Current differencing amplifiers

Three sets of cards deal with current differencing amplifiers of the LM3900 kind. This set covers signal processing applications, set 17 covers signal generation and a third set deals with various other circuits including test, measurement, detection, logic and driving circuits.

Typical performance
Supply: 15V
R₂: 5kΩ
Voltage swing: 2,800 (69dB)
Output swing: 0.1 to 14.2V
Output current: source 10mA, sink 1.3mA
(overdriving inverting input increases sink current up to >30mA)
Input current: 30nA
Unity-gain bandwidth: 2.5MHz
Slew rate: +0.5V/μs, −20V/μs
N.B. Data is for National

Semiconductor LM3900. A similar amplifier is available from Motorola and other manufacturers are expected to "second-source" such devices. Refer to manufacturers data sheets particularly for maximum ratings. While other current-differencing amplifiers may be expected to have similar performance in the circuits to be described it is important that the ratings of particular devices are not exceeded.

Circuit description
Transistors Tr₁, Tr₄ are a current mirror with the collector current of Tr₄, approximately equal to non-

Basic amplifiers—1

Typical performance
Supply: +15V
R₁: 100kΩ
R₂: 2.2MΩ
R₃: 1MΩ
C₃: 0.1μF
Direct output voltage ≈ 7V
Voltage gain ≈ −10

while the virtual earth at the inverting input (though ≈0.6V d.c.) gives a voltage gain of −R₂/R₁. The addition of reactive components modify the gain, so that a high-frequency roll-off is readily achieved by placing a capacitor across R₃ (corner frequency 1/2πR₃C).

Capacitive coupling may be required to the load, while the reactance of C₃ ≪ R₃ at lowest frequency. Maximum resistance values of up to 10MΩ may be used, but roll-off due to stray capacitances is likely.

Circuit description—1
No direct current flows in R₁ and hence R₃, R₄ determine the d.c. operating conditions. For perfect balance between the input circuit transistors they will carry equal current and for Vₛ<ε⁺, the direct output voltage is given by Vₒ = (ε⁺V), i.e. R₃ = 2R₄ is the usual condition for maximum available voltage swing with the output biased at supply mid point. As there is no significant alternating current in R₄, it is the alternating currents in R₃ and R₁ that are equal in magnitude

wireless world circard

Set 16: c.d.as—signal processing—2

Circuit description—2
The base-emitter voltage of the input transistor is ≈ 0.55V at room temperature, falling by ≈2.5mV for every 1% rise in temperature and by less than this (typically 0.5 to 1mV) for each 1-V increase in the supply voltage. This voltage is thus sufficiently stable to be used as the reference voltage for measuring the d.c. output conditions, and the technique may be called the "nVₘₑa" biasing method. It is identical in principle to that used in the d.c. feedback pair and the "amplified-diode". If the source has an internal resistance to ground ≈R₄ then direct coupling may be used with R₅ chosen to provide the required input resistance for the circuit and R₆ determining both the voltage gain and the direct output voltage. Resistor R₅ is omitted in this mode as it is the input coupling capacitor. Direct output voltage ≈(R₅)/(R₆+1)Vₑₑ. Voltage gain ≈ −R₄/R₅. The method requires modification both for high and low gains as the direct output voltage may not be convenient. By capacitive coupling to R₅ the d.c. and gain conditions can be made independent, with direct output voltage ≈(R₅)/(R₆+1)Vₑₑ and voltage gain ≈ −R₄/R₅.
inverting input current, subtracting from inverting input current at base of $T_R_2$. The net input current to $T_R_2$ is \((I_- - I_+)^2\) and this is amplified by $T_R_4$, with $T_R_3$ forming an improved emitter follower output stage. Constant-current generators define the operating conditions while $T_R_4$ comes into action on over-driving the input to maximize the sink-current. Output depends on the difference between two positive input currents with negative feedback taken to the inverting input when the gain is to be defined. The non-inverting input is outside the feedback loop, and behaves as a forward-biased p-n junction. With resistive negative feedback applied between output and inverting input, the direct currents in the two inputs will be equalized to within the accuracy of the current mirror. If the non-inverting input current is defined by a resistor to $+V$, the direct output voltage is then a fixed fraction of $+V$. Transistor $T_R_3$ base current is $\approx 30\mu A$, allowing very low bias/signal currents, and like the voltage gain and output current capabilities is controlled over wide temperature and supply variations. An internal regulator (not shown) ensures this by providing the constant currents while also biasing the set of transistors that clamp each input to $\approx -0.3V$ on negative input swings.

