

Current-Conveyor Basics and Applications

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11.1.1 Introduction

The current-conveyor is a versatile analog device which is intended to be used with other circuit components to implement many analog signal processing functions. It is an analog circuit building-block in much the same way as a voltage op-amp, but it presents an alternative method of implementing analog systems which traditionally have been based on voltage op-amps. This alternative approach leads to new methods of implementing analog transfer functions, and in many cases the conveyor-based implementation offers improved performance to the voltage op-amp-based implementation in terms of accuracy, bandwidth and convenience. Circuits based on voltage op-amps are generally easy to design since the behaviour of a voltage op-amp can be approximated by a few simple design rules. This is also true for current-conveyors, and once the appropriate design rules are understood, the application engineer is able to design conveyor based circuits just as easily.

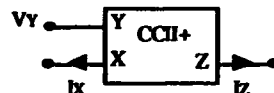
Although the current-conveyor concept has been around for a very long time, it is only in the past five years general purpose monolithic devices have been available. The emergence of these has resulted in a renewed awareness and an improved understanding of the benefits of designing analog systems using current-conveyors and it is likely that current-conveyors will become very much more common-place in the near future.

11.1.2 Background

The first generation current-conveyor or CCI was proposed by Smith and Sedra in 1968 [1] and the more versatile second generation current-conveyor, or CCII, was introduced by the same two authors in 1970 [2], as an extension of their first generation conveyor. The CCII is without doubt a much more valuable and adaptable building block and we will concentrate mostly on this device throughout this chapter. Figure 1a shows the voltage-current describing matrix for the CCII and Figure 1b the schematic normally used for the CCII with the power supply connections omitted.

$$\begin{bmatrix} I_Y \\ V_X \\ I_Z \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 \\ 1 & 0 & 0 \\ 0 & \pm 1 & 0 \end{bmatrix} \begin{bmatrix} V_Y \\ I_X \\ V_Z \end{bmatrix}$$

(a) Describing I-V matrix



(b) Schematic

Figure 1 The CCII current-conveyor.

The voltage at the low impedance input node X follows that at the high impedance input node Y , while the input current at node X is mirrored or 'conveyed' to the high impedance

output node Z. The \pm sign indicates the polarity of the output current with respect to the input current; by convention, a positive sign indicates that both the input and output currents simultaneously flow into or out of the device, thus Figure 1b illustrates a CCII+. For the first generation conveyor, or CCI, the input current at node X was reflected to input Y, that is, the two inputs had equal currents. In the case of the second generation conveyor input Y draws no current, and this second generation, or CCII formulation has proved to be much more adaptable and versatile than its first generation predecessor. Because of the combined voltage and current following properties, CCII's may be used to synthesise a number of analog circuit functions which are not so easily or accurately realisable using voltage op-amps.

At the time of the conception of the current-conveyor in 1968 the semiconductor industry was absorbed with the development of the monolithic voltage op-amp and without any clearly stated advantages there was no motivation for the development of a monolithic current-conveyor. However, since then several hundred papers have been published on the theory and applications of current-conveyors. Some of these application areas are shown in Figure 2 and circuit realisations are given throughout this chapter. As current-conveyors become more readily available and circuit designers become more familiar with the versatile of current-conveyors it is certain that further ingenious uses will be devised.

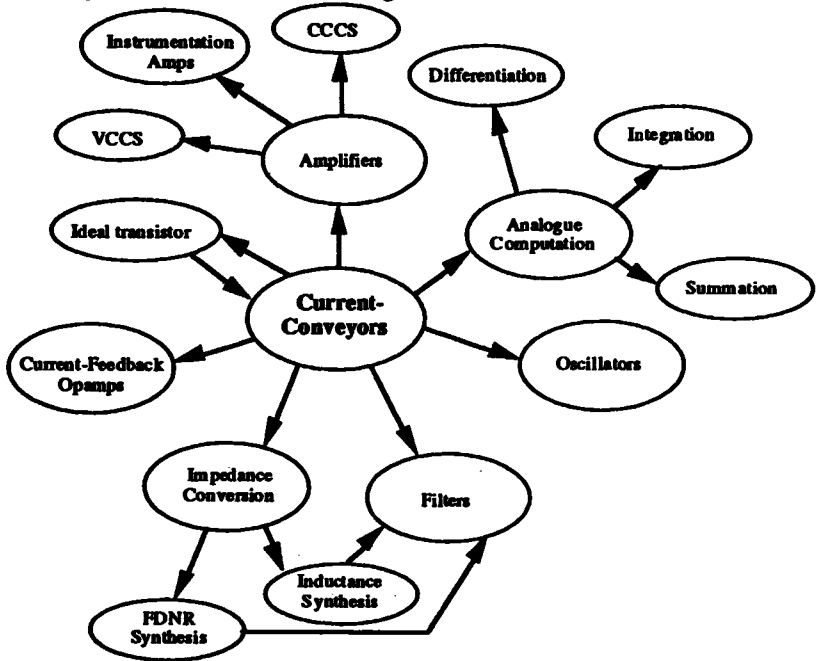


Figure 2 Current-conveyor applications.

11.1.3 Amplifiers built with current-conveyors

The current-conveyor has precise unity current gain between Z and X, rather than the high ill-defined open-loop gain of the voltage op-amp, therefore in amplifier applications the current-conveyor is generally used without any overall negative feedback. The advantage of this approach is that the traditional closed-loop gain-bandwidth conflict of negative feedback voltage op-amp circuits is avoided. Of course, the benefits of global negative feedback, for

example noise reduction, improvement in input and output impedance levels, can no longer be exploited, but the absence of overall negative feedback generally results in wider bandwidths at higher levels of gain. However, to maintain a high level of accuracy without the use of negative feedback, a high quality current-conveyor integrated circuit realisation is required.

