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January 1996, Rev.D FN7152

Current Mode Multiplier

élantec

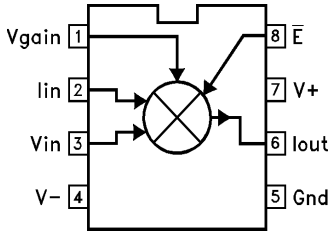
The EL2082 is a general purpose variable gain control building block, built using an advanced proprietary complementary bipolar process. It is a two-quandrant multiplier, so that zero or negative control voltages do not allow signal feedthrough and very high attenuation is possible. The EL2082 works in current mode rather than voltage mode, so that the input impedance is low and the output impedance is high. This allows very wide bandwidth for both large and small signals.

The I_{IN} pin replicates the voltage present on the V_{IN} pin; therefore, the V_{IN} pin can be used to reject common-mode noise and establish an input ground reference. The gain control input is calibrated to 1mA/mA signal gain for 1V of control voltage. The disable pin ($\overline{E}$) is TTL-compatible, and the output current can comply with a wide range of output voltages.

Because current signals rather than voltages are employed, multiple inputs can be summed and many outputs wire-or'ed or mixed.

The EL2082 operates from a wide range of supplies and is available in standard 8-pin plastic DIP or 8-pin SO.

EL2082
(8-PIN SO, PDIP)
TOP VIEW



Features

- Flexible inputs and outputs, all ground referred
- 150MHz large and small-signal bandwidth
- 46dB of calibrated gain control range
- 70dB isolation in disable mode @ 10MHz
- 0.15% diff gain and 0.05° diff phase performance at NTSC using application circuit
- Operates on ±5V to ±15V power supplies
- Outputs may be paralleled to function as a multiplexer

Applications

- Level adjust for video signals
- Video faders and mixers
- Signal routing multiplexers
- Variable active filters
- Video monitor contrast control
- AGC
- Receiver IF gain control
- Modulation/demodulation
- General “cold” front-panel control of AC signals

Ordering Information

PART NUMBER	TEMP. RANGE	PACKAGE	PKG. NO.
EL2082CN	0°C to +75°C	8-Pin PDIP	MDP0031
EL2082CS	0°C to +75°C	8-Pin SO	MDP0027

CAUTION: These devices are sensitive to electrostatic discharge; follow proper IC Handling Procedures.
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Absolute Maximum Ratings (T_A = 25°C)

V_S	Voltage between V_S+ and V_S- . . . . . +33V	P_D	Maximum Power Dissipation . . . . . See Curves
V_{IN}, V_{OUT}	Voltage . . . . . ±V_S	T_A	Operating Temperature Range . . . . . 0°C to +75°C
V_E, V_{GAIN}	Input Voltage . . . . . -1 to +7V	T_J	Operating Junction Temperature . . . . . 150°C
I_{IN}	Input Current . . . . . ±5mA	T_{ST}	Storage Temperature . . . . . -65°C to +150°C

CAUTION: Stresses above those listed in "Absolute Maximum Ratings" may cause permanent damage to the device. This is a stress only rating and operation of the device at these or any other conditions above those indicated in the operational sections of this specification is not implied.

IMPORTANT NOTE: All parameters having Min/Max specifications are guaranteed. Typical values are for information purposes only. Unless otherwise noted, all tests are at the specified temperature and are pulsed tests, therefore: $T_J = T_C = T_A$

DC Electrical Specifications

V_S = ±15V, V_G = 1V, V_E = 0.8V, V_{OUT} = 0, V_{IN} = 0, I_{IN} = 0

PARAMETER	DESCRIPTION	TEMP	MIN	TYP	MAX	UNITS
V_{IO}	Input Offset Voltage	Full	-20		20	mV
I_{OO}	Output Offset Current	Full	-100		100	µA
R_{INI}	I_{IN} Input Impedance; I_{IN} = 0, 0.35mA	Full	75	95	115	Ω
V_{CMRR}	Voltage Common-Mode Rejection Ratio, V_{IN} = -10V, +10V	Full	45	55		dB
I_{CMRR}	Offset Current Common-Mode Rejection Ratio, V_{IN} = -10V, +10V	Full		0.5	5	µA/V
V_{PSRR}	Offset Voltage Power Supply Rejection Ratio, V_S = ±5V to ±15V	Full	60	80		dB
I_{PSRR}	Offset Current Power Supply Rejection Ratio, V_S = ±5V to ±15V	Full		1	10	µA/V
I_{BVIN}	V_{IN} Bias Current	Full	-10		10	µA
R_{INV}	V_{IN} Input Impedance; V_{IN} = -10V, +10V	Full	0.5	1.0		MΩ
Nlini	Signal Nonlinearity; I_{IN} = -0.7mA, -0.35mA, 0mA, +0.35mA, +0.7mA	Full		0.10	0.4	%
R_{OUT}	Output Impedance V_{OUT} = -10V, +10V	Full	0.25	0.5		MΩ
V_{OUT}	Output Swing; V_{GAIN} = 2V, I_{IN} ±2 mA, R_L = 4.0k	Full	-11		+11	V
V_{IOG}	V_{OS}, Gain Control, Extrapolated from V_{GAIN} = 0.1V, 1V	Full	-15		15	mV
A_I	Current Gain, I_{IN} ±350µA	Full	0.9	1.0	1.1	mA/mA
Nling	Nonlinearity of Gain Control, V_{GAIN} = 0.1V, 0.5V, 1V	Full		2	5	%
I_{SO}	Input Isolation with V_{GAIN} = -0.1V	Full	-80	-96		dB
V_{INH}	$\overline{E}$ Logic High Level	Full	2.0			V
V_{INL}	$\overline{E}$ Logic Low Level	Full			0.8	V
I_{LH}	Input Current of $\overline{E}$, V_E = 5V	Full	-50		50	µA
I_{LL}	Input Current of $\overline{E}$, V_E = 0	Full	-50		50	µA
I_{ODIS}	I_{OUT}, Disabled $\overline{E}$ = 2.0V	Full			±10	µA
I_S	Supply Current	Full		13	16	mA