Further reading
Frederiksen, T. M., Howard, W. M., Sleeth, R. S., The LM3900-A New Current-Differencing Quad of $\pm$ Input Amplifiers, National Semiconductor application note AN72.
Motorola Linear Integrated Circuit Data Book, pp.7-446, 7-453, 7-456 and 7-463; data sheets on MC3301P and MC3401P amplifiers.
National Semiconductor, Linear Integrated Circuits, pp.226-33, data sheets on LM3900.

Typical performance
Supply: $+15V$
$R_1$: 47k$\Omega$
$R_2$: 100k$\Omega$
$R_3$: 1M$\Omega$
$C_1$: 0.1$\mu F$
$C_2$: 1$\mu F$
Direct output voltage $\approx 7.5V$
Voltage gain $\approx +9.5$

Circuit description—4
The value of the feedback resistor is limited to a few megohms for several reasons (bias instability, effect of stray capacitance, noise and hum). If it is required to have a high input resistance and high voltage gain than the a.c. and d.c. feedback must be different. As shown, $R_4$ and $R_5$ constitute a potential divider for the output signal while only $R_6$ is involved in the d.c. feedback. Output voltage has a quiescent value of $+V(R_2/R_4)$. Voltage gain is $\approx -(R_2/R_1)(R_3/R_4+1)$. Where $R_3=R_1$, a convenient condition, the voltage gain simplifies to $-(R_2/R_1+1)$. However when the ratio $R_3/R_4$ is large the feedback theory demands that the limited open-loop gain be taken into account. In practice, a ratio that should set the gain to $-20$ will do so to within about 1%, while a nominal gain of $-100$ would be nearer to $-95$. Cross references
Set 5, cards 5, 8, 9, 10.
Set 7, cards 4, 10.
Set 10, cards 1, 9.
Set 16, cards 2, 5, 6, 10.

Circuit description—3
A third variation on the biasing methods available, is to take the bias resistor $R_1$ to a separate reference voltage $V_{ref}$ which can be decoupled to make the output voltage much less dependent on supply ripple. A single reference voltage (here equal to $+V/2$) may be used for a number of separate amplifiers separate control of the quiescent output conditions is by variation of $R_4$ for each amplifier while adjustment of $R_1$, $R_3$ varies all of them simultaneously. The amplifier is shown with the signal applied to the non-inverting input. No feedback is available at this input and so the impedance of the input transistor affects the input current. At room temperatures, $r_i \approx 0.026/\beta B_i$, giving a value $>3k\Omega$ for the values shown. This reduces the gain to about 3% below the simple theoretical relationship $R_1/R_3$. This biasing method is equally applicable to the inverting amplifiers.
Basic amplifiers—2

Typical performance
Supply: +15V
R1, R2: 10kΩ
R3: 220kΩ
R4: 100kΩ
C1: 1μF
C2: 1μF
Direct output voltage ∼ 7V
Voltage gain ∼ −10
(transformer secondary output)

Circuit description—1
Common-mode signals are a problem when transmission over lines has to take place in a noisy environment. By coupling the signals through a transformer such common mode signals are minimized, i.e., the anti-phase inputs of the current differing amplifier offer a further improvement. Any common-mode voltage at the transformer secondary produces equal currents at the two inputs largely cancelling each other because the gain at the two inputs is equal and opposite. R1act is inserted to achieve the correct loading on the source with R1, R2 sufficiently larger not to affect that loading. Quiescent output voltage ∼ V(R2/R1); voltage gain ∼ −R4/R1.

