tr_2 and Tr_3 Tr_4 to give a current-transfer ratio of unity to within 0.4%, due mainly to their very close V_{BE} matching.

Current mirror applications

In many applications it is desirable to control the output current rather than the output voltage of a circuit, especially when driving reactive loads or current-activated transducers. For example, a controlled current is required to produce a defined magnetic field from an inductive coil. It is not always feasible to voltage drive the load through a high-values series resistor, particularly if a significant back e.m.f. is generated. (An appropriate example could be that of a recording head for magnetic tape and cassettes.)

Unfortunately, all the standard textbook circuits for producing controlled bipolar output currents from ordinary operational amplifiers using grounded sources and loads suffer from serious practical problems, usually due to the extremely tight matching required for the resistors controlling the balance of negative and positive feedback³. Circuits requiring non-critical resistor matching that produce superior results can be designed using four-transistor current mirrors. Both transconductance and current amplifier configurations are possible, normally termed voltage-controlled current sources (v.c.c.s.) and current controlled current sources (c.c.c.s.) respectively.

Figure 7 shows the circuit of a bipolar transconductance amplifier (v.c.c.s.) using both n-p-n and p-n-p current mirrors where the output will be proportional to the input voltage. The RCA CA 3096 AE transistor array contains three n-p-n and two p-n-p transistors, which means that two arrays are required to construct a positive and negative four-transistor current mirror pair. The current mirrors are used to sense the operational amplifier's supply currents which, apart from the nearly constant bias currents, are proportional to the output current⁴. A copy of the output current, whether positive or negative, is thus fed back to the inverting input terminal to be compared with the input voltage.

This forces the op.-amp. to generate an output current equivalent to the input voltage V_{in} divided by the transconductance gain setting resistor R_T . Output currents up to 20 mA pk-pk can be obtained with very low distortion independent of the output voltage. Below 1mA the harmonic distortion, mainly second harmonic, is almost constant at 0.03%, rising to 1% at 20mA. It is not necessary with this type of circuit to return the load resistor to ground: it can be terminated on any voltage as long as the resulting load voltage excursions are within the capability of the op.-amp. and the voltage supplies. The recommended op.-amp. frequency compensation should be followed, remembering that for a transconductance amplifier the equivalent voltage gain is given by R_L divided by R_T . Care must be taken when using high values of R_T (equivalent to a low transconductance gain) to ensure that adequate compensation is provided for the op.-amp., since the resulting voltage gain can turn out to be surprisingly low. The circuit can be treated as an ordinary operational amplifier circuit with a slightly restricted bandwidth caused by the shortfall in gain-bandwidth product of the p-n-p transistors in the RCA array. Any op.-amp. similar to a 741 or 301A can be used.

An inverting circuit can be obtained by changing the series-feed back connexion to the shunt-feedback arrangement of Fig. 8. Now the feedback current is balanced against the input current in R_T produced by the input voltage V_{in} : the inverting transconductance gain is still given by $1/R_T$. The distortion figures are marginally superior to the series-feedback case, since there is no voltage excursion

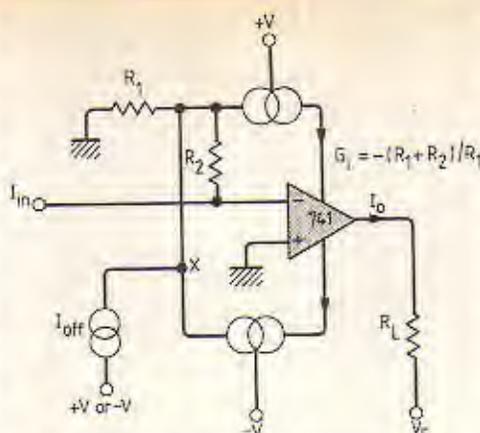


Fig. 9. Attenuating current feedback introduces gain into current-controlled current source, which is similar to v.c.c.s. but without input resistor R_T .