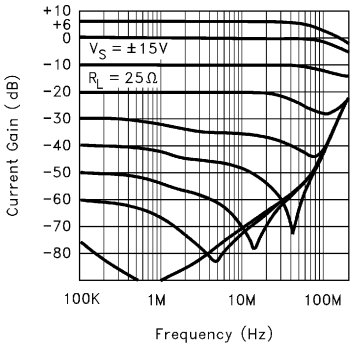
AC Electrical Specifications

$R_L = 25\Omega$, $C_L = 4pF$, $C_{IN} = 2pF$, $T_A = 25°C$, $V_G = 1V$, $V_S = \pm15V$

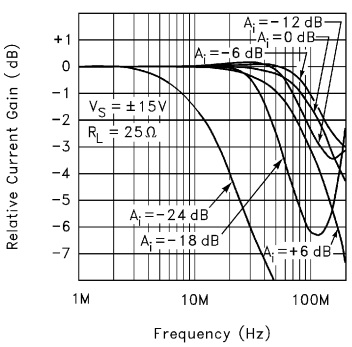
PARAMETER	DESCRIPTION	MIN	TYP	MAX	UNITS
BW1	Current Mode Bandwidth -3dB		150		MHz
BW2	±0.1dB		30		MHz
BWp	Power, $I_{IN} = 1mA_{P-P}$		150		MHz
BWg	Gain Control Bandwidth		20		MHz
SRG	Gain Control Slew Rate V_G from 0.2V to 2V		12		(mA/mA)/µs
T_{REC}	Recovery Time from $V_G < 0$		250		ns
T_{EN}	Enable Time from $\overline{E}$ Pin		200		ns
T_{DIS}	Disable Time from $\overline{E}$ Pin		30		ns
D_G	Differential Gain, NTSC with I_{IN} = -0.35mA to +0.35mA		0.25		%
D_P	Differential Phase, NTSC with I_{IN} = -0.35mA to +0.35mA		0.05		Degree

Typical Performance Curves

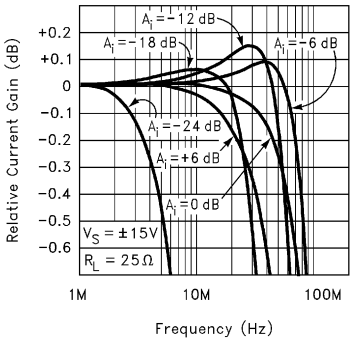
Current Gain vs Frequency for Different Gains



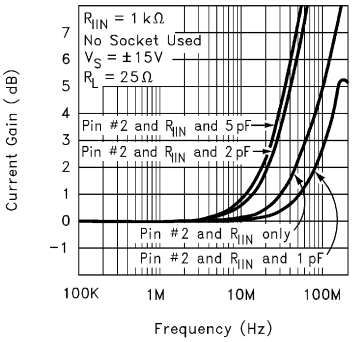
Current Gain vs Frequency



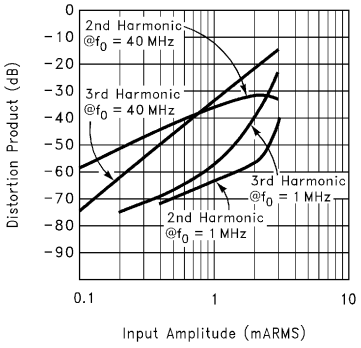
Current Gain Flatness



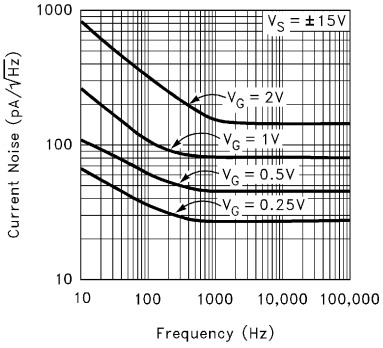
Frequency Response in Voltage Input Mode



Harmonic Distortion vs Input Amplitude

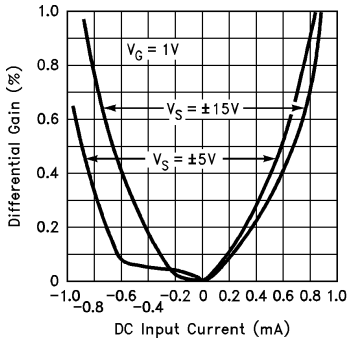


Output Current Noise vs Frequency

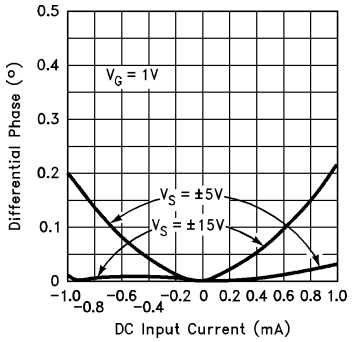


Typical Performance Curves (Continued)