Circuit description—2
Where the source is inductive or is to be transformer-coupled, a variant of the "n-small" biasing method provides a simple solution. Again there is the restriction that the direct output voltage and the gain of the amplifier itself are controlled by the same resistor ratio but decoupling part of R1 to ground can make the ratio for signal frequencies a high value at d.c. if required. In the extreme case, R1 can be completely decoupled giving the full open-loop gain of the amplifier.

This coupled with the step-up turns ratio of the transformer gives a very high overall gain. As shown, the overall voltage gain is ∼ (R/R1 + 1) and the quiescent output voltage is ∼ (V/2 + 1)VA.

Because the input current required by the amplifier is very small (< 1μA) the effective input impedance remains high regardless of the gain, and high step-up ratios are possible. This yields a very sensitive microphone amplifier though the noise performance is unlikely to allow use in audio applications.

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Logic gates

Typical performance
Supply: +20V
IC: ±LM3900
R1: 150kΩ
R2: 2kΩ
Output logic 0: 150mV
Output logic 1: 19.2V
For inputs commoned, output changes state for input voltage ∼ 20% of +V.

Circuit description
Availability of two current inputs simplifies the design of basic logic gates with these amplifiers. For example, a low current at one input can hold the output in one desired state while the other input receives the sum of the currents from two or more inputs. This sum can be set to overcome the bias when only one input is high or only if all are simultaneously high, leading to OR and AND-type circuits respectively. (The amplifier is working as a high-gain comparator and can also provide a majority-gate in which any two out of three inputs are enough to provide the required output. By extension some of the simpler forms of threshold logic are possible by scaling the values of resistors to assign a different weight to their importance in decision making.)

In the first configuration, if any input is high the current driven into the non-inverting input exceeds the inverting input current and the output is driven high, i.e. an OR gate. The remaining input resistors connected to logic 0 bypass a small portion of that current (<0.5V/R for each resistor) but unless the number of inputs is large and/or the supply voltage is low, this is not a problem. Speed of response is limited to ∼ 0.5V/μs for positive swings and up to 20V/μs for negative swings though the fall in voltage is slower as logic 0 is approached. By interchanging the inverting and non-inverting inputs with no change in component values, a NOR gate is produced. This flexibility of being able to produce different logic functions from the same package is very attractive. In addition, one or more of the amplifiers can be used to provide astable, Schmitt trigger functions, etc. for obtaining the appropriate waveforms with which to drive the gates.

Component changes
+V Normal voltage range is +4 to +36V, but some devices will operate to <3V without difficulty.
R1, R2 Ratio of the resistances is chosen to ensure that with the lowest expected value of logic 1 to any one input that, the resulting current flow into the non-inverting input is sufficient to overcome
High voltage amplifiers

Circuit description
Where a high output voltage swing is required the amplifier must be fed from a separate low-voltage supply, and a suitable high-voltage transistor employed to withstand the main supply voltage. The configuration depends on whether it is the output voltage or current that is to be defined. If the former, then the feedback is taken from in shunt with the load. For an inverting-gain amplifier and the transistor in the common-emitter mode, the inverting gain of the transistor necessitates that the feedback be applied to what is normally considered as the non-inverting input. For an input of 0V d.c., the current flow in R1 is small and that in R4 is forced by the feedback to equal that in R3. If in this condition the output is desired to be +HT then

\[ HT/R4 = V/R4 \]

gives the required value of R4. Overall voltage gain is given by

\[ (-R2/R1) \]

as the circuit is effectively a "see-saw" amplifier while the input resistance is approximately R1. Resistance R4 introduces a small amount of negative feedback into the output stage and raises the amplifier quiescent voltage well into its linear region. Capacitor C1 modifies the gain/frequency characteristic to maintain stability at the higher loop gain. Output voltage swing can be up to 95% of the supply voltage if lightly loaded and the negative feedback keeps the output impedance reasonably low.