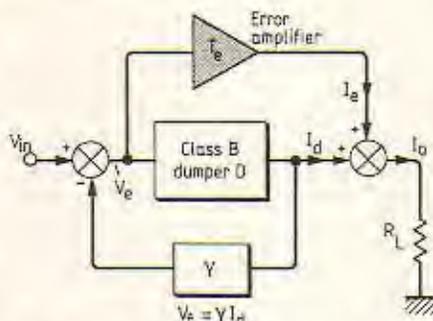
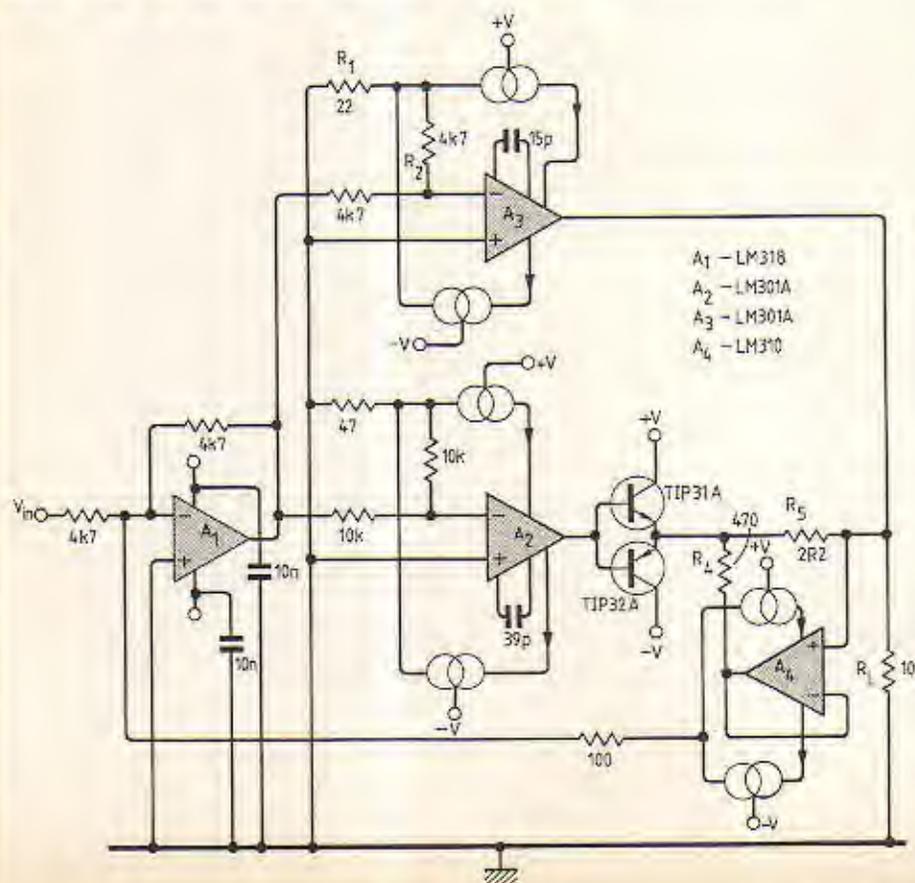


Fig. 10. Transconductance amplifier with feedback and error feedforward.

Fig. 11. Practical 1A Class B current-dumping v.c.c.s.



A1 - LM318
A2 - LM301A
A3 - LM301A
A4 - LM310

whatsoever at the virtual earth connexion⁵.

This topology also presents an opportunity for the design of a current amplifier (c.c.c.s.) simply by removing the input resistor, leaving an amplifier with 100% negative shunt feedback derived from the output current. Gain can be introduced into the circuit by attenuating the feedback current before it is summed at the op.-amp. input. The circuit of the bipolar current amplifier in Fig. 9 uses two resistors to produce the required current attenuation in a manner analogous to a potential divider. The current gain is then defined simply by:

$$G_1 = -(R_1 + R_2 / R_L)$$

Measurement of the input impedance of the circuit of Fig. 9 with a gain of 20 indicates 1Ω at 100Hz, rising to 25Ω at 10kHz. The output impedance varies in the opposite manner, being $150k\Omega$ at 100Hz dropping to $25k\Omega$ at 10 kHz. The output impedance figures could be improved if manufacturers provided a range op.-amps. with alternative output stages in place of the voltage output stages presently used.