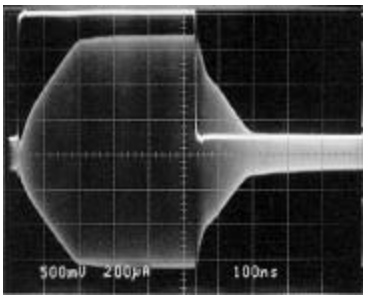
Differential Gain Error vs DC Offset Current



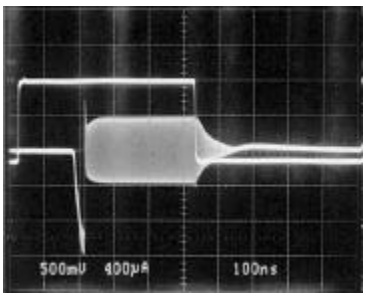
Differential Phase Error vs DC Offset Current



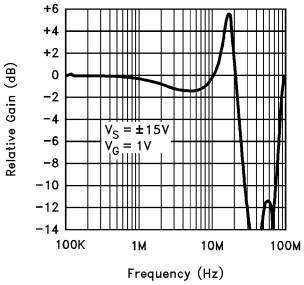
Gain Pin Transient Response



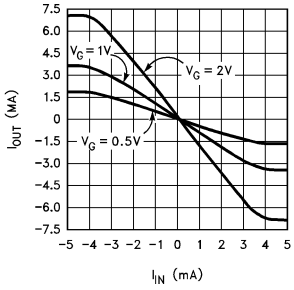
Gain Control Recovery From Vg = -0.1V



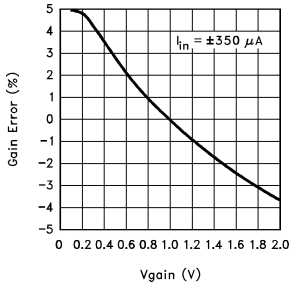
Gain Control Pin Frequency Response



I_{OUT} vs I_{IN}

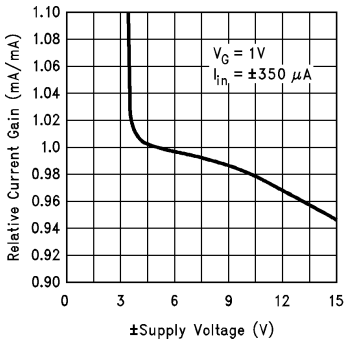


Normalized Gain Error vs V_{GAIN} Voltage

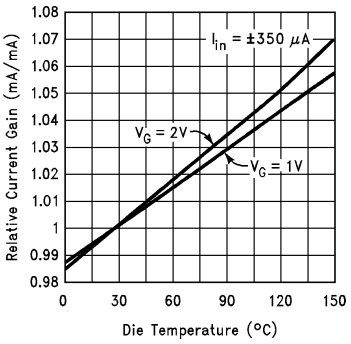


Typical Performance Curves (Continued)

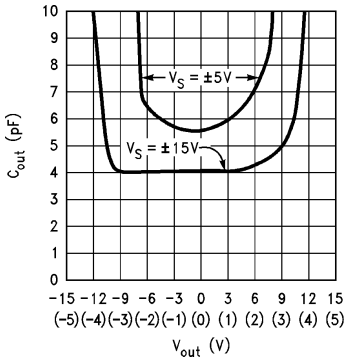
Current Gain vs Supply Voltage



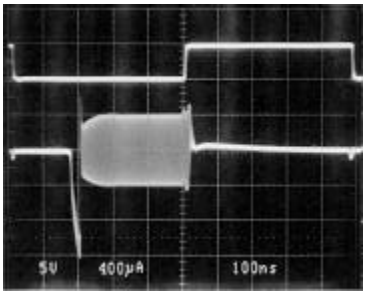
Current Gain vs Temperature



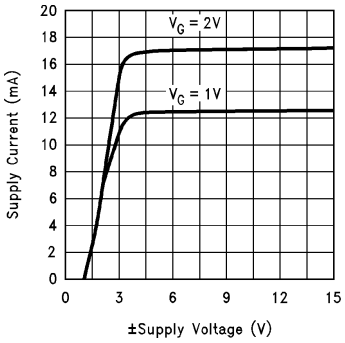
Output Capacitance vs Output Voltage



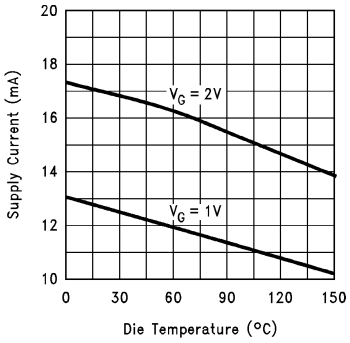
Enable Pin Response



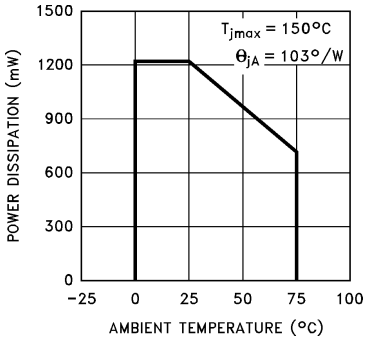
Supply Current vs Supply Voltage



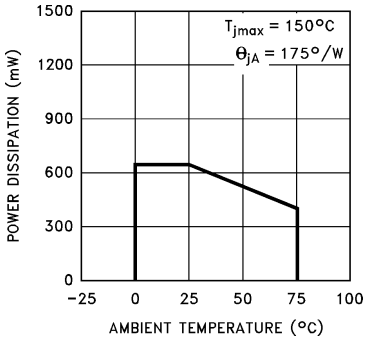
Supply Current vs Die Temperature



8-Pin Plastic DIP Maximum Power Dissipation vs Ambient Temperature



8-Pin SO Maximum Power Dissipation vs Ambient Temperature



Applications Information

The EL2082 is best thought of as a current-conveyor with variable current gain. A current input to the I_{IN} pin will be replicated as a current driven out the I_{OUT} pin, with a gain controlled by V_{GAIN}. Thus, an input of 1mA will produce an output current of 1mA for V_{GAIN} = 1V. An input of 1mA will produce an output of 2mA for V_{GAIN} = 2V. The useable V_{GAIN} range is zero to +2V. A negative level on V_{GAIN}, even only -20mV, will yield very high signal attenuation.