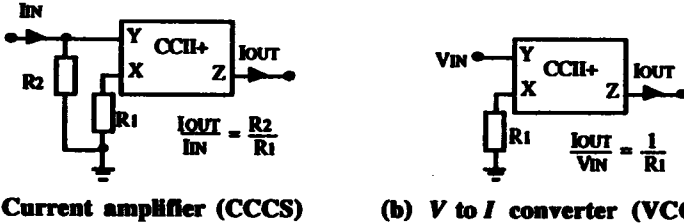


Figure 3 Current output amplifiers.

The CCII can easily be used to configure the two current output amplifiers, as shown in Figure 3. In Figure 3a, R_2 must be much less than the input impedance at node Y for current transfer accuracy, that is, $R_2 \ll R_{INY}$, and this factor will limit the maximum possible gain of the circuit. In both circuits R_1 forms a pole with parasitic capacitance to ground at node X , therefore the value of this resistor should be kept low to ensure that this parasitic pole does not dominate the frequency response.

The current-conveyor can be used to advantage in current-mode circuits and probably the most useful utility building-block being the unity current gain current amplifier, shown in Figure 4. The X node has a low input impedance and conveys the input current through to the high output impedance Z node. Hence the current gain is unity, negative for a CCII+ and positive for a CCII-. This circuit is the antithesis of a voltage-follower and is referred to as a current buffer or current-follower [3] and can be used to advantage as an interface between voltage and current-mode circuits. For example, adding a current-follower before the amplifier of Figure 3a provides the circuit with a greatly reduced input impedance and improves the integrity of the current drive into the Y node of the second current-conveyor.

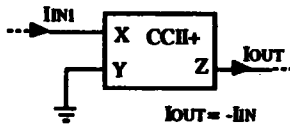


Figure 4 Current-buffer or current-follower.

Voltage amplification can be obtained by adding a voltage buffer to the VCCS of Figure 3b. The circuit of Figure 5 is such an amplifier, with a second CCII+ being used as a voltage-follower.

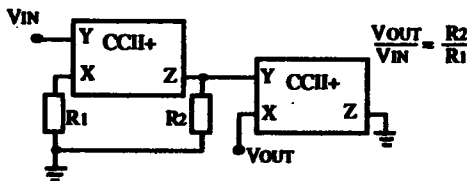


Figure 5 Voltage amplifier.

Using two CCII's, Toumazou and Lidgey [4] described a differential voltage-to-current converter as shown in Figure 6.

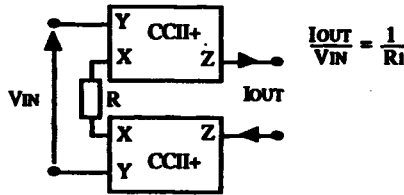


Figure 6 Differential V to I converter.

Any differential input voltage appears across resistor R_1 to generate a current which is then conveyed to the two outputs. The common-mode gain of the circuit will be zero provided that both devices are well matched and have a precise voltage-following action between their respective Y and X inputs. The circuit does not rely on any external resistor matching for high CMRR, however the CMRR will roll-off at higher frequencies due to parasitic capacitance to ground at the X nodes, which will then give rise to common-mode currents. The CMRR can be improved by increasing the differential gain, that is, by reducing the value of R. The ultimate limit on CMRR will be determined by the mismatch in open-loop bandwidth of the two conveyors.

By converting the output current back to a single-ended voltage, this differential transadmittance cell can be extended to produce a high performance instrumentation amplifier [5, 6], as shown in Figure 7. This instrumentation amplifier has a high CMRR, requires no external component matching and exhibits a bandwidth independent of gain. Experimental results of this architecture instrumentation amplifier are included later in section 11.1.10.

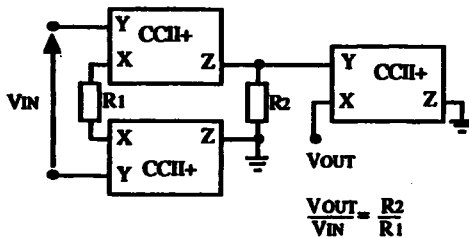


Figure 7 Instrumentation amplifier.

This instrumentation amplifier has proved to be particularly suitable for the implementation of EMG amplifiers with shutdown control [7]. Shutdown control is necessary to isolate the EMG amplifier during the period that an electrical stimulation is applied to the muscles, since this large stimulus will otherwise saturate the amplifier. The amplifier would then require a recovery period before it is able to measure the much weaker EMG muscle response. This shutdown control is most easily implemented by providing a switch control at the amplifier inputs. Conventional EMG amplifiers based on voltage op-amps require switching at a high impedance node, and this tends to cause transient spikes which themselves can saturate the amplifier. Using the instrumentation amplifier of Figure 6, the switching control can be moved to a low impedance node, so that resistor R_1 is either disconnected or connected between the two low impedance X inputs. This conveyor-based

EMG amplifier demonstrates no saturation or recovery problems, and is able to measure the EMG signal accurately, immediately after the stimulation pulse.

11.1.4 Analog computation

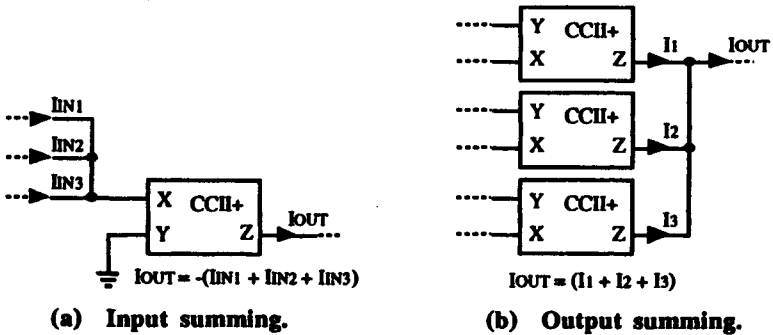


Figure 8 Current summing with current-conveyors.