Component changes
R1, R3 These set the input resistance and the voltage gain R2>R4 to minimize loading on

Power amplifiers

Circuit description
Addition of a Darlington-connected pair of transistors increases the output current capability from 10mA to 1-5A depending on the ratings of the transistors used. One restriction is that the Vbe's of the transistors limit the output voltage to around 2.5V below the supply level allowing for the amplifier internal saturation. The additional phase shifts that may occur even in an emitter follower make external compensation desirable. For supply, input and output to be all positive, the configuration shown is adequate, where with R1=R3 and V1 above the amplifier internal Vbe. The relationship is then

\[ V_{out} = V_{be} + (R4/R1) (V1 - V_{be}) \]

as the currents in R1 and R4 have to be equal. This means that as a d.c. amplifier it is of relatively low accuracy but is quite suitable for supplying small d.c. motors under the control of a phase-locked loop. A combination of the techniques for increasing the current ratings and voltage could allow the production of high-power amplifiers. Replacing the emitter follower stages by common emitter amplifier increases the available positive output swing to within a hundred millivolts of the positive supply.

Component changes
R1 This sets the voltage gain in conjunction with R4. Because of the Vbe offset, the output voltage becomes temperature dependent particularly for V1, i.e. high gains are not compatible with good stability in this configuration. R1 100k to 10MΩ.
R6, R7 These provide forward bias for each of the inputs allowing the output to be
the output. \( R_4 \) 1\,\text{M\Omega} \text{ to 22\,\text{M\Omega}}. \)

\( R_3 \) Load resistance, dictated by user requirements. For use as an r.c.-coupled amplifier, \( R_3 \) has to carry a quiescent current greater than peak load current required.

\( R_4 \) Sets d.c. conditions at output. If \( V_{oc} = +HT \) required for \( V_{in} = 0 \), then \( +V/R_4 = +HT/R_4 \). If output quiescent to be \( +HT/2 \) as when used for amplifying a.c. input signal, then \( +V/R_4 = +HT/2R_4 \). \( R_4 \) Not critical. Raises amplifier output quiescent voltage to 1 to 3V range, i.e. into linear region. Value dependent on output quiescent current but might be 50\,\Omega to 5\,\text{k\Omega}.

\(+V\) Chosen to suit amplifier, and available supplies (+4 to +36V), \( +HT \) Dictated by load requirements.

\( R_1 \) Must have voltage rating in excess of \( +HT \) particularly if inductive loading possible.

Circuit modifications
- A non-inverting amplifier

controlled for inputs down to zero. This is best achieved for \( R_1 = R_2 \) when \( V_{oc} = V_i \) is the first-order approximation, the \( V_{be} \) effects at the two inputs cancelling.

\( R_4 \) As suggested, \( R_4 \) may equal \( R_1 \) for optimum stability but with gain restricted to \( +1 \). \( R_4 > 2.0R_1 \) except for a.c. amplifiers where some drift in quiescent conditions is acceptable.

\( R_3 \) 1\,\text{k\Omega}. Together with \( C_1 \) is part of the suggested network for maintaining stability. Limits amplifier current under load short circuit conditions providing protection for amplifier and output stage. Not adequate unless proper heat-sinking used since limit of output current depends on transistors' current-gains and is ill defined.

\( C_1, C_2 \) Control high frequency performance. \( C_1 \): 330p to 2.2nF; \( C_2 \): 5.6 to 22pF. Choose lowest values giving stability under operating conditions.

\( R_4 \) Load resistance. This may drop across the primary is too small to allow of the d.c. feedback. The two \( 10\,\text{M\Omega} \) resistors define the potential at the transistor emitter as about \( 2V_{be} \) and allows the output current to be fixed reasonably accurately, falling with rise in temperature. By decoupling the emitter and taking overall a.c. feedback from load, the output impedance can still be made low if required.