The EL2082 in Conjunction with Op-Amps

This resistor-load circuit shows a simple method of converting voltage signals to currents and vice versa:

$$GAIN = \frac{V_{GAIN}}{1V}\left(\frac{R_L}{R_{IN} + 95\Omega}\right)\left(\frac{R_F + R_G}{R_G}\right)$$

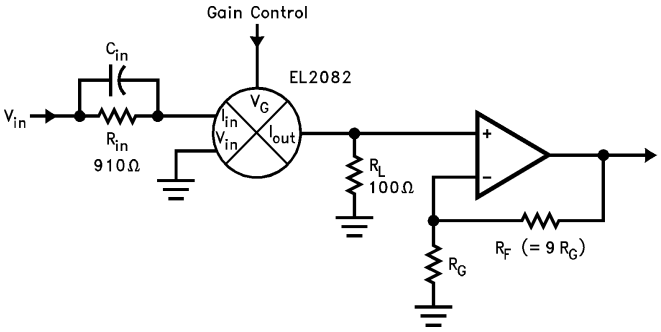


FIGURE 1. EL2082 + OP-AMP

R_{IN} would typically be 1kΩ for video level inputs, or 10kΩ for ±10V instrumentation signals. The higher the value of R_{IN} (the lower the input current), the lower the distortion levels of the EL2082 will be. An approximate expression of the nonlinearity of the EL2082 is:

$$Nonlinearity(\%) = 0.3*I_{IN}(mA)^2$$

Optimum input current level is a tradeoff between distortion and signal-to-noise-ratio. The distortion and input range do not change appreciably with V_{GAIN} levels; distortion is set by input currents alone.

The output current could be terminated with a 1kΩ load resistor to achieve a nominal voltage gain of 1 at the EL2082, but the I_{OUT}, load, and stray capacitances would limit bandwidth greatly. The lowest practical total capacitance at I_{OUT} is about 12pF, and this gives a 13MHz bandwidth with a 1kΩ load. In the above example a 100Ω load is used for an upper limit of 130MHz. The operational amplifier gives a gain of +10 to bring the overall gain to unity. Wider bandwidth yet can be had by installing C_{IN}. This is a very small capacitor, typically 1pF-2pF, and it bolsters the gain above 100MHz. Here is a table of results for this circuit used with various amplifiers:

OPERATIONAL AMPLIFIER	POWER SUPPLIES	R_F	R_G	C_{IN}	-3dB BANDWITH	0.1dB BANDWIDTH	PEAKING
EL2020	±5V	620	68	_	34MHz	5.6MHz	0
EL2020	±15V	620	68	_	40MHz	7.4MHz	0
EL2130	±5V	620	68	_	73MHz	11MHz	1.0dB
EL2030	±15V	620	68	_	93MHz	12MHz	1.3dB
EL2090	±15V	240	27	_	60MHz	10MHz	0.5dB
EL2120	±5V	220	24	_	57MHz	10MHz	0.4dB
EL2120	±15V	220	24	_	65MHz	11MHz	0.3dB
EL2070	±5V	200	22	2pF	150MHz	30MHz	0.4dB
EL2071	±5V	1.5k	240	2pF	200MHz	30MHz	0
EL2075	±5V	620	68	2pF	270MHz	30MHz	1.5dB

Maximum bandwidth is maintained over a gain range of +6 to -16dB; bandwidth drops at lower gains. If wider gain range with full bandwidth is required, two or more EL2082s can be cascaded with the I_{OUT} of one directly driving the I_{IN} of the next.

The EL2082 can also be used with an I ⟶ V operational circuit:

$$GAIN = -\frac{V_{GAIN}}{1V}\left(\frac{R_F}{R_{IN} + 95\Omega}\right)$$

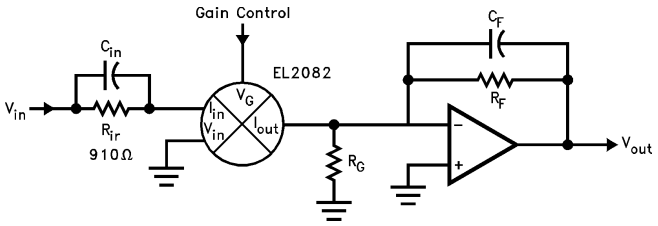


FIGURE 2. INVERTING EL2082 + OP-AMP

The circuit above gives a negative gain. The main concern of this connection involves the total I_{OUT} and stray capacitances at the amplifier's input. When using traditional op-amps, the pole caused by these capacitances can make the amplifier less stable and even cause oscillations in amplifiers whose gain-bandwidth is greater than 5MHz. A typical cure is to add a capacitor C_F in the 2pF-10pF range. This will reduce overall bandwidth, so a capacitor C_{IN} can be added to regain frequency response. The ratio C_F/C_{IN} is made equal to R_{IN}/R_F.