Current summation is readily achieved with a current-conveyor, either at the low input impedance X node or at the high output impedance Z node, as shown in Figure 8.

In their original papers presenting the current-conveyor, Sedra & Smith [1,2] outlined a number of current-mode analog computational elements, ranging from integrators and differentiators to more complex function generators. Current differentiators and integrators are easily implemented, as shown in Figure 9. In Figure 9a, resistor *R* should be fairly small to minimise the effect of stray capacitance at node X, meanwhile in Figure 9b, *R* should be limited in size to minimise current transfer errors, that is, $R \ll R_{INy}$.

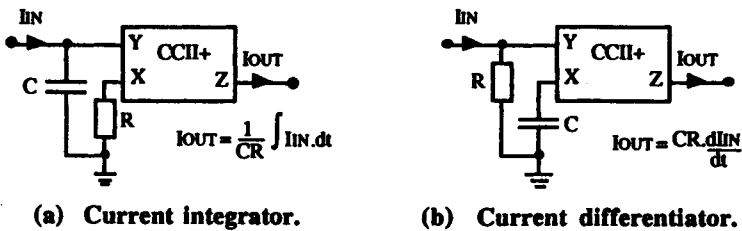


Figure 9

For voltage differentiation and integration, an additional output voltage buffer is required to supply load current, as shown in Figure 10.

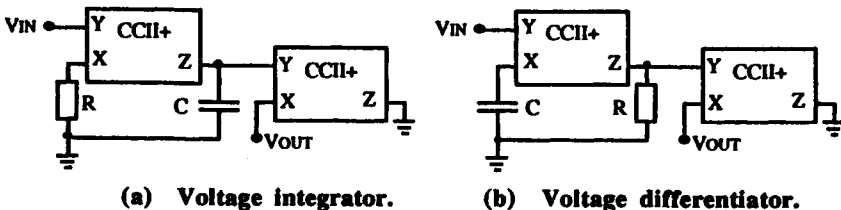


Figure 10

In many cases circuits containing voltage op-amps with capacitive feedback components must be designed very carefully since a capacitor contributes phase lag, which can result in circuit instability. However, because of the absence of global negative feedback, these conveyor-based integrator and differentiator circuits do not suffer the same potential instability problems. The circuits of Figures 9 and 10 also utilise grounded rather than floating capacitors, which is an advantage for integrated circuit realisations.

11.1.5 Impedance conversion

Much of the research in the area of current-conveyors has focused on the implementation of impedance converters. Many voltage op-amp-based impedance converters require complicated circuitry and tight component matching requirements. With its combined voltage and current following properties, the CCII is a more flexible alternative to implement these functions.

Impedance converters are used in many areas of active circuit design, for example in active filters and inductance simulation. The frequency dependent negative resistor, or FDNR, is a circuit element used in higher order filter design which is synthesised using impedance converters. Components with negative impedance are used to cancel out positive lossy components in certain circuits by generating an identical negative component and these components also are simulated using a negative impedance converter, [4, 5].

11.1.5.1 Grounded negative impedance converters (NIC)

Figure 11a shows a negative impedance converter presented by Sedra & Smith [2]. By terminating input Y with an impedance Z_T , as shown in Figure 11b, a current-controlled NIC, or negative admittance converter is obtained. Similarly, by terminating input X with an impedance of Z_T , as shown in Figure 11c and driving input Y with a voltage, a voltage controlled NIC is obtained.

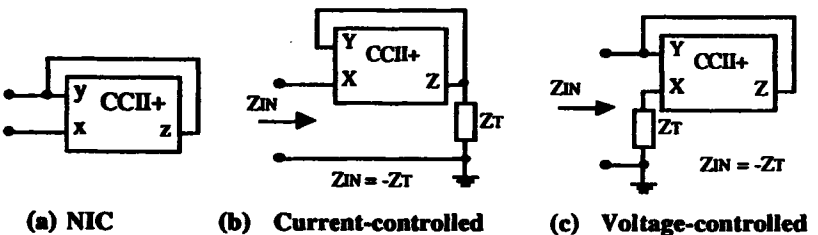


Figure 11 Grounded negative impedance converters.

11.1.5.2 Grounded generalised impedance converter (GIC)

Using two CCII's as shown in Figure 12, more general impedance conversion can be implemented, depending on the nature of the circuit components.

If both conveyors are of equal polarity, that is both CCII+ or CCII-, then a negative GIC is obtained. By terminating V_2 with a grounded impedance Z_3 , the resulting input impedance at V_1 , $Z_{IN} = -(Z_1Z_2/Z_3)$. Conversely, if the two conveyors are of opposite polarity then a positive GIC is obtained, and the addition of a grounded impedance Z_3 at V_2 will result in an impedance $Z_{IN} = (Z_1Z_2/Z_3)$ at V_1 . By choosing Z_1 and Z_2 as resistors and Z_3 as a capacitor as shown in Figure 13, a grounded inductor where $L = R_1R_2C$ can be implemented. Similarly a grounded FDNR can be obtained by choosing Z_1 and Z_2 as capacitive and Z_3 as resistive.

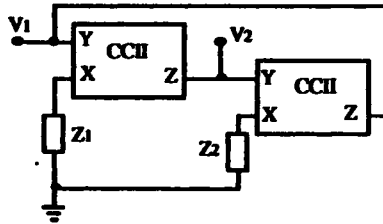


Figure 12 Generalised impedance converter (GIC).

More recently, systematic methods have been presented for the implementation of any required immittance function using current-conveyors. By extending the circuit of Figure 12, several researchers [8-11] present methods for synthesising series and parallel immittance combinations, and also implement immittance functions of any order by 'nesting' stages of the basic circuit. These complex immittance functions can then be used to implement filter circuits which are practically insensitive to active parameter variations.