- Increased output resistance is possible by using compound output stages of various kinds that can also increase the current capability, subject to device power limitations. An f.e.t. draws no current from the amplifier ensuring that the load and emitter resistor currents change together if the current \( i_{bas} \) is made constant, either because of the high supply voltage and correspondingly high resistance \( R_4 \) or by a separate low-voltage constant-current stage.

Cross references
Set 7, cards 5, 7.
Set 16, cards 1, 2, 6.

applied with a complementary-symmetry output (Fig. M2). Not recommended for low-distortion applications, since additional bias networks to overcome crossover problems would lose the advantage of simplicity normally offered by this amplifier.

- Although the supply voltage capability is good (up to 36V) the output swing can be extended to \( \pm 70\,\text{V} \) (Fig. M3) using a bridge configuration. The simplest arrangement for a.c. signals has the \( 1\,\text{M\Omega} \) and 470\,\text{k\Omega} resistors setting the output quiescent voltages to \( \mp V/2 \) for maximum undistorted voltage swing. Replacing the two input resistors by a potentiometer and applying the signal via the pot to opposite-phase inputs gives anti-phase outputs that can be set for equal magnitude with an overall voltage gain of about 18.

Cross references
Set 7, cards 2, 4, 7, 8, 10, 11.
Set 16, card 5.
High voltage amplifiers

Circuit description
Where a high output voltage swing is required the amplifier must be fed from a separate low-voltage supply, and a suitable high-voltage transistor employed to withstand the main supply voltage. The configuration depends on whether it is the output voltage or current that is to be defined. If the former, then the feedback is taken from in shunt with the load. For an inverting-gain amplifier and the transistor in the common-emitter mode, the inverting gain of the transistor necessitates that the feedback be applied to what is normally considered as the non-inverting input. For an input of 0 V d.c., the current flow in R1 is small and that in R2 is forced by the feedback to equal that in R4. If in this condition the output is desired to be $+HT$ then $+HT/R_2 = +V/R_0$ gives the required value of $R_4$. Overall voltage gain is given by $(-R_2/R_4)$ and the circuit is effectively a “see-saw” amplifier while the input resistance is approximately $R_1$. Resistance $R_4$ introduces a small amount of negative feedback into the output stage and raises the amplifier quiescent voltage well into its linear region. Capacitor $C_1$ modifies the gain/frequency characteristic to maintain stability at the higher loop gain. Output voltage swing can be up to 95% of the supply voltage if lightly loaded and the negative feedback keeps the output impedance reasonably low.

Component changes
$R_1$, $R_3$ These set the input resistance and the voltage gain $R_2 > R_4$ to minimize loading on

Power amplifiers

Circuit description
Addition of a Darlington-connected pair of transistors increases the output current capability from 10mA to 1-5A depending on the ratings of the transistors used. One restriction is that the $V_{BE}$’s of the transistors limit the output voltage to around 2.5V below the supply level allowing for the amplifier internal saturation. The additional phase shifts that may occur even in an emitter follower make external compensation desirable. For supply, input and output to be all positive, the configuration shown is adequate, where with $R_3 = R_5$, $V_0$ varies linearly with $V_1$ provided $V_1$ is above the amplifier internal $V_{BE}$. The relationship is then $V_0 \approx V_{BE} + (R_3/R_1) (V_1 - V_{BE})$ as the currents in $R_1$ and $R_3$ have to be equal. This means that as a d.c. amplifier it is of relatively low accuracy but is quite suitable for supplying small d.c. motors under the control of a phase-locked loop. A combination of the techniques for increasing the current ratings and voltage could allow the production of high-power amplifiers. Replacing the emitter follower stages by common emitter amplifier increases the available positive output swing to within a hundred millivolts of the positive supply.