Current-feedback amplifiers eliminate this difficulty. Because their -input is a very low impedance, capacitance at the summing point of an inverting operational circuit is far less troublesome. Here is a table of results of various current-feedback circuits used in the inverting circuit:

OPERATIONAL AMPLIFIER	POWER SUPPLIES	R_F	R_{IN}	R_G	-3dB BANDWIDTH	0.1dB BANDWIDTH	PEAKING
EL2020	±5V	1k	910	_	29MHz	4.3MHz	0
EL2020	±15V	1k	910	_	34MHz	5.3MHz	0
EL2130	±5V	1k	910	_	61MHz	9.7MHz	0
EL2030	±15V	1k	910	_	82MHz	12.3MHz	0
EL2171 (Note 1)	±5V	2k	1.8k	1k	114MHz	11MHz	1.2dB

NOTE:

1. With the EL2171, the EL2082 had ±15V supplies and the EL2171 required a 150Ω output load.

The EL2120 and EL2090 are suitable in this circuit but they are compensated for 300Ω feedback resistors. R_{IN} would have to be reduced greatly to obtain unity gain and the increased signal currents would cause the EL2082 to display much increased distortion. They could be used if the input resistor were maintained at 910Ω and R_F reduced for a -1/3 gain, or if R_F = 1k and an overall bandwidth of 25MHz were acceptable.

The EL2082 can also be used within an op-amp's feedback loop:

$$GAIN = -\frac{1V}{V_{GAIN}}\left(\frac{R_{F9} + 95\Omega}{R_{IN}}\right)$$

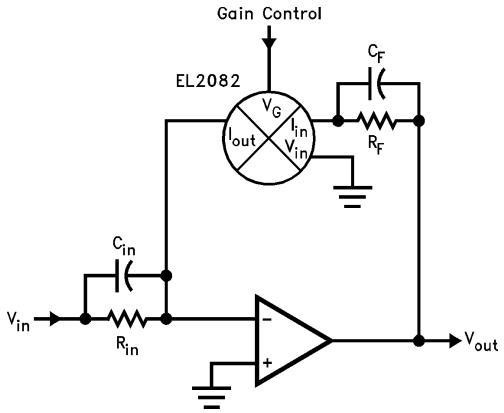


FIGURE 3. EL2082 IN FEEDBACK INVERTING GAIN

With voltage-mode op-amps, the same concern about capacitance at the summing node exists, so C_F and C_{IN} should be used. As before, current-feedback amplifiers tend to solve the problem. However, in this circuit the inherent

phase lag of the EL2082 detracts from the phase margin of the op-amp, and some overall bandwidth reduction may result. The EL2082 appears as a 3.0ns delay, well past 100MHz. Thus, for a 20MHz loop bandwidth, the EL2082 will subtract 20MHz x 3.0ns x 360 degrees = 21.6 degrees. The loop path should have at least 55 degrees of phase margin for low ringing in this connection. Loop bandwidth is always reduced by the ratio $R_{IN}/(R_{IN} + R_F)$ with voltage mode op-amps.

Current-feedback op-amps again solve the summing-junction capacitance problem in this connection. The loop bandwidth here becomes a matter of transimpedance over frequency and its phase characteristics. Unfortunately, this is generally poorly documented in amplifier data sheets. A rule of thumb is that the transimpedance falls to the value of the recommended feedback resistor at a frequency of $F_{-3dB}/4$ to $F_{-3dB}/2$, where F_{-3dB} is the unity-gain closed-loop bandwidth of the amplifier. The phase margin of the op-amp is usually close to 90° at this frequency.

In general, R_F is initially the recommended value for the particular amplifier and is then empirically adjusted for amplifier stability at maximum V_{GAIN}, then R_{IN} is set for the overall circuit gain required. Sometimes a very small C_F can be used to improve loop stability, but it often must be in series with another resistor of value around $R_F/2$.

A virtue of placing the EL2082 in feedback is that the input-referred noise will drop as gain increases. This is ideal for level controls that are used to set the output to a constant level for a variety of inputs as well as AGC loops. Furthermore, the EL2082 has a relatively constant input signal amplitude for a variety of input levels, and its distortion will be relatively constant and controllable by setting R_F. Note that placing the EL2082 in the feedback path causes the circuit bandwidth to vary inversely with gain.

The next circuit shows use of the EL2082 in the feedback path of a non-inverting op-amp:

$$GAIN = \frac{1V}{V_G}\left(\frac{R_F + 95\Omega}{R_G}\right)$$

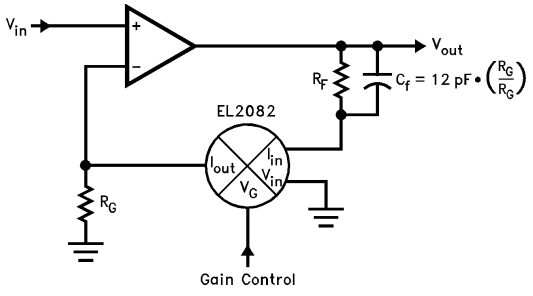


FIGURE 4. EL2082 IN FEEDBACK NON-INVERTING GAIN

This example has the same virtues with regards to noise and distortion as the preceding circuit; and its bandwidth shrinks with increasing gain as well. The typical 12pF sum of EL2082 output capacitance in parallel with stray capacitance necessitates the inclusion of C_F to prevent a feedback pole. Because of this 12pF capacitance at the op-amp -input, current-feedback op-amps will generally not be useable. As before, the loop bandwidth and phase margin must accommodate the extra phase lag of the EL2082.