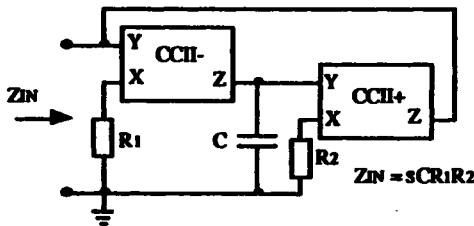


Figure 13 Simulated grounded inductance.

Himura et al [13, 14] have presented series and parallel impedance simulators which require only a single CCII and Figure 14 shows such a series impedance simulator.

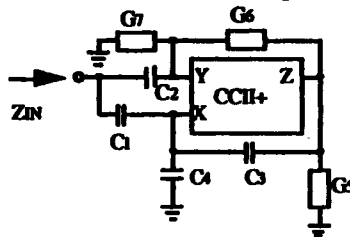


Figure 14 Series impedance simulator.

When the matching condition $C_4G_6 = 2C_3G_7$ is met, then the input impedance of the circuit becomes a series connection of inductance L_S , resistance R_S and capacitance C_S , where

$$L_S = (2C_2C_3) / [(C_1 + C_2)(G_7G_5 + G_6G_7 + G_5G_6)]$$

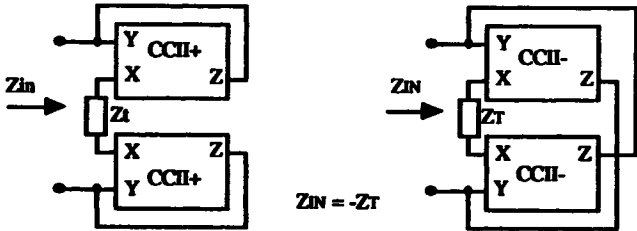
$$R_S = (G_5C_2 + G_6C_2 - G_6C_1) / [(C_1 + C_2)(G_7G_5 + G_7G_6 + G_5G_6)]$$

$$C_S = (C_1 + C_2).$$

By altering these component values, the value of R_S can be made positive, negative or zero, and thus the circuit of Figure 14 can implement either an $L-C-R$ series impedance, an $L-C$

series impedance (note that this configuration does requires tight component matching), or an $L-C$ -negative R series impedance. These series impedances are useful in designing circuits such as active ladder filters and oscillators.

11.1.5.3 Floating negative impedance converter



(a) Floating NIC using CCII+ (b) Floating NIC using CCII-

Figure 15 Floating negative impedance converters.

A floating NIC can be implemented by combining the grounded NIC of Figure 11c with the differential transadmittance cell of Figure 6. The two conveyors must have the same polarity; the choice of either CCII+ or CCII- will determine the nature of the feedback connections, as shown in Figure 15.

11.1.5.4 Floating generalised impedance converter

By combining two differential transadmittance cells, a floating GIC can be implemented as shown in Figure 16.

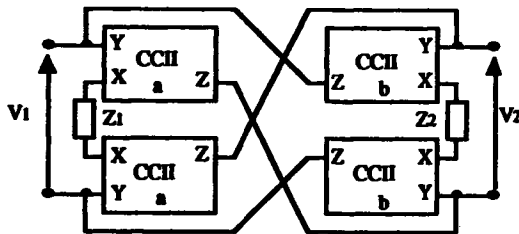


Figure 16 Floating GIC.

If all the conveyors are of equal polarity, that is, all CCII+ or all CCII-, then a positive GIC is obtained. By connecting an impedance Z_3 across V_2 , the resulting impedance at V_1 , is given by $Z_{IN} = (Z_1 Z_2 / Z_3)$. If the two conveyors marked 'a' are of equal polarity while those marked 'b' are of the opposite polarity, then a floating negative GIC is obtained. Alternatively, a negative GIC can be implemented by reversing one set of feedback connections in the circuit of Figure 16, as shown in Figure 17. This negative GIC implementation has the advantage that all the conveyors are of the same polarity. The circuit would therefore be especially suitable for monolithic fabrication, since four devices of the same type on the same silicon wafer would have a high degree of matching.

Floating inductors and FDNR implementations based on only two CCII have also been reported [8-10]. These two-conveyor implementations generally require an increased number of external components, many of which have to be tightly matched, and in some

cases they also suffer from high component sensitivities. One circuit with low component sensitivities is the two-conveyor implementation reported by Senani [15], and implemented practically by Wilson [16], which is shown in Figure 18.

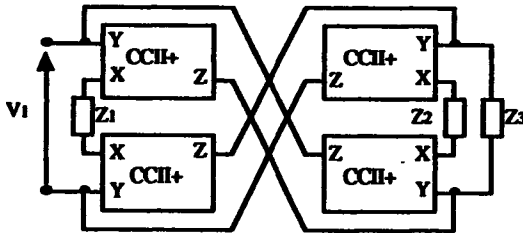


Figure 17 Floating negative GIC.

With either Z_2 or Z_4 capacitive and all other elements resistive, a floating inductor can be synthesised, where $L = R_1C_2R_3R_5/R_4$.

With two capacitive components (Z_1 and Z_3 or Z_1 and Z_5 or Z_3 and Z_5), and all other elements resistive, a floating FDNR is obtained, that is

$$Z_{IN} = -\omega^2 C_1 C_3 R_5 / R_2 R_4.$$

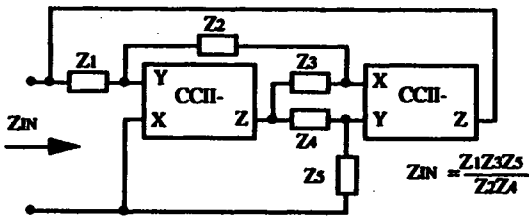


Figure 18 Two-conveyor floating GIC.

Since the circuit of Figure 18 requires only two CCII's it is more suitable for discrete implementation than the four-conveyor circuits of Figures 16 and 17. However in Figure 18, an increased number of external components are required.