In contrast to voltage-controlled circuits, current amplifiers are required to operate from high source impedances and into low load impedances. It is still desirable to null the op.-amp. input offset voltage for critical work to maintain a low output offset current for lower values of source impedance. The Fig. 9 circuit produces an output offset current of around $10\mu A$ with the input open circuit and the op.-amp. input nulled to better than a millivolt. This offset current, caused largely by the affects of op.-amp. bias currents being reflected through the current mirrors, can be drastically reduced by connecting an equivalent bleed current to the output of the current mirrors, point X in Figs 8 and 9. A single resistor to whichever supply rail is indicated will perform the task adequately. The most convenient method of determining the output offset current is by using a digital voltmeter to monitor the output voltage across a temporary high-valued load resistor. An output offset of less than $50nA$ can be easily obtained after adjustment. In this respect, current output amplifiers can be more accurate than voltage amplifiers since, under most conditions, their output offset signal represents a smaller fraction of their maximum output.

Current amplifier using error feedforward

The three previous designs, whilst being extremely useful at low currents, cannot readily be extended to high currents because of the restricted current handling capacity of the transistor arrays forming the mirrors. Class AB current boosters could be used but their well known thermal limitations make it desirable to operate a high-current output stage completely in Class B where there are no critical bias adjustments. Unfortunately, the crossover distortion produced by Class B output stages has traditionally made them unsuitable for applications requiring precision low-distortion waveform reproduction.

However, the technique of error feedforward around a Class B output stage, often referred to as "current dumping", previously employed for a voltage power amplifier⁶, can be applied to current output amplifiers with very good results⁷.

An outline of the proposed method is shown in Fig. 10. A feedback voltage is derived directly from the Class B dumper output current and compared to the input voltage of the system. The resulting error voltage drives both the dumper pre-amp and the error feedforward amplifier. By choosing a suitable gain for the error amplifier any non-linearities in the gain of the dumper and its pre-amp can be compensated by the amplified error signal added at the output connexion. The relevant equations for the sub-units are:

$$\begin{aligned} I_o &= I_d + I_e \\ V_f &= \gamma \cdot I_d \\ V_e &= V_{in} - V_f \\ I_e &= T_e \cdot V_e \\ I_d &= V_e \cdot D \end{aligned}$$

From these equations it can be shown that:

$$I_o = V_{in} \cdot T_e (1 + D/T_e) / (1 + \gamma \cdot D)$$

This equation can be made insensitive to D and its variations (non-linearities) by setting:

$$\gamma \cdot T_e = 1$$

The balance equation indicates that if the transconductance of the feedback network γ^{-1} is made equal to the transconductance gain T_e of the forward error loop, then the gain of the system becomes insensitive to non-linearities within the Class B output stage and its pre-amp. The ratio of current contributions from the Class B dumper and the error amplifier is determined by the ratio of their transconductance gains. By a suitable choice of open-loop gain and feedback factors it can be arranged that the error amplifier normally supplies only a small proportion of the output current, except during the crossover period of the dumper transistors when there is no feedback signal, and the error amplifier supplies all the output current. The transconductance of the system at balance is given by the transconductance of the error feedforward amplifier alone. The overall result of this is ideally zero distortion at the balance condition. However, in practice, the error amplifier and the floating current monitor A_4 contribute their own distortion, but this is quite small, since they only operate at low currents.