The V_{IN} pin can be used instead of the I_{IN} pin so:

$$G_M = \frac{I_{OUT}}{V_{IN}} = \frac{-V_G}{1V}\left(\frac{1}{R_G + 95\Omega}\right)$$

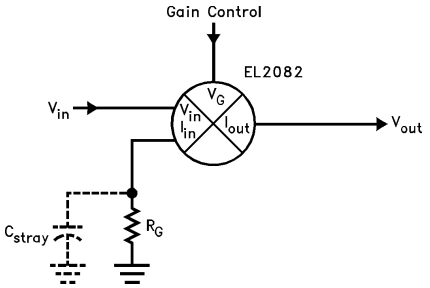


FIGURE 5. THE V_{IN} PIN USED AS SIGNAL INPUT

This connection is useful when a high input impedance is required. There are a few caveats when using the V_{IN} pin. The first is that V_{IN} has a 250V/µs slew rate limitation. The second is that the inevitable C_{STRAY} across R_G causes a gain zero and gain INCREASES above the $1/(2\pi C_{STRAY} R_G)$ frequency and can peak as much as 20dB with large C_{STRAY}. A graph of gain vs. frequency for several C_{STRAYS} is included in the typical performance curves. In general, if wide bandwidth and frequency flatness is desired, the I_{IN} pin should be used.

The V_{IN} pin does make an excellent ground reference pin, for instance when low-frequency noise is to be rejected. The next schematic shows the EL2082 V_{IN} pin rejecting possible 60Hz hum induced on an R_F input cable:

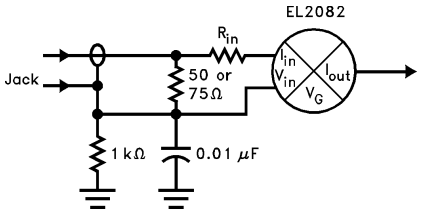


FIGURE 6. USING THE V_{IN} PIN AS A GROUND REFERENCE TO REJECT HUM AND NOISE

This example shows V_{IN} rejecting low-frequency field-induced noise but not adding peaking since the 0.01µF bypass capacitor shunts high-frequency signals to local ground.

Reactive Couplings with the EL2082

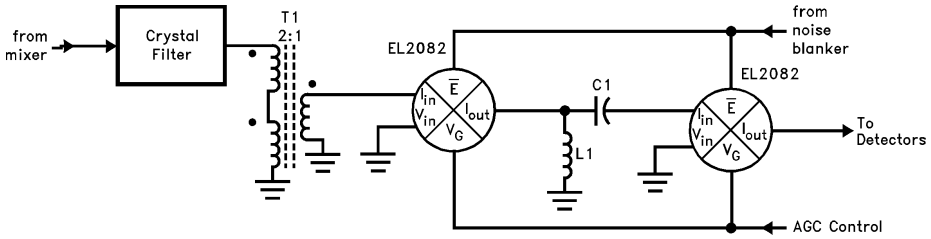


FIGURE 7. EXAMPLE REACTIVE COUPLINGS WITH EL2082'S

The above sketch is an excerpt of a receiver IF amplifier showing methods of connecting the EL2082 to reactive networks. The I_{IN} pin of the EL2082 looks like 95Ω well past 100MHz, and the output looks like a simple current-source in parallel with about 5pF. There is no particular problem with any resistance or reactance connected to I_{IN} or I_{OUT}. The mixer output is generally sent to a crystal filter, which required a few hundred ohm terminating impedance. The impedance of the I_{IN} pin of the first EL2082 is transformed to about 400Ω by the 2:1 transformer T1. The two EL2082s are used as variable-gain IF amplifiers, with small gains offered by each. The output of the first EL2082 is coupled to the second by the resonant matching network L1-C1. For a Q of 5, Xc1 = x11 = 5 x 95Ω, approximately. The impedance seen at the first EL2082's I_{OUT} will be about Q_2 x 95Ω, or 2.5k, and by impedance transformation alone the first gain cell delivers 28dB of gain at V_G = 1V. More gain cells can be used for a wider range of (calibrated) AGC compliance.

The $\overline{E}$ input can be used as a high-speed noise blanker gate.

Linearized Fader/Gain Control

The following circuit is an example of placing two EL2082s in the feedback network of an op-amp to significantly reduce their distortions:

$$V_{OUT} = K \bullet V_A + (1-K) \bullet V_B$$
where $0 \le K \le 1$

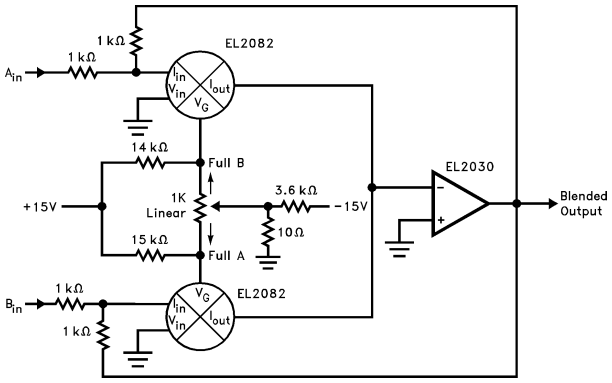
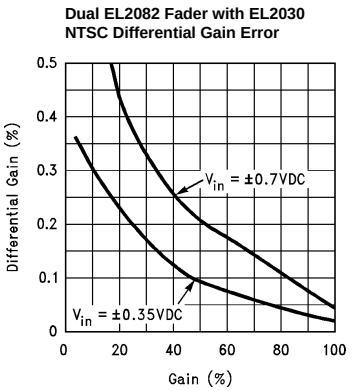
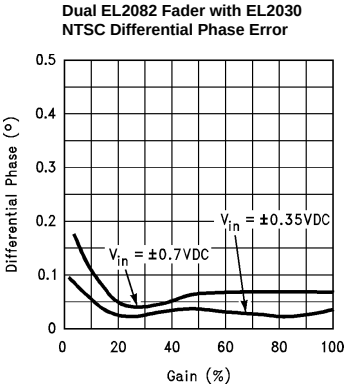