11.1.6 Filters

Current-conveyors have received considerable attention from designers concerned with the implementation of active filters, since voltage op-amp-based filters are bandwidth limited by the voltage op-amps themselves. Many of the earlier CCII-based filter designs required floating capacitors, had poor sensitivity values and were only able to realise one particular filter type [8-10]. The emergence of GICs described in the previous section allowed for a wider range of filters to be implemented based on simulated inductors, for example, elliptic ladder filters as well as various types of Butterworth filters.

A 3rd-order low-pass Butterworth filter reported by Nandi [17] which is particularly attractive, due to its use of grounded capacitors and equal-valued components, is shown in Figure 19.

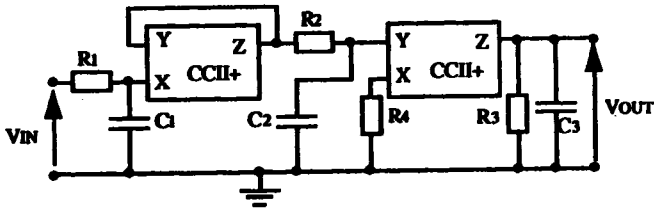


Figure 19 Butterworth lowpass filter.

For the circuit of Figure 19

$$\frac{V_{OUT}}{V_{IN}} = \frac{R_3 / R_4}{a_3 s^3 + a_2 s^2 + a_1 s + 1}$$

where $a_3 = R_1 C_1 R_2 C_2 R_3 C_3$, $a_2 = R_1 C_1 R_2 C_2 + R_2 C_2 R_3 C_3 + R_1 R_3 C_3 (C_1 - C_2)$,

and $a_1 = R_1 (C_1 - C_2) + R_2 C_2 + R_3 C_3$.

With equal-valued components, that is $R_1 = R_2 = R_3 = R$ and $C_1 = C_2 = C_3 = C$, a 3rd-order transfer function is obtained

$$\frac{V_{OUT}}{V_{IN}} = \frac{1}{R^3 C^3 s^3 + 2R^2 C^2 s^2 + 2RCs + 1}$$

The resulting filter is practically insensitive to active parameters, that is insensitive to the current tracking errors of the current-conveyors.

Toumazou and Lidgley [18] proposed and demonstrated the first universal CCII biquad filter, based on an earlier circuit reported by Nawrocki and Klein [19]. The implementation used only grounded capacitors and resistors, but required seven CCII's. Depending on how the CCII's are realised in practice, this relatively large number of current-conveyors used to achieve a 2nd-order filter response can be reduced significantly using hardware reduction techniques in the full filter implementation. A similar performance universal active filter requiring three CCII's was proposed by Singh and Senani [20], however this circuit needs additional voltage buffers to provide a high input impedance and low output impedance. More recently, Lui and Tsao [21] have presented a family of biquad filters with high input impedance which require only a single CCII and four external passive components. Figure 20 shows the general circuit implementation.

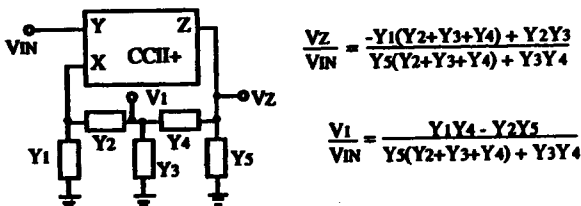


Figure 20 General biquad filter architecture.

The nature of the admittances Y_1 to Y_5 determines the filter response. For example, with $Y_1=0$, $Y_2=1/R_2$, $Y_3=sC_3$, $Y_4=sC_4$ and $Y_5=1/R_5$, a low-pass characteristic is obtained at V_1 and a band-pass characteristic is obtained at V_2 , with

$$\omega_0 = (C_3C_4R_2R_5)^{-1/2} \text{ and } Q = \frac{1}{C_3+C_4} (R_5C_3C_4/R_2)^{1/2}$$

High-pass, band-pass, all-pass and notch filters can also be implemented with selection of the appropriate admittances Y_1 to Y_5 , and all filters exhibit a low sensitivity to the CCII current tracking error, and practically zero sensitivity to the voltage tracking error.

The filter designs discussed so far have been voltage-processing circuits; that is, they are an alternative method of implementing voltage op-amp-based filter designs. However, because of its current-following capability, the CCII is also able to implement current-processing filter circuits [22-26]. Similar to their voltage-mode biquad filters, Lui et al [23] have presented a family of current-mode biquads, based on a single CCII with five external passive components. The general filter configuration is shown in Figure 21.

Again selection of the appropriate admittances Y_1 to Y_5 gives the required filter response. For example, with $Y_1=1/R_1$, $Y_2=\infty$, $Y_3=1/R_3$, $Y_4=sC_4$ and $Y_5=sC_5$ results in a low-pass filter with $\omega_0 = (R_1R_3C_4C_5)^{-1/2}$ and $\omega_0/Q = (C_4+C_5)/R_1C_4C_5$. Depending on the value of the admittances, an input current buffer may be necessary to prevent excessive loading of the current source input.

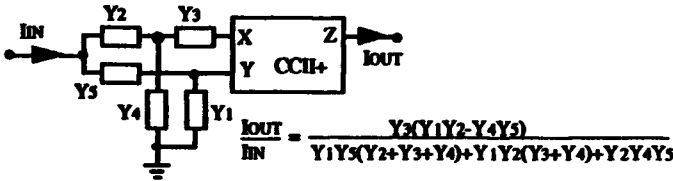


Figure 21 General current-mode biquad filter.

A universal active current filter based on a single CCII has been presented by Chang [25], as shown in Figure 22.