Component changes
$R_1$ This sets the voltage gain in conjunction with $R_4$. Because of the $V_{BE}$ offset, the output voltage becomes temperature dependent particularly for $V_1$ comparable to $V_{BE}$, i.e. high gains are not compatible with good stability in this configuration. $R_4$, 100k to 10MΩ.
$R_3$, $R_4$ These provide forward bias for each of the inputs allowing the output to be
the output. \( R_4 \) 1 MΩ to 22 MΩ.

- **Load resistance**, dictated by user requirements. For use as an r.c.-coupled amplifier, \( R_4 \) has to carry a quiescent current greater than peak load current required.

- **Sets d.c. conditions at output.** If \( V_{oc} = +HT \) required for \( V_{in} = 0 \), then \( +V/R_4 = +HT/R_4 \). If output quiescent to be \( +HT/2 \) as when used for amplifying a.c. input signal, then \( +V/R_4 = +HT/2R_4 \).

- **Not critical.** Raises amplifier output quiescent voltage to 1 to 3 V range, i.e. into linear region. Value dependent on output quiescent current but might be 50 Ω to 5 kΩ.

- **\(+V\)** Chosen to suit amplifier, and available supplies (+4 to +36 V).

- **\(+HT\)** Dictated by load requirements.

- **\(T_4\)**. Must have voltage rating in excess of \(+HT\) particularly if inductive loading possible.

**Circuit modifications**

- A non-inverting amplifier

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controlled for inputs down to zero. This is best achieved for \( R_4 = R_2 \) when \( V_o = V_1 \) is the first-order approximation, the \( V_{BE} \) effects at the two inputs cancelling.

- **As suggested, \( R_4 \) may equal \( R_2 \) for optimum stability but with gain restricted to \( +1 \). \( R_2 > 2.0 R_4 \) except for a.c. amplifiers where some drift in quiescent conditions is acceptable.

- **\( R_4 \) 1 kΩ.** Together with \( C_4 \) is part of the suggested network for maintaining stability. Limits amplifier current under load short circuit conditions providing protection for amplifier and output stage. Not adequate unless proper heat-sinking used since limit of output current depends on transistors' current-gains and is ill defined.

- **\( C_4 \) 330 pF to 2.2 nF.** \( C_1 \) 5.6 to 22 pF. Choose lowest values giving stability under operating conditions.

- **\( R_4 \) Load resistance.** This may drop across the primary is too small to allow of the d.c. feedback. The two 10-MΩ resistors define the potential at the transistor emitter as about \( 2V_{BE} \) and allows the output current to be fixed reasonably accurately, falling with rise in temperature. By decoupling the emitter and taking overall a.c. feedback from the load, the output impedance can still be made low if required.

- Increased output resistance is possible by using compound output stages of various kinds that can also increase the current capability, subject to device power limitations. An f.e.t. draws no current from the amplifier ensuring that the load and emitter resistor currents change together if the current \( I_{fet} \) is made constant, either because of the high supply voltage and correspondingly high resistance \( R_4 \) or by a separate low-voltage constant-current stage.

**Cross references**

Set 7, cards 5, 7.

Set 16, cards 1, 2, 6.

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applied with a complementary-symmetry output (Fig. M2). Not recommended for low-distortion applications, since additional bias networks to overcome crossover problems would lose the advantage of simplicity normally offered by this amplifier.

- Although the supply voltage capability is good (up to 36 V) the output swing can be extended to \( \approx \) 70 V (Fig. M3) using a bridge configuration. The simplest arrangement for a.c. signals has the 1 MΩ and 47 kΩ resistors setting the output quiescent voltages to \( +V/2 \) for maximum undistorted voltage swing. Replacing the two input resistors by a potentiometer and applying the signal via the pot to opposite-phase inputs gives anti-phase outputs that can be set for equal magnitude with an overall voltage gain of about 18.

**Cross references**

Set 7, cards 2, 4, 7, 8, 10, 11.