FIGURE 8. LINEARIZED GAIN CONTROL/FADER





The circuit sums two inputs A and B, such that the sum of their respective path gains is unity, as controlled by the potentiometer. When the potentiometer's wiper is fully down, the slightly negative voltage at the V_G of the B-side EL2082 cuts off the B signal to better than 70dB attenuation at 3.58MHz. The A-side EL2082 is at unity gain, so the only (error) signal presented to the op-amp's -input is the same (error) signal at the I_{IN} of the A-side EL2082. The circuit thus outputs -AIN. Since the error signal required by the op-amp is very small, even at video frequencies, the current through the A-side EL2082 is small and distortion is minimized.

At 50% potentiometer setting, equal error output signals flow from the EL2082s, since the op-amp still requires little net -input current. The EL2082s essentially buck each other to establish an output, and 50% gain occurs for both the A and B inputs. The EL2082s now contribute distortion, but less than in previous connections. The op-amp sees a constant 1k feedback resistor regardless of potentiometer setting, so frequency response is stable for all gain settings.

A single-input gain control is implemented by simply grounding B_{IN}.

Distortion can be improved by increasing the input resistors to lower signal currents. This will lower the overall gain accordingly, but will not affect bandwidth, which is dependent upon the feedback resistors. Reducing the signal input amplitude is an analogous tactic, but the noise floor will effectively rise.

Another strategy to reduce distortion in video systems is to use DC restoration circuitry, such as the EL2090 ahead of the fader inputs to reduce the range of signals to be dealt with; the -0.7V to +0.7V possible range of inputs (due to capacitor coupling) would be changed to a stabilized -0.35V to +0.35V span.

The EL2020, EL2030, and EL2120 (at reduced bandwidth since it is compensated for 300Ω feedback resistors) all give the same video performance at NTSC operation.

Variable Filters

This circuit is the familiar state-variable configuration, similar to the bi-quad:

$$F_O = \frac{V_G}{1V}\left(\frac{1}{2\pi(R + 95\Omega)C}\right)$$

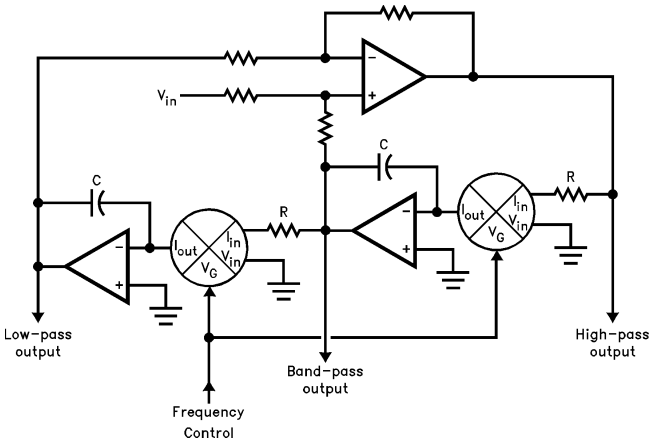


FIGURE 9. VOLTAGE TUNEABLE BI-QUAD FILTER

Frequency-setting resistors R are each effectively adjusted in value by an EL2082 to effect voltage-variable tuning. Two gain controls yields a linear frequency adjustment; using one gives a square-root-of-control voltage tuning. The EL2082s could be placed in series with the integrator capacitors instead to yield a tuning proportional to $1/V_G$.

The next circuit is one of a new class of “CCII” filters that use the current-conveyor element. Basic information is available in the April 1991, volume 38, number 4 edition of the IEEE Transactions on Circuits and Systems journal, pages 456 through 461 of the article “The Single CCII Biquads with High-Input Impedance”, by Shen-Iuan Liu and Hen-Wai Tsao.

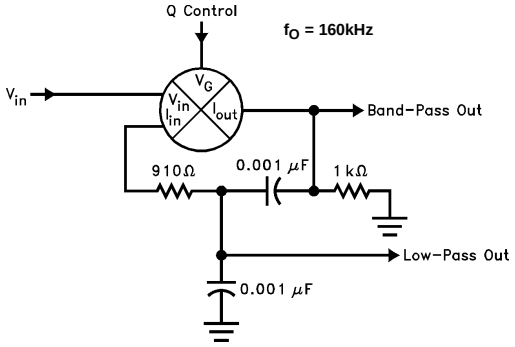


FIGURE 10. “CCII” CLASS FILTER

This interesting filter uses the current output of the EL2082 to generate a bandpass voltage output and the intermediate node provides a second-order low-pass filter output. Both outputs should be buffered so as not to warp characteristics,

although the V_{IN} of the next EL2082 can be driven directly in the case of cascaded filters. The V_{GAIN} input acts as a Q and peaking adjust point around the nominal 1V value. The resistor at I_{OUT} could serve as the frequency trim, and Q trimmed subsequently with V_{GAIN}.