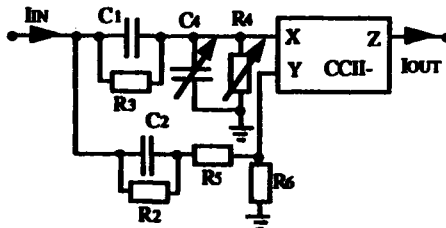


Figure 22 Universal current-mode filter.

With R_2 open circuit, this architecture can implement notch, all-pass and low-pass filter characteristics by altering the values of C_4 and R_4 . With resistor R_2 present, band-pass and high-pass characteristics can be obtained, again by tuning the values of C_4 and R_4 . The resulting filters have low passive sensitivities, and are almost insensitive to the current

tracking error of the conveyor. An alternative method of deriving current-filtering circuits from already existing voltage filters has been presented by Roberts and Sedra [26].

Many voltage-mode *active-RC* filters use op-amps configured as unity gain buffers to maximise the useable bandwidth of the circuit, and current-mode filters can be derived from these voltage buffer-based filters using the adjoint circuit method. To find the adjoint circuit of any network, a replica of the original circuit must be constructed first. All elements in this replica circuit are then replaced by their adjoint elements; passive components such as resistors and capacitors remain the same, while unity gain voltage buffers are replaced by unity gain current buffers, or current-conveyors. Voltage input and output signals are replaced by current output and input signals, so the two adjoint circuits are exactly equivalent in the voltage-mode and current-mode. The new current-mode circuit will have the same component sensitivities as the voltage-mode original, so low sensitivity current-mode circuits can be obtained from low sensitivity voltage-mode circuits which have already been developed. The Sallen-Key active biquad (SAB) of Figure 23a would be therefore transformed to the CCII-based circuit of Figure 23b by the adjoint circuit method.

It should be noted that in practice an input DC current path must be provided for all active devices employed as for all analog circuits and in the circuits described in the Impedance Converter and Filter sections above any such additional DC bias components have not been included in the circuit diagrams.

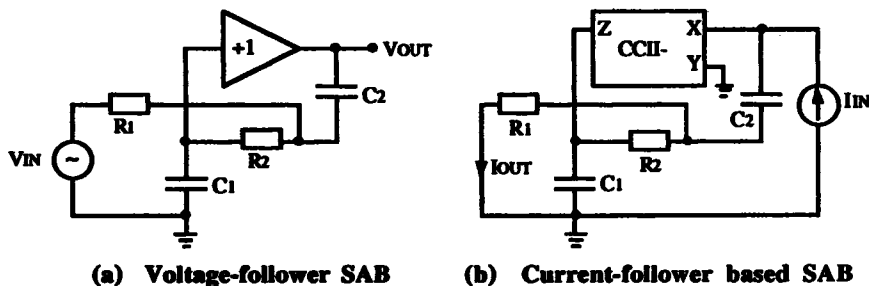


Figure 23 Sallen-Key active biquad filters.

11.1.7 The ideal transistor and the current-conveyor

So far a transistor-level realisation of the CCII has not been discussed. The current-voltage transfer relationship for the CCII+ is given by

$$V_X = V_Y, I_Y = 0 \text{ and } I_Z = I_X.$$

These equations show that there is a simple voltage-following action between input node *Y* and output node *X* and that there is a simple current-following action between input node *X* and output node *Z*. Also these characteristic equations tell us that the impedance relationship for the current-conveyor are

$$Z_{IN} = \infty, Z_X = 0 \text{ and } Z_{OUT} = \infty.$$

Figure 24 shows a schematic representation of a CCII- built with a single BJT and on reflection it is clear that the current-conveyor is effectively an ideal transistor, with infinite β and infinite g_m .

Driving into the base of a BJT gives almost unity voltage gain from input base to output emitter, with high input impedance and low output impedance, and driving into the emitter of a BJT gives almost unity current-gain from emitter input to collector output, with

low input impedance and high output impedance. Drawing the comparison further the high input impedance Y node corresponds to the base (or gate) of a transistor, the low input impedance X node corresponds to the emitter (or source) of a transistor, and the high output impedance Z node corresponds to the collector (or drain) of a transistor. Clearly one transistor cannot function alone as a complete current-conveyor as an unbiassed single transistor can only handle unipolar signals and the high accuracy unity voltage and unity current gain required for a high performance current-conveyor cannot be obtained. However the generic relationship between the current-conveyor is valid and it does provide valuable insight into the development and operation of monolithic current-conveyors described in the next section.

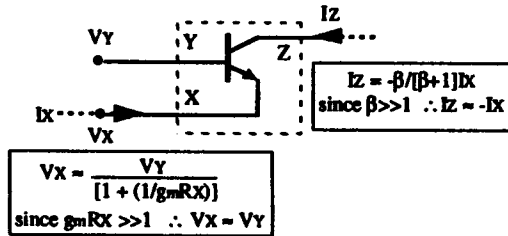


Figure 24 Single BJT CCII-.

11.1.8 Supply-current sensing, current-feedback op-amps and the CCII

Over the last decade current-converter research and current-conveyor research has proceeded in parallel with little cross reference, despite the fact that the two areas are very closely related. The difference between the two research areas is really only one of semantics and not substance, as the current-conveyor, or part current-conveyor, can be identified easily in nearly all the current-converter work. Toumazou and Lidgey worked extensively on current-converter realisations based on the technique of power supply-current sensing on voltage-mode op-amps, described in [3, 5], to achieve current output performance. The circuit performances obtained were extremely encouraging, with useful practical circuits reported on current-followers and follower-based amplifiers, high speed precision rectifiers, instrumentation amplifiers, and current amplifiers [5, 27].

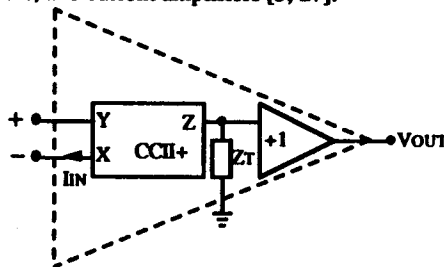


Figure 25 Current-feedback op-amp structure.