Set 16, card 5.
**Bandpass filters**

Typical performance
- \( R_1: 47k\Omega \)
- \( C: 10nF \)
- \( R_3, R_4: 1M\Omega \)
- \( R_2: 6.8M\Omega \)
- \( R_s: 100k\Omega \)
- Supply: +15V
- \( f_0: 320Hz \)
- For \( Q = 15 \)
- \( k = 0.640 \)

N.B. Onset of oscillation at \( k = 0.648 \) compared with theoretical value of 0.66.

Circuit description
Active filter techniques based on operational amplifiers are applicable to current-differencing circuits. That shown is a direct adaptation of "Circuit 1, Set 1." A Wien bridge is placed between source and output with the potential at its junction monitored by the amplifier input terminal. Resistor \( R_1 \) is a high-value resistance to minimize loading on the network. Resistor \( R_4 \) provides a variable amount of positive feedback to increase the Q of the circuit with minimal effect on the centre-frequency while \( R_s \) sets the quiescent output voltage to allow for maximum signal swing. Because \( R_s > R_1 \), \( \alpha \) would be required for stability, with \( R_4 \) taken directly to the output, and variable high-value resistors are inconvenient, the alternative is to tap \( R_4 \) onto a variable portion of the output voltage. The result is a band-pass filter with centre frequency given by \( f = 1 / 2\pi CR \) and with Q controlled by \( R_4 \). The limited gain and bandwidth capabilities of the amplifier does not allow the circuit to provide large stable Qs nor to operate successfully at frequencies \( > 10kHz \). The input impedance falls as the Q is increased because of the increased output swing (gain \( \propto \) Q to a first order). It is also true that the sensitivity of such circuits to variations in component values is proportional to the Q, i.e. for \( Q = 10 \) a 1% change in a critical resistor might.

**Notch filters**

Typical performance
- Supply: +20V
- \( R_1: 100k\Omega \)
- \( R_3, R_4: 1M\Omega \)
- \( R_s: 1k\Omega \)
- \( R_2: 47k\Omega \)
- \( C: 10nF \)
- Notch frequency: 319Hz

Achieved for \( m \approx 0.35 \)
for \( k = 0 \) and \( k = 1 \)

Circuit description
The circuit is again derived from a Wien Bridge to demonstrate the principles by which known circuits can be adapted to current-differencing amplifiers. It provides a notch or null in the response at a frequency set by the RC values \( (f = 1/2\pi RC) \) with \( R_3 \) providing a trimming action to get as true a null as possible. If \( R_4 \) is a low resistance potentiometer, then the depth of the null is unaffected as the tapping point is varied, since both ends of the potentiometer are at zero for this condition. However, the positive feedback introduced has the effect off-null of sharpening up the response so that the gain remains close to unity for frequencies close to the null. This makes the circuit useful for nulling out the fundamental (or particular harmonic) of a complex waveform with minimal effect on the other harmonics. A weakness of this particular circuit is that it relies on the matching of the inverting and non inverting input sensitivities making it prone to variation with supply/temperature changes, etc.

Component changes
- \( R, C \) The difficulty with this particular circuit is that these should have a low impedance compared with \( R_4 \) so that the latter does not load them, while not in turn being disturbed by the varying source resistance of \( R_4 \) as \( k \) is varied. Typically \( R = 100k\Omega \).
produce >10% variation in the Q itself.

Component changes
R, C The load on the output
should not fall much below
10kΩ at any frequency to
avoid loss of loop gain,
distortion, etc. R 10k to 1MΩ,
C 220p to 10μF.
R4 Used to set a convenient
level of current at amplifier
inputs; too low and loading on
Wien network disturbs Q, f0;
too high and stray capacitive
coupling disturbs response.
220k to 10MΩ.
R3 Convenient to set R3 = R4
using R4 to control feedback.
Alternatively for fixed Q
particularly if low, eliminate R4
taking R3 directly to output.
Then R3 > 1.5 R4.
R4 Not critical, but >10kΩ
and <R3.
R5 Sets quiescent output
everage for maximum swing.
Should be chosen after other
components have been
selected for desired response.
Typically 2 to 10 X R4, but for
a given value, small variations
in k and hence Q can be
accommodated without serious
bias changes.