Negative Components

The following circuit converts a component or two-terminal network to a variable and even negative replica of that impedance:

$$Z_{IN} = \frac{(Z + 95\Omega)}{(1 + -V_G / 1V)}$$

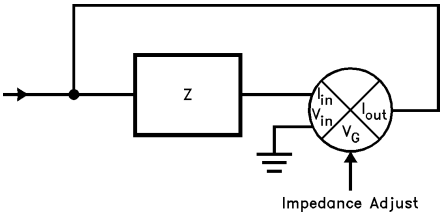


FIGURE 11. VARIABLE OR NEGATIVE IMPEDANCE CONVERTER

A negative impedance is simply an impedance whose current flows reverse to the normal sense. In the above circuit, the current through Z is replicated by the EL2082 and inverted (I_{OUT} flows inverted to the sense of I_{IN} in the EL2082) and summed back to the input. When $V_G = 0$ or $V_G < 0$, the input impedance is simply Z+95Ω. When $V_G = 1V$, the negative of the current through Z is summed with the input and the input impedance is “infinite”. When $V_G = 2V$, twice the negative of the current through Z is summed with the input resulting in an input impedance of -Z-95Ω.

Thus variable capacitors can be simulated by substituting the capacitor as Z. “Negative” capacitors result for $V_G > 1V$, and capacitance needs to be present in parallel with the input to prevent oscillations. Inductors or complicated networks also work for Z, but a net negative impedance will result in oscillations.

EL2082 Macromodel

This macromodel has been designed to work with PSPICE (copywritten by the Microsim Corporation). E500 buffers in the V_{IN} voltage and presents it to the R_{INI} resistor to emulate the I_{IN} pin. E501 supplies the non-linearity of the current channel and replicates the I_{IN} current to a ground referenced voltage. R500 and C500 provide the bandwidth limitation on the current signal. E502 supplies the V_{GAIN} non-linearity and drives the L501/R501/C501 to shape the gain control frequency response. E503 does the actual gain-control multiplication, and drives delay line T500 to better simulate the actual phase characteristics of the part G500 creates the current output, and R_{OUT} with C_{OUT} provide proper output parasitics.

The model is good at frequency and linearity estimates around V_G = 1V and nominal temperatures, but has several limitations:

- The V_G channel does not give zero gain for $V_G < 0$; the output gain reverses - don't use $V_G < 0$
- The V_G channel is not slew limited
- Frequency response does not vary with supply voltage
- The V_{IN} channel is not slew limited
- Noise is not modeled
- Temperature effects are not modeled
- CMRR and PSRR are not modeled
- Frequency response does not vary with V_G

Unfortunately, the polynomial expressions and two-input multiplication may not be available on every simulator.

Results have been confirmed by laboratory results in many situations with this macromodel, within its capabilities.

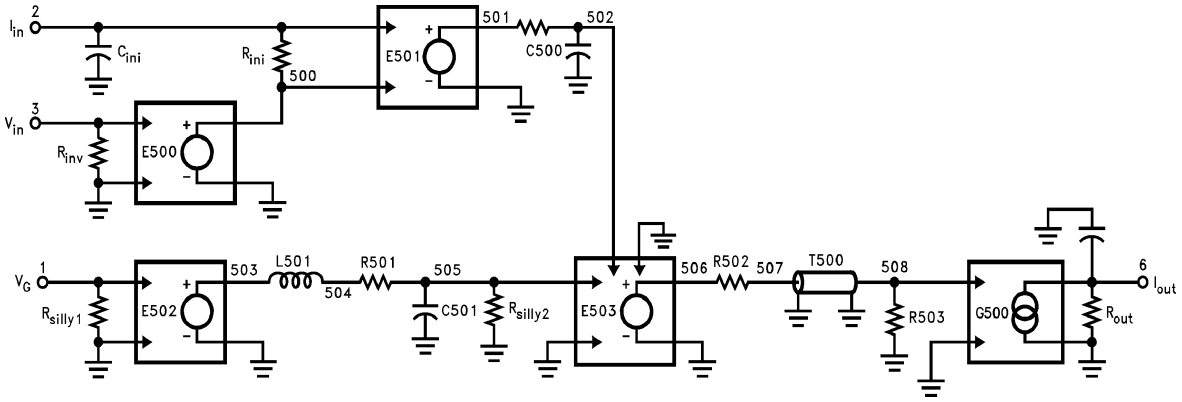


FIGURE 12. SCHEMATIC OF EL2082 MACROMODEL

EL2082 Macromodel (Continued)

```
*:                          Vgain
*                           |    Iin
*                           |    |    Vin
*                           |    |    |              Iout
*                           |    |    |              |
.SUBCKT EL2082macro (1    2    3    4    5    6    7    8)
***
*** I-to-I gain cell macromodel ***
***
******
Cini 2 0 2P
C500 502 0 0.9845P
C501 505 0 1000P
Cout 6 0 5P
******
L501 503 504 0.1U
******
Rsilly1 1 0 1E9
Rsilly2 505 0 1E9
Rini 2 500 95
Rinv 3 0 2Meg
Rout 6 0 1Meg
R500 501 502 1000
R501 504 505 5
R502 506 507 50
R503 508 0 50
******
E500 500 0 3 0 1
E501 501 0 POLY(1) (2,500) 0 2 0 -.8
E502 503 0 POLY(1) (1,0) 0 1.05 -.05
E503 506 0 POLY(2) (505,0) (502,0) 0 0 0 0 1
G500 6 0 508 0 -0.0105
T500 508 0 507 0 Z050 TD1.95N
******
.ENDS
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