Monolithic current-feedback op-amps have relatively recently become commercially available as high-speed alternatives to voltage op-amps [28]. Examination of the internal structure of the current-feedback op-amp is revealing as the device is essentially a CCII+ with the internal high impedance Z node connected to an output voltage-follower, as shown in Figure 25. Any current flowing at the low impedance inverting input is conveyed to the gain node (Z_T), and the resulting voltage is buffered to the output. Z_T is thus the open-

loop transimpedance gain of the current-feedback op-amp, which in practice is equal to the parallel combination of the CCII+ output impedance, the voltage buffer input impedance and any compensation capacitance at the gain or Z_T node.

Generally the gain node is not connected to an external pin and so the Z node of the CCII+ cannot be accessed. However, simply removing the output buffer from a current-feedback op-amp is all that is required to obtain a high performance CCII+.

11.1.9 CCII01 current-conveyor

The first commercially available current-conveyor based on the current-feedback op-amp is the CCII01 from LTP Electronics [29], shown in Figure 26. The circuit is fully symmetrical, with Q_1 to Q_4 comprising the input stage, Y input being high impedance and X input/output being low impedance. A voltage applied to the Y input is followed by the X input. The collector currents of Q_2 and Q_3 are reflected and recombined through current-mirrors CM_1 and CM_2 to the high output impedance Z node. Hence any input into the X node is conveyed with unity current gain through to the Z output node.

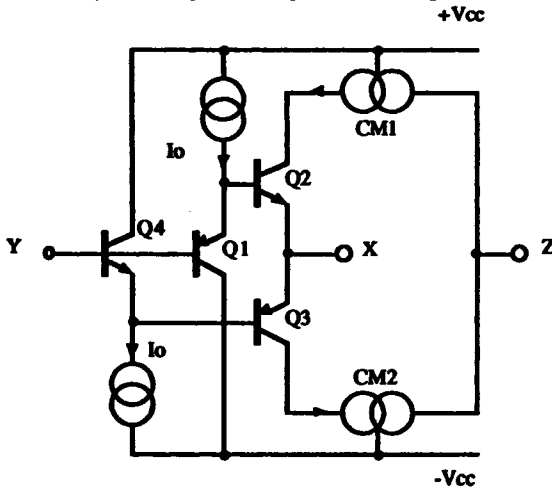


Figure 26 Internal schematic of the CCII01 current-conveyor.

The CCII01 is built on a high speed dielectric isolation fully complementary bipolar process and supplied as dual device in an 8-pin DIL package. The device features an equivalent slew-rate of 2000 V/ μ s and a 100 MHz bandwidth. The equivalent open-loop gain is 80 dB and the CMRR performance is better than 53 dB at 1 MHz. The maximum output current from the device is ± 10 mA and it operates from ± 5 V to ± 15 V supplies.

11.1.10 IA and PFWR application examples of the CCII01

11.1.10.1 CCII01 instrumentation amplifier

A particularly useful and elegant application of the current-conveyor is the current-mode instrumentation amplifier (IA) employing the differential V to I converter of Figure 6. The circuit schematic and measured CMRR performance is shown below in Figure 27. The key advantage of this architecture IA is the high CMRR achievable at high frequencies without any requirement for an accurately matched resistor network.

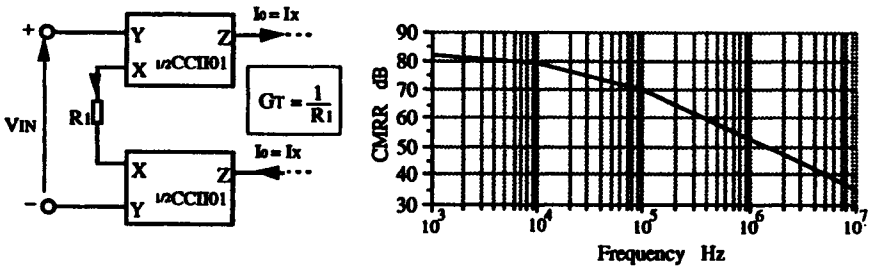


Figure 27 CCII01 IA and measured CMRR performance.

The output signal is taken from either Z node. Conversion of the output Z node current into a voltage is readily achieved using the conventional inverting transresistance stage shown in Figure 28. Alternatively the output signal current can be taken from the two conveyor Z nodes and fed into a differential transresistance amplifier to achieve double the gain.

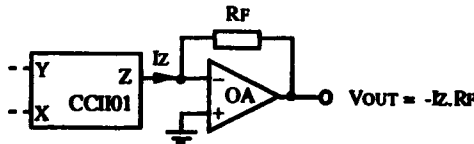


Figure 28 Transresistance conversion of I_z .

11.1.10.2 CCII01 precision full-wave rectifier

Another equally elegant use of the current-conveyor is for high speed precision rectification. The classical problem with conventional precision rectifiers based on diodes and op-amps is that during the non-conduction/conduction transition of the diodes the op-amps have to recover with a finite small-signal dV/dt resulting in significant distortion during the zero-crossing of the input signal. Using high slew-rate op-amps does not solve this fundamental drawback since it is a small-signal transient problem [30]. Conventional rectifiers are thus limited to a frequency performance well below the gain-bandwidth product of the amplifier.

Improvements [30] have been made to rectifier high frequency performance by the use of current-mode techniques primarily based upon employing the power supply rails of the op-amp as a current rectification path. However, a problem encountered with such schemes is that signal levels need to be significantly higher than the supply bias to guarantee precision rectification at high frequency and so again loss of precision occurs at signal zero-crossings. Even using high speed current-feedback amplifiers the performance is still limited to some tens of kiloHertz [30], which is significantly below the f_T of the current-feedback amplifiers used.