Circuit modifications
• Multiple feedback circuits
(Fig. M1) are the equivalent of
the virtual earth circuits used
with conventional operational
amplifiers. With correct scaling
of resistors, capacitors Q > 1
is achieved simultaneously
with centre-frequency gain of
unity. Again both are strongly
dependent on component
stability at high-Q values.
• A better approach to
filter design with current-

Use if fine frequency control is
possible at the oscillator, as
may be possible when used as
part of a distortion measuring
system.
• Other passive networks have
been described which give a
good null at a specified
frequency and though normally
used as shown with a separate
feedback attenuator and buffer,
the network can also be used as
above. By deriving the
feedback directly from the
output of the amplifier variable
resistance of the attenuator is
avoided and the need for the
buffer eliminated.

Further reading
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Cross references
Set 1, cards 1, 3, 6, 7, 8, 12.
Set 5, cards 6, 10.
Set 10, card 8.
Set 16, cards 8, 9.

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Gain-controlled amplifiers

Circuit description
The non-inverting input has a diode-connected transistor as part of a current-mirror. For a given feedback resistor to the inverting input, the current flowing in the non-inverting input gives a proportional output voltage. The non-linearity of the input impedance leads to output waveform distortion if any low-resistance parallel path is used to attenuate the signal. If the path consists of a suitably-biased silicon diode then its non-linearity is comparable with that of the input stage and the input signal division between the two paths has little variation with signal amplitude, i.e. little distortion results. This remains true while the signal current is well below the bias level.

As the direct current in the diode is increased, the corresponding fall in slope-resistance by-passes more signal current reducing the gain. Resistor $R_4$ provides sufficient bias current to set the output potential to a minimum of $V_d/3$. This applies when the bias voltage is low with the direct current in $R_3$ as well as the a.c. signal current via $R_2$ are both shunted away through $R_3$, $R_2$. The gain is virtually zero in this condition. As $V_B$ approaches one diode voltage, the direct currents in $D_1$, $D_2$ and the amplifier input become comparable. The slope resistances are also comparable and the input current divides equally between the amplifier and $D_2$. For any further increase in $V_B$ the signal current flows on through $D_2$ to the amplifier with negligible attenuation by $D_2$. At this extreme the direct current in $R_3$ contributes to the amplifier bias and the output d.c. level rises to $2V_d/3$.

Low pass/high pass filters

Circuit description
The most convenient configuration for most feedback circuits with these amplifiers is the virtual earth configuration; it is not possible to use series applied feedback as with standard Sallen & Key type filters. The network is not easy to analyse component-by-component because of the interactions between them but the overall transfer function is well defined particularly if the gain in the pass-band is low, e.g. unity. Then it is sufficient if the amplifier voltage gain is >100 to have a transfer function that is very close to the theoretical value. A second-order low-pass filter results where the cut-off frequency may range from 1Hz to >100kHz. A convenient means of adjustment of the filter properties is via $C_1$, $C_2$. It is their product that fixed the cut-off frequency (for $R_1$, $R_2$, $R_4$ fixed) while their ratio determines the shape of the transfer function (i.e. Butterworth, etc.). The output impedance is low and the input impedance may be up to 1MΩ if required allowing such filters to be cascaded with negligible loading.

As shown the output voltage is defined as $(R_6/R_1+1)V_{BE}$ provided the input has a quiescent value of zero, as would be achieved by capacitive coupling from the previous stage. Since $R_1=R_2$ is a convenient value for design of the filter characteristics this restricts the output to a quiescent value of about 1.2V regardless of the supply. This may be inconvenient where larger voltage swings are desired and alternative biasing schemes may be needed (see Circuit modifications).