One possible circuit for the combined feedforward/feedback approach is shown in Fig. 11. The error feedforward amplifier A_3 and the dumper pre-amplifier A_2 , intended for 25mA pk-pk maximum output, use four-transistor mirrors as previously described. The non-linear dumper consists simply of a pair of unbiased power transistors. A fractional copy of the dumper output current is obtained by A_4 and returned to the input summing

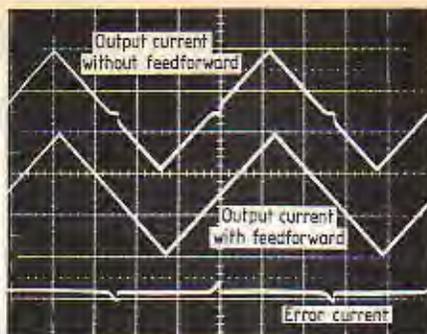


Fig. 12. Triangular wave at 2kHz with and without feedforward.

amplifier A_1 . The feedback factor R_4/R_5 is set equal to the forward error gain $(R_2 + R_1)/R_1$ to satisfy the balance condition.

The upper trace of Fig. 12 shows a 2kHz triangular voltage waveform across the 10Ω load resistor when the feedforward is disconnected, whilst the middle trace shows the effects of adding in the feedforward error at the output connexion. The error-cancelling effects of the balance condition can be clearly seen, there being no discernible disturbance in the linear waveform. The bottom trace shows the error current measured across a separate 10Ω resistor for comparison. Output currents up to 1Apk-pk. can be obtained with this circuit, although the photographs were taken at a low current (15mApk-pk.) where the effects of crossover distortion are more noticeable.

Distortion measurements indicate that the second harmonic is 70dB below the output at 100mA pk-pk., rising by approximately 10dB at 10mApk-pk. and 1Apk-pk. The third harmonic is also lowest at around 100mApk-pk., being 85dB below the output, rising to 75dB at 10mApk-pk. and 80dB at 1Apk-pk. Second-harmonic distortion is generated by the current mirrors in the error feedforward amplifier and the dumper current monitor, whereas the third harmonic is produced by the crossover behaviour of the dumper. Higher harmonics are also present, but are signifi-

cantly below the level of the second and third under similar conditions. Disconnecting the error feedforward loop increases both the second and third harmonic distortion by around 30dB in the critical low-level output region. The relative improvement in distortion performance due to the feedforward connexion is maintained at higher frequencies where the effects of uncompensated crossover distortion become more significant. A further reduction in distortion would require a specially optimized feedforward amplifier and current monitor using discrete components.

Current mirror circuits offer a versatile design tool that can be employed in most applications where a controlled current is required. In conjunction with op-amp supply current sensing they facilitate the design of a wide range of low-distortion transconductance and current amplifiers.

Literature Received

Six-page colour brochure from Crow of Reading gives an outline of the company's activities in the field of broadcast television engineering, which extends from the supply and installation of a single monitor to the design, construction and commissioning of large studios and switching centres. Brochure can be had from Crow at PO Box 36, Reading, Berks, RG1 2NB
WW401

Important characteristics and application information on a range of p.c.o.m.s and similar devices from a number of manufacturers is presented in convenient form on a wallchart, available from Microsystem Services, Duke Street, High Wycombe Bucks.
WW403

Small tools for use in the production of electronic equipment - wire strippers and cutters, board assembly tools and p.c.b. cleaning brushes - are featured in a leaflet published by Braser International Ltd, Unit A, Portway Industrial Estate, Andover SP10 3LU.
WW404

An extremely wide range of microwave aerials, cables and waveguides is fully covered in a weighty catalogue (around 200 pages) which can be had from Andrew Antennas, Lochgelly, Fife, KY5 9HG.
WW405

A range of silicon controlled rectifiers and triacs made by TAG Semiconductors is listed in a selection guide, with main characteristics and a cross reference to other makes. The guide is obtainable from TAG Semiconductors Ltd, 73/79 Rochester Row, London SW1P 2NX.
WW406

Publication HCG 1 from Highland describes the types of multiway connector currently available. Heavy and light-duty types are made, with from 2 to 128 poles and in ratings from 8A 250V to 35A 440V. Highland Electronics Ltd, Highland House, 8 Old Steine, Brighton, BN1 1EJ.
WW407

Large colour catalogue from Ross illustrates a very wide range of audio equipment and accessories, including headphones, test gear, intercom, audio and video leads and adapters and microphones. Ross Electronics, 49/53 Pancras Road, London NW1 2QB.
WW408

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