A full-wave precision rectifier can be configured easily using two CCII, as shown in Figure 29. Both the CCII form a differential voltage to current converter such that during the positive input cycle, the output currents of value V_{IN}/R flow out of the Z-node of CCII(a) and into the Z-node of CCII(b), thus making only D_4 and D_2 active. Because D_2 is active, the current from the Z-node of CCII(a) flows into the output resistor R making $V_{OUT} = V_{IN}$. During the negative input cycle, only D_3 and D_1 are active thus the output current of

CCII(b) is driven into R making $V_{OUT} = V_{IN}$. Clearly the magnitude of gain is R_2/R_1 and this can be increased from unity by making $R_2 > R_1$.

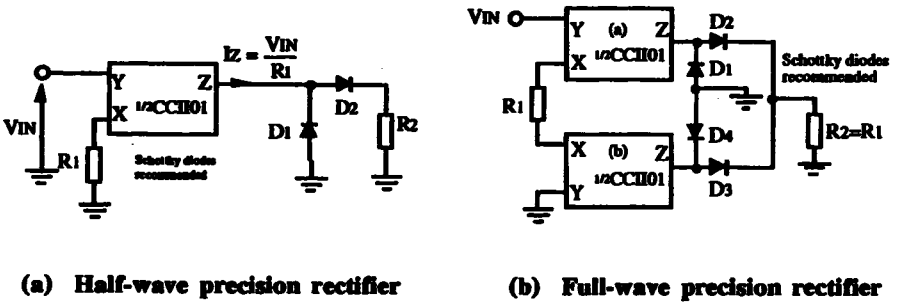


Figure 29 Current-conveyor precision rectifiers.

The circuit was built with $100\ \Omega$ resistors and Schottky diodes and Figure 30 shows typical performance for the half-wave precision rectifier shown in Figure 29a at two operating frequencies, 100 kHz and 1 MHz.

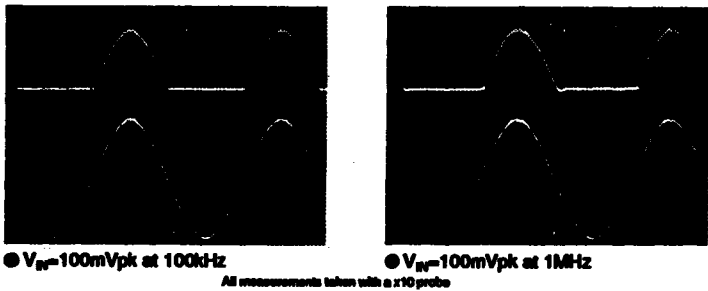


Figure 30 Precision half-wave rectifier.

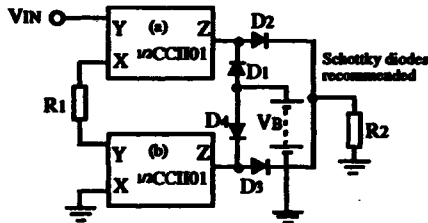


Figure 31 Improved current-conveyor PFWR

The performance is good but with the CCII01 exhibiting an f_T of approximately 100 MHz, one would naturally expect the circuit to work close to the f_T of the device. However, this is not the case since at very low signal levels all the diodes are off, and as a result the differential voltage to current converter is transformed into a high gain differential voltage

amplifier. Although the CCI01 exhibits a very high slew-rate, in the region of 2000 V/ μ s, it is the small signal dV/dt that limits performance at zero-crossings. The solution to this problem is to modify the circuit by offsetting the output of the conveyors by biasing all the diodes appropriately. The new scheme [31] is shown in Figure 31.

In the circuit, the voltage at the anode of diodes, D_1 and D_4 , is biased by a low impedance voltage source, V_B , at approximately 0.6 V, allowing 0.3 V forward bias for each Schottky diode. Thus when D_1 and D_4 conduct the voltage at the subsequent Z-terminal is approximately +0.3 V and the output voltage is zero. This condition ensures that the load impedance presented to the Z-terminal is kept low at all times, especially as the diode pairs D_1/D_3 and D_2/D_4 swap conduction roles. The net effect is that the current-conveyor outputs do not go into slew at each zero-crossing. Effectively the improved scheme provides a class AB voltage bias so that all the diodes are on the edge of conduction during the zero input condition. Note that the battery, V_B , may be replaced by an appropriately biased diode voltage source. The improved rectifier operates well up to a frequency of 30 MHz, as shown in [31]. Quite an acceptable rectifier performance is achieved and clearly is substantially less distorted than the non-diode biased circuit which begins to show significantly poor performance at about 4 MHz. Beyond 30 MHz, the rectifier integrity of the final circuit does start to deteriorate. This is due to the roll-off in V_X/V_Y with frequency. However the measured results are encouraging and represent a significant improvement.

11.1.11 The future of the current-conveyor

Although 20 years old, the current-conveyor is a circuit concept whose time has now come. There is no doubt that the current-conveyor is both versatile and convenient for many applications. The applications described demonstrate the ease with which complex analog functions can be realised using current-conveyors. Often fewer parts are needed than the op-amp counter part, due to the very attractive combined voltage/current capability of the current-conveyor. There are many applications where op-amp realisations are significantly less practical than the equivalent current-conveyor realisation. Further more, the current-conveyor can provide high performance operation, particularly in terms of speed and bandwidth, due to the inherent local feedback of the follower-based structure of the device.

Commercially available current-conveyors are a welcomed and valuable addition for the application engineer, complementing the ubiquitous op-amp. The semiconductor industry has now recognised the potential of current-mode techniques, attested by the successful current-feedback op-amp which is a close relative to the current-conveyor. All the ingredients for a high performance current-conveyor realisation are already accessible to the semiconductor industry, so it is only a matter of time before the current-conveyor takes its place as a standard part in the application designers tool-kit.

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